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TH-3 Microwave Radio System:

System Considerations

By R. M. JANSEN and R. C. PRIME

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This paper gives a general description of the TH-3 long-haul microwave radio system which operates in the 6 GHz common carrier band. Performance objectives are expressed, and the allocation of noise to the various contributors is developed. Particular attention is given to the balance between intermodulation noise and tertiary interference. In conclusion, the predicted performance is compared with measurements made on the initial installation.

I. INTRODUCTION

The TH-3 microwave radio relay system is a new long-haul facility designed to provide modern solid state equipment with improved performance for use in the 6 GHz band. When work began on the project in 1966, most of the Bell System radio traffic was carried on the 4 GHz TD-2 system. Also, at that time the first TD-3[1] radio route was being installed. Since TH-3 routes can be installed most economically as an "overbuild" on existing TD-2 and TD-3 (TD radio) routes, it was clear that TH-3 must be compatible with TD-2, and as much like the new TD-3 as possible. The features of TH-3 which differentiate it from its predecessor in the 6 GHz band, the TH-1 radio system,[2] and make it more like TD radio are the individual transmitter-receiver bays, and the 70 MHz IF. The latter increases certain difficulties with interchannel interference but allows TH-3 to use the FM terminals, switching systems, and some of the IF circuits developed for TD-3. Also, by adopting the 70 MHz IF, a combined TD and TH-3 facility with crossband protection switching looks feasible at this time.

Although TH-3 was originally conceived as a long-haul, multichannel system with a 4000-mile capability, the need to adapt it to shorter routes with a smaller cross section of circuits was quickly recognized. The so-called medium-haul TH-3 then came into being. It is discussed in later articles of this issue.[3,4]

Many of the design techniques which were assimilated during the TD-3 development were directly applicable to TH-3. Thus the designers of TH-3 had a headstart with regard to available circuit designs and to understanding how superior transmission performance could be achieved. However, TH-3 did bring significant advances to the art of radio repeater design. In particular, microwave integrated circuits, microwave envelope delay equalizers, and a means of preventing adjacent channel interference called RF squelch are key features of the TH-3 repeater.

The measuring capabilities and analysis tools that were available at the start of the project were a valuable asset. For example, the computer-operated transmission measuring set[5] was invaluable for its precise measurements of amplitude and delay distortion. Theory relating distortions to intermodulation noise was already available,[6–9] and computer programs developed for TD-3 were used initially to determine the distortion requirements. However, these programs could not conveniently handle some commonly occurring distortions (such as fourth and higher-order envelope delay distortion). As an important step in the TH-3 development, a computer program was written based on new mathematical techniques[10] which gave the designers adequate means to relate system distortion to intermodulation noise.

II. SYSTEM OBJECTIVES

The circuit objectives for noise, tones, and reliability are consistent with the current Bell System objectives for long-haul facilities.

(*i*) *"Worst circuit" noise of 41 dBrnc0 for a 4000-mile system during periods of nonfaded transmission.* The noise may increase to 55 dBrnc0 during fading, after which the channel will be switched automatically to a protection channel.

(ii) *Single-tone interference of −68 dBm0 maximum in any voice circuit of a 4000-mile (41 dBrnc0) system during nonfaded transmission.* Subjective tests have shown that if the noise-to-tone-power ratio in a message circuit is constant, the tone is less discernible when the noise power increases. The result is that the requirement for those baseband tones which increase dB for dB with fading is −47 dBm0 when the noise in the circuit is 55 dBrnc0 (during a 40 dB fade). Under normal conditions this corresponds to a −87 dBm0 requirement.

(iii) *Reliability of 0.02 percent per year for a two-way, 4000-mile system, as measured by accumulated yearly outage.*

The following are particular design objectives for TH-3:

(iv) *1800 message circuits per channel.*

(v) *10 MHz baseband.* The bandwidth is adequate to transmit 1800 message circuits or high-definition TV.

(vi) *Baseband response of ±0.25 dB flatness over the message band from about 5 kHz to 8.5 MHz for each radio channel on an IF protection switching section. Also, no more than 30 degrees phase difference (at baseband) between the radio channels of a protection switching section.* These objectives prevent hits on data transmission and disturbance to TV when the signal is transferred between regular and protection channels. To meet the above phase objective, the absolute delay of all radio channels in a switching section must be equal to within 10 ns, which corresponds to 6.5 feet of IF cable.

(vii) *75 ±20°F Operating Temperature Range.* In long-haul radio stations the temperature in the vicinity of the TH-3 radio bays will be maintained within these limits. As a safeguard against loss of temperature control, the system should continue to operate with little degradation at ambient temperatures between 40°F and 120°F. The latter temperature range is specified as the operating temperature range for medium-haul TH-3.

III. SYSTEM MODEL AND NOISE ALLOCATIONS

Besides the actual T-R bays, the components of the TH-3 and TD-3 systems are almost identical. Figure 1 is a block diagram of a TH-3 system. The number and sequence of the various components in the transmission path of a 4000-mile system has a first-order effect on the performance. Therefore to make system calculations, a system model was constructed which in the best judgment of the designers would

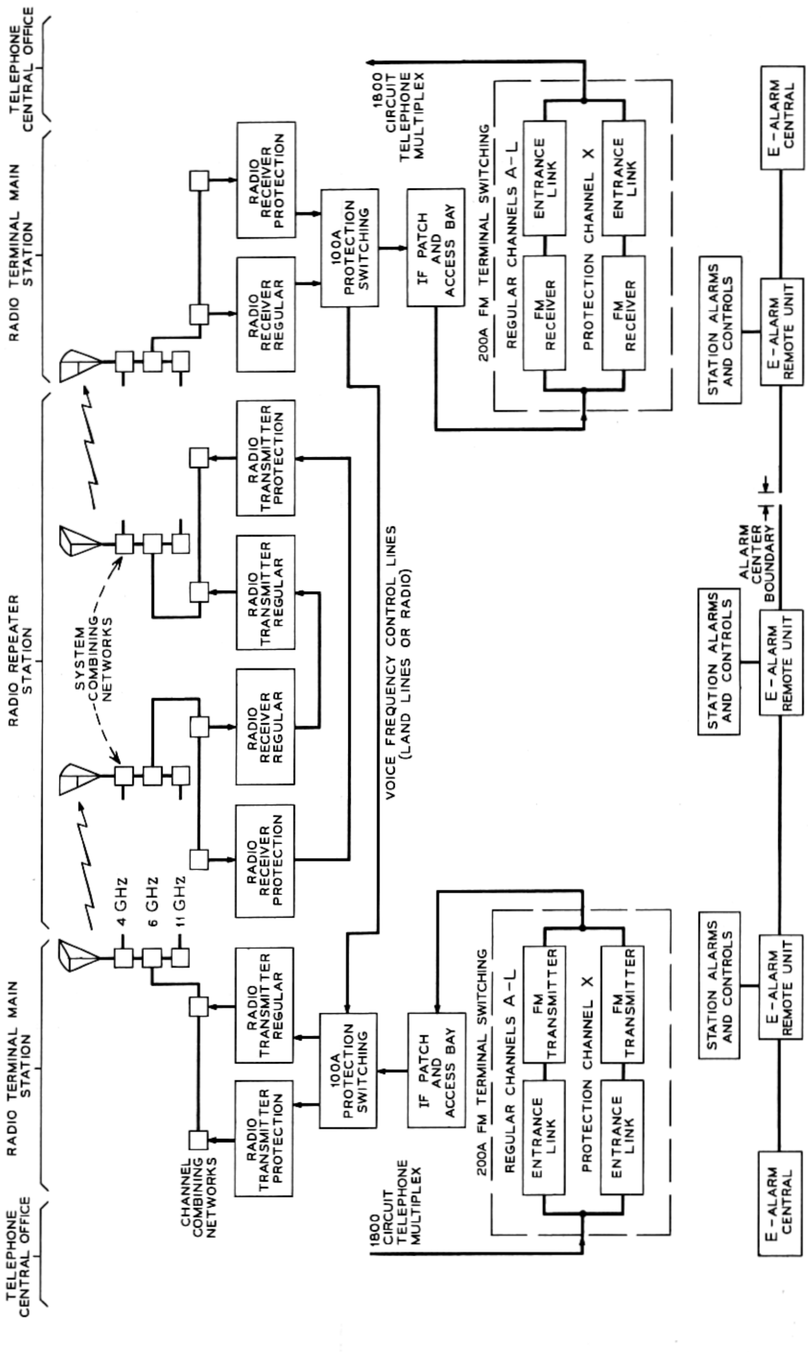

Fig. 1—Block diagram of TH-3 system.

represent the nominal TH-3 system. The principal assumptions made in the model were based on the conditions existing in the field.

(*i*) 150 hops in 4000 miles.
(*ii*) 51 main stations with IF switching, and 100 repeater stations.
(*iii*) 17 of the 51 switching main stations will also have L-Multiplex terminals. These stations, called terminal main stations, have by necessity FM terminals and wire line entrance links. Thus, the model 4000-mile system is divided into 16 parts and they are assumed to be connected at the channel, group, or supergroup level.
(*iv*) The path loss of a nominal hop is 63.2 dB from the transmitter test access on the TH-3 T-R bay to the receiver test access. (The locations of these access points in the T-R bay are explained in the following article.)[11]

During the course of the development the 41 dBrnc0 allocated to the total system was divided between the system components in several different ways. The final allocation is shown in Fig. 2.

3.1 *Multiplex*

MMX-2R mastergroup multiplex and LMX-2 multiplex terminals provide the 1800-message circuit load. The baseband signal extends from 564 kHz to 8.524 MHz.

3.2 *Wire Line Entrance Link*

The 3A entrance link was described in Ref. 1. Since that writing the link has been improved to handle 1800 circuits. Also, the maximum length of the entrance line has been increased to eight miles by employing an intermediate repeater.

3.3 *FM Terminals*

Before the advent of the 4A FM transmitter, 35 dBrnc0 had been allocated to the sixteen 3A FM transmitter-4A FM receiver pairs. The

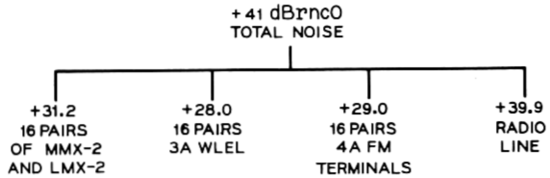

Fig. 2—Allocation of total noise objective.

4A FM terminals[12] have improved noise performance and sixteen pairs are allocated only 29 dBrnc0.

3.4 *Protection Switching*

The 200A protection switching system protects the FM terminals and WLELs but need not be dedicated to TH-3. The twelve working channels may be divided between TD-2, TD-3, and TH-3 systems and the single protection channel is provided with switched pads to match its transmission levels to those of the system it is protecting. The 100A protection switching system[13] protects three hops on the average, although ten-hop switching sections are not uncommon. Minor modifications were made to the 100A system to adapt it to TH-3, as explained in Section 4.1.4.

3.5 *IF Patch and Access Bay*

The terminal main stations have an IF patch and access bay. This bay is common to all heterodyne radio systems in the station which have a 70 MHz IF. In a typical large station, the IF trunks from several TH-3, TD-2, and TD-3 radio lines will fan into the IF patch and access bay where they will be assigned to FM terminal equipment within one of several 200A protection switching systems or, in the case of through channels, routed to outgoing radio lines. The purpose of the bay is to provide an easy means to reassign IF trunks and to have a common location to monitor all the IF trunks in the station. IF trunks carrying TV are also brought through the patch bay and, where necessary, connections are made to the television operating center through an FM terminal and special WLELs.

3.6 *Alarm System*

The C1 alarm system may still be used with TH-3, but the E-type alarm is now recommended. The E alarm uses rugged digital coding and has the potential to monitor remotely the meter readings on the T-R bays.

3.7 *Power*

Most of the subsystems which compose TH-3 are operated from a −24 volt power plant. However, there is a need for + 130V and +24V in some alarm and switching circuits. In particular, both −24V and +24V plants are used with the 100A protection switching system. Other subsystems needing +130V and +24V may be supplied from the −24V supply by means of dc-to-dc converters.

3.8 *Radio Line*

The radio line is that part of the system from the first microwave radio transmitter to the last radio receiver inclusive. Thus, it includes all of the radio T-R bays and the antenna systems. Also, when considering the total noise allocation for the radio line, interference picked off the air by the receiving antennas must be taken into account. The noise allocations for the radio line are shown in Fig. 3 and are discussed below.

(*i*) *Co-channel interference* is the same as for TD-3 and has been dealt with adequately.[14]
(*ii*) *Intersystem interference* has been included because of the high usage of the 6 GHz band. The allocation is of the same order as co-channel interference.
(*iii*) *RF echoes in the antenna systems* were allocated only 6 dBrnc0 per hop. The allocation is nevertheless realistic as a result of the antenna improvement program[15] which was started in 1964. This effort was directed mainly at reducing echoes due to mode conversion.
(*iv*) *Echoes in IF trunks* were given essentially the same allocation as in TD-3.
(*v*) *Tertiary interference and noise from the microwave T-R bays* are dealt with in the following sections.

IV. DESIGN ASPECTS OF THE TH-3 RADIO LINE

4.1 *General*

The first step in designing the system was to choose the basic frequencies: RF, IF, and baseband, on which the TH-3 design was founded. The factors involved in the decision-making process are discussed in the following paragraphs.

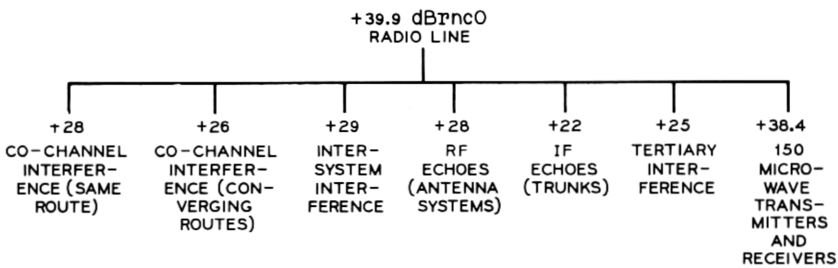

Fig. 3—Noise allocations for the radio line.

4.1.1 *RF Channel Assignments*

The TH-3 system was designed initially to operate on sixteen channel frequencies. These frequencies are consistent with the TH-1 channel assignments[2] because of the obvious need to coordinate with TH-1 and other users of the 6 GHz common carrier band.* However, the TH-1 auxiliary channel frequencies are not used since for most long-haul applications 4 GHz auxiliary channels will be available for order wire, switching, and alarm purposes. Later in the TH-3 development the need for TH-3 on the so-called staggered frequencies (which are interleaved with the CCIR recommended frequencies) was recognized, and the design was augmented with sixteen alternate channel frequencies. The staggered frequency plan will be used on routes where it is necessary to coordinate with systems already operating on the staggered frequency plan. Both of the RF frequency plans are shown in Fig. 4. The use of the staggered frequencies is not without some disadvantage. Channel 28S† cannot be squeezed into the common carrier band and consequently the route may grow to only seven instead of eight radio channels. It is some consolation that there is a channel 20S which, along with 18S, can be substituted for another channel in the event that this other channel is blocked. Unfortunately a channel cannot be made up from alternate hops of 18S and 20S because they are so close together in frequency that the mutual interference would be intolerable.

4.1.2 *Microwave Carrier Frequencies*

All of the microwave carrier‡ frequencies in TH-3 are generated below the RF signal frequency. This action was taken to avoid stability problems with the lower sideband upconverter. With the chosen microwave carrier arrangement it was computed that the RF selectivity was adequate to prevent image interference from radars below the common carrier band. Another matter requiring attention was the need to shield components sufficiently so that near-end image interference at high-low repeater stations would not be caused by RF leakage between bays.

* The TH-1 broadband channel assignments are recommended by the CCIR (REC 383-1, Documents of the XIIth Plenary Assembly, New Delhi, 1970, Vol. IV, Part I, p. 84) and they are an accepted standard within the United States.

† Staggered channels are denoted by the letter S, and regular channels by the letter T.

‡ The microwave carriers are generated by the microwave generators and shift oscillators in the T-R bays. These carriers convert IF to RF in the transmitter modulator and RF to IF in the receiver modulator.

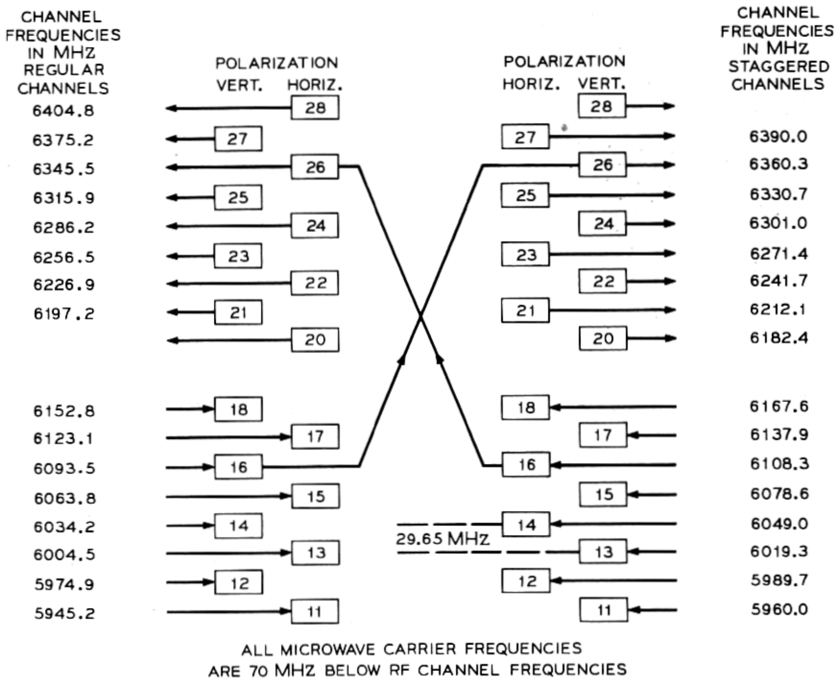

Fig. 4—TH-3 regular and staggered frequency assignments (low-high station).

4.1.3 *Growth Sequence of Radio Channels*

The standard growth sequence for the regular channels is 4T,† 8T, 2T, 6T, 3T, 7T, 1T, 5T. This sequence is the same as for TH-1. It was chosen because the preferred order of growth for TM-1 is the reverse of that for TH-1 and therefore conflicts with short-haul systems are more likely to be postponed. The even channels are on one polarization, with channel 4T nearest the antenna, and the odd channels are on the other polarization with channel 3T nearest the antenna. Four antennas are required at a repeater station, and each antenna is connected exclusively to either receivers or transmitters. Normally channels 8T and 1T will be the protection channels, although any two channels can be used. If the staggered frequency plan is used, the standard growth sequence will be 1S, 5S, 3S, 7S, 2S, 6S, 4S. The even channels are equipped last because there is no channel 8S.

† Channel 4T is short for 14T or 24T.

4.1.4 *Switching System Parameters*

The original 100A designed for TD systems has a protection pilot at 7 MHz and a noise monitoring slot at 9 MHz. Both of these frequencies were changed in TH-3 because the TH-3 baseband extends to 10 MHz. The noise slot was moved to 10.2 MHz to be above the signal, but well below a 10.7 MHz tone which will be present on most TH-3 channels. The protection pilot was placed at 11.88 MHz, well above the tone. The pilot frequency was chosen so that the second-order sidebands of the pilot would fall nominally between the second and third mastergroups on the adjacent channels. However, the level of the sidebands is such that it should not matter if the sidebands do fall into mastergroup 2 or 3.

4.1.5 *Modulation Parameters*

In the preliminary calculations of thermal noise it was assumed that TH-3 would use the same peak frequency deviation as TD-3. Experience with the system has shown no reason to change this parameter. The average busy hour speech load produces an rms frequency deviation of 798 kHz or the per circuit rms frequency deviation is 118 kHz. In Bell System terminology the equivalent sine wave power (explained in Ref. 16) of the 1800-message circuit load is +27.6 dBm0.

The TD-3 pre-emphasis (see Fig. 4 of Ref. 1) is widely used by TD-2 now that the capacity of that system has been expanded to 900 and 1200 message circuits. The same characteristic was chosen for TH-3 to simplify the 200A protection channels. (The pre-emphasis and de-emphasis networks are located within the 200A switching system, and a particular 200A system might include TD-2, TD-3, and TH-3 channels within the same switching group.) This pre-emphasis has a 3 dB advantage at the top message circuit (8.524 MHz) relative to the 4.6 MHz reference (or crossover) frequency.

4.2 *Thermal Noise*

The allocation for the 150 microwave transmitters and receivers was originally lower than specified in Fig. 3. As the development progressed it became evident that the thermal noise would exceed its portion of the allocation, principally because an optimistic view had been taken of the losses in waveguide networks and antenna systems. Relief came from two directions. Firstly, the improved performance of the 4A FMT permitted the bay allocation to be raised to 38.4 dBrnc0. Secondly, the improvements in the IF filter design made it possible to reduce the allocation for intermodulation noise, and as a consequence the allocation for thermal noise could be raised to 36.9 dBrnc0 (or 15.1 dBrnc0 per hop). This was a reasonable objective, still low enough to assure a 40 dB fade range.

4.3 *Intermodulation Noise*

Excellent intermodulation noise performance was achieved by adhering to the following principles in the design of the T–R bay:

(*i*) Using wideband active devices and achieving the required selectivity with passive networks;
(*ii*) controlling carefully the distortion of all networks and equalizing as close as possible to the source of the distortion. This minimizes the effects of AM-to-PM conversion and the need for mop-up equalization;
(*iii*) maintaining return losses above 30 dB in almost all cases; and
(*iv*) blocking harmonics of the IF where they could give rise to echolike distortions.

It was found that the RF channel combining and separating networks in the T-R bays caused significant envelope delay distortion (EDD) to those thru-channels on the same polarization which were closest in frequency. The total EDD introduced into a channel by the channel combining and separating networks of the channels 59.3 MHz above and below is shown in Fig. 5. This EDD is very well controlled and is essentially the same for all channels. Therefore, it was

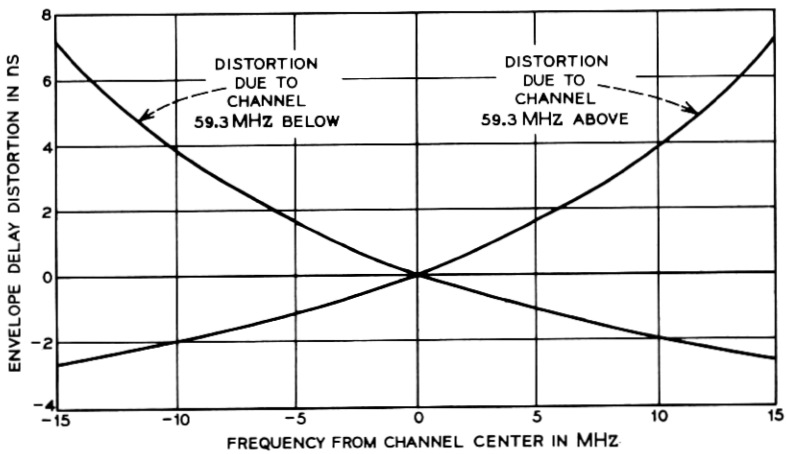

Fig. 5—Envelope delay distortion due to channel networks in channels 59.3 MHz above and below channel of interest.

possible to equalize it almost exactly at IF in every radio receiver. There are different situations according to which other channel networks are in the path of the channel of interest. The alternatives are: no equalization, equalizers to match either one of the EDD characteristics in Fig. 5, or an equalizer to compensate for both adjacent channels in which case the slope components cancel and only parabolic equalization is required.

As a consequence of these design techniques only a small amount of EDD slope is necessary to mop up TH-3 switching sections. No parabolic or other shape of EDD mop-up equalizer is required.

In the TH-3 system model it is assumed that a 4000-mile circuit is demodulated to baseband sixteen times. Since the reason for demodulating to baseband is to drop and pick up circuits, it was assumed that the composition of the circuit load is sufficiently different on each of the sixteen sections that intermodulation noise contributions from the individual sections add on a 10 log law* (power addition). Early measurements on preproduction T-R bays in the laboratory indicated that the intermodulation noise of a simulated multihop section was accumulating on an almost systematic 19 log law. Thus, a 14 log law represented the equivalent smooth accumulation of intermodulation noise for the overall system.

It was discovered that the 1009A† IF bandpass filter in the radio receiver was mainly responsible for the near systematic addition. Actually, the noise contribution of one filter was small compared to the thermal noise of a hop, but on long systems the noise from the filters was dominant. Further, it was clear that the original allocation of 34.2 dBrnc0 for the intermodulation noise of the T-R bays could not be met. The requirements for the filter were carefully reviewed and the conclusion was that too much selectivity was being specified at the expense of inband distortion to the transmitted signal.

The design of a new IF bandpass filter, designated the 1044A, was then undertaken. Its specifications were carefully related to intermodulation noise with the aid of the computer program referred to earlier. The filter was made as broad as possible so that the inband distortion was minimized. The objective was to hold the noise contribution of an individual filter to 0 dBrnc0 so that it would be satisfactory even if it caused systematic addition. However, the wide

* A 10 log law means that the total noise from N hops is 10 log N dB greater than the noise from one hop.

† The amplitude versus frequency characteristics of the 1009A and 1044A IF filters are shown in the following article.[11]

bandwidth of the filter gave rise to an additional source of noise by a mechanism known as tertiary interference.

The new filter was successful and the noise allocation for the intermodulation noise of the T-R bays was reduced to 33 dBrnc0 with confidence. However, tertiary interference, which had been negligible before, was allocated 25 dBrnc0. Although the net change in the allocations was small, the ability of the system to operate within them was vastly improved by the introduction of the 1044A filter.

4.4 *Tertiary Interference*

The mechanism of tertiary interference is well known. Ref. 2 explains how tertiary interference gave potential tone problems in the TH-1 system. In TH-3 the problem was not that of transferring a single tone from one channel to another, but of directly superimposing the message modulation of one channel onto the message sidebands of the next-to-adjacent channel. In previous systems the selectivity of the channels has been so great that the center channel greatly attenuates the tertiary path. However, in TH-3 a repeater will transmit some of the first-order sidebands of the adjacent channels, as shown in Fig. 6. Consequently, tertiary interference appears as a low level of noise in the disturbed channel. In the future, as more and more circuits are transmitted over systems using the established frequency plans, tertiary interference will be an increasingly important consideration. It is shown in the Appendix how the 25 dBrnc0 allocation was derived for tertiary interference.

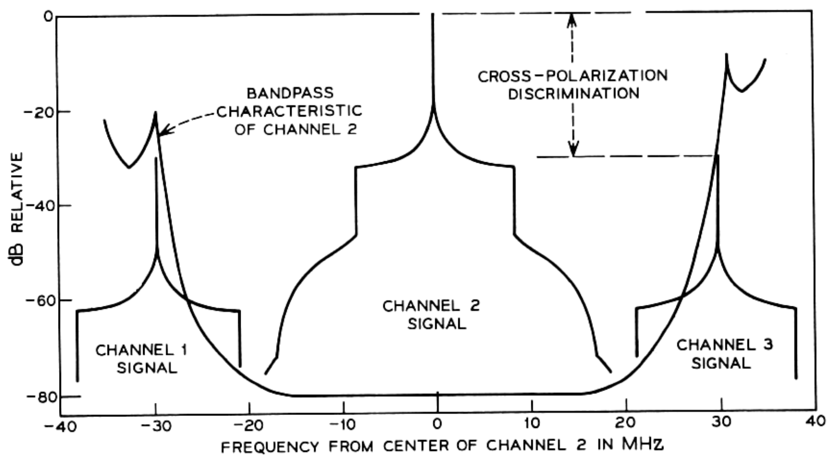

Fig. 6—Bandpass characteristic shown relative to spectra of adjacent channels.

4.5 *RF Squelch*

RF squelch is a technique for suppressing interference from a noisy channel into an adjacent channel. The principle of the squelch scheme is to monitor the IF at 86 MHz and attenuate the transmitted power by 29 dB when the noise is equivalent to a fade of 47 dB or greater. Briefly, the squelch initiator circuit detects high noise in the IF by monitoring the output of the limiter in the radio transmitter. During a deep fade or an open channel the squelch initiator changes the bias on the diode in the transmitter modulator so that the conversion efficiency is reduced by 29 dB. In most other systems a carrier resupply is used to quiet the channel by inserting a clean carrier. The resupply method has a disadvantage which the RF squelch was specifically designed to overcome; that is, when the carrier resupply operates, the channel is dead. On the other hand, when the squelch operates, it is equivalent to a 29 dB fade in the following hop. Thus, if the units operate inadvertently, the channel is completely disabled by the resupply, but it is still usable in the case of RF squelch. Also, it is argued that when a channel fades beyond about 50 dB, but cannot get protection, it is better to make the circuits unacceptably noisy for a fraction of a second than to have a break in the connection while the resupply is operated. Further, the carrier resupply quiets the message circuits, simulating the off-hook condition for all the idle circuits and consequently a massive seizure may occur in the tandem switching machine. In contrast, when the squelch is operated, the message circuits will be noisy and the off-hook condition will not be recognized by the terminal signaling equipment.

The tertiary noise interference described earlier will become much greater if the gain of a radio repeater in the center channel is increased. This would happen if the center channel suffered a selective fade, or the previous transmitter failed or was down for maintenance. To prevent excessive tertiary interference under these circumstances it is necessary that the RF squelch be operated in the first radio repeater following the fade or failure. Therefore, since the excess gain of the TH-3 repeater (which is actually due to the AGC amplifier and limiter) is approximately 50 dB, the RF squelch has to operate on less than a 50 dB fade.

If strong adjacent channel carriers are transmitted through the AGC amplifier and limiter, the repeater will not go to full gain when the desired RF signal is lost. The selectivity of the IF bandpass filter in the radio receiver was chosen to prevent the adjacent channel carriers from reducing significantly the excess gain of the repeater and thereby changing the trip point of the RF squelch by more than 0.3 dB. As already mentioned, the new IF filter was designed with minimum selectivity consistent with good intermodulation noise performance. Based on the minimum antenna cross-polarization discrimination (XPD) of 25 dB, and the receiver RF selectivity of 9 dB, the IF filter had to provide 25 dB of selectivity at ±29.65 MHz.

4.6 *Tone Problems*

The interference-to-carrier (I/C) requirements for tones at IF are shown in Fig. 7. Two requirements curves are given, one for tones where the I/C ratio does not change with fading, and one for tones where the I/C ratio increases dB for dB with the fade. The former is similar to the requirement specified for TD-3.[1] Generally there is a 19 dB difference between the two requirements curves as explained in the tone objectives in Section II. Below about 17 MHz the requirement is determined by interference into the channel under consideration, and above about 17 MHz by interference into the adjacent channels when the tone is retransmitted. The most severe requirements are imposed by tones which fall into the baseband regions occupied by the message circuit load and the 11.09 to 11.67 MHz auxiliary channel.[3] The 1800 circuits occupy the band 0.564 to 8.524 MHz, but the requirement shows this region extending to 0.312 MHz in recognition that an extra 60 circuits may be added at the low-frequency end if desired.

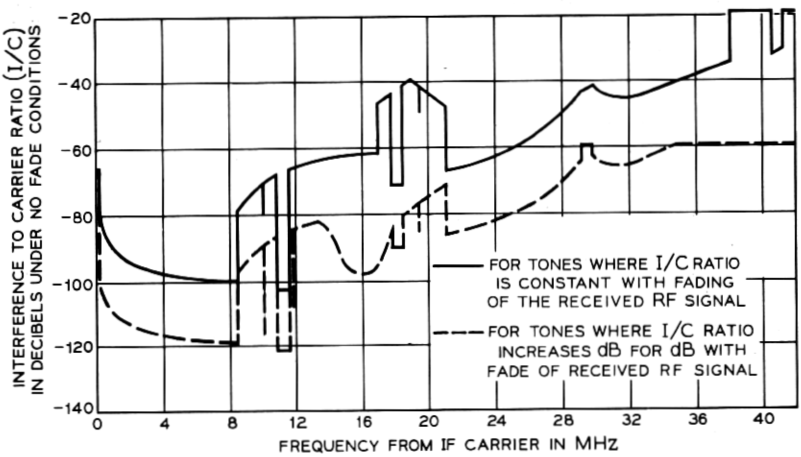

Fig. 7—Requirement for tone interference to carrier ratio at microwave receiver output.

There are special requirements at the 10.2 and 11.88 MHz to prevent malfunction of the 100A and 300A switching systems.[4] Also, the squelch initiator could be operated by interfering tones instead of noise, and at around 16 MHz the requirement for tones whose I/C ratio increases with fading is controlled by RF squelch considerations.

Some tones, such as those introduced by the dc-to-dc converter in the TWT power supply, are introduced at every repeater in the system. For these tones a multiple exposure factor was applied which decreased the permissible level of the tones. The factor was computed as 10 log of the number of tones expected to fall in a 3 kHz message circuit. However, if there were many tones the factor was computed so that the combined power of the tones plus thermal noise would not be more than 0.1 dB greater than the thermal noise alone. For the TWT power supply tones, the multiple exposure factor was calculated to be 13 dB based on either the tone or noise criteria, except in the auxiliary channel (because the auxiliary channel is limited to only ten hops) where the factor is 3 dB.

4.6.1 *Co-Channel Interference*

A carrier-to-interference (C/I) requirement of 66 dB is placed on co-channel interference consistent with what can be achieved with the front-to-back coupling loss of the horn antenna. A much more severe C/I requirement would be necessary if the beat frequency between the carriers of the co-channel signal and the desired signal exceeded the baseband frequency of the lowest message circuit. Although the multiplex normally used with TH-3 will put the lowest message circuit at 564 kHz, the system was engineered for a lowest circuit of 312 kHz so that an extra supergroup can be carried if desired. By carefully controlling the frequency of the transmitted signals the beat frequencies on long systems are held below 312 kHz. This is achieved by putting frequency stability requirements of ± 100 kHz on the FM transmitter, 5 ppm (± 30 kHz at 6 GHz) on the microwave generators, and ± 4 kHz on the shift oscillators. If the repeater station bays did not use a shift oscillator, the stability requirements for the FM transmitter and microwave generator would be much more severe.

4.6.2 *RF Selectivity and Interchannel Tone Interference*

The RF selectivity of the T-R bay was determined by the need to suppress interchannel tone interference and direct adjacent channel interference (DACI). The amplitude response of the RF networks is presented in Fig. 9 of Ref. 11. Field measurements of DACI are recorded later in this paper. Based on early measurements it was determined that about 10 dB of receiver RF selectivity at ±29.65 MHz was adequate to avoid DACI problems with the adjacent channels. The actual selectivity is approximately 9 dB.

For the purpose of calculating tone interferences certain minimum characteristics of the antenna system were assumed.

(*i*) The minimum XPD is 25 dB.
(*ii*) The minimum side-to-side coupling loss between antennas is 80 dB for the same polarization, and 83 dB for opposite polarizations.

As explained in Ref. 2, 74.13 MHz is the natural IF for TH-1 channel assignments. If a 74.13 MHz IF had been used, the interchannel tone interferences would have fallen nominally at zero and 14.83 MHz in the baseband. As a consequence of using the 70 MHz IF, similar interferences fall at 4.13, 6.57, 8.26, and 10.70 MHz. Three of these tones fall into the message portion of the baseband and therefore they must be very low in level. Nevertheless, the RF selectivity needed to suppress these tones is not considered excessive.

4.6.2.1 *The 4.13 MHz Tones.* Tones at 4.13 MHz are caused by near-end interference between transmitters and receivers of channels 16* and 20, 17 and 21, or 18 and 22. The coupling path from transmitter to receiver may be via the side-to-side coupling of the antennas or leakage between the bays. Certain tone mechanisms of this type control the transmitter RF selectivity with a requirement of 93 dB at ±70 MHz. Other mechanisms involving the transmitted signal itself put RF selectivity requirements of 45 dB at −74.1 MHz on the disturbed channel receiver. In the latter case, the RF selectivity is more than adequate. Also, there must be 160 dB loss in the leakage path between the transmitted and the received signals.

4.6.2.2 *The 6.57 MHz Tones.* Tones at 6.57 MHz in the baseband also are caused by near-end interference. The image interference from channels 14 through 18 into channels 20 through 24, respectively, falls into this category and is potentially the most serious in the system. This tone puts a requirement of 97 dB on the RF selectivity of the receiver at −133 MHz which again is achieved comfortably. However, the loss in the leakage path must be 180 dB which is by far the toughest

* In this description of tone interferences, the channel number, N, can refer to either the regular channel frequencies or the staggered frequencies, where appropriate.

leakage requirement. This requirement is met by designing the TWT to prevent excessive leakage and carefully sealing the receiver modulator.

4.6.2.3 *The 8.26 MHz Tones.* Tones at 8.26 MHz in the baseband are caused by far-end image interference from channels 11, 12, 13, 20, 21, 22, and 23 into 16, 17, 18, 25, 26, 27, and 28, respectively. This type of interference requires slightly less RF selectivity in the receiver (94 dB at 148 MHz) than the previous case and is of little concern.

4.6.2.4 *The 10.70 MHz Tones.* Tones at 10.70 MHz in the baseband are caused by both far- and near-end interference between adjacent channels (29.65 MHz apart) and next-to-adjacent channels (59.3 MHz apart). The most important interference is the far-end interference between next-to-adjacent channels. This mechanism sets a requirement of 47 dB on the RF selectivity of the receiver at ± 59.3 MHz for hops with nominal received signal power (-23 dBm). For those few hops with the highest allowed received signal level (-15 dBm), the requirement is 55 dB, which is missed typically by about 2 dB. As far as near-end interference is concerned, the worst case is due to beat oscillator leakage. The path loss requirement for the leakage is 131 dB which is small compared to other leakage requirements referred to above.

V. RESULTS OF LABORATORY AND FIELD TESTS

5.1 *Description of Laboratory and Field Facilities*

5.1.1 *Laboratory*

The first prototype bay was constructed in 1967 from preproduction models of the bay components. Three additional bays made according to the production drawings were added in early 1968. Four laboratory bays (two main station and two repeater station) were then available for system testing.

The four laboratory models were used extensively before the field testing began. Particular attention was given to laws of addition of intermodulation noise, echolike noise generated by harmonics of the IF, and tones associated with the TWT power supply.

5.1.2 *Field Trial Route*

The initial installation of TH-3 was a nine-hop route between Vega, Texas, and Dodge City, Kansas. Testing was originally scheduled for the period May through December 1969. Since the 1044A filter was not available until December 1969, some further testing was performed in 1970 to evaluate the system equipped with this filter.

The route consisted of two switching sections with an intermediate main station at Hooker, Oklahoma. The first switching section, Vega to Hooker, was the primary trial section.

The channels on the first switching section were chosen to examine the most important intrasystem interferences. At least three adjacent channels were necessary to check third-order interferences* and tertiary interference. Also channels were needed to check the special tone interferences (see Section 4.6.2). To accomplish these ends, channels 1T, 2T, and 3T were installed from Vega to Hooker and channels 3T, 7T, and 8T from Hooker to Vega. Channels 3T and 7T were installed in both directions on the second switching section to provide one regular and one protection channel. By interconnecting at IF, any number of hops up to thirty could be connected in tandem in the first switching section.

One new building and tower were constructed for the initial installation. The remaining stations were existing TD-2 stations and required only additions to the existing buildings. The antennas were re-oriented using the latest technique to reduce multimoding, and the cross-polarization discrimination was optimized at 4 GHz as is common practice for routes carrying both 4 and 6 GHz signals. The first switching section was 139.4 miles long or an average of 27.9 miles per hop. This conforms reasonably well with the system model which assumes 26.7 miles per hop. Also, since in the system model FM terminals are placed approximately every ten hops, measurements on ten hops (a loop from Vega to Hooker to Vega) can be extrapolated to the full system length on a power basis.

5.2 *Antenna System Performance*

A knowledge of the antenna coupling losses[1] that actually existed on the field trial route was necessary to properly evaluate the results of interference tests.

Swept measurements were made of the side-to-side and back-to-back coupling losses at all possible locations on the route. Generally, the measured coupling losses varied rapidly with frequency. The side-to-side coupling versus frequency response had a very fine grain, almost noiselike, structure which is attributed to energy coupling via many modes. The back-to-back coupling response also showed rapid variations although the structure was much coarser than for side-to-side coupling. The minimum side-to-side coupling losses followed an approximate normal distribution. For the same polarization, the mean was 92 dB and the standard deviation was 5.5 dB. For opposite polarization, the mean was 102 dB and the standard deviation was 8 dB.

* Third-order interferences are caused by $2A - B$ and $A + B - C$ type products between the RF signals of adjacent radio channels.

A distribution for back-to-back coupling could not be obtained because the test equipment lacked the necessary sensitivity. The minimum value measured was 115 dB.

Swept cross-polarization discrimination measurements were made on each hop of the first switching section. A signal was swept in 20 MHz increments over as much of the 6 GHz band as intersystem interference considerations would allow. It was found that the XPD of two antennas was less than 25 dB at 6 GHz. These two antennas were readjusted to meet the 25 dB system requirement. A typical measured XPD is shown in Fig. 8. The peak-to-peak variation was typically 1 to 6 dB within a 20 MHz band although larger variations were observed. The broad variation over a 250 MHz band is shown in Fig. 9. Figure 10 shows the cumulative distribution curve for lowest values of XPD measured in each 20 MHz band for every antenna. The inverse linear average needed for tertiary interference (see Appendix) is 29.4 dB.

Co-channel interference is dependent on the front-to-back coupling loss of the antennas. This ratio was spot-checked at a few stations and the lowest ratio measured was 67 dB, which is 1 dB better than the co-channel interference objective.

Swept envelope delay distortion measurements were made on each of the hops of the Vega-Dodge City route. The EDD test signal was applied to the antenna system at the 6 GHz system combining networks and received at the far-end combining network. A typical EDD measurement is shown in Fig. 11. The ripple observed in Fig. 11 can be attributed to the following two phenomena:

(*i*) dominant-mode echoes in the circular waveguide,
(*ii*) echoes due to mode conversion in a multimode medium such as the circular waveguide.

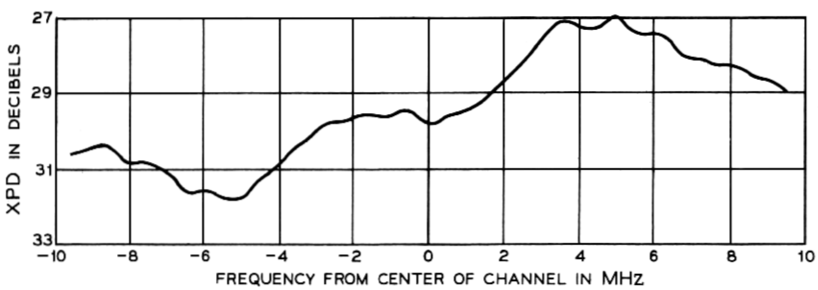

Fig. 8—Typical XPD as measured on the field trial route.

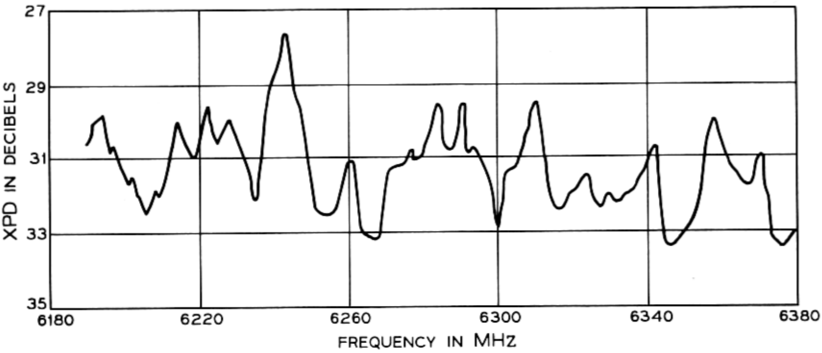

Fig. 9—Typical broadband XPD variation as measured on the field trial route.

To reduce multimoding, the first switching section was equipped with an improved circular flex (KS-20104) connecting the antenna to the rigid circular waveguide run. As a control, the second section was left with the original flex (KS-15690). The measurements indicated that the improved flex reduced the peak-to-peak ripple by a factor of two. The EDD ripples on the first section corresponded to antenna echoes

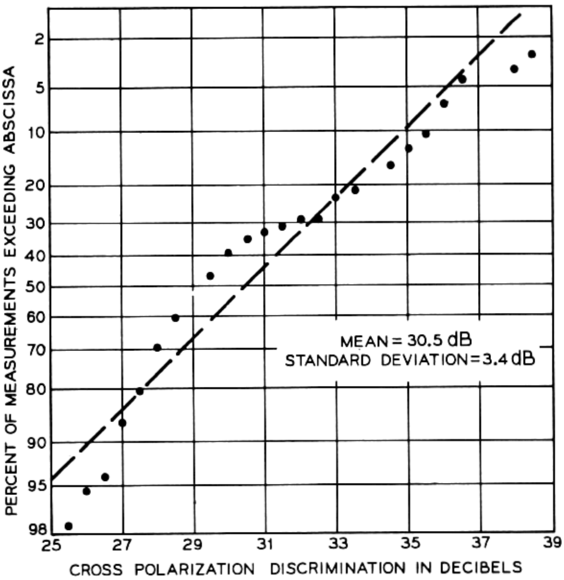

Fig. 10—Distribution of antenna system cross-polarization discrimination as measured on the field trial route.

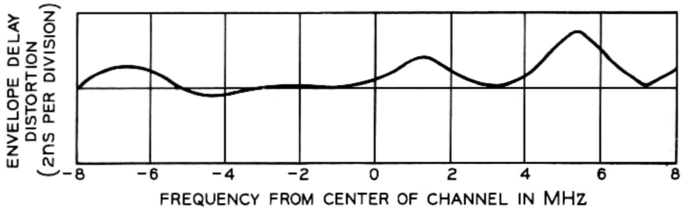

Fig. 11—Typical envelope delay distortion of antenna system equipped with KS-20104 circular flex.

which are typically 55–60 dB below the desired signal. The noise contributed by echoes of this level is computed to be less than 6 dBrnc0 per hop in the top channel, which will meet the antenna system allocation.

5.3 *Multihop Performance*

5.3.1 *Envelope Delay Distortion*

Since the 1044A filter was not available until December 1969, most of the field trial measurements were made on bays equipped with the 1009A filter. The envelope delay distortion measurements showed a systematic increase with the number of hops. After a few hops, a coarse (W-shaped) ripple became the dominant feature of the measurements. A ten-hop measurement typically had 6 ns peak-to-peak ripple over the center portion of the band. Little difference could be seen in the one- or two-hop results using either 1009A or 1044A filters. However, the ten-hop results with the 1044A filter, shown in Fig. 12,

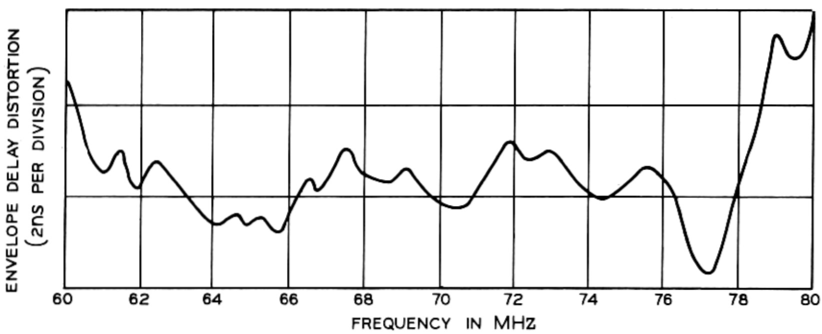

Fig. 12—Envelope delay distortion measured on a ten-hop loop equipped with 1044A IF filters. (The test set receiver had a 1.4 kHz lowpass filter before the oscilloscope and a sweep rate of 100 Hz.)

were markedly different. Although a W shape is still discernible, the peak-to-peak ripple is less than 3 ns. To achieve the EDD shown in Fig. 12, it was necessary to use a 0.25 ns/MHz sloper on every other hop. Most of the ripple is attributable to the antenna systems although the W shape is caused by the equalized RF networks.

5.3.2 *Noise Loading*

For all noise loading measurements the noise generator output was bandlimited from 0.3 to 8.204 MHz to simulate the 1800-circuit load that the system was designed to carry. This signal was then shaped by the pre-emphasis network.

The noise loading measurements showed that the intermodulation noise of the system related very closely to the measured EDD. When the new IF filter became available, a significant reduction in the system intermodulation noise was observed, as shown in Table I below, even though the law of addition was still high. A set of noise load "V-curves" obtained with the 1044A filter installed is shown in Fig. 13. These data were recorded on ten hops of the field trial route by looping channels at IF. The contribution of the FM terminals has been subtracted out so that the curves represent the noise due to the radio line portion of the system.

The noise of ten hops of the radio line is presented in Fig. 14 as a function of baseband frequency. The total noise at reference drive is shown along with the thermal and intermodulation noise components which were derived from the V-curves.

The thermal noise objective for ten hops of the radio line is 25 dBrnc0. As can be seen from Fig. 14, the requirement is exceeded by about 1 dB at 8.5 MHz. This is partially due to the baseband rolloff of the system. Since the ten hops exhibited about 0.6 dB rolloff, this would increase the thermal noise by 0.3 dB over what would be expected from a system with no rolloff. In addition, some of the first production components had noise figures slightly above normal.

The objective for cross-modulation noise caused by transmission

TABLE I—INTERMODULATION NOISE (AT 8.0 MHz) OF THE SYSTEM EQUIPPED WITH 1009A OR 1044A IF FILTERS (dBrnc0)

No. of Hops	With 1009A Filter	With 1044A Filter
2	16.0	9.0
6	23.5	15.4
10	26.0	19.2

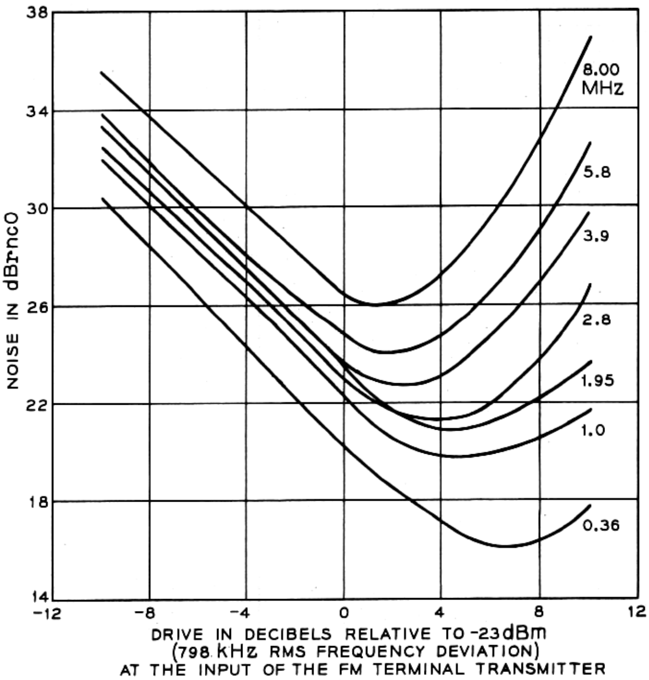

Fig. 13—Noise load V-curves measured on a ten-hop loop equipped with 1044A IF filters.

deviations and antenna system noise is 22.4 dBrnc0 for the radio line portion of a ten-hop system. As can be seen from Fig. 14, the measured cross-modulation noise meets the objective with 3 dB margin which is sufficient to allow the system to meet the total noise objective of 27.2 dBrnc0, even though the thermal noise objective was exceeded by about 1 dB.

Tests were also made comparing the chosen pre-emphasis and the pre-emphasis recommended by the CCIR for 1800-circuit loading.* The results of these tests indicate that the CCIR network gives about 1 dB less noise at low and high baseband frequencies, but about 1 dB more noise over the center of the baseband. On the whole, the performance of the two networks is about the same.

* CCIR REC 275-2, Documents of the XIIth Plenary Assembly, New Delhi, 1970, Vol. IV, p. 139.

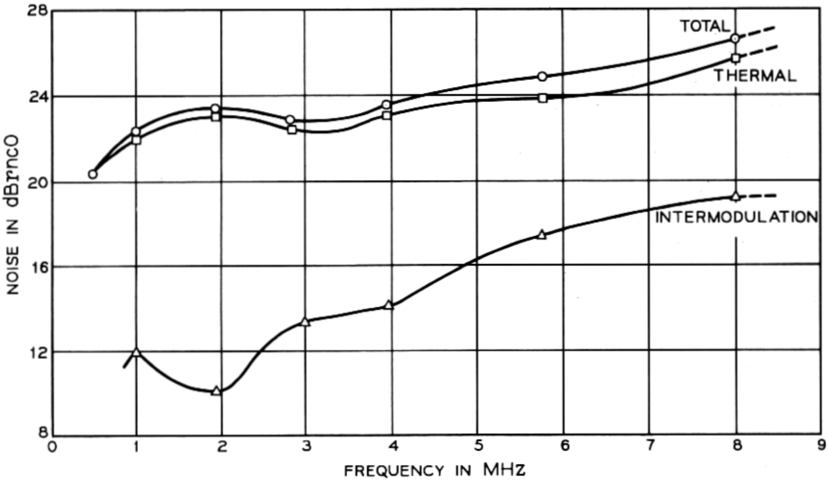

Fig. 14—Thermal and intermodulation noise versus baseband frequency as derived from noise load measurements on a ten-hop loop.

5.3.3 *Baseband Response*

The radio line exhibits a characteristic 0.06 dB/hop rolloff at 8.5 MHz and 0.12 dB/hop rolloff at 11.88 MHz. The rolloff is so consistent that the difference between any two radio channels will be less than 0.25 dB, so the wideband signals can be switched from regular to protection channels without significant level changes.

5.4 *Intrasystem Interference*

5.4.1 *Tertiary Interference*

The tertiary interference (see Section 4.4 and the Appendix) was measured by fading the center channel while one of the adjacent channels was carrying a noise load signal. Typical results of a measurement are shown in Fig. 15. The noise increases linearly with the fade depth* up to about 40 dB. Above this value, the total noise power approaches the carrier power and the amount of transfer in the limiter is reduced. This leads to the increase of the double adjacent noise at the same time that the transfer tertiary is beginning to roll off. At about a 47 dB fade, the squelch operates and the gain of the center channel repeater is reduced by 29 dB, thus reducing the tertiary noise by the same amount.

* Assuming a selective fade which affects only the center channel.

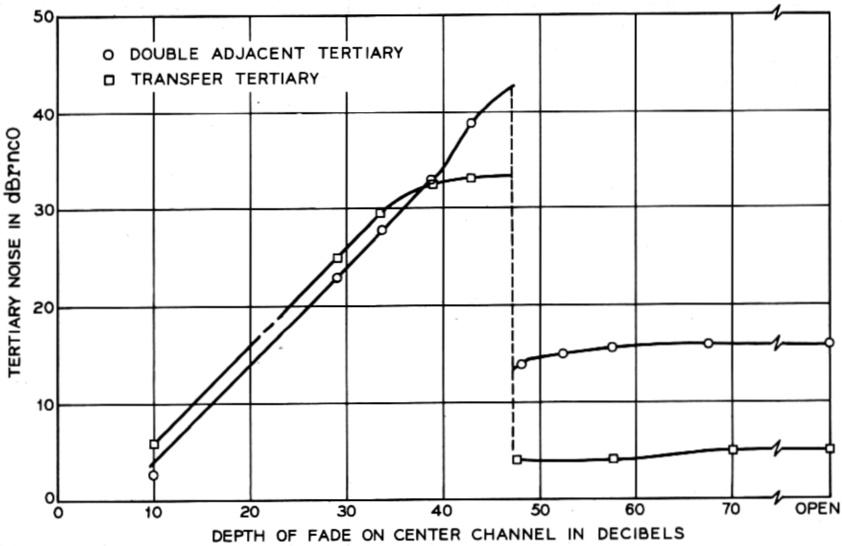

Fig. 15—Typical tertiary interference versus fade on center channel as measured on the field trial route.

Figure 16 shows a plot of the first- and third-order tertiary noise versus baseband frequency. The shape of the first-order noise is due to the selectivity of the center channel repeater (see Fig. 6). The shape of the third-order noise is sharper because it has the additional selectivity of two more repeaters. It can be seen that the difference between the two orders is not very large at the top baseband frequency as this is where the adjacent channel has the least selectivity.

Figure 17 shows the measured first- and third-order tertiary interference at 8 MHz after the noise was adjusted for XPDs of 30 dB.* Second-order noise is not shown because the XPDs involved in the path were very high. It was observed that with very high XPDs large variations in the value of the XPD were common. The predicted noise is also shown. The slope of the curves beyond the second-order path is due to selectivity of the center channel repeater at 21.7 MHz which is 5 dB per hop. When the center channel is open, the first repeater after the open channel is at full excess gain, and the noise power at the output is 38 dB below normal signal power (29 dB of loss is due to the squelch). The next repeater provides the additional 9 dB of gain to provide a squelched output, 29 dB below normal. As a result, the

* The XPDs were adjusted to 30 dB, the approximate mean of the measured XPD data.

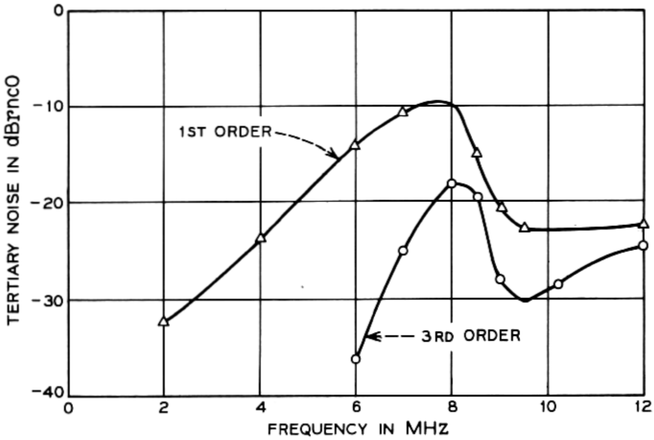

Fig. 16—Tertiary noise at a 0 dB fade versus baseband frequency as measured on the field trial route (XPD adjusted to 30 dB).

second-order tertiary noise is greater than the first-order noise. The measured values are in reasonable agreement with the values predicted from the model. The variation in the measured values is due to variations in the XPD with time and measurement error.

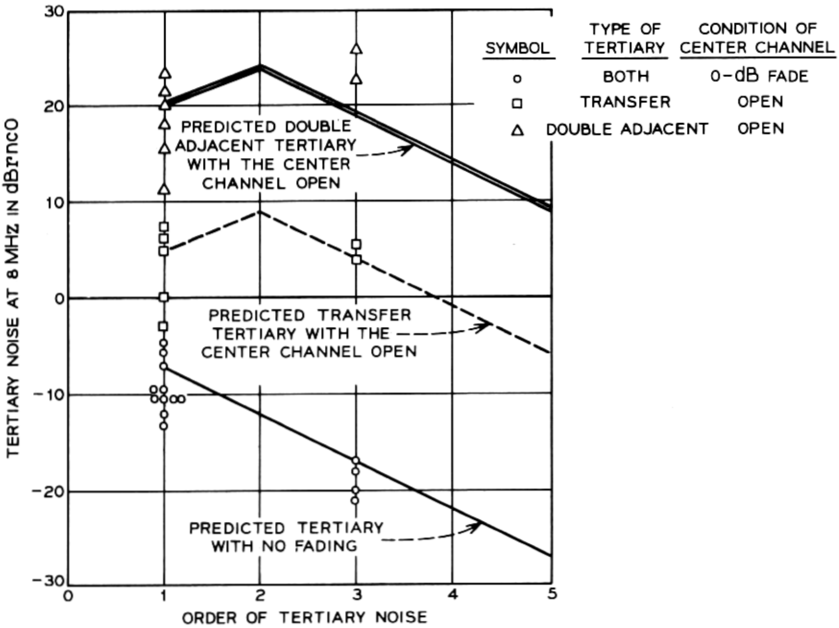

Fig. 17—Comparison of measured and predicted tertiary interference.

5.4.2 *Tone Interference*

Leakage measurements were a major part of the field tests. The results of tone measurements indicate that the bay meets the leakage requirements when correctly assembled; i.e., all covers in place and screws properly tightened.

Early measurements on the trial route showed that tones generated in the TWT power supply converter were out of limits both at the low end and the top end of the baseband. These tones fell at harmonics of the 24 kHz converter frequency. They were particularly strong in the auxiliary channel region. More filtering was added to the power supply and the tones were brought down to the objectives, except in the auxiliary channel region where they were about 7 dB above the objectives (which includes a 3 dB multiple exposure factor). A new power supply is being introduced into manufacture; one of its features is improved tone performance.

5.4.3 *Adjacent Channel Interference*

Adjacent channel interference was measured as a function of the depth of fade on the disturbed channel. The results of these measurements are shown in Fig. 18. The data has been adjusted to correspond to an XPD of 30 dB. If the faded hop had a lower XPD, the noise for deep fades would increase about 2 dB for every dB decrease in the XPD. Thus a combination of a 40 dB differential fade on a hop with minimum XPD would result in some circuits exceeding 55 dBrnc0. The increase in noise at 10.2 MHz would not be sufficient to cause the 100A system to switch the channel before a 40 dB fade. However, the likelihood of a 40 dB fade with no fade of the adjacent channels is extremely low.

The primary purpose of the RF squelch was to prevent the noise from an open channel from "spilling" into the adjacent channels centered 29.65 MHz away. When a channel was opened, causing the squelch to operate in every transmitter, it was found that the noise spillover into the adjacent channels increased their noise by less than 1 dB. Thus, even though a channel may be open, the effect on the adjacent channels is small.

VI. ACKNOWLEDGMENTS

The work of many individuals is reported in this article. In particular the authors acknowledge the accomplishments of T. G. Cross, W. D. Hutcheson, J. S. Torres, and D. S. Williams.

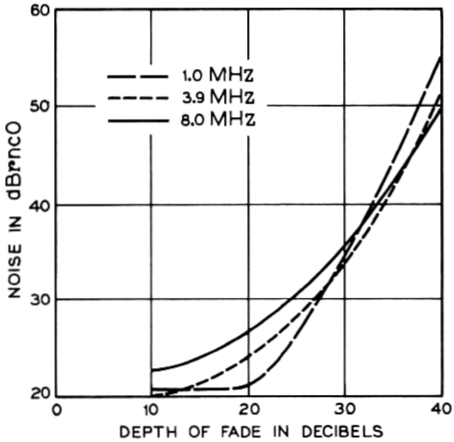

Fig. 18—Adjacent channel interference measured as a function of fade on disturbed carrier (XPD adjusted to 30 dB).

APPENDIX

Reference 2 described two types of tertiary mechanism which will be called first-order transfer tertiary and first-order double adjacent tertiary interference, respectively. These interferences are defined as first-order paths in Fig. 19, along with second- and third-order paths, etc. The figure shows the interference being introduced into channel 2 from channel 1 on the first hop. In practice, additional sets of paths will exist for interferences introduced into channel 2 from channels 1 and 3 on all hops but the last. It is clear from Fig. 19 that the interference will be transmitted by channel 2 and transfer back into channels 1 and 3 at every following hop. If channel 2 has a selectivity of a few dB to the outermost first-order sidebands of channel 1, the interference due to the second- and higher-order paths will rapidly decrease. However, if there is no selectivity in the center channel, there will be a vast increase in tertiary interference because the nth-order interferences will be no less than the first order.

It follows from the shape of the amplitude characteristic of the center channel (see Fig. 6) that the tertiary interference will be greatest in the top message circuit. The interference into the top message circuit due to transfer tertiary and double adjacent interference is expressed below in terms of the signal-to-noise ratios at baseband. In deriving these equations it was assumed that the signal-to-noise ratio in the first-order sideband regions (after a limiter) is the same as the signal-to-noise at baseband.

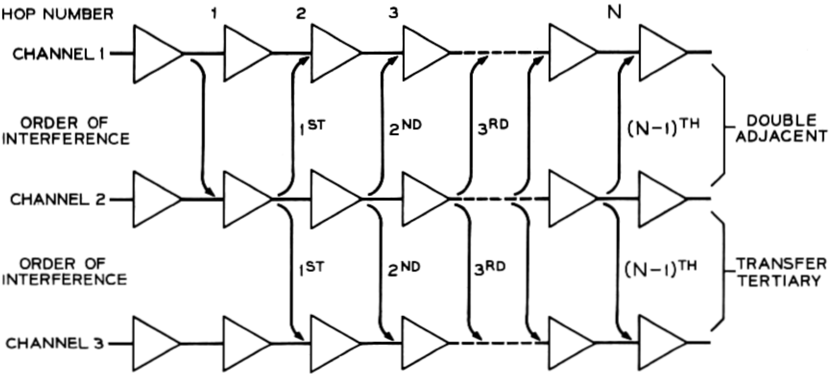

Fig. 19—Kth-order tertiary interference paths.

$$(S/N)_{\text{total}} = -10 \log \left\{ \sum_{k=1}^{N-1} (N - k) 10^{-(S/N)_k/10} \right\} - M \text{ dB} \tag{1}$$

where

N = number of hops in the system,
M = 6 dB due to the fact that the disturbed channel can be exposed to a transfer tertiary and a double adjacent interference on both sides amounting to four interference paths in all.

The term $(S/N)_k$ is the signal-to-noise ratio due to a kth transfer tertiary (or double adjacent) path, given by:

$$(S/N)_k = 2(\overline{XPD}) + kL + T + C + B \text{ dB} \tag{2}$$

where

$\overline{XPD}$ = cross polarization between channels in dB,
L = selectivity of the repeater at the frequency of the top message circuit in the adjacent channel in dB,
T = 6 dB due to limiter transfer in the channel adjacent to the disturbed channel,
C = 6 dB due to the advantage of the correlated sidebands in the disturbed channel over the interference,
B = 10 log 4/3 = 1.2 dB which is a factor for spreading interfering 3 kHz talkers over 4 kHz (1 kHz guard band and 3 kHz for talker) in the disturbed channel.

Power addition is assumed in this equation because there is several kHz change in the frequency difference between adjacent channels on successive hops. The equation is written as though the XPD were a constant value for all hops. In practice of course this is not true, but it can be shown that for a large number of hops $\overline{XPD}$ is the inverse linear average given by:

$$\overline{XPD} = -10 \log \left\{ \frac{\sum_{n=1}^{N} 10^{-X/10}}{N} \right\} \tag{3}$$

where X is the XPD of the individual hops in dB.

In equation (3), X (dB) is a random variable and if it is assumed that the XPD distribution measured on the field trial route is representative (see Fig. 10), then

$$\overline{XPD} = 29.4 \text{ dB}.$$

The RF selectivity of the repeater at 21.13 MHz (which is the frequency of the top message circuit in the adjacent channel) is approximately 5.0 dB; i.e., L in equation (2) is 5.0 dB. Substituting these values into equations (1) and (2) results in a signal-to-noise ratio of 47.6 dB for a 150-hop system which is equivalent to 24.4 dBrnc0.

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