Communications Symposium '83

Satellite and Broadband Technologies

Scientific Atlanta

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Overview of Commercial Satellite Communications

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The period from Arthur Clarke's 1945 prediction of geostationary satellite communications covering the entire planet till satellites were stationed over the Atlantic, Pacific and Indian Oceans was less than twenty-five years. In the following fifteen-year period, satellite communications has affected all of us. Most international calls are carried by satellite. Much of the television that we watch has been relayed, processed, or distributed by satellite. In fact, many of us who live in cabled cities can choose among fifty channels or more for viewing on a particular night, largely brought to us by satellite. Some morning papers, radio programs, and basic weather data appear coast-to-coast rapidly because of the satellite. Even while we sleep, computers in an increasing number of business offices are talking to each other by satellite. Virtually the whole world--from the busiest urban center to the most remote island--can be interconnected by satellite communications networks capable of providing economical and reliable transmission of communications signals, including voice, data, electronic mail, and video. The satellite's advantages of distance-insensitivity, point-to-multipoint capability and improved quality over long distance are unique. These facts have led Dr. Clarke to speculate on the future impact of satellite communications in The View From Serendip, 1977:

> "I submit...that the eventual impact of the communications satellite upon the whole human race will be at least as great as that of the telephone upon the so-called developed societies. In fact, as far as real communications are concerned, there are as yet no developed societies; we are all in the semaphore and smoke signal stage...I believe that the communications satellite can unite mankind."

Brief History of Communications Satellites

In the late 1950's, both the United States and the Soviet Union began developing satellites and the necessary launch vehicles to place them into orbit. The Soviet Union's successful launching of SPUTNIK 1 came in October of 1957, and was closely followed by the United States' EXPLORER 1 in January of 1958. Shortly thereafter, in December 1958, the world's first active communication satellite, the U.S. Army-built SCORE, was launched. The ECHO satellites that followed in the early sixties were passive. They merely reflected signals back to the earth. Active satellites TELSTAR and RELAY, which amplified signals from the earth and retransmitted them, were launched in 1962.

The rockets available in 1960 could boost satellites into orbits no higher than 10,000 km above the earth. The challenge was to increase the orbit to approximately 36,000 km, where the satellite's period is one day. When located over the equator, the satellite at that height becomes geosynchronous; i.e., it appears fixed in space relative to earth stations on the ground. This has the obvious advantage of eliminating tracking electronics and position drives for antennas, thereby greatly simplifying the earth stations. The first successful geosynchronous satellite was SYNCOM II, launched in 1963.

The commercial era opened with the Communications Satellite Act of 1962 which set up the Communications Satellite Corporation (COMSAT) as the United States' "carrier" for satellite telecommunications. This was followed by the formation of the International Telecommunications Satellite Organization (INTELSAT) and the launching of "Early Bird" or INTELSAT I in 1964.

Canada, because of its widely separated areas of population and often harsh terrain and environment, was the first country to realize a domestic (DOMSAT) geosynchronous satellite system. Canada established TELESAT Canada in 1969 and launched ANIK A1 in 1972. Also in 1972, the Federal Communications Commission authorized the United States common carriers to construct and operate satellite systems for domestic telecommunications in the "free enterprise" mode. This led to Western Union's launching of WESTAR 1 in 1974, RCA's launching of SATCOM F-1 in 1975 and AT&T's launching of COMSTAR D-1 in 1976.

Since SYNCOM, over 100 geosynchronous satellites have been launched, of which about 90 percent have been communications satellites.

INTELSAT

The increase in capacity and capabilities of the INTELSAT satellites illustrates the growth that is possible in satellite communications. INTELSAT I (Early Bird), which was first operational in 1965, had an equivalent voice circuit capacity of 240 channels or one TV channel. It provided service between Europe and North America only. INTELSAT II had the same capacity. The capacity for INTELSAT III jumped to 1,500 channels and INTELSAT IV to 4,000 channels plus two TV signals. INTELSAT IV-A, which was first operational in 1975, has a capacity of 6,000 channels plus two TV signals (Figure 1).

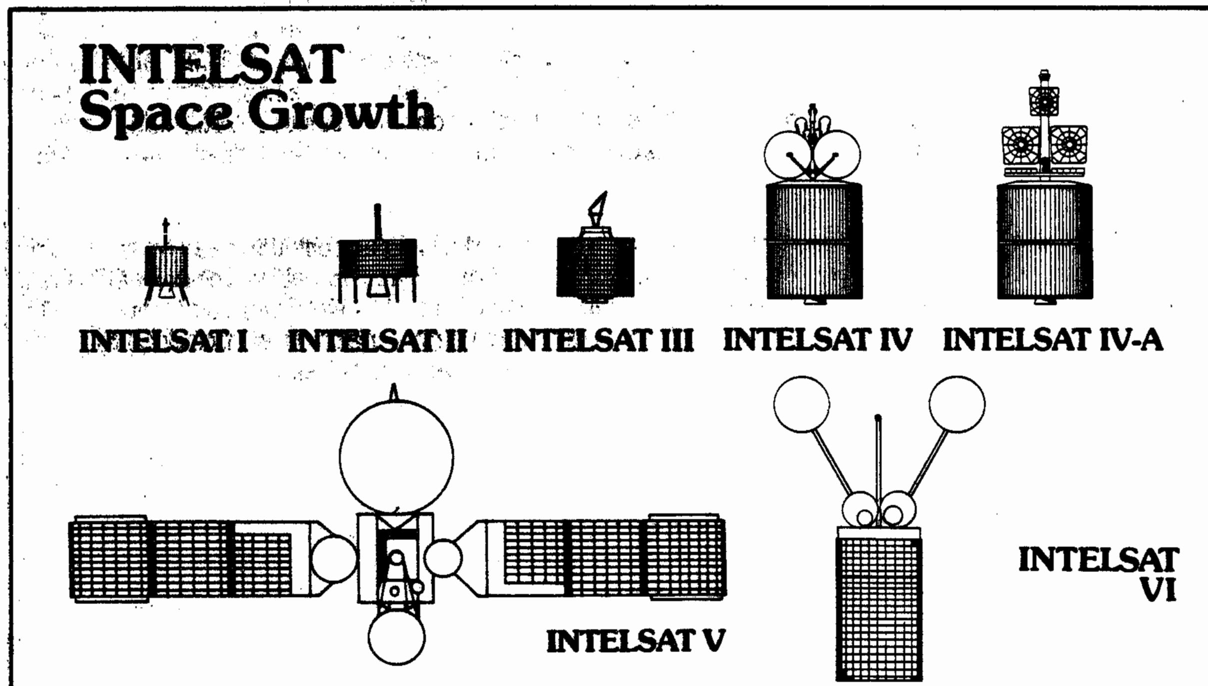


Figure 1. Growth of INTELSAT Satellites

INTELSAT V has a capacity of 12,500 voice channels plus two TV signals. It has global, zone and spot beams to supply different communications capabilities to different regions, and also uses the 11/14 GHz bands as well as the 4/6 GHz bands. Two hemispheric beams allow the use of the same frequencies in different areas and effectively double the traffic capacity for the frequency band. The use of orthogonal polarizations effectively doubles the available frequency spectrum. The combination therefore allows a fourfold use of the 500 MHz available at 4/6 GHz. A twofold use of the 11/14 GHz band is realized through the use of regional beams. INTELSAT V-A, which is to be launched in 1984, will have a capacity of 15,000 voice channels, and INTELSAT VI (1986) will have a capacity of more than 30,000 voice channels and four TV channels.

The size of earth stations for use in the INTELSAT system has been dramatically reduced. The system originally used 30-meter-diameter reflectors (standard "A" stations). With improvements in system technology, the use of smaller earth stations became possible. Antennas with diameters of 11 meters (standard "B" stations) are now in frequent use throughout the system.

The improvements in satellite and earth station technology have allowed a price reduction of a factor of four in leased international circuits since 1965. Discounting inflation, the real circuit cost has decreased by more than a factor of ten. In the years from 1965 to 1975, the INTELSAT network grew from 60 telephone circuits to about 10,000 circuits, more than doubling every two years. The number of each stations employed in this network now numbers about 300 in almost 150 countries.

A number of countries now lease whole or partial transponders from INTELSAT for intracountry communications. It should be noted that countries typically use the INTELSAT system for domestic communications prior to launching of their own domestic satellite. The principal advantage of a satellite system to a developing nation lies in the startling improvement in the reliability and quality of international communications. The relative immunity to signal degradation with poor weather conditions, multi-path and solar activity permits the establishment of communication links that are almost infinitely more reliable than conventional radio communication systems. In addition, good quality television and other services can be provided that were not previously possible.

The potential for competition in international satellite communications is on the horizon. Two companies, Orion Satellite Corporation and International Satellite, Inc., have each filed for Trans-Atlantic satellite systems. It appears that the international satellites business has the potential to become as competitive as the domestic U.S. satellite business.

Maritime Satellite Communications

Until the MARISAT system was initiated in 1976, radio communications with ships was in a primitive state, basically suffering from low capacity and poor reliability of conventional radio at the VLF, LF and HF bands. The MARISAT system opened a new era in communications by providing high-quality, highly reliable voice, data, facsimile and teleprinter service to ships at sea. The service is as simple to operate as ordinary telephone and telex and avoids the multiple hours that highly trained operators have to spend on the ship and on shore for a normal telegraph message. The system, now called INMARSAT, has revolutionized maritime communications and has enhanced the safety of ocean travel.

The INMARSAT commercial system consists of satellites over the Atlantic, Pacific and Indian Oceans and earth stations located on the coasts. Communication between the shore stations and the satellite is in the 4/6-GHz band. Communication between the ships and the satellite is in the 1.4/1.6-GHz region. The shipboard units are stabilized to keep the antennas pointed toward the satellite in the presence of roll, pitch and yaw motions of the ships.

Today there are over 1,900 INMARSAT terminals in operation on ships and offshore drilling rigs. The number of users is expected to exceed 10,000 by the mid-1990's.

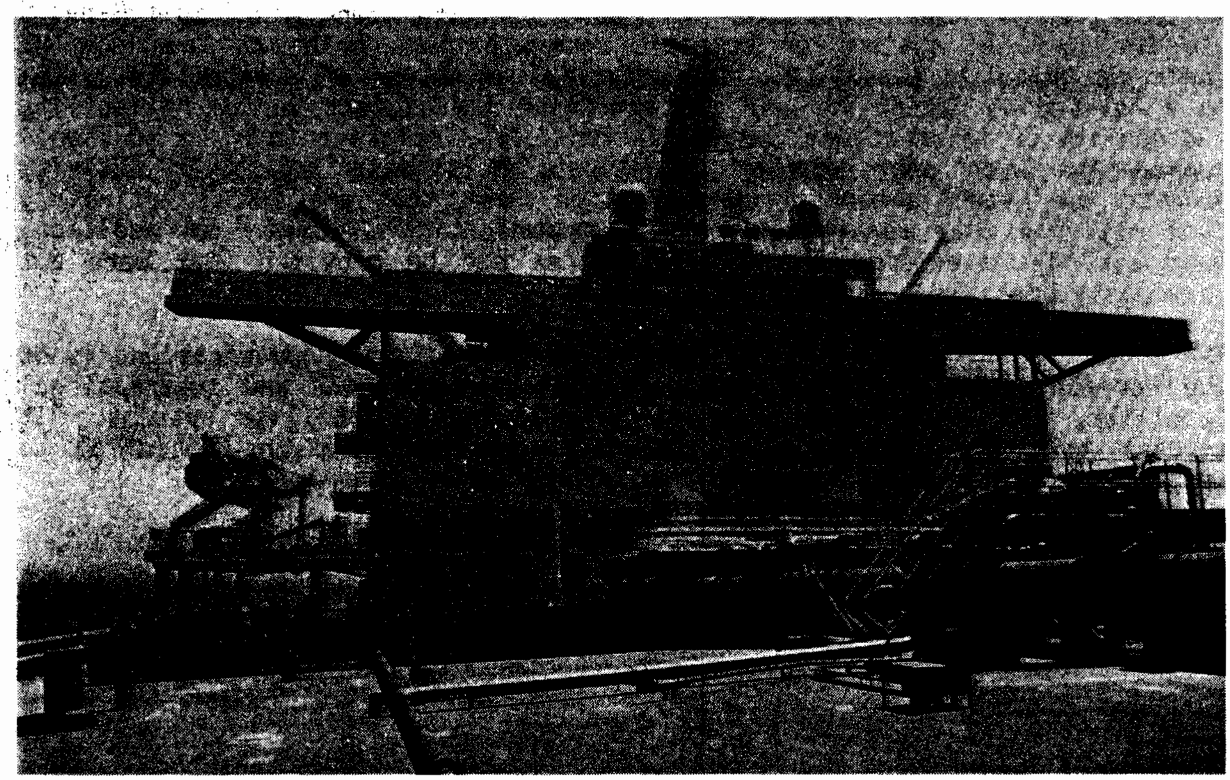


Figure 2. Twin MARISAT Antennas for Ship-to-Shore Communications via Satellite

United States Domestic Satellite Systems

The development of domestic satellite communications in the United States began with the establishment by the common carriers of earth station and satellite networks for transmission of voice, television and data to the large cities. The first U.S. domestic satellite communication system was placed in operation in 1973 by RCA American Communications (Americom), using the Canadian ANIK A2 Satellite. RCA used WESTAR in 1974 and has used its own satellite system since 1975. Most of RCA's transponders are being used for cable television distribution to a very large network of earth stations. RCA also provides leased private line service to a number of companies, performs specialized voice, television and data services for the U.S. government using dedicated earth stations, and leases transponders to common carriers.

Western Union, who launched the first commercial satellite for U.S. domestic communications, has installed medium/heavy route earth stations to provide metered private line, data and dedicated voice service to a number of cities. The earth stations are integrated into Western Union's extensive terrestrial microwave system to carry telex, mailgram, voice and data for Western Union itself. In addition, Western Union distributes television and radio for a number of users.

American Satellite Company, jointly owned by Fairchild Industries and Continental Telephone, shares the WESTAR system. American Satellite specializes in providing voice and data communications to 5- to 11-meter earth stations located on end-users' premises through their Satellite Data Exchange (SDX) service.

AT&T and GTE currently use the COMSTAR satellites owned by Comsat General, a subsidiary of COMSAT. AT&T launched its own satellite called Telstar in August of 1983. In 1984, GTE will launch its own satellite called GSTAR, which will operate at 12/14 GHz.

In 1980 Satellite Business Systems (SBS), a partnership among wholly-owned subsidiaries of Comsat General Corporation, IBM, and Aetna Life and Casualty Company, started implementing a satellite system in the 12/14 GHz frequency band. The first SBS satellite was launched in November of 1980, the second in September of 1981, and the third in November of 1982.

It is estimated that above mentioned common carriers have installed about 350 transmit/receive earth stations. Another 650 transmit/receive earth stations have been installed by others for such purposes as television uplink for cable television distribution, local television origination, and voice and data uses.

Alaskan Satellite Communications

Alaska is an ideal region for satellite communications. With the snow, ice, mountains and large territory, it is much easier to install satellite earth stations than to install repeater stations. Before satellites were available, HF radio provided the only communication, which was often unreliable due to frequent auroral disturbances. One use for satellite communications was planned in 1972 to provide voice and data communications for the Alaskan pipeline. The communication system was designed to be the world's most reliable, with a back-bone system of line-of-sight microwave and a satellite backup system. Ten-meter earth stations were installed in Valdez, Prudhoe Bay, and Fairbanks, by RCA Alaska Communications to complement a 30-meter earth station near Anchorage which had been taken over from Intelsat. The communication system has indeed become as reliable as planned.

An issue continually facing Alaska is the health and education of the people living in remote areas. Experiments were conducted in the early 1970's using ATS-1 and ATS-6 satellites for reaching the remote areas. In 1975, the State of Alaska appropriated funds for the purchase of 100 small 4.5-meter earth stations to serve 100 remote villages, using one of the new domestic satellite systems. Alaska now receives telephone and television service using the small earth stations in the remote villages and a number of 10-meter earth stations in the larger communities. Live television from the contiguous U.S. (CONUS) is transmitted to the large communities, using two television signals on one satellite transponder. Other services are offered, including instructional television, teleconferencing, facsimile and computer data transmission. Alaska has one of the largest SCPC voice networks in the world. Alaska now has its own satellite, AURORA (SATCOM V), located at 143°W.

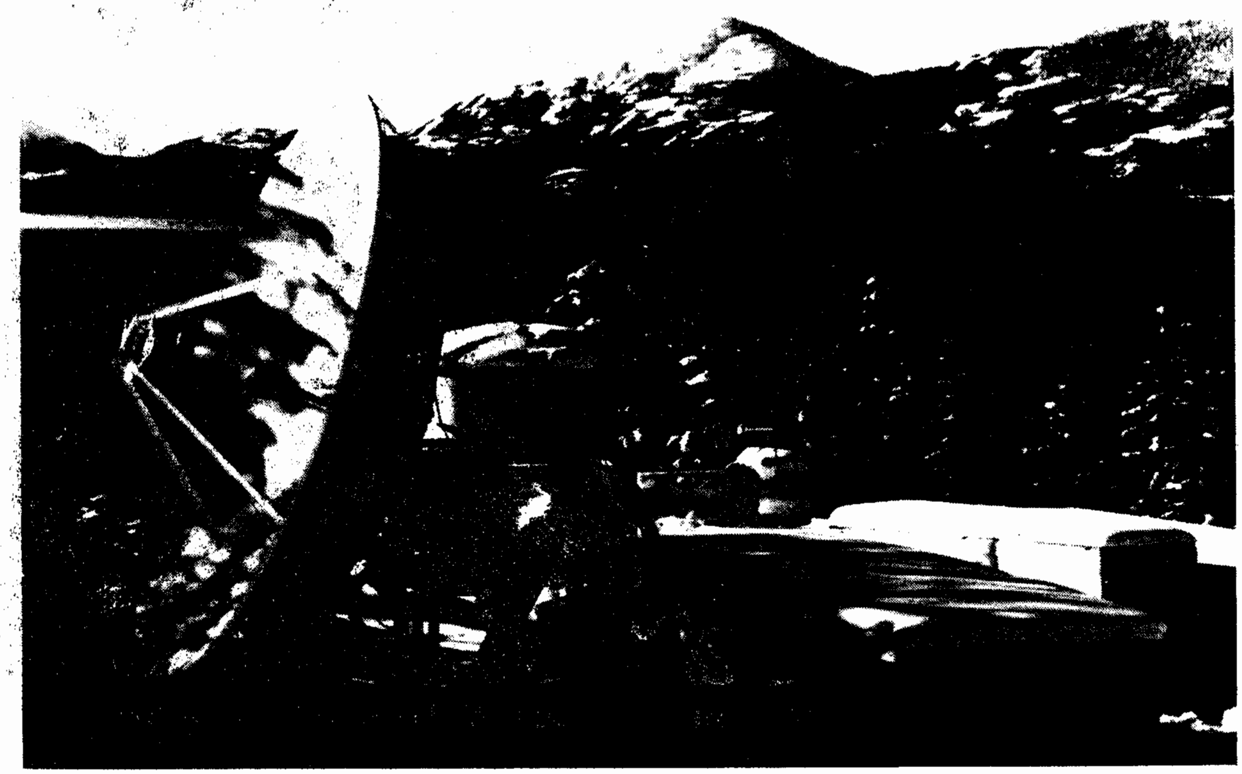


Figure 3. Alaskan Earth Station

Cable Television

Satellite broadcast to cable television systems is a classic example of a private enterprise system at its best. Equipment Manufacturers, program suppliers, satellite operators and CATV operators combined to bring significant strength to an industry that was otherwise rather stagnant.

The story goes as follows: In late 1972, Teleprompter Corporation became interested in using satellites to distribute television programs. They contracted with Scientific-Atlanta for a transportable earth station to be moved around the United States to demonstrate the excellence of TV distribution by satellite. This was successfully done. In 1975 a program supplier, Home Box Office (HBO), announced its intent to distribute programming by satellite. HBO signed an agreement with RCA, which in turn leased interim space on WESTAR I until its satellite was launched. Earth stations were purchased from Scientific-Atlanta by UA-Columbia and by ATC. Programming began on September 30, 1975, with the famous Ali-Frazer fight, the "Thrilla in Manila."

In 1976, HBO had about one-half million subscribers, with about one-eighth of these receiving the programs on 45 earth stations. HBO now has about 13 million subscribers on over 3,000 cable television systems, most of which use satellite earth stations for reception of the programming.

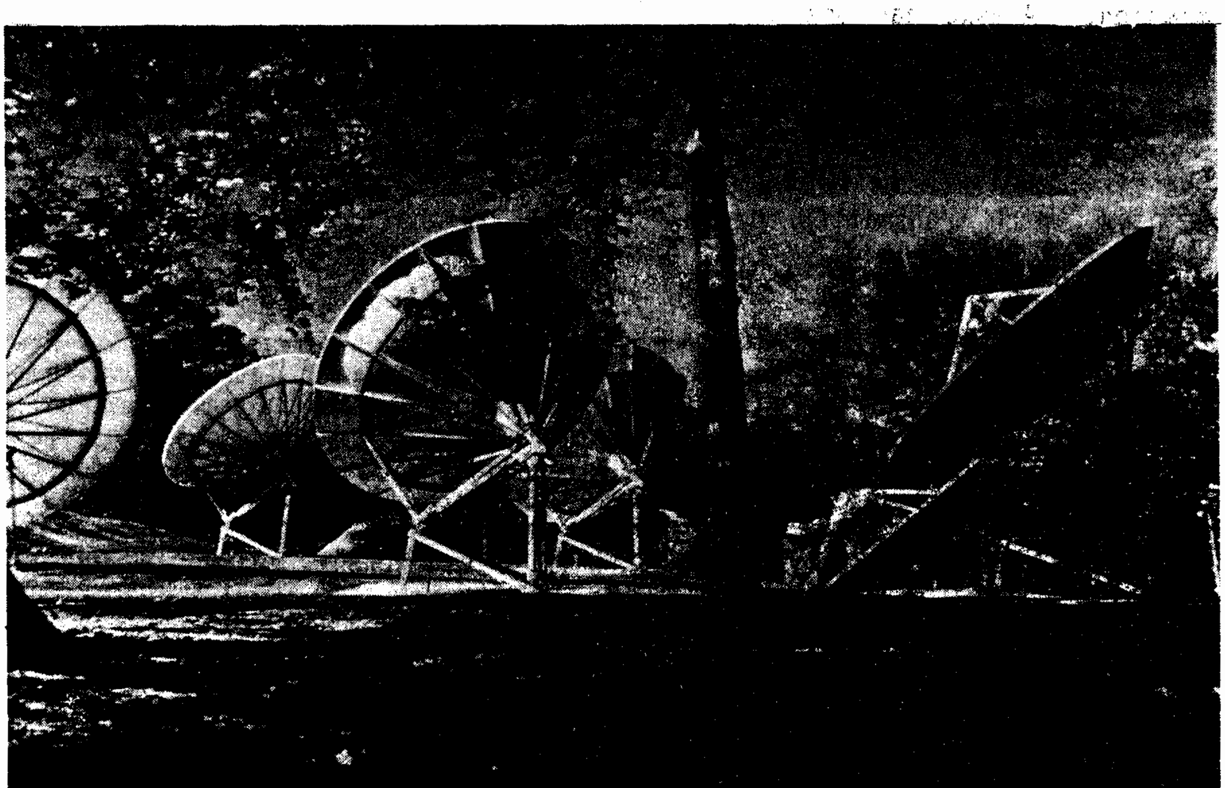


Figure 4. Cable News Network (CNN) Uses Multiple Earth Stations

Home Box Office was not the only success. In 1976 the FCC allowed Southern Satellite Systems, Inc. to provide satellite common carrier service for WTBS (formerly WTCG) in Atlanta. WTBS is now received by over 20 million subscribers on over 3,500 cable television systems. More than 4,500 cable systems are now equipped with earth stations. Several systems have two earth stations and some have four earth stations, making a total of about 10,000 cable television earth stations. A number of factors have led to this growth. One was obviously good TV programming. Another was the FCC's reduction of minimum size receive-only (TVRO) antenna diameter from 9 meters to 4.5 meters in 1976 and deregulation in 1979. Another was investment in technology and manufacturing facilities for earth stations. The first TVRO earth station sold for around $75,000. TVRO earth stations can now be bought for less than $10,000.

The cable television programming is varied and extensive. There are a number of independent TV stations offering advertising-supported programs transmitted by satellite. Movie and sports channels exist in abundance. In addition, there are children's channels, religious channels, ethnic channels, news channels, political channels, educational channels, music channels, business channels, women's channels, cultural channels and adult channels. By 1986 it is expected that cable programming will be carried by a number of satellites including RCA, SATCOM, Hughes Galaxy I, Western Union WESTAR, and Southern Pacific Spacenet.

Cable television has the potential to expand in Europe much as it has in the United States. The European Communications Satellite, ECS-1, began service in mid October of 1983, using the 12/14 GHz frequency band. The seven Eutelsat signatories using the satellite are Germany, Belgium, France, Italy, the Netherlands, the United Kingdom, and Switzerland.

Broadcast TV Stations

The first customer to transmit television for broadcast was the Robert Wold Company, which transmitted the Texas Rangers baseball game from Milwaukee to Dallas by WESTAR in 1975. The Robert Wold Company and others have transmitted many program hours of television since then.

There are a number of organizations, including Independent Television News Association, Robert Wold, Westinghouse's Vidstar, Spanish International Network, Christian Broadcast Network, PTL, Trinity Broadcasting and the Public Broadcasting System (PBS), which distribute programs to broadcast stations. The Public Broadcasting System (PBS) has now installed earth stations for the distribution of television to most of its affiliates.

ABC, CBS and NBC regularly use satellites to relay news, sports and special programs on a point-to-point basis. These networks have plans to distribute television by satellite to the local stations.

CBS will implement a C-band distribution to its affiliates using 7-meter and 4.6-meter antennas. Up to nine 10-meter antennas and a number of transportables will be used to uplink remote programming. The total system is to be in place by 1987. NBC, on the other hand, will implement a Ku-band distribution system. The system will begin using SBS satellites and then migrate to RCA's Ku-band satellites when launched.

Radio broadcast programming is being distributed by satellite. Mutual Broadcasting System and National Public Radio have satellite distribution systems. The ABC, CBS, NBC and RKO radio networks have selected Scientific-Atlanta digital audio earth stations for distribution of programs to their affiliates. The digital service offers superior audio signal-to-noise, dynamic-range, distortion and crosstalk quality. Up to twenty 15-kHz audio channels can be transmitted through one transponder and received by inexpensive 2.8-meter earth stations. Over 1,500 earth stations have been installed by the end of 1983 in this program. The number of earth stations in use by radio broadcasters will probably exceed 5,000 by 1985.

Figure 5. Digital Audio Earth Station

Print Media

Newspapers printed in various corners of the world are fed by signals that are transmitted by satellite. The Dow Jones Company was one of the first to use this transmission medium. Its initial domestic transmission began in 1976. The use of the satellite has now been expanded to include overseas printing of the Wall Street Journal.

The New York Times transmits its pages electronically from New York. It is now printed simultaneously in Chicago and Florida, as well as in New York. This is made possible through a satellite communications system using American Satellite digital earth terminals.

Hotel/Motel

A large market for satellite communications terminals is the hotel/motel industry. Major hotel/motel chains are providing improved television entertainment to guest rooms and are offering additional meeting services through teleconferencing. Holiday Inns were the first large hotel/motel chain to utilize the satellite. In July 1979, Holiday Inns signed contracts for approximately 200 earth stations to be used by the corporate-owned hotels. Many hundreds of earth stations have been installed at hotels and motels since then. A very common sight as one travels the highways is that of the satellite "dish" at the hotels.

Mini-Cable/SMATV, Home Market

In 1979 the FCC deregulated receive-only earth stations, freeing the small TV receive-only (TVRO) from restrictions and frequency coordination. This event signaled the opening of a mini-cable and satellite master antenna TV (SMATV) market for TVROs for condominium and apartment complexes and subdivisions. The earth stations, usually smaller than those at cable head-ends, can serve remote locations less expensively than service can be provided by extending the cable distribution system. A representative of the Society for Private and Commercial earth stations has estimated that there are about 400,000 TVROs installed and 20,000 a month being shipped.

Most of the pay-TV companies feel that scrambled TV by satellite is a necessity. The transmission of television in a scrambled format is, by definition, "non-standard". The use of a non-standard signal has led to the exploration of a format that would be ideal for transmission by satellite. One such transmission system, called MAC (multiplexed analog components) allows the delivery of superior quality video, two or more channels of high-fidelity sound, addressability and encryption.

An addressable satellite system allows the further expansion of pay-TV into markets that could not previously be explored. Combined with low-cost earth stations, pay-TV operators can address mini-cable systems for hospitals, apartment complexes, condominiums and suburban communities.

The investment in technology and manufacturing facilities and the large market volume have caused a dramatic decrease in the cost of earth station TVRO receivers. Scientific-Atlanta's first TVRO receiver, the Model 411, sold for about $16,000 in 1974. Approximately one-hundred of these video receivers were sold. The next generation receiver, the 414, sold for about $8,000 in 1977, and the total quantity sold was slightly over 2,000. The third generation receiver, the 6600, sold for about $3,500 in 1980; and about 30,000 receivers have been sold at this point. The fourth generation receiver, the 6650, sells for about $1,800 and a fifth generation receiver, the 9500, has just been introduced.

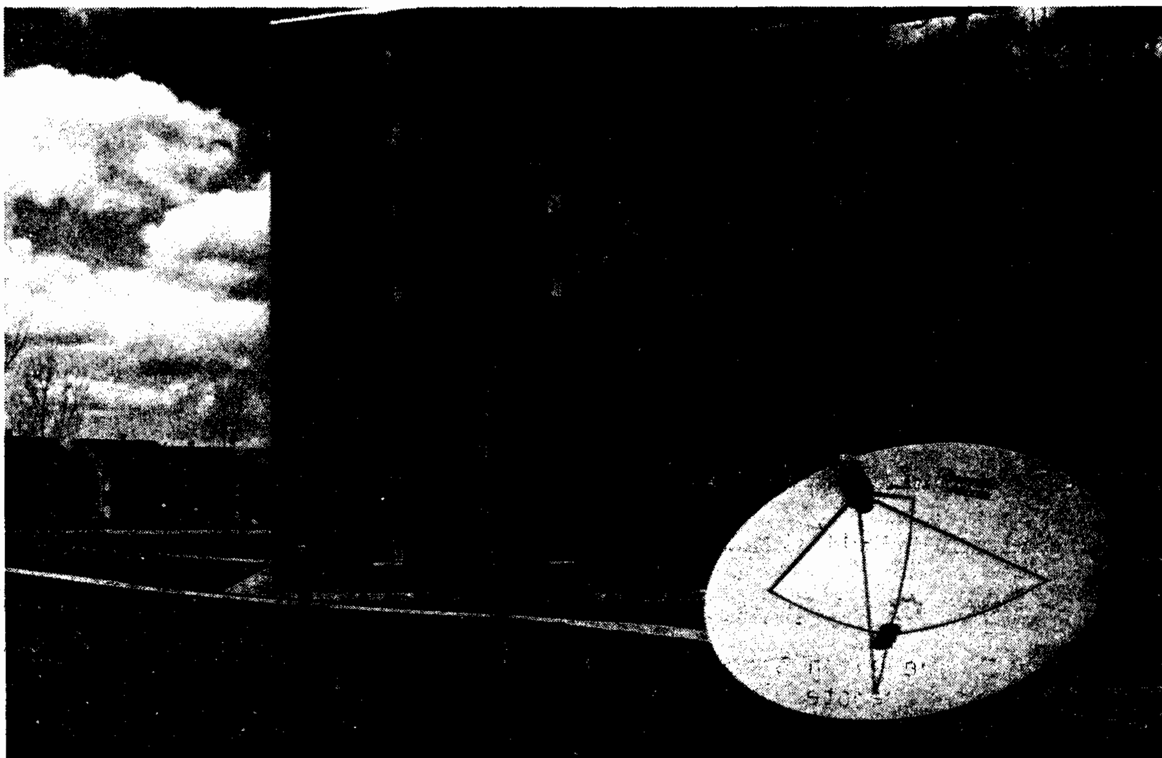


Figure 6. Earth Stations for Mini-cable Systems Serve Hospitals, Condominiums, Apartment Complexes and Suburban Communities

Business Telecommunications Earth Terminals

The business telecommunications market is very large. The use of digital processing for voice, image, and data is increasing dramatically. Digital earth terminals are expected to be on the rooftops or in the parking lots of virtually every major shopping center and office complex in the country, providing facsimile, voice and video messages. Executives in widely separated cities will be able to see and hear each other--and transmit hard-copy messages--in teleconferences with satellite hookups. Digital terminals allow computer-to-computer dialogue, entry to, computation, and retrieval from a central computer, and transmission of printed information, digitized voice and facsimile. A large, geographically-dispersed company in the U.S. can have instant full-time interconnectivity for all its locations.

In 1980 SBS, started implementing an extensive digital time-division-multiple-access (TDMA) system for transmitting voice, data, and image. The seven-billion-dollar intracompany business communications market is SBS's principal target. SBS offers complete voice, data, electronic mail, and teleconferencing services to large corporations. The earth stations operate in the 12/14-GHz frequency band so that they can be located in cities without frequency interference problems. The terminals are small (5.5-meter and 7.7-meter antennas) and can be located on customer premises.

The need of insurance companies to share data has stimulated the formation of a resale common carrier of SBS service. ISACOMM, which is majority owned by United Telecommunications, is selling communications services to smaller users, initially in the insurance industry. ISACOMM stated service through earth stations in Wausau, Wisconsin, and St. Louis, Missouri, in early 1981. The company anticipates a total network of 40 earth stations by 1984.

Telecommunications from shared earth stations can be distributed locally through cable that is being laid by the CATV system operators in the United States. New CATV operators are installing two-way business cables in addition to home entertainment cables. The new cables offer over 400 MHz of available communications capacity.

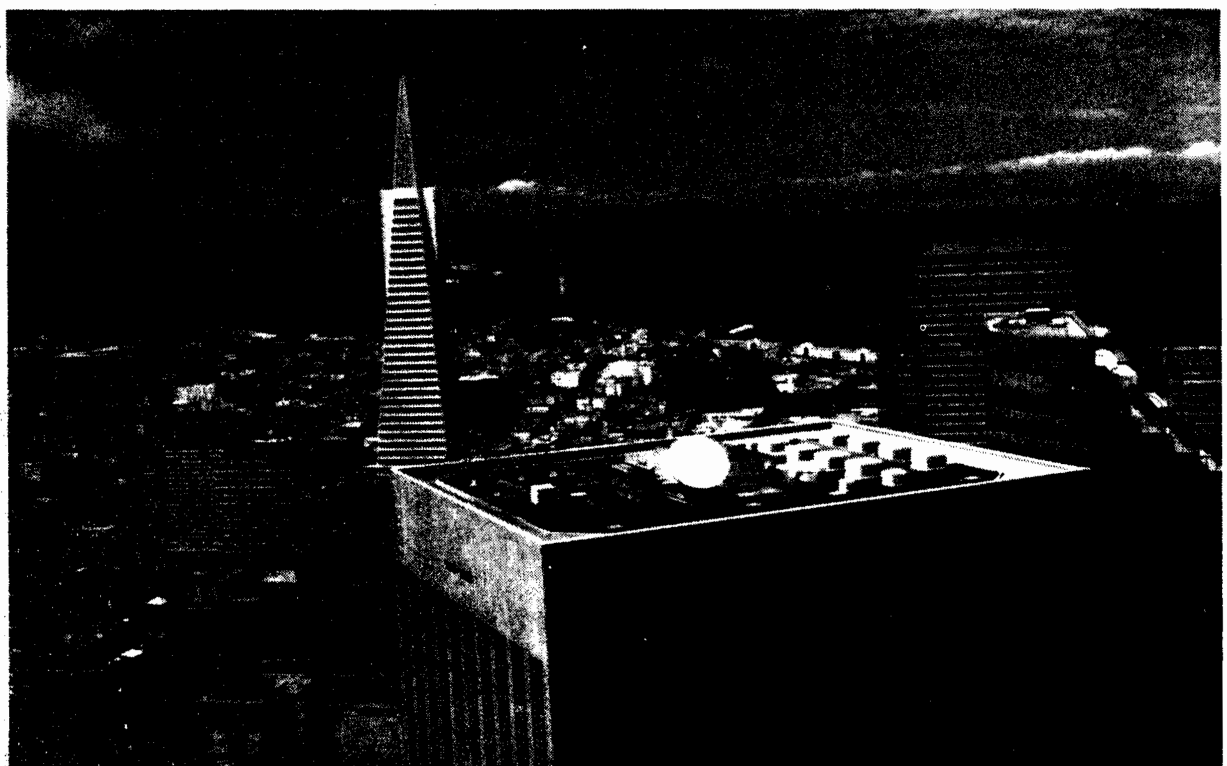


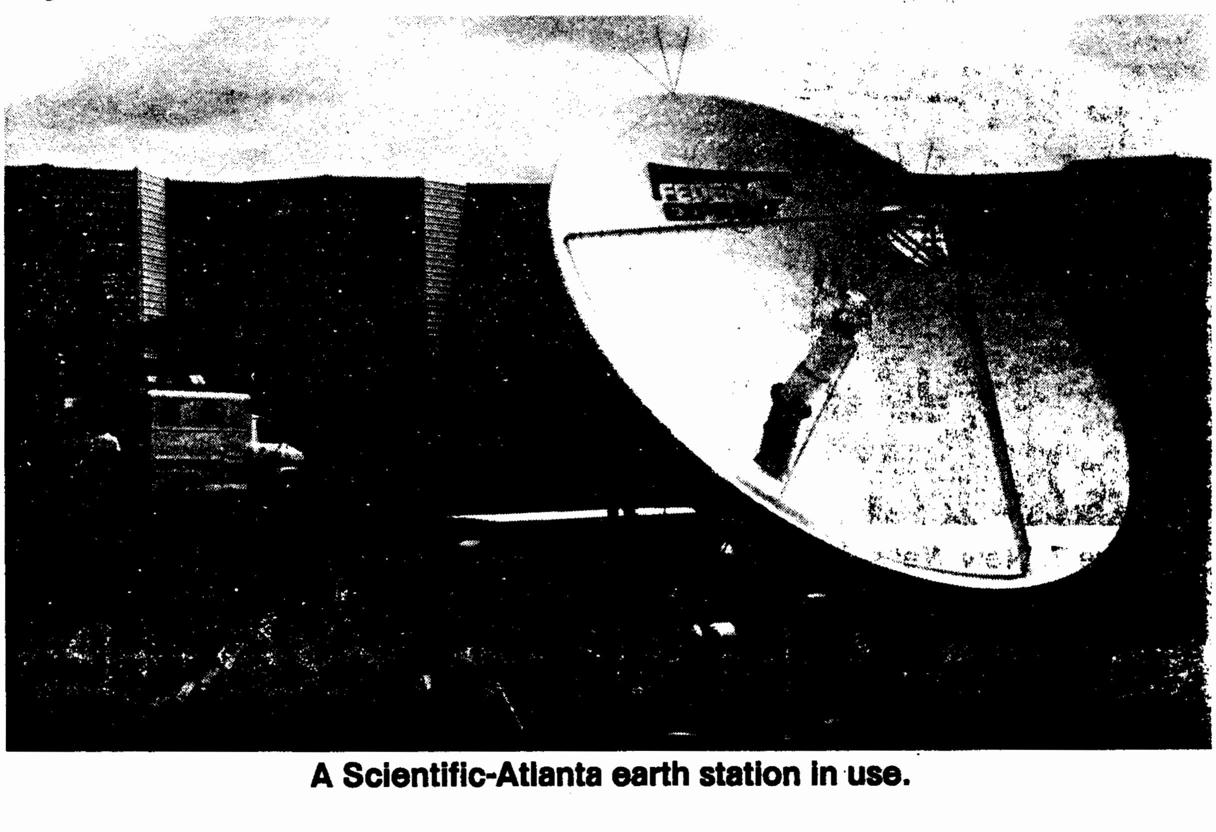
Figure 7. New Wells Fargo Bank in San Francisco Using SBS Earth Terminal

Future Directions

The lack of available orbital positions for new satellites serving the United States and the demand for new voice, data, and video services have caused the FCC to adopt closer satellite spacing. The FCC will immediately implement uniform 2° orbital spacings for newly launched 12/14 GHz satellites. The current U.S. satellites in orbit using 12/14 GHz (Ku-band) are SBS1, SBS2, and SBS3 located at 100°W, 97°W, and 94°W, respectively. These satellites will be moved to 99°W, 97°W, and 95°W to accommodate new Ku-band satellites as shown in Table 1.

The costs and difficulties of immediately implementing 2° spacings at 4/6 GHz (C-band) have caused the FCC to adopt an interim spacing plan providing a combination of 3°, 2.5°, and 2° orbital spacing. Never-the-less, 2° was adopted as the long-term orbital spacing for C-band. The assigned orbit positions and the satellites expected to occupy these positions are shown in Table 1 for both C- and Ku-bands. It is noted that for C-band, 2° spacing is used in the eastern and far western part of the arc, 2.5° in the middle of the arc, and 3° in the western part of the arc. Note that Satcom 3-R which has more earth stations pointing toward it than any other satellite, does not have to relocate to accommodate the new satellites in the interim spacing plan.

From Table 1 it is seen that only nine C-band slots are unassigned and only 16 Ku-band slots are unassigned from 55°W to 143°W. The number of new satellite applicants is expected to greatly exceed the available orbital slots. Already in line for the remaining slots are CableSat General (3 satellites), Ford Aerospace (3), GTE (2), and Hughes (4), National Exchange (5), SBS (1), and Western Union (5). The FCC is expected to have to decide from among the many applicants who will receive the new orbital slots, something that it has sought to avoid.



A Scientific-Atlanta earth station in use.

Table 1. U.S. Orbital Positions

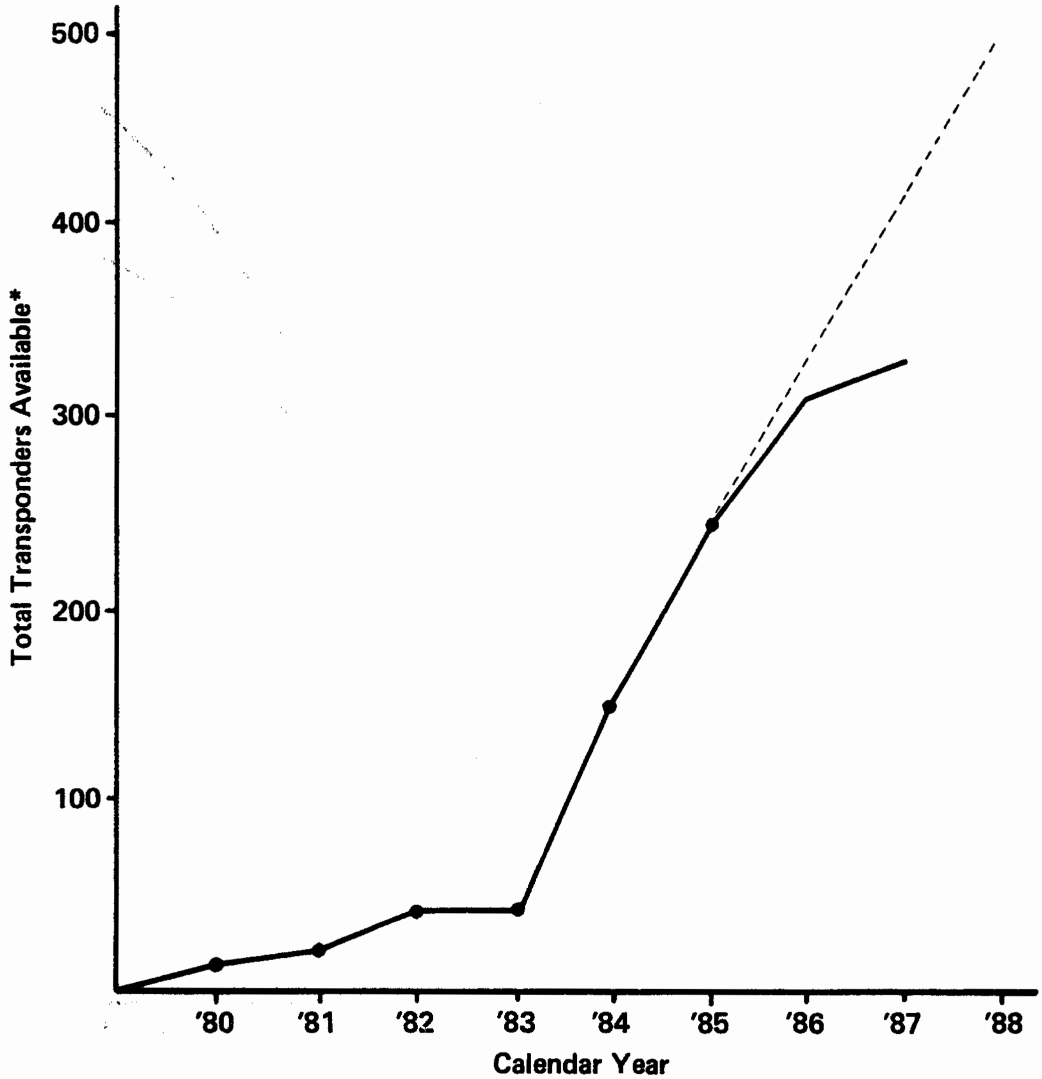
C-Band		Ku-Band	
55	-	55	-
57	-	57	-
59	-	59	-
61	-	61	-
63	-	63	-
65	-	65	-
67	Satcom 6	67	-
69	Spacenet 2*	69	Spacenet 2*
		71	-
72	Satcom 2R	73	-
74	Galaxy 2	75	-
76	Telstar 3	77	RCA K2
78.5	Westar 2,3	79	Rainbow 2
81	AmSat 2*	81	AmSat 2*
83.5	Satcom 4	83	ABCI 1
		85	USSSI 1
86	Westar 6	87	RCA K1
88.5	Telstar 2	89	SBS4
91	Spacenet 3*	91	Spacenet 3*
93.5	Galaxy 3	93	-
		95	SBS3
96	Telstar 1	97	SBS2
98.5	Westar 4	99	SBS1
101	-*	101	-*
		103	GSTAR 1
104.5	Canada	105	GSTAR 2
108	Canada	107.5	Canada
		110	Canada
111.5	Canada	112.5	Canada
113.5	Mexico*	113.5	Mexico*
116.5	Mexico*	116.5	Mexico*
		117.5	Canada
119.5	Westar 5	120	USSSI 2
122	Spacenet 1*	122	Spacenet 1*
		124	SBS5
125	Comstar 4	126	RCA K3
128	AmSat 1*	128	AmSat 1*
		130	ABCI 2
131	Satcom 3-R	132	Rainbow 1
134	Galaxy 1	134	-
		136	-
137	-	138	-
139	Satcom 1-R	140	-
141			
143	Satcom 5		

*Dual C/Ku

The satellites to be launched in the next few years are listed in Table 2 along with their expected launch dates. Hybrid satellites contain both C-band and Ku-band transponders. It is seen that a large number of Ku-band satellites are to be launched in the next three years. The number of Ku-band satellite transponders available for U.S. domestic communications is shown in Figure 8 versus calendar year. It is noted that the number of Ku-band transponders will quadruple during 1984.

Table 2. Near-Term Satellite Launch Schedule

C-Band	Date	Ku-Band	Date
RCA SATCOM 6	5/86	SBS4	9/84
		SBS5	2/86
AT&T TELSTAR 2	8/84	GTE GSTAR 1	5/84
AT&T TELSTAR 3	5/85	GTE GSTAR 2	8/84
		GTE GSTAR 3	1985
WU WESTAR 6	3/84	RCA K1	5/85
WU WESTAR 7	1984	RCA K2	1/86
WU WESTAR 8	1985	RCA K3	8/87
		WU WESTAR 9	1985
HUGHES GALAXY 3	6/84	WU WESTER 10	1985
		WU WESTAR 11	1986
		USSSI/USAT 1	2/84
		USSSI/USAT 2	8/84
		ABCI 1	12/86
		ABCI 2	2/87
		RAINBOW RSI 1	8/86
		RAINBOW RSI 2	11/86
Hybrid		Hybrid	
SPC SPACENET 1	5/84	AMERICAN SATELLITE 1	9/85
SPC SPACENET 2	10/84	AMERICAN SATELLITE 2	3/86
SPC SPACENET 3	3/85		



*Sum of all Transponder Types
(i.e., 43 MHz, 54 MHz, and 72 MHz)

Figure 8. Ku-Band Transponder Availability (Based on FCC 2° Assignments)

The number of C- and Ku-band transponders covering the domestic U.S. is shown versus calendar year in Figure 9. Beyond 1985 there is an optimistic as well as a pessimistic projection depending on whether one believes that transponders are being launched faster than they can be used and will be slowed down or that the companies will continue launching satellites, obtaining rights to valuable real estate and loading the transponders as much as possible. In 1986 there will be over 500 C-band transponders and over 200 Ku-band transponders aimed toward the United States

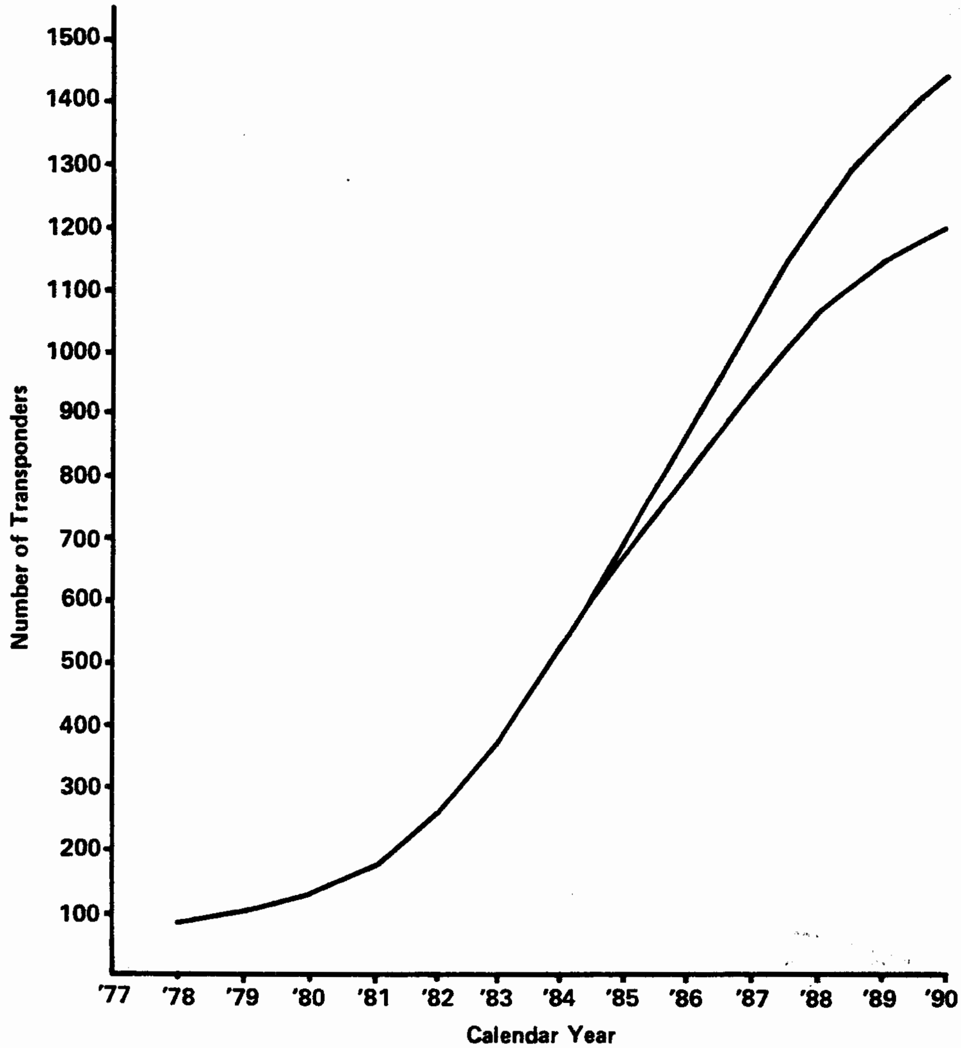


Figure 9. Number of Satellite Transponders — Domestic U.S.

The orbital arc is finite and the arc above the U.S. is likely to be saturated by the end of this decade even with 2° spacing. This means that efficient use of spectrum is a must. Hybrid C- and Ku-band satellites must use the spectrum as efficiently as C- or Ku-band satellites or the hybrid satellites should not be used. In the long run, however, techniques such as bandwidth compression, switching spot beams, Ka-band (20/30 GHz), and space platforms should provide a steady increase in satellite communications capacity well into the next century.

Video-Plus

An example of more efficient use of available transponders is embodied in the concept of video-plus. A transponder, which carries a video signal, can be shared with other carriers, such as 56 kb/s digital SCPC, without interference to the video signal or the digital carriers. Furthermore, this can be accomplished without modification of the modulation format of the video signal. Thus, transponders, which previously carried only one video signal, may be used to carry many additional voice and data signals.

Video-Plus, which is described in a later chapter, is not a subcarrier technique. SCPC carriers can be transmitted from uplinks completely separate from the video uplink. Hence, the signals do not have to be combined at one origination site. For example, a hotel using a 4.6-meter dish to receive entertainment programming from a satellite could add transmit capability to its earth station and share one of the video transponders. A typical application might be hotel reservations or intercity telephone service for guests.

Other applications of Video-Plus include:

- Nationwide Control of Addressable CATV Set-Top converters
- One-way Video Teleconferencing/Two-Way Voice and Data for Business Education
- Small Private Telecommunication Networks
- Interactive CATV Systems with One-Way Video and Two-Way Data and Voice

Direct Broadcast Satellites (DBS)

The capability of broadcasting a TV signal through a satellite directly to the home has been under development for some time. In fact, a number of U.S. households now receive satellite broadcasts from current C-band satellites with 3-meter antennas located at their homes. However, here we wish to examine DBS terminals with antennas whose diameters are less than 1-meter and therefore have lower cost and can be located on the roof of the home, if desired. It is not possible to use antennas with diameters smaller than 1 meter in the 4/6-GHz band because of the interference from adjacent satellites and because of the relatively low allowable EIRPs of satellites in this band. The DBS systems will make use of Ku-band and high-power amplifiers on board the satellites.

The satellites with very high power amplifiers will not be launched until late 1985 or early 1986. In the meantime, a number of entrepreneurs plan to offer interim DBS services. For the most part, these interim services require antennas larger than one meter in diameter because of the relatively low power of the satellites. However, Satellite Television Corporation (STC) plans to offer an interim DBS system using the high-power transponders of SBS4. These transponders will allow the use of antennas as small as 0.75 meter (2-1/2 foot) which can be located on home rooftops. The system will then migrate to higher power DBS when the satellites are launched in

1985 to 1986. Technically, this system makes a lot of sense. It allows small roof-mountable antennas. When the higher power DBS satellites are launched in 1986, extended definition televison can be received by the homeowner. When high-definition TV sets are available, HDTV can also be broadcast to more expensive receivers offering theater quality into the home.

COMSAT, RCA, Western Union, CBS, USSB (Hubbard Broadcasting), and DBSC (Pritchard) are among those applying for higher power DBS systems. Some of the applicants propose to provide pay programming, and other advertiser-supported programming. Some plan to originate programs and others to offer leased channels. High definition television (HDTV) and channels of stereo audio are among the options offered by some of the DBS applicants.

The technology is available to build a DBS system. There are, however, financial, programming and regulatory problems to be faced. The launching of a three- or four-satellite system and the production of programming will cost the DBS applicant upwards of three-quarters of a billion dollars. In addition, homeowners may not buy or lease a DBS terminal unless there are many channels of good programming unavailable elsewhere at lower cost. With only three to five channels per time zone and satellite, this means that the DBS home terminals must be able to receive programs from a number of satellites. In order to receive programs from more than one satellite, the technical parameters must be compatible from satellite to satellite.

There is competition for orbit space for DBS, as there is for C-band and Ku-band DOMSAT. A number of North American countries have been alloted DBS orbital slots during RARC-83. These include Canada, United States, Mexico, Cuba and the Bahamas. The U.S. obtained eight DBS orbital positions above the U.S. Approximately five to eight satellites can be placed "close together" in each orbit position so that they appear as the same satellite to the small DBS antennas. Efficient utilization of the spectrum dictates that each satellite has about four to five high-power transponders (plus backup). There could likely be 20 or more DBS satellites operating above the U.S. by 1988, allowing reception by the homeowner using a smaller than 1-meter diameter antenna.

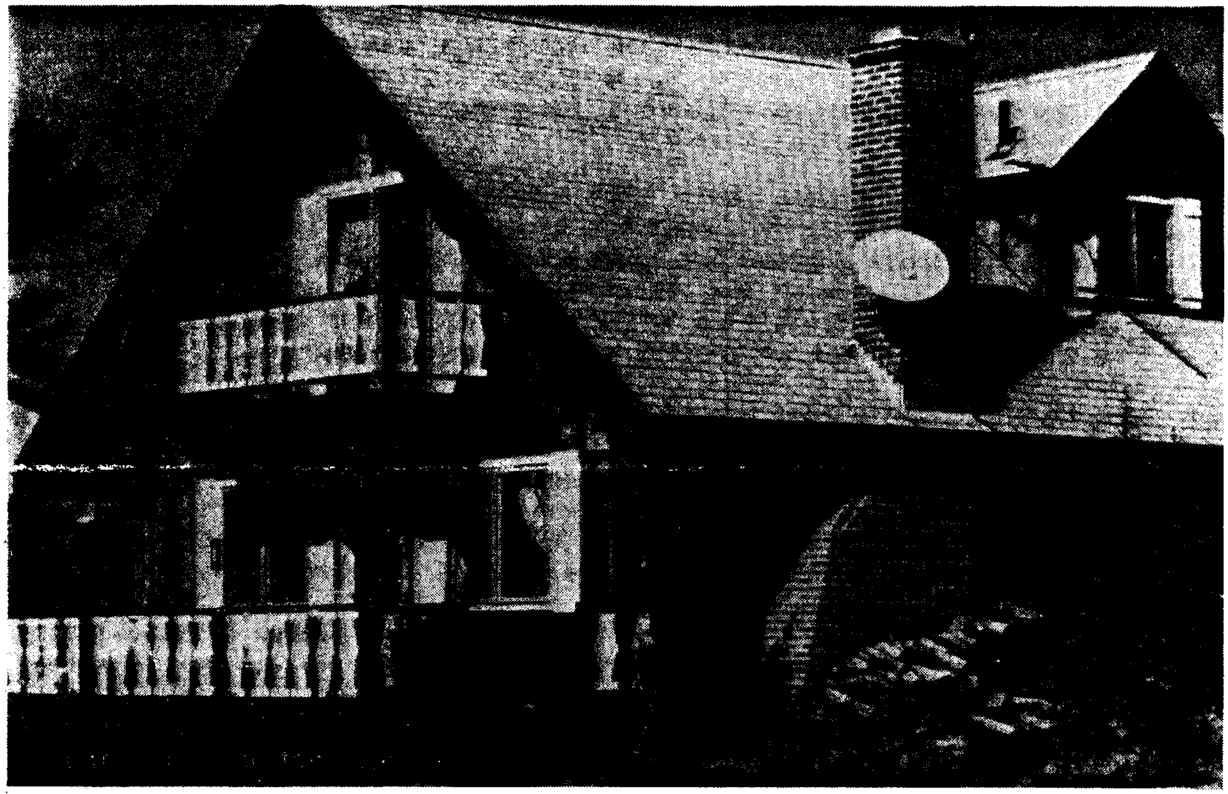


Figure 10. DBS Terminal (Photo courtesy Satellite Television Corporation)

Conclusion

The satellites to be launched in the next few years are reasonably well defined now. The principal improvements will be higher power TWTs, new antenna designs, switching flexibility and higher frequencies.

The push to higher frequencies will avoid the current terrestrial interference problem, making earth stations in the cities more practical. High frequency satellites also favor high power satellites because of a lack of interference with terrestrial microwave. The main disadvantage of the higher frequencies is susceptibility to rain outage. Techniques such as "diversity," the use of alternate transmission paths, or reduction of the data rate may be used to circumvent this problem.

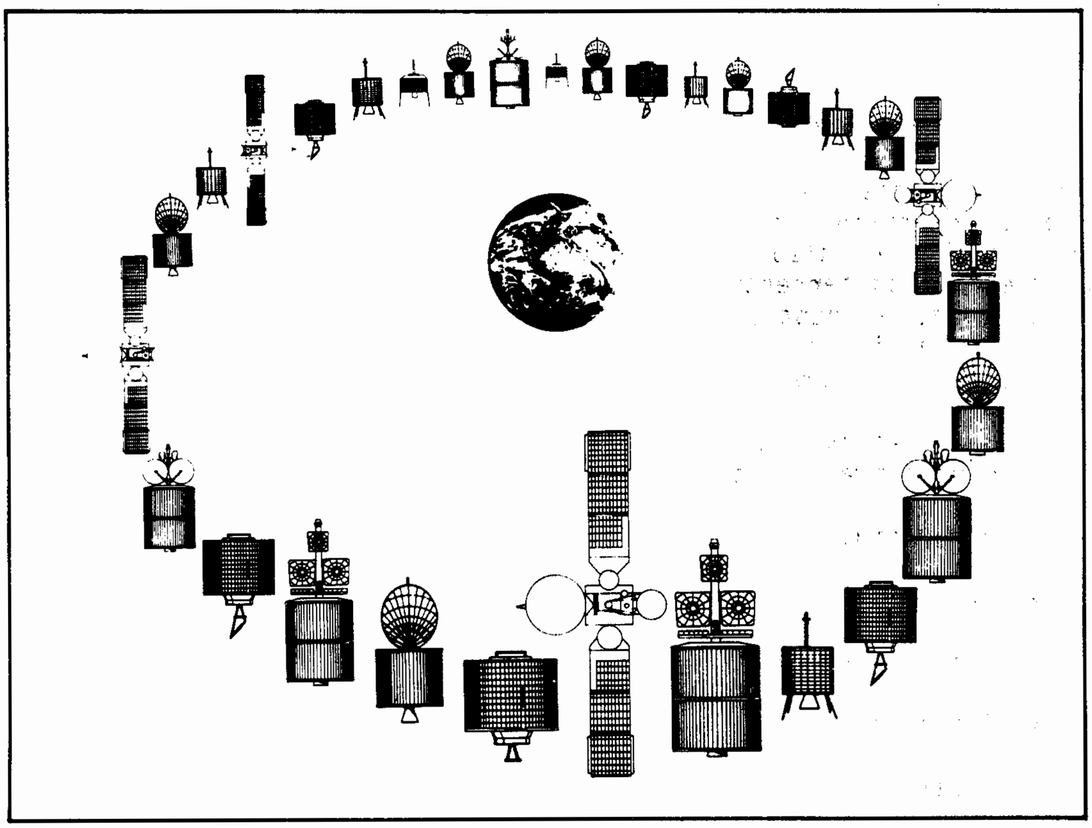
In the more distant future, a number of technological advances will permit growth and performance improvement in satellite communications.

- High-gain, multiple-beam antennas for the satellite
- Very efficient, lightweight solar arrays
- High-power, Solid-state transponders
- Improved satellite battery life
- Satellite beam switching

- Inter-satellite links that will allow placement of satellites at the extremes of the service arc
- Development of Ka-band (20/30/40 GHz) for higher communications capacity
- Satellite signal regeneration
- Large space platforms supporting several smaller satellites
- Advanced digital signal processing, improved data compression, and modulation techniques
- Improved time- and frequency-division multiple-access techniques

There also will be a trend toward integrating satellites, cable, fiber optics, and microwave into coherent networks with distributed control.

The shortage of energy is already changing our way of life. Rates for satellite communications will drop compared to the prices of paper, gasoline and transportation. The rising costs and inconvenience of business travel will be a strong incentive to substitute telecommunications for some travel. In addition, telephony will increase at a rapid rate. The percentage of telephony carried over the satellite will increase with the interconnection of business networks. Also, data communications, which is still in its infancy, will expand very rapidly as new devices are developed that use existing and projected transponder capacity. This has led one expert to forecast a worldwide need for 120 transponders for data, 1,000 transponders for voice, and 8,000 transponders for video conferencing by 1995. By year 2000, perhaps we will be beyond the semaphore and smoke signal stage in communications.



Introduction to Satellite Communications

J. Searcy Hollis

General

This symposium is directed largely toward the technical aspects of satellite communications and is attended by many who are experts in the field. This paper is to introduce the basics of satellite communications to readers who are not experts without going too deeply into technical details. The reader may be helped by reference to the Glossary in the Appendix for definitions of terms which are not defined. Detailed treatments of most of the topics discussed will be found in the other papers.

Almost all satellite communications in the free world are by satellites located in the "geosynchronous orbit". This is the circular orbit which lies at a height of about 22,300 miles above the earth in the plane of the equator, illustrated in Figure 1.

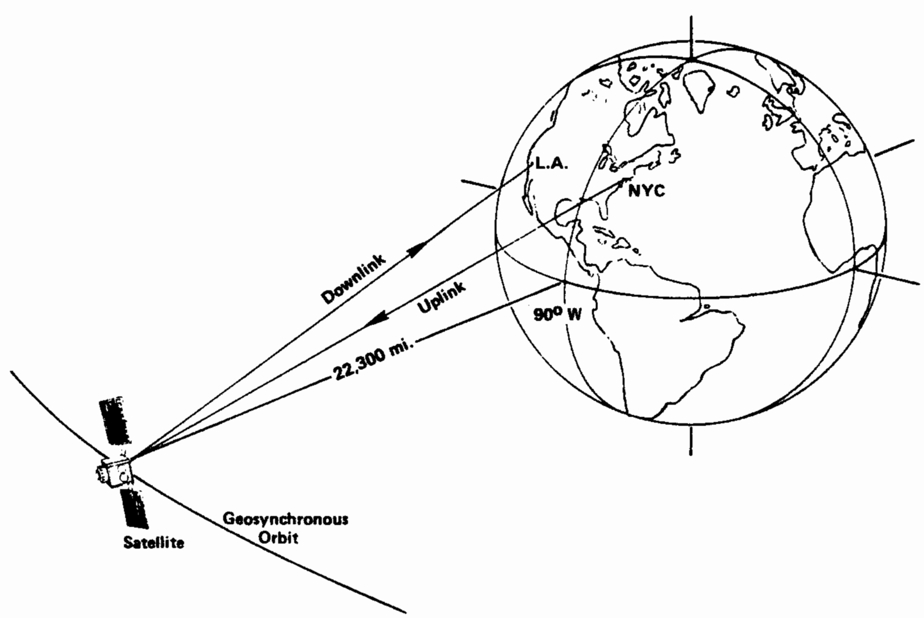


Figure 1. Satellite in Geosynchronous Orbit

Satellites in the geosynchronous orbit rotate from west to east. They appear fixed in space to earth stations on the ground because they orbit in synchronism with the earth's rotation. A satellite which is lower orbits faster; one that is higher orbits slower. Compare the 90-minute orbit of the Space Shuttle, which operates roughly 150 miles above the earth, with the 28-day orbit of the moon.

Because of the 22,300-mile height of the geosynchronous orbit, satellites in it have direct lines of sight to almost half the earth, as shown in Figure 1.

Except for small regions near the North Pole and the South Pole, widely separated earth stations can be seen from a single satellite. For example, Prudhoe Bay at the northern end of the Alaskan pipeline and villages farther north in Canada have television reception and voice communications with the world by satellite.

The INTELSAT international satellite communications network is shown in Figure 2, and the high population of satellites located on or authorized for the geosynchronous arc available to the United States, Canada and other countries of the Americas is illustrated in Figure 3.

Geosynchronous satellites are in effect unmanned relay stations. Communication by satellites was made possible by parallel advances in space technology and electronics. Arthur C. Clarke, the noted British scientist and science fiction writer, proposed relay stations in geosynchronous orbit for satellite communications in 1945. He proposed manned relay stations because the short life expectancies of vacuum tubes ruled out unmanned satellites.

Synchronous orbits had not been achieved in 1945. Indeed, a satellite was not put even into a nonsynchronous orbit until 1957 with Sputnik. and launching a satellite into synchronous orbit did not come until 1963. Nevertheless, advances in rocketry during World War II indicated to a man of Clarke's imagination that heavy payloads could be launched into synchronous orbit with sufficient research effort.

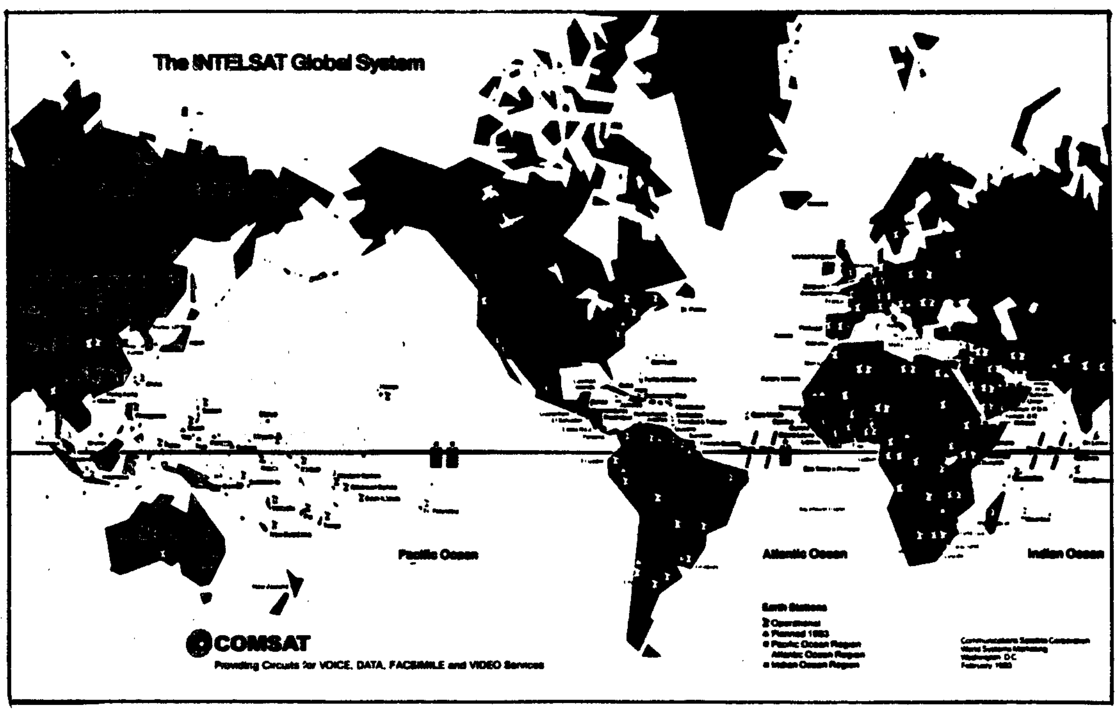


Figure 2. INTELSAT International Satellite Communications Network

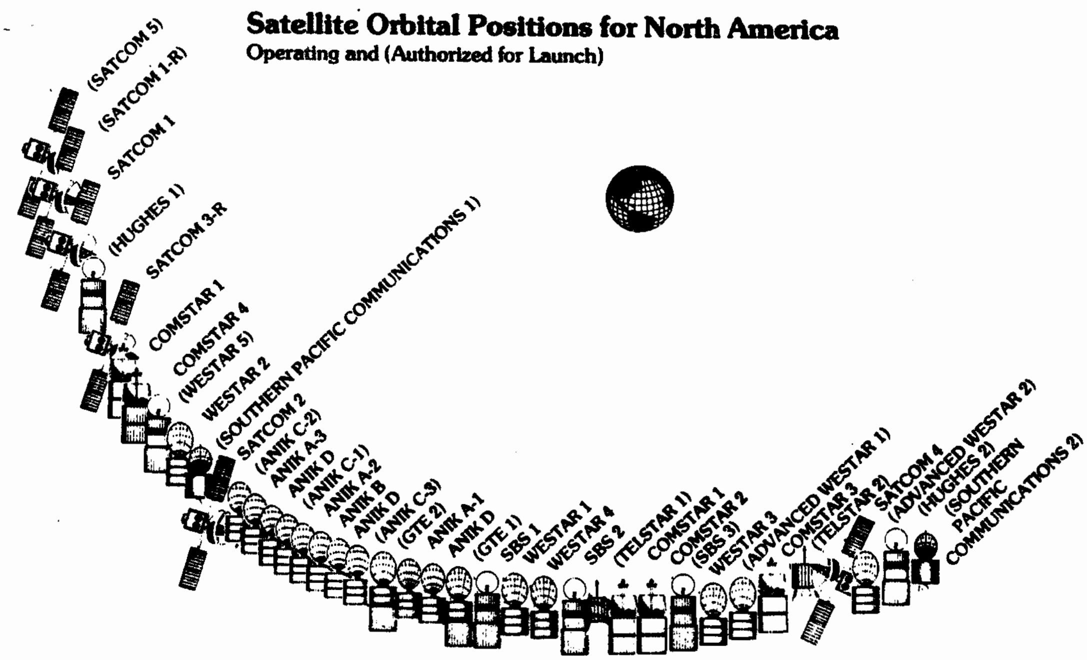


Figure 3. Satellite Population of Geosynchronous Arc Available to the United States

On the other hand, in spite of Clarke's vision, satellite communications as we know it today would not be possible without transistors, which are small, use little power, and have extremely long life expectancies. The invention of the transistor by scientists of Bell Telephone Laboratories in 1947 was one of the key factors that let the United States land men on the moon. It and other advances made relatively lightweight, unmanned satellites possible and economically feasible.

Communication by satellite is completely different from that by long-distance radio. Long-distance communication at radio frequencies is possible because the "ionosphere", produced by bombardment of the upper atmosphere by the sun, usually acts as a mirror to reflect certain radio waves back to earth.

As the frequency increases, a critical point is reached where the ionosphere ceases to act as a reflector, letting the waves pass through into space. We know, of course, that long-distance transmission of television signals, which are in the radio frequency range above 54 MHz. is not usually possible.

Communications satellites operate at much higher frequencies, in the microwave range, as shown in Figure 4. Here the ionosphere is always virtually transparent regardless of sunspot activity or time of day, permitting continuous, almost loss-free transmission to and from satellites in orbit.

Most current communications satellites operate in the split frequency band which is designated "C-Band" in Figure 4. The uplink band from the earth station to the satellite is at 6 GHz: the downlink band is at 4 GHZ. Taken together they are called variously "the 6/4 GHz" band, the "4/6 GHz" band or C-band.

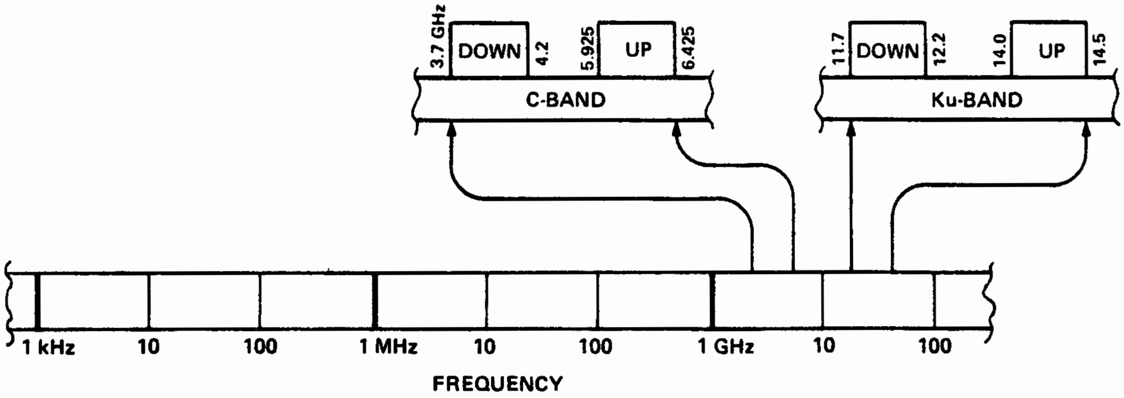


Figure 4. Frequencies in the Microwave Range of Communications Satellites

Our discussions here will be directed initially to the 6/4 band, although higher frequencies, especially the 14/12 GHz band (also called Ku-band) are coming into use as more spectrum is needed to handle the ever-increasing flood of information.

Stationkeeping

It was stated that the synchronous satellite appears "stationary" in space. Actually, a synchronous satellite is never perfectly stationary, because a number of forces including the pull of the sun and moon perturb its orbit. If left alone, it would eventually drift out of position. To overcome this, the position of the satellite is continuously monitored by an earth station, called a TT&C (telemetry, tracking and command) station, and small jets of a propellant such as hydrazine are used to keep it in position within a "station-keeping" box.

The station-keeping box is typically a square which is ±0.1 degree on each side and is oriented with the sides parallel with and perpendicular to the orbital plane.

Sufficient hydrazine must be carried on board the satellite to last for its predicted life, which is usually from 7 to 10 years. The TT&C stations must be highly accurate to make optimum use of the on-board propellant.

Footprint

The transmitting and receiving antennas on the satellite are designed to cover only desired regions of the earth's surface. This has several purposes. It concentrates the power radiated from the satellite into desired directions, increases the sensitivity of its receiving antennas, and helps prevent interference with signals from other satellites.

The part of the earth's surface covered by a satellite is called the satellite's "footprint". The footprint may cover one or more relatively localized regions or a complete hemisphere. A typical footprint is shown in Figure 5. The footprint is, or course, not sharply defined. The 3 dB contour represents the half power level. Signal strengths tend to peak near the center of the footprint and roll off fairly steeply past the 3 dB contour.

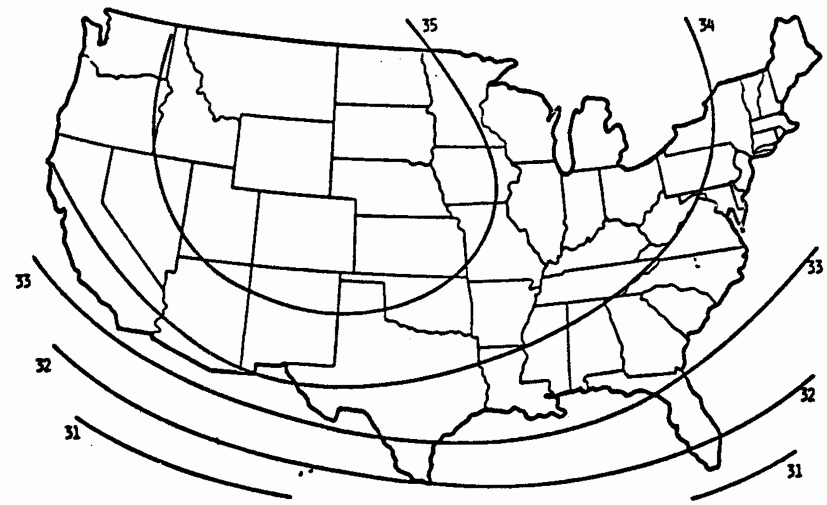


Figure 5. Satellite "Footprint"

The power levels radiated by typical satellites which operate in the 6/4 GHz band are of the order of a few watts. Those operating in the 14/12 GHz band usually radiate somewhat more power for reasons that will be explained later.

The effective power radiated toward the footprint (called the effective isotropic radiated power or EIRP) is typically increased to between 2,000 and 4,000 watts (33 to 36 dBW) by the beaming action (gain) of the antenna.

All the power radiated by the satellite is supplied by solar panels, which convert sunlight directly to electricity.

The solar panels have to face the sun to be effective, while the antennas have to be directed to keep the footprint in place. There are two basic types of satellites, based on the method of stabilization and control of the direction of the solar panels. These are the "body stabilized" type and the "spin stabilized" type, shown in Figure 6.

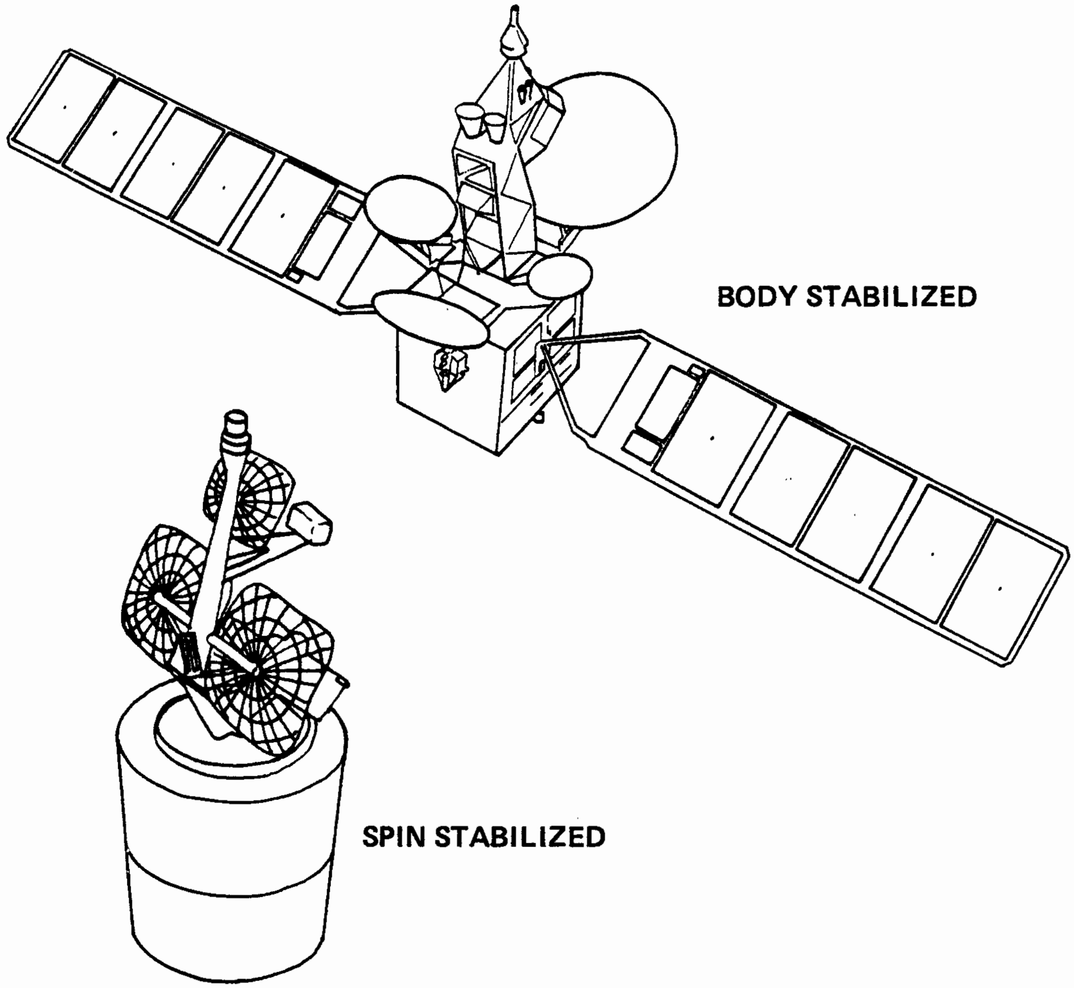


Figure 6. "Body Stabilized" and "Spin Stabilized" Satellites

The body-stabilized satellite is designed to keep the antennas pointing correctly while pointing all of the solar panels toward the sun. The spin-stabilized satellite is cylindrical and has its solar cells mounted around its periphery. The body spins about its axis for stabilization while the antennas are "despun" to point independently toward the earth. In this case about one-third of the cells effectively face the sun at one time.

Batteries are used in almost all satellites to take care of solar panel outages during times when the satellite is eclipsed by the earth. Monitoring and control of the attitude of the satellite is the responsibility of the TT&C station.

Polarization

Electromagnetic waves and antennas are always "polarized" in some manner. The polarization may be linear, circular or elliptical. For our purposes in this paper we will dismiss elliptical polarizations as being nonideal cases which are intended to be either linear or circular.

Linear polarizations and circular polarizations are illustrated in Figure 7. A linearly polarized antenna receives maximum power from an incident linearly polarized wave if the "tilt angles" of the wave and antenna polarizations are aligned in space as in Figure 7(a). The wave is then said to be "co-polarized" or "polarization matched".

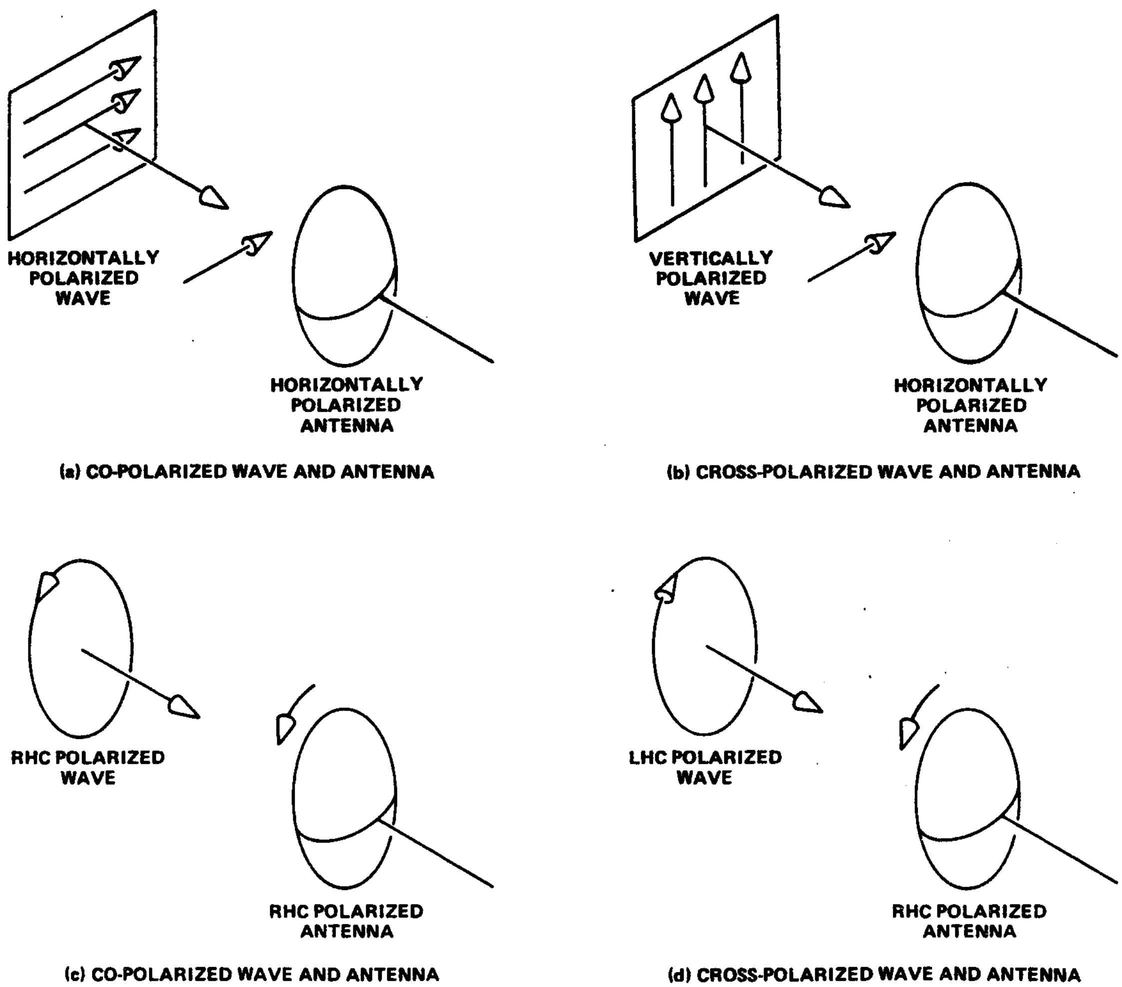


Figure 7. Linear and Circular Polarizations

As the tilt angle of the wave or antenna rotates from co-polarization, the received power decreases. When the tilt angles are 90 degrees apart as in Figure 7(b), the antenna is "cross polarized" to the wave and receives no power from it. The antenna and wave then have "orthogonal" polarizations. A given wave can have two orthogonal polarizations which exist simultaneously and carry different information without interference. It will be seen that this principle is used to increase the "information capacity" of satellites and of the geosynchronous orbit.

Circular polarizations have either right-hand (RHC) or left-hand (LHC) "senses". RHC and LHC polarizations are orthogonal. A circularly polarized satellite and a circularly polarized earth station are co-polarized if they have the same senses and are cross-polarized if they have opposite senses. The relative tilt angles of circularly polarized antennas and waves are of no consequence and are not even defined. This represents an advantage of circular polarization over linear polarization, since the tilt angle of the earth station does not have to be adjusted for a particular satellite. On the other hand, there are a number of trade-offs, expecially because circularly polarized antennas tend to cost more than linearly polarized ones. Most domestic satellites are linearly polarized while INTELSAT satellites are circularly polarized.

Satellite Information Capacity

One of the major reasons for the impact of satellites is their tremendous information carrying capacity. This is because of the large bandwidth available at microwave frequencies.

The information carried by any type of modulated RF or microwave "carrier" is contained in "sidebands" which spread out on each side of the carrier. Transmission of information at a high rate requires a large bandwidth to accommodate these sidebands.

A typical satellite has 24 transponders. Each transponder has a bandwidth of approximately 35 MHz and is capable of accommodating one high-quality television channel or about 2,000 voice-grade telephone channels. In contrast, the complete radio broadcast band has a bandwidth of only about 1 MHz.

As an example of a typical satellite, the transmit and receive frequency plans of an RCA Satcom satellite are shown in Figure 8. The numbered brackets represent each channel. The bandwidth of the channel is represented by the width of the bracket. The carrier frequency, which is shown above the channel number, is centered on each channel.

Note that the total bandwidth covered by the 24 transponders is 500 MHz. Squeezing 24 transponders into this amount of spectrum is accomplished by a process called "frequency reuse by polarization diversity", which we will call simply "frequency reuse".

Frequency reuse is implemented by staggering the microwave carriers of alternate transponders so that only sideband energy overlaps and by use of orthogonal polarizations.

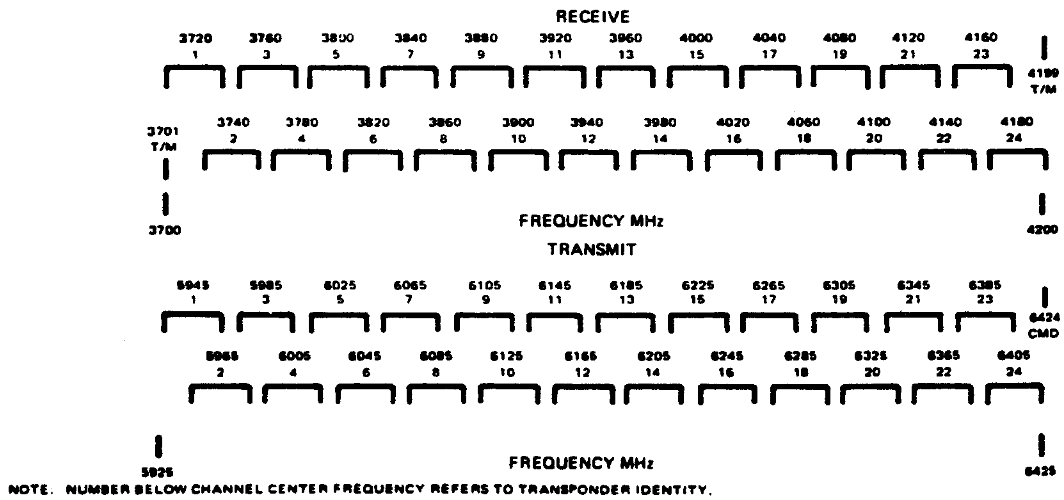


Figure 8. RCA Satcom Satellite Frequency Plans

The signals of alternate transponders in the frequency plan of Figure 8 are nominally orthogonal. If they were exactly orthogonal and the associated earth stations were ideal, there would be no interference caused by the overlapping sideband energy of adjacent transponders.

In practice, the polarizations of the antennas of the satellite and earth stations are not ideal. Some small amount of interference occurs, but the combination of nearly orthogonal polarizations and use of a staggered frequency plan provides for high quality transmission under almost all weather conditions. This permits 24 transponders in the same band that was used for 12 transponders in older satellites which do not employ frequency reuse, essentially doubling the information capacity of the satellite.

INTELSAT accomplishes frequency reuse by using right-hand (RHC) and left-hand (LHC) orthogonal polarizations on the alternate transponders.

Although the term frequency reuse by itself is used to mean frequency reuse by polarization diversity where the meaning is clear, it can also be applied to the reuse of available spectrum by other means, such as by using several spot beams on a satellite. INTELSAT V and certain of the other later-generation satellites make use of this technique. It will become more prevalent as satellites become larger and more complex.

Earth Station Antennas

One of the major advantages of using geostationary satellites is the simplicity of the earth stations which are used with them. An earth station has to have a relatively narrow beam to let it pick out a particular satellite and to increase its effectiveness in transmission and/or reception of signals. Since the satellite appears stationary, the complex electronics and drive mechanisms are eliminated which would be required to keep the beam on a moving satellite.

A number of different types and sizes of earth-station antennas are used in satellite communications depending on the application. High information capacity and high-quality link performance tend to demand larger antennas. The largest antennas routinely used in satellite communications are 30 meters in diameter. These are used in INTELSAT A stations and in certain extremely high performance domestic systems. The antenna shown in Figure 9 is 10 meters in diameter.

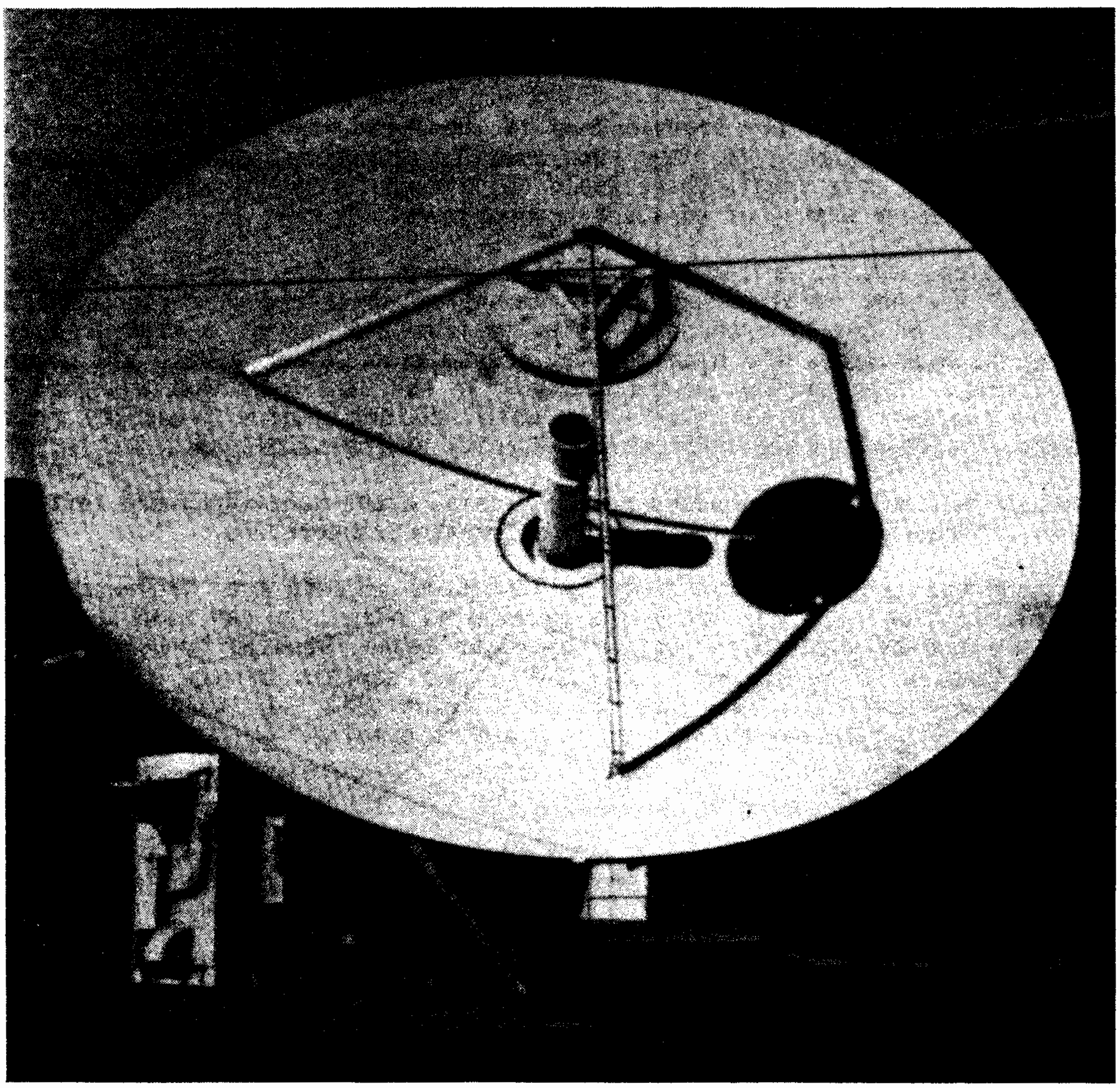


Figure 9. 10-Meter Antenna

At the low end of the scale, the smallest diameter antenna that can be used is determined by the spacing between satellites because the beamwidth of an antenna at a given frequency is essentially inversely proportional to its diameter.

At first glance it may seem that the geosynchronous orbit could yield an almost infinite information capacity by adding more satellites. This is not true, however, because the closer the spacing between satellites the narrower are the required beamwidths of the earth-station antennas.

The minimum orbital spacing between United States domestic satellites has been 4 degrees until recently. This spacing permits high-quality transmission of television signals, data and voice in the 6/4 GHZ band with antennas that are as small as 4.5 meters in diameter and usable signals with antennas as small as about 1.5 meters for certain applications.

Because of the pressure for more orbital capacity, the Federal Communications Commission has recently decided to ultimately reduce the minimum spacing between U.S. domestic satellites from 4 degrees to 2 degrees in the orbital arc between 55 degrees and 143 degrees west longitude. The closest spacing that exists at this writing is 3 degrees between the Galaxy I satellite and the SATCOM III-R satellite.

It will take some time for all the orbital slots to become filled, but decreasing the orbital spacing to 2 degrees will ultimately increase the inter-system interference to some extent for all types of service. Predictions of the resulting interference have been made, but the ultimate effect will not be known for a number of years.

Side Lobe Control

A communication antenna has a main beam as shown in Figure 10, but all antennas radiate some energy into unwanted directions or receive unwanted signals through "side lobes". Side-lobe energy of a transmitting earth station can interfere with other satellites which have orbital slots near the desired satellite and with terrestrial systems. Side lobes of a receiving earth station can receive interfering signals from other satellites and from terrestrial systems.

Well designed communication antennas have low side lobes. The FCC has lowered maximum allowable side-lobe levels of regulated earth station antennas as part of its program of decreasing the orbital spacing between satellites.

Noise and Sensitivity

"Noise" is the combination of disturbances that tends to obscure the information content of a signal. The sensitivity of a satellite communications system is limited by noise. Noise can enter a system from a number of sources, as shown in Figure 11.

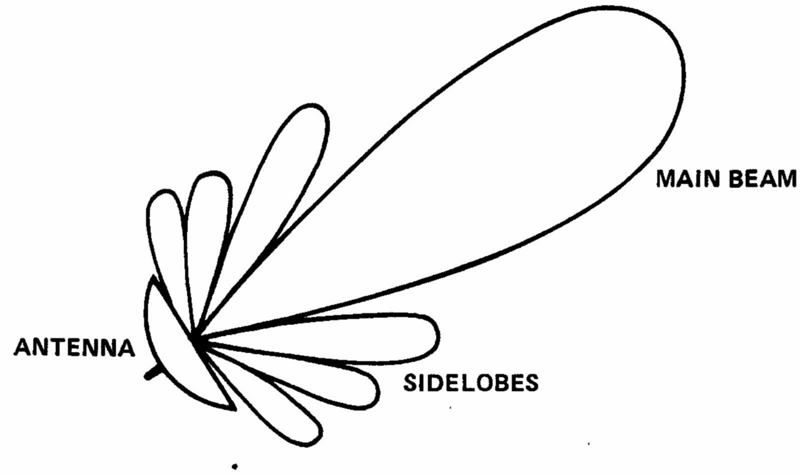


Figure 10. Typical antenna radiation pattern showing sidelobes

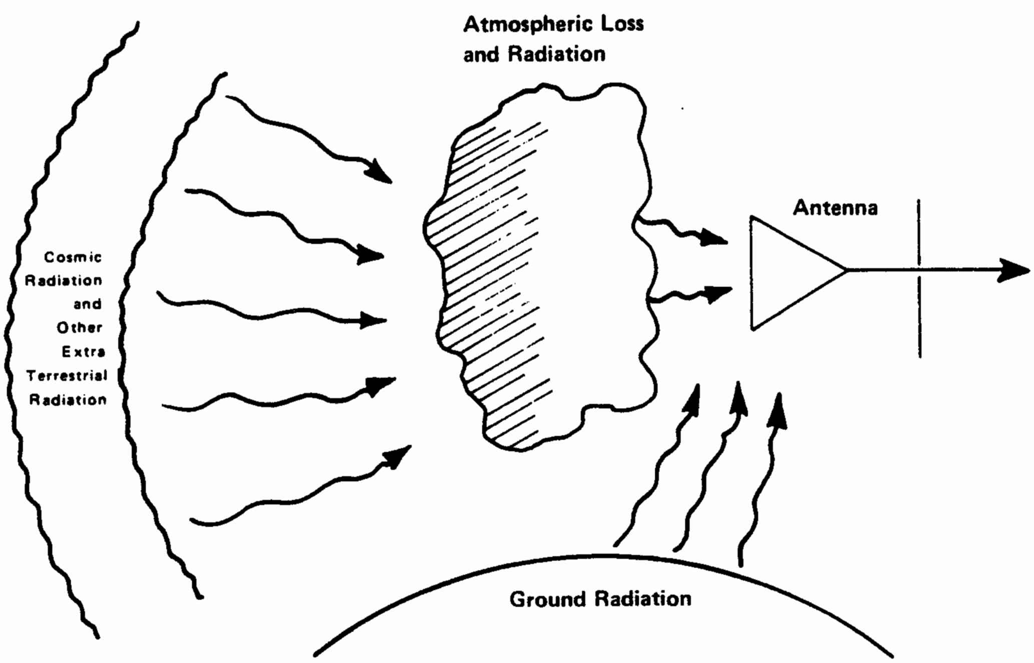


Figure 11. Noise Sources

The sensitivity of a receiving system is limited by its ability to discriminate against noise in favor of the desired signal. This is determined by the electrical size of the antenna and its ability to reject noise and the noise rejection of the earth-station receiver itself.

High power is radiated by an earth station to overcome the spreading loss of the signal in traveling the large distance to the satellite and to override the noise at the satellite receiver input.

Because the signal at the earth station from the satellite is very low and because large antennas are expensive, it is important for the earth-station receiver sensitivity to be as high as practical.

The most sensitive receivers use cryogenically cooled microwave amplifiers, but these amplifiers are expensive and are used only in the highest performance earth stations.

Relatively inexpensive uncooled field effect transistor amplifiers based on a semiconductor called gallium arsendide (GaAs FETs) have been developed which give adequate sensitivity for many systems. The low cost of GaAs FETs has been one of the major factors in the growth of satellite communications.

Modulation Formats and Access Techniques

The modulation formats employed for various satellite communication applications are determined by the requirements of the application. Video and voice signals are commonly transmitted in an "analog" format while computer data, for example, are transmitted in a "digital" format. Analog signals are electrical replicas of the information being transmitted. Digital signals are numerical codes which represent sampled analog signal levels or numerical values such as computer data.

Analog-to-digital converters are often used to transmit analog signals digitally. The digital signals are then reconverted to analog form at the receiving end of the system by digital-to-analog converters. Systems of this type are coming into increasing use as the costs of digital circuits decrease.

Video signals are likely to occupy a complete transponder, using frequency modulation (FM). Where many narrowband channels such as voice-grade circuits are required, the channels are combined by a technique called frequency division multiplexing (FDM) and used to frequency modulate a carrier (FDM/FM).

For full-transponder video or multiple-channel signals multiplexed on a single carrier, each transponder is accessed by a single earth station at a given time. For lower-capacity applications, a transponder will be shared by a number of earth stations, each of which may use a number of carrier frequencies. This technique is called frequency division multiple access (FDMA).

In one of the important applications of FDMA, a single voice-grade signal is transmitted on each carrier. This approach is designated single-channel-per-carrier (SCPC). It permits installation of stations which need only a limited capacity and permits easy addition of channels as required.

SCPC, using a technique called demand assignment multiple access (DAMA), permits an earth station to use a channel only as required, making the channel available to other earth stations when it is not needed. This greatly increases the use factor of a given transponder which is assigned to this service.

Digital modulation formats are becoming increasingly important as advances in digital technology continue. The cost of digital transmission is rapidly decreasing, and it is likely that in the future both analog and digital signals will be sent largely by digital systems.

There are a number of digital modulation formats. We will indicate a few. Popular formats vary the phase of a carrier or subcarrier in steps of 180 degrees (BPSK) or 90 degrees (QPSK). Time division multiplexing (TDM) is a technique in which a number of signals modulate a subcarrier sequentially in closely spaced time slots without interference. It is the digital equivalent of frequency division multiplex (FDM) in analog systems. A number of earth stations can access the same transponder by means of a controlled technique called time division multiple access (TDMA) to accomplish a function somewhat like that which FDMA accomplishes in analog systems.

The quality of any communications system is determined by the difference between the output signal and the input signal. In an analog system, the difference is measured by distortion and noise. In a digital system, it is measured by bit error rate (BER). Bit error rates of one error in 20,000 provide high-quality audio. Data transmission usually requires much lower error rates. Techniques such as forward error correction (FEC) can be used to decrease bit error rates by factors of 100,000 to a rate of one error in 10,000,000 bits.

Effect of Frequency on Satellite Communications Systems

Because of the virtually unlimited applications for satellite communications and the limited information capacity of the 6/4 GHz band, the higher frequency satellite bands, especially the 14/12 GHz band, will come into greater use in the future. It is therefore interesting to consider the effect of increasing frequency on the design and performance of satellite communications systems.

As the frequency increases, the major effects are (1) narrowing of the beam of an antenna of a given size, (2) increase in losses, and (3) increase in the required surface accuracy of the reflector. For a given satellite EIRP, the increase in frequency increases the earth station receiving antenna cost in at least three ways:

- by the tighter surface tolerance, which requires a more accurate and stiffer reflector,

- by the increased difficulty of pointing the antenna toward the satellite, which requires a costlier antenna mounting structure, and

- by the fact that the increase in atmospheric attenuation and the inherent increase in LNA noise force a larger antenna diameter, which acts to reinforce the difficulties associated with the first two factors.

The net result is that if earth station costs are to be kept low, the EIRP of the satellite must be greater at the higher frequency band so that smaller earth station antennas can be used. This increases the cost of the satellite because the increase in EIRP must come from either higher-power transponders or from footprints covering smaller areas.

The smaller footprints do not represent a disadvantage where a small area is to be covered, such as Japan, or a country in Europe. On the other hand, where a large area is to be covered such as the United States, the smaller footprints require multiple beams, each with some number of transponders determined by the level of service to be provided. The net effect is an increase in satellite solar-panel power requirements and in overall weight and complexity of the satellite. Typical system designs for low-cost receiving systems are based on four or five beams covering CONUS (continental USA).

In spite of the problems indicated above, a move to higher frequencies for new satellites is inevitable for the United States because of orbital crowding and use of the 6/4 GHz band by terrestrial and satellite systems. Systems are already being implemented for the 14/12 GHZ band. In fact, once sufficient cost is transferred to the satellite so that small antennas can be used and sufficient system information capacity is provided, many new applications such as direct broadcast satellite (DBS) open up. It is evident, however, that satellite systems at 6/4 GHz are here to stay because of their inherently low cost and because the 6/4 GHz frequency band exists.

Summary

A brief, semi-technical description has been presented of some of the elements of communication by geosynchronous satellites. Detailed presentations will be found in the succeeding papers of this notebook and in references listed in the bibliographies of the various papers.

The Broadband Cable/Satellite Connection for Business Communications

Dr. Guy W. Beakley and Alex B. Best

Introduction

The primary factor in the growth of cable television in the 1970's was the economical delivery of unique television programming by satellite-to-cable distribution systems. The success of this venture established a connection between cable and satellites for entertainment. In this chapter the connection between cable and satellites is explored for business communications. The cable/satellite connection for business offers growth potential for the 1980's expected to exceed that of the entertainment connection of the 1970's.

Information

Much has been said about the evolution of the United States from an industrial society to an information society. Just as the number of people working in industry started to outnumber the people working in agriculture in the early 1900's, the people working in information occupations now outnumber those working in industrial occupations. We have indeed encountered "The Third Wave."[1]

Information and the way it is used will be the key strategic variable in many businesses. For this to happen, effective information transfer can no longer be constrained by a communication system designed for voice transmission.

With the advent of the communications satellite, large amounts of information can be easily sent from one city to another. The satellite carriers have made a significant investment in earth stations for intercity distribution of information. The problem lies in distribution of this information to the local user. The cable that is now being laid for distribution of television can be used for a substantial part of this local distribution. The linking of satellite earth stations with cable television distribution systems offers many new opportunities as the information society develops.

Business Communications Needs

A number of large businesses are essentially information-based; included are banks, insurance companies, investment companies, and stock exchanges. Their business depends on gathering and assimilating large amounts of data. Office automation will allow the information to be processed and stored electronically with only summary information being recorded on paper. Communication and transfer of this information can be accomplished by the interconnected cable and satellite systems described in this paper.

1. Alvin Toffler, The Third Wave, William Morrow, New York, 1980.

Although productivity improvements have occurred in the agriculture, manufacturing and service industries, productivity improvement in the office has been slow. However, improvements are now being made to change this situation. Communicating word processors, facsimile devices, computer-to-computer links, voice messages stored in digital format, integrated voice and data PBX, and intrafacility communication networks are now available to upgrade office systems. A sophisticated business communication system allows charts, budgets, last minute schedule changes, press releases, contracts, and photographs to be transmitted quickly and reliably. Many people will be able to work at home connected by cable to their office. Executives attending out-of-town meetings and salespeople on the road will be able to have immediate and accurate access to information back at the home office. In the future, a person will be able to substitute teleconferencing for some business travel. The possibilities seem unlimited.

With office automation moving ahead, satellite common carriers have recognized business communications as a viable market. These satellite carriers are shaping intercity business communications with their concept of shared earth stations. Digital satellite carriers offer business customers a capability to transmit vast amounts of data, voice, electronic mail and other business signals from the earth station up to the satellite and back down to another earth station. Cable is a logical choice of a medium to distribute this data from the earth station to the business in each city.

Business Communications Services

Satellite and cable systems offer increased communications capabilities for text, facsimile, data transmission, voice, visual aids, and video conferencing. In the past, text transmission has been handled principally by telex, which is slow and has a rudimentary character set. Telex is being replaced with communicating word processors that allow typewritten material to be transmitted from one machine to another at high speed. Much of the transmission and storage is accomplished electronically, thus creating tremendous communications needs. With a cable/satellite communications system, a secretary can type a letter, attach the appropriate electronic "address" and send a copy directly to another terminal at a distant facility.

Graphs, pictures and photographs do not lend themselves to character transmission and are better sent by facsimile. In order to obtain high resolution and transmit at a rate of one page per second, the bit rate needs to be of the order of 256 kb/s. This high data rate, which is available on the satellite, creates the need for a high capacity local distribution system such as cable can provide.

The next step beyond communicating by voice and still pictures is conferencing using full-motion video. Point-to-point video conferencing now makes economic sense. Many national sales meetings have been set up with regional salespeople coming to local auditoriums where earth stations are installed or have been temporarily set up. The salespeople view the program material on large television screens and respond to the presenter by phone.

Point-to-point teleconferencing is very costly when using full-motion video because of the large bandwidth required. Analog video requires one-quarter to one full transponder of satellite. The use of this much capacity for a normal business meeting would be excessively expensive, hence there is a need to digitize the information and reduce the data.

NTSC color video can be digitized at the rate of about 90 Mb/s with no visible degradation. Removing redundancy allows a video signal to be sent at 20 Mb/s with insignificant loss in video quality. When the motion is limited and the camera is fixed, it is possible to reduce the bit rate to 6 Mb/s with limited loss in quality. Current digital technology has reduced the bit rate to 1.5 Mb/s while maintaining reasonable cost. As new devices for video coding are developed and as the cost of travel increases, business teleconferencing will become commonplace.

Consider next data transmission. Large computers can manipulate data at rates of one Mb/s or faster. Efficient resource sharing, computer backup, file protection, core diagnosis and other factors make it necessary for distant computers to communicate with each other. Transmission at Mb/s rates is required for computer communications in the future.

Since the office of the future will be integrated electronically, a number of scenarios for linking the equipment and people become evident. One scenario sees the hub of the automated office being an integrated voice and data PBX. The PBX permits users to transmit, switch, and store voice and data. Another possibility has local networks connecting office machines, digital telephones, and intelligent terminals or work centers. These terminals process and display data, text, pictures, and graphs. Regardless of the method, there will be a need for rapid communication of information within the office and from office to office.

Satellite Business Communications

As we discussed in the paper "Overview of Commercial Satellite Communications, " satellite business carriers have made great strides in the intercity portion of satellite business communications. Carriers such as SBS, American Satellite, RCA, and Western Union have provided increasingly sophisticated business communications facilities for their customers. American Satellite through its satellite data exchange (SDX) service has specialized in providing voice and data communications to small earth stations located on end-user's premises. American Satellite now has about one hundred earth stations operating. RCA Communications is providing data, voice, facsimile, slow scan TV and teleprint service called "56 Plus." Western Union also offers a similar data service to its customers. Voice channels can be provided over these digital links by using phase code modulation (PCM) or continuous variable slope delta modulation (CVSD) encoding.

Communication of business data by satellite offers a number of advantages. The telephone system normally restricts data transmission to maximum speeds of 9.6 kb/s. However, data rates of 50 Mb/s can be transmitted by satellite.

The telephone network can be expected to produce on the average, one error per 100,000 (10^5) bits transmitted. Elaborate error checking and correcting schemes, which reduce efficiency, have been developed to allow the use of the telephone channel. Error rates of fewer than one error per 10,000,000 (10^7) bits are easily accomplished using satellite communications with error rates typically lower than 10^{-8} with simple error correction devices.

Broadband Business Communications

Having achieved a long distance network, the carriers are then faced with the need for local distribution to their customers. The options available are the telephone plant, broadband cable, and microwave links. Prior to the development of pro-competitive policies by the FCC, local distribution was dominated by the local telephone companies. With the settlement of the AT&T antitrust suit and the development of alternative local distribution technologies, a new business opportunity "telephone bypass" has emerged. This bypass industry has been dominated by Digital Terminations Services (DTS), which use private microwave, cable systems, and radio common carriers, which use cellular radio. The local telco's offer data distribution primarily by twisted pair, although some progressive companies are working with fiber optics, broadband cable, and cellular radio.

Independent of whether the local distribution supplier is a bypass company or the Telco, broadband cable has some advantages over the other technologies for completing the satellite communications link. Like the twisted pair telephone plant, broadband cable is a proven existing technology. However, broadband cable, like satellite offers excellent bit error rate performance in addition to being very cost effective for high data rates. When compared to microwave and radio, the inherent closed system (shielded cable) nature makes broadband cable less susceptible to external problems such as line of sight and RF interference. Also, broadband cable systems do not require the sophisticated control schemes used in cellular radio and private microwave DTS.

Overall, broadband cable is flexible, available, and cost effective as a local data distribution medium. Broadband cable meets the service needs for intracity data communications for both the telco and the private network industry.

Conclusion

Cable television systems can play an important part in the intracity distribution of voice, data, electronic mail, teleconferencing and other business communications. In only a few years, cable has gone from an auxiliary system for delivering television to areas with poor television reception to the most versatile and economical means for mass distribution of many channels of video entertainment. The next equally dramatic step will involve cable operators finding new customers for using the cable to distribute business communications within the city. These customers will include satellite communications carriers, large corporations, financial institutions, municipal agencies, hospitals, and university complexes.

Two-Way Broadband Communications

J. Mauney/D.H. Slim

The Community Antenna Television (CATV) system had its inception in the early fifties. As originally conceived, the CATV system simply provided basic limited channel TV service to restricted off-the-air reception areas. The complete CATV system (Figure 1) commonly included three basic parts: (1) a master antenna system (tower, preamps and appropriate antennas); (2) a head-end system (single-channel processors and modulators); and (3) a broadband distribution system (trunk and feeder). Initially, the headend and broadband distribution systems provided for one-way only transmission; however, as operating techniques developed and equipment technology advanced, operators promoted and found interest in two-way transmission. The intent of this paper is to familiarize the reader with the types of two-way systems in use and the system components used to operationally construct a two-way broadband RF distribution system.

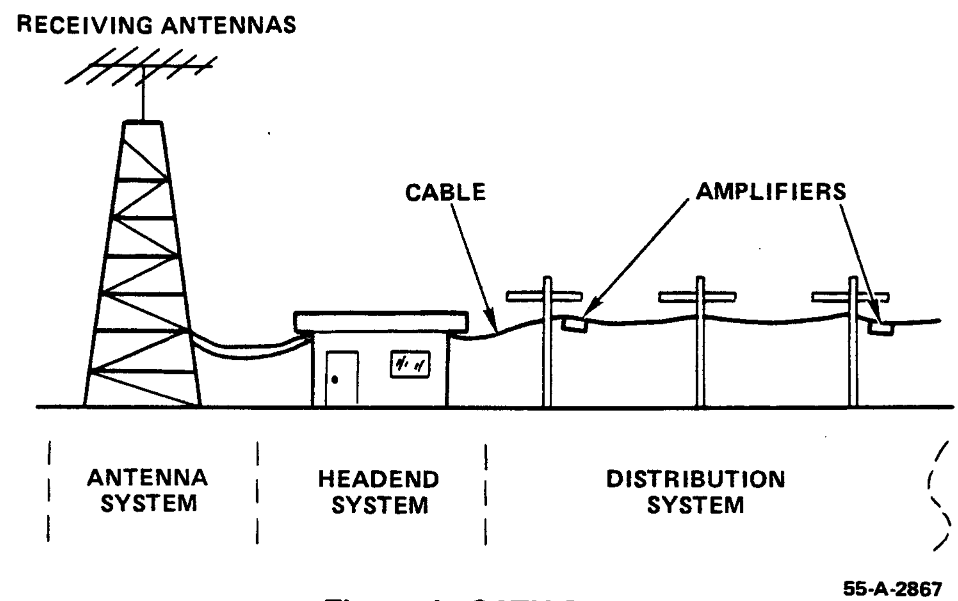


Figure 1. CATV System

Two-Way History and General Information

Presently, there are three specific types of two-way broadband RF distribution systems in use. The generic name for a two-way RF distribution system is split-band system; however, specific names are more commonly used for the three system types noted above. These systems are called sub-split, mid-split and high-split systems where the specific name refers to the frequency band division for forward and reverse transmission. Although there are differences between the various equipment manufacturers as to forward and reverse frequency band edges, in general, the bands occupy approximately the same relative frequency blocks in the spectrum and loosely fall under the sub-split, mid-split and high-split classifications.[1] Table 1 shows Scientific-Atlanta's sub-split, mid-split and high-split forward and reverse frequency band allocation.

System Type	Forward Frequency (MHz)	Reverse Frequency (MHz)
Sub-split	54 - X	5-30
Mid-split	174 - X	5-108
High-split	234 - X	5-174

Table 1. Two-Way Split Band Frequency Allocation

Of the three two-way systems noted above, the sub-split type is by far the most common in operation. The sub-split system was cable TV's first attempt at two-way operation and was developed initially because it easily fit into the frequency plan of the CATV system as it was originally conceived.

The original CATV subscriber system forward frequency band was selected based on a need to accommodate standard off-the-air channel frequencies and the existing channel select capability of the television receiver. The original forward subscriber system, therefore, began at channel 2 (54 MHz) and extended through channel 13 (216 MHz) excluding the presently defined mid-band frequencies from 88 to 174 MHz.[2] As knowledge of system operation developed and equipment technology advanced, operators began considering additional channel capacity above the 12 channels initially offered. Equipment manufacturers responded to the demand for additional subscriber system forward transmission bandwidth and extended the original upper 216 MHz limit to the present practical limit of 440 MHz.

Enterprising system operators looking for methods of increasing revenue began to consider the prospects of not only a link to the subscriber, but a link from the subscriber back to a central facility. Since the frequency spectrum below channel 2 was vacant, it seemed an ideal slot to provide the return link from the subscriber. With the standard forward system already constructed and the cable itself bidirectional and able to support information flow in either direction, it seemed logical that by adding a reverse signal

[1]To alleviate confusion as to exact frequency bands, one manufacturer has suggested that the more well-established sub-split term be retained, but mid-split and high-split be dropped in favor of the generic split-band classification.

[2]Originally, CATV repeater amplifiers used single-ended circuitry which resulted in a second-order distortion build-up in the mid-band frequency spectrum. Push-pull circuitry solved this problem and opened up the mid-band spectrum for added channel capacity.

path around each forward repeater amplifier, the existing system without major modification would support two-way transmission. CATV equipment manufacturers responded to this idea and engineered the proposed sub-band equipment and system; these became known as sub-split equipment and system, respectively. Immediately, sub-split reverse upgradeable equipment became a necessity, and few systems built since the early seventies are without sub-split reverse capability.

As applications for two-way transmission evolved, the need for a second type of two-way broadband RF distribution system developed. This system was to provide equal bandwidth in both forward and reverse directions and be used for interconnection between multiple common institutions. Because of the institutional application and the equal bandwidth requirement, the system was called interchangeably an institutional or mid-split system. Equipment developed for institutional system application was and is most commonly called mid-split equipment.

As originally conceived, the institutional system was to have its own separate cable and was to shadow the main subscriber cable. In this dual cable configuration, the standard subscriber cable became known as the "A" cable system, and the institutional cable as the "B" cable system.[3] Also, it was originally proposed that the sub-split reverse from the "A" cable system be added into the "B" cable system so that only a single common reverse path back to the headend would be required.[4] This was accomplished through a "seventh" port connection on the "A" and "B" cable repeater amplifier station housings. However, in recent years this common A/B cable interconnection has been questioned due to the potential of "B" cable reverse contamination from stray signal ingress into the "A" cable sub-split subscriber system. Sub-split reverse ingress will be discussed in more detail later. Block diagrams showing a basic two-way amplifier station, the standard sub-split subscriber system and the dual cable system are shown in Figures 2, 3 and 4, respectively.

When mid-split equipment was originally developed, the upper frequency limit for forward amplifiers was approximately 270 MHz, and the lower frequency limit for reverse amplifiers was 5 MHz. Allowing for a guard band (diplex filter high-pass/low-pass crossover region) between forward and reverse, the total 5 to 270 MHz band was approximately equally divided for two-way operation. However, since the initial development of mid-split, the forward amplifier upper frequency limit has stepped upwards from 270 to 300, 330, 360, 400, and finally, to the present limit of 450 MHz. As a result, the

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[3]The "B" cable designation no longer strictly implies institutional system operation. There are dual cable systems presently in operation where both cables are subscriber systems. In these dual cable subscriber systems, the cables are also identified as "A" and "B" cable.

[4]The concept of adding the "A" cable sub-split reverse into the "B" cable mid-split reverse at each main repeater station was orignally proposed to minimize the additive effects of "A" cable split-band filter group delay.

term mid-split no longer strictly implies equally divided frequency bands for forward and reverse. Although two-way mid-split and dual cable systems were much talked about during the early and mid-70's, not much demand was created and not many systems of this type were constructed.

Interest in the third type of two-way system grew from the 1980 development of the 400 MHz expanded bandwidth subscriber system. As 400 MHz technology advanced, system operators began looking for services that would interest

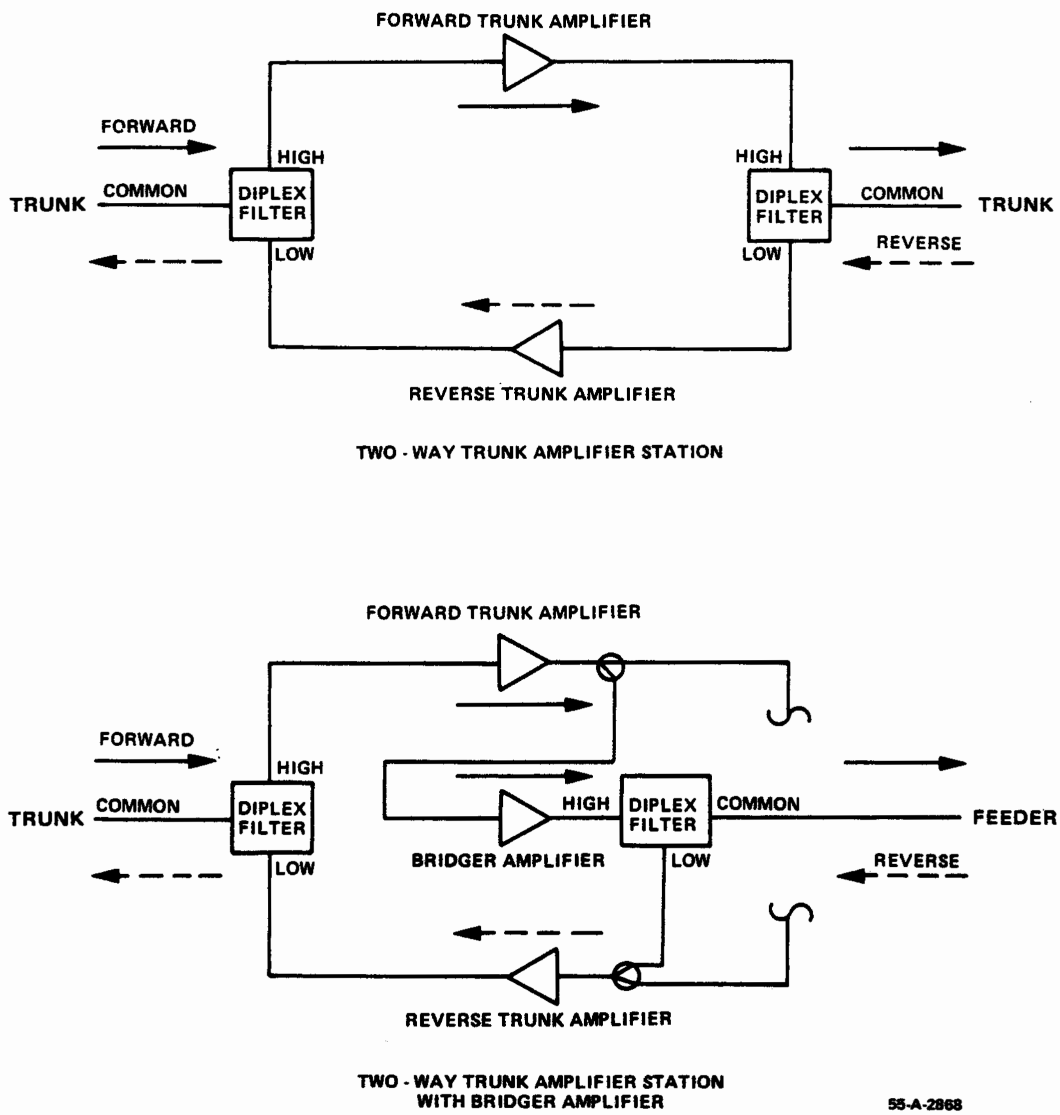

Figure 2. Basic Two-Way Trunk Amplifier Station

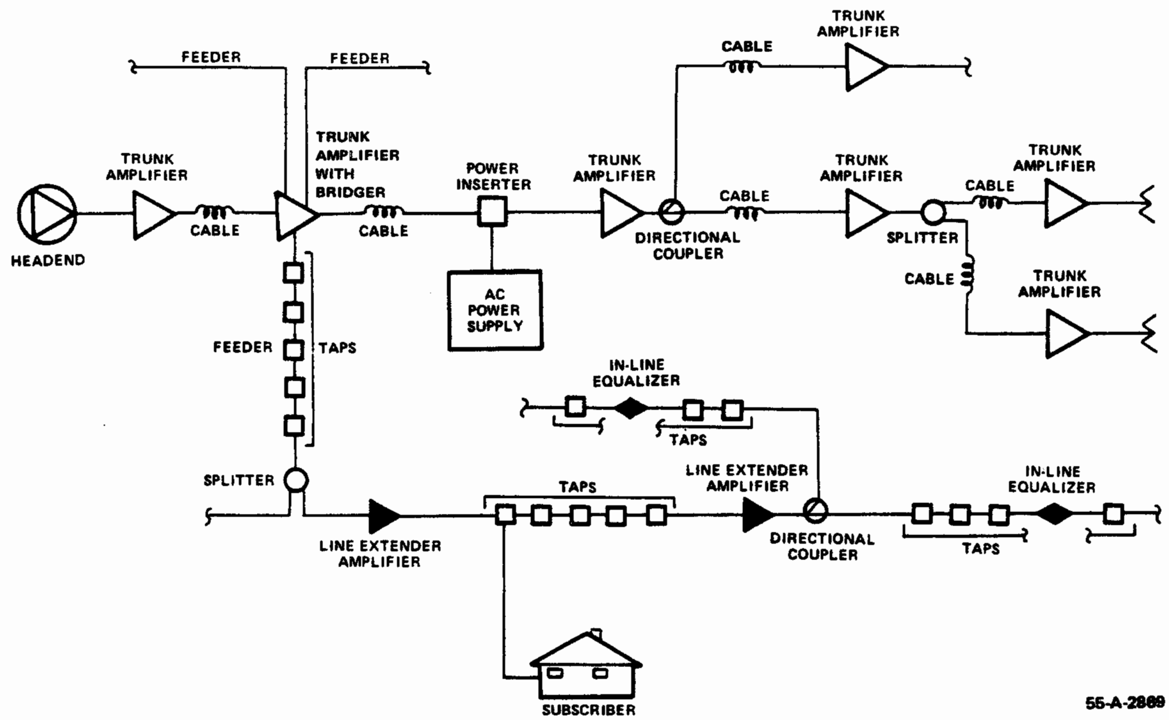

Figure 3. Sub-Split Subscriber System

franchising authorities and at the same time provide additional sources of revenue. As a result renewed interest in the institutional system began, and at the request of system operators, equipment manufacturers began to dust off mid-split and at the same time examine a new equal bandwidth split-band system for 400 MHz. Since mid-split was still commonly used to describe split-band systems above 30 MHz but below 300 MHz, the name given the 400 MHz equal bandwidth split-band system was high-split. However, as upper frequency expansion continued, the high-split term soon lost its original equal bandwidth connotation and, consequently, today no longer strictly implies equally divided frequency bands for forward and reverse. However, when readily available, it is expected that high-split will become the new standard for two-way institutional split-band RF distribution networks and will predominate in that role at least until further forward bandwidth extension occurs.

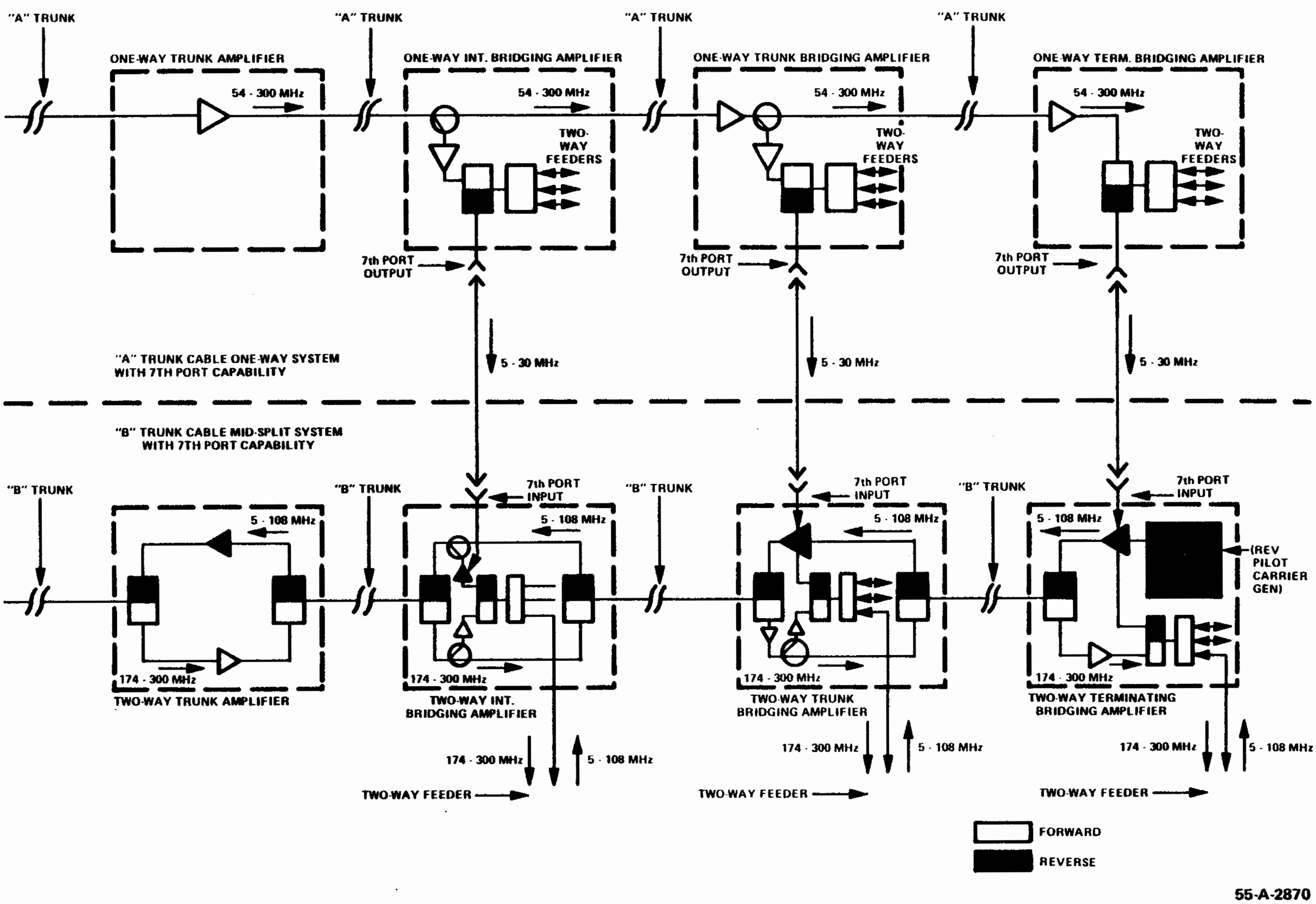

Figure 4. Dual Cable System

The Broadband Cable Distribution System

CATV RF broadband distribution systems typically fall under one of two classifications. One is called a subscriber system and is used to provide services to the general public. The subscriber system may or may not include sub-split reverse. The other is called an institutional system and includes reverse and connects banks, schools, government facilities and the like. The application of the institutional system is now expanding in the area of supplying the data communications links discussed in paper 1-2, Broadband/Satellite Connection for Business. The true institutional system always includes a reverse system with a frequency bandwidth greater than that of sub-split reverse. Occasionally, the individual functions are inter-mixed with the subscriber system providing a two-way link between institutions and the institutional system providing services to the general public. However, by historical definition, each still retains its identity in that the subscriber system forward frequency band always begins at 54 MHz and may or may not include sub-split reverse, while the institutional system forward frequency begins above 100 MHz and always includes a broadband reverse system.

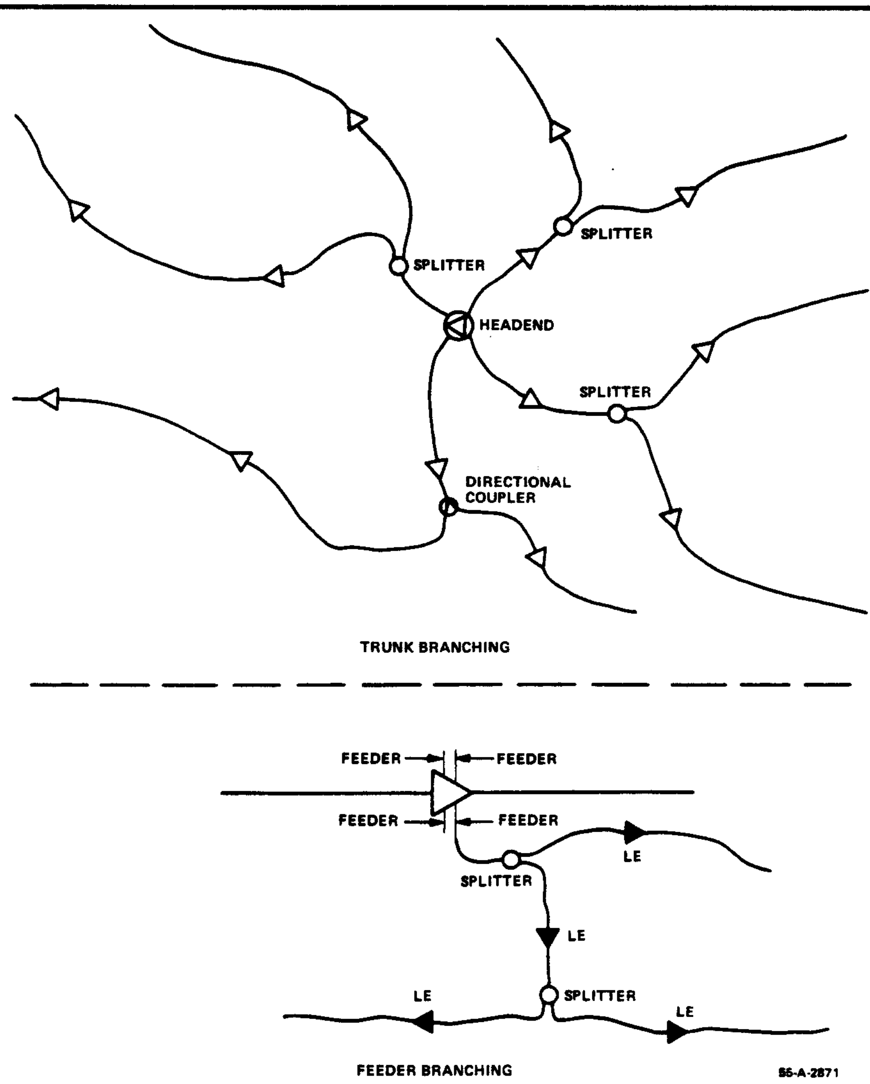

Figure 5. Trunk and Feeder Branching

Both the subscriber system and the institutional system are constructed in a similar fashion and typically include a trunk system and feeder system. The trunk system is the primary transportation system and, as such, branches in tree fashion throughout the area to be cabled. The feeder system branches from each trunk station in a similar tree-like manner and serves as the connecting link between the trunk system and the end user (subscriber or institutional). In some instances, the institutional trunk is routed directly to the end user and the feeder system is not included. The trunk and feeder system is shown in Figure 3 and trunk and feeder branching in Figure 5.

Large-diameter coaxial cable, repeater amplifiers, three-port power dividers (splitters and directional couplers), ac power supplies and ac power inserters are the main component parts used to construct the trunk system. A single series trunk path called a cascade can be miles in length and can contain many trunk-class repeater amplifiers. Consequently, individual trunk amplifiers must be high-quality, high-reliability amplifiers that exhibit low levels of noise and non-linear distortion. Power dividers having equal and unequal loss characteristics are used to branch the trunk system. The splitter or equal loss power divider is used at branch junctions where equal length trunk branches are required; the directional coupler or unequal loss power divider is used at branch junctions where unequal-length trunk branches are required. Careful system design and efficient use of the family of power dividers has a direct effect on the total number of cascaded trunk amplifiers and, consequently, on overall system performance. The trunk amplifier stations (also feeder amplifier stations) are ac powered through the same coaxial cable that carries the RF signal. Pole-mounted 60V ac power supplies are located at intervals throughout the system and are connected to the coaxial cable through power inserters. A block diagram showing typical trunk system operational construction is shown in Figure 6.

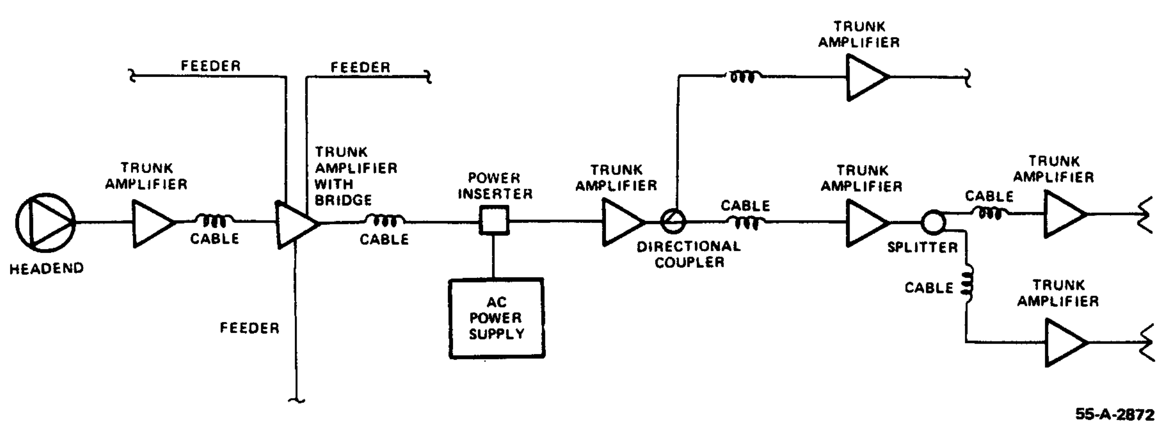

Figure 6. Trunk System Operational Construction

Small-diameter coaxial cable, repeater amplifiers, three-port power dividers (splitters and directional couplers), directional taps, and in-line equalizers are the main component parts used to construct the feeder system. Unlike the trunk, a single series feeder path or cascade is relatively short and typically contains not more than three line extender class repeater amplifiers. Consequently, individual feeder amplifiers do not require the same degree of sophisticated level control and frequency response circuitry characteristic of the individual trunk amplifier. Feeder amplifiers do, however, require the same high quality, high reliability and low levels of noise and non-linear distortion characteristic of the trunk amplifier. The first feeder amplifier is actually located within the trunk amplifier station housing and is called a bridging amplifier. The bridging amplifier receives signal from the trunk, raises the level and provides the source signal for up to four individual trunk station feeder cables. In addition, the bridging amplifier provides feeder to trunk isolation to minimize possible spurious signal leakage into the high-quality trunk transportation system. Feeder cables leaving the trunk are tapped with directional taps at regular intervals to provide connections for subscriber access to the system. Splitters and directional couplers are used to branch the feeder system just as described for the trunk. In-line equalizers are spliced into feeder cable spans to adjust for accumulated differences in signal level that occur due to the frequency-dependent loss variation of coaxial cable. Since each trunk station supports its own independent feeder system, there are generally as many independent feeder systems within a cable distribution system as there are trunk stations. Feeder system operational construction is shown in Figure 7.

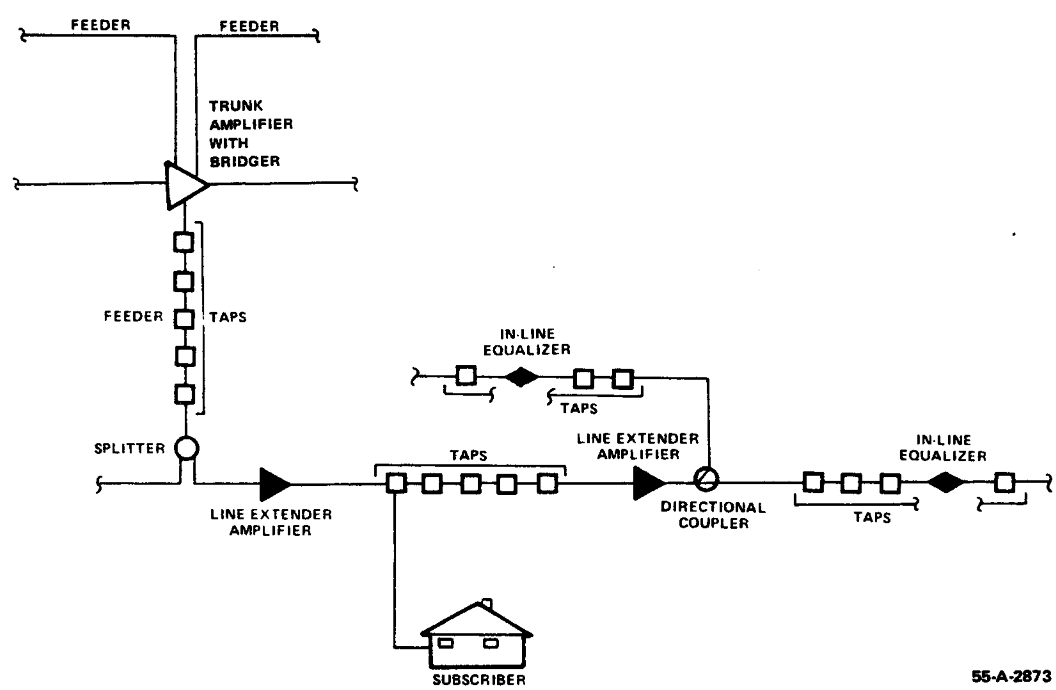

Figure 7. Feeder System Operational Construction

Subscriber and institutional trunk systems are constructed in exactly the same fashion and, except for trunk amplifier station diplex filters and reverse amplifiers, use exactly the same component parts. Subscriber and institutional feeder systems are, however, not always exactly similar in that the institutional feeder system does not typically provide connections for the general public. Consequently, the institutional feeder system may not include large quantities of directional taps and does not include sub-split only, in-line equalizers. In fact, in many cases, the institutional feeder system simply consists of the bridging amplifier and direct cable connection to the user.

In recent years, all components used to construct both the subscriber and institutional trunk and feeder system have been broadband bidirectional.[5] Consequently, for these systems to support reverse, it is only necessary to provide proper split-band diplex filters and reverse amplifier modules for each of the trunk and line extender stations. As it relates to components this is true; however, a proper working reverse system is guaranteed only if a separate reverse system design analysis is performed simultaneously with forward system design. Some sub-split systems that were designed for forward-only operation have encountered problems when activating the sub-split reverse upgradeable option.

Equipment for two-way broadband RF distribution systems has been available to the cable TV industry for almost a decade. Moreover, many of the proposed products that would utilize two-way transmission are available and have been successfully operating in two-way cable system for years. As a result, many of the initial problems have been eliminated, and much has been learned about the operational requirements and limitations associated with these types of systems. One problem long fought and still present, is spurious signal ingress into the sub-split.subscriber reverse system. Two factors have made this problem an especially difficult one to solve. One factor is the sub-split reverse frequency band itself and the infinitely large numbers and high levels of non-cable related 5 to 30 MHz signals always present in the environment. The other factor is related to subscriber system architecture and the potential problems due to loss of shielding integrity at the many subscriber connection points within the system.

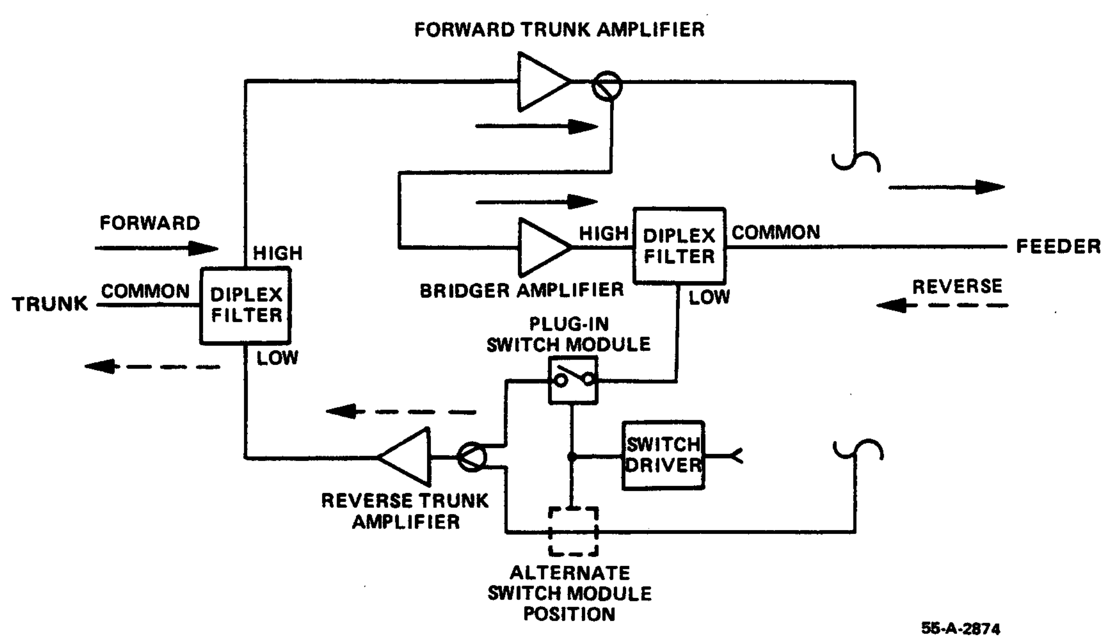

Figure 8. Two-Way Trunk Amplifier Station with Bridger Amplifier and Reverse Switch

[5] This may not be true for older systems and excludes the sub-split only, in-line equalizer.

A close look at the subscriber reverse system shows many reverse signal-source origination points all funneling back to the headend. A little ingress at all of the origination points or a large ingress at only a few can equally cause serious operational problems. Given that sub-split reverse ingress problems exist, the question arises as to how to locate the source or sources. One method would be to disconnect all reverse feeder connections to the system and independently turn each back on while monitoring for ingress. In fact, this approach, has proved to be the only effective method to date; computerized equipment to accomplish this has been developed and is presently available. A block diagram showing the location of the trunk station reverse system switch is presented in Figure 8.

The Digital Headend

Several articles that have been published recently refer to the "digital headend." The digital headend will process signals that are somewhat different from video signals and will use modulation methods that may be unfamiliar. Nevertheless, the analogy to the classical cable television headend is apparent.

On the customers' premises, racks of equipment consisting of digital multiplexers and standard bandwidth modems are installed. The output of the modems are frequency multiplexed on a two-way system with the output from other modems. At the earth station, there will be much larger "digital headends" receiving the modem RF signals and processing them to a format campatible with the satellite equipment. For installations where the digital earth station is not colocated with the cable headend, data translators may be required to allow full access to any location in the system. In many respects this "digital headend" is less complex than some of the very sophisticated cable headend systems which are being installed today.

The information that is to be sent by cable can be multiplexed using time-division multiplexing (TDM) or frequency-division multiplexing (FDM). Some combination of time and frequency division access will probably be used for most applications. Consider the data rates which are commonly used in satellite circuits (Table 2). If many low data rate ports are available at one location, the most cost-effective method of transmitting the signals is to time-division multiplex before modulation. On the other hand, the higher data rate services are usually sent single-channel-per-carrier (SCPC) using frequency-division multiplex on the coax or satellite.

Table 2. Commonly Used Data Rates for Satellite and Terrestrial Services

Characteristic	Specification
Multiples of 1.2 kb/s	1.2, 2.4, 4.8, 9.6, 19.2 kb/s
Multiples of 56 kb/s	56, 112, 224, 448 kb/s
T1	1544 kb/s
2T1	3088 kb/s
T1C	3152 kb/s
T2	6312 kb/s

Another item that needs to be mentioned is changing access according to the changing needs of different users (multiple access). Pure time-division systems may be made multiple access by allowing different users to occupy different time slots on demand (TDMA). Likewise, frequency-division multiplex systems may be extended to demand access (FDMA or DAMA). Another method that is commonly used is called carrier sense multiple access/collision detection (CSMA/CD). Basically, a station wishing to transmit using CSMA/DC listens to the circuit. If the link is idle, it transmits. If two stations should transmit simultaneously (collide), each attempts to transmit again after a random delay. Two other widely accepted schemes for multiaccess that are closely related are polling and token passing. In polling, a master controller polls each station on the network. If a station has a message to send, the controller gives the station the right to transmit. In token passing, the control is not centralized; the stations are given a sequence. After transmitting a message or no message, each station passes access rights (i.e., token) to the next station. This passing of the token continues in a cyclical fashion.

In addition to the multiplexing and accessing method, one also has to consider the modulation that is to be used. The basic possibilities are amplitude shift keying (ASK), phase shift keying (PSK), and frequency shift keying (FSK). An attractive method for transmitting on nonlinear systems (satellites) is PSK. PSK transmission can take place using any number of phases, e.g., two phases (biphase, BPSK), four phases (quadra-phase, QPSK), etc. BPSK and QPSK are widely used in satellite circuits. When transmitting at high data rates on cables, the frequency spectrum must be conserved. Since the cable is relatively linear and the signal-to-noise ratios are high, an attractive modulation method is a combination of amplitude and phase shift keying. The measure of transmission efficiency normally used is called bits/Hz. This is the number of bits per second that can be transmitted in one Hz bandwidth. The theoretical number of bits/Hz for BPSK is one, for QPSK it is two, and for a combination of amplitude and phase shift keying, it can be three or more. For data rates of T1 or larger, a combination of amplitude and phase shift keying can conserve spectrum on the cable.

On the other hand, consider the case of a large number of users at different locations, each user having a relatively low data rate, and each user transmitting occasionally. Here a modulation method can be employed that uses spectrum less efficiently, but yields less costly hardware, e.g., FSK. In addition, the system could employ CSMA/CD or token passing, giving all users access to the same channel. In most cases, it is evident which type of modulation method should be employed.

Application for Cable

The potentially large market for cable servicing as the local distribution facility connecting businesses with satellite common carriers has been discussed. In addition, some of the technical methods that are available for implementing this cable distribution system have been investigated. Now, let us consider an example of two current business communication needs. Business A wanted to establish a dedicated communication link to another city with eight voice channels, one 9.6 kb/s circuit, one 56 kb/s circuit, and one 224 kb/s circuit.

Business B wished to communicate with 40 voice channels through the same satelite communications earth station. Twenty-four voice channels can be digitized on a commercially available channel bank and sent on a T1 (1.544 Mb/s) circuit. Data multiplexers can be added to one channel bank in order to provide the eight voice channels and the various data circuits. The T1 output from the channel bank is then fed into a T1 modem and converted to the appropriate frequencies for transmission on the cable (Figure 9). The 40-voice-channel requirement for Business B can be satisfied using two channel banks and two T1 modems. Both customer requirements are satisfied economically using the configuration shown in Figure 9.

System Capacity

Before exploring the question of capacity of a broadband cable system, it would be beneficial to look at a couple of examples of data circuits on cable. Figure 10 is a schematic representation of a mid-split broadband distribution system.

Signal flow in the conventional (downstream) direction takes place in the band 174 to 440 MHz, while reverse (upstream) signals are carried in the band 5 to 108 MHz. This two-way capability is achieved by the use of diplexing filters and dual amplifiers at each signal amplification station.

TA: Trunk Amplifier. (Long-distance, high-quality transportation.)

LE: Line Extender Amplifier. (Short-distance, high-gain, for subscriber circuits.)

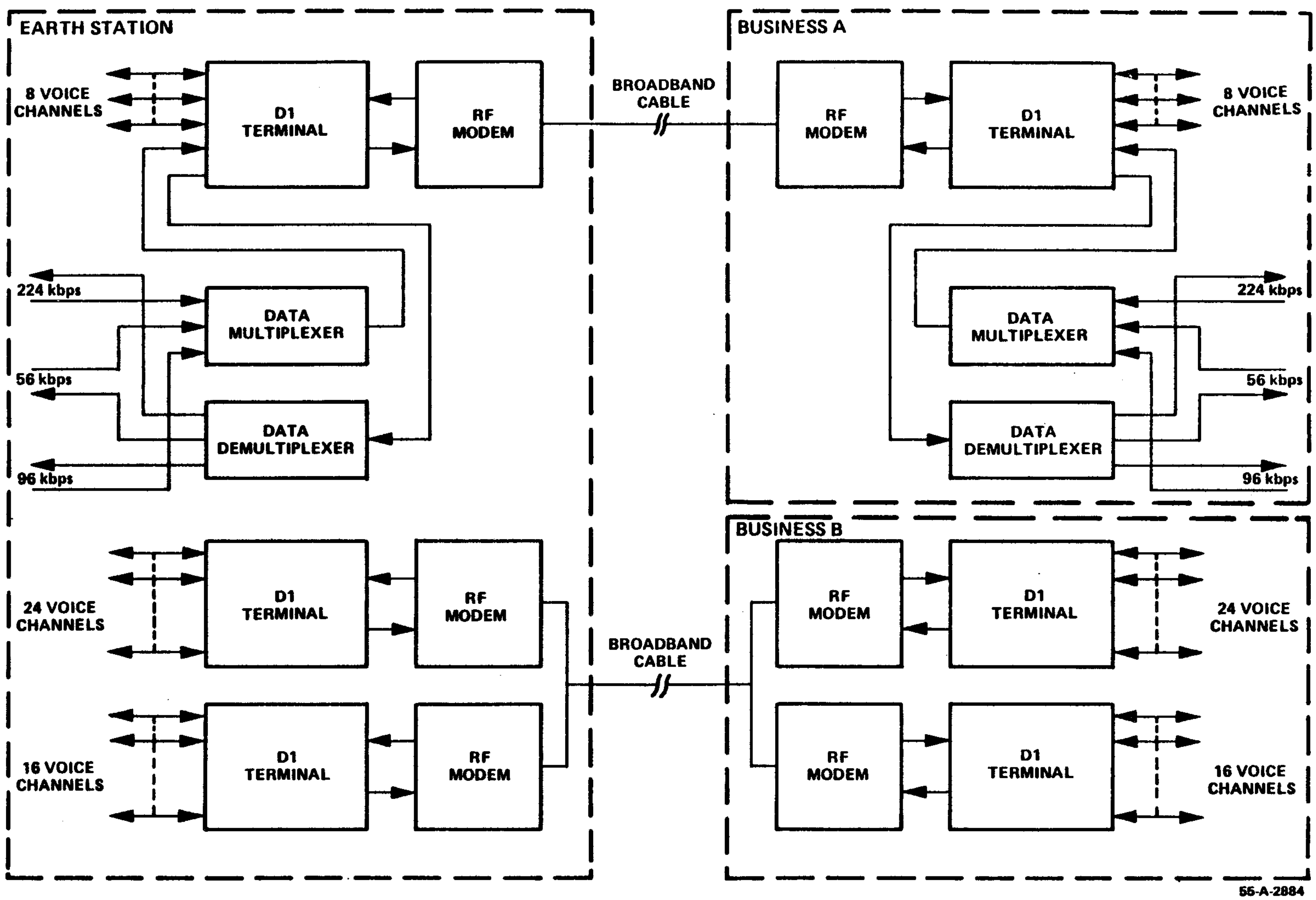

Figure 9. Cable Data Communications Link Equipment Configuration

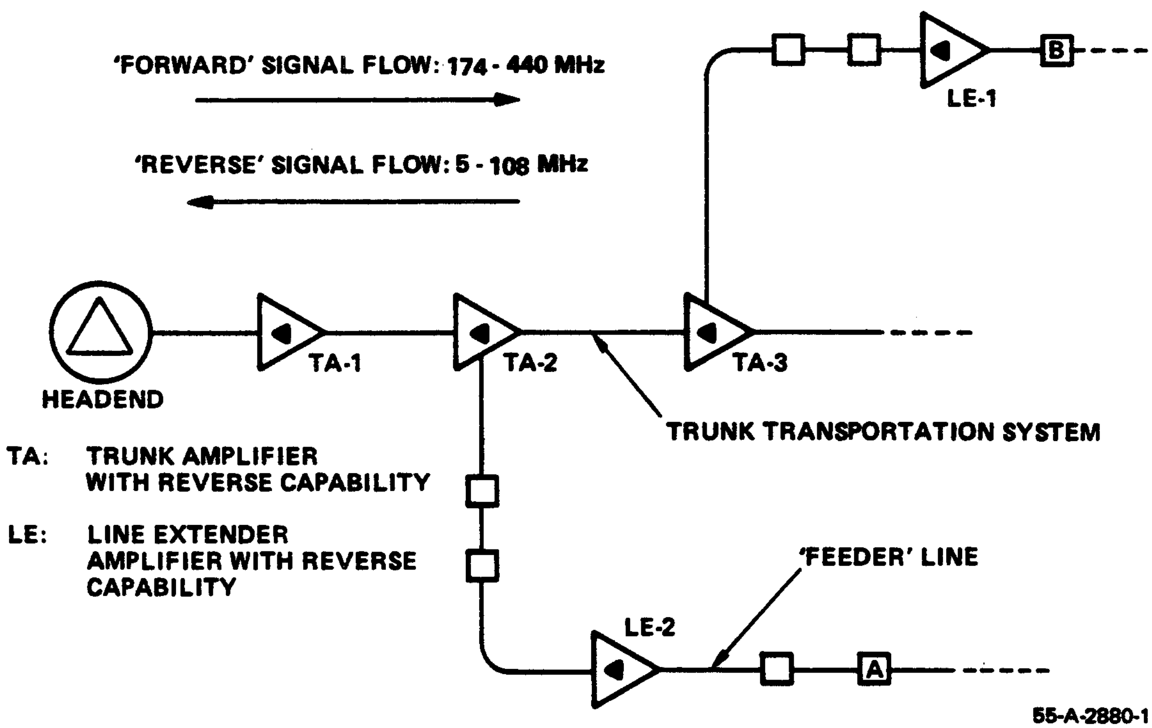

Figure 10. Signal Flow

Suppose that at point "A" there is a bank branch office equipped with a data terminal which must communicate with the master computer located in the head office at point "B". The data terminal at "A" is connected to a modem which can transmit onto and receive from the distribution system, and the master computer at "B" is similarly connected. It is clear that there is no way in which a signal can travel directly from "A" to "B" using the route LE-2, TA-2, TA-3, LE-I, or vice versa. Any signal in the range 5 to 108 MHz will travel only toward the headend, and any signal in the range 174 to 440 MHz will travel only away from the headend. However, the addition of one simple device at the headend will solve this problem, and allow great flexibility in connecting additional customers to the system. The device is a frequency translator, which receives signals at the headend in the "reverse" frequency range and "translates" them to the "forward" frequency range. It can be designed to translate a single 6 MHz channel or several channels, depending on the amount of traffic and the availability of unoccupied channels.

Therefore, when "A" wishes to send a message to "B," "A's" modem transmits at a frequency in the range 5 to 108 MHz. The signal returns to the headend, where the translator shifts it to an available channel in the "forward" spectrum. "B" can then receive the transmission.

Figure 11 is a schematic representation of one version of a mid-split private broadband network. The difference between this distribution network and the network of Figure 10 is that all communication is between the headend and the remote sites; i.e., no direct communication is allowed between remote sites. This agreement could arise with a company that would have a headquarters with multiple remote buildings or if a network was built solely to distribute communications to and from a satellite site. In this case, the modems at the remote sites, A and B, would transmit in the 5 to 108 MHz band and receive in the 174 to 440 MHz band. The modems located at the headend, C, can use an inverted frequency plan; that is, transmit in the 174 to 440 MHz band and receive in the 5 to 108 MHz band. This inverted frequency arrangement saves hardware by not requiring the frequency translator.

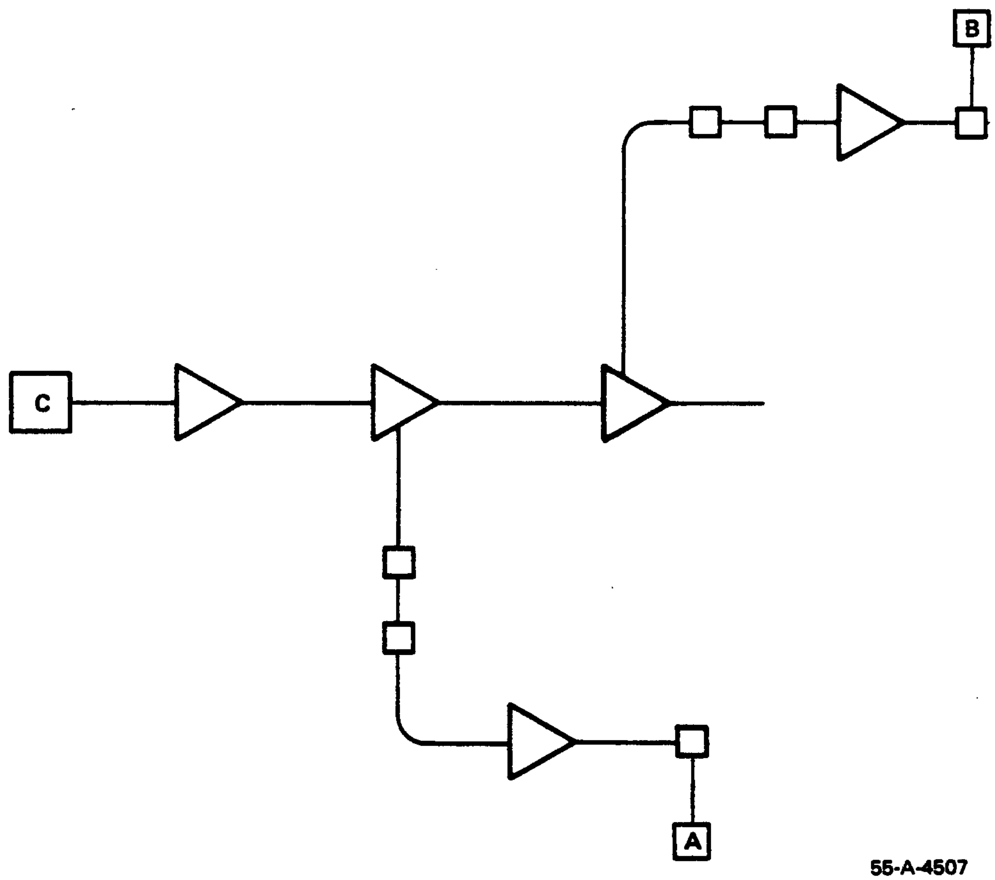

Figure 11. Schematic Representation of One Version of a Mid-Split Private Broadband Network

The data transmission capacity of a broadband system depends on the following factors:

a. Available bandwidth (expressed as the number of unoccupied 6 MHz channels). (Note that in an Institutional system, the whole cable spectrum may be given over to data transmission.)

b. Number of data circuits.

c. Modem spectral occupancy, which is a function of both the data rate and the bandwidth efficiency of the modem.

To help visualize the potential capacity, we will examine first the case of the intracity, point-to-point connection. Modems for broadband systems are available from a number of sources (including Scientific-Atlanta); therefore, for the purpose of this discussion, we will assume a hypothetical, medium-speed modem which can handle data at rates up to 19.2 kb/s. The figures which we shall use are reasonable and quite representative of state-of-the-art devices which meet the requirements of high reliability and low cost.

Using Bi-Phase Shift Keying (BPSK) as our modulation scheme, we may expect to transmit a 19.2 kb/s data stream in a bandwidth considerable less than 100 kHz. We wish to allow a certain safety margin, however, so we will suppose that 100 kHz represents the "channel" separation required by the modems. A quick calculation shows that a single 6 MHz TV channel could support transmissions from 60 such modems.

To determine the actual capacity of a typical system, refer to Figure 12. This diagram is an extension of Figure 10, since it illustrates the two-way communication between points "A" and "B." The cable system has a reverse channel, designated T7, available and a forward channel, designated H, currently unused. Each 6 MHz channel has been divided into 60 "sub-channels," numbered 1 through 60, so that we can readily describe the "slots" which the modems require for transmission and reception.

When modem "A" transmits a signal, it uses subchannel 1 of channel T7. This can be written as T7(1). The signal travels in the reverse direction to the headend, where it is intercepted by the frequency translator and converted to channel H, subchannel 1, or H(1). The transmission can now travel in the forward direction throughout the distribution system, and be received by modem "B," which is tuned to H(1).

Transmission in the opposite direction, from "B" to "A," must make use of a different subchannel. This is because both modems may be required to transmit and receive simultaneously (i.e., full duplex operation). Therefore, modem "B" transmits in T7(2), and modem "A" is tuned to receive H(2).

To summarize, a pair of modems operating in full duplex mode requires the allocation of two subchannels in the reverse frequency range, and two corresponding subchannels in the forward frequency range. In our hypothetical example, this is equivalent to 200 kHz in channel T7, and 200 kHz in Channel H. It follows that, if all the subchannels in T7 and H were allocated, the system could support full duplex communication between 30 pairs of modems.

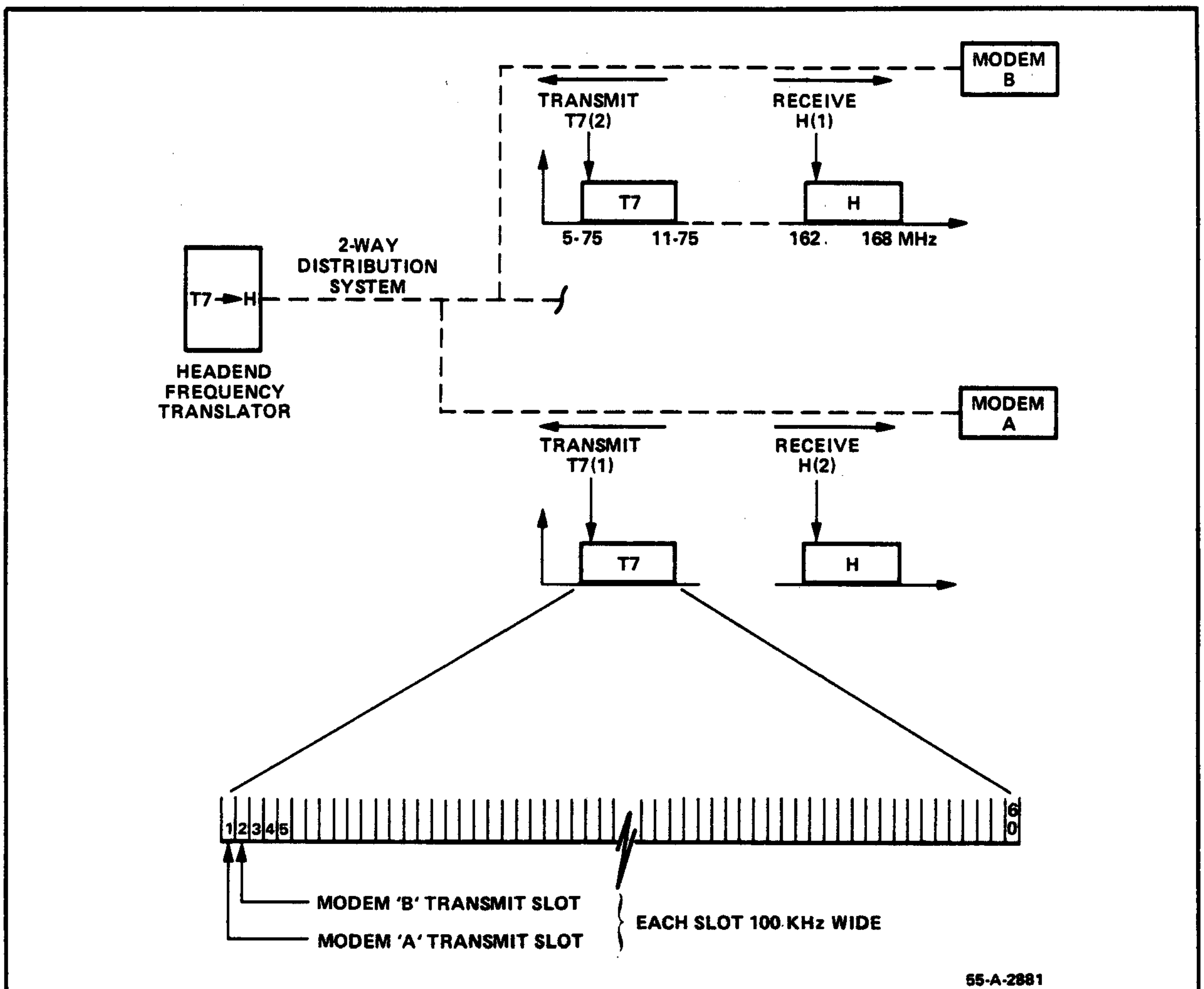

Figure 12. Frequency Assignments for a Typical CATV Modem Pair

A more efficient use of bandwidth can be obtained when a Multipoint service is required. In this configuration, a master computer communicates with several slave terminals by sequential polling. Since only one modem is transmitting at any instant, the same frequencies can be allocated to all modems, with a consequent drastic reduction in bandwidth. Typical of the terminals requiring this kind of service are bank teller machines; a master computer at the bank's main office interrogates teller terminals at each branch office in sequence. When this type of connection is established using telephone lines, low data rates are frequently employed in order to minimize the degree of line-conditioning required, and hence the cost. By switching to broadband cable, a higher data rate can be realized, and the polling time reduced.

High-Speed Services

The use of CATV systems for the transmission of high-speed (1.544 Mb/s) data, while employing the same priciples of forward and reverse signal flow as described above, necessitates a different approach to the subject of hardware. At these data rates, care must be taken to restrict the amount of bandwidth required by each data channel, lest the most valuable technical resource of coaxial cable, namely its prodigious bandwidth, be rapidly consumed.

A simple FSK or BPSK modulation scheme, which may be economically desirable and technically acceptable in a low-speed modem, could be disastrous when applied to T1 transmission equipment. The Scientific-Atlanta Model 6402 Modem achieves a bandwidth efficiency of better than 2 bits/Hz (i.e., a 1.544 Mb/s data stream is transported in a bandwidth less than 750 kHz) by using the QASK-16 modulation process. QASK-16 is a special case of the M-ary Amplitude Phase Shift Keyed (MAPSK) family of signal sets which provide enhanced bandwidth efficiency through efficient signal packing at the expense of bit-error probability in a noisy environment. However, since coaxial cable systems can provide high signal-to-noise ratios, these complex modulation schemes are applicable.

The Model 6402 is primarily intended for use with readily available PCM trunk terminal equipment ("channel banks"), using a Bell standard DS1 interface. In this configuration, telephone voice signals between office PBX (Private Branch exchange) equipments are converted to digital form in groups of up to 24 circuits, multiplexed into a single 1.544 Mb/s data stream, and transported over coaxial cable via the 6402 modem. This approach eliminates the need for multiple leased telephone lines between office buildings. Up to four such T1 links can be accommodated in a single 6 MHz TV channel. Most PCM trunk terminal manufacturers can supply data channel units to take the place of standard voice channels, thus offering the possibility of mixed voice/data over a single T1 link. Figure 13 is a schematic diagram of such an interconnection.

Other applications of the 6402 modem include high-speed computer graphics and compressed digital video (teleconferencing) interconnections, which make use of the T1 data rate. The hardware which has been briefly described in this article is, in principle, extremely simple: the method of connected translators and modems into an existing CATV system is straight-forward, and alignment procedures can be conducted using conventional test equipment. For a relatively small initial investment, the cable system owner can begin to serve the needs of data subscribers throughout his franchise area, thus opening up a new source of revenue.

Broadband technology offers the immediate incentive of reduced cost and improved quality of transmission, and the confidence that increasingly sophisticated networking capability and faster data transportation, necessitated by the growing demand for information transfer and the concomitant growth of data technology, are tasks that fall naturally with the scope of coaxial cable systems.

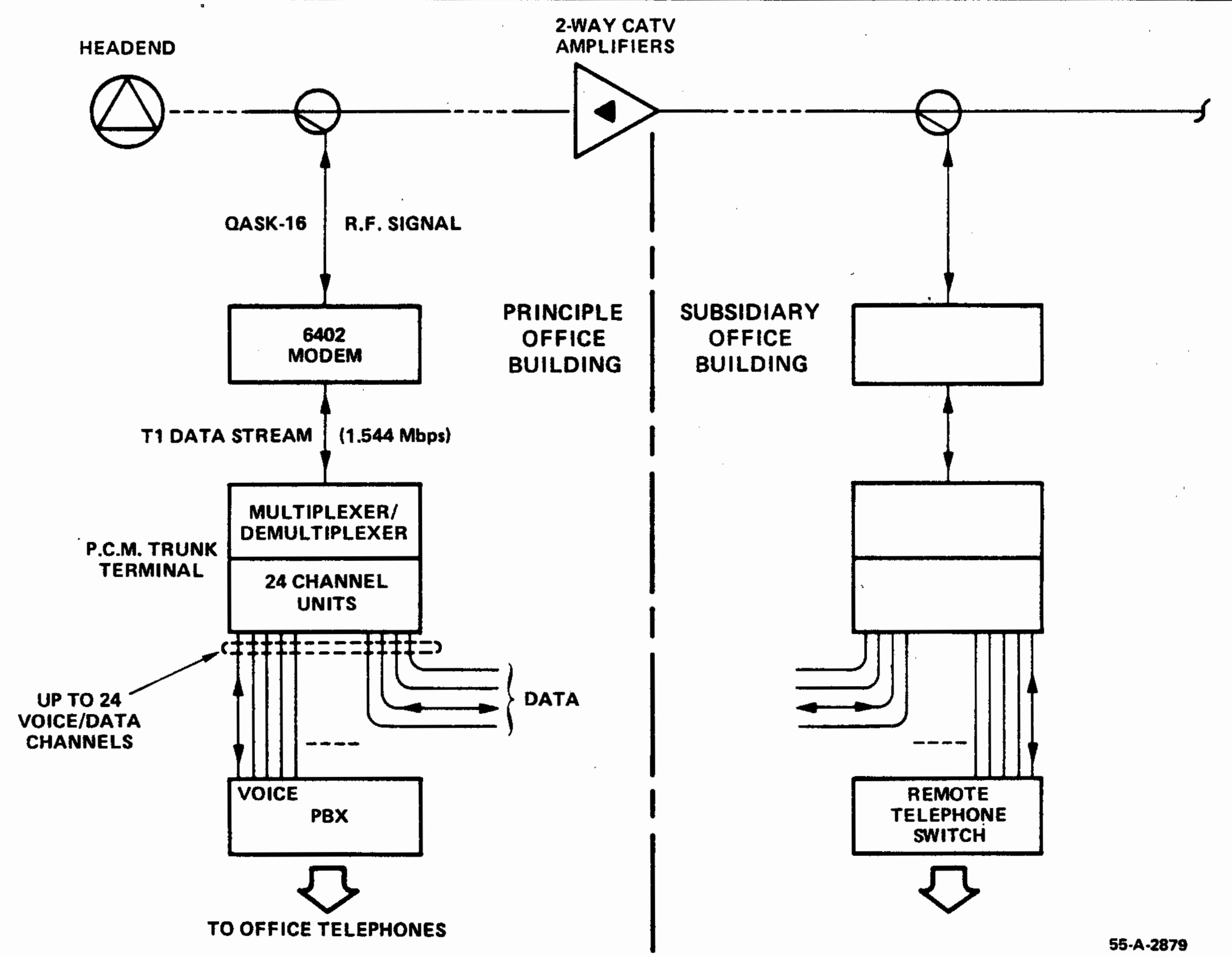

Figure 13. Voice/Data Interconnection Using T1

Conclusions

This paper has reviewed the basics of broadband communications networks. It has been shown how broadband cable can be configured to handle the distribution of voice, data, electronic mail, teleconferencing and other business communications. The inherent flexibility of bandwidth allocation, low cost of high data rate equipment, quality of service (low error rate), and high reliability/availability (proven technology) makes broadband cable an ideal medium for serving the local distribution needs of satellite communications carriers, large corporations, financial institutions, municipal agencies, hospitals, and university complexes.

Principles of Satellite Communications

J. Searcy Hollis

Introduction

The purpose of this paper is to present some of the elements of satellite communications. We will assume that the reader has a technical background in electronics and a general understanding of satellite communications.

There are many applications of satellite communications and many facets of the theory which we cannot cover in a single paper. The approach taken here is to describe a simple FM/video link as one example, using it as a vehicle to describe the background for the various link equations and giving references as we proceed. We will then discuss briefly certain of the more important topics. A Glossary of Terms is included in Section 5 to define terms which occur frequently. An appendix is included at the end of the paper, which discusses decibels and especially the units of certain quantities which are expressed in decibels.

Desription of a Simplified Satellite Link

A satellite is assumed to be at a longitude of 90 degrees W in the geosynchronous orbit at a height of 35,800 km above the equator (see Figure 1). The transmitting earth station is near New York City at 74 degrees W, 41 degrees N; the receiving earth station is near Los Angeles at 118 degrees W, 34 degrees N.

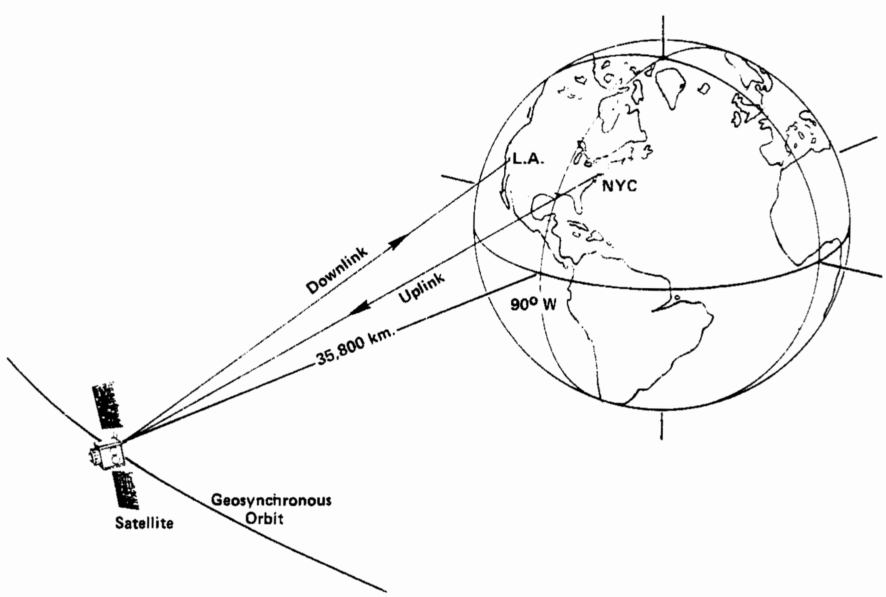

Figure 1. Satellite Link Geometry

It can be shown* that the satellite is about 40 degrees above the horizon (satellite elevation angle) as viewed from the transmitting earth station and lies generally to the south-southwest at an azimuth of about 204 degrees. From the receiving earth station, the satellite elevation angle is 40.5 degrees, and the azimuth is about 136.5 degrees.

Figure 2 is a block diagram of the assumed FM/video communications link. It consists of a transmit earth station, the satellite, a receiving earth station, and the propagation paths traversed by the signals. For simplicity, the satellite is assumed to have only one transponder, and the uplink and downlink frequencies are assumed to be 6 GHz and 4 GHz, respectively. In practice there are no FM/video carriers at exactly 6 GHz, and in the 6/4-GHz band** downlink carriers are offset 2.225 GHz instead of 2 GHz below the corresponding uplink carriers. We will assume that only one earth station accesses the transponder at a given time.

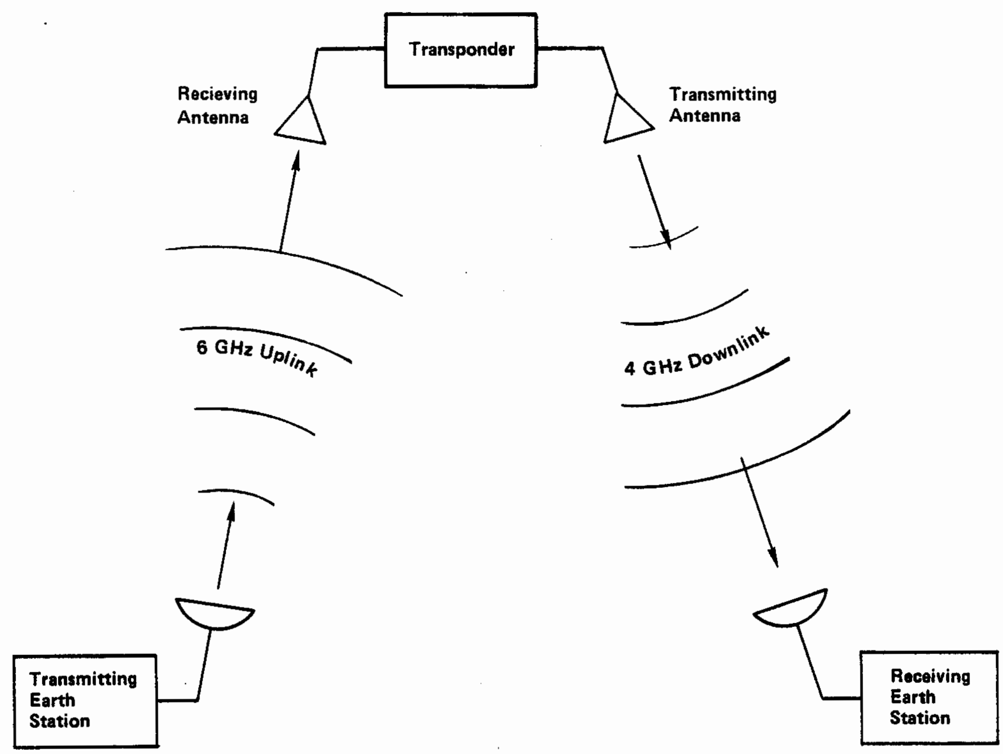

Figure 2. Simplified block diagram of a 6/4 GHz satellite communications link. Only one of the 12 or 24 transponders that make up a typical satellite is shown

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* See the paper, "Earth Station Geometry."
**The designation 6/4-GHz is used to indicate that the uplink is in the 5.925- to 6.425-GHz band, and the downlink is in the 3.7- to 4.2-GHz band.

The uplink consists of the earth station transmitter and its antenna, the uplink propagation path, the satellite receiving antenna, and the transponder receiver. The satellite transponder receives the incident signal, converts it from the 6-GHz band to the 4-GHz band, and transmits it back to earth.

The downlink consists of the transponder transmitter, the satellite transmitting antenna, the downlink propagation path, and the earth station receiving antenna and its associated receiving equipment.

The quality of any communications link is determined by the difference between the output signal and the input signal. In an analog system such as FM/video, the difference is represented by distortion and by added noise. In a digital system, it is represented by the bit error rate (BER). The satellite link under consideration must adhere to distortion and video signal-to-

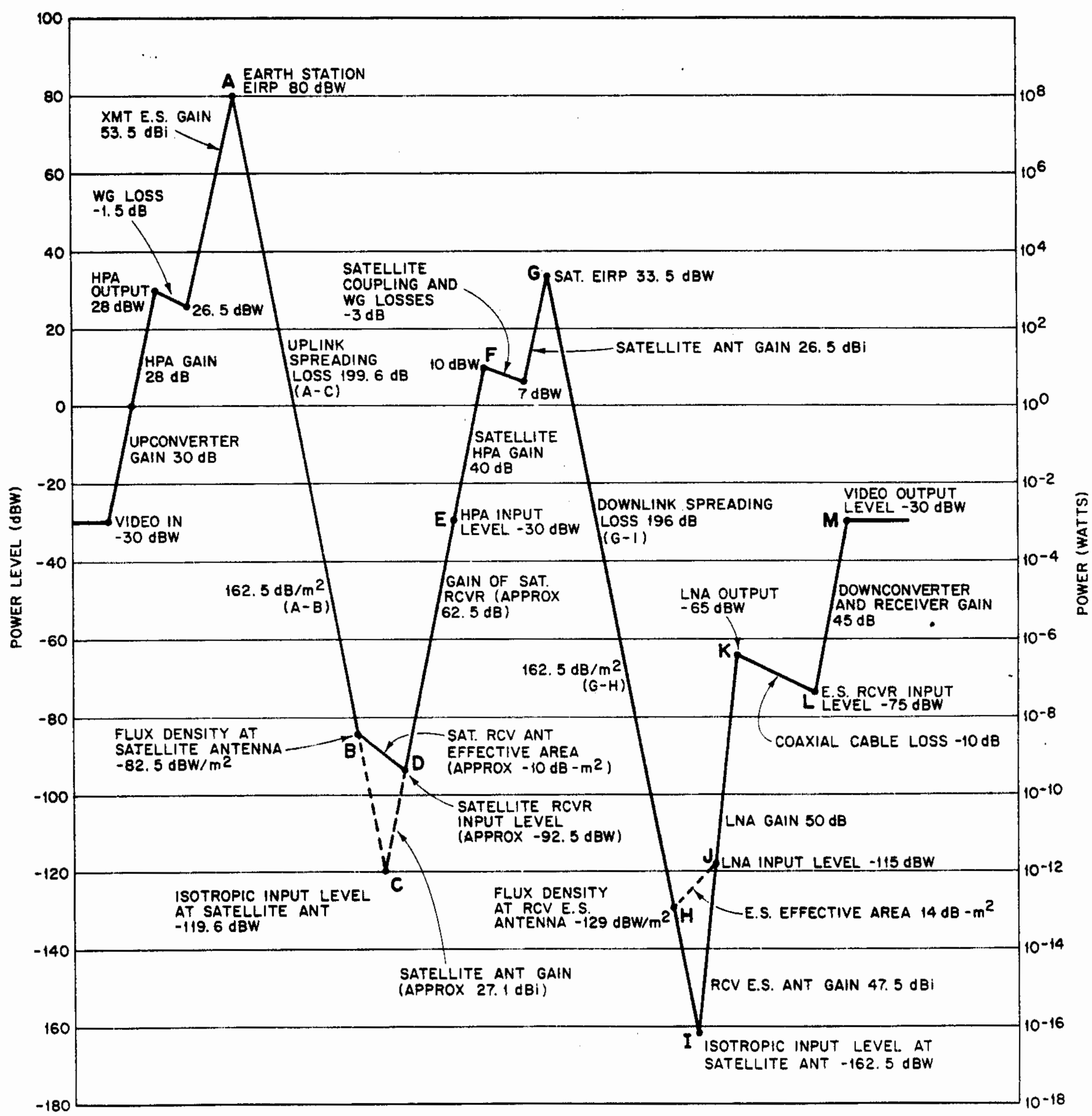

Figure 3. Signal Level Changes Through Hypothetical Satellite Link

noise criteria which are appropriate to the application. These criteria are given in detail in the several standards which are listed in the bibliography.

In our hypothetical system we will postulate a 54-dB,* clear-weather video signal-to-noise ratio (S/N) at the receiving end as defined in Reference 1, and will assume distortion limits which will be discussed in the paragraph "Link Performance."

Very large changes in signal levels take place as the signal passes through the complete communications link. Figure 3 follows these changes.

Uplink

Transmitter and Transmitting Antenna. Figure 4 is a simplified block diagram of the transmitting earth station. It consists of an FM modulator, an upconverter, a high-power 6-GHz transmitter, and a transmitting antenna.

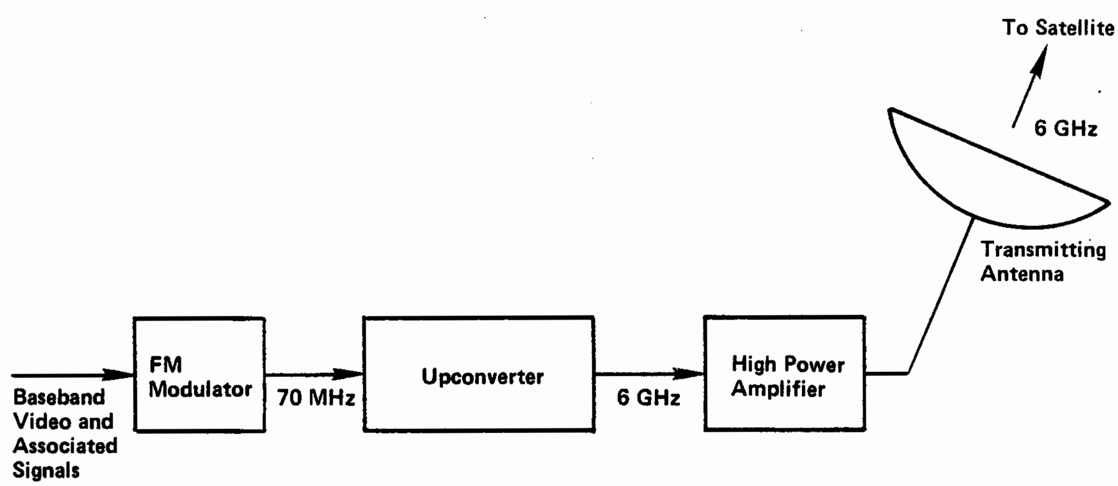

Figure 4. Simplified Block Diagram of Transmitting Earth Station

Before modulation the video signal is processed to preemphasize the higher frequency components, and an energy dispersal waveform is added. Preemphasis acts to improve the output video signal-to-noise ratio (S/N) by compensating for the increase in noise density with frequency (triangular noise) which is characteristic of the receiver demodulator. (The preemphasis is removed by a deemphasis network after the receiver discriminator.) The energy-dispersal signal frequency-modulates the carrier with a triangular waveform at the video frame rate to disperse the RF spectrum. This reduces interference with terrestrial microwave and other satellite links and reduces intermodulation among the multiple carriers which exist in a real satellite.

*A received video signal with an S/N of 54 dB is of "broadcast quality." Lower values of S/N are adequate for many applications.

The baseband video and the associated audio and energy-dispersal waveforms are frequency-modulated onto a 70-MHz carrier. The modulated signal, whose modulation bandwidth is approximately 36 MHz, is upconverted to 6 GHz, amplified and used to drive a klystron high-power amplifier (HPA) which feeds the transmitting antenna.

The transmitter and transmitting antenna are characterized by an effective isotropic radiated power (EIRP), given by

$$\mathrm{EIRP} = P_0 g_T \qquad \text{(watts)} \qquad (1)$$

where P_0 is the transmitter output power in watts and g_T is the transmitting antenna power gain.

The gain of an antenna acts to concentrate the transmitter output power into a narrow beam. In general, the narrower the pattern the higher the gain. The effective isotropic radiated power is the power that would have to be transmitted if the radiated power were to spread out uniformly in all directions from the source, implying an antenna gain of unity.

The transmitting antenna has to be relatively large in diameter to generate a narrow beam. The narrow beam pattern protects other satellites from interference from the transmitted wave, and at the same time produces high gain which is required to overcome propagation losses.

Equation (1) can be expressed in decibels* by

$$\mathrm{EIRP} = P_0 + G_T \qquad \text{(dBW)} \qquad (2)$$

where EIRP and P_0 are expressed in dBW and G_T is in dBi.

Our link will use the 10-meter-diameter antenna of Figure 5, which has a gain of 53.5 dB and a half-power beamwidth of about 0.32 degrees at 6 GHz. We will assume an EIRP of 80 dBW (10^8 watts), which is obtained from the 53.5 dB of antenna gain (about 224,000) and 26.5 dBW of transmitter power (about 450 watts). After discussing the uplink path and the satellite sensitivity, we will check to see if this assumed EIRP is adequate.

Uplink Propagation Path. On passage through the atmosphere, the EIRP is decreased slightly by atmospheric attenuation, which is very small at 6 GHz even in heavy rains (see Figure 6). On the other hand, a very large attenuation occurs because of the spreading of the spherical wavefront as the wave radiates outward toward the satellite.

It is evident from the definition of EIRP and the large distances involved that a very small fraction of the total EIRP of 100,000,000 watts intercepts the satellite.

The signal level of the wave incident on the satellite is measured by the power density S of the approaching wavefront expressed in watts per square meter.

*See Appendix A for discussion of decibels.

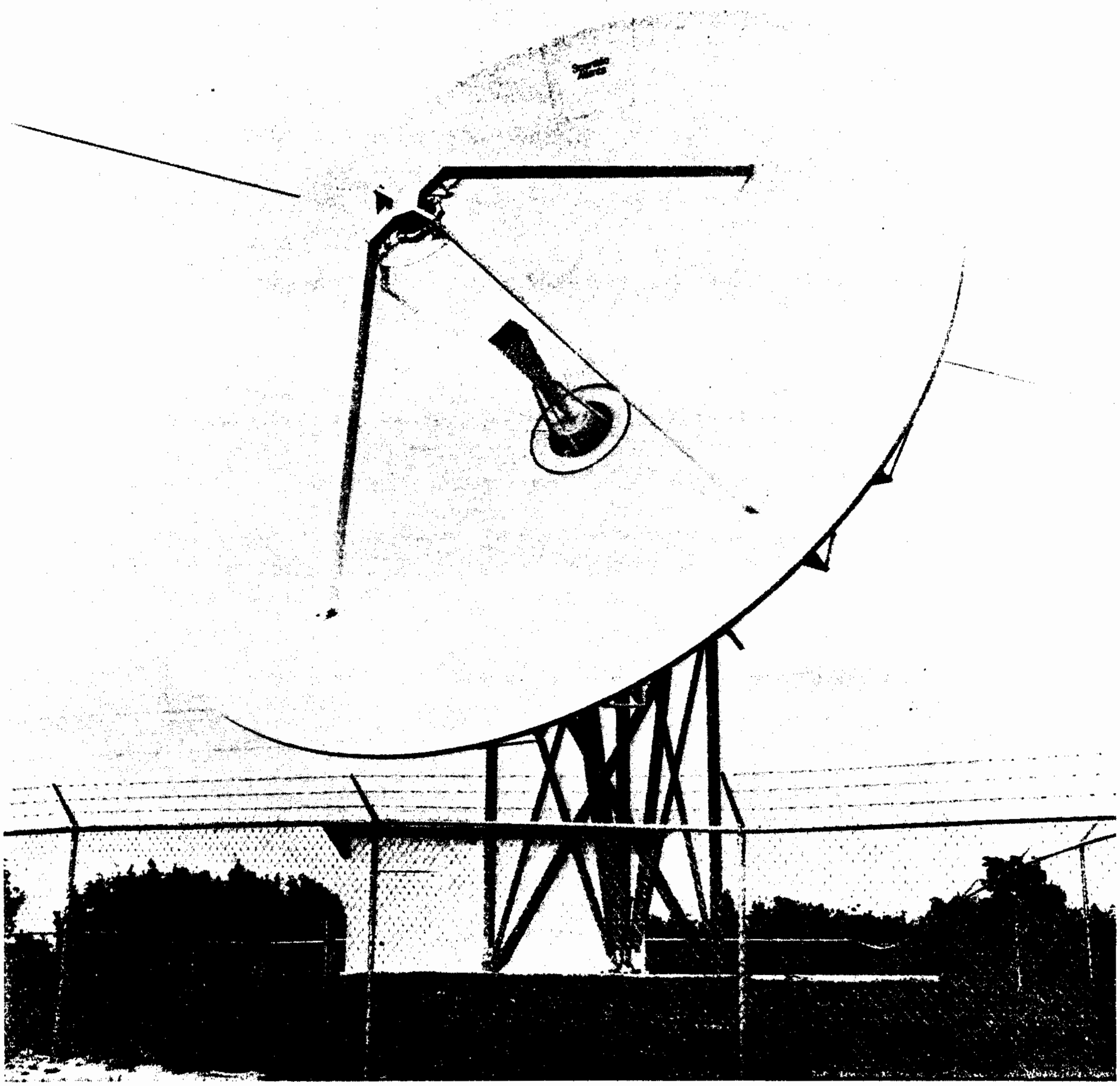

Figure 5. Scientific-Atlanta Earth Station Using 10-Meter-Diameter Antenna

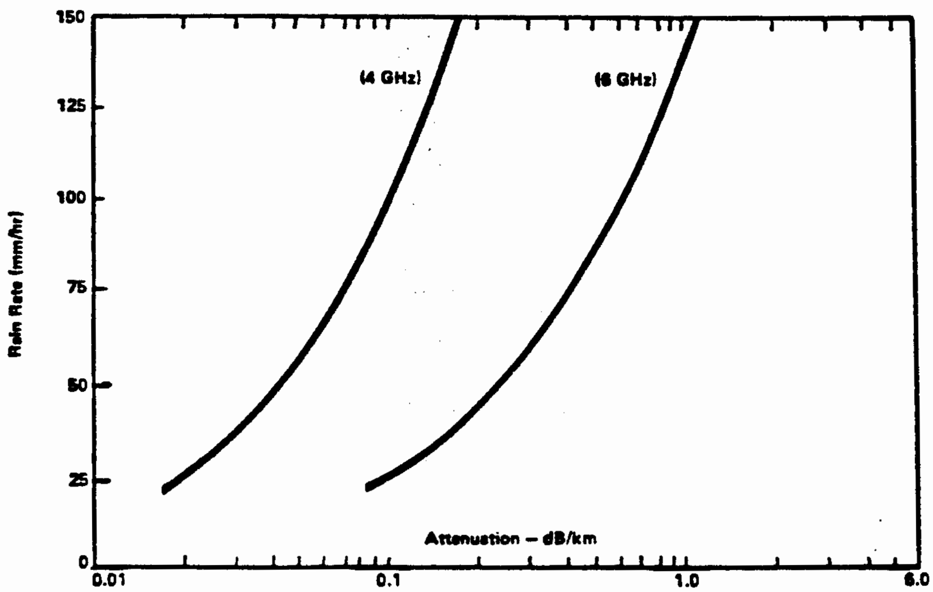

Figure 6. Attenuation vs. Rain Rate. The total atmospheric attenuation is obtained from the integrated effect of the attenuation over the total path. The total path length which contributes to rain attenuation is relatively short because the path to the satellite is slanted rather than horizontal. Typical values of attenuation due to rain, even in heavy rains, are likely to be less than 1 decibel at 6 GHz (1 dB of attenuation correstponds to k_A = 0.794 in (3))

It can be shown that the distance R_S from New York City to the satellite is approximately 37,750 kilometers.* The surface area of a sphere of this radius is given by

$$A_S = 4\pi R_S^2 = 1.791 \times 10^{16} \quad (\text{meter}^2) \qquad (3)$$

The flux (power) density at the satellite is calculated as if the EIRP uniformly covered the total surface area A_s. The resulting flux density is given by

$$S = \frac{\text{EIRP}}{4\pi R^2} k_A = \frac{\text{EIRP}}{A_s} k_A \quad (\text{watts/meter}^2) \qquad (4)$$

where k_A is the atmospheric attenuation factor, which is less than unity.

It is convenient to convert (4) to decibel form to obtain the flux density in decibels relative to one square meter:

$$S(\text{dBW/m}^2) = \text{EIRP(dBW)} - 10 \log A_S - L_A(\text{dB}) \qquad (5)$$

where L_A (= -10 log k_A) is the atmospheric attenuation in decibels. We can assume k_A to be unity in (4) with small error, making L_A(dB) in (5) zero.

*See the paper, "Satellite Link Analysis."

We will want S to saturate the satellite; that is, to cause it to transmit maximum power. We can arrange (5) to give

$$\text{EIRP} = 10 \log A_s + S \qquad (\text{dBW}) \qquad (6)$$

where S is the saturation flux density of the satellite. To confirm whether our assumed uplink EIRP is adequate, we must know the saturation flux density. This requires discussion of the characteristics of the satellite.

Satellite. Current 6/4-GHz satellites have either 12 or 24 transponders, which have bandwidths of about 36 MHz each. Figure 7 is a simplified block diagram of our assumed single-transponder satellite. A more detailed discussion of practical satellites is given later.

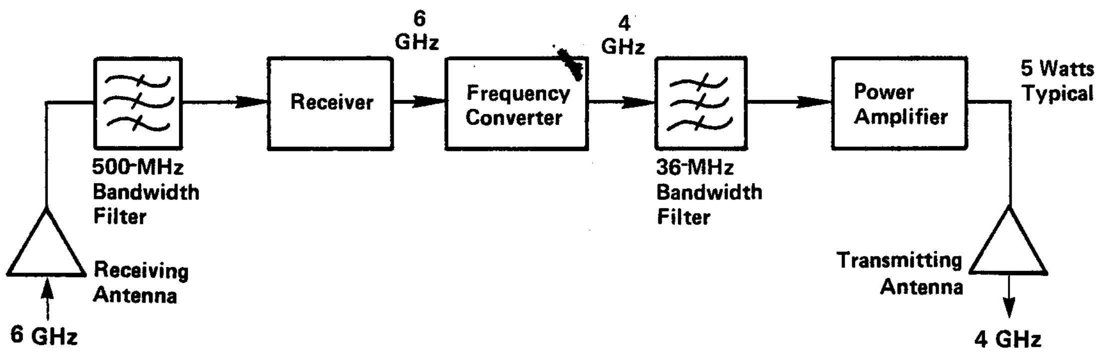

Figure 7. Block Diagram of Simplified Single-Transponder Satellite

Satellite Receiving Antenna and Receiver. To accommodate signals received from wide geographical areas, the satellite receiving antenna has a broad beam pattern. Figure 8 is a footprint showing contours of flux density which are required to saturate our assumed satellite transponder as a function of latitude and longitude of a transmitting earth station. The curves are also labeled in terms of a value G/T, which we will discuss later.

From (3) we obtain for the first term of the right side of (6) a value of 162.5 dBW/m^2. Reference to Figure 8 indicates that a flux density of -82.5 dBW/m^2 is required to saturate the satellite receiver from an earth station near New York City. Thus, we have from (6):

$$\text{EIRP} = 162.5 - 82.5 = 80.0 \ (\text{dBW}) \qquad (7)$$

confirming that our assumed value of 80 dBW is nominally correct for the EIRP of the earth station.

In practice several dB of additional power would be made available by use of a larger HPA to overcome waveguide and switching losses, antenna pointing error, atmospheric attenuation and other losses.

So far we have been concerned with the required flux density incident at the satellite. The power received by the satellite antenna is given by the incident flux density times the effective area $A_e(\theta_1, \phi_1)$ of the receiving antenna times the polarization efficiency p:

$$C = SA_e p = \frac{EIRP}{4\pi R^2} A_e(\theta_1, \phi_1)\, p \quad \text{(watts)} \tag{8}$$

where the letter C implies "carrier" and where in our case the direction (θ_1, ϕ_1) is the direction to New York from the satellite.* The polarization efficiency p is virtually unity in a properly designed and adjusted system.

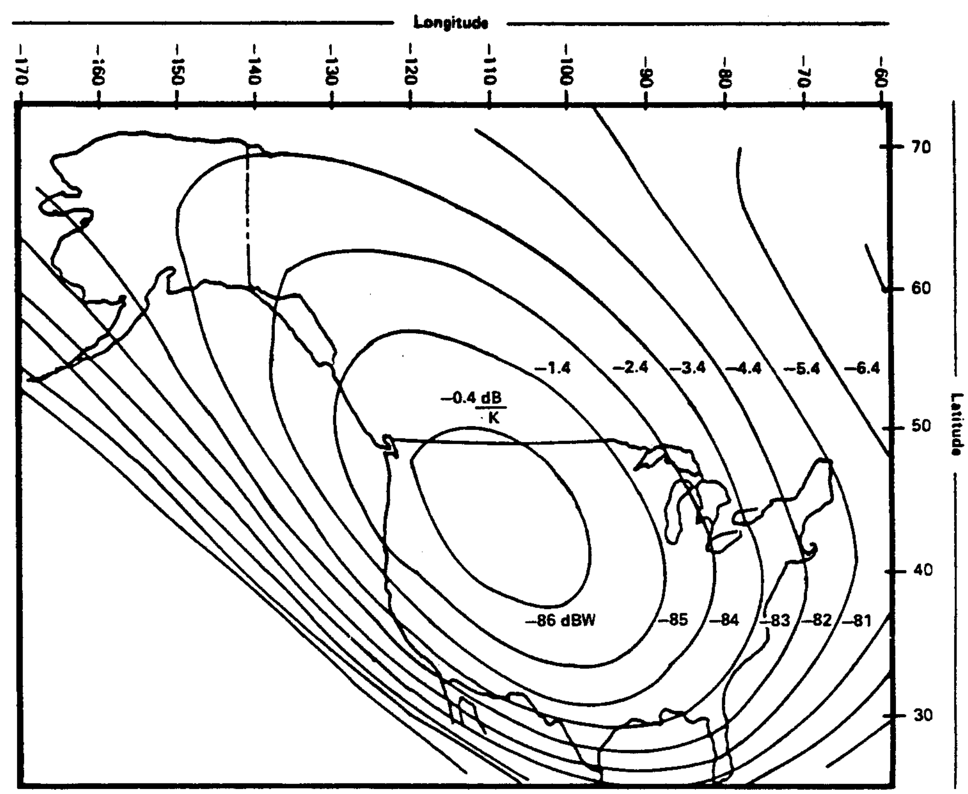

Figure 8. Typical Variation of Saturation Flux Density and G/T for a Domestic U.S. Satellite

*The gain $g(\theta, \phi)$ and the effective area $A_e(\theta, \phi)$ of an antenna are quantitative descriptions of its beam pattern. The coordinates (θ, ϕ) are angular spherical coordinates.[2] We will drop the coordinates θ and ϕ when the direction is obvious.

It can be shown that the effective area and the gain of any reciprocal antenna are related by the equation:

$$A_{eR}(\theta,\phi) = \frac{g(\theta,\phi)\lambda^2}{4\pi} \qquad (9)$$

where λ is the wavelength. This relationship will be needed a little later in the discussion, but for the moment we are interested in the directions θ and ϕ which translate into the latitude and longitude of the footprint of Figure 8.

The effective area of a satellite antenna which provides coverage of CONUS (the 48 contiguous states) at 6 GHz as in Figure 8 is quite small, about 0.1 meter^2. If we knew the effective area or the gain of the satellite antenna, we could calculate C, but neither the effective area nor the gain of the receiving antenna is called out in Figure 8.

Out of interest a value of C is calculated using an estimated value of $0.1 m^2$ for A_e in (8) and indicated in dBW at point C on Figure 3, but this value is only approximate and is not used in the link analysis.

It was indicated that the contours of Figure 8 are labeled both in terms of the saturation flux density in dBW and the figure of merit G/T of the satellite receiving system as a function of direction. This is done for good reason. It turns out that, instead of C, we need a quantity, C/N_0, the carrier-to-noise power-density ratio when the satellite is saturated. This ratio can be obtained from the G/T and saturation flux density curves of Figure 8.

Before calculating C/N_0 we must define G/T. To do this we will need to discuss noise power, noise temperature, and related subjects.

Noise Power and Noise Temperature. At the satellite a certain amount of electrical noise power N_A enters the satellite receiver via the antenna along with the carrier power C. An additional amount of noise power N_E is generated in the low-level stages of the receiver.[3]

The noise power is uniformly distributed in frequency. The noise power density N_0, is defined as the noise power per unit of bandwidth, usually megahertz or hertz. The latter is used especially where the application is for satellite communications systems that have bandwidths which are narrow compared with 1 MHz (see the paper "Satellite Link Analysis"). We will discuss later why the noise power density is used rather than the total noise power.

Because the noise is random, it adds on a power basis rather than on a voltage basis. The noise power density at a reference point in the receiver, shown in Figure 9, is given by:

$$N_{0S} = N_{0A} + N_{0E} \quad \text{(watts/MHz)}, \qquad (10)$$

where the subscript S indicates "system"; in this case referring to the uplink receiving system, A indicates "antenna", and E indicates "receiver."*

Additional noise is added on the downlink. In our case, most of the noise is added on the downlink, but it will be seen later that the video signal-to-noise ratio for the overall link is affected by both the uplink and downlink noise.

The noise entering or generated within a satellite receiver is from a number of sources, but it can be treated as if it were thermal noise, which is caused by the random motion of electrons in a conductor caused by their thermal energy.

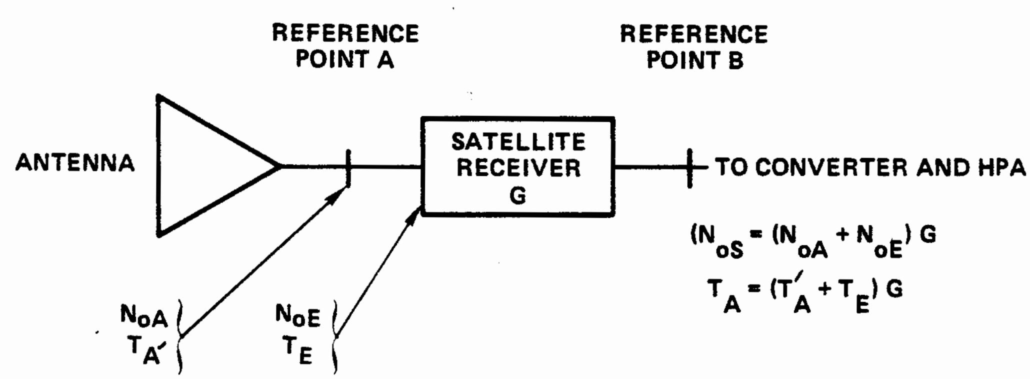

Figure 9. Simple Diagram Showing Noise Power Density and Noise Temperature at Reference Points in a Satellite Receiving System.

It is convenient to measure the noise power or noise power density in terms of an effective noise temperature. It is well known that the thermal noise power N_X at a given reference point of a receiving system** is given by:

$$N_X = kT_XB_N \quad \text{(watts)} \tag{11}$$

where

T_X is the effective noise temperature at the reference point in Kelvins (K),

k is Boltzmann's constant (1.38×10^{-23} Joule/K), and

B_N is the noise bandwidth in Hertz, and where

x identifies the noise power N_S, N_A, N_E.

*The subscript E is used instead of R to adhere to convention; the letter E will stand for "effective" when we convert to noise temperature.
**See the paper "Noise Temperature and G/T."

If we let B_N = 1 Hz, we have from (11)

$$T_x = \frac{N_{ox}}{k} \qquad (12)$$

By division of (10) by k and use of (12) we have

$$T_S = T_A + T_E \quad \text{(Kelvins)} \qquad (13)$$

T_S is called the <u>operating temperature</u> or the <u>system temperature</u>, T_A is the <u>antenna temperature</u>, and T_E is the <u>effective input noise temperature</u> of the receiver measured at the antenna terminals as shown in Figure 9. The designation "system" again refers here to the uplink receiving system.

From (13) and Figure 9, it can be seen that the effective input noise temperature of the receiver defined at a given reference point is the system temperature which would exist if the antenna temperature at the given reference point were zero.

Carrier to Noise-Power Density Ratio. It is customary to express (8) in the form:

$$C = \text{EIRP}\ g_R \left[\frac{\lambda}{4\pi R}\right]^2 p \qquad \text{(watts)} \qquad (14)$$

by using (9), where the direction associated with g_R has been dropped.

The carrier-to-noise power ratio is obtained by dividing (14) by (11), giving:

$$C/N_o = \frac{g_R}{T_S}\left[\frac{\text{EIRP}}{k}\left(\frac{\lambda}{4\pi R}\right)^2 p\right] \qquad \text{(Hz)} \qquad (15)$$

Note that C/N_o in (15) is proportional to g_R/T_S. This ratio, usually converted to decibels and designated simply G/T, is at least to some extent under the control of the satellite designer as opposed to the factors within the brackets. As has been indicated, the ratio G/T, expressed in dB/K, is called the <u>figure of merit</u> of the satellite receiving system.

Conversion of (15) to decibels gives

$$C/N_o(\text{dB-Hz}) = G/T(\text{dB/k}) + \text{EIRP} - 228.6 - L_S - L_p \tag{16}$$

where

EIRP is expressed in dBW,

$$L_p = -\,10 \log p \qquad (\text{dB}) \tag{17}$$

$$L_S = 20 \log (4\pi R/\lambda) \qquad (\text{dB})^* \tag{18}$$

$$228.6 = -10 \log (1.38 \times 10^{-23}) \quad (\text{dBW/Hz-K}) \tag{19}$$

In (16) C/N_O (dB-Hz) and G/T (dB/K) are symbolic notations, respectively.

$$C/N_o\ (\text{dB-Hz}) \equiv C\ (\text{dBW}) - N_o\ (\text{dBW/Hz}) \tag{20}$$

$$G/T\ (\text{dB/K}) \equiv G_R\ (\text{dBi}) - T_S\ (\text{dB-k}) \tag{21}$$

They also represent the numerical ratios C/N_O and g_R/T_S expressed in decibels.

Equation (16) can be written**

$$C/N_O = \text{EIRP} + G/T + 228.6 - L_S - L_P \qquad (\text{dB-Hz}) \tag{22}$$

or

$$C/N_O = \text{EIRP} + G/T + 168.6 - L_S - L_p \qquad (\text{dB-MHz}) \tag{23}$$

where in (23) a constant term 10 log 10^6 (= 60 dB) has been subtracted because C/N_O is defined in dB-MHz.

It is often convenient to eliminate EIRP and the distance from the earth station to the satellite from (15) and express it in terms of the flux density incident on the satellite and g_R/T_S. Using (4) with $k_A = 1$ in (15) gives

$$C/N_o = S\,\frac{g_R}{T_S}\,\frac{\lambda^2}{4\pi}\,\frac{p}{k} \tag{15A}$$

or

$$C/N_o = S\,\frac{g_R}{T_S}\,\frac{0.3^2}{4\pi}\,\frac{1}{f^2}\,\frac{p}{K} \tag{15B}$$

---

*The quantity L_S is usually called "space loss" or, more appropriately, "spreading loss."

**For simplicity we set the atmospheric loss L_A(dB) zero in (5). Inclusion of atmospheric attenuation results in a term $-L_A$ on the right side of (16).

where $\lambda = \frac{c_o}{f}$,

c_o is the velocity of light (3×10^8 meter/sec), and

f is expressed in GHz in (15B).

Conversion of (15B) to decibels gives

$$C/N_o = S + G/T - 21.45 - 20 \log f(\text{GHz}) = 228.6 \text{ (dB-Hz)} \qquad (22A)$$

or

$$C/N_o = S + G/T - 21.45 - 20 \log f(\text{GHz}) + 168.8 \text{ (dB-MHz)} \qquad (23A)$$

It is important to note that both C/N_o and G/T are independent of the reference point in the receiving system amplifier chain at which they are measured. This is because C, N_o, G and T are all changed by the same factor as the reference point is changed.*

In Figure 8 the contours of saturation flux density already discussed coincide with contours of G/T. For our case, G/T equals -4 dB/K for the signal path to the satellite from New York City. We will assume L_p to be zero and we will calculate L_s to be 199.6 dB for the distance from New York City to the satellite of 37,750 km by using (18).

From (23) we have:

$$C/N_o = 80 + 168.6 - 4 - 199.6 = 45 \quad \text{(dB-MHz)} \qquad (24)$$

As a check we have from (23A)

$$C/N_o = -82.5 - 4 - 21.5 - 15.6 + 168.6 = 45 \text{ (dB-MHz)} \qquad (24A)$$

As has been indicated, we are not given the gain or the effective area of the satellite antenna in Figure 8. We are given the value of flux density which is required to saturate the satellite receiver and the G/T of the satellite receiving system. In our calculations we did not find either C or N_o, only the quantity C/N_o, but this was the value needed.

It will be shown later that for this value of C/N_o, the uplink contributes only a small fraction of the total noise of the overall link, and that the downlink, as in typical cases, contributes most of the noise. This is as it should be in a well designed uplink because if the uplink is noisy, it is not possible to realize a high-quality overall link, regardless of the quality of the downlink.

*See the paper "Noise Temperature and G/T."

Downlink

Satellite TWT. As shown in Figure 7, the receiver feeds a downconverter, which offsets the frequency and drives the downlink TWT output power amplifier. In keeping with our simplifying assumptions on page 1, the frequency offset is shown as 2 GHz instead of 2.225 GHz.

The transponder TWT maximum output is relatively low, typically between 5 and 10 watts. In Figure 3 we use 5 watts (7 dBW). Figure 10 shows a typical TWT power input-output curve.

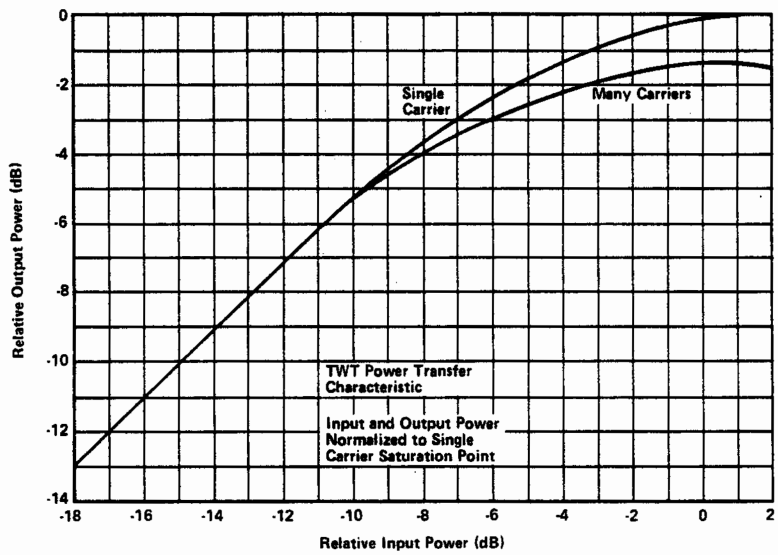

Figure 10. TWT Power-Transfer Characteristic, Showing Linear and Saturation Regions

When only one carrier exists, as in our case, the TWT is usually operated under saturated conditions. When more than one carrier is present, as is often the case in many applications, the TWT has to be operated in its linear region to control intermodulation distortion. This requires a backoff in input power, and this produces a decrease in both uplink and downlink C/N_0.

Satellite EIRP. The satellite output is measured by the EIRP, defined by (1) or (2). The satellite EIRP is lower than that of the uplink earth station because of two factors. First, the power obtained from the solar panels which power the satellite is limited by satellite size and launch costs. Second, downlink coverage is provided over a wide area (usually somewhat the same as the uplink G/T footprint), and this limits the gain of the satellite transmitting antenna. A typical downlink footprint, described by contours of constant EIRP, is shown in Figure 11.

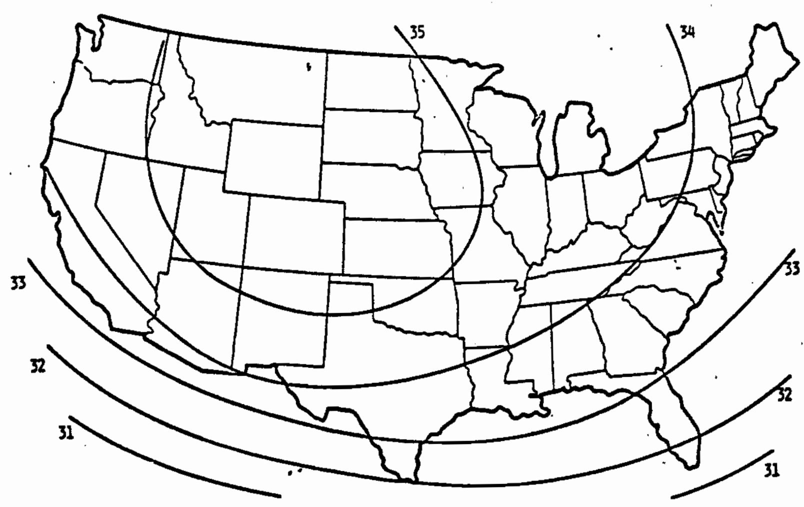

Figure 11. Footprint Showing Assumed Satellite EIRP in dBW at 4 GHz. (Figure actually shows average EIRP of RCA SATCOM F-1 at 135°W for transponders 3,7,11,15,19, and 23)

The operation of the downlink is basically similar to that of the uplink except for the lower EIRP produced by the satellite.

Receiving Earth Station. A simplified block diagram of the receiving earth station is shown in Figure 12. It consists of the earth-station antenna, a low-noise amplifier (LNA), a receiver, and ancillary equipment.

As in the case of the satellite, the receiving earth station is characterized by a figure-of-merit G/T, whose required value is determined by the originally postulated video signal-to-noise ratio of 54 dB. We will arrive at this required value in the following paragraphs.

The downlink carrier-to-noise ratio is again given by (23), where the various terms represent downlink parameters.

It may be interesting to discuss why C/N_0 is used instead of C/N in analyzing the link performance. The relationship between C/N_0 and C/N is given by:

$$C/N_0 \text{ (dB-MHz)} = C/N \text{ (dB)} + 10 \log B_N \text{ (MHz)} \qquad (25)$$

where B_N is the IF bandwidth of the earth station receiver.

C/N_0 is used because the video signal-to-noise ratio for a given video-signal deviation is determined by C/N_0 and the noise bandwidth of the baseband filter function[4] of the receiver, and not by the predetection carrier-to-noise power ratio C/N. For a U. S. television signal with CCIR weighting and with peak deviation of 10.75 MHz, the video signal-to-noise ratio as defined in reference 4 is given by:

$$S/N = C/N_0 \text{ (dB-MHz)} + 22.6 \text{ (dB)} \tag{26}$$

For the postulated video signal-to-noise ratio of 54 dB, the required value of C/N_0 is thus 31.4 dB-MHz. This is actually the required overall-link C/N_0. We will show that the required downlink C/N_0 is slightly higher, 31.6 dB-MHz when the uplink noise is taken into account. We will use this latter value in our downlink calculations.

Rearranging (23) gives:

$$G/T = C/N_0 - EIRP + L_S - 168.6 \quad \text{(dB/K)} \tag{27}$$

where L_p is assumed to be zero for the downlink, as it was for the uplink.

For our example we will read a satellite EIRP of 33.5 dBW in the direction to Los Angeles from Figure 11. For the distance from Los Angeles to the satellite, L_S at 4 GHz is found to be 196 dB.

Substitution of the appropriate parameters in (23) gives:

$$G/T = 31.6 - 33.5 + 196 - 168.6 = 25.5 \quad \text{(dB/K)} \tag{28}$$

This is the figure of merit of the earth station receiving system which is required to provide a downlink C/N_0 of 31.6 dB-MHz under the assumed conditions.

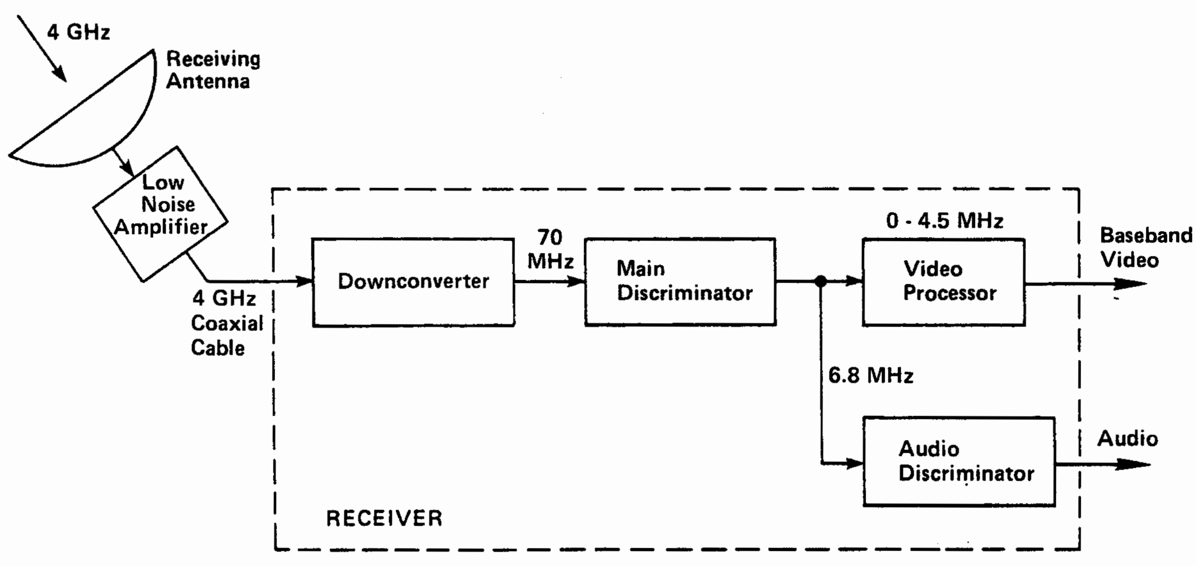

Figure 12. Simplified Block Diagram of Assumed Satellite Receiving Earth Station. The LNA is bolted directly to the output waveguide of the receiving antenna to eliminate losses which would otherwise increase the system noise temperature

Receiving-System Noise Temperature. As in the case of the uplink, T_S is given by (13). The antenna temperature T_A of earth stations can be made very low because the earth station beam looks at the cold sky.[5]* T_A for the chosen antenna varies from 46K at an elevation angle of 5 degrees to 18K at zenith. At an elevation angle of 34 degrees, it is approximately 20K.

From the definition of G/T given in (21), the required G/T given by (28) is obtained by subtracting the receiving system noise temperature from the receiving antenna gain. Thus, we can trade off one against the other. This is illustrated by the graph of Figure 13.

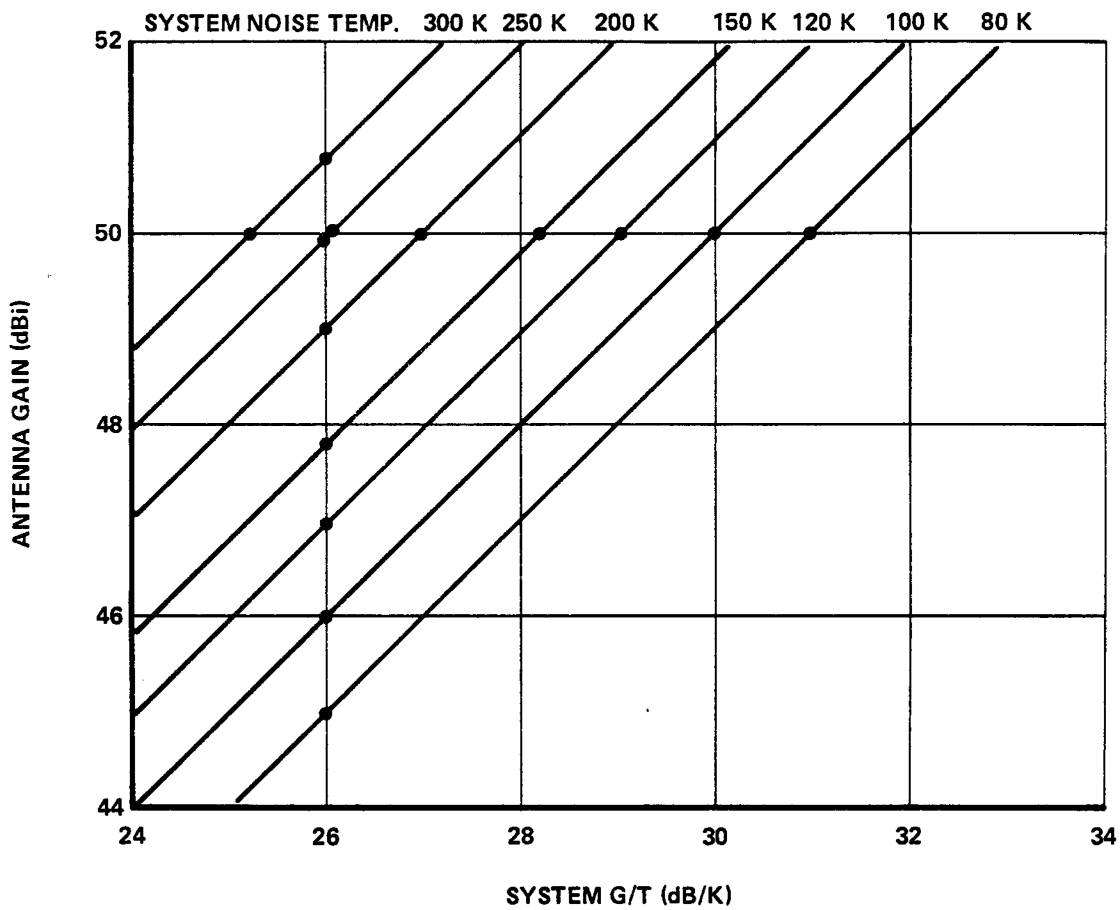

Figure 13. System G/T vs. Antenna Gain

In the figure the horizontal row of dots at an antenna gain of 50 dBi represents an increase in system G/T of 5.74 dB, from 25.23 dB/K, to 30.97 dB, as the system noise temperature decreases from 300K (24.77 dB/K) to 80K (19.03 dB/K) if the antenna gain is maintained constant at 50 dBi.

The vertical row of dots at a G/T of 26 dB/K represents a permissible decrease in antenna gain from 50.77 dBi at 300K to 45.03 dBi at 80K to maintain the indicated G/T.

*See the paper "Noise Temperature and G/T.

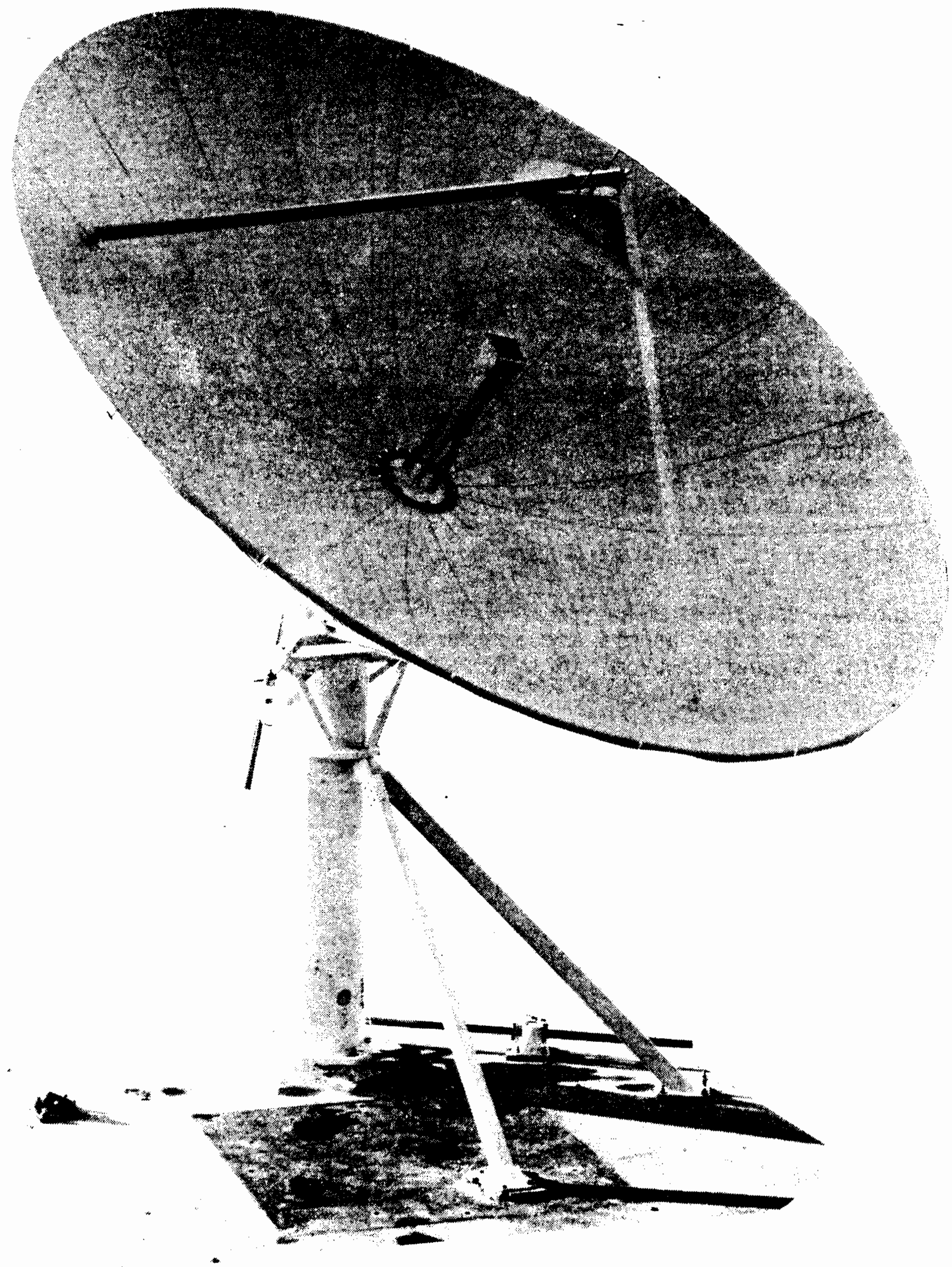

Figure 14. Scientific-Atlanta Model 8010 7-Meter Satellite Communications Antenna

The Scientific-Atlanta Model 8010 7-Meter Antenna shown in Figure 14 has a gain of 47.5 dB at 4 GHZ. We will choose this antenna, which results in a maximum allowable system noise temperature T_S of 22.0 dB/K for the G/T of 25.5 dB/K. This T_S translates to 158 Kelvins [= 10 exp(22.0/10)]. We will find that this maximum temperature is easily realizable at moderate cost.

If we were to choose a 5-meter antenna with a gain of 44.5 dB, the resulting required system noise temperature would be 19.0 dB/K or 79 Kelvins. From (13) the antenna temperature is subtracted from the maximum allowable system noise temperature to determine the maximum allowable receiving system noise temperature.*

It will be found that T_E in (13) for a well-designed earth station is almost identically equal to T_{LNA}, the temperature of the low-noise amplifier. In our case the error is less than 1 Kelvin in the allowable noise temperature. If we assume an antenna temperature of 20 degrees, the required LNA temperature is 59 Kelvins. This noise temperature is not currently realizable with an uncooled GaAs FET LNA.

A 5-meter-diameter antenna would give an S/N of 51 decibels with a T_S of 158 Kelvins. This S/N is adequate for many applications.

As indicated, we will choose a 7-meter antenna and a moderate-cost 100K LNA. For an antenna noise temperature of 20K, the downlink system noise temperature is 120K, or 20.8 dB/K. From (21) the resulting G/T is given by:

$$G/T = 47.5 - 20.8 = 26.7 \text{ dB/K} \qquad (29)$$

This choice represents a margin of 1.2 dB over the value of 25.5 dB/K given by (28). This would permit the system noise temperature to rise as high as 158 Kelvins without compromising the system S/N specifications.

LNA

Low-noise amplifiers (LNAs) used in receiving earth stations (Figure 12) are GaAs FET (Gallium-Arsenide-Field-Effect-Transistor) amplifiers. Research in recent years has reduced the noise figures** of GaAs FETs to the point that uncooled amplifiers as low as 80K are available. The LNA is bolted directly to the output waveguide flange of the receiving antenna, where T_A is defined, to eliminate noise contributions which would be introduced by waveguide losses if the LNA were separated from the antenna.

In our example, the LNA has a gain of 50 dB. The receiver is located in a convenient position, some distance from the antenna, and is connected to the LNA output by a coaxial cable whose loss is assumed to be 10 dB. The high gain of the LNA almost completely overrides losses and noise contributions arising in the cable and later circuits.

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* See the paper "Noise Temperature and G/T."

**Noise figure is given by $NF(\text{dB}) = 10 \log \left(\frac{T_E}{290} + 1\right)$. Thus, a 290K effective noise temperature T_E is equivalent to a 3-dB noise figure.

Receiver

The receiver (Figure 12) converts the frequency of the incoming signal to an intermediate frequency (IF). It rejects unwanted signals in the process, demodulates the IF signals by means of the main IF discriminator and processes the composite baseband video to restore it to is pre-transmission format. The audio subcarrier is separated from the main IF and is demodulated by the subcarrier discriminator. Baseband video and audio are available at separate output coaxial connectors.

Overall-Link C/N_0 and C/N

From (26) it was determined that the required value of overall-link C/N_0 was 31.4 dB-MHz. Because of the contribution of the uplink noise to the overall-link noise, the required downlink ratio must be slightly greater than 31.4 dB-MHz to give an overall-link C/N_0 of 31.4 dB-MHz. This is shown below.

The overall-link C/N_0 results from adding the noise-power densities of the uplink and downlink referred to the same carrier levels:

$$(N_0/C)_0 = (N_0/C)_U + (N_0/C)_D \tag{30}$$

where the subscripts 0, U and D represent "overall-link", "uplink" and "downlink", respectively, and where the terms in parentheses are numerical ratios (not decibels).

Thus, for given overall and uplink carrier-to-noise power densities,

$$(N_0/C)_D = (N_0/C)_0 - (N_0/C)_U \tag{31}$$

In our example,

$$(N_0/C)_0 = \frac{1}{10^{31.4/10}} = \frac{1}{1380} \tag{32}$$

and

$$(N_0/C)_U = \frac{1}{10^{45/10}} = \frac{1}{31623} \tag{33}$$

From (31),

$$(N_0/C)_D = \frac{1}{1380} - \frac{1}{31623} = \frac{1}{1443} \tag{34}$$

The required downlink carrier-to-noise-power density C/N_0 is thus given by:

$$(C/N_0)_D = 10 \log 1443 = 31.6 \text{ (dB-MHz)} \qquad (35)$$

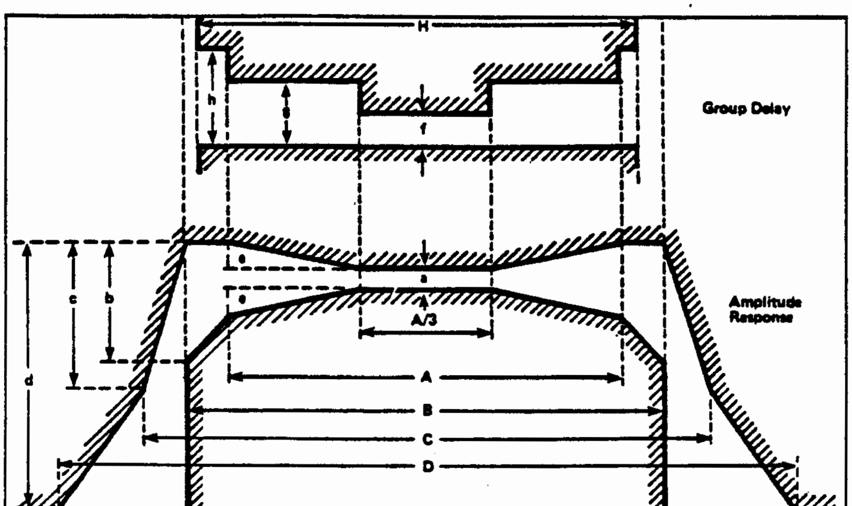
The final system C/N is determined from (25), where B_N is the earth-station receiver IF bandwidth (except in the unlikely event that the satellite transponder bandwidth is narrower than the receiver bandwidth and determines the system bandwidth).

Link Performance

Distortion. At the beginning of our example, we set a specification of 54 dB for the video signal-to-noise ratio (S/N) and indicated that distortion limits would be discussed.

The video S/N criterion is satisfied by the carrier-to-noise power-density ratio of the overall-link for the FM deviation used as indicated by (26). Linear and nonlinear distortions of the output waveform are introduced by variations in the amplitude and group-delay responses of the transmitter, satellite and receiver. To maintain distortions within acceptable limits, specifications have been established which set limits on the group delay and amplitude responses of the receiver and the transmitter. These specifications are defined by a mask within which the responses must fall. An example of such a mask is shown in Figure 15.

With reference to the distortion limits for our example, after the complete link has been installed, the uplink amplitude and group-delay responses are adjusted to fall within limits set by masks such as those of Figure 13. If the receiver has been compensated separately such that its response falls within the same mask, the total variation in group-delay response will be approximately twice that defined by the mask.



	BW (MHz)	A (MHz)	B (MHz)	C (MHz)	D (MHz)	a (dB)	b (dB)	c (dB)	d (dB)	e (dB)	H (MHz)	f (ns)	g (ns)	h (ns)
131630	36.0	28.8	36.0	45.25	60.0	0.6	2.5	10.0	25	0.3	33.1	3	5	15

Figure 15. INTELSAT 36-MHz Group-Delay and Amplitude-Response Masks

In addition to the RF tests described above, real-time baseband measurements are made as in testing any standard video link to determine total input-to-output link distortion.

Threshold

In satellite FM/video systems, as indicated in our example, a high deviation (10.75 MHz) and, hence, a wide occupied bandwidth (approximately 36 MHz) are usually used to gain an increase in video signal-to-noise ratio (S/N) over the IF carrier-to-noise power-density ratio C/N_0.

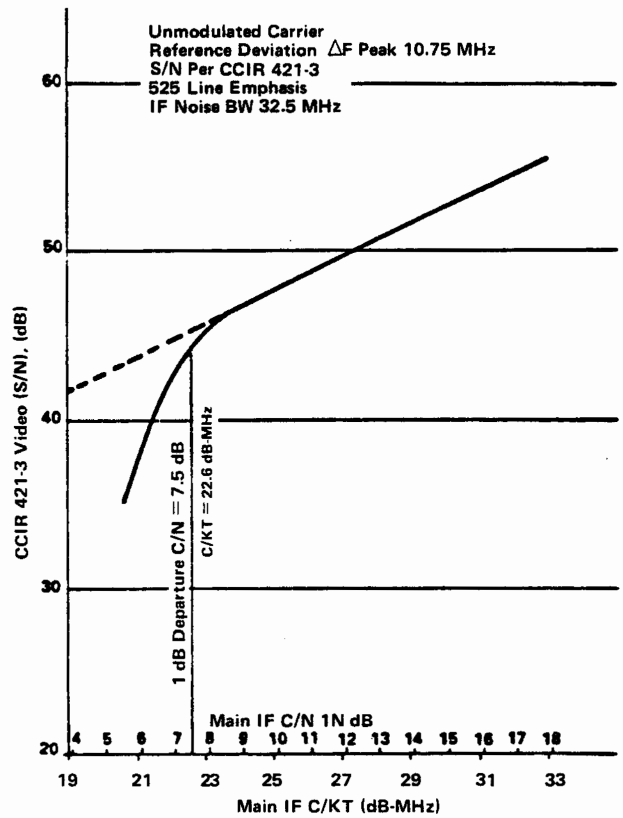


Figure 16. Illustration of FM/Video Receiver Threshold

In the example, we discussed noise in relation to the value of C/N_0 which was required to produce a desired value of S/N. This enhancement in S/N occurs only if C/N is above a certain threshold level. The effect of threshold is shown by Figure 16, which is a graph of S/N as a function of C/N of a typical video receiver.

When the C/N ratio at the discriminator is above the threshold level, there is a one-to-one ratio between video signal-to-noise power and IF carrier-to-noise power. As the IF C/N ratio passes through threshold, random noise peaks begin to cause the instantaneous envelope of the carrier-plus-noise voltage to pass through zero. When this occurs sudden phase excursions of the IF signal occur, which the discriminator interprets as being caused by large instantaneous frequency changes. This results in impulse noise spikes in the baseband signal, causing white-to-black and black-to-white streaks on the video display. The rate of occurrence increases as C/N decreases. As C/N decreases still further, loss of sync occurs, and the picture is lost. Satellite communications systems of other types, such as digital links, may or may not exhibit pronounced thresholds, depending on the demodulation process.

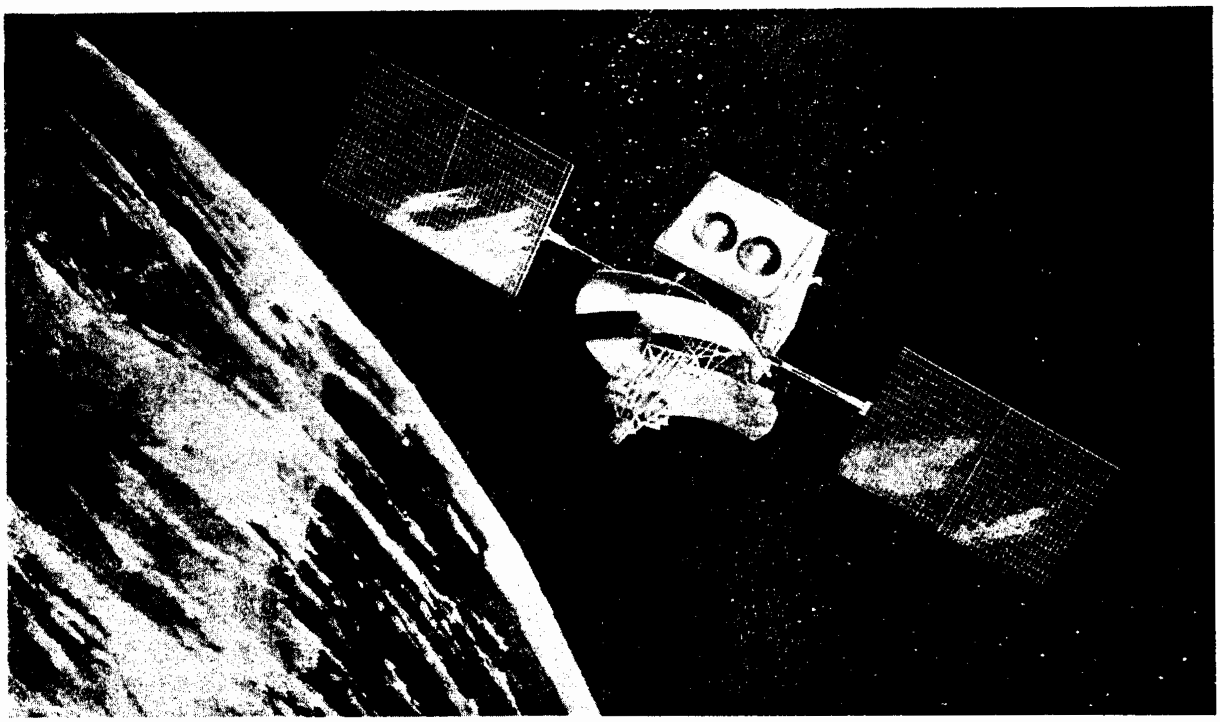
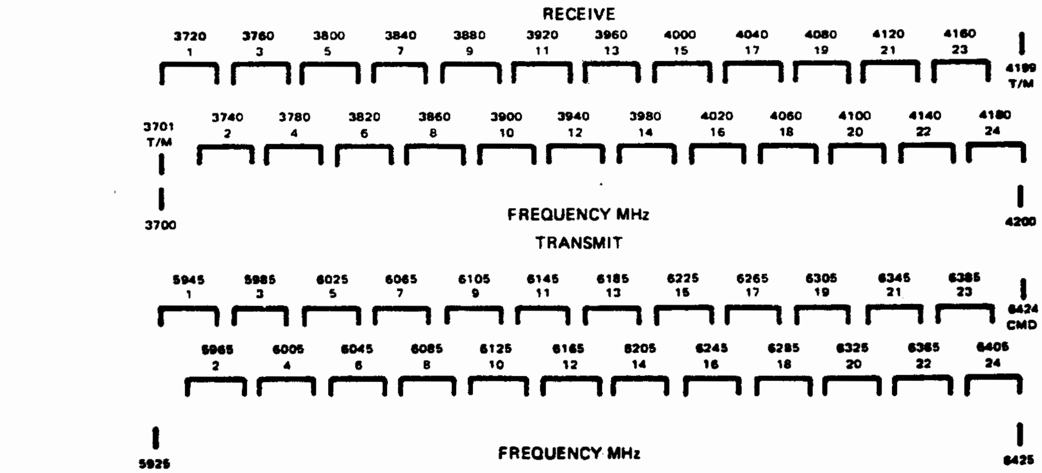


Figure 17. RCA Satcom Satellite

Satellites

The hypothetical satellite in our example was assumed to contain a single transponder for simplicity. As already stated, operational 6/4-GHz satellites in use over the U.S. have either 12 or 24 transponders.[6] They of course contain block converters, diplexers, and other circuits not shown in the simple satellite of Figure 8.

Figure 17 is a picture of an RCA SATCOM satellite, as an example, and Figure 18 is a transmit and receive frequency plan of its 24 transponders. The COMSTAR satellite also has 24 transponders, while the older generation WESTAR and ANIK satellites have 12 transponders. The 24 transponders in the newer satellites are made possible by the technique of frequency reuse, which is discussed in the next section.



NOTE: NUMBER BELOW CHANNEL CENTER FREQUENCY REFERS TO TRANSPONDER IDENTITY.

Figure 18. Transmit-Receive Frequency Plan of RCA Satcom Satellite

If left alone, a synchronous satellite would finally drift out of position. To overcome this, it is continuously monitored by a telemetry tracking and control (TT&C) station. Small jets of a propellant such as hydrazine are used to keep it within a "station-keeping box," and sufficient propellant is carried on board the satellite for its predicted life, usually 7 to 10 years. Station-keeping boxes for 6/4-GHZ satellites serving the USA are ±0.1 degree on each side. This amount of drift is small enough that antennas which are less than about 10- or 11-meters in diameter can usually be left stationary. Larger antennas may contain a sensing mechanism to keep the antenna beam peaked on the satellite.

The solar panels, which power the satellites by converting sunlight directly to electricity, have to face the sun to be effective. The attitude of the satellite also has to be stabilized to keep its antennas pointing in the desired direction or directions.

Some satellites are designed so that the solar panels are continuously pointed toward the sun with their antennas independently pointing toward the earth, as in the case of SATCOM and INTELSAT V. Many satellites, including INTELSAT's I through IVA, WESTAR, ANIK, and COMSTAR are cylindrical in shape, and the solar cells are mounted around the periphery. The bodies of these satellites spin about their axes for stabilization, with the antennas always pointing to the earth. In this way, about one-third of the cells effectively face the sun at a given time. Batteries are used in almost all satellites to take care of solar panel outages during eclipses of the satellite by the earth.

Frequency Reuse

Consideration of the SATCOM frequency plan of Figure 18 shows that the odd-channel sidebands overlap those of the even channels. This overlap occurs because the newer satellites are designed to transmit and receive signals simultaneously on two orthogonal polarizations, increasing from 12 to 24 the number of transponders in the available 500-MHz band.

The polarizations of all of the U.S. and Canadian domestic satellites are linear. Linear polarizations which are at right angles to each other are electrically orthogonal. If the polarization of the receiving antenna is orthogonal to that of the incident wave, the polarization efficiency p in (6) is zero, and no power is received from the wave. The orthogonal polarization can then be used to receive independent information.

Frequency reuse is made practical by staggering the carriers of the odd and even channels so that only sideband energy overlaps. The carriers are staggered because it would be difficult to design systems which reject the unwanted signal by polarization discrimination alone. Staggering the carriers reduces the required system polarization discrimination to about 25 dB, a value which is easily achievable except when exceptionally hard rains cause additional depolarization. Even under conditions of severe depolarization, which occur only occasionally, the interference to the orthogonal channel is usually not extremely severe because of the protection afforded by the staggered-carrier frequency plan.

Carson's Rule

Virtually all satellite communications systems use some form of angle modulation (such as, for example, frequency modulation). To generate angle modulation, the phase or the frequency, which is proportional to the time derivative of the phase, is varied linearly with the modulating signal.

The occupied bandwidth for frequency modulation is determined by both the frequency deviation of the carrier and by the frequencies contained in the modulating signal. Specifically, the FM bandwidth (that occupied by sidebands of significant level) is given by Carson's rule:[7]

$$B_C = 2(f_m + \Delta f) \tag{36}$$

where:

B_C is the Carson's-rule bandwidth,

f_m is the highest modulation frequency of interest, and

Δf is the peak deviation of the carrier.

As already indicated, in domestic-satellite FM/video systems a high deviation, hence, a wide occupied bandwidth is used to gain a significant increase in video signal-to-noise ratio over the IF carrier-to-noise density ratio C/N_0.

Modulation Formats and Access Techniques

The modulation techniques employed for various satellite communications applications depend on the particular requirements of the application. Video and voice signals are commonly transmitted by analog systems such as we have described, while computer data, for example, are transmitted by digital systems.

Analog-to-digital converters are often used to transmit analog signals by digital systems. The digital signals are then reconverted to analog form at the receiving end of the system by digital-to-analog converters. Such systems are coming into increasing use as the costs of digital circuits decrease. In digital systems, modulation formats are usually used which give a low-bit error rate for a given ratio E_b/N_0, where E_b is the energy per bit.

Existing satellites and associated earth terminal equipment are designed with modulation formats for a wide range of data rates from television signals or high-data-rate streams of digital data to low data rates such as are represented by individual audio channels.

Where many narrowband channels such as voice-grade circuits are required, the narrowband channels are frequency-division-multiplexed (FDM), and these signals are used to frequency-modulate a carrier. This modulation method is called FDM/FM.

For full-transponder video or multiple-channel voice-grade circuits, each transponder is accessed by a single earth station at a given time. For lower-capacity voice, data and half-transponder video applications, a transponder will be shared by a number of earth stations, each of which may use a number of carrier frequencies. This technique is called frequency-division multiple access (FDMA).

In one of the important applications of FDMA, a single voice-grade channel is transmitted on each carrier. This approach, designated single-channel-per-carrier (SCPC), uses carriers spaced at 60 kHz or less.

Although SCPC requires a high degree of frequency stability of the individual carriers, it permits installation of stations which need only a limited number of voice-grade channels and permits the easy addition of more channels as required.

SCPC with Demand Assignment Multiple Access (DAMA) is a technique in which each earth station uses a channel only as required. When the earth station is not using a channel, it is available for use by other earth stations. DAMA greatly increases the average number of voice channels that can be carried by one transponder. It is already in operation in the INTELSAT SPADE system and on the INMARSAT/MARISAT system, and it is beginning to find increasing application in domestic and foreign thin-route systems.

Biphase and quadriphase digital modulation (PSK and QPSK) formats are becoming increasingly important as advances in digital technology continue, and as the costs of digital circuitry decrease.

Systems with time-division-multiplex multiple access (TDMA) and SSTDMA (TDMA with satellite switching) are also in use or under development. Satellite Business Systems (SBS), a partnership among wholly-owned subsidiaries of COMSAT General Corporation, IBM, and Aetna Life & Casualty Company, has implemented an extensive TDMA multipoint communications system using a single, time-shared carrier per transponder with transmission using QPSK in a burst mode. The system operates at 12 and 14 GHz, and the earth stations can be located directly on customers' premises.

SSTDMA uses high-gain spot beams in a switching matrix to provide interconnectivity of multiple antenna beams. The high gain provided by the spot beams helps to reduce the required satellite EIRP or to overcome problems caused by rain attenuation where the systems are designed for 12 and 14 GHz.

These and various other techniques are used to take advantage of available spectrum and to provide system flexibility to accommodate the various categories of users and applications.

Effect of Frequency on Satellite Communications Systems

Because of the virtually unlimited applications for satellite communications and the limited information capacity of the 6/4-GHz band, the higher frequency satellite bands, especially the 14/12-GHz band, will come into greater use in the future. It is therefore interesting and important to consider the effect of increasing frequency on the design and performance of satellite communications systems.

The major plus-factors resulting from going to higher frequencies are (1) the availability of more spectrum, (2) the fact that less interference exists from terrestrial microwave systems, and (3) the production of higher gain by antennas of a given aperture size.

Negative factors are (1) the increase in atmospheric attenuation with frequency (especially that due to rain), (2) the decrease in sensitivity (increase in noise temperature) of receivers, (3) the narrower beamwidths produced by antennas of given aperture size, and (4) the smaller surface tolerances required to prevent degradation of antenna performance at the shorter wavelengths.

Consider the effect of frequency on a receiving earth station antenna. It is well known that the gain of an antenna of a given diameter increases as the square of the frequency. This increase in gain can lead one to think that an antenna of a given size receives more power for a given satellite EIRP as the frequency increases. This is, of course, not true because the received power is given by (8), which is independent of frequency if the effective area of the antenna is constant. The realizable effective area is, in fact, essentially independent of frequency for reflector-type antennas of a fixed size.

The result is that, as the frequency increases, the major effects are (1) narrowing of the beam, (2) increase in losses and (3) increase in the required surface accuracy of the reflector. For a given satellite EIRP, the increase in frequency increases the earth station receiving antenna cost in at least three ways:

- by the tighter surface tolerance, which requires a more accurate and stiffer reflector,
- by the increased difficulty of pointing the antenna toward the satellite, which requires a costlier antenna mounting structure, and
- by the fact that the increase in atmospheric attenuation and the inherent increase in LNA noise temperature force a larger antenna diameter, which acts to reinforce the difficulties associated with the first two factors.

The net result is that if earth station costs are to be kept low, the EIRP of the satellite must be greater at the higher frequency band so that smaller earth station antennas can be used. This increases the cost of the satellite because the increase in EIRP must come from either higher-power transponders or from footprints covering smaller areas.

The smaller footprints do not represent a disadvantage where a small area is to be covered, such as Japan, or a country in Europe. On the other hand, where a large area is to be covered such as the United States, the smaller footprints require multiple beams, each with some number of transponders determined by the level of service to be provided. The net effect is an increase in satellite solar-panel power requirements. Typical system designs for low-cost receiving systems are based on four or five beams covering CONUS (continental USA).

In spite of the problems indicated above, a move to higher frequencies for new satellites is inevitable for the United States because of orbital crowding and use of the 6/4 GHz band by terrestrial and satellite systems. Systems are already being implemented for the 14/12-GHz band. In fact, once sufficient cost is transferred to the satellite so that small antennas can be used and sufficient system information capacity is provided, many new applications such as direct broadcast satellite (DBS) open up, for example. It is evident, however, that satellite systems at 6/4 GHz are here to stay because of their inherently low cost and because of the existing frequency spectrum.

Summary and Conclusions

This paper has introduced satellite communications by describing a hypothetical FM/video link and discussing certain of the more important topics. As indicated, the field of satellite communications is complex and multifaceted, and the surface has only been skimmed. Details of various theoretical and practical aspects, such as antenna and equipment design and fabrication, installation of earth stations, provision for redundancy and verification of performance are contained in other papers.

Appendix A

Decibels and Decibel Designations

The decibel, abbreviated dB, is a logarithmic unit used to measure the ratio between power levels. By definition,

$$N(dB) = 10 \log P_1/P_2, \tag{1-A}$$

where N is the number of decibels and P_1/P_2 is a power ratio.

When P_1/P_2 is greater than unity, N is positive; when P_1/P_2 is less than unity, N is negative. The fraction is often inverted when the power ratio is less than unity, and the ratio is expressed as a decibel loss.

Decibels are advantageous primarily because they transform multiplication and division into addition and subtraction. If n_1 and n_2 are power ratios whose values in dB are N_1 and N_2, the product $n_1 n_2$ is represented by $(N_1 + N_2)$ dB.

To permit use of decibels in specifying power levels, P_2 of equation (1-A) is set equal to a specific value of power. The resulting power levels are usually expressed in dBW or dBm, where P_2 is set equal to one watt or milliwatt, respectively.

In satellite communications and in other applications, it is often convenient to convert numbers which are not dimensionless power ratios to decibel form. The resulting numbers are not true decibels, and it is customary to indicate this by their decibel designations.

The process of determining the designation is to include the residual dimension which is not a power ratio in the designation suffix. Some examples used in satellite communications are:

T (Kelvins) → dB-K

Frequency (Hz) → dB-Hz

C/N_0 [carrier power (watts)/noise power (watts/Hz)] → dB-Hz

Noise power density (watts/Hertz) → dBW/Hz

Flux density (watts/meter^2) → dBW/m^2

Watts → dBW, Milliwatts → dBm

Gain → dBi (decibels relative to the gain of an isotropic radiator)

A dash and a slash are commonly used to indicate whether the dimension represented by a designation is in the numerator or the demoninator.

It is a common practice to not use a dash in dBW, dBm and dBi. The suffix i on dBi is often omitted because the isotropic reference is accepted as standard in satellite communications. It must be retained, however, if its omission introduces an ambiguity. For example, it is essential to know whether the isotropic gain of a sidelobe or the sidelobe suppression below the peak of the antenna beam is meant.

Where factors are employed which are raised to powers, the decibel value can be multiplied by the power and the power reduced to unity. Note the definition of space loss L_S given with (13) and compare with (8). See also decibels in the Glossary, Section 5C-1.

Example of Decibel Designations

Equation (12) is rewritten below as (2-A) with the residual dimensions of each factor indicated by (3-A) and the corresponding decibel designations by (4-A):

$$C/N_0 \quad = \quad g_R \quad \frac{1}{T_s} \quad \mathrm{EIRP} \quad \frac{1}{k} \quad \left(\frac{\lambda}{4\pi R}\right)^2 \quad p \qquad (2\text{-A})$$

$$[\mathrm{Hz}] \quad \leftrightarrow \quad [\mathrm{none}] \quad [1/\mathrm{K}] \quad [\mathrm{W}] \quad \left[\frac{\mathrm{Hz\ K}}{\mathrm{W}}\right] \quad [\mathrm{none}] \quad [\mathrm{none}] \qquad (3\text{-A})$$

$$(\mathrm{dB\text{-}Hz}) \quad \leftrightarrow \quad (\mathrm{dBi}) \quad (\mathrm{dB/K}) \quad (\mathrm{dBW}) \quad \left(\mathrm{dB} - \frac{\mathrm{Hz\ K}}{\mathrm{W}}\right) \quad (\mathrm{dB}) \quad (\mathrm{dB}) \qquad (4\text{-A})$$

The dimensions on both sides of (3-A) are the same, since dimensions in terms 2 and 3 of the right side cancel with the K and W in term 4. The corresponding designations of the decibel equations also cancel in the same manner. The terms whose designations are dB or dBi represent dimensionless factors.

The reader is warned that the procedures involving decibel designations outlined herein are not universally adhered to. Some writers of articles on satellite communications choose to ignore decibel designations completely. We believe, however, that the procedures indicated are in keeping with general practice in the field.

References

1. Lorimer Clayton, Jr., "FM Television Signal-to-Noise Ratio," IEEE Transactions on Cable Television, October 1976, pp. 25-30.

2. Hollis, Clayton and Lyon, Microwave Antenna Measurements, Scientific-Atlanta, Inc., 1970, Chapter 2.

3. Mischa Schwartz, Information Transmission, Modulation, and Noise, McGraw-Hill, 1959.

4. L. Clayton, op. cit.

5. J.D. Kraus, Radio Astronomy, McGraw-Hill, 1966, p. 99.

6. See Communications Satellite Systems: An Overview of the Technology, edited by R.G. Gould and Y.F. Lum, IEEE Press, 1976, pp. 37-53.

7. BTL Staff, Transmission Systems for Communications, Bell Telephone Laboratories, 4th Edition 1971, p. 115.

Bibliography

A comprehensive bibliography of the literature on satellite communications is given in the following book:

Harry L. Van Trees (editor), Satellite Communications, IEEE Press, 1980.

Noise Temperature and G/T of Satellite Receiving Systems

J. Searcy Hollis

Introduction

Noise is the combination of unwanted disturbances that tend to obscure the information content of a signal.

The limiting sensitivity of a receiving system is determined by the ratio of received signal power to noise power which just satisfies an established criterion. The criterion may be that the signal is barely detectable or that it exceeds some specified signal-to-noise ratio. The latter is the case in most practical communications systems.

Although certain types of noise, for example ignition noise and radar signals, are periodic in nature, noise which is of routine importance to satellite communications systems can usually be assumed random with continuous spectra. It will be seen that noise temperature is a convenient measure of noise power with continuous spectra.

Sources of Noise

The major sources of noise in the indicated context can be separated into external noise and internal, or circuit, noise. These sources of noise are illustrated in Figure 1.

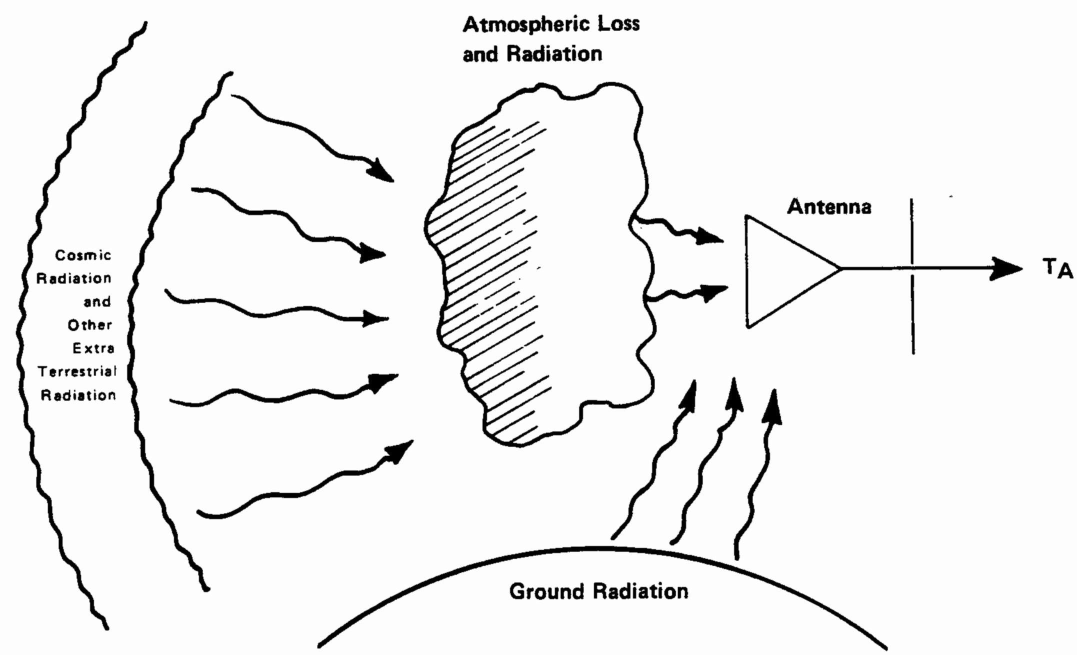


Figure 1. Sketch Showing Elements Which Contribute to the Noise Temperature of a Communications Receiving System

Internal noise can be broken into thermal noise and other forms of circuit noise, such as shot noise in vacuum tubes, current noise in semiconductors and movements of domain boundaries in ferromagnetic devices. Thermal noise is random electrical noise caused by the motion of free electrons in conductors. Thermal noise power is proportional to absolute temperature, which is a measure of the thermodynamic energy of the conductor.

External noise is largely due to extra-terrestrial sources and thermal radiation from the atmosphere and the earth. Cosmic noise is a low level of extra-terrestrial radiation that appears to come from all directions. It is considered to be residual radiation due to the events that occurred during the origin of the universe.

The sun is an extremely strong source of noise, which can interrupt satellite communications when it passes behind the satellite being used and thus lies in the main lobe of an earth station's receiving antenna pattern. The moon is a much weaker source, which is relatively innocuous to satellite communications. Its radiation is due to its own temperature and to reflected radiation from the sun.

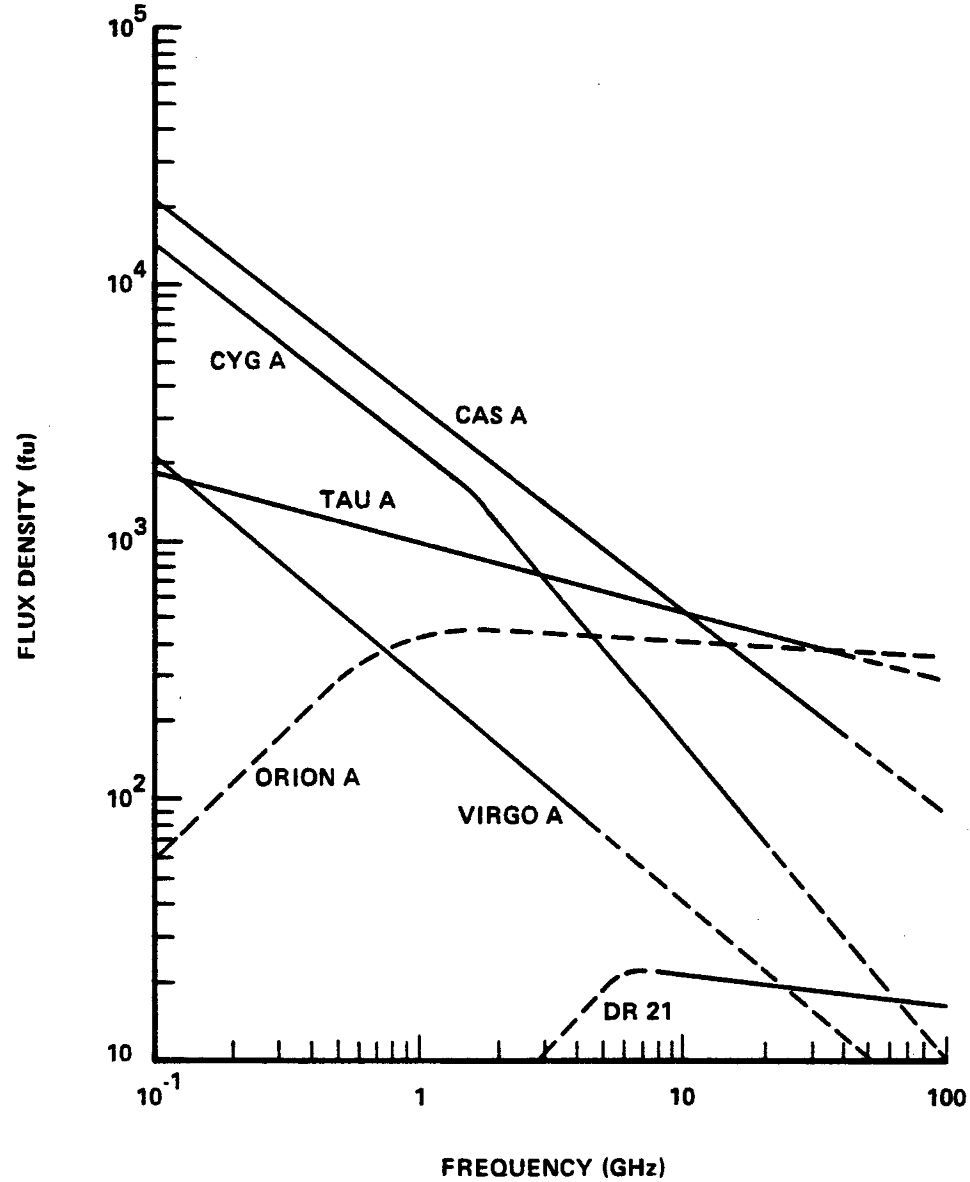
The atmosphere affects external noise in two ways. It attenuates noise passing through it, and it generates thermal noise because of the energy of its constituents. Ground radiation, which includes radiation from objects of all kinds, is also thermal in nature.

Radio Stars. Mapping the sky with radio telescopes has disclosed "radio stars" in addition to the background cosmic noise. These are discrete sources of noise which emit energy in the radio and microwave regions of the spectrum. Figure 2 is a graph which shows the flux density of several radio stars, and Table 1 gives technical information about each.

Table 1. Radio Source Data

		Position		Size	Visibility	
Radio Star	Type*	RA^h	Dec°	RA' x DEC'	NL°	to SL°
Cas A	SR	23.4	58.6	4 x 4	90	11
Tau A	SR	5.5	22.0	3.3 x 4	90	48
Orion A	EN	5.5	-5.4	3.5 x 3.5	65	75
Cyg A	RG	20.0	40.6	1.6 x 1	90	29
Virgo A	RG	12.5	12.7	1 x 1.8	73	57
DR 21	EN	20.6	42.2	< 0.3	90	28

*SR = Supernova Remnant
EN = Emission Nebula
RG = Radio Galaxy



From NBSIR 74–382

Figure 2. Flux Density Spectra of Several Radio Stars

The flux density incident from these stars is not high enough to represent a problem in satellite communications, but certain stars, especially Cas A, which is in the constellation Cassiopeia, are strong enough for use as measurement sources for the larger satellite communications antennas.

Cas A has been extensively mapped, and a contour map, showing its brightness, is shown in Figure 3. Figure 3, Figure 1 and Table 1 are from National Bureau of Standards Report NBSIR 74-382 A Study of the Measurement of G/T Using Cassopeia A by D. F. Wait, et.al.

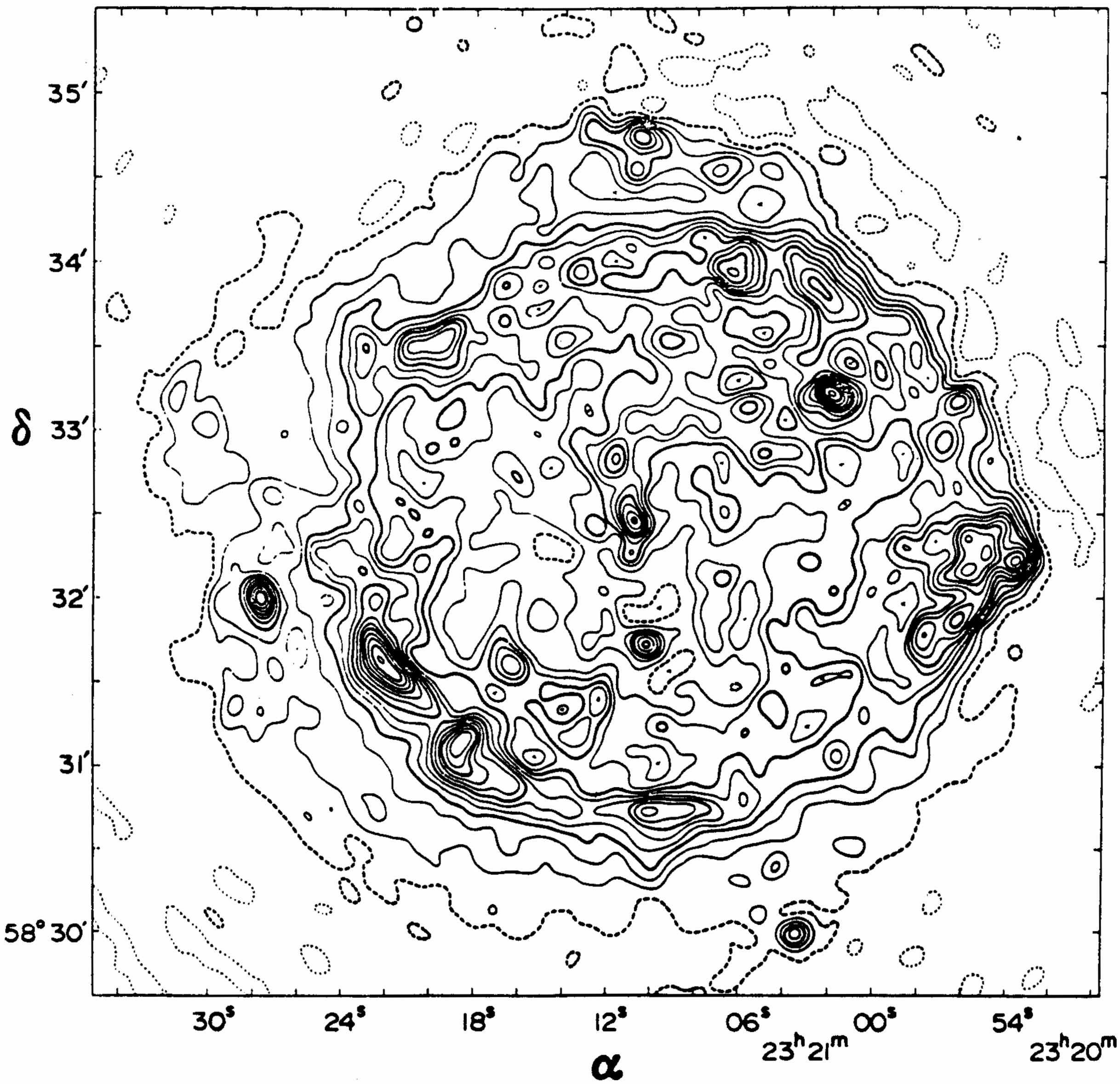


Figure 3. Brightness Temperature Contour Map of Cas A
(Coordinates for Epoch AD 1950.0, 5.0 GHz)
From NBSIR 74-382
(Courtesy of Monthly Notices of the Royal Astronomical Society, Vol. 151, No. 1, p. 112, 1970)

Noise Fundamentals

Noise Power and Noise Power Spectral Density. Boltzmann's constant k relates temperature to the thermal energy of motion of matter. Its value is approximately 1.3805×10^{-23} Joules/K. The random acceleration of electrons in any type of matter produces electrical noise power, which is proportional to temperature in Kelvins.*

This power is uniformly distributed in frequency from zero through the microwave region of the spectrum. It has a noise power spectral density** defined by

$$N = dP_N/df \quad \cdot \qquad \text{(Watts/Hertz)} \qquad (1)$$

The Fourier components which describe the motion of electrons in matter begin to roll off in amplitude toward the millimeter-wave region of the spectrum. Quantum noise, caused by random changes of energy states of electrons, increases with frequency and begins to dominate above this frequency region.

The available noise power density N_O is that noise-power density delivered from a noise source to a conjugate load. It is given by

$$N_O = kT \quad , \qquad \text{(Watts/Hertz)} \qquad (2)$$

We will tend to omit the modifier "available" in this paper and define noise power density to be synonymous with available noise power density on the assumption that we are considering only conjugately-matched devices.

Noise Temperature. The noise temperature T_X is a measure of the noise power produced by a communications system, subsystem, component, or noise source which is designated by the subscript x; it is always an "effective" temperature rather than a physical temperature, since it is a measure of all of the noise, both thermal and non-thermal.

T_X is the ratio of the available noise-power density to Boltzmann's constant:

$$T_X = \frac{N_O}{k} \qquad \text{(K)} \qquad (3)$$

It is a convenient measure because it falls in the range of a few degrees to a few thousand degrees in satellite communications circuits.

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* See Glossary. "Kelvins" is abbreviated to "K."

**The designation "spectral density" is in contrast to "flux density." The latter refers to noise power per square meter, which may be incident on an antenna, for example. Noise flux density has the units of watts/meter^2 or watts/Hz/meter^2. In the discussions that follow, we will omit the modifiers "flux" and/or "spectral" except where the omission leads to ambiguity.

Noise Sources. A noise source is a device which generates a continuous spectrum of electromagnetic energy. A standard noise source is a noise source whose output noise-power density or noise temperature has been calibrated to a specified accuracy.

Most standard noise sources used in noise measurements of satellite earth station receiving systems fall into one of the following categories:

a. Passive loads in waveguide or coaxial cable cooled to the boiling point of a specific liquified gas, such as Helium, Nitrogen or Freon.

b. Passive loads heated to an accurately controlled temperature above the ambient.

c. Noise sources obtained by electrical discharge in rarefied gases, such as argon, neon or xenon.

d. Solid-state noise sources.

e. Receiving antennas. An antenna is a noise source which derives its noise power largely from the noise flux which is incident upon it.

Antenna Noise Temperature. An antenna was defined in (e) above as a noise source. A typical antenna used in satellite communications can be considered a noise source combined with a nearly lossless, linear two-port* as shown in Figure 4, where the two-port represents the ohmic losses of the antenna. In the following development, we will consider these losses to be zero. The effects of losses will be taken into account by the techniques described under the section entitled System Noise Temperature.

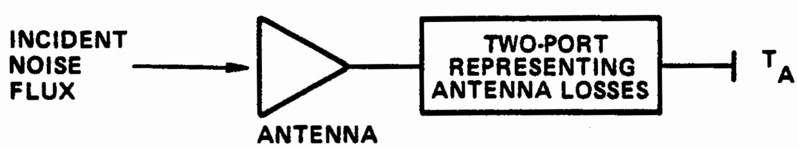

Figure 4. Antenna, Showing Incident Noise Flux

*See Glossary (Section 5C-1) and paragraph on power gain in this paper.

It can be shown[1] that if a lossless antenna is enclosed in a perfectly absorbing box whose temperature is T, the antenna appears at its terminals to be a noise source of temperature $T_A = T$. It "sees" a temperature of T from all directions.

If the box is removed, the antenna will see different values of incident noise flux, represented by different noise temperatures, as a function of direction about the antenna. The value of T_A is the antenna noise temperature of the antenna in a particular environment.

The contribution of the incident noise flux to T_A is weighted by the radiation pattern of the antenna. The total antenna noise temperature can be approximated by the equation:

$$T_a = \int_{\Omega_1} G_1[T_C G_\tau + (1-G_\tau)T_\tau] \frac{d\Omega_1}{4\pi} + \int_{\Omega_2} G_2[\rho T_C + (1-\rho)T_G] \frac{d\Omega_2}{4\pi} + T_s \quad (4)$$

where

T_a = antenna temperature,

G_1 = the gain of the antenna as a function of the directions from which the antenna is receiving direct radiation from the sky.

G_2 = the gain of the antenna as a function of the directions from which the antenna is receiving direct radiation from the earth,

Ω_1 = the region of solid angle of the antenna pattern that is above ground level (steradians),

Ω_2 = the region of solid angle of the antenna pattern that is toward the earth (steradians),

T_C = the background noise temperature of the sky,

T_τ = the equivalent atmospheric noise temperature,

T_G = noise temperature of the ground (approximately 290K),

T_s = noise temperature of the sun*

G_τ = the atmospheric transmission factor (<1) through the troposphere and lower atmosphere including the effects of moisture, and

ρ = the power reflection coefficient of the earth (<1).

Figure 5 is a simple illustration showing a power pattern superimposed on an antenna, which illustrates the concept of antenna temperature. In Figure 5 the antenna beamwidth is shown much wider than that of a typical satellite antenna. In practice the beamwidth will almost always be less than about two degrees and will often be less than 0.5 degree.

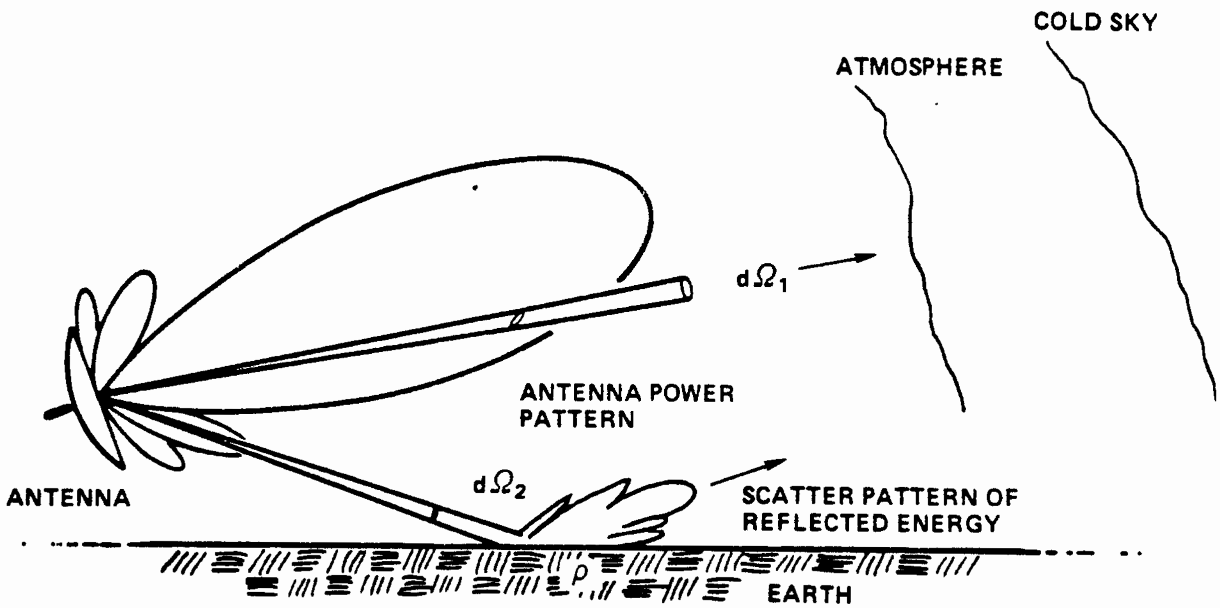

Figure 5. Simple Illustration Showing Basis for Equation Approximating Antenna Noise Temperature

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*T_S is the value of an integral Ω_3 over the sun's disk. It is approximately given by

$$T_s = K\ \bar{G}\ T_{QS}\ G_\tau\ 4.75 \times 10^{-6} \qquad \text{(for a quiet sun)}$$

where

the factor 4.75×10^{-6} is the fraction of the celestial sphere subtended by the sun's disk,

$\bar{G}$ = the average antenna gain over the sun's disk,

T_{QS} = equivalent noise temperature of the quiet sun ($10^5K < T_{QS} > 10^6K$)

C is a correction factor which applies when the beam pattern of the antenna weights the flux from the sun. Its maximum value is unity. When the main beam is directed toward the sun, its value is less than unity if the beamwidth approaches or becomes smaller than the angle subtended by the sun.

The antenna temperature is usually minimum at zenith, typically 15 to 20 degrees for a low-loss antenna with low wide-angle sidelobes. As the elevation angle decreases, the antenna temperature increases because more of the high-level sidelobes look at the earth, which has a temperature of about 290K. For smooth earth or water ρ tends to be small, especially at small grazing angles, decreasing the contribution of the second term of the integral over Ω_2. A typical curve of the variation of noise temperature with elevation angle is illustrated in Figure 6.

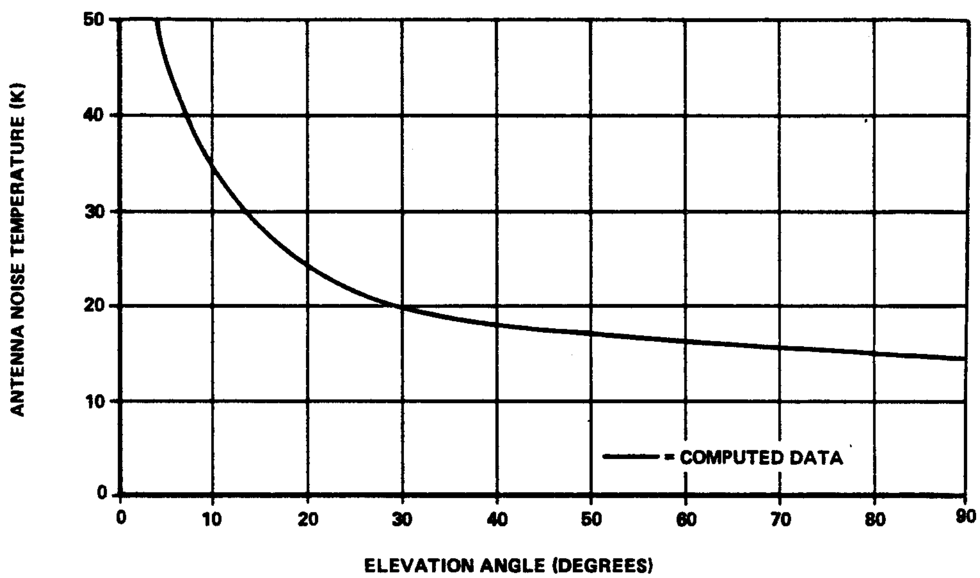

Figure 6. Typical Variation of the Noise Temperature of a Satellite Antenna as a Function of Elevation Angle

Available Gain. The available (power) gain G (available transmission factor) between an input port and output port of a linear transducer (two-port) is the ratio of the available signal power at the output port to the available signal power at the input port, as shown in Figure 7. It implies conjugate matches at the input and output of the two-port. For signal-to-noise ratios greater than unity, the gain for the noise will be equal to that for the signal.

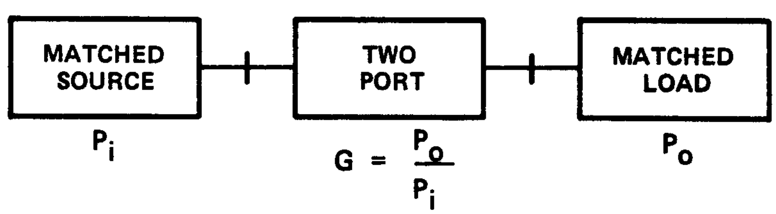

Figure 7. Illustration of Available Gain, G

The two-port can be an active device, such as an amplifier, or a passive device, such as an attenuator or waveguide. The gain of a passive device is always less than unity. Often gains less than unity are inverted and designated "losses." In the following sections we will not invert gains which are less than unity. This simplifies equations involving gains of cascades of transducers.

Gain can be expressed in dB or as a gain factor. For example, a matched amplifier with a gain of 20 dB has a gain G = 100; a section of cable with a 3-dB insertion loss has a gain G = 0.5. It is customary to omit the word "factor" when it is understood whether the gain is a factor or a number in decibels.

Gain is a point function of frequency. It involves power ratios in the case of a single-frequency signal; it involves power density ratios in the case of a continuous spectrum, such as noise.

Effective Input Noise Temperature of a Two-Port. The effective input noise temperature of a two-port is a fictitious temperature which, when added to the noise temperature of a matched source connected to a matched noise-free two-port with the same gain as the actual two-port, will produce the same output noise-power density as the actual two-port, as illustrated in Figure 8.

The effective output temperature of a two-port is the effective input temperature multiplied by its gain G.

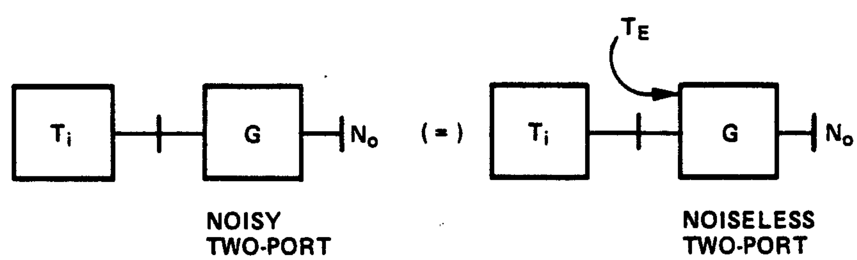

Figure 8. Illustration of Effective Input Noise Temperature

The Effective Input Noise Temperature of a Passive Two-Port. It can be shown that the effective input noise temperature of a passive two-port at a uniform physical temperature T_p is given by

$$T_E = \left(\frac{1 - G}{G}\right)T_p \quad \cdot \qquad (K) \qquad (5)$$

where G is less than unity for practical devices. For the theoretical case of a passive, lossless two-port, G = 1 and $T_E = 0$.

When G is zero, (5) becomes indeterminate. This is logical because the effective input noise temperature of a device with zero gain (infinite loss) is meaningless. In this case (5) is multiplied by G, and the effective output noise temperature is seen to the physical temperature T. The two-port is then simply a noise source of temperature T.

Noise Figure. The noise figure of a two-port can be defined as the ratio of the output noise power of the two-port with a 290K noise source connected to its input to the output noise power of a noiseless two-port with the same input noise source. From Figure 8 the noise figure is seen to be given by

$$NF = \frac{G(290 + T_E)}{G(290)} = 1 + \frac{T_E}{290} \qquad (6)$$

The two-port can be active or passive, with a gain greater than or less than unity. Note that if the two-port adds no noise, the noise figure is unity. A two-port with a T_E of 290K has a noise figure of 2 (3 dB).

Noise figure is sometimes called noise factor. Some people prefer to use noise factor for the numerical ratio of (6) and noise figure for the then defined noise factor converted to decibels. The IEEE[2] has decided that the two terms are synonymous and can imply either the numerical ratio or the ratio in decibels.

Solution of (6) for T_E gives

$$T_E = (NF-1)\ 290 \qquad (K) \qquad (7)$$

The numerical noise figure ranges from unity to infinity as the effective temperature ranges from zero to infinity. The noise figure in decibels ranges from zero to infinity.

The effective temperature expressed in decibels is not a true ratio; it has the dimensions of dB/K. See Appendix A of paper 1-2. It ranges from $-\infty$ to ∞, although an effective temperature of less than 1K (negative dB/K) is rare.

Noise Bandwidth. The noise bandwidth B of a receiving system is defined to be the bandwidth B of a rectangularly-shaped, noise-free filter with an available power gain G_O whose noise-power output is the same as that of the actual filter (see Figure 9). This is on the assumption that the noise-power density is constant with frequency. Thus

$$B = \frac{1}{G_O} \int_0^\infty G(f)df \quad . \qquad (\text{Hertz}) \qquad (8)$$

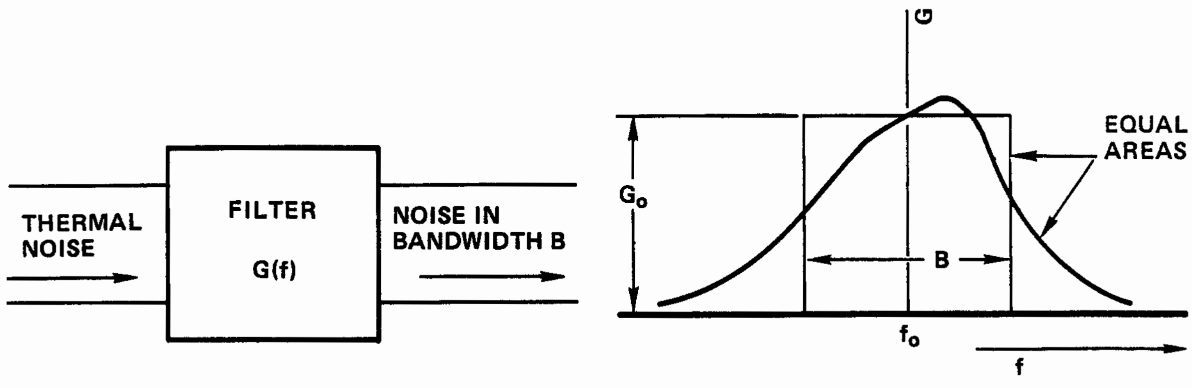

Figure 9. Illustration of Noise Bandwidth for Flat Noise

The noise power output of the filter is given by

$$P_{NO} = kT_iBG_o = kT_i \int_0^\infty G(f)df \qquad \text{(watts)} \qquad (9)$$

where T_i is the available noise temperature at the input of the filter.

The noise bandwidth B is not a constant parameter of a receiving system. Its value depends on the choice of f_o, which in turn defines G_o in Figure 9. The definition of noise bandwidth assumes that the filter contributes no noise of its own. In practice every circuit contributes some noise. The definition of noise bandwidth assumes that sufficient preamplification exists to cause the filter noise contribution to be negligible. (See the next section, which discusses the noise of amplifier cascades.)

System Noise Temperature

A satellite-communications receiving system consists of active and passive devices which, for the purpose of system noise temperature calculations, can be considered a cascade of linear two-ports. All these devices generate noise which combine to form the effective noise temperature of the cascade.

Consider Figure 10. It shows a cascade of two-ports fed by a source, T_{io}, which is the antenna in operational situations. Every two-port in the cascade is characterized by a gain and an effective input noise temperature.

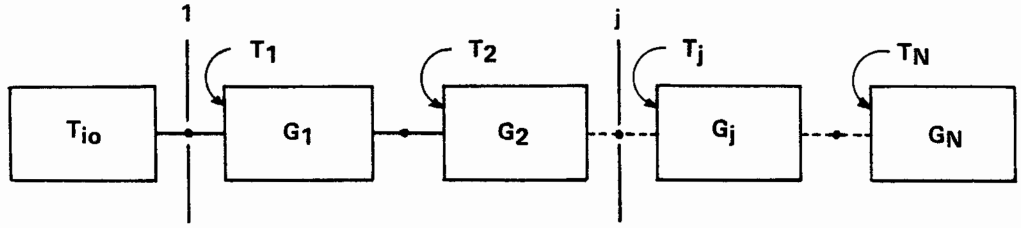

Figure 10. System Block Diagram Showing Cascade of Two-Ports fed by a Noise Source of Temperature T_{io}

For the cascade of two ports of Figure 10, the total noise power at the output is given by:

$$P_0 = kB[(T_{i0} + T_1)\ G_1G_2 \cdots G_N + T_2G_2 \cdots G_N + \cdots T_NG_N] \qquad (10)$$

where

k and B have been defined and

T_{i0} is the source temperature.

The total noise power referred to the input of the cascade (reference plane 1) can be obtained by dividing equation (10) by the total system gain, i.e.:

$$P_I = kB[T_{i0} + T_1 + T_2/G_1 + T_3/G_1G_2 + \cdots + T_N/G_1G_2 \cdots G_{N-1}] \quad (11)$$

The factor within the brackets of (11) is the operating noise temperature or system noise temperature T_{S_1} of the cascade* referred to plane 1:

$$T_{S1} = T_{i0} + T_1 + T_2/G_1 + T_3/G_1G_2 + \cdots + T_N/G_1G_2 \cdots G_{N-1} \qquad (12)$$

The effective input noise temperature T_{E_1} of the cascade of Figure 10 is defined by the system temperature at reference plane 1 with the source temperature equal to zero:

$$T_{E1} = T_1 + T_2/G_1 + T_3/G_1G_2 + \cdots + T_N/G_1G_2 \cdots G_{N-1} \qquad (13)$$

It is often convenient to define a source temperature, an effective input noise temperature and a system temperature with reference to the k^{th} reference plane, as shown in Figure 11, instead of the first. For example, the k^{th} device may be the low-noise amplifier (LNA). With respect to this port, the cascade of devices to the left is the source defined at the k^{th} plane. The cascade to the right can be thought of as a sink.

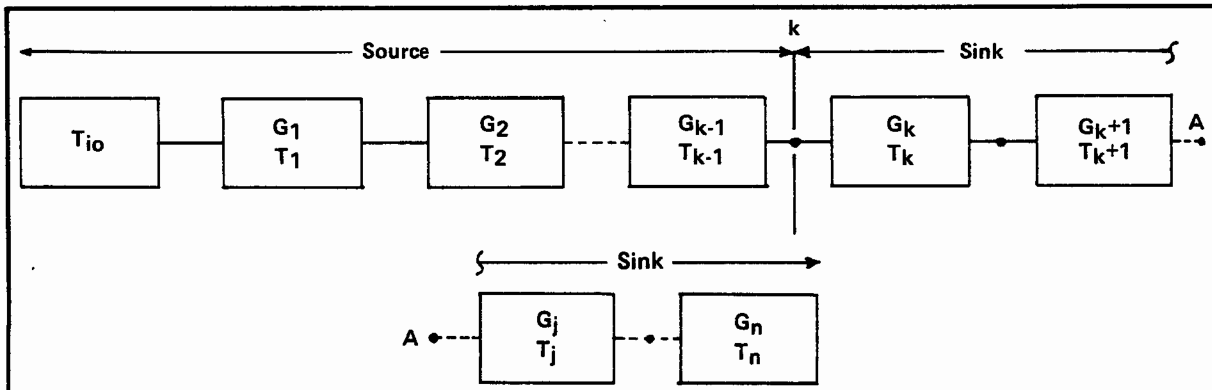

Figure 11. System Block Diagram Showing Definition of Noise Temperature at Reference Plane k

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*We prefer writing T_{op} instead of T_S, but will stick to convention and use T_S. The subscript "op" prevents confusion with "source", which can occur when T_S is used. The subscript "i" is used in this paper for "source" for the same reason.

The source reference plane is transferred from a given plane toward the output of the cascade by (1) multiplying the source temperature defined at the first plane by the gains of the intervening stages and (2) adding the temperatures of each new source multiplied by the gain of the stages intervening between it and the new reference plane.

The source temperature defined at the k^{th} plane is then given by:

$$T_{ik} = (T_{i0} + T_1)\, G_1 G_2 \cdots G_{k-1} + T_2 G_2 \cdots G_{k-1} + T_{k-1} G_{k-1} \qquad (13A)$$

The effective input noise temperature T_{Ek} of the cascade representing the sink is given by:

$$T_{Ek} = T_k + T_{k+1}/G_k + \cdots + T_N/G_k G_{k+1} \cdots G_{N-1} \qquad (13B)$$

As shown in Figure 12, the total system noise temperature T_{Sk} with reference to the k^{th} plane is given by:

$$T_{Sk} = T_{ik} + T_{Ek} \quad . \qquad (14)$$

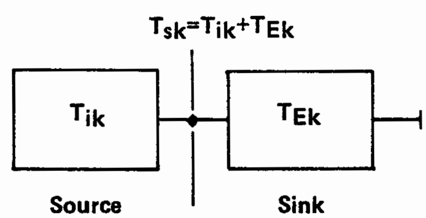


Figure 12. Cascade of Two-Ports Fed by Noise Source

Application To Satellite Receiving Systems

System Temperature. In a satellite receiving system, the antenna can be regarded as a source.

The system temperature is given by:

$$T_S = T_A + T_E \qquad (15)$$

If there are waveguide and/or other devices between the antenna and the LNA, as shown in Figure 13, they constitute part of the source with respect to the reference plane k.

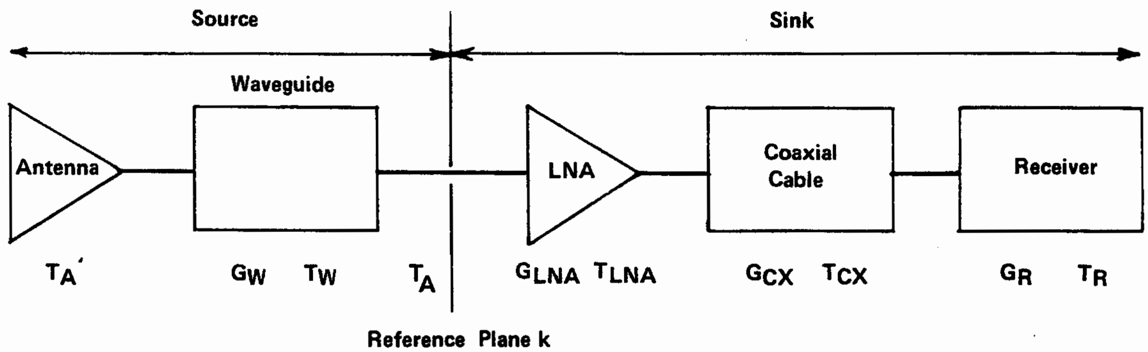


Figure 13. Block Diagram Representing Source and Sink Regions of an Antenna System.
Source and Sink are with Respect to Reference Plane k.

At reference plane k, the sink region consists of the LNA and all the devices which follow the LNA. With the exception of the antenna, all the devices are represented by their effective input noise temperatures. The antenna temperature can be defined at any reference plane. For example, it can be defined as a source temperature T_A' at the antenna side of the waveguide run of Figure 13 or as a source of temperature T_A' at the LNA input.

If it is defined at the antenna terminals by T_A', the antenna temperature T_A defined at the LNA input is given by

$$T_A = T_{ik} = (T_A' + T_W)G_W = T_A'G_W + (1-G_W)T_p \tag{16}$$

where T_W and G_W are T_E and G of (5). The effective noise temperature of the sink referred to reference plane k is given by:

$$T_E = T_{LNA} + T_{CX}/G_{LNA} + T_R/G_{LNA}G_{CX} \tag{17}$$

The system noise temperature at the input to the LNA is given by the sum of (16) and (17). In the above, both G_W and G_{CX} are less than unity and G_{LNA} is greater than unity.

The Figure of Merit, G/T. The symbol G/T (dB/K) is called the <u>figure of merit</u> of a satellite communications system. It is defined by

$$G/T\ (dB/K) = G\ (dBi) - T_S\ (dB\text{-}K) \tag{18}$$

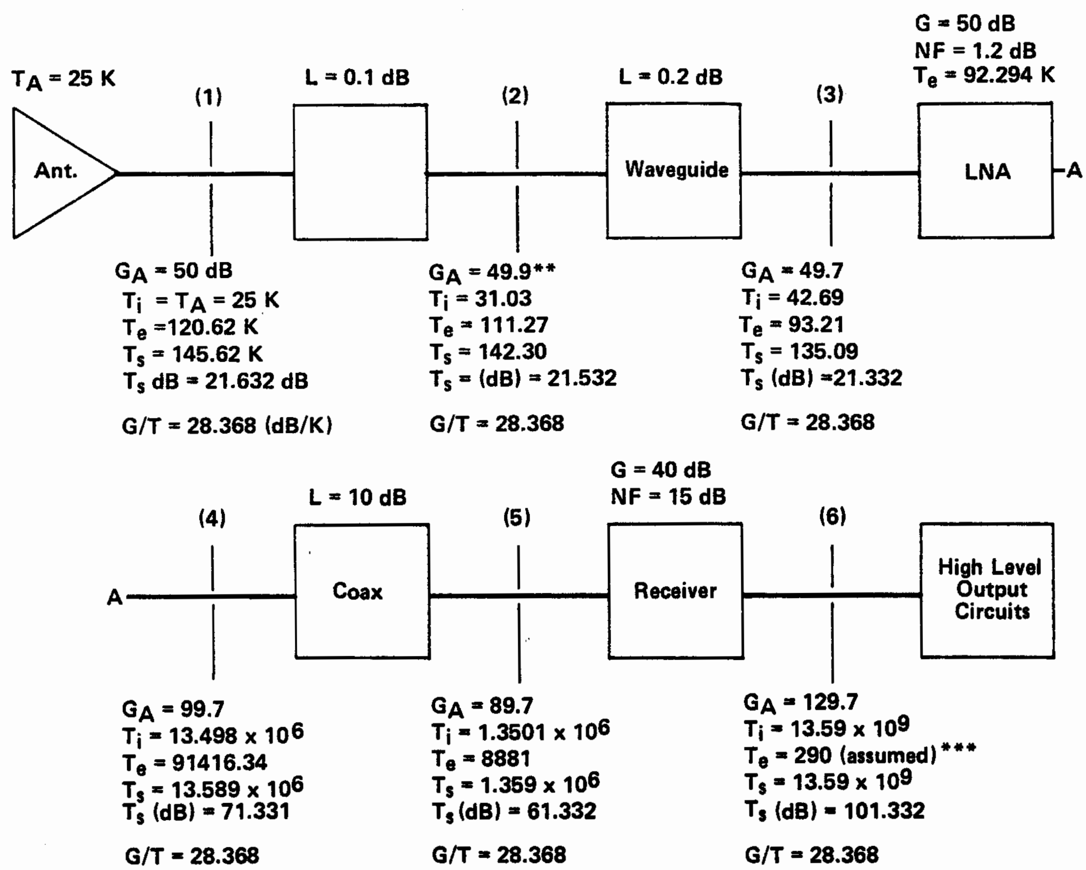
where G (dBi) is the gain of the receiving antenna defined at a given reference plane and T_S (dB-K) is the system temperature at the same reference plane. The reason for its definition as the figure of merit is given in the discussion which accompanies equation (15) in the paper "Principles of Satellite Communication."

An example of a complete receiving system is shown in Figure 14 in which both noise temperature and G/T calculations are illustrated. The following observations can be outlined:

a. The value of T_S is different at every junction.

b. The value of G/T (where $T \equiv T_S$) is the same at every junction.

c. The system noise temperature at the input to the LNA is influenced largely by the noise temperature of the components which precede the LNA and the LNA itself. The components which follow the LNA have a negligible contribution on the system noise temperature at the LNA input junction if the LNA gain is high.



*G/T = G_A/T_s

**Dimensions are same as at position (1)

***T_e for the output circuits is likely to be inconsequential in comparison with T_i at (6) and can usually be neglected.

Figure 14. Example of Noise-Temperature and G/T Calculations for Cascade of Two-Ports.

Note that G/T is independent of position in cascade.

The parameters which have a significant influence on G/T are the following:

a. Antenna Gain and Antenna Noise Temperature--See the discussion in papers 1-2 and 2A-1.

b. Antenna Elevation Angle--The lower the elevation angle, the higher the antenna noise temperature, and hence the lower the G/T for a given antenna gain.

c. Feed and Waveguide Insertion Loss--The lower the loss of these devices, the higher the G/T.

d. LNA--The lower the noise temperature of the LNA, the higher the G/T. The LNA must have a sufficiently high gain to suppress the noise contribution due to the components which follow the LNA. For example, in Figure 14, if the LNA has a gain of only 40 dB, then, at the input to the LNA, the value of T_S will increase to 144.1K. This means that system G/T will be reduced by about 0.26 dB. For a 30 dB LNA gain, the system G/T will drop an additional 1.96 dB.

Measurement of Noise Temperature

Measurement of T_E. The effective input noise temperature of any sink can be made by use of calibrated hot and cold noise sources as in Figure 15. In practice the sink is likely to consist of an LNA followed by a receiving system. The measurements are usually made with a radiometer or a commercial noise test set. If the device under test operates at microwave frequencies (such as a 4-GHz antenna or LNA), the noise is converted to 70 MHz and is bandlimited by a bandpass filter with a width of about 5 MHz.

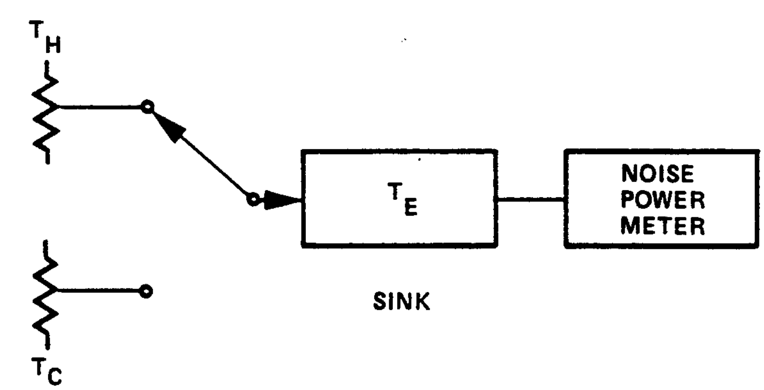


Figure 15. Setup for Measurement of the Effective Noise Temperature of a Sink

When a sink is connected to two different noise sources, the ratio of the noise power outputs is called the Y factor.

If the effective source temperatures of two calibrated noise sources are T_H and T_C, and T_E is the effective input noise temperature of the sink, the resulting Y factor can be written

$$Y_1 = \frac{T_H + T_E}{T_C + T_E} \qquad (21)$$

From (19) T_E is seen to be given by

$$T_E = \frac{T_H - Y_1 T_C}{Y_1 - 1} \quad . \qquad (K) \qquad (20)$$

Measurement of T_A. Measurement of the antenna temperature T_A can be made by the Y factor method, where the antenna replaces the cold load (see Figure 16).

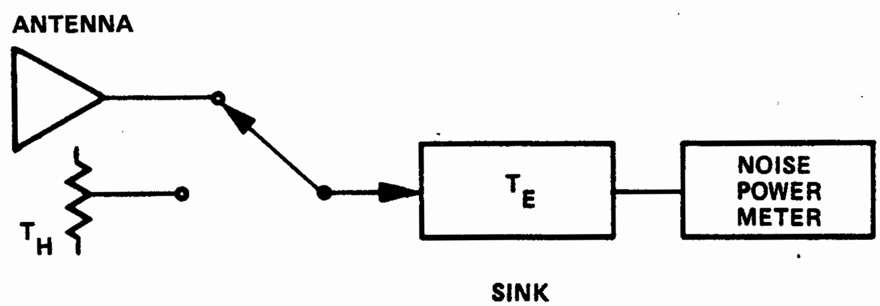


Figure 16. Setup for Measurement of Antenna Temperature

The Y factor, which we will call Y_2, is given by

$$Y_2 = \frac{T_H + T_E}{T_A + T_E} \quad . \qquad (21)$$

T_A is then given by

$$T_A = \frac{T_H - T_E(Y_2 - 1)}{Y_2} \quad , \qquad (22)$$

where T_E is known from (20).

Measurement of System Noise Temperature. The system noise temperature, also called the operating noise temperature, is given by (15); therefore, it is the sum of T_E from (20) and T_A from (22):

$$T_S = T_A + T_E \tag{23}$$

If a hot and cold load are both switched with the antenna as in Figure 17,

$$Y_1 = \frac{T_C + T_E}{T_A + T_E} \tag{24}$$

and

$$Y_2 = \frac{T_H + T_E}{T_A + T_E} \quad . \tag{25}$$

Simultaneous solution of (24) and (25) gives

$$T_S = \frac{T_H - T_C}{Y_2 - Y_1} \quad . \tag{26}$$

In this method it is not necessary to determine T_E directly.

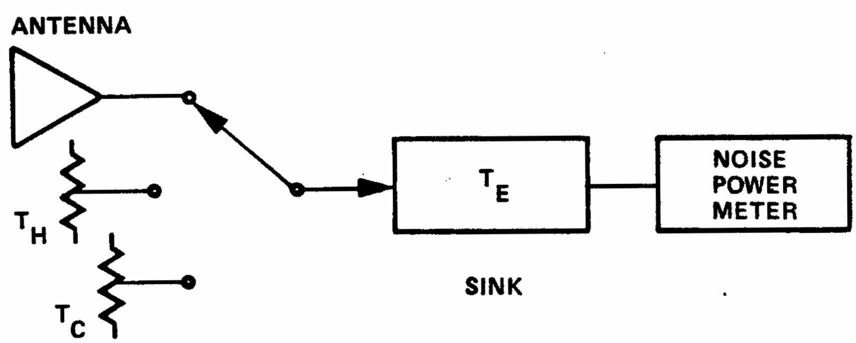


Figure 17. Setup for Measurement of System Temperature

Measurement of G/T

For some applications, the G/T of an antenna system is required to be verified experimentally. There are several methods which are available for experimental verification. Some of the commonly known methods are as follows:

a. Both the antenna gain and the system noise temperature are measured separately and then algebraically combined to determine the system G/T. The gain is determined using a known gain standard on a suitable antenna range, and the system noise temperature is measured using known calibrated noise sources.

b. Using a radio star with a known flux density, the antenna G/T can be measured directly,[3] or the gain and noise temperature can be measured separately.

c. The sun, the moon and some of the planets within our solar system can be used for direct G/T measurement.

d. A geosynchronous satellite whose EIRP is known can be used to measure G/T directly.

The indirect method requires that the gain of the antenna be measured at the same reference plane where T_S is measured. In this measurement, the uncertainty in the value of gain of the standard gain horn can prove to be a major limitation in the accuracy.

For satellite antennas used in the 4/6 GHz band which have diameters no greater than 11 meters, method (a.) is likely to be the most practical of the methods listed. Method (b.) can be used for antennas as small as 11 meters, but it requires extreme care and is expensive to implement. Methods (c.) have been described in the literature, but have a number of problems which tend to make them impracticable. For example, the extended size of the sun and moon and the variability of their radiated flux densities are specific problems. Method (d.) can be used as a rough performance check in comparing one receiving system against another, but it generally suffers in accuracy because of lack of precise calibration of the EIRP of most satellites.

Pertinent details regarding the G/T measurement using radio stars are explained in reference 10. For small antennas the use of a radio star for a direct G/T measurement can lead to a considerable error. Kreutel, et. al.,[8] have indicated that with Cas A the antenna system should have a G/T in excess of 35.4 dB/K in order to have a probable error of ±0.2 dB. Recently, NBS[9] has measured an antenna system (Scientific-Atlanta 10-meter antenna) with a 20.24 dB/K G/T at 2.26 GHz. Using Cas A, it was shown that the estimated error is ±0.3 dB. This type precision is in conflict with Kreutel's[8] predictions. However, NBS personnel are able to achieve a high measurement accuracy by utilizing an automated measurement system developed around a highly accurate power measurement bridge known as the NBS type II self-balancing bridge.

The accuracy of the determination of T_S and G/T depends on several factors. In measurements using noise sources, uncertainties in the values of T_H and T_C, uncertainties in the Y-factor due to source level errors, amplifier nonlinearity, instability and jitter, and cascade mismatch errors are some of the prime factors which affect the overall accuracy. The various error sources are discussed in detail in several of the references, especially in 1, 3, 10, 11 and 12.

References

1. John D. Kraus, Radio Astronomy, McGraw-Hill Company, 1966.

2. ANSI/IEEE STd. 100-1977, IEEE Standard Dictionary of Electrical and Electronics Terms, Institute of Electrical and Electronics Engineers, Inc., 1977.

3. D. F. Wait, et.al., "A Study of the Measurement of G/T Using Cassiopeia A," NBS Report NBSIR 74-382, June 1974.

4. L. V. Blake, "Antenna and Receiving-System Noise-Temperature Calculation," Research Report AD-265414, September 1961.

5. R. Brown and A. C. Lovell, The Exploration of Space by Radio, John Wiley and Sons, Inc., 1958.

6. A. D. Kuz'min and E. A. Salomonovich, Radioastronomical Methods of Antenna Measurements, Academic Press, 1966.

7. "Influence of the Non-Ionized Regions of the Atmosphere on the Propagation of Waves," CCIR Plenary Assembly Report No. 234-1, 1966.

8. R. W. Kreutel and A. O. Pacholder, "The Measurement of Gain and Noise Temperature of a Satellite Communications Earth Station," Microwave Journal, October 1969.

9. D. F. Wait, "Satellite Earth Terminal G/T Measurements," Microwave Journal, April 1977.

10. "Measurement of the Radio G/T with the Aid of Radio Stars," Fixed Service Using Communication Satellites, CCIR Plenary Assembly, Geneva, Vol. 4, 1974.

11. International Electrotechnical Commission (IEC) Standard S10-2-1, 1978.

12. C. T. Stelzried, "The Deep Space Network - Noise Temperature Concepts, Measurements and Performance," Publication 82-33, Jet Propulsion Laboratory, Pasadena, CA, 21982.

13. C. T. Stelzried, "Noise Adding Radiometer Performance Analysis," TDA Progress Report 42-59, Jet Propulsion Laboratory, Pasadena, CA, Oct. 15, 1980, pp. 98.

Satellite Link Analysis

Introduction

This paper is intended as a general review of the principles which control radio communications links, with specific applications to geostationary satellite links as currently achieved in domestic U.S. systems. The interpretation of the results of the link analysis in terms of end-to-end performance is discussed for television, message circuits, and digital applications. With the concepts presented herein, the effects of various earth station parameters can be examined for a variety of fixed satellite applications spanning the requirements of the communications industry.

Because of the extended scope of this presentation, some portions of the discussion are necessarily abbreviated. In these cases, references are provided to more extensive treatments of the individual topics. An attempt has been made, as much as possible, to present a discussion which may be of interest to an audience of diverse backgrounds and varied technical expertise.

The discussion is naturally divided into two parts:

1. The analysis of the radio link, resulting in a determination of the net carrier-to-noise ratio (C/N), and,

2. The interpretation of the link C/N for each of several services (video, message, digital, etc.) in terms of end-to-end performance characteristics.

This separation of the problem into two parts emphasizes the fact that, regardless of the details of the application, the C/N analysis follows the same basic pattern. Conversely, the interpretation of the impact of a particular C/N obtained for the link is highly dependent on the application and must therefore be discussed separately for each case.

The fundamental relationships used for a radio link calculation will be reviewed to establish a starting point fo the discussion of satellite system links. The parameters for a typical domestic satellite link will then be introduced, and the effects of earth-station parameters on the radio link will be discussed. These results will be extended to video, video subcarrier, message, and digital applications to provide examples of the analysis techniques required for these services.

Radio Link Analysis Fundamentals

The purpose of a radio link analysis is to determine the transmission quality that can be expected for signal carriers relayed from one point to another. This is expressed as the ratio of the received radio carrier power to the noise power density (C/N_0).

Consider a transmitting station illustrated in Figure 1 with transmit power P_T and transmit gain g_T. The effective isotropic radiated power (EIRP) for the station (along the main beam of the antenna) is the product $g_T P_T$. At a distance r from the transmitter, the radiated flux density, S, along the beam axis is:

$$S = \frac{g_T P_T}{4\pi r^2} K_A P = \frac{EIRP}{4\pi r^2} K_A P \text{ in power/m}^2 \qquad (1)$$

where:

K_A = the atmospheric attenuation factor

P = the polarization efficiency

If an antenna with an effective area in square meters, A_e, is receiving this flux density, the received carrier level at the antenna output is:

$$C = SA_e = \frac{P_T g_T A_e}{4\pi r^2} K_A P \qquad (2)$$

For the remainder of this analysis, we will assume K_A and P are unity.

The effective noise power density at this point is given by:

$$N = KT_s \qquad (3)$$

where:

K = Boltzmann's constant = 1.38×10^{-23} Joules/K or 228.6 dB in logarithmic form

T_s = System noise temperature (a measure of the noise contributed by the receiving system per unit bandwidth)

The details of the determination of the system noise temperature from the equipment characteristics are explained in a separate paper. At this point, we will assume this parameter is a basic characteristic of the receiving system.

Using these results, the carrier-to-noise density ratio C/N_o is:

$$C/N_o = \frac{P_T g_T A_e}{4\pi r^2 KT_s} = \frac{(EIRP)\ (A)}{(4\pi r^2)(KT_s)} \qquad (4)$$

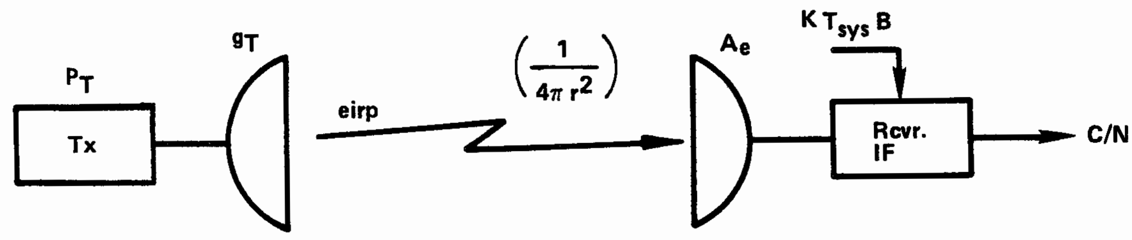


Figure 1. RF Link Diagram

The effective aperture of the receiving system A_e can be calculated from antenna fundamentals and the gain at their frequency:

$$A_e = \frac{g_R \lambda^2}{4\pi} \tag{5}$$

where:

g_R = power gain of the receiving antenna on its boresight axis.

Substituting this relation into the expression for C/N_o, we get:

$$C/N_o = \frac{P_T g_T g_R \lambda^2}{(4\pi r)^2 KT_s} \tag{6}$$

If the various factors in the C/N expression are grouped as follows, then we get the following for C/N_o:

$$C/N_o = \left(\frac{\lambda}{4\pi r}\right)^2 (EIRP) \left(\frac{g_R}{T_s}\right) \tag{7}$$

The factor $\left(\frac{\lambda}{4\pi r}\right)^2$ is often inverted and defined as the spreading loss or space loss factor.

The spreading loss can also be expressed as:

$$L_s = \left(\frac{4\pi r f}{c_o}\right)^2 \tag{8}$$

where:

c_o = the speed of light = 3 X 10^8 meter/second

f = the frequency in Hz

Link calculations are usually carried out in dB rather than directly from the above relations because of ease of working in common logarithms. The C/N_0 expression thus becomes:

$$(C/N_0)\ dB = 10 \log (C/N_0)$$

$$(C/N_0)\ dB = EIRP - L_s + (G/T) + 228.6 \qquad (9)$$

where:

$$EIRP = 10 \log (EIRP) \text{ in dBW}$$

$$L_s = 20 \log \left(\frac{4\pi rf}{c_0}\right)$$

$$= 92.45 + 20 \log r\ (Km) + 20 \log f\ (GHz) \qquad (10)$$

$$(G/T) = 10 \log (g_R/T_s) \qquad (11)$$

Alternately, C/N_0 can be expressed in terms of the flux density, S,

then:

$$C/N_0 = S + G/T - 20 \log f_{GHz} - 21.45 + 228.6 \qquad (13)$$

G/T

The importance of the term G/T in equation (9) cannot be overstated. Examination of the C/N expression shows that for a given available transmitting system power and information format (and thus bandwidth), the only available method of controlling the received signal quality that can be used by the downlink operator is through the system G/T. Note that the G/T provides a dB direct relationship with C/N_0. Thus, a receiver system with a 30-dB/K G/T will have a 10-dB better C/N_0 than a receive system with a 20-dB/K G/T. It will be shown later in the analysis of various modes of transmission that this can be an extremely important consideration.

G/T is the figure-of-merit for a receive system. It is a function of the gain of the antenna and first amplifier along with the antenna noise temperature, first amplifier noise temperature, and losses located ahead of the first amplifier. Since the antenna gain is a function of aperture size which is related to total cost of the installed antenna including manufacturing cost of the parts, foundation, transportation, and erection, raising the G/T by increasing the antenna gain will raise the total system cost. Higher G/Ts achieved by lowering the system noise temperature are accomplished by using lower noise LNAs or achieving a low antenna noise temperature. For a detailed discussion of G/T and noise temperature, see the paper "Noise Temperature and G/T of Satellite Receiving Systems" in this publication.

Figure 2 shows a block diagram of a typical receive system. Each device in the RF path has an associated gain (or loss) and a noise temperature. The G/T is the ratio of this system gain to system noise temperature in dB/K. The G/T will be the same regardless of the reference point in the system used for the calculation.

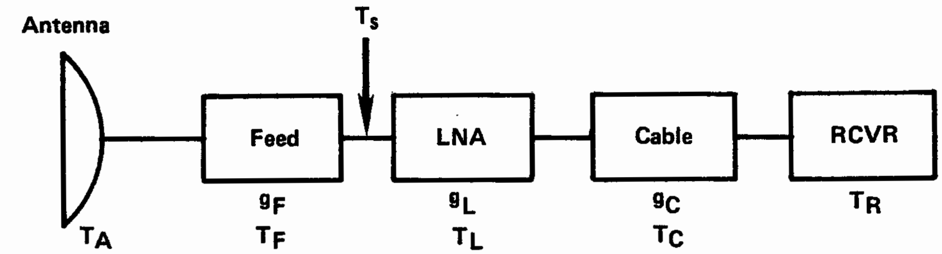


Figure 2. Basic System Diagram for G/T Calculation

These contributions are combined to reflect the noise power weighted by the gain distribution through the chain. The noise temperature of the system is determined primarily by the antenna, the low-noise amplifier, and the coupling waveguide or cable run between the LNA and the antenna. Loss between the antenna and the LNA can result in significant degradation of G/T and, consequently, in signal quality for a very low-noise system.

On the other hand, the noise temperature (or noise figure) of the video receiver is of little consequence, since this factor is masked by the LNA gain, which is usually greater than 45 dB.

The IF noise bandwidth is normally chosen to be the smallest passband compatible with the FM deviation used for the transmission. For domestic full-transponder video, this is about 34 MHz. Half-transponder video normally requires a bandwidth of 17.5 MHz. International full-transponder video through INTELSAT requires 30 MHz.

For an earth station with transmit and receive capability, the receive performance is characterized in the same way as for the receive-only station. The figure-of-merit must be adjusted as required to account for the addition of the transmit/receive diplexing filter and the transmit reject filter which are inserted between the antenna terminals and the LNA input.

Cascaded RF Links

Since a satellite link consists of two cascaded RF links--uplinks and downlinks--the C/N_0 ratios must be combined to determine the net effect of transmission through the total path. If each has a carrier-to-noise ratio $(C/N_0)_1$ and $(C/N_0)_2$, respectively, then the resulting ratio (C/N_0) total is:

$$(C/N_0)_{total} = \frac{1}{\frac{1}{(C/N_0)_1} + \frac{1}{(C/N_0)_2}} \qquad (14)$$

This combining operation cannot be accomplished directly with ratios expressed in dB. The ratios must be expressed in non-logarithmic form and then summed according to the previous format, then re-expressed in dB. This summing rule may be extended to any number of contributions such as multiple satellite hops or the addition of C/I (carrier-to-interference ratios) effects on the path performance.

For example, consider a satellite link with an uplink C/N_0 of 105 dB-Hz and a downlink C/N_0 of 95 dB-Hz. The total in C/N_0 is expressed as:

$$C/N_0 = \frac{1}{\frac{1}{10^{105/10}} + \frac{1}{10^{95/10}}}$$

$$C/N_0 = 94.6 \text{ dB-Hz}$$

The resulting C/N_0 is always lower than the lower of the two ratios. If $(C/N_0)_1 = (C/N_0)_2$, the resulting (C/N_0) total will be 3 dB lower than either C/N_0.

Satellite Transponder

For commercial satellite communications, the orbiting spacecraft provides a one-hop carrier relay over a wide geographic area. For the domestic systems presently in use, the uplink signal is transmitted in a 36-MHz bandwidth near 6 GHz. This signal is received by the satellite, amplified, translated in frequency, filtered, and re-transmitted in a 36-MHz band near 4 GHz. In a typical satellite of this class, there are 12 to 24 transponders assigned within each frequency band.

Since the satellite serves as a transmit/receive station, it must be characterized by a G/T for the uplink side, and by an EIRP for the downlink side.

The G/T is typically +1 dB/K to -6 dB/K for domestic satellites. Figure 3 is a typical G/T curve for an existing satellite. The EIRP is normally specified at the saturation point for the transponder power amplifier. The EIRP for a typical domestic communications satellite is 32 to 36 dBW in the prime-coverage areas. The EIRP varies slightly with geographic location; contour maps (called "footprints") for relative EIRP are usually available for assessing these variations. An example of a footprint is shown in Figure 4.

To couple the uplink and downlink signal strengths, the uplink RF flux density required at the satellite to saturate the transponder is also specified. As shown in Figure 5, typical domestic satellites have a required saturation flux density less than -82 dBW/m^2 over most of the country.

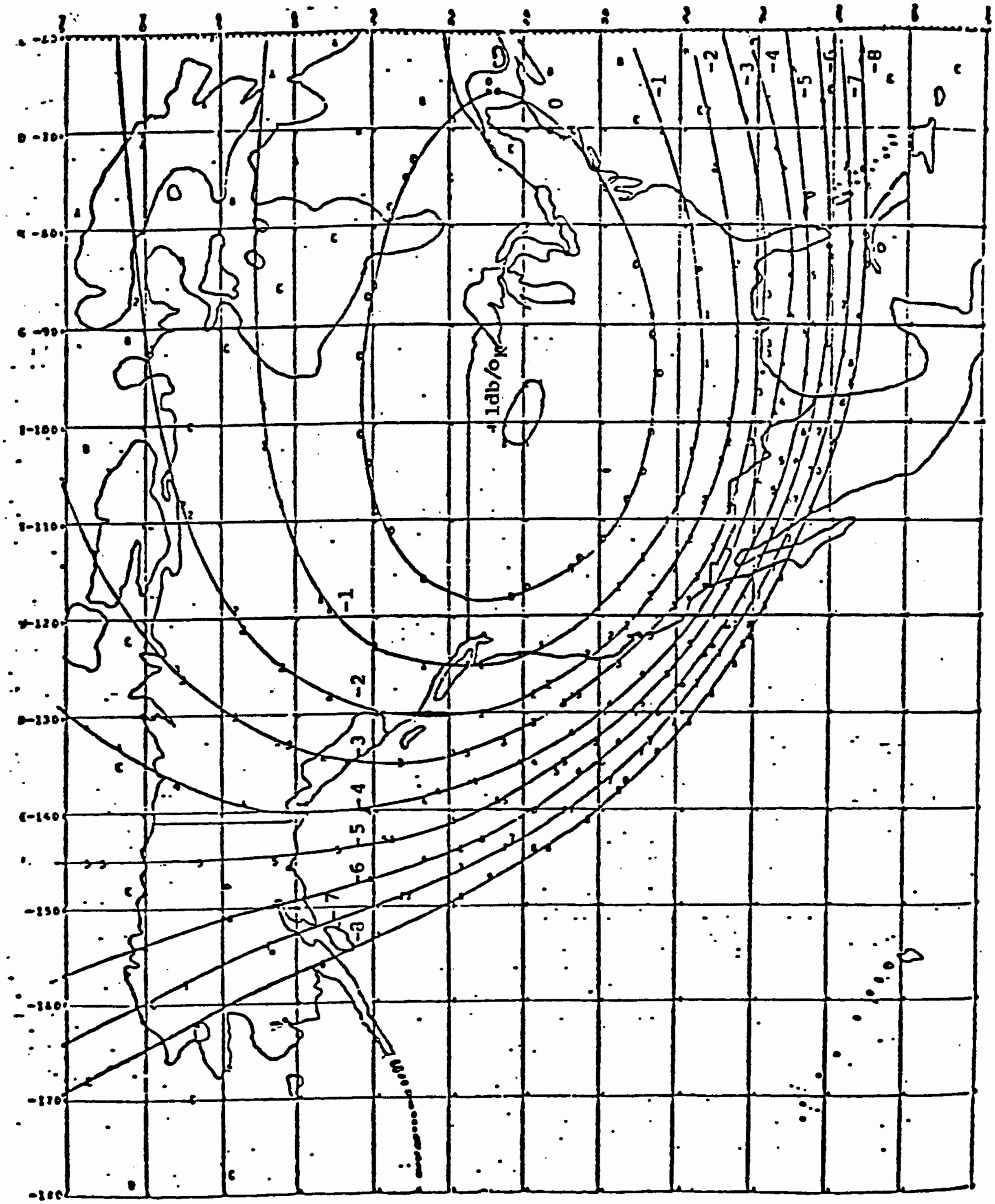


Figure 3. Typical Values of Satellite G/T

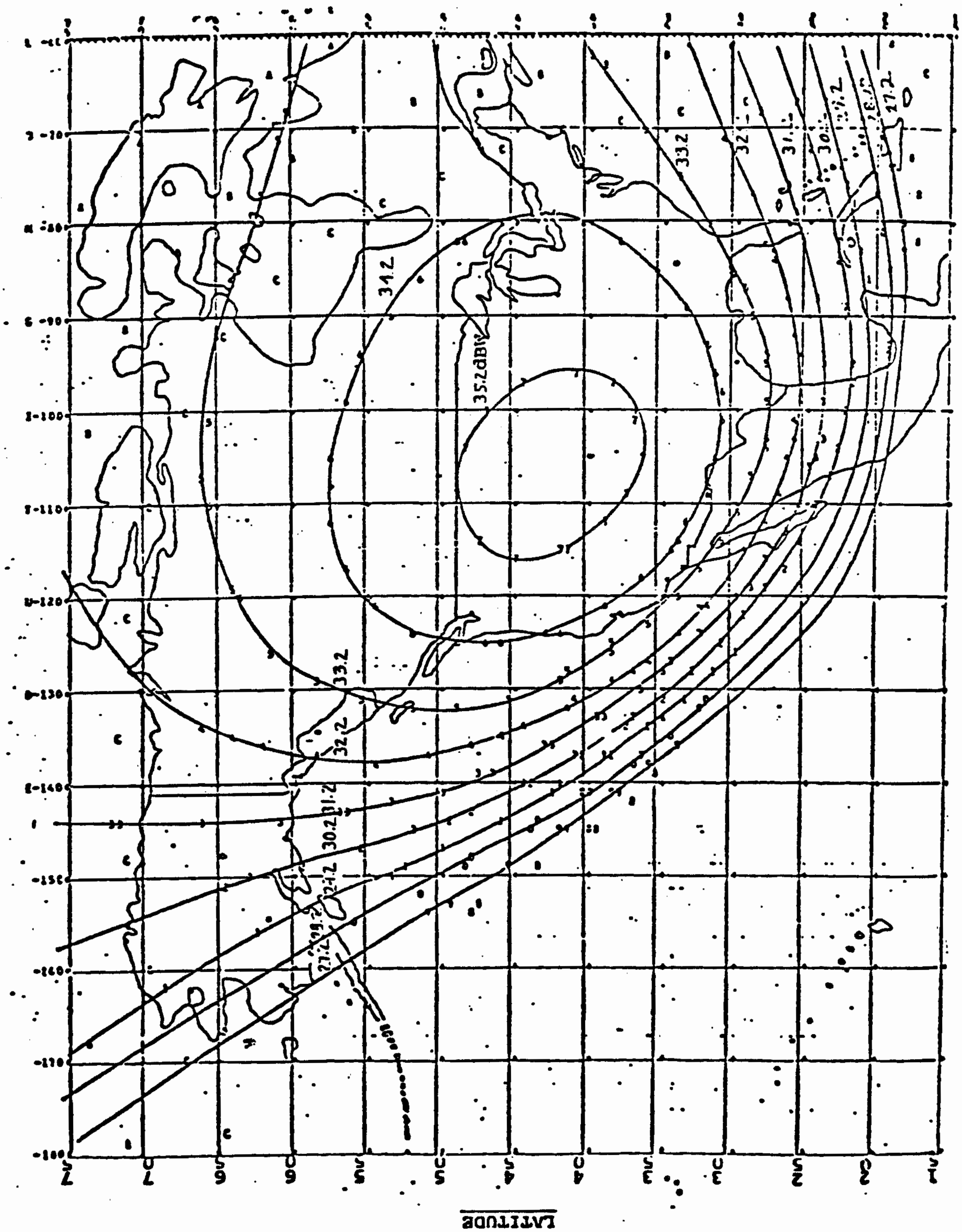


Figure 4. Typical EIRP Values

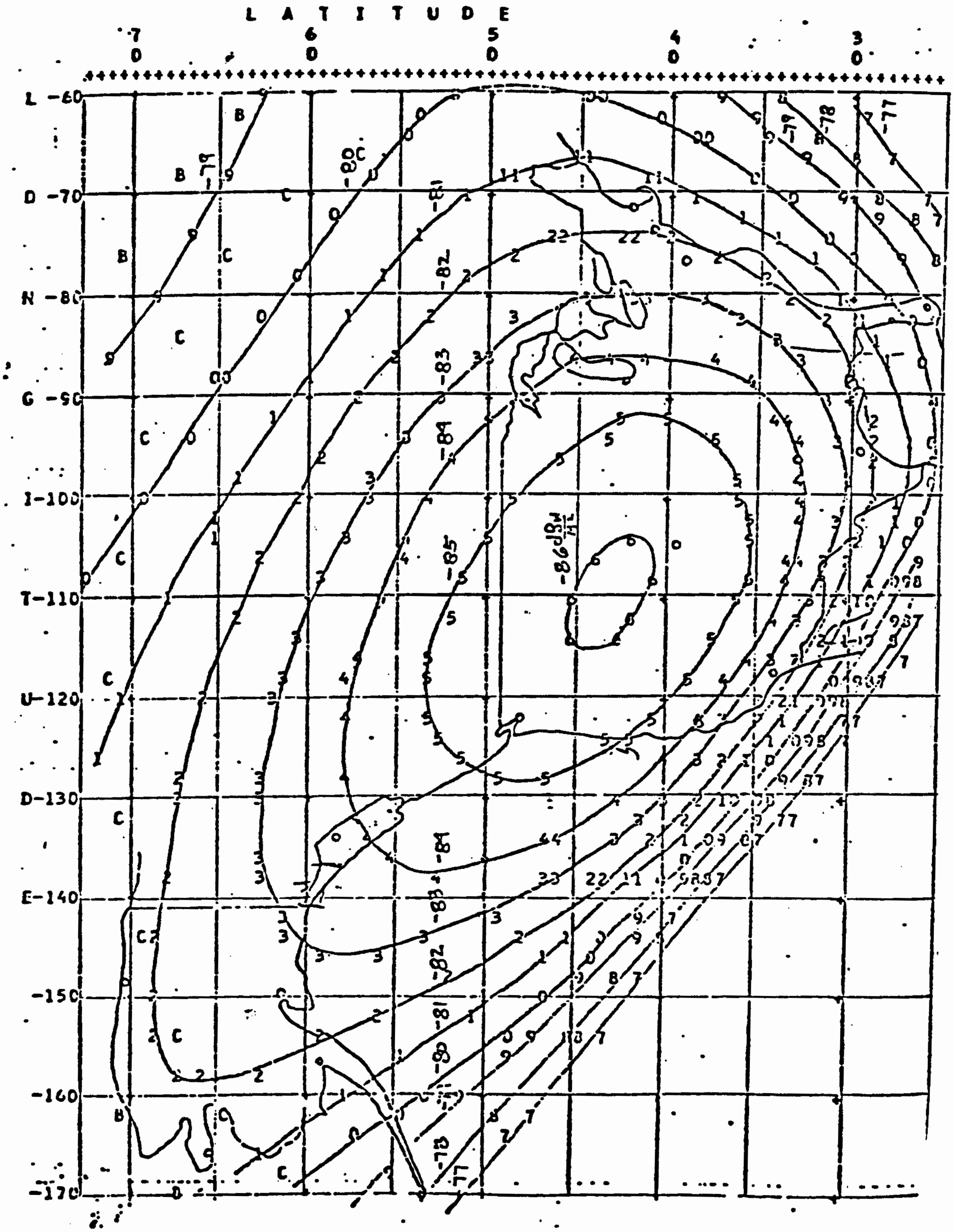


Figure 5. Typical Values of Flux Density Required to Saturate Transponders

The intermodulation performance of the transponder amplifier greatly affects multi-carrier operation of a single satellite transponder. Figure 6 shows typical intermodulation curves for a transponder. Curve (A) is the intermodulation performance for a large number of equally spaced carriers within a transponder. Curve (B) shows the intermodulation characteristics for two carriers.

In order to bring the intermodulation products to an acceptable level and thus not degrade the link C/N_0 with poor C/Im performance, the total power level in the uplink is reduced until the C/Im ratio is 20 dB or better. This means, or course, that the total downlink EIRP from the satellite has been reduced. The transponder output power reduction is shown in Figure 7. To increase the C/Im ratio to 24 dB for the two-carrier case, we see from Figure 6, that the input power must be "backed off" 7 dB. This results in a total downlink power reduction of 3 dB. In addition, the resulting downlink power must be shared between the carriers actually transmitted. For example, if two half-transponder carriers are transmitted, each will have a transmitted power 3 dB lower than the total backed-off output power noted above.

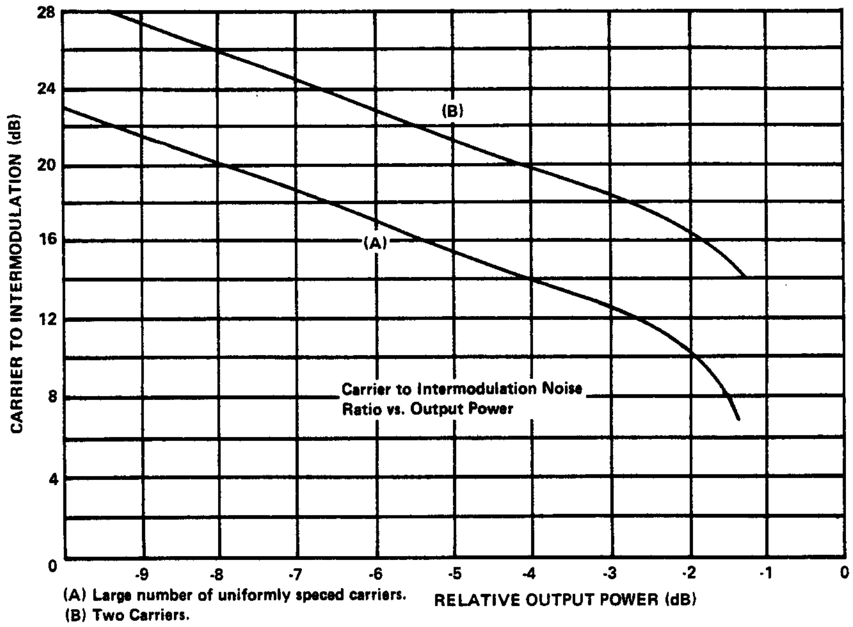


Figure 6. Output Power Normalized to Single Carrier Saturation Point

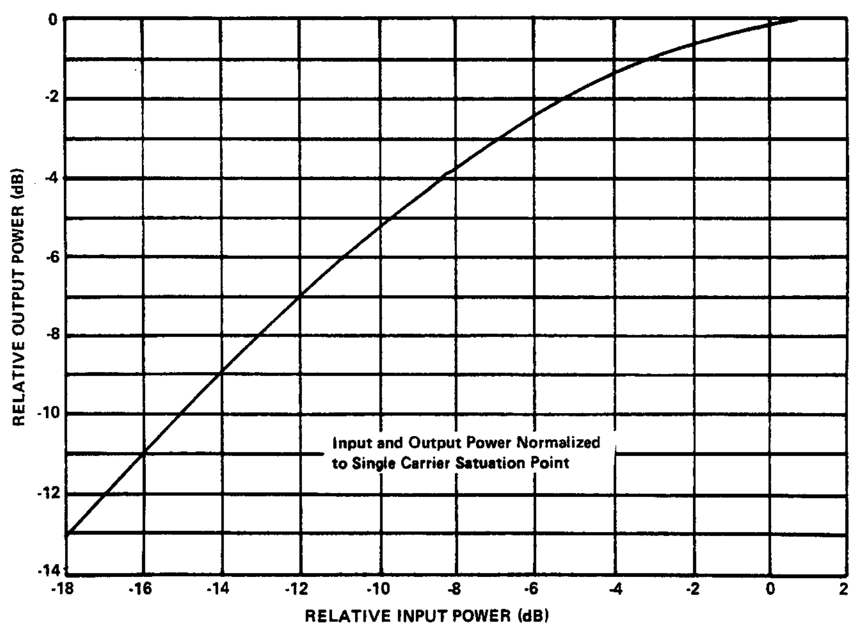


Figure 7. TWT Power Transfer Characteristics

Satellite Link Analysis
General

With the preliminary procedures and specifications described above, we can proceed to the actual link calculations. First we must determine the distance, or slant range, from the satellite to the earth station, to find the corresponding space loss factor. In Appendix A, the orbit geometry is described and the "space loss" is found to be:

$$L_S = 185.05 + 10 \log (1 - 0.295 \cos H \cos \Delta L) + 20 \log f \qquad (15)$$

where:

L_S = Space loss factor in dB

H = Latitude of earth station

ΔL = Difference in longitude for earth station and satellite

f = frequency in GHz

Using this factor, we will now develop the link analysis for the up and downlink, including some refinements from the basic process described earlier.

Uplink

For the uplink, we assume a flux density S at the satellite. In the case of full-transponder video, this is the saturation flux density specified for the satellite. Then, the uplink C/N_0 is given by:

$$C/N_0 = S + \left((G/T)_{sat} - 20 \log f - 21.45\right) + 228.6 \qquad (16)$$

where:

S = Flux density in dBW/m^2

f = Uplink frequency in GHz

$(G/T)_{sat}$ = Satellite receiver G/T in dB/°K

228.6 = $10 \log 1.38 \times 10^{-23}$ (Boltzmann's Constant)

Example:

Using the typical data given for domestic satellites, we have for the uplink:

$$C/N_0 = -82 + [-6 - 20 \log 6 - 21.45] + 228.6 = 103.6 \text{ dB-Hz}$$

The required EIRP for the transmitting station is directly obtained from the flux S and the spatial loss factor $L = 10 \log (4\pi r^2)$:

$$EIRP = S + L + L_1 \qquad (17)$$

The loss factor L_1 (~1 dB) is included to account for aging, pointing error accumulation in the transmitting station, and a small amount of clear weather atmospheric absorption. Then, using the transmit antenna gain G_T, the HPA power required for the station is found from the EIRP.

Example:

For a typical slant range of 39,500 km we have L = 163, so:

EIRP = (-82 + 163 + 1) dBW = +82 dBW

For a 10-meter transmitting antenna, the gain at 6 GHz is about 53.4 dBi, so the HPA required (accounting for approximately 1-dB loss estimated for the waveguide from the HPA to the antenna) is:

P_T = (82 - 53.4 + 1) dBW = 29.6 dBW

or

$P_T \cong 1$ kW

Small Aperture Antenna Transmissions

Transportable earth stations are generally configured with five-meter (or smaller) antennas even when their primary purpose is to uplink full transponder video signals. Since the gain of a five-meter antenna is about 6 dB less than that of a ten-meter antenna, calculations show that we are 6 dB - (34.5 - 29.6) dB = 1.1 dB below saturation for having assumed a 2.8-kW HPA and a required saturation flux density of -82 dBW/m^2 as used in the previous example. If we refer to Figure 7, which shows typical TWT power amplifier transfer characteristics, it can be shown that the actual output power loss of a typical transponder may not be operating at saturation, only a small loss in received signal strength will be observed. This loss of signal will be unimportant at all receive earth stations except those that are extremely marginal (operating at or near FM threshold). Since the general application for transportable earth stations is to transmit programming back to a major city, (thus a large earth station with a high G/T), where it is integrated into programming and retransmitted, the slight loss of signal due to the small-aperture uplink will be negligible at the final receiving earth station.

It should also be noted that the saturation flux density required for most transponders is less than -82 dB/m^2 over much of the country (see Figure 5). For applications in these locations, complete saturation of the transponder will be achieved using the small transportable earth station.

Downlink

The downlink C/N is given by:

$$C/N_0 = (EIRP)_{sat} + L_s + (G/T)_{ES} + 228.6 + L_m \qquad (18)$$

where:

L_m = Link margin

L_s = "Space Loss" factor given in the above section.

$(G/T)_{ES}$ = Earth station, G/T

Example:

For a typical video receive-only station, the G/T is about 26.8 dB/°K. Then, for 34-dBW satellite EIRP, we get, using r = 39,580 mi and a 1 dB link margin:

$$(C/N_0)_{down} = (34 - 196.4 + 26.8 + 228.6 - 1) \text{ dB} = 92 \text{ dB-Hz}$$

The loss factor (L_1)is again used to account for pointing error, aging, and clear air absorption.

Interference

The uplink and downlink contributions are the principal factors of concern in the link calculation. However, there are often sources of interfering signals which must be considered in the analysis. For signals which are small relative to the carrier level (which is the usual case), the interference energy is usually added to the thermal noise in the usual way, as a power ratio C/I. This is the normal technique for handling interfering signals from an adjacent transponder for frequency-reuse satellites. (See detailed discussion of Interference Analysis by Cook in this digest.)

Total Link C/N and C/I

The uplink, downlink, and interference contributions must be combined by power addition to obtain the final C/N_0 ratio.

Example:

Using the results obtained for the uplink and downlink examples above, the receive carrier-to-noise density is:

$$C/N_0 - 103.6 \oplus 92 = 91.7 \text{ dB-Hz}$$

where $\oplus$ refers to power addition.

The carrier-to-thermal noise in a 36-MHz bandwidth is

$$C/N_{th} = C/N_0 - 10 \log B = 16.1 \text{ dB}$$

Assuming a level of 20 dB, the net C/N is

$$\begin{aligned} C/N &= \frac{C}{N_{th}} \oplus \frac{C}{I} \\ &= 16.1 \oplus 20 \\ &= 14.6 \text{ dB} \end{aligned}$$

Since $C/N + 10 \log B_{IF} = C/N_0$, this is equivalent to a carrier-to-noise power density:

$$C/N_0 = 90.2 \text{ dB-Hz} \qquad (19)$$

Video Performance

The video performance must be examined for two district degradations:

a. Thermal noise in the baseband, which is similar to the noise encountered in broadcast TV, and,

b. Impulse noise, which is a phenomenon associated with FM threshold effects.

Above the FM threshold region, only the first factor is of significance. In this case, the video signal-to-noise is given by:

$$(S/N)_t = (C/N_o)\,\frac{12\,(\Delta F_s)^2}{b_n^{\,3}} \tag{20}$$

where:

C = Carrier power (watts)

N_o = Noise power density at that point in the receiver where C is measured = kT_s (watts/MHz)

k = Boltzmann's constant
= 1.3806×10^{-17} W/MHz/°K

T_s = System operating noise temperature referred to that point in the system where C is measured.

F_s = Half of the peak-to-peak deviation produced by that part of the video waveform which is defined to be the "signal" (MHz)

b_n = Noise bandwidth of the baseband filter function (representing the combination of the deemphasis network, measurement bandlimiting filter, and weighting network (when used)) with respect to "triangular" noise (MHz).

$$b_n = \left[\,3\int_0^F f^2\,|H(f)|^2\,df\,\right]^{1/3}$$

$H(f)$ = Product of deemphasis, bandlimiting filter, and weighting (if used) transfer functions.

F = An integration limit frequency high enough so that H(f) may be considered zero above F.

The definition of $(S/N)_t$ and a derivation are given in Reference 1.

In decibels, $(S/N)_t$ is 10 times the logarithm of the numeric ratio expression.

Note that $(S/N)_t$ is not a function of IF bandwidth or modulating frequency. The relation (or any of its variants) is applicable, however, only when the IF bandwidth is at least adequate to support the signal.

Since the video signal is a complex waveform and a variety of emphasis and noise weighting functions can be used in principle, specific standards have been established to provide guidance in evaluating this S/N ratio. The S/N equation presented is sufficiently general to accommodate any standard by specifications of the appropriate ΔF_s and b_n. The standard presently used for most domestic satellite video transmissions is set by CCIR Rec. 421-3, Transmission Advisory Committee, a joint committee of television network broadcasters and the Bell System. For this standard, the "signal" is taken to be the luminance part of the video waveform, such that:

$$F_s = 0.714\ F_v$$

where:

F_v = Half the peak-peak deviation produced by the video waveform including sync tips

For this standard, the unweighted bandwidth is:

$$b_n = 3.357 \text{ MHz}$$

and the bandwidth with noise weighting is:

$$b_n' = 1.574 \text{ MHz}$$

The subjective weighting advantage for FM transmission is thus:

$$30 \log \frac{b_n}{b_n'} = 9.9 \text{ dB}$$

For a more complete discussion of the standards and recommendation, see Reference 1.

Another method of expressing the video S/N ratio in terms of the system:

$$(S/N)_v = C/N + 10 \log 3 \left(\frac{(\Delta F)}{fm}\right)^2 + 10 \log \left(\frac{B_{IF}}{2B_V}\right) + W + CF \qquad (21)$$

where:

$(S/N)_v$ = peak-to-peak luminance signal-to-noise ratio

ΔF = peak composite video deviation

fm = highest baseband frequency

B_v = video noise bandwidth

B_{IF} = IF noise bandwidth

W = emphasis plus weighting improvement factor (12.8 dB)

CF = rms to peak-to-peak luminance signal conversion factor (6.0 dB)

Example:

For full-transponder video with F_V = 11 MHz transmitted through the satellite link analyzed in the previous section, we will compute the S/N projected for the received signal.

$$S/N = 14.6 + 10 \log 3 \left(\frac{11}{4.2}\right)^2 + 10 \log \left(\frac{36}{8.4}\right) + 18.9$$

$$S/N = 52.9 \text{ dB}$$

As the C/N ratio is reduced, the performance of an FM discriminator begins to deviate from the predictions obtained from the video S/N equation. This is shown in the measured performance data shown in Figure 8, the region where the departure from theoretical occurs is called "the threshold region."

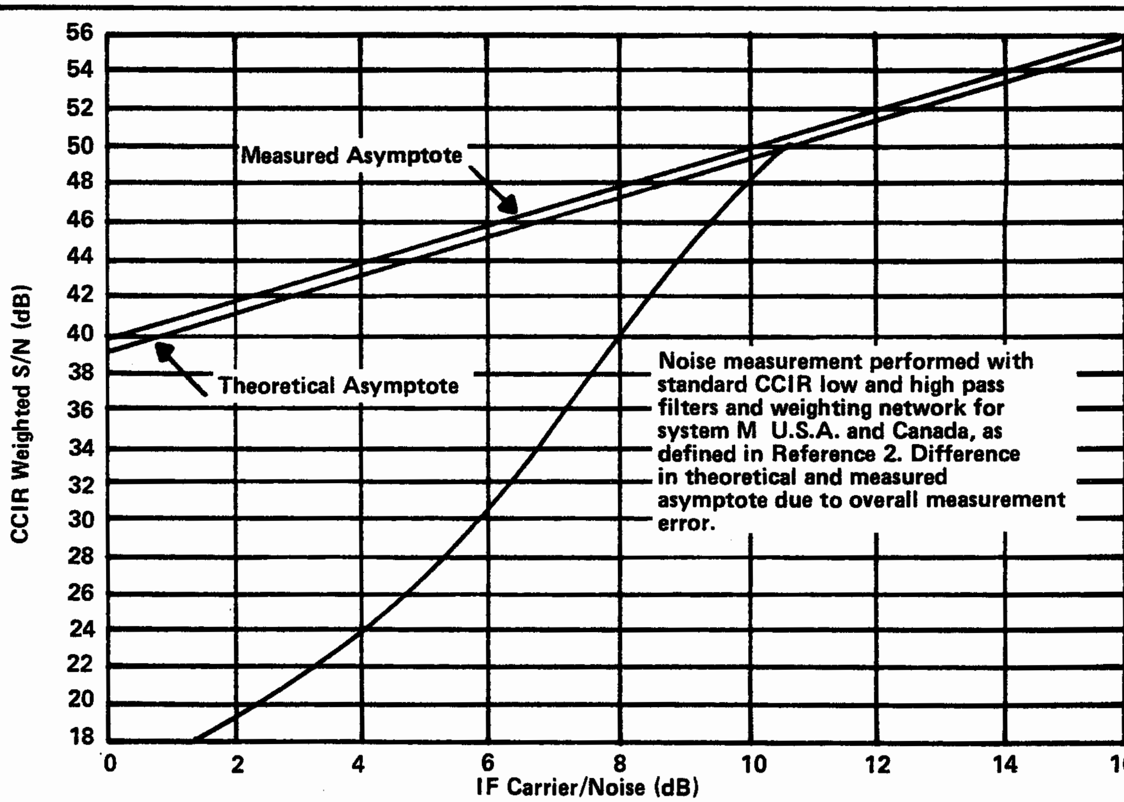


Figure 8. Model 411WB Receiver S/N vs. C/N

Near threshold, the preceding analysis is not sufficient in itself to describe the quality of the received signal. In this region, the thermal noise peaks in the IF band have a finite probability of exceeding the carrier level, resulting in an apparent phase inversion in the composite signal. The FM discriminator then reacts to this with a rapid transition, which results in a baseband transient or impulse disturbance. The threshold for the onset of this phenomenon depends on the nature and design of the video demodulator (The term "threshold" is used in a variety of meanings for video demodulators, and this must be examined carefully for the particular problem under consideration). For standard equipment presently in use in commercial satellite communications, the impulse noise becomes noticeable first at about 11 dB C/N. Threshold extension equipment is available which suppresses the impulse noise down to a C/N ratio of 7 to 9 dB. This topic will be discussed further in one of the following papers.

Half Transponder Video

The demand for transponder space has increased dramatically during the last few years, thus greatly exceeding the capacity of existing satellites. One natural solution to this problem is to operate two simultaneous video signals in one transponder. This system is generally known as half-transponder video. Half-transponder video has been used very successfully by INTELSAT for several years. Usually, the video carrier shares a transponder with a carrier of some other service such as an FDM message, or SCPC carriers.

The problems associated with half-transponder video signals can be identified quickly by examining a link analysis for this format.

Generally the signal-to-noise performance of half-transponder video will be about 10 dB lower than that for full-transponder video assuming that the receive earth station has the same G/T in both cases. There are two reasons for this reduction in received signal quality. First, the allowed transmission bandwidth is less than that allowed for full-transponder video; thus, the deviation of the FM carrier due to the video signal is lower and, correspondingly, the FM improvement is lower. Second, since there are two carriers using the same transponder, the total available EIRP for each signal is lower than for one carrier per transponder, and the transponder power must be reduced to avoid intermodulation problems.

As an example, assume a saturated transponder output of 34 dBW at the receive earth station location. In order to avoid intermodulation problems, the total power output of the transponder must be backed off about 5 dB depending on the specific transponder characteristics. Since the total output power must be shared between two equal video carriers, each will have a total power 3 dB lower than the total output power of the transponder or 8 dB lower than the saturated output power of the transponder.

The downlink C/N can be stated as:

$$C/N = (EIRP)_{sat} - L_s + G/T - 10 \log B - L_m \qquad (22)$$

If we assume the same earth station G/T of 26.8 dB/°K, a 1-dB margin and a saturated transponder EIRP of 34 dB, then for our example, we find a received C/N of:

$$C/N = (34 - 8) - 196.4 + 26.8 - 72.4 + 228.6 - 1$$

$$C/N = 11.6 \text{ dB}$$

The actual C/N will be somewhat less than this since the effects of interference and uplink C/N have been ignored.

The peak deviation of the video signal for half transponder carriers is usually 6.7 MHz and the IF filters are 17.5 MHz. These parameters can be used to convert to received video signal-to-noise as follows:

$$S/N = 11.6 + 10 \log 3 \left(\frac{6.7}{4.2}\right)^2 + 10 \log \left(\frac{17.5}{8.4}\right) + 18.9$$

$$S/N = 42.5 \text{ dB}$$

Notice that this is about 10 dB lower video signal-to-noise than was calculated previously in the full-transponder carrier example with the same earth station G/T.

Several conclusions about half-transponder operation can be reached from the example. First, the demodulated received signal performance is much poorer than with full-transponder video. Second, small aperture antennas will not provide enough C/N to operate the FM demodulator above threshold. Third, since the effects of subcarriers were not considered in the example, their addition to the main carrier will result in further degradation to the final video signal-to-noise performance.

Audio Subcarrier

The audio subcarrier $(S/N)_{sc}$ ahead of the subcarrier discriminator can be shown to be:

$$(S/N)_{sc} = C/N_0 + 10 \log \left[1/2 \left(\frac{\Delta F_c^{\,2}}{f_{sc}^{\,2} B_{sc}} \right) \right] - 10 \log \left[1 + \frac{1}{12} \left[\frac{B_{sc}}{f_{sc}} \right]^2 \right] \qquad (23)$$

where:

F_c = peak deviation of the main carrier by the subcarrier

f_{sc} = subcarrier frequency

B_{sc} = subcarrier filter noise bandwidth

The last term in the expression for $(S/N)_{sc}$ accounts for the slope of the baseband noise power density across the subcarrier filter bandwidth. It is usually less than 0.01 dB and may be neglected.

This expression may be rewritten after the correct substitution as:

$$(C/N)_{sc} = C/N + 10 \log \left(\frac{B_{IF}}{2\, B_{sc}} \right) + 10 \log \left(\frac{\Delta F_c}{f_{sc}} \right)^2 \qquad (24)$$

where:

B_{IF} = IF noise bandwidth

Once the $(C/N)_{sc}$ ratio has been determined, the actual audio signal-to-noise ratio $(S/N)_a$ may be determined by looking at the FM improvement, the bandwidth improvement and the pre/deemphasis curve improvement. This reduces to:

$$(S/N)_a = (C/N)_{sc}\ (3)\left(\frac{\Delta F_{sc}}{f_m}\right)^2\left(\frac{B_{sc}}{2\ B_a}\right) + E \qquad (25)$$

or written in terms of logs:

$$(S/N)_a = (C/N)_{sc} + 10 \log 3 \left(\frac{\Delta F_{sc}}{f_m}\right)^2 + 10 \log \left(\frac{B_{sc}}{2\ B_a}\right) + E \qquad (26)$$

where:

f_m = maximum audio frequency

B_a = audio noise bandwidth

ΔF_{sc} = peak subcarrier deviation

E = audio pre/deemphasis advantage = about 12 dB

Example:

With an audio subcarrier frequency of 6.8 MHz, a subcarrier deviation of 2-MHz peak, and an audio frequency bandwidth of 15 kHz, we will calculate the (S/N):

Assume C/N_o = 90 dB

thus:

C/N = 14.4 dB

$$(C/N)_{sc} = 14.4\text{ dB} + 10 \log\left[\frac{36 \times 10^6}{(2)\ (600 \times 10^3)}\right] + 10 \log\left[\frac{2 \times 10^6}{6.8 \times 10^6}\right]^2$$

$(C/N)_{sc}$ = 14.4 dB + 14.8 dB - 10.63 dB

$(C/N)_{sc}$ = 18.6 dB

Notice that the $(C/N)_{sc}$ is greater than the C/N for the main carrier. This reflects the bandwidth advantage of the narrow subcarrier bandwidth.

Now the $(S/N)_a$ can be calculated from equation (26) as follows:

$$(S/N)_a = 18.6 \text{ dB} + 10 \log \left[3\left(\frac{75 \times 10^3}{15 \times 10^3}\right)^2\right] + 10 \log \left[\frac{(600 \times 10^3)}{(2)\ (15 \times 10^3)}\right] + 12$$

$$(S/N)_a = 18.6 \text{ dB} + 18.75 \text{ dB} + 13 \text{ dB} + 12 \text{ dB}$$

$$(C/N)_a = 63 \text{ dB}$$

Cue Multiplex

Additional voice grade audio channels may be multiplexed above the program audio in the audio baseband. The performance of the "cue" channels may be analyzed in much the same manner as the program subcarrier. Final audio signal-to-noise ratio, $(S/N)_{cue}$ is related to the subcarrier $(C/N)_{sc}$, and thus C/N_o, since these channels are carried as FM on the subcarrier.

The expression for $(S/N)_{cue}$ is:

$$(S/N)_{cue} \text{ (weighted)} = C/N_o + 10 \log \left[\frac{1}{4}\left(\frac{\Delta F_c}{f_{sc}}\right)^2 \frac{(\Delta FM)^2}{f_{mc}{}^2 B_{nm}}\right] \qquad (27)$$

where:

ΔFM = peak deviation of subcarrier by multiplex signal

f_{mc} = arithmetic mean center of multiplex slot

B_{nm} = noise bandwidth of cue baseband filter function (including weighting) with respect to flat noise

ΔF_c = peak deviation of the main carrier by the subcarrier

Example:

If we assume a multiplex slot extending from 20.3 kHz to 23.4 kHz then fmc - 21.85 kHz. Without weighting, for an ideal filter extending from 300 to 3400 Hz, Bnm = 3.1 kHz. This approximation can be used with good results. Substituting the previous data we get:

$$(C/N)_{cue} = 90 + 10 \log \left[\frac{1}{4}\left(\frac{2 \times 10^6}{6.8 \times 10^6}\right)^2 \frac{(50 \times 10^3)^2}{(21.85 \times 10^3)^2 (3.1 \times 10^3)}\right]$$

$$(S/N)_{cue} = 45.6 \text{ dB}$$

FDM/FM

FDM/FM is a method of mlultiplexing many voice grade (3.1 kHz) audio channels on to one FM carrier. This is accomplished by converting each channel to an assigned frequency as a SSBSC signal in the baseband frequency range. The total baseband then modulates an FM carrier. Demodulation and de-multiplexing are accomplished in the reverse order at the receive station. Details of this system wll be presented in a later paper.

FDM/FM performance is measured in terms of picowatts of noise per FDM channel. The noise-per-channel is related to the total S/N ratio in the total baseband with a test tone signal. This, in turn, may be related back to C/N ratio in the receive system IF. Table 1 shows typical operating parameters for the INTELSAT IV system.

The relationship between C/N in the IF and S/N in the baseband is:

$$S/N = C/N + 20 \log \left[\frac{\Delta F_{TT}}{f_{ch}}\right] + 10 \log \left[\frac{B_{IF}}{B_{ch}}\right] + P + W \qquad (28)$$

where:

ΔF_{TT} = rms test tone deviation

f_{ch} = highest voice channel frequency

B_{ch} = voice channel bandwidth

P = top channel emphasis improvement factor

W = psophometric weighting improvement factor = 2.5 dB

Once the test tone S/N ratio has been determined, the noise per channel in picowatts may be determined from:

$$\text{Noise} = \log^{-1} \left[\frac{90 - (S/N)}{10}\right] \text{pWp0} \qquad (29)$$

Example:

Consider a 1200-channel system with a top baseband frequency of 5260 kHz and an rms test tone deviation of 650 kHz. The top slot emphasis advantage is 4 dB. C/N will be assembled to be calculated level of 20 dB.

$$S/N = 20 \text{ dB} + 20 \log \left[\frac{650 \times 10^3}{5260 \times 10^3}\right] + 10 \log \left[\frac{36 \times 10^6}{3.1 \times 10^3}\right] + 2.5 \text{ dB} + 4 \text{ dB}$$

$$S/N = 20 \text{ dB} - 18.2 \text{ dB} + 40.7 \text{ dB} + 2.5 \text{ dB} + 4 \text{ dB}$$

$$S/N = 49.0 \text{ dB}$$

Table 1. Global Beam INTELSAT IVA and V Transmission Parameters (Regular FDM/FM Carriers)

Carrier Capacity (No. of Channels)	Top Base-band Freq. (kHz)	Allocated Satellite BW Unit (MHz)	Occupied Bandwidth (MHz)	Deviation (rms) for 0 dBm0 Test Tone (kHz)	Multi-channel rms Dev. (kHz)	Carrier-to-Total Noise Temp. Ratio at Operating Point (8000 + 200 pWOp from RF Sources) (dBW/K)	Carrier-to-Noise Ratio in Occupied BW (dB)	Ratio of Un-modulated Carrier Power to Maximum Carrier Power Density Under Full Load Condition (dB/4 kHz)
n	f_m	b_a	b_o	f_r	f_{mc}	C/T	C/N	
12+	60.0	1.25	1.125	109	159	154.7	13.4	20.0
24	108.0	2.5	2.00	164	275	153.0	12.7	22.3
36	156.0	2.5	2.25	168	307	150.0	15.1	22.8
48	204.0	2.5	2.25	151	292	146.7	18.4	22.6
60	252.0	2.5	2.25	136	276	144.0	21.1	22.4
60	252.0	5.0	4.00	270	546	149.9	12.7	25.3
72	300.0	5.0	4.50	294	616	149.1	13.0	25.8
96	408.0	5.0	4.50	263	584	145.5	16.6	25.6
132	552.0	5.0	4.40	223	529	141.4	20.7	24.2*(X 1)
96	408.0	7.5	5.90	360	799	148.2	12.7	27.0
132	552.0	7.5	6.75	376	891	145.9	14.4	27.5
192	804.0	7.5	6.40	297	758	140.6	19.9	25.8*(X 1)
132	552.0	10.0	7.50	430	1020	147.1	12.7	28.0
192	804.0	10.0	9.00	457	1167	144.4	14.7	28.6
252	1052.0	10.0	8.50	358	1009	139.9	19.4	27.0*(X 1)
252	1052.0	15.0	12.40	577	1627	144.1	13.6	30.0
312	1300.0	15.0	13.50	546	1716	141.7	15.6	30.2
432	1796.0	15.0	13.0	401	1479	136.2	21.2	27.6*(X 2)
432**	1796.0	17.5	15.75	517	1919	138.5	18.2	30.8
432	1796.0	20.0	18.0	616	2276	139.9	16.1	31.5
612	2540.0	20.0	17.8	454	1996	134.2	21.9	78.9*(X 2)
432	1796.0	25.0	20.7	729	2688	141.4	14.1	32.2
792	3284.0	25.0	22.4	499	2494	132.8	22.3	30.0*(X 2)
972	4028.0	36.0	36.0	802	4417	135.2	17.8	34.5
1092***	4892.0	36.0	36.0	701	4118	132.4	20.7	32.2*(X 2)

** Contingency Carrier, used only with INTELSAT IVA.
***Not used with INTELSAT V.
+ Approved for use with INTELSAT V only.

$$\text{Noise} = \log^{-1}\left[\frac{90 - 49.0}{10}\right]$$

Noise = 12,589 pWpO

SCPC

SCPC (single-channel-per-carrier) is another method of multiplexing many channels on a single transponder. In this case, however, the channels are not multiplexed on to a signal FM carrier, rather each one has its own carrier and separate frequency assignment within the transponder. SCPC systems may have many operating parameters and modes. Frequency bandwidths can vary from strictly voice grade 3.4 kHz to program channel 15 kHz. Actual transmission may be FM analog or digital. Currently, the largest production systems are FM analog; therefore, only this analysis will appear. Psophometric weighting and emphasis are also used in SCPC. Another performance advantage used in SCPC is called companding. This includes compressing the amplitude range of signal at the transmit system and expanding it at the receive system. The analysis of SCPC performance proceeds along the same lines as any other FM system. The test tone-to-noise ratio depends on the IF C/N ratio along with FM improvement factors, bandwidth improvement, and weighting/emphasis improvement, if any.

The relationship for S/N versus C/N is:

$$S/N = C/N + 10\ \log 3\left(\frac{\Delta F}{f_m}\right)^2 + 10\ \log\left[\frac{B_{IF}}{2B_a}\right] + W + C \qquad (30)$$

where:

ΔF = peak deviation

f_m = highest baseband frequency

B_a = audio noise bandwidth

B_{IF} = IF noise bandwidth

W = emphasis plus weighting improvement factor

c = companding advantage

Example:

Consider voice grade SCPC channel with a peak deviation of 7.3 kHz, an IF bandwidth of 25 kHz and a weighting/emphasis advantage of 7 dB, and a companding advantage of 17 dB.

Assume the C/N to be 10 dB in the IF.

$$S/N = 10 + 10\ \log\left[3\left(\frac{7.3 \times 10^3}{3.4 \times 10^3}\right)^2\right] + 10\ \log\left[\frac{25 \times 10^3}{2(3.4 \times 10^3)}\right]$$

+ 7 dB + 17 dB

S/N = 10 dB + 11.4 dB + 5.6 dB + 7 dB + 17 dB

S/N = 51 dB

This is a test tone-to-noise ratio which can be converted to picowatts (weighted):

$$\text{Noise} = \log^{-1}\left[\frac{90 - 51}{10}\right]$$

Noise = 7943 pWp0

Digital Applications

Digital systems may be incorporated in the satellite link in much the same way as SCPC carriers. A modem takes the raw digital data and modulates it on to an IF carrier in the 70-MHz IF band. This signal can then be upconverted to the 6-GHz satellite uplink band. This system will be more fully explained in a later paper in the digest.

The usual method of performance comparison between digital systems is to compare their bit error rates (BER), which are the rates at which errors in bits are made during transmission. For example, a BER of 10^{-5} would mean that for every 10^5 bits sent, there would be one error. If the transmission rate were 10^5/second, then one error would be made every second.

The digital data may be modulated on to the RF carrier in one of several different ways. Some of these are FSK (frequency shift keying), PSK (phase shift keying), BPSK (biphase PSK) and OOK (on-off keying). Each type of transmission has advantages and disadvantages.

The BER analysis of a particular mode of transmission must proceed from a noise model since incoming data bits will have a signal-to-noise ratio which depends on the link parameters. Since electrical noise is often the summation of the effects of a large number of randomly moving electrons, it can be considered to have a Gaussian distribution. A probable distribution function (PDF) for a Gaussian-distributed variate is shown in Figure 9.

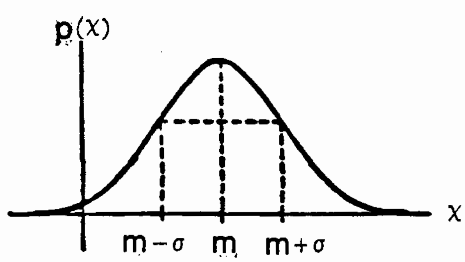


Figure 9. Gaussian PDF

This function is described by:

$$p(\chi) \frac{1}{\sigma \sqrt{2\pi}} \exp\left[-(\chi-m)^2/2\sigma^2\right] \tag{31}$$

where:

m = mean value of χ

σ = standard deviation of χ

Notice in the figure that χ will be most likely found, at m. To find the probability that χ is within a particular range of values, we integrate $p(\chi)$ over that area.

Now if the demodulated digital data is considered to be 0 and 1 at levels -A and +A, we will get two levels with Gaussian noise impressed on them as shown in Figure 10. If the threshold level (or decision level) is set at 0 as shown in the figure, then the probability of mistaking a 0 for a 1 is shown as area P_{e1} in the figure. Likewise, the probability of mistaking a 1 for a 0 is shown as area P_{eo}. The total probable area is then $P_{eo} + P_{e1}$.

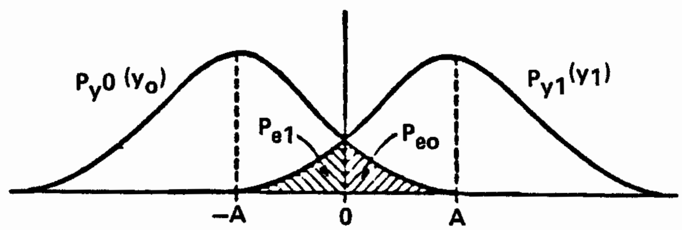


Figure 10. Signal-plus-noise PDF's for Polar Binary Signalling

If the correct integrals are taken:

$$P_{eo} = P_{e1} = \frac{1}{\sigma \sqrt{2\pi}} \int_0^{\infty} \exp\left[-(\lambda_o + A_1)^2/2\sigma^2\right] d\lambda_o \tag{32}$$

If we let $A = \frac{Y_o + A_1}{\sigma}$, then,

$$P_{eo} = P_{e1} = \frac{1}{\sqrt{2\pi}} \int_{A_1/\sigma}^{\infty} \exp\left(-A^2/2\right) dA \tag{33}$$

Now if we take the signal power to be A^2 and the noise power σ^2 = nB where n is the noise density and B is the bandwidth, then:

$\frac{A}{\sigma}^2$ is the S/N ratio

Then the probability of error can be taken to be:

$$P_e = \left[\frac{2}{\sqrt{2\pi(S/N)}}\, e^{-(S/N)}\right]/2 \qquad (34)$$

Of course, this is a very simplified analysis for a polar binary baseband signal. However, the same techniques may be used to generate BER curves for any mode of transmission to be used. Sample curves are shown in Figure 11. Here the BER is shown as a function of received signal-to-noise ratio in dB.

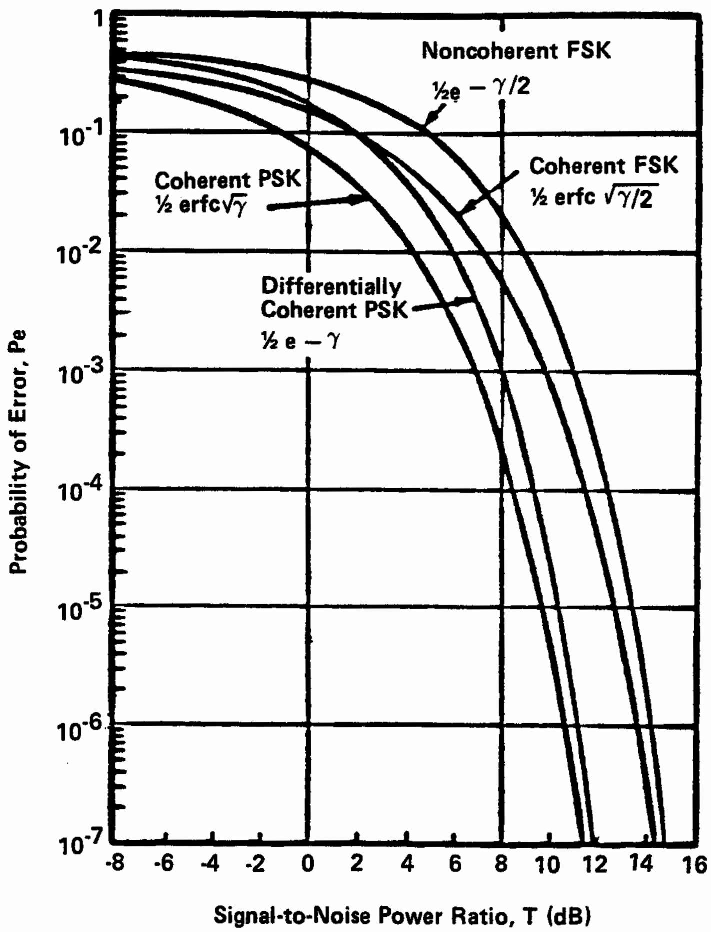


Figure 11. Error-Rates for Several Binary Systems

Other factors can be used to influence the BER for a particular signal level. For example, coding may cause the BER to be several orders of magnitude lower that the theoretical curves shown in the last figure.

Example:

Consider the BER curves shown in Figure 12. This curve shows the BER versus energy per bit of a 56 Kb/s system which uses 84 kHz of bandwidth. The energy per bit is related to C/N by the following relationship:

$$E_b/N_o = C/N - 10 \log BR + 10 \log B_{IF} \tag{35}$$

where:

E_b/N_o = energy/bit per Hz

BR = bit rate

B_{IF} = IF transmission bandwidth

If we have a C/N = 9.3 dB, then:

$$E_b/N_o = 9.3 \text{ dB} - 10 \log (56 \times 10^3) + 10 \log (84 \times 10^3)$$

$$E_b/N_o = 9.3 \text{ dB} - 47.5 \text{ dB} + 49.2 \text{ dB}$$

$$E_b/N_o = 11.0 \text{ dB}$$

From Figure 12, this corresponds to a BER of 3 x 10^{-6}.

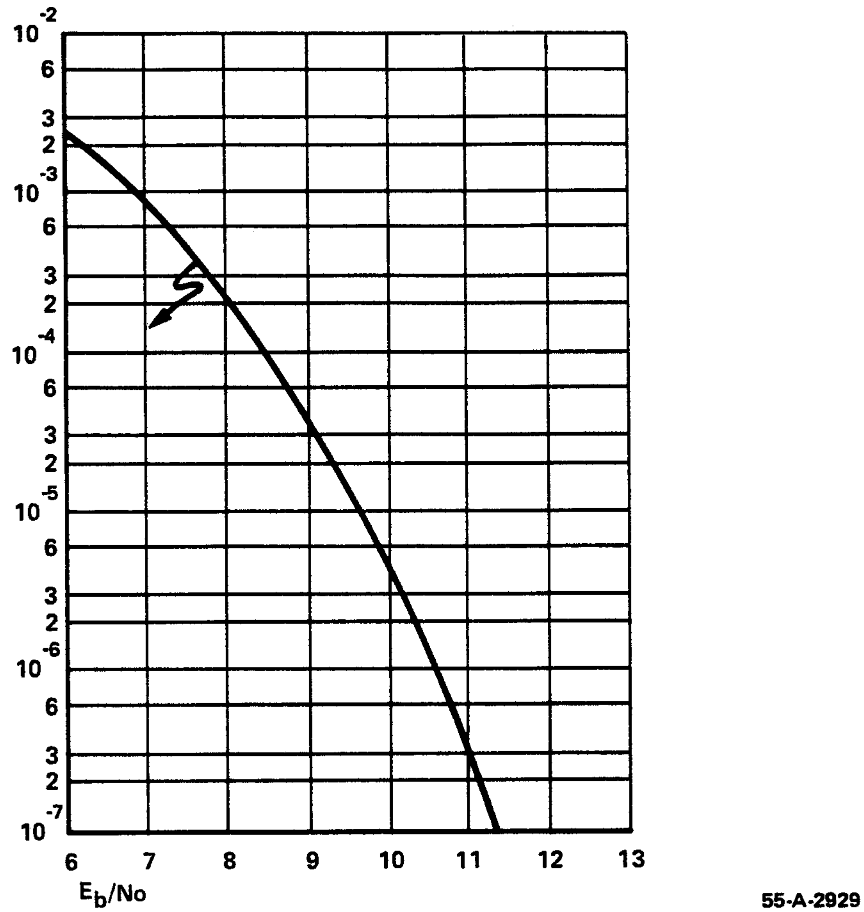


Figure 12. Plot of Theoretical $E_{b/No}$ vs. BER

Summary

In this paper we have looked at satellite link performance and the effect of earth station characteristics on that performance. In addition, various modes of transmission have been examined briefly to see how link performance affects their respective performance criteria. Finally, examples have been examined to illustrate typical numbers which occur in satellite link analysis. A more complete link budget is shown in Appendix B to this paper. Appendix A contains several relationships which may be of interest in satellite link analysis.

References

1. L. Clayton, "FM Television Signal-to-Noise Ratio," IEEE Transactions on Cable Television

2. C.C.I.R. (International Radio Consultative Committee), "Transmission of Sound Broadcasting and Television Signals over Long Distances (CMTT)", XIIIth Plenary Assembly, Vo. XII, International Telecommunication Union, Geneva, 1975, pp. 21-41.

3. J.P. Aasteund, "Aids for Domsat Communications System Performance Calculations, NTC 1974.

4. A. Bruce Carlson, Communication Systems, McGraw-Hill, 1975.

5. Stein and Jones, Modern Communications Principles, McGraw-Hill, 1967.

Appendix A

Geostationary Satellite Orbit Geometry

For a geostationary satellite, the orbit is a circle lying in the equatorial plane which, from simple mechanics has a radius:

$$R - 6.611\ R_0$$

Where R_0 is the radius of the earth (3963 statute miles). The location of the satellite is specified by the corresponding longitude coordinate, L_{sat}. For a station at latitude H and longitude L, the slant range r from the satellite to the station is found to be:

$$r = 26485\ [1 - 0.295 \cos(H) \cos(L-L_{sat})]^{1/2} \quad \text{(statute mi.)}$$

The corresponding spatial loss factor is:

$$L = 10 \log (4\pi r^2) = 163.6 + 10 \log [1 - 0.295 \cos(H) \cos(L-L_{sat})] \text{ dBm}^2$$

The path time delay is:

$$r = 142 \log (1 - 0.295 \cos B) \text{ dB}$$

Appendix B

Sample Link Budgets

Link Parameters	Units	FDM-FM	Video
UPLINK			
Saturation Flux Density	dBW/m^2	-82.0	-82.0
Input Backoff	dB	4.0	0
Power Sharing	dB	0	0
Operating Flux Density	dBW/m^2	-86.0	-82.0
Isotropic Antenna Area	dB/m^2	-37	-37
Satellite Received Power	dBW	-123.0	-119.0
Path Loss (Clear Weather)	dB	200.1	200.1
Operating EIRP	dBW	+77.1	+81.1
Satellite G/T	dB/°K	-7.4	-7.4
10 Log K	dBW	-228.6	-228.6
10 Log B (36 MHz)	dB	75.6	75.6
C/N (Uplink)	dB	22.6	26.6
DOWNLINK			
Satellite Saturated EIRP	dBW	+34.0	+34.0
Output Backoff	dB	1.0	0
Power Sharing	dB	0	0
Operating EIRP	dBW	33.0	+34.0
Path Loss (Clear Weather)	dB	196.4	196.4
Earth Station G/T	dB/°K	26.8	26.8
10 Log K	dB	-228.6	-228.6
10 Log B (36 MHz)	dB	75.6	75.6
C/N (downlink)	dB	16.4	17.4
LINK PERFORMANCE			
C/N (uplink)	dB	22.6	26.6
C/N (downlink)	dB	16.4	17.4
C/Int	dB	-	20.0
C/IM	dB	-	-
C/N (system)	dB	15.4	14.6
Threshold C/N	dB	10.0	10.0
Fade Margin	dB	5.4	4.6

System Performance

System Parameters	Units	FDM-FM	Video
TV VIDEO			
C/N	dB	-	14.6
Max. Video Frequency	MHz	-	4.2
Overdeviation	dB	-	-
Peak Operating Deviation	MHz	-	10.7
FM Improvement	dB	-	13.2
BW Improvement	dB	-	6.3
Weighting/Emphasis Improvement	dB	-	12.8
P-rms Conversion Factor	dB	-	6.0
Total Improvement	dB	-	38.3
S/N (peak-to-peak/rms-luminance signal)	dB	-	52.9
TV PROGRAM CHANNEL (SUBCARRIER)			
Peak Carrier Deviation	MHz	-	2.0
Subcarrier Frequency	MHz	-	7.5
FM Improvement	dB	-	-11.5
BW Improvement	dB	-	14.0
Total Improvement	dB	-	2.5
C/Nsc (subcarrier)	dB	-	16.9
Peak Subcarrier Deviation	kHz	-	75
Max Audio Frequency	kHz	-	15
FM Improvement	dB	-	18.8
BW Improvement	dB	-	13.8
Emphasis Improvement	dB	-	12.0
Total Improvement	dB	-	44.6
S/N (audio)	dB	-	59.2
FDM/FM			
Number Channels	-	1200	-
Test Tone Deviation (rms)	kHz	650	-
Top Baseband Frequency	kHz	5260	-
FM Improvement	dB	18.2	-
BW Improvement	dB	40.7	-
Weighting Improvement	dB	2.5	-
Emphasis (top slot) Improvement	dB	4.0	-
Total Improvement	dB	29.0	-
TT/N	dB	49.0	-
Noise	pWpO	12589	-

An Introduction to FDM/FM Communications Via Satellite

Introduction

Early in the days of telephone, each voice channel was carried by a pair of wires running between exchange offices. As the number of telephones grew (20 million in 1920), the need for combining channels together for more efficient transmission became apparent.

The first multiplexing schemes used what is called frequency division multiplexing; each voice channel is assigned a particular slot in a composite baseband signal. FDM systems are now in wide use in practically all transmission media; coaxial cable, microwave radio, and satellite communications. These systems have a capacity of twelve- to thirty-six-hundred channels.

How FDM Works

An FDM composite baseband is formed by generating a single sideband AM signal with each voice channel. Every channel has its own carrier frequency, and carrier frequencies are spaced 4 kHz apart. In other words, the voice channels are stacked end-on-end in frequency. Figure 1 shows a typical arrangement of 12 voice channels, multiplexed together to form what is called a group (type A). Note that the actual voice band is 300 to 3400 Hz, but 0 to 4 kHz is reserved per channel. The extra bandwidth is used for filter guard bands.

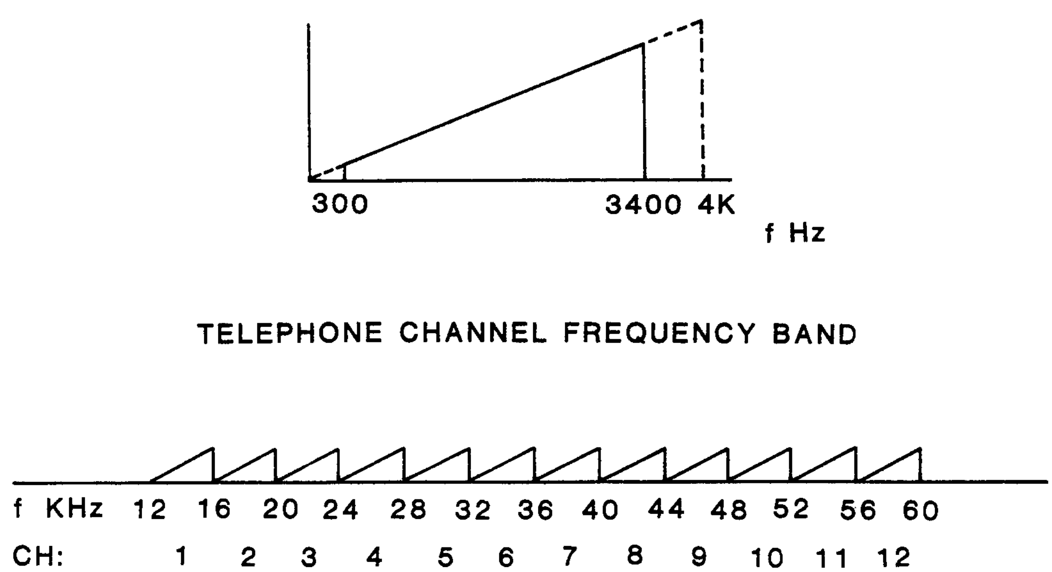


Figure 1. Type A Group

Once the voice channels are multiplexed into twelve channel groups, they are further combined to form higher capacity systems. Five groups are multiplexed together to form a supergroup. Again, single sideband modulation is used to combine the groups together, much as the voice channels are multiplexed together to form a group.

Various multiplex hierarchies are used by different organizations. Any number of 12 channel groups can be multiplexed together to generate a wide range of channel capacities.

Characteristic of the FDM Signal

Design of equipment for FM/FDM signals requires a good knowledge of the characteristics of an FDM signal. Since the FDM signal is composed of a number (12 to 3600) of telephone voice channels multiplexed together, it is instructive to first examine the characteristics of a single telephone voice circuit.

A telephone circuit designed to carry only one voice channel must be able to pass the peak signal level of speech without significant distortion. The peak to rms ratio of the speech signal varies with each talker, and the mean power level of speech varies with each talker as well; therefore, the single channel amplifier must be able to handle the peak signal level of the loudest expected talker.

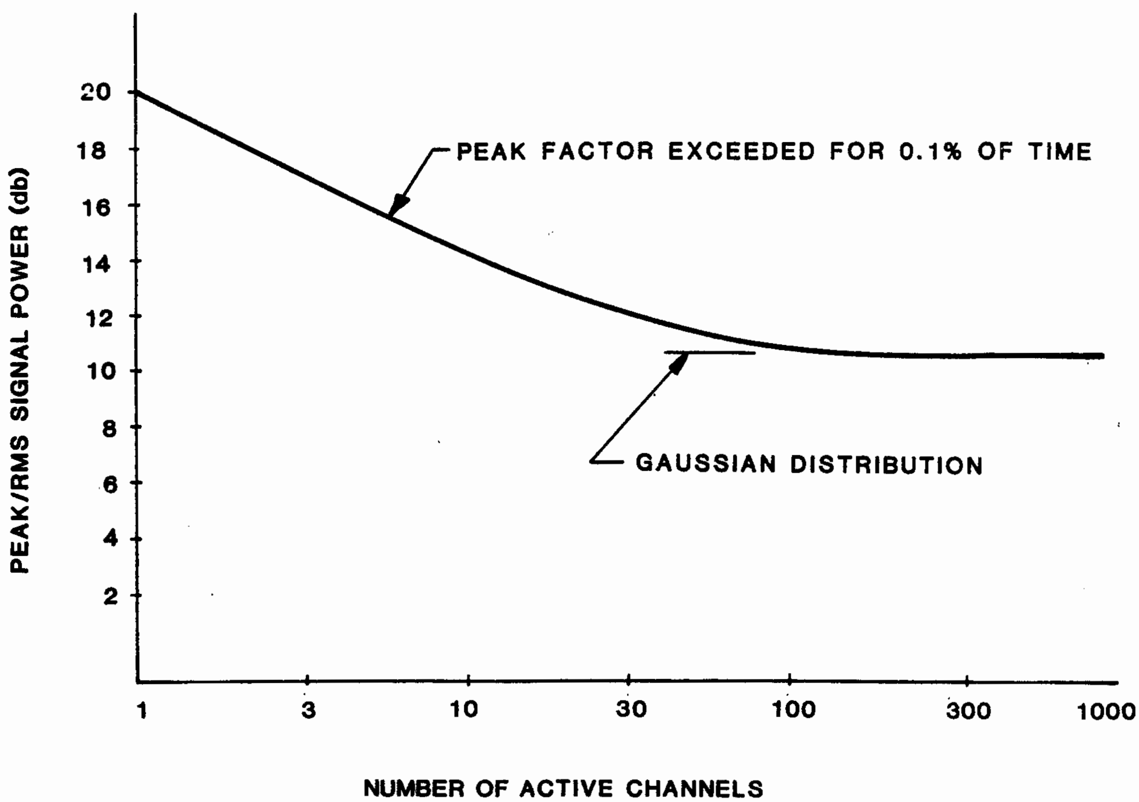


Figure 2. Variation of Peak Factor with Number of Active Channels

When many voice channels are added together, as in an FDM system, the composite signal has characteristics which depend not only on the talker characteristics but also on the number of channels in the system. Many studies have been done to examine the amplitude distributions of FDM baseband signals, the earliest of which were done by Holbrook and Dixon of Bell Laboratories in 1939. Results of these studies show the relationship between the number of active voice channels and the probability that a certain peak to rms ratio will be exceeded for some small (e.g., 0.1%) percentage of time. For a large number of channels (>240), this amplitude distribution is seen to be essentially Gaussian; that is, the composite baseband signal loads similar to white noise that has been passed through appropriate bandlimiting filters. Figure 2 shows the variation of peak factor with number of active channels. In designing FDM equipment, the peak factors and the percentage of active channels (activity factor) must be taken into account. The activity factor is defined as the fraction of total system capacity which accommodates all the system traffic for all but 1% of the busiest hour. The activity factor tends toward 0.27 for higher channel capacities, and 0.58 for low capacities.

This knowledge allows us to predict the peak power level of a multichannel baseband. Such information is essential when designing FDM equipment, since it is the inability to handle peak signal levels that causes distortion of the FDM signal.

As was noted above, the amplitude distribution of an FDM signal with more than 240 channels looks Gaussian. This makes it possible to simulate, for testing purposes, the FDM signal with noise whose power spectral density is constant with frequency; i.e., "white" noise. Common practice is to simulate an FDM signal with bandlimited white noise with a mean power of

$$-15 + 10 \log N \text{ dBm0} \tag{1}$$

for channel capacities $N > 240$. This takes into account the average talker power, activity coefficient, and contribution of signaling to the total baseband power.

It is also common practice to simulate basebands of less than 240 channels with bandlimited white noise with a mean power of

$$-1 + 4 \log N \text{ dBm0} \tag{2}$$

for $12 < N < 240$.

The amplitude distribution of a 12- to 240-channel baseband is not Gaussian; in fact, the mean power of the noise test signal departs (above) the actual mean power of the baseband signal by about 5 dB for a 12-channel system. A happy coincidence, however, is that the peak signal levels of the Gaussian test signal and the actual FDM baseband coincide within about 0.5 dB. White noise simulation of the baseband is then a valid method for testing low-capacity FDM systems.

Equations (1) and (2) are used to calculate the noise load ratio, or multichannel load factor. Figure 3 shows the relationship between actual mean and peak levels for the baseband signal and the white noise signal used for testing.

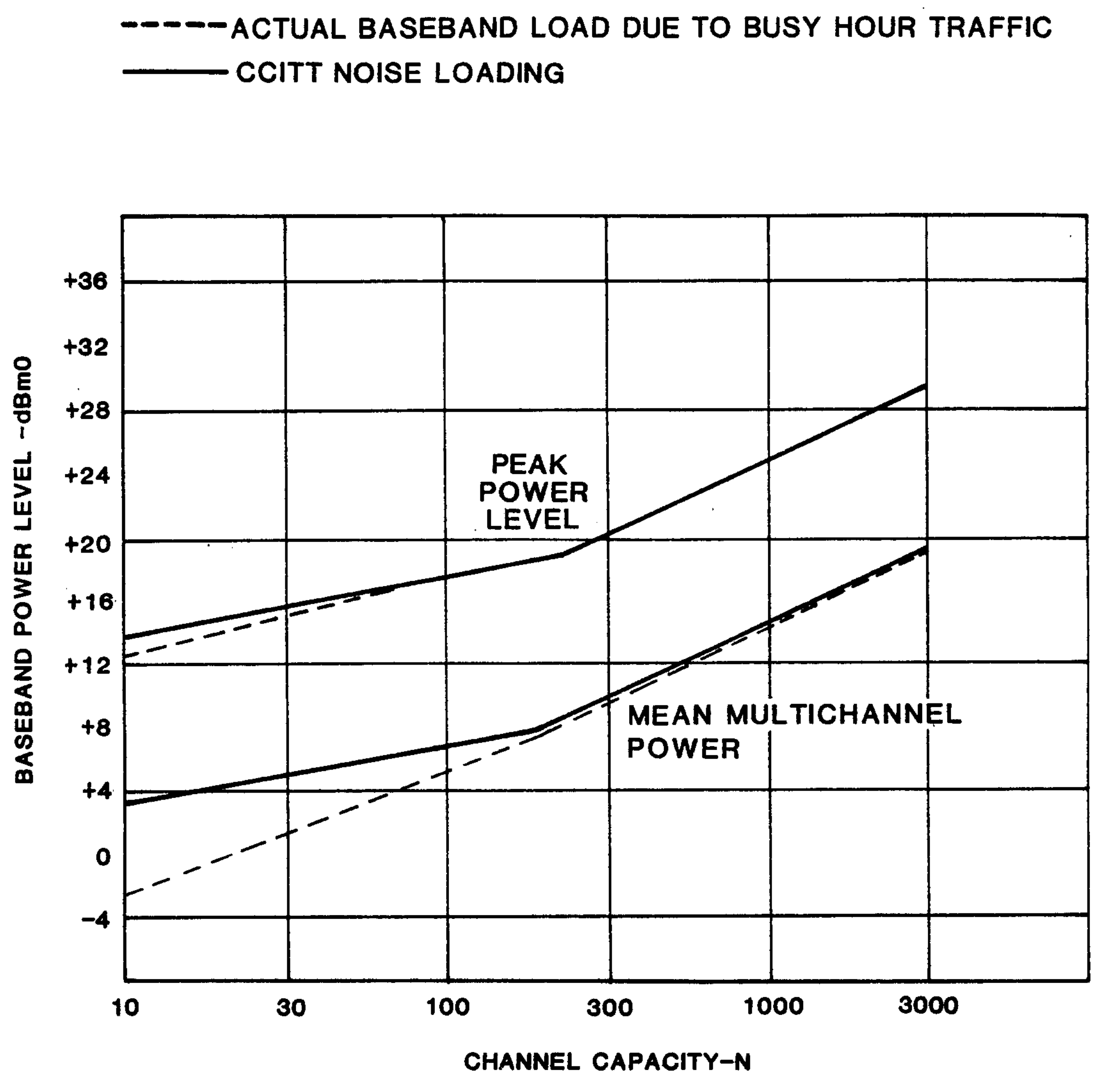


Figure 3. Baseband Power Levels as in Function of Channel Capacity

Testing the FDM Channel: NPR and S/N

The use of white noise testing in developmental work and proof of performance testing is universal in FDM communication systems. The objective is to ensure good quality telephone circuits through the system. A typical requirement for a satellite link is 51 dB weighted* signal-to-noise.

To test the FDM channel, bandlimited white noise with mean power given by equations (1) or (2) is applied to the system. Also, bandstop, or notch, filters are inserted in line. When such a signal is passed through an FDM system, the depth of the notch in the noise spectrum will decrease due to

*Most satellite systems are calculated and measured using CCITT psophometric weighting. The Bell system uses C-weighting. The psophometric weighting advantage is 2.5 dB; it is 1.5 dB for C-weighting.

thermal and intermodulation noise. The ratio of the power spectral density of noise in the flat area of the spectrum to that in the notch is called the noise power ratio (NPR), and is a direct measure of the quality of the FDM system. For CCITT loading, the signal-to-noise ratio of the channel where the notch is placed is given by

$$S/N = NPR + 2 + 6 \log N + W \quad N < 240 \tag{3}$$

$$S/N = NPR + 16 + W \quad N > 240 \tag{4}$$

where N is the system channel capacity, and W is the weighting advantage.

Transmission of the FDM Signal: The FM Satellite Link

In satellite systems, the FDM baseband frequency modulates an RF carrier, usually at 70 MHz, which is then translated to the appropriate frequency for transmission. This resulting carrier is called an FDM/FM carrier. Several such carriers may use portions of a satellite transponder simultaneously, depending on their size. This is known as frequency-division-multiple-access, or FDMA. Such systems are extensively used by INTELSAT and the domestic common carriers.

Signal-to-Noise Ratio in an FDM/FM System

It is important in the design of an FDM/FM satellite system to be able to predict the signal-to-noise ratio of the worst-case channel in the FDM baseband. This is easily done by examining the basic FM equations.

For an FM system operating in the linear region above threshold, the power spectral density of the demodulator output noise can be shown to be

$$S(f) = \frac{f^2}{2\left(C/N_o\right)} \text{ for } f < {}^{B}/_{2} \tag{5}$$

where f is the baseband frequency.

C/N_o is the input carrier-to-noise power density and B is the predetection (IF) bandwidth.

The signal power at the demodulator output is

$$P_s = \frac{\Delta f^2}{2} \tag{6}$$

where Δf is the peak-per-channel deviation.

To find the signal-to-noise ratio in a particular baseband channel, we find the total noise power in that channel by integrating (5), then forming the ratio of P_s to the noise power.

$$P_n = 2 \cdot \int_{fm - \frac{b_{ch}}{2}}^{fm + \frac{b_{ch}}{2}} S(f)\, df \tag{7}$$

where fm is the center of the desired channel, b_{ch} is the telephone channel bandwidth (3100 Hz)

This integral can be approximated by assuming that S(f) is constant over the band of interest. Then

$$P_n = \frac{b_{ch} \cdot fm^2}{(C/N_o)} \tag{8}$$

So, the output S/N ratio is

$$S/N = C/N_o \cdot \frac{1}{2\, b_{ch}} \cdot \left(\frac{\Delta f}{fm}\right)^2 \tag{9}$$

or, in logarithmic form,

$$S/N\ dB = C/N_o - 10 \log (2 \cdot b_{ch}) + 20 \log \left(\frac{\Delta f}{fm}\right) \tag{10}$$

where C/N_o is now a dB ratio, rather than a power ratio. This equation does not include the effects of weighting or preemphasis, discussed below.

Preemphasis

A quick look at equation (5) shows that the output noise power spectral density is parabolic with baseband frequency. If we did nothing to change this situation, the channels at the high baseband frequencies would pay a large signal/noise penalty. Preemphasis is used in FM transmitters to correct this problem. The preemphasis network has increasing gain with increasing frequency; it is inserted in the baseband chain before the FM modulator. The effect is to increase the deviation of the higher frequency channels with respect to the lower channels, thus improving the output signal-to-noise ratio for those channels. A corresponding deemphasis network in the receiver restores all channels to their proper amplitude. Now, a more complete version of equation (10) is

$$S/N\ (dB) = C/N_o - 10 \log (b_{ch}) + 20 \log \left(\frac{\Delta f_{tt}}{fm}\right) + P + W \tag{11}$$

where P is the preemphasis gain (or loss)in dB relative to the gain at the test tone frequency, and W is the weighting advantage (2.5 dB for psophometric weighting). In this equation, Δf_{tt} is the rms test-tone deviation at the test-tone frequency, which is the 0-dB gain frequency of the preemphasis/deemphasis networks. The pre/deemphasis networks in general use today are specified by CCIR Recommendation No. 464.

Degradation of the Output S/N: Imperfect Electronics

Although the signal-to-noise ratio of a voice channel in an FDM/FM satellite system is largely a function of the received carrier-to-noise power-density ratio, the earth station and satellite electronics can contribute to channel noise. In fact, at high C/N_0 levels, this noise will dominate. Prudent design requires a thorough understanding of the distortion mechanisms involved. A block diagram discussion of the major parts of a message exciter and receiver will show potential problem areas, and what effect they have on system performance. Figure 4 is a block diagram of the Scientific-Atlanta Model 7560 Message Exciter; Figure 5 is the complementary Model 7510 Message Receiver.

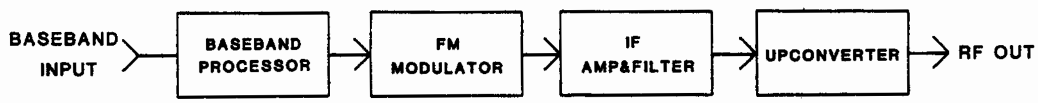


Figure 4. Message Exciter Block Diagram

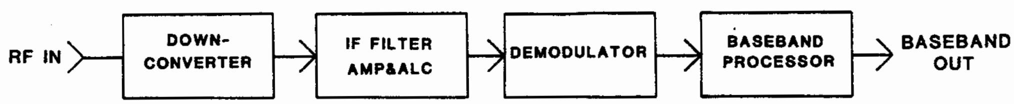


Figure 5. Message Receiver Block Diagram

a. Baseband Processor. In the message exciter, the baseband processor amplifies the baseband to the proper level, preemphasizes the signal, and prevents over-deviation of the carrier under heavy loading conditions. Two sources of noise arise in the baseband circuits--thermal and intermodulation noise. Thermal noise becomes significant at low baseband signal levels. Intermodulation noise is caused by nonlinear distortion in amplifiers, and becomes important at high baseband levels; i.e., when the baseband is heavily loaded. The baseband circuits must be characterized carefully for both thermal and intermodulation noise performance to ensure minimal contribution to channel noise. The baseband circuits in the receiver perform complementary functions to those in the exciter.

b. Wideband Modulator and Demodulator. In the exciter, the wideband modulator frequency modulates an RF carrier with the FDM baseband. If the modulator voltage to frequency transfer characteristic is nonlinear, distortion will result. This distortion is equivalent to nonlinear distortion in baseband amplifiers, and results in intermodulation noise. In addition, residual phase noise of the modulator can degrade the system noise floor. Similar comments apply to the receiver demodulator.

c. IF Filters. Bandlimiting filters are used in both the receiver and exciter. Linear distortion in these filters can cause intermodulation noise in the system. Linear distortion is any amplitude or group delay variation (phase nonlinearity) across the RF bandwidth of the carrier. Linear delay distortion results in second-order intermodulation products in the multiplex signal whose total power is proportional to the square of the linear delay component. Parabolic delay distortion results in third-order intermodulation products, with power again proportional to the square of the parabolic delay. Delay distortion can result not only from IF filters, but also from any other bandpass element in the system. Satellite transponders exhibit delay distortion, and are typically equalized at the transmit site.

d. Noise Due to Up/Downconverters. Up/downconverters contribute to system noise in two ways--residual phase noise of local oscillators and group delay distortion. Residual phase noise of local oscillators directly adds to the thermal noise floor of the system. Typically, this is only significant in the lowest channels of an FDM baseband.

There are other sources of distortion and noise in a typical earth station. Multiple carriers in a high-power amplifier give rise to intermodulation products. Group delay distortion can be caused by RF reflections in cables or interfaces with poor impedance match.

Good design practice generally allocates about 10% of the total noise power in a voice channel to earth station equipment noise. Equipment designers must be familiar with all these distortion mechanisms in order to design cost effective, high-performance communications equipment.

References

1. M. J. Tant, "Multichannel Communications Systems and White Noise Testing," White Crescent Press, London, 1974

2. "Reference Data for Satellite Communications Earth Stations," ITT Space Communications, 1973

3. M. R. Freeling and W. H. Braun, "Maximum Transponder Capacity for Transmission of FDM/FM Channels," RCA Review, September 1980

4. B. W. Brinegar, "Evaluation of Intrinsic Noise in FDM/FM Systems," Satellite Communications Symposium, 1982

FM SCPC Systems

Tommy Brigman and Richard Harris

Introduction

FM SCPC is a communication system which provides efficient and flexible use of available satellite transponder space. It is particularly suited to international and domestic markets using a large number of remote locations with relatively low traffic requirements. An SCPC system can be initially configured with small number of manually controlled channels and later expanded to hundreds of channels utilizing Demand Assignment Multiple Access (DAMA). This flexibility provides a means of implementing a communication system cost effectively upfront and expanding economically as the requirements increase. A typical FM SCPC system is shown in Figure 1.

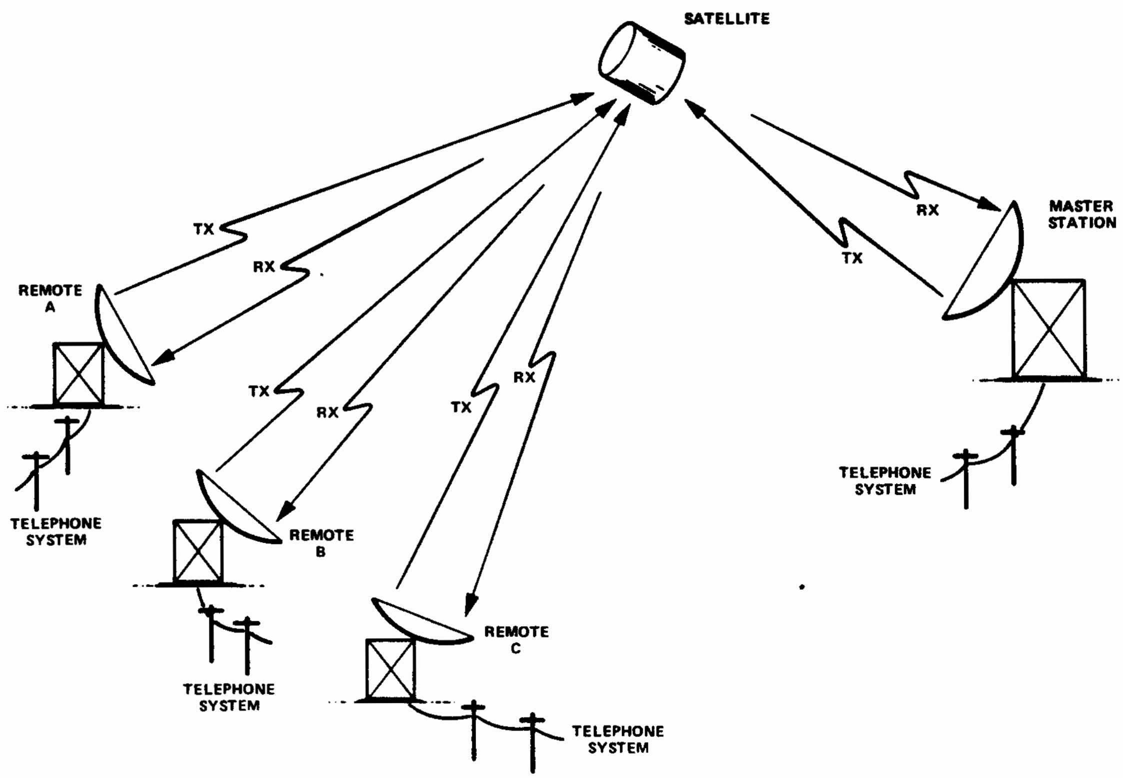


Figure 1. Typical FM SCPC System

One such system is the Australian Outback System described in the 1980 Scientific-Atlanta Satellite Symposium. This is a particularly good example of how an SCPC system architecture can be structured to allow growth from a simple beginning to a complex system with 3,000 channels. A system design such as this provides a very cost-effective domestic communication system which can be expanded to provide extensive national as well as international telephone service.

The initial conception of an SCPC system may involve limited economic resources and limited access satellite transponder space. In many situations much of the system is predefined and maximum use of the available resources is the major consideration. A typical SCPC earth station is shown in Figure 2. Again, using the Australian Outback System as a reference, a cost tradeoff analysis was presented and is included in Appendix A.

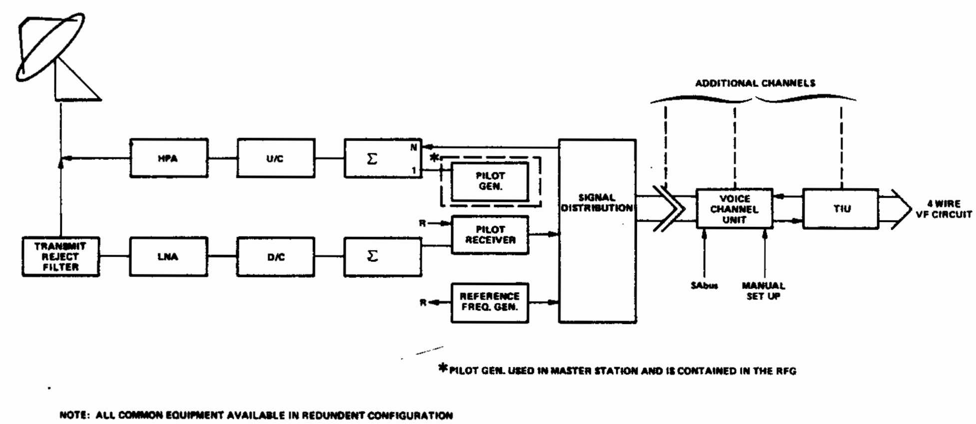


Figure 2. Typical SCPC Earth Station

FM SCPC System Design

To illustrate techniques and principles of a typical SCPC system, an example system will be analyzed. This example system has a frequency plan illustrated in Figure 3 with half-transponder video and 140 channels of SCPC. The video will be uplinked from a different station than the SCPC.

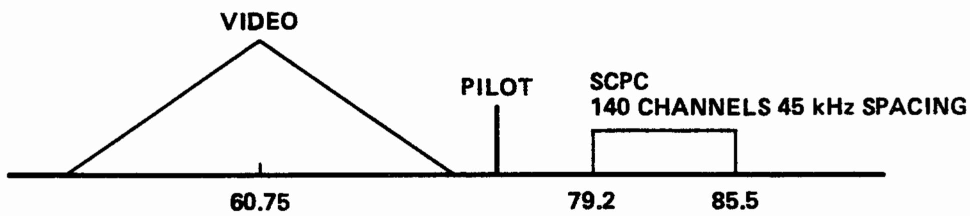


Figure 3. FM SCPC System Design

The satellite to be used has the following transponder characteristics for beam center operation:

Table 1. Transponder Characteristics

Specification	Characteristic
Satellite Antenna Gain	
Receive	18.2 dBi
Transmit	18.2 dBi
Uplink Saturation Flux Density	-77.8 dBW/m^2
Downlink EIRP (Saturated Carrier)	26.1 dBW
Satellite G/T	-11.8 dB/K

The earth stations in this system consist of a MAIN STATION with an 11-meter antenna which is geographically located -2.1 dB from the satellite beam peak, and a REMOTE STATION with a 7-meter antenna which is geographically located -1.3 dB from the satellite beam peak. Table 2 lists the station parameters.

Table 2. Section Parameters

	ANTENNA	TX ANT GAIN	RX ANT GAIN	G/T
MAIN STATION	11-Meter	54.1 dBi	51.7 dBi	31.7 dB/K
REMOTE STATION	7-Meter	49.3 dBi	47.5 dBi	26.5 dB/K

Since the transponder contains both SCPC and video signals, the available transponder power must be shared between the two signals. The design consideration used to divide the power is the receive S/N ratio requirements of the two signals. Working backwards from desired S/N, the power required for each signal can be determined. If this power is more than the available power from the satellite, tradeoffs must be made on the desired video S/N versus the desired SCPC S/N. In this example, the video signal will demand nearly equal to 95 percent of the satellite power leaving 5 percent for the SCPC. This is primarily due to the large pre-detection bandwidth necessary to demodulate the wideband FM video signal.

Another consideration in this design is the satelite backoff determination. The satellite backoff is the reduction of uplink transmit power necessary to place the satellite TWT in an acceptably linear operating region. When the transponder is saturated, the TWT is obviously in an extremely non-linear region. Multicarrier operation at saturation will produce severe intermodulation products that can degrade the link performance drastically. In this example, an input backoff of -8.4 dB and an output backoff of -3.4 dB is used.

Tables 3 and 4 list the transmission and satellite parameters, respectively.

Table 3. Transmission Parameters

Description of Carriers	Parameters	
Characteristic	TV Video	Telephone
Number of carriers	1	56 (40% activity on 140 channels)
Modulation	FM	SCPC
Rx EIRP (Main Station)	20.45 dBW	-3.33 dBW per channel
Rx EIRP (Remote Station)	21.27 dBW	-2.22 dBW per channel
Pre-detection Noise Bandwidth	15.75 MHz	36 kHz per channel
Baseband Noise Bandwidth	5.6 MHz	3.4 kHz per channel
Peak Deviation (ΔF)	7.5 MHz	9.2 kHz per channel
Carson's Rule Bandwidth ($2\ \Delta F + FM$)	25 MHz	24.6 kHz
Highest Baseband Frequency (Fm)	5 MHz	3.1 kHz
Preemphasis Advantage	↓	6.3 dB
Weighting Advantage	16.3 dB (combined)	2.5 dB
Companding Advantage	-	17.0 dB

Table 4. Satellite Parameters

Description		Parameters
$EIRP_{SATELLITE}$ = 26.1 dBW		Beam Center Saturated
$EIRP_{SATELLITE}$ = 24.0 dBW		Main Station Saturated
$EIRP_{SATELLITE}$ = 24.8 dBW		Remote Station Saturated
Input Backoff = -8.4 dB		
Output Backoff = -3.4 dB		
Total Available Downlink EIRP =	22.7 dBW	Beam Center
Total Available Downlink EIRP =	20.6 dBW	Main Station
Total Available Downlink EIRP =	21.4 dBW	Remote Station
Saturation Flux Density =	-77.8 dBW/m^2	Beam Center
Saturation Flux Density =	-75.7 dBW/m^2	Main Station
Saturation Flux Density =	-76.5 dBW/m^2	Main Station

Link Calculations (SCPC per channel)

Main Station

1) Uplink C/N_{ou}

$$C/N_{ou} = S + G/T_{SATELLITE} - 20 \log F_{GHz} - 21.45 + 228.6$$

where:

S = transmit flux density dBW/m^2

G/T_{SAT} = -11.8 dB/K

F_{GHz} = 6

S is the transmit per carrier flux density. It can be calculated from the receive EIRP and going backwards to the transmitter.

S = S' - input backoff - (EIRP' - EIRP)

S' = flux density required to saturate transponder from uplink site

EIRP' = downlink EIRP for saturated carrier

EIRP = downlink EIRP per SCPC carrier

$$S = -75.7 - 8.4 - (20.6 - (-3.33))$$

$$S = -108 \text{ dBW/m}^2$$

$$C/N_{ou} = -108 - 11.8 - 15.56 - 21.45 + 228.6$$

$$C/N_{ou} = 71.79 \text{ dB-Hz}$$

2) Downlink C/N_{od}

$$C/N_{od} = EIRP - L_s + G/T + 228.6$$

L_s = free space loss from satellite to earth station - 196.66

EIRP = satellite EIRP for SCPC channel

G/T = earth station figure of merit

$$C/N_{od} = -3.33 - 196.66 + 31.7 + 228.6$$

$$C/N_{od} = 60.31 \text{ dB-Hz}$$

3) Total Link C/N_o (Assume C/I, carrier-to-interface ratio, is negligible for this case. In many cases C/I is an important consideration.)

$$C/N_o = \frac{1}{\frac{1}{C/N_{ou}} \oplus \frac{1}{C/N_{od}}}$$

$\oplus$ denotes power addition

$$C/N_o = 10 \log \frac{1}{\frac{1}{10^{71.79/10}} + \frac{1}{10^{60.31/10}}}$$

$$C/N_o = 60.01 \text{ dB-Hz}$$

4) Channel S/N Ratio

$$S/N = C/N_o + 10 \log 3 \left(\frac{\Delta F}{f_M}\right)^2 - 10 \log 2B_a + P + C$$

where:

ΔF = peak deviation = 9.2 kHz

f_M = highest baseband frequency = 3.1 kHz

B_a = baseband noise bandwidth = 3.4 kHz

P = preemphasis advantage = 6.3 dB

C = companding advantage = 17.0 dB

$$S/N = 60.01 + 10 \log 3 \left(\frac{9.2\ E3}{3.1\ E3}\right)^2 - 10 \log 2(3.4E3) + 6.3 + 17.0$$

$$S/N = 60.01 + 14.22 - 38.33 + 6.3 + 17.0$$

$$\underline{\underline{S/N = 59.2\ dB}}$$

Remote Station

1) Uplink C/N_{ou}

$C/N_{ou} = S + G/T_{SATELLITE} - 20 \log F_{GHz} - 21.45 + 228.6$

$S = -76.5 - 8.4 - (20.6 - (-2.22))$

$S = -107.72$

$C/N_{ou} = -107.72 - 11.8 - 15.56 - 21.45 + 228.6$

$C/N_{ou} = 72.07$ dB-Hz

2) Downlink C/N_{od}

$C/N_{od} = EIRP - L_s + G/T + 228.6$

$C/N_{od} = -2.22 - 196.66 + 26.5 + 228.6$

$C/N_{od} = 56.22$ dB-Hz

3) Total C/N_o

$$C/N_o = \frac{1}{\frac{1}{C/N_u} \oplus \frac{1}{C/N_{od}}}$$

C/N_o = 56.10 dB-Hz

4) Channel S/N Ratio

$$S/N = C/N_o + 10 \log 3 \left(\frac{\Delta F}{fM}\right)^2 - 10 \log 2 B_a + P + C$$

$$S/N = 56.1 + 14.22 - 38.33 + 6.3 + 17.0$$

$$\underline{\underline{S/N = 55.29 \text{ dB}}}$$

Appendix A

Australian Outback System Tradeoff Analysis

The main system parameters for four channel spacings have been calculated and are shown in Table 1. The systems all meet the specifications for the model. All systems are calculated for the transponder-bandwidth-limited case and three different earth-station system temperatures. The traffic capacity has been calculated using a 2-percent reduction to allow for DAMA control traffic on satellite.

Table 1. Main System Parameters

Parameter	System 1	System 2	System 3	System 4
Channel Spacing (kHz)	60	45	30	22.5
Number of Channels (Duplex)	300	400	600	800
Channel Peak Deviation	21.7	12.2	8.6	5.4
Required C/N_0 dB/Hz	56.2	56.0	59	63
One Hop S/N Ratio (dB)	55.2	50	50	50
Call Blocking Ratio	1 in 100	1 in 100	1 in 100	1 in 100
Downlink $EIRP/_{ch}$ dBW	7.2	6.0	4.2	3.0
Uplink $EIRP/_{ch}$ dBw	46.7	45.5	43.7	42.5
G/T min. dB/K	16.9	17.9	22.7	27.9
Antenna Diameter 1*	2.8	3.11	5.38	9.79
(Meters) 2*	3.2	3.55	6.21	11.3
3*	3.6	4.0	7.0	12.7
HPA $Power/_{ch}$ 1*	3.46	2.13	0.5	0.11
(Watts) 2*	2.65	1.63	0.35	0.08
3*	2.09	1.29	0.27	0.06
C/N dB	10	10	15.84	21.4
Traffic Capacity (Erlangs)	268	371	567	765
Traffic/Station (Erlangs)	0.089	0.124	0.189	0.25

*These items show the parameter values versus the station system temperature. 1, 2, 3 refer to system noise temperatures of 150, 200, and 250K, respectively.

For each system, the channel deviation was chosen for the optimum system. Inspection of the table shows that system 2 meets the S/N and C/N criteria, simultaneously. System 1 betters the S/N specification by 5.2 dB while just meeting the C/N specification. System 3 and 4 better the C/N specification by 5.84 and 11.4 dB, respectively, while just meeting the S/N specification. This is just as predicted in Appendix B. Let us compare System 1 and 2. System 2 carries 38 percent more traffic than System 1. The antennas are close in size, using the rule of thumb cost approximation that the ratio of cost is proportional to the square of the diameter ratio, the System 2 antennas would be about 25 percent more than those in System 1. On the other hand, System 1 requires higher HPA power. There is a good possibility that the station costs would come out equal or very close. In that case, System 1 has little to recommend it, and would not be used. System 2 would be preferred and used up to its maximum traffic load. Assume the traffic load is between 0.124 and 0.189 erlangs per station. This level can be carried by System 3 on one transponder or System 2 using two transponders.

The antennas in System 3 would cost about three times as much as those in System 2. The power amplifier cost would be significantly less in System 3 than in System 2. However, System 2 using two transponders would need frequency-agile up/downconverters under DAMA system control which would add cost. Some simple arithmetic would show which alternative is best. If the costs of both alternatives are close, then the second transponder might be chosen since System 2 with two transponders can carry 35 percent more traffic than System 3 and one transponder. Assume now the traffic load is between 0.189 and 0.25 erlangs power station. The choice is between System 2 and two transponders and System 4 and one transponder. The antenna costs in System 4 would be many times greater than those in System 2. The diameter square rule does not apply here since at 3 meters labor saving manufacturing techniques exist which are not usable at 10 meters. The multiplier may well be 25 or better. For illustration, let us assume: the cost differential between an installed 10-meter antenna and an installed 3-meter antenna is $50,000. This cost difference multiplied by 3,000 is 150 million dollars. System 4 is therefore unlikely to be used.

System Calculations — Downlink

Assume that the SCPC channel has the following parameters: audio range 300 to 3400 Hz, pre/deemphasis break-points 600 and 5000 Hz, 2-for-1 companding, signal-to-noise 50 dB unweighted. The signal-to-noise performance is given by Equation A-1:

$$S/N = C/N_o + 10 \log 3 \left(\frac{f_{pk}}{F_M}\right)^2 - 10 \log 2 B_a + P + C \qquad \text{(A-1)}$$

where:

F_{pk} is the peak deviation in Hertz

B_a is the baseband noise bandwidth

P is the improvement due to pre/deemphasis

C is the improvement due to companding

The emphasis improvement with the characteristics defined in above is 5.2 dB. The subjective companding improvement varies with individual listener and speaker. Experimental results give a number of 12.3 dB observed by 90 percent of listeners. We shall use this number for this system model. The C/N_0 needed to meet the 50 dB specified signal-to-noise is obtained by substituting the specified parameters in Equation A-1. This is the theoretical link performance. To allow for degradation due to uplink performance, intermodulation noise, pointing errors, etc., the C/N_0 is increased by 1 dB. Knowing C/N_0, Equations A-2 and A-3 may be used to calculate various trade-offs between number of channels, satellite EIRP, and earth station G/T.

$$EIRP_{ch} = C/N_0 + S.L. - G/T - 228.6 \qquad (A\text{-}2)$$

$$EIRP_{SAT} = B.O. - EIRP_{ch} = 10 \log N \qquad (A\text{-}3)$$

where:

$EIRP_{ch}$ is the power radiated per channel in dBW

S.L. is a "space loss" factor given by:

$$S.L. = 185.04 + 10 \log (1 - 0.3 \cos H \cos OL) + 20 \log F \qquad (A\text{-}4)$$

where:

H = Latitude of earth station

OL = Difference in longitude between earth station and satellite

F = Frequency in GHz of downlink

G/T = Figure-of-merit of earth station antenna/LNA system

N = Maximum number of simplex SCPC channels in the transponder

B.O. = Backoff from transponder saturation in dB

$EIRP_{SAT}$ = Transponder saturated radiated power in dBW

The maximum number of simplex SCPC channels in the transponder (N) is related to the number of usable duplex trunks (N_T) by the relationship shown in Equation A-5.

$$N_T = \frac{N}{2 \times .4} \qquad (A5)$$

where .4 is the activity factor for VOX operated SCPC.

For a given system, the number of required trunks is determined by the number of stations in the system and the traffic originated at those stations. Once the total Traffic in Erlangs is known, the number of trunks used to carry this traffic at a defined grade of service can be found either from lookup tables or by solving the Erlang B or Poisson traffic equations depending on usage in the particular country. This trunk number is N_T; using Equation A-5 N can be derived. Substituting N in Equation A1-3, $EIRP_{ch}$ can be found. Substituting this value in Equation A-2 gives the necessary G/T.

Uplink Calculation

The uplink $EIRP_u$ can be found from Equation A-6.

$$EIRP_u = \theta\ 228.6 - 10 \log N - B.O._L + L \qquad \text{(A-6)}$$

where

θ = Flux density needed at satellite to produce saturation output EIRP.

(-82 dBW/m^2 for RCA type)

L = spatial loss factor ($L = 10 \log 4 \pi f^2$)

r = slant range station to satellite

$B.O._L$ = Extra backoff needed to allow for non linear input/output curve of the TWT. $B.O._L$ = 5 dB

The amplifier power-out needed is given by:

$$P_{OUT} = EIRP_u - G_1 + 1 \text{ dB}$$

where G_1 is the antenna gain at the uplink frequency and 1 dB is the assumed loss between amplifier and antenna.

Appendix B

Derivation of Optimum SCPC Channel Deviation

1. Voice Channel

Assume that the channel must meet a 50 dB signal-to-noise and a C/N of 10 dB, simultaneously. Assume further that the channel IF bandwidth is equal to the Carson's Rule bandwidth B.

then $B = 2(f_{pk} + 3400)$ (f_{pk} in Hertz)

and $C/N_o = 10 \log 2(f_{pk} + 3400) + 10$ (dB-Hz)

Substituting this equation in Equation A-1 from Appendix A gives:

$$50 = 10 \log 2\,(f_{pk} + 3400) + 10 + 10 \log \frac{3f_{pk}^2}{2 \times 3400^3} + I_p + I_c$$

Substituting for I_p and I_c the values used in Appendix A and reorganizing the equation gives:

$$90 + 10 \log (f_{pk} + 3400) + 20 \log f_{pk} = 123.7$$

Solving for f_{pk} gives the optimum value for this case f_{pk} = 12.2 kHz, giving an IF bandwidth of 31.2 kHz which fits well with 45 kHz spacing. The C/N_o required is 55 dB. If the peak deviation is increased, the C/N_o must be increased to meet the C/N specification and the S/N performance exceeds the specification. This appears when using larger channel spacing such as 60 kHz.

Satellite Link Analysis
Summary

R.B. Harris

Introduction

This paper is intended to provide a reference document for individuals to use in the calculation of satellite link performance. Because of this intent, the discussions of the equations are abbreviated. Hopefully this will be a useful document that can be easily accessed and accurately implemented.

The paper will be divided into two general areas. First is an analysis of the radio link performance resulting in the net carrier-to-noise ratio (C/N_O). Second is an interpretation of the effect of this C/N_O on the end-to-end performance of several services (video, message, SCPC, digital, etc.). This separation emphasizes the fact that the C/N_O analysis follows the same pattern, regardless of the particular service being transmitted.

Link Analysis

The purpose of radio link analysis is to determine the transmission quality that can be expected for signal carriers relayed from one point to another. In satellite communications, two relays are actually performed. One from the transmit earth station to the satellite and the other from the satellite to the receive earth station. For this reason, separate carrier-to-noise calculations will be made on the uplink and downlink.

Uplink $(C/N_O)_u$

The uplink carrier-to-noise ratio can be determined using the following formula:

$$\left(C/N_O\right)_u = S + (G/T)_S - 20 \log F(\text{GHz}) - 21.45 + 228.6 \text{ (dB-Hz)}$$

where:

S = radiated flux density

$(G/T)_S$ = satellite uplink receive system (from contour map)

$F(\text{GHz})$ = uplink frequency in GHz

21.45 dB = $20 \log \left(\frac{4\pi R}{\lambda}\right)$

228.6 dB = $10 \log \left(1.38 \times 10^{-23}\right)$ Boltzmans constant

The radiated flux density can be determined from,

$$S = P_o + G_T - 162.5$$

where:

P_o = HPA output power in dBW

G_T = transmit antenna gain in dBi

162.5 dB = spreading loss = $10 \log 4\pi R^2$

R = distance from earth station to satellite in meters

Downlink $(C/N_o)_D$

The downlink $(C/N_o)_D$ can be calculated using the following equation:

$$\left(C/N_o\right)_D = \left(G/T\right)_e + \left(EIRP\right)_s - 228.6 - L_s - L_p \quad \text{(dB-Hz)}$$

where:

$(G/T)_e$ = earth station figure of merit

$(EIRP)_s$ = satellite downlink EIRP (from contour map)

L_s = $20 \log \left(\frac{4\pi R}{\lambda}\right)$

L_p = $-10 \log P$

λ = downlink signal wavelength

P = polarization efficiency

Overall $(C/N_o)_o$

The overall $(C/N_o)_o$ can be calculated from the $(C/N_o)_D$ and $(C/N_o)_u$. Notice this equation contains a $\oplus$ symbol. This means power addition and requires the objects of the addition to be in watts and not dBW.

$$\left(C/N_o\right)_o = \frac{1}{\frac{1}{\left(C/N_o\right)_u} \oplus \frac{1}{\left(C/N_o\right)_D}} \quad \text{(dB-Hz)}$$

Sometimes it is useful to express the carrier-to-noise in a bandwidth wider than 1 Hz. This is usually expressed as C/N and can be calculated from C/N_0 in the following way:

$$C/N = C/N_0 - 10 \log B_n$$

where:

B_n = noise bandwidth of system where C/N is desired

C/N_0 Summary

The previous equations provide a means to calculate the performance of a satellite link. This performance is expressed as a carrier-to-noise ratio in a 1 Hz bandwidth, providing a number useful in S/N analysis.

S/N Analysis

There are many different communication schemes used in satellite communications. FM techniques have been heavily utilized in such systems as message, video and SCPC. Digital transmissions have recently gained popularity utilizing PM techniques. This section will provide equations to predict S/N or BER performance of these systems, based upon the C/N_0 performance of the satellite link.

NOTE

The following equations provide accurate results for systems operating 2 dB above the demodulation threshold.

Message S/N

Message transmission uses FDM-FM techniques to multiplex many voice channels into one baseband signal. Due to the triagular noise spectrum inherent with FM demodulation, the S/N is dependent on the channel frequency. The following equation can be used to calculate the S/N performance:

$$S/N = C/N_0 - 10 \log \left(2B_{ch}\right) + 20 \log \left(\frac{\Delta F_{tt}}{F_c}\right) + P + W$$

where:

C/N_o = link carrier-to-noise ratio in dB-Hz

B_{ch} = channel noise bandwidth

ΔF_{tt} = peak deviation of IF signal produced by test tone signal

F_c = channel frequency

P = preemphasis advantage

W = weighting advantage

The preemphasis advantage is also dependent on the channel frequency and can be calculated using the following formula:

$$P = 5 - 10 \log \left\{ 1 + \frac{6.90}{1 + \frac{5.25}{\left(\frac{F_r}{F_c}\right) - \left(\frac{F_c}{F_r}\right)^2}} \right\}$$

where:

F_c = channel frequency

F_r = 1.25 x maximum baseband frequency

The weighting advantage can be calculated using,

$$W = 2.5 + 10 \log \left(\frac{B_{ch}}{3,100}\right)$$

SCPC S/N

SCPC stands for "single channel per carrier," implying each channel has its own FM carrier. The following equation can be used to analyze SCPC S/N:

$$S/N = C/N_o + 10 \log 3 \left(\frac{\Delta F}{f_m}\right)^2 - 10 \log 2B_a + P + C$$

where:

C/N_o = link carrier-to-noise ratio in dB-Hz

ΔF = peak deviation of IF signal produced by test tone signal

F_m = highest baseband frequency

B_a = audio filter noise bandwidth

P = preemphasis advantage

C = companding advantage

Video S/N

Baseband video signals are modulated on a single IF carrier, generating a wideband FM signal. This baseband signal includes video information and sync tips for horizontal line synchronization. Since the sync tips do not contain video information, they are excluded from the following S/N calculation:

$$S/N = C/N_o + 10 \log 12 + 20 \log \Delta F_s - 30 \log B_v + P + W$$

where:

C/N_o = link carrier-to-noise ratio in dB-Hz

ΔF_s = peak deviation of IF signal produced by the "video" portion of the baseband signal

B_v = noise bandwidth of the baseband video filter

P = preemphasis advantage

W = weighting advantage

Audio Subcarrier S/N

The audio portion of a television signal is usually transmitted as an FM subcarrier added to the video baseband, forming a composite baseband. On the receive side, the S/N from the video demodulator is the C/N input to the subcarrier demodulator. The following equations present techniques to calculate $(C/N)_{sc}$ and S/N_{prgm}:

$$(C/N)_{sc} = C/N_o + 20 \log \left(\frac{\Delta F_{sc}}{F_{sc}}\right) - 10 \log 2B_{sc}$$

where:

$(C/N)_{sc}$ = carrier-to-noise of the audio subcarrier input to the subcarrier demodulator

C/N_o = link carrier-to-noise in dB-Hz

ΔF_{sc} = peak deviation of IF carrier produced by subcarrier signal

F_{sc} = frequency of subcarrier

B_{sc} = noise bandwidth of predetection subcarrier filter

$$S/N_{prgm} = C/N_o + 20 \log \left(\frac{\Delta F_{sc}}{F_{sc}}\right) + 20 \log \left(\frac{\Delta F_a}{B_a}\right) - 10 \log B_a - 1.25 + P$$

where:

C/N_o = link carrier-to-noise ratio

ΔF_{sc} = peak deviation of IF carrier produced by the subcarrier signal

F_{sc} = frequency of subcarrier signal

ΔF_a = peak deviation of subcarrier produced by audio signal

B_a = noise bandwidth of audio baseband filter

P = preemphasis advantage

Digital BER

Digital systems use slightly different parameters to define the performance of link. A term E_b/N_o or energy per bit is used instead of C/N_o. They are related by the following formula:

$$E_b/N_o = C/N_o - 10 \log BR$$

where:

C/N_o = link carrier-to-noise ratio in dB-Hz

BR = bit rate being transmitted

The equations that quantify the performance of a digital link are complex. Instead of providing these equations, sample performance charts will be shown to indicate performance (Figures 1 and 2).

Notice, instead of S/N the digital systems use Bit Error Rate (BER) to define the performance of the system.

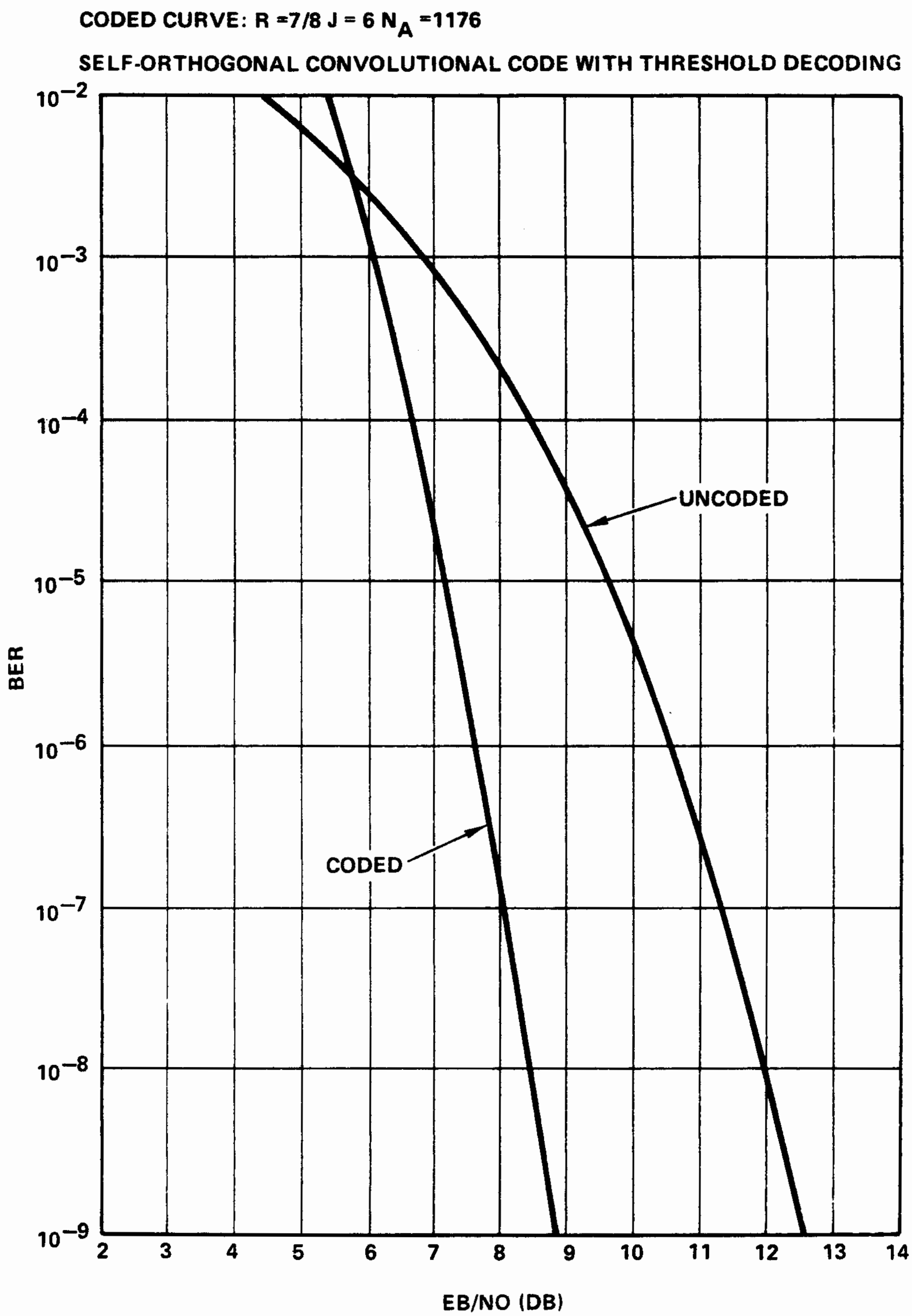


Figure 1. Theoretical BER Performance for BPSK/QPSK

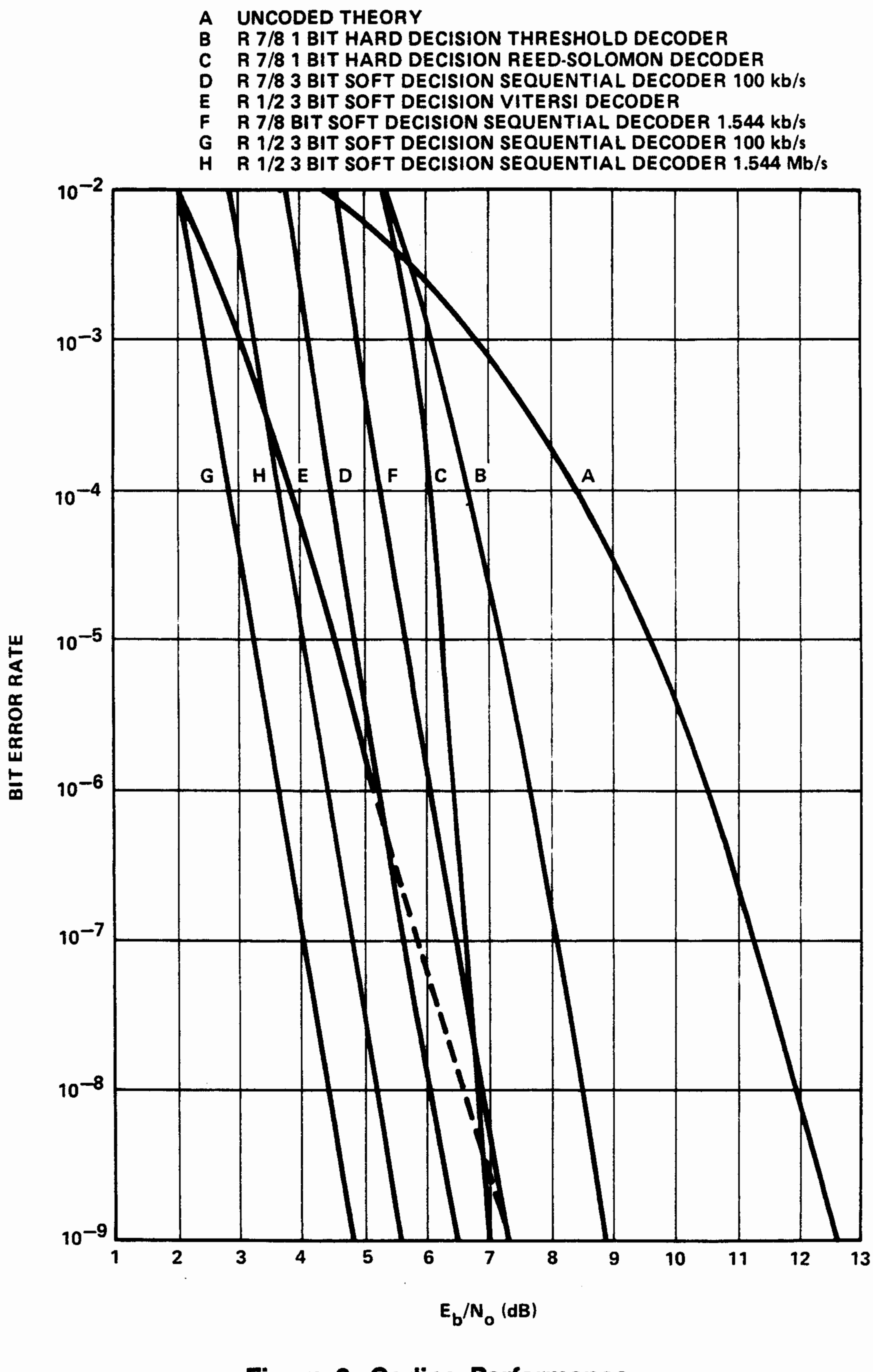


Figure 2. Coding Performance

Summary

Uplink

$$EIRP_e = P_o + G_t$$

$$S = EIRP_e - 162.5$$

$$N_o = k\ T_o\ B_n$$

$$\left(C/N_o\right)_u = S + G/T - 20 \log F(GHz) - 21.45 + 228.6 \quad (dB\text{-}Hz)$$

Downlink

$$\left(C/N_o\right)_D = G/T + EIRP_s - 228.6 - L_s - L_p$$

$$L_s = 20 \log \left(\frac{4\pi R}{\lambda}\right)$$

$$C/N = C/N_o - 10 \log B_n$$

Overall

$$\left(C/N_o\right)_o = \frac{1}{\frac{1}{\left(C/N_{o\ u}\right)} \oplus \frac{1}{\left(C/N_o\right)_D}} \quad \text{or} \quad \left(N_o/C\right)_o = \left(N_o/C\right)_u \oplus \left(N_o/C\right)_D$$

where:

$\oplus$ = denotes power addition

SCPC

$$S/N = C/N_o + 10 \log 3\left(\frac{\Delta F}{F_m}\right)^2 - 10 \log 2B_a + P + C$$

Video

$$S/N = C/N_o + 10 \log 12 + 20 \log \Delta F_s - 30 \log B_v + P + W$$

Message

$$S/N = C/N_o - 10 \log \left(2B_{ch}\right) + 20 \log \left(\frac{\Delta F_{tt}}{f_a}\right) + P + W$$

Digital

$$E_b/N_o = C/N_o - 10 \log BR$$

Glossary of Symbols

EIRP	-	Effective Isotropic Radiated Power
P_o	-	HPA output power in dBW
G	-	Transmit antenna gain in dBi
S	-	Flux density in dBW
N_o	-	Noise power density per unit bandwidth
k	-	Boltzmans constant = 228.6 dB-Hz or $1.38 * 10^{-23}$ Joule/K
T_o	-	Equivalent noise temperature in K
B_n	-	Filter noise bandwidth
C	-	Carrier power in watts
G/T	-	Earth Station Figure of Merit dB/K
L_s	-	Spreading loss
L_p	-	Polarization efficiency loss
R	-	Distance from earth station to satellite in meters
λ	-	Wavelength of signal
ΔF	-	Peak deviation (If video signal, this includes sync tips)
f_m	-	Highest baseband frequency
B_a	-	Audio filter noise bandwidth
P	-	Preemphasis advantage in dB
C	-	Companding advantage in dB
ΔF_s	-	Peak deviation produced by that part of the video signal defined to be the "signal"
B_v	-	Video filter noise bandwidth
W	-	Weighting advantage
B_{ch}	-	Message baseband channel bandwidth
ΔF_{tt}	-	Peak deviation due to test tone

Interference Analysis

James H. Cook. Jr.

Introduction

The consideration of interference in a satellite communications system is important, not only for being interfered with, but generating interference into existing systems. It is mandatory for a proposed transmit system in the United States to submit to the Federal Communications Commission (FCC) a coordination filing which includes an interference analysis. This analysis must show the impact of the proposed system on existing operational systems and must meet the allowable interference requirements of the FCC. Coordination for receive-only systems is not mandatory and is only necessary when the system desires interference protection from future transmitting systems.

To demonstrate the analysis procedure, a typical 7-meter broadcast FM/TV system is considered. The model is described below with the supporting calculations for the interference study.

Model for Interference Analysis

The interference analyses presented in the following sections are based on models for the antenna characteristics, the geostationary satellite systems, and the spectral characteristics for the desired and interfering carriers. In this section the key assumptions upon which these models are based are described and examined. The single-entry model for adjacent satellite interference is also discussed.

Antenna Characteristics

The primary characteristics of the antenna which affects the interference analysis is the angular discrimination, the gain differential between the on-axis peak to an off-axis angle for an interference source. For this analysis the radiation patterns of all earth station antennas are characterized by the standard envelope (32-25 log θ) dBi out to 48 degrees and -10 dBi from 48 degrees to 180 degrees. (See Introduction to Earth Station Antennas in this digest.) In actuality, the antenna radiation patterns fall below this envelope for most of the region between 1 and 180 degrees, so this is a limiting case assumption. Table 2-1 shows the angular discrimination calculations for various earth station antenna sizes. No angular discrimination is assumed for the satellite antennas in the adjacent satellite interference analysis.

Table 2-1. Angular Discrimination of Earth Station Antennas

Antenna Diameter (meters)	Transmit Gain (dBi)	Transmit Discrimination 4.2°	Transmit Discrimination 5.25°	Receive Gain (dBi)	Receive Discrimination 4.2°	Receive Discrimination 5.25°
3.06				39.5	23.1	25.5
4.6	46.3	29.9	32.3	43.5	27.1	29.5
5.0	47.4	31.0	33.4	44.5	28.1	30.5
7.0	49.4	33.0	35.4	47.5	31.1	33.5
10.0	53.5	37.1	39.5	50.9	34.5	36.9
13.0	56.0	39.6	42.0	52.0	35.6	38.0
30.0	63.3	46.9	49.3	60.3	44.5	46.9

Satellite Characteristics

The characteristics of the present geostationary communications satellites illuminating the U.S. are shown in Table 2-2. These are used as guidelines for analyses of the adjacent satellite interference effects.

Based on a stationkeeping accuracy of ±0.1, the worst-case orbital separation is 3.8 degrees. The resulting effective angular separation of the earth station for worst-case viewing angle for the 48 contiguous states is computed to be 4.2 degrees. (See Appendix A). In all calculations presented, an angular offset of ±4.2 degrees is used with the exception of the case of the Anik satellites, which have a 5-degree orbital separation and an effective 5.25-degree angular separation.

The analyses in the following sections assume the nominal values for EIRP shown in Table 2-2. For those cases which do not refer to specific satellite, an EIRP of 34 dBW is used.

Table 2-2. Interfering Satellite Parameters

Parameter	WESTAR	COMSTAR	SATCOM	ANIK
EIRP Range (dBW)	33-36	33-37	33-37	33-36
EIRP Nominal (dBW)	34	35	34	34
Transponder Saturation Flux Density (dBW/m^1)	-82	-74	-83	-82
Longitude (degrees)	I 99° II 123.5° III 91.0°	A. 128° B. 95° C. 87°	I 135° II 119° III*(132°)	AI 104° AII 109° AIII 114°

*Not presently in orbit.

Polarization Effects for Adjacent Satellite

The RCA satellite system utilizes a 20-degree polarization twist for a 24-transponder dual-polarized frequency plan. This results in a polarization discrimination effect for interference into and from other domestic satellite systems. The polarization discrimination angles between the RCA system and all other domestic systems are listed in Table 2-3.

The polarization twist produces an added isolation for interference signals from other satellites, in addition to that given by the angular discrimination in Table 2-1, to the extent that the transmit or receive polarization is maintained in the sidelobes for the antennas in the satellite system.

Table 2-3. Polarization Discrimination Angles

	WESTAR & ANIK		COMSTAR			
RCA	up	down	up even	up odd	down even	down odd
Up even	20°	70°	70°	20°	20°	70°
Up odd	70°	20°	20°	70°	70°	20°
Down even	70°	20°	70°	20°	20°	70°
Down odd	20°	70°	20°	70°	70°	20°

For an ideal antenna which maintains the desired polarization through the sidelobe region, this provides an isolation of 9.3 dB (20 log cos 70°) for transponders with a 70-degree relative polarization, or 0.54 dB (20 log cos 20°) for those with a 20-degree relative polarization.

In reality, the cross-polarization isolation is probably less than 10 dB in the close-in sidelobe region for adjacent satellite signals. The isolation factors computed above for the 20 twist angle are thus modified by the imperfect polarization of the sidelobes. Since for the 20-degree factor, the effect of the twist angle is relatively insignificant even in the ideal case, this effect is not included in the analyses which follow. However, it may be possible to reduce a potentially disruptive interference carrier from an adjacent satellite by selecting a 70-degree twist transponder for the service. This may yield 4- to 6-dB additional isolation depending on the antenna performance.

Spacecraft Polarization. The spacecraft antenna has a minimum cross-polarization isolation of 36 dB. For these calculations, it is assumed that the satellite transponder of interest transmits and receives only signal components of the desired polarization.

Cross-Polarization Interference. The interference signals from the cross-polarized transponders must also be considered in this analysis. Table 2-4 presents a summary of the polarization isolation factors for uplink and downlink. Since the characteristics of the cross-polarized transponders (uplink saturation flux density and downlink EIRP) are considered identical for this analysis, the uplink and downlink factors can be combined into a net isolation factors, as shown.

Cross-polarized transponders are centered on 20-MHz offset frequencies, so that the center frequency for one polarization falls between those of the adjacent cross-polarized transponders. The isolation provided by the carrier offset combined with the spectral characteristics (produced by modulation) must be used to determine the interference signal level for this source.

Table 2-4. Internal Satellite Cross-Polarization Isolation Summary

Factor	Isolation (dB)
UPLINK	
E.S. Antenna, 7 meter; 0.13° rms pointing for 45 mph winds gusting to 60 mph plus 0.1° satellite station keeping error = 0.23° pointing error	30.0
Rain*	31.0
Faraday Rotation*	35.0
Satellite Antenna	36.0
Net Uplink Isolation**	26.3
DOWNLINK	
E.S. Antenna (0.23° rms pointing)	30.0
Rain*	34.0
Faraday Rotation*	27.0
Satellite Antenna	36.0
Net Downlink Isolation**	24.4
Net Link Isolation**	22.2

* Degradation not exceeded 99.9% of the time.
** Based on power summation.

Single-Entry Adjacent Satellite Model

The computation of interference from other satellites is accomplished in this analysis by the single-entry method, to account for the effects associated with signals from the two adjacent satellites and the two semi-adjacent satellites. For this method, the single-entry interference is computed for an adjacent satellite at 3.8° (orbital arc) from the desired satellite. Then it is assumed, as a limiting or worst-case situation, that the same interference is produced from each of the other interfering satellites. These are combined using power summation and the angular discrimination (32-25 log θ); the net result is an interference level which is 4 dB higher than the single-entry level.

The equations to calculate the adjacent satellite interference are given below:

$$(C/I)_u = (EIRP)_{ES} - \sum_{i=1}^{N} \oplus \left\{ (EIRP)_i - (G_i - G(\Theta_i)) + F_i + P_i \right\} \text{ dB} \quad (1)$$

where:

$\Sigma \oplus$ = power summation

$(EIRP)_{ES}$ = Earth station radiated power in dBW

$(EIRP)_i$ = Effective radiated power of interfering earth station (dBW)

G_i = Peak gain of interfering earth station (dBi)

$G(\Theta)_i$ = Gain of interfering earth station in direction $(\Theta)_i$ (dBi)

F_i = Frequency discrimination factor for i^{th} earth station

P_i = Polarization discrimination factor for i^{th} earth station

$$(C/I)_D = (EIRP)_{SAT} + G_{ES} - \sum_{i=1}^{N} \oplus \left\{ (EIRP)_i + G_{ES}(\Theta_i) + F_i + P_i \right\} \quad (2)$$

where:

$(EIRP)_{SAT}$ = Effective radiated power of satellite in the direction of receive earth station in dBW

G_{ES} = Gain of the receive earth station, dBi

$G_{ES}(\Theta_i)$ = Gain of the receive earth station in the direction Θ_i

F_i = Frequency discrimination factor

P_i = Polarization discrimination factor

$$(C/I)_{ADJ.SAT} = (C/I)_U \oplus (C/I)_D \qquad (3)$$

If the single-entry calculation is done for the worst-case interfering signal, the results yield a limiting value for the adjacent satellite interference.

Interference Analysis for FM/TV Service

System Parameters

The interference analysis for the FM/TV service is based on the following parameters:

Parameter	Specification
Transponder EIRP	34 dBW (saturated)
Antenna Size	
Transmit	7-meter
Receive	3-meter, 4.6-meter, 5-meter, or 10-meter
Uplink EIRP	80 dBW
Transmit Power	4.5 kW

Interference into the Proposed System

Interference into the proposed system can originate from the following sources:

- Adjacent satellite signals
- Internal cross-polarized signals
- Terrestrial microwave signals

These are analyzed separately in the following paragraphs and then combined to determine the total interference into the system. The results can then be incorporated in the link analysis.

Adjacent Channel Interference

The worst-case interference from an adjacent satellite occurs when all the transponder energy is concentrated within the desired signal bandwidth. The isolation is then produced solely by the angular discimination of the transmit (for uplink) or receive (for downlink) antenna. Thus for a 3.8-degree orbital separation and a 10-meter antenna, the single entry carrier-to-interference (C/I) is computed as follows:

Uplink:	W_w, flux density wanted signal		-83 dBW/m^2
	W_i, flux density interfering signal		-82 dBW/m^2
	ΔG, 10-meter		37.1 dB
	C/I up		36.1 dB
Downlink:	E_w, EIRP wanted signal		32 dBW
	E_i, EIRP Interfering signal		34 dBW
	ΔG, 7-meter		31.1 dB
	C/I down		29.1 dB
Total		$C/I_{up} \oplus C/I_{down}$	28.3 dB

$\oplus$ Denotes power summation.

This analysis ignores polarization effects, which would improve the downlink contribution somewhat depending on the relative polarization. The net adjacent satellite interference is then 24.3 dB, according to the single-entry method described above.

Adjacent Satellites

Consider a typical frequency reuse system with 24 transponder channels. If every channel carries traffic, for any one channel, the other 23 channels act as interference. However, only the four nearest channels in frequency need to be considered. The four channels consist of two co-polarized adjacent channels and two cross-polarized adjacent channels. It is assumed that all the interferences are incoherent. The amount of interfernce can be computed by convolving the power spectra of the wanted and unwanted signals.

Several types of signals may occupy the adjacent channels such as HBR data, SCPC, FDM-FM, and FM/TV. It can be shown that the adjacent co-polarized channels have negligible effect compared to the two adjacent cross-polarized channels (see Appendix B). The interfering power from these four types of service is shown below:

Signal	Polarization	Interfering Power
FM/TV	Co-	negligible
HBR data	Co-	negligible
SCPC	Cross-	-39 dB below saturation
FDM-FM	Cross-	-6.45 dB below saturation

The interfering power from the cross-polarized SCPC channel assumes that the carriers in the 14-MHz shared-frequency band are directly interfering with the FM/TV channel. The SCPC carriers are assumed to be equally spaced, and the effects of intermodulation (intermod) are neglected, since the level of the intermod power spectrum is many dB below the power spectrum associated with the SCPC carriers.

The C/I due to the adjacent channels can be calculated by comparing the wanted to interfering powers. The cross-polarized FDM/FM, carrier-to-interference ratio for the uplink is

$$\left(\frac{C}{I}\right)_u = 6.45 + (XPI)_u$$
$$= 6.45 + 26.3$$
$$= 32.75 \text{ dB}$$

The cross-polarized FDM/FM, carrier-to-interference ratio for the downlink is:

$$\left(\frac{C}{I}\right)_D = 6.45 + (XPI)_D$$
$$= 6.45 + 24.4$$
$$= 30.85 \text{ dB}$$

The cross polarized SCPC, carrier-to-interference ratio for the uplink is:

$$\left(\frac{C}{I}\right)_u = 3.9 + (XPI)_u + \text{Backoff}$$

Assuming a 4-dB backoff, the $(C/I)_u$ is:

$$\left(\frac{C}{I}\right)_u = 7.9 + 26.3$$
$$= 34.2 \text{ dB}$$

The cross-polarized SCPC carrier-to-interference ratio for the downlink is:

$$\left(\frac{C}{I}\right)_D = 7.9 + (XPI)_D$$
$$= 7.9 + 24.2$$
$$= 32.1 \text{ dB}$$

The total adjacent satellite carrier-to-interference ratio is given by:

$$\left(\frac{C}{I}\right)_u = \left(\frac{C}{I}\right)_{u \text{ FDM-FM}} \oplus \left(\frac{C}{I}\right)_{u \text{ SCPC}}$$
$$= 30.4 \text{ dB}$$

$$\left(\frac{C}{I}\right)_D = \left(\frac{C}{I}\right)_{D \text{ FDM-FM}} \oplus \left(\frac{C}{I}\right)_{D \text{ SCPC}}$$
$$= 28.4 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{S \text{ ADJ. SAT.}} = \left(\frac{C}{I}\right)_u \oplus \left(\frac{C}{I}\right)_D$$
$$= 26.3 \text{ dB}$$

Terrestrial Interference

Terrestrial microwave carriers are centered on frequencies offset by 10 MHz from the satellite carriers. To analyze the effect of terrestrial carriers on the FM/TV system, it is necessary to determine the power level of the interfering signal and the spillover of terrestrial carrier spectra into the passband of the receiver. The first factor involves site details, such as angular discrimination and distance to the interfering transmitter. The second factor can be computed from the spectral distribution projected for the terrestrail carrier and the filter characteristic of the receiver. For the purpose of this analysis, it is assumed that the C/I due to terrestrial microwave is 25 dB minimum.

Summary

A summary of the contributions to interference into the FM/TV system is presented in Table 3-1. The net carrier-to-interference ratio for this service is computed from the power sum of the interfering signals.

Table 3-1. Summary of Interference into FM/TV System, in Terms of Carrier-to-Interference Ratio

Adjacent Satellite	24.3 dB
Adjacent Transponder	26.3 dB
Terrestrial	25.0 dB
Net C/I, system*	20.3 dB

*Power summation.

Interference Into Other Services

This section presents calculations for interference produced by the 7-meter FM/TV transmit system into other services. The uplink EIRP is well within the coordination limits for the transmit stations.

The following analysis of the interference into adjacent satellite services is based on a satellite EIRP of 34 dBW. The EIRP, space loss, and other miscellaneous losses are assumed to be identical for adjacent systems so that the C/I is dependent only on relative EIRP, angular discrimination, and bandwidth factors. Although there can be potential additional discrimination associated with polarization differences between the adjacent systems, this has been omitted to obtain a worst-case analysis.

Interference into FM-FDM Service

Table 4-1 gives the analysis of interference of the indicated FM/TV system into a saturated carrier FM-FDM service. The analysis yields a worst-case C/I of 29.3 dB, resulting in a negligible effect to the FM/FDM service. Both the uplink interfering antenna and the downlink receiving antenna are assumed to be 7-meter diameter for the worst-case. An improvement of approximately 2.4 dB occurs when the downlink receiving antennas are 13 meters in diameter.

Table 4-2 gives an analysis of interference into a multi-carrier FM-FDM service. The input and output backoffs are assumed to be 11 dB and 6 dB, respectively. These levels represent typical worst-case backoffs resulting in the lowest energy densities expected for this type system. The nominal C/I into multi-carrier FM-FDM systems is calculated based on full-transponder utilization for both desired and interfering carriers.

Table 4-1. Interference into a Saturated Carrier FM-FDM Service

DOWNLINK

	Satellite System					
	WESTAR		COMSTAR		ANIK	
	(13m)	(7m)	(30m)	(7m)	(30m)	(7m)
EIRP, wanted signal (dBW)	34.0	34.0	34.0		34.0	
EIRP, interfering signal (dBW)	34.0	34.0	34.0		34.0	
ΔG, antenna (dB)	35.6	31.1	44.5	31.1	46.9	31.1
$(C/I)_D$	35.6	31.1	44.5	31.1	46.9	31.1

UPLINK

Flux Density, wanted signal(dBW/m^2)	-82.0		-74.0		-82.0	
Flux Density, interfering signal(dBW/m^2)	-83.0		-83.0		-83.0	
ΔG, antenna (dB)	33.0		33.0		33.0	
$(C/I)_u$	34.0	34.0	42.0	42.0	34.0	34.0
System (C/I), dB*	31.7	29.3	40.0	30.8	33.8	29.3

*Power summation.

Table 4-2. Interference into Multi-Carrier FM-FDM Service

DOWNLINK

	Satellite System					
	WESTAR		COMSTAR		ANIK	
Receive Antenna Diameter (m)	(13)	(7)	(30)	(7)	(30)	(7)
EIRP, wanted signal (dBW)	28.0	28.0	28.0	28.0	28.0	28.0
EIRP, interfering signal (dBW)	34.0	34.0	34.0	34.0	34.0	34.0
ΔG, dB	35.6	31.1	44.5	31.1	46.9	33.5
$(C/I)_D$, (dB)	29.6	25.1	38.5	25.1	40.9	27.5

UPLINK

	WESTAR		COMSTAR		ANIK	
Flux Density, wanted signal(dBW/m^2)	-93.0		-85.0		-93.0	
Flux Density, interfering signal(dBW/m^2)	-83.0		-83.0		-83.0	
ΔG	33.0		33.0		35.4	
$(C/I)_u$, (dB)	23.0	23.0	31.0	31.0	25.4	25.4
System C/I, (dB)*	22.1	20.9	30.3	24.1	25.3	23.3

*Power summation.

Interference into FM/TV Service

Table 4-3 presents the analysis of interference into a saturated full-transponder FM/TV service. This results in a worst-case C/I of 22.2 dB for the 3-meter receive antenna, which is in excess of the FCC objective of 22 dB for these services. For the 4.6-meter and larger receive antenna the C/I is greater than 25 dB.

Table 4-3. Interference into a FM/TV Service

	Receive Antenna Diameter				
DOWNLINK	3.0	4.6	5.0	7.0	10.0
EIRP, wanted signal	34.0	34.0	34.0	34.0	34.0
EIRP, interfering signal	34.0	34.0	34.0	34.0	34.0
ΔG, (dB)	23.1	27.1	28.1	31.1	34.5
$(C/I)_D$	22.5	25.6	26.3	28.0	29.4
UPLINK					
Flux Density, wanted signal (dBW/m^2)			-82.0		
Flux Density, intefering signal (dBW/m^2)			-83.0		
ΔG, (dB)			-33.0		
$(C/I)_u$			34.0		
System C/I (dB)*	22.2	25.0	25.6	26.1	28.1

*Power Summation.

Interference into SCPC System

Many possible types of multi-carrier SCPC systems exist. In order to access the impact of the FM/TV system on these systems, the number of carriers is varied from 20 to 1200. The parameters of these systems are listed below in Table 4-4.

Table 4-4. Parameters of SCPC Systems

Uplink to Westar Number of Carriers	20	40	800	1200
Parameter				
Noise Bandwidth (kHz)	250	250	35	22.5
Backoff (dB)	4	4	7	7
EIRP (dBW)	60	57	51*	49*
Satellite Gain (dBi)	26	26	26	26
Satellite EIRP, per carrier (dBW)	17	13	+2	0

*Assumes 40% occupancy.

Parameter	WESTAR	SATCOM II	ANIK
Transmit Antenna	13M	7M	30M
EIRP, total power (dBW)	83.0	80.0	84.0
Transmit Antenna Gain (dBi)	54.3	49.4	63.3
Satellite Gain (dBi)	26.0	26.5-39.5	26.0
Satellite EIRP (dBW)	34.0	34.0	36.0

The interference is calculated in Table 4-5 through 4-7 for three cases, assuming a 5-meter or 13-meter receive antenna with the WESTAR satellite, and a 30-meter receive antenna with the ANIK satellite.

Table 4-5. Interference into SCPC Systems; 13M

DOWNLINK	WESTAR (13M)			
Number of Carriers	20	40	800	1200
EIRP, wanted signal, (dBW)	17.0	13.0	2.0	0.0
EIRP, interfering signal, (dBW)	14.2	14.2	5.6	3.7
ΔG, (dB)	35.6	35.6	35.6	35.6
$(C/I)_D$, (dB)	38.4	34.4	32.0	31.9
UPLINK				
Flux Density, wanted signal (dBW/m^2)	-102.0	-105.0	-114.0	-116.0
Flux Density, interfering signal (dBW/m^2)	-112.8	-112.8	-111.4	-113.3
ΔG, (dB)	33.0	33.0	33.0	33.0
$(C/I)_u$, (dB)	43.8	40.8	30.4	30.3
System C/I, (dB)*	37.3	33.5	28.1	28.0

*Power summation.

Table 4-6. Interference into SCPC Systems; 5M

DOWNLINK	WESTAR (5M)			
Number of Carriers	20	40	800	1200
EIRP, wanted signal, (dBW)	17.0	13.0	2.0	0.0
EIRP, interfering signal, (dBW)	14.2	14.2	5.6	3.7
ΔG, (dB)	28.1	28.1	28.1	28.1
$(C/I)_D$, (dB)	30.9	26.9	24.5	24.4
UPLINK				
Flux Density, wanted signal, (dBW/m^2)	-102.0	-105.0	-114.0	-116.0
Flux Density, interfering singal (dBW/m^2)	-112.8	-112.8	-111.4	-113.3
ΔG, (dB)	33.0	33.0	33.0	33.0
$(C/I)_u$, (dB)	43.8	40.8	30.4	30.3
System C/I, (dB)*	30.7	26.7	23.5	23.4

*Power summation.

Table 4-7. Interference into SCPC System; 30M

DOWNLINK	ANIK (30M)			
Number of Carriers	20	40	800	1200
EIRP, wanted signal, (dBW)	17.0	13.0	2.0	0.0
EIRP, interfering signal, (dBW)	14.2	14.2	5.6	3.7
ΔG, (dB)	46.9	46.9	46.9	46.9
$(C/I)_D$, (dB)	49.7	45.7	43.3	43.2
UPLINK				
Flux Density, wanted signal, (dBW/m^2)	-102.0	-105.0	-114.0	-116.0
Flux Density, interfering signal (dBW/m^2)	-112.8	-112.8	-111.4	-113.3
ΔG, (dB)	33.0	33.0	33.0	33.0
$(C/I)_u$, (dB)	43.8	40.8	30.4	30.3
System C/I, (dB)*	42.8	39.6	30.1	30.0

*Power summation

As can be seen from the analysis, the worst-case C/I was 24.3 dB for the 1200-channel, 5-meter receive system. This level of interference presents no problem to the system since operating C/Ns of 10 to 12 dB are typical.

Summary

Interference into other systems from a 7-meter FM/TV broadcast system has been analyzed in the preceding paragraphs. The results obtained from these caluculations indicate that the proposed system is compatible with the postulated existing services in the 4- and 6-GHz bands, to the extent that the C/I contributions can be treated as additive link noise. The worst-case C/I values for each of the services considered is listed in the Table 4.8 below:

Table 4-8. Summary of Interference from a 7-Meter FM/TV Broadcast Service into Postulated Existing Services

Service	Worst-Case C/I (dB)
FM-FDM	29.3
Multi-Carrier FM-FDM	20.9
FM/TV, 3M receive	22.2
4.6M receive	25.0
5.0M receive	25.6
7.0M receive	26.1
10.0M receive	28.1
SCPC, 1200 channels, 5M receive	23.4
13M receive	28.0
30M receive	30.0

Comments on New Reduced Satellite Spacing

The FCC in August of 1983 has finalized the new satellite orbital assignments based on a frequency and polarization plan to allow satellite spacing to be reduced from the present 4° to 2° with an average spacing of 2.5° at C-band. The implementation of this plan depends on several important technical achievements including:

a. Sidelobe reduction of transmit earth station antennas from present (32-25 log Θ) dBi envelope to (29-25 log Θ) dBi for $1° < \Theta < 7°$.

b. All frequency reuse satellites.

c. Adjacent satellite, same frequency transponders orthogonally polarized.

d. Homogeneity of satellite EIRP and saturation flux density characteristics for minimum spacing.

e. Implementation dates dependent upon life of present and next generation satellites.

The rationale of this document follows this scenario. The reduction of close-in sidelobes by 3 dB will allow C/I to remain constant when reducing the satellite spacing from 4° to 3° without any other changes. To reduce from 3°, a change in operating characteristics other than the earth station antenna patterns must be controlled. To allow further reduction, polarization and frequency must be controlled such that additional suppression of interference be achieved through cross-polarization discrimination characteristics. This will allow acceptable carrier-to-interference for spacings approximately 2.5°, and with the desired homogeneity of satellite characteristics acceptable C/I for spacings of 2°.

Since the FCC has implemented a plan based on these conditions, the interference analysis becomes more complicated in that the single-entry model cannot be used for the calculations. Instead, the calculations must use equations (1). (2) and (3).

Antenna Characteristics

The primary characteristic of the antenna which affects the interference analysis is the angular discrimination, the gain differential between the on-axis gain, and the gain of an off-axis angle for an interfering source. For this analysis, the copolarized radiation patterns of the assumed earth station antennas are characterized by the standard envelope 32-25 log Θ or the proposed envelope 29-25 log Θ. The cross-polarized radiation patterns are characterized by 22-25 log Θ for the standard envelope and 19-25 log Θ for the proposed envelope. In actuality, the antenna radiation sidelobe may fall below or above these reference envelopes by some predetermined acceptable level. Any sidelobes which are below the reference envelope and at the appropriate pointing angles of adjacent satellites would reduce the interference and, conversely, any sidelobes above the envelope pointing at adjacent satellites would increase the interference.

The cross-polarization discrimination of 10 dB is assumed to apply for clear weather conditions. During periods of rain, the depolarization of the incoming signal may reduce this number to 0 dB.

Satellite Characteristics

The analyses in these comments are based on a satellite deployment model with co-frequency transponders on adjacent satellites being cross-polarized with each other. This model does not exist today and cannot exist for a period of years. Nevertheless, the calculations are performed with this model to demonstrate the expected result some years in the future. Three cases of this model have been examined:

1. A homogeneous model in which interfering and desired satellites have the same saturation flux density and radiated EIRP. (The radiation patterns yield the same signal strength at any given location on the ground.)

2. A model in which the interfering satellite EIRP exceeds the desired satellite EIRP by 2 dB.

3. A model in which the interfering satellite EIRP exceeds the desired satellite EIRP by 4 dB.

EIRP is a very important consideration. Even at the time of launch, antenna and transponder characteristics of satellites are such that their initial EIRP contours on the earth's surface are not identical. Differences in the initial EIRP contours and differences in transponder aging must be considered in a practical system. An orbital spacing plan that is predicated on differntial EIRPs of less than 2 dB represents an impractical burden, both on the satellite manufacturers and on the FCC in assuring compliance with a more stringent specification. It is suggested that the calculations for the first case (equal EIRPs) not be taken as representing a practical case.

The calculations do not include station-keeping inaccuracies and are based on geosynchronous rather than topocentric angles. An average topocentric angle for the continental United States (CONUS) can be estimated by multiplying the geocentric angle by 1.08.

Existing Satellite Parameters. In order to consider the reasonableness of a homogeneous satellite model, a compilation of EIRPs at various locations within CONUS was made (Table 1) for the existing C-Band satellites. In the eastern section of the orbital arc, 91°W to 99°W, the average EIRP difference between WESTAR I, WESTAR II, COMSTAR D1, and COMSTAR D2 at the ten sample sites is 0.94 dB with a maximum EIRP difference of 2.5 dB. In the western section of the orbital arc, 119°W to 135°W, the average EIRP difference between COMSTAR D4, WESTAR II, RCA F1, and RCA F2 is 3 dB with a maximum EIRP difference of 4.3 dB. The EIRPs used in this comparison are published values and do not take into consideration any decrease in performance due to aging.

Table 1. EIRP (dBW) for Selected Sites in CONUS for Various Satellites

Site	WESTAR III 91°W 4160H	COMSTAR D-1 95°W 3920W	COMSTAR D-2 95°W 3940H	WESTAR I 99°W 4160H	RCA F-2 119°W 3720H	WESTAR II 123.5°W 4160H	COMSTAR D-4 128°W 3980H	RCA F-1 135°W 3720V
Atlanta, GA	35.5	34.8	34.8	35.4	32.3	34.9	35.6	33.2
Bangor, MA	33.1	33.7	34.2	33.5	30.8	34.2	35.1	33.8
Brownsville, TX	31.5	34.0	33.9	32.2	30.7	32.9	34.1	30.2
Denver, CO	35.5	35.7	35.6	35.5	34.5	34.9	36.3	34.2
Kansas City, MO	35.5	35.8	36.0	34.5	34.1	34.9	36.3	34.2
Los Angeles, CA	34.1	34.6	34.3	34.0	33.1	34.9	35.7	31.8
Miami, FL	33.0	33.2	33.0	33.7	28.5	31.9	31.8	29.4
Minneapolis, MN	34.6	35.0	35.3	34.1	33.7	34.5	34.9	34.2
New York, NY	34.8	34.5	34.5	34.9	31.4	34.7	35.7	33.7
Seattle, WA	35.1	34.5	34.5	34.6	34.6	33.5	34.7	33.5

Interference into the New System

Interference into the new system can originate from the following sources:

- Adjacent satellite signals
- Internal cross-polarization signals
- Terrestrial microwave signals

These are analyzed separately in the following paragraphs and then combined to determine the total interference into the system.

Adjacent Satellite Interference. Interference from adjacent satellites occurs in two ways: uplink interference from earth stations transmitting to adjacent satellites and downlink interference from adjacent satellite transmission into the desired earth station. The interference in both the uplink and downlink consists of many signals, but is primarily caused by the co-frequency channels/or transponders and the two 20-MHz offset-frequency channels in a frequency reuse system. For the system propsed by the Commission, the primary interferers are:

a. The co-frequency, cross-polarized channel on the first adjacent satellite on each side.

b. The two 20-MHz offset-frequency, co-polarized channels on the first adjacent satellite on each side.

c. The co-frequency, co-polarized channel on the second adjacent satellite on each side.

d. The two 20-MHz offset-frequency, cross-polarized channels on the second adjacent stallite on each side.

The contribution to interference of satellites at positions greater than four degrees from the desired satellite tends to be somewhat noise-like in that it is the result of a number of small relatively non-coherent signals.

The equations for calculation of the adjacent satellite interference are given below:

$$(C/I)_u = (EIRP)_{ES} - \sum_{i=1}^{N} \oplus \left| (EIRP)_i - (G_i - G(\Theta_i)) + F_i + P_i \right| \quad dB$$

where:

$\Sigma \oplus$ = Series power summation

$(EIRP)_{ES}$ = Earth station radiated power in dBW

$(EIRP)_i$ = Effective radiated power of interfering earth station (dBW)

G_i = Peak gain of interfering earth station (dBi)

$G(\Theta)_i$ = Gain of interfering earth station in direction $(\Theta)_i$ (dBi)

F_i = Frequency discrimination factor for i^{th} earth station

P_i = Polarization discrimination factor for i^{th} earth station

$$(C/I)_D = (EIRP)_{SAT} + G_{ES} - \sum_{i=1}^{N} \oplus \left| (EIRP)_i + G_{ES}(\Theta_i) + F_i + P_i \right|$$

where:

$(EIRP)_{SAT}$ = Effective radiated power of satellite in the direction of receive earth station in dBW

G_{ES} = Gain of the receive earth station, dBi

$G_{ES}(\Theta_i)$ = Gain of the receive earth station in the direction Θ_i

F_i = Frequency discrimination factor

P_i = Polarization discrimination factor

$$(C/I)_{ADJ \cdot SAT} = (C/I)_U \oplus (C/I)_D$$

where:

$\oplus$ = Power summation

The polarization discrimination factor in the above equations is the system discrimination rather than that of the receive or transmit antenna alone. A well-designed dual linearly polarized antenna can achieve excellent cross-polarization discrimination on or near the main beam axis (greater than 30 dB relative to the copolarized energy) and reasonable rejection of the cross-polarized signals in the close-in sidelobe regions. The adjacent satellite signals are received through the sidelobes of the earth station antenna; therefore, the proposed 19-25 log Θ envelope is assumed in the analysis. This assumption, rather than being conservative, is optimistic when one considers the interactions of the ionosphere and atmosphere on the transmitted and received signals, the purity of the initial transmitted signals (from the earth station and/or the satellites), and the polarization angle alignment between satellites. A more conservative and, probably, more realistic analysis would assume the same levels of copolarized and cross-polarized energy in the off-axis regions.

The frequency discrimination factor is related to the spectra of the desired and undesired signals. This factor can range from 10 to 0 dB dependent upon the interfering power from different services. For example, the F_i term where an AM-FM/TV signal is interfering with an FM/TV signal occupying the same bandwidth, would be 0 dB. For a 20-MHz offset-frequency FM/TV signal 30 MHz bandwidth) interfering with an FM/TV signal, the F_i term would range between 3 and 10 dB depending on the characteristics of the video signals. It is suggested that for typical FM/TV signals a value of 6.5 dB is more realistic than the 8.0 dB assumed in the FCC analysis.

Internal Interference. The internal interference in a satellite system is primarily due to the two adjacent 20-MHz offset-frequency, cross-polarized channels. The interfering power from different services has been calculated by convolving the power spectra of the individual services and is given in Appendix B.

Terrestrial Interference. Terrestrial microwave carriers are centered on frequencies offset by 10-MHz from the satellite carriers. To analyze the effect of terrestrial carriers on the FM/TV system, it is necessary to determine the power level of the interfering signal and the spillover of the terrestrial carrier spectra into the passband of the receiver. The first factor involves site details, such as angular discrimination and distance to the interfering transmitter. The second factor can be computed from the spectral distribution projected for the terrestrial carrier and the filter characteristic of the receiver. For the purpose of this analysis, it is assumed that the C/I due to terrestrial microwave is 25 dB.

Interference Analysis for FM/TV Service. The interference for the FM/TV service is based on the following parameters:

Parameter	Specification
Transponder EIRP	34 dBW (saturated)
Antenna Size	
Transmit	10 meter
Receive	3 meter, 4.6 meter, 5 meter, 7 meter, or 10 meter
Uplink EIRP	80 dBW
Transmit Power	4.5 kW

Each of the antennas listed above is presently used in FM/TV systems. Many are licensed and regulated and therefore "protected" from interference in certain respects. Many receive-only stations are unlicensed and not protected. In the new satellite environment, the antennas of unlicensed stations which serve many motels, churches, apartment complexes, CATV and MATV systems, will receive additional interference although it may not result in unacceptable operation.

The results of the analysis are presented in Figures 1 through 8; two figures are included for each antenna. All assume a polarization discrimination factor of -10 dB and frequency discrimination of 6.5 dB; one is for the proposed sidelobe envelope and the other for the existing sidelobe envelope. The frequency discrimination factor was assumed to be 6.5 dB. Each figure includes three cases of desired signal EIRP relative to interfering signal EIRP, and the effect of variable terrestrial interference is shown.

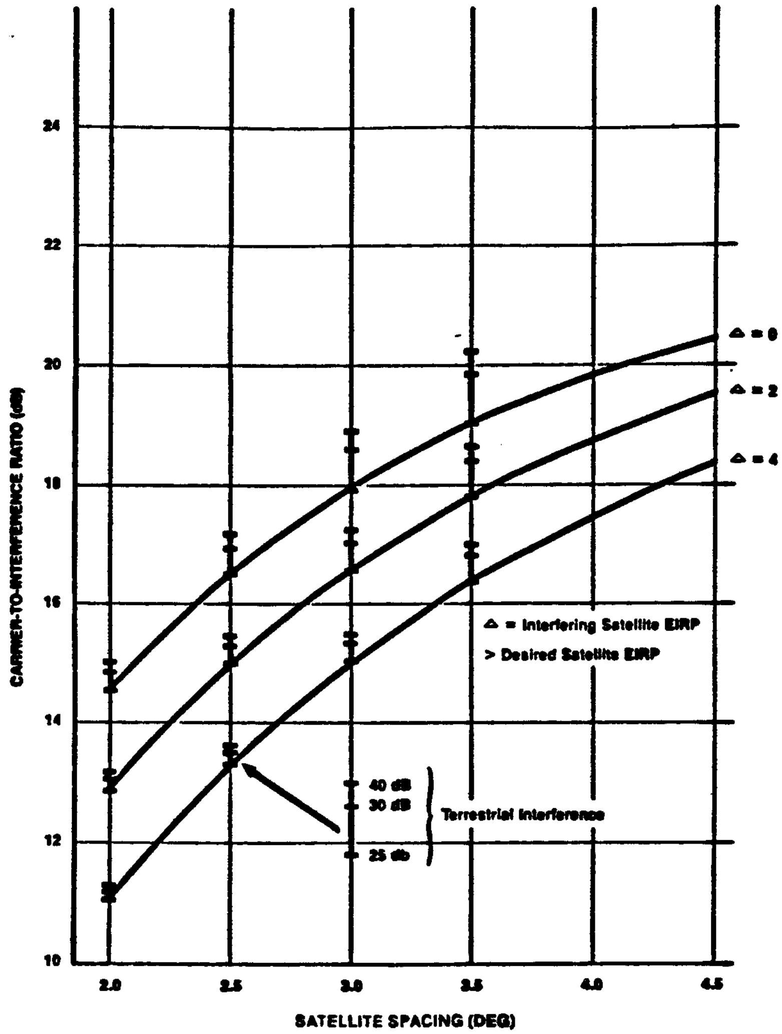


Figure 1. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink and a 3-Meter Downlink. FX = 6.5 dB PX = 10 dB. Proposed 29-25 Log Θ Envelope and Adjacent Satellite Polarization Interleaving.

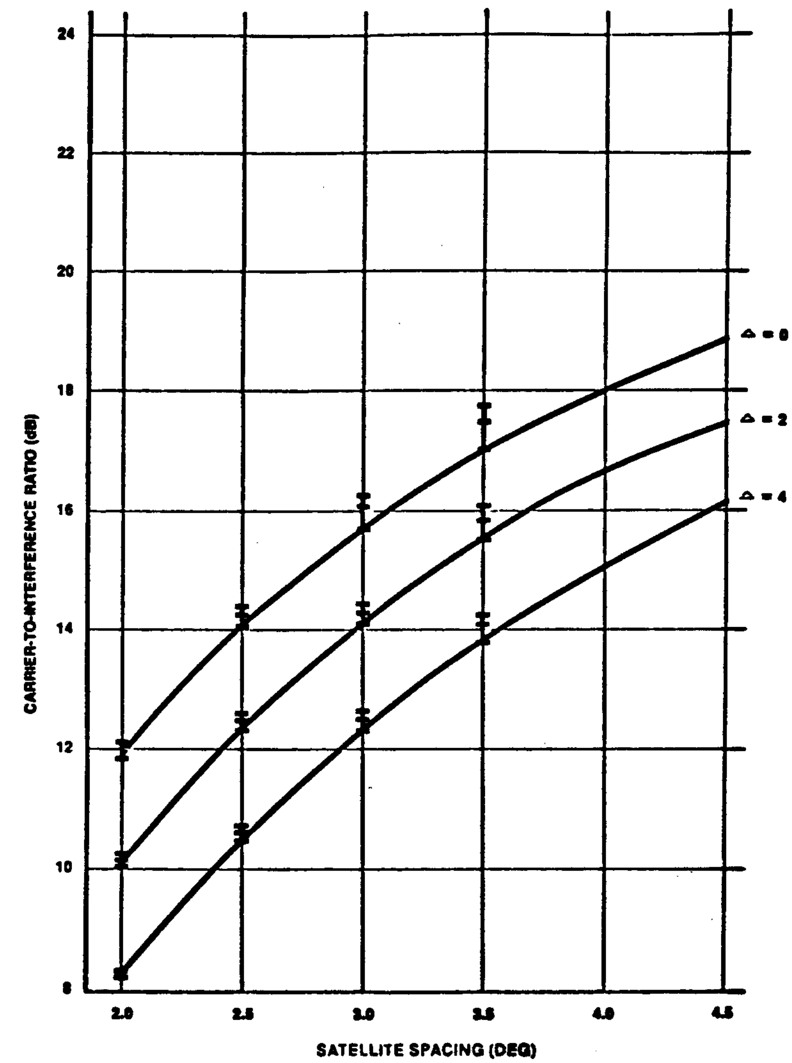


Figure 2. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink an 3-Meter Downlink. FX — 6.5 dB PX = 10 dB. 32-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

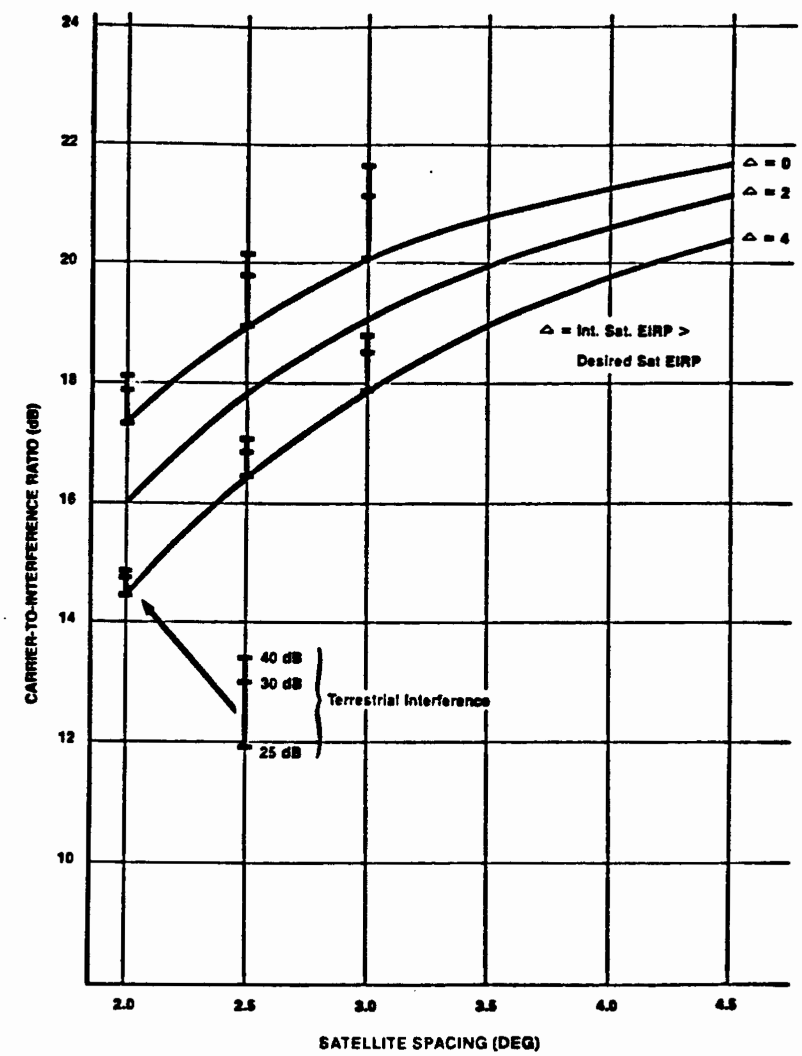


Figure 3. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink and a 4.5-Meter Downlink. FX = 6.5 dB PX = 10 dB. Proposed 29-25 Log Θ Envelope and Adjacent Satellite Polarization Interleaving.

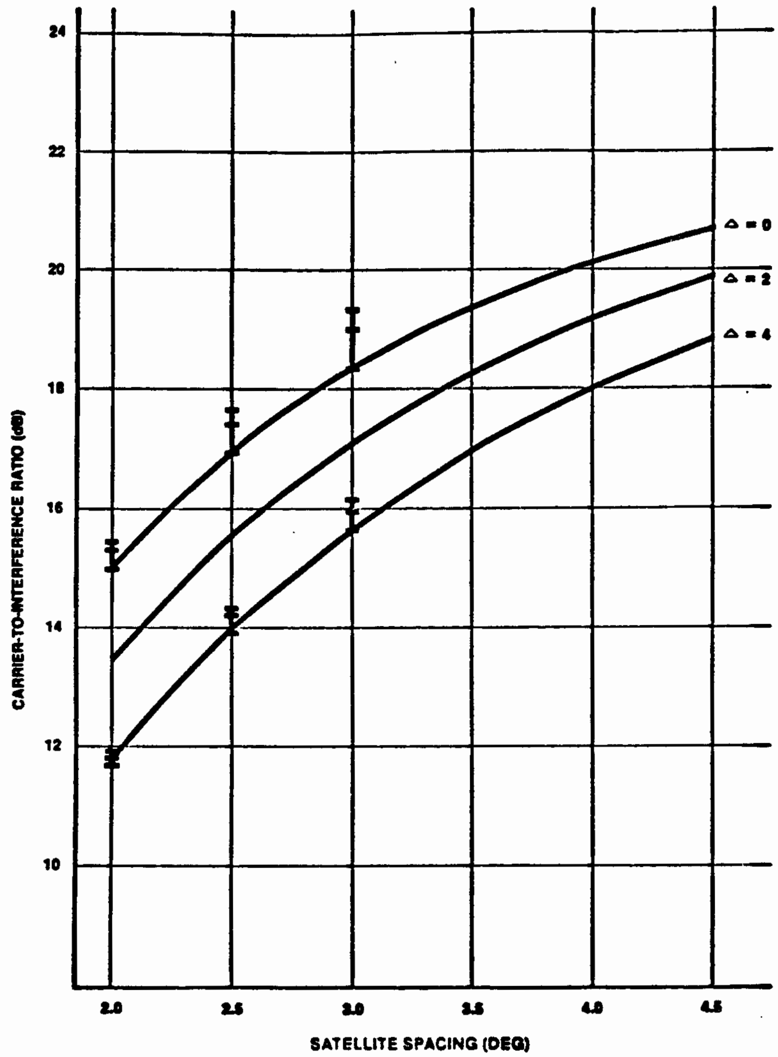


Figure 4. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink and a 4.5-Meter Downlink. FX – 6.5 dB PX = 10 dB. 32-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

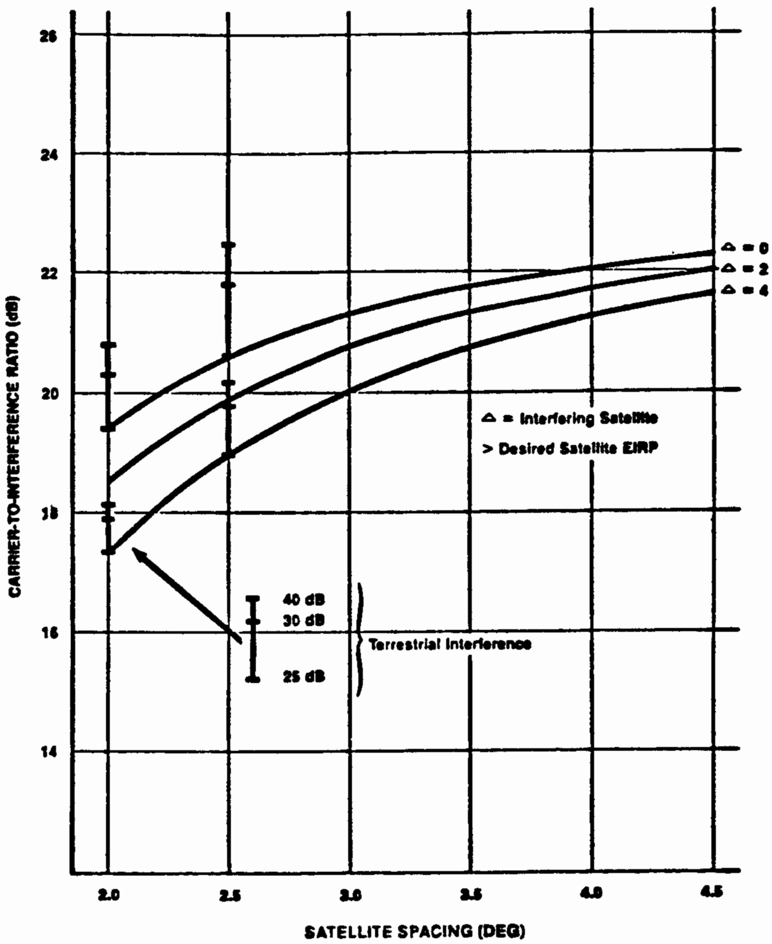


Figure 5. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink ana 7-Meter Downlink. FX = 6.5 dB PX = 10 dB. Proposed 29-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

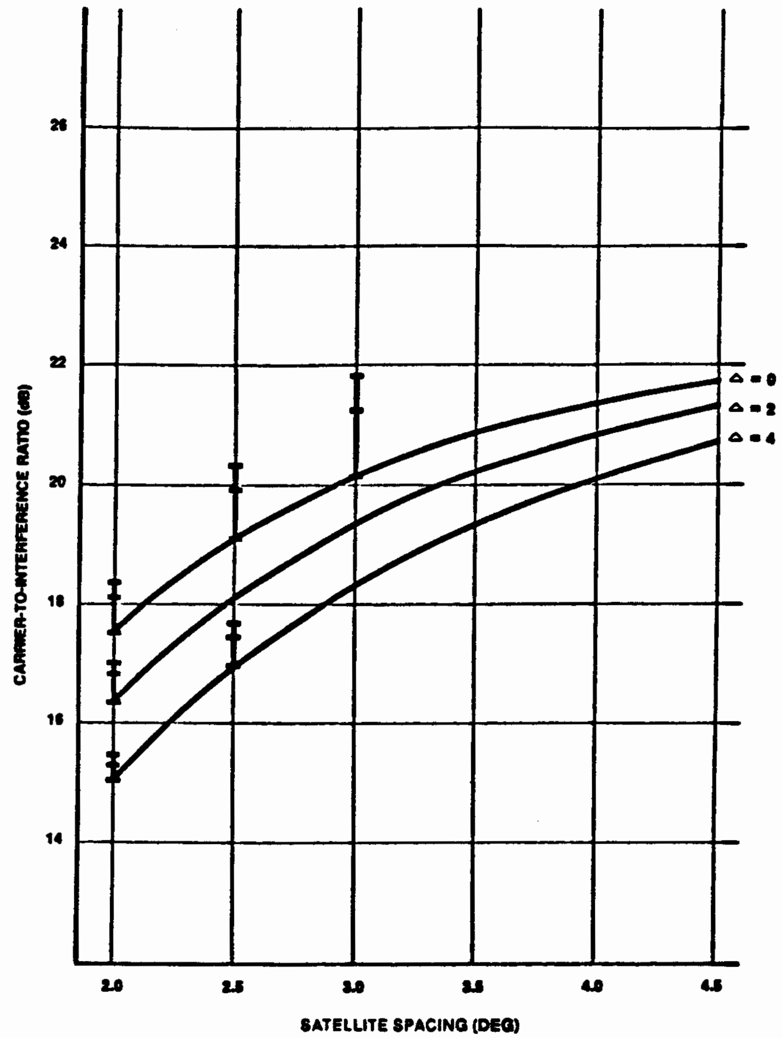


Figure 6. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink and 7-Meter Downlink. FX = 6.5 dB PX = 10 dB. 32-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

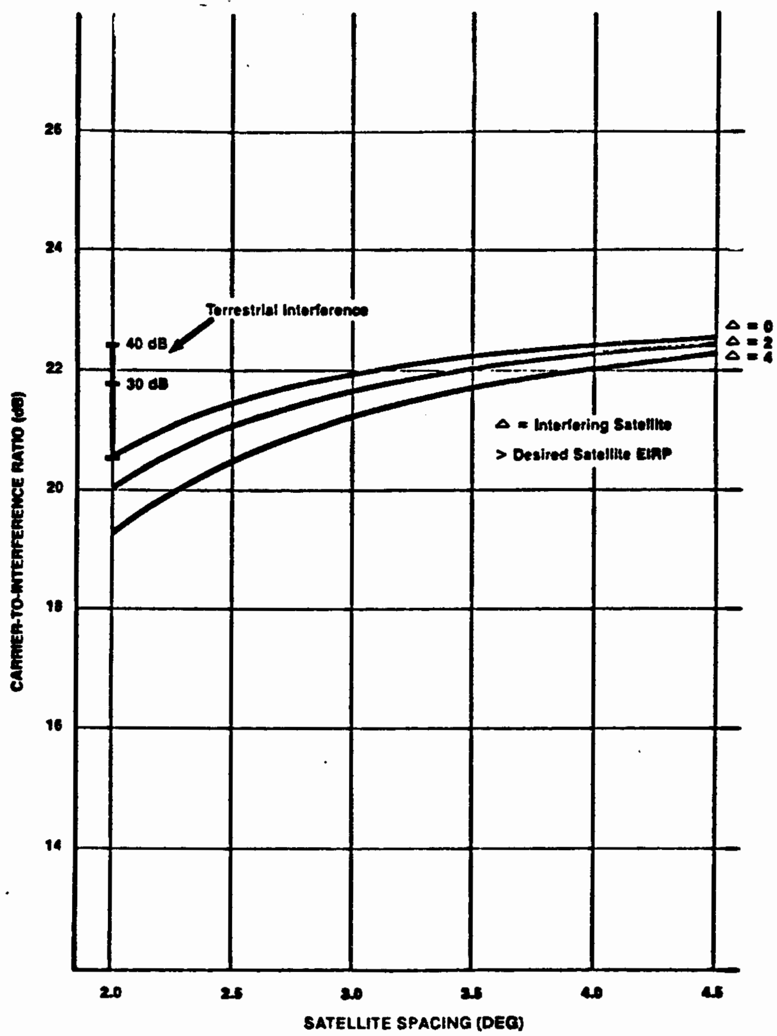


Figure 7. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink and a 10-Meter Downlink. FX = 6.5 dB PX = 10 dB. Proposed 29-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

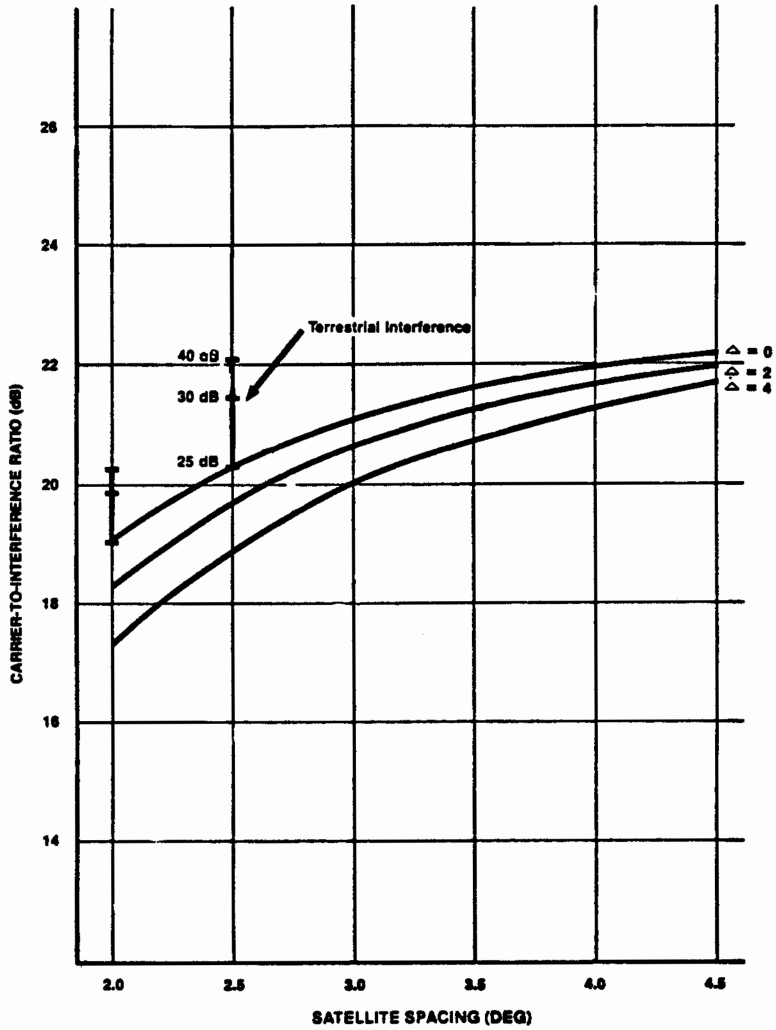


Figure 8. Carrier-to-Interference Ratio vs. Satellite Spacing for a 10-Meter Uplink an 10-Meter Downlink. FX = 6.5 dB PX = 10 dB. 32-25 Log Θ Pattern Envelopes and Adjacent Satellite Polarization Interleaving.

Appendix A

The geostationary orbit for a satellite is a circle lying in the equatorial plane which has a radius of:

$$R_g = 6.611\ R_o \qquad (1)$$

where:

R_o is the radius of the earth (3959 statute miles).

The location of the satellite is specified by the corresponding longitude coordinate, L_{sat}. For a station at a given latitude and longitude, the slant range r from the station to the satellite is given by:

$$r = R_o\left(6.611^2 + 1\text{-}2\ (6.611)\ \cos\ (H)\ \cos\ (\Delta L)\right)^{1/2} \text{ statute miles} \qquad (2)$$

where:

ΔL = station longitude — satellite longitude

H = station latitude

It is apparent that the angular separation between two satellites as seen by an earth station is dependent on the station latitude and longitude and the longitudes of the two satellites.

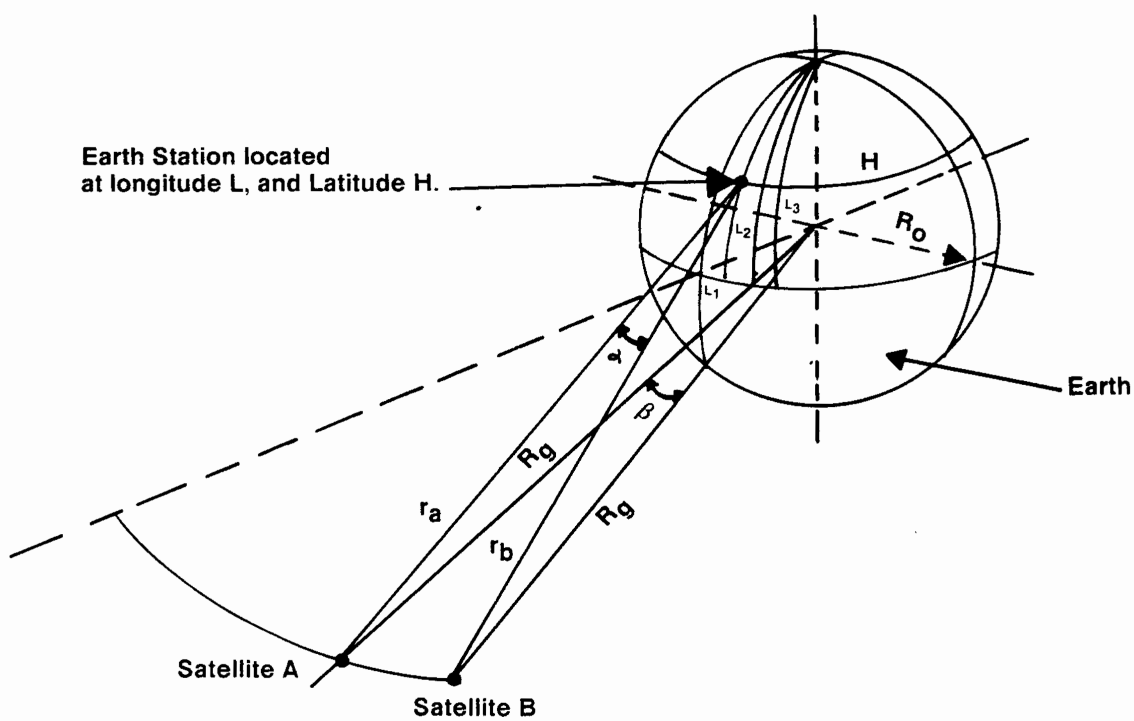


Figure A1. Geometry of Two Satellites as Viewed from Earth Station

The separation distance, $\overline{AB}$, between satellites A and B can be calculated from the law of cosines for the triangle OAB in Figure A1.

$$\overline{AB} = \left(2R_g^2 \ (1 - \cos \beta)\right)^{1/2} \qquad (3a)$$

$$\overline{AB} = R_0 \left(2 \ (6.611)^2 \ (1 - \cos \beta)\right)^{1/2} \qquad (3b)$$

where:

β is the orbital arc separation between satellites.

The triangle SAB in Figure A1 can be used to calculate $\overline{AB}$ also.

$$\overline{AB} = (r_a^2 + r_b^2 - 2r_a \, r_b \cos \alpha)^{1/2} \qquad (4)$$

where:

α is the effective angular separation between satellites A and B as viewed from the earth station, S.

Equating (3b) and (4) yields an equation relating α and β.

$$R_0^2 \left(2 \ (6.611)^2 \ (1 - \cos \beta)\right) = r_a^2 + r_b^2 - 2r_a r_b \cos \alpha \qquad (5)$$

$$\cos \alpha = \left[r_a^2 + r_b^2 - 2R_g^2 \ (1 \ - \cos \beta)\right] / 2r_a r_b$$

Substituting for r_a and r_b (from equation 2) yields:

$$2(6.611)^2 \ (1 - \cos \beta) = \left[\left((6.611)^2 + 1 - 2(6.611) \cos H \cos \Delta L_A\right)\right]$$
$$+ \left[\left((6.611)^2 + 1 - 2(6.611) \cos H \cos \Delta L_B\right)\right] - 2 \left[\left((6.611)^2 + 1\right.\right.$$
$$\left.\left.- 2(6.611) \cos H \cos \Delta L_A\right)^{1/2}\right]\left[\left((6.611)^2 + 1 - 13.322 \cos H \cos \Delta L_B\right)^{1/2}\right] \cos \alpha$$

Since several constants recur in this equation, it is convenient to identify them as Cs.

$$C_1 = (6.611)^2 = 43.70532$$
$$C_1 = C_1 + 1 = 44.70532$$
$$C_3 = 2(6.611) = 13.222$$

Substituting Cs and solving for cos α yields:

$$\cos \alpha = \frac{2C_2 - C_3 \cos H \ (\cos \Delta L_A + \cos \Delta L_B) - 2C_1 \ (1 - \cos \beta)}{2(C_2 - C_3 \cos H \cos \Delta L_A)^{1/2} \ (C_2 - C_3 \cos H \cos \Delta L_B)^{1/2}} \qquad (6)$$

It is now of interest to calculate the effective angular separation for adjacent satellite orbit positions as viewed by earth stations located in the continental United States for the orbital segment, 70°W to 135°W for β = 3.8°; equation 6 becomes:

$$\cos \alpha = \frac{2C_2 - 0.19218 - C_3 \cos H(\cos \Delta L_A + \cos \Delta L_B)}{2(C_2 - C_3 \cos H \cos \Delta L_A)^{1/2} (C_2 - C_3 \cos H \cos \Delta L_B)^{1/2}} \quad (7)$$

Appendix B: Adjacent Channel Interference Analysis

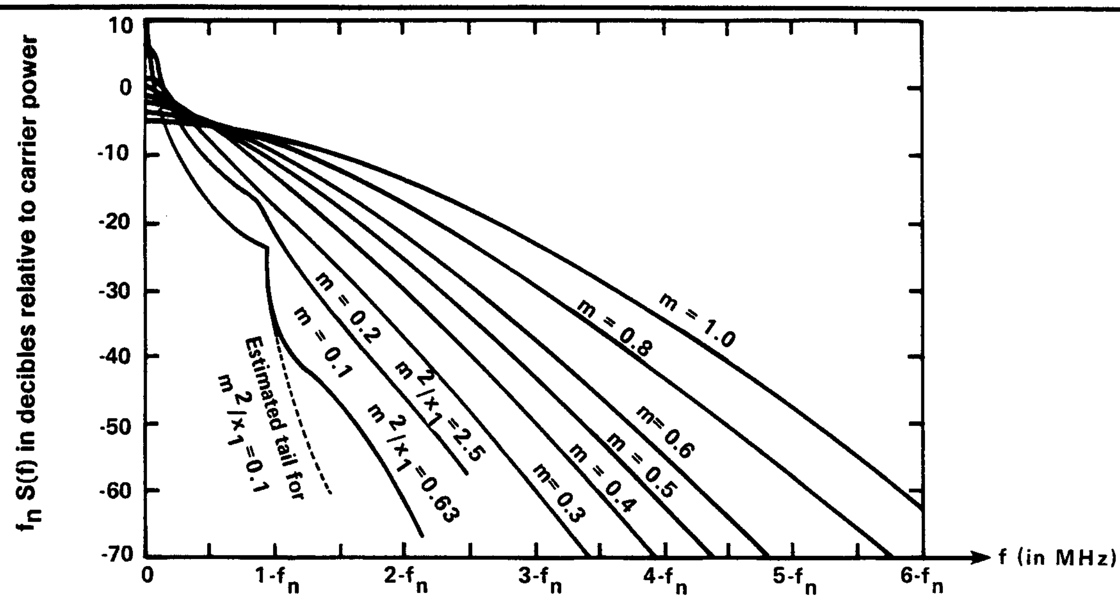
Introduction

Adjacent channel interference can be computed by convolving the power spectra of the wanted and unwanted signals. In the following analysis it is assumed that all the interferences are incoherent. The power spectra of each signal is described. The individual signals are Frequency-Division-Multiplex-Frequency Modulation (FDM-FM), Television-Frequency Modulation (TV/FM), High Bit Rate Data (HBR, i.e., 4 ϕ PSK), and FM Single-Channel-per-Carrier (SCPC).

Power Spectra of an FDM-FM Carrier

The spectrum of a carrier that is frequency modulated by a multiplexed telephony baseband is, in general, a complicated function which depends on many parameters. When the baseband signal consists of many single-sideband, frequency-multiplexed telephone channels, it is often convenient to simulate the baseband signal by an equivalent band of random noise. The determination of the power spectrum when the modulating signal consists of random noise involves considerable analysis. A particular case, often assumed in the analysis of a radio system, is that of FM by a random noise signal of uniform power sensitivity. (Proc. IEEE, Part B, Vol. 108, pp 75-89, Jan. 61). The shape of the power spectrum in this case largely depends on the modulation index (rms modulation index is useful since the modulating signal is a random noise voltage).

When the rms modulation index is very small and the lowest modulation frequency is not zero, a bounded continuous spectrum results, together with a residual carrier at the mean carrier frequency as shown in Figure B1 (the case of $m = 0.1$, $m^2/x_1 = 0.1$, $x_1 = f_1/f_n$). The residual carrier corresponds to the carrier component of the spectrum with a single modulating frequency.



f_n: Highest modulating frequency in MHz.
S(f): Power spectrum in dB/MHz.
m: Multi-channel RMS deviation.

Figure B1. Power spectrum of a FDM-FM carrier for RMS modulation index m=0.1 ~ 1.0. S(f)f_n is expressed in dB relative to unmodulated carrier power (normalized for unit carrier power)

For intermediate values of rms modulating index, power spectra based on measurements are believed to be the most reliable. Normalized spectrum curves obtained from measurements are shown in Figure B1 for eight values of m between 0.1 and 1.0.

When the rms modulation index is larger (>1.5), the mean power spectrum normalized for unit carrier power is of the form:

$$S_\phi(f) = \frac{1}{\sqrt{2\pi}\sigma} \exp\left(-f^2/2\sigma^2\right)$$

where:

σ = multichannel rms deviation in MHz

f = frequency relative to carrier frequency in MHz

$\phi(t)$ = multiplexed telephone baseband signal, sumulated by a random noise signal of uniform power spectra.

Figure B2 shows the power spectrum when rms deviation is 4 MHz. In order for rms modulation index, m, to be greater than 1.5, the highest modulation frequency has to be less than 2.96 MHz resulting in Carson's bandwidth of 34 MHz. In this case, 705 voice channels can be multiplexed.

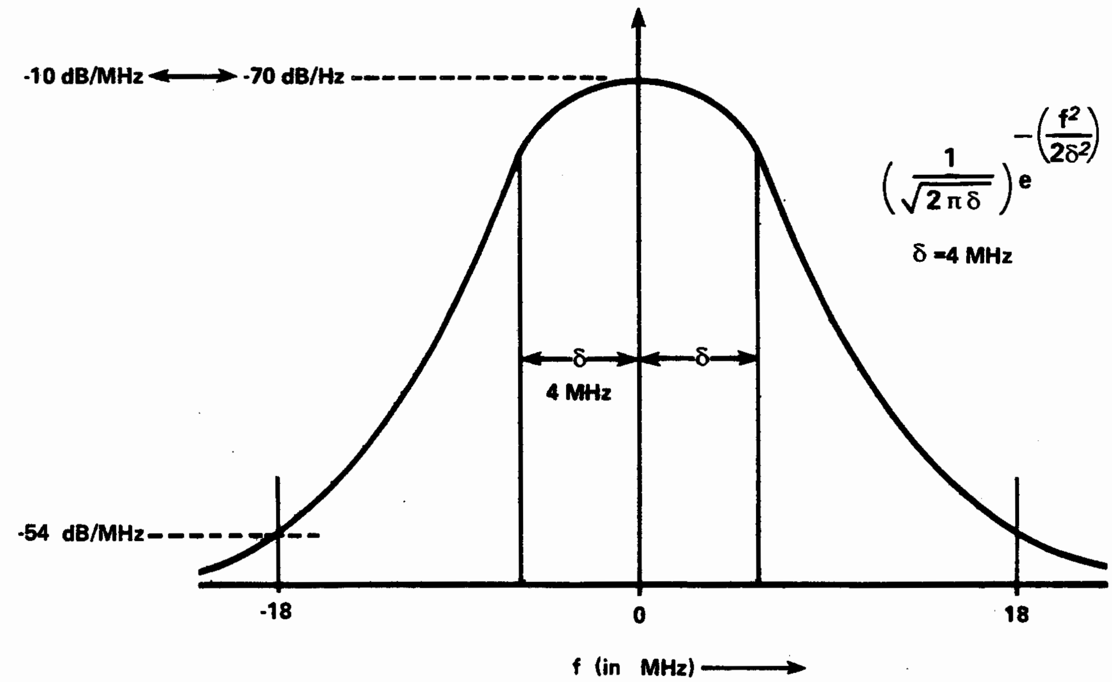


Figure B2. Power spectrum of a FDM-FM carrier when rms modulation index is large

Power Spectrum of a TV/FM Carrier

According to H.W. Evans, Bell Laboratories, the power in any 4-kHz band is at least 30 dB below (i.e., 66 dB/Hz) the power of the unmodulated carrier when the peak frequency deviation ratio is three, using Bell System standard pre-emphasis. These calculations assumed the FM spectrum of a band of white noise is similar to the spectra of preemphasized FM/TV signals near the carrier where the density is highest.

According to the power spectra shown on page 461 of Bell Laboratories "Transmission Systems for Communications," the power spectra is almost flat over the bandwidth of $2f_T$ and drops outside of this range with the rate of change dependent upon the rms phase deviation. The power spectral shape reported by COMSAT is similar. It is flat over ±12.5 MHz from the carrier and

$$B_{IF} = 2(\Delta f_1 + f_m)$$
$$= 2(12.5 + 5.5) = 36 \text{ MHz}$$

If we assume most of the power is contained in this 25-MHz band at this frequency deviation, the power in any 1-MHz band is 15 dB below the power of the unmodulated carrier (10 log (1/25) = -14 dB). See Figures B3 and B4.

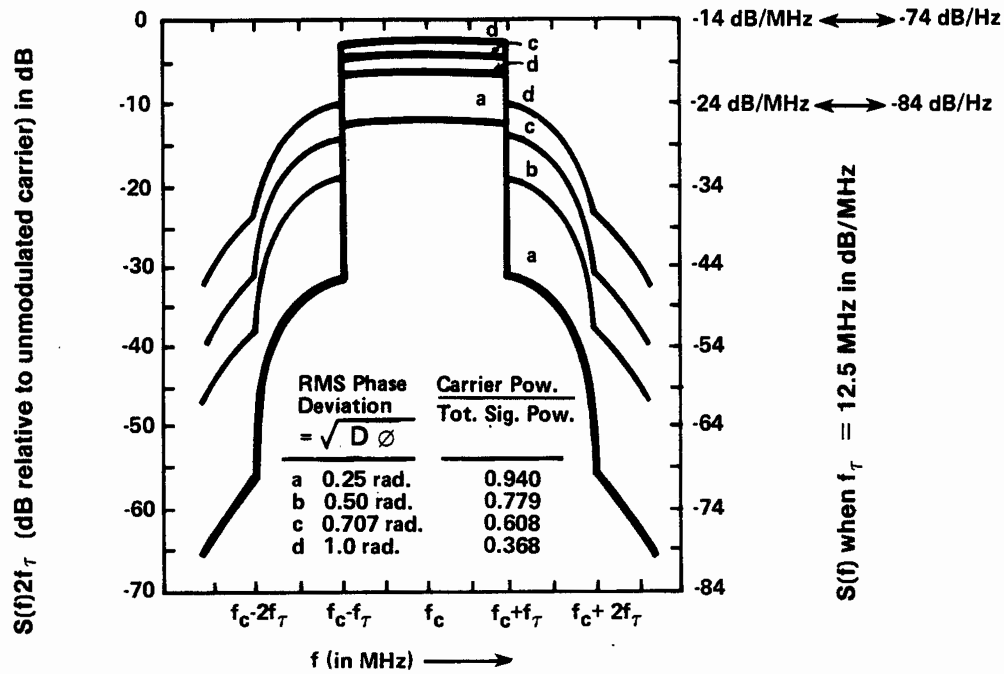


Figure B3. Sideband Spectra of a Carrier Phase-Modulated by a Baseband Signal Consisting of a Flat Band of Random Noise which Extends from 0 to f $_T$ Hz

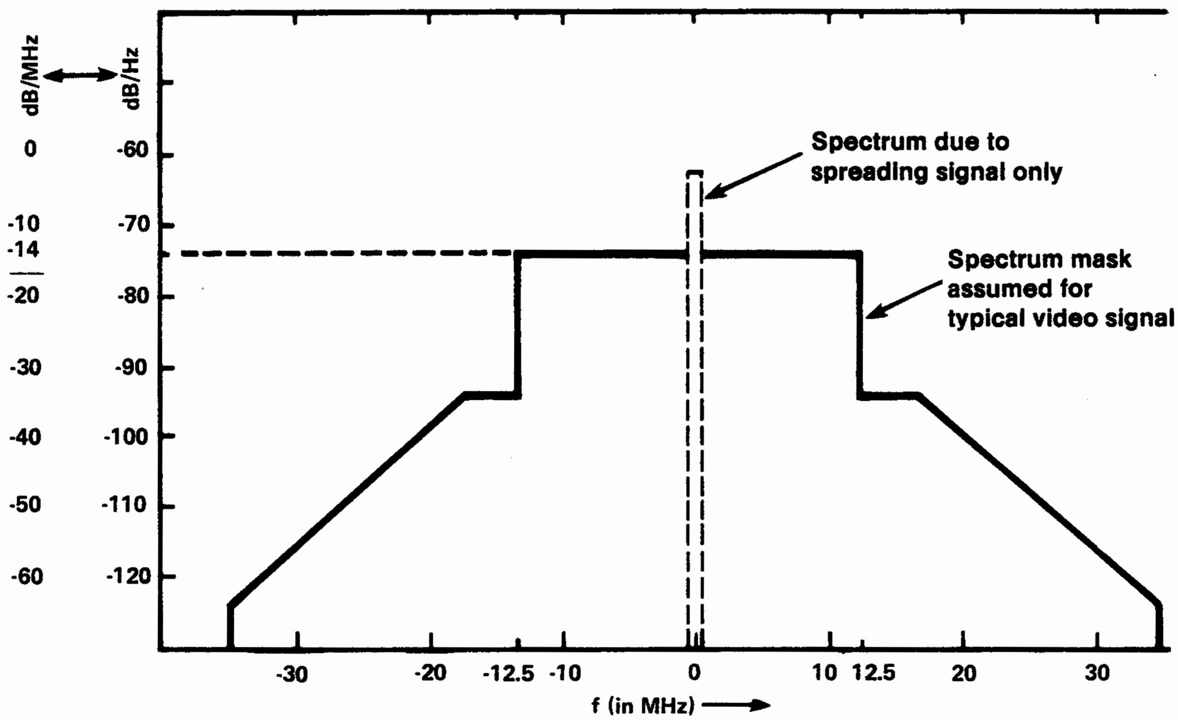


Figure B4. Spectral "Mask" for FM/TV Carrier

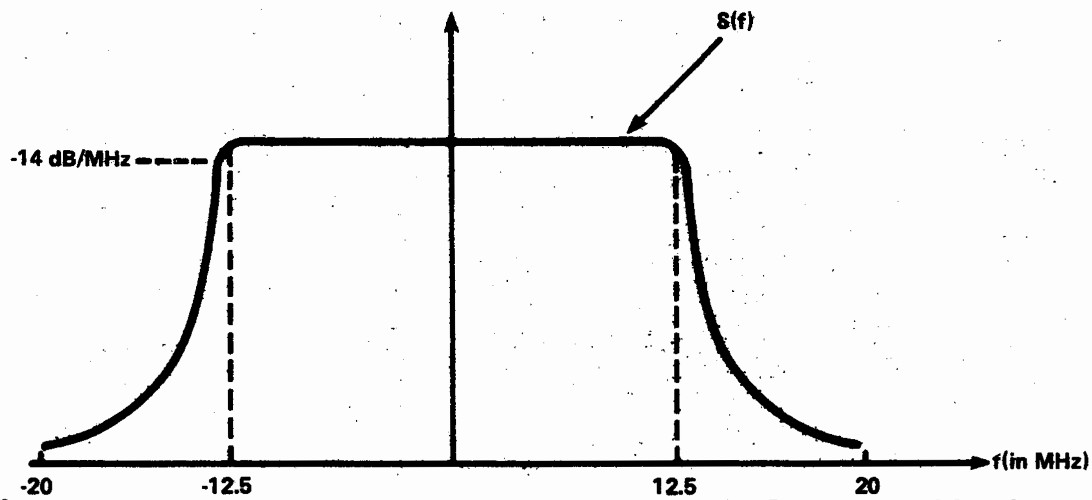


Figure B5. Approximate Power Spectrum of a FM/TV Carrier used for the Analysis

Power Spectrum of a 4Ø-PSK Carrier (High Bit Rate Data)

The primary cause of adjacent channel interference from PSK carriers is the power spectrum spreading due to TWT non-linearities. Power spectrum spreading is discussed extensively by Lyons, "Effects of PSK Spectral Spreading in a Satellite Transponder," IEEE International Comm. Conf., pp 363-1, June 1974, and will not be discussed here.

The power spectrum shown in Figure B6 below will be used for this analysis.

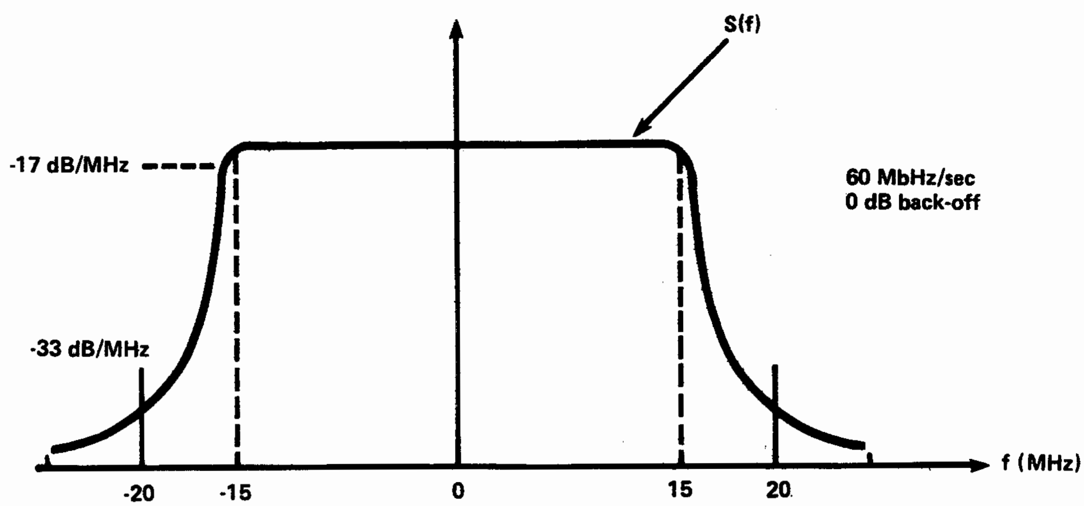


Figure B6. Approximate Power Spectrum of a 4Ø-PSK Carrier (high bit rate data)

Power Spectrum of SCPC Carriers and Associated Intermod

The primary interference of SCPC carriers is due to the carriers themselves and not to the associated intermod power spectrum. Theoretical investigation has shown that the intermod spectrum peak is 16 below the carrier level and therefore will be neglected in this analysis.

Interference to a FM/TV Channel

Let I denote the interfering power without taking the polarization isolation into consideration, expressed in dB below saturation.

$$I = 10 \log \left(S(f_1) + 10 \log (B)\right), \text{ dB below saturation}$$

where:

$S(f_1)$ = power spectral density at f_1 (MHz)

f_1 = frequency from the center of the power spectrum of an interfering signal to the FM/TV carrier

B = RF noise bandwidth associated with an FM/TV carrier (MHz)

For FDM/FM-Co-Polarized:

$$I = 10 \log \left[\frac{1}{\sqrt{(2\pi)(16)}} \exp\left(-40^2/(2 \times 4^2)\right)\right] + 10 \log (B)$$

$$= \log(1.9339 \times 10^{-23}) + 10 \log B$$

$$= -227 + 10 \log B$$

$$= \text{negligible}$$

For HBR Data:

$$I = -17 \text{ dB/MHz} + 10 \log B$$

$$I = \frac{-17 \text{ dB}}{\text{MHz}} (25 \text{ MHz}) + 10 \log B$$

$$= -425 + 10 \log B$$

$$= \text{negligible}$$

FDM-FM Cross Polarized:

$$I = 10 \log \left[\int_{-\infty}^{-3} \frac{1}{\sqrt{2\pi}\,\sigma} \exp\left(-f^2/\,2\sigma^2\right) df \right]$$

where:

σ = 4 MHz

I = 10 log (0.2266)

= -6.45 dB below saturation

SCPC Cross-Polarized

By assuming that SCPC carriers in 14-MHz band (17 MHz - 3 MHz) are directly interfering with FM/TV channel, the interfering power will be:

I = 10 log (14 MHz/34 MHz)

= -3.9 dB below saturation

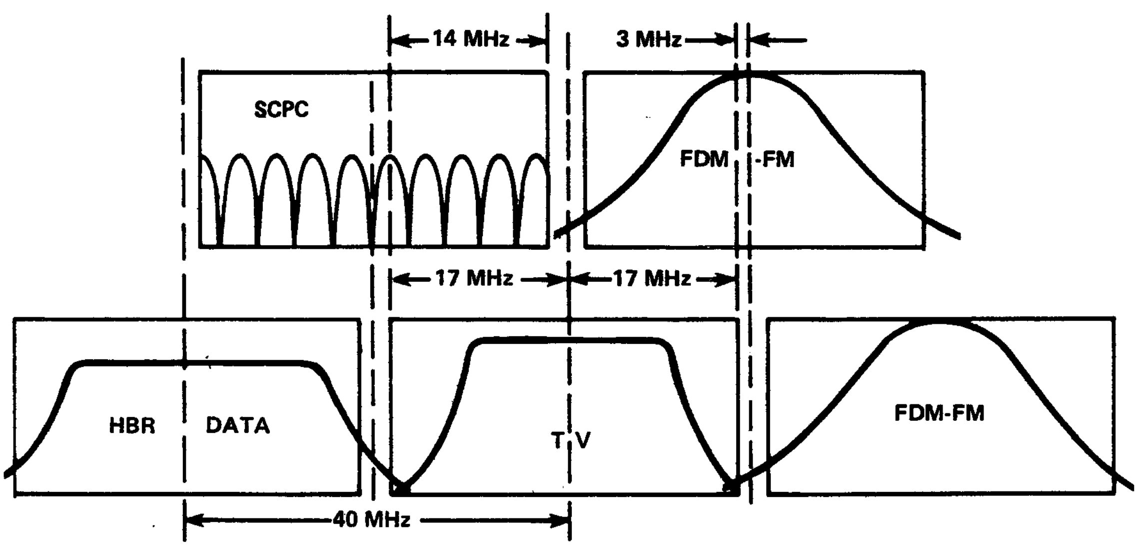


Figure B7. Interference Model to a TV Channel for Typical Reuse Satellite

C/I Due to Cross-Polarized FDM/FM Channel:

For the offset cross-polarized FDM/FM channel, the interfering power I is -6.45 dB below saturation. This means that the interfering power into the FM/TV channel on the uplink is:

$$P_I = -81.5 \text{ dBW/m}^2 + (-37 \text{ dBm}^2) + G_{sat} - 6.45 - (XPD)_u$$

The power of the FM/TV carrier on the uplink is:

$$P_C = -81.5 \text{ dBW/m}^2 + (-37 \text{ dBm}^2) + G_{sat}$$

saturation flux density = -81.5 dBW/m^2

effective area of isotropic = -37 dBm^2

Therefore, the carrier-to-interference ratio on the uplink is:

$$\left(\frac{C}{I}\right)_u = 6.45 + (XPD)_u$$

where:

XPD is the polarization discrimination to the cross-polarized signal.

For the downlink, the interfering power is:

$$P_I = 32 \text{ dBW} - \text{Path loss} + G_{ES} \quad -6.45 - (XPD)_D$$

The FM/TV carrier power on the downlink is:

$$P_C = 32 \text{ dBW} - \text{Path Loss} + G_{ES}$$

Therefore, the carrier-to-interference ratio on the downlink is:

$$\left(\frac{C}{I}\right)_D = 6.45 + (XPD)_D$$

Then, the total carrier-to-interference ratio will be:

$$\frac{C}{I} = \left(\frac{C}{I}\right)_u \oplus \left(\frac{C}{I}\right)_D$$

$$= (6.45 + (XPD)_U) \oplus (6.45 + (XPD)_D)$$

where:

$\oplus$ denotes power summation

C/I Due to Cross-Polarized SCPC Channel:

For the offset cross-polarized SCPC channel, the interfering power, I is -3.9 dB below saturation. This means that the interfering power into the FM/TV channel on the uplink is:

$$P_I = (-81.5\ \text{dBW/m}^2 - \text{input backoff}) + (-37\ \text{dBm}^2) + G_{sat}$$

$$-3.9 - (XPD)_u$$

The power of the FM/TV carrier on the uplink is:

$$P_C = -81.5\ \text{dBW/m}^2 + (-37\ \text{dBm}^2) + G_{sat}$$

Therefore, $\left(\frac{C}{I}\right)_u = + 3.9 + \text{input backoff} + (XPD)_u$

Typically input backoff = 4 dB

$$\left(\frac{C}{I}\right)_u = +7.9 + (XPD)_u$$

On the downlink, the interfering power is:

$$P_I = (32\ \text{dBW} - \text{output backoff}) - \text{Path Loss} + G_{ES} - 3.9 - (XPD)_D$$

and the power of the FM/TV carrier is:

$$P_C = 32\ \text{dBW} - \text{Path loss} + G_{ES}$$

Therefore, $\left(\frac{C}{I}\right)_D = 3.9 + \text{output backoff} + (XPD)_D$

Output backoff = 2.6

$$\left(\frac{C}{I}\right)_D = 6.5 + (XPD)_D$$

Total carrier to interference ratio will be:

$$\frac{C}{I} = \left|7.9 + (XPD)_u\right| \oplus \left|6.5 + (XPD)_D\right|$$

Total C/I for System

$$\left(\frac{C}{I}\right)_S = \left(\frac{C}{I}\right)_{FDM/FM} \oplus \left(\frac{C}{I}\right)_{SCPC}$$

Polarization: Typical Site Central United States

I. Uplink

Satellite	35 dB	
Ground Station	35 dB	28.0 dB
Faraday	35 dB	

II. Downlink

Satellite	35 dB	
Ground Station	35 dB	26.0 dB
Faraday	29 dB	

III. Atmospheric Effect - 25° Elevation Angle

% Time	Rain Rate	4 GHz	6 GHz
99.0	1/2"/hr	33.0	30.5
99.9	1-1/2"/hr	25.0	21.0
99.99	3"/hr	20.0	16.5

IV. Polarization Discrimination

% Time	4 GHz	6 GHz
99.0	22.8	23.1
99.9	19.5	17.8
99.99	16.5	14.5

V. Carrier-To-Interference TV/FM

% Time	$(XPD)_u$	$(XPD)_D$	$(C/I)_u$	$(C/I)_D$	$(C/I)_s$
99.0	23.1	22.8	27.10	26.56	23.81
99.9	17.8	19.5	21.2	21.26	18.22
99.99	14.5	16.5	18.6	17.96	15.26

A more thorough examination of this subject is contained in a RCA American Communications transmission engineering report prepared by Dr. M.K. Lee. This report is entitled, "Overall System Cross-Polarization Isolation and Interference from Cross-Polarized Channels in the Spectrum Reuse System," Report No. TER-008-75. The discussion and analysis of this paper contains concepts and details that are presented in this report. Appreciation is noted for Dr. Lee's work.

Introduction to Earth Station Antennas

James H. Cook. Jr.

Introduction

An earth station antenna system is made up of many component parts such as the receiver, low-noise amplifier, antenna, etc. All of the components have an individual role to play and their importance in the system should not be minimized. The antenna, of course, is one of the more important component parts since it provides the means of transmitting signals to the satellite and/or collecting the signal transmitted by the satellite. The antenna not only must provide the gain necessary to allow proper transmission and reception but also must have radiation characteristics which discriminate against unwanted signals and minimize interference into other satellite or terrestrial systems. The antenna also provides the means of polarization discrimination of unwanted signals. The individual communication system operational parameters dictate to the antenna designer the necessary electromagnetic, structural, and environmental specifications necessary for the antenna.

General Requirements

Antenna requirements can be grouped into several major categories, namely, electrical or RF, control systems, structural, pointing and tracking accuracy, environmental, and miscellaneous requirements such as radiation hazard, primary power distribution, etc. Only the electrical or RF requirements will be dealt with herein.

Frequency

The World Administrative Radio Conference (WARC) has the responsibility of frequency assignments for communication and other radiating services. WARC was convened in the latter months of 1979. Frequency allocations for all types of satellite communications were set and entered into force as of January 1, 1982 after ratification by member Administrations of the International Telecommunications Union. Tables 1 through 5 are a tabulation of the satellite communication services frequency allocations in the WARC 79 frequency allocations. These services are the fixed-satellite service (FSS), the inter-satellite service (ISS), the broadcasting-satellite service (BSS), the mobile-satellite service (MSS) and the maritime mobile- and aeronautical mobile-satellite services. The tables indicate the regional extent of the allocation for each band and the bandwidth. Earth-to-space (uplink) bands and space-to-Earth (downlink) bands are listed separately, with commonly paired bands on the same line. For allocation purposes, the world is divided into three geographical regions: Region 1 contains Europe, Africa, the USSR, and the Peoples Republic of Mongolia; Region 2 consists of the Americas and Greenland; and Region 3 includes Asia (except USSR and Mongolia), Australia, New Zealand, etc.

Table 1. WARC 79 Fixed Satellite Service Allocations

Earth-to-Space (GHz)	Region	Bandwidth (MHz)	Space-to-Earth (GHz)	Region	Bandwidth (MHz)
2.655-2.690	2^b, 3^b	35	2.50-2.535	2^b, 3^b	35
			2.535-2.690	2^b	155
5.725-5.850	1	125	3.40-4.20	1, 2, 3	800
5.850-7.075	1, 2, 3	1225	4.50-4.80	1, 2, 3	300
7.90-8.40	1, 2, 3	500	7.25-7.75	1, 2, 3	500
12.50-12.7	1	200			
12.70-12.75	1, 2	50			
12.75-13.25	1, 2, 3	500			
			10.70-11.70	1, 2, 3	1000
14.00-14.50	1, 2, 3	500			
27.00-27.50	2, 3^c	500	11.70-12.30	$2^{b,c}$	600
			12.20-12.50	3^b	300
			12.50-12.75	1, 3	250
27.50-31.00	1, 2, 3	3500	17.70-21.20	1, 2, 3	3500
42.5-43.50	1	1000			
47.2-49.20	2^a	2000	37.50-40.5	1, 2, 3	3000
49.2-50.2	1	1000			
50.40-51.40	1	1000			
71.0-74.0	1, 2, 3	3000	81.0-84.0	1, 2, 3	3000
74.0-75.5	1, 2, 3	1500			
92.0-95.0	1, 2, 3	3000	102.0-105.0	1, 2, 3	3000
202.0-217.0	1, 2, 3	15000	149.0-164.0	1, 2, 3	15000
265.0-275.0	1, 2, 3	10000	231.0-241.0	1, 2, 3	10000

a. Intended for use by, but not limited to, BSS feeder links
b. Limited to national and sub-regional services
c. Upper band limit (12.3 GHz) may be replaced by a new value in the range 12.1 to 12.3 GHz at the 1983 RARC for Region 2.

Table 2. WARC 79 Intersatellite Service Allocations

Band (GHz)	Bandwidth (GHz)
22.55-23.55	1
32.00-33.00	1
54.25-58.20	3.95
59.0-64.0	5
116.0-134.0	18
170.0-182.0	2
185.0-190.0	5

Table 3. Broadcasting-Satellite Service Allocations

Earth-to-Space (GHz)	Region	Bandwidth (MHz)	Space-to-Earth (GHz)	Region	Bandwidth (MHz)
Feeder links for the BSS may, in principle, use any of the FSS (Earth-to-Space) bands listed in Table 2 with appropriate coordination. However, the following bands were set aside for exclusive or preferential use by such feeder links.			0.62-0.79[a]	1, 2, 3	170
			2.50-2.69[b]	1, 2, 3	190
10.70-11.70	1	1000	11.70-12.1	1, 3	400
14.50-14.80	1[c], 2, 3	300	12.10-12.2	1, 2, 3	100
17.30-18.1	1, 2, 3	800	12.20-12.5	1, 2	300
			12.50-12.7	2, 3[b]	200
			12.70-12.75	3[b]	50
27.00-27.50	2, 3	500	22.50-23.00	2, 3	500
47.20-49.20	1, 2, 3	2000	40.50-42.50	1, 2, 3	2000
			84.00-86.00	1, 2, 3	2000

a. Limited to TV
b. Limited to community reception
c. Excluding Europe

The allocations recognized the need for substantial increases for commercial FSS systems in the vicinity of the 6/4- and 14/11-GHz bands, for maritime mobile- and aeronautical mobile-satellite systems in the 1.6-GHz band, and for domestic broadcasting- and fixed-satellite systems in the 11.7- to 12.7-GHz band for the Americas. As can be seen from the tables, the primary frequency bands for the Americas for domestic satellite communications will continue to be the 6/4-GHz bands and 14/12-GHz bands. For the international satellite systems the 6/4-GHz and 14/11-GHz bands will continue to be the predominant FSS and BSS frequencies.

Table 4. Mobile Satellite Services Allocations

Earth-to-Space (MHz)	Region	Bandwidth (MHz)	Space-to-Earth (MHz)	Region	Footnote
121.45-121.55[a,b,h]	1, 2, 3		--		3572A
242.95-243.05[a,b,h]	1, 2, 3		--		3572A
235.0-322.0[a,h]	1, 2, 3	87			3618
335.4-399.9[a,h]	1, 2, 3	64.4			3618
405.5-406.0[c]	Canada	0.5	--		3533A
406.0-406.1[b]	1, 2, 3	0.1	--		3633A
406.1-410.0[c]	Canada	3.9	--		3634
608.0-614.0[d,e]	2	6.0	--		
806.0-890.0[c,h]	2, 3	3.0			3662C
	Norway				3662CA
	Sweden				3670B
942.0-960.0[c,h]	3,Norway	18			3662C
	Sweden				3662CA
1645.5-1646.5[f]	1, 2, 3	1.0	1554.0-1545.0[f]	1, 2, 3	3695A
(GHz)			(GHz)		
7.90-8.025[a]	1, 2, 3	125	7.250-7.375[a]	1, 2, 3	3764B
14.00-14.50[e,g]	1, 2, 3	500	--		
29.50-30.0[e]	1, 2, 3	500	19.70-20.20[e]	1, 2, 3	
30.00-31.00	1, 2, 3	1000	20.20-21.20	1, 2, 3	
43.50-47.00[h]	1, 2, 3	3500	39.50-40.50	1, 2, 3	3814C
50.40-51.40[e]	1, 2, 3	1000	--		
66.00-71.00[h]	1, 2, 3	5000			3814C
71.00-74.00	1, 2, 3	3000	81.00-84.00	1, 2, 3	--
95.00-100.00[h]	1, 2, 3	5000			3814C
135.00-142.00[h]		7000			3814C
190.00-200.00[h]		10000			3814C
252.00-265.00[h]		13000			3814C

a. Footnote allocation
b. Emergency position indicating radio-beacons only
c. Footnote allocation excludes aeronautical mobile-satellite services
d. Excludes aeronautical mobile-satellite services
e. Secondary allocation
f. Distress and safety operations only
g. Footnote allocation to land mobile-satellite service only
h. No direction specified

Table 5. Maritime and Aeronautical Mobile-Satellite Service Allocations

MARITIME MOBILE Earth-to-Space	MARITIME MOBILE Space-to-Earth	AERONAUTICAL MOBILE Earth-to-Space	AERONAUTICAL MOBILE Space-to-Earth
1626.5-1645.5 MHz	1530.0-1544.0 MHz	1646.5-1660.0 MHz	1545.0-1559.0 MHz
		1610.0-1625.5 MHz[a,*] 5000.0-5250.0 MHz[a,*] 15.40-15.70 GHz[a,*]	

a. No direction specified
* Footnote allocation

Radiation Envelope

The primary electrical specifications of an earth station antenna are gain, noise temperature, VSWR, power rating, receive/transmit group delay, radiation pattern, polarization, axial ratio, isolation, and G/T. All of the parameters except the radiation pattern are determined by the system requirements. The radiation pattern should meet the minimum requirements set by International Radio Consultative Committee (CCIR) of the International Telecommunications Union (ITU) and/or the national regulatory agencies such as the Federal Communication Commission (FCC). Earth stations which operate in a regulated environment in the United States domestic system must meet the requirements set forth in the FCC regulations pertaining to sidelobes. (See Part 25, Paragraph 25.209 of FCC Regulations and RM-2725, Amendment of the Commission's Rules and Regulations or Policies Relative to Satellite Earth Station Antennas to Permit Receive-Only "Small Earth Stations".) The FCC regulation specifies a sidelobe envelope of:

$$(32-25\log\theta)\,\text{dBi} \qquad 1^\circ<\theta<48^\circ$$
$$-10\ \text{dBi} \qquad 48^\circ<\theta<180^\circ$$

where θ is the angle in degrees from the axis of the main lobe, and dBi refers to dB relative to an isotopic radiator. For the purposes of this envelope, the peak gain of an individual sidelobe may be reduced by averaging its peak level with the peaks of the nearest sidelobes on either side, or with the peaks of two nearest sidelobes on either side, provided that the level of no individual sidelobe exceeds the gain envelope by more than 6 dB. The antenna sidelobe envelope must conform to this specification in the 6- or 14-GHz transmit band, and it is recommended that the antenna meet the envelope in the 4- or 12-GHz receive band. In the small antenna ruling, the Commission stated that, "it appears that the carrier-to-interference objective will be satisfied for present domestic satellites if the earth station sidelobe levels do not exceed the envelope defined in Section 25.209 of the Rules and Regulations." (See Figure 1).

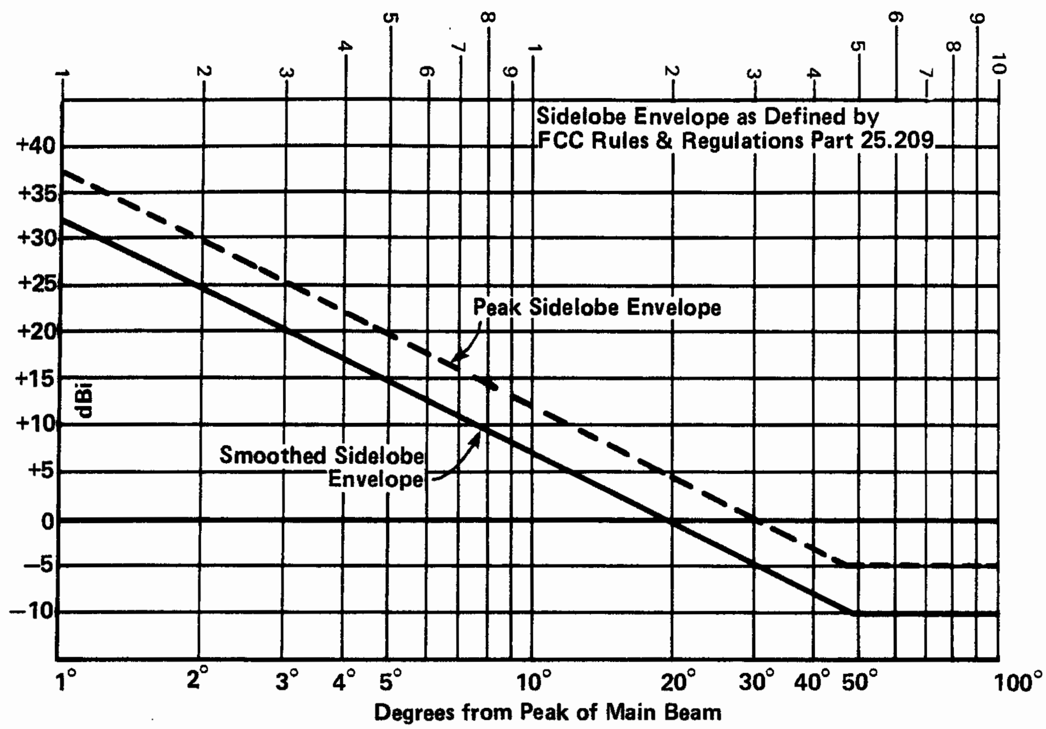


Figure 1.

In 1979 the FCC deregulated receive-only (RO) earth stations. This ruling permits a receive-only system to be installed and operated without the frequency coordination process to determine if the antenna system will have interference problems caused by existing terrestrial or satellite systems which share the same frequency band. Waiving coordination also denies the operating system the protection from future transmitting terrestrial or satellite system interference. Of course, the deregulation ruling also allows the use of antennas which do not meet the 32-25 log θ envelope, since the interference environment is the responsibility of the user and protection is no longer guaranteed by the FCC.

Earth station antennas operating in international satellite communications must have sidelobe performance as specified by INTELSAT Standards or by CCIR Recommendation 483 and Report 391-2 (see Figure 2). The CCIR Standard is somewhat more lenient than the FCC document in that it allows any peak sidelobe to exceed the envelope by 10 dB rather than 6 dB. Nevertheless, it should be pointed out that both documents refer to the (32-25 log θ) dBi envelope. The CCIR Standard specifies the envelope in terms of allowing 10% of the sidelobes to exceed the envelope and also permits the envelope to be adjusted for antennas whose aperture is less than 100λ. CCIR Report 391-2 allows the envelope to be determined by:

$$G = (52 - 10 \log(D/\lambda) - 25 \log \theta)\text{dBi}$$

when $D/\lambda < 100$. This envelope takes into consideration the theoretical limitations of small antenna design and is representative of the actual measured patterns of well-designed antennas.

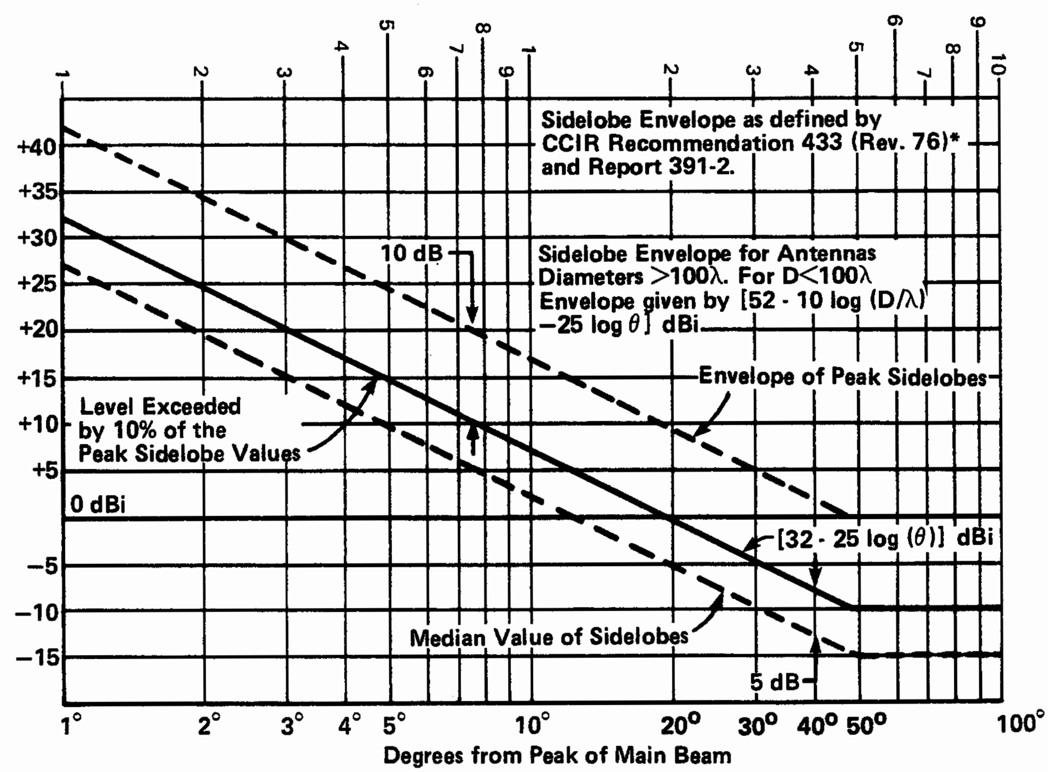


Figure 2.

On August 16, 1983, the FCC amended the Rules and Regulations Part 25.209 pertaining to the antenna gain envelope. The new standard is to apply to all new transmit antennas installed after July 1, 1984 and to all transmit antennas after January 1, 1987. The new standard is as follows:

a. The gain of any antenna to be employed in transmission from an earth station in the fixed-satellite service shall lie below the envelope defined below:

1. In the plane of the geostationary satellite orbit as it appears at the particular earth station location:

$29-25 \log_{10} \Theta$ dBi	$1° < \Theta < 7°$
$+ 8$ dBi	$7° < \Theta < 9.2°$
$32-25 \log_{10} \Theta$ dBi	$9.2° < \Theta < 48°$
$- 10$ dBi	$48° < \Theta < 180°$

where Θ is the angle in degrees from the axis of the main lobe, and dBi refers to dB relative to an isotropic radiator. For the purpose of this section, the peak gain of an individual sidelobe may not exceed the evelope defined above for Θ between 1° and 7°. For Θ greater than 7°, the envelope may be exceeded by 10% of the sidelobes, but no individual sidelobe may exceed the envelope by more than 3 dB.

2. In all other directions:

Outside the main beam, the gain of the antenna shall lie below the envelope defined by:

$32-25 \log_{10} \Theta$ dBi	$1^\circ \leq \Theta \leq 48^\circ$
-10 dBi	$48^\circ < \Theta \leq 180^\circ$

where Θ is the angle in degrees from the axis of the main lobe, and dBi refers to dB relative to an isotropic radiator. For the purpose of this section, the peak gain of an individual sidelobe may be reduced by averaging its peak level with the peaks of the nearest sidelobes on either side, or with the peaks of two nearest sidelobes on either side, provided that the level of no individual sidelobe exceeds the gain envelope given above by more than 6 dB.

b. The off-axis cross-polarization isolation of any antenna to be employed in transmission at frequencies between 5925 and 6425 MHz from an earth station to a space station in the domestic fixed-satellite service shall be defined by:

$19-25 \log_{10} \Theta$ dBi	$1.8^\circ < \Theta \leq 7^\circ$
-2 dBi	$7^\circ < \Theta \leq 9.2^\circ$

c. Any antenna licensed for reception of radio transmission from a space station in the fixed-satellite service shall be protected from radio interference caused by other space stations only to the degree to which harmful interference would not be expected to be caused to an earth station employing an antenna conforming to the standards defined in paragraphs a. and b. of this section.

d. The standards specified in paragraphs a. and b. of this section shall apply to all new antennas installed after July 1, 1984 and to all antennas after January 1, 1987.

e. The operations of any earth station with an antenna not conforming to the standards of paragraphs a. and b. of this section shall impose no limitations upon the operation, location, and design of any terrestrial station, any other earth station, or any space station.

f. Section 25.251(d)(4) is revised to read as follows:

The FCC further acknowledged within the text of the Federal Register Statement 47 CFR Port 25, Vol. 48, No. 173, September 6, 1983, that the envelope defined above is only a reference envelope in the receive band. Receiving antennas do not have to conform to this envelope to be eligible for licensing. Facilities with performance worse than the reference must, of course, accept correspondingly potential, higher interference levels. The interference levels should be calculated based on typical measured radiation patterns, site location and for a desired satellite or satellites. This analysis may result in acceptable receive-only carrier-to-interference performance for antennas meeting the present 32-25 log Θ envelope even with orbital spacings as small as 2°. Since discrimination, that is, peak on axis gain to sidelobe gain is the important determining factor, not an arbitrary sidelobe gain performance envelope relative to isotropic.

Figure of Merit, G/T

A very useful indication of the performance of an earth station is the figure of merit defined as:

$$G_T = 10 \log \left[\frac{\text{receiving antenna power gain}}{\text{system noise temperature (K)}} \right]$$

where the antenna power gain and system noise temperature are referred to some convenient point in the system; i.e., the low-noise amplifier input. Note that G/T is the figure of merit of a receiving earth station and does not apply to the transmit properties of an earth station. G/T is an important parameter of an earth station in that C/N, S/N, E_b/N_o, etc. are directly related to it. For a detailed explanation of G/T, the reader is referred to Hollis' paper of G/T and Noise Temperature. The relationships between G/T and other important system measures of performance are given by McKimmey's treatment of Link Analysis.

Types of Earth Station Antennas

Several types of earth station antennas are now in use within the United States and abroad. These antennas can be grouped into two broad categories--single-beam antennas and multiple-beam antennas. A single-beam earth station antenna is defined as an antenna which generates a single beam which is pointed toward a satellite by means of a positioning system. A multiple-beam earth station antenna is defined as an antenna which generates multiple beams by employing a common reflector aperture with multiple feeds illuminating that aperture. The axes of the beams are determined by the location of the feeds. The individual beam indentified with a feed is pointed toward a satellite by positioning the feed without moving the reflector. The majority of the earth station antennas now in use are single-beam antennas.

Single-beam antenna types used as earth stations are paraboloidal reflectors with focal point feeds (prime focus antenna), dual reflector antennas such as the Cassegrain and Gregorian configurations, horn reflector antennas, offset-fed paraboloidal antennas, and offset-fed, multiple reflector antennas. Each of these antenna types has its own unique characteristics, and the advantages and disadvantages have to be considered when choosing them for a particular application.

Axisymmetric Dual Reflector Antennas

The predominant choice of most system operators has been the dual-reflector Cassegrain antenna. Cassegrain antennas can be divided into three primary types:

a. The classical Cassegrain geometry[1,2] employing a paraboloidal contour for the main reflector and a hyperboloidal contour for the subreflector (see Figure 3). The paraboloidal reflector is a point focus device with a diameter D_p and a focal length f_p. The hyperboloidal subreflector has two foci. For proper operation, one of the two foci is the real focal point of the system and is located coincident with the phase center of the feed; the other focus, the virtual focal point, is located coincident with the focal point of the main reflector.

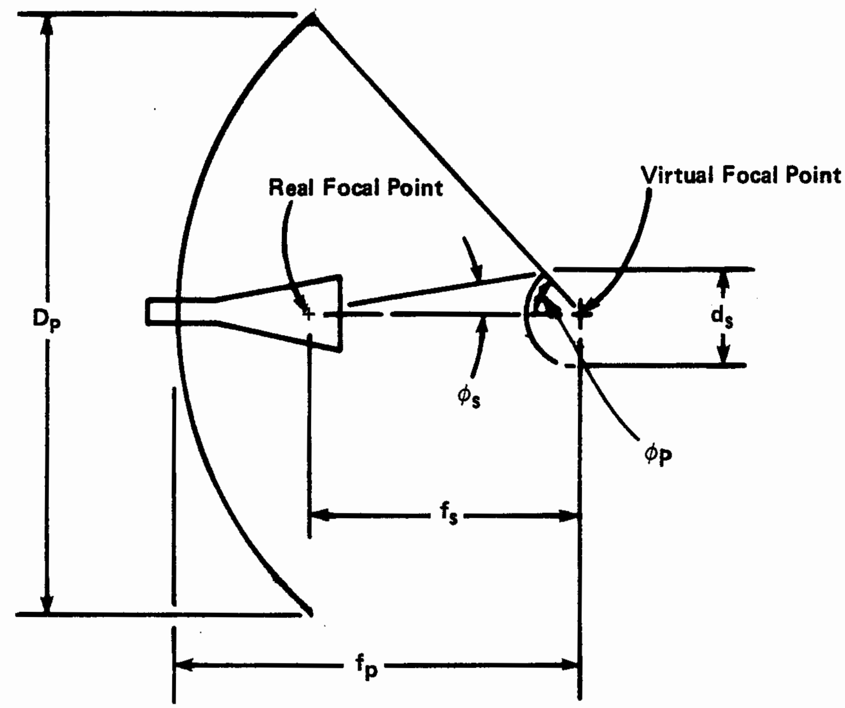


Figure 3. Geometry of the Cassegrain Antenna System

b. A geometry consisting of a paraboloidal main reflector and a special-shaped, quasi-hyperboloidal subreflector.[3,4] The geometry in Figure 3 is appropriate for describing this antenna. The main difference between the classical Cassegrain mentioned above is that the subreflector has been designed such that the overall efficiency of the antenna has been enhanced, thereby yielding improved gain performance. This technique is especially useful with antenna diameters of approximately 30 to 100 wavelengths; for example, a 5-meter antenna in the 6/4-GHz frequency band.

c. A generalization of the Cassegrain geometry consisting of a special-shaped, quasi-paraboloidal main reflector and a shaped, quasi-hyperboloidal subreflector.[5,6,7,8] Green[9] observed that in dual reflector systems with high magnification--essentially a large ratio of main reflector diameter to subreflector diameter--the distribution of energy (as a function of θ) is largely controlled by the subreflector curvature. The path length or phase front is dominated by the main reflector (see Figure 4). Kinber[5] and Galindo[6,7] found a method for simultaneously solving for the main reflector and subreflector shapes to obtain an exact solution for both the phase and amplitude distributions in the aperture of the main reflector of an axisymmetric, dual reflector antenna. Their technique, based on geometrical optics, was highly mathematical and involved solving two simultaneous, non-linear, first-order, ordinary, differential equations. These methods have been applied by Scientific-Atlanta and others to design axisymmetric, dual reflector antennas when maximum gain is needed for a given size reflector antenna. The method of solution allows the specification of an arbitrary amplitude distribution, whereby one can make compromises between sidelobes and antenna efficiency (see Figure 4).

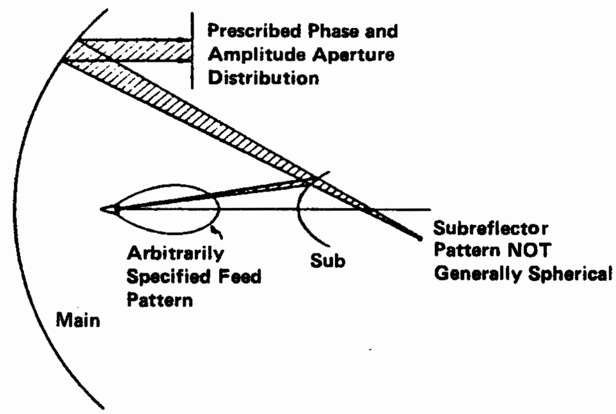


Figure 4. Circularly Symmetric Dual Shaped Reflectors

Prime-Focus-Fed Paraboloidal Antenna

The other type of antenna most often employed in the United States for a receive-only application is the prime-focus-fed paraboloidal (see Figure 5) reflector. For large aperture sizes, this type of antenna has excellent sidelobe performance in all angular regions except the spillover region around the edge of the reflector. Even in this area a sidelobe suppession can be achieved which satisfies FCC pattern requirement. The antenna efficiency for apertures greater than 100 wavelengths is in the 60% region; therefore, it represents a good compromise choice between sidelobes and gain.

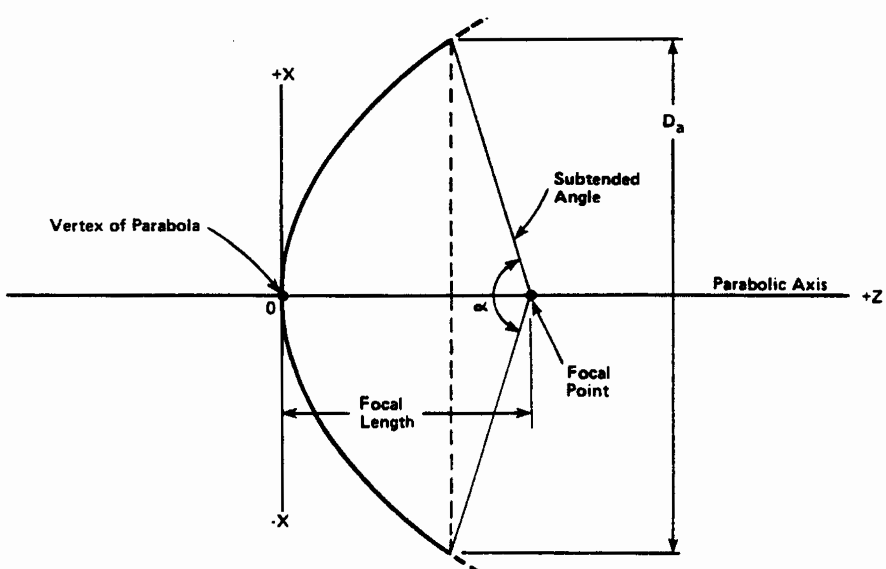


Figure 5. Geometry of Prime Focus Antenna

For aperture sizes less than approximately 40 wavelengths, the blockage of the feed and the feed support structure raises the sidelobes with respect to the peak of the main beam such that it becomes exceedingly difficult to meet the FCC sidelobe specification.

The CCIR sidelobe specification can be met since it contains a modifier which is dependent on the aperture size.

Horn Reflector Antenna

Two other types of single-beam antenna are used in earth stations. These are horn reflector and offset-fed reflectors. The horn reflector antenna can be grouped into two primary types, although other variations of the basic design concept exist:

Pyramidal Horn Reflector

The pyramidal horn reflector has been used for many years in the terrestrial microwave field, primarily by AT&T. This antenna offers excellent efficiency and sidelobe performance especially in the near-in region and in the back region. Its performance is a result of its closed configuration and lack of aperture blockage that is inherent in the axisymmetric antennas mentioned above. The polarization response off-axis is an area of concern especially when used with a frequency-reuse satellite.

Conical Horn Reflector

The conical horn reflector is similar in design to the rectangular horn reflector mentioned above. It has similar advantages and disadvantages, with some exceptions. The first sidelobe is improved due to the circular aperture. The conical horn also has less wind resistance than the pyramidal horn, again resulting from its basic shape. The pyramidal and conical horn reflector have somewhat lower noise temperature than axisymmetric reflector antennas. On the other hand, both also have unique mounting requirements which are somewhat restrictive and cumbersome. The overall length of the antenna, for example, is typically twice the aperture size. (See Figure 6 for the dimensions of a typical 4-meter conical horn reflector). Problems with transportation can also occur due to the large physical dimensions of the antenna.

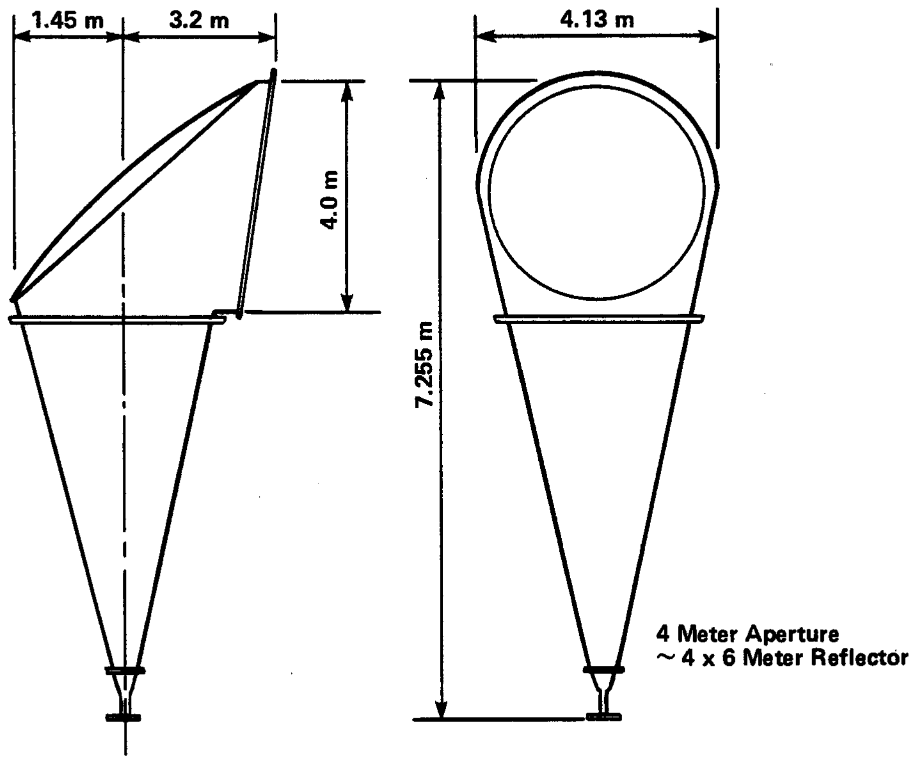


Figure 6. 4 Meter Conical Horn Reflector Antenna

Offset-Fed Reflector Antennas

To date, the offset-fed reflector antenna has not been widely used as an earth station antenna in the United States because of its cost. This probably will remain the case in the near future, but this antenna's sidelobe performance should make it a viable candidate for use at some time in the future. The offset, front-fed reflector antenna can employ a single main reflector or multiple reflectors, with two reflectors the more prevalent of the multiple reflector designs. The offset, front-fed reflector, consisting of a section of a paraboloidal surface (Figure 7), minimizes diffraction scattering by eliminating the aperture blockage of the feed and feed support structure. Sidelobe levels of (32-30 log θ) dBi can be expected from this type of antenna. Aperture efficiencies of 60-70% can also be expected. The increase in aperture efficiency as compared to an axisymmetric, prime-focus-fed antenna is due to the elimination of direct blockage. The geometry of this antenna type may present some unusual and interesting mount configurations. For a more detailed discussion of this type of antenna, see C. Mentzer[10].

Offset-fed dual reflector antennas exhibit sidelobe performance similar to that of the prime-focus-fed offset reflector. Two offset-fed dual reflector geometries have been used for earth station antennas: the double-offset geometry shown in Figure 8(a) and the open Cassegrain geometry introduced by Cook, et al[11] of Bell Labs shown in Figure 8(b). In the double-offset geometry, the feed is located below the main reflector, and no blocking of the optical path occurs. In contrast to this, the open Cassegrain geometry is such that the primary feed protrudes through the main reflector; thus it is not completely blockage-free. Nevertheless, both of these geometries have the capability of excellent sidelobe and efficiency performance.

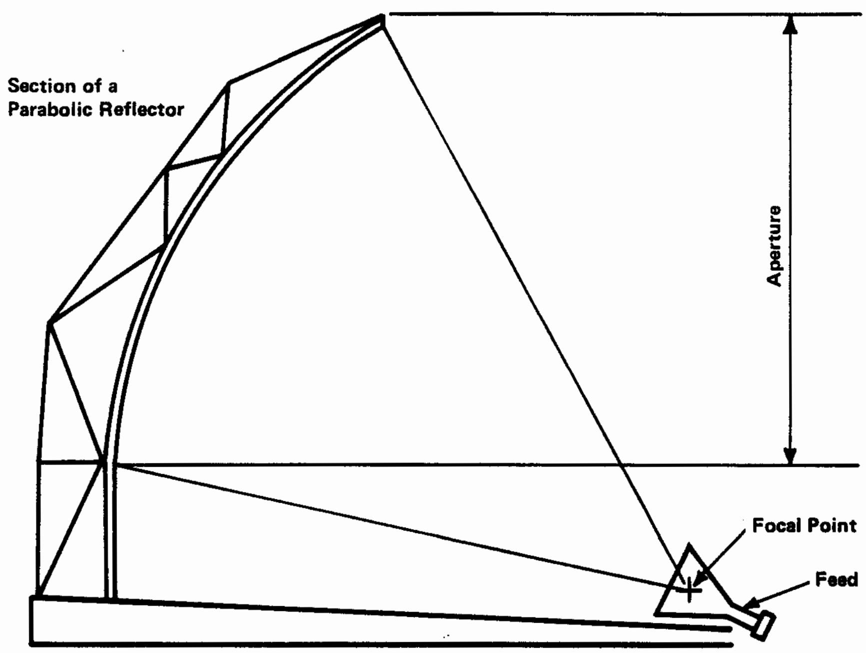
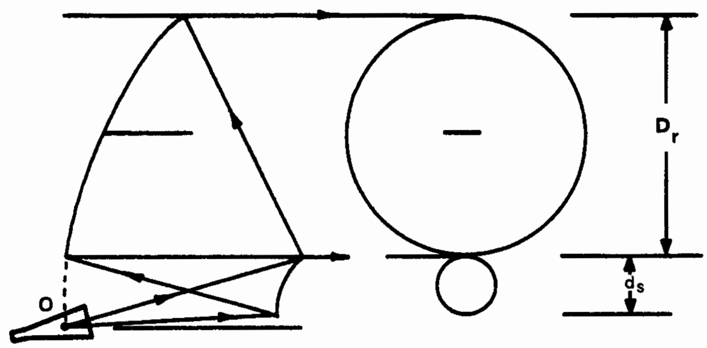
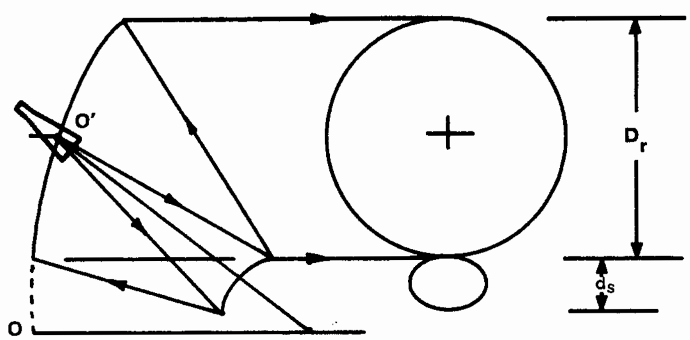


Figure 7. Basic Offset Fed Parabolic Antenna

The disadvantage of the offset geometry antennas is that they are asymmetric. This leads to increased costs, since there exists only one plane of symmetry, and the geometry also has some effect on the electrical performance. The offset geometry, when used for linear polarization, has a depolarizing effect on the primary feed radiation and produces two cross-polarized lobes within the main beam in the plane of symmetry. When used with circular polarization, the geometry introduces a small amount of beam squint whose direction is dependent upon the sense of polarization.



(a.) **Double-offset Geometry (Feed Phase Center and Paraboloidal Vertex at O)**



(b.) **Open Cassegrainian Geometry (Feed Phase Center located at O'; Paraboloidal Vertex at O)**

Figure 8. Offset Dual Reflector Geometries

During the past few years considerable analysis has been performed by Galindo-Israel, Mittra, and Cha[12] for this geometry for high aperture efficiency applications. Their analytical techniques are reported to result in efficiencies in the 80-90% range. Mathematically, the offset geometry, when formulated by the method used in designing axisymmetric dual reflectors results in a set of partial, differential equations for which there is no exact geometrical-optics solution. The method of Galindo-Israel, Mittra, and Cha is to solve the resultant partial differential equations as if they were total differential equations. Then, using the resultant subreflector surface, the main reflector surface is perturbed until a constant aperture phase is achieved.

Multiple Beam Antennas

During the past few years there has been an increasing interest in receiving signals simultaneously from several satellites with a single antenna. This interest has prompted the development of several multibeam antenna configurations which employ fixed reflectors and multiple feeds. The antenna engineering community, of course, has been investigating multibeam antennas for many years. In fact, in the middle of the seventeenth century Christian Huygens and Sir Isaac Newton first studied the spherical mirror, and the first use of spherical reflector as a microwave antenna occurred during World War II. More recently, the spherical reflector, the torus reflector, and the dual reflector geometries, all using multiple feeds, have been offered as antennas with simultaneous, multibeam capability. Chu[13] in 1969 addressed the mutiple-beam spherical reflector antenna for satellite communication; Hyde[14] introduced the multiple-beam torus antenna in 1970; Ohm[15] presented a novel multiple-beam Cassegrain geometry antenna in 1974. All three of these approaches are discussed below, as well as variations of scan techniques for the spherical reflector.

Spherical Reflector

The properties, practical applications, and aberrations of the spherical reflector are not new to microwave antenna designers. Its popularity is primarily due to the large angle through which the radiated beam can be scanned by translation and orientation of the primary feed. This wide-angle property results from the symmetry of the surface. Multiple beam operation is realized by placing multiple feeds along the focal surface. In the conventional use of the reflector surface, the minimum angular separation between adjacent beams is determined by the feed aperture size. The maximum number of beams is determined by the percentage of the total sphere covered by the reflector. In the alternate configuration described below, these are basically determined by the f/D ratio and by the allowable degradation in the radiation pattern.

In the conventional use of the spherical reflector, the individual feed illuminates a portion of the reflector surface such that a beam is formed coincident to the axis of the feed. The conventional multibeam geometry is shown in Figure 9. All of the beams have similar radiation patterns and gains, although there is degradation in performance in comparison to that of a paraboloid. The advantage of this antenna is that the reflector area illuminated by the individual feeds overlaps, reducing the surface area for a given number of beams in comparison with individual single beams antennas.

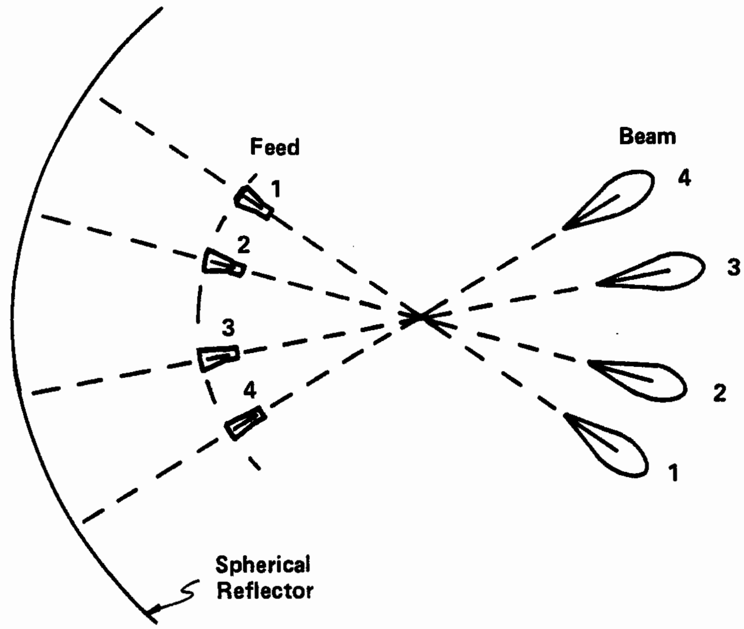


Figure 9. Conventional Spherical Multibeam Antenna Using Extended Reflector and Multiple Feeds

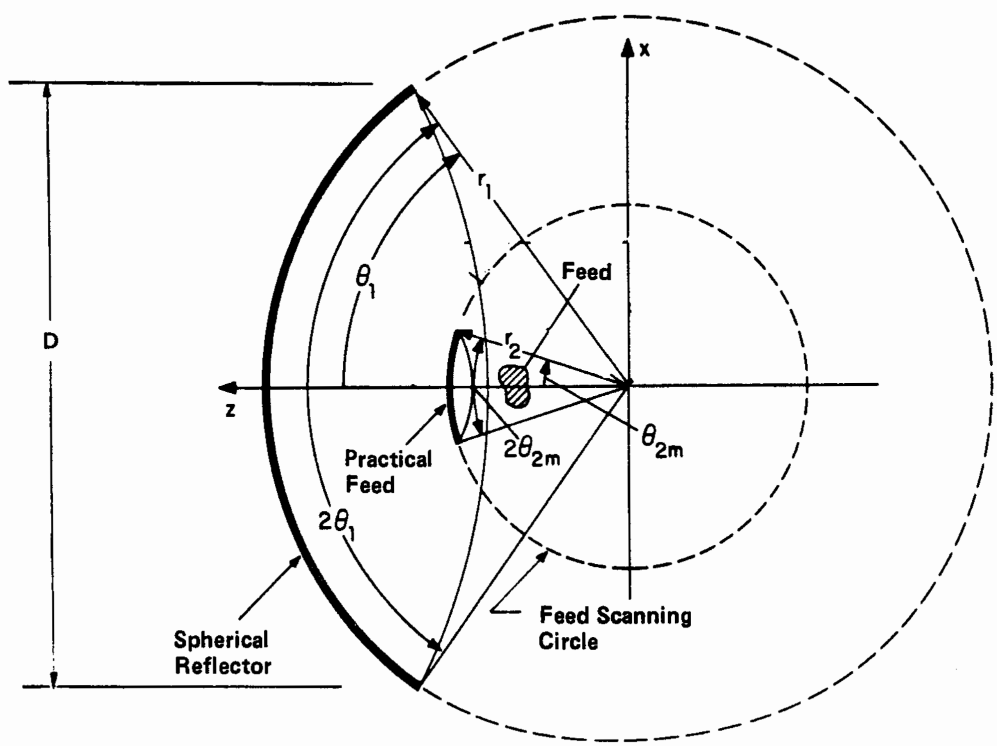


Figure 10. Convention Spherical Reflector and Feed Geometry

The alternate multibeam spherical reflector geometry is shown in Figure 11. For this geometry each of the feed elements points toward the center of the reflector with the beam steering accomplished by the feed position. This method of beam generation leads to considerable increase in aberration, including coma; therefore, the radiation patterns of the off-axis beams are degraded with respect to the on-axis beam. This approach does not take advantage of the sperical reflector properties that exist in the conventional approach. In fact, somewhat similar results could be achieved with a paraboloidal reflector with a large f/D. An example of the pattern distortion and gain loss as the feed is scanned off the reflector symmetry axis is shown in Figure 12 for a paraboloid reflector antenna[16] with an f/D of 0.4. The reflector diameter was 2.75 m operating at a frequency 7.9 GHz. The on-axis beam has a half-power beamwidth of 1° and first sidelobe of 25 dB. The coma effect on sidelobe and the loss of gain vs. scanning is given in Table 6.

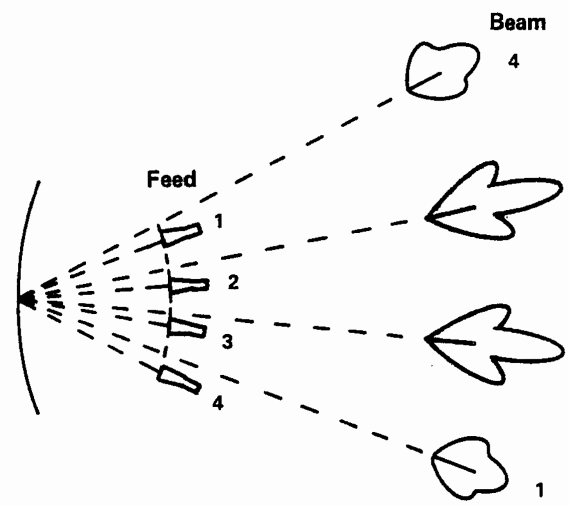


Figure 11. Alternate Spherical Multibeam Antenna Using Minimum Reflector Aperture with Scanned Beam Feeds

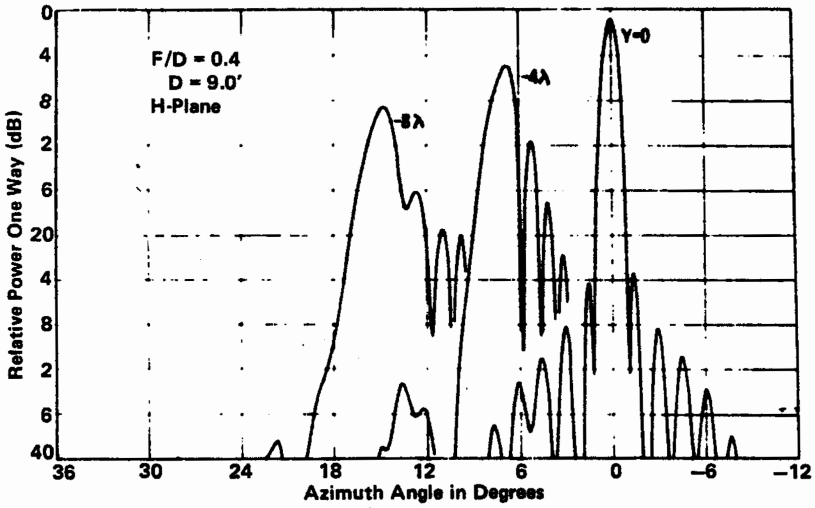


Figure 12. Measured Secondary Patterns as Function of Lateral Primary Feed Displacement for a Paraboloidal Reflector (from Reference 16)

Table 6. Gain Loss and First Sidelobe Level as a Function of Lateral Primary Feed Displacement for a Paraboloidal Reflector

Beam Scan (degrees)	Gain Loss (dB) Relative to On-Axis Beam	First Sidelobe (dB) Relative to Peak of Main Beam
0°	0.0	25.0
7°	4.+	7.0
14°	7.+	10.0
21°	11.+	2.0

An important factor to consider for this antenna is that the gain loss due to scanning is more dependent upon the actual angular scan than on the scan measured in half-power beamwidths.

Torus Antenna

The torus antenna is a dual curvature reflector, capable of multibeam operation by feeding it with mutiple feeds similar to those of the conventional spherical reflector geometry. The feed scan plane can be inclined to be in the orbital arc plane, allowing the use of a fixed reflector to view geosynchronous satellites. The reflector has a circular contour in the scan plane and a parabolic contour in the orthogonal plane (see Figure 13). It can be fed either in an axisymmetric or an offset-fed configuration. The offset geometry for use as an earth station antenna has been sucessfully demonstrated by COMSAT LABS[17]. The radiation patterns meet the (32-25 log θ) dBi envelope.

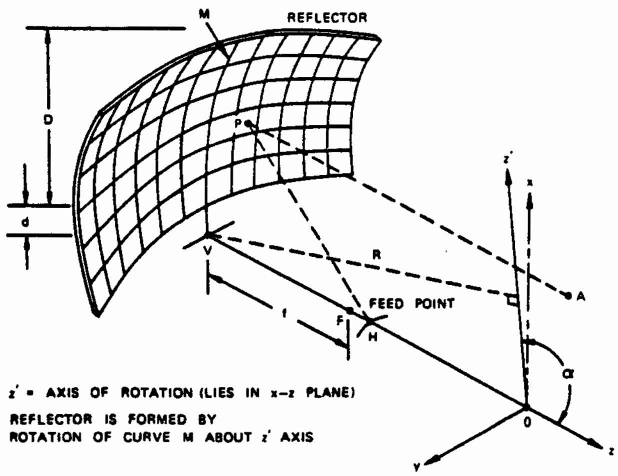


Figure 13. Torus Antenna Geometry. Comsat Technical Review Vol. 4, No. 2, Fall 1974

The offset-fed geometry results in an unblocked aperture, which gives rise to low wide-angle sidelobes as well as provides convenient access to the multiple feeds.

The torus antenna has less phase aberration than the spherical reflector because of the focusing in the parabolic plane. Because of the circular symmetry, feeds placed anywhere on the feed arc form identical beams. Therefore, no performance degradation is incurred when multiple beams are placed on the focal arc. Point focus feeds may be used to feed the torus up to aperture diameters of approximately 200 wavelengths. For larger apertures it is recommended that aberration correcting feeds be used.

The scanning or multibeam operation of a torus requires an oversized aperture to accommodate the scanning. For example, a reflector surface area of approximately 214 square meters will allow a field of view (i.e., orbital arc) of 30° with a gain of approximately 50.5 dB at 4 GHz (equivalent to the gain of a 9.65-meter reflector antenna). This surface area is equivalent to approximately three 9.65-meter antennas. Therefore, for a field of view of 30°, the aperture of three 9.65-meter antennas allows an operator the capability of receiving signals from eight satellites spaced at 4° intervals.

The main disadvantage of the antenna is in its stringent installation requirements. The mounting structure has to be customized for the site latitude and longitude, as the reflector plane of scan has to be accurately located relative to the orbital plane. The mutiple feeds also require a secondary structure or building to support them and their related electronics.

Offset-Fed, Multibeam Cassegrain Antenna

The offset-fed, multibeam Cassegrain antenna is comprised of a parabolic main reflector, a hyperbolic oversized subreflector, and mutiple feeds located along the scan axis, as shown in Figure 14. The offset geometry essentially eliminates beam blockage, thus allowing a significant reduction in sidelobes and antenna noise temperature. The Cassegrain feed system is compact and has a large focal-length-to-diameter ratio (F/D) which reduces aberrations to an acceptable level, even when the beam is moderately far off-axis. The low sidelobe performance is achieved by using a corrugated feed horn which produces a Gaussian beam.

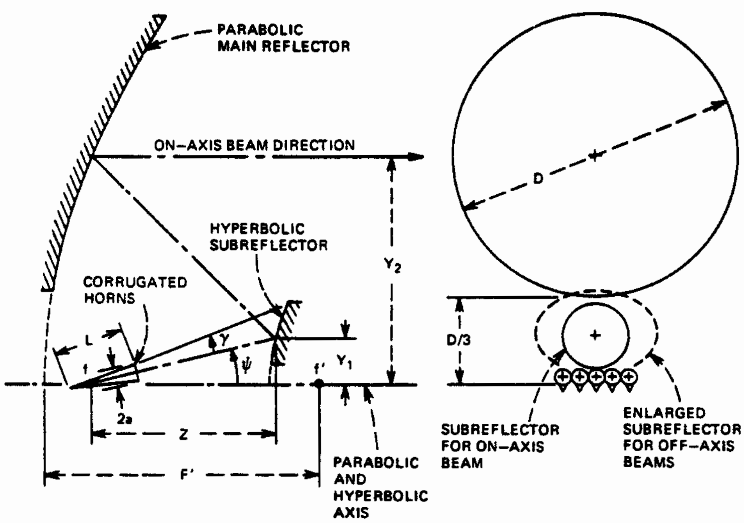


Figure 14. Geometry of the Offset-Fed, Multibeam Cassegrain Antenna. The Bell System Technical Journal, October 1974

A typical antenna design consisting of a 10-meter projected aperture would yield half-power beam widths and gain commensurate with an axisymmetric 10-meter antenna, 0.5° HPBW and 51-dB gain at 4 GHz. The subreflector would need to be approximately a 3 x 4.5-meter elliptical aperture. The feed apertures would be approximately 0.5 meter in diameter. The minimum beam separation would be less than 2°, more than sufficient to allow use with synchronous satellites with orbit spacings of 3° or greater. For the ±15° scan the gain degradation would be approximately 1 dB, and the first sidelobe would be approximately 20-25 dB below the main beam peak. The wide-angle sidelobes would meet the (32-25 log θ) dBi envelope. Figure 15 is a model of the antenna configuration.

The reflector and subreflector require separate structures. This necessitates care in the alignment on site, since plane orientation is dependent on the longitude and latitude of the site.

Typical Multibeam Orbit Coverage

For a multibeam antenna to receive signals from the existing U.S. domestic satellites, a field of view of 30° is necessary. Within the orbital arc of 86° W to 136° W are eleven domestic satellites operating in the 6/4-GHz frequency band. These include the three Canadian ANIK satellites. Figure 16 shows the azimuth and elevation look angles to these satellites from a typical site in the Southeastern United States. For the multibeam configurations, the mutiple feeds would have to be placed on a scan axis such that the resultant beams are pointed to these azimuth and elevation coordinates. It is obvious that the feeds would necessarily be mounted at considerable height above the ground to traverse the required arc for the spherical reflector. For the torus antenna the offset reflector and the feed would be tilted with respect to the earth such that the feeds can traverse the required arc. The offset Cassegrain's reflector, subreflector and feed scan arc would also be tilted with respect to the earth to provide the desired arc coverage.

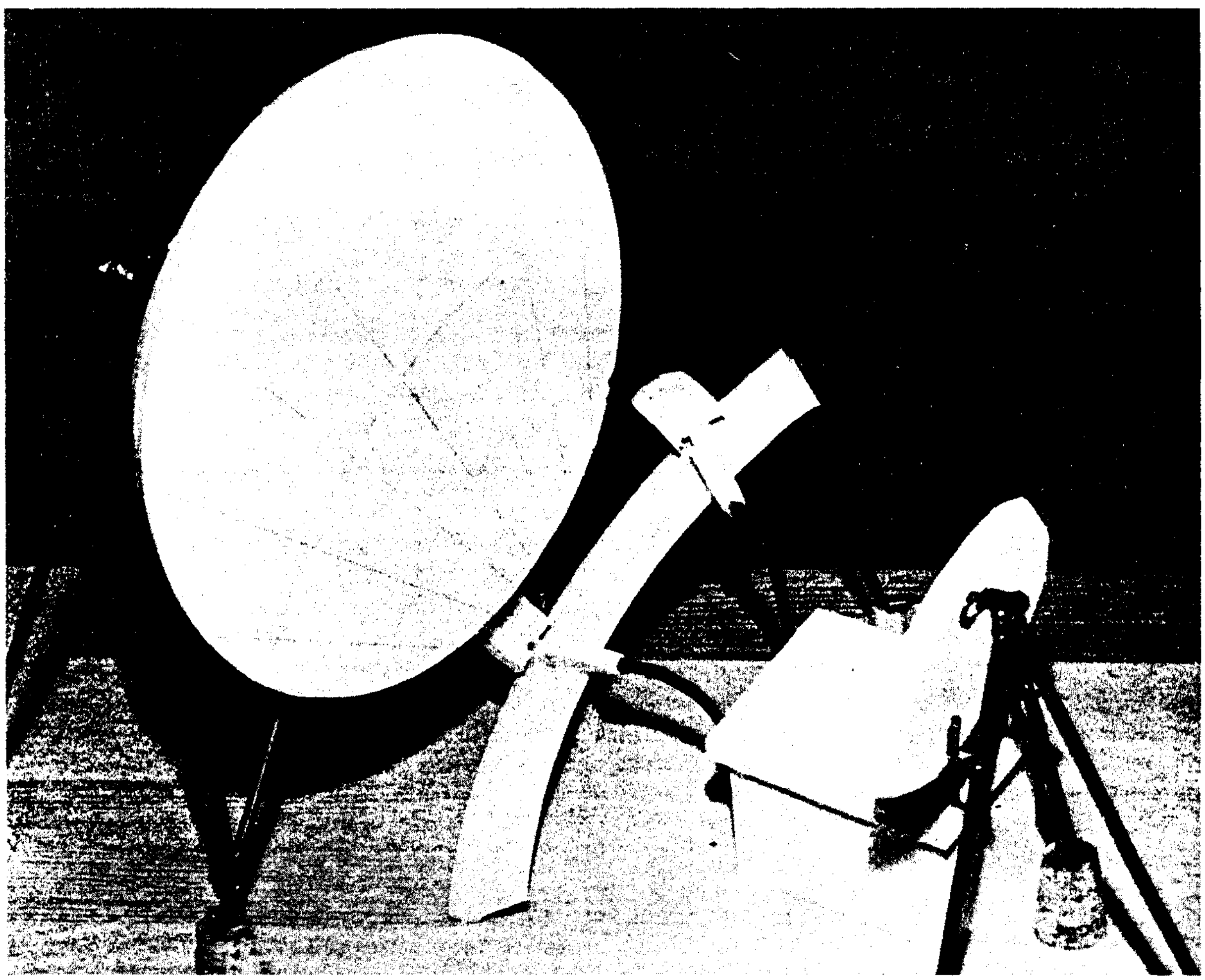


Figure 15. Model of Offset-Fed, Multibeam Cassegrain Antenna Configuration

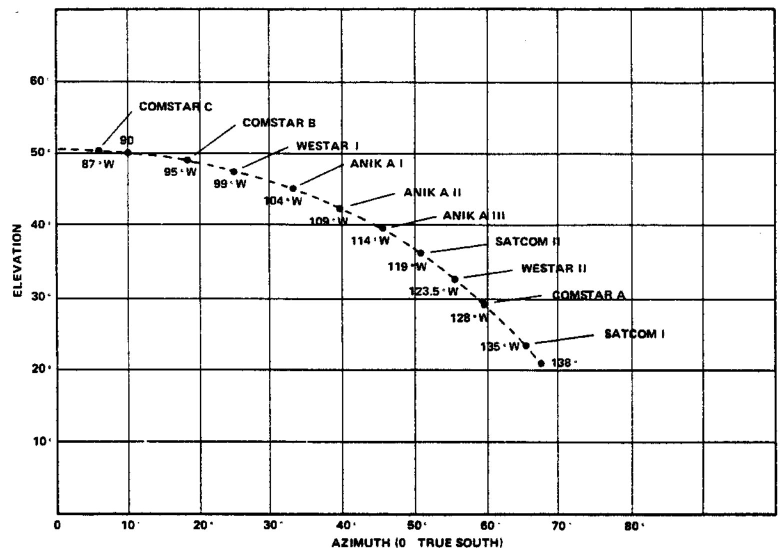


Figure 16. Orbital Arc as Viewed from Typical Site in Southeast 90°W to 136°W Longitude

Angle Tracking

Automatic tracking of satellite position is required for many earth station antennas. Monopulse, conical scan, sequential lobing, single-channel monopulse and steptrack techniques are applicable for this purpose. The primary advantages and principal use of the steptrack, conical scan, single-channel monopulse and three-channel monopulse configuration are given in Figures 17 through 20. The most common of these techniques for small- and medium-aperture earth stations are the steptrack and single-channel (pseudo-monopulse) monopulse.

Steptrack (Antenna or Feed)

- Primary Use
  - Geosynchronous satellites

- Principal Advantages
  - Low cost
  - Requires only a single beam feed
  - Requires one RF channel

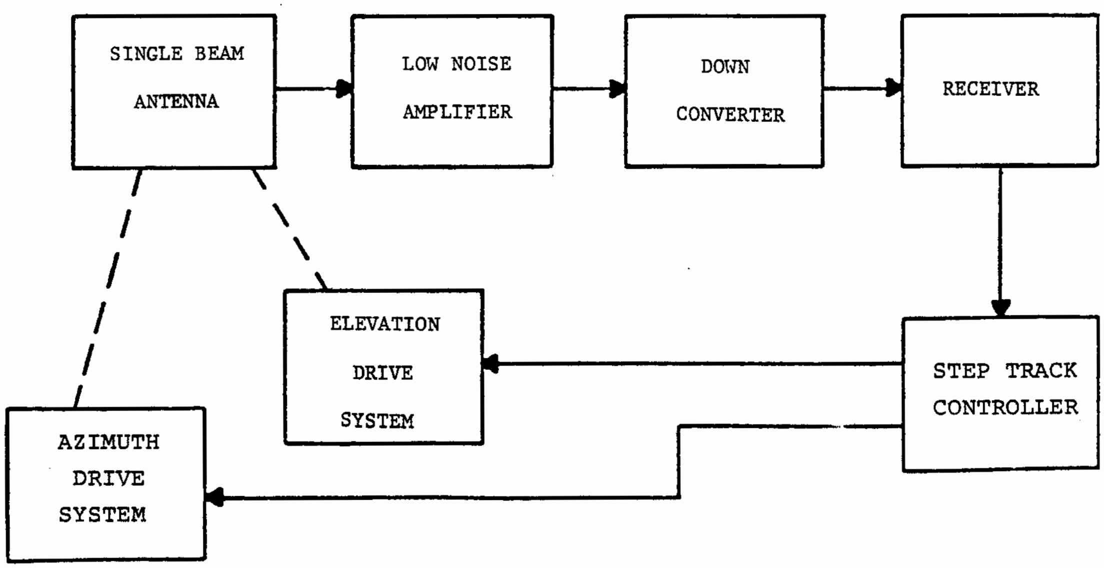


Figure 17. Steptrack (Antenna or Feed)

Conical Scanning

- Primary Use
  - Polar or geosynchronous satellites
  - Low or high dynamic targets

- Principal Advantages
  - Low cost
  - Uses a single beam feed
  - Uses one RF channel

- Principal Disadvantages
  - Requires four pulses or short duration continuous signal to obtain tracking information (typically 33 ms)
  - Subject to mechanical reliability problems due to continuous feed rotation
  - Subject to tracking loss due to propeller and other low rate modulation sources

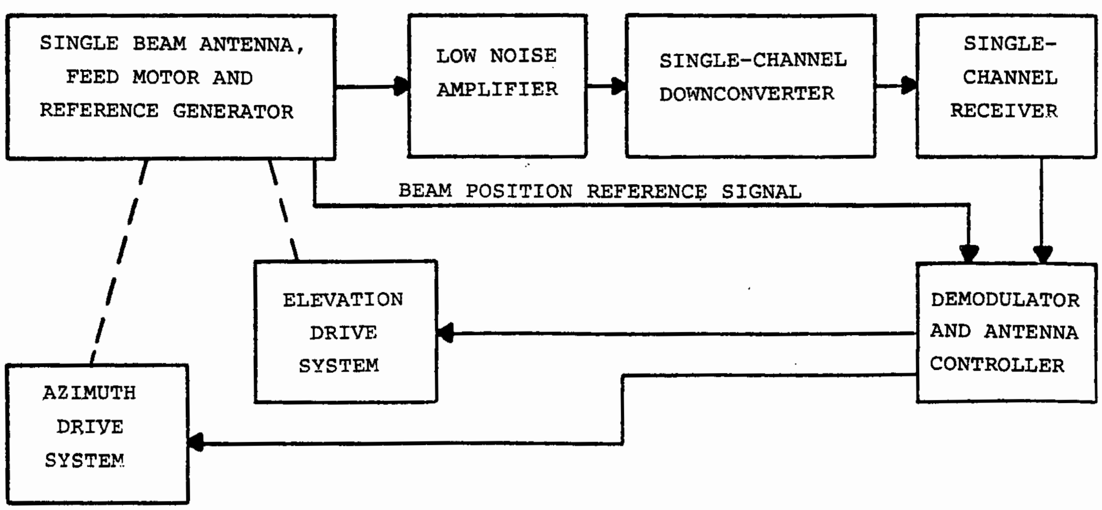


Figure 18. Conical Scanning

Single-Channel Monopulse

- Primary Use
  - Polar or geosynchronous satellites
  - Low or high dynamic targets

- Principal Advantages
  - Use only one RF channel
  - Lower cost than three-channel monopulse
  - Has low angular tracking jitter
  - Uses fast, random rate electronic beam steering to prevent tracking loss due to propeller and other low-rate modulation sources

- Principal Disadvantages
  - Requires four pulses or short duration continuous signal to obtain tracking information (approximately 1.5 ms)
  - Requires three beam feed

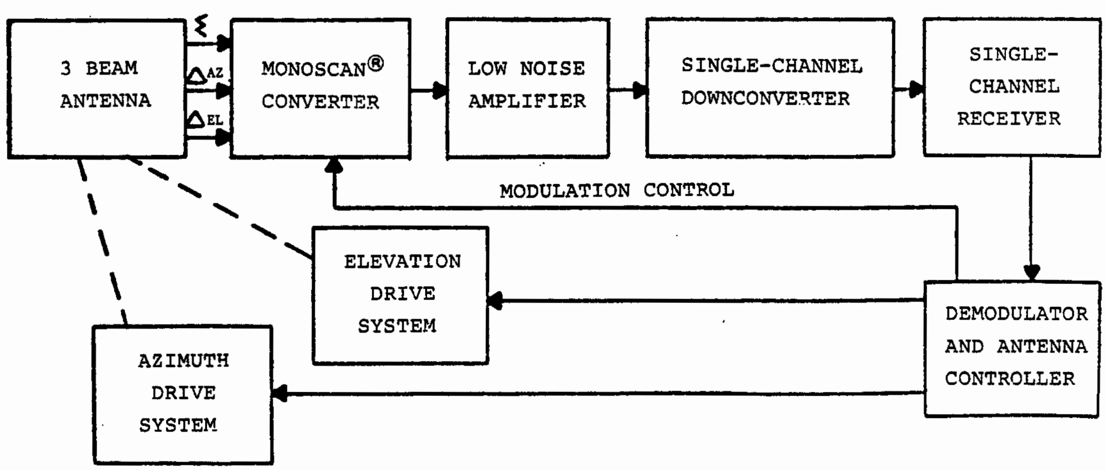


Figure 19. Single-Channel Monopulse

Three-Channel Monopulse

- **Primary Use**
  - Orbiting or geosynchronous satellites
  - Radar tracking
  - Low or high dynamic targets such as aircraft or missiles

- **Principal Advantages**
  - Provides instantaneous angle information
  - Can track targets with rapidly fluctuating signal levels
  - Has very low angular track jitter

- **Principal Disadvantages**
  - Requires three phase-matched channels from the antenna to the receiver
  - Highest cost system approach
  - Requires three beam feed

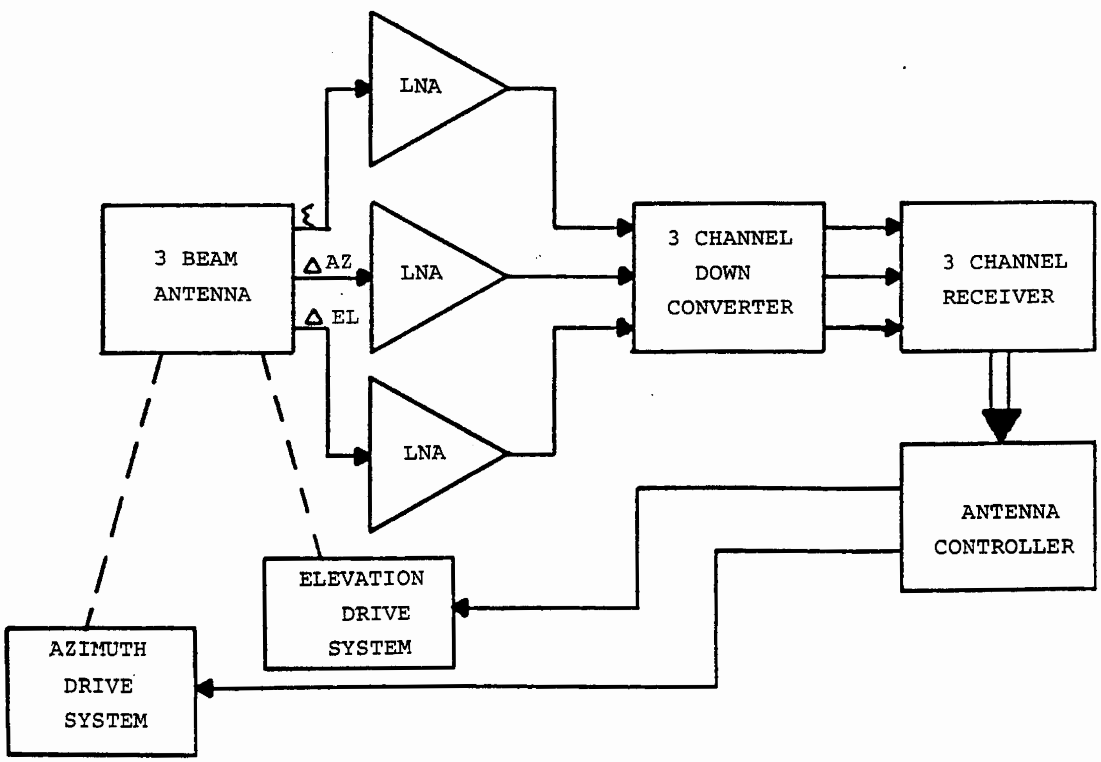


Figure 20. Three-Channel Monopulse

Steptrack

Steptracking is a technique employed primarily for maintaining the pointing of an earth station's antenna beam toward a geosynchronous satellite. Since most geosynchronous orbiting satellites have some small angular box of stationkeeping, a simple peaking technique can be used to keep the earth station beam correctly pointed.

The signal peaking or steptrack routine is a software technique that maneuvers the antenna toward the signal peak by following a path along the steepest signal strength gradient or by using some other algorithm that accomplishes the same result [18,19]. Any method of direct search for maximization of signal is applicable; one such method calculates the local gradient by determining the signal strength at three points (A, B, C) (Figure 21.) From the signal strength at points A, B, C, an angle α_2 is computed representing the unit's gradient vector angle relative to the azimuth axis. Once C_1 is determined, A_2 is set equal to C_1, and another angle θ_2 us defined by $\theta_2 = \theta_1 + \alpha_2$ ($\theta_1 = 0$). The angle is used to relate pedestal movements to a fixed pedestal coordinate system. The procedure is repeated a number of times until the antenna boresight crosses throughout the peak. At this time the step size is reduced by one-half and for each time the peak is crossed, it is again reduced by one-half until the peaking resolution is accomplished.

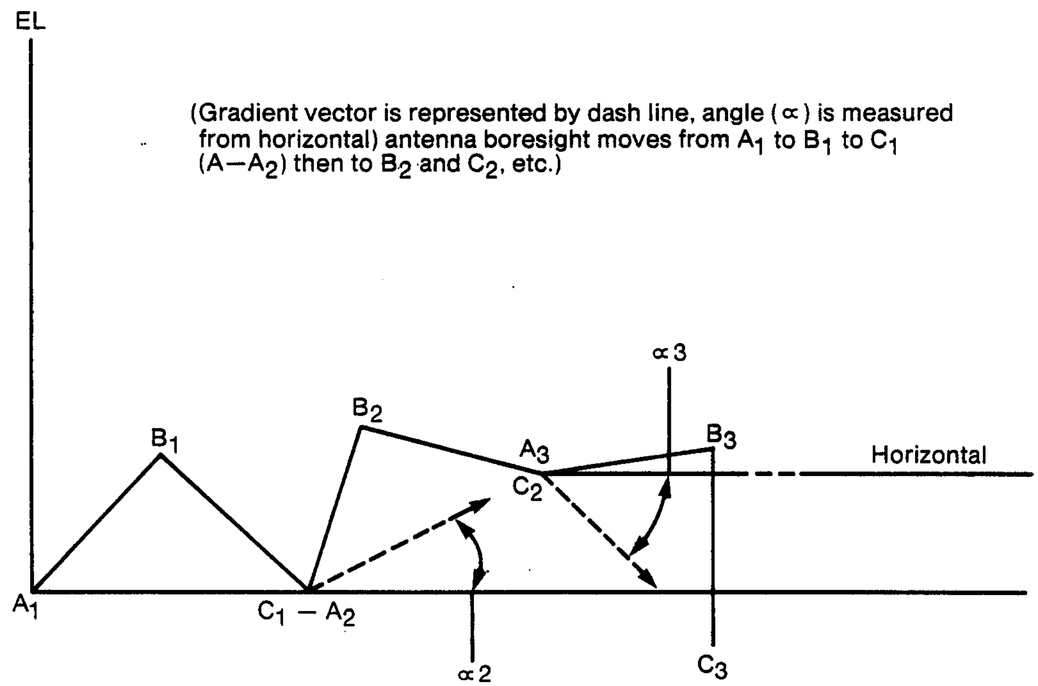


Figure 21. Signal Strength Gradients

Single-Channel Monopulse

This technique utilizes multiple elements or modes to generate a reference signal, an elevation error signal, and an azimuth error signal. The two error signals are then combined in a time-shared manner by a switching network which selects one of two-phase conditions (0° and 180°) for the error signal. The error signal is then combined with the reference signal, with the resultant signal then containing angle information. This allows the use of a single receiver for the tracking channel. This technique is equivalent to sequential lobing where the lobing is done electrically and can be adjusted to any desired fixed or variable scan rate.

Polarization

Many satellite communication systems are dual polarized (frequency reuse), and are therefore susceptible to interference from the intended cross-polarized signals when the medium of propagation is such that the ellipticity of the signals are changed. This condition can occur during atmospheric conditons such as rain and Faraday rotation in the ionosphere. Therefore, in order to maintain sufficient polarization discrimination, special precautions must be taken in regard to polarization purity of the dual-polarization antennas; indeed, adaptive polarization correcting circuits may be necessary.[20,21]

Several types of polarziation discrimination enhancement schemes may be used. They include:

1. Simple rotation of the polarization major ellipse axis to correct for rotation due to the ionosphere (applicable for linear polarizations). Transmit and receive rotations are in opposite directions.

2. Sub-optimal correction of depolarization, ellipticity correction with respect to one of the signals, but not orthogonalizing the two signals (differential phase).

3. Complete adaptive correction including orthogonalization (differential phase and amplitude).

The polarization enhancement schemes may be implemented at the RF frequencies or they may follow a downconversion stage and be implemented at IF frequencies. In either case the circuitry must operate over the full bandwidth of the communication channel. Kreutel, et al[22] treat the implementation of RF frequencies in detail. The reader is also referred to Gianatasio[23,24] for both RF and IF circuitry and experiments to verify performance of the circuitry.

Summary

Many kinds of antennas are used as earth station antennas. For overall performance the prime-focus-fed paraboloidal and the dual-reflector Cassegrain antennas have been the predominant types of antenna used. Their choice has been based on the best tradeoffs between electrical and mechanical performance commensurate with cost. It is expected that these types of antennas will continue to be the earth station antennas for many years to come, although some increase in the use of offset-reflector geometries is expected. Interest in multiple-beam antennas will increase, but it will tempered by the increased cost and installation complexity. The recent FCC assignments of orbital arc positions between 70° W. and 143° W. longitude should also encourage multiple antenna installations for a given site.

References

1. P.W. Hannah, "Microwave Antennas Derived from the Cassegrain Telescope", IRE Transactions on Antennas and Propagation, March 1961, pp. 140-153.

2. P.A. Jensen, "Designing Cassegrain Antennas", Microwave, December, 1962, pp. 10-16.

3. G.W. Collins, "Shaping of Subreflectors in Cassegrainian Antennas for Maximum Aperture Efficiency", IEEE Trans. Ant. & Prop., Vol. AP-21, No. 3, May 1973, pp. 309-313.

4. S. Parekh, "On the Solution of Best Fit Paraboloid as Applied to Shaped Dual Reflector Antennas", IEEE Trans. Ant. & Prop., Vol. AP-28, No. 4, July 1980, pp. 560-562.

5. B. Ye Kinber, "On Two Reflector Antennas", Radio Eng. Electron, Physic., Vol. 6, June 1962.

6. V. Galindo, "Design of Dual Reflector Antenna with Arbitrary Phase and Amplitude Distributions", PTGAP Int. Symp., Boulder, Colo., July 1963.

7. V. Galindo, "Synthesis of Dual Reflector Antennas", Elec. Research Lab., Univ. of Calif. Rept. No. 64-22, July 30, 1964.

8. W.F. Williams, "High Efficiency Antenna Reflector", Microwave Journal, July 1965, pp. 79-82.

9. K.A. Green, "Modified Cassegrain Antenna for Arbitrary Aperture Illumination", IEEE Trans. Ant. & Prop., Vol. AP-11, No. 5, September 1963.

10. C.A. Mentzer, "Analysis and Design of High Beam Efficiency Aperture Antennas", Doctoral Thesis 74-24, 370, Ohio State Univ., 1974.

11. J.S. Cook, E.M. Elam, H. Zucker, "The Open Cassegrain Antenna: Part I. Electromagnetic Design and Analysis", The Bell System Tech. Jour., Sept. 1965, pp. 1255-1300.

12. V. Galindo-Israel, R. Mittra, A. Cha, "Aperture Amplitude and Phase Control of Offset Dual Reflectors", USNC/URSI and IEEE Int. Symp. on Ant. & Prop., May 1978.

13. T.S. Chu, "A Multibeam Spherical Reflector Antenna", IEEE Antennas and Propagation Int. Symp., Program and Digest, Dec. 9, 1969, pp. 94-101.

14. G. Hyde, "A Novel Multiple-Beam Earth Terminal Antenna for Satellite Communication", Int. Conf. on Comm. Record, June 1970, Conf. Proc. Paper No. 70-CP-386-COM, pp. 38-24 to 38-33.

15. E.A. Ohm, " A Proposed Multiple-Beam Microwave Antenna for Earth Stations and Satellites", Bell Sys. Tech. J., Vol. 53, pp. 1657-1665, Oct. 1974.

16. W.A. Imbriale, P.G. Ingerson, and W.C. Wong, "Large Lateral Feed Displacements in a Parabolic Reflector", IEEE Trans. Ant. & Prop., Vol. AP-22, pp. 742-746, Nov. 1974.

17. G. Hyde, R.W. Kreutel, L.V. Smith, "The Unattended Earth Terminal Multiple-Beam Torus Antenna", COMSAT Technical Review, Vol. 4, No. 2, Fall 1974, pp. 231-264.

18. M.J.D. Powell, "An Efficient Method for Finding the Minimum of a Function of Several Variables without Calculating Derivatives," Computer Journal, Vol. 7, 1964, pp 155.

19. J. Kowalik and M.R. Osborne, Methods for Unconstrained Optimization Problems, American Elsevier Publishing Co., 1968.

20. T.S. Chu, "Restoring the Orthogonality of Two Polarizations in Radio Communication Systems, I and II," BSTJ, Vol. 50, No. 9, November 1971, pp. 3063-3069, and Vol. 52, No. 3, March 1973, pp. 319-327.

21. R.W. Kreutel, D.F. DiFonzo, W.J. English, R.W. Gruner," Antenna Technology for Frequency Reuse Satellite Communications," Proc. IEEE, Vol. 65, No. 3, March 1977, pp. 370-378.

22. A.J. Gianatasio, "Broadband Adaptively-Controlled Polarization-Separation Network," 7th European Microwave Conference Proceedings, Copenhagen, Sweden, September 1977, pp. 317-321.

23. A.J. Gianatasio, "Adaptive Polarization Separation Experiments," Final Report, NASA CR-145076, November, 1976.

Digital Audio for Satellite Distributed Network Radio — System Overview

Introduction

Scientific-Atlanta recently signed contracts with ABC, CBS, and NBC as the supplier of 3-meter digital earth terminals for their network radio stations. The resulting terminal is a time-division-multiplexed (TDM) digital system with a transmission rate of 8.78 Mb/s. This system has been tested and qualified with rooftop reception in downtown Manhattan. All digital audio testing was performed directly over the satellite in the presence of severe terrestrial interference. Excellent performance was recorded on all tests. It is forecast that by early 1984, over 3,000 of these 3-meter terminals will be installed throughout the nation, receiving high-quality program feeds from the major radio networks.

Scientific-Atlanta's 3-meter TDM digital audio terminal is designed to receive high-speed TDM digital data at 8.78 Mb/s, demodulate, decode, and then demultiplex the data into the desired audio and data channels. The terminals will support data rates equivalent to twenty 15 kHz audio channels (384 kb/s each) or an equivalent larger number of lower bandwidth channels. Each 384 kb/s channel can support either one 15 kHz program, two 7.5 kHz programs, twelve 32 kb/s auxiliary channels (voice cue or data) or one 7.5 kHz program channel, and six 32 kb/s auxiliary channels. One 32 kb/s auxiliary channel slot is reserved for system synchronization.

Using this TDM terminal, all radio stations have access simultaneously to any of the twenty 15 kHz channels. This allows the station to receive their time zone network feed, news services, data, voice cue, as well as other program material. This simultaneous access is obtained by adding plug-in cards to the mainframe equipment.

Digital transmission was selected instead of analog transmission for its data and channel use flexibility, the outstanding quality of the received program audio and the efficient usage of the satellite transponder. To obtain analog-transmitted high-quality audio into a 3-meter network requires the use of analog compression techniques. Unlike digital compression, a residue compression is retained in the expanded analog audio. Digital transmission provides expanded dynamic range (about 15 dB more than analog) and high signal-to-noise ratio of a very high level that has never been experienced in network reception.

Two types of analog radio transmission are presently used for network transmission--the single channel per carrier (SCPC) and the video subcarrier or diplex method. For quality audio reception into a 3-meter network, only the SCPC provides acceptable performance for a large number of audio channels.

Analog earth terminals are less expensive than digital, but their satellite transmission capacity is reduced. For high quality reception only about ten (10) channels of 15 kHz SCPC analog transmission can be used in the satellite 3-meter network reception. Sixteen is probably the maximum number which could be supported. TDM digital transmission, on the other hand, allows twenty high-quality 15 kHz channels in the satellite.

Other competing digital approaches such as the 5-carrier T1-rate transmission were usually more expensive for receiving full network service. For example, to simultaneously access digital audio or data channels in two T1-rate transmitted carriers requires additional RF and demodulating equipment. To access data simultaneously in all five T1-rate carriers, the terminal cost will be prohibitive. Using the TDM approach allows access simultaneously to all audio and digital data on the full transponder. The TDM approach is also more immune to terrestrial interference than is the T1-rate approach.

Scientific-Atlanta's TDM digital earth terminal provides:

- Efficient use of the satellite
- Inexpensive full transponder service
- Very high quality audio reception
- Built-in expansion capability for future services
- One-way data service capability
- Relative immunity to terrestrial interference

Scientific-Atlanta will manufacture and sell the terminals and will offer installation and 24-hour maintenance service.

Equipment Description

The 15 kHz program signal is sampled at 32 kilosamples per second and digitized to a 15-bit word. Digital commanding techniques are used to instantaneously compress the 15-bit word to an 11-bit word. A parity bit is added, resulting in a word length of 12 bits. The parity bit is used in error concealment encoding that allows bit error rate to degrade to 10^{-5} with "just perceptible" audio degradation in the 15 kHz audio channel unit. The noise performance of the terminal is enhanced by forward-error-correction (FEC) encoding. The combined data is bi-phase modulated (BPSK) at 70 MHz, then upconverted to the 6 GHz band for transmission to the satellite. Figure 1 is a block diagram of the 3-meter receive-only terminal.

The outside equipment for the basic terminal consists of Scientific-Atlanta's 3-meter antenna, a 120 Kelvin low-noise amplifier (LNA), and 200 feet of 1/2-inch foam-filled coaxial cable. This outside equipment configuration will be satisfactory for 96% of the radio stations located within the Continental USA. Larger antennas or lower temperature LNAs are recommended for radio stations in Florida, Southern Texas and some areas in the upper New England States.

The inside equipment is contained in two chassis that are mounted in the radio stations' equipment rack. The wideband receiver converts 4 GHz received transmission down to 70 MHz where it is bi-phase demodulated, the forward error correction (FEC) coding is decoded, and the resulting 7.68 Mb/s data stream is supplied to the data processing unit (DPU) mainframe.

The 7.68 Mb/s data is demultiplexed in the DPU chassis where the full transponder service is available with plug-in cards. A summary specification of the terminal is provided in Table 1.

The optional plug-in units are as follows:

Dual 15 kHz channel unit --------Two independent 15 kHz channels on one card

Dual 7.5 kHz channel unit -------Two independent 7.5 kHz channels on one card

Voice cue channel unit ----------A 32 kb/s delta modulated (CVSD) 3 kHz voice unit on one card

Data channel unit ---------------Three-port addressable asynchronous for slow and medium rate terminal equipment

A three-transponder select switch will also be offered as optional equipment. Scientific-Atlanta will also offer a full line of outside equipment including lower temperature LNAs, larger antennas, and a low sidelobe antenna for improved performance for small orbital satellite spacing now being considered.

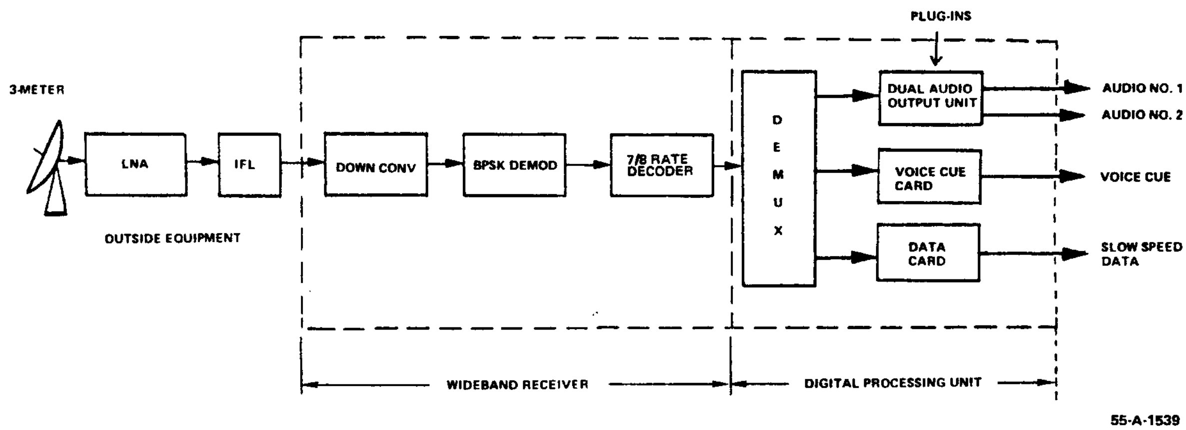


Figure 1. Three-Meter Digital Audio Terminal

Table 1. 3-Meter Digital Audio Terminal Specification Summary

Item	Specifications
Antenna	
Diameter	3 meters
Gain	39.5 dB
Low-Noise Amplifier	
Noise Temperature	$120^{o}K$
Gain	50 dB
Wideband BPSK Receiver	
Downconversion	Dual
Received Data Rate	8.78 Mb/s
FEC Decoding	Rate 7/8 hard decision threshold decoding
Bit Error Rate	Better than 10^{-7} on or above 33 dBW satellite power contours
Digital Audio (15-kHz unit)	
Sample Frequency	32 kHz
Digital Compression	15 to 11
Sine wave Overload Level	+28 dBm0*
Peak Unaffected Signal	+24 dBm0
Signal-to-Quantizing Noise Ratio	≥56 dB (at +22 dBm)
Idle Channel Noise	≥81 dB (below +24 dBm)
Total Harmonic Distortion	0.3%

* 0 dBm0 = 0 dBm into 600 ohms

Receive-Only Terminal Link Performance

Performance Analysis

In this section, the RF performance of Scientific-Atlanta's 3-meter TDM digital earth terminal is reviewed. For all practical purposes, the effect of bit errors on the digital audio performance is eliminated completely whenever the bit error rate (BER) is equal to or better than 10^{-7}.

The next paragraph reviews the location of radio stations, illustrating that a majority of these stations are located on or above the 33 dBW power contour of the satellite. Link fade margin is addressed, followed by a tabulation of the link calculations. The last paragraph discusses the effects of bit errors on digital audio performance.

Satellite EIRP Contours vs. Terminal Location

Performance of the receive-only earth station depends on the earth station equipment configuration and the satellite power or EIRP contours. The majority of the end users for digital audio will be the existing radio stations, and fortunately most of these radio stations are located on the higher EIRP contours of the satellite. Table 2 shows a count of the top 300 Areas of Dominant Influence (ADI) for radio market as to their location in the various contours of a particular satellite.

Table 2. Number of Terminals Located in EIRP Contours for a 600 Terminal Network

EIRP (dBW):	36-35	35-34	34-33	33-32	32-31	TOTAL
No. of Terminals:	92	254	228	16	10	600

These estimates of terminal count versus satellite power contours are based on the EIRP contours of SATCOM F1.

Link Fade Margin

The margins required for a satellite link are substantially different than for a terrestrial microwave system. The margins required for a satellite link depend on system availability and must include allocations (margins) for:

a. Atmospheric absorption

b. Rain fades

c. Antenna pointing errors (including wind loading effects)

d. Transmitter power level variations

One major advantage of a system which uses a saturating carrier in the transponder is that uplink fades can be neglected, since the transponder is operated beyond saturation by the expected amount of the uplink fade. Thus, this guarantees that during maximum uplink fades, the transponder does not come out of saturation. The downlink fade margins at 4 GHz required for a BER availability of 99.99% are shown in Table 3.

Table 3. Link Margin for 99.99% BER Availability

Parameter	Margin (dB)	Note
Atmospheric Absorption	0.1	a
Rain Fades	0.6	b
Antenna Pointing	0.1	c
Transmit Power Variation	0.0	d
	0.8 dB	

NOTES:

a. Atmospheric absorption: 0.1 dB.

b. Rain fades: This is a function of rain rate, earth station elevation angle, and vertical depth of rain cells. The baseline contour (32 dBW) for the RCA system (cutting through the Southern U.S.) is at higher elevation angles (resulting in lower fades) while the better EIRP contours are at lower elevation angles so that the increased rain fade is compensated by the higher available EIRP. For 99.99% of the time, the downlink rain fades (heavy rain over 60mm/hr) will be under 0.6 dB (note that a typical rain cell might extend 30 km or more in length, but only about 2 km in depth). A satellite link will experience a fade corresponding to the 2 km depth times the sin(x) of the elevation angle x, whereas a microwave system would experience a fade corresponding to the entire horizontal extent of the rain cell.

c. Antenna pointing errors: This is dependent on the antenna aperture and is well under 0.1 dB for 3-meter antennas.

d. Transmit power level variations: For saturated transponder, this is negligible since it only affects uplink losses.

Interference

The interference into the receive-only digital audio terminal is a composite of interference from various sources. The interference model used in the calculations is shown in Table 4.

The top two C/I ratios in Table 4 are typical values while the C/I ratios for adjacent satellite and terrestrial interference are obtained from analysis.

Table 4. TDM Digital Interference Model

Source	C/I (dB)
Uplink	31
Cross Polarization	33
Adjacent Satellite	18.1
Terrestrial	30
Composite C/I*	17.5

*Composite C/I calculation: $10^{-1.75} = 10^{-3.0} + 10^{-3.3} + 10^{-1.81} + 10^{-3.0}$

Terrestrial Interference

Scientific-Atlanta employs a notch filter designed as part of its bandpass-matched filter which reduces the effect of terrestrial interference while enhancing the desired signal detection.

The frequency response on an ideal matched filter for 9 Mb/s BPSK modulated signal obeys the $(\sin x/x)^2$ response with its first nulls at ±9 MHz. This ideal filter is approximated by a filter with nulls at ±10 MHz, which significantly reduces the effect of most terrestrial interference while enhancing the detection of the desired signal.

Scientific-Atlanta's 3-meter TDM digital terminal has been tested in severe terrestrial interference environment. Figures 2 and 3 are photos of frquency spectrum measured during the ABC-Radio test with a 3-meter rooftop antenna at 1926 Broadway, New York, N.Y. The top photo shows the wideband modulated BPSK signal between the two TD-2 terrestrial interfering signals (arrows). The interfering signal was about the same level as the unmodulated satellite carrier (Figure 3), or about -123 dBW. Excellent reception was obtained during the test in the presence of this interference.

The notched matched filter provides a protection of about 38 dB from two equal-level TD-2 levels of -116 dBW, which provides a terrestrial C/I ratio of 30 dB.

Link Calculations

The link calculations for Scientific-Atlanta's 3-meter digital audio terminal are illustrated in Table 5. Direct addition of the positive and negative numbers in the top part of the table establishes the ideal downlink E_b/N_o. Power adding the composite C/I ratio results in the effective E_b/N_o. Subtracting link implementation margin of 1.5 dB results in the realized E_b/N_o into the rate 7/8 threshold decoder. The last table entries are the margins above BER = 10^{-7} for the various satellite contours. In all cases, adequate margin is available for excellent digital audio reception.

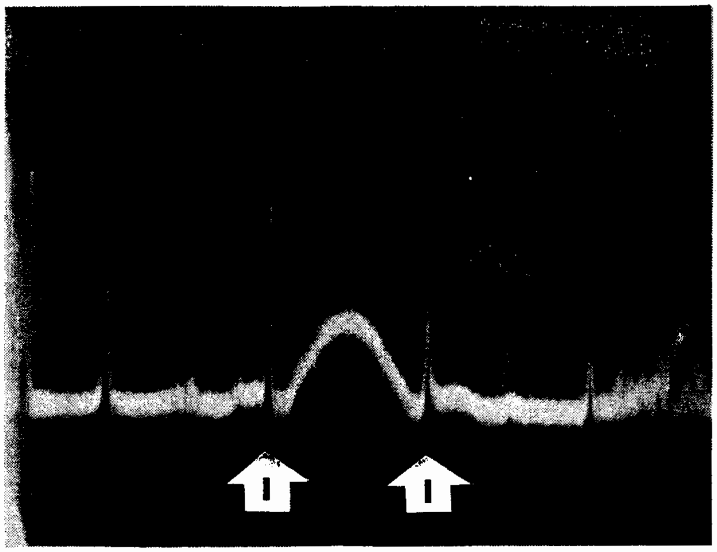


Figure 2. **Wideband BPSK Signal Received by 3-Meter Rooftop Antenna at 1926 Broadway, New York, NY. Terrestrial Interfering Signal at ±10 MHz from Transponder Center**

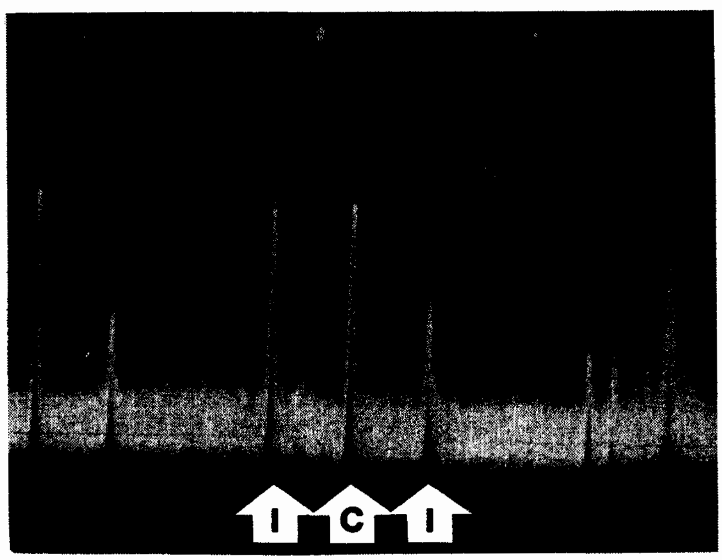


Figure 3. **Center Spike is Unmodulated Carrier Received from Satellite into 3-Meter Rooftop Antenna**

Table 5. Three-Meter Network Link Calculation

Parameters	Values			Notes
Satellite Saturated EIRP (dBW)	32.0	33.0	34.0	
Multiple Carrier Loss (dBW)	0.0	0.0	0.0	
Space Loss (dB)	-196.8	-196.8	-196.8	
Boltzmann's Constant (dB)	+228.6	+228.6	+228.6	
Information Rate (7.68 Mb/s)	- 68.9	- 68.9	- 68.9	
3-Meter, 120° LNA G/T	+ 17.5	+ 17.5	+ 17.5	(1)
Ideal Downlink E_b/N_o (dB)	12.4	13.4	14.4	
Fade Margin (99.99%) (dB)	- 0.8	- 0.8	- 0.8	(2)
Downlink E_b/N_o (dB)	11.6	12.6	13.6	(5)
Composite C/I (dB) (4.0° orbital spacing)	17.5	17.5	17.5	(3)
Effective E_b/N_o (dB)	10.6	11.4	12.1	
Link Implementation Loss (dB)	- 1.5	- 1.5	- 1.5	(4)
FEC Decoder E_b/N_o	9.1	9.9	10.6	
Decoder E_b/N_o for BER = 10^{-7}	8.1	8.1	8.1	(5)
Extra Margin above BER = 10^{-7}	1.0	1.8	2.5	

NOTES:

(1) The G/T varies with the elevation angle of the antenna: G/T = 17 dB/K at 5° elevation up to 17.9 dB/K at 40° elevation. An average value of 17.5 dB/K is used in the calculations.

(2) 99.99% fade margin discussed in link fade margin section above.

(3) Composite C/I discussed in interference section above. For 3° orbital spacing C/I reduces to 15 dB, which reduces the FEC decoder E_b/N_o by 0.6, 0.8, and 0.9 dB on the 32, 33, and 34 dBW contours respectively.

(4) Typical measured performance over simulated satellite link using Scientific-Atlanta's BPSK modulator and demodulator is better than 1 dB from theory.

(5) The decoder is rate 7/8 1-bit hard decision threshold decoder. Value is measured value.

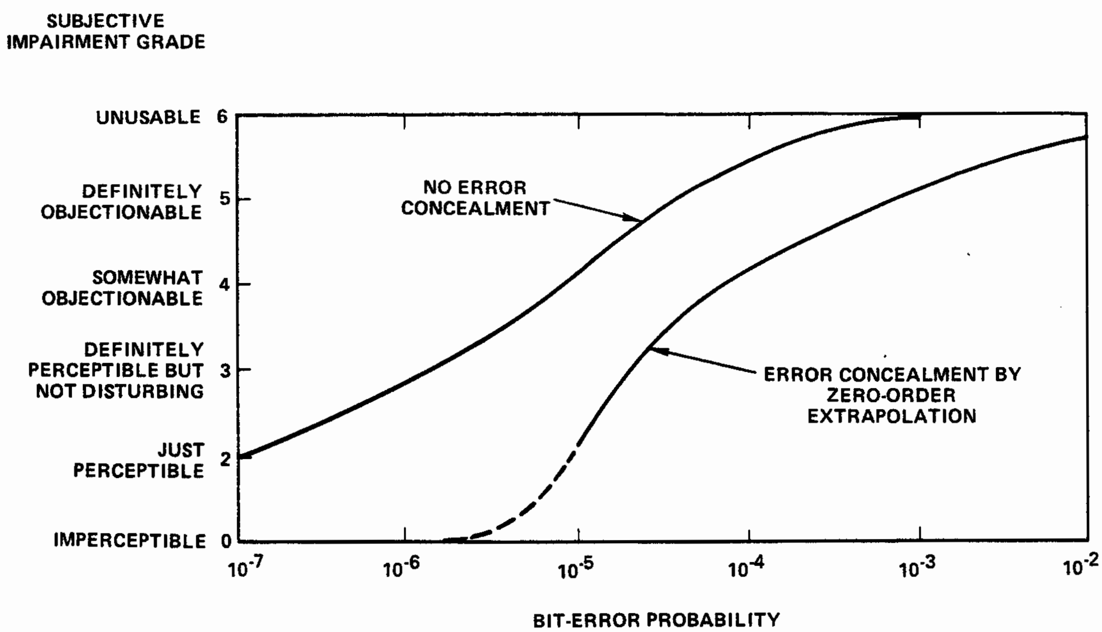
If additional margin is required at some stations, it can be obtained by increasing the terminal G/T. Typical G/T increases obtained by modifying the outside equipment are listed in Table 6.

Table 6. Methods of Increasing G/T

Modification Antenna Dia.	LNA (Kelvin)	Typical G/T Increase (dB)
3.0 M	120	-0-
3.0 M	100	0.5
3.0 M	90	0.9
3.0 M	80	1.2
3.6 M	120	1.0
3.6 M	100	1.5
3.6 M	90	1.9
3.6 M	80	2.2
4.6 M	120	3.7

Interpretation of the Effect of BER on Digital Audio Performance

The bit error rate (BER) is the average rate at which bit errors occur; it is the estimate of the bit error probability. The BER remains constant whether we consider the high-speed 7.68 Mb/s data stream or a single 15 kHz channel (384 kb/s) demultiplexed from this high-speed data stream. Scientific-Atlanta uses a single bit error concealment encoding to further enhance the performance of the digital audio terminal. The subjective performance of this terminal is illustrated in Figure 4 on a subjective impairment grade. Since the RF link is designed to operate at 10^{-7} error rate or better, the effects of bit errors will not degrade the performance.



SUBJECTIVE EFFECT OF BIT ERRORS IN PRESENCE OF PROGRAM (SOLO TRUMPET)

([1]) J. R. CHEW AND M.E.B. MOFFAT, "PULSE-CODE" MODULATION FOR HIGH-QUALITY SOUND-SIGNAL DISTRIBUTION: PROTECTION AGAINST DIGIT ERRORS," BRITISH BROADCASTING CORPORATION REPORT NO. 1972/18 UDC 534.86-626376.56

Figure 4. Error Concealment Performance

Theory of Digital Communications

Dr. James S. Gray

Why Digital Communications

There is an ever increasing number of digital sources that need to communicate with each other. Computer mainframes need to talk to computer mainframes, files need to be transferred from one memory bank to another memory bank. Word processors, "dumb" terminals, "smart" terminals, and other types of business machines all need to talk with other business machines. With the advent of electronic mail, documents can be transmitted using digital facsimile. This will significantly change the way information is routed around the nation in the future.

There are also advantages to the digitization of analog sources. For example, if we digitize audio, then one can have a mixed voice data network. Toll quality voice can be achieved at 32 kb/s using CVSD (Continuous Variable Slope Delta Modulation) at the present time. Some laboratories are claiming toll quality voice at 16 kb/s, and various companies are working on acceptable quality voice at 9.6 kb/s. Another area of considerable research effort at the present time is digital video. One application is video conferencing by corporations to eliminate the high cost of individuals traveling to meet with each other. Various companies are working on techniques such that acceptable video with motion is achieved at a T-1 data rate. With the ever increasing cost of fuel related to the OPEC situation, it is logical to assume that with time this is going to become the mechanism by which different organizations or different groups within a given organization communicate with each other. Broadcast quality video can be transmitted at the present time at 22 Mb/s using the proper source encoding equipment. The problem, however, is that this equipment is too expensive. If one could achieve low-cost source encoding equipment, then one could transmit two video signals over one transponder with an aggregate data rate of 44 Mb/s. This allows enough excess bandwidth to add sophisticated error correction coding. One should use error correction coding because redundancy reduction techniques make the effects of bit errors more important. Thus, with the proper source encoding equipment, one could alleviate the current transponder shortage problem by transmitting two video programs over one transponder, as opposed to the one presently being transmitted using FM techniques.

Another advantage of using digital techniques to transmit analog information is that one one accepts the inherent distortion due to quantization, then by using the proper combination of modulation technique, forward error correction coding, power, bit rate, etc., one can guarantee a low enough error rate so that there is no more degradation in the signal over the initial quantization error. When one transmits an analog signal using analog techniques with repeaters, the signal-to-noise ratio continuously degrades. With the digital situation, one can keep the error rate low enough that the effect of going through multiple repeaters in a system does not degrade the overall quality of the transmitted signal. Finally, digital techniques allow one to easily encrypt the signal to maintain the privacy of information. Most analog encryption techniques are easily defeated and do not achieve much protection. Digital encryption techniques, however, are extremely powerful and can ensure the privacy required in business communication networks.

How Do We Transmit Digital Information Via Satellite

Consider an RF carrier A $\cos(\omega_c t + \emptyset)$. One can transmit digital information via the satellite by modulating the amplitude A of the RF carrier, by modulating the frequency ω_c of the carrier, or by modulating the phase $\emptyset$ of the carrier. Figure 1 shows a comparison of performance between modulating the amplitude, frequency, and phase of the carrier. The OOK curve represents gating the amplitude on and off the carrier. For example, a One is the carrier being gated on with the amplitude equal to A, and a Zero represents the carrier being gated off or A=0. One sees that this technique takes the most power for a given bit error rate. The second technique consists of modulating the frequency of the carrier so that a One corresponds to one frequency for the RF carrier, and a Zero corresponds to a second frequency. This technique is more efficient than On/Off Keying, but still not as efficient as modulating the phase. The third technique consists of actually modulating the phase of the carrier according to the data state. As shown, this is the most efficient of the three in terms of requiring the minimum energy for a given bit error rate. It should be noted that the normalized signal-to-noise used in comparing these techniques is a quantity called E_b/N_o. E_b is energy per bit, and N_o is the spectral density of the noise. Measuring performance versus E_b/N_o allows us to normalize the performance and make it independent of the given data rate. For those who are more used to thinking in a C/N sense, let's examine E_b/N_o. Energy per bit is power-times-time, so take carrier power C times bit period T to end up with CT/N_o. But CT/N_o can be written as C divided by No/T. However, T is equal to bit period, so 1/T is equal to R or the bit rate; therefore, one ends up with E_b/N_o equal to C/N_oR. But N_oR is the noise power in a bit rate bandwidth, so E_b/N_o is C/N where N is measured in a bit rate bandwidth. This expression is shown in Figure 1. The PSK curve shown in Figure 1 is equally valid for biphase modulation where we take each bit and modulate the carrier into two phase states, or by quadraphase modulation where we take every two bits or dibits and modulate the carrier into four phase states. The reason for this will be explained shortly. How do we modulate the phase of an Rf carrier? In Figure 2, the vector representation of this is also presented for one symbol period or bit period of the data stream. BPSK can be represented as $Y(t) = A\cos(\omega_c t + \emptyset(t))$; where $\emptyset(t)$ is zero for a One and $\emptyset(t)$ equals π for a Zero.

Another important representation is to rewrite this as $A(t)\cos(\omega_c t + \emptyset)$ where A(t) is +1 for a One and -1 for a Zero. Thus we can treat BPSK modulation as double-sideband suppressed-carrier AM-type modulation where one is mutiplying the RF carrier by a waveform that has a plus 1 normalized value for a zero-degree phase state and a minus-one value for a 180° phase state. Also shown in Figure 2 is the vector representation for quadraphase shift key modulation. As one can see, the 00 vector is zero degrees, the 01 vector is 90°, the 11 vector is 180°, and the 10 vector is 270°. One must judiciously pick phase states so that a 90° phase error only causes one bit error and not two bit errors. The time domain representation of the RF carrier for the different phase states is also shown. One can represent QPSK as $Y(t) = (\omega_c t + \emptyset(t))$ where $\emptyset(t)$ takes on the different phase values according to the expression shown. There are obviously other codings. This expression can be rewritten as two BPSK modulated quadrature carriers; that is, $Y(t) = a(t)\cos(\omega_c t + \emptyset) + b(t)\sin(\omega_c t + \emptyset)$ where a(t) is a ±1 normalized bit stream and b(t) is also a ±1 normalized bit stream. Thus, the odd bits in the original data stream are put into a(t) and the even bits into a b(t). As discussed previously, BPSK and QPSK have the same theoretical curve as shown

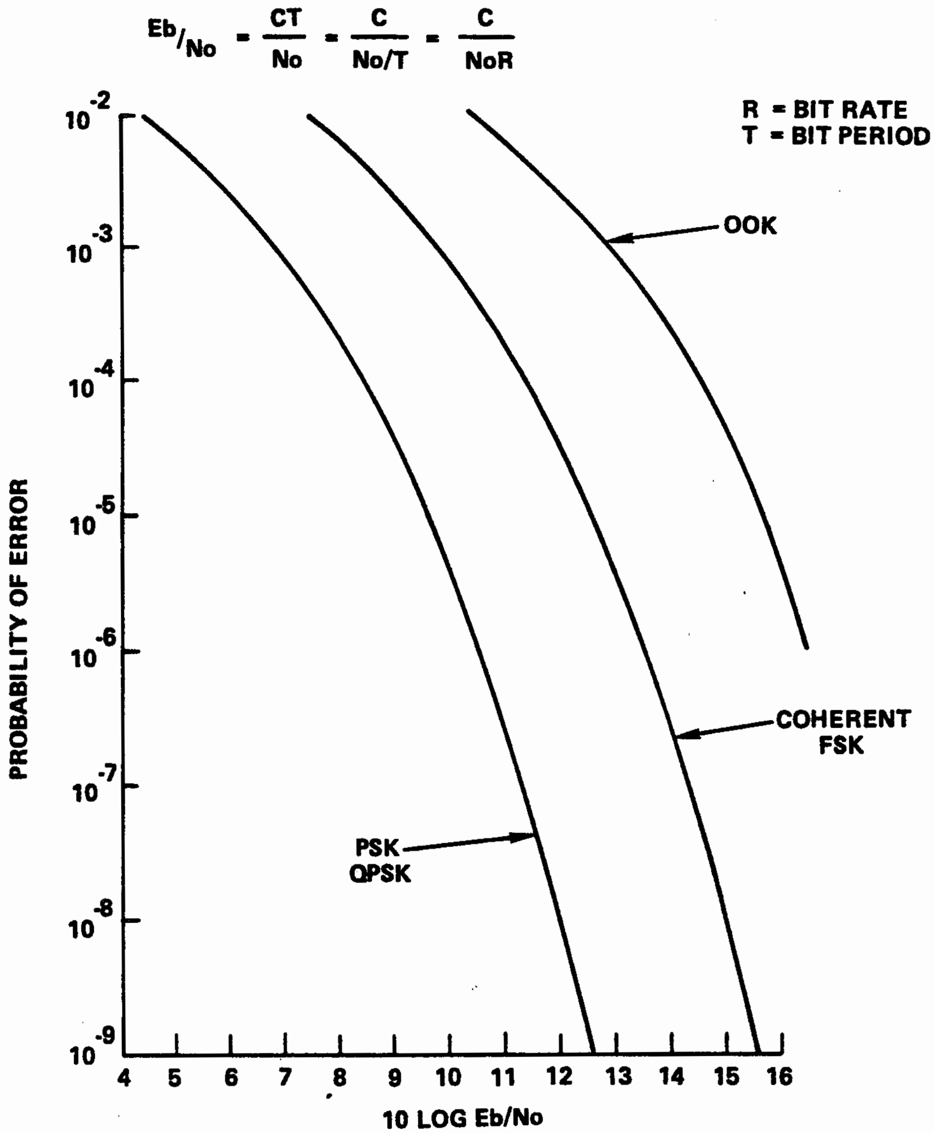


Figure 1. Comparison of Modulation Techniques

in Figure 3. The reason for this follows: One can think of QPSK as modulating two quadrature carriers at half the data rate; thus, we have used half the RF power for each data stream. However, the bit period for each data stream now is twice as long. Therefore, when one takes half the power times twice the bit period, one ends up with the same E_b/N_o on each carrier as one originally had for a BPSK carrier and thus can achieve an overall bit error rate for each quadrature carrier that is the same as the orginal BPSK bit error rate.

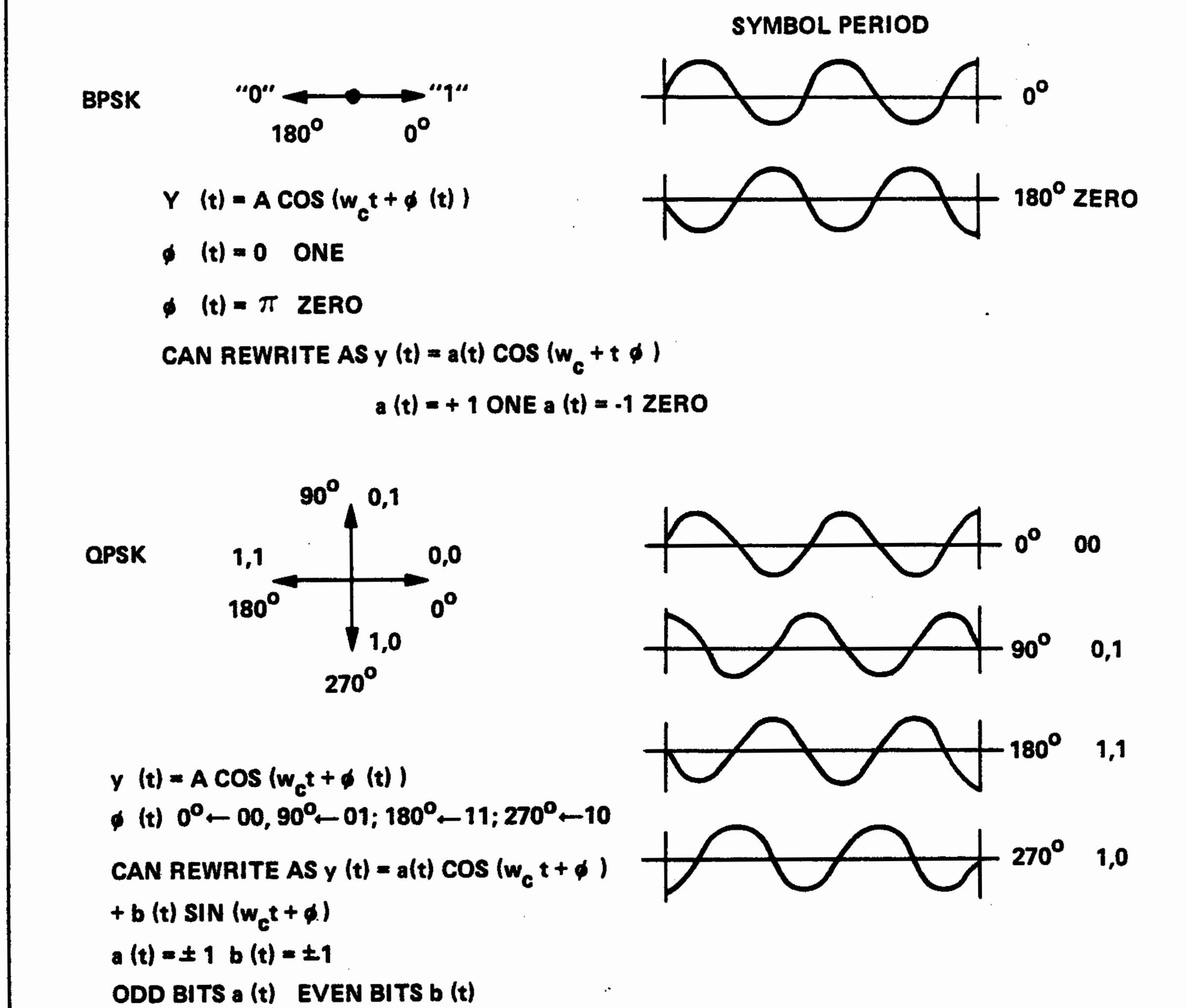


Figure 2. BPSK and QPSK Modulation

Transponder Limitations

Transponders are inherently limited in their bandwidth and their power. For example, a typical C-band transponder has a 36-MHz bandwidth. A typical saturated EIRP is about 36 dBW, and allowing a 4-dB output backoff for linear operation with multiple carriers results in effective EIRP of about 32 dBW. Thus, one is limited by the bandwidth used to transmit over the satellite and also by the power one can use. How can we combat both these effects? First, let us consider filtering. The spectrum for non-bandlimited NRZ (non-return to zero code) is shown in Figure 4. The spectrum roll-off has a $(\sin x)/x)^2$ characteristic. The nulls of the spectrum are at the symbol rate, where for BPSK the symbol rate is equal to the bit rate and for QPSK the symbol or baud

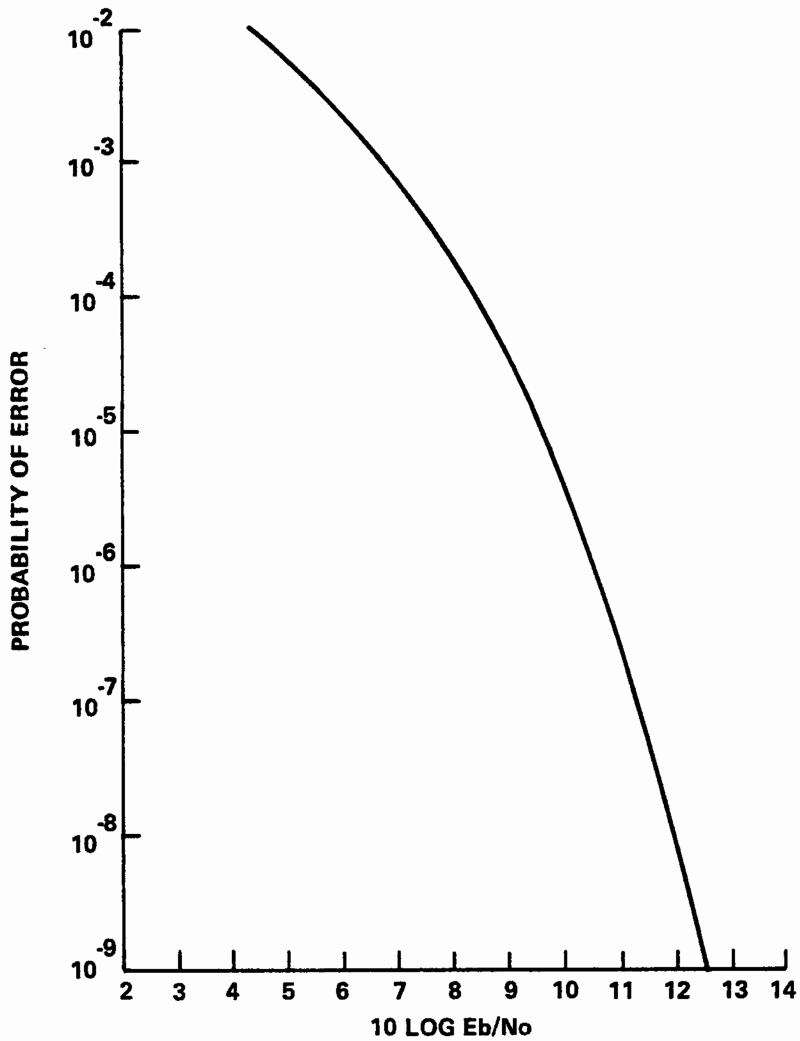


Figure 3. Pe vs. Eb/No for PSK and QPSK

rate (or the rate at which one changes modulation parameters) is at one-half the bit rate. This is because in QPSK one changes the phase of the four phase states every two bit periods of the original data stream. Let us take several QPSK spectrums that are non-bandlimited and place them at 1 1/2 times the baud rate spacing. This is shown in Figure 5. As one can see, there are considerable tails from adjacent carriers that spread into the mainlobe of the desired carrier. The power from adjacent channels acts like noise for the desired channel and is called adjacent channel interference (ACI). Thus, one's effective signal-to-noise ratio is lowered, and it takes more power over the satellite to achieve a given bit error rate.

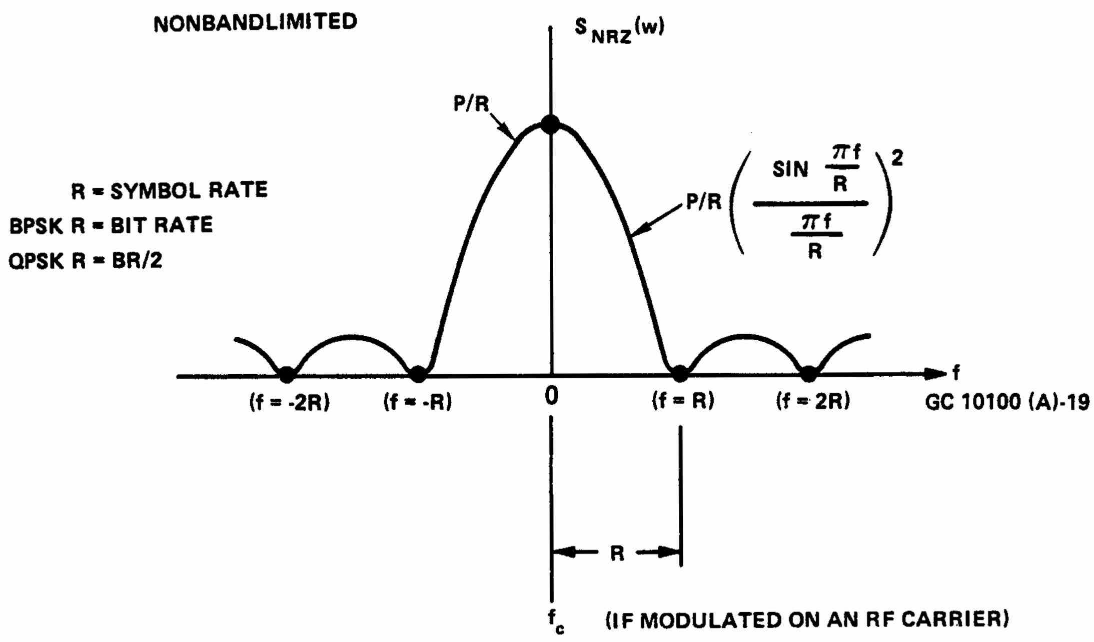


Figure 4. Power Density Spectrum of an NRZ Code (Random bit pattern)

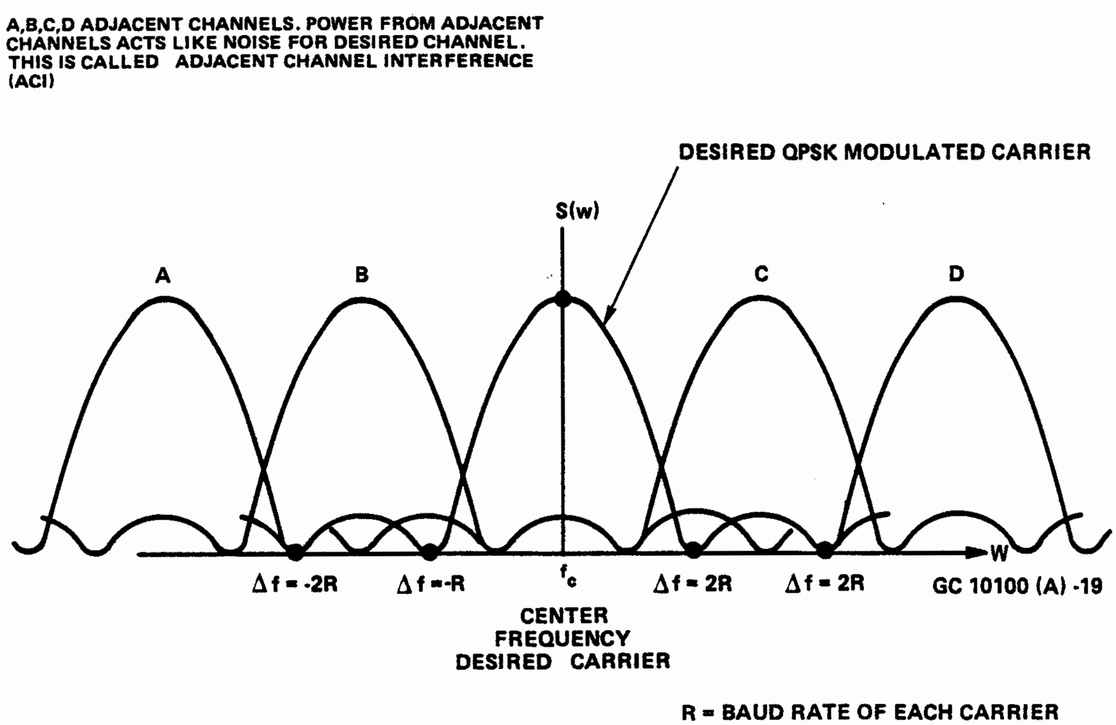


Figure 5. Overlapping Spectrums of Non-Bandlimited QPSK Carriers Spaced at 1.5 Times Baud Rate or 0.75 Times Bit Rate

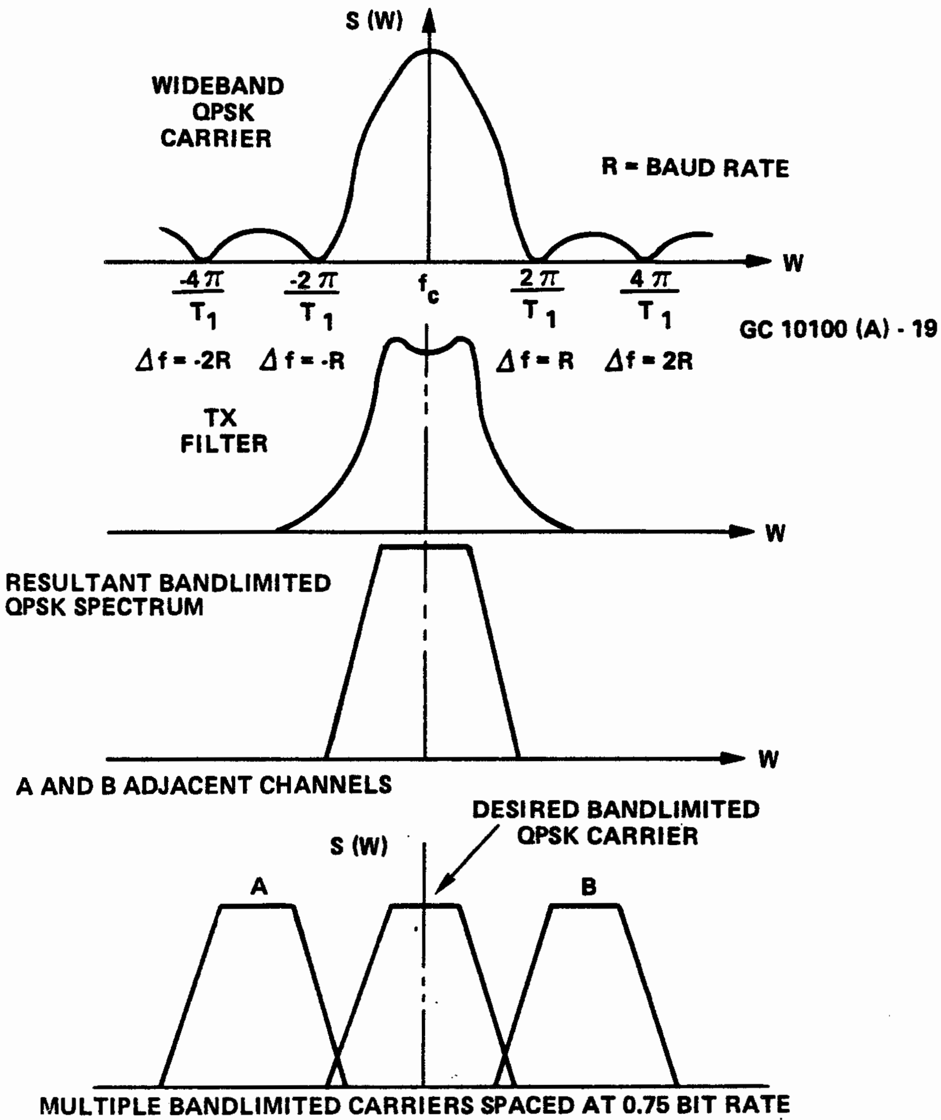
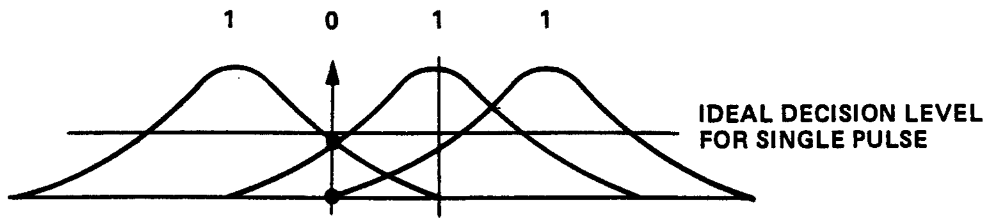


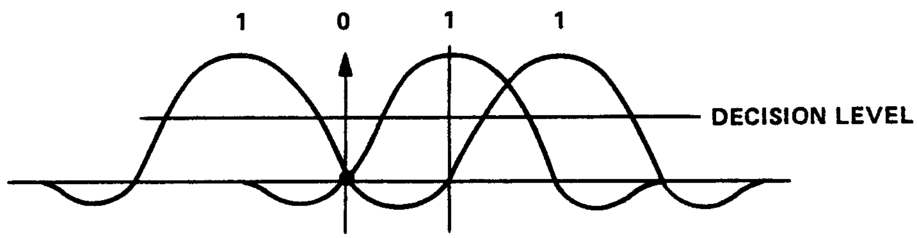
Figure 6. Bandlimiting QPSK Modulated Carrier to Minimize Adjacent Channel Interference

In Figure 6, the original $(\sin x)/x)^2$ spectrum is shown, then a typical transmit filter spectrum is shown. The resultant combination spectrum is then presented in the second part of the figure. If one takes this band-limited spectrum and stacks such spectrums at 1-1/2 times the baud rate spacing, one gets the figure in the bottom of the figure. Now it is seen that there is minimal overlap between the spectrums. Therefore, by filtering one can eliminate the power from adjacent channels and minimize the effects of ACI. Not just any filtering can be used, however, because the effects can be quite insidious.



For a Zero or No pulse the desired signal value is zero. The preceding and following bits have values of 0.4 at the optimum sample time. Thus, the summed interference from adjacent bits equals 0.8 which is greater than the decision threshold of 0.5. This results in a state estimation that a One was sent whereas a Zero was actually transmitted.

Such interference from adjacent bits acts like noise for the desired bit and is called Intersymbol Interference (ISI).



Same pattern with Nyquist Filtering. All other bit amplitudes go through Zero at the sample time for any given bit.

Figure 7. Innersymbol Interference and Nyquist Filtering

Filtering

To illustrate the effects of filtering, let us consider an On/Off Key-type system, where there is a pulse for a One and there is no pulse for a Zero. Thus, one is trying to distinguish between the absence or presence of pulses. Ideally, the threshold is at half the peak value of a pulse. Shown in the top part of Figure 7 is a bit stream with a non-optimum type of filtering. There are considerable tails to the pulses. At the point where one is trying to make a decision on a given pulse, it is seen that there are considerable tails from adjacent pulses. These pulse tails add up to cause confusion, just as noise causes confusion around the sampling point, and thus act like noise on the decision process. This effect is called inter-symbol interference (ISI). Thus, while filtering can be used to eliminate the power from adjacent channels, filtering itself can cause an equivalent noise due to the inter-symbol interference and can cause an increase in the power required to achieve a certain bit error rate. Is there a method of filtering such that this degradation is not caused? If one uses Nyquist Filtering, where one has the proper odd type of symmetry around the 6-dB point of the overall filtering through the whole system, then one can achieve the condition where all other pulses are going through zero at the time a given pulse is being sampled. Thus, one can optimally control the tails of the pulses so that these are all going through zero at the time one makes a state-estimation on a given pulse. This is shown in the bottom diagram where a set of Nyquist pulses is presented for the same bit pattern as that at the top of the figure. Thus, there is a very important trade between ISI and ACI--one has to filter in a very special way and trades eliminating power from the adjacent channel versus the inter-symbol interferences caused by filtering a given channel. Of course, one can only approximate the Nyquist waveforms. Therefore, there is always a trade between filter complexity, how much ISI one is creating by filtering, and how much ACI one is eliminating. The time domain response of the overall filtering in the whole system, that is, modulators, upconverters, transponders, downconverters, demodulators, all determine the ultimate bit error rate performance of the system. It should be noted that amplifiers have other limitations in the sense that they have finite slew rates. There are only finite energy sources available and, therefore, the time domain response is not always the inverse Fourier transform of the frequency domain response. The inverse Fourier transform assumes there is always adequate current or voltage available. In reality this is not always the case; therefore, one must also carefully consider the slew rate limitations of equipment. In summary, with proper filtering one can bandlimit the spectrum, which allows one to stack up many more carriers on the transponder than would be possible without filtering. It is also to be emphasized that the filtering must be a very special kind, and there is a definite trade in the final analysis between minimizing adjacent channel interference, minimizing inter-symbol interference, and maximizing signal-to-noise ratio at the decision time. For example, with the proper filtering, one can place 800 each 56-kb/s carriers on a 36-MHz-wide transponder due to bandwidth limitations alone.

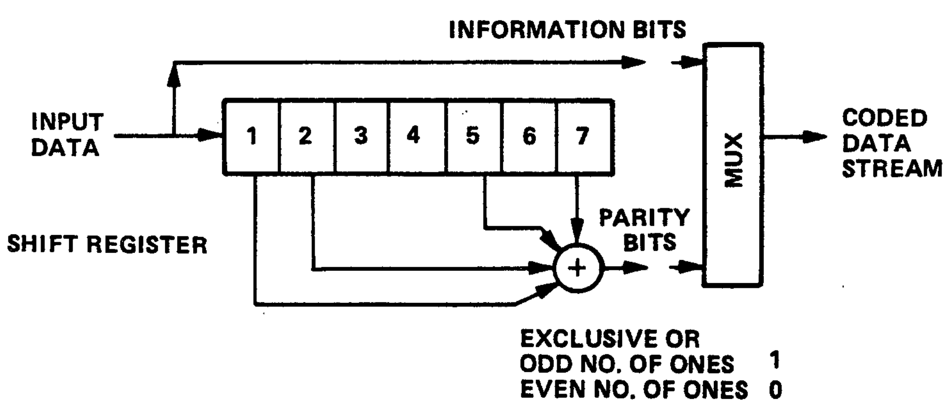


Figure 8. Coding Example r = 1/2 Convolutional Encoder

Coding

Let us next consider power limitations and how one combats this limitation. The non-bandlimited PSK curve was shown in Figure 3. Is there a mechanism by which a given bit error rate performance can be achieved at a lower power level? Thankfully, there is and this is known as forward error correction (FEC) coding. In forward error correction coding, a controlled redundancy is added to the data necessitating an increase in the actual transmit bit rate, but at the same time achieving a net reduction in the amount of power required in transmission. In general, one talks about rate K/N codes. For K information bits, there are N coded output bits. The output coded bit rate is known as the symbol rate. One should be careful not to confuse the baud rate or the symbol rate of a modulator (the rate at which one changes modulation parameters) with the symbol rate of the FEC Codec (the coded output data rate). The output data rate of the encoder is N divided by K times the information data rate R. It should also be noted that when one encodes, one increases the data rate, thereby decreasing the energy per bit. One has to take that into count in calculating the coding gain. In other words, the net coding gain is the code overall gain minus the loss due to spreading the energy by increasing the data rate. An example of a good cost-effective coder is a convolutional coder with threshold decoding. An example of a simple convolutional coder with threshold decoding is used to illustrate the idea of coding. In Figure 8, a rate 1/2 convolutional encoder is presented. Here the input data is routed through a 7-bit shift register. The input data is also routed around the register into the output MUX. As one can see, different stages of the register are Exclusive-Or-ed together to produce what is known as a parity bit. An Exclusive-Or logic circuit produces a One output when there is an odd number of One inputs. If there is an even number of Ones, the output is a zero. One can multiplex the two data streams together into one coded stream at twice the original information data rate. If one is using QPSK modulation, one can alternatively put the original information bits on one quadrature carrier and put the parity bits in the other quadrature carrier. In Figure 9, the corresponding threshold decoder is shown. The data that is received with errors is demultiplexed into two data streams. One has to achieve branch synchronization such that one has the information bits in the proper data stream and parity bits in the other data stream. The information bits are routed through a parity generator just like the one in the encoder where the parity bits are recreated. These recreated parity bits are then compared with the parity bits received over the satellite. The comparison output, which is a module-two comparison (an Exclusive-Or), is called a syndrome. Now, if there is a single parity error, this will cause a 1000000 syndrome pattern as shown in the figure. However, if there is a bit error while the parity bit is correct, then as the bit error goes through the shift register, every time it is in one of the places where there is a tap, it will cause the recreated parity bit to be wrong and will generate a One in the syndrome. Therefore, a 1100101 syndrome pattern will be generated as shown. The syndromes are fed into a 7-bit shift register where this pattern is the input to a threshold circuit. Thus, with a single bit error, a threshold of four could be used. Whenever a threshold of four is achieved by summing the proper places in the syndrome register, then the threshold circuit can be used to correct the information bit and eliminate the error. What is known as syndrome feedback can also be used so that at the time the information bit is corrected, the syndrome register contents are inverted. This allows one to achieve somewhat more coding gain.

Thus, one can see by this simple coding example how certain syndrome patterns can be recognized and used to discover and correct the bit errors by adding some controlled redundancy to the orginal data. In Scientific-Atlanta's commercial digital modem-forward error correction codec, we are presently using a rate 7/8 FEC coder that will take a 10^{-4} channel bit error rate and correct it to 10^{-9} bit error rate. Thus, five orders of magnitude improvement in error rate are achieved by coding. At the same time, being a high rate code, that is 7/8, the spectrum is only spread by a factor of 8/7 times the original information data rate. Ideally, one would like a high rate code in which one spreads the spectrum very little while at the same time achieving a lot of coding gain. Presently, most links are more power-limited than bandwidth-limited. For example, with 5-meter antennas and 56-kb/s data streams using rate 7/8 coding, one can typically put up 200 carriers on a single transponder if one allows properly for margin, pointing error, rain, etc., whereas the bandwidth limitation of the filters would allow 800 carriers. If one uses 10-meter antennas, however, where one basically has a 6-dB gain over a 5-meter antenna, the net number of carriers due to power limitations has approached 800 carriers and the number of carriers due to bandwidth limitations is approximately 800 carriers. This is usually an optimal operating point for a link. However, it cannot be overemphasized that the fundamental limitation is economics and what is most feasible or actually used is very much driven by the economics of the situation.

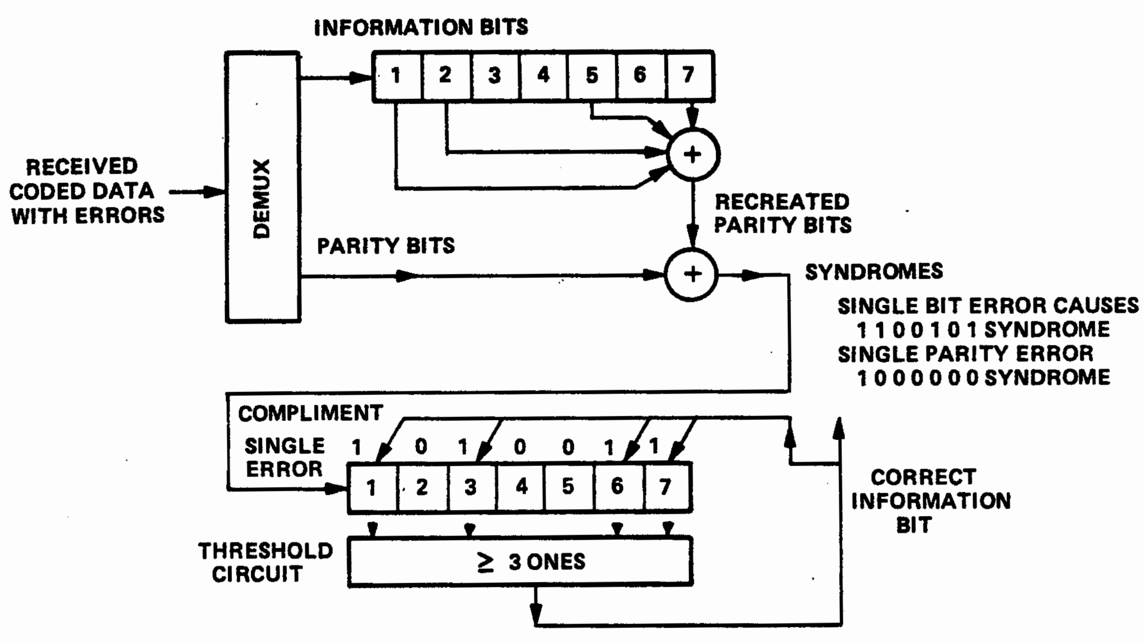


Figure 9. Coding Example: Threshold Decoder for r-1/2 Code

Scientific-Atlanta Digital Earth Terminal

To allow cost-effective digital transmission over satellites, Scientific-Atlanta has developed a complete line of digital earth terminal products. Nyquist quadraphase and biphase modems with corresponding forward-error correction codecs allow one to transmit at low power levels and yet occupy a minimum spectrum. The associated low-phase noise modem synthesizers have also been developed that allow one to vary modem transmit and receive frequencies. 1 For N switches allow the user to back up N modems and replace a failed modem with a backup unit, thereby increasing the availability of the link. There have been similar developments in the area of up- and downconverters with respect to achieving phase noise performance that is adequate for data transmission. A monitoring and control unit, MCU, monitors the overall earth station, switches out any component that has failed, calls up an overall master station, and verifies by telephone line that there has been a replacement of a failed unit, thus allowing one to dispatch service personnel. These units have all been designed to provide a cost-effective solution to the problem of transmitting information digitally via satellite.

Summary

The previous sections have given an overview of how one can transmit information over earth satellites using digital techniques and why one might want to do so. It should be emphasized, as previously stated, that the final driving force is economics and not technology. That is to say, one can always use a powerful enough FEC coder or large antenna such that one can get the link to be bandwidth-limited as well as power-limited. This, however, might not be a cost-effective solution for a given business network. As also discussed previously, there could be great advantage to transmitting broadcast video over the satellite using digital techniques. The source encoding equipment at this time, however, is too expensive for this to be economically viable. Thus, digital earth station equipment must use technology to provide cost-effective solutions for the transmission of digital data via satellite.

Error-Correction Coding

John F. Schimm, Jr.

Introduction

Error correcting coding is a signal processing technique used to improve the reliability of messages on digital channels. Many different types of digital channels exist--satellite data links, digital telphone links, microwave channels, computer mainframe memories, computer magtape, holographic and photographic film. Each digital channel medium has its own set of characteristics and idiosyncrasies. Many methods and approaches to error correction exist. This discussion is limited to error correction for satellite data links. Even in this limited area, many diverse approaches to error control exist.

Automatic request repeat (ARQ) is an error control technique that repeats the transmission when an error is detected in the reception. File transfers of computer tape to computer tape use this method for overall error control. Clearly, this method becomes very inefficient when the channel errors become excessive or when transmission delay times are large.

FEC Coding

For satellite data links, feed forward error correction (FEC) is a proven efficient method for improving data reliability. This method allows transmission of data over an imperfect channel with the data reliability restored after reception. Typically, an encoder and decoder are added to the digital data link. A simple block diagram is shown below:

Although individual FEC coding schemes take on many different forms, two ingredients are common to all. One is redundancy. Coded digital messages always contain extra or redundant symbols. These symbols are used to accentuate the uniqueness of the message, and are chosen so that channel noise will not corrupt enough symbols in a message to destroy its uniqueness. The second ingredient is noise averaging. Averaging is obtained by making the redundant symbols depend on a span of information symbols. Since the amount of noise that affects an individual bit varies in a probabilistic fashion with large noise excursions being rather infrequent, one can ameliorate their effect by averaging over several received symbols.

Bit Error Rate and E_b/N_o

One of the quality measures of a digital channel is bit error rate (BER). The BER of a channel is measured by counting the number of errors E which occur in a block of B bits. Then BER = E/B. Thus, a BER of 10^{-7} means we would expect to have an average of one error in every 10^7 bits. Another way to see this is to say that the probability that an individual bit is in error is 10^{-7}. Clearly, the BER tells nothing about the probability distribution of the errors, which can be important. For satellite data links, an excellent model for this distribution is the white Guassian noise channel.

An important quantifying measure of a digital channel is the energy-per-bit-to-noise-density ratio (E_b/N_o). In essence, the received E_b/N_o of a channel represents the received signal-to-noise ratio. Thus,

$$\frac{E_b}{N_o} = \frac{C}{N} \cdot \frac{NB}{R}$$

where:

C = Received carrier power.
N = Received noise power.
NB = Received noise power measured in B rectangular bandwidth.
R = Channel symbol rate.

At this point, one might ask, "Given a specified E_b/N_o, what is the best BER we can expect?" The answer to this question depends on the modulation method and type of receiver used. One modulation method, phase-shift keying (PSK), has shown itself to give good error performance with reasonable receiver hardware complexity. This modulation technique is also somewhat impervious to certain types of interference. A plot of the theoretical BER performance of PSK modulation used with a coherent receiver is shown in Figure 1 as the uncoded curve. The BER or P_e for this curve is given by:

$$P_e = Q \sqrt{2 E_b/N_o}$$

where the Q function is given by:

$$Q(\alpha) = \frac{1}{\sqrt{2\pi}} \int_{\alpha}^{\infty} \exp(-B^2/2)\, dB$$

A table of BER values versus E_b/N_o is given in the Appendix. Note that by increasing E_b/N_o we may decrease the BER. The goal of FEC coding is to allow one to achieve a given BER at a small E_b/N_o. In Figure 1, we see an uncoded system requires a minimum of E_b/N_o of 11.3 dB to achieve a BER of 10^{-7}. The coded curve shows us that this same BER can be achieved in a coded system at an E_b/N_o of only 8.0-dB or a 3.3-dB savings. However, keep in mind that the coded system requires adding redundant symbols to the input message. Thus, our transmission data rate (or baud rate) and required channel bandwidth have increased according to the number of redundant symbols added. The <u>rate</u> (R) of the coded system is inversely related to the bandspreading required. In Figure 1, the code rate is R = 7/8, meaning for every seven input symbols, eight output symbols are produced. Thus, the bandspreading of this code is 8/7.

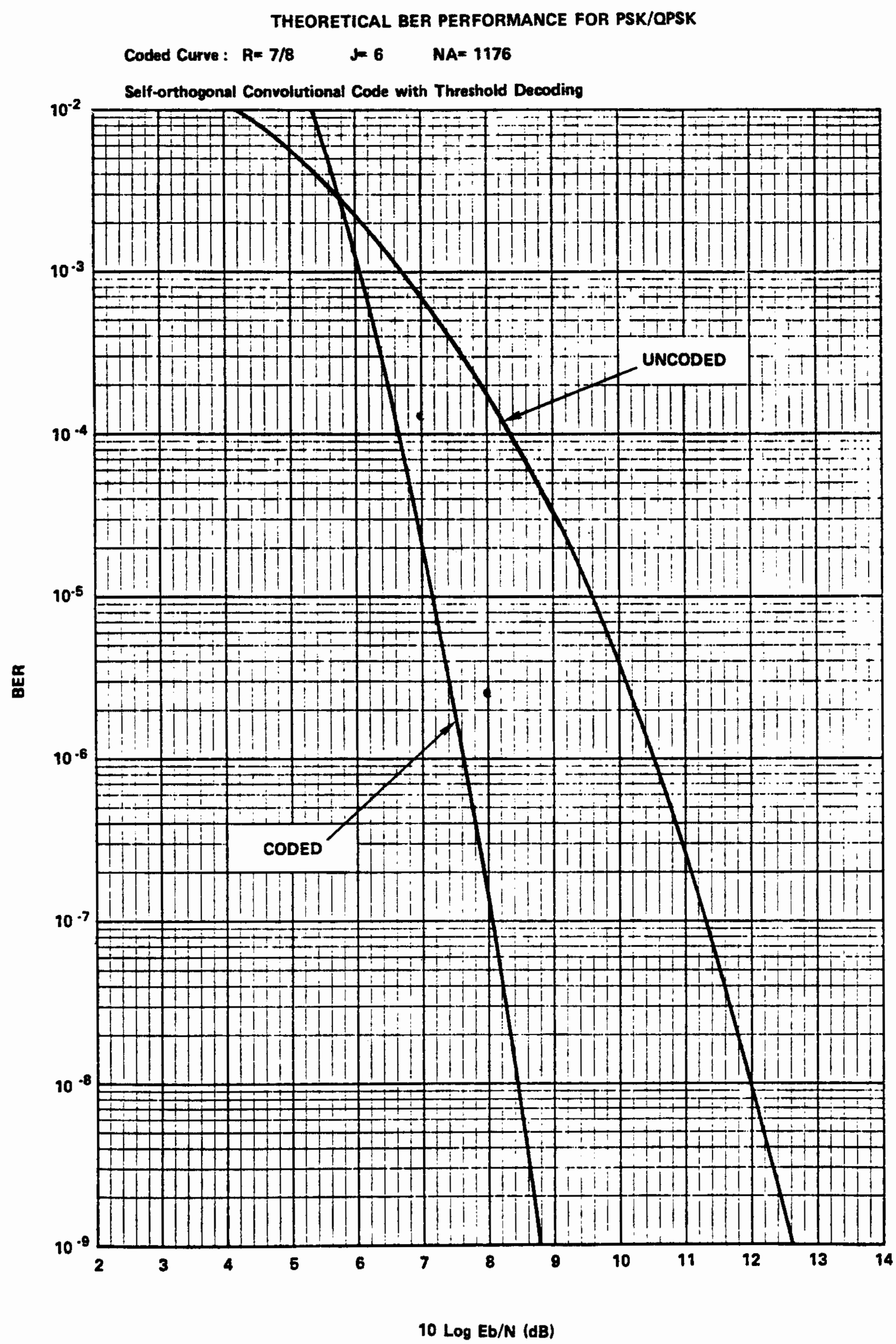


Figure 1. Theoretical BER Performance for PSK/QPSK

Economic Implications

The 3.3-dB gain of the above-coded system may be viewed in several different ways. For example, if one were comparing antenna sizes, a 7-meter-dish antenna provides 3-dB gain over a 5-meter dish. Or, if one is comparing LNAs, an 80°K LNA provides about 1.2 dB over a 120°K LNA. By comparing the cost versus the gains, one will find that the coder is the most cost-effective method to achieve link margin gain.

In many systems, it is economically feasible to justify all three methods of improving link margin. Figure 2 shows a plot of percent bandwidth versus percent power. In leasing satellite transponder space from the common carriers, costs are based on the percent of total power used or percent of total bandwidth used, with the largest percent being the governing cost. A power-limited link (in our definition of usage) is one which has a larger percent power requirement than percent bandwidth requirement. A bandwidth-limited link is just the opposite. For the power-limited link, the use of coding, larger antenna and/or higher performance LNA are justified. These improvements move the power-bandwidth operating point as shown in Figure 2 from point A to point B or C. For a bandwidth-limited link, these changes are unnecessary, as the constraining factor or cost is the bandwidth usage.

Most satellite links tend to be power-limited, although there are efforts to conserve bandwidth as well. In the Rate 7/8 coded system discussed previously, a 3-dB gain will improve a 100% power-requiring link to a 50% power requiring link. If the bandwidth requirements were 25%, the rate 7/8 coded system will require 28.6% of the available bandwidth. In other words, we have cut the cost of operating the link by 50% with only 3.6% more of the available bandwidth required. Because an individual user's link is unique, each must be optimized for antenna size, LNA and coding. Other factors not discussed are the number of transmit sites, number of receive sites and configuration of the network. Optimization of these factors is beyond the scope of this discussion.

In the next section we discuss codes and coding technique which can achieve various amounts of link margin gain.

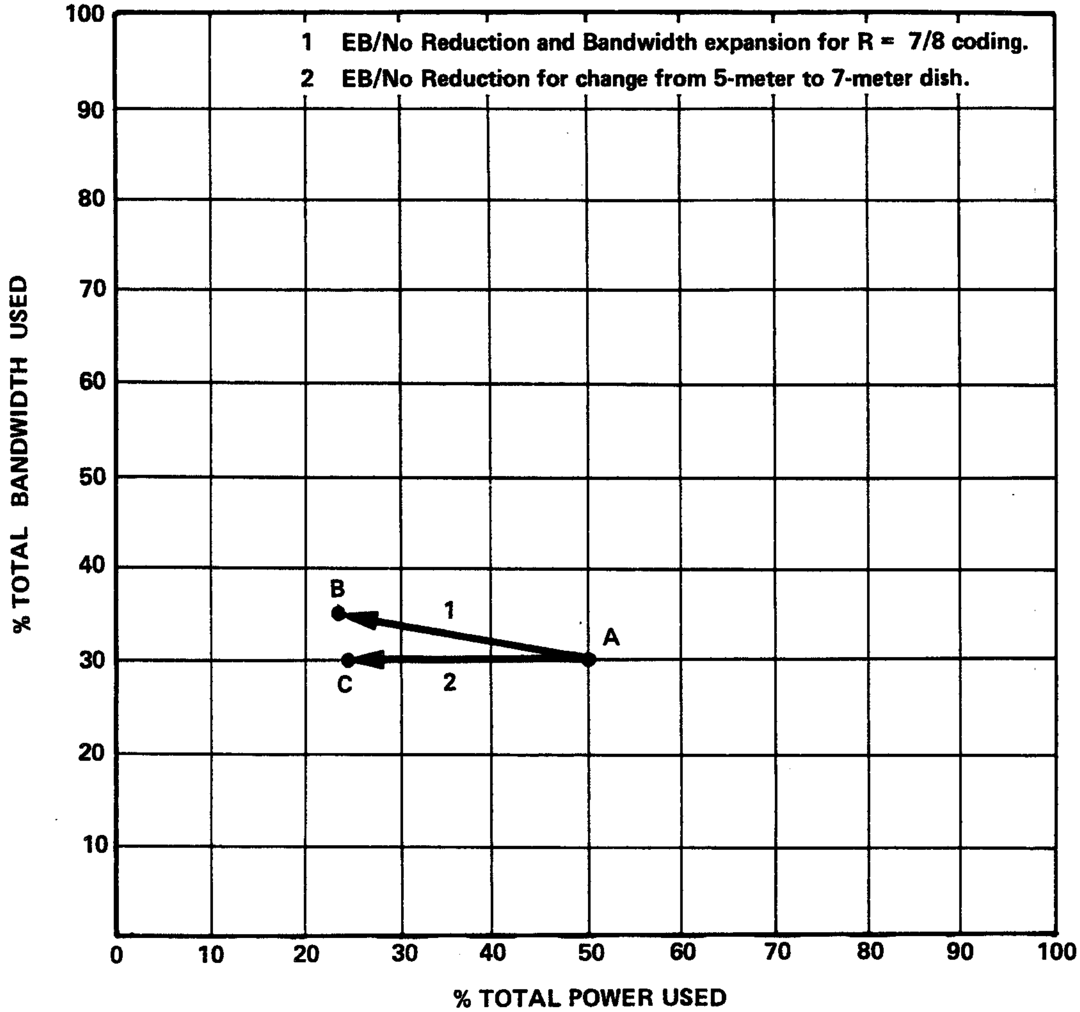
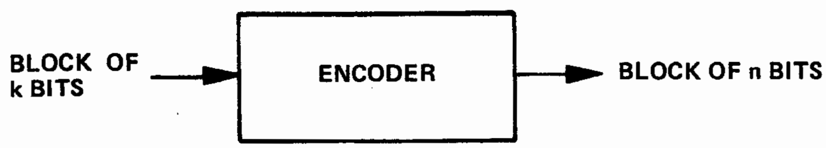


Figure 2. % BW Used vs. % Power Used

Coding Technique

Many good-to-excellent codes and coding methods exist. The primary performance limitation in most coded systems is the complexity of the decoder. Thus, in discussing coding it is very important to look at the decoding method and its hardware requirements.

There are two basic methods of coding--block and convolutional. A block code, as the name imples, groups the input data into blocks of k bits. For each input block of k bits, the encoder produces an output block of n bits. Thus, n-k additional bits are added.



The rate of this code is then R = k/n. Now, if the k input bits are part of the n output bits, then the code is termed systematic. Codes not possessing this property are termed non-systematic. For block codes, there is no memory between blocks. In other words, an output block is dependent only on its input block and not on any previous input blocks.

Systematic codes have the obvious advantage that the information bits appear as part of the codeword. This reduces complexity and requires less buffering and data manipulation. Also, the BER of the link may be checked with error correction inhibited. For BER, performance-wise, there is no difference for systematic versus non-systematic linear block codes. Because of this, systematic codes are used almost exclusively. For convolutional codes (to be discussed), there is typically a BER performance advantage for a non-systematic code when used with certain types of decoders.

Well-known linear block codes include:

- Hamming — Single error correcting
- BCH — Multiple error correcting
- Reed Solomon — Multiple symbol correcting

This list is by no means inclusive, as many other types of block codes exist.

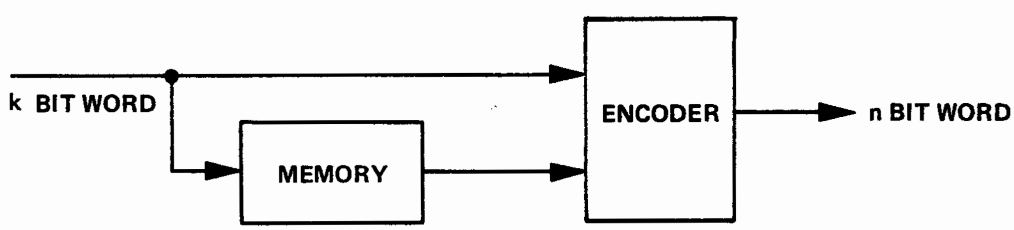
Although there are also many ways to decode block codes, three of the better known techniques are threshold decoding, Meggitt decoding, and algebraic decoding. Meggitt decoders can, in principle, decode any cyclic code (of which cyclic Hamming and BCH codes are members). Practical hardware considerations limit use of these decoders to three random error correcting codes or less. Performance gains are in the 1- to 2.5-dB range.

Threshold decoders are very similar to Meggit decoders, but require the code to have a certain structure in order to work. For these codes threshold decoding is a simple and powerful algorithm. Performance gains of up to 3.5 dB are possible with reasonable hardware. Also, threshold decoding can be implemented fairly easily at high data rates where other methods may require hardware faster than present state-of-the-art.

Algebraic decoding is the most powerful decoding method of the three discussed. It is also the most complex in hardware. This method uses the cyclic and algebraic properties of the code to construct a set of simultaneous equations in several unknowns. An efficient solution to these equations can be found by using Berlekamp's algorithm.[1] Then, using Chien's search method[2] the error locations can be found. Additionally, Burton's method[3] may be used to replace an inversion step. Algebraic decoding is generally used with BCH codes or Reed-Solomon codes and can provide some impressive gains. This method's major drawback is its lack of ability to accommodate soft-decision information. As a result, algebraic decoders (especially Reed-Solomon decoders) are often used as the outer decoder in a concatenated code scheme. The inner decoder is generally a soft-decision decoder utilizing a short-constraint-length code. Performance gains of 6 to 7 dB are not uncommon for this configuration.

Convolutional Codes

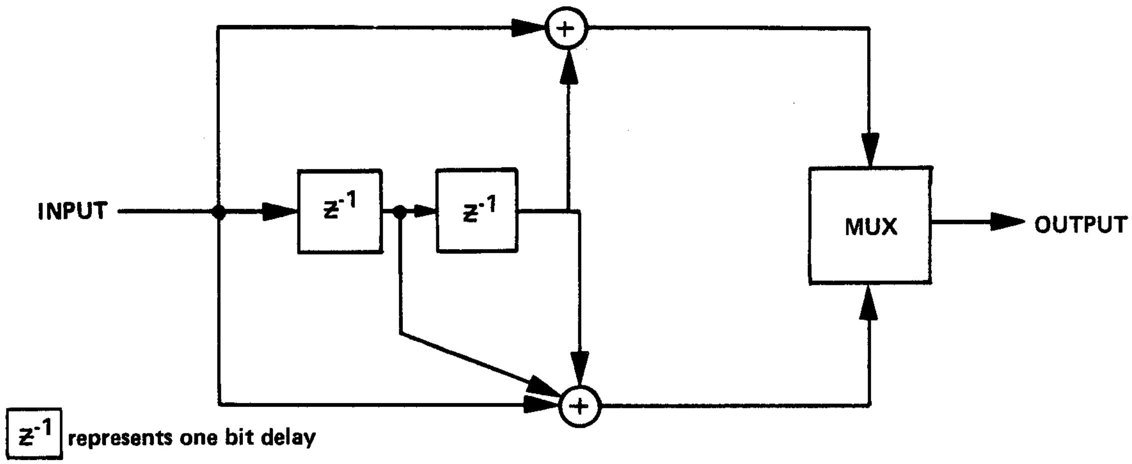
Convolutional codes, like block codes, produce an n-bit output codeword for every k-bit input message. For a convolutional code, however, the output codeword is not only a function of the present k-bit input message, but can also be influenced by previous input messages. Thus, a convolutional encoder has memory of previous input messages.



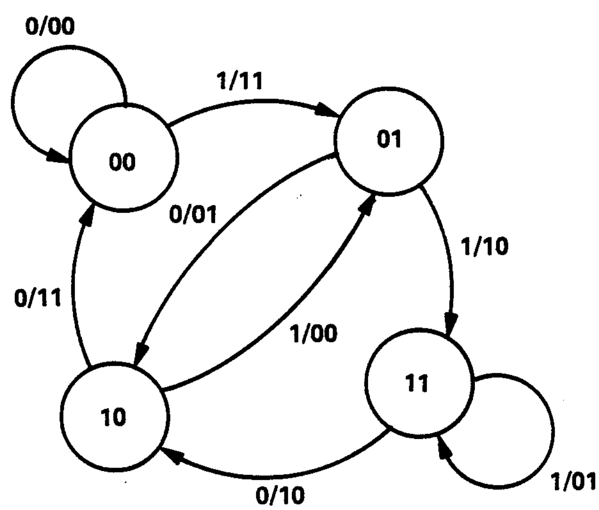
The constraint length of the code is usually defined as simply the number of bits in the encoder memory plus the number of inputs.

Because it contains memory, a convolutional encoder may be regarded as a state machine, with the state being determined by the memory contents.

RATE 1/2 ENCODER CONSTRAINT LENGTH = 3



The k-bit input message modifies the state of the machine on the next clock cycle. Effectively, the k-bit input defines one of 2^k possible branches to a new state. With each branching operation, an n-bit codeword is output for transmission over the channel. In the example above, there are four possible states 00, 01, 10, and 11. The input message, either 0 or 1, defines one of two branches. For each branch operation, the encoder outputs a 2-bit codeword. A state diagram can be defined as follows:



The binary bits inside the circles define the state. The symbol along the top of a transition line indicates the input symbols at the encoder that cause that transition, and the symbols below the transition line show the resulting channel symbols at the encoder output. A sequence of information symbols define a path through the state diagram, and the channel symbols encountered along this path produce the resulting encoded channel sequence.

An alternative method of describing a convolutional code is in terms of a coding tree. Figure 3 shows a coding tree for the example above. An input of "0" chooses an upper branch from a node, while an input of "1" chooses a lower branch. The output symbols are indicated above each branch. From this diagram it is clear that beyond three branches (or one constraint length) into the tree, the structure becomes repetitive. This effect occurs because of the finite number of states in the encoder. An alternate tree drawing that eliminates the duplicate states is shown in Figure 4 and is called a trellis, a tree-like structure with merging branches. Code branches produced by an input 0s are shown as solid lines, while input 1s produce a dashed line. Once again, the output symbols are shown along the top of each line. Each set of four vertically aligned points (or nodes) represents the four possible states of the encoder.

To decode a convolutional code, there are basically two approaches. One approach relies on the tree structure of the code and estimates the original path taken by the encoder. Decoders of this type are maximum likelihood (Viterbi) decoders and sequential decoders. The other approach relies only on the distance properties between codewords over its constraint length. A decoder of this type is the threshold decoder.

Maximum likelihood decoders (MLD) estimate the path through the trellis by correlating the actual received symbol sequence with all possible transmitted symbol sequences of a given length (typically, four or five constraint lengths). For code of constraint length 6, this method would require approximately $2^{30} = 1.07 \times 10^9$ comparisions. In 1967, Viterbi[4] published an algorithm which dramatically reduces the number of comparisons required. This algorithm is essentially a dynamic programming solution to finding the path with the highest correlation. For each of the possible states of the encoder, a MLD decoder stores a path metric and path specification of the best path to that state. At a new iteration, each state is extended to the next 2^k nodes in the trellis.

A branch metric is calculated for each extension by correlating the actual received symbol with the hypothesized transmitted symbol for that extension. The branch metric is added to the old path metric to produce a new path metric at the new state. At this point, 2^k paths (with their associated path metric) point to each new state. The path specification with best metric is chosen and the remaining paths are deleted. The algorithm then repeats itself extending these states to new states.

After the decoder has accumulated a path specification to a depth of several constraint lengths, the initial part of the path estimated may be output to the user. A traceback algorithm selects the path with the best metric at the deepest depth and traces back through four or five constraint lengths. The oldest constraint length of data estimates on this path form the output to the user.

Viterbi decoders (or MLDs) achieve respectable coding gains (5 dB at 10^{-5} BER, rate 1/2 code, constraint length = 6, 3-bit soft decision) for reasonable hardware complexity. Practically, these decoders are limited to use with codes of constraint lengths less than 8 due to the storage requirements for the states. (Storage requirements increase exponentially with constraint length.) Also, most MLDs are built for code rates of 1/2 to 3/4. An MLD built for a code rate of 7/8 would require fairly complex hardware. However, methods exist to trick a 1/2 decoder to operate at a higher rate at the expense of reduced coding gain.[5]

Sequential decoding is another error correction technique which relies on the trellis (tree) structure of convolutional codes. Sequential decoding is very similar to MLD decoding in that branch metrics, path metrics and path specifications are stored. However, sequential decoding does not keep track of these quantities for all possible encoder states. Rather, only the path and states that appear to be the most probable are extended. Because of this limited search, the decoder is never completely certain that the present path is the best.

This approach can be viewed as a trial-and-error search for the correct path in the code tree. The search is performed in a sequential manner, operating always on a single path. At each step, the decoder moves forward by extending the most probable branch from the current node or state. The decoder is allowed to back up and change previous branch decisions. If an incorrect branch is taken, subsequent extension of this path will be wrong. The decoder is able to recognize this situation by examining the path metric. However, it is possible to extend an incorrect path for several branches. These computational requirements for these sequential decisions are variable and dependent on the severity of the channel noise. During periods of low noise, the decoder advances quickly through the tree, seldom following an incorrect branch. However, when the channel noise increases due to its probabilistic nature, the decoder will explore many incorrect paths before finding the correct path.

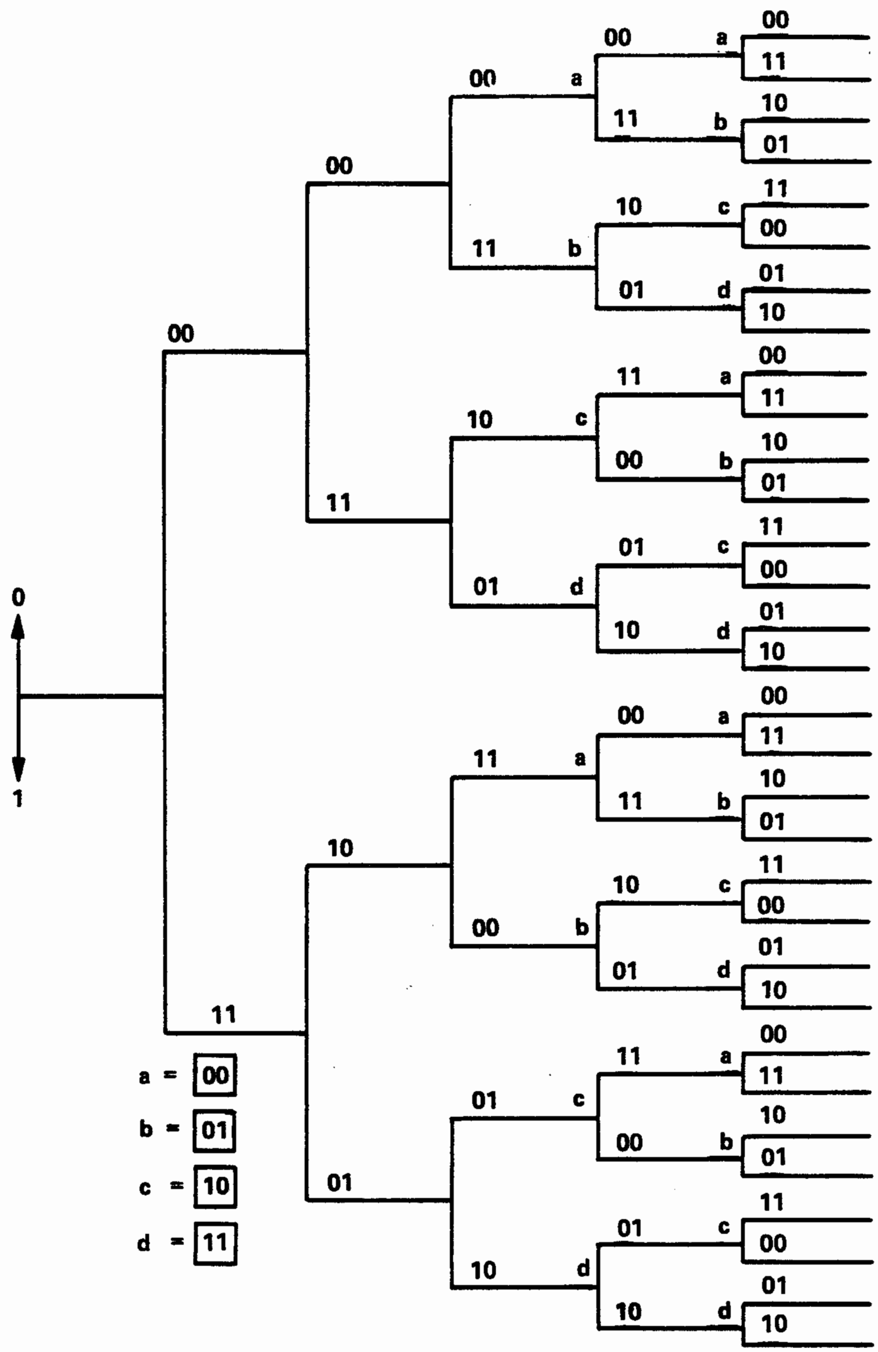


Figure 3. Tree-Code Representation for Encoder

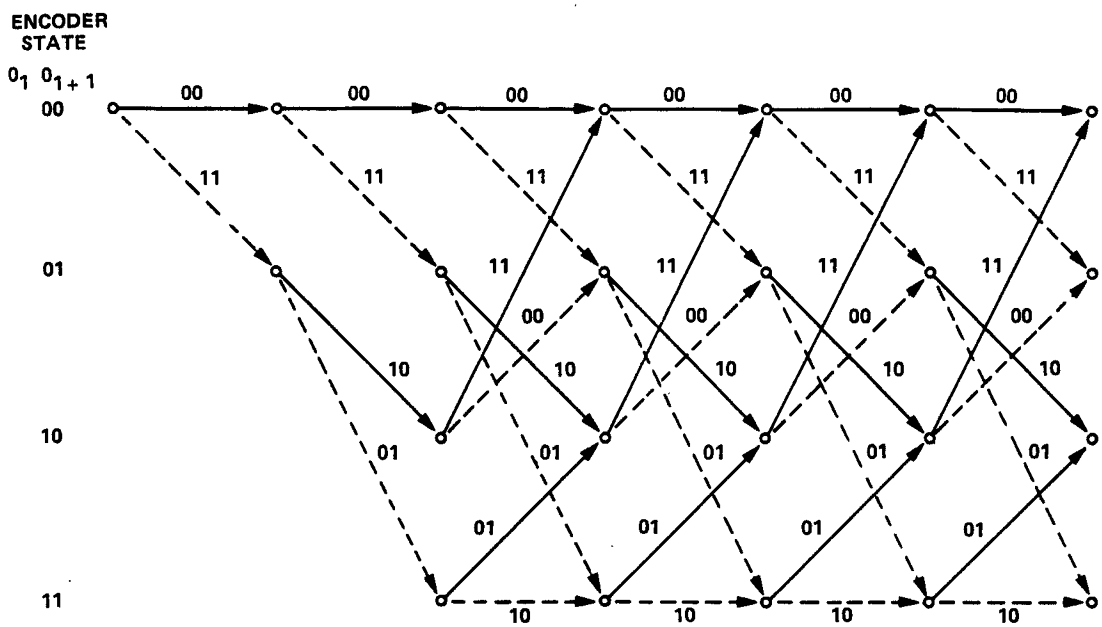


Figure 4. Trellis Representation for the Code Generated by $g_1(d) = 1+D+D^2$, $g_2(D) = 1+D^2$. (0 input=solid line; 1 input=dashed line)

Because of the variable speed with which the decoder advances through the tree, input and ouput buffers are required to keep the data flow continuous. Indeed, there is a finite probability that the decoder will overflow the input buffer before a good path can be found. In this case, the decoder is forced to reset itself. This action causes the data buffer to be output uncorrected and is known as an erasure of the data bits. The two best known sequential decoding methods, the Fano algorithm and the stack algorithm, are both not erasure free. A batch algorithm development by Chevilatt and Costello[6] called the multiple-stack algorithm (MSA) is erasure free. However, adaptation of this algorithm to medium speed continuous operation involves solving challenging memory management problems. For all types of sequential decoders, the probability of buffer overflow can be minimized through proper code selection and adjustment of other parameters.

Threshold Decoding Using CSOC Codes

Convolutional self-orthogonal codes (CSOC) form a large class of systematic convolutional codes of various code rates and error correcting abilities. These codes are called self-orthogonal since an additional orthogonalization step is not required in decoding. Figure 5 shows a block diagram of a rate 7/8 decoder. An encoder, G(D), processes the information bits and computes a parity bit which is affixed to the encoder output bit stream. The decoder processes the received bits which have been corrupted by channel noise. A re-encoder G´(D) encodes the receive bits in the same manner as did the encoder. By comparing the received parity bits with the regenerated parity bits, a syndrome bit is formed and stored in the syndrome register. The threshold circuits look for various patterns of "ones" in the syndrome bits to estimate the occurrence of errors. The resets to the syndrome register remove the effect of a given error estimate on following error estimates.

For a rate 7/8, J=6, d_{min} = 7 CSOC code, the encoder and syndrome registers are 146 bits long. Figure 6 shows a block diagram of the encoder register, and Figure 7 shows the syndrome register. This code can correct any three errors occurring in its constraint length of 1176 bits. The performance of this code was shown earlier in Figure 1, along with uncoded theory for a PSK ideal receiver. At 10^{-7} output BER, this code provides 3.3-dB gain. Thus, using this coder, a 56K bits/s communications link could be operated at less than half the power. The transmitted singal would, of course, be bandspread to 8/7 x 56 baud/s = 64K baud. At 10^{-9} output BER, this code provides 3.8-dB gain.

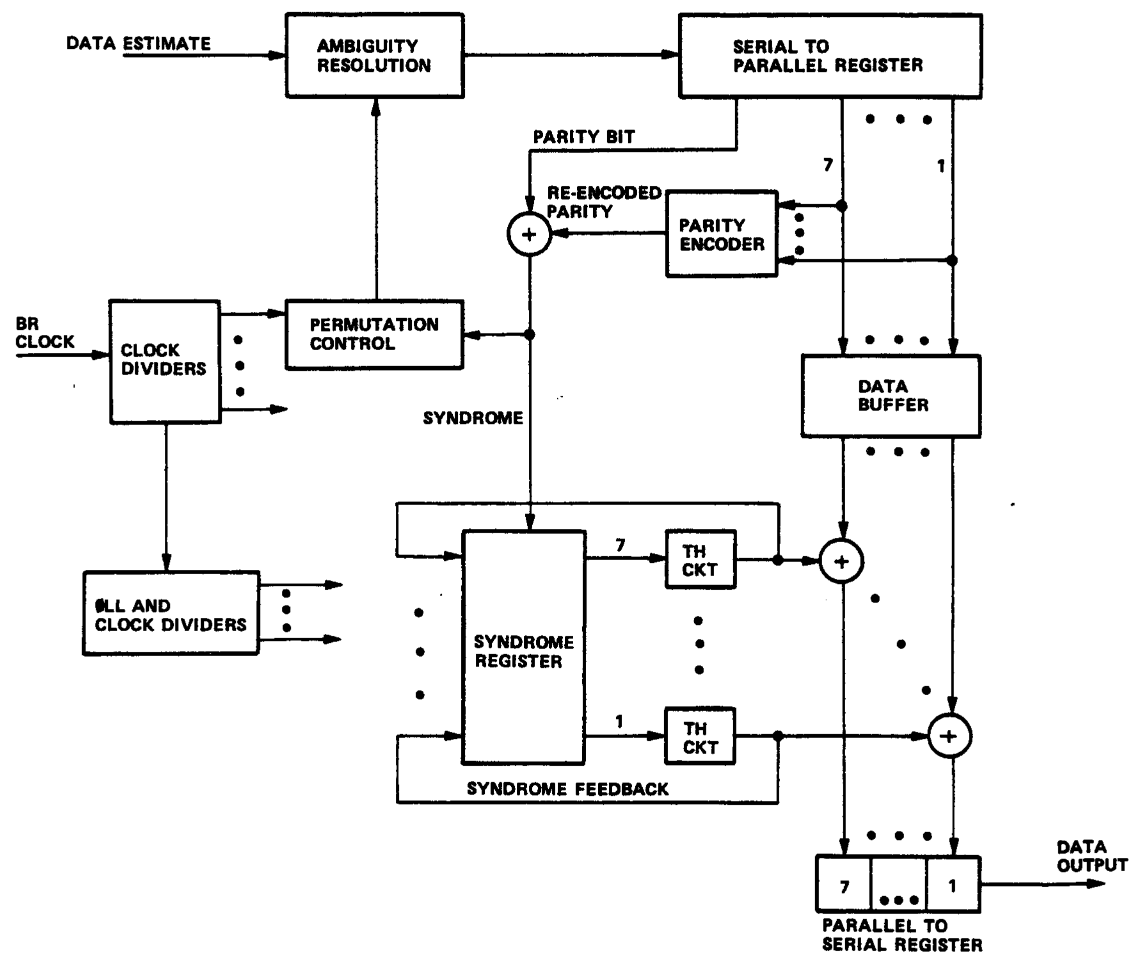


Figure 5. Decoder Block Diagram

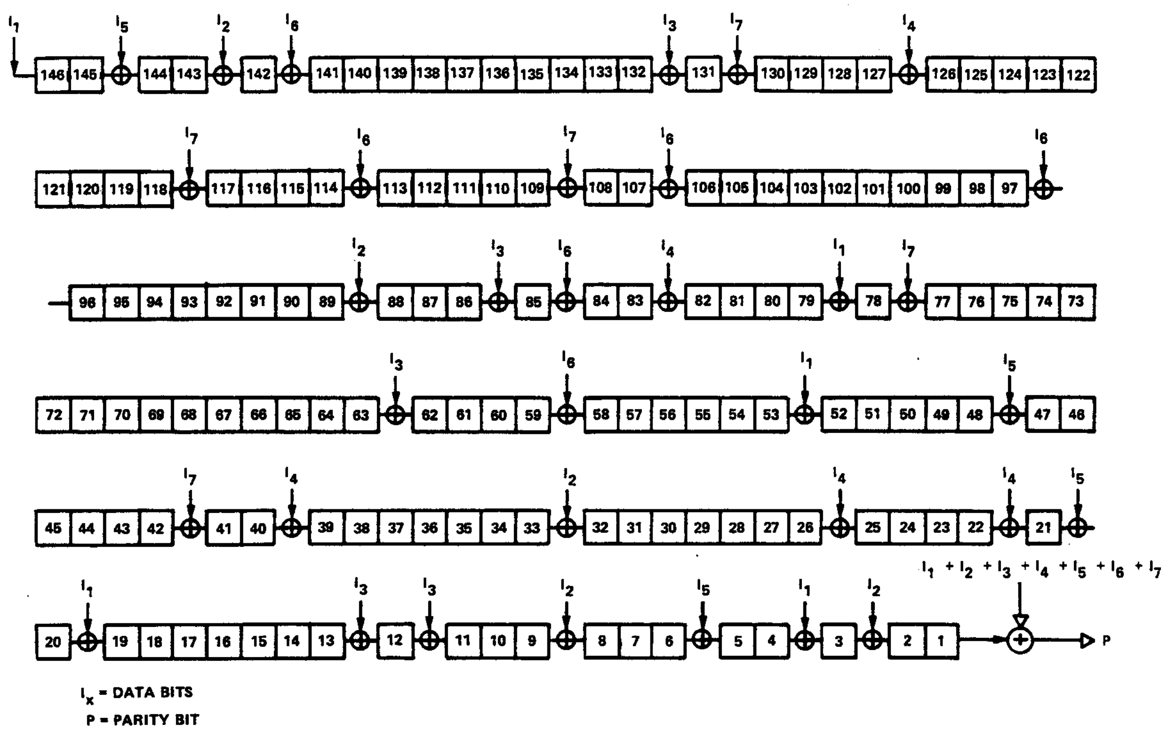


Figure 6. Parity Encoder Register

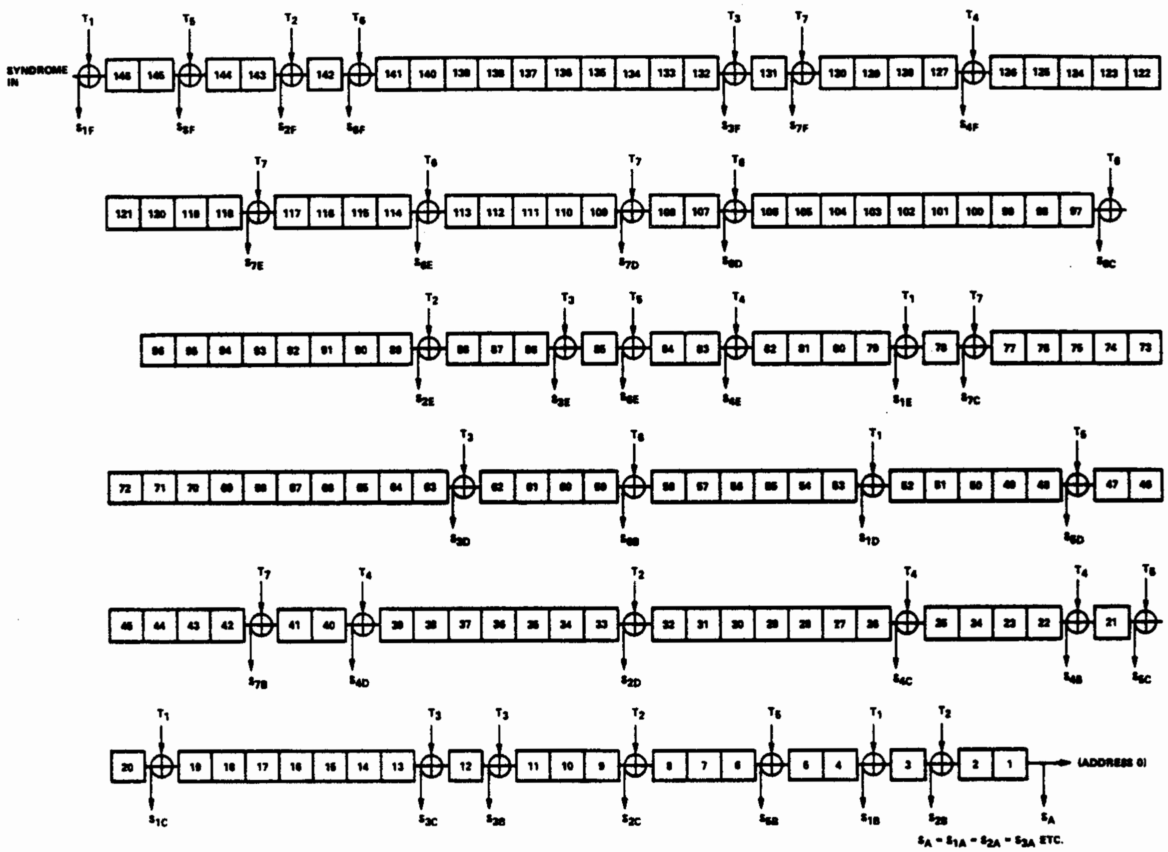


Figure 7. Syndrome Register Block Diagram

Conclusion

Coding is a cost-effective means of increasing the energy efficiency of a power-limited digital data communications link. Various system parameters and characteristics must be taken into account when optimizing the data link for the users' needs. Many different methods of coding have been developed and are available for meeting the unique requirements of various digital communication link configurations.

Appendix

Eb/No dB	P(e) 0.00	0.01	0.02	0.03	0.04
0.0	7.8650E-02	7.8411E-02	7.8172E-02	7.7934E-02	7.7696E-02
0.1	7.6274E-02	7.6038E-02	7.5802E-02	7.5567E-02	7.5332E-02
0.2	7.3927E-02	7.3694E-02	7.3461E-02	7.3228E-02	7.2996E-02
0.3	7.1609E-02	7.1379E-02	7.1149E-02	7.0920E-02	7.0690E-02
0.4	6.9322E-02	6.9094E-02	6.8868E-02	6.8641E-02	6.8415E-02
0.5	6.7065E-02	6.6841E-02	6.6618E-02	6.6395E-02	6.6172E-02
0.6	6.4841E-02	6.4620E-02	6.4400E-02	6.4180E-02	6.3961E-02
0.7	6.2650E-02	6.2432E-02	6.2216E-02	6.1999E-02	6.1783E-02
0.8	6.0492E-02	6.0278E-02	6.0065E-02	5.9852E-02	5.9639E-02
0.9	5.8369E-02	5.8159E-02	5.7949E-02	5.7739E-02	5.7530E-02
1.0	5.6282E-02	5.6075E-02	5.5869E-02	5.5663E-02	5.5457E-02
1.1	5.4231E-02	5.4028E-02	5.3825E-02	5.3622E-02	5.3420E-02
1.2	5.2216E-02	5.2017E-02	5.1818E-02	5.1619E-02	5.1421E-02
1.3	5.0239E-02	5.0044E-02	4.9849E-02	4.9654E-02	4.9459E-02
1.4	4.8301E-02	4.8109E-02	4.7918E-02	4.7727E-02	4.7536E-02
1.5	4.6401E-02	4.6213E-02	4.6026E-02	4.5839E-02	4.5652E-02
1.6	4.4541E-02	4.4357E-02	4.4174E-02	4.3991E-02	4.3808E-02
1.7	4.2721E-02	4.2541E-02	4.2362E-02	4.2183E-02	4.2004E-02
1.8	4.0942E-02	4.0766E-02	4.0591E-02	4.0416E-02	4.0241E-02
1.9	3.9203E-02	3.9031E-02	3.8860E-02	3.8690E-02	3.8519E-02
2.0	3.7506E-02	3.7339E-02	3.7172E-02	3.7005E-02	3.6839E-02
2.1	3.5851E-02	3.5688E-02	3.5525E-02	3.5363E-02	3.5201E-02
2.2	3.4238E-02	3.4079E-02	3.3921E-02	3.3762E-02	3.3605E-02
2.3	3.2668E-02	3.2513E-02	3.2359E-02	3.2205E-02	3.2051E-02
2.4	3.1140E-02	3.0990E-02	3.0840E-02	3.0690E-02	3.0541E-02
2.5	2.9655E-02	2.9509E-02	2.9363E-02	2.9218E-02	2.9073E-02
2.6	2.8214E-02	2.8072E-02	2.7930E-02	2.7789E-02	2.7649E-02
2.7	2.6815E-02	2.6678E-02	2.6540E-02	2.6404E-02	2.6268E-02
2.8	2.5460E-02	2.5326E-02	2.5194E-02	2.5061E-02	2.4930E-02
2.9	2.4147E-02	2.4019E-02	2.3890E-02	2.3762E-02	2.3635E-02
3.0	2.2878E-02	2.2754E-02	2.2630E-02	2.2506E-02	2.2383E-02
3.1	2.1652E-02	2.1532E-02	2.1412E-02	2.1293E-02	2.1174E-02
3.2	2.0469E-02	2.0353E-02	2.0237E-02	2.0122E-02	2.0007E-02
3.3	1.9328E-02	1.9216E-02	1.9105E-02	1.8994E-02	1.8883E-02
3.4	1.8229E-02	1.8122E-02	1.8015E-02	1.7908E-02	1.7802E-02
3.5	1.7172E-02	1.7069E-02	1.6966E-02	1.6864E-02	1.6761E-02
3.6	1.6157E-02	1.6058E-02	1.5959E-02	1.5861E-02	1.5763E-02
3.7	1.5183E-02	1.5088E-02	1.4993E-02	1.4899E-02	1.4805E-02
3.8	1.4249E-02	1.4158E-02	1.4067E-02	1.3977E-02	1.3887E-02
3.9	1.3355E-02	1.3268E-02	1.3181E-02	1.3095E-02	1.3009E-02
4.0	1.2501E-02	1.2417E-02	1.2335E-02	1.2252E-02	1.2170E-02
4.1	1.1685E-02	1.1605E-02	1.1526E-02	1.1448E-02	1.1369E-02
4.2	1.0907E-02	1.0831E-02	1.0756E-02	1.0681E-02	1.0606E-02
4.3	1.0167E-02	1.0094E-02	1.0023E-02	9.9515E-03	9.8805E-03
4.4	9.4623E-03	9.3939E-03	9.3258E-03	9.2581E-03	9.1907E-03
4.5	8.7938E-03	8.7289E-03	8.6643E-03	8.6001E-03	8.5362E-03
4.6	8.1600E-03	8.0985E-03	8.0373E-03	7.9765E-03	7.9160E-03
4.7	7.5601E-03	7.5019E-03	7.4440E-03	7.3865E-03	7.3293E-03
4.8	6.9930E-03	6.9381E-03	6.8835E-03	6.8292E-03	6.7752E-03
4.9	6.4580E-03	6.4062E-03	6.3547E-03	6.3035E-03	6.2527E-03
5.0	5.9539E-03	5.9051E-03	5.8567E-03	5.8086E-03	5.7607E-03

Appendix

Eb/No dB	P(e) 0.05	0.06	0.07	0.08	0.09
0.0	7.7458E-02	7.7221E-02	7.6984E-02	7.6747E-02	7.6510E-02
0.1	7.5097E-02	7.4862E-02	7.4628E-02	7.4394E-02	7.4160E-02
0.2	7.2764E-02	7.2533E-02	7.2301E-02	7.2070E-02	7.1839E-02
0.3	7.0461E-02	7.0233E-02	7.0005E-02	6.9777E-02	6.9549E-02
0.4	6.8189E-02	6.7964E-02	6.7739E-02	6.7514E-02	6.7289E-02
0.5	6.5949E-02	6.5727E-02	6.5505E-02	6.5283E-02	6.5062E-02
0.6	6.3741E-02	6.3522E-02	6.3304E-02	6.3085E-02	6.2867E-02
0.7	6.1567E-02	6.1351E-02	6.1136E-02	6.0921E-02	6.0706E-02
0.8	5.9426E-02	5.9214E-02	5.9003E-02	5.8791E-02	5.8580E-02
0.9	5.7321E-02	5.7113E-02	5.6904E-02	5.6697E-02	5.6489E-02
1.0	5.5252E-02	5.5047E-02	5.4842E-02	5.4638E-02	5.4434E-02
1.1	5.3219E-02	5.3018E-02	5.2817E-02	5.2616E-02	5.2416E-02
1.2	5.1223E-02	5.1026E-02	5.0828E-02	5.0632E-02	5.0435E-02
1.3	4.9265E-02	4.9072E-02	4.8878E-02	4.8686E-02	4.8493E-02
1.4	4.7346E-02	4.7156E-02	4.6967E-02	4.6778E-02	4.6589E-02
1.5	4.5466E-02	4.5280E-02	4.5095E-02	4.4910E-02	4.4725E-02
1.6	4.3626E-02	4.3444E-02	4.3263E-02	4.3082E-02	4.2901E-02
1.7	4.1826E-02	4.1648E-02	4.1471E-02	4.1294E-02	4.1118E-02
1.8	4.0067E-02	3.9893E-02	3.9720E-02	3.9547E-02	3.9375E-02
1.9	3.8349E-02	3.8180E-02	3.8011E-02	3.7842E-02	3.7674E-02
2.0	3.6673E-02	3.6508E-02	3.6343E-02	3.6179E-02	3.6015E-02
2.1	3.5039E-02	3.4878E-02	3.4717E-02	3.4557E-02	3.4397E-02
2.2	3.3448E-02	3.3291E-02	3.3134E-02	3.2978E-02	3.2823E-02
2.3	3.1898E-02	3.1746E-02	3.1594E-02	3.1442E-02	3.1291E-02
2.4	3.0392E-02	3.0244E-02	3.0096E-02	2.9949E-02	2.9802E-02
2.5	2.8929E-02	2.8785E-02	2.8642E-02	2.8498E-02	2.8356E-02
2.6	2.7509E-02	2.7369E-02	2.7230E-02	2.7091E-02	2.6953E-02
2.7	2.6132E-02	2.5997E-02	2.5862E-02	2.5727E-02	2.5593E-02
2.8	2.4798E-02	2.4667E-02	2.4537E-02	2.4406E-02	2.4277E-02
2.9	2.3508E-02	2.3381E-02	2.3255E-02	2.3129E-02	2.3003E-02
3.0	2.2260E-02	2.2137E-02	2.2016E-02	2.1894E-02	2.1773E-02
3.1	2.1055E-02	2.0937E-02	2.0819E-02	2.0702E-02	2.0585E-02
3.2	1.9893E-02	1.9779E-02	1.9666E-02	1.9553E-02	1.9440E-02
3.3	1.8773E-02	1.8664E-02	1.8554E-02	1.8446E-02	1.8337E-02
3.4	1.7696E-02	1.7590E-02	1.7485E-02	1.7381E-02	1.7276E-02
3.5	1.6660E-02	1.6558E-02	1.6457E-02	1.6357E-02	1.6257E-02
3.6	1.5665E-02	1.5568E-02	1.5471E-02	1.5374E-02	1.5278E-02
3.7	1.4711E-02	1.4618E-02	1.4525E-02	1.4433E-02	1.4341E-02
3.8	1.3797E-02	1.3708E-02	1.3619E-02	1.3531E-02	1.3443E-02
3.9	1.2923E-02	1.2838E-02	1.2753E-02	1.2669E-02	1.2584E-02
4.0	1.2088E-02	1.2007E-02	1.1926E-02	1.1845E-02	1.1765E-02
4.1	1.1291E-02	1.1214E-02	1.1137E-02	1.1060E-02	1.0983E-02
4.2	1.0532E-02	1.0458E-02	1.0385E-02	1.0312E-02	1.0239E-02
4.3	9.8099E-03	9.7397E-03	9.6698E-03	9.6003E-03	9.5311E-03
4.4	9.1237E-03	9.0570E-03	8.9907E-03	8.9247E-03	8.8591E-03
4.5	8.4726E-03	8.4094E-03	8.3465E-03	8.2840E-03	8.2218E-03
4.6	7.8559E-03	7.7960E-03	7.7365E-03	7.6774E-03	7.6186E-03
4.7	7.2725E-03	7.2160E-03	7.1597E-03	7.1038E-03	7.0483E-03
4.8	6.7216E-03	6.6682E-03	6.6152E-03	6.5625E-03	6.5101E-03
4.9	6.2021E-03	6.1519E-03	6.1019E-03	6.0523E-03	6.0029E-03
5.0	5.7131E-03	5.6659E-03	5.6189E-03	5.5722E-03	5.5259E-03

Appendix

Eb/No dB	P(e) 0.00	0.01	0.02	0.03	0.04
5.0	5.9539E-03	5.9051E-03	5.8567E-03	5.8086E-03	5.7607E-03
5.1	5.4798E-03	5.4340E-03	5.3885E-03	5.3432E-03	5.2983E-03
5.2	5.0346E-03	4.9916E-03	4.9490E-03	4.9065E-03	4.8644E-03
5.3	4.6173E-03	4.5771E-03	4.5371E-03	4.4974E-03	4.4580E-03
5.4	4.2269E-03	4.1893E-03	4.1520E-03	4.1149E-03	4.0780E-03
5.5	3.8623E-03	3.8272E-03	3.7923E-03	3.7578E-03	3.7234E-03
5.6	3.5224E-03	3.4897E-03	3.4572E-03	3.4250E-03	3.3931E-03
5.7	3.2061E-03	3.1757E-03	3.1455E-03	3.1156E-03	3.0859E-03
5.8	2.9123E-03	2.8842E-03	2.8562E-03	2.8285E-03	2.8009E-03
5.9	2.6401E-03	2.6140E-03	2.5882E-03	2.5625E-03	2.5370E-03
6.0	2.3884E-03	2.3643E-03	2.3404E-03	2.3166E-03	2.2931E-03
6.1	2.1559E-03	2.1337E-03	2.1117E-03	2.0898E-03	2.0681E-03
6.2	1.9419E-03	1.9214E-03	1.9012E-03	1.8811E-03	1.8612E-03
6.3	1.7452E-03	1.7264E-03	1.7078E-03	1.6894E-03	1.6711E-03
6.4	1.5648E-03	1.5476E-03	1.5306E-03	1.5137E-03	1.4970E-03
6.5	1.3998E-03	1.3841E-03	1.3686E-03	1.3532E-03	1.3379E-03
6.6	1.2492E-03	1.2349E-03	1.2207E-03	1.2067E-03	1.1928E-03
6.7	1.1121E-03	1.0991E-03	1.0862E-03	1.0734E-03	1.0608E-03
6.8	9.8751E-04	9.7572E-04	9.6403E-04	9.5247E-04	9.4102E-04
6.9	8.7466E-04	8.6399E-04	8.5342E-04	8.4296E-04	8.3261E-04
7.0	7.7267E-04	7.6304E-04	7.5351E-04	7.4408E-04	7.3474E-04
7.1	6.8075E-04	6.7208E-04	6.6350E-04	6.5502E-04	6.4662E-04
7.2	5.9812E-04	5.9034E-04	5.8264E-04	5.7503E-04	5.6750E-04
7.3	5.2404E-04	5.1708E-04	5.1019E-04	5.0338E-04	4.9665E-04
7.4	4.5782E-04	4.5161E-04	4.4546E-04	4.3939E-04	4.3338E-04
7.5	3.9880E-04	3.9326E-04	3.8780E-04	3.8240E-04	3.7706E-04
7.6	3.4634E-04	3.4143E-04	3.3658E-04	3.3179E-04	3.2706E-04
7.7	2.9986E-04	2.9551E-04	2.9123E-04	2.8699E-04	2.8281E-04
7.8	2.5880E-04	2.5497E-04	2.5119E-04	2.4746E-04	2.4377E-04
7.9	2.2264E-04	2.1928E-04	2.1596E-04	2.1268E-04	2.0944E-04
8.0	1.9091E-04	1.8796E-04	1.8505E-04	1.8218E-04	1.7935E-04
8.1	1.6315E-04	1.6057E-04	1.5803E-04	1.5553E-04	1.5306E-04
8.2	1.3894E-04	1.3670E-04	1.3449E-04	1.3232E-04	1.3017E-04
8.3	1.1791E-04	1.1597E-04	1.1406E-04	1.1217E-04	1.1031E-04
8.4	9.9706E-05	9.8028E-05	9.6374E-05	9.4745E-05	9.3140E-05
8.5	8.4000E-05	8.2555E-05	8.1132E-05	7.9731E-05	7.8351E-05
8.6	7.0501E-05	6.9262E-05	6.8042E-05	6.6841E-05	6.5659E-05
8.7	5.8943E-05	5.7885E-05	5.6843E-05	5.5818E-05	5.4809E-05
8.8	4.9086E-05	4.8185E-05	4.7299E-05	4.6427E-05	4.5570E-05
8.9	4.0712E-05	3.9948E-05	3.9198E-05	3.8459E-05	3.7733E-05
9.0	3.3627E-05	3.2983E-05	3.2349E-05	3.1727E-05	3.1115E-05
9.1	2.7658E-05	2.7116E-05	2.6584E-05	2.6061E-05	2.5548E-05
9.2	2.2650E-05	2.2197E-05	2.1752E-05	2.1315E-05	2.0885E-05
9.3	1.8467E-05	1.8089E-05	1.7719E-05	1.7355E-05	1.6997E-05
9.4	1.4989E-05	1.4675E-05	1.4368E-05	1.4066E-05	1.3770E-05
9.5	1.2109E-05	1.1850E-05	1.1597E-05	1.1348E-05	1.1104E-05
9.6	9.7362E-06	9.5236E-06	9.3152E-06	9.1110E-06	8.9108E-06
9.7	7.7905E-06	7.6166E-06	7.4463E-06	7.2794E-06	7.1160E-06
9.8	6.2027E-06	6.0613E-06	5.9227E-06	5.7871E-06	5.6542E-06
9.9	4.9135E-06	4.7990E-06	4.6869E-06	4.5772E-06	4.4698E-06
10.0	3.8721E-06	3.7799E-06	3.6896E-06	3.6014E-06	3.5150E-06

Appendix

Eb/No	P(e)				
dB	0.05	0.06	0.07	0.08	0.09
5.0	5.7131E-03	5.6659E-03	5.6189E-03	5.5722E-03	5.5259E-03
5.1	5.2536E-03	5.2093E-03	5.1652E-03	5.1214E-03	5.0778E-03
5.2	4.8225E-03	4.7810E-03	4.7396E-03	4.6986E-03	4.6578E-03
5.3	4.4188E-03	4.3799E-03	4.3413E-03	4.3029E-03	4.2648E-03
5.4	4.0414E-03	4.0051E-03	3.9690E-03	3.9332E-03	3.8976E-03
5.5	3.6893E-03	3.6554E-03	3.6218E-03	3.5884E-03	3.5553E-03
5.6	3.3613E-03	3.3298E-03	3.2985E-03	3.2675E-03	3.2367E-03
5.7	3.0565E-03	3.0272E-03	2.9982E-03	2.9693E-03	2.9407E-03
5.8	2.7736E-03	2.7465E-03	2.7196E-03	2.6929E-03	2.6664E-03
5.9	2.5118E-03	2.4867E-03	2.4618E-03	2.4371E-03	2.4126E-03
6.0	2.2697E-03	2.2466E-03	2.2236E-03	2.2009E-03	2.1783E-03
6.1	2.0467E-03	2.0253E-03	2.0042E-03	1.9833E-03	1.9625E-03
6.2	1.8414E-03	1.8218E-03	1.8024E-03	1.7832E-03	1.7641E-03
6.3	1.6530E-03	1.6351E-03	1.6173E-03	1.5996E-03	1.5821E-03
6.4	1.4804E-03	1.4640E-03	1.4478E-03	1.4316E-03	1.4156E-03
6.5	1.3228E-03	1.3078E-03	1.2929E-03	1.2782E-03	1.2636E-03
6.6	1.1790E-03	1.1654E-03	1.1518E-03	1.1385E-03	1.1252E-03
6.7	1.0483E-03	1.0359E-03	1.0236E-03	1.0115E-03	9.9943E-04
6.8	9.2968E-04	9.1845E-04	9.0734E-04	8.9634E-04	8.8544E-04
6.9	8.2236E-04	8.1222E-04	8.0218E-04	7.9224E-04	7.8241E-04
7.0	7.2550E-04	7.1636E-04	7.0732E-04	6.9837E-04	6.8951E-04
7.1	6.3832E-04	6.3010E-04	6.2198E-04	6.1394E-04	6.0598E-04
7.2	5.6005E-04	5.5269E-04	5.4541E-04	5.3821E-04	5.3108E-04
7.3	4.8999E-04	4.8341E-04	4.7690E-04	4.7047E-04	4.6411E-04
7.4	4.2745E-04	4.2158E-04	4.1579E-04	4.1006E-04	4.0439E-04
7.5	3.7178E-04	3.6657E-04	3.6142E-04	3.5633E-04	3.5131E-04
7.6	3.2239E-04	3.1777E-04	3.1321E-04	3.0870E-04	3.0425E-04
7.7	2.7868E-04	2.7460E-04	2.7058E-04	2.6660E-04	2.6267E-04
7.8	2.4014E-04	2.3655E-04	2.3300E-04	2.2950E-04	2.2605E-04
7.9	2.0625E-04	2.0310E-04	1.9999E-04	1.9692E-04	1.9390E-04
8.0	1.7656E-04	1.7380E-04	1.7108E-04	1.6840E-04	1.6576E-04
8.1	1.5062E-04	1.4822E-04	1.4585E-04	1.4352E-04	1.4121E-04
8.2	1.2805E-04	1.2597E-04	1.2391E-04	1.2188E-04	1.1988E-04
8.3	1.0848E-04	1.0667E-04	1.0489E-04	1.0314E-04	1.0141E-04
8.4	9.1559E-05	9.0001E-05	8.8467E-05	8.6955E-05	8.5466E-05
8.5	7.6991E-05	7.5653E-05	7.4335E-05	7.3037E-05	7.1759E-05
8.6	6.4495E-05	6.3349E-05	6.2221E-05	6.1111E-05	6.0018E-05
8.7	5.3816E-05	5.2839E-05	5.1878E-05	5.0932E-05	5.0001E-05
8.8	4.4726E-05	4.3896E-05	4.3080E-05	4.2278E-05	4.1488E-05
8.9	3.7020E-05	3.6318E-05	3.5628E-05	3.4950E-05	3.4283E-05
9.0	3.0513E-05	2.9922E-05	2.9341E-05	2.8770E-05	2.8209E-05
9.1	2.5043E-05	2.4547E-05	2.4060E-05	2.3582E-05	2.3112E-05
9.2	2.0464E-05	2.0050E-05	1.9643E-05	1.9244E-05	1.8852E-05
9.3	1.6647E-05	1.6303E-05	1.5965E-05	1.5633E-05	1.5308E-05
9.4	1.3480E-05	1.3195E-05	1.2916E-05	1.2642E-05	1.2373E-05
9.5	1.0864E-05	1.0630E-05	1.0400E-05	1.0174E-05	9.9530E-06
9.6	8.7145E-06	8.5222E-06	8.3337E-06	8.1489E-06	7.9679E-06
9.7	6.9558E-06	6.7989E-06	6.6452E-06	6.4946E-06	6.3472E-06
9.8	5.5242E-06	5.3968E-06	5.2721E-06	5.1500E-06	5.0305E-06
9.9	4.3647E-06	4.2619E-06	4.1612E-06	4.0627E-06	3.9664E-06
10.0	3.4305E-06	3.3479E-06	3.2671E-06	3.1881E-06	3.1108E-06

Appendix

Eb/No dB	P(e) 0.00	0.01	0.02	0.03	0.04
10.0	3.8721E-06	3.7799E-06	3.6896E-06	3.6014E-06	3.5150E-06
10.1	3.0352E-06	2.9613E-06	2.8891E-06	2.8184E-06	2.7493E-06
10.2	2.3663E-06	2.3074E-06	2.2499E-06	2.1936E-06	2.1387E-06
10.3	1.8346E-06	1.7879E-06	1.7423E-06	1.6978E-06	1.6543E-06
10.4	1.4142E-06	1.3774E-06	1.3416E-06	1.3065E-06	1.2723E-06
10.5	1.0838E-06	1.0550E-06	1.0270E-06	9.9954E-07	9.7281E-07
10.6	8.2572E-07	8.0329E-07	7.8143E-07	7.6011E-07	7.3933E-07
10.7	6.2523E-07	6.0788E-07	5.9096E-07	5.7449E-07	5.5843E-07
10.8	4.7048E-07	4.5713E-07	4.4413E-07	4.3147E-07	4.1915E-07
10.9	3.5178E-07	3.4157E-07	3.3164E-07	3.2198E-07	3.1258E-07
11.0	2.6131E-07	2.5356E-07	2.4603E-07	2.3870E-07	2.3158E-07
11.1	1.9281E-07	1.8697E-07	1.8129E-07	1.7577E-07	1.7041E-07
11.2	1.4130E-07	1.3692E-07	1.3267E-07	1.2854E-07	1.2453E-07
11.3	1.0283E-07	9.9569E-08	9.6410E-08	9.3344E-08	9.0369E-08
11.4	7.4294E-08	7.1889E-08	6.9557E-08	6.7296E-08	6.5104E-08
11.5	5.3287E-08	5.1524E-08	4.9815E-08	4.8160E-08	4.6556E-08
11.6	3.7934E-08	3.6651E-08	3.5409E-08	3.4206E-08	3.3042E-08
11.7	2.6798E-08	2.5871E-08	2.4975E-08	2.4108E-08	2.3269E-08
11.8	1.8783E-08	1.8119E-08	1.7477E-08	1.6857E-08	1.6258E-08
11.9	1.3060E-08	1.2588E-08	1.2132E-08	1.1692E-08	1.1267E-08
12.0	9.0060E-09	8.6735E-09	8.3525E-09	8.0427E-09	7.7438E-09
12.1	6.1585E-09	5.9261E-09	5.7019E-09	5.4857E-09	5.2773E-09
12.2	4.1752E-09	4.0141E-09	3.8589E-09	3.7094E-09	3.5653E-09
12.3	2.8058E-09	2.6951E-09	2.5886E-09	2.4861E-09	2.3874E-09
12.4	1.8686E-09	1.7933E-09	1.7208E-09	1.6511E-09	1.5841E-09
12.5	1.2330E-09	1.1822E-09	1.1334E-09	1.0865E-09	1.0414E-09
12.6	8.0599E-10	7.7203E-10	7.3943E-10	7.0814E-10	6.7811E-10
12.7	5.2178E-10	4.9931E-10	4.7776E-10	4.5709E-10	4.3727E-10
12.8	3.3447E-10	3.1974E-10	3.0564E-10	2.9212E-10	2.7917E-10
12.9	2.1224E-10	2.0269E-10	1.9355E-10	1.8480E-10	1.7643E-10
13.0	1.3329E-10	1.2716E-10	1.2130E-10	1.1569E-10	1.1034E-10
13.1	8.2830E-11	7.8935E-11	7.5215E-11	7.1662E-11	6.8270E-11
13.2	5.0917E-11	4.8469E-11	4.6134E-11	4.3907E-11	4.1782E-11
13.3	3.0954E-11	2.9433E-11	2.7984E-11	2.6603E-11	2.5287E-11
13.4	1.8606E-11	1.7671E-11	1.6782E-11	1.5935E-11	1.5129E-11
13.5	1.1055E-11	1.0487E-11	9.9473E-12	9.4343E-12	8.9467E-12
13.6	6.4904E-12	6.1498E-12	5.8263E-12	5.5191E-12	5.2275E-12
13.7	3.7646E-12	3.5627E-12	3.3711E-12	3.1894E-12	3.0172E-12
13.8	2.1566E-12	2.0383E-12	1.9263E-12	1.8202E-12	1.7197E-12
13.9	1.2198E-12	1.1514E-12	1.0867E-12	1.0255E-12	9.6767E-13
14.0	6.8102E-13	6.4199E-13	6.0512E-13	5.7029E-13	5.3739E-13
14.1	3.7518E-13	3.5320E-13	3.3246E-13	3.1290E-13	2.9445E-13
14.2	2.0389E-13	1.9168E-13	1.8018E-13	1.6934E-13	1.5913E-13
14.3	1.0927E-13	1.0258E-13	9.6286E-14	9.0367E-14	8.4800E-14
14.4	5.7727E-14	5.4115E-14	5.0722E-14	4.7535E-14	4.4542E-14
14.5	3.0055E-14	2.8134E-14	2.6330E-14	2.4639E-14	2.3053E-14
14.6	1.5416E-14	1.4409E-14	1.3465E-14	1.2581E-14	1.1753E-14
14.7	7.7873E-15	7.2670E-15	6.7804E-15	6.3254E-15	5.9000E-15
14.8	3.8725E-15	3.6081E-15	3.3612E-15	3.1306E-15	2.9154E-15
14.9	1.8952E-15	1.7629E-15	1.6396E-15	1.5246E-15	1.4175E-15
15.0	9.1240E-16	8.4730E-16	7.8672E-16	7.3035E-16	6.7790E-16

Appendix

Eb/No dB	P(e) 0.05	0.06	0.07	0.08	0.09
10.0	3.4305E-06	3.3479E-06	3.2671E-06	3.1881E-06	3.1108E-06
10.1	2.6818E-06	2.6158E-06	2.5513E-06	2.4882E-06	2.4266E-06
10.2	2.0850E-06	2.0325E-06	1.9813E-06	1.9312E-06	1.8823E-06
10.3	1.6119E-06	1.5704E-06	1.5300E-06	1.4905E-06	1.4519E-06
10.4	1.2390E-06	1.2064E-06	1.1746E-06	1.1436E-06	1.1134E-06
10.5	9.4673E-07	9.2130E-07	8.9650E-07	8.7231E-07	8.4872E-07
10.6	7.1907E-07	6.9933E-07	6.8008E-07	6.6133E-07	6.4305E-07
10.7	5.4279E-07	5.2756E-07	5.1272E-07	4.9827E-07	4.8419E-07
10.8	4.0715E-07	3.9547E-07	3.8410E-07	3.7303E-07	3.6226E-07
10.9	3.0344E-07	2.9454E-07	2.8588E-07	2.7746E-07	2.6927E-07
11.0	2.2465E-07	2.1792E-07	2.1137E-07	2.0501E-07	1.9882E-07
11.1	1.6520E-07	1.6014E-07	1.5522E-07	1.5044E-07	1.4580E-07
11.2	1.2064E-07	1.1686E-07	1.1320E-07	1.0964E-07	1.0618E-07
11.3	8.7482E-08	8.4682E-08	8.1965E-08	7.9329E-08	7.6773E-08
11.4	6.2978E-08	6.0917E-08	5.8919E-08	5.6983E-08	5.5106E-08
11.5	4.5002E-08	4.3497E-08	4.2039E-08	4.0626E-08	3.9259E-08
11.6	3.1914E-08	3.0823E-08	2.9767E-08	2.8745E-08	2.7755E-08
11.7	2.2458E-08	2.1673E-08	2.0913E-08	2.0179E-08	1.9469E-08
11.8	1.5678E-08	1.5118E-08	1.4577E-08	1.4053E-08	1.3548E-08
11.9	1.0856E-08	1.0460E-08	1.0077E-08	9.7074E-09	9.3505E-09
12.0	7.4553E-09	7.1770E-09	6.9085E-09	6.6494E-09	6.3995E-09
12.1	5.0763E-09	4.8826E-09	4.6959E-09	4.5159E-09	4.3424E-09
12.2	3.4265E-09	3.2928E-09	3.1641E-09	3.0401E-09	2.9207E-09
12.3	2.2924E-09	2.2009E-09	2.1130E-09	2.0284E-09	1.9469E-09
12.4	1.5197E-09	1.4577E-09	1.3982E-09	1.3409E-09	1.2859E-09
12.5	9.9809E-10	9.5650E-10	9.1655E-10	8.7819E-10	8.4136E-10
12.6	6.4929E-10	6.2163E-10	5.9509E-10	5.6963E-10	5.4521E-10
12.7	4.1828E-10	4.0006E-10	3.8260E-10	3.6587E-10	3.4983E-10
12.8	2.6678E-10	2.5490E-10	2.4353E-10	2.3264E-10	2.2222E-10
12.9	1.6842E-10	1.6075E-10	1.5342E-10	1.4641E-10	1.3971E-10
13.0	1.0522E-10	1.0032E-10	9.5645E-11	9.1177E-11	8.6908E-11
13.1	6.5031E-11	6.1939E-11	5.8988E-11	5.6171E-11	5.3482E-11
13.2	3.9756E-11	3.7824E-11	3.5981E-11	3.4225E-11	3.2550E-11
13.3	2.4033E-11	2.2839E-11	2.1702E-11	2.0619E-11	1.9588E-11
13.4	1.4363E-11	1.3633E-11	1.2939E-11	1.2279E-11	1.1652E-11
13.5	8.4832E-12	8.0428E-12	7.6244E-12	7.2268E-12	6.8491E-12
13.6	4.9507E-12	4.6880E-12	4.4386E-12	4.2021E-12	3.9776E-12
13.7	2.8539E-12	2.6990E-12	2.5523E-12	2.4132E-12	2.2815E-12
13.8	1.6245E-12	1.5345E-12	1.4492E-12	1.3685E-12	1.2921E-12
13.9	9.1294E-13	8.6119E-13	8.1227E-13	7.6603E-13	7.2232E-13
14.0	5.0633E-13	4.7699E-13	4.4929E-13	4.2315E-13	3.9847E-13
14.1	2.7705E-13	2.6064E-13	2.4517E-13	2.3059E-13	2.1684E-13
14.2	1.4952E-13	1.4047E-13	1.3195E-13	1.2392E-13	1.1637E-13
14.3	7.9564E-14	7.4640E-14	7.0011E-14	6.5660E-14	6.1570E-14
14.4	4.1730E-14	3.9091E-14	3.6613E-14	3.4287E-14	3.2104E-14
14.5	2.1566E-14	2.0172E-14	1.8865E-14	1.7640E-14	1.6492E-14
14.6	1.0978E-14	1.0253E-14	9.5735E-15	8.9381E-15	8.3435E-15
14.7	5.5024E-15	5.1307E-15	4.7834E-15	4.4589E-15	4.1557E-15
14.8	2.7146E-15	2.5272E-15	2.3523E-15	2.1892E-15	2.0371E-15
14.9	1.3177E-15	1.2247E-15	1.1381E-15	1.0574E-15	9.8232E-16
15.0	6.2911E-16	5.8374E-16	5.4154E-16	5.0231E-16	4.6584E-16

Bibliography

[1]E.R. Berlekamp, Algebraic Coding Theory, New York: McGraw-Hill, 1968.

[2]R.T. Chien, Cyclic Decoding Procedures for Bose-Chaudhuri-Hocquenghem Codes," IEEE Transactions on Information Theory, IT-10, No. 4, October 1964, pp. 357-363.

[3]H.O. Burton, "Inversionless Decoding of Binary BCH Codes," IEEE Transactions on Information Theory, IT-17, No. 4, July 1977, pp. 464-466.

[4]A.J. Viterbi, "Error Bound for Convolutional Codes and an asymptotically Optimum Decoding Algorithm," IEEE Transactions on Information Theory, IT-13, No. 2, April 1967, pp. 260-269.

[5]J.B. Cain, G.C. Clark, and J.M. Geist, "Punctured Convolutional Codes of Rate (n-1)/n and Simplified Maximum Likelihood Decoding," IEEE Transactions on Information Theory, IT-25, No. 1, January 1979, pp. 97-100.

[6]P.R. Chevillat and D.J. Costello, Jr., "A Multiple Stack Algorithm for Erasurefree Decoding of Convolutional Codes," IEEE Transactions on Information Theory, CMD-25, No. 12, December 1977, pp. 1460-1470.

The SAbus Interface — A System Solution

G. Hammond/B. Reifman

Introduction

In recent years, more satellite earth station owners are requiring remote monitor and control capability to effectively manage their systems. The most common approach has been to provide individual contact-closure interfaces from each piece of equipment to a central control panel. Although this approach is simple to design, it is generally inflexible and requires significant effort in system implementation due to the wiring effort alone.

The SAbus is an interface designed explicitly as an economical method of providing complex remote monitor and control capability in satellite earth station equipment. Although it requires a limited degree of "intelligence" in each earth station component, this provides comprehensive monitor and control capabilities and a degree of flexibility not available in customized system solutions. The SAbus provides a powerful base for the design of both small and large earth station systems, and in either case significantly reduces the effort required for future system changes.

Electrical Specifications

Electrically, the SAbus is compatible with EIA RS-422, which is an interface specification that will eventually replace RS-232 as the industry standard for computer interconnection. RS-422 is a unipolar, balanced, 5-volt serial interface designed to connect equipment which must exchange data over considerable distances with high-noise immunity and high speed. Standard IC drivers and receivers are available for RS-422 that convert to and from TTL logic levels. The SAbus subset of RS-422 allows up to 64 devices to be connected in parallel with up to 4,000 feet between any master and group of slaves.

Physical Specifications

The physical implementation of the SAbus interface takes the form of a single 9-pin "D" connector located on the rear panel of a compatible device. This connector and its wiring is compatible with EIA RS-449, which is the mechanical specification for RS-422/423-compatible equipment. The 9-pin connector chosen for the SAbus connector is described as the secondary interface in RS-449 and has only the four data lines and circuit, common and shield. No hardware handshaking is used in the SAbus protocol, so all the control lines specified for the standard 37-pin connector are not needed. An SAbus compatible device that is only capable of operating as a slave has a female connector, whereas masters have male connectors. All SAbus devices can operate in electrical parallel with only a single 9-conductor cable required to connect all devices controlled by a master. Figure 1 illustrates the connection of a master and multiple slaves.

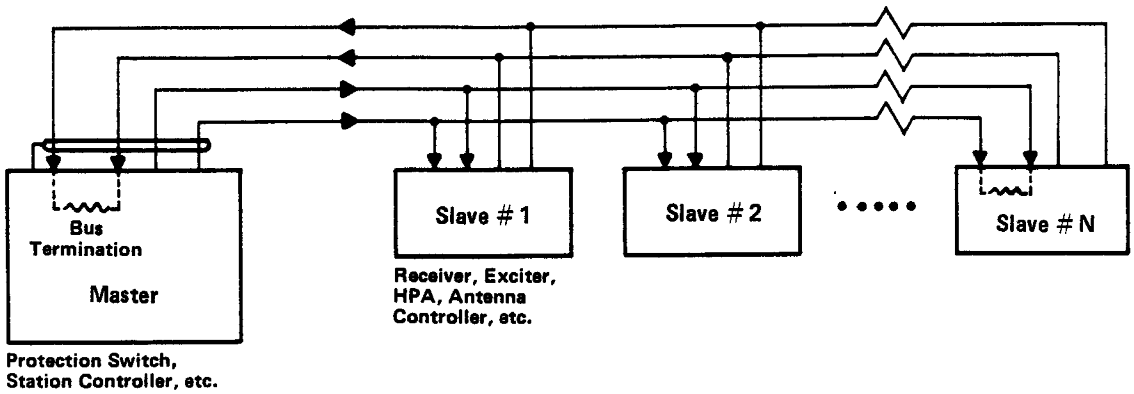


Figure 1. SAbus Master and Slaves

SAbus Protocol

The SAbus interface is a multi-drop, balanced line, asynchronous, full-duplex communications link designed to interconnect equipment for remote control and switching applications. Products that are SAbus compatible can be linked together over a parallel-connected 4-wire circuit without regard to their particular function.

Each SAbus configuration can have one master and up to 63 slave devices. Each slave device is internally configured to respond to a unique address. A master could be a protection switch, earth station controller, or any micro- or mini-computer that is electrically and operationally compatible with the SAbus. Since the electrical specifications are very similar to EIA standards RS-422 and RS-449, virtually any computer that meets these standards is capable of controlling remote devices over the SAbus.

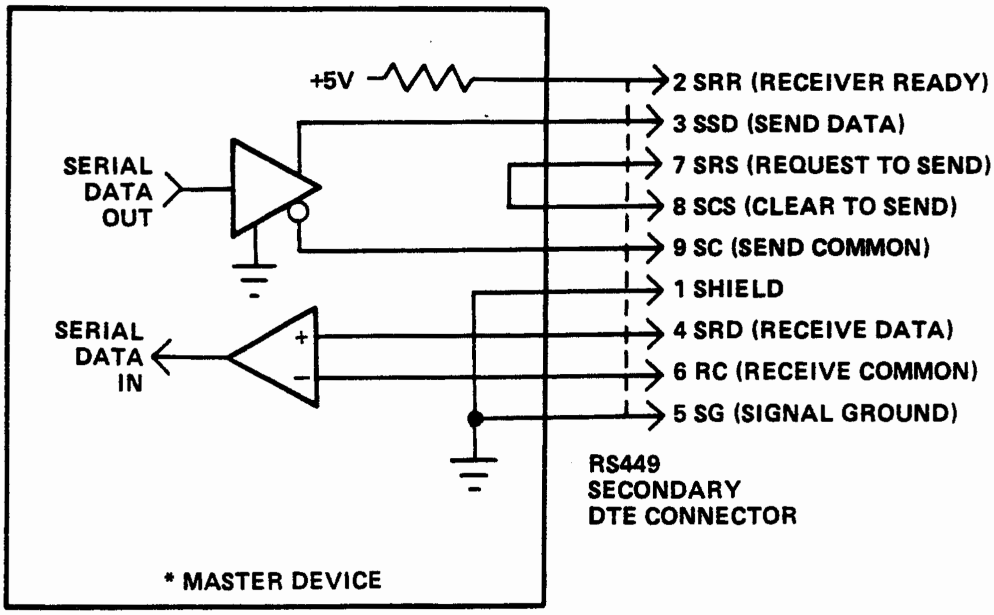
Data Format

SAbus data format supports industry's standard asynchronous ASCII format with one start bit, eight data bits (7-bit ASCII with 8th sent as even parity), and one stop bit. The ASCII control character subset 00-1F (hex) are reserved for message control. The printable ASCII characters 20-7F (hex) are used for address, command and data characters. The standard bus data rate via direct connect (up to 4,000 ft) is 9,600 BAUD, the data rate for devices connect to a master via modem is 1,200 BAUD.

Message Protocol

Message format and protocol over the SAbus is a derivative of IBM's binary synchronous communications protocol (BISYNC). The master station sends a command over the bus to all remote stations. The station, whose address is contained in the second byte of the command message, carries out the requested commands and then replies with a response message containing its own address and status information relating to its present condition. A remote station only sends a response following a command containing its unique address from the master. This prevents bus contention caused by more than one remote device communicating over the SAbus as the same time.

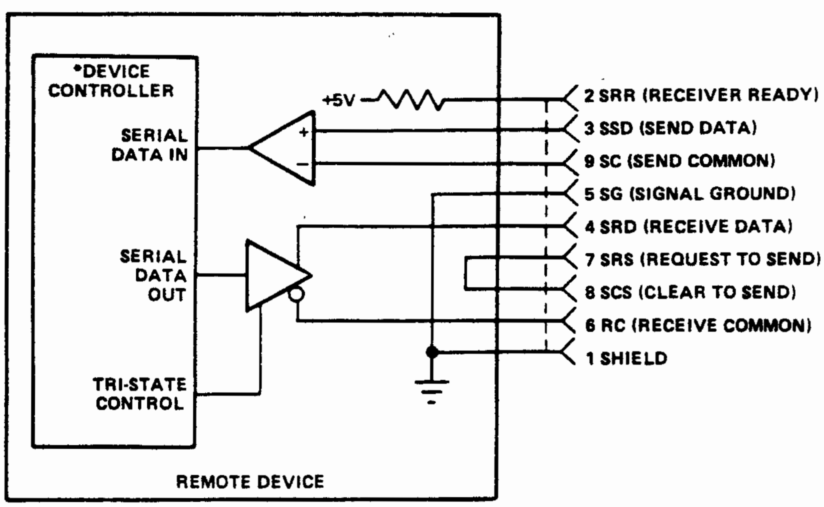
A remote device ignores all commands that contain parity or checksum errors, protocol errors, a wrong address, or message overrun errors. A remote device replies with a not-acknowledged (NAK) character if it receives an invalid command or data.



* MASTER DEVICE MAY BE A PROTECTION SWITCH, STATION CONTROLLER OR ESSENTIALLY ANY GENERAL PURPOSE MINI OR MICRO-COMPUTER.

55-A-4586

Figure 2. SAbus Master Device Connections



* THE DEVICE CONTROLLER WILL TYPICALLY BE SINGLE CHIP MICROCOMPUTER INSTALLED IN EXISTING PRODUCTS OR MAY BE A FUNCTION OF A MICROPROCESSOR IN NEW MICROPROCESSOR BASED EQUIPMENT (SUCH AS ANTENNA CONTROLLER).

55-A-4587

Figure 3. SAbus Slave Device Connections

Message Format

Command messages (see Figure 4) begin with Start-of-text byte, STX, followed by a remote address, a command byte and multiple data bytes. The End-of-text byte, ETX, is sent following the last data byte, and the message is terminated by a checksum character. Response messages are identical to command messages in format with the exception of the ACK (Acknowledge) or NAK (Not Acknowledge) character at the start of the message instead of STX. Figure 4 illustrates the format of the command and response messages. A command or reply message may have a variable length up to a maximum of 132 bytes, including delimiters and checksum. Although most currently implemented SAbus devices require no or very few data bytes, the capability for long messages is built into the protocol, so that future applications requiring the transfer of large amounts of data can be accommodated. SAbus devices should observe the length of all messages, predefined by their communication protocol, and NAK messages longer than permitted.

COMMAND MESSAGE:

STX	ADDRESS	COMMAND	D_1	D_2	D_n	D_{n+1}	ETX	CHKSUM

RESPONSE MESSAGE: COMMAND ACKNOWLEDGED

ACK	ADDRESS	COMMAND	D_1	D	D_n	D_{n+1}	ETX	CHKSUM

RESPONSE MESSAGE: COMMAND NOT ACKNOWLEDGED-UNABLE TO EXECUTE OR INCORRECT COMMAND

NAK	ADDRESS	COMMAND	ETX	CHKSUM

55-A-4588

Figure 4. SAbus Message Format

Message Delimiters

A command message begins with STX (02 hex), the ASCII Start-of-text control character. A message-acknowledged reply begins with ACK (06 hex), the ASCII Acknowledge control character and a message-not acknowledged reply begins with NAK (15 hex), the ASCII Not Acknowledge control character. All messages end with the ETX (03 hex), the ASCII End-of-text control character, followed by the checksum byte.

Address Character

The device address must be a valid ASCII printable character between "0" and "o", or 30 through 6F in hex; thus, 64 SAbus addresses are possible. Address 0 (ASCII 30) is reserved as an "all call" address and will cause all devices on the bus to execute a command without generating a reply.

Command Character

The command character (CMD) immediately follows the device address and specifies one of a possible 80 different commands for a particular device. Values from 30 to 7F (hex) are allowed. Commands may be completely device dependent with the exception of command 30 (hex) which must cause a device to return its six character device type and command 31 (hex) which is a status poll.

Command and Reply Data

A command or device reply may contain from 0 to 128 data characters and is restricted only to printable ACII characters 20-7F (hex).

Check Character

The last character of any SAbus message is the check character (CHK). This character is simply the bit-by-bit exclusive OR of all characters in the message starting with the STX character through the ETX character. This forms a Longitudinal Redundancy parity check over the entire message.

Message Timing

Different devices will require varying times to execute commands from a master. A receiver, for example, may be instructed to change frequency and may require up to a second for the synthesizer to lock. This should not, however, prevent it from immediately acknowledging the command. The NAK or ACK reply does not signify that a function has actually taken place, but only that the message was received and understood. A status reply should indicate if a device is executing a time-consuming function.

A remote device must begin responding to a command within 100 milliseconds after receiving the last character of the command and no more than 10 milliseconds must pass between each character. If remote device does not respond within this time, the master-controller should attempt to re-establish communication by repolling this device at least once. Figures 5 and 6 show a remote device SAbus state table and timing requirements.

At least a 10-bit time delay must be inserted between command messages in order to "wake up" a remote device. Once device is awakened by data on the bus, it looks for STX followed by its address. If it does not see its own address, it ignores the rest of the message by going to "sleep" and remains in such state until the serial data line idles for at least 10-bit times or approximately 10 milliseconds.

SAbus Command Restrictions

All Sabus-compatible devices must respond to a command "0", 30 (hex), with 6 data bytes of ASCII characters in the following form:

ACK ADDR 30 D1 D2 D3 D4 D5 D6 ETX CHSUM

where D1-D4 are four ASCII characters representing the model number and D5-D6 are two ASCII characters representing a software version number.

If more than one command is required to obtain status information of device's functions that can cause setting of a change bit, then the device must implement a clear change bit command and this must be the only command which causes the change bit to be cleared. If several commands have to be executed in order to set all the information that can cause a change bit to be set, then multiple change bits may be used to reduce the bus traffic.

Wherever possible, SAbus numeric data should be sent encoded as ASCII data characters, and only in cases where it cannot be avoided, numeric data should be sent in Binary or BCD packed format. Status bits in data bytes, i.e. change bits, alarm bits, etc., should occupy no more than four bits in the low-order nibble. The high-order nibble should be set to 03 to guarantee that the byte will contain a printable ASCII character.

Slave Device State Diagram

Introduction

General Description. The Slave State diagram (see Figure 5) presents the required protocol implementation at the Slave device that guarantees the proper transfer and processing of communication messages sent by a Master-controller over SAbus.

State Diagram Notation. Each state that a slave device can assume is represented graphically as a circle. A single-digit number is used within the circle to identify the state.

All permissible transitions between states are represented graphically by arrows between them. Each transition is qualified by a condition that must be true in order for the transition to occur. The device will remain in its current state if conditions which qualify transitions leading to other states are false or conditions that qualify pseudo-transitions are true. A pseudo-transition is a transition that occurs within the same state and is represented graphically by arrows leaving from and arriving at the same state. Table 1 describes mnemonics used to identify transitions in the state diagram.

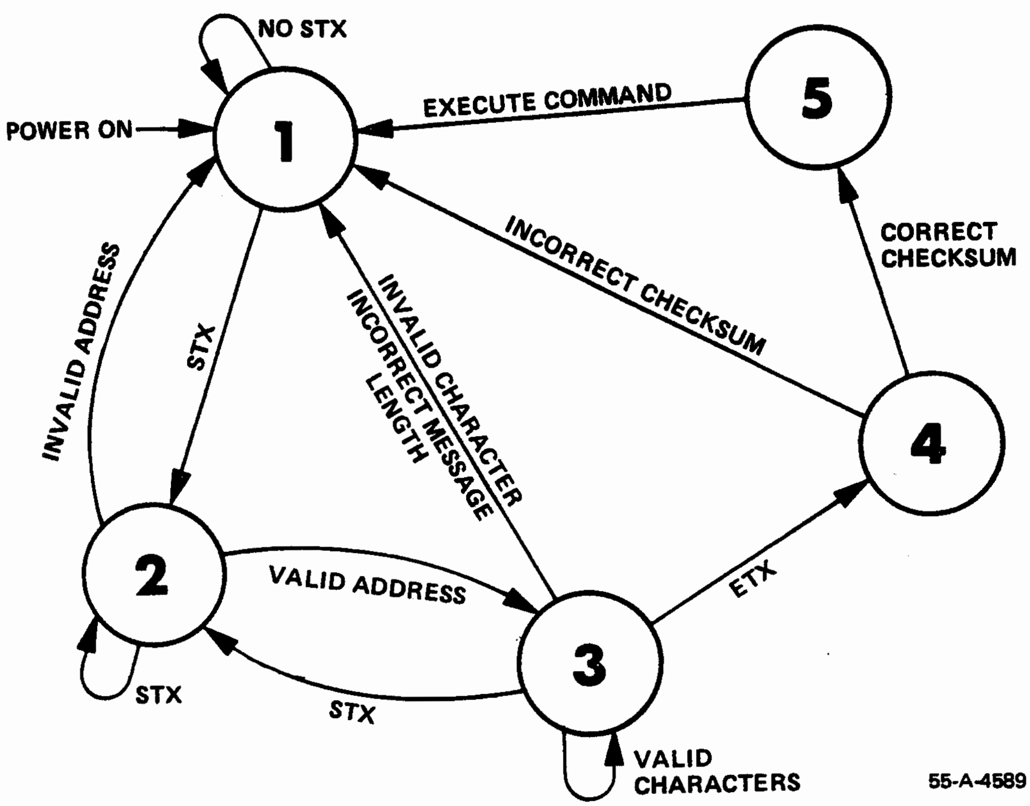


Figure 5. Slave State Diagram

Table 1. State Diagram Mnemonics

Mnemonics	Description
STX	Start-of-Text ASCII control character, used as a header in SAbus command message to identify the beginning of a new message.
ETX	End-of-Text ASCII control character, used as a termination character in SAbus messages to identify the end of data.
Checksum	LRC byte (Longitudinal Redundancy Check) is a last byte in the SAbus message data block. The value of LRC byte is the exclusive OR of all message bytes including the STX and the ETX bytes and is used to detect errors during transmission of data.

States Description

State 1 (Device Idle State). In State 1, a device is ready to receive a new message, and therefore, must complete any previous message reception. A device always powers on in State 1.

A device will exit State 1 and enter State 2 (Device Addressed State) only if STX byte is received.

State 2 (Device Addressed State). In State 2, a device is waiting to receive the address byte, the second byte of SAbus command message.

A device will exit State 2 and enter:

a. State 3 (Device Data State) if received address byte equals a device's address.

b. State 1 (Device Idle State) if received address byte does not equal a device's address.

c. State 2 (remain in current state) if STX byte is received, which may be the beginning of a new message data block.

State 3 (Device Data State). In State 3, a device is engaged in receiving the command and associated data bytes sent by a master-controller.

A device will exit State 3 and enter:

a. State 4 (Device Data Error State) if ETX byte is received signifying the end of data in the message.

b. State 1 (Device Idle State) if invalid command, or data character, or incorrect number of data bytes is received.

State 4 (Device Data Error State). In State 4, a device is waiting to receive a Checksum byte which tests the transmitted message for errors.

A device will exit State 4 and enter:

a. State 5 (Command Execute State) if a Checksum byte is true -- received LRC value of Checksum byte equals to LRC value computed by a device during message reception.

b. State 1 (Device Idle State) if a Checksum byte is false -- received LRC value of Checksum byte does not equal to LRC value computed by a device during message reception.

State 5 (Command Execute State). In State 5, a device, having completed a reception of SAbus message, executes a device's function specified by a command byte. A device will send an appropriate response message to a master-controller within 100 milliseconds after receiving the last character of the message.

A device will always exit State 5 and enter Device Idle State, State 1.

Video Plus

J.M. Hooper/F.L. McCormick

Introduction

Perhaps the most difficult part of putting together a satellite network is finding available space segment. Even with the launch of new satellites, transponders, or portions of transponders, are not readily available to the small business user. This problem will never be completely eliminated because of the finite availability of geostationary orbital arc. Thus, we must find ways to use this limited resource more efficiently.

Video-Plus, a new and exciting technique being developed by Scientific-Atlanta, may provide a significant amount of previously unused satellite capacity in existing transponders. The concept is very simple. A transponder which carries a video signal can be shared with other carriers, such as 56 kb/s digital SCPC and analog FM SCPC, without interference to the video signal of the SCPC carriers. Furthermore, this can be accomplished without any modification to the modulation format of the video signal. Thus, transponders which have previously carried only one video signal may be used to carry many additional voice and data signals without interference.

The significance of the Video-Plus technique is obvious; with this technique much more transponder space is available than we have previously realized! Video-Plus provides new space segment availability for the user and more potential revenues for the transponder owners.

Video-Plus is not a subcarrier technique. SCPC carriers can be transmitted from uplinks completely separate from the video uplink. Hence, the signals do not have to be combined at one origination site. For example, a hotel using a 4.6-meter dish to receive entertainment programming from SATCOM 3R could add transmit capability to its earth station and share one of the video transponders. A typical application might be hotel reservations or intercity telephone service for guests. Thus, the transponder is utilized from many uplink locations.

The Video-Plus technique is not without its limitations. These are explained in detail in the next section. However, simply stated, when carriers are added to a transponder which is carrying a saturated video signal, the power of the transponder is shared among all carriers. Therefore, a limit exists at which point degradation of the video signal begins to occur. This limit must be carefully controlled so that picture quality is preserved at all receiving sites. Analysis, lab testing, and satellite testing show that a significant amount of transponder capacity is available for digital and FM SCPC carriers to be added to video.

The next section will discuss the technical details of the Video-Plus technique and provide supporting calculations. The last section of this paper will describe some applications where Video Plus can be used effectively.

Video-Plus System Parameters

Frequency Plan/Bandwidth Requirements

A typical single-transponder video uplink earth station transmits a carrier at the center frequency of the satellite transponder at a level that provides a saturated downlink carrier. Figure 1 illustrates the frequency plan for this arrangement. The modulated video carrier with a 10.75 MHz peak deviation requires a bandwidth of about 30 MHz to provide low distortion transmission of program material or test waveforms. Of the usable 36 MHz transponder bandwidth, 3 MHz is available on either side of the video carrier. Intermodulation products restrict the placement of SCPC carriers to only one side of the transponder.

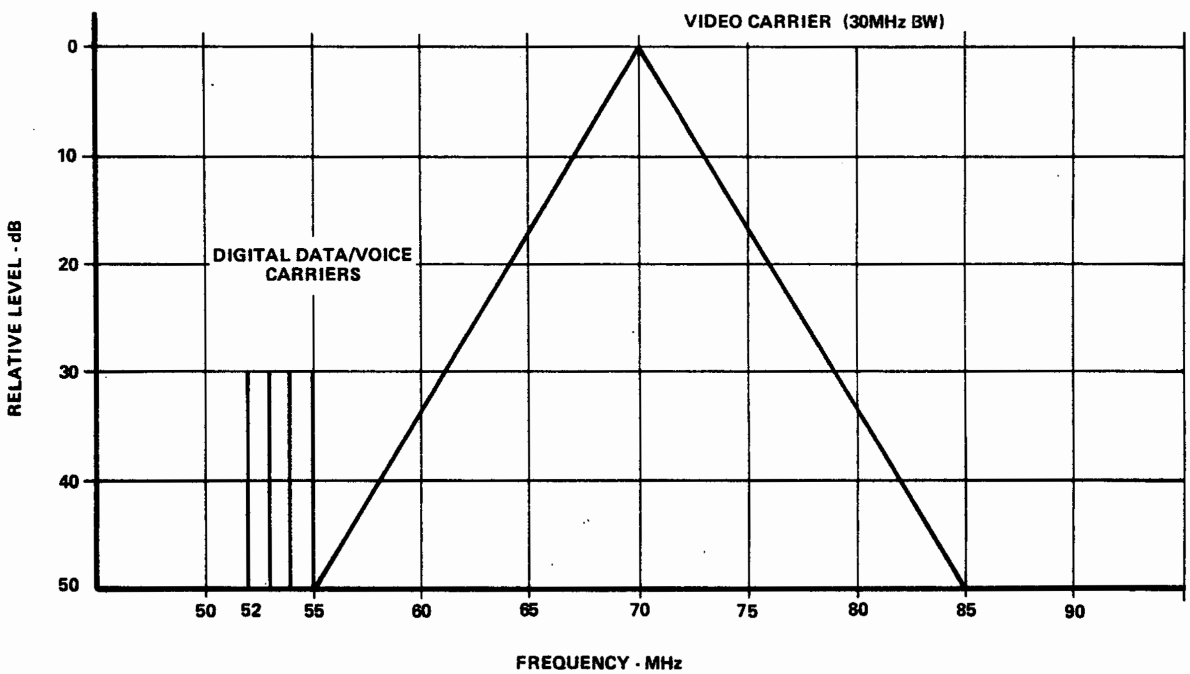


Figure 1. Transponder Frequency Plan (Centered Video)

Channel spacings for typical SCPC-type carriers are as follows:

SCPC Carrier Type	Channel Spacing	Channel Capacity (Bandwidth Limit)
30 kHz FM-SCPC	30 kHz	100
45 kHz FM-SCPC	45 kHz	66
56 kb/s QPSK	50 kHz	60
56 kb/s BPSK	100 kHz	30
1.544 Mb/s QPSK	1.5 MHz	2

The bandwidth-limited channel capacity is also shown. The channel capacity may be limited to a lesser number of channels by interference, G/T, EIRP, Video C/N or other constraints. These limitations are discussed in more detail in the following sections.

Figure 2 details a frequency plan with the video offset by 3 MHz. This frequency plan provides more available bandwidth and reduced interference. This plan has possible applications to new video services in which the video carrier can initially be offset from the transponder center frequency.

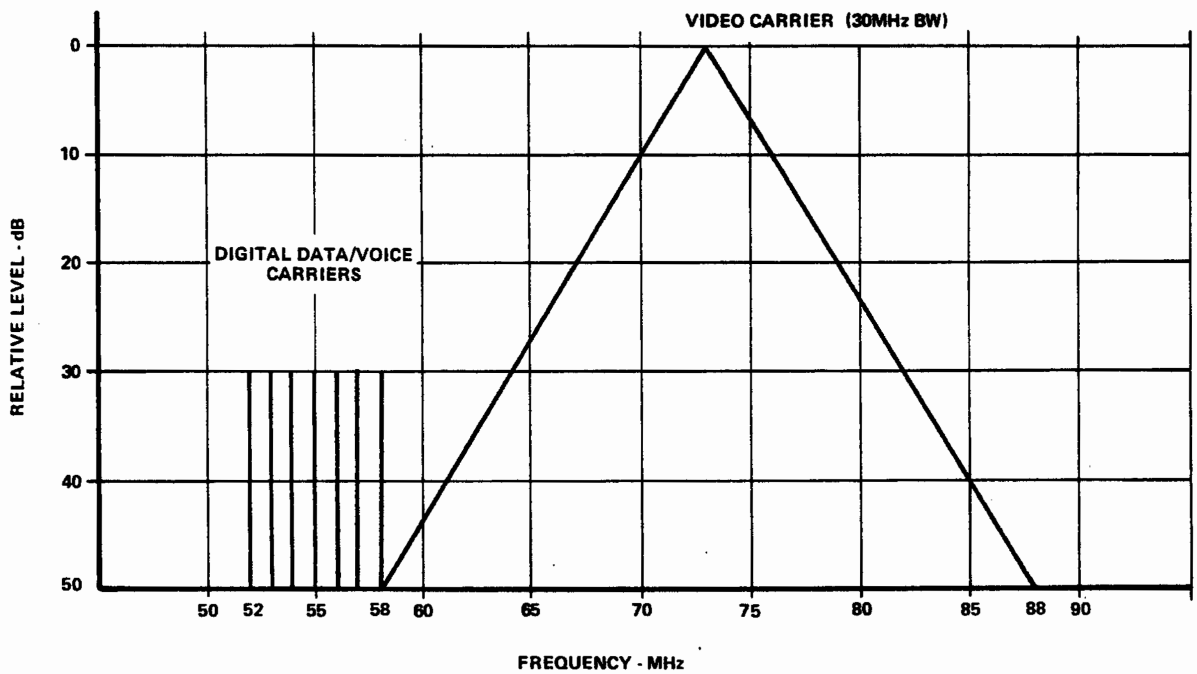


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Figure 2. Transponder Frequency Plan (Offset Video)

Transponder Operating Point/Power Sharing

The transponder may be operated at saturation or at some other output backoff. It is usually preferable to operate the transponder with a saturated video carrier to provide the highest possible C/N for low G/T video receive-only earth stations. Fortunately, the required-per-carrier SCPC carrier power level is up to 30 dB below the video carrier level, and many SCPC carriers can be added before significant power is "used" by the SCPC carriers, and before interference to the video occurs.

Operating Constraints

In-Band Intermodulation Products

Intermodulation (IM) products are generated when multiple carriers are passed through a non-linear device, i.e., a transponder operating at or near saturation. The level and frequency of the IM products are a function of the level and frequency of the multiple carriers. Figure 3 shows IM products that are generated by a CW saturated carrier and seven SCPC carriers each at a level 20 dB below the CW carrier. Note the following:

a. Third-order two-carrier IM products above the CW carrier are about 5 dB below the SCPC carriers.

b. Third-order three-carrier IM products around the CW carrier are about 30 dB below the CW carrier.

c. Third-order two-carrier products around the SCPC carriers are more than 30 dB below the SCPC carriers.

d. Other third- and fifth-order products are outside the 40 MHz transponder frequency allocation.

The high-level third-order IM products of a. above prevent the transmission of carriers on both sides of the saturated carrier. Modulation of the saturated carrier will lower the level and widen the frequency band of the IM products.

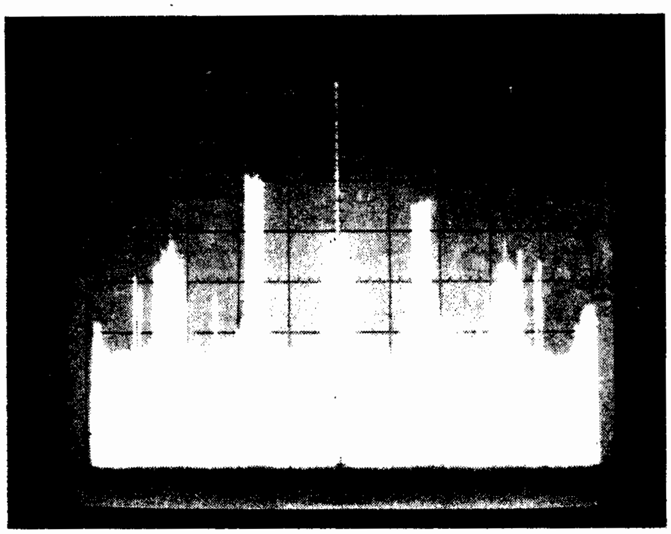

Figure 3. In-Band IM Products

Video-to-SCPC Interference

Video-to-SCPC interference is a function of:

- Transmitted video spectrum bandwidth
- Relative SCPC carrier levels

- Transponder output backoff
- Video carrier frequency stability

Figures 4 and 5 illustrate a color bar spectrum and seven SCPC carriers. The transmit filter bandwidth is 36 MHz, and peak deviation is 10.75 MHz. Figures 6 and 7 illustrate color bar spectrums with 30 MHz and 25 MHz bandwidths, respectively. Note that the color bar spectrum (in a 30 kHz bandwidth) is at least 12 dB below the SCPC carrier level. Figure 8 illustrates the color bar spectrum with a 36 MHz bandwidth filter and with the video carrier shifted in frequency by 1 MHz.

SCPC interference tests have shown that video interference in a 30 kHz bandwidth starts to interfere with the SCPC carriers when it is 5 dB and 0 dB, respectively, below 56 kb/s QPSK digital data (coded) and 30 kHz FM-SCPC carrier levels.

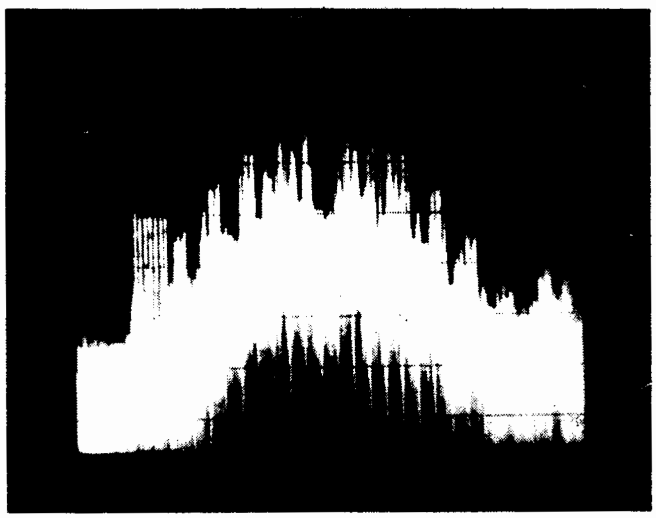

Figure 4. Color Bar/SCPC Spectrum

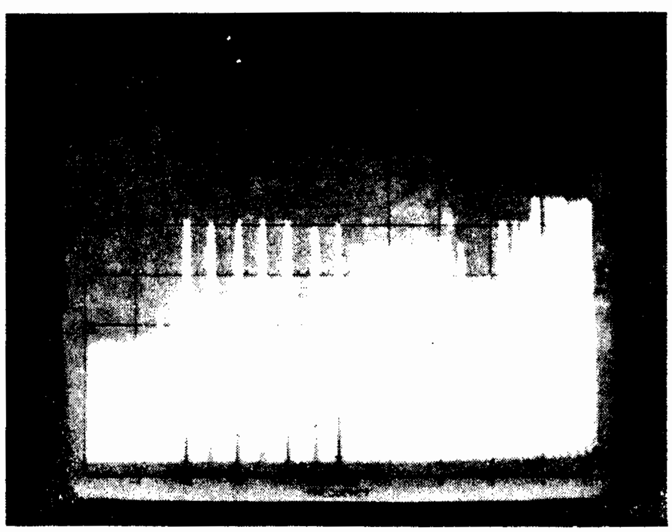

Figure 5. Color BAR/SCPC Spectrum

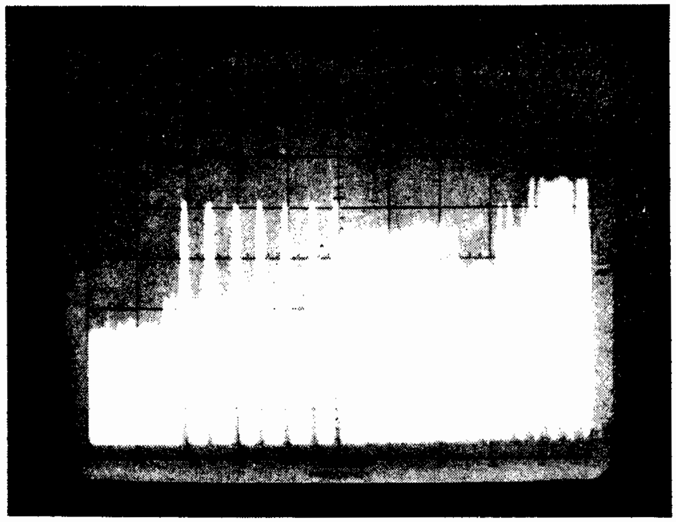
Figure 6. Color Bar/SCPC Spectrum

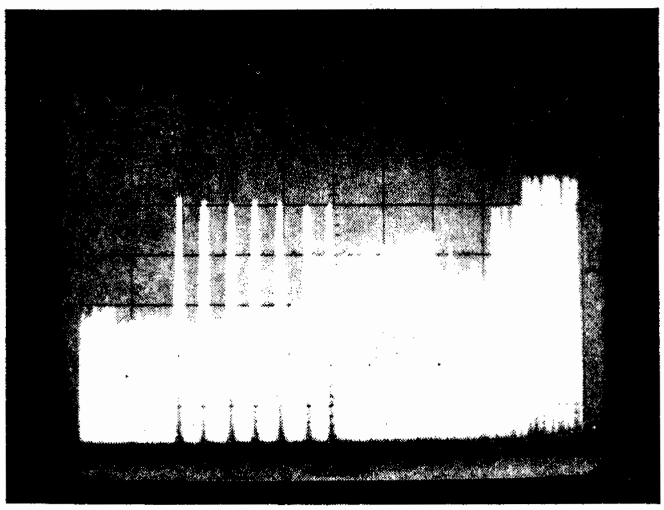
Figure 7. Color Bar/SCPC Spectrum

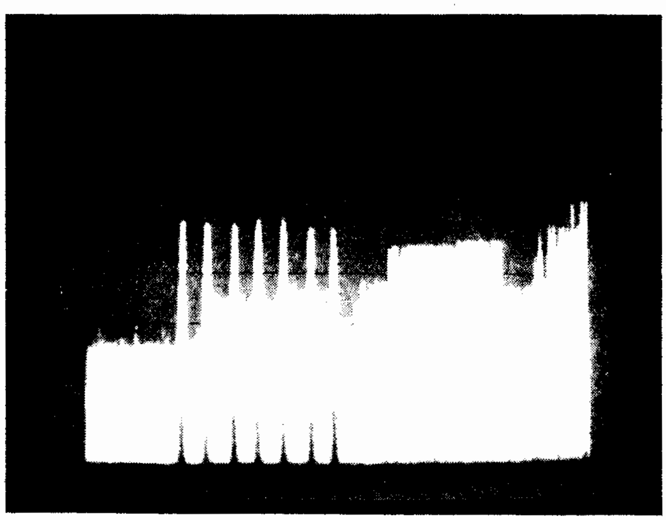
Figure 8. Color Bar/SCPC Spectrum

Other Potential Interference Sources

Although not unique to the video plus transmission plan, other potential interference sources to the low level SCPC carriers must be considered. These include:

- Adjacent copolarized transponder interference
- Adjacent cross-polarized transponder interference
- Adjacent satellite interference
- Terrestrial microwave interference

The center frequency of a C-band cross-polarized transponder is offset by 20 MHz from the adjacent transponder. Thus, transmitted carriers in this transponder will overlap the frequency band occupied by the SCPC carriers. The worst-case situation is probably when the cross-polarization transponder is occupied by a saturated video carrier. Tests show that to prevent interference to SCPC carriers, the SCPC carrier level must be from 3 to 5 dB above the cross-polarization isolation as seen at the receiving earth station.

Terrestrial microwave carriers should not interfere with the SCPC carriers since they are offset from the SCPC occupied bandwidth. This assumes, of course, that the terrestrial microwave has been sufficiently frequency coordinated to prevent interference to the received video.

In summary, before operating in the video-plus mode, an interference analysis must be made, just as an interference analysis must be made for any full or partial transponder operating mode.

SCPC-to-Video Interference

SCPC-to-video interference is a function of:

- Transponder output backoff
- Receiver IF bandwidth
- Video C/N ratio
- Relative total SCPC carrier power

Video interference tests show that color bars are more susceptible to SCPC interference than regular program material (as might be suspected).

Also, a video site with a low C/N ratio is more susceptible since the video receiver is operating closer to threshold.

SCPC carriers begin to interface with video color bars when the total SCPC carrier power is about 15 dB below a saturated video carrier when measured, using a 30 MHz bandwidth video receiver operating at a 15 dB C/N ratio. Following are other measured interference levels (30 MHz bandwidth, 15 dB C/N ratio):

Subjective Interference	Relative Total Interference Levels
Color Bars	-15 dB
Live Program	- 8 dB

Video/Audio Performance

Video and audio performance meets NTC-7 recommendations when using either 36, 30 or 25 MHz bandwidth transmit and receive filters. The differential phase and gain degrade only slightly when using 25 MHz bandwidth filters. Tests on the above were conducted with a 10.75 MHz peak video deviation and one 6.8 MHz audio subcarrier. SCPC interference does not affect linear or non-linear waveform distortion.

From the above interference and video/audio performance considerations, it appears that most system configurations should use 30 MHz bandwidth filters in both the transmitter and receiver.

Carrier Suppression/Carrier Level Variation

When different level carriers (in this case 10 dB or more) are amplified by a non-linear device (transponder TWT), the larger carrier will suppress the level of the low-level carrier. The small carrier suppression factor at saturation for typical TWT's is 4 dB. This, of course, means that the uplink EIRP of the low-level SCPC carriers must be 4 dB higher than that necessary for linear transponder operation (6 dB output backoff).

Figure 9 illustrates typical TWT (transponder) transfer characteristics. Input and output backoffs are relative to the saturated power output. For a typical satellite transponder, the saturated power output (i.e., at 0 dB output backoff) is +33 dBW. Small carrier suppression and small carrier level change are shown versus backoff. Note that for a ±2 dB input backoff change (i.e., the saturating carrier has changes ±2 dB), the small carrier level changes -3/+2 dB, respectively. This means that the video-plus system must be designed to accommodate uplink level variations of the high-level saturating carrier. These variations are caused by uplink transmitter power changes, transmit antenna gain changes due to wind conditions or snow accumulations, and uplink path loss changes mainly due to rain storms.

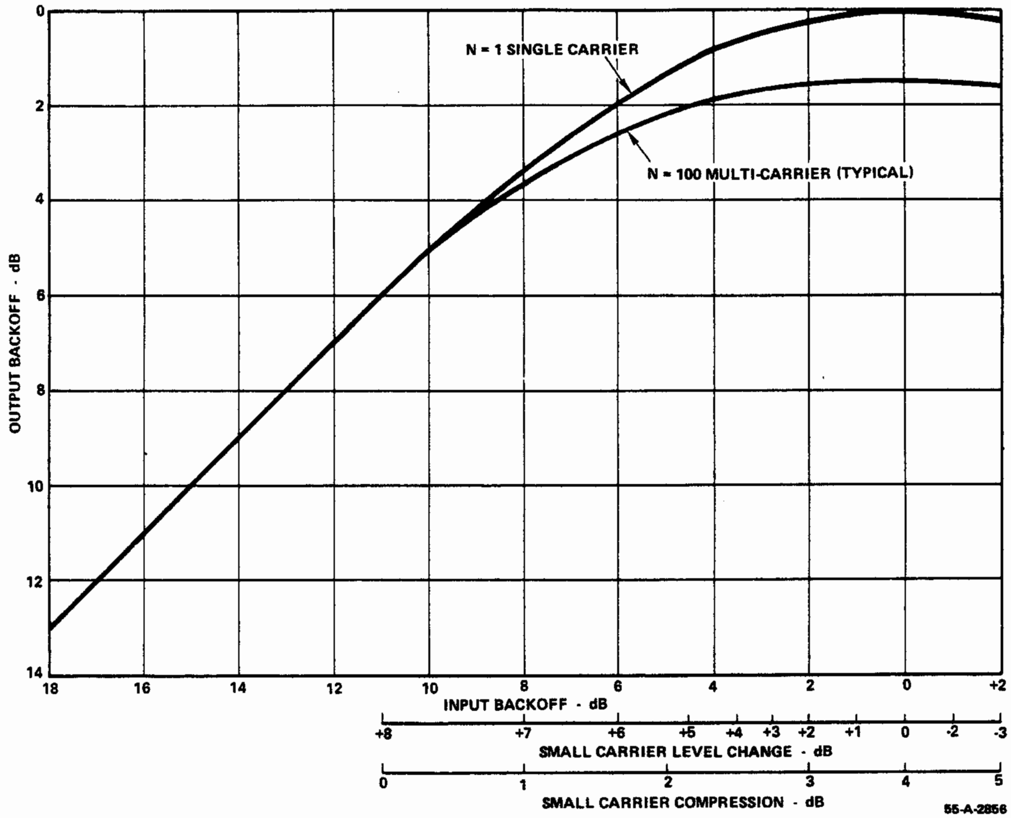

Figure 9. Typical TWT Transfer Characteristics

Additionally, as the total small carrier power approaches the level of the large carrier, suppression of the large carrier will occur. This suppression is due to a combination of non-linear and power-sharing effects. A 1 dB large carrier suppression occurs when the total power of the small carriers is about 7 dB below the large carrier power.

Channel Capacity/Earth Station G/T System Performance

The SCPC channel capacity is a function of various factors including satellite EIRP, earth station G/T, link margin, link performance and interference criteria. Table 1 details typical system performance for video, 56 kb/s QPSK digital data, and 30 kHz FM-SCPC links. Note that the number of SCPC channels varies with the earth station G/T. The system performance calculations assume negligible "outside" interference, color bar interference criteria, and 2 dB video-plus interference margins.

Table 1. Typical System Performance

	Characteristic	Performance	
1.0	Satellite Characteristics		
	Type	Typical Domestic	
	EIRP (saturated)	+33.0	dBW
2.0	Earth Station Characteristics		
	Location	CONUS	
	Elevation Angle	20	deg.
3.0	Equipment Configuration		
	TX IF Filter Bandwidth	30	MHz
	RX IF Filter Bandwidth	30	MHz
	Deviations (pk)		
	Video	10.75	MHz
	Subcarrier	2.0	MHz
	Audio	200	kHz
	Energy Dispersal	1.0	MHz
4.0	Video-Plus Parameters		
	Video Operating Point	Saturated (0 dB output backoff)	
	Frequency Plan		
	Video Carrier	Centered	
	SCPC Carriers	Lower 3 MHz	
	In-Band Interference-Free Criteria		
	Video Waveform	Color Bars	
	Relative Interference Level		
	To Video (For C/N)		
	11 dB	-17	dB
	14 dB	-15	dB
	20 dB	-14	dB
	To 56 kb/s QPSK Digital Data	-30	dB
	To SCPC Voice	-32	dB
	To SCPC Pilot	-32	dB
	Threshold Criteria		
	Digital Data	BER = 10^{-7} (Min)	
	FM-SCPC Voice	S/N = 50 (Min)	dB
	Link Margin	2.0 (Min)	dB
	Channel Spacing/Bandwidth Channel Limit		
	Video	30	MHz
	56 kb/s QPSK Digital Data	50 kHz/60 channels	
	FM-SCPC Voice	30 kHz/100 channels (including pilot)	

Table 1. Typical System Performance (continued)

Characteristic	Performance			
5.0 System Performance				
Earth Station G/T	22	26	30	dB/K
Video Performance				
C/N	11.5	15.5	19.5	dB
S/N	49	53	57	dB
56 kb/s QPSK Digital Performance				
Relative Operating Carrier Level	-27	-28	-28	dB
Number Channels (one way)	10	20	25	
Threshold BER		10^{-7}		
30 kHz FM-SCPC Voice Performance				
Relative Operating Carrier Level	-30	-30	-30	dB
Number Channels (one-way)	20	32	40	
Threshold S/N		50		dB

Applications

As shown in the previous section, significant additional space segment can be made available with the Video-Plus technique. On a 24 transponder satellite used exclusively for video, at least 240 SCPC carriers per satellite (10 SCPC carriers times 24 transponders) can be transmitted. This excess capacity gives rise to several interesting applications.

Hotel/Motel Networks

An ideal application for Video-Plus is a Hotel/Motel voice and data network. As a greater number of hotels and motels install earth stations for reception of entertainment, the feasibility of networking becomes more realistic. By adding uplink capability to a receive-only earth station, a network node is created. By utilizing the Video-Plus technique, the earth station could be used for reception of entertainment programming as well as two-way voice and data traffic.

Nationwide Control of Addressable CATV Set-Top Converters

The CATV Multiple System Operator judges the economics of addressable converters on the basis of offsetting the cost of an addressable system with savings in operating expenses and additional revenues form Pay Per View. A significant part of the cost of addressability, however, is associated with the part of the system that controls addressability; this cost is only minimally affected by the size of the system controlled. Not only the cost of the hardware and software driving the system, but also the additional staff of operators and technicians must be offset by the benefits. It would therefore be beneficial to the MSO to control some number of systems from a single control point, thereby spreading the costs across a larger revenue base. The Scientific-Atlanta Series 8500 Set-Top Terminal was system designed to accommodate this need while utilizing Video-Plus as the data link.

In a Video Plus-System, the Multiple System Operator's addressability control system can be located anywhere in the United States, where it uplinks to a transponder which is viewed by all of its cable systems. From this Master Control Center, control data is transmitted via the digital carriers to all cable systems, where it can then be distributed to the converters to be controlled. The equipment at the cable system headend necessary to accomplish the data transfer to the cable system is dramatically less expensive than a typical headend addressable control system, and it can easily be maintained by the existing CATV technical staff.

Other Applications

Some of the other applications that exist for Video Plus are:

- One-way video teleconferencing/two-way voice and data for business or education
- Interactive CATV systems with one-way video and two-way data and voice
- Small private telecommunication networks
- Digital/SCPC Program Audio Distribution

Conclusion

Video-Plus provides an exciting new source for additional space segment which has been previously unused. New opportunities now exist for networks which may not have been economically feasible on a dedicated transponder basis in the past.

Low-Noise Amplifiers

Ken Johnson

Introduction

The Low-Noise Amplifier (LNA) is an important earth station component because its noise figure and gain along with the antenna gain virtually set the system G/T. In early earth stations, parametric amplifiers were used to provide low noise temperatures at high costs, or bipolar amplifiers and, in some cases, mixer/IF amplifiers were used where lower performance could be tolerated. Today, with the development of the low noise Gallium Arsenide Field Effect Transistor (GaAs FET), most LNAs use these devices in the first two gain stages. Noise temperatures of 4-GHz GaAs FET LNAs range from 75K to 120K for uncooled amplifiers to as low as 50K for a thermoelectrically cooled amplifier. Such low noise temperatures rival those of parametric amplifiers, but at a fraction of the cost.

LNA Functions

Besides its primary purpose of providing a low-noise temperature, we can identify four primary functions of the LNA as it is used in Scientific-Atlanta Earth Stations:

- Provide high gain along with low noise to establish high system G/T. The LNA is generally mounted as close to the antenna feed as possible so that transmission line losses to the LNA will be at an absolute minimum. Sufficient gain must be provided by the LNA to overcome losses in the transmission line from the LNA to the receiver and to override noise which originates after the LNA.

- Provide transition from antenna waveguide to TEM coaxial line. Since long waveguide runs are expensive, the LNA is designed to accept a waveguide input and provide a coaxial line output.

- Provide adequate mechanical strength to permit bolting directly to the antenna waveguide and to allow connection of a long coaxial cable to the unit.

- Provide RFI/EMI tight weatherproof housing for circuitry. The LNA is usually exposed to the elements, and consequently the circuitry must be enclosed in a weathertight enclosure. Also, in any particular location, the LNA may be exposed to electromagnetic interference (EMI) or radio frequency interference (RFI) and must, therefore, be designed to completely shield the delicate transistors from any such interference.

Scientific-Atlanta LNAs are designed to perform all the required functions, as well as meeting other critical specifications required from the LNA. Without going into detail at this point, the critical specifications will be listed and described as follows:

Typical Specifications

Frequency Range: 3.7 - 4.2 GHz
Noise Temperature: 80K - 120K. The exact noise temperature is set by the required system G/T.
Gain: 50-dB minimum.* Gain must be sufficient to overcome any cable losses to the receiver.
Gain Flatness: ±0.5 dB/500 MHz/±0.25 dB/40 MHz. Gain flatness is required to prevent cross modulation between channels and assure low signal distortion.
Power Out 1-dB Gain Compression: 0 dBm minimum. A sufficiently large power output at gain compression assures low distortion at large input signal levels.
Input VSWR: 1.25 maximum. A low input VSWR is inherent in achievement of a low noise in an LNA which uses an isoadaptor.
Output VSWR: 1.5 maximum. A low input VSWR assures that there will not be resonances and instabilities or signal level variation caused by reflections in the long cable run to the receiver.
Temperature: -30°C to +50°C typical. LNAs are exposed to a variety of outdoor temperatures and must function well at all of these temperatures.

Other specifications such as group delay, AM-PM conversion, etc., are often specified, but it has been found that if the LNA meets the critical specifications, it will also meet the required AM-PM conversion and group delay.

All of the specifications were considered in the design of the LNA. In this paper we will trace through the design techniques for meeting the specifications and then show how the Scientific-Atlanta LNA meets these requirements. Finally, we will look briefly at future LNA improvements through cooling and device improvement. First, we will show what the LNA looks like, consider the block diagram of the LNA and its bias circuitry, and then describe the GaAs FET used in the first two stages.

LNA Description

Figure 1 is a photograph of the Scientific-Atlanta Series 300 LNA, showing the waveguide input to the LNA which bolts to the antenna feed structure. On top of the waveguide section is the isolator which isolates the transistor amplifier from any antenna mismatch reflections and provides a transition from the waveguide transmission mode to the coaxial mode of operation required for feeding the signal to the amplifier. The isolator, or isodaptor, is followed by the housing for the transistor amplifier and its bias circuitry. The amplifier output is taken from a coaxial connector, which in this particular LNA also serves to provide the dc power input to the LNA.

* Gain of 60 dB can be provided for special applications.

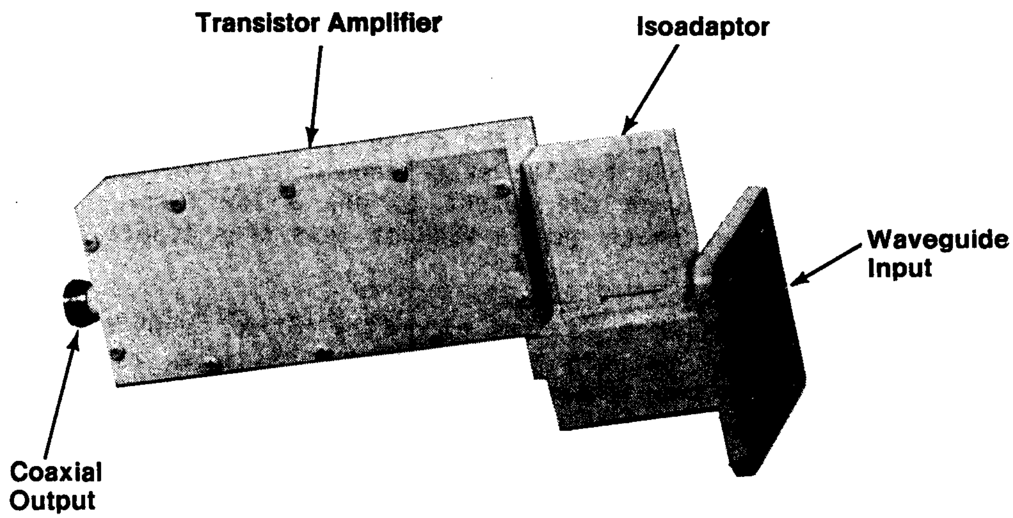

Figure 1. Photograph of Scientific-Atlanta Series 300 LNA

LNA BLOCK DIAGRAM

To show what are the component parts of a typical LNA, consider the block diagram of Figure 2.

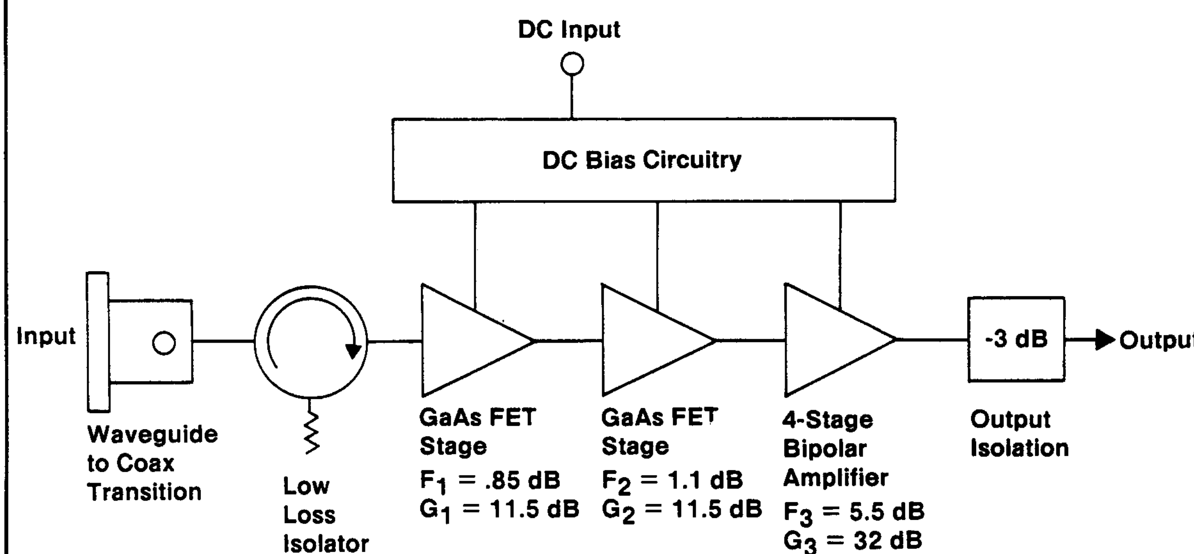

Figure 2. LNA Block Diagram

The input to the LNA has the isoadaptor which consists of a waveguide to coax transition in conjunction with a very low-loss isolator. The isoadaptor typically has less than 0.15 dB loss. Since this loss adds directly to the noise figure, it must be kept as low as possible. Following the isoadaptor are six stages of transistor amplification. The first two stages must use FET devices in order to meet the required noise figure. The next four stages are bipolar, although possible two FET stages could have been used. Finally, an isolation block is used on the output to get a low VSWR. In this LNA design a 3 dB pad was used which gave a VSWR of less than 1.35:1 over the 3.7 to 4.2 GHz band.

The block diagram also shows a dc bias circuitry block. For bipolar bias a simple resistive feedback network may be used. FET bias is more complex since positive and negative voltage are required to bias the FET if the source terminals are to be grounded. A voltage regulator is incorporated in the dc bias circuit to permit a wide range of dc voltages to be supplied to the LNA without affecting its performance.

The regulated voltage is also used to power an oscillator which is rectified and filtered to produce a negative voltage needed for the GaAs FET bias. An active bias-transistor network is used for the GaAs FET stages to assure good temperature stability.

Scientific-Atlanta also has available LNA's which accept ac input. In this case, a separate connector is provided to accept the ac power which is transformed to a lower voltage level and rectified. No degradation in amplifier performance results whether ac or dc is supplied.

GaAs FET

The GaAs FET is so important that it is worthwhile spending some time describing the device and how it is characterized for amplifier use.

Device Description

The construction of the GaAs FET used in the LNA is surprisingly simple. Unlike microwave bipolar transistors, it requires no special diffusions to achieve n or p type layers of semiconductor. Rather, as shown in Figure 3, it consists of a layer of semi-insulating GaAs (Gallium Arsenide) on which a lightly doped n+ layer of GaAs, called an epitaxial layer, is grown. Then, in a vacuum chamber under proper temperature conditions, a layer of metal film, usually gold, is evaporated on the epitaxial layer to form a Shottky-barrier junction. Source, gate, and drain contacts are etched by photolithographic techniques. Finally, a passivation layer is added.

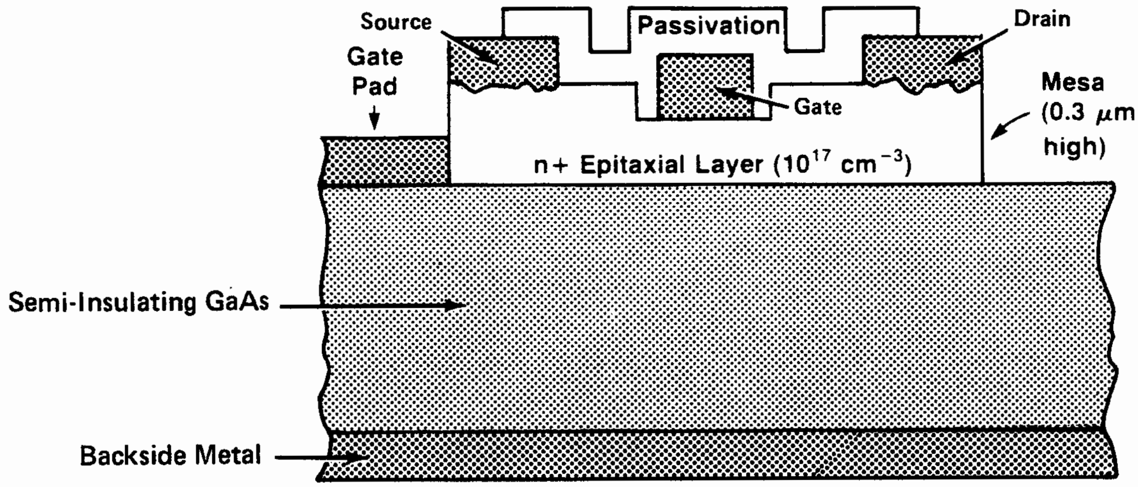

Figure 3. Cross section of FET geometry.

To achieve the lowest noise FETs, the epitaxial layer under the gate must be etched into a recess, as shown in Figure 3. For a microwave FET, the gate "finger" is extremely narrow, as narrow as 0.5 micron. This 0.5 micron width is equal to a wavelength of blue light. A typical pattern for a FET is shown in Figure 4, which is the pattern for a 1.0 micron-wide gate stripe.

Electrically, the FET resembles a vacuum triode with the gate corresponding to the grid, the source corresponding to the cathode and the drain corresponding to the anode. Figure 5 shows an equivalent circuit for the GaAs FET. Since it corresponds to the triode, it has a relatively high input impedance. It is this fact which makes the FET so useful at microwave frequencies. Bipolar transistors are difficult to use as low-noise amplifiers above 4 GHz because their matching impedance is so low.

GaAs FETs are currently capable of noise figures as low as 0.6 dB at 4.0 GHz with corresponding gains of greater than 14 dB, although in production a spread of somewhat poorer values of gain and noise figure must be accepted.

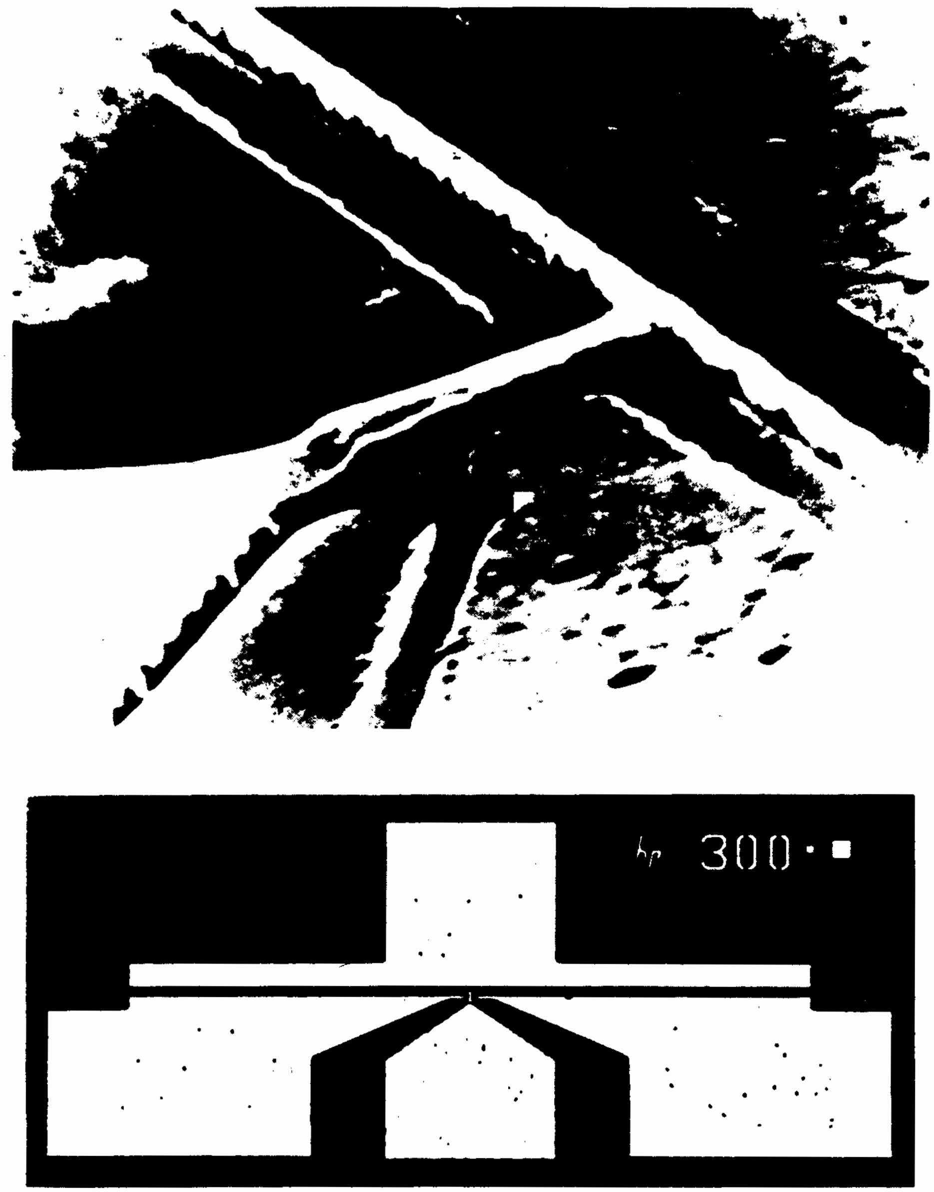

Figure 4. HP TC-300 Gallium Field-Effect Transistor Chip. Scanning Electron Microscope Photo Shows 1- m-wide Gate Stripe

GaAs FET Characterization

At microwave frequencies, use of the equivalent circuit shown in Figure 5 is extremely difficult because the package in which the device is placed has a great effect on its electrical characteristics. The tiny wire used between the gate bonding pad and the gate lead on the package has a significant inductance at 4 GHz. Consequently, such devices as the FET are best characterized by S-parameters.

S-parameters are two port parameters like h-parameters, except that instead of being in terms of voltages and currents, they represent the transistor in terms of transmission and reflection coefficients. To put it another way, they represent normalized input impedance (S_{11}), normalized output impedance (S_{22}), forward gain (S_{21}), and reverse gain (S_{12}) when the transistor is placed in a 50-ohm line. These parameters are used in computer programs in designing an amplifier.

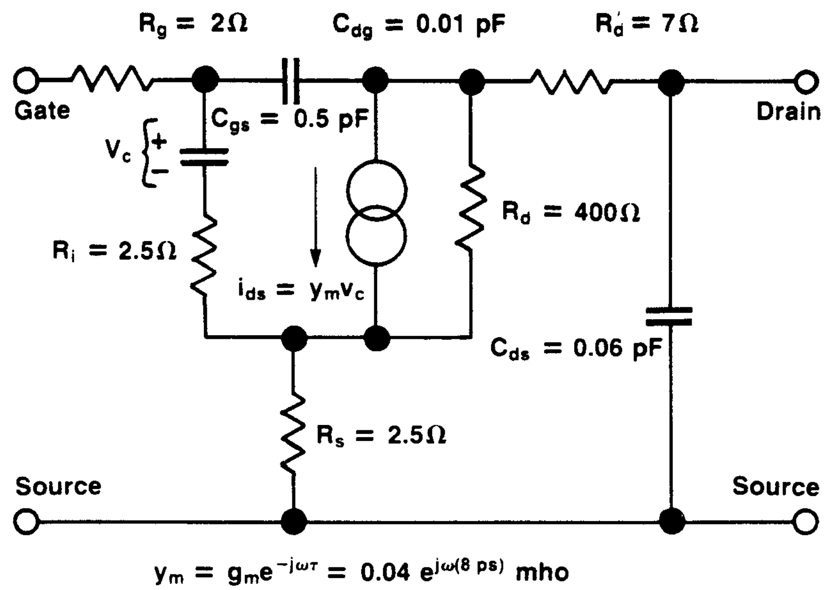

Figure 5. GaAs MESFET Equivalent Circuit

Amplifier Design

In approaching the design of an amplifier, the first step is to prepare a block diagram showing each transistor with its contribution to gain and noise figure of the overall amplifier. An example has already been given in Figure 2. It is then necessary to compute the overall gain and noise figure considering each stage to ensure that design specifications are met. Let us consider, for example, the design of a 100K LNA.

Transistor Stages

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots \frac{F_N - 1}{\prod_{i=1}^{N-1} G_i}$$

$$F = 1.216 + .0198 + .0091 + \cdots .00375$$

$$= 1.2488 \ (.9649 \text{ dB})$$

$$F = 64.8\text{K} + 5.94\text{K} + 2.72\text{K} + 1.126\text{K}$$

$$= 74.6\text{K}$$

Transistor Stage Values

N.F.(dB)	Gain (dB)
F_1 = .70	G_1 = 11.5
F_2 = .90	G_2 = 11.5
F_3 = 4.5	G_3 = 7.0
F_4 = 6.0	G_4 = 8.5

Etc.

Summary of Noise Figure Contributors

Transistor Stages	= .965 dB
Isoadaptor Loss	= .150 dB
Coupling Loss	= .050 dB
4.2 GHz Degradation	= .100 dB
	1.265 dB ($\approx$100K)

Figure 6. Noise Figure Calculations for a 100K LNA

Minimum Noise Figure Design

Using measured data on the FET and bipolar transistors, one can compute the noise figure and gain of each stage and their contribution to the entire amplifier as shown in Figure 6. The overall amplifier noise figure is given by the top equation in the figure and values of each stage noise figure and gain are tabulated to the right in the figure.

In terms of contribution to overall noise temperature, the first stage contributes 64.8K, the secondary only 6K. The reason, of course, is that the first stage gain reduces the effect of the second stage noise figure. Notice, though, a good low-noise FET must be used in the second stage, for if a state-of-the-art bipolar transistor were used, the second stage contribution could be as much as 20K instead of 6K. Care must be taken to ensure that the stages are properly matched for minimum noise figure as well.

Toward the bottom of Figure 6 all the contributions to the amplifier noise figure are summarized. Transistor stages contribute 0.965 dB (74.6K), and isoadaptor losses contribute 0.15 dB. Coupling losses to the transistor include bypass capacitor and line losses; these contribute 0.05 dB. Finally, the transistors are specified for noise figure at 4.0 GHz, but the amplifier must operate at 4.2 GHz. This contributes another 0.10 dB to the noise figure, giving a total of 1.265 dB - a value corresponding to a noise temperature of slightly less than 100K.

Once the number of stages has been chosen and transistors selected, the next step is the actual circuit modeling and design.

Circuit Modeling and Computer Design

Scientific-Atlanta LNAs are all developed using computer-aided-design and printed circuit techniques. While it is possible to design a single stage

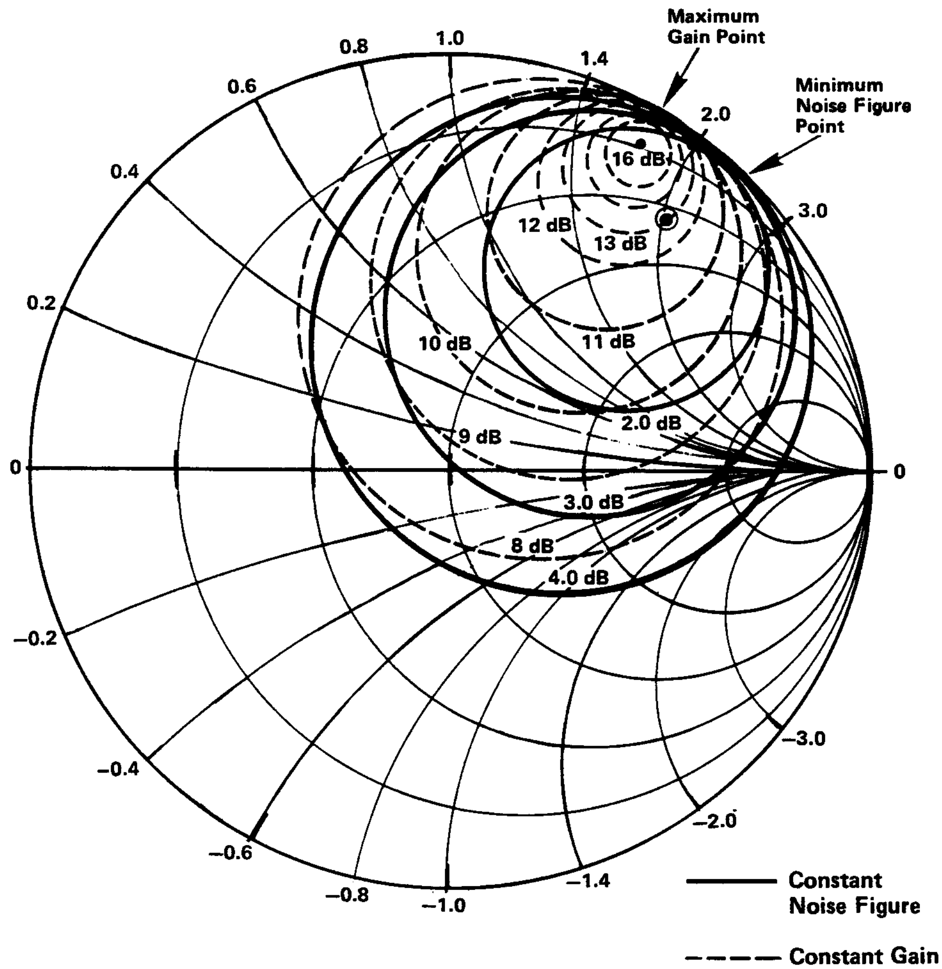

Figure 7. Smith Chart Representation of Constant Gain and Noise Figure Circles for GaAs FET

amplifier without the aid of a computer, the task becomes monumental when several stages in cascade must be optimized over a band of frequencies. This is especially so when minimum noise figure is sought for the second and third stages as well as the first stage.

In any case, one invariably starts with one stage, optimizes it for noise figure and gain, adds the second, optimizes it, adds the third and so on. First, a circuit must be selected for matching into and out of the transistor, and values computed that are somewhat near the desired values. Use of the Smith Chart allows the designer to plot the family of the impedance values presented to the transistor, thus giving a constant gain or constant noise figure. These appear on the Smith Chart as circles of constant gain or constant noise figure as shown in Figure 7. Notice that there is only one point which gives minimum noise figure and another point giving maximum gain. Thus, for the transistor of Figure 7, the gain available at minimum noise figure is 13 dB. Since we are primarily concerned with minimum noise figure, circuit elements must be chosen to match this impedance (namely 30 + j100). The elements chosen consist of lengths of line which form open circuited stubs, short circuited stubs, etc.

In a somewhat similar fashion, matching elements are selected for the output of the transistor. These choices of matching elements are then used as starting points in a computer program. The computer program used at Scientific-Atlanta optimizes for gain and noise figure over the design band in an iterative process. The program takes the initial design values, computes gain and noise figure at a number of frequencies, and compares these values with the desired values of gain and noise figure. An error function is then generated. By varying selected parameters such as capacitance, line length, stub impedance, a new error function is computed. If the new error is less, the circuit values are readjusted to the new values and the process is repeated. After the iteration error comes within acceptable limits, the computer prints out the final circuit element values. The second stage is then added and the process repeated, adding the third, fourth, fifth, and sixth stages.

Design for High Gain Compression Level

Another factor entering into amplifier design is amplifier linearity, which is uaually specified in terms of the output power level at which the gain compresses 1 dB. A large value of gain compression, as we said before, is also important in order to have low cross-modulation products and low signal distortion in cases where the returned signal from the satellite might be somewhat large. This is a factor which really only affects the last stage or perhaps last two gain stages, which have been preceded by some 45 dB of gain. To achieve a high compression level, three things are required. First, a transistor must be selected which has a high compression level. Such transistors usually have somewhat lower gain and often cost more than the bipolar devices used in the preceding stages. Second, the biasing of the transistor is modified to provide the necessary higher current and voltage for a higher compression level. Finally, the transistor output must be well matched so that none of the signal output is reflected back into the transistor.

It must be remembered that it is not always desirable to design for the highest compression level since some sacrifice in gain, output VSWR and increased dc power results. The standard Scientific-Atlanta LNA is specified to have a minimum 0 dBm output at 1-dB gain compression, although most units have +5 dBm. LNAs with higher gain compression levels are available on request.

Amplifier Biasing and Stabilizing

One of the well-known problems encountered in amplifier design at microwave frequencies is the suppression of oscillation. Achievement of an unconditionally stable amplifier requires careful attention to the mounting of transistors, use of resistors to isolate bias points and use of absorbing material in the amplifier housing. It also requires choosing the correct matching elements in the initial design. Some matching elements tend to have parasitic resonances which cause oscillation at undesired frequencies.

Scientific-Atlanta LNAs are designed using these techniques and consequently are stable under all temperature conditions.

LNA Performance

Having discussed the LNA design in some detail, we will now show some performance characteristics of Scientific-Atlanta LNAs.

Noise Figure Performance

Of primary interest is the noise figure behavior of the LNA. Figure 8 is a typical plot of the noise figure and noise temperature of two LNAs supplied by Scientific-Atlanta as a function of frequency. One is a 100K unit and the other an 80K LNA. Charcteristic of the 100K LNA is that the noise temperature at the low frequency end of the band is the lowest and increases slightly toward the high end of the band. This is the type of behavior that would be expected from an LNA designed for minimum noise figure across the band. The noise figure would (as it does) follow the shape of the noise figure curve for the device alone. From 3.7 to 4.2 GHz the device alone varies some 0.2 dB, as does the amplifier. There is a slight dip or deviation from a straight line curve in the middle of the band where the slight reactive mismatches are a minimum.

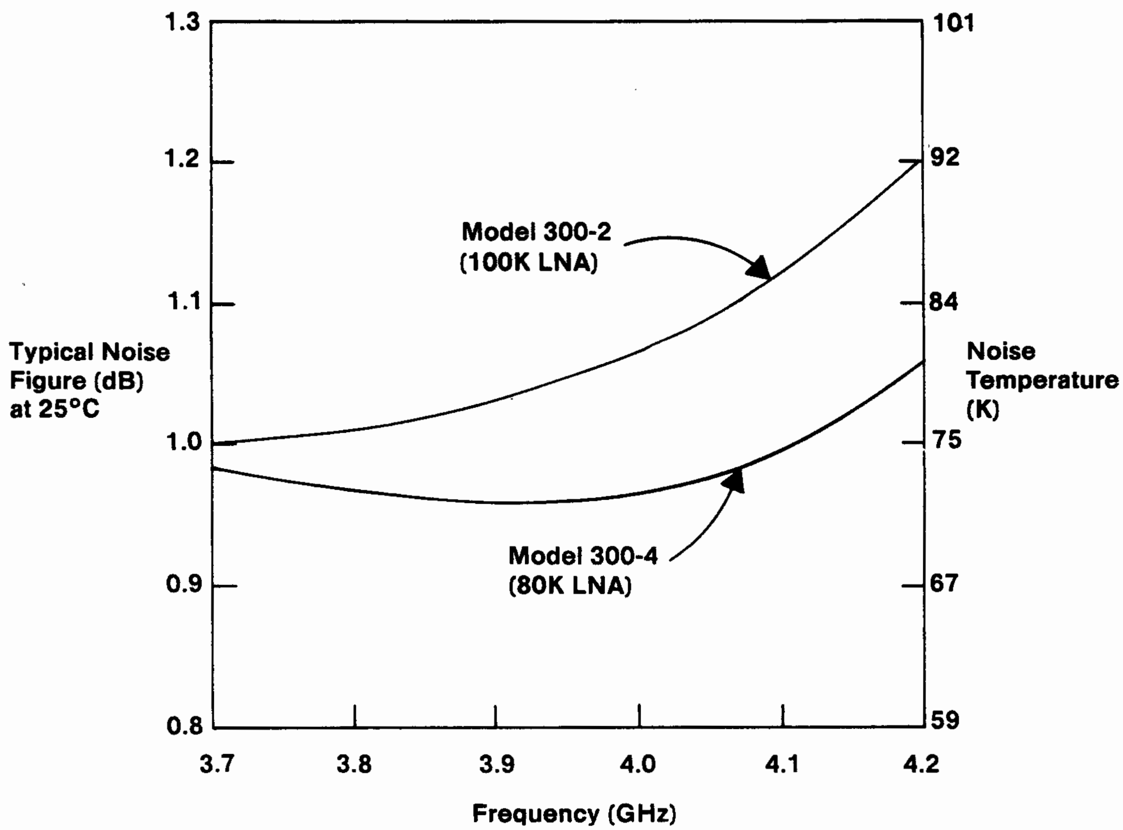

Figure 8. Noise Figure vs. Frequency for Two LNA Models

For the 80K LNA there is a slight difference, primarily because greater tuning time must be expended to make the high frequency noise figure as low as possible. Achievement of the 80K noise temperature requires special selection of FETs and isoadaptors as well as longer tuning time.

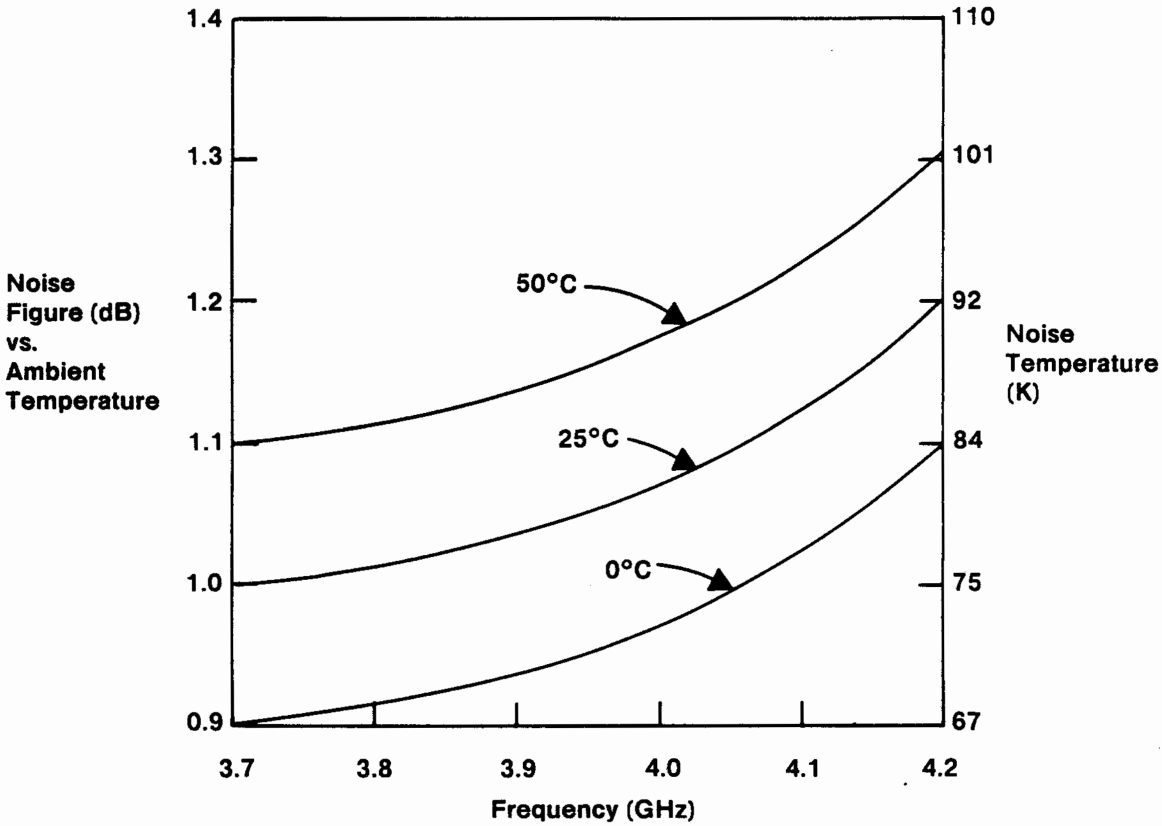

Figure 9. LNA Noise Figure vs. Frequency for Three Ambient Temperatures

Figure 9 shows a 100K LNA noise figure and noise temperature versus frequency at three ambient temperatures. Notice that at 50°C a 10K increase in amplifier noise temperature results. Similarly, at 0°C, an 8K decrease in amplifier noise figure is measured. These results are very close to what is predicted analytically for these amplifiers. It should be kept in mind that LNA noise temperatures are always specified only for room temperature operation. Thus, when computing system G/T, care must be taken to include any noise temperature degradation that occurs if the LNA is expected to operate at temperatures above 25°C.

Gain Characteristics

Although the LNA is followed by a receiver with an automatic gain control circuit, it is important that the LNA gain not vary excessively with temperature. A typical gain-versus-frequency characteristic for Scientific-Atlanta LNAs is given in Figure 10 for three different temperatures. Notice that the gain is constant within approximately ±0.25 dB at any one temperature and that the gain increases approximately 3 dB from 0°C to 50°C. This gain variation is well within receiver AGC operating ranges. Notice that no significant change in the shape of the gain curve takes place. This also is an important feature of Scientific-Atlanta LNAs, since it means that there will be no changes in gain flatness which could result in intermodulation distortion.

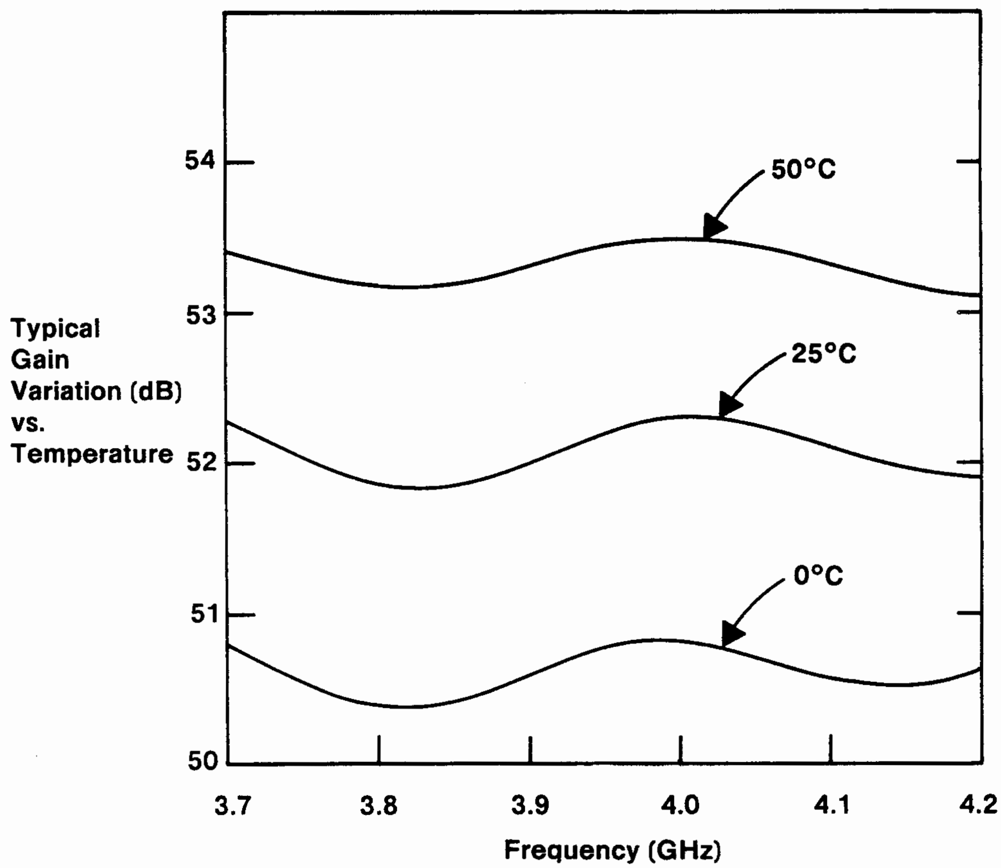

Figure 10. LNA Gain Variation vs. Frequency for Three Temperatures

LNA Data Sheet

To summarize the performance of the Scientific-Atlanta LNA, we have included a typical data sheet in the Equipment section under Electronic Products. All of the critical specifications are met by this unit, and other special requirements can be provided on request. Figure 11 is a photograph of a finished production unit. Scientific-Atlanta has produced hundreds of such units to date with noise temperatures less than 100K.

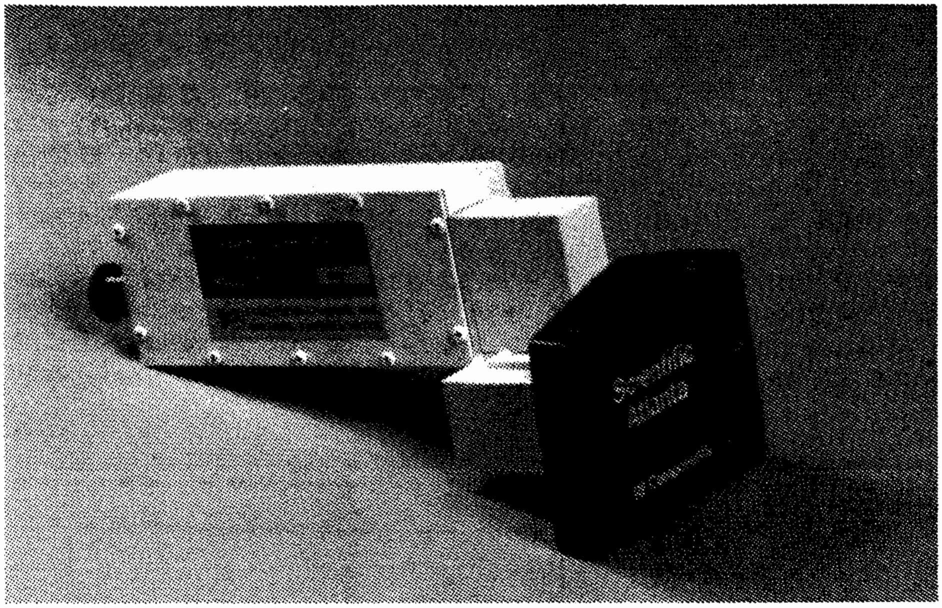

Figure 11. Scientific-Atlanta Production LNA

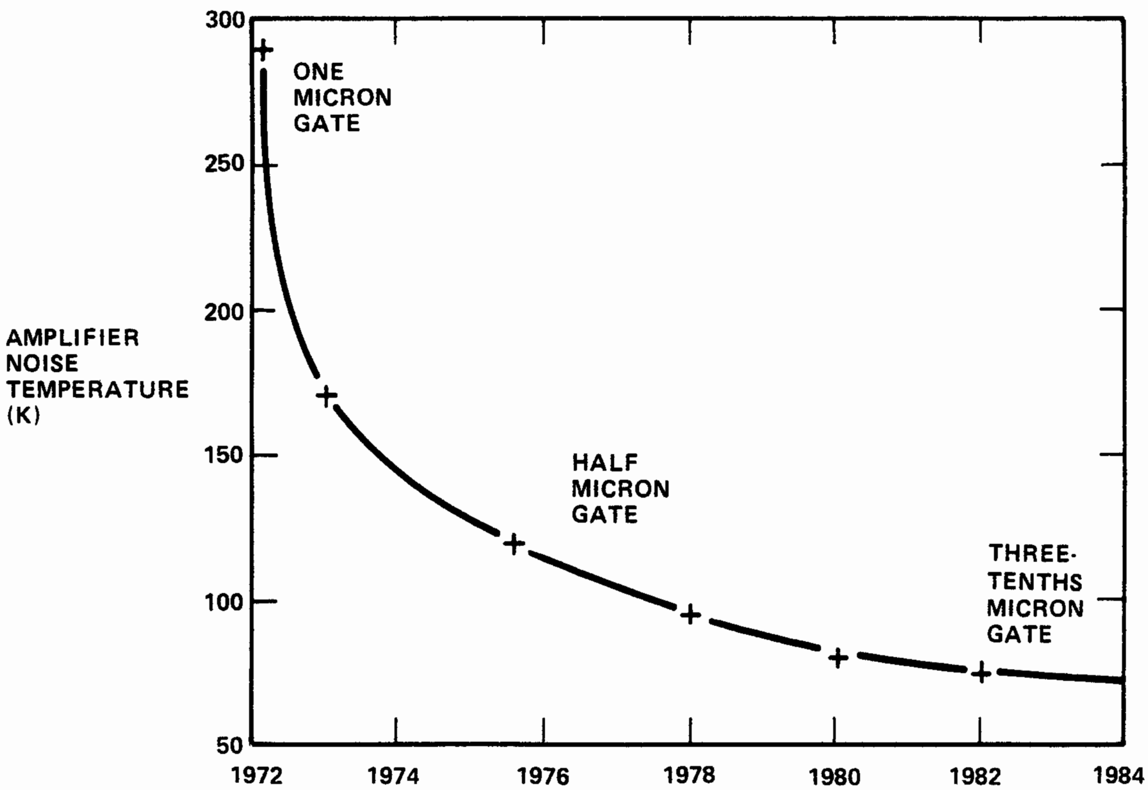

Figure 12. LNA Noise Temperature Vs. Time

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Future LNA Noise Figure Improvements

Since the development of the GaAs FET and its introduction to the market in 1972, there has been a steady decrease in the noise temperature of commercially available devices. This improvement in noise temperature is illustrated in Figure 11. Early FETs had one-micron gate widths, which gave noise temperatures of 290K (3-dB N.F.) in 1972. By the end of 1973, amplifiers with 180K noise temperatures were available. Half-micron gate-width devices were introduced in 1975, making possible 120K LNAs. Further improvement in device fabrication techniques, such as the use of a recessed gate, yielded devices capable of providing an uncooled LNA noise temperature of 80K. In 1982 a three-tenths micron device became commercially available which permitted achieving a 75K noise temperature. This new device was designed more particularly for 12 GHz than for 4 GHz, so the amplifier noise temperature improvement was not as dramatic as might be expected.

Referring to the summary noise figure calculations of Figure 6, it can be seen that certain losses are fixed regardless of the device used. These losses correspond to a noise temperature of 20K for an uncooled amplifier, with a perfect GaAs FET. If one extrapolates to the future using the noise temperature achieved versus time for an uncooled LNA, as shown in Figure 12, we can expect eventually to have LNAs available with noise temperatures of 60K or less.

Significant reduction in noise temperature is possible by cooling the first stage GaAs FET. An LNA noise temperature of 50K is attainable by use of a thermoelectric cooler to cool the FET to -40°C.

Summary

This paper has shown some of the design techniques used to meet the requirements for an LNA. While the LNA design is not simple, LNAs using the GaAs FETs have proven to be the most reliable and cost-effective way to achieve low system noise temperature earth stations. Device improvements of the future may result in uncooled LNAs with noise temperatures equal to those of parametric amplifiers at a fraction of the cost.

Dielectric Stabilized Local Oscillators for Block Converters

C.A. Bishop

Introduction

Scientific-Atlanta, as part of its ongoing effort to improve the quality and reliability of its low-noise block converters (LNCs), is developing various approaches to the local oscillator. Part of the effort in this area is directed toward developing the dielectric resonator oscillator for use in block converters. In this paper the requirements for a good local oscillator (LO) are discussed along with the advantages and disadvantages of several approaches to implementing this component. Information is also presented describing the dielectric stabilized oscillator (DSO) and the dielectric resonators used to stabilize these oscillators.

Why Are DSOs Used as Local Oscillators? Local Oscillator Configurations for Block Converters

The local oscillator is an important component in any heterodyne receiver including satellite block converters and satellite receivers. The local oscillator allows frequency conversion from a higher frequency (or range of frequencies) to a lower frequency (or range of frequencies). The characteristics of an LO are important since a poor oscillator can degrade the signal-to-noise ratio or the picture quality from a satellite receiver. A good LO should have low phase-noise, low drift, low harmonics, high-power output, and low cost.

Low local oscillator phase noise is an important requirement since phase noise on the LO will be downconverted along with the desired signal and will look like FM noise on the FM signal. When the signal is demodulated, the FM noise from the LO will also be demodulated and will appear as noise on the output video. In the demodulated video, the phase noise will appear as streaking that blurs picture detail; therefore, for high-quality video, low LO phase noise is essential.

Local oscillator drift can cause the desired signal to be downconverted off frequency and, if the drift is great enough, the IF signal will fall outside of the receiver passband. When this happens, a good demodulation cannot be expected since part of the IF signal is amplified less and phase shifted differently than the rest of the signal causing distortion and loss of signal-to-noise ratio. The drift specification can be difficult to achieve for the temperature variations seen by an antenna-mounted block converter; but due to the effect drift has on picture quality, the drift specification is clearly one of the most important LO requirements.

The harmonic output from the LO should generally be minimized since harmonics can cause excess production of spurious mixer products. The spurious mixer products are caused by mixing of desired signals plus their harmonics with the LO plus its harmonics. Therefore, it is important to minimize the LO harmonics and to bandpass filter the desired frequency band to minimize these spurious products.

The output of the LO drives a mixer which accomplishes the downconversion. Mixers typically require 3-7 dBm of power from the local oscillator, and the LO should be able to deliver this power plus a margin to allow for manufacturing variations. Extra power is also desirable since it allows the insertion of a resistive pad which helps to control load variations presented by the mixer.

Low cost is, of course, important since it allows Scientific-Atlanta to produce lower-cost downconverters for all of its satellite communications customers.

Type of LOs and their Advantages and Disadvantages

Several types of local oscillators could be used in satellite block downconverters including phase-locked loops, cavity stabilized oscillators, and dielectric stabilized oscillators. Phase-locked loops have good close-in phase noise and good stability, but they may have phase-noise problems at frequencies far from the carrier. Phase-locked loops also suffer from greater complexity than some of the other types of oscillators partly due to the need for frequency multiplication of the output signal.

Cavity oscillators have good stability and phase noise, but these oscillators are only used where size is not a constraint since the cavity can be quite large. Size, however, is certainly a constraint in antenna-mounted block converters. Manufacturing of the cavity to tight specifications may also be a problem with cavity stabilized oscillators. Futhermore, due to the dependence of cavity oscillators on the physical dimensions of the cavity (which change with temperature), it is difficult to maintain low-temperature drift with these oscillators.

Dielectric stabilized oscillators have all the advantages of cavity oscillators, but the size is much smaller since the cavity is replaced by a dielectric resonator. Additionally, DSOs can be easily mated to microstrip circuits which allows the oscillator to be built on printed circuit board. The manufacturing problems which exist with the cavity oscillator are not as severe with the dielectric stabilized oscillator since the dielectric resonators are made of a ceramic material. Since ceramic manufacturing processes are well developed, the ceramic resonators can be made to tight tolerances at low cost.

How is a DSO Contructed?

There are two major classes of DSOs--the feedback type DSO and the negative resistance type DSO. Both types of oscillators use the same size and shape resonator, but the circuit layout is somewhat different. In this section the properties of dielectric resonators will be discussed along with a description of both types of DSOs.

Properties of Dielectric Resonators

A dielectric resonator is a cylindrical-shaped piece of high-dielectric constant (er=38) ceramic with special properties. The resonators are typically made from either barium tetratitanate [Ba2Ti9020] or zirconium titanate [(Zr,Sn) Ti04]. These materials are chosen because they have a very small and controllable temperature coefficient of dielectric constant along with a low coefficient of thermal expansion. These characteristics are important since they cause the electro-magnetic properties of the resonator to change very little over large temperature variations. This helps a dielectric resonator stabilized oscillator to maintain low frequency drift over a large variation in ambient temperature. Dielectric resonators have another very important characteristic--very high "Q". High "Q" implies that the resonator has an extremely sharp resonance and very low loss. High "Q" resonators allow oscillators to be built that exhibit low phase noise, another characteristic of a good local oscillator. The low phase noise is a result of the sharp resonance of the dielectric resonator which allows the conditions for oscillation to be met at essentially only one, well-defined frequency.

The free space resonant frequency is determined by the physical dimensions and the dielectric constant of the dielectric cylinder (puck). The frequency is proportional to the height and diameter of the puck and proportional to the square root of the dielectric constant. The height-to-diameter aspect ratio is also important since it affects spurious resonances. An h/D aspect ratio of about 0.4 has been found to be optimum. When the puck is placed in a housing on a glass-teflon substrate, the dimensions of the housing also effect the resonant frequency. However, if the sides of the enclosure are far away (>2 X puck diameter), the only significant effects are caused by the height of the metal top above the puck and the thickness of the substrate. The only constraint on the top is that it must not be allowed to come to close to the puck or the "Q" of the dielectric puck will be degraded. The sensitivity to the metal top is fortunate, however, since this provides a convenient tuning mechanism (see Figure 1).

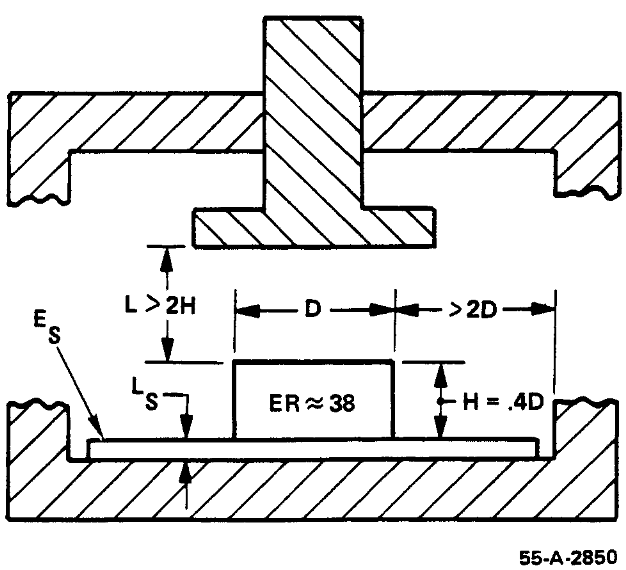


Figure 1.

The dielectric resonator when coupled to a microstrip line exhibits a parallel resonance, but this resonance can be changed to a series resonance with a quarter-wave section. By varying the distance to the puck from the transmission line, the coupling to the puck is changed and the magnitude of the reflection coefficient looking into the line can be changed. By changing the length of the transforming section, the angle of the reflection coefficient can be varied. Therefore, by changing these two parameter, coupling and position along the line, a wide range of reflection coefficients can be realized. The actual coupling mode from the microstrip line to the puck is magnetic as shown in Figure 2. This means that the puck should be placed at a high current point along the microstrip line for most efficient coupling.

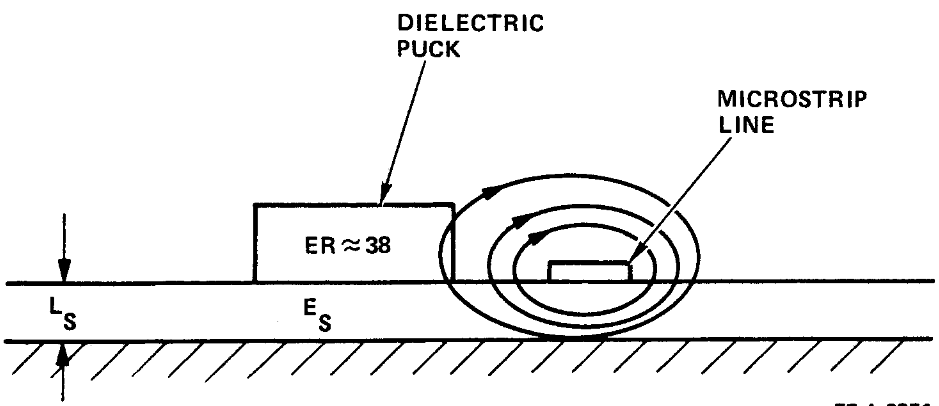


Figure 2.

One other important parameter of the dielectric puck is its temperature coefficient which sets the drift of the stabilized oscillator with respect to temperature. Furthermore, with pucks made from modern resonator materials, the designer can specify the desired temperature coefficient within certain limits. This allows the production of oscillators with even lower drift since the temperature coefficient of the puck can be chosen to offset any temperature variations encountered in the rest of the oscillator circuit. For example, if the unstabilized oscillator has a tendency to drift down in frequency with increasing temperature, the puck temperature coefficient can be chosen such that the puck drifts up on frequency with increasing temperature. The two drifts, then, tend to cancel each other and the oscillator frequency remains more nearly constant with temperature variations.

Feedback Type DSO

There are in general two classes of oscillators--the feedback oscillator and the negative resistance type oscillator. Both types of oscillators can be stabilized with a dielectric resonator. A feedback oscillator is implemented by feeding part of the output signal back to the input port in phase with the input signal to achieve positive feedback. This oscillator can be shown in schematic form (Figure 3) as an amplifier and a feedback network.

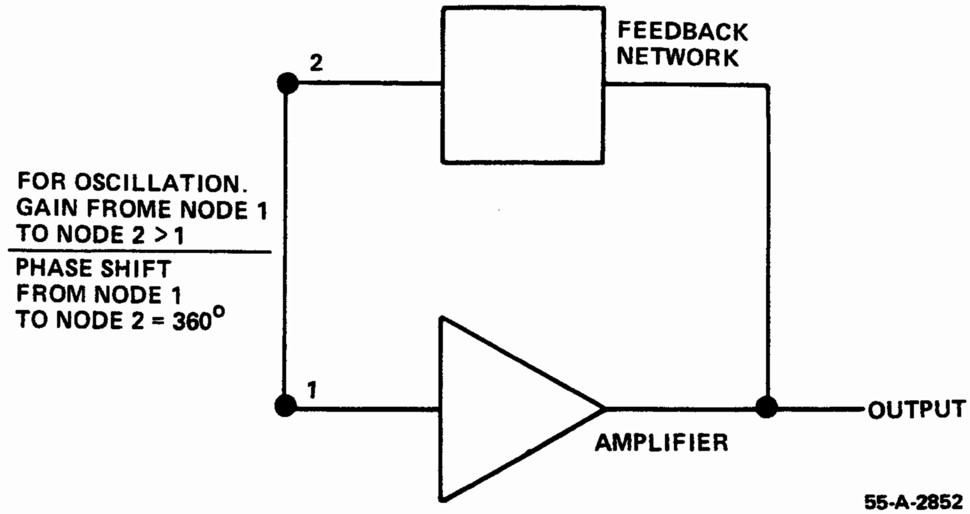


Figure 3.

Oscillation will occur if the loop gain is greater than unity and the total phase shift around the loop is 360 degrees. The power output will depend on the maximum output from the amplifier, and the stability and phase noise will depend on the phase variations around the loop.

A dielectric resonator stabilized feedback oscillator uses a dielectric puck as part of the feedback network. The puck along with the coupling structure provide a feedback network with a very sharp phase slope; thus, the 360-degree phase position is very sharply defined. Since the resonant frequency of the puck shows little variation with respect to temperature, drift is minimized; and since the phase slope of the puck is very sharp, very little phase noise can exist. There is, however, some loss of signal in the feedback network due to coupling losses, so to sustain oscillation, this coupling loss must be less than the gain of the amplifier. There is also a tradeoff here because the sharpness of the phase slope depends in part on how loose the coupling is to the puck; therefore, for the sharpest phase slope (lowest phase noise), loose coupling is desirable. But, loose coupling means greater loss, so a higher gain amplifier is required to maintain oscillation.

In a practical implementation, the amplifier is laid out in microstrip with transmission lines from the input and output running near each other (Figure 4). The dielectric puck is placed between these two lines and positioned in the center to realize a bandpass filter feedback network. With this configuration, a small amount of noise begins the oscillation. The noise is amplified by the amplifier, then fed back to the input through the dielectric resonator. The feedback signal is then amplified and oscillations build up until the amplifier reaches its maximum output.

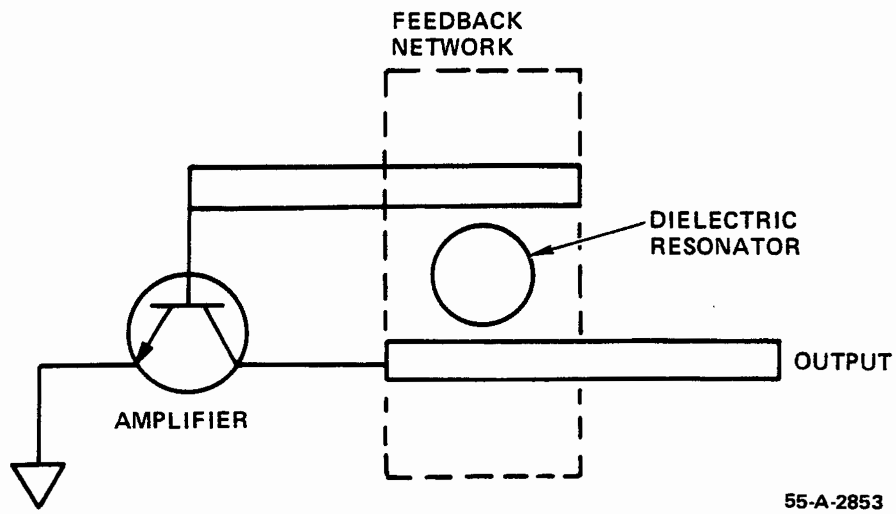


Figure 4.

Negative Resistance Type DSO

A negative resistance oscillator utilized an active negative resistance component such as a Gunn diode or a transistor configured to appear as a negative resistance. With this type of oscillator, the frequency of oscillation is limited by the range in which the device appears as a negative resistance and the actual oscillation frequency is determined by a resonant network placed in the circuit. A negative resistance component may be modeled as a negative resistance in series with a reactance (Figure 5). Oscillation will occur when the negative resistance equals the load resistance and the reactances are tuned out by a resonant circuit so that the net reactance is zero. The resonant circuit is often placed in the output circuit where it sets the output impedance at which oscillation will occur. If a Gunn diode is the active device, the resonator is placed in the output circuit (since a diode has only two terminals), but if a transistor is used for the active device, the resonant circuit may be coupled to any of the available ports.

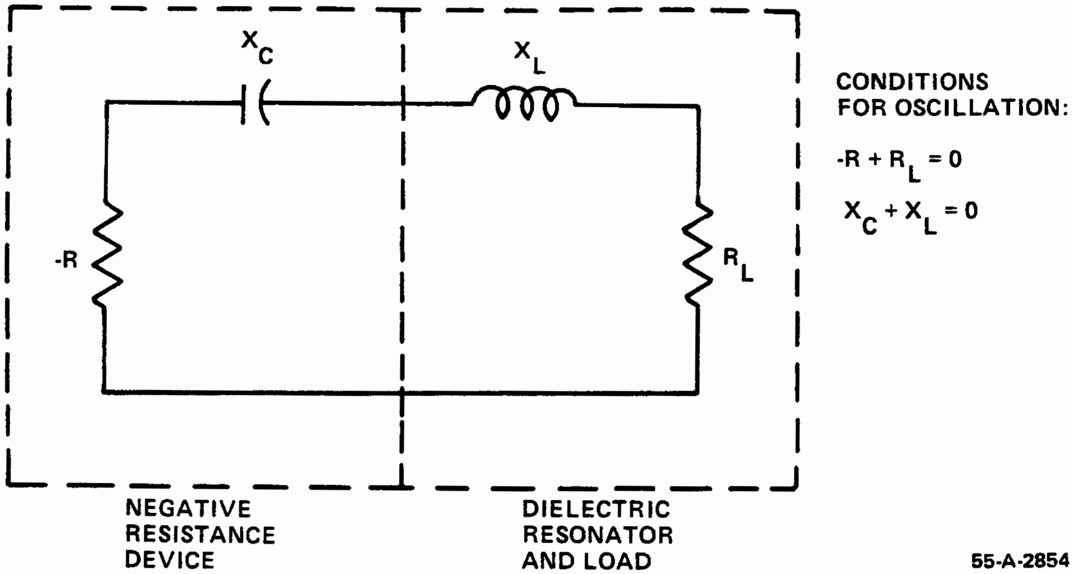


Figure 5.

The negative resistance oscillator can be stabilized by using a dielectric puck for the resonator. The puck is coupled to a microstrip line in the output circuit (or other port for the transisitor) to set the impedance seen looking into the microstrip line (Figure 6). The effective impedance looking into the line can then be varied by coupling the puck to the line differently and by moving the puck along the line (see the section on dielectric resonators). When the puck causes the conditions for oscillation to be satisfied, oscillation will begin. The impedance set by the puck will not change much with temperature, so drift is minimized since the oscillation will only exist when the proper impedance exist. Furthermore, since the "Q" of the puck is so high, the impedance curve is very sharp near resonance. This minimizes phase noise. The output power of the oscillator is determined by the power that the active device can produce into the impedance set by the dielectric puck at the frequency of oscillation.

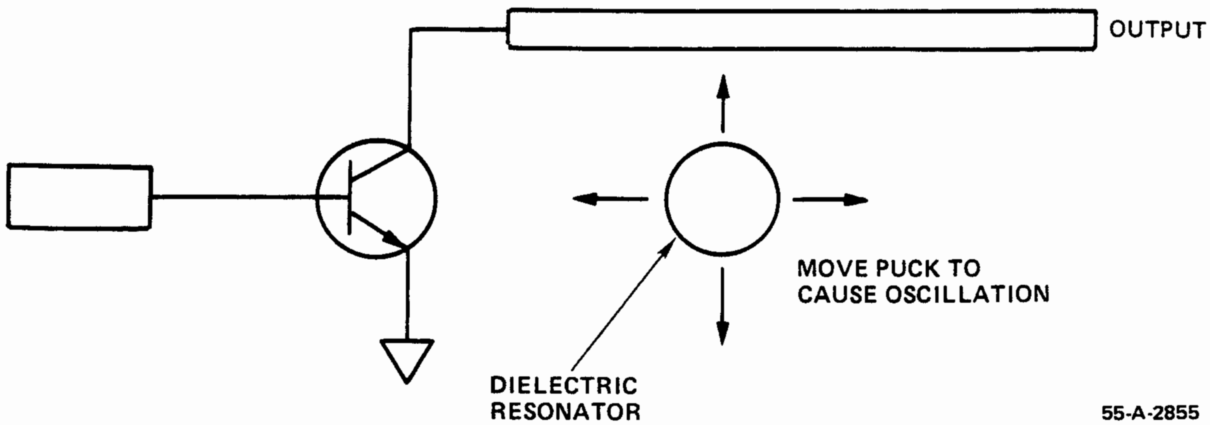


Figure 6.

Conclusion

The local oscillator is an important component in a satellite receiving block converter; therefore, Scientific-Atlanta is developing technologies which will lead to higher performance, more cost-effective microwave oscillators. One type of local oscillator that promises to be superior and more cost effective for many applications is the dielectric stabilized oscillator. DSOs have excellent properties when used as local oscillators, and they are reliable and economical to produce. In this paper the important properties of local oscillator as they relate to DSOs have been discussed. The properties of dielectric resonators have also been examined, and two general classes of DSOs have been described.

Digital Modulation: Characteristics and Performance of Representative Modulation Techniques

L. Montreuil

Introduction

The difference between analog and digital communications systems is that digital communications sends only a finite set of waveforms, in contrast with analog communications which can send an infinite set of waveforms. The objective of the receiver is not to reproduce the waveform but to determine from a noise perturbed signal which of the finite set of waveforms had been sent by the transmitter. The purpose of this paper is to describe the characteristics and performance of some modulation techniques used for digital transmission.

Amplitude Modulation (AM) Techniques

The simplest digital AM technique is Double Sideband (DSB) AM modulated by a binary signal. This modulation is represented by:

$$s(t) = \frac{A}{2} [1 + m(t)] \cos \omega_c t$$

where m(t) is the modulating signal and ω_c is the carrier frequency. For the case of 100-percent modulation by a nonreturn-to-zero (NRZ) binary data waveform $[m(t) = \pm 1]$, we have an On-Off-Keying (OOK) modulation. The two symbols used for this modulation can be written:

$$s_0(t) = 0$$

$$s_1(t) = A \cos \omega_c t$$

In such modulation technique, the carrier is present and demodulation can be done without a coherent demodulator.

Since the carrier conveys no information, efficiency can be improved by the use of double sideband suppressed carrier (DSB-SC) AM. The general form of DSB-SC signal is:

$$s(t) = Am(t) \cos \omega_c t$$

where $m(t) = \pm 1$ and the two signals are:

$$s_0(t) = A \cos \omega_c t$$

$$s_1(t) = -A \cos \omega_c t$$

The modulation is also called Binary-Phase-Shift-Keying (BPSK) which will be discussed under PM techniques.

The DSB techniques involve the transmission of a redundant sideband. For applications in which spectral efficiency is important, the bandwidth can be reduced by a factor of two by using only one sideband. The Single Sideband (SSB) modulation can be written as·

$$s(t) \quad = \quad A\left[m(t) \cos \omega_c t + \hat{m}(t) \sin \omega_c t\right]$$

where $\hat{m}(t)$ is the Hilbert transform of $m(t)$. SSB are often generated by filtering out one sideband; the sharp cutoff required presents some implementation problems; thus, a bandpass with smooth roll-off is often used. This procedure results in a Vestigal Sideband (VSB) signal.

Quadrature Amplitude Modulation is yet another AM alternative. This technique involves summing two DSB-SC in quadrature (90° apart) and can be written:

$$s(t) \quad = \quad A\left[m_i(t) \cos \omega_c t + m_q(t) \sin \omega_c t\right]$$

When $m_i(t)$ and $m_q(t)$ are equal to ±1, a QPSK signal is produced; if $m_i(t)$ and $m_q(t)$ are equal to +1, -1 or 0, and are correlatively coded, a Quadrature Partial Response (QPRS) modulation is produced (also called Duobinary technique). The number of possible states can be increased on each axis to increase the bandwidth efficiency resulting in a M-ary QAM, where M is the number of possible states in the constellation. In practice, the number of states on each axis M is a power of two, $M = 2^n$, the total number of states in the constellation is M^2 or 2^{2n} (figure 1e). Each symbol can be written as:

$$s_i(t) \quad = \quad \alpha_i \cos\left[\omega t + \theta_i\right]$$

where:

$$(i \quad = \quad 1, 2, 3, \ldots \quad M^2$$

Each symbol is characterized by an amplitude α_i, and a phase, θ_i, of the carrier; thus, QAM technique [also named Amplitude Phase Keying (APK)] is a hybrid of amplitude and phase modulation. Because of the ever increasing need for bandwidth conservation, this technique is becoming more popular; each symbol can convey many bits of information, thus reducing the symbol rate for a given bit rate.

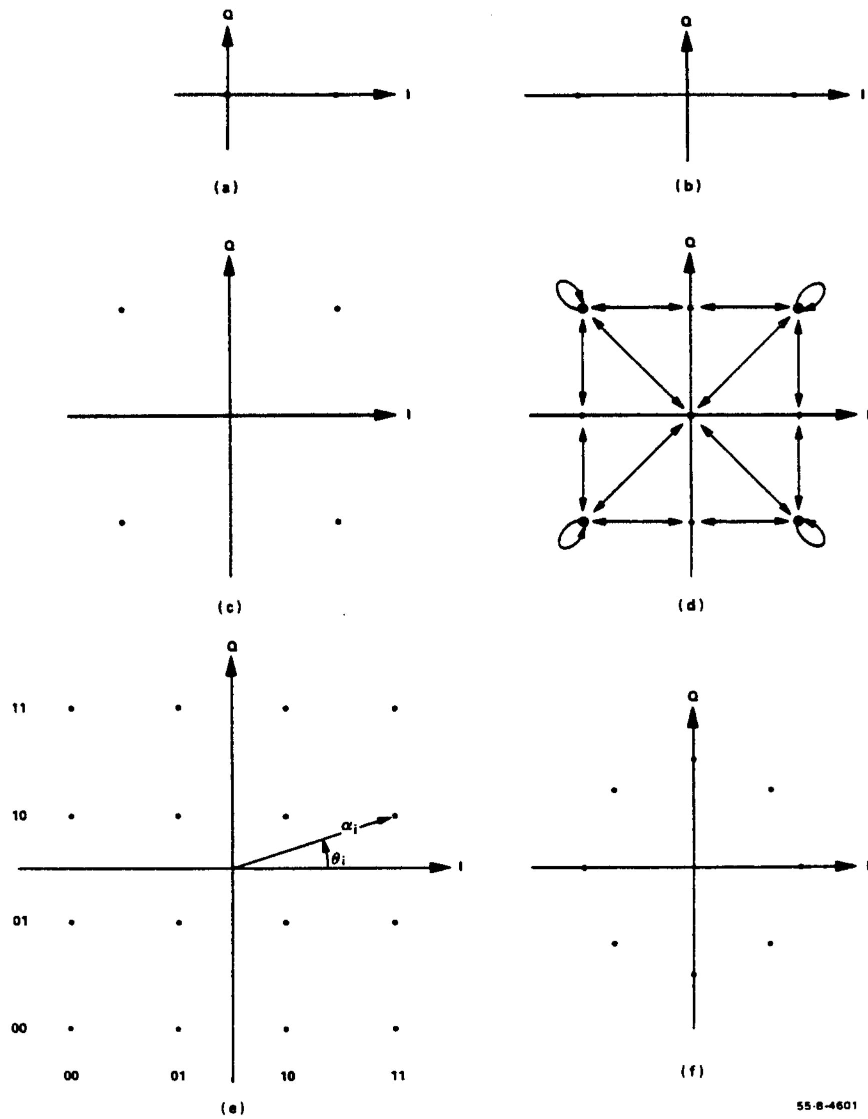


Figure 1. Phasor Diagrams (a) On-Off-Keying, (b) Binary-Phase-Keying, (c) Quadrature-Phase-Shift-Keying, (d) Quadrature-Partial-Response, (e) 16-Quadrature-Amplitude-Modulation and (f) 8-Phase-Shift-Keying

Frequency Modulation (FM) Techniques

The simplest FM techniques is Frequency-Shift-Keying (FSK) involving binary signaling by the use of two frequencies separated by Δf, where Δf is the frequency deviation. The binary signaling can be written:

$$s_1(t) = A \sin(\omega_1 t + \theta)$$

$$s_0(t) = A \sin(\omega_0 t + \theta)$$

where A is the amplitude of the signal, ω is frequency in radians/seconds and θ an arbitrary phase. For FSK schemes, the modulation index, d, is defined in terms of the frequency spacing, where:

$$d = \Delta f T = \frac{\Delta\omega}{2\pi} T$$

and T is the symbol duration. Recently, considerable interest has arisen in continuous phase FSK (CP-FSK) and in it's coherent detection. This modulation scheme permits fast spectral roll-off by eliminating abrupt phase changes at the bit transition, and coherent detection allows the use of a lower modulation index with improved performance.

The correlation between the two signal pair give a measure of their difference (or likeness). When two signals are opposed or antipodal, their correlation γ is -1; when two signals are orthogonal, their correlation is 0; and when two signals are indentical, their correlation is +1. The correlation of two signals is defined as:

$$\gamma = \frac{1}{E} \int_{0}^{T} s_1(t)\ s_2(t)\ d_t$$

and for FSK modulation:

$$\tau_{FSK} = \frac{\sin\left[(\omega_1 - \omega_0)T\right]}{\left[\omega_1 - \omega_0\right]T}$$

The coefficient γ is plotted in Figure 2 as function of normalized frequency difference $[\omega_1 - \omega_0]T$. The value of γ never reaches -1 so it's not possible to have an antipodal signaling. However, it can be seen that the most negative correlation value occurs when $(\omega_1 - \omega_0)T \approx 3\pi/2$ or when

$$d = 0.71$$

With that modulation index, the highest performance possible with a binary FSK is obtained, and the performance will be 1-γ = 1.21 or 1.59 dB better than orthogonal signaling. That is, a modulation index of 0.71 is considered a slight advantage over wide frequency deviation.

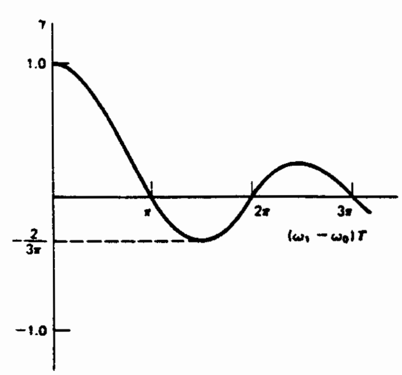


Figure 2. FSK Signal Correlation Coefficient Versus Frequency Separation

When d = 0.5, the FSK system operates at the first zero crossing of the correlation coefficient γ, which is the minimal deviation that can be used for orthogonal signaling. The common name for that signaling is Minimum-Shift-Keying (MSK). This technique achieves performance identical to coherent PSK and has better spectral properties; more power in the main lobe and less in the sidelobes. The signal is given by:

$$s(t) = \cos\left(\frac{\omega_c t + m_k \pi t + \theta_k}{2T}\right)$$

where:

ω_c = Carrier frequency (rad/sec)

m_k = Data input (± 1)

T = Bit period

θ_k = Constant phase valid for $kT<t<(k+1)T$

In the MSK signal function, excess phase function $\theta(t)$ is given by:

$$\theta(t) = \left(\frac{m_k \pi}{2T}\right) t + \theta_k$$

The phase transition between each states is done continuously (Figure 3), and the actual phase transmitted is a function of the previous phase transmitted, as shown in the phase trellis diagram (Figure 4).

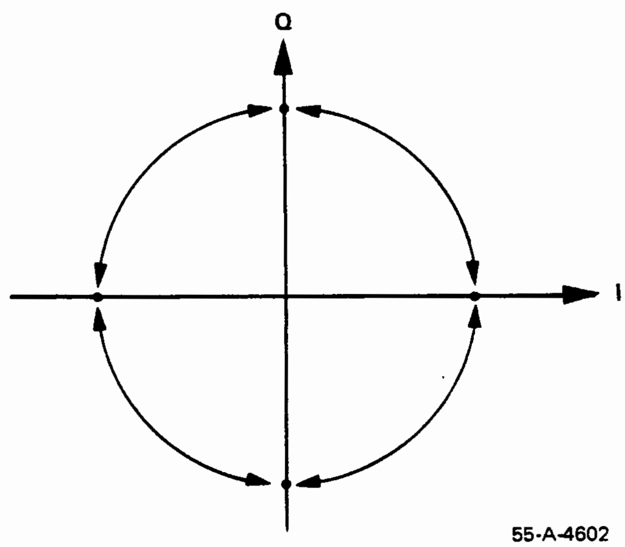


Figure 3. Phasor Diagram for MSK

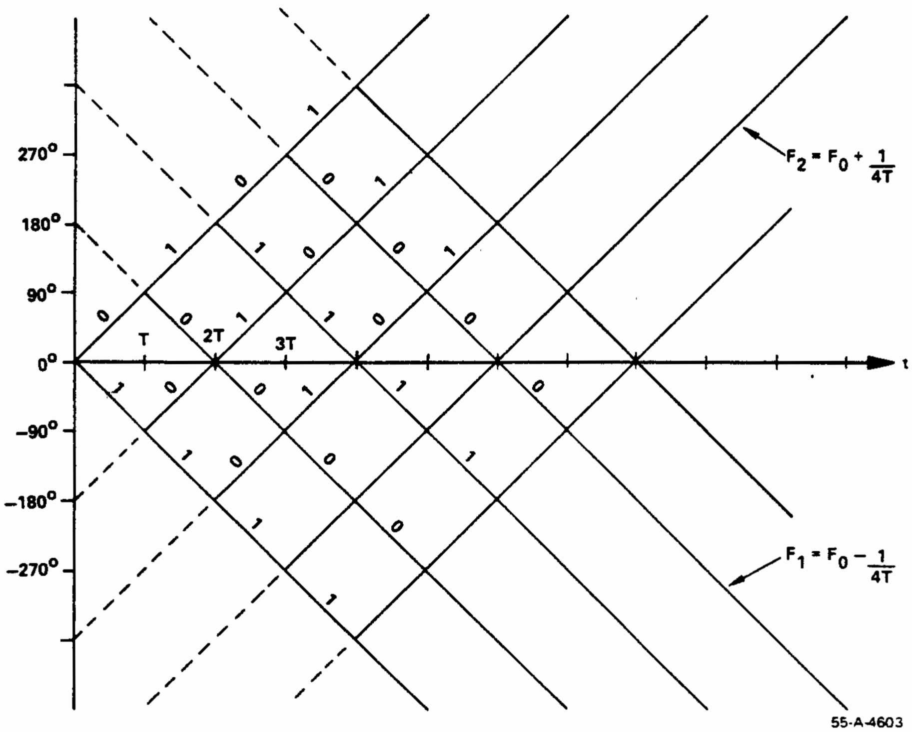


Figure 4. MSK Phase Trellis

Phase Modulation (PM) Techniques

Phase-Shift-Keying (PSK) modulation are commonly used phase modulation techniques; some are similar to AM techniques, like BPSK and QPSK, and have previously been discussed in AM techniques. These techniques need precise phase reference for coherent detection, which are usually obtained by nonlinear operation on the received signal like X^2 or X^4. These carrier-recovery techniques exhibit phase ambiguities of 180° or 90°. To overcome this problem, we Differentially Encode (DE) the data and the informations are not anymore carried by the absolute value of the phase of the carrier but by it's relative value from the previous symbol. Since DE-PSK techniques use two symbols to decide what information have been transmitted, the error rate performance is slightly inferior to PSK techniques. A modified version of QPSK, Offset-Keyed QPSK (OK-QPSK), has come in use in recent years because of his spectral proprieties. A conventional QPSK have, after filtering, large amplitude variation who, after passing through a nonlinear channel, regenerate the filtered sidelobes. In contrast, OK-QPSK does not have large amplitude variation because the data on the Q channel is shifted by T/2 second in respect to the I channel. When I and Q channels are added together, the resulting carrier vector can only shift by 90° instead of the 90° and 180° of the conventional QPSK.

Like multi-level AM system, multi-phase PSK (M-PSK) can be used to improve the spectral efficiency, but this technique is not popular for M over eight because of the non-optimal use of the power available and the complexity of the MODEM. Instead, techniques like M-QAM seem to be the best choice for future bandwidth conservative systems.

Comparison of Modulation Schemes

Systems Performance

In digital transmission systems, a measure of performance is the required ratio of energy by bit over the noise spectral density, E_b/N_o, to achieve a required bit error rate.

$$\frac{E_b}{N_o} = \frac{P}{N_o R}$$

where:

P = Signal power (watts)

N_o = Noise power spectral density (W/Hz)

R = Bit rate (bits/sec)

Table 1 and 2 give the theoretical performance of a representative modulation scheme without bandlimiting, and Figure 5 compares the efficiency and the performance of few modulation schemes with ideal filtering to the Shannon limit.

Table 1. Ideal Performance of Representative Modulation Scheme

Type	Modulation Scheme	$*E_b/N_o$ (dB)
AM	OOK coherent detection	11.4
	OOK envelope detection	11.9
	QPRS	10.7
	16-QAM	12.4
	64-QAM	16.5
FM	FSK noncoherent detection (d = 1)	12.5
	MSK	8.4
	DE-MSK	9.4
PM	BPSK	8.4
	DE-BPSK	8.9
	BPSK differential detection	9.3
	QPSK	8.4
	QPSK differential detection	10.7
	OK-QPSK	8.4
	8-PSK	11.8
	16-PSK	16.2

* For an error rate of 10^{-4}

Table 2. Probability of Error for Representative Modulation Scheme

Modulation	P_E
OOK Coherent Detection	$\frac{1}{2}\ \mathrm{erfc}\ [(E_b/2\ N_o)^{1/2}]$
BPSK, QPSK	$\frac{1}{2}\ \mathrm{erfc}\ [(E_b/N_o)^{1/2}]$
BPSK Noncoherent Detection	$\frac{1}{2}\ \exp\ (-E_b/N_o)$
FSK Noncoherent Detection (d = 0.5)	$\frac{1}{2}\ \exp\ [-E_b/2\ N_o]$
M-PSK	$\approx \mathrm{erfc}\left(\sin \frac{\pi}{M}\left(\frac{(\log_2 M)\ E_b}{2\ N_o}\right)^{1/2}\right)$

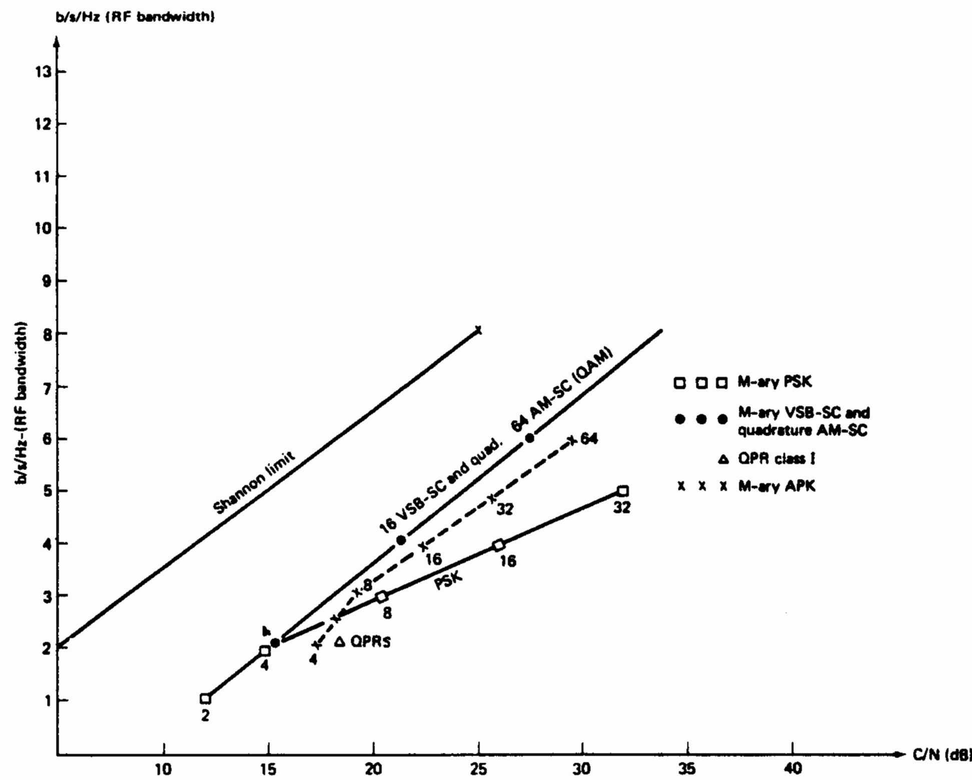


Figure 5. Spectral Efficiency and Performance of Various Modulation Scheme Configured to the Shannon Limit

Bandwidth of Digital Modulation

One important parameter of a modulation is it's bandwidth efficiency; the ratio of the bits rate by the occupied RF bandwidth. The implications of bandwidth can vary considerably from context to context, and consequently no single definition suffices.

Spectral Characteristics

For PSK or AM systems driven by nonfiltered NRZ data, it can be shown that the power spectral density is given by:

$$G(f) = \frac{2P}{R_S} \frac{\sin^2 [2\pi f/R_S]}{[2\pi f/R_S]^2}$$

where:

P = Modulated signal power

R_S = Symbols rate

That expression can be applied to BPSK, QPSK, MPSK, and QAM. For a bit rate R, then $R_S = R$ for BPSK; $R_S = R/2$ for QPSK; $R_S = R/3$ for 8-PSK; and $R_S = R/4$ for 16 QAM. For a M-ary modulation where M is an integer of 2 we have:

$$R_S = R/ (\ln M/\ln 2)$$

For binary continuous phase FSK systems, the spectral density is given by:

$$G(\omega) = \frac{2A^2 \sin^2 [(\omega - \omega_1)/2]T \sin^2 [(\omega - \omega_2)/2]T}{T[1- 2\cos (\omega - \propto) T \cos \beta T + \cos^2 \beta T]} \left(\frac{1}{\omega - \omega_1} - \frac{1}{\omega - \omega_2} \right)^2$$

$$+ \frac{2A^2 \sin^2 [(\omega + \omega_1)/2]T \sin^2 [(\omega + \omega_2)/2]T}{T[1- 2\cos (\omega + \propto) T \cos \beta T + \cos^2 \beta T]} \left(\frac{1}{\omega + \omega_1} - \frac{1}{\omega + \omega_2} \right)^2$$

where:

T = Bit period

A = Signal amplitude

ω_1, ω_2 = Signaling angular frequencies

$\propto$ = $(\omega_2 + \omega_1)/2$

β = $(\omega_2 - \omega_1)/2$

For the special case of MSK (d = 1/2), the power spectral density, with respect to the center frequency f_0, where $f_0 = (f_1 + f_2)/2$ and f_1, f_2 are the signaling frequencies, is given by:

$$G(f) = \frac{8\,P_c}{\pi^2\,R}\;\frac{1 + \cos(4\pi f/R)}{[1-(16/R^2)f^2]^2}$$

where:

R = Bit rate

Figure 6 gives the spectral density normalized to the bit rate for BPSK, QPSK, and MSK.

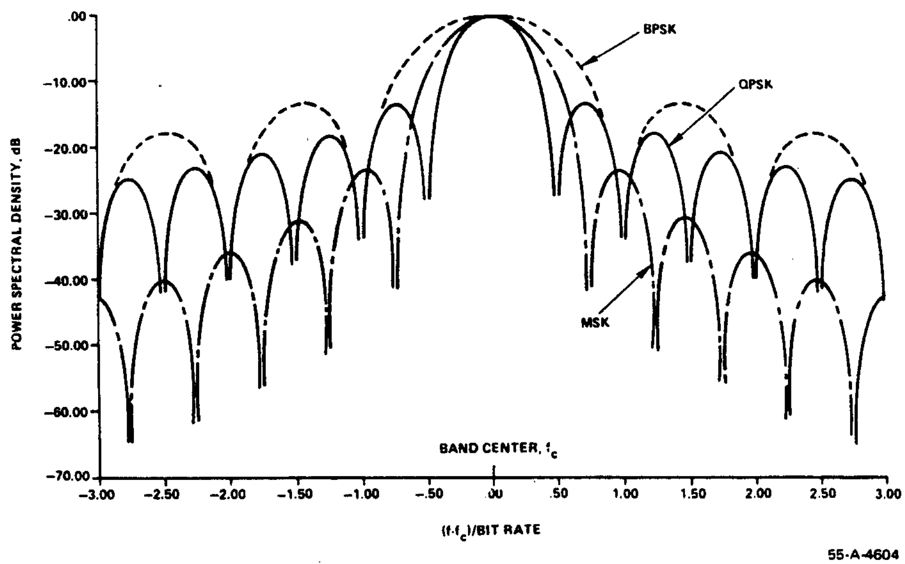


Figure 6. Spectral Densities of BPSK, QPSK and MSK Modulation

Null-to-Null Bandwidth

A simple and popular measure of bandwidth is the null-to-null bandwidth delimiting the main lobe. That definition of bandwidth is well defined for BPSK, QPSK, and MSK, but that is not the case for all modulation; some do not have well defined spectral null. For some modulation schemes considered here, the null-to-null normalized bandwidths are:

BPSK	2.00 Hz/R
QPSK	1.00 Hz/R
8-PSK	0.66 Hz/R
16-QAM	0.50 Hz/R
MSK	1.50 Hz/R

Fractional Power Containment Bandwidth

One definition of bandwidth (FCC Rules and Regulations Section 2.202) states that occupied bandwidth is the band which leaves exactly 0.5 percent of signal power on each side of the band limit. Thus 99 percent of the signal power is inside the occupied band. Figure 7 gives the fractional out-of-band power bandwidth versus the RF bandwidth normalized to the bit rate for various non-bandlimited modulation schemes.

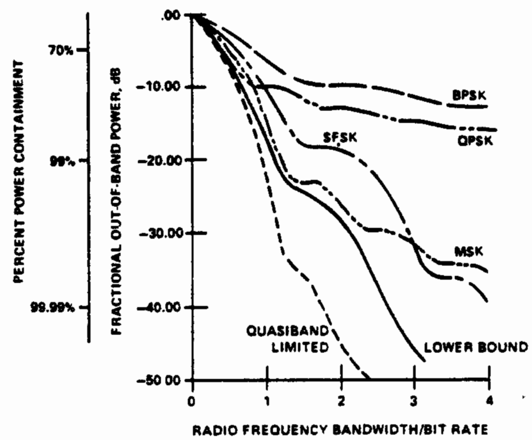


Figure 7. Fractional Out-of-Band Power for Various Modulation Schemes

The bandwidth for various modulation schemes using the 99 percent criterion are:

BPSK	20.56 Hz/R
QPSK	10.28 Hz/R
8-PSK	6.85 Hz/R
16-QAM	5.14 Hz/R
MSK	1.18 Hz/R

The high values for BPSK, QPSK, 8-PSK, and 16-QAM are due to the slow rate of spectral roll off f^{-2}, compare to MSK f^{-4}.

Half-Power Bandwidth

The half-power bandwidth is the bandwidth where the power spectral density has dropped 3 dB from it's peak value. Some values for various modulation schemes are:

BPSK	0.88 Hz/R
QPSK	0.44 Hz/R
8-PSK	0.29 Hz/R
16-QAM	0.22 Hz/R
MSK	0.59 Hz/R

Spectral Efficiency for Filtered Bandwidth

Table 3 lists the bandwidth efficiency and the required signal-to-noise ratio to achieved an error rate of 10^{-4} for some typical bandlimited systems. The results presented here were derived from various sources, thereby using different filters.

Table 3. Relative Signaling Speeds of Representative Modulation Schemes

Type	Modulation Scheme	Speed (b/s for Hz)	$^*E_b/N_o$ (dB)
AM	OOK coherent detection	0.8	12.5
	QPRS	2.25	11.7
	16-QAM	3.1	13.4
	64-QAM	4.5	--
FM	FSK noncoherent detection (d = 1)	0 8	11.8
	MSK	1.9	9.4
	DE-MSK	1.9	10.4
PM	BPSK	0.8	9.4
	DE-BPSK	0.8	9.9
	BPSK differential detection	0 8	10.6
	QPSK	1.9	9.9
	QPSK differential detection	1.8	11.8
	8-PSK	2.6	12.8
	16-PSK	2.9	17.2

*For an error rate of 10^{-4}

Ratio of the Peak Power to the Mean Power

For power limited transmission systems, such as satellite transmitter, RF amplifiers are often operated near saturation to enhance the efficiency and to raise the output power of the transmitter. For this application, modulation techniques having a constant power are more appropriate. For applications where the mean power is not equal to the peak power, it should be kept in mind that the power amplifier must not saturate on the peak state; if so, the performance will be degraded.

Table 4 gives the ratio of the peak power to mean power for a representative modulation scheme.

Table 4. Ratio of the Peak to the Mean Power

Type	Modulation Scheme	Ratio of the Peak to Mean Power (dB)
AM	OOK	3.01
	QPRS	3.01
	16-QAM	2.55
	64-QAM	3.68
FM	FSK	0.0
PM	BPSK	0.0
	QPSK	0.0
	8-PSK	0.0

Summary

In selecting a modulation technique, the following goals are considered:

1. Maximize the transmission rate, R
2. Minimize probability of bit error rate, P_E
3. Minimize required energy by bit, E_b
4. Minimize the required bandwidth.

However, goals (1) and (2) are in conflict with goals (3) and (4)--in a pratical system, there are several constraints and limitations that necessitate the trading-off of any one requirement with each of the others. Some of the constraints are: the Nyquist theoretical minimum bandwidth requirement, the Shannon capacity theorem, the Shannon limit, government regulations, technological limitations and the cost. Thus, before choosing a modulation technique, the goals, requirements and trade-offs must be considered before making a decision. An operator should never choose a modulation technique and afterwards the channel, but characterize the channel first then choose the modulation fulfilling objectives, leaving some room for the design margin.

Bandwidth-Efficient, High-Speed Modems for Cable Systems

J.M. Rozmus

Introduction

The advent of well-designed coaxial cable networks covering large, metropolitan or suburban areas has led to a high degree of interest in the use of these networks for data communications. With the requirement for commercial systems of C/N >36 dB, i.e. C/N_0 >102 dB-Hz, and the ability for good designs to exceed this by as much as 9 dB [1], complex, wideband digital modulation formats can be supported. Because commercial systems are designed to distribute primarily video modulation formats, the transmission system must be maintained in a highly linear mode, thus modulation schemes which may suffer when confronted with the nonlinearities inherent in satellite communications, for example, may be employed with relative inpunity on the coaxial medium. All of the above factors became apparent resulting in the current industry drive to develop data communications capability on existing, planned, and future coaxial systems.

In September 1981, Scientific-Atlanta installed a demonstration data link in order to show the feasibility of providing wideband data services via coaxial cable. This initial installation provided local distribution of 96 voice channels between ISACOM, Inc., operating Satellite Business Systems (SBS) digital earth terminals in Atlanta and Houston, and one of their customers, National Data Corporation (NDC). This link was established with a dedicated cable pair (for redundancy) installed and maintained by South Media, Inc. The link provided NDC with voice and data services between their offices in Atlanta and Houston. Two crucial features of data communications over coaxial systems were demonstrated. The first was that transmission performance was limited only by terminal equipment. The cable medium performed transparently. The second feature involved communication economics. Coaxial cable was unrivalled when cost was a consideration.

Since the installation of this first link, a third fundamental principle was uncovered. While the cable offers unparalled cost and performance for high-speed data communications, its bandwidth must be treated as a highly prized commodity. The installation of wideband full duplex links at T1 (1.544 Mbps) and T2 (6.312 Mbps) rates employing modulation techniques with moderate bandwidth efficiency, e.g. 0.5 bits/Hz, would only permit 19 T1 links or 4 T2 links on a midsplit coaxial cable network as shown in Figure 1. Therefore, the capacity of the network is greatly enhanced when more efficient modulation techniques are applied. At 2.0 bits/Hz, 76 T1 links or 19 T2 links can be supported. Application of current digital communications technology is essential in order to effect more efficient utilization of the available bandwidth. It was therefore determined that an approach was necessary which would maximize spectral efficiency without driving manufacturing costs to prohibitive levels. A technique satisfying these requirements has been developed and is the subject of the remainder of the paper.

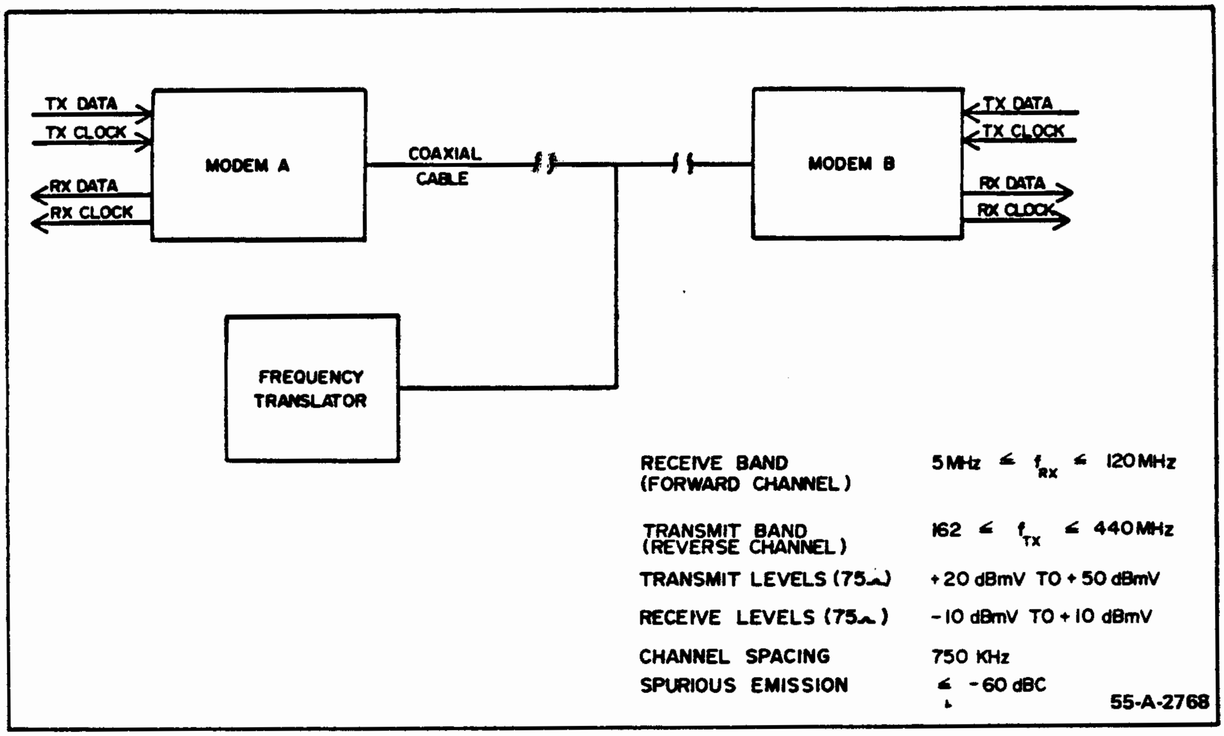


Figure 1. Communication System Configuration

System Design

In order to meet the necessary bandwidth efficiency requirements for high-speed data transmission, the detailed design of the modulation format was essential. While spectral efficiencies of 0.5 bits/Hz to 1.0 bits/Hz are readily achievable, it was determined that 2 bits/Hz could be realized without undue additional manufacturing cost. This was possibly due to the ease of implementation of the selected QASK-16 signal constellation, in both the modulation and demodulation processes. QASK-16 is a special case of the M-ary Amplitude-Phase-Shift-Keyed (MAPSK) family of signal sets which provide enhanced bandwidth efficiency through efficient signal packing at the expense of bit error probability, P_b, in the noisy environment. However, since coaxial systems provide high signal-to-noise ratios, these complex modulation formats are applicable.

A performance goal of $P_e < 10^{-9}$ was established for the data link thereby constraining the modulation complexity due to the fixed limit of $C/N_o > 102$ dB-Hz. A large body of literature exists on studies and hardware implementations covering many of the MAPSK signal sets [2-11]. After an evaluation of the myriad possibilities, it was determined that QASK-16 offered the best compromise of spectral efficiency, P_b, and cost of implementation. The average symbol signal-to-noise ratio, $R_d \triangleq ST/N_o$, is easily found to be $R_d > 46$ dB for a coaxial system operating at marginal performance levels or better and a T1 bit rate (QASK-16) transmits four bits per symbol). However, data signals cannot be permitted to operate at power levels in excess of the video signals with which they must co-exist. Therefore, it was determined that the power of each digitally modulated carrier be 15 dB below video levels for T1 and 9 dB for T2. This results in a total channel power (6-MHz channels) 6 dB below a corresponding video channel when the channel is fully packed with T1 or T2 data carriers at 0.75- MHz and 3-MHz spacings, respectively. This requirement reduced the worst case to R_d >31 dB for both rates. In order to achieve a symbol error probability, Ps, of 10^{-9}, R_d >26 dB is a lower bound when implementation losses are included. Thus, 5 dB of system margin exists for the QASK-16 modulation scheme.

The QASK modulation format selected from the family of MAPSK signal sets does not represent the optimum signal set; however, it has been shown to be degraded from the optimum by only tenths of a dB [3,5]. This penalty is insignificant when faced with the complexities of implementation of the alternatives. Through the combination of several decision-directed or decision-feedback techniques, all of the functions necessary for demodulation of the QASK signal set will be shown to be readily implemented.

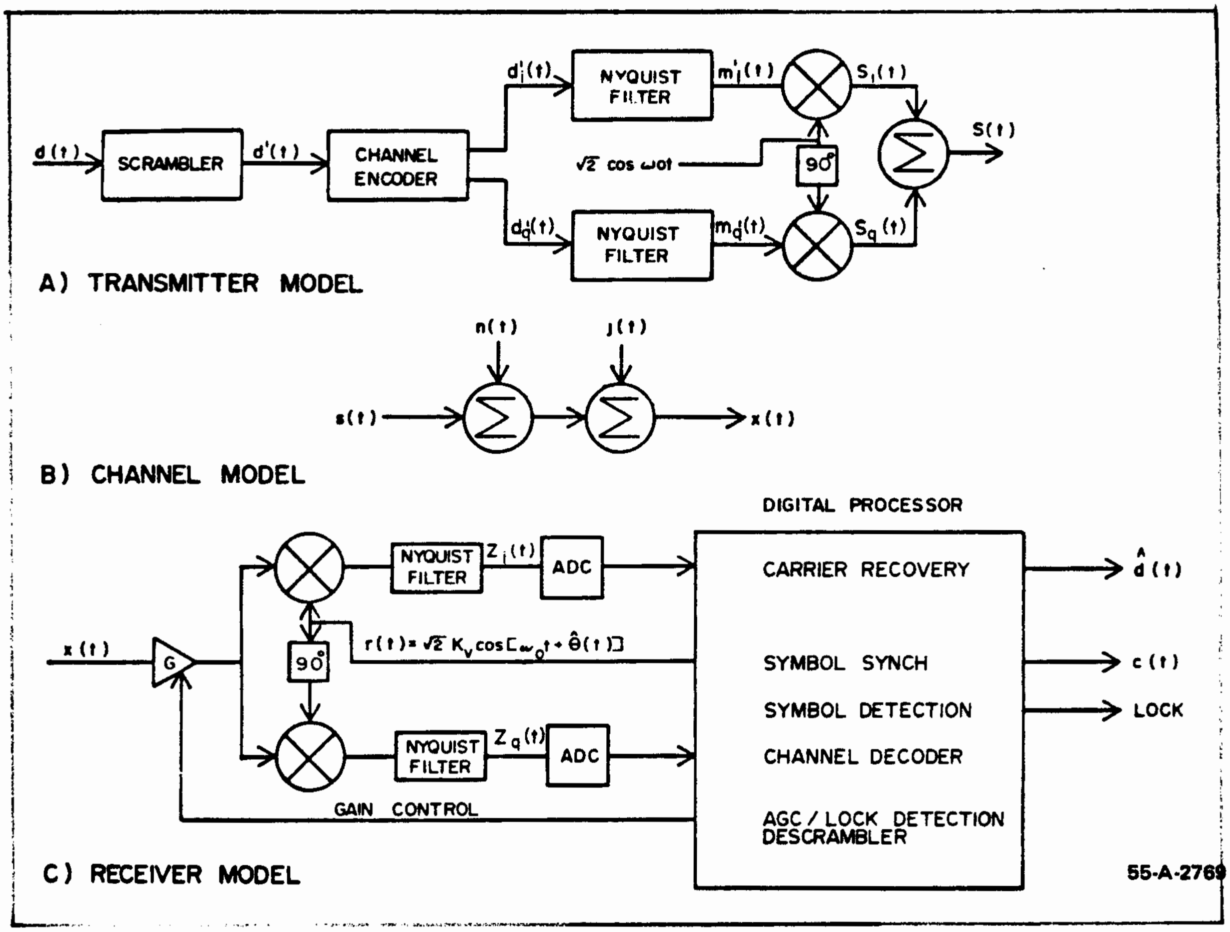


Figure 2. System Models

System Model

The basic models for the QASK communication link are shown in Figure 2. The incoming data stream, d(t) is scrambled to ensure adequate symbol transitions, then taken four bits at a time and differentially encoded [12], then filtered to provide minimum bandwidth and inter-symbol interference (ISI) and impressed on quadrature carriers. The transmitted signal s(t), and M-ary quadrature-amplitude-shift-keyed (QASK-M) signal with a symbol interval of T-seconds, can be represented mathematically as

$$S(t) = \sqrt{2}\,[m_i'(t)\cos w_o t + M_q'(t)\sin w_o t] \qquad (1)$$

where m_i (t) and M_q (t) are scrambled, encoded and filtered pulse trains. These quadrature pulse trains take on equally likely values $j\delta$ with $j = \pm 1, \pm 3, \ldots, \pm(K-1)$ in each channel. Thus for K = 4, the case of interest here, $m_i'(t)$ and $m_q'(t)$ are the filtered versions of amplitude-shift-keyed (ASK) inputs with equally likely values of $\pm\delta$, $\pm 3\delta$, resulting in the QASK-16 signal set with two amplitudes and two phases in each quadrature channel. The average signal power of the transmitted signal set is

$$S = 2/3(K^2 - 1)\delta^2. \qquad (2)$$

This transmitter model is depicted in Figure 2a.

The channel shown in figure 2b is assumed to be an additive white Gaussian noise (AWGN) channel where the noise n(t) has a two-sided spectral density $N_o/2$W/Hz. In addition, the channel adds a random phase shift to the signal s(t) such that the received signal is of the form,

$$\begin{aligned} x(t) &= s[t, \theta(t)] + n(t) + J(t) \\ &= \sqrt{2}\, m_i'(t)\cos[w_o t + \theta(t)] \\ &\quad + m_q'(t)\sin[w_o t + \theta(t)] \\ &\quad + n(t) + J(t)\,\} \end{aligned} \qquad (3)$$

where $\theta(t) \triangleq \theta_o + \Omega_o t$, with θ an uniformly distributed phase shift and Ω_o the frequency shift from its nominal value of w_o. The additional interence signals, though present in practice, are assumed negligible in the discussion to follow.

RF Section

The RF Processor is comprised of the transmit and receive IF assemblies as well as the transmit power amplifiers, receive amplifiers, and the synthesized local oscillators (LO's). A mid-split diplexer is used to interface the transmit and receive assemblies to the cable. While filtering is performed in both the transmitter and receiver of the 6402, none of these filters affects the modulated signal. The spectral efficiency is achieved through baseband filtering in the baseband processor. The design philosophy adopted for the RF Processor was to employ standard components in a way such that highly reliable RF signal processing was possible without generating interference which would affect the performance of other signals on the coaxial network.

Transmitter

The transmitter is composed of several elements: The quadrature modulator, synthesized transmit LO, and power amplifiers. The modulator employs a 145-MHz TCXO as an IF, which is then modulated by the in-phase and quadrature ASK symbols, $m_i'(t)$ and $M_q'(t)$ as described previously. These quadrature signals are summed resulting in the QASK-16 signal. This signal is filtered with a broad IF filter, amplified, and then translated to the transmit frequency by the transmit LO which results in 0.75-MHz channel spacing across the reverse channel, 5 to 102 MHz. Filtering in the transmitter amplifier stages rejects the TX LO such that the worst-case spurious output is -60 dBc.

Receiver

The receiver assemblies are composed of broadband amplifiers, the synthesized receive LO, and the quadrature demodulator. In order to prevent leakage of the RX LO onto the cable, each receiver broadband amplifier is preceded by a compensating attenuation, such that the net gain is essentially 0 dB. This inserts the necessary isolation to keep the spurious levels due to the RX LO at <-60 dBc. The RX LO then provides the necessary conversion frequency, with 0.75-MHz resolution, to translate the receive channel to the 150-MHz receive IF. In the IF, the signal is amplified and filtered with a broad channel filter with 10-MHz bandwidth to reject undesired channels. A pin diode attenuator is employed in the receive IF for processor control of the input levels. This signal is then applied to two mixers using the reconstructed carrier references as supplied by the Baseband Processor to demodulate the incoming signals to baseband. At this point the quadrature baseband signals are applied to the Baseband Processor inputs for digitizing.

The receiver model is shown in Figure 2c where the input signal, $X(t)$ is multipled by a locally generated quadrature reference,

$$r(t) = \sqrt{2} \cos \left[w_0 t + \theta(t) \right] \tag{4}$$

where $\hat{\theta}(t)$ is the local estimate of $\theta(t)$. Because the case of interest here is that of very high R_d, the noise will be neglected in further discussions; the case of low-to-moderate R_d is treated adequately in the references.

The quadrature signals multiplied by the reference and again Nyquist filtered are represented as

$$z_i(t) = m_i'(t) \cos \phi(t) + m_q'(t) \sin \phi(t)$$

$$z_q(t) = m_i'(t) \sin \phi(t) + M_q'(t) \cos \phi(t)$$

where $\phi \triangleq \theta(t) - \hat{\theta}(t)$ is the carrier recovery loop phase error. The baseband signals, $z_i(t)$ and $z_q(t)$ are then quantized in an analog-to-digital converter, and processed in order to recover the carrier phase, process symbol synchronization, detect the transmitted symbols, control the gain and detect lock. The algorithms necessary to provide these functions are implemented in a digital processor, the details of which are left to a subsequent section.

The carrier recovery algorithm employs a decision-feedback technique analyzed by Simon and Smith [13]. The symbol synchronization algorithm is a generalized data transition tracking loop, similar in concept to that analyzed by Simon [14-15]. The AGC algorithm is a decision-directed technique as analyzed by Weber [16], with the lock detection algorithm employing the AGC error signal as its decision criterion. The channel encoding is essentially differential encoding and is necessary in order to remove the quadrant ambiguity in the received symbols. The filtering for bandwidth efficiency, or Nyquist filtering, compresses the transmitted spectrum to achieve the 2 bits/Hz spectral efficiency while minimizing intersymbol interference (ISI). The filter is partitioned between the transmitter and receiver in order to minimize adjacent channel spillover and adjacent channel interference, respectively.

The modem developed for the mid-split cable system is designated the 6402 High Speed Modem. A functional block diagram of the 6402 is given in Figure 3. Each of the elements in the diagram are described in more detail in later paragraphs. The salient features are the frequency agility provided by transmit and receive synthesizers, low spurious emissions allowing for reliable coaxial network operation, and a high-performance baseband detection technique resulting in $P_b = 10^{-9}$.

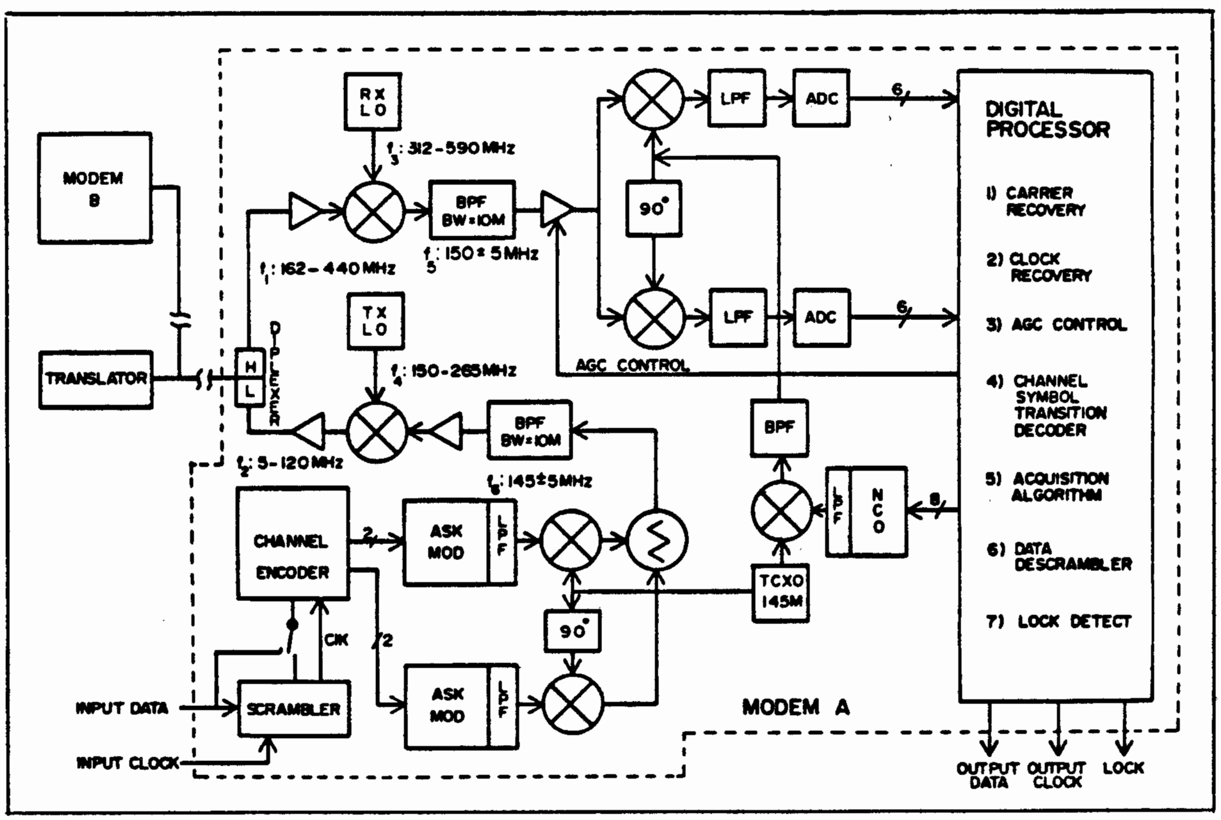


Figure 3. QASK Modem Functional Block Diagram

Baseband Processor

Technology

With careful use of a few state-of-the-art components, it was possible to build most of the baseband processor with standard MSI-TTL technology rather than more expensive ECL. An example of critical section is the numerically controlled oscillator. The 5-MHz NCO requires the fastest TTL PROM and TTL resisters commercially available.

Algorithms

The major functions of the baseband processor are shown in the block diagram of Figure 4. The carrier synchronization section keeps the 150-MHz local oscillator in phase with the incoming carrier by controlling the phase of the 5-MHz NCO output which is mixed to produce the local oscillator. The symbol synchronization section keeps the symbol clock synchronized with the symbol periods of the incoming baseband signals. The automatic gain control (AGC) controls the amplitude of the baseband signals being sampled by the analog-to-digital converters. The lock detection provides a positive indication when the processor is synchronized to a legitimate data transmission. The transmit section consists of a scrambler to ensure a sufficient rate of symbol state changes and an encoder to resolve the ambiguity that derives from the rotational symmetry of the symbol vectors in the I-Q plane. (See detailed explanations below.) The decoder and descrambler in the receive section provide the inverse functions of the encoder and scrambler.

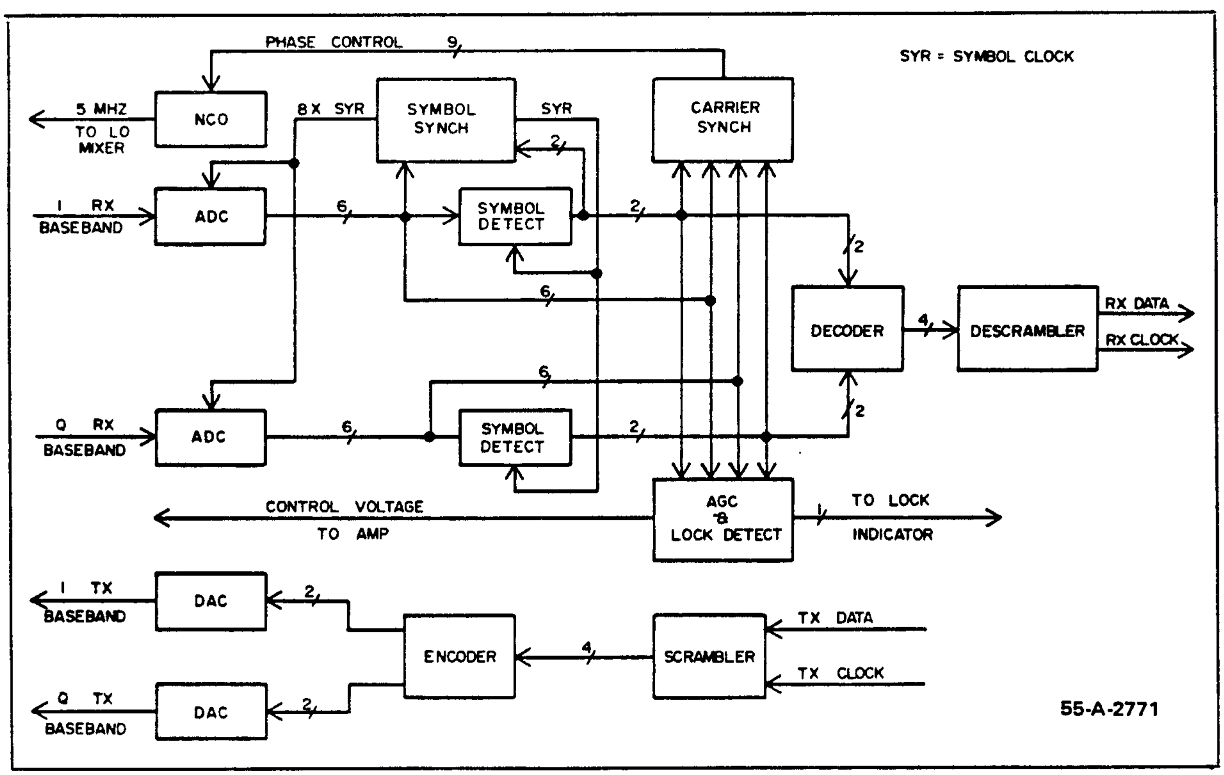


Figure 4. Baseband Processor

The input to all of the algorithms of the baseband processor is the series of 6-bit digital samples of the analog baseband signals of the I and Q channels. The baseband signals are sampled eight times per symbol period. One of the eight samples occurs in the middle of the symbol period. This is the symbol sample. It is a digital measure of the symbol level which is decoded by the symbol detect circuitry. The four levels of a perfect baseband signal produce digital values of +24, +8, -8, -24 for the +3, +1, -1, and -3 levels, respectively.

To understand the algorithms of the baseband processor, it is best to visualize the state of the baseband signals as one of 16 vectors in the I-Q plane (see Figure 5). For example, if the I level is +3 and the Q level is +1, then the corresponding vector points from the origin to the point in the lower right-hand corner of the (+, +) quadrant. Its phase is $\tan^{-1}(1/3)$.

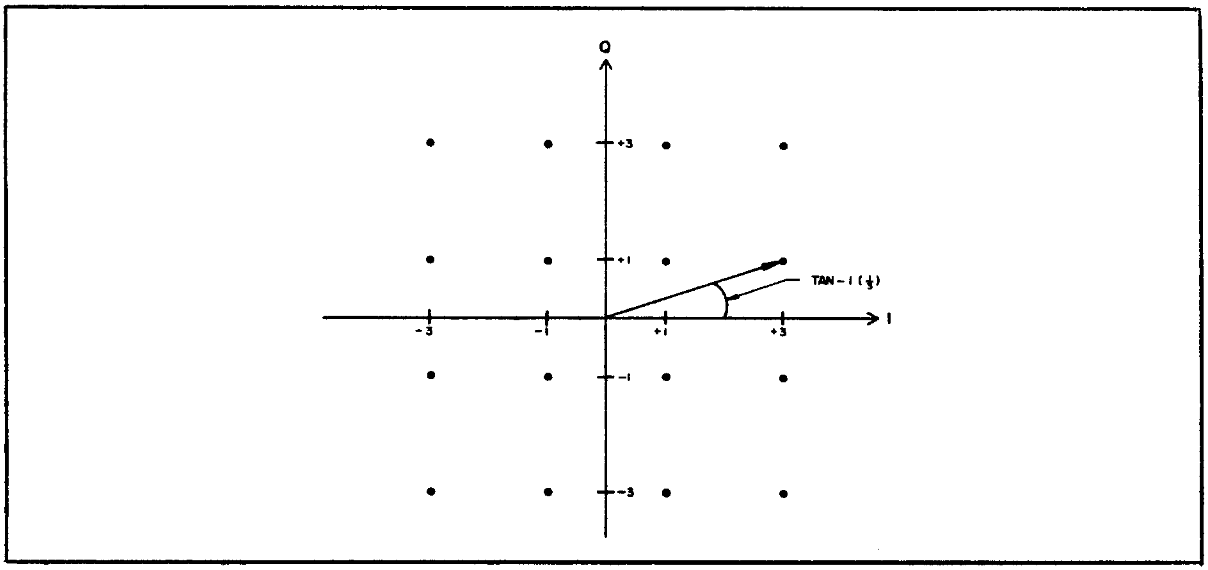


Figure 5. Symbol Vectors

The carrier synchronization algorithm compares the phase angle of the observed symbol vector with the phase of the expected ideal vector for the current symbol. For example, if the I and Q symbol samples are +27 and -6 respectively, the level detect circuits will detect (I,Q) equals (+3, -1) and the carrier synch circuitry will compare $\tan^{-1}$ (-6/+24) to $\tan^{-1}$(-1/+3). The difference is scaled and used to control the phase of the 150-MHz local oscillator. The resulting phase-locked loop has a loop noise bandwidth of 50 kHz and a damping factor of 1.0. It is a perfect second-order loop.

The symbol synchronization algorithm functions are as follows. Samples of the I channel baseband are summed from the center of one symbol period to the next. If a large symmetrical transition (i.e., +3 to -3, or -3 to +3) occurred during that time, the result should be zero. If the result is not zero, then the sign of the sum, along with the direction of the transition, indicates whether the symbol clock is lagging or leading the received symbol rate. These indications are averaged overtime, and the output controls the frequency of the symbol rate clock.

The automatic gain control computes the difference between the absolute value for the symbol sample and the absolute value of the expected ideal sample value. This difference is accumulated for 16 symbol periods. The result is summed with the current gain control level to change the gain of the variable-gain amplifier.

The lock detection circuitry computes the same difference as the AGC, but accumulates the absolute value of this difference for 16 symbol periods. If the modem is not in the lock state, a result less than the threshold will increment the lock counter. When the lock counter reaches a trigger value, the lock state is entered. If the modem is in the lock state, a result greater than the threshold will increment the unlock counter. If the unlock counter reaches its own trigger value, the modem leaves the lock state.

The 4-bit symbols are encoded as shown in Figure 6[12]. The upper two bits are encoded differentially by indicating the change in quadrant from the previous symbol. The lower two bits are coded by position within the quadrant. The differential encoding is necessary because the demodulator cannot distinguish among the four phases of the carrier. For example, if the previous transmitted vector was in the (+,+) quadrant and the current data is 0110, then the encoder will cause a (+1, -3) vector to be transmitted. The receiver may distinguish any of the following combinations:

Previous Quadrant	Current Vector	Quadrant Change
(+, +)	(+1, -3)	+90°
(+, -)	(-3, -1)	+90°
(-, -)	(-1, +3)	+90°
(-, +)	(+3, +1)	+90°

In all four cases, the correct data (0110) is decoded.

The scrambling algorithm (17) is essentially an exclusive OR combination of

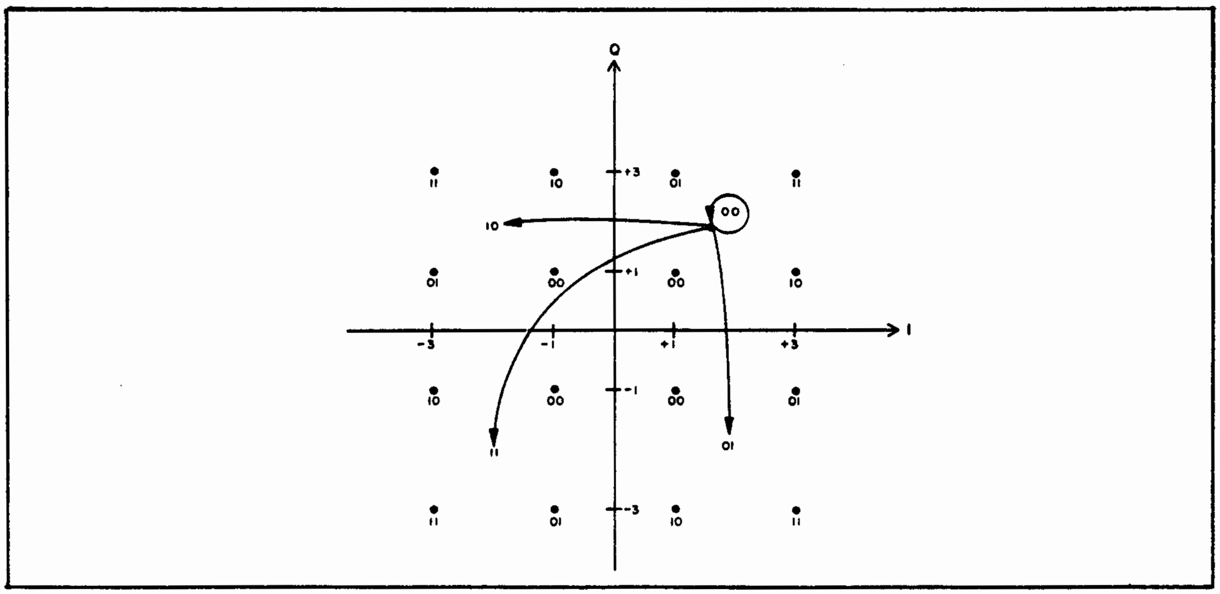


Figure 6. Symbol Coding

the 20th previous, 3rd previous, and current data bits. In addition, the polarity of the output is inverted whenever four consecutive 8-bit output sequences are identical.

Conclusion

The design of the 6402 High Speed Modem was accomplished using state-space simulation techniques for performance verification. Each of the elements in the receiver and modulator were modeled and tested completely prior to the hardware development. Laboratory tests have provided further verification using Scientific-Atlanta's 400-MHz headend and distribution system. The frequency-agile, point-to-point modem represents a strong choice for high-speed communications over cable, with its bandwidth efficiency maximizing the spectral utilization of the coaxial cable system. The modulation scheme, QASK-16, selected for the 6402, is an excellent compromise between bandwidth efficiency and performance without leading to prohibitive manufacturing costs. Through novel design, the 6402 achieves 2 bits/Hz spectrum efficiency using reliable, low-cost components.

References

1) R.E. Pidgeon, "Performance of a 440 MHz; 54 Channel, Cable Television Distribution System," 30th Annual NCTA Convention Technical Papers, 1981.

2) C.M. Thomas, "Amplitude-Phase-Keying with M-ary Alphabets: A Technique for Bandwidth Reductions," Proc. Int. Telemetering Conf., pp. 289, October 1972.

3) J.G. Smith, "On the Feasibility of Efficient Multi-Amplitude Communication," National Telecommunications Conf. Rec., pp 20E-I, 1972.

4) G.J. Foschini, R.D. Gitlin, and S.B. Weinstein, "On the Selection of a Two-Dimensional Constellation in the Presence of Jitter and Gaussian Noise," Bell Systems Tech. J., Vol. 54, pp. 927-965, July-Aug. 1973.

5) M.K. Simon and J.G. Smith, "Hexagonal Multiple-Phase-and-Amplitude-Shift-Keyed Signal Sets," IEEE Trans. Comm., Vol. COM-21, pp. 1108-1115.

6) G.J. Foschini, R.D. Gitlin, and S.B. Weinstein, "Optimization of Two-Dimensional Signal Constellation in the presence of Guassian Noise," IEEE Tran. Comm., Vol. Com 22, pp. 28-38, January 1974.

7) J.G. Smith, "A Review of Multiple Amplitude-Phase Digital Signals," IFTJ., Vol 1, pp. 6-14, Jan.-Feb. 1974.

8) C.M. Thomas, M.Y. Weidner, and S.H. Durrani, "Digital Amplitude-Phase-Keying with M-ary Alphabet," IEEE Trans. Comm., Vol. COM-22, pp. 168-180, Feb. 1974.

9) M. Washio and T. Itaya, "Experimental Study on a New 1.6 Gbps 16-Level APSK Modem," National Tel. Conf. Proceedings, pp. 05:6-1 - 05:6-6, 1977.

10) A.P. Clark and M.J. Fairfield, "Detection Processes for a 9600 bit/s Modem," The Radio and Electronic Engineer, Vol. 51, No. 9, pp. 455-465, Sept. 1981.

11) W.J. Weber, III, P.H. Stanton, J.T. Surnida, "A Bandwidth Compressive Modulation System Using Multiple-Amplitude Minimum Shift Keying MAMSK)," IEEE Trans. on Comm., Vol. COM-26, pp. 543-551, May 1978.

12) W.J. Weber, III, "Differential Encoding for Multiple Amplitude and Phase Shift Keying Systems," IEEE Trans. on Comm., Vol. COM-26, pp. 385-391, March 1978.

13) M.K. Simon and J.G. Smith, "Carrier Synchronization and Detection of QASK Signals," IEEE Trans. on Comm., Vol. COM-23, pp. 606-620, June 1975.

14) M.K. Simon, "Data-Derived Symbol Synchronization of MASK and QASK Signals," IEEE Trans. on Comm., Vol. COM-23, pp. 606-620, June 1975.

15) M.K. Simon, "Data-Derived Symbol Synchronization of MASK and QASK Signals," Jet Propulsion Laboratory, Pasadena, Cal., JPL Tech. Memo 33-720, Dec. 15, 1974.

16) W.J. Weber, III, "Decision-directed Automatic Gain Control for MAPSK Systems," IEEE Trans. on Comm., Vol. COM-23, pp. 510-517, May 1975.

17) C.C.I.T.T. Recommendation V.35, App. 1.

Reliability Analysis

S.N. Baker

Introduction

The Quality Department of the Satellite Communications Division has established a program of reliability/availability prediction and measurement for all standard products. Mean-time-between-failure (MTBF) calculations are done at the component and unit assembly level. These MTBF predictions are then used to predict system availability. Computer programs have been written that process the bulk of these predictions.

Computer software is also on-line that provides actual failure rates for each model. The feedback loop is completed by concentrating engineering resources on differences that exist between predictions and actual field failure rates.

The purpose of this paper is twofold: 1) to present the methods used in determining reliability predictions and measurements, and 2) present the effect on availability of several system configurations.

MTBF Calculations

Predictions of MTBF are obtained using the parts count method of MIL-HDBK-217D, Section 5.2. The failure rate of an electronic assembly is the summation of the failure rates of each component that make up that assembly (e.g., resistors, integrated circuits, etc.) and the failure rates of the connections (e.g., solder joints). These generic failure rates depend on the use environment. The "Ground, Fixed" environment is used for earth station equipment. This environment assumes conditions less than ideal, such as installation in permanent racks with adequate cooling air and possible installation in unheated buildings.

The base failure rate of each component must be multiplied by a quality factory. The quality factor is a function of the purchased quality level or designator (e.g., RL and RLR for resistors) and in-house quality procedures (e.g., burn-in and testing). The general expression for equipment failure rate is:

$$\lambda_{EQUIP} = \sum_{i=1}^{i=n} N_i \left(\lambda_G \pi_Q\right)_i$$

where:

N_i = Quantity of i^{th} generic part

λ_G = Generic failure rate for the i^{th} generic part

π_Q = Quality factor for the i^{th} generic part

n = Number of different generic part categories

A computer program has been written in BASIC that contains the military handbook data. An MTBF prediction is obtained by categorizing each component on the parts list, tallying up the generic parts, and then entering them in the program. All components are included whether critical or not to unit failure. An example printout is shown in Figure 1. The failure rates are failures per one million hours of operation. In addition to the predicted field failure rate (last column), the program also computes the purchased and ideal failure rates. The purchased column contains the failure rates of components as purchased, assuming no in-house screening, burn-in, or testing. This column is the summation of in-house and field failure rates. To compute the ideal failure rate column, the software automatically substitutes components to arrive at the theoretically highest MTBF for a particular design and chooses the highest quality levels for each part. For example, ceramic ICs are substituted for plastic. The only way to improve the ideal failure rate is to reduce the parts count. This approach gives the designer a theoretical MTBF ceiling.

Satellite Communications Division

Reliability Prediction-Ground Fixed Environment

VIDEO EXCITER

CODE #	DESCRIPTION	QTY	PREDICTED FAILURE RATES PURCH	IDEAL	FIELD
1	RESISTOR-COMPOSITION,RCR,MIL-R-39008	471	0.30	0.02	0.10
6	RESISTOR-FILM,RN,MIL-R-10509	120	2.16	0.01	0.22
23	POT-COMPOSITION,RV,MIL-R-94	32	4.48	0.06	0.45
34	CAPACITOR-PLASTIC,CRH,MIL-C-83421	15	0.04	0.00	0.01
36	CAPACITOR-MICA,CM,MIL-C-5	27	0.30	0.00	0.03
41	CAPACITOR-CERAMIC,CK,MIL-C-11015	357	6.43	0.06	0.64
43	CAPACITOR-TANTALUM,SOL,CSR,MIL-C-39003	68	0.29	0.03	0.10
50	CAPACITOR-VAR,AIR,TRIMMER,CT,MIL-C-92	3	5.70	0.11	0.57
53	TRANSFORMER-AUDIO POWER OR PULSE	1	0.11	0.04	0.04
56	COIL-RF,FIXED	88	2.90	0.97	0.97
57	COIL-RF,VARIABLE	8	0.55	0.18	0.18
65	SWITCH-TOGGLE OR PUSH BUTTON	6	0.35	0.02	0.05
68	CONNECTOR-CIRCULAR,RACK & PANEL	24	1.22	0.41	0.41
69	CONNECTOR-PWB	1	0.07	0.02	0.02
70	CONNECTOR-COAXIAL	2	0.10	0.03	0.03
71	PWB-2 SIDED	14	0.41	0.04	0.04
73	TRANSISTOR-SI,NPN	61	9.76	0.10	0.20
74	TRANSISTOR-SI,PNP	8	1.92	0.02	0.04
77	TRANSISTOR-SI,FET	6	12.60	0.13	0.25
79	DIODE-SI,GENERAL PURPOSE	37	1.15	0.01	0.02
81	DIODE-ZENER OR AVALANCHE	12	1.44	0.01	0.03
83	DIODE-SI,MICROWAVE DET.	3	5.46	1.17	2.34
87	DIODE-VARACTOR,STEP,RECOVERY,OR TUNNEL	9	90.00	0.90	1.80
88	LED	30	9.90	0.10	0.20
91	DIODE - PIN	3	39.00	0.39	0.78
94	CONNECTION - SOLDERWAVE	4241	2.58	2.58	2.58
95	CONNECTION - HAND SOLDER	236	1.29	1.29	1.29
101	DIGITAL BIPOLAR, HERMETIC, 1-20 GATES	1	0.21	0.01	0.02
113	DIGITAL BIPOLAR, NON-HERMETIC, 1-20 GATES	6	4.20	0.05	0.12
114	DIGITAL BIPOLAR, NON-HERMETIC, 21-50 GATES	8	5.60	0.08	0.16
132	DIGITAL MOS, HERMETIC, 3001-5000 GATES	1	2.60	0.10	0.80
137	DIGITAL MOS, NON-HERMETIC, 1-20 GATES	18	11.34	0.14	1.30
138	DIGITAL MOS, NON-HERMETIC, 21-50 GATES	6	4.83	0.06	0.55
139	DIGITAL MOS, NON-HERMETIC, 51-100 GATES	2	2.17	0.03	0.25
176	MOS ROM, PROM, EPROM, NON-HERM, 1-320 BITS	1	1.05	0.01	0.12
183	MOS ROM, PROM, EPROM, NON-HERM, 17001-38000 BIT	1	15.40	0.10	1.76
192	BIPOLAR RAM, NON-HERM, 1-320 BITS	1	2.10	0.03	0.06
214	MOS RAM, NON-HERM, 11001-17000 BITS	1	42.00	0.22	4.80
217	LINEAR, 1-32 TRANSISTORS, HERMETIC	10	4.03	0.16	1.24
220	LINEAR, 1-32 TRANSISTORS, NON-HERMETIC	16	27.44	0.25	3.14
240	POWER SUPPLY	3	36.00	8.40	12.00
241	CRYSTAL	4	2.40	0.56	0.80
243	MIXER	3	24.00	4.20	6.00
244	FREQUENCY SOURCE	1	40.00	9.20	13.20
245	VACUUM FLOURESCENT DISPLAY	1	1.50	0.35	0.50
246	BANDPASS FILTER	2	60.00	14.00	20.00
247	SWITCH - DIP HEXIDECIMAL	8	12.81	0.03	0.20
248	FRONT PANEL	1	1.16	0.06	0.17
	TOTAL PREDICTED FAILURE RATES		501.34	46.74	80.57
	TOTAL PREDICTED MTBFs (hours)		1995	21396	12412

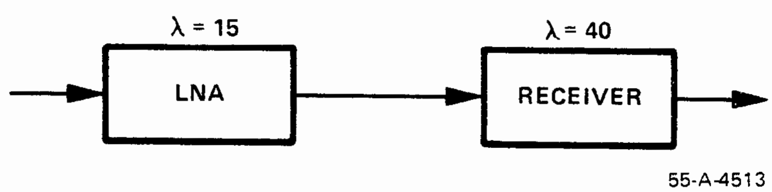
Figure 1. MTBF Prediction

Availability Calculations

After the field failure rates have been calcualted for each unit, this information can be used to determine system availability. Availability is defined as the probability of finding a system or subsystem operating at any arbitrary future time. The key inputs to system availability calculations are:

- System configuration
- Mean-time-to-repair
- Mean-time-to-change
- Quantity of shelf spares

The first step is to define a reliability model that corresponds to the actual system. The simplest model is the single thread which is represented as follows:



This model implies that both pieces of equipment must be working for the system to be considered operational.[1]

First, the availability for each block is calculated using the equations:

$$\text{MTBF} = \frac{10^6}{\lambda}, \quad A = \frac{\text{MTBF}}{\text{MTBF+MTTR+MTTC}}$$

For an unmanned station it is assumed that the time to diagnose and change a failed unit is 24 hours (MTTC) and the time to have the unit repaired at the factory is two weeks or 336 hours (MTTR). The MTBFs and availabilities for each block in the system are as follows:

[1]A redundant LNA power supply is assumed. Because its effect on system availability is neglibile, it is ignored.

LNA: $\quad MTBF_L = \frac{10^6}{15} = 66,667 \text{ hrs.} \qquad A_L = \frac{66,667}{66,667+336+24} = 0.9946$

RECEIVER: $\quad MTBF_R = \frac{10^6}{40} = 25,000 \text{ hrs.} \qquad A_R = \frac{25,000}{25,000+336+24} = 0.9858$

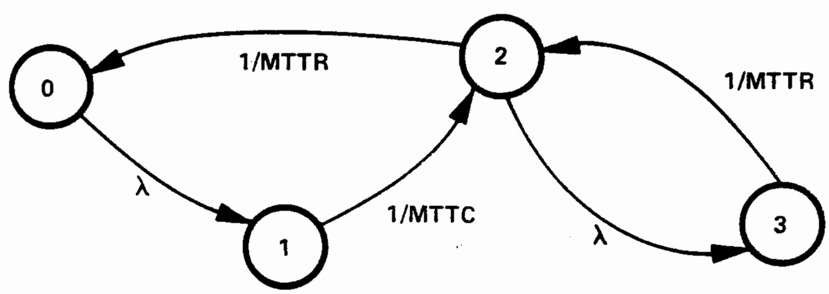
Single thread system availability is obtained by multiplying the availabilities for each block together:

$$A_{SYSTEM} = A_S = \prod_{i=1}^{N} A_i$$

Thus for this example:

$$A_S = (A_L)(A_R) = (0.9946)(0.9858) = 0.9805$$

What if a shelf spare or complete set of replacement modules is maintained at the site for the LNA and receiver? Here, the system will be down during the time to diagnose and replace the failed unit and also in the event the shelf spare fails during the factory repair cycle of the original unit. This can be represented by the following state diagram:



55-A-4511

STATE		CONDITION		SYSTEM STATUS
0	-	None failed	-	Up
1	-	On-line unit fails	-	Down
2	-	Shelf spare installed	-	Up
3	-	Shelf spare fails	-	Down

The availability of each block is:

$$A = \frac{MTBF + MTTR}{MTBF + MTTC + MTTR + \frac{(MTTR)^2}{MTBF}}$$

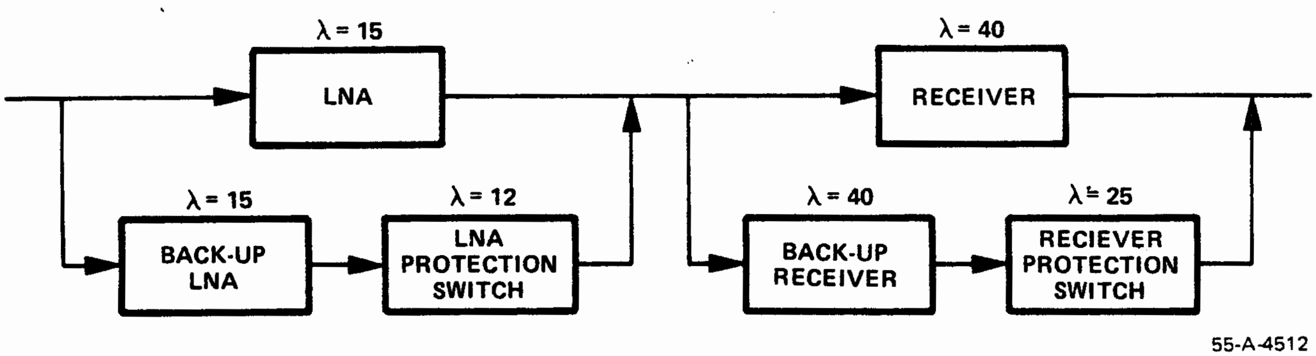
Thus, for this single thread system with a complete set of shelf spares for each unit, the availabilities are:

$$A_L = \frac{66,667 + 336}{66,667+24+336+\frac{(336)^2}{66,667}} = 0.9996$$

$$A_R = \frac{25,000 + 336}{25,000+24+336+\frac{(336)^2}{25,000}} = 0.9989$$

$$A_{SYSTEM} = (0.9996)\ (0.9989) = 0.9985$$

The last basic reliability model is a system that is configured with backup units that are automatically brought on-line by a protection switch. In this instance the MTTC can be considered negligible. However, the failure of the protection switch must be included. Scientific-Atlanta protection switches are designed such that if the protection switch alone fails, the system remains operational. Hence, the failure rate of the switch is a factor only when it is called upon to switch in a backup unit. This can be represented by the following reliability model:



At the receiver the signal has two alternate paths. For the system to stay up, the receiver on-line must be up or the backup receiver and the protection switch must be up. To find the availability of the redundant pair of receivers, a Boolean Algebra truth table approach is used. In the truth table (Figure 2), a 1 or 0 entry in each column indicates success or failure respectively of each unit. All possible combinations of all units working and failing are taken into account. The probability that a unit is failed is one minus its availability. To obtain the probability for a row in the table, the probabilities of each column are multiplied. The system availability is then the sum of all probabilities where the system is up.

Figure 2. Truth Table Method for Redundant Systems

$A_R = 0.986738$ $A_P = 0.991670$

On-line Receiver	Back-up Receiver	Protection Switch	Receiver Subsystem	Probability
0	0	0	0	-
0	0	1	0	-
0	1	0	0	-
0	1	1	1	$(1-A_R)(A_R)(A_P) = 0.012977$
1	0	0	1	$(A_R)(1-A_R)(1-A_P) = 0.000109$
1	0	1	1	$(A_R)(1-A_R)(A_P) = 0.12977$
1	1	0	1	$(A_R)(A_R)(1-A_P) = 0.008111$
1	1	1	1	$(A_R)(A_R)(A_P) = 0.965542$

$$A_{REDUNDANT} = \sum A_i = 0.999715$$

Correspondingly for a pair of redundant LNAs:

$A_L = 0.9950$ $A_P = 0.9960$ $A_{REDUNDANT} = 0.999955$

Thus, for a system with redundant hot standbys, the system availability is:

$$A_{SYSTEM} = (0.999955)(0.999715)$$

$$= 0.999670$$

When more units than a simple 1:1 are configured redundantly, calculating the availability becomes more complex. For example, the equation for the availability of six units on-line operating in parallel with two hot spares on standby via an automatic protection switch is:

$$A = A_i{}^8 + 8(A_i) + 28(A_i)^6\ (1-A_i)^2$$

where:

A_i = the availability of an individual unit

The coefficients of this equation are the first portion of a row from Pascal's triangle. A computer program has been written in BASIC that computes any R:N configuration. An example printout is shown in Figure 3.

Probabilities and Outage Time

The probability (P_S) of a unit operating without failure for a specific period of time (t) can be found using:

$$P_S = e^{-\lambda t}$$

For a period of 8,736 hours (1 year), the probability of successful operation of a unit with an MTBF of 36,400 hours (4.2 years) is:

$$\lambda = \frac{10^6}{36,400} = 27.5 \text{ failures}/10^6 \text{ hours}$$

$$P_S = e^{-(27.5)(8736)/10^6} = 0.79$$

The probability of failure (P_F) for a time period of 8,736 hours is the probability of success subtracted from 1 which in this case is 21%:

$$P_F = 1 - P_S = 1 - 0.79 = 0.21$$

The predicted outage time per year is found by multiplying one minus the availability by the number of hours per year:

$$\text{OUTAGE} = (1-A)(8736)$$

Satellite Communications Division
Subsystem Availability Calculations

SCPC-Channel Synthesizers

ON-LINE= 48 MTBF= 66666.7 HRS MTTR= 720 HRS

# SPARES	AVAILABILITY
0	0.59712988
1	0.90337457
2	0.98354111
3	0.99781691
4	0.99976169
5	0.99997779
6	0.99999819
7	0.99999987
8	0.99999999
9	1.00000000
10	1.00000000

Figure 3. R:N Availability Analysis

For a unit with an availability of 0.9990, the outage is:

$$\text{OUTAGE} = (1 - 0.9990)(8736) = 8.74 \text{ hours}$$

How Redundancy Affects Availability

Three basic reliability models have been discussed--single thread, single thread with shelf spares, and on-line redundancy with automatic protection switches. The models and failure rates correspond to typical video downlink system configurations for unmanned equipment. Table 1 summarizes relevant data for these systems along with the manned station case.

Table 1. Summary of Basic Receive System Configurations

SYSTEM	UNMANNED, MTTC=24 HOURS			MANNED, MTTC=4 HRS (LNA) MTTC=1 HR (RECEIVER)	
	COST FACTOR*	AVAIL-ABILITY	PREDICTED OUTAGE TIME PER YEAR	AVAIL-ABILITY	PREDICTED OUTAGE TIME PER YEAR
Single Thread	1.0	0.9805	170.4 hrs	0.9817	159.9 hrs
Single Thread with Shelf Spares	2.0	0.9985	13.1 hrs	0.99970	2.6 hrs
Redundant	4.1	0.99967	2.9 hrs	0.99967	2.9 hrs

*GCE equipment only.

Table 1. Summary of Basic Receive System Configurations

Several conclusions can be drawn from this data table:

1. An unmanned station with a protection switch achieves the same reliability level as attended equipment with shelf spares.

2. For a single thread system with no shelf spares, only a 6.2% improvement is achieved by manning the station.

3. Significant improvement in system reliability is attained by:

   a. Providing shelf spares

   b. Adding protection switches

   c. Manning the equipment (when spares are provided)

Further improvement in availabilities at the systems level could be achieved by:

1. Increasing the quantities of standby and shelf spares

2. Reducing MTTR through special maintenance contracts

At the hardware design level, MTBF could be improved by:

1. Reducing parts count

2. Specifying components with higher reliability levels

3. Increased screening of purchased parts

4. Increased unit burn-in

The cost effectiveness of each change that is contemplated should be evaluated from the system availability viewpoint.

Reliability Measurement

An MTBF or availability prediction is just that--a prediction. Predictions are useful during the design phase and when evaluating the cost effectiveness of various system configurations. An invaluable feedback tool to the prediction process is the measurement of actual failure rates. The report that provides this information is the Product Performance Summary (PPS). Figure 4 shows a sample report for one model. Cost figures are omitted due to their proprietary nature.

Every piece of electronic equipment that is returned to the factory is tracked with a service record. The service record provides the failure information that goes into PPS. Shipping logs provide the quantity shipped information.

The unit of measure or index that PPS uses for failure rate is failures per unit month (F/UM). The reciprocal of F/UM is the average months between failures. The index is calculated for a time interval of one year using the data currently available. As an example, if the latest month is October and 10 units were shipped in January, then this product has experienced 100 unit months of operation. If one failure occurs in April and another in September, then the F/UM would be:

$$\frac{2 \text{ failures}}{100 \text{ unit months}} = 0.020 \text{ F/UM}$$

PRODUCT PERFORMANCE SUMMARY

MODEL 00

	7/83	6/83	5/83	4/83	3/83	2/83	1/83	12/82	11/82	10/82	9/82	8/82
SHIPPED	8	20	31	22	16	19	8	17	11	7	39	58
COST WTY	$	$	$	$	$	$	$	$	$	$	$	$
S & R	$	$	$	$	$	$	$	$	$	$	$	$
FAILURES	5	8	3	8	7	8	5	4	9	6	19	6
F / UM	0.046	0.048	0.043	0.029	0.029	0.028	0.028	0.027	0.000	0.000	0.000	0.000
DISCREPs	9	9	5	20	14	11	9	5	11	9	25	8

	7/83	6/83	5/83		7/83	6/83	5/83		7/83	6/83	5/83
COMP	22%	67%	20%	WRKSHP	22%	0%	20%	NOPROB	22%	11%	20%
DESIGN	22%	0%	40%	SHIP	11%	11%	0%				
ADJUST	0%	0%	0%	MGMT	0%	0%	0%	OTHER	0%	11%	0%

SUMMARY	SHIPPED	WTY COST	INT COST	TOT COST	COST/F	FAILURES	DISCRP	AVE AGE	D/U
	256		$	$	$	88	135	7.5	0.53

Figure 4. Reliability Measurement

The failure rate (λ) can easily be obtained by multiplying F/UM by the constant 1374. For this example:

$$(0.020 \text{ F/UM})(1{,}374) = 27.5 \text{ failures}/10^6 \text{ hours}$$

Or the MTBF can be obtained by dividing the constant 728 by F/UM:

$$\frac{728}{0.020 \text{ F/UM}} = 36{,}400 \text{ hours between failures}$$

It is important to note that calculating λ and MTBF from F/UM provides a 12 month "snapshot" index that could vary from the actual long-term failure rate.

Terms for Glossary

Availabilty (A)	- Probability of finding a system or subsystem operating at any arbitrary future time.
Failure Rate (λ)	- Average number of failures per one million hours of operation.
Mean-Time-Between-Failure (MTBF)	- Average number of hours between failures (the hours of operation divided by the number of failures).
Mean-Time-To-Change (MTTC)	- Average time in hours to diagnose a failure and then automatically switch in a hot spare, manually install a shelf spare, or manually perform the function of the failed unit.
Mean-Time-To-Repair (MTTR)	- Average time in hours to repair a failed unit of equipment or obtain a replacement from the manufacturer or a spare depot.

References

M. Sasaki, S. Kaburaki, S. Yanagi, "System Availability and Optimum Spare Units," IEEE Transaction on Reliability, V. R-26, No. 3, August 1977.

MIL-HDBK-217D, Reliability Prediction of Electronic Equipment, January 15, 1982.

MIL-HDBK-217C, Reliability Prediction of Electronic Equipment, "Appendix A - System Reliability Modeling," April 9, 1979.

P. O'Connor, Practical Reliability Engineering, Heyden, 1981.

S. Gelfond, M. Gerber, A. Kirillov, N. Konstantinov, A. Kushnirenko, Sequences, Combinations, Limits, M.I.T. Press, 1969.

C. Prier, "Pseudo Reliability Measure from Field Data," internal paper, 1980.

Earth Station Antennas, RF Considerations

Rick Barker and John Friesz

Introduction

The earth station antenna is a vital link between the satellite and earth station electronic equipment. It must provide reliable high receiving gain under sometimes severe environmental conditions. In addition, its far-field pattern must have sufficiently low wide-angle sidelobes to suppress interfering signals within the very congested operating bands. The intent of this paper is to outline the antenna requirements and consider the reflector-type antenna which is most often used to meet these requirements.

Earth Station Antenna Requirements

Earth station antenna requirements are dependent on many factors, such as channel capacity, receiving equipment, signal modulation, reliability requirements, geographical location relative to the satellite, site interference profiles and site environmental conditions. This discussion will consider the antenna requirements for typical small earth terminals.

As mentioned previously, antenna requirements are to some extent site related; because of this, the following table may not be applicable for all sites. It does, however, outline typical requirements for an average continental U.S. site. These antenna requirements are given as flexible guidelines which may vary with total system performance requirements.

The antenna characteristics given in Table 1 are applicable primarily to reflector-type antennas which are from 5- to 11-meters in diameter. The reflector antenna is the most widely used and accepted type of antenna because it satisfies system antenna requirements and is desirable from the viewpoints of mechanical simplicity, cost and availability.

Table 1. Earth Terminal Antenna Requirements

Characteristic	Specification
Frequency of Operation	3700 to 4200 MHz Receive 5925 to 6425 MHz Transmit
Beamwidth	Consistent with gain requirements
First Sidelobe	-14 dB Maximum
VSWR	1.30:1 Receive band 1.25:1 Transmit band

Table 1. Earth Terminal Antenna Requirements (continued)

Characteristic	Specification
Polarization Discrimination	-30 dB (relative to peak of co-polarized beam)
Maximum Transmit Power	5 kW average 10 kW peak
Feed System Pressurization	0.5 psig Nominal
Polarziation Rotation	±90°
Isolation (Transmit-to-Receive)	35 dB
Pattern Envelope	Compliant with FCC Regulation 25.209

Electrical and Electromagnetic Requirements

Antenna design requirements can be grouped into several major categories--electrical or RF, control systems, pointing and tracking accuracy, and miscellaneous requirements such as radiation hazard protection and other environmental safety factors. The electrical or RF parameters are of primary concern in this section. An excellent overview of the electrical and mechanical characteristics of antennas for earth stations is given in EIA Standard RS-411.

Gain and Directivity

The power gain (g) of an antenna in a specified direction is 4π times the ratio of the power radiated per unit solid angle in that direction to the net power accepted by the antenna from its generator. This quantity is an inherent property of the antenna and does not involve system losses arising from a mismatch of impedance or polarization.

The directivity (D) of an antenna in a specified direction is 4π times the ratio of the power radiated per unit solid angle in that direction to the total power radiated by the antenna. This term differs from power gain because it does not include antenna dissipation losses.

Gain and directivity are directly related by:

$$g = \eta D$$

where η is the ohmic efficiency, which is less than unity; that is,

$$\eta = P_T/P_O$$

where P_O is the power accepted by the antenna from the RF source, and P_T is the power radiated into space.

The maximum directivity of an aperture is given by:

$$D_m = \frac{4\pi A}{\lambda^2}$$

where:

A is the maximum projected area of the aperture

λ is the wavelength of operation.

For a circular aperture, the maximum directivity is:

$$D_m = \left(\frac{\pi d}{\lambda}\right)^2$$

where:

d = diameter of the aperture

λ = wavelength of operation.

The aperture diameter and wavelength are expressed in the same units.

For a non-uniform aperture distribution, the directivity is given by:

$$D = k_1 \left(\frac{\pi d}{\lambda}\right)^2$$

where:

k_1 is a function of the distribution and is less than unity. For practical reflector-type antennas, k_1 is typically between 0.5 and 0.9. The gain of the antenna is then given by:

$$g = \eta k_1 \left(\frac{\pi d}{\lambda}\right)^2$$

The quantity ηk_1 will be referred to as "aperture efficiency" herein. The aperture efficiency, ηk_1, of typical reflector type antennas varies from 0.5 to 0.65 for focal point-fed, axisymmetric antennas and from 0.60 to 0.90 for dual-reflector and offset-fed antennas. For a discussion of the predominant types of earth station antennas, see "Introduction to Earth Station Antennas" in this digest.

Input Impedance and Voltage Standing Wave Ratio

The input impedance of an antenna affects interaction between the antenna and its associated components. Antenna impedance can be an important factor in the consideration of power transfer, noise, and stability of active circuit components such as the low-noise amplifier.

For most aplications a conjugate match is desired between the antenna and the transmission line or device to which it is connected. When a conjugate match does not exist, some of the available power is lost by reflection.

Most antennas are connected to the electronic networks or circuits by a transmission line. In practice, it is usually advantageous to match the antenna to the transmission line at the antenna terminals. This minimizes the line losses and the voltage peaks on the line, thus facilitating high power transmission. Imperfect matching of the antenna to the transmission line creates a reflected wave in the line; the reflected power relative to the incident power is:

$$\frac{P_{refl}}{P_{inc}} = \left| \frac{Z_{ant} - Z_o}{Z_{ant} + Z_o} \right|^2$$

where:

Z_0 is the characteristic impedance of the transmission line

Z_{ant} is the impedance of the antenna.

This ratio is related to the voltage reflection coefficient, Γ, and to the voltage standing wave ratio, VSWR, by the standard transmission line relationships:

$$\frac{P_{refl}}{P_{inc}} = |\Gamma|^2 = \left| \frac{VSWR - 1}{VSWR + 1} \right|^2$$

For most station antennas the VSWR is usually specified to be 1.3:1 or less. This corresponds to a reflection coefficient of 0.1304. The reflection coefficient is important since the mismatch loss of the antenna and the VSWR are determined by it. The mismatch loss is given by:

$$L_m = [-10 \log (1 - \Gamma^2)]$$

The reflection coefficient is also related to the return loss which is the logarithmic form of the VSWR expressed in (dB) and given by:

$$\text{Return loss (dB)} = (-20 \log \Gamma)$$

A typical plot of return loss versus frequency for an earth station feed is shown in Figure 1.

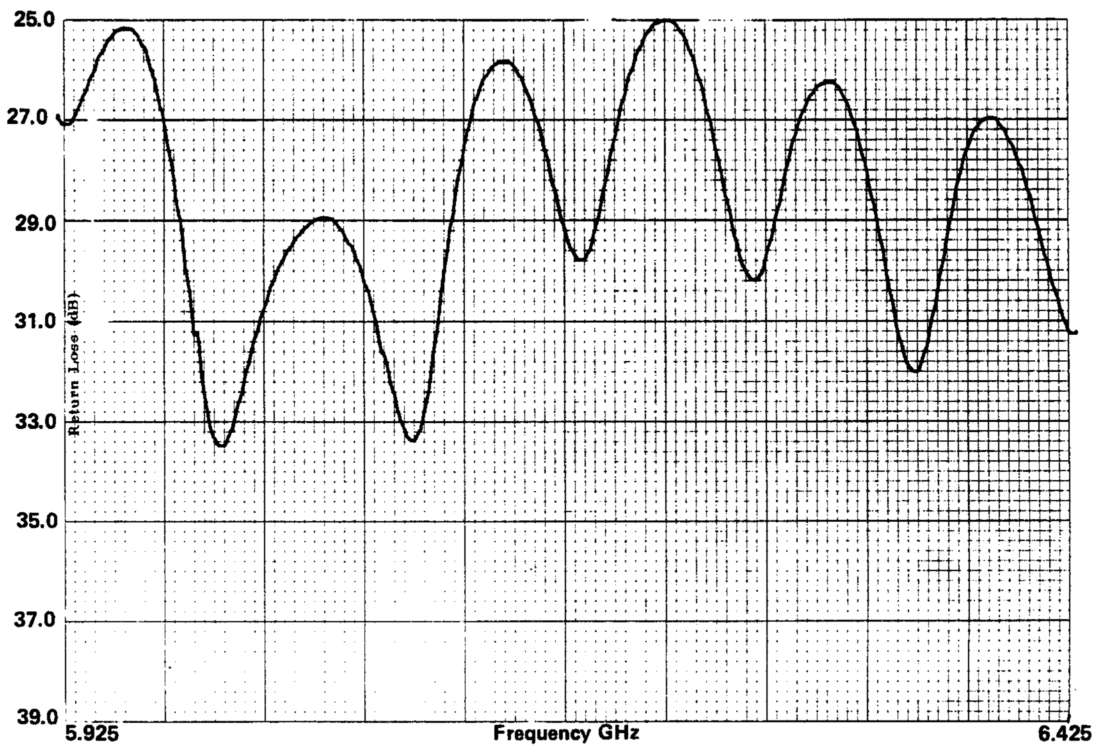


Figure 1. Feed Horn Return Loss vs. Frequency

To minimize degradation of the antenna noise temperature caused by waveguide losses between the antenna and low-noise amplifier (LNA), the LNA is usually connected directly to the antenna. It should present a nearly flat load across the frequency band, for example, the 3.7 to 4.2 GHz receive band.

Polarization

In general, the electric field vector of a uniform plane wave can be resolved into two orthogonal electric vectors. If the fields have the same relative phase, then they are said to be linearly polarized. If the fields are 90° out-of-phase, and have equal amplitudes, they are said to be circularly polarized. All waves are actually elliptically polarized which is the degenerated case of circular or linear polarization. U.S. domestic satellites are linearly polarized. INTELSAT uses circular polarization. A more thorough treatment of the implications of polarization and polarization isolation is given in the paper, "Reuse Earth Station Antennas" in this digest.

Surface Tolerance

Deviations of the surface of a reflector antenna from the desired surface can be grouped into two categories, systematic and random. Both types of deviations lead to degradation of the antenna's gain and sidelobe performance. By careful design and fabrication, the systematic deviations can usually be controlled such that they have a small effect on performance. In practice some amount of random deviations will always occur. The gain loss due to random deviations has been statistically analyzed by Ruse[1]. Ruse considered random phase errors caused by random distortions of the aperture surface. His analysis led to an equation for the loss of gain due to ε, the rms surface tolerance of the reflector.

$$\delta = e^{-(4\pi\varepsilon/\lambda)^2}$$

or, expressing in dB,

$$\delta = 4.92\ (\varepsilon f)^2$$

where:

ε = effective rms surface tolerance in inches

f = frequency in GHz.

The effective rms surface tolerance can be calculated from measured deviations normal to the reflector surface or axially from the surface. The correction factor vs. f/D is shown in Figure 2.

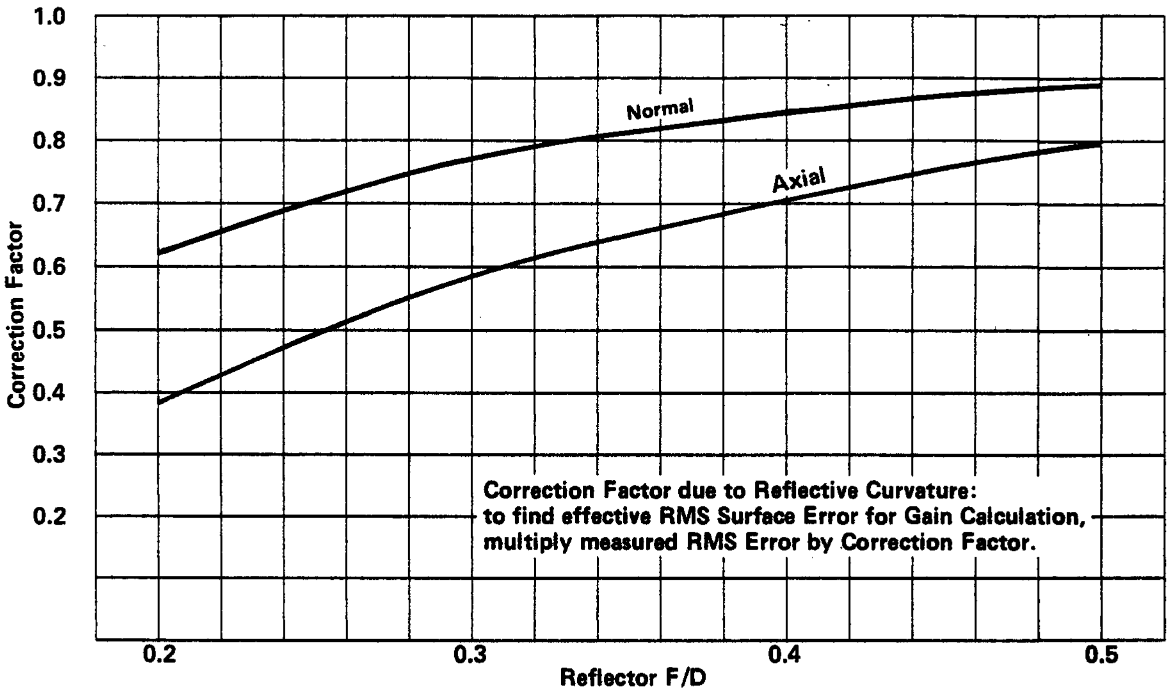


Figure 2.

Antenna Design Considerations

An earth station reflector antenna is shown graphically in Figure 3. The reflector shown is a paraboloid of revolution, which has an associated focal point. The distance from the reflector vertex to the focal point is called the focal length. The focal point position defines a subtended angle between the focal point and the edge of the reflector.

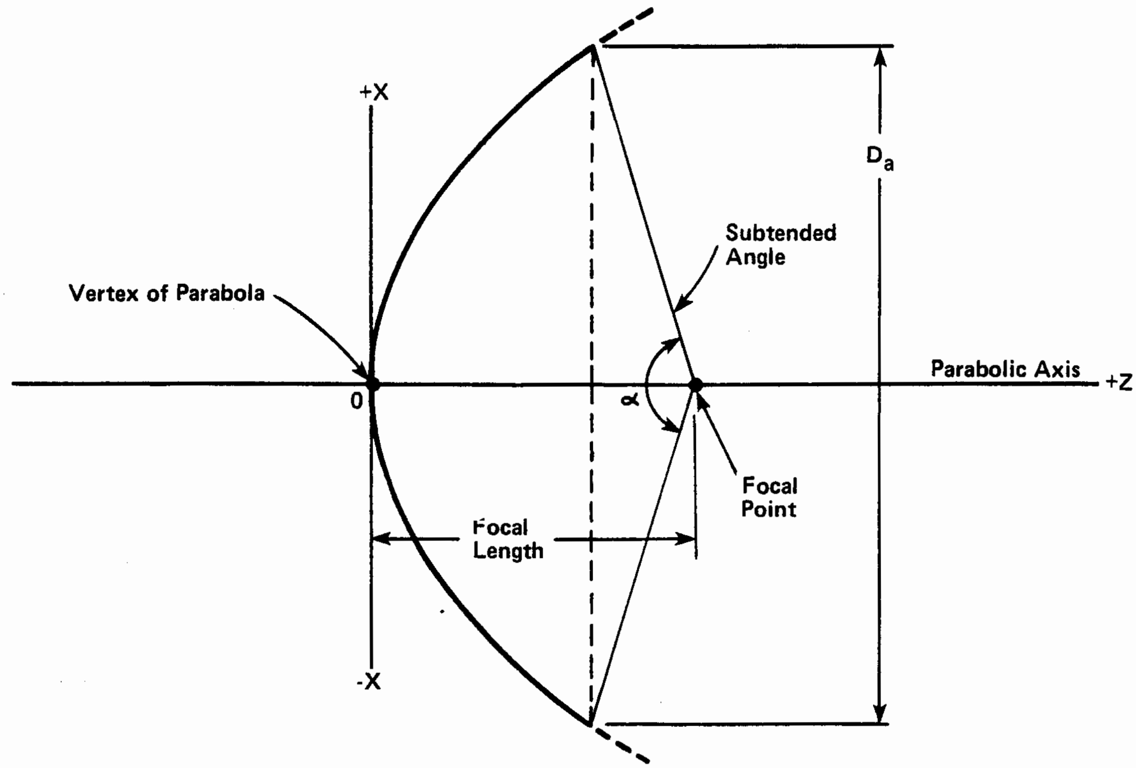


Figure 3. Geometry of Parabolic Reflector

The energy captured by the reflector is convergent on the focal point (focused) where it is received by a feed. The feed is connected to the receive system through some form of transmission line which is almost always a waveguide.

Another approach to focusing is to place a subreflector in the path of the convergent beam, thus focusing the incident energy onto a feed located near the vertex of the main reflector. The resulting antenna is classified as a dual-reflector type. When the subreflector is a hyperboloid, the antenna is classified as a Cassegrain type.

The basic requirements for efficient operation of any reflector antenna are proper amplitude illumination and constant phase across a relatively unobstructed reflector aperture. The problems associated with optimizing these factors are usually complex and require analytical analysis and antenna range testing. Some of the considerations in achieving an optimum design are presented in the following sections.

Prime Focus Antenna Geometry

The primary feed pattern of the prime focus configuration, as shown through ray tracking in Figure 4, is key to the antenna characteristics such as gain, beamwidth, sidelobes, wide-angle radiation and noise temperature. If an ideal feed pattern, as shown in Figure 5, could be achieved, it would be possible to achieve almost 100 percent aperture efficiency. This pattern would uniformly "illuminate"* a reflector out to the periphery then drop abruptly to zero, capturing all the RF energy in the field impinging on the reflector aperture. This also assumes ideal conditions for other contributing factors such as feed phase taper, reflector surface tolerance and aperture blockage.

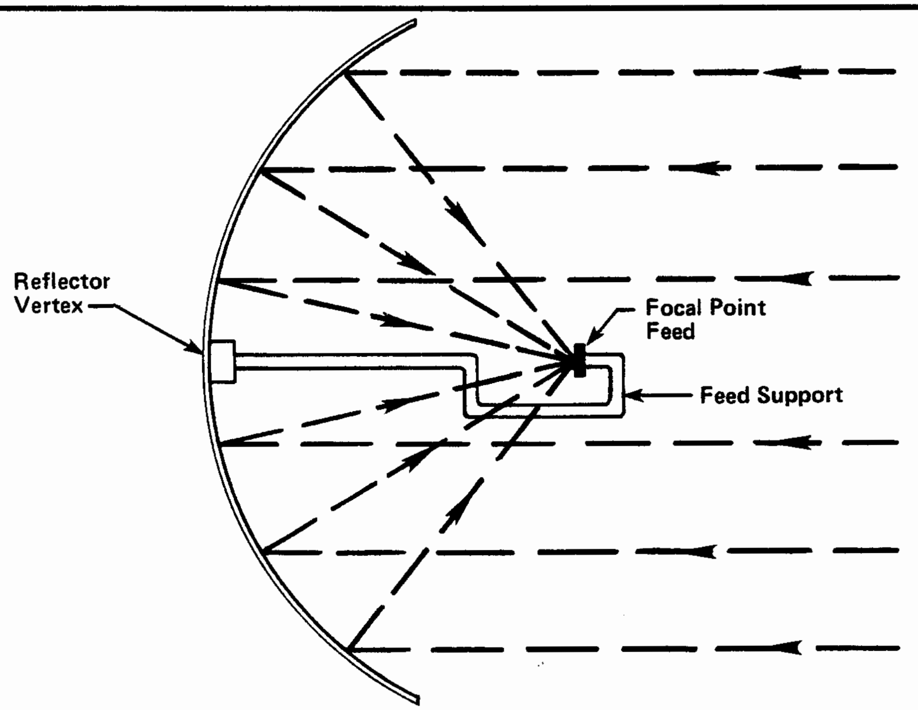


Figure 4. Geometry of Prime Focus Feed Reflector Configuration

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* The term "illuminate" is a transmitting term used because the antenna operates reciprocally on transmitting and receiving.

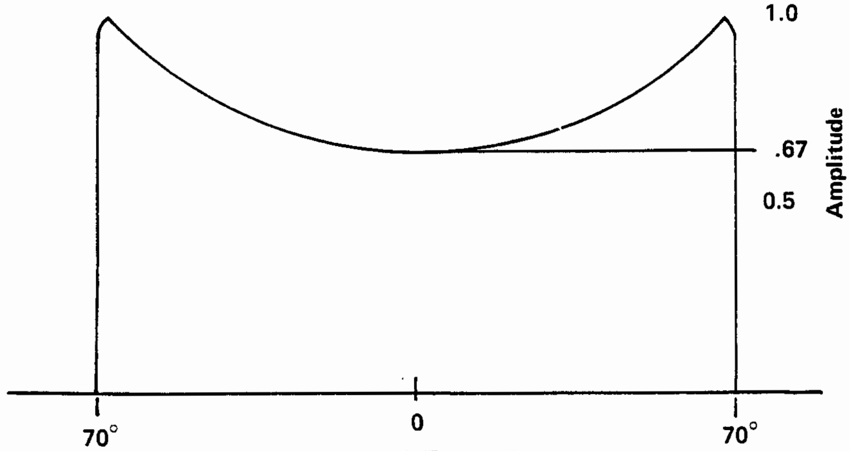


Figure 5. Ideal Feed Pattern

The feed pattern of Figure 5 is plotted in rectangular coordinates. The increase in pattern level toward the reflector edge is to compensate for the greater distance from the focal point to the reflector in comparison with that from the focal point to the vertex. The value at the center relative to that at the edge is given by:

$$a = \cos^2 (\theta/2)$$

where:

θ = half-angle of the paraboloid.

For the example of Figure 5, a = 0.67.

A feed pattern of the type shown in Figure 5 is not achievable. However, there are feeds which are capable of approaching the ideal pattern, as shown in Figure 6. This typical pattern is from a corrugated horn which is used as the feed for the Scientific-Atlanta Model 8001, a 10-meter earth station antenna. One can achieve illumination efficiencies of the reflector as high as 80 percent with a circular symmetric pattern of this type.

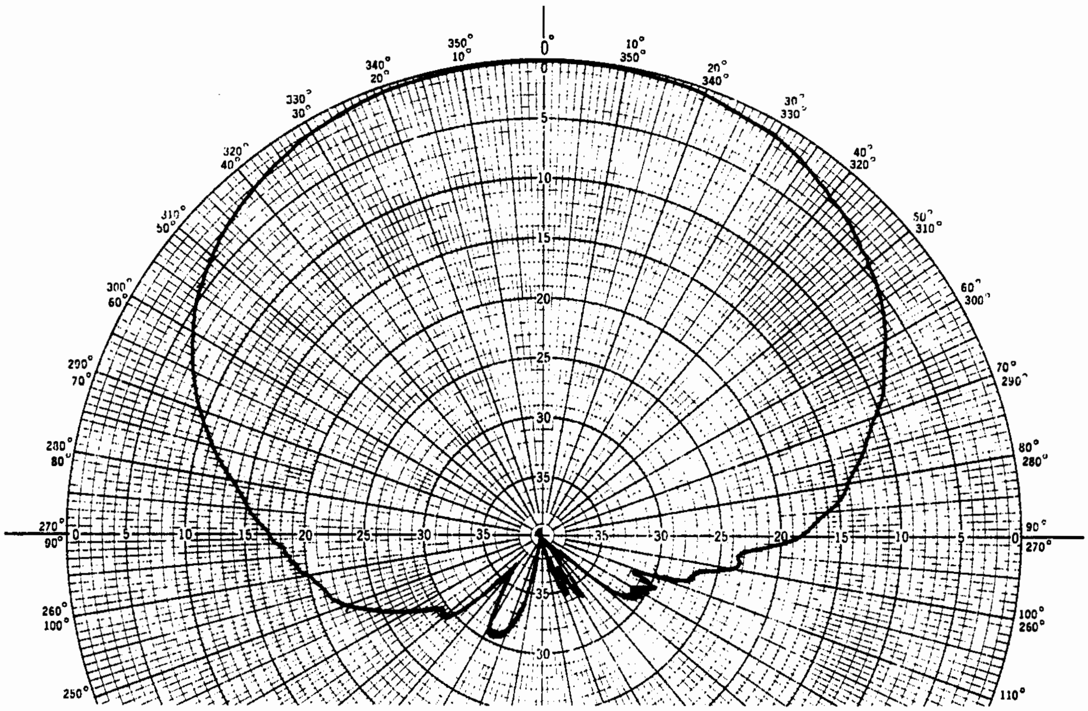


Figure 6. Typical Feed Pattern of Prime Focus Corrugated Horn

Figure 7 shows the two patterns of Figure 5 and 6 superimposed. The shaded areas are the two areas of concern which degrade antenna efficiency. The upper shaded area denotes the departure from a uniform taper of illumination out to the reflector half angle. This particular pattern, because of its 10-dB taper, has the effect of reducing the illumination efficiency from 100 percent to 80 percent. The lower shaded area is the feed radiation which is spilled over the edge of the reflector. The term used to describe its effect on total antenna efficiency is, appropriately enough, spillover efficiency. This typical feed pattern has a spillover efficiency of 90 percent. Also the phase taper and cross-polarization efficiencies are in the 97 percent range. The resultant of these efficiencies yields a feed illumination of approximately 74 percent. The gain of the antenna is reduced still further by random phase errors due to reflector surface tolerances and aperture area blocked by the feed and its support. These two effects reduce the total antenna efficiency to 64 percent at the feed aperture. This means that 64 percent of the power density in the incident field over the reflector aperture would be captured by the feed. This relatively high efficiency from a prime focus antenna can only be achieved by employing design techniques which include primary pattern characteristics similar to those of the pattern shown in Figure 6, minimizing the blockage area from the feed and its support structure to less than 2 percent of the total reflector area and maintaining the reflector rms surface tolerance to within $\lambda/30$.

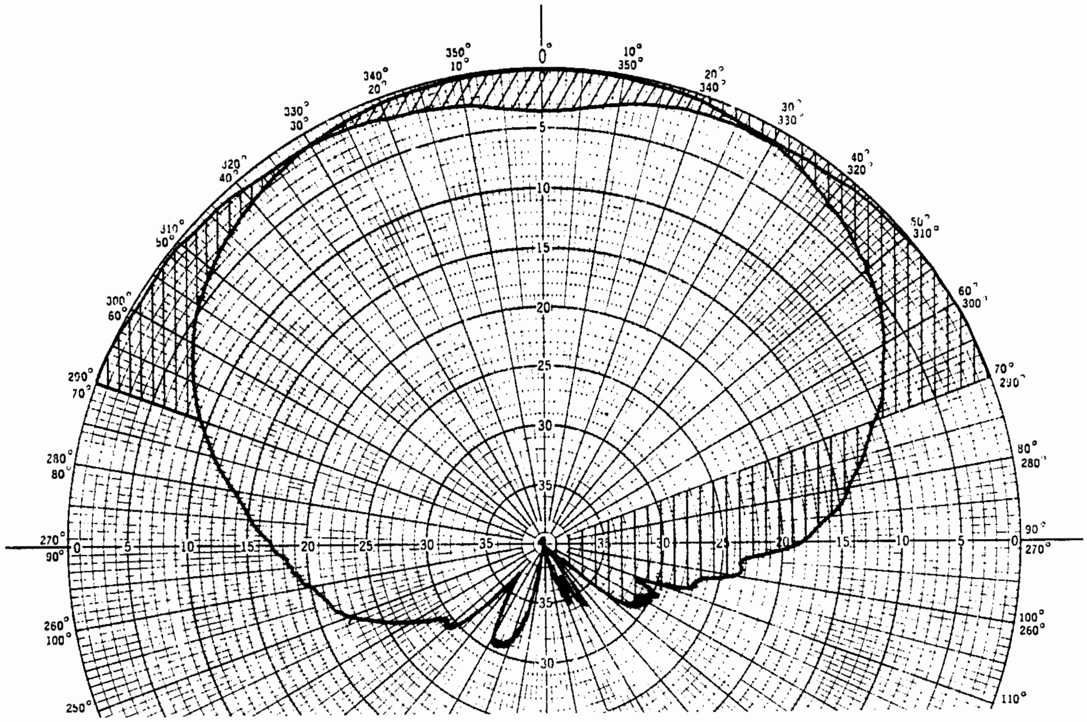


Figure 7. Actual Feed Pattern to Ideal Feed Pattern Comparison

The patterns shown in Figure 8 and 9 are typical patterns of a 10-meter prime focus antenna at 4 and 6 GHz. This antenna has been designed to minimize wide-angle sidelobes. The pattern envelope of FCC regulation 25.209 is plotted as a comparison. This antenna pattern is achieved by placing the feed at the focal point of the main reflector to limit the spillover and edge-diffraction contributions to one area of the pattern rather than two, as in the case of the dual-reflector antenna. The pattern area affected in the prime focus configuration is in the 100 to 110° region of the pattern. Also, the blockage area from the feed and its support is limited to less than 2 percent of total reflector area. This is important because any blocked area within a reflector has associated with it a broad, low-level pattern which affects the resultant far-field patterns. Obviously, the larger the area blocked, the greater the effect on the pattern. Keeping the blockage to a minimum reduces the scattering or reradiation of energy which would increase the wide-angle sidelobes.

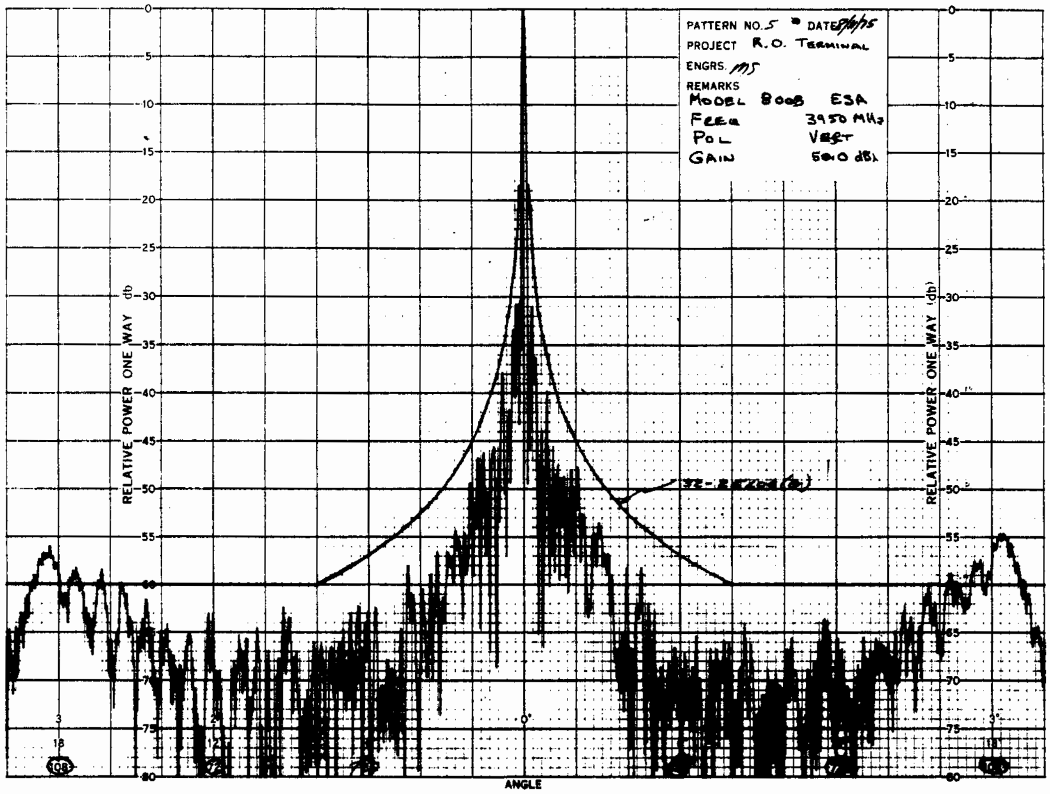


Figure 8. Typical 4 GHz Pattern of 10 Meter Prime Focus Antenna

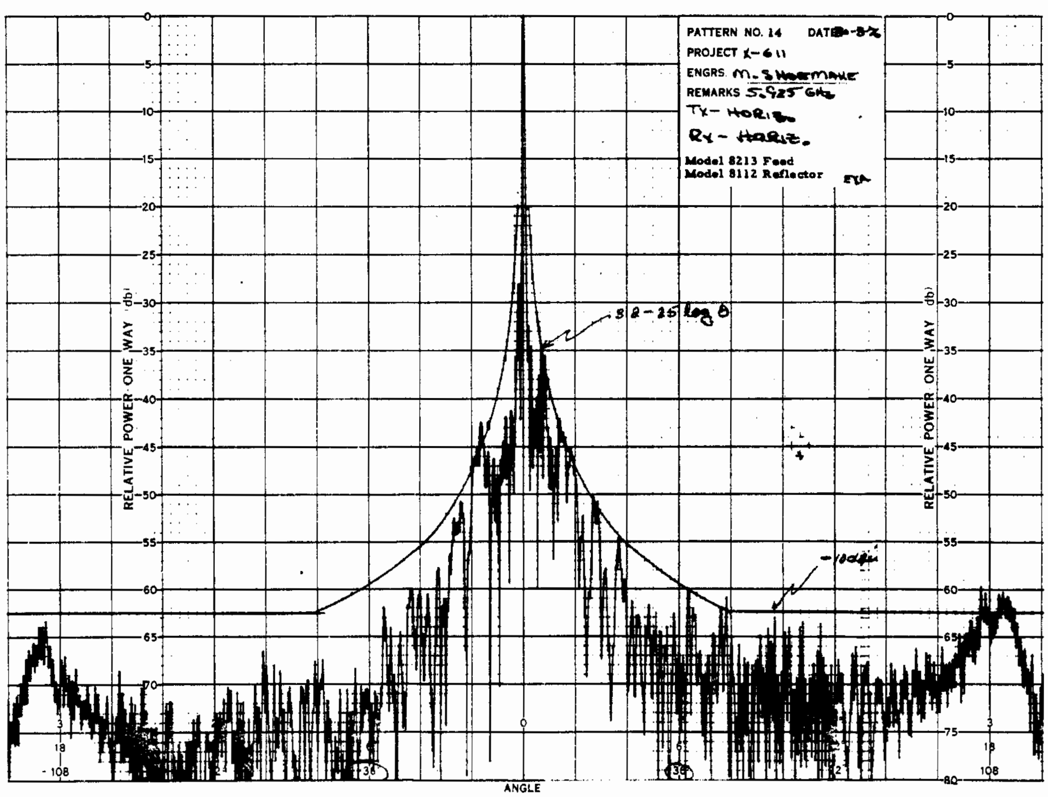


Figure 9. Typical 6 GHz Pattern of Prime Focus Antenna

The cross-polarization, far-field pattern of the earth station antenna is important to providing the polarization discrimination characteristics required when operating with a frequency-reuse satellite. If the antenna is designed to achieve the characteristics previously described, the cross-polarization pattern will be determined largely by the corresponding feed pattern. The excellent cross-polarization characteristics of a corrugated horn are well known. The cross-polarization pattern of a 10-meter reflector shown in Figure 10 was achieved by using this type of feed.

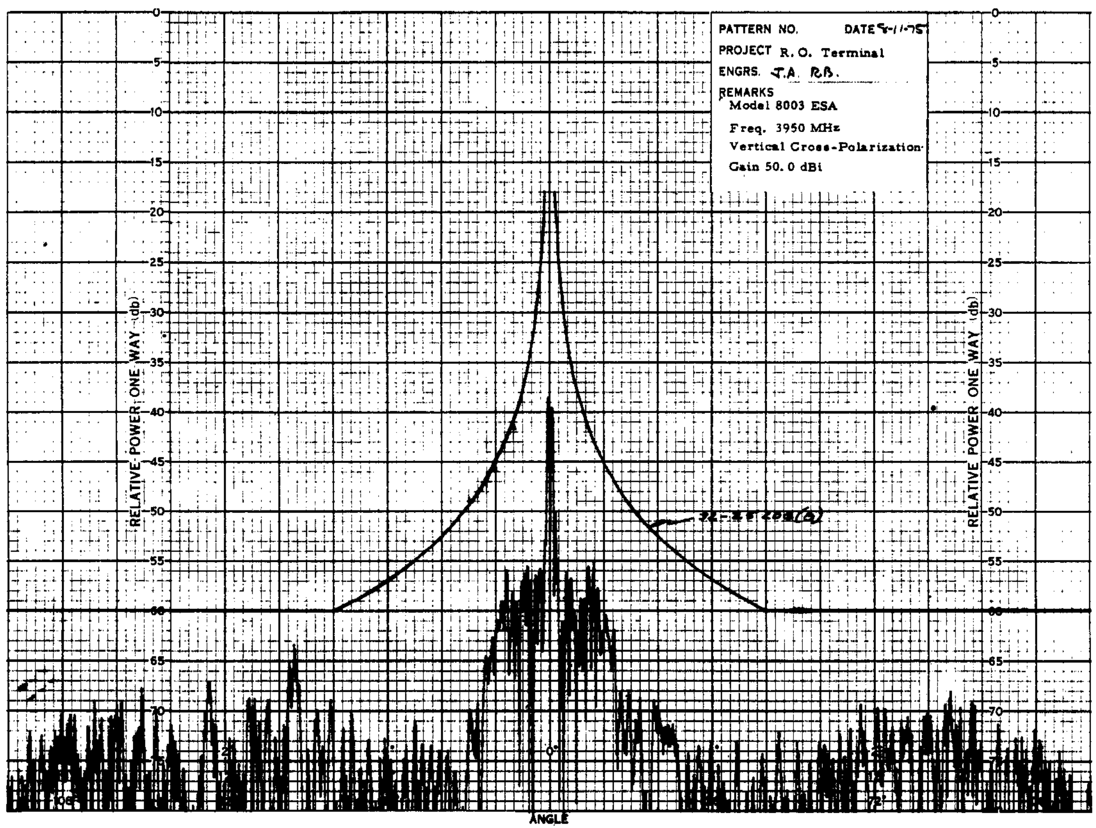


Figure 10. Typical Cross Polarization Pattern of 10 Meter Prime Focus Antenna

Cassegrain Antenna Geometry

The Cassegrain antenna design is derived from the telescope design of William Cassegrain. It is a dual-reflector antenna employing what are normally termed a "main reflector" and a "subreflector." In the true Cassegrain geometry, the main reflector is a paraboloid and the subreflector is a hyperboloid. Variations include the Gregorian design with an ellipsoidal subreflector and so-called "shaped" designs where the surfaces are generated numerically to realize a specified amplitude and phase distribution.

As shown in Figure 11 and 12, the Cassegrain feed horn/subreflector combination replaces the feed in the prime focus configuration. This leads to several advantages and disadvantages as will be discussed. Consider a plane wave impinging on the main reflector. Being paraboloidal, the main reflector directs all the rays toward its focal point just as in the prime focus geometry. Now, however, the hyperboloidal subreflector reflects the rays to its second focal point. The primary feed horn is placed so that its phase center is coincident with the second subreflector focus and, therefore, it receives the incoming energy. By reciprocity, the process can simply be reversed on transmission.

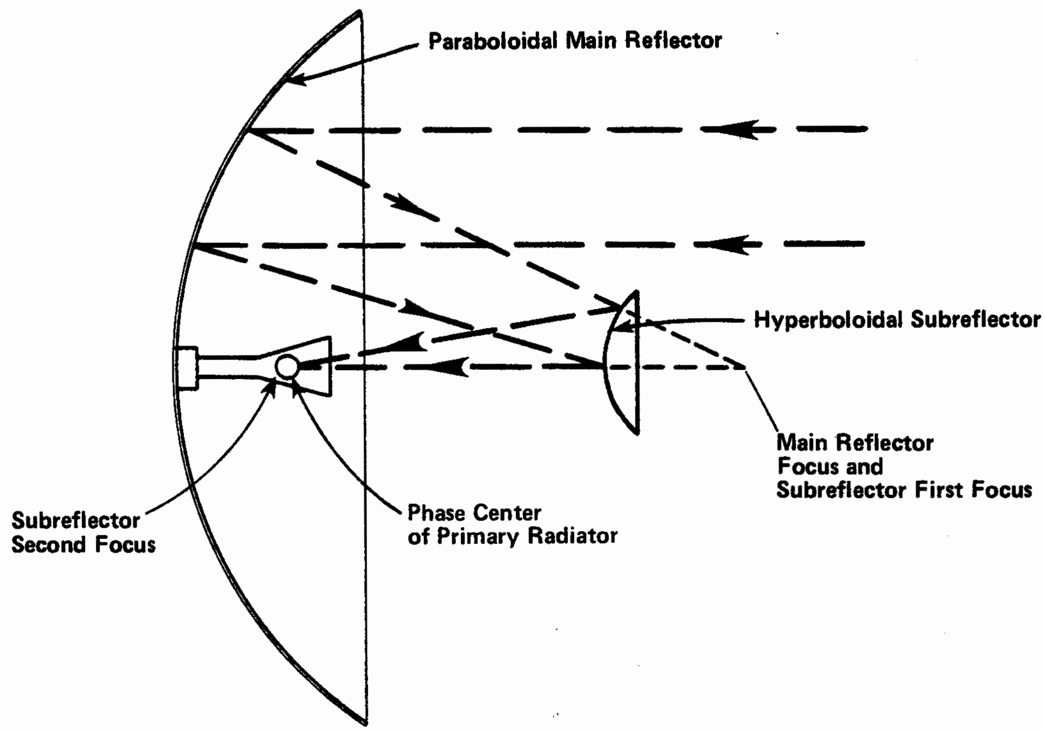


Figure 11. Cassegrain Antenna Geometry

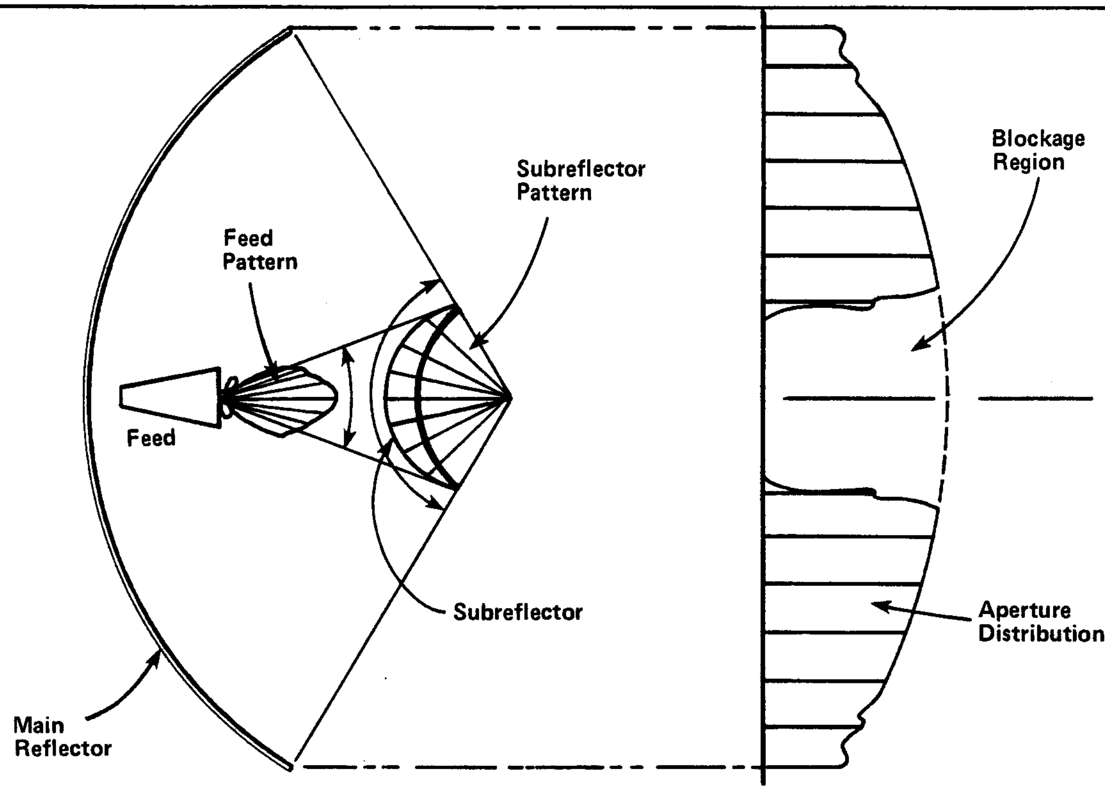


Figure 12. Aperture Distribution of a Cassegrain Antenna

This simplified discussion is based on geometrical optics, realizable only by working with zero wavelength or infinite structures. In real antennas the performance is affected by diffraction effects, particularly from the reflector and subreflector edges and the subreflector support structure. The subreflector diffraction alters the pattern produced by the subreflector as it illuminates the main reflector (considering the antenna as a transmitter). A typical scatter pattern is shown in Figure 13. This is the pattern formed by the feed/subreflector combination. Notice the ripple caused by energy diffracting at the subreflector edges.

The subreflector diffraction effects limit the use of dual-reflector antennas to main reflectors of about 40 λ and larger, with the exact limit depending on specifications. A typical standard Cassegrain antenna of 80 λ in diameter will give 60 to 65% overall efficiency.

The Cassegrain design ensures uniform phase, as can be seen by verifying that all rays travel equal distances from the feed to the subreflector to the main reflector and back to the antenna aperture plane.

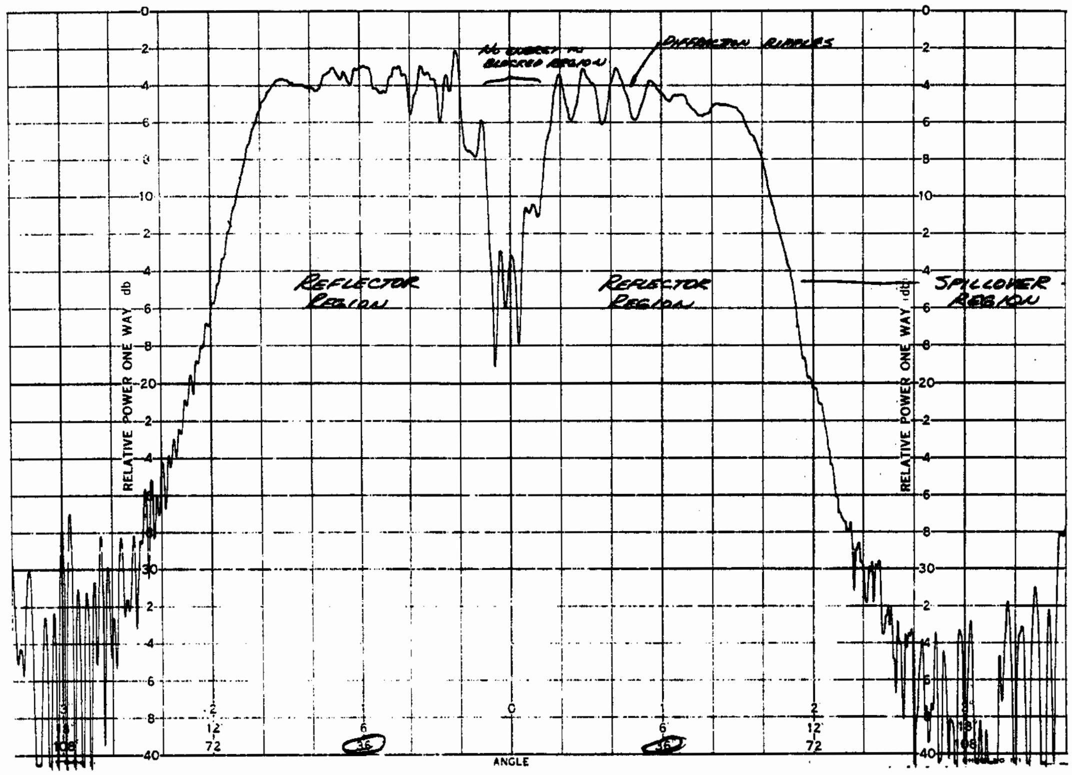


Figure 13. Typical Feed/Subreflector Scatter Pattern

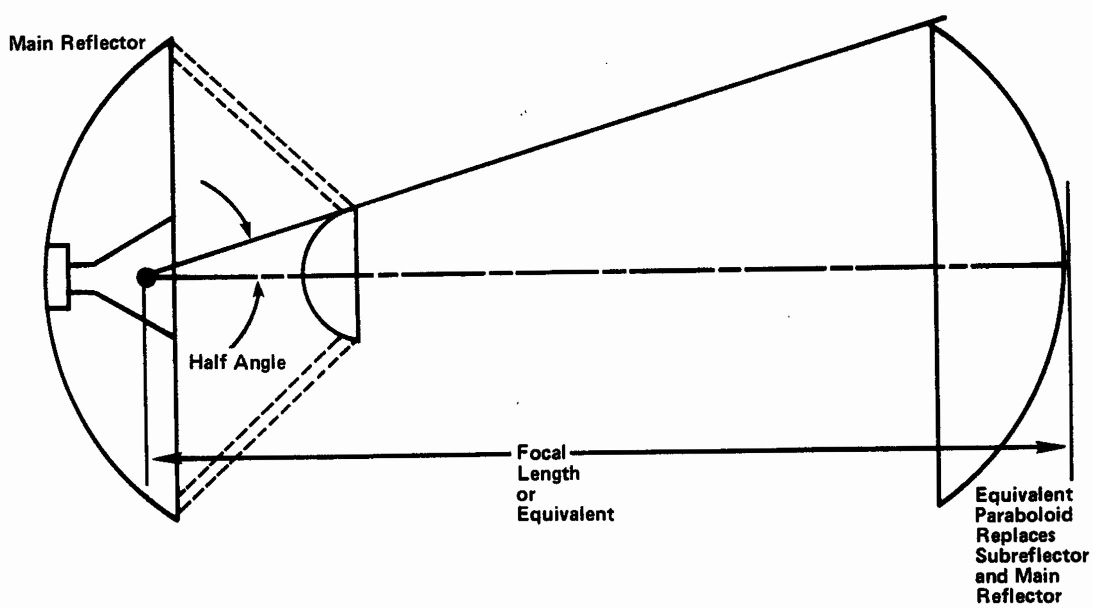


Figure 14. Equivalent Paraboloid for a Cassegrain Antenna

After ensuring uniform phase, we investigate the aperture amplitude distribution. This is straightforward if we think of the Cassegrain antenna in its equivalent prime focus geometry. Figure 14 shows the concept. In essence, the main reflector is mirrored in the subreflector. This permits replacing the subreflector and main reflector by an equivalent paraboloidal reflector with a very long focal length. The ratio of this length to the true main reflector focal length is termed the magnification factor. The equivalent antenna consists of this paraboloid being fed by the unaltered Cassegrain feed horn.

In a prime focus system, the feed pattern determines the aperture amplitude distribution; since the Cassegrain has an associated equivalent prime focus system, the same is true for a Cassegrain system. The subreflector and main reflector will essentially reproduce the feed pattern in the aperture plane. Therefore, the feed horn performance is critical to antenna performance.

Several types of feed horns are currently in use in earth station antennas--diagonal horns, corrugated horns, and multimode horns. The selection is dictated by RF specifications, weight, size, fabrication techniques and economics.

A diagonal horn feed is often chosen when dissimilar transmit and receive patterns are desirable. This is the case for an antenna needing very low 6-GHz sidelobes and a very high 4-GHz gain. Figures 15 and 16 show the radiation patterns of such a diagonal horn. The large taper (24 dB) at 6 GHz produces low sidelobes and the 14-dB taper at 4 GHz produces a near uniform aperture distribution and, therefore, high efficiency.

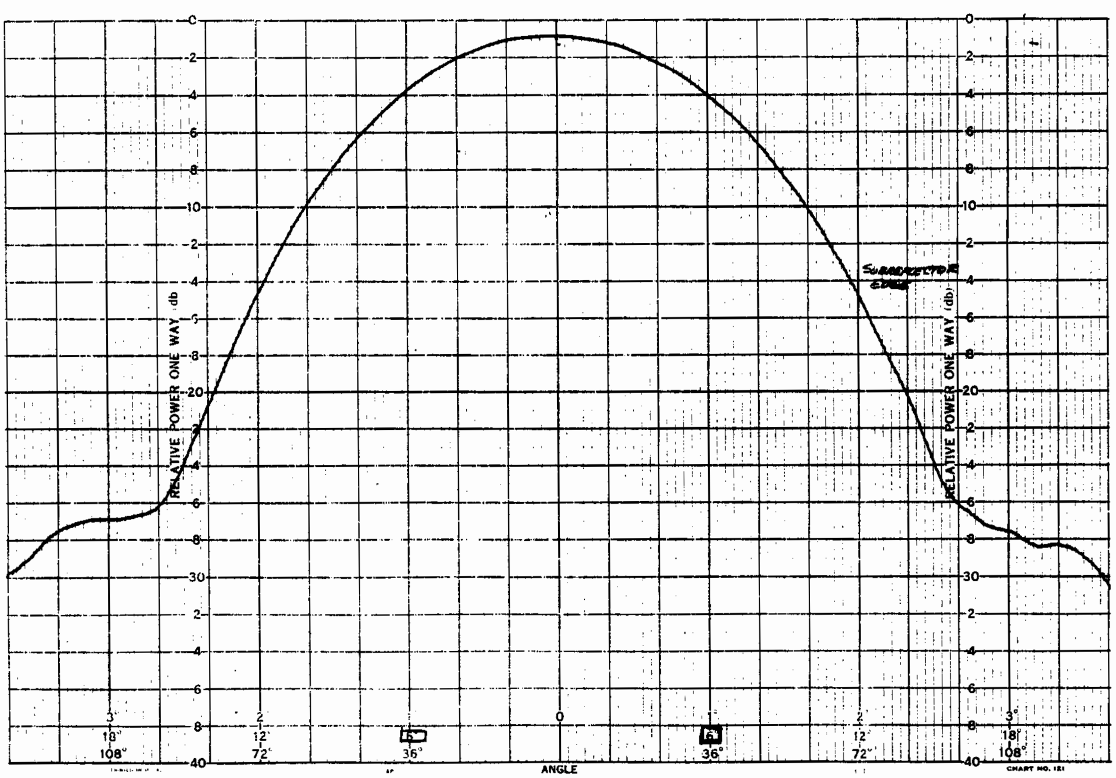


Figure 15. 4.2 GHz Diagonal Horn Pattern

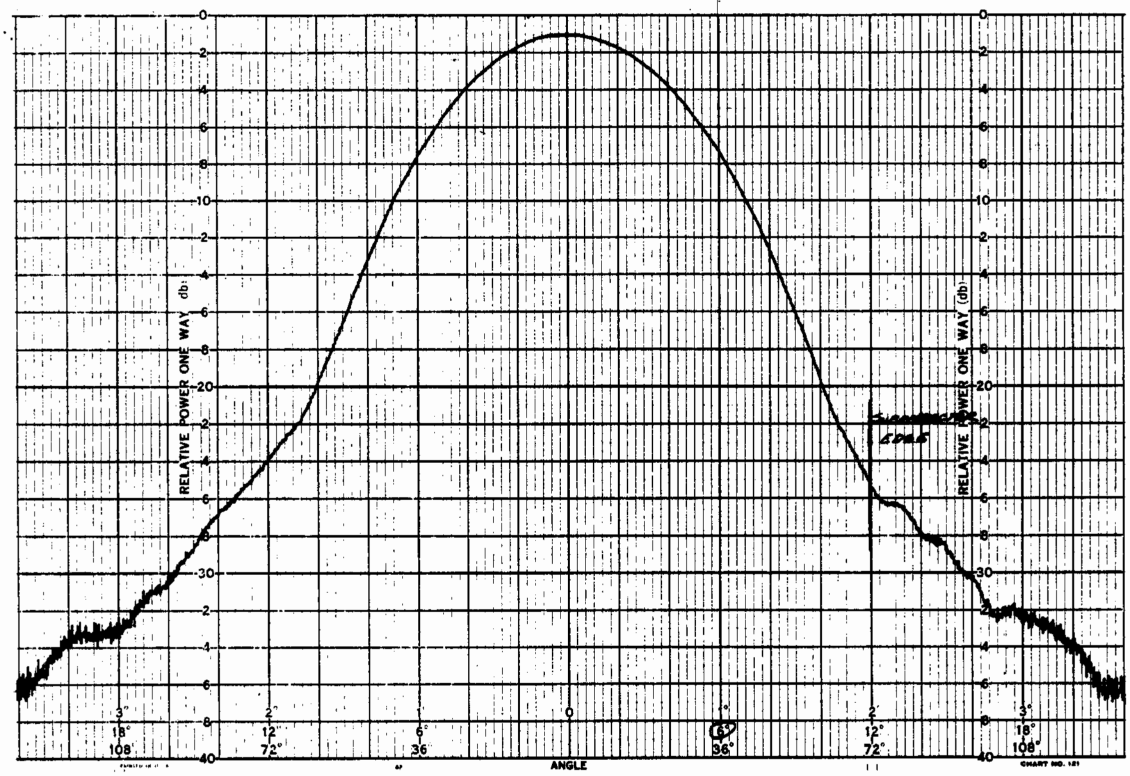


Figure 16. 6.175 GHz Diagonal Horn Pattern

A corrugated feed horn is often used when the transmit and receive specifications are similar because it maintains a constant beamwidth over large frequency bands. The receive and transmit distributions will be similar, yielding similar efficiencies and sidelobes. Typical corrugated horn patterns are shown in Figure 17 and 18.

The multimode-feed horn is also employed as an earth station feed. It has characteristics similar to the corrugated horn except that its radiation characteristics are much more frequency sensitive.

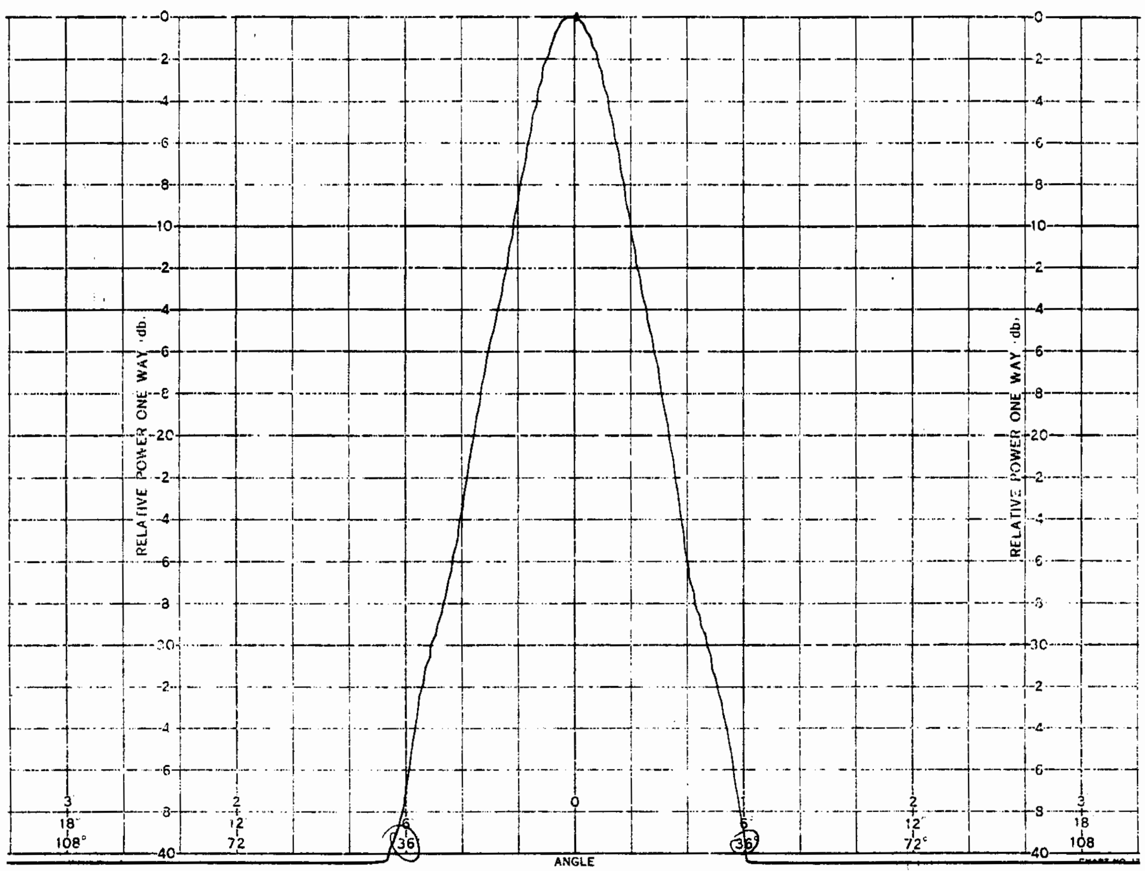


Figure 17. 3.95 GHz Corrugated Horn Pattern

In addition to affecting the aperture amplitude distribution, the feed pattern determines the amount of energy spilled over past the subreflector. From Figure 19 it is obvious that to keep this energy low, it is important that the feed pattern drop sharply beyond the subreflector. Not only will this improve the antenna gain, but it will also prevent large sidelobe increases in the angular region around the subreflector. Notice from the patterns of Figure 20 and 21 that the sharper dropoff of the diagonal horn means lower sidelobes in the subreflector spillover region. Thus, the Cassegrain design forces the designer into a compromise: either use a very high taper (low energy) on the subreflector edge to reduce the spillover, or suffer the subreflector spillover in order to have a more uniform distribution on the main reflector.

Shaped-reflector systems were devised to overcome this Cassegrain compromise. The idea is to shape the reflectors such that a large energy taper on the subreflector will still produce an aperture distribution which is nearly uniform in both phase and amplitude. The subreflector and main reflector coordinates are usually calculated numerically by computer, with no closed form solution for the surfaces ever being determined. Both reflectors are still surfaces of revolution, but they are no longer a hyperboloid and a paraboloid.

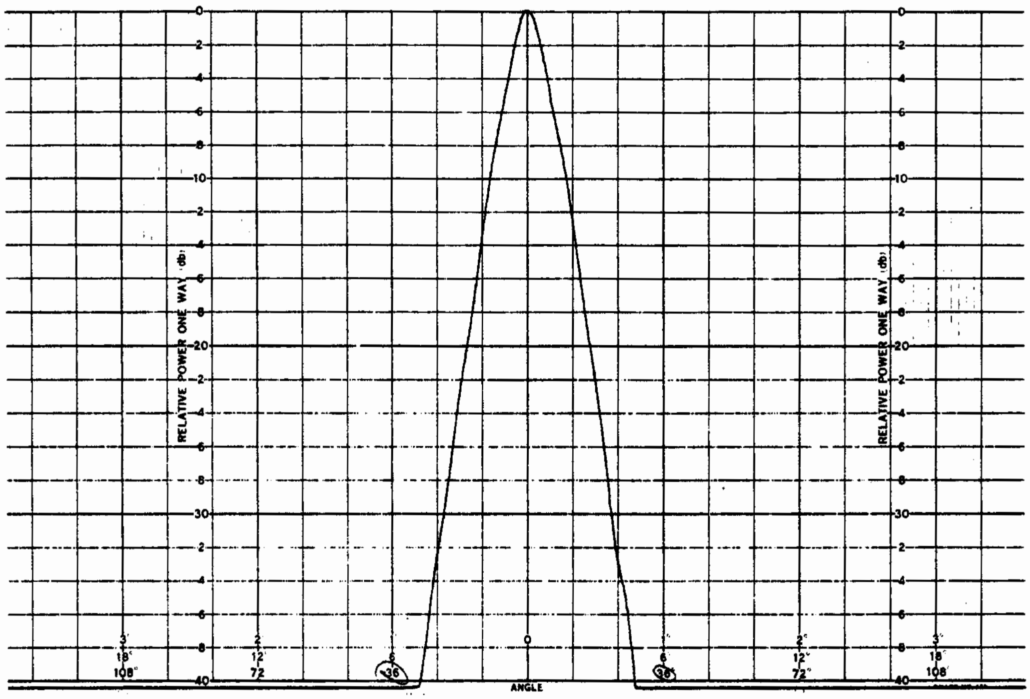


Figure 18. 6.175 GHz Corrugated Horn Pattern

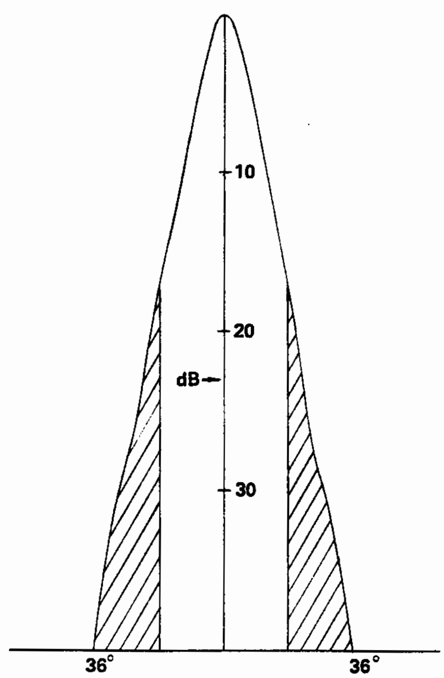


Figure 19. Typical Feed Pattern Showing Spillover Energy

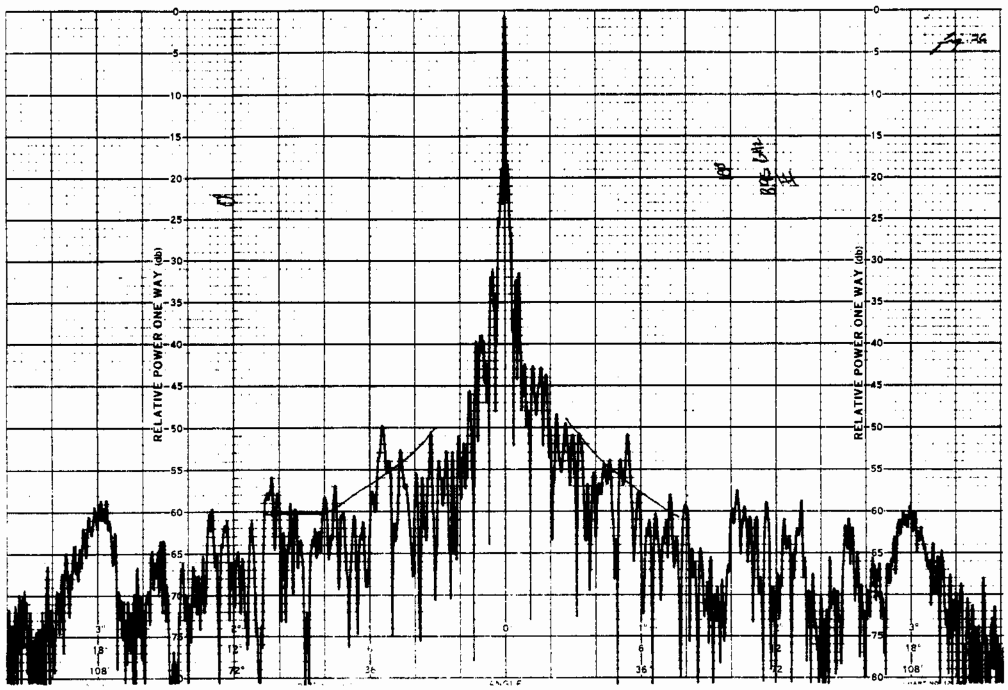


Figure 20. Pattern of a 10 Meter Earth Station with Diagonal Horn Feed

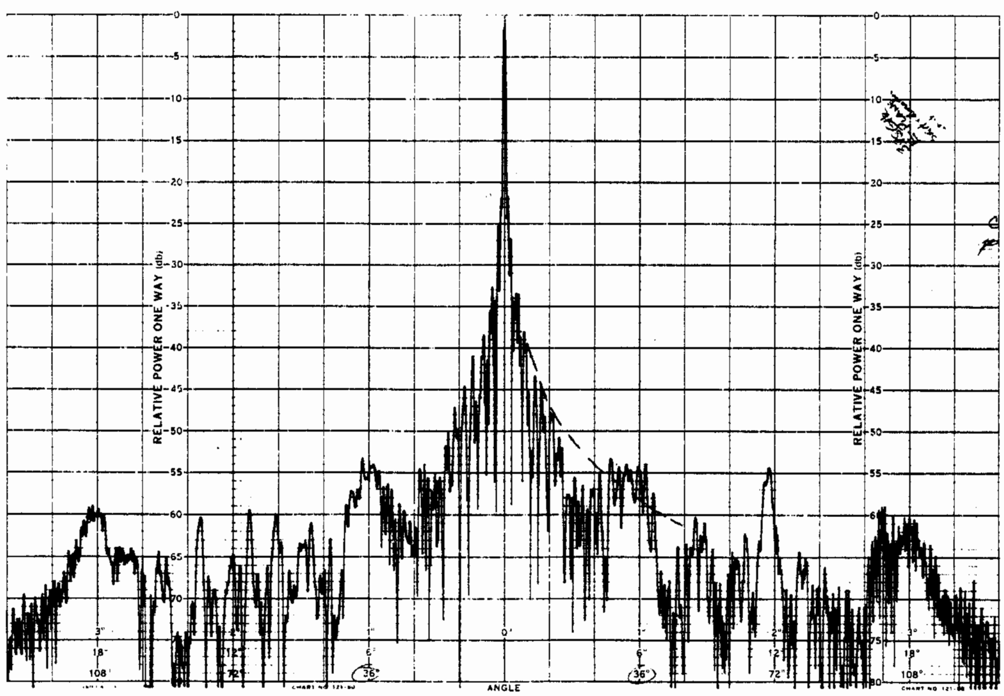


Figure 21. Pattern of Antenna of Fig. 20 with Corrugated Horn Feed

While shaped-system coordinates are actually the solution of several simultaneous equations, it is helpful to imagine the shaping process as follows: we wish to keep the spillover energy low, so we use a narrow feed pattern. However, this means a large subreflector edge taper, and so a large main dish taper. The main dish is being utilized inefficiently since its outer area has very little energy on it. So we shape the subreflector to redistribute the energy, directing energy from the center toward the outer portion of the main reflector. At the same time, we correct the main reflector to restore uniform phase across the aperture with the shaped subreflector.

A typical partial computer printout for a shaped system's coordinates is shown in Figure 22. The input parameters in this case are subreflector size, main reflector size, half angle to the subreflector, feed location, feed pattern shape and aperture distribution.

```
? 30.,222.,20.,0.,80,0,.631,222,0,158.

SHAPED REFLECTOR PROGRAM    MODIFIED FOR CORRUGATED HORN

MAIN REFLECTOR DIAM. =      444.00 INCHES

SUBREFLECTOR DIAM. =      60.00 INCHES

HALF ANGLE TO SUBREFLECTOR =      20.00 DEGREES

FEED DISTANCE FROM APERTURE PLANE =      0.     INCHES

TAPER ON SUBREFLECTOR =      -21.61 DB

TAPER ON MAIN REFLECTOR =      -4.00 DB
```

X2	Y2	X1	Y1	THETA1	THETA2	Y0-Y2
222.0000	-.0000	30.0000	0.	20.0000	66.7665	.0000
221.0000	.6600	28.2329	.7739	19.0742	66.8778	.0409
220.0000	1.3205	26.9956	1.3267	18.4114	66.8762	.0783
219.0000	1.9805	26.0352	1.7613	17.8883	66.8154	.1129
218.0000	2.6395	25.2460	2.1218	17.4524	66.7178	.1454
217.0000	3.2971	24.5736	2.4310	17.0767	66.5950	.1760
216.0000	3.9535	23.9550	2.7175	16.7276	66.4607	.2047
215.0000	4.6078	23.3270	3.0094	16.3694	66.3290	.2324
214.0000	5.2605	22.6875	3.3081	16.0009	66.2005	.2585

Figure 22. Computer Printout for Dual Shaped Reflector Coordinates

A dual-shaped system is capable of efficiencies of 65 to 80%. A gain budget for such a system designed for 68% efficiency is shown in Figure 23. This particular system has a 9-dB "uptaper," that is, the geometry of the system uses a 13-dB subreflector taper to produce a 4-dB main dish taper. Subreflector diffraction must then, of course, be accounted for.

Gain Factors	4000 MHz	6000 MHz
Gain of Ideal 10 Meter Antenna	*52.50 dBi*	*56.02 dBi*
Aperture Illumination Efficiency	*-0.09 dB*	*-0.50 dB*
Gain of Ideal 10 Meter Antenna	*52.50 dBi*	*56.02 dBi*
Aperture Illumination Efficiency	*-0.09 dB*	*-0.50 dB*
Spillover for Main Reflector	*-0.10*	*-0.06*
Spillover for Subreflector	*-0.60*	*-0.50*
Blockage Due to Subreflector and Spars	*-0.35*	*-0.64*
Surface Tolerance	*-0.13*	*-0.30*
Primary Pattern Phase Error	*-0.12*	*-0.20*
Cross-Polarization Loss	*-0.06*	*-0.10*
Loss in Horn—Orthomode Assembly (including mismatch)	*-0.20*	*-0.20*
Gain at the Output of the Orthomode Transducer	*50.85 dBi*	*53.5 dBi*

Figure 23. Gain Budget for a 10 Meter Dual Shaped Antenna

To illustrate how main reflector taper affects gain, Figure 24 lists illumination efficiencies for various $1-\rho^2(1-A)$ tapers. As with all antennas, the more uniform distributions represent generally higher first sidelobes.

Edge Taper (A) (dB)	Voltage	Illumination Efficiency (%)	Gain Loss (dB)
0	0	100	0
1	.891	.999	.005
2	.794	.996	.019
3	.708	.990	.042
4	.631	.983	.074
5	.562	.975	.112
6	.501	.965	.157
7	.447	.954	.207
8	.398	.942	.206
9	.355	.930	.317
10	.316	.918	.374
15	.178	.860	.654
20	.100	.818	.875

Figure 24. Illumination Efficiency vs. Taper for a $1-p^2$ (1-A) Voltage Distribution

Shaping techniques offer several advantages in addition to a more uniform distribution. For instance, a standard Cassegrain system has a gain loss due to subreflector blockage firstly because that area of the aperture is unusable, and secondly because the system puts energy into that wasted area. Shaped designs allow the designer to direct energy away from the central blocking region, allowing blockage to reduce gain largely by a loss in area. In addition, this means that reduced energy is directed back toward the feed horn so that the reaction of the subreflector on the feed impedance is geatly reduced.

Another shaping technique is possible with reflectors which are equal to or greater than about 150 wavelengths in diameter. This technique maintains a high energy level on most of the main reflector but drops sharply near the periphery. The result is to produce low sidelobes in the reflector spillover region and yet a reasonably high efficiency.

In all reflector antennas, including dual reflectors, objects in the aperture affect sidelobes. This is shown in Figure 25 where two spars of the antenna of Figure 20 were replaced with small guy wires. Notice that the pattern is affected in nearly all regions.

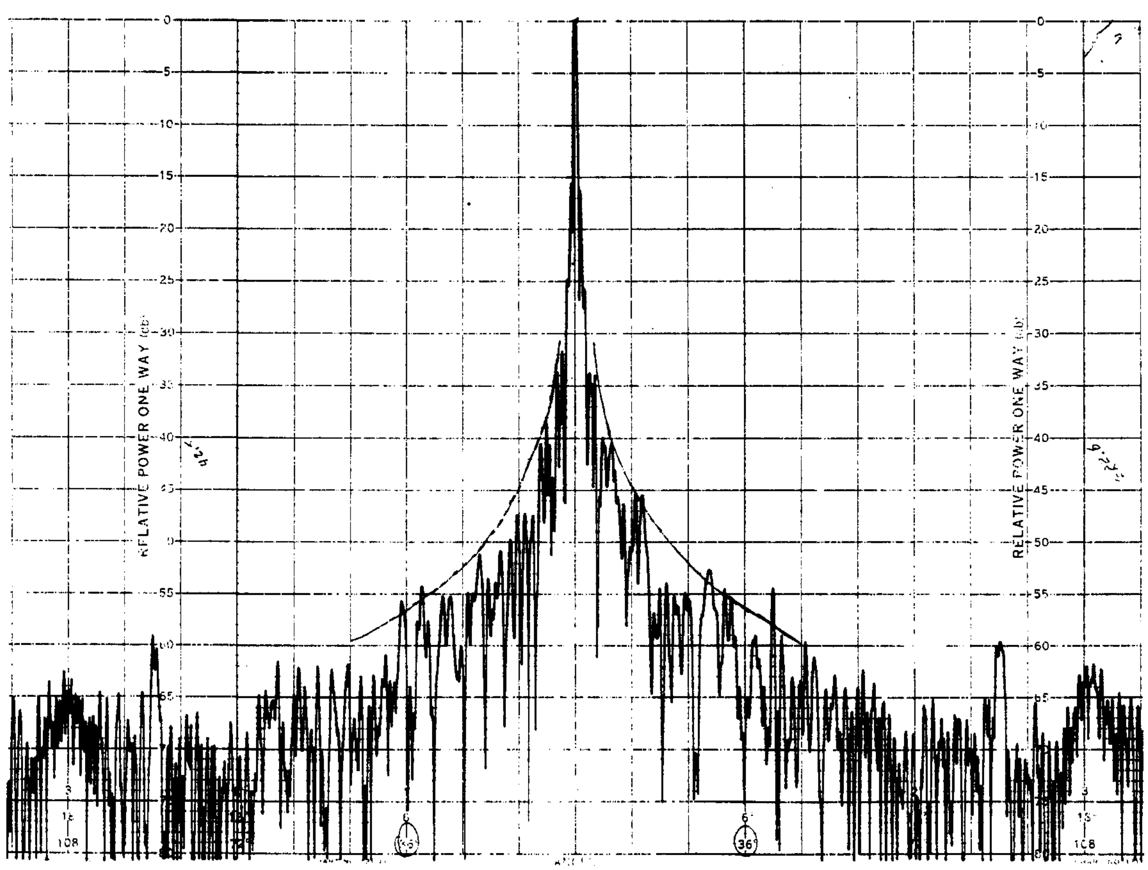


Figure 25. Same Pattern as Fig. 26 but with 2 Spars Replaced by Guy Wires

Comparisons

The following is a general rule of thumb for earth stations in the 10-meter range: if patterns are of prime importance, use a prime focus antenna; if gain is the critical parameter, use a dual-shaped system. For example, a comparison of the pattern of Figures 26, 27 and 8, 9 shows the prime focus design to have superior sidelobes, facilitating antenna coordination and site selection. This is primarily due to the reduction of blockage, diffracting and scattering contributors (spars and subreflector).

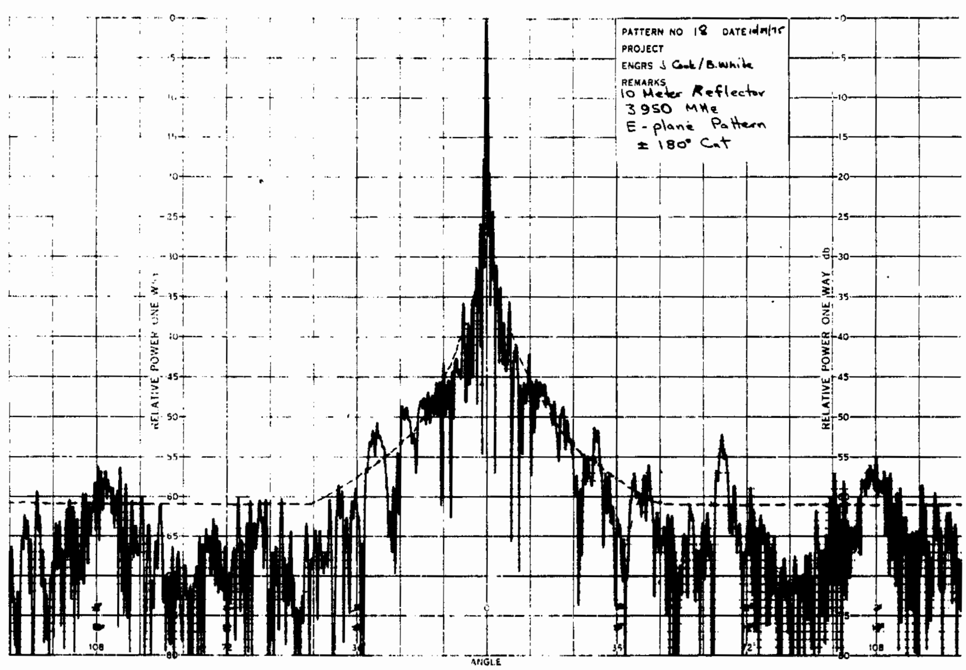


Figure 26. Typical 4 GHz Pattern of 10 Meter Dual Reflector Antenna

The improved sidelobes of a prime focus design are not gratis, however, they are paid for with reduced gain. The antenna of Figure 26 provides 0.65 dB more receive band gain than the prime focus antenna of Figure 8. 0.4 dB is attributable to aperture efficiency increase and 0.25 dB to a reduced length of waveguide, assuming that the LNA is located in the reflector hub.

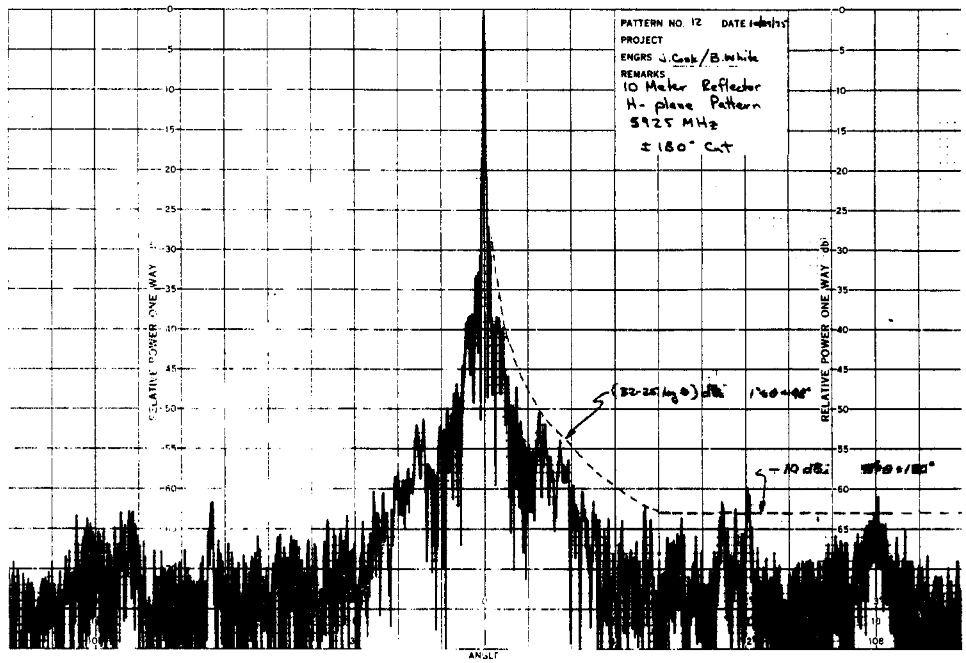


Figure 27. Typical 6 GHz Pattern of 10 Meter Dual Reflector Antenna

This 0.25-dB resistive loss not only reduces gain, but it adds 17K of noise to the system. This noise increase along with the 0.65-dB gain decrease means a 1.5-dB G/T reduction in a typical system with a 55K LNA. EIRP can similarly be about 0.8 dB higher with dual-reflector systems.

Polarization requirements often dictate the use of dual-reflector antennas. Both the prime focus and dual-reflector antennas of Figures 8, 9 and 26, 27 are equipped with feeds which are rotatable from the reflector hub, so they are both suitable for simple polarization adjustment. However, such schemes as polarization tracking, whether it be programmed Faraday correction or polarization autotracking, call for the vertex-mounted feed of a dual-reflector system.

The complexity of an autotracking feed (i.e., feed for an AZ-EL autotrack system) also normally requires a dual-reflector geometry.

References

1. "A Simplified Approach to Cassegrain Antennas," William T. Whistler, Electronic Design, January 20, 1964, p. 44.

2. "RF Design of Communication Satellite Earth Stations," Cuccia, Williams, Cobb, Rahilly, Microwaves, July 1967.

3. "Microwave Antenna Theory and Design," by Samuel Silver, 1964, Dover Publications, N.Y.

Re-Use Earth Station Antennas

Rick Barker

Introduction

The commercial satellite communications industry has grown substantially in the past decade. In the early satellites, the power supply system was the chief factor limiting the communications capacity. This restriction was determined by the payload limit that the launch vehicle could carry. Recently, the limitations are not power availability, but the available bandwidth that can be used in existing communications link.

This problem of limited spectrum has numerous solutions. New frequency band allocations such as 14/12 GHz and 30/20 GHz are in use and offer additional bandwidth to ease the load. However, these require entirely new satellite and earth station systems. Another method, is to "reuse" the common C-Band (3.7- to 4.2-GHz downlink and 5.925- to 6.425-GHz uplink) by making use of the inherent isolation characteristics of orthogonal polarizations.[1]

The polarization state of the electromagnetic field allows two independent signals to be processed through the same communications link in the same frequency band. This method effectively doubles the bandwidth capacity of a satellite communications system. Polarization frequency reuse should not be confused with spatial reuse in which the frequency spectrum is reused with a multiple-beam antenna isolated by the spatial separation of the radiation patterns of each beam. As an example of a frequency-reuse satellite, the RCA SATCOM satellite has twelve transponders that are horizontally polarized and twelve transponders that are vertically polarized, thus permitting the reuse of the entire 500-MHz bandwidth in both transmit and receive applications. The carrier-to-interference ratio is enhanced by offsetting the center frequencies of each orthogonal channel by half the normal channel spacing. (See paper included in this book by J. Searcy Hollis, "Principles of Satellite Communications.")

Cross-Polarization Discrimination in Frequency Reuse Systems

The success of frequency-reuse systems relies ultimately on maintaining high isolation between orthogonal channels. Frequency offsetting and the use of high quality antenna components in both satellite and earth station systems minimize the polarization coupling between channels which would otherwise result in interference and consequent signal-to-noise reduction. The measurement of polarization discrimination characterizes the ability of an antenna system to provide isolation between orthogonal channels. Polarization discrimination is defined as the ratio of power received (or transmitted) in the principal polarization to the power received in the orthogonal polarization.

Depolarization Factors

Factors in the antenna system, as well as the anisotropic behavior of the propagation medium, cause depolarization of the desired signal. The cross-polarization discrimination provided by the satellite antenna varies depending on the quality of the antenna and the relative location of the earth station. High quality antennas achieve 30- to 40-dB polarization discrimination along the antenna boresight. Off-axis, the polarization discrimination deteriorates. The worst case is usually in the 45° plane of the radiation pattern.

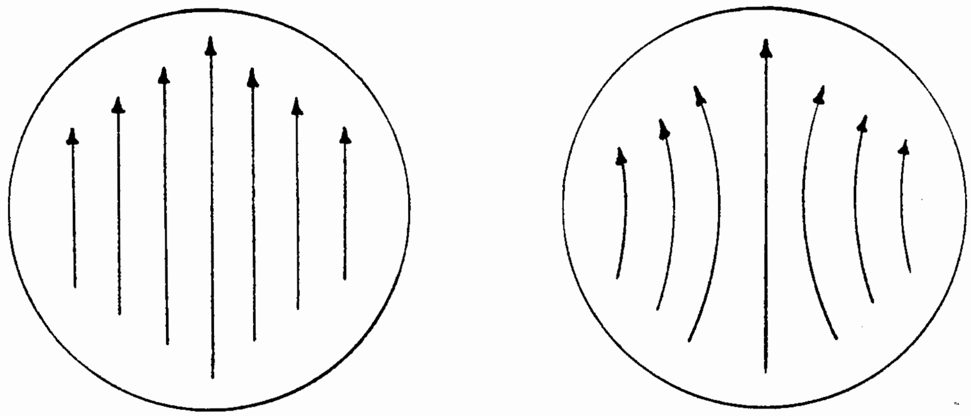
High winds or satellite drift can affect the pointing of the antenna. If the satellite and earth station antennas are not perfectly aligned, cross-polarization degradation can result. Further deterioration of the signal occurs from various phenomena in the propagation path, such as Faraday rotation in the ionosphere and depolarization due to the presence of rain. A summary of the sources of depolarization is given in Table 1.

Table 1. Depolarization Requirements for Good Dual Polarization Antenna Performance

Sources of Depolarization	Typical Objectives
OMT-Polarizer Isolation	40 dB
OMT-POlarization Axial Ratio	
a) Linear Polarization	40 dB
b) Circular Polarization	0.2 dB
Input VSWR to Feed Horn	1.3:1
Feed Horn Axial Ratio	
a) Linear Polarization	40 dB
b) Circular Polarization	0.2 dB
Subreflector Alignment	0.7 mm offset
Asymmetric Surface Distortion of Main Reflector	0.5 mm
Sureflector Support Spars	Tripod or Quadrapod
Light Rainfall (Compensation not Mandatory)	27 mm/hr. (27 dB)

Antenna Properties Related to Depolarization

Cross-polarization is produced by the nature of the field distribution in the antenna aperture. Variations of the electric field over the aperture result in radiated cross-polarization, as illustrated in Figure 1. A perfectly symmetric and smooth reflector does not generate a depolarizing component on-axis or in the principal planes (parallel or perpendicular to a linear polarized vector), but would generate a depolarization component in other planes. Asymmetry and surface errors in the reflector cause further reduction in an antenna's polarization discrimination. Additional depolarization can occur from the subreflector in a dual-reflector type antenna.



A. Parallel fields would produce no cross-polarization.

B. Anti-parallel fields produce off-axis cross-polarization.

Figure 1. Linear Field Distribution on a Circular Antenna Aperture

Although several guidelines can be stated, it is difficult to compare the performance of different classes of antennas; for example, prime-focus and dual reflector, or Cassegrain and Gregorian. Polarization discrimination performance improves as the reflector diameter increases relative to the wavelength (both main reflector and subreflector). As the reflector approaches a planar surface (large f/D), the field distribution in the aperture more closely resembles the parallel field-line condition shown in Figure 1, thereby reducing the cross-polar contribution. Also, a shaped, high-efficiency subreflector (used in quasi-Cassegrain antennas) contributes to reduction in cross-polarization over the traditional hyperboloidal/parabolodial configuration.[2]

The primary feed polarization property is the dominant factor in the determination of an antenna's polarization discrimination. Feeds which produce symmetric radiation patterns in all planes are ideal. Corrugated horns are more balanced than diagonal horns in this respect. Diagonal horns, though, are preferable to square, rectangular, or conical horns. Small-aperture,

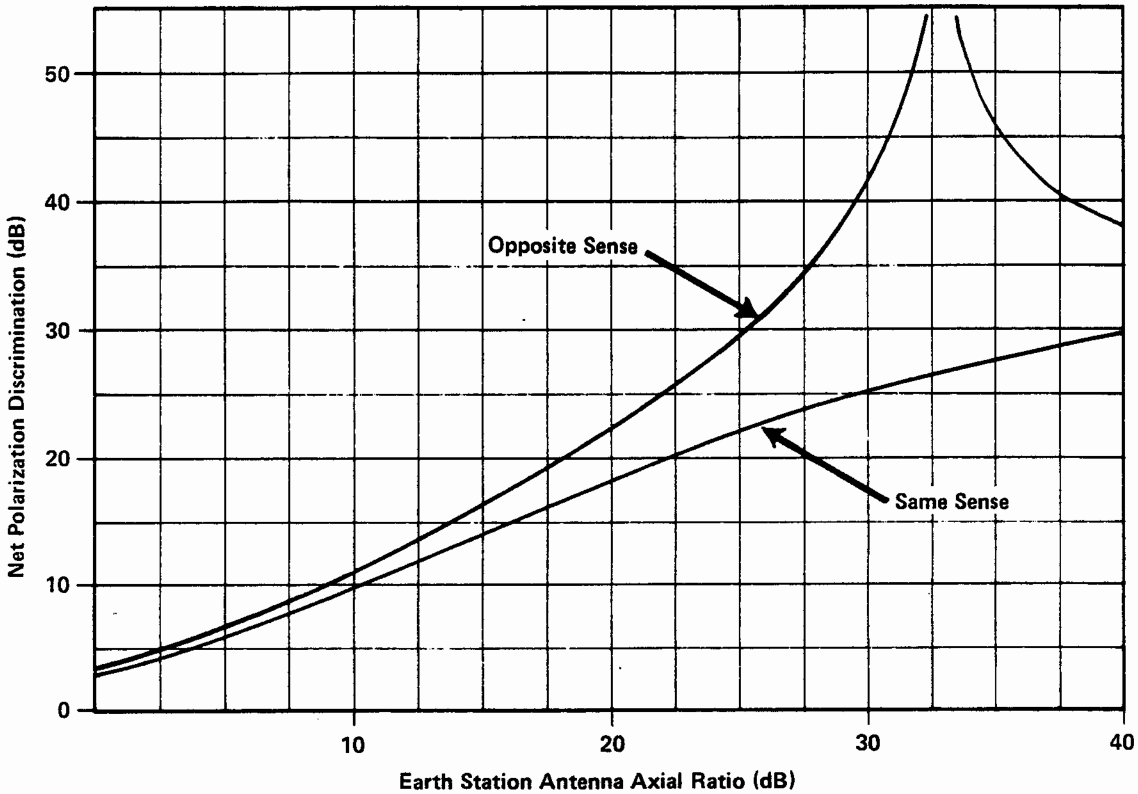


Figure 2. Net Polarization Discrimination for Earth Station Links with a Typical (33 dB Axial Ratio) Satellite Antenna

circular horns radiate less cross-polarized content than large-aperture horns because the fundamental TE_{11} mode field lines curve less in the small-aperture horns. However, with small-aperture horns a large half-angle is usually subtended by the main reflector (or subreflector). Problems associated with subtending a larger reflector angle are a low f/D ratio in the prime-focus configuration or increased blockage and VSWR in the dual-reflector configuration.

Two antennas with different values of polarization discrimination interact to produce a combined effect for a given satellite earth station link. Figure 2 illustrates the net effect for a range of earth station antenna axial ratios assuming a satellite antenna with 33-dB axial ratio.

Ionospheric Depolarization

The ionosphere is a neutral gas containing charged particles (called a plasma). A plasma in the presence of a magnetic field becomes an anisotropic medium. The earth's magnetic field provides the condition necessary for the ionosphere to be anisotropic. The effect of the ionosphere is to rotate the polarization of a radio wave passing through this layer of the atmosphere. Faraday observed the same occurrence in light traveling through gyrotropic crystal with an applied magnetic field. Consequently, the phenomenon is called "Faraday rotation".

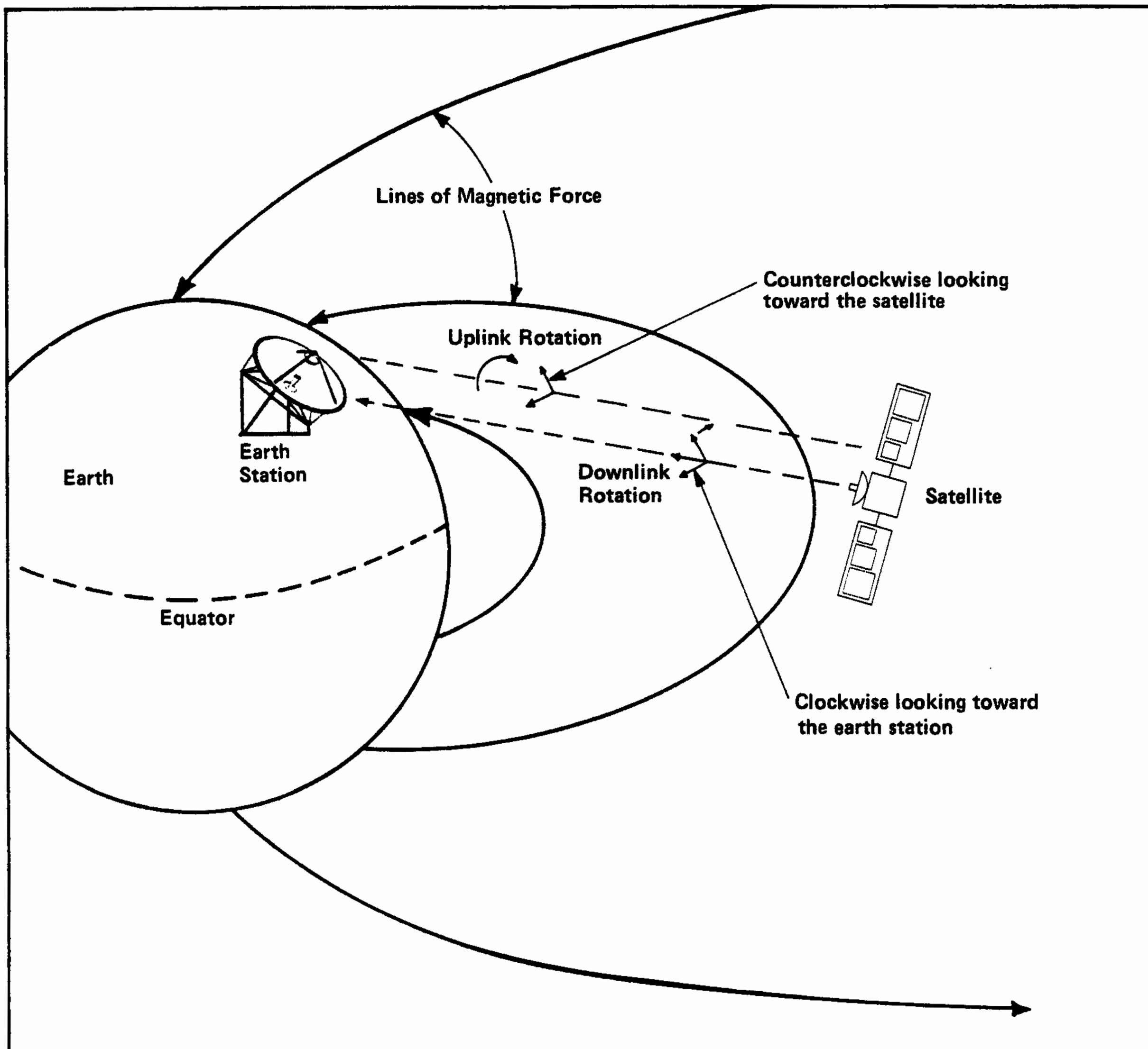


Figure 3. Faraday Rotation of Satellite Communication Signals in the Ionosphere

The rotation of an electromagnetic wave passing through the ionosphere depends on the direction of propagation (definition of anisotropic). For an earth station in the northern hemisphere looking toward a synchronous satellite, an uplink linear polarization is rotated counterclockwide (CCW), and a downlink linear polarization is rotated counterclockwise (CCW) (as viewed from the ground looking toward the satellite). Therefore, to correct for the polarization rotation in the receive signal, the earth station feed should be rotated CCW (as one faces the satellite). However, since the uplink is rotated CCW, the earth station transmit portion of the feed should be rotated CW in the opposite direction of the receive feed. Figure 3 illustrates the sense of depolarization experienced by microwave signals in the ionosphere for both uplink and downlink communications.

The magnitude of Faraday rotation depends on the electron density of the ionosphere, the magnetic field intensity, and the component of the wave traveling along the magnetic field vector. The amount of rotation varies hourly and seasonally depending on the concentration of electrons in the ionosphere. Generally, the ionization is highest during the day shortly after noon, during the winter or spring (for the northern hemisphere), and during periods of heavy solar activity. Also, the amount of Faraday rotation is inversely proportional to the square of the frequency. Whereas ionospheric depolarization may be a serious handicap at 4 and 6 GHz, at Ku-band and higher frequencies the problem is negligible. Typical peak values are near 4° at 4 GHz and 2° at 6 GHz for northern continental United States.[3] Figure 4 correlates the amount of polarization rotation with the resulting reduction in cross-polarization discrimination. The polarization discrimination attributed to 4° of Faraday rotation is approximately 23 dB.

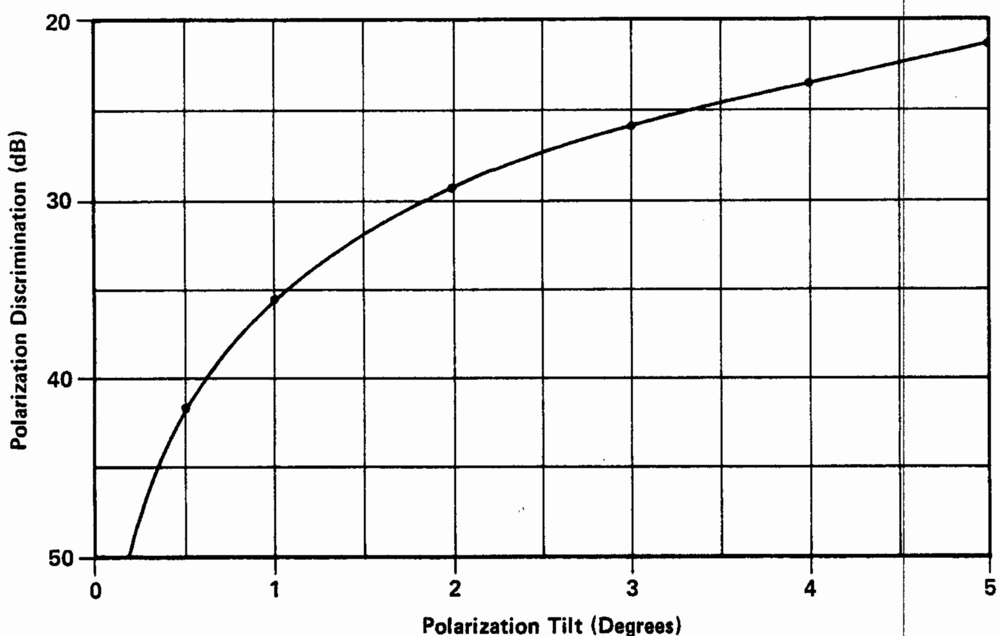


Figure 4. Effect of Polarization Rotation on Polarization Discrimination

Rain Depolarization

Rainfall along the transmission path from the earth station to the satellite depolarizes the signal. The depolarization comes about because the falling raindrops are elliptical in shape and cause amplitude and phase distortions in the two orthogonal linear polarization components of any incident polarization. The result of the propagation through the rain cell is therefore a change in the polarization of the signal. The resultant depolarization is a function of frequency, rain rate, rain cell size and look angle to the satellite. The amount of depolarization caused by the rain cell is dependent upon the incident polarization, but all polarizations (linear, elliptical, or circular) are affected; for example, a pure linear polarization (axial ratio equal to infinity) will be changed to an elliptical polarization (axial ratio will be greater than 1) as it passes through a rain cell.

The amplitude and phase distortions are, as mentioned above, frequency dependent. In the 6/4-GHz bands, accepted theory points out that the differential phase between the two orthogonal linear compoenents is the dominant factor of the two. At the higher frequency bands, 14/12 or 30/20, both factors are significant. Therefore, rain depolarization for a given rain cell and rain rate is more serious for the higher frequency bands. Typical depolarization due to rain for the 6/4-GHz band is shown in Figures 5 and 6.

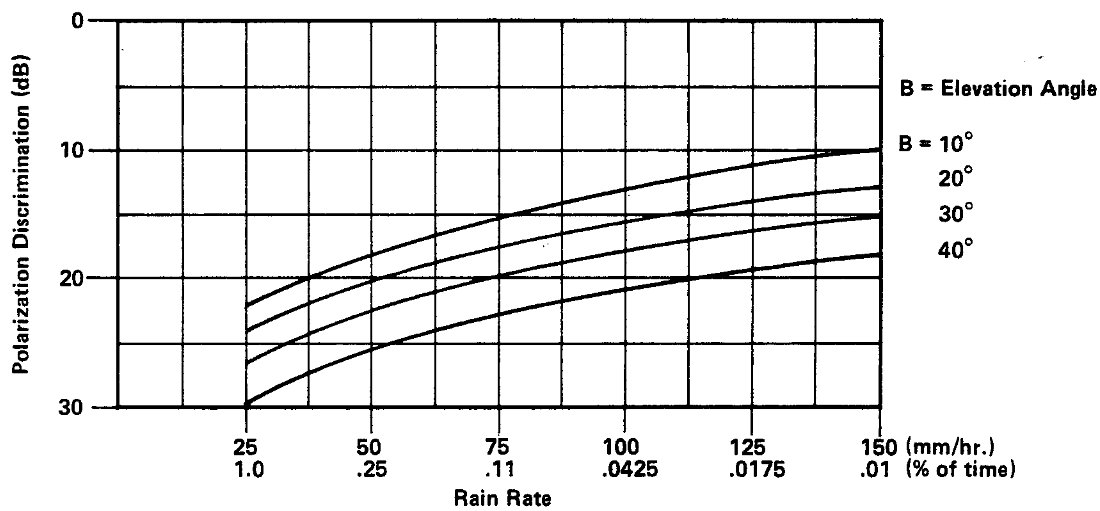


Figure 5. Reduction in Polarization Discrimination Attributed to Rainfall at 4 GHz (Constant Canting Angle = 30°)

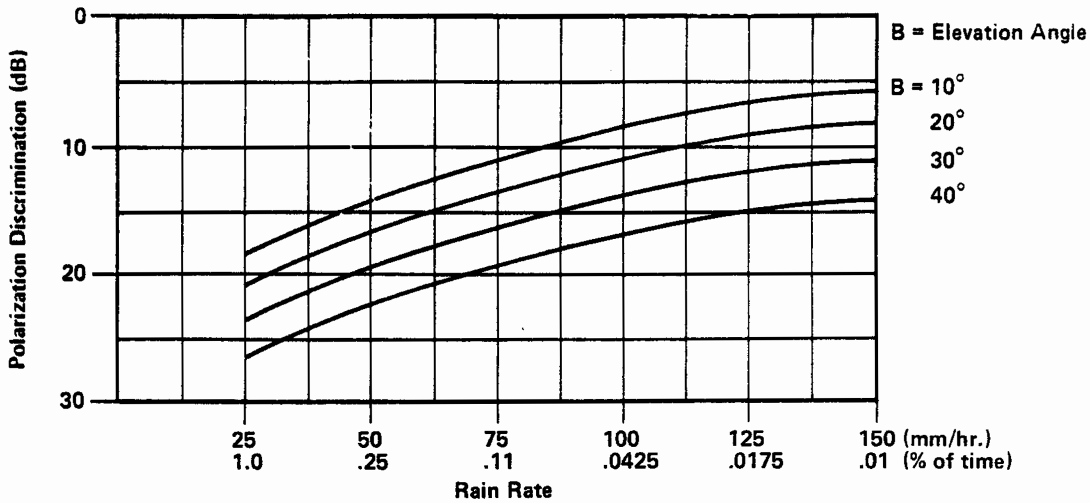


Figure 6. Reduction in Polarization Discrimination Attributed to Rainfall at 6 GHz (Constant Canting Angle = 30°)

Compensating for Depolarization

Beyond the initial system design, adaptive control systems can be used if necessary to automatically correct for signal depolarization resulting from Faraday rotation and rainfall. Analytical models can predict the behavior of the ionosphere and the resultant Faraday rotation that would be produced. Controls can be programmed to adjust the signal polarization accordingly. A more accurate method is to monitor a beacon signal from the satellite and generate error signals based on the amount of detected depolarization. The error signals drive compensation polarizers in the reuse feed. The same information can be used to "precompensate" the uplink signal. In the future, satellite telemetry data may contain feedback to assist the generation of accurate uplink compensation control voltages.

A quarter-wave polarizer can compensate for the rain-induced differential phase shift at 4 and 6 GHz in either linearly or circularly polarized signals. Differential attenuation caused by rain is negligible at 4 to 6 GHz, but can be substantial between 10 to 35 GHz. Differential attenuation applies only to linearly polarized signals and destroys the orthogonality of the wave. If desired, a compensating attenuator may be used in the feed to restore the orthogonality of the signal.

References

1) P.J. Wood, "Depolarization with Cassegrainian and Front-Fed Reflectors," Electronics Letters, pp. 181-183, March 26, 1973.

2) T. Murakami and G.S. Wickizer, "Ionospheric Phase Distortion and Faraday Rotation of Radio Waves," RCA Review, pp. 475-503, September 1969.

3) T.S. Shu, "Rain-Induced Cross-Polarization at Centimeter and Millimeter Wavelengths," Bell System Technical Journal, pp. 1557-1579, October 1974.

4) S. Parekh, "Frequency Reuse - Cross-Polarization Discrimination and Feed Components," Earth Station Symposium '78, November 8-10, 1978.

Earth Station Geometry

F.M. Fonda

Introduction

The subject of earth station geometry deals with the aiming of a very narrow pencil beam of the antenna towards a geostationary satellite.

The fraction of hemisphere area of sky covered by the pencil beam of an earth station antenna is quite small. The fraction is:

$$\left(\frac{1}{\frac{1}{1-\cos \phi /2}}\right)$$

where ϕ is the half-power beamwidth of the antenna.

A 5-meter antenna having a nominal beamwidth of 1° at 4 GHz covers one part in 26,000 of the hemisphere. The 10-meter coverage is one part in 104,000. Thus, locating a satellite requires accuracy in positioning relative to known geographical coordinates. Reliable communication performance requires that the antenna mount structure have sufficient rigidity to maintain this position under operating environments.

Satellite Earth Geometry

Figure 1 is a space view showing some of the domestic satellites at their respective longitudinal positions. The orbital radius of each satellite is equal to 6.611 earth radii. At this distance the period of orbital rotation for a satellite is equal to the period of earth rotation. The orbital plane of these satellites passes through the equatorial plane of the earth. Because of these two constraints--a circular orbit about the earth with a radius of 6.611 earth radii and the plane of the orbit lying in the equatorial plane of the earth--the satellites appear stationary with respect to any point on the earth.

Antenna Positioning Systems for Geostationary Satellite Coverage

There are two broad classes of positioning systems used for satellite earth station antennas. One class consists of orthogonal two-axis configurations; the other is the one-axis or single-axis configuration. The two-axis systems are characterized by the orientation of the lower-most axis with respect to the ground. Thus, a two-axis system having its lower axis perpendicular to the ground (Figure 2a) is called "elevation-over-azimuth."

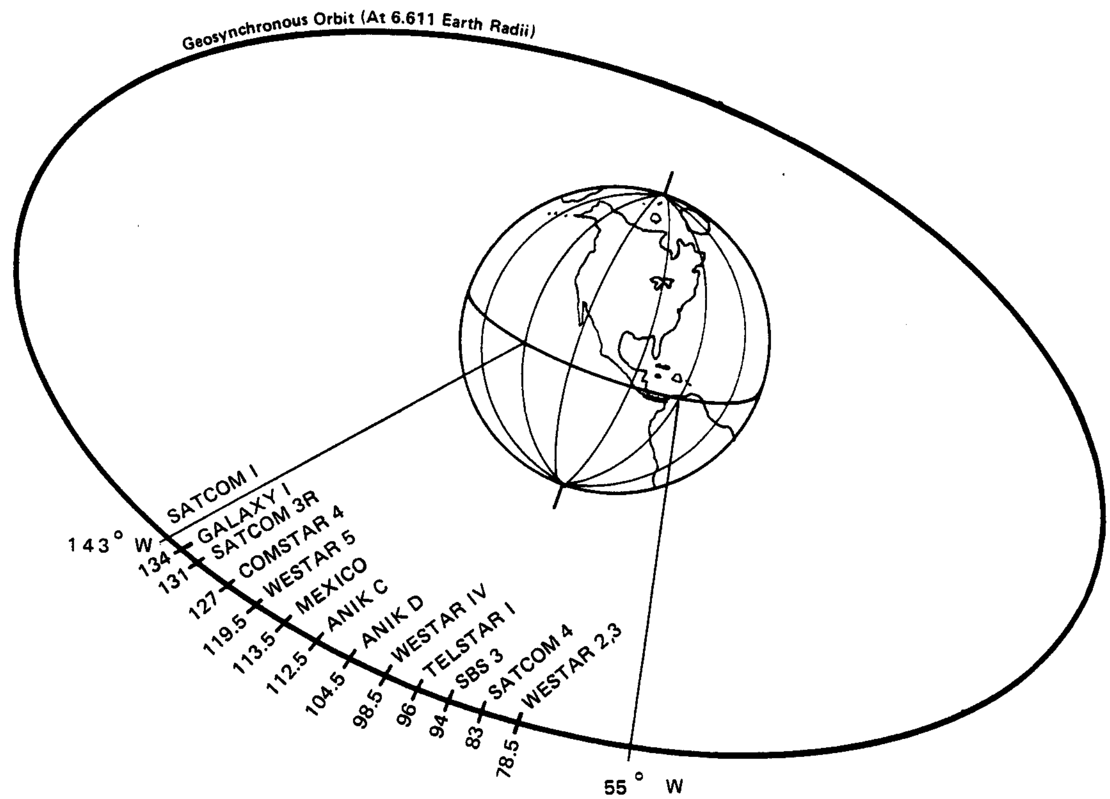


Figure 1. Satellite-Earth Geometry

One that has its lower axis parallel to the ground (Figure 2b) is called "X-Y ." One that has its lower axis parallel to the earth's axis of rotation (Figure 2c) is called "hour angle-declination" ("HA-DEC") or "polar." Each of the three positioning systems has the beam-axis or pointing direction perpendicular to the upper axis. Providing there are no physical rotational limitations, all three types can theoretically point in any direction.

Development of single-axis antenna mounts was brought about by efforts to reduce the costs and increase the versatility of earth stations. For a comparison of characteristics, consider a hypothetical X-Y mount of Figure 3b installed in Atlanta, Georgia. The lower axis, which is parallel to the ground, is placed normal to the azimuth direction 176.69 degrees. In this orientation the beam axis can traverse from any satellite to any other satellite between 87° and 135° west with only 0.8 degrees change in the X1 axis. Alternatively, with the X1 axis fixed, the maximum elevation error is ±0.4 degrees. The two-axis mounts of the last example can obviously be replaced by the single-axis system shown in Figure 3a, provided the inclination and orientation of the single axis is the same as the upper axis of the previously described in X-Y mount. The single-axis system can thus scan any satellite between 70° and 135° west with the same error since the two systems are identically equivalent.

The declination-corrected single-axis mount, shown in Figure 4, can be considered an adaptation of the polar mount. The axis of the declination-corrected mount is inclined along a north-south line like the polar mount, but the angle of inclination is slightly different. Unlike the single-axis system of figure 2b, which has the beam axis normal to the axis of rotation, the declination-corrected system has the beam direction depressed a fixed angle from perpendicular so that the axis of the antenna beam sweeps along the surface of a cone centered about the axis of rotation. It permits a predetermined arc of the geosynchronous orbit to be scanned with minimum pointing error. It will be shown that less than 0.05 degrees pointing error results in pointing to any satellite in the 70- to 135-degree synchronous arc from any position in CONUS (contiguous United States).

The following sections will discuss aspects of these various positioning systems and show how they solve the satellite-viewing geometry problem.

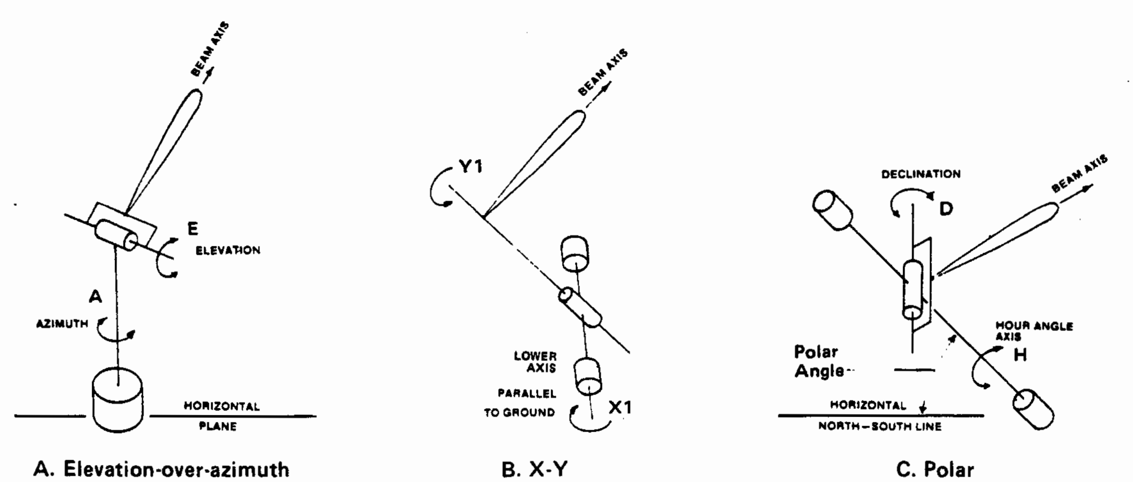


Figure 2. Schematic Illustrations of Two-Axis Earth Station Positioning Systems

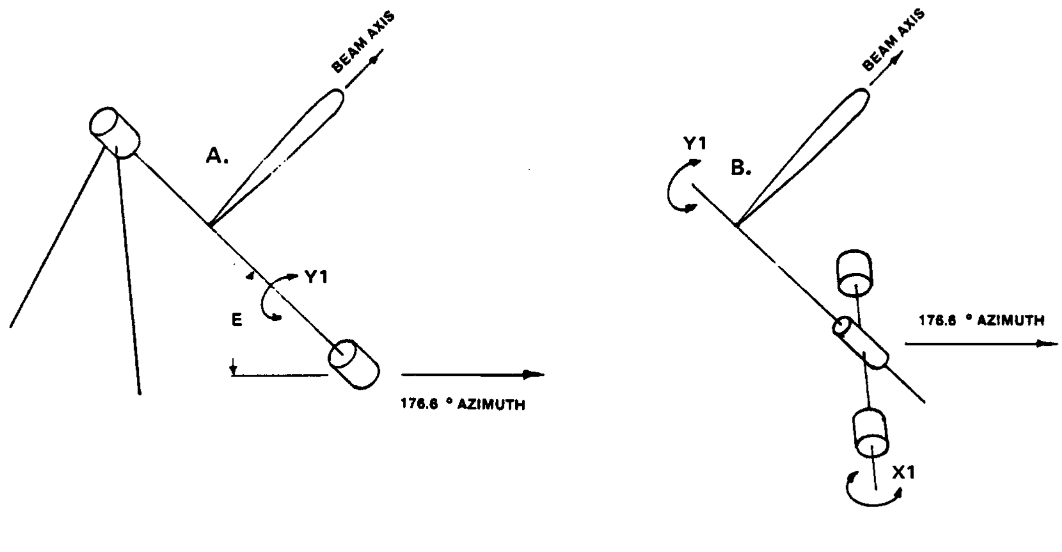


Figure 3. Comparison of Single-Axis Antenna Mount (a) with X-Y Mount (b) Having Same Orientation

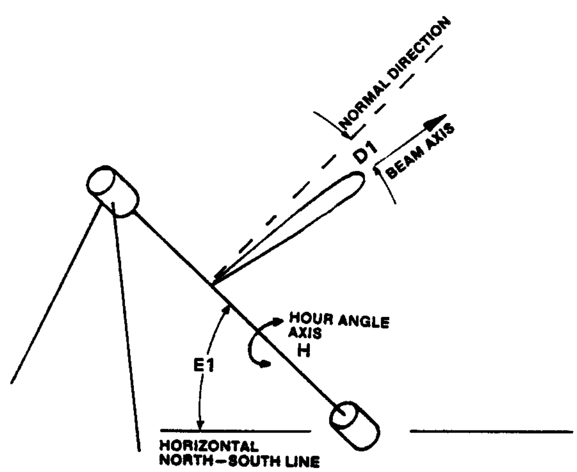


Figure 4. Declination-Corrected Single-Axis Mount

Elevation-over-Azimuth Pointing System

A universal method of describing the direction of a point on earth as well as points in space is by using the elevation-over-azimuth coordinate system. Azimuth is defined as an angle produced by rotation about an axis which is perpendicular to the local horizontal plane. The zero reference for measuring true azimuth is north. East is then 90°, south is 180° and west is 270°. Elevation is defined as an angle produced by rotating a line about an axis which lies in the local horizontal plane. In an elevation-over-azimuth pointing system, an azimuth rotation causes the elevation axis to rotate in the local horizontal plane. If the beam axis of an antenna is considered attached and normal to the elevation axis, an elevation rotation will always cause the beam axis to rotate in a local vertical plane. When the beam axis is parallel to the ground, the elevation angle is zero. A 90° elevation rotation points the beam to the zenith.

Since gravity level is the reference for elevation in this two-axis system, a simple bubble-level protractor or inclinometer can be used to establish the correct elevation angle to a particular satellite. Then, an azimuth sweep can be made to point the antenna beam at the satellite. Figure 5 shows elevation-over-azimuth geometry for geosynchronous satellite sighting. The azimuth axis of the mount is shown perpendicular to the local horizontal plane. An extension of this axis intersects the center of the earth. The elevation axis is parallel to the local horizontal ground plane. The elevation angle of the line-of-sight is measured in a local vertical plane.

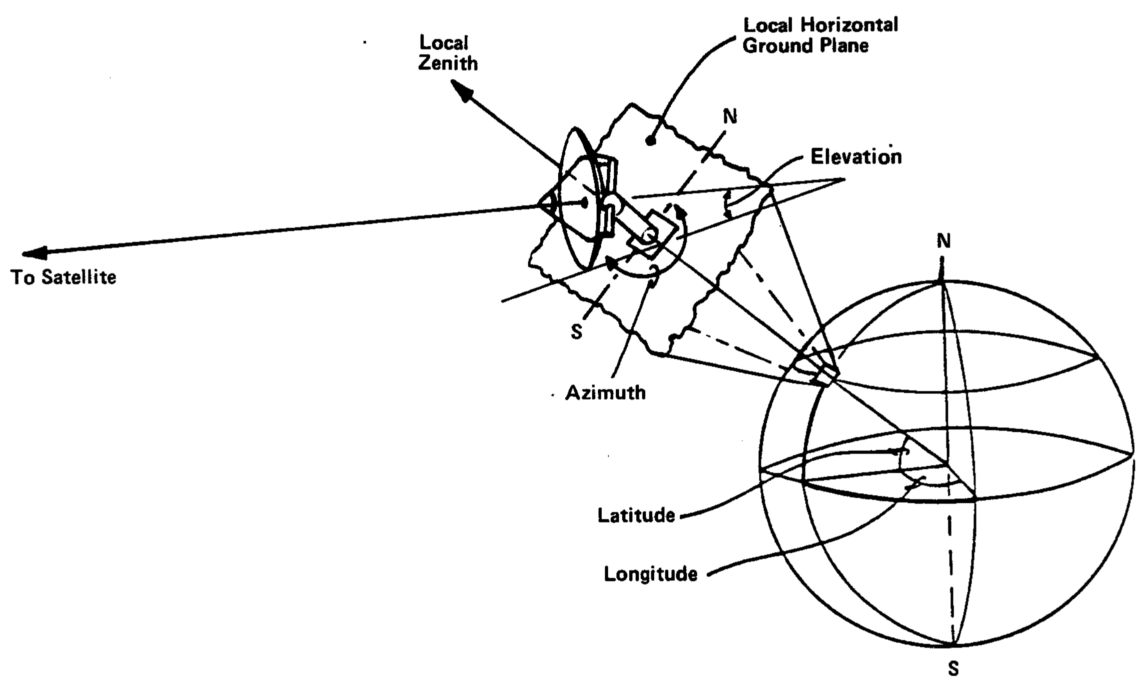


Figure 5. Elevation-over-Azimuth Geometry

The appendix gives a derivation of the elevation and azimuth angles to a satellite and also shows the geometry of these angles in more detail.

The initial installation of an elevation-over-azimuth mount is not critical. The azimuth axis should be nearly vertical so that an azimuth sweep results in a minimum change in elevation angle. The Scientific-Atlanta 5- and 10-meter mounts, if installed with their three feet within 1/8 inch of being level, will have an elevation pointing error of less than 0.1 degree for a 50-degree change of azimuth position. Minimum and maximum elevation angles for covering satellites between 70 and 135 degrees west longitude are given in Figure 6 for the United States. Note that except for the northeastern New England states and the northwest tip of Washington, the minimum required elevation look angle is greater than 15 degrees. The numbers enclosed in boxes and located every 5 degrees in latitude between 25° and 45° are the maximum elevation angles at those latitudes. These maximum elevation angles will occur when the satellite and the site are at the same longitude.

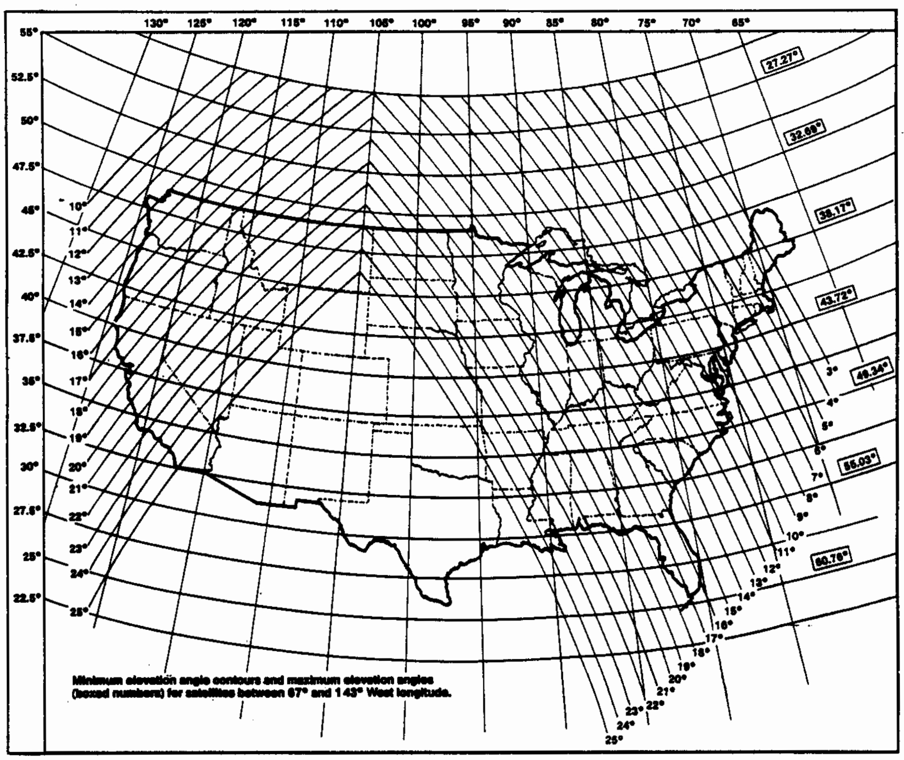


Figure 6. Minimum Elevation Angle Contours for Satellite between 67° and 143°

The azimuth center heading of the mount should be selected to ensure coverage of satellites located between 70 and 135 degrees west longitude. Figure 7 gives the total azimuth range necessary in the United States for this coverage. Note that the required range increases for lower latitudes. Figure 7 also shows the azimuth center heading of the positioner for equal clockwise and counter-clockwise pointing angles to cover the 70- to 135-degree orbital arc.

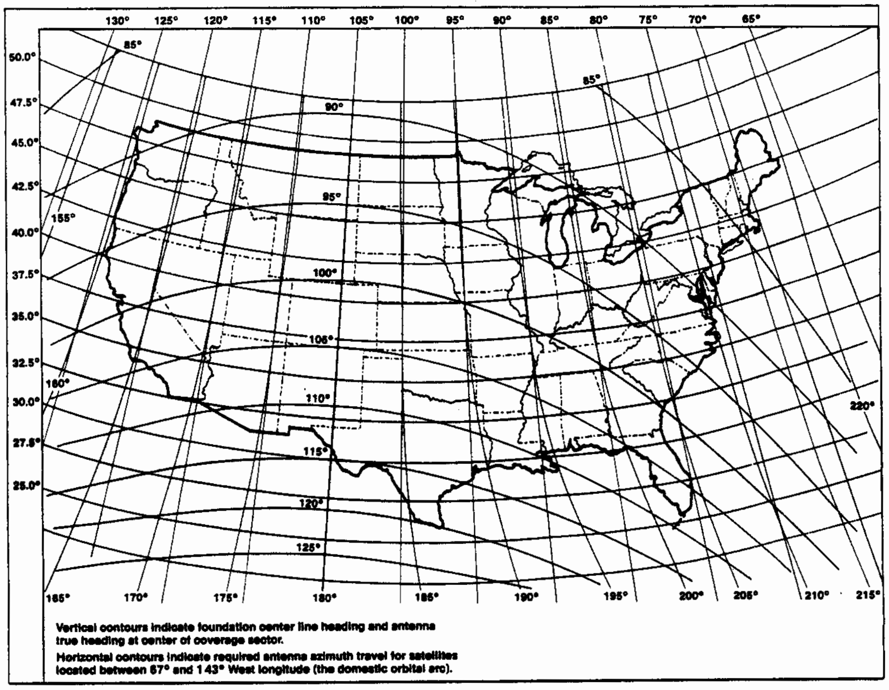


Figure 7. Total Azimuth Angle Contours of Satellites between 67° and 143°

Figure 8 is a graph of azimuth and elevation angles versus a particular site latitude and the longitudinal difference between the satellite and the site. The horizontal and vertical rectangular coordinates are site latitude and difference longitude, respectively. The curved lines running toward and labeled at the top of the graph are the required azimuth angles (add to 180 degrees if satellite is west of site, subtract from 180 degrees if satellite is east of site). The curved lines running toward and labeled at the left margin (down to 15 degrees) are required elevation angles. (The elevation lines of 10 degrees and below are labeled at the top of the graph.) To determine required azimuth and elevation angles, find the satellite site longitudinal difference and move vertically on this line until it intersects the horizontal site latitude line. At this intersection interpolate between bounding azimuth and elevation curves for the requiring angles. (Make the 180-degree azimuth correction as cited above.) This graph and its derivation is contained in the appendix.

The azimuth and elevation angles to a particular geostationary satellite can be calculated using the satellite longitude, Z; the site longitude, Y; the site latitude, X; and the below listed equations:

$$C = Z - Y$$

$$A(\text{azimuth}) = 180° + \tan^{-1}\left(\frac{\tan C}{\sin X}\right)$$

$$E(\text{elevation}) = \tan^{-1}\left(\frac{(\cos C \cos X) - 0.15126}{\sqrt{\sin^2 C + \cos^2 C \sin^2 X}}\right)$$

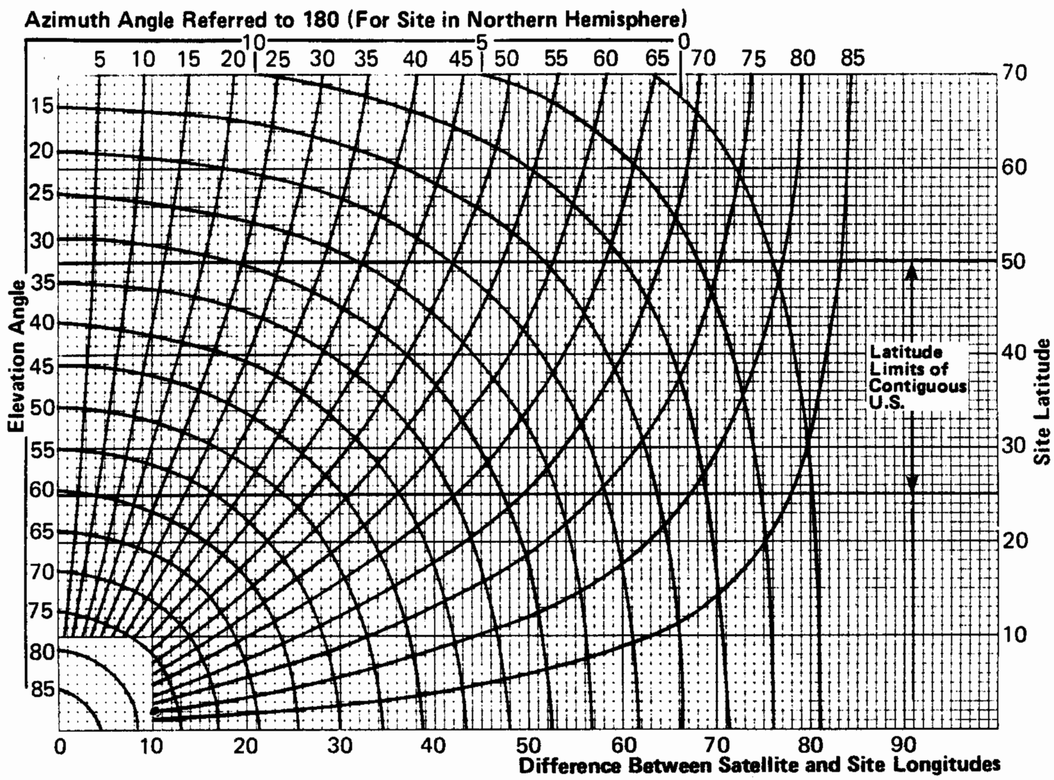


Figure 8. Universal Azimuth-Elevation Look Angles

The process of establishing the angular references for the earth stations angular readout system is called "boresighting." If the earth terminal foundation and anchor bolts could be installed to a known heading with no error, or if the installed foundation mount heading is surveyed accurately, then boresighting is simply a matter of referencing the readout of the mount axis with respect to the foundation. If the mount heading is not known with accuracy sufficient to directly calibrate the angular readout system, then peaking up on a known satellite and making a gravity-based angular measurement of an axis will permit system boresighting. To do this, a particular satellite must be first acquired as described below for the elevation-over-azimuth system:

a. A particular satellite's local azimuth and elevation angles are calculated based upon the site's longitude and latitude.

b. An inclinometer is set to the proper elevation angle. It is placed on the feed plate and the antenna is elevated until the bubble shows level.

c. The antenna is scanned in azimuth until the satellite is acquired and peaked.

d. The azimuth and elevation measuring system is set to read the correct azimuth and elevation angles to the satellite. At this point the earth station is boresighted.

X-Y Pointing System

The X-Y axes configuration is structurally less complicated than elevation-over-azimuth positioning support systems. The X-Y mount is sometimes called an azimuth-over-elevation mount. This is really a misnomer because azimuth is defined as an arc of the horizon. Figure 9 shows the lower-most axis, designated X1. This axis is parallel to the ground and in this example passes through the rear two feet of the mount. Rotation about the axis (by changing the length of the front foot or the lengths of the two rear feet) moves the antenna in elevation. The Y1 axis lies in a vertical plane which is perpendicular to X1 axis. The position of the Y1 axis in the vertical plane can range from vertical to horizontal depending upon rotation of the X1 axis. Note that the mount heading is designated A1. This angle, measured from north, is to the projection of the Y1 axis onto the horizontal plane. It is established during installation of the foundation.

The X-Y mount is generally less complex than an elevation-over-azimuth mount because the ground and the foundation provide direct support for the lower-most axis (the X1 axis). The lower-most axis (azimuth) of the elevation-over-azimuth mount is perpendicular to the ground, and so structure must be provided to support its upper end. This difference is offset somewhat in practical systems when ground clearance for the reflector is considered. The elevation-over-azimuth mount tends to raise the reflector as a direct consequence of establishing the supported vertical azimuth axis. The X-Y configuration requires attention to ground clearance, and compromises in satellite arc coverage frequently must be made.

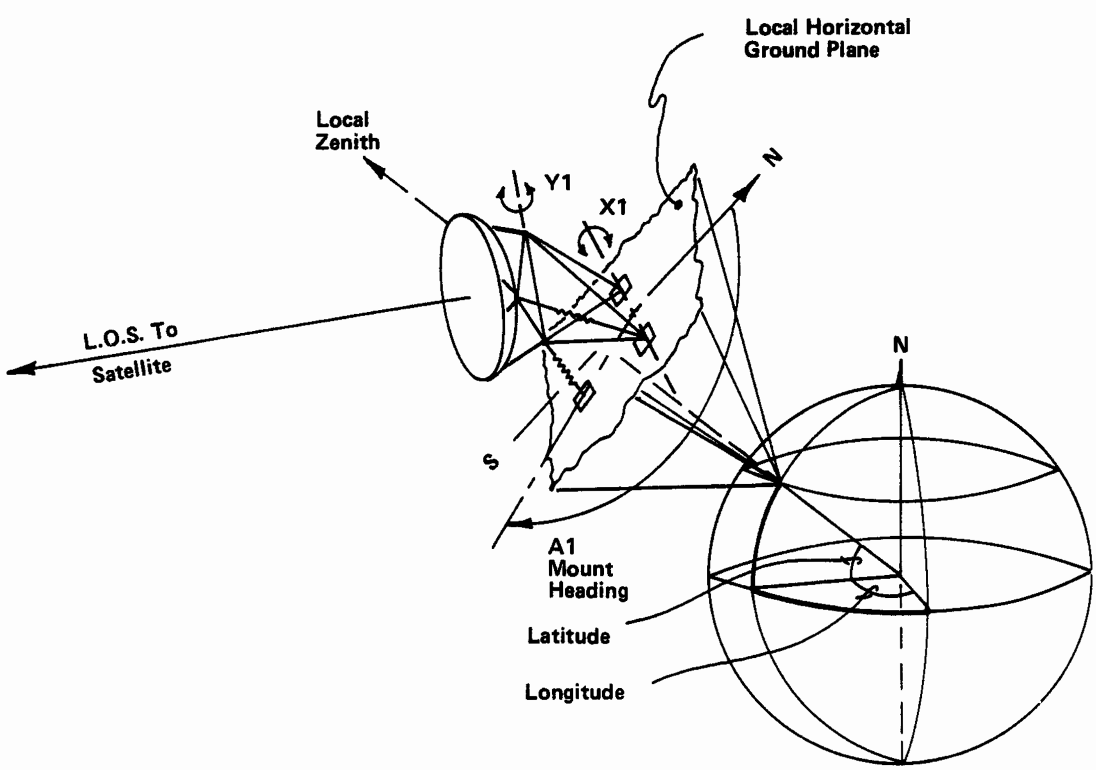


Figure 9. X-Y Mount Geometry

Each manufacturer of X-Y mounts has his own particular recommendation for establishing the mount heading during foundation pouring and anchor bolt setting. In general, the heading is chosen so that a specific satellite arc coverage can be obtained in a continuous sector. Whatever the recommendation, the actual installed heading must be determined accurately so that the system can be boresighted and the X1-Y1 angles calculated to position the antenna to various satellites. As is the case for the elevation-over-azimuth mount, the mount heading for the X-Y mount may be determined by accurate direct survey or by an indirect process using an acquired satellite and gravity-level-referenced angular measurement of one of the axes as starting points. This latter process is mathematically achieved as follows:

1) $A1 = A - A2$

Where: (A is satellite azimuth angle, A1 is mount-heading azimuth angle)

2) $A2 = \sin^{-1}\left(\frac{\sin Y1}{\cos E}\right)$; use inclinometer to measure Y1

(Y1 is upper-axis angle, E is satellite elevation angle)

or 3) $A2 = \cos^{-1}\left(\frac{\tan E}{\tan X1}\right)$; use inclinometer to measure X1

(X1 is lower-axis angle)

or 4) $A2 = \tan^{-1}\left(\frac{\tan Y1}{\cos X1}\right)$; use inclinometer to measure Y1 and X1

When A1 is determined, it then is a constant for the particular site in all future determinations of the X-Y mount angles X1 and Y1.

These angles can be determined for a particular satellite by first determining its elevation angle E and then proceeding as follows:

1) $X1 = \tan^{-1}\left(\frac{\tan E}{\cos A2}\right)$

2) $Y1 = \tan^{-1}(\tan A2 \cos X1)$

or 3) $Y1 = \sin^{-1}(\cos E \sin A2)$

For this discussion the following axis and angle convention is employed:

The X1 lower axis angle is measured the same way as elevation E; i.e., when the lower axis has the upper axis vertical the antenna is pointing to the horizon and X1 is 0. The Y1 upper axis angle is zero at peak ascension, and is positive toward the west and negative towards the east. For example, when X1 is 90°and Y1 is zero, the antenna points to the zenith.

A discussion of pointing and alignment techniques that relate to the X-Y mount is presented in Appendix A.

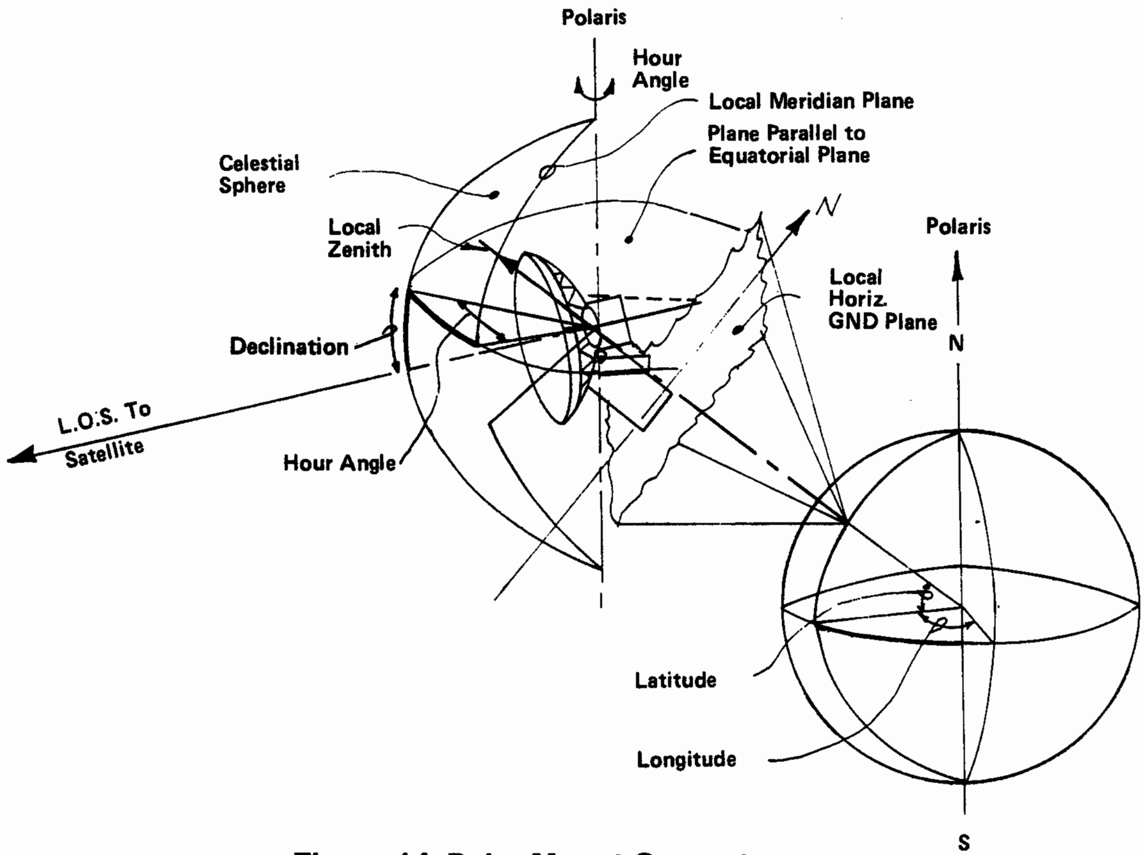


Figure 14. Polar Mount Geometry

Polar Mount Pointing System

A polar mount has two axes of rotation, hour-angle and declination. Figure 14 shows its geometry. The hour-angle axis is aligned parallel with the earth's axis. The hour angle is inclined in a north-south direction from local horizontal through an angle equal to the site latitude. Thus, this axis is parallel to the ground at the equator and perpendicular to the ground at either pole. When this alignment is achieved, it is possible by rotating about this axis at one revolution per day to keep the line-of-sight fixed at a point on the celestial sphere. Most astronomical telescopes have the polar axis configuration because of this characteristic. With zero declination angle (the antenna line-of-sight perpendicular to the hour-angle axis) sweeping about the hour angle axis could scan the antenna beam in a plane parallel to the plane through the equator. To point to a synchronous satellite on the equatorial plane, the antenna must be depressed in declination because of the finite orbital radius of the satellite. The amount of declination required is a function of (a) satellite longitude, (b) site longitude, and (c) site latitude. A polar mount located in Philadelphia for instance has a 0.41° required change in declination covering satellites between 70° and 135° west longitude. It is possible to reduce this required declination change by more than an order of magnitude by introducing a correction tilt to the hour-angle axis during initial system installation. The possibility then exists that such a system could point with negligible error to any satellite between 70° and 135° by rotating about one axis only. The hour-angle and declination angle-equations are given in the Appendix B.

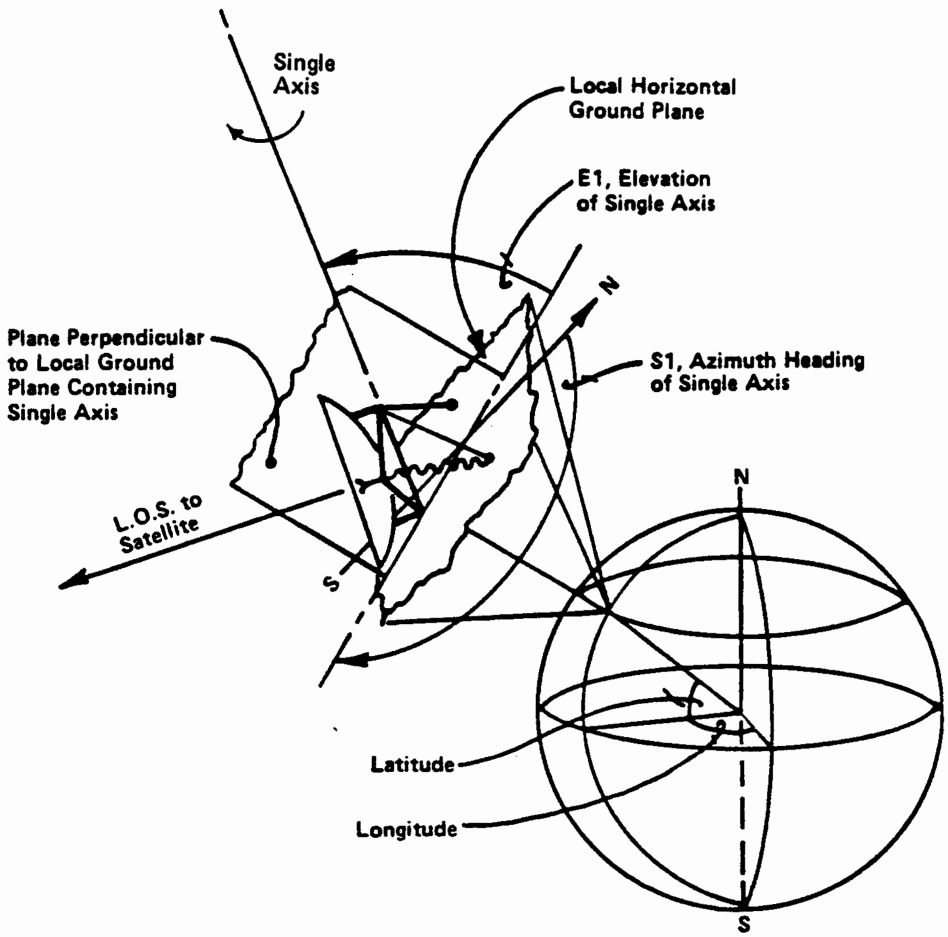


Figure 15. Single Axis Mount Geometry

Single-Axis Pointing System

Scientific-Atlanta has developed the least complicated of all positioning systems for satellite earth station antennas. This class of positioning system is defined by the term single-axis. There are two sub-groups of positioners within this class: the basic single-axis positioning system and the declination-corrected single-axis or DEC-corrected single-axis positioning system. A single-axis system which has the beam axis perpendicular to the axis of rotation may be considered to be an X-Y mount whose lower axis is fixed after the installation of a particular site. The pointing system is based on the fact that an axis of rotation which is normal to the plane passing through the station site and two satellite points will aim the antenna to the two satellites with zero pointing error. Figure 15 shows the geometry of the single-axis pointing system. The pointing errors for satellites between and beyond the two satellites is generally small. Figure 16 shows that the error spread for U.S. eastern satellites is less than 0.1° when establishing COMSTAR III and WESTAR IV for zero error. Figure 17 similarly shows an error spread of less than 0.1° for western satellites using SATCOM I and II for zero error. A good compromise for U.S. satellite coverage is shown in Figure 18. The two satellites chosen for zero error are COMSTAR I and II. All satellites from 87° to 135° are within 0.4° of the antenna LOS.

Unlike conventional HA-DEC mounts, Scientific-Atlanta single-axis mounts are relatively insensitive to errors in foundation heading. This is because the lower end of the axis is attached to the front foot with what is essentially a ball joint. The upper end of the axis is similarly supported by an equivalent ball joint connection to two supporting legs. These legs are telescoping and therefore allow the upper axis to pivot about the lower end. The proper axis orientation can be achieved by telescoping the two supporting legs. Furthermore, this orientation can be maintained even with considerable error in the location of the foundation mounting feet. Since the legs are telescoped to create the proper distance from the upper end of the axis to the displaced mounting feet, the mount compensates for foundation installation errors. It may be seen that there is no need to have a level foundation. Each of the three mounting feet are independent and need only to be placed such that (1) a stable structure results and (2) the two telescoping legs and the single-axis strut adjustor can be operated within their physical-length constraints.

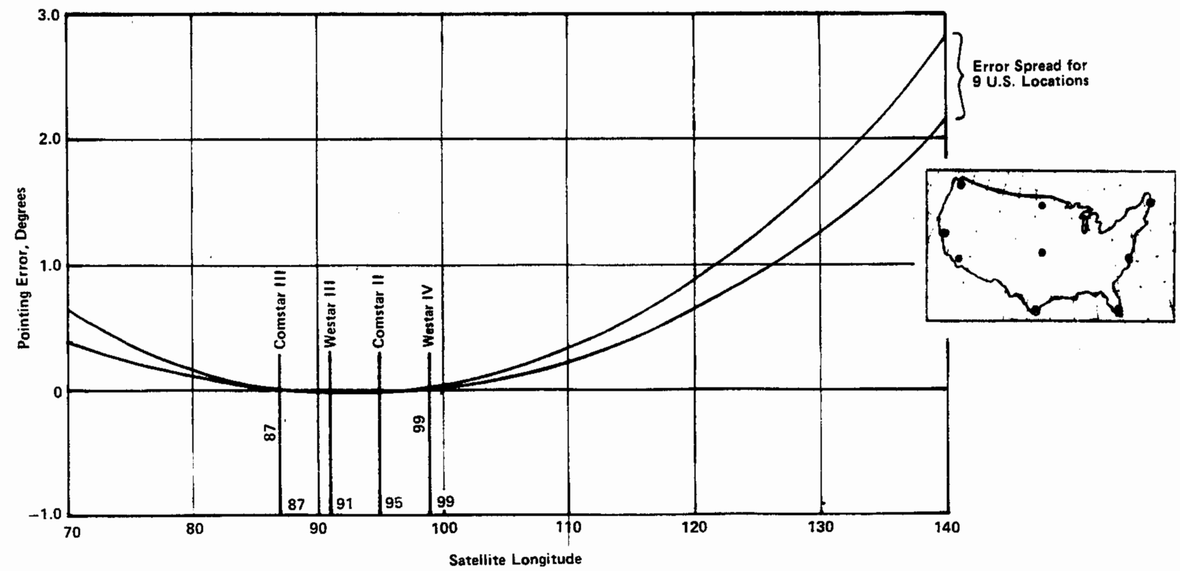


Figure 16. Pointing Error for U.S. Eastern Satellites (single axis zero error at 87° and 99°W longitude)

A convenient way to describe the orientation of the single-axis is to give its orientation in local azimuth and elevation angles. Figure 19 shows the single-axis geometry in a local coordinate system. A vertical plane is passed through the axis. The azimuth and elevation angles of the axis are measured by measuring the azimuth direction of this plane and the elevation of the axis above the horizontal.

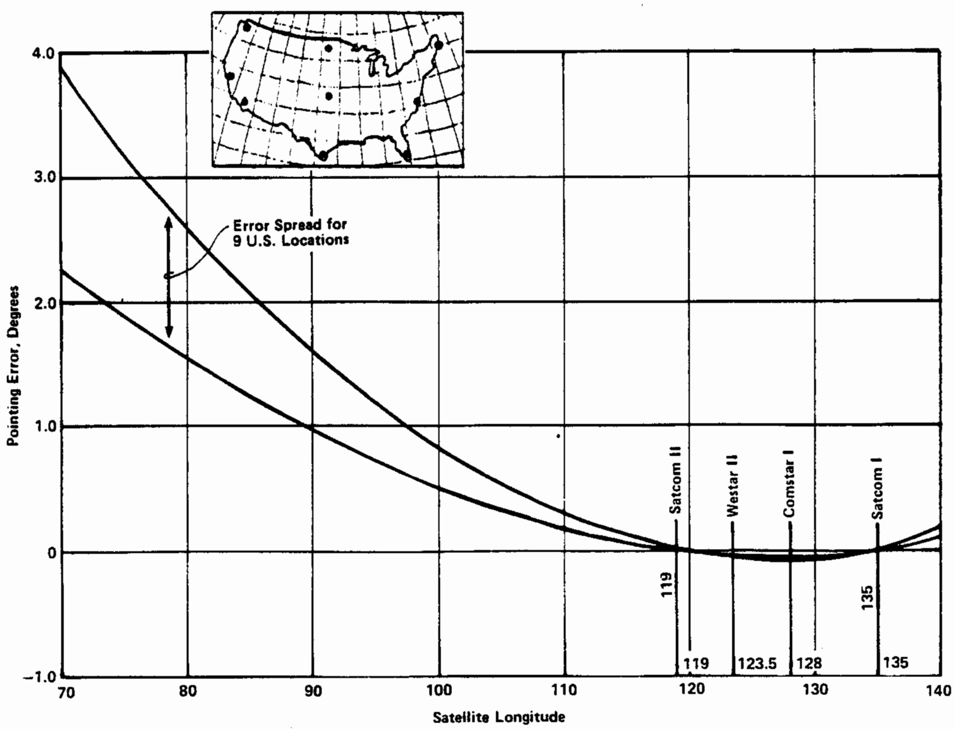


Figure 17. Pointing Error for U.S. Western Satellites (single axis zero error at 119° and 135°W longitude)

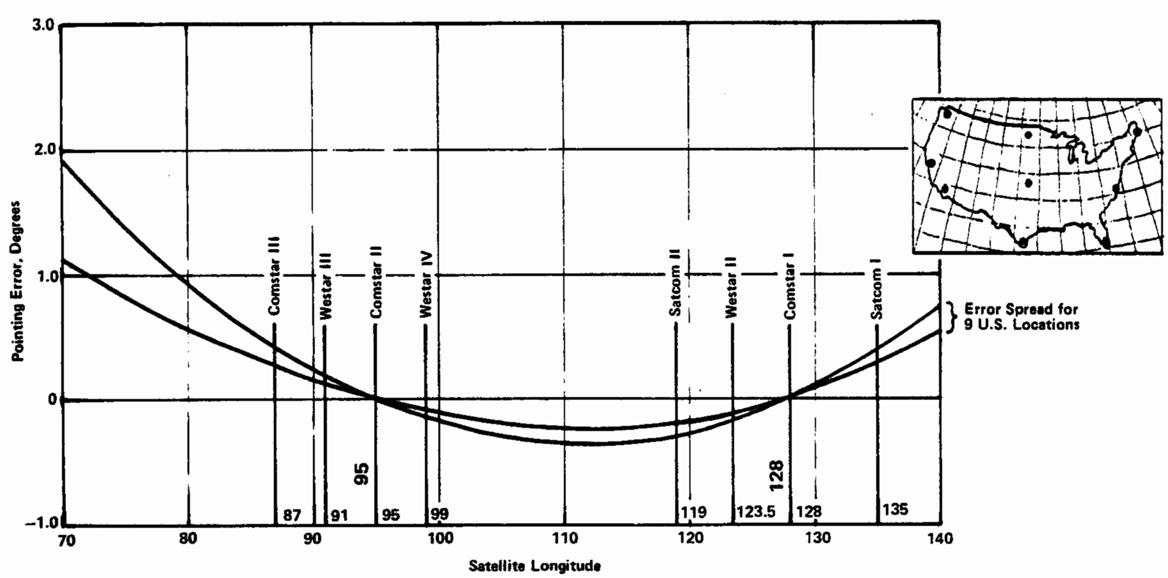


Figure 18. Pointing Error for U.S. Eastern and Western Satellites (single axis zero at 95° and 128°W longitude)

The single-axis pointing geometry can be described in terms of X-Y geometry. To do this, the following conventions are established:

1) The single axis is the Y1 axis.

2) This axis is elevated with respect to the horizontal through a fixed angle, E1. (E1 = 90 - X1)

3) If X1 = 90°, the axis lies on the ground.

4) With Y1 = 0, the antenna looks at the zenith.

5) The azimuth direction of the vertical plane containing the single axis is S1. (The bottom end of the axis is the pointing end of the plane.)

When two satellites with azimuth and elevation angles of A_1, E_1, and A_2, E_2, respectively, are selected for zero pointing error, S1 and E1 can be determined from the following equations:

$$S1 = A_1 - \tan^{-1}\left\{\frac{\cos(A_2 - A_1) - \left[\frac{\tan E_2}{\tan E_1}\right]}{\sin(A_2 - A_1)}\right\}$$

$$A2_1 = A_1 - S1; \quad A2_2 = A_2 - S1$$

$$X1 = \tan^{-1}\left(\frac{\tan E_1}{\cos A2_1}\right) = \tan^{-1}\left(\frac{\tan E_2}{\cos A2_2}\right)$$

$$E1 = 90 - X1$$

The Y1 look angles to the satellite can be determined by:

$$Y1_1 = \tan^{-1}(\tan A2_1 \cos X1)$$

$$Y1_2 = \tan^{-1}(\tan A2_2 \cos X1)$$

Establishing the proper orientation of the single axis is most conveniently done using a plumb line, a horizontal line and a site compass. The horizontal line is aligned along the proper azimuth heading using the site compass. The two rear legs are adjusted so that the plumb line dropping from the reference hole in the single-axis structural member of the mount just touches the horizontal line at the proper distance. The telescoping legs are drilled and bolted at this position, and proper installation is completed. If there is any uncertainty in the actual azimuth heading, S1 of the installed single-axis mount, the same procedure described for the X-Y mount, may be used to boresight the system with a single exception. The single-axis X1 angle (90 - E1) is fixed because there is only one X1 angle which provides zero error at the two selected satellites for a given installation. Therefore, the rear legs are differentially adjusted to correct the heading error. The single axis is rotated through an angle Y1 a corresponding amount shown in Figure 11 to correct for the heading error, A2.

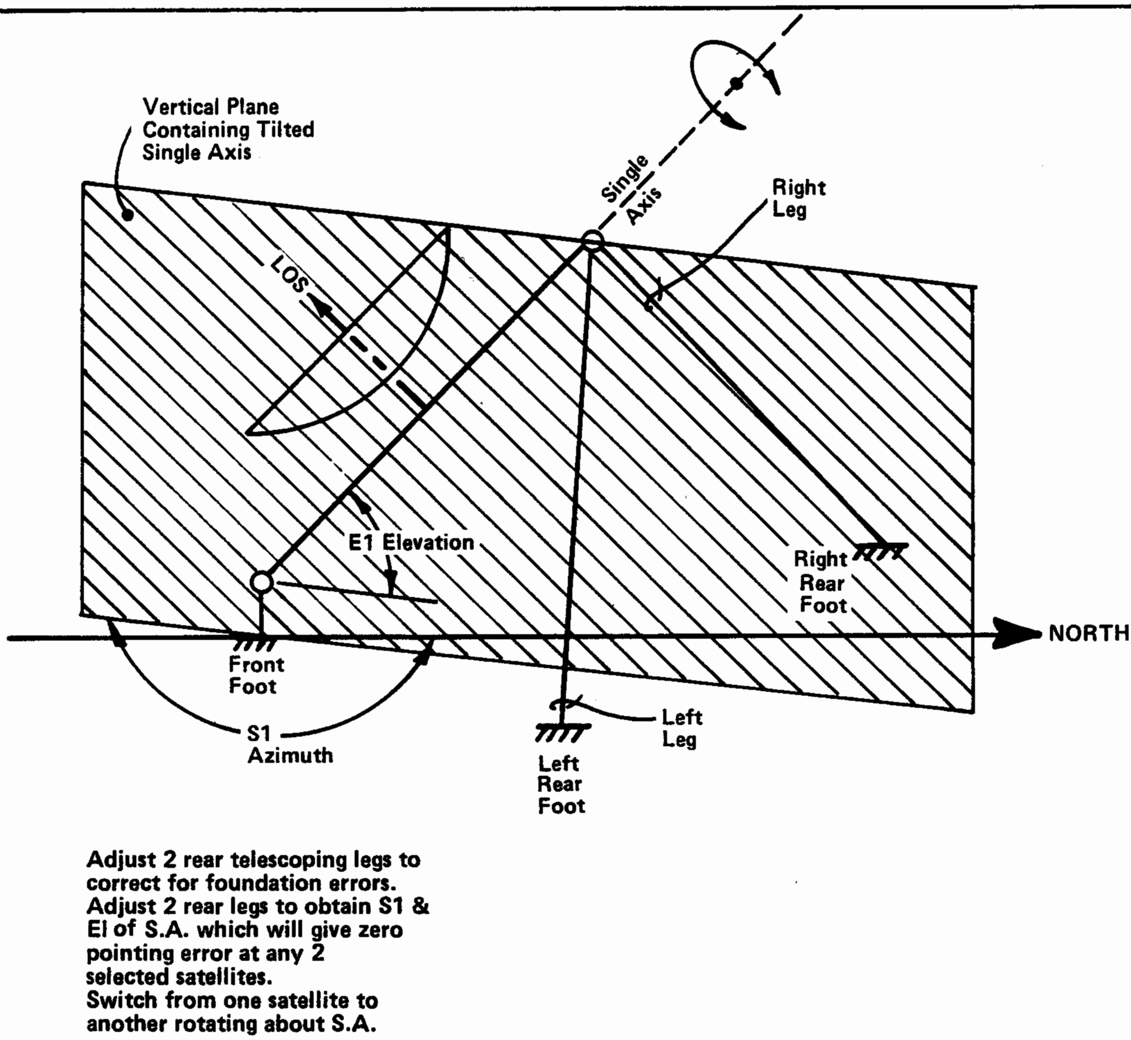


Figure 19. Single Axis Geometry

Declination-Corrected Single Axis Pointing System

Figure 21 gives the geometry of the declination-corrected single-axis pointing system. The angle, D1, reflector tilt is similar to the declination angle, D, of a polar mount. The single-axis elevation angle E2 is similar to the polar angle of the polar mount. The single axis is aligned in a true north-south plane. With angles D1 and E2 properly established (during initial installations), the declination-corrected single-axis mount will position the antenna to within 0.005 degree of the geosynchronous orbital arc for the range of satellite positions from 70° to 135° west longitude for earth stations located anywhere in the continental United States. Figure 22 is an error-spread plot of this pointing system for latitudes of 25° and 50°. The correction factor, K (used in pointing equations presented later), is a single value for this range of latitudes. (A different value of K is used for latitudes below 25° and above 50°. The resulting error plots are similar and show no greater spread.) This error plot is symmetrical about zero LHA (Local Hour Angle); therefore, it is equally applicable for positive LHAs required at eastern sites and negative LHAs required at western sites.

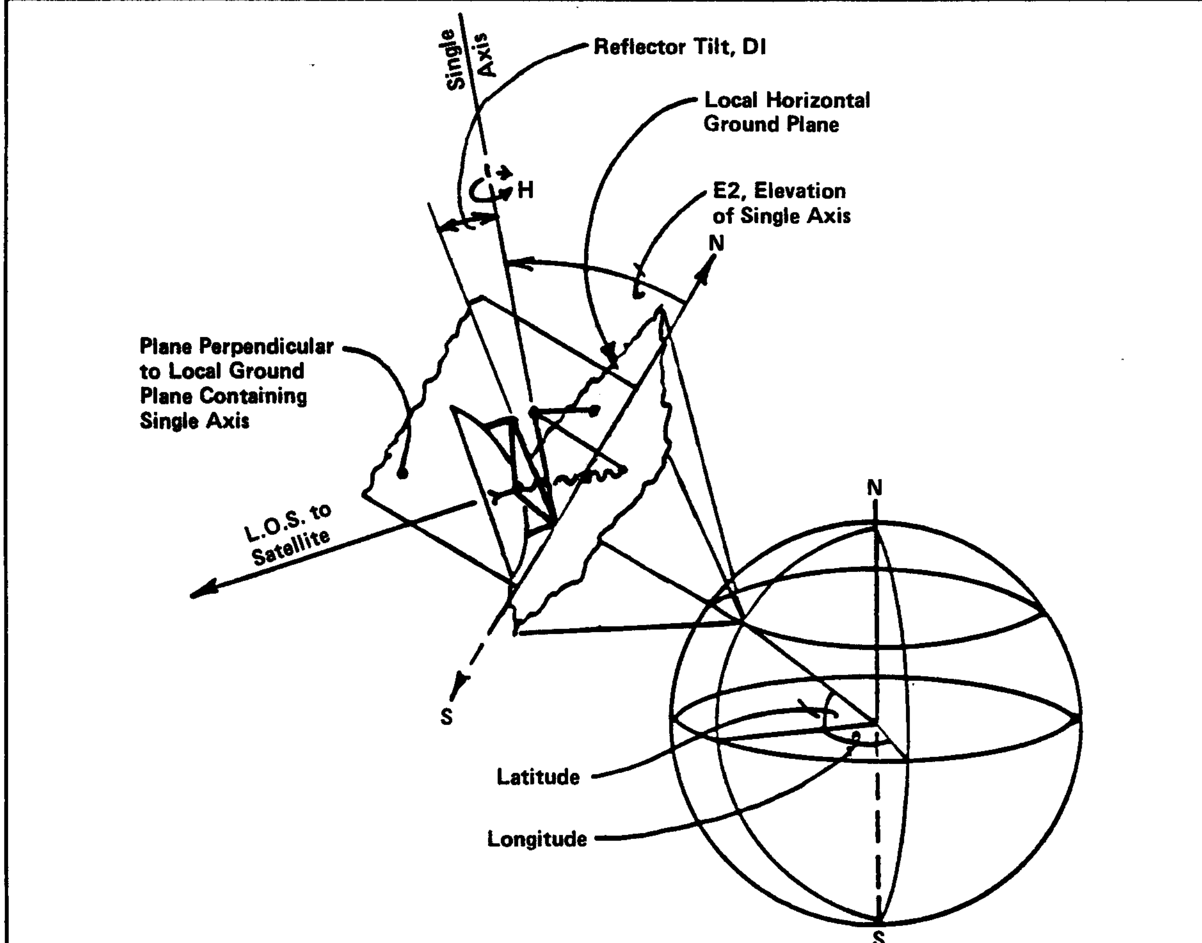


Figure 21. Declination Corrected Single Axis Mount Geometry

The single axis is aligned in a true north-south plane regardless of the site location. It is theoretically possible to reduce even further the pointing error by aligning the S.A. at a heading similar to the noncorrected single-axis system previously described. However, the error associated with north-south alignment is practically zero already, and there is an advantage in the uniformity of installations all having the same heading.

The installation equations for the declination-corrected single-axis mount are:

$$D1 = K + \tan^{-1}\left(\frac{\sin x}{6.611}\right)$$

$$E2 = X - D1 + \tan^{-1}\left(\frac{\sin X}{6.611 - \cos X}\right)$$

The pointing angle about the single axis for any particular satellite is nearly the same as LHA for a HA-DEC mount:

$$H \approx \tan^{-1}\left(\frac{\sin C}{\cos C - 0.15126 \cos X}\right)$$

The declination-corrected single-axis positioning system has the same foundation inaccuracy tolerance as does the single-axis positioner previously described. The single axis of rotation is supported by the equivalent of ball joints. The two supporting rear legs are telescoping to allow achieving the proper single-axis elevation angle, E2, and at the same time, compensate for foundation anchor bolt placement errors. This physical design allows the careful setting of the single axis to well within 1 arc minute of the required north-south heading and the proper elevation angle, E2. When this installation alignment has been completed, the telescoping legs are drilled and bolted. The system then can sweep the entire U.S. assigned satellite orbital arc about the single axis with essentially zero error.

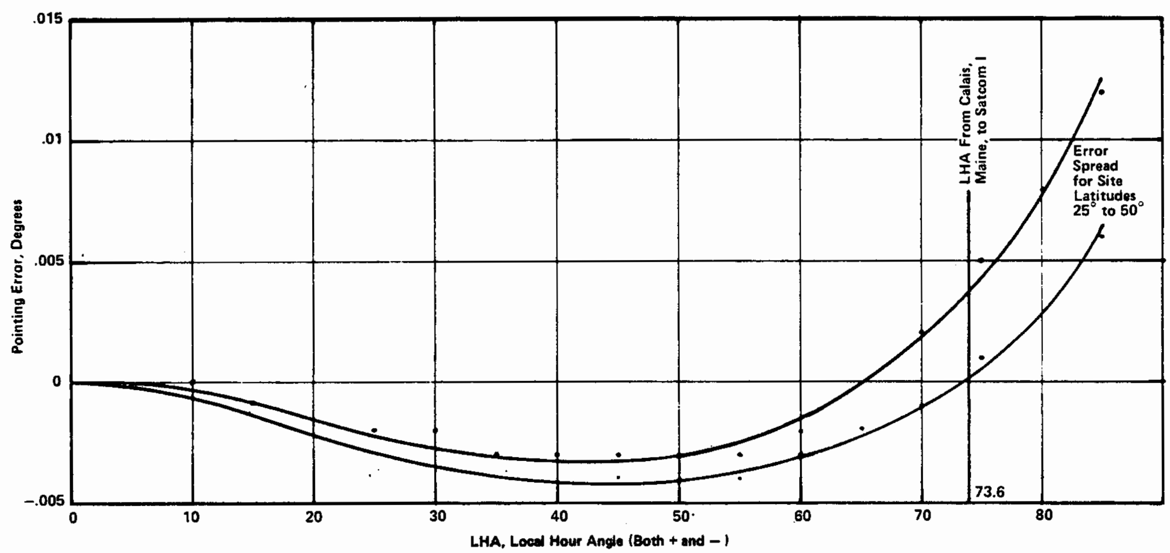


Figure 22. Pointing Error for Declination Corrected Scientific-Atlanta Mount

Several benefits accrue. The reliability of those installations which are remotely positioned is doubled over two-axis systems. The entire range of U.S. satellites is covered in one continuous sector. The polarization of the antenna is maintained constant with respect to the orbital plane and is unaffected by antenna position. As a consequence, no polarization peaking of the antenna feed is required for those satellites which maintain their polarization vectors in or orthogonal to the orbital plane. Satellite acquisition is greatly simplified because the search is one dimensional instead of two. Rotating the antenna about the single axis either in an easterly or westerly direction will intercept the next geosynchronous satellite to the east or west of the present position.

APPENDIX A

Pointing and Alignment Techniques

When an X-Y mount is boresighted using the indirect satellite reference procedure, the extent to which the mount heading is unknown can add to satellite acquisition difficulties. The following procedure will aid this initial satellite acquisition process.

1) Select a satellite which has an azimuth angle A as close as possible to the estimated mount heading A1 so that A2 = (A - A1) is as close to zero as possible.

2) Calculate X1 and Y1 using the best estimate of A2. Adjust the X and Y axes to the calculated X1, Y1 angles using the mount's degree scales, if supplied, or an inclinometer.

3) Scan through an angle of ±ΔY1 corresponding to the uncertainty in A2. Correct X1 as necessary during this scan so that the acquisition beamwidth is not exceeded by the cross-coupled motion in elevation. Figures 10, 11, 12 and 13 are provided to give perspective to this procedure.

Figure 10 is a plot of Y1 as a function of satellite elevation, E, and satellite-mount heading azimuth difference, A2. It can be seen that the amount of Y1 rotation required for a given A2 is always less than A2. This graph can aid the satellite acquisition process. An example is given at the end of this section using Figure 12.

Figure 11 is a plot of ΔX1 = X1 - E, as a function of A2 and E. It can be used to roughly determine the required mount X1 angle by moving vertically along the particular A2 value until the proper E curve is reached and then horizontally to read ΔX1. The proper value of X1 is then ΔX1 + E.

The purpose of Figure 11 is to indicate the amount of change in X1 as a function of a change in A2 to aid in the initial acquisition of the satellite when there may be an error in mount heading.

The following example will explain the use of these figures as aids in boresighting an X-Y mount.

For this example, the nearest satellite is 18.5 degrees away from the assumed mount heading. The mount heading can be off as much as ±5 degrees. Therefore, 13.5°<A2<23.5°. Elevation angle, E, to this satellite is 40.0°.

Refer to Figure 12 to determine range of values of Y1 for sweep search. Figure 12 is made from Figure 10 as follows:

1) Draw horizontal reference line at elevation E = 40°.

2) Determine the location of A2 = 18-1/2° between the A2 curves of 15° and 20° on the line E = 40°.

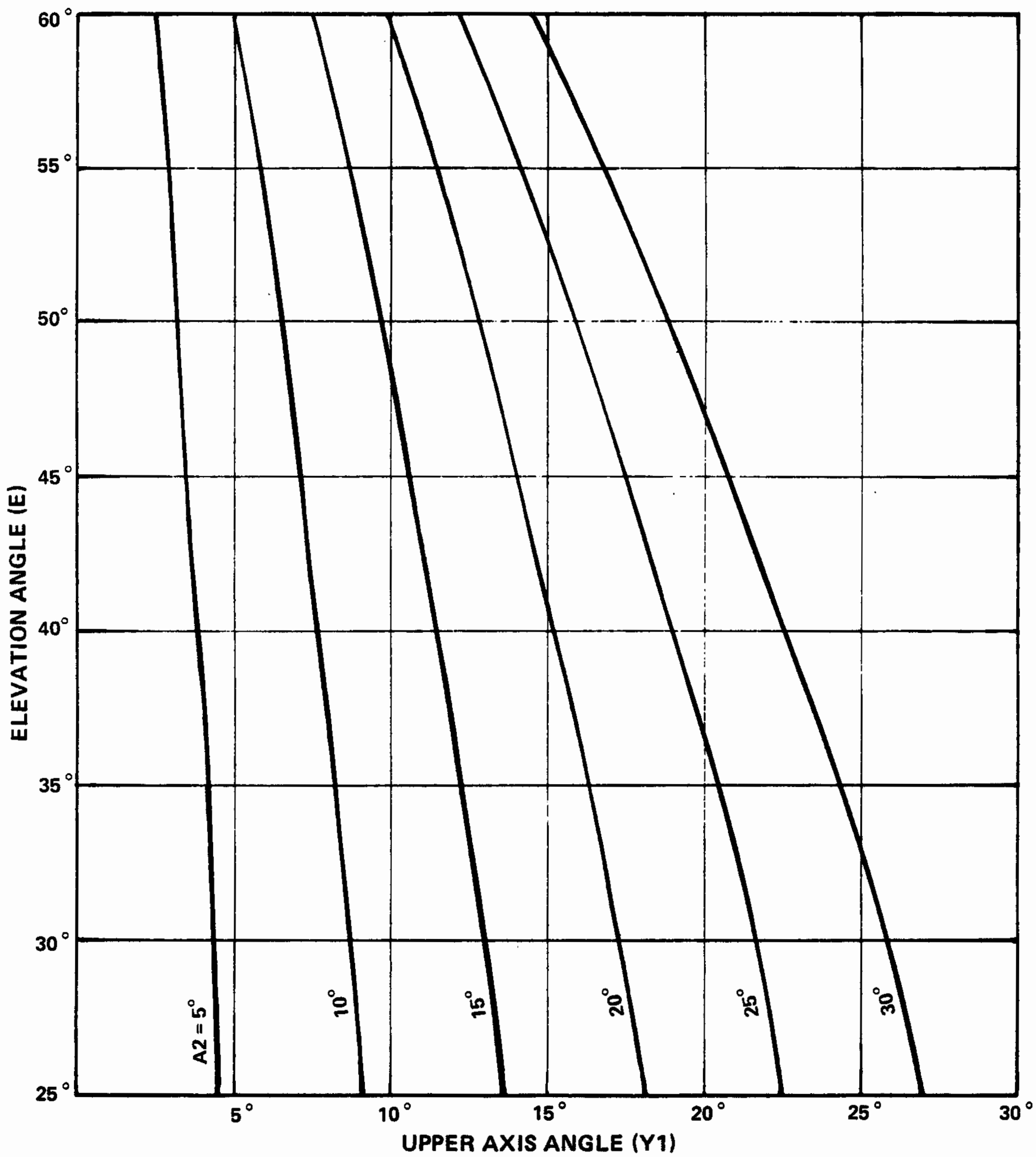


Figure 10. Y1 = f (A2,E) for an X-Y Pointing System

3) Also locate 18 1/2° ±5° = 13-1/2° and 23-1/2° on the line E = 40°.

4) Drop vertical lines from the three values of A2 down to the Y1 scale and read corresponding Y1 values. (For A2 = 13-1/2°, Y1 = 10-1/4°; for A2 = 18-1/2°, Y1 = 14°; for A2 = 23-1/2°, Y1 = 17-3/4°)

Thus, for the nominal A2 value of 18-1/2°, the corresponding Y1 is 14°. The algebraic sign of Y1 is the same as A2. If the satellite is east of the foundation heading, A2 is (-) and Y1 is (-). Note that the differential travel of the Y axis is ±3-3/4° for accommodating the A2 uncertainty of ±5°. A Y-axis search for the satellite under these circumstances should be limited to ±3-3/4° about the nominal Y1 value of 14°. See ①.

Refer to Figure 13 to determine the corresponding changes in X1. The elevation angle of 40° falls within the range of the E = 30° to 55° curve in Figure 10. Therefore, just this curve is reproduced in Figure 13. The change in X1 angle is determined as follows:

1) Locate A2 value of 13-1/2°, 18-1/2°, and 23-1/2° on the horizontal A2 scale, Draw vertical lines from these points until they intercept the E = 30° to 55° curve.

2) From the intercepts on the curve, draw horizontal lines to intercept the ΔX1 scale.

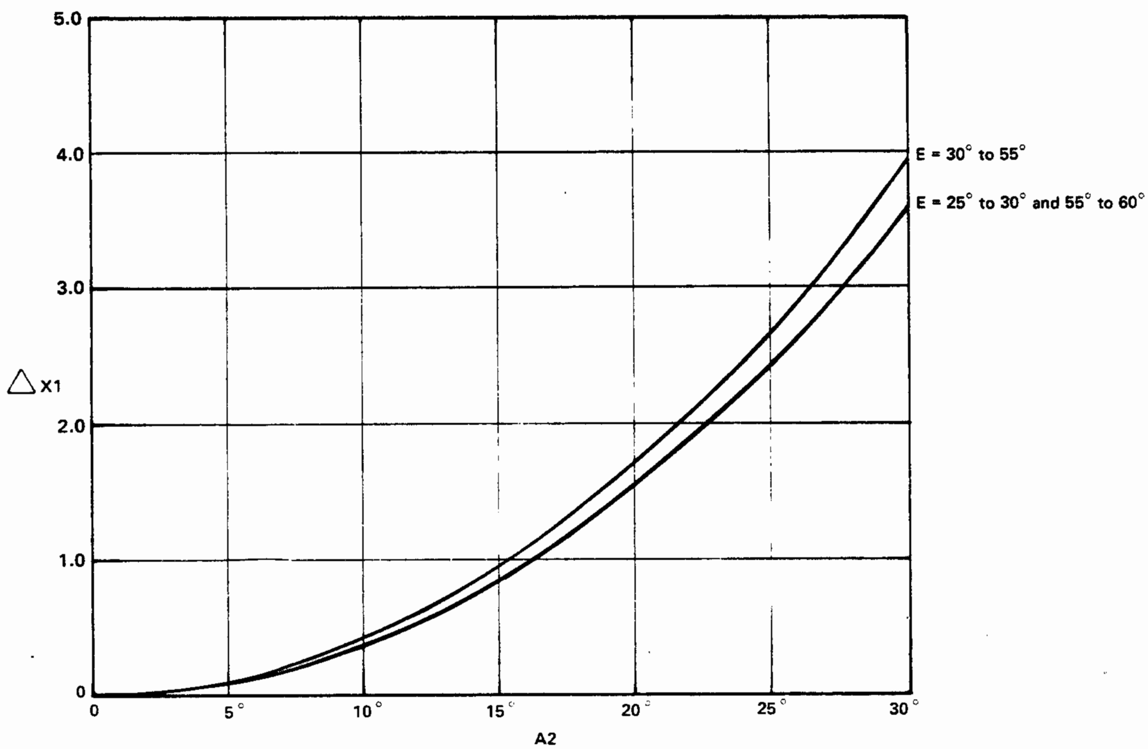


Figure 11. ΔX1 = f (A2,E)

3) Read the ΔX1 values for corresponding ΔA2 values.

Note that a decrease in the absolute value of A2 gives a decrease in X1 and occurs with a decrease in the absolute value of Y1. Thus, if the foundation heading is closer to the satellite than the best guess, the X1 angle must be decreased during the search. For this example, X1 decreased by 0.65° for a 5° error decrease in actual A2, and X1 increases by 0.95° for a 5° error increase in actual A2. See 2.

(1) The nominal value of Y1 should be calculated using Y1 = $\sin^{-1}$ (cos E sin A2) instead of using Figure 10. Using this equation, Y1 = 14.07°. Similar calculations using A2 = 13-1/2° and 23-1/2° yield Y1 values of 10.30° and 17.79°. Figure 10 and the subsequent example of Figure 12 are intended to give a broad view of the relationship of Y1 as a function of A2 and E.

(2) The nominal value for X1 should be calculated using

$$X1 = \tan^{-1}\left(\frac{\tan E}{\cos A2}\right)$$

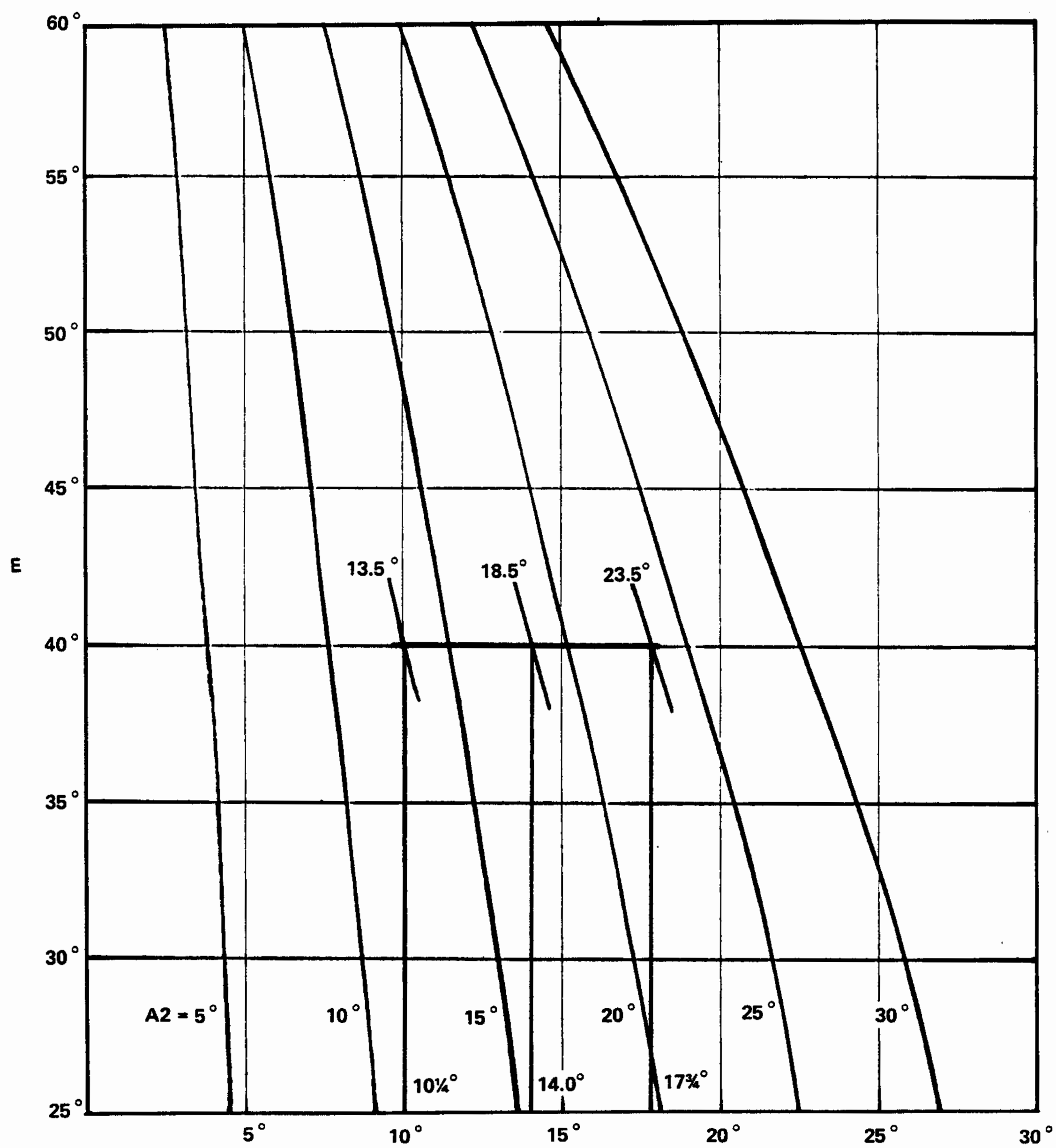


Figure 12.

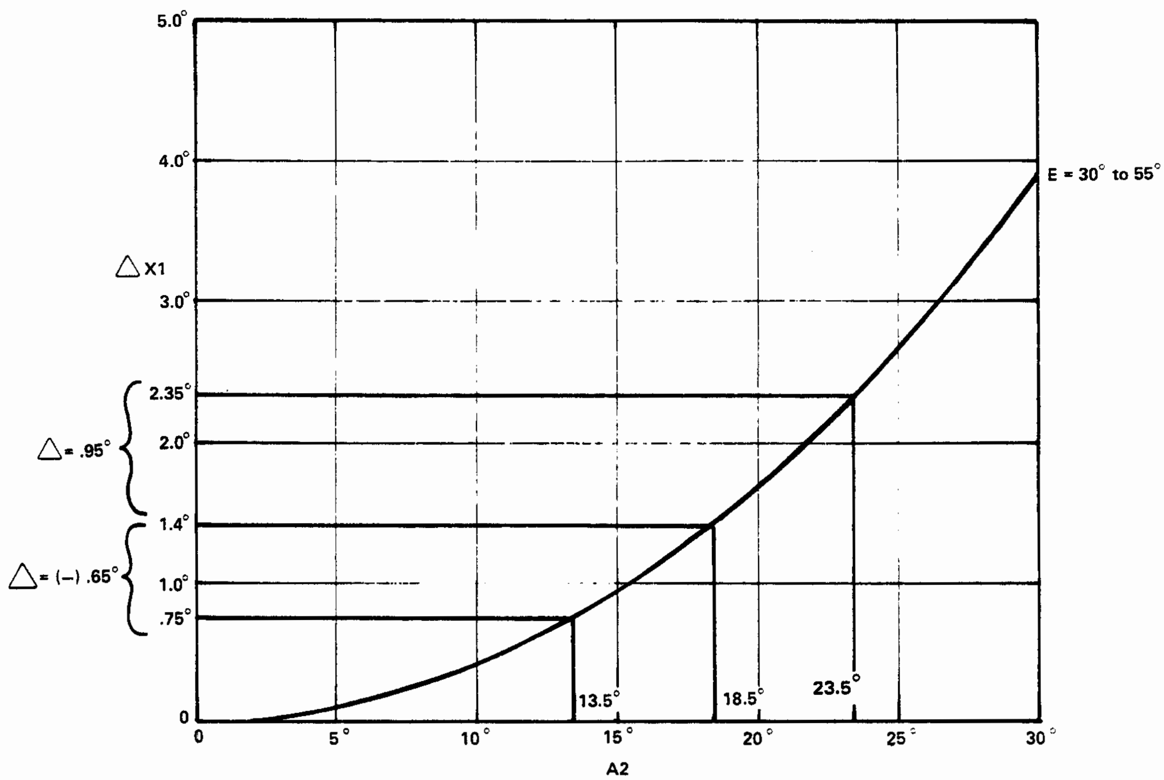


Figure 13.

For this example, X1 = 41.50°. As for ΔY1, the ΔX1 values are easily calculated. Figures 11 and 12 should be viewed in the same light as Figures 10 and 12.

After the satellite is acquired and the pointing of the antenna is peaked, A2 can be accurately determined by measuring either X1 or Y1 and using the appropriate A2 formula. When this is done, the boresighting of the X-Y mount is complete.

APPENDIX B

Derivation of Elevation and Azimuth Pointing Angles to Geostationary Satellites

Refer to Figure 1A, 1B, and 1C. Figure 1B is a bottom view of the earth and the geosynchronous orbital arc. It shows a satellite located at an angle C where:

$$C = \text{satellite longitude} - \text{site longitude} \tag{1}$$

It also shows the geosynchronous arc at 6.611 earth radii. A rectangular coordinate system is established in the orbital plane with the local meridian as a coordinate axis. In this coordinate system, the coordinates of the satellite are:

$$(6.611 \sin C,\ 6.611 \cos C) \tag{2}$$

Figure 1A is a side view of the earth with the equatorial plane bounded by the geosynchronous orbital arc shown as a horizontal line. A local horizontal earth plane at a latitude angle, X, is shown as a tangent line touching the earth. The azimuth and elevation angles to a satellite are developed by projecting the satellite position (from 1B) onto this horizontal earth plane.

Figure 1C is a projection of the orbital arc into the local horizontal plane. It is the view of the orbital arc by an observer at a great distance above his local position on the earth. The task of generating the local AZ-EL angles is essentially accomplished by projecting the rectangular coordinates of the satellite from the orbital plane to the local horizontal plane. It is helpful to consider that the local horizontal plane starts out parallel to the orbit plane at the north pole and is rotated about an axis X-X by an angle 90 - X(90° - local latitude). This rotation has not changed the coordinate 6.611 sin C. However, it is seen that the rotation has shortened the coordinate 6.611 cos C by sin X. Thus the coordinates of the satellite in the local horizontal plane are:

$$(6.611 \sin C,\ 6.611 \cos C \sin X) \tag{3}$$

The angle C' measured in the local horizontal plane is then:

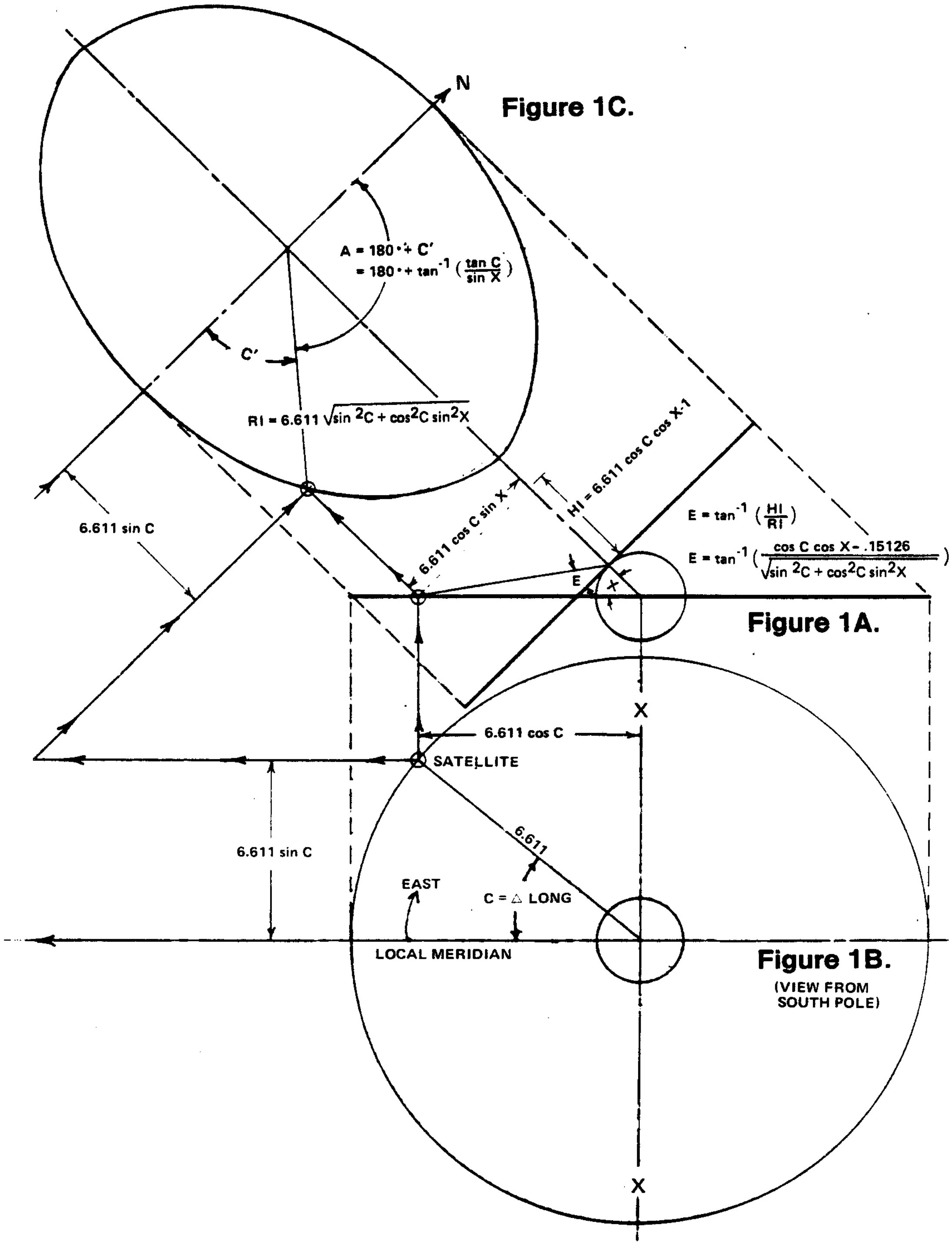
$$C' = \tan^{-1}\left(\frac{6.611 \sin C}{6.611 \cos C \sin X}\right) = \tan^{-1}\left(\frac{\tan C}{\sin X}\right) \tag{4}$$

Note that in Figure 1B and 1C the satellite is east of the site, and therefore, C and C' are negative (C = satellite longitude - site longitude). Azimuth is measured from north and thus is equal to 180° + C': or

$$A = 180^\circ + \tan^{-1}\left(\frac{\tan C}{\sin X}\right) \tag{5}$$

The distance to the satellite in the local horizontal plane is the hypotenuse of the right triangle having legs equal to the retangular coordinates:

$$R1 = 6.611\sqrt{\sin^2 C + \cos^2 C \sin^2 X} \tag{6}$$



Figures 1A, 1B, and 1C. Earth Station Satellite Geometry

Referring to Figure 1A, the height of the satellite above the local horizontal plane is:

$$H1 = 6.611 \cos C \cos X - 1 \tag{7}$$

The elevation angle of the satellite is then:

$$E = \tan^{-1} \frac{H1}{R1} \tag{8}$$

$$E = \tan^{-1}\left(\frac{(\cos C \cos X) - 0.151263}{\sqrt{\sin^2 C + \cos^2 C \sin^2 X}}\right) \tag{9}$$

List of Symbol Definitions for Pointing Angles and Boresight Reference

X = Site Latitude
Y = Site Longitude
Z = Satellite Longitude

FOR EL-OVER-AZ POINTING ANGLES
C = Z - Y
A = True Local Azimuth of Satellite
E = Local Elevation of Satellite

FOR X-Y MOUNT POINTING ANGLES
A1 = True Local Azimuth of Perpendicular to Lower Axis (Mount Azimuth Heading)
A2 = A - A1
X1 = Lower Axis Angle with Respect to Horizontal(zero when Y1 axis is vertical)
Y1 = Upper Axis Angle Measured With Respect to Peak Ascension

For HA-DEC (POLAR MOUNT) POINTING ANGLES
H = Local Hour Angle Measured with Respect to Peak Ascension (Angle Positive for Angles West of Site)
D = Declination (A Depressing Declination is Considered Negative)

FOR SINGLE AXIS POINTING ANGLE AND BORESIGHT REFERENCE
A_1 = True Local Azimuth of Eastern Satellite For Zero Error
A_2 = True Local Azimuth of Western Satellite For Zero Error
E_1 = Local Elevation of Eastern Satellite For Zero Error
E_2 = Local Elevation of Western Satellite For Zero Error
S1 = True Local Azimuth of Single Axis
E1 = Local Elevation of Single Axis(zero when single axis is horizontal)
$A2_N$ = A_N - S1 (A_N Azimuth of Satellite, N)
X1 = 90 - E1
$Y1_N$ (Upper Axis Angle Measured with Respect to Peak Ascension to Satellite N)

FOR DECLINATION-CORRECTED SINGLE-AXIS POINTING ANGLE AND BORESIGHT REFERENCE
D1 = Reflector Tilt from Single Axis (A Depressing tilt is Considered Positive)
E2 = Declination-Corrected Single-Axis Elevation
K = Zero Error Approximation Constant
H = Declination-Corrected Single-Axis Pointing Angle to Satellite (Angle Positive for Angles West of Site)

Pointing Angle and Boresight Equations

ELEVATION-OVER-AZIMUTH POINTING ANGLES

$$A(AZ) = 180° + \tan^{-1}\left(\frac{\tan C}{\sin X}\right) \quad \text{(Northern Hemisphere)}$$

$$E(EL) = \tan^{-1}\left(\frac{\cos C \cos X - 0.15126}{\sqrt{\sin^2 C + \cos^2 C \sin^2 X}}\right)$$

Y-OVER-X LOOK ANGLES AND BORESIGHT

$$X1 \text{ (Lower Axis)} = \tan^{-1}\left(\frac{\tan E}{\cos A2}\right)$$

$$Y1 \text{ (Upper Axis)} = \tan^{-1}(\tan A2 \cos X1)$$

$$Y1 = \sin^{-1}(\cos E \sin A2)$$

$$A2 = A - A1$$

$$A2 = \sin^{-1}\left(\frac{\sin Y1}{\cos E}\right)$$

$$A2 = \cos^{-1}\left(\frac{\tan E}{\tan X1}\right)$$

$$A2 = \tan^{-1}\left(\frac{\tan Y1}{\cos X1}\right)$$

DEC OVER HA (POLAR) POINTING ANGLES

$$H \text{ (Hour Angle)} = \tan^{-1}\left(\frac{\sin C}{\cos C - 0.15126 \cos X}\right)$$

$$D \text{ (Declination)} = \tan^{-1}\left(\frac{-0.15126 \sin X \sin H}{\sin C}\right)$$

SINGLE-AXIS POINTING ANGLE AND BORESIGHT

$$S1 \text{ (Single-Axis Azimuth)} = A_1 - \tan^{-1}\left\{\frac{\cos(A_2 - A_1) - \left[\frac{\tan E_2}{\tan E_1}\right]}{\sin(A_2 - A_1)}\right\}$$

$$A2_N \text{ (Satellite Azimuth Minus SA Azimuth)} = A_N - S1$$

$$E1 \left(\begin{matrix}\text{single axis}\\ \text{elevation}\end{matrix}\right) = 90 - \tan^{-1}\left(\frac{\tan E_1}{\cos A2_1}\right) = 90 - \tan^{-1}\left(\frac{\tan E_2}{\cos A2_2}\right)$$

$$Y1_N \left(\begin{matrix}\text{single axis}\\ \text{pointing angle}\\ \text{to satellite N}\end{matrix}\right) = \tan^{-1}(\tan A2_N \sin E1)$$

Declination Corrected Single Axis Pointing Angle and Boresight

$$D1 \left(\begin{array}{c}\text{reflector tilt}\\ \text{from single axis}\end{array}\right) = K + \tan^{-1}\left(\frac{\sin X}{6.611}\right)$$

$$E2 \left(\begin{array}{c}\text{declination-corrected}\\ \text{single-axis elevation}\end{array}\right) = X - D1 + \tan^{-1}\left(\frac{\sin X}{6.611 - \cos X}\right)$$

$$H \left(\begin{array}{c}\text{declination-corrected}\\ \text{single-axis pointing}\\ \text{angle to satellite}\\ \text{(approximate)}\end{array}\right) \approx \tan^{-1}\left(\frac{\sin C}{\cos C - 0.15126 \cos X}\right)$$

$$H \left(\begin{array}{c}\text{declination-corrected}\\ \text{single-axis pointing}\\ \text{angle to satellite}\\ \text{(exact)*}\end{array}\right) = \cos^{-1}\left(\frac{\sin E + \sin E2 \sin D1}{\cos D1 \cos E2}\right)$$

$$A\ (AZ) = 180 + \tan^{-1}\left(\frac{\sin H}{\cos E2 \tan D1 + \sin E2 \cos H}\right)$$

$$E\ (EL) = \sin^{-1}(\cos H \cos D1 \cos E2 - \sin E2 \sin D1)$$

*The algebraic sign of H must be assigned (+ for west, - for east)

Earth Station Antennas
Mechanical and Structural Considerations

Roger Peirce

Introduction

An earth station antenna is both an electrical receiver/transmitter and a mechanical/structural system. The proper functioning of a reflector-type antenna requires that it maintain a precision surface while exposed to a sometimes harsh environment. The location and size of the earth station antenna usually make it subject to local building codes. Since most earth station antennas cannot be shipped assembled, it is necessary to have a method of mechanical assembly that assures adequate surface accuracy after assembly on-site. This paper discusses the major mechanical and structural considerations required in the design of an earth station antenna.

Structural Design Criteria

In most areas of the United States, the reflectors of antenna earth stations are classified as structures that must be designed, fabricated and installed to comply with local building codes. The code which is almost universally accepted is the American National Standard Building Code Requirements for Minimum Design Loads in Buildings and Other Structures, ANSI A58.1. Paragraph 1.3 of that standard states:

> Buildings or other structures, and all parts thereof, shall be designed and constructed to support safely all loads, including dead loads, without exceeding the allowable stresses (or ultimate strengths when appropriate load factors are applied) for the materials of construction in the structural members and connections. When both wind and earthquake loads are present, only that one which produces the greater stresses need be considered, and both need not be assumed to act simultaneously.

The code does not specifically define the term allowable stresses, but virtually all engineering societies concerned with structures and decisions of the courts have agreed that the allowable stresses for steel and aluminum are those specified by the American Institute of Steel Construction Manual of Steel Construction and the Aluminum Association Aluminum Construction Manual.

The loads which must be safely supported by an earth station antenna system are the weight of the antenna and the attached equipment, the expected snow and ice load, earthquake loads and the wind load.

The margin between the allowable stress and the yield point of the material is sometimes called the "safety-factor". However, it should be observed that this term is never used by the building code or the Manual of Steel Construction. The "safety-factor" of most structures cannot be defined by a single value or a meaningful minimum because the allowable stress for steel (or aluminum) cannot be defined by a single value. The AISC Manual of Steel Construction requires over 900 pages to describe construction standards and define the allowable stress for steel.

All reputable earth station manufacturers follow the policy of designing the reflector and supporting structure so that at the catalog-specified survival wind the system is safely supported.

Wind Loads

Although the wind load is not the only load that must be considered in the reflector and mount design, wind is usually the largest single contributor to the stress and deflection in the structure. Thus, a considerable effort has been expended in getting accurate, reliable information on wind loads.

Earth station antennas have a specification that is variously called maximum wind, survival wind or withstand wind. These terms should be considered synonymous. They represent a part of the total load that must be safely supported. We will use survival wind.[1]

A comparison of the typical survival wind and the wind load stipulated by the building codes will not be attempted in this discussion. It is sufficient to state that at the manufacturer's specified survival wind the system must be safely supported without exceeding the allowable stresses for the materials. Since antenna reflectors are usually constructed of steel and aluminum, the preceding sentence can be paraphrased as follows: at the manufacturer's specified survival wind the system must be safely supported without exceeding the allowable stresses as defined by the manual of Steel Construction or the Aluminum Construction Manual.

The wind loads for the design of Scientific-Atlanta's reflectors, mounts and foundations are derived from wind tunnel tests performed on scale-model paraboloidal reflectors. In 1964 Jet Propulsion Laboratory carried out a comprehensive study of reflectors with various configurations at the wind tunnel facility at California Institute of Technology. The reports provide data on pressure differences, force and moment coefficients for reflectors as influenced by wind direction, F/D ratio, surface porosity, spoilers, simulated support structure and Reynolds number.

Drag, lift and moment coefficients provided by the JPL report are referenced to the vertex of the reflector. These data are used in determining the reactions at the mount foundations and stress in the structural elements of the mount. The reports provide a detailed map of pressure differences across the reflector surface. These data are used to calculate the internal stresses in the reflector panels and panel support structure. The appropriate data can be used to make a computer-generated table of wind load information for a unique set of conditions being studied or considered. Figure 1 is a typical printout showing forces and moments on a typical reflector. As expected, the largest drag force occurs at a relative wind of zero degrees; that is, where the wind is parallel to the boresight line and blowing into the concave aperture.

[1]Survival wind as used herein when combined with ice and dead weight results in the "Design Load" as defined in EIA Standard RS-222C.

WIND FORCE ON SOLID SURFACE PARABOLOIDAL REFLECTOR
BASED ON JPL WIND TUNNEL TESTS. (PROGRAM REVISED 2/2/77)

MODEL NUMBER	8-METER	PEDESTAL
REFLECTOR DIAMETER	26.3	FEET
REFLECTOR F/D RATIO	0.330	
VERTEX AXIAL OFFSET	0.00	FEET
VERTEX RADIAL OFFSET	0.00	FEET
PIVOT AXIS TO REF PLANE	12.00	FEET
AMBIENT TEMPERATURE	0	DEGR F
WIND VELOCITY	100.0	MPH
DYNAMIC PRESSURE	28.87	LBS/SQ-FT

ANT AXIS FROM WIND DEGREES	TORQUE ON PIVOT AXIS LB-FT	DRAG LBS	LIFT/OR SIDE FORCE LBS	MOMENT ABOUT REF PLANE, LB-FT
0	0	23436	0	281237
10	-4717	22968	-3906	270896
20	-10459	21256	-8125	244617
30	-14765	19858	-12187	223537
40	-16405	17324	-15624	191482
50	-19071	14970	-19218	160568
60	-35272	11216	-22343	99317
70	5742	7489	-18749	95612
80	38963	3948	-8593	86338
90	51882	2760	0	85006
100	51882	3850	5937	98080
110	52703	3665	781	96685
120	56394	4480	781	110157
130	52908	6041	2500	125398
140	47986	8844	3750	154116
150	39373	11140	3906	173057
160	28710	13583	3281	191705
170	15555	16318	2031	211401
180	0	16874	0	202491

FINISHED

Figure 1. Wind Loads on 8-Meter Reflector

It is important to note, however, that the maximum loads in parts of the support structure will occur when the wind is blowing on the back of the reflector, as shown in Figure 2. For the reflector position illustrated, the wind direction is 120 degrees relative to the antenna boresight direction (antenna elevation 60 degrees). This condition is usually the most severe load for the elevation jack and the elevation pivot axis. At the 60-degree elevation position, the elevation jack on an elevation-over-azimuth mount is at or near the extreme extended length or longest column condition where column stability must be carefully evaluated.

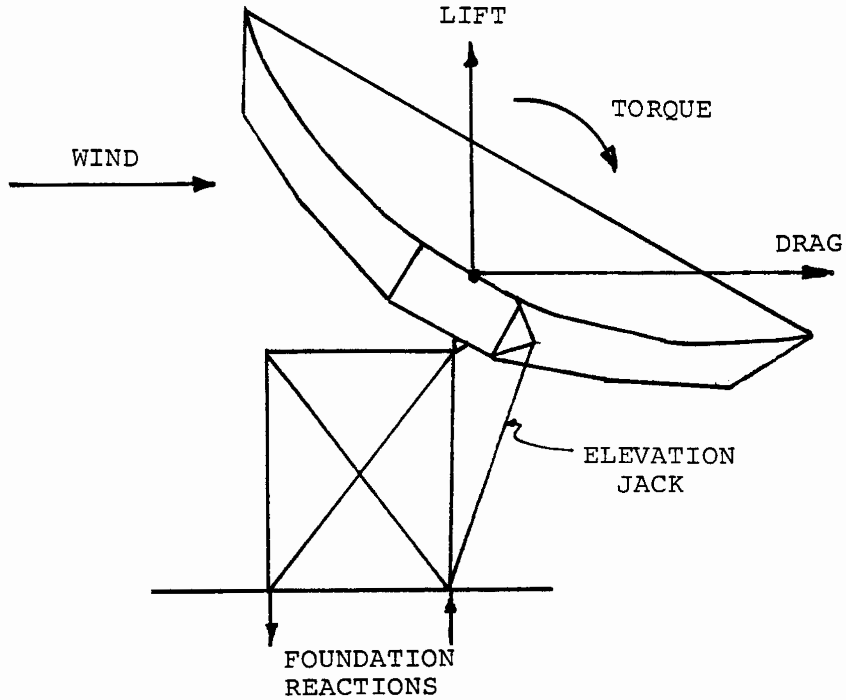


Figure 2. Reflector at 60 Degree Elevation Angle. Wind is 120 Degrees from Boresight Direction

Figure 3 shows maximum wind drag forces developed at survival wind velocity. This figure is intended to give the reader a grasp of the magnitude of wind forces generated by earth station antennas.

Antenna	Survival Wind Speed (mi/h)	Drag Force (lbs.)
3.0 Meter	100	3,200
4.6 Meter	110	9,300
5.0 Meter	125	14,000
7.0 Meter	125	28,000
7.7 Meter	125	32,000
10.0 Meter	125	55,000
11.0 Meter	125	66,000

Figure 3. Maximum Reflector Drag Forces

Analytical Methods

Extensive use of computer-aided analytical techniques is employed in the design of earth station antenna systems. These computer programs are based on the finite-element analytical approach to structural analysis and have proprietary names such as STRUDL, NASTRAN, EASE 2, TRUSS 1B and FRAME.

The use of these programs involves first generating a mathematical model of the structure to be analyzed. This includes describing the members' geometrical relationship, section and material properties, and joint-constraint conditions. Loads are then applied at joints or nodal points to model specific environmental conditions. Twenty to thirty load cases are often analyzed to determine both survival and operational responses of the structure. The effects of dead load, ice load, operational wind and survival wind load conditions, and differential temperature and dynamic loading of members are analyzed.

The results are reviewed to determine the maximum load conditions in each member and joint. All members and joints are analyzed using methods approved by the American Steel Construction Institute or the American Aluminum Association to verify that the allowable stresses are not exceeded. The reactions to the fixed frame of reference (usually the earth) are reviewed and tabulated for use in foundation design.

The structural stiffness of earth station antennas is also important to assure acceptable pointing stability and surface tolerances during operational environmental conditions. Boresight pointing accuracy specifications of from 0.15° to 0.02° can be required during 60 mph winds and 1-inch radial ice. In many instances, especially when the antenna is to be used at the higher frequency bands, the stiffness of the assembly is the determining factor for the design parameters of structural members. The members supporting the reflector surface are also carefully analyzed for stiffness, since the deflection of these members under load results in degradation of the RF performance. Figures 4 through 7 illustrate typical models used in computer analyses of reflector structures and a sample printout of the results.

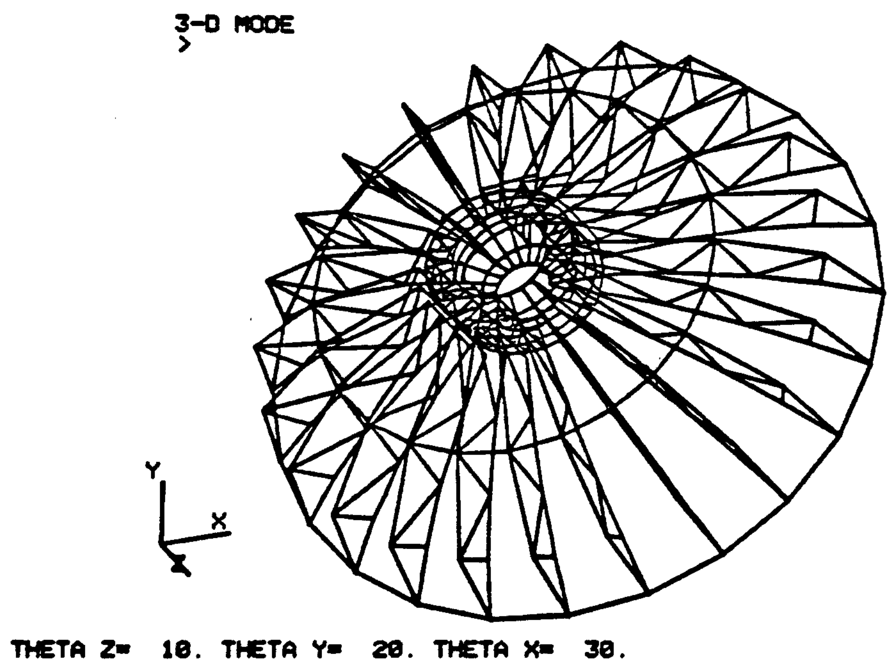


Figure 4. Computer Generated Plot

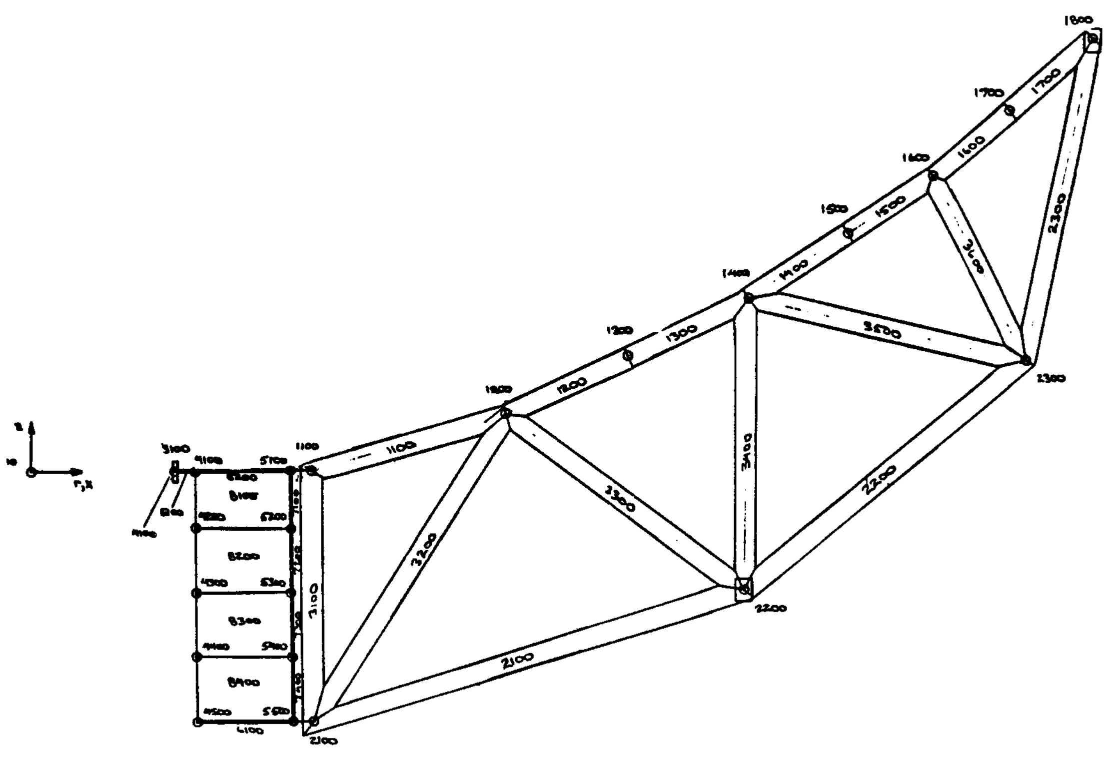


Figure 5. Computer Model Diagram, Truss

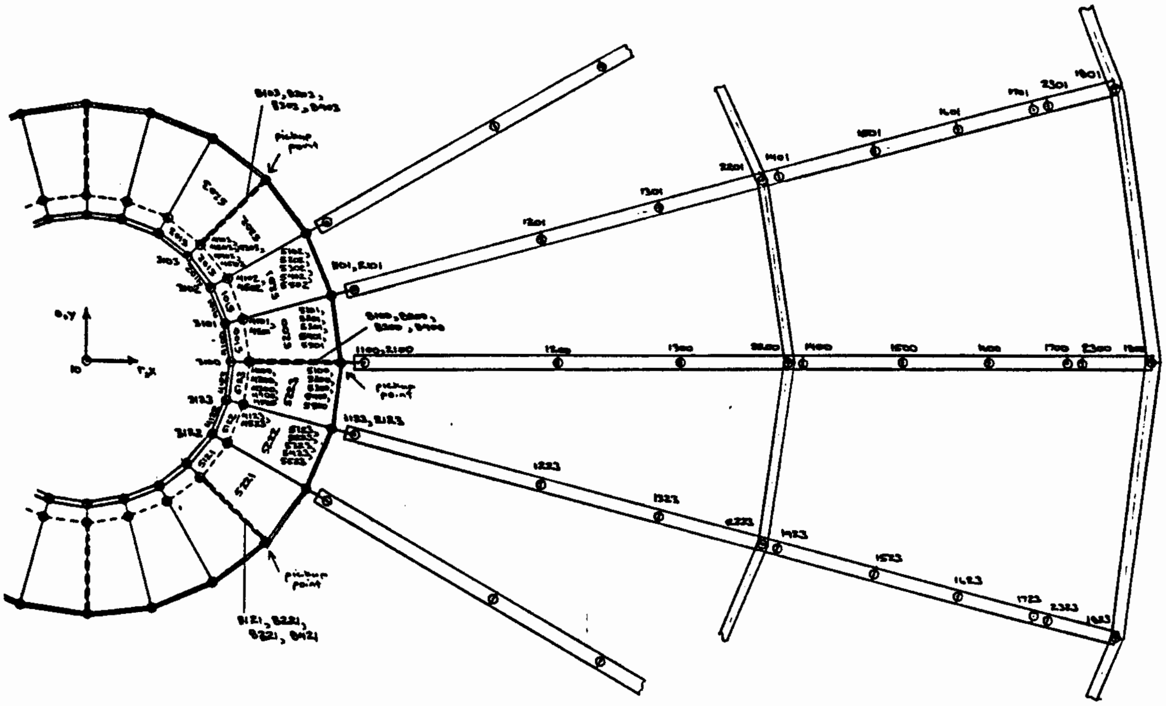


Figure 6. Computer Model Diagram, Hub, and Truss

SBS 7.7 METER TRUSS/HUB ANALYSIS
THREE DIMENSIONAL MODEL NUMBER 1

BEAM NUMBER	LOAD CASE	NODE NUMBER	AXIAL FORCE	LOCAL-Y SHEAR	LOCAL-Z SHEAR	AXIAL TORQUE	LOCAL-Y MOMENT
4416	1	5116	-34.953	24.088	-.268	-.984	5.189
		5117	34.528	-24.264	-.530	.984	-6.418
4417	1	5117	-13.361	24.657	.297	.898	1.305
		5118	12.904	-24.718	-1.094	-.898	-7.840
4418	1	5118	27.714	21.547	3.217	5.975	-7.720
		5119	-28.170	-21.487	-4.014	-5.975	-26.258
4419	1	5119	95.208	20.497	-1.936	7.331	10.994
		5120	-95.634	-20.320	1.139	-7.331	3.457
4420	1	5120	170.369	12.209	20.438	-2.078	-64.604
		5121	-170.734	-11.928	-21.235	2.078	-131.216
4421	1	5121	218.156	4.719	-15.794	15.024	90.346
		5122	-218.437	-4.353	14.997	-15.024	54.339
4422	1	5122	255.411	-6.459	-.629	6.893	2.434
		5123	-255.588	6.885	-.168	-6.893	-.268
4423	1	5123	123.205	-1.770	-.122	.095	.297
		5100	-123.220	1.884	-.077	-.095	-.087
4500	1	5500	10.441	3.598	4.769	9.087	-12.421
		5501	-10.381	-3.830	-5.658	-9.087	-36.574
4501	1	5501	-16.243	-6.073	7.315	-10.899	-33.459
		5502	16.419	5.818	-8.182	10.899	-39.360
4502	1	5502	-27.171	19.079	20.804	-36.697	-94.382
		5503	27.451	-19.377	-21.629	36.697	-105.008
4503	1	5503	-413.290	-29.559	-9.866	21.156	71.844
		5504	413.656	29.201	9.100	-21.156	17.278
4504	1	5504	-232.820	-2.427	11.427	-7.667	-47.167
		5505	233.245	1.994	-12.120	7.667	-63.476
4505	1	5505	-107.344	-8.199	7.320	-2.186	-20.061
		5506	107.800	7.684	-7.931	2.186	-51.604
4506	1	5506	-50.733	-6.530	3.719	-.460	-4.412
		5507	51.190	5.929	-4.246	.460	-33.014

Figure 7. EASE2 Program Printout

Mechanical Design for Surface Accuracy

Earth station antenna reflectors and mounts must be assembled from several or more parts, and the assembled system must have a high degree of surface accuracy for proper operation. There are two distinctly different design approaches that are currently employed to achieve the required surface accuracy. One of the techniques results in a custom-fitted reflector; the other results in a reflector which can be assembled from interchangeable parts.

Custom-fitted assembly (not used by Scientific-Atlanta) can be accomplished in several ways. For example, some manufacturers of earth station reflectors use a factory-assembly procedure for each reflector. A backing structure, consisting of a central hub and radial trusses, is assembled in a prepared area. The radial trusses are all match-marked to the hub and numbered. Next, a sweep template or swing arm with dial indicators is hung on a special central hub shaft so that it rotates above the assembly. The surface panels are then hand-fitted or adjusted to a position that will provide the best fit to the template. After all panels are clamped in place and checked with the template, tight-fitting bolt holes are hand-drilled through the panel flanges and the truss tabs. The panels are also match-marked and coded so that the reflector can be disassembled and reassembled at the earth station site with all the parts in the original position. This method of fabrication is suitable only to low production rates. Since each part is custom fitted, delays can occur if parts are lost or damaged at any time after the reflector leaves the manufacturer's plant.

Another custom-fitted assembly method employs adjustable panels that are surveyed and individually fitted into positions using a theodolite as an assembly tool. Each panel is supported by six or eight built-in screw-type adjustors. A theodolite, which is axially mounted at the reflector hub, is sighted at surface-mounted targets on the panels while the panels are adjusted to the proper shape and position. The assembly takes place at the earth station site and is typically performed by a highly skilled team. This assembly method has been frequently employed for reflectors having diameters of more than 20 meters.

Another approach to achieving a high degree of reflector surface accuracy is to design the antenna to have interchangeable parts. This is the design used by Scientific-Atlanta. The design employs a concept that results in a predictable surface conformity and surface tolerance when the reflector is assembled from precision parts. The concept is simple, but the execution requires a high degree of dimensional accuracy from parts that can be rather large. The accuracy of the interchangeable parts is accomplished by the use of precision tools, jigs and fixtures.

All of the parts that are used for the reflector assembly such as the hub, back structure, panels and subreflector spars are fabricated and inspected to ensure adherence to the required tolerances. When these parts are assembled on the site, the completed reflector will meet the surface conformity specifications. Scientific-Atlanta has demonstrated on many occasions that when a random selection of parts is assembled, the surface accuracy and RF performance are achieved within specification.

Fabrication Methods

Scientific-Atlanta uses two methods for fabrication of panels for paraboloidal reflectors. Both methods result in precision panels which can be assembled without using special tools or instruments resulting in minimum installation time, guaranteed surface accuracy and interchangeability of parts in case of damage.

Panels used for reflectors which are about 7-1/2 meters or larger in diameter and are produced in limited volume are built up using sheet aluminum and custom-formed reinforcing sections. This method consists of dividing the panel surface into sections and fabricating aluminum panels for each section. These panel sections are placed face down on the panel fixture in their appropriate locations and held in position with clamps and weights. Radial and circumferential stiffeners are laid over the panels. The stiffeners are slotted approximately every two inches to allow them to conform to the panel sections. The stiffeners are then systematically welded and riveted to the panels and each other. Figure 8 shows a panel being assembled in the panel fixture. The permanent attachment of the panels and stiffeners results in a stiff structure which maintains its shape after removal from the panel fixture (see Figure 9). All required bolt holes are drilled in the panel parts using drill bushings that are accurately located in the panel fixture. After fabrication, each panel is carefully inspected using gauge bars with dial indicators to assure surface conformity.

Figure 8. Fabrication of Reflector Panel in Panel Fixture

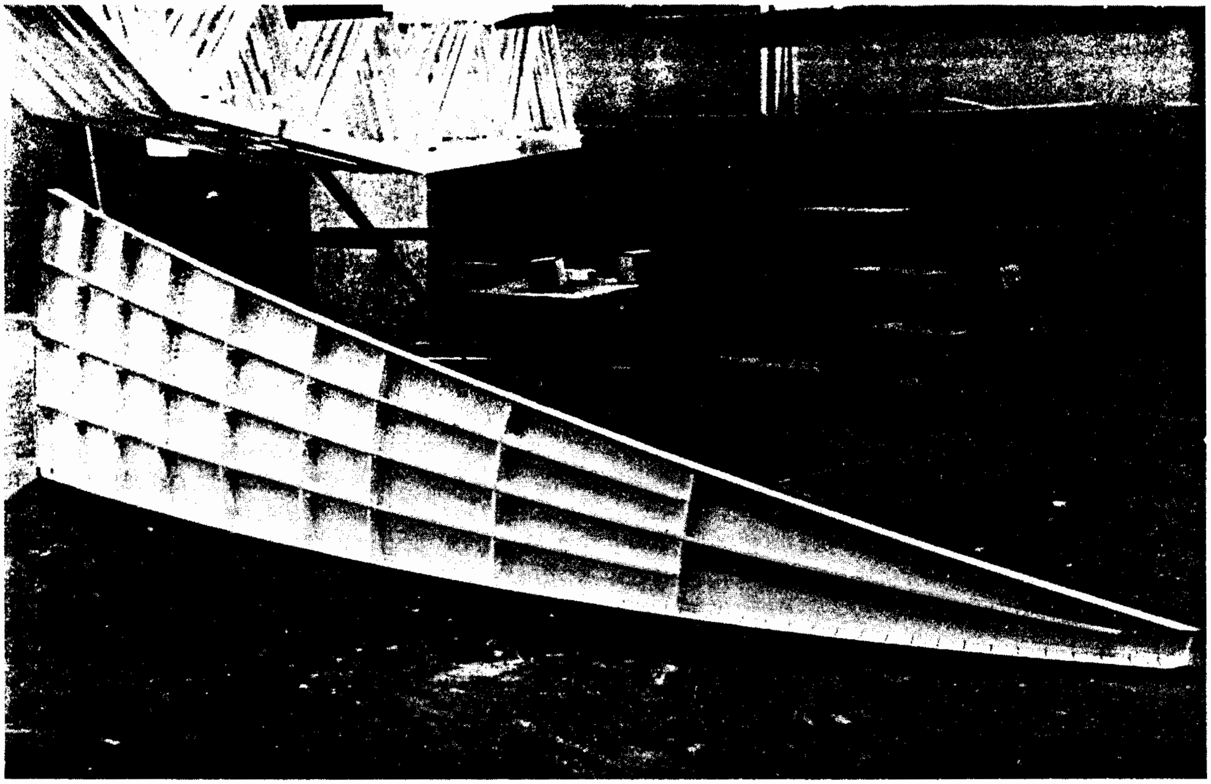


Figure 9. Fabricated Panel for 10-Meter Reflector

The panel fixture consists of radial and circumferential templates, which are assembled on massive tooled surfaces to maintain precision. The RF shape of the reflector is transformed by a computer into coordinates of the numerically-controlled machine which manufactures the template. Tooling holes are machined into the templates relative to the machined surface contour to facilitate accurate assembly. After assembly of the fixture, the surface is checked using optical measurement techniques.

The second fabrication method is to stamp the panels using matched dies. The application of this process to fabrication of reflector panels was developed by Scientific-Atlanta and has been used successfully on reflectors from three to seven meters in diameter. Reflectors with a surface accuracy of 0.015-inch rms have been assembled using stamped panels. The process consists of shearing a panel blank from aluminum sheet and then placing it between the matched forming dies. A single stroke of a giant press forms the concave reflector surface part of the panel, both side flanges and the outer edge flange.

Checking fixtures are used after stamping to verify surface accuracy. The formed panels are placed in drill jigs, which locate the precision attachment holes with respect to the panel surface (see Figure 10.)

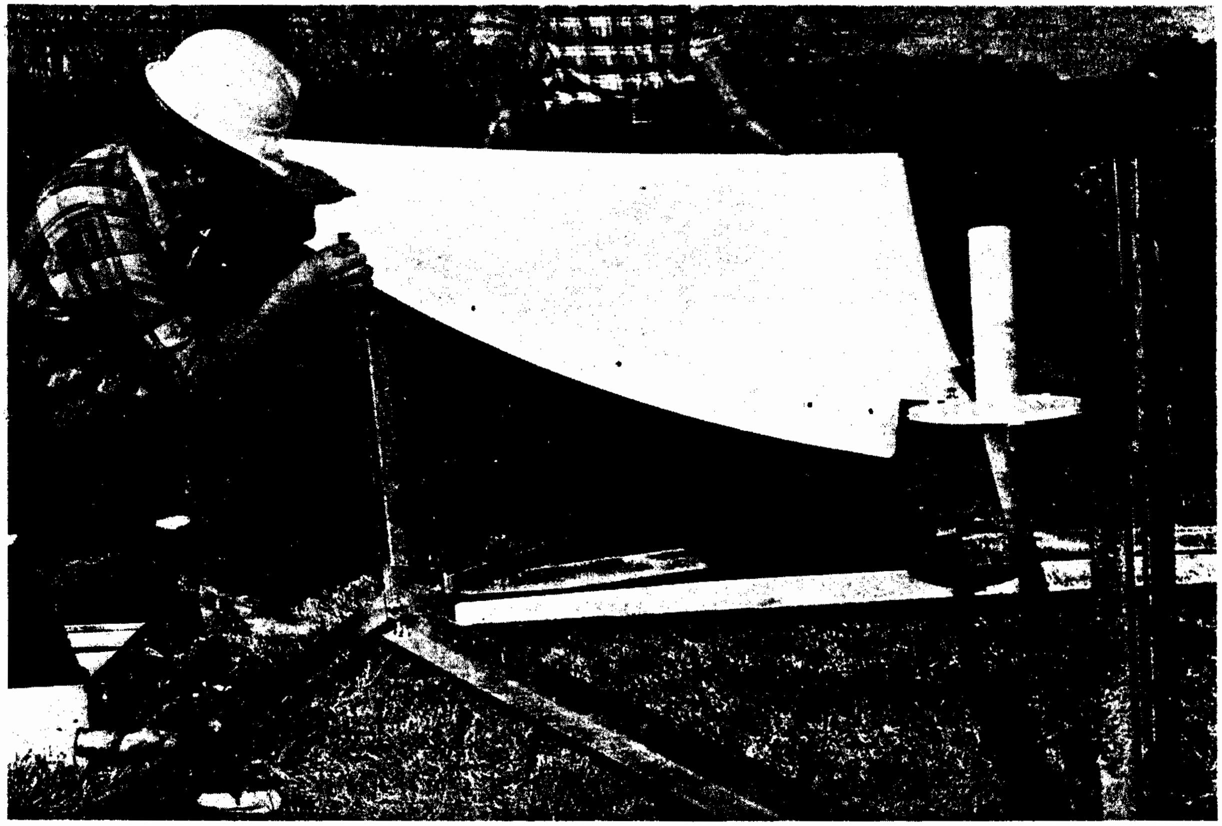


Figure 10. Stamped Panel

The radial truss and space frame are two approaches used by Scientific-Atlanta for making reflector backing structures. The radial truss system consists of precision-machined trusses attached to a machined central hub. The radial trusses are used when maximum stiffness and surface accuracy are desired (see Figure 11.) The trusses consist largely of lightweight tubular or H-section aluminum extrusions. Cover-plates reinforce the joints, and machined aluminum bar stock forms the terminal interface that will be bolted to the hub. All of these parts are clamped in a welding fixture to assure consistent dimensional control during welding. The welded trusses are allowed to cool for twenty-four hours to assure temperature equilibrium before machining.

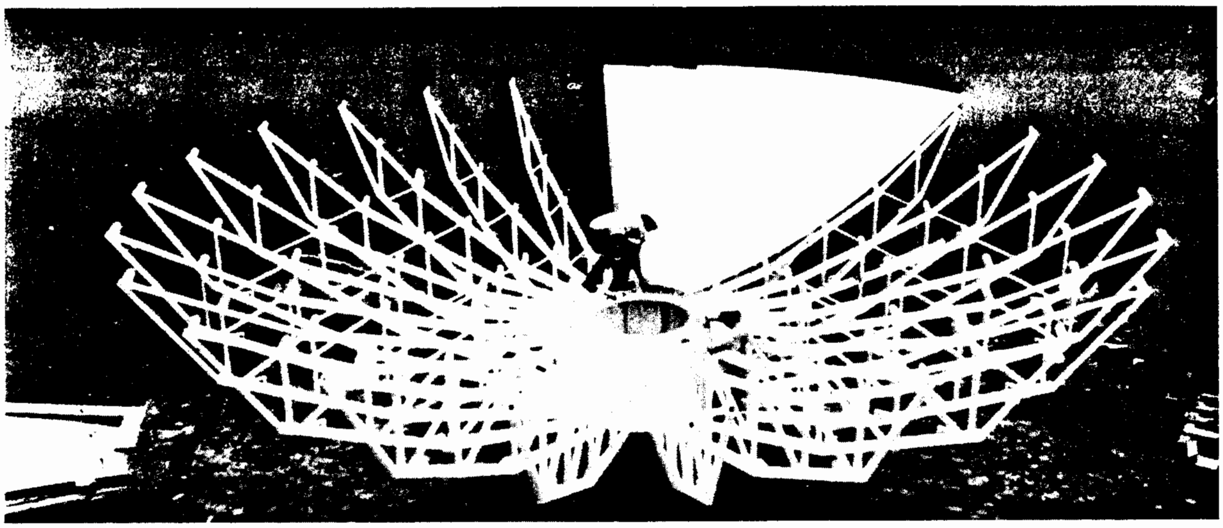


Figure 11. Hub and Truss Type Reflector Assembly

The truss-machining jig consists of a large steel table with precision-machined tooling points (see Figure 12). The welded truss is bolted by the hub pick-up points to the fixture, and the panel tab holes are precision-drilled. Each day a master truss is positioned in the jig to test it for accuracy.

Figure 12. Jig for Machining Trusses

The space-frame concept is based on the principle of interlocking tetrahedrons, with the geometry accurately controlled by the lengths of the three legs in each triangle. Precise effective lengths of the structural parts are maintained by accurate location of the close-tolerance holes. Accurate drill fixtures and hole-punching tools are employed for all operations.

This concept results in reflector backing structure nodes which are true within a few thousandths of an inch. Figure 13 illustrates a space-frame backing structure.

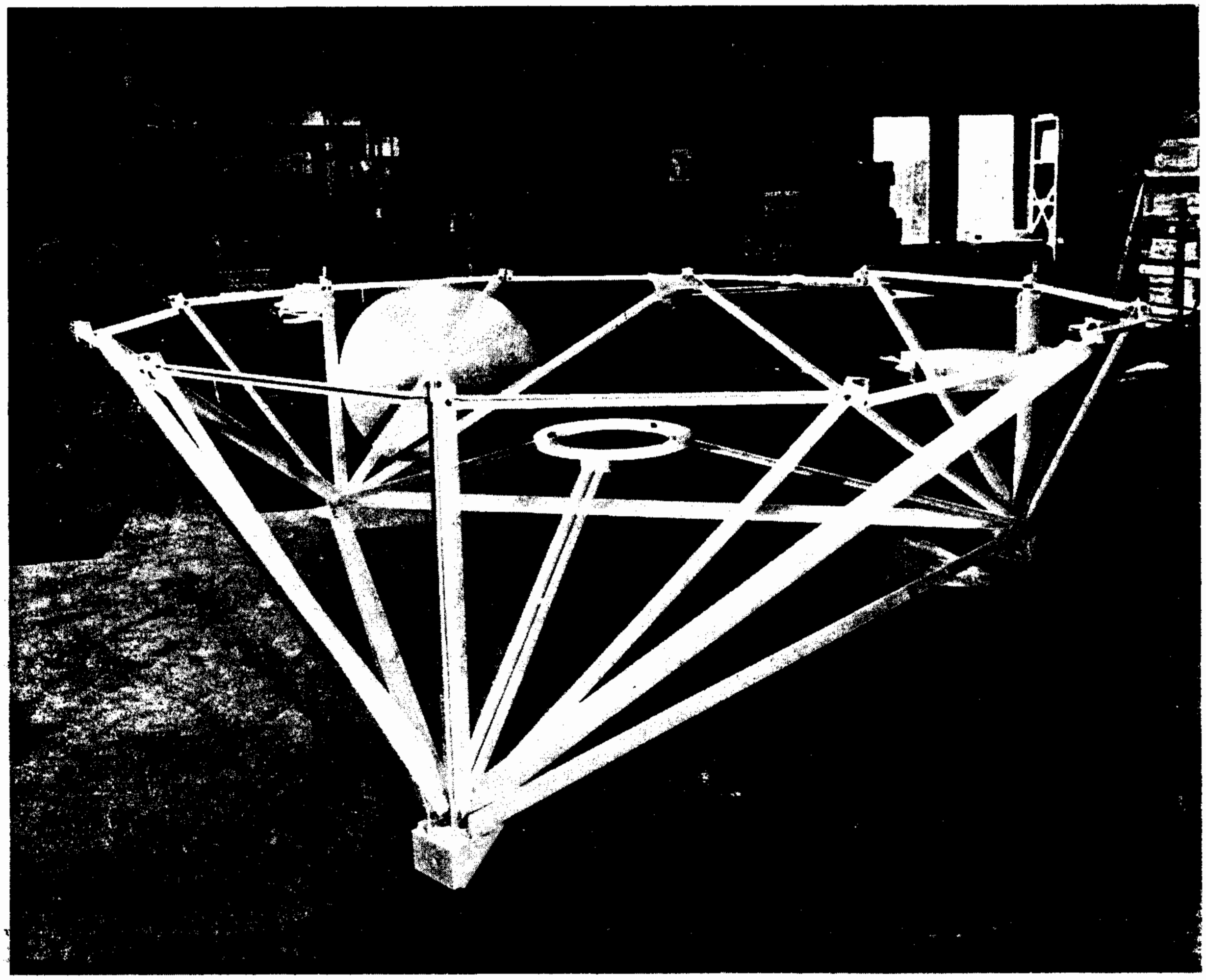
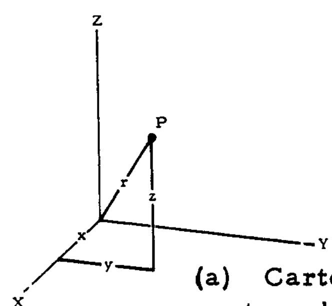


Figure 13. Space-Frame Structure for 4.6-Meter Reflector

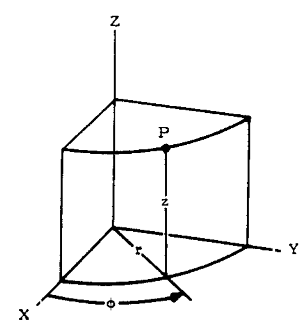
Reflector Surface Conformity Surveys

Assurance of proper antenna performance requires that the reflector surface conform to a theoretical surface within a specified tolerance. This assurance can be obtained by a surface conformity survey (surface mapping). Such surveys are important to the design engineers to show whether or not a new design meets the surface tolerance specification. A conformity survey is also periodically performed on reflectors assembled from production parts as a step in the quality control program. The survey yields the root-mean-square (rms) deviation of the actual surface computed normal to the theoretical best-fit surface. The survey also provides the magnitude, character and location of the principal deviations.

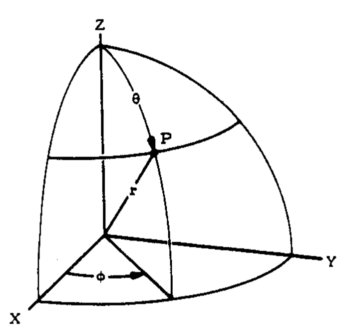
Surface conformity surveys have been performed with a variety of measurement instruments. Crude surveys are made using mechanical sweep templates. Some surveys have been performed using stereo photographic triangulation. Scientific-Atlanta currently uses clinometer bars for manual measurements and an automatic surface measurement system (ASMS) for high-speed measurements. Each method must supply sufficient raw data to permit a set of measured points to be described in three-dimensional space with respect to a coordinate reference frame.



(a) Cartesian or rectangular coordinates



(b) Cylindrical coordinates



(c) Spherical coordinates

Three Coordinate systems defining the same position in space. All three systems are right-hand systems, since rotation from X toward Y produces right-hand screw motion in the positive Z direction.

Figure 14. Three Coordinate Systems

Position in three-dimensional space is usually described by one of the three sets of orthogonal[1] coordinates shown in Figure 14. These are rectangular or Cartesian coordinates, cylindrical coordinates and spherical coordinates. The coordinate system that is most convenient for the surveying system and subsequent calculations is usually employed, since the conversion of data sets from one coordinate system to another is a trivial problem.

The survey and analysis of the data usually involve the consideration of two or more separate sets of data, each in its own coordinate frame. The survey instrument has its own coordinate frame. The theoretical paraboloidal or shaped surface is typically described by a set of equations, referenced to another coordinate frame. The required solution may develop a best-fit mathematical model referenced to a third coordinate frame. The following discussion assumes that the survey and the calculations are in spherical coordinates (Figure 14c) where each point on the surface of the reflector is defined by the angles $\emptyset$ and θ and a distance r, relative to the same origin.

Figure 15 illustrates the application of the spherical coordinate system to the conformity survey. Points p_1, $p_2 \ldots p_n$ are selected locations on the reflector surface along radials $\emptyset = Cj$, for $j = 1$ to K_p, where $k_p = 360/C$. The survey system determines $(\theta_n, r_n)_j$ for $n = 1$-to-N, where $r_n \sin\theta_n$ is the radius normal to Z of the Nth survey point. These points are fitted to the theoretical design curve describing the reflector surface.

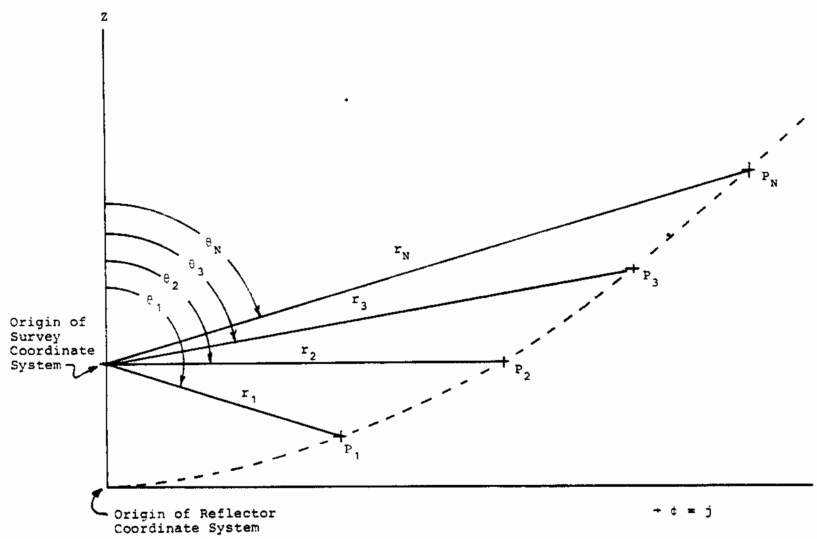


Figure 15. Reflector Surface Conformity Survey

[1]A three-space coordinate system is mathematically orthogonal if, at every point, the three-component vector directions are mutually orthogonal.

Clinometer-Bar Method

The surface conformity survey with clinometer bars is performed using a high-resolution, high-precision clinometer with a suitable quantity of especially fabricated bars of graduated lengths. Figure 16 illustrates the clinometer-bar method of surveying.

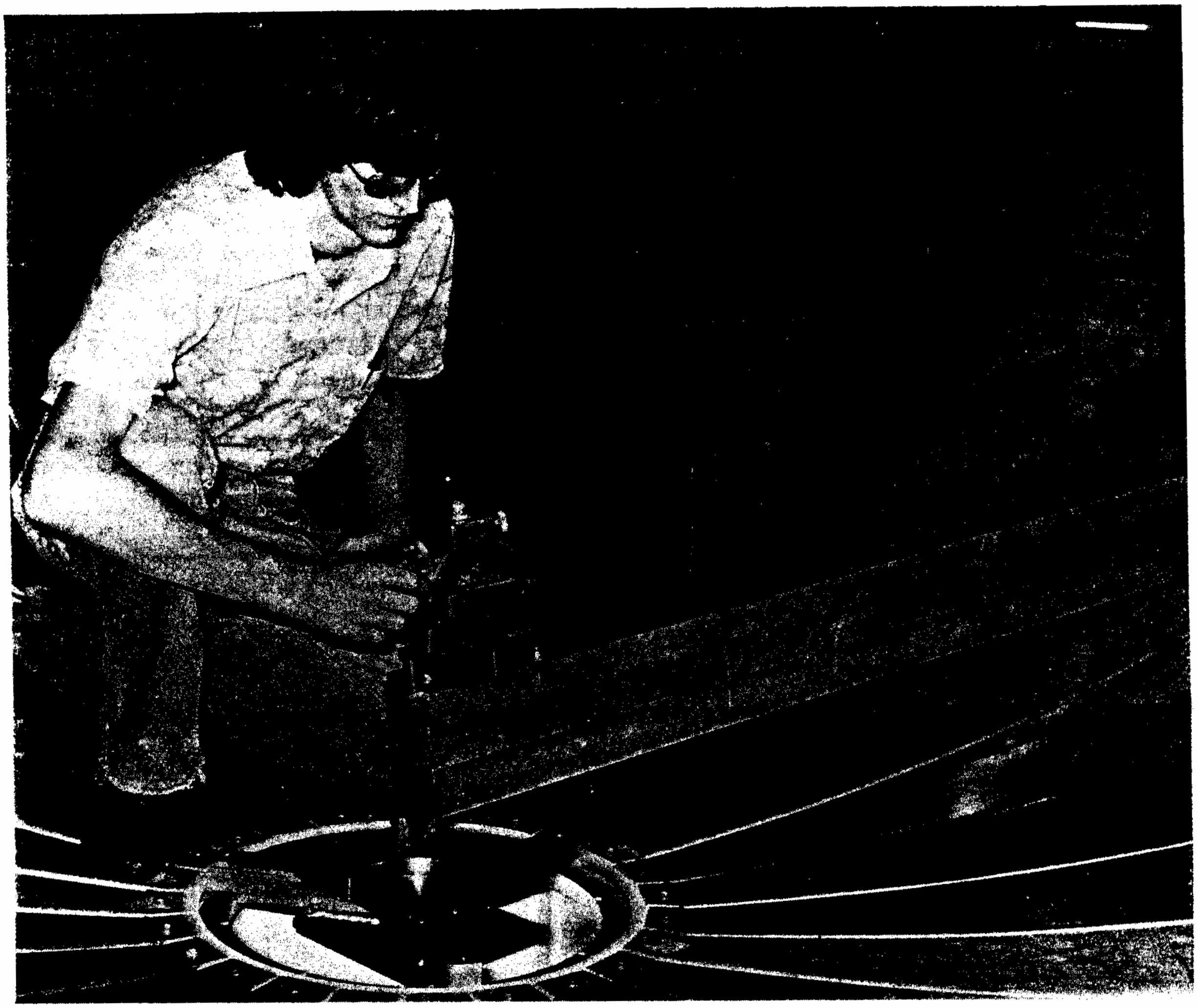


Figure 16. Measurement by Clinometer Bar

The selected bar is seated on the central pivoting fixture attached to the reflector hub, and the outboard contact is placed in contact with the reflector surface. After the clinometer is adjusted, the angle reading is recorded. The outdoor contact point is then rotated to a new angle Ø for the next reading. Six to twelve different bar lengths are usually employed to adequately cover the reflector surface.

Each bar is individually calibrated by a tilt adjustment on the clinometer seat, so that the clinometer reading accurately measures the inclination of the line from the contact point p to the center "O" of the cylindrical support.

The clinometer-bar system of measurement is different from most other systems in that the elevation angles are relative to a gravity-referenced horizontal plane. The reflector being measured is set approximately level before measurements begin. Reorientation of the gravity-based elevation angles to the reflector coordinate system is accomplished by the computer that is used to analyze the results.

Automated Surface Measuring System

The automated surface measuring system (ASMS) is designed to survey the surface deviations of an antenna reflector relative to a defined surface. Selected points on a reflector surface are surveyed by means of calibrated, adjustable-length arms attached to the turntable of a 19-bit encoder. The encoder is mounted on an axis of rotation that is approximately coincident with the reflector geometric axis (Figures 17 and 18).

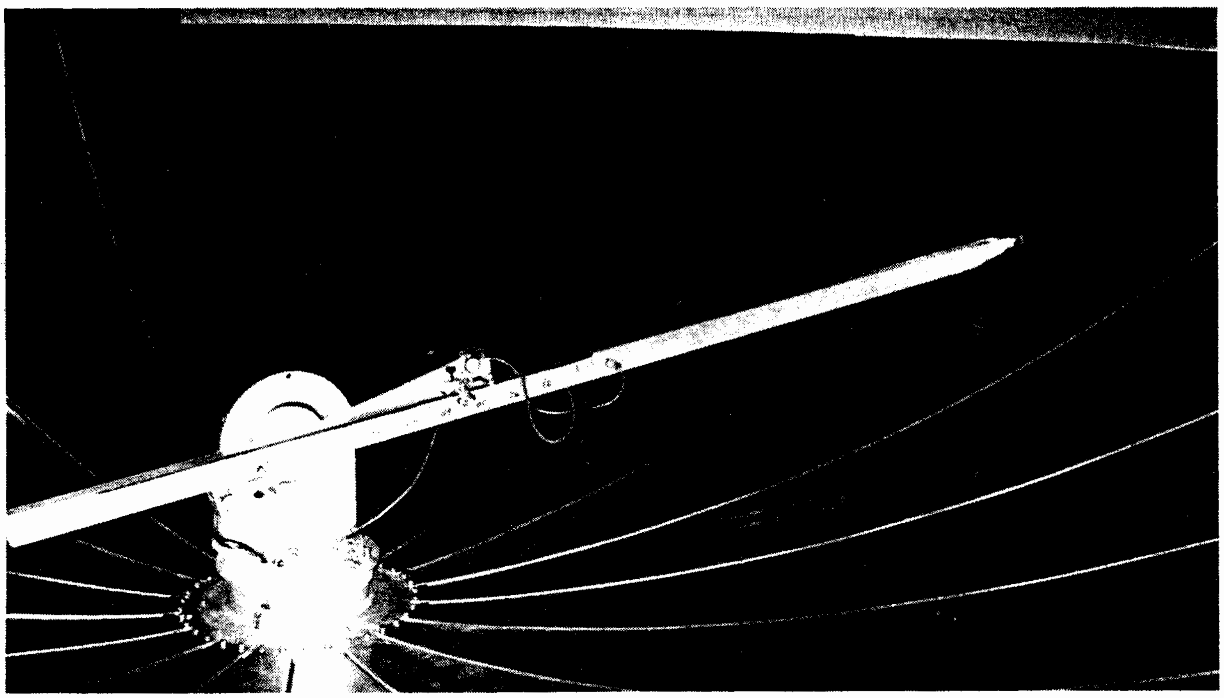


Figure 17. Automatic Surface Measurement System (ASMS) in a 7.7 Meter Reflector

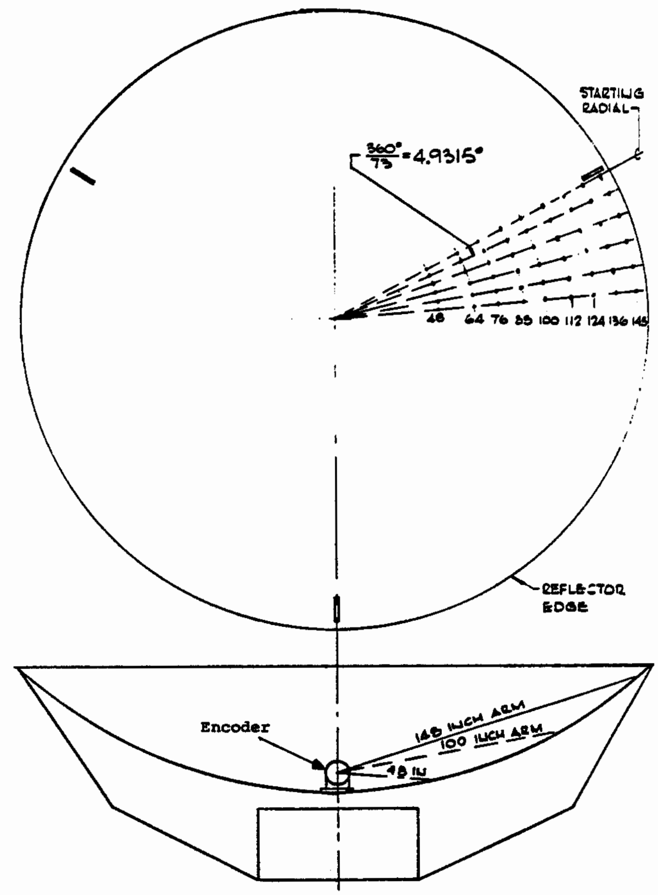


Figure 18. ASMS Installation in 5.5-Meter Reflector

A calibrated arm is caused to travel in a circle about the reflector axis by a motor-driven wheel which rolls along the reflector surface. As the wheel and arm circumscribe the reflector, variations in the reflector surface cause variations in the angle of inclination of the arm and variations in the encoder read-out. A synchro in the azimuth rotation axis provides a readout reference signal which is used to space the readings on the circle. Both the encoder and synchro outputs are coupled to a computer for decoding and processing (Figure 19).

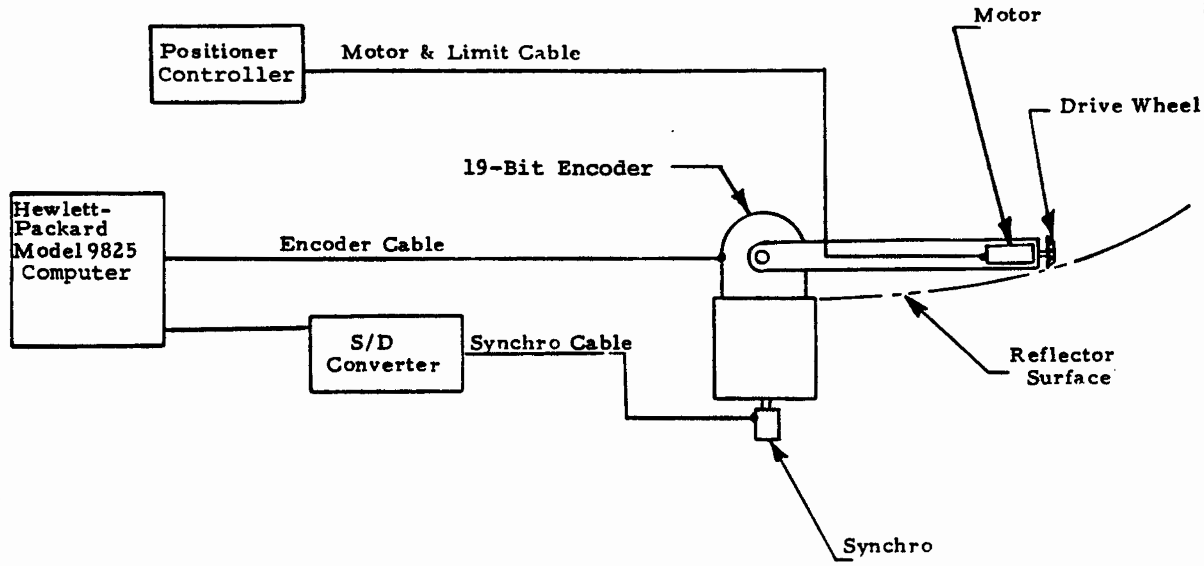


Figure 19. ASMS Block Diagram

Position in three-dimensional space is measured by the ASMS in spherical coordinates where:

θ = angle measured by the 19-bit encoder

ϕ = angle measured by the aximuth synchro

r = length of the calibrated arm

The measured data are utilized to compute the reflector surface deviations (rms, normal to the defined surface).

The advantages of the ASMS are the speed and accuracy inherent in the system. For example, on a single cut that takes less than 90 seconds, up to 300 elevation angles are automatically read by the 19-bit encoder and transmitted to the computer memory for storage. The possibilities of human reading and transcribing mistakes are eliminated.

SCIENTIFIC-ATLANTA

** 5.5 REFLECTOR RMS MEASUREMENTS **

S-A RUN # 5-0002R HUGHES S/N: 6/17/80 1:36P

	---BAR DATA---		-TILT FIT PARAMETERS-		DEVIATIONS NORMAL TO SURFACE ARE (inches):		
Bar	Length	Radius	Tilt Deg	Axis Az Deg	Prof	Irreg	Total
1	103.509	101.439	0.028	161.7	-0.002	0.017	0.017
2	91.512	90.575	0.029	131.6	0.003	0.012	0.013
3	32.029	81.687	0.029	161.2	-0.005	0.011	0.012
4	70.044	70.036	0.024	165 2	0.008	0.008	0.012
5	58.032	57.844	0.023	159.9	0.002	0.011	0.011
6	46.006	45.015	0.022	180.4	0.014	0.012	0.018
7	38.047	36.041	0.018	173.0	-0.007	0.012	0.014
8	34.048	31.310	0.019	162.9	-0.021	0.009	0.023
			Average:		For Total Surface		
			0.027	163.0	0.008	0.012	0.015

PROFILE FIT PARAMETERS
Focal Length = 65.234
Clinometer Z Origin = 10.165

SIGNED

INSPECTOR-BENSON

****

Figure 20. Tabulation of Computer Runs

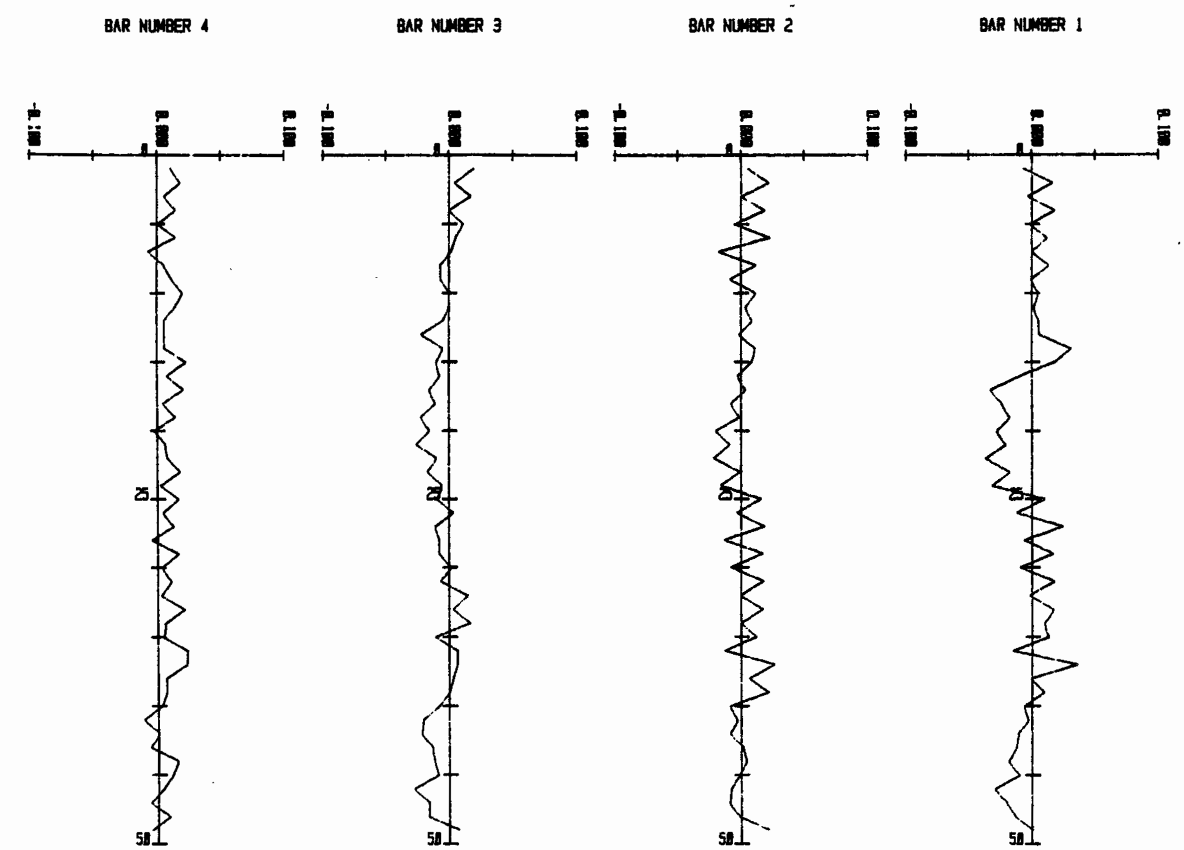


Figure 21. Tabulation of Computer Runs

SURFACE DATA

BAR # RAD	1	2	3	4	5	6	7	8
1	-0.006	0.006	0.019	0.011	-0.005	0.010	-0.007	-0.015
2	0.016	0.022	0.004	0.018	0.015	0.033	0.014	-0.003
3	-0.003	0.001	0.017	0.006	-0.007	0.013	-0.015	-0.021
4	0.018	0.019	0.000	0.015	0.018	0.031	0.004	-0.022
5	-0.001	-0.004	0.011	0.001	-0.025	-0.006	-0.032	-0.036
6	0.012	0.023	0.005	0.015	0.009	0.020	-0.004	-0.020
7	0.000	-0.017	0.001	-0.007	-0.026	-0.001	-0.030	-0.040
8	0.014	0.012	-0.007	0.006	0.013	0.025	-0.002	-0.020
9	-0.000	-0.008	-0.007	0.012	0.000	0.017	-0.013	-0.023
10	0.005	0.012	0.000	0.020	0.017	0.029	-0.002	-0.032
11	0.002	0.004	0.000	0.014	-0.012	0.006	-0.022	-0.039
12	0.006	0.009	-0.005	0.006	0.009	0.024	0.002	-0.016
13	0.006	-0.001	-0.022	0.006	-0.008	0.008	-0.009	-0.017
14	0.031	0.011	-0.005	0.006	0.006	0.016	-0.011	-0.026
15	0.016	0.008	-0.010	0.023	0.005	0.010	-0.010	-0.023
16	-0.009	-0.003	-0.007	0.008	0.009	0.023	-0.003	-0.023
17	-0.032	0.004	-0.016	0.021	0.011	0.027	0.004	-0.016
18	-0.024	-0.008	-0.011	0.005	0.009	0.026	0.012	-0.005
19	-0.017	-0.001	-0.022	0.014	0.007	0.021	0.007	-0.008
20	-0.028	-0.020	-0.016	-0.002	-0.006	0.013	0.005	-0.009
21	-0.021	-0.009	-0.026	0.006	0.000	0.016	0.007	-0.007
22	-0.036	-0.021	-0.010	0.008	0.006	0.013	-0.003	-0.018
23	-0.017	-0.001	-0.017	0.018	0.011	0.017	0.001	-0.017
24	-0.031	-0.016	-0.006	0.003	-0.000	0.013	-0.012	-0.024
25	0.010	0.016	-0.009	0.016	0.010	0.026	0.002	-0.015
26	-0.012	-0.003	0.003	0.005	-0.002	0.013	-0.008	-0.020
27	0.024	0.018	-0.011	0.013	0.011	0.022	0.006	-0.011
28	-0.006	-0.013	-0.008	-0.004	-0.002	0.010	-0.016	-0.027
29	0.017	0.017	-0.008	0.016	0.020	0.025	0.003	-0.017
30	-0.009	-0.008	0.002	0.004	-0.002	0.012	-0.014	-0.024
31	0.018	0.018	-0.007	0.011	0.012	0.026	0.013	-0.009
32	-0.002	-0.000	0.014	0.003	-0.011	0.009	-0.017	-0.023
33	0.017	0.017	0.003	0.021	0.013	0.023	0.003	-0.014
34	0.010	0.000	0.016	0.006	0.001	-0.003	-0.034	-0.044
35	0.013	0.012	-0.010	0.005	0.015	-0.019	-0.011	-0.024
36	-0.015	-0.013	0.007	0.024	-0.006	-0.013	-0.038	-0.044
37	0.036	0.026	0.006	0.022	0.021	0.016	-0.008	-0.022
38	-0.000	0.007	0.003	0.007	-0.012	0.011	-0.003	-0.018
39	0.010	0.022	-0.000	0.007	0.019	0.024	-0.004	-0.021
40	-0.006	-0.008	-0.008	0.003	-0.006	0.011	-0.008	-0.015
41	-0.003	-0.003	-0.021	-0.011	-0.001	0.008	-0.008	-0.022
42	-0.011	-0.009	-0.022	0.001	-0.011	-0.003	-0.013	-0.017
43	-0.012	0.002	-0.013	-0.006	0.004	0.018	0.001	-0.016
44	-0.018	0.005	-0.011	0.015	-0.001	0.009	-0.010	-0.017
45	-0.010	-0.001	-0.008	0.011	0.015	0.019	-0.002	-0.022
46	-0.029	-0.008	-0.028	0.004	-0.009	-0.006	-0.023	-0.038
47	-0.019	-0.009	-0.016	-0.006	-0.010	-0.002	-0.019	-0.033
48	-0.013	-0.001	-0.016	0.009	-0.004	0.002	-0.010	-0.021
49	0.001	0.022	0.008	-0.005	-0.001	0.023	-0.003	-0.018

Figure 22. Tabulation of Computer Runs

Communications, Protocols, and Polling

D. Mann

Communications for Distributed Networks

This paper is intended as a review of the general methods employed today with regards to communications for distributed networks. Discussion will center on the use of satellite networks and local networks. As these topics are quite involved, only a general overview will be presented. References cited at the end of this paper contain a more in-depth discussion of the subject matter.

The different types of ground-based communications available will be touched on lightly, with the use of examples from techniques currently being employed by Scientific-Atlanta whenever possible. Discussion of satellite-based communications techniques will center mostly on current theory employed in the research community.

Protocols that are discussed will include the usage of SAbus protocol for local network communications, and the use of variations of a theoretical type protocol for satellite networking.

A discussion of methods employed by Scientific-Atlanta for the polling devices in a local network will be presented, with examples provided on the capabilities from a recent project. In addition, satellite networking will be discussed with regard to the differences in methodology required in polling earth stations and sending commands.

Ground-Based Communications

Distributed local area networks can be organized in many different fashions. Some of the approaches used for network configurations are:

- Star network - each device is connected separately to a central controlling point in the network.

```
                O
            O   |
             \  |  O
              \ | /
               \|/
O---------------O--------------------O
              /  \
             /    \
            /      \
           O        \
                     \
                      O
```

- Bus network - or multidrop places all the devices in the network together on a single shared line.

```
     O      O      O        O
     |      |      |        |
-----+--+---+--+---+--+---+-+--+---
        |      |      |   |    |
        O      O      O   O    O
```

- Ring network - or loop where each device is connected to each other in a closed circular fashion, providing two path ways around the ring, going in either direction. Breaking the ring at one spot would not affect the network's functionality.

```
   O----------O
  /            \
 /              \
O                O
 \              /
  \            /
   O----------O
```

- Mesh network - any combination of any of the above networks, with each connecting node representing either a single device or a local network comprised of a star, bus, or ring design. Typically used over large geographical areas.

```
                      O
    O---O              \   O
   /     \              \ /
  O       O-------------O-----O
   \     /             / \
    O---O             O   \
                           \       O        O
                            \      |        |
                             O-------+----+---+-------
                                          |
                                          O
```

Devices in a network configuration can be connected in either a point-to-point fashion, or they may be multiplexed at some point in the network (generally to remote sites). Most often multiplexing is done over long distances to reduce cost and/or complexity, by concentrating the many lines used for devices into one line. Multiplexors are either time-division (TDM) or frequency-division (FDM) dependent techniques for splitting up the many incoming signals, over a single physical line.

Types of line connections made between devices are generally made via one or a combination of the following:

- Locally by means of a cable--either coaxial, twisted-wire pairs, or fiber optic.
- Line-of-site by either microwave, or laser.
- Via the telephone company (TelCo) on either a regular phone line or on a dedicated leased line connection.

Satellite-Based Communications

As a commercial user's needs grow, and as networks of earth station sites evolve, the need arises for less dependence on the common carrier's (TelCo) lines of communication. More emphasis is placed on taking advantage of satellites not only for the distribution of video and audio signals, but also for controlling the flow of signal traffic and using the satellite facilities as a means of interconnecting users of the network. By using the satellite as part of a shared distributed network, earth station affiliates may now obtain all of their programming via satellite, instead of land lines, thus reducing the portion of cost imposed by the TelCo's. Additionally, users can receive instructions on daily operation via satellite as well. If the network user is a receive-only earth station, then they must still rely on land lines to respond to other users in the network. However, transmit and receive earth stations may respond via satellite as well.

Some disadvantages to using satellites are that one can only cover a third of the earth's surface with a single satellite (in actuality this is generally less). No privacy is available in a satellite-based network without the use of data encryption techniques. Satellite channels may be inadvertently jammed or blasted by malicious or ignorant uplink earth stations. The channel can be saturated when peak traffic times occur. All of these problems occur in an equivalent fashion over terrestrial lines, so it should be recognized that they are not unique to satellites.

Some of the advantages to using satellite-based communications are that packets for more than one user are broadcast once. New users can enter the network without need to rewire any portion of the existing network. Rapid reconfiguration of the network is possible as well as being easily expandable. Mobile users are handled without troubles. Packet switching is handled without the need for packet switches.

Uplink Facilities. A major component of any satellite-based network is the uplink earth station facilities. These facilities serve as a hub for a local or regional area network. All receive-only stations will generally connect to the closest uplink via a ground line, so that they may respond to polling of status and commands if need be. With regard to packets of data sent over the satellite network, an uplink site has the ability to monitor satellite traffic for collisions on the transmissions it makes with other uplink sites.

This would generally be avoided by a scheme of round robin discrete time intervals, in which each uplink station would transmit commands, or status of receive-only users in its local network. Collision avoidance can also be handled by putting different uplink transmissions on different channels on the same satellite, or putting the transmission on different satellites altogether. Being able to monitor other receive-only stations and control them with commands from the uplink allows greater control over the functioning of a network, and the routing of packets through the network.

Downlink Facilities. Downlink facilities are earth stations equipped only with the ability to receive signals, be they video or audio. A facility must still use a ground link to communicate with other users in the satellite based network. Receive-only earth stations are far less costly than uplink facilities and are therefore more affordable for the smaller affiliate stations.

Protocols

Various different types of protocol exist today, but this paper will discuss two of interest--the multidrop protocol used for local networks (an example of which is the SAbus protocol) and a theoretical protocol used for satellite networks.

Multidrop Protocol

The multidrop approach to local networking follows what is typically known as "statistical multiplexing using centralized polling."[2] The polling controller or master device on a multidrop bus will request status from each of the slave devices on the bus (twisted-wire pair cable for an SAbus). Upon receiving a poll for status from the master device, a slave will respond within a set number of ms with data describing briefly the slave device's current state. If a state change has occurred since the last poll, then the master device can poll for additional information. The master device then moves on to the next device on the bus. Each time a poll is made the slave device has to be identified via an address specification, which is unique for each slave device. Overhead required for a poll includes information on the start of text, an identifying slave device address, the data itself, an end-of-text identifier, and a checksum value. The data sent in a poll, and responded to by the slave device, generally consists of a command followed by any pertinent parameters needed. According to R. D. Rosner, "the key point is that polling is a prime example of a system which uses dynamic, rather than fixed, assignment of the capacity of the transmission line or shared medium. The available capacity is shared, following some formalized assignment procedures among the active members of the user community, and little or no capacity is wasted by the users that are currently inactive."[2] In the case of the multidrop bus, slave devices simply listen to each poll message on the line and ignore any messages that are not addressed to itself.

Typical master devices can include the following:

- A Master Control Computer - Has the ability to centrally control all elements of remote and local earth stations. Can easily be adapted to control various equipment protocols, most commonly via an RS-232C cable.
- An Earth Station Controller
- A Video Protection Switch
- A Uplink Protection Switch

Typical slave devices can include the following:

- A Video Protection Switch
- A LNA Protection Switch
- A Uplink Protection Switch
- A Receiver
- A Exciter
- A Antenna Controller
- A Klystron Tuner
- A High-Power Amplifier
- A Waveguide Switch Interface
- A device connected to the bus via software and/or hardware driver interface.

Satellite Protocol

A feasible approach to providing a satellite network protocol would be a slotted aloha protocol technique. To describe briefly its approach, one should know that the following must be employed:

- Establish a synchronized time interval over a satellite channel, according to the time to transmit the length of the largest packet message.
- Broadcast a packet only at the beginning of this discrete time interval, for example every 20 ms.

- Provide a round robin polling approach, with the ability to reprogram the order of poll and the priority.

- The packet itself would contain a (hello, message, goodbye) type of structure.

- Collisions occasionally occur due to quarter second (250 ms) delays in transmissions.

- Uplinks would listen to their own messages and compare with what was transmitted; if a collision has occurred, then it would retry the transmission when its time slice comes up again.

- "Packet narrowcasting" would be employed, which implies that the packet contains source and destination addresses, as well as a repetition counter for repeat transmits of the same packets.

- "Capture effect" could also be used, where different uplinks would broadcast signals at differing power levels, thereby providing a pseudo-priority basis for resolving collision conflicts. In other words, stronger signals to the satellite provide a higher priority by increasing the probability that its packet would get through. If the highest priority makes it through the satellite transponder successfully, then you have one less uplink packet to retransmit in the event of a collision.

Polling and Commands

Local SAbus Networks

In a local area network, polling performs a request for status in a round robin fashion, over a twisted-wire pair cable. Each poll expects a response, within a specified time frame, from the device. In order that future requirements be accommodated, the current polling module being used for projects has been designed to allow for the polling of SAbuses (or in fact any type bus interfaceable through a DEC compatible interface board, RS-232C, RS-422, etc.,) in any order, as well as poll the devices on a bus in any order (Figure 1). The frequency with which devices or buses are polled can be handled in the same variable fashion. The method used was to place bus numbers in a variable size array which can have the buses arranged in any order; in addition, one is able to repeat the same number in the array several times (effectively increasing the frequency of polling that particular bus). Therefore, the buses can be prioritized and/or have the frequency of polling an individual bus increased at the expense of others. The same approach was applied to the polling of devices on each bus, providing the ability to vary both the order in which devices are polled and the frequency with which individual devices are polled.

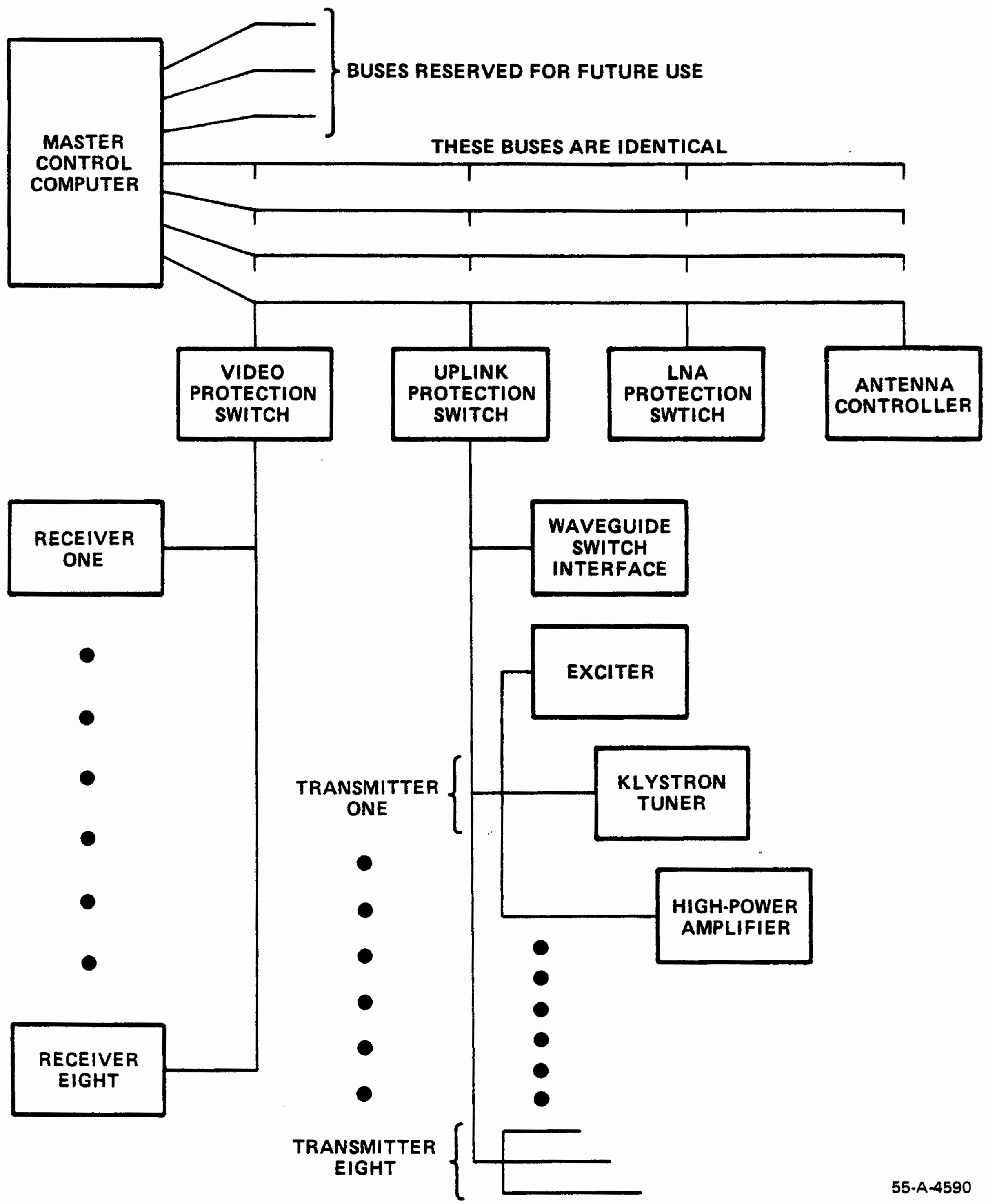


Figure 1. Local Network Diagram

Satellite Networks

Satellite-based network polling must be handled in a different fashion than ground-based networking, due to the nature of communications. Because of delays in transmission and the fact that not all earth stations have an uplink facility, a more applicable approach might be the use of a slotted aloha protocol. In polling the stations, a response would not be required if the master earth station or controller broadcast what it assumed to be the current state of each station, and then received a response from the slave stations only when the state changed. This change notice could be supplied via land lines for receive-only earth station, and via satellite for transmit and receive stations.

Commands over satellite networks can be handled in similar fashion, whereby a response is given from the slave earth station only when a problem exists in meeting the demands of the command. Additionally, commands can be issued which are either global (pertaining to all stations), group (pertaining to a subset, i.e., regional), or specific (pertaining to a single earth station). Commands could also be submitted to all stations hours in advance of their acutal execution to ensure their proper execution at the proper time. This would also assist in cutting down on unnecessary traffic during peak times. See section 2D-4 regarding the event scheduler. In addition, one can prioritize traffic such that the most critical packets of commands get through fastest, with precedence over polling packets and other low priority commands.

To provide privacy to all, the transmission of polling and command data packets encryption may be employed. It is possible to protect, via software encryption, both the status report of stations and the global, group, or specific commands to the network. Data can be protected from outside eavesdroppers on the network, as well as providing protection for stations within the network itself. This is achieved by providing as many access keys and/or encryption methods as is necessary to keep outside, group, or specific user eavesdropping from occurring. Electronic signatures can be changed on a frequent basis to provide protection against the accidental (or otherwise) disclosures of the access keys.

References

[1]Uyless D. Black, Data Communications, Networks, and Distributed Processing, Reston Publishing, TK5105.B576, 1983

[2]Roy D. Rosner, Distributed Telecommunications Networks via Satellites and Packet Switching, Lifetime Learning Pub., TK5105.R66, 1982

[3]James Martin, Telecommunications and the Computer, Prentice-Hall, TK5101.M326, 1976

[4]James Martin, Teleprocessing Network Organization, Prentice-Hall, TK5102.5.M293, 1970

[5]Communication Satellite Systems: An Overview of the Technology, IEEE Press, TK5104.C6, 1975

Glossary of Terms

Checksum - A method of verifying data correctness by summing the bits in a message and creating a unique sum. This sum is placed at the end of a data message sent by one device, and then verified later on by the receiving device.

Collision - When data is transmitted by one device in a network at the same time another device transmits over the same channel.

Local Network - High-speed communications channels interconnected in a network for a limited geographical area.

Multiplex - Connecting many devices in a network into the same channel, and sharing the use of that common channel.

Packet Switching - Real Time processing and routing of data messages between devices.

Point-to-Point - Connecting two devices in a network directly to each other, allowing no other devices on the channel.

Protocol - Definition of how the devices in a network communicate with other devices, or the rules devices are required to follow while talking over network channels.

Round Robin - Communicate with each device in a network by following a set order or sequence, and then continually repeat that same sequence.

Command Language or Menu-Driven Control

S.N. Cole

Introduction

As the sophistication of satellite earth-station equipment increases, it becomes correspondingly more difficult for the operator to monitor and control such equipment without a central work station. For example, a recently installed major uplink facility consisted of four subsystems (expandable up to seven subsystems), each of which is designed to contain the following controllable hardware modules:

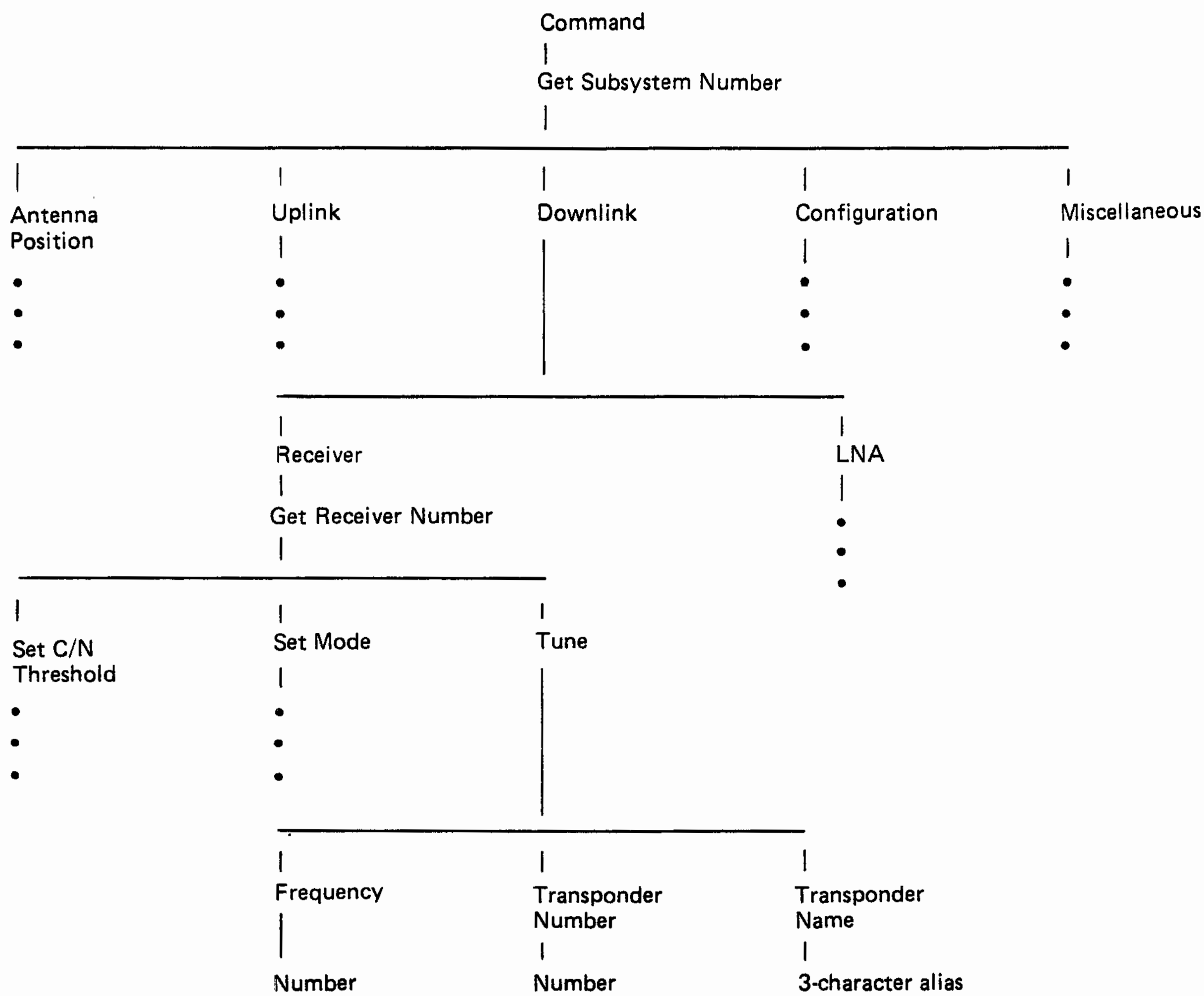
- 1 Antenna Positioner
- 1 Uplink Protection Switch
- 1 Downlink Protection Switch
- 1 LNA Protection Switch
- Up to 8 Receivers
- 3 LNAs
- Up to 8 HPAs
- Up to 8 Exciters
- Up to 23 Waveguide Switches

A digital computer with a keyboard/display unit makes it possible for the operator of systems such as the above to perform the monitoring and control functions easily and reliably. The associated software provides facilities in several distinct functional areas. This paper focuses on one of those--the facility to read and interpret commands from the operator.

The following sections discuss various issues pertaining to the specification of the operator-interface facility. Two equally important points of view are represented: (1) the operator's point of view (ease of use), and (2) the programmer's point of view (cost). Fortunately, these two points of view are in agreement with regard to one important issue--unnecessary complexity imposes a burden both upon the operator and upon the programmer and should be avoided.

Tree Structure of Command Syntax

The syntax of the commands used by the operator to express a complete instruction can be described abstractly as a tree. For example, the tree might assume the following form:



An important fact to observe about the tree is that the operator-interface program can learn the operator's intent by successive stages of refinement. This holds true both for menu-driven and for command-language systems as long as the syntax can be represented by a tree structure. A tree is also a finite-state machine, and this fact can be exploited to design the operator-interface program. The initial state of the finite-state machine coincides with the root of the tree (the node labeled "Command"). The program uses the first input after the subsystem number to determine which major branch to traverse; if the operator indicates "Downlink", the finite-state machine undergoes transition to a state that coincides with the node labeled "Downlink". Transitions continue in this way until the finite-state machine reaches a final state, which coincides with one of the leaves of the tree; at this point the operator-interface program "knows" the operator's intent and can act accordingly.

Menu-Driven System Control

A menu-driven system is characterized by a series of query-response exchanges between the computer and the operator to express one command. The computer's query is usually in the form of a menu displayed on the operator's console, which lists the valid responses and their meanings. The operator's response is supposed to be one of the possibilities listed in the menu (usually a digit or a letter).

For example, a menu-driven system based on the tree in the previous section would start with the query

```
ENTER SUBSYSTEM NUMBER (1 .. 7) >
```

Assume that we have a sophisticated menu-driven system, which displays (in addition to the queries) the status of the subsystem, module, or component that is currently selected. Then after the operator types in a digit designating the subsystem, the computer would display the overall status of the subsystem followed by

```
A  ANTENNA POSITIONER
U  UPLINK MODULE
D  DOWNLINK MODULE
C  CONFIGURATION
M  MISCELLANEOUS
Z  RETURN TO FIRST-LEVEL MENU

ENTER COMMAND TYPE >
```

Note the last possibility. Menu-driven systems should permit the operator to return to the first-level menu (ENTER SUBSYSTEM NUMBER >) quickly and easily in response to any query. (This facility could also be provided by means of a special key -- see the section entitled "Deletion" below.) Suppose that the operator responds by typing 'D'. Then the computer would display status information about the Downlink Module of the given subsystem followed by another menu inviting the operator to select a receiver, to select an LNA, or to return to the first-level menu. If the operator selects a receiver, the computer would continue by asking for the receiver number. After the operator supplies the receiver number, the computer would display detailed information about the specified receiver followed by another menu inviting the operator to select an action -- set C/N threshold, set mode, tune the receiver, or return to first-level menu. If the operator elects to tune the receiver, the computer would ask whether the operator wishes to tune it to a specified frequency, to tune it to a transponder designated by number (1..24), to tune it to a transponder designated by alias (for example "1H" or "12V"), or to return to the first-level menu. Finally, if the operator elects to tune it to a frequency, the computer would ask for the frequency and check the response for validity.

There are two levels of sophistication with regard to the status displays. The simpler approach would be to paint the status information only after the operator completes his response to a query. The more sophisticated approach would be to update the display (while waiting for the operator's response) when pertinent information changes. The first approach is usually unsatisfactory, because the information on the display can be out-of-date and, therefore, misleading. For example, if several minutes have elapsed since the most recent operator response, and if the carrier-to-noise ratio has dropped significantly, the operator would not be aware that remedial action is needed.

A menu-driven system can also be used to display an array of values and permit the operator to update one or more of them. For example, the configuration branch of the tree may contain a screen on which the operator would display or update the coordinates of several satellites. In order to update information, the operator needs a method to express which element of the array to change. A simple solution would be a dialog such as the following. The computer asks the operator for the number of a satellite to change (or a code that indicates "return to first-level menu"). The operator types the satellite number n. The computer asks for the coordinates of satellite n (or a code that indicates "return to first-level menu"). The operator types the coordinates. The computer updates the display of satellite coordinates and repeats the query for the number of the next satellite to change. An alternative solution would permit the operator to move a cursor on the screen by means of function keys. The operator would then move the cursor to the satellite coordinates that need to be updated and overwrite them. The first solution can be cumbersome and annoying to the operator. The second solution is considerably more expensive to program. The trade-off decision should take into account how often the operator needs to update the array of information.

Command-Language System Control

A command-language system is characterized by a single exchange between the computer and the operator to express a complete command. The computer's query usually consists of one or two characters. For example, the query "C>" could be used to indicate that the computer is expecting an ordinary console command, whereas the query "G>" would indicate that the computer is expecting a command from a set of commands that could include graphics instructions. The operator's response consists of a line of text that expresses the complete intent of the command.

Each command-language command is mnemonic. An example of a command derived from the tree illustrated above might be of the form "SUBSYSTEM 5 DOWNLINK RECEIVER 3 TUNE TRANSPONDER NUMBER 9". Immediate improvements in readability can be obtained by employing abbreviations: (1) let the subsystem number be denoted as a command prefix terminated by a colon (:); (2) let the keyword "RECEIVER" imply the downlink branch of the tree. Then, the abbreviations yield "5: RECEIVER 3 TUNE TRANSPONDER NUMBER 9". Rearranging the tree yields further improvements in readability: let the first branch after the subsystem be an imperative verb ("MOVE", "SET", "TUNE", "CONFIGURE", "ENABLE", etc.). Then, the above command could be written in the form "5: TUNE RECEIVER 3 TRANSPONDER NUMBER 9".

In menu-driven systems it was noted that status could be displayed for the currently selected subsystem, module, or component with successively greater detail as the dialog proceeds from the root of the tree toward one of the leaves. The counterpart for this in command-language systems (to display the subsystem, module, and component in succession as the operator types the command) does not make much sense. Instead, the command language typically includes specific commands that the operator can use to indicate the type of information to be displayed. The area of the screen for echoing keystrokes and for displaying error messages can be as small as two lines. This leaves the remainder of the screen for status displays.

Implementation Techniques

The process of interpreting a command (or, for menu-driven systems, a series of commands) is called parsing. After parsing a command, the computer will know (1) whether the command is valid and, if valid (2) the operator's intent. The Parser routine is not directly concerned with the execution of the command. Indeed, one of the possibilities is that the command is not to be executed immediately; instead, the command is to be inserted into a list of commands (sometimes called an "event schedule") for execution at a specified subsequent time. In this case, the parser is used merely to verify that the command is valid.

A straightforward way to implement the parser is to let the program structure mirror the structure of the tree. Branches in the tree would be implemented by means of corresponding multiway branches in the program. High-level subroutines could enclose the code for each of the major branches attached to the root of the tree. These subroutines could, in turn, call lower level subroutines that correspond to branches closer to the leaves.

Another way to implement the parser is by using tables. This approach takes advantage of the fact that the tree is a finite-state machine, and that there are well-known techniques for using a table to represent a finite-state machine. The table will generally contain transition rules and other instructions to the parser. An example of a transition rule would be an entry in the table that instructs the parser to perform one of the following three actions: (1) read the keyword "ENABLE", and proceed to state 31; (2) read the keyword "DISABLE", and proceed to state 75; (3) read an integer between 1 and 7, and proceed to state 19. If the operator does not type "ENABLE", "DISABLE", or an integer between 1 and 7, the parser should print an error message. An example of a command other than a transition rule would be to record the subsystem number from the most recent integer read, and then proceed to the next higher-numbered state. This rule, for example, might occur at state 19, in which case the subsystem number in the range one through seven is known to the parser. These examples illustrate that the table-driven parser is essentially an emulator which executes a higher-level program (the language description) built into the table itself.

It is not obvious which of these two approaches is easier to program. The first approach appears to be simpler. However, if the syntax is sufficiently complicated, a parser using the first approach will be larger, and it is well known that the management of large programs introduces unexpected problems, especially if memory limitation becomes an important factor.

Desirable Features

Uniformity

A uniform set of conventions should be employed for computer cues and operator response. This will make it easier for the operator to become familiar with the program and (even for experienced operators) will reduce the risk of operator error.

In menu-driven systems, the menu at each level should be displayed at the same place on the screen. Moreover, employing page numbers and uniform conventions in screen titles will be of benefit to the operator.

In command-language systems, parallel structure should be designed into command format. Consider, for example, two typical commands in an earth-station system:

(1) to tune the third receiver of subsystem five to the ninth transponder of the current satellite;

(2) to set the C/N threshold of the third receiver of subsystem five to 15 dB.

If the first command were expressed in the form

"5: TUNE RECEIVER 3 TRANSPONDER 9".

It would be inappropriate to express the second command in the form

"5: SET C/N THRESHOLD 3 15".

Instead, a preferred form would be

"5: SET RECEIVER 3 C/N 15".

Batch Files

Batch files are large collections of commands on disk, which are used to retain information that the operator modifies only infrequently (e.g. configuration parameters or the event schedule). It is possible to provide batch files for menu-driven systems, but the programming is not easy. A partial solution is for the computer to record a file copy of the commands while the operator is keying them in. However, this solves only the problem of initial data entry and does nothing to alleviate the difficulty of updating the batch file. Editing the file is either a very sophisticated operation -- both for the operator and the programmer, or the operator will have to use an ordinary text editor to modify a very cryptic file.

Abbreviations or Shortcuts

The complete forms of keywords such as "RECEIVER" and "TRANSPONDER" are useful in the operator's manual and, perhaps, for the inexperienced operator. However, after some experience in using the system the operator will soon become impatient with the amount of typing that is required and will demand some form of abbreviation or shortcut to expedite command entry. The need for abbreviation is a problem only in command-language systems, since menu-driven systems are inherently brief.

Ideally, the computer should accept any abbreviation using the first n characters of a keyword as long as the keyword (or satellite name) is unambiguous. The computer should take advantage of the complete context of the keyword (i.e. the previous text in the command line) to interpret the abbreviation. A somewhat simpler (albeit less satisfactory) alternative would be to require, say, the first three characters in abbreviating a keyword or name. However, such compromises might restrict the vocabulary of the command language, since it would require all keywords and names to be unique in the first three characters. For example, if the abbreviation were based on the first three characters, one of the two keywords "DISPLAY" and "DISABLE" would have to be eliminated from the vocabulary.

A useful feature to provide with abbreviation is the display of the unabbreviated keyword. For example, after the operator has typed "TUNE R", the computer knows that the second keyword is "RECEIVER". Therefore, if the operator types a space after the "R" (to indicate end-of-abbreviation), the computer could echo "ECEIVER".

Validation Based on Current Antenna Direction

Errors in a command should be caught as early as possible. Therefore, if the command refers to a transponder not present on the current satellite (the satellite closest to the antenna direction), that command should be rejected. More correctly the computer should use the direction expressed by the last MOVE ANTENNA command instead of the current antenna direction, because the operator may want to tune the receiver to a transponder while the antenna is slewing. The recommended way to compute the closest satellite is to calculate the great-circle angle difference between the (azimuth, elevation) coordinates of the antenna and the (azimuth, elevation) coordinates of each satellite; then choose the satellite whose angle is smallest. The following formula gives the cosine of the great-circle angle.

$$\cos(\text{angle}) = \cos(EA - ES) - \cos(EA)\cos(ES)\,[1 - \cos(AA - AS)],$$

where

EA = elevation angle of the antenna
ES = elevation angle of the satellite
AA = azimuth angle of the antenna
AS = azimuth angle of the satellite

The closest satellite is the satellite for which the above expression is maximum.

Early and Clearly Expressed Error Messages

If the operator makes an error during command entry, and the computer can detect the error, it should print a diagnostic as soon as it is feasible to do so. This kind of close interaction with the operator is inherent in menu-driven systems. On the other hand, programs based on the command-language approach often neglect to provide such interaction, because they do not retain control during keystroke entry. Most high-order languages provide a keyboard/display driver, which performs functions such as keystroke echoing, responding to the delete-character or delete-line function key, and numeric conversion. When the application program uses the driver, it does not regain control until the operator has completed keyboard entry and until the driver has already partially digested the command line. On the other hand, the operating system or the high-level language usually provides some kind of mechanism that allows the application program to process each keystroke (although such a mechanism is often obscure and difficult to ferret out of the documentation). In exchange for the greater degree of control, the programmer must forego the automatic features provided by the driver; thus echoing, deleting, conversion, etc. must be performed in the application program.

The computer should print a diagnostic as soon as an error is detected and require the operator to correct the erroneous portion of the command before proceeding with the remainder of the command. Immediate error reporting is recommended for the following reasons.

- It improves the readability of diagnostics and relieves burden on the operator of interpreting ambiguous messages such as "WORD 4 OF COMMAND NOT RECOGNIZED".

- It makes it unnecessary for the operator to retype the entire command just to fix an error in one portion.

- The operator can be confident that the initial portion of the command (already typed) is correct.

There are two levels of sophistication. The simpler of the two would be to check for validity only after every word of a command (e.g. keyword, name, or numeric parameter). The more sophisticated approach would be to check for validity after every keystroke. With regard to keywords, these two methods are essentially the same for menu-driven systems, because keywords are normally represented by one-character codes. However, checking numeric parameters or satellite names while they are being typed increases the cost of the programming for menu-driven systems as well as for command-language systems. Numeric parameters are often diagnosed with respect to a valid range of values. In this case, checking a numeric parameter while it is being entered does make sense; it amounts to checking whether the partially typed number exceeds the upper limit of the range. On the other hand, it would be considerably more difficult (and probably not worthwhile) to check uplink and downlink frequencies, because only a discrete set of frequencies is valid. In this case, it makes more sense to read all of the digits of a frequency parameter before checking whether the value is valid for the current satellite.

Help

Menu-driven systems automatically provide help by displaying a cue while soliciting data entry. The counterpart in a command-language system would be a facility in which the operator requests help by typing a special key (for example "?"). The computer would answer this request for help by displaying a list of the operator's valid responses at this stage of command entry. Possible types of cue would include the name of a keyword, the range of values for a numeric parameter, an indication that the response could be a satellite name, an indication that the response could be an exciter frequency, etc. If the computer is checking for validity after every keystroke, then each keystroke should reduce this list of cues. In other words, if the operator requests help in the middle of a command word, the computer should answer by displaying only the abbreviated list of words that would be obtained by completing the partially typed word.

Deletion

The operator needs the ability to correct a typing error by deleting all or part of a command. He typically expresses his intent to delete all or part of a command by using one of the special keys: the [delete] key, the [backspace] key, or one of the control characters (entered by holding the [ctrl] key down while typing one of the letter keys). Although control characters are harder to type, this is usually a virtue, because it reduces the risk of the operator's deleting part of a command by mistake.

In command-language systems three delete keys are recommended:

<delete character> -- delete the last character from the end of the partially typed command, and restore the display to the state that it was in prior to this last character. For example, assume that the operator had typed "TU R " and that the computer had echoed "TUNE RECEIVER". Then, in response to <delete character>, the computer should erase "ECEIVER" and revert to the state that it was in prior to reading the last space.

<delete word> -- delete one or more characters from the end of the partially typed command until the entire command is deleted, or until a nonspace character preceded by a space is deleted. (This definition may be extended to include separator characters other than space, e.g. comma, colon, equal sign, etc.) Restore the display to the state that it was in prior to this last word.

<delete line> -- delete all characters from the partially typed command.

Of course, the computer should accommodate a combination of multiple <delete character>s and <delete word>s in succession.

Three levels of deletion, similar to the scheme described above, would also be appropriate for menu-driven systems.

<delete character> -- delete the last character typed, unless this last character terminated a portion of the command. For example, assume that the computer displayed a menu that solicited a response in the form of a number between one and five. After the operator types "3", he can use the <delete character> key to erase the "3". On the other hand, after the operator types "3" followed by [RETURN], the computer displays the menu at the next node of the tree, and the <delete character> key could not be used to return to the previous menu.

<return to previous level> -- if any responses have been typed in for the currently displayed menu, erase all such responses; otherwise, return to the previously displayed menu.

<return to initial level> -- return to the highest level menu. Using a function key to express "RETURN TO 1st LEVEL MENU" is recommended in preference to including this option as a separate line on each menu.

Flexibility in Syntax

The operator's context may determine the vocabulary of commands that should be valid.

- VIEW commands, e.g. VIEW RECEIVER 3, are valid only at a color graphics terminal;
- the INSERT command is valid only while the operator is editing the event schedule;
- the command EDIT EVENT SCHEDULE is not valid while the operator is editing the event schedule;
- when the computer encounters a TUNE EXCITER command, it should print the cue ARE YOU SURE?, except when during the execution of a command from the event schedule.

The table-driven implementation is particularly appropriate in this situation. Four tables are maintained: (1) for the syntax at the operator's terminal; (2) for the syntax at the color graphics terminal; (3) for the syntax during the editing of the event schedule; (4) for the syntax during the execution of commands from the event schedule.

Documentation

Drafting an easy-to-read users' manual is a significant portion of the effort in producing an operator-interface program. The cost of this effort should not be forgotten when the program is being designed. Brevity of expression should be exercised wherever possible, because this will reduce cost as well as improve ease of use. Unfortunately it is often difficult to be concise in describing a menu-driven system, because the prompts displayed to the operator change as the program traverses the tree. Command-language systems can be described concisely, but the operator needs to learn syntax conventions. For example, "Language elements in square brackets represent items that may optionally be omitted," and "Language elements separated by vertical strokes represent alternative choices." In many cases, the meaning of a command is not self explanatory, and it is necessary to describe the effects of the various alternatives.

Graphics

C.A. Gregory

Graphics System in an Earth Station Control Environment

Computer graphics is a user friendly and effective means of communication between user and computer. Currently, most remote control earth stations are monitored by an alphanumeric display viewed on a video display terminal. Earth station control can be enhanced by the inclusion of a computer graphics system in the control environment. Data can be presented on a graphics display terminal in a manner that is easily and quickly understood by the user. Even though it is the intent of alphanumeric displays to be user friendly, it is simpler for a user to comprehend data presented as colors and patterns than it is to interpret a table of numbers. By presenting earth station status as a diagram or view on a graphics display terminal, the earth station operator can better assimilate the displayed information required for real-time decisions.

Computer graphics offers more than the ability to encode information into color and pattern. It is also possible to draw a representation of almost any object on a graphic screen in whatever form desired. Another feature is zoom, which allows the user to 'blow up' a portion of the screen to a larger size. When these techniques are employed with color and patterns in designing the graphics system required by the user, a good human interface for control system users can be created.

Graphics Requirements in Earth Station Control

The graphics system which displays information to the operator is subject to guidelines and requirements as is any software system. Simplicity, completeness, and performance are requirements that are closely related. Simplicity and completeness are important when displaying information to the user. Information must be concisely presented to the operator and it must also be complete. The operator should not have to guess at details because they are not present on an earth station diagram.

Simplicity in a graphics control environment is always desirable. Simplicity of design leads to simplicity in implementation and use. If a graphics system is encumbered with a complicated design, the system will be difficult to implement and risks being unmanageable for the user. Simplicity can be undone by completeness if the user desires a detailed system. Complete data is necessary but must be presented so it does not burden the user with unnecessary details. Data to be displayed must be chosen for its relevance and importance so that an excess of detail is not presented.

There is a balance to be achieved between completeness and simplicity. If simplicity is overemphasized, details needed for the operator to do the job may be lacking. If completeness is overemphasized, the operator may be burden with the task of deciphering details presented on the terminal screen before changes in system status can be recognized. Clearly, an equal measure of simplicity and completeness is necessary when designing a graphics system.

Performance can be affected by completeness and simplicity. If the graphics system has the good balance of simple and complete displays, performance should be adequate. But if completeness is overemphasized, processing time may be increased in order to update the detailed graphics representation of earth station status. Since operators in an earth station environment frequently have the need to make timely decision based on the reported earth station status, it is desirable to minimize the time needed to refresh the graphics screen or process the data.

The system design requirements of simplicity, completeness, and performance, when implemented effectively, result in a system easy for the operator to use and understand. Graphics displays are used to enhance communication between operator and the earth station in a control environment. Meeting the criteria which defines an effective graphics system assures a system that is easy to use and understand.

Implementation of Graphics Systems in a Control Environment

The earth-station graphics system is a simple and straightforward presentation of earth station information. There are two views representing the uplink (Figure 1) and downlink (Figure 2), either of which can be displayed on the graphics terminal. On the left side of the downlink view are three antennas with the name of the satellite which the antenna has acquired. To the right of the antennas are three columns each representing one separate downlink module. At the head of the column is a protrayal of a Scientific-Atlanta 7620 Video Protection Switch. Under the Video Protection Switch are eight boxes representing each of the eight receivers associated with that Video Protection Switch. The uplink is drawn similarly, the difference being a protrayal of a Scientific-Atlanta 7640 Uplink Protection Switch with up to eight transmitters beneath.

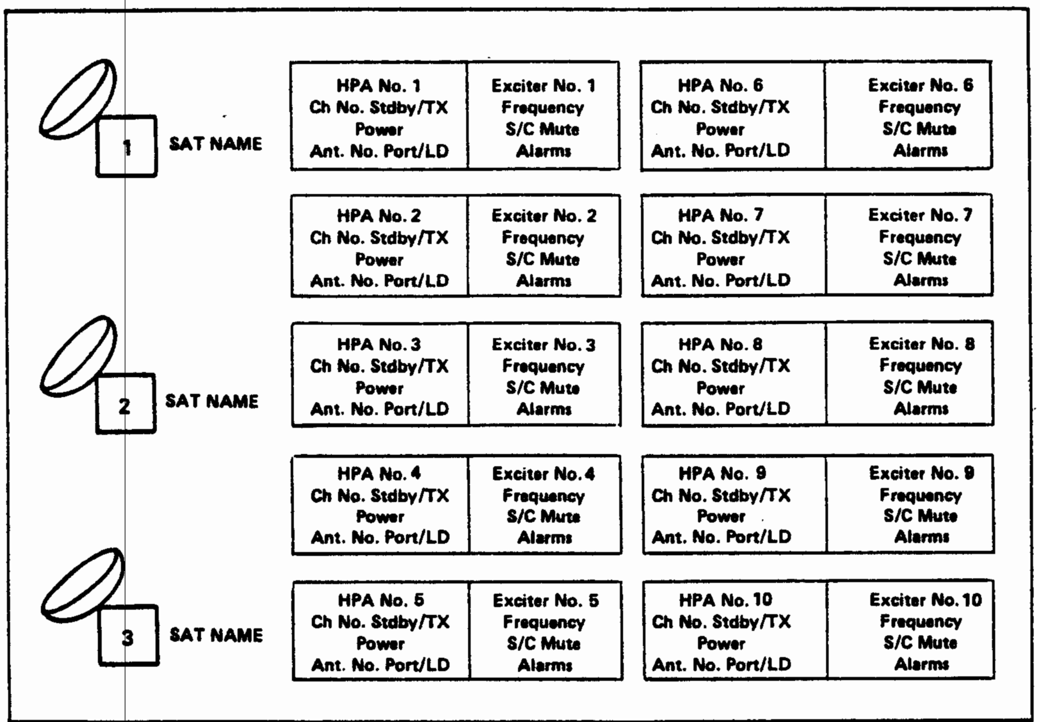


Figure 1. Uplink Display

RED →	ALARM
GREEN →	NORMAL
YELLOW →	PROTECTED

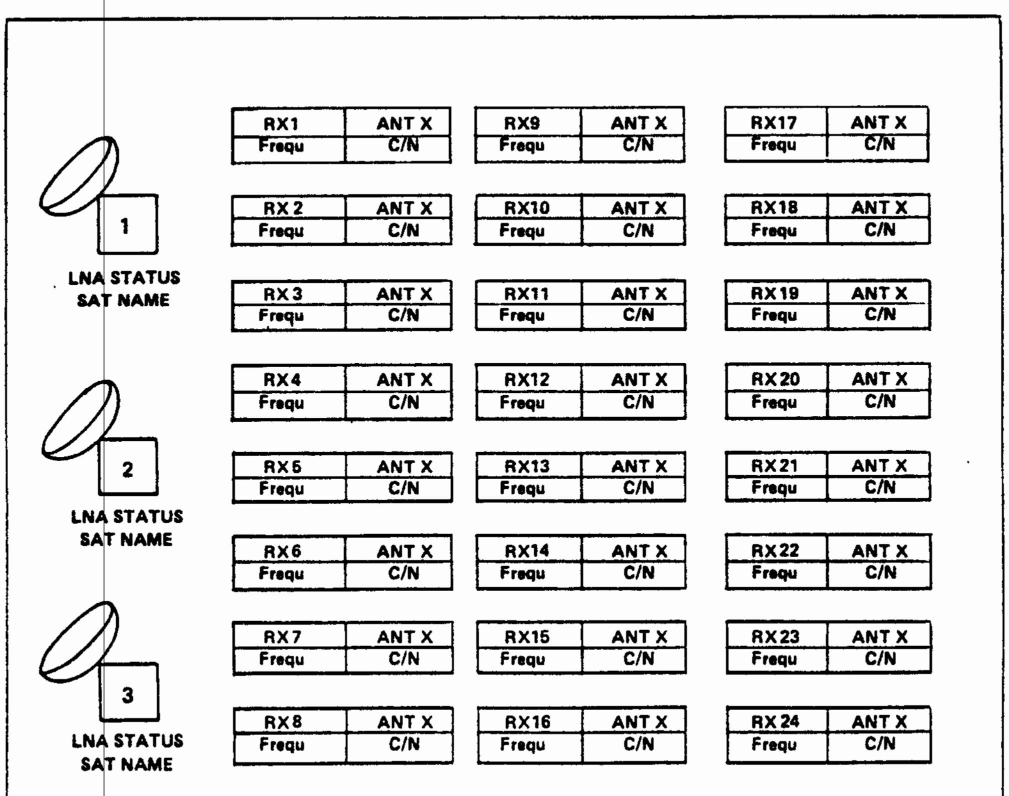


Figure 2. Downlink Display

RED →	ALARM
GREEN →	NORMAL
YELLOW →	PROTECTED

There are four colors utilized: green, red, yellow, and blue. Green represents a normal state; alarm states are either red or yellow. Blue is used for a background color. A component is drawn in the background color if it does not exist for that downlink/uplink module.

An operator has the information needed displayed on either one of the two views. By glancing at the view the operator can quickly tell the status of the entire earth station. Any changes made by the operator will be displayed on the view. Earth station information is updated continuously when one of the views is displayed giving the operator timely changes in earth station status.

Earth station graphics provided for one project has more detail due to the requirements specified by the user. By requesting a higher level of detail, the user sacrificed simplicity in view design and performance of the graphics system in order to gain a complete and precise picture of the earth station.

The graphics system has view hierarchy. At the uppermost level of hierarchy is a concise picture of the entire earth station called the Master Control View (Figure 3). The Master Control View pictures each of the seven subsystems with the three members of the middle level, the uplink module (Figure 4), the downlink module (Figure 5), and the antenna control module (Figure 6). The lowest level of hierarchy contains the individual components presented in the middle level. The components are the Receiver (Figure 7), the Exciter, and the High Power Amplifier.

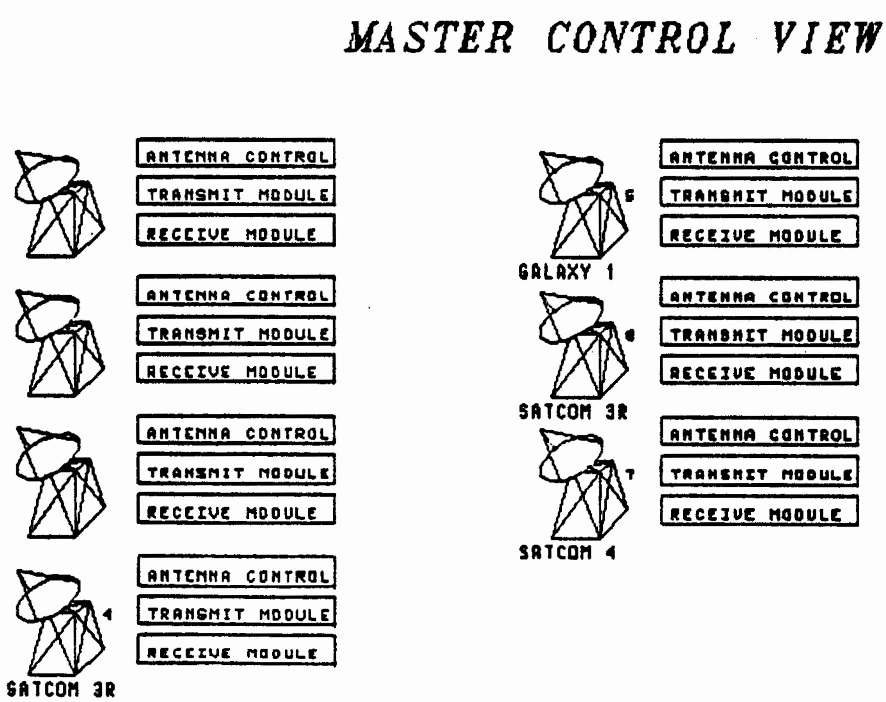


Figure 3. Master Control View

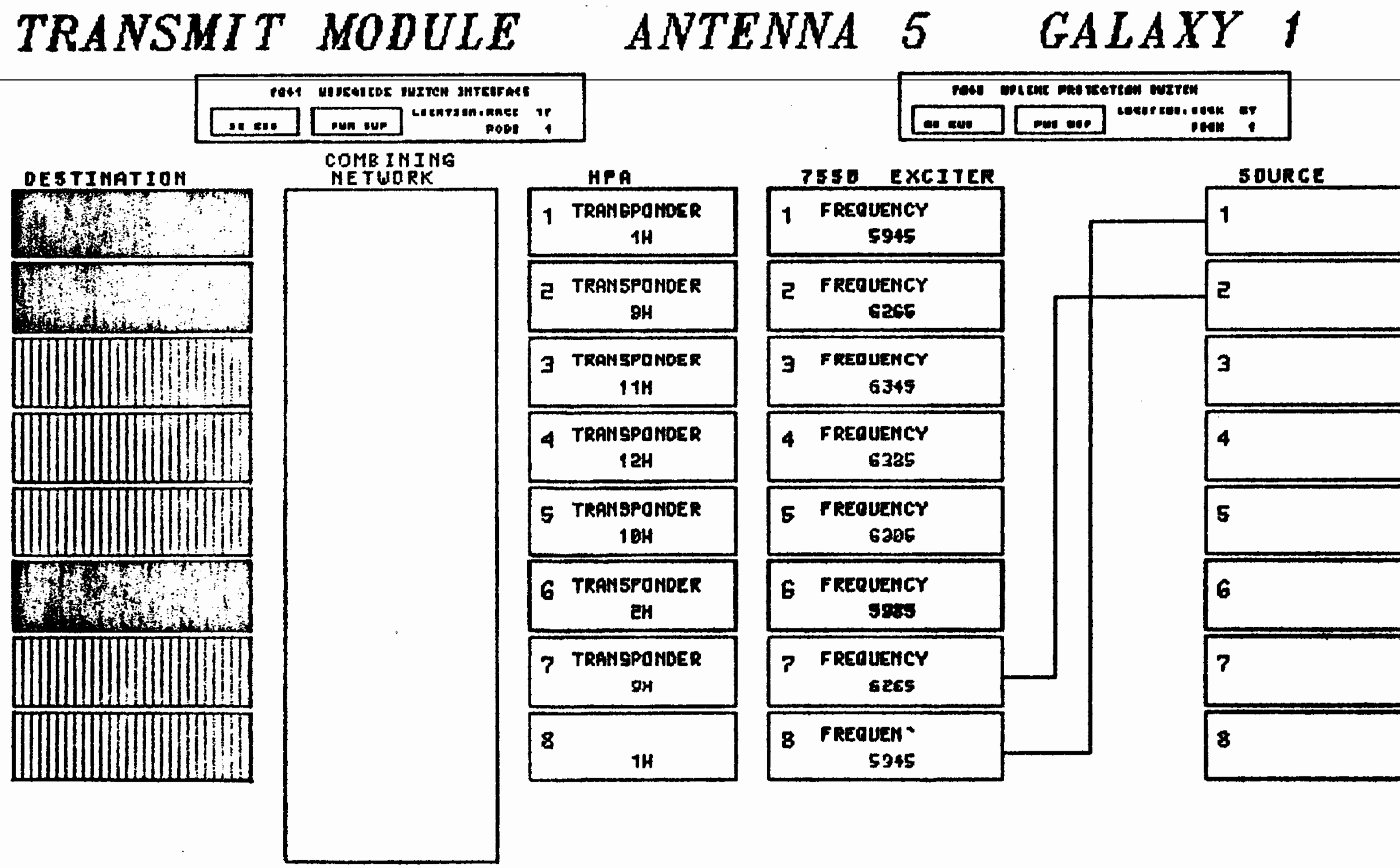
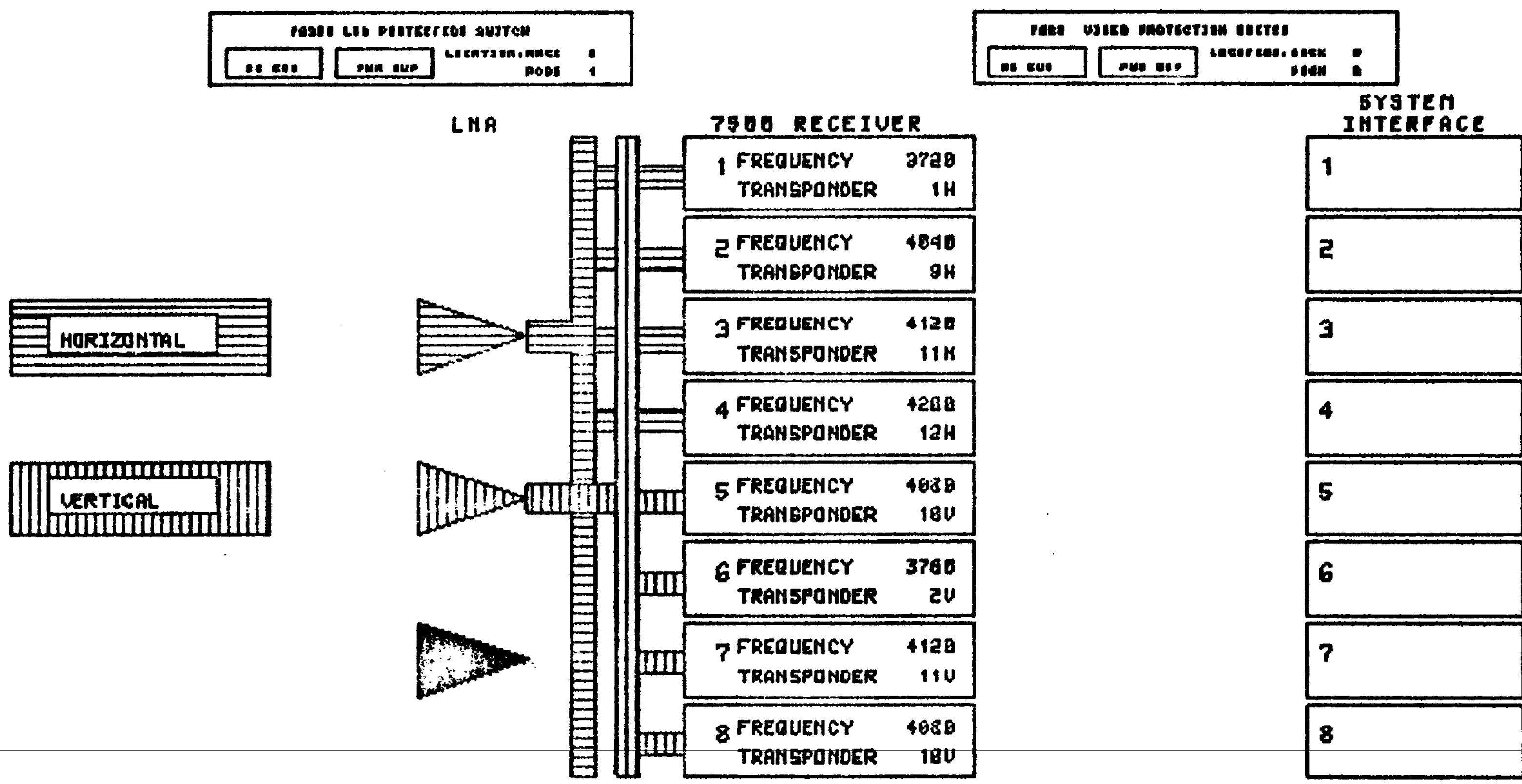


Figure 4. Uplink Module

RECEIVE MODULE ANTENNA 5 GALAXY 1



ANTENNA CONTROL   ANTENNA 5   SATCOM 4

SA BUS

IN MOTION

PAN OVR

PED OVR

MOT LIM

WESTAR	GALAXY 1	SATCOM 3R	SATCOM 4	SATCOM 4
INTELSAT 5	INTELSAT 4	SATCOM 4	SATCOM 3R	SATCOM 4
SATCOM 4	SATCOM 4	SATCOM 4	GALAXY 1	GALAXY 1
WESTAR	WESTAR	GALAXY 1	SATCOM 5	SATCOM 4

AZ 342.33

EL 26.87

PDL 52.78

LOCATION: RACK 34

POSN 8

Figure 6. Antenna Control Module

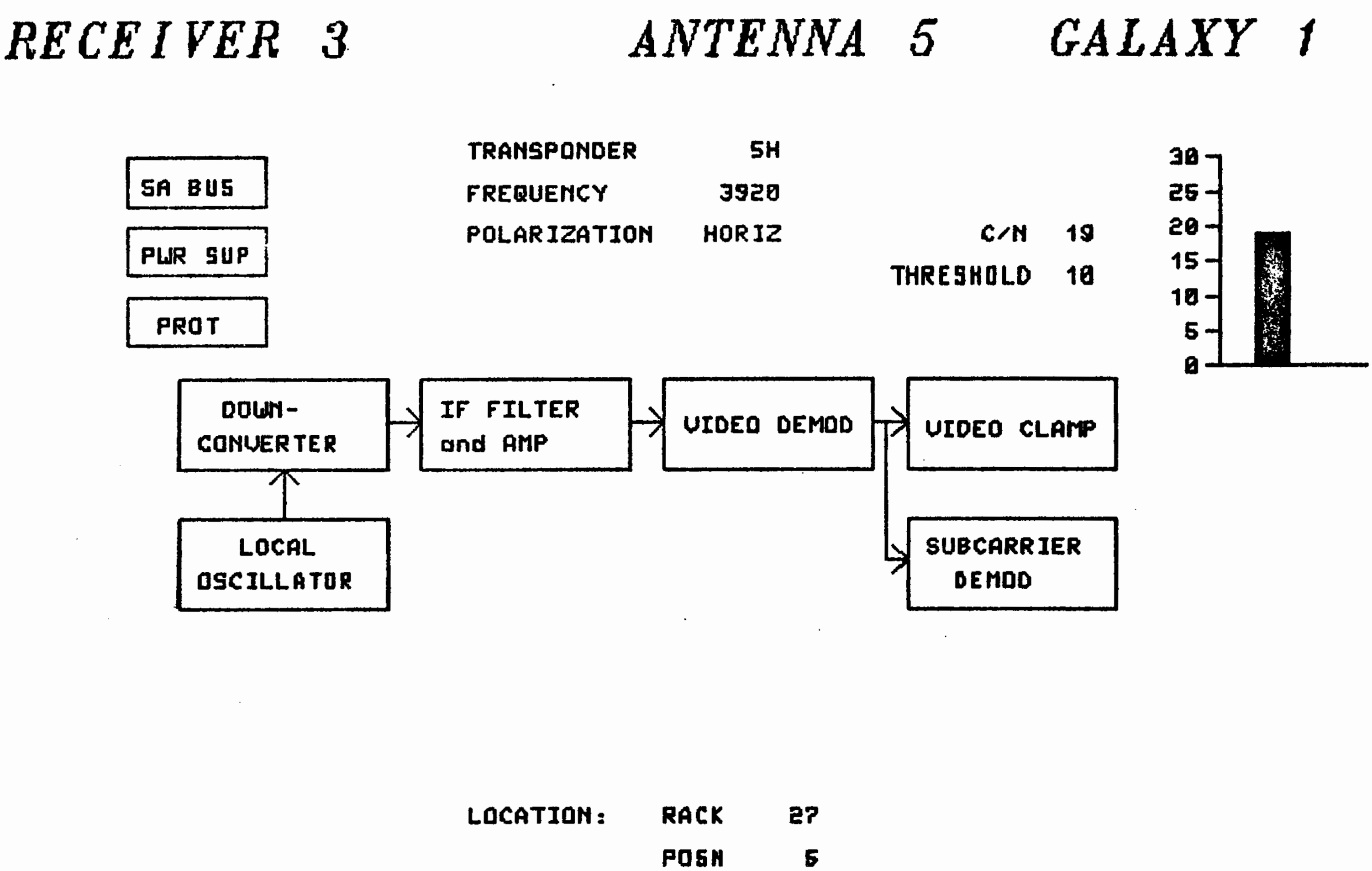


Figure 7 Receiver

Each view (except the Master Control View) has an overview in the upper left corner of the screen. The overview is a small representation of the Master Control View. There are three horizontal lines associated with a subsystem number, with each of the horizontal lines representing the antenna control module, uplink module and downlink module, respectively. The overview is used to alert the operator when alarms occur in the earth station if the operator is looking at a view other than the Master Control View.

Earth station information is presented as simple block diagrams with color, pattern and highlighting employed to alert the operator to changes in earth station status. Color represents normal and alarm states. Patterns represent, in the form of hash marks, the polarization of some components. Highlighting, the process of making items on the screen blink, is used to notify the operator of alarms. All views are updated continuously so as to present the most current status of the earth station to the operator.

Summary

As demonstrated in these two examples of earth station graphics, the inclusion of computer graphics in a remote control earth station environment enhances the operator's ability to recognize changes in earth station status. The earth station operator is able to preceive alarms quickly as communicated to him by the graphics display. Through the use of a computer graphics system at the earth station site, the earth station operator has complete, concise and well displayed information before him, contributing to a dependable and reliable remote control earth station.

Event Scheduling

G. Hartwick

Introduction

Controlling large, sophisticated satellite earth stations or large networks of these earth stations often requires the operator to submit a considerable number of commands at precise times. Examples of commands that have to be issued are ones that move antennas, tune receivers or transmitters, set polarization, and set waveguide switches. It would be helpful if sequences of these commands could be stored at the earth station for automatic execution at designated times. Such a stored schedule could, by allowing advanced entry and editing of commands, reduce operator error and could, by reducing the number of commands to be entered at one time, improve operator efficiency.

This paper describes an implementation of event scheduling recently completed at Scientific-Atlanta. The implementation consists of two major tasks: the editing and storing of commands in the schedule and the processing of stored commands from the schedule. The first task is called EDIT EVENT SCHEDULE and the second task is called PROCESS EVENT SCHEDULE. Both of these tasks rely heavily on a command language parser that is described in the paper "Command-Language or Menu-Driven Control."

Event Schedule Features

- Commands can be stored in the event schedule after causing a call to the Edit Event Schedule routine by issuing the command EDIT EVENT SCHEDULE.

- The editor has its own command set.

- The advantage of immediate syntax checking of commands is realized by using a command language parser (paper "Command-Language or Menu-Driven Control") to obtain both edit and control commands. Thus, only valid control commands are entered into the event schedule.

- The user is prompted when help is requested.

- Commands are logically stored in the event schedule in their fully expanded (nonabbreviated) form.

- Control commands are scheduled by the day, hour, minute, and second using a 24-hour clock.

- The commands that may be entered into the event schedule are a subset of the control command set. A diagnostic is displayed if the user attempts to enter a command that is not a member of this subset. For example, the EDIT EVENT SCHEDULE command would not be allowed in the event schedule.

- Encoded commands are submitted for subsequent processing at the appropriate time by the Process Event Schedule task. This task is always "asleep" unless it is time for a command in the event schedule to be processed. At that time, it is "awakened;" the command is extracted from the event schedule; the command is passed one character at a time for encoding and processing; the command is purged from the event schedule; and, the task puts itself back to "sleep."

Event Schedule Editor Features

The event schedule editor is a "current line" editor; that is, its commands affect or are relative to the last line seen by the user. The features of this editor are:

- The editor is called from another program or task.

- The editor has five commands. They are LINE, INSERT, DELETE, LIST and EXIT. The LINE command is used to move the "current line" forwards or backwards in the event schedule file. INSERT is used to add a new command or a comment to the file. DELETE removes the "current line." The LIST command displays the entire or a specified range of the event schedule on the user's terminal. And, EXIT returns contol to the calling program.

- Only one terminal at a time is allowed to be in the EDIT EVENT SCHEDULE mode.

- The "current line" is always located at the first command in the event schedule when the EDIT EVENT SCHEDULE command is issued.

- A command may be inserted BEFORE, OVER, or AFTER the "current line." The default value is to insert before the "current line."

- The LIST command displays the command in its full, nonabbreviated form.

- The user edits a copy of the event schedule.

- After issuing the EXIT command, the user will be prompted for permission to replace the master copy of the event schedule provided an insert or delete operation had been performed.

- If, when the EXIT command is issued and permission to replace has been obtained, commands exist in the copy that have scheduled execution times which are elapsed, a warning will be issued to the user. The user will then have the choice to have the commands executed or to remove them from the schedule.

- The INSERT command may be implied by typing a valid control command with the appropriate time. The time field could cause the definition of "current line" to change.

- Some commands may imply other commands. For example, tuning a receiver to a transponder on a specific satellite may imply a move antenna command.

Edit Event Schedule Commands

The command syntax described below illustrates the user inputs required to effect changes in the event schedule. Parameters enclosed in angle brackets, <>, indicate operator inputs. Parameters enclosed in square brackets, [], indicate options. And, a "|" indicates that either parameter may be used.

When the EDIT EVENT SCHEDULE command is issued, control of the user's terminal is passed to the Edit Event Schedule routine. At that time, only the following commands will be accepted for processing from that terminal:

LINE [<n>] | EOF]

This command causes the "current line" to be defined as the next line--or command, as only one command per line is allowed--in the event schedule. The user may move the "current line" <n> lines, where <n> is a valid, signed integer. EOF causes the "current line" to be defined at the end-of-file and not at the last line. If <n> is greater than or less than the number of lines remaining in the file, the "current line" will be, correspondingly, positioned at the last or first line in the file.

INSERT [BEFORE | OVER | AFTER]

This command causes a new item to be entered into the event schedule. If BEFORE is specified, the new item is entered immediately prior to the "current line." If OVER is specified, the "current line" is replaced by the new item. And, if AFTER is specified, the new item is entered immediately after the "current line." The default is BEFORE. After issuing the insert command, the user is prompted for the new item which is to be terminated by a carriage return.

DELETE

The "current line" is removed from the file when this command is issued.

LIST [<n> [:<m>]]

Causes the next line in the event schedule to be displayed. Parameters <n> and <m> are valid, signed integers which cause either the nth or the nth through the mth lines(s) to be displayed. In all cases of this command, the "current line," is redefined as the last line displayed.

EXIT

Returns control to the calling program. If an INSERT or DELETE command has been processed, asks the user for permission to replace the event schedule. If commands with elapsed times exist in the new copy of the event schedule, ask the user to resolve the situation.

Inputs. Inputs to this process consist of edit commands, the event schedule, and commands and comments to be entered into the event schedule.

The syntax of a command to be entered into the event schedule is:

YYMMDD HHMMSS <command> [;<comment>]

where, <comment> is a string of characters not exceeding the current line length, YYMMDD is a valid future date, HH ranges from 00 to 23, MM ranges from 00 to 59, and SS ranges from 00 to 59.

Outputs. The editor outputs a file of ASCII characters which correspond to the commands that would be entered at the console keyboard. This file is available for execution after leaving the editor.

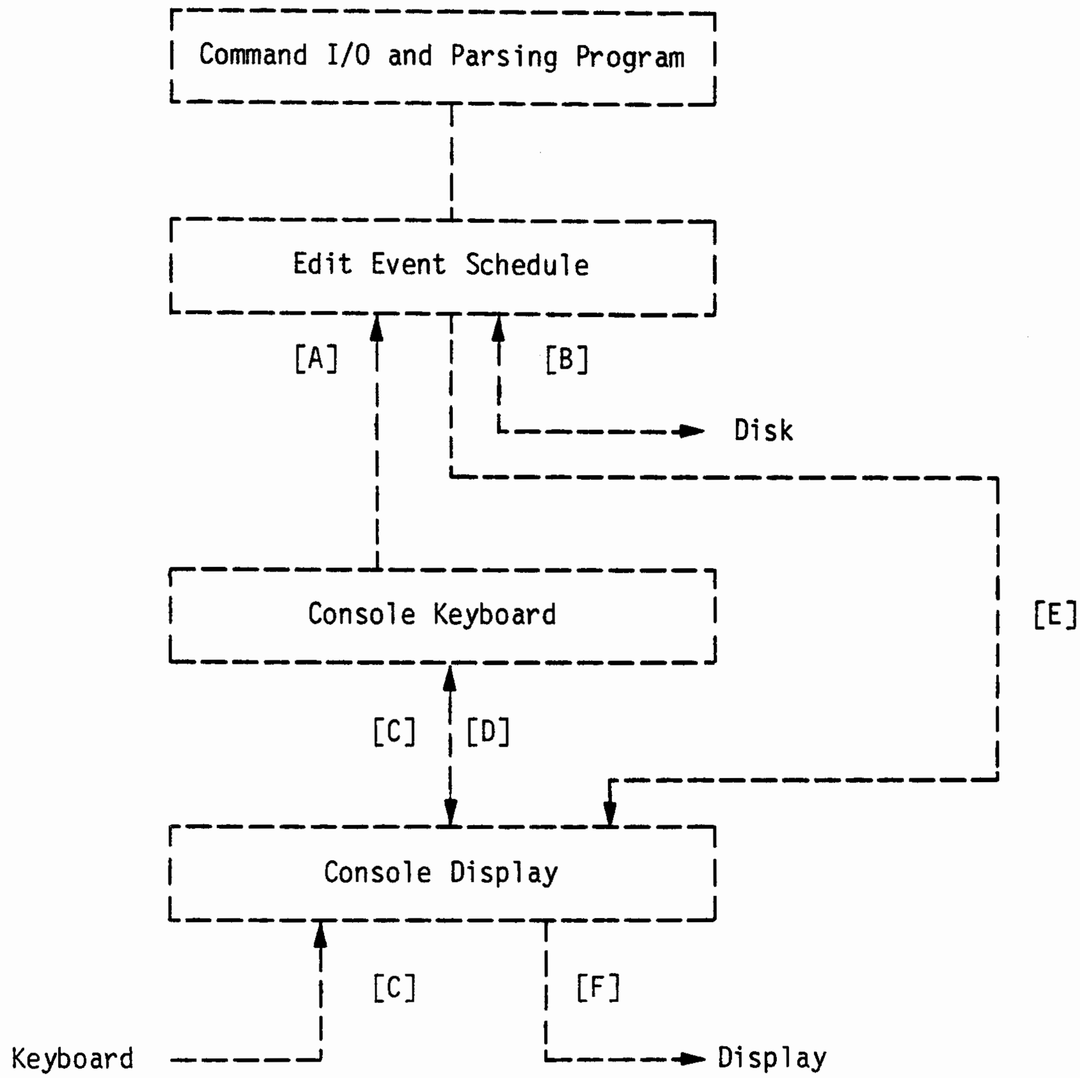
Structure of Tasks

The structure is described by the interface diagrams as shown in Figures 1 and 2. Each box represents a compilation unit (main program or routine). A line between two boxes represents a routine call--the compilation unit in the higher box calls the routine in the box beneath it. An arrow represents data flow: a letter in brackets (e.g. "[A]") identifies the type of data; legends at the bottoms of the diagrams provide definitions.

Edit Event Schedule Structure

The program or task charged with command input keeps track of the current activity being performed at the console. If the activity is to edit the event schedule (triggered by the EDIT EVENT SCHEDULE command), it invokes the Edit Event Schedule routine (Figure 1).

The Edit Event Schedule routine edits and stores commands in the event schedule. It gets successive commands from the Console Keyboard routine.



Legend

[A] encoded keyboard command

[B] valid keyboard command or comment in text form and, if command, the encoded command

[C] keystroke

[D] echoed keystroke, help report, or response to "delete" request

[E] displays

[F] echoed keystroke, cursor control, or status report

Figure 1. Edit Event Schedule Structure

The Console Keyboard routine collects keystrokes from Console Display and parses the commands. At the completion of the command, it returns an encoder command to the Edit Event Schedule routine.

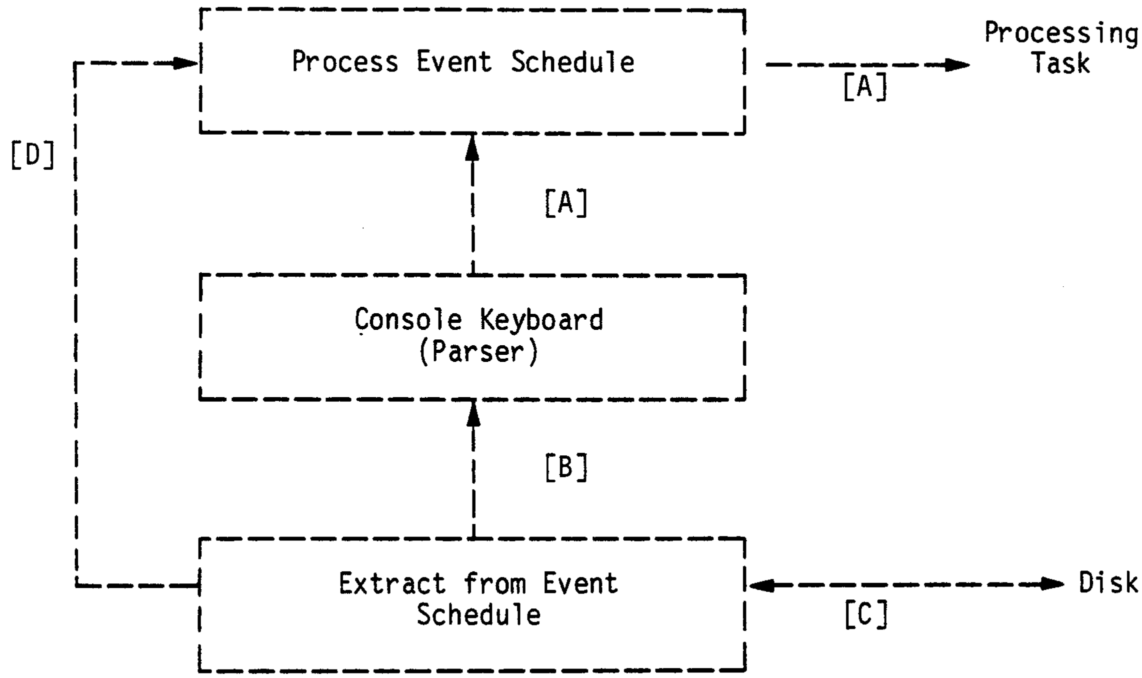
The Console Display routine is the interface to the keyboard and the display CRT or printer. It relays information to the Console Keyboard routine one character at a time.

Process Event Schedule Structure

The Process Event Schedule routine (Figure 2) gets successive commands from the Event Schedule and relays them to the processing task. It also performs initialization, which consists of reading the parsing tables. Process Event Schedule remains dormant until the time-of-day arrives for the next command in the event schedule to be processed. It is also awakened when Edit Event Schedule ends because of the possibility that there is a new next command as a result of the recent editing.

The Console Keyboard routine collects stored keystrokes from the Extract from Event Schedule routine and parses the commands. At the completion of the command, it returns an encoded command to the Process Event Schedule task.

The Extract from Event Schedule routine gets encoded event-schedule commands from the disk in text form. It relays this information to the Console keyboard routine one character at a time. It ignores requests from the Console Keyboard routine to echo the characters.



Legend

[A] encoded keyboard command

[B] characters extracted from the event schedule

[C] commands extracted from the event schedule, deletion marking

[D] date and time of next command in the event schedule

Figure 2. Process Event Schedule Structure

An Introduction to Broadband Multi-Access Techniques

C.M. Ermer

Introduction

The purpose of this paper is to give an overview of the various techniques that can be used to allow multiple users to share the broadband cable. The emphasis of the paper will be placed on techniques used in packet communication systems. While the current interest in Local Area Networks has generated a great deal of interest in packet networks and thus a proliferation of techniques, this paper will restrict itself to the more commonly used protocols (for a review of the many other available multi-access protocols, see the Tobagi paper[1]). In particular, the discussion will focus on how the protocols work, what are the key parameters that affect performance, how do each of the protocols perform, and what general applications are best suited for each protocol.

Broadband Cable

The broadband cable medium is discussed in more detail in other papers; in particular, see the paper "Two-Way Broadband Communications and an Introduction to Broadband System Design." For the purposes of this paper, it only needs to be recognized that the broadband system is of a tree structure, is basically a broadcast medium, has both an upstream (to the root) path and a downstream (toward the branches) path, and is a closed system. Thus. in order to communicate between any two arbitrary points n the system will require an upstream signal from the sending point and a downstream broadcast of the signal from the root (headend) to the receiving point. Therefore, the available bandwidth would be about 100 MHz in a Mid-Split system and about 170 MHz in a High-Split system.

The resource of the broadband cable can be viewed as having two dimensions--frequency and time. Modulation of an RF carrier allows for information to be constrained to a specific amount of bandwidth (frequency) and time. Various modulation techniques allow for more or less efficient use of frequency and time (see the paper "Digital Modulation: Characteristics and Performance of Representative Modulation Techniques"). Obviously, a modulation technique that conserves bandwidth, allowing either a smaller amount of frequency to be used for the same data rate or a higher data rate (a smaller amount of time) for the same frequency usage, would increase the cost of the hardware. Since the broadband cable is a finite resource, it must be efficiently utilized: this requires both a choice of modulation technique and frequency and time assignment to the various users.

The simplest multi-access technique is to frequency-divide the spectrum (FDM). In this approach each communication link is assigned a piece of the spectrum (see Figure 1). Thus, each communication link is assigned a unique piece of bandwidth, and the various communications terminals are free to communicate at any time without interfering with any other terminals. When using FDM, frequency division multiplexing, conservation of the frequency resource requires that for any desired communication rate, bits per sec (bps), an emphasis must be placed on selection of a bandwidth efficient modulation technique. That is, a high bits per hertz (bpHz) will conserve bandwidth and allow more circuits.

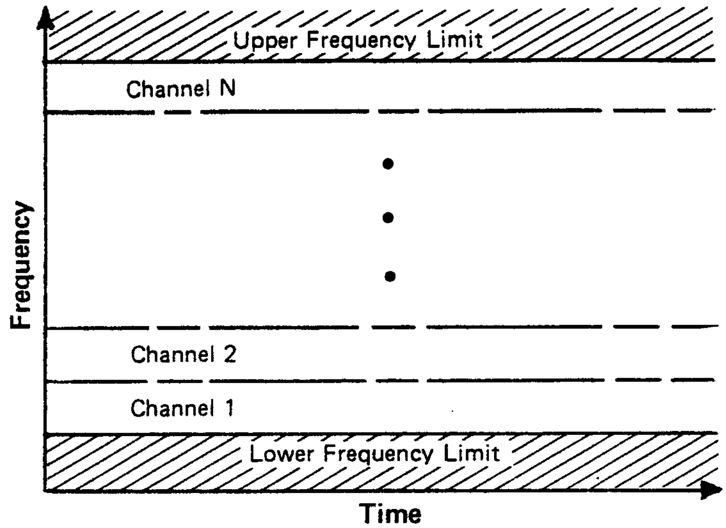


Figure 1. FDM Resource Allocation

The other common fixed assignment technique is Time Division Multiplexing (TDM). In this technique, each communication link is assigned a specific time slot (see Figure 2). Here each unit is assigned a sending and receiving time slot. For example, if A is to communicate with B, A could be assigned to transmit in Time Slot 1 and receive in Time Slot 2. Obviously, B would then transmit in Time Slot 2 and receive in Time Slot 1. For the TDM technique, conservation of resources requires that a modulation technique be used that has a high data rate (bps) in order to allow for maximum users. Since each end user only transmits for short bursts, the other constraint on the modulation scheme is that acquisition times (receiver/transmitter synchronization) must be short.

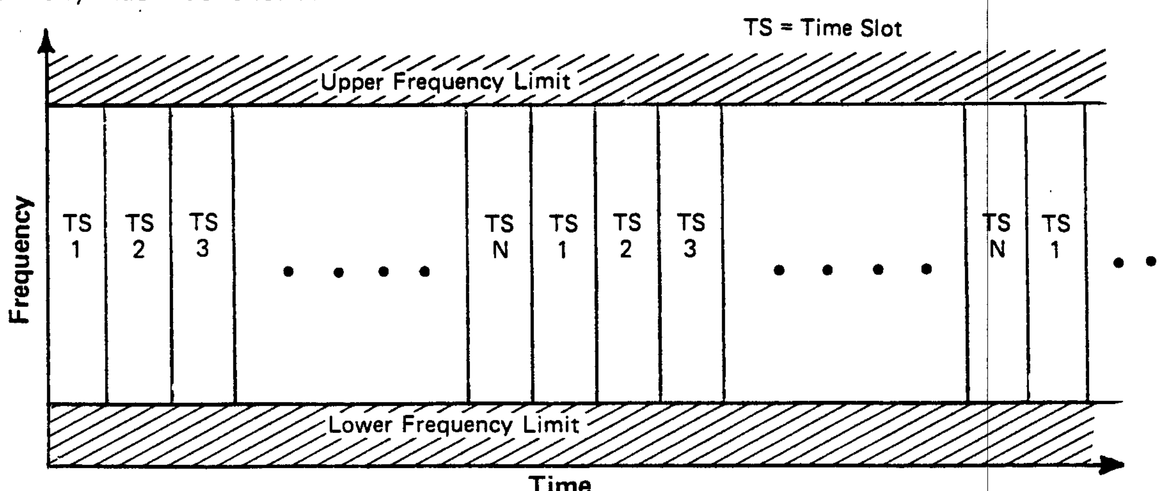


Figure 2. TDM Resource Allocation

In general, high-speed data links require the transmission of large amounts of information and are generally best served with bandwidth efficient FDM modems. On the other hand, low-speed data links generally do not require large data transfers and tend to be bursty in nature. Cost is important and bandwidth efficient modulation cannot be justified. When FDM is used with low-speed modems, a large number of carriers are required; this can be an administrative problem particularly if crystal-controlled modems are used. Thus, for low-speed data circuits, it is often convenient to use TDM techniques.

Finally, any real communication system will have a combination of both high- and low-speed circuits. In this case, the broadband cable can be divided using a combination of both FDM and TDM (see Figure 3).

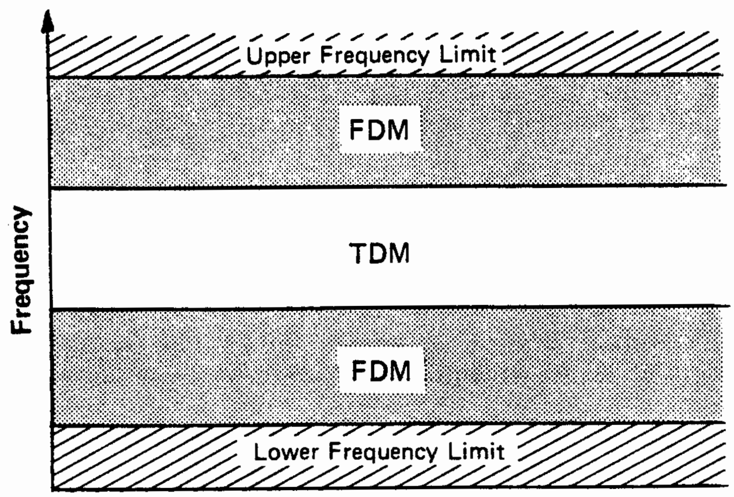


Figure 3. Hybrid Resource Allocation

Up to this point the access techniques are fixed assignment; that is, the frequency or time slot are assigned to a single user. In many instances users may not have sufficient data/information to fill the total channel capacity. This means that the broadband resource will not be fully utilized. In particular, computer traffic tends to be bursty; that is, infrequent and random with message size small relative to the channel capacity. For computer networks, FDM and TDM would be a very inefficient use of the broadband cable.

The desire to have efficient methods of handling the communication needs of bursty users has led to the development of packet-switched communications systems. The basic idea behind a packet communication system is to assign the total resource of the communication system to a single user for a short time. Thus, if proper control algorithms can be defined, many users can share the channel (multi-access) and the channels resources can be efficiently utilized. These control algorithms, usually referred to as multi-access protocols are the subject of the remainder of this paper.

Multi-Access Protocols

Multi-access protocols fall into two general categories--demand assignment and random access. Demand assignment protocols control the resource by explicitly determining the requirements of the various users and deterministically assigning the resource to avoid conflicts. These techniques give very predictable results and are well controlled under all conditions. Common demand assignment techniques are polling, token passing, and Broadcast-Recognizing Access Method (BRAM). Random access protocols are those that allow a user to transmit on the channel when ever the user has information to communicate. That is, there is no central contoller that decides the allocation of the channels resource. Each individual terminal is free to access the channel at any time. Obviously, collisions will occur; each protocol has its own technique for handling collision and taking action to complete the information transmission. Two of the common random access techniques are Aloha and Carrier Sense Multiple Access with Collision Detection (CSMA/CD).

Before describing the above multi-access protocols and discussing their performance characteristics, it would be best to define a few standard terms.

- Channel Throughput (S): The rate of message transmittal; that is, the actual number of messages that are successfully transmitted per unit of time.

- Channel Capacity (C): The maximum value of S that can be achieved for a specific access protocol.

- Expected Dealy (D): The average time between message generation and successful transmission. Successful transmissions means the reception of the message at the destination terminal.

- Propagation Delay (d): The time it takes to send a message between the most remote pair of terminals on the system; i.e., transmission time.

- Offered Traffic (G): The total amount of messages requiring transmittal both new and repeated.

- Overhead: The information (bits) in addition to the actual data required to make the network function. This would include: headers, deliminators, addresses, control bits, etc.

Before beginning an evaluation of the various protocols, it would be best to get a feel for the limits of system performance. The theoretical performance of any protocol is bounded, and the bound can be determined from Queuing Theory. The best any multi-access protocol can perform is if the system had perfect knowledge. That is, each terminal would need to be ordered at the instant it had a message to send and each terminal would send its message as soon as it could without causing any collisions. This system of Perfect Scheduling can be modeled by using the Single-Channel, Single-Phase Queuing Theory Model. Because all peripheral multi-access protocols require overhead and real-time delays, it is impossible to achieve perfect scheduling; however, the M/D/1* model does provide the optimal performance boundary.

*In the M/D/1 queuing theory model M defines an exponential distribution function for customer-interarrival time; D defines a deterministic (fixed packet size) customer-service time distribution; and 1 defines a single server model.

Demand Assignment Protocols

Polling

Polling is one of the most straight-forward forms of a multi-access protocol. It is simple, stable, predictable, and easy to implement. In a polling scheme, a central controller has ultimate authority for the network. The central controller sequentially polls each remote terminal to inquire if the terminal has any messages to send. The terminal responds with a simple no (short message) or transmits its message (a maximum message length may be imposed). When the transmitting terminal completes its message, the central controller polls the next terminal. The process continues in a cyclical fashion. The strength of this technique is its obvious stability. The process is centrally controlled with relatively unintelligent remote terminals. Because of the deterministic cyclical nature, a maximum delay time can be guaranteed. The weaknesses of the system are the expected delay and the channel throughput. The expected delay is relatively long when compared to random access techniques for cases when the system loading is small. This is true since each terminal must be polled on every cycle even if very few terminals have messages to transmit. Also, the high overhead (the polling messages) and trasmission delays seriously impact the throughput of the channel.

Token Passing

Token passing is closely related to polling. In token passing, the control function is decentralized; that is, each terminal is given the intelligence to determine when it has control of the network. The system works by establishing a logical ring on the broadcast medium; that is, each terminal is assigned a sequence number. The process begins by the first station transmitting its message, at the end of its message, it transmits the token (a special code) to the next terminal. Now that terminal two holds the token, it controls the network. Terminal two can either send its message or pass the token to terminal three. As with polling, this cycle continues constantly. The most obvious advantage of this system is the improved performance relative to polling. By passing the token, the delay associated with the polling sequence is reduced. In addition, reliability can be improved since the system is not totally dependent on a central controller. Like polling, token passing has the advantage of being stable and can be designed to give a guaranteed maximum delay time. As with polling, token passing has a relatively high expected delay, for low levels of system usage, since the token is constantly passed around the ring. Token passing is more complicated to implement than polling. Conditions that increase the complexity are: how to add a new terminal to the network, how to remove a terminal from the network without stalling the token, and how to recover if one or more bogus tokens are created.

Broadcast-Recognizing Access Method (BRAM)

Broadcast-Recognizing Access Method (BRAM)[2] is a demand assignment protocol that uses a carrier sense instead of a token to give a terminal control of the network. BRAM requires terminals to start a transmission at times that are integral multiples of the channel propogation delay d. Collisions are avoided by a scheduling algorithm; S(i,j), where i is the terminal ready to transmit and j is the terminal that transmitted last. The scheduling algorithm is a linear algebraic relationship of i, j, and k (the number of nodes on the network; S(i, j)) is designed to give a unique value to each terminal (i) for any j. The algorithm can either give priority to the most recent terminal to transmit or lowest priority to the terminal that transmitted last. In either case, the algorithm is designed to give a rotating priority to all terminals, thus on average giving equal access to the network to all terminals. Collisions are avoided in BRAM by requiring each ready terminal to compute S(i, j) (note that a transmitting terminal must identify itself during all transmissions), and wait d•S(i, j) time units after a transmission stops before trying to access the network. Since S(i, j) is unique to each terminal, each terminal will have a unique access time. When a terminal's access time occurs, the terminal must sense the cable for a carrier, if the cable is idle the terminal is free to transmit. If the cable is not idle, the terminal defers. Everytime a new terminal gains access all terminals compute a new S(i, j).

Random Access Protocols

Aloha

ALOHA* is a random access technique that allows a terminal to transmit anytime it has a message to send. After transmitting, a terminal waits for a time-out period to receive an acknowledgement that the message was received.

If no acknowledgement is received, the terminal assumes that a collision occurred and retransmits the message. Since no attempt is made to determine if the channel is busy, the likelyhood of a collision is very high for even moderate loading; therefore, ALOHA has a very low capacity.

Carrier Sense Multiple Access (CSMA)

Carrier Sense Multiple Access (CSMA) is a logical extension of ALOHA. In this technique, the terminal must first determine if the channel is busy (carrier presence). If no carrier is present, the terminal is free to transmit. If a carrier is present, the terminal waits a time-out period. In order to assure that any terminals that sense the channel being busy at the same time, will try again at different times, the time out is generated from a probability distribution function. Terminals continue to sense the channel and wait the time-out until the channel is free. Due to the channel's propogation delay, carrier sensing is not adequate to prevent collisions. Thus, acknowledgements are required to guarantee that messages are received. Performance can be improved by adding collision detection (CSMA/CD). In this case, the terminal must listen while it transmits; if a collison occurs (incoherent data), the colliding terminals stop transmitting and reset their random time-outs before trying again. Collision detection thus decreases the amount of time a collision can occur and frees up the channel sooner improving overall performance. The advantage of CSMA/CD is its basic simplicity and disadvantage is the lack of pure control.

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*ALOHA was developed at the University of Hawaii in 1970 for a packet switching radio network.

Performance Comparisons

The protocols discussed thus far are representative of the demand assignment and the random access types. By using various gimicks and adaptive techniques that overcome the inherent weaknesses of each protocol, some performance improvement can be made to each. However, none of the protocols can be forced into any dramatic improvement. Of course, any improvement comes at the expense of implementation. For an in-depth review of each protocol and its various improved versions, see Franta and Chlamtac[3].

The remainder of this paper will discuss the performance characteristics of the various protocols. The objective will be to show the relative differences. Precise performance of each protocol depends on a multitude of variables such as: number of terminals on the network, the size of the packets, input rate, and the propogation delay. These variables are characteristic of any particular network application and are fixed for all protocols. How each protocol will perform can then be modeled either analytically or by simulation[3].

The key measures of performance are throughput and delay as a function of offered traffic. Figure 4 shows the relationship between throughput and offered traffic. For low levels of traffic, CSMA/CD performs very close to the perfect (M/D/1) model. Polling and token schemes, because of their higher overhead (control information), will have a slightly inferior throughput performance, but for all the relationship is essentially linear.

The major difference in the protocols shows up as traffic increases. The M/D/1 model continues its linear relationship until the channel capacity is reached, at this point no more traffic can be handled. The demand assignment protocols, polling and token passing, will asymptotically approach a throughput level less than the channel capacity. The difference between this level and the channel capacity is due to a function of the overhead required to control the network. To understand why CSMA/CD crashes, it must be remembered that offered traffic consists of new traffic plus repeat traffic. When traffic is low relative to channel capacity, collisions are few and the ratio of repeat to new traffic is small. As traffic grows, collisions increase and the ratio of repeat to new traffic increases. Eventually the repeat traffic gets so large that the throughput actually decreases. Thus, all protocols work well with respect to throughput at low levels of traffic, however, the demand assignment techniques are more stable.

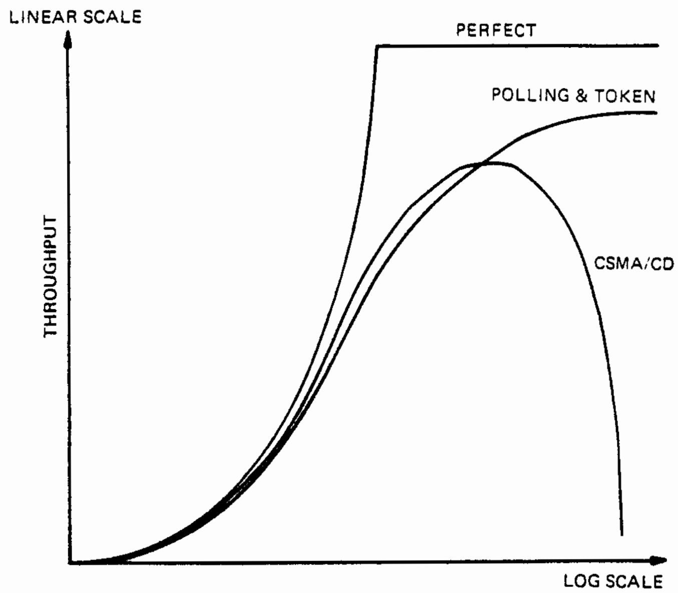


Figure 4. Throughput vs. Offered Traffic

While all perform well with respect to throughput at low levels of traffic and demand assignment, protocols are clearly superior at high traffic levels. It would be incorrect to claim that demand assignment is a better approach. The next performance measure, expected delay, must be considered. Figure 5 shows the relationship between delay and throughput. As can be seen, demand assignment techniques have a greater delay at low throughput levels. However, as offered, traffic increases and thus throughput expected delay increases for all protocols. As was seen in Figure 4, CSMA/CD will crash sooner, but all protocols develop excessive delay as channel capacity is reached.

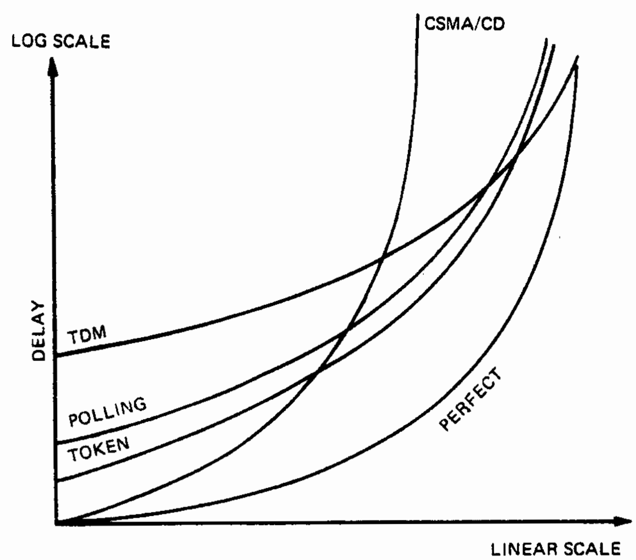


Figure 5. Delay vs. Throughput

Conclusion

This paper has presented the basic principles of broadband multi-access protocols. Three general types were reviewed: fixed assignment, demand assignment, and random access. Selection of a particular protocol is dependent on the application. As a general rule, fixed assignment techniques are best suited to applications that require an almost constant transfer of information. Both demand assignment and random access are both best for applications with bursty information, transferal requirements. Random access techniques will generally be preferred when offered traffic is small relative to channel capacity, and propogation delay is small. In these cases, expected delay will be small. Random access is the most common technique applied to Local Area Networks (LANs), since LANs exhibit the above characteristics; Ethernet is a prime example of CSMA being applied to an LAN. Demand assignment techniques are superior if the offered traffic is high, relative to channel capacity, and/or propogation delay is high. For Metropolitan Area Networks (MANs), propogation delay is high and, since the broadband cable is a valuable resource, achieving a high throughput is desirable; thus, demand assignment techniques will be the preferred protocol.

References

[1]Fouad A Tobagi, "Multi-Access Protocols in Packet Communications Systems," IEEE Transactions on Communications, Vol. COM-28, No. 4, April 1980

[2]I. Chlamtac et al, "BRAM: The Broadcast Recognizing Access Method," IEEE Transactions on Communications, Vol. COM-27 pp. 1183-1190, August 1979

[3]Franta WR and Chlamtac I, Local Networks, Motivation, Technology, Performance, Lexington Books, D.C. Heath and Company, Lexington, Massachusetts, 1981

Video Broadcast Receiver, Series 7500

T.C. Mock

Introduction

Designed specifically for the broadcast and common carrier industry, the Scientific-Atlanta Series 7500 Video Receiver offers unmatched flexibility and performance at a low cost. Reception of satellite TV signals meeting RS-250B and NTC-7 performance requirements is made possible with the 7500.

For flexibility in control and interface with protection switches and remote control systems, the 7500 incorporates a microprocessor for monitor and control. It utilizes the SAbus control bus for remote interface.

General Description

The 7500 Receiver is all modular in design; all circuit cards (IF amplifier, video demodulator, video clamp, and subcarrier demodulators) plug in from the top of the unit. In addition, the downconverter is removable as a unit from the left side of the recevier. Three auxiliary slots are provided for additional subcarrier demodulators. Monitor and control of the 7500 is provided by a front-panel keyboard and display. Frequency is displayed with a 6-digit 7-segment display. Alarms are indicated by large back-lit LED indicators.

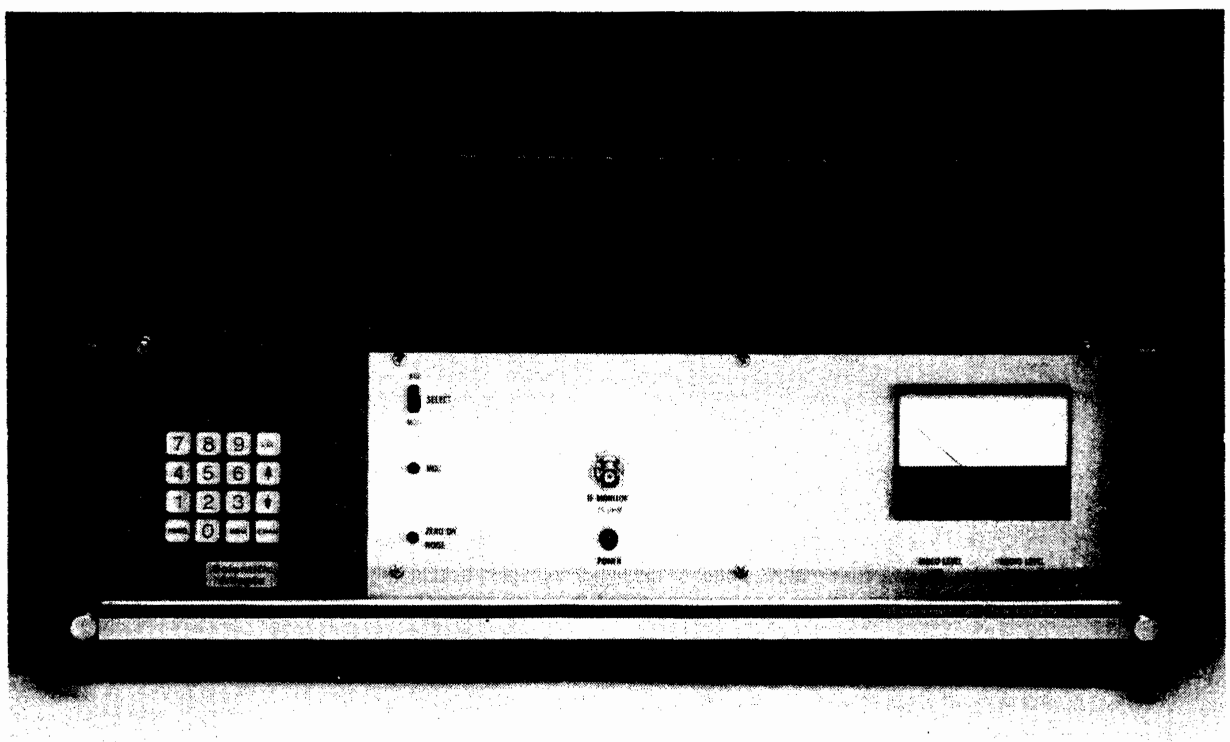


Figure 1. Series 7500 Video Broadcast Receiver

The 7500 employs a microprocessor for logic control and remote interface. Incorporating this intelligence in the downconverter enables the 7500 to communicate with the SAbus standard RS-422 data bus. The bus is a byte-serial, parallel-connected bus which allows simple interconnection of receivers and protection switches, remote control systems, and the like.

Figure 2 is a block diagram of the Series 7500 Video Broadcast Receiver.

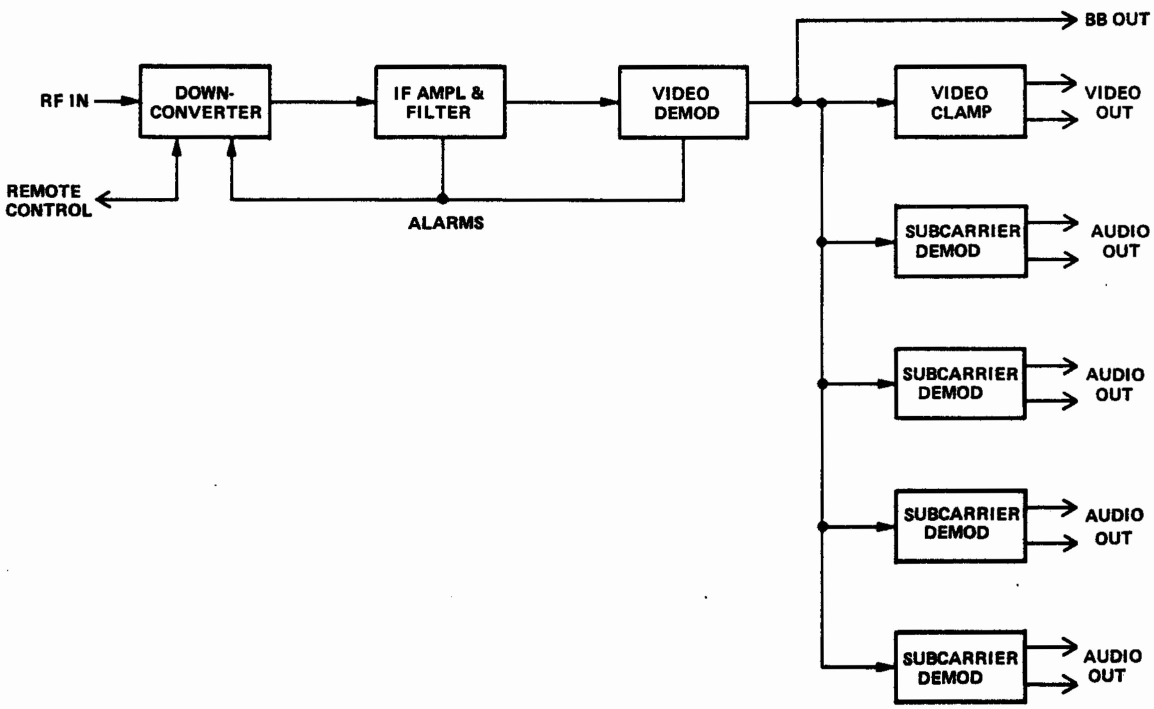


Figure 2. Series 7500 Block Diagram

Downconverter

The 7500 utilizes a dual-conversion downconverter with a frequency-synthesized local oscillator for transponder selection. A block diagram of the downconverter is shown in Figure 3.

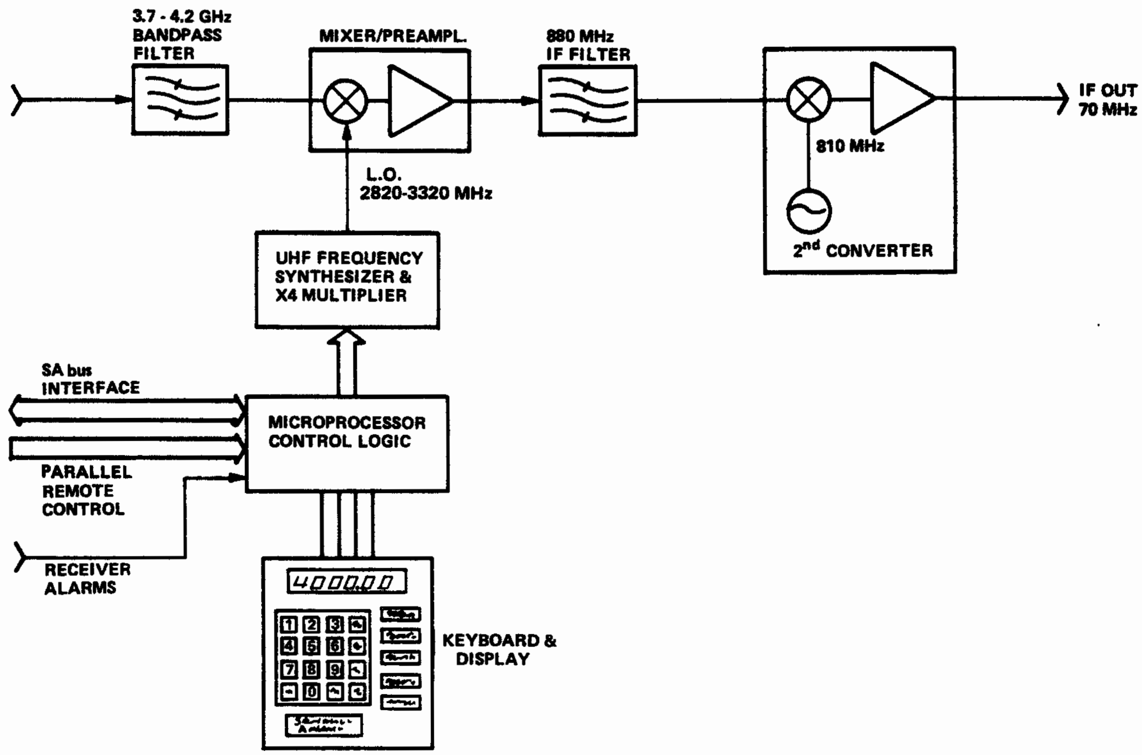


Figure 3. 7500 Downconverter Block Diagram

Signals in the 3.7- to 4.2-GHz band enter the receiver and are bandpass filtered to reject image and other unwanted signals. The desired frequency is converted to the 880-MHz first IF and amplified by the mixer-preamplifier. The local oscillator signal for the first conversion is generated by a UHF frequency synthesizer and times-4 multiplier, which provides the required 2.82- to 3.32-GHz LO signal. Frequency can be selected in 20-MHz steps for normal full-transponder video. With the optional 250-kHz synthesizer, half-transponder video and other special formats can be accommodated.

The 800-MHz IF signal passes through a bandpass filter and is then converted to the final IF frequency of 70 MHz in the second converter assembly. Total gain of the downconverter is 20 dB.

The downconverter assembly also contains the microprocessor control logic. These circuits take data from the front-panel keyboard, receiver module alarms and remote control interface, and also display frequency and status on the front panel. Remote control of the 7500 is accomplished in two ways, via the SAbus standard RS-422 control bus or by conventional parallel remote control (BCD or 1-of-6 modes). In addition to allowing remote frequency control, the SAbus also provides for remote status telemetry. The current recieve frequency, carrier-to-noise ratio and alarm status may be obtained over the SAbus.

IF Amplifier/Filter

Two functions are performed by the IF amplifier module. First, it amplifies the IF signal to the -5 dBm level required by the video demodulator, and secondly, it bandpass filters the IF signal to reject adjacent transponder signals and establish the proper predetection bandwidth for video. The IF amplifier also provides automatic gain control to maintain a constant -5 dBm output signal with input signals from -50 to -10 dBm (-70 to -30 dBm at the downconverter input).

Characteristics of the IF filter are extremely important in establishing video performance of the receiver. Accordingly, the IF filter in the 7500 is optimized for video. Particular importance is placed in maintaining good group delay characteristics past the -3 dB response points of the filter.

Video Demodulator

The function of the video demodulator is simply to extract the frequency-modulated information from the RF carrier. An important parameter in demodulators is threshold performance. The threshold of a demodulator is that point at which the output S/N begins to fall faster than the input carrier-to-noise ratio (more precisely, it is where 1-dB deviation between a straight line relationship between input C/N and output S/N occurs). Threshold performance is important where the satellite EIRP is low, because it may allow the use of smaller antennas. The 7500 employs a patented threshold extension technique which is automatically switched in when the input carrier-to-noise ratio falls below 12 dB. Deemphasis of the video baseband signal also occurs in the demodulator.

Video Clamp

After demodulation, the video baseband still contains the 30-Hz triangular energy dispersal waveform. Removal of this triangular waveform is accomplished in the video clamp. This is accomplished by sampling the video waveform during the horizontal sync pulse and applying the sampled voltage to a subtractive feedback loop. This technique results in greater than 40 dB of dispersal rejection with no distortion of the sync tips.

Also included in the clamp is a low-pass filter used to remove the audio subcarrier(s) from the video baseband.

Two versions of the clamp are available; the standard clamp provides two clamped video outputs and filtering for standard 525L NTSC video, and the optional clamp provides an additional clamped unfiltered video output and provisions for field-changeable video roofing filters. These filters are used for special applications--PBS, INTELSAT, and PAL Video.

Subcarrier Demodulator

Program audio transmitted on an FM subcarrier is demodulated in the subcarrier demodulator with conventional limiter/descriminator techniques. The 7500 Receiver contains four module positions which can be used for audio demodulators or other special functions. Two balanced 600 Ω outputs are provided for two of the positions, and one for the remaining two.

7500 Specifications and Features

RF Frequency	3.7 to 4.2 GHz
Channel Spacing	20 MHz or 250 kHz, frequency-synthesized
Input Level	-35 dBm to -70 dBm
Control	Front-panel keyboard and display
Alarms	C/N, video, LO, power supply
IF Bandwidth	30, 25, 17.5 MHz
Subcarrier Demods	Up to 4, 5.8, 6.2, 6.8, 7.4 MHz, others available
Remote Control	SAbus serial data and parallel BCD or 1 of 6

Video Performance

Differential Phase	<1°	10-90% APL
Differential Gain	<2%	10-90% APL
Line Time Distortion	<1%	
Field Time	<1%	
Dispersal Rejection	>40 dB	
2T Short Time Distortion	±1%	
Output Level	1V peak-to-peak/75 Ω Nominal	

Audio Peformance

Frequency Response	50-15000 Hz ±0.5 dB
Distortion	<1% @ 150-kHz subcarrier deviation
Output Level	Nominally 0 dBm/600 Ω at 75-kHz test tone deviation

Figures 4 and 5 illustrate video threshold performance and audio threshold performance, respectively.

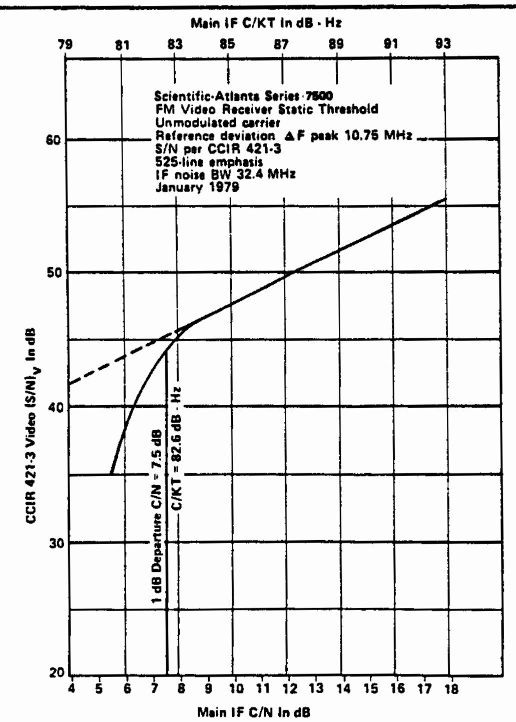


Figure 4. Video Threshold Performance

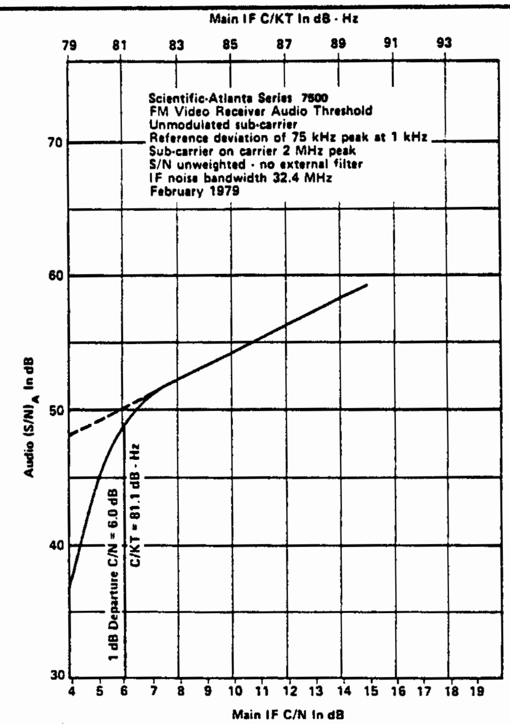


Figure 5. Audio Threshold Performance

Video Exciter, Model 7550

T.C. Mock

Introduction

The 7550 Video Exciter is a new generation exciter designed for high performance and flexibility in modern broadcast and common carrier systems. It incorporates new circuit techniques which can improve video performance over a satellite link by a significant amount. Modular in design, it can accommodate a wide variety of plug-in modules, including up to three subcarrier modulators and an auxiliary group delay equalizer module.

General Description

Housed in a standard 5-1/4-inch rack-mounted chassis, the 7550 Exciter is designed with a clean mechanical style which provides maximum accessibility to circuit modules. A removable upconverter assembly with an integral front panel occupies the left-most quarter of the exciter mainframe. Functional circuit modules (such as the wideband modulator) plug into a card frame assembly located to the right of the upconverter. Each of these modules has front-panel status indicators and controls as appropriate to its function. During normal operation, the card frame and modules are hidden behind a blank panel which hinges downward for maintenance and adjustment.

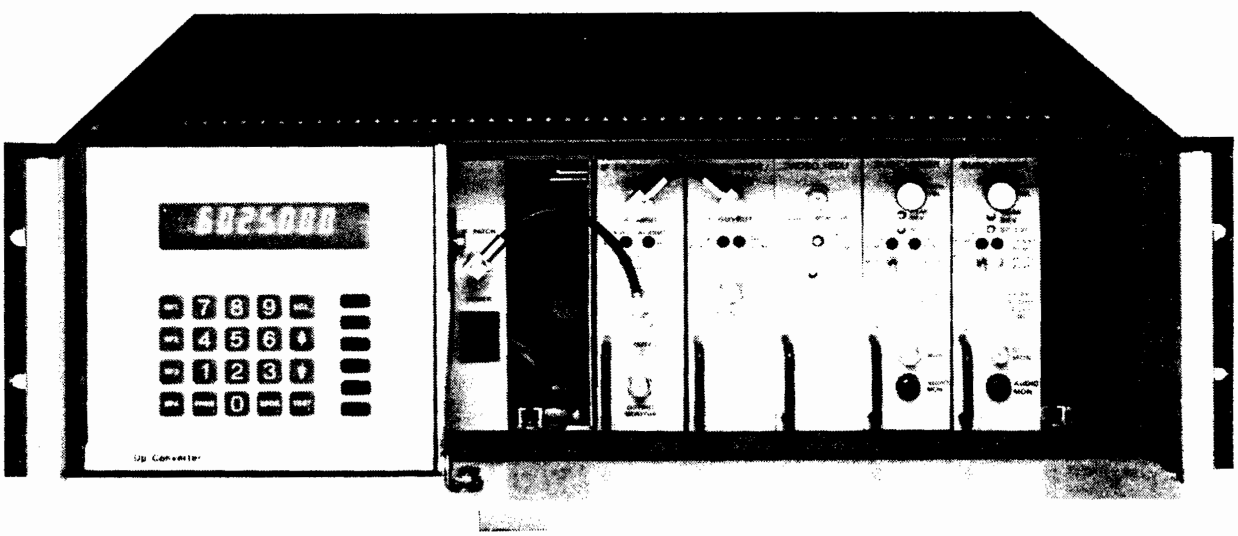


Figure 1. Model 7550 Video Exciter with Front Panel Lowered

Control of frequency and display of exciter status is provided by a front-panel keyboard and numeric 9-digit display, similar to that used in the 7500 Receiver. This capability is provided by a microprocessor in the upconverter, through which commands (both local and remote) are entered. Also, commands and status information from and to the exciter are communicated over the SAbus by the microprocessor.

Extra module slots are provided in the card frame for versatility; the 7550 can support up to three subcarrier modulators (program audio, cue or data). Additionally, one module position is uncommitted, for use with auxiliary group delay equalizers.

A block diagram of the 7550 is shown in Figure 2, and technical characteristics are listed in Table 1.

Table 1. 7550 Video Exciter Features and Specifications

Characteristic	Specifications
Frequency	5.925 to 6.425 GHz 125 kHz step size synthesizer L.O.
RF Level	0 to +10 dBM
IF Bandwidth	36 MHz (full-transponder video)
	17.5 MHz (half-transponder video)
	Others for special formats
Video Modulator	Sync-referenced AFC
Subcarrier Modulators	Three internal, auxiliary external input
	5.0 to 8.0 MHz, synthesized
	<1% THD, 50 to 15000 Hz
	Cue input
	600 Ω balanced audio
Video Distortion	<1% differential phase, 1% differential gain measured back-to-back with 7500 Video Receiver
Remote Interface	SAbus; parallel frequency
Size	5-1/4 by 19-inch standard rack mount
Power Consumption	<100W

Upconverter

The upconverter section of the 7550 is completely modular and is removable from the exciter mainframe as a unit. In addition to RF circuitry, the upconverter houses the monitor and control microprocessor for the entire exciter. A complete block diagram of the upconverter is shown in Figure 3.

IF signals at 70 MHz enter the upconverter and are amplified in the mixer-preamplifier module, then converted to the 1115-MHz second IF frequency. A post amplifier at 1115 MHz with a gain of 20 dB provides a high-level RF signal to drive the output mixer. Gain of this module is adjustable, and sets the overall converter gain in the range of -5 to +30 dB. For normal video systems, gain is set at 17 dB.

A phase-locked oscillator at 1045 MHz provides the LO for the first upconversion. The 5-MHz reference oscillator is provided by a TXCO in the second LO module.

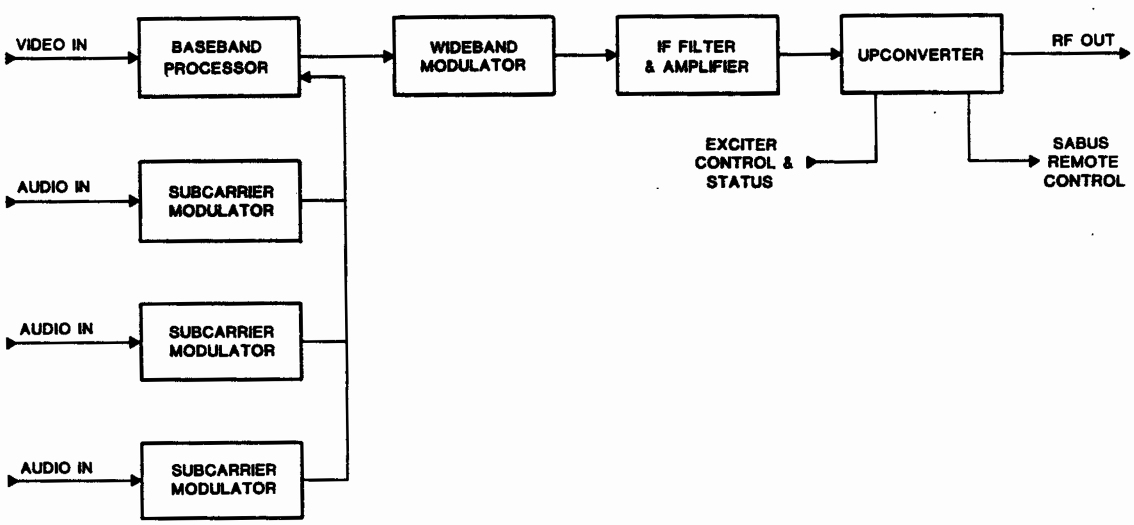


Figure 2. 7550 Block Diagram

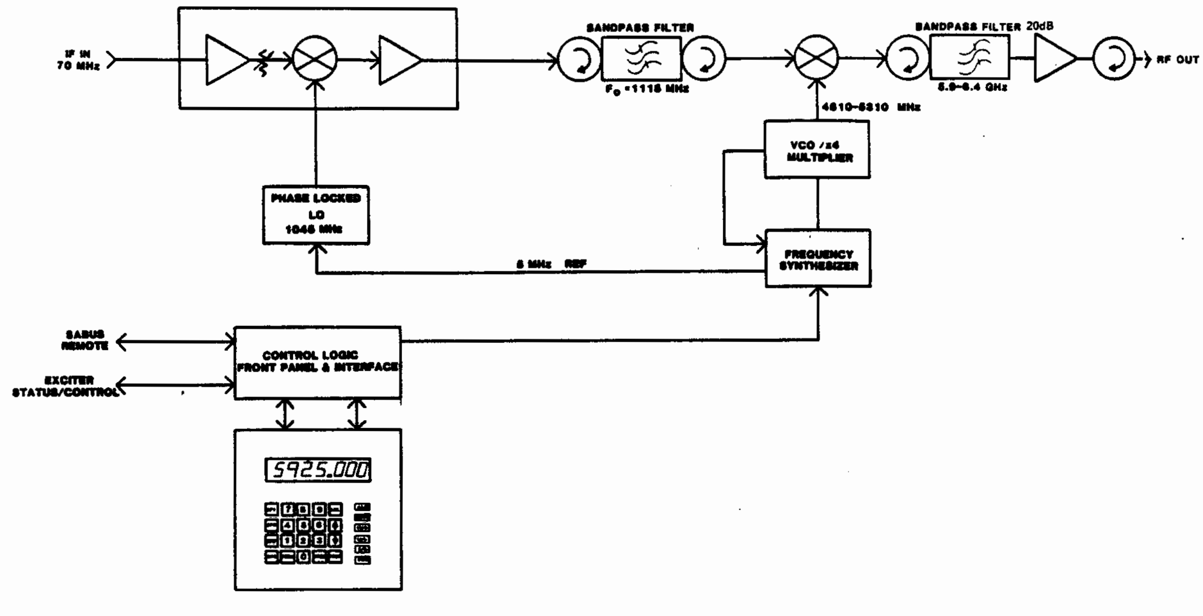


Figure 3. Upconverter Block Diagram

After conversion to 1115 MHz, the IF signal is bandpass-filtered to remove unwanted mixing products, and then upconverted to the final output frequency of 5.9 to 6.4 GHz in a high-level mixer. A frequency-synthesized oscillator operating in the 4810- to 5310-MHz range provides the local oscillator for the final upconversion. Frequency step size of 125 kHz is provided in the synthesizer. A 5-MHz TCXO is built into the synthesizer module. This signal is also sent to the 1045-MHz LO. Additionally, provision is made for use of an external 5-MHz reference oscillator, if desired. The RF signal at the 6-GHz output frequency is again bandpass-filtered to remove unwanted mixing products. An amplifier with 20 dB gain provides power output of up to +10 dBM.

Video Modulator

The heart of the 7550 Exciter, and the key to its high performance, is the video modulator. Advanced circuit design techniques in the modulator improve video performance over a satellite link by reducing video distortion to almost imperceptible levels. This is accomplished in two ways: use of an extremely linear FM modulation technique, and with an AFC system that ignores the APL of the incoming video. Linear frequency modulation of the carrier is generated by modulating two voltage-controlled oscillators in opposite directions. The two oscillator outputs are mixed, resulting in the 70-MHz IF output frequency. This technique results in typical modulator nonlinearity of less than 0.5% over a 60-MHz modulation bandwidth.

Low video distortion at varying APL is guaranteed by the unique AFC system in the 7550. This AFC samples the carrier frequency only during the horizontal sync pulses and locks the sampled carrier frequency in a phase-locked loop frequency synthesizer. This results in an AFC system that ignores the APL of the video, thereby keeping the modulated waveform optimally positioned in the IF filter bandwidth. Greater utilization of the available bandwidth results. In addition, differential phase and gain are not dependent on APL, and bounce test signals do not disturb the modulator AFC.

The carrier frequency at the sync tips can be set to any desired frequency in the IF bandwith on-board switches. In the event the modulator loses sync, it reverts to a conventional averaging AFC mode.

IF Filter/Amplifier

Spectrum control filtering of the modulated carrier is performed in the IF filter/amplifier module. Normal full transponder video requires a 36-MHz bandwidth filter. In addition, this module performs the exciter ALC function, ensuring a constant IF output power.

Baseband Processor Module

All functions involving processing of the video baseband are contained in the baseband processor module. Video entering the module is buffered, preemphasized, and passed through a low-pass roofing filter. The module contains preemphasis for both 525L and 625L video, and two roofing filters. A switch selects 525- or 625-line operation; this switch can also select a flat response.

Both horizontal sync and vertical sync are detected in the baseband processor. Horizontal sync is provided for the modulator AFC system. Vertical sync is used to synchronize the energy dispersal waveform. The triangular energy dispersal waveform is synchronized with the incoming video vertical interval. Its level is set to deviate the carrier 1-MHz peak. If video is absent, the level of the energy disperal waveform is increased automatically.

Audio subcarriers are summed with the video baseband and energy dispersal waveform, and the composite baseband signal routed to the wideband modulator.

Subcarrier Modulator

The 7550 Mainframe is capable of supporting three audio subcarrier modulator plug-ins. Designed for maximum versatility, the 7550 Subcarrier Modulator is frequency-synthesized and can be set anywhere between 5.0 and 8.0 MHz. With switchable preemphasis (J17, flat, 75µs) and wide deviation capability (50-kHz to 500-kHz peak, it is designed to accommodate all subcarrier formats. Audio processing includes 15-kHz low-pass filter, adjustable deviation limiting and peak level metering. Audio level is displayed in a peak reading bar graph display.

Video Receiver, Series 6600

C. Sirles/A.P. Best

Introduction

The increased employment of video receivers in geostationary satellite microwave links during the past few years has been phenomenal. The continued high volume applications in cable television systems has been joined by a vast array of new markets. These include: broadcasters, hotels and motels, home installations, condominium complexes, religious groups, educational facilities, hospitals, and many more.

In order to take advantage of the obvious economies and lower receiver costs made possible by high volume production, the overriding design objectives in developing the 6600 Series receivers were twofold. First, the receiver had to be of sufficiently high quality and reliability to meet the performance demands required by the most stringent users. Second, the unit must have the packaging flexibility coupled with the optional features which ensure application compatability with emerging markets and non-obsolescence for existing markets.

The Series 6600 Receiver line, consisting of the Model 6601 Single-Channel Unit, the Model 6602 Frequency-Agile Unit with remote tuning capability, and the Model 6603 Frequency-Agile Unit with no remote tuning capability, have proven to be very successful in the marketplace in satisfying customer requirements. The remainder of this presentation will discuss the receiver in detail.

Design Goals

Approximately five years ago, FM video receivers were of the type designed for a low volume market in its beginning stages. Features and design techniques were aimed at an application which was somewhat undefined. As a result, features and performance characteristics included many which were unnecessary, and many necessary ones were omitted.

Early in development stages of the Series 6600 Receiver line, the following basic features were defined:

- The receivers must be modular in construction to facilitate production, and to provide the versatility required by the market.
- Providing optional agility or single-channel reception must be made as simple as possible to reduce system costs and offer flexibility to a growing market.
- Powering must be made modular in order that ac or dc optional power units may be utilized.
- Space must be provided and wired in an economical manner to provide for future expansion.

- Design from a manufacturing point of view was considered important for economical reasons. Assembly methods which interface with the latest automated production methods was to be foremost as a design team goal.

- The power module would have a 50% excess capacity for powering external equipment such as low-noise amplifiers and modulators.

- Remote tuning ability should be available with command inputs compatible with video protection switch logic and simple contact closures for "cherry-picking" applications.

These goals led to a sheet metal chassis with plug-in downconverter and power supply modules. The downconverter is installed from the receiver front panel. The power pack installs from the receiver rear panel. Figures 1 and 2 are Models 6601 and 6602, respectively.

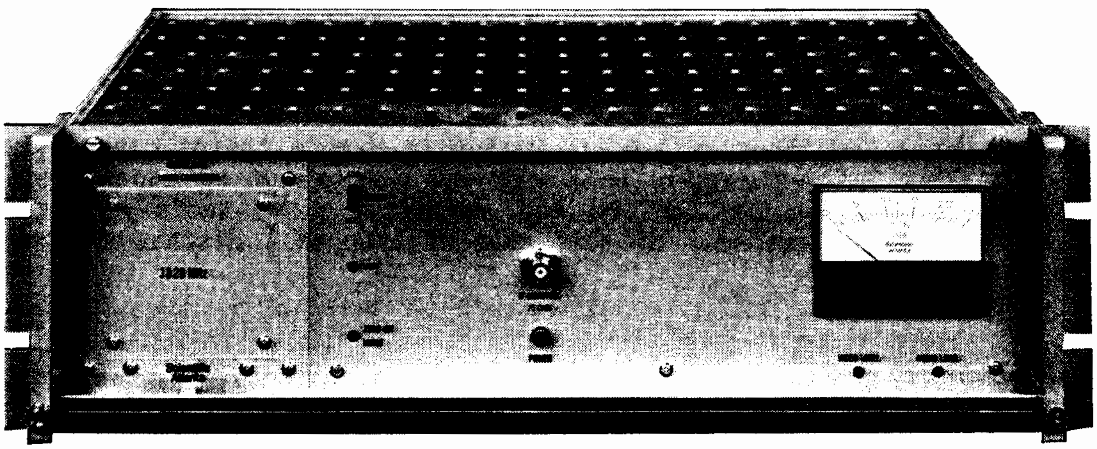


Figure 1. Model 6601 Single-Channel Receiver

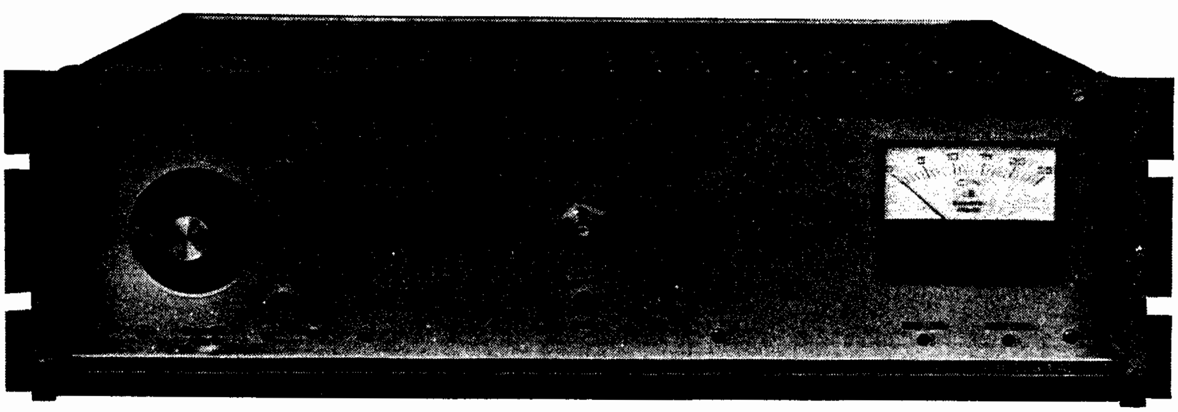


Figure 2. Model 6602 Agile Receiver

All remaining electronic modules are plug-in edge cards tied together electrically by a printed wiring board. Figure 3 shows the card rack area with interlaced sheet metal partitions. The Model 6603 looks identical to the 6602 except the remote tuning indicators are removed from the front panel.

Modular construction with interconnecting cards increases reliability, decreases cost, reduces hand-wiring labor to a minimum, and leads to efficient production and trouble location.

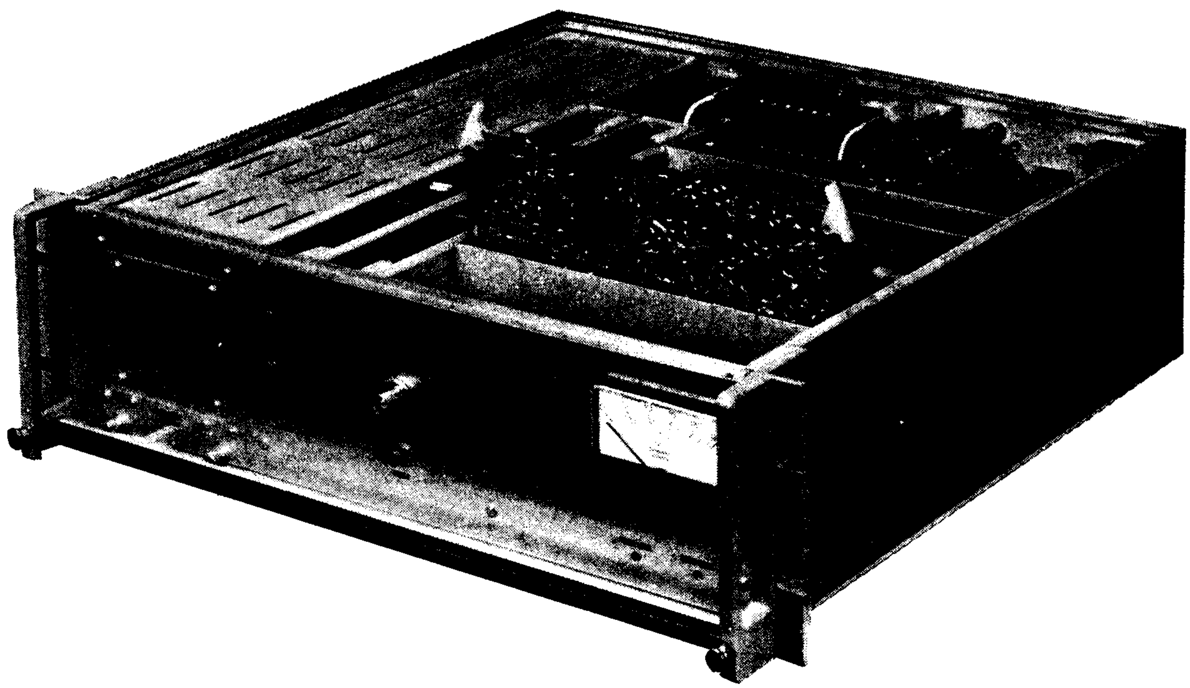


Figure 3.

General Description

A block diagram of the Series 6600 Receivers is shown in Figure 4. Received RF signals are converted to a 70-MHz IF frequency by the downconverter module A2. Amplification, filtering, and AGC follows in IF module A5. Module A6 demodulates the 70-MHz IF, and retrieves the entire baseband spectrum from 1 Hz to 10 MHz. Video and all subcarriers are present at its output terminals. The baseband signal is passed to five module slots--clamp Module A7, program audio demodulator Module A9, and three auxiliary modules A8, A10, and A11.

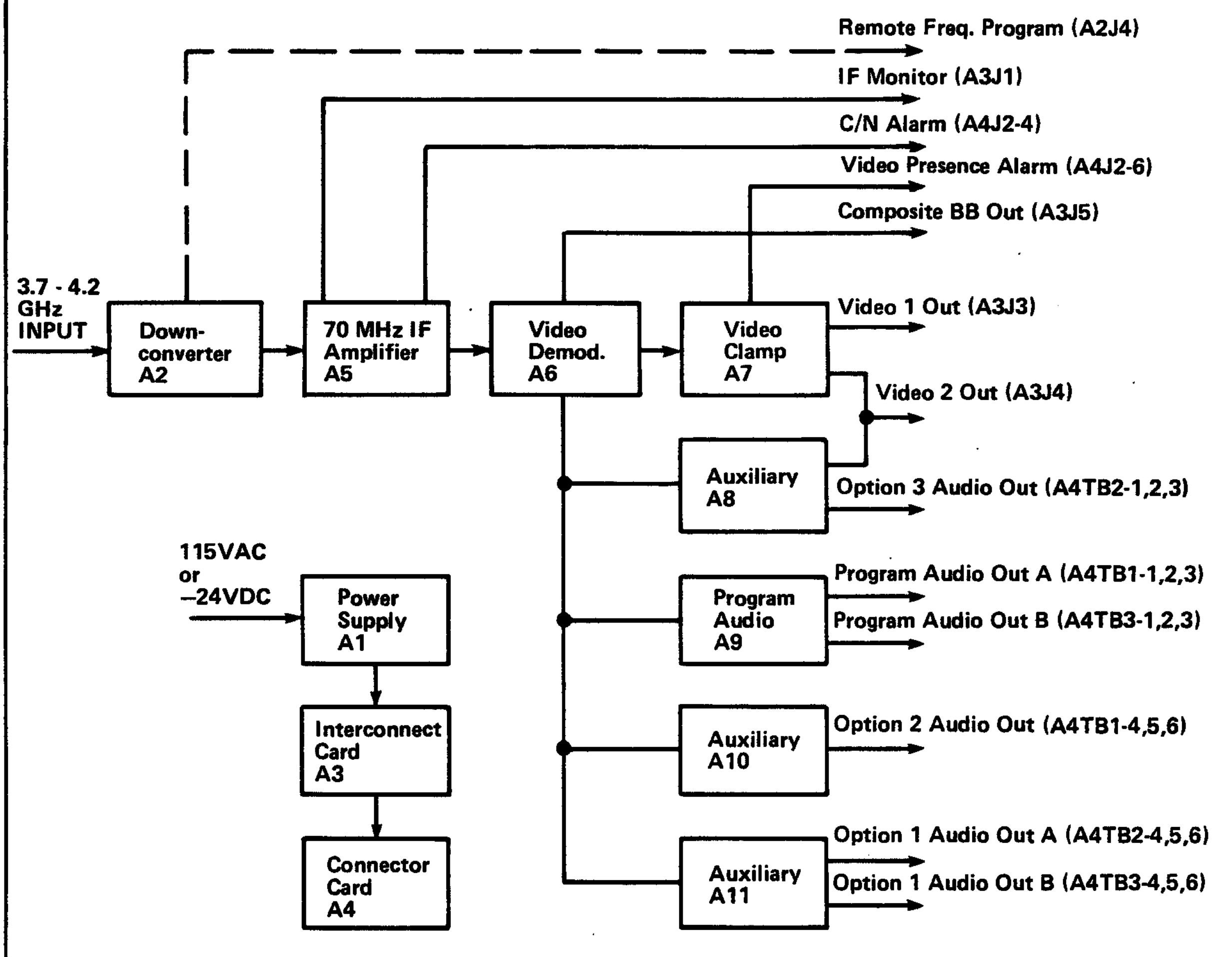


Figure 4. Series 6600 Video Receiver Block Diagram

Downconversion

RF signal frequencies of 3.7 to 4.2 GHz are first converted to an 880-MHz IF frequency by a microwave front-end. IF bandwidth at this point is 80 MHz. A second converter module heterodynes this 880-MHz IF down to 70 MHz, and the bandwidth remains 80 MHz. Overall gain in the downconverter is 20-dB nominal which means noise figure is primarily determined by the downconverter front-end.

Dual-conversion downconverters have the advantage that channel selection can be achieved by only proper selection of the first LO frequency. No tuning of the input bandpass filter is required. By selecting a first IF frequency which is sufficiently high (880 MHz), the entire 500-MHz spectrum (3.7 to 4.2 GHz) can be fed through the input image filter at the downconverter input. The received frequency is determined only by the first LO frequency which is required to center the desired transponder in the 880-MHz IF bandpass filter. The IF bandwidth at 880 MHz is sufficiently wide so that no appreciable envelope delay or amplitude roll-off is incurred.

Three different models of downconverters are available. In the Model 6601 the first LO frequency is determined by a plug-in crystal which is located inside a front-panel cutout. No tuning or adjustments are necessary. This model lends itself to the reception of any standard or non-standard channel frequency plan.

Figure 2 shows the front-panel control on the Model 6602 Agile Receiver. This control selects any one of 24 transponder frequencies by tuning the first LO in 20-MHz increments between adjacent transponders. Keyed to the control knob is a relay driver circuit which controls the position of an external polarization relay. This relay is included as a standard part on all 6602 Receivers. In addition to channel selection by the front-panel knob, the received frequency can also be programmed by six lines on the rear panel. The code required is user-selectable by internal programming jumpers. One code format is compatible with previous Scientific-Atlanta receivers. Once the remote enable line is grounded, the receiver goes to the frequency programmed on the six programming lines. The lower yellow indicator on the front panel indicates programming is remote when the remote enable line is grounded.

Another code format permits selection of one of any six of the transponder frequencies. Internal jumpers can make available any six of the 24 transponder frequencies such that each one of the code lines calls up a different frequency. This arrangement is useful in programming the receiver from an external control unit programmed to select different transponders at different times of the day.

A front panel indicator in the upper right corner is useful in determining a remotely programmed frequency. Simple rotation of the front-panel control until the REMOTE FREQUENCY lamp illuminates tells the operator what frequency program exists on the remote lines. This indication is valid regardless of the code format used as long as the receiver is being tuned remotely.

The Model 6603 is identical to the 6602 except that the 6603 cannot have its frequency tuned remotely. Also the yellow REMOTE indication lamp and the green REMOTE FREQUENCY lamp have been removed from the front panel. The polarization relay must be ordered as an option on the Model 6603 Receiver.

IF Amplification

The 70-MHz IF out of the downconverter is coupled to the input of the IF amplifier through a 75-ohm BNC jumper cable which is accessible on the rear panel of the receiver. A final IF frequency of 70 MHz was chosen for several reasons. First, 70 MHz has long been an industry standard in microwave receivers. Most all microwave link analyzer equipment is designed for 70-MHz systems. Some head-end sites are remotely located from the satellite terminal. Special arrangements can permit downconversion at an antenna site and transmission of the 70-MHz IF over long distance to a final processing site. This split receiver configuration is also compatible with fiber optic transmitters and receivers.

Under conditions of severe interference from terrestrial microwave transmitters, it is also possible to improve picture performance by inserting conventional notch filters in this 70-MHz interface line. Interfering carriers at ±10 MHz and ±20 MHz from the 70-MHz center frequency can be removed using this technique with no acceptable or degradation to the video waveform performance.

IF gain and filtering at 70 MHz is reasonably low in cost and simple to achieve. Modern filter synthesis techniques lead to uniform, predictable filter responses with good selectivity and envelope delay characteristics. Construction techniques at these frequencies can be conventional plated-through-hole printed wiring boards requiring very little labor to assemble. Parts costs are also low for operation at these frequencies.

A simplified Series 6600 IF block diagram is shown in Figure 5. Five pole pairs of a modified Tchebycheff low-ripple filter, a single-section delay equalizer, and two PIN diode attenuator sections are sandwiched between nine stages of gain.

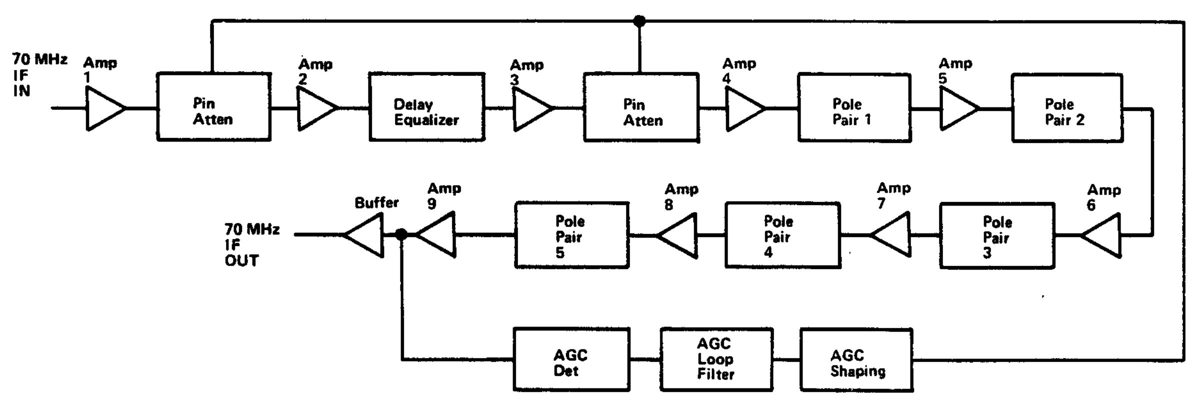


Figure 5. IF Amplifier

Each section of the IF can be aligned on its own, and when all sections are connected, little or no overall alignment is necessary. The effective noise bandwidth selected for these receivers was an INTELSAT 30 MHz. Effective noise bandwidth is 32.4 MHz. This bandwidth has become somewhat standard in the video receiver industry and is a good compromise for threshold performance and video distortion under heavily loaded modulation conditions, such as three subcarriers.

AGC control is very important in high performance FM receivers in that threshold performance and to some extent, video distortion is affected by IF level into the limiter/demodulator. This receiver series has been designed to operate with -5 dBm out of the IF. Filter shape and IF level are consistent throughout the 40-dB dynamic range of AGC. The AGC control loop contains linearization circuitry which permits C/N (carrier-to-noise) at the IF output to be read to within ±1 dB over an 8 to 25-dB reading range. A front-panel meter provides this reading. A zero-on-noise control is provided and must be set for the particular installation to ensure accurate C/N readings. An IF monitor port is also provided for external monitoring, and a separate buffer amplifier delivers -5 dBm of level into 75 ohms.

It should be pointed out that the front-panel C/N meter is only accurate for the particular transponder on which the receiver was calibrated. Once the receiver is tuned to a different channel, a change in the meter reading is only an indication that the received signal strength has changed. To measure C/N, the meter must be recalibrated for that particular transponder.

Demodulation

FM demodulation techniques have improved during the past few years, especially their operation at low carrier-to-noise ratios. With decreasing antenna sizes, threshold performance is one of the most important considerations in video FM receivers, especially with heavy subcarrier loading.

Other important considerations in video demodulators are linearity, AM to PM conversion, envelope delay, and intermodulation distortion occurring in the output video amplifiers.

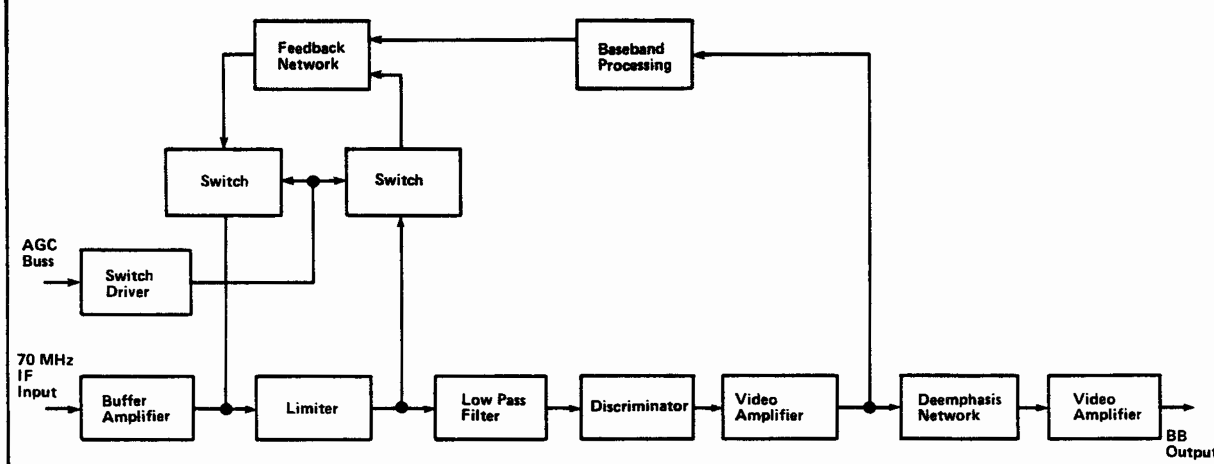


Figure 6. Video Demodulator Block Diagram

Figure 6 is a simplified block diagram of the Series 6600 Video Demodulator. The IF amplifier 70-MHz output enters a buffer amplifier which provides a return loss better than 25 dB. A limiter follows which converts the entering IF signal to a constant amplitude squarewave. A low-pass filter follows which removes third harmonic content from the waveform.

A discriminator followed by two video amplifiers and de-emphasis demodulates the IF signal. Baseband information ranging from 1 Hz to 10 MHz (including video, a program audio subcarrier, dispersal, and in some cases, additional subcarriers) is delivered by the video demodulator output stages to the remaining receiver modules.

Because of the trend to smaller antenna sizes, an early decision was made to incorporate threshold extension into all Series 6600 Receivers. An electronic switch is incorporated into the demodulator to switch threshold extension in at C/N ratios less than 10 dB. Slightly better video performance is realized above 10-dB C/N ratios when threshold extension is out. The threshold extension switch is driven from the same circuit as the C/N meter, and in most instances, can be switched on and off by adjusting the Zero on Noise Control. This will not affect the normal operation of the receiver.

Video Processing

International agreement requires energy dispersal modulation on satellite transmissions to prevent concentration of energy at one frequency. This waveform is triangular with apexes located at vertical intervals. This triangular waveform must be removed after demodulation by a video clamp.

Figure 7 is a block diagram of the Series 6600 Clamp. Video baseband and sync information are separated. A low-pass filter removes aural subcarriers and bandlimits the video to 5 MHz and below. A sample-and-hold circuit driven by the separated sync pulses samples the video waveform and clamps the video sync tip to a dc level of -0.25 volts.

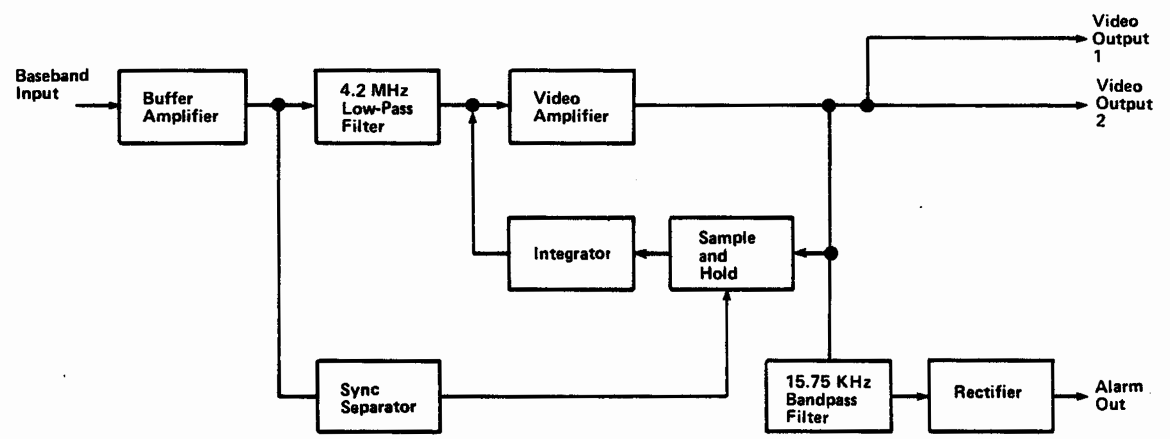


Figure 7. Video Clamp

Another circuit detects the presence of 15-kHz sync information and gives an alarm if sync is absent. An alarm level of 0 to -0.2 volts is present when sync is absent, and a level of -1 to -5 volts is present during normal operation. This line is available on the rear panel together with a C/N line. The C/N line presents 0 volts at 0-dB C/N ratios and -1 volt/10 dB for other C/N ratios. These two alarm lines are useful for protection services.

Two 75-ohm video outputs are provided by the clamp on rear panel connectors. Level at these two outputs is adjustable by a front-panel VIDEO LEVEL control.

Program Audio Demodulator

Program audio demodulation is an often overlooked important consideration in video receivers. Because audio information is also frequency modulated on a subcarrier, threshold must also be considered for audio. Modulation levels are normally chosen so that with proper subcarrier bandpass filtering, threshold occurs at slightly less main IF carrier-to-noise ratios than does video. In Series 6600 Receivers, video threshold occurs near 7.5-dB C/N ratios while audio thresholds near 6 dB.

Program audio is usually located on a 6.8-MHz subcarrier. Some satellites use other subcarriers. Accommodation of other subcarriers is only a matter of frequency-scaling a few parts.

Baseband bandwidths of 15 kHz for audio are normal for satellite transmissions. Total harmonic distortion of less than 1% coupled with a frequency response of better than ±1 dB provide high quality audio.

Figure 8 is a block diagram of the Series 6600 Audio Demodulator. Subcarrier is delivered by the video demodulator to an input buffer amplifier. A bandpass filter with a noise bandwidth of approximately 500 kHz follows. A limiter/discriminator demodulates the subcarrier. The audio baseband is bandlimited by a 15-kHz low-pass filter. Deemphasis and balanced 600-ohm outputs are provided on rear-panel terminals.

Over the past few years there has been a proliferation of the use of additional subcarriers carrying a multitude of different type information. In most instances the technical requirements necessary to recover this information vary as to audio bandwidth, pre-detection bandwidth, and deemphasis requirements. Subcarrier demodulators to handle most of these present applications exist for the 6600 Receiver, however, care should be taken in selecting the appropriate units to ensure compatibility with requirements.

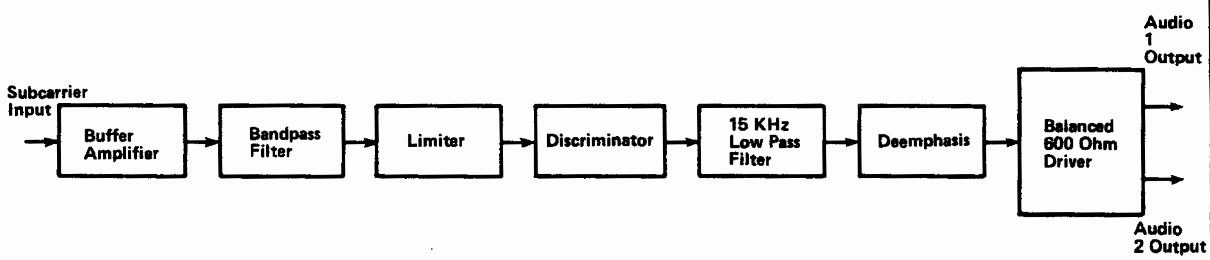


Figure 8. Audio Demodulator

Auxiliary Modules

Auxiliary Module A8 can be utilized for one of two functions. A card which modulates the audio out of A9 onto a 4.5-MHz subcarrier and adds this carrier to Video Output 2 can be provided in this position. Microwave systems can be driven direct from the radio without an additional external audio modulator. An audio demodulator can also be placed in this position for demodulation of additional subcarriers. One balanced 600-ohm output is provided on the rear panel from this position.

Auxiliary Modules A10 and A11 provide two additional slots for audio subcarrier demodulators. Module A10 provides a single 600-ohm output while A11 provides two 600-ohm outputs. Modules A8, A10, A11 together provide three subcarrier demodulators, in addition to program audio (A9). Additional links on the interconnecting card between A8, A9, A10, and A11 permit pilot carriers from the Program Audio Module to be demodulated. Queing and remote receiver control may be added in the future.

Two versions of the Power Supply Module exist. One is a conventional series regulator type which plugs into a 115 volt outlet and provides ±15 volts to the receiver. The other is a dc converter module which accepts as an input -19V dc to -32V dc and provides as an output ±15 volts. Both of these units provide short-circuit-current foldback and overvoltage protection. They also have sufficient excess current capacity to power two LNAs, a 416 Modulator, an external polarization relay and three additional subcarrier demodulators. These two units are easily interchangeable in the field.

Specifications

Table 1 is a listing of Series 6600 specifications. As compared to most overall link specifications, it can be seen that the receiver is relatively transparent.

Figure 9A and 9B give the threshold performance of the Series 6600 Receivers. The demodulator operates without threshold extension down to a C/N of 10 dB. Threshold extension is automatically switched in at a C/N of 10 dB and below.

Summary

Evolution of video receivers has occurred rapidly during the past five years. Volume production is now the goal in this expanding market. Versatility and lower cost to the customer will be the main benefit brought about by this increase in volume.

Table 1. Series 6600 Video Receiver Technical Characteristics

Characteristic	Specification
RF Input	
Maximum Level	-34 dBm
Frequency	3700 to 4200 MHz
Channel Selector	
Model 6601	Selection of channels is accomplished by changing crystals. No tuning is necessary.
Model 6602	Switch selectable - 24 channels
Impedance	50 ohms
Return Loss	>20 dB
Noise Figure	15 dB Max
Image Rejection	>60 dB
LO Leakage	<-70 dBm
IF	
Intermediate Frequency	70 MHz
Effective Noise Bandwidth	32.4 MHz Nominal
Impedance	75 ohms, unbalanced
Return Loss at IF Monitor Ports	>20 dB
Dynamic Operating Range	40 dB
Baseband	
De-emphasis	525-Line (CCIR Rec. 405-1)
Deviation Range	6 to 12 MHz peak at de-emphasis cross-over frequency
Video	
Video Level	1V peak-to-peak ±3 dB adjustable
Response (15 Hz to 4.2 MHz)	
Standard	±0.5 dB
With TED	±1.0 dB
Impedance	75 ohms, unbalanced
Return Loss	>26 dB
Polarity	Black-to-white: positive-going
Clamping	40-dB dispersal rejection
Line-Time Waveform Distortion	<1% tilt
Field-Time Waveform Distortion	<1% tilt
Differential Phase	<±1° 10 to 90% APL
Differential Gain	<±2.5% 10 to 90% APL
Audio	
Subcarrier Frequency	6.8 MHz standard, other frequencies available
Frequency Response	30 Hz to 15 kHz ±0.5 dB
De-emphasis	75 μs
Output Level	Continuously variable, -10 to +10 dBm
Impedance	600 ohms, balanced
Harmonic Distortion	<1%
Operating Temperature	0° to 50°C (32° to 122°F)
Mechanical	
Height	133.4 mm (5.25 inches)
Width	482.6 mm (19 inches)
Depth	482.6 mm (19 inches)
Power Requirements	
At 105 to 125V ac	Approximately 50 watts
At -19 to -32V dc (with optional power module)	Approximately 3.2 amperes

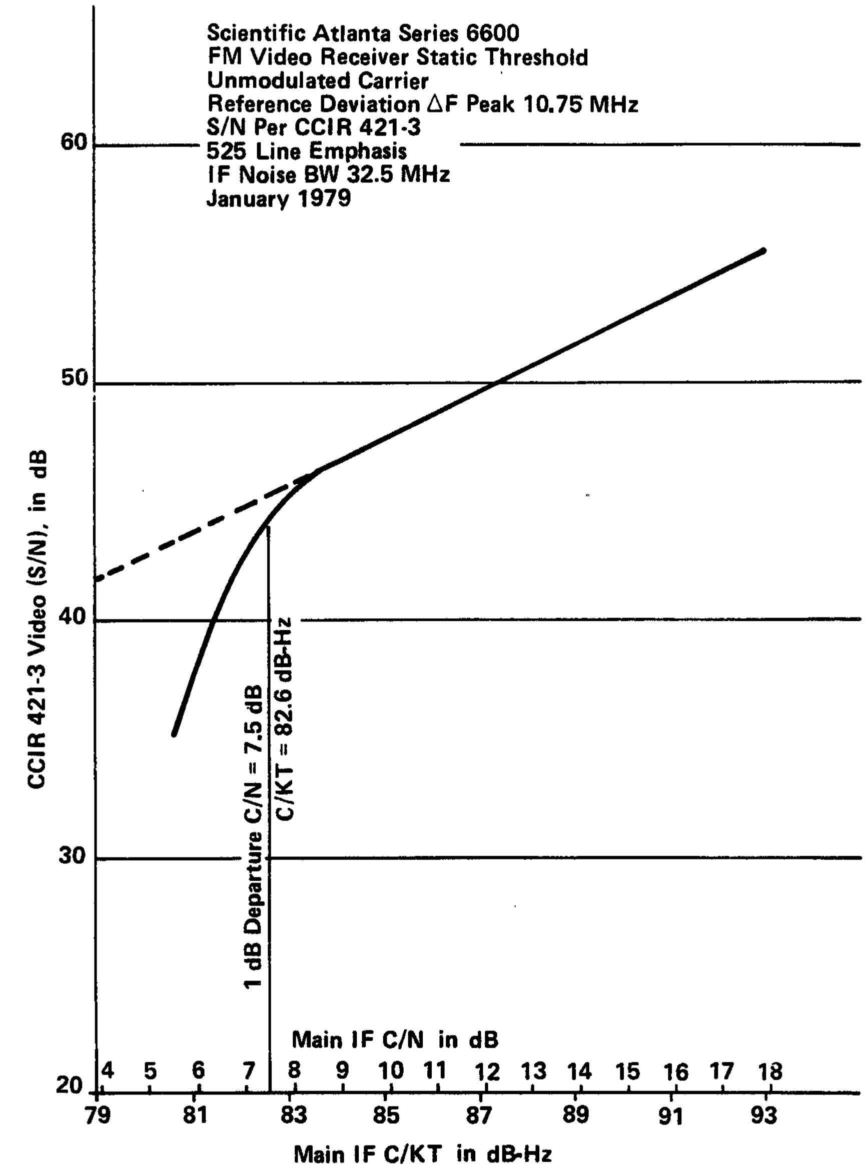


Figure 9A. Video Threshold

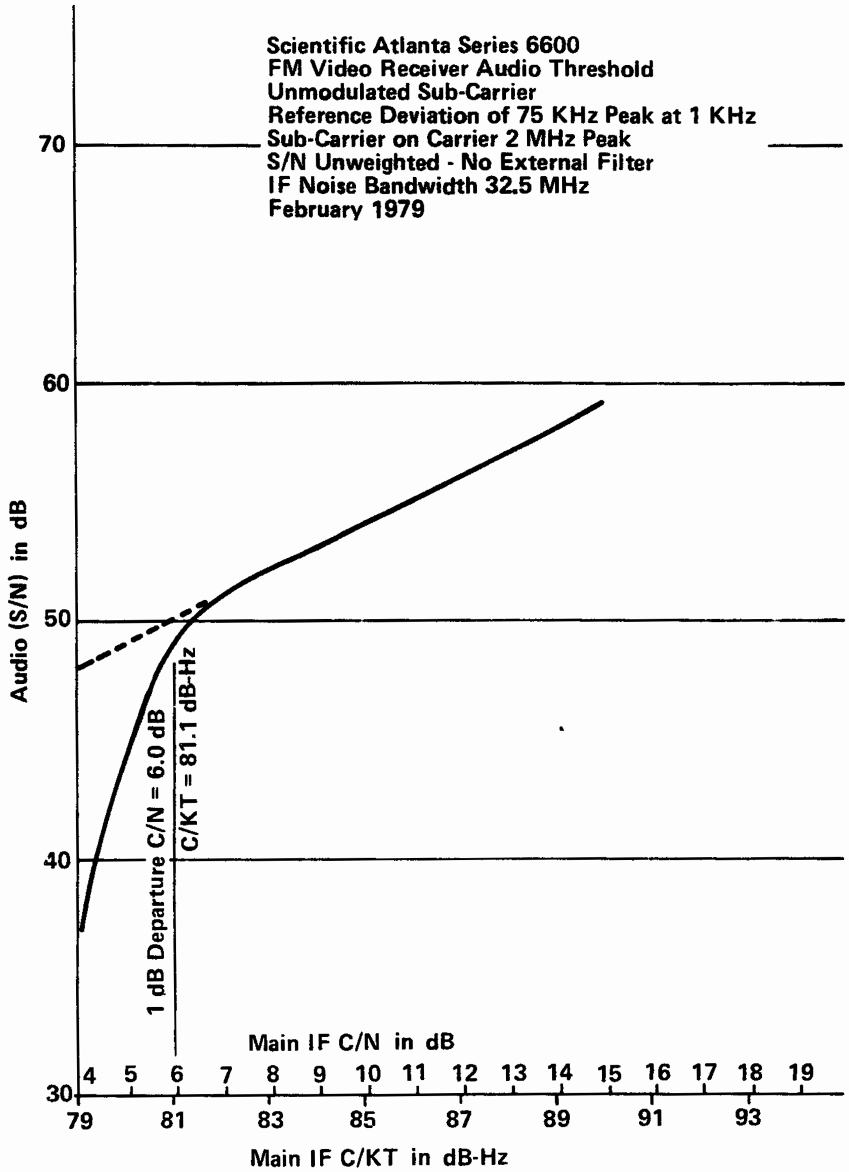


Figure 9B. Audio Threshold

Low Noise Converter, Series 360

Luis Rovira

Introduction

Traditionally, earth station receive electronics consist of a low-noise amplifier at the antenna and a length of microwave cable connecting it to one or more indoor receivers. The receivers are equipped with microwave down-converters to translate the 3.7- to 4.2-GHz satellite downlink frequencies to a lower intermediate frequency (IF) band.

The low-noise converter (LNC) offers an alternative to this approach. The LNC combines the low-noise amplifier and a block downconverter into one antenna-mounted package. The IF signal is then cabled to the indoor receiver, where channel selection and further signal processing occur.

This paper discusses the advantages of frequency block conversion at the antenna, and describes the design trade-offs and circuits used in the Scientific-Atlanta Series 360 Low-Noise Converter.

The Block Conversion Approach

Most earth station receivers in use today get their inputs from the antenna electronics directly at the 3.7- to 4.2-GHz satellite downlink band (Figure 1a). Professional quality receivers are usually of the double-conversion type and must include a microwave local oscillator (LO), mixer and filter in their first converters. Channel selection within the band is done in a receiver by choosing the first LO frequency to mix only the desired channel down to the first intermediate frequency (IF).

The block converter approach (Figure 1b.) takes the first frequency conversion of the double-conversion system and moves it out to the antenna with the low-noise amplifier (LNA). However, channel selection can no longer occur at the first converter* since that converter may feed more than one receiver; instead, channel selection must be done by selecting the LO frequency in the remaining converter in each receiver.

The function of the block converter is, as the name implies, to convert the entire block of frequencies (channels) to the input frequencies of the receiver (270 to 770 MHz in the Series 360 LNC and 6650 Video Receiver).

*Some home earth stations are appearing on the market which do select a channel at the antenna, but obviously such a system would not feed multiple receivers.

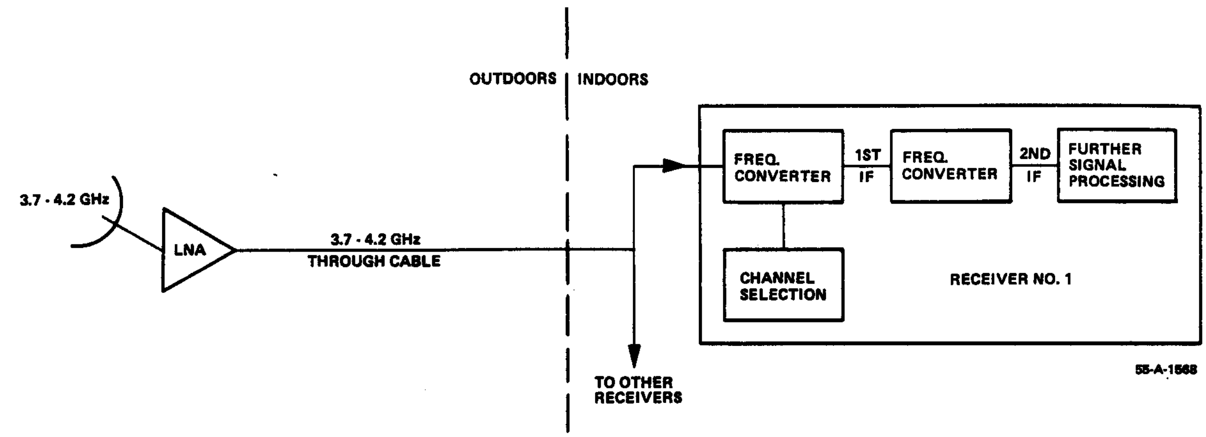


Figure 1a. A Typical Earth Station

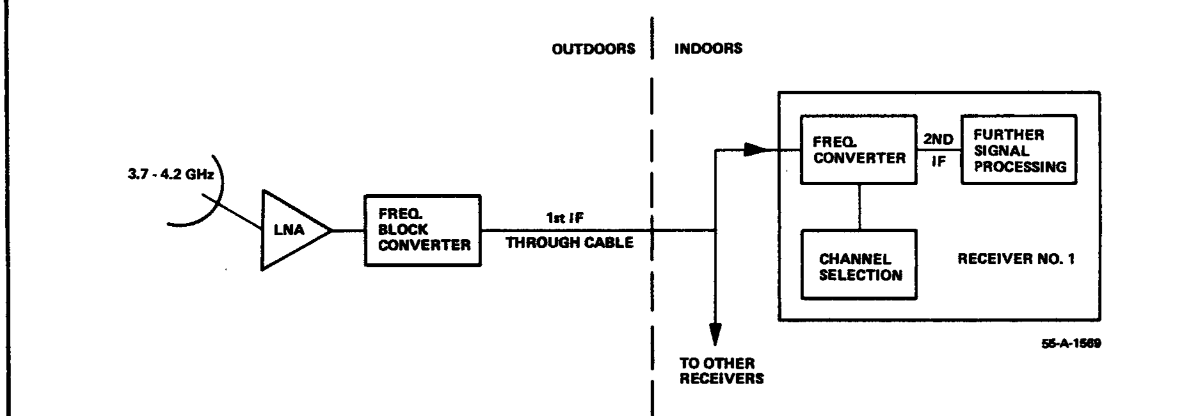


Figure 1b. Block Conversions at the Antenna

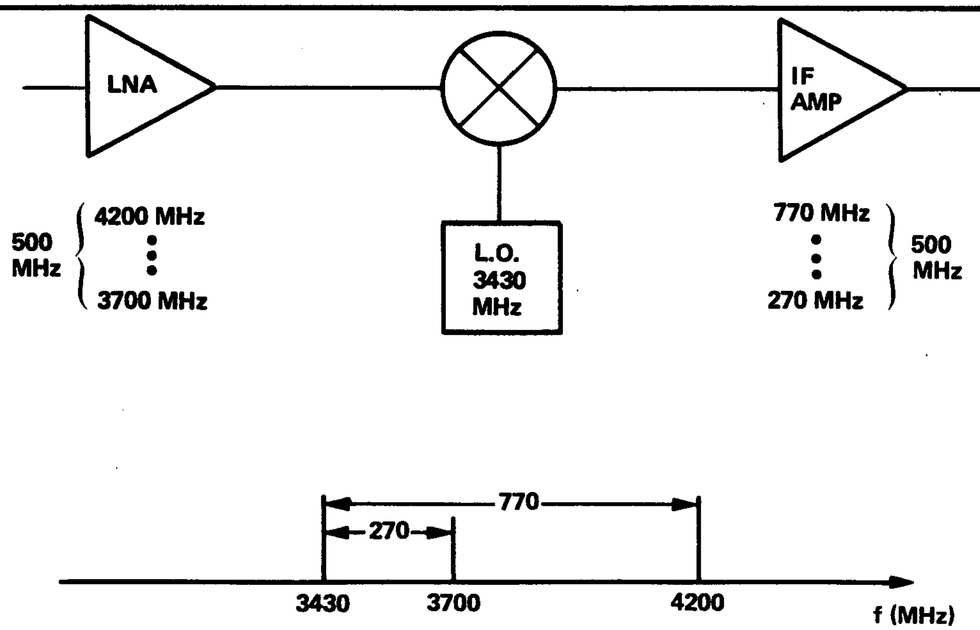


Figure 2. A Simplified Block Converter and its Frequency Scheme

Figure 2 shows a simplified block diagram of the converter. After amplification, the 3.7- to 4.2-GHz signal is mixed with a fixed-frequency local oscillator (LO). The IF band which results then depends on the difference between the input frequencies and the LO.

Advantages

The design and testing of equipment which must operate over a wide temperature range is more difficult and time consuming than that of equipment which operates indoors in a controlled environment. Yet, there are significant advantages to moving the first frequency converter outdoors.

Any given type of coaxial cable becomes increasingly lossy as the frequency through it is increased. Adding length to a cable run also increases the loss, for a given type of cable. Also, cable types with lower loss characteristics are more expensive than lossier coax. Considering these factors it is easy to see the advantages in cabling cost and flexibility of converting to a lower frequency band at the antenna. Lower cost cable can be used for a given distance, or much longer runs than previously possible (without a line-amp) can be designed into the system using a more expensive, lower loss cable.

Another important advantage to the block conversion scheme is that multiple receivers can now share more of the electronics. An LNC at the antenna is more expensive than just an LNA, but only one is needed for several receivers. The receivers contain no expensive microwave electronics and only need to perform one frequency conversion, instead of the usual two.

Choice of IF Frequencies

The choice of IF frequencies for such a converter, although essentially a systems problem, has some important implications for the electronics designer. The most important of these fall into the following categories:

- Image Rejection
- % Bandwidth at IF
- Spurious Inteference
- Technology Used
- Cable Slope and Loss

Image Rejection

Figure 3 expands on the frequency chart of Figure 2. By analogy to a mirror placed at the LO frequency in Figure 3, frequencies on the opposite side of the LO from the desired band, and separated from the LO by the same amount as the desired band, are referred to as image frequencies.

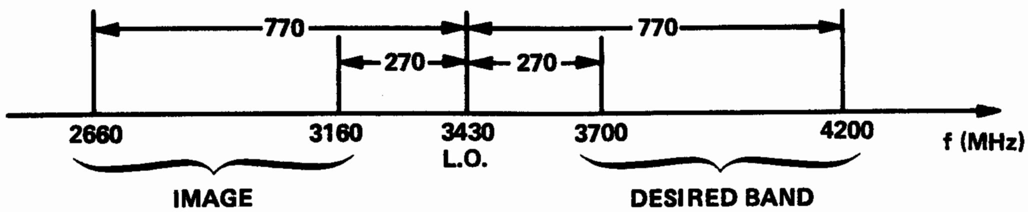


Figure 3. Location of Image Frequencies

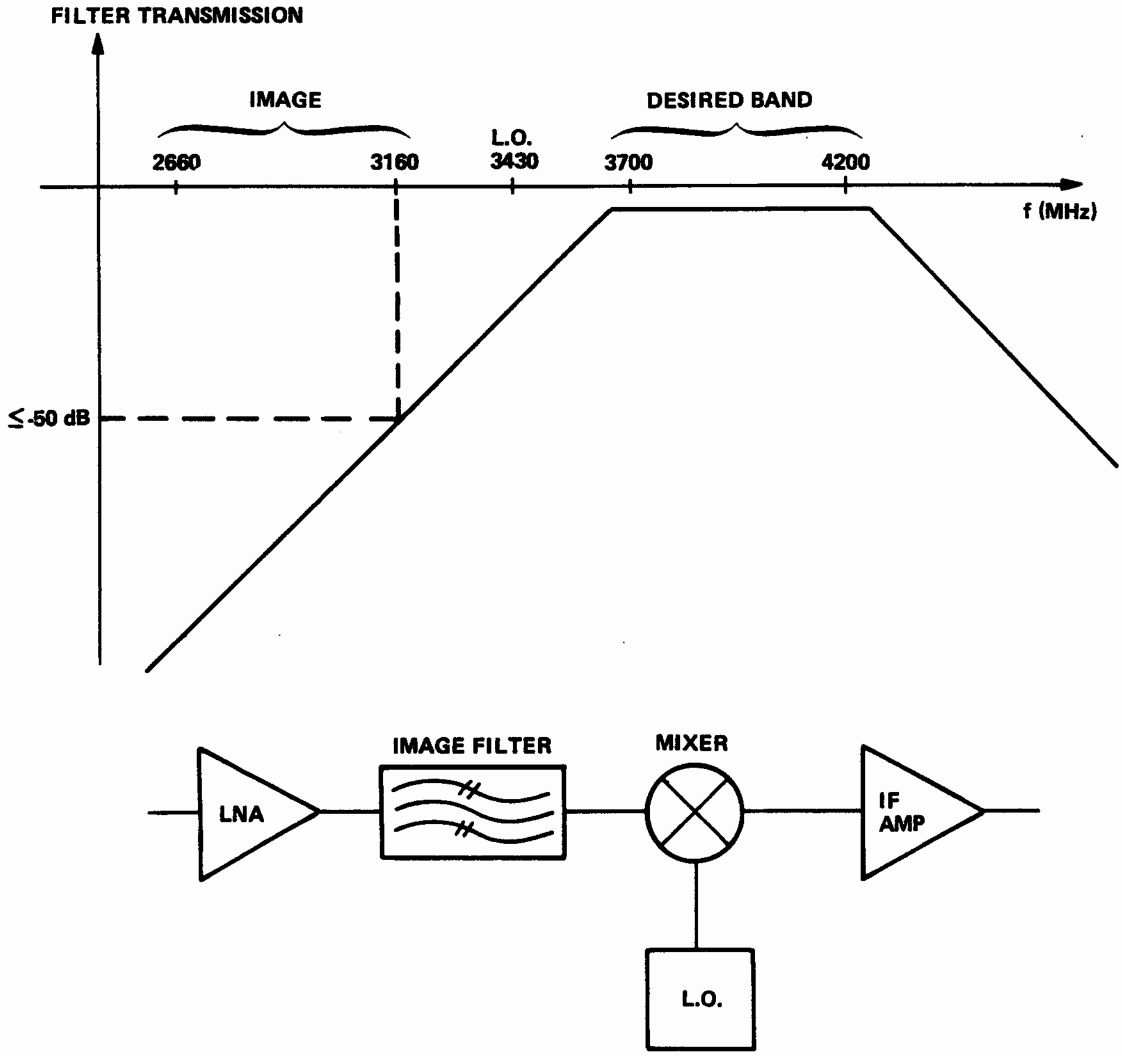


Figure 4. Location and Response of Image Filter

If any signal is present inside the image band, ranging from 2660 MHz (3430 - 770) to 3160 MHz (3430 - 270) in our case, that signal will also be converted down to the IF band and will cause interference to a desired channel.

Once the interfering signal is in the IF band, filtering cannot be used to remove it. For this reason, the image must be rejected before it can be frequency-converted. In the Series 360 LNC, this is accomplished using an image filter which passes the desired band and rejects the image before it can get to the mixer (see Figure 4). More will be said about this filter in a later section.

As one can see from Figures 3 and 4, the lower the choice of IF frequencies, the closer the required LO moves to the desired band. At the same time, the image moves closer to the desired band twice as quickly. The image rejection of a given filter decreases as the image approaches the desired band, so either more interference must be allowed or a better filter is required.

Image rejection considerations, then, would favor an IF band as high in frequency as possible, to move the image far from the desired band and alleviate the filtering requirements.

Percent Bandwidth at IF

Electronic circuits tend to be frequency selective, and circuits that force this selectivity to either extreme--highly selective or very broadband--require special design considerations.

A useful measure of this selectivity is the percent bandwidth, as defined below:

$$\% \text{ BW} = \frac{f_H - f_L}{f_C} \times 100$$

where:

f_H is highest frequency

f_L is lowest frequency

f_C is center frequency

In a block downconverter scheme, since the entire band is converted, the numerator in the above equation remains constant, but the denominator decreases. As a result, the percent bandwidth necessarily increases. The lower the IF center frequency chosen, the larger the percent bandwidth will be. This consideration would favor a high IF.

Spurious Interference

In some schemes, and for reasons to be discussed later, the LO is not simply an oscillator at the LO frequency, but can be a signal generated by combining various other frequencies. Whatever generation scheme is chosen, it is necessary that both the IF frequencies and the LO generation method be chosen so that unwanted frequency products do not fall in the input band or the IF band and cause interference. Each scheme under consideration must be examined carefully.

Technology

As far as the technology is concerned (i.e., the types of components, structures and circuit materials used), the radio frequency spectrum splits up into three very arbitrary categories.

Under 100 MHz, components with wire leads are usually used. Distributed components, such as quarter-wavelength transmission lines and capacitive or inductive stubs, are seldom used because they take up too much room on circuit boards. Circuit board materials with closely controlled parameters are not necessary, since the boards are used only as a way of supporting and interconnecting components. Silicon semiconductor devices are inexpensive and, in the lower part of this frequency range, integrated circuits are commonly used.

Above 1000 MHz, microwave techniques are used. Wire lead lengths on components are significant fractions of a wavelength and thus are avoided. Instead, leadless "chip" resistors and capacitors (not to be confused with integrated circuit "chips") are common. Lumped inductors are rarely employed. Distributed or transmission line components and cavities are used for reactive matching and filtering. The circuit boards themselves, being part of the distributed components, are made of materials with closely controlled losses, dielectric constants, and dimensional parameters. Materials such as teflon-fiberglass and alumina ceramics, as well as more exotic ones, are often used. Semiconductor devices used are specialized and expensive.

Between 100 and 1000 MHz a technological "in-between" region exists. Distributed components are often too long for the size constrained packages that go at an antenna feedpoint. Lumped components are difficult to deal with. Component leads act as inductances, and stray capacitances may cause problems. Connecting paths on printed circuit boards cannot be considered simple nodes. Components exhibit self-resonances and signals cross couple to other circuits. These problems are not insurmountable, but care must be exercised in circuit design and board layout.

On the positive side, semiconductor devices for use at these frequencies are common and less expensive than microwave semiconductors. Wire-leaded components are less expensive and easier to work with than chip components. The result is that once the design and layout difficulties are overcome, the final product is usually not as expensive as a microwave circuit.

Cable Loss and Slope

As mentioned previously, the coaxial cable used between the antenna electronics and the indoor receiver has increasing loss at higher frequencies. A corollary effect to this is that a higher percent bandwidth (lower IF frequencies) would exhibit more slope in the signal level versus frequency at the input of the receiver.

This is not a serious problem and is easily remedied using a slope equalizer. An equalizer would have to be used anyway at almost any IF, and getting more slope compensation from the equalizer is not difficult.

To summarize this section, we emphasize that the choice of the comparatively low 270- to 770-MHz IF was certainly not without its disadvantages in circuit design considerations. However, it was felt that the disadvantages of 96% bandwidth and difficult image rejection, as well as the time-consuming problem of circuit design at these frequencies, were well worth the substantial savings in cable costs and in simpler receivers without microwave front ends, particularly in multiple receiver installations.

Block Diagram Description

Having looked at some of the "systems" considerations and their general effects on circuit designs, let us now turn to a more specific description of the Series 360 LNC. Figure 5 shows a block diagram of the converter.

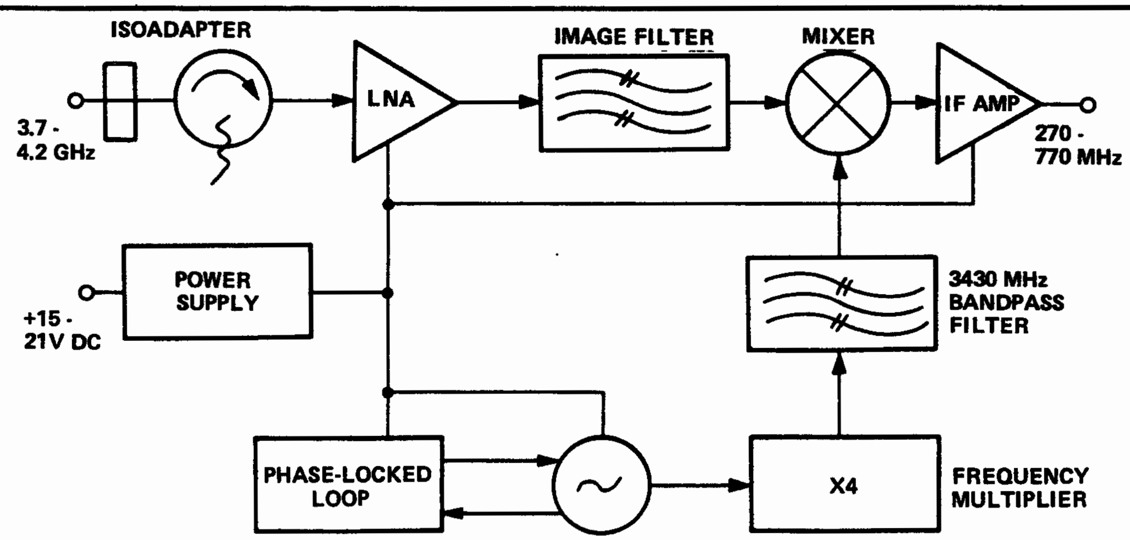


Figure 5. Series 360 LNC Block Diagram

Isoadapter

The isoadapter performs three main functions. It provides a transition from the waveguide input to the TEM coaxial mode of propagation required to feed the low-noise amplifier input, properly terminates the antenna feed, and provides the LNA with the constant, resistive source impedance from which to derive the minimum noise figure match for the critical first stage of the amplifier. The isoadapter accomplishes the latter two functions using a ferrite circulator with one port terminated into 50 ohms. This circulator acts as an isolator which allows a wave to pass from the input port to the LNA, but not in the reverse direction. Any energy reflected back from the LNA is dissipated in the 50-ohm termination. Thus, the antenna feed sees no reflected wave, and by definition, this is a good termination. Similarly, the LNA sees only the incident wave and no reflection from the isolator; thus its source impedance is fixed.

LNA

The 3.7- to 4.2-GHz low-noise amplifier used in the converter is essentially the first four stages of the series 300 LNA described in Ken Johnson's "Low-Noise Amplifiers" paper in the symposium collection. The reader is referred to that paper for further details. The amplifier consists of two Gallium Arsenide Field-Effect Transistor (GaAs FET) stages cascaded with two silicon bipolar transistor stages. Its purpose is to establish the noise temperature of the converter and to provide sufficient gain to overcome the losses and noise temperature contributions of the image filter, mixer and IF amplifier. These losses and noise contributions are not as great as those of a cable at microwave frequencies, and hence the greater gain of the Series 300 LNA is not required.

Image Filter

The need for an image filter has already been discussed in a previous section. It should be noted that, once the filter is designed, a good filter that provides more image rejection is no more costly than a simple one. The reason is its microstrip construction.

A microstrip transmission line can be constructed by taking a piece of double-sided copper-clad circuit board material, leaving one side fully copper-clad, and etching all the copper off of the other side except for a flat line of specified width (see a cross-sectional view in Figure 6). Assuming the copper is thin enough, a TEM mode transmission line is formed, whose characteristic impedance is determined by the width W, the thickness of the board H, and the dielectric constant of the board material, ε. Various lengths and interconnections of these lines can be used to make the distributed elements and resonators needed to build interstage matching networks and filters used in microwave circuits.

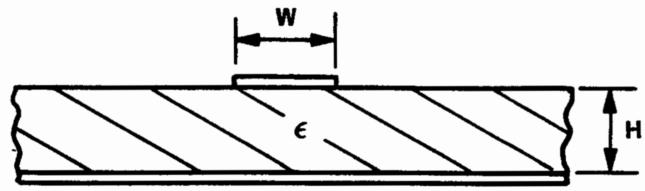


Figure 6. Cross Section of a Microstrip Transmission Line

Figure 7 shows a top view of a simple microstrip "coupled-resonator" filter and its lumped element equivalent. Coupling of the resonator occurs across the gaps between the quarter wavelength sections of etched lines on the circuit board. In our case, where a low IF requires an image filter with a steep roll-off on the low frequency side (see Figure 4), all that is required to increase the order of the filter is the extra board area for additional resonators.

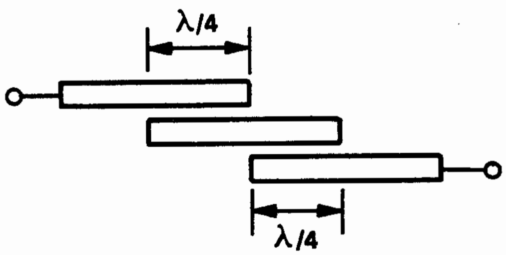
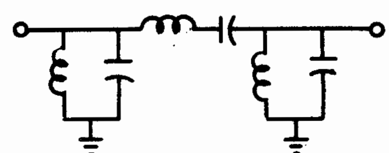


Figure 7. Microstrip Filter



Lumped Element Equivalent

It is much easier to interface to transistors and other lumped components with this type of construction than with resonant cavities and the various other types of transmission lines. Circuits constructed on a single dielectric substrate using a combination of microstrip transmission line components and lumped components are referred to as "microwave integrated circuits," or MICs.* All of the radio frequency circuits in the LNC are made using MIC technology.

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* Not to be confused with "monolithic microwave integrated circuits or MMICs" which are made entirely on one semiconductor substrate and are similar to the IC "chips" used at lower frequencies.

Mixer

The mixer uses two Schottky Barrier diodes which are switched alternately ON and OFF by the local oscillator. When the 3.7- to 4.2-GHz signal is passed through these diodes, the waveform is chopped by the switching of the diodes. It can be shown that this chopped waveform has components in the IF frequency band, and this IF band is extracted from the input and LO components by a microstrip low-pass filter at the output of the diodes. The diodes are fed by a microstrip 90° coupler which isolates the two input ports and switches the diodes in a phase relationship that results in only the IF signal being in-phase at the output of the two diodes. This balanced arrangement also helps reduce the levels of the input signals at the output of the mixer.

IF Amp

Requirements for the IF amplifier include properly terminating the mixer and its filter through a broad (96%) bandwidth, providing enough broadband gain (30 dB) to overcome the losses and noise temperature of the cable and receiver, and matching to the cable. The amplifier must also be tolerant of loading errors, such as short or open circuits, without suffering damage or going unstable.

Three feedback transistor stages of 10-dB gain each are used in the LNC IF amplifier. Each stage uses a high-quality microwave transistor at these relatively low IF frequencies to obtain a nearly ideal "gain block". This is then surrounded by heavy negative feedback to make the resulting stage unconditionally stable and independent of device variations from lot to lot. Broadband input and output matching is also accomplished by proper choice of feedback resistances. An attenuator is added at the output to protect the last stage against extreme load variations. Extra dc current is also used in the last stage to handle high levels without distortion.

L.O.

What has previously been referred to as simply the "local oscillator" actually encompasses four of the blocks shown in Figure 5--the phase-locked loop (PLL), voltage-controlled oscillator (VCO), 4X multiplexer, and 3430-MHz bandpass filter.

Considering that only one LO frequency is required for a block converter, the reason for this complexity bears some consideration. The two most important specifications on a signal source or oscillator are frequency stability and phase or frequency noise.

Poor frequency stability in a local oscillator could cause the IF signal to drift partially off the "edge" of an IF filter and possibly cause distortion, loss of signal-to-noise ratio or loss of information. Drift in the converter oscillator could be corrected by using automatic frequency control (AFC) in the receiver LO, but this can add complexity to each of the receivers in a multi-receiver installation. It would be more efficient to design a more stable LO in the LNC.

Noise is often thought of as being strictly an amplitude variation, but in fact has a frequency component as well. This frequency component of noise is particularly important in an FM system, where it can be discriminated and can reduce the ultimate signal-to-noise ratio. This noise can be generated as a phase or frequency "jitter" in any of the local oscillators of a system. High-frequency oscillators tend to have more such jitter. This frequency noise in the local oscillator is added to the desired signal and is not reduced in the downconversion process.

Choices for the type of LO in a converter such as the Series 360 LNC include cavity-controlled oscillators, dielectric resonator oscillators and automatic frequency control of the oscillator.

The cavity oscillator is difficult to integrate with MIC technology. It has good phase noise performance at frequencies far from the carrier and not so good close in. It also has some tendency to drift with thermal changes of cavity dimensions.

The dielectric resonator oscillator is very similar to the cavity, except that the resonant "cavity" is actually a solid made of a high Q dielectric. Work on DSOs is presently going on at Scientific-Atlanta.

Automatic frequency control (AFC) can control drift, but requires a control signal from the receiver and its associated cabling problems. It does not have good phase noise performance.

The quartz crystal-controlled oscillator has excellent phase noise characteristics and the best resistance to drift of any of the practical approaches at this time. The problem is that the fundamental frequency of the quartz crystal is limited to less than about 25 MHz. However, it is possible to phase-lock a higher frequency oscillator to a crystal oscillator and get excellent frequency stability and noise from the high-frequency oscillator.

A problem with phase-locking to a quartz crystal oscillator is that the phase-locking circuitry is itself limited in frequency. The solution to this is to oscillate at a lower frequency and frequency-multiply. In the 360 LNC, the oscillator operates at 857.5 MHz and this is multiplied by N in a step-recovery diode circuit (see Figure 5), where N is an integer. The 3430-MHz bandpass filter then removes all but the N=4 signal.

The phase-locking scheme used to stabilize and reduce the phase noise of the 857.5-MHz oscillator is shown in Figure 8.

The output of the 857.5-MHz voltage-controlled oscillator (VCO) is sampled by the phase-locked loop (PLL) board.* This signal is divided by 64 and compared with a quartz crystal-controlled reference oscillator. An error signal is generated by the phase/frequency comparator that is proportional to the difference in frequency or phase between the reference and the divided signal. This signal is then filtered and amplified to produce a dc control voltage, which causes the VCO output to change in the direction required to correct the error.

*Standard nomenclature refers to the VCO as part of the phase locked loop.

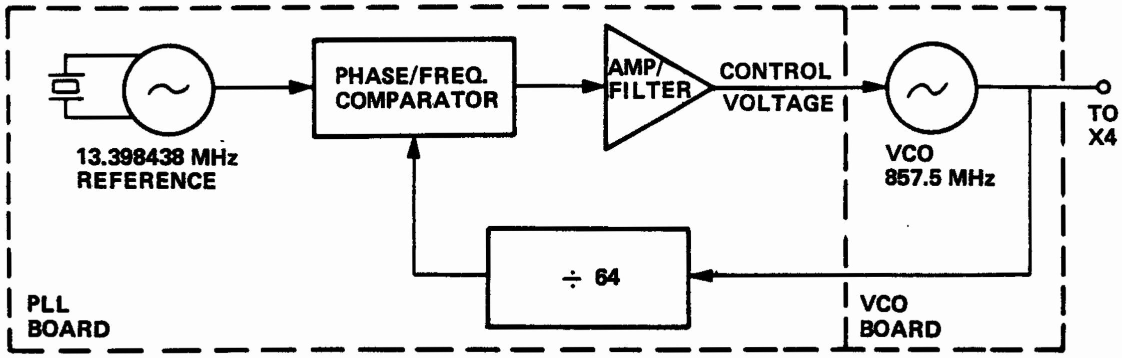


Figure 8. Phaselocked Oscillator Used in Series 360 LNC

Not only does this scheme reduce drift to a minimum, but VCO phase jitter within the bandwidth of the phase-locked loop is corrected out, and what remains is the much lower phase noise of the reference oscillator.

Power Supply

The LNC receives unregulated dc power either through the center conductor of the cable, or through terminals on the housing. The power supply board takes this voltage and converts it to the regulated supply voltages required by the circuits.
The power supply board also uses a voltage inverter to convert the positive dc voltage into a negative voltage to reverse-bias the FET gates in the LNA.

Specifications

The following pages show a data sheet for the Series 360 LNC.

Conversion gain is specified for the LNC instead of the usual simple gain to indicate that the input and output frequencies are different.

Return loss is a measure of how close the actual input or output impedance is to the specified nominal. For example, the nominal input impedance of the LNC is 50 ohms. Actually, the input impedance is close enough to 50 ohms that a perfect 50-ohm source would see a reflected wave from the input that is at least 20 dB below the incident wave.

Image rejection, noise figure, and the other specifications have already been mentioned or are self-explanatory.

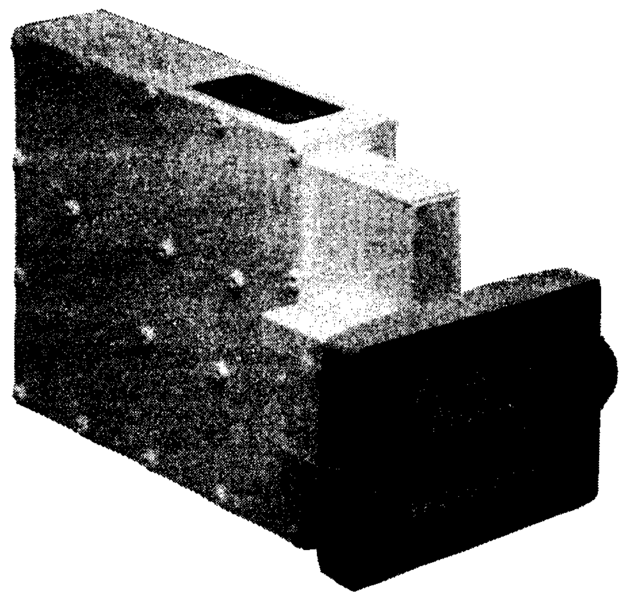
Specifications

The following pages detail the specifications for the Series 360 LNC.

Conversion gain is specified for the LNC instead of the usual simple gain to indicate that the input and output frequencies are different.

Image rejection, noise figure, and the other specifications have already been mentioned or are self-explanatory.

Frequency Range
3.7 to 4.2 GHz
Input Level
-75 dBm to -95 dBm per channel
Input Impedance
50 ohm
Return Loss
20 dB Min
Conversion Gain
56 dB ±3 dB
Image Rejection
50 dB Min
IF Frequency Band
270 to 770 MHz
IF Output Impedance
75 ohm
IF Return Loss
17 dB Min
Temperature
-40°C to +60°C full specification
Input Connector
CPR 229G flanged
Output Connector
Type F
Supply Voltage
+15V to +21V
Power Requirements
400 mA Max at 15 to 21 volts



Model No.	Noise Temperature	Noise Figure (25°C)
360-1	120K	1.5 dB Max
360-2	100K	1.3 dB Max
360-3	90K	1.2 dB Max

Model No.	Noise Temperature	Noise Figure (60°C)
360-1	140K	1.7 dB Max
360-2	120K	1.5 dB Max
360-3	110K	1.4 dB Max

Conclusion

A low-noise block converter has been described. Its benefits for a particular type of receiving subsystem for video communications has been presented. Such systems have the advantages of eliminating microwave receiver front ends and reducing the duplication of circuits in a multiple receiver installation. Cabling the lower frequencies in from the antenna is also easier and less costly.

The Scientific-Atlanta Series 360 LNC and 6650 Video Receiver use a particularly low-frequency IF band to maximize the benefits of frequency conversion at the antenna. This choice of IF frequencies requires some important considerations in the design of circuits for the LNC. These considerations were discussed and the resulting circuits used in the LNC were described.

12 GHz Low-Noise Converter

K.M. Johnson

Introduction

Satellite communications at frequencies around 12 GHz are expected by many analysts to form a significant, if not dominant, portion of the TV reception market in the decades to come. This is especially true in direct satellite to home reception known as Direct Broadcast Satellite (DBS) reception.

An earth station designed to operate in the 12 GHz frequency band (usually designated Ku-band) consists of three parts: the Antenna, the Outdoor Unit and the Indoor Unit. The Outdoor Unit is generally visualized as some type of frequency converter mounted directly to the antenna which converts the 12 GHz frequency band to some lower frequency band. A 12 GHz LNA might be used directly at the antenna, but usually cable losses at 12 GHz are so great that it is more preferable to perform frequency downconversion as physically close to the antenna as possible. The Indoor Unit or Receiver then takes the downconverted signal, performs possibly a second conversion, signal demodulation, video and audio processing and remodulation before going to the user's TV set.

This article describes the Scientific-Atlanta 361 Series Low-Noise Converter (LNC) which is designed to interface with the low-cost 6650 Series Receiver currently being produced in large quantities at Scientific-Atlanta. This antenna-mounted LNC block converts the 11.7 to 12.2 GHz frequency band to an output band of 270 to 770 MHz. It has a gain of 56 dB and a noise figure of 3.1 dB.

Typical Specification

Besides providing gain and frequency conversion, it is necessary for the LNC to meet a number of specifications in order to give good TV reception and to properly interface with the receiver. Without going into detail at this point, the critical specifications for the LNC will be listed and described in Table 1.

Table 1. Low-Noise Converter Specifications

Characteristic	Specification
Input Frequency Range	11.7 to 12.2 GHz This, of course, is set by the satellite system.
Output Frequency Range	270 to 770 MHz The output range is set by the receiver being used. Generally this is chosen as low as possible in order to get low cable losses, to permit use of low-cost cabling and to minimize the amount of microwave circuitry in the receiver.
Gain	56 dB Nominal; 50 dB Min Gain must be sufficient to overcome any cable losses or noise contribution from the receiver. Gain is usually set within ±1 dB of the nominal 56 dB of gain.
Gain Flatness	±0.35 dB/40 MHz Gain flatness is required to prevent cross modulation between channels and to assure low signal distortion, especially when low C/N ratios are present. Specifying gain flatness over the entire 500 MHz band is not critical as long as the gain stays within the specified variation over temperature and within the specified incremental flatness of ±0.35 dB (40 MHz).
Gain Variation over Temperature	±3 dB This is specified in order to stay within the AGC range of the receiver considering also other factors such as cable loss.

Table 1. Low-Noise Converter Specifications - Continued

Characteristic	Specification
Power Output	+5 dBm Minimum at 1 dB gain compression. A sufficiently large power output at gain compression assures low distortion and intermodulation products at large input signal levels. This is particularly important at Ku-band where a large link margin is required due to large attenuation from rain fade. A large link margin means that larger signals are received by the LNC on clear days.
Input VSWR	1.25 Max A low input VSWR is inherent in an LNC which uses an isoadaptor.
Output VSWR	1.5 Max A low output VSWR assures that there will be no resonances and instabilities or signal level variations caused by reflections in the long cable run to the receiver.
Spurious Signals In-Band	Below noise level in a 100 kHz detection bandwidth.
Spurious Signals Out-of-Band	-20 dBm Max This specification is required to prevent possible intermodulation signals in the receiver. Normally out-of-band signals are filtered in the receiver.
Phase Noise of Converted CW Signal	In a 3,000 Hz bandwidth, the noise level below the carrier measured 250 kHz from the carrier will be -65 dBc maximum. This specification assures that the LNC will not degrade the signal-to-noise ratio of the system.

Table 1. Low-Noise Converter Specifications - Continued

Characteristic	Specification
Temperature	-40°C to +60°C typical LNCs are exposed to a variety of outdoor temperatures and must function within specification at all temperatures.

Other requirements in the LNC design include designing the LNC to be weatherproof, providing adequate mechanical strength for bolting directly to the antenna, and providing for supplying dc power to the LNC. DC power is supplied to the LNC through the RF output connector. The Scientific-Atlanta Series 361 LNCs meet all of the preceding specifications.

LNC Description

Figure 1 is a photograph of the Scientific-Atlanta Series 361 LNC showing the waveguide input to the LNC which bolts directly to the antenna feed. On top of the waveguide section is the isolator which isolates the transistor amplifier from any antenna mismatches and provides a transition from the waveguide transmission mode to the microstrip mode required for the 12 GHz low-noise amplifier in the LNC.

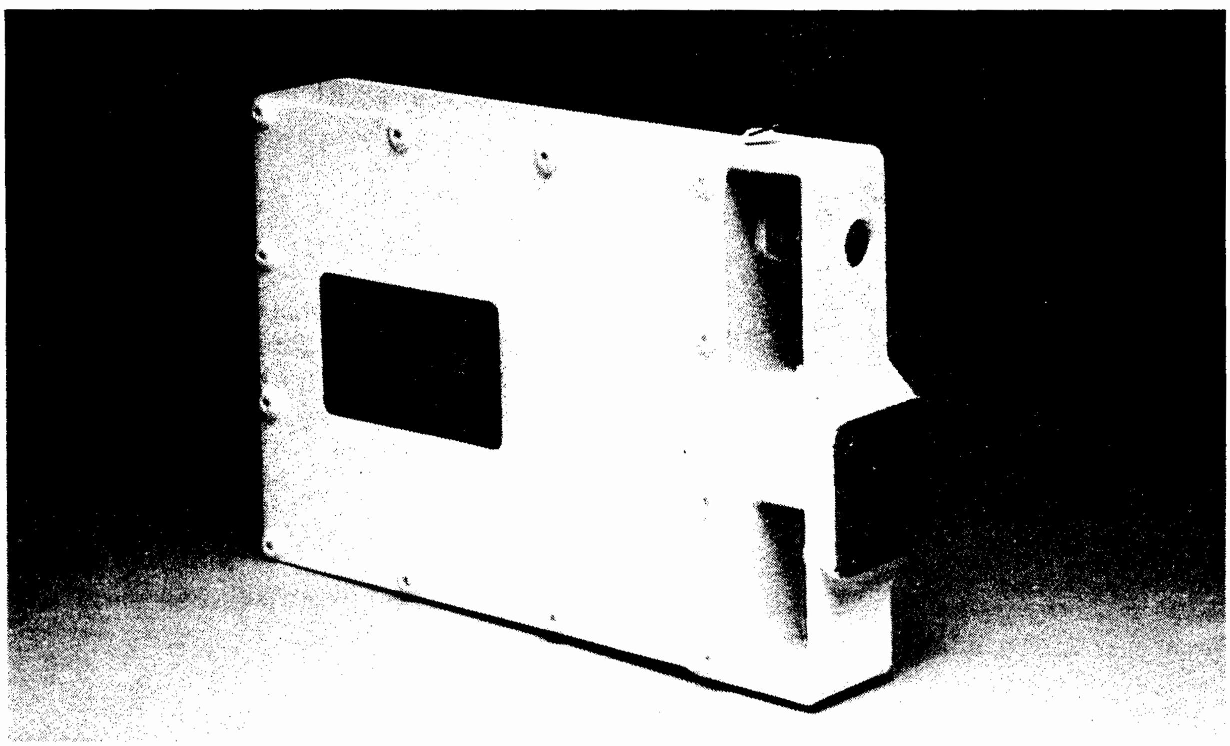


Figure 1. Series 361 Low-Noise Converter

The isolator is followed by the LNC housing and contains all the LNC amplifier, mixer and local oscillator circuits. Output from the LNC is through a coaxial "F" connector shown just above the isolator on the LNC. This connector is also where the dc power is fed to the LNC circuits. A support bracket is used to provide mechanical strength to the isolator and waveguide to coax adaptor which bolts directly to the antenna.

LNC Block Diagram

The circuit blocks which are in the 361 LNC are shown in the block diagram of Figure 2. Input to the LNC begins at the isoadaptor consisting of the waveguide-to-coax adaptor followed by a low-loss isolator. Isoadaptor loss is generally less than 0.24 dB which must be kept low since it adds directly to the amplifier noise figure. Following the isoadaptor is a three-stage GaAs FET amplifier. This amplifier has 24 dB of gain which is sufficient to permit the noise figure contribution of the following circuitry to add at most 0.26 dB to the overall LNC noise figure.

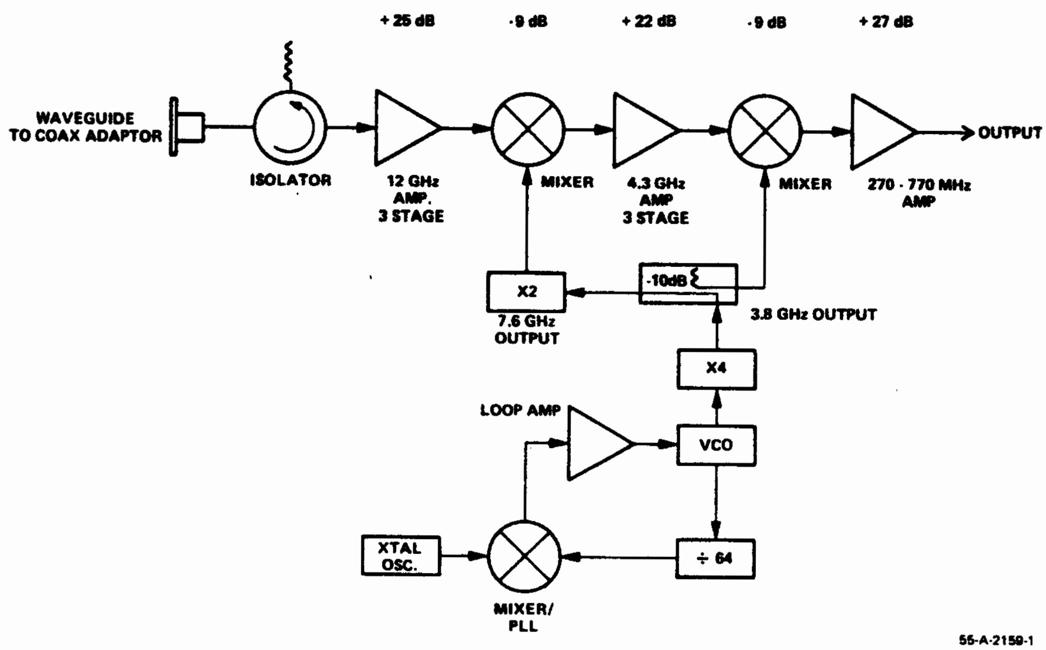


Figure 2. 12 GHz LNC Block Diagram

Following the amplifier is a mixer with an image filter as part of the mixer structure. This filter gives greater than 40 dB image rejection. The image filter-mixer conversion loss is 9.0 dB. LO frequency for this mixer is at 7.62 GHz. The IF output frequency band of the mixer is 4.08 to 4.58 GHz and is amplified in a three-stage low-noise amplifier having two FETs and one bipolar transistor. The IF amplifier provides 22 dB of gain and has a maximum noise figure of 5.0 dB.

The first IF amplifier is followed by a second mixer which is in turn followed by a second IF amplifier at the output frequency of 270 to 700 MHz. The second mixer is preceded by a filter to reject any unwanted out-of-band signals. The loss through the filter and mixer combination is 9.0 dB. LO frequency for this mixer is just half that of the first mixer and is at 3.81 GHz. The final amplifier, a feedback structure, has a gain of 27 dB over the 270 to 770 MHz band.

As may be seen from the block diagram, LO for the converter is generated by a VCO at 952.5 MHz which is phaselocked to a crystal at 14,8828 MHz. The VCO output is amplified to a level of 150 mW, then times-four multiplied in a step recovery diode multiplier to 3.81 GHz with an output power of 30 mW. Some of this output is coupled off to provide the LO to the second mixer while the rest is multiplied times two in a Shottky-Barrier diode multiplier to give the LO signal at 7.62 GHz for the first mixer.

Not shown in the block diagram is the bias circuitry and voltage regulator. DC voltage to the LNC is regulated and some of it is used to generate a negative voltage internally for biasing the FETs. Active bias circuits as well as thermister temperature compensation is used to get good stability over temperature from the LNC.

Description of Circuits

While conventional techniques are used in the design of most of the circuitry, it is worthwhile to describe some of the circuit techniques used especially in relation to critical specifications and unique techniques required to obtain good performance at Ku-band. For lowest manufacturing cost and ease of tunability, microstrip circuitry is used throughout. The type used, for the most part, is a single, low-dielectric constant board made of teflon fiberglass (TFG) material with a ground plane on one side and the transmission line circuitry on the other.

12 GHz Low-Noise Amplifier

As depicted in the block diagram, three amplifiers are used in the converter. The first amplifier, i.e., the 12 GHz LNA, was designed using conventional design techniques as described in the paper "Low-Noise Amplifiers." Briefly, after selecting appropriate transistors, a preliminary circuit is synthesized using device S-parameters. A computer optimization routine then optimizes selected circuit elements such as stubs, line lengths and capacitor values to get optimum gain and noise performance from the amplifier. The circuit is laid out using these optimized values, a circuit mask cut and the circuit fabricated. Some experimental "tweeking" is required to get the final circuit.

Some of the features of the 12 GHz LNA are shown pictorally in Figure 3 which shows a stage of the Ku-band amplifier as it is in the TFG board. The FET is actually mounted in a square hole in the board with source leads connected to ground through metal ribbons in the hole. Lines leading to and from the FET are generally 70 ohm lines to minimize radiation and junction effects. Bias is supplied through a transmission line low-pass filter structure consisting of quarter-wave stubs corresponding to capacitors to ground and quarter-wave, high-impedance lines corresponding to series inductors. In addition, 100 ohm series resistors are used for bias stabilization as shown. These techniques or slight variations are used all the way down to 2 GHz, although at lower frequencies a simpler low-pass bias structure is used.

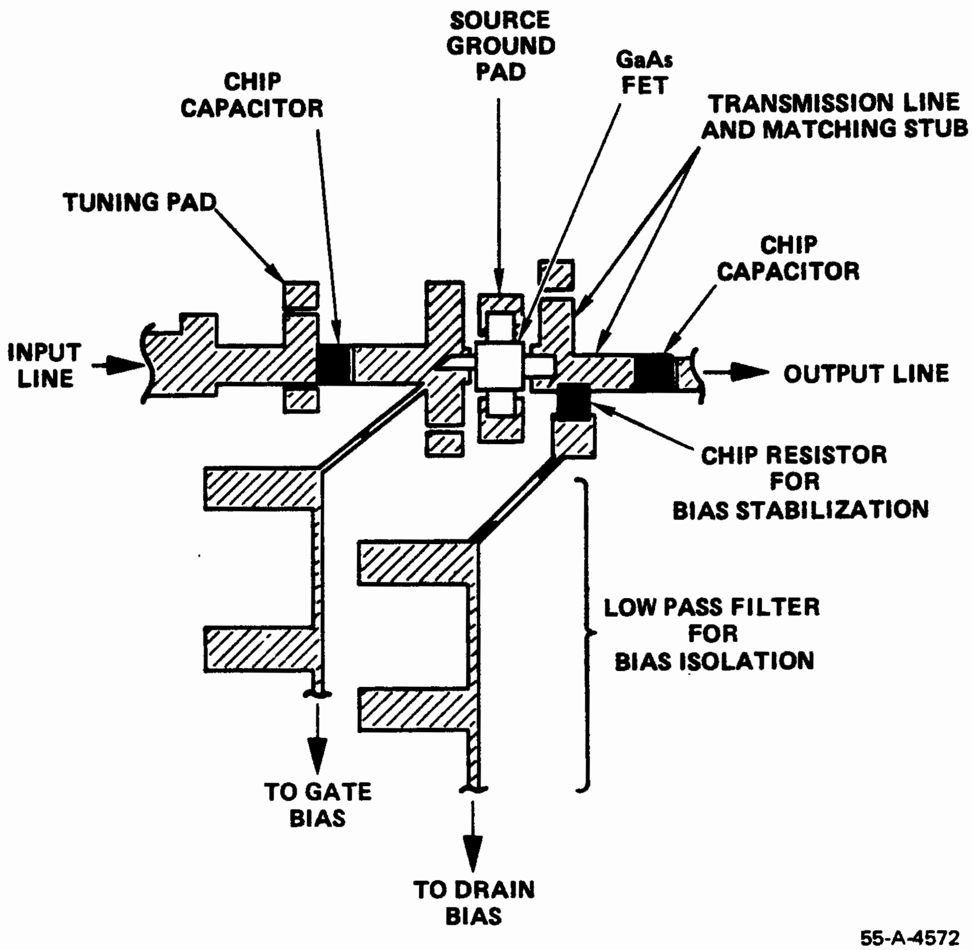


Figure 3. Portion of 12 GHz LNA Circuit

The noise figure of the 12 GHz LNA for the most part determines the overall noise figure of the LNC. This amplifier has three stages of gain with an overall gain of 25 dB. The first FET in the amplifier is selected to have the lowest noise figure. Figure 4 summarizes the noise figure calculations for this three-stage FET amplifier including isoadaptor losses. The first two transistors used have a typical noise figure of 2.0 dB. Over the 500 MHz band, the associated gain for the transistor is close to 8.3 dB per stage. F_M in the calculation is the 15 dB mixer/IF amplifier noise contribution. The result, as shown in the figure, is that the LNA would have a noise figure of 2.28 dB not including mixer/IF contributions and 2.53 dB including them. To this it is necessary to add the isoadaptor loss of 0.25 dB, the line loss and coupling loss leading up to the first FET of 0.20 dB, and the 0.10 dB degradation due to the transistor being specified at 12.0 GHz instead of 12.2 GHz. The result is that the LNC would have an overall noise figure of 3.09 dB. Of course, in some cases, somewhat lower noise figures would be obtained from slightly lower noise FETs.

Transistor Stages		Summary of Noise Figure Contributors	
$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 \; G_2} + \frac{F_M}{G_1 \; G_2 \; G_3}$		Transistor Stages	2.54 dB
		Isoadaptor Loss	0.25 dB
$F = 1.585 + 0.086 + 0.0216 + 0.10$		Coupling Loss	0.20 dB
$F = 1.7926$ (2.54 dB)		12.2 GHz Degradation	0.10 dB
			3.09 dB

Figure 4. Noise Figure Calculations for 12 GHz LNC

Performance for a typical three-stage FET amplifier is shown in Figure 5. Here over the 11.7 to 12.2 GHz frequency band a gain of 24.6 ±0.2 dB is measured with a noise figure maximum of 2.73 dB excluding the mixer/IF amplifier contribution but including the isoadaptor and coupling losses. This is close to the value predicted by the noise figure calculations.

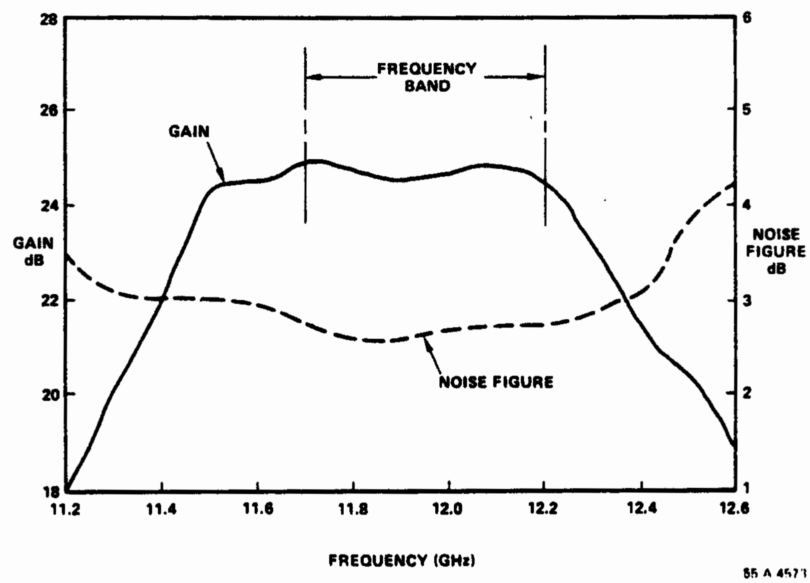


Figure 5. Gain and Noise Figure of Three-Stage LNA Used in Ku-Band LNC

12 GHz Mixer

The mixer for converting the 11.7 to 12.2 GHz band to the intermediate frequency band of 4.08 to 4.58 GHz is a fairly simple single diode mixer structure. The one unique feature of this mixer is that 12 GHz image filter and the LO filters are designed with a characteristic impedance of 80 ohms rather than 50 ohms. This is done since an edge coupled filter in TFG microstrip with a 50 ohm characteristic impedance would propogate higher order modes due to the wide line width. With the 80 ohm impedance, the line width is only 0.040 inches instead of the 0.084 inches for a 50 ohm line.

A sketch of this mixer structure along with conversion loss measurements is shown in Figure 6. The sketch shows the two edge-coupled transmission line filters for both signal band and LO frequencies as well as the low-pass filter for the IF band. Conversion loss for this mixer is less than 8.8 dB over the entire signal band including filter losses.

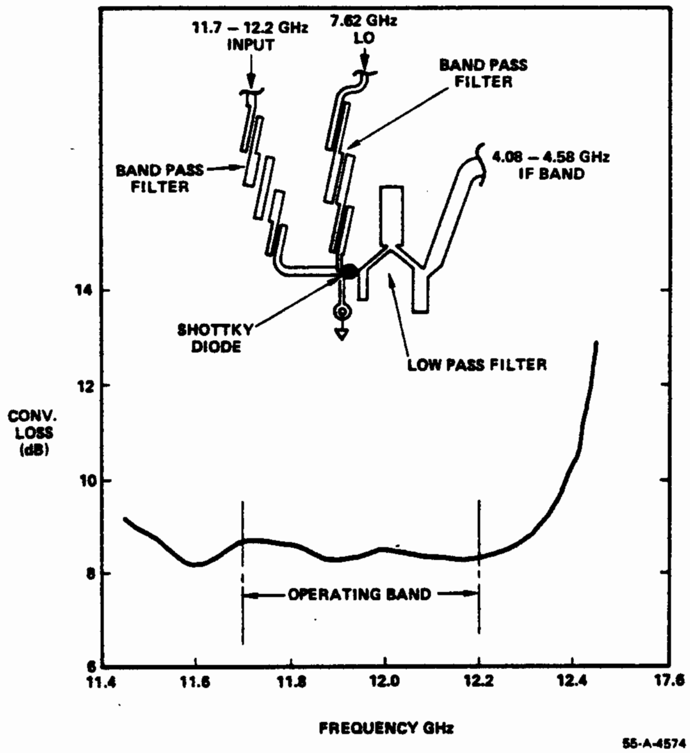


Figure 6. Series 361 LNC 12 GHz Mixer

IF Amplifiers

Two IF amplifiers are used in the 361 LNC. The first operating in the 4.08 to 4.58 GHZ frequency band has two FET stages and a bipolar transistor stage. This combination gives the desired low-noise figure performance. The design of this amplifier follows that described in the paper "Low-Noise Amplifiers," so it will not be repeated here. The amplifier has a 3 dB resistive pad in the output to provide a low output VSWR to the following mixer. The overall gain of the amplifier was 22 dB including the resistive pad.

The second IF amplifier operates over the entire 270 to 770 MHz range and has three bipolar transistors. To achieve a flat low-noise response over the entire band, it was necessary to use a feedback type of structure as shown in Figure 7. In this type of circuit, if the S21 (forward transmission S-parameter) is very large, the gain is essentially the ratio of the feedback resistor R_F to the source resistance of 50 ohm. This corresponds to a gain of 9.5 dB per stage or 28.5 dB for the three-stage amplifier. The IF amplifier actually measured 27 dB gain including a 3 dB because the source resistance for the second and third transistors is actually the output resistance of the preceding transistor. This impedance is lower than S22 (the output impedance S-parameter) since it is reduced by the feedback mechanism across the preceding transistor. Notice that a small inductance is used in series with the feedback resistor. This provides a reduced feedback and hence higher gain at high frequencies. The result is that the amplifier achieved a gain of 27.2 ±0.15 dB over the entire frequency band. Noise figure for the amplifier remained below 5.8 dB over the entire band.

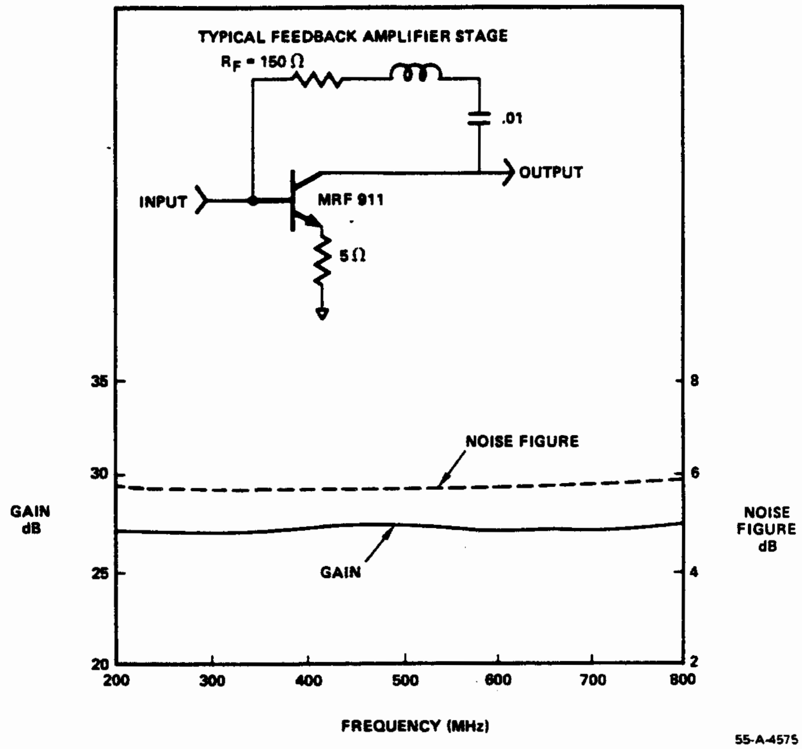


Figure 7. IF Amplifier Gain and Noise Figure vs. Frequency

Frequency Multipliers

LO for the LNC is generated by a VCO at 952.5 MHz. A portion of the VCO output is divided down to 14.8828 MHz by a divide-by-64 prescaler and phase locked to a crystal reference oscillator. The 952.5 MHz signal is amplified to a level of 150 to 200 mW and multiplied times 4 to 3,810 MHz to provide LO to the second mixer. A second multiplication times 2 is done to get the first LO signal at 7,620 MHz.

The times 2 multiplier consists simply of a printed low-pass filter on the input, a shottky barrier diode for resistive multiplication and a bandpass filter at the second harmonic. Again, to get good filter performance, this output filter has an impedance level of 80 ohms. A shottky diode was used rather than a step-recovery diode since it is presented a good resistive termination to the times 4 multiplier which drove it, and since no frequency breakup occurs in shottky doublers. Conversion loss for this multiplier was measured to be 11 dB including all filter losses. Fundamental, third and fourth harmonic signals were all attenuated at least 37 dB.

The times 4 multiplier, which uses a step recovery diode, is somewhat more complex since fundamental, second and fourth harmonics must all be resonated in the circuit. Circuit configuration for this multiplier is shown in Figure 8 along with a plot of its conversion loss versus frequency. This multiplier was fabricated on high dielectric constant substrate in order to reduce the size of the circuit. Input signals are filtered in the low-pass filter structure, and the output bandpass filter selects the fourth harmonic. A shorted stub, one-quarter wavelength long, is appropriately placed in conjunction with the location of the output filter so that fundamental, second and fourth harmonic frequencies are all resonated. The result as the figure shows is that the conversion loss to 3,810 MHz is only 6 dB.

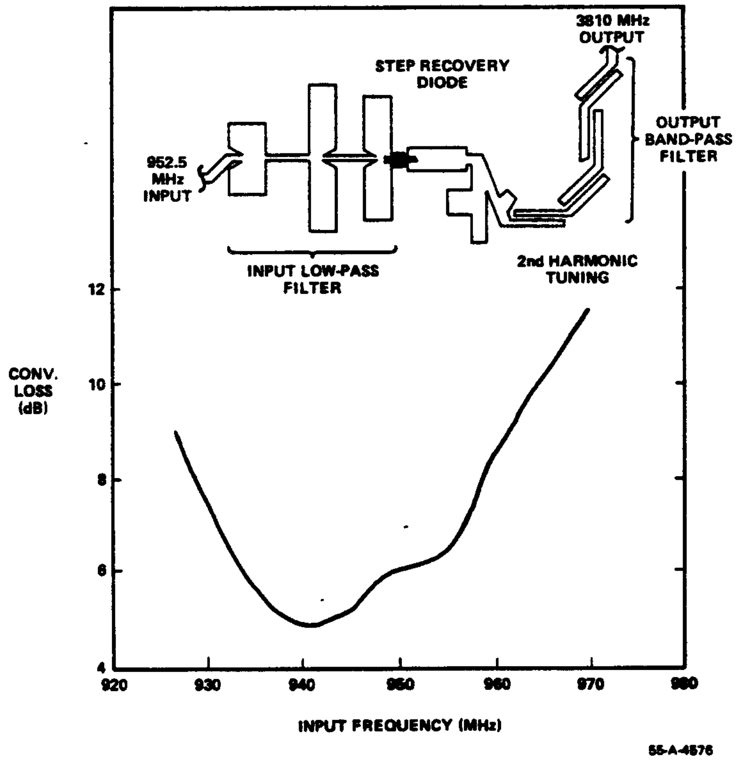


Figure 8. Times 4 Frequency Multiplier for Series 361 LNC

VCO Circuit and PLL

The VCO circuit, essentially a modified Colpitts oscillator, is shown in Figure 9. Feedback capacitors were selected using a computer program which computes the values necessary to obtain maximum output power. As mentioned before, a portion of the VCO output is prescaled in a divide-by-64 and phase locked to a 14.8878 MHz crystal reference oscillator. A similar phase-locked loop circuit is discussed in the paper "Low-Noise Converter," Series 360 by Rovira. The loop bandwidth of the PLL is only 10 kHz, so that the close-in phase noise is very low being essentially that of the crystal reference oscillator.

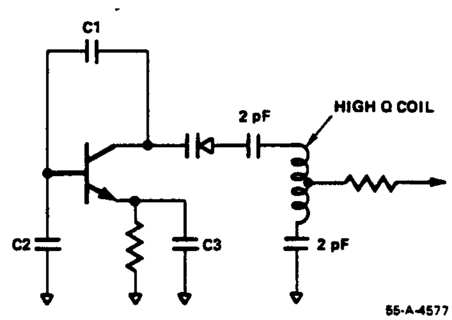


Figure 9. VCO Circuit Showing Circuit for Increasing Oscillator Q

Outside the 10 kHz locking bandwidth, the phase noise becomes that of the VCO. This noise is in turn increased by the frequency multiplication factor in the LNC which for the double conversion used increases the noise about 22 dB. For a typical free-running VCO, this phase noise would be excessive and would result in a somewhat noisy TV picture. A straightforward computation of the free-running oscillator noise can be made from the following equation:

$$\frac{N_{OUT}}{P_o} = \frac{N^2 kTF\Delta f}{P_o}\left[\left(\frac{f_o}{2Q_L f_m}\right)^2 + 1\right]\left[1 + \frac{f_\alpha}{f_m}\right]$$

where N_{OUT} is the noise out

P_o is the power out (0.010W for VCO used)

Δf is the video bandwidth (3,000 Hz in measurement)

N is the multiplication ratio (N = 12)

F is the transistor noise figure (F = 2.24)

f_o is the oscillator frequency (950 MHz)

QL is the loaded Q (This is computed from the oscillator computer program to be 1.46)

f_m is the frequency at which noise measurement is made (measured at 100 kHz)

f_α is the frequency at which the flicker noise equals the additive white noise ($f_\alpha \approx 3 \times 10^5$)

Using these values, the noise-to-power-out ratio computes to -47.7 dB. This is quite close to the measured value of -46 dB but is more than could be tolerated without reducing the system signal-to-noise ratio. In order to reduce the oscillator noise, the VCO Q was increased by adding two 2pf capacitors, in series with the high Q inductor as shown in Figure 9. The additional 1pf capacitance in the series circuit requires that a much larger inductor be used to obtain resonance and hence, larger Q results. The original circuit capacitance was 3.6pf. Adding the two pf series capacitors reduces this to an effective series resistance of 0.78pf. This resulted in a circuit Q of 6.7 instead of 1.46, hence, a noise improvement of 13.2 dB. In addition, by using a transistor with an output power of 30 mW instead of 10 mW, this gave an additional 5 dB of noise-to-signal power ratio. Resulting phase noise for the 361 LNC which uses the higher power, higher Q VCO is shown in Figure 10. Here, the noise level 100 kHz from the carrier is 65 dB as would be expected. This performance is more than adequate so that no degradation in picture quality results.

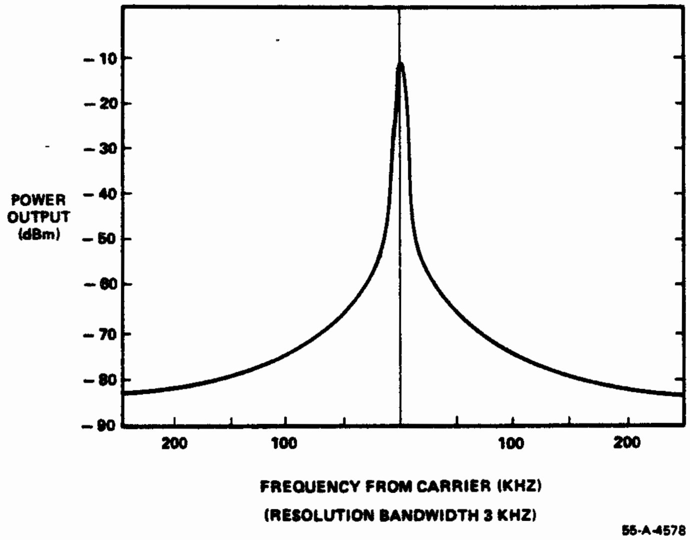


Figure 10. Series 361 LNC Spectrial Purity of Signal Converted from 11.76 GHz to 330 MHz

LNC Performance

Having discussed the specifications, design and circuit details of the LNC, characteristics of a typical Scientific-Atlanta LNC will be shown.

Gain Performance

Gain of a typical LNC, one taken directly off the production line, is shown in Figure 11. The gain versus fequency is plotted for three different ambient temperatures. At room temperature, the gain is flat within ±0.7 dB at a nominal gain of 56 dB. At +60°C, the gain remains flat within ±1 dB but the nominal gain falls about 1.5 dB. As the ambient temperature is reduced to -40°C, the gain begins to increase in value until it reaches a nominal maximum gain of 58 dB and the gain shows some tilt in its slope with the gain flatness of ±2.2 dB. At no point does incremental flatness exceed the specified ±0.35 dB/40 MHz. Since the primary concern of gain slope is at low C/N ratio, and it is at -40°C that the noise figure actually improves by almost 0.6 dB, no system degradation would be observed at -40°C.

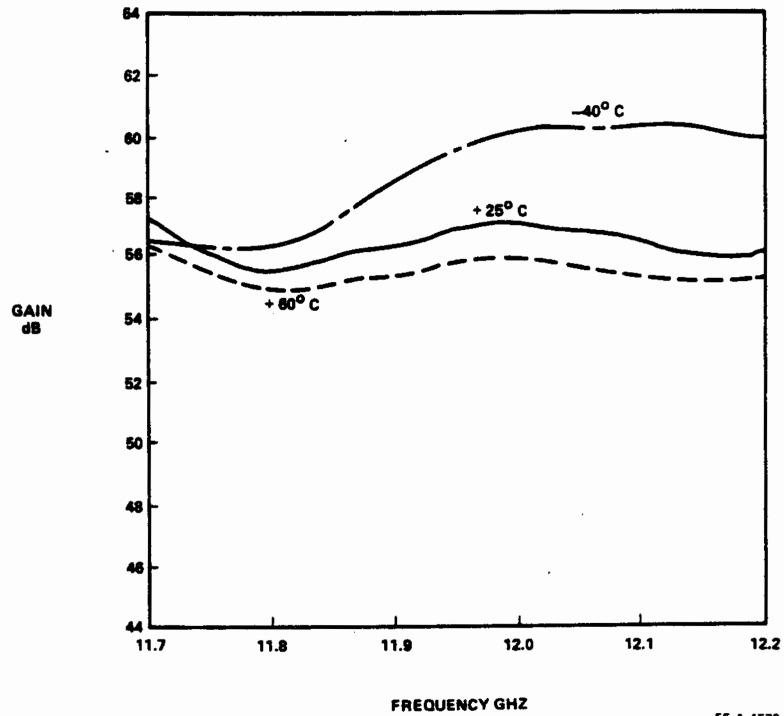


Figure 11. Series 361 LNC Gain vs. Frequency for Three Ambient Temperatures

Noise Figure Performance

Noise figure versus frequency performance for the same typical LNC is shown in Figure 12. This figure provides useful information for those needing to know the degradation in noise figure performance at high temperatures. As the figure shows, at 25°C the LNC has a noise figure of 3.1 dB. AT 60°C, it has increased to 3.5 dB. At the low ambient temperature of -40°C, the noise figure has reduced to a maximum value of 2.73 dB.

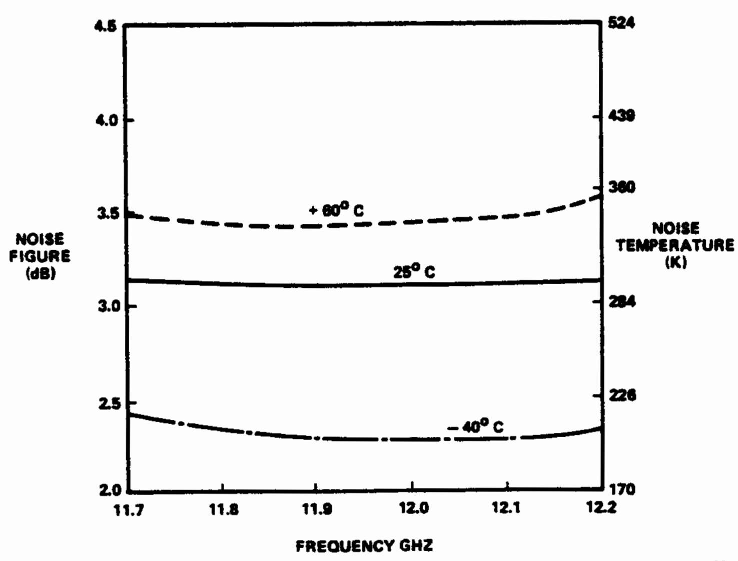
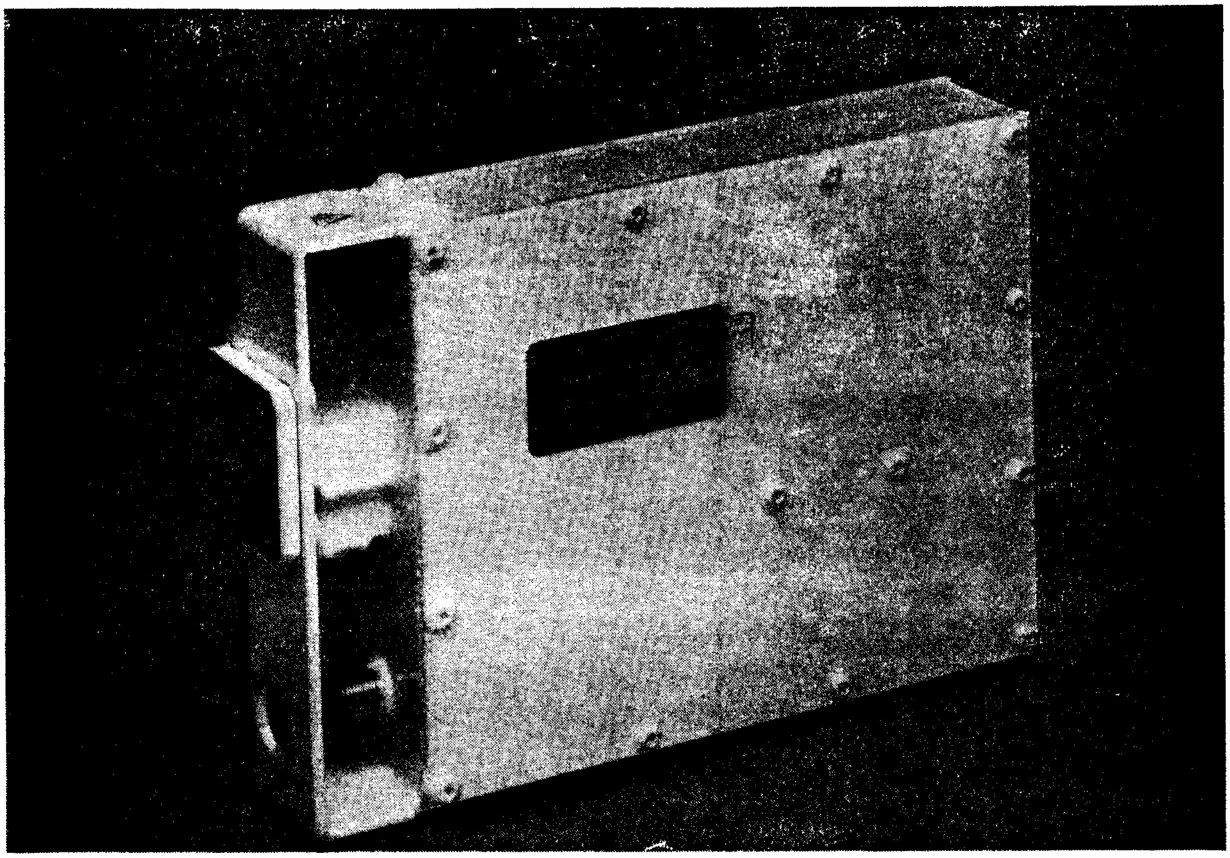


Figure 12. Series 361 LNC Noise Figure vs. Frequency for Three Ambient Temperatures

LNC Data Sheet

To summarize the performance of the Scientific-Atlanta Series 361 LNC, a typical data sheet is included with this paper. To date, Scientific-Atlanta has produced hundreds of such LNCs.

11.7 to 12.2 GHz Low Noise Converter
Series 361



The Series 361 Low Noise Converter (LNC) is designed for use with the Model 6651 video receiver to provide low cost, high performance satellite television reception. The LNC is a combination of a low noise GaAs FET amplifier and a block downconverter for optimum noise performance. The LNC converts the entire 500 MHz satellite band to UHF frequencies between 270-770 MHz at the antenna feed. This eliminates the need for microwave components in the receiver as well as expensive coaxial cable from the antenna to the headend. The result is a low-cost earth station electronics subsystem.

The LNC utilizes internal output isolation and an integral isolator to protect against antenna mismatch. A precision cast housing contains a low noise amplifier, microwave oscillator, microwave filters and power supply regulator. Protection is provided to withstand exposure to all weather conditions. Standard units are powered through the center conductor of the coaxial cable. The Series 361 LNC is available in 440K, 360K and 300K noise temperature ranges.

Specifications

Frequency Range
11.7-12.2 GHz
Input Level
−75 dBm to −95 dBm per channel
Input Impedance
50 ohms
Input Return Loss
20 dB min.
Conversion Gain
56 dB +5 dB, −3 dB
Image Rejection
50 dB min.
IF Frequency Band
270-770 MHz
IF Output Impedance
75 ohms
IF Return Loss
17 dB min.
Temperature
−40°C to +60°C full specification
Input Connector
WR75 flanged
Output Connector
Type F
Supply Voltage
+15V to +21V
Power Requirements
400 mA max. at 15 to 21 volts

Model No.	Input Frequency	Output Frequency	Noise Temperature	Noise Figure (25°C)
361-1	11.7-12.2 GHz	270-770 MHz	440K	4.0 dB Max
361-2	11.7-12.2 GHz	270-770 MHz	360K	3.5 dB Max
361-3	11.7-12.2 GHz	270-770 MHz	300K	3.1 dB Max
361-4	11.62-11.8 GHz	440-620 MHz	440K	4.0 dB Max
361-5	11.62-11.8 GHz	440-620 MHz	360K	3.5 dB Max
361-6	11.62-11.8 GHz	440-620 MHz	300K	3.1 dB Max

(All specifications subject to change without notice.)

The Model 6650 Video Receiver

Kenneth C. Wunner

Introduction

The increased use of satellite earth stations by cable operators, broadcasters, hotel/motel operators, and many others, coupled with an increase in the number of subcarrier services per transponder, has created a need for a lower cost, more compact and better performing receiver system. The 6650 Video Receiver was designed to meet this need.

The major design objectives in developing the 6650 Receiver were fourfold. First, the receiver was to take advantage of block downconversion at the antenna. This would eliminate costly pressurized heliax runs and microwave components from the unit. Second, two receivers were to fit into a standard 19-inch rack, thus reducing the amount of space needed for mounting the units. Third, the video demodulator must perform well in the presence of multiple subcarrier signals. This would decrease the problem of reduced threshold performance on multiple subcarrier transponders. Fourth, the receiver must be of a high enough quality and reliability to meet all the demands placed upon it by Scientific-Atlanta's customers.

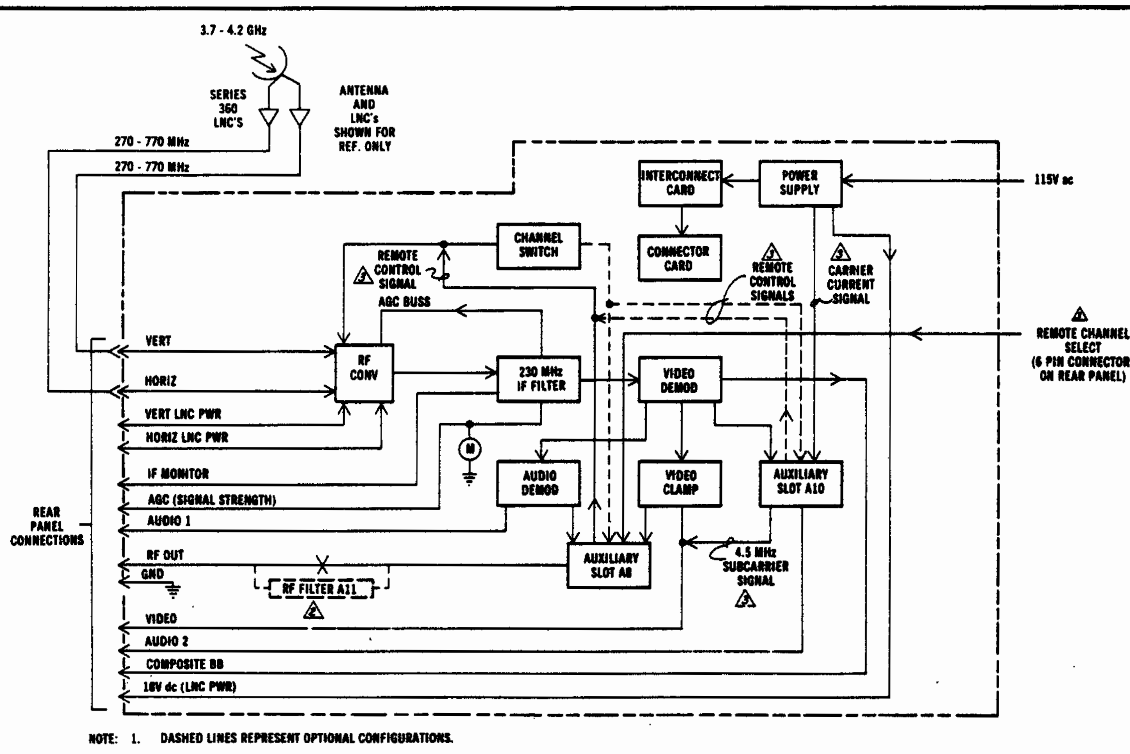
The 6650 Video Receiver (Figures 1 and 2) along with all of its options meets and/or exceeds these four objectives. The remainder of this paper will discuss the receiver in detail.

Figure 1. Model 6650 Video Receiver

General Description

A block diagram of the 6650 Video Receiver connected to two Series 360 Low-Noise Converters (LNCs) is shown in Figure 5. The 270- to 770-MHz signals from the LNCs enter the 6650's RF Converter, where first the proper polarization for the desired channel is selected. Next, the desired signal is converted to an IF frequency of 230 MHz. The signal then passes through the IF filter, where amplification, filtering and AGC is performed. The signal, now ready for demodulation, is converted into a baseband spectrum of 1 Hz to 10 MHz by the video demodulator. The baseband signal is passed to four points:

- The video clamp which filters and restores the dc level of the video contained within the baseband signal.
- The audio demodulator which selects the desired subcarrier and demodulates the audio present on it. This audio signal is available on the rear panel as AUDIO 1.
- Auxiliary slot A10, which may be used for an additional audio demodulator.
- The rear panel, for monitoring purposes or for connection to other equipment (i.e., stereo decoders, data demodulators, etc.).



NOTE: 1. DASHED LINES REPRESENT OPTIONAL CONFIGURATIONS.

2. A11 USED ONLY WHEN OPTIONAL TV MODULATOR IS USED IN A8.

3. OPTIONAL MODULES ARE NEEDED FOR THESE FUNCTIONS.

Figure 5. Model 6650 Video Receiver Block Diagram

Block Downconversion

The RF signals in the frequency band of 3.7 to 4.2 GHz are received by the antenna and routed to one of two Series 360 LNCs through an orthogonal mode transducer (OMT). The OMT separates the vertically and horizontally polarized signals into two 500-MHz bandwidth segments, both at 3.7 to 4.2 GHz. The LNCs take each of these segments, amplify and block-downconvert each of them to two 500-MHz bandwidth signals, now in the 270- to 770-MHz band (see Table 1). From here each signal is sent via its own coaxial cable to the RF inputs of the 6650 Receiver.

The 270- to 770-MHz band was chosen for several reasons:

- Lower-cost cable may be used to connect the 6650 Receiver to the LNC. RG-59 or RG-6 is recommended for most applictions.

- Low-cost UHF power splitters may be used to distribute the signals to multiple receivers. These are readily available and cost much less than conventional microwave power splitters.

- Convenient and easily obtainable F-type connectors are used. F-type connectors are lower cost than the conventional N-type needed for microwave signals.

- Microwave components are eliminated from the receiver. This allows the use of less complicated and more reliable circuitry in the RF converter part of the 6650 Receiver.

RF Conversion

The RF converter module (Figure 6) consists of five PWBs which provide the necessary signal processing to convert the dual 270- to 770-MHz input from the LNCs to a single-channel 230-MHz IF at a fixed -35 dBm level.

The Voltage-Controlled Oscillator (VCO) and synthesizer boards contain a varactor-tuned local oscillator (LO) running at 520 to 980 MHz, phase-locked to a 10-MHz crystal source. The six input code lines from the front-panel switch, or from the remote control card in slot A8, program frequency dividers to select the LO frequency. This type of LO is highly stable; therefore, fine tuning is not required for proper operation.

Table 1. Satellite Transponder Frequencies

Single Polarization Satellite Transponder Number (Note 1)	Dual Polarization Satellite Transponder Number (Note 2)	Center Frequency (MHz)	Block Converted Frequency (MHz)
1	1 (V)	3720	290
	2 (H)	3740	310
2	3 (V)	3760	330
	4 (H)	3780	350
3	5 (V)	3800	370
	6 (H)	3820	390
4	7 (V)	3840	410
	8 (H)	3860	430
5	9 (V)	3880	450
	10 (H)	3900	470
6	11 (V)	3920	490
	12 (H)	3940	510
7	13 (V)	3960	530
	14 (H)	3980	550
8	15 (V)	4000	570
	16 (H)	4020	590
9	17 (V)	4040	610
	18 (H)	4060	630
10	19 (V)	4080	650
	20 (H)	4100	670
11	21 (V)	4120	690
	22 (H)	4140	710
12	23 (V)	4160	730
	24 (H)	4180	750

NOTES:

1. Transponder assignments for 12-channel satellites are all horizontally polarized.
2. Transponder assignments for 24-channel satellites are horizontally polarized for the even-numbered transponders and vertically polarized for the odd-numbered transponders.

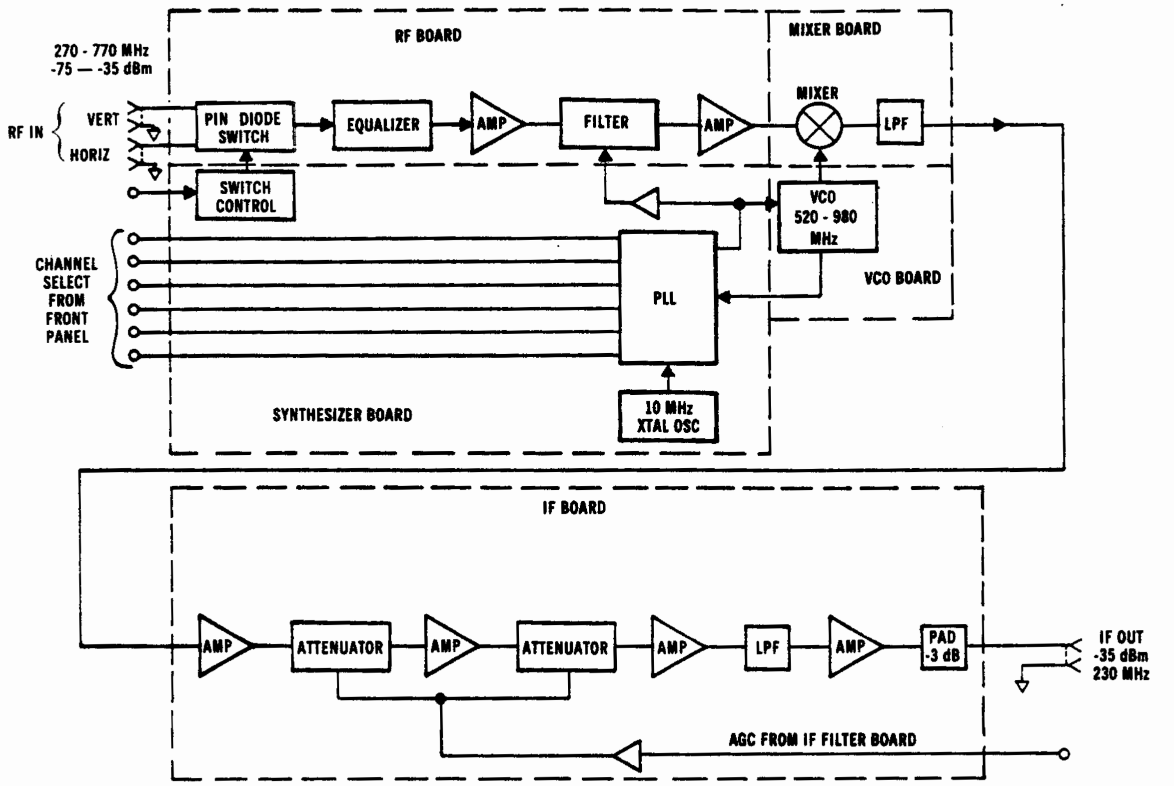


Figure 6. RF Converter Block Diagram

The RF board selects between the vertical and horizontal inputs from the LNC with a PIN diode switch. The signal then passes through an amplitude equalizer with 7 dB of slope to equalize for cable loss. It is then amplified and filtered to remove image frequencies and LO radiation.

The RF signal from the RF board and the LO signal from the VCO board are applied to a mixer on the mixer board. Low-pass filtering of the resultant 230-MHz IF signal is used to attenuate any LO signal leakage through the mixer.

The IF board amplifies, filters, and attenuates the 230-MHz IF signal. Amplifier stages provide 40 dB of gain, and PIN diode attenuator stages provide 0 to 40 dB of attenuation for automatic gain control (AGC). The AGC control signal for the attenuator comes from the IF filter board.

IF Filtering and Amplification

The IF filter board (Figure 7) contains a bandpass filter centered at 230 MHz with an effective noise bandwidth of 30 MHz, a delay equalizer and the AGC detector and loop filter. The IF filter consists of a front fourth-order bandpass filter section, delay equalizer section, and a rear fourth-order bandpass filter section isolated by fixed attenuators and buffer IF amplifier stages.

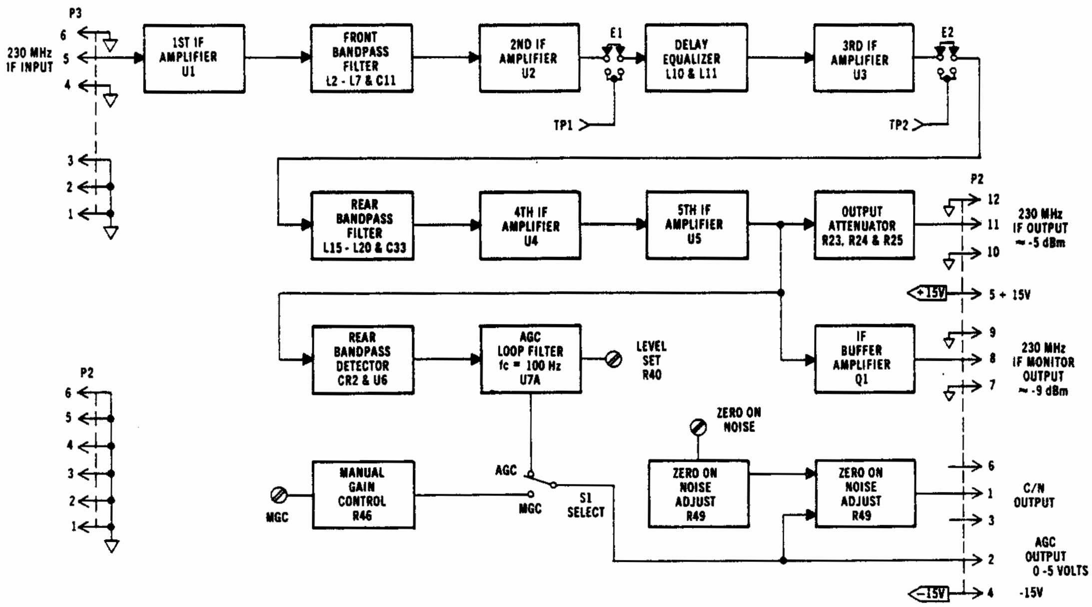


Figure 7. IF Filter Block Diagram

The two bandpass filter sections were derived from fourth-order Cauer-Chebyshev low-pass prototypes. The designs are such that they exhibit an amazing degree of arithmetic symmetry in both amplitude and group delay about the 230-MHz center frequency. The front bandpass filter section has transmission notches at 204 MHz and 256 MHz. The rear bandpass filter section has transmission notches at 210 MHz and 250 MHz which help to eliminate adjacent channel cross-polarized interference.

This delay equalizer is formed by a single second-order network. This section compensates for some of the non-flat group delay produced by the bandpass filter sections.

The AGC detector and loop filter are formed by a diode detector and op-amp integrator. AGC level for the 230-MHz signal is -5 dBm ±1 dB for an input dynamic range of 40 dB. The AGC signal from the loop filter is buffered and sent to three places: the RF converter, the front-panel meter, and the AGC terminal on the rear of the receiver. The AGC voltage is an indication of the received signal strength and can be used for proper pointing of the antenna and, with proper calibration, to measure approximate received signal C/N.

Demodulation

The video demodulator board (Figure 8) contains a phase-locked loop oscillator, video amplifier and deemphasis circuitry. The video demodulator receives a frequency-modulated 230-MHz IF signal from the IF filter card. This signal is used to phase-lock an internal voltage-controlled oscillator. The voltage required to hold lock to the FM signal is proportional to the instantaneous frequency of the input signal. This voltage thus represents the modulation on that signal.

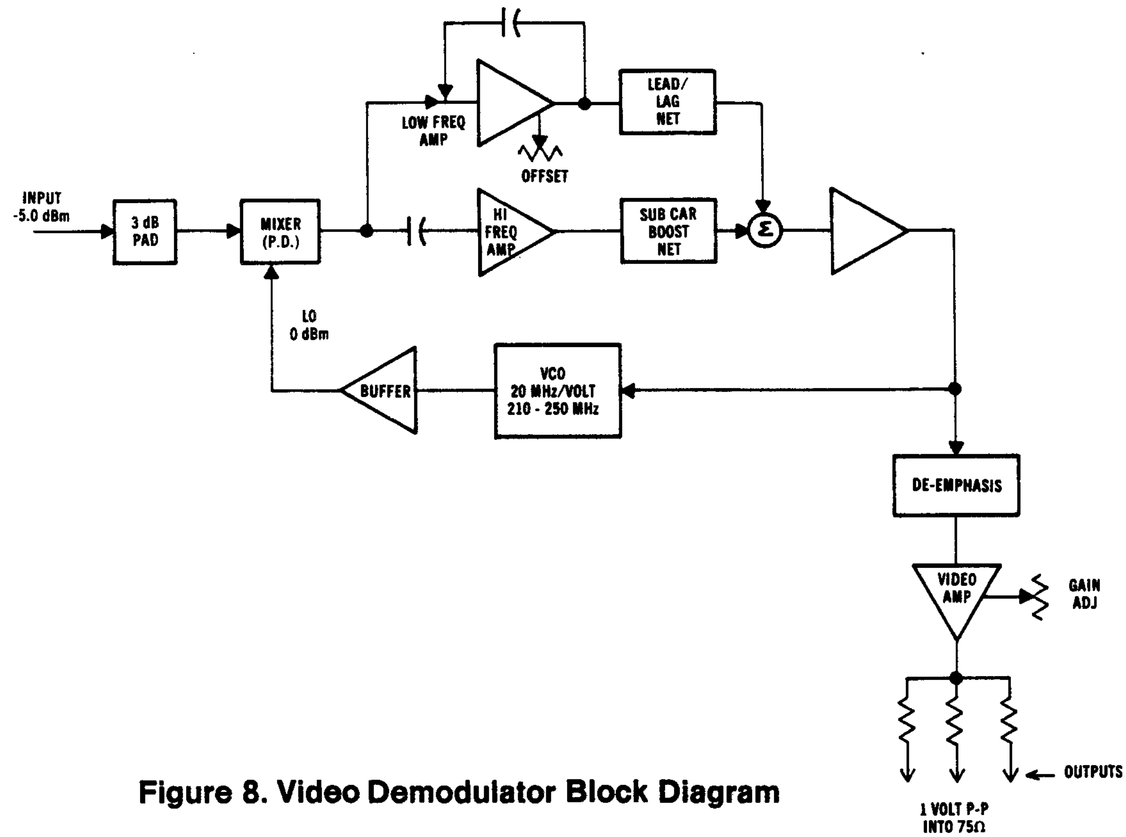


Figure 8. Video Demodulator Block Diagram

The voltage-controlled oscillator's output drives a balanced mixer. The second input to the mixer is the IF signal. The output from the mixer is therefore proportional to the phase difference between the input IF signal and the VCO signal. This output signal is split to drive an ac-coupled amplifier and a dc-coupled amplifer. The ac-coupled amplifier has two stages connected by a compensation network. This coupling network serves to boost the gain of the phase-locked servo loop, thus improving the tracking performance to the audio subcarrier and other subcarriers which are be present in the composite signal. The dc-coupled amplifier is configured as an integrating amplifier to provide high gain down to dc. A lead-lag network couples the dc path into the last stage of the ac amplifier. The loop is then completed by connecting the loop amplifiers to the VCO.

Output from the phase-locked loop is passed through a deemphasis network and then to a video amplifier to boost the level to 1 volt for a deviation of 10.75 MHz. The composite baseband signal is outputted to three different points. One output goes to the video clamp. A second output goes to the program audio demodulator, and a third output goes to a rear-panel connector.

Video Processing

International agreement of satellite transmissions requires energy dispersal modulation to prevent concentration of energy at one frequency. This waveform is triangular, with apexes located at the vertical intervals. This triangular waveform must be removed after demodulation by a video clamping circuit.

The video clamp board (Figure 9) contains an equalized low-pass filter, video clamp and output amplifier and driver.

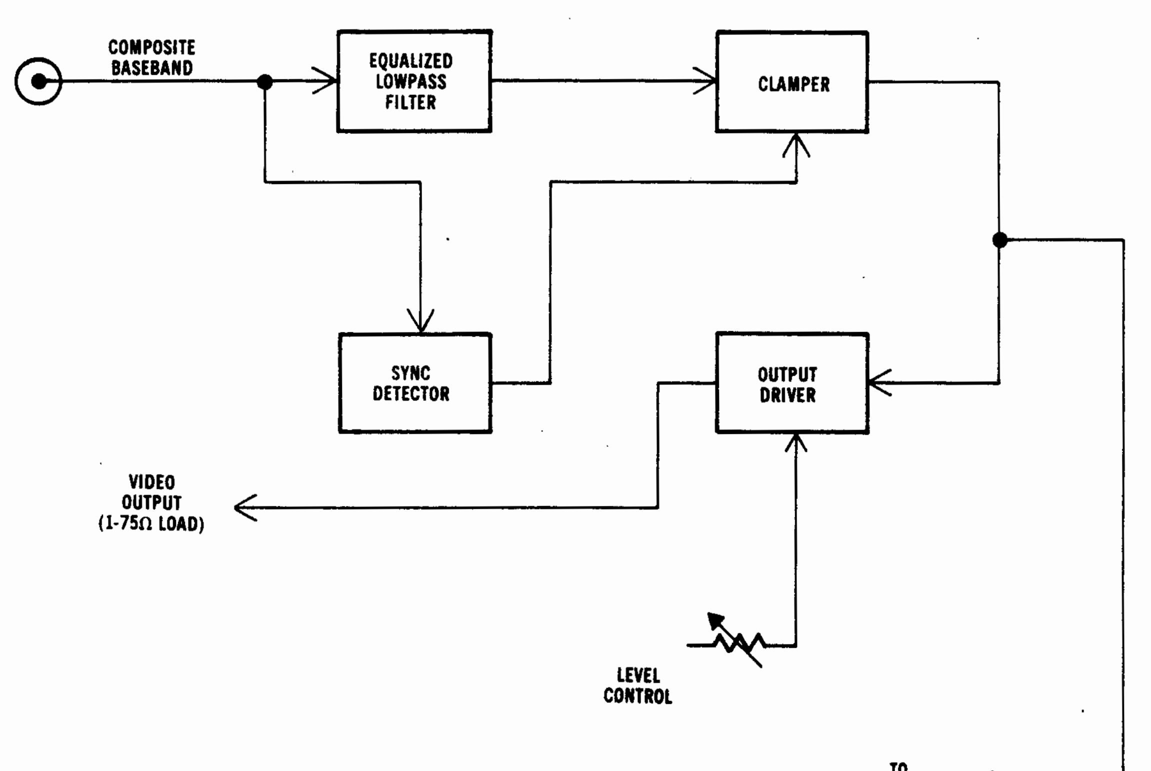


Figure 9. Video Clamp Board Block Diagram

The signal received from the video demodulator is split into two paths; one path goes through a 4.5-MHz equalized low-pass filter which removes the audio subcarriers from it; the other path goes to a sync detector which generates a signal corresponding to the video sync pulse. The sync detector commands the clamp to clamp the sync tips of the now filtered video signal to ground, thus removing the dispersal waveform.

From the clamper, the video signal is amplified by a combination dc-coupled negative feedback amplifier and line driver. One 75-ohm video output is provided on the rear panel. Level at the output is adjustable by the front-panel VIDEO LEVEL control.

Audio Demodulation

The program audio is usually located on a 6.8-MHz subcarrier. Some satellites use a subcarrier at 6.2 MHz. Presently, the 6650 Receiver can demodulate 6.2- or 6.8-MHz subcarriers by placing a card for the particular frequency into the audio demodulator slot.

Audio baseband bandwidths of 15 kHz are normal for satellite transmission. The 6650 Receiver can demodulate this bandwidth fully with less than one percent total harmonic distortion and with a frequency response of better than or equal to ±1 dB.

The audio demodulation card (Figure 10) contains a subcarrier select filter, FM demodulator, 20-kHz low-pass audio filter, 75 μs deemphasis circuit, and a balanced 600-ohm line driver.

The composite signal from the video demodulator is passed through the subcarrier select filter which selects the desired subcarrier and rejects all other signals. The filter is a 400-kHz-wide inductively coupled bandpass type. From the filter, the desired subcarrier is demodulated by a limiter/quadrature detector integrated circuit, and baseband audio is obtained.

The audio is passed through a 20-kHz low-pass active filter which eliminates any unwanted ultrasonic components present. The now "clean" audio signal is deemphasized and amplified. A balanced 600-ohm output is provided on the rear panel. Its level can be controlled from the front-panel AUDIO LEVEL control.

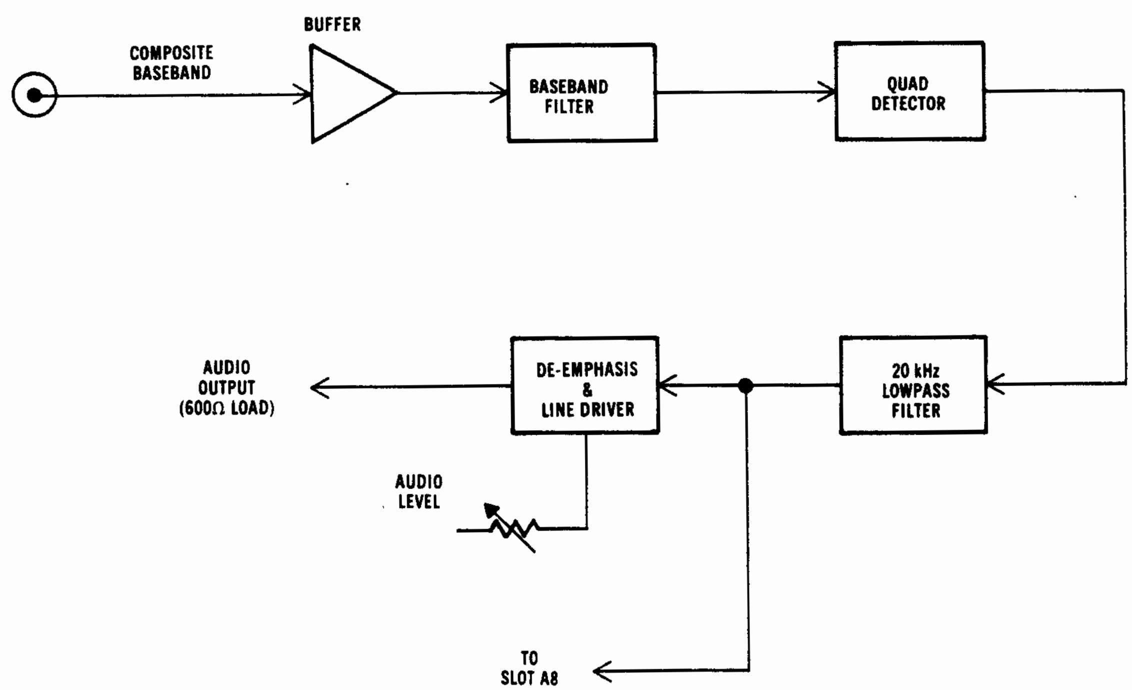


Figure 10. Audio Demodulator Board Block Diagram

Auxiliary Slot A10

This slot is provided for the addition of an auxiliary subcarrier demodulator, or a 4.5-MHz subcarrier modulator.

The auxiliary audio from the demodulator is available on the rear panel through the AUDIO 2 terminals. All functions are the same as for the program audio demodulator, except the output level is now controlled from an on-card level control.

The subcarrier modulator produces a 4.5-MHz FM signal modulated with program audio. The peak deviation is 25 kHz, and 75 µsec preemphasis is used. The 4.5 MHz is summed with the video and is available at the VIDEO OUT connection on rear panel.

The composite 4.5-MHz signal can be used in applications where the receiver is located a distance from the head-end. It allows audio and video to be easily sent via terrestrial microwave link from the earth station site to the head-end for distribution.

Auxiliary Slot A8

This slot is provided for the use of either a remote control interface card, or a Class I TV modulator.

The interface card allows the user to remotely control the channel selection of the receiver via the remote control inputs on the rear panel. When in the remote mode, the REMOTE light on the front panel will come on. If the channel select switch is rotated until the REMOTE FREQUENCY light comes on, the channel indicated by the switch is the channel selected by the remote control signal.

All 24 channels can be remotely selected, or six channels can be selected by "cherry-picking" (grounding individual pins), depending on how the interface card is programmed. Programming is accomplished by setting the select switches on the card.

The TV modulator card is a cost-effective device designed for MATV or monitoring applications. It is not intended for cable applications. It produces an RF signal on one of two low band channel (3-4) of approximately -46 dBm, modulated with both audio and video. A separate RF OUT connection is provided on the rear panel for obtaining the modulator signal. This output can be connected directly to a TV receiver. The design meets all the FCC requirements for a Class I TV device.

Power Supply

The power supply is a conventional 3-terminal regulator type which plugs into a 115V ac ±15V ac outlet, and provides ±15V dc and +18V dc to the receiver. The ±15V dc is used to power all the modules inside the 6650 Receiver. The +18V dc is used to power a Series 360 LNC, either by a separate power cord or via the coaxial cable connecting it to the receiver. Both dc supplies have current foldback and overvoltage protection.

Specifications

Table 2 is a listing of the Model 6650 Video Receiver technical specifications. As compared to most link specifications, it can be seen that the 6650 Receiver is relatively transparent.

Summary

The 6650 Video Receiver is the latest in Scientific-Atlanta's line of satellite receivers. It has been designed to be a low-cost, high performance, highly versatile device useful in a wide variety of earth station applications.

Table 2. Model 6650 Video Receiver Technical Specifications

Characteristic	Specification
RF Input	
Level	-75 dBm to -35 dBm
Frequency	270 MHz to 770 MHz
Impedance	75 ohms
Return Loss	>14 dB
Noise Figure	<12 dB (at CH24, with AGC attenuator set for minimum loss)
Image Rejection	>45 dB
IF	
Intermediate Frequency	230 MHz
Effective Noise Bandwidth	30.0 MHz
Impedance	75 ohms
Return Loss at IF Monitor Part	>15 dB
Dynamic Operating Range	40 dB
Video	
Baseband	Deemphasis 525-line CCIR Rec. 405-1
Deviation Range	6 to 12 MHz peak at deemphasis crossover frequency
Video Level	1V peak-to-peak ±3 dB, adjustable
Response (15 Hz to 4.2 MHz)	±1.0 dB
Impedance	75 ohms, unbalanced
Return Loss	>26 dB
Polarity	Black-to-white: positive going
Clamping	30 dB dispersal rejection
Line-Time Waveform Distortion	<1% tilt
Field-Time Waveform Distortion	<1% tilt
Differential Phase	<±1.5° 10% to 90% APL
Differential Gain	<±3% 10% to 90% APL

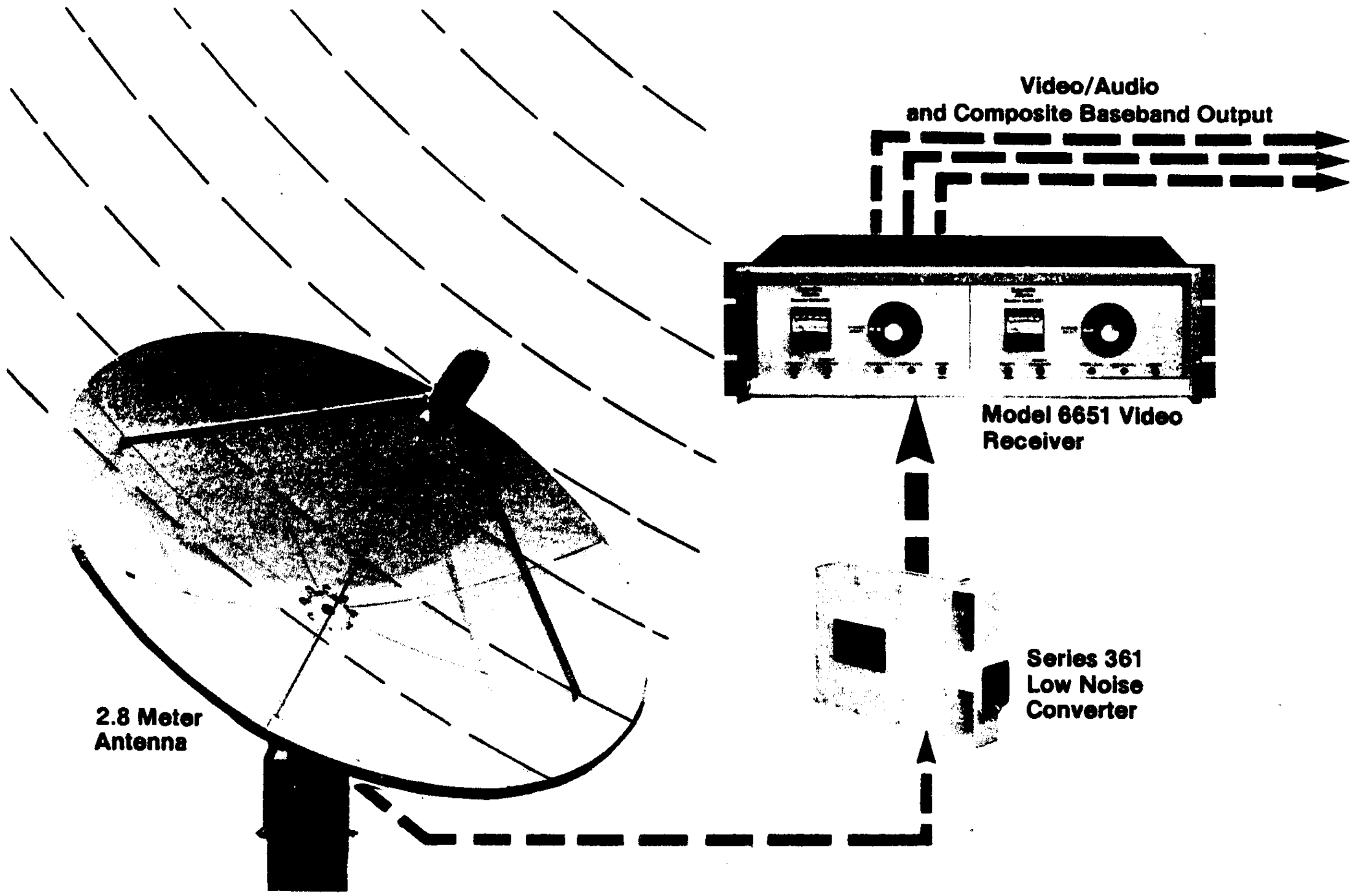
Table 2. Model 6650 Video Receiver Technical Specifications — continued

Characteristic	Specification
Audio	
Subcarrier Frequency	6.8-MHz standard, other frequencies available
Frequency Response	30 Hz to 15 kHz ±1.0 dB
Deemphasis	75 μs
Output Level	Continuously variable, -10 to +10 dBm
Impedance	600 ohms, balanced
Harmonic Distortion	<1%
Controls	
Rear Panel Power	On/Off
Front Panel	
Video Level	Adjusts video output level to 1V peak-to-peak ±3 dB
Audio Level	Adjusts audio output level -10 to +10 dBm
Local Frequency Switch	In 20-MHz increments, 24 channels
Top Panel	
AGC/MGC	Selects automatic or manual gain control of IF filter/amplifier
MGC	Adjusts gain of IF filter/amplifier
Zero-on-Noise	Allows calibration of meter for C/N measurements
General	
Operating Temperature	0°C to +50°C (32°F to 122°F)
Mechanical	
Height	133.4 mm (5.25 inches)
Width	209.8 mm (8.26 inches)
Depth	344.4 mm (13.56 inches), excluding front-panel knob
Power Requirements	75 watts Max at 100 to 130V ac, 60 Hz

Options

6.2 MHz Subcarrier Demodulator
6.8 MHz Subcarrier Demodulator
Interface Logic board remote control
Class I TV Modulator
Rack Adapter (Figure 2)
4.5 MHz Audio Modulator

Ku-Band Receive-Only System



Scientific-Atlanta offers earth stations to receive satellite television transmissions from 11.7 to 12.2 GHz.

A typical Ku-band video receive terminal consists of the following:

- A Series 9000 Ku-band earth station antenna, with elevation-over-azimuth mount and single or dual polarized feed
- A Series 361 low noise converter
- A Model 6651 video receiver, with ANIK-C specifications (similar configurations for SBS, OTS, and other satellites are available)

Each system also includes 100 feet of coaxial cable to connect the low noise converter (LNC) to the video receiver, and installation and operation instructions. The Model 6651 receiver and Series 631 LNC specifications are listed on the following pages.

Message Receiver Model 7510

T.C. Mock

Introduction

The Scientific-Atlanta Model 7510 Message Receiver is designed to fit into any FDM/FM configuration, either domestic or international. The Model 7510 offers unmatched flexibility and ease of reconfiguration. In addition, the 7510 uses the SAbus for remote control and status monitoring, allowing it to be used with a variety of controllers and protection switches. Finally, the microprocessor controller internal to the 7510 facilitates self test, local monitoring, and local control of the receiver.

General Description

The 7510 Receiver is modular in design (Figure 1)--all modules (IF filter, AGC amplifier, demodulator, baseband processor, out-of-band noise monitor) plug in from the front of the receiver, behind a drop-down front panel. In addition, the downconverter is also easily removable from the front of the receiver.

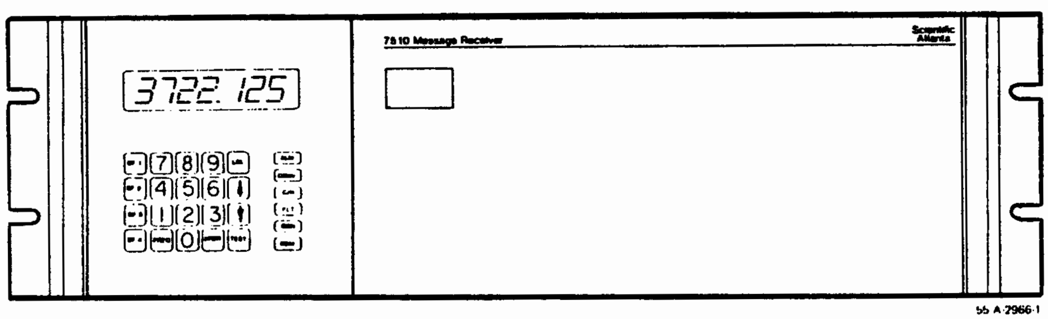


Figure 1. Front-Panel Drawing of the Model 7510 Message Receiver

Monitor and control of the 7510 is accomplished via a front-panel keyboard and display. Frequency, failed module, carrier-to-noise ratio (C/N), out-of-band noise (OBN), and pilot level are displayed on an 8-digit, 7-segment display, while the alarms are displayed on large, back-lit LED indicators. Each of the modules has front-panel status indicators and controls as necessary.

The design of the 7510 facilitates reconfiguration--any change that may be required for a different configuration may be made by changing a switch setting or replacing a plug-in printed circuit board on a module. In fact, much of the reconfiguration is accomplished automatically or can be accomplished from the downconverter front panel. For example, the loop filter for the threshold extension demodulator and the monitored out-of-band noise slot can be changed from the receiver front panel.

Parts of this reconfiguration can be made automatic. Based on the IF filter and the baseband processor wich are installed in the mainframe, the microprocessor in the downconverter will select the proper threshold extension demodulator loop filter and the proper out-of-band noise slot and reconfigure the 7510 accordingly. The microprocessor can also inform the operator if an invalid combination of demodulator, IF filter, and baseband processor has been installed in the mainframe.

Figure 2 is a block diagram of the 7510, and Table 1 contains the technical specifications of the 7510.

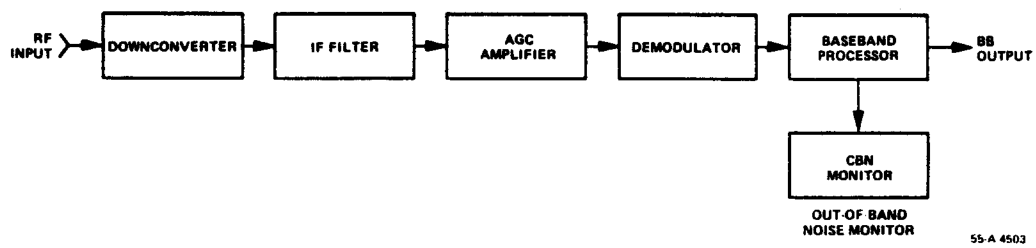


Figure 2. Block Diagram of the Model 7510 Message Receiver

The Downconverter

The downconverter section of the 7510 is completely modular and is removable from the mainframe as a unit. In addition to the RF circuitry housed in the downconverter, this module also contains the microprocessor and control circuitry for the receiver. A block diagram of the downconverter appears as Figure 3.

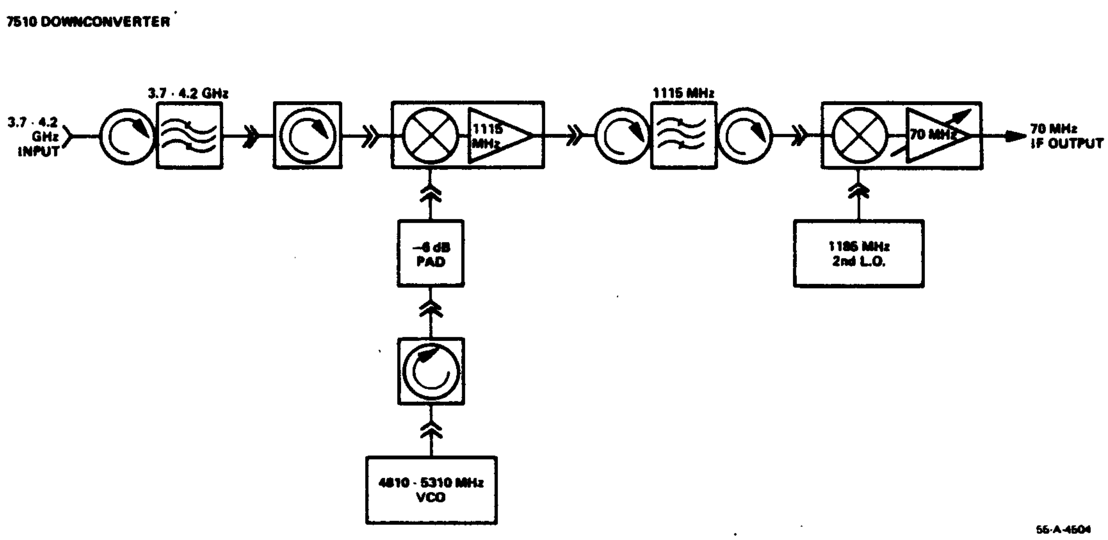


Figure 3. Block Diagram of the Downconverter of the Message Receiver

RF signals between 3.7 and 4.2 GHz enter the downconverter and are filtered by a 3.7 to 4.2 GHz bandpass filter to remove the image and other unwanted signals. The output of this filter is mixed with the synthesized first local oscillator. This synthesized LO allows the receiver to tune the received band in 125 kHz steps (the LO tunes from 4815 MHz to 5315 MHz). The signal which results from this mixing (1115 MHz) is amplified in the first mixer preamplifier, then fed to a 1115 MHz bandpass filter to remove undesired mixing products.

The filtered signal is mixed again with a fixed 1185 MHz second local oscillator in the second mixer preamplifier. The signal resulting from this mixing is the desired 70 MHz IF which drives the IF filter module. The gain of this module is variable and is set such that the overall conversion gain of the downconverter is 20 dB.

The reference for the first and second local oscillators is a temperature-controlled crystal oscillator contained in the first LO (synthesizer). Should stability greater than that provided by the temperature-controlled crystal oscillator be desired, then provisions are made for accepting an external 5 MHz reference.

IF Filter Module

The IF filter module contains the IF filter and a gain block, and is a modular unit which is installed behind the drop-down front panel. INTELSAT-compatible filters in bandwidths from 1.25 MHz to 36 MHz are available as plug-in units for this module. In addition, other bandwidths for special applications are also available.

AGC Amplifier Module

The AGC module contains a variable-gain amplifier which accepts signals from the IF filter module in the range of -45 to -5 dBm and provides the amount of gain necessary to produce an output at -5 dBm. This module also provides an accurate indication of the carrier-to-noise ratio (C/N) of the incoming IF signal. A bar graph display is provided on the module front panel calibrated for C/N and adjustable threshold alarms are also provided. Finally, an analog voltage corresponding to C/N is available on the rear of the 7510 and is sent to the controller in the downconverter for remote status monitoring.

Demodulator Module

There are two demodulator modules which are available for use in the 7510: a narrowband (threshold extension) demodulator for use in 12 through 432 channel systems, and a wideband demodulator for use in systems carrying more than 432 channels. The wideband demodulator is designed for use with an IF bandwidth up to 40 MHz and will provide linearity of 1% or better.

The threshold extension demodulator is a phase-locked loop demodulator designed to provide maximum threshold extension for narrowband systems. Six different loop filters are contained in this module and are electronically selected manually on the module, manually from the downconverter front panel, or automatically from the downconverter controller. The controller can decide, based on the baseband processor and IF filter installed in the main-frame, which loop filter is best suited to a particular configuration.

Baseband Processor

The baseband processor module contains the baseband amplifiers, the deemphasis networks, and the pilot monitor/alarm circuitry. Deemphasis networks for all INTELSAT and domestic message formats are available as plug-in cards for this module. Two baseband outputs are available with levels adjustable for -15 to -35 dBm test tone. The 60 kHz continuity pilot monitor provides a pilot level alarm as well as an analog voltage corresponding to the pilot level. This analog voltage is also sent to the controller in the downconverter so that the pilot level may be monitored remotely. Finally, this module also contains the high-pass filter for removal of energy dispersal.

Out-of-Band Noise Monitor Module

The 7510 has a fully synthesized out-of-band noise (OBN) module to maximize the receiver's flexibility. This monitor may be tuned in 1 kHz steps to monitor any OBN slot between 66 kHz and 15 MHz. There are alarm indicators on the front panel as well as a bar graph display which indicates the OBN level in picowatts. The module provides an OBN alarm and an analog voltage corresponding to the out-of-band noise at a connector on the rear of the receiver. This analog voltage is also fed back to the controller in the downconverter so that the OBN level can be monitored remotely.

The out-of-band noise module can be tuned manually on the module, manually from the downconverter front panel, or automatically from the microprocessor controller. The controller can decide, based on the baseband processor installed in the receiver mainframe, which out-of-band noise slot to tune to.

Table 1. Model 7510 Message Receiver Technical Specifications

Characteristic	Specification
Message Performance Parameters	
Channel Capacities	12 through 2892 channels
Deviation Range	100 kHz to 1 MHz 0 dBm0 test-tone deviation
Formats	All standard and expanded formats for INTELSAT IV, IV-A, and V
Deemphasis	Per CCIR Rec. 464
Baseband Output Level	-15 to -25 dBm test tone
Output Impedance	75 ohms, unbalanced
Output Return Loss	26 dB Min
Energy Dispersal Rejection	50 dB Min
Intermodulation and Equipment Noise	300 pWp0 Max
RF Characteristics	
Frequency Range	3.7 to 4.2 GHz
Input Impedance	50 ohms, unbalanced
Input Return Loss	23 dB Min
Noise Figure	14 dB Max
Input Level	-88 to -33 dBm
RF-IF Amplitude Response	± 0.1 dB/f_0 ± 18 MHz
Group Delay Distortion	0.03 ns/MHz linear, 0.01 ns/MHz/MHz parabolic, and 1 ns ripple over f_0 ± 18 MHz
Third-Order Intermodulation	60 dB below 2 -30 dBm signals
LO Leakage	-70 dBm Max

Message Exciter Model 7560

T.C. Mock

Introduction

The Model 7560 Message Exciter is part of a new generation of message and video exciters and receivers designed for high performance and maximum flexibility. This exciter incorporates new techniques which will improve performance while increasing ease of operation and status monitoring. The 7560 message exciter is modular in design (Figure 1) and can accommodate a wide variety of plug-in modules, including an auxiliary delay equalizer/attenuator module which will allow delay equalization via plug-in equalizer boards and remote control of output level via a 0.5 dB step attenuator.

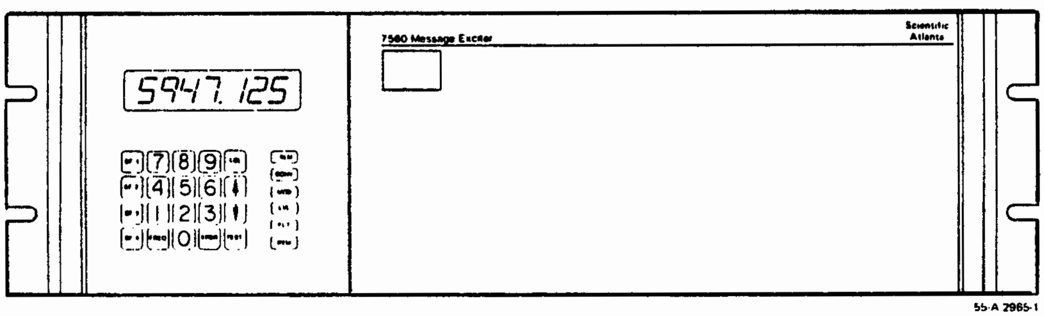


Figure 1. Front-Panel Drawing of the Model 7560 Message Exciter

General Description

The 7560 message exciter is housed in a 5-1/4-inch rack-mount chassis which provides maximum accessibility to the individual modules. The upconverter occupies the left-most quarter of the mainframe and is removeable from the front of the exciter. Each of the modules occupy a slot within a card frame assembly located to the right of the upconverter behind a drop-down front panel. Each of these modules has controls and status indicators as appropriate to its function.

The frequency and the module failure status are displayed on an 8-digit, 7-segment display on the front panel of the upconverter. The keypad on the front panel of the upconverter controls the output frequency of the exciter as well as controlling several of the module functions. Remote control and status monitoring of the exciter is accomplished using the SAbus.

The 7560 message exciter is capable of supporting any INTELSAT or domestic message format. All IF filters, roofing filters, and preemphasis networks plug into the appropriate module and are therefore easily changed. In addition, the controller in the upconverter monitors the modules installed in the mainframe and can warn the operator of invalid module combinations.

Table 1 lists the technical specifications of the 7560 message exciter while Figure 2 is a block diagram of the 7560.

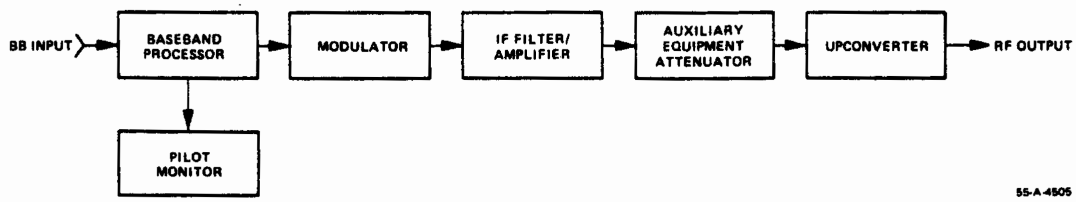


Figure 2. Block Diagram of the Model 7560 Message Exciter

Baseband Processor

All processing functions performed on the incoming baseband signal are performed in this module. These functions include preemphasis, addition of energy dispersal, baseband automatic level control, and lowpass filtering. Preemphasis and roofing filters are available as plug-in units for all INTELSAT and domestic message formats. The amount of added energy dispersal is automatically controlled to conform to CCIR recommendations. In addition, a baseband automatic level control is provided which will allow up to 10 dB of overdrive without additional distortion or overdeviation.

Modulator

Advanced circuit design techniques improve modulator performance through the use of an extremely linear FM modulation method. Linear frequency modulation of the carrier is accomplished by modulating two voltage-controlled oscillators in opposite directions and mixing their outputs to obtain the required 70 MHz output. This technique typically results in less than 0.5% modulator nonlinearity over a 60 MHz modulation bandwidth.

IF Filter/Amplifier

Spectrum control filtering of the modulated (70 MHz) carrier is performed by the IF filter/amplifier. INTELSAT-compatible filters with bandwidths from 1.25 to 36 MHz are available as plug-in cards for this module. Other filters are available for special applications. In addition to filtering, this module also provides automatic level control to ensure constant IF output power to the upconverter.

Upconverter

The 7560 upconverter is a modular unit and is removable from the front of the exciter. In addition to the RF circuitry, the upconverter module also houses the monitor and control microprocessor for the exciter. A block diagram of the upconverter appears as Figure 3.

Table 1. Model 7560 Message Exciter Technical Specifications

Characteristic	Specification
Message Performance Characteristics	
Channel Capacities	12 through 2892 channels
Deviation	100 kHz to 1 MHz 0 dBm0 test tone deviation
Formats	All standard and expanded formats for INTELSAT IV, IV-A, and V
Preemphasis	Per CCIR Rec. 464
Baseband Frequency Response	±0.25 dB 12 kHz to f_{max}
Baseband Input Impedance	75 ohms, unbalanced
Baseband Return Loss	26 dB Min
Energy Dispersal Frequency	20 Hz to 150 Hz
Energy Dispersal Level	Automatically adjusted as a function of input level
Intermodulation Equipment Noise	Less than 300 pWp0, 12 through 2892 channels
RF Output Characteristics	
Output Frequency Range	5.925 to 6.425 GHz
Output Level	-10 dBm to +7 dBm
Output Level Stability	±0.25 dB per day
Output Impedance	50 ohms, unbalanced
Output Return Loss	20 dB Min
Frequency Agility	Synthesized in 125 kHz steps
IF-to-RF Amplitude Response	0.25 dB, $F_0 \pm 18$ MHz
Delay Distortion	0.03 ns/MHz linear, 0.01 ns/MHz/MHz, and 1 ns peak-to-peak ripple over f_0 ±18 MHz
Modulator Linearity	Less than 1%
Spurious Output	Less than -90 dBm 4 kHz

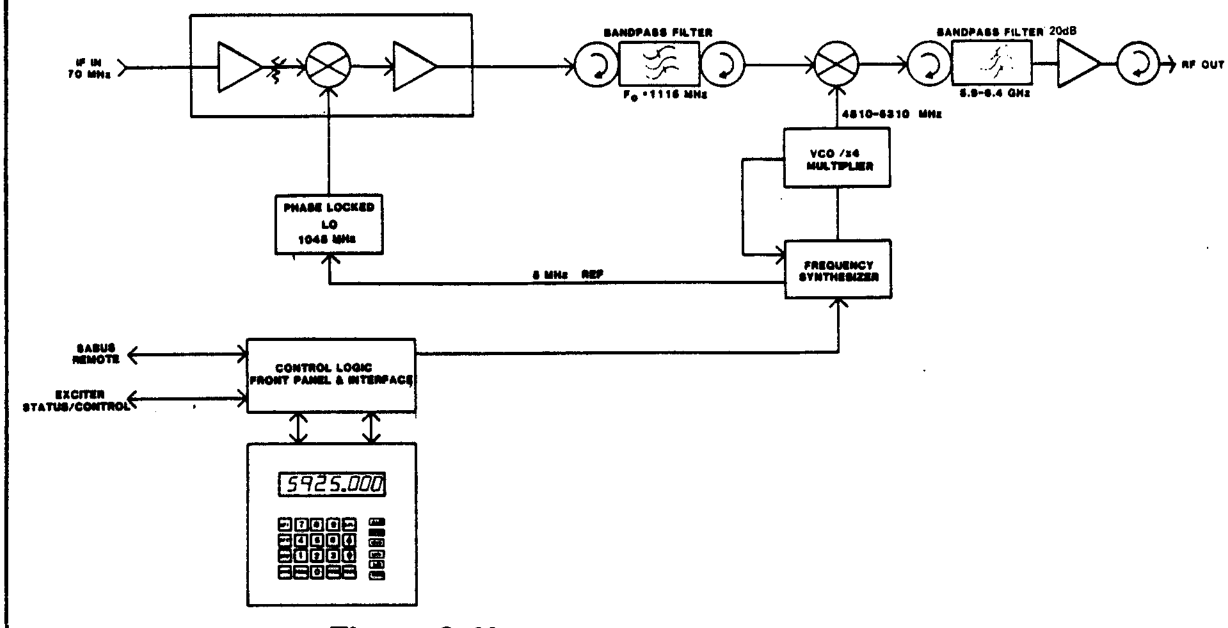


Figure 3. Upconverter Block Diagram

IF signals at 70 MHz enter the upconverter and are amplified in the mixer preamplifier module, then mixed with a 1045 MHz first local oscillator. The result of this mixing is a 1115 MHz second IF frequency. This signal is filtered to remove undesired products and amplified. A gain of 20 dB is required to provide the high level signal for the output mixer. The gain of this module is variable and it allows a conversion gain between -5 and +30 dB.

A phase-locked oscillator at 1045 MHz is the first local oscillator. This oscillator derives its reference from a 5 MHz temperature-controlled crystal oscillator (TCXO) in the second local oscillator synthesizer.

The 1115 MHz second IF signal is mixed again with the output of a synthesized second local oscillator. This synthesizer allows the output frequency to vary in 125 kHz increments. This module also contains the 5 MHz TCXO used as a reference by both the synthesizer and the first LO. If a more stable reference is desired, provisions for the insertion of an external 5 MHz source are available.

The result of this mixing process is then bandpass-filtered to remove unwanted mixing products. An output amplifier with 20 dB of gain provides output levels up to +10 dBm.

Pilot Monitor

The pilot monitor module monitors the level of the incoming 60 kHz continuity pilot. This level is displayed on an LED bar-graph indicator on the module front panel. Should the level of this pilot be either too low or too high, the pilot monitor will initiate an alarm. In addition, an analog voltage corresponding to the pilot level is fed to the rear panel of the exciter and to the controller in the upconverter so that the pilot level can be monitored remotely.

Series 8300 SCPC Equipment

T. Brigman

Introduction

One technique for increasing the effective number of channels throughout a satellite transponder is called Single-Channel-per-Carrier, or SCPC. In this technique each voice channel is modulated directly on a separate RF carrier, which is then offset from other such carriers on a frequency grid that spans the satellite transponder available bandwidth. These frequencies are carefully controlled to allow close, accurate spacing of the individual carriers generated at different stations. For transmission to the satellite, all carriers are combined into a composite IF, which is upconverted to the 5.925 to 6.425 GHz band and transmitted to the satellite. The received signal from the satellite in the 3.7 to 4.2 GHz band is converted back to IF and passed through common processing equipment prior to the selection and demodulation of the individual carriers in the channel equipment. An immediate savings is realized here by sharing the expensive microwave equipment among all the channels.

With the SCPC techniques, it is possible to allocate the communications capacity of a station to each of several destinations on a channel-by-channel basis. It is also possible to add capacity, one channel at a time, as required. This is made possible by the multipoint nature of the communications satellite, coupled with the independent channel nature of the SCPC scheme. These characteristics are very useful in networks with many stations that require light to medium traffic between most or all of these stations, or those that have light traffic initially with an uncertain growth pattern and rate. Ultimately, the SCPC system can provide a very large overall communications capacity from the satellite transponder.

Scientific-Atlanta has designed an SCPC equipment system which can be used efficiently in both light and heavy traffic systems. The following paragraphs provide an overall system description, the equipment bay configurations for both heavy and light traffic systems, and module-by-module equipment descriptions.

SCPC Systems Description

The proposed SCPC terminal equipment is intended for use in satellite communications earth stations utilizing domestic or international synchronous communications satellites. The SCPC equipment provides duplex high-grade voice channels for interface at the four-wire point. A block diagram of the SCPC terminal is shown in Figure 1.

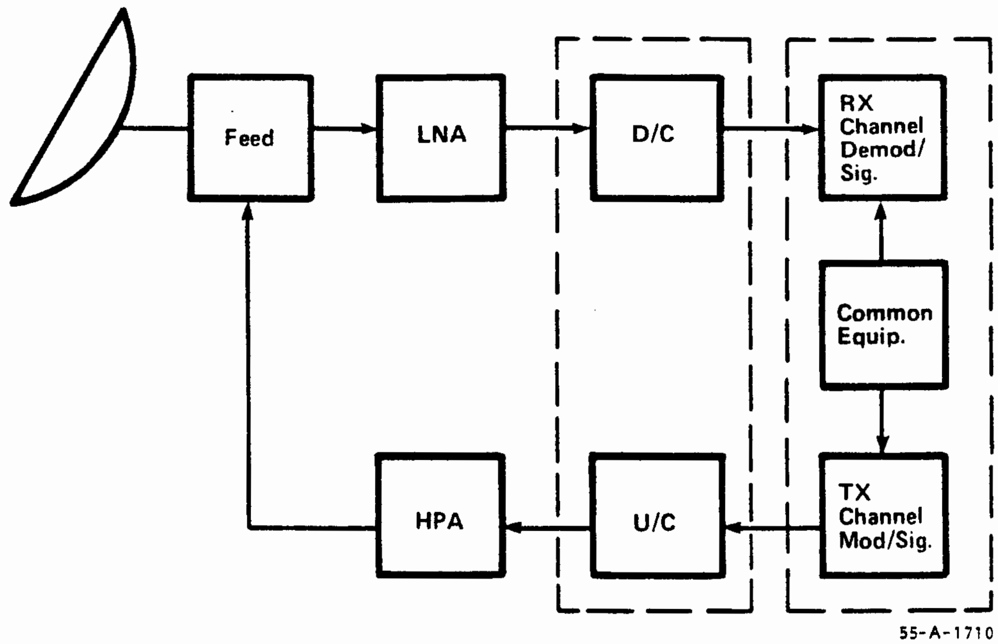


Figure 1. SCPC Terminal Block Diagram

The function of the SCPC terminal is to accept VF signals from interface circuits, process the signals, and modulate each signal onto an individual IF carrier. All IF carriers are then combined to result in a single composite IF output. This combined IF signal is then translated to the appropriate uplink frequency band near 6 GHz and fed to the station high-power amplifier (HPA). On the receive side of the transmission path, the signal from the station low-noise amplifier (LNA) comes into the SCPC terminal, where it is translated in frequency down to the terminal common IF frequency. This composite IF signal is processed to provide frequency correction (AFC) and then distributed to all channel units in the terminal. Each carrier is then demodulated individually, and the resulting baseband signal processed to provide a flat VF output for each channel. The interface at baseband is four-wire 600-balanced.

An important general advantage of SCPC communications is the relative ease of changing terminal capacity as requirements for the network change. Within the limitations of available transponder bandwidth, the station capacities within a system can be increased directly by adding new channel equipment and increasing HPA power.

The proposed equipment is intended for normal operation at C/k T = 54 to 58 dB Hz, and is intended to provide maximum utilization of transponder capacity, while providing toll-quality transmission by the use of companding circuitry. The baseband response is essentially flat from 300 to 3,400 Hz, and the equipment offers excellent delay distortion characteristics. These characteristics, together with the threshold extension demodulator, afford good voice communications as low as C/k T = 50 dB Hz, and excellent data transmission performance at the normal operating points. The proposed standard product equipment utilizes 22.5 kHz, 30 kHz, or 45 kHz channel spacing depending on customer requirements.

Table 1 lists the main equipment performance specifications.

Table 1. Single-Channel-per-Carrier Equipment Performance Specifications

Characteristic	Specification
	General
Emphasis Characteristic	Dual break 600/5,000 Hz
VOX/Echo Suppression	Per CCITT G.161
Companding	Per CCITT G.162
Receiver Squelch	Operational for C/N_0 = 49 dB-Hz
	VF Interface
Input/Output	4-wire, 600 ohms, balanced
Return Loss	20 dB, 300 to 3,400 Hz
Transmit Level (0 dBm0)	+7 to -16 dBm
Receive Level (0 dBm0)	+7 to -16 dBm
	IF Interface
Frequency	52 to 88 MHz
Input/Output	75 ohms, unbalanced
	IF Looped Terminal Performance
Notched Noise (without Compander) (1) at 55.4 dB-Hz	
With Emphasis	-36.0 dBm0p
Without Emphasis	-31.0 dBm0 (flat weighting)
Channel Total Harmonic Distortion (1,000 Hz, 0 dBm0)	<2.5%

(1) Measured in presence of -3 dBm0 1,000 Hz tone.

Characteristic	Specification
IF Looped Terminal Performance - continued	
Channel Intermodulation (Two-tone Pairs, at 860 Hz and 1,380 Hz with Composit Level -7 dBm0)	Second Order: -35 dB Third Order: -40 dB
Channel Frequency Response	400 to 3,000 Hz ±0.8 dB 300 to 3,400 Hz +1, -1.5 dB
Channel Delay Distortion (Ref. 0 to 188 Hz)	600 to 3,000 Hz 200 microseconds 500 to 3,000 Hz 300 microseconds
Channel Crosstalk (Uncompanded) (adjacent Channel Modulated by 0 dBm0 1,000 Hz Tone)	50 dBm0
Level Stability	±0.5 dBm/month, at 25°C
Compander Tracking	±1 dB
CONVERTER SYSTEM PARAMETERS **Transmit Chain Performance**	
Frequency Band	5.925 to 6.425 GHz
Output Third Order Intercept Point	+5 dBm
Gain	-5 dB to +15 dB (adjustable)
Spurious	
In-Band	5.9 to 6.4 GHz -65 dBm
Out-of-Band	3.7 to 4.2 GHz -80 dBm
Other	-65 dBm
Output Impedance	50 ohms
Frequency Stability	±1 part per 10^8 per month
Power	115V ac ±10%, 60 Hz ±5%

Characteristic	Specification
	Receive Chain Performance
Frequency Band	3.7 to 4.2 GHz
Noise Figure (Downconverter Input)	14 dB (typical)
Input Levels	-75 dBm to -90 dBm per carrier
Frequency Stability	±1 part per 10^8 per month
Input Impedance	50 ohms
Power	115V ac ±10%, 60 Hz ±5%
Third-Order Intermodulation (Two -10 dBm Output Signals)	-40 dBc
Spurious Outputs	-75 dBm

Equipment Configuration

The SCPC equipment is divided into two types--common equipment and channel unit equipment.

Common Equipment

This consists of those units which serve all the channel units in the station. These units are:

a. High-power amplifier

b. Low-Noise Amplifier

c. Up/downconverter

d. Channel Equipment Bay

e. Common Equipment Bay

f. Reference Generator Unit (RFG)

g. Pilot Receiver Unit (PRU)

Channel Unit Equipment

This consists of:

a. Channel Unit Cards

   (1) Modem

   (2) Synthesizer

   (3) Voice Processor

b. Channel Unit Card Cage - This unit holds four channel units

c. Channel Unit Equipment Bay - This bay can hold up to 32 channel units

d. Channel Unit Power Supply.

Figures 2 through 4 depict various SCPC equipment configurations.

Figure 2 shows a rack elevation for one configuration of SCPC equipment. As shown, the rack contains redundant common equipment along with redundancy switching.

As shown, the rack can hold up to 16 channel units as well as the common equipment.

Figure 3 shows a rack elevation for the channel unit equipment bay. The bay contains up to eight card cages which can hold a total of 32 sets of channel unit cards. The card cages have 13 card slots. The RFG and Pilot Rx are capable of servicing all channel units in one channel unit bay. The bay is shown with redundant RFGs. The channel unit power supply shown is is capable of driving 16 channel units. Thus two power supplied are needed to power a bay with 32 channel units.

The channel unit power supply consists of two power supplies, each of which is capable of powering two card cages. There is also space in the power supply chassis for a third power supply also capable of driving the two card cages along with switching to allow 1 for 2 redundancy switching.

The bay is air cooled by blowers mounted above the power supplies. The card cage presents low impedance to the cooling air. The equipment bay is a cabinet with rear door access and side panels which confines the blower output and channels it up through the card cages to provide very efficient cooling. Typically, the temperature rise from bottom to top of the rack is less than 10 degrees Fahrenheit.

Figure 4 shows a rack elevation of a combined common and channel unit equipment bay. This layout is used in smaller stations. The bay can hold up to 16 channel units. The common equipment is redundant and includes redundancy switching. The channel unit equipment is blower cooled as is the conversion equipment. The HPAs are set cooled using internal blowers.

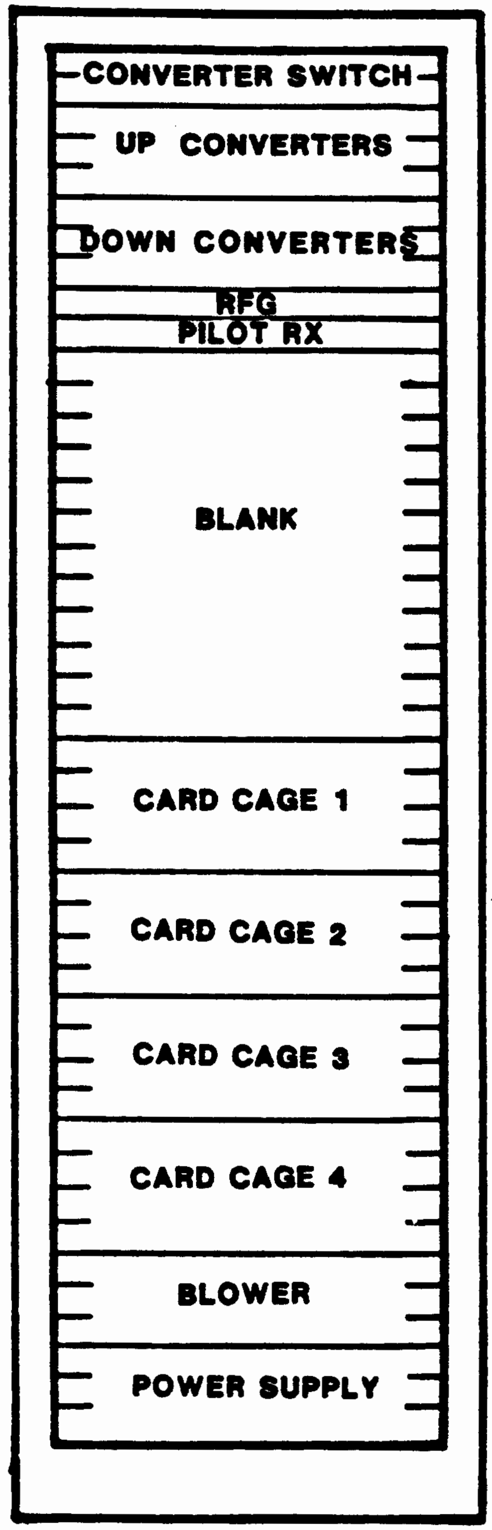


Figure 2. Rack Elevation Common Equipment Plus Channel Units

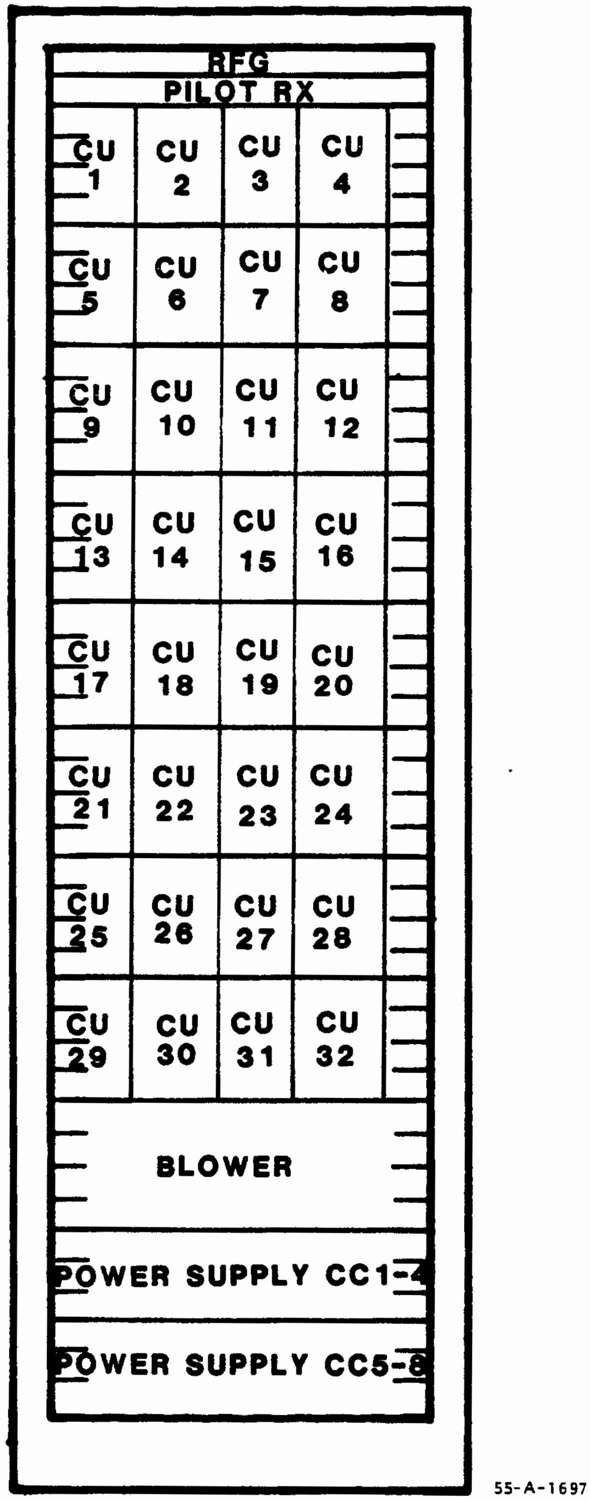


Figure 3. Rack Elevation Channel Unit Bay Holds Up to 32 Channel Units

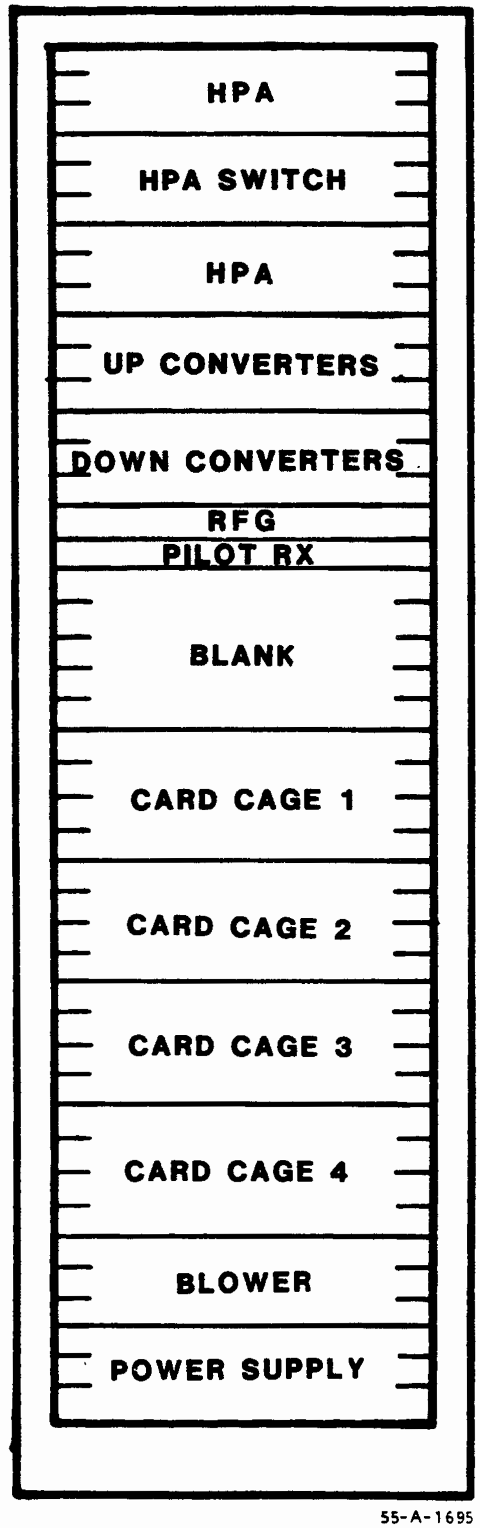


Figure 4. Rack Elevation Redundant Common Equipment Including HPA and Provisions for 16 Channel Units

Equipment Descriptions

Model 7570, 7520 Converters

General. The Model 7570, 7520 Converters are specifically designed for SCPC operation. SCPC operation requires higher frequency stability, lower FM residual noise and higher linearity than either message or video system operation. The converters are of the dual-conversion type which allow tuning to any transponder by changing the local oscillator frequencies. The converters are synthesizer-tuned and have three modes of operation; Frequency mode, DOMSAT transponder mode and INTELSAT transponder mode. In the Frequency mode, the converters tune in 5 MHz increments. In the DOMSAT mode, the converters tune to the center frequencies of the 24 DOMSAT satellite transponders. In the INTELSAT mode, the converters tune to the center frequencies of the twelve INTELSAT transponders. The microprocessor automatically accounts for the 50 MHz frequency step between transponder 6 and 7 in the INTELSAT mode. Entry to the microprocessor is by means of a front-panel-mounted keyboard. The frequency of operation is entered into the keyboard after the appropriate operation code has been entered. The frequency is displayed in MHz, and the transponder is displayed by number on the front-panel display. The converters are packaged in three systems as follows:

a. Nonredundant System Model 7575. An up- and a downconverter are packaged in a 5-1/4-inch-high, 19-inch, rack-mountable package. The package also contains the power supplies and a 5 MHz reference oscillator used by the converters. The front-panel layout is shown in Figure 5.

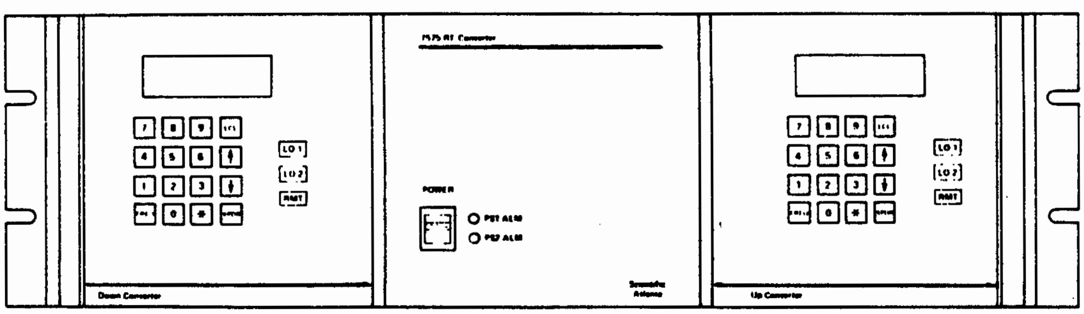


Figure 5. Non-Redundant System Front-Panel Layout

b. Dual Upconverter System Model 7570-01. Two upconverters are packed in a 5-1/4-inch-high, 19-inch, rack-mountable package which also contains two power supplies, one for each converter, and a 5 MHz reference oscillator. The front-panel layout is shown in Figure 6.

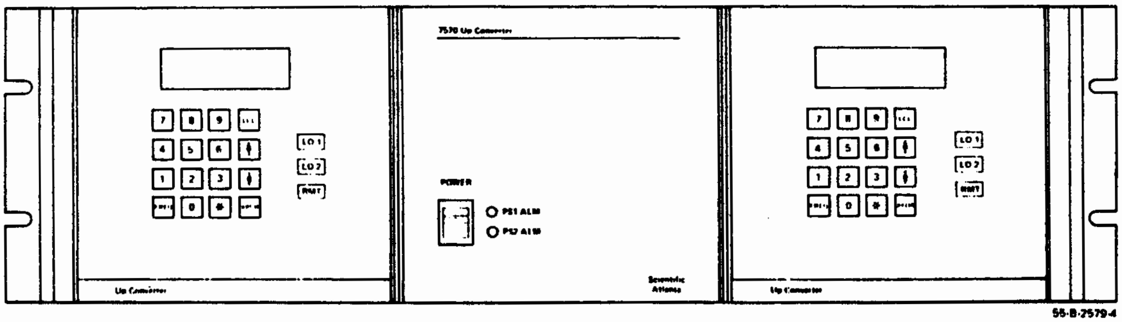


Figure 6. Redundant Upconverter System Front-Panel Layout

c. Redundant Downconverter Model 7520-01 System. Two downconverters, along with the redundancy switch, two power supplies, and a 5 MHz reference oscillator, are mounted in the 5-1/4-inch-high, rack-mountable package. The necessary switching logic is also mounted in this package. Status display and manual switching control are mounted in the center of the front panel. Figure 7 shows the front-panel layout.

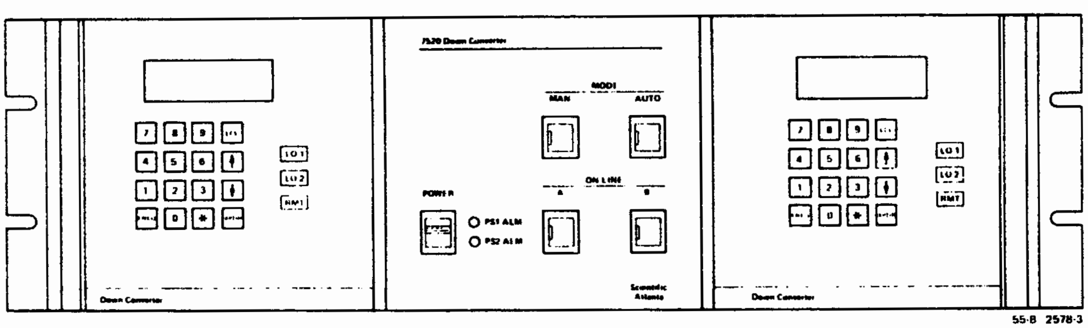


Figure 7. Redundant Downconverter System Front-Panel Layout

All local oscillators in the converter systems are derived from and have the same stability as a highly stable, low-noise 5 MHz proportional oven-controlled crystal oscillator. In the nonredundant system, one oscillator controls both converters. The redundant upconverter system and the redundant downconverter system are controlled by a pair of 5 MHz oscillators which are connected to drive an up/downconverter pair.

The converter operation is described in the next section. Table 2 shows the upconverter specifications. Table 3 gives the specifications for the down-converter.

Table 2. Model 7570 Upconverter Specifications

Characteristic	Specification	
Input Frequency Range	52 to 88 MHz	
Input Impedance	75 ohm (BNC female)	
Input Return Loss	26 dB	
Maximum Composite Level	SCPC	-17 dBm
	Message	-5 dBm
	Video	-5 dBm
	Digital	-10 dBm
Output Frequency Range	5.925 to 6.425 GHz	
Frequency Control	Synthesizer (5 MHz STEP size)	
Output Impedance	50 ohm (type N female)	
Output Return Loss	20 dB	
Upconverter Gain	-5 dB to +15 dB (adjustable)	
Output Third-Order Intercept Point	+5 dBm (at U/C gain = +5 dB)	
Reference Frequency Stability	$\pm 1.0 \times 10^{-8}$/month	
Stability Translation Error	1217 Hz on output/Hz on reference	
Output Frequency Stability	±250 Hz	
Output Frequency Response	±0.5 dB/36 MHz	
Spurious Outputs		
5.9 to 6.4 GHz	-65 dBm	
3.7 to 4.2 GHz	-80 dBm	
Other	-65 dBm	
Total Converter Noise	55 Hz rms in band from 300 to 3,400 Hz	
Specified Temperature Range	15 to 35°C	
Operational Temperature Range	0 to 41°C	

Table 3. Model 7520 Downconverter

Characteristic	Specification	
Input Frequency Range	3.7 to 4.2 GHz	
Input Impedance	50 ohm (type N female)	
Input Return Loss	>20 dB	
Maximum Composite input Level	SCPC	-27 dBm
	Message	-35 dBm
	Video	-35 dBm
	Digital	-34 dBm
Output Frequency	52 to 88 MHz	
Frequency Control	Synthesizer (5 MHz step size)	
Output Impedance	75 ohm (BNC female)	
Output Return Loss	>24 dB	
Downconverter Gain	10 to 30 dB (adjustable)	
Output Third-Order Intercept Point	+2 dBm (at D/C gain = 10 dB)	
	+10 dBm (at D/C gain = 20 dB)	
	+15 dBm (at D/C gain = 30 dB)	
Reference Oscillator Stability	$\pm 1 \times 10^{-8}$/month	
Stability Translation Error	826 Hz on output/Hz on reference	
Output Frequency Response	±1.0 dB/36 MHz	
Spurious Outputs	-75 dBm	
L.O. Leakage	-55 dBm	
Total Converter Noise	50 Hz rms in band from 300 to 3,400 Hz	
Specified Temperature Range	15 to 35°C	
Operational Temperature Range	0 to 41°C	
Noise Figure	14 dB typical 16 dB Max	
Image Rejection	50 dB	

Converter Operation. Operational descriptions and block diagrams of the up- and downconverter appear in the following paragraphs.

Upconverter. Figure 8 is a block diagram of the upconverter. The input IF signal (52 to 88 MHz) is applied to a broadboard linear amplifier with variable gain housed in the first mixer preamp. The gain range is 0 to 20 dB. This signal is then converted to 1,115 MHz in the first mixer. The L.O. signal is 1,045 MHz (see diagram for other frequencies) which is phase locked to the highly stable 5 MHz reference oscillator. The mixer output is amplified again by a fixed 18 dB gain stage. This signal is then applied to a 40 MHz wide bandpass filter which uses input and output circulators for spurious prevention. After filtration, the signal is applied to the second mixer where it is upconverted to the 5.925 to 6.425 Hz band. Here, the L.O. signal is synthesizer controlled and is also phase locked to the stable 5 MHz reference. The output of the synthesizer VCO is multiplied by four and then filtered before being applied to the mixer. The mixer output signal is then filtered by a 500 MHz wide bandpass filter which also uses input and output circulators for spurious prevention. The filter output is then taken to the rear of the chassis to the type-N output connector.

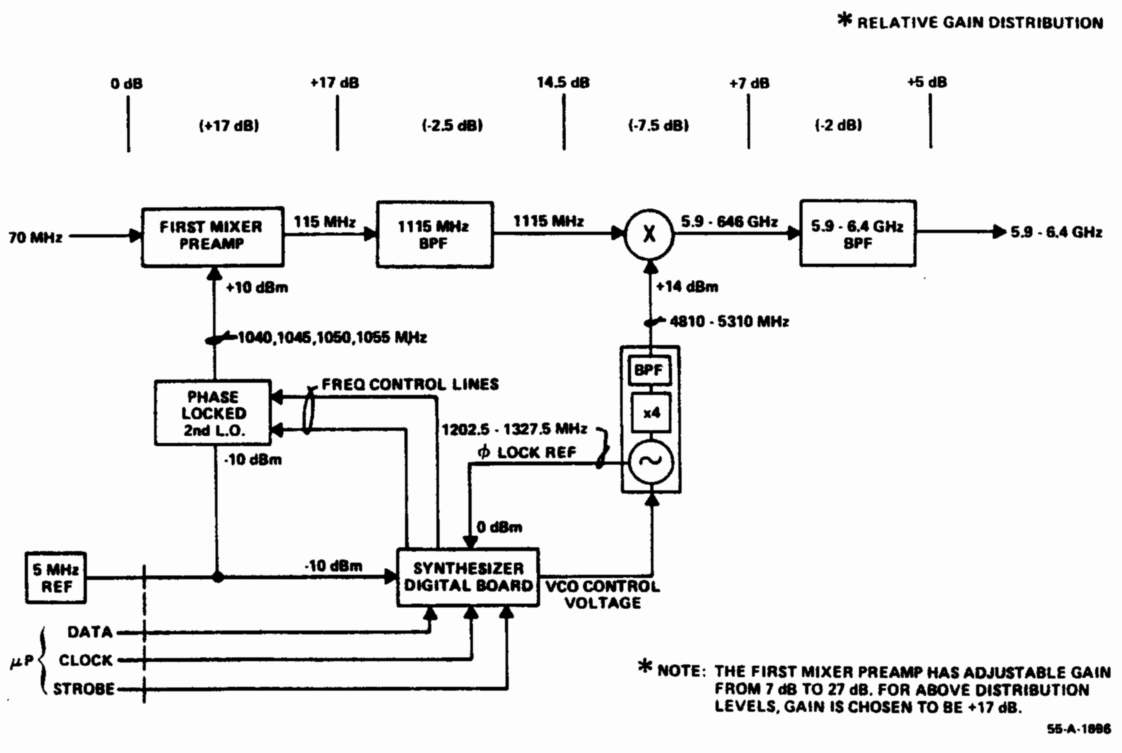


Figure 8. Upconverter Block and Level Diagram

Downconverter. Figure 9 is a block diagram of the downconverter. The input RF signal (3.7 to 4.2 GHz) is applied to a 500 MHz wide bandpass filter. The signal then goes into the first mixer where it is converted to 880 MHz. The L.O. signal is a synthesizer-controlled VCO phase locked to the stable 5 MHz reference. The VCO output is multiplied by four and then bandpass filtered before being applied to the mixer. The mixer output is applied to a broad-board linear amplifier with a fixed gain of about 24 dB. This signal then goes to a 70 MHz wide bandpass filter and on to the second mixer. The second mixer converts the signal to the 52 to 88 MHz band. This L.O. signal is also a synthesizer-controlled VCO phase locked to the stable 5 MHz reference. Following the mixer is a variable gain broadband amplifier with 20 dB of adjustment range. Then the signal is filtered again by a 36 MHz wide band-pass filter and applied to the output amplifier. The amplifier output then goes to the 70 MHz IF out BNC connector at the rear panel.

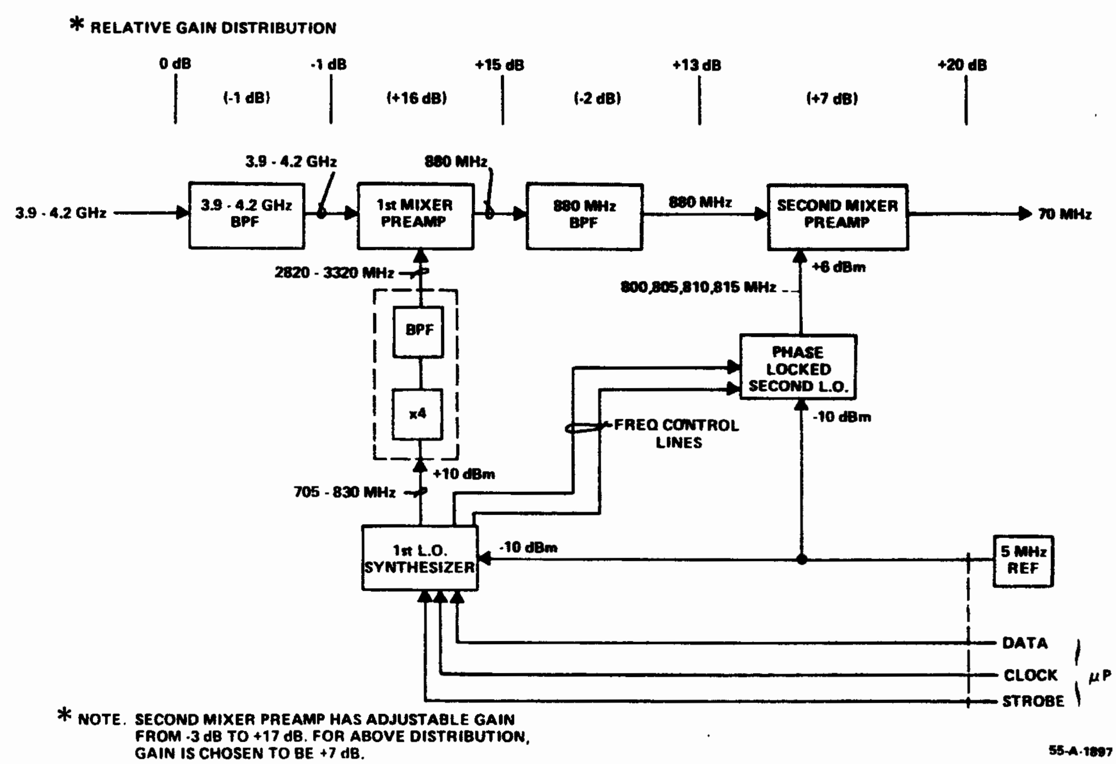


Figure 9. Downconverter Block and Level Diagram

Reference Frequency Generator RFG. The RFG provides reference signals for the SCPC channel units. The signals provided are:

- 3.6 MHz Synthesizer Reference
- 286.859375 kHz Modulator and Voice Processor Reference
- Pilot Signal
- Signaling Reference (2,600 or 3,825 Hz)

The first two of the signals are provided on all RFGs. The last two are optional plug-in boards. Typically, the pilot generator is only used in one or two stations in a given SCPC system. The RFG module is plugged into a motherboard mounted above the channel unit card cages. It is powered by the channel unit card cage. The RFG can drive up to 32 channel units.

Operation. Figure 10 is a block diagram of the RFG configured for redundant operation. Operation is as follows.

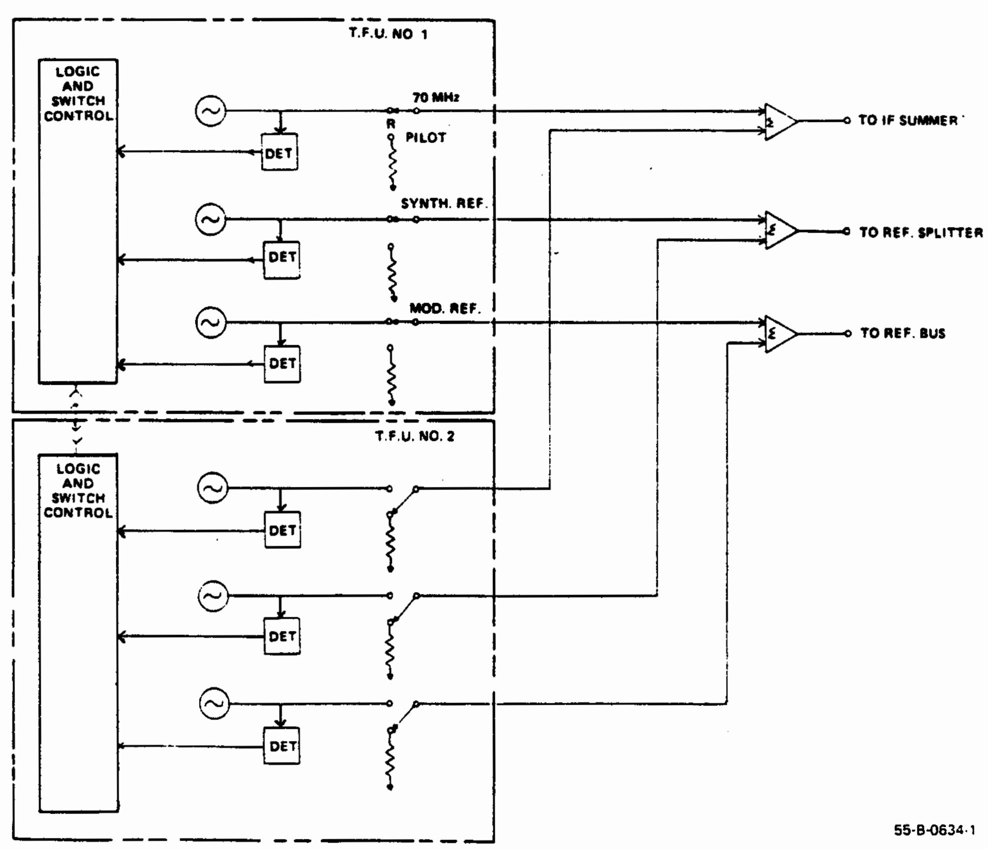


Figure 10. Redundant RFG Block Diagram

a. Synthesizer Reference. The 3.6 MHz signal is provided by a highly stable temperature-controlled crystal oscillator (TCXO). The oscillator is buffered. A variable attenuator sets the output level to +7 dBm for all racks where the number of channel units is 1 or 32. The frequency stability of this synthesizer reference signal is 1 part in 10^{-8} per day.

b. Modulator and Voice Processor Reference. The 286.859375 kHz signal is derived by dividing a 4.58975 MHz TCXO signal by sixteen. The divided signal is bandpass filtered, and the signal output level is set using the variable attenuator. Frequency stability of the source is the same as that of the synthesizer reference.

c. Pilot Reference. The pilot signal is derived from a synthesizer locked to the 3.6 MHz synthesizer reference. The synthesizer is switch programmable to provide pilot signals of 52 to 88 MHz in channel increments. These signals allow the pilot to be centered in the full transponder, either half transponder or any of the four quarter-full transponder segments.

   The synthesizer output is buffered and the level set by a variable attenuator. Frequency stability is the same as that of the synthesizer reference.

d. Signaling Reference. This signal is derived from the 3.6 MHz TCXO by division in a programmable divider. Division by 941 gives a signal at 3825.72 Hz and division by 1,385 gives a signal at 2599.28 Hz. The divider output is low-pass filtered, and the level is set via a variable attenuator. Frequency stability is the same as that of the synthesizer reference.

e. Alarms. All signals are level detected and compared to a reference level. Failure will give an alarm indication to an OR gate which drives a front-panel red LED. This LED is illuminated when any signal exceeds a pre-determined limit. The pilot generator also has a loss-of-lock alarm which will also cause the alarm LED to turn on.

Redundant Operation. When the RFG is used in a redundant system, a switching and logic board is added to the RFG motherboard. Both RFG modules are identical in the redundant pair.

The operation is as follows. Consider the "synth ref" redundant pair. As shown, RFG Module No. 1 is on-line. Its output is fed via the hybrid summer shown to the synthesizer reference splitting system. At this time RFG Module No. 2 is off-line, and the other leg of the summer is terminated. When there are no faults, this mode of operation is set up by switches on each TFU switching and logic card. RFG Module No. 1 is "on-line."

Consider a failure of the synthesizer reference signal. If the synthesizer reference level is within limits in RFG Module No. 2, the switch in RFG Module No. 1 operates, removing any signal from RFG Module No. 1 and terminating its leg of the summer. At the same time, the switch in RFG Module No. 2 operates and connects the synthesizer reference signal to the summer.

All signals are made redundant in the same manner except the signaling reference signal.

This signal uses the same logic as the others but does not feed a signal summer. It feeds a "bus" system and so thus does not need a summer.

The system logic is arranged so that if either RFG module is removed the other remaining RFG module drives the rack.

Pilot Receiver Model 7720. The pilot receiver system performs the following functions.

- Provides receive-side AFC for the station via the 40.18 MHz (45, 22.5 kHz system) second local oscillator in the channel unit demodulators for 45 kHz channel space or via 40.12 MHz (nominal) for 30 kHz channel space.

- Provides automatic gain control output which may be used to control a steptrack antenna pointing system.

- Provides frequency agility using a frequency-agile, first-local oscillator, thus allowing great flexibility in the placement of the SCPC pilot.

- Automatic acquisition of pilot if phase-lock loss is indicated. Sweep search circuit is activated on loss of lock and deactivated when lock is acquired.

- Operation may be either single-ended or redundant.

Figure 11 is a top view of the complete pilot receive system, showing the two pilot receiver modules plugged into the motherboard.

The unit is a rack-mountable chassis, 1-3/4 inches in height. It contains two pilot receiver modules which plug into a motherboard containing the redundancy switching circuitry. The unit is powered by external plus and minus 19-volt supplies. The pilot receiver modules plug in from the front of the unit. The 40.18 MHz input signal is sufficient to drive a channel unit bay containing up to 32 channel units.

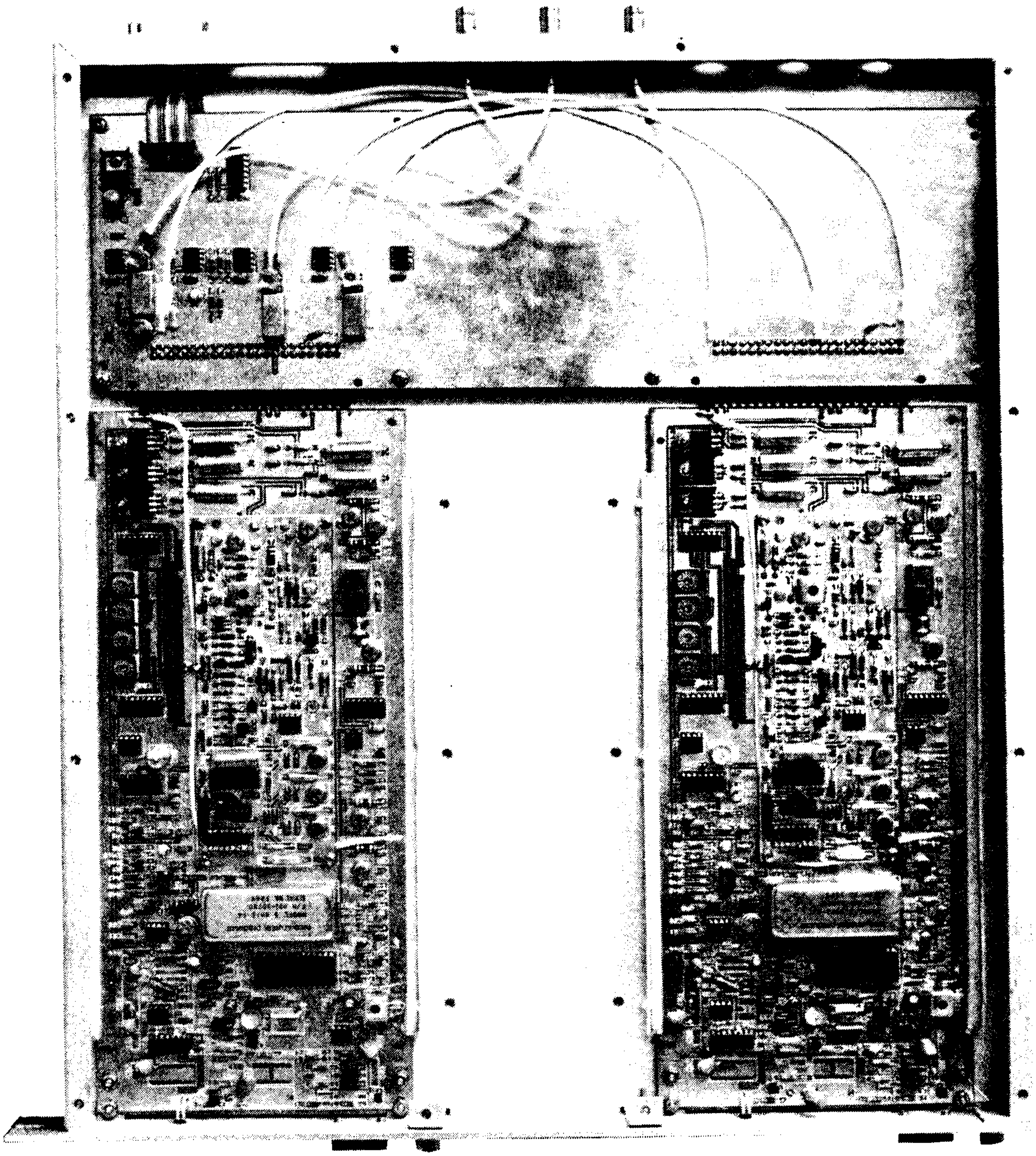


Figure 11. Top View of the Complete Pilot Receive System

Pilot Receiver Module Operation. Figure 12 is a block diagram of the pilot receiver module. Figure 13 is a more detailed block diagram of the AGC system and Figure 14 is a block diagram of the phase-locked loop threshold extension system. The receiver operation is as follows:

a. Overall Description. The pilot receiver performs the AFC function by comparing the received pilot to a very stable 3.6859375 MHz* crystal oscillator in a phase detector and changing the frequency of a 40.18 MHz (nominal) voltage-controlled oscillator until the frequency difference at the phase detector is reduced to zero. This process transfers an equal frequency error to the 40.18 MHz oscillator which cancels out the satellite translation error at the second mixer in the channel unit demodulator.

b. Signal Flow Description. Refer to Figure 12, SCPC Pilot Receiver block diagram. The input signal may be in the range of 52 to 88 MHz and is fed to the receiver via the receive IF distribution system at a nominant level of -55 dBm. The signal is mixed down to 46 MHz in the broadband first mixer. The local oscillator signal is derived from a synthesized oscillator, the frequency of which is set by switches which define the synthesizer divide ratio. This synthesized LO allows frequency-agile receiver operation. The synthesizer is identical to the units used in the channel unit synthesizer Model 7740. After the first mixer, the signal is bandpass-filtered in a filter centered at 46.0775 MHz** with a 3 MHz bandwidth. The signal is then buffer-amplified and applied to the second mixer. The local oscillator for this mixer is provided by the 40.18 MHz VCO. This mixer output is buffer-amplified and filtered by a 34-kHz-wide, bandpass crystal filter centered at 5.8975 MHz***. Following this filter, the signal is applied to the automatic gain control (AGC) circuitry. (Refer to Figure 13).

   The crystal filter output is applied to a variable gain amplifier via a 0 dB gain buffer. The variable gain amplifier is level detected compared to a reference voltage, and the difference in signal is amplified and integrated and used to control the variable

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NOTE: *3.6859375 MHz for 45 kHz channel increment
3.728125 MHz for 30 kHz channel increment

**46.0775 MHz for 45 kHz channel increment
46.085 MHz for 30 kHz channel increment

***5.8975 MHz for 45 kHz channel increment
5.965 MHz for 30 kHz channel increment

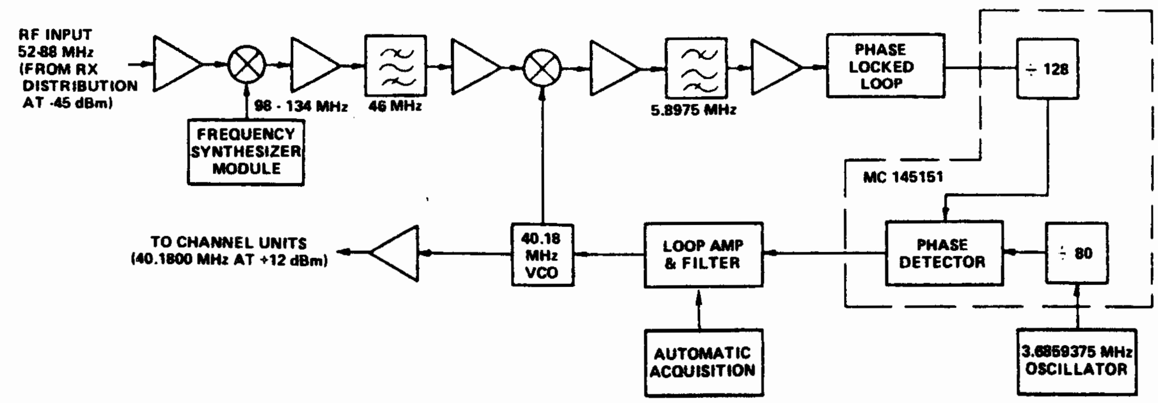


Figure 12. SCPC Pilot Receiver

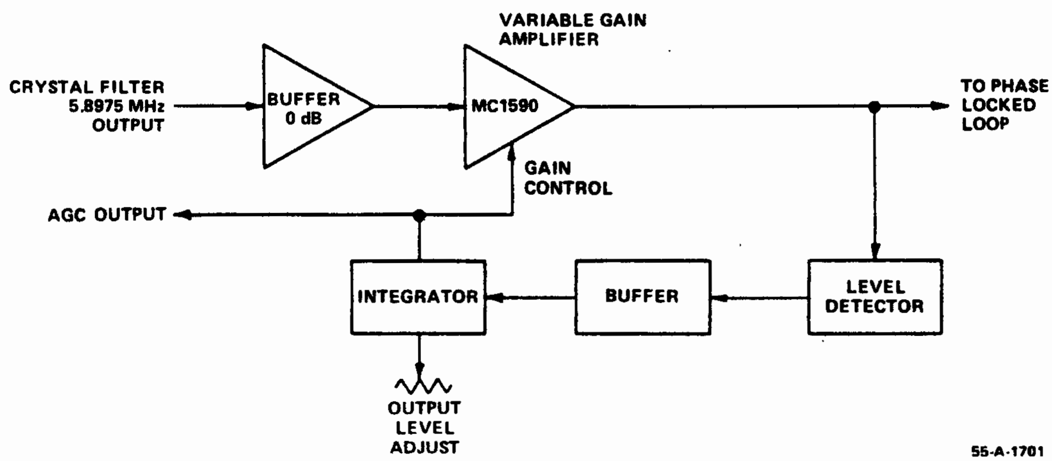


Figure 13. Pilot Receiver AGC

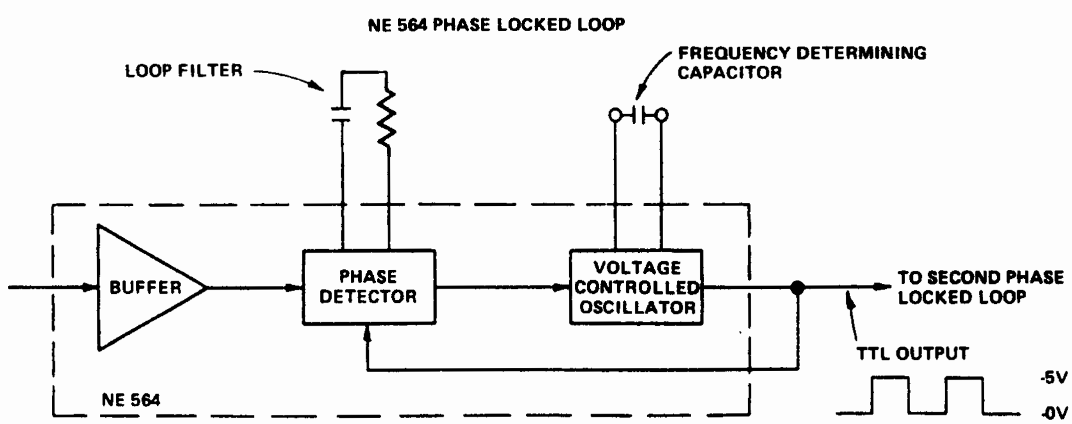


Figure 14. Phase-Locked Loop - NE 564

55-A-1698-1

gain amplifier is level detected compared to a reference voltage, and the difference in signal is amplified and integrated and used to control the variable gain amplifier to close the AGC loop. A sample of this voltage is signal-conditioned to provide a low impedance output and is available at the rear panel connector.

The AGC amplifier provides a constant output signal which is needed to drive a phase-locked loop system which provides signal conditioning prior to the digital PLL.

The output of this phase-locked loop is applied to a second phase-lock loop. In this loop the 5.8975 MHz signal is divided by 128 and applied to a phase detector. The other input is a 3.6859375 MHz signal divided by 80. The phase detector output after amplification and filtering controls the frequency of the 40.18 MHz VCO. The loop action is such as to reduce to zero the frequency difference between the inputs to the phase detector. When this occurs and the loop is locked, the satellite translation error carried on the pilot is transferred to the 40.18 MHz oscillator in such a sense as to cancel the error in the received carriers in the demodulation process.

Redundancy Operation. (Refer to Figure 15). The incoming receive input containing both the pilot and channel carriers is split in a passive splitter and the outputs are fed to the two pilot receivers. If the pilot receivers are operational and phase lock has been acquired by both, the control logic places the output of the receiver A on line. If receiver A loses lock, but receiver B retains lock, then the ouput of receiver B is placed on-line. If both receivers lose lock, then no switching occurs.

The control and switching circuitry is mounted on the motherboard. The 40.18 MHz output level is set to +12 dBm and is filtered to reduce harmonics.

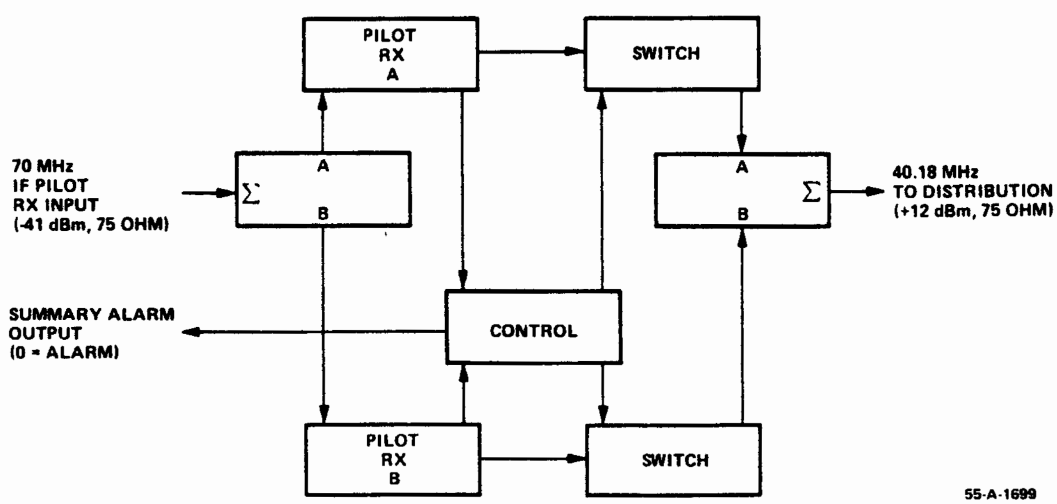


Figure 15. Model 7710 Pilot Receiver System—Redundant Operation

Channel Unit Equipment

General. This section describes the Series 8300 SCPC voice channel equipment, which consists of:

- Voice Processor
- Dual-frequency synthesizer; no AFC
- FM Modem

The channel unit accepts VF signals from the terrestrial network, processes these signals through filters, compressors and preemphasis, and modulates the desired transmit carrier in the 52 to 88 MHz IF band. On the receive side, it accepts a composite IF signal, selects and demodulates the required carrier, and processes the resulting VF signal through deemphasis, expander and band-limiting filtering.

The principal features of the channel equipment are:

- Threshold extension demodulator
- Excellent audio channel flatness and delay distortion
- Excellent adjacent channel rejection
- Accurate compander tracking over a wide range
- Fast-operating VOX detector over the full dynamic range
- Fast synthesizer locking (for DAMA applications)
- Direct channel readout (for pre-assigned operation)
- Dual synthesizers for independent transmit/receive frequency
- Linear limiting for deviation control, to prevent distortion
- Plug-in modularity.

Other features are listed for each of the modules, along with the discussions which follow.

The channel unit modules are housed in a channel equipment card cage, which provides module interconnection as well as IF and reference signal splitting and distribution.

Model 8324 Voice Processor (Figure 16). The purpose of the voice processor is to provide all necessary audio processing and control functions. The voice processor sets the overall voice frequency bandwidths, provides signal delay and gain, emphasis and companding. Echo suppression is also accomplished in the Model 8324 Voice Processor.

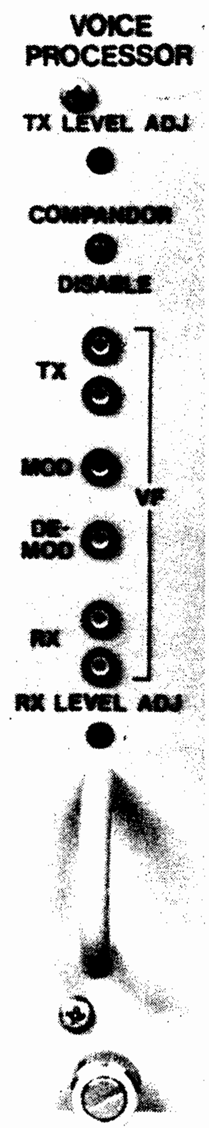


Figure 16. Voice Processor Controls, Indicator, and Monitoring Points

The voice processor features include:

- 2:1 companding (compatible with CCITT G162)
- Emphasis
- Choice of order for emphasis and companding
- Echo suppression

- Data over voice option
- Linear limiting
- Voice-operated carrier control switch (VOX)

The voice processor is composed of a motherboard, an audio delay board, a receive rectifier and holdover board, and an optional data-over-voice board. (Figure 17).

The printed wiring boards plug into a receptacle on the backplane board at the rear of the channel unit shelf assembly.

The echo suppression option is implemented with the addition of the receive rectifier and holdover board and some components on the motherboard. These components include the level comparator, certain logic components, and those components necessary to implement the 6 dB receive echo suppression pad.

The emphasis and companding circuits can be strapped in any order, as desired. The emphasis, then, can precede or succeed the companding function.

No manipulation of controls is required for the voice processor during normal operation. If a change in RX or TX interface level is required, the 0 dBm0 level for these interfaces can be adjusted by means of the TX LEVEL ADJ and RX LEVEL ADJ controls on the front panel, refer to Photo No. 10103. The 0 dBm0 level for the transmit interface is variable from 0 to -16 dBm. For the receive signal interface, 0 dBm0 is variable from 0 to +10 dBm.

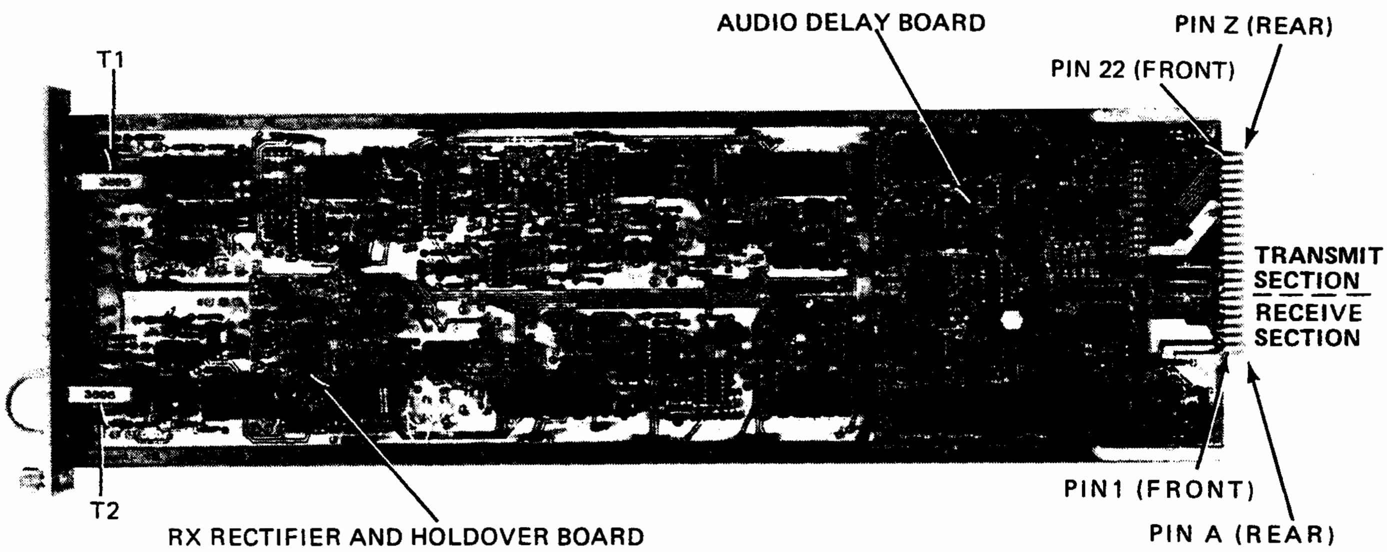


Figure 17. Model 8324 Voice Processor—Interior

Voice Processor Operation. The following paragraphs detail the operation of the Model 8324 Voice Processor.

a. Transmit Section. The transmit circuitry is shown in the top half of the block diagram (Figure 18). Signal flow is from left to right.

b. Input Amplifier. The input amplifier, which has a gain control accessible from the front panel (TX LEVEL ADJ), permits the 0 dBm0 level to be varied from 0 to -16 dBm.

c. Two-Pole Low-Pass Filter. After the input amplifier, the signal enters a two-pole, 20 kHz, low-pass filter where high-frequency signal components are removed.

d. Three-Pole High-Pass Filter. A three-pole Chebyshev high-pass filter follows the low-pass filter. The 200 Hz, high-pass filter establishes the lower band edge of the voice processor and rolls off at the rate of 18 dB per octave. Ripple in this filter is 0.15 dB.

e. Voice-Operated Switch (VOX) System. The VOX system uses a sample of the transmit audio signal taken at the output of the three-pole high-pass filter. This system compares the peak audio level to a preset reference and provides that result as one input (VOX command) to the logic that controls the carrier. (The comparator output is high if the transmit audio level exceeds the reference.)

   The Transmit Disable and Transmit Enable inputs take precedence over the VOX command when they are both held high.

   If the carrier has been activated by the VOX signal, the carrier will be held on after the cessation of audio for an adjustable period determined by the holdover circuit. This holdover time is nominally 125 ms.

f. Compressor. The compressor is designed for unity gain at 0 dBm0 and provides a compression ratio of 1 dB output for each 2 dB variation of input signal. (The 0 dBm0 point is defined as that point at which the output level is equal to the input level.) This compression ratio is maintained down to a level -50 dBm0 at the input of the voice processor. Input signals more than 50 dB below 0 dBm0 are not compressed.

   The compression may be disabled via the Compander Enable input of the voice processor. A logic low at this terminal will clamp the compressor to unity gain and eliminate the compression function.

g. Preemphasis. After the compressor, the signal is passed into the preemphasis circuitry, which has unity gain at 1 kHz and breakpoints at 600 Hz and 5 kHz. From here, the preemphasized signal is routed to the compressor.

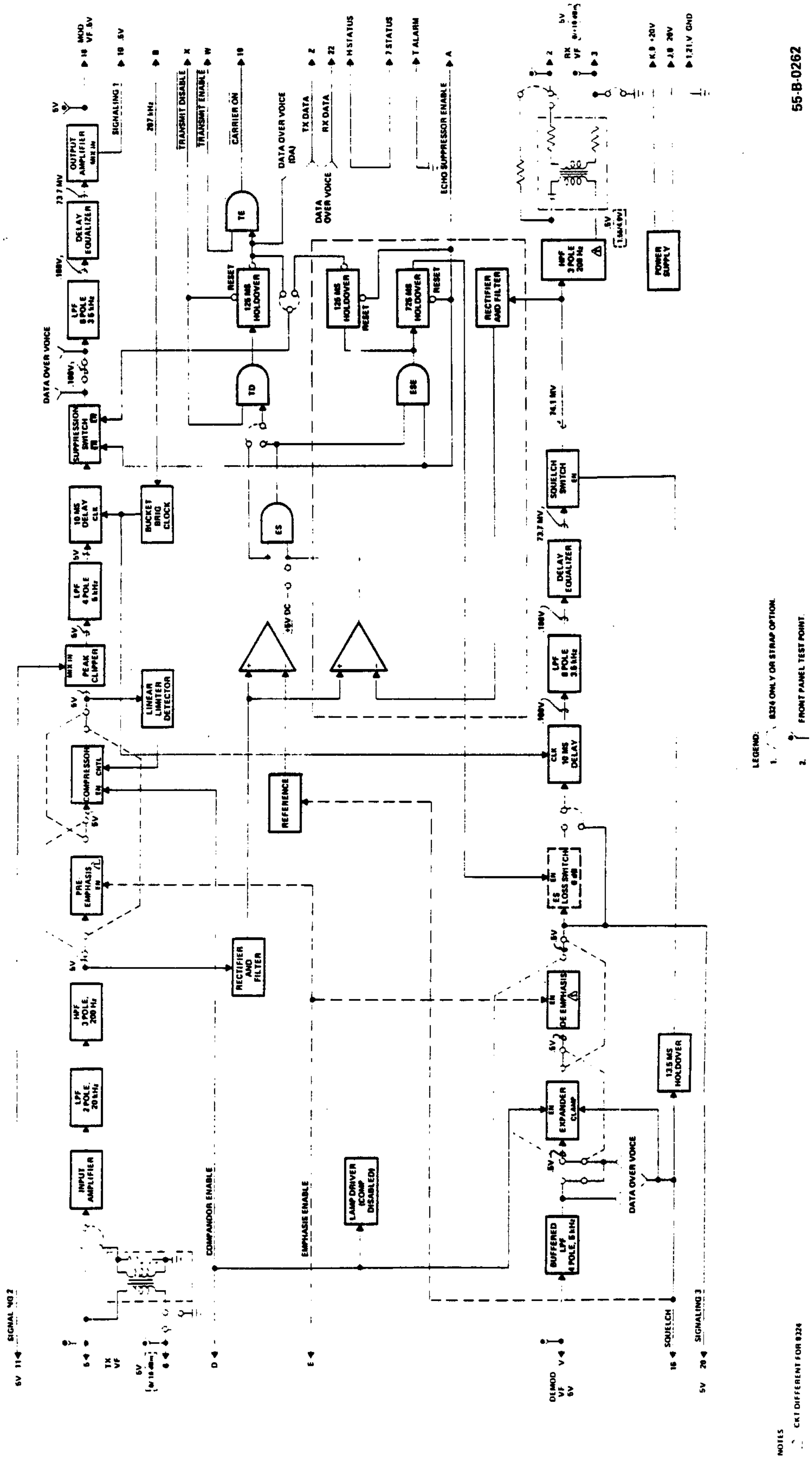


Figure 18. Model 8324 Voice Processor Block Diagram

h. Emphasis Switching. Analog switches for the preemphasis and deemphasis circuits are arranged such that when the Emphasis Enable input to the voice processor is brought low, the frequency selective components of the preemphasis and deemphasis blocks are moved, thereby changing these elements into unity gain amplifiers.

i. Linear Limiter and Peak Clipper. In order to avoid overdeviation, the transmit side of the voice processor is required to limit the maximum level of signal sent to the modulator. The linear limiter detector is a circuit which operates with the gain control element in the compressor and is designed to reduce the gain of the voice processor when signals exceeding +4 dBm0 at the transmit input are present. While this device eliminates overdeviation without introducing distortion, it cannot act instantaneously. Thus, in addition, a peak clipper is cascaded after the compressor to prevent initial peaks from passing to the modulator. The linear limiter begins its action at a level of +2 dBm0 while the peak clipper acts a +2.5 dBm0.

   An additional input is brought from the edge connector of the voice processor and mixed in after the compressor at the input to the peak clipper. This port is designed for use with the terrestrial interface unit and takes advantage of the 10 ms delay circuitry while bypassing the preemphasis and compression functions. This port is not used in Model 8324, since it does not have a terrestrial interface unit.

j. Four-Pole, Low-Pass Filter. After the peak clipper, the signal passes through a four-pole, 5 kHz, Chebyshev low-pass filter provided to eliminate high-frequency signals which might mix with the clock frequency of the 10 ms delay circuit. The filter has unity gain and ripple of 0.15 dB or less. Rolloff is at a rate of 24 dB per octave.

k. 10 ms Audio Delay Circuit. Sampling and storage is performed with a bucket brigade device opeating with a 24 kHz dual clock. This results in a 48 kHz sample rate and, with the 1,024 storage elements, provides a 10.67 ms delay. The bucket brigade device has a nominal insertion loss of 8.5 dB, which is made up later in the voice processor. The clock for the delay unit is developed from an external 288 kHz reference signal. This reference signal is provided to the voice processor at a level of 1V peak-to-peak. It is then buffered, divided in frequency by 12 and used to drive the bucket brigade device. If the data-over-voice option is used, it is inserted between the 10 ms delay and the six-pole, low-pass filter.

l. Suppression Switch. Echo suppression is implemented on the transmit side of the processor with a digitally controlled gate. The gate enables the audio path whenever either of the two control lines is brought low. One line is driven from the echo suppressor logic and the other is used to disable suppression. When disabled, the gate introduces more than 60 dB of loss.

m. Six-Pole, Low-Pass Filter. The six-pole, low-pass filter is used to establish the upper band edge of the voice processor and also to filter out the clock noise of the bucket brigade. This filter is a Chebyshev design with a rolloff rate of 36 dB per octave and a ripple of 0.15 dB or less. The filter is designed for unity gain.

n. Delay Equalizer. The six-pole, low-pass filter is cascaded with a delay equalizer, which has a nominal loss of about 8 dB. The delay equalizer minimizes group delay distortion in the transmit channel to allow high speed data to be sent through the voice processor.

o. Output Amplifier. The 74 mV signal from the output of the delay equalizer passes to the output amplifier and is boosted to 0.5V rms.

p. Receive Section. The receive circuitry is shown in the bottom half of the block diagram (Figure 18). Signal flow is from left to right.

q. Four-Pole, Low-Pass Filter. The signal from the demodulator is brought in at a level of 0.5 rms (0 dBm0) and is buffered and passed through a four-pole, 5 kHz, low-pass filter. This filter is identical to the one on the transmit side and is used to reduce the noise on the unprocessed demodulated audio. A connection to the optional data-over-voice circuitry is provided after this filter. If the data-over-voice option is not present, this connection is bypassed.

r. Deemphasis. After the expander, the signal is passed into the deemphasis circuitry, which has unity gain at 1 kHz and breakpoints at 600 Hz and 5 kHz.

s. Expander. The expander, which has a 1 dB to 2 dB expansion characteristic, reestablishes the dynamic range reduced by the compressor on the transmit side. The expander may be enabled and disabled with the Compander Enable signal. If disabled, it is fixed to unity gain and operates as a linear amplifier. Also provided on the expander is a clamp input which operates from the squelch input to the voice processor. This function mutes audio to the control circuit of the expander and prevents noise which is present in the absence of a carrier from loading the control circuit. The variable gain element of the expander is simultaneously disconnected from the control circuit and clampd to unity gain.

t. Echo Suppression Loss Switch. On the receive side echo suppression is supplied with a 6 dB digitally-controlled pad. A logic high inserts the 6 dB loss.

u. 10 ms Audio Delay. The audio is then sent to the 10 ms delay circuitry. This delay is identical to that used on the transmit side and serves to prevent noise which is present at the end of the transmission from getting through the voice processor before the squelch can be activated.

v. Six-Pole, Low-Pass Filter. After delay, the audio signal is passed through a six-pole, 3.5 kHz, low-pass Chebyshev filter identical to that used on the transmit side. This filter establishes the noise bandwidth of the receive side and also eliminates the clock noise associated with the bucket brigade delay. The filter is delay-equalized for data transmission.

w. Squelch. From the low-pass filter, the audio signal is routed through the squelch switch, which is used to eliminate most of the noise present when a carrier is absent from the demodulator. Some of the noise, however, is allowed to pass through the squelch switch, and thus simulate the background noise which is present when the carrier is up.

The squelch signal from the demodulator is not used directly to operate the squelch switch. It is first delayed approximately 13 ms to allow previously received noise to empty out of the bucket brigade before the squelch switch is opened. The squelch signal is only delayed in the transition from carrier off to carrier on. In the opposite direction, the squelch switch is operated immediately.

x. Three-Pole, High-Pass Filter. After squelching, the audio signal is passed through a three-pole, high-pass, 200 Hz Chebyshev filter which establishes the lower band edge of the receive side and is identical to the three-pole filter used on the transmit side. The gain of this block may be adjusted to set the 0.5V rms 0 dBm0 receive output interface level. This gain control (RX LEVEL ADJ) is accessible from the front panel.

y. Lamp Driver. The lamp driver is provided to illuminate a yellow LED on the front panel of the voice processor when the compander is disabled.

Model 8320 FM Modem (Figure 19). The purpose of the FM Modem is to perform the frequency modulation of the channel carrier, using the processed transmit VF signal from the voice processor, and to select and demodulate the receive channel carrier from the composite received IF spectrum.

The FM Modem consists of a motherboard and four plug-in printed wiring board assemblies, a crystal filter, buffer amplifier, and voltage regulator circuits (Figure 20). Three of the plug-in assemblies (Demodulator Mixer, Filter/AGC, and TED) and the Crystal Filter perform the demodulation function. The buffer amplifier and the fourth plug-in PWB assembly (Modulator) perform the modulation function. The voltage regulators regulate +20V dc and -20V dc to +15V dc and -15V dc for all the circuits requiring dc power.

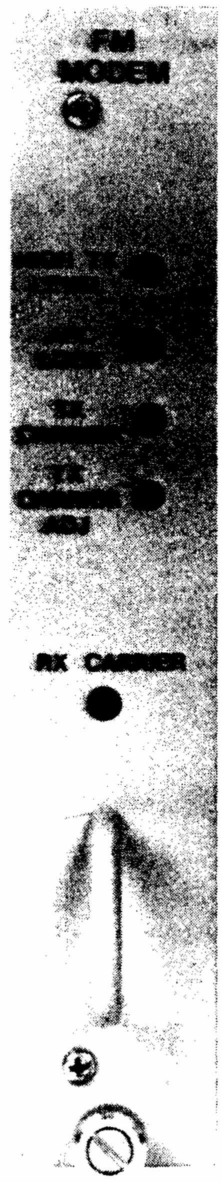


Figure 19. FM Modem, Indicators

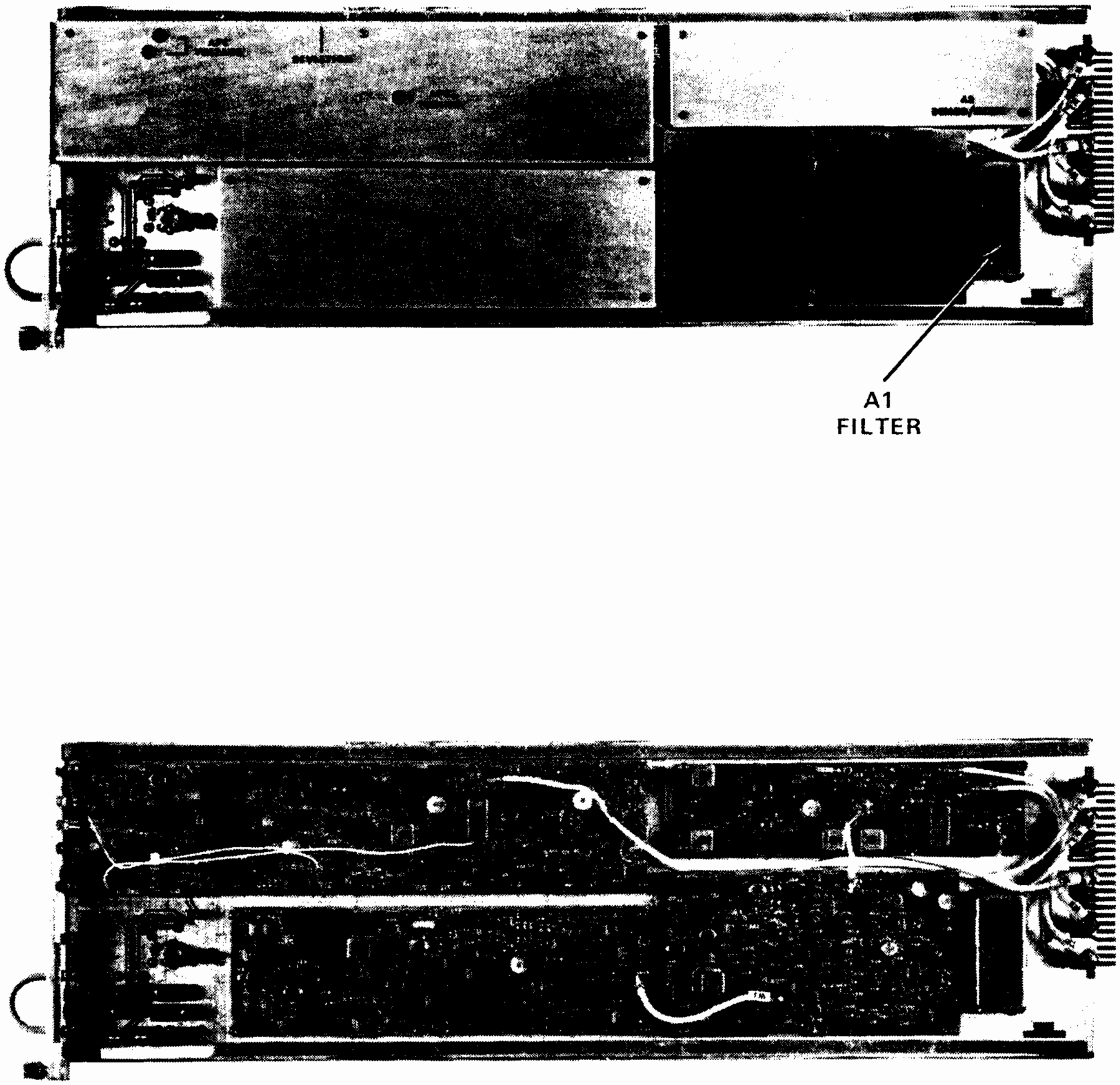


Figure 20. FM Modem—Interior

The FM modem features include:

- Threshold extension
- Excellent adjacent-channel rejection
- Deviation sensitivity independent of transmit channel frequency
- High TX carrier level stability through ALC circuitry
- Good temperature compensation over the range 0° to 50°C
- Wide-range squelch operation
- Low distortion
- Two-level (switched) transmit IF output

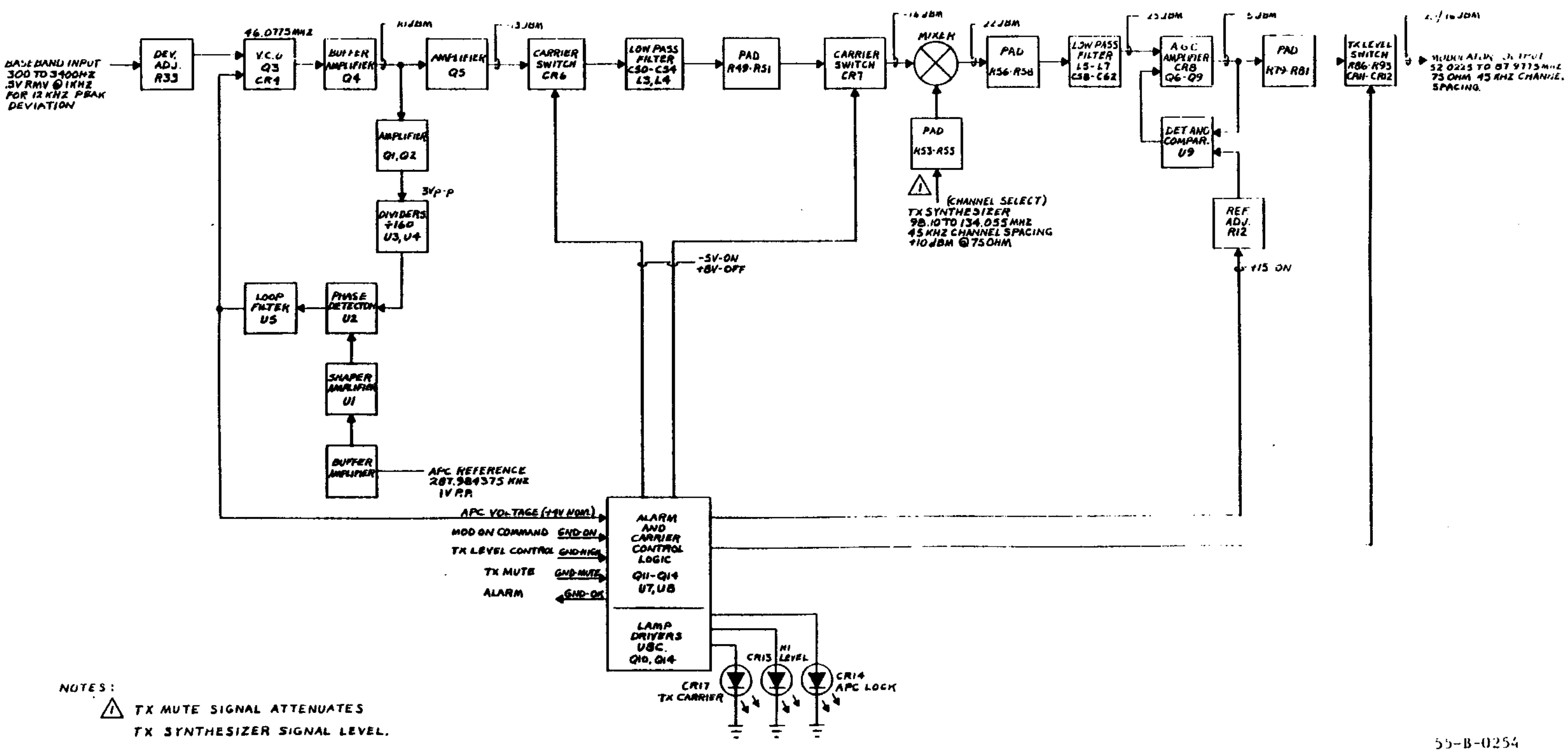
FM Modem Operation. The following paragraphs detail the operation of the Model 8320 FM Modem.

a. Modulator (Refer to Figure 21). The modulator consists of a VCO which contains a single varactor for both modulation and automatic phase control. The VCO operates at 46.985 MHz and is divided to 288.03125 kHz for phase lock to the 288.03125 kHz reference signal furnished from the system Reference Frequency Unit.

   The 46.085 MHz IF signal is amplified and filtered before channel selection occurs via a mixing process with the TX synthesizer. The output of the mixing process is filtered to remove the image frequency while passing the desired signal in the range of 52 to 88 MHz. The desired signal then passes through an AGC amplifier which maintains extreme flatness across the 52 to 88 MHz IF band. The output section of the modulator consists of the TX level switch (controlled by DAMA or manual operation) which allows the IF signal level to be boosted by 4 dB.

   The carrier control logic consists of logic gates, transistor switches, and pin diode switches. The carrier is normally controlled by the MODEM ON command; however, in the event of a TX synthesizer failure (TX mute) signal or an APC failure (Modem Alarm) signal, the carrier control logic will turn the carrier OFF regardless of the state of the MODEM ON command.

b. Demodulator (Refer to Figure 22). The input to the demodulator is low-pass filtered before channel selection occurs via a mixing process with the RX synthesizer. The output of this mixing process (45.965 MHz) is amplified and filtered before it is again mixed (with 40 MHz) to create the final IF frequency of 5.965 MHz. The 5.965 MHz signal is narrow-band filtered and AGC'd. Also, the squelch information is derived at this frequency by detecting the out-of-band noise present on the IF signal. The AGC'd IF signal is then presented to the threshold extension demodulator which utilizes a regenerative voltage-controlled feedback loop around a limiter. The signal is demodulated and presented to the Voice Processor for signal conditioning.



NOTES:

1. TX MUTE SIGNAL ATTENUATES TX SYNTHESIZER SIGNAL LEVEL.

55-B-0254

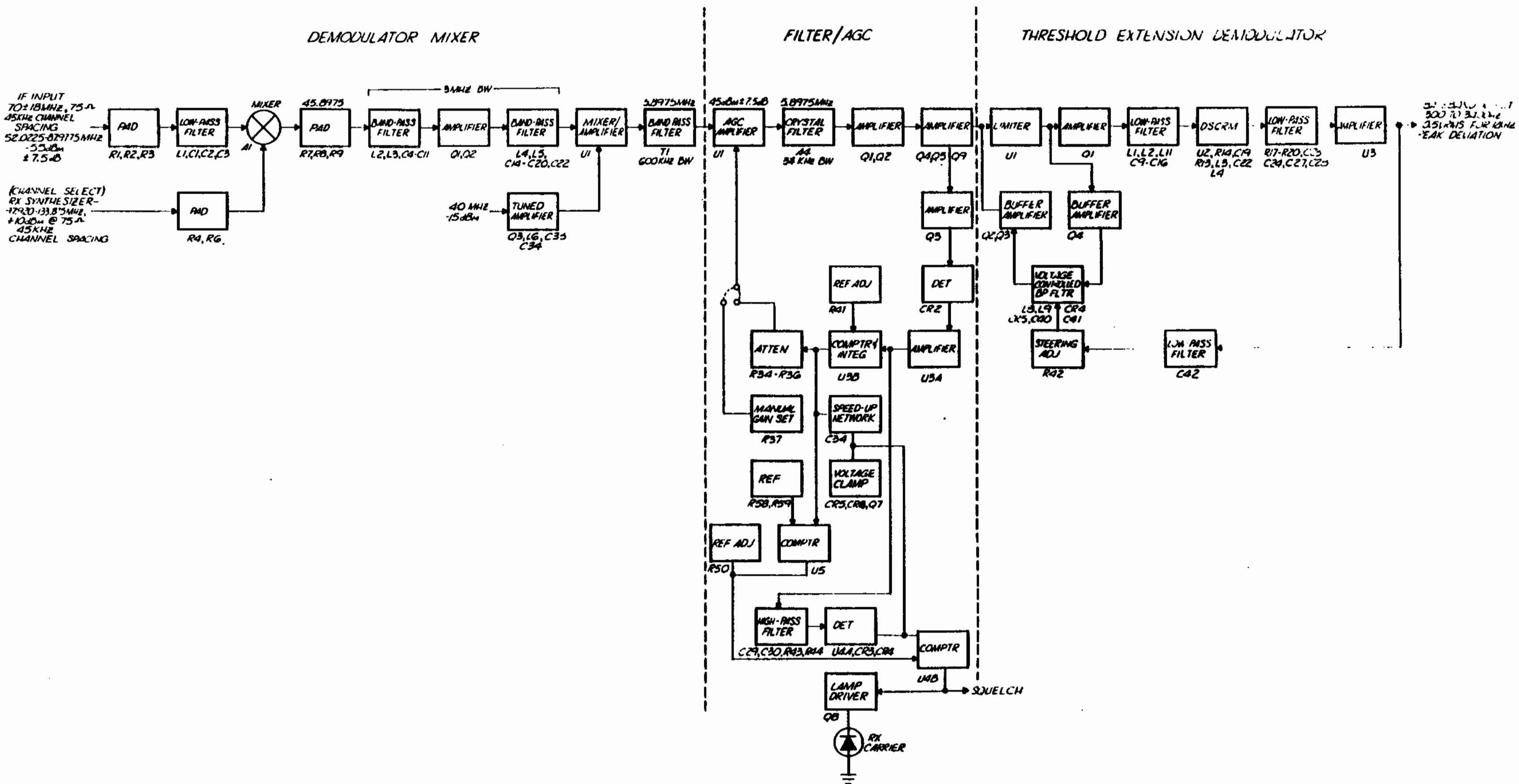


Figure 22. Series 8300 Demodulator Functional Block Diagram

Model 7740 Frequency Synthesizer. The Model 7740 Frequency Synthesizer is a microprocessor-controlled unit which performs the following functions: (See Figure 23, Synthesizer Block Diagram.)

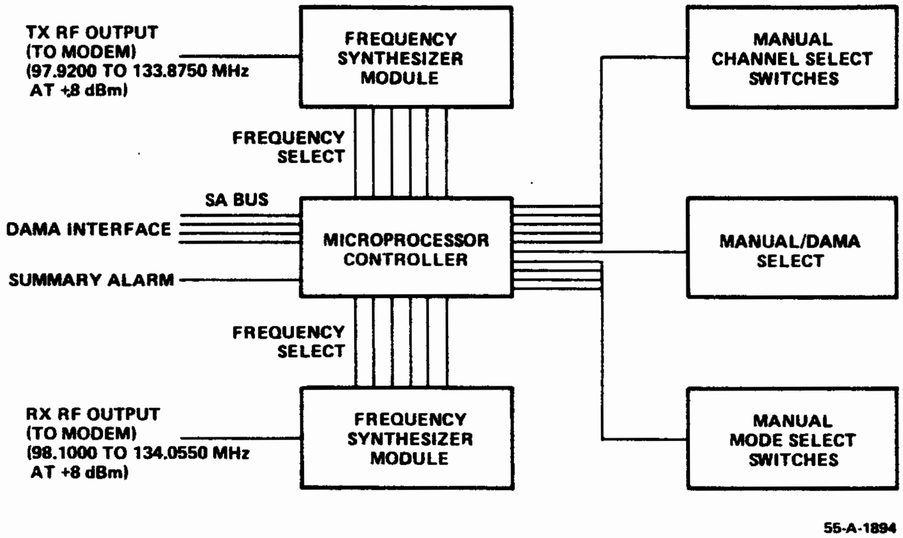


Figure 23. Synthesizer Block Diagram

- Provides a synthesized low-noise local oscillator for the transmit side of the channel unit. The signal's frequency range is 98 to 134 MHz and output level is +8 dBm.

- Provides a synthesized low-noise local oscillator for the receive side of the channel unit. The signal's frequency range is 98 to 134 MHz and output level is +8 dBm.

- Provides summary alarm output for the channel unit.

- Provides control of various functions in the channel including the following:

  a. Pre- and Deemphasis enable/disable

  b. Echo suppressor enable/disable

  c. Transmit enable/disable

  d. Compander enable/disable

- Provides for DAMA control via the SAbus interface in the unit. The SAbus also allows for remote control of the channel unit and extension of alarms to a central unit if required.

- Provides front-panel LED indication of phase lock on both synthesizers.

The unit is a plug-in module 7 inches high and 21 inches long.

Frequency Synthesizer Operation. The following paragraphs detail the operation of the Model 7740 Frequency Synthesizer.

a. Synthesizer. Figure 24 is a block diagram of the synthesizer module used in the unit. The Voltage-Controlled Oscillator (VCO) is controlled by a phase-locked loop. A portion of the VCO output is counted down in a programmable counter and compared to a synthesizer reference in the phase detector. Any phase difference causes a change in frequency or phase of the VCO until the loop is locked. The main output of the VCO is disabled whenever the loop is not phase locked. The loop is designed to achieve rapid acquisition and phase lock for DAMA operation. The synthesizer frequency is controlled by entering the channel number (via manual switches) in the microprocessor. The microprocessor then loads the required division into the programmable counter. The synthesizers may also be remotely controlled via the SAbus. The step size between channels is variable and can be 22.5 kHz, 30 kHz or 45 kHz. The unit is also capable of one-half channel offset in the 30 kHz and 45 kHz modes. This mode of operation is effective in lowering intermodulation energy falling in the channel unit passband.

b. Manual Control. The manual control switches Synthesizer Block Diagram perform the following functions: (See Figure 23, Synthesizer Block Diagram.)

- Set manual enable/disable
- Set the RX channel assignment
- Set the TX channel assignment
- Set emphasis enable/disable
- Set echo suppressor enable/disable
- Set transmit enable/disable
- Set compandor enable/disable

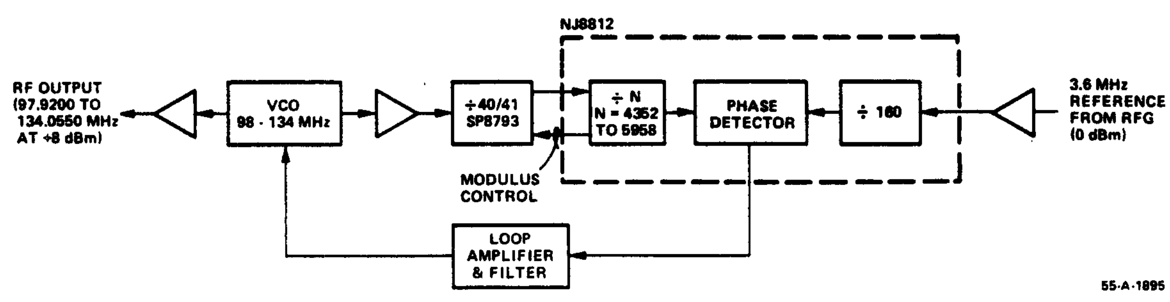


Figure 24. Synthesizer Module Block Diagram

c. Alarms. The synthesizer extends any channel unit alarms after summary. The synthesizer alarms are:

- RX Synthesizer Phase-Lock Alarm
- TX Synthesizer Phase-Lock Alarm

d. DAMA Interface. The synthesizer interfaces with the DAMA system via the SAbus. In DAMA operation all of the functions already described may be controlled by the DAMA system. The synthesizer microprocessor carries out the DAMA commands and report successful accomplishment of all such commands.

Model 8331 Channel Unit Card Cage (Figure 25). The purpose of the Model 8331 Channel Unit Card Cage is to provide a housing and all necessary interconnecting circuitry for four channel units. Each channel unit consists of a Voice Processor, and FM Modem, and a Frequency Synthesizer. One extra slot, located on the extreme right-hand side, is provided to accept an additional module if required.

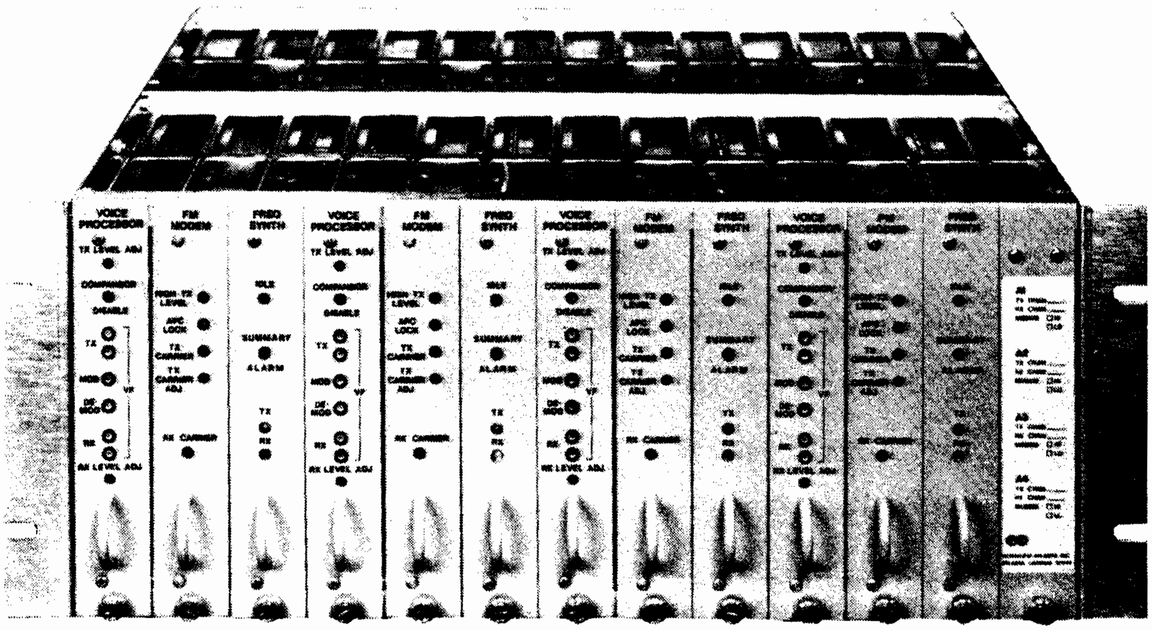


Figure 25. Channel Unit Shelf Assembly

A four-layer backplane provides mating connectors and interconnecting circuitry for each of the channel unit modules. Another printed circuit board, mounted on a hinged panel at the rear of the assembly provides the splitting and combining functions for the channel unit reference and IF signals. These signals are distributed to each channel unit via discrete coaxial cables and the backplane edgecard connectors.

All audio, control, reference frequency, and IF interfaces are available on connectors at the rear of the unit.

The Channel Unit Card Cage provides all the necessary connections between the VF Processor, the FM Modem, and the Frequency Synthesizer, as well as all external connections for power input, controls, audio input/output, reference signal input, alarms and IF input/output.

Since the channel unit cage itself is designed for electrical interconnect and distribution of signals only, no active devices are employed.

Channel Unit Card Cage Operation. The DAMA input is a parallel data bus provided for external digital frequency assignment and control of the channel units. Connection is made to each of the four synthesizer locations.

The alarm bus connector provides a status alarm loop through each channel unit. The loops may be wired externally for either individual or summarized alarm functions.

The Audio Input/Output connector is the interface for terrestrial baseband input to (and output from) the Voice Processor.

The Channel Unit control connector provides for direct external control, by means of logic levels, of such channel unit functions as transmit enable/disable, echo suppression, emphasis, and companding.

The Receive/Transmit Data connector provides one transmit and one receive path per channel unit Voice Processor for data operation.

The dc power supply inputs for +20 and -20 volts dc is on a single multi-pin connector.

The 70 MHz IF outputs from the four channel unit modems are summed by a power combiner and appear at a single connector.

The incoming 70 MHz Receive signal is connected at a single connector and distributed to each channel unit modem via a power divider.

The 40 MHz Demodulator Reference from the Pilot Receive Unit is also a single connection. The signal is split by a power divider and distributed to each channel unit modem.

The 3.6 MHz Synthesizer Reference from the RFG is connected at a single connector and distributed to each channel unit synthesizer by a power divider.

A 288 kHz Demodulator Reference generated by the RFG is connected to a single connector and applied to the primary winding of a transformer. The transformer is used to step-up the reference level and to provide impedance matching. From the secondary winding of the transformer, the stepped-up 288 kHz signal is fed in parallel to each FM modem and each voice processor.

Digital Earth Station Equipment

This paper describes equipment typically provided with a digital earth station. Equipment consists of both high frequency and digital units. Since the high-frequency equipment is described elsewhere, the major emphasis is on digitally related equipment.

Station Description

Equipment for a digital earth station is comprised of digital equipment as well as high frequency equipment. A typical earth station equipment complement is shown in block diagram of Figure 1. Basically, high frequency equipment is located in a shelter in close proximity to the antenna. Digital equipment is rack-mounted in the customer facilities.

Signal flow through the station is as follows: assume that a number of digital signals are transmited on one transponder with each digital signal modulating a separate carrier. The received signal is amplified by a low-noise amplifier (LNA) located at the receive port of the antenna feed. This signal is then downconverted to a frequency range where it can be handled more conveniently. The downconverter output is fed to the digital equipment via coaxial cable and coupled to an appropriate set of demodulators. Each demodulator is tuned to the desired receive carrier by an internal crystal or synthesizer. The digital output of the demodulator is then typically passed to the customer via a protection switch. The switch provides for backup demodulation equipment in the event of an on-line equipment failure. See "Theory of Digital Communications" paper for further detailed explanation.

In like manner, the transmit digital signals are fed via the protection switch to the modulators. The modulator transmit frequency is selected by either an internal crystal or synthesizer. The output of each of the modulators is combined and fed via coaxial cable to the upconverter. Here the signal is converted to the desired frequency for transmission to the satellite. The high-power amplifier (HPA) boosts the upconverter output to the desired level for transmission.

It should be noted that the high frequency equipment is usually provided in a redundant configuration. Single-thread equipment is shown for simplicity.

Additional equipment is provided with the indoor units which permit local and remote monitoring of station status as well as control. Remote operation is usually performed over telephone lines to a central station.

High Frequency Equipment

Antennas

The antenna used by a given terminal is dictated by a combination of several conditions. The following conditions are typical major influences on selection of antenna size:

- Conservation of satellite power
- Local interference
- Space availability
- Cost

Typical antennas used with digital terminals are 10, 7, 5, 4.6, and 3 meters in diameter. A 10-meter antenna is shown in Figure 2. Larger antennas require less satellite power and obviously are more expensive. However, circuit costs are dependent on the satellite power used. Thus, the trade-off between initial equipment cost and recurring circuit costs is an important factor in antenna size selection.

In areas with high interference, an antenna with a low profile is sometimes a controlling parameter. Thus, a smaller antenna must be used. In congested areas, such as industrial parks or cities, antennas are often located on roof tops, and this may dictate a small size. In another case, for example, a large number of specialized stations using a single carrier and having low-bit-rate requirements can use very small antennas, thus minimizing station cost. Another application of the small antenna is for low-cost audio distribution.

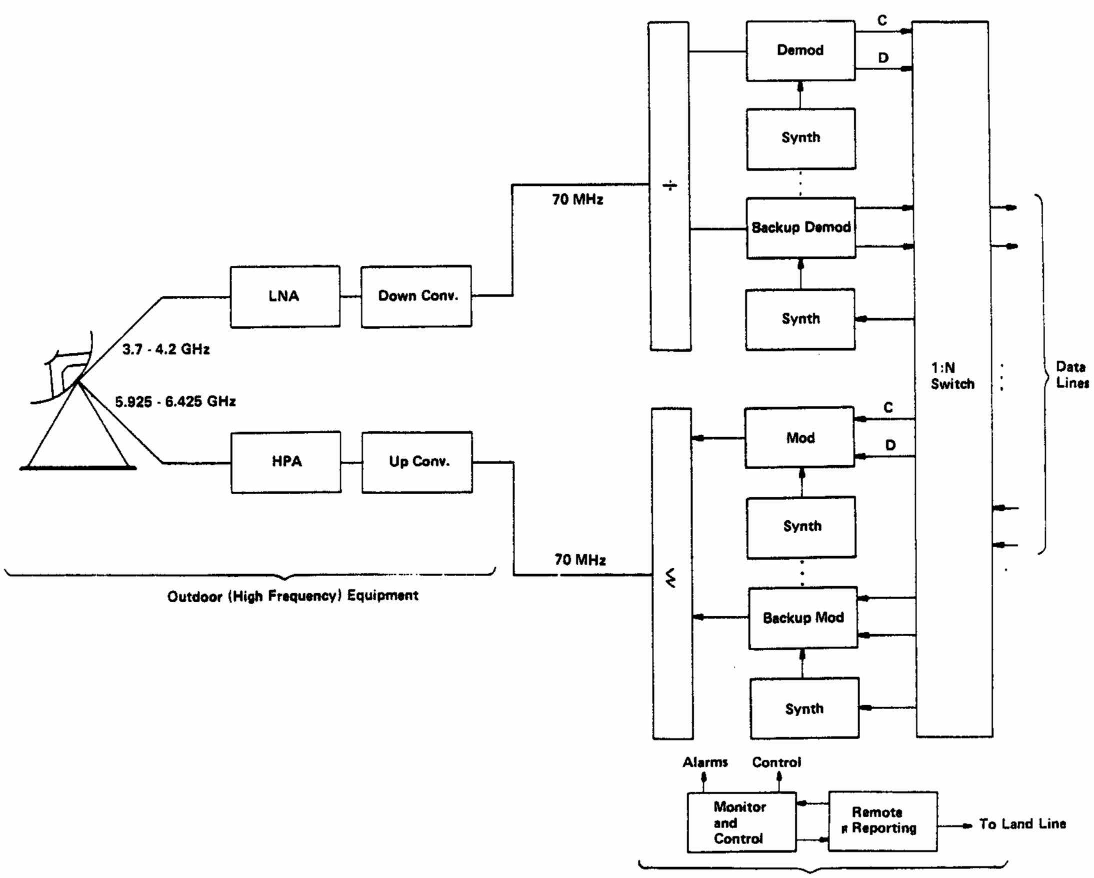


Figure 1. Simplified Digital Terminal Block Diagram

Figure 2. 10-Meter Antenna

Low Noise Amplifier (LNA)

The low-noise amplifier, in conjunction with the antenna, determines the G/T or figure-of-merit of the station. A higher G/T requires less satellite power for given performance characteristics. The quality of a low-noise amplifier is a function of its noise temperature; the lower--the better. Two types of low-noise amplifiers are utilized--the parametric amplifier and GaAs FET amplifier in Figure 3.

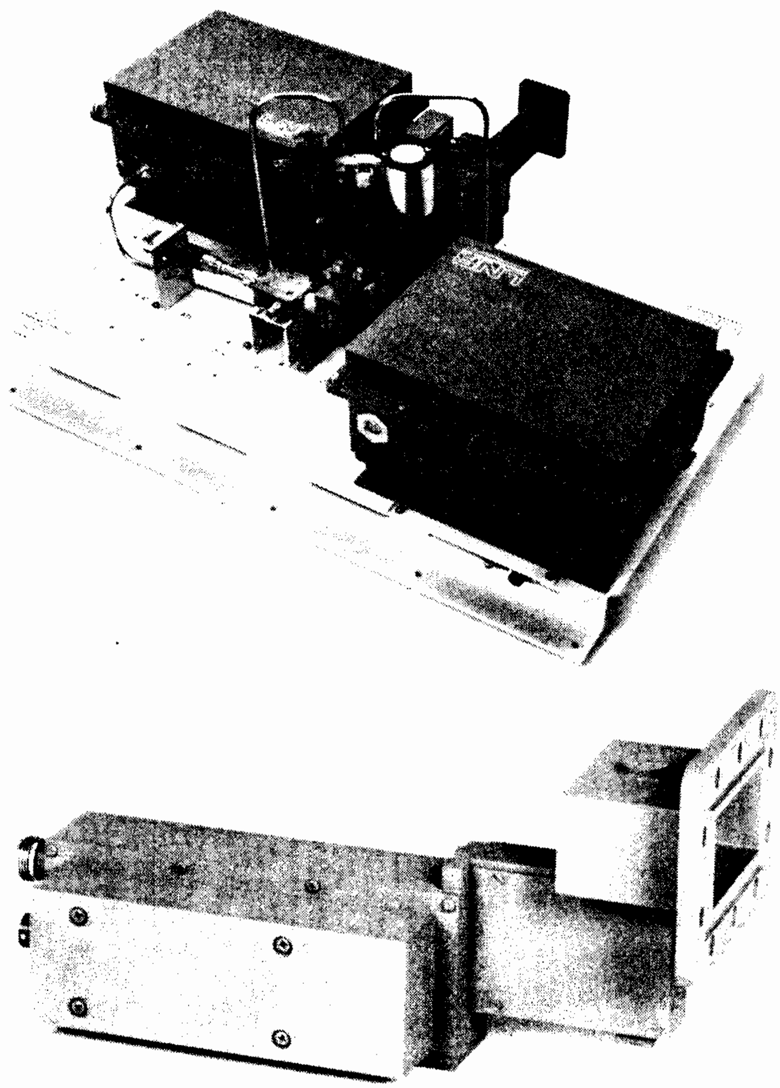


Figure 3. GaAs FET and Parametric Amplifier

Parametric amplifier noise temperatures range from 30K to 70K. The GaAs FET family of amplifiers range in temperature of 80 to 120K. A parametric amplifier is approximately ten times more expensive than the GaAs FET. Thus, the FET is the most common amplifier used with digital terminals.

High Power Amplifiers (HPA)

High-power amplifiers, for digital applications, range in power from 1 watt to 600 watts. (A 75-watt HPA is shown in Figure 4.) HPAs in the 1- to 10-watt range are available in solid-state configurations while the 15- to 600-watt units are traveling wave tubes.

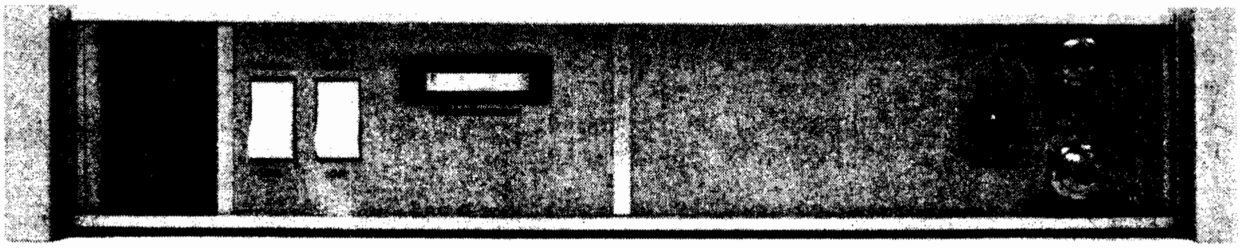


Figure 4. 75 Watt TWT HPA

The HPA size is dictated by the following parameters:

- Transmit antenna size
- Receive station G/T
- Transmit station location relative to satellite
- Number of carriers
- Carrier Bit Rate
- Required receive performance

Converters

Up- and downconverters are available in a dual-conversion configuration. The dual-conversion models permit easy and rapid change in operating frequency. This feature is critical, for example, when rapid restoration of service is required in the event of transponder failure.

The dual-conversion converter is shown in Figure 5. This unit can be tuned to any of six frequencies by a front-panel switch. This converter can also be tuned to any one of the six frequencies remotely.

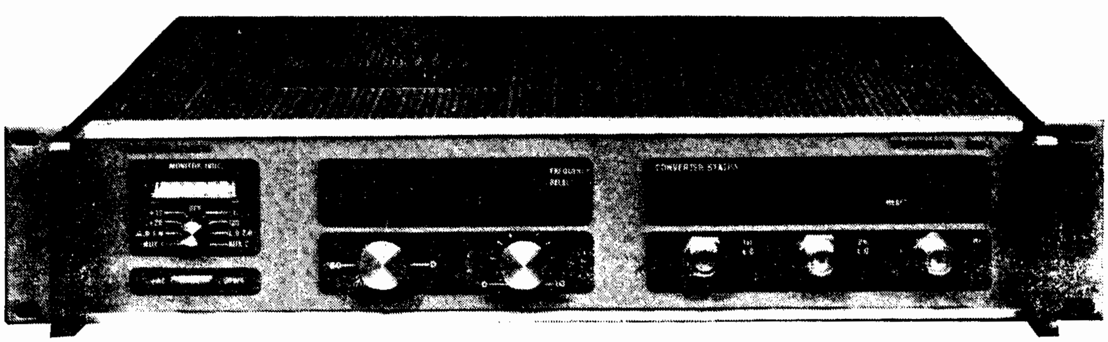


Figure 5. Dual-Conversion Converter

Digital Equipment

General

The digital equipment complement permits the transmission and reception of a number of digital carriers simultaneously. Capability is available to handle from 30 kb/s to 10 Mb/s. The most common equipment in use to date operates at 56 kb/s. Each 56-kb/s data stream can be the output of a conventional multiplexer with input data streams at rates of 32.0, 19.2, 9.6, 4.8 2.4, 1.2, or 300 baud. For example, up to ten 4.8-kb/s data streams can be multiplexed for transmission over one 56-kb/s carrier.

Equipment can be configured with a single modem. Alternatively, a 1:N redundant configuration is frequently provided where high reliability is required. A typical rack arrangement is shown in Figure 6. The rack on the left is a common equipment rack with an expansion rack shown on the right.

The common equipment consists of a monitor and control unit (MCU), baseband patch panel, a 1:N protection switch, an IF patch panel, redundant modulators, redundant demodulators and redundant synthesizers. The racks can be configured to accommodate up to eight duplex data links by adding modem equipment in additional pre-wired equipment racks.

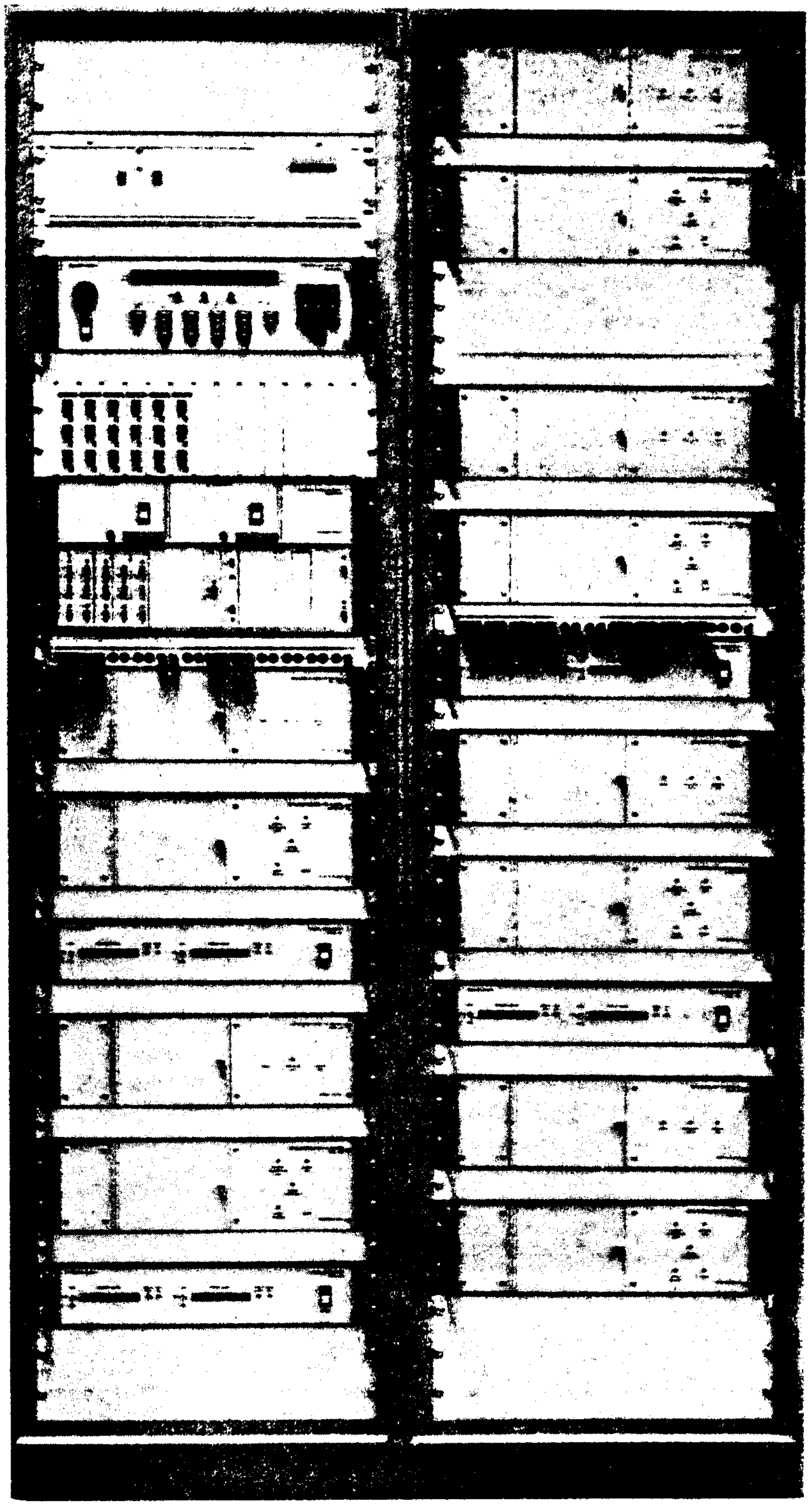


Figure 6. Typical Digital Equipment Rack Layout

MODEM/CODECS

Introduction. Modems are available in both QPSK and BPSK versions. The QPSK version is the most common and is discussed in detail in this section. The Model 8800 series of bandlimited QPSK modems/codecs combines two very desirable properties: bandwidth efficiency and power efficiency. Bandwidth efficiency is achieved through the use of QPSK modulation and Nyquist filtering. Power efficiency is achieved by using QPSK modulation combined with forward-error-correction coding.

These modem/codecs are available in simplex packaging (a transmit drawer and a receive drawer) or in a duplex package (modem/codec in one drawer), thereby allowing the user to optimally configure his equipment.

The data rate range for these units spans 30 kb/s to 10 Mb/s. There are standard data rates such as 56 kb/s, T1, etc.; but other rates are available.

Modulator. A block diagram of the bandlimited QPSK modulator is presented in Figure 7. The encoded data is shifted into the demultiplexer by the 8/7 R clock. The encoded data stream is demultiplexed in two data streams with half-bit delay between them (thereby generating offset QPSK). The two fixed level NRZ data streams with a $(\sin X/X)^2$ spectrum are filtered by identical low-pass filters such that an overall square root Nyquist, $\sqrt{N(W)}$, frequency response is produced. The filtered data streams double-sideband-modulate two quadrature sine waves at the transmit frequency f_1. The modulated outputs are summed to produce bandlimited QPSK. Special circuitry keeps the two outputs equal before summation. Equality is important for generating true QPSK modulation and achieving near theoretical modem statistical performance. The bandlimited QPSK is processed by the bandlimited AGC to produce a constant -10 dBm output from the AGC. Filtering within the AGC loop greatly reduces modulated harmonics of transmit frequency f_1. Since f_1 is directly modulated, no spurious outputs are produced within the transponder bandwidth. A printed circuit (PC) board plug-in attenuator reduces the AGC output to the desired customer output level. A front-panel potentiometer allows one to continuously vary the output level by ±2/3 dB. The front panel of the modulator is shown in Figure 8.

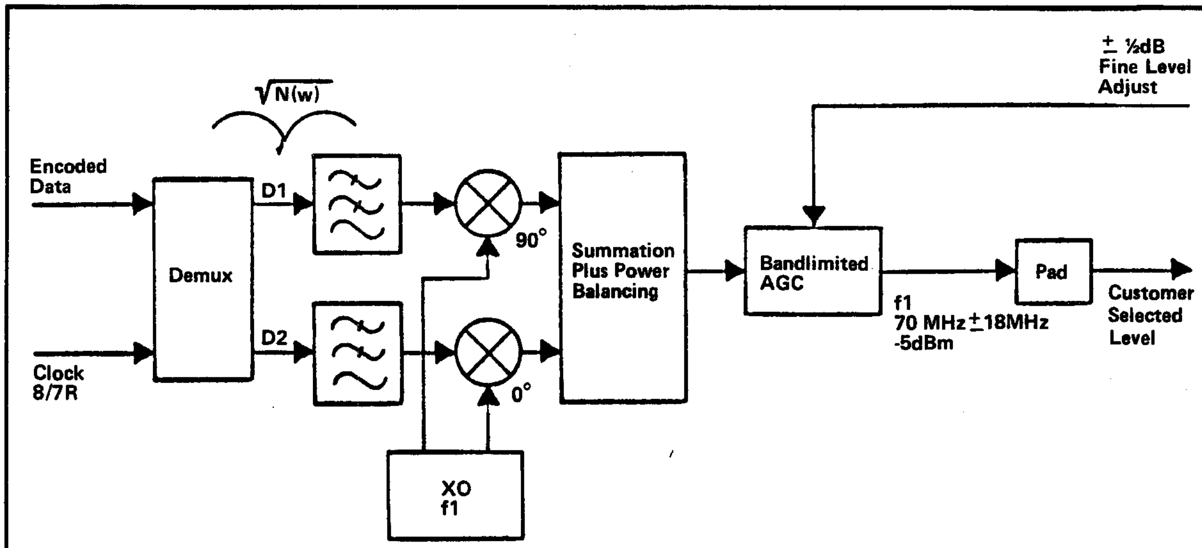


Figure 7. Bandlimited QPSK Modulator

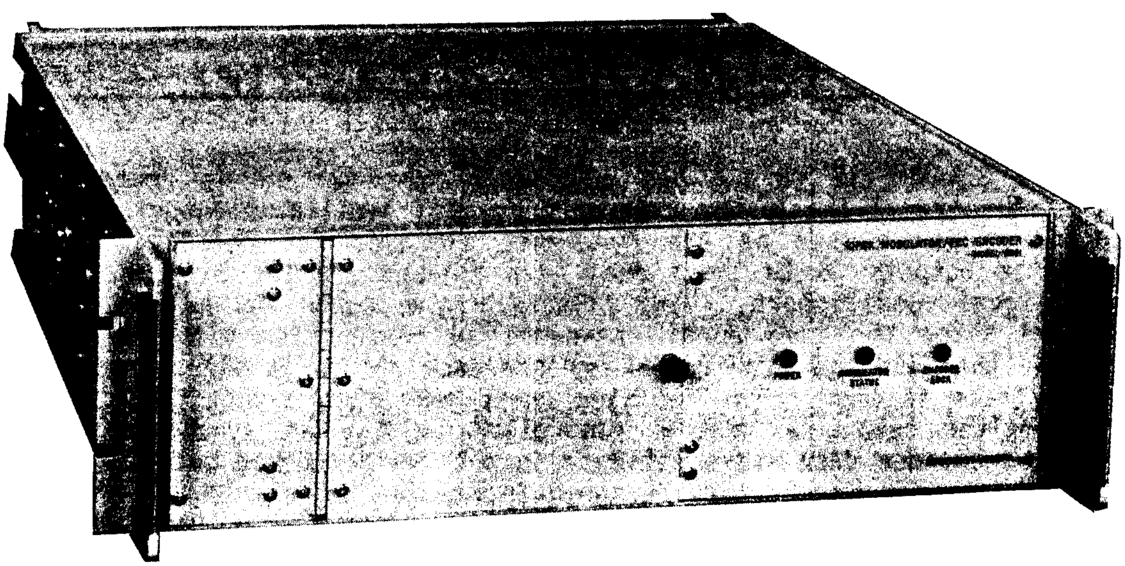


Figure 8. QPSK Modulator FEC Encoder

Demodulator Description. A block diagram of the bandlimited QPSK demodulator is presented in Figure 9. The input downconverter is switched to receive either the assigned receive frequency f_2 or its modulator output center frequency f_1. The input is downconverted to 46 MHz. The 46-MHz input is amplified, filtered, and then downconverted to the second IF of 2.5 MHz by the second mixer and a 43.5-MHz VCO. The 2.5-MHz IF AGC AMP provides two outputs. One, which is bandlimited, drives a X4 (times four) circuit to produce an unmodulated carrier. This regenerated $4f_0$ carrier reference is compared to a 10-MHz crystal oscillator in the phase detector. The phase detector output drives the phase-lock loop filter that controls the 43.5-MHz VCO. This so called long-loop-for-carrier regeneration keeps the narrowband spectrum centered in the demod filters.

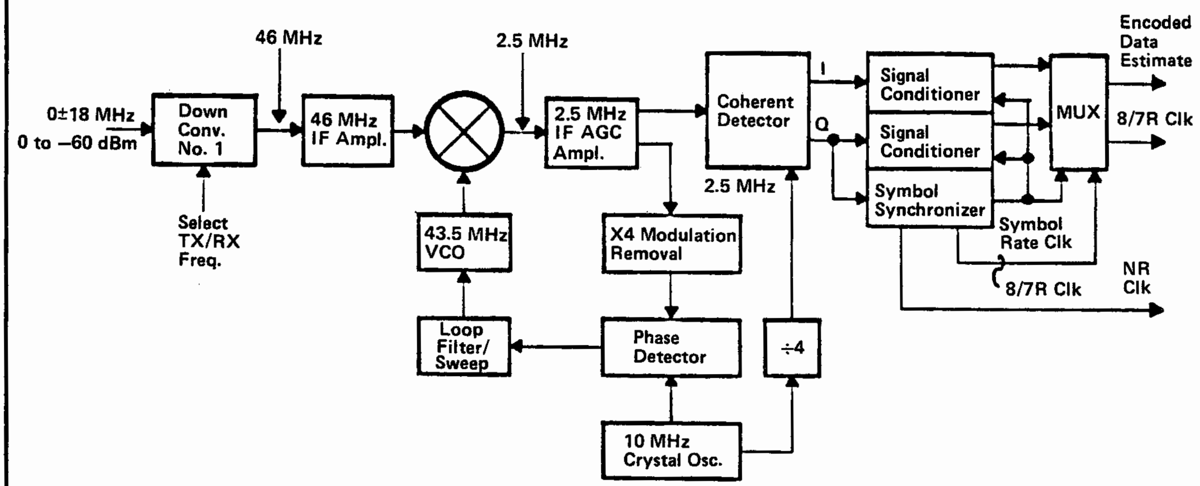


Figure 9. Bandlimited QPSK Demodulator

The other 2.5-MHz IF output has a broader spectrum and is routed to the coherent detector. The 10-MHz reference crystal oscillator output divided by four and proper phased is the coherent reference for demodulation. The demodulated outputs go to the signal conditioners and the bit synchronizer. In the signal conditioners, $\sqrt{N(W)}$ filters remove noise and adjacent channel power and shape the time domain response in conjunction with the transmit filter, so that minimum intersymbol interference occurs at the time each data state estimate is made. The decision units make such a data state estimate on each filter output at a time controlled by the bit synchronizer. The multiplexed encoded data state estimate is the overall demod output. A block diagram of the bit synchronizer is shown in Figure 10.

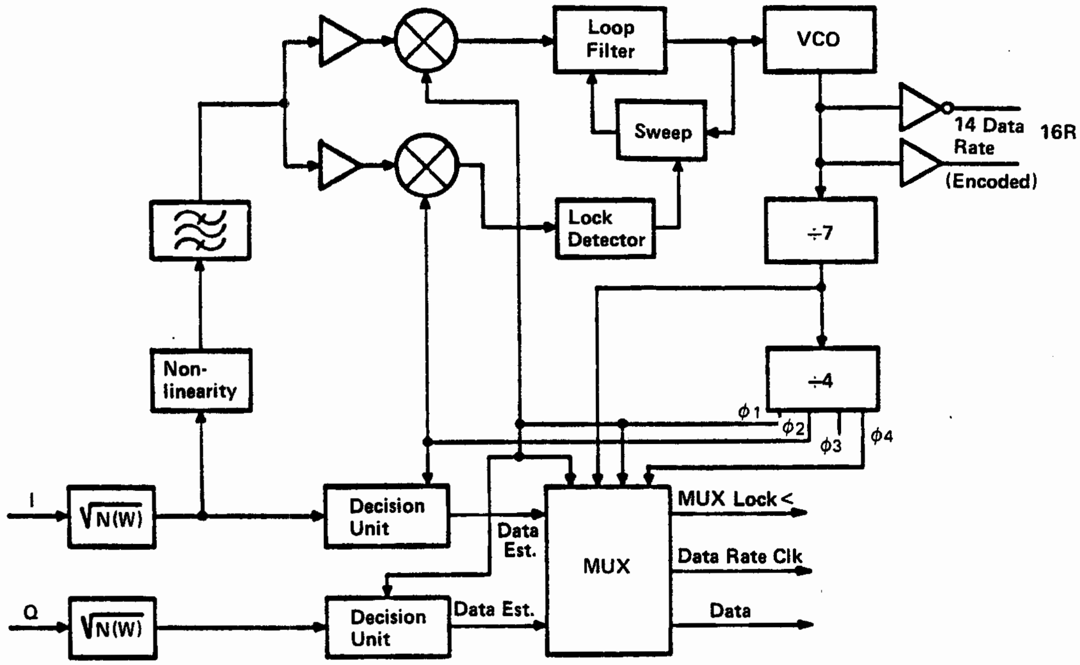


Figure 10. Bit Synchronizer Block Diagram

In the bit synchronizer, one of the coherent detectors is filtered and non-linearly processed to produce a bit-rate spectral component which is bandpass filtered. The bandpass filter output is the reference for a phase-lock loop. The 8/7 bit rate VCO output of the phase-lock loop is routed to the signal conditioner and is also provided as an output. An N-R clock is also provided to the decoder for its internal use.

In the long loop previously described, bandpass limited in the X2 circuitry and adaptive sweep circuitry in the phaselock filter are used to minimize false locking to data-induced spurs if the V.35 scrambling is bypassed. Careful attention has to be directed to such false locking in carrier regeneration circuitry.

The front panel of the demodulator is shown in Figure 11.

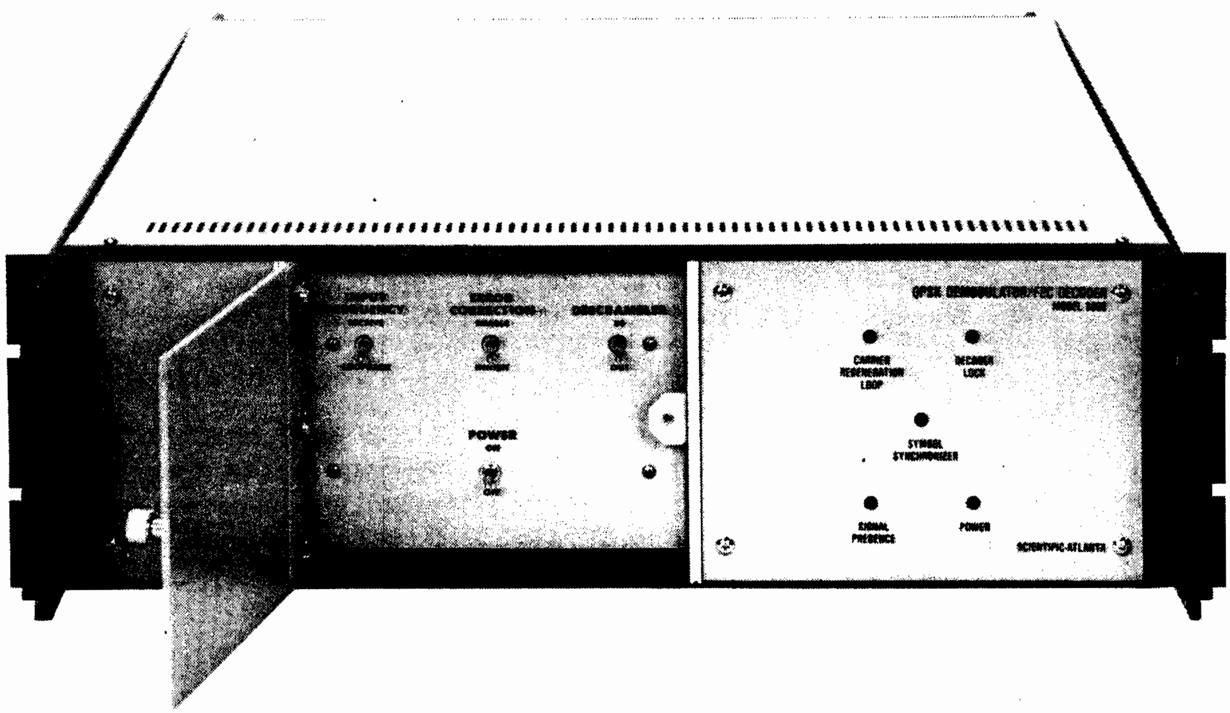


Figure 11. QPSK Demodulator/FEC Decoder

Codec Description. The rate 7/8 FEC codec uses a convolutional self-orthogonal code. The code has a minimum distance of 7 and is triple-error correcting. The constraint length of the code is 1176 bits. The encoder is systematic and one parity bit plus seven information bits are the output in an alternating fashion. The encoder has an internal phase-lock loop that uses the input information rate, R, clock as a reference to generate the output, 8/7 R, symbol rate clock, and all internal clocks that are required.

The decoder uses threshold decoding with syndrome feedback. The decoder resolves the lockup ambiguity of the QPSK demodulator as well as achieving its own branch synchronization. The decoder receives an N x 8/7 R clock from the demodulator which is divided to generate all its internal clocks and the output bit rate, R, clock. The correction capability of the FEC codec is such that for random uncorrelated errors an input BER of 10^{-3} is converted to an output BER of $2x10^{-6}$, input BER of 10^{-4} to output BER of $1x10^{-9}$, and input BER of 10^{-5} to output BER of 10^{-13}.

Specification. A specification for the QPSK modem/codec is given in Table 1. Certain data rates and interfaces are evolving as standards, but other rates and interfaces can easily be provided. Please contact Scientific-Atlanta regarding any desired differences from the specification. The theoretical back-to-back modem/codec bit error rate performance with no implementation loss is presented. This is a conservative bound allowing for aging, environment, etc. Typical measured coded performance is well within 1 dB of theory.

Measured Performance. Typical measured performance is provided in Figures 12 and 13. The uncoded performance in Figure 12 with a pseudo-random data input is seen to be within 1 dB of theory for E_b/N_os as high as 12 dB. This measured performance is the average of all four possible lockup phases of the demodulator. It should be noted that one can optimize one phase at the expense of the other phase lockup states. The performance with scrambling is found by multiplying each measured error rate by three since scrambling introduces an error propagation of three errors for each bit error.

Table 1. QPSK Modem FEC Codec Performance Specifications

Characteristic	Specification
	Prime Power
Voltage	120V ac ±10%
Frequency	50 to 60 Hz
	Service Conditions
Operation	Continuous
Temperature	32°F to 120°F
	IF Interface
Modulator	
Power Level	-20 dBm Max. Order any level less than or equal to -20 dBm.
Impedance	75 Ω unbalanced
Frequency Stability	±1 kHz
Output Frequency	52 to 88 MHz center frequency
Type Modulation	Offset QPSK
Bandwidth	3-dB bandwidth equals symbol rate 99% of signal power with 1.5 x symbol rate or 0.75 x bit rate
Connector	BNC
Demodulator	
Power Level	-50 dBm Nominal, -60 to -40 dBm
Impedance	75 Ω unbalanced
Acquisition Range	±25 kHz from center frequency
Input Frequency	52 to 88 MHz center frequency
Type Modulation	Offset QPSK
Bit Error Performance	Within 1.5 dB of attached theoretical curve for modem/codec with random data. (Typical performance within 1 dB of theoretical)
Connector	BNC

Table 1. QPSK Modem FEC Codec Performance Specifications (continued)

Characteristic	Specification
Baseband	
Data Rate	30 kb/s to 10 Mb/s
Data Format	NRZ-L
Transmit Clock	Internal to Modem, $\pm 10^{-5}$ stability OR external clock input, ±0.1% stability. Higher stability options available. Internal clock is optional.
Receive Clock	Recovered from received signal.
Timing Jitter	Transmit <10% of a clock period Receive <5% of a clock period
Scrambler	per CCITT V.35
Interface	per CCITT V.35; others on request
Connector	per Bell No. 41450; others on request
FEC Codec	
Code	Convolutional self-orthogonal, minimum distance 7.
Rate 7/8	
General Polynominal	0, 2, 8, 32, 88, 142 0, 3, 19, 52, 78, 46 0, 11, 12, 62, 85, 131 0, 21, 25, 39, 82, 126 0, 5, 20, 47, 84, 144 0, 58, 96, 106, 113, 141 0, 41, 77, 108, 117, 130
Decoder	Threshold decoding with syndrome feed-back.
Performance	random errors 10^{-3} corrected to 2×10^{-6} 10^{-4} corrected to 1×10^{-9} 10^{-5} corrected to 6×10^{-13}

Table 1. QPSK Modem FEC Codec Performance Specifications (continued)

Characteristic	Specification
Mechanical	
Size (for Modem Codec) Full Duplex package	19 inch width (48.63 cm) (standard rack mounting) 7 inch height (17.78 cm) 20 inch depth (50.8 cm)
Modulator/Encoder	19 inch width (48.63 cm) 5-1/4 inch height (13.34 cm) 20 inch depth (50.8 cm)
Demodulator Decoder	19 inch width (48.63 cm) 5-1/4 inch height (13.34 cm) 20 inch depth (50.8 cm)
Indicators	
Demod carrier reference synchronization	Front-panel green indicator is activated if carrier reference loop is locked.
Demod Symbol Timing	Front-panel green indicator is activated if symbol synchronizer has locked.
Input power level to Demod	Front-panel green indicator is activated if input power is within AGC range.
Demod Status	A form-C relay closure is provided at the rear panel when branch synchronization is achieved in the FEC decoder. Such synchronization also indicates that the carrier reference loop and the symbol timing loop are locked.
Modulator power	Front-panel green indicator is activated if modulator output power is present.
Modulator status	A form-C relay closure is provided at rear panel if modulator output power is present.

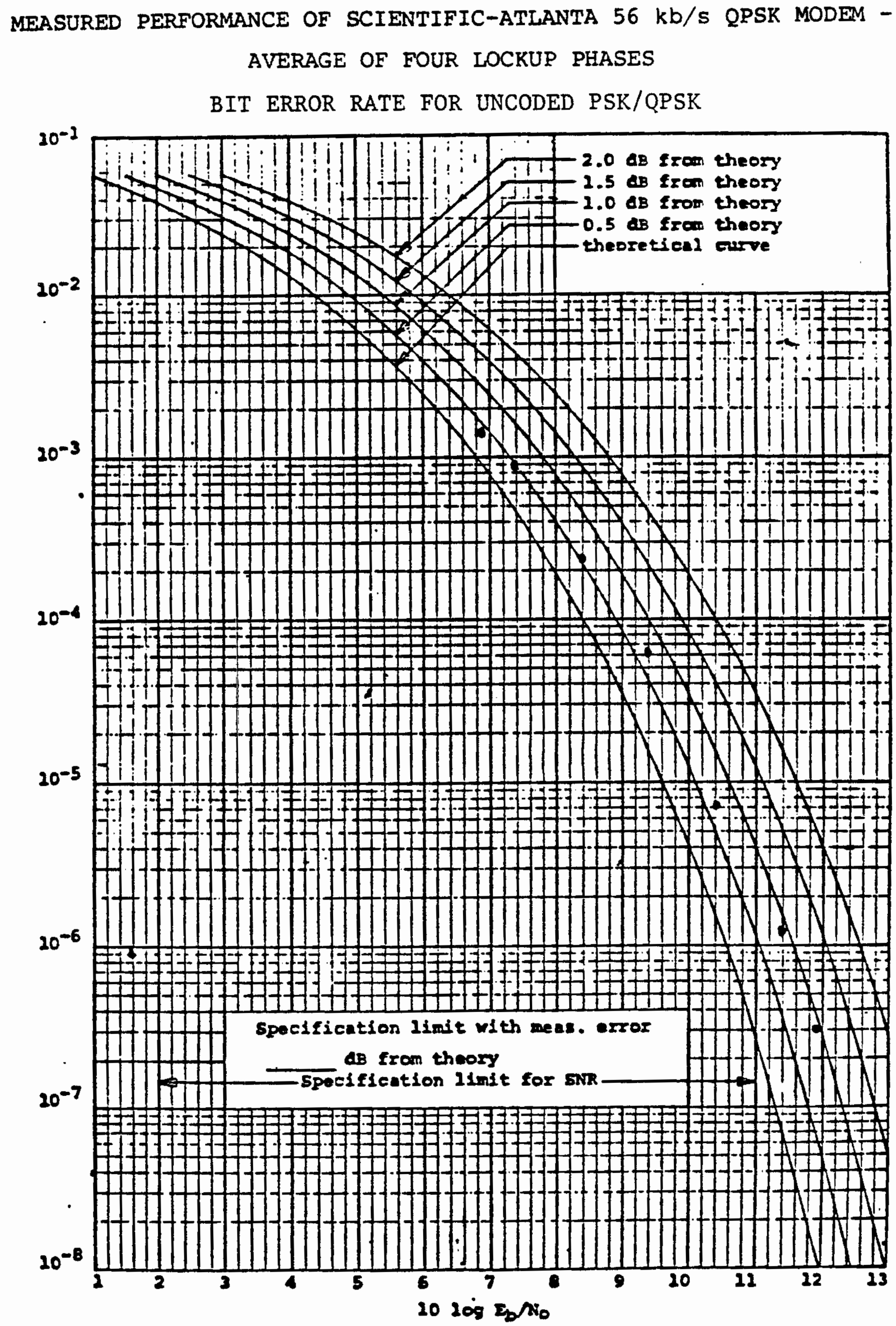


Figure 12. Typical Measured Uncoded Performance

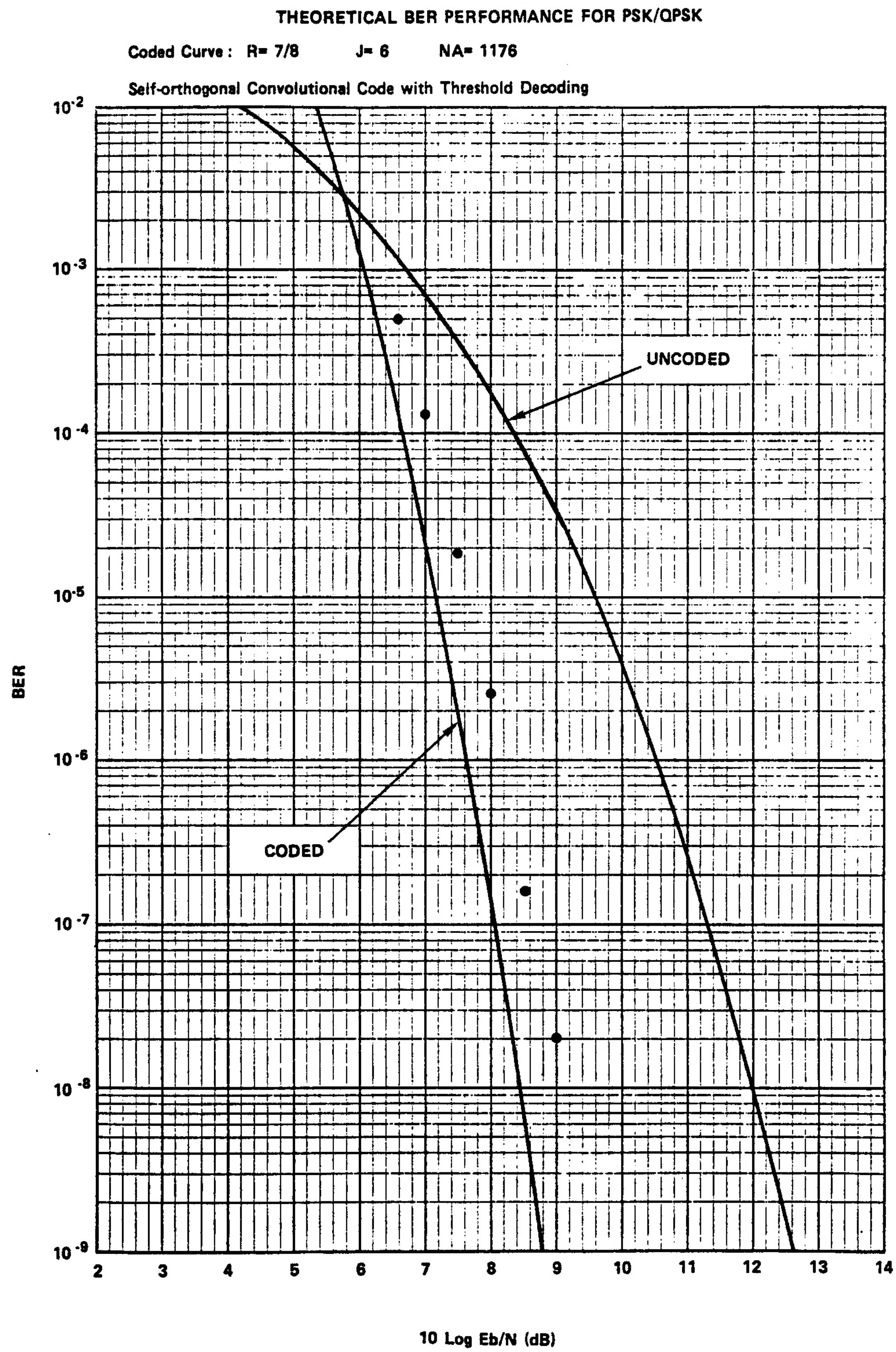


Figure 13. Typical Measured Coded Performance

Typical measured coded performance is given in Figure 13 and is within 1/2 dB of theory for a pseudo-random input. Again, performance is averaged over all four phase lockup states. Performance with scrambling is again found by multiplying each measured BER point by three to derive scrambled performance.

MODEL 8851 1:N MODEM PROTECTION SWITCH

General. The Model 8851 1:N Modem Protection Switch (shown in Figure 14) is a microprocessor-based device which provides redundancy protection for a Series 8800 Digital Modem Subsystem containing from one to eight on-line units, with one frequency-agile backup unit. The switch is expandable to accommodate the desired number of on-line modem/codec units, by adding the required plug-in interface modules. The switch can be configured for transmit-only, receive-only, or full-duplex applications; in the duplex mode, transmit and receive switching operates independently. An option for a redundant internal clock is provided in the switch. This unit is contained in a 8-3/4-inch-high (22.23 cm) by 19-inch-wide (48.63 cm) rack-mounted package. Redundant internal power supplies are included.

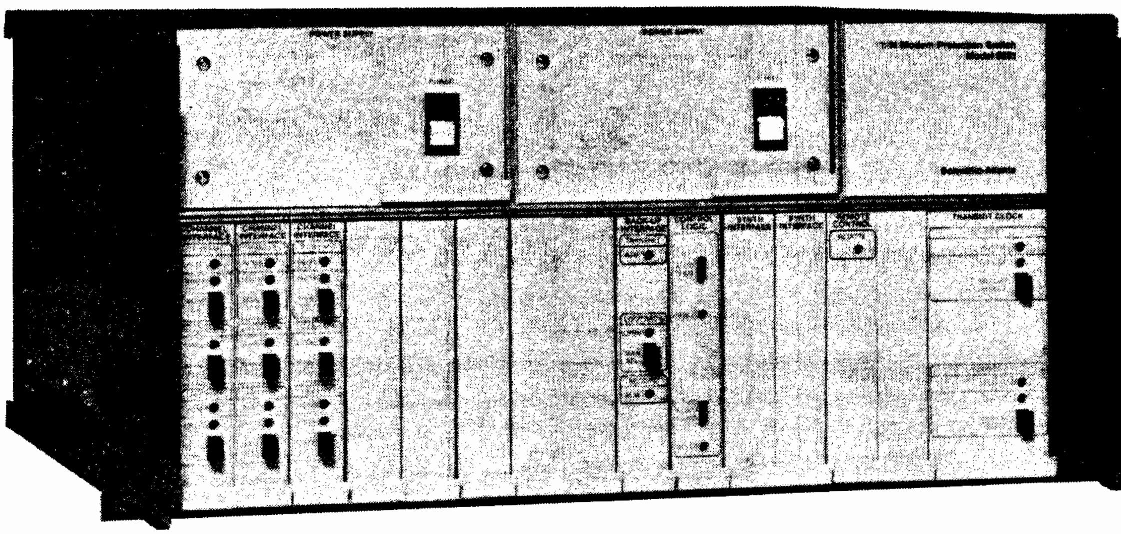


Figure 14. Model 8851 1:N Modem Protection Switch

1:N Modem Subsystem. A typical 1:N Modem Subsystem is shown in Figure 15. A general block diagram of the 1:N Modem Protection Switch is shown in Figure 16. The protection switch provides both a Modem interface and a Customer (user) interface. Up to eight transmit and eight receive channels can be implemented by adding plug-in channel interface modules. Various standard interface cards are available (i.e., V.35, RS-422, DS-1). A preemptible, ninth data channel is available by using the frequency-agile backup unit. However, this channel is preempted if an on-line unit fails.

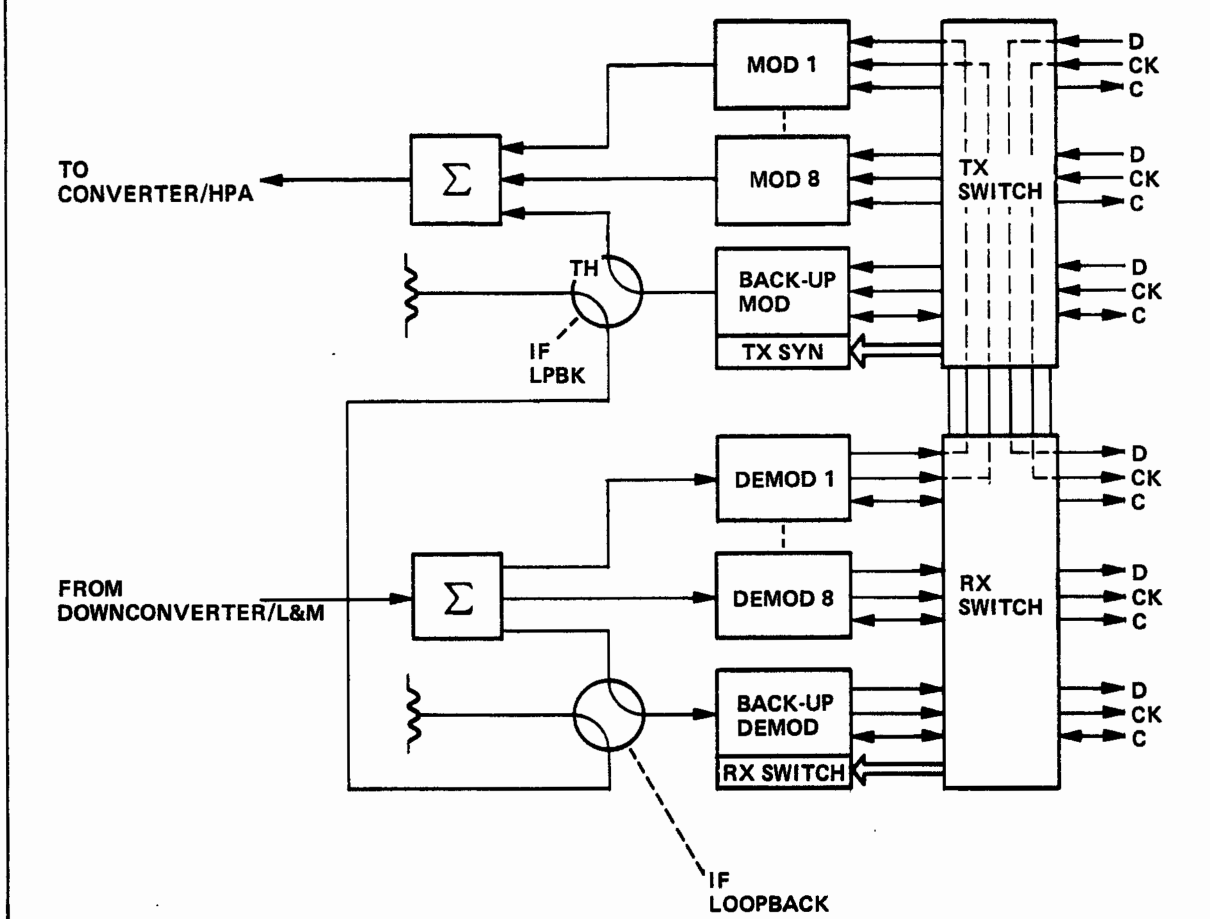


Figure 15. Typical 1:N Modem Subsystem

The protection switch may operate as either a simplex (transmit or receive) or a duplex (transmit and receive) unit. All switching functions are detected and commanded independently for modulators and demodulators. This prevents a far-end, uplink failure from causing an unnecessary local, uplink protection switch attempt.

Modulator outputs are combined in a power combiner and are given an output inhibit commanded by the protection switch if a failure occurs. This arrangement eliminates the need for a mechanical switching matrix on the transmit signals.

Synthezisers for the frequency-agile backup units are controlled by the protection switch via parallel control lines. A synthesizer interface module contains miniature rotary switches which allow field programming of ten backup frequencies. Two modules are required for duplex operation.

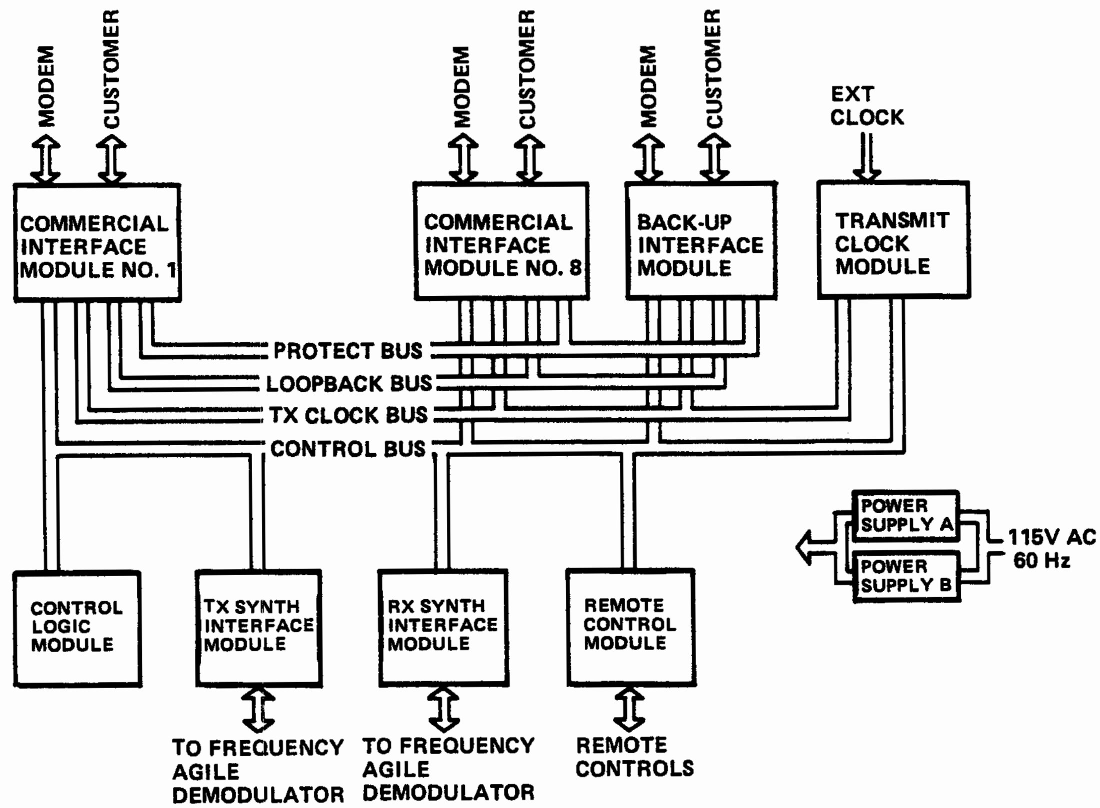


Figure 16. 1:N Modem Protection Switch Block Diagram

Table 2. Loopback Modes

Name	Source	Destination	Signals	Modems
<u>Mode 1</u>				
Data (Baseband)	Customer Data Input	Customer Data Output	TX Data - RX Data TX Clock - RX Clock	All On-line units Backup
	Demodulator Data Output	Modulator Data Input	RX Data - RX Data RX Clock - TX Clock	
<u>Mode 2</u>				
RF (Satellite)	RF Uplink	RF Downlink	TXRF - RXRF	All On-line units

Loopback. There are three loopback modes in the Modem Subsystem: Data loopback, IF loopback, and RF loopback. These are summarized in Table 2.

Data loopback sends data and clock from a demodulator output to the input of the modulator on a selected data channel. At the same time, customer data and clock inputs are sent the customer data and clock ouputs of the selected data channel. This mode is selectable manually from the front panel or by remote control.

IF loopback applies only to the agile backup unit shown in Figure 15; an IF signal path is provided by a coax switch under the control of the protection switch. This loopback signal is normally used for continuous monitoring of the backup unit. An early indication of failure can thus be reported before the backup unit is required for service.

RF loopback is accomplished by tuning a demodulator to the transmit frequency of the modulator of the same data channel. Control of the demodulator is provided by the protection switch. This mode is selectable at the switch by remote control only; however, it may be selected manually at the front panel of the modem.

Remote Control. Due to the large number of control and monitor points within the 1:N modem subsystem, remote control and monitoring of the protection switch is provided via a serial ASCII data interface. This permits comprehensive remote control with very little hardware and wiring.

Transmit Clock. An optional Transmit Clock module is available which provides a 1:1 protected transmit clock signal to all customer interface ports. The module contains two TCXOs as the clock sources. However, an externally supplied signal can be substituted for one of the TCXOs.

Power Supplies. The protection switch contains redundant power supplies to ensure failsafe operation. Power supplies are monitored by the Control Logic Module and failures are reported via the remote control interface. A failed power supply may be replaced without disrupting data traffic. Power supplies operate from 115V ac, 60 Hz.

Model Protection Procedure

Transmit. Figure 17 is a top-level flow chart of the protection procedure followed by the switch when a failure occurs in an on-line modulator.

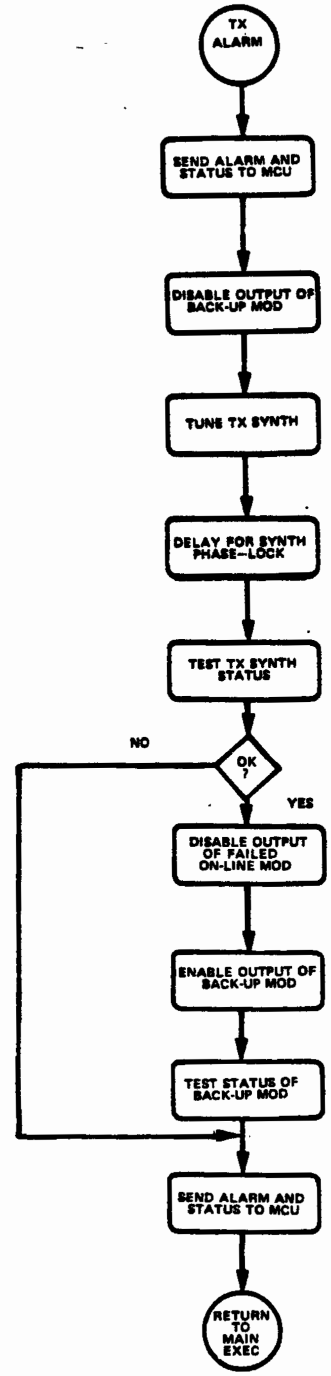


Figure 17. Modulator Protection Procedure

The protection switch scans the alarm inputs from each modulator. When an alarm is detected, a status update is sent to the Remote Control Interface. The switch then disables the output of the backup modulator and sends a frequency tune command to the transmit synthesizer. After a delay to allow for the synthesizer's phase-locked loop to settle, the switch tests the synthesizer status. If the status is bad, the attempt to switch is aborted. An updated status is again sent to the Remote Control Interface. If the synthesizer status is good, the switch disables the output of the failed on-line modulator and enables the backup modulator. A final test is then made on the backup modulator and an updated status is sent to the Remote Control Interface.

Receive. Figure 18 is a top-level flow chart of the protection procedure followed by the switch when a failure occurs in an on-line demodulator. The protection switch scans the alarm inputs from each demodulator. When an alarm is detected, a status update is sent to the Remote Control Interface. A frequency tune command is then issued to the receive synthesizer. After a delay to allow for phase lock, the synthesizer status is tested. If the status is bad, the switch attempt is aborted and an updated status report is sent to the Remote Control Interface.

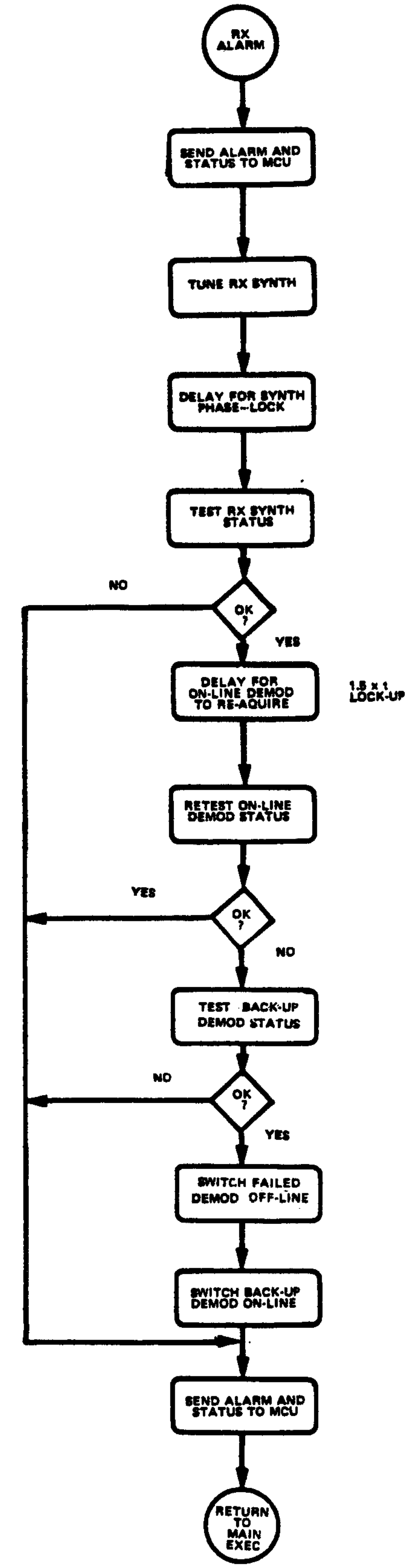


Figure 18. Demodulator Protection Procedure

If the status is good, the switch delays for approximately 1.5 times the normal lockup time of the demodulator. This gives the previously failed on-line demodulator a chance to reacquire the incoming signal. Thus, an unnecessary switch is prevented in case the failure was caused by a short-term, transient condition.

If the on-line demodulator reacquires the signal within the specified time, the switch attempt is aborted and an updated status report is sent to the Remote Control Interface. However, if the on-line demodulator does not reacquire the signal, the backup demodulator is tested for acquisition.

If the backup demodulator has not acquired the signal at this time, the problem is assumed to be elsewhere in the system and no switch attempt is made. If the backup demodulator has acquired the signal, the failed demodulator is switched off-line and the backup demodulator is switched on-line. An updated status report is then sent to the Remote Control Interface.

Characteristics Summary. The 1:N modem protection switch characteristics are summed in Table 3.

Table 3. 1:N Modem Protection Switch Characteristics

Characteristic	Specification
Customer Interface	
Data, Clock, Controls	V.35 compatible (other levels available)
Connector	25 pin "D"
Modem Codec Interface	
Data, Clock, Controls	V.35 compatible (other levels available)
Connector	25 pin "D"

Table 3. 1:N Modem Protection Switch Characteristics (continued)

Characteristic	Specification
Protected Signals	Transmit Data Transmit Clock Request to Send Clear to Send Receive Data Receive Clock Receive Line Signal Detect Data Set Ready
Loopback Signals	Data Clock (no controls)
Remote Control Interface	Serial ASCII, RS-232, NRZ 2400 Baud (other interfaces available)
Operating Environment	
Temperature	0°C to 50°C (32°F to 120°F)
Relative Humidity	0 to 95% without condensation
Power Requirements	108 to 132V ac, 50 to 60 Hz, 80W
Dimensions	Standard EIA panel 483 mm (19 inches) wide, 222 mm (8.75 inches) high, 406 mm (16 inches) deep
Net Weight	13.6 kg (30 lb)

MODEM FREQUENCY SYNTHESIZER

General. The modem frequency synthesizer (shown in Figure 19) is available in both single and dual configurations. In the single configuration, the Model 8861 is a single low-band synthesizer for use with transmit modems; Model 8862 is a single high-band synthesizer for use with receive modems.

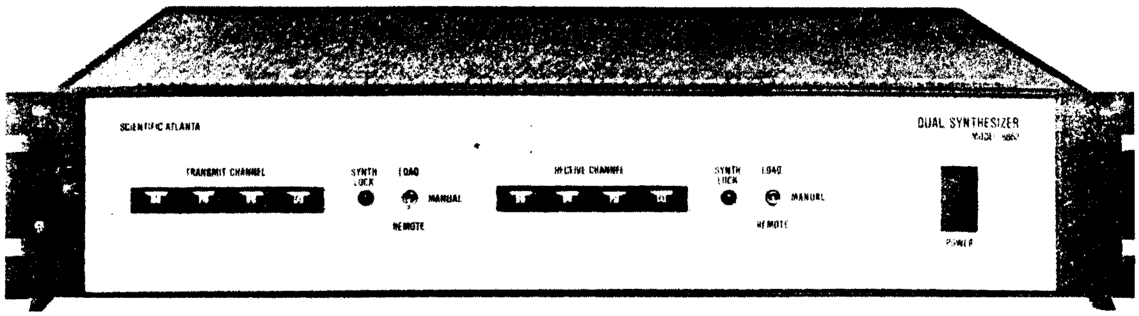


Figure 19. Model 8863 Dual Synthesizer

In the dual configuration, the Model 8863 is available in three versions: receive, transmit, transmit/transmit, and receive/receive. These versions are identified by dash numbers as shown in Table 4 below.

Table 4. Dual Synthesizer Configurations

Model	Configuration
8863-1	RX/TX
8863-2	TX/TX
8863-3	RX/RX

Description. Figure 20 shows a functional block diagram of one of the RF synthesizers. A 4.5-MHz crystal oscillator provides stable clock for the synthesizer and a modulus 180 or 200 counter divides this clock to generate 25.0-kHz or 22.5-KHz reference frequencies, respectively. An internal jumper selects the proper frequency. The reference frequency determines the channel frequency spacing. By using the fine frequency adjust and reference frequency test point, the reference frequency can be adjusted to precisely the correct value.

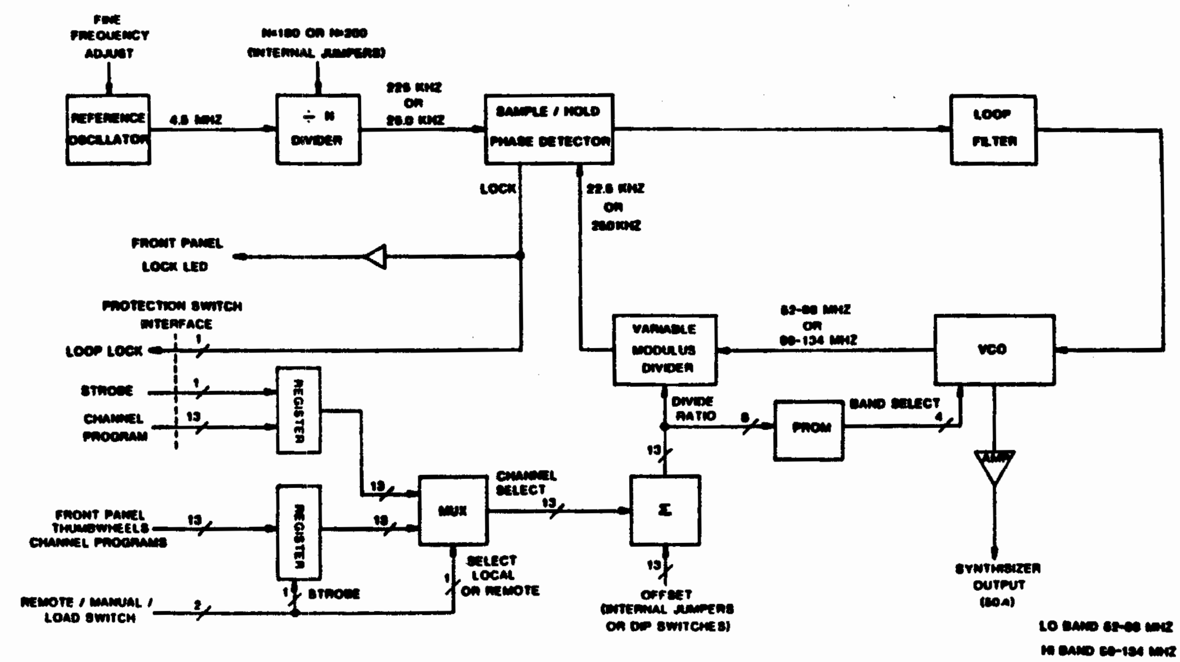


Figure 20. Model 8863 Synthesizer Block Diagram

The sample-and-hold phase detector generates a sawtooth waveform at the reference frequency rate. A double sample-and-hold samples the waveform at a rate determined by the variable modulus divider output. When the loop is locked and no phase error exists, the sawtooth waveform is sampled in the center of its rising ramp. For a positive or negative phase error, the waveform is sampled earlier or later, producing a signed voltage proportional to the phase effect. From this error voltage, the loop filter derives an output voltage to drive the VCO to the correct frequency and phase.

A PROM selects the correct band by switching in inductors in the VCO. The VCO produces an output frequency ideally proportional to its input voltage. (The VCO is actually somewhat non-linear.) An AGC circuit levels the analog output of the VCO, and output filtering suppresses spurious products.

An RF amplifier with a high third-order intercept drives the 50-ohm output at +13 dBm. The variable modulus counter divides the VCO logic output by a fixed divide ratio to produce either a 25.0-kHz or 22.5-kHz output as required. To generate the divided ratio, a digital summer adds a fixed offset number to the channel number.

The low- and high-band synthesizers each use a different offset number which is programmed by internal jumpers. For the channel number, a mux selects either the front-panel thumbwheels or an internal holding register. A front-panel switch controls the mux selection. A channel number may be strobed into the holding register from a remote device by using the remote interface. Alternatively, the channel number may be strobed into a holding register by using the LOAD switch on the front panel. A loop-lock indicator is provided as status to the remote device for control purposes.

Specification. Specifications for the modem frequency synthesizer are summarized in Table 5 below.

Table 5. Modem Frequency Synthesizer Specifications

Characteristic	Specification
	Transmit Synthesizer Output
Power Level	+13 dBm, ±2 dB
Return Loss	>20 dB
Impedance	50 ohms, unbalanced
Frequency Range	52 to 88 MHz
Frequency Spacing	22.5 or 25.0 kHz
Spurious Output	<-60 dBc to 30 MHz beyond band edges
Acquisition Time	<50 ms
Connector	BNC
	Receive Synthesizer Output
Power Level	+13 dBm, ±2 dB
Return Loss	>20 dB
Impedance	50 ohms, unbalanced
Frequency Range	98 to 134 MHz
Frequency Spacing	22.5 or 25.0 kHz
Spurious Output	<-60 dBc to 30 MHz beyond band edges
Acquisition Time	<50 ms
Connector	BNC
	Reference Oscillator
Settability	1 ppm (parts per million)
Range	±30 ppm
Stability	±5 ppm (0°C to 50°C)
Aging	5 ppm/year

Table 5. Modem Frequency Synthesizer Specifications (continued)

Characteristic	Specification
Controls and Indicators	
Four Decade BCN Thumbwheels	Programs Output Channels
Remote/Manual/Load	Selects thumbwheels or remote. Loads thumbwheel channel number into synthesizer Input-for-output channel
Lock LED	Indicates synthesizer loop is locked
Remote Interface	
Levels and Impedance	Open collector TTL with pull-up resistors
Channel Input	Programs Output Channel - 13 parallel lines
Strobe	Clocks channel input into internal memory
Loop Lock	Status indicator to remote unit
Connector (2)	25 pin subminiature -D
Operations	
Duty Cycle	Continuous
Operating Temperature	0°C to 49°C (32°F to 120°F)
Power Requirements	108 to 132V ac, 50 to 60 Hz, 45W
Dimensions	Standard EIA Panel 483 mm (19 inches) wide 88.9 mm (3.5 inches) high 508 mm (20 inches) deep

MONITOR AND CONTROL UNIT

General Description. The Monitor and Control Unit (MCU) is a microprocessor-based device which provides the following four functions for a satellite earth station:

a. As its name implies, the MCU provides monitor and control capabilities for earth station equipment connected to it.

b. It provides a summary alarm of any and all faults present in the earth station equipment.

c. It provides protective switching of faulted on-line units with operational backup units.

d. It provides a parallel or serial remote interface.

The MCU is designed to detect faults in various devices in a satellite earth station, make decisions as to what action should be taken to correct the faults, and command the external devices to correct these faults. In addition, the MCU may be placed in the manual mode, and control of the various connected earth station equipment can be exercised via the front-panel keyboard. The MCU can be used in earth stations of various complexities by simple reprogramming. A remote interface is provided for connection to a remote unit of a central reporting system.

A block diagram of the MCU is shown in Figure 21. The MCU has a basic configuration which may be expanded by the addition of plug-in I/O cards.

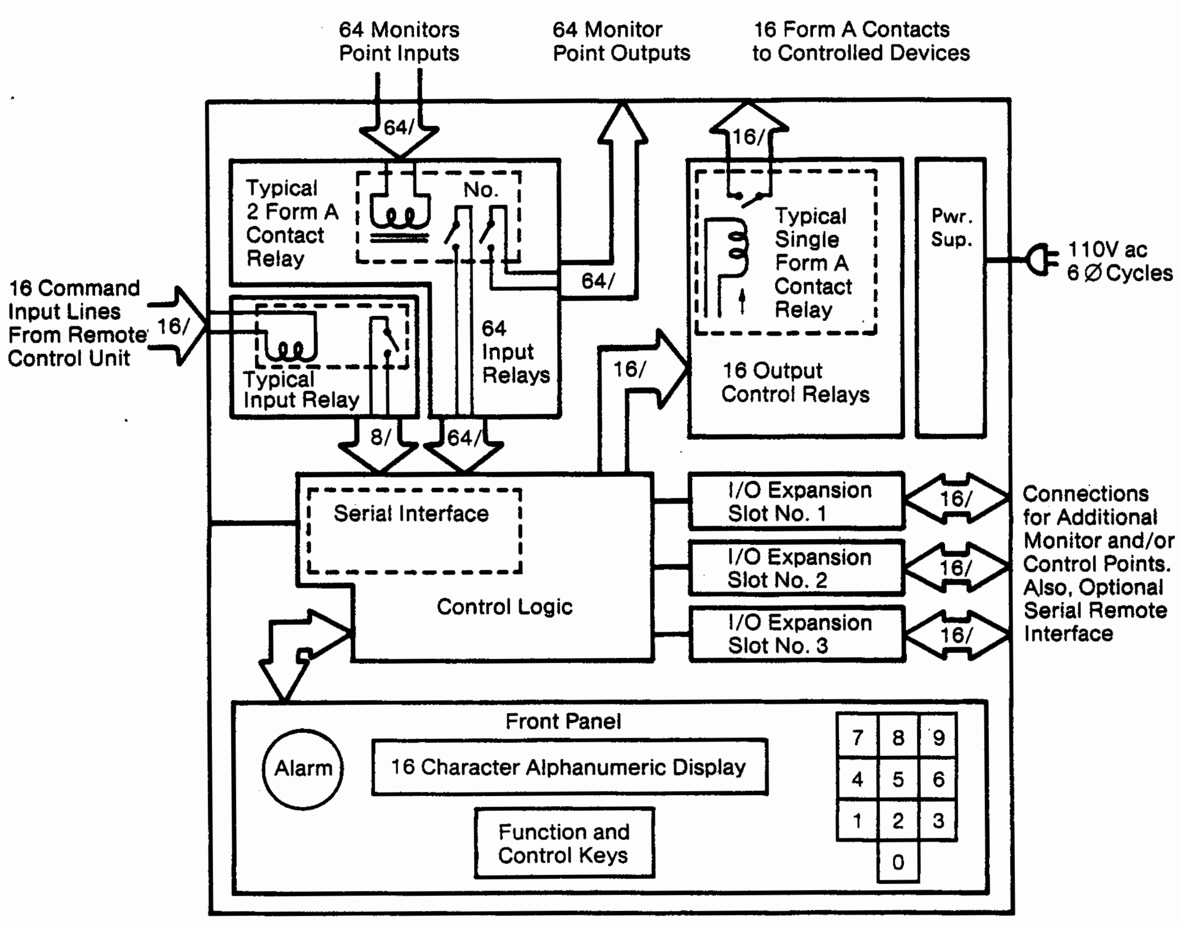


Figure 21. Monitor and Control Block Diagram

Input Relays. There are sixty-four input relays each having two form-A contacts. One set of contacts is connected to the microprocessor and control logic while the other set of contacts is routed back out to the rear of the chassis. These sixty-four isolated contact closures are available for connection to any external remote interface unit or communication device.

In addition to these sixty-four equipment status relays, there are sixteen input relays to receive commands from a remote control unit.

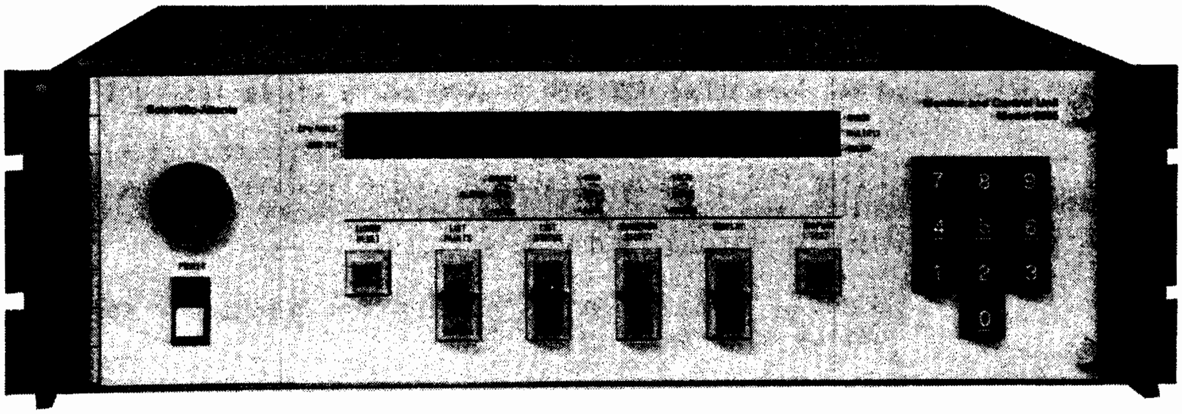


Figure 22. Monitor and Control Unit Front Panel

Output Relays. Sixteen output relays with form-A contacts are available for connection to various MCU controlled points at the earth station. These relays have low-power contacts and are designed to energize an intermediate relay to control the desired device.

Control Logic. The control logic consists of microprocessor, memory, and necessary support circuits. The operating program resides in a pluggable memory that can be custom programmed (at the factory) for various earth station configurations.

I/O Expansion Slots. The number of points monitored by the MCU can be expanded from the basic sixty-four points to eighty points by the addition of an input expansion card. Similarly, the number of points contolled by the MCU can be increased from the basic sixteen to thirty-two points by plugging in an output expansion card.

An optional remote serial interface card can be plugged in instead of using the standard parallel input and output lines of the MCU.

Front Panel. Principle features of the front panel are shown in Figure 22 and are explained in the following sections.

Display. The display consists of sixteen alphanumeric characters and four discrete LEDs. The alphanumeric display is used for indicating the status of a requested device, indicating which devices have faults, indicating which devices had faults at one time but have since "self-healed" and for prompting an operator for next keystrokes. The discrete LEDs are used for indicating MAJOR or MINOR faults, MULTIPLE FAULTS and ALARM DISABLED.

Keyboard. The keyboard is used to examine the status of all connected devices, to examine all current and recently "self-healed" faults, and to command various devices to be on-line or off-line for maintenance or trouble-shooting purposes. The key functions are explained as follows:

Key	Function
0 - 9	Used for selecting which device to monitor and control.
ENTER	Used to signal the end of the entry of an instruction.
CLEAR	Used to clear an incorrectly entered instruction or to clear the previous keyboard entry from the display.
MONITOR	Used to display a device's status.
COMMAND	Used to initiate an instruction for bringing a specified device on-line or off-line.
LIST STATUS ↑↓	Used for scrolling through the list of statuses of all devices.
LIST STATUS ↑↓	Similar to the list status scrolling keys but only lists those devices with faults.
ALARM RESET	Silences the audible alarm even though faults are present. Allows alarm to sound if a new fault should occur.
LAMP TEST	Causes a "block" to move across the alphanumeric display to test for any burned-out LEDs in the display. Also tests the four discrete LEDs.

Toggle Switches. There are three toggle switches and their functions are explained as follows:

Switch	Function
AUTO/MANUAL	Selects whether the MCU automatically scans the connected devices for faults and backs up the faulted device, or whether scanning and backing up is to be accomplished manually via the front-panel keyboard.
REMOTE/LOCAL	This switch selects whether MCU will be controlled locally (via the front-panel keyboard or remotely via the remote interface).
ALARM DISABLE	This switch completely disables the audible alarm. The ALARM DISABLED LED will illuminate to indicate that the alarm is disabled.

Alarm. The audible alarm will sound whenever faults are present, when the alarm switch is not in the Disabled position, and when the alarm has not been reset. If the alarm is not disabled, it will only sound when a new fault occurs.

Power Switch. The power switch is a paddle-type ON/OFF switch with an LED for indicating the power is on.

Operation Modes. In the automatic mode, the MCU continuously monitors all equipment fault lines and the configuration of all switches on a rotating basis. If the REMOTE/LOCAL switch is in the LOCAL position, the MCU may be placed in the automatic mode by placing the AUTO/MAN switch in the AUTO position. If the REMOTE/LOCAL switch is in the REMOTE position, the automatic mode may be entered by an appropriate command from the remote unit.

If an equipment fault is detected, the MCU determines if it is an on-line unit and, if so, takes the faulted device off-line and brings a backup unit on-line to replace it. The display then indicates the type of failed unit, its number, its status (on-line or off-line), whether it is a major or minor fault, and whether there are now multiple faults present.

A minor fault is a fault in an off-line unit or a fault in an on-line unit with an operational backup unit. A major fault is a fault in an on-line unti with no operational backup unit. The main difference between a major and minor fault is that a major fault takes the station off the air until fixed while a minor fault is one that may temporarily take the station off the air but can be fixed by the MCU to bring the station back on the air. A fault in an off-line unit is also considered a minor alarm.

If a switch is found to be in an incorrect position, the MCU will attempt to switch it to the correct position. If the switch is still in the incorrect position, it will be indicated as a major or minor fault depending on whether the earth station is taken off the air or not.

When any fault occurs, the audible alarm will sound until the fault is cleared or the alarm is disabled or reset.

Statuses and faults can be examined while the MCU is in the automatic mode by using the DISPLAY function or the LIST STATUS or LIST FAULTS keys. However, if a new fault occurs while statuses or faults are being examined, the new fault will override and reset these functions, and the new fault will automatically be displayed.

Manual Mode. In addition to being able to monitor the status of the station as described above, the manual mode allows the operator to control the configuration of the station. This control includes the ability to switch devices on-line or off-line, to turn HPA power on or off, and to initiate an IF or RF loop test.

If a failure occurs while the MCU is in the manual mode, the fault information will still be displayed (as in the automatic mode), but no automatic switching of an off-line unit for a faulty on-line unit will occur. When the MCU is placed back in the automatic mode though, it will switch on an operational back-up unit in for a faulty on-line unit. When the MCU is switched from the automatic mode to the manual mode, no change in the station's configuration will occur.

Remote Control Mode. The remote control mode allows transfer of the total status and control capability of the station to a remote console if requested by the remote unit. When the remote control mode is selected, the station will initially be in the automatic mode regardless of the position of the AUTO/MANUAL switch. Once in the remote control mode, a command can be entered from the remote console to place the MCU in a remotely controlled manual mode. The front-panel keypad and function switches are effectively disabled while the MCU is in the remote control mode.

In either the automatic or manual mode, the sixty-four status lines and sixteen control lines are available for connections to the remote control unit. It is only in the remote control mode that the remote console can exercise control of the MCU. If the serial port option is implemented, additional information, such as the station configuration, can be sent.

Interface Characteristics. The interface characteristics are summarized in Table 6 below:

Table 6. Interface Characteristics

Characteristic	Specification
Power	115V ac @ 60 Hz
Status Inputs	
Parallel	Form-A contact closures. These contacts pull in a low-power relay in the MCU and are used for isolation purposes. A closure is interpreted as a FAULT.
Serial	Standard RS-232 interface from the 1:N Modem Protection Switch.
Control Outputs	All control output lines are form-A relay closures.
Power (Max)	10 watts dc
Current (Max)	0.5 amps dc
Voltage (Max)	200 volts dc
Remote Control Interface	
Status Outputs	All sixty-four status input contact closures are available as separate, isolated contact closures to an external remote reporting unit of the customer's choosing.
Control Inputs	The MCU can be commanded from a remote console via a remote interface by providing up to sixteen contact closures.
Remote Control Interface (Optional Serial)	Two serial interface options are available if the parallel input and output lines of the remote control interface described above are not used. 1. Standard RS-232 Serial Interface Port. 2. Serial Telco/Auto-Dial/Auto-Answer Port

Digital Audio Receive Only Terminal
Outside Equipment and BPSK Receiver/FEC Decoder

Dr. James S. Gray

The basic Digital Audio Receive-Only Terminal consists of the outside installed equipment, Wideband BPSK (bi-phase shift key) Receiver/FEC (forward error correcting) Decoder chassis and the Digital Processing Unit.

Three Meter Antenna

The 3-meter antenna combines excellent performance features with high-volume manufacturing techniques to produce the most cost-effective, small-aperture terminal in the industry. These features include:

a. Corrugated aperture, focal-point feed system with:

   1) High-aperture efficiency

   2) Radiation patterns at 4 GHz which meet requirements of FCC 25-209

   3) Single or dual (optional) polarization operation

   4) Weather-resistant housing for LNA protection and

   5) Convenient polarization adjustment

b. Precision die-stamped reflector panels which provide:

   1) Long-term durability in harsh environments

   2) Dimensional repeatability, consistent high performance

   3) Minimized weight and volume for ease of shipping and installation

   4) Simplified logistic support

c. A rugged reflector support and mount providing:

   1) Stiffness required for operation under severe environmental conditions

   2) Structural safety per guidelines established by the AISC and EIA

   3) Convenience in shipping and handling

   4) Ease and minimum cost of installation

   5) Simple pointing adjustment mechanism.

LNA

The LNA receives the incoming downlink signal directly from the feed, amplifies it, and routes it to the receive electronics via coaxial cable. The LNA is completely solid-state, and covers the entire 3.7- to 4.2-GHz receive bandwidth. The electrical power for the LNA is provided by a separate power supply which is located in the station receive equipment rack. No separate power lines are run to the LNA, since +15V dc power is supplied on the coaxial signal cable.

Each GaAs FET LNA produces a maximum noise temperature of 120K over the receive band while producing a nominal 60 dB of gain. To achieve this performance, the amplifier is designed with six stages of gain. The first two stages employ GaAs FET devices to get the minimum possible noise figure, while the next four stages utilize bipolar transistors. Active biasing is used on some of the stages to provide good gain stability with temperature. The entire microwave circuit is constructed on microstrip for ease of tuning and low-cost printed circuit production.

All LNA Units are supplied with an integral isolator to ensure a low input VSWR. The isolator package includes a waveguide input so that the LNA can be directly connected to the feed receive port. Provision for output isolation is provided on the microstrip board after the six stages of gain. Each LNA unit includes an integral power supply, permitting the dc voltage to vary from +15V to +25V without performance degradation. All units are fully weatherproofed, and all may be pressurized as part of the waveguide assembly.

Wideband BPSK Demodulator

The wideband BPSK receiver as shown in Figure 1 receives the 4-GHz signal from the IFL cable. The 4-GHz input signal is downconverted to 70 MHz in a dual-conversion downconverter and amplified. The downconverter output at 70 MHz is then routed to the BPSK demodulator card where coherent demodulation takes place. The output of the demodulator is 8.777-Mb/s data and the reconstructed clock. These initial data state-estimates and clock are then routed to a forward-error-correction decoder where much better state-estimates are

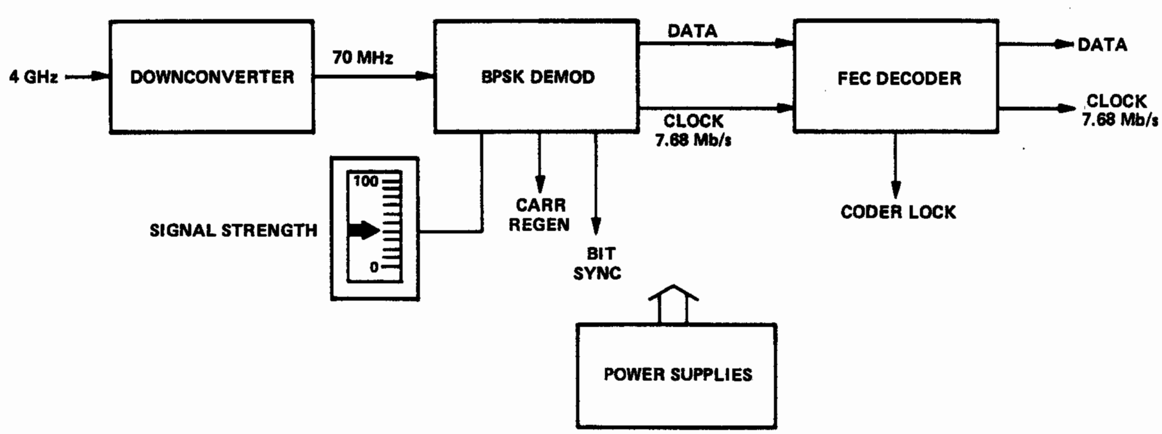


Figure 1. Wideband BPSK Receiver/FEC Decoder

made. The output of the decoder is a 7.68-Mb/s data stream and clock at 7.68 Mb/s. More detail for each portion of the Wideband BPSK Receiver/FEC Decoder is now presented.

As seen in the overall block diagram of the Wideband BPSK Receiver/FEC Decoder in Figure 1, there is a signal strength meter which is useful in ascertaining that adequate signal strength is available from the receive signal from the satellite. There are also carrier regeneration and bit synchronization LEDs on the front panel that show that these processes are properly functioning in the biphase demodulator. The decoder also has an LED lock indicator that indicates that it has achieved branch synchronization.

Downconverter

A block diagram for the downconverter is shown in Figure 2. A dual-conversion downconverter is used so that all satellite transponders can be easily acessed. The first downconverter converts the 4-GHz input signal, which actually lies in the range of 3.7- to 4.2-GHz to an 880-MHz IF. The local oscillator for this downconverter lies in the 710- to 825-MHz region, and its frequency is changed by plugging in a different crystal. The output frequency spacing of this LO varies by 5 MHz, which then gets multiplied by 4 in the downconverter #1. The 880-MHz signal out of the first downconverter is filtered and then routed to the second downconverter where it is mixed with an 810-MHz crystal source and mixed down to the 70-MHz final output. The overall gain of the downconverter is approximately 25 dB. The components in this downconverter are the same as those used in the 6601 Video Demodulator made by Scientific-Atlanta. Being able to use the same downconverter components for the digital audio terminal has a very beneficial effect on the overall cost of the terminal since these components are made in high volume, as are the antennas, LNAs and interconnect facility cables.

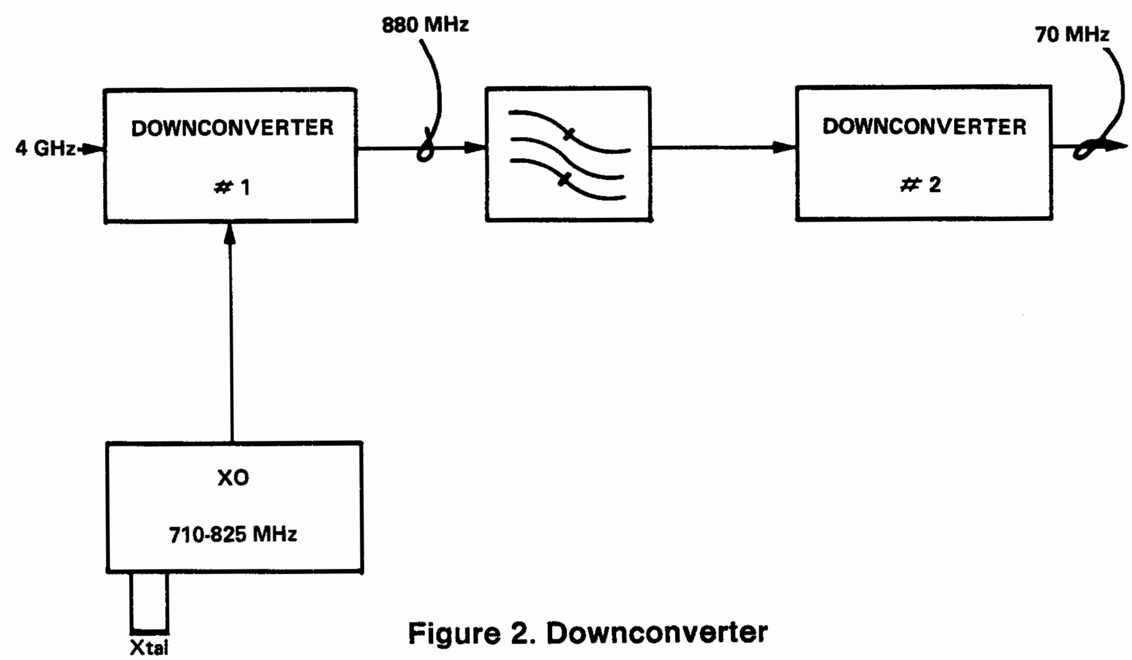


Figure 2. Downconverter

Wideband BPSK Receiver

A block diagram for the BPSK demodulator card is shown in Figure 3. The 70-MHz signal plus noise is routed initially through the matched filter which is implemented at IF by using an arithmetically symmetrical intermediate frequency filter. The analytic low-pass equivalent of this filter is an approximation to the sliding integral matched filter for non-bandlimited PCM NRZ PCM. Placing the filter at the input to the demodulator maximizes rejection of interfering signals and enhances the signal-to-noise in the carrier regeneration and bit synchronization circuitry. The output of the filter is automatic-gain-controlled and then split and routed into several paths, one of which is the bit synchronization path where envelope variations at the bit rate due to matched filtering are routed through an even-order nonlinearity to recreate a bit rate spectral component which is then narrowband-filtered for memory and signal-to-noise improvement. The output of the bit synchronizer controls the decision-making process in the signal conditioner, where at the end of each bit period one makes a state-estimate on whether a one or zero is sent.

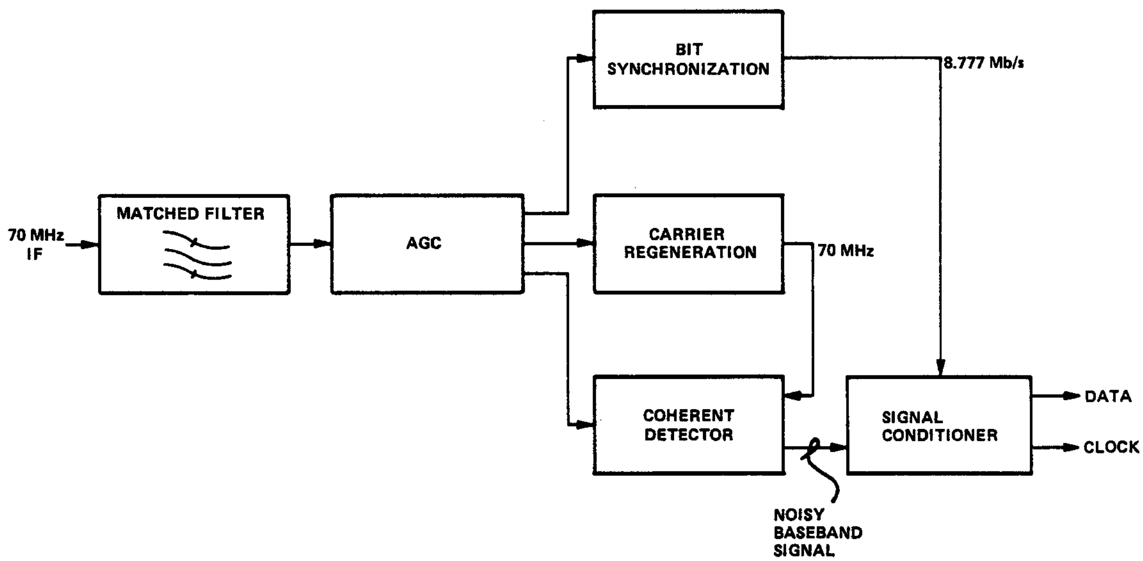


Figure 3. BPSK Demodulator

In the carrier regeneration circuitry, again, the output of the even-order nonlinearity contains a times-two component which modulo 2π wipes off the carrier modulation since the original carrier modulation was either zero or π which multiplied by two will give multiples of 2π or essentially unmodulated carrier at twice the carrier frequency. This is filtered for signal-to-noise enhancement and then divided by two and routed to the coherent detector with the proper phasing to provide coherent reference for demodulation. the other output of the AGC is routed to the coherent detector where it is mixed with the reconstructed coherent reference to produce a baseband output which contains the desired baseband signal and noise. The ouput of the coherent detector is filtered in the signal conditioner to remove carrier reference leak-through and double-frequency terms and then compared to a reference level. At the end of each bit period, a state-estimate is made under control of reconstructed clock from the bit synchronizer. The output data estimate at 8.777 Mb/s data rate is then routed to the FEC decoder along with the reconstructed bit rate clock.

Figure 4 provides further information on the carrier regeneration and bit synchronization process. The ouput of the AGC goes through a times-two or even-order nonlinearity and is then routed into two paths. The envelope variations created by the IF matched filter when going through the even-order nonlinearity produce a bit rate spectral component which is then passed through a narrowband temperature-compensated filter at 8.777 MHz to pick off the bit rate spectral component. This filter provides signal-to-noise improvement and memory for periods where there are no transitions. This signal is then routed with the proper phasing to become the bit rate reference for the signal conditioner. In the upper path the combination of the filtered signal and the times-two component of the nonlinearity produce an unmodulated carrier at twice the input frequency of 70 MHz. This signal is processed by a temperature-compensated narrowband filter for signal-to-noise improvement. The output of the filter is limited and then divided by two and filtered. The resulting 70-MHz signal with the proper phasing becomes the coherent reference for the coherent detector.

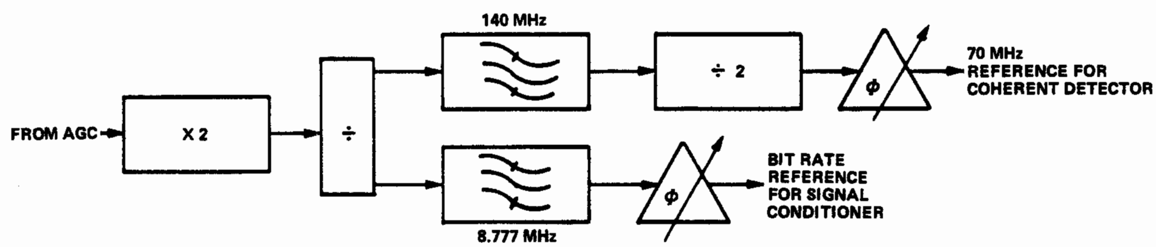


Figure 4. Carrier Regeneration and Bit Synchronization

Signal Conditioner

A block diagram for the signal conditioner portion of the BPSK demodulator card is shown in Figure 5. As shown, the coherent detector baseband signal plus noise output is lowpass-filtered to remove double-frequency terms and coherent reference carrier feedthrough. The analog ouput of the lowpass filter is then compared to a reference level in the limiting amplifier and converted to a logic level according to whether the filter output is above or below the reference level. This logic level then is sampled by a high-speed master-slave D flip-flop which makes a state-estimate as to whether a One or a Zero was transmitted. The bit rate clock from the bit synchronizer is likewise converted in a limiting amplifier to a logic level and, with the proper phasing, controls the sampling process of the flip-flop in making the state-estimate at the end of each bit period. The data estimate and the clock at 8.777 Mb/s are then routed to the forward-error-correction decoder for further processing.

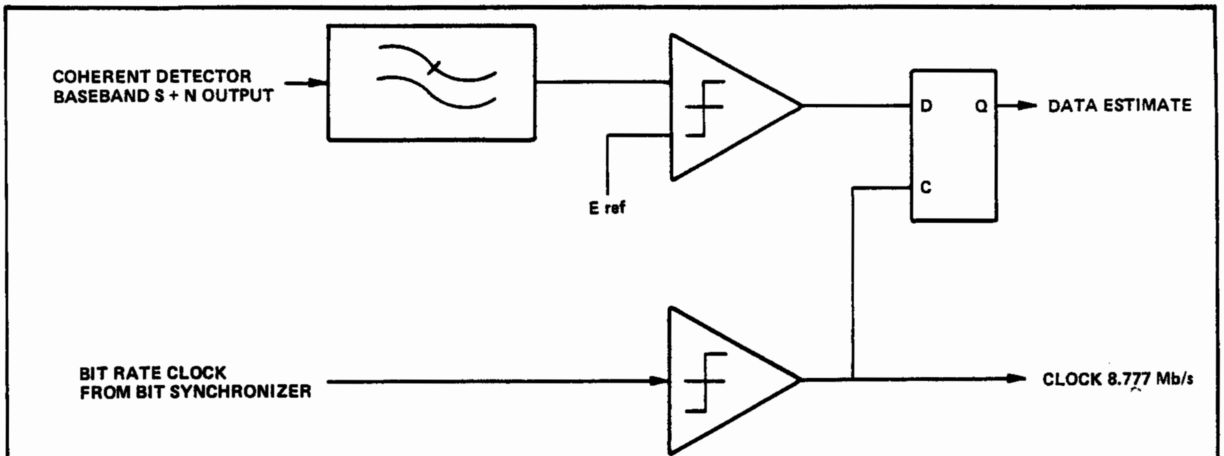


Figure 5. Signal Conditioner

FEC Decoder

The FEC Decoder processes the output of the Wideband BPSK Receiver to correct errors that occur in transmission. A rate 7/8 convolutional self-orthogonal code with a minimum distance of 7 is used. The FEC Decoder uses threshold decoding with syndrome feedback and achieves a 3.45-dB link gain improvement at a 10^{-8} output bit error rate. This decoder performs the following functions:

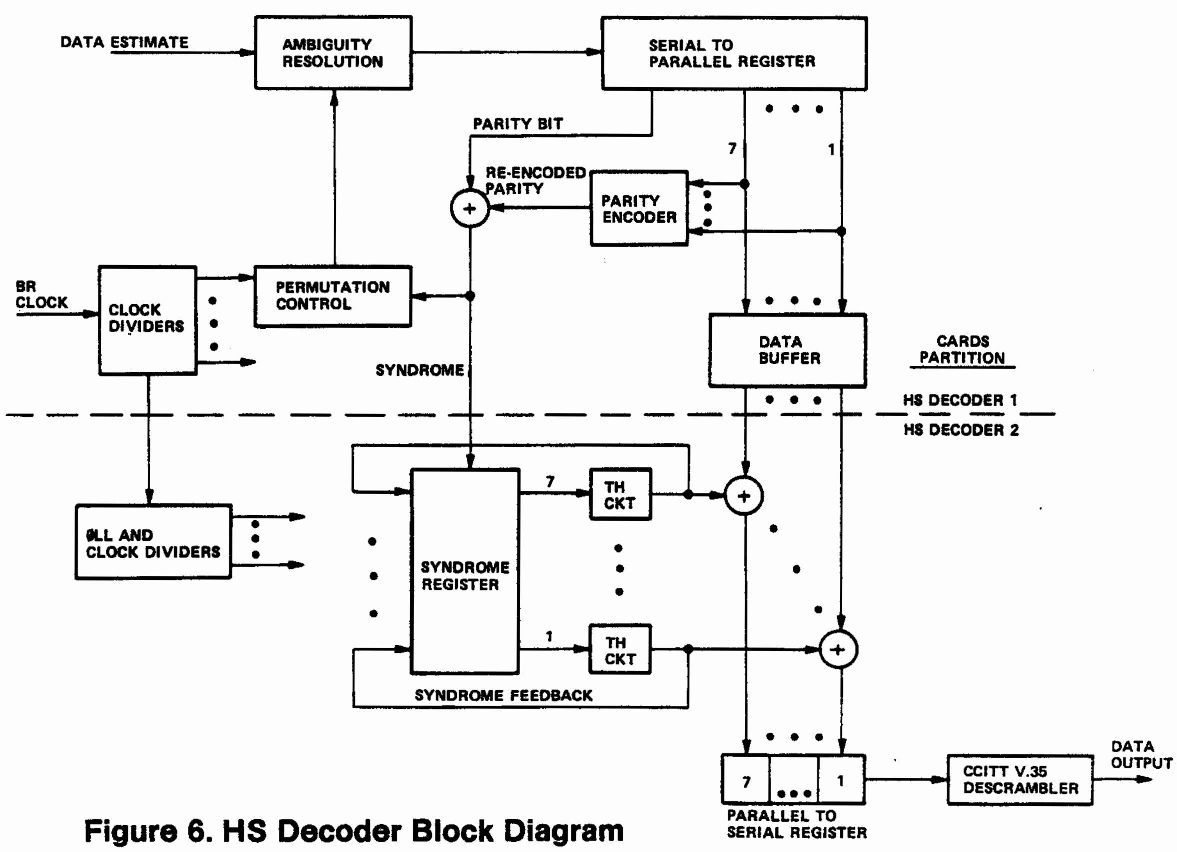


Figure 6. HS Decoder Block Diagram

a. Generates a BR clock from the 8/7 BR clock from the BPSK demodulator clock.

b. Resolves phase ambiguities (data inversion) that result as a function of the modulation/demodulation process.

c. Provides error correction to reduce the overall bit error rate (BER).

d. Provides a descrambler to remove the scrambling action used by the transmit modem for energy dispersal.

Figure 6 shows a block diagram for the high-speed decoder. An ambiguity resolution circuit corrects any data inversion introduced in the demodulation process. The serial-to-parallel register demuxes the parity bit and provides seven parallel data bits to the parity encoder and the data buffer. The parity re-encoder (Figure 7) encodes the receive data bits by the same process as that used by the transmit modem. Thus, if no error occurs in the transmission of the data, the received and transmitted bits are the same; thus, the received parity and re-encoded parity generated by the received data are the same. In this case, the syndrome bits, which represent a comparison or receive parity and re-encoded parity, are all "Zeroes." Syndrome bits "Ones" occur when the two parity bits do not agree, indicating that one or more transmission errors have occurred.

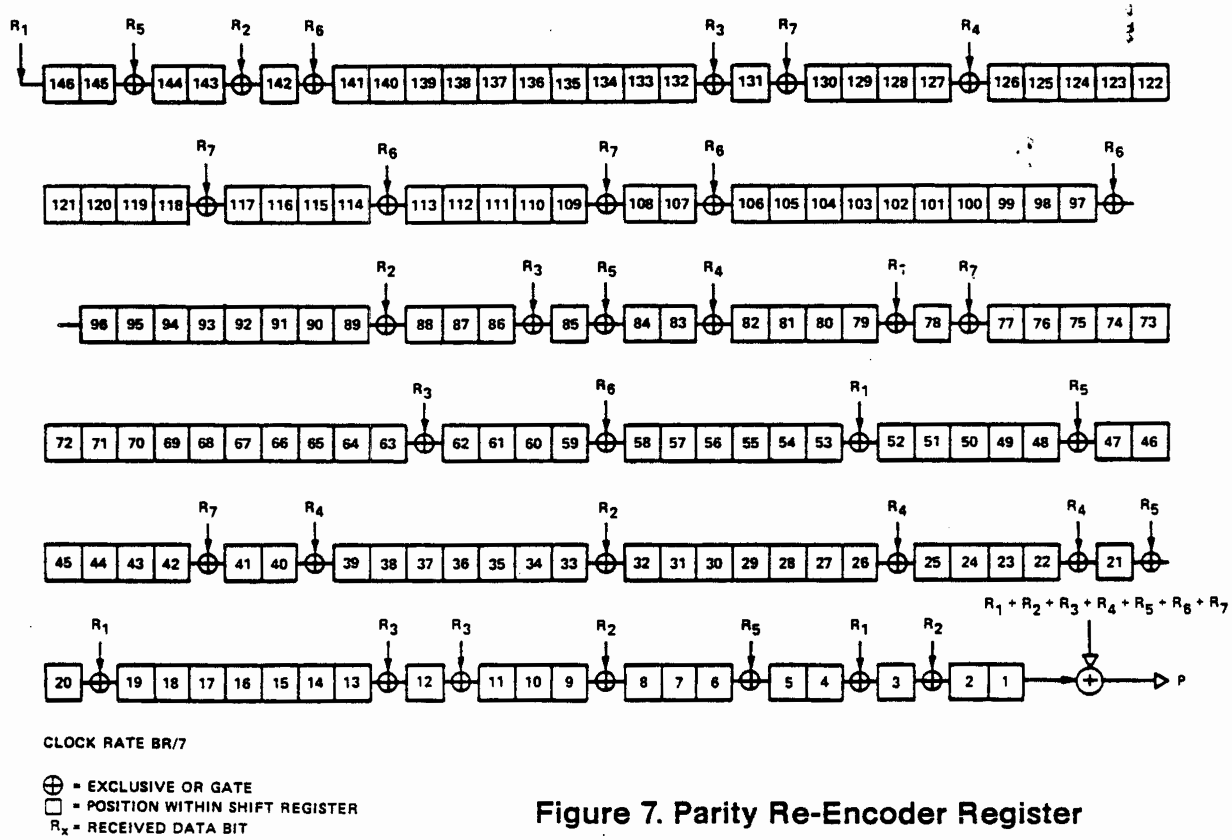


Figure 7. Parity Re-Encoder Register

The permutation control circuit monitors the number of syndrome bit "Ones" produced. If the number of syndrome "Ones" averages 50%, then the permutation will advance the block framing state and/or advance the state sequence of the ambiguity resolution circuit. When a state is found which has less than 25% syndrome bit "Ones" (an uncoded BER of 5.9×10^{-3}), the permutation control inhibits permutation and declares lock. After lock is achieved, error correction may begin.

The syndrome register (Figure 8) stores the last 146 syndrome bits. The seven threshold circuits each search for specific syndrome patterns to determine if an error has occurred. When an error is detected, the threshold circuit outputs a correction signal to the exclusive-or gate connected to the data buffer output. The threshold circuit also inverts the syndrome bits in the syndrome register to remove the error bit effect on future bits to be corrected. The data buffer delays the data bits (approximately 146 bit periods in each path) so that the correction signal will be aligned with the proper bit to be corrected. A parallel-to-serial register converts the seven

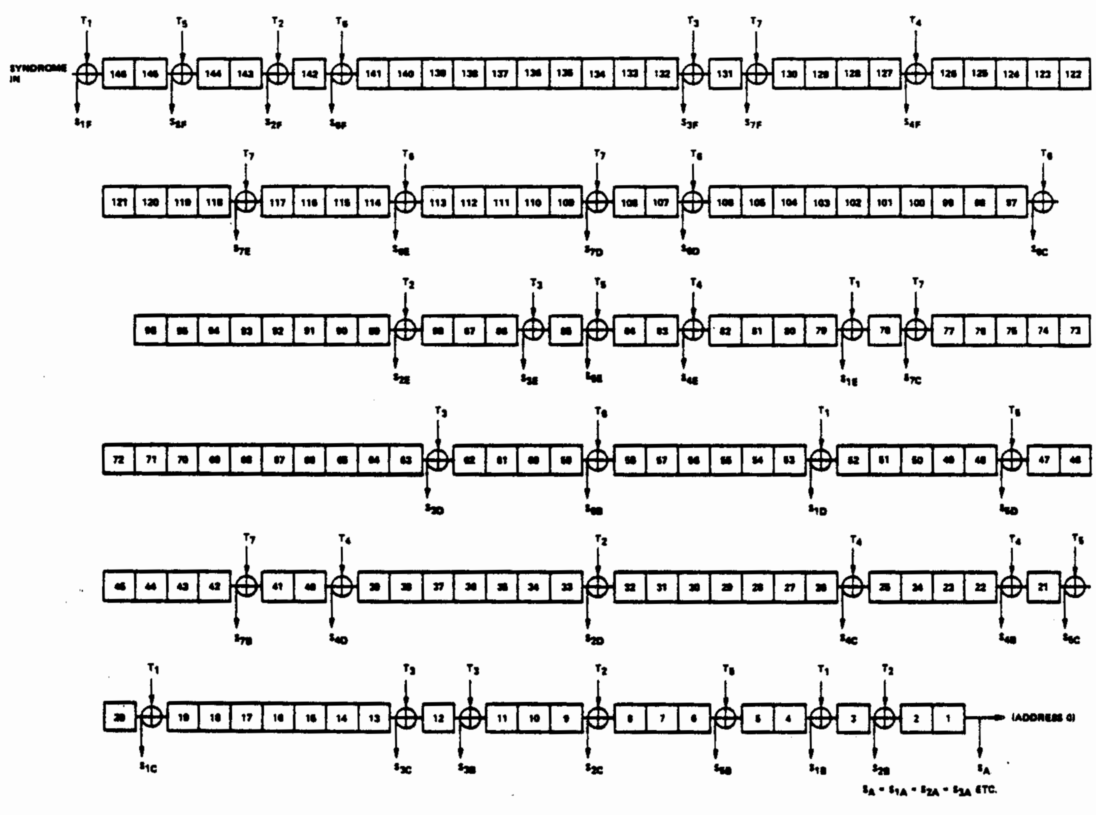


Figure 8. Syndrome Register Block Diagram

corrected data bits into a bit-serial stream. A CCITT V.35 descrambler (Figure 9) then removes the scrambling action introduced by the transmit modem for energy dispersal.

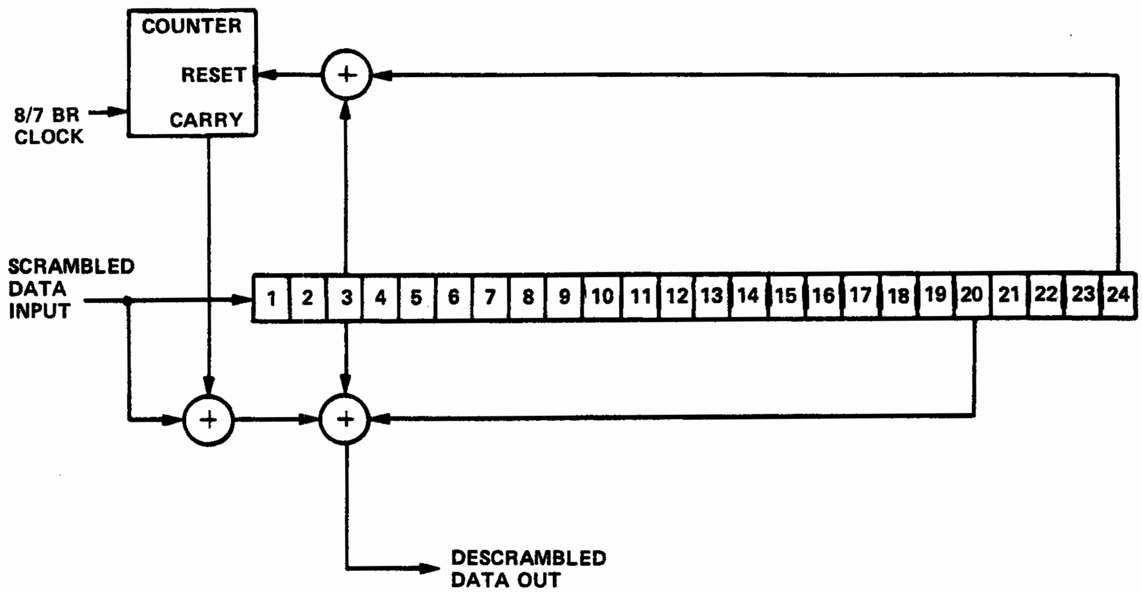


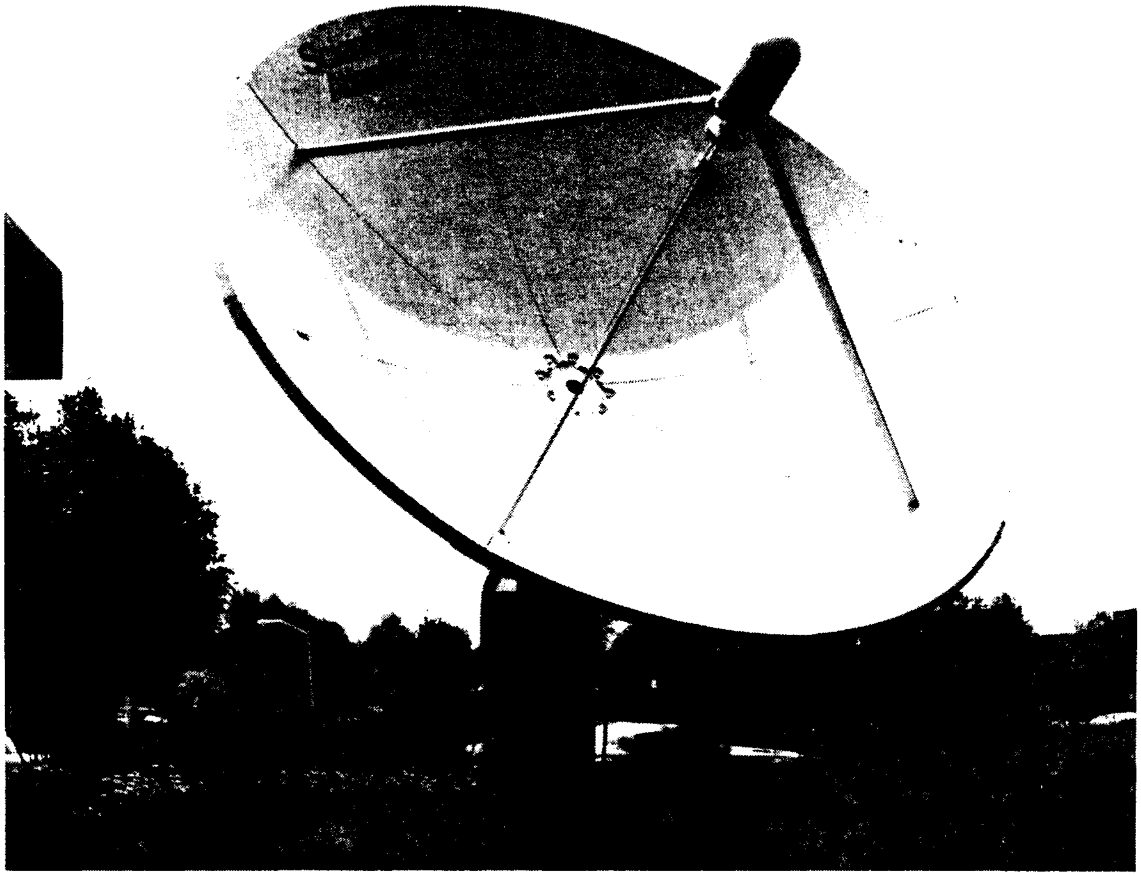
Figure 9. Descrambler Block Diagram

Various clock divider and sync circuits on the HS decoder cards provide the BR and BR/7 clock required. A phaselock loop on the high-speed decoder 2 card generates a BR clock from the BR/7 clock produced by the high-speed decoder 1 card. The BR clock drives the parallel-to-serial register and V.35 descrambler. The overall improved data state estimate at 7.68 Mb/s and the bit rate lock at the same rate are the overall output of the chassis. FEC coding provides a powerful means of enhancing the output data quality. For example, a 10^{-4} satellite channel error rate is corrected to 10^{-9}, or five orders of magnitude improvement.

Summary

The Wideband BPSK Receiver/FEC Decoder has been optimized for both performance and cost. Measured performance has been 1 dB from theoretical limits which is excellent. The design maximizes the use of high-volume, low-cost assemblies as, for example, in the downconverter. Much effort has also been expended in using simple but elegant circuitry to reduce parts count for reliability and cost. Likewise, adjustments have been eliminated whenever possible. The result speaks for itself in cost and performance.

Figure 10 shows a photograph of the engineering prototype Receive-Only Digital Audio Terminal. The wideband receiver is in one drawer and the digital processing unit is in a second drawer. The wideband receiver receives 4-GHz signal plus noise and has as its output data and clock. The DPU takes the wideband data and clock and has as its output either high-quality audio, voice cues, or data.



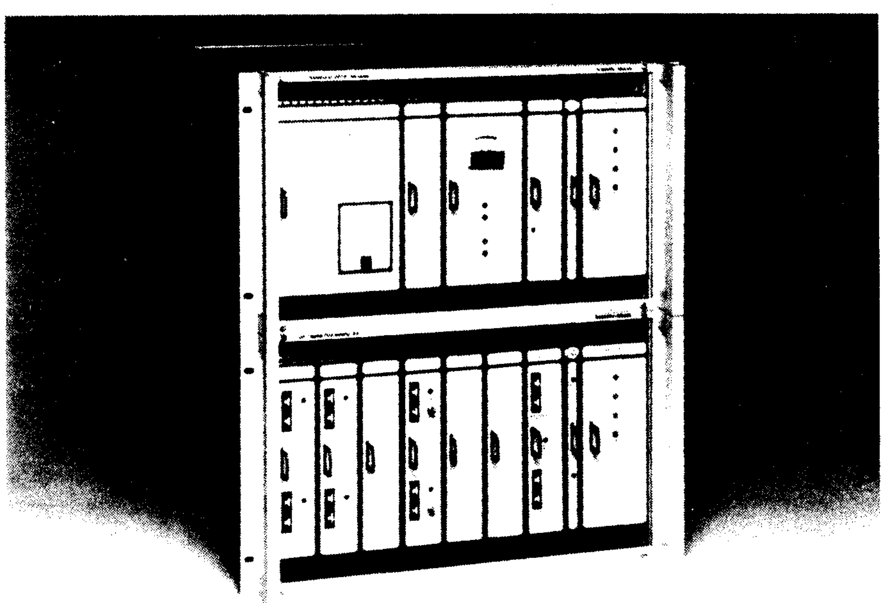


Figure 10. Digital Audio 3-Meter Earth Station

Digital Audio System Multiplexing Equipment

J.W. Chamberlin

Introduction

This paper presents the equipment which makes up the Network Radio Audio Satellite Distribution System. The discussion involves a description of how each piece of equipment functions as part of the complete system. The presentation is made in an incremental fashion, in which the basic system concept is considered first and additional requirements are dealt with in subsequent sections.

Basic System

The block diagram in Figure 1 shows the basic system which will accept program audio, cue and data inputs at the network studio, and distribute them to the network affiliates. The equipment detailed in this diagram is provided by Scientific-Atlanta. The terrestrial communications link and the uplink RF equipment are provided by the common carrier. The network source material is inserted into the digital audio system at the network studio. The material is digitized, then digitally transmitted over a common-carrier-supplied terrestrial communications link to the transmit satellite earth station. At the earth station, the data streams, often from several networks, are combined together into one composite data stream which is modulated and then transmitted to the satellite via the common-carrier-supplied uplink RF equipment. At the network-affiliated radio station, the data stream is received from the satellite and demodulated, divided into data streams corresponding to each channel, and then converted back to analog waveforms.

The network studio contains the studio link multiplexer, the channel units and a stable clock source. The studio link multiplexer accepts four 384-kb/s inputs and combines them with synchronization information to form a T1 (1.544-Mb/s) output, which is compatible with readily available commercial equipment used in the terrestrial microwave link.

The channel units interface the program audio, voice cue and data inputs with the multiplexer. Four types of channel units exist. Program audio units are available as dual 15-kHz and dual 7.5-kHz units. Auxiliary units are available as voice cue and data units. The channel units are housed in the same chassis with the multiplexer. When the number of channels desired exceeds the capacity of one studio link multiplexer, additional studio link multiplexers and channel units are added at the studio.

The 384-kb/s inputs each can be used in any one of four fashions:

a. Support one 15-kHz program audio channel

b. Support up to two 7.5-kHz program audio channels

c. Support up to twelve 32-kb/s auxiliary channels

d. Support one 7.5-kHz program audio channel and up to six 32-kb/s auxiliary channels.

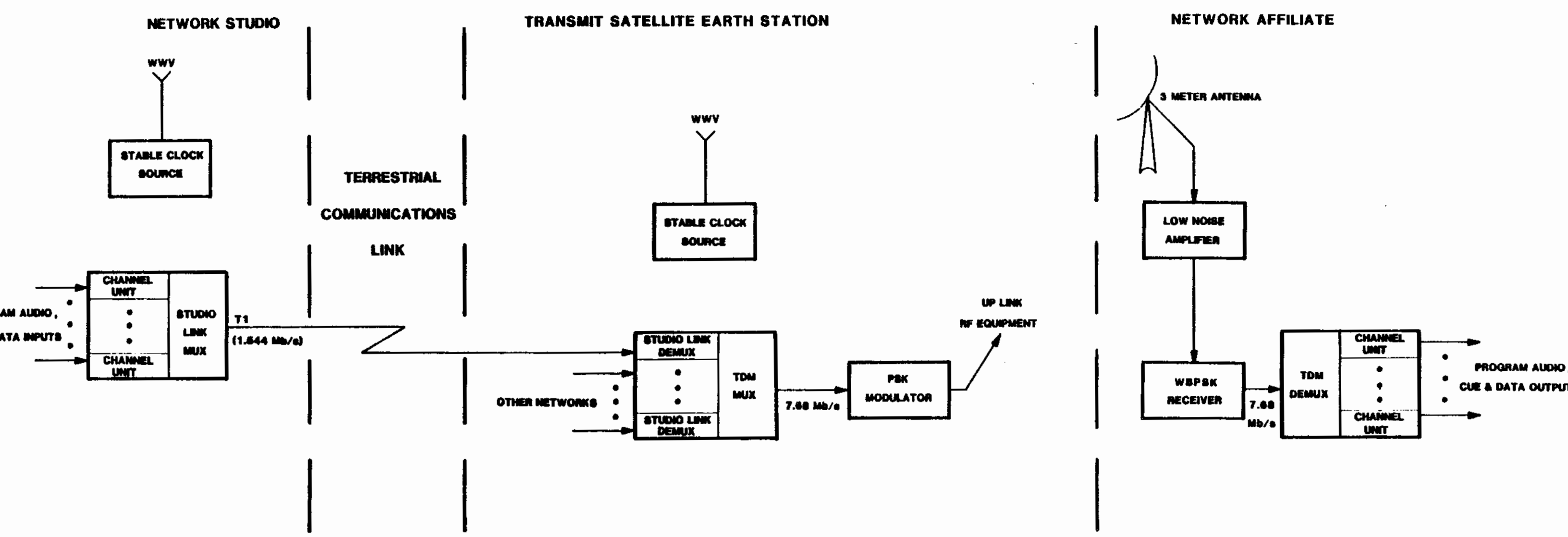


Figure 1. Digital Audio System Concept Simplified Block Diagram

The terrestrial communications link supplied by the common carrier provides a T1 input port at the studio and a T1 output port at the earth station. It is envisioned that the link will be implemented with a combination of cable and microwave services. The terrestrial communications link must have a maximum BER of 1×10^{-8} with a 0.99999 availability (including power outages).

The T1 data stream is routed to a studio link demultiplexer at the transmit satellite earth station. This unit separates the T1 data stream into its component channels. This demultiplexed data is sent to the TDM multiplexer, where it is combined with the outputs of other studio link demultiplexers to form one 7.68-Mb/s data stream. The TDM multiplexer and the studio link demultiplexers which it serves are contained in the same chassis. The TDM multiplexer has the capacity to handle nineteen complete 384-kb/s channels and one additional 384-kb/s channel, with one 32-kb/s slot reserved for system synchronization.

The T1 inputs to the studio link demultiplexers will often originate at several geographically separated network studios. This constraint requires that the data rates of each T1 link be closely matched, which requires that each network studio and the transmit satellite earth station have a very stable clock. This is provided by the stable clock sources which derive their long-term stability from the WWV broadcasts provided by the National Bureau of Standards.

The 7.68-Mb/s output of the TDM multiplexer is routed to the PSK modulator, which adds error correction coding information and then converts the data stream to a biphase-modulated signal in the 70-MHz I.F. band. This is routed to the common-carrier-supplied-RF uplink equipment for transmission to the satellite.

The return signal from the satellite is received at each network affiliate earth station by the 3-meter antenna. The received signal is amplified by a low-noise amplifier (LNA) which is mounted at the focal point of the antenna. The LNA output passes through an inter-facility link cable (IFL) to the wide-band PSK (WBPSK) receiver. This unit downconverts the high-frequency signal from the satellite to a signal in the 70-MHz band and then demodulates to 70-MHz signal and performs error correction decoding to recover the 7.68-Mb/s data stream. The 7.68-Mb/s data stream is applied to the TDM demultiplexer where it is separated into individual data streams which are routed to the receive channel units. These channel units correspond in type to the transmit channel units at the network studio. The channel units convert the data streams back to the original program audio, voice cue and data signals which were digitized at the network studio.

Headend Equipment with Monitoring

Properly functioning "head-end" equipment (the equipment located at the network studio and the transmit satellite earth station) is essential to the operation of the network. Thus, it is crucial that any equipment failures be detected as rapidly as possible. This detection can be best accomplished by providing the capability to continuously monitor all transmissions at the network studio and at the transmit satellite earth station. The block diagram in Figure 2 shows the head-end equipment portion of Figure 1, with the addition of the program monitoring equipment. The network studio is provided

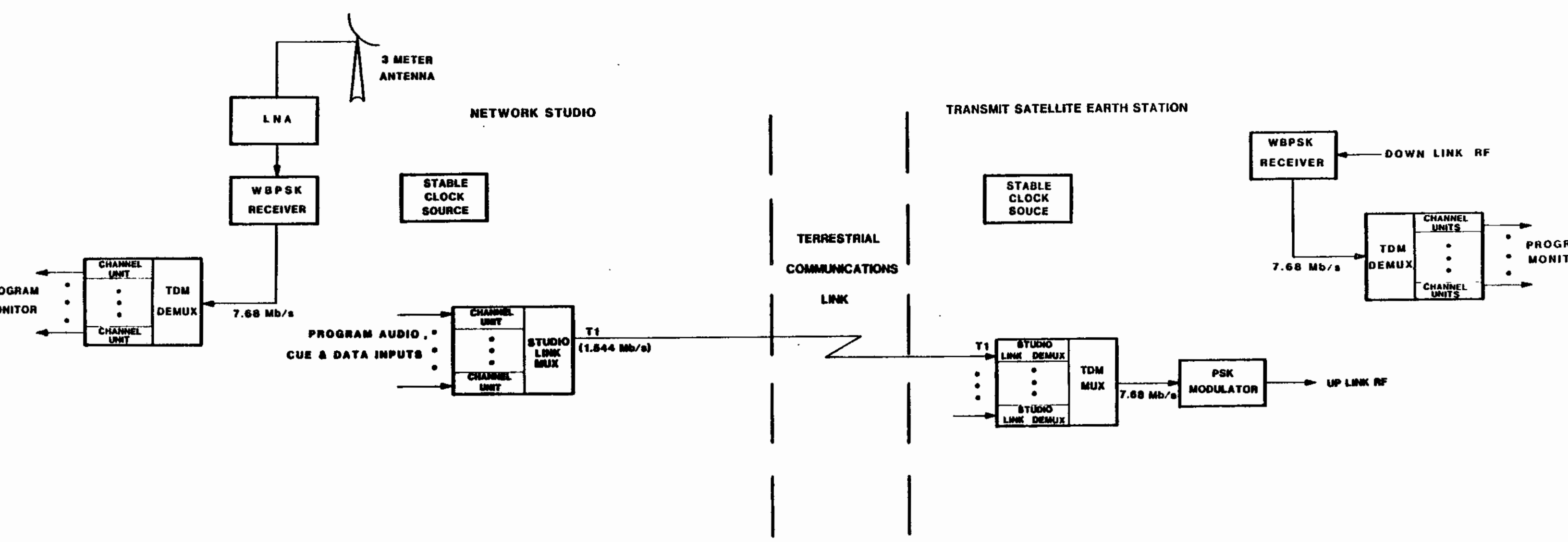


Figure 2. Digital Audio Transmit Systems Concept with Monitoring

with the same receive-only earth station equipment found at the network affiliate radio stations. The transmit satellite earth station has the same monitoring equipment with the exception of the antenna and the LNA, which are already in place.

Headend Equipment with Redundancy Protection

The availability of the system can be greatly improved by adding duplicates of key equipment, which can be switched on-line in the event of primary unit failure. In the block diagram of Figure 3, a second studio link multiplex unit has been added to the network studio.

The outputs of both studio link multiplexers are routed to a studio link multiplexer protection switch which will select the output of one of the multiplexers on command. A studio link demultiplex unit (which contains a studio link demultiplexer and receive channel units) has been added to the network studio equipment to allow maintenance to be performed on the multiplexer which is off-line. In order to further improve the availability of the system, an uninterruptible power source should be supplied by the common carrier for the studio. This unit should be capable of supplying eight hours of power for some studio equipment (e.g., tape decks), the Scientific-Atlanta studio equipment and the common carrier equipment located at the studio.

At the transmit satellite earth station, the same strategy has been used. A second TDM multiplex unit has been added in parallel with the first. The outputs of the TDM multiplexers are routed to a TDM multiplexer protection switch that functions in the same fashion as the studio link multiplexer protection switch. The dual outputs of the protection switch are routed to a pair of BPSK modulators, which drive separate uplink RF chains. A studio link demultiplex unit with channel units is provided to allow monitoring of the T1 lines for maintenance purposes. As in the satellite earth station, an uninterruptible power source is provided by the common carrier.

Headend Equipment with Back Haul

The system which will initially be installed is not expected to have a back-haul link, but will be compatible with the later addition of this capability. The back-haul signal is received at the transmit satellite earth terminal, in order to take advantage of the earth terminal's large antenna in minimizing the usage of satellite power. As shown in Figure 4, the output of the down-link RF equipment is applied to the back-haul demodulator which recovers the T1 data stream. The T1 output of the demodulator is routed through a patch panel to the terrestrial communications link. The patch panel is provided to allow the back-haul to be directly retransmitted without going to the studio (in case of emergency). At the network studio, the data stream is applied to a studio link demultiplex unit which contains the studio link demultiplexer and receive channel units required to regenerate the program audio, voice cue and data signals.

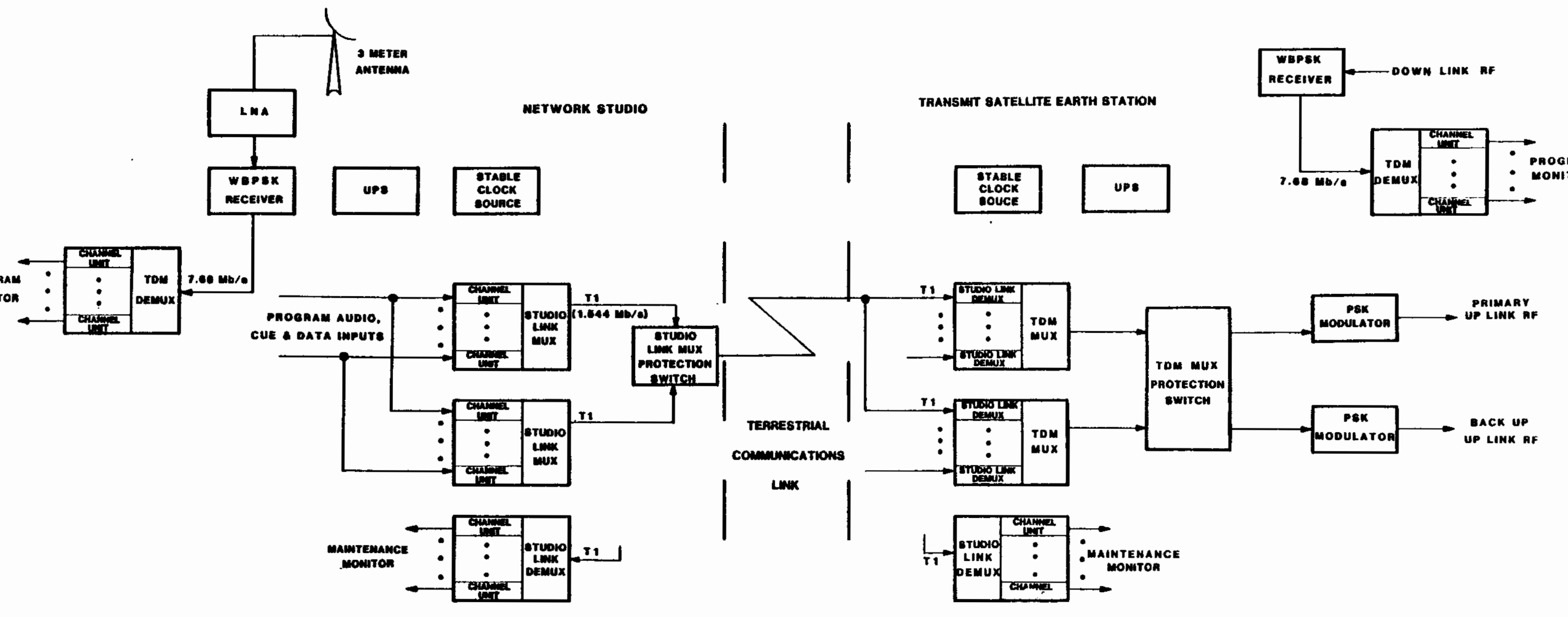


Figure 3. Digital Audio Transmit Systems Concept—With Redundancy

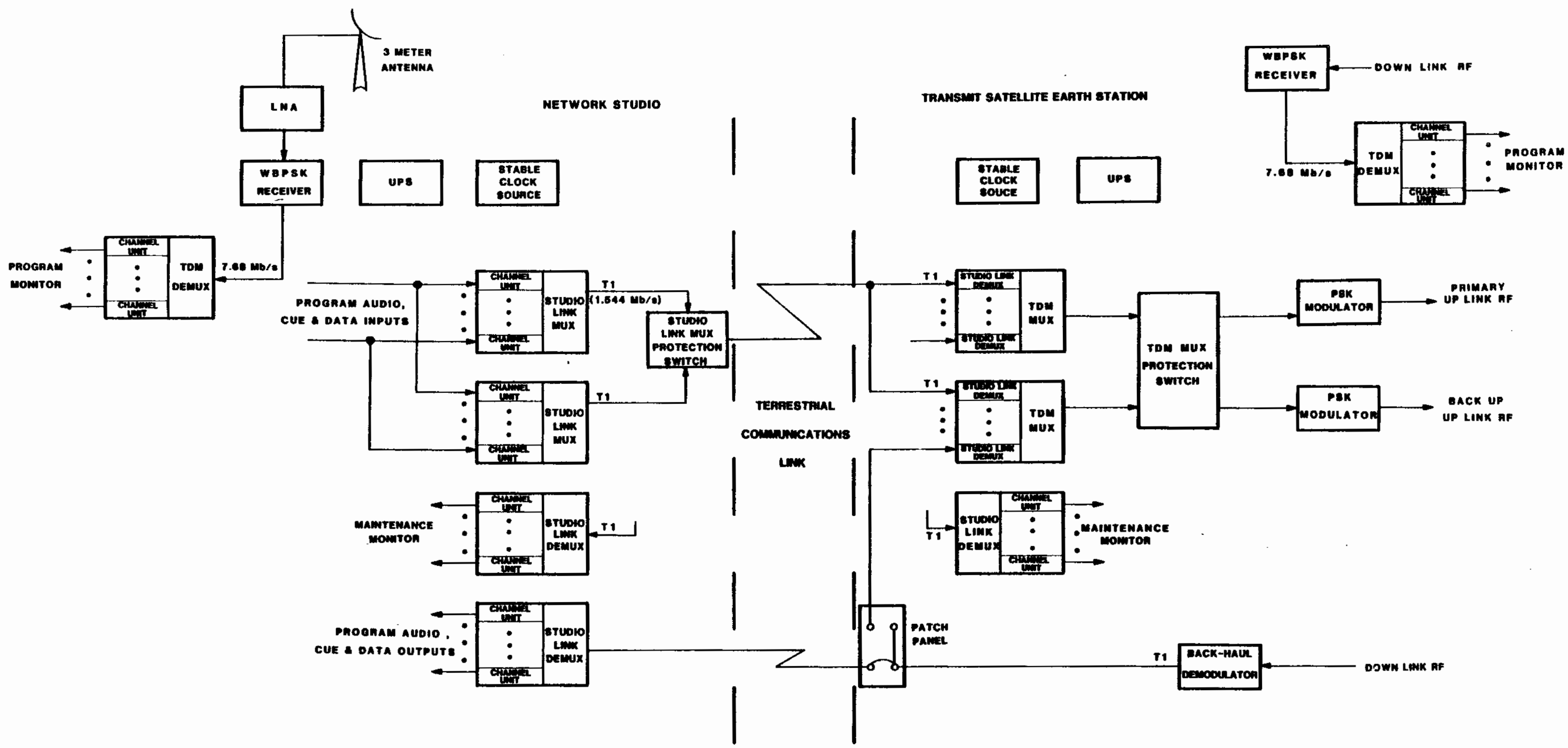


Figure 4. Digital Audio Transmit Systems Concept—With Back Haul

Digital Audio Source Encoding and Decoding for Satellite Communications

Dr. Peter G. Schreiner, III

Introduction

Digital program audio distribution via satellite offers the user the digital communications channel characteristics of superior interference immunity and regeneration without degradation, along with a new standard of audio performance. The performance limits of the system, which are fixed by the source encoding used in the channel, are tailored to the current and future requirements of network audio distribution, while striving for cost-effective excellence in a highly competitive marketplace.

The source encoding and decoding equipment is designed for the broadcast industry. Long term reliability, minimum maintenance, and special environmental conditions (i.e., transmitter site operation) have all been carefully considered in the product development.

Source Coding

To realize the advantages of a digital communications channel, an audio input source must be digitally encoded. The analog processing prior to encoding and the manner in which the analog signal source is encoded into digital words determines the maximum attainable signal-to-noise ratio (S/N), the S/N at any given input level, and the minimum data rate which is required in the channel.

The objectives in determining an appropriate source encoding for a digitally implemented program audio channel are (1) that the coded channel must perform, as well as, if not better than, a completely analog channel intended for the same purpose, and (2) that the cost of realization be minimized.

Source encoding with 15-bit linear pulse code modulation (PCM) digitally compressed to 11 bits provides system performance better than currently used analog channels in a cost-effective manner. Digital compression allows the full dynamic range of a 15-bit linear analog-to-digital (A/D) converter to be used in encoding the source, while requiring only an 11-bit representation of the input sample to be transmitted. This reduces the bandwidth required to transmit the sampled input, thereby making more efficient use of available satellite capacity.

The compression algorithm used achieves a binary coded digitally linearizable 13-segment approximation of a logarithmic compression characteristic. Figure 1 shows the compression characteristic with an 11-bit $\mu = 255$ characteristic shown for reference. Peak signal-to-quantization noise distortion (S/N_D) is 56 dB, and a dynamic range of greater than 82 dB is achievable with 15-bit converter equipment.

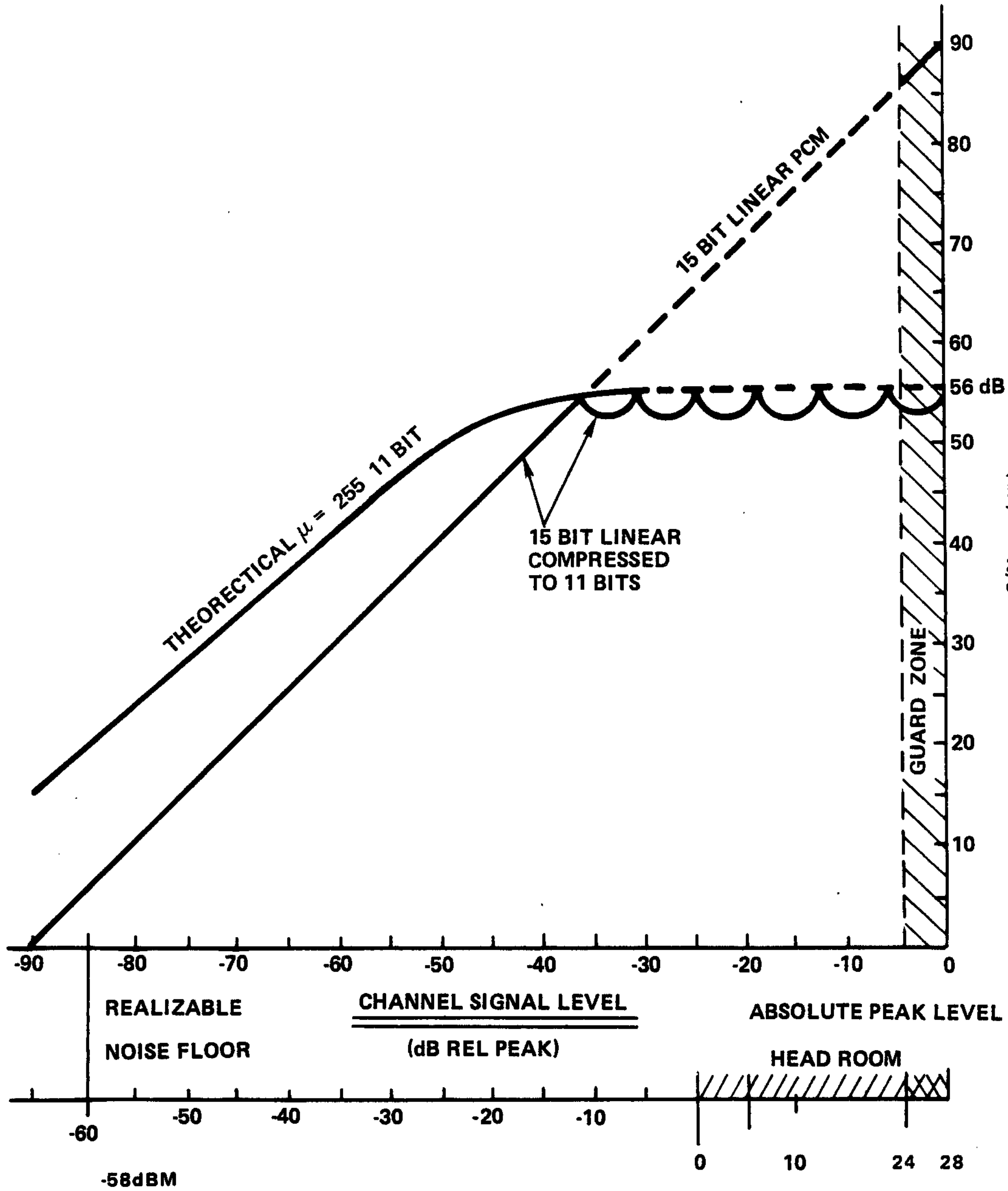


Figure 1. 15-Bit Linear PCM Compressed to 11-Bit Code Sine Wave Signal to Quantization Noise

The digital compression is implemented by converting the parallel, 15-bit, offset binary, digital output of the A/D converter to an 11-bit code with inverted, folded, binary structure. The code mapping which is implemented is presented in Table 1.

Table 1. 15-Bit Linear to 11-Bit Compressed Code

15-Bit Linear Code																11-Bit Compressed Code										
Sign 1	2	3	4	5	6	7	8	9	10	11	12	13	14	15		Sign 1	2	3	4	5	6	7	8	9	10	11
1	1	a	b	c	d	e	f	g	x	x	x	x	x	x	Cord 7	1	0	0	0	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	1	a	b	c	d	e	f	g	x	x	x	x	x	6	1	0	0	1	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	1	a	b	c	d	e	f	g	x	x	x	x	5	1	0	1	0	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	0	1	a	b	c	d	e	f	g	x	x	x	4	1	0	1	1	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	0	0	1	a	b	c	d	e	f	g	x	x	3	1	1	0	0	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	0	0	0	1	a	b	c	d	e	f	g	x	2	1	1	0	1	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	0	0	0	0	1	a	b	c	d	e	f	g	1	1	1	1	0	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$
1	0	0	0	0	0	0	0	a	b	c	d	e	f	g	0	1	1	1	1	$\bar{a}$	$\bar{b}$	$\bar{c}$	$\bar{d}$	$\bar{e}$	$\bar{f}$	$\bar{g}$

Table 1 shows the compressed code assigned to positive, 15-bit, input words (Sign bit, MSB = 1) in each of the cords of the seven-segment approximation to the positive half of the logarithmic companding characteristic.

Parallel conversion provides a reduction in the hardware and timing complexities required to implement the design. Because of the high speed of operation of the compressor, it may be shared by many channels and is therefore economically incorporated into the channel unit multiplexer.

It is noted that the compressed code limits the signal to quantizing noise ratio (S/N_D) at high signal levels (cords 2 through 7) to 56-dB peak. This is because the output word at these levels is limited to nine bits of resolution: $S/N_D = (20 \log 2^9) + 1.8$ dB = 56 dB.

However, in the first two cord segments (cords 0 and 1), the low level resolution of the 15-bit PCM code is fully preserved. This provides a low idle channel noise floor and allows the wide dynamic range required for high quality audio reproduction.

The compression characteristic provided by this algorithm is particularly appropriate for audio applications. Even for input signal levels 30 dB below the peak signal level, a S/N_D of greater than 50 dB is maintained, and the masking characteristics of the ear cause the signal to be perceived as essentially noise-free and undistorted.

With digital compression, there are no attack or release times and so the pumping effects and distortion products of analog compressors are avoided. There are also no gain errors as are commonly caused by tracking problems in analog companding systems.

The use of digital compression causes an increase in circuit complexity and introduces some noise modulation because of the gain ranging which occurs with signals whose levels transverse the higher level cords of the compression characteristic. The increase in circuit complexity is easily justified by the additional number of channels which can be transmitted through a single satellite transponder because of the bit-rate reductions which the digital compression provides.

The noise modulation which the digital compression causes is left as an artifact of the processing. The compression algorithm chosen, however, achieves signal-to-quantizing noise ratios of greater than 52 dB for signals with amplitudes in the upper 30 dB of the dynamic range of the converter. These relatively high signal-to-quanitizations noise ratios are effective in causing masking of the quantization noise in the hearing process and minimizing the perception of noise modulation.

Operation with Channel Errors

While the channel will normally operate at very low bit error rates (BER less than 10^{-8}), provision must be made to assure acceptable performance under adverse conditions.

The use of forward error correcting coding reduces the channel error rate. But ultimately, for an audio channel, it is desirable to provide additional means of ensuring that the perceptual effects of channel errors, however infrequent, are minimal.

At high channel bit error rates, digit errors may be manifested as relatively large error voltage spikes at the analog output. These voltage spikes are perceived as clicks after transduction because of their short duration and their large energy content at high audio frequencies. The use of deemphasis aids in reducing the objectionable effects of the clicks caused by channel errors and improves the channel signal-to-noise ratio. Also, the use of "error concealment" provides a signficant subjective improvement in the reconstructed audio channel at high BER. With error concealment, a parity bit is added to the data word to be transmitted and if, upon receipt, a parity error is detected, the previous data word is retained.

The structure of the digital audio source encoding affects the magnitude of errors that may be produced by bit errors. It can be shown that folded binary coding, which is used in this system, has properties which reduce the error voltages produced at high BER, especially during the critical periods of low signal levels. The folded binary code is used in many PCM applications to reduce the effects of channel errors. In this scheme, the signal is encoded into N bits, with a sign bit and N-1 bits for amplitude. This limits the maximum excursion associated with a single uncorrected bit error within a word to a 1/4 scale excursion. Errors in the Most Significant Bit MSB-(Sign Bit) result in an excursion limited to 2 X the sample amplitude.

Program Channel Encoder and Decoder Design

A detailed description of the design criteria for a dual 15-kHz channel illustrates the design philosophy employed in developing the source encoding and decoding circuitry for various channel configurations desired by the audio broadcast industry.

15-kHz Audio Channel Source Coding

The source encoding block diagram of Figure 2 will be used as an outline for the description of the signal flow and the design discussion for the channel source encoding.

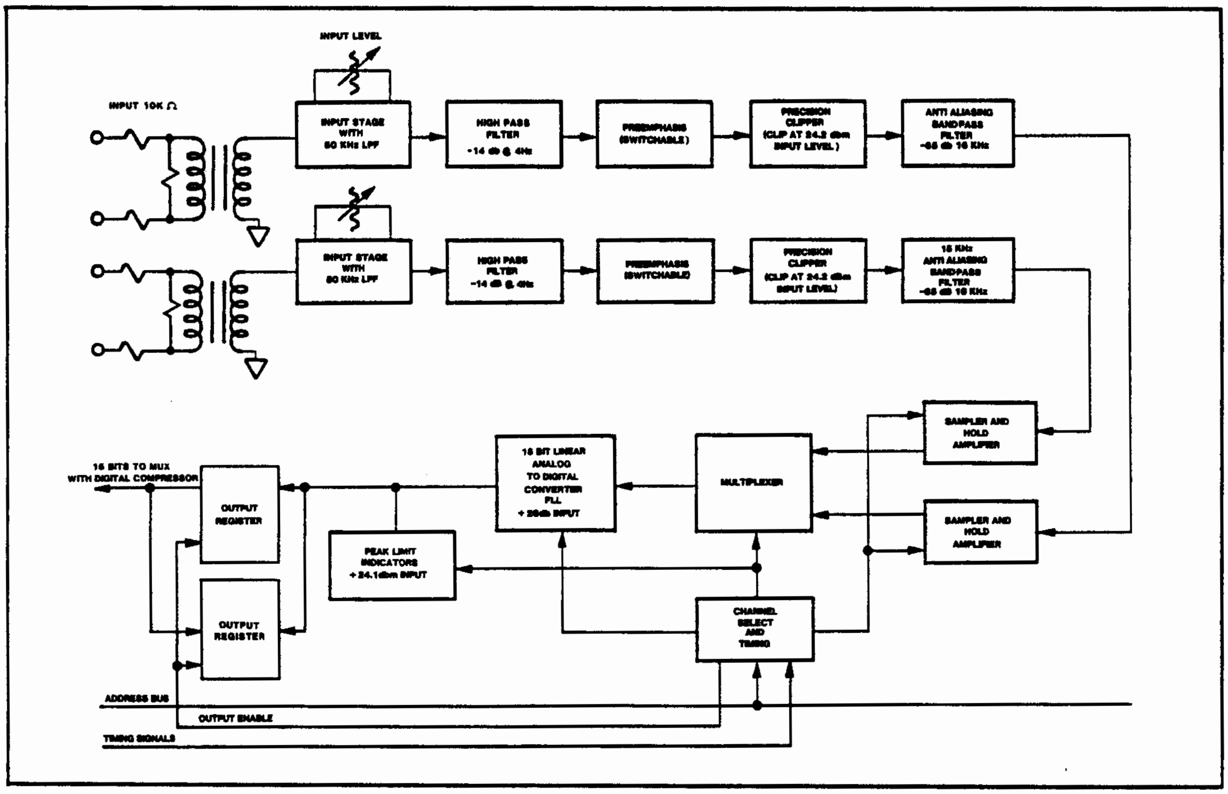


Figure 2. Source Encoding, Channel Card Block Diagram

a. Audio Input Stage. The channel unit audio input provides a balanced and floating 10K ohm termination to the audio source. The input transformer is interfaced to the audio source through a resistive pad to reduce the level of the signal passed through the input transformer. This reduction in signal level allows a smaller transformer core size while still meeting the required specifications for low total harmonic distortion (THD) with peak level low-frequency signals. The input pad also lessens the effects of interactions between reactive sources and the input impedance of the transformer.

The alloy core input transformer provides very low distortion and flat broadband frequency response. The magnetic and electrostatic shielding of the transformer provide for excellent reduction in crosstalk between adjacent input transformers and effective shielding from radiated interference present at transmitting sites. The secondary of the input transformer is single-ended and fed into the passive first pole of a 3-pole, low-pass filter. This input stage filter serves to provide an initial bandwidth limit to any out-of-band signals, such as RFI or tape recorder bias, and also to reduce the performance required of later filter stages.

The last function of the input stage is to provide the proper operating level for the channel. The only gain adjustment for the channel source encoding circuitry is in the input stage. This gain adjustment is used to set the maximum level of the input signal that can be passed through the channel.

The voltage levels in the input stage amplifier and in the stages to follow are maintained at a level that ensures that slew rate limiting does not occur and that the wide dynamic range (>120 dB) of the operational amplifiers is effectively utilized.

b. High-Pass Filter. The input low-pass filter is followed by a second-order high-pass filter with 3-dB point at 8.76 Hz. This filter is used to limit the low-frequency response of the channel so that infrasonic (subsonic) signals such as record warps, tone arm resonances and ventilation resonances cannot cause converter overload or output transformer saturation. The better than 16 dB of attenuation at 4 Hz provided by this filter is adequate protection for the channel and is perceptually undetectable.

c. Preemphasis. Switchable preemphasis conforming to CCITT recommendation J17, with -6.5 dB at 800 Hz, is provided so that the improvements in signal-to-noise and click reduction which are provided by the corresponding deemphasis network in the decoder may be utilized as a high-quality option.

d. Precision Clipper. To guarantee that the A/D converter is not driven into saturation when the input to the channel is overdriven, the conditioned input signal is precisely limited at +0.2 dB above the peak input level for the channel. A precision diode bridge clipper is used to provide symmetrical clipping with fast recovery. The clipper prevents converter saturation and has less measured or perceived distortion than a number of the more complex alternative techniques employed for FM deviation limiters.

e. 15-kHz Anti-Aliasing Filter. To bandwidth limit the audio signal to less than one-half the 32-kHz sampling frequency used for a 15-kHz audio channel, a 0.15-dB ripple, 12th-degree filter with 65-dB minimum stop band rejection is used.

The 65 dB minimum attenuation of the stop band is adequate to guarantee that the aliased components will be below the converter noise floor for all but certain rare full-level high-frequency synthetically-generated tones.

f. Sample-and-Hold Amplifier. To prevent the input voltage to the A/D converter from varying during the time it takes to complete a conversion, the filtered audio signal is sampled and held. The sample-and-hold amplifiers of the dual channel cards sample the input signals simultaneously.

The sample-and-hold and the A/D converters are operated at a 32 kHz clock rate. This sampling rate conforms to the internationally recommended standard for broadcast digital audio.

A highly reliable hybrid circuit module, which is compatible with the A/D converter, has been chosen for the sample-and-hold amplifier.

g. Multiplexer. The outputs of the sample-and-hold amplifiers are presented to the A/D converter, one at a time, under the control of the timing logic. The multiplexer allows the A/D converter to be shared by two channels.

h. Analog-to-Digital Conversion. The sample-and-hold input signal must be converted to a 15-bit digital word in less than 12 μs. It is essential the that precision of the 15-bit words, especially at low signal levels, be maintained throughout the useful life of the converter. The precision of the 15-bit word decision levels, their differential linearity, determines the ultimate noise floor and the total harmonic distortion (THD) of the system. The sample-and-hold, the A/D converter, and the multiplexer operate under the control of timing signals from the terminal multiplexer.

Since the differential linearity specification of the A/D converter is so critical for audio applications, it must be conservatively guaranteed in the converter design.

Because the soft failure of a data converter, i.e., its failure to meet specifications, is difficult to detect without involved system performance tests, and requires expensive converter replacement if it occurs, special consideration is given to the selection of this part.

i. Peak Limit Indicator. In audio transmission and reproduction, the dynamic range of the channel is used to best advantage when the peaks of the program material are adjusted just to the level where distortion begins to increase. In the digital audio channel, this level corresponds to one which causes the signal amplitude peaks to cause A/D converter saturation codes to be produced (all 1's or all 0's out of the converter). Because of the amplitude peaking of the clipped input signals occurring during input overdrive, the actual level which needs to be detected is at least 3 dB below the converter saturation level. The peak limit indicator circuits detect the PCM codes which exceed this level during any 31.25 μs sample period, and signal their presence via an LED on the front panel.

The limit indicators provide a highly precise reference level for all channel signal measurements. As the indicator signals the maximum level at which a signal will pass through the channel, it is very useful in setting up the headroom allocation and provides the reference level for S/N_D, THD and dynamic range measurements.

Peak indicators are becoming more common on all forms of audio equipment and the reasoning is clear; to obtain the maximum undistorted dynamic range of the recording or transmission equipment, it is essential to know if the normal signal level range of the equipment is being exceeded. During normal operation, the optimum drive level of the channel may easily be determined by increasing or decreasing the level presented to the channel until the limit indicator flashes infrequently on program audio peaks.

j. Output Registers and Data Bus. The 15-bit PCM words from the A/D converter are passed to holding registers and then over the parallel data bus to the terminal multiplexer and digital compressor. The digital compressor assigns an 11-bit representation for each 15-bit PCM word from the A/D converter according to the previously described algorithm.

15-kHz Source Decoding

The source decoding block diagram of Figure 3 will be used to structure the design discussion and signal flow description for a dual 15-kHz decoder.

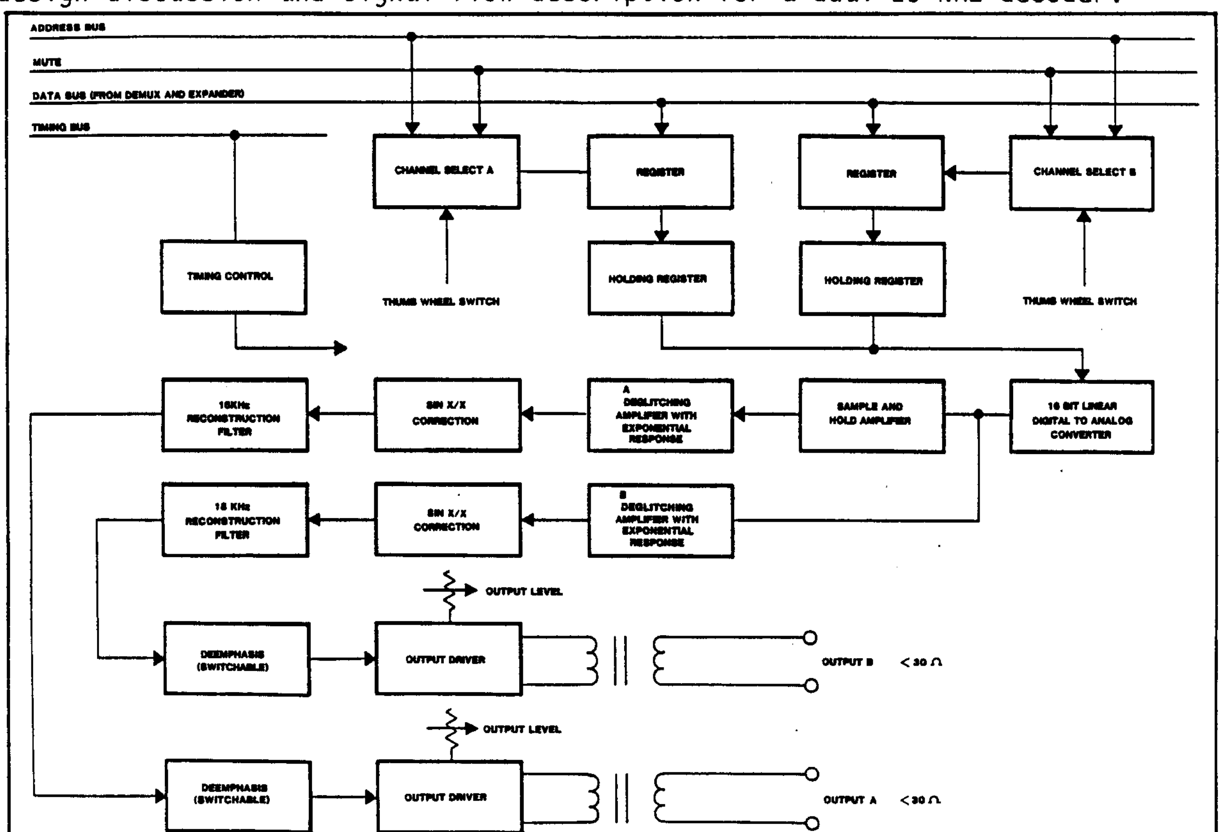


Figure 3. Source Decoding Channel Card Block Diagram

a. Channel Selection and Timing Control. Channel selection thumbwheels on the front panel of the decoder card set the codes which specify the address of the channel. The channel select logic, timing control logic, and storage registers are used to sequentially process the digitally expanded data removed from the data bus so that the two analog samples on the card may be output simultaneously.

b. Error Concealment. The mute control line from the terminal demultiplexer signals if a parity error occurred in the word received for the channel. If an error was detected, the previous valid word from the demultiplexer is retained in a register. If no error is detected, the new word is entered into the register. As previously mentioned, this form of "error concealment" serves to reduce the perceived effects of channel bit errors.

c. Digital-to-Analog Converter. The D/A converter operates on the 15-bit word from the digital expander in the demultiplxer to produce a precise analog voltage output which corresponds to a sampled value of the input signal.

   The same constraints of long-term reliability and stringent differential linearity which were required of the A/D are necessary for the D/A. The reliability of this module is at least as critical as that of the A/D, since many more D/A modules will be used in the system, and the users cannot be expected to have the technical skill required to diagnose a converter soft failure.

d. Sample-and-Hold Amplifier. For the high-speed D/A converter to be shared by two channels and allow simultaneous output of analog samples, one of the converted samples must be "held" while the other is being converted by the D/A converter.

e. Deglitching Amplifier (Distortion Suppressor). The output of the D/A contains switching transients and data transitions with rise and fall times which can cause slewing distortions. It is, therefore, necessary to condition the output steps of the D/A converter to obtain a low distortion reconstruction of the input signal.

   The rise time of the data transitions is limited by an RC time constant incorporated in the switched amplifier of the deglitching amplifier. This fixed rise time (RC = 3.4 μs) limits the slew rate of the signal and reduces the slew rate requirements of the deglitching amplifier and following stages so that slewing distortion is not encountered. The frequency response roll-off caused by the RC filter is corrected in the (Sin X)/X correction filter.

   The timing for the deglitching amplifier is obtained from the terminal demultiplexer. The amplifier samples the D/A output after its switching transients have decayed to a suitable level (<6 μs). The deglitching amplifier must be allowed to track the D/A output for at least three time constants (3RC = 10.2 μs) so that flat frequency response and high S/N are maintained through the deglitching amplifier. The output of the deglitching amplifier takes on the stepped form of a zero order hold with exponential rise and fall times.

f. (Sin X)/X Correction. The output frequency response of the sampled data systems using a zero order hold has a (sin X)/X shape. This, and the rolloff caused by the lowpass filter effect of the RC time constant employed in the deglitching amplifier, must be compensated to achieve the flat frequency response required for the digital audio system.

   A two-pole active filter provides the precise compensation necessary to correct the (Sin X)/X and RC low-pass responses to within 0.05 dB over the 15-kHz bandwidth.

g. 15-kHz Reconstruction Filter. A bandpass filter, similar to that used for the input anti-aliasing filter, is used for baseband signal reconstruction. The requirements for the reconstruction filter are nearly as severe as those for the anti-aliasing filter. High amplitude components of the source signal located at frequencies near the upper cutoff frequency of the channel are folded back in the sampled data spectrum to be presented as nonharmonic, out-of-band, audible components in the signal presented to the reconstruction filter. That is, 15 kHz at + 24 dBm input also appears as a 17-kHz component at +22 dBm. The reconstruction filter must reduce these components and other artifacts below the level of the system noise floor. It is also necessary that the reconstruction filter guarantee that no digital processing artifacts be present in the output to disrupt other equipment to which the signal is routed.

   This filter is constructed to provide reliable, EMI-immune performance.

h. Deemphasis. When the switchable deemphasis is enabled, it compensates for the use of pre-emphasis in the encoder, improves the signal-to-noise ratio for the channel, and reduces the perception of noise modulation.

i. Output Stage. The fully reconstructed baseband signal is routed to the output stage for level adjustment and buffering. The output stage has the slew rate and current drive capacity required to deliver a low-distortion replication of the reconstructed signal at +24 dBm into a 600-ohm load. An internally accessible level adjustment for the output is provided so that the peak output level may be varied. This is normally set to provide unity gain through the channel.

   To achieve low distortion at peak signal levels and low frequencies (20 Hz), a specially designed output transformer is required. The transformer frequency response is flat within 0.1 dB from 20 Hz to 50 kHz. The output amplifier is formed with a bridging amplifier and which provides a low-impedance source to the transformer. The voltage source drive for the transformer assures low crosstalk between adjacent output transformers and low distortion.

   Adequate heat sinking is provided for the output stages so that no cooling fans are required in the equipment. The output stages are unconditionally stable and protected against short circuit loading and overvoltage transients.

   The low impedance ($<30\Omega$) output is balanced and floating.

Front-Panel Controls, Indicators, and Test Points

Encoder Front Panel

The encoder front panel is shown in Figure 4. The indicators are described below.

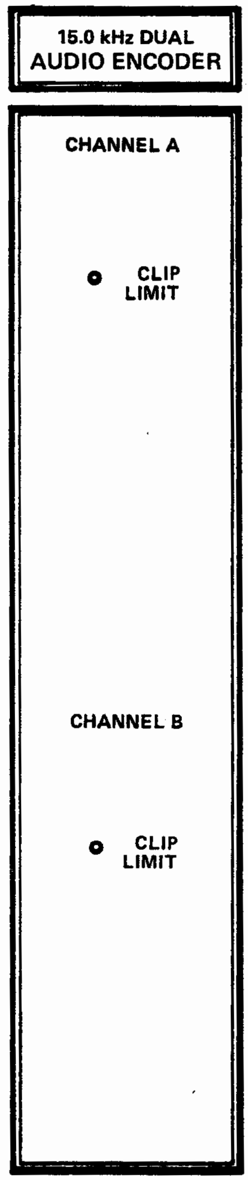


Figure 4. Encoder Channel Card Front Panel

A and B Clip Limit. LEDs indicate when the input signal possesses peaks which are greater than the clipper threshold. The gains in the channel are set so that the limit indicators are illuminated for inputs >+24.1 dB.

Decoder Front Panel

The decoder front panel is shown in Figure 5. The individual controls and indicators are described below.

a. A and B Channel Select. The channel select thumbwheels allow independent selection of any of the channels available to the Digital Program Terminal.

b. A and B Audio Present Indicators. LEDs are provided to indicate the presence of time-varying signals above a preset threshold. This actively indicates the presence of audio signals within the channel.

The operation of the error concealment and mute functions is also signaled by the Audio Present indicators. If an error occurs in a channel and is concealed, the Audio Present indicator is turned off momentarily. Muting of a channel is signaled by the LED being turned off for the duration of the mute command.

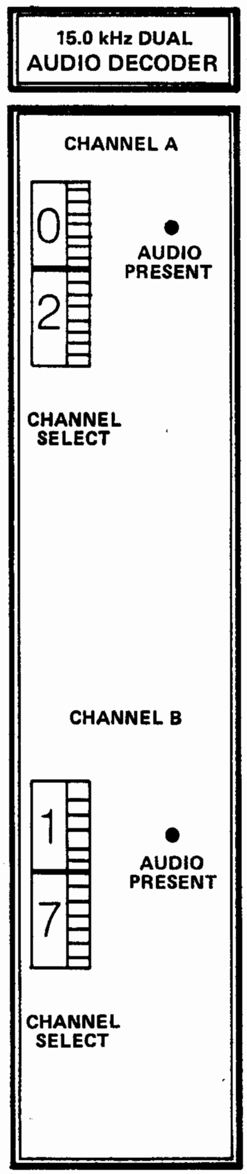


Figure 5. Decoder Channel Card Front Panel

Summary

Thorough consideration of the communication system requirements led to the implementation of the 15-bit pcm to 11-bit compressed transmission code source coding for network program audio distribution. This coding selection and a conservative hardware design provide a reliable high quality audio interface for a digital communications channel to be used for satellite audio distribution.

Landsat-D Wideband Unbalanced QPSK Demodulator/Bit Synchronizer-Signal Conditioner

Dr. James S. Gray

Theory

The LANDSAT-D satellite transmits unbalanced QPSK (UQPSK) with 84.903 Mb/s on the I channel which has 80% of the total satellite power and 15.062 Mb/s on the Q channel with 20% of the satellite power. The I channel transmits data from the thematic mapper, and the Q channel transmits the data from the multi-spectral scanner.

In Figure 1, the equations for UQPSK are presented. As seen in the expression, S(t) is the same as that for QPSK except there are factors A and B for the two quadrature carriers. In straight QPSK, A and B are equal. a(t) is the data on the higher power or I channel; b(t) represents the data on the Q channel. As shown, the A and B may be written in terms of the total satellite power, P, and ρ the fraction of the total power on the I channel. One then gets the equations on the bottom in terms of total satellite power, P, and also ρ, the fraction of power on the I channel.

$$S(t) = A\, a(t) \cos(\omega_c t + \phi) + B\, b(t) \sin(\omega_c t + \phi)$$

Where:

a(t) = ±1 IS I CHANNEL DATA - HIGHER DATA RATE WITH HIGHER POWER.

b(t) = ±1 IS Q CHANNEL DATA.

LET TOTAL POWER BE P AND ρ BE FRACTION OF P ON I CHANNEL THUS,

$$\frac{A^2}{2} = \rho P \rightarrow A = \sqrt{\rho}\,\sqrt{2P}$$

$$\frac{B^2}{2} = (1-\rho)P \rightarrow B = \sqrt{1-\rho}\,\sqrt{2P}$$

SO THAT

$$S(t) = \sqrt{2P}\,[\sqrt{\rho}\, a(t) \cos(\omega_c t + \phi) + \sqrt{1-\rho}\, b(t) \sin(\omega_c t + \phi)]$$

Figure 1. Unbalanced QPSK (UQPSK)

The vector or phasor diagram for UQPSK is presented in the left half of Figure 2. With 80% of the power in the I channel, the I channel vector is twice as long as the Q channel vector; therefore, the resultant vectors for the I and Q channel combination are 26.565° from the I axis. Also shown in Figure 2 are the vectors after the input spectrum is doubled. Notice that the four states are collapsed in two states. A very interesting question is, "Can one do carrier regeneration by squaring?" With normal QPSK, one cannot. One has to use multiplication by four or the equivalent fourth-order Costas loop. To answer this question, consider Figure 3 where the average phase vector of UQPSK after squaring is calculated. Assume that the data rate on the I channel is greater than the Q data rate and that both are much greater than the noise bandwidth B_N of the filter following the squaring. This could also be the noise bandwidth of the phaselock loop. Let P_I be the probability of ones on the I channel, and P_Q be the probability of ones on the Q channel; derive the expression shown for the average phase vector after squaring. To provide a coherent reference for demodulation, this vector is divided by two which results in the average carrier reference phase vector expression shown in the figure. If the data is random on either channel, the average phase vector is zero and, therefore, the average phase error is zero. Thus, one can use squaring as a mechanism of recovering the carrier for UQPSK. As will be shown later, this has some very significant advantages under conditions of modulator phase imbalance. This technique greatly reduces the crosstalk produced by phase imbalance and the resulting degradation of performance from theory.

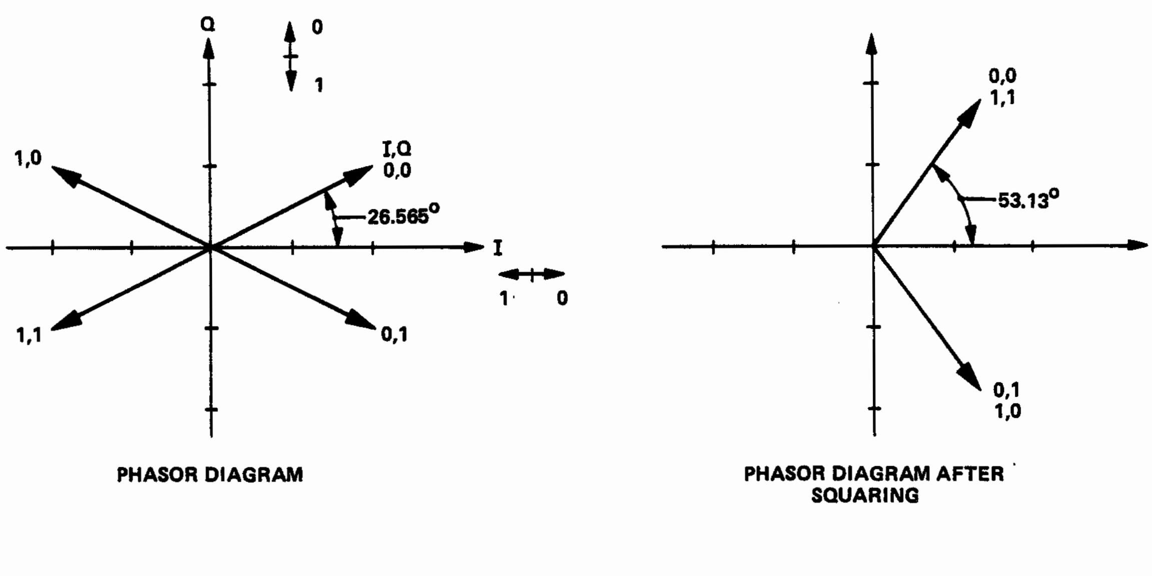


Figure 2. UQPSK Vector Diagrams

ASSUME $R_I > R_Q >> B_N$

B_N NOISE BANDWIDTH OF FILTER AND/OR QLL FOLLOWING SQUARING.
WHAT IS AVERAGE PHASE ERROR IF STATE PROBABILITIES ON I AND Q CHANNELS DO NOT EQUAL 1/2?

LET P_I BE PROBABILITY OF ONES ON I CHANNELS

P_Q BE PROBABILITY OF ONES ON Q CHANNEL

AVERAGE PHASE VECTOR AFTER SQUARING IS

$$[2(2P_I P_Q - P_I - P_Q) +1] (53.13) \text{ DEG}$$

AFTER DIVIDING BY TWO, THE AVERAGE PHASE VECTOR IS

$$[2(2P_I P_Q - P_I - P_Q) + 1] (26.565) \text{ DEG}$$

IF $P_I = 1/2$ $\quad 2(P/_Q - 1/2 - P/_Q) + 1 = 0$

IF $P_Q = 1/2$ $\quad 2(P/_I - P/_I - 1/2) + 1 = 0$

THUS, IF EITHER CHANNEL IS UNBIASED, AVERAGE PHASE ERROR IS ZERO.

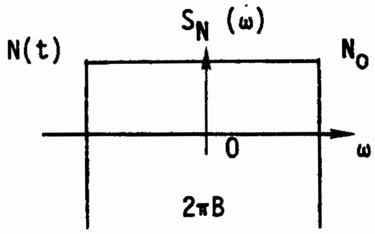
Figure 3. Average Phase Vector after Squaring of UQPSK

To examine the effects of noise on the carrier regeneration process, consider Figure 4. In Figure 4, at the top of the figure, the expression for UQPSK plus bandpass noise centered in a filter B is given. The figure shows the model for carrier regeneration. The input signal-to-noise ratio goes through a filter in front of the nonlinearity to constraint the bandwidth of the noise, gets multiplied by two in the times-two circuitry, and then is routed through a narrowband filter B_N where B_N represents the noise bandwidth. It is desirable to derive an expression for the signal-to-noise ratio at the output of the narrow filter B_N. The noise is modeled as shown in the diagram where it is assumed to be a constant noise spectral density N_O over a bandpass bandwidth noise bandwidth B around the center frequency of the carrier. One can then derive the expression for the output signal-to-noise ratio as shown in the figure. If one defines the input signal-to-noise ratio in terms of the total power divided by N_O times the noise bandwidth B for the input filter, and then lets the input bandwidth B be approximately equal to R_I (the I channel data rate) to get a reduced complexity expression, one may derive the two expressions shown at the bottom of the page. The absolute expression is exactly the same as that for biphase PSK (BPSK) with squaring except for two additional terms. The numerator has $(2\rho-1)^2$ factor multiplying the input signal-to-noise ratio. This takes into account the fact that not all the power is in the I channel. The denominator of the expression looks like the normal BPSK squaring expression except there is an extra term which is the $\rho(1-\rho)$ $(1-R_Q/3R_I)S_i/N_i$ term. This term is the result of signal-times-signal crosstalk introduced by the UQPSK process. Rewriting everything in decibel form, one then derives the expression on the bottom of the page which again looks like the standard squaring type expression with BPSK except there is the 10 log $(2\rho-1)^2$ expression, and, in the final term, there is the extra factor in the denominator that has to do with the SXS crosstalk. It is very important to note from the original expression derived that the desired double-frequency carrier term is proportional to P^2 and that the self-noise term introduced by signal-times-signal products is also proportional to P^2. Several papers have been written that show that as the signal-to-noise improves, the jitter at the output of a phaselock loop with noise bandwidth B_N actually increases.

With a phaselock loop as the desired signal term increases due to P^2 increasing, the noise bandwidth of the phaselock loop is increased. Similarly, the self-noise spectral density due to the signal-times-signal products is also increasing. Thus, the output noise variance of the loop increases. If one uses a limiter in front of the loop, however, one sees that the signal-to-noise ratio coming out of the limiter will be constant since both the desired signal term and the noise term are proportional to P^2. Therefore, a limiter in front of the phaselock loop not only keeps the loop parameters constant, but also keeps the signal-to-noise constant into the loop and, therefore, the variance of the loop output jitter constant under high signal-to-noise conditions.

$$\sqrt{2P}\,\sqrt{\rho}\;a(t)\;\cos\,(\omega_c t + \phi) + \sqrt{2P}\,\sqrt{1-\rho}\;\sin\,(\omega_c t + \phi) + \sqrt{2}\;N(t)\;\cos\,(\omega_c t + \phi)$$

$$SNR_i \longrightarrow \boxed{B} \longrightarrow \boxed{X2} \longrightarrow \boxed{B_N} \longrightarrow SNR_o$$



$$SNR_o = \frac{1/2\;P^2(2\rho-1)^2}{2P\rho N_oB_N + 2P(1-\rho)N_oB_N + N_o{}^2\;BB_N + 2P^2\rho(1-\rho)(1-R_Q/3R_I)(B_N/R_I)}$$

IF DEFINE $SNR_i \triangleq \frac{S_i}{N_i} = \frac{P}{N_oB}$, THEN FOR $B \approx R_I$ ONE MAY WRITE

$$SNR_o = \frac{S_o}{N_o} = \frac{1}{4}\cdot\frac{B}{B_N}\cdot\frac{\frac{S_i}{N_i}\,(2\rho-1)^2}{1 + \frac{1}{2\;S_i/N_i} + \rho(1-\rho)\left(1-\frac{R_Q}{3R_I}\right)\frac{S_i}{N_i}}$$

$$10\;\log\;SNR_o =$$

$$10\;\log\frac{B}{B_N} - 6\;\mathrm{dB} + 10\;\log\;(2\rho-1)^2 + 10\;\log\frac{S_i/N_i}{1 + \frac{1}{2\;S_i/N_i} + \rho(1-\rho)\left(1-\frac{R_Q}{3R_I}\right)\frac{S_i}{N_i}}$$

Figure 4. Carrier Regeneration by Squaring for UQPSK

In Figure 5, the conditions for LANDSAT-D, that is ρ equals 0.8 and $\frac{R_Q}{R_I}$ approximately equals $\frac{15}{85}$ are inserted into the equations previously derived for the signal-to-noise ratio out of the carrier regeneration process. The resulting expression is that shown. At high signal-to-noise ratios,

$$10 \log SNR_o = 10 \log \frac{B}{B_N} - 2.23 \text{ dB}$$

FOR $\rho = 0.8 \quad \frac{R_Q}{R_I} = \frac{15}{85}$

$$10 \log SNR_o = 10 \log \frac{B}{B_N} - 6 \text{ dB} - 4.43 \text{ dB} + 10 \log \frac{S_i/N_i}{1 + \frac{1}{2 S_i/N_i} + 0.1504 S_i/N_i}$$

AT HIGH S_i/N_i THIS BECOMES

$$10 \log SNR_o = 10 \log \frac{B}{B_N} - 2.23 \text{ dB}$$

Figure 5. Carrier Regeneration Conditions for LANDSAT D

This equation allows one to calculate how narrow the noise bandwidth B_N must be in order to have an adequate coherent reference signal-to-noise ratio for coherent demodulation. These results are plotted in Figure 6. The normalized expression $10 \log S_o/N_o - 10 \log B/B_N$ is plotted on the ordinate versus $10 \log S_i/N_i$ on the abscissa. This plot limits at minus 2.23 dB due to the SXS self-noise term. This result is used as a guideline to choose the noise bandwidth B_N such that an adequate carrier reference for coherent demodulation is obtained so that the UQPSK demodulator performs close to theory. Noise creates variance of the carrier reference zero crossings, which causes signal variation and crosstalk between the I and Q channels. This degrades the resulting bit error rate performance.

$\rho = 0.8$; $R_Q/R_I = 15/85$;

$$10 \log SNR_O - 10 \log B/B_N = -6 \text{ dB} + 10 \log (2\rho-1)^2 +$$

$$\left[+10 \log \frac{S_i/N_i}{1 + \frac{1}{2\, S_i/N_i} + \rho(1-\rho)\left(1 - \frac{R_Q}{3\, R_I}\right) S_i/N_i} \right]$$

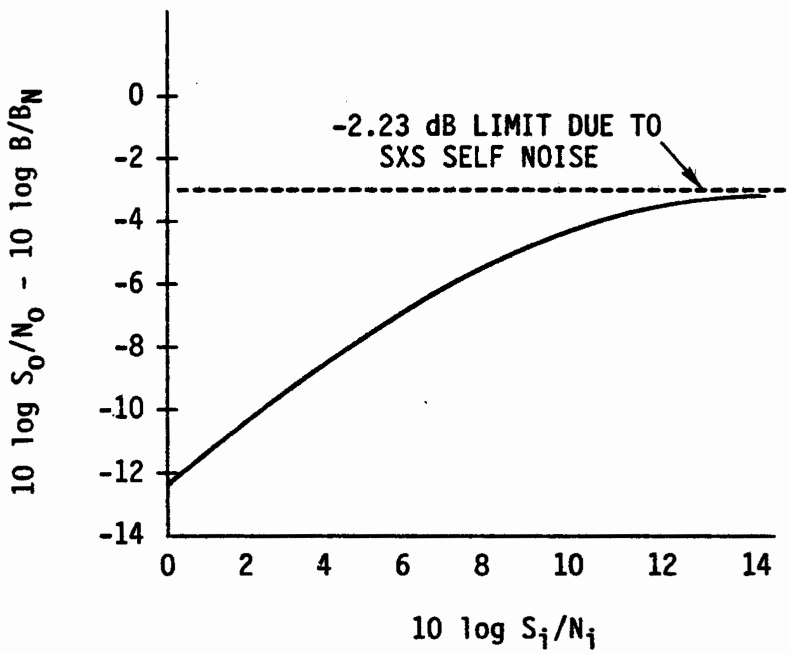


Figure 6 Plot of LANDSAT D Carrier Regeneration Equations

In Figure 7, plots are provided of the required carrier reference signal-to-noise ratio for a given loss from theory. The signal-to-noise ratio of the carrier reference, is defined as $10 \log \frac{1}{\bar{\theta}^2}$ where $\bar{\theta}^2$ is the variance of the phase noise of the reconstructed coherent reference for demodulation. As presented for BPSK, QPSK and Offset QPSK, these curves can serve as guidelines. For example, the plot on the right shows that signal-to-noise is required at various error rates for a loss due to carrier regeneration phase noise of 0.2 dB. The middle plot is for a 0.10 dB loss, and the plot to the left is for a 0.05 dB loss. As seen, conventional QPSK and Offset QPSK require much higher signal-to-noise ratios than BPSK. This is due to the crosstalk between the two channels. A similar curve can be derived for UQPSK.

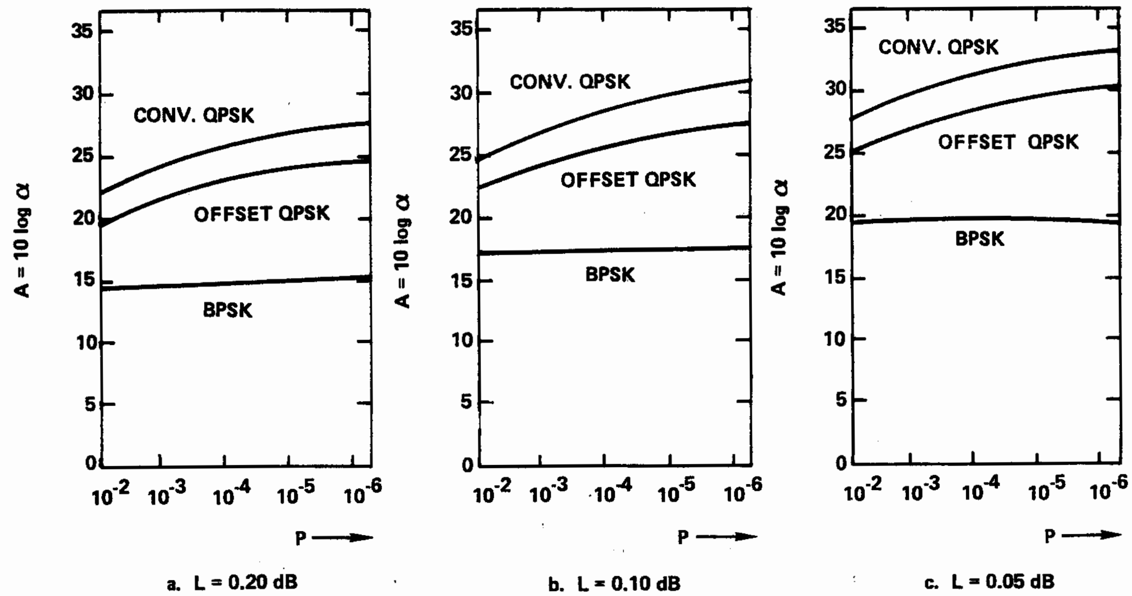
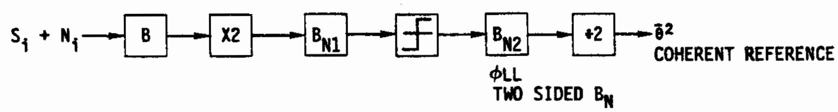


Figure 7. Required SNR $\triangleq$ 10 Log $\frac{1}{\theta^2}$ for BPSK, QPSK, and UQPSK

For UQPSK, the crosstalk of the large I channel over into the small Q channel due to noise or phase imbalance is the most disastrous effect. Curves derived for UQPSK have to show the parameter ρ, as well as the bit error rate, since the relationship is very strongly a function of ρ. Obviously, the closer ρ is to one, the less the large channel will be affected by crosstalk from the small channel; but, more disastrously, the small channel will be very affected by crosstalk from the large channel. One can use the QPSK curves as a starting point and add in extra margin as a function of the value of ρ to handle UQPSK or, correspondingly, if one has the time, one can derive a complete family of curves of the required carrier reference signal-to-noise ratio for a certain performance loss as a function of input signal-to-noise ratio (or equivalently bit error rate) and ρ. A sample carrier regeneration analysis for UQPSK using the LANDSAT-D parameters is presented in Figure 8. Assuming that the minimum bit error rate is 10^{-2}, this implies approximately 5 dB $\frac{E_b}{N_o}$ on the I channel if one allows some implementation loss. Assuming a filter bandwidth, B, roughly proportional to the bit rate, R_I, about a 6-dB signal-to-noise ratio out of the filter B results. From our previous curves, a signal-to-noise ratio for the coherent reference of 25 dB is assumed for a 0.1-dB loss due to phase reference noise. If one inserts this in the previous equations, noting that there is a divide-by-two following the phaselock loop or the narrowband filter B_N, then one derives for the LANDSAT parameters that the noise bandwidth B_N needs to be approximatley 446 kHz. This noise bandwidth for the phaselock loop implies an f_N for the loop of approximately 56 kHz with a damping of one. To meet other modem performance requirements, and to allow for effects of UQPSK operation, the final f_N of the loop used is more like 8 to 10 kHz.



10^{-2} BER $\rightarrow \sim 5$ dB $E_b/N_{0/I}$ $\rightarrow \sim 6$ dB S_i/N_i

at 10^{-2} REQUIRE $10 \log \frac{1}{\bar{\theta}^2} = 25$ dB FOR 0.1 dB LOSS DUE TO PHASE REF.

$$25 \text{ dB} = -10.43 \text{ dB} + 10 \log \frac{S_i/N_i}{1 + \frac{1}{2\, S_i/N_i} + \rho(1-\rho)\left(1 - \frac{R_Q}{3R_I}\right) S_i/N_i} + 10 \log \frac{B}{B_{N1}} + 3 \text{ dB}$$

$$+ 10 \log \frac{B_{N1}}{B_{N2}} + 6 \text{ dB}$$

$$= -6.8 \text{ dB} + 10 \log \frac{B}{B_{N1}} + 9 \text{ dB} + 10 \log \frac{B_{N1}}{B_{N2}}$$

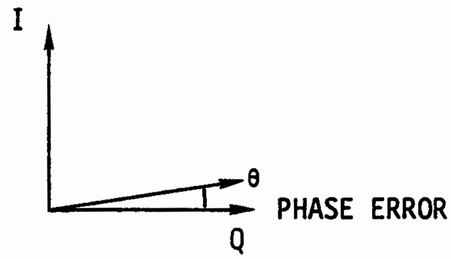
$$22.8 \text{ dB} = 10 \log \frac{B}{B_{N2}} = 10 \log R_I/B_{N2}$$

$$R_I/B_{N2} = 190.546$$

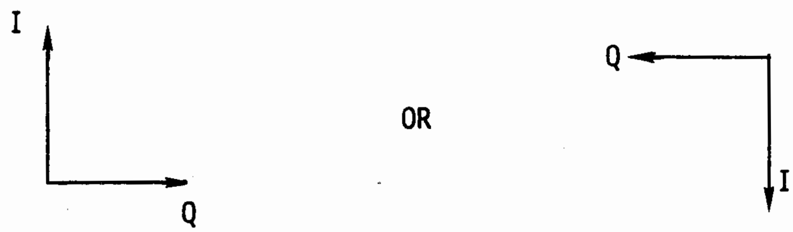
$$R_I = 85 \text{ Mb/s} \longrightarrow B_{N2} = 446 \text{ kHz}$$

Figure 8. Sample Carrier Regeneration Angles for UQPSK

It has been previously alleged in this paper that using the squaring technique for carrier regeneration for UQPSK has advantages for demodulation when there is a phase error in the modulator. This is illustrated in Figure 9. The transmitted vectors are such that the I vector can be considered to be lined up on the actual I axis, but assume that the Q vector is several degrees from the actual quadrature condition so that there is a phase error θ as shown. The reconstructed vectors, however, will be 90° apart. Since we are locking with the squaring technique strictly to the I channel, the I channel vector lies on the I channel axis and has zero phase error with the transmitted I vector. The resultant demodulated output will be I′ on the I channel which will be the desired ±I term with a ±Q sin θ crosstalk term. Thus, there is a small crosstalk of the small channel into the large channel. The resulting Q′ output will be ± 0, that is no crosstalk from the I channel plus ± Q Cos θ or a small reduction in amplitude in the Q channel. The really important result here is that half the crosstalk terms one normally gets when there is a phase error have gone away. That is, the I channel desired term is not reduced in amplitude and the crosstalk from the other channel is from the small Q channel into the strong I channel. On the channel with the smaller amount of power, that is the Q channel, there is no crosstalk from the larger power I channel into the Q channel. Normally there is ± I sin θ crosstalk term in the Q channel. Since the I channel is a much larger value than the Q channel in UQPSK, this can have very disastrous results. Therefore, since this technique basically locks onto the I channel, half the normal crosstalk terms go away, and the performance of this system is quite excellent with modulator phase imbalance. Using fourth-order techniques for carrier regeneration, either times-four or fourth-order Costas loop, one can get up to 3-dB loss for a 3° modulator phase imbalance depending on the assumptions one makes. As will be seen in the measured results section, the performance for this system is excellent under this condition of imbalance.



TRANSMITTED VECTORS



RECONSTRUCTED VECTORS FOR DEMODULATION

RESULTING OUTPUT

$$I' = \pm I \quad \pm Q \sin\theta$$
$$Q' = \pm 0 \quad \pm Q \cos\theta$$

IF x4 OR 4TH ORDER COSTAS LOOP

$$I' = \pm I \cos\theta \pm Q \sin Q$$
$$Q' = \pm I \sin\theta \pm Q \cos\theta$$

Figure 9. Resulting Output Due to Modulator Phase Imbalance

Hardware Description

A block diagram of the UQPSK Demodulator/Bit Synchronizer-Signal Conditioners (BSSCs) is presented in Figure 10. In the demodulator portion, the IF signal-plus-noise is automatic-gain-control amplified, bandpass-filtered, and then routed in parallel to the carrier regeneration circuitry and the coherent detector. The carrier regeneration circuitry nonlinearly reconstructs a coherent reference for demodulation which is routed to the coherent detector. In the coherent detector, in-phase and quadrature demodulation occurs. The noisy baseband outputs are the inputs to the 84.903-Mb/s bit synchronizer-signal conditioner (BSSC) and the 15.062-Mb/s BSSC. In the 84.903-Mb/s signal conditioner, signal-plus-noise is optimally filtered and then state-estimated. The matched filter output is also routed to the bit synchronizer where an even-order nonlinearity produces a bit rate spectral component. A phaselock loop is locked to this component, and the VCO output is both an overall BSSC output plus controls the state-estimation in the signal conditioner. On the Q channel similar processing occurs on the 15.062-Mb/s data stream plus noise. More detailed block diagrams and descriptions of the various components of the overall block diagram are now presented.

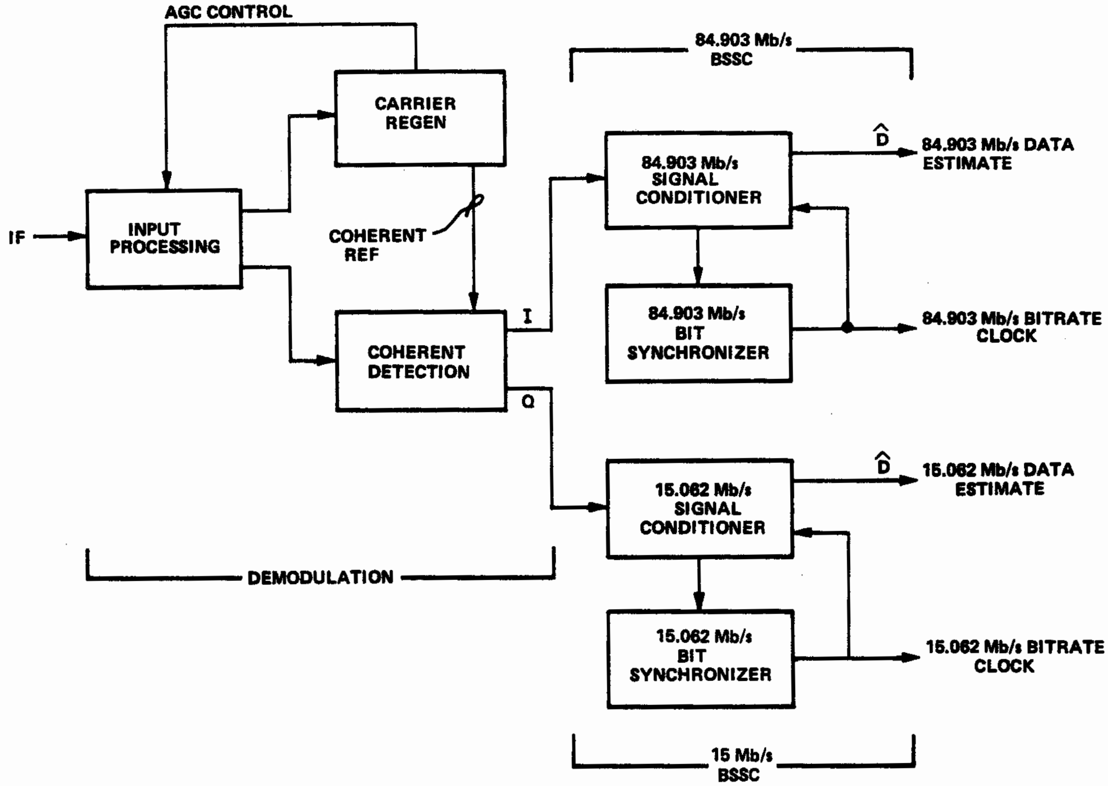


Figure 10. UQPSK Demodulator/BSSC's

Input Processor

The input processor is shown in Figure 11. The IF signal-plus-noise is wideband-bandpass-filtered to eliminate spurious signals and then variable-gain-amplified. The envelope detection for automatic gain control is accomplished in the carrier regeneration circuitry where filtering is narrower, and thus signal-to-noise higher. Non-coherent ACG makes output level independent of phase noise. The variable gain amplifier output is power-divided and sent to the carrier regeneration circuitry and to the coherent demodulator.

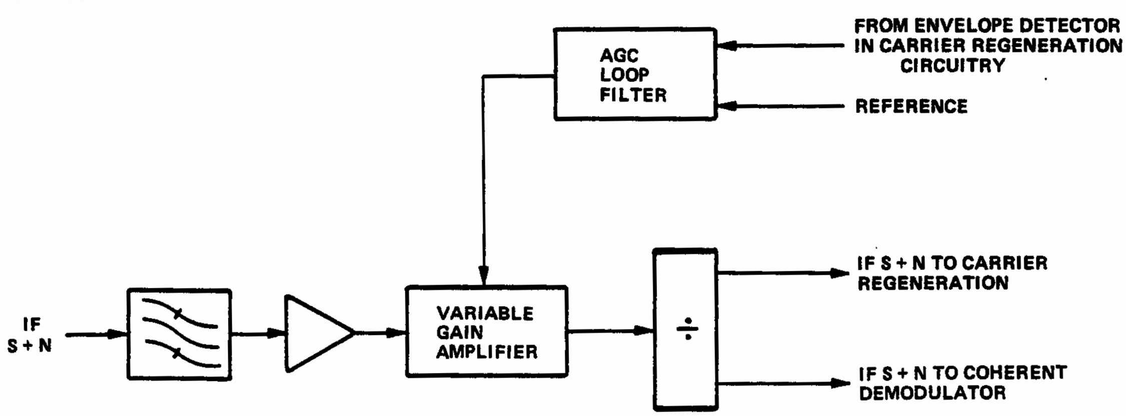


Figure 11. Input Processor

Carrier Regeneration

In Figure 12, the input to the carrier regeneration circuitry is narrowband-filtered in F1 and then routed to the AGC envelope detector and to an X2 non-linearity. The output of the X2 circuit contains an unmodulated carrier at twice the desired reference frequency for either biphase or unbalanced QPSK (UQPSK). The F2 filter improves the signal-to-noise ratio which is then processed by a limiter. The limiter performs two important functions. The self noise 2fc term of UQPSK is directly proportioned to signal power as is the desired 2fc term. Without a limiter, the variance of loop phase noise will increase as SNR increases while a limiter produces a constant SNR into the phaselock loop due to self noise. The limiter also keeps loop dynamics constant. The quadrature phase detector output drives the loop filter which controls the VCO. One VCO output is the coherent reference for demodulation while the other output is doubled and filtered by another F2 like the one before the loop. With two identical filters, phase shift due to doppler is cancelled. The output of F2 drives the mixers with quadrature signals after passing through the quadrature hybrid. The in-phase detector output is monitored by the lock detector which disables the sweep input into the loop filter when lock is detected.

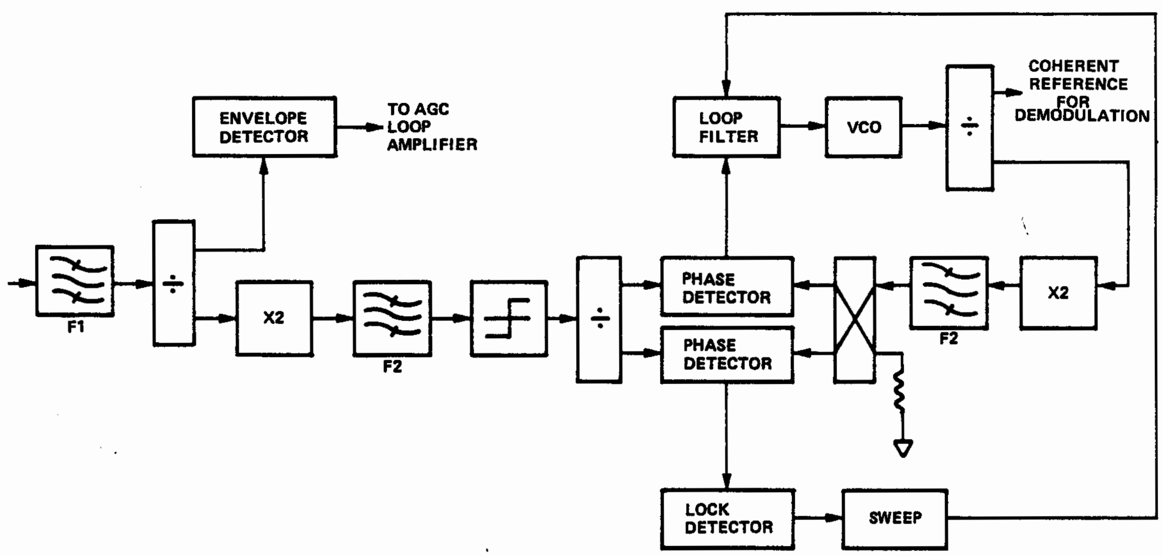


Figure 12. Carrier Regeneration

Coherent Detector

The coherent detector is shown in Figure 13. The IF S + N from the input processor is power-divided and routed through attenuators to the phase detectors. The L ports of these detectors are driven by quadrature components of the coherent reference. The coherent reference input passes through a variable phase shifter to optimize timing and then split into quadrature components by the quadrature hybrid. The hybrid outputs drive the phase detector L ports through attenuators. Both the I and Q baseband outputs are filtered by constant-resistance lowpass filters to remove sum frequencies and carrier reference feedthrough. The I output connects to the 84.903-Mb/s BSSC while the Q output connects to the 15.062-Mb/s BSSC.

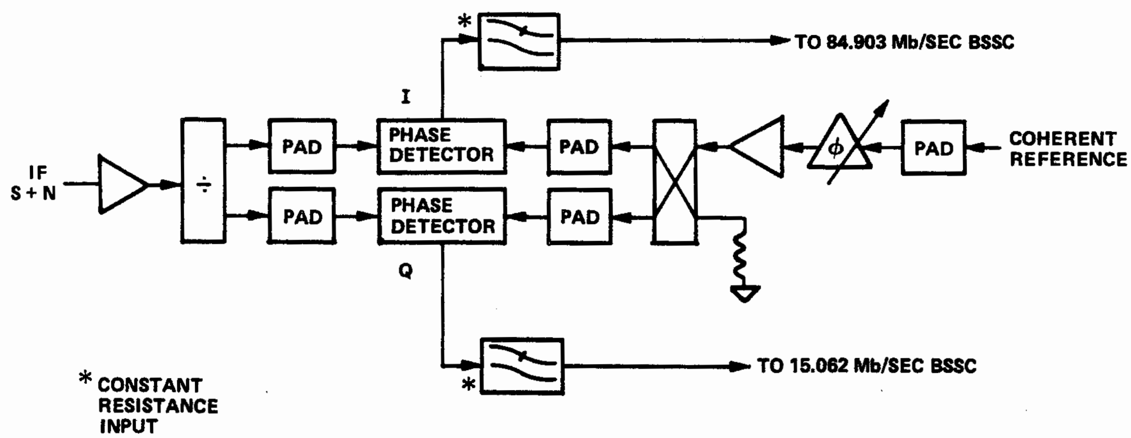


Figure 13. Coherent Detector

A block diagram for either the 84.903-Mb/s or 15.062 Mb/s-BSSC is presented in Figure 14. The baseband S + N input is optimally filtered by the time-invariant-matched filter which is a sliding integral approximation. The output y(t) of the matched filter is the input to the decision unit and also to an even-order nonlinearity. In the decision unit, y(t) is compared to a reference level E at the end of each period, and a state estimate, $\hat{D}$, for each bit is made. The timing for the decision unit is provided by the bit synchronizer portion of the BSSC.

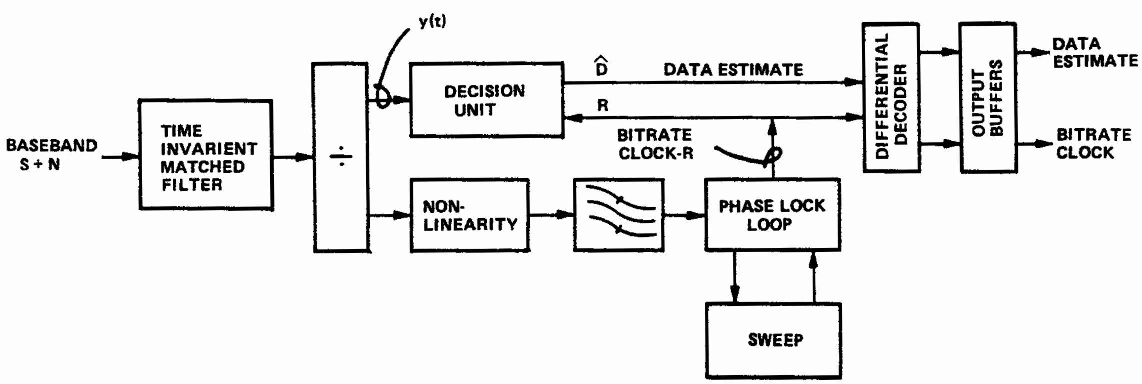


Figure 14. Bit Synchronizer-Signal Conditioner (BSSC)

The combination of the even-order nonlinearity and input filtering produces a bit rate spectral component at the nonlinearity output. This component is bandpass filtered to provide memory since bit transitions are intermittent and to improve SNR. The bandpass filter output is the reference for a type II phaselock loop similar to that already described in the carrier regeneration section. The bit rate VCO output, R, is routed to the decision unit as previously described and is also routed along with the data state-estimate $\hat{D}$ to the differential decoder which may be switched in or out. The decoder output clock and data are processed by output buffers to provide the overall output data estimate and bit rate clock. Functionally, the 84.903-Mb/s and 15.062-Mb/s BSSCs are the same but will vary slightly in their implementation. The 15.062-Mb/s input is essentially non-bandlimited while the 84.903-Mb/s input is somewhat bandlimited. The use of time-invariant-matched filters allows the matched filter response to be easily altered to optimize performance with bandlimiting.

In Figure 15, a photograph of the resulting hardware is shown. The UQPSK demodulator is packaged in one chassis, and each bit synchronizer-signal conditioner is packaged in its own chassis. That is, the 84.903-Mb/s BSSC is in one chassis, and the 15.062-Mb/s BSSC is in a separate chassis. The units shown are first generation production units. In the second generation production units, the bit synchronizer-signal conditioners have shrunk to a 3-1/2-inch chassis as opposed to the original 5-1/4-inch chassis.

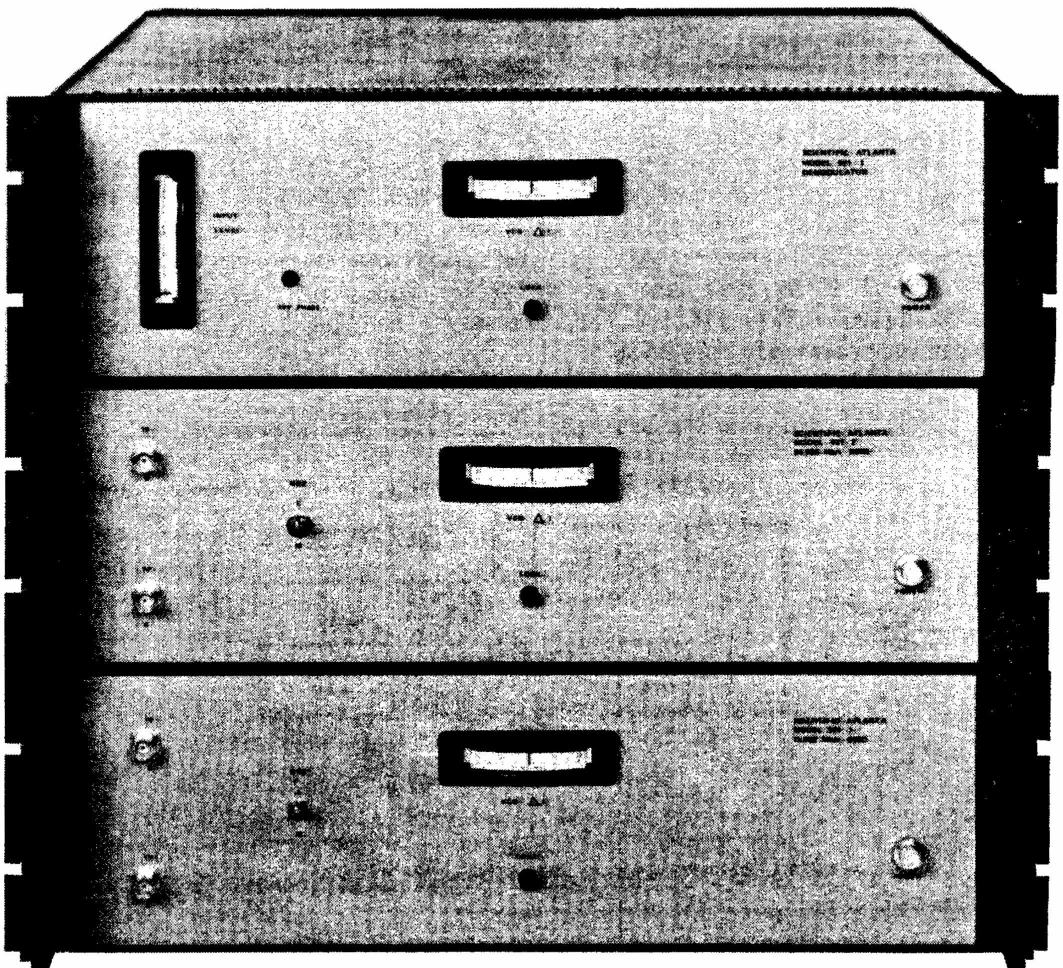


Figure 15.

Performance Results

The measured performance of the resulting UQPSK demodulator BSSC hardware is presented in Figure 16. In this particular figure, there are $(2^{23}-1)$ PN sequence data on each channel. Performance on both the I and the Q channel is within 1.2 dB from theory over a wide range of bit error rates. Specification limits of this hardware were 2.5 dB from theory.

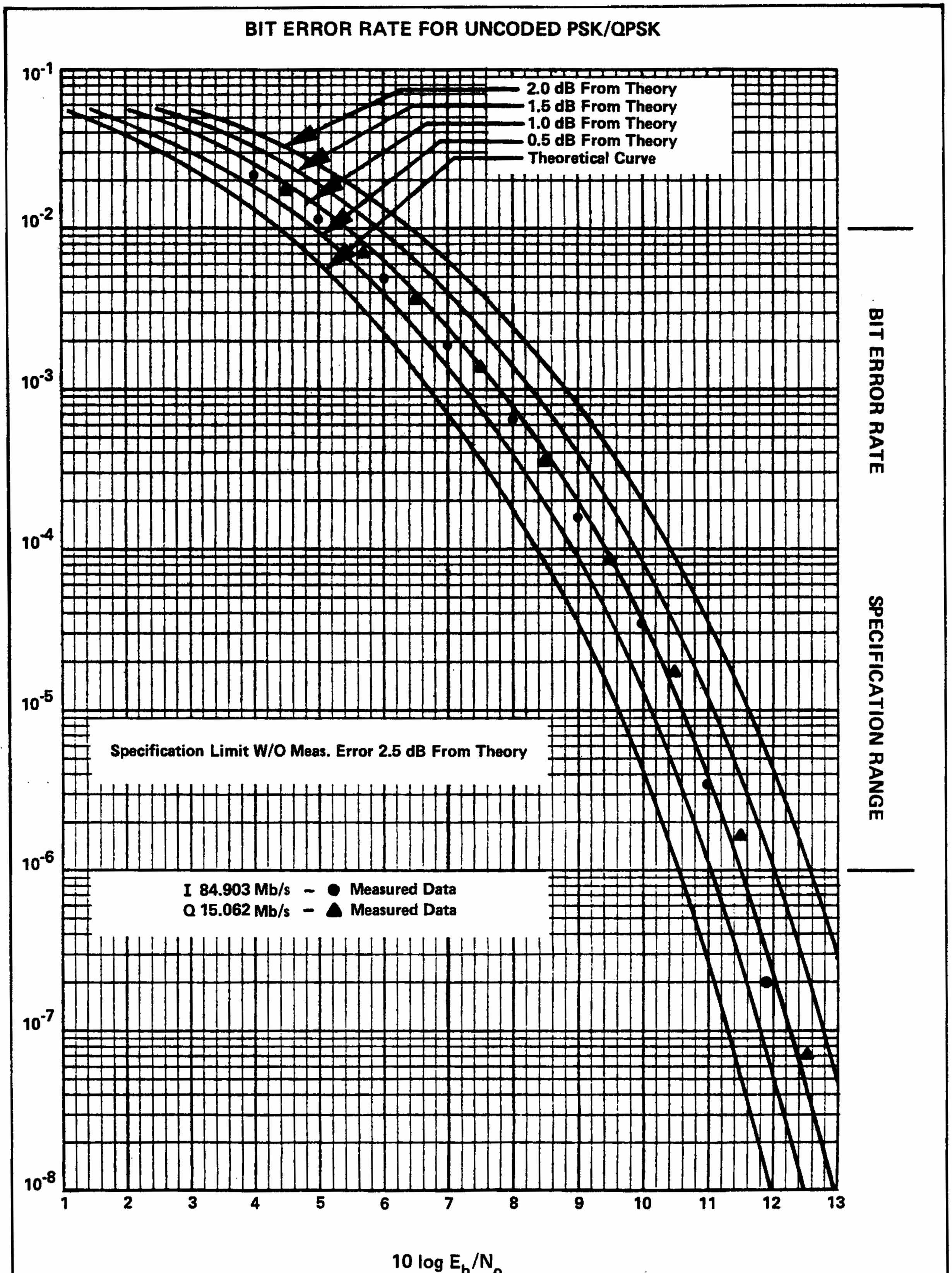

BIT ERROR RATE FOR UNCODED PSK/QPSK
2.0 dB From Theory
1.5 dB From Theory
1.0 dB From Theory
0.5 dB From Theory
Theoretical Curve
Specification Limit W/O Meas. Error 2.5 dB From Theory
I 84.903 Mb/s – ● Measured Data
Q 15.062 Mb/s – ▲ Measured Data
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
10^{-6}
10^{-7}
10^{-8}
1 2 3 4 5 6 7 8 9 10 11 12 13
BIT ERROR RATE
SPECIFICATION RANGE
10 log E_b/N_o

Figure 16.

Measured performance is essentially identical for other data combinations such as zero substitutions on the Q channel and unusual data patterns on the Q channel. The data presented is that measured on a production unit delivered to a customer, and all production units have measured essentially the same. Such performance is excellent and represents minimum implementation loss.

It has been previously discussed that the squaring technique of regenerating a coherent reference for demodulation is relatively impervious to the effects of modulator phase imbalance. In Figure 17, the measured performance of the UQPSK demodulator-BSSCs with a 3-degree modulator imbalance is shown. It is seen that there is essentially no degradation, and the curves look the same as when the modulator is well balanced. Data has also been run on the effects of amplitude imbalance on the modulator with similar results achieved. The most serious offender is the effect of phase imbalance. As mentioned previously the author's past experience with hardware and simulations on the effect of modulator imbalance for UQPSK have shown that many schemes suffer several dB degradation under this condition of imbalance.

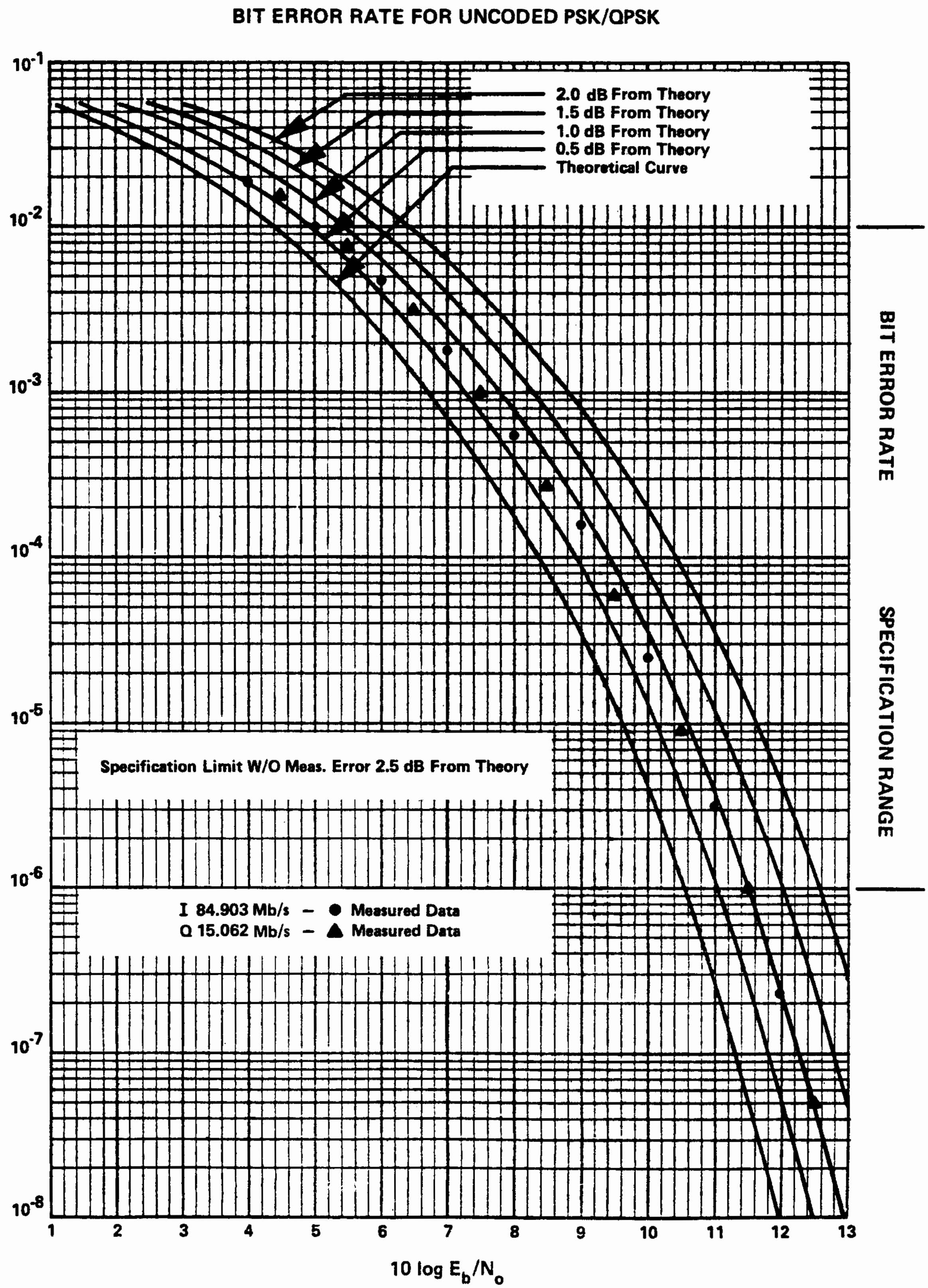


Figure 17. Measured Performance UQPSK Demodulator/BSSCs with 3 Degrees of Modulator Phase Imbalance

Summary

In summary, this hardware has been delivered to multiple companies and countries for earth stations to receive the LANDSAT data from the new LANDSAT-D earth resources satellite. The LANDSAT-D satellite provides more spectral bands and greater resolution than LANDSAT A through C, thereby enhancing the already extremely useful information received from the current LANDSAT satellites. The hardware discussed here performs very close to theory, which maximizes the quality of the LANDSAT data received.

Digital Frame Synchronizer, Phaser IV, V, and VI

Scientific-Atlanta, through its subsidiary, Digital Video Systems, offers a complete line of frame synchronizers for use in cable TV, LPTV, and broadcast applications. The Phaser IV, V, and VI series of synchronizers provide state-of-the-art technology in video synchronization.

The Phaser synchronizes input signals from an NTSC or black and white video source, including satellite, studio, network, remote camera, off-air, etc.

The Phaser lets the operator ignore time and distance, synchronizing signals automatically—such as Electronic News Gathering (ENG) programming transmitted via microwave—as if they originated in-house.

The Phaser eliminates complex signal routing and timing techniques. Signals from inaccessible cameras, remote vans, or moving vehicles are timed and synchronized into a fixed based control where they can be mixed, dissolved or used in chroma key or special effects with other sources.

With the Phaser, the operator can easily mix any live video inputs with studio programming, microwave, satellite, and VTR signals to produce a completely synchronized output without gen-locking.

The Phaser replaces the complex system of delay lines and pulse delay compensators which may now be used to time several studios through master control.

And for infinite switcher re-entry, which previously required very costly equipment, the Phaser is a breakthrough.

The Phaser IV with a 1.5 million bit RAM fieldstore memory is the basic "building block". Phaser V and Phaser VI employ the same modular design concept.

The Phaser V synchronizer builds onto the exceptional features of the Phaser IV. It has a framestore memory to provide double the memory capacity with half the number of components.

The Phaser VI takes the concept of synchronizing a step further, incorporating a new picture adaptive digital comb filter to provide the highest quality NTSC chroma inversion ever. It ensures virtually perfect decoding of the NTSC color TV signal as required in freeze frame applications or for studio-to-studio synchronization and switcher re-entry.

Rugged Packaging
The Phaser synchronizers are compact and rugged. Measuring only 3-1/2" high, they fit any surroundings, from studios and master control rooms to compact mobiles, and in standard headend equipment racks.

Simplicity Designed-In
Designed for ease of use, Phaser controls are conveniently placed on the front louvres in a logical arrangement. Functional controls located on the circuit boards are easily accessed and work at the flick of a switch. Adjustments are rarely necessary as one of the main benefits of the Phaser is the reduced need for service. Signals are phased automatically.

Easy Maintenance
LED's on the Phaser front louvres indicate operational status and simplify fault identification. If service is required, boards are quickly replaced to reduce downtime to a minimum.

Reliability
Digital Video Systems' units are checked for quality control and tested to ensure reliable service and long life.

The Smart Synchronizers
The Phaser IV has a 1.5 million bit 16K RAM fieldstore memory while the Phaser V and the Phaser VI have a 4 million bit memory provided by 64K RAMs.

Automatic Diagnostics
Memory fault diagnostics and fault concealment is automatic and continuous. Memory chip failure is detected and hidden and an LED readout code on the board identifies which chip is faulty.

Total Control
The Phaser's unique integral digital processing amplifier ensures precision signal level control. It replaces sync, burst, and blanking and adjusts video gain, black level, chroma gain and hue. All of the functions of a processing amplifier are obtained, plus a digital frame synchronizer, in a single compact unit. The Phaser eliminates picture shifts in switching and editing by converting all incoming NTSC signals to RS-170A with a sync to burst relationship that doesn't vary.

Automatic Hot-Switching
The Phaser automatically hot-switches between unrelated non-synchronous signals, holding the last complete field until the next valid vertical interval appears at the input. An internal mode selection switch provides the alternatives of: passing the input video regardless of input signal status; hot-switch mode with freeze and drop to black after a few frames if the input is not restored; and hot-switch mode with freeze until the input is restored.

Instant Performance Evaluation
The internal digital calibration signal generator provides for precision alignment of the system analog output. This test signal can also be routed from analog output to analog input to verify system status by side-by-side comparison every 16 lines of system throughput against the digital test signal output.

Any misalignment of the analog input is then easily and accurately corrected. This feature gives instant system performance evaluation and precision alignment.

No Noise or Jitter
The 5 TV line hysteresis function overrules the motion discontinuities which may occur when frames are deleted or inserted, even when the incoming signal is very noisy or has jitter.

Videotext/Teletext
With full bandwidth and precise digital blanking the Phaser synchronizers pass VIRS, closed caption subtitling, and teletext without degradation.

Chroma Inversion
For the first time, chroma inversion is a totally acceptable mode of operation, even for the highest quality prime time programming. The Phaser VI picture adaptive digital comb filter sets new standards for an NTSC decoder, with virtually undetectable chroma inversion.

Wide Application
The Phaser IV is primarily designed for use in broadcast and industrial TV, cable systems and production houses which require basic, low-cost synchronizing capability.

The Phaser V will meet the sophisticated requirements of those needing a full frame synchronizer while the Phaser VI meets the most stringent specifications, including an output picture that never shifts.

Cable TV
Using the Phaser in cable TV has measurable benefits. Synchronizing incoming demodulated feeds prior to cable distribution substantially reduces co-channel interference and improves the signal-to-noise ratio. With a synchronous system, the operator can insert commercials, key in messages, provide the viewer with stable channel changes, and reduce the cost of set-top data distribution equipment.

The Phaser synchronizes signals from headend demodulators, satellite receivers, and microwave links to your local programming facility, providing highest quality processed video with new digitally-generated synch, burst, and blanking for your cable distribution system.

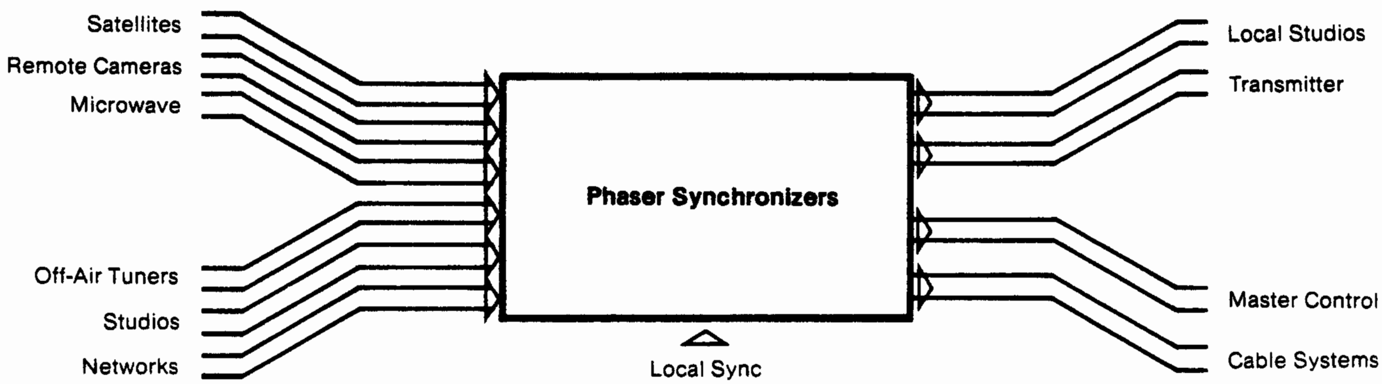
Infinite Switcher Re-Entry
When used for switcher re-entry, the Phaser VI gives your existing switcher a flexibility which was previously available only with the most expensive equipment. The Phaser makes it possible to feed the output of any mix effects buss into the input of any other.

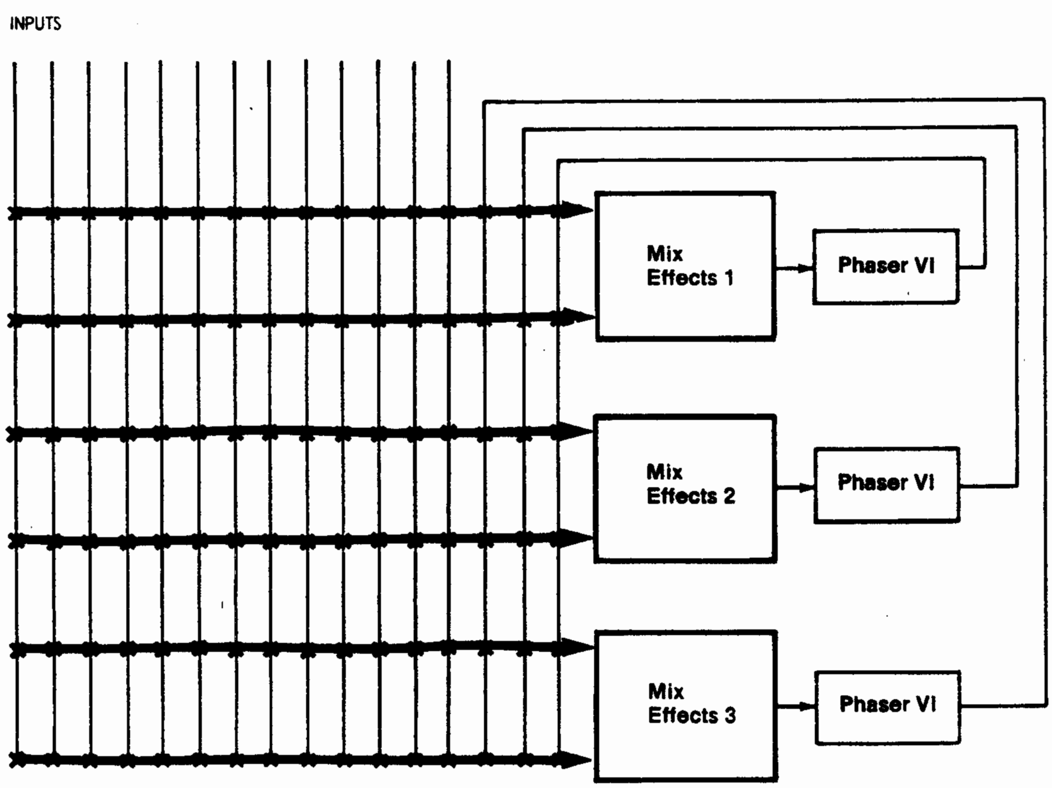
This infinite signal re-entry multiplies the capabilities of your switcher while providing a quality never before obtainable. (All analog timing delay lines can be eliminated if you choose.) This can be especially beneficial in local origination studios.

Features

- Digital Quality and Reliability
- Microprocessor Control
- Integral Digital Processing Amplifier
- Automatic Memory Fault Diagnostics and Concealment
- Internal, Digital Calibration Generator
- Modular Design Provides for Easy Maintenance
- Instant Performance Evaluation

The Phasers — Multi-Functional Synchronizers





Specifications

Mainframe

Digital Sampling
  Phaser IV
    8 bits at 14.3 MHz (256 levels at 4 times NTSC subcarrier)
  Phaser V
    8 bits at 14.3 MHz sampling on the R-Y, B-Y vectors (256 levels at 4 times NTSC subcarrier)

Bandwidth
  ±0.5 dB to 4.2 MHz, down 3 dB at 5.5 MHz

Signal-to-Noise Ratio
  Better than 56 dB (rms noise referenced to peak-to-peak quantizing range from dc to 4.5 MHz bandwidth)

Differential Phase
  $< 2°$

Differential Gain
  $< 1\%$

Processing Amplifier
  Digital sync, burst, and blanking are added to the output video RS-170A specification

Front Panel Operating Controls
  Video level, set-up, hue, chroma gain, systems horizontal phase, subcarrier phase, and bypass

Input Signals
  Video
    1.0V = 3 dB composite video at 75 ohms
  Reference Video
    High impedance loop-thru input
    Any color video signal with 40 IRE units of synchronization and burst = 3 dB, where burst has a non-varying phase relationship to sync.

Output Signals
  Outputs 1, 2, and 3
    1.0V composite analog at 75 ohms

Bypass
  In the event of power failure, VIDEO IN is directly connected to VIDEO OUTPUT 1. Also switch activated.

Calibration Generator
  1.0V composite video switched into all three video outputs digitally generated

Test Signals
  Full field color bars and multiramp

  In "wrap-around" mode, output and throughput are compared on a 16-line alternate basis

Color/Black and White Operation
  Automatic burst sensing and switchover to black and white mode; burst is turned off on the output

Oscilloscope Trigger
  Input source video and output video trigger at horizontal or vertical rates

Phaser IV

Correction Range
  262.5 TV lines

Hysteresis
  5 TV lines

Dimension
  89H x 406W x 572D mm (3.5H x 16W x 22.5D in.)

Weight
  22.7 kg (50 lb)

Power Requirements
  112V ac (nominal), 60 Hz, 250W max

Operating Temperature Range
  0°C to +40°C (+32°F to +104°F)

Phase V

Correction Range
  512 TV lines

Hysteresis
  5 TV lines

Power Requirements
  115V ac (nominal), 60 Hz, 400W max

Phaser VI

Same as Phaser V, plus 3-line adaptive comb filter

Mini-Cable Systems

Steve Turner

Introduction

Of the approximately 12,000 operating cable systems in the United States today, it is estimated that over one-half are mini-cable systems. Mini-cable systems have experienced tremendous growth during the past three years fueled by the growing desire of consumers to avail themselves of the wide variety of entertainment programming delivered via satellites.

Mini-cable systems are contrasted from major franchised cable systems in that they generally have very little coaxial cable plant and do not cross public streets. The FCC definition of cable television systems specifically states that the term shall not include any such facility that serves fewer than 50 subscribers or any such facility that serves or will serve only subscribers in one or more multiple-unit dwellings under common ownership, control or management. The implication of the FCC definition is that mini-cable systems do not necessarily fall under the rules and regulations governing cable systems.

Although mini-cable systems can be installed in almost any kind of business enterprise, the main applications that have developed are in multiple-unit dwellings and lodging establishments.

Multiple-Unit Dwellings

Multiple-unit dwellings generally include apartments, condominiums and trailer parks. There are currently over 100,000 establishments in this category, of which approximately 3,000 have installed private cable systems. Mini-cable systems are generally installed in multiple-unit dwellings in order to provide premium entertainment programming to residents. This programming (movies, sports, news, ethnic programming, etc.) is viewed by many as a necessary amenity for maintaining or increasing their occupancy rate.

Owners or managers of multiple-unit dwellings often choose to install a mini-cable system instead of hooking up to a franchised cable system for one or more of the following reasons:

a. A desire to provide satellite programming to their residents prior to the time cable is scheduled to pass their location;

b. An interest in the additional income associated with owning and operating a private system; or

c. Development of poor relations with the local cable company.

Multiple-unit dwelling owners must examine each situation to determine if a mini-cable installation is the appropriate vehicle for satisfying their business needs.

Lodging

Of the 40,000 hotels/motels in the United States today, approximately 3,000 have installed a private cable system. The first cable systems were installed in lodging properties in 1979. Since then, studies have shown that premium television has guest appeal and, in fact, is an amenity that many travelers have come to expect.

In addition to providing satellite programming to their guests, lodging owners are also able to participate in teleconferencing networks. As business travel becomes more expensive, meeting planners are turning in growing numbers to the satellite teleconference. Property owners with a satellite receiving system are able to offer their facility as a site for participation in a satellite teleconference, thus bringing in meeting room, overnight guest and food service dollars that might otherwise pass them by.

Other Applications

In addition to multiple-unit dwellings and lodging, other applications for mini-cable systems are universities, prisons, hospitals, and private networks. Although applications may vary, system components are essentially the same. Products developed specifically for mini-cable systems will be described in the following section.

System Description

The system diagram (Figure 1) of a mini-cable system is very similar to that of a major CATV system in everything but size (Figure 2). Both systems strive to provide a viewer at the extremities of the distribution system with the highest quality signal for the maximum possible time. However, in considering the individual components of the mini-cable system and their contribution to the overall network, some important differences become apparent.

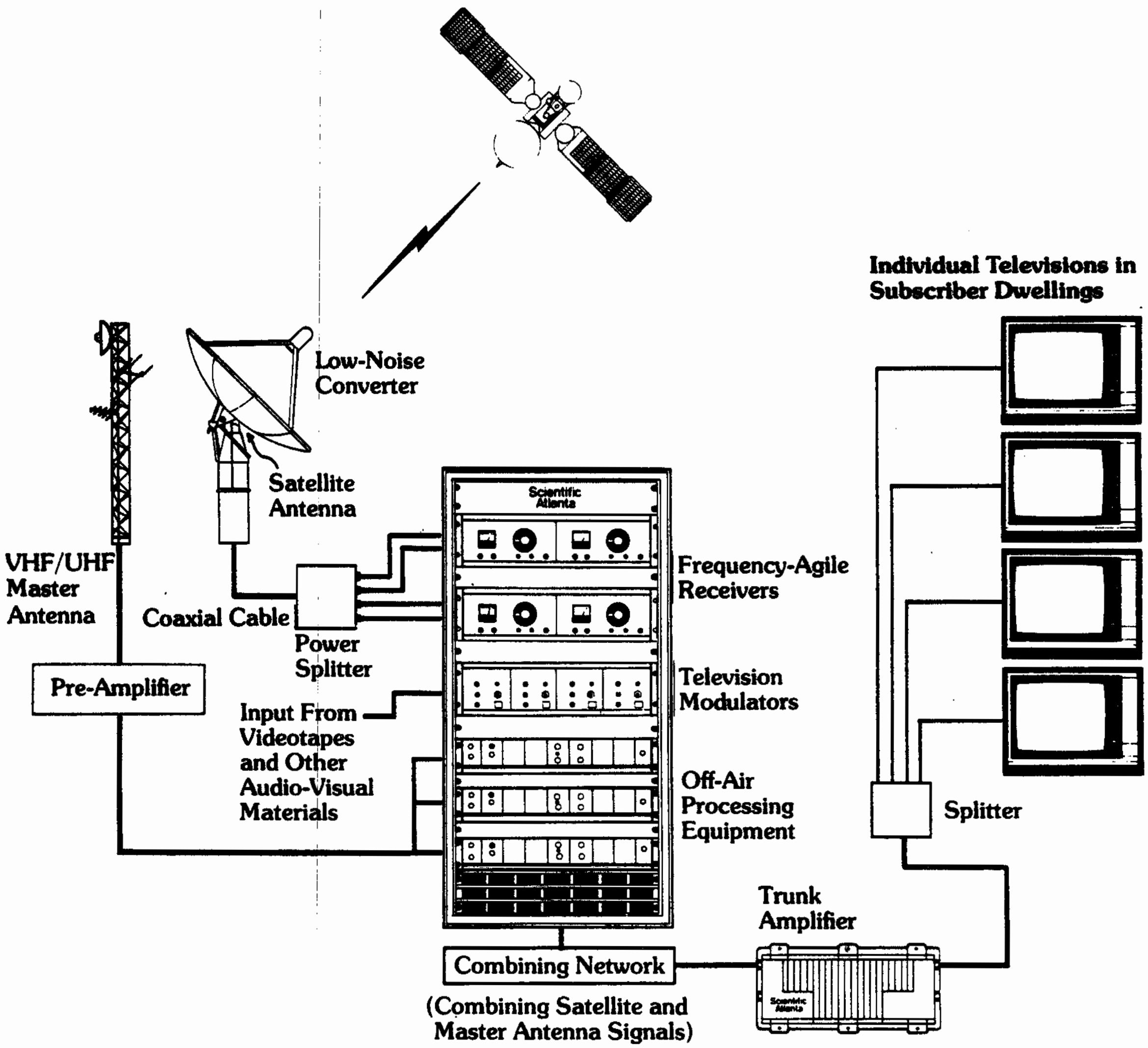


Figure 1. Mini-Cable System

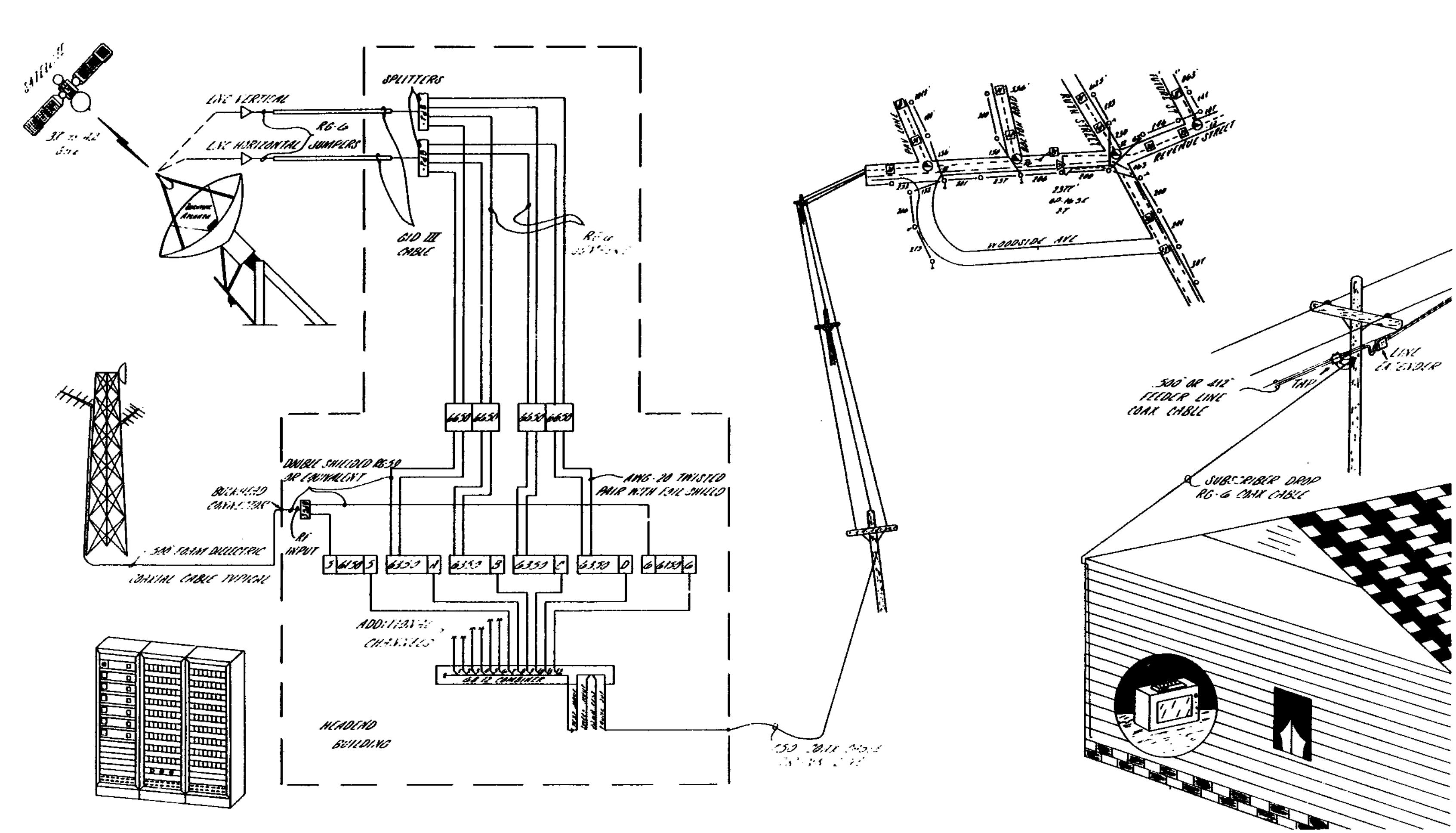


Figure 2. CATV System

In a CATV system the subscriber may be receiving one of up to approximately 108 available channels delivered to him through many miles of coaxial cable and several dozen amplifiers. With the potential for system degradation and equipment faults, a CATV operator must introduce a very high quality signal into the system to be certain of satisfying his paying customer. In a mini-cable system, the complexity of the network supplying a dozen or so channels tailored to the specific needs of 100 households without the burden of franchise demands is much reduced. Cable lengths are measured in hundreds of yards, and amplifier cascades are rarely necessary. The potential for complaints should a problem arise is also reduced, and service can be speedily restored without extensive troubleshooting.

The selection of the equipment to feed these mini-cable systems can thus take advantage of the reduced requirement for system margin while still maintaining satisfactory subscriber performance.

The most visible result of this reduced requirement is a reduction in antenna size. Instead of seeking a minimum video signal-to-noise (S/N) ratio of 50 dB in order to provide the household with 46 dB or better, we can think in terms of a typical 48 dB which may deteriorate to 46 dB under unfavorable weather conditions and yet still please a critical end user. In terms of satellite carrier-to-noise ratio (C/N), the corresponding reduction is from 12.3 dB (50 dB S/N) minimum to 10.3 dB (48 dB S/N) typical. This relates to a reduction in antenna size from 4.6 meters (Figure 5) to 2.8 meters (Figure 3) or 3.2 meters (Figure 4); with the appropriate low-noise device, antennas of this size would give satisfactory service over most of the contiguous U.S. with current satellites.

Additional requirements of an antenna suited for a mini-cable application are that it should be easy to install. A choice of surface or in-ground mount with minimal concrete requirements would suit most situations and could be easily adapted for rooftop configurations.

The mount geometry is important too. If the unit is to be installed by personnel unfamiliar with cable television earth station installations, the mount should be independent of the foundation orientation to eliminate errors. For application where reception from a single known satellite is required, an elevation-over-azimuth mount geometry provides the lowest cost and is the easiest to orient toward the satellite.

Some applications require that the antenna be moved occasionally (or frequently) to gain access to other satellites. For such situations, a declination-corrected polar geometry is most suited, particularly if it can be easily and simply oriented and aligned and can be motorized at very low cost with a programmable controller while still maintaining commercial specifications. Added benefit might be gained in the future if the antenna had a surface good enough to support highly-efficient Ku-Band operation upon the addition of a suitable feed.

The Series 9000 antennas (Figures 3 and 4) meet all of these objectives and, furthermore, at extremely low cost. While a separate paper discusses these products in detail, it is important to consider the continued application of Scientific-Atlanta's full range of earth station antennas.

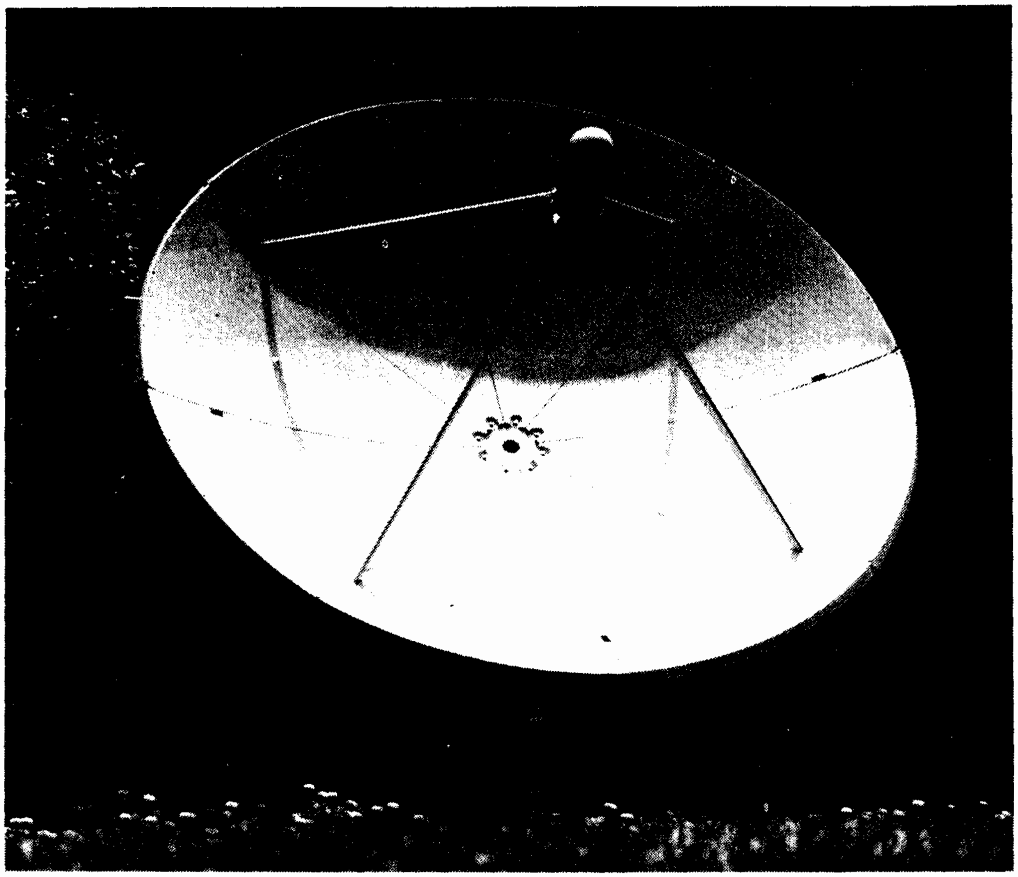


Figure 3. 2.8 Meter Series 9000 Antenna

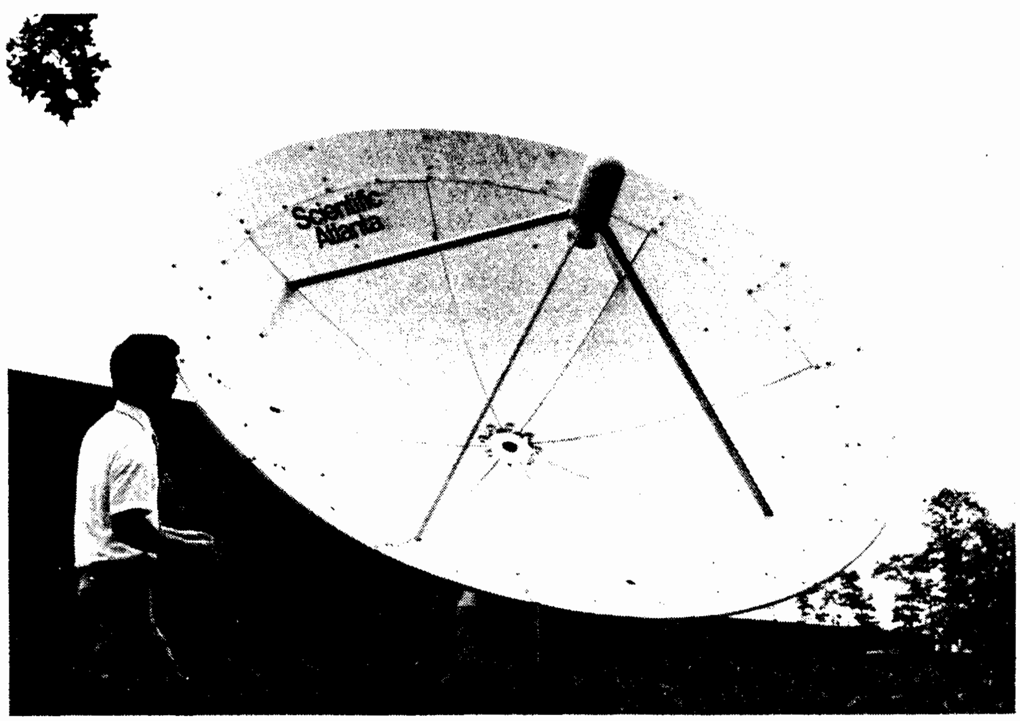


Figure 4. 3.2 Meter Series 9000 Antenna

Where an FCC license is required, a Model 8346 4.6-Meter Antenna (Figure 5) might be preferred and might also be the choice, along with the Model 8008 5-Meter Antenna for data applications, particularly for transmission of data. Both the Model 8346 and 8008 are available in economical prime-focus form.

Figure 5. Model 8346 4.6 Meter Antenna

When installing an earth station, the location must be very carefully considered. The antenna must naturally have a clear line-of-sight to the satellite of interest, but the look angles to all other satellites must also be considered. With the changing state of the communications industry, it is a wise precaution to keep all available options open. The ideal location for an antenna is on the ground where protection from terrestrial microwave interference is maximized. While interference from the telephone company microwave routes can be minimized with filters, the protection of trees and buildings may be essential in downtown locations. Rooftop installations are often more complex than those on the ground because of the significant uplift forces on the antenna due to even moderate wind forces. The majority of roof structures are designed to support downward forces of a few hundred pounds, not upward forces of several thousand.

Moving back from the antenna, we encounter the receive electronics, a combination of amplifier and receiver. Already field-proven in the CATV industry, block conversion is also the preferred choice of the mini-cable operator. By converting the received satellite signals in a block from 4 GHz to UHF at the antenna, significant savings may be realized. Low-cost coaxial cable now feeds a series of UHF video receivers. No microwave components need be used in the headend, and automatic assembly of the receivers, which yields lowest costs, can be maximized. All of this must be accomplished without sacrificing performance or flexibility as CATV and mini-cable systems alike demand the best from the receiver. For this reason the product selected by both operators is likely to be the same. The Model 6650 Video Receiver (Figure 6) and associated Series 360 Low-Noise Converter have been the choice of cable systems and mini-cable systems across the nation, with over 15,000 being installed since introduction in March, 1981. This receiver enables a system operator to receive the wide variety of services carried on subcarriers of the transponder via the composite baseband output or from an on-board subcarrier demodulator in addition to the audio associated with the video. The receiver has options which enable remote switching by a number of means; a full description of these options is found in a separate paper.

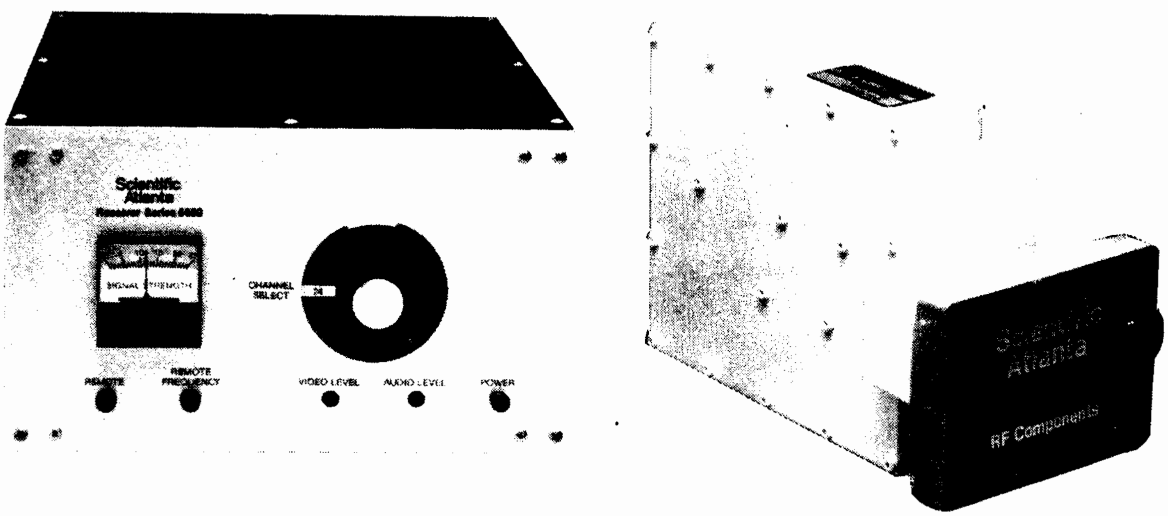


Figure 6. Model 6650 Video Receiver and Model 360 Low-Noise Converter

The LNC amplifies the microwave signals at 4 GHz with minimal noise contribution. A phase-locked oscillator and mixer generates a band of frequencies from 270 to 770 MHz corresponding exactly to the 3.7 to 4.2 GHz input frequencies. The signals are carried via low-cost 75 ohm cable to conventional UHF splitters to feed the tuneable receivers. A power inserter enables dc power to be transmitted from the receiver to the LNC for powering protected from atmospherically induced surges and eliminating the need for separate cabling. In a dual-polarized configuration, two cables and LNC's are necessary as the signals from the two polarizations must be kept separate.

One further advantage of the block conversion to low frequencies is the ability to separate the antenna from the receive electronics by a greater distance than is possible at 4 GHz. If the distance to be run is a few hundred feet, then CATV distribution or trunk cable may be used. For longer distances an equalizer may be sufficient to handle the loss before requiring an inexpensive post amplifier.

Baseband video and audio from each receiver is introduced into a television modulator. This piece of equipment generates first a 4.5 MHz subcarrier containing the program audio, and then an RF signal on a recognized TV channel in conventional format. Traditional CATV products have been extremely modular to be flexible enough to be of use in the many diverse system configurations in use today. Such flexiblity is seldom required in a mini-cable system, and instead of selecting the Model 6350 Modulator (Figure 7), a mini-cable system is more likely to use the modulator designed expressly for this application, the Model 6330 (Figure 8). This represents the latest in a long tradition of designing high-quality modulators. Surface Accoustic Wave (SAW) vestigial sideband filtering is standard, enabling an operator to fully use all the available channels without leaving guard bands to relieve spurious signal problems. Available in VHF (2-13), Midband (A-I) and Superband (J-W) channels, this modulator exceeds all performance expectations for smaller cable systems. Utilizing integrated circuitry for reliability and cost savings, the 6330 Modulator includes the features necessary to enable a system operator to easily confirm optimum system performance. The 6330 Modulator occupies only one quarter of a rack width and shares rack space efficiently with the 6650 Receiver, thus minimizing the space requirements for a multi-channel system. Output levels of more than 60 dBmV reduces the need for much additional amplification within the system. In systems using the mid-band range of channels, converters are often used in each household. While providing the operator with no security, they do allow the use of an additional nine channels at low cost. These low-cost converters invert the spectrum as they translate signals back to VHF and require a spectrum inversion option on the Model 6330, a factory-fitted inexpensive feature.

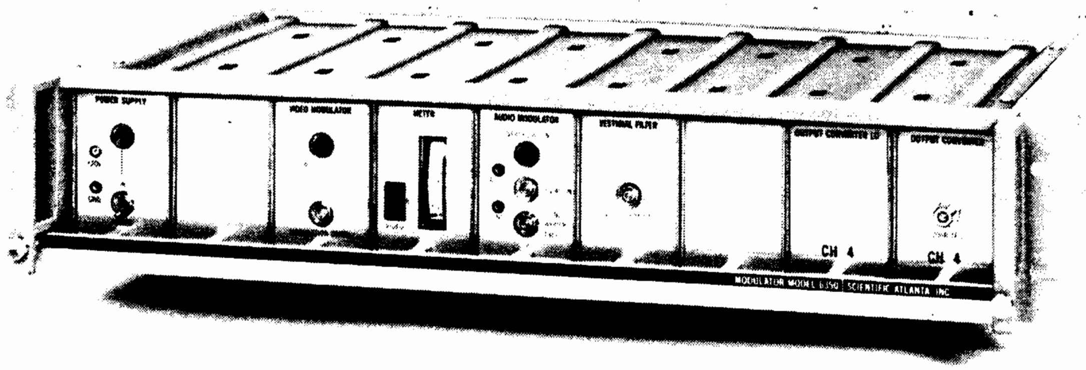


Figure 7. Model 6350 Video Modulator

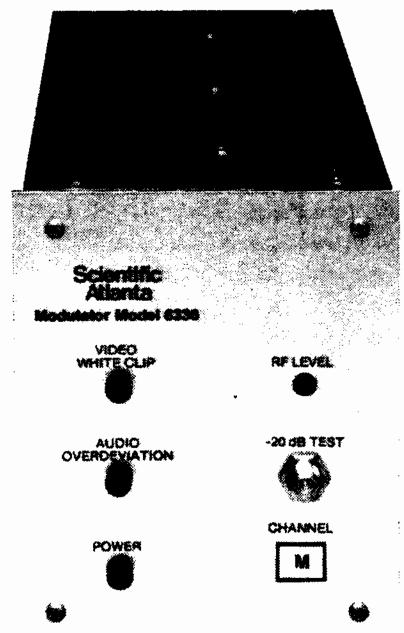


Figure 8. Model 6330 Video Modulator

The satellite portion of the mini-cable system is just a part of the whole. It must be combined with off-air signals from conventional UHF and VHF antennas. These signals are filtered and amplified, the UHF channels being translated to a lower frequency.

The Model 6130 Signal Processor (Figure 9), as a companion to the Model 6330 Modulator, is also designed to occupy one-quarter rack space. Important standard features include SAW filters, the ability to permit adjacent channel rejection, automatic gain control, adjustable sound carrier level and a high-level IF switch. Front-panel indicators and adjustments allow effective monitoring of unit operation and status of signal input. Extensive signal processing offers standard VHF, midband and superband output channels as well as VHF, midband, superband and UHF input channels.

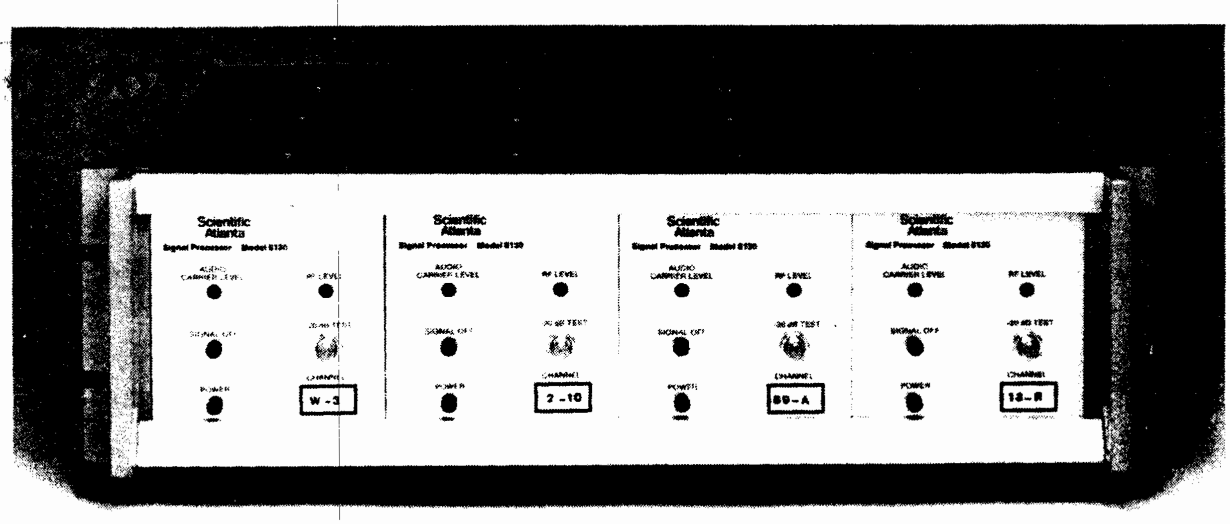


Figure 9. Model 6130 Signal Processor

Building a System

Off-air signals are combined with the satellite signals and possibly signals from a character generator or video tape player through a combining network or combining amplifier and, once on a single coaxial cable, distributed throughout the system. Further amplification may be provided by CATV line extenders or other low-cost broadband amplifiers. The "headend" electronics, as it is now known, is to be mounted in standard 19-inch racks with adequate spacing for optimum cooling and with ordered and labeled wiring. This will assist in an understanding of the system, as well as aiding in maintaining and troubleshooting the system.

If the rackup is designed and configured in the factory, the benefit of years of design experience may be utilized, and a record of the design is maintained to aid in system expansion at a later date. In addition, the system is tested and calibrated as a complete unit requiring only the addition of ac power.

The distribution system may take many forms. The older variety of daisy chain provides no security and may not always be capable of handling more than the VHF channels. A simple form of security can be provided by using "home-run" cabling. In this scheme, the household may be turned off simply by disconnection, although the cable costs are higher. For extensive systems or those with many channels or tiers, a conventional CATV network is most applicable.

A 36-channel converter is available from Scientific-Atlanta which provides security and flexiblity. For expanded needs, a 64-channel converter is also available in a programmable or addressable model. The option of addressability is open for pay-per-view and multi-tier systems. Thus, it is apparent that there are indeed similarities between the CATV and mini-cable network; the same quality of components, the same reliability backed by a vendor with extensive experience, technical expertise and a full field support program are required in both cases.

Economics of Mini-Cable Systems

The economics of mini-cable are influenced primarily by the same variables found in the cable industry. The difference between the two is largely a matter of scale.

The economic parameters for mini-cable systems are of three categories: system design/cost considerations, revenue projections and financial parameters. Each is discussed in detail below.

a. System Design/Cost Considerations. Mini-cable systems generally consist of a single satellite antenna, a downsized headend (4 to 12 channels) and a distribution system. For the purpose of this analysis, a distribution system is assumed to already be in place in the form of a master antenna system.

   The system configuration utilized to develop the economic models in this article consists of one Series 9000 Antenna, two Model 360-1 Low-Noise Converters and five each Model 6330 Modulators and Model 6650 Receivers.

b. Revenue Projections. Revenue projections for this model were based on the assumptions that appear listed in Table 1. This five-channel system consists of two pay channels and three other satellite-delivered channels.

Table 1. Mini-Cable System Revenue Assumptions

Subscriber Rates	Subscriber Fees
Basic - 50%	Basic - $9.00/month
Pay - 50% of Basic	Pay - $14.00/month

Expenses

1. Programming Costs

   Two Pay Channels at $4.50 per subscriber

   Three Basic Services:

   CNN $0.15 per subscriber

   ESPN $0.04 per subscriber

   WTBS $0.10 per subscriber

2. Franchise Fees/Owner Commissions

   5% of Programming Income

3. General and Administrative Expenses

   5% of Programming Income

c. Financial Parameters. The financial parameters used in the mini-cable economic model are as follows:

1. Tax rate = 46%

2. Investment tax credit (first year only) = 10%

3. Depreciation of plant and equipment was determined using the same scale allowed for cable operations.

Economic Model Output

The output of the mini-cable economic model provides for our analysis of net income, cash flow and return on investment.

a. Net Income. Monthly net income after the first year of operation for the mini-cable system outlined earlier ranges from $66 for a 150-unit service area to 1785 for an 800-unit service area. Figure 10 shows the relationship of monthly net income to size of area serviced in units. Breakeven after the first year of operation in terms of potential units serviced for the model system was 70 units.

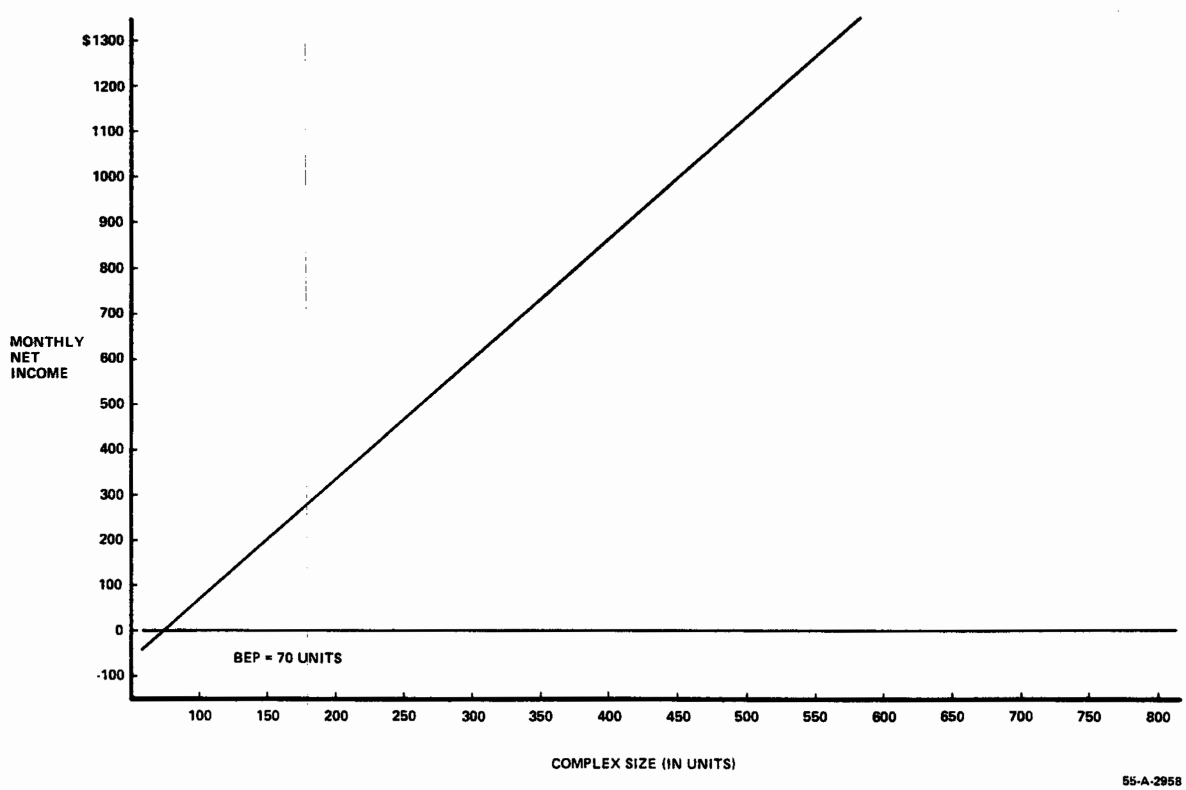


Figure 10. Mini-Cable Net Income/Break-Even Analysis

b. Cash Flow Analysis. Figure 11 illustrates the number of months to positive cash flow relative to the number of potential units being serviced. This graph indicates that all mini-cable systems of 475 potential subscribers or more reach a positive cash flow for the system owner in less than one year.

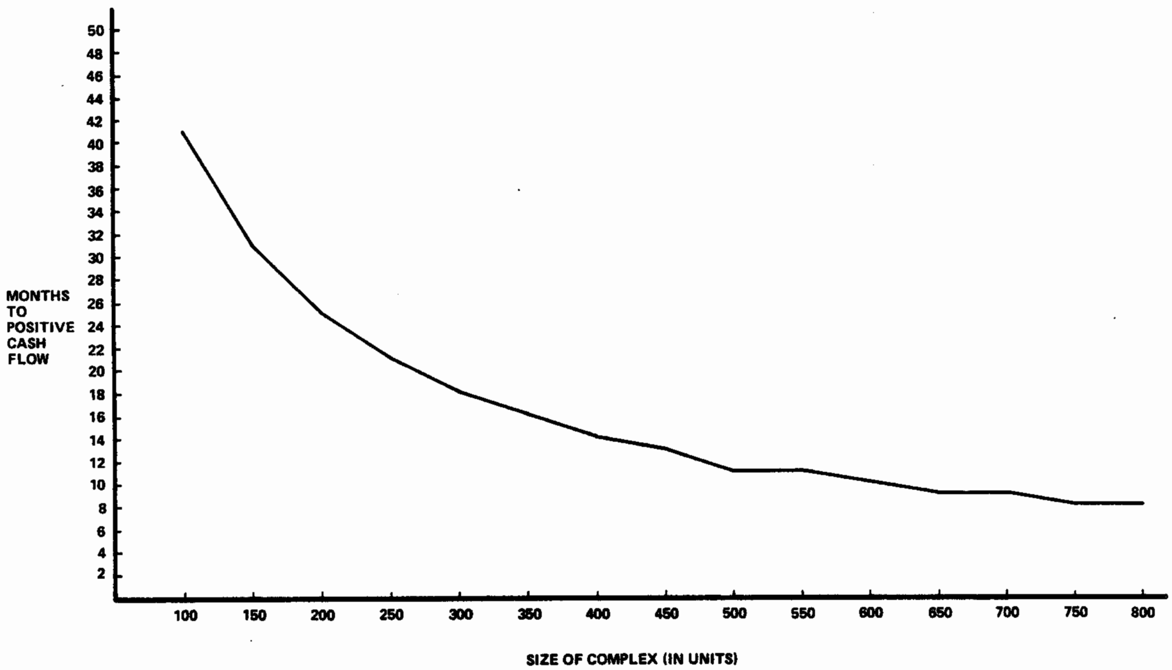


Figure 11. Mini-Cable Cash-Flow Model

c. Return on Investment. Net cash return on total investment for the first year of operation is graphically illustrated in Figure 12. Of particular significance here is the fact that even small service areas have a high ROI the first year, and larger systems (500+) are capable of regenerating the total investment in one year or less.

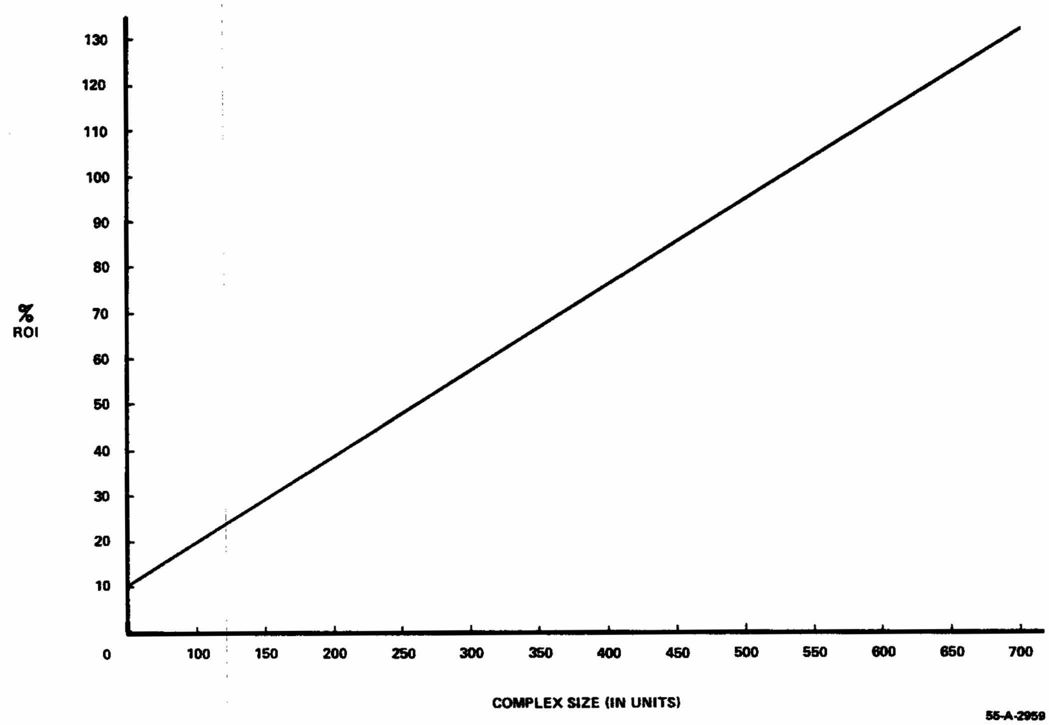


Figure 12. Mini-Cable Return-on Investment

Conclusions

The above model portrays a relatively conservative scenario of business opportunity in the mini-cable system market. The relatively low breakeven point, prompt recovery to positive cash flow and strong ROI performance indicates that mini-cable has a very good potential to generate substantial cash flow for its owners. This opportunity becomes even more attractive when one considers the potential cash flow that may be generated from several mini-cable systems.

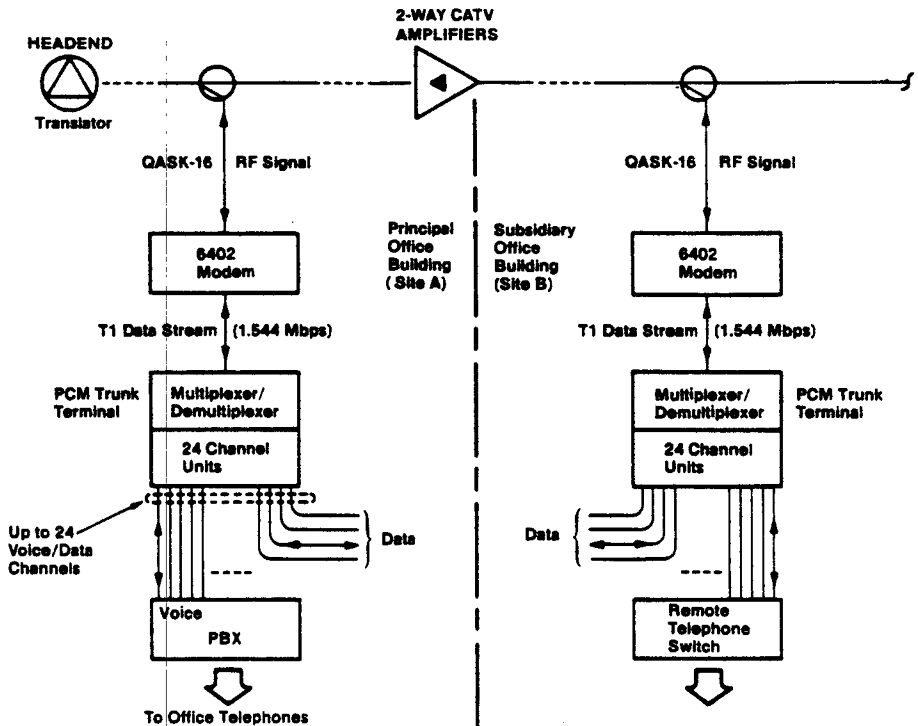


Figure 1. Voice/Data Interconnection Using T1 Trunks

With the use of the high-speed broadband data modem, voice and data communications via CATV coaxial cable has become a reality. Coaxial cable is a broadband rather than a narrowband medium, meaning that it has an inherent high information-carrying capacity or bandwidth, whereas data communications via conventional links are structured around the more limited voice-grade network. The broadband data modem adapts voice and data signals into a format for transmission over this high-speed, high-capacity medium.

Broadband capacity is currently available close to many business locations in the form of cable television. Many of these systems have unused bandwidth on their entertainment network, and some have second cable systems called institutional, or "B" cable, networks. These second networks provide two-way capacity for local interconnection and offer ideal voice and data communications alternatives.

To demonstrate how data communications is accomplished via a CATV system, refer to Figure 1. A modem located at business site "A" transmits data onto the cable where it travels to the headend along the cable return path. At the headend, a translator changes frequency of the signal and sends it back out onto the same cable along its forward path. The signal is received by an authorized modem at business site "B" and response from business "B" is transmitted back to the headend along the cable return path. The translator again redirects the signal along the forward path to business "A." The modem simultaneously transmits and receives information in a bandwidth efficient manner, utilizing very little of the available cable spectrum, or bandwidth. One translator is sufficient for multiple transmission channels.

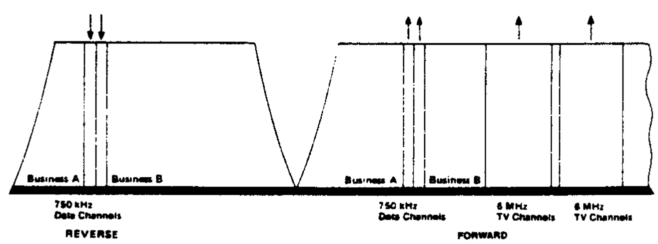
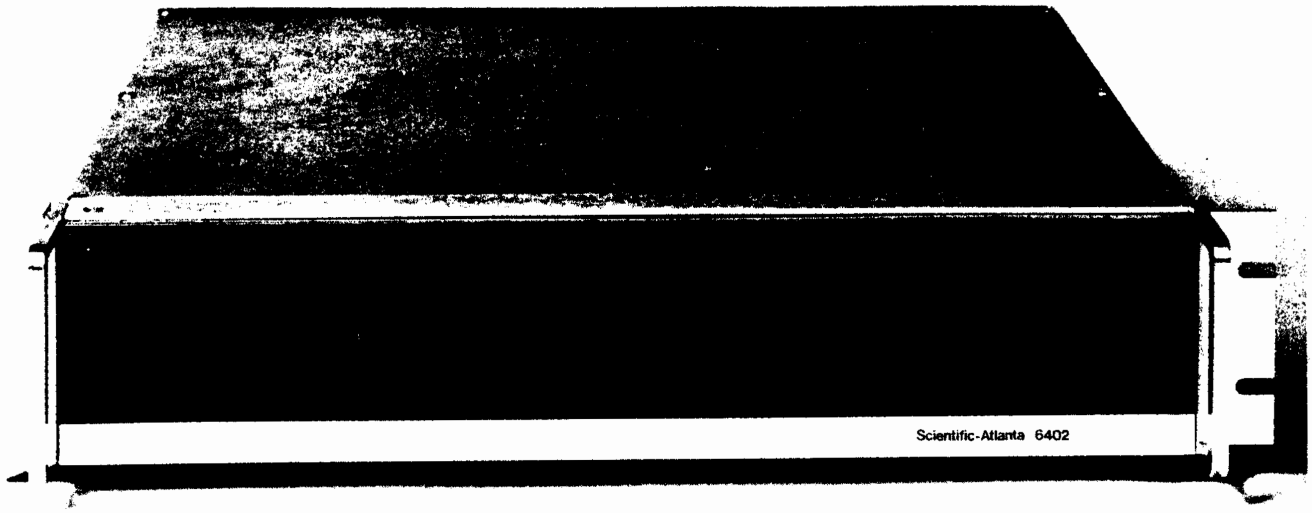


Figure 2. Sub-Split CATV System

While most modern cable systems have anywhere from 220 MHz to 450 MHz of bandwidth available on the cable, high-speed data communications may use only a little over one megahertz for transmission and reception. For example, in this diagram of bandwidth in a sub-split cable television system (figure 2), communications between Business "A" and Business "B" uses only 1.5 megahertz in the forward direction for a 1.544 Mbps data stream and 1.5 megahertz in the reverse direction, as compared to the 6 megahertz needed for transmission of an entertainment program. A mid-split cable system offers even greater full duplex communications capacity.

A combination of CATV distribution products and coaxial cable coupled with broadband data products provide the high-speed, cost-effective communications network for optimum use by the CATV operator and the commercial user.

Broadband Data Modem, Model 6402



The Model 6402 broadband data modem is a high-speed modem designed to facilitate point-to-point data communications via coaxial cable. It enables the cable operator to lease to businesses bandwidth on entertainment or institutional systems. Because of the modem's bandwidth efficiency, many businesses can be accommodated in a small amount of cable spectrum. The Model 6402 is frequency agile and can be manually adjusted at the business site on both the transmit and receive frequencies.

Features

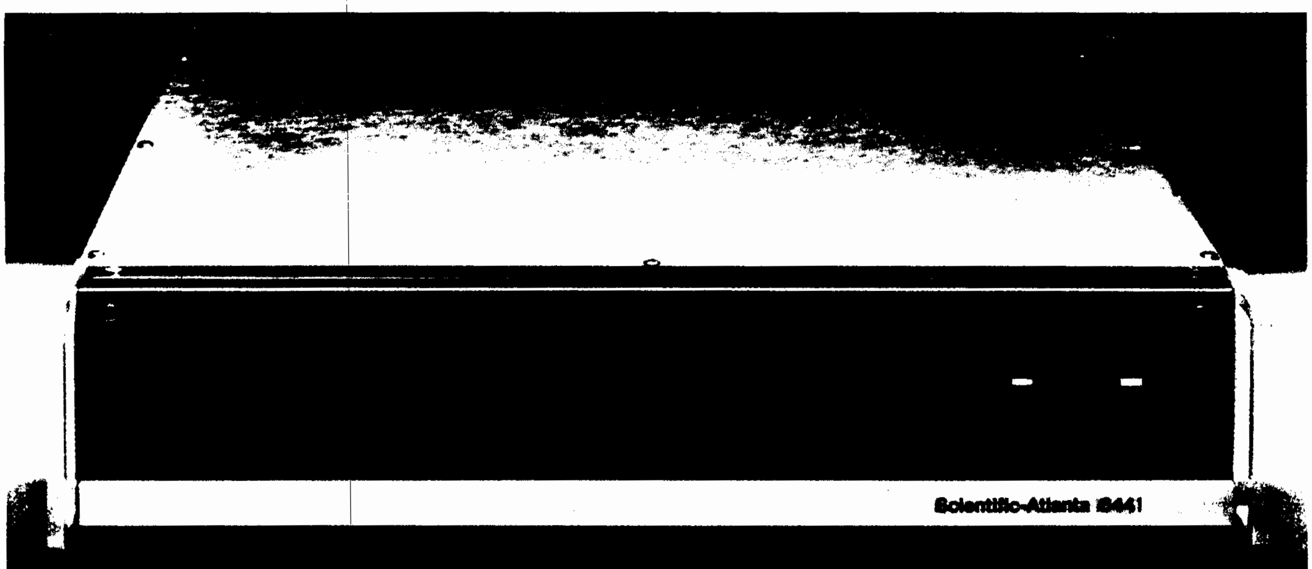
- High-Speed Standard T1 Data Rate
- Point-to-Point Data Communications
- Bandwidth Efficient
- Synchronous, Full Duplex Operating Mode

Specifications

General

Power
  115V or 230V ac ±10%
  100 watts, 50/60 Hz or −48V dc
Temperature
  +10°C to +50°C (+50°F to +122°F)
Size
  Standard 19-inch rack mount chassis
  Optional stand alone package
Modulation
  QASK-16
Transmitter
  Level
    +20 dBmV to +50 dBmV
  Frequency Range
    Standard 5 to 120 MHz
    Optional 162 to 440 MHz
Receiver
  Level
    −10 dBmV to +10 dBmV
  Frequency Range
    Standard 162 to 440 MHz
    Optional 5 to 120 MHz
Frequency Adjustment Resolution
  250 kHz (Permits transmission in both HRC and IRC formats)
Performance
  Operational BER $< 10^{-9}$ at C/N $\geq$ 33 dB (NCTA)
Spectral Efficiency
  750 kHz spacing
Operating Mode
  Synchronous, full-duplex
CATV System Compatibility
  Operates with conventional 2-way systems (Sub-split or mid-split)
Interface
  Standard Digital
    RS-442 (RS-449 DT)
  Optional Digital
    DS-1, V.35, RS-449 SR
RF
  BNC connector, 75 Ω impedance
Scrambling
  In accordance with CCITT V.35
Controls
  Power
  Baseband Loopback
  Local Loopback
Indicators
  Power
  Baseband Loopback ON
  Local Loopback ON
  Lock Detect
Data Rate
  Standard T1 (1.544 Mbps)
Accessories
  Frequency synthesizer programmer

Frequency Translator, Model 6441



The Model 6441 frequency translator is designed to facilitate two-way communications via standard CATV systems in a mid-split configuration. The translator enables the cable operator to utilize any unoccupied channel in the entertainment or institutional system in a highly reliable, bandwidth-efficient manner. The Model 6441 functions as a modular headend unit which may be configured for the operator's specific bandwidth and channel allocation.

Features

- Flexible Frequency Translation
- 1,2, or 3-Channel Capacity
- Frequency Agile
- HRC and IRC Compatible
- Optional Fault Protection

Specifications

General

Power
  Standard 100V to 130V ac, 60 Hz
Temperature Range (to maintain specifications)
  0°C to +50°C (+32°F to +122°F)
Dimensions
  88.9H x 482.6W x 406.4D mm (3.5H x 19W x 16D in.)

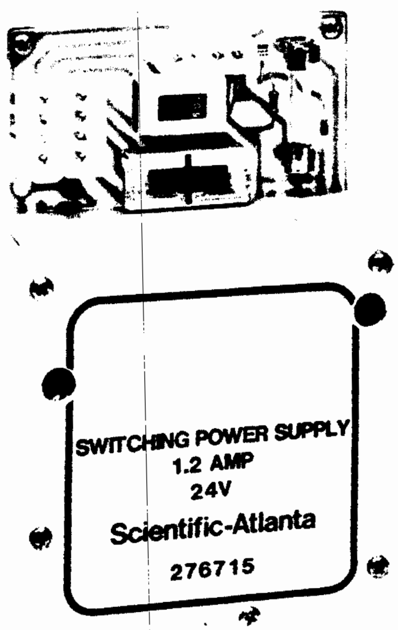
Operational

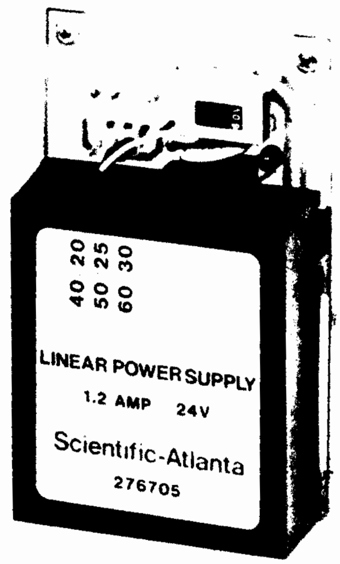
Input Frequency Range
  5 to 120 MHz
Output Frequency Range
  162 to 440 MHz
Frequency Adjustment Resolution
  250 kHz
Translation Channel Capacity
  1, 2, or 3 adjacent channels
Effective Bandwidth
  5.5 MHz, 11.5 MHz, or 17.5 MHz
Frequency Stability (over temperature range)
  ±865 Hz
Input Level
  −10 dBmV to +10 dBmV
Output Level
  +20 dBmV to +50 dBmV
Spurious Outputs
  <−60 dBc
Adjacent Channel Rejection
  60 dB
Input/Output Impedance
  75 Ω nominal

Connectors

RF
  75Ω BNC
Test
  25-pin D-type
Controls
  Power
  Level adjust
Indicators
  Power
  Input channel frequency
  Output channel frequency
Accessories
  Frequency synthesizer programmer
  1:1 protection switch

6501/6502 Power Supply Modules 276705 and 276715





Power supply modules 276705 and 276715 are used in the 6501/6502 distribution amplifier station to convert station ac input voltage to a well-regulated dc voltage for station powering. The switching regulated supply (276715) provides improved efficiency over the standard linear power supply (276705) which results in cost savings in operating expense. The switching regulated supply is a constant power device, meaning that it automatically adjusts its internal operating parameters for most efficient use of different levels of voltage and current within a system. The switching regulated supply automatically determines load power requirements and adjusts its operation so that only that amount of power is provided.

6501/6502 Power Supply Modules 276705 and 276715

Specifications

	Switching Regulated 276715	Linear 276705
Output Voltage	24V ±1V	24V ±1V
AC Input Voltage (1) (2)	60V - 37V min., 60V system typical	60V - 37V min., 60V system typical
	30V - 19V min., 30V system typical	30V - 19V min., 30V system typical
Output Current	1.2A max	1.2A max
Output Ripple	10 mV p-p max	10 mV p-p max
Switching Frequency (3)	5 kHz to 20 kHz	Not applicable
Tsus (4)	20 ms min.	20 ms min.
Efficiency at 60V/30V Input and full Load	64% min.	50% min.
Output Fuse	1.5A	1.5A
Output Overvoltage Protection	32V threshold for SCR Crowbar	32V threshold for SCR crowbar
Operational Temperature	−40°C to +60°C (−40°F to +140°F)	−40°C to +60°C (−40°F to +140°F)
Weight		
Net	0.34 kg (0.75 lb)	0.32 kg (0.7 lb)
Shipping	0.68 kg (1.5 lb)	0.64 kg (1.4 lb)

Ordering Information

Switching Regulated Power Supply Module Part No. 276715
Linear Power Supply Module Part No. 276705

NOTES:

1. Transformer selection for 30 or 60V operation made via switch
2. 60 Hz or 50 Hz transformers available
3. Load dependent
4. Tsus = Minimum time that the supply can sustain stated output voltage when ac is interrupted at full load.

330 MHz Forward Specifications

	Forward Sub-Split	Forward Mid-Split
Frequency Response (1)	5-30 MHz ±0.5 dB	174-330 MHz ±0.5dB
Minimum Full Gain (1) (2)		
Model 6501	24 dB	24 dB
Model 6502	30 dB	30 dB
Gain Control Range		
Selectable (6)	0. to 15 dB	0 to 15 dB
Equalization Range		
Selectable (7)	0 to 30 dB	0 to 30 dB
Thermal Compensation	(9)	(9)
Noise Figure (1)	13 dB	13 dB
Hum Modulation (10)	70 dB	70 dB
Output Level (5)	+46 dBmV	+46 dBmV
Distortion Specifications (2)		
Channel Loading	40	26
Cross Modulation	58 dB	61 dB
Composite Triple Beat (3) (10)	58 dB	61 dB
Second Order (4) (10)	71 dB	71 dB
Return Loss	16 dB	16 dB
Max. AC Thru Current	6A	6A
Test Point	−20 ±1 dB	−20 ±1 dB
Current Requirements at 24V dc	.35A	.35A

Operating Temperature
−40°C to +60°C (−40°F to +140°F)

Weight
Net: 0.34 kg (0.75 lb)
Shipping: 0.68 kg (1.50 lb)

NOTES:
1. Includes all losses with amplifier in normal operating configuration and 2 dB loss of interstage trim
2. At 68°F (20°C)
3. Carrier to average composite triple beat, unmodulated carriers, falling on any channel
4. Any $f_1 \pm f_2$ falling on any channel
5. Operational tilt = 7 dB forward only and sub-split, 3.5 dB mid-split
6. In 1 dB steps
7. In 1.5 dB upper frequency cable equivalent steps
8. Optional — Cable equivalent at upper frequency
9. 10 dB of cable plus 10 taps
10. Carrier to distortion ratio

Distribution Amplifier, Models 6501 and 6502

450 MHz Forward Specifications

	Forward Sub-Split	Forward Mid-Split	Forward High-Split
Frequency Response (1)	54-450 MHz ±0.5 dB	174-450 MHz ±0.5 dB	243-450 MHz±0.5 dB
Minimum Full Gain (1) (2)			
Model 6501	27 dB	27 dB	27 dB
Model 6502M	33 dB	33 dB	33 dB
Model 6502A	30 dB	30 dB	30 dB
Gain Control Range			
Selectable (6)	0 to 15 dB	0 to 15 dB	0 to 15 dB
Equalization Range			
Selectable (7)	0 to 30 dB	0 to 30 dB	0 to 30 dB
Thermal Compensation	(9)	(9)	(9)
Noise Figure (1)			
Model 6501	8 dB	8 dB	8 dB
Model 6502	6.5 dB	6.5 dB	7 dB
Hum Modulation (10)	70 dB	70 dB	70 dB
Output Level (5)	+46 dBmV	+46 dBmV	+46 dBmV
Distortion Specifications (2)			
Channel Loading	62	46	36
Cross Modulation	55 dB	56 dB	59 dB
Composite Triple Beat (3) (10)	55 dB	56 dB	59 dB
Second Order (4) (10)	70 dB	70 dB	70 dB
Return Loss	16 dB	16 dB	16 dB
Max. AC Thru Current	6A	6A	6A
Test Point	−20 ±1 dB	−20 ±1 dB	−20 ±1 dB
Current Requirements at 24V dc			
Model 6501	.43A	.43A	.43A
Model 6502M	.49A	.49A	.49A
Model 6502A	.58A	.58A	.58A

Operating Temperature

−40°C to +60°C (−40°F to +140°F)

Weight

Net: 0.34 kg (0.75 lb)

Shipping: 0.68 kg (1.50 lb)

NOTES:

1. Includes all losses with amplifier in normal operating configuration and 2 dB loss of interstage trim
2. At 68°F (20°C)
3. Carrier to average composite triple beat, unmodulated carriers, falling on any channel
4. Any $f_1 \pm f_2$ falling on any channel
5. Operational tilt = 7 dB sub-split, 5 dB mid-split, 4 dB high-split
6. In 1 dB steps
7. In 1.5 dB upper frequency cable equivalent steps
8. Optional — Cable equivalent at upper frequency
9. 10 dB of cable plus 10 taps
10. Carrier to distortion ratio

Reverse Specifications

	Reverse Low Gain Sub-Split	Reverse High Gain Sub-Split	Reverse Mid-Split	Reverse High-Split
Frequency Response (1)	5-30 MHz ±0.5 dB	5-30 MHz ±0.5 dB	5-108 MHz ±0.5 dB	5-174 MHz ±0.5 dB
Minimum Full Gain (1) (2)	20 dB	25 dB	30 dB	29 dB
	20 dB	25 dB	30 dB	29 dB
Gain Control Range				
Selectable (5)	0 to 15 dB	0 to 15 dB	0 to 15 dB	0 to 15 dB
Equalization Range				
Selectable (6)	0 to 15 dB	0 to 15 dB	0 to 15 dB	0 to 15 dB
Noise Figure (1)	6 dB	5.5 dB	7.5 dB	7 dB
Hum Modulation (7)	70 dB	70 dB	70 dB	70 dB
Input Level	+17 dBmV, flat	+17 dBmV, flat	+17 dBmV, flat	+17 dBmV, flat
Distortion Specifications (2)				
Channel Loading	4	4	12	22
Cross Modulation	57 dB	91 dB	87 dB	76 dB
Composite Triple Beat (3) (7)	–	–	87 dB	86 dB
Second Order (4) (7)	73 dB	73 dB	70 dB	75 dB
Return Loss	16 dB	16 dB	16 dB	16 dB
Test Point	−20 ±1 dB	−20 ±1 dB	−20 ±1 dB	−20 ±1 dB
Current Requirements at 24V dc	.11A	.55A	.46A	.46A

Operating Temperature

−40°C to +60°C (−40°F to +140°F)

Weight

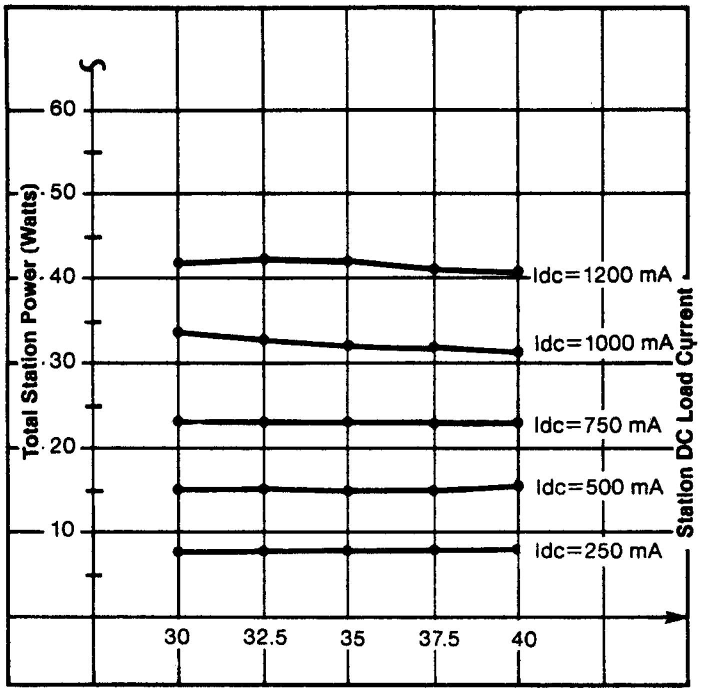
Net: 0.34 kg (0.75 lb)

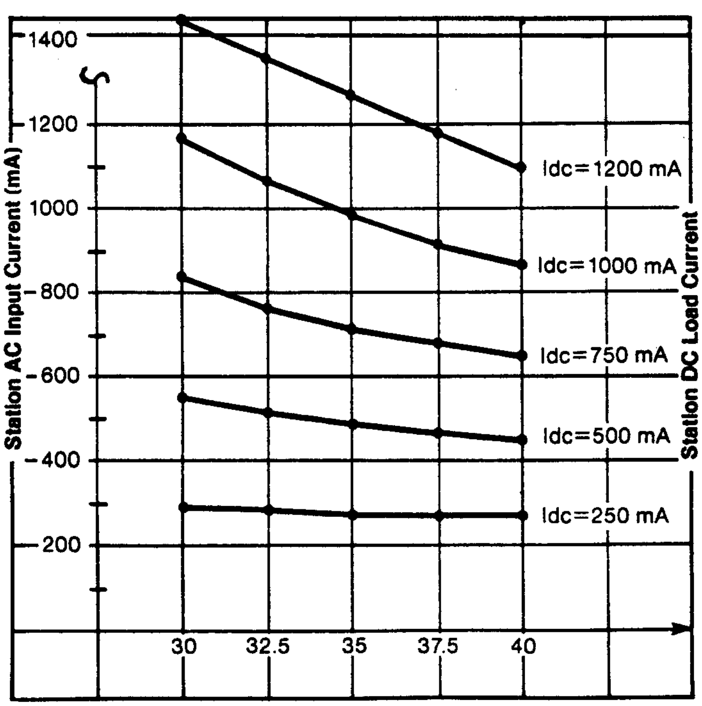
Shipping: 0.68 kg (1.50 lb)

NOTES:
1. Includes all losses with amplifier in normal operating configuration
2. At 68°F (20°C)
3. Carrier to average composite triple beat, unmodulated carriers, falling on any channel
4. Any $f_1 \pm f_2$ falling on any channel
5. In 1 dB steps
6. In 1.5 dB upper frequency cable equivalent steps
7. Carrier to distortion ratio

Distribution Amplifier, Models 6501 and 6502

Model 6501/6502 Switching Regulated Power System

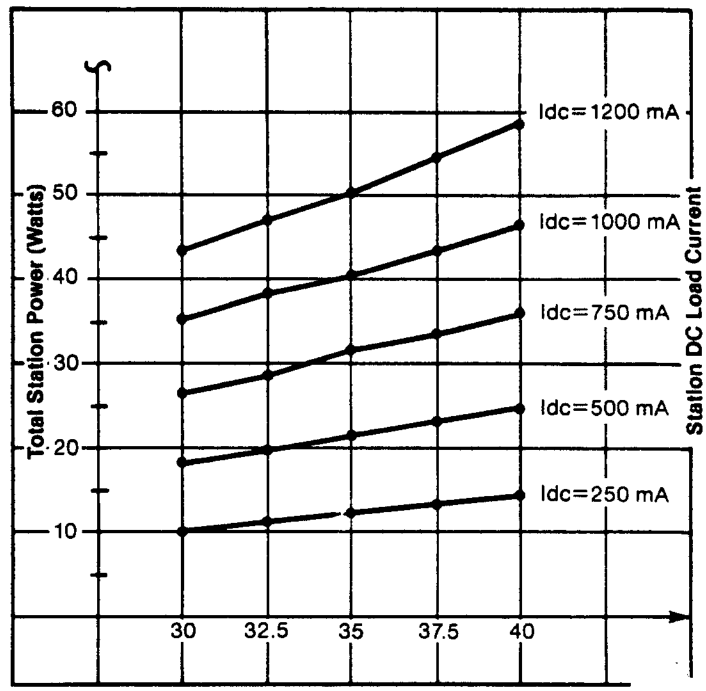


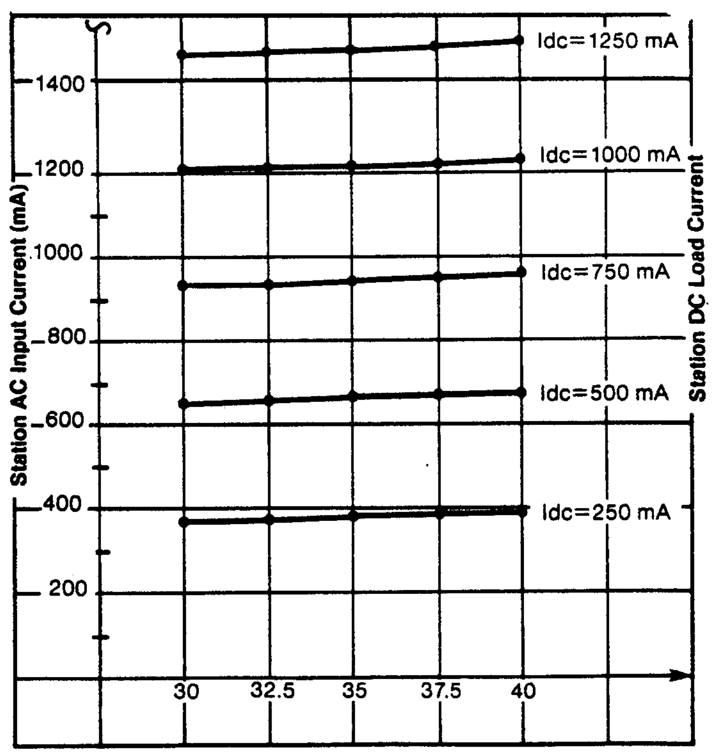


Station AC Input Voltage (volts)
40 Volt Tap

NOTE: AC Voltage and Current are True RMS Values

Model 6501/6502 Linear Power System



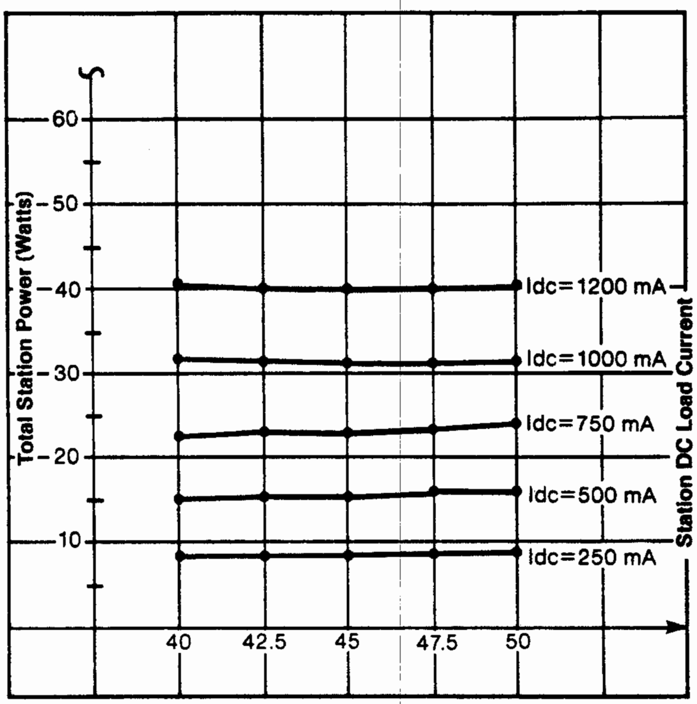


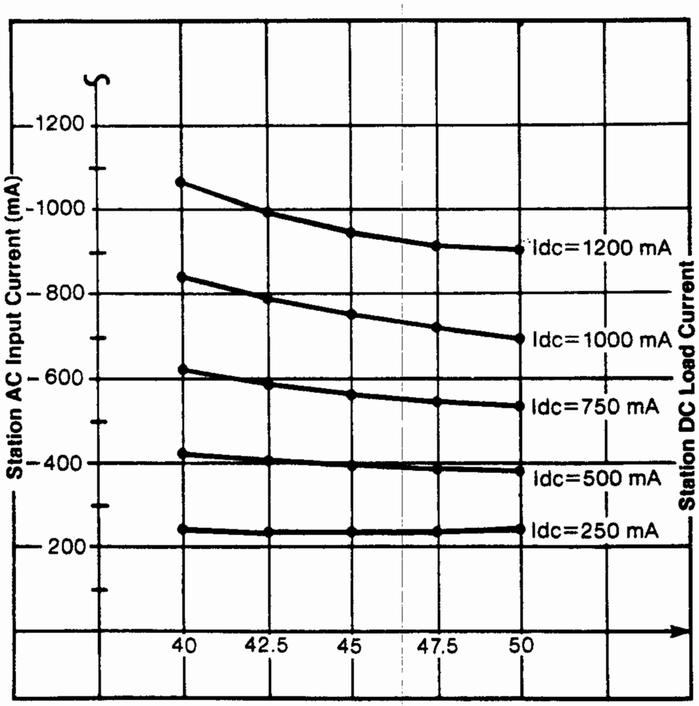
Station AC Input Voltage (volts)
40 Volt Tap

NOTE: AC Voltage and Current are True RMS Values

Distribution Amplifier, Models 6501 and 6502

Model 6501/6502 Switching Regulated Power System

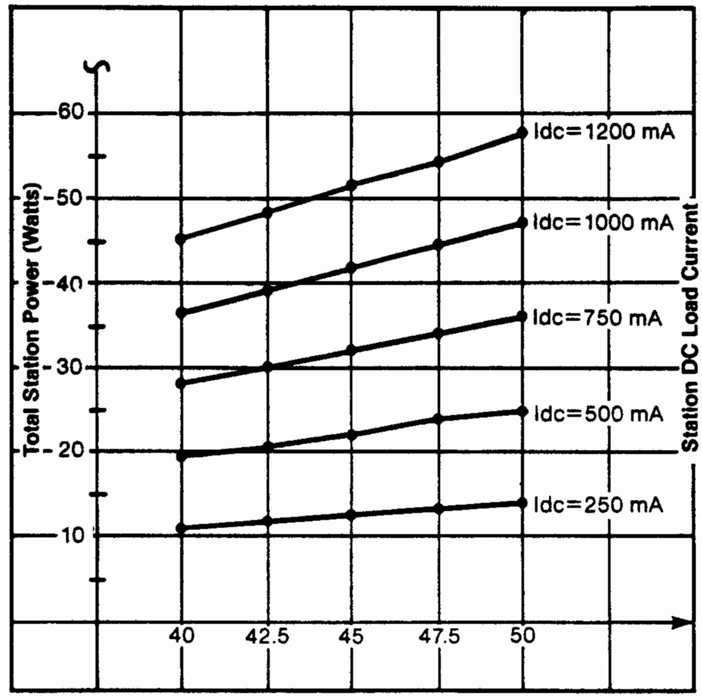


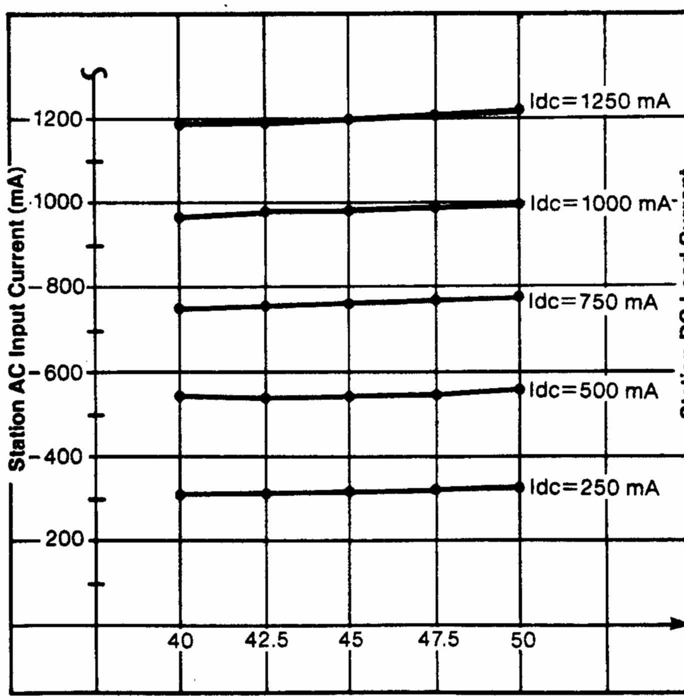


Station AC Input Voltage (volts)
50 Volt Tap

NOTE: AC Voltage and Current are True RMS Values

Model 6501/6502 Linear Power System

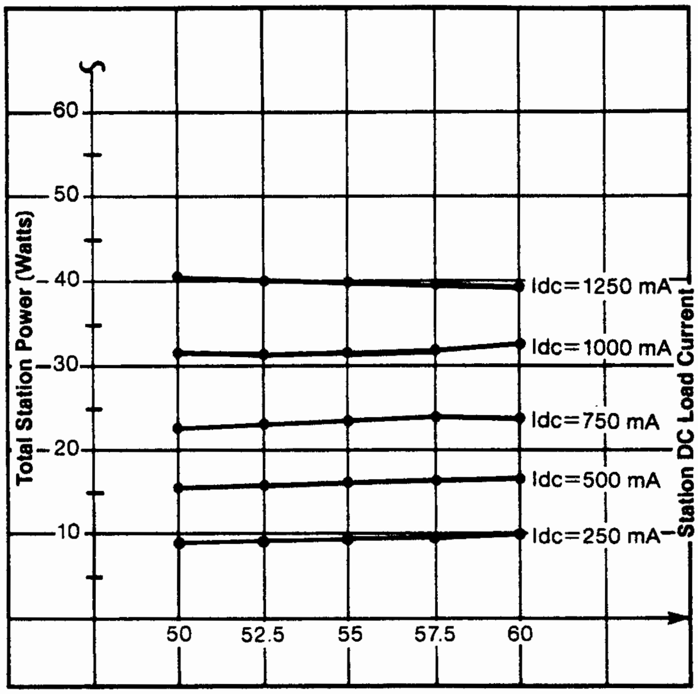


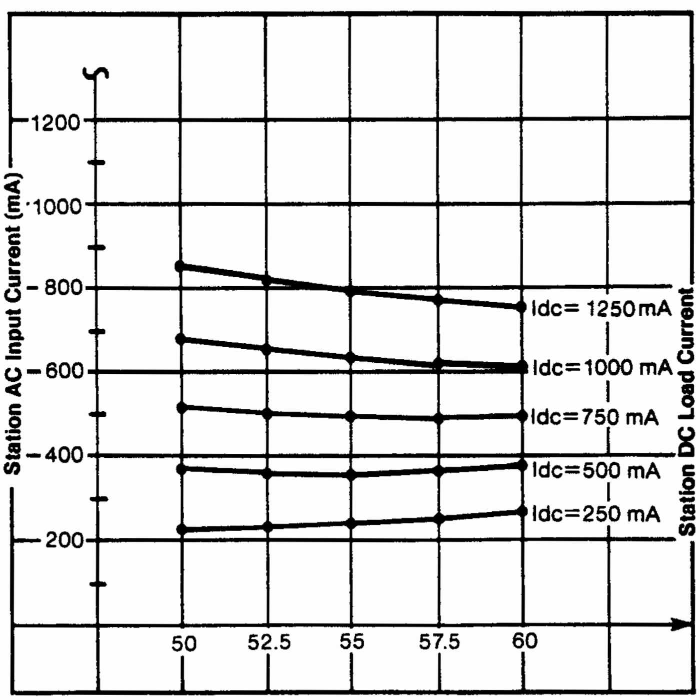


Station AC Input Voltage (volts)
50 Volt Tap

NOTE: AC Voltage and Current are True RMS Values

Model 6501/6502 Switching Regulated Power System

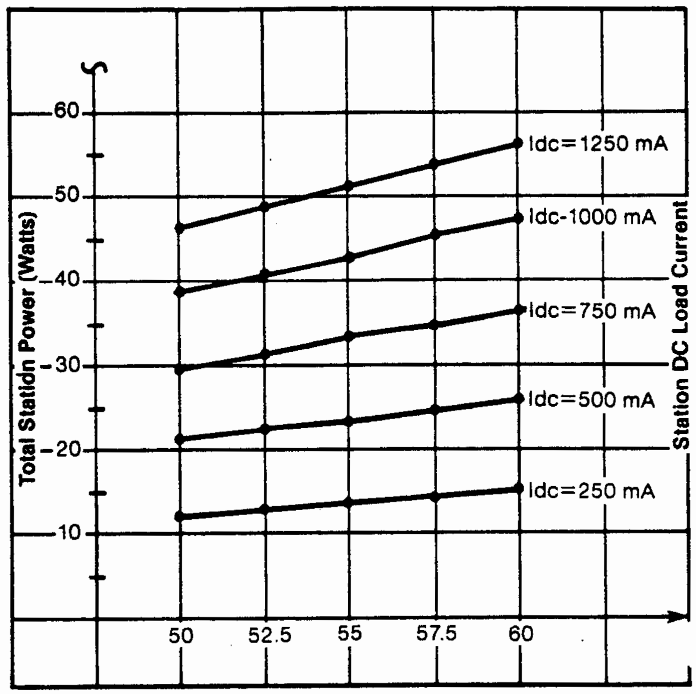


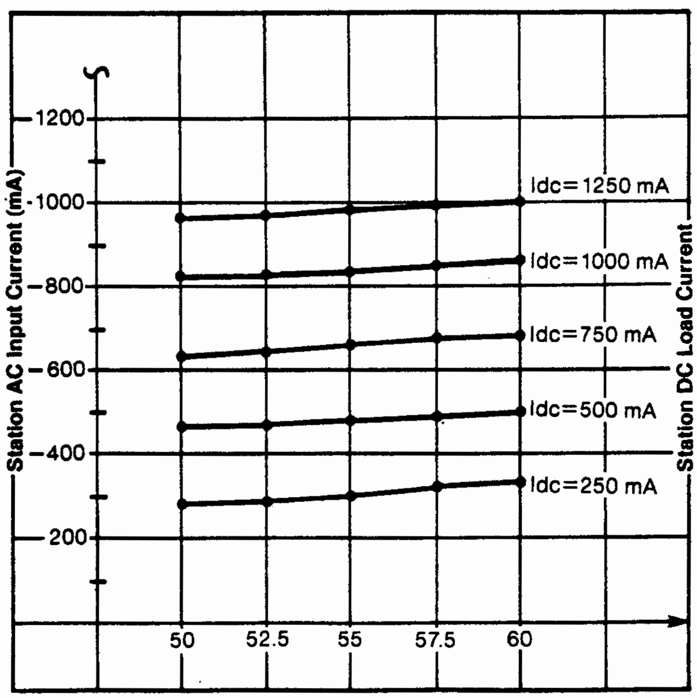


Station AC Input Voltage (volts)
60 Volt Tap

NOTE: AC Voltage and Current are True RMS Values

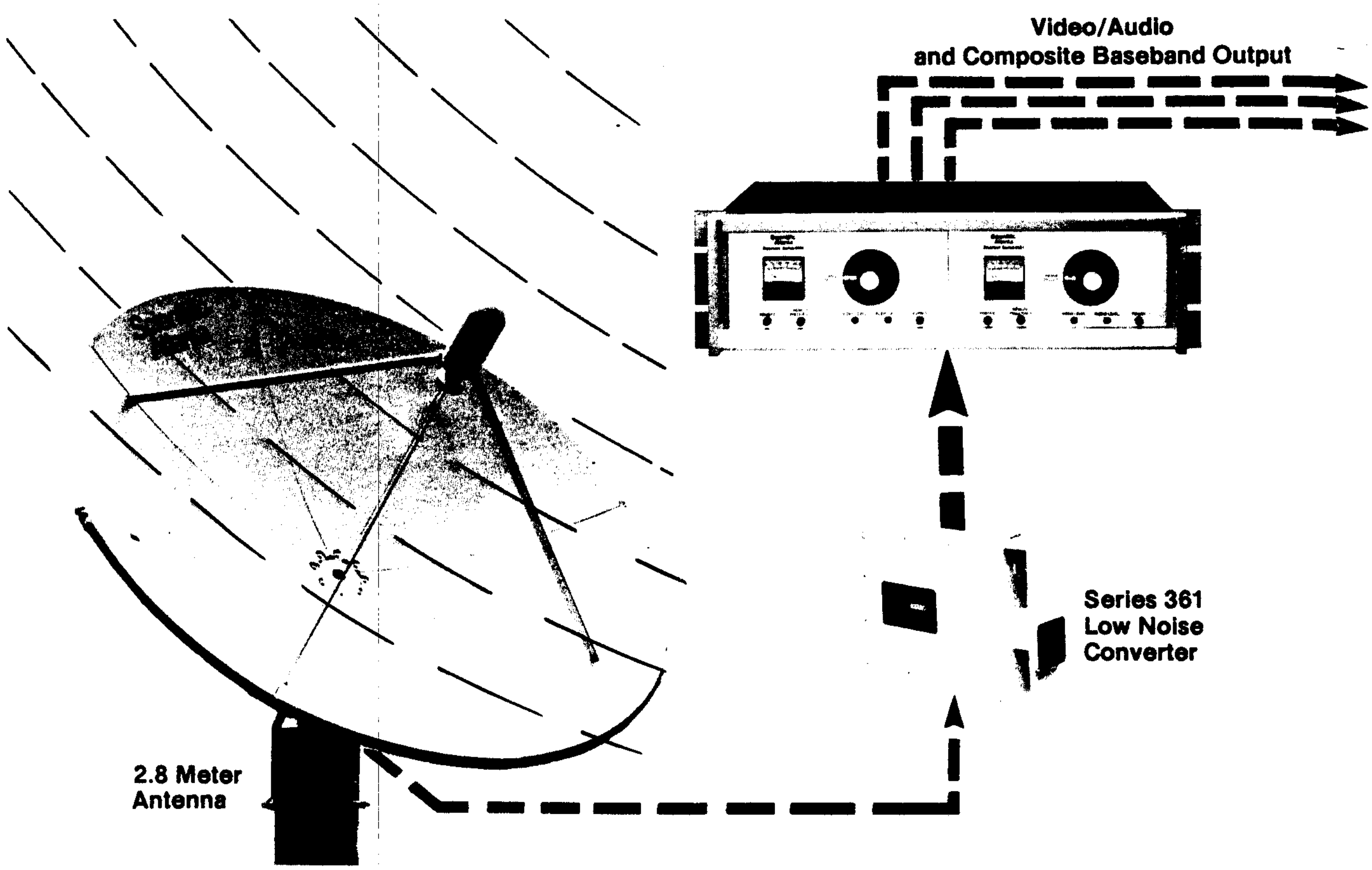
Model 6501/6502 Linear Power System





Station AC Input Voltage (volts)
60 Volt Tap

NOTE: AC Voltage and Current are True RMS Values



Ku-Band Receive-Only System

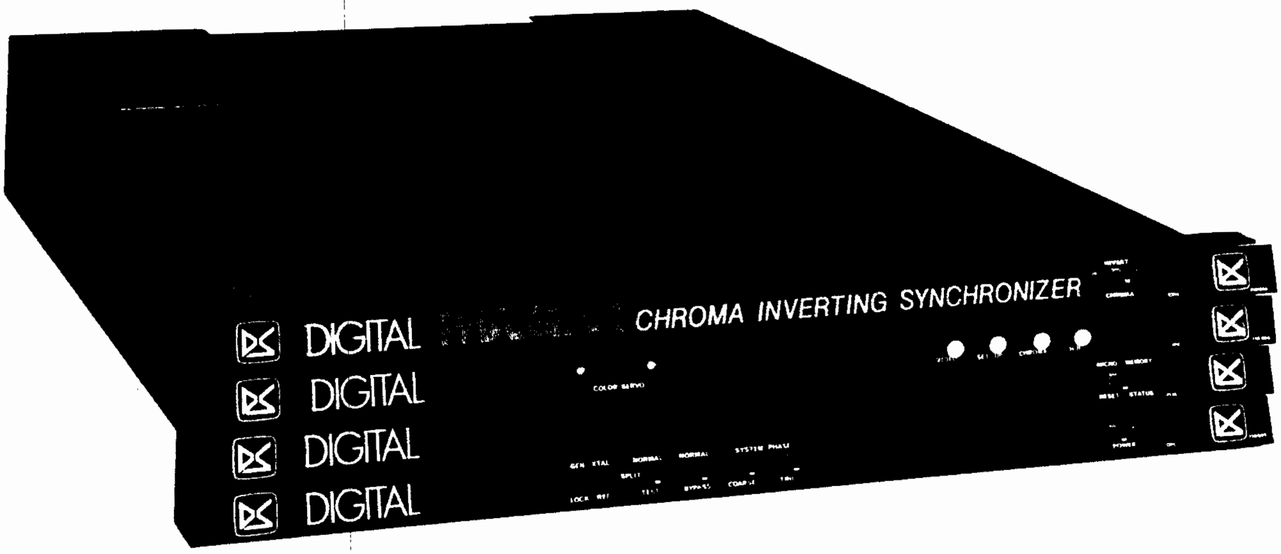
Scientific-Atlanta offers earth stations to receive satellite television transmissions from 11.7 to 12.2 GHz.

A typical Ku-band video receive terminal consists of the following:

- A Series 9000 Ku-band earth station antenna, with elevation-over-azimuth mount and single or dual polarized feed
- A Series 361 low noise converter
- A Model 6651 video receiver, with ANIK-C specifications (similar configurations for SBS, OTS, and other satellites are available)

Each system also includes 100 feet of coaxial cable to connect the low noise converter (LNC) to the video receiver, and installation and operation instructions.

Digital Frame Synchronizer, Phaser IV,V, and VI



Scientific-Atlanta, through its subsidiary, Digital Video Systems, offers a complete line of frame synchronizers for use in cable TV, LPTV, and broadcast applications. The Phaser IV, V, and VI series of synchronizers provide state-of-the-art technology in video synchronization.

The Phaser synchronizes input signals from an NTSC or black and white video source, including satellite, studio, network, remote camera, off-air, etc.

The Phaser lets the operator ignore time and distance, synchronizing signals automatically—such as Electronic News Gathering (ENG) programming transmitted via microwave—as if they originated in-house.

The Phaser eliminates complex signal routing and timing techniques. Signals from inaccessible cameras, remote vans, or moving vehicles are timed and synchronized into a fixed based control where they can be mixed, dissolved or used in chroma key or special effects with other sources.

With the Phaser, the operator can easily mix any live video inputs with studio programming, microwave, satellite, and VTR signals to produce a completely synchronized output without gen-locking.

The Phaser replaces the complex system of delay lines and pulse delay compensators which may now be used to time several studios through master control.

And for infinite switcher re-entry, which previously required very costly equipment, the Phaser is a breakthrough.

The Phaser IV with a 1.5 million bit RAM fieldstore memory is the basic "building block". Phaser V and Phaser VI employ the same modular design concept.

The Phaser V synchronizer builds onto the exceptional features of the Phaser IV. It has a framestore memory to provide double the memory capacity with half the number of components.

The Phaser VI takes the concept of synchronizing a step further, incorporating a new picture adaptive digital comb filter to provide the highest quality NTSC chroma inversion ever. It ensures virtually perfect decoding of the NTSC color TV signal as required in freeze frame applications or for studio-to-studio synchronization and switcher re-entry.

Rugged Packaging
The Phaser synchronizers are compact and rugged. Measuring only 3-1/2" high, they fit any surroundings, from studios and master control rooms to compact mobiles, and in standard headend equipment racks.

Simplicity Designed-In
Designed for ease of use, Phaser controls are conveniently placed on the front louvres in a logical arrangement. Functional controls located on the circuit boards are easily accessed and work at the flick of a switch. Adjustments are rarely necessary as one of the main benefits of the Phaser is the reduced need for service. Signals are phased automatically.

Easy Maintenance
LED's on the Phaser front louvres indicate operational status and simplify fault identification. If service is required, boards are quickly replaced to reduce downtime to a minimum.

Digital Frame Synchronizer, Phaser IV, V, and VI

Reliability
Digital Video Systems' units are checked for quality control and tested to ensure reliable service and long life.

The Smart Synchronizers
The Phaser IV has a 1.5 million bit 16K RAM fieldstore memory while the Phaser V and the Phaser VI have a 4 million bit memory provided by 64K RAMs.

Automatic Diagnostics
Memory fault diagnostics and fault concealment is automatic and continuous. Memory chip failure is detected and hidden and an LED readout code on the board identifies which chip is faulty.

Total Control
The Phaser's unique integral digital processing amplifier ensures precision signal level control. It replaces sync, burst, and blanking anc adjusts video gain, black level, chroma gain and hue. All of the functions of a processing amplifier are obtained, plus a digital frame synchronizer, in a single compact unit. The Phaser eliminates picture shifts in switching and editing by converting all incoming NTSC signals to RS-170A with a sync to burst relationship that doesn't vary.

Automatic Hot-Switching
The Phaser automatically hot-switches between unrelated non-synchronous signals, holding the last complete field until the next valid vertical interval appears at the input. An internal mode selection switch provides the alternatives of: passing the input video regardless of input signal status; hot-switch mode with freeze and drop to black after a few frames if the input is not restored; and hot-switch mode with freeze until the input is restored.

Instant Performance Evaluation
The internal digital calibration signal generator provides for precision alignment of the system analog output. This test signal can also be routed from analog output to analog input to verify system status by side-by-side comparison every 16 lines of system throughput against the digital test signal output.

Any misalignment of the analog input is then easily and accurately corrected. This feature gives instant system performance evaluation and precision alignment.

No Noise or Jitter
The 5 TV line hysteresis function overrules the motion discontinuities which may occur when frames are deleted or inserted, even when the incoming signal is very noisy or has jitter.

Videotext/Teletext
With full bandwidth and precise digital blanking the Phaser synchronizers pass VIRS, closed caption subtitling, and teletext without degradation.

Chroma Inversion
For the first time, chroma inversion is a totally acceptable mode of operation, even for the highest quality prime time programming. The Phaser VI picture adaptive digital comb filter sets new standards for an NTSC decoder, with virtually undetectable chroma inversion.

Wide Application
The Phaser IV is primarily designed for use in broadcast and industrial TV, cable systems and production houses which require basic, low-cost synchronizing capability.

The Phaser V will meet the sophisticated requirements of those needing a full frame synchronizer while the Phaser VI meets the most stringent specifications, including an output picture that never shifts.

Cable TV
Using the Phaser in cable TV has measurable benefits. Synchronizing incoming demodulated feeds prior to cable distribution substantially reduces co-channel interference and improves the signal-to-noise ratio. With a synchronous system, the operator can insert commercials, key in messages, provide the viewer with stable channel changes, and reduce the cost of set-top data distribution equipment.

The Phaser synchronizes signals from headend demodulators, satellite receivers, and microwave links to your local programming facility, providing highest quality processed video with new digitally-generated synch, burst, and blanking for your cable distribution system.

Infinite Switcher Re-Entry
When used for switcher re-entry, the Phaser VI gives your existing switcher a flexibility which was previously available only with the most expensive equipment. The Phaser makes it possible to feed the output of any mix effects buss into the input of any other.

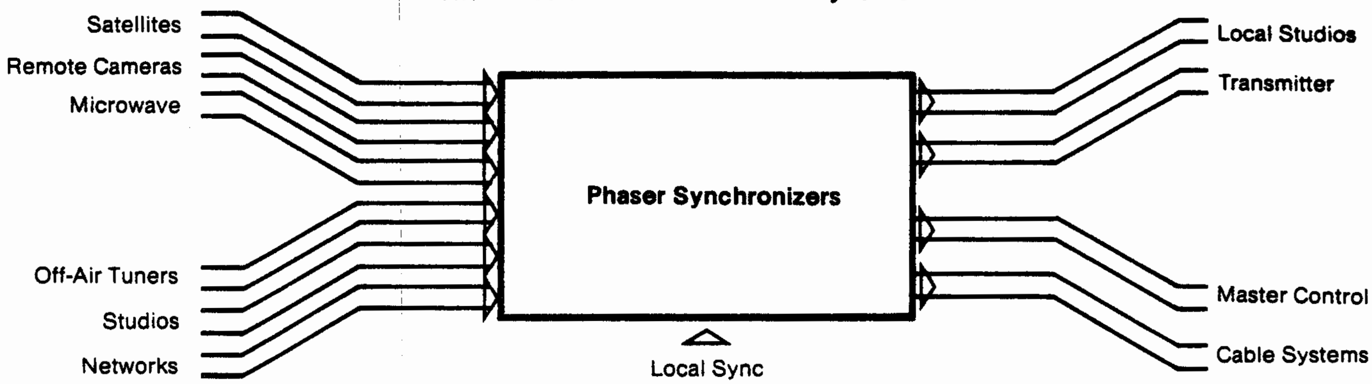
This infinite signal re-entry multiplies the capabilities of your switcher while providing a quality never before obtainable. (All analog timing delay lines can be eliminated if you choose.) This can be especially beneficial in local origination studios.

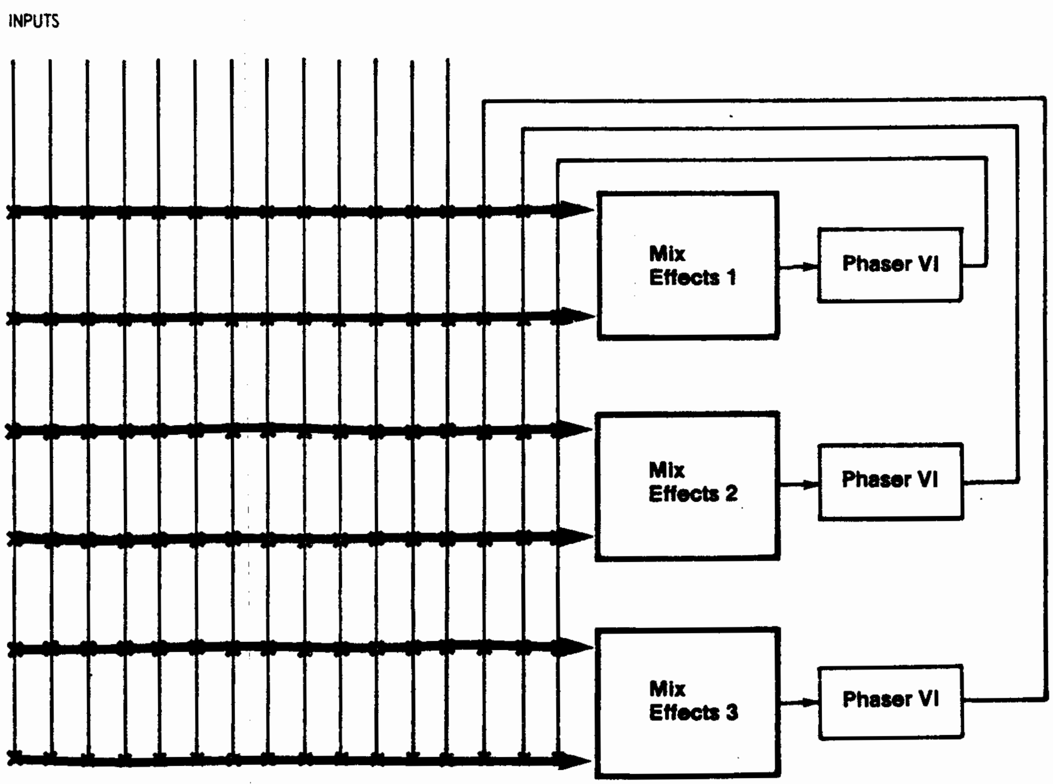
Digital Frame Synchronizer, Phaser IV, V, and VI

Features

- Digital Quality and Reliability
- Microprocessor Control
- Integral Digital Processing Amplifier
- Automatic Memory Fault Diagnostics and Concealment
- Internal, Digital Calibration Generator
- Modular Design Provides for Easy Maintenance
- Instant Performance Evaluation

The Phasers — Multi-Functional Synchronizers





Digital Frame Synchronizer, Phaser IV, V, and VI

Specifications

Mainframe

Digital Sampling
  Phaser IV
    8 bits at 14.3 MHz (256 levels at 4 times NTSC subcarrier)
  Phaser V
    8 bits at 14.3 MHz sampling on the R-Y, B-Y vectors (256 levels at 4 times NTSC subcarrier)

Bandwidth
  ±0.5 dB to 4.2 MHz, down 3 dB at 5.5 MHz

Signal-to-Noise Ratio
  Better than 56 dB (rms noise referenced to peak-to-peak quantizing range from dc to 4.5 MHz bandwidth)

Differential Phase
  < 2°

Differential Gain
  < 1%

Processing Amplifier
  Digital sync, burst, and blanking are added to the output video RS-170A specification

Front Panel Operating Controls
  Video level, set-up, hue, chroma gain, systems horizontal phase, subcarrier phase, and bypass

Input Signals
  Video
    1.0V = 3 dB composite video at 75 ohms
  Reference Video
    High impedance loop-thru input
    Any color video signal with 40 IRE units of synchronization and burst = 3 dB, where burst has a non-varying phase relationship to sync.

Output Signals
  Outputs 1, 2, and 3
    1.0V composite analog at 75 ohms

Bypass
  In the event of power failure, VIDEO IN is directly connected to VIDEO OUTPUT 1. Also switch activated.

Calibration Generator
  1.0V composite video switched into all three video outputs digitally generated

Test Signals
  Full field color bars and multiramp

  In "wrap-around" mode, output and throughput are compared on a 16-line alternate basis

Color/Black and White Operation
  Automatic burst sensing and switchover to black and white mode; burst is turned off on the output

Oscilloscope Trigger
  Input source video and output video trigger at horizontal or vertical rates

Phaser IV

Correction Range
  262.5 TV lines

Hysteresis
  5 TV lines

Dimension
  89H x 406W x 572D mm (3.5H x 16W x 22.5D in.)

Weight
  22.7 kg (50 lb)

Power Requirements
  112V ac (nominal), 60 Hz, 250W max

Operating Temperature Range
  0°C to +40°C (+32°F to +104°F)

Phase V

Correction Range
  512 TV lines

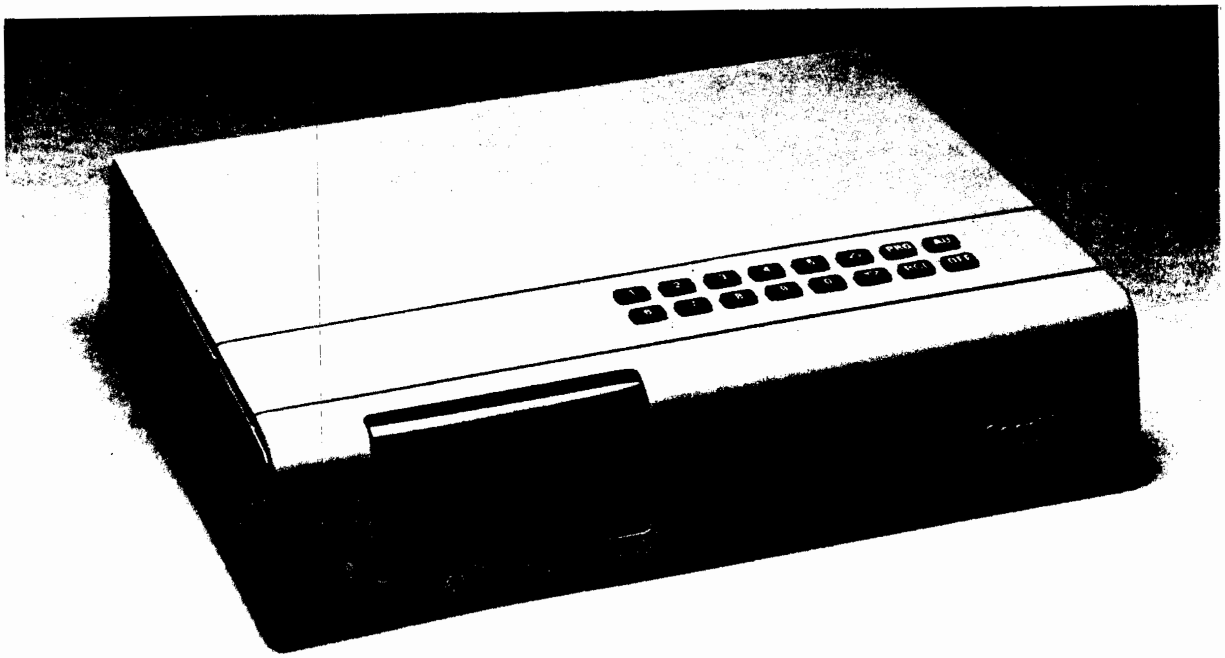
Hysteresis
  5 TV lines

Power Requirements
  115V ac (nominal), 60 Hz, 400W max

Phaser VI

Same as Phaser V, plus 3-line adaptive comb filter

Set-Top Terminal, Series 8500



The Series 8500 Scientific-Atlanta set-top terminals are available in three unique set-top models, all offering fully electronic, microprocessor-based control. Series 8500 set-top terminals operate at frequencies up to 440 MHz and can deliver 128 channels in a dual cable system.

The Series 8500 Programmable set-top terminal includes a long list of standard product line features, including programmable frequency allocation and a built-in remote control receiver.

The Series 8500 Programmable set-top terminal with Descrambling uses dynamic switched sync suppression for signal security. Dynamic switched sync suppression is an important Scientific-Atlanta innovation that provides economical security that is extremely difficult to defeat.

The Series 8500 Addressable set-top terminal is an advanced electronic set-top terminal providing head-end control for service level authorizations and pay-per-view events. The addressable CATV system requires computer capability, software programs, several headend pieces, and personnel orientation. Scientific-Atlanta offers two versions of addressable control systems, with complete hardware and software, site preparation, and personnel training.

- The System Manager II is a comprehensive, stand-alone computer system providing connections for multiple CRT's and disc memory with a data base for subscriber information and inventory control. The System Manager II can operate independently as a fully automated central computer system and can interface with a host billing system.
- The Addressable Control Unit (ACU) is a microcomputer-based, addressable headend system. The ACU data base is keyed to converters only, and is designed to interface with a host computer containing both business and billing software. It can also be used in a partially automated addressable control system where subscriber data base information is managed from a manual filing system.

Control Key Functions — Both Set-Top Terminal and Series 8500 Remote Control Unit.

Increment/Decrement Keys — Steps up or down consecutively through all authorized channels.

RCL Key — Steps through up to 20 favorite channels in memory.

OFF Key — Switched ac feature permits on/off control of the television from the set-top unit.

AUTH Key — To request channels subject to parental discretion, press this key and enter a 5-digit secret code.

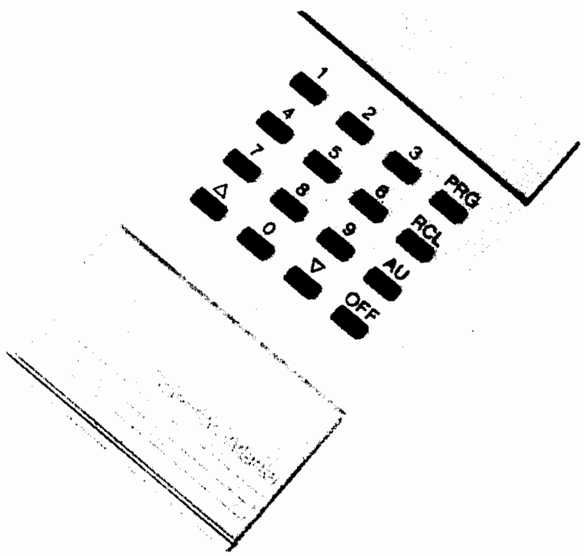
Program Key — To program favorite channel memory, press this key and enter desired channel number.

Set-Top Terminal, Series 8500

Features

Series 8500 Set-Top Terminal

- Handsome, High-Tech Exterior
- Bright LED Display
- Extra Surge Protection
- Lightning Path Protection
- Coded Serial Communications Protects Against Internal Tampering
- Advanced Automatic Manufacturing Technique
- Well Ventilated
- UL Listed
- Protective Pads



Series 8500 Remote Control Unit

- Remote Control Receiver Standard on all Series 8500 Set-Tops
- Duplicates all Keypad Functions
- Dual Message Pulse Infrared System Eliminates False Activation
- Ultra Lightweight

Series 8500 Programmable

The many standard features of the Series 8500 set-top terminals provide important operational benefits and subscriber conveniences. The Series 8500 Programmable set-top terminal incorporates all of these standard features, is totally electronic and authorization instructions are PROM controlled. Scientific-Atlanta also offers a PROM programmer which can be used routinely by non-technical personnel. Because all terminals include a remote control receiver, subscriber requests for hand-held remote control units can be easily filled by mail. A service call to replace converters is not necessary.

A subscriber-preferred feature of the Series 8500 Programmable is PROGRAMMABLE FREQUENCY ALLOCATION which enables the set-top terminal to associate any carrier frequency with any channel number. This unique Scientific-Atlanta feature reduces subscriber confusion by simplifying in-home set-top usage and satisfies FCC requirements for carrying off-air stations on channel.

Programmable frequency allocation can also be used to assign offsets wherever necessary, and to avoid "holes" in the channel lineup caused by FAA interference. By simply eliminating the frequencies of channels that cannot be carried on a given CATV system, all remaining channels can be assigned consecutively.

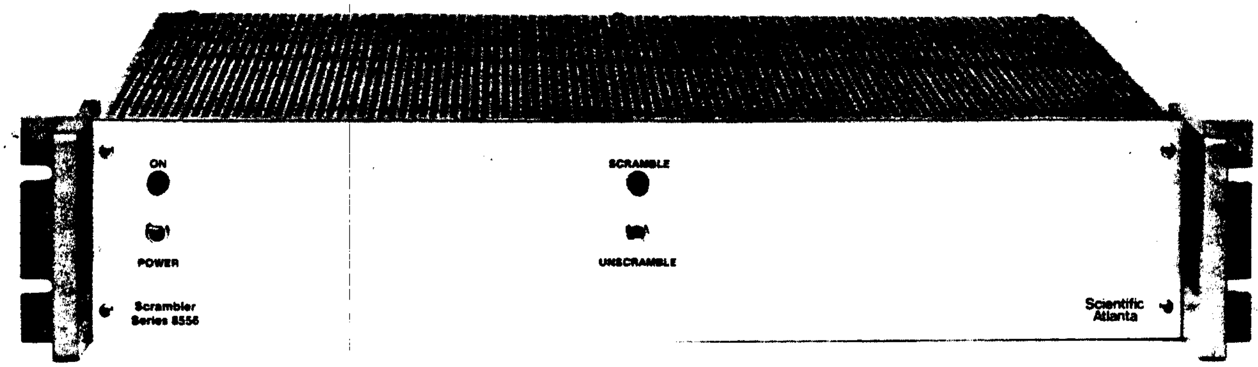
The Programmable set-top terminal can be configured to tune to a predetermined DEFAULT FREQUENCY when a subscriber attempts to tune a channel that is not included in his service level. The subscriber, therefore, neither sees scrambled video nor hears unauthorized audio. The default frequency can be used as a "barker" channel to promote premium services or pay-per-view events.

PARENTAL CONTROL can be applied to any channel on the Series 8500 set-top terminal. If a channel subject to parental discretion is requested without entering the correct authorization code, the set-top terminal will tune to the "barker" channel frequency. No objectionable language will enter the home, since audio is also denied. Access is provided by a five-digit code.

The Series 8500 set-top terminal offers an advantage in dual cable systems because the EXTERNAL DUAL CABLE SWITCH is located at the point the cable enters the home. Switching between the A and B cable is automatic. Channels can be numbered consecutively from 1 to 128, therefore subscribers do not have to distinguish between A and B cables when changing channels.

Subscriber convenience features include INCREMENT and DECREMENT keys. The Series 8500 set-top terminal will step up or down, one at a time, through all authorized channels. The unit will skip over any unauthorized channels completely, showing neither blank screens nor snow.

The FAVORITE CHANNEL MEMORY feature enables subscribers to program up to 20 frequently watched channels into the set-top terminal, and changed at any time. By pressing the recall key (RCL), the unit will step through all favorite channels in the order they were stored in memory.



Series 8500 Programmable with Descrambling

The Scientific-Atlanta Series 8500 Programmable with Descrambling set-top terminal introduces virtually unbreakable signal security using broadband transmission, while incorporating all standard Series 8500 features. Scientific-Atlanta offers DYNAMIC SWITCHED SYNC SUPPRESSION, an innovative and now indispensible scrambling/descrambling method to secure premium programming and pay-per-view channels.

Dynamic switched sync suppression increases signal security by introducing random timing elements into the scrambling process. Because of the complexity of the scrambling technique, internal tampering and reproducible scrambling defeats are extremely difficult.

The Series 8500 descrambler uses a custom-manufactured timing chip to delay the restoration of the video sync tip a few microseconds after the reference pulse is received. The timing delay for every video frame is changed randomly. Because of this Scientific-Atlanta development, the CATV operator benefits from superior signal security.

Features

- Dynamic Switched Sync Suppression
- Full Line Headend Scrambling Accessories
- Custom-Manufactured Timing Chips that cannot be Purchased Commercially
- All Series 8500 Standard Product Features

Series 8500 Addressable

The Scientific-Atlanta Series 8500 Addressable set-top terminal is part of an integrated addressable control system which automates both CATV business management and the way changes in subscriber service levels are carried out. The features of the Series 8500 Addressable have been developed to maximize the benefits of addressability with user-friendly and efficient software, large capacity for premium programming and pay-per-view events, additional security measures, and an efficient interface with a business management system.

CENTRAL CONTROL OF AUTHORIZATIONS is the basic function of addressability. All additions and deletions of program tiers, single channels, pay-per-view events, and program promotions can be accomplished from the headend without service calls, thus reducing tremendous overhead costs. Execution of pay-per-view events also eliminates costly field work to place and remove traps.

GLOBAL AUTHORIZATION AND DEAUTHORIZATION of pay-per-view events is a far-reaching Scientific-Atlanta development for addressable systems. A single command simultaneously authorizes or deauthorizes all Scientific-Atlanta set-top terminals entitled to receive a particular program event, enabling the cable operator to present back-to-back pay-per-view programs on the same channel. A global command can also be used to authorize all addressable set-top terminals in a system, regardless of service level, for a pay-per-view preview or other promotional offering.

Global commands can also be applied to SERIES EVENTS. A customer might request all segments of a series of special programs, or only particular segments. A global command authorizes all set-top terminals at the outset of each segment and deauthorizes them at the end. Instructions to authorize set-top terminals requesting additional segments are stored in memory until a series deauthorization command is issued.

ADVANCE PAY-PER-VIEW PROGRAM LOADING makes the global command format practical. Authorization instructions for up to 25 pay-per-views, including multiple-segment series, can be entered at any time before a program begins. This capability eases the problem of handling last-minute order entry rushes. At the precise starting and ending time of a pay-per-view event, all pre-loaded set-top terminals are turned on or off simultaneously by a global command.

An ELECTRONICALLY ALTERABLE READ-ONLY MEMORY (EAROM) stores all authorization information. The EAROM stores all channel authorization and scrambled channel information, and it can be changed from the headend. It is not affected by power outages or by unplugging the set-top terminal in the home. This feature makes advance program loading possible by protecting stored information and by eliminating the need to readdress the units repeatedly.

Two important security checks can be performed on the Series 8500 Addressable set-top terminal. The LEGAL TERMINAL TEST discourages theft of terminal units, migration of set-top terminals from one CATV system to another, and helps control delinquent accounts. The headend stores the digital address of all set-top terminals authorized in a particular system and instructs all unauthorized set-top terminals to shut down.

The REFRESH TIMER SIGNAL prevents program theft by thwarting attempts to trap out deauthorization instructions from the headend. A software-based timer in each Series 8500 Addressable set-top terminal will expire and disable the unit unless it is periodically refreshed from the headend. The refresh signal can be set at the headend for various time intervals ranging from one-half hour to sixty-four hours, according to the CATV operator's needs. This capability ensures that each home set-top terminal is operating at the service level for which it is billed.

Feature Summary

Feature	Programmable	Programmable With Descrambling	Addressable
Programmable Frequency Allocation	X	X	X
Unauthorized Channel Default to Barker	X	X	X
Parental Control Function	X	X	X
Automatic Dual Cable Switching	X	X	X
Increment/Decrement	X	X	X
Favorite Channel Memory	X	X	X
Remote Control Receiver	X	X	X
Switched AC for TV	X	X	X
Dynamic Switched Sync Suppression Scrambling		X	X
Central Control of Authorizations			X
Central Execution of Pay-Per-View Functions			X
Global Authorization and Deauthorization			X
Pay-Per-View Preview			X
Advance Pay-Per-View Program Load			X
Non-Volatile Memory			X
System Security-Legal Terminal Test			X
System Security-Refresh Timer Signal			X

Specifications

Environmental

Temperature
0°C to 45°C (32°F to 113°F)
Relative Humidity
5 to 95%

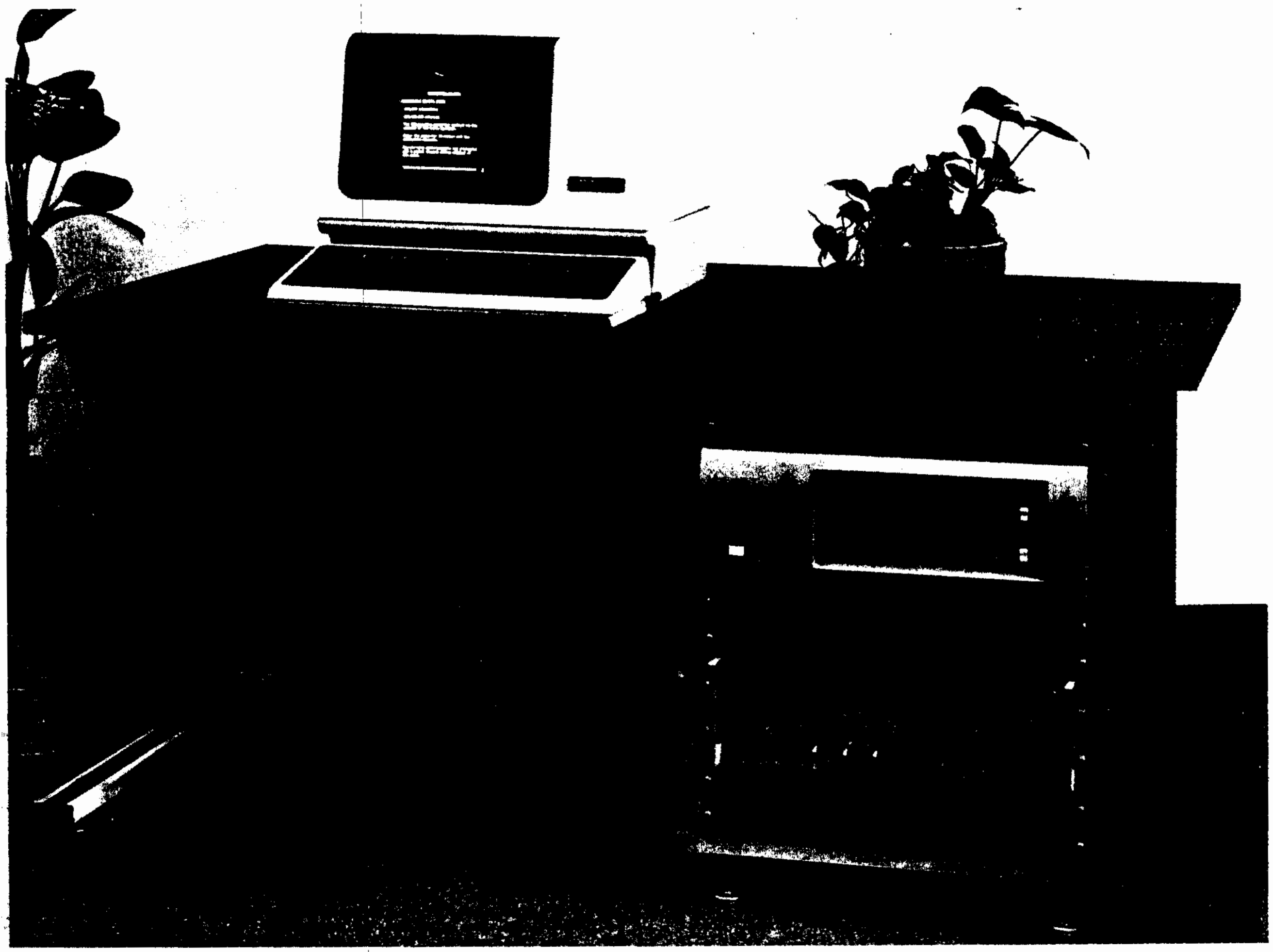
Electrical

Bandwidth
54 to 440 MHz
Number of Channels
64 with single cable
128 with dual cable
Output Channel
3 or 4
Flatness (over channel bandwidth)
±2 dB
Gain
0 to 9 dB
Noise Figure
13 dB typical
Return Loss
Input: 8 dB minimum on tuned channel
Output: 12 dB minimum
Isolation Input/Output
60 dB
Spurious
Input: −37 dBmV
Output: −57 dBmV
Frequency Accuracy
±100 kHz
Frequency Stability (0°C to +50°C)
±100 kHz
AC Input Range
105V to 125V ac
Power Consumption
20 watts maximum
Power Supply Surge Protection
Yes
Distortion at +15 dBmV; 60-Channel Load
Flat Input
Second Order Intermodulation: −57 dB
Cross Modulation: −57 dB
Compositie Triple Beat: −57 dB
Input Level
−7 to +20 dBmV

Mechanical

Dimensions
264.2L x 203.2W x 53.3H mm (10.4L x 8W x 2.1H in.)
Weight
2.5 kg (5.5 lb)
Keyboard Type
16-position, X-Y matrix
Display Type
LED, 12.7 x 6.6 mm (0.5 x 0.26 in.)

Addressable Control Unit, Series 8552



The Series 8552 Addressable Control Unit (ACU) offers the CATV operator reliable computer hardware and software for managing systems using Scientific-Atlanta Series 8500 set-top terminals. The ACU is a device controller. It can function as an intelligent peripheral to a host computer, integrating converter-based addressable commands with a data processing and billing system. It can also serve as a stand-alone system in which subscriber information is cross-referenced in a manual filing system.

Scientific-Atlanta will assist the cable operator in evaluating his needs for addressable control systems. Extensive customer support includes site preparation, hardware installation, personnel training, full documentation, user-friendly software, and an interface document for a host computer system.

The ACU is based on an Intel 8085 microprocessor. The ACU stores all information necessary to control each Series 8500 Addressable set-top terminal individually, or to control addressable terminals in the cable system by global commands. The ACU has a non-volatile memory and retains all stored information until specifically updated. This prevents the loss of a data base in a power failure, or because the communications link with a host computer is temporarily broken.

The Addressable Control Unit can be configured to control up to 120,000 set-top units in a single cable system or up to 70,000 units in a dual cable system. The ACU uses a Digital Equipment Corporation (DEC) Series VT100 CRT and requires a standard RS-232C electrical interface. In stand-alone systems, all order entry is accomplished on this keyboard. In a host computer system, a CRT is used primarily for system status checks and system redundancy.

The ACU system also includes an Addressable Transmitter (ATX). The ATX, located at the headend, accepts converter control data from the ACU and transmits the data over the cable system. It operates at a standard frequency of 108.2 MHz. The unit also generates converter control signals needed to keep all legal Series 8500 Addressable set-top terminals active. Output data is transmitted as an FSK-modulated digital signal at a data rate of 19,200 baud. The ACU can serve multiple headends, with one ATX required at each headend. Two ATX's can be installed in tandem for system redundancy at each site.

The Series 8500 addressable set-top terminal contains a non-volatile memory which can be changed only from the headend. Advance pay-per-view program authorization and global channel authorization functions are possible because the non-volatile memory can store authorization instructions indefinitely, even through power outages.

A set-top terminal can be pre-authorized to receive a certain pay-per-view event at any time prior to the beginning of the program. This eases the problem of handling last minute order entry rushes. At the beginning and end of a pay event, all Series 8500 set-top terminals entitled to receive the program can be turned on or off simultaneously with one global command from the headend. A global command can be issued system-wide to authorize a channel temporarily for a pay-per-view preview or other program promotion.

The Series 8552 ACU system provides an innovative format for central control of authorizations, pay-per-view management, and additional system security. The ACU data base is keyed to converter identification, providing control of individual converters and of system parameters. Individual converters can be addressed to begin service, change service levels, authorize single channels or tiers, or to disable the set-top unit. Scientific-Atlanta's unique global command format increases the efficiency of its pay-per-view management system, enabling the cable operator to run back-to-back pay-per-view programs on the same channel.

ACU software provides a legal terminal test, and ATX software provides a refresh timer signal for additional system security. These two security features are designed to discourage theft of terminals and migration of set-top units from one CATV system to another, to help control delinquent accounts, and to thwart attempts to circumvent deauthorization instructions from the headend.

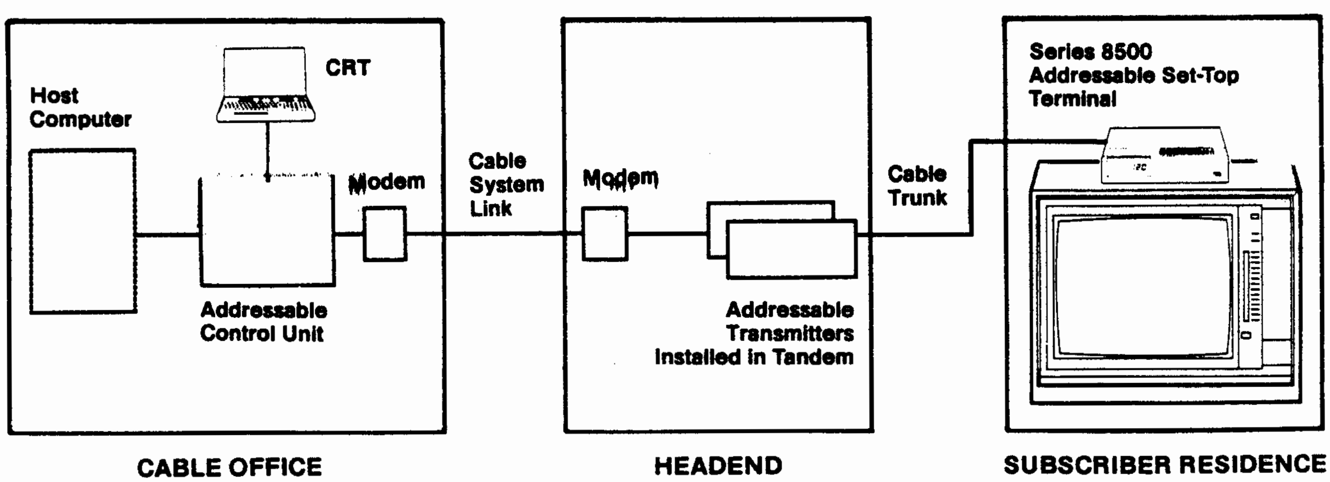
The legal terminal test checks to see if converters are operating at the correct service level. This test is conducted system-wide using a global command. If a set-top unit does not respond correctly to the legal terminal test, it is automatically shut down.

The refresh timer signal is an on-going function of the Addressable Transmitter. The ATX continually transmits an encrypted message which the set-top terminal must decode in order to stay active. Any attempt to interrupt communications between the headend and the converter prevents the converter from receiving the refresh signal, and the unit will be disabled. The refresh signal can be set from the ACU for various time intervals ranging from one-half hour to sixty-four hours.

Features

- Individual Converter Enable/Disable
- Individual Converter Channel Authorization/Deauthorization (change of service level)
- Advance Pay-Per-View Event Loading
- Global Authorization of a Pay-Per-View Event (only pre-loaded converters)
- Refresh Timer Signal
- Legal Terminal Test
- Display Service Levels for Each Device

Addressable Control System



ATX Specifications

ATX Input
Data Rates
110, 150, 300, 600, 1200, 2400, 4800, 9600 baud

ACU Interface
Data Type
Asynchronous, Synchronous
Transmission Mode
Simplex
Electrical Interface
RS-232C

ATX Output
Carrier Frequency
108.2 MHz standard—Other frequencies available by special order
Frequency Accuracy
.0010%

Specifications

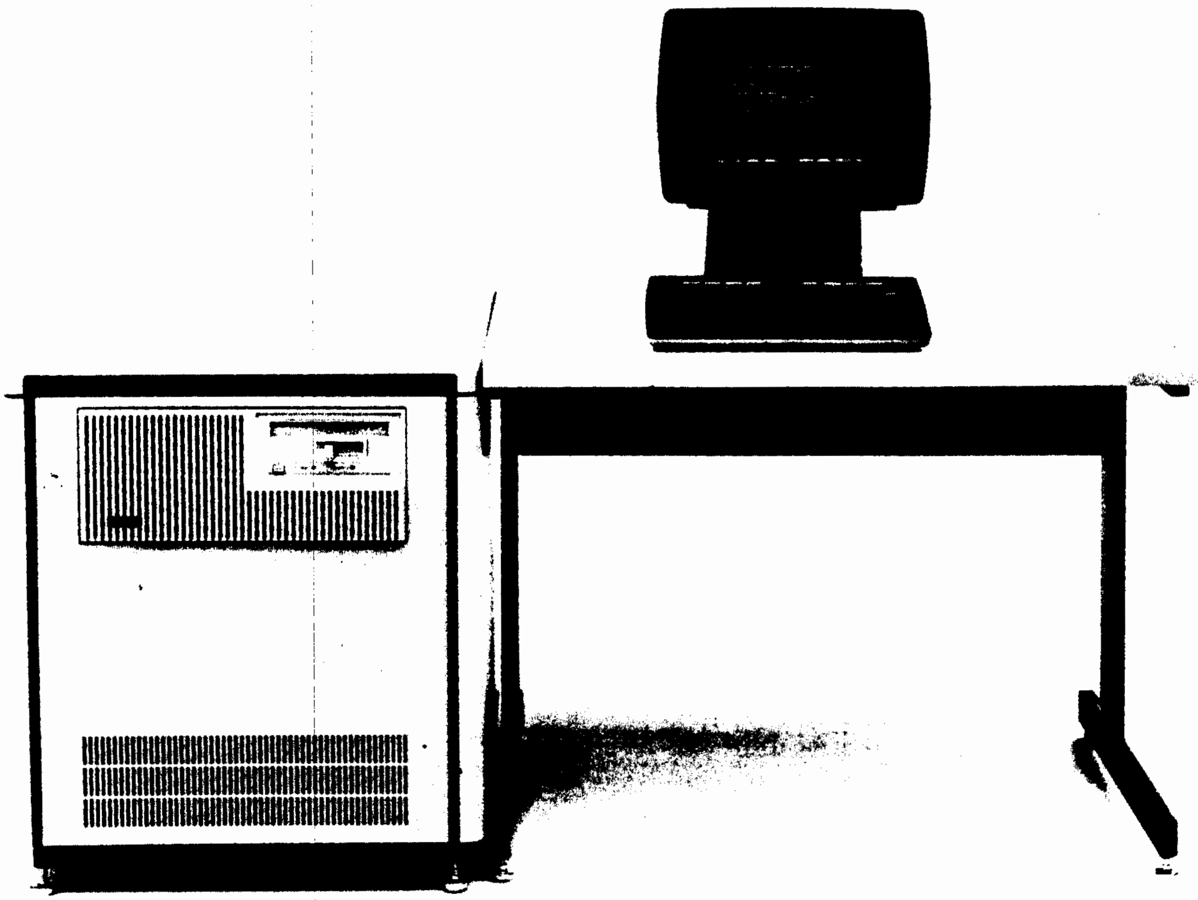
ACU Computer Input
Data Rates
110, 150, 300, 600, 1200, 2400, 4800, 9600 baud
Asynchronous Protocol
Custom
Synchronous Protocol
IBM Bi-sync (via protocol converter)
Electrical Interface
RS-232C

ACU Local CRT Input
Data Rates
110, 150, 300, 600, 1200, 2400, 4800, 9600 baud
Data Type
ASCII Asynchronous
CRT Type
DEC VT-102
Electrical Interface
RS-232C

ACU Output
Data Rates
110, 150, 300, 600, 1200, 2400, 4800, 9600 baud
Data Type
Asynchronous, Synchronous
Transmission Mode
Simplex
Electrical Interface
RS-232C

ACU Power Requirements
Input Voltage
115V ac nominal
Power Consumption
240 watts maximum
Temperature Limits
0°C to 50°C (32°F to 122°C)
Channel Spacing
400 kHz
Signal Level
+25 dBmV to +45 dBmV (adjustable)
Modulation
FSK
Peak Deviation
20 kHz nominal
Data Format
Bi-phase coded
Data Rate
19,200 baud

Power Requirements
Input Voltage
115 volts ac
Power Consumption
50 watts
Temperature Limits
0°C to 50°C (32°F to 122°F)



The System Manager II offers the CATV operator a comprehensive, user friendly computer hardware and software package for managing systems using Scientific-Atlanta 8500 Addressable set-top terminals. The System Manager II features integrated device control/subscriber information software and can perform subscriber management, pay-per-view management, inventory management, report, and test functions. The software uses pre-formatted CRT display screens for efficient keyboard entries.

The System Manager II can operate independently of a host billing computer or it can be linked to a host billing computer. As a stand-alone system, the System Manager II can generate subscriber reports that provide organized information to be entered into a billing computer. When the System Manager II is linked to a host computer, a single entry can be established.

Scientific-Atlanta will assist the cable operator in evaluating his needs for addressable control systems. Extensive customer support includes site preparation, hardware installation, personnel training, fully-documented software programs, and software updates.

System Manager II Components

The central component of the System Manager II is a Hewlett-Packard 1000 A600 disc-based computer. Subscriber disc capacity is determined by the cable operator's data base requirements. The disc storage can be backed up with a magnetic tape cartridge. The Hewlett-Packard 1000 A600 provides multiple-CRT ports for convenience and to help distribute work load effectively. These peripheral devices require a standard RS-232C electrical interface.

An addressable converter system incorporating the System Manager II also includes an Addressable Transmitter (ATX). The ATX, located at the headend accepts control data from the System Manager II computer and transmits the data over the cable system. The unit also generates converter control signals needed to keep all Series 8500 Addressable set-top terminals active. Output data is transmitted as an FSK-modulated digital signal at a data rate of 19,200 baud. It operates at a standard frequency of 108.2 MHz. The System Manager II can serve multiple headends, with one ATX required at each headend.

Two ATX's can be installed in tandem for system redundancy at each site.

The Series 8500 System Manager II provides an innovative format for subscriber account management, pay-per-view management, converter inventory management, test and report functions, and additional system security. The System Manager II data base integrates device control data and subscriber information to present a logical and easy-to-operate, English-language software command structure. To safeguard the system, the System Manager II requires a password and code number for computer access.

The System Manager II computer automatically associates subscriber account numbers with set-top terminal identification codes. Subscriber records can be easily retrieved by any of several access keys, including last name, street address, telephone number, and account number. Individual set-top terminals can be addressed to begin service, change service levels, authorize single channels or tiers, or to disable the set-top unit. Scientific-Atlanta's unique global command format increases the efficiency of its pay-per-view management structure, enabling the CATV operator to run back-to-back pay-per-view programs on the same channel.

The System Manager II provides a flexible pay-per-view structure. An initial computer entry is made to define the date, start time, stop time, and event identification code for each program. Related programs can then be combined into tickets. A ticket is a marketable combination of program events which can be ordered individually or as a package.

Subscriber orders can be taken any time after events and tickets are defined, easing the problem of last minute order entry rushes. All events and tickets are displayed on a CRT screen. To complete an order entry transaction, the clerk has only to enter a single character on the pre-formatted screen. The task of repeatedly entering lengthy coded information for each order is eliminated, reducing the incidence of clerical error.

At the beginning and end of a pay event, the set-top terminals of all subscribers who have ordered the event are authorized or deauthorized simultaneously by an automatic global command. This is controlled by a real-time clock in the system. The global command can be conducted manually if desired. A global command can also be issued system-wide to authorize all set-top terminals to receive a channel temporarily for a pay-per-view preview or other promotional program.

The System Manager II provides inventory management functions to track set-top terminals starting from the time they are received. The serial numbers and addresses of all new set-top units are transferred to the inventory files from tape cartridges provided by the factory. The cable operator has the option to define location and status codes which give him appropriate information for his system. Examples of "locations" are: receiving dock, warehouse, with installer, or at a particular subscriber's residence. "Status" examples are: in stock, in service, being tested, or lost.

The System Manager II provides the capability to test individual converters. Testing can be performed when the converters are received from the manufacturer or when the converters are installed in a subscriber's residence. This function contains a feature which allows the cable operator to define specific test sequences.

The System Manager II can generate extensive printed reports of subscriber, converter, and pay-per-view information. These reports provide a valuable tool for efficiently operating an addressable converter system, monitoring and tracking converter resources, and marketing addressable services. The cable operator can select from a list of standard System Manager II reports, or he can easily define special reports to meet his own unique requirements.

The System Manager II and ATX software provides a legal terminal test and a refresh timer signal for additional system security. These two security features are designed to discourage theft of terminals and migration of set-top units from one CATV system to another, to help control delinquent accounts, and to thwart attempts to circumvent deauthorization instructions from the headend.

The legal terminal test checks to see that only converters recorded in the ACU or System Manager II data base are operational. This test is conducted system-wide using a global command. If a set-top unit is installed in the cable system but is not included in the test of valid terminals, it is automatically shut down.

The refresh timer signal is an on-going function of the Addressable Transmitter. The ATX continually transmits an encrypted message which the set-top terminal must decode in order to stay active. Any attempt to interrupt communications between the headend and the converter prevents the converter from receiving the refresh signal, and the unit will be disabled. The refresh signal can be set from the System Manager II for various time intervals ranging from one-half hour to sixty-four hours.

Additional internal security is provided by a printed transaction summary which lists all changes in the data base and indicates which operators made the changes. The cable operator can trace illegal company entries and identify who might be tampering with the system.

Features

- Uses Hewlett-Packard 1000 A600 Computer
- Software Automatically Integrates Subscriber Information and Device Control Data
- Can be Interfaced with a Host Billing Computer
- Software uses Pre-Formatted CRT Display Screen Format
- Multiple-CRT Ports
- Efficient Structure for Pay-Per-View Management
- Inventory Control
- Test and Report Capability
- Added System Security
- Extensive Customer Support

Specifications

AMS Hardware

HP 1000 A600 Computer with 512 kbps memory
28 Mbps or 64 Mbps Winchester disc with integral tape backup unit
720 (28.35 in.) cabinet with 16 I/O slots

Electrical

Power Input
Line Voltage
108V to 126V ac
Line Frequency
58 Hz to 62 Hz
Power Requirements
Minimum of dual 20A, grounded wall-mounted receptacle
Power Consumption
1500 watts maximum

Physical

Temperature
10°C to 40°C (50°F to 104°F)
Relative Humidity
20% to 80%, non-condensing
Dimensions
718.8H x 635W x 812.8D mm (28.3H x 25W x 32D in.)
Weight
163.6 kg (360 lb)

Printer

Dot-Matrix Serial Impact Printer
180 characters per second
68/136/227 characters per line
Number of Copies
One original and up to five copies with carbon-print paper

Console

Interactive Character-Mode CRT Terminal
960 characters per second
80 characters x 24 lines per page

Terminals

Display CRT Terminal
960 characters per second
80 characters x 24 lines per page

Addressable Transmitter

ATX Input

Data Rates
110, 150, 300, 600, 1200, 2400, 4800, 9600 baud

AMS Interface

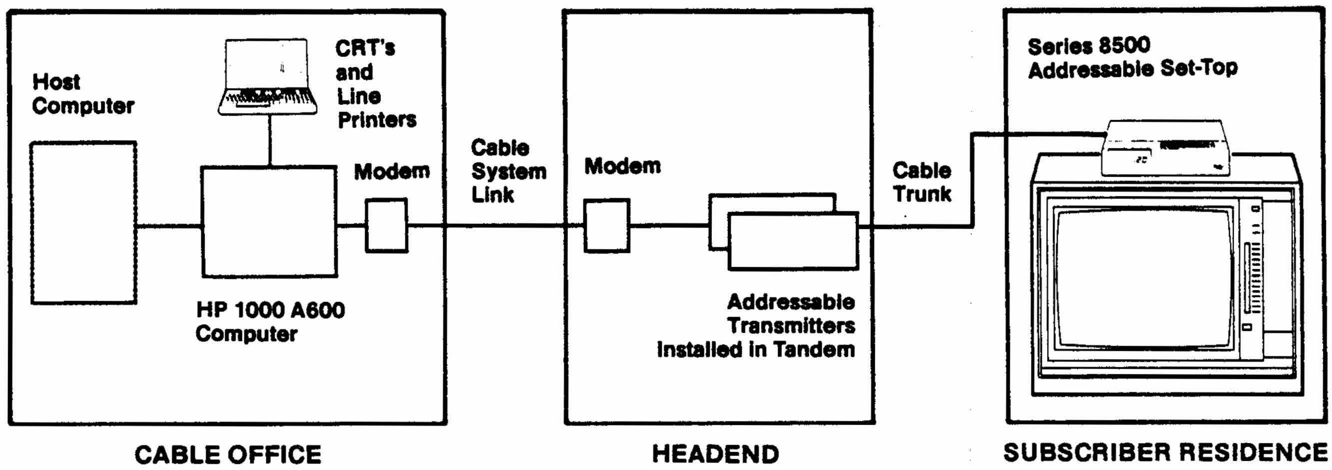
Data Type
Asynchronous, Synchronous
Transmission Mode
Half Duplex
Protocol
Custom

ATX Output

Carrier Frequency
108 MHz standard. Other frequencies available by special-order
Frequency Accuracy
0.0010%
Channel Spacing
400 kHz
Signal Level
+25 dBmV to +45 dBmV (adjustable)
Modulation
FSK
Peak Deviation
20 kHz nominal
Data Format
Bi-phase coded
Data Rate
19,200 baud

Power Requirements

Input Voltage
115V ac nominal
Power Consumption
50 watts maximum
Temperature Limits
0°C to 50°C (32°F to 122°F)



CRT's and Line Printers
Host Computer
Modem
HP 1000 A600 Computer
Cable System Link
Modem
Addressable Transmitters Installed in Tandem
Cable Trunk
Series 8500 Addressable Set-Top
CABLE OFFICE
HEADEND
SUBSCRIBER RESIDENCE

Series 9000 Antennas for C- and Ku-Band Operation

Robert Fitzgerald

Applications

The Series 9000 Antenna is designed to be a high-performance, low-cost antenna for audio, video and digital receive-only applications. The product is specifically designed to meet the needs of Digital Audio Broadcasting, mini-cable systems, teleconferencing, and the private earth-station owner.

Within the contiguous United States, a carrier-to-noise (C/N) ratio of greater than 10.5 dB and a signal-to-noise (S/N) ratio of 48.5 dB is easily achievable for video receive-only applications.

Introduction

A major goal of the design of the Series 9000 antenna was to meet both the current and future needs of the Satellite Communications industry. To achieve this objective, the antenna was designed in a modular fashion which allows a customer to configure an antenna to meet the specific needs of the application. In addition, retrofit kits will be available to allow modification of the antenna in the field as these requirements change. All modules are designed to meet possible future needs such as the sidelobe requirements for 2° satellite spacing. Below are some examples of the flexibility available with the Series 9000 antenna.

- 2.8-meter reflector with EL/AZ mount
- 3.2-meter reflector with EL/AZ mount
- Retrofit kit to upgrade 2.8-meter to 3.2-meter reflector
- C-Band or Ku-Band operation
- Single or dual polarization
- Polar mount with motorized option
- Electronic polarization option

The Series 9000 antenna is truly the antenna of today and tomorrow.

Antenna Design and Construction

Another major goal of the design was to produce a very high-quality, low-cost product. To achieve this objective, a major investment in high-volume production tooling was required. This antenna is built using steel stampings, aluminum extrusions, plastic injection molds, aluminum die and investment castings, and steel roll-formed parts. The design of each part was functionally optimized in conjunction with the constraints of the production process, resulting in high-quality, low-cost parts.

The Series 9000 Antenna is a commercial-grade product designed to meet AISI, AISC and RS-222-C specifications, thereby meeting or exceeding building code requirements.

The Series 9000 antenna has four main subsystems, namely:

1. Feed

2. Reflector

3. Mount

4. Base

Also available is a panel extending kit which extends the reflector diameter from 2.8 meters to 3.2 meters.

Feed

The feed subsystem for single polarization reception consists of four feed supports, a feed housing, a prime focus feed, a cover, and a rod and gear for polarization rotation (see Figure 1). An optional dual-polarized feed provides simultaneous reception from both polarization transmissions.

A unique feature of this subsystem is the polarization rotation mechanism. The feed can be manually rotated by an operator standing alongside the dish via the turn of a screwdriver. This feature simplifies the polarization rotation process, as there is no need to climb into the dish or to obtain a stepladder to reach the feed. Also, the operator is kept out of the path of the satellite signal and, therefore, it is easier to peak the signal.

There is also a Ku-band feed available for the American and European markets. In the near future, an electronic polarization rotator with a position controller, will be offered for remote polarization rotation.

Reflector

The reflector subsystem consists of 8 panels, 8 braces, and a hub (see Figure 2). The surface tolerance of the panels allows for either C- or Ku-Band reception with a 0.025 inch rms maximum. The reflector panels and the hub are steel stampings manufactured using techniques originally developed by the automotive industry. These techniques were enhanced to provide very close surface tolerances for the panels.

Galvanized steel is used to provide corrosion protection and, in addition, a special, highly protective electroplate primer is applied. A finish coat of high-quality industrial paint is then applied to ensure a long-life protective finish.

Mount

The first mount in the Series 9000 line to be produced is the EL/AZ mount. This mount consists of two pivot axis brackets, a beam, four beam supports, a cone, an actuator, actuator attachment brackets, and a clamp (see Figure 2).

To point the antenna at a satellite, one simply sets the elevation angle as specified in a table based on antenna site location and the satellite to be viewed. He then rotates the antenna about the azimuth axis until a signal is received.

In the near future a polar mount will be available which allows a customer to set up the antenna one time, and then, by adjusting one actuator, go from satellite to satellite. This one time setup is very simple because all adjustment axes are independent of each other. Therefore, adjusting one axis does not misalign another axis as is the case for most polar mounts available today. Both the EL/AZ mount and polar mount can be installed without regard for foundation heading.

Base

There are two different styles of bases available--in-ground and surface. The difference between an in-ground base and a surface base is the method of interface with the concrete foundation. The in-ground base is poured into the concrete and becomes a part of the foundation whereas the surface base has a flange with bolt holes on the bottom which allows the base to be bolted to eight anchor bolts which are embedded in the concrete.

The flanged surface base is flexible in that the base can be removed from the foundation if it becomes desirable to relocate the antenna. The flanged base is easier to install than the in-ground base, whereas, the in-ground base has the advantage of lower cost.

Extending Panels

The Extending Panel Kit consists of 16 extending panels, 16 extending panel braces, and 8 panel braces which are identical to the braces used for the 2.8-meter configuration. The extending panels are steel stampings manufactured using the same techniques as used on the main panels. The surface tolerance for 3.2-meter systems is 0.040 inches rms maximum.

Installation

The Series 9000 Antenna was designed to minimize the time required for assembly. A special effort was made to eliminate parts that are different but look similar, to make all parts easily identifiable by their shape. Also, wrench clearances were designed into each part to obviate the use of special tools. The antenna is designed to be assembled by two people with a small stepladder without the need for a crane. The antenna system assembles in a building block fashion with one part following the next until the entire assembly is erected. Even the reflector is assembled on top of the mount one section (petal) at a time.

Specifications

The table below lists the main specifications for the Series 9000 Antenna with an EL/AZ mount.

Table 1. Series 9000 Antenna Specifications

Characteristic	Specification 2.8 Meter	 3.2 Meter
	General	
Surface Tolerance	0.025° rms Max	0.040° rms Max
Pointing Accuracy	0.12° rms @ 45 mi/h with gusts to 60 mi/h	0.25° rms at 45 mi/h with gusts to 60 mi/h
Feed Type	Prime Focus	Prime Focus
Wind Loading		
Operational (mi/h)	60	60
Survival (mi/h), No Ice	100	100
Operational Temperature Range	-40°F to +149°F	-40°F to +149°F
Shipping Weight	350 lbs	440 lbs
Azimuth Range	360°	360°
Elevation Range	5° to 60°	5° to 60°

Table 1. Series 9000 Antenna Specifications (continued)

Characteristic	Specification 2.8 Meter	Specification 3.2 Meter
	Electrical	
Operating Frequency	3.7 to 4.2 GHz	3.7 to 4.2 GHz
Gain (Midband at OMT Port)	39.5 dBi	41.0 dBi
VSWR (Max)	1.3:1	1.3:1
Polarization	Linear	Linear
First Sidelobe Level	-20 dB	-20 dB
Cross Pol Discrimination (on Axis)	35 dB	35 dB
Beamwidth	1.9°	1.7°

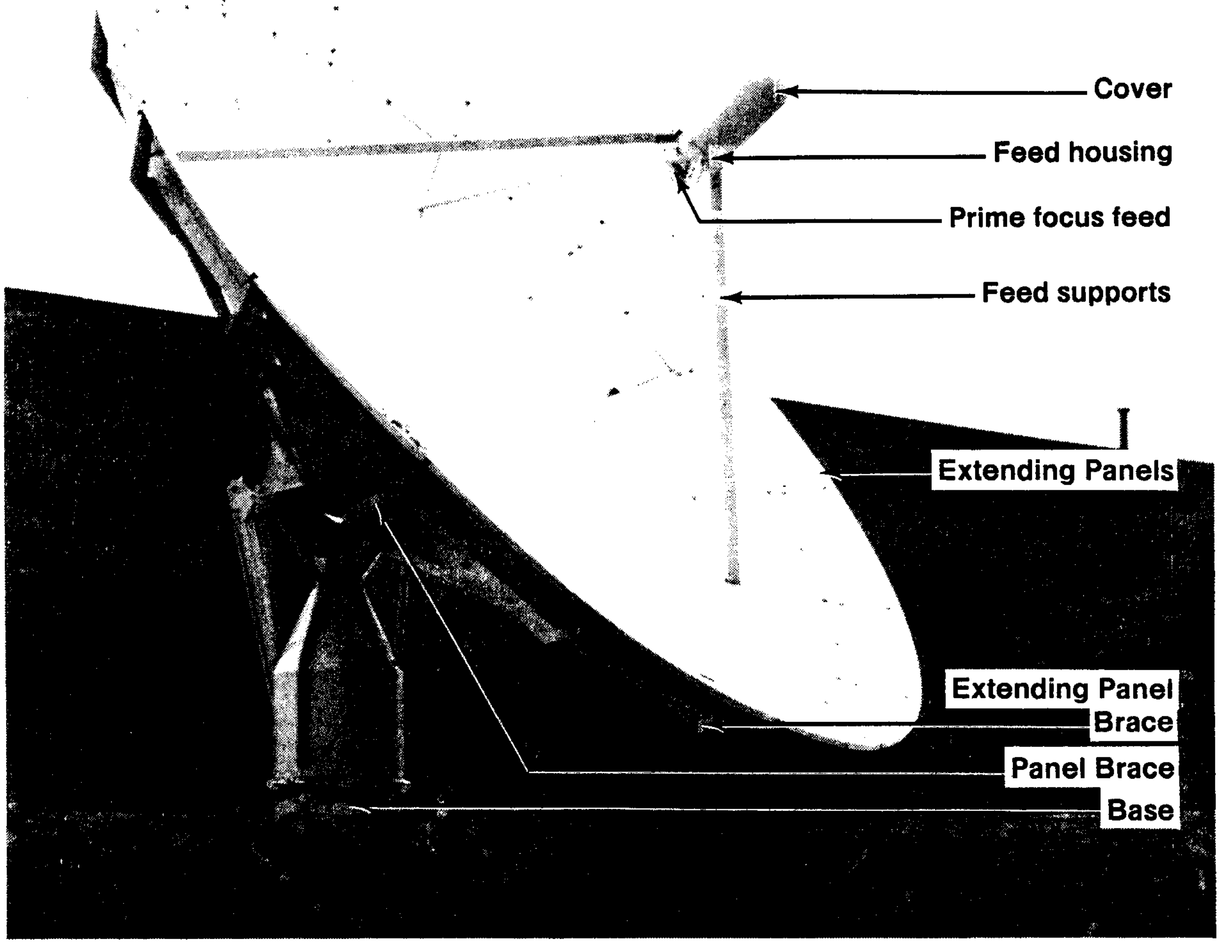
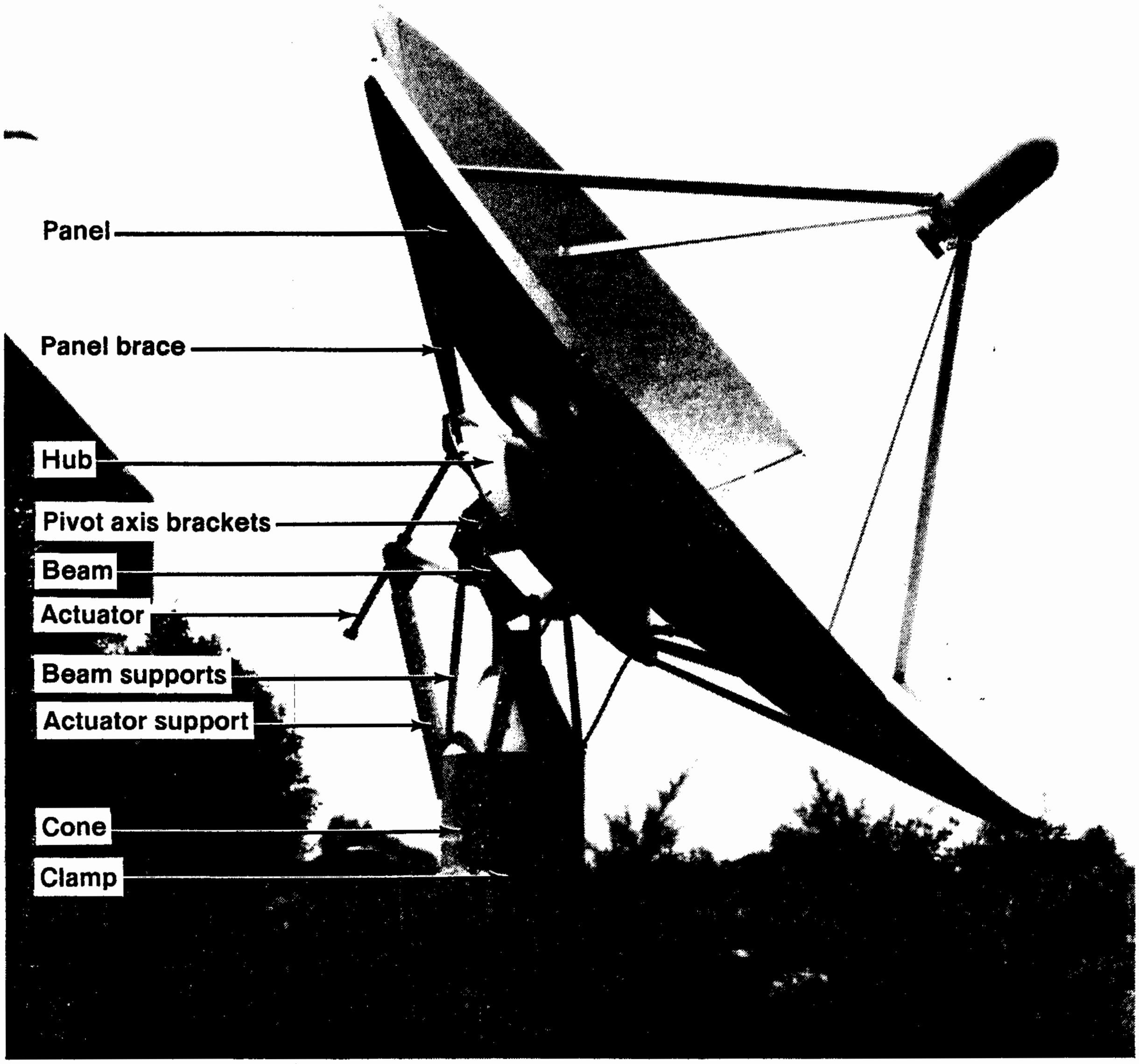


Figure 1. Feed Subsystem



Panel
Panel brace
Hub
Pivot axis brackets
Beam
Actuator
Beam supports
Actuator support
Cone
Clamp

Figure 2.

Small Earth Station Antennas for C-Band Operation

G. C. Browning

Introduction

This paper addresses the subject of earth station antennas having a diameter of 5 meters or less. It includes descriptions of Scientific-Atlanta 4.6- and 5-meter receive-transmit systems.

General Design Considerations

Small antennas of various sizes have distinct cost, electrical, and structural advantages. These distinctions should be investigated thoroughly before a choice is made of one antenna size over another. All Scientific-Atlanta antennas are designed with certain objectives in mind. These include high electrical performance, structural integrity, quick and easy assembly without field adjustment, interchangeable parts, reduced shipping volume, minimum maintenance and low cost.

Electrical performance and proper structural strength and rigidity in antennas is achieved by extensive analysis and testing. These subjects are reviewed in other papers and, therefore, will not be discussed at length here. The other common design features are very important, particularly in small antennas, and warrant further discussion.

Shipping volume is important because, if the shipping container volume is kept small, shipping costs are reduced and the antenna can be transported to virtually any possible site. For example, small containers can be loaded in freight elevators for rooftop mounting or loaded in aircraft for remote site locations.

The shipping container size dictates that the main antenna assemblies, such as, the reflector, mount, etc., be designed for assembly from smaller component parts. As a result all component parts of Scientific-Atlanta's small antennas can be carried by one person, and the entire antenna can be assembled readily by three people with no special equipment and with no special training or skills. All production parts are fabricated with specially-designed hard tooling which guarantees interchangeable parts.

Reflector

The reflector is made up of a reflector surface, a backing structure and a hub. The reflector surface is developed from assembled panels, and the backing structure which supports the panels is composed of various structural members. On the 4.6- and 5-meter antennas, the hub is a circular or rectangular enclosure to which the panels and backing structure attach.

The number of panels making up the reflector surface varies from 12 to 24 depending on the antenna size. Each panel is formed by matched metal-stamping dies using fully annealed aluminum sheet, ranging from 0.090 to 0.125 inches thick. These sheets are sheared and notched in a precision blanking die and then formed to a paraboloidal contour in a large press. To ensure the consistency of the stamped panels during the stamping process, sample panels are periodically placed in a checking fixture to inspect the panel contour and side flange angles. After the panels are stamped, they are located in a drill fixture that accurately drills the side flange and hub mounting holes. The panels are dipped in an inhibitor (a chromate conversion) for corrosion protection and good paint adhesion. Finally, they are sprayed with a special high-reflectance paint that disperses infrared and optical radiation to protect the feed while leaving the RF energy unaffected. In addition, the paint is formulated to protect the panels and give the antenna extended life.

All of the antenna parts including the backing structure are made with precision tooling, including drill and welding fixtures, punching and shearing dies and forming dies.

As a final check to ensure that all reflector parts are made to the design specifications, random reflector panels and backing structure parts are periodically assembled and surface rms tests are performed to verify that the surface tolerance is within specification. These rms tests are made with a proprietary computer-controlled automatic surface measurement system (ASMS) developed at Scientific-Atlanta.

Mount

In any antenna system, the function of the mount is twofold; it must safely transfer all reflector loads to the foundation and permit the reflector to be pointed toward a desired location in space. These two functions form the basis for Scientific-Atlanta mount designs, although other considerations such as low cost and ease of pointing are also important design objectives.

The 4.6-meter and 5-meter mounts are of the elevation-over-azimuth type. These mounts are designed to rigidly support a reflector at specified wind loads with low pointing errors.

The 5-meter elevation-over-azimuth mount is easy to point since its azimuth angles are directly read from local true north and its elevation angles are read from the local horizon. One person can readily move the antenna in azimuth and elevation to direct the beam from one satellite to another.

The 4.6-meter elevation-over-azimuth mount is engineered to provide continuous satellite arc coverage from any location in the contiguous United States. Pointing the antenna is rapid and accurate. Complete 360° azimuth coverage does not require alignment of the foundation to a specific heading, eliminating the possibility of installation errors associated with foundation centerlines.

4.6-Meter Earth Station Antenna

The 4.6-meter antenna is designed for receive or transmit applications in the 4- to 6-GHz range serving video, audio or data circuits from any satellite located above 5° elevation. Like other small Scientific-Atlanta antennas, it can be easily installed in a wide variety of locations with a minimum of site preparation.

Figure 1. Typical Model 8346 4.6-Meter Earth Station Antenna Installation

Reflector

The 4.6-meter reflector surface is made up of twelve separate 5052-0 aluminum panels which are 0.090 inches thick. The panels bolt to each other and to a common hub through precisely located side-flange and hub holes.

The backing structure that supports the reflector is twelve steel tube struts which attach from the hub to panels. The struts are formed and punched with hard tooling designed to assure consistent surface tolerances and interchangeable parts. The struts and hub are hot-dipped galvanized for maximum protection against corrosion. The design transmits reflector wind loads directly to the hub with negligible bending moments, resulting in a strong, low-compliance antenna structure.

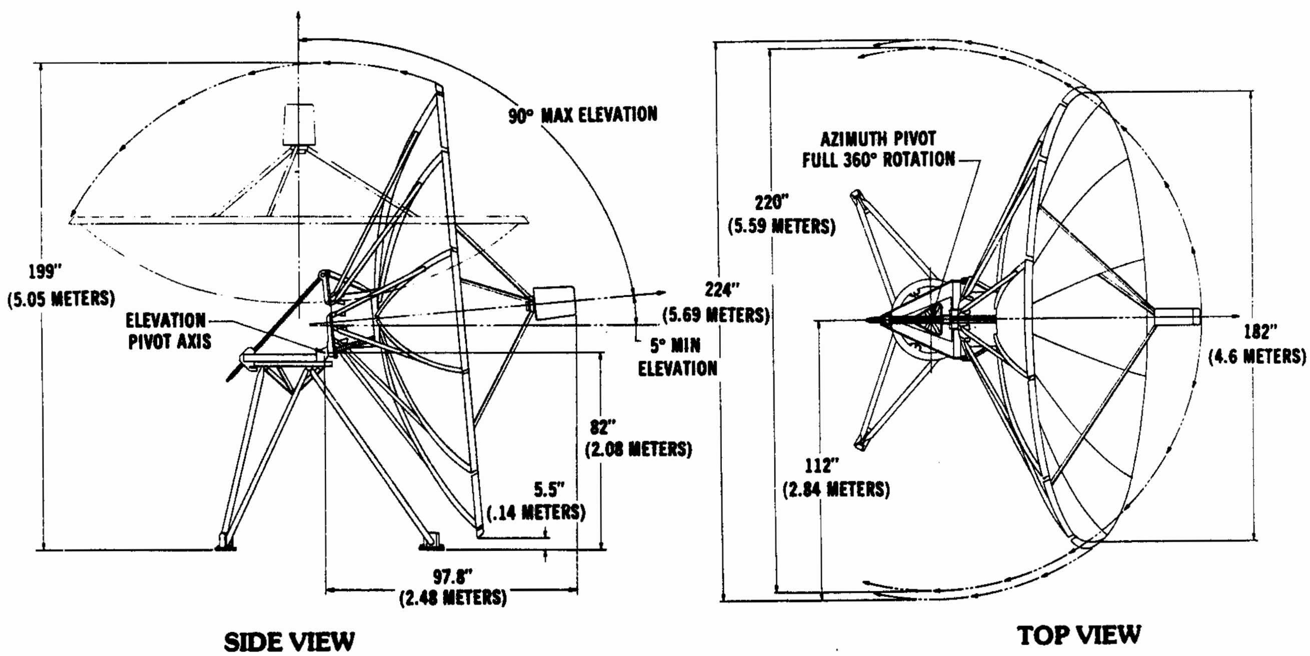


Figure 2. 4.6-Meter Reflector Outline Drawing

Mount

The 4.6-meter elevation-over-azimuth mount is made up of an azimuth ring which is supported by six legs and three cast feet. The feet are for use when the antenna is mounted on a monolithic slab. Atop the azimuth ring sits an A-frame which is the pivoting member in the azimuth direction and also the attaching point of the elevation screw.

The mount is designed so that no one part weighs in excess of 100 pounds (45 kg). After a foundation has been prepared, two men can install the antenna in less than one day.

All members of the mount are hot-dipped galvanized with the exception of the actuator screw and pins which are zinc-electroplated.

Feed System

The 4.6-meter antenna uses a standard economical prime focus Feed or an optional Cassegrain system where higher gain is required.

Both the feed horn in the prime focus system and the subreflector in the Cassegrain system are accurately related to the main reflector by three support struts.

The prime focus system meets the 29-25 log θ sidelobe specification for reduced satellite spacing.

The Cassegrain subreflector is a shaped quasi-hyperboloid which is formed on spinning equipment using a contoured template that maintains close surface tolerances. Each subreflector contour must conform to a checking template to within certain minimum tolerances.

Polarization adjustment is easily accomplished in the receive-only prime focus system and is 360° continuous.

The subreflector is accurately located relative to the main reflector by three spars that attach at three separate points on the reflector and at a common junction connection behind the subreflector. The subreflector is fastened to each spar through a support stud.

There is also a dual feed available for the 4.6-meter and 5-meter antennas.

Model 8008 5.0-Meter Earth Station Antenna

The 5-meter antenna is designed to provide receive or transmit capabilities in the 4/6-GHz range for applications using geosynchronous satellites. It is a high-performance antenna that offers an increase of 1 dB in gain over that of the 4.6-meter antenna. A shrouded version, Model 8008LS, is offered which provides low sidelobes and is recommended for sites experiencing heavy terrestrial microwave congestion. Figures 5 and 6 show the two versions of the 5-meter antenna.

Reflector

The 5-meter reflector employs a 24-panel, bolt-together design which minimizes transportation and installation problems. The panels are fabricated utilizing a cost-effective die-stamping technique that ensures part interchangeability. The panel blanks are made of 0.125-inch, 5052-0 aluminum-alloy sheet. Integral radial and rim flanges are formed when the panels are stamped. These are drilled in a tooling fixture to provide precise registration and structural integrity when the panels are assembled. The 24 panels form a paraboloid of revolution with an f/D of 0.375. A machined central hub provides an interface for each panel that precisely locates them radially with respect to the vertex. The hub is sufficient in size to house electronic equipment and is enclosed to provide protection from the environment.

Mount

The 5-meter mount features an elevation-over-azimuth configuration. A tripod design, utilizing the front leg as a kingpost for azimuth travel, provides independent adjustment of elevation and azimuth. Course and fine adjustment capability is built into each mechanism, making pointing and peaking easy. The standard mount provides continuous elevation coverage from 15° to 60°. Optional elevation adjustment mechanisms make it possible to achieve a maximum of 0° to 70° total elevation travel in two sectors. The two rear corners of the mount serve as attachment points for the azimuth adjustment mechanisms, providing full arc coverage (70° to 135°W) in two sectors. Since the azimuth axis is vertical, there is no weight imbalance that would cause rotation. Sector change-over and azimuth adjustments are easily accomplished.

Figure 4. Typical Model 8008 5.0 Meter Earth Station Antenna

The antenna loads are transferred by the mount to a three-point foundation that forms a 96-inch equilateral triangle. The mount utilizes structural steel and aluminum parts. For corrosion protection the steel parts are hot-dip galvanized, and the aluminum parts are iridited, primed with yellow zinc-chromate primer and painted.

Feed System

As with the 4.6-meter antenna, the 5-meter is available with a standard low-cost prime focus Feed or with an optional Cassegrain system.

The prime focus Feed offers unusual economy in a mid-sized earth station antenna while meeting the 29-25 log θ sidelobe specification for reduced satellite spacing. Typical first sidelobes are 22 dB below the beam peak.

Several options are available in the high-gain, dual-reflector Cassegrain feed system. The dual-frequency band feed is designed to operate in the 3.7 to 4.2 GHz receive band and the 5.925 to 6.425 GHz transmit band. A diagonal horn illuminates a shaped subreflector which provides efficient illumination of the 5-meter reflector aperture.

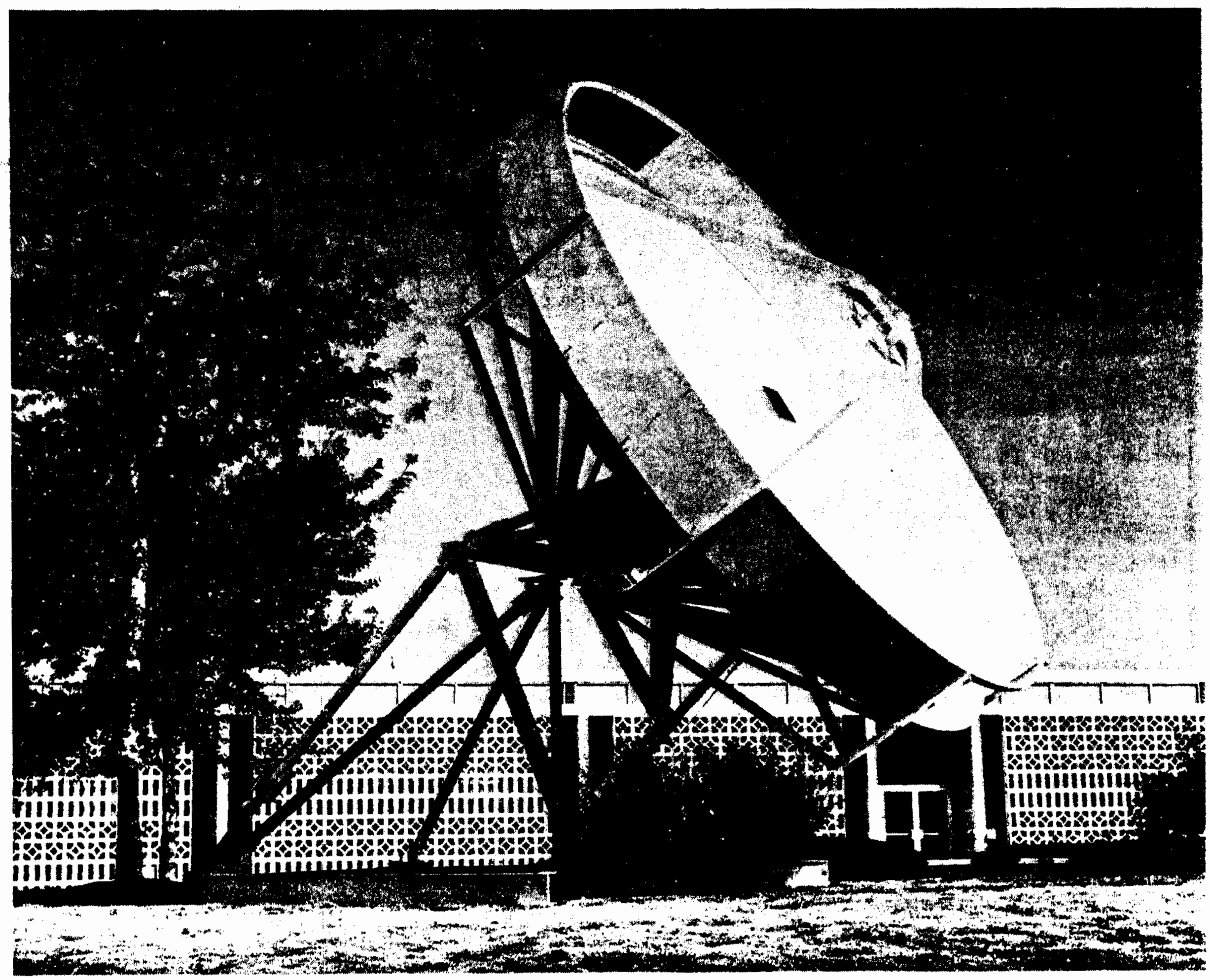


Figure 5. Typical Model 8008LS 5.0 Meter Earth Station Antenna

An orthomode transducer provides the WR229 and WR137 waveguide feed outputs for the receive and transmit bands, respectively. A waveguide transition with WR229 waveguide output is used for the single-polarized receive-only feeds. A dual-polarized orthomode coupler with two WR229 waveguide outputs can be provided for simultaneous orthogonal receive requirements.

The subreflector is structurally supported from the reflector by three equally-spaced spars which form a tripod. The location of the subreflector relative to the supporting structure is preset. All parts are made of aluminum and are painted to provide protection and a pleasing appearance.

Antenna Specifications

The specifications for the 4.6-meter and 5-meter antennas are provided below in tabulated form for easy cross reference for comparing the antennas.

Characteristic	4.6 Meter	5.0 Meter
Feed System	Prime Focus [Cassegrain]	Prime Focus [Cassegrain]
Mount Configuration	EL over AZ	EL over AZ
Arc Coverage (Std.)	70° to 140°W	70° to 135°W
Operating Frequency Range Receive (GHz) [Transmit (GHz)]	 3.7 to 4.2 [5.925 to 6.425]	 3.7 to 4.2 [5.925 to 6.425]
Antenna Gain (dB) 3.95 GHz	43.0 [43.5]	44.1 [44.5]
VSWR	1.3:1	1.3:1
Polarization	Linear, 360°	Linear, 360°
Wind Loading Operational (mi/h) Survival (mi/h), No Ice	 60 100	 60 125
Temperature Range Operational Survival (No Wind)	 -20°C to 55°C -35°C to 65°C	 -20°C to 55°C -35°C to 55°C
Pointing Accuracy(60 mi/h Wind)	.20° rms	.11° rms
Shipping Weight (lbs)	2,300	2,500
Net Weight (lbs)	1,200	1,500
Shipping Volume (ft^3)	200	220

[ ] Indicates Cassegrain System

7-Meter Earth Station Antenna for C-Band Operation Model 8010C

Thomas A. Akin and Carlton T. Ethridge

Introduction

The Scientific-Atlanta Model 8010C 7-Meter Earth Station Antenna has been designed for the TV broadcast industry's current and projected future satellite TV relay service. Under average conditions using conservatively rated components, it will provide broadcast quality video reception from U.S. domestic satellites located anywhere within the FCC recommended arc. Its radiation distribution envelope meets FCC requirements for transmitting to the satellite as well as receiving. The coverage is continuous, and provides for periodic minor adjustment for satellite drift. The Scientific-Atlanta 8010C 7-meter Earth Terminal combines broadcast quality and broadcast reliability in a design which should well serve the TV broadcast industry.

Scientific-Atlanta 7-Meter Antenna, Model 8010

The Scientific-Atlanta Model 8010CM 7-Meter Antenna is shown in Figures 1 and 2. The antenna is a dual-shaped Cassegrain-type antenna with a high-efficiency corrugated conical horn feed illuminating the subreflector. The optics of the antenna are judiciously chosen to give a high efficiency operation and low sidelobe performance.

The reflector consists of eighteen match-metal die-stamped aluminum panels. The panels start as sheets of 1/8-inch-thick 5052-0 fully annealed aluminum. These sheets are sheared to the correct blank size and then notched in a 100-ton press. The blanks are stamped in an 800-ton press, which simultaneously forms the compound-curved surface, the two side flanges and the end flange. The stamped panel is placed in a combination checking and drilling fixture where an inspector verifies that the panel meets the surface and edge tolerances. Then it is drilled using drill bushings in the fixture as guides. This process ensures that all panels are alike within the tolerance established and that they can be assembled in any random sequence into a reflector. The use of eighteen panels was selected as a compromise between minimizing part size for handling ease and maximum part size to minimize cost and tolerance build-up of any non-random errors. This choice was also based on satisfying the 120° symmetry required for the three-spar subreflector support system.

The panels are supported by 18 aluminum struts connected between the panel flange joints and the rear end of the hub. These struts are made from special extrusions to provide effective structural support for the panels while being very easy to assemble into the antenna structure. The struts are anodized for protection and appearance; the hard surface ensures that the parts will not be marred in handling and assembly, and require no maintenance.

The antenna hub consists of two machined cast aluminum end pieces welded to either end of a cylindrical shell. The shell is 1/8-inch aluminum sheet, rolled to 37-inch diameter and seam welded. The hub provides a suitable housing for RF components and a motorized feed rotator when the antenna is so equipped.

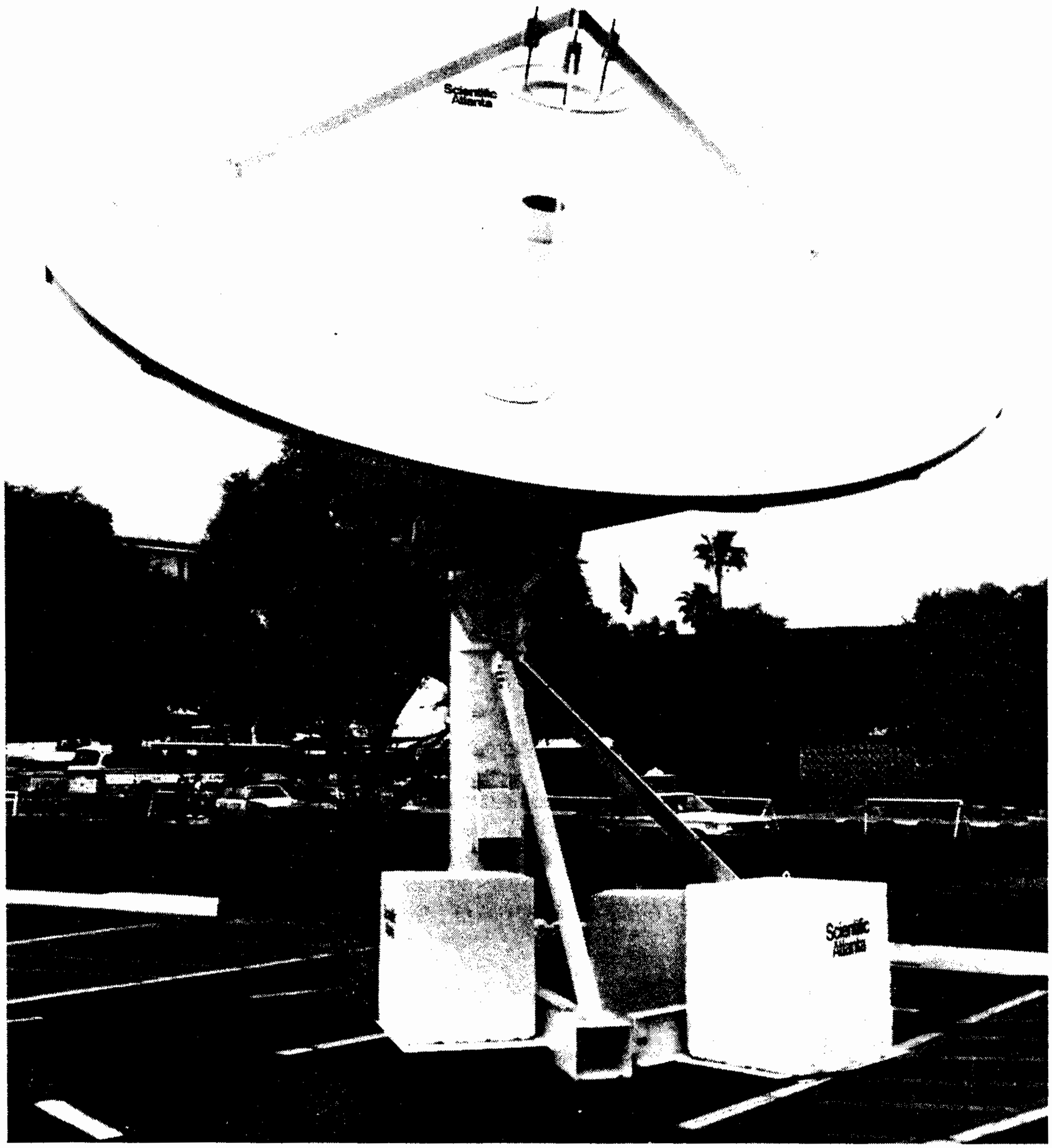


Figure 1. Model 8010C 7-Meter Earth Station Antenna

The hub is designed with covered access openings in the side and rear end. The rear cover provides suitable holes for passage of waveguide, RF and electrical cables. This arrangement permits field assembly without the necessity for drilling holes. The side opening is designed for easy access to the hub for passage of components and for assembly work. This access is especially useful during antenna assembly, because it permits placement of

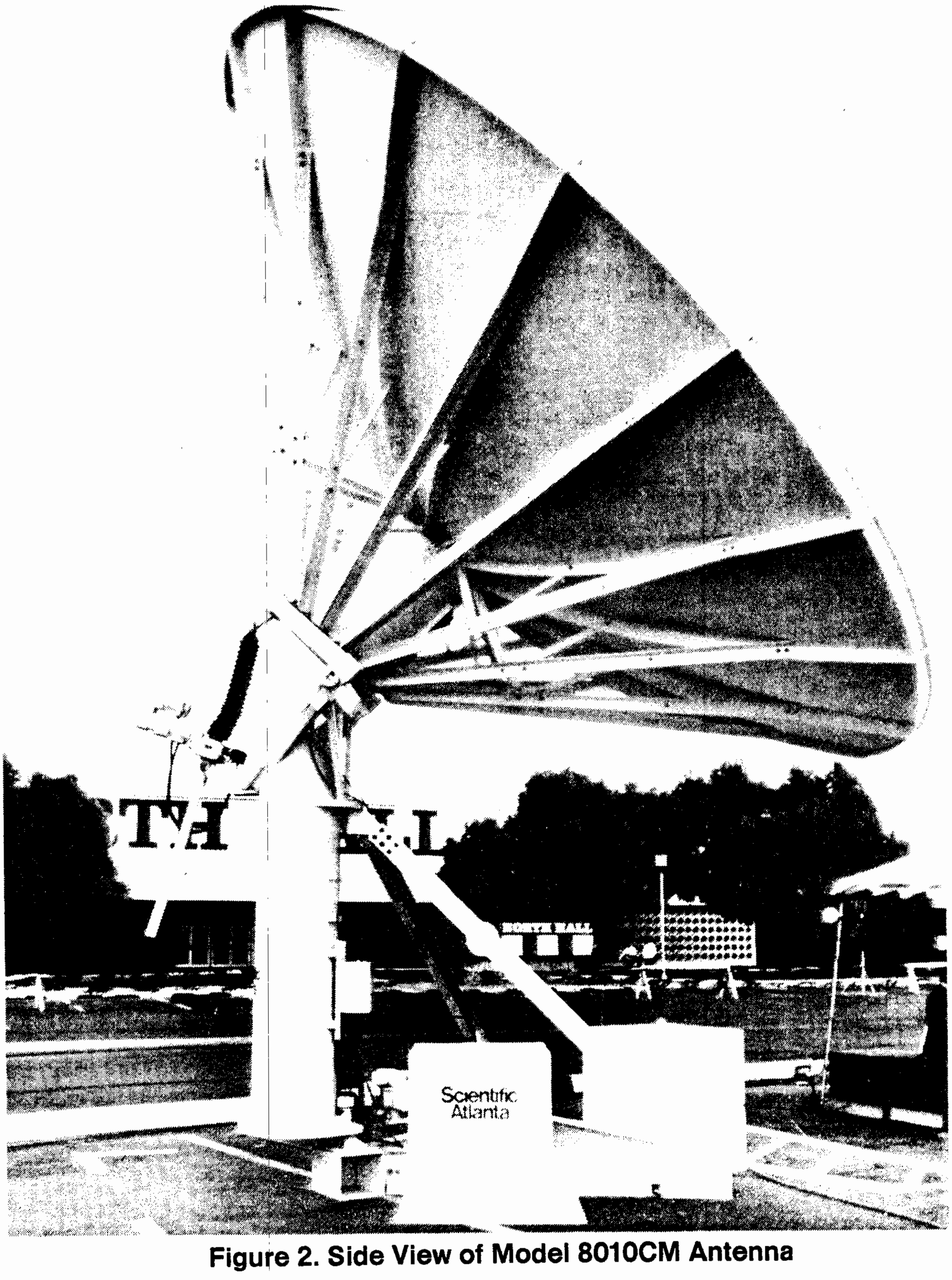


Figure 2. Side View of Model 8010CM Antenna

the hub at the vertical (zenith) position as required for assembly of the reflector, while still allowing personnel access through the hub. This eliminates the necessity for climbing over the edge of the reflector during assembly. After completion of the installation, the side opening provides easy access for servicing the components within the hub.

The hub is produced using precision N/C machinery and specially-designed tooling, ensuring accuracy of manufacture and complete interchangeability.

The feed assembly is mounted on the forward end of the hub assembly.

The feed horn mounting ring is machined with respect to the aperture and rear waveguide of the horn so that the ring seating surface is perpendicular to the RF axis of the horn. The feed is supported by a series of bearings in the front hub ring casting to allow polarization rotation of the feed about the fixed RF axis. The precision extruded square section waveguide of the horn supports orthogonally polarized RF energy.

The feed horn is terminated in one of several configurations. These include a single-polarized receive-only transition, a dual-polarized receive-only orthomode transducer (OMT), a transmit/receive OMT, or a three-port, dual-polarized receive, single-polarized transmit transducer. In operation as a receiving earth terminal, low-noise GaAs FET amplifiers are bolted to the OMT. This entire subsystem is easily rotatable by hand by removing the hub-access cover, loosening the feed-clamp screw inside the hub and turning the assembly. For the motorized version of the antenna, the feed polarization motion is motor driven and can be remotely programmed to turn to any pre-selected position.

The motorized version of the antenna, designated Model 8010CM, features superior versatility and positioning accuracy.

The elevation and azimuth motions are driven by dual-motor drive units which provide high-speed motion for rapid traverse of the arc, and low-speed motion for accurate control of the antenna stopping position. The drive unit consists of a main motor with an integral brake/clutch, a "micro-motor" with an integral brake, and a 10:1 ratio gearbox connecting the two motors. The unit is a self-contained, maintenance-free compact package in which the mechanical functions of braking and clutching occur automatically.

The two drive units and the feed polarization drive system are controlled by an antenna controller consisting of a Model 8840A Remote Control Unit and a Model 8841A Motor Control Unit. With this control system, the high-speed and low-speed motors on each axis, as well as the feed polarization drive motor, can be operated or "jogged" independently, or automatic operation may be selected. In the automatic operation mode, the antenna controller may be commanded to drive the antenna to point to any preselected satellite with a repeatable accuracy of 0.06°. Display resolution is 0.03°.

To achieve this high accuracy of pointing in the automatic mode, the antenna is driven at rapid speed to within 0.3° of the target angle, at which point the slow-speed motor takes over to approach the stopping position. This design greatly reduces position errors normally caused by system dynamics, and results in highly accurate repeatable positioning.

7-Meter EL/AZ Mount

The mount is of the torque tube or king post type. Because of the large vertical tube which is the dominant structural member, the mount is structurally and mechanically very efficient, as is evident from the following list of features, many of which are characteristic of the vertical post design:

a. The entire mount is exceptionally rigid. The pointing error in 45 mph winds gusting to 60 mph is 0.08° rms. This value was established by full-scale load tests in which the worst possible wind direction and pointing angles were simulated.

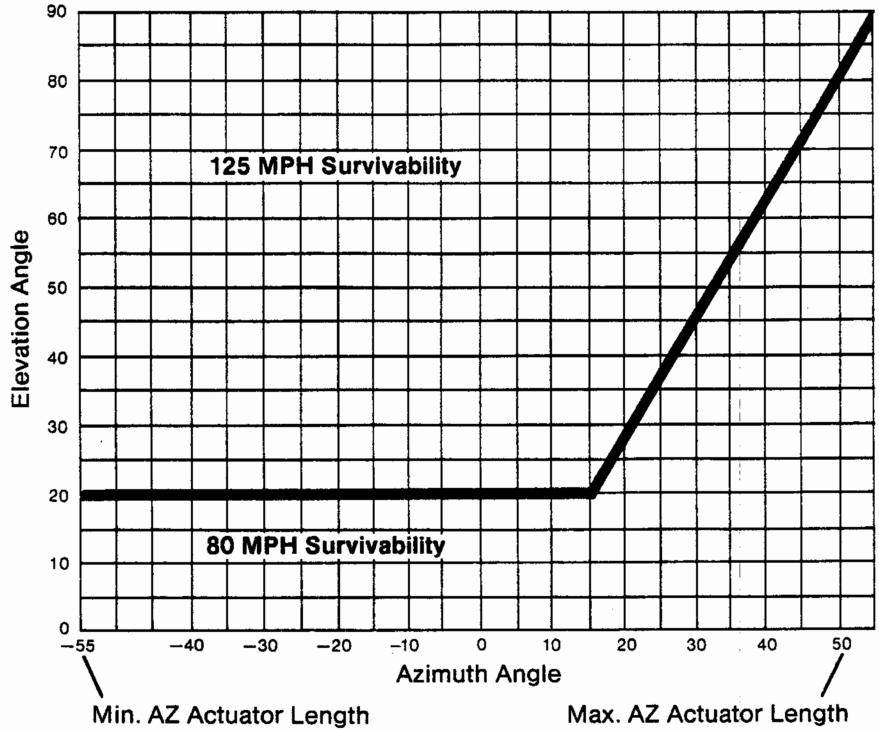
b. The manual (non-motorized) version of the mount will survive winds in excess of 125 mph regardless of antenna pointing angles. This capability was confirmed by actual load tests on the prototype mount. In the motorized version, column load capability of the azimuth ballscrew actuator limits the survival wind to 80 mph with the antenna pointed at any angle (unstowed). However, by driving the antenna a few degrees at most in either elevation or azimuth the antenna can be placed in the "stowed" pointing angle envelope where the survival wind speed is 125 mph. Refer to Figure 3.

c. The mount may be converted from manual to motorized configuration by replacement of actuators. Four pin joints are involved in the conversion.

d. Installation of the mount is not site-sensitive: Foundation heading accuracy is not critical with an elevation-over-azimuth mount, and the 180° azimuth coverage in 110° overlapping sectors affords a wide variation in foundation heading for coverage of a given satellite arc.

e. The pointing procedure is greatly simplified by use of continuously-threaded rods as the positioning struts for elevation and azimuth antenna angles. Hand nuts running on these rods provide fine adjustment and rigid locking at any angular position when tightened on both sides of the pivot trunnion. Angular position indicators at both elevation and azimuth pivot axes provide a ready reference for pointing.

f. The mount installation will tolerate a reasonably unlevel or rough foundation without loss of structural integrity because the mounting feet are "grouted in" to the foundation when installed. The required grout is supplied with the antenna.



To stow the antenna move the mount in azimuth and/or elevation until a configuration within the 125 mph survivability envelope on the chart is achieved.

Figure 3. Stowing Chart

g. Assembly and installation procedures have been simplified greatly. All critical or difficult assembly operations are done prior to shipment, and no holes are required to be drilled in the field. Only 20 fasteners (all standard) are required to completely assemble the mount in the field, exclusive of foundation attachments. Spherical bearings at the azimuth pivots preclude necessity for precise alignment of the structure when the mount is erected. No special tools or equipment are required for assembly and installation.

h. Required maintenance is minimal. For the manual mount, which uses self-lubricated bearings at all moving points, no periodic maintenance is required; it is necessary only to control corrosion at critical points and replace damaged components if required. The only additional requirement for the motorized version is an annual inspection of the drive units with lubricant replenishment in the jackscrew assembly as required.

Model 8010C 7-Meter Antenna Specifications

Technical Characteristics

Characteristic	Specification	
General		
Antenna Type	Cassegrain, dual-reflector, shaped	
Antenna Diameter	7 meters (23 feet)	
Mount Type	Elevation-over-Azimuth	
Coverage		
Azimuth	180° in 3 overlapping 110° sectors	
Elevation	0 degrees to 90 degrees	
Antenna Travel Rate (motorized option)*	Azimuth	Elevation
High Speed	1.9°/s	1.0°/s
Low Speed	0.18°/s	0.10°/s
Electrical		
Operating Frequency		
Receive	3.7 to 4.2 GHz	
Transmit	5.925 to 6.425 GHz	
-3 dB Beamwidth		
Receive	0.7° Nominal	
Transmit	0.46° Nominal	
Gain at OMT Output (midband)		
Receive	47.5 dBi	
Transmit	50.5 dB	
Voltage Standing Wave Ratio		
Receive and Transmit	1.3:1 Max	
Polarization	Linear	
Polarization Adjustment		
Manual	360° continuous	
Motorized	180° reversible	
Isolation between Ports		
Transmit/Receive	35 dB Min	
Receive/Receive	30 dB Min	
Cross-Polarized Suppression	35 dB Min on axis	

*Two-speed drive is standard with motorized option.

Technical Characteristics (continued)

Characteristic	Specification	
Electrical - Continued		
Radiation Pattern Sidelobe Envelope per FCC RFG 25-209	29-25 log θ dBi	1° ≤ θ ≤ 7°
	+8 dBi	7° < θ ≤ 9.2°
	32-25 log θ dBi	9.2° < θ ≤ 48°
	-10 dBi	48° < θ
Noise Temperature at Elevation Angle	EL-Angle	Kelvin
	5°	45
	10°	30
	15°	25
	20°	22
Environmental (Manual Option)		
Wind Resistance		
Survival	125 mph any direction or 87 mph with 2-inch radial ice at -40°F and above	
Operating	Holds pointing to within 0.08° rms in 45 mph winds gusting to 60 mph	
Ice	2-inch radial ice with 87 mph wind	
Environmental (Motorized Option)		
Wind Resistance at 0° (+32°F) Air Temperature		
Operational	Winds to 60 mph	
Survival with Antenna Stowed[1]	125 mph with no ice 87 mph with 2-inch radial ice	
Survival with Antenna at[1] Any Position (not stowed)	80 mph with no ice	
Temperature Range		
Operational	-29°C to +65°C (-20°F to +150°F)	
Survival	-40°C to +93°C (-40°F to +200°F)	
Shipping		
Weight (approximate)		
Shipping	5,000 lbs	
Net	3,750 lbs	
Shipping Volume (approximate)	600 cu ft	

[1] For definition of "stowing", see Figure 5.

10 and 11-Meter Earth Station Antennas for C-Band Operation

Nathan Knutson

Introduction

The antenna system is one of the important parts of an earth station. In a receive-only application, the antenna receives the desired signals transmitted from the satellite and must provide sufficient discrimination to unwanted signals which occupy the same congested frequency bands. In a transmit and receive broadcast application, the antenna not only receives signals, but also must transmit signals in a different frequency band and sometimes do both simultaneously. It must be highly reliable and capable of withstanding and operating under severe environmental conditions.

Earth station antenna requirements are dependent on many factors such as channel capacity, receiving equipment, signal modulation, reliability requirements, geographical location relative to the satellite, site interference profiles, and site environmental conditions. This discussion will be limited to the characteristics and specifications of 10- and 11-meter earth terminals with one or two video downlinks and/or uplinks.

Series 8000 10- and 11-Meter Antennas

The Scientific-Atlanta Series 8000 10- and 11-meter earth station antennas are presently being used in hundreds of installations throughout the world to meet the requirements of high-reliability and high-quality service for television, voice and data communications. This accumulated experience in design, manufacturing, installation, and continuing reliability of operation has proven the concepts and the hardware.

The Model 8002A-HP 10-meter earth station antenna provides receive/transmit and receive-only capabilities in the 4/6-GHz bands to fulfill voice, message, and video requirements in communications systems. This earth station antenna is a dual-reflector, Cassegrain system that exhibits highly efficient RF performance as well as low sidelobes making it suitable for use in congested terrestrial microwave areas.

Designed specifically for INTELSAT-B applications, the Model 8007 11-meter earth station antenna is a cost-effective, high-efficiency antenna that exhibits high G/T and low sidelobes. This antenna, capable of receive/transmit operation in the 4/6-GHz range, is also well suited for use in areas of low satellite signal strength such as Puerto Rico, Hawaii, and Alaska.

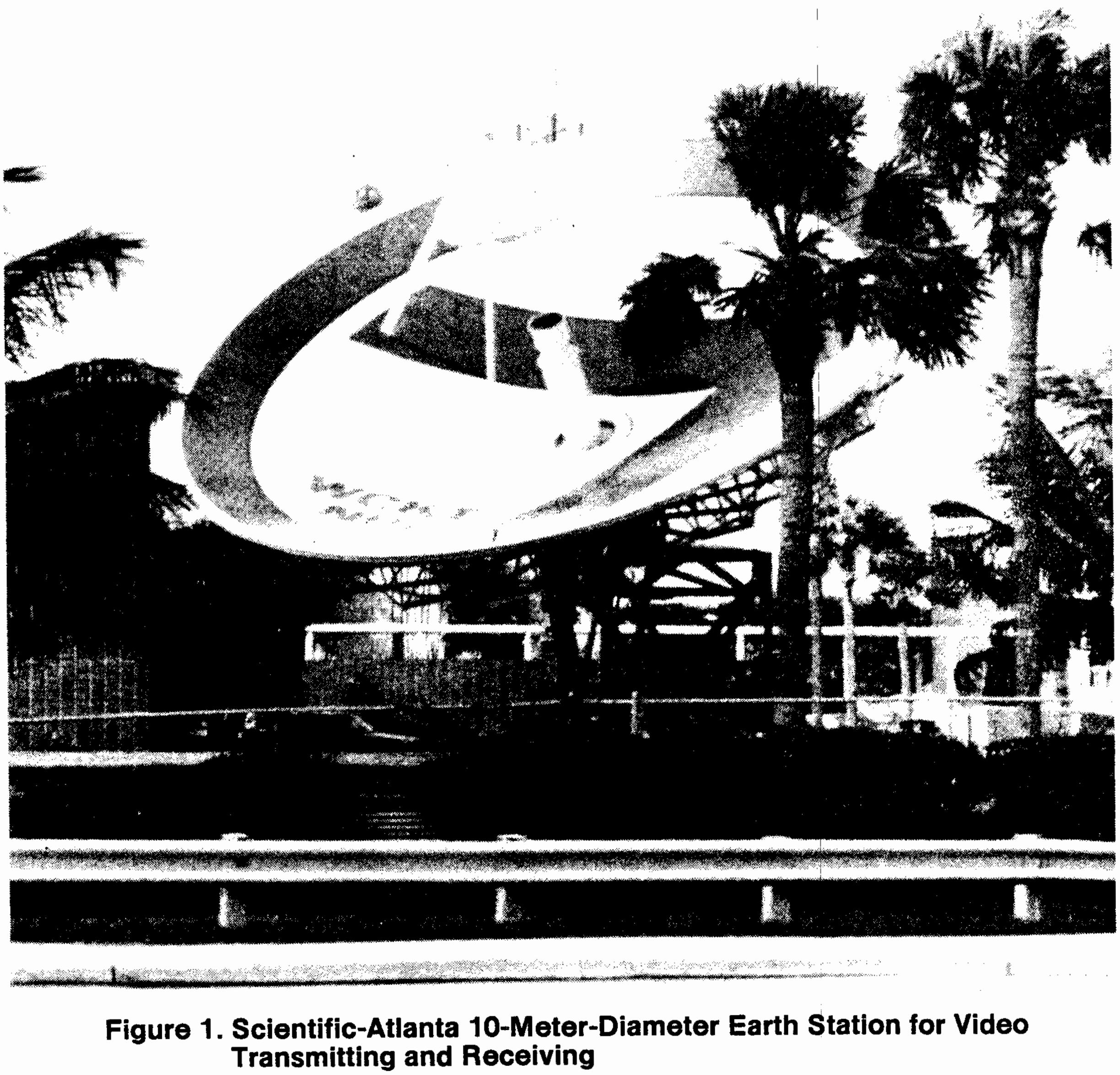
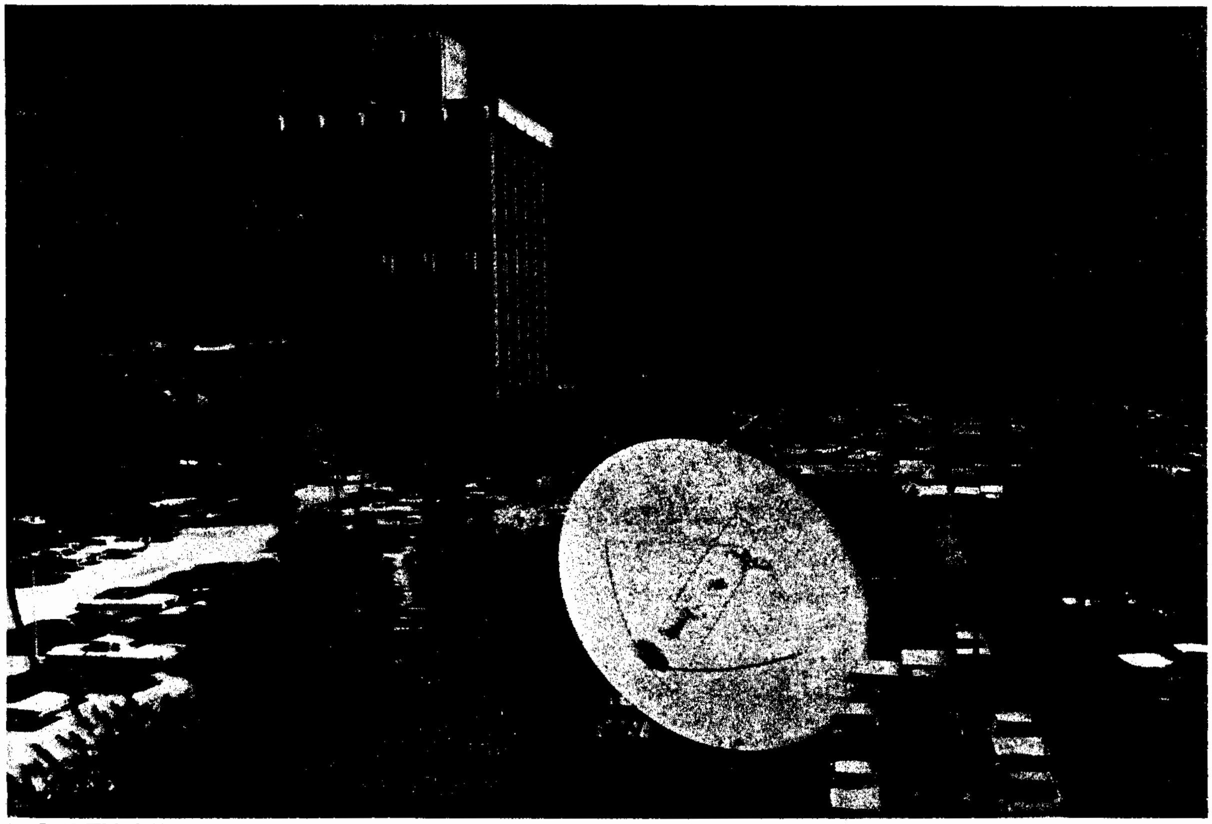


Figure 1. Scientific-Atlanta 10-Meter-Diameter Earth Station for Video Transmitting and Receiving


WCIX-TV Miami, Florida

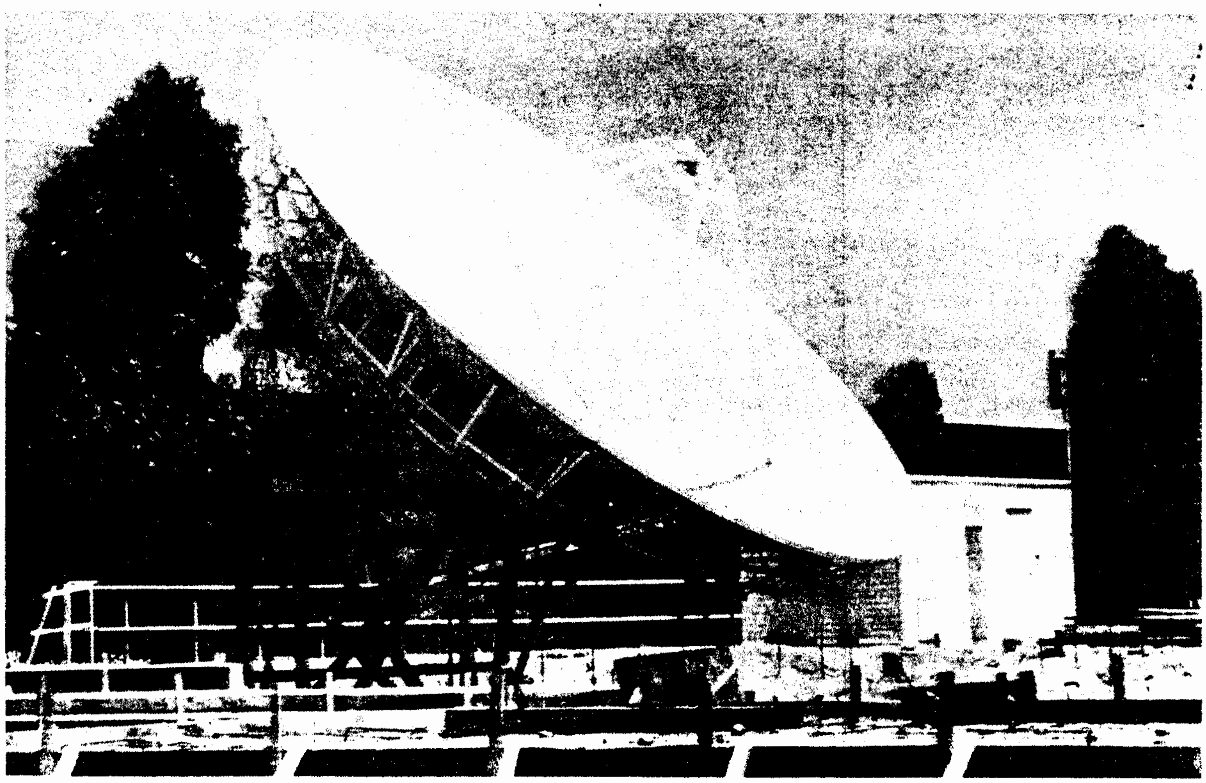
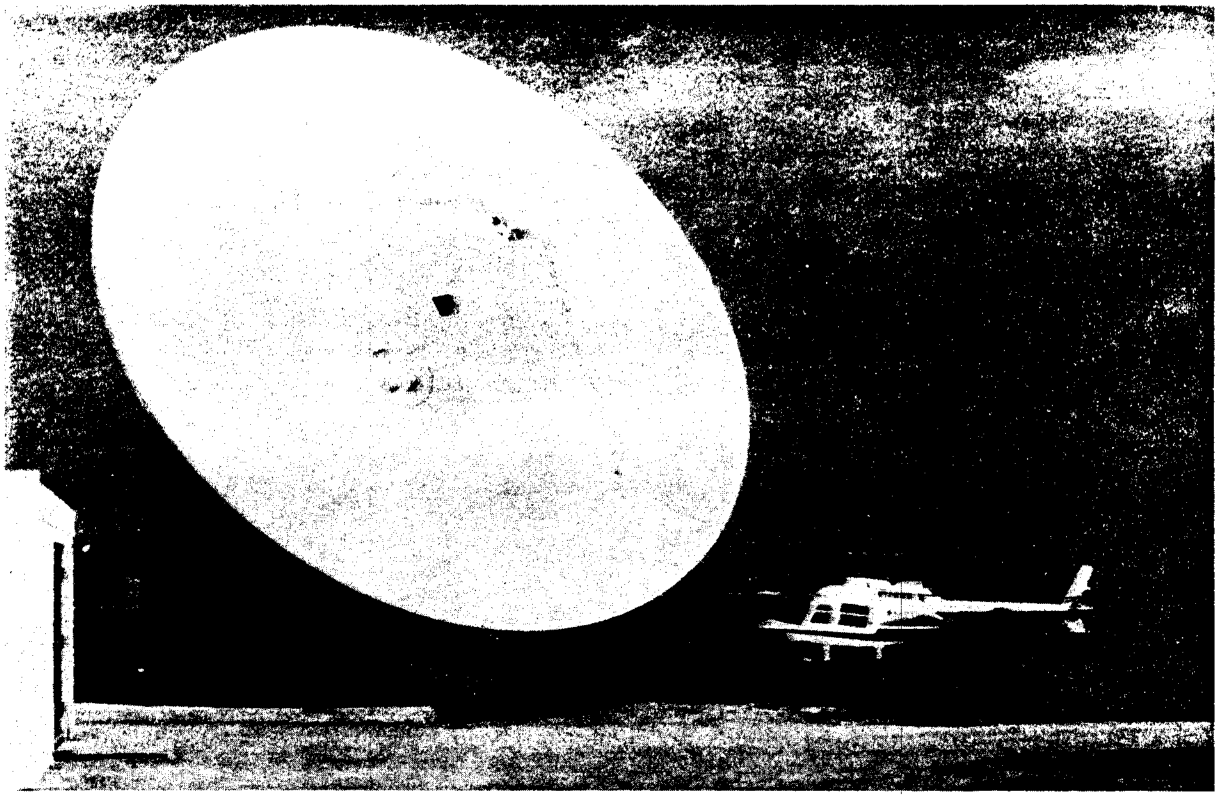
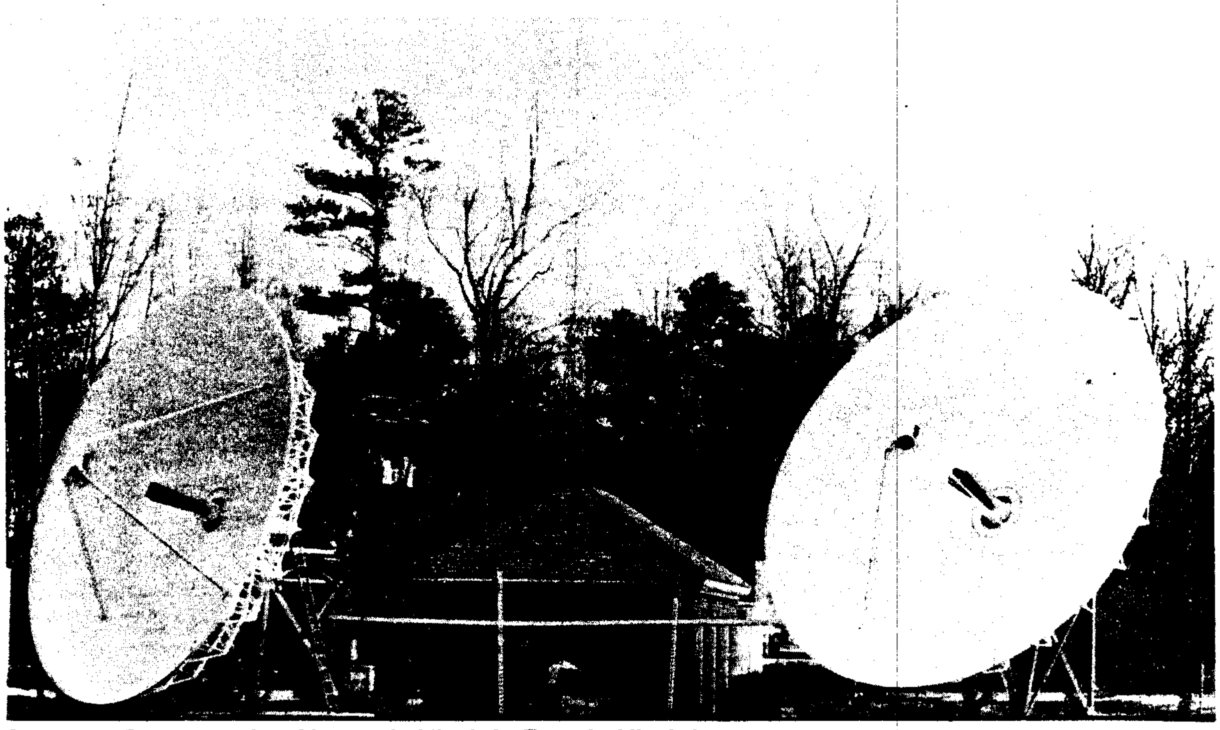

Trinity Broadcasting Network Santa Ana, California

Figure 2. Scientific-Atlanta Satellite Earth Stations for Television Broadcasters



WTOG-TV St. Petersburg, Florida



Christian Broadcasting Network Virginia Beach, Virginia

Figure 3. Scientific-Atlanta Satellite Earth Stations for Television Broadcasters

Shaped Reflectors

The Scientific-Atlanta Series 8000 dual-shaped Cassegrain reflector systems are optically designed to yield high-aperture efficiency.

The reflector assembly used in all configurations consists of a machined central hub, to which are attached the radial trusses and the reflector panels. Ample space is provided inside the hub for the mounting of redundant low-noise amplifiers and switching equipment. Each reflector panel is a single piece manufactured on precision tooling to ensure accuracy and to allow interchangeability. Machined tabs on the radial trusses support the panels and lock-in the overall shape and antenna surface tolerance. Optical alignment of panels during installation is not required.

The reflector assemblies are made entirely of aluminum. The choice of a single material for the structure has the advantage of having a single coefficient of thermal expansion for all parts. Aluminum has a thermal conductivity that is approximately three times that of steel. An all-aluminum antenna will more quickly reach thermal equilibrium when differentially heated by the sun or wind chill. Differential expansion or contraction is then a transient condition which is minimized by high thermal conductivity of the all-aluminum antenna reflector.

The reflector is produced on precision jigs and fixtures. All attachment points are machined or set on precision tooling. The 6-foot-diameter hub has 48 machined attachment pads for the 24 structural backup rib/trusses. The rib/trusses each have six machined mounting bars which locate the reflector surface panels. Each panel is fabricated on a special shaping fixture that consists of a number of aluminum templates manufactured on a numerically controlled machine. This fixture creates the desired surface shape in the reflector panels and locates reference attachment holes and edges to within 0.003 inch of the desired surface.

The panels are reinforced by a number of serrated rib members which are both welded and riveted to the panel facings in a manner developed by Scientific-Atlanta to minimize shrinkage and distortions which would occur in an all-welded assembly. The reflector has no field adjustments for locating discrete points into difficult-to-establish proper position.

The reflector needs no expansion joints or compliant attachment points to accommodate differential expansion of dissimilar materials. The Series 8000 10- and 11-meter reflectors have substituted machining accuracy at the factory for correctional assembly in the field. When assembled, the reflectors are a strong, rigid structure that is ready for operation.

Analytical procedures (including computer simulations) and load tests were both employed in the design of the reflector and mount. Assembled reflectors were load tested and the deflections recorded with the measurements to confirm the calculations.

The Series 8000 reflector designs have been proven in some of the most extreme environmental conditions recorded, and the reliability achieved continues to demonstrate the strength, stiffness, accuracy, and thermal stability of the design.

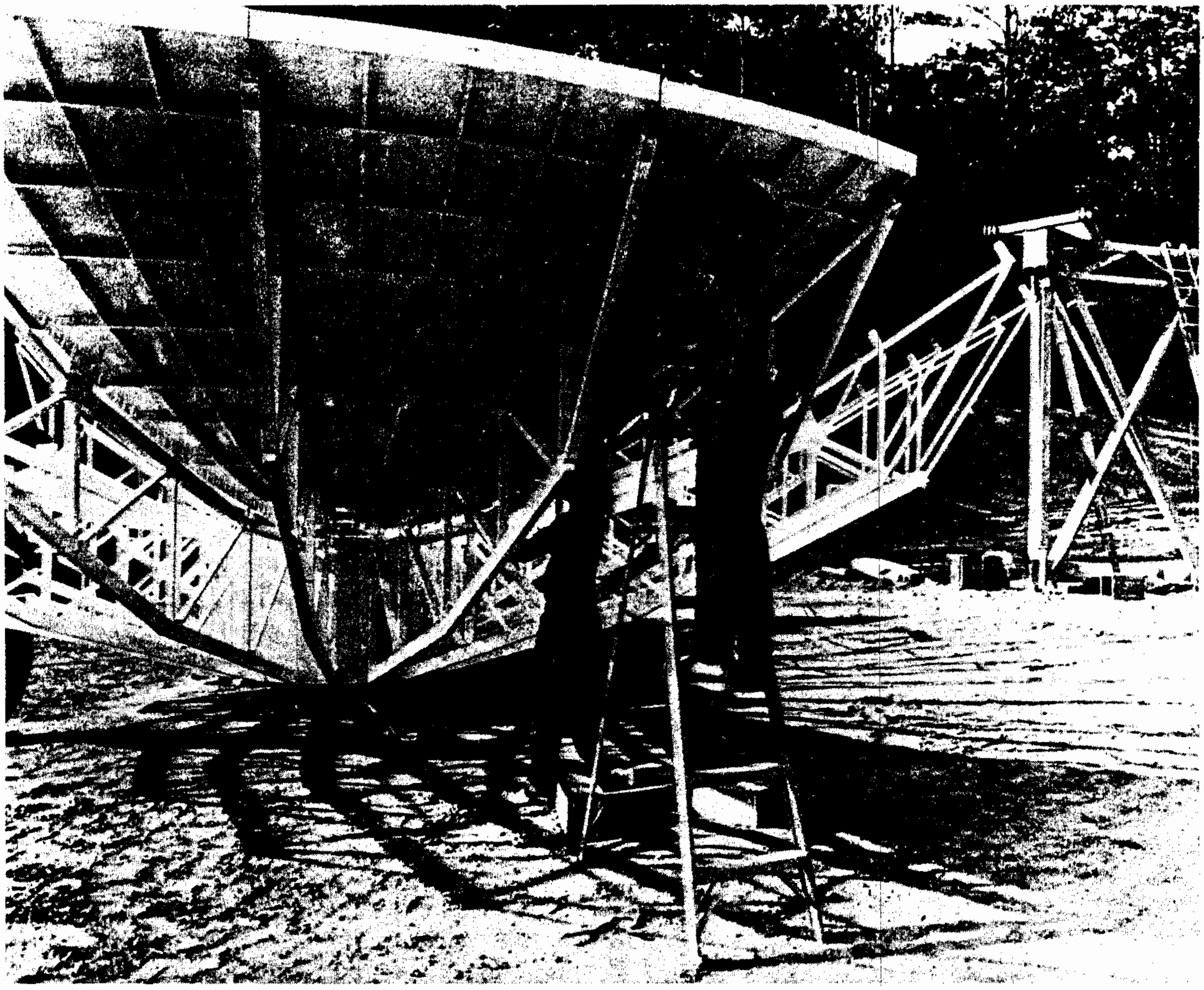


Figure 4. Installation of Series 8000 Earth Terminal Reflector

Feed Systems

The 10- and 11-meter feed systems consist of three major parts: (1) a radiating horn, (2) a diplexer or orthomode transducer (OMT), and (3) a specially-shaped subreflector which complements the shaped reflector. The feed system design is directed toward achieving high efficiency while maintaining an average sidelobe envelope beneath the 32-25 log θ gain curve in both frequency bands.

The feed system employed on the 10- and 11-meter antennas is similar to a conventional Cassegrain type which uses a directional feed element and a hyperboloidal subreflector to illuminate a paraboloidal reflector. The basic difference results from the use of specially-shaped reflectors and subreflectors to increase aperture efficiency without degrading sidelobe performance.

The radiating feed horn of a shaped Cassegrain configuration is designed to have a large amplitude taper across the subreflector in order to reduce spillover energy which affects the sidelobe performance. The subreflector is then shaped to redistribute the energy to provide a near uniform amplitude distribution across the main reflector for high aperture efficiency. The shape of the main reflector is designed to be complementary to the subreflector and corrects the quadratic phase error of the illuminating function. The result is efficiency exceeding 70%.

The Model 8002A-HP, high-performance 10-meter antenna and the Model 8007 11-meter antenna utilize corrugated conical horns as the radiating feed element (see Figure 6). Although more costly to produce, these horns provide optimum radiation patterns for high efficiency-low sidelobe performance and exhibit excellent cross-polarization characteristics needed for frequency-reuse operation.

The 10-meter high-performance antenna and the 11-meter antenna use a precision machined aluminum cast subreflector. They are 56 inches and 60 inches in diameter, respectively. They are supported by a tripod spar arrangement made from aluminum extrusions which minimize the effect on sidelobe performance.

The 10- and 11-meter antennas may be equipped with diplexers or orthomode transducers to meet the requirements of any application. The available options are as follows:

a. Receive-Only

- Single polarization
- Dual orthogonal polarization, linear/circular.

b. Transmit/Receive

- One transmit port, one receive port with orthogonal polarization, linear/circular.
- One transmit port, one receive port with coplanar polarization, linear.
- One transmit port, two orthogonal receive ports, linear.
- Two orthogonal transmit ports and two orthogonal receive ports (fixed), linear/circular.
- Two orthogonal transmit ports and two orthogonal receive ports (4-GHz and 6-GHz ports independently adjustable), linear/circular.

The waveguide interfaces are CPR 229G and CPR 137G flanges for receive and transmit, respectively. The waveguide interface is conveniently located on the inside of the reflector hub, thus providing a weather-resistant enclosure with ample space for redundant amplifiers with switching and filtering, as shown in Figure 7.

The feed horn is covered with a 0.005-inch-thick LLUMAR® or polyester film material which allows pressurization of the horn/transducer assembly up to 1 psig.

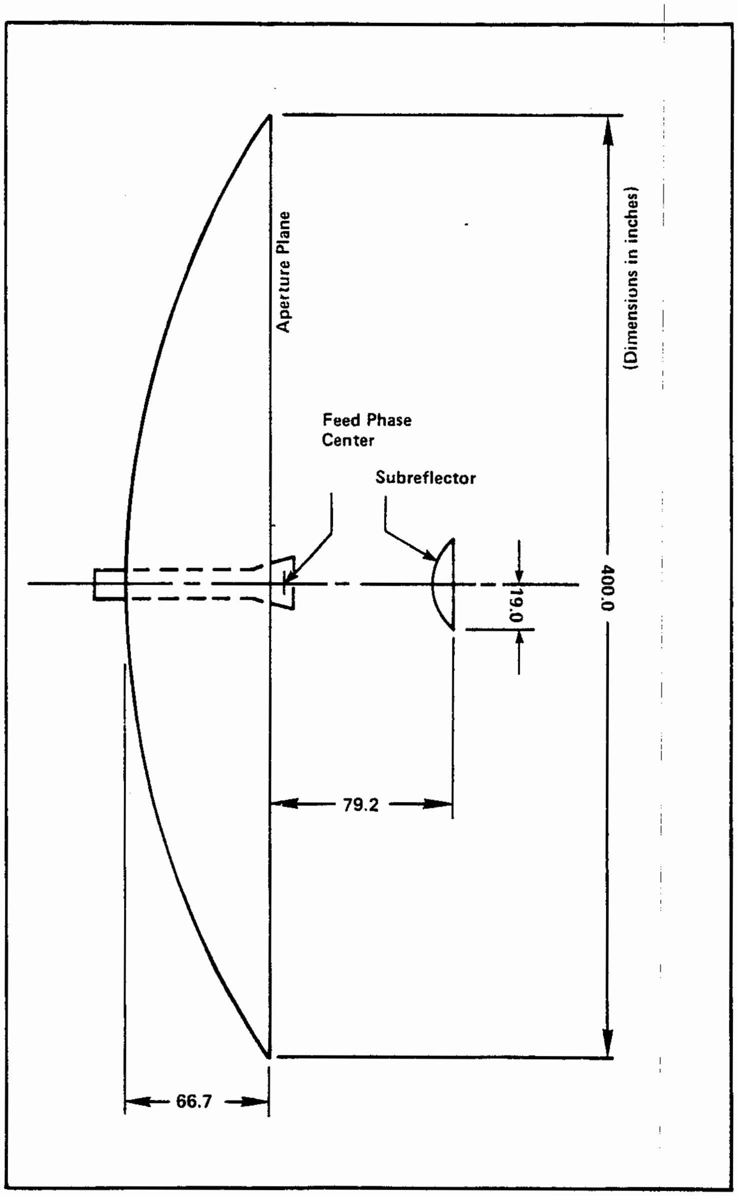


Figure 5. Model 8002, 10-Meter Earth Terminal Antenna Basic Antenna Geometry

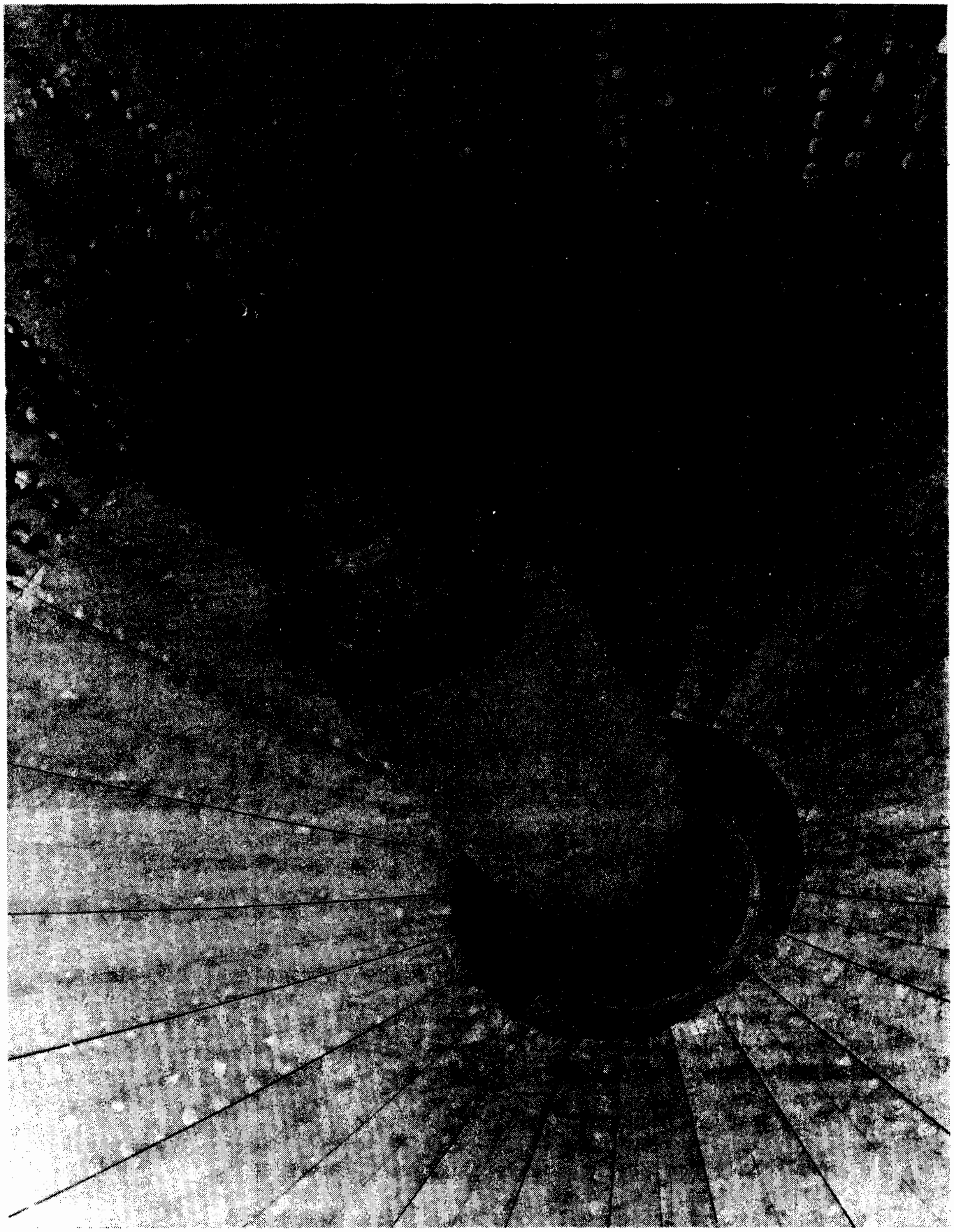


Figure 6. Corrugated Horn for 4- and 6-GHz in 10-Meter Antenna System

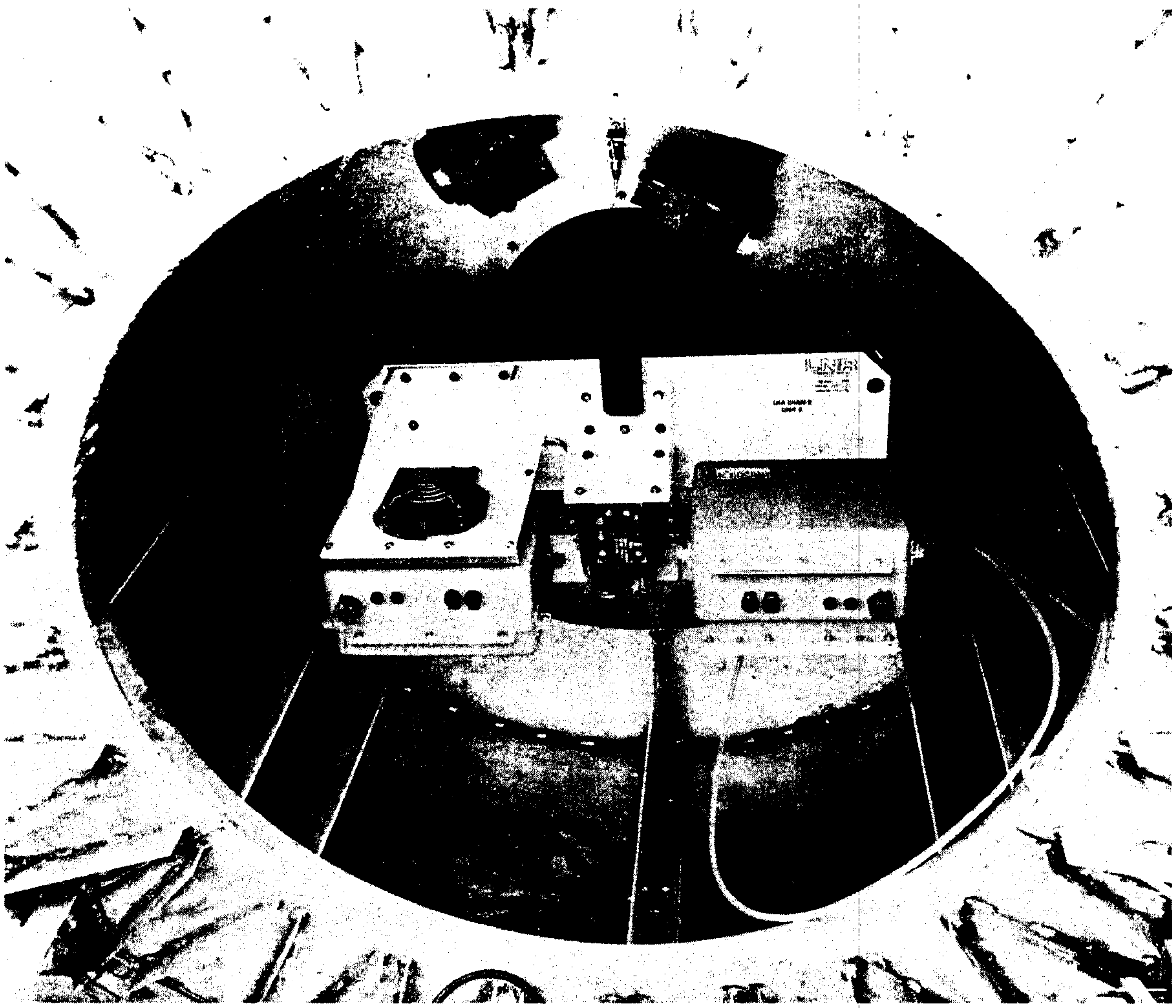


Figure 7. Redundant 4-GHz Parametric Amplifier System Installed in the Model 8111 10-Meter Hub (hub access covers removed)

Mount and Positioning System

The Models 8002A-HP and 8007 antennas use the conventional elevation-over-azimuth mounting configuration. The advantages of this configuration are the following:

- Easily understood coordinate system
- Tolerance of foundation sighting errors
- Applicable for use at any latitude and longitude
- One standard design with one foundation interface plan
- Capable of self-erection and assembly of reflector in place on mount
- Capable of having a horizontal work platform to conveniently service antenna hub mounted equipment.

Two mounts are available to support and position the 10-meter antennas. Constructed of structural steel, the principal difference between the two mounts is their height; the Model 8024 is 3.7m (12 ft.) high, while the Model 8021 is 5.2m (17 ft.) high. The taller mount permits the reflector to be depressed below +15° elevation to enable boresighting geostationary satellites from high northern latitudes.

The Model 8025 11-meter mount is generally similar to the Model 8021 except that heavier structural members are used throughout.

The Scientific-Atlanta elevation-over-azimuth mount is an equilateral triangular tower structure. The triangular tower rises to a height of 17-feet (or 12 feet) and terminates in a horizontal top which is triangular in shape and 10 feet on a side. The tower has three vertical legs, each perpendicular to the local horizontal. The design objective of the tower is to obtain maximum torsional stiffness to resist windup caused by azimuth wind moments acting on the reflector.

An optimization study was performed to determine the most efficient method to carry shear loads across the three vertical faces of the tower. The X-brace structure which ties two upper nodes of the tower to two lower nodes was found to give maximum stiffness per pound of material. The azimuth jack reaction force forming a couple with the upper-azimuth bearing housing (attached to the top of the front leg) is resisted primarily by the three sets of X-braces of the tower. One may visualize the structure as a three-sided prism having walls 12 or 17 feet high and 10 feet wide. The 10-foot width of the three walls gives the required degree of torsional stiffness.

The 10-foot spread between the three mounting feet of the tower eases the alignment problems of the azimuth axis of the mount. Shims are provided to go over the foundation anchor bolts and space between the concrete and the mounting feet of the mount. Bringing the height of the three mounting feet to within 1/16 of an inch of each other is an easy installation task. When this is done, the peak tilt of the azimuth axis is less than 0.04 degrees and the attendant cross-coupling in elevation for motion in azimuth becomes negligible in acquisition of the satellite from tabulated data. There is no further alignment or shimming required. All members of the mount are bolted together in an orderly sequence. Because of the close tolerance in hole locations of the various members, successful assembly is assured, giving a square, precise structure.

The mounts are made with ASTM A-36 steel throughout. All parts are hot-dip galvanized, except for the lower azimuth bearing housing and the azimuth turntable which are cadmium plated with a chromate-conversion coating. All hardware is A325 per AISC and is also hot-dip galvanized.

Antenna positioning about two axes is provided by two precision jackscrew assemblies. These units permit rapid reorientation of the antenna during changeover from one satellite to another. The standard manual 10-meter configuration, consisting of Models 8035A Azimuth and 8036A Elevation Actuators, can also be driven by a hand-held drill motor.

Positioning of the 11-meter antenna is provided by standard motor-driven azimuth and elevation jackscrews. Digital position readout, automatic position programming, remote polarization adjustment, and local control are included in the available optional control equipment.

Currently available satellites now span an orbital arc of from 67° to 143° west longitude. For most of the U.S., this equates to a requirement of 110° of azimuth coverage for an elevation-azimuth configured antenna mount. The Scientific-Atlanta 10- and 11-meter antennas have the 110° azimuth coverage capability in either the manual or motorized actuator options. For the manual actuators this coverage is accomplished by two overlapping 57° sectors which require repositioning the azimuth actuator for the respective sector coverage.

To reposition the antenna quickly and efficiently, it may be desirable to have a motor-actuated antenna to take full advantage of all the programming available on the various geosynchronous satellites. Scientific-Atlanta has introduced a 110°-continuous coverage motorized azimuth actuator. Designated the Model 8031-110 for the 10- and 11-meter antennas, these actuator options are offered on new antennas and also offered as retrofit kits for current owners of Scientific-Atlanta 10- and 11-meter antennas.

Recently introduced is an actuator system for high-speed antenna rotation. Models 8031SHS-110 azimuth and 8032SHS elevation actuators will reposition the 10- or 11-meter antennas along 110° of orbital arc within one minute. With the recent surge in available programming on various satellites, this rapid satellite change capability greatly increases the value of the earth station. Now, tight programming schedules need not be limited by the time constraints of satellite switching as before. These high-speed systems are also offered as retrofit kits to current owners of Scientific-Atlanta 10- and 11-meter antennas.

A wide variety of options is available for providing motorized actuators and remote position control for the Series 8000 10- and 11-meter antennas.

Each of the combinations of actuators and control units requires a Model 8841A Local Contactor at the antenna mount for independent but simultaneous control of all axes. These units are housed in weatherproof NEMA enclosures and provide the interface between primary power, actuator motors, position data potentiometers, and the Model 8840A programmable remote control unit.

System electrical grounding is ensured by a ground strap between the reflector and mount (therefore across the mount bearings). The mount has high-conductivity metal joints throughout. Holes are provided in the mount rear legs for attachment to suitable ground.

Figure 8. 11-Meter INTELSAT Standard B Earth Station

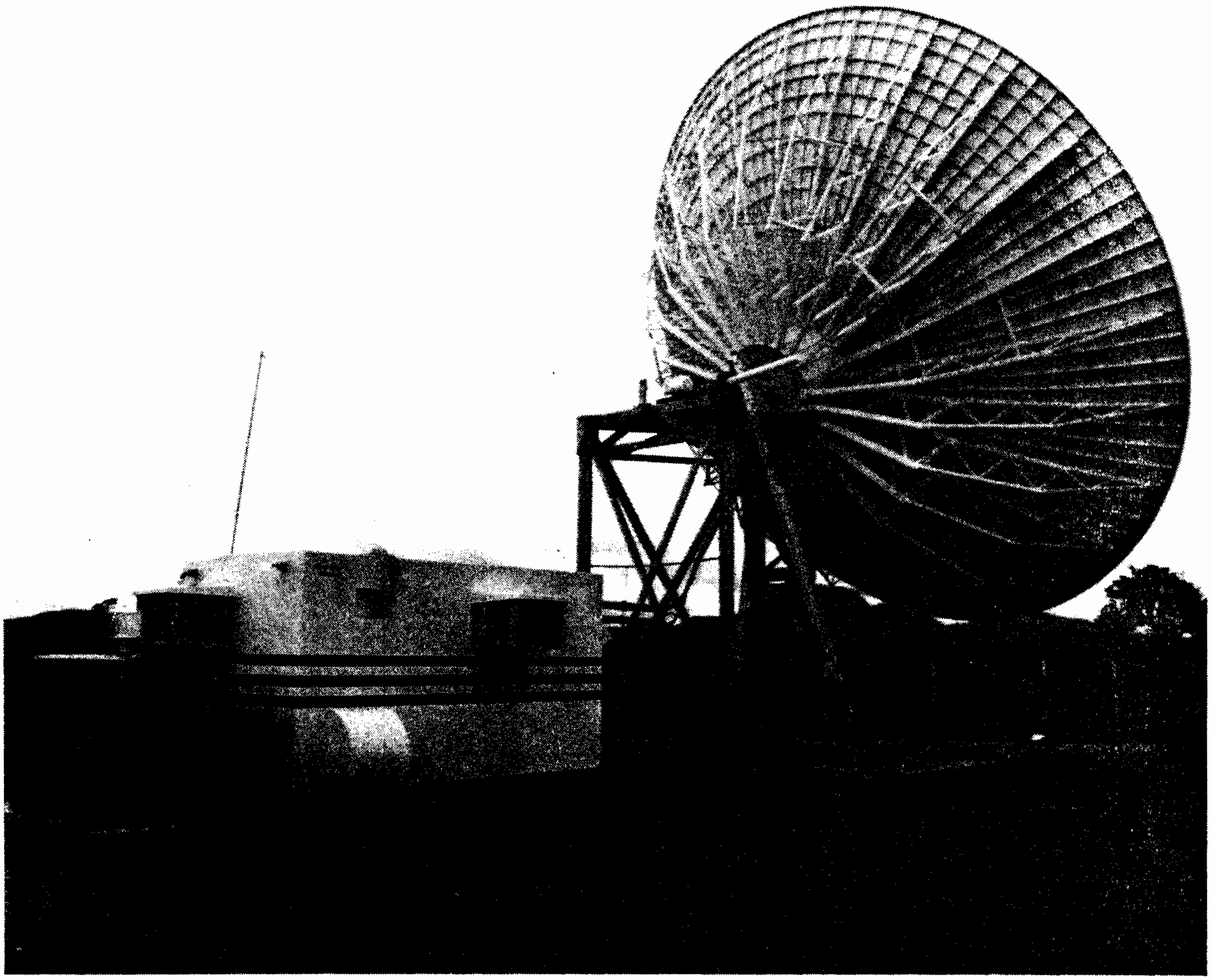


Figure 9. 11-Meter Intelsat Standard B Earth Station in Nigeria, Africa.

Shipping and Installation

The design parameters of the Series 8000 antennas and the manufacturing processes employed were chosen to minimize costs of shipping and installation. The antenna, when crated, weighs less than 22,000 pounds and occupies volume of less than 2,600 cubic feet. They can be conveniently loaded on a 40-foot trailer for transportation to the installation site.

The installation procedures, as mentioned previously, require no field alignment of panels, trusses, or feed, which greatly simplifies the task. An erection kit is also optionally available for installation of the antenna without the use of a crane.

The typical foundation requires approximately 44 cubic yards of concrete. Scientific-Atlanta recommends that a professional engineer be employed to evaluate foundation requirements at each site, based on local soil conditions.

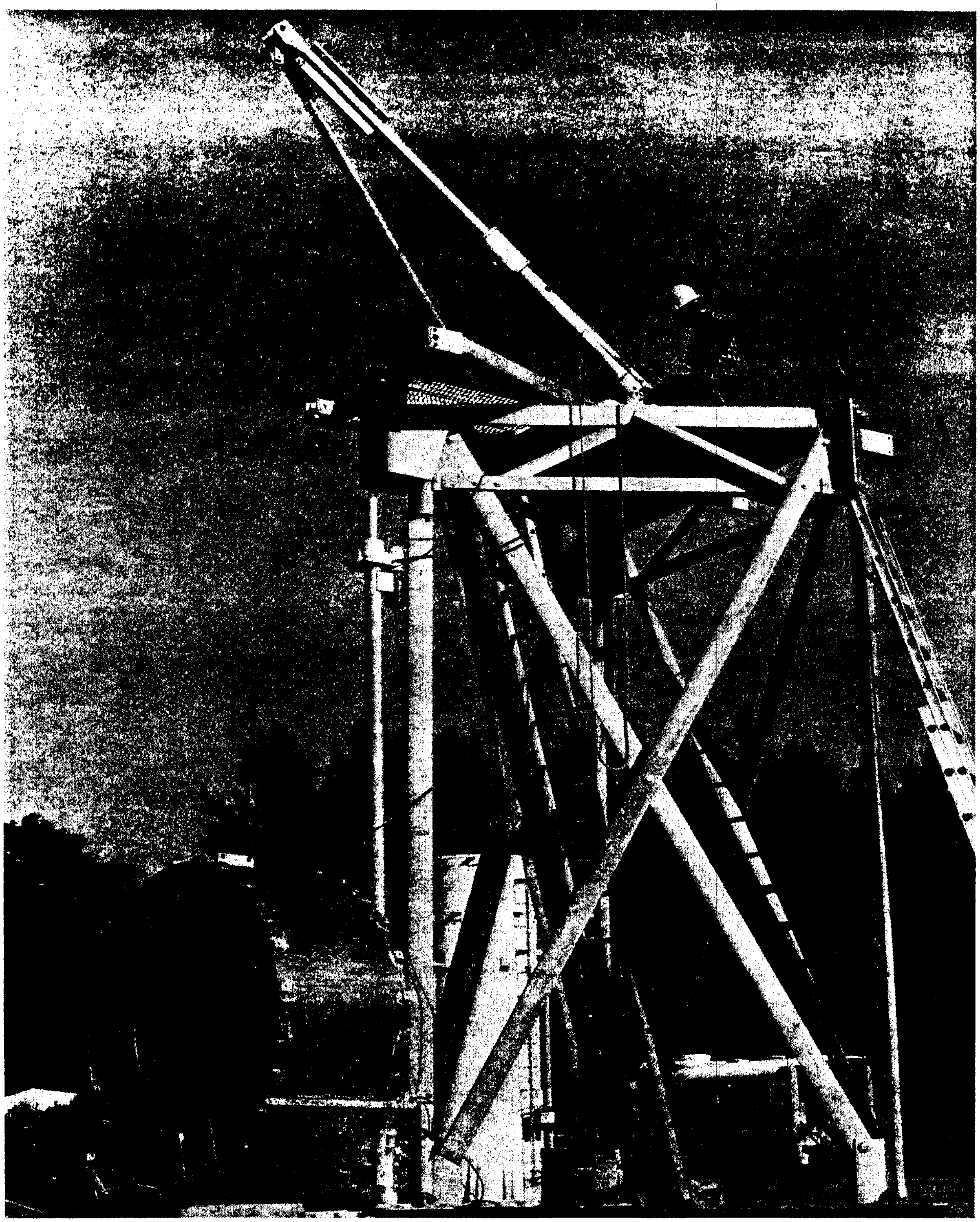


Figure 10. Scientific-Atlanta 10-Meter Antenna Installed Without Using Crane

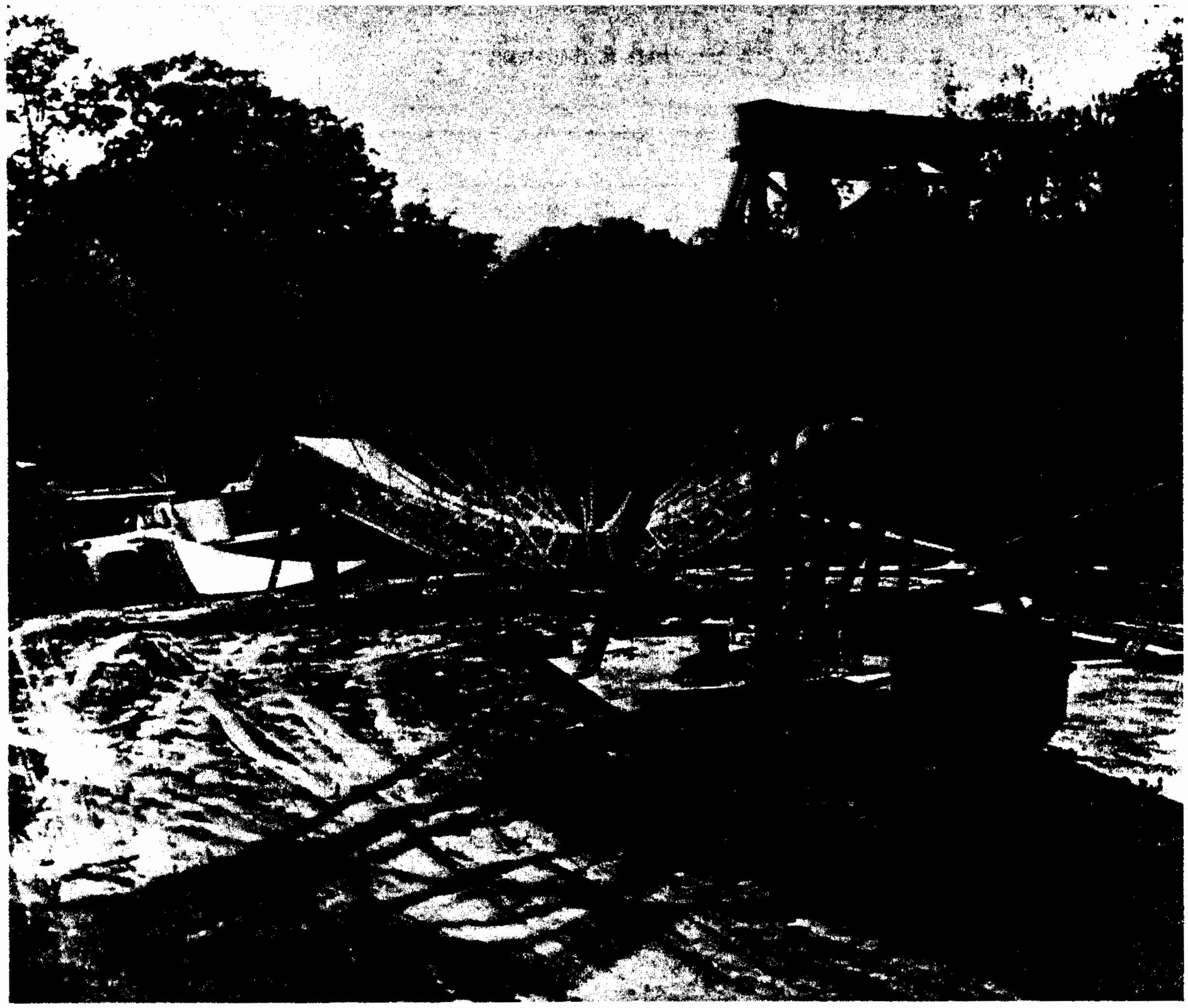


Figure 11. Installation of Series 8000 Earth Terminal Mount

Industry Standards

The antenna systems were designed to meet or exceed the following industry standards:

American National Standards Institute	A58.1
American National Standards Institute	A95.1
Electronic Industries Association	RS-195-B
American Institute of Steel Construction	
The Aluminum Association	

Models 8002A-HP and 8007 Earth Station Antennas

Technical Characteristics

Characteristic	Specification	
GENERAL	8002A-HP	8007
Antenna Type	Cassegrain, Dual Shaped Reflector	Same
Antenna Diameter	10 meter (32.8 feet)	11 meter (37 feet)
Pointing Range		
Azimuth Standard and Optional Motorized	110° in 2 overlapping 57° sectors	--
Azimuth, Standard Motorized	--	110° in 2 overlapping 60° sectors
Azimuth, Optional Extended Coverage Motorized	110° in one continuous sector	
Elevation	Tall Mount, 0° to +90° Short Mount, +15° to +90°	+5° to +90°
Polarization Adjustment	360° Manual Standard ±90° Motorized Remote*	Same
Antenna Travel Rate (motorized version)		
Azimuth		
Standard	0.01°/s	0.03°/s
High Speed	0.1°/s	Same
Super High Speed	1.87°/s	Same
Elevation		
Standard	0.01°/s	0.03°/s
High Speed	0.07°/s	--
Super High Speed	1.1°/s	Same
Power Requirements (actuators only)**	208V ac, 3ϕ, 4-wire, 60 Hz	Same
Surface Tolerance	1 mm (0.040 inch) static	Same
Pointing Accuracy	0.041° rms in 48-km/h (30-mi/h) wind gusting to 72-km/h (45-mi/h)	Same
	0.091° rms in 72-km/h (45-mi/h) wind gusting to 105-km/h (65-mi/h)	Same
ELECTRICAL		
Operating Frequency		
Receive	3.7 to 4.2 GHz	Same
Transmit	5.925 to 6.425 GHz	Same
Feed Type	Corrugated Horn	Corrugated Horn
Gain (Ref. OMT Port)		
Receive	50.85 dBi	52.00 dBi
Transmit	53.50 dBi	54.80 dBi

Models 8002A-HP and 8007 Earth Station Antennas

Technical Characteristics (continued)

Characteristic	Specification	
	8002A-HP	8007
VSWR	1.30:1 Max	Same
Polarization	Linear (Optional Circular)	Same
Axial Ratio (on-axis)		
Linear	35 dB Min (standard)	Same
Circular	3 dB Max 0.5 dB Max (optional)	Same
Isolation (port-to-port)		
Linear Receive	30 dB Min	Same
Transmit-Receive	35 dB Min	Same
Half-Power Beamwidth (nominal)		
Receive	0.46°	0.40°
Transmit	0.32°	0.26°
First Sidelobe Level	-14 dB	Same
ENVIRONMENTAL		
Operational	-20°C to +55°C	Same
Survival***	-35°C to +60°C	Same
Wind Loading at 0° Air Temperature		
Operational	72-km/h (45-mi/h) gusting to 105-km/h (65-mi/h)	Same
Survival***	200-km/h (125-mi/h) with no ice; 140-km/h (87-mi/h) with 51 mm (2 in.) radial ice	
Solar Radiation	1.1 mW/mm^2 (350 Btu/ft^2-hr)	
Atmospheric Conditions	Salt, Pollutants, and corrosive contaminants encountered in coastal and industrial areas	Same
SHIPPING		
Weight (approximate)		
Shipping	17,500 lbs.	21,500 lbs.
Net	11,000 lbs.	15,000 lbs.
Shipping Volume (approximate)	1,900 ft^3	2,600 ft^3

*Optional remote control, motor-driven polarization drive.
**Other voltages/frequencies available.
***Survival conditions considered separately. Adjustable components securely clamped. Antennas are to be driven to favorable stowing orientations for survival wind conditions.

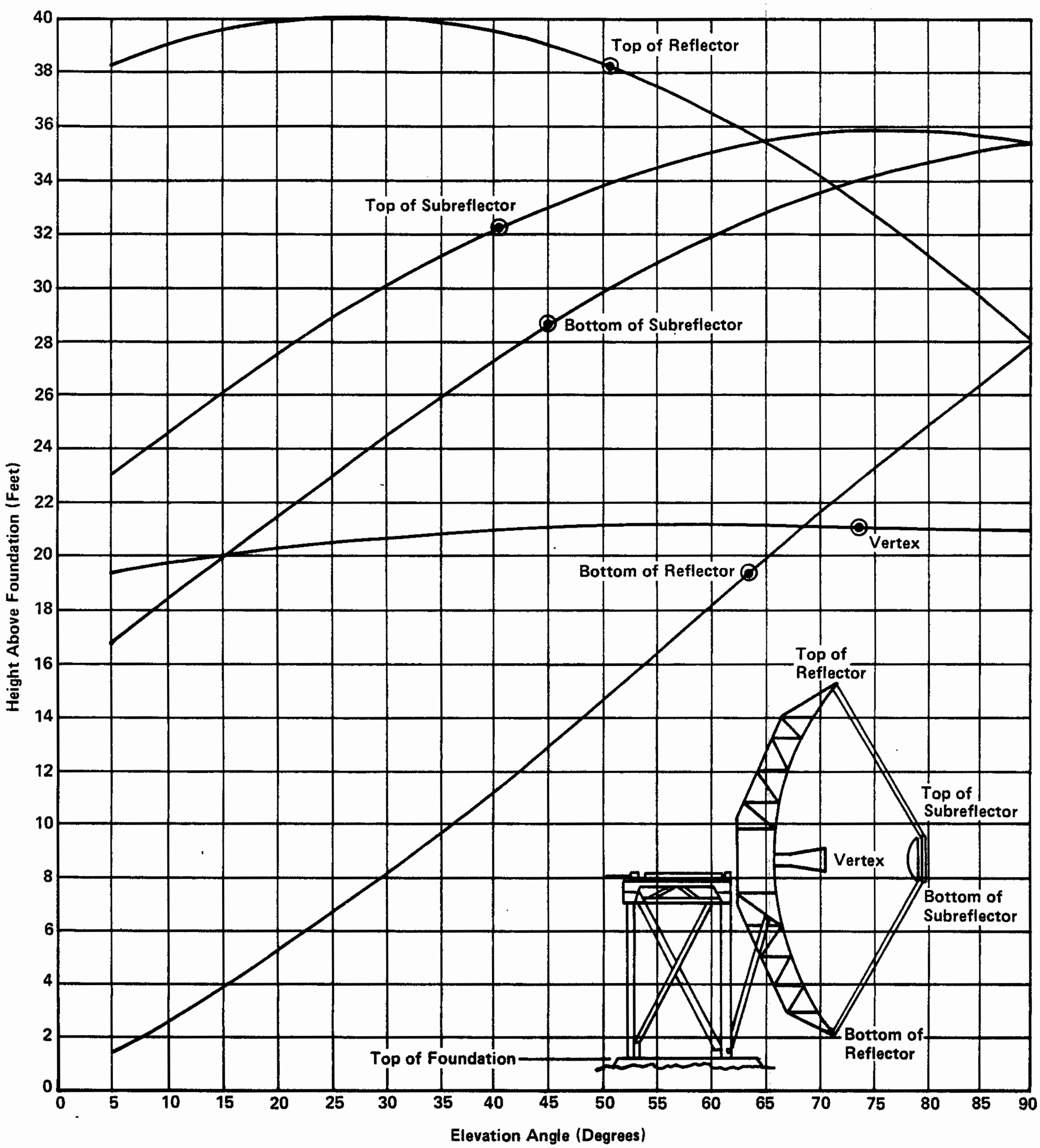


Figure 12. Model 8007 Outline Heights above Foundation vs. Elevation Angle (11 meter)

Models 8002A-HP and 8007 Earth Station Antennas

Technical Characteristics (continued)

Characteristic	Specification 8002A-HP	8007
VSWR	1.30:1 Max	Same
Polarization	Linear (Optional Circular)	Same
Axial Ratio (on-axis)		
Linear	35 dB Min (standard)	Same
Circular	3 dB Max 0.5 dB Max (optional)	Same
Isolation (port-to-port)		
Linear Receive	30 dB Min	Same
Transmit-Receive	35 dB Min	Same
Half-Power Beamwidth (nominal)		
Receive	0.46°	0.40°
Transmit	0.32°	0.26°
First Sidelobe Level	-14 dB	Same
ENVIRONMENTAL		
Operational	-20°C to +55°C	Same
Survival***	-35°C to +60°C	Same
Wind Loading at 0° Air Temperature		
Operational	72-km/h (45-mi/h) gusting to 105-km/h (65-mi/h)	Same
Survival***	200-km/h (125-mi/h) with no ice; 140-km/h (87-mi/h) with 51 mm (2 in.) radial ice	
Solar Radiation	1.1 mW/mm^2 (350 Btu/ft^2-hr)	
Atmospheric Conditions	Salt, Pollutants, and corrosive contaminants encountered in coastal and industrial areas	Same
SHIPPING		
Weight (approximate)		
Shipping	17,500 lbs.	21,500 lbs.
Net	11,000 lbs.	15,000 lbs.
Shipping Volume (approximate)	1,900 ft^3	2,600 ft^3

*Optional remote control, motor-driven polarization drive.
**Other voltages/frequencies available.
***Survival conditions considered separately. Adjustable components securely clamped. Antennas are to be driven to favorable stowing orientations for survival wind conditions.

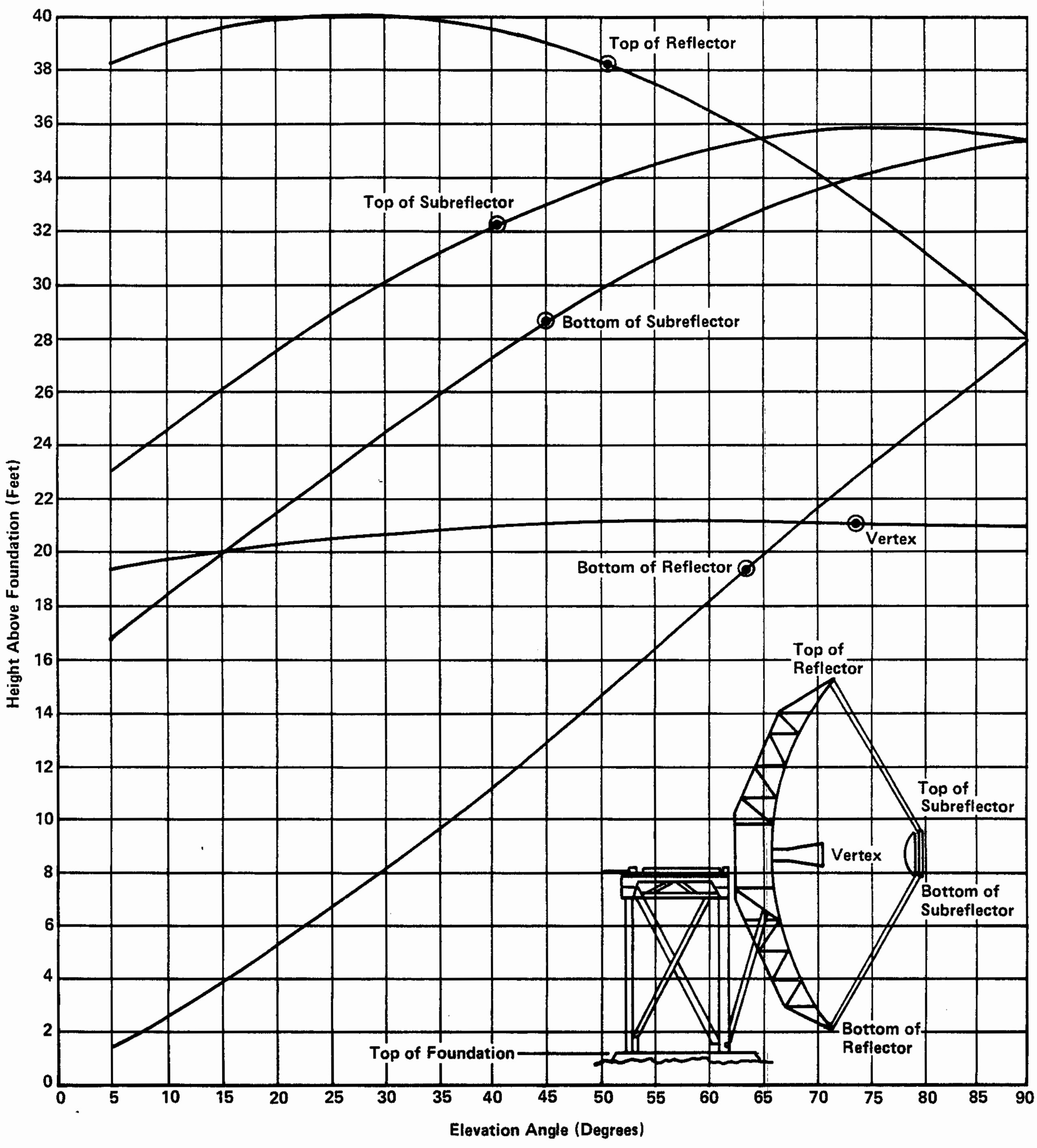


Figure 12. Model 8007 Outline Heights above Foundation vs. Elevation Angle (11 meter)

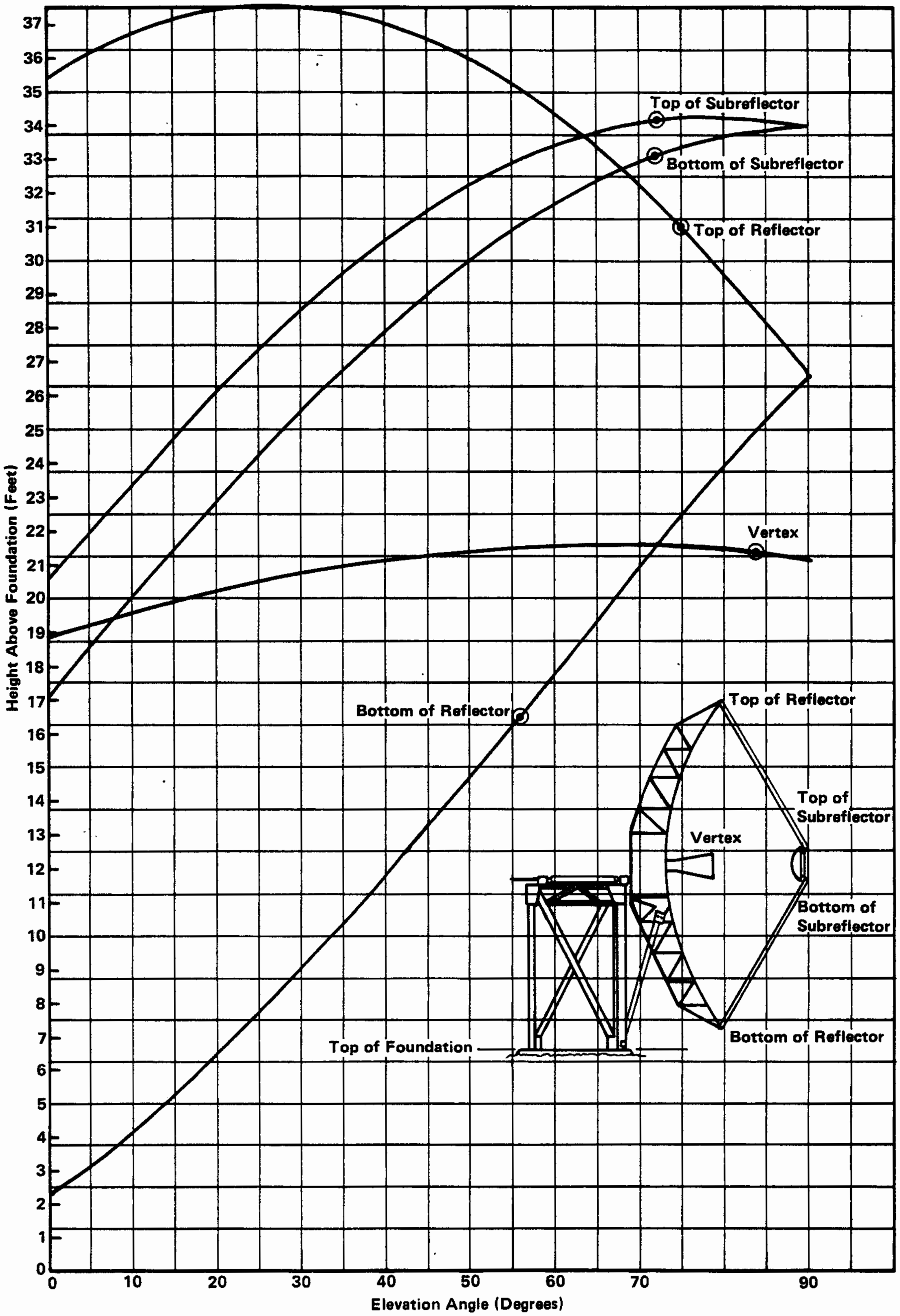


Figure 13. Model 8002 Outline Heights above Foundation vs. Elevation Angle, Standard Mount (10 meter)

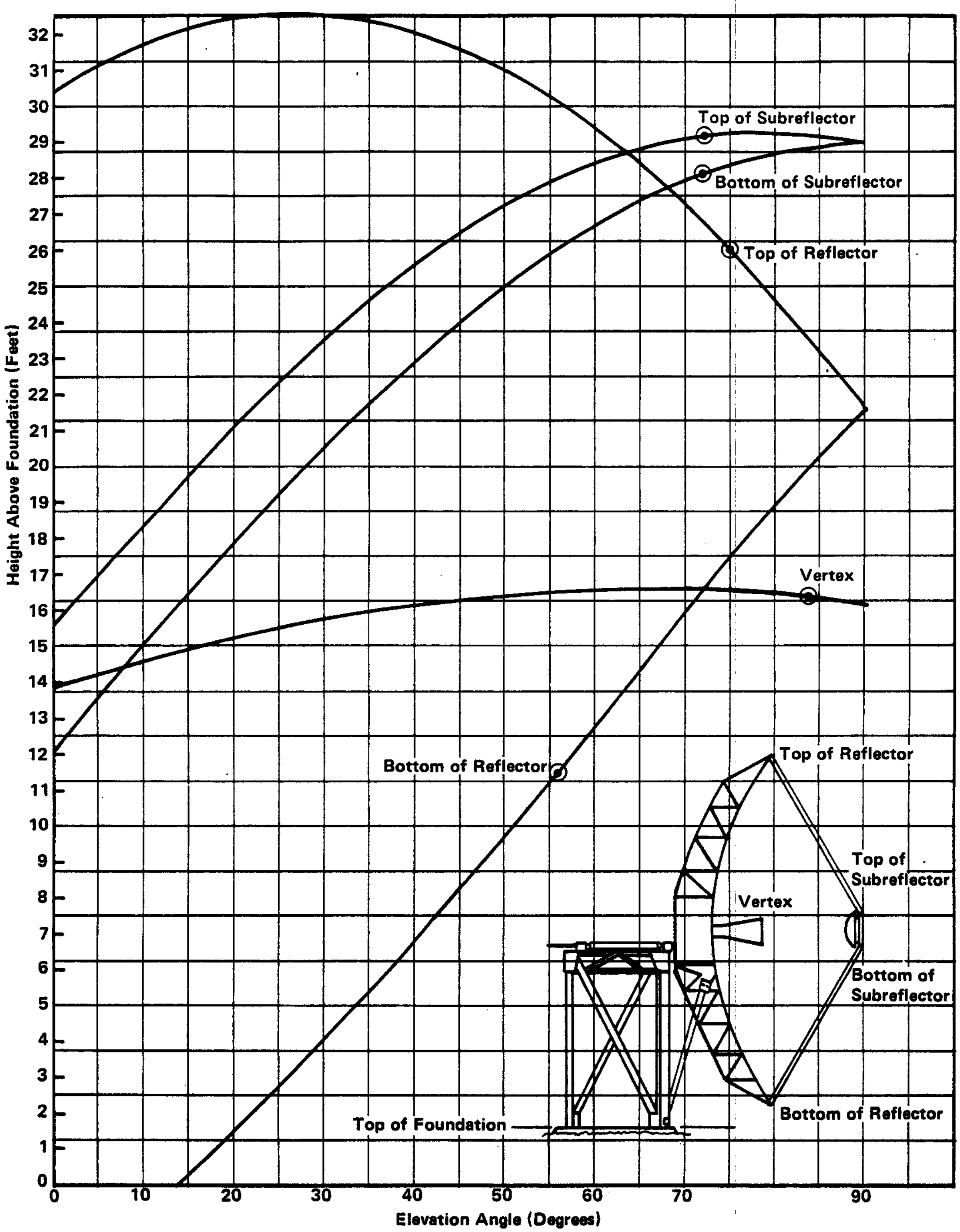


Figure 14. Model 8002 Outline Heights above Foundation vs. Elevation Angle, Short Mount (10 meter)

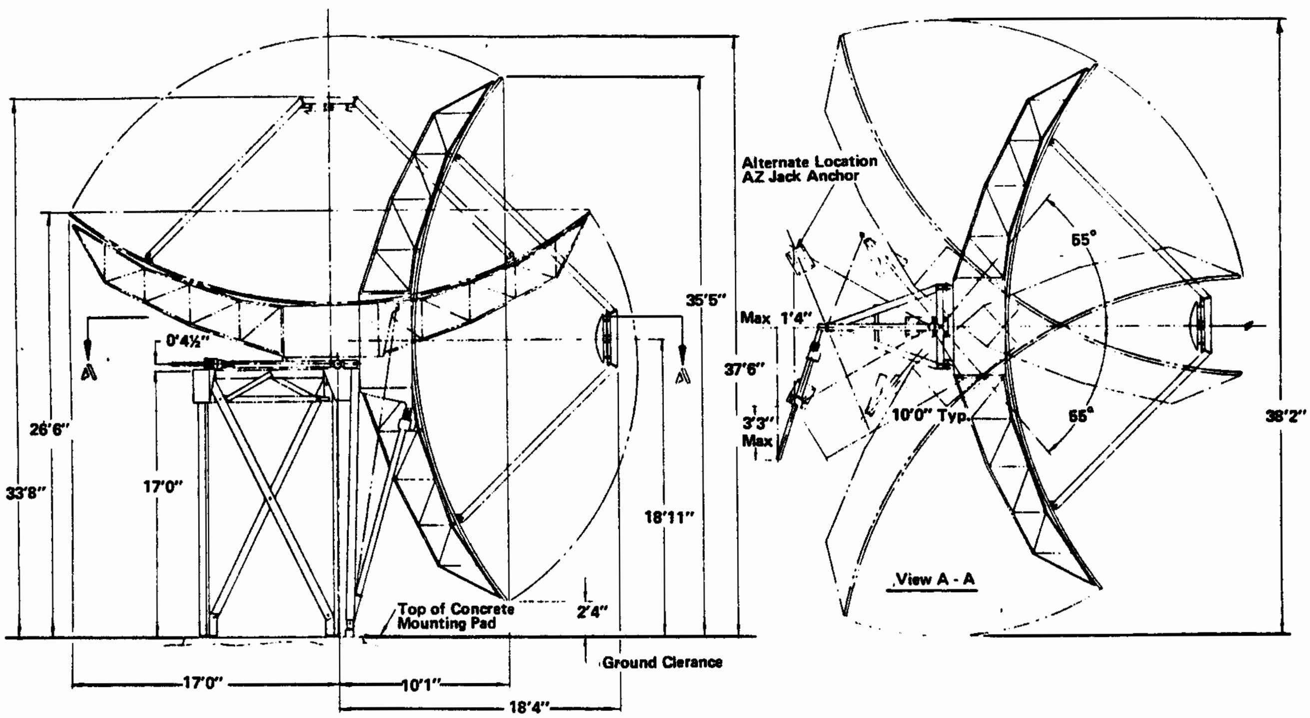


Figure 15. Model 8002 10-Meter Earth Terminal Outline and Basic Dimensions (tall mount)

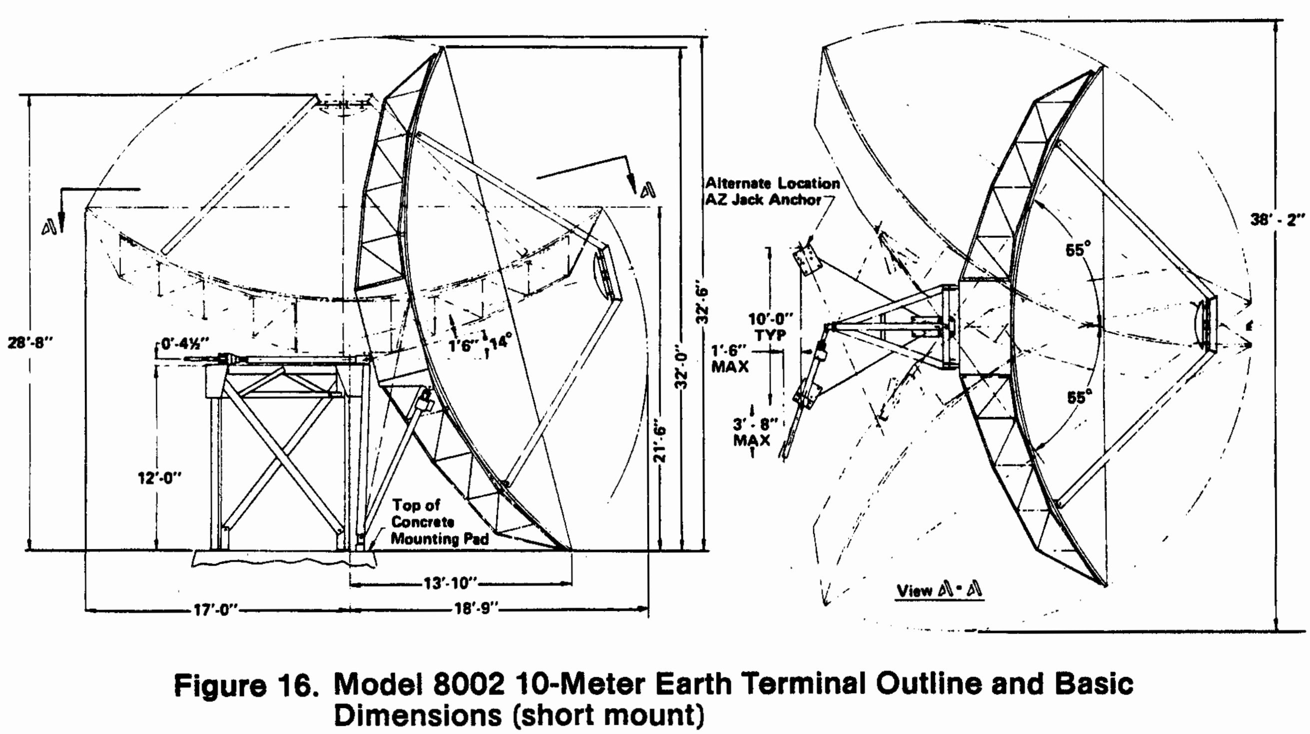


Figure 16. Model 8002 10-Meter Earth Terminal Outline and Basic Dimensions (short mount)

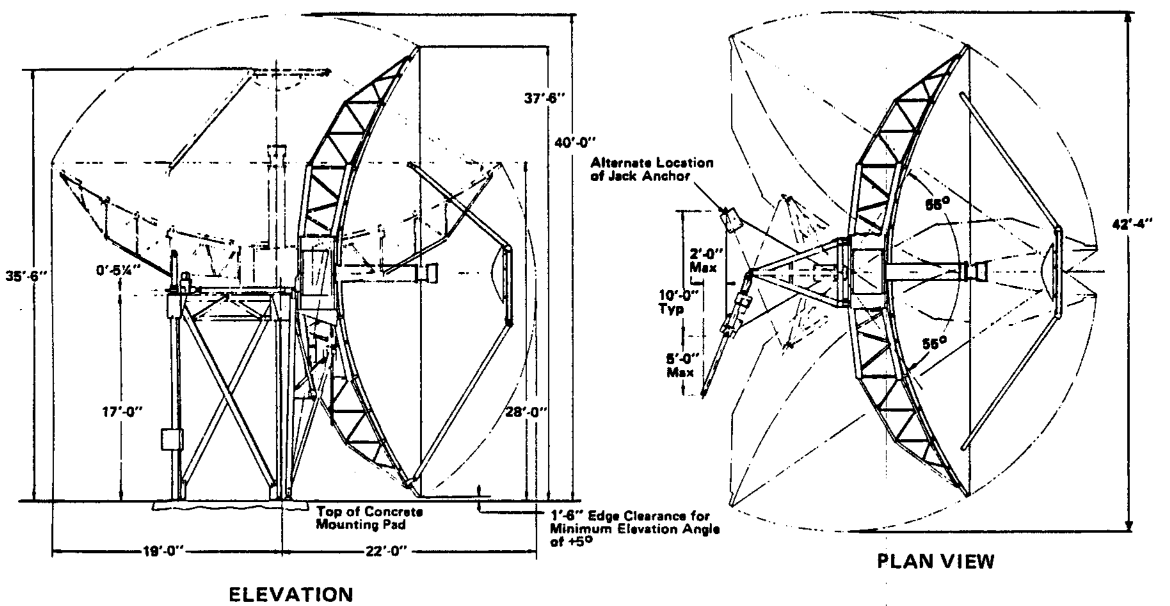
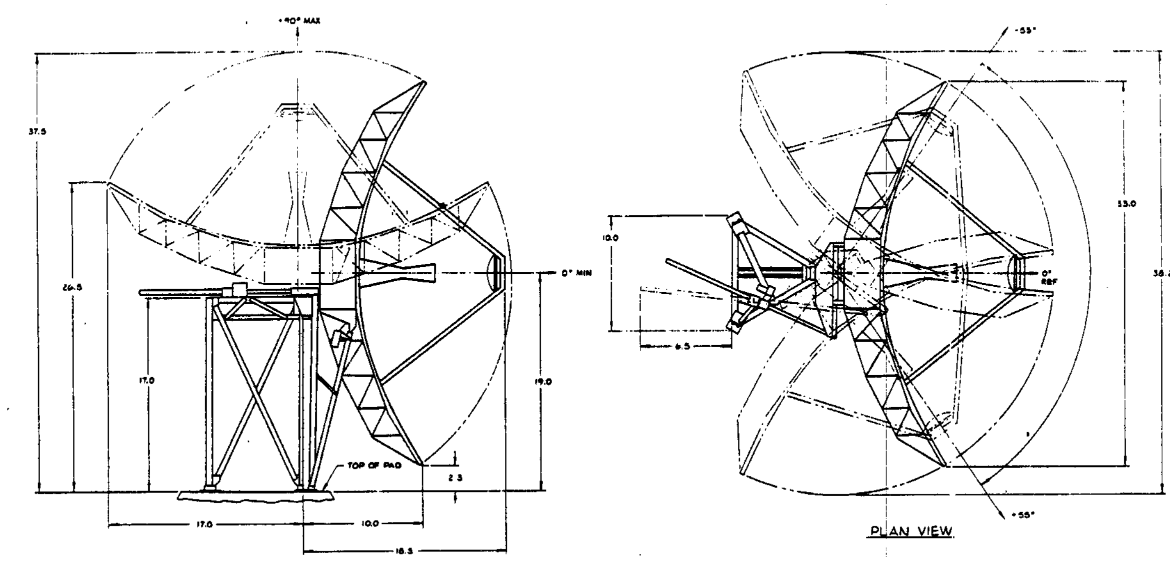


Figure 17. Model 8007 11-Meter Earth Terminal Outline and Basic Dimensions



NOTES:
1. ALL DIMENSIONS ARE IN FEET.

Figure 18. Outline and Basic Dimensions, 8000 110° Series (Tall)— 10-Meter Earth Terminal

High Performance Ku-Band Antennas

Melvin K. Cooker, Jr.
Raymon A. Heaton

Abstract

Two high performance Ku-band antennas have been developed by Scientific-Atlanta as a new line of standard products. The first application of these antennas was for SBS earth stations. Dual-shaped Cassegrain designs were chosen to achieve the very high gain, low sidelobes and precise pointing accuracy required. Measurements and analysis of RF efficiency, reflector surface accuracy and pointing error are presented. Mechanical design and manufacturing techniques which produce precision and interchangeable reflector components in high volumes are described. RF and mechanical tradeoffs are discussed.

Introduction

Satellite Business Systems (SBS), a partnership of IBM, Comsat General, and Aetna Life & Casualty, is a specialized common carrier which offers large U.S. corporations a total communications service via satellite. The SBS network requires unattended Ku-band earth terminals on the customers' premises which can handle high data rate traffic with very high availability during a wide range of severe environmental conditions.

For this system, the earth station antenna must provide very high gain with low sidelobes, precise pointing accuracy and precision remote control. In addition, the antennas must be nearly maintenance-free and designed for high production volumes.

Scientific-Atlanta has supplied antennas to Hughes Aircraft Company, which is producing an initial order of 102 SBS earth terminals. More than twelve months were required to design, develop and test the new state-of-the-art Ku-band earth station antennas.

Two high-performance antenna configurations with motor drives are now in production. A 5.5-meter-diameter version (see Figure 1) is suitable for rooftop installations where size and wind loading are important considerations. A larger, 7.7-meter-diameter reflector (see Figure 2) is preferable for applications in extreme northern or southern latitudes of the continental U.S. where illumination by the satellite will be lower. The 7.7-meter antenna is also applicable in areas that are otherwise difficult to frequency-coordinate.

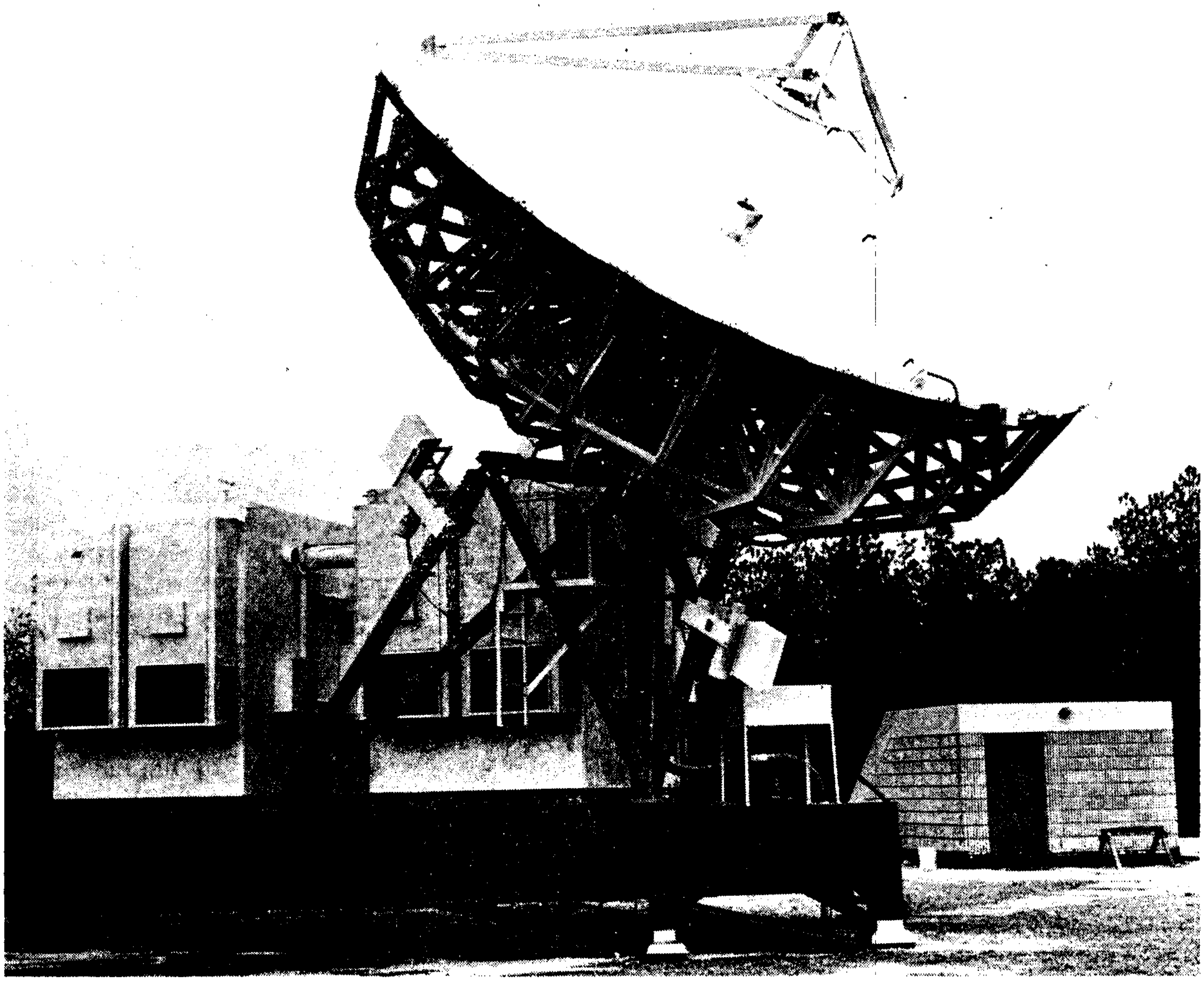


Figure 1. 5.5 Meter Earth Station Antenna

Table 1. Antenna Technical Characteristics

Characteristic	5.5-Meter	7.7-Meter
General		
Antenna Type	Dual-shaped Cassegrain system with corrugated feed horn.	
Antenna Size	5.5-meter-diameter shaped main reflector and 0.58-meter-diameter shaped subreflector	7.7-meter-diameter shaped main reflector and 0.81-meter-diameter shaped subreflector
Reflector Construction	All aluminum 24 precision die stamped panels, truss supports, central hub	All aluminum, 24 faceted panels, truss supports, central hub
Mount Configuration	Elevation-over-azimuth	
Electrical		
Operating Frequency Band		
Transmit	14.0 to 14.5 GHz	
Receive	11.7 to 12.2 GHz	
Antenna Gain (Ref. OMT port)		
Transmit	56.0 dBi at 14.25 GHz	59.0 dBi at 14.25 GHz
Receive	54.8 dBi at 11.95 GHz	57.8 dBi at 11.95 GHz
VSWR (Ref. OMT port)	1.3:1 Max	
Polarization	Linear; transmit-receive orthogonal	
Isolation	35 dB Min (transmit to receive)	
Polarization Adjustment	360°	
Power Capability	3 kW CW	
First Sidelobe Suppression	-15 dB	
Radiation Patterns* Transmit Band and Receive Band	(32-25 log θ) dBi	1° <θ <48°
	-10 dBi	48° <θ <120°
	-15 dBi	120° <θ <180°
Cross-Polarized Suppression	35 dB or better on boresight axis and 30 dB within 3-dB beamwidth	

*Average Sidelobe Envelope (per FCC Reg. 25-209 and CCIR Report 391-2)

Table 1. Antenna Technical Characteristics (continued)

Characteristic	5.5-Meter		7.7-Meter	
Antenna Noise Temperature	Elevation	°K	Elevation	°K
	5°	75	5°	74
	10°	46	10°	45
	15°	34	15°	33
	20°	27	20°	26
	30°	20	30°	19
	40°	17	40°	16
Mechanical				
Main Reflector Surface Accuracy	0.028 inches (0.71 mm) rms normal to surface; measured static (0.25-inch rms effective electrical) 0.035 inches (0.89 mm) rms normal to surface; dynamic, operational environment (0.31-inch rms effective electrical)			
Antenna Pointing Accuracy (per EIA STD. RS-411)	0.029° rms		0.020° rms	
Antenna Pointing Range				
Azimuth	55° in one sector; 87° in two overlapping sectors		64° in one sector; 110° in two overlapping sectors	
Elevation	4.5° to 64.5°		4.5° to 90°	
Antenna Slew Rate	0.01° per second			
Environmental				
Windloads				
Operational	75 km/h gusting to 100 km/h			
Survival, no ice	200 km/h			
Survival with 25 mm radial ice	110 km/h			
Temperature Range				
Operational	-30°C to 50°C (-22°F to 122°F)			
Survival	-40°C to 60°C (-40°F to 140°F)			
Ice				
Operational	6 mm radial			
Survival	25 mm radial			
Solar	1.1 mW/mm^2 (710 mW/in^2)			
Seismic Loadings				
Survival	Mercalli VIII (0.3 g horizontal and vertical with 50 km/h winds gusting to 75 km/h)			

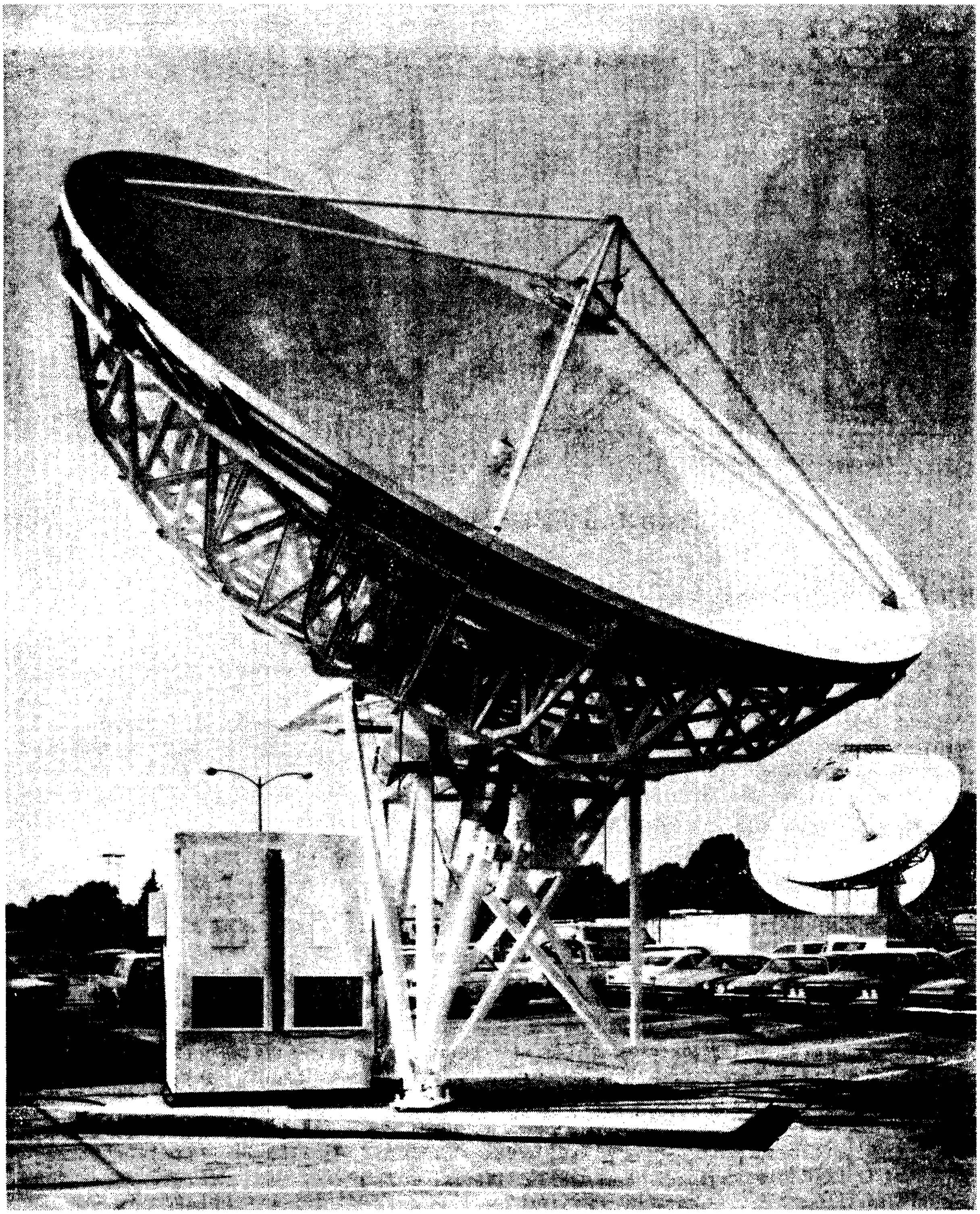


Figure 2. 7.7 Meter Earth Station

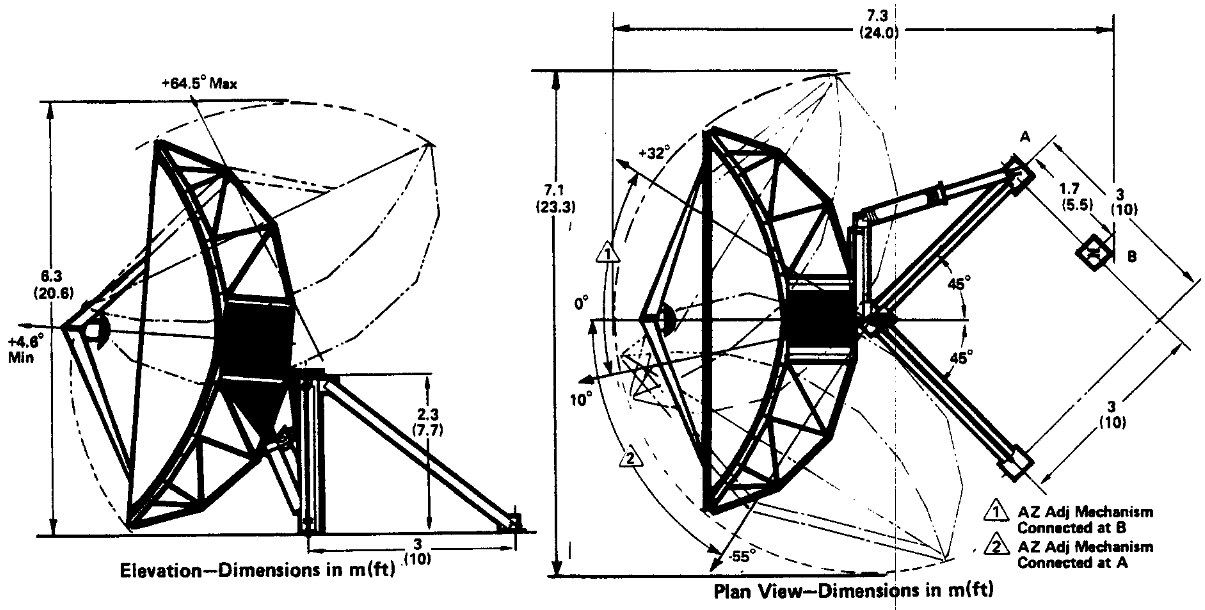


Figure 3. 5.5 Meter Antenna Outline

RF Optics Design

Scientific-Atlanta considered many technical approaches for the 5.5- and 7.7-meter antennas. Among these were offset-fed reflectors, focal-point-fed reflectors, and axisymmetric dual-reflector geometries. Each of these configurations has particular advantages, but the axisymmetric, shaped dual-reflector geometry provides the most flexibility in design and is the most effective compromise.

The offset-fed, dual-reflector approach offers significant sidelobe improvement in the azimuth plane over axisymmetric antennas since this plane does not have blockage from the feed or subreflector. Overshadowing this advantage, however are the disadvantages of high fabrication cost, and the possibility of decreased off-axis cross-polarization discrimination.

The advantages of a focal point design are feed cost and a moderate reduction in blockage and sidelobe levels. However, the design has relatively low aperture efficiency compared to a shaped Cassegrain antenna. Thus, for equivalent gain a larger reflector is required which offsets the lower feed cost. In addition, the focal point feed interface is inconvenient and results in additional loss, particularly for transmit applications.

For this application the dual-shaped, Cassegrain design offered the best combination of performance and cost. Using computer-aided design techniques, the primary feed pattern and shaping of the reflectors was optimized for gain and sidelobe performance.

A corrugated conical horn was chosen as the feed element for its pattern symmetry, cross-polarization characteristics, and nearly constant beamwidth over the operating frequency band. Using a technique developed by Scientific-Atlanta for 4/6-GHz corrugated horns, the horn was impedance-matched to achieve a VSWR less than 1.04:1 in both transmit and receive frequency bands.

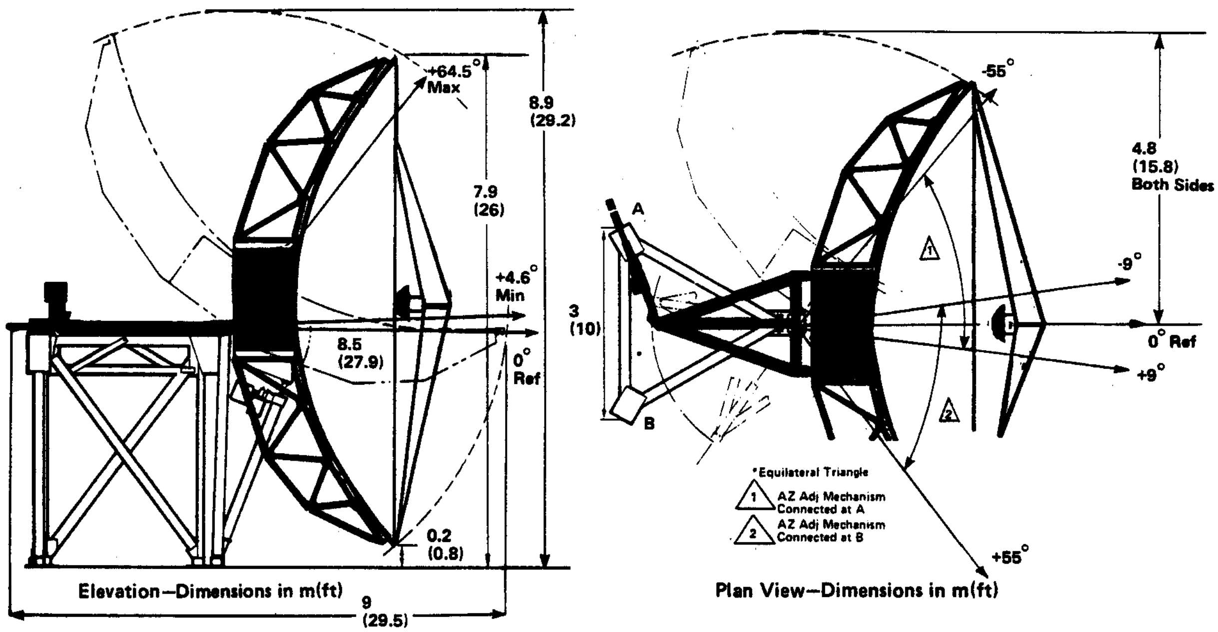


Figure 3. 7.7 Meter Antenna Outline

The precision machined subreflectors for the 5.5-meter and 7.7-meter antennas are 23 inches and 32 inches in diameter, respectively. The relatively large diameter-to-wavelength ratios provide well-behaved scatter patterns which contribute to the excellent gain and sidelobe performance. A three-spar subreflector support provides the rigidity required for dynamic wind conditions while minimizing blockage effects.

Mechanical Considerations

The main reflectors of both antennas consist of 24 panels attached to a central hub and supported by 24 trusses. Aluminum is used for all the reflector components to minimize weight and corrosion and obtain maximum strength and rigidity. The use of a single material with high thermal conductivity also minimizes panel surface distortion due to unequal thermal expansion.

The 0.035-inch rms surface deviation due to dynamic conditions is based on computer analysis and deflections obtained by sand bag loading prototype antenna panels. The results confirm that the reflectors have the mechanical integrity required for both surface accuracy and beam pointing during severe weather conditions.

Achieving high surface accuracy on routine production units is essential to the design. The precision of the antenna hub is key to obtaining surface accuracy. The hub is an aluminum cylinder capped by front and rear diaphragms. The diaphragms are precisely machined for attachment of the 24 trusses.

Reflector panels are held in place by the radial trusses attached to the hub. The trusses are fabricated from special aluminum extrusions and welded in a fixture to maintain overall dimensions. Panel mounting tabs on the truss are then drilled and routed in a precision hard tool to ensure accuracy and interchangeability of the trusses. The truss tabs provide support and proper alignment of the panels.

The 5.5-meter panels are precision die stamped from 0.125-inch aluminum sheet. The edges of each panel formed during the stamping process act as stiffeners to retain the panel's contour. After stamping, the panels are positioned in a drill fixture which precisely locates assembly holes in the panel edge that mate with mounting holes on the truss tabs. The die-stamping production process assures repeatable panel dimensions and complete panel interchangeability. The technique is also very cost-effective for the planned high volume production of the 5.5-meter antenna. A faceted panel fabrication process is used for the 7.7-meter antenna since it is being produced in smaller quantities. Both manufacturing processes used by Scientific-Atlanta produce accurate and interchangeable antenna panels.

The factory production techniques used for the 5.5- and 7.7-meter reflectors eliminate time-consuming panel alignment during field installation. The surface precision is tooled into the interlocking reflector components. In the field a highly accurate surface of revolution is produced by simply assembling the reflector without special tools or highly skilled personnel.

The ease of assembly, surface accuracy, and interchangeability of reflector components were verified during acceptance tests of the antennas. The process involved assembling the 7.7-meter reflector twice and making a surface measurement after each assembly. A manual clinometer bar measurement technique was used to survey 511 points on the reflector surface. RMS surface accuracy was calculated from the clinometer data.

For the first measurement, the panels and trusses of the reflector were arranged without any prior considerations. The measured rms of the configuration was 0.027 inches normal to the surface. The reflector was disassembled, the arrangement of the panels and trusses was randomized and the reflector reassembled for the second rms measurement. The randomized reflector configuration measured 0.028 inches rms normal to the surface. The difference between the two measurement results are statistically insignificant, confirming that reflector components are interchangeable and that an accurate surface is produced without panel alignment.

The same tests were performed on the 5.5-meter reflector. The first surface measurement was 0.020. After disassembly and random reassembly, the reflector surface measured 0.023, again confirming the tooled-in precision of the antenna components.

Very precise antenna pointing accuracy even during severe weather conditions is required for Ku-band operations. The rms pointing accuracy of the 5.5- and 7.7-meter antennas is better than 0.1 beamwidth, 0.029° and 0.020°, respectively.

Computer-aided design and analysis were utilized extensively to obtain the necessary mount and reflector stiffness for the required pointing accuracy. Truss-tab pick-up points for the reflector panels were computer-located to achieve the best stiffness in regions of maximum wind load. Similarly, special truss extrusions were contoured radially from the hub to achieve the best trade-off of weight and stiffness.

Backlash contribution to pointing error also received special attention. Trade-offs of manufacturing tolerance versus ease of field assembly were made. Special rod end caps were designed for the actuator connections and the antenna clevis joints.

In summary, the Scientific-Atlanta 5.5- and 7.7-meter earth station antennas satisfy the demanding Ku-band performance requirements. They are cost-effective, producible in high volume, and are easily assembled in the field without requiring critical alignment procedures. Many of the antennas have been installed and are currently demonstrating performance exceeding the design objectives.

LANDSAT D Tracking Antenna

W. K. Dishman

The LANDSAT-D antenna can receive and autotrack a source transmitting in both S-Band (2.2 to 2.3 GHz) and X-Band (8.0 to 8.4 GHz). The antenna configuration is a 10-meter reflector with a Cassegrain subreflector and feed. A dual-frequency monopulse feed is used which consists of four square X-Band horns surrounded by four diagonal S-Band horns. Normally, the feed receives right-hand circularly-polarized (RHCP) signals, however conversion to left-hand circular polarization (LHCP) or vertical polarization is possible. This system is normally equipped with a 50°K S-Band parametric amplifier and a 90°K X-Band parametric amplifier. System G/T measurements averaged 21 dB/°K at S-Band and 31.5 dB/°K at X-Band. Antenna range tests show that the antenna has excellent sum and difference patterns suitable for LANDSAT-D and similar satellite data acquisition applications.

Description of 10 Meter Antenna

The optics of the antenna design were chosen to optimize the overall performance of the dual-frequency band feed. Illumination efficiency and spillover efficiency design tradeoffs were made to yield acceptable gain, sidelobe and tracking performance. The dimensions of the optics are shown in figure 1.

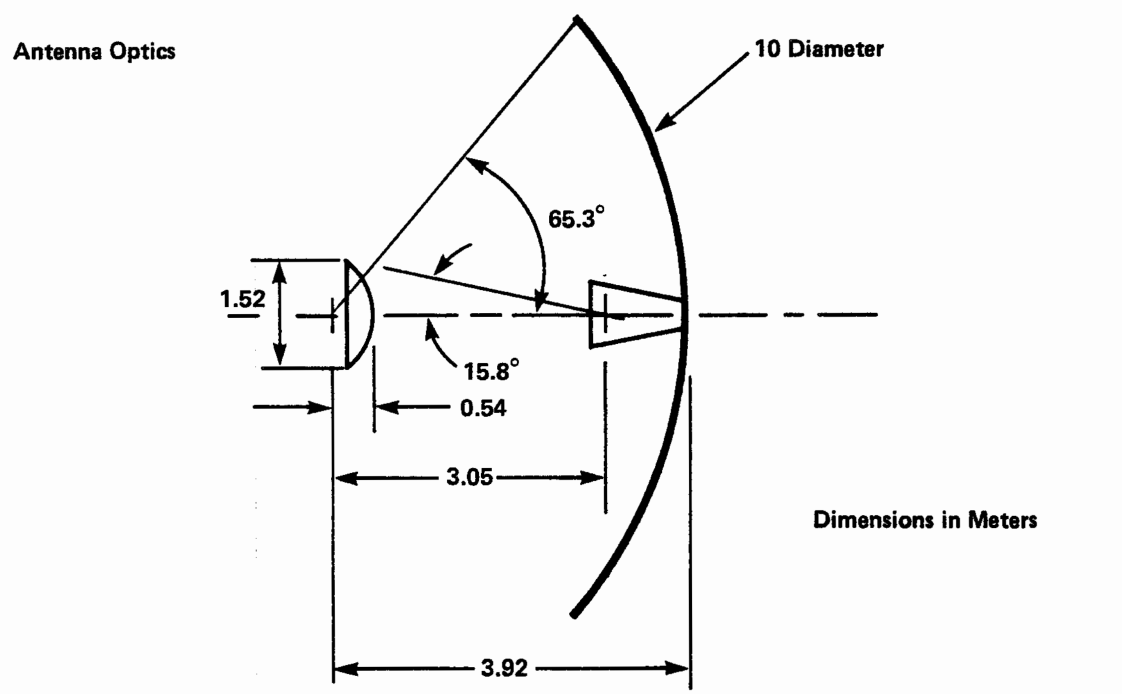


Figure 1. LANDSAT D Antenna Optics

The main reflector is a 10-meter-diameter paraboloidal surface with a focal-to-diameter (F/D) ratio of 0.39. The reflector fabrication techniques are the same as described in the "10- and 11-Meter Antenna" paper.

The subreflector is a 1.524-meter-diameter hyperboloidal surface with an eccentricity of 1.55 and a surface tolerance of 0.18 mm rms. The subreflector is supported by four 7.6 x 12.7 cm ogival cross-section spars which are attached directly to the reflector truss members. This allows the loads from the subreflector to be transferred through the truss members into the central hub without applying these forces to the main reflector surface.

LANDSAT-D Feed

The LANDSAT-D feed assembly, shown in figure 2, consists of four X-Band waveguide horns surrounded by four S-Band horns. The X-Band horns are 5.16 cm square apertures which taper to 2.85 cm square waveguide. A dielectric slab polarizer is placed diagonally across a section of the square waveguide followed by a horizontal resistive card and a square-to-rectangular waveguide transition. The dielectric polarizer converts a RHCP wave to a vertically polarized wave. Any residual horizontally polarized signal that would reflect off the transition is attenuated by the resistive card. The square-to-rectangular waveguide transition is a 4-step Tchebyschef transformer which produces a broadband impedance match. The ouput from each feed is a standard WR112 waveguide flange (UG 138 cover).

The four feed output flanges are in a square pattern and mate to the waveguide monopulse comparator located directly to the rear of the X-Band horns. This comparator provides a sum output by combining the four horn signals in phase and difference outputs by combining signals from each side of boresite 180° out-of-phase in the elevation and the azimuth planes.

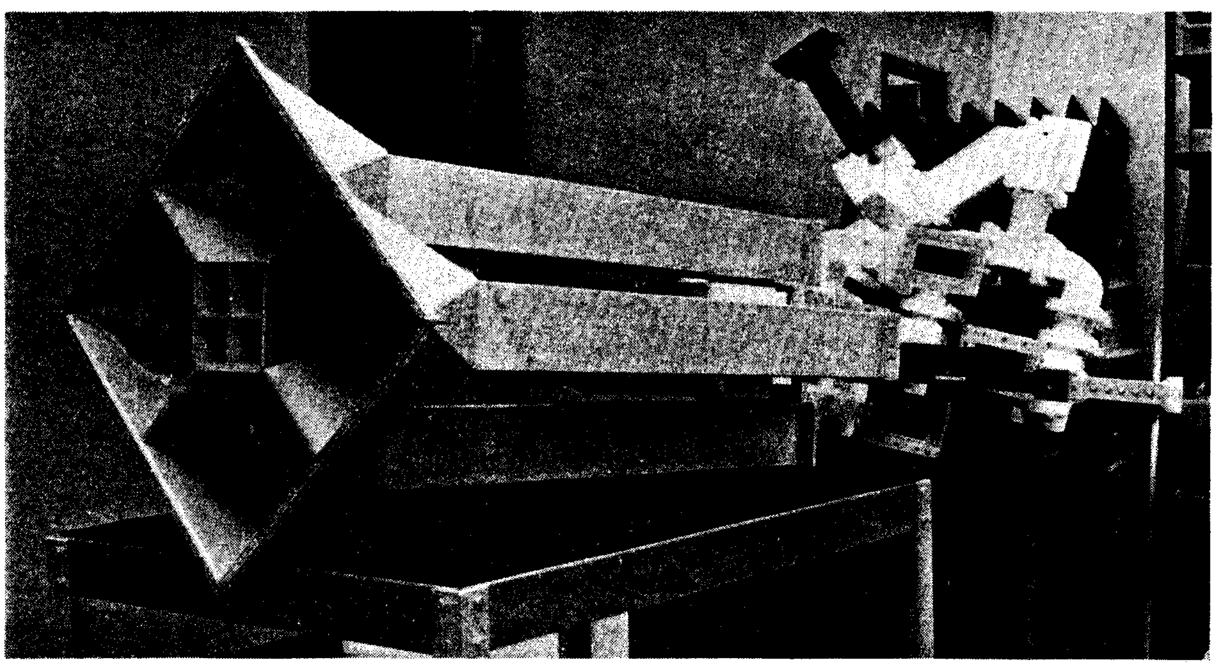


Figure 2.

The S-Band feed horns function identically to the X-Band horns except that they are diagonal waveguide horns arranged in a diamond-configured array instead of a square-configured array. The S-Band apertures are approximately 18.8 cm square, and they taper to a 8.64 cm square waveguide. The X-Band horns occupy only a small corner of the S-Band aperture and therefore do not seriously effect its radiation properties. The polarizers and transitions are identical in function to the X-Band units. A WR340 waveguide comparator is used to combine the diamond-array output ports. This large comparator bolts to the horn outputs which lie in the plane of the reflector vertex, so the comparator occupies the central area of the reflector hub. Enclosures for RF electronics components are located on the hub wall around the comparator.

The feed unit is contained within a square feed can which attaches to the reflector hub. A LUMAR® radome seals the waveguide apertures. Thermostatically controlled heating pads on the S-Band horns prevent ice accumulation on the radome. The feed and components are pressurized with dry air to prevent the ingress of moisture, therefore maintaining overall antenna performance.

RF Electronics

There is a set of functionally identical RF electronics for each frequency band. They produce both a low-noise data channel output and a psuedo-monopulse tracking channel output. Each channel is downconverted to an IF frequency by multiple downconverter units. A feed block diagram is shown in figure 3.

The sum channel output from the monopulse comparator is sent to a parametric amplifier through a 30-dB test coupler and a 5-pole bandpass filter. The test coupler allows injecting a test signal for system evaluation. Amplified signal from the paramp is divided into a data channel and a tracking channel. To minimize losses, most components preceding the amplifiers are waveguide type. Azimuth and elevation difference channel outputs from the monopulse comparator are diplexed into one line by a MONOSCAN® converter. The MONOSCAN® converter allows a reduction in the complexity of the feed by requiring only one receiving channel for both the azimuth and elevation tracking error signals. In addition, the MONOSCAN® converter has a phase shifter section that allows the insertion of a 0° phase shift or a 180° phase shift in the selected channel.

Output signal from the MONOSCAN® converter is amplified by a 35-dB gain GaAs FET preamplifier. This signal is sent through a remote-controlled phase shifter and coupled into the tracking portion of the sum channel through a 10-dB directional coupler.

The sum and difference channels are carefully phase-matched through this directional coupler output. This allows the coupled difference channel to amplitude-modulate the sum channel as the MONOSCAN® converter is switched between the two azimuth/elevation inputs and the two phase states. The scan sequence (AZ 0°, EL 0°, AZ 180°, EL 180°) is generated by the antenna control unit. The effect of the scan sequence produces a small deflection of the tracking channel antenna beam into four positions around the antenna boresite at the scan rate. The antenna control unit utilizes the relative ampli-

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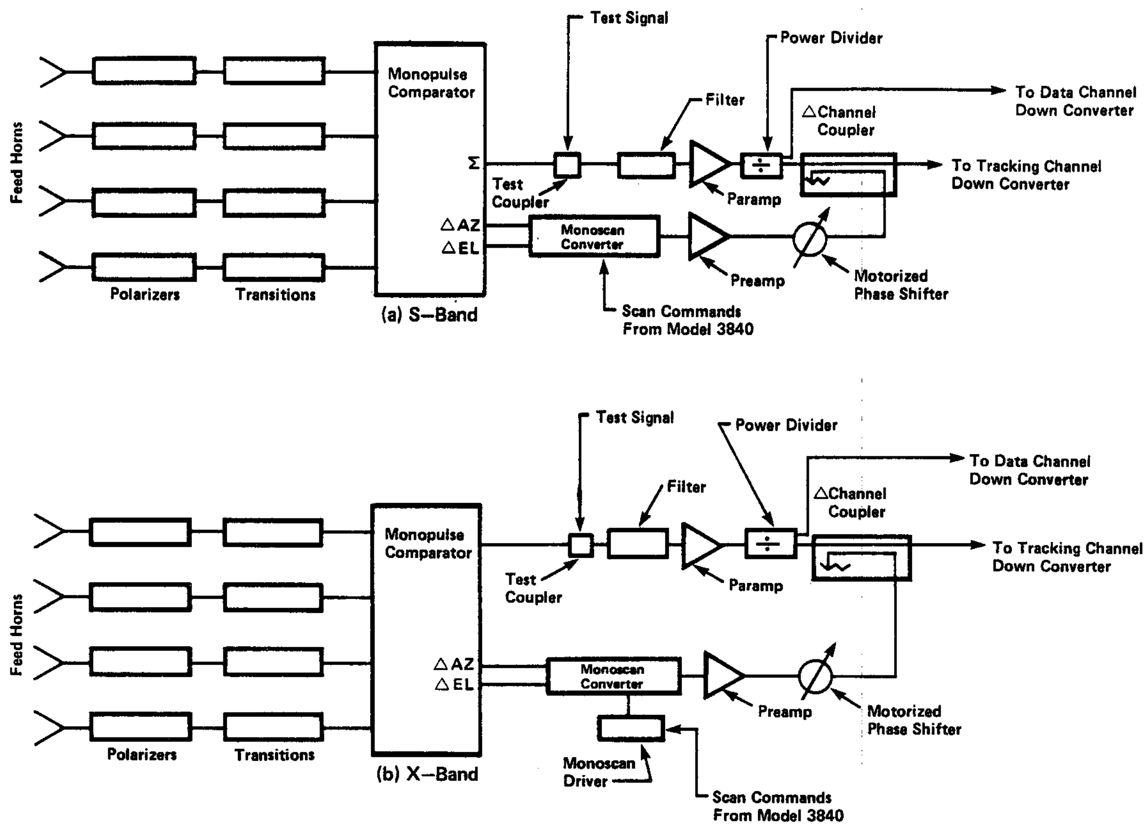


Figure 3. S and X Band Feed Block Diagram

tude information of the four beam positions to reposition the antenna boresite at the signal source (the amplitude difference is nulled by the control unit). The remote-controlled phase shifter is included to allow sum and difference channel phase-match readjustment in case any components change phase length with age.

Antenna Efficiency Calculations and Gain Budget

The antenna aperture illuminations are optimized for S- and X-Band sum patterns. The feed horns and subreflector are designed for an amplitude taper between -10 and -12 dB over the operating frequency bands. Measured feed patterns for mid-frequency S- and X-Band are shown in figures 4 and 5. The subreflector half-angle from the feed phase center is approximately 15.8°. As is well known, a four-element monopulse feed requires uniform excitation of the elements and results in rather high sidelobes in some planes. This yeilds rather high subreflector spillover losses. The subreflector scatter patterns were computed using the Geometrical Theory of Diffraction (GTD) analysis technique; the reflector illumination efficiency was then computed from these patterns. A typical gain budget was tabulated from these computed values for each band (Table 1). The gain budget includes estimates for all expected losses. Average values for measured antenna gain at the comparator sum port are listed for comparison.

Table 1. Gain Budget

Item	Gain/Loss (dB) 2.25 GHz	Gain/Loss (dB) 8.2 GHz
Gain of 100% efficient aperture	47.50	58.50
Feed Efficiency	-3.82	-3.46
Surface tolerance	-0.04	-0.51
VSWR (1.3:1)	-0.07	-0.07
Feed resistive loss	-0.05	-0.05
Polarizer	-0.05	-0.05
Comparator Network	-0.20	-0.20
Gain at the sum port of the monopulse comparator	43.27	54.16
Average measured gain at the sum port of the monopulse comparator	43.9 dBi	55.3 dBi

System G/T Analysis

The G/T analyses for the X- and S-Band systems are shown in tables 2 and 3. The calculations required are described in a previous section of the symposium. The loss and gain numbers are obtained from component specifications and worst-case estimates for manufactured parts.

Table 2. System G/T Analysis — S-Band (5° Elevation, 290°K Ambient Temperature)

Junction	Device	Type	Gain/Loss (dB)	Noise Temperature (K)	System Temperature (K)
1	Antenna	-	43.6	85.0	195
2	Feed	Passive	-.05	3.36	193
3	Polarizer	Passive	-.05	3.36	191
4	Comparator	Passive	-.20	13.67	182
5	Waveguide	Passive	-.10	6.76	178
6	Loop Coupler	Passive	-.03	2.01	177
7	Filter	Passive	-.15	10.19	171
8	Paramp	Active	35.0	55.0	540638
9	Power Divider	Passive	-3.26	324.33	255215
10	Downconverter	Active	20.9	4099.33	3.1×10^7
11	Cable Run	Passive	-14.1	7164.15	1.2×10^6
12	Power Divider	Distribution Amplifier	-3.26	324.33	576642
13	Isolation Amplifier	Distribution Amplifier	0.	2610.0	576642
14	Receiver		0.	5496.0	576642
		System G/T:	20.7 dB/K		

G/T measurements were made using an extraterrestrial source (solar flux) and an injected signal technique. Effort was taken to minimize the sources of error. Reasonable accuracy was achieved at S-Band (±0.4 dB) and X-Band (±0.75). The higher uncertainty at X-Band was due to the flux density uncertainty.

Measured Antenna Performance

The LANDSAT-D antenna performance was measured on a 1500 foot elevated range. Typical patterns are shown in figures 6 through 9 and data is summarized in table 5.

Table 3. System G/T Analysis — X-Band (5° Elevation, 290°K Ambient Temperature)

Junction	Device	Type	Gain/Loss (dB)	Noise Temperature (K)	System Temperature (K)
1	Antenna	-	54.3	80.0	221
2	Feed	Passive	-.05	3.36	218
3	Polarizer	Passive	-.05	3.36	216
4	Comparator	Passive	-.13	8.81	210
5	Waveguide	Passive	-.05	3.36	207
6	Loop Coupler	Passive	-.03	2.01	206
7	Filter	Passive	-.15	10.19	199
8	Paramp	Active	35.0	90.0	628680
9	Power Divider	Passive	-3.16	310.34	303689
10	Downconverter	Active	20.4	4201.57	3.3 x 10
11	Cable Run	Passive	-16.6	12965.56	728500
12	Power Divider	Distribution Amplifier	-3.26	324.33	343898
13	Isolation Amplifier	Distribution Amplifier	0.	2610.0	343898
14	Receiver	Active	0.	5496.0	343898
		System G/T:	31.1 dB/K		

Table 4. G/T Calculations and Measurements

	S-Band	X-Band
Calculated System G/T (Indirect) (Elevation Angle = 5°)	20.7 dB/K	31.1 dB/K
Measured Using Solar Flux Technique	21.23 dB/K	31.26 dB/K
Measured Using Injected Signal Level Technique	20.7 dB/K	32.96 dB/K

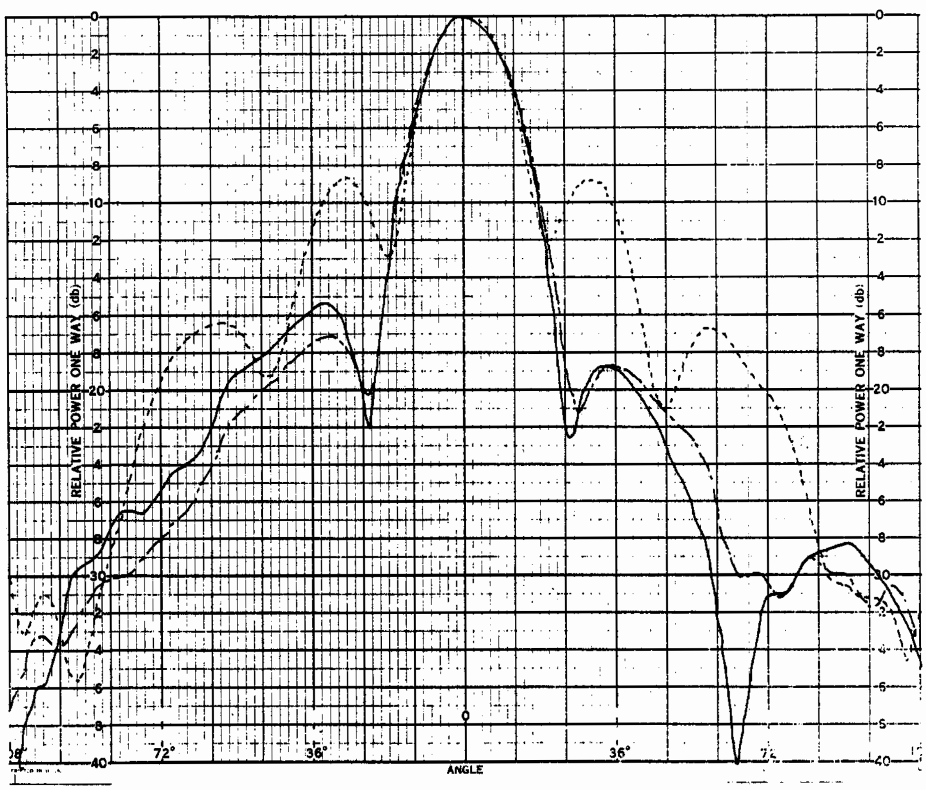


Figure 4. S-Band, 2.25 GHz Sum Channel Feed Pattern

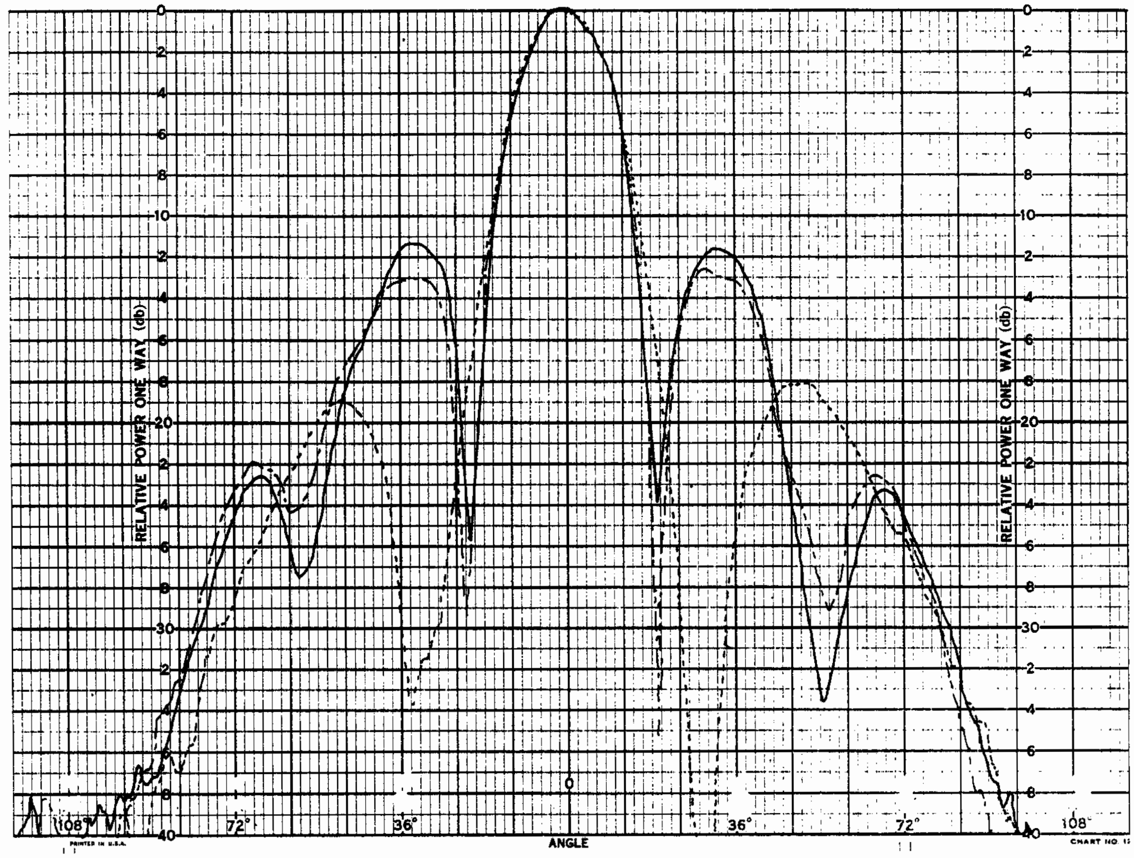


Figure 5. X-Band, 8.20 GHz Sum Channel Feed Pattern

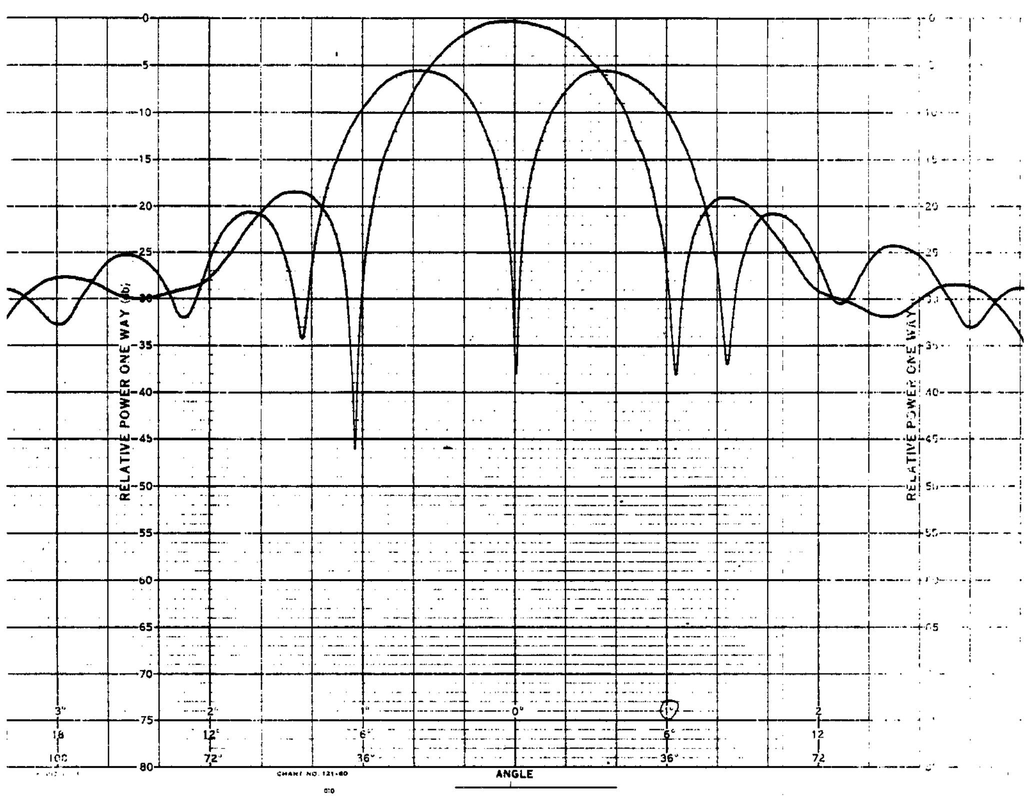


Figure 6. 2.25 GHz Sum and Azimuth Difference Pattern, ±5°

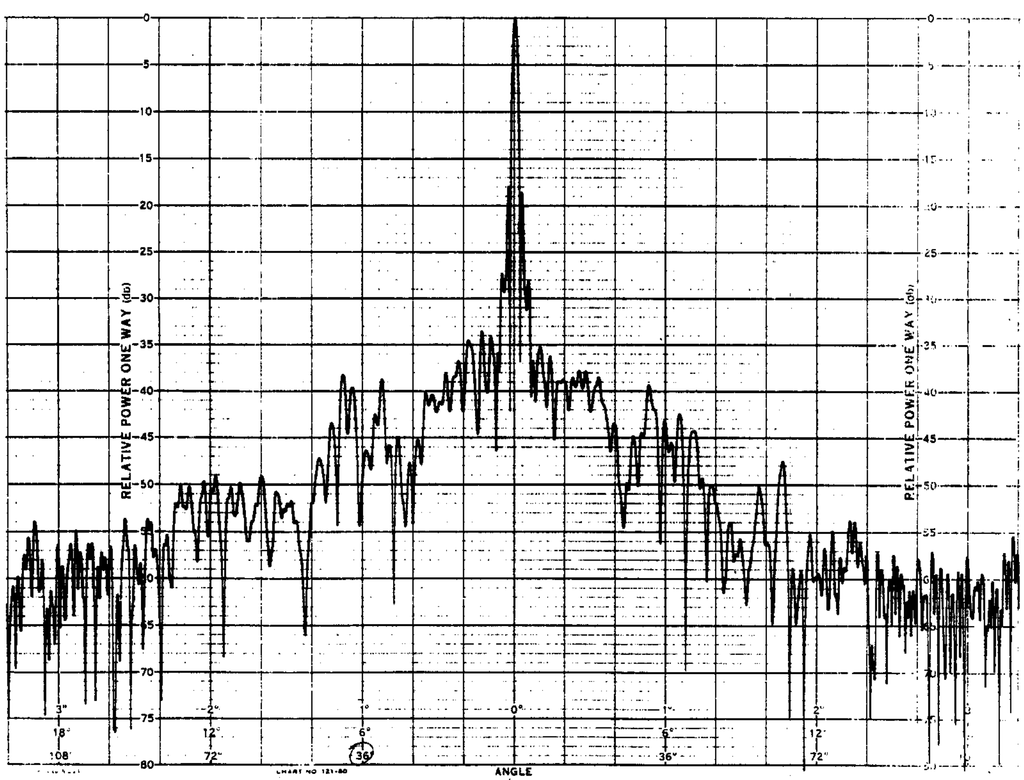


Figure 7. 2.25 GHz Sum and Azimuth Difference Pattern, ±180°

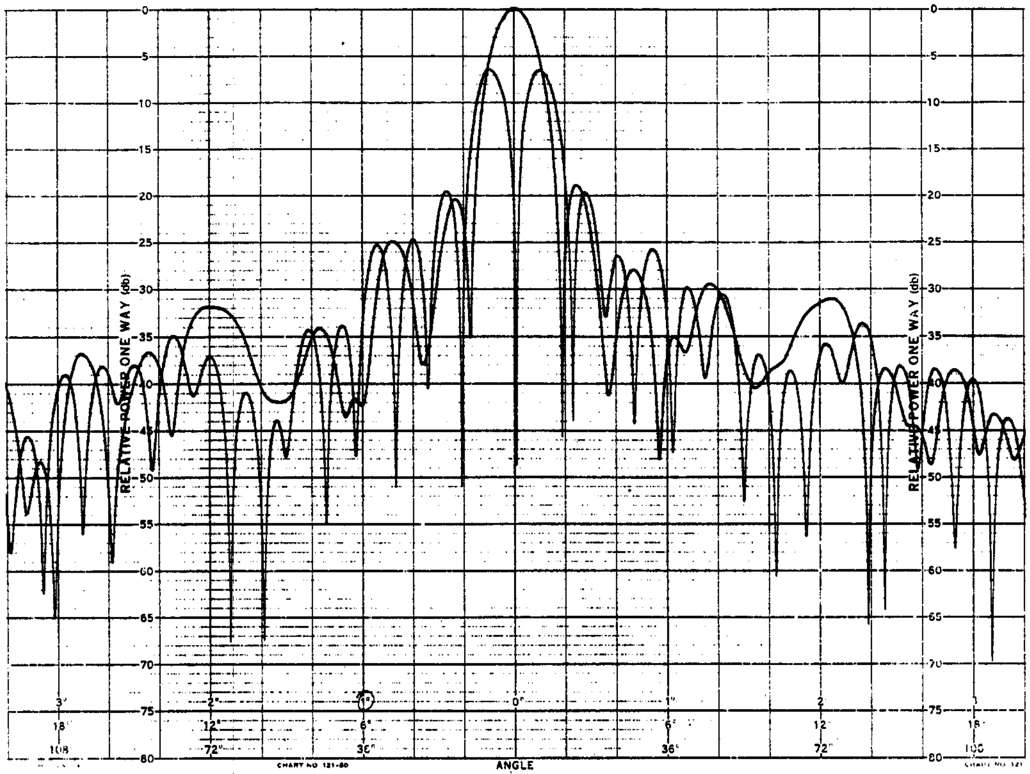


Figure 8. 8.212 GHz Sum and Azimuth Difference Pattern, ±5°

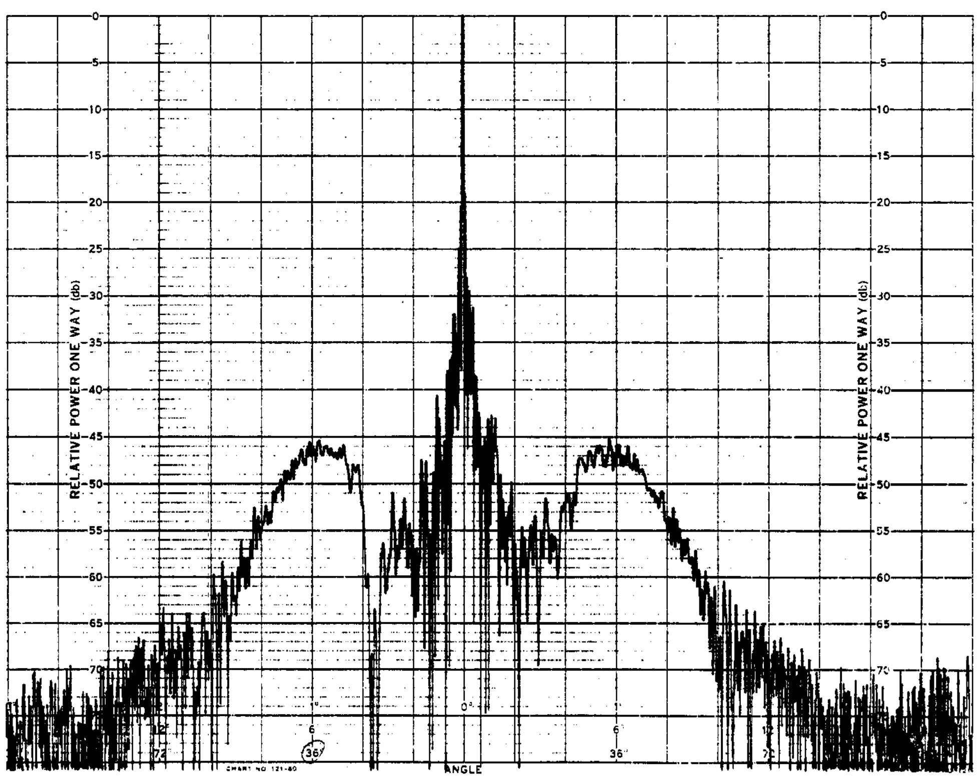


Figure 9. 8.212 GHz Sum and Azimuth Difference Pattern, ±180°

Table 5. Measurement Summary GE/NASA LANDSAT D Antenna Y401

PARAMETER		S-Band			X-Band		
		2.20 GHz	2.25 GHz	2.30 GHz	8.025 GHz	8.213 GHz	8.40 GHz
Gain at comparator output (dBi)		42.52	43.6	43.1	55.8	55.5	54.7
First sidelobe suppression (dB below Σ-peak)	AZ cut	18.0	18.5	18.2	19.5	18.8	20
	EL cut	18.2	18.0	18.8	19.2	18.8	20
Null depth (dB below Σ-peak)	AZ cut	51	42	36	47	47	44
	EL cut	34	37	36	44	43	45
Axial Ratio (dB)		0.2	0.1	0.3	0.6	0.7	0.9
Beamwidth	AZ cut	0.90°	0.89°	0.88°	0.26°	0.24°	0.25°
	EL cut	0.90°	0.91°	0.88°	0.25°	0.24°	0.25°
Σ/Δ misalignment	AZ cut	<0.03°	<0.03°	<0.02°	<0.01°	<0.01°	<0.01°
	EL cut	<0.02°	<0.03°	<0.03°	<0.01°	<0.01°	<0.01°
Max. in-band system VSWR at comparator output		1.25:1			1.20:1		

Note:
Pattern checks of S-Band and X-Band patterns after focus at ∞ show slight improvement in sidelobes at X-Band (-19.2 dB @ 8.212 GHz) and significant improvement of S-Band sidelobes (-21 dB @ 2.25 GHz)

Frequency Reuse Feeds

Scientific-Atlanta has implemented the most advanced technology into the Series 8230 Frequency-Reuse Feeds for applications with earth stations operating in the INTELSAT or domestic communications systems. The feeds can be supplied with linear or circular polarization and are available for use with the Model 8007 11-Meter, the 8002 10-Meter, or the 8010C 7-Meter Antennas. The feed package is compact and can be purchased separately for upgrading existing stations.

Construction and Operation

The heart of the feed (common to all models) is the unique directional junction or diplexer (Figure 1). A circular waveguide section is coupled to a coaxial waveguide by using a corrugated waveguide section between the two. The circular waveguide supports both transmit and receive frequencies in the TE_{11} mode. The coaxial waveguide supports the TE_{11} mode in coaxial guide, and the combination center conducter and 6-GHz waveguide supports the TE_{11} 6-GHz mode. The corrugated section is used to direct the 4-GHz fields into the coaxial guide while also coupling the transmit fields from the 6-GHz guide to the common circular waveguide. When the frequency separation is complete, the 4-GHz fields are coupled to orthogonal waveguide ports attached

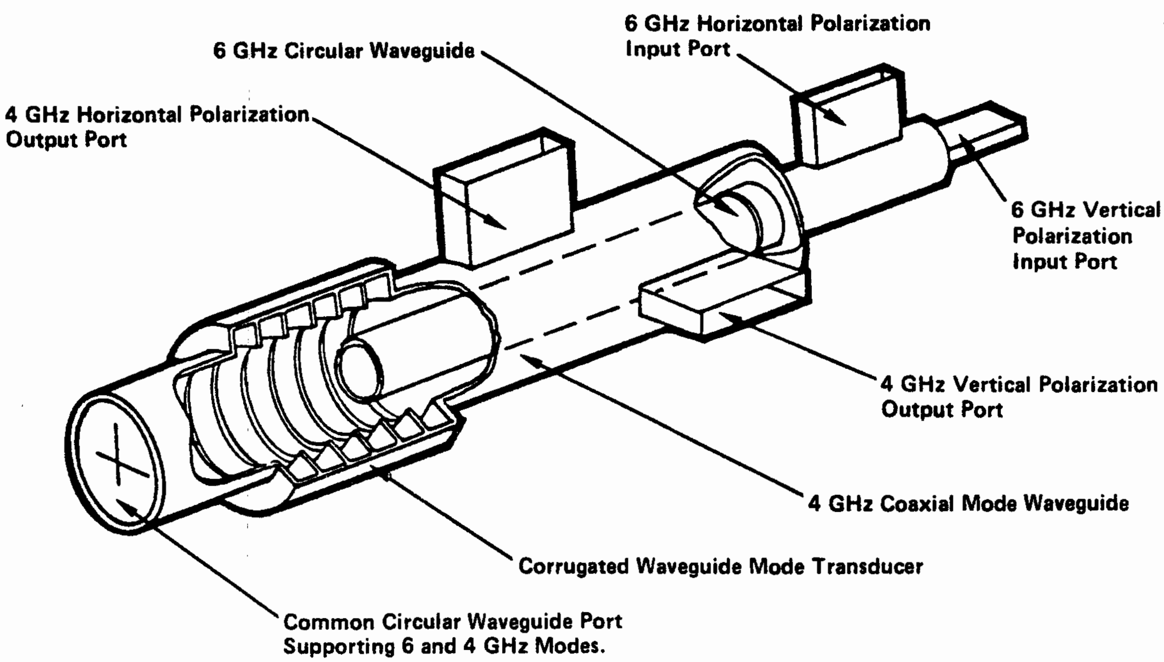


Figure 1. Cutaway View of Frequency Reuse Diplexer

to the outer conductor of the coaxial section. The 6-GHz signals are obtained through an orthomode transducer attached to the central waveguide. Keeping the 4-GHz signals from interfering with the 6-GHz signals is not a problem, since the 6-GHz waveguide is below cut-off at 4 GHz. However, the 6 GHz could be coupled into the 4-GHz ports. Isolation is offered first by the corrugated section itself, and then by 6-GHz band reject filters which are attached to the 4-GHz output ports.

Finally, there are eight inductive posts surrounded by capacitive irises which in combination act as additional filtering to the 6-GHz signals. These posts are also used to support the center waveguide to the outer conductor of the coaxial section.[1],[3],

Model 8230 Feed

The 8230 Feed is the most basic of the reuse feeds and consists of a corrugated horn attached directly to the diplexer unit. A 6-GHz OMT is then attached to the back of the diplexer at the 6-GHz circular waveguide flange. The 4- and 6-GHz waveguide ports are positioned so that they are fixed in-line with each other and the entire feed can be rotated manually or by motorized drive to compensate for the correct polarization angle. Technical characteristics and a typical feed arrangement of the Model 8230 Feed are listed in Table 1 and Figure 2.

Table 1. Model 8230 Reuse Feed Technical Characteristics

Characteristic	Specification
Frequency of Operation	
Receive	3700 to 4200 MHz
Transmit	5925 to 6425 MHz
Voltage Standing Wave Ratio (Antenna)	
Receive	1.3:1 Max
Transmit	1.25:1 Max
Axial Ratio (On-axis)	35 dB Min
Insertion Loss	
Receive	0.20 dB typical, 0.25 dB Max
Transmit	0.20 dB typical, 0.25 dB Max
Isolation	
Receive to Receive	35 dB Min
Transmit to Transmit	35 dB Min
Transmit to Receive	75 dB Min
Power Handling Capability	5 kW CW, 20 kW peak
Polarization Adjustment (Manual or Remote Control)	±90°, receive and transmit polarization rotate together
Radiator Type	Corrugated horn
Pressure (Dry Air)	Pressurizable to 0.5 psig
Feed Interface Flanges	
Receive	CPR-229-G
Transmit	CPR-137-G

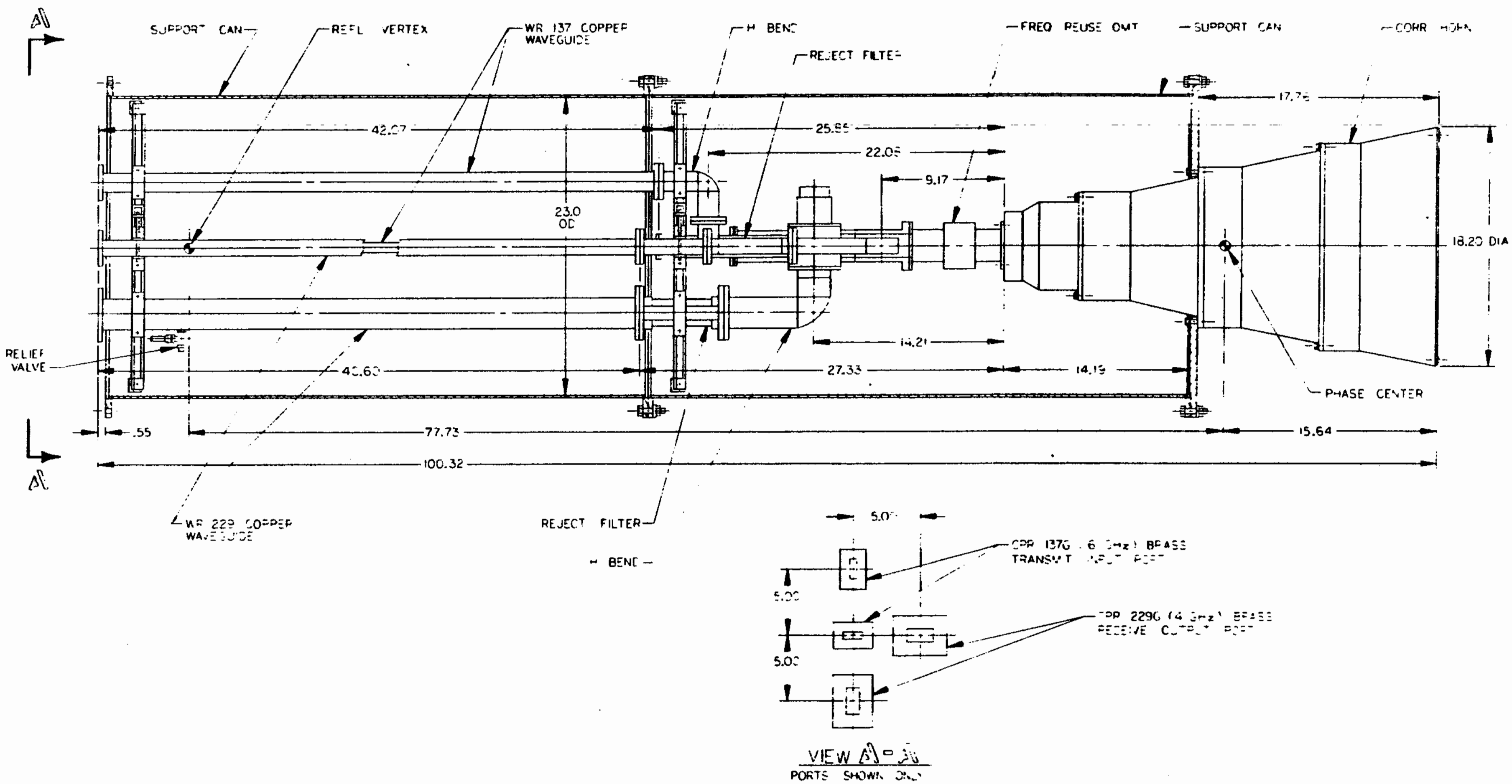


Figure 2. Model 8230 11-Meter Frequency Reuse Feed Outline Drawing

Model 8230C Feed

The circularly polarized version of the four-port frequency reuse feed has been specifically designed for the very low axial ratio requirements of the INTELSAT Global Communication System and other similar circular applications.

The mechanical layout and structural integrity of the feed is similar to its linear co-part with the exception of two polarizing devices. In the 4 GHz coaxial waveguide section, a precision-tuned 90° pin polarizer is added to provide the proper phase shift necessary to create left- and right-hand circular polarization, respectfully, at the orthogonal 4 GHz output ports. Similarly, a 6 GHz 90° pin polarizer is used in the common-round transmit waveguide section to provide left- and right-hand circular polarization transmit when the signal is input at the orthogonal waveguide ports.

As with all Scientific-Atlanta feeds developed for C-Band, standard CPR-229-G and CPR-137G interface flanges are used making field retrofits feasible.

Circular feed versions are available for the 7-meter and 11-meter antennas but can also be implemented easily into other antennas with specific requirements.

Feeds are pressure-sealed, primed, and painted for maximum corrosion resistance in most world environments. (See technical characteristics and outline drawings in Table 2 and Figure 3, respectfully.)

Table 2. Scientific-Atlanta Model 8230C Circularly Polarized Frequency Reuse Feed

Characteristic	Receive	Transmit
Operating Frequency	3.7 to 4.2 GHz	5.925 to 6.425 GHz
VSWR (Antenna)	1.3:1	1.25:1
Axial Ratio (On-axis)	0.5 dB	0.5 dB
Isolation		
Port-Port	20 dB	20 dB
Transmit-Receive		75 dB
Insertion Loss	0.25 dB Max	0.25 dB Max
Power Handling (Each Port)		5 kW cw 20 kW peak
Feed Interface	CPR-229-G	CPR-137G

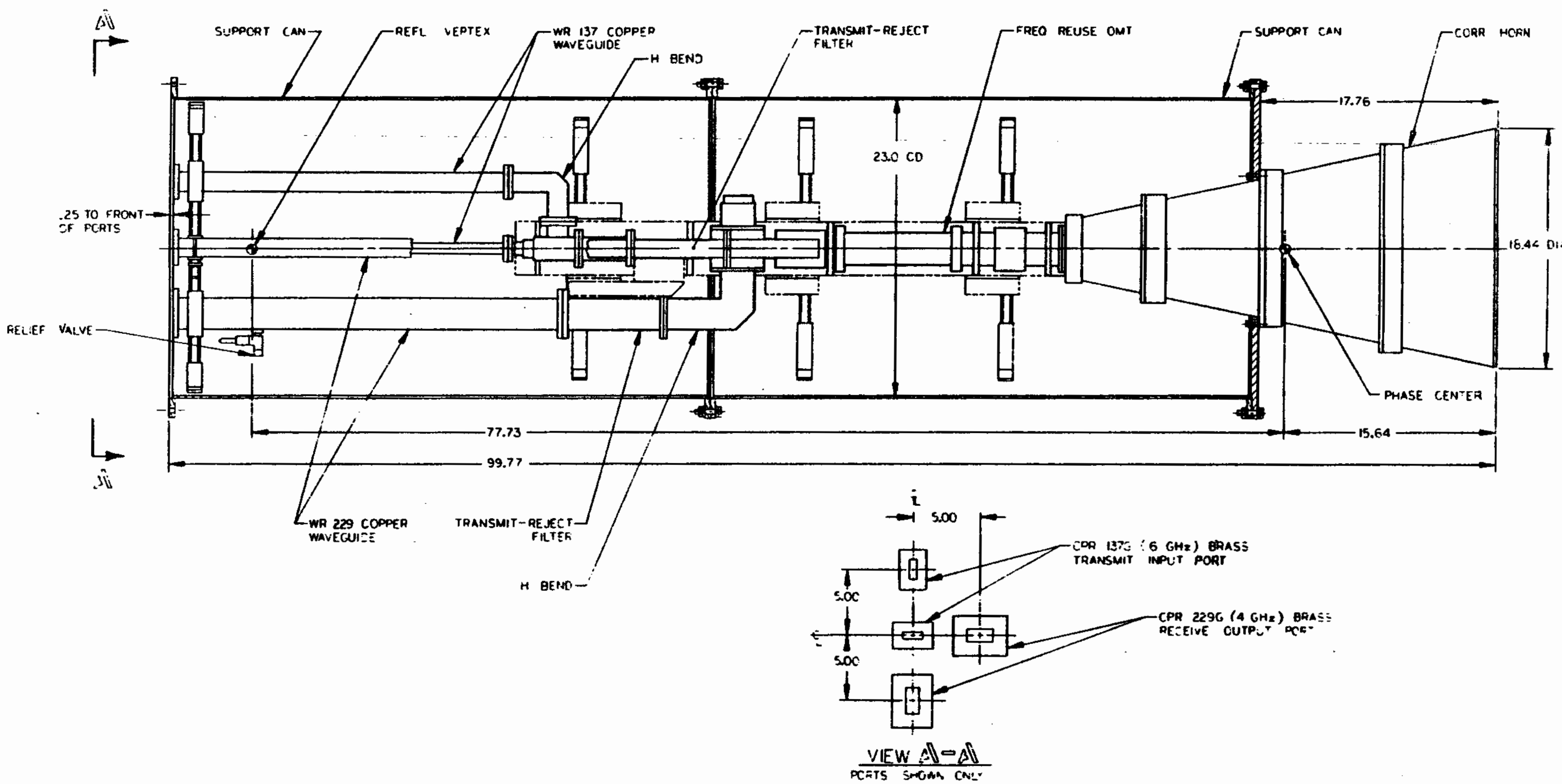


Figure 3. Model 8230C 11-Meter Frequency Reuse Feed Outline Drawing

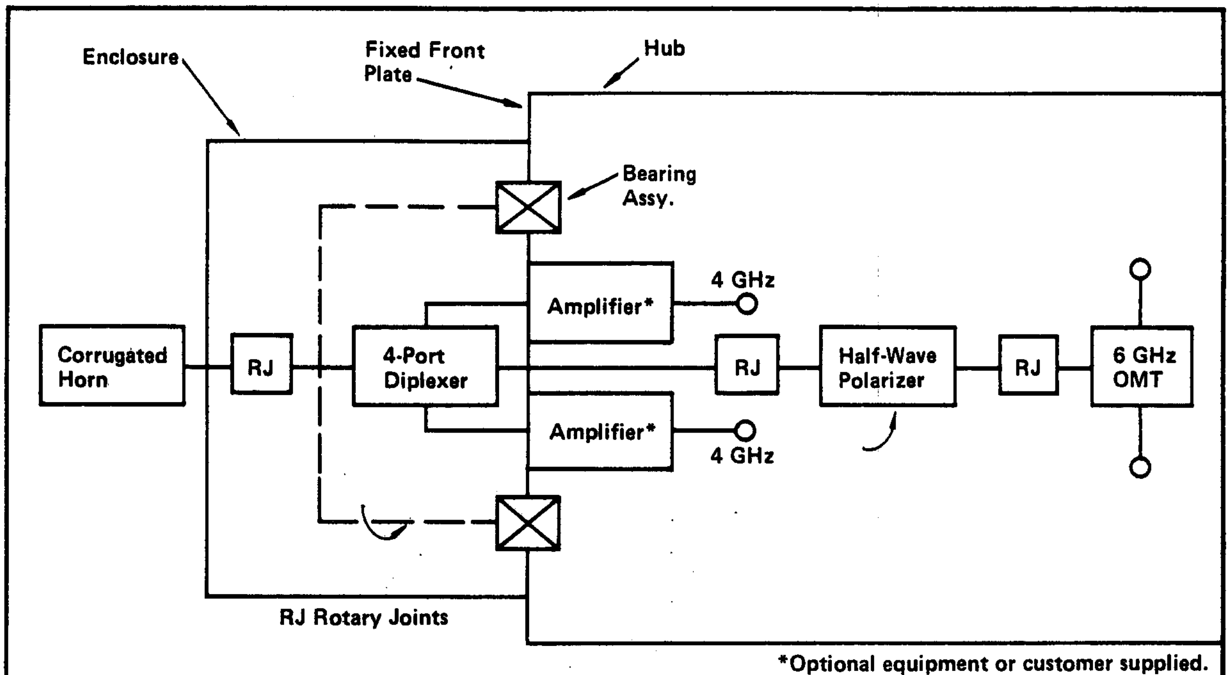


Figure 4. Model 8231 Frequency Reuse Feed Simplified Block Diagram

Model 8231 Feed

The 8231 Feed allows independent rotation of the polarization angle in the transmit and receive ports (Figures 4 and 5). This allows a means of compensation for Faraday effects which rotate the polarization angle of the 4 and 6 GHz in opposite directions. Each feed component is mechanically isolated from the others by rotary joints. A 6-GHz half-wave, linear-polarizer section is added between the diplexer body and the 6-GHz OMT. This component provides 180 degrees more phase shift for one polarization than it does for the orthogonal polarization. When rotated, it effectively matches the polarization angle of the 6-GHz OMT with the incoming polarization angle. The advantage of using this component is that the 6-GHz OMT which is rigidly mounted with waveguide need not be rotated. Also, the half-wave polarizer only needs to be rotated physically 45 degrees to go from horizontal to vertical polarization. Both the diplexer body (4 GHz) and the polarizer (6 GHz) are internally motor-driven and controlled remotely with the Model 8045 Control Unit. Limit switch protection is provided to prevent over-rotation and allows ±90° polarization rotation at both 4 and 6 GHz. Technical characteristics of the Model 8231 Reuse Feed are listed in Table 3.

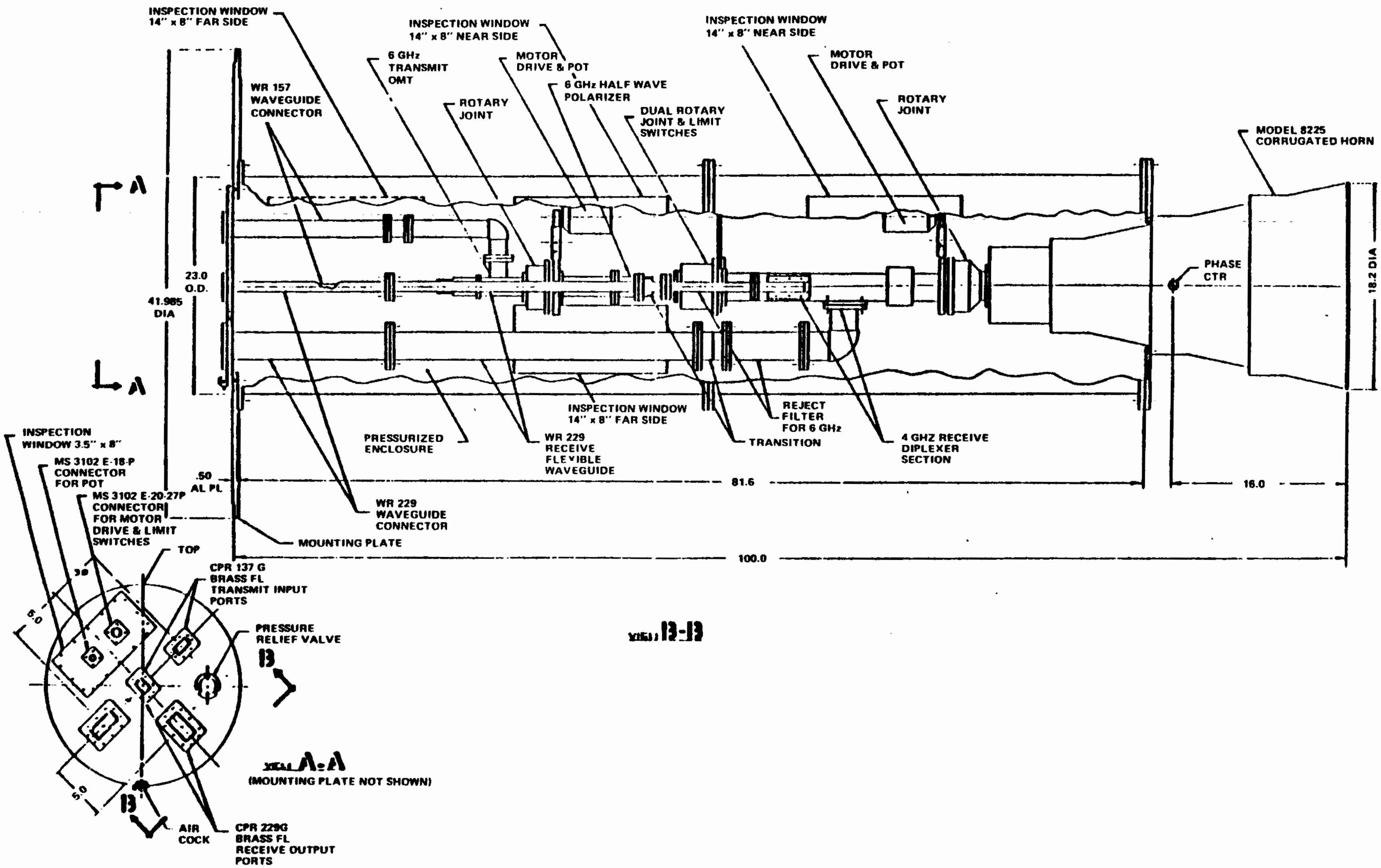


Figure 5. Model 8231 Frequency Reuse Feed Outline Drawing

Table 3. Model 8231 Reuse Feed Technical Characteristics

Characteristic	Specification
Frequency of operation	
Receive	3700 to 4200 MHz
Transmit	5925 to 6425 MHz
Voltage Standing Wave Ratio	
Receive	1.25:1 Max
Transmit	1.20:1 Max
Axial Ratio (On-axis)	35 dB Min
Insertion Loss	
Receive	0.20 dB typical, 0.25 dB Max
Transmit	0.20 dB typical, 0.25 dB Max
Isolation	
Receive to Receive	35 dB Min
Tranmsit to Transmit	35 dB Min
Transmit to Receive	75 dB Min
Power Handling Capability	5 kW CW, 20 kW peak
Polarization Adjustment (Remote controlled)*	±90° in both bands, receive and transmit polarization independently adjustable
Radiator Type	Corrugated horn
Pressure (Dry air)	Pressurizable to 0.5 psig
Waveguide Output (2 each)	
Receive	WR-229
Transmit	WR-137

*The Model 8231 Frequency Reuse Feed includes a Model 8045 Remote Control Unit.

Testing

Each component of every feed is electrically tested before assembly and again after assembly to ensure proper performance. Tests such as VSWR, isolation between ports, cross-polarization isolation, insertion loss and pressurization are all recorded and maintained in permanent files.

References

1. G. Hyde, R.W. Kruetel, Jr., L.V. Smith, "The Unattended Earth Terminal Multiple-Beam System Technical Review, Volume 4, No. 2, Fall 1974.

2. S. Parekh, "Frequency Reuse - Cross-Polarization Discrimination and Feed Components," Earth Station Symposium '78, November 8-10, 1978.

3. R.W. Gruner, "6-Port Directional Orthogonal Mode Transducer Having Corrugated Waveguide Coupling for Transmit/Receiver Isolation," U.S. Patent 3.922.621, June 3, 1974.

Planning a Receive - Only Earth Station

Mike Smith

Introduction

A formal written plan is recommended when undertaking the installation of an earth station. Such a plan aids in identifying all tasks that must be accomplished, and then facilitates accomplishing them in the most effective manner. Parallel activities will then be identified as well as items that must follow sequentially. A good plan will provide an accurate assessment of total project costs as well as estimates of the time it will take to complete the project. Items to be addressed herein are those necessary for the planning process as well as an identification of major tasks that must be performed as part of an earth station project.

The first step in any project is to determine the business purpose of the project. There are many questions to be answered, such as:

- Will the earth station be used for video, voice, or data service? Perhaps all three?
- Will the station be receive-only or transmit/receive?
- Are the current plans for the earth station valid for the foreseeable future, or is it known now that in one or two years additional uses (e.g., receive-only now; transmit later) will be made of the station? If the answer is yes to future uses, plan for it now.
- For cable television video, is pay television being added? Are non-premium video services being added? Perhaps both? At the same time or time phased? Is a subscriber rate increase to be tied to the addition of satellite-delivered signals?

Site Selection

Naturally, the first consideration in selecting an earth station site is locating the required amount of space in a location convenient to the rest of one's operation. For a broadcaster--next to the studio; for a cable television operator--next to the head-end. Several other factors influence what is a good site:

- Will the site be free of interference from existing terrestial microwave systems?
- Is the site already owned or available for purchase or lease?
- If not on existing premises, how close is the proposed site, and is interconnect possible via cable or microwave?
- Is the site accessible in varying weather conditions all hours of the day, etc?

Generally, several desired sites are selected according to the preceding criteria and then checked for frequency coordination to determine which are operationally feasible. Site selection is a local function done primarily by the earth station owner/operator, with suitability confirmed by the frequency coordination process. Note that site elevation over mean terrain is not of particular advantage in earth station operation. Path loss between the satellite and earth station is not significantly reduced by a mountain top site as compared to a lower elevation. Usually, a higher elevation results in increased exposure to interfering terrestrial signals.

Frequency Coordination

Frequency coordination is an outside service that is purchased by the owner/operator of the proposed earth station. The process consists of computer analyses of a data base containing the transmission paths of all terrestrial microwave systems within a given radius. Many times an actual on-site RF measurement of interfering signals is required before a site is cleared as suitable for formal coordination with other carriers, for licensing with the FCC for protection from future interference, and for actual operation.

Performance Standards

At some time early in the project, a technical and economic decision must be made concerning the level of technical performance that must be met by the earth station. Technical performance must be sufficient to provide reliable quality transmission and/or reception of planned services now and in the future. However, it must be pointed out that valuable investment funds can be wasted by designing into an earth station an excessive operating margin that will never be used.

Even though there are many parameters that describe the operation of the earth station and its components, for video receive-only stations the parameters most often considered are the carrier-to-noise ratio (C/N) and signal-to-noise ratio (S/N). For a cable TV operator the carrier-to-noise ratio desired is that which will allow operation at an RF level sufficiently above the FM threshold level of the video receiver so as to deliver impulse noise-free pictures at the required reliability (expressed as a percent of time). The signal-to-noise ratio desired is that which will deliver quality video signals free of noticeable thermal noise. For a broadcaster, tighter RS250B or NTC-7 specifications apply.

Earth Station Configuration

All the preceding considerations have an effect on what equipment will be used in the earth station. These decisions are made either with or without the help of an equipment manufacturer or frequency coordinator. Equipment decisions to be made are:

- LNA with cable interconnect at 4 GHz or LNC (block downconverter) with cable interconnect at UHF.

- Antenna size: 2.8M, 3.2M, 4.6M, 5M, 5.5M, 7M, 7.7M, 10M, 11M (Scientific-Atlanta antenna sizes).
- LNA or LNC gain: 50 dB, 55 dB, 60 dB.
- LNA or LNC noise temperature: 80K, 90K, 100K, 120K.
- Length, size (7/8", 1/2" RG6) and type (foam or air dielectric) of coax between LNA or LNC and receiver.
- For video applications, the number of video receivers needed, as well as a choice between grades of receiver (cable TV grade or broadcast grade).
- Automatic failure protection of LNAs and/or video receivers.
- Remote control of the earth station.
- Any other interface electronics such as RF modulator, microwave transmission equipment, etc.
- This is a good time to be thinking about test equipment needed for maintaining the earth station.

Filing For and Receiving Proper Authorizations

For receive-only earth stations, FCC construction permits are no longer required and FCC licensing is optional. However, it is pointed out that licensing does protect the frequency-coordinated earth station against interference from new terrestrial microwave paths. Therefore, licensing is recommended in order to protect the investment in place and to prevent operating difficulties in the future.

Other concerns are local zoning restrictions and electrical and building permits as appropriate.

Procurement of Equipment

In general, this is done at about the time that results of the frequency coordinations are known and the FCC application (if applicable) is being prepared.

Site Layout

In general, the site layout is determined by the size of the earth station and the amount of available space for the earth station site. The antenna and electronics equipment should be co-located, if possible, to prevent long cable runs. If it is not possible to locate the electronics equipment with the antenna, it is advisable to run larger-diameter RF cable from the LNA or LNC located in the antenna hub to the receiver rack located in the equipment shelter.

For excessively long cable runs, amplifiers must be placed in line with the cable and spaced at regular intervals. Failure to do this will result in receiver performance degradation (G/T). Placement of the amplifiers is extremely critical if this problem is to be avoided.

The antenna should be located such that no objects (trees, buildings, etc.) are positioned in the look angle of the antenna. If the RF path to the satellite arc is not clear, signals will be reduced and cross-polarization alignment may be affected.

Figure 1 shows a typical site layout for a small TV receive-only 5-meter earth station. Generally, the TV R/O electronics include only video receivers, an antenna motor controller, and possibly protection switches. Thus, the equipment shelter size can be kept to a minimum.

If it becomes necessary to operate with more than one satellite simultaneously, two antennas may be necessary. Spacing between the antennas should be sufficient to ensure that one antenna does not block the view of the other antenna. The separation used will depend on the possible look angles of the two antennas. The equipment shelter should be located approximately between the antennas to provide as short a coaxial run as possible.

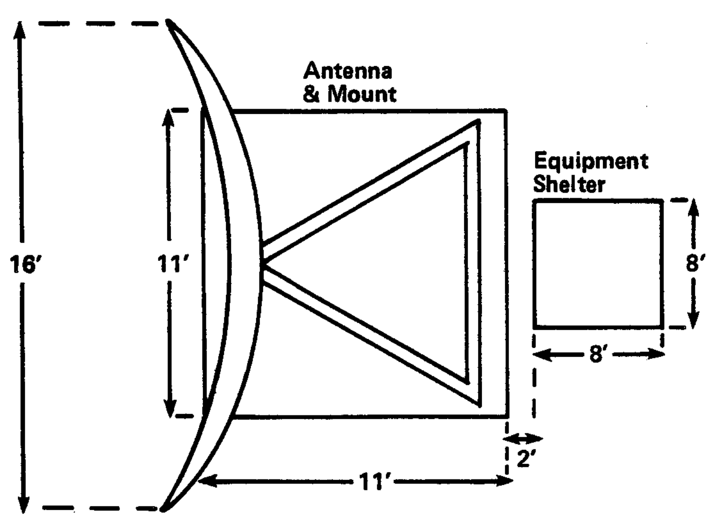


Figure 1. Typical TV R/O Site Layout

Planning a Transmit/Receive Earth Station

J. Hebert

Introduction

The critical areas and design considerations involved in the installation of a transmit/receive earth station must be determined and solutions identified prior to the actual installation. The purpose of this paper is to provide some insight into the design and installation of a transmit/receive earth station. Related topics to be discussed are radiation hazards of the antenna and periodic maintenance of all equipment.

To avoid any confusion between message systems and video systems, only equipment types will be discussed. Any special item pertinent to either system to aid in the installation effort will be noted.

General

The basic transmit/receive earth stations consist of the following subsystems:

- Antenna subsystem
- Low-noise amplifier subsystem
- Downlink subsystem
- Uplink subsystem
- Earth station controller (optional)

Associated with the planning and installation of the transmit/receive earth station are many details, attention to which should provide safe and profitable operation of the earth station for many years. Many considerations associated with receive-only earth stations are applicable to transmit/receive earth stations. Items such as foundation centerline, antenna-motion clearance, low-noise amplifier subsystems, air-dielectric cable, power drives and video receivers are installed and operated in the same manner in both stations. Since these are covered in a separate paper, their discussion will not be repeated here.

Several other items must be considered when planning and installing the transmit/receive earth station that would not normally apply to the receive-only station. The purpose of this paper is to discuss some of the installation details of these items.

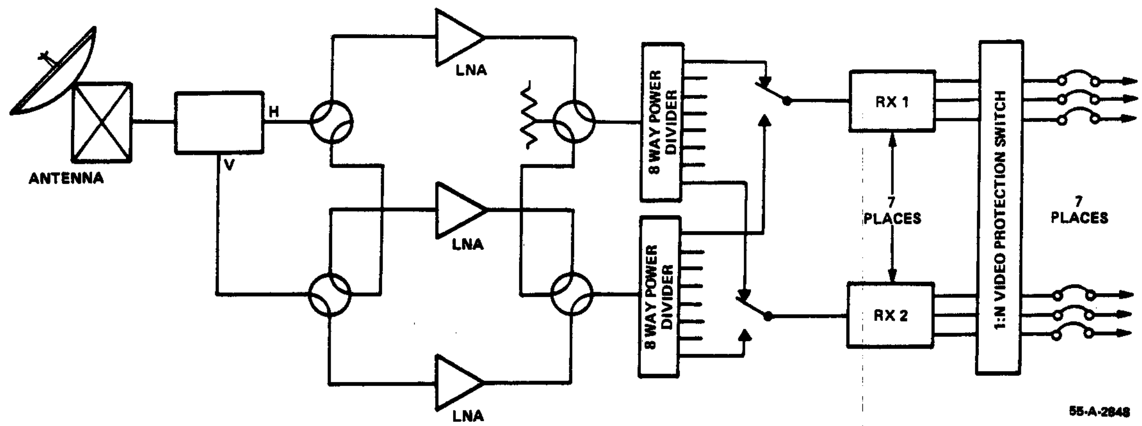


Figure 1. Downlink Subsystem

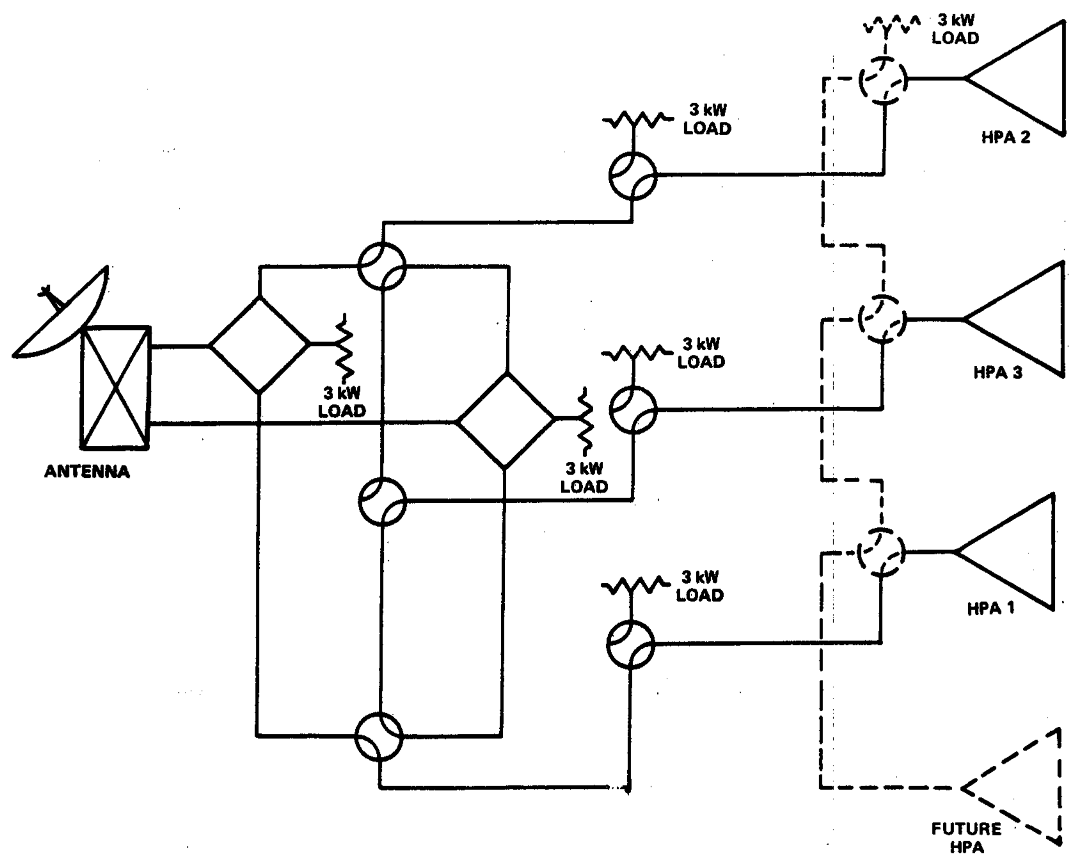


Figure 2. Uplink Switching Subsystem

Frequency Coordination and Licensing

Unlike receive-only earth stations, which may be installed at the owner's risk, uplink stations require complete frequency coordination and FCC licensing because of their potential for interfering with other services. A formal, detailed request must be made to the Federal Communications Commission. The request should include the following:

- Nature of the request and public interest consideration
- Legal, technical and financial qualifications
- Construction proposal and schedule
- Environmental considerations
- Technical proposal
- Applicant's certification
- Frequency coordination and interference analysis report

The interference analysis report should include the following:

- Conclusion of the study
- Summary of the results
- Earth station coordination data
- Certification of the applicant

Other items to be included with the proposal are as follows:

- Recent finance balance sheet
- Description and quotation on proposed earth station
- FCC Form 403 - Application for Radio Station License or Modifications thereof
- List of officers and directors of the applicant
- Copies of the applicant's radio/TV station licenses and permits

Site Selection and Layout

For obvious reason, site selection is an important consideration in the earth station plan. Due to the physical size of the antenna, usually ten or eleven meters, the terrain must be able to withstand the heavy construction equipment needed for the antenna erection, and soil bearing pressure must be adequate to meet the requirement of the antenna.

Access to remote antenna locations during adverse weather conditions is a consideration that may require construction of an access road prior to site construction. Adequate area surrounding the antenna must be available for antenna motion. The equipment shelter must be located at a point where it does not hamper antenna motion and where the transmit waveguide length can be kept to a minimum.

Design of the antenna foundation is covered in detail in the following paper, "Antenna Foundations for a Satellite Earth Station."

The foundation should be designed and approved by a local engineering firm before the date of installation.

The antenna foundation can be installed and allowed to set before the antenna parts are received. To allow adequate time for installation and curing of the foundation, Scientific-Atlanta will ship antenna anchor bolts and template, foundation drawings, and foundation centerline heading information to customers. If an equipment shelter is planned for the site, the shelter foundation should be poured at this time according to plans agreed upon by the customer and Scientific-Atlanta.

Power Requirements

The ac power requirements of a transmit/receive earth station are substantially greater than the requirements of a receive-only station. Typical ac requirements for a receive-only station include video receivers, low-noise amplifiers, compressor-dehydrator and control. The total power required to operate this equipment is less than 10 kW.

The major power consumer of a transmit/receive station is the high-power amplifier (HPA). Video uplink stations typically have one or more 3 kW HPA's, each requiring 12 kW of three-phase, ac power. Message earth station typically use HPA's in the 125 to 400 watt range, requiring up to 3 kW of single phase power. Most transmit stations are configured for automatic protection with a hot standby HPA protecting one or more on-line units. The 3 kW HPA's require 208V ac, 3-phase, 4-wire primary power with ±10% line voltage regulation and phase imbalance less than 2%.

Another large power consumer, although used seasonally, is the antenna deicing equipment. Feed/subreflector deicing requires up to 3 kW. Half-reflector deicing requires up to 27 kW of power. Full reflector deicing requires up to 51 kW. In areas where severe icing or snow does not occur, feed/ subreflector deicing is usually sufficient.

Additional uplinks add approximately 13 kW each, including power required for the HPA, exciter and additional switching equipment. Other options such as high-speed antenna drives and auto or steptracking on the antenna drives add further to the power requirements, and must be considered in planning the primary power subsystem.

Air conditioning is usually needed in a transmit station. This may require substantial power at southern sites.

In applications where prolonged service interruptions due to primary power failure cannot be tolerated, an engine-generator set is provided. These sets are usually diesel or gas (natural or bottled) powered. Since the duration of the outage may range from minutes to days, the engine-generator must be rated to carry the entire earth station load, including air-conditioning and deicing (though not necessarily both simultaneously) and antenna drives. Fuel capacity must be adequate to allow continuous operation during any period when bad weather or other conditions may delay replenishment.

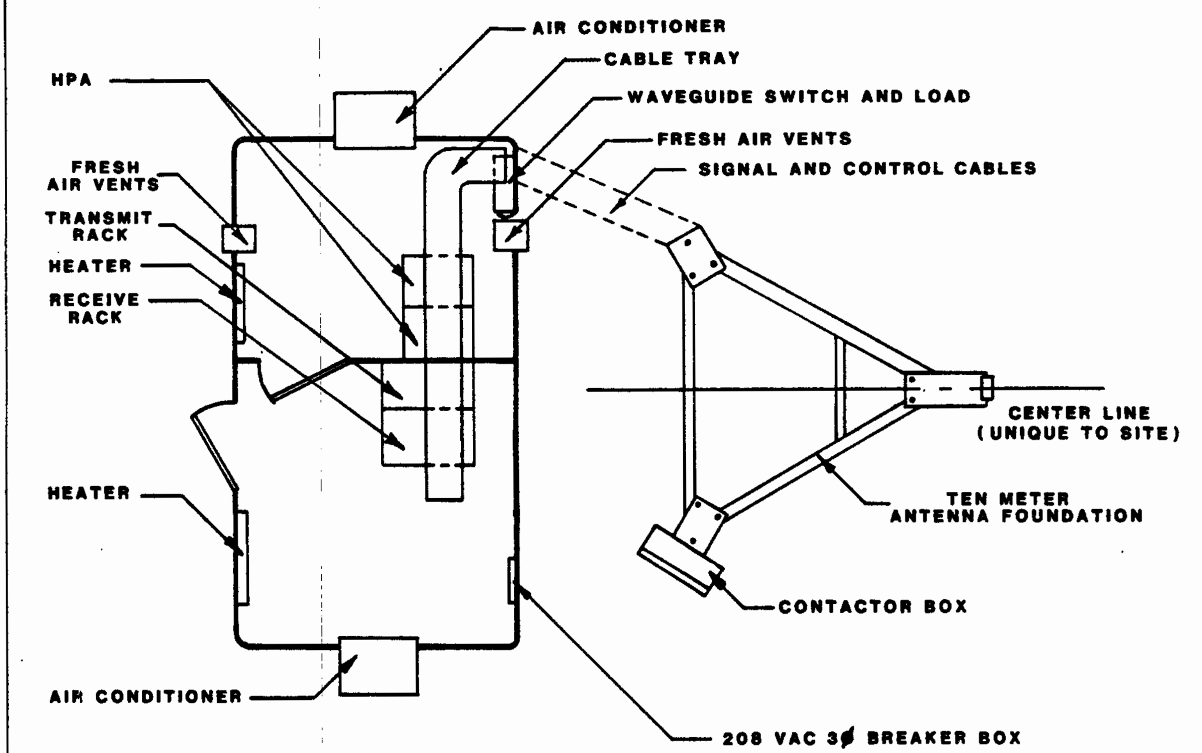


Figure 3. Typical Earth Station Layout

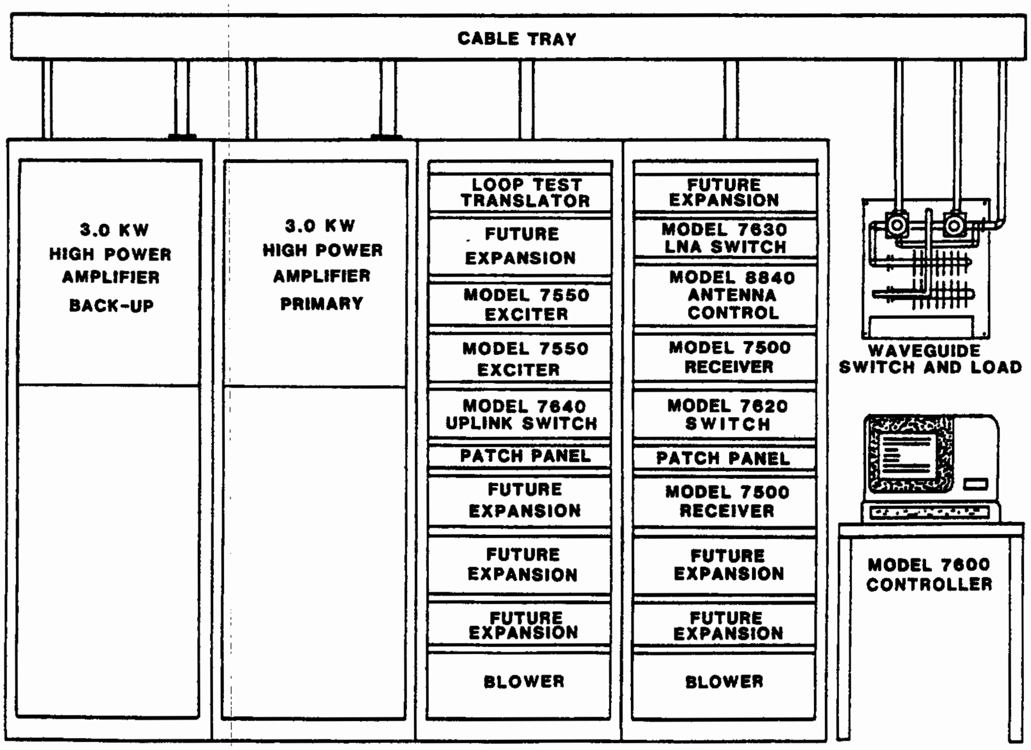


Figure 4. Typical Ground Control Equipment Redundant Transmit/Receive Television Earth Station

Applications requiring continouous operation during even brief power interruptions will require an uninterruptible power system (UPS) consisting of storage batteries, a charger, and an inverter to produce the required ac power. UPS capcity is governed by electrical load and required operating time. To minimize cost, the capacity should be limited to carrying only the critical loads, i.e., electronics, through a period long enough for the engine-generator to start and take the load. A time of five minutes is usually sufficient.

A summary table of typical power requirements is included as Appendix A.

Shelter Requirements

The equipment space needed for a complete transmit/receive earth station is much greater than that for a receive-only station. Each 3.35 kW HPA requires shelter space that is approximately 28 inches wide by 32 inches deep by 78 inches high. For the typical earth station with redundant up/downlinks and options such as motorized antenna positioning and deicing, two standard 19-inch equipment racks are normally required to accommodate the GCE and control electronics. Additionally, space for waveguide runs and switching must be provided in the shelter. A typical shelter layout is shown in Figure 3, with a typical rack installation shown in Figure 4.

The above size requirements suggest that the minimum shelter size for a redundant transmit/receive earth station is approximately 10 feet by 18 feet. Smaller sizes may be used, but they are quite cramped, especially when testing and maintenance of the earth station are in progress.

In the same manner as an equipment room layout is developed before an installation, a shelter layout must be developed and submitted to the shelter manufacturer. Manufacturers of shelters usually offer standard shelter configurations, but some customization is usually required. Some items to consider when laying out equipment in a shelter are door opening and location, cable tray placement, power panel location, waveguide penetration and waveguide switch location, air-conditioning and venting, HPA exhaust and air intake panels, lighting type, ac outlet placement and various security features such as protection and alarm for fire and entry.

If the exciters are to be housed separately from the HPA's, the distance between HPA and exciter should be kept to a minimum to avoide excessive RF losses. If the distance is over 30 feet, it may be necessary to add additional amplification to the exciter output.

Transmit Waveguide and RF Switching

Other major components used in the transmit/receive earth station are the transmit waveguide and the RF switching matrix. Both are unique to transmit stations and will be discussed separately.

The transmit waveguide is a low-loss transmission path or line connecting the output of the HPA to the transmit port of the antenna. The attenuation of the transmission path should be less than 1 dB. The 1 dB requirement specifies that overall length and number of connectors or switches be kept to a minimum.

When designed properly, the transmit waveguide system should fit together exactly with little or no air leaks. If a system is poorly designed, it will be a plumbing nightmare. Normally, rigid waveguide is used between the HPA and RF switch matrix. Proper planning should guarantee precise connections. From the switch matrix, a short piece of twistable-flexible waveguide is used for connection to the semi-rigid elliptical waveguide run out to a point on the antenna. Another twistable-flexible waveguide section is used to connect the elliptical waveguide to the antenna port. This should provide flexibility to allow for any rotation in the polarization or antenna motion (azimuth or elevation) without damaging the waveguide.

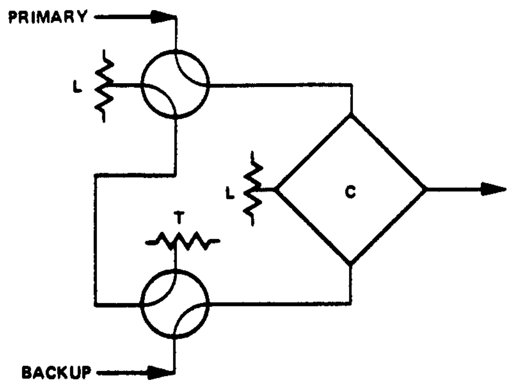
The waveguide is firmly secured using various techniques. Inside the shelter or equipment room it is either braced to a cable tray or suspended from the ceiling using cable hangers. From the equipment area to the antenna, either a cable tray or large diameter conduit is used. The minimum bend radius of the elliptical waveguide is 12 inches in the E-plane and 32 inches in the H-plane. If conduit is used to house the waveguide, it should be at least four inches inside diameter and have no more than two 90° bends.

Due to the enormous power differences at the feed between transmit and receive ports, a transmit reject filter must be used between the receive feed port and the input to the LNA. This filter will minimize any transmit power leaking into the LNA package and cause out-of-band saturation.

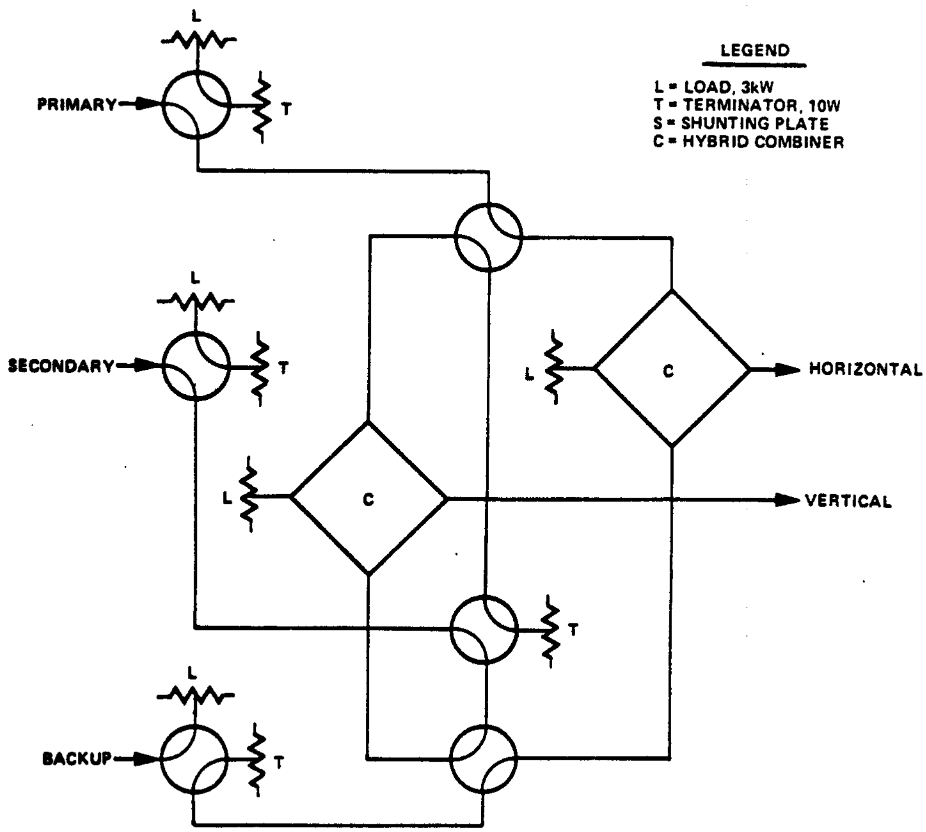
The RF switching matrix is an integral part of the waveguide runs and allows protection of HPA's. By switching a backup HPA to an antenna port during a failure of the normal HPA, earth station operation is only interrupted momentarily. Existing waveguide runs to the antenna port are used since the switching occurs as close to the HPA output as possible. In some applications the backup is kept in transmit mode with the output of the HPA into a 3 kW dummy load. This method allows for hot standby switching if both HPA's are tuned to the same frequency. It is sometimes useful for individual 3 kW dummy loads to be dedicated to each HPA through a waveguide switch for ease in testing and alignment as well as immediate uplink capability. Some typical switch and load panels are shown in Figure 5.

Cooling and Ventilation

Air-conditioning requirements should be tailored to the size of the earth station and the particular configuration at the site. The total number of HPA's (both in transmit and hot standby), exciters, receivers, any UPS control equipment and any other large heat-generating equipment should be included in the cooling budget.



1:1 WITH COMBINER
2- OR 3- PORT FEED



1:2 WITH COMBINER
4- PORT FEED

Figure 5. Typical HPA Waveguide Switch and Load Panels

The HPA's and associated equipment represent the largest single source of heat in a transmit station. Each HPA is vented through the rear wall of the equipment shelter. Two 8-inch-diameter vents are provided: one for fresh air (input), the other to exhaust the heated air from the equipment shelter. Each HPA has two blowers to accomplish this task. Approximately 8 kW of heat are removed from each HPA in this manner. This equates to approximately 27,000 BTU/hr.

Using the guidelines specified in the above paragraphs, approximately 1 kW of heat per HPA will remain in the shelter to be cooled by the air-conditioning system. Additionally, the air-conditioner must handle the heat released by the dummy load to which a hot standby HPA is connected.

As a safety precaution, the shelter or equipment room where the transmitters are to be located should also contain a fresh-air system to circulate outside air into and out of the shelter in the event the normal air-conditioning system fails. This system can be fully automatic with a temperature switch to activate the air circulator, or a manual switch control.

Safety

When designing, operating and maintaining a transmit earth station, safety precautions should always be considered. Video transmit earth stations frequently exceed EIRP levels of 80 dBW. This produces a microwave radiation hazard in the area directly in front of the main reflector. Since the beamwidths of the 10-meter and 11-meter antennas are extremely narrow, the majority of the radiated microwave energy is confined to the area within an imaginary cylinder extending from the front of the main reflector. Locations with high elevation look angles experience little or no real hazard from microwave radiation. Although U.S. locations on domestic satellites have high elevation angles, care should be taken in site planning to ensure that any nearby tall buildings are well out of the main beam.

The main beam should not be blocked by any object. Blocking the path will affect earth station performance (G/T) and cross-polarization rejection, as well as safety. Precautions taken to keep this area clear include enclosing at least the area surrounding the antenna and mount within a fence. This enclosure protects the antenna from vandals and from people (i.e., children) who like to climb. Climbing on the main reflector when the station is transmitting is extremely dangerous. The area between the subreflector and main reflector has a very high microwave radiation density.

Appendix A
TYPICAL EARTH STATION POWER REQUIREMENTS

		208V/3Φ (kVA)	110V/1Φ (VA)
Antenna Motor Drives (Full Load/Run Condition			
7-Meter:	High Speed	9	
10-Meter:	Standard	5	
	High Speed	10	
11-Meter:	Standard 60°	8	
	Standard 110°	6.5	
	High Speed 110°	12	
8840A Remote Control			50
Antenna Deicing			
7-Meter:	Feed and Sub	1	
	Half Main	10	
	Full Main	19	
10-Meter:	Feed and Sub	2	
	Half Main	26	
	Full Main	50	
11-Meter	Feed and Sub	3	
	Half Main	27	
	Full Main	51	
Single Electronic Units (Double for Redundant Systems)			
3.35 kW HPA		12	
7500 Receiver			100
7550 Exciter			100
Other Electronics			
7630 LNA Protection Switch			100
7620 Receiver Protection Switch			100
7640 Uplink Protection Switch			100
7600 Earth Station Controller			300
7605 Earth Station Interface			100
Test Equipment, Rack, Misc.			2,000
Shelter: A/C, Heat, Lights		2-9	600-1,800

Full Load Current (Per Leg) $= \dfrac{\text{VA}}{\text{P* x Voltage}}$

*For:

$1\Phi : P = 1$

$3\Phi : P = \sqrt{3}$

Antenna Foundations for Satellite Earth Station

Roger A. Peirce

Introduction

A satellite earth terminal antenna typically utilizes a concave, solid-surface reflector, which must be maintained in a fixed position while being subjected to wind loads. The forces imposed on an antenna by a survival wind loading are significant because of the reflector's large surface area. Because the center of the reflector may be many feet above the ground, the foundation must resist not only horizontal drag forces but also large overturning moments as well. The foundation must safely transfer these forces and moments into the ground, providing a stable base that will keep antenna deflections within allowable limits.

This paper will describe foundation types which are suitable for earth terminal antennas and will discuss the effect of soil types on foundation capacity. Site requirements and construction techniques are also presented. All information included is for reference and is not intended to be a recommendation for a particular site.

Foundation Loadings

A satellite earth terminal antenna must be designed to safely withstand any load which could reasonably be expected to occur during its useful life. The weight of the antenna will always have to be supported. It is called the dead load. The antenna will be exposed to all local weather phenomena during its operation. Conditions may occur which allow the formation of a sheet of ice on the antenna. This additional weight of ice is called the ice load. The antenna will also be subjected to wind forces called the wind load. Due to the large surface area of the reflector, forces generated by wind will be considerable. The design wind speed should be based upon the 50-year mean recurrence interval at the site. These wind speeds may be obtained from the American National Standards Institute (ANSI) publication A58.1 - 1972, which also defines a method for computing the magnitude of wind forces. Design wind forces may also be obtained from Electronic Industries Association (EIA) Standard RS-222-C, which is a structural code for antenna supporting structures. From the map in Figure 1 it can be seen that the maximum 50-year recurring wind in the continental United States is 110 mph. It is recommended that 80 mph be used as a minimum design wind speed, using a higher speed if required by site location.

Wind load should be included in all loading combinations because of its relative magnitude. Suggested load cases are as follows:

- Case 1 - Dead Load + Wind Load
- Case 2 - Dead Load + Ice Load - 1/2 Wind Load

All possible antenna orientations with respect to the foundation should be considered to ensure that maximum wind-force reactions are calculated. The horizontal forces caused by wind vary as the square of the wind speed. As an example, if 110 mph was used in calculating the force in equation 1, approximately 80 mph ($110 \div \sqrt{2}$) should be used in equation 2. For most site localities, load-case 1 will control the foundation size.

Site Requirements

Selection of a suitable site for an earth terminal antenna will help ensure trouble-free operation. The following are some of the guidelines which should be used in selecting a site.

a. The site should have a clear line of sight to the satellite. The antenna should be clear of any trees, buildings, or any other obstructions which might interfere with satellite signals.

b. The site should be relatively level.

c. The site should be free of underground obstructions that would interfere with foundation construction.

d. The site should be in a well-drained area. Do not select a location that is damp or subject to drainage runoff.

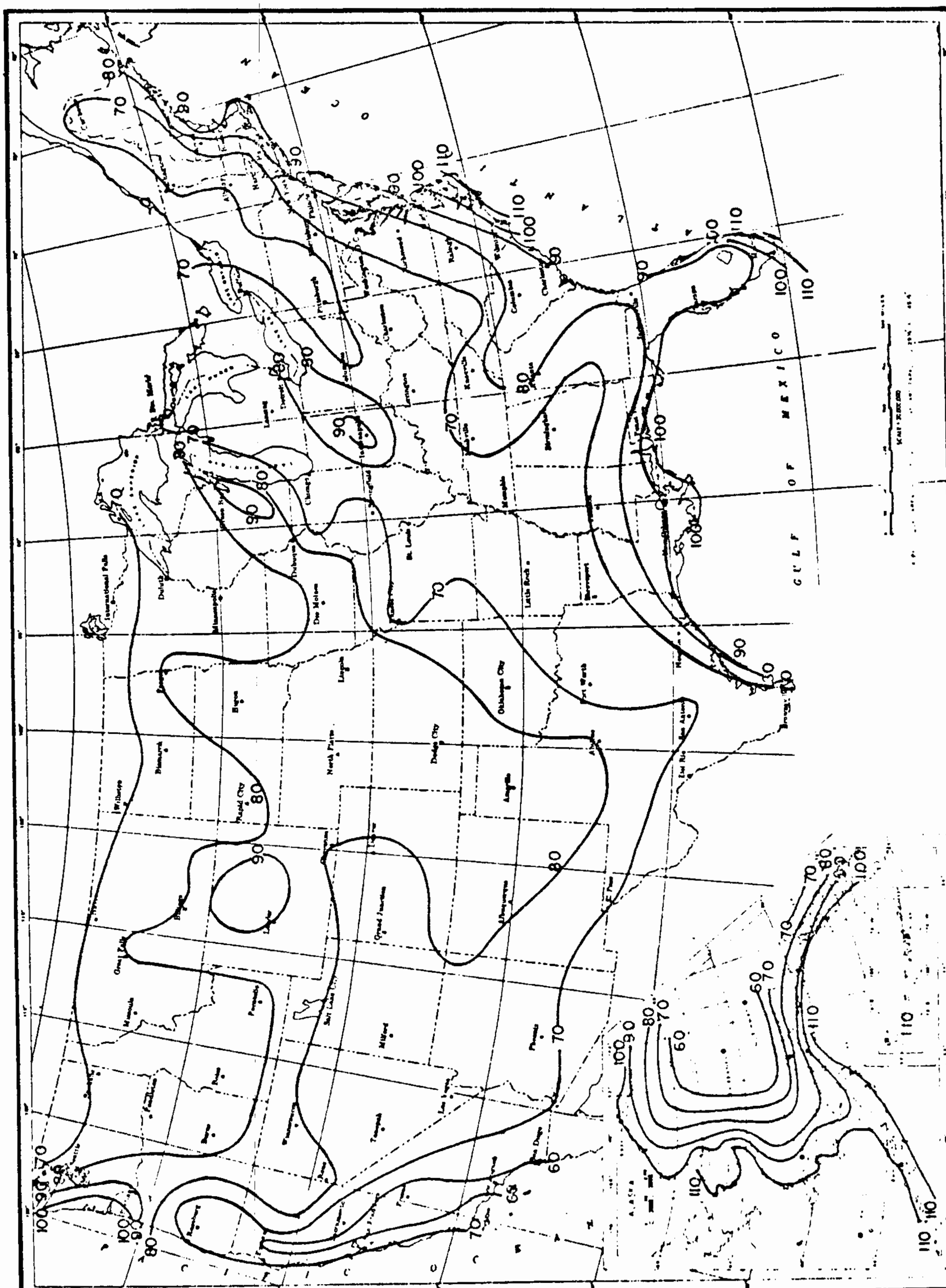
e. The site should conform to all local building codes.

Local Soil Conditions

Before selecting a foundation for an earth station antenna, a soil investigation should be made at the intended site. Soil is a general term used to describe any unconsolidated material composed of discrete solid particles with gases or liquids between. Soil can be made up of any number of different substances, and as a consequence, its engineering properties can vary considerably.

For the purpose of engineering study, soils are usually classified into two general groups. Soils in which absorbed water and particle attraction form a mass which holds together and deforms plastically are known as cohesive or clay soils. Soils that do not exhibit the ability to stick together (dry sands and gravel) are called cohesionless soils. Many soils are mixtures of bulky grains and clay minerals and exhibit some degree of varying consistency with changes in moisture. These too are termed cohesive soils if the effect is significant. Typical soil properties are shown in Figures 2 and 3.

Given an exact loading, the use of a single standard antenna foundation in a variety of soil types may not be economical. Before selecting a foundation type, the soil conditions at the site must be identified. Although there is no sharp dividing line between cohesionless and cohesive soils, most soils can be safely classified as being either one or the other. Proper engineering values can then be selected. There are, however, some cases in which these two groups cannot accurately describe actual conditions, and anyone planning to erect an antenna should check with a local authority who is knowledgeable concerning any special soil problems at the site. If unusual conditions exist, a proper foundation can then be specially designed.



AMERICAN NATIONAL STANDARD A58.1-1972

Figure 1. Basic Wind Speed in Miles per Hour Annual Extreme Fastest-Mile Speed 30 Feet above Ground, 50-Year Mean Recurrence Interval

The results of a soil investigation will provide the foundation engineer with the necessary soil properties to complete a proper foundation design.

Frost penetration is a special problem which must be evaluated at each site. Freezing of water in soil can result in upward expansion of the overlying ground. Any foundation whose base does not extend below this frozen zone may be subject to upheaval. To prevent this, the bottom of the antenna foundation should be at least as deep as the maximum anticipated frost penetration. Frost depth will be the determining factor for foundation depth at most northern United States sites. Frost penetration varies from 5 inches of the Gulf and Pacific coasts to 100 inches in Minnesota and Main. Frost penetration for the United States is shown in Figure 4.

All earth station antennas should be provided with a suitable grounding system that meets the requirements of local codes. This system should be designed by a local engineer, since the grounding system required will depend on the resistive characteristics of the soil. Scientific-Atlanta provides all necessary grounding straps required on the mount.

Typical Soil Properties Relating to Foundation Design

Massive Crystalline Bedrock: Granite, Gneiss	200,000
Foliated Rock: Schist, Slate	80,000
Sedimentary Rock: Hard Shales, Siltstones, Sandstones	30,000
Exceptionally Compacted Gravels or Sands	20,000
Compact Gravel or Sand-Gravel Mixtures	12,000
Loose Gravel: Compact Coarse Sand	8,000
Loose Coarse Sand or Sand-Gravel Mixtures:	
Loose Fine Sand or Wet, Confined Fine Sand	4,000
Stiff Clay	8,000
Medium Stiff Clay	4,000
Soft Clay	2,000

Figure 2. Allowable Bearing Capacity on Soils, Pounds per Square Foot (PSF)

Soil Type	Cohesion (PSF)	ϕ (Degrees)
Silts, Wet	0	10°
Sand, Dry	0	34°
Sand, Immersed	0	34°
Clay, Liquid	100	0°
Clay, Very Soft	200	2°
Clay, Soft	400	4°
Clay, Farily Stiff	1000	6°
Clay, Very Stiff	2000	12°
Cemented Sand & Gravel	500	34°
Cemented Sand & Gravel, Wet	1000	34°

Figure 3. Typical Cohesion and Angle of Internal Friction Values for Soil

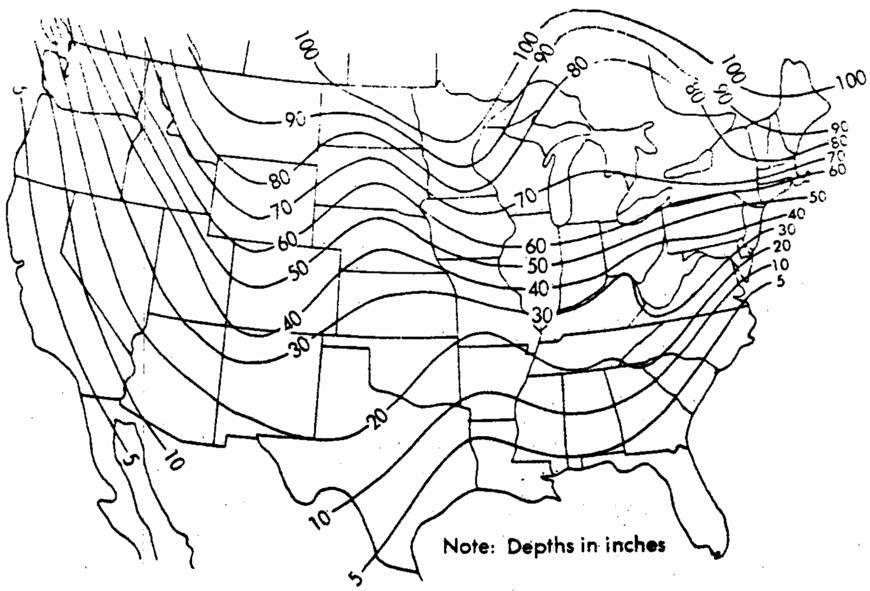


Figure 4. Maximum Depth of Frost Penetration in the United States

Foundation Types

Two different types of foundations are commonly used for earth terminal antennas. The first is the mat or spread footing which utilizes a single monolithic reinforced concrete mat to resist all loadings (see Figure 5). Mat footings have the advantage of distributing downward forces over a large surface area, resulting in low soil bearing pressures. Overturning forces due to wind loads are resisted by the weight of the mat. Antennas which produce significant overturning moments will require sizeable mats for stability. The second type of footing is the drilled pier which utilizes reinforced concrete cylinders which are placed in auger-drilled holes. Downward loads are resisted through end bearing and by friction between the pier and surrounding soil. Overturning forces will produce additional downward loads on some piers and uplift on others, depending on wind direction. Uplift forces are resisted by friction between the pier and soil and by the weight of the pier. The tops of the piers are attached to a rigid triangular frame. This enables them to act as a unit to resist horizontal forces. Pier foundations require much less concrete than do mat footings because overturning moments are not resisted by concrete weight alone. This saving of concrete makes drilled piers economically attractive where soil conditions are favorable for their construction. The following section describes elements of the design theory for both foundation types.

Foundation Theory

The load-carrying capacity of a given foundation depends upon the properties of the soil in which it rests. In order to design an adequate foundation, one must understand the different effects of cohesive and cohesionless soils on both mat and drilled pier foundations.

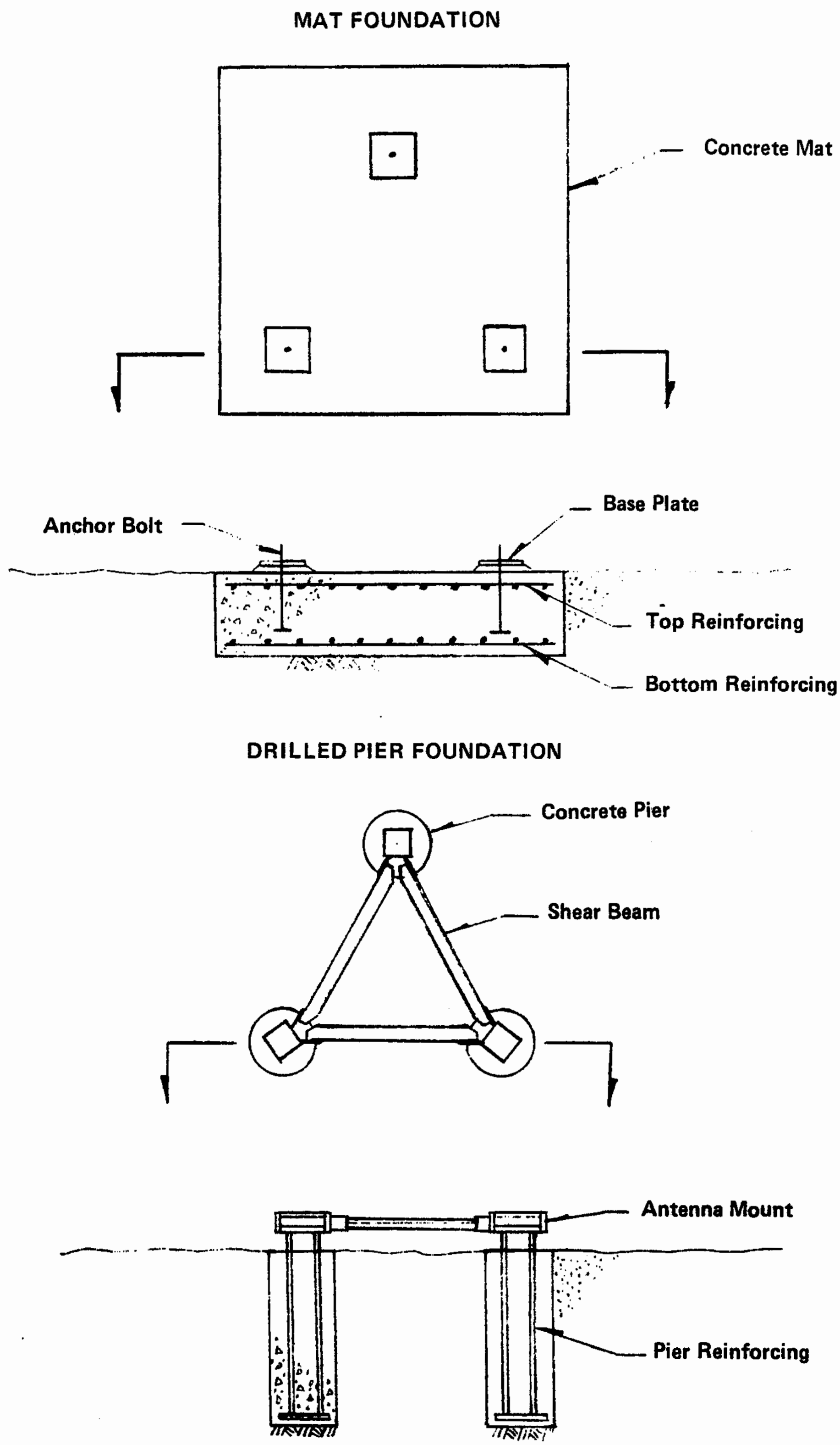


Figure 5. Typical Antenna Foundations

Mat Foundation

The load capacity of a mat foundation depends on the bearing capacity of the soil. Compressive loads are distributed over the entire footing bearing area as shown in Figure 6-A. Dense sands and stiff clays have a bearing capacity, of about 8000 pounds per square foot (PSF). Most soils (both cohesive and cohesionless) have a bearing capacity of at least 3000 PSF and will safely support a mat foundation.

When the footing is subjected to overturning wind loads, as in Figure 6-B, some piers may have an upward load. Soil pressure is no longer uniform, and care must be taken not to exceed the allowable soil bearing capacity. Also, the overturning moments should be checked at point A. Stabilizing moments caused by P_2 and the footing and the soil weights must be 1.5 times greater than the overturning moment caused by P_1. Lateral forces are resisted by friction between the footing and soil and to a lesser extent by the passive resistance of the soil.

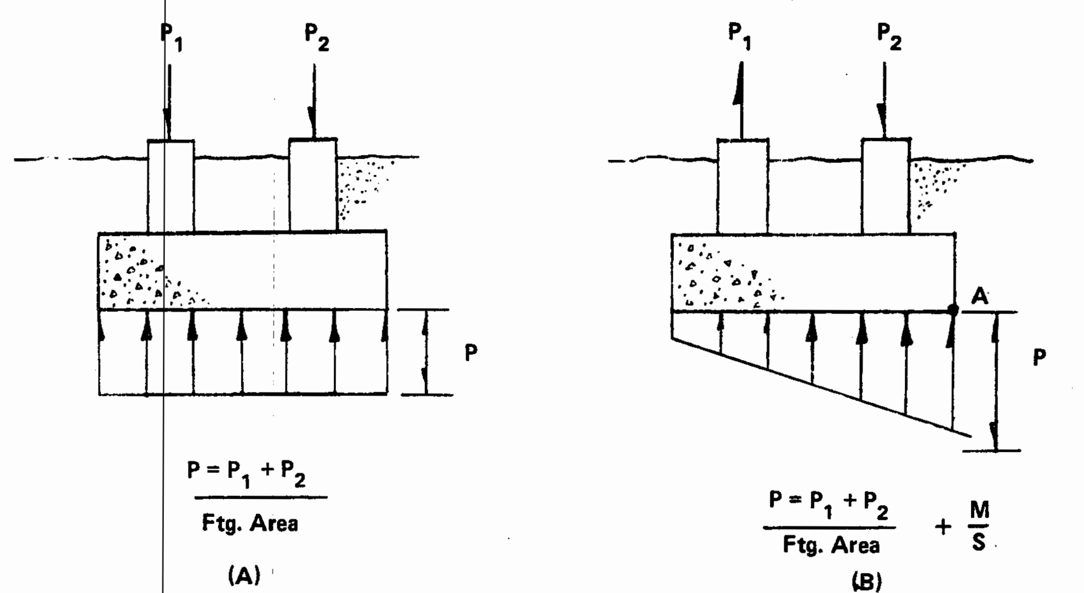


Figure 6.

Drilled Piers

Drilled piers resist downward loads with skin friction acting on the sides of the piers and with end bearing acting at the base. Wind uplift is resisted by skin friction and pier weight. The amount of skin friction resistance which can be developed for a given foundation is dependent on the shear strength properties of the soil.

Cohesive soils resist pier pullout by the direct cohesion of the soil to the pier. Simply put, clay soils act as a glue, sticking to the pier surface. This strength is measured by the undrained shear strength or cohesion of the soil. A stiff clay may have a cohesive strength of 2000 PSF, while an average clay will have a cohesion of about 1000 PSF. Thus, skin friction resistance of a given pier equals its side area multiplied by the soil cohesion. This figure should be reduced by a factor of 2/3 to 1/2 for safety.

Cohesionless soils resist pullout in a different manner. The bulky grains of sandy soil do not stick to the pier surface. The frictional resistance of sandy soils on the pier is analogous to sliding a block along a horizontal surface.

The frictional resistance of the block in Figure 7-A with the horizontal force P is equal to the weight of the block times the coefficient of friction that is effective between the block and the surface. In Figure 7-B the pier is analogous to the friction surface in Figure 7-A. The horizontal soil pressure on the pier (which increases with depth) is analogous to the weight of the block in Figure 7-A. The coefficient of friction between the soil and the pier is given by a property of the soil called the angle of internal friction (ϕ). The useable coefficient value is equal to Tan ϕ. The resistance to the vertical force P in Figure 7-B is equal to the sum of the horizontal soil weight times pier side area times Tan ϕ. From this relationship it can be seen that pullout resistance in cohesionless soils increases with depth. Shallow piers in sand have little pullout capacity, while deeper piers will begin to develop sizeable tension capacities.

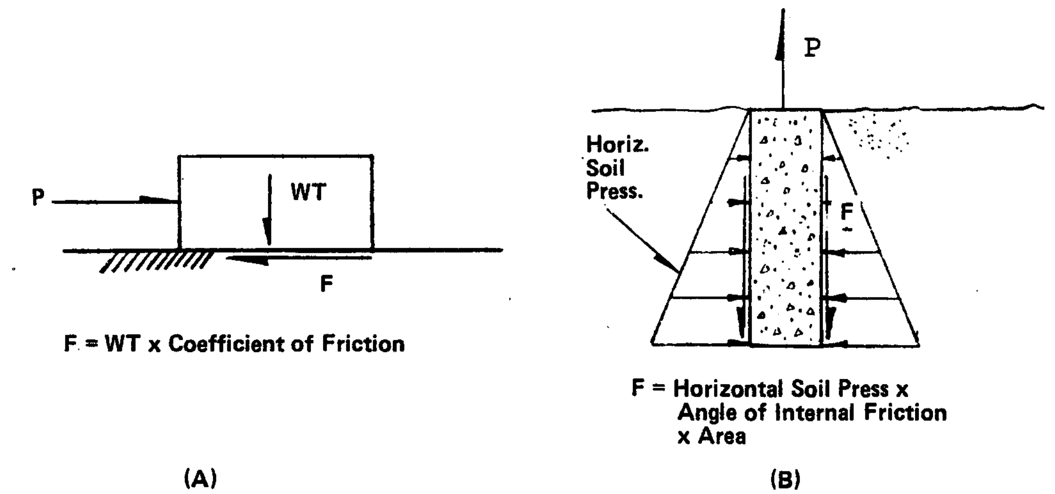


Figure 7.

Empirical equations have been developed for the deep-bearing capacities of both cohesive and cohesionless soils. These capacities increase with depth and may be considerably larger than surface bearing capacities for the same soil (used in mat foundation design). This increase in bearing capacity along with skin friction resistance explain why a drilled pier can be smaller than an equivalent mat foundation.

Resistance to lateral forces on piers in cohesionless soils is a function of the passive resistance of the soil. Lateral load resistance in cohesive soils is a function of cohesion. Empirical design methods are available which make the determination of lateral pier capacity a simple matter.

The main advantage of pier foundations is their economy. Quite often the factor which determines the required pier size is resistance to uplift. This resistance may be increased or by belling out (enlarging) the bottom of the piers. This will require drilling a deeper hole and using more concrete. Another way in which pier uplift capacity can be increased is by the use of earth anchors. Anchors can usually be installed quickly if soil conditions are normal. In some cases it may be more economical to use shorter piers with earth anchors than to use deeper piers (see Figure 8).

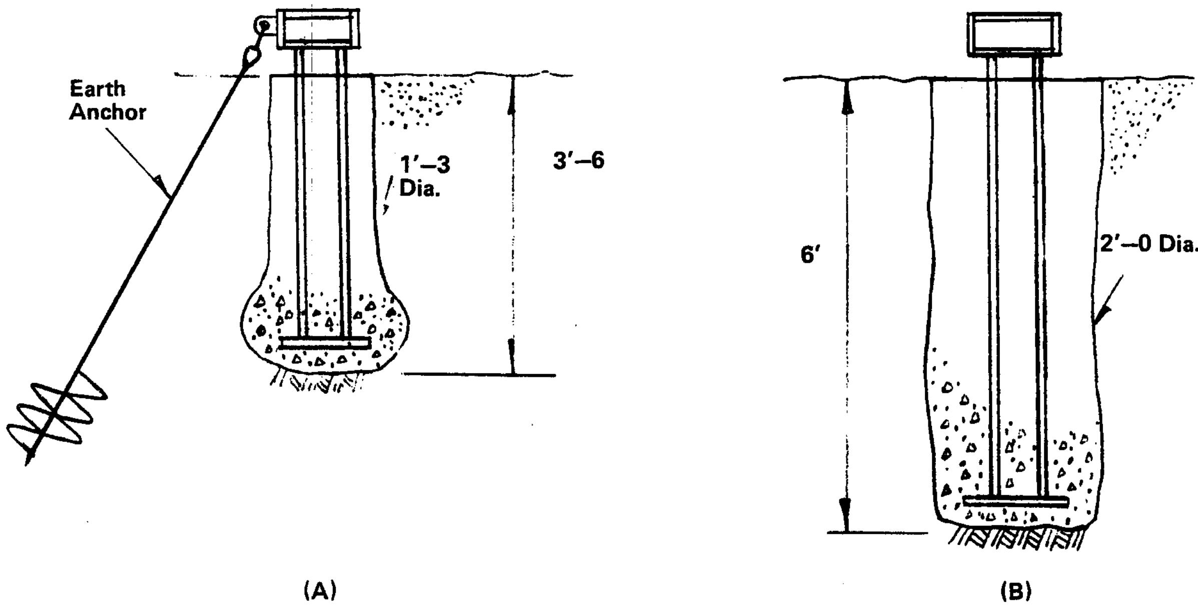


Figure 8. These Two Piers Have Equivalent Uplift Capacities

Foundation Headings

Foundation headings are very important to proper antenna operation. Generic considerations and recommendations for aligning the foundation and pointing the reflector at satellites are treated in detail in the article entitled "Earth Terminal Geometry."

Many earth station antennas can only scan a certain portion of the sky (typically 110° in azimuth) because of mechanical limitations. If the foundation heading is established incorrectly, the antenna will not be able to scan the portion of sky originally intended. An inaccurate foundation heading may require the construction of another foundation to point the antenna at the desired satellite. The installer of an earth station should be sure that his foundation contractor is aware of the importance of foundation heading.

There are several methods that can be used to establish a foundation heading. A known reference survey line near the antenna site can be transferred to the foundation by normal surveying techniques. The accuracy of the existing survey line should be established. If there is no convenient reference line existing, one can be established by a polaris (North Star) observation made by a competent surveyor. Alternatively, the heading can be established by using a compass. A compass does not point to true north but to magnetic north. The angle between true north and magnetic north at a given place is called the magnetic declination. The magnetic declination for any point in the United States can be determined from the Magnetic Declination Map[1] published by the U.S. Geological Survey and from other reliable maps. The magnetic declination should be added or subtracted, depending on the local

[1] Maps are published every five years, most recently in 1980.

direction of declination, when a compass direction is used to determine the foundation heading. Magnetic disturbances caused by large metal structures and electrical fields will affect compass reading accuracy. Compass readings should not be used when these disturbances are present.

The heading accuracy required for a particular foundation depends on the following factors:

- The range and domain of the satellites considered.
- The latitude and longitude of the earth station.
- The type of earth station mount.

Once a particular type of antenna and look angles are selected at a given site location, the allowable foundation heading error can be determined. The allowable heading error is defined as the maximum misalignment of the foundation that will still allow the antenna to function as planned.

Foundation heading errors can be caused by:

- Survey errors.
- Misalignment of anchor bolts.
- Improper definition of "heading" resulting in 180° misalignment.
- Other construction errors such as tilt, etc.

As an example, if an allowable foundation heading error is 6°, some error should be budgeted to survey inaccuracy and some to construction errors. Thus, the allowable survey error would be less than 6°, as would the allowable construction error with the sum of the two equal to 6°. If a total heading error of ±5° or less is required for a foundation, it is recommended that a reliable existing survey or a polaris observation be used in establishing the foundation heading. If the contractor or surveyor elects to use a compass, the reading should be made with a high-quality compass using the best possible techniques.

Foundation heading accuracy requirements may become more stringent when a satellite is placed at one of the extremes of the antenna arc coverage. In this case, a slight heading error may place the desired satellite out of the arc of coverage. For this reason it is good practice to leave adequate allowance for foundation error between a desired satellite and the end-of-the-arc of coverage.

Correct foundation headings are very important, and mistakes in orientation can result in lost time and additional construction.

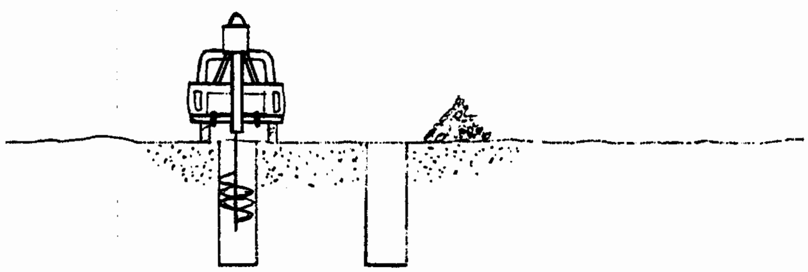
Construction Methods

Pier Foundation

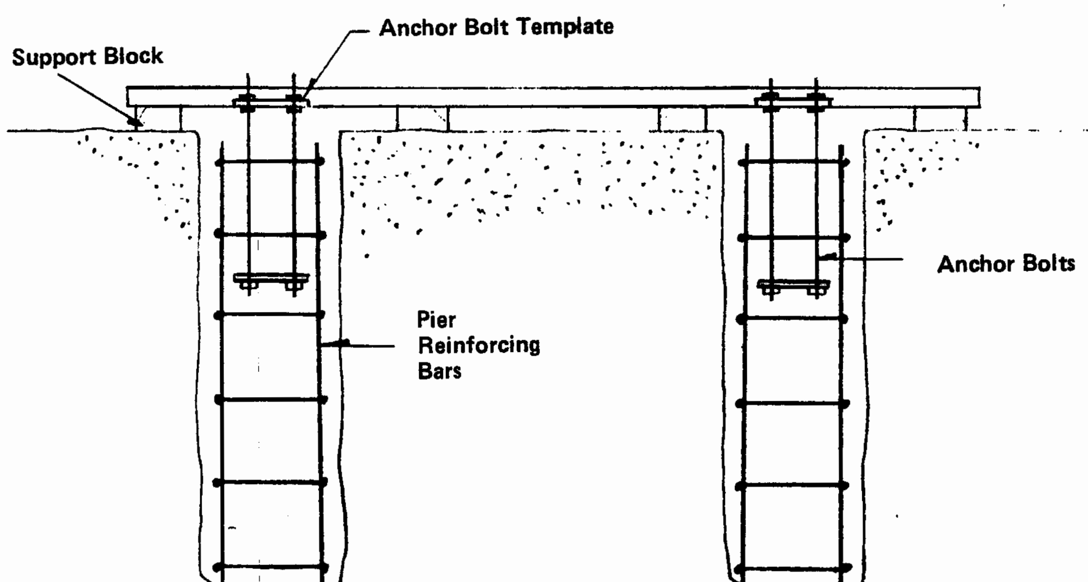
The following are two methods by which pier foundations can be constructed. The first method shows the usual building methods for a pier foundation while the second method shows the Scientific-Atlanta Pier Foundation Kit (available for the 4.6-meter and 5-meter antennas) which is easier to construct.

Method A - Pier Foundation Construction

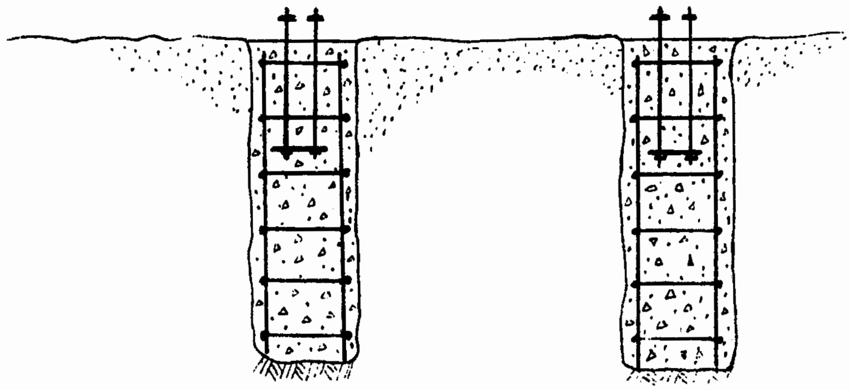
Step 1 - Holes are drilled by either a truck-mounted or hand-held power auger. The bottom of the hole should be tamped. The bottom of the hole should also be enlarged if possible.



Step 2 - Pier reinforcement is lowered into drilled holes and anchor bolts are positioned using a template.



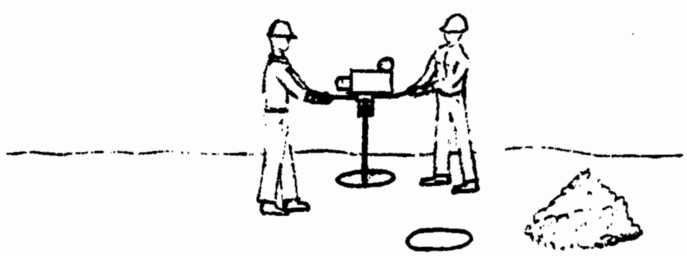
Step 3 - Concrete is placed into drilled hole, template is removed and concrete is allowed to set. The pier is then ready for base plate and antenna mount.



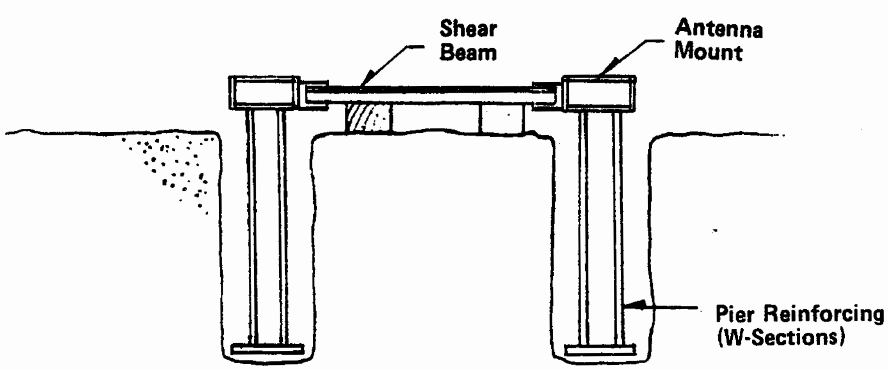
Method B - Pier Kit Installation for 3-Meter and 4.6-Meter Antennas

A unique steel pier foundation kit is available for the 4.6- and 5-meter earth stations that provides cost savings and reduced installation times, as compared to a monolithic pad. This foundation kit utilizes a design concept that incorporates steel pier members which are inserted into cylindrical holes augered into the earth. Savings are realized by the decreased concrete requirement (about one-tenth of typical pad requirements) and the limited site preparation (a concrete form is not required).

Step 1 - Holes are dilled by either a truck mounted or handheld auger. The bottom of the hole should be tamped and enlarged if possible.



Step 2 - After the holes are drilled and tamped, a preassembled load frame consisting of w-sections, shear beams and antenna mounts is positioned in the holes and supported by leveling blocks.



Step 3 - Concrete is placed into the holes.

Step 4 - Once the concrete has hardened, the antenna may be erected.

General Requirements for Pier Construction

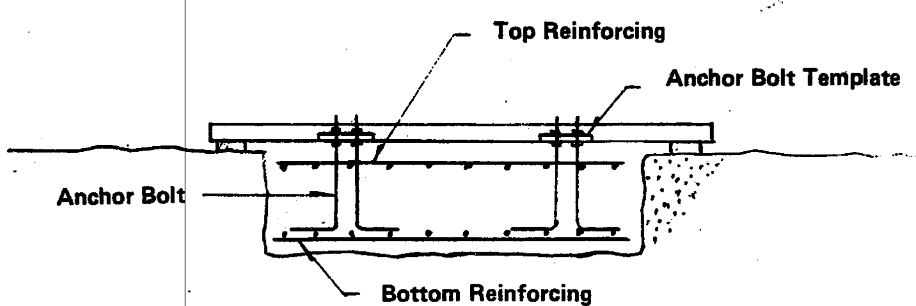
- The pier should always be reinforced, as it is subject to tension and bending as well as compression.
- Pier should extend below frost depth.
- Pier hole should be free of loose dirt and bottom should be tamped before placing concrete.

Mat Foundations

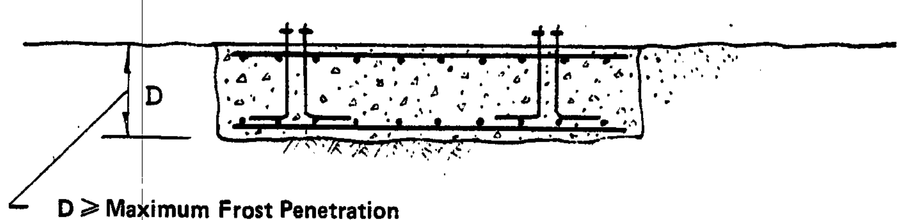
Mat foundations have been used to support many earth terminal antennas. The following is a general procedure for mat construction.

Step 1 - Soil is excavated down to the elevation of the bottom of the mat. Loose dirt should be removed.

Step 2 - Reinforcing bars and anchor bolts are positioned in excavation. Anchor bolts are supported by a template.



Step 3 - Concrete is placed and allowed to set. The mat is then ready for base plates and antenna mounts.



Selection of Foundation Type

The selection of a foundation type is sometimes influenced by special problems at the site. If the soil is very rocky or if the site is solid rock, pier holes cannot be drilled by lightweight equipment. In this case, a mat foundation can be placed directly on the rock, thereby reducing construction time and costs. In some cases pier holes cannot be drilled because the soil is so loose that the holes collapse before concrete can be placed. Again, a mat foundation can be used because it would require a shallower excavation. Piers are often used at sites with deep frost penetration because they require less excavation than an equivalent mat constructed at the same depth.

Several factors may influence the cost effectiveness of different foundation designs. Sometimes the economical advantages of a pier foundation are offset because of the cost required to bring in special drilling equipment to a remote or inaccessible site. If the owner is planning to install many earth stations, it may be cheaper to buy his own drilling equipment and install pier foundations using his own trained crew. Many small contractors are better equipped to construct mat foundations than they are to build piers, which means they will build a mat for a lower cost. If no soil problems are present at the site, either type can be used, and the selection of the type of foundation can be based on cost.

In some locations the foundation design must bear the seal of a professional engineer registered in the state in which the earth station will be built.

Considerations for Rooftop Mounts

Buyers of earth stations may want to locate the antenna on the roof of an existing structure. There are many potential problems associated with rooftop mounts that must be investigated for each installation.

The building must be strong enough to support the large wind loads generated by the antenna. A review should be performed of the design of the structure upon which the antenna will be mounted. Many smaller structures such as motels and small office buildings will not support the wind loads of an antenna. Tall buildings are subject to differential thermal expansion and swaying due to wind loads. Movements due to these two cases could exceed the maximum allowable pointing deviation of the antenna.

In most cases a special load frame must be designed to connect the antenna mount to building columns. The load frame must be sufficiently rigid that its added deflection will not cause pointing problems. Most rooftop installations are expected to be considerably more costly than a typical ground installation.

The antenna wind survival capability will have to be progressively derated as the antenna is moved higher above the ground. The American National Standard Institute publication A58.1 - 1972 gives information on effect of height on wind velocity pressures affecting buildings shown in Table 6. A wind of a given velocity (which is measured at ground level) will produce a certain pressure against a resisting surface. To account for less ground surface resistance to wind flow, this design pressure is increased as the height of the resisting surface is increased. The design wind pressure exerted by a 120 mi/h wind varies from 34 pounds per suare foot at ground level to 96 pounds per square foot at a height of 800 feet. The increase in wind pressure on tall buildings will result in larger design loads on rooftop mounts than on corresponding ground-level foundations.

Rooftop mounts for antennas are much more expensive than are conventional ground-level foundations and are subject to additonal deflections and survival deratings. The use of rooftop mounts is not recommended in most applications, although as satellite communications become more widespread, an increasing number of new buildings may have antenna mounting areas included in their designs.

Material Requirements

Materials used in earth terminal antenna foundation construction should conform to the following specifications:

a. Concrete. All concrete should have a minimum 28-day compressive strength of 3000 psi. When pier foundations are used, concrete slump should be six or seven inches. This will ensure that the concrete will flow freely around reinforcing bars and completely fill the space outside the cage. Maximum aggregate size should be one inch.

b. Reinforcing Bars. All reinforcing steel should be intermediate grade steel bars with 40,000 psi minimum yield strength. Reinforcing bars should be deformed. The deformations should conform to specifications for deformations of deformed steel bars for concrete reinforcement (ASTM-A615-75).

c. Anchor Bolts. Anchor bolts should be sized to safely carry tension, shear, and compressive loads. Anchor bolt templates should be used.

d. Grout. Use non-shrink type of grout for grouting base plates.

e. Soil Bearing Capacity. The minimum recommended soil bearing capacity is 3000 PSF. For pier foundations in cohesive soils, the minimum recommended cohesion is 1000 PSF. For pier foundations in cohesionless soils, the minimum recommended angle of internal friction is 30°.

Exact Design Procedures

This article has attempted to explain some of the basic principles involved in earth terminal antenna foundation design and construction. For a more complete understanding, the reader is directed to the following sources:

Foundation Engineering, edited by G.A. Leonards, McGraw-Hill Book Company, 1962.

American National Standards Institute (ANSE), Publication A58.5 - 1972.

Electronic Industries Associated (EIA) Standard RS-222-C Structural Standards for Steel Antenna Towers and Antenna Supporting Structures.

Receive Only Earth Station Installation

Stephen E. Havey

Introduction

This paper outlines the general considerations and procedures for the installation of Scientific-Atlanta receive-only earth stations. Routine maintenance and fault isolation is also discussed.

Site Location

Location of a suitable site is discussed in the previous article, "Planning a Receive-Only (or Transmit/Receive) Earth Station." As a reminder, some of the considerations are:

1. Is the land available for lease or purchase?

2. Is the site free from potential terrestrial microwave interference?

3. Is the line of sight to the desired satellite(s) clear of obstructions such as trees and buildings?

Foundations

The foundation is a critical element of the earth station installation. Each antenna offered by Scientific-Alanta is designed to cover the complete satellite arc of 67° to 143° west longitude from the continental U.S. (except for the southern tip of Texas) without modification to the foundation. The foundation must withstand the specified maximum wind loads while maintaining the proper pointing angle.

The instruction manual for each Scientific-Atlanta earth station antenna contains wind loading data and typical foundation drawings. Refer to these drawings and to the previous paper, "Antenna Foundations for Satellite Earth Stations."

It is imperative that the owner of an antenna satisfy himself through the use of competent engineering assistance that the foundation is properly designed for his particular application and the local building codes. Typical foundation plans are provided by Scientific-Atlanta; however, Scientific-Atlanta does not represent or warrant that a particular design or size of foundation is appropriate for any locality or installation.

Foundation Heading

For any selected earth station site, a foundation heading (pad center line) is supplied by Scientific-Atlanta that allows the antenna to access any satellite within 67° to 143° west longitude geostationary arc without a foundation change. This foundation heading is given in degrees from true north. It is unique for each site and for each antenna type. A change in either location <u>or</u> antenna type will alter the foundation heading.

Prior to pouring the pad or foundation, provisions should be made for providing ac voltage at the pad (if applicable) and an RF cable interface between the pad and the headend building. This can easily be accomplished through the use of 4-inch conduit.

Establishing the Centerline

Scientific-Atlanta provides pointing information consisting of the foundation heading (centerline) and look angles. For most locations and antenna types, the foundation heading may be established through the use of a good quality magnetic compass. When using a magnetic compass, you must correct for the local magnetic declination; that is, the difference between true north and magnetic north. On most Scientific-Atlanta earth station foundations, the centerline heading is on a line which evenly bisects the two rear mounting feet and passes directly through the front mounting foot (Figure 1).

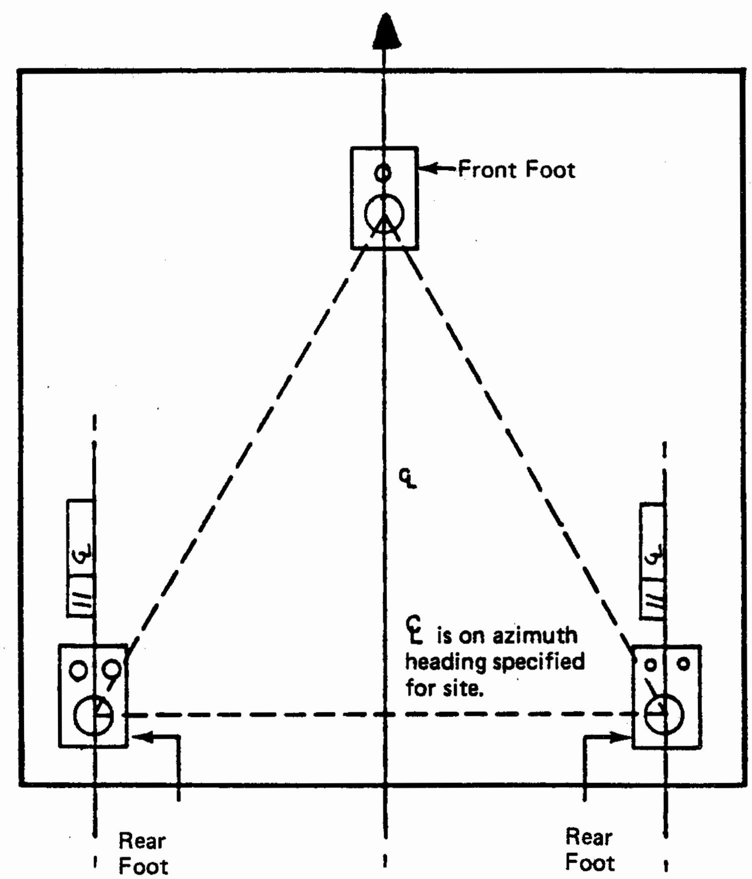


Figure 1. Typical Antenna Foundation Heading

Antenna Construction

Each model of earth station antenna manufactured by Scientific-Atlanta contains parts and construction procedures which are unique to that model. For that reason, we will not attempt to provide detailed instructions for each antenna. Each antenna comes with a detailed instruction manual containing step-by-step instructions, a complete parts listing and a full set of drawings. Refer to the respective manual for your antenna and read it thoroughly before attempting to construct the antenna.

Installation Techniques

This section is intended to assist in the installation of those earth station components located between the earth station antenna and the video receiver or receivers. In addition, there are instructions for antenna pointing, polarization, satellite acquisition and troubleshooting guidelines.

Introduction

Figure 2 illustrates a typical TVRO (television receive-only) earth station. This particular system is equipped for operation on both polarizations, with outputs for four (4) receivers on each polarization. In sequential order, from the antenna to the receiver, we will cover the following items:

1. Polarization
2. LNA (Low-Noise Amplifier) or LNC (Low-Noise Converters)
3. Coaxial Jumpers
4. Coaxial Cable and Connectors
5. Power Dividers
6. Automatic Polarization Relay (for 6602, 7500, and 414 Receivers only)
7. Pressurization Unit
8. Signal Attenuation
9. Antenna Pointing
10. Satellite Acquisition

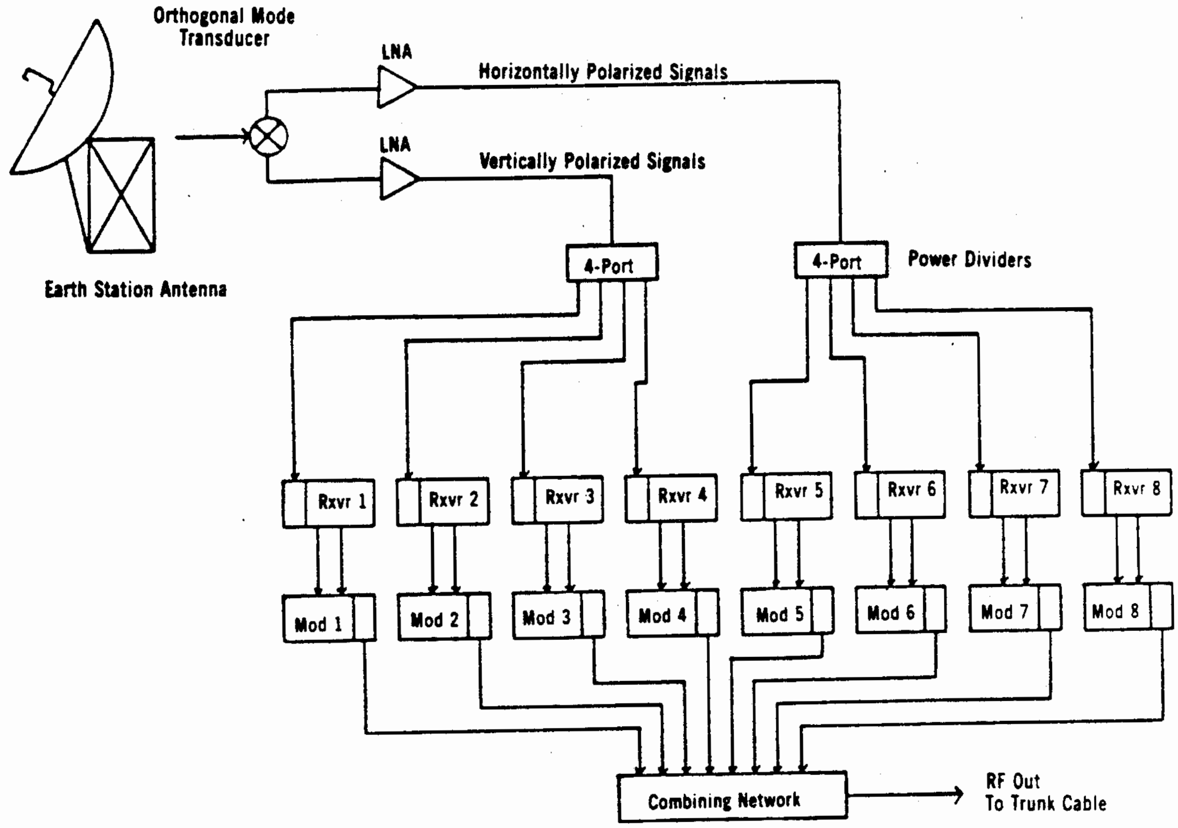


Figure 2. Earth Station Block Diagram

Polarization

All signals from current DomSats (domestic satellites) are polarized in either the horizontal or vertical plane (as referenced to the equator). Most satellites transmit in both polarizations, thereby achieving a 24-video-channel capacity in the 500 MHz bandwidth available (3.7 to 4.2 GHz). Figure 3 lists the frequencies, transponder number and services for the three satellites currently providing video programming to the cable TV industry. In planning to carry programming from each polarization simultaneously, it is necessary to install two (2) LNAs and run two (2) separate cable runs to the receiver location.

The signals from each polarization may be neither combined nor mixed. Separate inputs to the two (2) LNAs are provided by the OMT (orthogonal mode transducer) available as standard equipment on all Scientific-Atlanta earth station antennas and assembled to the feed horn of the antenna. Figure 4 is a photograph of an OMT properly oriented for maximum signal on each polarization. This position will vary slightly for other locations in the United States and a small amount of "peaking" will be required for optimum performance at each site.

	SATCOM III R 131° W LONGITUDE		SATCOM IV 83° W LONGITUDE		WESTAR V 123° W LONGITUDE		GALAXY I 134° W LONGITUDE	
Frequency (MHz)	TRP#	Program	TRP#	Program	TRP#	Program	TRP#	Program
3720	1	Nickelodeon/ARTS	1	SIN	1D	Bluemax	1	HBO
3740	2	PTL	2	Bravo FNN	1X	CBS/Fox	2	Group W/ Westinghouse
3760	3	WGN	3	SPN	2D	WOR-TV	3	HBO
3780	4	Spotlight	4	Home Sports Ent.	2X		4	
3800	5	The Movie Channel	5		3D		5	Times Mirror
3820	6	WTBS	6	Eros	3X	Dow Jones	6	SIN
3840	7	ESPN	7	NCN	4D	CBS Television Net.	7	Turner Broadcasting
3860	8	CBN Cable Net.	8		4X	SNC (A)	8	Group W/ Westinghouse
3880	9	USA Cable Net.	9		5D		9	
3900	10	Showtime (W)	10		5X	Disney Channel (W)	10	Times Mirror
3920	11	MTV	11		6D	SNC (National)	11	
3940	12	Showtime (E)	12	The Playboy Chan.	6X	Disney Channel (E)	12	Group W/ Westinghouse
3960	13	HBO (W)	13		7D		13	
3980	14	CNN	14		7X	SNC (B)	14	Viacom International
4000	15	CNN Headline News	15	Biznet	8D	SNC 2	15	
4020	16	ASCN HTN Plus/NJT	16		8X	SNC (Backhand)	16	Viacom International
4040	17	Cable Health Net.	17	TBN	9D	The Nashville Net.	17	HBO
4060	18	Reuters Monitor Service/EWTN	18	Time Video Info. Services	9X	SNC (C)	18	Turner Broadcasting
4080	19	C-SPAN	19		10D		19	HBO
4100	20	Cinemax (E)	20		10X	The American Net.	20	SIN
4120	21	The Weather	21		11D	Spotlight (W)	21	HBO
4140	22	MSN/Daytime	22	ABC Remote Feeds	11X	Selec TV	22	Group W/ Westinghouse
4160	23	Cinemax (W)	23	Galavision	12D	Daytime/ARTS	23	HBO
4180	24	HBO (E)	24	NBC Remote Feeds	12X	B.E.T.	24	

NOTE: Polarization on SATCOM IIIR and IV:
Odd transponders are vertical; even transponders are horizontal.
Polarization on Westar V or Galaxy I is opposite of above:
Odd or "D" transponders are horizontal; even or "X" transponders are vertical.

Figure 3. Transponder Assignments

LNA (Low-Noise Amplifier)

NOTE: Scientific-Atlanta, Inc. manufactures both low-noise amplifiers (LNAs) and low-noise converters (LNCs). The coaxial cable, coaxial connectors, and power dividers discussed in this chapter apply to systems using LNAs. For specific information regarding LNCs and the coaxial cable, connectors and signal splitters used with them, refer to the section on low-noise converters later in this paper.

Purpose. The purpose of the LNA is to amplify the signal from the antenna to a level adequate to overcome attenuation from cable and splitter losses and still be within the input range of the video receiver. At the same time, it must introduce a minimum level of noise to the system. Therefore, the two main specifications for a LNA are gain and noise figure. The standard LNA provided by Scientific-Atlanta has a gain of 50 dB (±3 dB) and a noise figure of 1.5 dB (also expressed as a noise temperature of 120K). Optional LNAs are also available with the following noise characteristics: 100K (1.3 dB) 90K (1.2 dB) and 80K (1.1 dB). All LNAs manufactured by Scientific-Atlanta are available in either 50-dB or 60-dB gain versions.

Installation. The LNA is bolted directly to the OMT, with no "right side up" orientation necessary. All Scientific-Atlanta antennas with a Cassegrain feed system are equipped with a feed horn which must be pressurized. Because the feed cavity of the LNA is airtight, there is no need for a mylar "window" between the LNA and the OMT. The LNA gasket supplied must be properly installed in order to achieve a proper seal. Figure 5 shows two (2) LNAs being installed on a 3M feed equipped with an OMT.

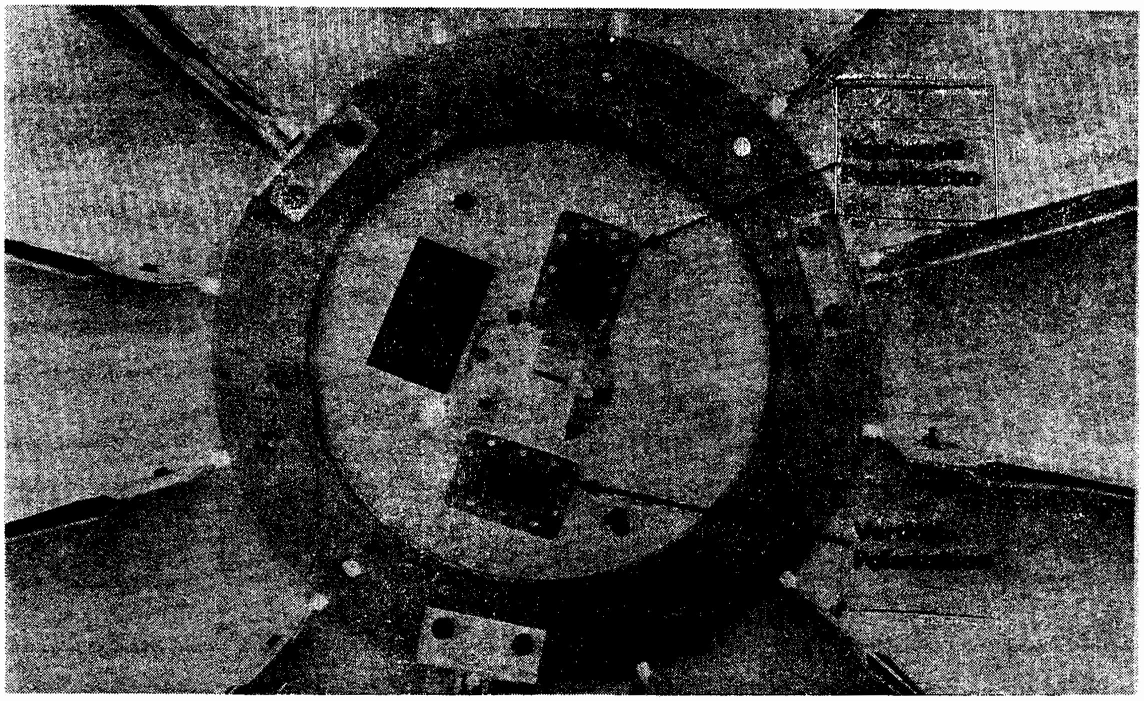


Figure 4. OMT Oriented for Maximum Signal at a Particular Site

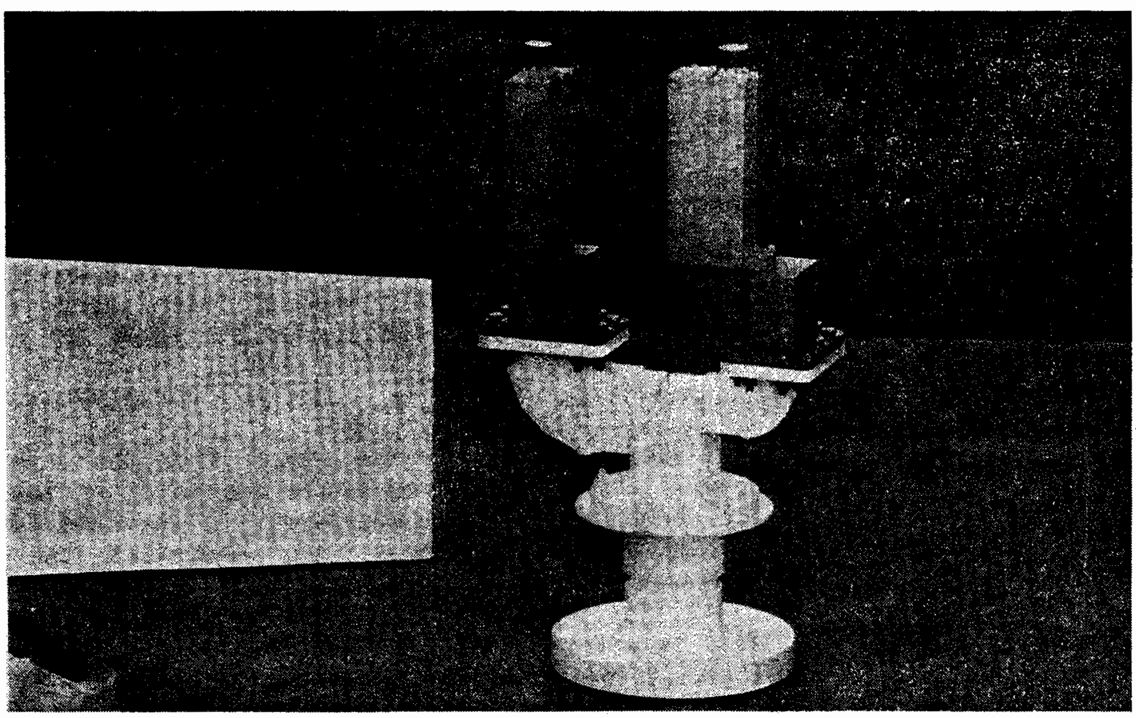


Figure 5. Two LNAs Installed on 3-Meter Feed

LNA Power. Currently there are two methods for powering the LNAs. They are "DC only" and "Tri-Power." The "DC only" LNA is the simplest and most commonly used method. The LNA is powered by +15V dc which is introduced through the RF coaxial cable. A Power Inserter is placed in-line with the RF output of the LNA (generally inside the headend building) and on the antenna side of any power dividers and/or coax switches. The Power Inserter is wired into the AUXILIARY connector on the back of any Scientific-Atlanta 6602 or 7500 Series Video Receiver (PIN 2:15V dc, PIN 3: GND). The receiver provides the +15V dc to power the LNA through this connection. If desired, the Power Inserter may be connected directly to an external power supply. Figure 6 is a detailed schematic of a "DC only" wiring configuration.

Tri-Powered LNAs. The tri-powered LNA is powered by either 110V ac, +15 to + 28V dc, or -15 to -28V dc. The selected voltage is introduced to the LNA through a separate power connector on the LNA. This necessitates the installation of a separate power cable.

Coaxial Jumpers

The RG-214 4.5-foot coaxial jumpers illustrated in Figure 6 are used to interconnect the RF signal from the LNA to the various parts of the system such as the main coaxial cables, power dividers, etc. They permit increased flexibility over the low-loss coaxial cables and facilitate the wiring of multiple receivers. While Scientific-Atlanta also offers these jumper cables in lengths of 9 feet and 15 feet, it is important to note that the signal attenuation is as high as 25 dB per 100 feet at the 4 GHz frequency being used.

Coaxial Cable/Connections

Purpose of Cable. The coaxial cable is used to transport the signal from the LNA to the receiver location (such as a CATV headend). The two types of cable currently supplied by Scientific-Atlanta are 7/8-inch O.D. air dielectric coaxial cable and 1/2-inch O.D. foam core coaxial cable. Both of these cables have an impedance of 50 ohms. The attenuation of the 7/8-inch cable is 3 dB per 100 feet, while the 1/2-inch cable has a loss of 8.5 dB per 100 feet. Keeping in mind that combined cable/splitter losses should not exceed 20 dB, 1/2-inch coax is not recommended for cable runs in excess of 100 feet. When using 7/8-inch cable, which has an air dielectric, the cable must be pressurized using a suitable unit such as Scientific-Atlanta's pressurization unit (described later). The 7/8-inch coaxial cable may be used for cable runs of up to 250 feet in most circumstances before additional amplifiers are required.

Purpose of Connectors. The coaxial connectors provide a transition from the coaxial cable to an "N" type female connector which matches the "N" type male connectors on the 4.5-foot coaxial jumper cables. In the case of 7/8-inch air dielectric coax, the connector also has provision for a 1/8-inch NPT air fitting to permit pressurization of the cable.

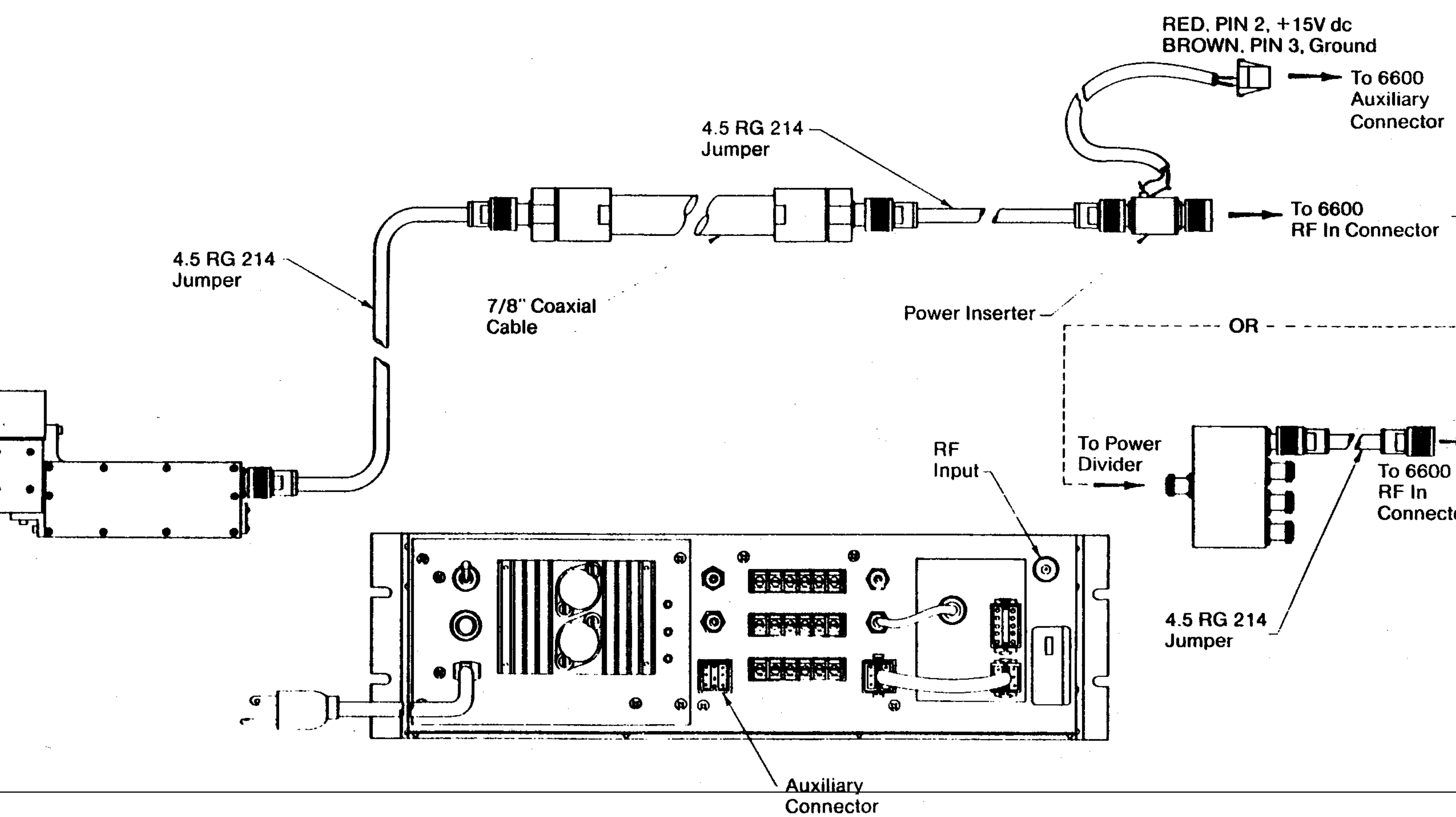


Figure 6. LNA DC Only Wiring Configuration

Installation of Cable. The cable may be installed using standard installation techniques, with the following precautions. Due to the high frequencies involved, extreme caution is required to avoid any sharp bends, kinks, dents, or other damage to the cable. A minimum bending radius of 10 inches for 7/8-inch cable and 5 inches for 1/2-inch cable must be observed. For these reasons we recommend that the cable be installed in 4-inch diameter PVC conduit and that, regardless of conduit diameter, only wide sweep bends be used.

Installation of Connectors. Installation instructions are provided with each connector. Please read these instructions carefully before attempting to install the connector. Due to the high frequencies being utilized, proper connector installation is essential for optimum system performance.

Power Dividers

Purpose. The power divider for your earth station is basically a RF splitter. Despite its name, a "power divider" is not a power-passing device in that no ac or dc voltage should be fed into or through the device. The purpose of the divider is to provide inputs to multiple receivers from one LNA. If you are planning to operate more than one receiver on each polarization, you will need a separate power divider on each polarization. Power dividers are available from Scientific-Atlanta in three configurations.

P/N	Description	Signal Attenuation
143036	4-Port Power Divider	7.5 dB
143035	8-Port Power Divider	10.5 dB
143034	16-Port Power Divider	13.5 dB

These devices provide outputs for 4, 8 and 16 receivers, respectively. Figure 7 is a photograph of a 4-port power divider.

Installation. Figure 2 shows the location of the power divider in the TVRO system. The input and output connections are "N" type female and accommodate the 4.5-foot coaxial jumper cable. The power divider should be secured in such a manner as to provide a minimum of strain to the coax jumpers attached to it. Due to the high isolation between ports on the power divider, no termination is necessary for any unused ports.

Automatic Polarization Relay

The automatic polarization relay will operate with Scientific-Atlanta Model 6602 and 7500 Video Receivers, providing automatic switching between horizontal and vertical polarization feeds. The unit has three "N"-type female RF connections and comes pre-wired to a multi-pin plug that connects to the 10-pin program plug on the rear of the receiver.

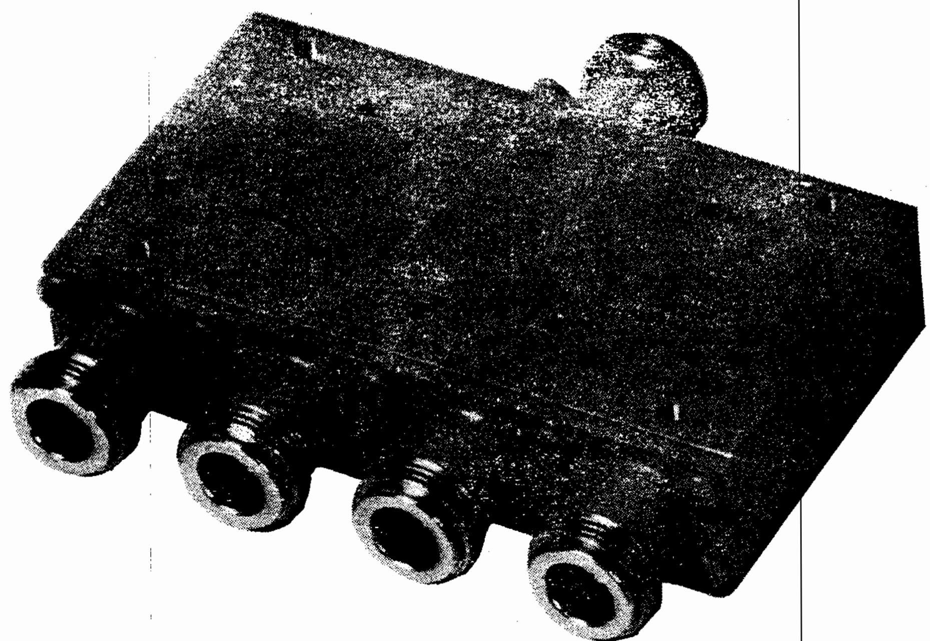


Figure 7. Four-Port Power Divider

Figure 8 illustrates the switch and identifies the proper connections.

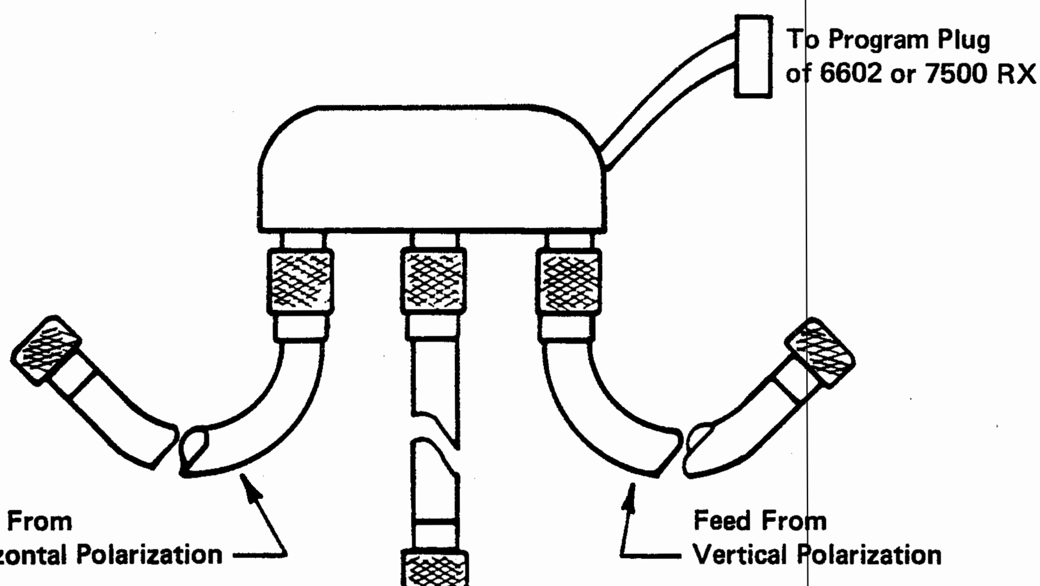


Figure 8. Relay Connections

In operation the switch selects between the two polarization feeds alternatively and in accordance with the frequency selecting dial on the front of the receiver, thus providing the correct input for the frequency selected. In the case of the 6602 and 7500 receivers, the appropriate input is also provided for a frequency selected via the remote interface.

Signal Attenuation

Due to the extremely low signal levels involved in earth station applications and the high attenuation at the frequencies used, care must be exercised to avoid excessive signal losses. As a general rule of thumb, signal losses should not exceed 20 dB between LNA and receiver, when using a 50-dB gain LNA. Using the schematic in Figure 2 and assigning the 7/8-inch cable a length of 100 feet, we can calculate our signal losses as follows:

Qty	Item	Attenuation Loss
1	4.5-foot Coax Jumper (25 dB per 100 feet)	1.125 dB
100 feet	7/8-inch Coax Cable (3 dB per 100 feet)	3 dB
1	4.5-foot Jumper	1.125 dB
1	4-Port Power Divider	7.5 dB
1	4.5-foot Jumper	1.125 dB
	Total Loss	13.875 dB

As can be seen, the losses add up rapidly. As a matter of fact, just placing a 16-port power divider in the system instead of a 4-port power divider will put us just over our "General" 20-dB maximum permitted loss.

If excessive losses are unavoidable due to physical requirements such as antenna location, a post amplifier may be installed with minimum effect on system performance.

LNC (Low Noise Converter)

Purpose. The purpose of the LNC subsystem is identical to that of an LNA system, yet with advantages for cable TV operators.

The principal advantage of a block conversion system is lower cost. Converting from a microwave frequency to UHF at the antenna significantly reduces the use of expensive microwave components and cable. A conventional system requires a separate microwave downconverter for each channel received. In the block downconversion system, this process is performed only once for all twelve channels of a given polarization, resulting in reduced microwave component costs. Less expensive UHF cable and "F" connectors can be used between the LNC and the 6650 Receiver, resulting in substantial cost savings in this area also.

Installation. The LNC and 6650 Receivers are installed in much the same way as the LNA subsystem. A block diagram is shown in Figure 9.

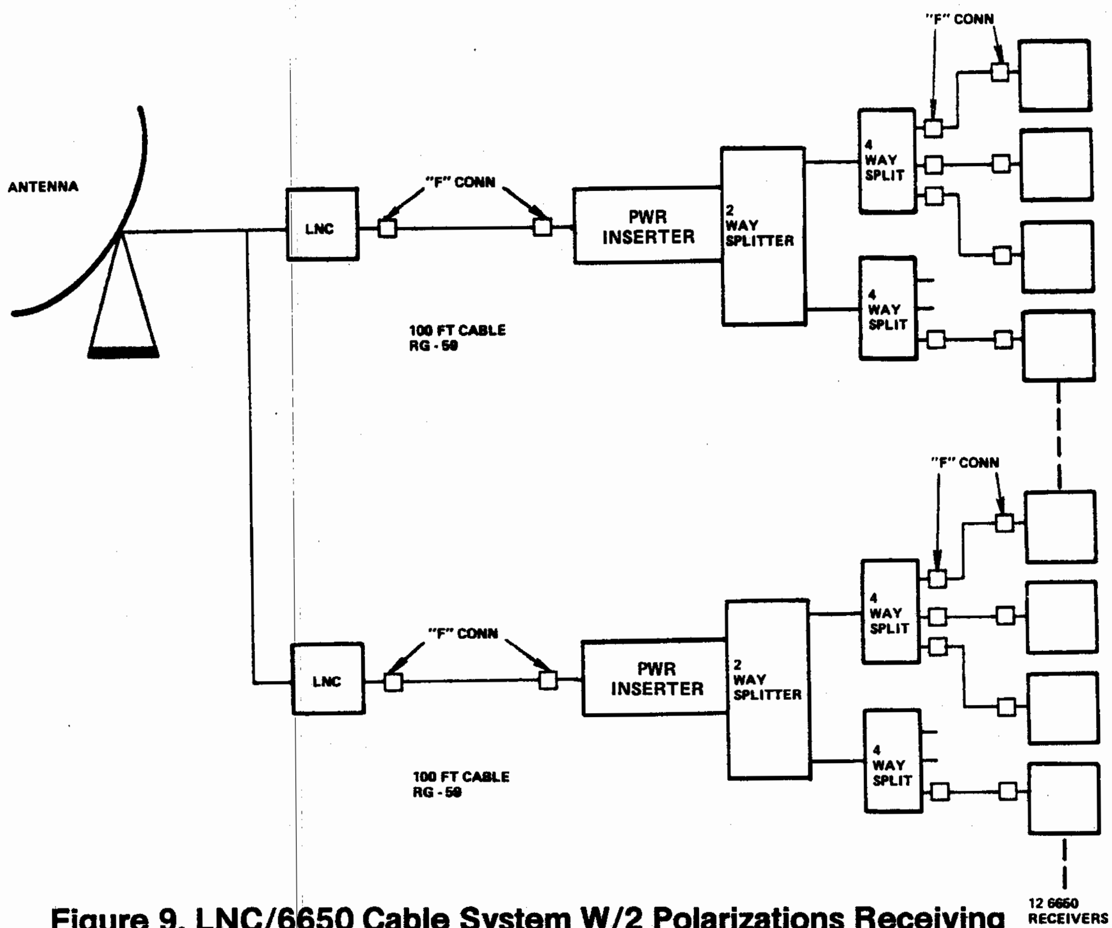


Figure 9. LNC/6650 Cable System W/2 Polarizations Receiving 6 Channels Each

Pressurization Unit

Purpose. The pressurization unit is used to prevent moisture from accumulating in the feed horn and, if applicable, the air dielectric cable in the earth station system. This is accomplished by first drying the air and second by maintaining a positive pressure of 1 PSIG in the system. Due to the low signal levels involved, a small amount of moisture can have a significant effect on system performance. Care should be taken to ensure that your pressurization system is operating properly. Pressurization is not required for antennas with prime focus feed systems unless air dielectric coax is used.

Operation. Figure 10 illustrates the pressurization unit available from Scientific-Atlanta. Air is introduced to the system via the compressor unit and stored in the holding tank, Item 2. The pressure in the holding tank is maintained at between 20 PSIG and 40 PSIG by means of the pressure switch, Item 20, which activates the compressor motor. Pressurized air is then fed from the holding tank to the desiccant tank, Item 3. The desiccant crystals remove the moisture from the air and should be replaced or dried as necessary. A color chart located on the desiccant bowl indicates when desiccant replacement becomes necessary. From the desiccant bowl, dried air is passed through the adjustable pressure regulator, Item 4, and into the system.

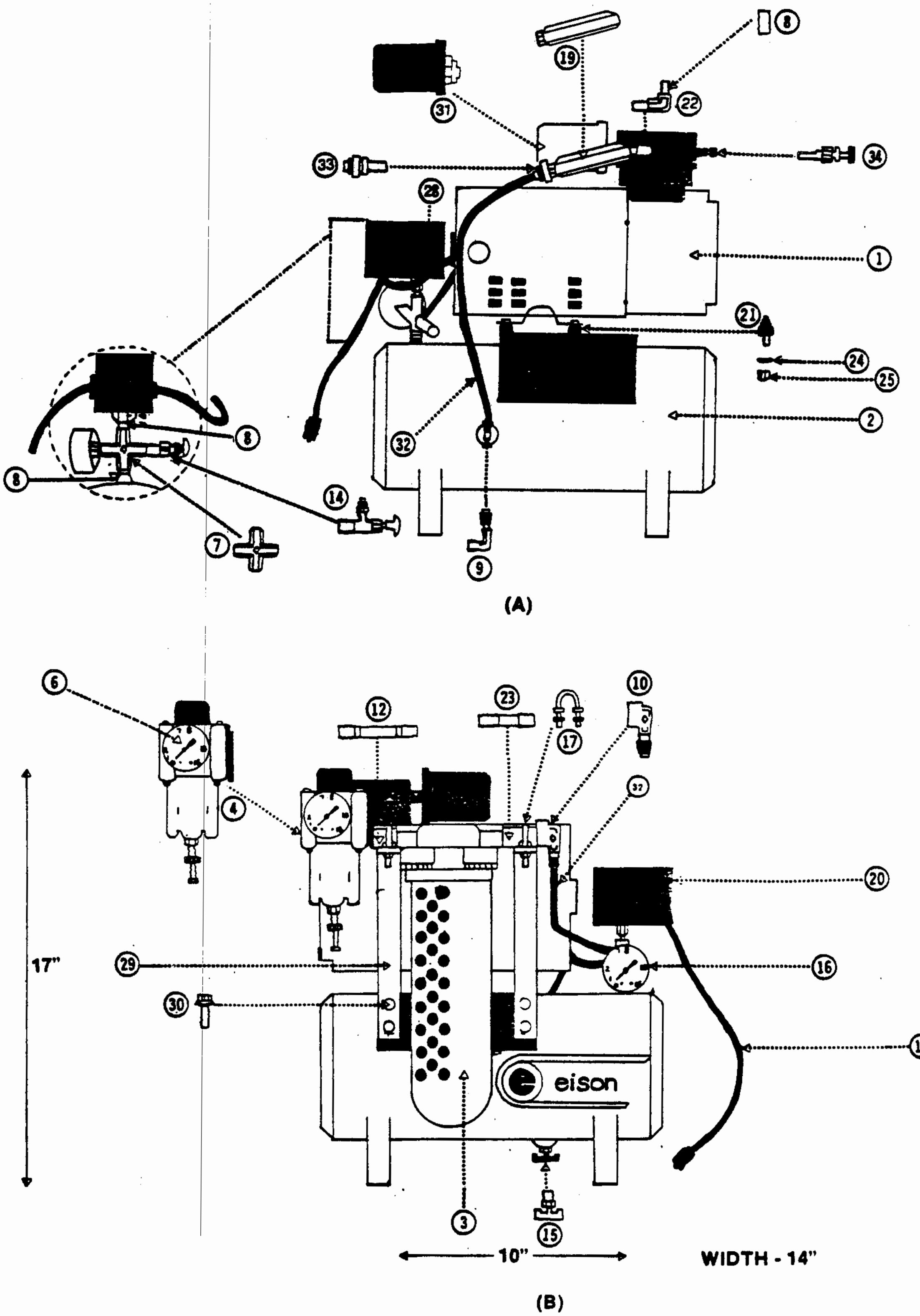


Figure 10. Scientific-Atlanta Pressurization Unit (View A & B)

Installation. Remove the unit from the shipping carton and check for any visible damage. Should any damage be evident, notify the carrier and save the shipping carton.

1. Remove the desiccant bowl from holder and fill with the desiccant packed separately in a one-pound can. (There is more than enough desiccant supplied so care should be taken to avoid overfilling the container.)

2. Replace bowl after making sure that the "O" ring, located on the holder, is clean and correctly seated.

3. Close the check valve, Item 14, and insert the ac power cord into a 110V ac, 60-Hz outlet. The compressor motor should activate and fill the storage tank to a pressure of approximately 40 PSIG. This can be monitored by observing the pressure gauge, Item 16. The compressor should shut down at this point.

4. Establish a temporary seal by placing a finger over the output of the regulator (Item 4) and open the check valve. Adjust the pressure, as indicated on the regulator pressure gauge, to 1 PSIG by means of the adjustment screw.

5. Remove and replace the finger several times and make sure that pressure returns to 1 PSIG.

6. Lock down the adjustment screw by means of the lock nut and connect the regulator output to the cable and/or feed horn with the air line and connectors provided. Figure 11 shows a connection diagram for a dual earth station installation. Connections for a single antenna would be identical, with the omission of the connections to the second earth station.

7. Check the system for leaks and repair as necessary. Compressor pump should cycle no more frequently than once every four hours.

8. Approximately once a month, drain the holding tank of accumulated moisture by means of the pet cock, Item 15.

Antenna Pointing

Having installed the items described, one is now ready to align the antenna to the satellite. This can be a simple operation as long as some basic steps are followed. Keep in mind that one is trying to "aim" at a target over 23,000 miles away, and that a "hit or miss" approach will be frustrating at best. For additional information on antenna pointing, refer to the paper, "Earth Station Geometry."

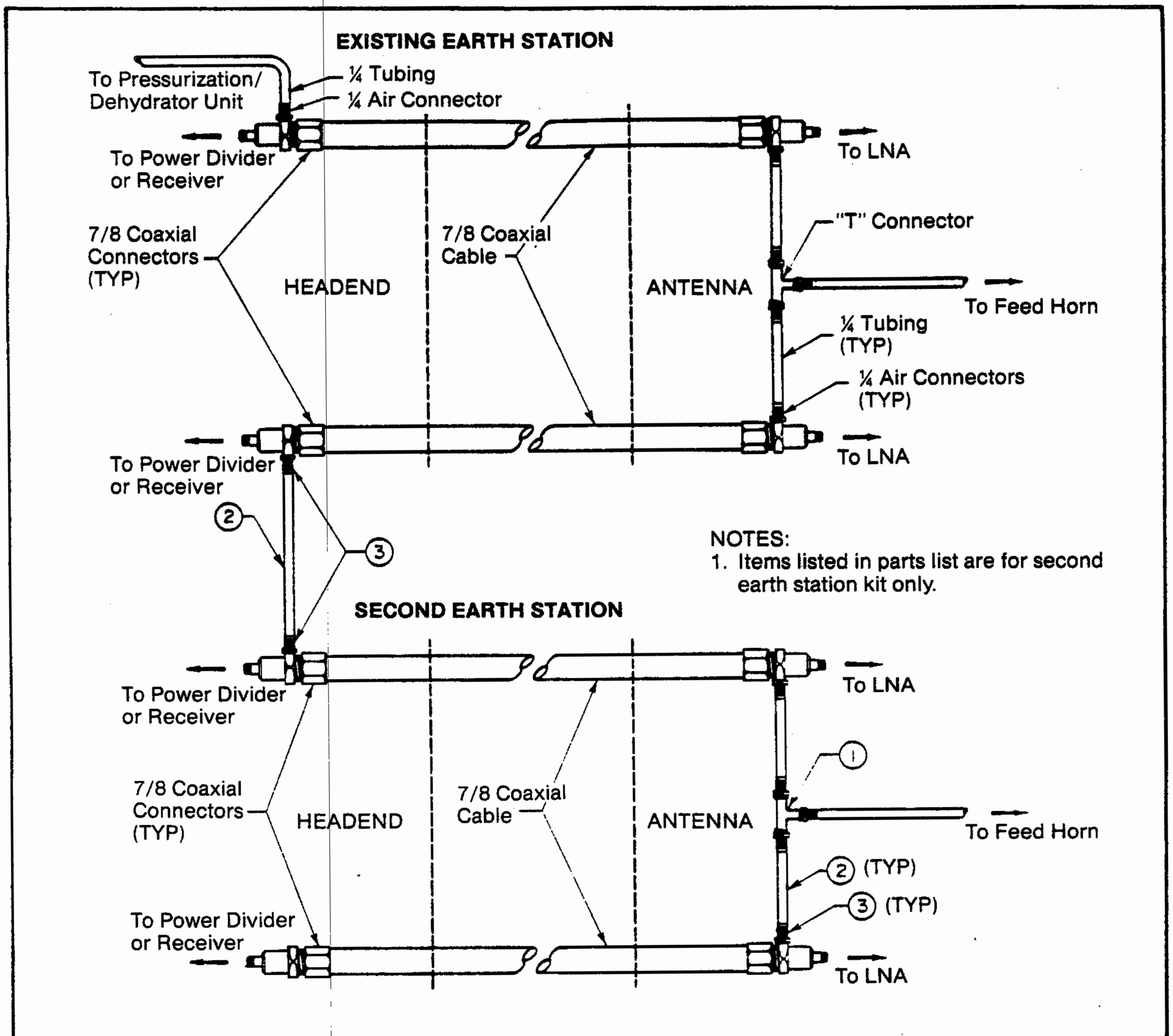


Figure 11. Pressurization Interconnection Diagram

From the Computers of:
SCIENTIFIC-ATLANTA, INC.
Video Communications Division

22-Jul-82 17:54

Look angle and centerline calculation for a 5-meter antenna to be installed in: Atlanta, Georgia.

Site Coordinates:	33 deg	45 min	0 sec	latitude
	84 deg	25 min	0 sec	longitude

Earth station look angles for: Atlanta, Georgia

Satellite Name	Satellite Longitude	Elevation Angle	Azimuth Heading True
SPACENET 1	70	47.82	155.18
GALAXY 2	74	49.2	161.7
ADV. WESTAR	79	50.33	170.32
SATCOM 4	83	50.73	177.46
COMSTAR D3	87	50.67	184.65
WESTAR 3	91	50.13	191.74
COMSTAR D2	95	49.14	198.59
SBS 2	97	48.49	201.9
WESTAR 4	99	47.74	205.1
ANIK-1	104	45.49	212.64
SBS 1	106	44.44	215.46
ANIK-2	109	42.75	219.47
ANIK-3	114	39.62	225.62
F2/SPACENET 2	119	36.19	231.14
WESTAR 5/2	123	33.28	235.15
COMSTAR D4	127	30.24	238.85
SATCOM 3	131	27.12	242.27
F1/Galaxy 1	135	23.92	245.46
SATCOM 1-R	139	20.67	248.44
SATCOM 2-R	143	17.39	251.26

5-METER FOUNDATION CENTER LINE IS 213.22 DEGREES FROM TRUE NORTH.

Figure 12. Look Angles

Look Angles. Regardless of the type of mount on your antenna, the desired result is to have the antenna pointing correctly at the satellite. This direction is known as the look angle and is expressed by two coordinates: elevation angle, in degrees above the horizon; and azimuth angle (compass heading), in degrees from true north. These look angles are provided for each earth station order and are usually mailed separately and in advance of the antenna. From any one location and for any one satellite, these look angles remain the same for any type of antenna and can therefore be used to cross-check other types of pointing information which may have been supplied for a particular antenna. Figure 12 shows a computer run for the look angles in Atlanta, Georgia.

Specific pointing information for each antenna is provided in the manual for that particular antenna. However, the following "tips" may prove useful.

1. Any pointers, counters, indicators, etc., provided with your antenna will only be as accurate as the center line of your foundation. Therefore, you should use care to make sure that the foundation center line for your site as provided by Scientific-Atlanta is determined as accurately as possible.

2. A gravity-operated device for determining elevation above the horizon will greatly facilitate setting the elevation angle. (This device can range from a protractor with a piece of string and a rock, to an inclinometer available at most hardware stores.)

3. If the pad center line is being determined by a surveyor, he can also identify and mark the azimuth headings for various satellites, given the look angle information available from Scientific-Atlanta. This usually adds little, if any, cost to his service and will aid in initial pointing.

4. Figure 13 is a chart for determining the look angles from any given location. Instructions for the chart follow.

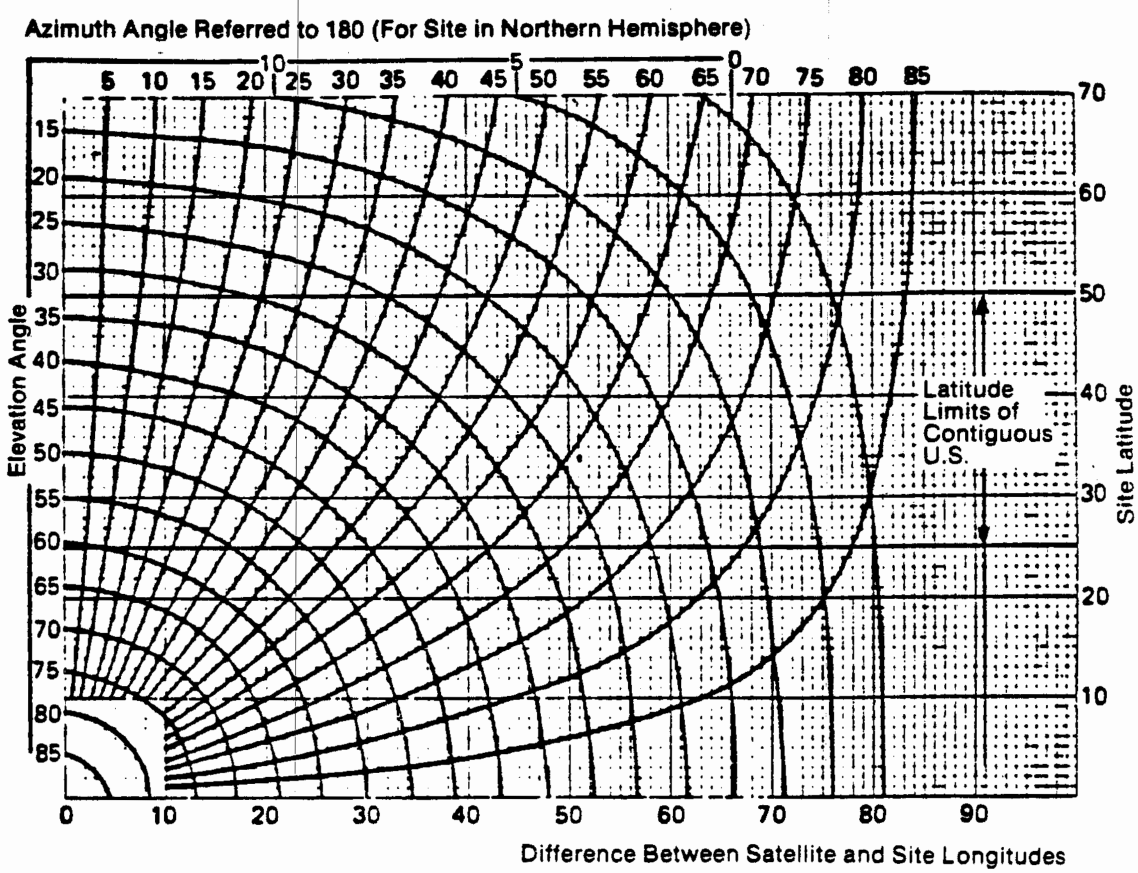


Figure 13. Universal Azimuth-Elevation Look Angles

Instructions.

1) Subtract site longitude from desired satellite longitude and plot this value (A) on the bottom scale of the chart.

2) Plot the site latitude (B) using the scale on the right side of the chart.

3) At the point where A and B intersect, determine the elevation angle and difference in azimuth using their respective graphs (the curved lines are used for this process).

4) If the value obtained in step 1 is a positive number, add the difference in azimuth from step 3 to 180° to obtain the azimuth heading from true north. If the value is a negative number, subtract the difference in step 3 from 180°.

Satellite Acquisition

If the antenna has been correctly pointed and all components have been properly installed, one should be receiving signals from the satellite. In practice, a small amount of antenna adjustment will be necessary to achieve optimum performance.

Presence of signal from the satellite may be observed on the C/N meter on the front of any 6600 or 7500 Series Receiver. In order for this meter to respond to variations in signal strength, the AGC/MGC switch on the receiver must be in the AGC position.

Presence of video on the satellite can be detected by connecting a video monitor to the video output of the receiver or by connecting a TV set via an RF modulator to the receiver.

Adjust the antenna as necessary for maximum signal using the instructions provided with each antenna.

Troubleshooting

This chapter is intended to provide assistance if there is little or no signal from the satellite-installed equipment in accordance with the instructions described herein. It is assumed that a 6600 Series Receiver is used and that it is in the AGC mode; also needed are a video monitor, or a modulator and TV set for viewing the actual picture.

Symptom	
No Signal No Picture	a. Check polarization of feed horn and ensure that frequency selected is on same polarization as input from the antenna.
	b. Check the ac input to receiver and dc input to LNA.

c. Adjust zero on noise control on front of 6600 RX for a positive indication (e.g., +5), and remove input to downconverter. Meter should then move full scale to left (less than "0"). If meter remains at +5 or cannot be set at +5, check the IF amplifier module in the receiver and make sure it is properly seated. If condition persists, replace LNA and/or check all cable connections from LNA to receiver.

Positive Signal Level No Picture	a. If using a 6602 RX with only one No Picture polarization input to the receiver, change frequency to a transponder on the opposite polarization. If C/N meter drops 7 to 10 dB, check connections to TV, modulator or video monitor. If little change occurs in signal strength, pan antenna and check again for change in signal level. If no change is detected, it is likely the antenna is not peaked on the satellite. Check antenna look angle and/or pan the antenna while observing for increase in signal level. (Note: Make sure AGC/MGC selector switch on receiver is in AGC position.)
Positive Signal Level Noisy Picture	a. Make sure AGC/MGC switch is in AGC position and adjust zero on noise control for a reading of 10 to 12 dB on C/N meter. b. Check to ensure that antenna is positioned for maximum signal level. As antenna is panned, the change in C/N meter level should indicate a single main beam with small peaks on either side. If two equal beams are indicated, polarization is incorrectly adjusted, or antenna is peaked on a sidelobe. c. Peak polarization adjustment for maximum signal level. d. Change receiver (if available) and repeat step a. e. Check all cable connections. (If cable becomes suspect, move receiver and monitor to the LNA via a 4.5-foot jumper.) f. Change LNA (if available) and repeat step a.

For additional troubleshooting guides, consult the instruction manual for your receiver.

Video Receive-Only Earth Station Performance Verification

Alex Best

Abstract

The proof of performance of a typical satellite video receive station consists of a series of RF and video tests which are compared to the specified parameters. The proof-of-performance test records provide a valuable guideline for routine maintenance and troubleshooting. Troubleshooting a satellite video receive station requires a basic understanding of the operation of the system components, receiver, LNA, etc., and how the component parameters affect the system performance. This paper discusses the performance evaluation tests, troubleshooting and maintenance.

Performance Verification

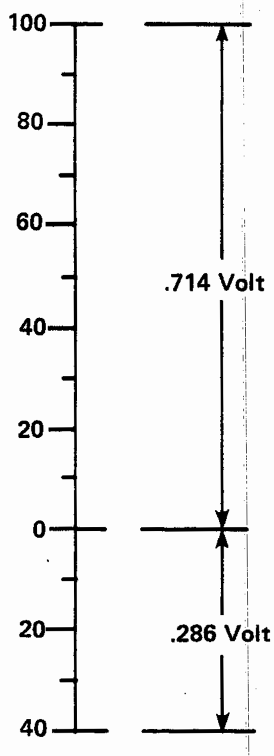
Introduction

Once the antenna is assembled, the LNA and cable installed, the system turned on, and the satellite located, the system engineer is faced with the task of determining that the system is operating properly; in other words, evaluating the performance of the station. This evaluation can be as simple as looking at the receiver output on a TV monitor or as complex as available test equipment will allow.

The procedures which follow describe the tests currently performed by Scientific-Atlanta field personnel to evaluate the performance of a satellite video receive station.

Test Waveforms

All waveform measurement techniques described in this paper are based on the IRE scale units of measurement (see Figure 1). The waveform technology used throughout this paper is in accordance with the definition shown in Figure 2, wherein the standard composite color video signal is defined.



Waveform Terminology

- A — The peak-to-peak amplitude of the composite color video signal
- B — The difference between black level and blanking level (set-up)
- C — The peak-to-peak amplitude of the color burst
- L — Luminance signal - nominal value
- M — Monochrome video signal peak-to-peak amplitude (M=L+S)
- S — Synchronizing signal - amplitude
- T_b — Duration of breezeway
- T_{si} — Duration of line blanking period
- T_{sy} — Duration of line synchronizing pulse
- T_u — Duration of active line period

Figure 1. The IRE Scale Units

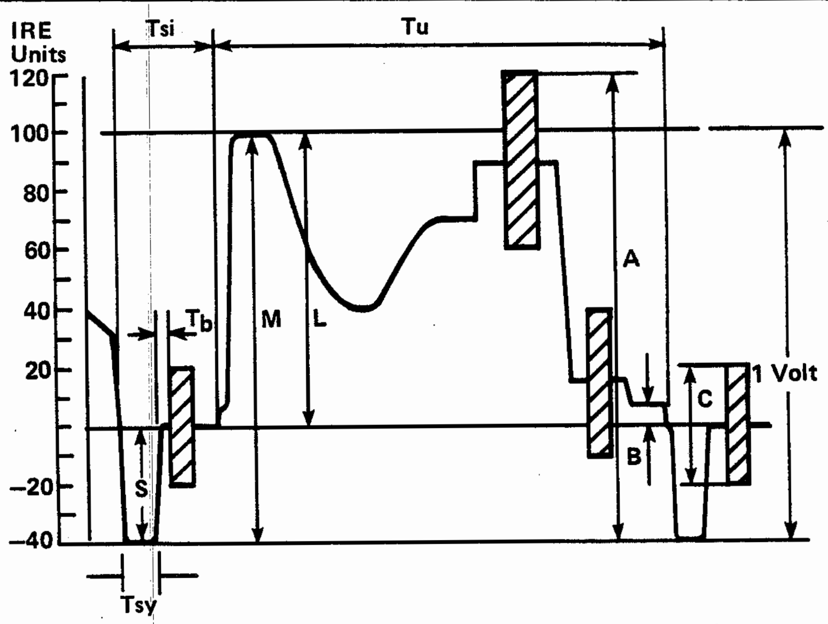


Figure 2. The Standard Composite Color Video Signal

The two principal test signals that are required to conduct the various measurements described in this report are:

a. The composite test signal shown in Figure 3, which consists of a line bar, a 2T pulse, a modulated 12.5T pulse, and a 5-riser modulated staircase signal.

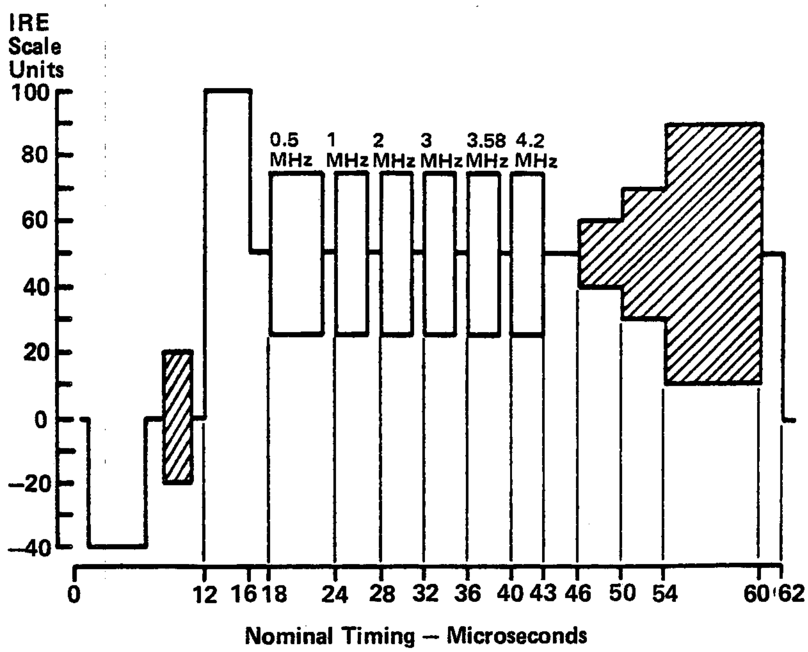
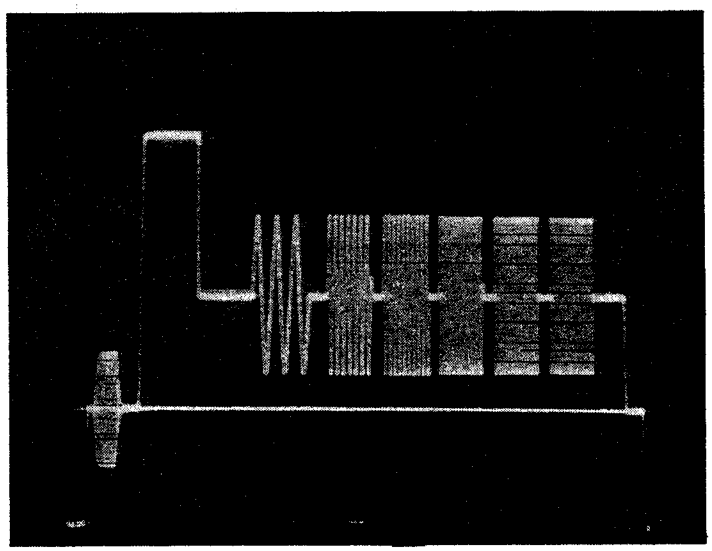


Figure 4. The Combination Test Signal



It should be noted that except where full-field test signals are essential to the measurement of a particular parameter, the measuring technique for each parameter is the same for both vertical interval test signals (VITS) and full-field test signals. Furthermore, the performance objectives apply irrespective of the average picture level (APL) within the APL range of 10% to 90%. This is an important point to remember when making VITS measurements, particularly during program transmission periods where control cannot be exercised over the APL value of picture signal. Many of the parameters can be markedly affected by APL variations, and accordingly, the operator should allow sufficient time when making VITS measurements to ensure a good portion of the APL range is explored by the picture signal.

Test Equipment Required

Tektronix 520A Vector Scope or, equivalent
Tektronix 1480 Waveform Monitor, or equivalent
Tektronix 147A NTSC Signal Generator, or equivalent
Tektronix CCIR Random Noise Low Pass Filter, or equivalent
Hewlett-Packard 435A RMS Power Meter, or equivalent
Hewlett-Packard 8558 Spectrum Analyzer, or equivalent
Hewlett-Packard 334A Audio Distortion Analyzer, or equivalent

Tests Performed

Main IF Carrier-to-Noise Ratio
CCIR Weighted Signal-to-Noise Ratio
Reduced Amplitude Video Response
"K" Factor (K_2T)
Luminance-Chrominance Delay
Differential Gain
Differential Phase
Audio Test Tone Signal-to-Noise Ratio
Audio Test Tone Distortion

Test Procedures

Main IF Carrier-to-Noise Ratio. This test verifies the performance of the RF and IF portions of the system; this is apparent from the expression for C/N:

$$C/N = (EIRP - LP + G/T - D) - 10 \log B_{IF} \qquad [dB]$$

where:

EIRP = Satellite effective isotropic radiated power in dBW

L_s = Path loss (free space) in dB

K = Boltzmann's constant = -168.6 dBW/MHz/K

B_{IF} = Effective IF Noise Bandwidth in MHz

G/T = System figure of merit dB/K = Gain -10 log t_s

G_a = Antenna gain in dB
t_s = system noise temperature in K
t_s = $t_a + t_1 + t_c/g_1 + t_r/(g_1 \times g_c)$
t_a = Antenna noise temperature in K
t_1 = LNA noise temperature in K
t_c = Cable noise temperature in K
t_r = Receiver noise temperature in K
g_c = Cable gain ratio
g_1 = LNA gain ratio

Examining the parameters which affect C/N, it becomes obvious that this is perhaps the single most important measurement to be considered. The C/N ratio is measured as follows:

a. Set the receiver AGC/MGC switch to MGC position.

b. Connect spectrum analyzer to filtered 70-MHz IF monitor port.

c. Raise antenna elevation until there is no input signal to the receiver as shown by the spectrum analyzer.

d. Connect the HP435 power meter to the 70-MHz filtered monitor port. Measure the noise power at this port and record it.

e. Lower the antenna elevation until an absolute peak is reached as shown on the power meter, and record this reading as carrier plus noise power.

f. (Carrier + Noise Power) - (Noise Power), where each quantity in parenthesis is in dBW is the (carrier + noise)-to-noise ratio $C+N/_N$ in dB.

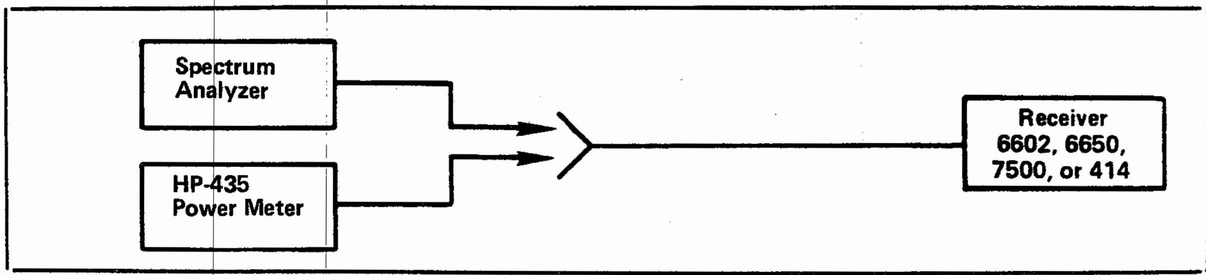
g. Compute carrier-to-noise ratio (C/N) as shown in the following example:

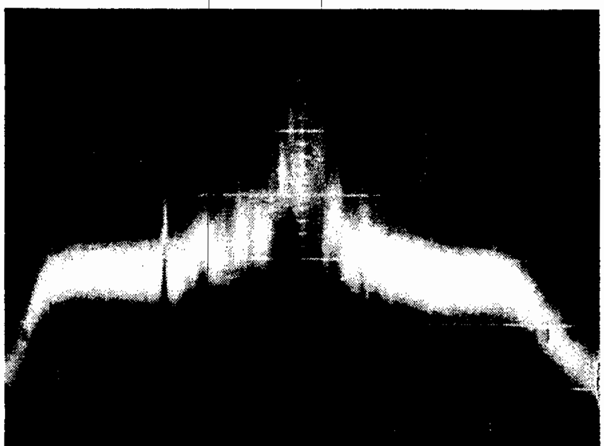
$$C+N/_N = 16 \text{ dB}$$

$$C/N = 10 \log \left[(\log^{-1}(16/10) - 1\right]$$

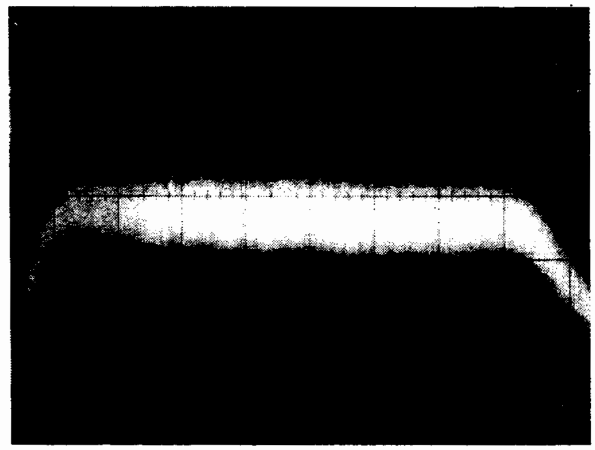
$$C/N = 15.89 \text{ dB}$$

Equipment set up for C/N measurement:





Spectrum Analyzer Display
70 MHz Center Frequency
Showing Received FM Signal



Spectrum Analyzer Display
70 MHz Center Frequency
Showing No Received Signal

CCIR Weighted Video Signal-to-Noise. This test verifies the performance of the up- and downlink video equipment and that of the program data 17 VITS--vertical internal test signals are not inserted at the uplink. The expression used for CCIR signal-to-noise (FM) is:

$$S/N\ (\text{video}) = \frac{C}{N_o} \frac{12(\Delta F_s)^2}{b_n{}^3}$$

where:

C = Carrier power in watts.

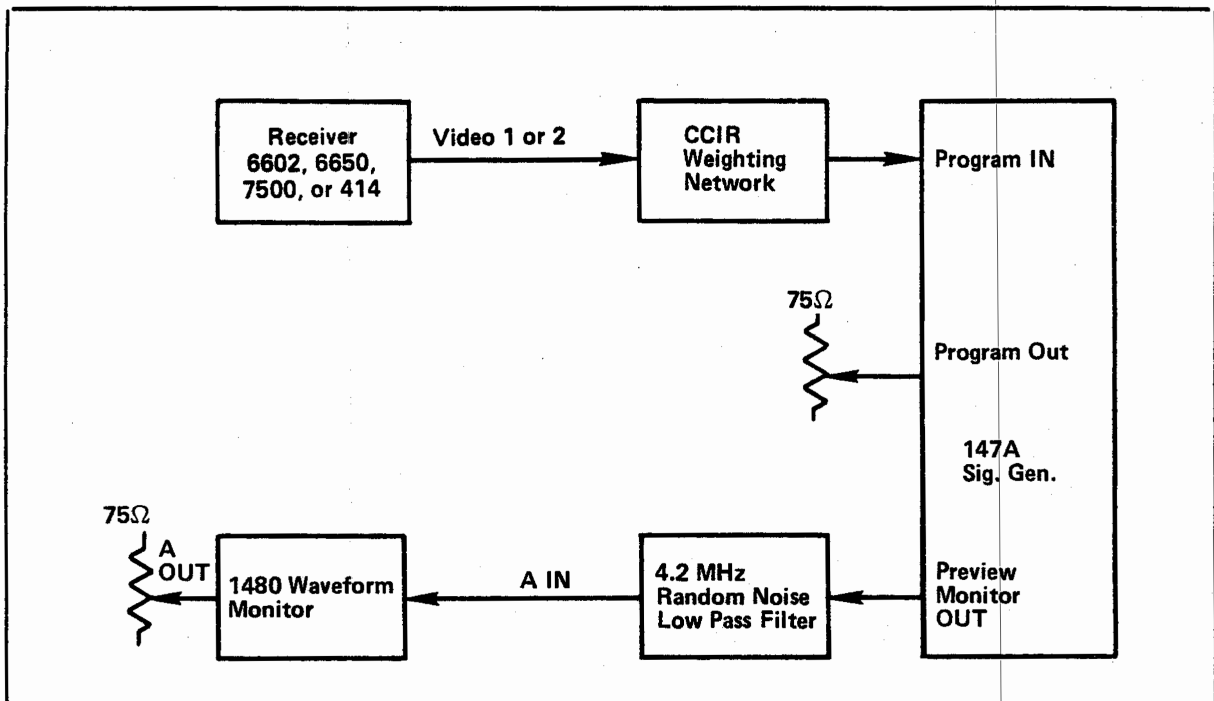
N_o = Noise power density at that point in the system where C is measured.

ΔF_s = Half the peak-to-peak deviation caused by that portion of the video waveform defined as the "signal".

b_n = Noise bandwidth of the baseband filter function which represents the combination of the deemphasis network, measurement bandlimiting filter, and weighting network with respect to triangular noise.

Examining the parameters which affect video signal-to-noise ratio, one can readily see that the uplink operating parameters EIRP, deviation, etc., and the receive station equipment performance--antenna through video amplifiers--are verified. The video signal-to-noise ratio (CCIR) is measured as follows:

a. Interconnect the receiver and test gear as shown in the block diagram.



b. Set 147A controls as listed below:

Function Switch	-	Noise
Local-Remote	-	Local
Gain-Variable/Unity	-	Unity
Program/Preview	-	Preview
Noise Attenuators	-	40 dB
Delete half line/full line	-	half line

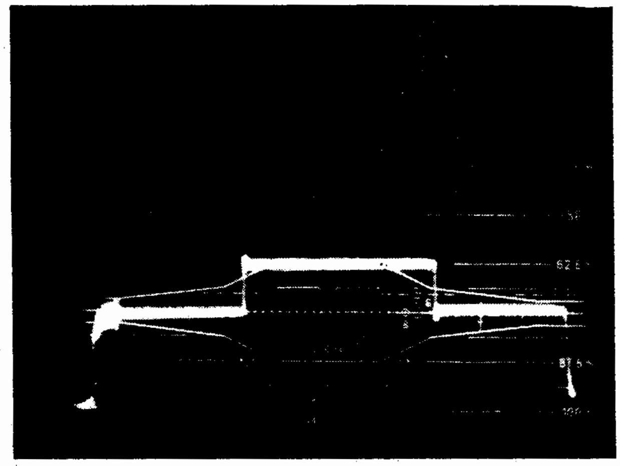
c. Program 147A to insert noise on line 17 both fields - see 147A Operation Manual for details.

d. Adjust 1480 waveform monitor to display line 17, field 1.

e. Adjust pedestal height (147A) until the inserted noise is at the same APL as the BAR in the composite test waveform.

f. Adjust the noise attenuators until the inserted noise is the same amplitude as the noise on the BAR.

g. Read and record the S/N ratio from the noise attenuators on the 147A Signal generator.

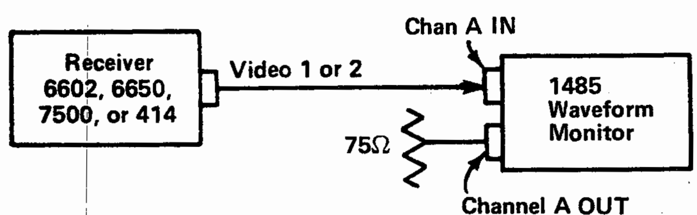


Waveform Monitor Display Showing Flatfield & Random Noise Test Signal.

Reduced Amplitude Video Response. This test verifies the frequency response of the up- and downlink video processing circuitry at six discrete frequencies from 0.5 MHz.

A subjective indication of phase response is shown by the degree of distortion or "rounding off" of the discrete frequency bursts. Reduced amplitude video response is measured as follows:

a. Interconnect the receiver and test equipment as shown in the block diagram.



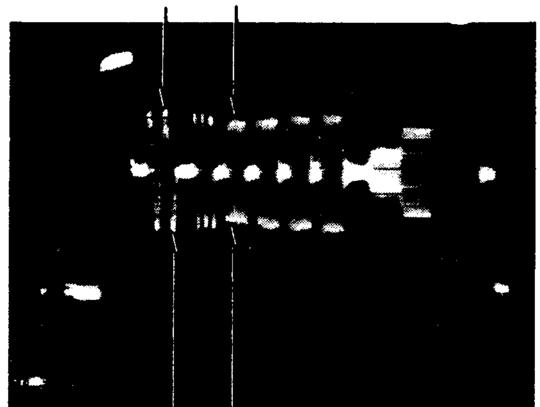
b. Adjust waveform monitor to display the multiburst test waveform (line 17, field 2 on HBO programming).

c. Adjust the "volts per full scale" control and "calibrate" control to a point such that the highest amplitude burst covers the area between 0 and 100 on the display graticule. This sets the amplitude of the higher amplitude burst at 20 "units".

d. Count the number of units covered by the lowest amplitude burst.

e. Reduced amplitude video response in
   dB = (20 log highest burst amplitude ÷ lowest burst amplitude)

   EXAMPLE:
   highest burst = 20 "units"
   lowest burst = 16.5 "units"
   reduced amplitude video response = $20 \log \frac{20}{16.5} = 1.67$ dB



20 Div. 16.5 Div.

"K" Factor K_{2t}. K factor is a measure of the short time response of the up- and downlink video processing circuitry. This measurement also gives subjective indications of phase and delay distortions as indicated by the distortion of the 2t pulse's height and width. "K" factor is measured as follows:

a. Interconnect receiver and test equipment as shown for reduced amplitude video response.

b. Obtain composite video test waveform on waveform monitor display. Center 2t pulse in the "K" factor window on the display graticule and adjust volts per full scale calibrate control so that the 2t pulse covers the area between the 0 or baseline point to the 100 unit.

c. Set mag. control to X25 and center the 2t pulse in the "K" factor window.

d. When volts per full scale = 1 volt, the window is 5%
   = 0.5 volt, the window is 2.5%
   = 0.2 volt, the window is 1%

e. Adjust volts per full scale until the ringing after and before the 2t pulse is just contained in the window limits, or the 2t pulse reaches the window shape factor limits. Interpolate and record the results as "K" factor (K_{2t}).

Luminance-Chrominance Delay. Luminance-chrominance delay is a measure of the system delay characteristics at chrominance frequencies with respect to luminance frequencies. L/C delay is measured as follows:

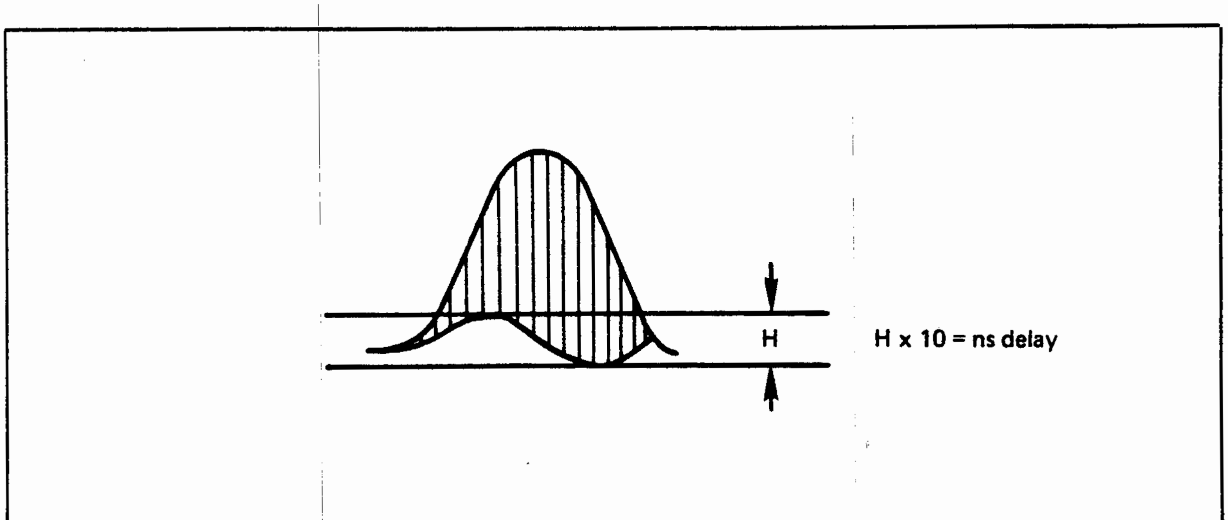
a. Interconnect equipment as shown for reduced amplitude video response.

b. Obtain the composite video test waveform on the waveform monitor display.

c. Adust mag. control for best resolution of 12.5T modulated $sine^2$ pulse.

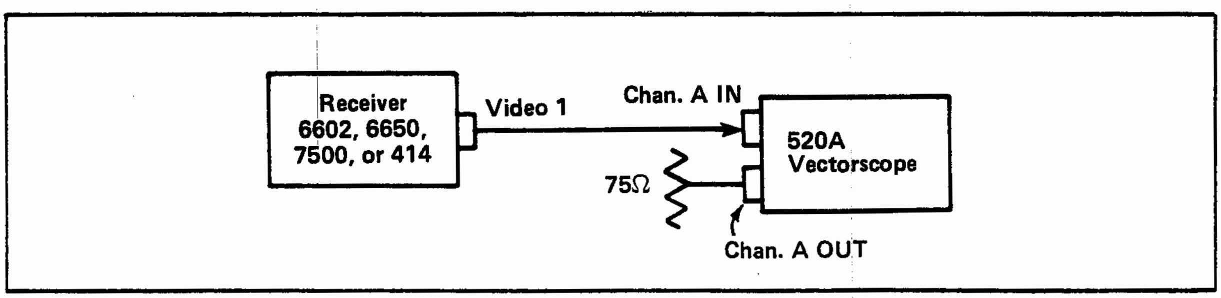
d. Measure the peak-to-peak amplitude of the sinusoidal base line distortion of the 12.5T pulse in IRE units.

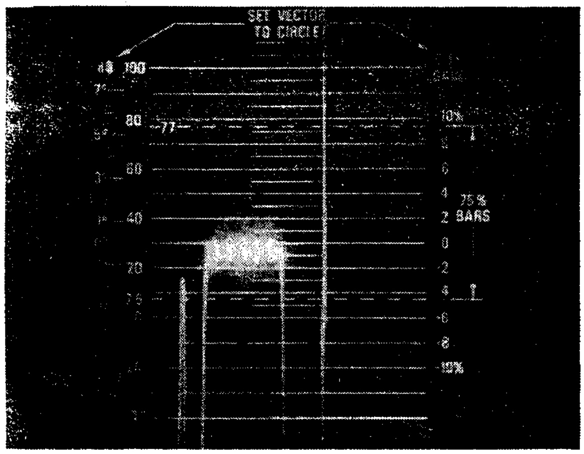
e. For 12.5T pulse delay nsec = 10 x peak-to-peak IRE from above.



Differential Gain. Differential gain is defined as the change in amplitude of the subcarrier portion of the modulated staircase as the luminance portion of the staircase is varied from blanking level to white level. Differential gain's most notable effect is observed as misregistered shade in the color television picture. Differential gain is measured as follows:

a. Interconnect the equipment as shown in the block diagram.





b. Set the vector scope line selector switch for the line number which contains the composite video test signal.

c. Depress the Channel A, vector, and VITS Field 1 or VITS Field 2 selectors as appropriate.

d. Adjust channel gain and phase controls until the vectors lie on the 180° line and coincide with the outer circle on the graticule.

e. Depress the differential gain selector and read differential gain in % from the vectorscope graticule.

Differential Gain Cal Factor

0.1 dB	1.2%
0.2 dB	2.3%
0.3 dB	3.4%
0.4 dB	4.5%
0.5 dB	5.6%
0.6 dB	6.7%
0.7 dB	7.8%
0.8 dB	8.8%
0.9 dB	9.8%
1.0 dB	10.9%
1.1 dB	11.9%
1.2 dB	12.9%
1.3 dB	13.9%
1.4 dB	14.9%
1.5 dB	15.8%

Differential Phase. Differential phase is defined as the change in phase of the subcarrier portion of the modulated staircase as the luminance portion of the modulated staircase is varied from blanking level to white level. Differential phase's most notable effect is observed as misregistered hue in the color television picture. Differential phase is measured as follows:

a. Interconnect equipment as shown in the block diagram for differential gain.

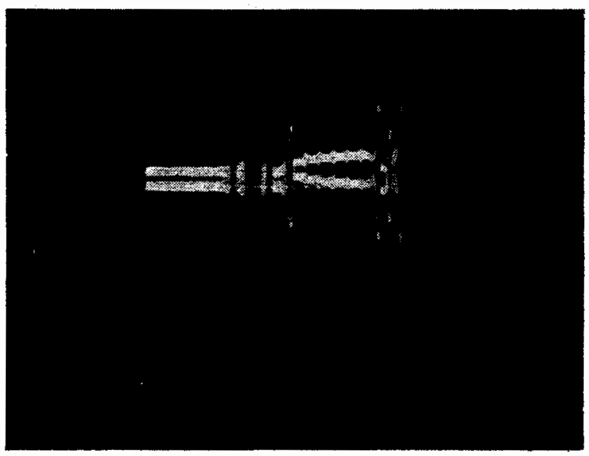
b. Set vector scope line selector switch for the line number which contains the composite video test signal.

c. Depress the Channel A, vector, and VITS Field 1 or Field 2 selectors as appropriate.

d. Set calibrated phase to 0°.

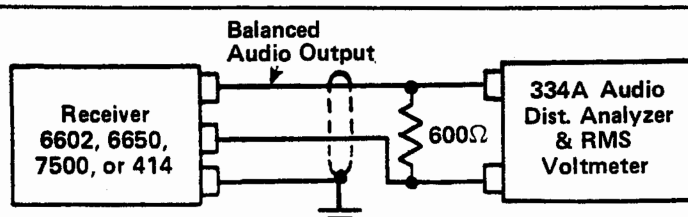
e. Adjust channel gain and phase until the vectors lie on the 180° line and coincide with the outer circle on the graticule.

f. Depress the differential phase button and adjust the channel phase control until either the left extremity or the right extremity of the double phase display is made to coincide. Now, using the calibrated phase shifter, bring together the extremities of the double phase display which were not made to coincide in the previous adjustment. Note the calibrated phase control reading and record as differential phase.



Audio Test Tone Signal/Noise. Audio test tone signal/noise is measured as follows:

a. Interconnect the equipment as shown in the block diagram.



b. Obtain 1-kHz audio test tone from source.

c. Set 334A for rms voltmeter mode of operation. Record signal + noise power from meter. Have the test tone removed and the input terminated at the source. Record the noise power indicated on the meter. The audio signal/noise ratio in dB is the signal plus noise power in dBW minus the noise power in dBW.

Audio Test Tone Distortion. Audio test tone distortion is measured as follows:

a. Interconnect the equipment as shown for audio signal/noise measurement.

b. Set 334A for distortion analyzer mode of operation.

c. Obtain 1-kHz audio test tone from source.

d. Adjust and optimize analyzer notch filter for the lowest % reading on the meter.

Maintenance

Periodically inspect the receiver mainframe for signs of damage from mechanical abuse. Also, look for evidence of overheating, especially in the power supply. An accumulation of dust on the power supply heat sink will have an insulating effect and prevent efficient heat dissipation. Keep both the inside and the outside of the mainframe clean. A low pressure air hose and a small paint brush can be used to clean connectors. The module edgeboard connectors can be cleaned with an art gum eraser. Spare modules should not be subjected to excessive heat, humidity or vibration. The antenna feed and coaxial cable pressurization unit should be checked periodically for proper operation. If a "dry air" system is used, the desiccant should be changed when two-thirds of the cylinder shows moisture contamination.

Fault Isolation

The information provided in this section will aid the operator in locating a faulty module so that he may replace it with a spare and resume operation. These module-level fault-isolation procedures are performed primarily with the receiver's front-panel meter and a volt-ohm meter. IF measurements, when necessary, can be made with a power meter such as the Hewlett/Packard Power Meter Model 435A.

If a spectrum analyzer is available, RF input measurements as well as more precise IF measurements can be made. The extender module which is included in the optional service kit is not required for module isolation.

The more detailed troubleshooting information in the module sections of the appropriate "Instruction Manual" will permit an experienced technician to further isolate a problem and make repairs. The equipment required for those procedures is listed in the "Instruction Manual."

CAUTION

Scientific-Atlanta recommends that, during the warranty period, any malfunctioning module be returned to the factory for repair. Damages sustained during the course of repair may void the warranty.

In the event of an apparent malfunction, the operator should make the following preliminary checks before tracing signals:

a. If the POWER switch does not illuminate, the input fuse may have blown or the ac power line may not be properly connected.

b. Using the front-panel meter, or voltmeter, check for the presence of the plus and minus dc supplies.

c. Make sure all modules are firmly inserted in the proper locations.

d. Make sure rear panel connections are complete.

e. Make sure the downconverter frequency determining elements (Switch X'tal, etc.) are set to the proper frequency.

f. Check the antenna for proper orientation.

g. Check the level and modulation settings of external equipment connected at the receiver's output. Incorrect settings can cause distortion of the video signal.

If there are other receivers at the site, problems which are external to the receiver may be identified by comparing the video output of the apparently faulty receiver to one which is operating properly. To do this, adjust the channel selector on the receiver which is operating properly to the same channel being viewed on the faulty unit. If this channel also appears to have problems on the good unit, then there is a high probability that the problem is not with either receiver but with the incoming signal on that particular channel.

If, however, the malfunction persists, it will be worthwhile to check the external equipment connected at the output of the suspected receiver against equipment performing satisfactorily with another receiver. With the channel selector still set on the frequency known to be operational, connect the good output signal processing equipment to the receiver's output connector(s). If the resulting output is still unacceptable, the receiver itself must be at fault.

Many apparent receiver failures, especially those involving a noisy video signal, can be traced to a poor quality uplink signal, improper antenna orientation, or a malfunction in station equipment external to the receiver. These items should be checked very carefully before tests involving the receiver are undertaken.

Video Uplink Earth Station Performance Verification

L. S. Hermann

Scope

The following sections describe seven tests which may be performed on a video uplink station to verify the performance of the equipment and ensure providing the high-quality signals required in video systems.

Output Power

1. Connect the test equipment as shown in Figure 1.

2. Check exciter frequency for mid-HPA channel.

3. Set Exciter for -10 dBm output.

4. Adjust HPA input attenuator for maximum output.

5. HPA output power may be calculated from output power meter reading by adding coupler value plus any attenuation between coupler and power meter.

Specification >34.4 dBW.

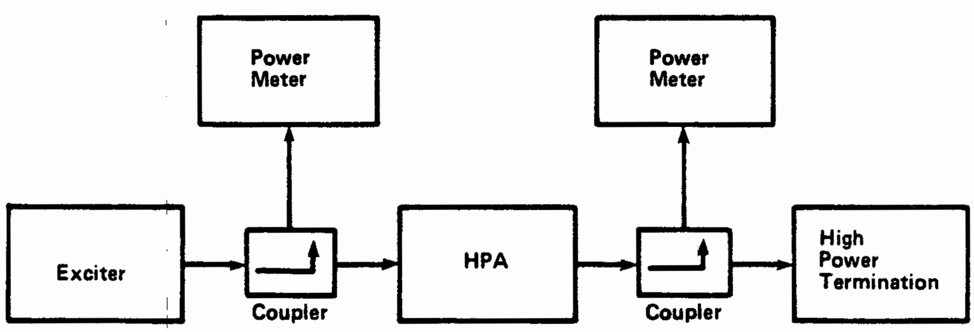


Figure 1. Output Power — HPA Gain

EIRP

EIRP may be calculated by measuring the output power at the OMT flange and adding the known antenna gain.

If the waveguide and switching system losses are known, then:

$$EIRP = G_{Antenna} + P_{HPA} - L_{WG} - L_{SW}$$

HPA Gain

1. Connect the test equipment as shown in Figure 1.

2. Check exciter frequency for mid-HPA channel.

3. Set HPA input attenuator for nominal output power.

4. HPA gain is HPA output power (dBW) - exciter output power (dBW)

Specification >70 dB.

IF-RF Gain and Gain Flatness

1. Connect the equipment as shown in Figure 2.

2. Adjust the sweep oscillator for 70 MHz ±20 MHz.

3. Check the exciter output for correct HPA channel output.

4. Calibrate scope output level and gain using attenuation and known input.

5. Measure IF-RF gain and gain flatness directly from scope display (Figure 3).

Specification: Total of 0.5 dB in center third of BW.
Total of 1 dB over remaining two-thirds of BW.

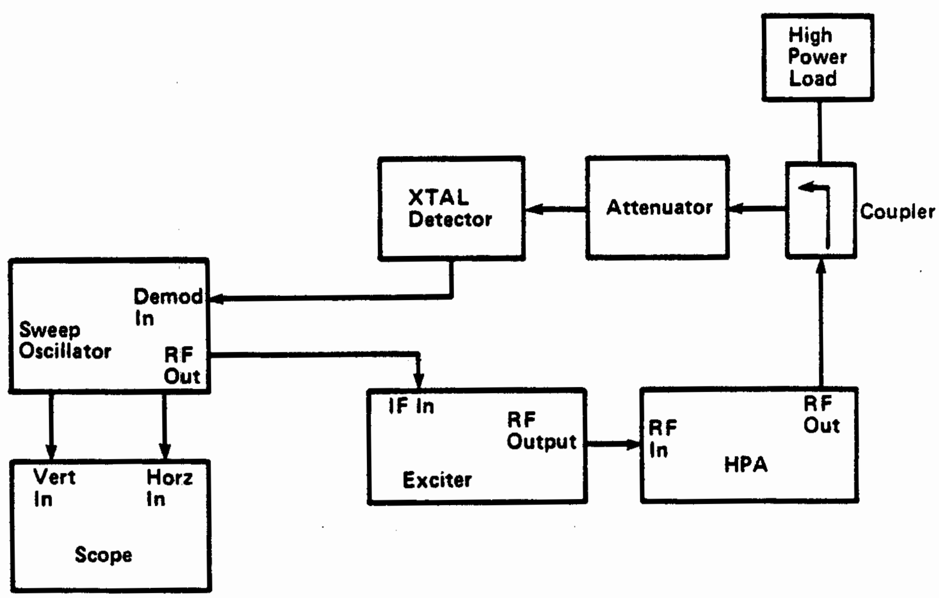


Figure 2. IF—RF Gain — Gain Flatness

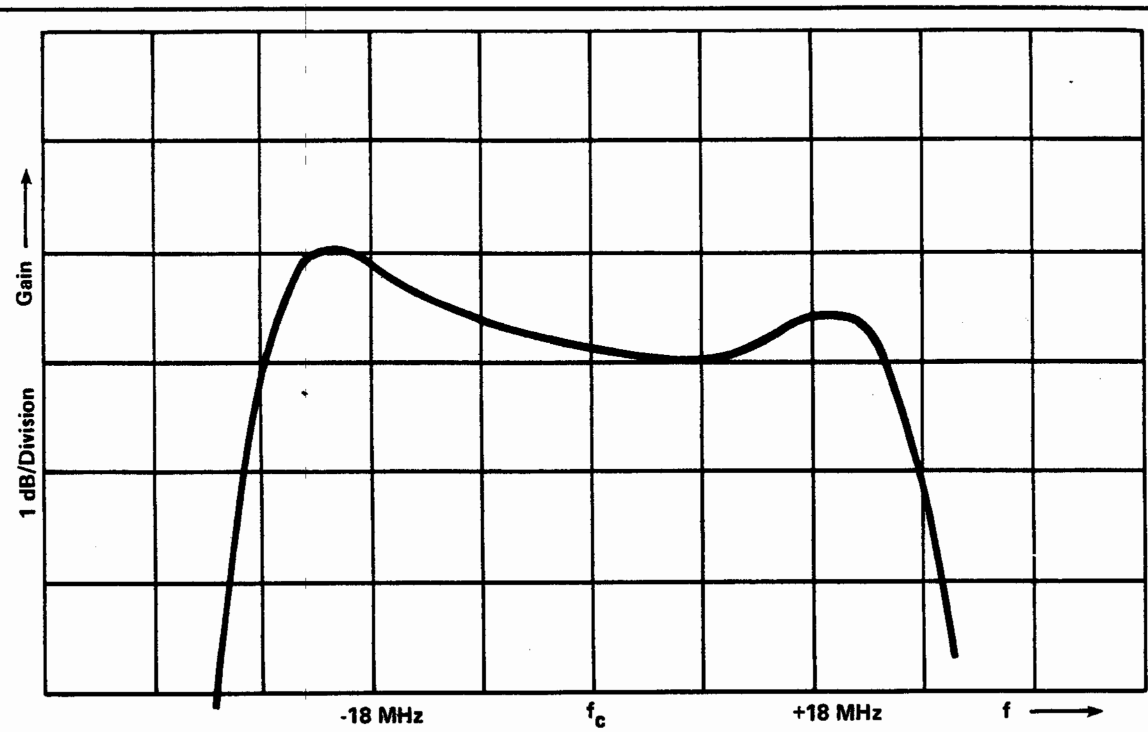


Figure 3. Typical HPA Channel

Group Delay

1. Connect test equipment as shown in Figure 4.
2. Check the exciter frequency for correct HPA channel.
3. Calibrate the MLA for 1 ns/cm group delay reading.
4. Measure the group delay across the channel (Figure 5).

Specification: Linear - 0.25 ns/MHz
Parabolic - 0.05 ns/MHz
Ripple - 2 ns/peak-to-peak

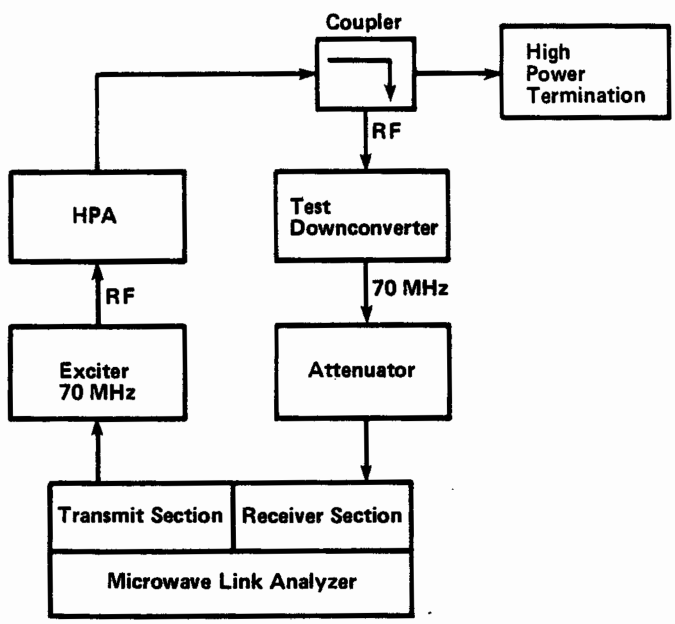


Figure 4. Group Delay

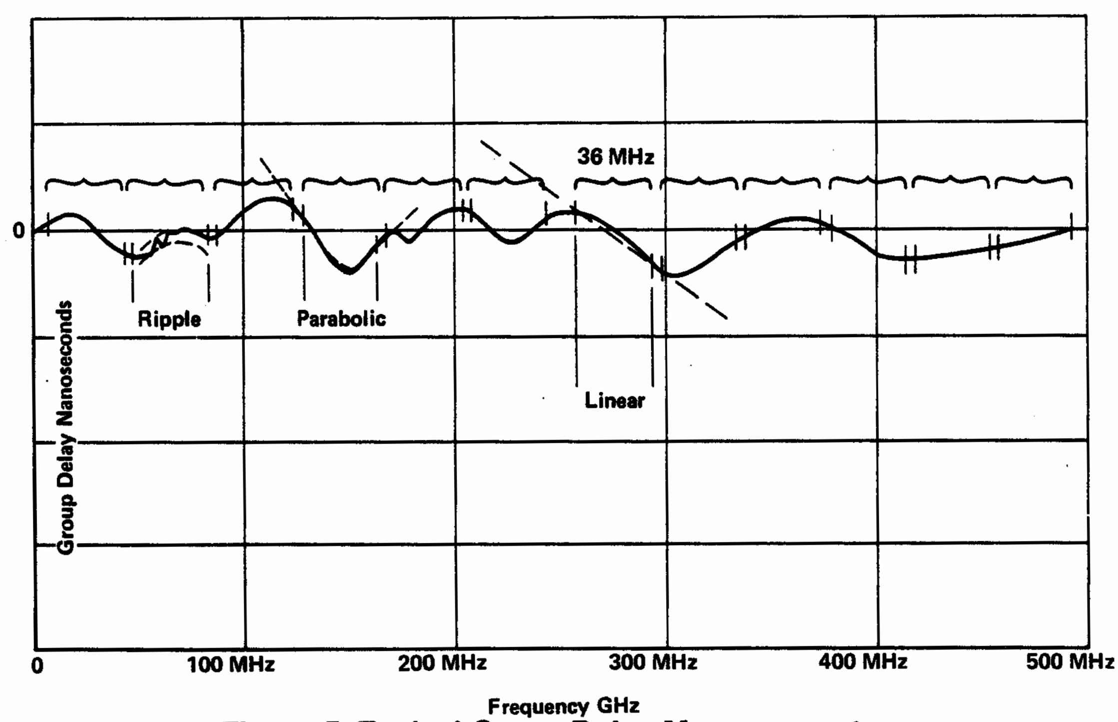


Figure 5. Typical Group Delay Measurement

Intermod

1. Connect the test equipment as shown in Figure 6.

2. The Exciter is driven by two in-band signals spaced such that the third-order intermodulation products are shown on the spectrum analyzer.

3. Measure the intermodulation performance per the graph shown in Figure 7.

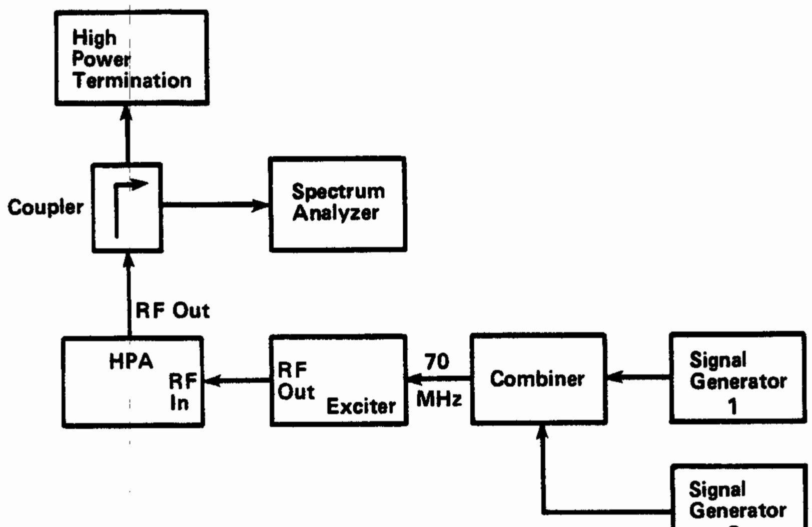


Figure 6. Intermod

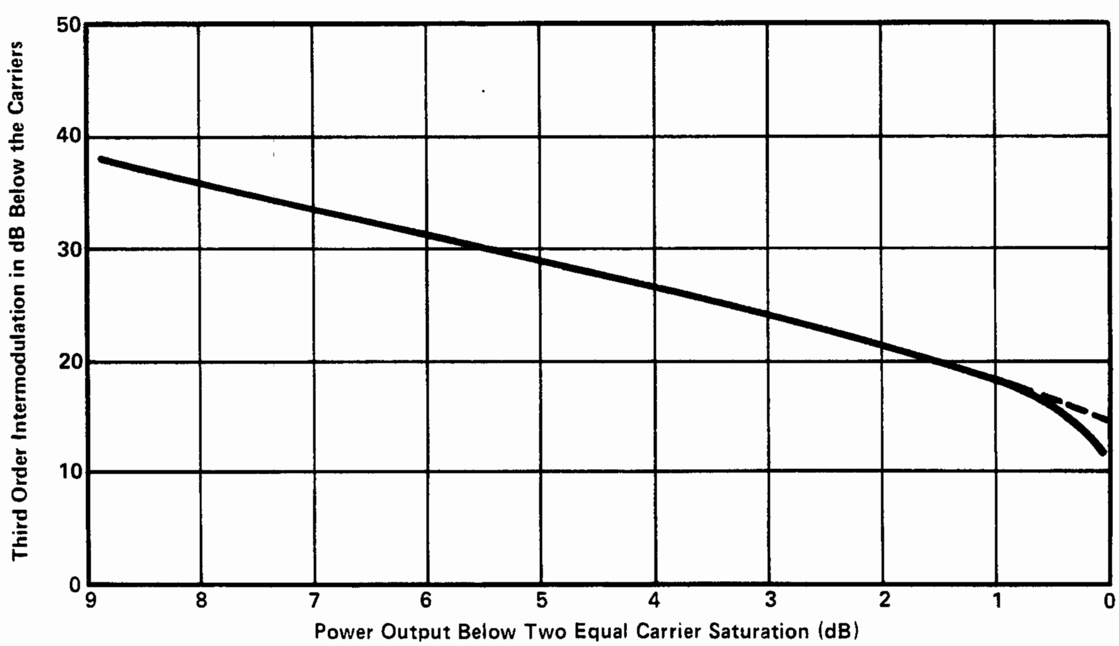


Figure 7. Third Order Intermodulation Distortion Under Two Equal Carrier Conditions

Video Tests

1. Connect test equipment as shown in Figure 8.

2. With all equipment set up for 1V peak-to-peak levels, measure baseband flatness using baseband signal generator and rms voltmeter or calibrated display.

3. Using video signal generator and waveform monitor with vector scope, measure all video parameters.

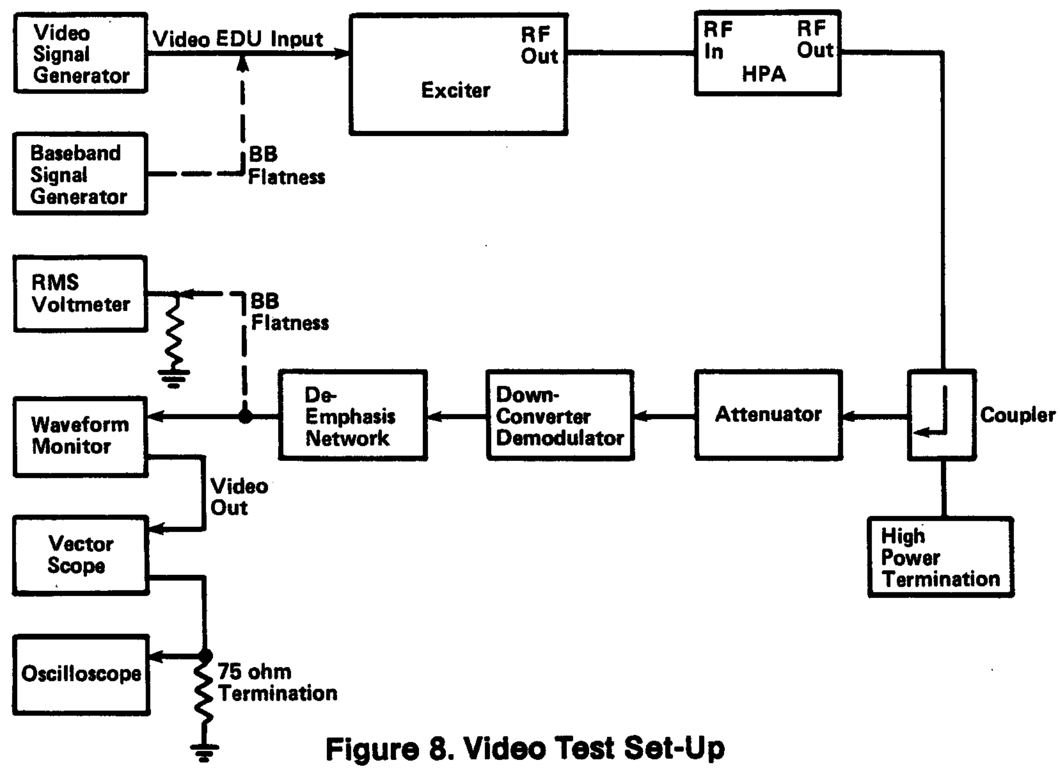


Figure 8. Video Test Set-Up

Typical measurements are:

- Differential Phase <±0.5°
- Differential Gain <±2%
- Field Time Distortion <1.0%
- Short Time Distortion <0.5%
- Chrominance to Luminance Delay <30 ns
- Non-Linear Distortion <3%

Operating a Transmit/Receive Earth Station

L. S. Hermann

Introduction

The major aspects of operating a satellite transmit/receive earth station will be addressed in this paper. Costs and revenues will be covered in a general sense, with particular emphasis on personnel as the major cost component. Transponder acquisition will then be discussed. Finally, equipment operation, troubleshooting, and maintenance of the station will be covered.

Transmit/receive operations fall into three categories: full-time operators, intermittent transmitters, and transportables. Aspects peculiar to each category will be addressed as necessary.

Costs

The costs of operating an earth station may be listed in the following order of relative magnitude:

- Personnel
- Depreciation of fixed assets
- Utilities
- Maintenance hardware
- Insurance
- Taxes

The first four of these will be discussed in detail.

In addition, the cost of the satellite transponder may be very large. This cost may be incurred as part of the transmission package (e.g., for a dedicated user), or may be billed separately to the end-user (e.g., with a resale operator). Transponder acquisition and costs are considered under a separate section of this paper.

Personnel

The personnel required to operate an earth station fall in these categories:

- Engineer
- Technician(s)

- Sales
- Accounting
- Legal
- Custodial

A full-time operator usually employs a full-time engineer and several technicians depending on the number of shifts and transmitters. An intermittent operator would use the part-time services of an engineer and full-time services of one or more technicians. A transportable operator requires a combination technician/engineer and usually a driver who may also be part technician.

These technical personnel should be first and foremost skilled in the media in which they operate, and secondly in RF microwave. The operators for a video station should be skilled in video technology; digital operators should be skilled in computer operation; SCPC/message operators should be skilled in telephony.

A technician familiar with the operating media may be trained in the RF operation of the station. He should soon be able to handle normal operation as well as minor troubleshooting and repair.

A person with engineering qualifications is needed to manage the technical aspects of an earth station. While being most familiar with the operating media, he would also have experience with microwave RF (e.g., terrestrial microwave). He whould then have little trouble learning satellite RF operation. This person should be capable of handling all major emergencies. As the engineer's time may only be partially used in operating the earth station, he may also manage other technical areas (e.g., a TV station or computer center), or manage the business side of the earth station.

A transportable earth station operator should be equally familiar with his media as with satellite RF operation. His customers will be specialists in the media they are broadcasting, and the operator must deal effectively with them. Yet, the operator must be self-sufficient and capable of handling any RF emergency. He should have extensive field training in satellite RF, preferably having been involved in a previous earth station installation and proof-of-performance.

The remaining personnel will be needed in various degrees depending on the nature of the service provided. An accounting staff is required to bill customers for usage. Legal help would be required for new FCC filings (e.g., for a change in tariff, or for an increase in power from an additional uplink or antenna).

Depreciation of Fixed Assets

Building space and support equipment (e.g. air conditioners) should be included along with the cost of the earth station in determining fixed assets. Otherwise, standard accounting procedures apply.

Utilities

Electricity will be the largest utility expense. The electrical load will depend on the configuration of the earth station as well as its use. Power requirements are listed in the previous paper, "Planning a Transmit/Receive Earth Station." Power required during various HPA usage cycles is discussed later in this paper under Equipment Operation.

Telephone charges may be substantial for an intermittent or transportable operator who must check with the satellite Technical Operations Center (TOC).

Water and other facility charges must be considered, especially in the case of a remote site.

Maintenance Hardware

Spare parts may be either stocked or purchased as needed. This measure of insurance is determined by the level of reliability required and the level of redundancy installed in the system.

A good rule-of-thumb is that maintenance hardware will cost three to five percent of the system cost per year. The power amplifier will require the greatest amount of maintenance, as it operates at high voltage and high temperatures. In the case of a Klystron power amplifier, the Klystron tube will need to be rebuilt every two to three years. A third tube is usually purchased when the first one fails to allow the operator to rotate the tubes for refurbishment.

Revenues

The most important decision in determining price levels is whether or not to apply for common carrier status. In order to charge specifically for the satellite uplink service, a common carrier license is needed. Tarriffed rates must then be filed with the FCC. Most intermittent or resale operators are licensed as common carriers.

If you are providing an end-to-end service of which the satellite uplink is only a part, you need not apply for common carrier status. Dedicated uplink operators would fall into this category.

Another determinant of the prices charged is the amount of backup protection offered the customer. As with satellite rates, uplink rates may be billed as:

- Fully protected (with a hot backup)
- Unprotected (single uplink)
- Unprotected and preemptible (if another uplink fails)

The type of service has an effect on HPA operating costs, as discussed later in this paper under Equipment Operation.

Transponder Acquisition

Full-time operators will either purchase or lease a full or partial transponder. Full transponders may sell for ten million dollars and up. Lease rates for a full transponder would depend on the kind of service and the length of the lease. Typical tariffs are listed in Table 1.

Table 1. Typical Full Transponder Lease Rates

Type of Service	Monthly Lease
Protected	\$150,000 to \$230,000
Unprotected	\$83,500 to \$149,500
Unprotected and Interruptible	\$66,000 to \$115,000

Source: Western Union Telegraph Co.; Tariff FCC No. 261; Satellite Transmission Service, page 33, effective June 10, 1982.

Full transponder service is also available in segments of as short as a half hour through numerous resellers. Charges for this transponder time vary according to time of day and volume. Typical charges run \$300 to \$700 per hour.

There are a number of companies which own tranponders and which resell time on various satellites. A representative sample of these resellers are shown in Table 2.

Table 2. Transponder Space Brokers

Argo Communications
100 Long Island Place, NW
Atlanta, GA 30328
Bill Papa

ASN, Inc.
310 14th Ave. S.
St. Petersburg, FL 33701

Bonneville Satellite Corp.
179 Social Hall Ave.
Salt Lake City, UT 84111
(801) 237-2450
Bruce Hall

Compact Video, Inc./
NETCOM Enterprises
Dept. of Satellite Services
2901 W. Alameda Ave.
Burbank, CA 91505
(213) 841-8887
Priscilla Davis

Equatorial Communications Service
300 Ferguson Dr.
Mountain View, CA 94043
Ed Parker

General Communications, Inc.
2550 Danali St.
Suite 505
Anchorage, AK 99503
(907) 338-6888
Bill Walp

Metropolitan Communications
Network Co.
P.O. Box 915
6861 Elm St.
McLean, VA 22101
(703) 734-2724
Dean Popps

Southern Satellite Systems
P.O. Box 45684
Tulsa, OK 74145
(918) 481-3275
Ruben Gant

United Video, Inc.
3801 S. Sheridan Rd.
Tulsa, OK 74145
(800) 331-4806
Tom McKeenzy

Western Telecommunications, Inc.
Box 22595
Wellshire Station
Denver, CO 80222
(303) 771-8200

Wold Communications, Inc.
10880 Wilshire Blvd.
Los Angeles, CA 90024
(213) 474-3500
Robert Wold

Partial transponder time is usually required for digital or voice transmission. This service is leased on the basis of number of carriers, bandwidth, and power required, as shown in Table 3.

These parameters are dependent on the system configuration and are determined by a system link analysis. A typical duplex voice FM-SCPC channel may cost $2,500 per month.

Table 3. Typical Partial Transponder Lease Rates

Basic	Monthly Lease (Sum)
Per Carrier (channel)	$390
Per Watt of Power	$131
Per kHz of Bandwidth	$ 3

Source: Western Union Telegraph Co.; Tariff FCC No. 261; Satellite Transmission Service; page 38; effective May 11, 1982.

Equipment Operation

Initiating a Satellite Transmission

Having secured transponder space, standard procedure for any operator when first transmitting is as follows:

a. Position the antenna on the satellite (if different than the previous transmission).

b. Using the receivers, check the null on adjacent transponders to assure the feed polarization angle is correct.

c. Tune exciter(s), HPA(s), and receiver(s) to the desired transponder frequency. On 6- or 12-channel klystron power amplifiers, this may require retuning the klystron tube.

d. Run power amplifier into a dummy load to check power levels.

e. Telephone the Satellite Technical Operations Center (TOC). This may be either the transponder owner or the satellite TT&C earth station. Ask for the Communications Technician (ComTech).

f. Transmit a CW carrier at a low power level. The TOC will check cross polarization and may ask you to adjust your feed polarization.

g. The TOC will give permission for full transmission.

h. Increase power to saturation.

i. Add deviation/modulation.

j. The TOC will check for adjacent transponder interference, then hang up.

k. Add the source.

Full-time operators who are continually going up and down on the same transponder need not repeat all these steps. Yet any operator who changes frequency and/or satellite should follow this procedure to avoid double-illumination of the wrong transponder.

When ceasing a transmission, the operator of an intermittent station should inform the TOC that he has done so for record and billing purposes.

It should also be noted that transponder costs will be incurred for both pre-broadcast testing and for other maintenance testing.

Time Required to Initiate a Transmission

Approximate times required in the various stages listed above are shown in Table 4.

Table 4. Time Required to Initiate a Transmission

Procedure	Approximate Time (Minutes)
Change Satellites	1 to 40 (depending on antenna motor speeds)
Retune HPA Frequency (klystron PAs only)	
Preset frequency	1 to 5
Requiring tuning	10 to 60 (depending on test equipment setup)
TOC Verification	5 to 20

Power Amplifier Operating Modes

Of the three kinds of power amplifiers currently available, the solid-state and TWT amplifiers with large bandwidths and long life spans are the most simple to operate. In addition, they are generally used in dedicated systems where they are seldom turned off.

Conversely, the Klystron High-Power Amplifier (HPA or KPA), is usually much larger, requires retuning to change transponders, and may experience intermittent operation. These conditions pose an interesting set of operating parameters.

Table 5 outlines the major characteristics of four HPA operating states. There are tradeoffs between the speed at which an HPA can be turned on, the amount of energy consumed (and heat given off), and the life expectancy of the tube.

Table 5. Major HPA Operating States

State	Life Expectancy of Klystron Tube	Power Required (kVA)	Heat Released	Time to Power Up (Minutes)
Full Filament/ High-Voltage	100%	12	100%	0
Cold	90%*	0	0%	7 to 10
Filament Foldback/ No High Voltage (newer models)	Intermittent use: 95% Continuous use: 2 hrs	1	3%	2 to 3
Full Filament/ No High Voltage (older models)	Intermittent use: 95% Continuous use: 30 min	1	3%	2 to 3

*Cold HPAs should be run with full load four times per year, at minimum, to prevent gas buildup inside the klystron tube. During long cold periods, the tube should also be stored externally on a Degassing Test Bed.

The normal operating mode of both the active and backup HPA is in the full-filament/high-voltage state. This provides the quickest transmission with the longest life. When an intermittent station is not in use, the cold state is normally used (provided no transmission is needed in less than ten minutes). However, if a station is used consistently more than a few hours each day, it becomes economic to leave all HPAs fully powered; this lengthens the tube life even though it uses more power.

The remaining two states of filament with no high voltage are used only when changing frequencies or performing internal tests. These states are used as infrequently as possible due to the severe impact on tube life.

Transportable Operation

Operation of a transportable earth station differs from that of a fixed station mostly in the area of setup. At least half a day should be allowed for the operator to position his vehicle, link in to his source, find the satellite, and fully test his system.

An added consideration with transportable earth stations is the frequency coordination required prior to each transmission. The frequency coordination companies provide a package for this service which allows multiple site clearings at a lower cost.

Troubleshooting

All earth station electronics include alarms in one or more forms, usually lights or LEDs. These should be monitored continuously during operation by the technician. In a system with automatic redundancy, a failure will cause the transmission to shift to the backup link.

The technician on duty should be capable of diagnosing the failure down to the board level, correcting the failure if possible, and bringing the system back on-line. If further troubleshooting is required, the engineer should be called in. With proper training, the engineer should be able to troubleshoot most failures down to the component level.

Preventive Maintenance

The amount of time and money invested in a transmit/receive earth station justifies the establishment of a preventive maintenance schedule. Weekly or monthly preventive maintenance inspections, depending on earth station size, should alert the technical or operations staff to any impending component degradation or failures.

Routine inspections will point out the blower about to experience bearing failure, air filters that are clogged, antenna motors that are noisy, and meter readings approaching out-of-tolerance conditions. Once discovered, impending problems can be corrected, and the chance for failure is reduced.

All operational manuals provided by Scientific-Atlanta specify the level of maintenance on each item. Motorized antennas require periodic lubrication and tightening of bolts; all should be checked to ensure no binding areas nor rust spots are obvious or that the feed window is not cracked and leaking. The transmit waveguide should be checked periodically to ensure that no air leaks have occured and no sections have been kinked or crimped. Positive pressure using dry air should be maintained in both the transmit and receive waveguide at all times to reduce line losses and prevent possible arcing in the transmit waveguide.

The HPAs are the most sensitive part of the system due to the high-power levels at which they operate.

The HPAs should be checked daily if they are operated continuously. The air circulation in the HPA cabinet is critical for operation. Loss of a blower will alarm the system, and the HPA will shut down until the fault is corrected. The design of most HPAs is such that a potentially self-destructive fault results in shutdown.

Maintenance cannot be overemphasized in owning and operating a transmit/-receive earth station.

Glossary

A/D or ADC — Analog to digital. Analog-to-digital converter.

AFC — Automatic frequency control.

AGC — Automatic gain control.

AISC — American Institute of Steel Construction.

ALC — Automatic level control.

ANSI — American National Standards Institute.

Aperture — (directive antenna) 1. The clear diameter of the parabolic reflector of a microwave antenna. 2. A surface, near or on a directive antenna, through which most of the radiated energy passes and to make assumptions regarding the field values for the purpose of computing fields at external points.

Aperture Efficiency — The ratio of the directivity of an antenna to that which would be obtained if the aperture illumination were uniform. (See Aperture, Directivity.)

APL — 1. (television) Average picture level. Average value of the picture signal, integrated over one frame. The picture signal does not include blanking signals. 2. (audio) Average program level.

ASCII — American Standard Code for Information Interchange.

ASMS — Automatic surface measuring system.

ASTM — American Society for Testing and Materials.

Asynchronous Communications — Transmission and reception of data in a mode where there is no continuous synchronization between the terminals.

AZ — Azimuth angle (of an object). The arc of the horizon, between true north and the vertical plane passing through the object, usually measured toward east.

Baud — A unit of signal speed equal to the number of signal elements or symbols transferred in each second; a unit of transmission rate of digital signals. It is the reciprocal of the length in seconds of the shortest element of the digital code.

BER — Bit error rate. The probability that a bit will be received in error, numerically equal to the number of bit errors divided by the number of bits transmitted.

Bit — Binary digit.

Bits per Second — The number of bits of data transferred in each second.

Blanking — The process of cutting off the electron beam in a camera or picture tube during the retrace period.

Blanking Level — 1. That level of a composite picture signal which separates the range containing picture information from the range containing synchronizing information. 2. The level of the front and back porches of the composite video signal.

BPSK — Bi-phase shift keying. Carrier modulation technique in which the phase of the transmitted carrier is shifted ±180 degrees with respect to a reference.

Byte — A group of adjacent binary digits operated on as a unit and usually shorter than a computer word. By currently accepted definitions, a byte is 8 bits.

CATV — Cable television or community antenna television.

CCIR — International Ratio Consultative Committee.

Chrominance — That property of light which produces a sensation of color in the human eye apart from any variation in luminance that may be present.

C/I — Carrier-to-interference ratio. The power in the desired carrier divided by the power in the interfering signal or signals.

C/N — Carrier-to-noise ratio. Ratio of carrier power to noise power in a defined frequency band.

C/N_0 — Carrier-to-noise-power-density ratio. (See C/N, Noise Power Density.)

Color-Bar Signal — A test signal, typically containing six basic colors: yellow, cyan, green, magenta, red, and blue, which is used to check certain functions of color TV systems.

Color Burst — In NTSC color systems, this normally refers to a burst of approximately 10 cycles of 3.579545-MHz subcarrier frequency on the back porch of the composite video signal. It serves as a color synchronizing signal to establish a frequency and phase reference for a chrominance signal.

Color Phase — The phase, with respect to the chrominance-carrier reference, of the component of the carrier-chrominance signal which corresponds to one of the chrominance primaries.

Color Signal — The chrominance and luminance components of the NTSC color television signal.

Color Subcarrier — In color systems, this is the carrier signal whose modulation sidebands are added to the monochrome signals to convey color information; in NTSC, it is a 3.579545-MHz sinewave.

Composite Baseband — For color video satellite links, this consists of the Composite Video, Energy Dispersal, and Audio Subcarrier signals.

Composite Video — For color, this consists of blanking, field, and line synchronizing signals, color burst, chrominance, and luminance picture information. These are all combined to form the complete color video signal.

Contrast — The ratio between the maximum and minimum brightness values in a picture.

CONUS — Contiguous (48) United States.

Crosstalk — Undesired energy appearing in one signal path as a result of coupling from other signal paths.

CVSD — Continuous variable slope delta modulation.

D/A or DAC — Digital to analog. Digital-to-analog converter.

Data — 1. Any representations, such as characters or analog quantities, to which meaning might be assigned. 2. Information in a form which can be transmitted by a communications link or processed by computers or machines; e.g., analog data or digital data.

DEC — 1. Declination angle. Angle between the celestial equator and an extra-terrestrial object, measured in a plane which is normal to the celestial equator. 2. Digital Equipment Corporation.

Decibel (dB) — A number denoting the ratio of the two amounts of power, being ten times the logarithm to the base 10 of this ratio. The abbreviation dB is commonly used for the term decibel. With P_1 and P_2 designating two amounts of power and n the number of decibels denoting their ratio,

$$n = 10 \log_{10}(P_1/P_2) \text{ decibels.}$$

When the conditions are such that ratios of currents or ratios of voltages (or analogous quantities) are the square roots of the corresponding power ratios, the number of decibels by which the corresponding powers differ is expressed by the following equations:

$$n = 20 \log_{10}(I_1/I_2) \text{ decibel}$$

$$n = 20 \log_{10}(V_1/V_2) \text{ decibel}$$

where I_1/I_2 and V_1/V_2 are the given current and voltage ratios, respectively. See Appendix A of paper 1-2.

Deemphasis — The restoration of a preemphasized frequency spectrum to its original form. The introduction of loss at the higher frequencies in the receiver of a communication system to compensate for earlier preemphasis in the transmitter. (See Preemphasis.)

Delay Distortion — That form of distortion which occurs when the envelope delay of a circuit or system is not constant over the frequency range required for transmission.

Differential Gain — A form of non-linear distortion. As applied to NTSC or PAL color television, it is the change in the amplitude of a small sinusoid at the color subcarrier frequency caused by variation in the accompanying luminance level. It is expressed as the difference in amplitude between the maximum amplitude and the minimum amplitude of the sinusoid when the maximum amplitude is normalized to 100 percent. The luminance signal shall explore the range from black to white.

Differential Phase — A form of non-linear distortion. As applied to NTSC or PAL color television, it is the change in the phase of a small sinusoid at the color subcarrier frequency caused by variation in the accompanying luminance level. It is expressed as the greatest difference in phase observed for luminance levels between black and white.

Digital Data — Information coded in digital electronic form from analysis or computation. Any data which is expressed in digits. Binary digits are usually implied.

Directional Coupler — 1. A transmission coupling device for separately sampling the forward (incident) and/or the backward (reflected) wave in a transmission line. 2. A device which can be inserted in a 75 ohm CATV cable to tap off the signal for local use.

Directivity — (antenna) The directivity of an antenna in a specified direction is 4π times the ratio of the power radiated per unit solid angle in that direction to the total power radiated by the antenna. This term differs from power gain in that it does not include antenna dissipation losses.

DMUX — Demultiplexer. Demultiplex. Equipment for or process of separating two or more messages which are interleaved in frequency or time for transmission on a single channel. (See FDM, TDM.)

Downlink — The circuit between a satellite and a receiving earth station, including the satellite transmitter and antenna, the satellite-earth propagation path, and the earth station antenna and receiver.

E_b/N_o — Energy-per-bit-to-noise-power-density ratio. Signal-to-noise ratio normalized with respect to bit rate.

$$E_b/N_o = \frac{P_s/R}{N_o}$$

where:

P_s = signal power

R = bit rate

No = noise power density

Echo — 1. A signal which has been reflected at one or more points during transmission. 2. A wave that has been reflected or otherwise returned with sufficient magnitude and delay to be perceived in some manner as a wave distinct from that directly transmitted. 3. An attenuated reflection of a talker's voice, separated from the primary wave by reflection at an electrical discontinuity in the circuit, and returned to him over the circuit on which he is talking and listening.

Effective Area — 1. In a given direction, the ratio of the power P_r available at the terminals of a receiving antenna to the power density S per unit area of a plane wave incident on the antenna from that direction, the incident wave being polarization matched to the receiving antenna:

$$A_E = \frac{P_r}{S},$$

where A_E is the effective area. (See Partial Effective Area, Realized Effective Area.)

EIA — Electronic Industries Association.

EIRP — Effective isotropic radiated power. Product of transmitted power times transmitting-antenna gain, usually expressed in dBW.

EL — Elevation angle. (of an object) The vertical angle from the local horizon to the object.

EMI — Electromagnetic interference.

End-to-End — Reference to the totality of the circuitry, devices and transmission media from input terminals to the output terminals of a communications link.

Energy Dispersal Waveform — For video satellite links, a triangular waveform with apexes located at the vertical sync intervals which is modulated onto the RF carrier to avoid concentration of energy at one frequency.

Energy Dispersal Scrambling — A method of making the output power spectrum of a PSK modulator independent of the data stream by scrambling the data at the transmit end and descrambling it at the receiver. Energy dispersal scrambling is required by the FCC to avoid interference with other signals.

EPROM — Erasable programmable read-only memory.

Equatorial (Polar) Coordinate System — A spherical coordinate system whose polar axis is aligned with the earth's poles.

ES — Earth station.

ET — Earth terminal. Extra-terrestrial.

FDM — Frequency-Division Multiplex — A method of multiplexing or combining more than one (usually many) voice or data channels for transmission on a single RF carrier. The channels are separated in frequency and are carried on subcarriers.

FDMA — Frequency-division multiple access (of a satellite transponder by different earth stations).

FEC — Forward error correction. An encoding technique for improving the BER in a data system. In a typical rate 3/4 FEC system, the BER may be improved by a factor of ten thousand.

Feed — That portion of an antenna which is coupled to the terminals and which functions to produce the aperture illumination on transmission or to couple energy concentrated at the focal point to the antenna terminals on receiving.

Field — One half of a complete picture (or frame) interval, containing all of the odd, or all of the even, lines of the picture. One of the two (or more) equal parts into which a frame is divided in interlaced scanning.

Field Frequency — The rate at which one complete field is scanned, 59.54 times a second in the NTSC color system.

Figure of Merit — (receiving system). (See G/T.)

Flux Density — Power density. Power per unit area normal to the direction of propagation of a propagating electromagnetic field.

Frame — One complete television picture, consisting of two fields of interlaced scanning lines. (See Field.)

Frequency Modulation (FM) — Angle modulation in which the instantaneous frequency of a sine-wave carrier is caused to depart from the carrier frequency by an amount proportional to the instantaneous value of the modulating wave.

Frequency Response — In a linear system, the frequency-dependent relation, in both gain and phase difference, between steady-state sinusoidal inputs and the resultant steady-state sinusoidal outputs. A plot of gain in decibels and phase angle in degrees versus logarithmic frequency is commonly called a Bode diagram.

Frequency Reuse — 1. A technique in which independent information is transmitted on orthogonal polarizations to "reuse" a given band of frequencies. 2. A technique in which independent information transmitted to different areas by the use of "spot" beams to "reuse" a given band of frequencies. Note: The techniques of 1 and 2 are sometimes clarified by addition of the phrases "by polarization diversity" and "by space diversity."

FSK — Frequency shift keying.

Full Duplex — The simultaneous 2-way flow of data between data terminals. A method of operation which provides simultaneous 2-way communications between two points.

Gain — (antenna) 1. The power gain of an antenna, in a specified direction, is 4π times the ratio of the power radiated per unit solid angle in that direction to the net power accepted by the antenna from its generator. The terms "gain" and "power gain" are synonymous. See: directivity; power gain; partial effective area (gain); realized gain. 2. The ratio of the signal level received (or transmitted) by an antenna to the signal level which would be received (or transmitted) by an isotropic antenna at the same location and under the same conditions, and with each antenna conjugately matched to the receiver (transmitter).

Gain — (amplifier) (See Power Gain.)

Gain Stability — Variation of the gain of a device, cascade of devices or system as a function of the family of possible causes of gain perturbation, such as temperature, humidity, line voltage and ageing of components. Ageing results in long-term changes in gain caused by gradual changes of components. Gain stability can be categorized into short-period (e.g., 1 second), medium-period (e.g., 1 hour), and longer variations. Gain stability is often specified as the maximum variation in decibels which a device experiences in a 24-hour test period. To avoid confusion it should be specifically stated whether the variation is the peak-to-peak variation, the peak plus-minus variation from a mean value, the rms variation from the mean, etc.

GCE — Ground communications equipment. This term relates to the earth station electronic equipment, such as receivers and exciters.

G/T — Figure of merit of a receiving system, expressed in dB/K. G/T is defined as the ratio of the receiving system gain, including the antenna gain, at a specified reference point in the receiving system (preceding the demodulator) to the receiving system noise temperature in Kelvins referred to the same point. For a given system and a given antenna elevation look angle, the value of G/T is independent of the point in the receiving system at which it is measured. In relation to its sensitivity of detecting the signals from a satellite, the higher the figure of merit the higher the sensitivity of the receiving system in detecting satellite signals.

HA — Hour angle. Angles about the polar axis of an equatorial coordinate system, traditionally measured in hours, minutes and seconds. The hour angle of an object is the polar angle between the meridian of the observer and the hour circle passing through the object. (See Hour Circles.)

Half Duplex — The non-simultaneous 2-way flow of data between communications terminals. Description of a system in which communication can be in either direction, but in only one way at a time.

Hour Circles — Great circles of the celestial sphere passing through the poles.

HPA — High-power amplifier. In a transmitting earth station, this is the final RF amplifier between the modulator/exciter and the antenna.

Idle Channel Noise — The total rms noise measured at the output of a channel with the input to the channel terminated with system reference impedance.

Insertion Gain — On insertion of a transducer in a transmission system, the insertion gain is the ratio of (1) the power delivered to the port of the system following the transducer to (2) the power delivered to the same port before insertion. The insertion gain may be equal to, more than or less than unity. It can be expressed as a power ratio or in decibels.

K — Kelvin. Temperature of a device in Kelvins. Zero K equals -273.15°C. The Kelvin scale is the same as the Celsius scale except for the offset of 273.15°C.

LNA — Low-noise amplifier. This is the preamplifier between the antenna and the earth station receiver. For maximum effectiveness, it must be located as near the antenna as possible, and is usually attached directly to the antenna receive port.

LNC — Low noise converter. Integrated LNA and downconverter.

Luminance — An attribute of light or color. Luminous intensity or brightness as measured by a photometer instead of by the human eye.

MATV — Master antenna television.

Modem — A contraction of modulator/demodulator. Usually a device that combines the modulation and demodulation functions in a single unit.

MTBF — Mean time between failure. A statistical determination of the time in hours-of-use between failures.

Multipoint (or Multi-drop) — A method of connecting several voice or data terminals to a single channel in a communications system.

MUX — Multiplexer or multiplex. Equivalent for or process of interleaving in frequency or time two or more messages for transmission on a single channel.

NF — Noise Figure. A figure of merit of a device, such as an LNA or receiver, which compares the device with a perfect device. NF is the ratio of (a) the noise power output of a matched transducer to (b) the portion of that noise which is attributable to thermal noise if the input port of the transducer is at the standard noise temperature of 290 Kelvins.
The noise figure of a device, expressed in dB, is related to its noise temperature T_E by

$$NF(dB) = 10 \log (1 + T_E/290)$$

where T_E is measured in Kelvins.

Noise Power (Spectral) Density — Noise power per unit of bandwidth.

Noise Temperature — More correctly called "effective noise temperature", this is a measure of the noise power referred to a given reference point of a matched receiving system, normalized with respect to bandwidth.

$$T_E = \frac{N}{kB} \equiv \frac{N_o}{k}$$

where
N is the noise power at the reference point in the noise bandwidth B, k is Boltzman's constant, and N_o is the noise power density.

NTSC — National Television Systems Committee. An industry-wide engineering group which, during 1950-1953, developed the color television specifications now established in the United States.

OMT — Orthomode transducer. A waveguide device attached to an antenna feed that permits using the antenna for simultaneous transmission and reception or orthogonal polarizations.

Orthogonal Polarization(s) — (See Polarizations, Orthogonal.)

Parity Check — A simple test for detecting errors in the transmission of ASCII characters. 1. (electronic computation) A summation check in which the bits in a character or block are added (modulo 2) and the sum checked against a single, previously computed parity digit; that is, a check that tests whether the number of ones is odd or even. 2. A method of checking the accuracy of transmission of digital data by adding a "parity bit" so that the total number of "ones" in each character is always odd.

Partial Effective Area (Gain) — That part of the effective area (gain) of an antenna which is in a specified polarization. The effective area (gain) of an antenna is the sum of two partial effective areas (gains) which are defined with respect to two orthogonal but otherwise arbitrary polarizations: e.g.: the effective area (gain) of any linearly polarized antenna is the sum of two equal partial effective areas (gains) which are respectively right-hand and left-hand circularly polarized. (See Polarizations, Orthogonal.)

Peak Unaffected Signal — (audio) The maximum sinewave input level which will be passed through a channel without clipping or limiting.

Polar Coordinate System — (See Equatorial Coordinate System.)

Polarization — 1. (wave) The polarization in a given direction of a single-frequency propagating electromagnetic wave is described by the amplitude and direction of its electric field (normal to the direction of propagation) is called the plane of polarization. The tip of the electric vector representing the field describes an elliptical locus called the polarization ellipse. Circular polarization and linear polarization are degenerate cases of elliptical polarization. The polarization is defined by the shape of the polarization ellipse (axial ratio), its orientation (tilt angle) and the direction in which the locus is traversed (sense). The sense of polarization is defined to be right-handed or left-handed. It is right handed if the electric vector rotates in a clockwise direction as viewed from behind the outgoing wave.
2. (antenna) The polarization of an antenna is defined by the polarization of the wave it radiates.

Polarization Efficiency — (of an antenna with respect to an incident wave) The polarization efficiency is the ratio of the incident power of a wave which is received by an antenna to the power which would be received if the wave and the antenna were polarization matched. A reciprocal antenna is polarization matched to an incident wave if the axial ratio, tilt angle and sense of polarization of the incident wave are coincident with those of the wave which would be radiated by the antenna.

Polarizations, Orthogonal — Any single-frequency propagating wave can be separated into orthogonal polarization components. The polarization components are orthogonal if at a point the total power density in the wave is equal to the sum of the power densities in the two components. Two polarizations are orthogonal if their tilt angles are mutually perpendicular, the magnitude of their axial ratios are equal and their senses are opposite. (See Polarization.)

Power Density — 1. (free-space propagating wave) Power per square meter in a plane normal to the direction of propagation (watts/meter^2).
2. (circuit) Power per unit bandwidth (usually in watts/Hz or watts/MHz).

Power Gain — 1. (antenna) See Gain. 2. (amplifier or transducer) The power gain (usually called the gain) is the ratio of the output power to the input power. The available gain is the gain realized under conjugately matched conditions. The gain can be expressed as a numerical ratio or in decibels. The numerical gain of a transducer can be less than, equal to or greater than unity. If it is less than unity, the numerical gain (fraction) is sometimes inverted and called a loss. The gain is called the saturated gain if the output power is constant with increasing input power.

Preemphasis — The intentional alteration of the spectrum of a signal by emphasizing one range of frequencies with respect to another, usually to gain an advantage with respect to noise or non-linear distortion. (See Deemphasis.)

Protocol — (data communications) A formal set of conventions which governs the flow and timing of data between communication devices.

PSK — Phase shift keying.

PWB — Printed wiring board.

QPSK — Quadriphase shift keying. A carrier modulation technique in which the phase of the carrier is shifted to any of 4 phases, whose separations are multiples of 90 degrees.

RAM — Random access memory.

RCVR — Receiver.

Realized Gain (Effective Area) — (antenna or transducer) The realized gain is the gain reduced by the losses due to mismatch.

Reflector — (aperture antennas) A conducting surface (or one of a set of such surfaces) which is (are) shaped to collimate a transmitted wave into a beam or focus an incident wave onto a feed for coupling to a transmission line.

ROM — Read only memory.

RMS — Root mean square.

RS232C — A standard created by the Electronic Industries Association (EIA) for the transmission of data by wire over short distances (less than 50 feet).

SAbus — A serial data bus developed by Scientific-Atlanta for monitor and control of earth station equipment.

SCPC — Single channel per carrier. A satellite transmission system that employs a separate carrier for each channel. This method is usually used at earth stations that have a low volume of traffic compared to that which can be carried by a full transponder.

Signal-to-Quantization Noise Ratio — (digital audio) The ratio (dB) of the amplitude of a single frequency output signal to the rms quantizing noise appearing at the terminated channel output.

Simplex — Data or voice transmission in one direction only at all times.

Sinewave Overload — (digital audio) The sinewave input signal level which causes the converter saturation codes (all 1's or all 0's) to occur on the peaks of the signal.

SMATV — Satellite master antenna television. A master antenna distribution system that distributes satellite delivered signals as well as off-air signals and locally originated programming. SMATV systems are mainly found in multi-unit dwellings and lodging establishments.

SMPTE — Society of Motion Picture and Television Engineers.

S/N — Signal-to-noise ratio. The ratio of signal power and noise power. A video S/N (Signal-to-Noise Ratio) of 54 to 56 dB is considered to be an excellent S/N, that is, of best broadcast quality. A video S/N of 48 to 52 dB is considered to be a good S/N at the headend for Cable TV.

Sync. — An abbreviation for the words "synchronization," "synchronizing," etc. Applies to the synchronization signals, or timing pulses, which, for example, lock the electron beam of the picture monitors in step, both horizontally and vertically, with the electron beam of the pickup tube. The color sync signal (NTSC) is known as the color burst.

Synchronous Communication — Transmission and reception in a mode where there is continuous synchronization between the terminals.

TDM — Time division multiplex. The process or device in which each modulating wave modulates a separate pulse subcarrier, the pulse subcarriers being spaced in time so that no two pulses occupy the same time interval. Time division permits the transmission of two or more signals over a common path by using different time intervals for the transmission of the intelligence of each message signal. (See also FDM.)

TDMA — Time-division multiple access (of a satellite transponder by different earth stations).

TED — Threshold extension demodulator. A circuit designed to lower the threshold of FM (frequency-modulation) demodulators, permitting operation of the system with a lower C/N without impulse noise showing on the picture.

Termination — A one-port load that terminates a transmission line in a specified manner, usually to prevent reflection.

THD — Total harmonic distortion. The ratio of the rms energy contained in all harmonic components of a single-frequency input signal to the fundamental component amplitude.

TVRO — Television receive-only (earth station).

TX — Transmit or transmitter.

Uplink — The circuit between a transmitting earth station and a satellite, including the earth transmitter and antenna, the earth-satellite propagation path, and the satellite antenna and receiver.

Vestigial Sideband — 1. Reference to a type of amplitude-modulated ratio or television signal in which most of one sideband is eliminated, leaving the other sideband intact. 2. The transmitted portion of the sideband which has been largely suppressed by a transducer having a gradual cutoff in the neighborhood of the carrier frequency.

Video — A term pertaining to the bandwidth and spectrum position of the signal resulting from television scanning.

Video Plus — A technique whereby SCPC carriers (analog or digital) share a transponder with a full-transponder video signal without mutual interference.

VSWR — Voltage standing-wave ratio (in a waveguide or transmission line). The ratio of the magnitude of the transverse electric field in a plane of maximum strength to the magnitude at the equivalent point in an adjacent plane of minimum field strength.

zenith — The zenith of an observer is the point of the celestial sphere which is vertically overhead. The nadir is the point vertically beneath.

Evaluation of Intrinsic Noise in FDM/FM Systems

B.W. Brinegar

The evaluation of FDM/FM message systems can be greatly enhanced by comparing theoretical calculations versus empirical data. This paper presents equations required to calculate the conventional signal-to-noise curves for FDM/FM systems using a limiter/discriminator as a demodulator. Also, equations used for occupied bandwidth and linearity bandwidth are presented.

Examples of several message configurations utilizing these calculations are included. Signal-to-noise curves are calculated by the equation:

$$S/N_{pwp} = (S/N\ Normal_{dB} - S/N\ Threshold_{dB})_{pwp} + S/N_{1/Fpwp}$$

This equation has three distict areas: the S/N 1/F region, the S/N normal region and the S/N threshold region. Observing the equations on the following pages, it becomes obvious which parts of the curve are affected by the different functional elements of the FDM/FM system (see Figure 1).

a. The 1/F noise region is generally degraded by power supply noise, station equipment noise, ground loops, baseband noise figure versus the discriminator constant, noise figure of the modulator baseband amplifier and various other man-made noise.

b. The S/N normal region is defined by the system's parameters and the noise figure of the receiver.

c. The S/N threshold is dependent on the IF noise bandwidth and the system parameters; however, the purity of the limiting and demodulation process can adversely affect this portion of the curve.

The S/N curves were calculated with the following parameters:

- 1/F noise floor = -110 dBm
- Baseband test tone = -25 dBm per channel
- Receiver noise figure = 13 dB
- Other parameters as shown on the curves

In actual practice satellite FDM/FM systems operate in the low C/N region, thereby allowing the modulator and demodulator to be optimized for a wider linearity bandwidth. Therefore, the 1/F noise floor as shown in the attached curves is set to -110 dBm for clarity and is not achieved in actual practice. The 1/F noise floor in practice runs approximately -87 dBm for 12-channel operation.

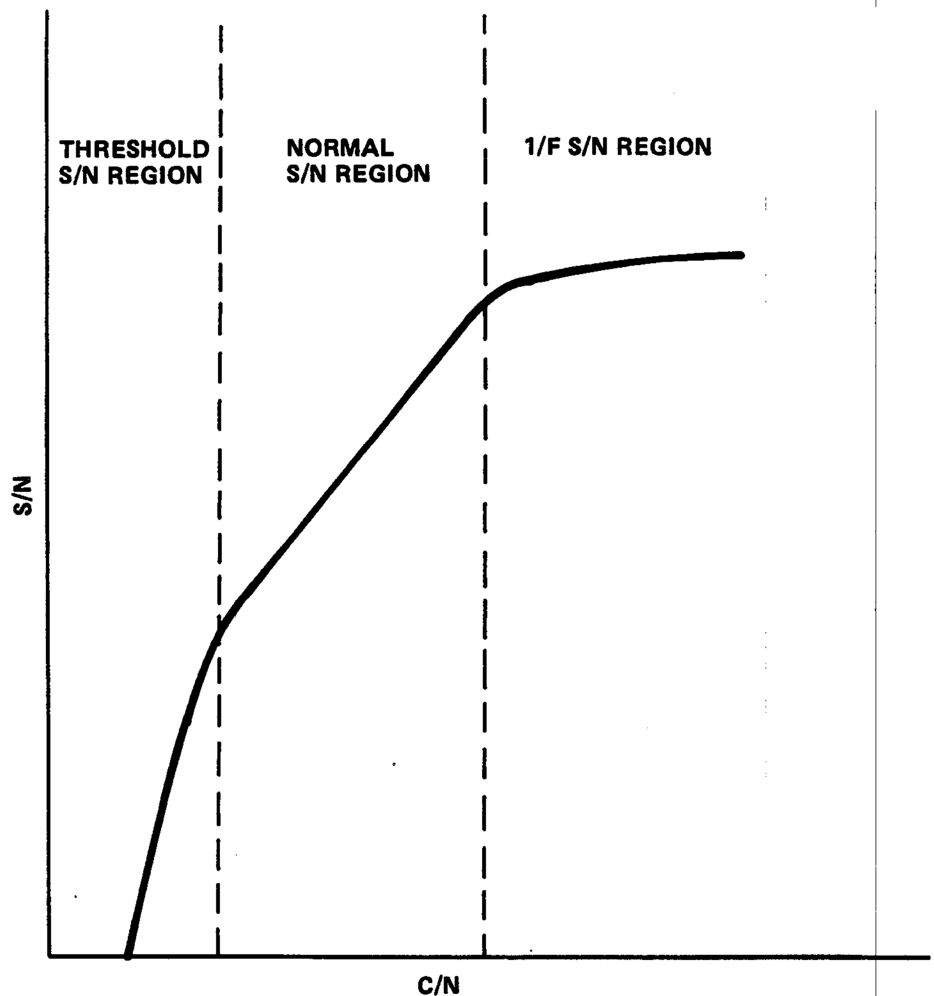


Figure 1. Conventional S/N vs. C/N Curve for FDM/FM Message Systems

The columns labeled NPR are actually NPR residual and do not include the intermodulation distortions generally associated with NPR measurements.

Signal-to-Noise Calculations

$$S/N_{normal} = \Big(10\ \log_{10}\ (2*BW/1E\text{-}3) + 10*\log_{10} T_k/290 - C_{dBm} + N_{fdB}$$

$$- 20\ \log_{10}\ (D_p/BW) - PE(X) - PW - 174\Big)$$

$$S/N_{threshold} = 10\ \log \left[1/3e^{-c/n} \left(\frac{B_{if}}{f_{ch}}\right)^2 \sqrt{1 + 24 \left(\frac{\sigma}{B_{if}}\right)^2 \left(\frac{C}{N}\right)^2} + 1\right]$$

$$S/N_{pwp} = 10^{\left(\frac{87.5 - S/N_{dB}}{10}\right)}$$

$$S/N_{dB} = 87.5 - 10\ \log S/N_{pwp}$$

$$C/N = C/KTB_{if}\ N_f$$

$$\sigma = 10^{\left(\frac{NLR}{20}\right)} \cdot D$$

$$C = \frac{10^{\left(\frac{C_{dBm}}{10}\right)}}{1000}$$

$$N_f = 10^{\left(\frac{N_{fdB}}{10}\right)}$$

σ = RMS Composite Deviation

C = Received Carrier in Watts

C_{dBm} = Received Carrier in dBm

K = 1.3804×10^{-23}

T_k = Temperature in Kelvin

f_{ch} = Notch Frequency

Log_{10} = Ln (X)/Ln (10)

PW = Psophometric Weighting

Dp = Peak Deviation of the Channel

$$PE = \text{Preemphasis} = 5 - 10 \log_{10} \left\{ \left[1 + \frac{\frac{6.90}{1 + 5.25}}{\left(\frac{f_r}{f_{ch}} - \frac{f_{ch}}{f_r} \right)^2} \right] \right\}$$

F_r = 1.25 f_{max}

f_{max} = Highest B/B Frequency

f_{low} = Lowest B/B Frequency

f_{ch} = Notch Frequency

S/N = NPR + BWR - NLR

$$BWR = 10 \log \frac{f_{max} - f_{low}}{3100}$$

NLR = 2.6 + 2 log N for 0 through 48 channels

NLR = -1 + 4 log N for 49 through 239 channels

NLR = -15 + 10 log N for 240 through 3600 channels

$$S/N = C/N + 10 \log \frac{BW_{if}}{6200} + 20 \log \frac{D_p}{f_{ch}}$$

Bandwidth Calculations

BW_o = Occupied bandwidth

BW_ℓ = Linearity/delay bandwidth

fm = Upper test cahnnel in MHz

Dv = Test tone rms deviation in MHz

Fp = Peak factor; 12 dB normally accepted

$$BW_o = 2\ fm + \log^{-1}\left[\frac{FP}{20}\right] Dv \log^{-1}\left(\frac{NLR}{20}\right)$$

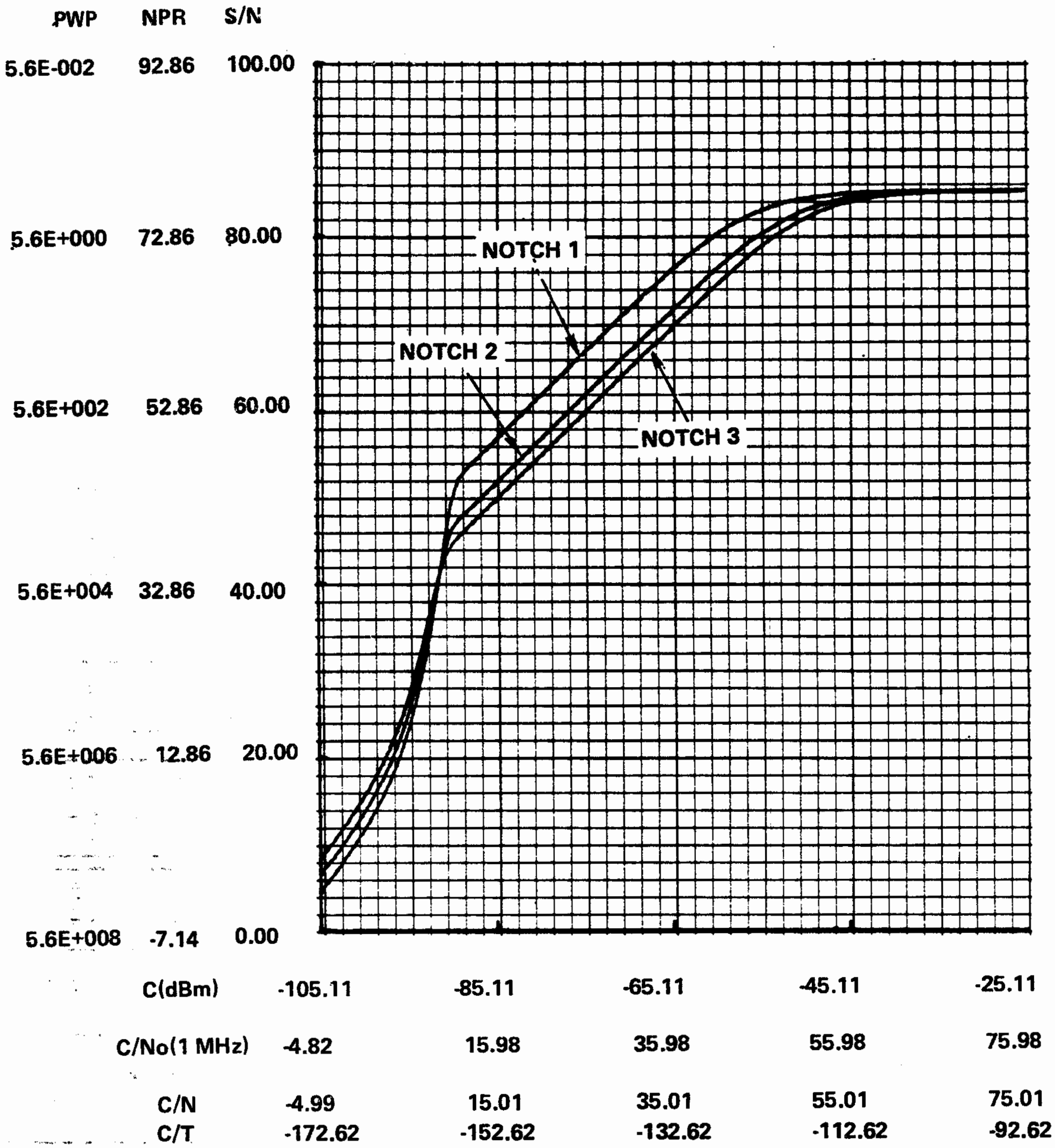
$BW_\ell = \pm 1/2\ BW_o$ setting $Fp = 0$ ($\log^{-1} Fp = 1$)

References

1. "Radio Relay Systems," International Telecommunications Union; CCIR, Volume 4, page 83.
2. Tarvin, J.C., "Microwave Signal to Noise Vs. Received Carrier Power Calculations for FM-FDM Systems," Rockwell International, WP-44.
3. "Psophometric Noise and Power," ITT Reference Data for Radio Engineers, 1968, Page 2-3.
4. Rice, S.O., "Noise in FM Receivers," Proceedings of the Symposium on Time Series Analysis, Wiley, 1963, page 395.
5. "Evaluation of S/N in the Threshold Region of an FM Receiver." Notes from M. Ramadan.
6. "Signal-to-Noise Computer Program." Notes from R. Harris.

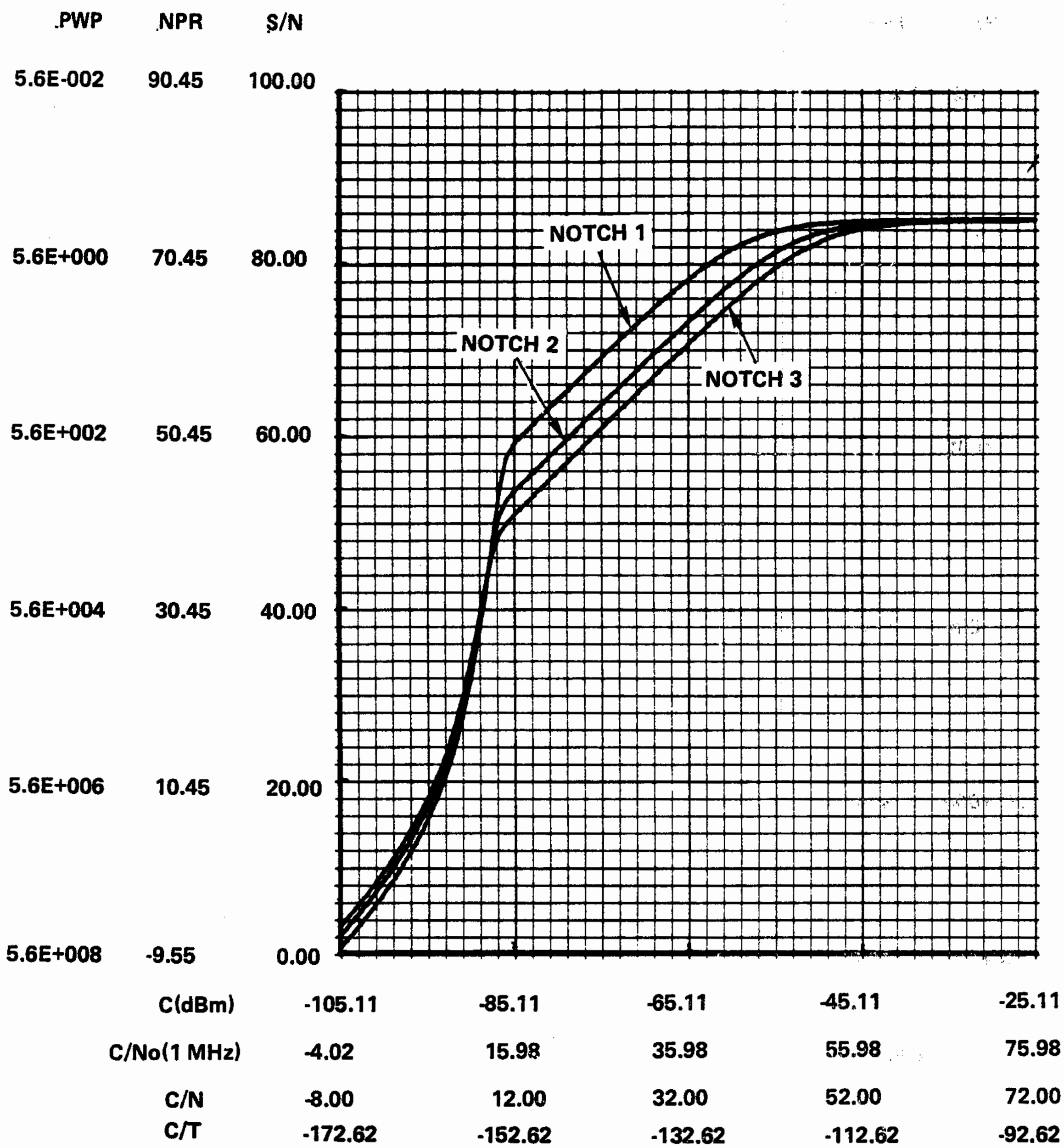
SCIENTIFIC-ATLANTA
MESSAGE DATA CHART

Channel Capacity	BW (MHz)	Top BB Freq. (kHz)	Test Tone Freq. (kHz)	Noise Load Ratio (dB)	RMS Test Dev. (kHz)	RMS MC Dev. (kHz)	Level into S-A Mod (140 kHz/-40 dBm) TT Dev. (dBm)	RMS Dev. (dBm)	1st Car. Null@ TT Freq.	S/N = NPR + (dB)	RX Max Sig. Level (dBm)	8000 pwp0 C/T	8000 pwp	50,000 pwp C/T	N 50, p
12	1.25	60	36.48	3.32	109	159	-42.17	-38.85	-47.07	8.57	-47	-154.7	39.90	-160.7	31
24	2.5	108	65.66	+ 4.52	164	275	-38.63	-34.11	-41.96	10.39	-45	-153.0	38.11	-159.0	30
36	2.5	156	94.85	+ 5.23	168	307	-38.42	-33.18	-38.77	11.44	-45	-150.0	37.06	-156.0	29
60	2.5	252	153.2	6.11	136	276	-40.25	-34.14	-34.61	12.78	-45	-144.0	35.72	-150.0	27
60	5.0	252	153.2	6.11	270	546	-34.30	-28.19	-34.61	12.78	-42	-149.9	35.72	-155.9	27
72	2.5	300	182.4	6.43	125	261	-40.98	-34.55	-33.09	13.25	-45	-141.7	35.25	-147.7	27
72	5.0	300	182.4	6.43	294	616	-33.56	-27.12	-33.09	13.25	-42	-149.1	35.25	-155.1	27
96	5.0	408	248.1	6.93	263	584	-34.52	-27.59	-30.42	14.13	-42	-145.5	34.37	-151.5	26
96	7.5	408	248.1	6.93	360	799	-31.80	-24.87	-30.42	14.13	-40	-148.2	34.37	-154.2	26
132	5.0	552	335.6	7.48	223	529	-35.96	-28.48	-27.79	14.93	-42	-141.4	33.57	-147.4	25
132	7.5	552	335.6	7.48	376	891	-31.42	-23.94	-27.79	14.93	-40	-145.9	33.57	-151.9	25
132	10.0	552	335.6	7.48	430	1020	-30.25	-22.77	-27.79	14.93	-39	-147.1	33.57	-153.1	25
192	5.0	804	488.8	8.13	180	459	-37.82	-29.69	-24.53	15.94	-42	-136.3	32.56	-142.3	24
192	7.5	804	488.8	8.13	297	758	-33.47	-25.34	-24.53	15.94	-40	-140.6	32.56	-146.6	24
192	10.0	804	488.8	8.13	457	1167	-29.72	-21.59	-24.53	15.94	-39	-144.4	32.56	-150.4	24
252	7.5	1052	639.6	9.01	260	733	-34.62	-25.61	-22.19	16.25	-40	-137.1	32.25	-143.1	24
252	10	1052	639.6	9.01	358	1009	-31.84	-22.83	-22.19	16.25	-39	-139.9	32.25	-145.9	24
252	15	1052	639.6	9.01	577	1627	-27.70	-18.69	-22.19	16.25	-37	-144.1	32.25	-150.1	24
312	10	1300	790.4	9.94	320	1005	-32.82	-22.88	-20.35	16.25	-39	-137.1	32.25	-143.1	24
312	15	1300	790.4	9.94	546	1716	-28.17	-18.23	-20.35	16.25	-36	-141.7	32.25	-147.7	24
432	15	1796	1092	11.35	401	1479	-30.86	-19.51	-17.55	16.25	-37	-136.2	32.25	-142.2	24
432	17.5	1796	1092	11.35	517	1919	-28.65	-17.30	-17.55	16.25	-37	-138.5	32.25	-144.5	24
432	20	1796	1092	11.35	616	2276	-27.13	-15.78	-17.55	16.25	-36	-139.9	32.25	-145.9	24
432	25	1796	1092	11.35	729	2688	-25.67	-14.32	-17.55	16.25	-35	-141.4	32.25	-147.4	24
612	20	2540	1544	12.87	454	1996	-29.78	-16.91	-14.54	16.25	-36	-134.2	32.25	-140.2	24
792	20	3284	1997	13.99	356	1784	-31.89	-17.90	-12.30	16.24	-36	-129.9	32.26	-135.9	24
792	25	3284	1997	13.99	499	2494	-28.96	-14.97	-12.30	16.24	-35	-132.8	32.26	-138.8	24
972	25	4028	2449	14.88	410	2274	-30.67	-15.79	-10.53	16.24	-35	-129.4	32.26	-135.4	24
972	36	4028	2449	14.88	802	4417	-24.84	- 9.96	-10.53	16.24	-34	-135.2	32.26	-141.7	24
1092	36	4892	2974	15.38	701	4118	-26.01	-10.62	- 8.84	16.59	-34	-132.4	31.91	-138.4	21
1332	36	5884	3577	16.25	591	3834	-27.49	-11.24	- 7.24	16.52	-34	-129.3	31.98	-135.3	21
1872	36	8120	4937	17.72	419	3181	-30.48	-12.75	- 4.44	16.46	-34	-123.5	32.04	-129.5	24



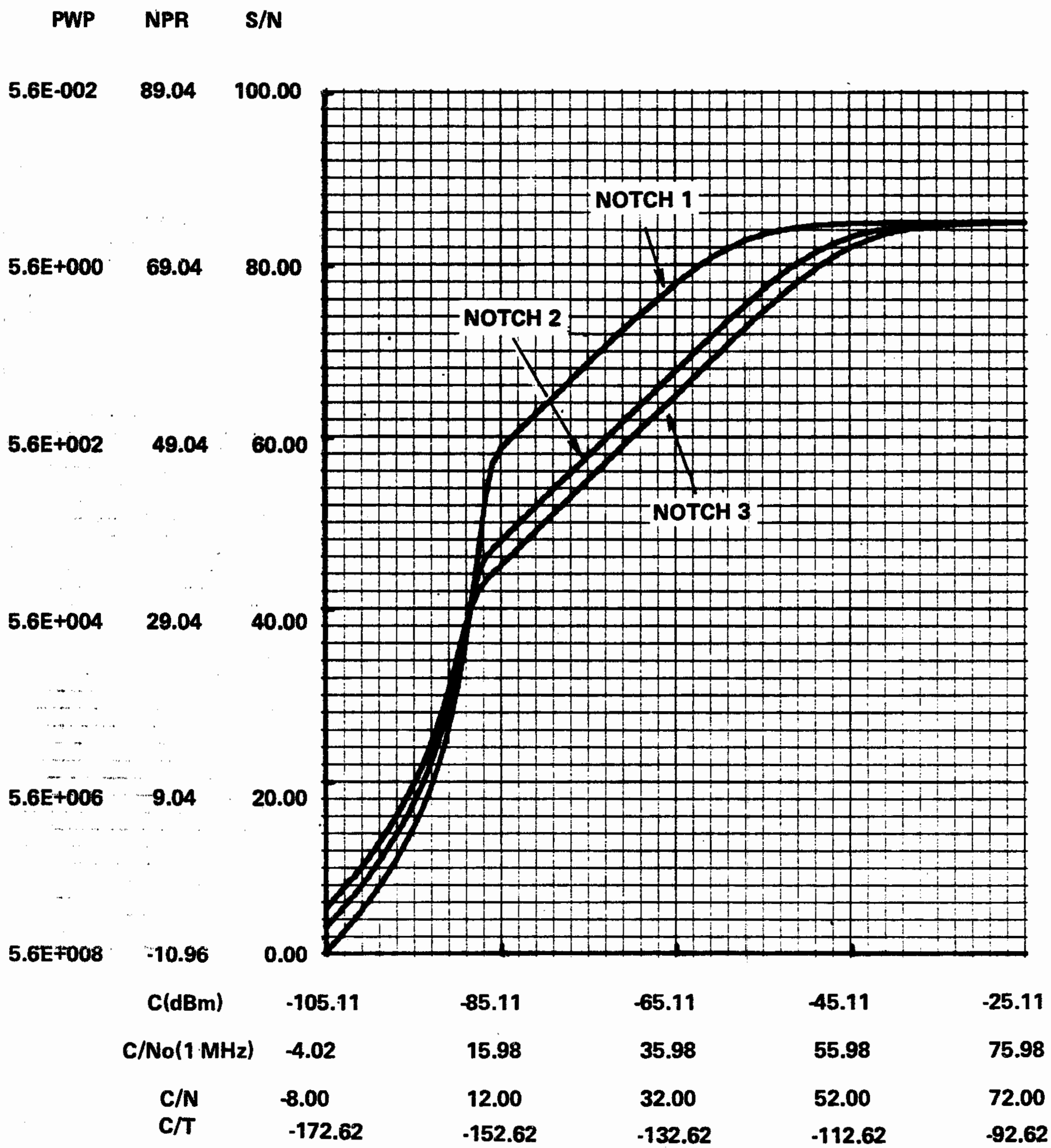
N. F. (dB)	13
BW (I. F.)	1250000
CHANNEL LOADING	12
NOTCH 1	16000
NOTCH 2	36000
NOTCH 3	56000
OCCUPIED BW	1620978.36
LINEARITY BW	497028.88
T. T. DEVIATION (RMS)	109000
P. E.	YES
P. S.	NO

Figure 3. S/N vs. C/N for 12 Channel Loading



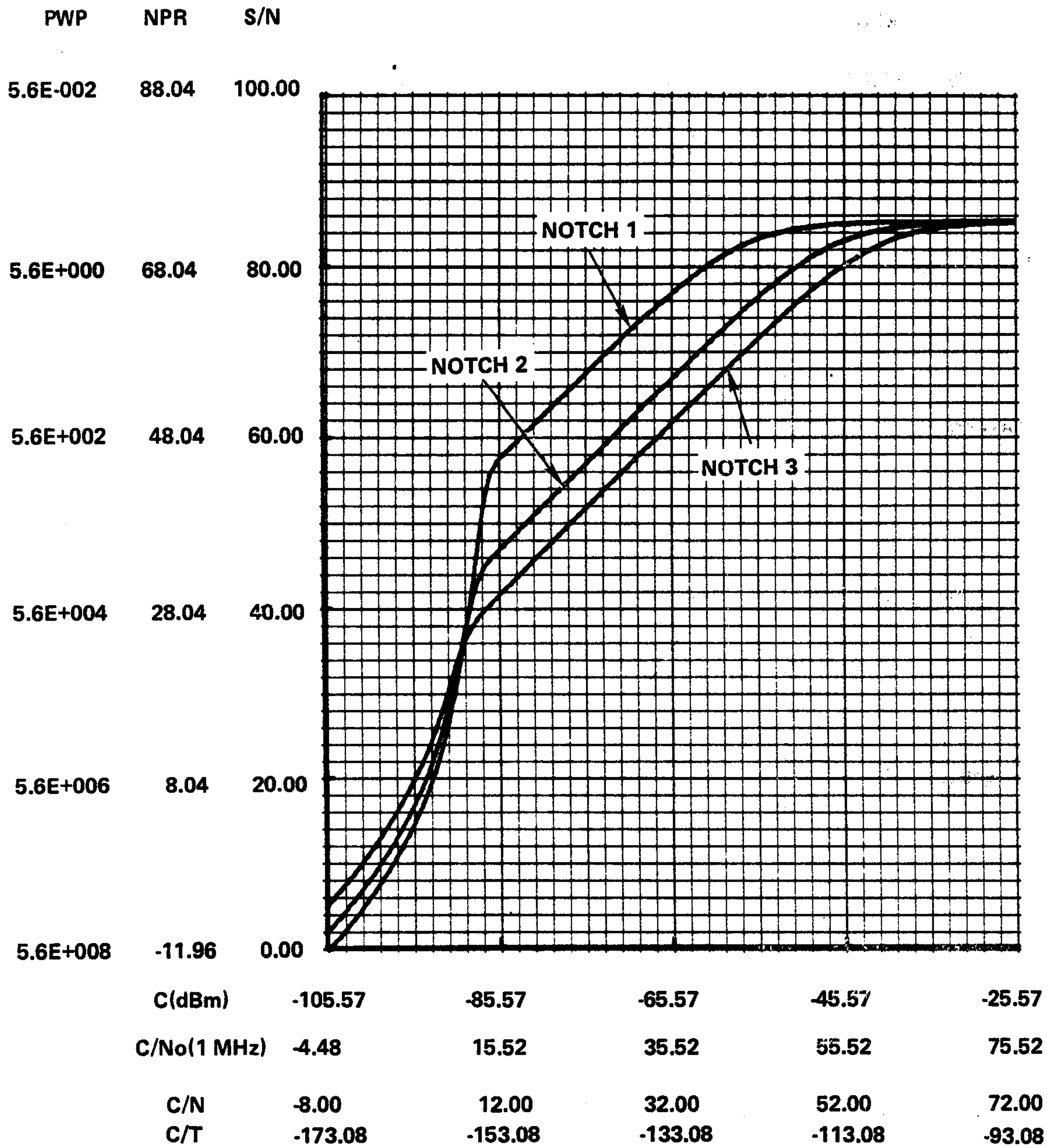
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	24
NOTCH 1	16000
NOTCH 2	36000
NOTCH 3	56000
OCCUPIED BW	2636442.57
LINEARITY BW	823887.94
T. T. DEVIATION (RMS)	164000
P. E.	YES
P. S.	NO

Figure 4. S/N vs. C/N for 24 Channel Loading



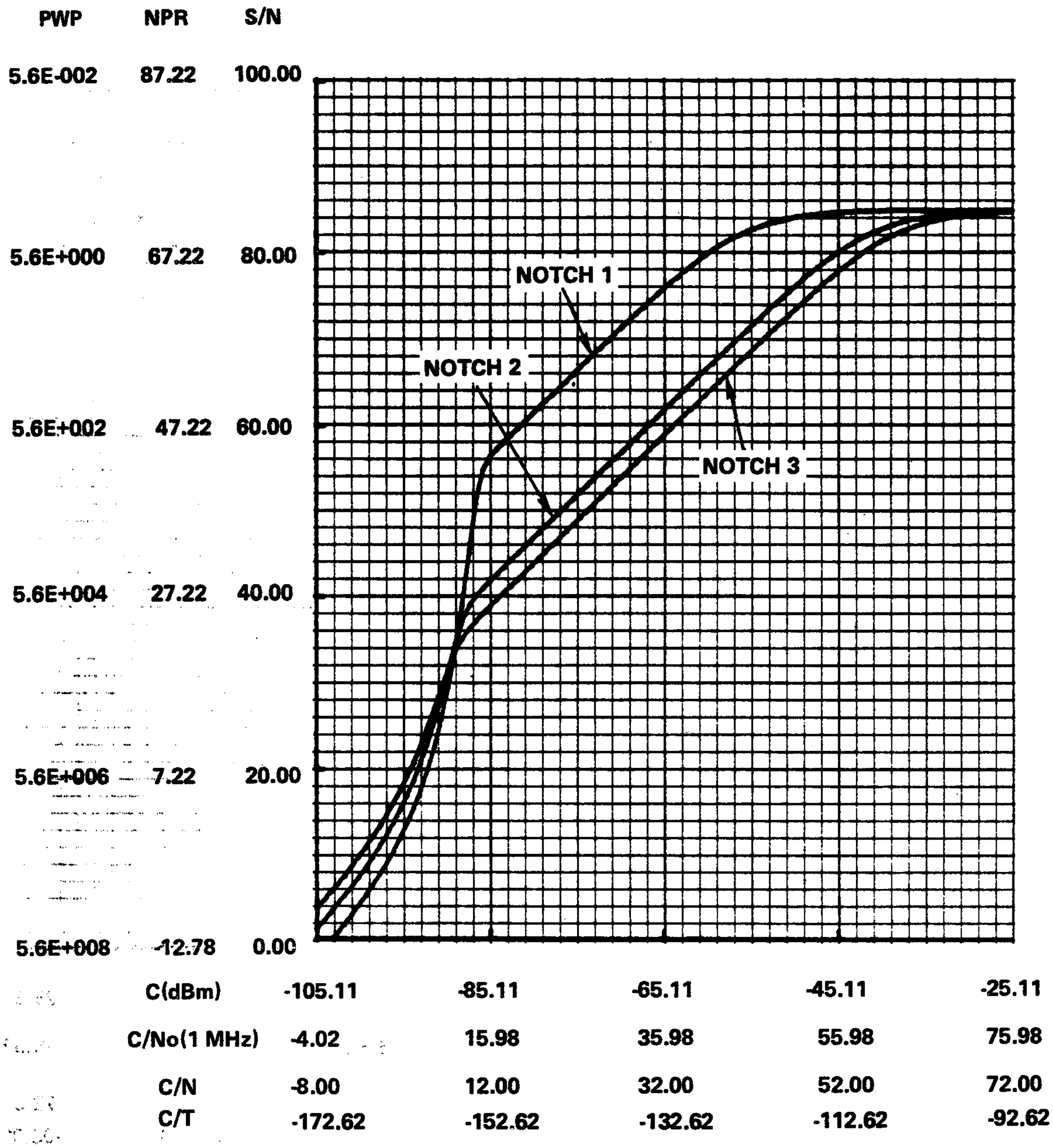
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	36
NOTCH 1	16000
NOTCH 2	78000
NOTCH 3	140000
OCCUPIED BW	2894077.90
LINEARITY BW	960588.92
T. T. DEVIATION (RMS)	168000
P. E.	YES
P. S.	NO

Figure 5. S/N vs. C/N for 36 Channel Loading



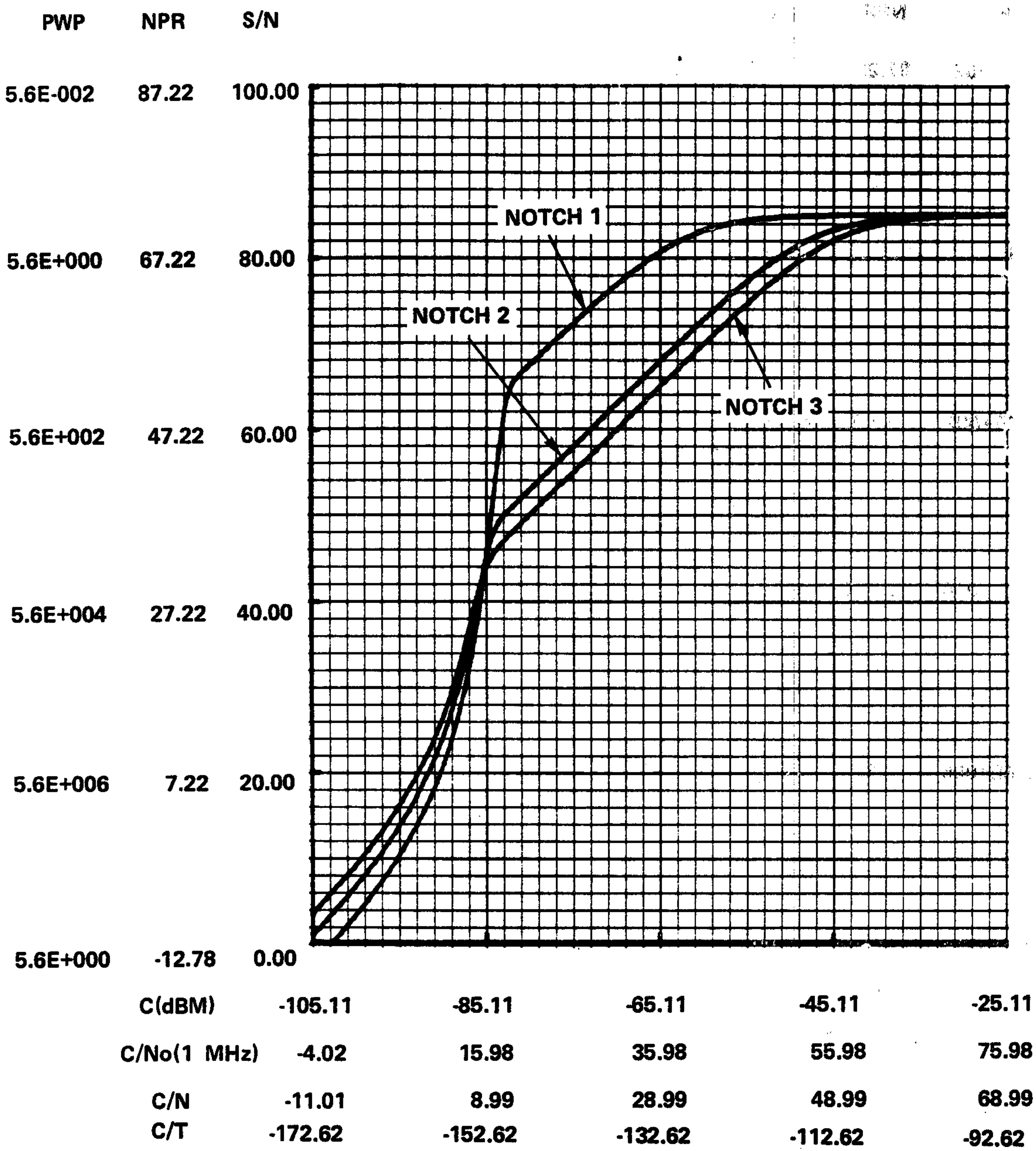
N. F. (dB)	13
BW (I. F.)	2250000
CHANNEL LOADING	48
NOTCH 1	16000
NOTCH 2	70000
NOTCH 3	185000
OCCUPIED BW	2796530.88
LINEARITY BW	1007872.12
T. T. DEVIATION (RMS)	151000
P. E.	YES
P. S.	NO

Figure 6. S/N vs. C/N for 48 Channel Loading



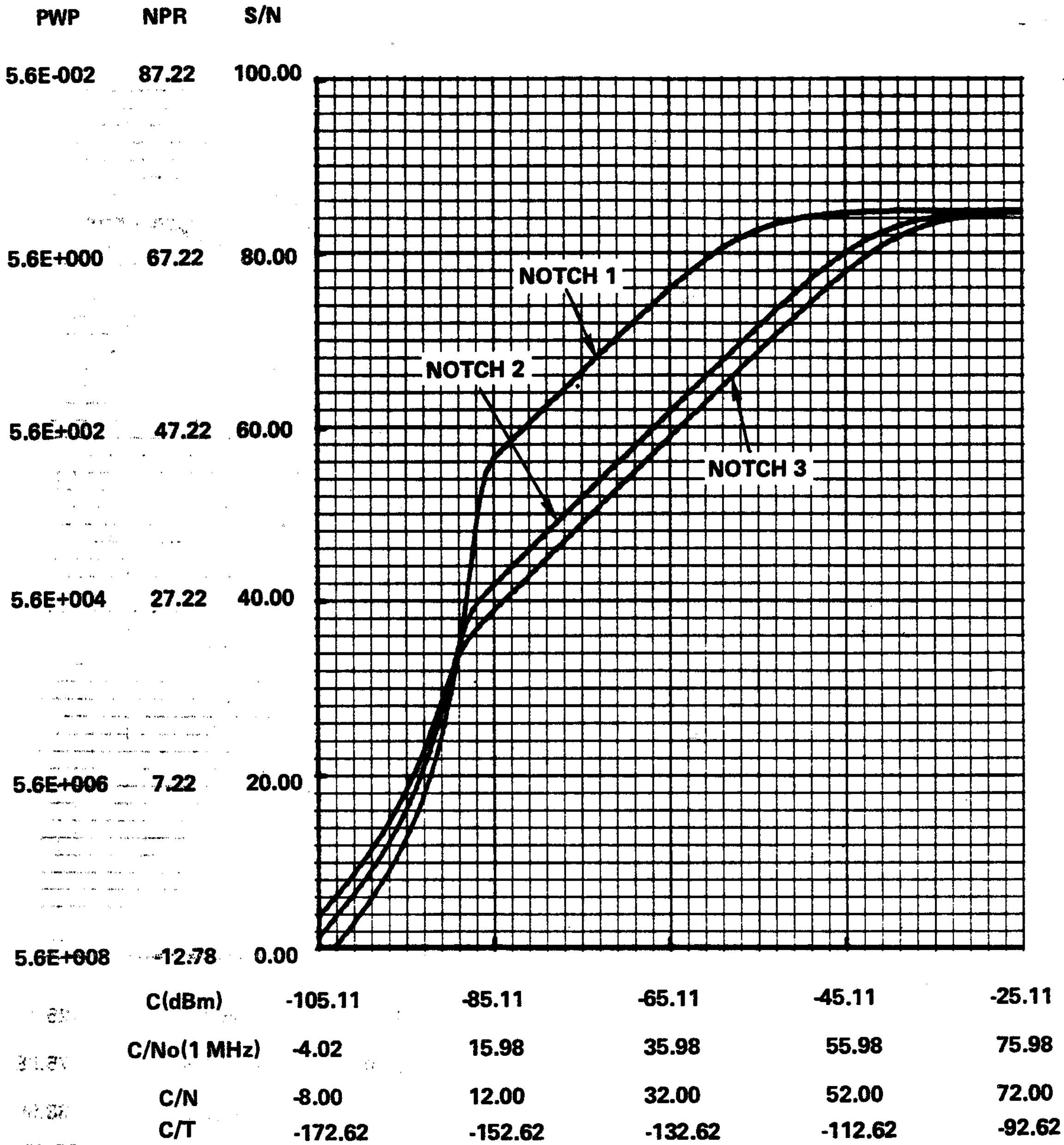
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	128000
NOTCH 3	240000
OCCUPIED BW	2692764.16
LINEARITY BW	1053792.93
T. T. DEVIATION (RMS)	136000
P. E.	YES
P. S.	NO

Figure 7. S/N vs. C/N for 60 Channel Loading



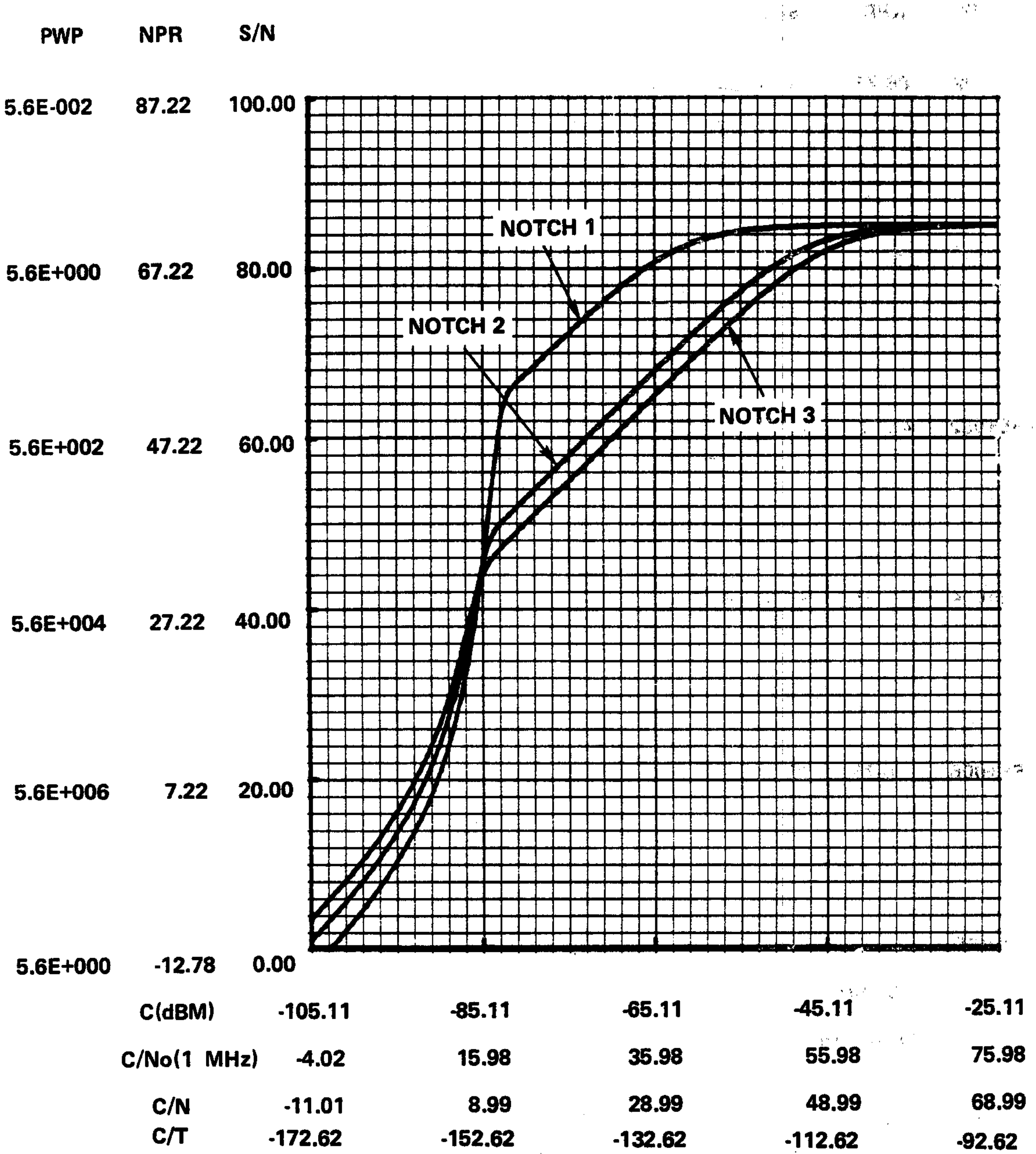
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	120000
NOTCH 3	240000
OCCUPIED BW	4849430.61
LINEARITY BW	1595500.68
T. T. DEVIATION (RMS)	270000
P. E.	YES
P. S.	NO

Figure 8. S/N vs. C/N for 60 Channel Loading



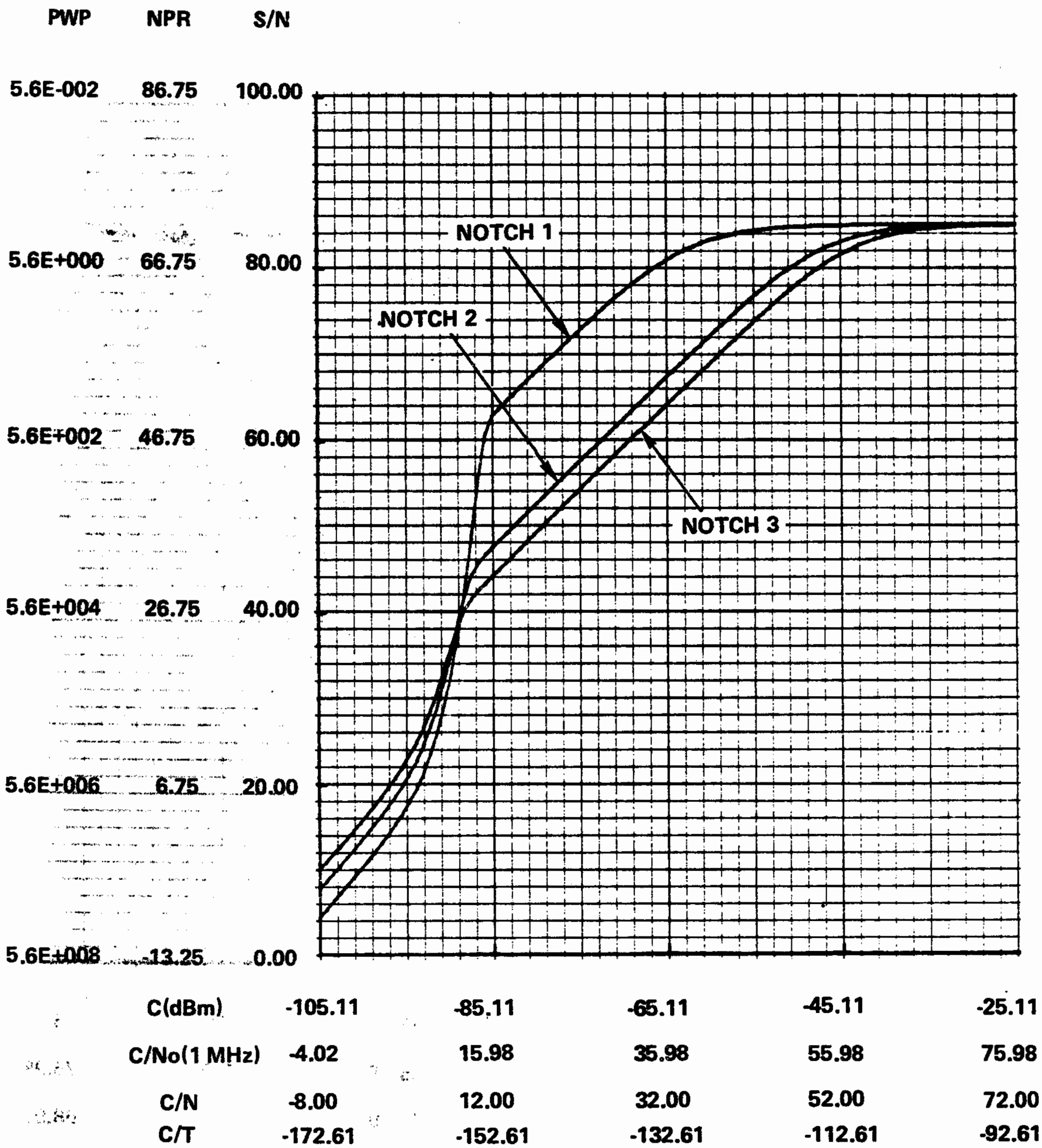
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	128000
NOTCH 3	240000
OCCUPIED BW	2692764.16
LINEARITY BW	1053792.93
T. T. DEVIATION (RMS)	136000
P. E.	YES
P. S.	NO

Figure 7. S/N vs. C/N for 60 Channel Loading



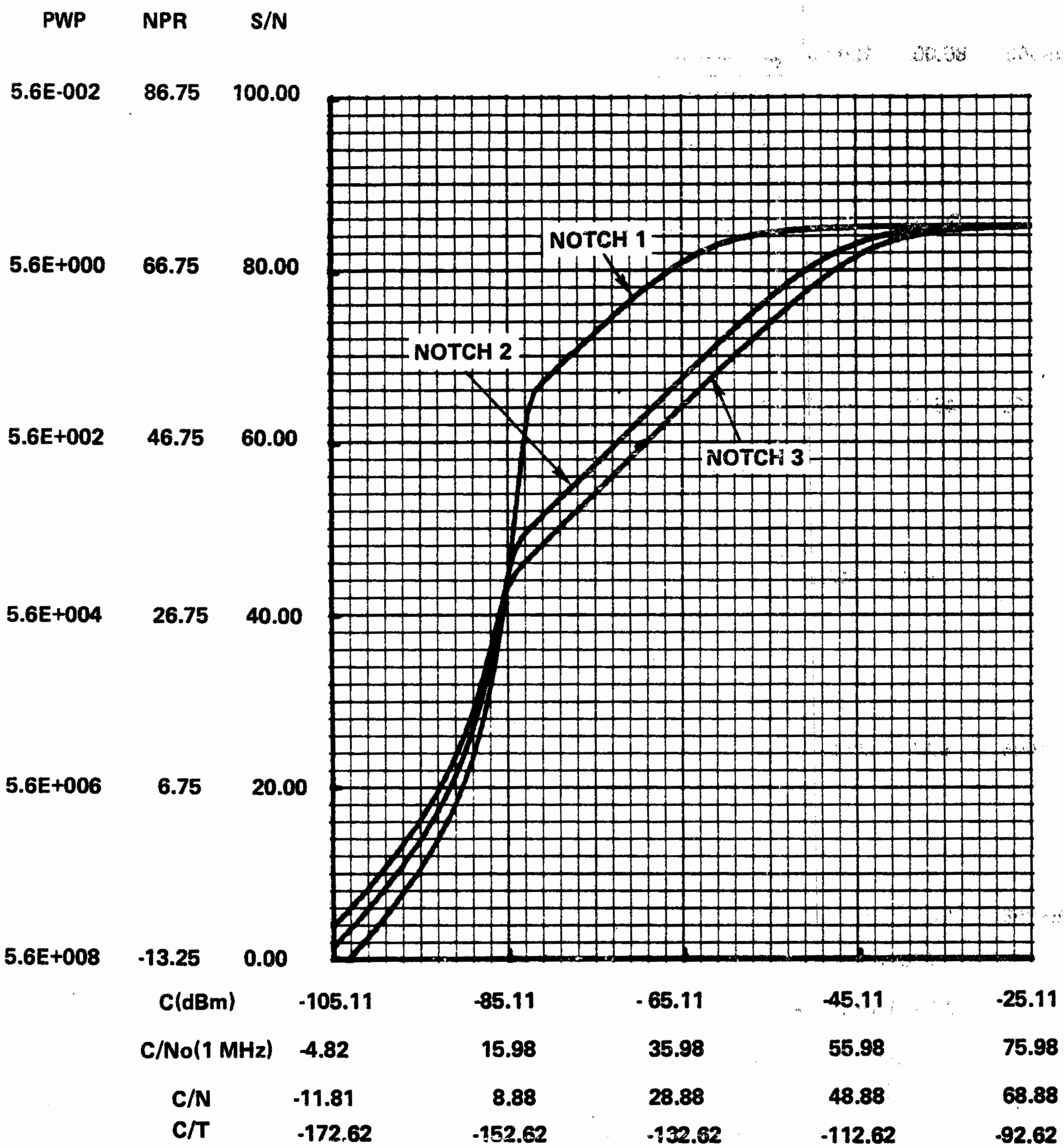
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	120000
NOTCH 3	240000
OCCUPIED BW	4849430.61
LINEARITY BW	1595500.68
T. T. DEVIATION (RMS)	270000
P. E.	YES
P. S.	NO

Figure 8. S/N vs. C/N for 60 Channel Loading



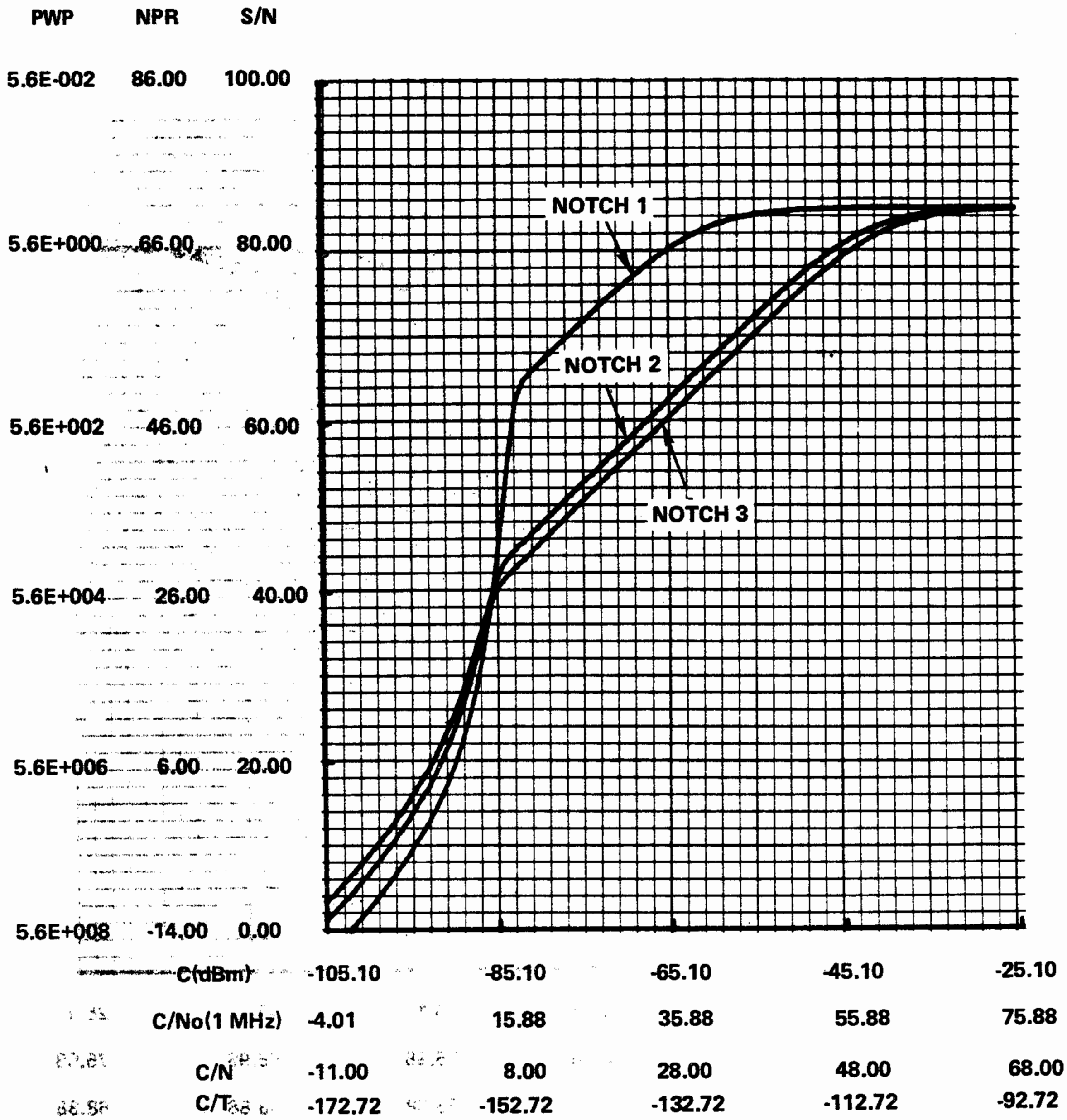
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	72
NOTCH 1	16000
NOTCH 2	140000
NOTCH 3	270000
OCCUPIED BW	5507311.65
LINEARITY BW	1832661.48
T. T. DEVIATION (RMS)	294000
P. E.	YES
P. S.	NO

Figure 9. S/N vs. C/N for 72 Channel Loading



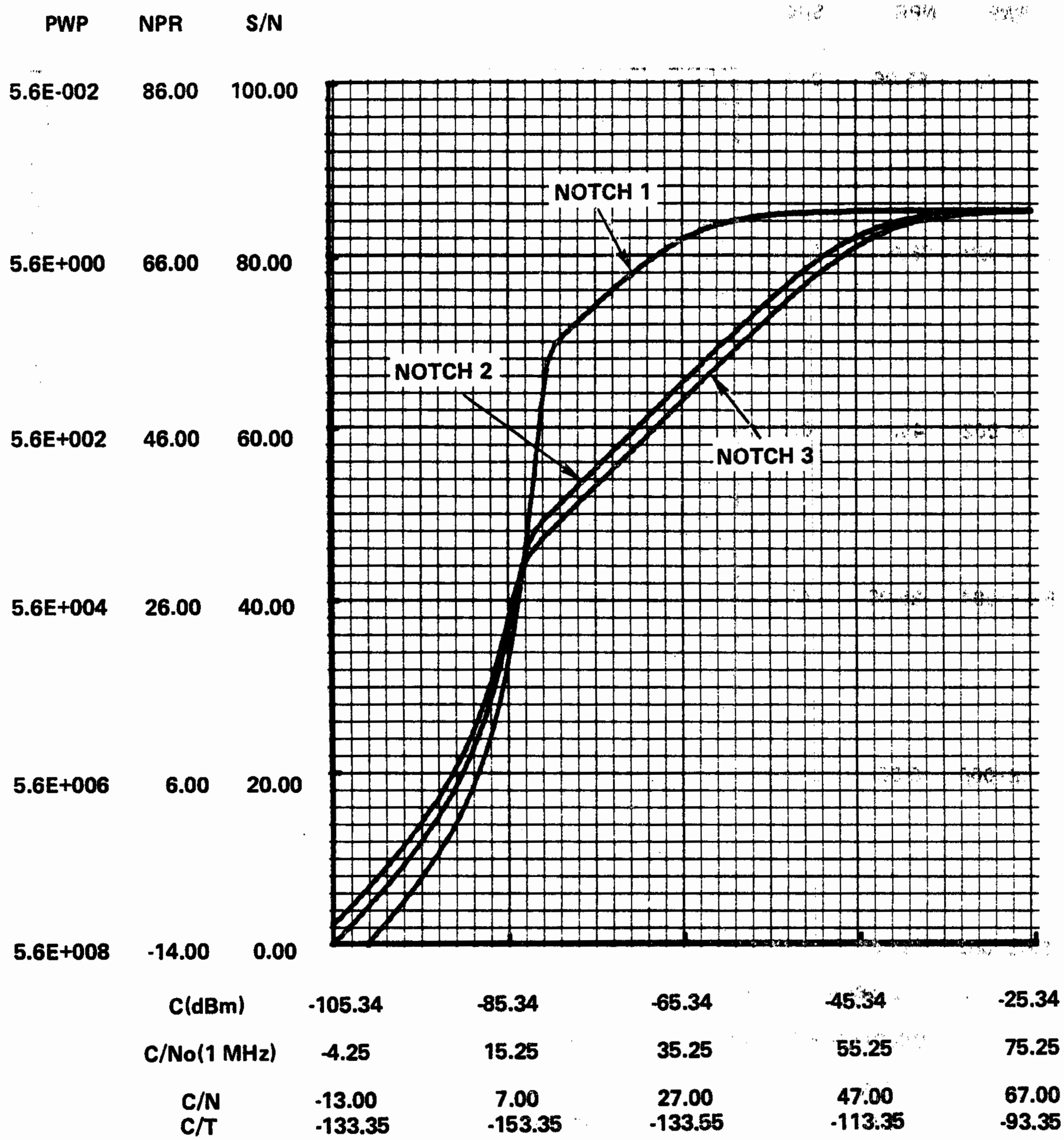
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	72
NOTCH 1	16000
NOTCH 2	140000
NOTCH 3	270000
OCCUPIED BW	5507311.65
LINEARITY BW	1832661.48
T. T. DEVIATION (RMS)	294000
P. E.	YES
P. S.	NO

Figure 10. S/N vs. C/N for 72 Channel Loading



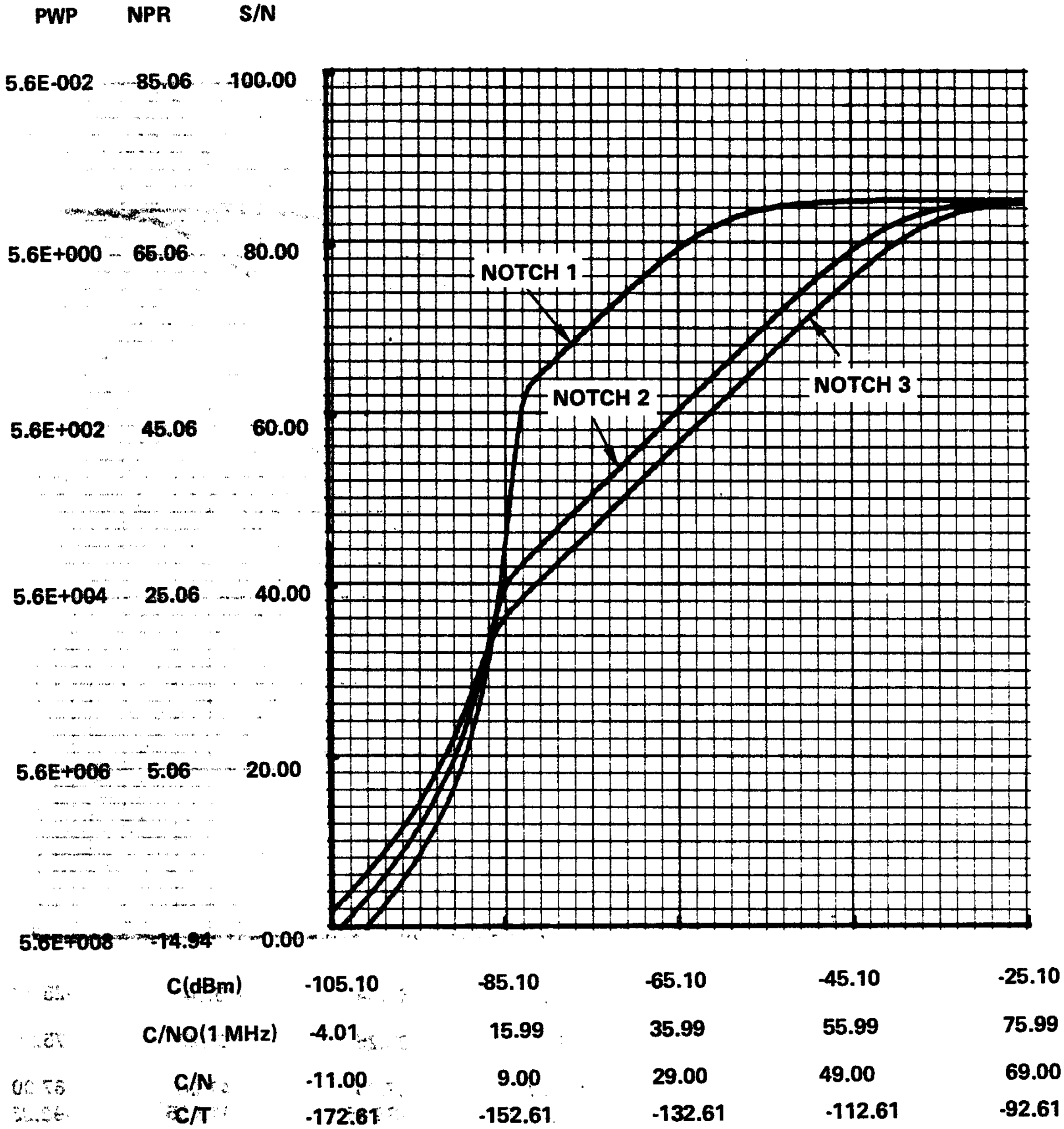
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	96
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	394000
OCCUPIED BW	5441858.22
LINEARITY BW	18588882.33
T. T. DEVIATION (RMS)	263000
P. E.	YES
P. S.	NO

Figure 11. S/N vs. C/N for 96 Channel Loading



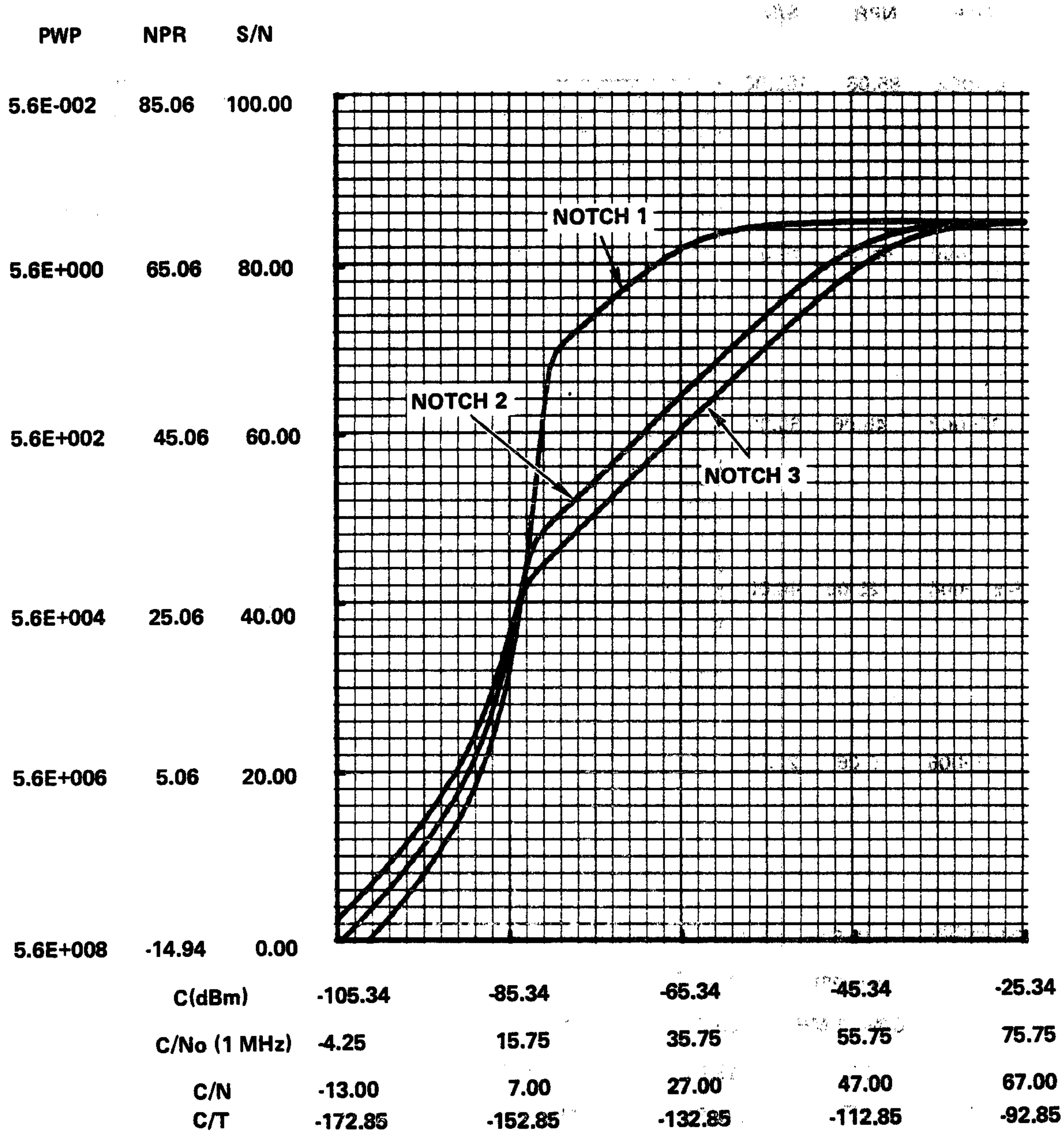
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	96
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	394000
OCCUPIED BW	7156826.31
LINEARITY BW	2388772.77
T. T. DEVIATION (RMS)	360000
P. E.	YES
P. S.	NO

Figure 12. S/N vs. C/N for 96 Channel Loading



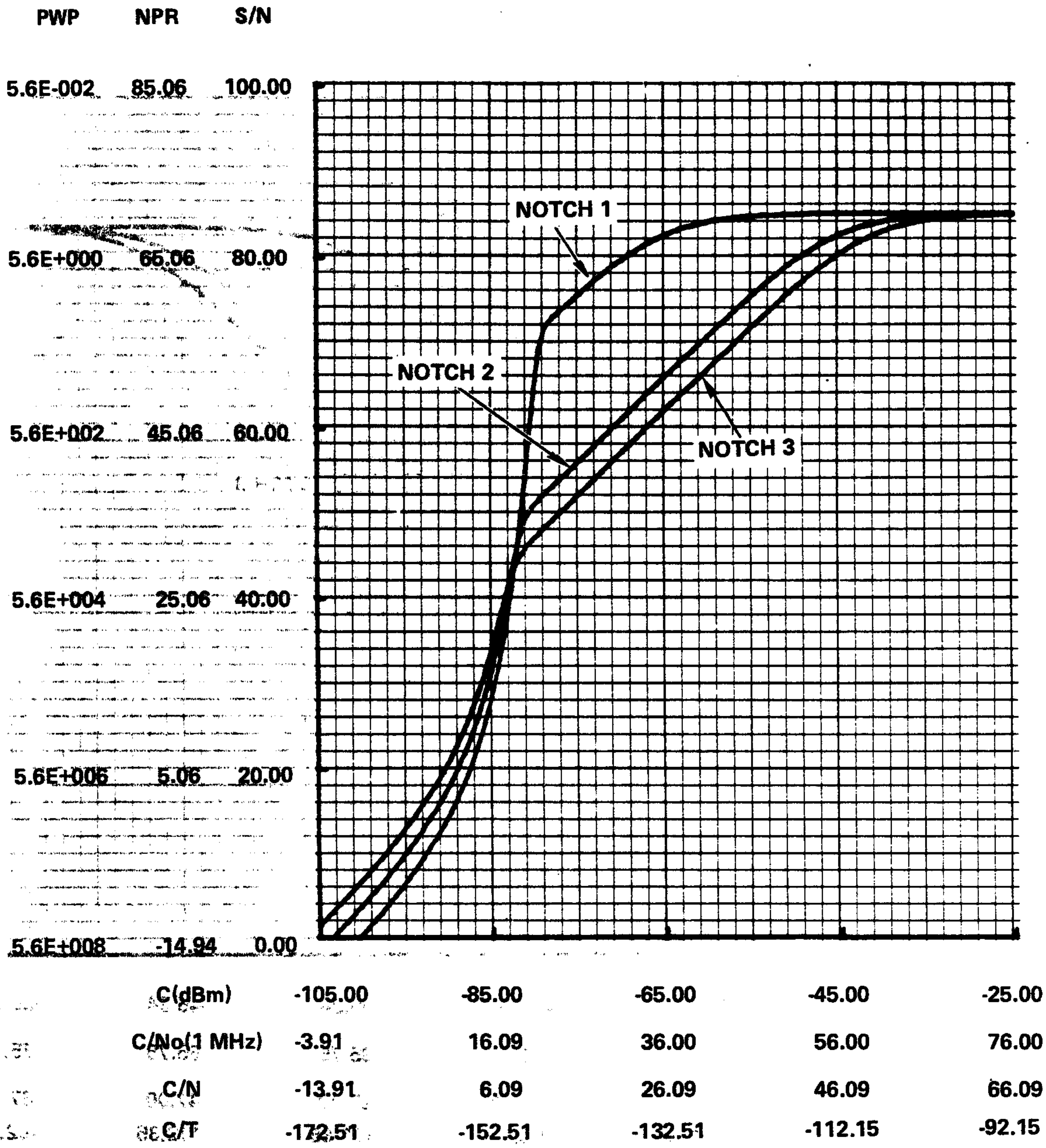
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	5308336.25
LINEARITY BW	2161879.12
T. T. DEVIATION (RMS)	223000
P. E.	YES
P. S.	NO

Figure 13. S/N vs. C/N for 132 Channel Loading



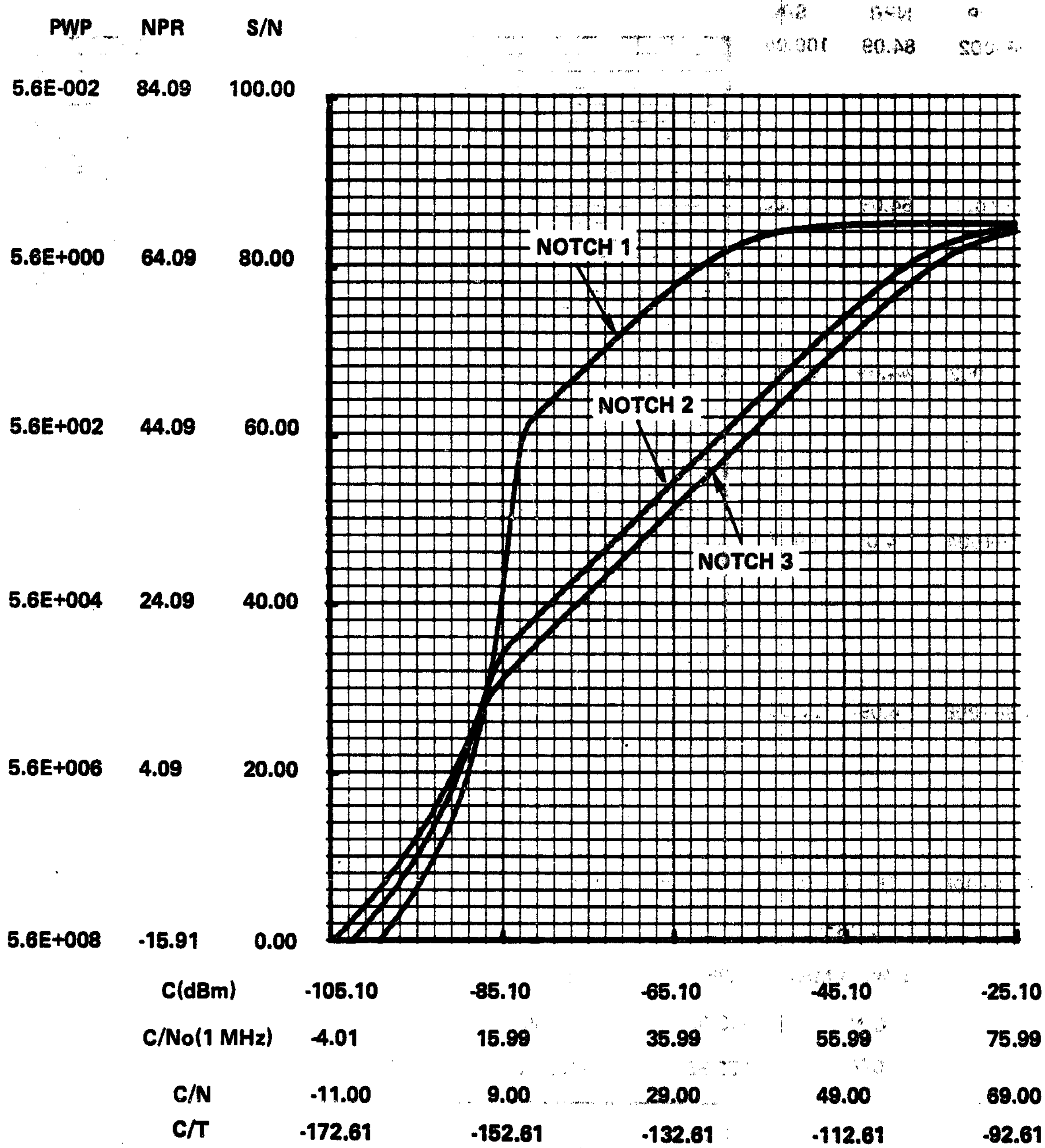
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	8191279.05
LINEARITY BW	2886041.92
T. T. DEVIATION (RMS)	376000
P. E.	YES
P. S.	NO

Figure 14. S/N vs. C/N for 132 Channel Loading



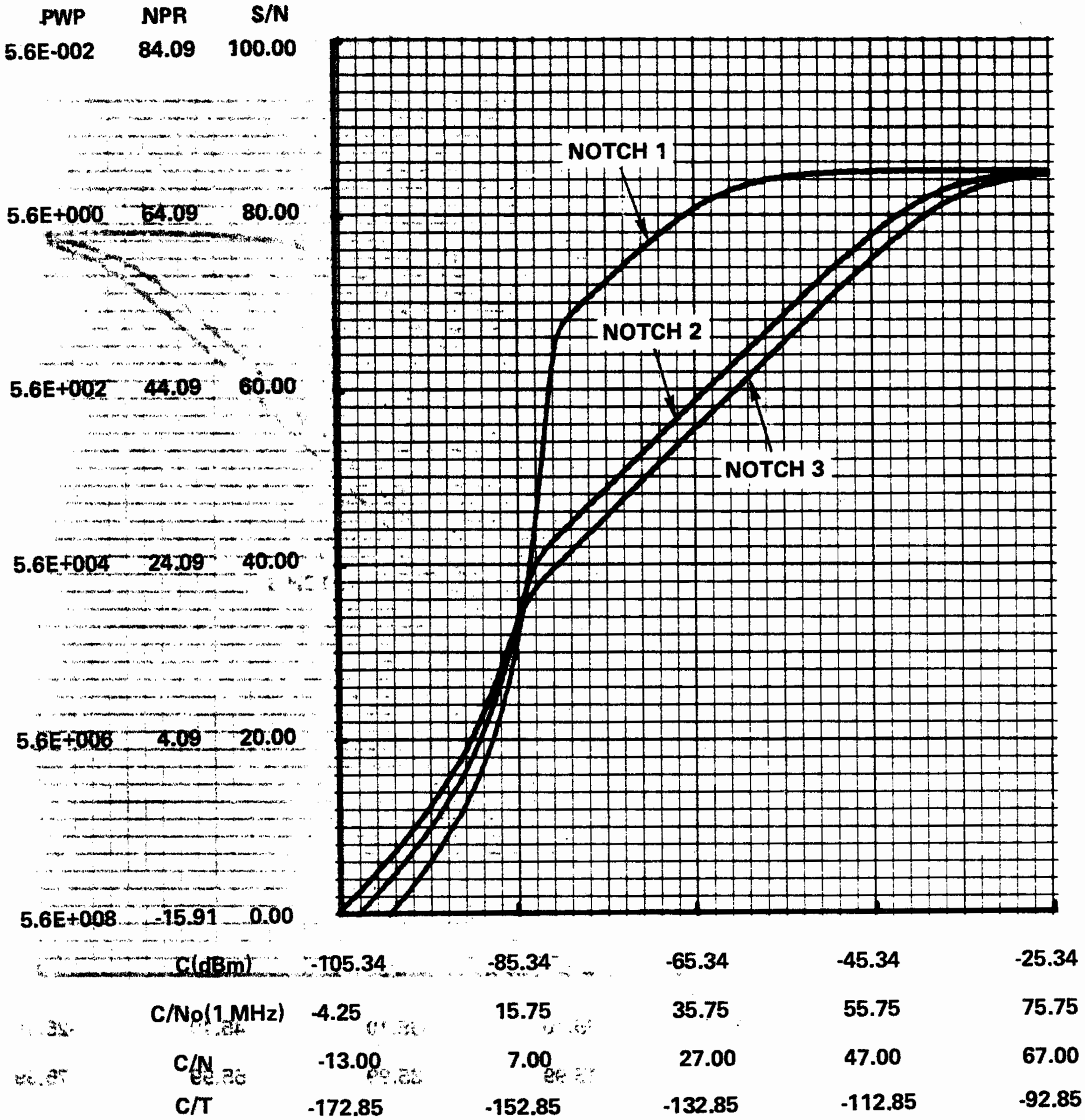
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	9288788.28
LINEARITY BW	3141628.79
T. T. DEVIATION (RMS)	430000
P. E.	YES
P. S.	NO

Figure 15. S/N vs. C/N for 132 Channel Loading



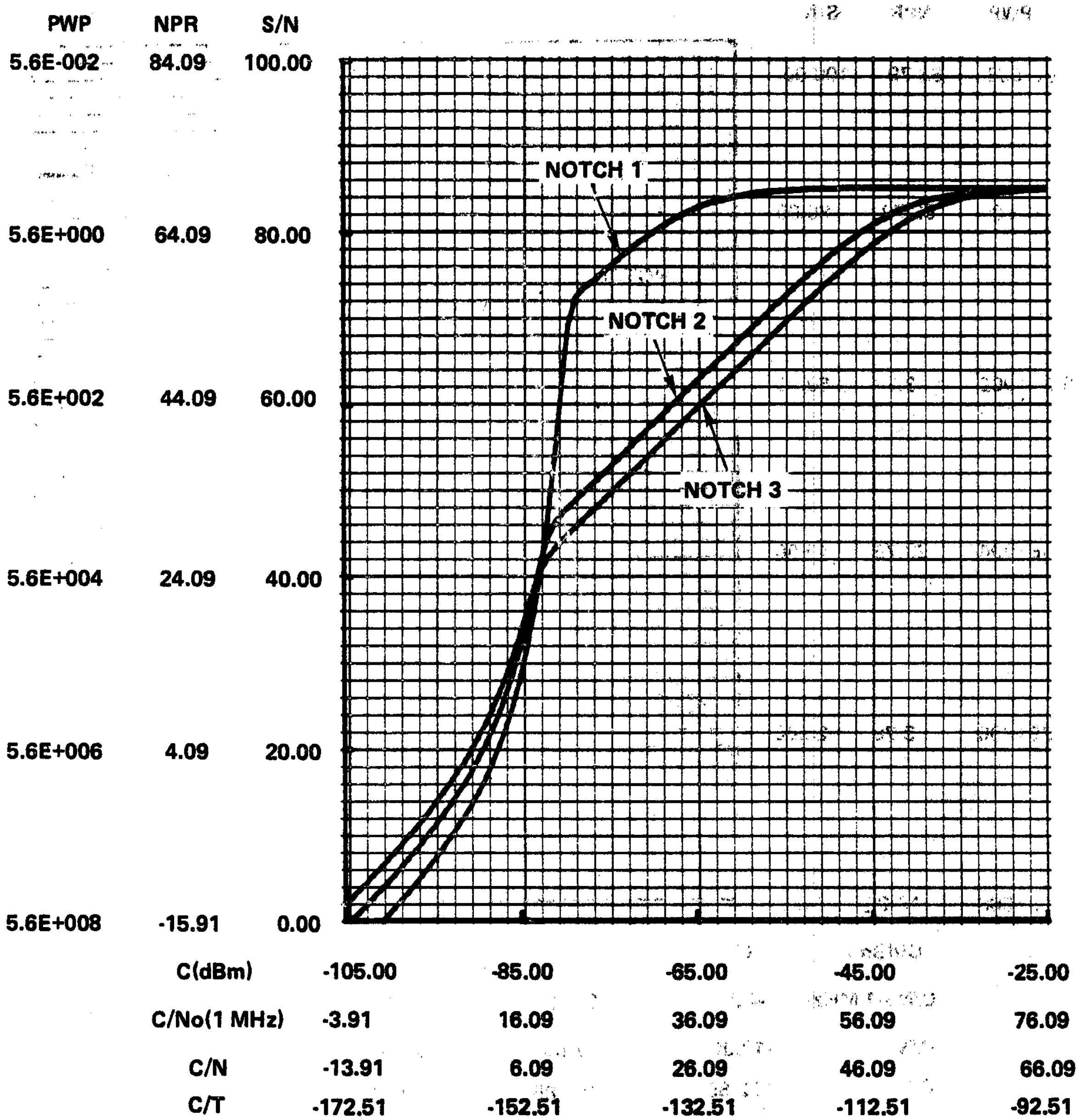
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	5254032.76
LINEARITY BW	2516853.83
T.T. DEVIATION (RMS)	180000
P. E.	YES
P. S.	NO

Figure 16. S/N vs. C/N for 192 Channel Loading



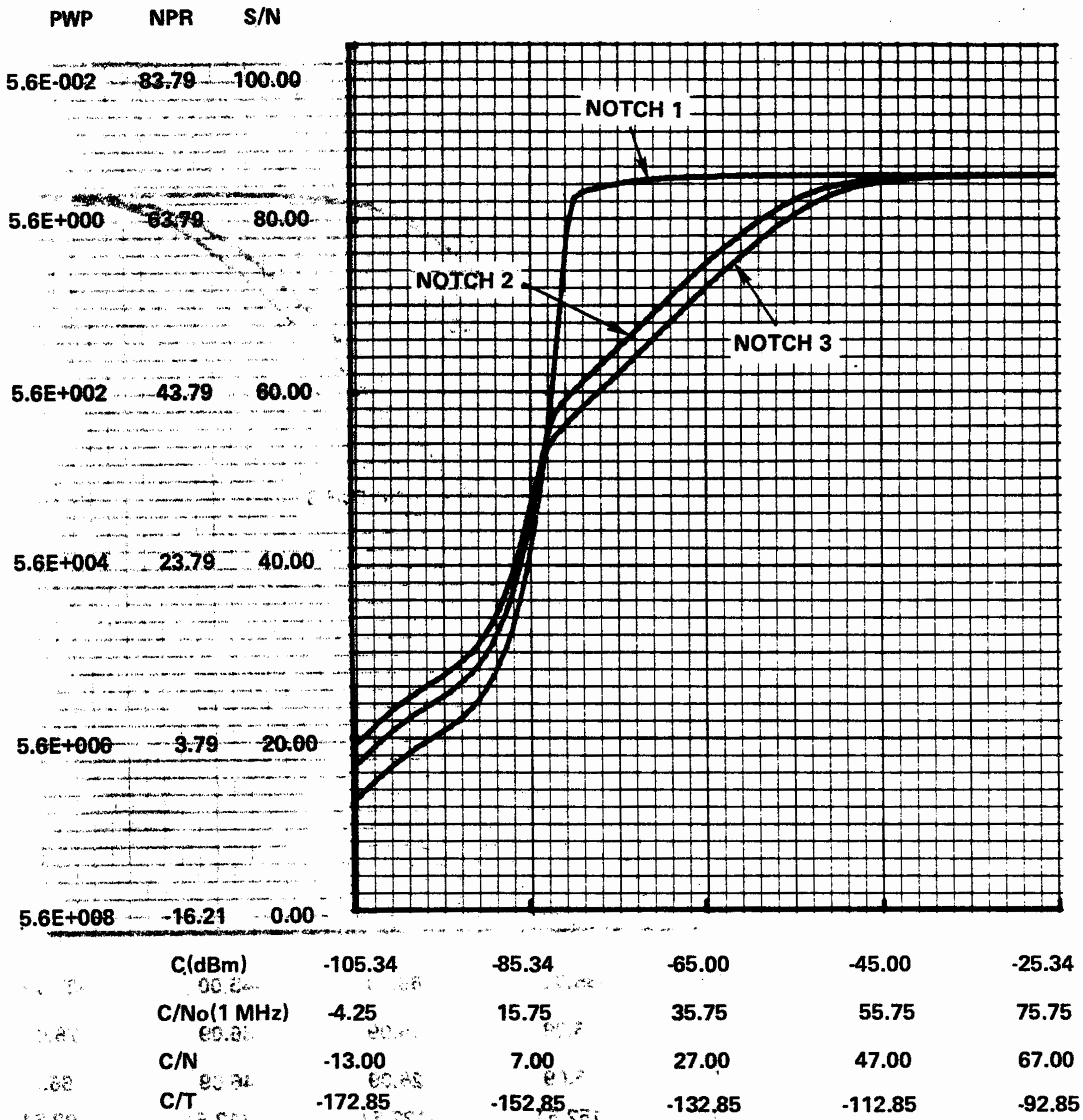
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	7630194.05
LINEARITY BW	3113518.81
T. T. DEVIATION (RMS)	297000
P. E.	YES
P/S.	NO

Figure 17. S/N vs. C/N for 192 Channel Loading



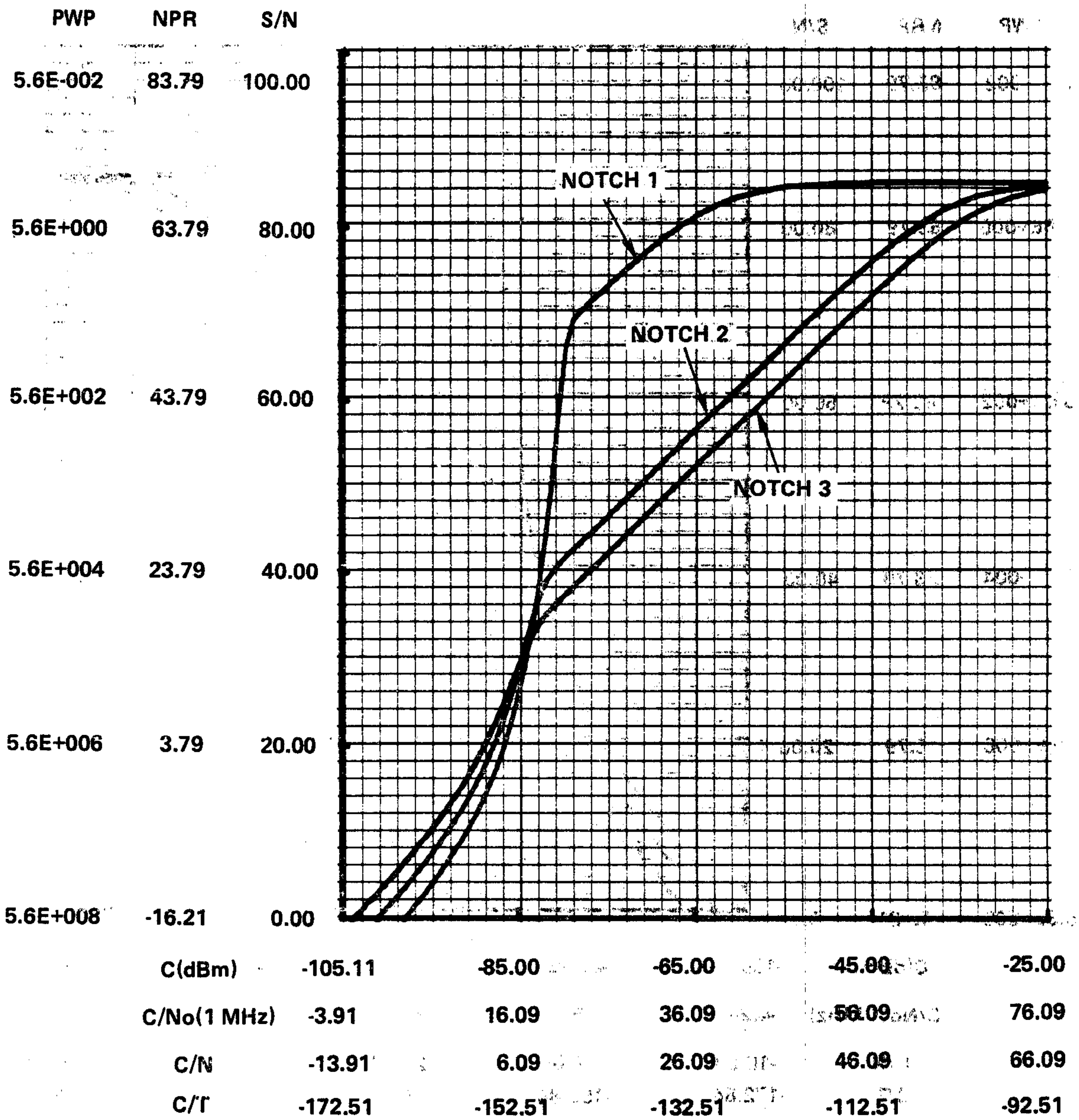
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	10879645.38
LINEARITY BW	3929744.44
T. T. DEVIATION (RMS)	457000
P. E.	YES
P. S.	NO

Figure 18. S/N vs. C/N for 192 Channel Loading



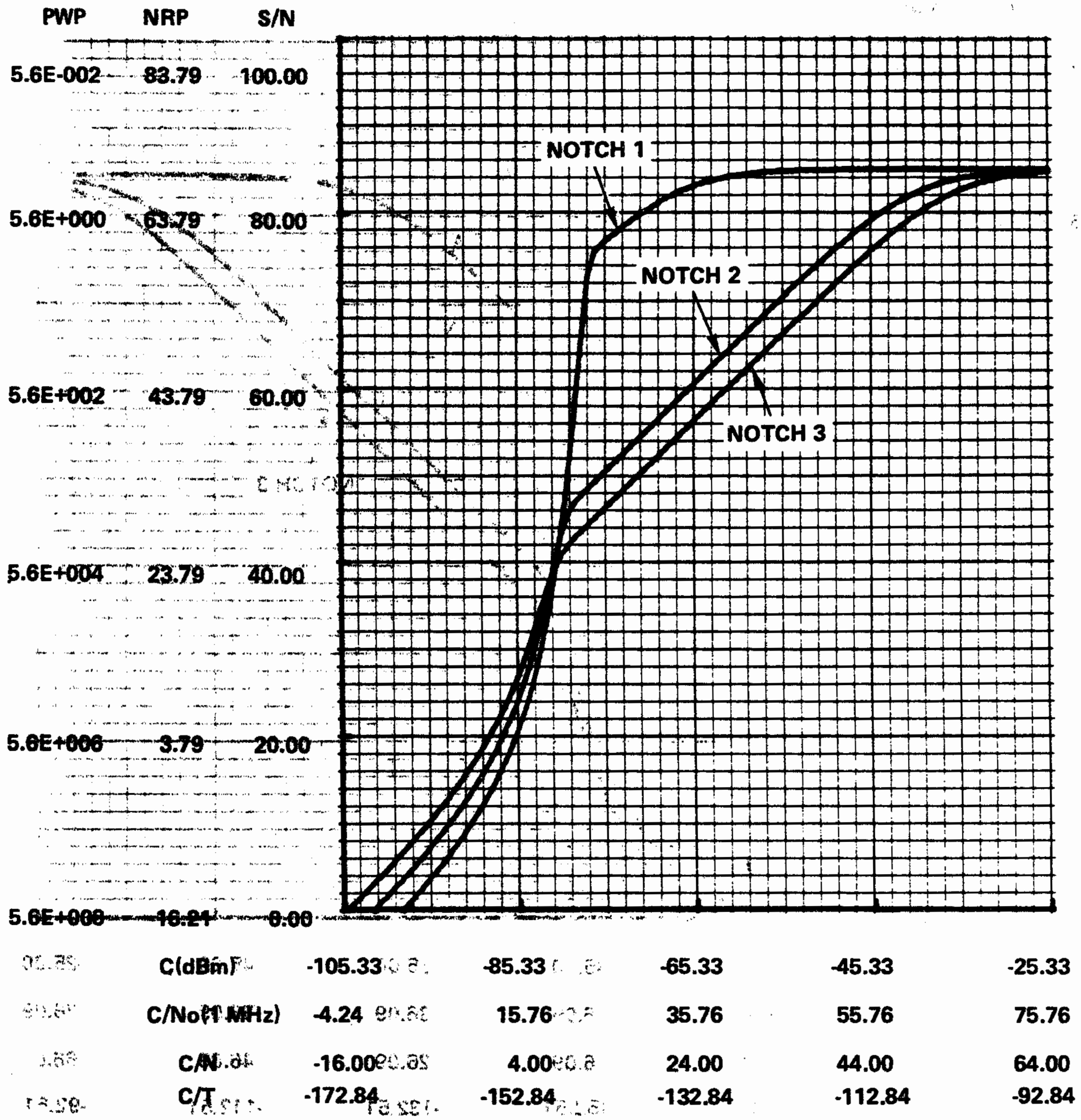
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	252
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1002000
OCCUPIED BW	68529487.76
LINEARITY BW	16769641.45
T.T. DEVIATION (RMS)	2600000
P. E.	YES
P. E.	NO

Figure 19. S/N vs. C/N for 252 Channel Loading



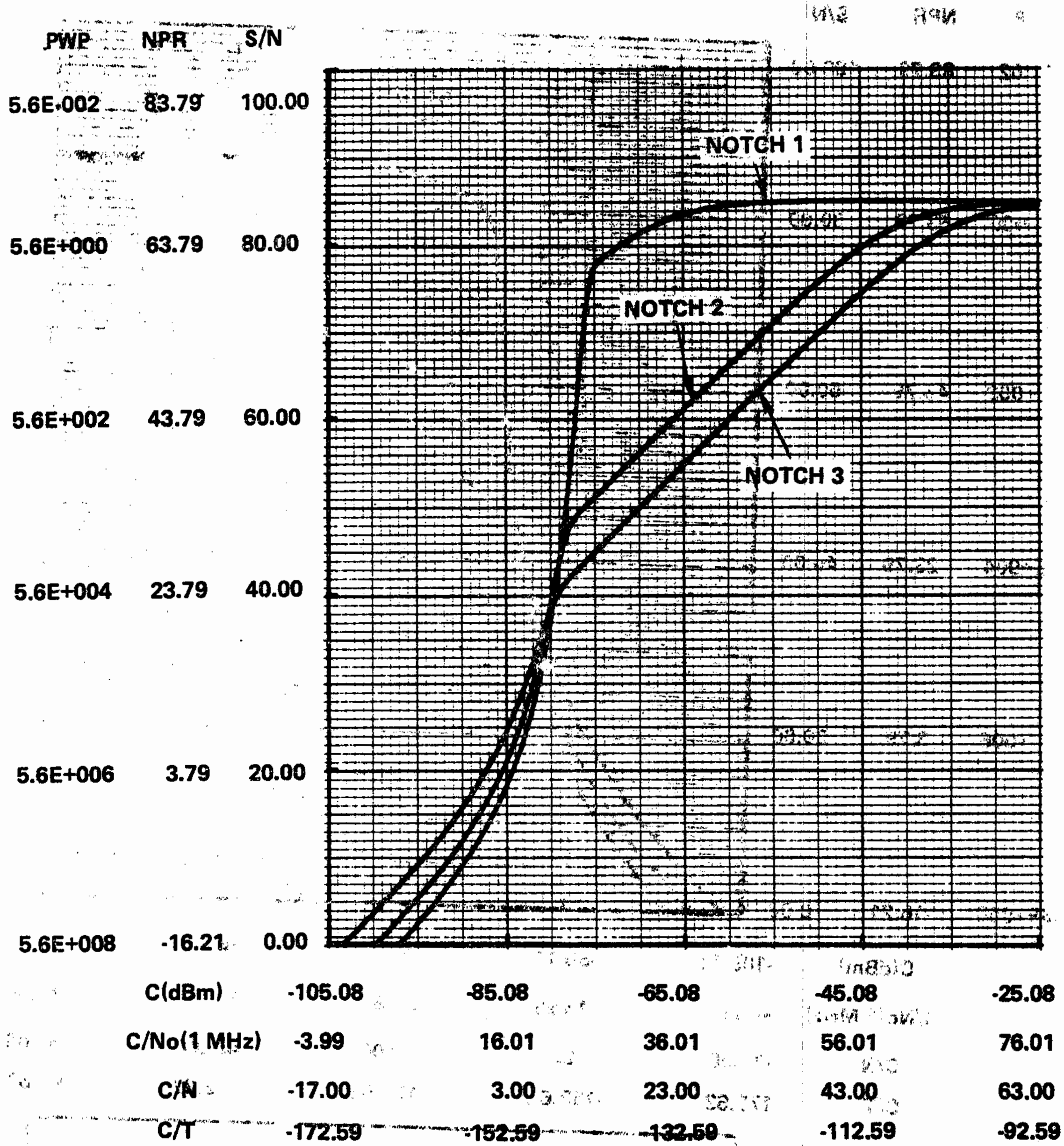
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	312
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1248000
OCCUPIED BW	18585478.65
LINEARITY BW	4592683.33
T. T. DEVIATION (RMS)	320000
P. E.	YES
P. S.	NO

Figure 22. S/N vs. C/N for 312 Channel Loading



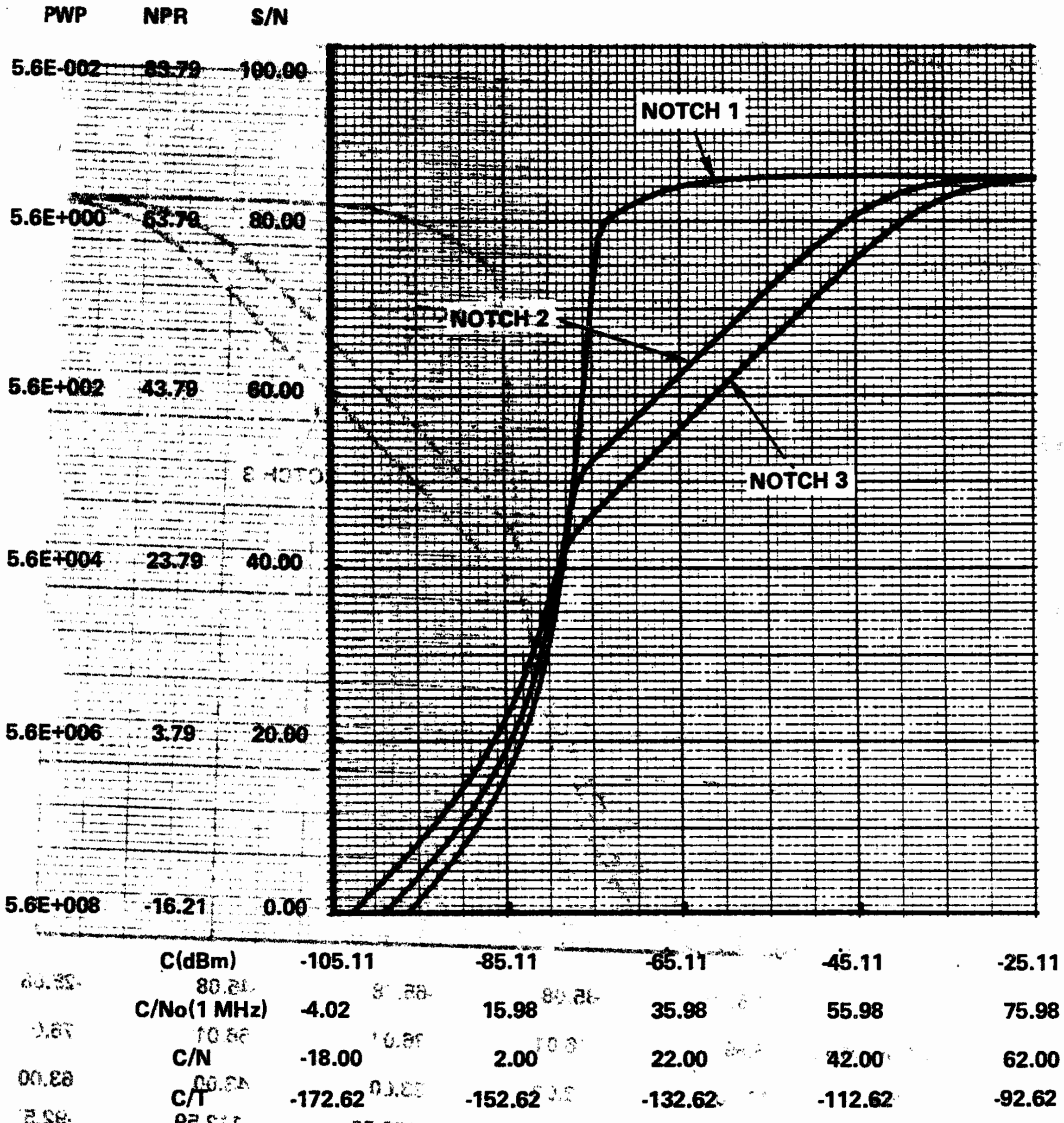
N. F. (dB)	13
BW (I. F.)	1.5E+7
CHANNEL LOADING	312
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1240000
OCCUPIED BW	16237652.95
LINEARITY	6012445.83
T. T. DEVIATION (RMS)	546000
P. E.	YES
P. S.	NO

Figure 23. S/N vs. C/N for 312 Channel Loading



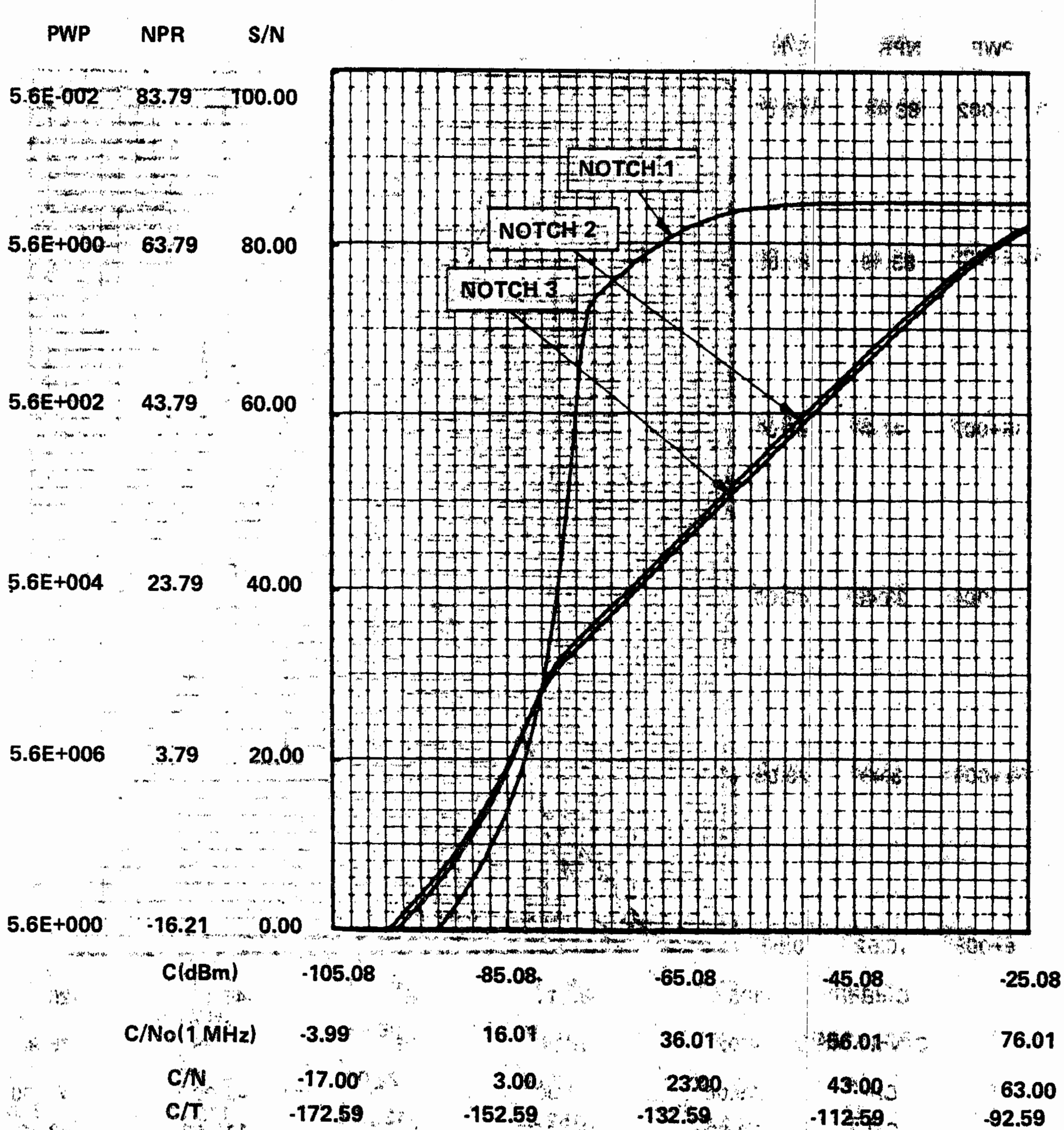
N. F. (dB)	13
BW (I. F.)	2.0E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	21694504.46
LINEARITY BW	8119975.91
T. T. Deviation (RMS)	616000
P. E.	YES
P. S.	NO

Figure 26. S/N vs. C/N for 432 Channel Loading



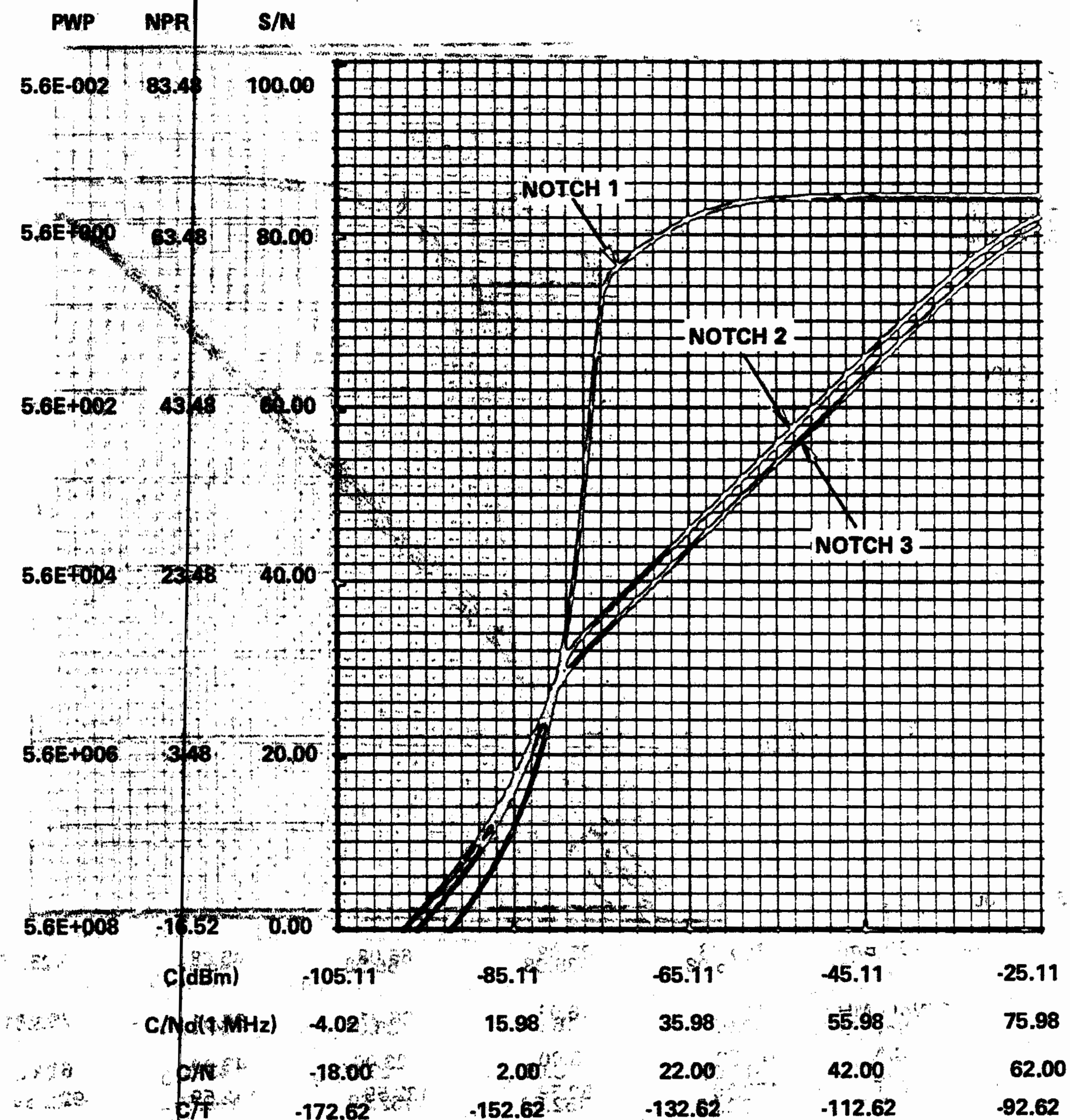
N. F. (dB)	13
BW (I. F.)	2.5E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	25019952.20
LINEARITY BW	8955290.98
T. T. DEVIATION (RMS)	729000
P. E.	YES
P. S.	NO

Figure 27. S/N vs. C/N for 432 Channel Loading



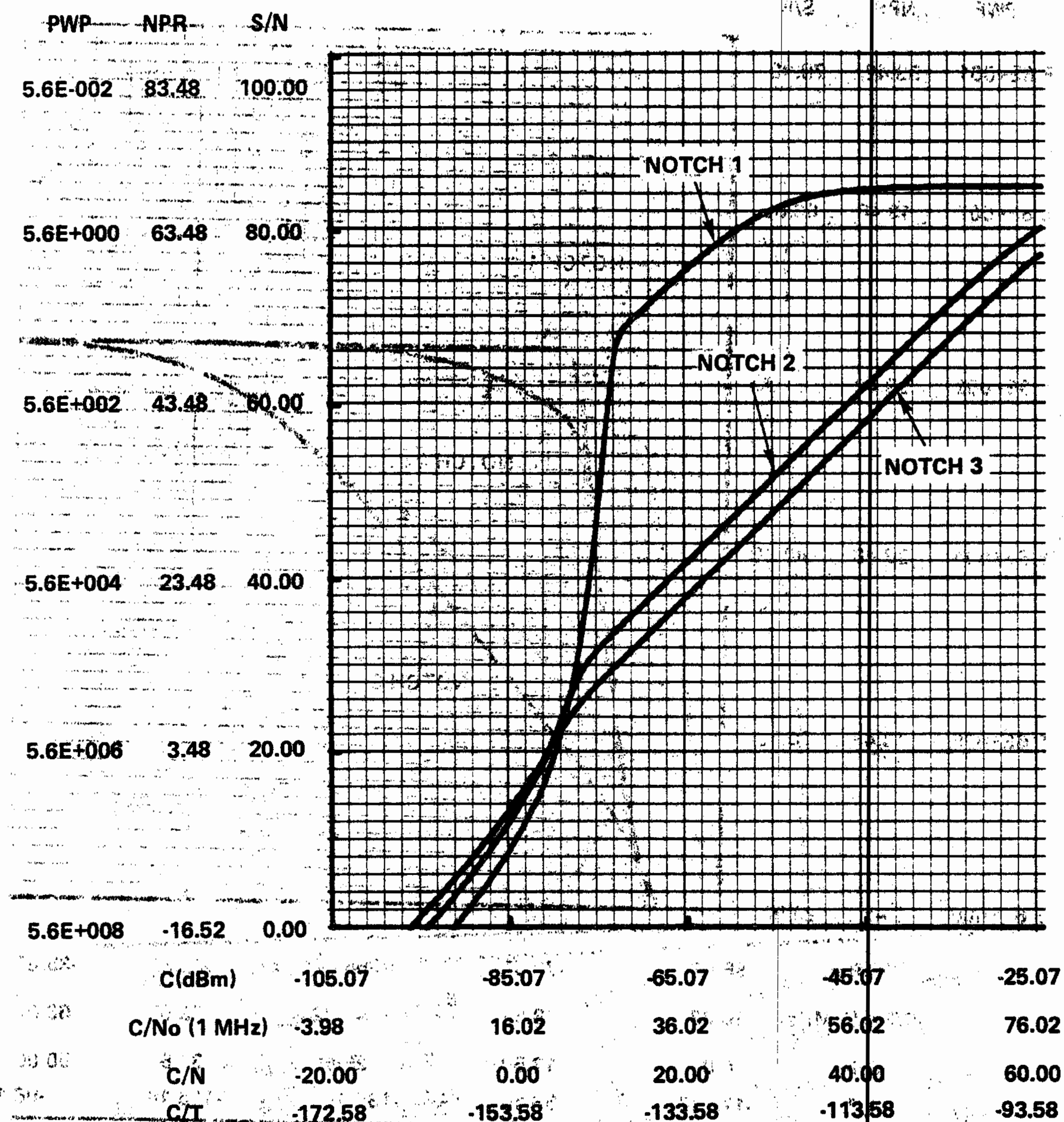
N. F. (dB)	13
BW (I. F.)	2.0 E+7
CHANNEL LOADING	792
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3150000
OCCUPIED BW	20703826.52
LINEARITY BW	10081619.57
T. T. DEVIATION (RMS)	356000
P. E.	YES
P. S.	NO

Figure 30: S/N vs. C/N for 792 Channel Loading



N. F. (dB)	13
BW (I. F.)	2.5E+7
CHANNEL LOADING	972
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3886000
OCCUPIED BW	26676275.72
LINEARITY BW	13123783.75
T. T. DEVIATION (RMS)	410000
P. E.	YES
P. S.	NO

Figure 31. S/N vs. C/N for 972 Channel Loading



N. F. (dB)	13
BW (I. F.)	4.0E+7
CHANNEL LOADING	2892
NOTCH 1	70000
NOTCH 2	5340000
NOTCH 3	1.17E+7
OCCUPIED BW	72606005.23
LINEARITY BW	37312729.99
T. T. DEVIATIO N (RMS)	619000
P. E.	YES
P. S.	NO

Figure 36. S/N vs. C/N for 2892 Channel Loading

Video Uplink Earth Station Performance Verification

L. S. Hermann

Scope

The following sections describe seven tests which may be performed on a video uplink station to verify the performance of the equipment and ensure providing the high-quality signals required in video systems.

Output Power

1. Connect the test equipment as shown in Figure 1.

2. Check exciter frequency for mid-HPA channel.

3. Set Exciter for -10 dBm output.

4. Adjust HPA input attenuator for maximum output.

5. HPA output power may be calculated from output power meter reading by adding coupler value plus any attenuation between coupler and power meter.

Specification >34.4 dBW.

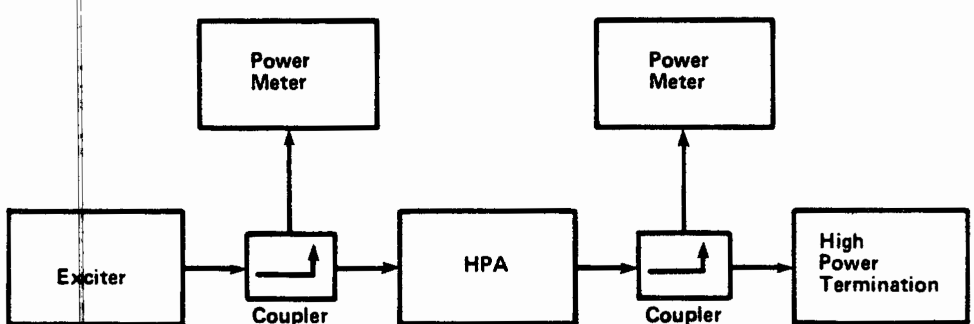


Figure 1. Output Power — HPA Gain

EIRP

EIRP may be calculated by measuring the output power at the OMT flange and adding the known antenna gain.

If the waveguide and switching system losses are known, then:

$$EIRP = G_{Antenna} + P_{HPA} - L_{WG} - L_{SW}$$

HPA Gain

1. Connect the test equipment as shown in Figure 1.
2. Check exciter frequency for mid-HPA channel.
3. Set HPA input attenuator for nominal output power.
4. HPA gain is HPA output power (dBW) - exciter output power (dBW)

Specification >70 dB.

IF-RF Gain and Gain Flatness

1. Connect the equipment as shown in Figure 2.
2. Adjust the sweep oscillator for 70 MHz ±20 MHz.
3. Check the exciter output for correct HPA channel output.
4. Calibrate scope output level and gain using attenuation and known input.
5. Measure IF-RF gain and gain flatness directly from scope display (Figure 3).

Specification: Total of 0.5 dB in center third of BW.
Total of 1 dB over remaining two-thirds of BW.

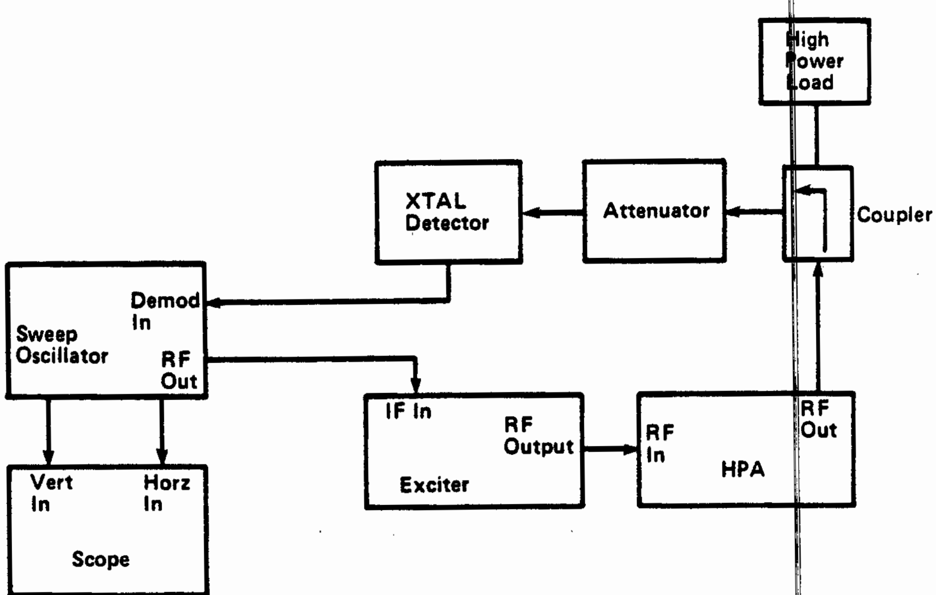


Figure 2. IF—RF Gain — Gain Flatness

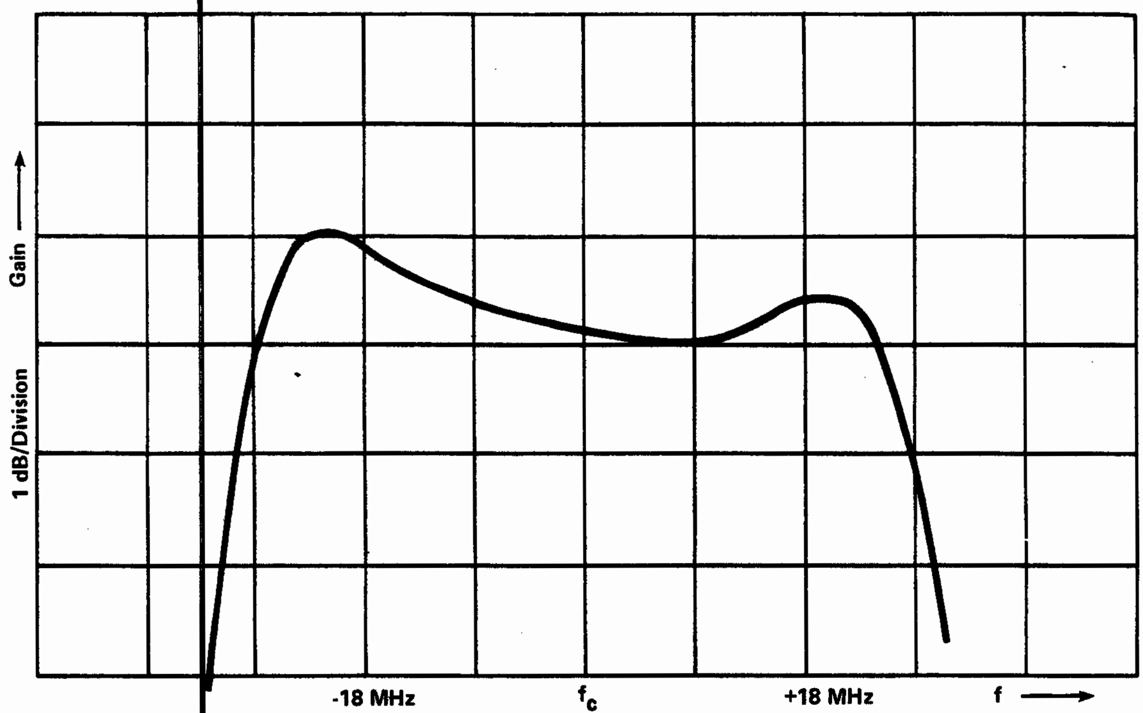


Figure 3. Typical HPA Channel

Group Delay

1. Connect test equipment as shown in Figure 4.
2. Check the exciter frequency for correct HPA channel.
3. Calibrate the MLA for 1 ns/cm group delay reading.
4. Measure the group delay across the channel (Figure 5).

Specification: Linear - 0.25 ns/MHz
Parabolic - 0.05 ns/MHz
Ripple - 2 ns/peak-to-peak

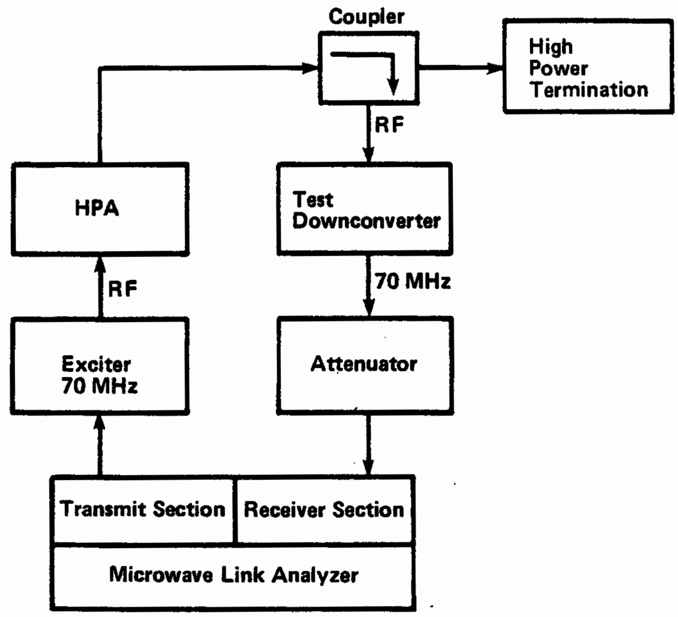


Figure 4. Group Delay

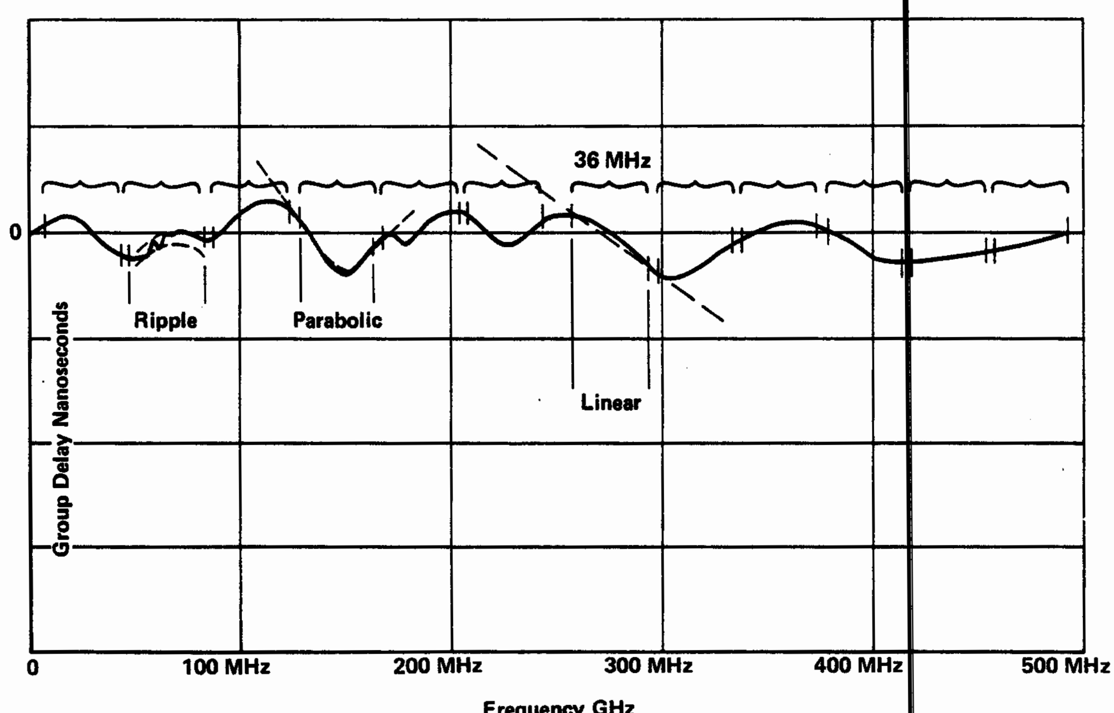


Figure 5. Typical Group Delay Measurement

Intermod

1. Connect the test equipment as shown in Figure 6.

2. The Exciter is driven by two in-band signals spaced such that the third-order intermodulation products are shown on the spectrum analyzer.

3. Measure the intermodulation performance per the graph shown in Figure 7.

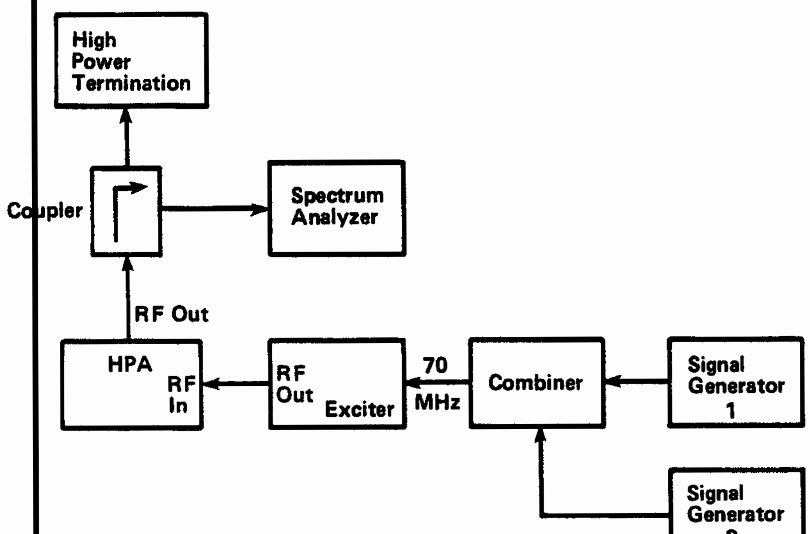


Figure 6. Intermod

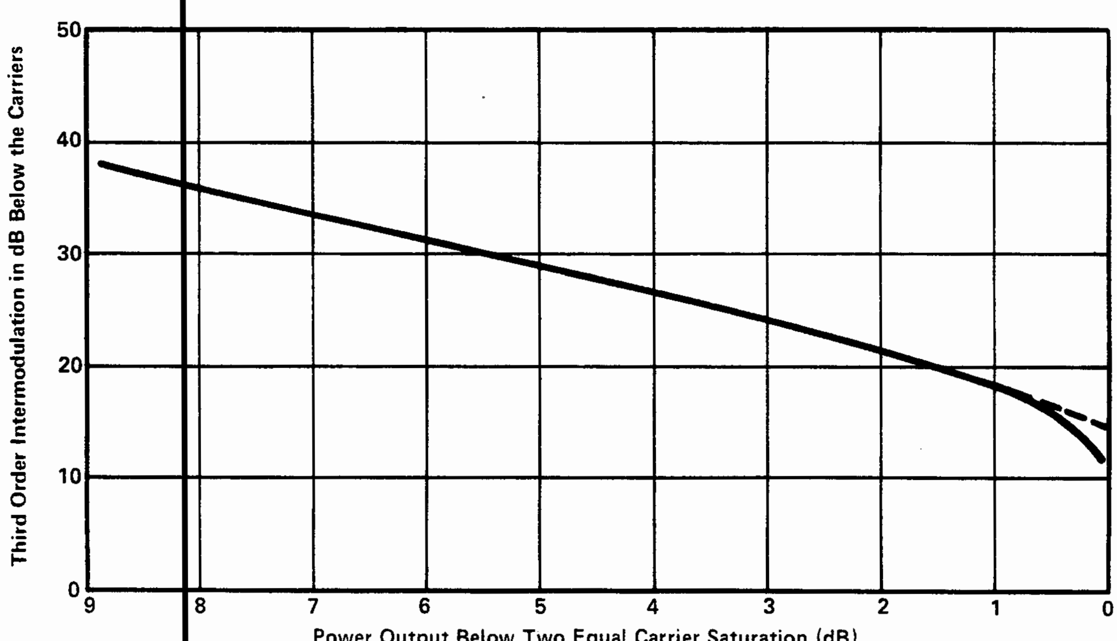


Figure 7. Third Order Intermodulation Distortion Under Two Equal Carrier Conditions

Video Tests

1. Connect test equipment as shown in Figure 8.

2. With all equipment set up for 1V peak-to-peak levels, measure baseband flatness using baseband signal generator and rms voltmeter or calibrated display.

3. Using video signal generator and waveform monitor with vector scope, measure all video parameters.

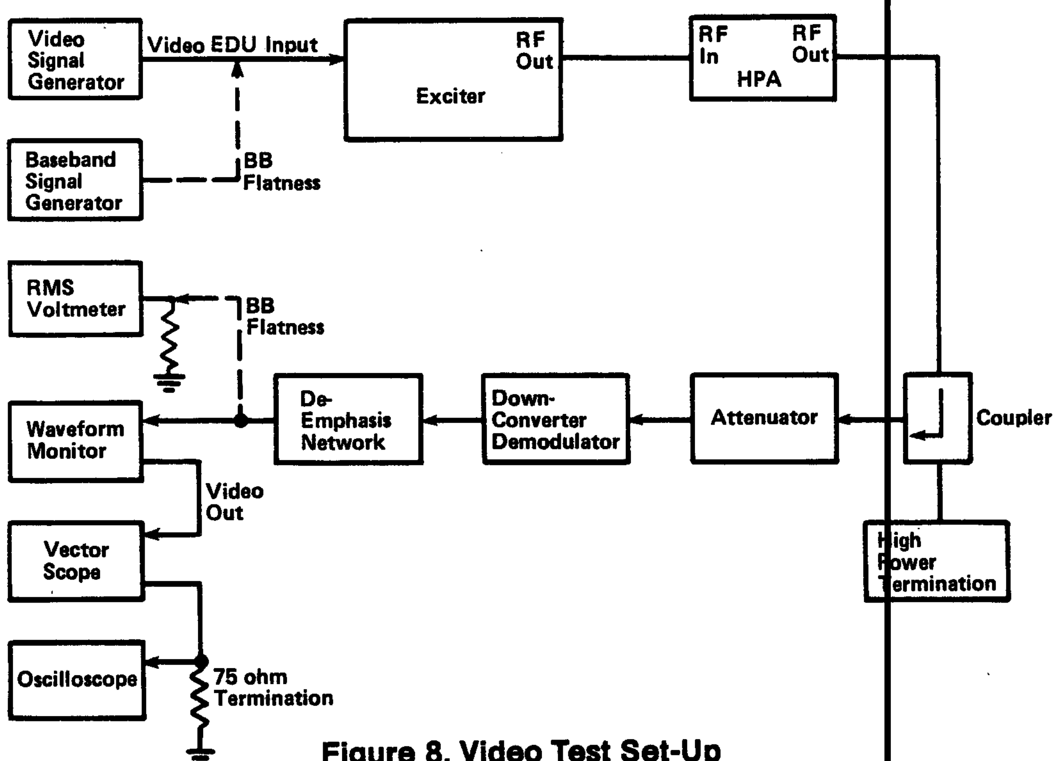


Figure 8. Video Test Set-Up

Typical measurements are:

- Differential Phase <±0.5°
- Differential Gain <±2%
- Field Time Distortion <1.0%
- Short Time Distortion <0.5%
- Chrominance to Luminance Delay <30 ns
- Non-Linear Distortion <3%

Operating a Transmit/Receive Earth Station

L. S. Hermann

Introduction

The major aspects of operating a satellite transmit/receive earth station will be addressed in this paper. Costs and revenues will be covered in a general sense, with particular emphasis on personnel as the major cost component. Transponder acquisition will then be discussed. Finally, equipment operation, troubleshooting, and maintenance of the station will be covered.

Transmit/receive operations fall into three categories: full-time operators, intermittent transmitters, and transportables. Aspects peculiar to each category will be addressed as necessary.

Costs

The costs of operating an earth station may be listed in the following order of relative magnitude:

- Personnel
- Depreciation of fixed assets
- Utilities
- Maintenance hardware
- Insurance
- Taxes

The first four of these will be discussed in detail.

In addition, the cost of the satellite transponder may be very large. This cost may be incurred as part of the transmission package (e.g., for a dedicated user), or may be billed separately to the end-user (e.g., with a resale operator). Transponder acquisition and costs are considered under a separate section of this paper.

Personnel

The personnel required to operate an earth station fall in these categories:

- Engineer
- Technician(s)

- Sales
- Accounting
- Legal
- Custodial

A full-time operator usually employs a full-time engineer and several technicians depending on the number of shifts and transmitters. An intermittent operator would use the part-time services of an engineer and full-time services of one or more technicians. A transportable operator requires a combination technician/engineer and usually a driver who may also be part technician.

These technical personnel should be first and foremost skilled in the media in which they operate, and secondly in RF microwave. The operators for a video station should be skilled in video technology; digital operators should be skilled in computer operation; SCPC/message operators should be skilled in telephony.

A technician familiar with the operating media may be trained in the RF operation of the station. He should soon be able to handle normal operation as well as minor troubleshooting and repair.

A person with engineering qualifications is needed to manage the technical aspects of an earth station. While being most familiar with the operating media, he would also have experience with microwave RF (e.g., terrestrial microwave). He whould then have little trouble learning satellite RF operation. This person should be capable of handling all major emergencies. As the engineer's time may only be partially used in operating the earth station, he may also manage other technical areas (e.g., a TV station or computer center), or manage the business side of the earth station.

A transportable earth station operator should be equally familiar with his media as with satellite RF operation. His customers will be specialists in the media they are broadcasting, and the operator must deal effectively with them. Yet, the operator must be self-sufficient and capable of handling any RF emergency. He should have extensive field training in satellite RF, preferably having been involved in a previous earth station installation and proof-of-performance.

The remaining personnel will be needed in various degrees depending on the nature of the service provided. An accounting staff is required to bill customers for usage. Legal help would be required for new FCC filings (e.g., for a change in tariff, or for an increase in power from an additional uplink or antenna).

Depreciation of Fixed Assets

Building space and support equipment (e.g. air conditioners) should be included along with the cost of the earth station in determining fixed assets. Otherwise, standard accounting procedures apply.

Utilities

Electricity will be the largest utility expense. The electrical load will depend on the configuration of the earth station as well as its use. Power requirements are listed in the previous paper, "Planning a Transmit/Receive Earth Station." Power required during various HPA usage cycles is discussed later in this paper under Equipment Operation.

Telephone charges may be substantial for an intermittent or transportable operator who must check with the satellite Technical Operations Center (TOC).

Water and other facility charges must be considered, especially in the case of a remote site.

Maintenance Hardware

Spare parts may be either stocked or purchased as needed. This measure of insurance is determined by the level of reliability required and the level of redundancy installed in the system.

A good rule-of-thumb is that maintenance hardware will cost three to five percent of the system cost per year. The power amplifier will require the greatest amount of maintenance, as it operates at high voltage and high temperatures. In the case of a Klystron power amplifier, the Klystron tube will need to be rebuilt every two to three years. A third tube is usually purchased when the first one fails to allow the operator to rotate the tubes for refurbishment.

Revenues

The most important decision in determining price levels is whether or not to apply for common carrier status. In order to charge specifically for the satellite uplink service, a common carrier license is needed. Tarriffed rates must then be filed with the FCC. Most intermittent or resale operators are licensed as common carriers.

If you are providing an end-to-end service of which the satellite uplink is only a part, you need not apply for common carrier status. Dedicated uplink operators would fall into this category.

Another determinant of the prices charged is the amount of backup protection offered the customer. As with satellite rates, uplink rates may be billed as:

- Fully protected (with a hot backup)
- Unprotected (single uplink)
- Unprotected and preemptible (if another uplink fails)

The type of service has an effect on HPA operating costs, as discussed later in this paper under Equipment Operation.

Transponder Acquisition

Full-time operators will either purchase or lease a full or partial transponder. Full transponders may sell for ten million dollars and up. Lease rates for a full transponder would depend on the kind of service and the length of the lease. Typical tariffs are listed in Table 1.

Table 1. Typical Full Transponder Lease Rates

Type of Service	Monthly Lease
Protected	$150,000 to $230,000
Unprotected	$83,500 to $149,500
Unprotected and Interruptible	$66,000 to $115,000

Source: Western Union Telegraph Co.; Tariff FCC No. 261; Satellite Transmission Service, page 33, effective June 10, 1982.

Full transponder service is also available in segments of as short as a half hour through numerous resellers. Charges for this transponder time vary according to time of day and volume. Typical charges run $300 to $700 per hour.

There are a number of companies which own tranponders and which resell time on various satellites. A representative sample of these resellers are shown in Table 2.

Table 2. Transponder Space Brokers

Argo Communications
100 Long Island Place, NW
Atlanta, GA 30328
Bill Papa

ASN, Inc.
310 14th Ave. S.
St. Petersburg, FL 33701

Bonneville Satellite Corp.
179 Social Hall Ave.
Salt Lake City, UT 84111
(801) 237-2450
Bruce Hall

Compact Video, Inc./
NETCOM Enterprises
Dept. of Satellite Services
2901 W. Alameda Ave.
Burbank, CA 91505
(213) 841-8887
Priscilla Davis

Equatorial Communications Service
300 Ferguson Dr.
Mountain View, CA 94043
Ed Parker

General Communications, Inc.
2550 Danali St.
Suite 505
Anchorage, AK 99503
(907) 338-6888
Bill Walp

Metropolitan Communications
Network Co.
P.O. Box 915
6861 Elm St.
McLean, VA 22101
(703) 734-2724
Dean Popps

Southern Satellite Systems
P.O. Box 45684
Tulsa, OK 74145
(918) 481-3275
Ruben Gant

United Video, Inc.
3801 S. Sheridan Rd.
Tulsa, OK 74145
(800) 331-4806
Tom McKeenzy

Western Telecommunications, Inc.
Box 22595
Wellshire Station
Denver, CO 80222
(303) 771-8200

Wold Communications, Inc.
10880 Wilshire Blvd.
Los Angeles, CA 90024
(213) 474-3500
Robert Wold

Partial transponder time is usually required for digital or voice transmission. This service is leased on the basis of number of carriers, bandwidth, and power required, as shown in Table 3.

These parameters are dependent on the system configuration and are determined by a system link analysis. A typical duplex voice FM-SCPC channel may cost $2,500 per month.

Table 3. Typical Partial Transponder Lease Rates

Basic	Monthly Lease (Sum)
Per Carrier (channel)	$390
Per Watt of Power	$131
Per kHz of Bandwidth	$ 3

Source: Western Union Telegraph Co.; Tariff FCC No. 261; Satellite Transmission Service; page 38; effective May 11, 1982.

Equipment Operation

Initiating a Satellite Transmission

Having secured transponder space, standard procedure for any operator when first transmitting is as follows:

a. Position the antenna on the satellite (if different than the previous transmission).

b. Using the receivers, check the null on adjacent transponders to assure the feed polarization angle is correct.

c. Tune exciter(s), HPA(s), and receiver(s) to the desired transponder frequency. On 6- or 12-channel klystron power amplifiers, this may require retuning the klystron tube.

d. Run power amplifier into a dummy load to check power levels.

e. Telephone the Satellite Technical Operations Center (TOC). This may be either the transponder owner or the satellite TT&C earth station. Ask for the Communications Technician (ComTech).

f. Transmit a CW carrier at a low power level. The TOC will check cross polarization and may ask you to adjust your feed polarization.

g. The TOC will give permission for full transmission.

h. Increase power to saturation.

i. Add deviation/modulation.

j. The TOC will check for adjacent transponder interference, then hang up.

k. Add the source.

Full-time operators who are continually going up and down on the same transponder need not repeat all these steps. Yet any operator who changes frequency and/or satellite should follow this procedure to avoid double-illumination of the wrong transponder.

When ceasing a transmission, the operator of an intermittent station should inform the TOC that he has done so for record and billing purposes.

It should also be noted that transponder costs will be incurred for both pre-broadcast testing and for other maintenance testing.

Time Required to Initiate a Transmission

Approximate times required in the various stages listed above are shown in Table 4.

Table 4. Time Required to Initiate a Transmission

Procedure	Approximate Time (Minutes)
Change Satellites	1 to 40 (depending on antenna motor speeds)
Retune HPA Frequency (klystron PAs only)	
Preset frequency	1 to 5
Requiring tuning	10 to 60 (depending on test equipment setup)
TOC Verification	5 to 20

Power Amplifier Operating Modes

Of the three kinds of power amplifiers currently available, the solid-state and TWT amplifiers with large bandwidths and long life spans are the most simple to operate. In addition, they are generally used in dedicated systems where they are seldom turned off.

Conversely, the Klystron High-Power Amplifier (HPA or KPA), is usually much larger, requires retuning to change transponders, and may experience intermittent operation. These conditions pose an interesting set of operating parameters.

Table 5 outlines the major characteristics of four HPA operating states. There are tradeoffs between the speed at which an HPA can be turned on, the amount of energy consumed (and heat given off), and the life expectancy of the tube.

Table 5. Major HPA Operating States

State	Life Expectancy of Klystron Tube	Power Required (kVA)	Heat Released	Time to Power Up (Minutes)
Full Filament/ High-Voltage	100%	12	100%	0
Cold	90%*	0	0%	7 to 10
Filament Foldback/ No High Voltage (newer models)	Intermittent use: 95% Continuous use: 2 hrs	1	3%	2 to 3
Full Filament/ No High Voltage (older models)	Intermittent use: 95% Continuous use: 30 min	1	3%	2 to 3

*Cold HPAs should be run with full load four times per year, at minimum, to prevent gas buildup inside the klystron tube. During long cold periods, the tube should also be stored externally on a Degassing Test Bed.

The normal operating mode of both the active and backup HPA is in the full-filament/high-voltage state. This provides the quickest transmission with the longest life. When an intermittent station is not in use, the cold state is normally used (provided no transmission is needed in less than ten minutes). However, if a station is used consistently more than a few hours each day, it becomes economic to leave all HPAs fully powered; this lengthens the tube life even though it uses more power.

The remaining two states of filament with no high voltage are used only when changing frequencies or performing internal tests. These states are used as infrequently as possible due to the severe impact on tube life.

Transportable Operation

Operation of a transportable earth station differs from that of a fixed station mostly in the area of setup. At least half a day should be allowed for the operator to position his vehicle, link in to his source, find the satellite, and fully test his system.

An added consideration with transportable earth stations is the frequency coordination required prior to each transmission. The frequency coordination companies provide a package for this service which allows multiple site clearings at a lower cost.

Troubleshooting

All earth station electronics include alarms in one or more forms, usually lights or LEDs. These should be monitored continuously during operation by the technician. In a system with automatic redundancy, a failure will cause the transmission to shift to the backup link.

The technician on duty should be capable of diagnosing the failure down to the board level, correcting the failure if possible, and bringing the system back on-line. If further troubleshooting is required, the engineer should be called in. With proper training, the engineer should be able to troubleshoot most failures down to the component level.

Preventive Maintenance

The amount of time and money invested in a transmit/receive earth station justifies the establishment of a preventive maintenance schedule. Weekly or monthly preventive maintenance inspections, depending on earth station size, should alert the technical or operations staff to any impending component degradation or failures.

Routine inspections will point out the blower about to experience bearing failure, air filters that are clogged, antenna motors that are noisy, and meter readings approaching out-of-tolerance conditions. Once discovered, impending problems can be corrected, and the chance for failure is reduced.

All operational manuals provided by Scientific-Atlanta specify the level of maintenance on each item. Motorized antennas require periodic lubrication and tightening of bolts; all should be checked to ensure no binding areas nor rust spots are obvious or that the feed window is not cracked and leaking. The transmit waveguide should be checked periodically to ensure that no air leaks have occured and no sections have been kinked or crimped. Positive pressure using dry air should be maintained in both the transmit and receive waveguide at all times to reduce line losses and prevent possible arcing in the transmit waveguide.

The HPAs are the most sensitive part of the system due to the high-power levels at which they operate.

The HPAs should be checked daily if they are operated continuously. The air circulation in the HPA cabinet is critical for operation. Loss of a blower will alarm the system, and the HPA will shut down until the fault is corrected. The design of most HPAs is such that a potentially self-destructive fault results in shutdown.

Maintenance cannot be overemphasized in owning and operating a transmit/-receive earth station.

Glossary

A/D or ADC — Analog to digital. Analog-to-digital converter.

AFC — Automatic frequency control.

AGC — Automatic gain control.

AISC — American Institute of Steel Construction.

ALC — Automatic level control.

ANSI — American National Standards Institute.

Aperture — (directive antenna) 1. The clear diameter of the parabolic reflector of a microwave antenna. 2. A surface, near or on a directive antenna, through which most of the radiated energy passes and to make assumptions regarding the field values for the purpose of computing fields at external points.

Aperture Efficiency — The ratio of the directivity of an antenna to that which would be obtained if the aperture illumination were uniform. (See Aperture, Directivity.)

APL — 1. (television) Average picture level. Average value of the picture signal, integrated over one frame. The picture signal does not include blanking signals. 2. (audio) Average program level.

ASCII — American Standard Code for Information Interchange.

ASMS — Automatic surface measuring system.

ASTM — American Society for Testing and Materials.

Asynchronous Communications — Transmission and reception of data in a mode where there is no continuous synchronization between the terminals.

AZ — Azimuth angle (of an object). The arc of the horizon, between true north and the vertical plane passing through the object, usually measured toward east.

Baud — A unit of signal speed equal to the number of signal elements or symbols transferred in each second; a unit of transmission rate of digital signals. It is the reciprocal of the length in seconds of the shortest element of the digital code.

BER — Bit error rate. The probability that a bit will be received in error, numerically equal to the number of bit errors divided by the number of bits transmitted.

Bit — Binary digit.

Bits per Second — The number of bits of data transferred in each second.

Blanking — The process of cutting off the electron beam in a camera or picture tube during the retrace period.

Blanking Level — 1. That level of a composite picture signal which separates the range containing picture information from the range containing synchronizing information. 2. The level of the front and back porches of the composite video signal.

BPSK — Bi-phase shift keying. Carrier modulation technique in which the phase of the transmitted carrier is shifted ±180 degrees with respect to a reference.

Byte — A group of adjacent binary digits operated on as a unit and usually shorter than a computer word. By currently accepted definitions, a byte is 8 bits.

CATV — Cable television or community antenna television.

CCIR — International Ratio Consultative Committee.

Chrominance — That property of light which produces a sensation of color in the human eye apart from any variation in luminance that may be present.

C/I — Carrier-to-interference ratio. The power in the desired carrier divided by the power in the interfering signal or signals.

C/N — Carrier-to-noise ratio. Ratio of carrier power to noise power in a defined frequency band.

C/N_0 — Carrier-to-noise-power-density ratio. (See C/N, Noise Power Density.)

Color-Bar Signal — A test signal, typically containing six basic colors: yellow, cyan, green, magenta, red, and blue, which is used to check certain functions of color TV systems.

Color Burst — In NTSC color systems, this normally refers to a burst of approximately 10 cycles of 3.579545-MHz subcarrier frequency on the back porch of the composite video signal. It serves as a color synchronizing signal to establish a frequency and phase reference for a chrominance signal.

Color Phase — The phase, with respect to the chrominance-carrier reference, of the component of the carrier-chrominance signal which corresponds to one of the chrominance primaries.

Color Signal — The chrominance and luminance components of the NTSC color television signal.

Color Subcarrier — In color systems, this is the carrier signal whose modulation sidebands are added to the monochrome signals to convey color information; in NTSC, it is a 3.579545-MHz sinewave.

Composite Baseband — For color video satellite links, this consists of the Composite Video, Energy Dispersal, and Audio Subcarrier signals.

Composite Video — For color, this consists of blanking, field, and line synchronizing signals, color burst, chrominance, and luminance picture information. These are all combined to form the complete color video signal.

Contrast — The ratio between the maximum and minimum brightness values in a picture.

CONUS — Contiguous (48) United States.

Crosstalk — Undesired energy appearing in one signal path as a result of coupling from other signal paths.

CVSD — Continuous variable slope delta modulation.

D/A or DAC — Digital to analog. Digital-to-analog converter.

Data — 1. Any representations, such as characters or analog quantities, to which meaning might be assigned. 2. Information in a form which can be transmitted by a communications link or processed by computers or machines; e.g., analog data or digital data.

DEC — 1. Declination angle. Angle between the celestial equator and an extra-terrestrial object, measured in a plane which is normal to the celestial equator. 2. Digital Equipment Corporation.

Decibel (dB) — A number denoting the ratio of the two amounts of power, being ten times the logarithm to the base 10 of this ratio. The abbreviation dB is commonly used for the term decibel. With P_1 and P_2 designating two amounts of power and n the number of decibels denoting their ratio,

$$n = 10 \log_{10}(P_1/P_2) \text{ decibels.}$$

When the conditions are such that ratios of currents or ratios of voltages (or analogous quantities) are the square roots of the corresponding power ratios, the number of decibels by which the corresponding powers differ is expressed by the following equations:

$$n = 20 \log_{10}(I_1/I_2) \text{ decibel}$$

$$n = 20 \log_{10}(V_1/V_2) \text{ decibel}$$

where I_1/I_2 and V_1/V_2 are the given current and voltage ratios, respectively. See Appendix A of paper 1-2.

Deemphasis — The restoration of a preemphasized frequency spectrum to its original form. The introduction of loss at the higher frequencies in the receiver of a communication system to compensate for earlier preemphasis in the transmitter. (See Preemphasis.)

Delay Distortion — That form of distortion which occurs when the envelope delay of a circuit or system is not constant over the frequency range required for transmission.

Differential Gain — A form of non-linear distortion. As applied to NTSC or PAL color television, it is the change in the amplitude of a small sinusoid at the color subcarrier frequency caused by variation in the accompanying luminance level. It is expressed as the difference in amplitude between the maximum amplitude and the minimum amplitude of the sinusoid when the maximum amplitude is normalized to 100 percent. The luminance signal shall explore the range from black to white.

Differential Phase — A form of non-linear distortion. As applied to NTSC or PAL color television, it is the change in the phase of a small sinusoid at the color subcarrier frequency caused by variation in the accompanying luminance level. It is expressed as the greatest difference in phase observed for luminance levels between black and white.

Digital Data — Information coded in digital electronic form from analysis or computation. Any data which is expressed in digits. Binary digits are usually implied.

Directional Coupler — 1. A transmission coupling device for separately sampling the forward (incident) and/or the backward (reflected) wave in a transmission line. 2. A device which can be inserted in a 75 ohm CATV cable to tap off the signal for local use.

Directivity — (antenna) The directivity of an antenna in a specified direction is 4π times the ratio of the power radiated per unit solid angle in that direction to the total power radiated by the antenna. This term differs from power gain in that it does not include antenna dissipation losses.

DMUX — Demultiplexer. Demultiplex. Equipment for or process of separating two or more messages which are interleaved in frequency or time for transmission on a single channel. (See FDM, TDM.)

Downlink — The circuit between a satellite and a receiving earth station, including the satellite transmitter and antenna, the satellite-earth propagation path, and the earth station antenna and receiver.

E_b/N_o — Energy-per-bit-to-noise-power-density ratio. Signal-to-noise ratio normalized with respect to bit rate.

$$E_b/N_o = \frac{P_s/R}{N_o}$$

where:

P_s = signal power

R = bit rate

No = noise power density

Echo — 1. A signal which has been reflected at one or more points during transmission. 2. A wave that has been reflected or otherwise returned with sufficient magnitude and delay to be perceived in some manner as a wave distinct from that directly transmitted. 3. An attenuated reflection of a talker's voice, separated from the primary wave by reflection at an electrical discontinuity in the circuit, and returned to him over the circuit on which he is talking and listening.

Effective Area — 1. In a given direction, the ratio of the power P_r available at the terminals of a receiving antenna to the power density S per unit area of a plane wave incident on the antenna from that direction, the incident wave being polarization matched to the receiving antenna:

$$A_E = \frac{P_r}{S},$$

where A_E is the effective area. (See Partial Effective Area, Realized Effective Area.)

EIA — Electronic Industries Association.

EIRP — Effective isotropic radiated power. Product of transmitted power times transmitting-antenna gain, usually expressed in dBW.

EL — Elevation angle. (of an object) The vertical angle from the local horizon to the object.

EMI — Electromagnetic interference.

End-to-End — Reference to the totality of the circuitry, devices and transmission media from input terminals to the output terminals of a communications link.

Energy Dispersal Waveform — For video satellite links, a triangular waveform with apexes located at the vertical sync intervals which is modulated onto the RF carrier to avoid concentration of energy at one frequency.

Energy Dispersal Scrambling — A method of making the output power spectrum of a PSK modulator independent of the data stream by scrambling the data at the transmit end and descrambling it at the receiver. Energy dispersal scrambling is required by the FCC to avoid interference with other signals.

EPROM — Erasable programmable read-only memory.

Equatorial (Polar) Coordinate System — A spherical coordinate system whose polar axis is aligned with the earth's poles.

ES — Earth station.

ET — Earth terminal. Extra-terrestrial.

FDM — Frequency-Division Multiplex — A method of multiplexing or combining more than one (usually many) voice or data channels for transmission on a single RF carrier. The channels are separated in frequency and are carried on subcarriers.

FDMA — Frequency-division multiple access (of a satellite transponder by different earth stations).

FEC — Forward error correction. An encoding technique for improving the BER in a data system. In a typical rate 3/4 FEC system, the BER may be improved by a factor of ten thousand.

Feed — That portion of an antenna which is coupled to the terminals and which functions to produce the aperture illumination on transmission or to couple energy concentrated at the focal point to the antenna terminals on receiving.

Field — One half of a complete picture (or frame) interval, containing all of the odd, or all of the even, lines of the picture. One of the two (or more) equal parts into which a frame is divided in interlaced scanning.

Field Frequency — The rate at which one complete field is scanned, 59.54 times a second in the NTSC color system.

Figure of Merit — (receiving system). (See G/T.)

Flux Density — Power density. Power per unit area normal to the direction of propagation of a propagating electromagnetic field.

Frame — One complete television picture, consisting of two fields of interlaced scanning lines. (See Field.)

Frequency Modulation (FM) — Angle modulation in which the instantaneous frequency of a sine-wave carrier is caused to depart from the carrier frequency by an amount proportional to the instantaneous value of the modulating wave.

Frequency Response — In a linear system, the frequency-dependent relation, in both gain and phase difference, between steady-state sinusoidal inputs and the resultant steady-state sinusoidal outputs. A plot of gain in decibels and phase angle in degrees versus logarithmic frequency is commonly called a Bode diagram.

Frequency Reuse — 1. A technique in which independent information is transmitted on orthogonal polarizations to "reuse" a given band of frequencies. 2. A technique in which independent information transmitted to different areas by the use of "spot" beams to "reuse" a given band of frequencies. Note: The techniques of 1 and 2 are sometimes clarified by addition of the phrases "by polarization diversity" and "by space diversity."

FSK — Frequency shift keying.

Full Duplex — The simultaneous 2-way flow of data between data terminals. A method of operation which provides simultaneous 2-way communications between two points.

Gain — (antenna) 1. The power gain of an antenna, in a specified direction, is 4π times the ratio of the power radiated per unit solid angle in that direction to the net power accepted by the antenna from its generator. The terms "gain" and "power gain" are synonymous. See: directivity; power gain; partial effective area (gain); realized gain. 2. The ratio of the signal level received (or transmitted) by an antenna to the signal level which would be received (or transmitted) by an isotropic antenna at the same location and under the same conditions, and with each antenna conjugately matched to the receiver (transmitter).

Gain — (amplifier) (See Power Gain.)

Gain Stability — Variation of the gain of a device, cascade of devices or system as a function of the family of possible causes of gain perturbation, such as temperature, humidity, line voltage and ageing of components. Ageing results in long-term changes in gain caused by gradual changes of components. Gain stability can be categorized into short-period (e.g., 1 second), medium-period (e.g., 1 hour), and longer variations. Gain stability is often specified as the maximum variation in decibels which a device experiences in a 24-hour test period. To avoid confusion it should be specifically stated whether the variation is the peak-to-peak variation, the peak plus-minus variation from a mean value, the rms variation from the mean, etc.

GCE — Ground communications equipment. This term relates to the earth station electronic equipment, such as receivers and exciters.

G/T — Figure of merit of a receiving system, expressed in dB/K. G/T is defined as the ratio of the receiving system gain, including the antenna gain, at a specified reference point in the receiving system (preceding the demodulator) to the receiving system noise temperature in Kelvins referred to the same point. For a given system and a given antenna elevation look angle, the value of G/T is independent of the point in the receiving system at which it is measured. In relation to its sensitivity of detecting the signals from a satellite, the higher the figure of merit the higher the sensitivity of the receiving system in detecting satellite signals.

HA — Hour angle. Angles about the polar axis of an equatorial coordinate system, traditionally measured in hours, minutes and seconds. The hour angle of an object is the polar angle between the meridian of the observer and the hour circle passing through the object. (See Hour Circles.)

Half Duplex — The non-simultaneous 2-way flow of data between communications terminals. Description of a system in which communication can be in either direction, but in only one way at a time.

Hour Circles — Great circles of the celestial sphere passing through the poles.

HPA — High-power amplifier. In a transmitting earth station, this is the final RF amplifier between the modulator/exciter and the antenna.

Idle Channel Noise — The total rms noise measured at the output of a channel with the input to the channel terminated with system reference impedance.

Insertion Gain — On insertion of a transducer in a transmission system, the insertion gain is the ratio of (1) the power delivered to the port of the system following the transducer to (2) the power delivered to the same port before insertion. The insertion gain may be equal to, more than or less than unity. It can be expressed as a power ratio or in decibels.

K — Kelvin. Temperature of a device in Kelvins. Zero K equals -273.15°C. The Kelvin scale is the same as the Celsius scale except for the offset of 273.15°C.

LNA — Low-noise amplifier. This is the preamplifier between the antenna and the earth station receiver. For maximum effectiveness, it must be located as near the antenna as possible, and is usually attached directly to the antenna receive port.

LNC — Low noise converter. Integrated LNA and downconverter.

Luminance — An attribute of light or color. Luminous intensity or brightness as measured by a photometer instead of by the human eye.

MATV — Master antenna television.

Modem — A contraction of modulator/demodulator. Usually a device that combines the modulation and demodulation functions in a single unit.

MTBF — Mean time between failure. A statistical determination of the time in hours-of-use between failures.

Multipoint (or Multi-drop) — A method of connecting several voice or data terminals to a single channel in a communications system.

MUX — Multiplexer or multiplex. Equivalent for or process of interleaving in frequency or time two or more messages for transmission on a single channel.

NF — Noise Figure. A figure of merit of a device, such as an LNA or receiver, which compares the device with a perfect device. NF is the ratio of (a) the noise power output of a matched transducer to (b) the portion of that noise which is attributable to thermal noise if the input port of the transducer is at the standard noise temperature of 290 Kelvins.
The noise figure of a device, expressed in dB, is related to its noise temperature T_E by

$$NF(dB) = 10 \log (1 + T_E/290)$$

where T_E is measured in Kelvins.

Noise Power (Spectral) Density — Noise power per unit of bandwidth.

Noise Temperature — More correctly called "effective noise temperature", this is a measure of the noise power referred to a given reference point of a matched receiving system, normalized with respect to bandwidth.

$$T_E = \frac{N}{kB} \equiv \frac{N_o}{k}$$

where
N is the noise power at the reference point in the noise bandwidth B, k is Boltzman's constant, and N_o is the noise power density.

NTSC — National Television Systems Committee. An industry-wide engineering group which, during 1950-1953, developed the color television specifications now established in the United States.

OMT — Orthomode transducer. A waveguide device attached to an antenna feed that permits using the antenna for simultaneous transmission and reception or orthogonal polarizations.

Orthogonal Polarization(s) — (See Polarizations, Orthogonal.)

Parity Check — A simple test for detecting errors in the transmission of ASCII characters. 1. (electronic computation) A summation check in which the bits in a character or block are added (modulo 2) and the sum checked against a single, previously computed parity digit; that is, a check that tests whether the number of ones is odd or even. 2. A method of checking the accuracy of transmission of digital data by adding a "parity bit" so that the total number of "ones" in each character is always odd.

Partial Effective Area (Gain) — That part of the effective area (gain) of an antenna which is in a specified polarization. The effective area (gain) of an antenna is the sum of two partial effective areas (gains) which are defined with respect to two orthogonal but otherwise arbitrary polarizations: e.g.: the effective area (gain) of any linearly polarized antenna is the sum of two equal partial effective areas (gains) which are respectively right-hand and left-hand circularly polarized. (See Polarizations, Orthogonal.)

Peak Unaffected Signal — (audio) The maximum sinewave input level which will be passed through a channel without clipping or limiting.

Polar Coordinate System — (See Equatorial Coordinate System.)

Polarization — 1. (wave) The polarization in a given direction of a single-frequency propagating electromagnetic wave is described by the amplitude and direction of its electric field (normal to the direction of propagation) is called the plane of polarization. The tip of the electric vector representing the field describes an elliptical locus called the polarization ellipse. Circular polarization and linear polarization are degenerate cases of elliptical polarization. The polarization is defined by the shape of the polarization ellipse (axial ratio), its orientation (tilt angle) and the direction in which the locus is traversed (sense). The sense of polarization is defined to be right-handed or left-handed. It is right handed if the electric vector rotates in a clockwise direction as viewed from behind the outgoing wave.
2. (antenna) The polarization of an antenna is defined by the polarization of the wave it radiates.

Polarization Efficiency — (of an antenna with respect to an incident wave) The polarization efficiency is the ratio of the incident power of a wave which is received by an antenna to the power which would be received if the wave and the antenna were polarization matched. A reciprocal antenna is polarization matched to an incident wave if the axial ratio, tilt angle and sense of polarization of the incident wave are coincident with those of the wave which would be radiated by the antenna.

Polarizations, Orthogonal — Any single-frequency propagating wave can be separated into orthogonal polarization components. The polarization components are orthogonal if at a point the total power density in the wave is equal to the sum of the power densities in the two components. Two polarizations are orthogonal if their tilt angles are mutually perpendicular, the magnitude of their axial ratios are equal and their senses are opposite. (See Polarization.)

Power Density — 1. (free-space propagating wave) Power per square meter in a plane normal to the direction of propagation (watts/meter^2).
2. (circuit) Power per unit bandwidth (usually in watts/Hz or watts/MHz).

Power Gain — 1. (antenna) See Gain. 2. (amplifier or transducer) The power gain (usually called the gain) is the ratio of the output power to the input power. The available gain is the gain realized under conjugately matched conditions. The gain can be expressed as a numerical ratio or in decibels. The numerical gain of a transducer can be less than, equal to or greater than unity. If it is less than unity, the numerical gain (fraction) is sometimes inverted and called a loss. The gain is called the saturated gain if the output power is constant with increasing input power.

Preemphasis — The intentional alteration of the spectrum of a signal by emphasizing one range of frequencies with respect to another, usually to gain an advantage with respect to noise or non-linear distortion. (See Deemphasis.)

Protocol — (data communications) A formal set of conventions which governs the flow and timing of data between communication devices.

PSK — Phase shift keying.

PWB — Printed wiring board.

QPSK — Quadriphase shift keying. A carrier modulation technique in which the phase of the carrier is shifted to any of 4 phases, whose separations are multiples of 90 degrees.

RAM — Random access memory.

RCVR — Receiver

Realized Gain (Effective Area) — (antenna or transducer) The <u>realized gain</u> is the gain reduced by the losses due to mismatch.

Reflector — (aperture antennas) A conducting surface (or one of a set of such surfaces) which is (are) shaped to collimate a transmitted wave into a beam or focus an incident wave onto a feed for coupling to a transmission line.

ROM — Read only memory.

RMS — Root mean square.

RS232C — A standard created by the Electronic Industries Association (EIA) for the transmission of data by wire over short distances (less than 50 feet).

SAbus — A serial data bus developed by Scientific-Atlanta for monitor and control of earth station equipment.

SCPC — Single channel per carrier. A satellite transmission system that employs a separate carrier for each channel. This method is usually used at earth stations that have a low volume of traffic compared to that which can be carried by a full transponder.

Signal-to-Quantization Noise Ratio — (digital audio) The ratio (dB) of the amplitude of a single frequency output signal to the rms quantizing noise appearing at the terminated channel output.

Simplex — Data or voice transmission in one direction only at all times.

Sinewave Overload — (digital audio) The sinewave input signal level which causes the converter saturation codes (all 1's or all 0's) to occur on the peaks of the signal.

SMATV — Satellite master antenna television. A master antenna distribution system that distributes satellite delivered signals as well as off-air signals and locally originated programming. SMATV systems are mainly found in multi-unit dwellings and lodging establishments.

SMPTE — Society of Motion Picture and Television Engineers.

S/N — Signal-to-noise ratio. The ratio of signal power and noise power. A video S/N (Signal-to-Noise Ratio) of 54 to 56 dB is considered to be an excellent S/N, that is, of best broadcast quality. A video S/N of 48 to 52 dB is considered to be a good S/N at the headend for Cable TV.

Sync. — An abbreviation for the words "synchronization," "synchronizing," etc. Applies to the synchronization signals, or timing pulses, which, for example, lock the electron beam of the picture monitors in step, both horizontally and vertically, with the electron beam of the pickup tube. The color sync signal (NTSC) is known as the color burst.

Synchronous Communication — Transmission and reception in a mode where there is continuous synchronization between the terminals.

TDM — Time division multiplex. The process or device in which each modulating wave modulates a separate pulse subcarrier, the pulse subcarriers being spaced in time so that no two pulses occupy the same time interval. Time division permits the transmission of two or more signals over a common path by using different time intervals for the transmission of the intelligence of each message signal. (See also FDM.)

TDMA — Time-division multiple access (of a satellite transponder by different earth stations).

TED — Threshold extension demodulator. A circuit designed to lower the threshold of FM (frequency-modulation) demodulators, permitting operation of the system with a lower C/N without impulse noise showing on the picture.

Termination — A one-port load that terminates a transmission line in a specified manner, usually to prevent reflection.

THD — Total harmonic distortion. The ratio of the rms energy contained in all harmonic components of a single-frequency input signal to the fundamental component amplitude.

TVRO — Television receive-only (earth station).

TX — Transmit or transmitter.

Uplink — The circuit between a transmitting earth station and a satellite, including the earth transmitter and antenna, the earth-satellite propagation path, and the satellite antenna and receiver.

Vestigial Sideband — 1. Reference to a type of amplitude-modulated ratio or television signal in which most of one sideband is eliminated, leaving the other sideband intact. 2. The transmitted portion of the sideband which has been largely suppressed by a transducer having a gradual cutoff in the neighborhood of the carrier frequency.

Video — A term pertaining to the bandwidth and spectrum position of the signal resulting from television scanning.

Video Plus — A technique whereby SCPC carriers (analog or digital) share a transponder with a full-transponder video signal without mutual interference.

VSWR — Voltage standing-wave ratio (in a waveguide or transmission line). The ratio of the magnitude of the transverse electric field in a plane of maximum strength to the magnitude at the equivalent point in an adjacent plane of minimum field strength.

zenith — The zenith of an observer is the point of the celestial sphere which is vertically overhead. The nadir is the point vertically beneath.

Evaluation of Intrinsic Noise in FDM/FM Systems

B.W. Brinegar

The evaluation of FDM/FM message systems can be greatly enhanced by comparing theoretical calculations versus empirical data. This paper presents equations required to calculate the conventional signal-to-noise curves for FDM/FM systems using a limiter/discriminator as a demodulator. Also, equations used for occupied bandwidth and linearity bandwidth are presented.

Examples of several message configurations utilizing these calculations are included. Signal-to-noise curves are calculated by the equation:

$$S/N_{pwp} = (S/N\ Normal_{dB} - S/N\ Threshold_{dB})_{pwp} + S/N_{1/Fpwp}$$

This equation has three distict areas: the S/N 1/F region, the S/N normal region and the S/N threshold region. Observing the equations on the following pages, it becomes obvious which parts of the curve are affected by the different functional elements of the FDM/FM system (see Figure 1).

a. The 1/F noise region is generally degraded by power supply noise, station equipment noise, ground loops, baseband noise figure versus the discriminator constant, noise figure of the modulator baseband amplifier and various other man-made noise.

b. The S/N normal region is defined by the system's parameters and the noise figure of the receiver.

c. The S/N threshold is dependent on the IF noise bandwidth and the system parameters; however, the purity of the limiting and demodulation process can adversely affect this portion of the curve.

The S/N curves were calculated with the following parameters:

- 1/F noise floor = -110 dBm
- Baseband test tone = -25 dBm per channel
- Receiver noise figure = 13 dB
- Other parameters as shown on the curves

In actual practice satellite FDM/FM systems operate in the low C/N region, thereby allowing the modulator and demodulator to be optimized for a wider linearity bandwidth. Therefore, the 1/F noise floor as shown in the attached curves is set to -110 dBm for clarity and is not achieved in actual practice. The 1/F noise floor in practice runs approximately -87 dBm for 12-channel operation.

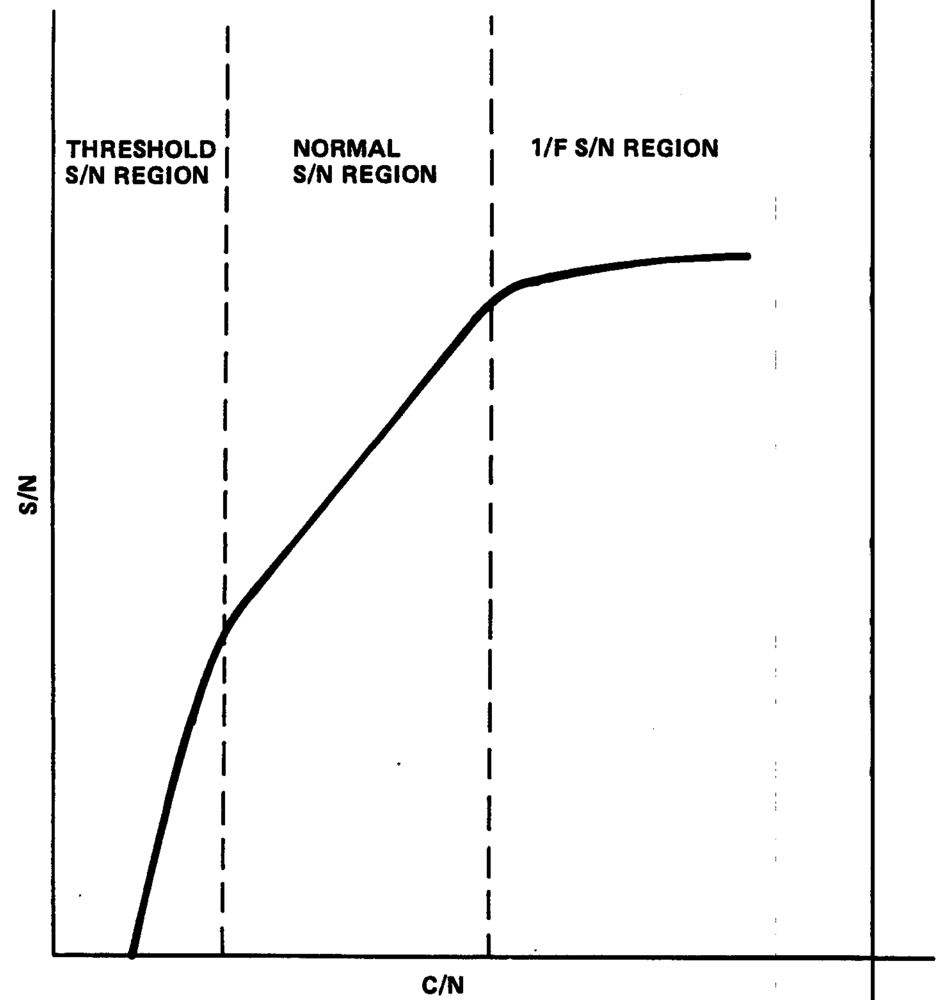


Figure 1. Conventional S/N vs. C/N Curve for FDM/FM Message Systems

The columns labeled NPR are actually NPR residual and do not include the intermodulation distortions generally associated with NPR measurements.

Signal-to-Noise Calculations

$$S/N_{normal} = \Big(10 \log_{10} (2*BW/1E-3) + 10*\log_{10} T_k/290 - C_{dBm} + N_{fdB}$$

$$- 20 \log_{10} (D_p/BW) - PE(X) - PW - 174\Big)$$

$$S/N_{threshold} = 10 \log \left[1/3e^{-c/n} \left(\frac{B_{if}}{f_{ch}}\right)^2 \sqrt{1 + 24 \left(\frac{\sigma}{B_{if}}\right)^2 \left(\frac{C}{N}\right)^2} + 1\right]$$

$$S/N_{pwp} = 10^{\left(\frac{87.5 - S/N_{dB}}{10}\right)}$$

$$S/N_{dB} = 87.5 - 10 \log S/N_{pwp}$$

$$C/N = C/KTB_{if} N_f$$

$$\sigma = 10^{\left(\frac{NLR}{20}\right)} \cdot D$$

$$C = \frac{10^{\left(\frac{C_{dBm}}{10}\right)}}{1000}$$

$$N_f = 10^{\left(\frac{N_{fdB}}{10}\right)}$$

σ = RMS Composite Deviation

C = Received Carrier in Watts

C_{dBm} = Received Carrier in dBm

K = 1.3804×10^{-23}

T_k = Temperature in Kelvin

f_{ch} = Notch Frequency

Log_{10} = Ln (X)/Ln (10)

PW = Psophometric Weighting

Dp = Peak Deviation of the Channel

$$PE = \text{Preemphasis} = 5 - 10 \log_{10} \left\{ 1 + \frac{6.90}{1 + \frac{5.25}{\left(\frac{f_r}{f_{ch}} - \frac{f_{ch}}{f_r} \right)^2}} \right\}$$

F_r = 1.25 f_{max}

f_{max} = Highest B/B Frequency

f_{low} = Lowest B/B Frequency

f_{ch} = Notch Frequency

S/N = NPR + BWR - NLR

$$BWR = 10 \log \frac{f_{max} - f_{low}}{3100}$$

NLR = 2.6 + 2 log N for 0 through 48 channels

NLR = -1 + 4 log N for 49 through 239 channels

NLR = -15 + 10 log N for 240 through 3600 channels

$$S/N = C/N + 10 \log \frac{BW_{if}}{6200} + 20 \log \frac{D_p}{f_{ch}}$$

Bandwidth Calculations

BW_o = Occupied bandwidth

BW_ℓ = Linearity/delay bandwidth

fm = Upper test cahnnel in MHz

Dv = Test tone rms deviation in MHz

Fp = Peak factor; 12 dB normally accepted

$$BW_o = 2\ fm + \log^{-1}\left[\frac{FP}{20}\right] Dv \log^{-1}\left(\frac{NLR}{20}\right)$$

$BW_\ell = \pm 1/2\ BW_o$ setting Fp = 0 ($\log^{-1}$ Fp = 1)

References

1. "Radio Relay Systems," International Telecommunications Union; CCIR, Volume 4, page 83.

2. Tarvin, J.C., "Microwave Signal to Noise Vs. Received Carrier Power Calculations for FM-FDM Systems," Rockwell International, WP-44.

3. "Psophometric Noise and Power," ITT Reference Data for Radio Engineers, 1968, Page 2-3.

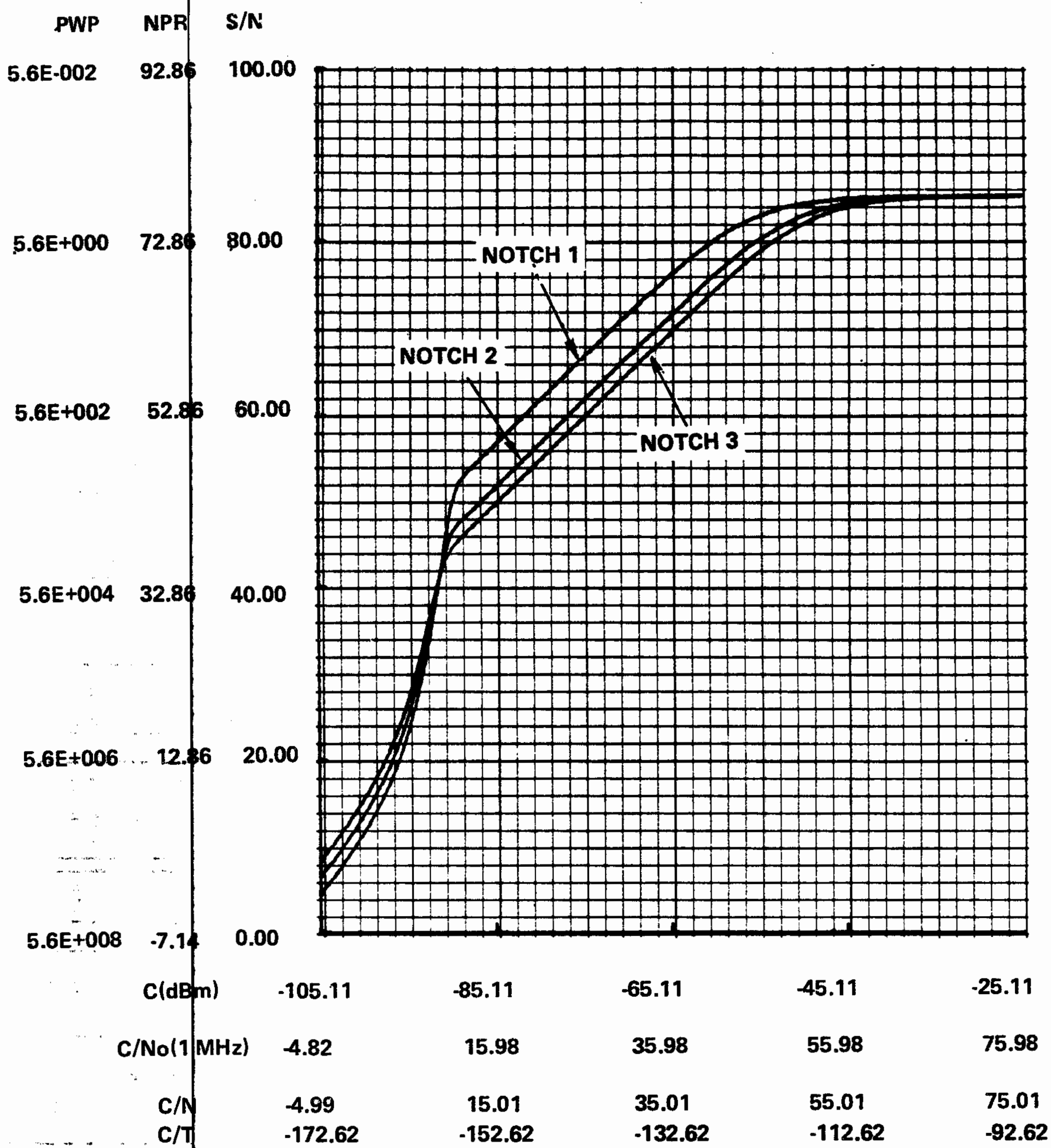
4. Rice, S.O., "Noise in FM Receivers," Proceedings of the Symposium on Time Series Analysis, Wiley, 1963, page 395.

5. "Evaluation of S/N in the Threshold Region of an FM Receiver." Notes from M. Ramadan.

6. "Signal-to-Noise Computer Program." Notes from R. Harris.

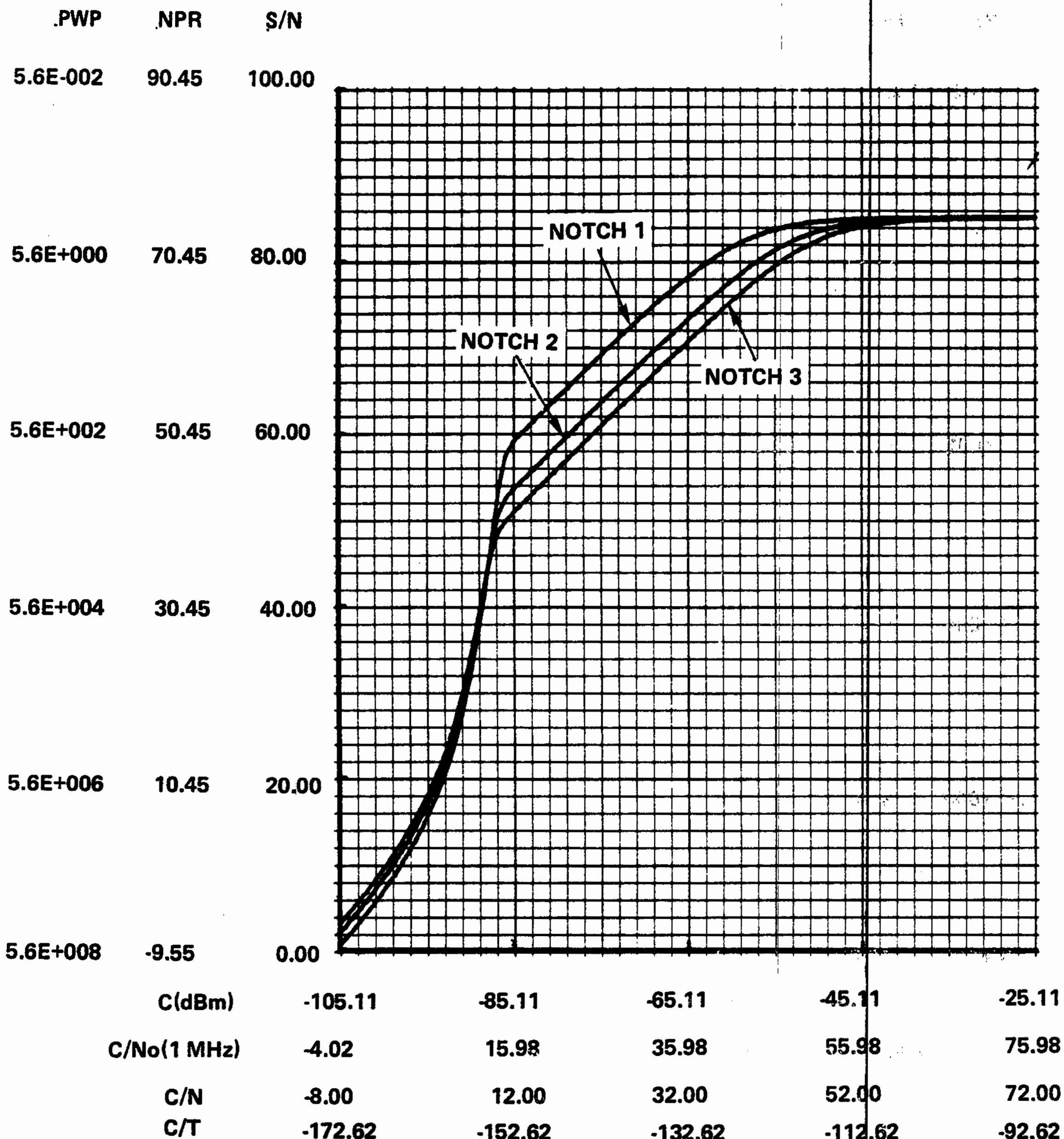
SCIENTIFIC-ATLANTA
MESSAGE DATA CHART

Channel Capacity	BW (MHz)	Top BB Freq. (kHz)	Test Tone Freq. (kHz)	Noise Load Ratio (dB)	RMS Test Dev. (kHz)	RMS MC Dev. (kHz)	Level into S-A Mod (140 kHz/-40 dBm) TT Dev. (dBm)	RMS Dev. (dBm)	1st Car. Null@ TT Freq.	S/N = NPR + (dB)	RX Max Sig. Level (dBm)	8000 pwp0 C/T	8000 pwp	50,000 pwp C/T	N 50, p
12	1.25	60	36.48	3.32	109	159	-42.17	-38.85	-47.07	8.57	-47	-154.7	39.90	-160.7	31
24	2.5	108	65.66	+ 4.52	164	275	-38.63	-34.11	-41.96	10.39	-45	-153.0	38.11	-159.0	30
36	2.5	156	94.85	+ 5.23	168	307	-38.42	-33.18	-38.77	11.44	-45	-150.0	37.06	-156.0	29
60	2.5	252	153.2	6.11	136	276	-40.25	-34.14	-34.61	12.78	-45	-144.0	35.72	-150.0	27
60	5.0	252	153.2	6.11	270	546	-34.30	-28.19	-34.61	12.78	-42	-149.9	35.72	-155.9	27
72	2.5	300	182.4	6.43	125	261	-40.98	-34.55	-33.09	13.25	-45	-141.7	35.25	-147.7	27
72	5.0	300	182.4	6.43	294	616	-33.56	-27.12	-33.09	13.25	-42	-149.1	35.25	-155.1	27
96	5.0	408	248.1	6.93	263	584	-34.52	-27.59	-30.42	14.13	-42	-145.5	34.37	-151.5	26
96	7.5	408	248.1	6.93	360	799	-31.80	-24.87	-30.42	14.13	-40	-148.2	34.37	-154.2	26
132	5.0	552	335.6	7.48	223	529	-35.96	-28.48	-27.79	14.93	-42	-141.4	33.57	-147.4	25
132	7.5	552	335.6	7.48	376	891	-31.42	-23.94	-27.79	14.93	-40	-145.9	33.57	-151.9	25
132	10.0	552	335.6	7.48	430	1020	-30.25	-22.77	-27.79	14.93	-39	-147.1	33.57	-153.1	25
192	5.0	804	488.8	8.13	180	459	-37.82	-29.69	-24.53	15.94	-42	-136.3	32.56	-142.3	24
192	7.5	804	488.8	8.13	297	758	-33.47	-25.34	-24.53	15.94	-40	-140.6	32.56	-146.6	24
192	10.0	804	488.8	8.13	457	1167	-29.72	-21.59	-24.53	15.94	-39	-144.4	32.56	-150.4	24
252	7.5	1052	639.6	9.01	260	733	-34.62	-25.61	-22.19	16.25	-40	-137.1	32.25	-143.1	24
252	10	1052	639.6	9.01	358	1009	-31.84	-22.83	-22.19	16.25	-39	-139.9	32.25	-145.9	24
252	15	1052	639.6	9.01	577	1627	-27.70	-18.69	-22.19	16.25	-37	-144.1	32.25	-150.1	24
312	10	1300	790.4	9.94	320	1005	-32.82	-22.88	-20.35	16.25	-39	-137.1	32.25	-143.1	24
312	15	1300	790.4	9.94	546	1716	-28.17	-18.23	-20.35	16.25	-36	-141.7	32.25	-147.7	24
432	15	1796	1092	11.35	401	1479	-30.86	-19.51	-17.55	16.25	-37	-136.2	32.25	-142.2	24
432	17.5	1796	1092	11.35	517	1919	-28.65	-17.30	-17.55	16.25	-37	-138.5	32.25	-144.5	24
432	20	1796	1092	11.35	616	2276	-27.13	-15.78	-17.55	16.25	-36	-139.9	32.25	-145.9	24
432	25	1796	1092	11.35	729	2688	-25.67	-14.32	-17.55	16.25	-35	-141.4	32.25	-147.4	24
612	20	2540	1544	12.87	454	1996	-29.78	-16.91	-14.54	16.25	-36	-134.2	32.25	-140.2	24
792	20	3284	1997	13.99	356	1784	-31.89	-17.90	-12.30	16.24	-36	-129.9	32.26	-135.9	24
792	25	3284	1997	13.99	499	2494	-28.96	-14.97	-12.30	16.24	-35	-132.8	32.26	-138.8	24
972	25	4028	2449	14.88	410	2274	-30.67	-15.79	-10.53	16.24	-35	-129.4	32.26	-135.4	24
972	36	4028	2449	14.88	802	4417	-24.84	- 9.96	-10.53	16.24	-34	-135.2	32.26	-141.7	24
1092	36	4892	2974	15.38	701	4118	-26.01	-10.62	- 8.84	16.59	-34	-132.4	31.91	-138.4	23
1332	36	5884	3577	16.25	591	3834	-27.49	-11.24	- 7.24	16.52	-34	-129.3	31.98	-135.3	23
1872	36	8120	4937	17.72	419	3181	-30.48	-12.75	- 4.44	16.46	-34	-123.5	32.04	-129.5	24



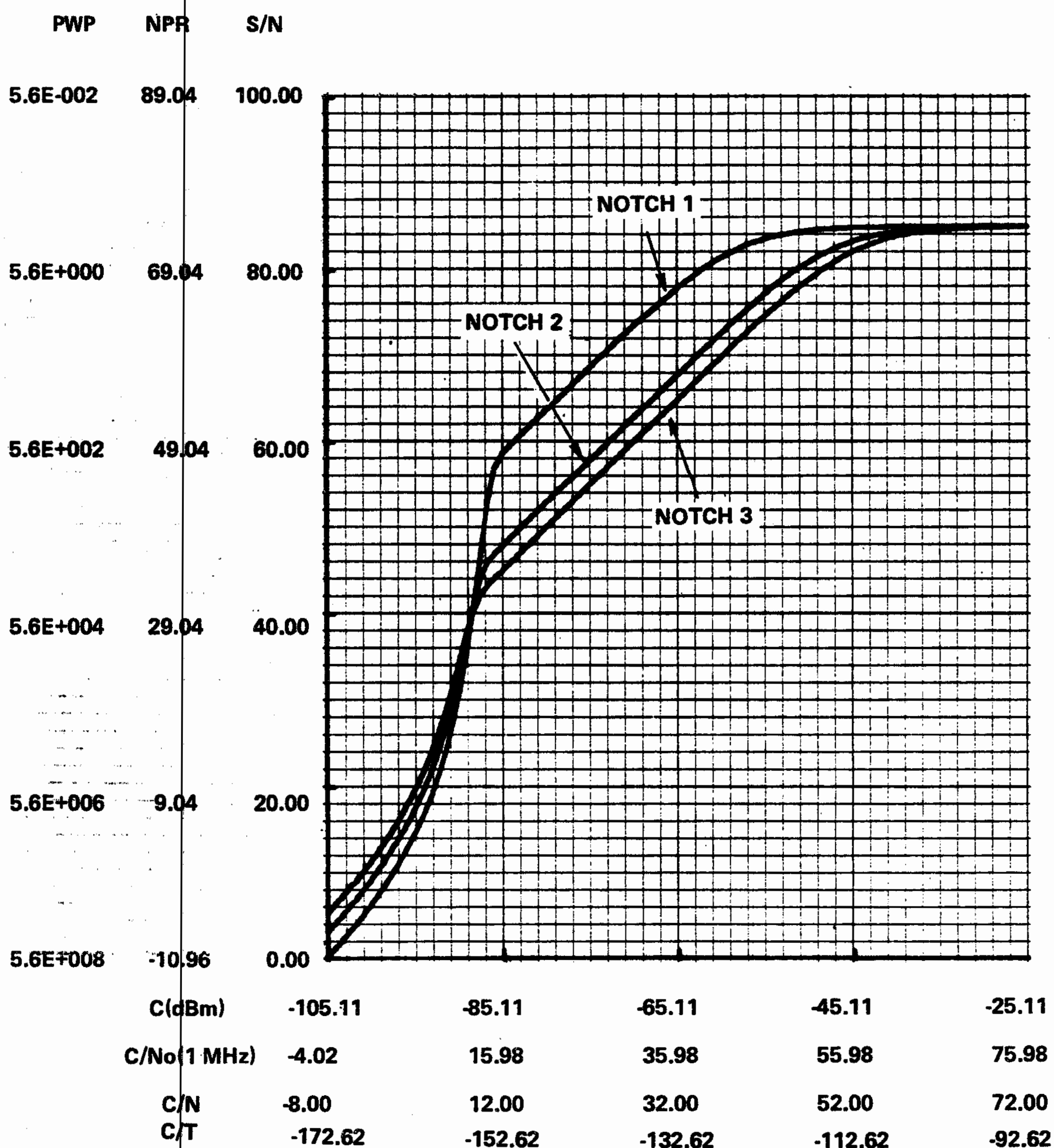
N. F. (dB)	13
BW (I. F.)	1250000
CHANNEL LOADING	12
NOTCH 1	16000
NOTCH 2	36000
NOTCH 3	56000
OCCUPIED BW	1620978.36
LINEARITY BW	497028.88
T. T. DEVIATION (RMS)	109000
P. E.	YES
P. S.	NO

Figure 3. S/N vs. C/N for 12 Channel Loading



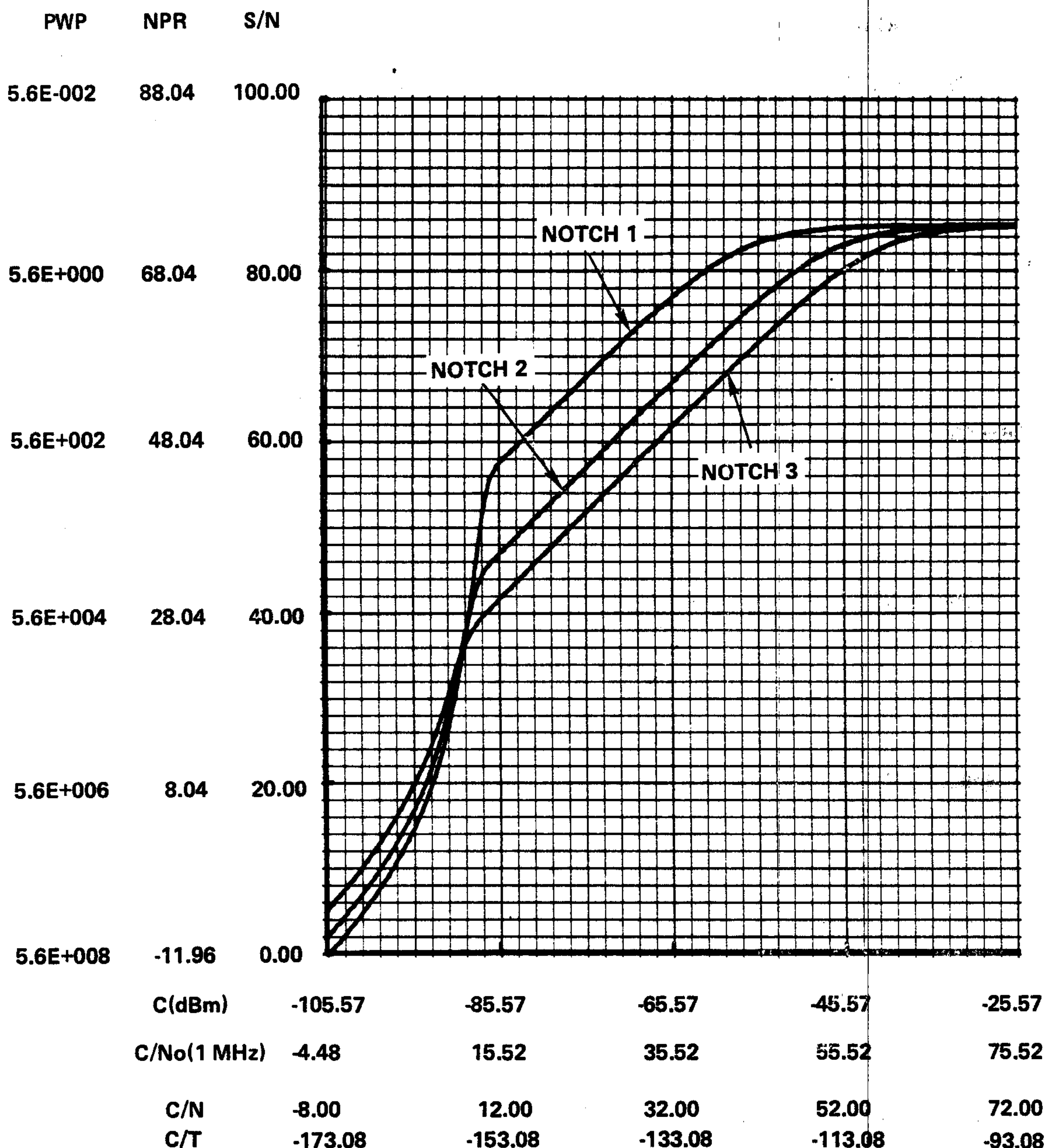
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	24
NOTCH 1	16000
NOTCH 2	36000
NOTCH 3	56000
OCCUPIED BW	2636442.57
LINEARITY BW	823887.94
T. T. DEVIATION (RMS)	164000
P. E.	YES
P. S.	NO

Figure 4. S/N vs. C/N for 24 Channel Loading



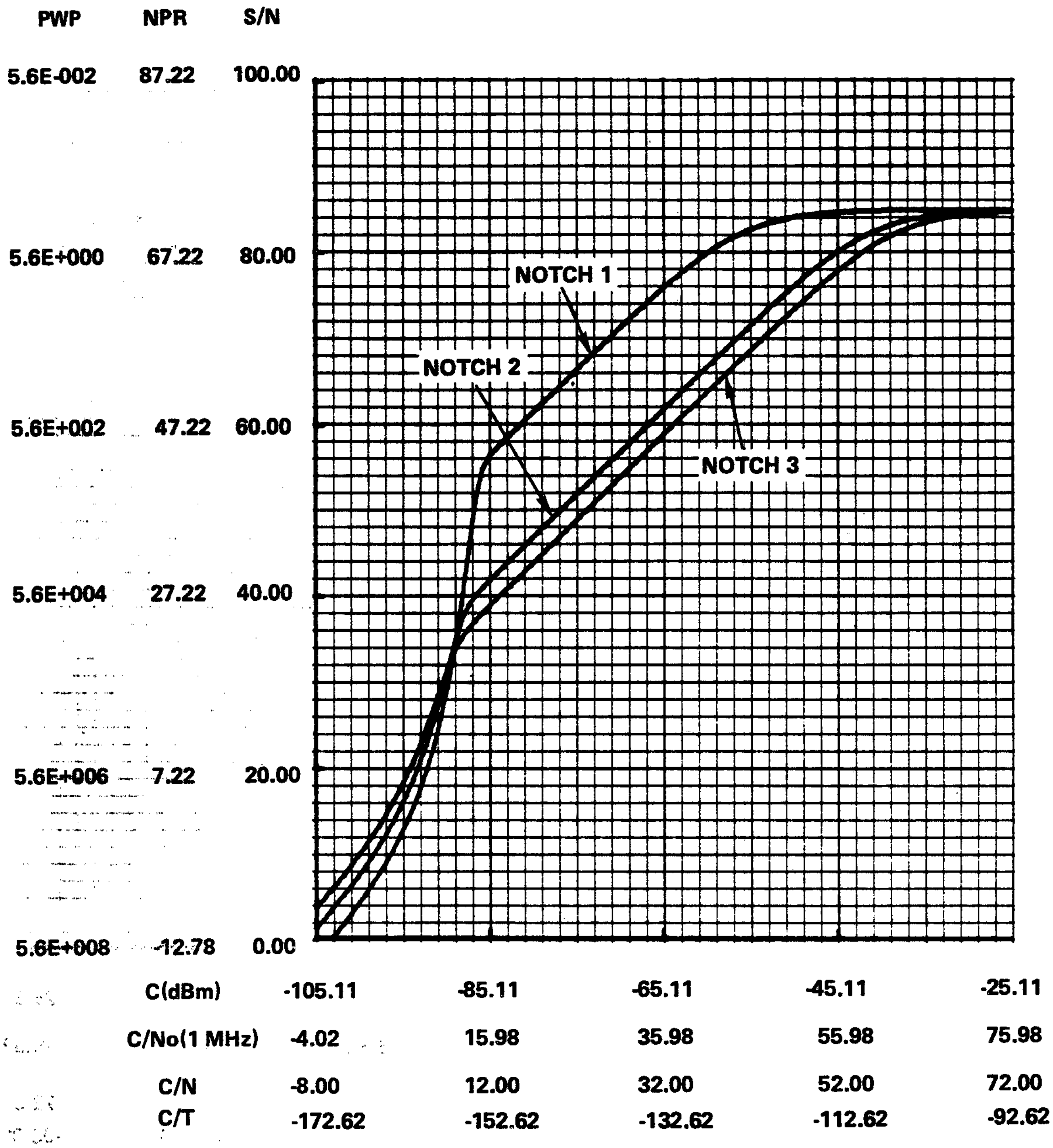
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	36
NOTCH 1	16000
NOTCH 2	78000
NOTCH 3	140000
OCCUPIED BW	2894077.90
LINEARITY BW	960588.92
T. T. DEVIATION (RMS)	168000
P. E.	YES
P. S.	NO

Figure 5. S/N vs. C/N for 36 Channel Loading



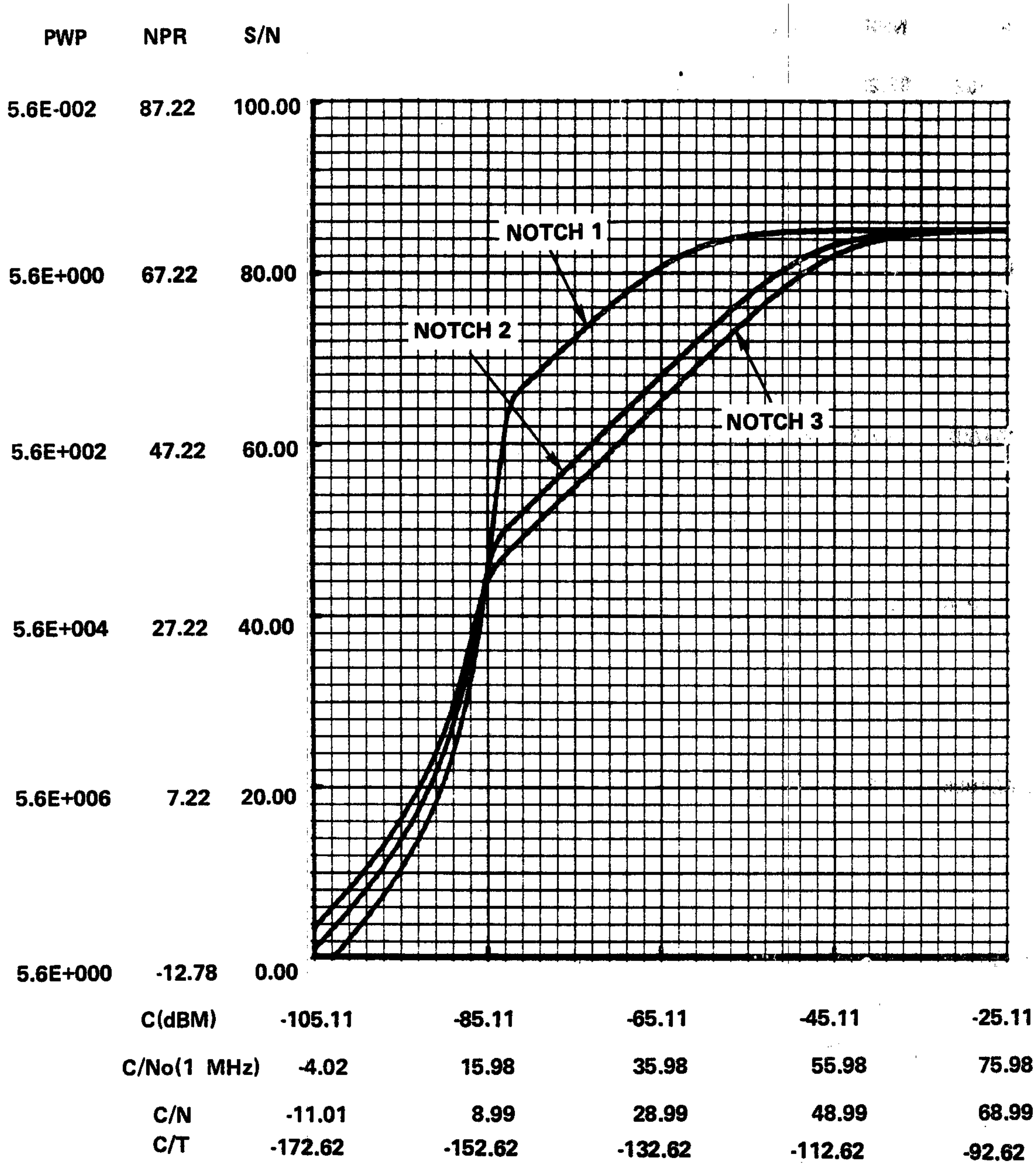
N. F. (dB)	13
BW (I. F.)	2250000
CHANNEL LOADING	48
NOTCH 1	16000
NOTCH 2	70000
NOTCH 3	185000
OCCUPIED BW	2796530.88
LINEARITY BW	1007872.12
T. T. DEVIATION (RMS)	151000
P. E.	YES
P. S.	NO

Figure 6. S/N vs. C/N for 48 Channel Loading



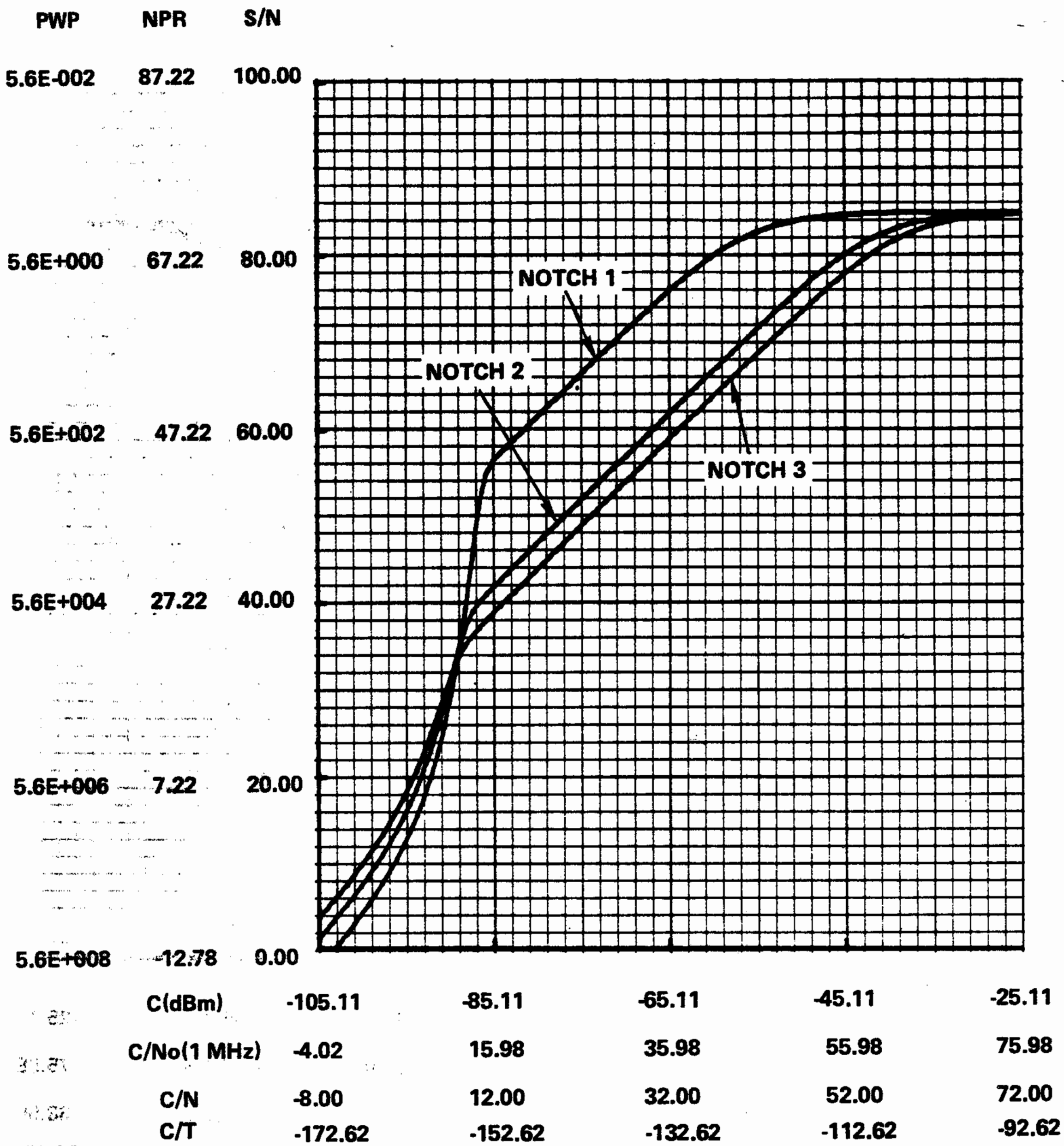
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	128000
NOTCH 3	240000
OCCUPIED BW	2692764.16
LINEARITY BW	1053792.93
T. T. DEVIATION (RMS)	136000
P. E.	YES
P. S.	NO

Figure 7. S/N vs. C/N for 60 Channel Loading



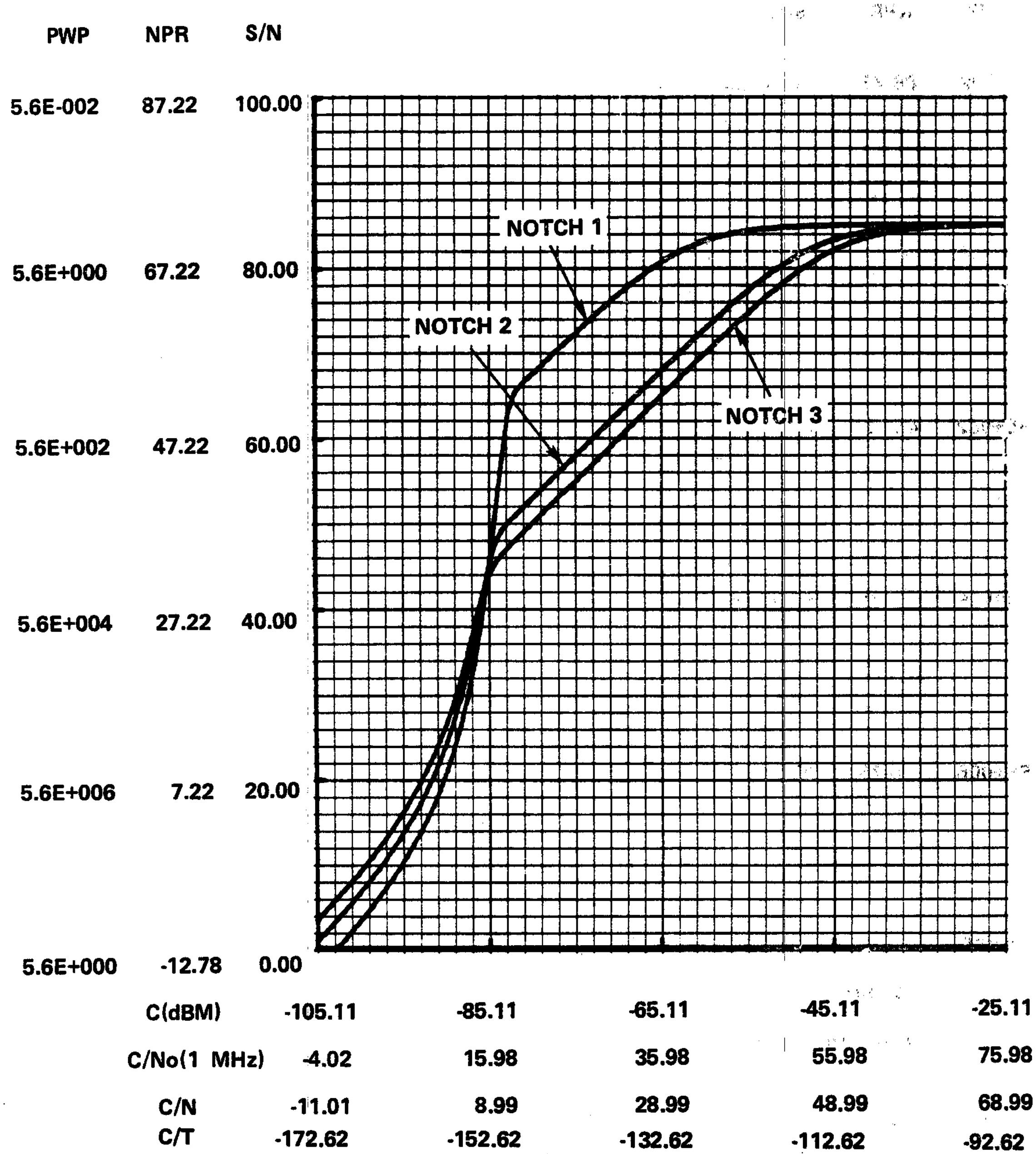
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	120000
NOTCH 3	240000
OCCUPIED BW	4849430.61
LINEARITY BW	1595500.68
T. T. DEVIATION (RMS)	270000
P. E.	YES
P. S.	NO

Figure 8. S/N vs. C/N for 60 Channel Loading



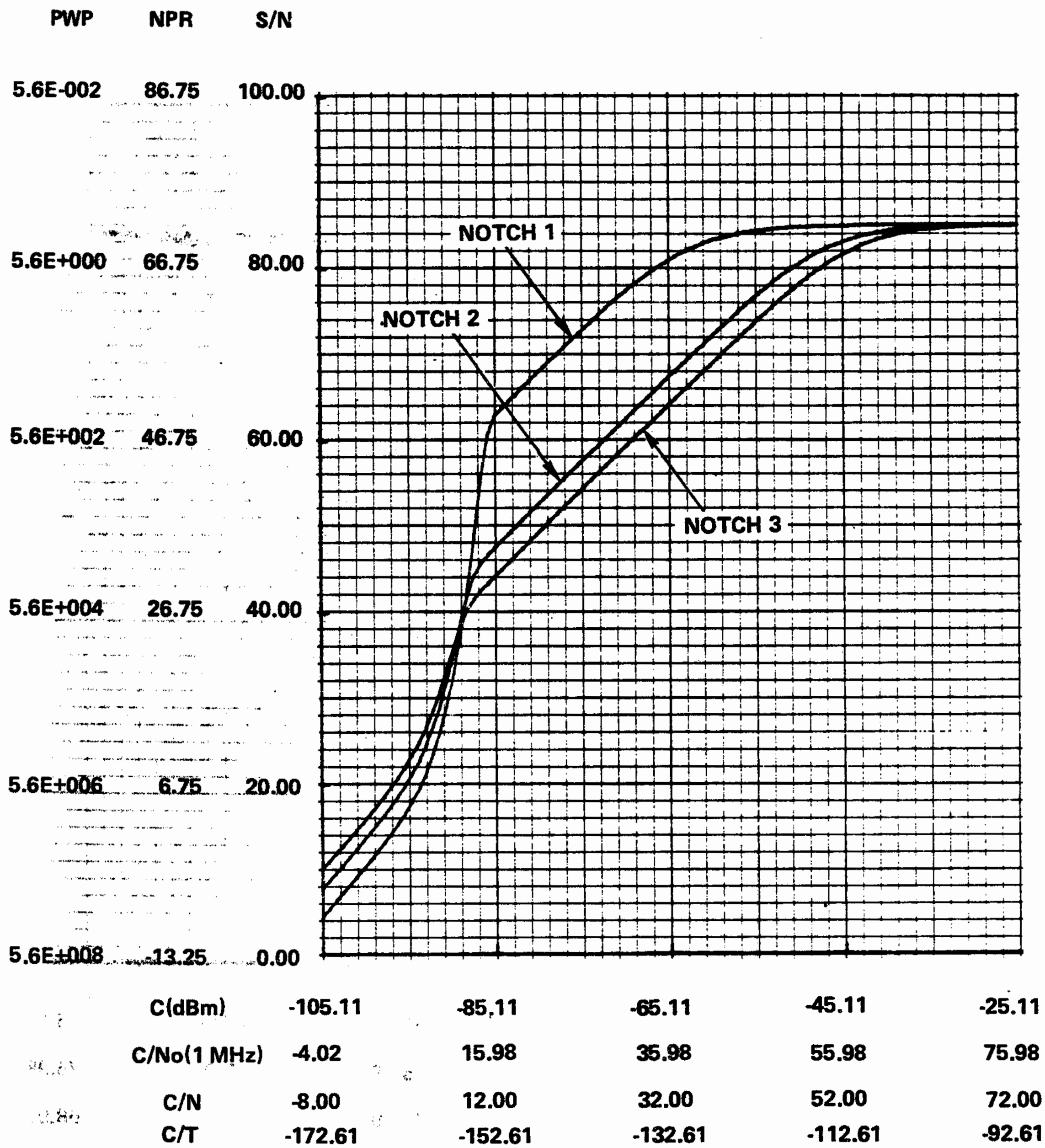
N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	128000
NOTCH 3	240000
OCCUPIED BW	2692764.16
LINEARITY BW	1053792.93
T. T. DEVIATION (RMS)	136000
P. E.	YES
P. S.	NO

Figure 7. S/N vs. C/N for 60 Channel Loading



N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	60
NOTCH 1	16000
NOTCH 2	120000
NOTCH 3	240000
OCCUPIED BW	4849430.61
LINEARITY BW	1595500.68
T. T. DEVIATION (RMS)	270000
P. E.	YES
P. S.	NO

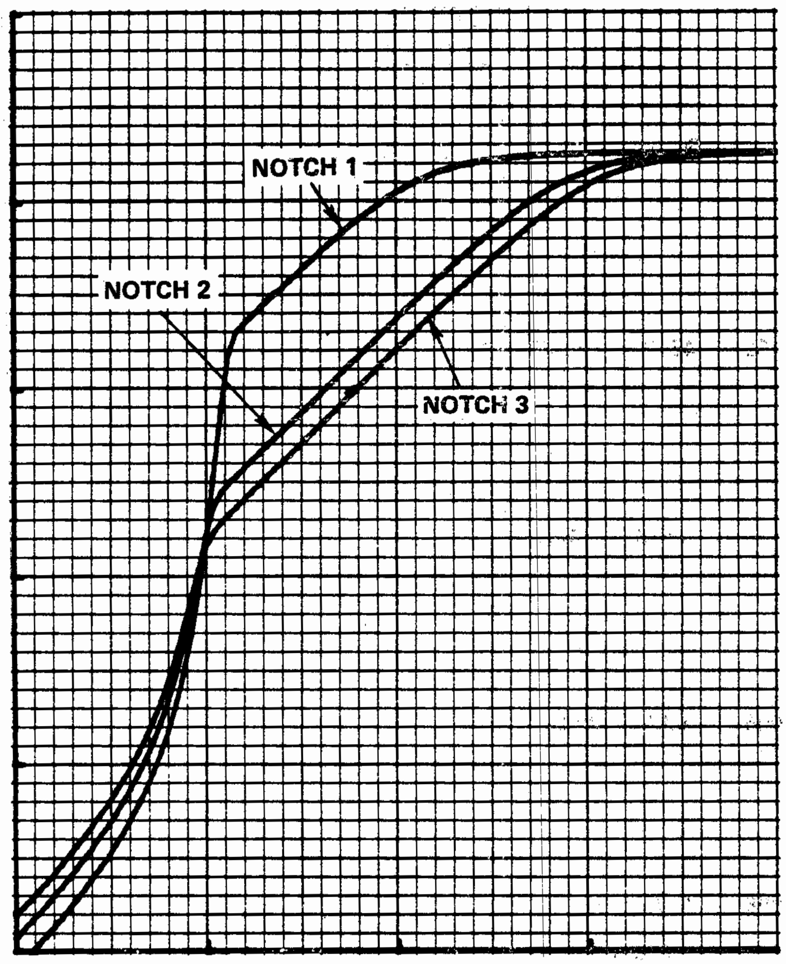
Figure 8. S/N vs. C/N for 60 Channel Loading



N. F. (dB)	13
BW (I. F.)	2500000
CHANNEL LOADING	72
NOTCH 1	16000
NOTCH 2	140000
NOTCH 3	270000
OCCUPIED BW	5507311.65
LINEARITY BW	1832661.48
T. T. DEVIATION (RMS)	294000
P. E.	YES
P. S.	NO

Figure 9. S/N vs. C/N for 72 Channel Loading

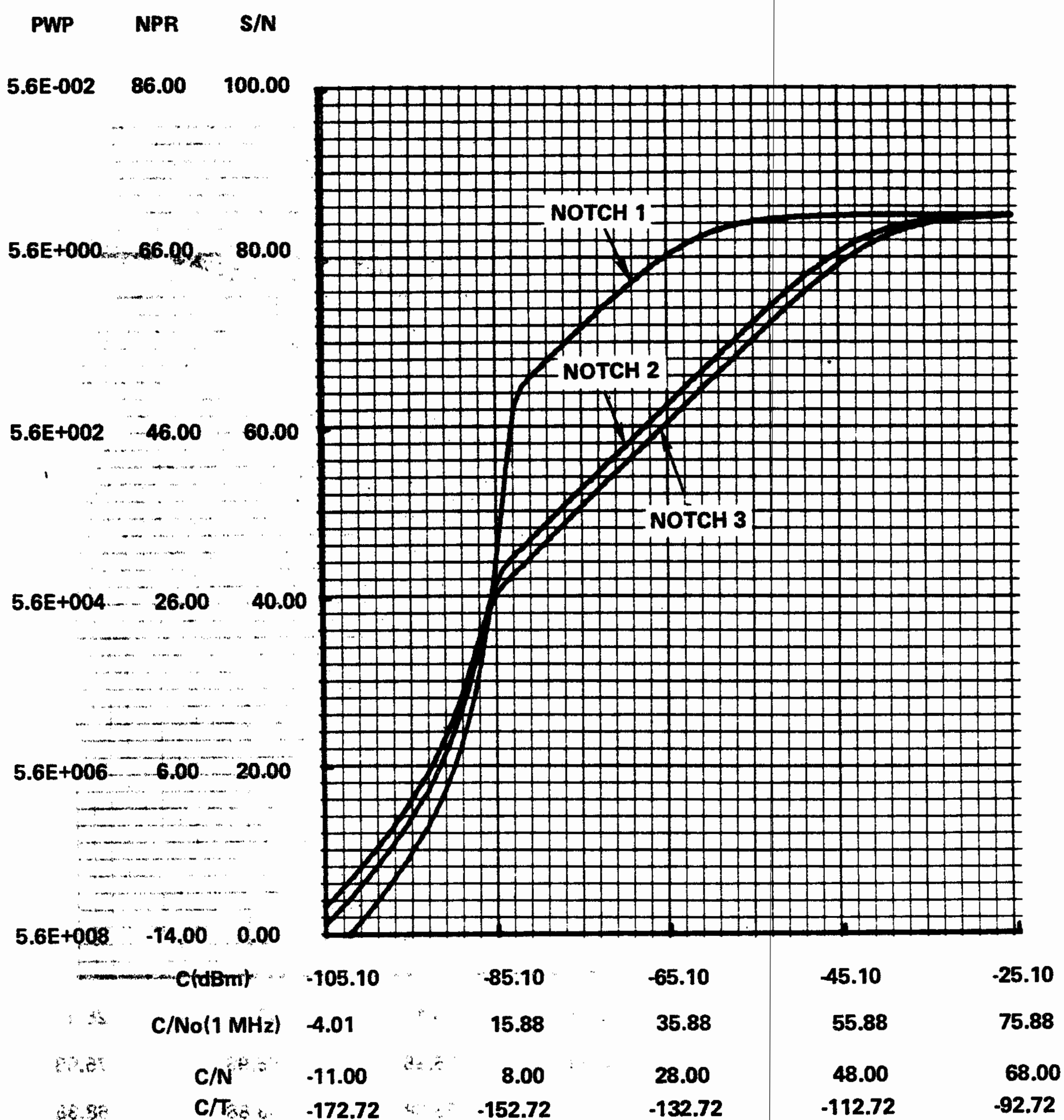
PWP	NPR	S/N
5.6E-002	86.75	100.00
5.6E+000	66.75	80.00
5.6E+002	46.75	60.00
5.6E+004	26.75	40.00
5.6E+006	6.75	20.00
5.6E+008	-13.25	0.00



C(dBm)	-105.11	-85.11	- 65.11	-45.11	-25.11
C/No(1 MHz)	-4.82	15.98	35.98	55.98	75.98
C/N	-11.81	8.88	28.88	48.88	68.88
C/T	-172.62	-152.62	-132.62	-112.62	-92.62

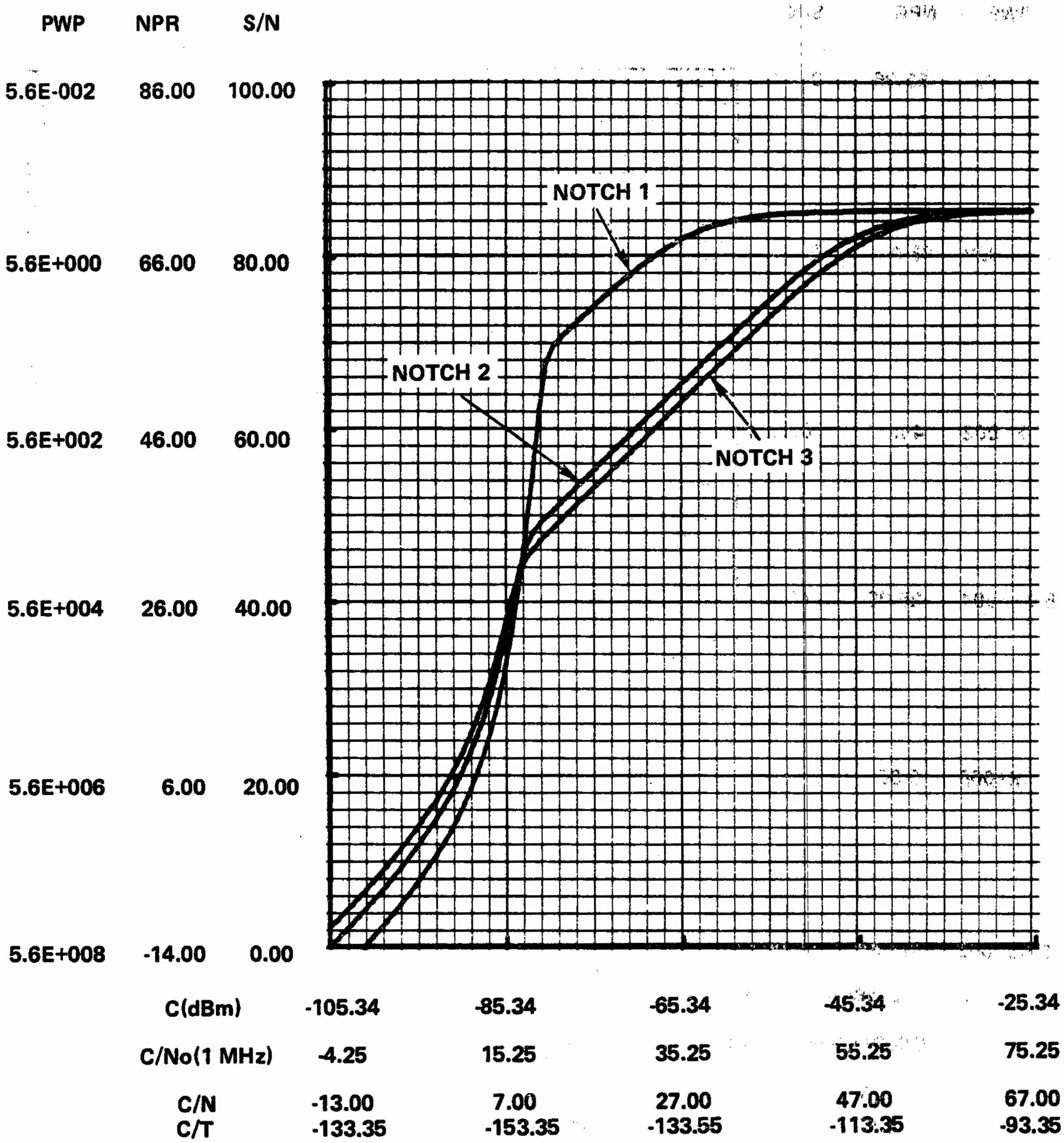
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	72
NOTCH 1	16000
NOTCH 2	140000
NOTCH 3	270000
OCCUPIED BW	5507311.65
LINEARITY BW	1832661.48
T. T. DEVIATION (RMS)	294000
P. E.	YES
P. S.	NO

Figure 10. S/N vs. C/N for 72 Channel Loading



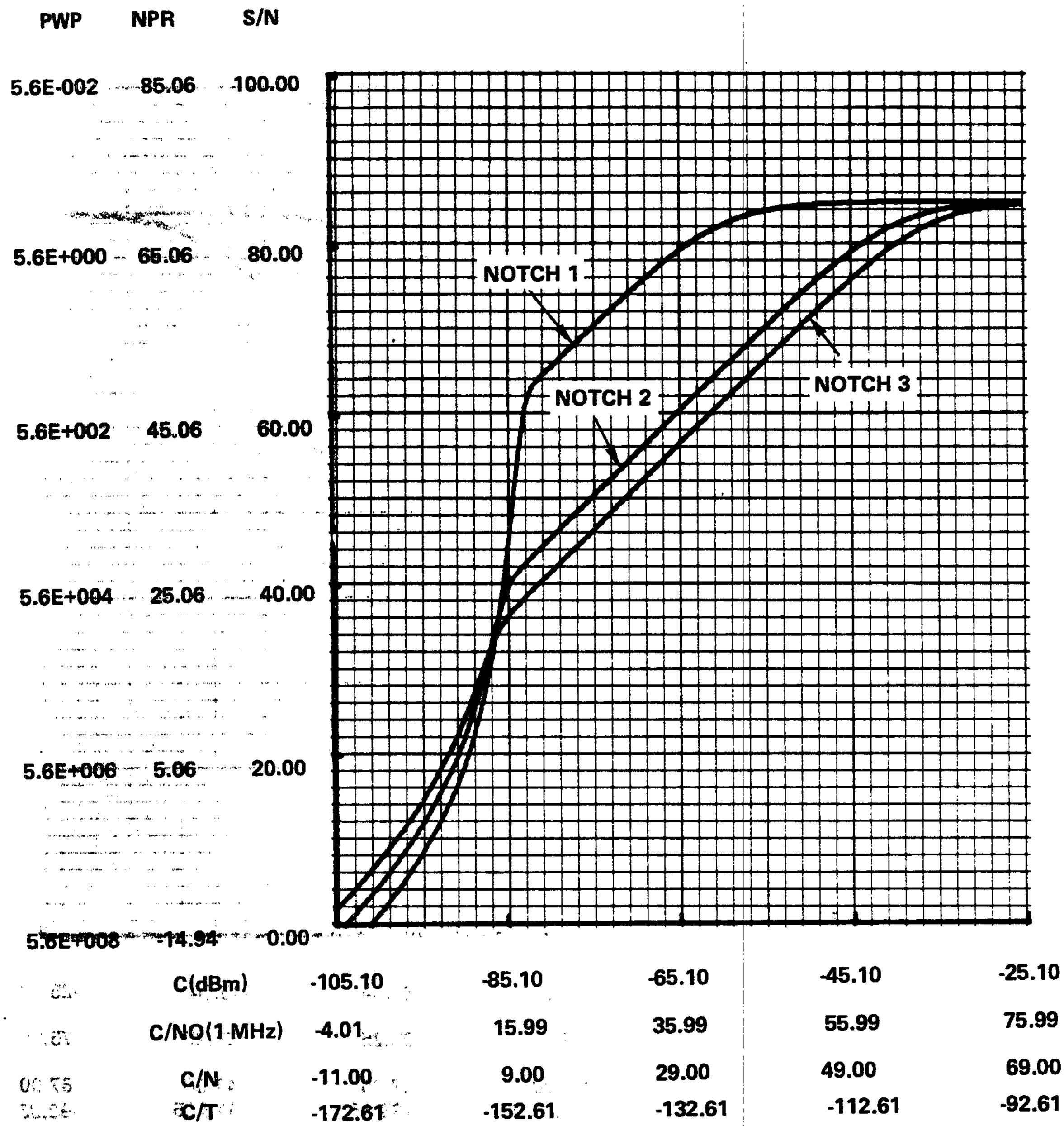
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	96
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	394000
OCCUPIED BW	5441858.22
LINEARITY BW	18588882.33
T. T. DEVIATION (RMS)	263000
P. E.	YES
P. S.	NO

Figure 11. S/N vs. C/N for 96 Channel Loading



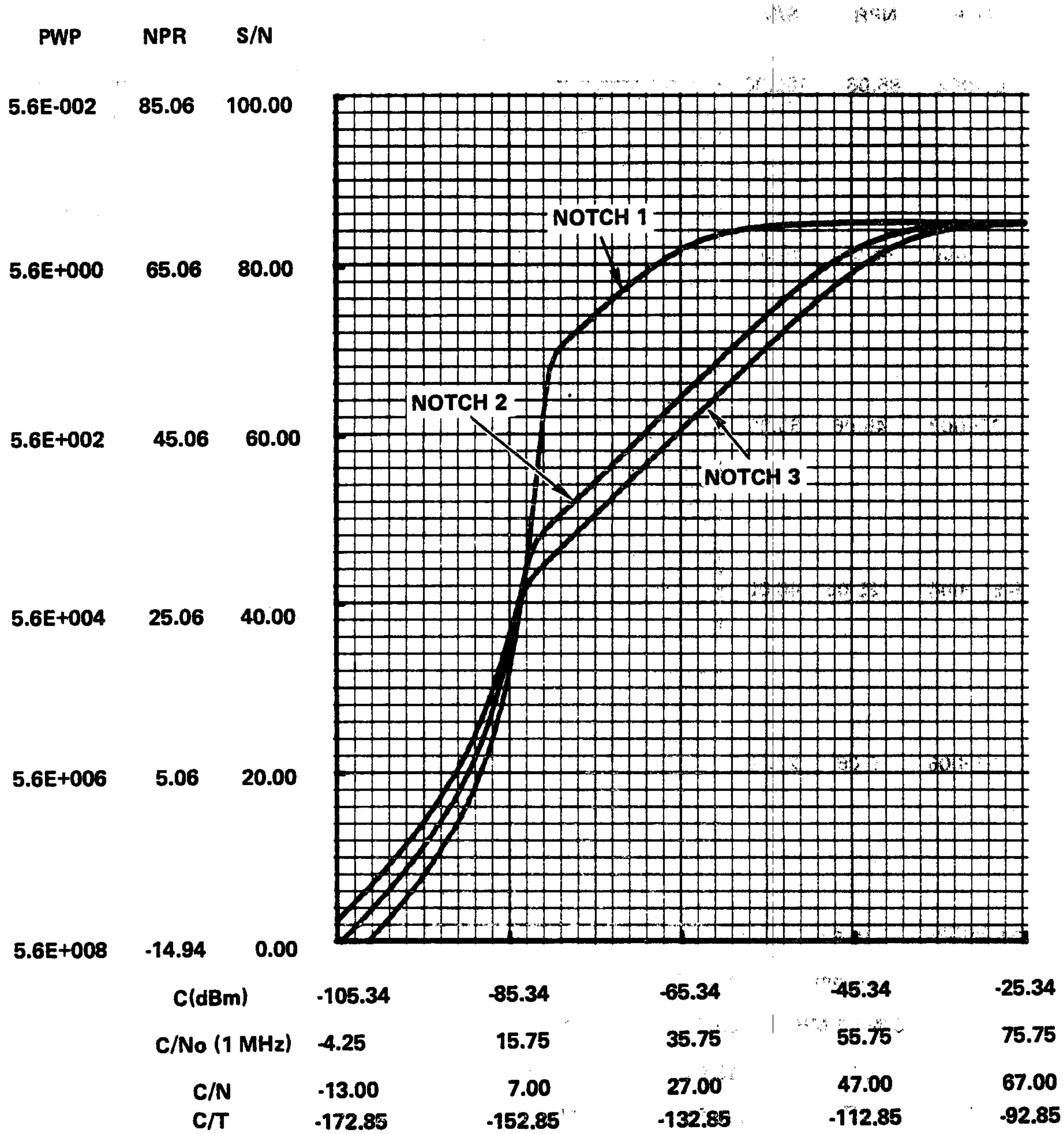
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	96
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	394000
OCCUPIED BW	7156826.31
LINEARITY BW	2388772.77
T. T. DEVIATION (RMS)	360000
P. E.	YES
P. S.	NO

Figure 12. S/N vs. C/N for 96 Channel Loading



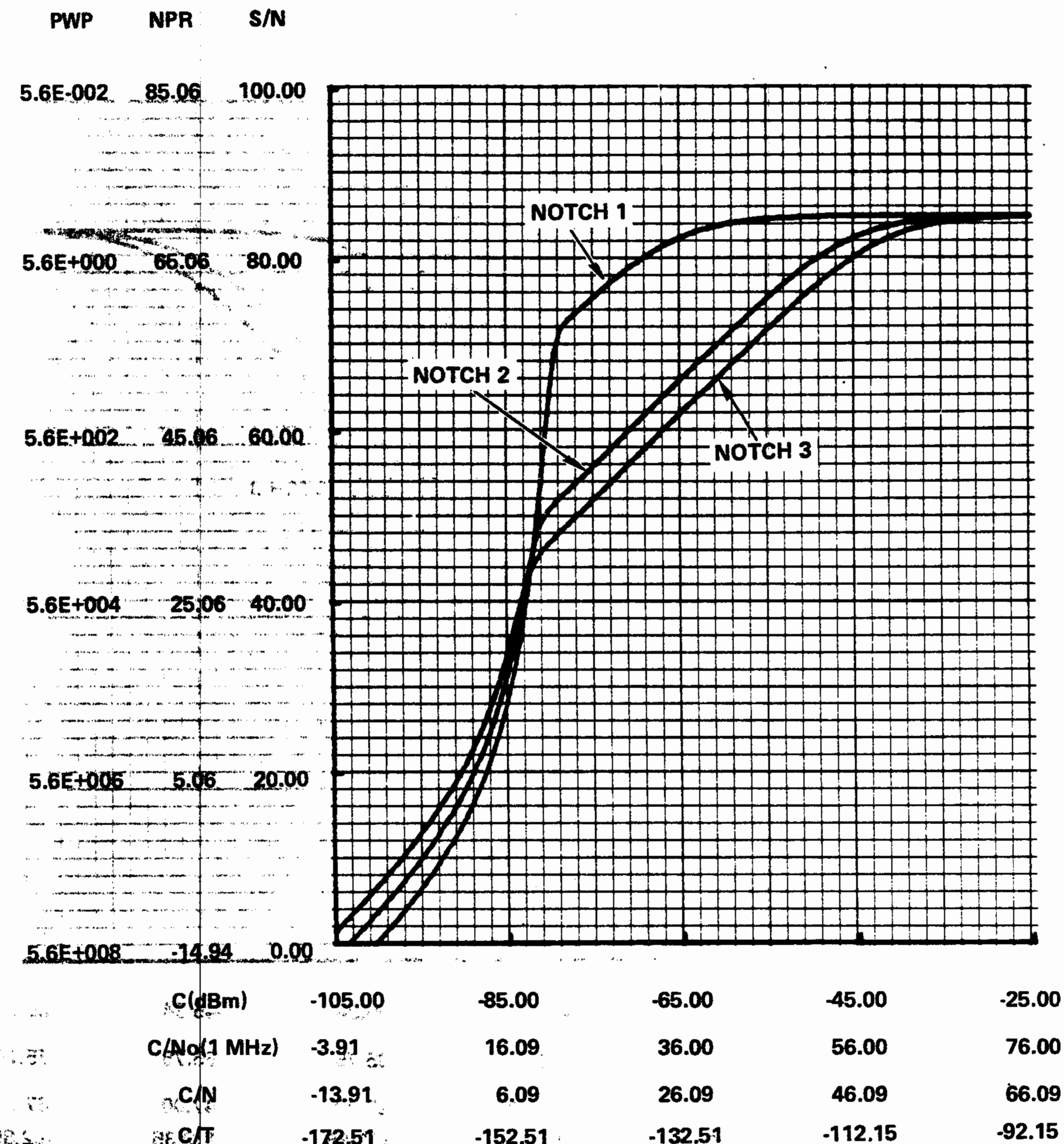
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	5308336.25
LINEARITY BW	2161879.12
T. T. DEVIATION (RMS)	223000
P. E.	YES
P. S.	NO

Figure 13. S/N vs. C/N for 132 Channel Loading



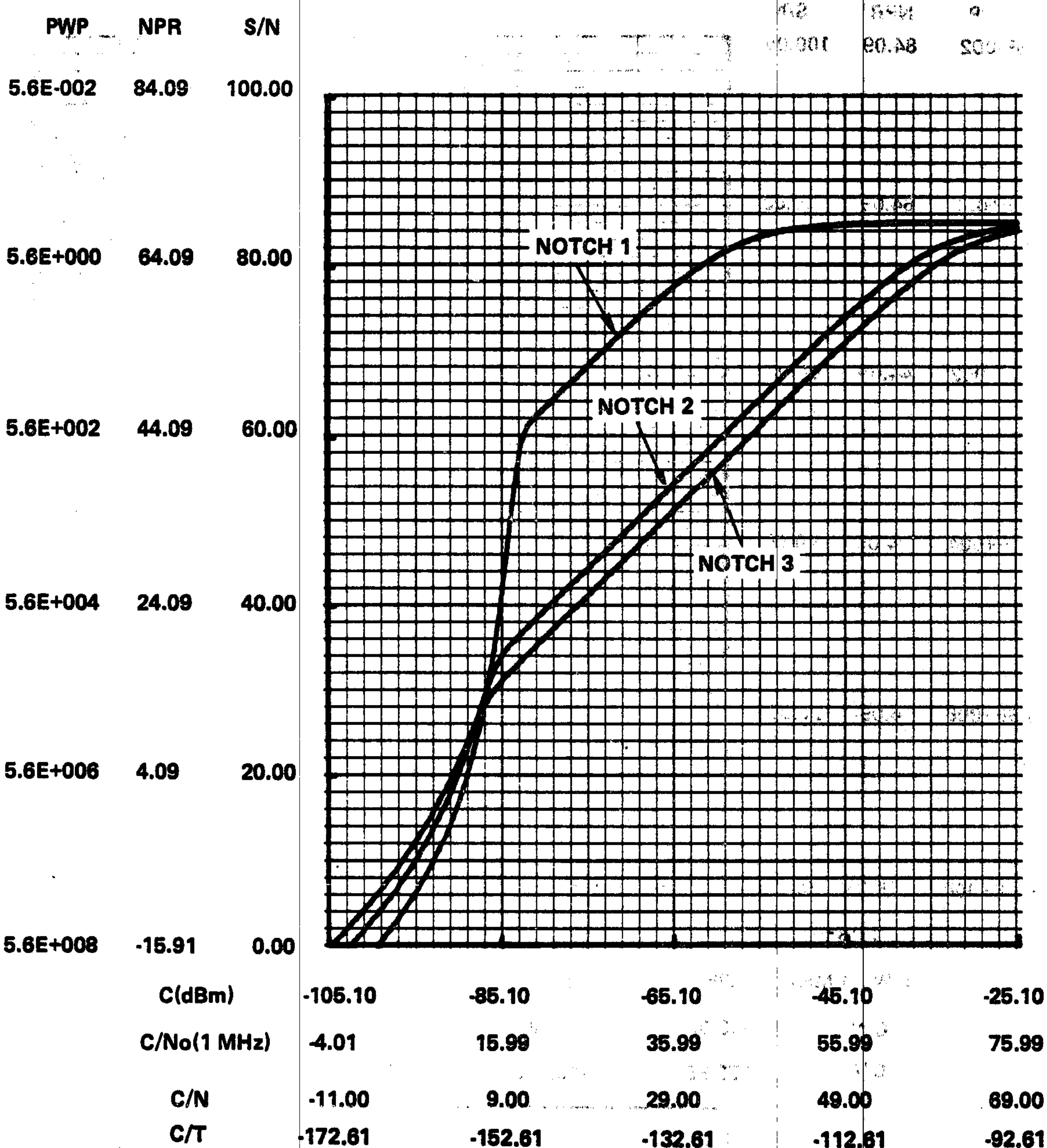
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	8191279.05
LINEARITY BW	2886041.92
T. T. DEVIATION (RMS)	376000
P. E.	YES
P. S.	NO

Figure 14. S/N vs. C/N for 132 Channel Loading



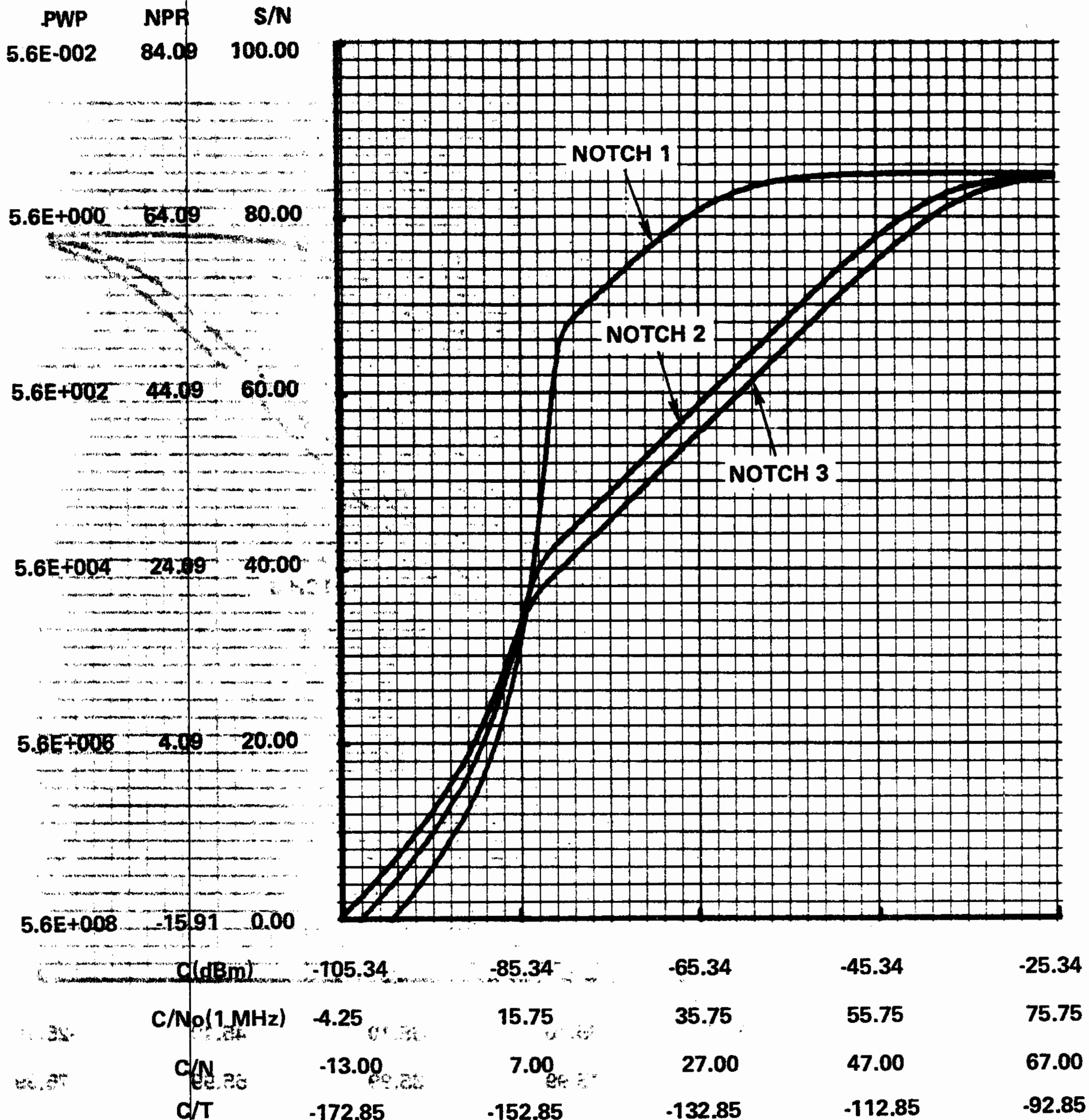
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	132
NOTCH 1	16000
NOTCH 2	240000
NOTCH 3	534000
OCCUPIED BW	9288788.28
LINEARITY BW	3141628.79
T. T. DEVIATION (RMS)	430000
P. E.	YES
P. S.	NO

Figure 15. S/N vs. C/N for 132 Channel Loading



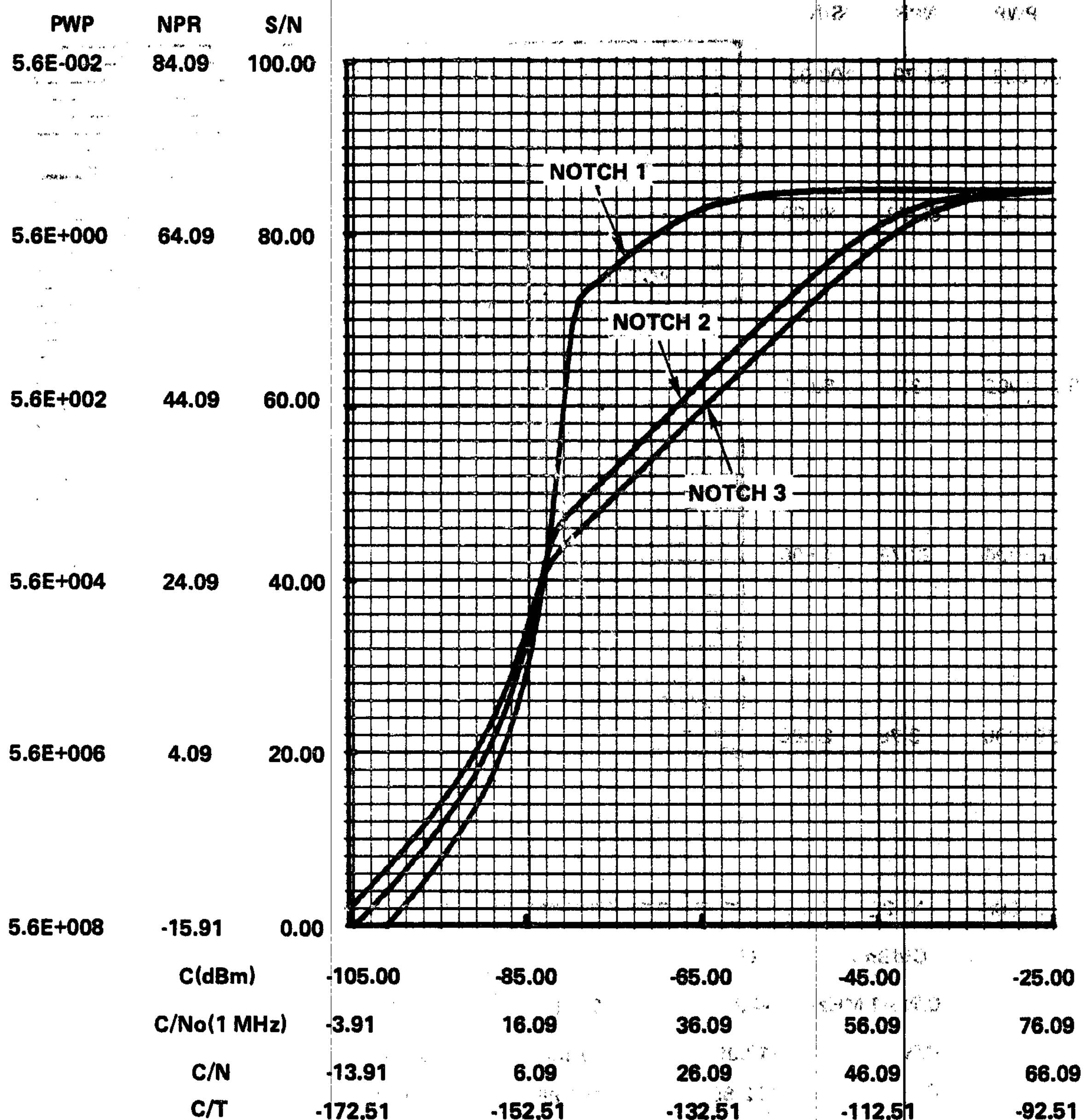
N. F. (dB)	13
BW (I. F.)	5000000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	5254032.76
LINEARITY BW	2516853.83
T.T. DEVIATION (RMS)	180000
P. E.	YES
P. S.	NO

Figure 16. S/N vs. C/N for 192 Channel Loading



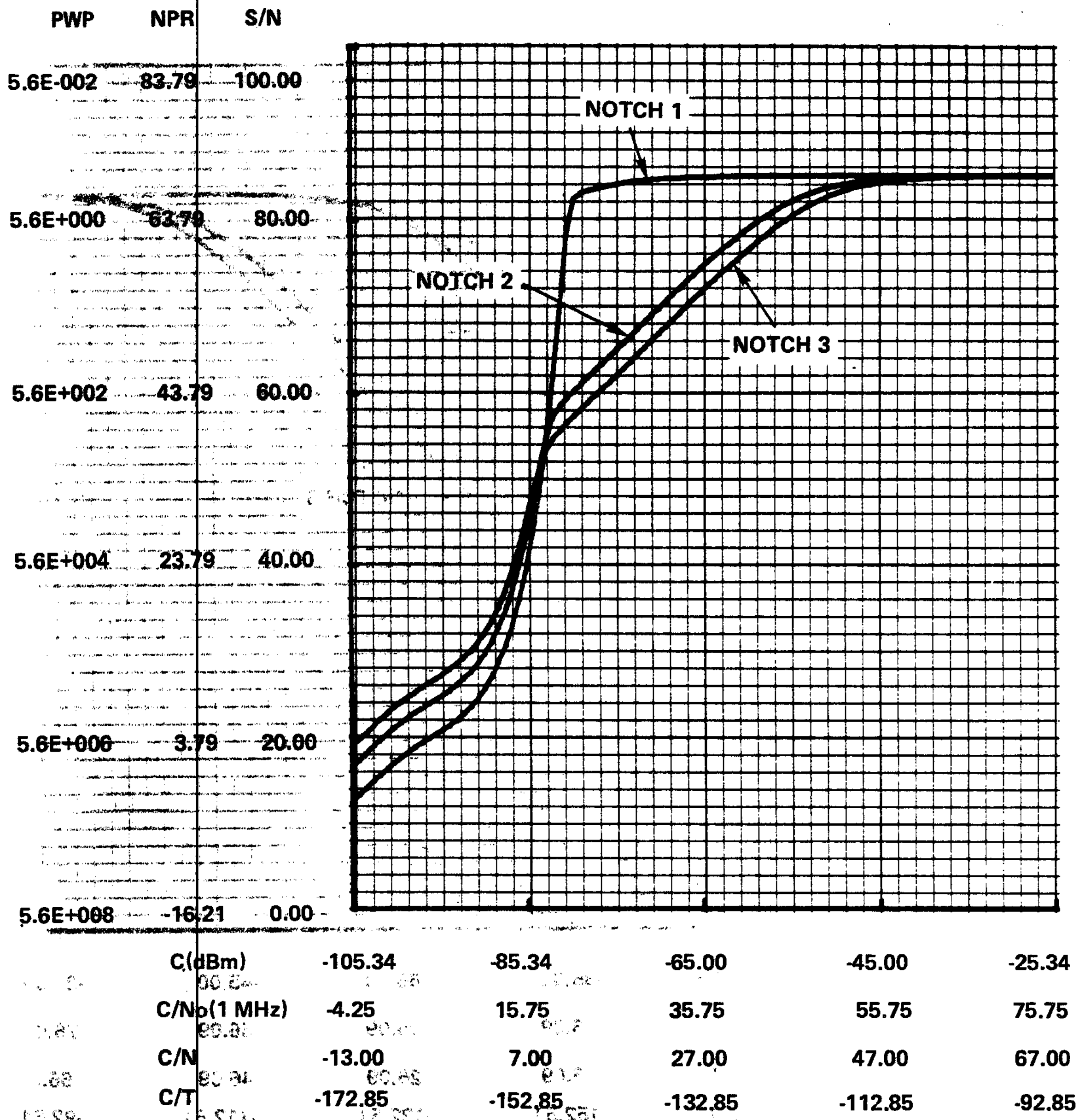
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	7630194.05
LINEARITY BW	3113518.81
T. T. DEVIATION (RMS)	297000
P. E.	YES
P/S.	NO

Figure 17. S/N vs. C/N for 192 Channel Loading



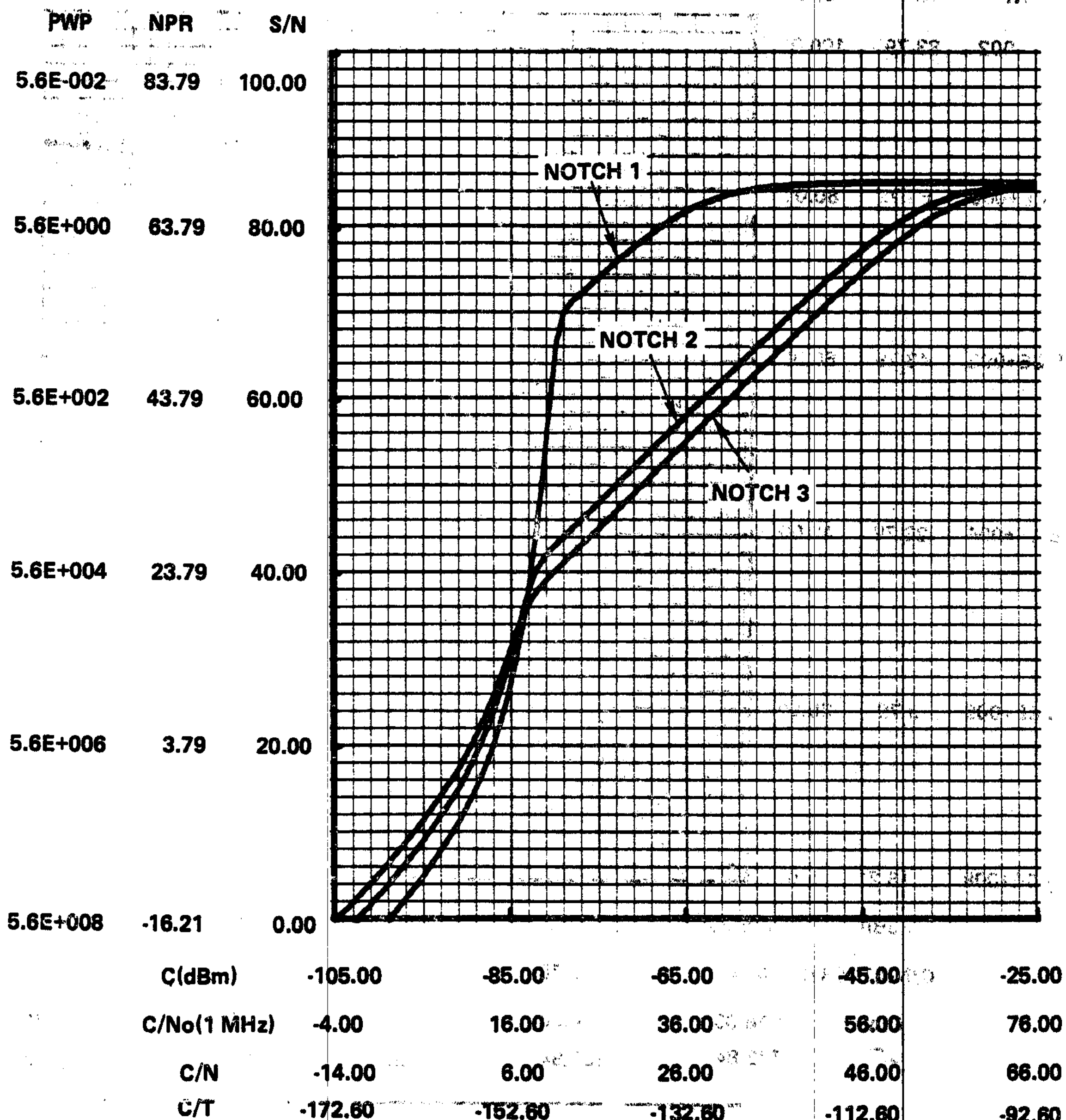
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	192
NOTCH 1	16000
NOTCH 2	394000
NOTCH 3	770000
OCCUPIED BW	10879645.38
LINEARITY BW	3929744.44
T. T. DEVIATION (RMS)	457000
P. E.	YES
P. S.	NO

Figure 18. S/N vs. C/N for 192 Channel Loading



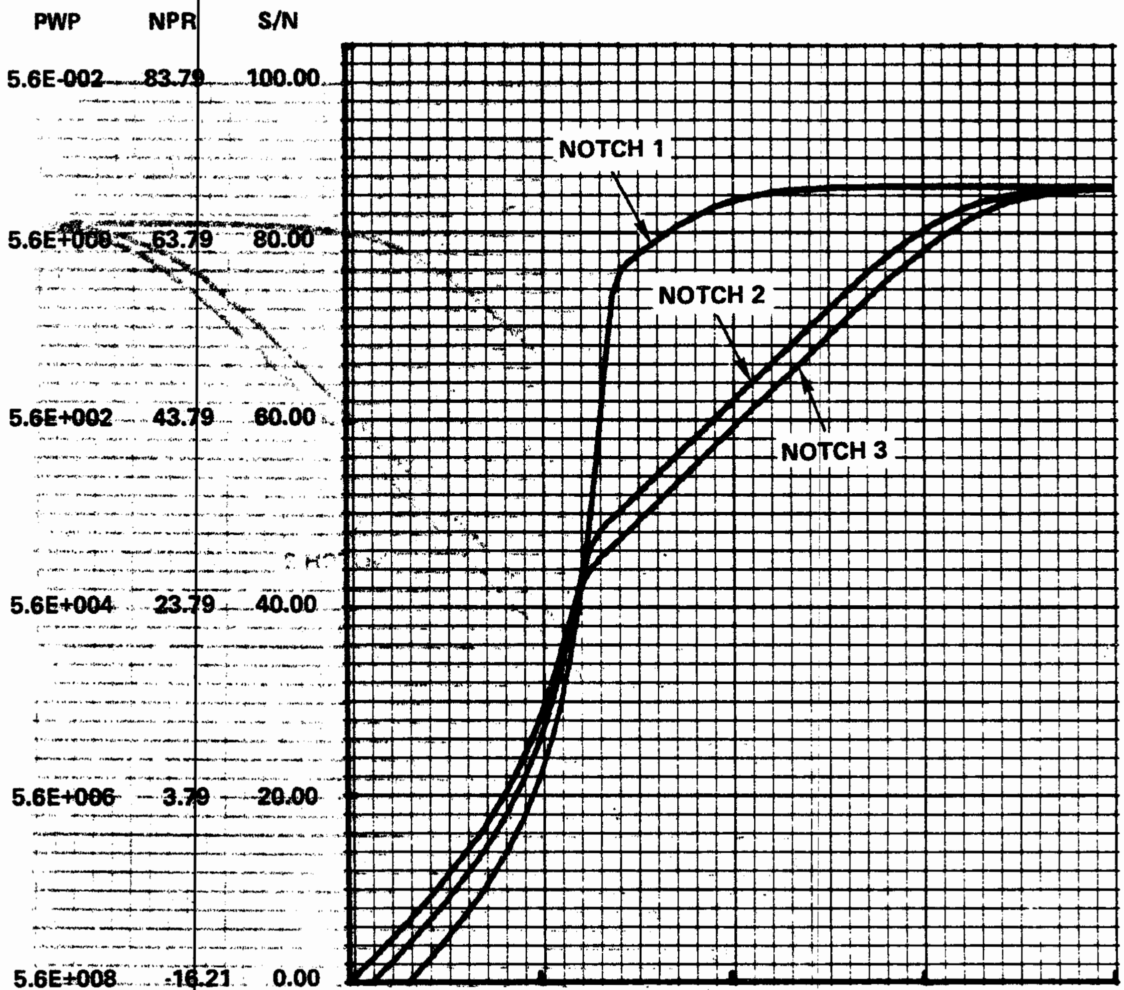
N. F. (dB)	13
BW (I. F.)	7500000
CHANNEL LOADING	252
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1002000
OCCUPIED BW	68529487.76
LINEARITY BW	16769641.45
T. T. DEVIATION (RMS)	2600000
P. E.	YES
P. E.	NO

Figure 19. S/N vs. C/N for 252 Channel Loading



N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	252
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1002000
OCCUPIED BW	10137012.85
LINEARITY BW	4111618.63
T.T. DEVIATION (RMS)	358000
P. E.	YES
P. S.	NO

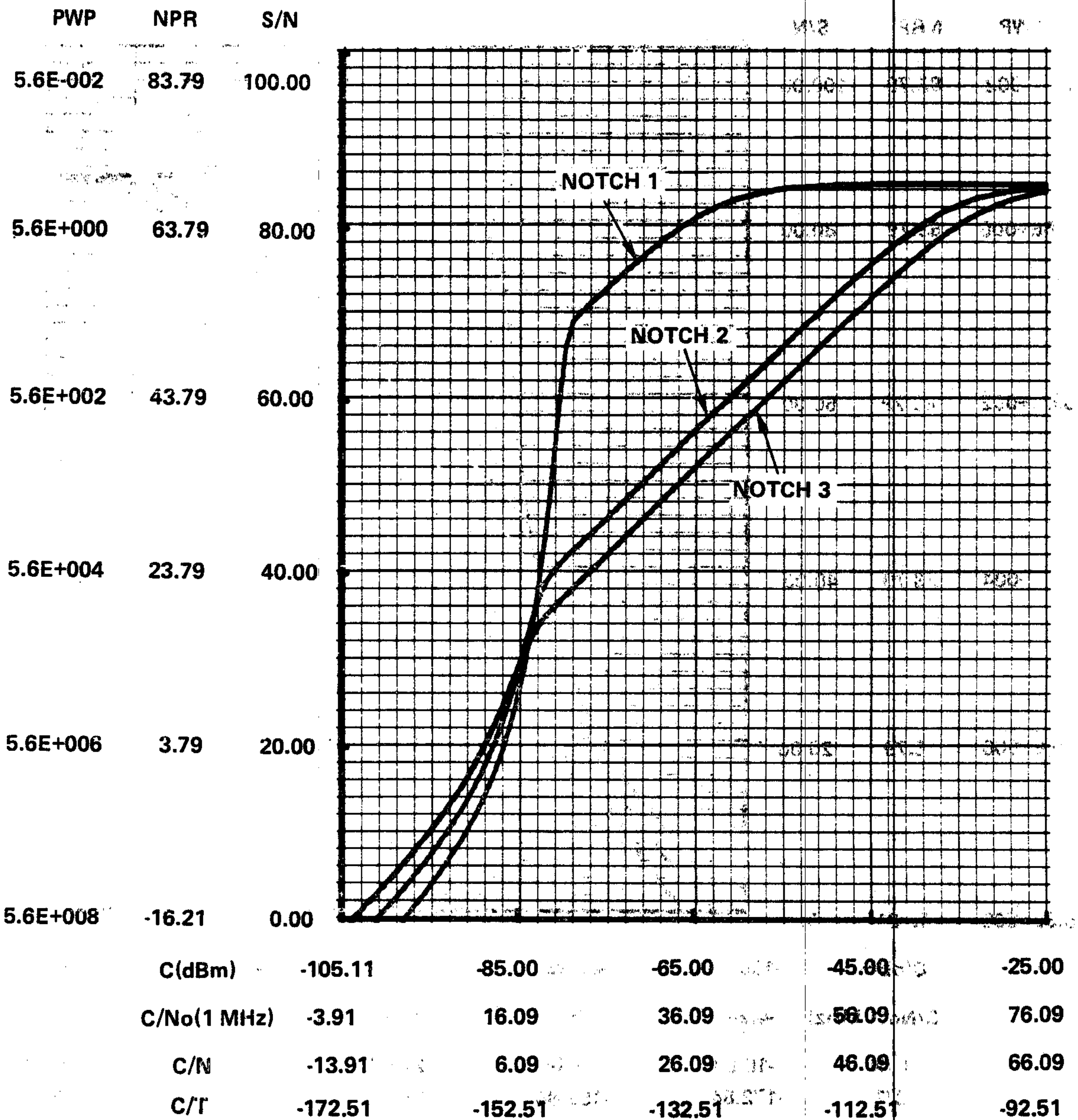
Figure 20. S/N vs. C/N for 252 Channel Loading



C(dBm)	-105.33	-85.33	-65.33	-45.33	-25.33
C/No(1MHz)	-4.24	15.76	35.76	55.76	75.76
C/N	-16.00	4.00	24.00	44.00	64.00
C/T	-172.84	-152.84	-132.84	-112.84	-92.84

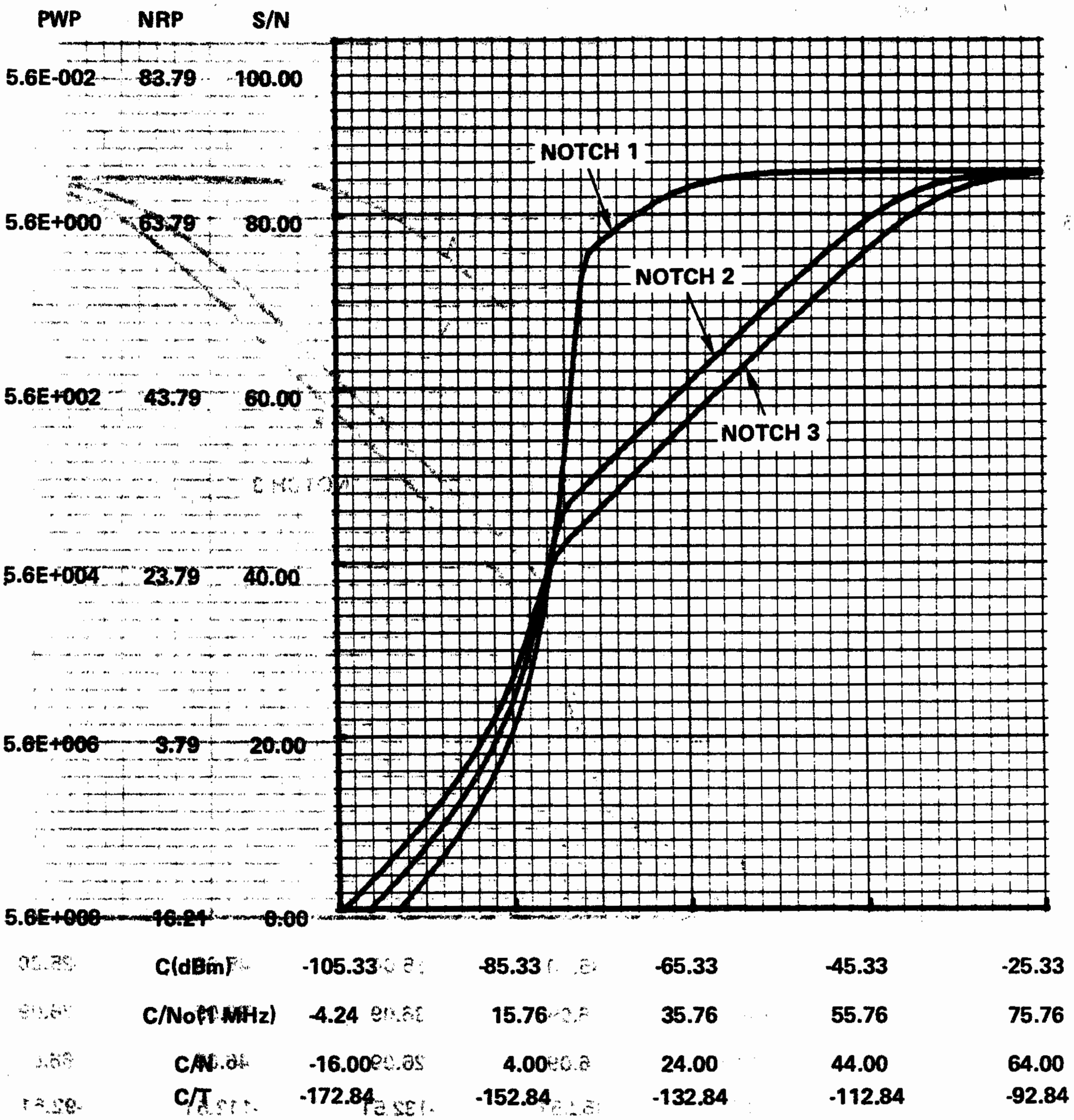
N. F. (dB)	13
BW (I. F.)	1.5E+7
CHANNEL LOADING	252
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1002000
OCCUPIED BW	15850382.17
LINEARITY BW	5348062.43
T. T. Deviation (RMS)	577000
P. E.	YES
P. S.	NO

Figure 21. S/N vs. C/N for 252 Channel Loading



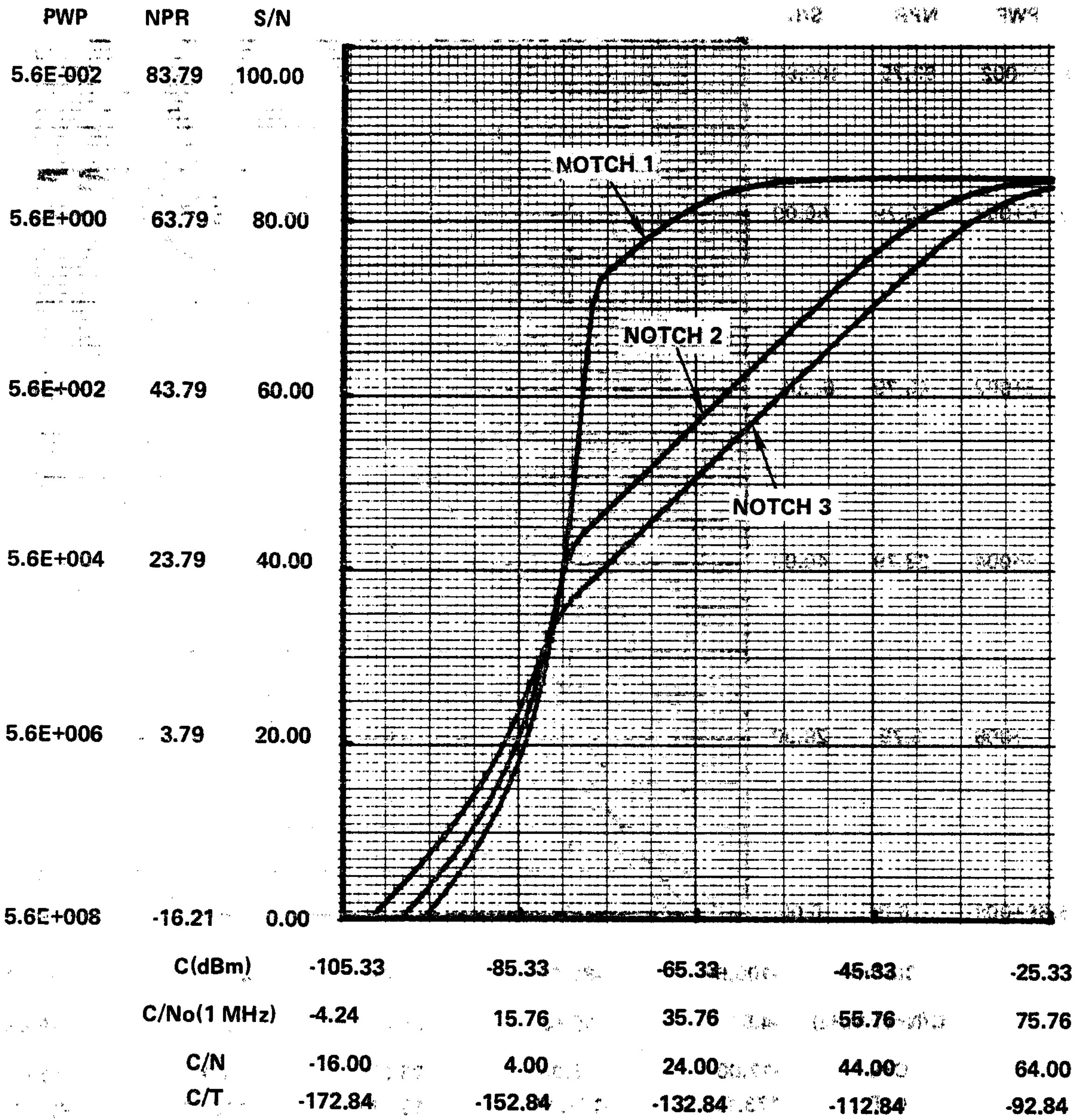
N. F. (dB)	13
BW (I. F.)	10000000
CHANNEL LOADING	312
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1248000
OCCUPIED BW	18585478.65
LINEARITY BW	4592683.33
T. T. DEVIATION (RMS)	320000
P. E.	YES
P. S.	NO

Figure 22. S/N vs. C/N for 312 Channel Loading



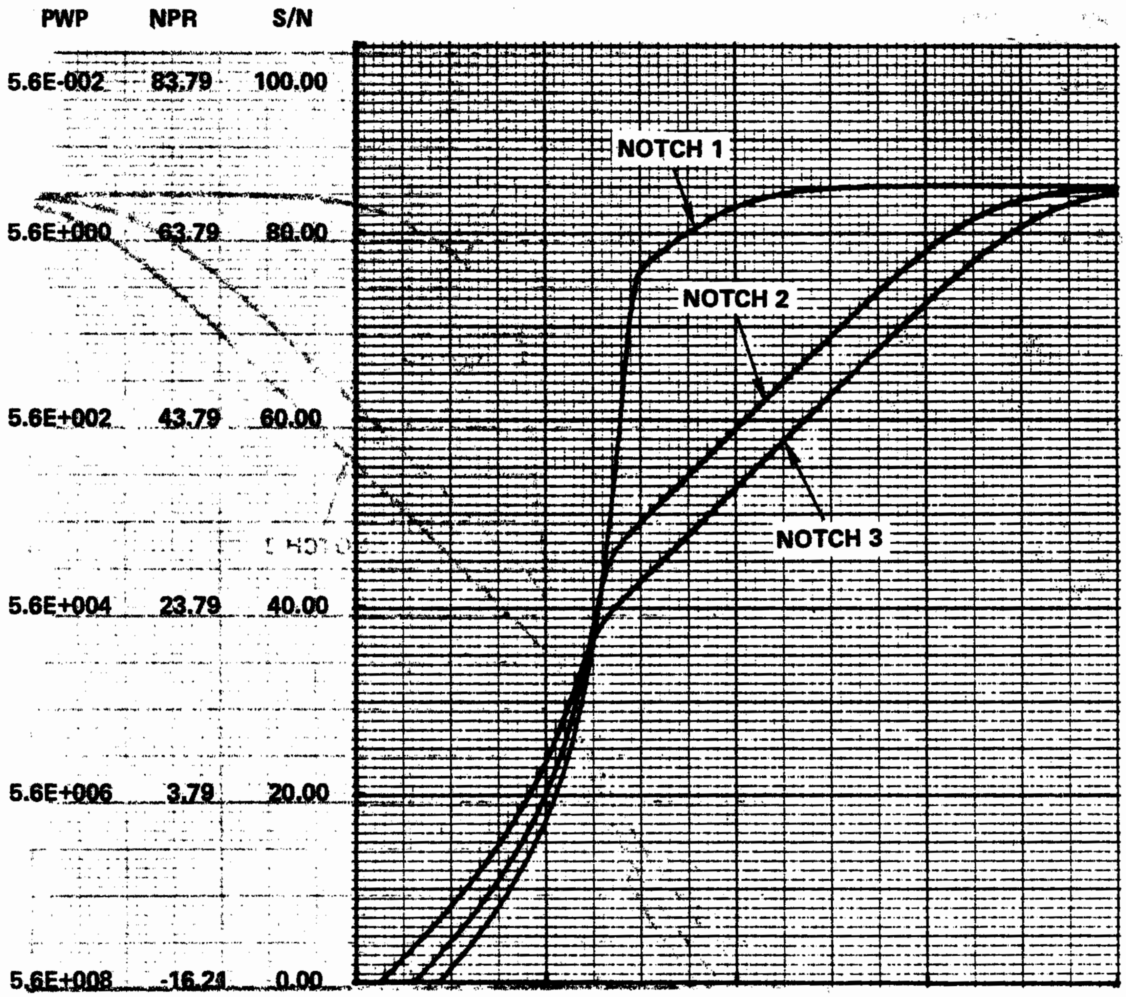
N. F. (dB)	13
BW (I. F.)	1.5E+7
CHANNEL LOADING	312
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1240000
OCCUPIED BW	16237652.95
LINEARITY	6012445.83
T.T. DEVIATION (RMS)	546000
P. E.	YES
P. S.	NO

Figure 23. S/N vs. C/N for 312 Channel Loading



N. F. (dB)	13
BW (I. F.)	1.5E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	15367325.15
LINEARITY BW	6530659.65
T. T. DEVIATION (RMS)	401000
P. E.	YES
P. S.	NO

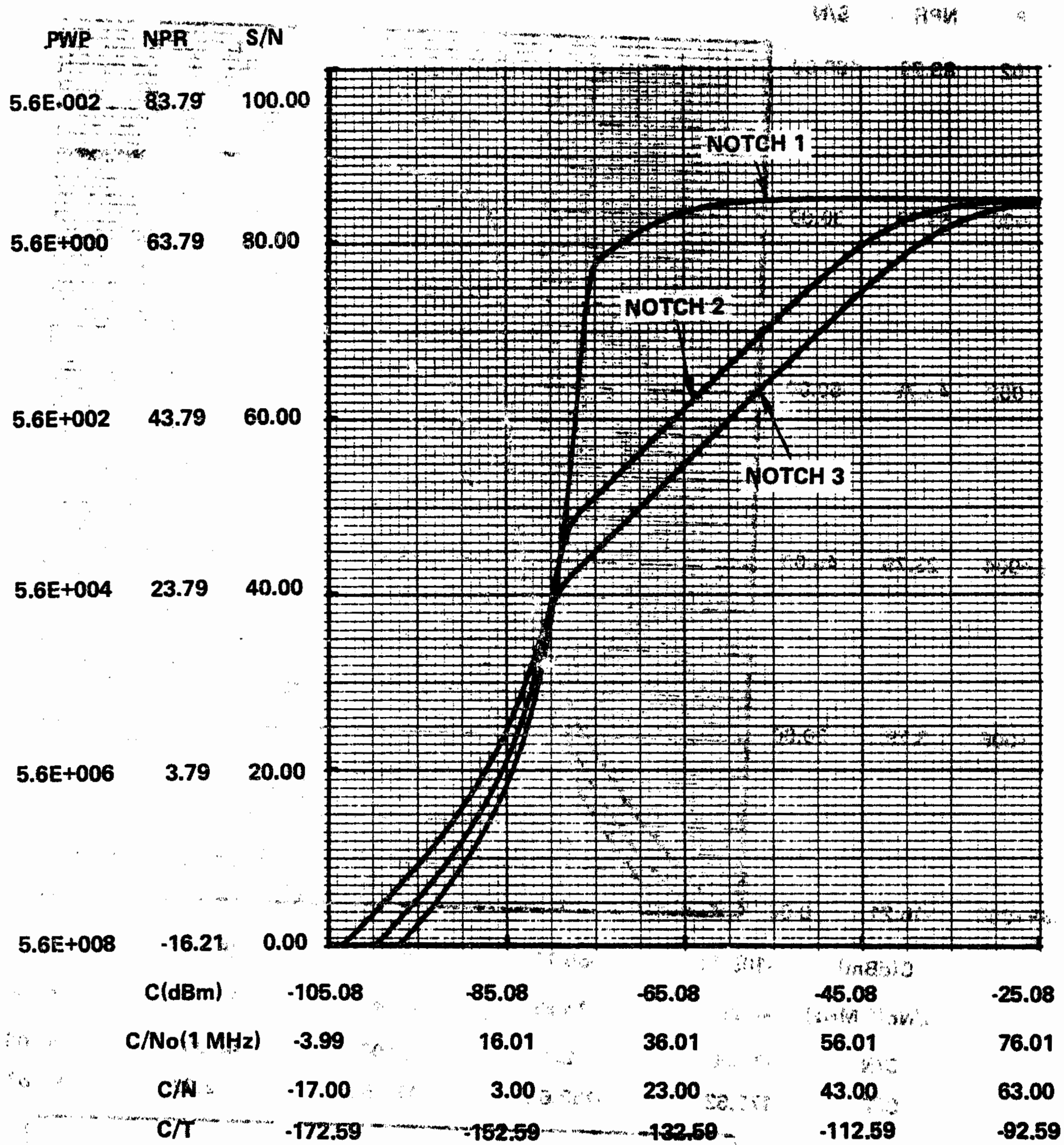
Figure 24. S/N vs. C/N for 432 Channel Loading



C(dBm)	-105.66	-85.66	-65.66	-45.66	-25.66
C/No(1 MHz)	-4.57	15.43	35.43	55.43	75.43
C/N	-17.00	3.00	23.00	43.00	63.00
C/T	-173.17	-153.17	-133.17	-113.17	-93.17

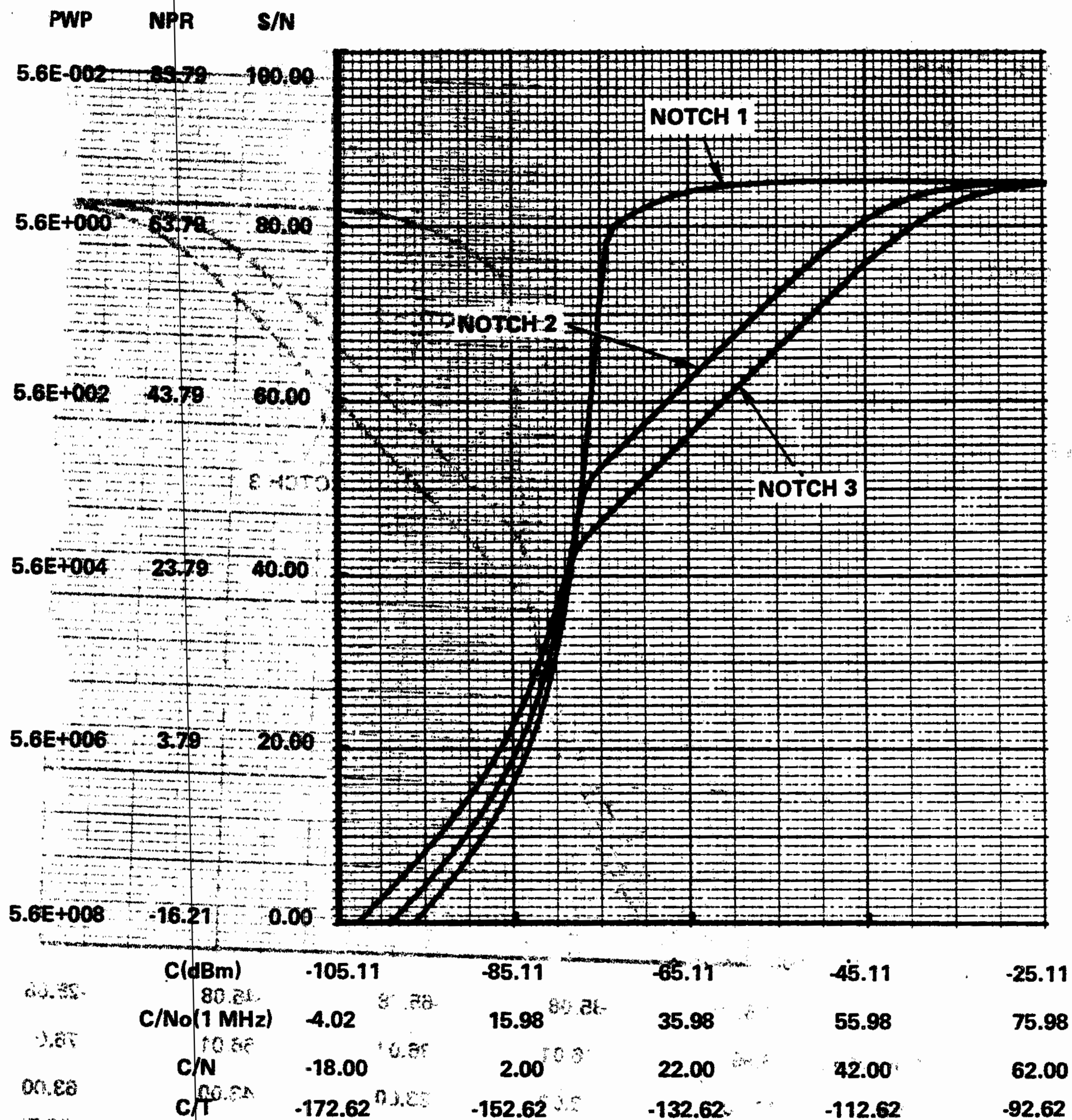
N. F. (dB)	13
BW (I. F.)	1.75E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	18781059.10
LINEARITY BW	7388151.21
T. T. DEVIATION (RMS)	517000
P. E.	YES
P. S.	NO

Figure 25. S/N vs. C/N for 432 Channel Loading



N. F. (dB)	13
BW (I. F.)	2.0E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	21694504.46
LINEARITY BW	8119975.91
T. T. Deviation (RMS)	616000
P. E.	YES
P. S.	NO

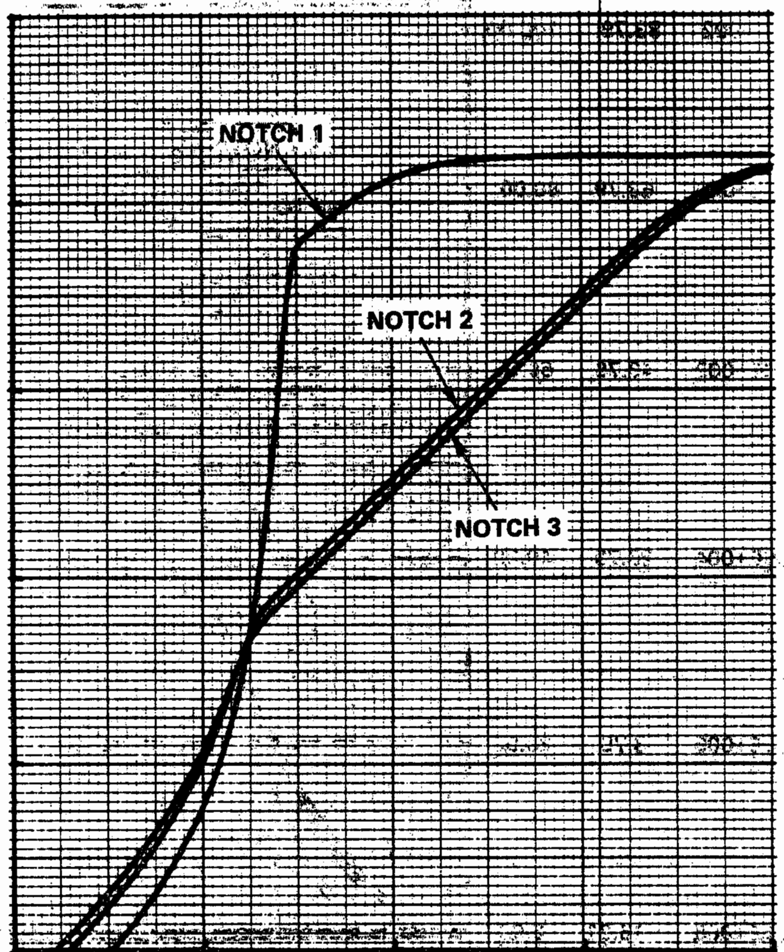
Figure 26. S/N vs. C/N for 432 Channel Loading



N. F. (dB)	13
BW (I. F.)	2.5E+7
CHANNEL LOADING	432
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1730000
OCCUPIED BW	25019952.20
LINEARITY BW	8955290.98
T. T. DEVIATION (RMS)	729000
P. E.	YES
P. S.	NO

Figure 27. S/N vs. C/N for 432 Channel Loading

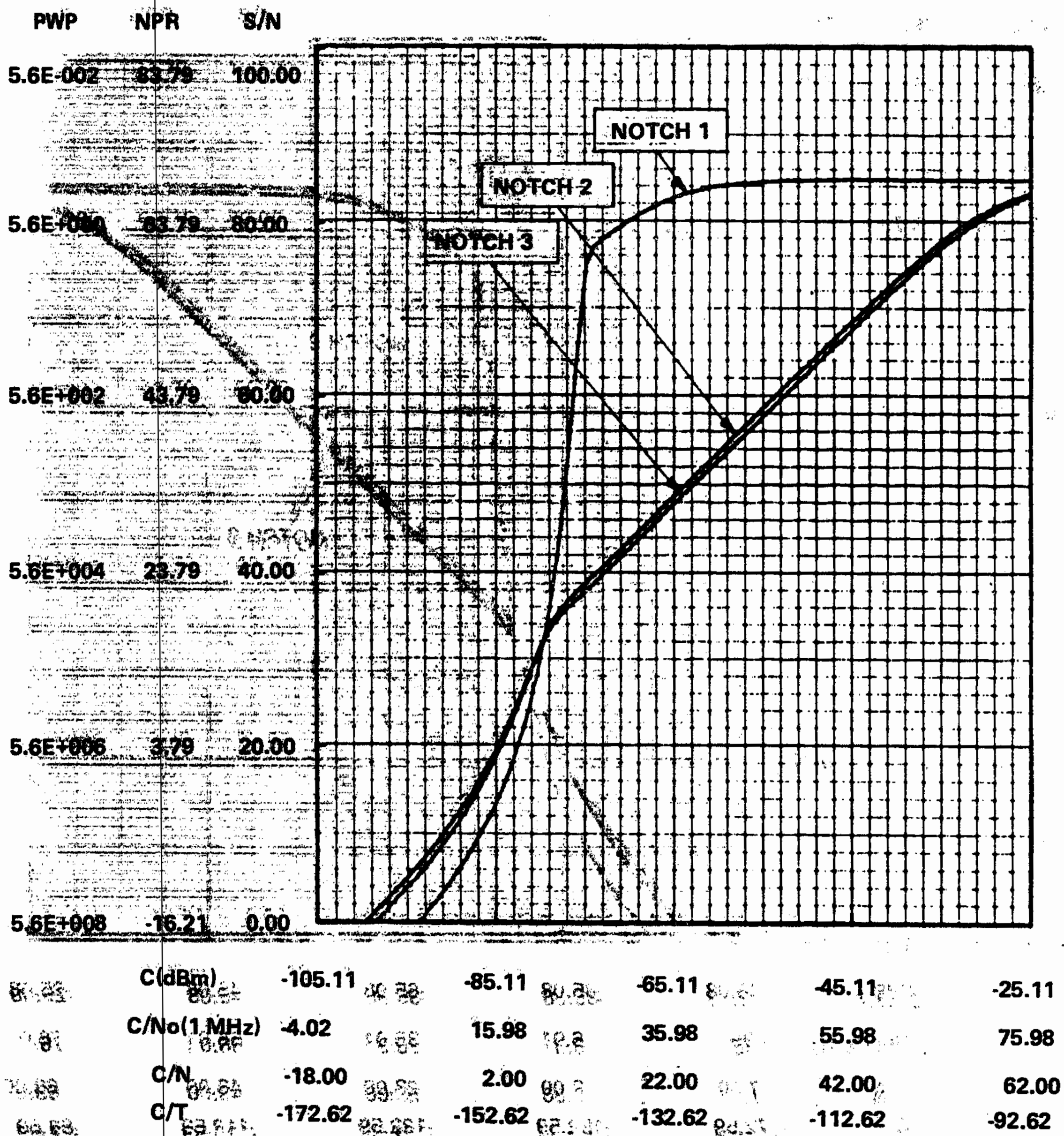
PWP	NPR	S/N
5.6E-002	83.79	100.00
5.6E+000	63.79	80.00
5.6E+002	43.79	60.00
5.6E+004	23.79	40.00
5.6E+006	3.79	20.00
5.6E+008	-16.21	0.00



C(dBm)	-105.08	-85.08	-65.08	-45.08	-25.08
C/No(1 MHz)	-3.99	16.01	36.01	56.01	76.01
C/N	-17.00	3.00	23.00	43.00	63.00
C/T	-172.59	-152.59	-132.59	-112.59	-92.59

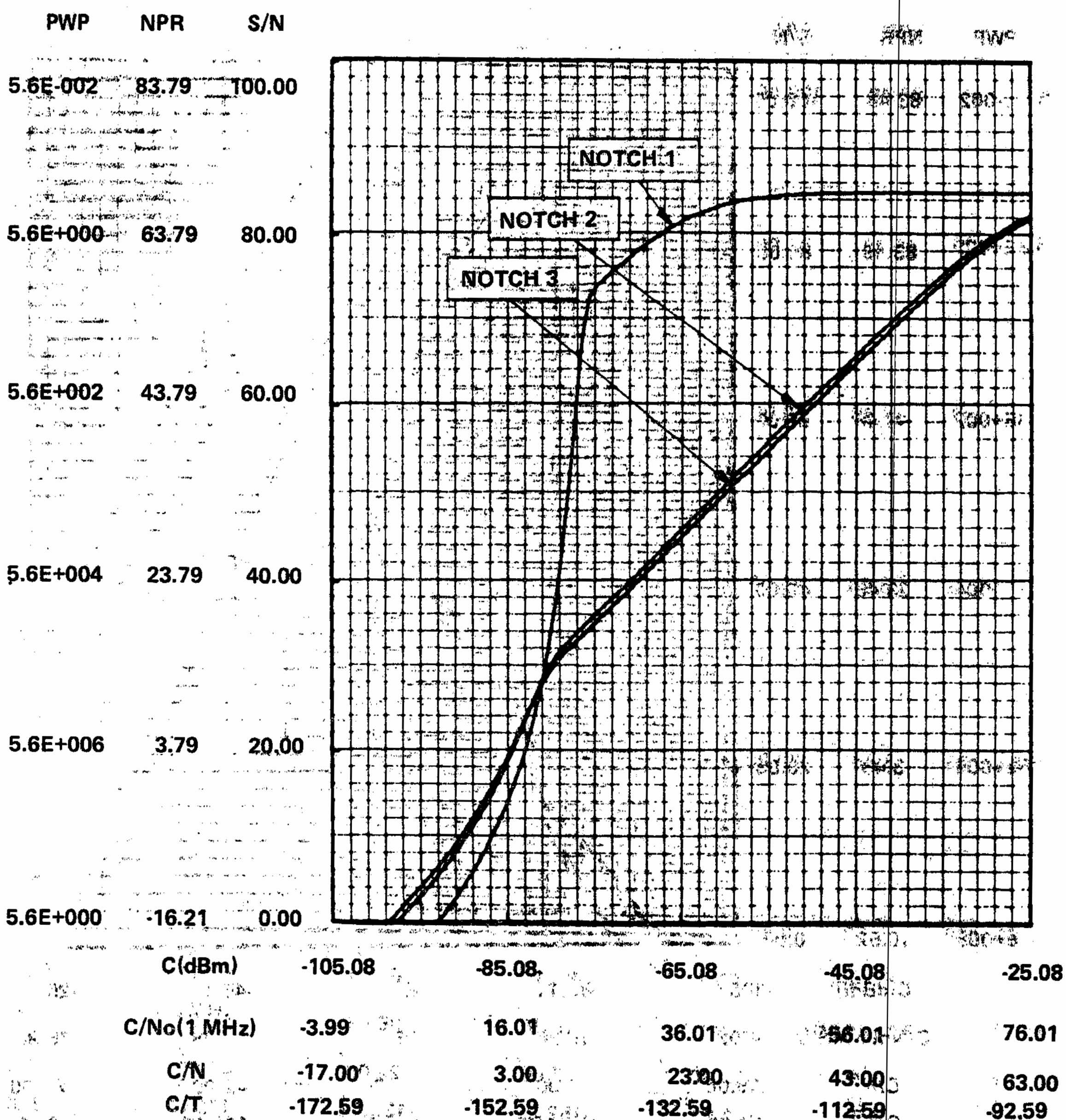
N. F. (dB)	13
BW (I. F.)	2.0E+7
CHANNEL LOADING	612
NOTCH 1	16000
NOTCH 2	1730000
NOTCH 3	2438000
OCCUPIED BW	21014806.82
LINEARITY BW	9054488.92
T. T. DEVIATION (RMS)	456000
P. E.	YES
P. S.	NO

Figure 28. S/N vs. C/N for 612 Channel Loading



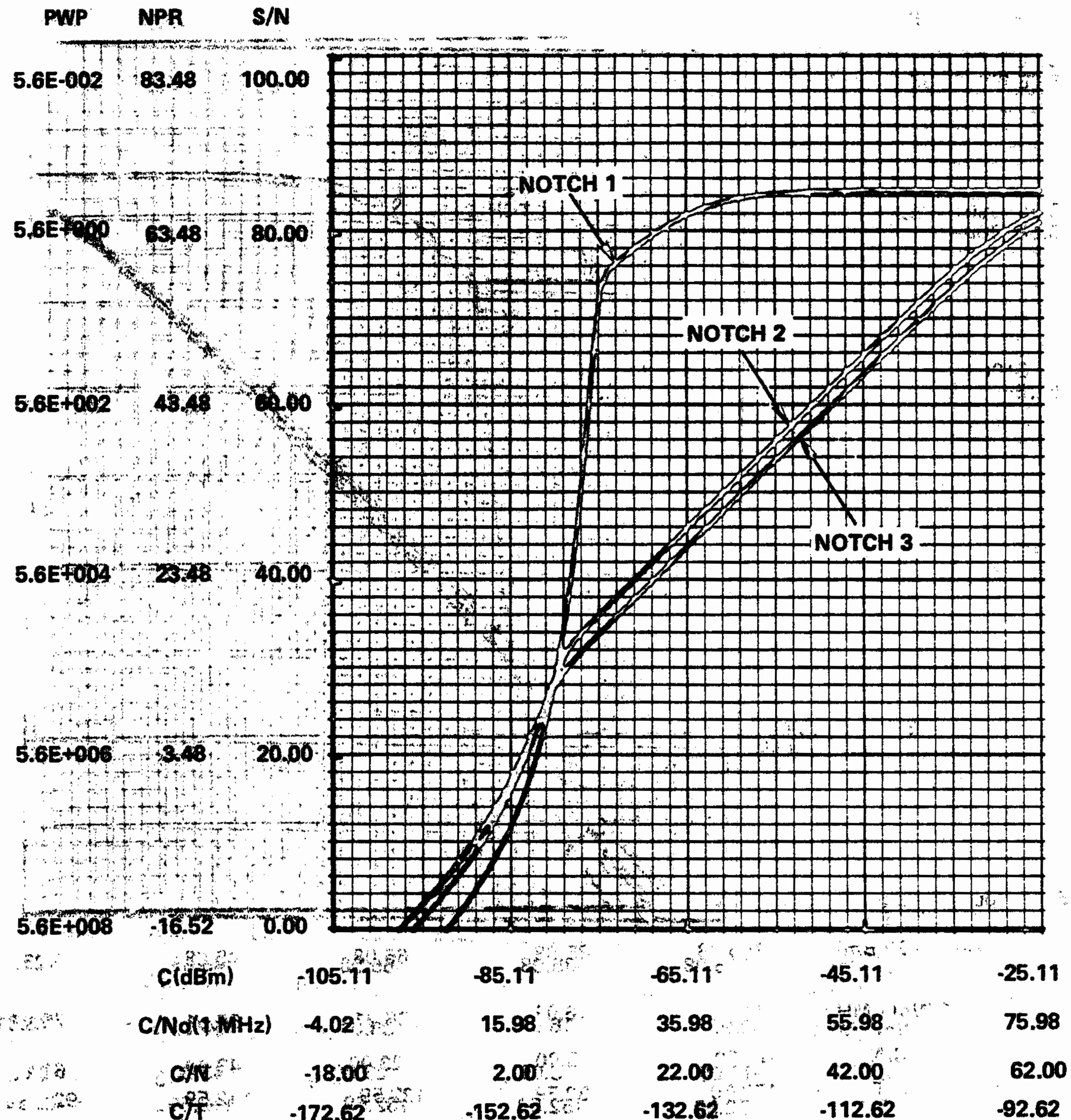
N. F. (dB)	13
BW (I.F.)	2.5E+7
CHANNEL LOADING	792
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3150000
OCCUPIED BW	26401905.15
LINEARITY BW	11512912.82
T. T. Deviation (RMS)	499000
P. E.	YES
P. S.	NO

Figure 29. S/N vs. C/N for 792 Channel Loading



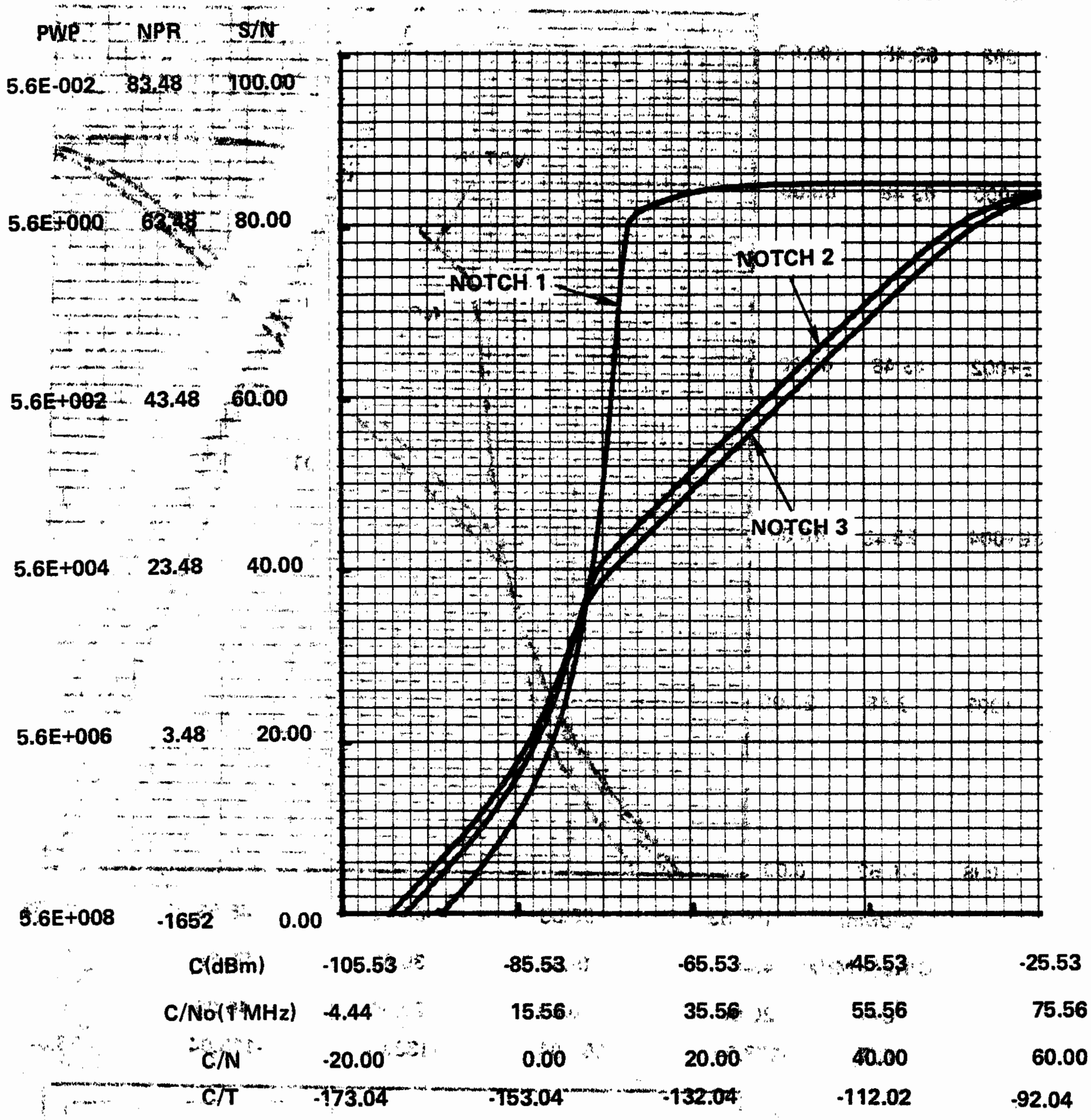
N. F. (dB)	13
BW (I. F.)	2.0 E+7
CHANNEL LOADING	792
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3150000
OCCUPIED BW	20703826.52
LINEARITY BW	10081619.57
T. T. DEVIATION (RMS)	356000
P. E.	YES
P. S.	NO

Figure 30. S/N vs. C/N for 792 Channel Loading



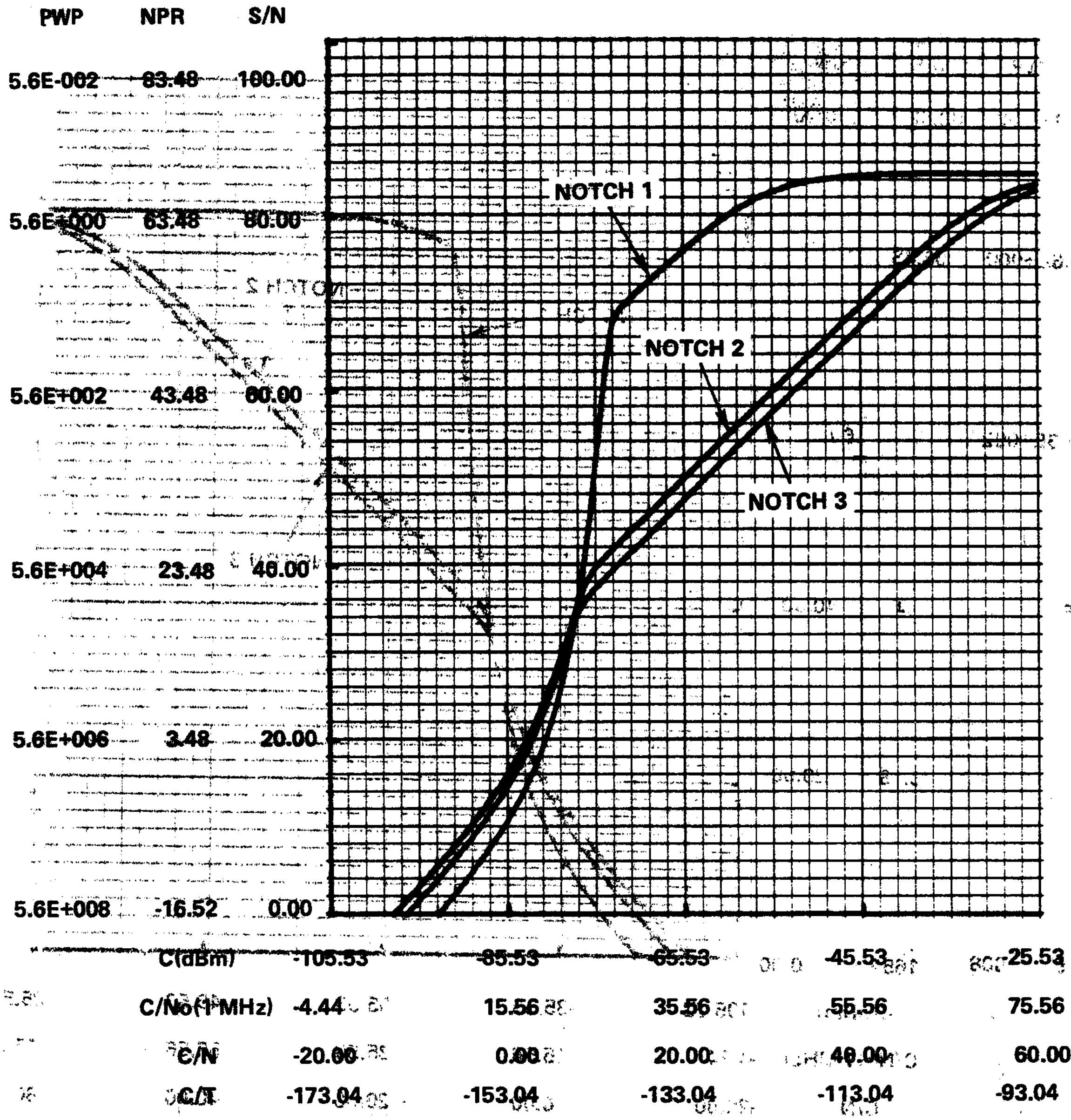
N. F. (dB)	13
BW (I. F.)	2.5E+7
CHANNEL LOADING	972
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3886000
OCCUPIED BW	26676275.72
LINEARITY BW	13123783.75
T. T. DEVIATION (RMS)	410000
P. E.	YES
P. S.	NO

Figure 31. S/N vs. C/N for 972 Channel Loading



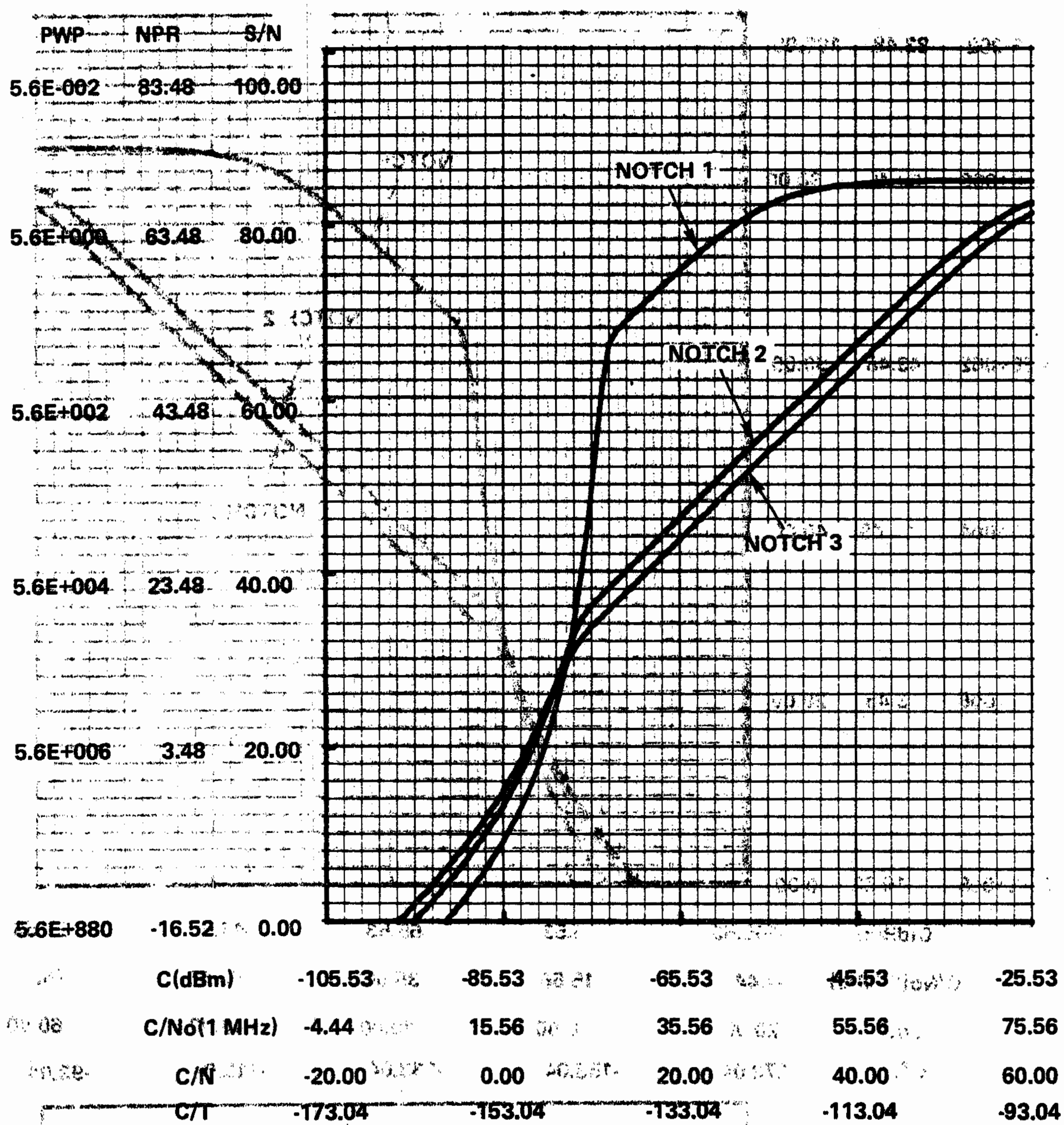
N. F. (dB)	13
BW (I. F.)	3.6E+7
CHANNEL LOADING	972
NOTCH 1	16000
NOTCH 2	2438000
NOTCH 3	3886000
OCCUPIED BW	43980375.44
LINEARITY BW	17470378.94
T. T. DEVIATION (RMS)	802000
P. E.	YES
P. S.	NO

Figure 32. S/N vs. C/N for 972 Channel Loading



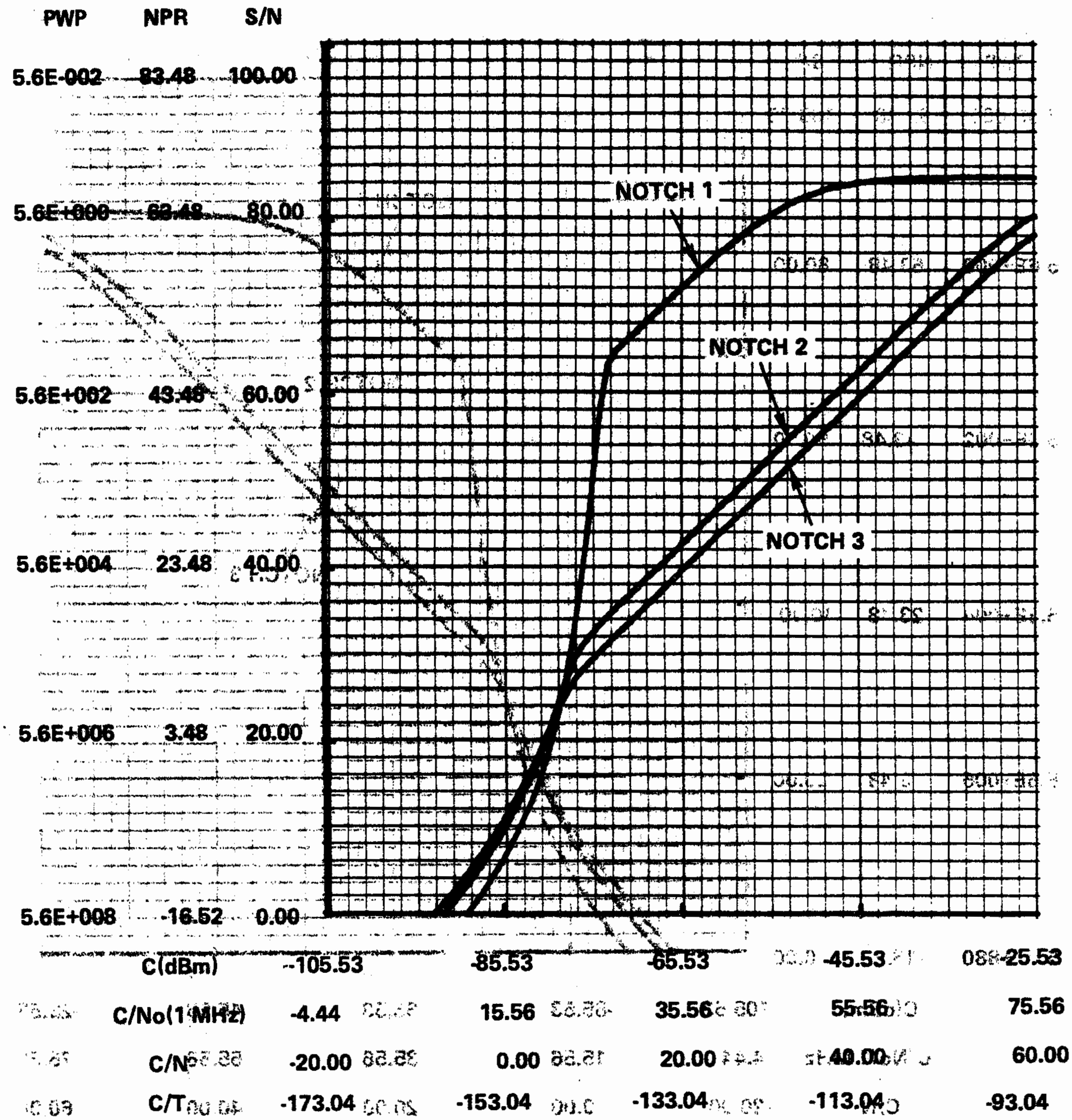
N. F. (dB)	13
BW (I. F.)	3.6E+7
CHANNEL LOADING	1092
NOTCH 1	70000
NOTCH 2	2438000
NOTCH 3	3886000
OCCUPIED BW	42528490.03
LINEARITY BW	17968312.21
T. T. DEVIATION (RMS)	701000
P. E.	YES
P. S.	NO

Figure 33. S/N vs. C/N for 1092 Channel Loading



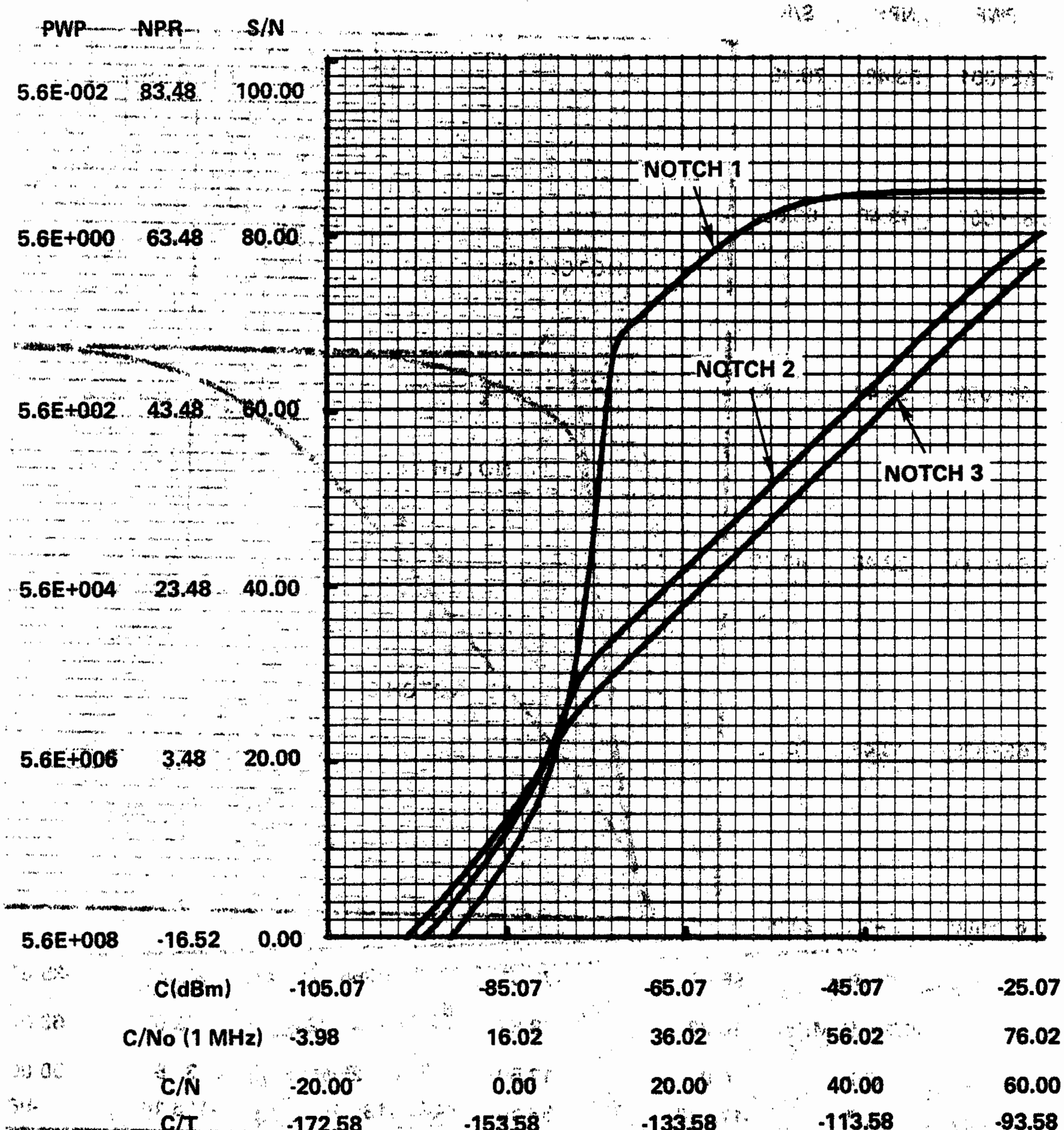
N. F.	13
BW (I. F.)	3.6E+7
CHANNEL LOADING	1332
NOTCH 1	70000
NOTCH 2	3150000
NOTCH 2	5340000
OCCUPIED BW	42381626.03
LINEARITY BW	19512910.99
T. T. DEVIATION (RMS)	591000
P. E.	YES
P. S.	NO

Figure 34. S/N vs. C/N for 1332 Channel Loading



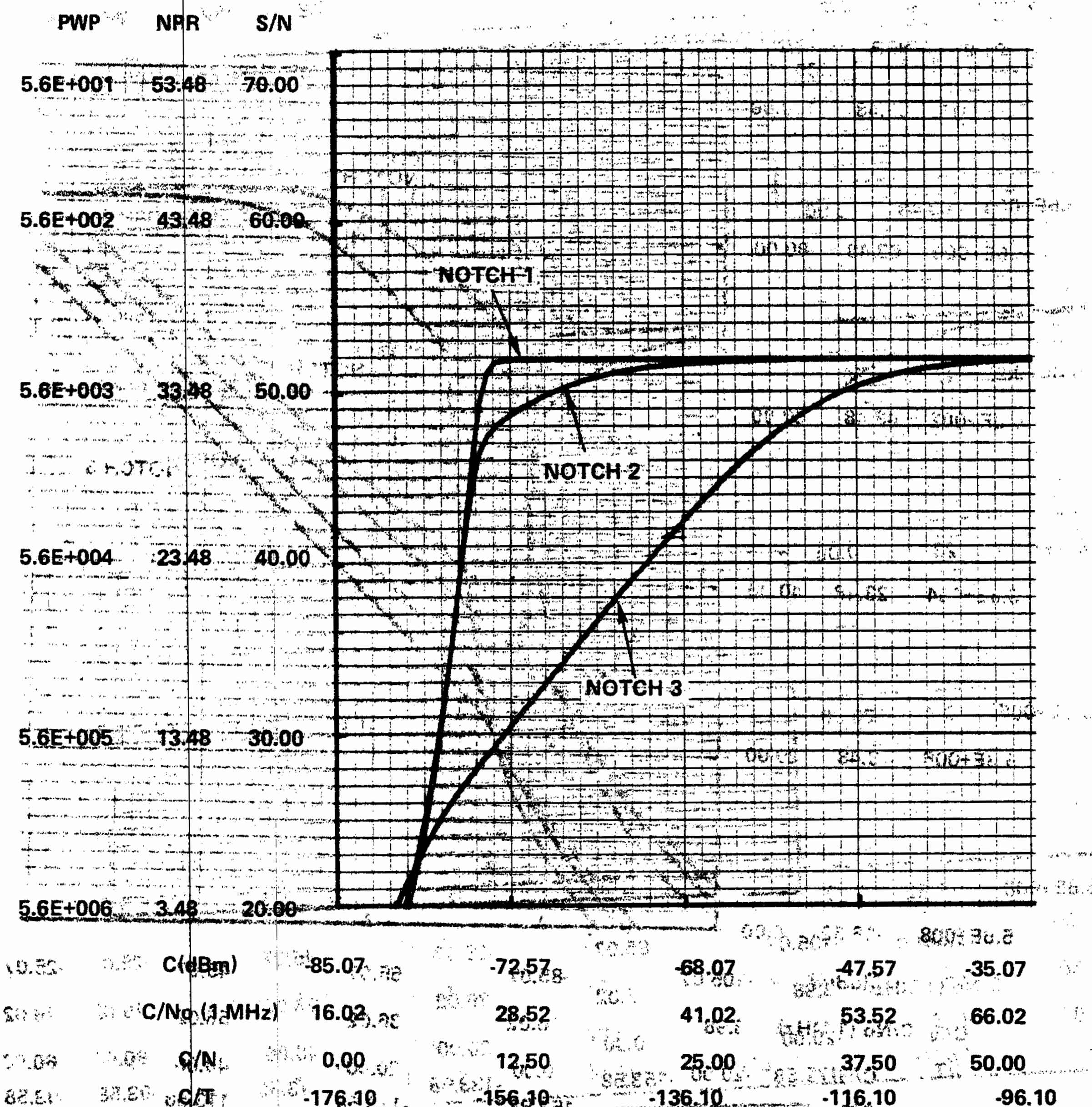
N. F. (dB)	13
BW (I. F.)	3.6E+7
CHANNEL LOADING	1872
NOTCH 1	70000
NOTCH 2	3150000
NOTCH 3	5340000
OCCUPIED BW	42261879.21
LINEARITY BW	23041182.99
T. T. DEVIATION (RMS)	419000
P. E.	YES
P. S.	NO

Figure 35. S/N vs. C/N for 1872 Channel Loading



N. F. (dB)	13
BW (I. F.)	4.0E+7
CHANNEL LOADING	2892
NOTCH 1	70000
NOTCH 2	5340000
NOTCH 3	1.17E+7
OCCUPIED BW	72606005.23
LINEARITY BW	37312729.99
T. T. DEVIATIO N (RMS)	619000
P. E.	YES
P. S.	NO

Figure 36. S/N vs. C/N for 2892 Channel Loading



N. F; (dB)	13
BW (I. F.)	4.0E+7
CHANNEL LOADING	2892
NOTCH 1	16000
NOTCH 2	534000
NOTCH 3	1.17E+7
OCCUPIED BW	72590776.66
LINEARITY BW	37308904.75
T. T. Deviation (RMS)	618800
P. E.	YES
P. S.	NO

Figure 37. S/N vs. C/N for 2892 Channel Loading

Service Arcs for North, Central and South America

(Degrees Longitude)