THE BELL SYSTEM

TECHNICAL JOURNAL

Volume 47 September 1968 Number 7

TD-3 Microwave Radio Relay System

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(Manuscript received November 15, 1967)

This paper describes the over-all system and, briefly, its major components. The system objectives and the design plan used to achieve these objectives are covered. Results of laboratory and field tests are presented. The paper concludes with a discussion of additional development work and system studies now under way or planned.

I. GENERAL

By early 1962 TD-2 radio relay routes were carrying more than 80,000 miles of intercity television circuits and about 35,000,000 miles of telephone circuits. These figures represented over 90 per cent of all intercity video and more than 40 per cent of all long distance telephone circuit mileage in the Bell System. Aided by the use of interstitial channels, which had begun about two years earlier,[1] the percentage of total video and telephone long distance mileage provided by TD-2 was increasing rapidly.

Nevertheless, as it was then being used, TD-2 had several shortcomings. It was an electron tube system with the maintenance problems associated with electron tubes. It required a multivoltage power source. With a message load of 600 telephone circuits per radio channel, a full route capacity of 6000 circuits* was realized using the 3700-4200 MHz common carrier frequency band. The TH system,[2] introduced in 1961, provided nearly twice the capacity in a band

* Based on use of 10 working and 2 protection channels on a fully equipped route.

of the same width, 5925-6425 MHz. In addition, the noise performance of the TD-2 system and the reliability as measured by yearly outage time, were considered to be only marginally satisfactory. Therefore it was decided that a new system, the TD-3, should be developed for use in the 3700 to 4200 MHz band. This, and the other articles in this issue, describe the new system.

The total TD-3 development program included several major items. The first to be completed was a new intermediate frequency protection switching system designed for use with both TD-2 and TD-3. Commercial use of this system, the 100A Protection Switching System, started in 1965 on TD-2 routes.[3]

The second item of the development program was new frequency modulation terminal equipment. Like the 100A, the FM terminals were designed for use with both TD-2 and TD-3.[4]

The third development category was comprised of the equipment and facilities for the "radio line." This included the microwave radio transmitting and receiving equipment, power equipment, test facilities, and building and antenna tower arrangements.[5–8]

Of at least equal importance to the development work was the effort spent on obtaining a better theoretical quantitative measure of the effects of several types of design deficiencies on the performance of a system such as TD-3. During the 1950's analytical methods[9, 10] were developed for computing the intermodulation noise resulting from some of the simpler types of transmission deviations which might exist in a phase modulation (PM) or frequency modulation (FM) system carrying a large number of message channels. Also, during that time, methods were devised for estimating the effects on the system's baseband amplitude response of amplitude-to-phase conversion produced by nonlinear elements in the system.

However, no methods were available for predicting the effects on intermodulation noise of combinations of even the simpler types of amplitude and phase distortion, or of predicting the magnitude of noise caused by an arbitrary amount of transmission distortion interacting with a unit (of the system) having an arbitrary amplitude-to-phase conversion characteristic. Before the TD-3 development was started, it was known that achieving the high performance objectives being set for the system might well depend on the ability to predict, quantitatively, the effects of various possible transmission distortions. Therefore, concentrated effort has been applied to solving these problems using the latest mathematical analysis techniques and computers to solve specific problems. The results of some of this work have been reported previously.[11, 12] The work proved to be of great value in finding early solutions to problems of excessive intermodulation noise encountered during early tests of the first TD-3 field installation.

II. SYSTEM PERFORMANCE OBJECTIVES

The TD-3 system is intended for new routes and for additions on existing TD-2 routes. At the beginning of the development program, a set of objectives was specified reflecting the designers' best judgment of the features and capabilities necessary for the new system.

The major objectives were:

(*i*) Compatibility with the TD-2 system. This dictated using the existing TD-2 frequency plan, antenna and outdoor waveguide arrangements, and repeater spacing.

(*ii*) Message capacity of 1200 circuits for each radio channel assigned to message service. For a fully equipped route of ten working channels this would provide a route capacity of 12,000 circuits.

(*iii*) "Worst circuit" noise of 41 dBrnc0 for a 4000 mile system during periods of normal (nonfaded) transmission. During fading, the system thermal noise will increase. Under such conditions, the "worst circuit" noise objective is 55 dBrnc0 maximum. At this noise level the radio channel normally will be switched to a standby protection channel.

(*iv*) Single tone interference of −70 dBm0 maximum in any voice circuit of a 4000 mile system during periods of normal (nonfaded) transmission. A tone of this level is barely discernible with a background noise of 41 dBrnc0 (which equals −47 dBm0 of 3 kHz flat noise). The tone interference, if dependent on the received carrier power, is permitted to increase during fading since the message circuit noise also will increase under this condition. A somewhat arbitrary decision was made to permit any carrier-dependent tone interference to increase to −42 dBm0 at the fade depth at which the radio channel noise has increased to 55 dBrnc0 (−33 dBm0 of 3 kHz flat noise) in the worst message circuit.

(*v*) Baseband amplitude response of ±0.25 dB flatness over the message band from about 500 kHz to 6 MHz for each radio channel in an IF protection switching section. This extremely stringent objective stems from the consideration of hits on data signals which may be caused by instantaneous level changes when the radio channels are switched between working and protection channels. Since data hits also can be caused by instantaneous phase changes, an objective of 30 degrees maximum phase difference (at baseband) between the radio channels of a protection switching section and between FM terminals has been established.*

Television transmission necessitates that the baseband amplitude response of the system, between one pair of FM terminals, remain approximately flat to frequencies as low as 6 Hz and be down no more than 3 dB at 2 Hz.

(*vi*) Reliability, as measured by yearly outage time, of 0.02 per cent per year for a two-way 4000 mile system. Experience indicated this would require that, for a fully equipped route, two of the 12 radio channels be reserved for protection.

Several comments are in order concerning these objectives. The use of the TD-2 frequency plan, which interleaves transmitting and receiving channels in the band, aggravates interchannel interference problems. A better plan from an interference standpoint, but one which might result in reduced system reliability, would group all transmitting frequencies in one half of the band (at a particular station), and all receiving frequencies in the other half. This is the plan used for the TH system.

However, the interference difficulties to be expected in using TD-3 and TD-2 on the same route, or at route crossings and junctions, would be so great as to preclude a change in frequency plan. In addition, the Bell System is not the only user of the 4 GHz common carrier band. Over the years, careful coordination has been required with other users to reduce interferences to tolerable levels. Therefore, the TD-2 frequency plan has been adopted for TD-3.

The noise objective of 41 dBrnc0 necessitates tight control of thermal and interference noise and precise control of transmission characteristics. The ability to exercise such tight control depends on the ability to make precise measurements of noise, amplitude, and delay deviations, and amplitude-to-phase conversion factors. The measurement precision obtainable with existing technology is barely adequate. During the TD-3 development, therefore, considerable effort was spent on devising special test facilities. In addition to the

* To meet this objective, all radio channels in a protection switching section are built out to approximately the same electrical length using IF cables. This build-out cable has been called DADE cable (differential absolute delay equalization). At 6 MHz, 30° phase shift is equal to about 14 ns delay difference which, in terms of IF cable, is equal to about 9 feet of the Western Electric Co. type 728A coaxial cable used at IF in the system.

test facilities[7] designed for maintaining operating systems, precision delay and amplitude-to-phase measuring equipment was constructed for laboratory use; these were of incalculable value during development. During field tests of the first system, a newly-developed automatic noise load scanner[13] greatly facilitated noise load testing.

III. SYSTEM MODEL

With the performance objectives for the system established, it was necessary to define the system arrangement in order to apportion the objectives among the various parts. In effect, a system model was synthesized. The model for TD-3, illustrated by the block diagram in Fig. 1, includes the voice circuit "stacking" or multiplex equipment, entrance link or baseband connecting facilities, baseband modulating and demodulating or FM terminal equipment, and the radio line. Protection switching equipment for the FM terminals and the radio line are also properly parts of the model, but except for their contributions to reliability and baseband frequency response, they have little influence on the breakdown of objectives.

The division of the 41 dBrnc0 total system noise objective among each of the major portions of the system is illustrated in Fig. 2 and discussed in more detail in the sections which follow.

3.1 *Multiplex, Wire Line Entrance Links and FM Terminals*

Standard mastergroup multiplex equipment forms the 1200 message circuit load that can be carried by each radio channel in the system. The baseband signal extends from 564 kHz to 5.772 MHz.

A wire line entrance link is used to connect the multiplex terminals to the FM terminals. Each entrance link includes the system pre- or de-emphasis network, equalization for the entrance link cable, and one or two baseband amplifiers for setting the required transmission levels. The distance between the multiplex and FM terminal equipment dictates the type of cable used for the entrance link and the number and locations of the entrance link amplifiers. Entrance links for use at distances up to 4 miles (a distance which requires amplifiers at each end of the link and the use of 0.375-inch air dielectric coaxial cable) have been designed for the system. The majority of entrance links are considerably shorter (2600 feet or less), require only a single amplifier, and use solid dielectric coaxial cable.

The TD-3 system model assumes that a 4000-mile message circuit

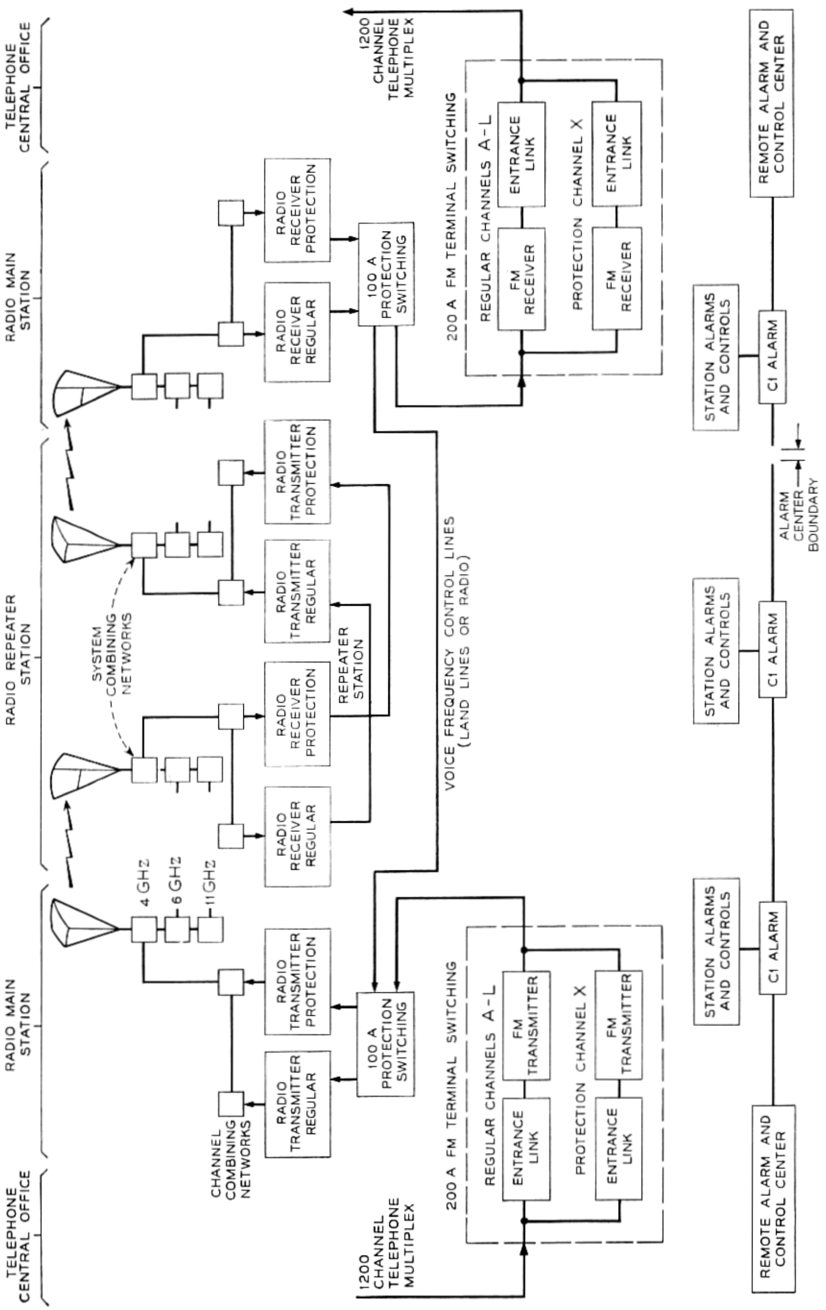

Fig. 1 — Block diagram of TD-3 system.

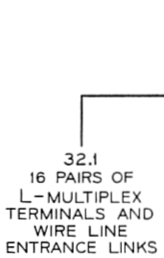

Fig. 2 — Allocation of total noise objective.

will pass through 16 pairs of FM terminals, wire line entrance links, and L-multiplex terminals. An allocation of 33 dBrnc0 total noise for 16 pairs of FM terminals is made as shown in Fig. 2. This allocation was based largely on previous system experience and on anticipated improvements that could be made with newer circuit designs. This total noise objective is made up, of course, of both thermal and cross-modulation noise. The division of the total noise allocation between these two components is given in another paper in this series.[4]

Also based on experience, 32.1 dBrnc0 were allocated for the total noise (thermal plus cross modulation) contributed by 16 pairs of L-multiplex terminals and wire line entrance links in the 4000-mile system model.

Several constants required in the system analysis relate to the 1200-message circuit load. The Holbrook and Dixon analysis[14] shows that 1200 circuits can be represented by an equivalent sine wave power of 26.0 dBm0. This represents the power of a complex, message-derived signal that is exceeded only a very small percentage of the time during the busy hour. It was assumed for the system model that this sine wave power will produce a peak frequency deviation of 4 MHz, the same deviation used in TD-2 and TH.

The average busy hour speech load is 14.8 dBm0, obtained by assuming an average talker power of -10 dBm0 and 25 per cent activity. This average busy hour speech load produces an rms frequency deviation of about 0.78 MHz. The difference of 11.2 dB between the equivalent sine wave power and the average busy hour speech load is defined as the multichannel load factor. Thus, this factor represents the difference between the average (rms) busy hour speech load and the maximum (rms) busy hour speech load which is exceeded only a very small percentage of the time.

3.2 *Radio Line*

A radio line is considered to start at the input to a microwave transmitter and to end at the output of a distant microwave receiver, one or more radio hops away. The 4000-mile system model for TD-3 assumes that the radio line portion of the system is composed of 140 radio hops, resulting in an average repeater spacing of 28.6 miles. As shown in Fig. 2, the radio line portion of the 4000-mile system is given a total noise allocation of 39.5 dBrnc0. This allocation is based largely on previous system experience and the anticipated performance of newer circuits. As discussed in Section IV, the total noise allocation for the radio line must be divided into allocations for thermal noise, cross-modulation noise, and radio channel interferences.

The system model also assumes that the radio line is broken into a number of IF protection switching sections. Based on TD-2 experience, an average switching section length of about 3 hops is assumed. More specifically, the radio line model is composed of 98 repeater-type stations and 42 main stations with IF switching. As noted earlier, the IF switching equipment makes a negligible contribution to the system noise performance. However, as noted in the paper that follows, the use of two microwave generators at the main stations, compared with only one at the repeater stations, is an important factor in determining the generator noise contribution.

The TD-3 system had to use the TD-2 frequency allocation plan shown in Fig. 3. In addition, TD-3 had to be designed to work with the TD-2 antenna and waveguide system. The antenna system, which uses the horn-reflector type antenna, circular waveguide, and polarization separation networks, has been described in earlier papers.[15, 16] Early in the TD-3 design period a model of the antenna system was assumed, and from this model an expected average received carrier power of -29.3 dBm was computed. This power was assumed throughout most of the development, and was used to estimate the thermal noise contribution of the radio line (see Section 4.2). Later studies have shown that the original antenna system model assumed too high a loss in the waveguide portions of the system, and that a more typical received carrier power is about -28 dBm. This power has been used in both this paper and the next[5] when assessing actual repeater performance.

Reliability studies indicate that each microwave receiver must have a fade margin of about 40 dB.[17] Fade margin is defined as the depth of fade in one hop that will cause the total system noise to in-

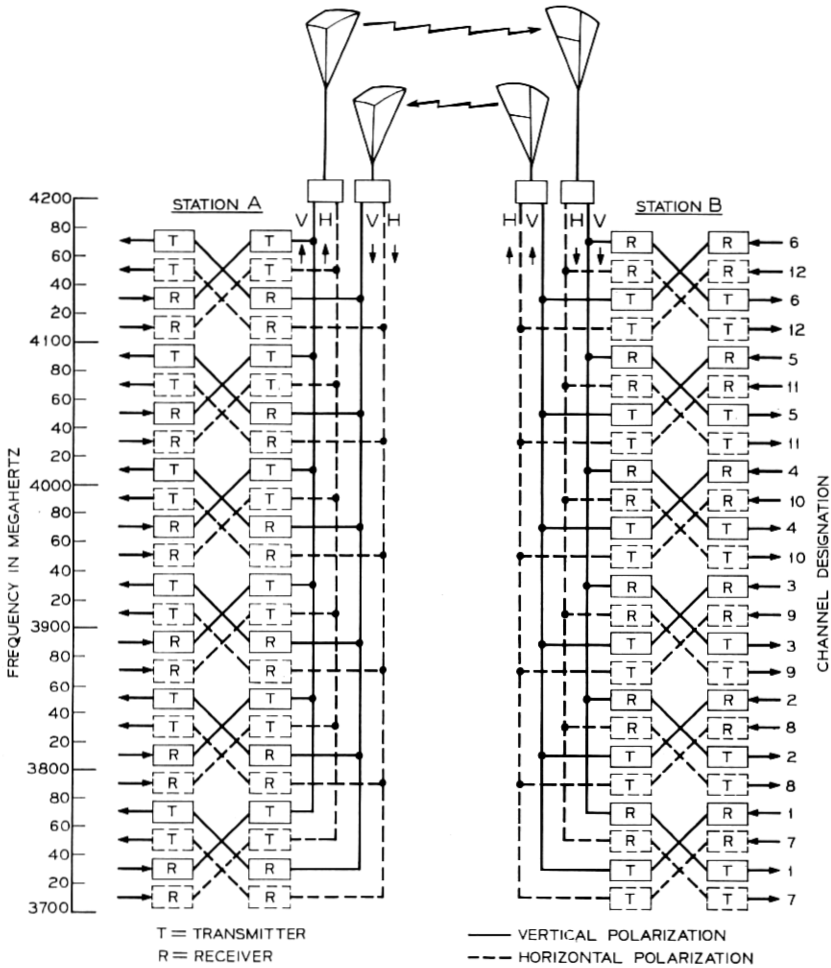

Fig. 3 — TD frequency allocation plan.

crease to 55 dBrnc0 in the worst message circuit. This is the noise level at which the channel is considered noncommercial and at which a switch to a protection channel normally is made.

IV. DEVELOPMENT PLAN FOR THE RADIO LINE

Once the broad objectives for the radio line had been set, a program was initiated to specify its basic components and parameters.

The procedure used to translate the desired transmission performance of the radio line into requirements for its major components was far from straightforward. Many of the requirements were related, so that in order to start iterative processes it was necessary to rely both on TD-2 and TH experience, and on feasibility studies conducted by the circuit design groups. In general, planning was directed as follows:

(*i*) Preliminary estimates of thermal noise were based on the system model, device performance, and system constants such as modulation index and pre-emphasis. It was assumed that the noise contribution from the microwave generator would be negligible and that the controlling factors were the available transmitted power and the achievable repeater noise figure.

(*ii*) Allocations of cross-modulation noise were made to co-channel interference,[18] and to waveguide and IF cable echoes.

(*iii*) The remainder of the radio line noise allocation was assigned to the cross-modulation noise of the microwave transmitters and receivers. For the most part this noise resulted from transmission deviations (amplitude and phase).

(*iv*) A large part of the planning was directed towards meeting the objective for spurious baseband tones while meeting the cross-modulation noise allocation for the microwave transmitters and receivers. To exclude spurious tones from the message band, it was necessary to build high selectivity into the repeater. This led to high in-band amplitude and envelope delay distortion. By comparing this in-band distortion with that which could be tolerated if the cross-modulation noise objective was to be met, the equalization requirements were defined.

In following these steps, the errors involved in estimating the noise from thermal sources, interferences, and transmission deviations were generally reasonable. Existing analytical methods and previous experience could be used at the start of the development program. However, as Section I mentions, no analytical methods were initially available for predicting the magnitude of noise caused by AM to PM conversion. The approach taken, which later proved to be wrong, was to assume that noise from this source would be small provided AM to PM conversion could be kept reasonably low.

4.1 *Capacity Limitations and System Constraints*

It was assumed that TD-3 would be able to handle two multiplex mastergroups (1200 circuits). The baseband load for two mastergroups extends from 0.564 to 5.772 MHz. At RF the first order sidebands occupy a band of ± 5.772 MHz about the carrier. Therefore, the second order sidebands of a channel do not overlap the first order sidebands of an adjacent channel located 20 MHz away. By keeping the FM index low, the power in the third and higher order sidebands is very small and the effect of overlap of these sidebands is insignificant. Also, by concentrating the signal power in the first order sideband region the transmission characteristics of the channel need be carefully controlled only within the center 12 MHz of the nominal 20 MHz band.

Since TD-3 had to be compatible with the massive TD-2 plant already in the field, it had to follow the same microwave frequency plan as TD-2 to avoid interferences and to allow partially equipped TD-2 routes to be filled out with TD-3. Also, TD-3 had to use a heterodyne repeater with an intermediate frequency of 70 MHz. This approach facilitates emergency restoration between TD systems and permits a combined TD-2 and TD-3 "intermix" system to use a common 100A switching system.

The noise resulting from waveguide echoes was essentially fixed by the antenna system already developed for long haul microwave radio.[15] The only influence that TD-3 had on this noise contribution was the impedance its transmitter and receiver presented to the waveguide runs. A constraint was also imposed by the section loss, which is a function of the antenna system and repeater spacing.

4.2 *Thermal Noise*

A very good estimate of the thermal noise performance of the system is easily calculated from the received carrier power, and the radio repeater noise figure. The transmitter output power and the section loss determine the received carrier power, and the repeater noise figure depends heavily on the noise of the first stage of the receiver.

Before work commenced on TD-3, a relatively inexpensive 5-watt traveling wave tube had been constructed for use at 6 GHz as the result of an experimental development project. The design incorporated periodic focusing and conduction cooling. It appeared that a similar TWT could be constructed for the 4 GHz band. Also, work on parametric amplifiers had progressed to the point where it was possible to predict that by using this device at the front end of the microwave receiver, the over-all repeater noise figure could be reduced to 7 dB.

Based on the likelihood that these two devices would be available, it was possible to proceed with the design of a 41 dBrnc0 system. As it turned out, TD-3 was able to do without the parametric amplifier because of the recent development of a low noise Schottky barrier diode receiver modulator and an improved noise figure IF preamplifier. Thus, whereas much of the design for TD-3 assumed the use of the parametric amplifier, the production TD-3 microwave receiver uses a Schottky barrier diode receiver modulator and no parametric amplifier.*

Actually, the thermal noise of the TD-3 radio line originates from two sources, the devices in the transmission path, and the microwave generators (beat oscillators). If TD-3 did not use pre- and de-emphasis, the noise at baseband from the transmission devices would have the familiar triangular spectrum. The thermal noise from the microwave generators is introduced through the up-converters and down-converters, which are located in the microwave transmitters and receivers, respectively. The output spectrum from each microwave generator consists of a carrier surrounded by noise sidebands. The angle modulated sideband components are transferred by the converters onto the signal carrier, without a change in their modulation index.

When the initial noise allocations were made it was assumed that the thermal noise from the microwave generators could be neglected. It was later determined that such high performance microwave generators were not feasible[5] and that although the noise contribution of a practicable generator was less than 0.3 dB in the top message circuit, the generator noise dominated at the lower baseband frequencies.

Pre-emphasis and de-emphasis networks shape the baseband signal at the input and output of the FM terminals. These networks give the first order signal sidebands the same shaping as the noise and as a result they keep the signal-to-noise ratio approximately constant over the baseband. However, balancing the system noise across the baseband is not simple because pre-emphasis affects both the thermal and cross-modulation noise components from the radio line and FM terminals. A pre-emphasis characteristic was arrived at early in development by scaling the TD-2 pre-emphasis network characteristic (see Fig. 4). Its performance proved to be satisfactory with the final

* Except for about 260 receivers. These receivers have no parametric amplifier and use the original design of receiver modulator which has 3.5 dB worse noise figure than the Schottky type.

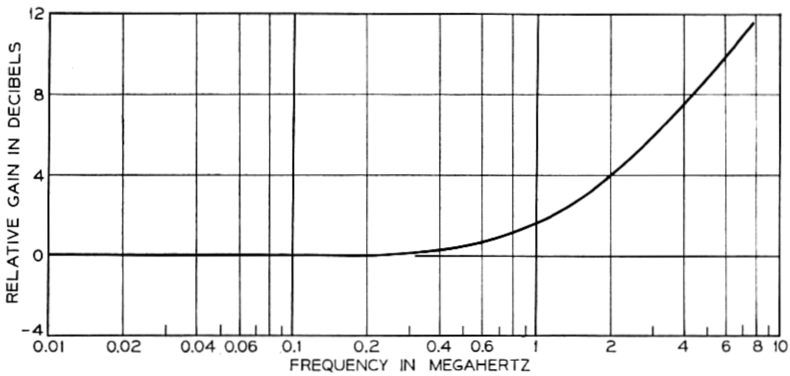

Fig. 4 — Pre-emphasis characteristic—message service.

repeater design so that this characteristic has been adopted for TD-3.

Of the 41 dBrnc0 system noise objective, 39.5 dBrnc0 was allocated to the radio line. The radio line allocation was, in turn, divided among various contributors, as shown in Fig. 5. Based on the 5-watt TWT and a 7 dB repeater noise figure, the thermal noise allocation for the radio line was calculated to be +34.9 dBrnc0. At the time this allocation was made, the pre-emphasis characteristic had not been specified, and a figure of 3.9 dB was used as an estimate of the noise improvement caused by pre-emphasis at the top message frequency. The actual pre-emphasis used (Fig. 4) provides about 1 dB less improvement than originally assumed.

4.3 *Interference and Echo Noise*

Important contributions to noise come from the co-channel interference mechanism. Its source is a channel on the same nominal fre-

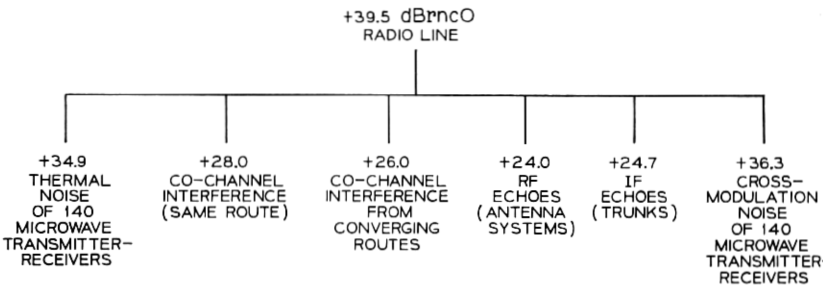

Fig. 5 — Allocation of noise objective for radio line.

quency, so that selective filtering in the repeater does not help. Instead, it is controlled by antenna couplings and route layout. The subject has been dealt with at length[18] and it is enough to say that the TD-2 and TH noise allocations of 28 dBrnc0 for the main route, and 26 dBrnc0 for converging routes, were used.

An allocation of 27.4 dBrnc0 was given to the cross-modulation noise which results from echoes in the antenna system and in IF trunks. To compute these noise contributions, the lengths of the echo paths and the return losses of the discontinuities were first converted to echo delays and echo amplitude ratios. Then the theory presented in Ref. 9 was applied. IF echoes were given 24.7 dBrnc0 based on the assumed return losses of the station equipment and the lengths of interconnecting trunks. RF echoes were allocated 24.0 dBrnc0, based on the dimensions of a typical TD-2 antenna waveguide run, and the magnitude and location of waveguide discontinuities which commonly occur at existing installations.

4.4 *Cross-Modulation Noise*

After the thermal, interference, and echo noise allocations had been determined it was obvious that 36.3 dBrnc0 remained for allocation to the cross-modulation noise for 140 transmitter receiver pairs. The next step was to divide this figure into allocations for each transmitter-receiver pair, and then to convert to requirements on the permissible amplitude and envelope delay distortion of the equalized microwave transmitter and receiver. To recognize the magnitude of the equalization problem, it was first necessary to determine the in-band distortion before equalization. This required a study of the selectivity (selective filtering) requirements for the transmitter and receiver.

4.5 *Selectivity and Interfering Tone Requirements*

Selectivity is required in the microwave transmitter and receiver to control inter- and intrasystem interferences, and to satisfy FCC regulations on the emission of spurious signals. Interference from outside the 500 MHz common carrier band has to be adequately suppressed. It was assumed that intersystem interference from systems within the common carrier band would be controlled by physically separating routes, and where necessary, avoiding same-channel conflicts. However, some of the co-channel noise allocation, mentioned earlier, took account of converging routes.[18]

The selectivity was determined primarily by considering intrasystem interference, and analysis was, for the most part, devoted to a study of the many mechanisms which cause spurious tones in the baseband signal. One of the system objectives was to prevent tones which fall into message-circuits from exceeding −70 dBm0 under no fade conditions. Based on this figure, a requirements curve was derived for the IF carrier to spurious tone interference ratio (C/I) at the output of the microwave receiver. The curve, Fig. 6, shows the requirement as a function of the resulting baseband interference frequency.

The most severe requirements are imposed when the interference falls in the baseband region occupied by the message circuit load. Incidentally, the curve shows this region extending from 312 kHz to 5.772 MHz, anticipating that an extra 60 circuits might be added at the low frequency end. In the bands 0 to 312 kHz, and 5.772 to 11.544 MHz (excluding a small band about 9 MHz), the requirement was based on the possibility of a 0 dBm0 test tone in the message band intermodulating with the interfering tone to generate a −70 dBm0 product tone elsewhere in the message band. The theory used to generate the requirements curve in the three frequency regions is developed in the Appendix. The derivation in the appendix pertains to an FM system; therefore the test tone and interference levels must be adjusted to take care of the system pre-emphasis shown in Fig. 4.

The 87 dB C/I requirement at 9 MHz in the baseband (see Fig. 6) was necessary because the initiator of the 100A switching system monitors noise at this frequency. An interfering tone must not be allowed to alter, appreciably, the fade depth at which a protection switch is requested. The requirement in the region between 11.544 MHz and 20 MHz was set by considering that tones in this band

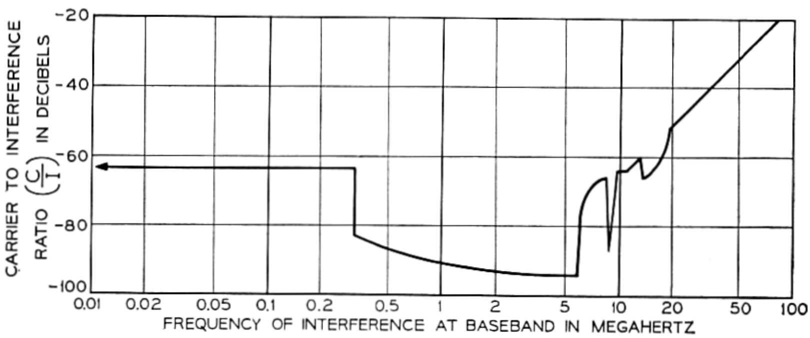

Fig. 6 — Requirement for carrier to single tone interference at microwave receiver output.

would be retransmitted and that they would fall into the baseband of an adjacent channel. Above 20 MHz the requirement was also determined by retransmitted tones, along with such factors as loss of fade range and capture of the FM receiver.

The requirements curve applies to interference tones whose carrier-to-interference ratios are not affected by fading. For those ratios which decrease dB for dB with carrier fade depth, the requirements are more stringent. Section II states that at a 40 dB fade, the tone interference in a baseband message-circuit is allowed to increase by 28 dB, to -42 dBm0. Therefore, the requirement on such tones was made 12 dB stiffer. However, for RF interference tones which will reduce the fade range, cause capture, and are retransmitted into adjacent channels, the requirements curve applies at any fade depth.

Many of the signals which cause interferences owe their frequency stability to an FM terminal transmitter. In the past, the FM transmitters have generated a "burbly" carrier, typical of klystron sources, and it has been possible to reduce the interference requirements by a "burble" factor of about 10 dB based on subjective measurements. With the 3A FM terminal transmitter the carrier is much more stable, and no such "burble" factor can be applied.

V. RESULTS OF LABORATORY AND FIELD TESTS

The first system-type measurements of the TD-3 system began in mid-1965 with four models of the repeater bay set up in the laboratory. Field testing commenced in April 1966 on the initial installation route. An extensive series of tests on the individual microwave transmitters and receivers, the individual hops, and the over-all multihop system were conducted on this route during an eight month period to obtain as complete an evaluation as possible of the system performance. Another series of tests was started in April 1967 on a second installation which involved the addition of several TD-3 radio channels to an existing, partially equipped TD-2 route. This second series primarily was intended to evaluate certain interferences between the two systems. Throughout field testing, the laboratory was used continually to supplement the field program and to perform certain tests that could be controlled better in the laboratory.

5.1 *Description of Laboratory and Field Test Installations*

5.1.1 *Laboratory*

The four laboratory repeater bays, both individually and in combination, were used almost continuously for various tests before the field tests on the initial installation began in 1966. The most important of these early tests involved the use of these repeater bays for FCC type acceptance tests, simulated multihop system tests, and adjacent channel interference tests. Later, the laboratory equipment was used to help track down the sources of excessive cross-modulation noise encountered at the beginning of initial installation tests, and to make further transmission and interference tests as part of continued general studies.

5.1.2 *Initial Installation*

The TD-3 initial installation was a new radio route consisting of a single 5-hop switching section with a length of about 140 miles between Alexander, Ark. (near Little Rock), and Arkabutla, Miss. (near Memphis, Tenn.). This switching section was the first portion constructed for a new TD-3 route that was planned by the A.T.&T. Long Lines Department for service, in late 1967, between Warrior, Ala. and Noble, Okla.

The selection of the radio channels for the initial installation was based almost entirely on the following test considerations:

(*i*) To permit field evaluation of transmitter modulators using either the upper or lower sidebands, it was desirable to have at least one channel with the local oscillator frequency above, and at least one with the local oscillator frequency below, the channel center frequency. This required installing at least one channel below the center of the 4 GHz band (channel 1, 2, 3, 7, 8, or 9) and at least one channel above the center (channel 4, 5, 6, 10, 11, or 12).

(*ii*) The intrasystem interference test considerations required that at least three adjacent one-way radio channels be installed to permit measuring the most important of the near-end and far-end adjacent channel type interferences. When combined with the local oscillator consideration, it was evident that the group of adjacent channels had to be located around the center of the 4 GHz band. Therefore it was decided to install channels 3, 4, 9, and 10 in each direction of transmission. This gave two channels with the local oscillator frequencies above, and two below, the channel center frequencies. By interconnecting these channels at IF at each end of the switching section, as many as 20 hops of each type of channel, and a maximum of 40 hops total, were available for measuring.

Six new buildings and towers of the integrated building-tower design[8] were constructed for the initial installation. The 100A protection switching system and 3A FM terminals were installed at each end of the system. The individual hops ranged from about 24 to 30 miles, and averaged 27.3 miles; towers ranged from 87.5 to 287.5 feet high, and the antenna waveguide system losses were between 2.0 and 2.8 dB per antenna. The computed section losses, from the input to the channel-combining network of the microwave transmitter to the output of the channel-combining network of the receiver, ranged from 61.6 to 64.2 dB. The computed rms average section loss was approximately 63 dB, about 3 dB lower than assumed in the system model described in Section III. Thus, for +37 dBm input to the channel combining network of the microwave transmitter, the average received carrier power on this route, at the input to the receiver channel band-pass filter, was about −26 dBm.

5.1.3 *Intermix Installation*

The second TD-3 installation involved adding three two-way TD-3 radio channels to an existing TD-2 route between Ennis, Texas (near Dallas) and Seguin, Texas (near San Antonio). This route comprised a single, 8-hop, 225-mile switching section. At the time of the tests, TD-2 radio channels 1, 2, 3, and 4 were in commercial service in both directions of transmission, and channel 6 was in north-to-south service. Channel 1 was assigned as the protection channel.

The selection of the TD-3 channels for the Ennis-Seguin route was based partially on normal route expansion and partially on test considerations. It was intended that the route represent a typical TD-2 route which had been partially equipped with channels of one polarization, and which was to be expanded by adding TD-3 channels of the opposite polarization. This meant that all of the TD-3 channels had to be selected from channels 7 through 12. Studies showed that by using TD-3 channels 7 and 8 with TD-2 channels 1 and 2, all of the major interferences that might exist between the two systems could be evaluated. To permit the various tests to be made without interrupting the TD-2 commercial service, TD-3 channel 12 also was installed.

The Ennis-Seguin route had conventional TD-2 repeater stations with separate buildings and towers. The 100A protection switching system (for use with both the TD-2 and TD-3 channels) and 3A FM terminals were installed at each end of the route. Hops ranged from about 20 to 34 miles and averaged 28.0 miles. Towers were from 162.5 to 275.0 feet high. Section losses ranged from 63.3 to 66.7 dB. The rms average section loss was about 65 dB, which is considered more typical than the 66 dB assumed in the system model. Thus, the average received carrier power on this route was approximately −28 dBm.

5.1.4 *Types of Receivers Tested*

Initially, a parametric amplifier in the microwave receiver appeared to be the most promising means of meeting the 7 dB repeater noise figure objective. Thus, a parametric amplifier with about 4 dB noise figure and 12 dB gain was developed. The first two field installations were equipped with these amplifiers. However, the amplifier proved to be too costly to continue in manufacture and too difficult to adjust and maintain in proper operation. Furthermore, it was found to be a source of AM to PM cross-modulation noise. For these reasons, it was decided in late 1966 to remove the amplifiers from the first two field installations before commercial service started and to discontinue their use in bay production. Except for the effect on the cross-modulation noise described in Section 5.2.2, this series of papers gives no further information on the parametric amplifier-equipped channels.

The elimination of the parametric amplifier was possible in part because of improved, low noise, semiconductor devices developed several years after system design began. In 1966 a new, low noise transistor with adequate power handling capacity became available for the IF preamplifier. With this transistor it was possible to obtain, with only minor circuit changes in the IF preamplifier, a noise figure of about 10.5 dB for the receiver modulator-IF preamplifier* and about 11 dB for the entire microwave receiver. The resulting over-all repeater noise figure was about 11.5 dB.† This improvement in the IF preamplifier was made on the first two field installations when the parametric amplifiers were removed. About 175 more microwave receivers of this type were manufactured during the first part of 1967 for other TD-3 routes. As described in Section 5.2.2, the total noise

* The original receiver modulator-IF preamplifier circuit had a noise figure of about 13.5 dB.

† Repeater noise figure includes the noise contributions from all circuits in the microwave transmitter and receiver, including the microwave generator. The repeater noise figure values given in this section apply to frequencies several megahertz or more either side of the carrier frequency (that is, frequencies corresponding to the upper baseband frequencies) where the contribution of the microwave generator is negligible. Receiver noise figure includes only the noise contributions of the circuits in the microwave receiver, without the microwave generator. The repeater noise figure is used when evaluating the noise contribution of the repeater under normal (nonfaded) conditions. The receiver noise figure is used when evaluating the repeater noise contribution under a deep fade condition.

performance of the system with this type of receiver departs only slightly from the original system objectives.

In the third quarter of 1967, a new receiver modulator-IF preamplifier circuit with a single Schottky barrier diode in the modulator and a noise figure of about 7 dB, was introduced for new production. This improved circuit, which reduces the microwave receiver noise figure to about 7.5 dB and the over-all repeater noise figure to about 8.5 dB, provides a system which meets the original total noise and fade margin objectives. Laboratory models of the new receiver modulator-IF preamplifier circuit were tested on two of the channels of the initial installation route. Section 5.2.2 discusses the performance measured on these channels.

5.2 *Test Results*

5.2.1 *Antenna System Performance*

Near-end interference between adjacent transmitting and receiving channels at a repeater station is dependent on the coupling loss between the antennas at that station. Two losses that must be considered are the side-to-side coupling loss, measured between a transmitting antenna and a receiving antenna facing the same direction, and the back-to-back coupling loss, between the transmitting and receiving antennas facing in opposite directions.

Side-to-side coupling loss is important when considering interferences between a transmitting channel and an adjacent, oppositely directed receiving channel. Back-to-back loss is important when considering interferences between a transmitting channel and an adjacent, similarly directed, receiving channel. Normally the side-to-side coupling loss is considerably lower than the back-to-back coupling loss and therefore presents the controlling near-end interference condition.

An earlier paper presented a cumulative distribution curve for side-to-side coupling losses measured on 97 pairs of horn-reflector antennas.[1] This information was used extensively in the selectivity determinations and general interference studies during TD-3 development.

Several factors made it desirable to obtain similar coupling loss information on the Alexander-Arkabutla initial installation. First, the original data represented single frequency measurements. No information was available on the variation of the coupling loss with frequency. Second, the original data were taken for the same carrier polarization. Since adjacent channels are oppositely polarized, it was desirable to determine any increase in coupling loss from cross polarization. Third, knowledge of the actual coupling losses on the route was needed to plan certain interference tests and to interpret the interference test results. For these reasons, swept measurements were made of the side-to-side coupling losses at all stations, and of the back-to-back coupling losses at all intermediate repeater stations, as part of the initial installation test program.

The swept measurements were made using the four available microwave transmitters at each station. Thus all of the data collected were obtained at frequencies around the center of the 4 GHz band. Each microwave transmitter carrier was swept ± 10 MHz about its center frequency by supplying a 60 to 80 MHz swept IF signal to the transmitter input. At each transmitted frequency, the coupling loss was measured for both the normal and cross-polarized components of the transmitted carrier. All coupling loss measurements include the loss of the transmitting and receiving antenna waveguide runs.

Cumulative distribution curves for the side-to-side and back-to-back coupling losses are given in Fig. 7. The coupling losses measured between a given pair of antennas exhibited a ripply characteristic but showed no broad frequency dependency. The peak-to-peak amplitude of the ripple averaged about 3 dB for the side-to-side and about 10 dB for the back-to-back coupling loss. At least part of the back-to-back ripple was observed to be caused by direct leakage from the microwave transmitter to the test receiver. This leakage was not significant when measuring the side-to-side coupling losses, which typically were 20 dB lower than the back-to-back losses. The minimum coupling loss measured in each of the four swept frequency bands involved in these tests for each antenna pair was used in forming the distributions of Fig. 7.

Figure 7 shows that cross polarization results in about 6 dB higher side-to-side coupling loss than measured for the same polarization. The previously published side-to-side coupling data[1] agree rather closely with the opposite polarization distribution curve. Since the earlier, single frequency data were taken for the same polarization, this discrepancy may be caused in part by having used for Fig. 7 the lowest coupling loss observed in each 20 MHz swept band and in part simply by a different and substantially smaller sample.

One antenna pair contributed all of the side-to-side coupling losses

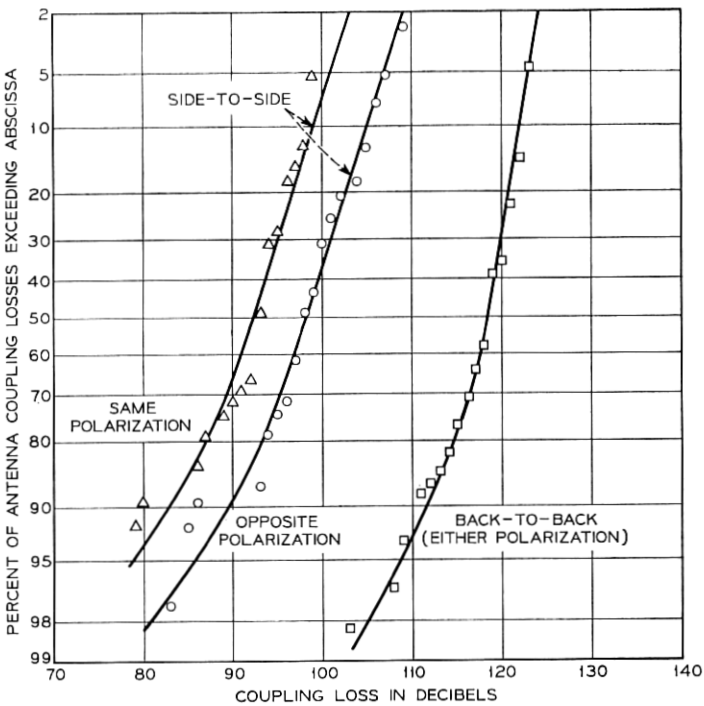

Fig. 7 — Distribution of antenna system coupling losses as measured on the initial installation.

below 88 dB. This station has a low tower (87.5 feet), and later measurements indicated that reflections from a nearby wooded area appeared to be significantly affecting this loss and causing it to vary appreciably from day to day.

The back-to-back coupling loss data showed little difference between the same and opposite polarizations. Therefore, the back-to-back distribution shown in Fig. 7 includes all measurements for both polarizations. The results may be several dB pessimistic because of the direct leakage into the test receiver.

Far-end interference between adjacent channels is dependent on the cross polarization discrimination (XPD). Swept XPD measurements were made on each hop of the Alexander-Arkabutla initial installation. These data were taken to obtain general information on the variation in the XPD across the 4 GHz band and specific information for planning and interpreting certain interference tests.

The swept measurements were made over 20 MHz wide bands centered at each of six uniformly spaced frequencies from 3710 to 4110 MHz.

In general, the XPD varied rapidly with frequency, with many peaks and valleys across each 20 MHz band. Peak-to-peak variations of 2 to 15 dB were typical. An extreme case was one antenna pair which exhibited more than 30 dB peak-to-peak variation in each of two swept bands in which the lowest XPD within each band was about 25 dB. There was no apparent correlation between the peak-to-peak variation and the lowest value of XPD in each band. For each antenna pair there was often a substantial difference between the lowest XPD measured within each 20 MHz band for the two polarizations. However, when compared across the entire 4 GHz band, the XPD's for the two polarizations of a given antenna pair showed no significant differences.

Figure 8 is a cumulative distribution curve showing the lowest XPD measured in each of the 20 MHz bands for all of the antenna pairs. The data for both transmitted polarizations has been used for this figure. One antenna pair accounts for approximately 80 per cent of the values below 30 dB.

Swept amplitude and envelope delay distortion were measured, on each of the hops of the initial installation, from the IF input of each microwave transmitter to the IF output of the far receiver. All showed a ripply characteristic which was traced to multimoding in the circular waveguide runs. The principal ripple-producing mechanism was found to involve a round trip echo of energy traveling in the TM_{01} mode in the circular waveguide. Generation of the TM_{01} mode and its reconversion to dominant mode (TE_{11}) occurred near the antenna feedhorn. The circular waveguide lengths used on the initial installation resulted in ripple periods of from 1 to 3 MHz. Typically, the echoes in each waveguide run were 50 to 60 dB below the desired signal. The effect of these echoes on the system cross modulation noise is discussed in Section 5.2.2. Means of reducing the TM_{01} and other higher order modes that may be generated in the antenna system are being studied.

5.2.2 *Noise Loading and EDD Measurements*

The first noise loading measurements on a multihop system were made in late 1965 using the radio repeater bays installed in the laboratory. During the field tests, many more noise loading measurements were

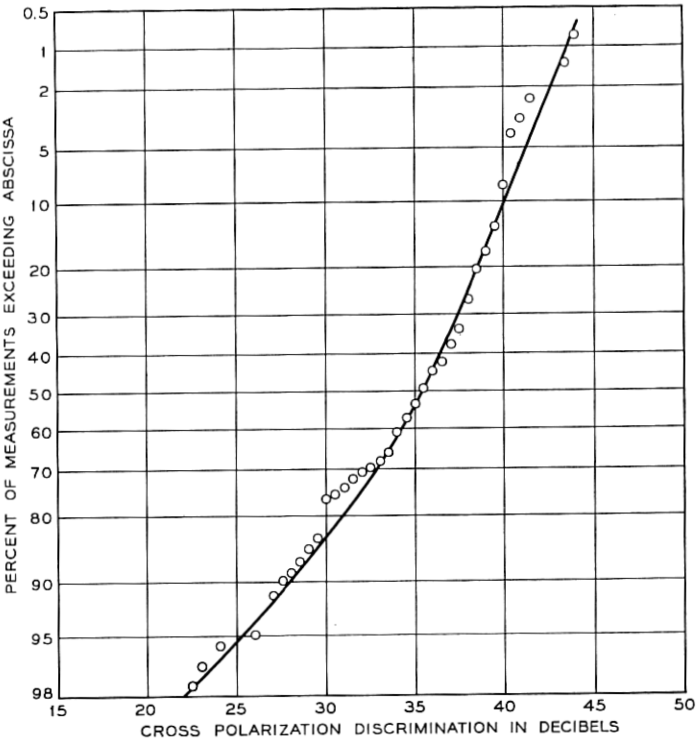

Fig. 8 — Distribution of antenna system cross polarization discrimination as measured on the initial installation.

made on each of the two-way radio channels, and on tandem combinations of these channels, on the Alexander-Arkabutla and Ennis-Seguin routes. All of these measurements used the automatic noise load scanner mentioned in Section II that provided a meter display of the noise at 0 dB TL simultaneously at each of six test frequencies across the baseband.[13] The test frequencies were 0.36, 1.0, 1.95, 2.8, 3.9, and 5.4 MHz.

Networks were used in the noise generator to form a pre-emphasized noise load signal extending from approximately 0.3 to 5.8 MHz to simulate the pre-emphasized 1200-channel message load which the system was designed to carry. The pre-emphasis network provided the shaping normally used in the system and which is shown in Fig. 4. Most of the noise load measurements were made from 15 dB below to 10 dB above reference (±4 MHz peak frequency) system deviation. Reference deviation corresponded to a total noise load power of −23.2 dBm at the input to the FM terminal transmitter.

The results of the early multihop system tests in the laboratory were extremely encouraging. For these tests, the repeater bays were interconnected to form a 4-hop system operating on channel 7. Noise load measurements showed low cross-modulation noise, particularly at the top end of the message band, and indicated that the system objectives were being met. In contrast to these early results, excessively high cross-modulation noise was measured during the beginning of the field tests on the initial installation. The field measurements rather quickly indicated a multiplicity of excess noise sources, at least some of which were dependent on the received carrier power.

Additional laboratory tests indicated that the noise was arising from two basic mechanisms: AM to PM conversion at both microwave and intermediate frequencies, and harmonic generation at IF. The rapidity with which the AM to PM problem was recognized, and corrective measures taken, resulted largely from the completion at about that same time of a theoretical analysis.[12]

Briefly, this is what was found and how it was corrected:

(*i*) *IF main amplifier:* Originally the repeater equalizer was located after the IF main amplifier.* As a result, all of the unequalized amplitude and envelope delay distortion in the signal path of the microwave receiver and the preceding microwave transmitter appeared ahead of the main amplifier. This distortion, which is contributed principally by the microwave networks and filters in the signal path, slightly altered the amplitude and phase of the sidebands relative to the carrier, causing the carrier to appear to be both amplitude- as well as frequency-modulated by the baseband signal. At normal input power, the IF main amplifier has a small but not insignificant AM to PM conversion (typically about 0.4 degrees/dB). The laboratory measurements showed that this amount of AM to PM conversion, in combination with the AM on the signal caused by the amplitude and delay distortion in the signal path, was sufficient to produce approximately 6 dBrnc0 of cross-modulation noise per hop in the top message circuit. This was corrected by locating the repeater equalizer ahead of the main amplifier.

(*ii*) *IF preamplifier:* In the original microwave receiver circuit, the second harmonic of the IF signal, generated in the IF preamplifier, was delivered to the IF main amplifier via the IF band-pass filter. This filter has loss peaks at 50 and 90 MHz but relatively low loss

* Reference 5 has a block diagram showing the major elements in the signal path of the microwave transmitter and receiver.

at 140 MHz. In addition, the filter has about 100 ns more delay at 70 MHz than at 140 MHz. As a result, the fundamental and second harmonic, when recombined by nonlinearity in the IF main amplifier, formed a delayed fundamental product equivalent to an IF echo. The resultant cross-modulation noise was estimated from the laboratory tests to be about 6 dBrnc0 per hop in the top message circuit at normal received carrier power. Up-fades, which were very prevalent during the summer on the initial installation route, markedly increased this noise. The problem was corrected by locating the IF low-pass filter ahead of the IF main amplifier to attenuate the second harmonic. Originally this filter had been located after the IF main amplifier to prevent this same noise problem from occurring in the main amplifier-equalizer-limiter portion of the circuit.

(*iii*) *TWT amplifier and parametric amplifier:* As previously mentioned, the original system included a parametric amplifier in the microwave receiver. It was determined that AM to PM conversion in both the TWT amplifier and the parametric amplifier, combined with the transmission deviations that preceded each of these units, produced cross-modulation noises which tended to cancel each other at normal received carrier power. The AM to PM conversion of the parametric amplifier increased with increased received carrier power. Thus, during up-fades, the parametric amplifier noise contribution tended to predominate, while during down-fades the TWT contribution was controlling. Removal of the paramertic amplifier from the system left the uncancelled TWT amplifier contribution, which was about 13 dBrnc0 per hop in the top message circuit. This noise was eliminated by using a relatively wide band, sideband selecting filter at the output of the transmitter modulator to reduce the distortion that preceded the TWT amplifier. The high selectivity filter originally used ahead of the TWT was moved to the TWT output to retain the over-all selectivity required in the microwave transmitter.

In retrospect, it became apparent why these noise sources were not observed during the original 4-hop system tests in the laboratory. Up-fading, which would have accentuated the IF main amplifier, IF preamplifier, and parametric amplifier contributions, was not performed. Down-fades in an individual hop were introduced, but the increase in thermal noise from the faded hop masked the change in the cross-modulation noise of the system. Most significant, however, was the voltage-type addition which this excess noise exhibited. The excess noise from the 4-hop system was found to have been cancelling residual cross-modulation noise in the standard 3A FM terminal pair used in the measurements. This led to an overly optimistic picture of the radio system cross-modulation noise when the terminal contribution was subtracted from the over-all measurement. The initial installation permitted putting many more hops in tandem, and by this means the radio system noise was made to predominate over the terminal contribution. Until this noise-addition problem was recognized, many false conclusions were reached, particularly in the field, on the cross-modulation noise contributions of the individual portions of the repeater. This experience demonstrates the importance of a field trial, involving a large number of units, for any new system.

A typical set of noise load "V-curves," obtained after making the above changes in the microwave transmitter and receiver, is shown in Fig. 9.* These data were recorded on channel 9 on the Alexander-Arkabutla initial installation route. The noise load set was located at Alexander, and the two-way channel was interconnected at IF at Arkabutla to form a 10-hop loop. Thus, the loop involved eight intermediate repeater station bays, two main station microwave transmitters and two main station microwave receivers.

For this test the microwave receivers were equipped with the 7 dB noise figure receiver modulator-IF preamplifier. The data shown in Fig. 9 represent the radio line portion of the loop; the contribution of the FM terminals has been subtracted. The measurements made at the 1.0 MHz test frequency have not been included in this figure. Below reference drive (the 0 dB point on the abscissa) the 1.0 MHz performance is approximately coincident with the 2.8 MHz curve. Above reference drive, the cross-modulation noise of the radio line, at both the 1.0 and 0.36 MHz test frequencies, tends to partially cancel the cross-modulation noise of the FM terminals. Since the

* These "V-curves" present the total noise of the system at 0 dB TL, at each of the indicated baseband frequencies, as a function of baseband drive into the FM terminal transmitter. Reference drive corresponds to the normal system deviation (which as has been noted is ±4 MHz peak for TD-3). To the left of reference drive, where the curves approach a one-to-one relationship of noise vs drive, the baseband signals are sufficiently low that the thermal noise predominates. If there were no sources of cross-modulation noise in the system, increasing the drive would result in a straight line, dB for dB, decrease in the noise versus drive. However, in the usual case, cross-modulation noise will begin to appear as the drive is increased and generally will predominate at the high drives (to the right of reference drive). At any drive, a "V-curve" ordinate value is the sum of the thermal and cross-modulation noise components. By linearly extrapolating the noise measured at very low drives, the thermal noise at higher drives can be computed. The cross-modulation noise is then obtained by subtracting the computed thermal noise from the measured total noise.

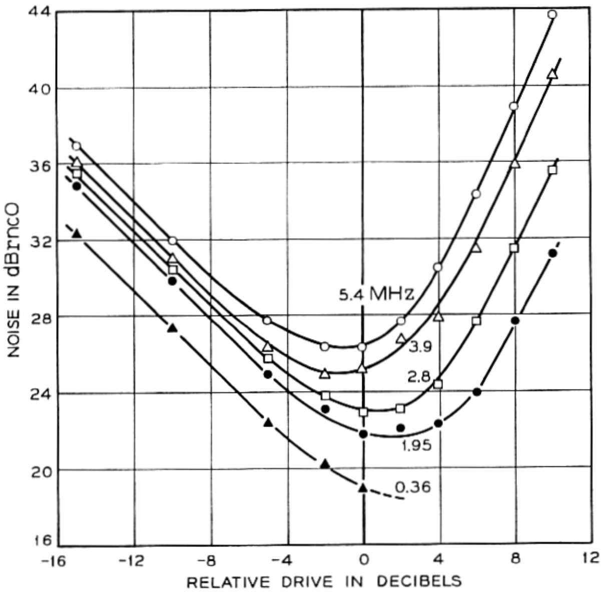

Fig. 9—Noise load V-curves measured on a 10-hop loop of channel 9 of the initial installation. The channel was equipped with Schottky barrier diode receiver modulators.

law of addition of the radio line and FM terminal noise is not known, an accurate estimate of the cross-modulation noise of the radio line cannot be made for high drives at either of these test frequencies.

The noise of the radio line portion of the 10-hop loop of channel 9 is shown in Fig. 10 as a function of baseband frequency. The thermal noise across the baseband was measured point by point with a tunable narrow band power meter. The cross-modulation noise was derived from the noise load measurements at reference drive.

The thermal noise objective for the radio line portion of a 10-hop system is 23.4 dBrnc0. This objective was based on a 7 dB noise figure *repeater*. Figure 10 shows that this objective appears to be just met at the top end of the message band. However, the received carrier power on the Alexander-Arkabutla route is approximately 2 dB higher than the average based on TD-2 system experience. Therefore, the thermal noise in the upper portion of the baseband should be increased by about 1.5 dB to make it representative of an average system.* If this adjustment is made, the thermal noise objective then

* The 2 dB decrease in received carrier power results in a somewhat smaller increase in the top message circuit thermal noise because of the noise contributions from the microwave transmitter circuits.

is exceeded in the top message circuit by about 1.5 dB and is the direct result of having exceeded the repeater noise figure objective by the same amount.

The thermal noise below about 2 MHz is contributed principally by the microwave generators. The ripple in the noise characteristic, which is particularly evident at the lower baseband frequencies, results from the addition and cancellation of the microwave generator noise at the intermediate repeater stations, as described in Ref. 5.

The objective for cross modulation noise caused by transmission deviations, echoes, and same-route cochannel interference is 25.8 dBrnc0 for the radio line portion of a 10-hop system. Figure 10 shows that this objective is met at all frequencies across the message band. The slope of the V-curves to the right of reference drive in Fig. 9 shows that third order cross modulation noise is predominant above about 2 MHz. Noise load measurements of 20, 30, and 40 hops on the initial installation showed that the cross modulation noise adds approximately on a power basis at all frequencies.

Calculation of the cross modulation noise caused by the residual transmission deviations, measured on equalized radio hops set up in

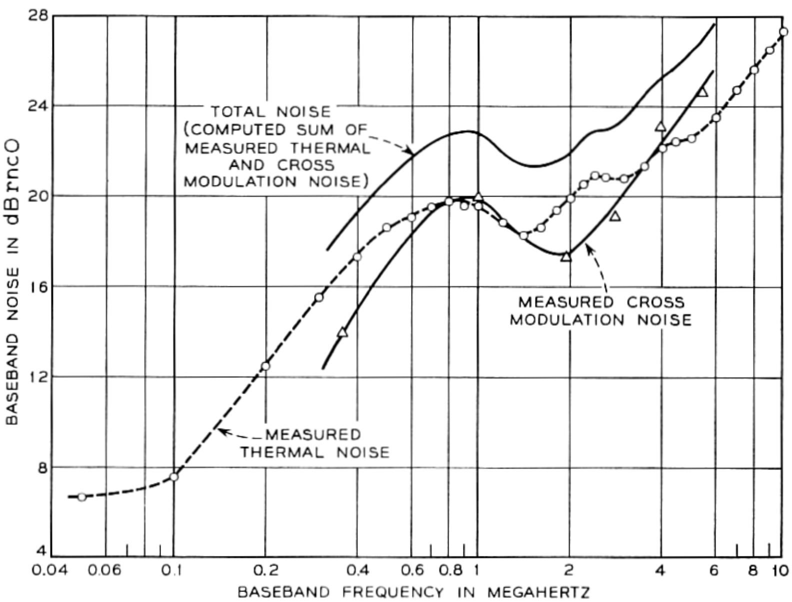

Fig. 10 — Thermal and cross modulation noise versus baseband frequency measured on a 10-hop loop of channel 9 of the initial installation. The channel was equipped with Schottky barrier diode receiver modulators.

the laboratory, indicated that the noise from this source should be approximately 17 dBrnc0 at the top end of the message band for a 10-hop system, which is about 9 dB lower than the original allocation. However, the field measurements were about 7 dB higher than this, and further studies are being made to determine the cause of this discrepancy. A large part of the difference appears to be caused by the multimoding problem in the antenna waveguide system. Calculations based on the analysis given in Ref. 9 indicate that the echoes caused by multimoding may contribute about 22 dBrnc0 noise in the top channel of a 10-hop system.

The envelope delay distortion characteristic measured on the 10-hop loop of channel 9 is shown in Fig. 11 and is typical of the initial in-

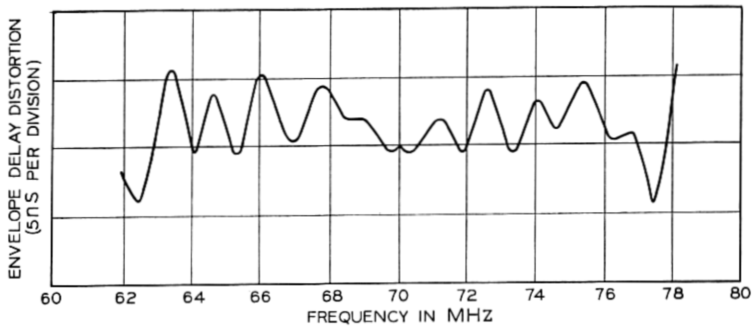

Fig. 11 — Envelope delay distortion measured on a 10-hop loop of channel 9 of the initial installation.

stallation channels. This characteristic was obtained after removing an envelope delay distortion slope component of about 2 ns/MHz using delay slope mop-up equalizers in the fifth and tenth microwave receivers of the 10-hop loop.* The predominant component is the ripple structure caused by the multimoding in the antenna waveguide system.

The results of the field measurements show that the total noise

* Some residual delay slope, principally from the channel separation networks in the immediately preceding repeater bay in the bay line-up, is expected in each channel. On the initial installation, the effect of the immediately adjacent bay was simulated, in the case of channel 9, by having the channel 8 separation networks installed ahead of the channel 9 bay in each line-up. As described in Ref. 5, provision is made in each microwave receiver for a mop-up delay slope equalizer to permit distributing the required delay slope equalization along the radio line.

objective for the system has been met for microwave receivers equipped with the Schottky-barrier diode receiver modulator. Some trade-off has been made, however, between the thermal and cross-modulation noise of the radio line compared with the original allocations. Table I estimates the noise in the top message circuit contributed by the radio line portion of a 4000-mile system. The radio line is assumed to be composed of 140 radio hops. The thermal noise estimate is based on the initial installation measurements increased by 1.5 dB to allow for a typical received carrier power. The cross-modulation noise estimate assumes power addition of the noise and is based on the initial installation measurements. Table I also shows the performance estimated for a radio line equipped with the 10.5 dB noise figure receiver modulator—IF preamplifier.

It is somewhat unrealistic to assume that a voice circuit traversing 4000 miles might be located at the top of the message band for the entire distance. Some frequency frogging is almost certain to occur at one or more of the intermediate multiplex-equipped terminal stations. Since the noise performance of the top message circuit usually is the poorest of any frequency in the message band, the actual noise performance of a typical 4000-mile voice circuit may be somewhat lower than that indicated in Table I.

5.2.3 *Baseband Amplitude Response*

A typical baseband amplitude response for a 10-hop radio channel is shown in Fig. 12. This measurement was made point-by-point on channel 9 on the Alexander-Arkabutla initial installation route. The test equipment was located at Alexander, and the two-way channel was interconnected at IF at Arkabutla to form a 10-hop loop. The baseband test tone was applied to the input of the FM terminal trans-

TABLE I—ESTIMATED NOISE IN TOP MESSAGE CIRCUIT FOR THE RADIO LINE PORTION OF A 4000 MILE SYSTEM

Condition	Noise (dBrnc0)		Total
	Thermal	Cross modulation	
8.5 dB noise figure repeaters (Schottky-barrier diode modulators)	36	36	39
11.5 dB noise figure repeaters (balanced receiver modulators)	39	36	41

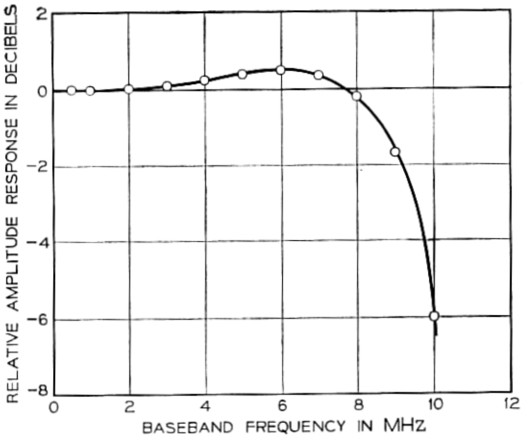

Fig. 12 — Baseband amplitude response measured on a 10-hop loop of channel 9 of the initial installation.

mitter at approximately 14 dB below that required for normal (±4 MHz peak) deviation. This test level, which was chosen primarily on the basis of test equipment considerations, is sufficiently low to make negligible the power in the second and higher order sidebands. The contribution of the FM terminals and the test equipment has been subtracted from the over-all result to obtain the characteristic shown.

As previously mentioned, random amplitude variation across the message band must be kept small to prevent instantaneous level changes from producing hits when the IF signal is switched between the regular and protection channels. The field measurements show an average variation of about 0.3 dB peak-to-peak over the message band for the radio line portion of a 10-hop system. There is no evidence of any systematic ripple or broad shape over this frequency range. This performance, although somewhat worse than the ±0.25 dB objective given in Section II, is about the same as the best performance that can be obtained in TD-2 and TH under very careful and frequent maintenance conditions.

Roll-off of the baseband response commences above 6 MHz. For 10 hops, the response is down, on the average, by about 0.5 dB at 8 MHz, 2 dB at 9 MHz, and 6 dB at 10 MHz. This roll-off is caused principally by the amplitude distortion of the microwave networks and IF band-pass filters of each hop and adds systematically in the system.

5.2.4 *Intrasystem Interference*

The Alexander-Arkabutla initial installation was tested to evaluate all of the basic types of interferences that might be found within the system: co-channel interference, adjacent channel interference, direct adjacent channel interference, and image channel interference. Also measured was the interference arising from the presence of tones 10 MHz away from the channel center frequency, which originates as a consequence of using a hetrodyne type repeater. In addition, third-order carrier intermodulation (2A-B and A+B−C type) in the microwave receivers was checked.

The tests showed that all of these basic types of interference are negligible provided that at least 85 dB antenna side-to-side coupling loss and 20 dB cross polarization discrimination are obtained. Lower values of coupling loss or cross polarization discrimination can result in excessive near-end or far-end interference, respectively, during fading. This excessive interference is of two types: (*i*) a form of direct adjacent channel interference which appears in the message band, and (*ii*) adjacent channel interference caused by the second order sidebands of the adjacent channel which disturb the 9 MHz region monitored by the 100A protection switching system noise detector. The side-to-side coupling loss specified is that required for an average hop having a normal received carrier power of −28 dBm, corresponding to 65 dB section loss. For the same performance, the recommended minimum side-to-side coupling loss must be increased 1 dB for each dB that the section loss exceeds 65 dB to maintain a minimum carrier-to-interference ratio of 20 dB.

The other possible interferences in the system are well under control, in many cases for appreciably lower values of side-to-side coupling loss or cross polarization discrimination. This good performance in many cases results from the selectivity that has been designed into the microwave transmitter and receiver, the suppression of local oscillator signals by modulator balance and filtering, relatively tight frequency control on the various frequency determining oscillators in the system, and the considerable attention that was given during equipment design to minimizing leakages. The 60 dB or greater front-to-back ratio normally obtained with the horn-reflector antenna is the principal factor in controlling intrasystem co-channel interference.[18]

5.2.4.1 Direct Adjacent Channel Interference. Direct adjacent channel interference is defined as that form of interference in which the signals transmitted on the adjacent channel appear in the interfered-with channel as intelligible crosstalk at the same baseband frequencies that these signals occupy in their own channel. Characteristics of this type of interference for tone modulation on the adjacent channel have been described for the TH[2] and TD-2 systems.[1] The TH measurements showed that the greatest interference occurred from tones transmitted at the top of the message band in the adjacent channel.

Similar results were obtained in TD-3. For example, laboratory measurements showed that when both the normal received carrier and an equal power adjacent channel carrier were present at the input to the channel separation network of a microwave receiver, (a condition that would exist if the side-to-side coupling loss were equal to the section loss) a 0 dBm0 6 MHz tone on the adjacent channel would appear at about −65 dBm0 at 6 MHz in the disturbed channel.* On the other hand, for the same carrier power condition, a 0.5 MHz tone on the adjacent channel produced about −85 dBm0 at 0.5 MHz in the disturbed channel.

A fade in the disturbed channel, relative to the adjacent channel, results in increased interference. For example, when the disturbed channel was faded 20 dB, the interference at 6 MHz caused by the 0 dBm0 6 MHz tone on the adjacent channel increased to about −50 dBm0. For deeper fades, the interference increased 2 dB for each dB that the fade exceeded 20 dB. Thus, for a 40 dB fade, the interference at 6 MHz reached a completely intolerable level of about −10 dBm0. Similar characteristics and increases were measured at 0.5 MHz when the 0.5 MHz 0 dBm0 tone was transmitted on the adjacent channel.

From the above results it is evident that the tone interference caused by a 0 dBm0 tone in the adjacent channel can become unacceptably high when the disturbed channel is faded. As explained, these results refer to a starting condition in which the disturbed and adjacent channel carriers are of equal power at the input to the channel separation network of the disturbed channel receiver. To reduce the interference, the power of the adjacent channel carrier at the receiver input must be reduced. If, for example, during the non-

* Reference 1 expresses the direct adjacent channel interference results in terms of an equal level crosstalk ratio. This ratio is the difference, in dB, between the tone measured on the disturbed channel and the tone applied to the adjacent channel when both tones are expressed in terms of the same transmission level point in the baseband of the system. Thus, the test result given here defines an equal level crosstalk ratio of −65 dB.

faded condition the disturbed channel carrier power is 20 dB greater at the receiver than the adjacent channel carrier power, a 40 dB fade on the disturbed channel will result in the same interference at 6 MHz as previously noted for a 20 dB fade (that is, about −50 dBm0 or 40 dBrnc0). Since for a 40 dB fade the noise of the faded channel is approaching 55 dBrnc0, it follows that the recommended 20 dB disturbed-to-adjacent-channel carrier ratio for the nonfaded condition is sufficient to make negligible any interference from 0 dBm0 tones transmitted on the adjacent channel.

A noise load, simulating a 1200 channel message and transmitted at normal (±4 MHz peak frequency) deviation on the adjacent channel, produced much more significant interference. It was found that the interference in the disturbed channel was substantially greater at the low end of the message band than at the high end, in contrast with the observations made with tone modulation on the adjacent channel. A nosie load causes an interference which cannot be interpreted in the measurement either as intelligible crosstalk (as direct adjacent channel interference is intended to define), or as unintelligible crosstalk. The interference at the low end of the message band can become extremely great, but it is doubtful that it would constitute intelligible crosstalk. It seems reasonable, however, to refer to this interference as a form of direct adjacent channel interference to distinguish it from ordinary adjacent channel interference. Further work is required to explain the observations that have been made and to define the nature of the interference.

Figure 13 shows results of a noise load type interference measurement made at 0.5 MHz in the baseband of a 2-hop laboratory system. In this test, a noise-loaded, adjacent channel carrier was coupled into the input of the channel separation network for the first hop's microwave receiver. The measured interference is shown as a function of the fade depth introduced on the disturbed channel carrier in the interfered-with hop. For perspective, notice that at normal deviation the noise load on the adjacent channel simulates in each message circuit a "signal" power of approximately 71 dBrnc0. Fig. 13 shows that if the ratio between the nonfaded disturbed carrier and the adjacent channel carrier is 10 dB or less, a 40 dB fade on the disturbed carrier results in an interference power in that channel which is higher than the "signal" power on the adjacent channel.

The results of similar measurements made at 5.6 MHz in the baseband of the disturbed channel are shown in Fig. 14. Comparison with Fig. 13 shows clearly that for low disturbed-to-adjacent channel car-

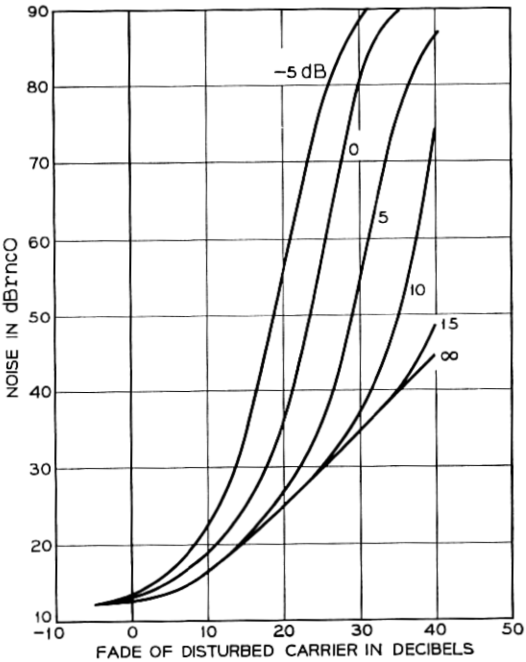

Fig. 13 — Interference, measured at 0.5 MHz in the baseband of the disturbed channel, resulting from a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier power.

rier ratios, considerably more interference exists at the low end of the message band than at the high end, when deep fades occur on the disturbed channel.

The 100A protection switching system initiator, which monitors the noise at 9 MHz in the baseband of the radio channel, is adjusted to initiate a switch of the channel to a standby protection channel if the noise at the top of the message band (5.772 MHz) reaches 55 dBrnc0. This adjustment assumes that:

(*i*) For deep fades, the thermal noise contributed by the faded receiver will be the controlling noise in the baseband of the faded channel.

(*ii*) As a result of (*i*), the noise in the message band will be greatest at the top of the message band during deep fades.

(*iii*) There is a predictable relationship between the noise at the top of the message band and the noise monitored at 9 MHz by the switching system initiator.

It is apparent from Figs. 13 and 14 that for low carrier ratios, interference rather than the receiver thermal noise can be the controlling noise in the message band during a deep fade of the disturbed carrier. Thus, if the noise at 9 MHz were to be determined only by the receiver thermal noise for these low carrier ratio conditions, excessively high interference could occur in the message band before a switch were requested. However, at these low carrier ratios, there is significant adjacent channel type interference appearing at 9 MHz in the disturbed channel resulting from the second order sidebands from the adjacent channel. This interference can appreciably reduce the fade depth for triggering the switch to a protection channel and thereby can prevent excessive direct adjacent channel type interference.

5.2.4.2 Adjacent Channel Interference. The adjacent channel interference (measured at 9 MHz in the disturbed channel) that results from the second order sidebands of a noise load signal on the adjacent

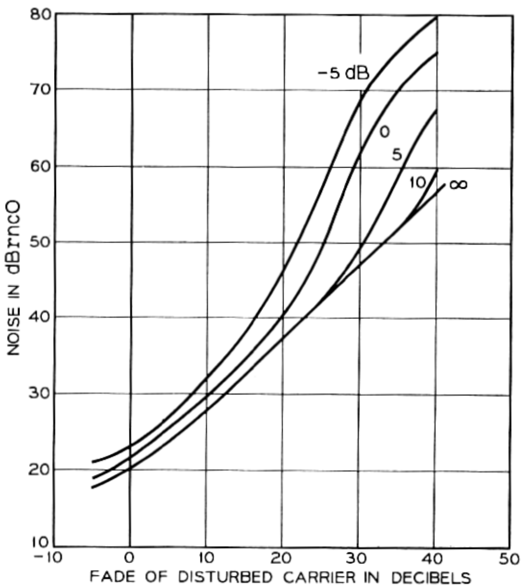

Fig. 14 — Interference, measured at 5.6 MHz in the baseband of the disturbed channel, caused by a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier power.

channel is illustrated in Fig. 15. The data for this figure were obtained during a series of controlled coupling loss measurements between adjacent channels on the initial installation and agree closely with calculated values. The noise increase, measured at 9 MHz at the end of a 10-hop loop of one of the radio channels, is shown in this figure as a function of fade depth introduced on one of the hops. The controlled coupling loss between this channel and the noise loaded adjacent channel was introduced at the input to the microwave receiver of the faded hop.

To illustrate the significance of the adjacent channel interference, assume that the initiator has been adjusted to request a switch for what corresponds to a 40 dB fade in the interfered-with hop for an ideal, no interference, condition. Figure 15 shows that for this condition (that is, infinite carrier ratio) the switch request would occur when the noise at 9 MHz increased 27 dB. If, however, on this hop a 0 dB ratio were to exist between the nonfaded disturbed carrier power and the adjacent channel carrier power, only a 22 dB fade would be required to produce the 27 dB increase in the 9 MHz noise and thereby initiate the switch request. In other words, the adjacent channel interference results in an 18 dB reduction in the fade margin of this hop.

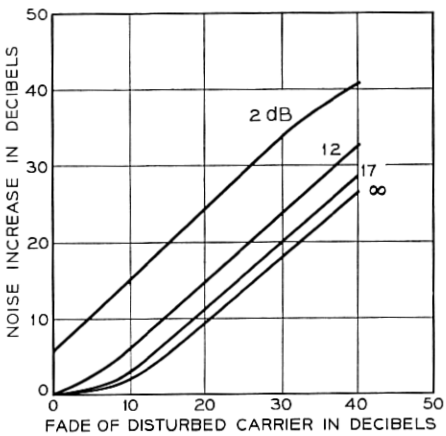

Fig. 15 — Interference, measured at 9 MHz in the baseband of the disturbed channel, resulting from the second order sidebands of a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier power.

Figure 13 shows that for a 0 dB carrier ratio and 22 dB fade, the direct adjacent channel type interference will have reached about 40 dBrnc0, which is well below the 55 dBrnc0 at which the message circuit would be considered noncommercial. Thus, as previously mentioned, it somewhat fortuitously turns out that because of the adjacent channel interference, a switch will be requested by the disturbed channel before excessive interference develops in the message band. However, the apparent benefit obtained in this way is offset somewhat by the increased switching that will occur whenever moderate fading develops on the interfered-with hop. The resultant usage of the protection channel will undoubtedly increase the probability that protection will not be available to a channel that might really need it during a deep fade. The net effect could be reduced system reliability.

This and the preceding section clearly indicate that one solution to a potentially serious interference problem is to make the interference negligible by engineering the antenna system for adequate coupling loss between the disturbed and adjacent channel carriers. The test results show that the interferences become negligible when the disturbed (nonfaded) to adjacent channel carrier ratio is about 20 dB or greater. This corresponds to 20 dB cross-polarization discrimination and an antenna side-to-side coupling loss 20 dB greater than the section loss. In general, typical antenna installations can meet these objectives. On some paths, however, careful attention must be paid to avoiding excessive foreground reflections if the side-to-side coupling loss objective is to be met.

5.2.5 *TD Intermix System Tests*

The primary reason for adding TD-3 channels to the existing Ennis to Seguin TD-2 route was to determine if the systems would interfere with each other. The channel assignments, mentioned earlier, interleaved TD-2 and TD-3 channels. The tests were made with the TD-2 microwave transmitters operating with one watt output.

No significant interferences were detected between the two systems owing to basic design features. Initially, some high level tones which appeared at baseband frequencies near 10 MHz were found in both systems. These tones were traced to leakage at defective, loose, or improperly assembled waveguide or coaxial connections, again demonstrating the need to control such leakage.

The noise and transmission performance of the TD-3 channels also were measured. The results were virtually the same as those for the Alexander-Arkabutla route reported in Sections 5.2.2 and 5.2.3.

5.2.6 *TV Transmission Performance*

The TD-3 radio system is expected to be used principally for transmitting message, data, and similar signals that can be multiplexed to use the approximately 6 MHz baseband capacity of the system efficiently. However, some occasional use for television undoubtedly will occur. A series of tests, made on the initial installation route to evaluate the system's TV transmission performance, included transmitting standard TV test signals, measuring differential gain and phase, and measuring the weighted signal-to-noise ratio.* All tests were made with 3A FM terminals and the same pre- and de-emphasis networks that are used for TV transmission in TD-2.[19] These networks provide approximately 13 dB shape across the video band. The tests were made at the Alexander station on 10-hop loops of each of the radio channels, and on 20-, 30-, and 40-hop loops formed by connecting the individual channel loops.

The TV test signals used were the standard multiburst and sine-squared signals. No significant distortion or axis shift was observed with the multiburst signal, nor was there any significant distortion with the $\frac{1}{4}$ or $\frac{1}{8}$ microsecond sine-squared pulses. A $\frac{1}{16}$ microsecond sine-squared pulse showed measurable widening and amplitude reduction, but this is to be expected from the relatively narrow bandwidth of the radio system compared with the pulse spectrum. The power spectrum in a $\frac{1}{16}$ microsecond pulse is down about 6 db at 8 MHz.

Differential phase measurements showed that for transmission through at least as many as 40 hops of radio, 1° or less of differential phase can be obtained at normal system deviation provided the radio system is delay slope equalized. The differential gain was approximately 0.15 dB for a 10-hop loop at normal system deviation. There was no evidence of systematic addition; for example, the differential gain measured for 30 hops was 0.16 dB. Objectives for the Bell System portion of a 4000-mile TV circuit are 3° differential phase and 2 dB differential gain. The field measurements indicate that the TD-3 system meets both objectives, assuming random addition of these distortions in the 4000-mile system.

* The weighted signal-to-noise ratio is defined as the ratio of the peak-to-peak signal to the weighted rms noise, where the weighting is a function of the frequency of the noise. A detailed discussion of noise weighting in TV performance evaluation is given in Ref. 20.

Low frequency (below 4 kHz) weighted noise measurements showed no contribution from the radio system up through 40 hops of radio. In other words, the 3A FM terminals are completely controlling in this frequency range. Monochrome weighting measurements, extrapolated to a 4000-mile (140 repeaters, 16 FM terminal pairs) system gave a weighted signal-to-noise ratio of 68 dB. For color weighting, the extrapolated weighted signal-to-noise ratio was 65 dB. Both of these results easily meet the 55 dB ratio presently required for the long haul portion of a TV network.

It was concluded from the results of the various TV transmission tests that the TD-3 radio repeater system, in combination with the 3A FM terminals, is capable of high quality television transmission.

VI. CONCLUSION

By early 1968 it is expected that about 2500 radio channel miles of TD-3 will be in commercial service. Considerable growth is anticipated during the next few years, and it is expected that the TD-3 facility eventually will carry a substantial part of the Bell System long distance circuit load. It is important, therefore, to use this new system with maximum efficiency, and to maintain a high level of performance at minimum cost. These objectives can be realized only with continuing effort in all pertinent areas.

In this era of rapid technological progress, new tools are continually becoming available. Unless these tools are properly exploited, a system such as TD-3 may become obsolete rapidly. Such a new tool is the Schottky barrier diode receiver modulator. This new modulator permitted removal of the parametric amplifier, accompanied by major reductions in receiver cost and complexity. Other devices and circuit designs are certain to appear which should be substituted where an economic advantage is realized.

With efficient use of the frequency spectrum becoming increasingly important, it is essential that radio systems be operated at maximum load capacity. If the side-to-side coupling losses of the antennas can be improved and if excellent cross-polarization discrimination can be achieved, it is possible that a moderate increase in circuit capacity (over the 1200 per radio channel) can be achieved. Work is already under way toward such antenna improvements. Related to load capacity is the selectivity of the microwave transmitter and receiver circuits. The effects of selectivity (and its distribution within the receiver and transmitter) on adjacent channel interference are not completely understood. Further studies are being made.

There is continuing effort to improve reliability (reduction of "outage" time) of radio systems. The use, in TD-3, of solid state devices for active circuit elements should greatly reduce outage time caused by active element failures. The system failure rate probably can be further reduced by developing new facilities for preventive maintenance. For example, it is believed that maintenance can be aided by automatically transmitting to maintenance centers detailed information on the condition of the equipment located in unattended stations. Recent development of a new alarm and order wire system* makes such an approach appear attractive, and its feasibility is being studied. Outage resulting from excessive path loss or fading is influenced by atmospheric conditions, path engineering, antenna characteristics, and the protection switching arrangement used. Possible improvements in path layout, antennas, and protection arrangements are being studied.

ACKNOWLEDGEMENTS

The development of the TD-3 system involved the contributions of many people. The authors of the articles in this issue, serving as reporters of the accomplishments, acknowledge the contributions of individuals not identified by name.

APPENDIX

Baseband Interference Caused by the Addition of a Tone to an FM Signal

Consider the desired FM signal,

$$S(t) = A_c \cos [\omega_c t + \phi(t)] \tag{1}$$

where

A_c = peak carrier amplitude
ω_c = carrier frequency in radians per second
$\phi(t)$ = angle modulation in radians

and an RF interfering tone,

$$I(t) = A_n[\cos (\omega_c + \omega_n)t + \theta] \tag{2}$$

where

A_n = peak interference amplitude
$\omega_c + \omega_n$ = frequency of interference in radians per second
θ = phase constant.

* The E1 status reporting and control system.

For the condition $A_c \gg A_n$, the combination of the FM signal and the RF interfering tone is given by[19]

$$M(t) \cong A_c\left\{1 + \frac{A_n}{A_c}\cos\left[\omega_n t + \theta - \phi(t)\right]\right\}$$
$$\cdot \cos\left\{\omega_c t + \phi(t) + \frac{A_n}{A_c}\sin\left[\omega_n t + \theta - \phi(t)\right]\right\}. \tag{3}$$

The carrier has been modulated by a phase function $\psi(t)$, which can be expressed as

$$\psi(t) = \phi(t) + \psi_1(t) \tag{4}$$

and it includes the original desired modulation, $\phi(t)$, plus the interference term

$$\psi_1(t) = K \sin\left[\omega_n t + \theta - \phi(t)\right] \tag{5}$$

where K is the RF interference-to-carrier ratio.

For this analysis it will be assumed that the FM signal consists of a test tone such that

$$\phi(t) = X \sin(qt + \alpha) \tag{6}$$

where

X = modulation index of test tone in radians
q = baseband frequency of test tone in radians/sec.
α = phase constant.

Equation (5) becomes

$$\psi_1(t) = K \sin\left[\omega_n t + \theta - X \sin(qt + \alpha)\right]. \tag{7}$$

In order to determine the separate frequency components of the interference term, the Bessel function is introduced.

$$\psi_1(t) = K \sum_{m=-\infty}^{\infty} J_m(X) \sin\left[\omega_n t + \theta - m(qt + \alpha)\right]. \tag{8}$$

After FM demodulation, the baseband voltage output is proportional to the frequency deviation given by

$$\frac{d\psi_1(t)}{dt} = K \sum_{m=-\infty}^{\infty} (\omega_n - mq) J_m(X) \cos\left[\omega_n t + \theta - m(qt + \alpha)\right]. \tag{9}$$

It will be assumed that $X \ll 1$, so that the only significant baseband interference tones occur when $m = 0$ and ± 1. That is

$$\begin{aligned}\frac{d\psi_1(t)}{dt} \cong K(\omega_n J_0(X) \cos [\omega_n t + \theta] \\ + (\omega_n - q)J_1(X) \cos [(\omega_n - q)t + \theta - \alpha] \\ + (\omega_n + q)J_{-1}(X) \cos [(\omega_n + q)t + \theta + \alpha]. \qquad (10)\end{aligned}$$

Introducing the approximations

$$J_0(X) \cong 1$$

$$J_1(X) \cong \frac{X}{2}$$

$$J_{-1}(X) \cong -\frac{X}{2}$$

$$\begin{aligned}\frac{d\psi_1(t)}{dt} \cong K\omega_n \cos [\omega_n t + \theta] \\ + K(\omega_n - q) \frac{X}{2} \cos [(\omega_n - q)t + \theta - \alpha] \\ - K(\omega_n + q) \frac{X}{2} \cos [(\omega_n + q)t + \theta - \alpha)].\end{aligned}$$

This expression shows that there is a baseband interference tone at the "beat" frequency between an RF interference tone and the signal carrier, and also that there are baseband interference tones located q radians per second on either side of the "beat" tone.

Figures 16 and 17 show the relative frequencies and magnitudes

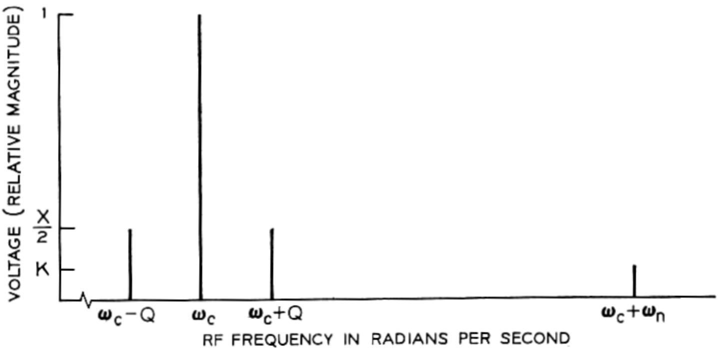

Fig. 16 — Spectrum at RF.

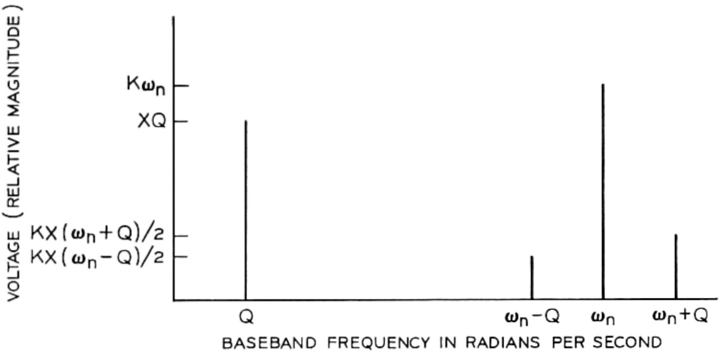

Fig. 17 — Spectrum at baseband.

of the tones at RF and baseband, respectively. Second order side bands have been neglected.

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