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## TD-3 Microwave Radio Relay System

By S. D. HATHAWAY, W. G. HENSEL, D. R. JORDAN,  
and R. C. PRIME

(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,<sup>1</sup> 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,<sup>2</sup> introduced in 1961, provided nearly twice the capacity in a band

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\* 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.<sup>3</sup>

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.<sup>4</sup>

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.<sup>5-8</sup>

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<sup>9, 10</sup> 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.<sup>11, 12</sup> 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 dBm0 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 dBm0 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 dBm0 (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 dBm0 ( $-33$  dBm0 of 3 kHz flat noise) in the worst message circuit.

(v) Baseband amplitude response of  $\pm 0.25$  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 dBnc0 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

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\* 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<sup>7</sup> 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<sup>13</sup> 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 dBmC0 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

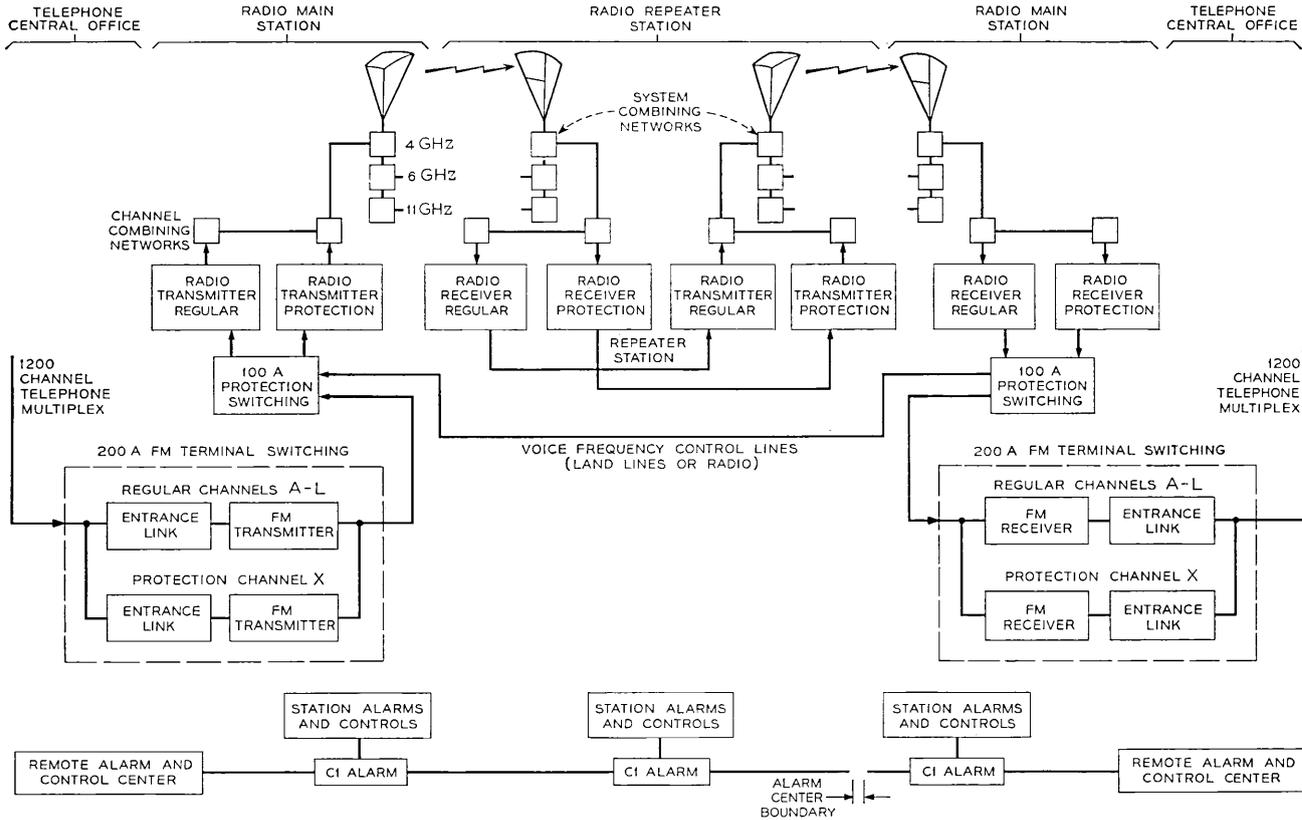


Fig. 1 — Block diagram of TD-3 system.

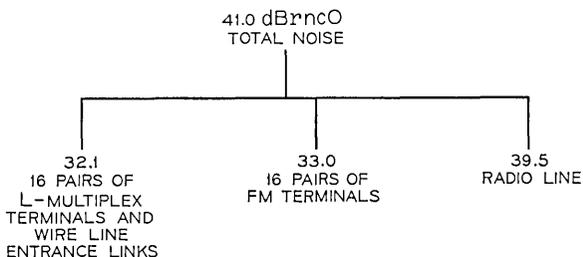


Fig. 2 — Allocation of total noise objective.

will pass through 16 pairs of FM terminals, wire line entrance links, and L-multiplex terminals. An allocation of 33 dBrc0 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.<sup>4</sup>

Also based on experience, 32.1 dBrc0 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<sup>14</sup> 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 dBm. 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.<sup>15, 16</sup> 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<sup>5</sup> when assessing actual repeater performance.

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

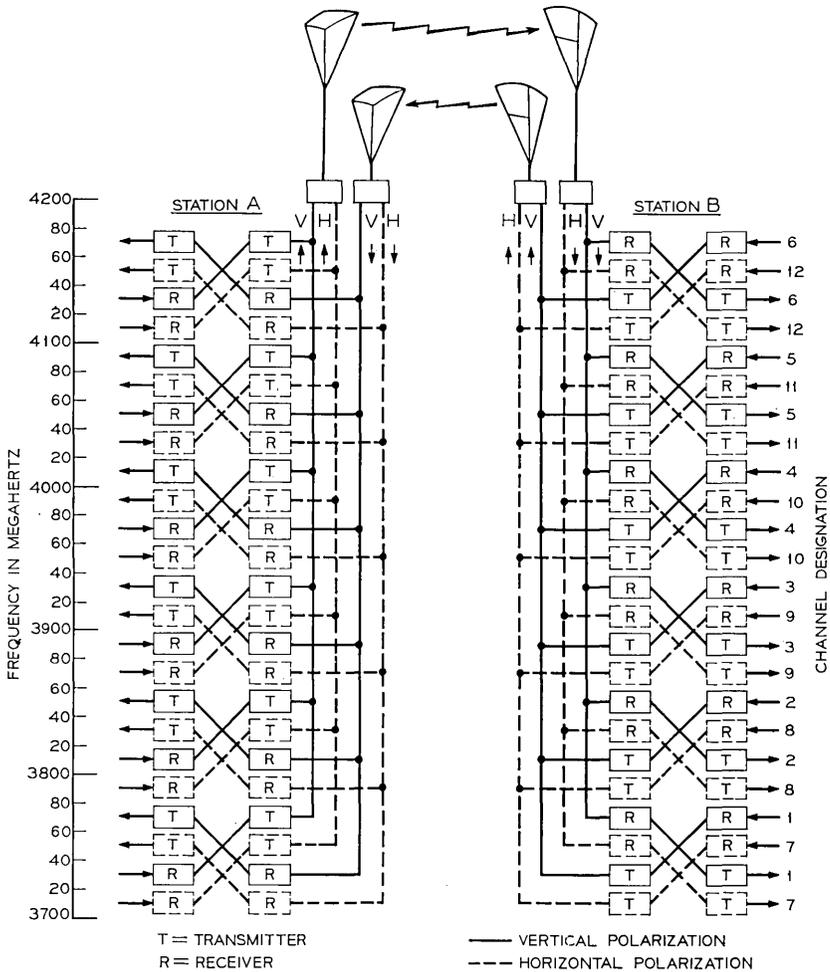


Fig. 3 — TD frequency allocation plan.

crease to 55 dBrcn0 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,<sup>18</sup> 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  $\pm 5.772$  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.<sup>15</sup> 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 dBmc 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<sup>5</sup> 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

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\* 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.

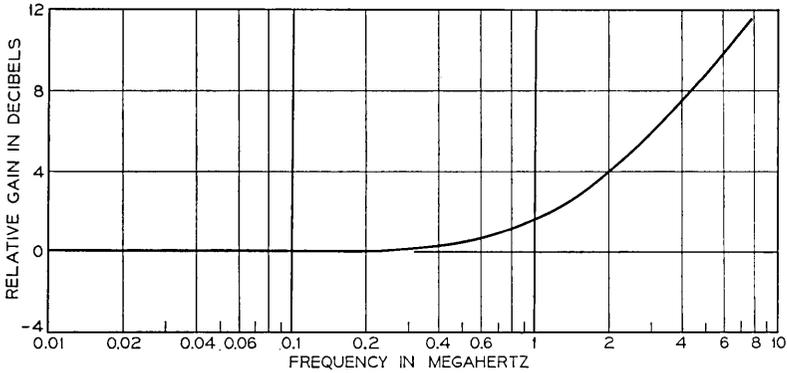


Fig. 4 — Pre-emphasis characteristic—message service.

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

Of the 41 dBrc0 system noise objective, 39.5 dBrc0 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 dBrc0. 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-

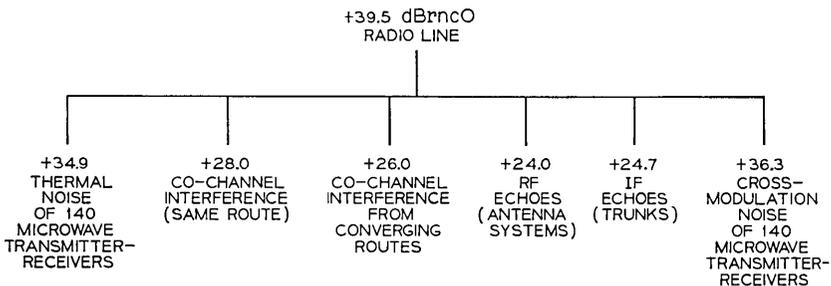


Fig. 5 — Allocation of noise objective for radio line.

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<sup>18</sup> 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.<sup>18</sup>

The selectivity was determined primarily by considering intra-

system 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

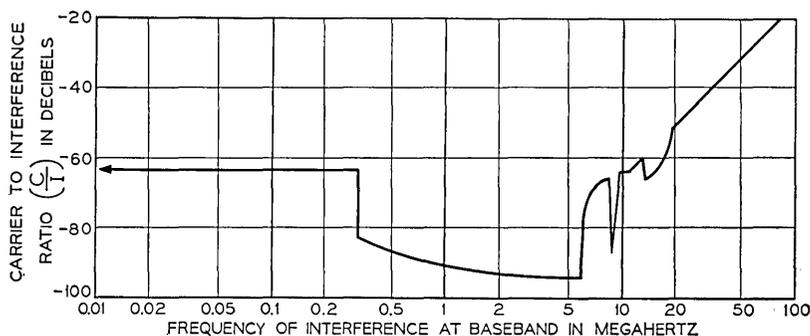


Fig. 6—Requirement for carrier to single tone interference at microwave receiver output.

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<sup>8</sup> 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

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\*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.<sup>1</sup> 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 fre-

quency. 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  $\pm 10$  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<sup>1</sup> 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

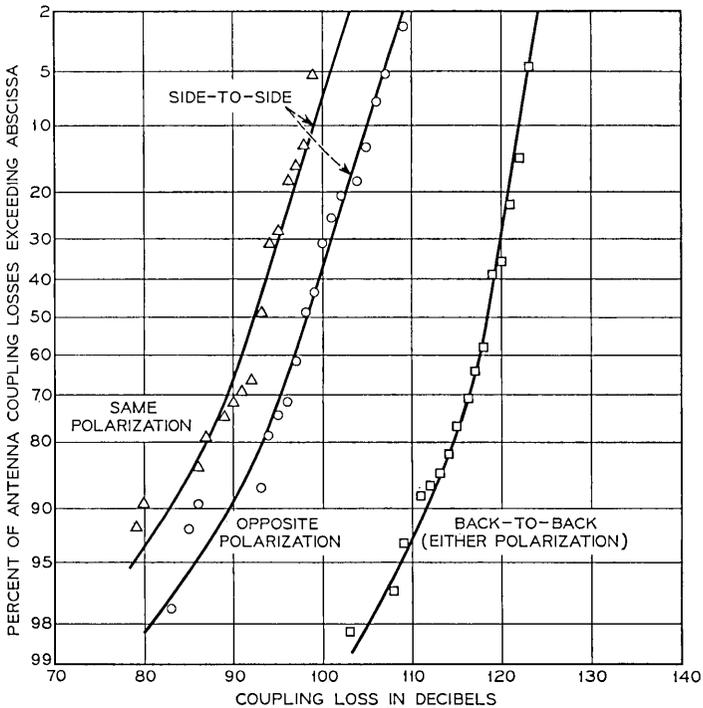


Fig. 7—Distribution of antenna system coupling losses as measured on the initial installation.

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

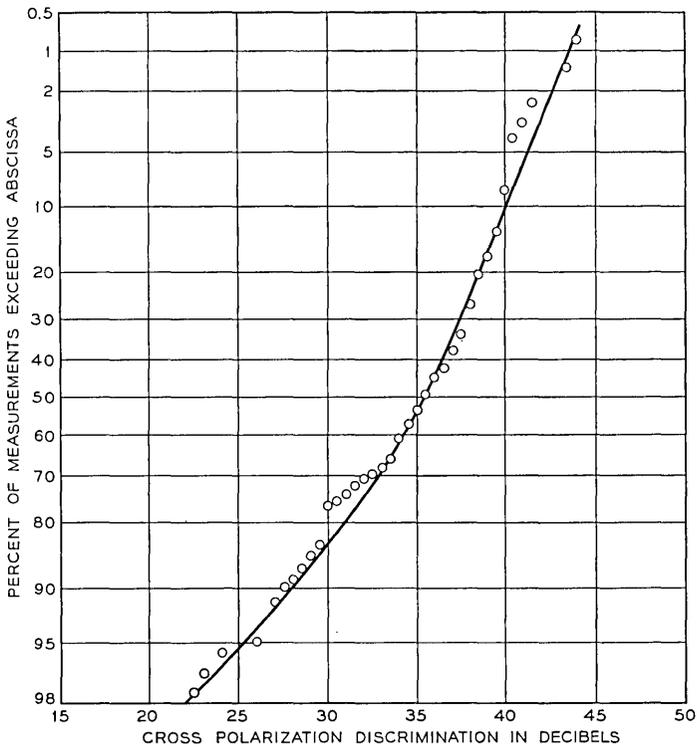


Fig. 8—Distribution of antenna system cross polarization discrimination as measured on the initial installation.

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.<sup>13</sup> 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 ( $\pm 4$  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.<sup>12</sup>

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 dB 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

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\* 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 dBrc0 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 parametric amplifier from the system left the uncancelled TWT amplifier contribution, which was about 13 dBrc0 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

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\*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  $\pm 4$  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.

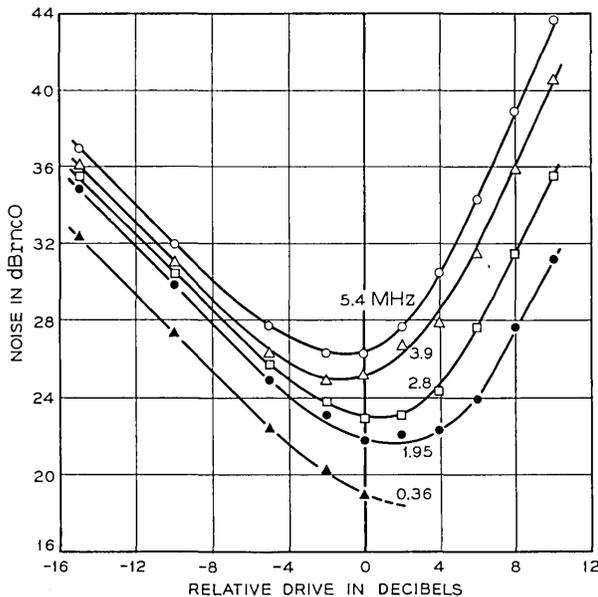


Fig. 9—Noise load V-curves measured on a 10-hop loop of channel 9 of the initial installation. The channel was equipped with Schottky barrier diode receiver modulators.

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 dBm/0. 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 dBmC0 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

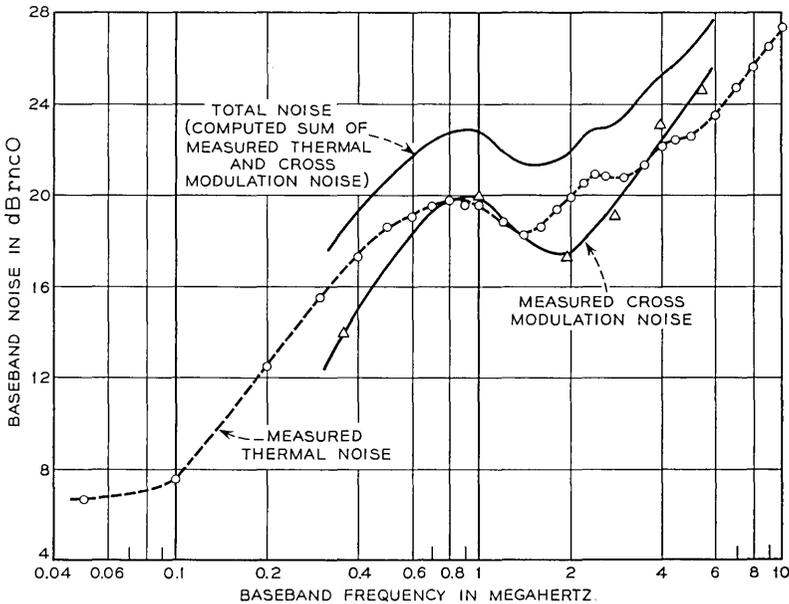


Fig. 10—Thermal and cross modulation noise versus baseband frequency measured on a 10-hop loop of channel 9 of the initial installation. The channel was equipped with Schottky barrier diode receiver modulators.

the laboratory, indicated that the noise from this source should be approximately 17 dBrc0 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 dBrc0 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-

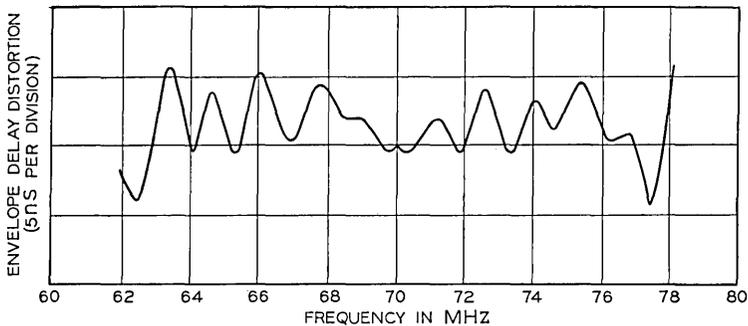


Fig. 11 — Envelope delay distortion measured on a 10-hop loop of channel 9 of the initial installation.

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 (dBmco)		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

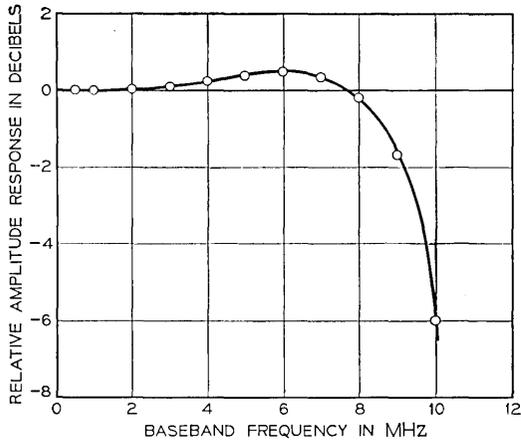


Fig. 12— Baseband amplitude response measured on a 10-hop loop of channel 9 of the initial installation.

mitter at approximately 14 dB below that required for normal ( $\pm 4$  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  $\pm 0.25$  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 heterodyne 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.<sup>18</sup>

*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<sup>2</sup> and TD-2 systems.<sup>1</sup> 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-

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\* 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  $-50$  dBm<sub>0</sub> or 40 dBrc<sub>0</sub>). Since for a 40 dB fade the noise of the faded channel is approaching 55 dBrc<sub>0</sub>, 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 dBm<sub>0</sub> tones transmitted on the adjacent channel.

A noise load, simulating a 1200 channel message and transmitted at normal ( $\pm 4$  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 noise 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 dBrc<sub>0</sub>. 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-

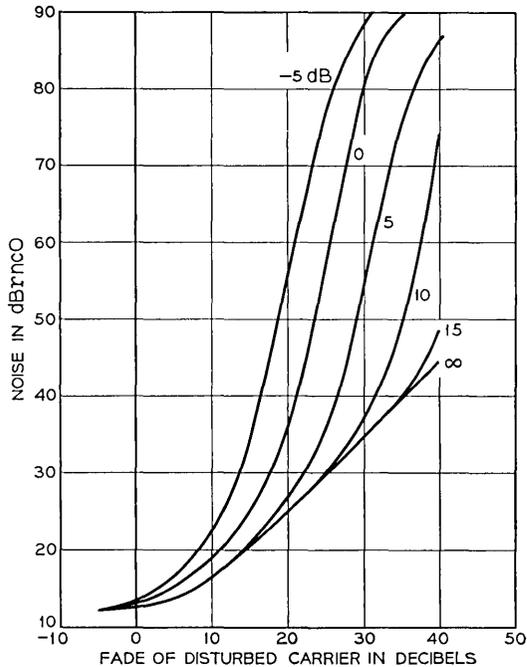


Fig. 13 — Interference, measured at 0.5 MHz in the baseband of the disturbed channel, resulting from a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier power.

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 dBm. 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

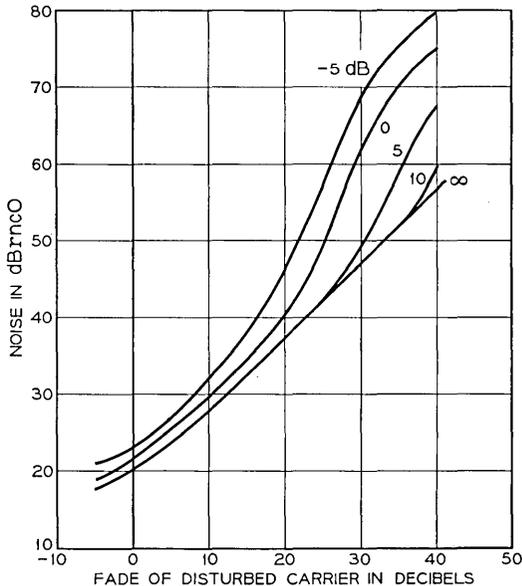


Fig. 14 — Interference, measured at 5.6 MHz in the baseband of the disturbed channel, caused by a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier power.

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.

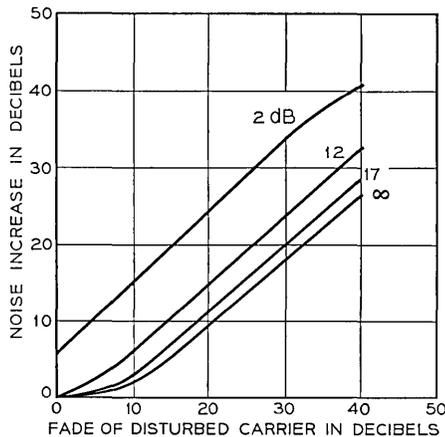


Fig. 15—Interference, measured at 9 MHz in the baseband of the disturbed channel, resulting from the second order sidebands of a 1200 channel noise load signal on the adjacent channel. The interference is shown for various ratios of the nonfaded disturbed carrier power to the adjacent channel carrier 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.<sup>19</sup> 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,  $1^\circ$  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  $3^\circ$  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.

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\* 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,

$$S(t) = A_c \cos [\omega_c t + \phi(t)] \quad (1)$$

where

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

and an RF interfering tone,

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

where

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

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\* 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<sup>19</sup>

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

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

$$\psi(t) = \phi(t) + \psi_1(t) \quad (4)$$

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

$$\psi_1(t) = K \sin [\omega_n t + \theta - \phi(t)] \quad (5)$$

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

$$\phi(t) = X \sin (qt + \alpha) \quad (6)$$

where

- $X$  = modulation index of test tone in radians
- $q$  = baseband frequency of test tone in radians/sec.
- $\alpha$  = phase constant.

Equation (5) becomes

$$\psi_1(t) = K \sin [\omega_n t + \theta - X \sin (qt + \alpha)]. \quad (7)$$

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

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

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

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

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

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

Introducing the approximations

$$\begin{aligned} J_0(X) &\cong 1 \\ J_1(X) &\cong \frac{X}{2} \\ J_{-1}(X) &\cong -\frac{X}{2} \end{aligned}$$

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

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

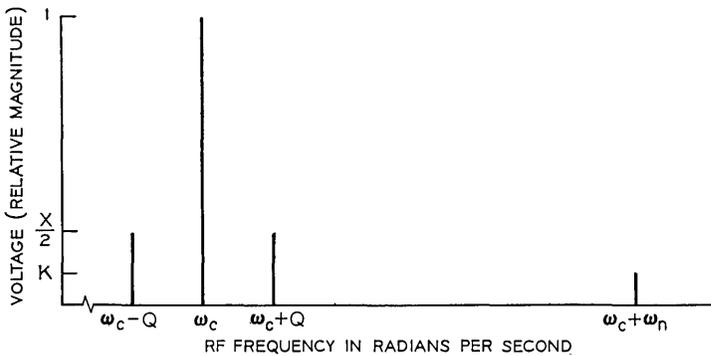


Fig. 16 — Spectrum at RF.

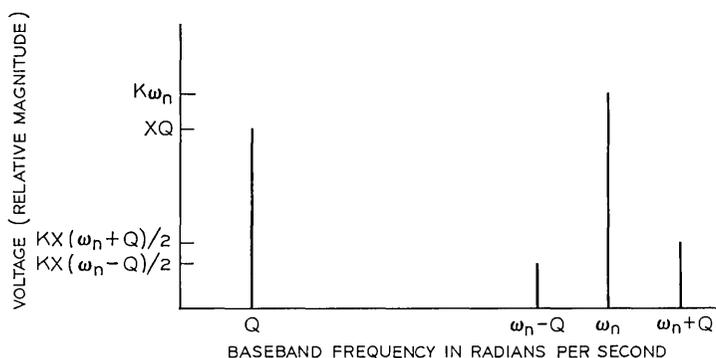


Fig. 17 — Spectrum at baseband.

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

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# Microwave Transmitter and Receiver

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(Manuscript received December 13, 1967)

*The basic blocks of the TD-3 radio system are the microwave transmitter and receiver units. Since up to 140 of these units may be connected in tandem, each must meet rather stringent requirements in order to satisfy the over-all system objectives described in the previous paper. The transmitter and receiver units use solid state semiconductor devices (except for the traveling wave electron tube amplifier in the transmitter), and operate from a single -24 volt battery source. The detailed design of these units is presented here with particular attention to thermal noise, cross modulation noise, selectivity, and equalization.*

## I. INTRODUCTION

The microwave transmitter and receiver units constitute the basic building blocks for the TD-3 radio system. They provide the gain necessary to compensate for the transmission loss of the station-to-station air path, and means for changing that gain during periods of radio signal fading. These units also include the selectivity required to keep the radio channels of a route properly separated.

As previously pointed out,<sup>1</sup> the Schottky barrier diode modulator is used in microwave receivers now being manufactured; therefore, most of our derivations pertain to this type of receiver. However, our analysis of receiver selectivity requirements pertains to a receiver equipped with a parametric amplifier, whose use was expected when the selectivity requirements were originally established.

## II. DESIGN FEATURES AND PERFORMANCE OBJECTIVES

The distinction between a design feature and a performance objective is not always clear because they may be related; therefore, somewhat arbitrary distinctions are made.

## 2.1 *Design Features*

The electronic circuits in the microwave transmitter and receiver, with one exception, use solid-state semiconductor devices. The exception is the traveling wave electron tube amplifier in the transmitter used to deliver the high power required at microwave frequencies. Modern, solid-state low voltage battery plants were in use in the telephone plant at the beginning of the TD-3 development. Since these low voltage sources were compatible with transistors, the microwave transmitter and receiver have been designed to operate from a single  $-24$  volt battery plant. A regulator<sup>2</sup> closely controls the voltages applied to solid-state devices and thereby eliminates degradation of performance with changes in battery voltage. An inverter-regulator,<sup>2</sup> powered from the  $-24$  volt source, provides the several voltages required by the traveling wave tube.

At intermediate frequencies, compatibility with the TD-2 system and with the 100A protection switching system<sup>3</sup> is achieved by using a 70 MHz frequency, by providing 75 ohm impedance circuits, and by using signal powers consistent with the switching system.

## 2.2 *Performance Objectives*

### 2.2.1 *Thermal Noise*

The first paper in this issue established the system thermal noise objective and allocated 34.9 dBrnC0 to the radio line.<sup>1</sup> The system model for the radio line assumed 140 repeaters; therefore, the per-repeater (or transmitter-receiver pair) thermal noise objective is:  $34.9 - 10 \log_{10} 140 = 13.4$  dBrnC0.

As the input signal to a microwave receiver decreases because of fading, the thermal noise contribution from that receiver increases and eventually controls total system noise. Noise of 55 dBrnC0 in a message channel has been established as the maximum permissible in a commercially acceptable circuit. Based on the best information available, a fade range of 40 dB for each receiver should insure that yearly outage time because of fading will provide acceptable circuit reliability if a 2 for 10\* protection switching arrangement is used. For this fade range, the allowable nonfaded thermal noise requirement for the receiver becomes  $55 - 40 = 15$  dBrnC0, which is slightly more lenient than the 13.4 dBrnC0 above.

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\* Two protection channels for ten working channels.

### 2.2.2 Cross Modulation Noise

The first paper in this issue allocated 36.3 dBrnC0 of cross modulation noise to the radio line.<sup>1</sup> Early in the development, this allocation was divided into two equal parts, one assigned to equalization misalignment resulting from changes in ambient temperature or humidity (discussed in Section 3.4.3), and one assigned to noise caused by residual amplitude and delay distortion in the signal path after equalization. Assuming that the noise adds on a power basis, the per-repeater allocation for the latter part is:  $33.3 - 10 \log_{10} 140 = 11.8$  dBrnC0.

With 11.8 dBrnC0 allotted to a transmitter-receiver pair, the related magnitudes of amplitude and delay distortions were needed. To simplify finding these magnitudes, we assumed that:

(i) The transmission characteristics would need to be well controlled only over the frequency region  $f_c \pm 6$  MHz, that is, the region occupied by the carrier and the first order sidebands of the signal.

(ii) The noise resulting from residual amplitude distortion would be negligible.

(iii) The residual delay distortion would include only slope and ripple components.

Using methods described in Refs. 4 and 5, the delay slope or ripple amplitudes which would produce 11.8 dBrnC0 of cross modulation noise were determined to be:

Delay slope, 0.15 nanoseconds per MHz.

Delay ripple, 0.2 nanoseconds peak amplitude for ripple periods of 10 MHz and greater; 0.6 to 0.2 nanoseconds peak amplitude for ripple periods from 2 to 10 MHz.

It is obvious that, if both delay slope and delay ripple distortions are present, power addition of the noise must not exceed 11.8 dBrnC0.

In addition, parabolic amplitude distortion must be controlled to achieve satisfactory system baseband amplitude response. An objective of  $\pm 0.05$  dB at  $f_c \pm 6$  MHz, relative to the response at  $f_c$ , was specified.

Late in the development period, methods became available which permitted predictions to be made of the cross modulation noise produced by simple types of amplitude and delay distortions for the type of pre-emphasis used.<sup>6</sup> It was determined that either the original delay slope or the ripple allowances would consume the 11.8 dBrnC0 repeater allocation. Also, the noise from these two sources generally

TABLE I—MAGNITUDE OF DISTORTION\*

Distortion type	Magnitude of distortion at $\pm 6$ MHz
	(nanoseconds)
Linear delay	0.9 (slope of 0.15 nanoseconds/MHz)
Parabolic delay	4.6
Cubic delay	1.1
	(decibels)
Parabolic gain	5.0
Cubic gain	0.12
Quartic gain	0.3

\* Distortion causing 11.8 dBmC0 of cross-modulation noise in the top message channel.

could be expected to add on a power basis. The magnitudes of specific types of distortion, each of which would account for 11.8 dBmC0 of cross modulation noise, are given in Table I.

Notice that no allocation was made to cross modulation noise caused by AM to PM converting devices. For that reason, noise from this source had to be kept small relative to that from the other sources if system objectives were to be met.

### 2.2.3 Interferences

Spurious tones can be generated in a voice channel by interferences from inside and outside the microwave transmitter and receiver. To reduce these tones to acceptable levels, the interferences must be controlled by selective filters, by directional devices, and by reduction of leaks between circuits.

The objective for a single frequency tone interference in a voice channel is a maximum of  $-70$  dBm0 under "no fade" conditions.<sup>1</sup> For interferences dependent on the received carrier power, the objective is relaxed to  $-42$  dBm0 at the fade depth corresponding to a radio channel noise of 55 dBmC0, the protection switch request point. In addition to the  $-70$  dBm0 objective, spurious tones at intermediate frequencies must be kept sufficiently low so as not to materially affect the automatic gain control range of the microwave receiver, or the switch-on point of the carrier-resupply circuit used to replace the carrier in a failed or very deeply faded radio channel.

With the TD-3 frequency plan, two types of adjacent channel interference occur: spillover of higher order sidebands from the disturbing channel, and direct transfer of modulation from the disturbing to the disturbed channel. Direct transfer is usually called direct

adjacent channel interference. The magnitudes of these interferences are determined primarily by the antenna side-to-side coupling loss, the degree of cross polarization discrimination achieved, and by the repeater selectivity. The design objective was to keep noise contributions from these sources negligible compared with the total noise objective.

The microwave transmitter and receiver must meet requirements for spurious radiation imposed by FCC regulations, and spurious radiation must not interfere with systems operating in the other common-carrier bands at 6 and 11 GHz.

#### 2.2.4 Baseband Amplitude Response

The baseband signals applied to the FM terminal transmitter at the input to the radio line extend approximately from 0.5 to 6 MHz. The first paper specifies an objective of  $\pm 0.25$  dB for baseband flatness for each radio channel in an IF switching section.<sup>1</sup> Experience has indicated that this objective can be met if the variation in amplitude response of the microwave transmitter-receiver over the range of  $f_c \pm 6$  MHz is held to  $\pm 0.1$  dB.

### III. CIRCUIT DESIGN

#### 3.1 Circuit Description

##### 3.1.1 Microwave Transmitter

Figure 1 shows the microwave transmitter with typical signal powers. It accepts a  $-7$  dBm signal at 70 MHz IF from the microwave receiver at a repeater station or from the FM terminal transmitter at a main station. It produces an output of  $+37$  dBm at one of the 24 channels in the 3710–4170 MHz frequency range.

The IF limiter<sup>7</sup> removes unwanted amplitude modulation from the input IF signal and furnishes an IF control signal to the IF carrier resupply circuit. The IF carrier resupply<sup>7</sup> furnishes a carrier should the normal IF signal be lost.

If the IF signal at the limiter input is lost, the carrier resupply furnishes a tone-modulated FM signal to the transmitter modulator. In addition, an operated carrier resupply provides a dc output to bias the limiter to a high insertion loss state and thus attenuates noise at the limiter input during the lost carrier period. The reinserted carrier prevents AGC circuits in subsequent repeaters from driving

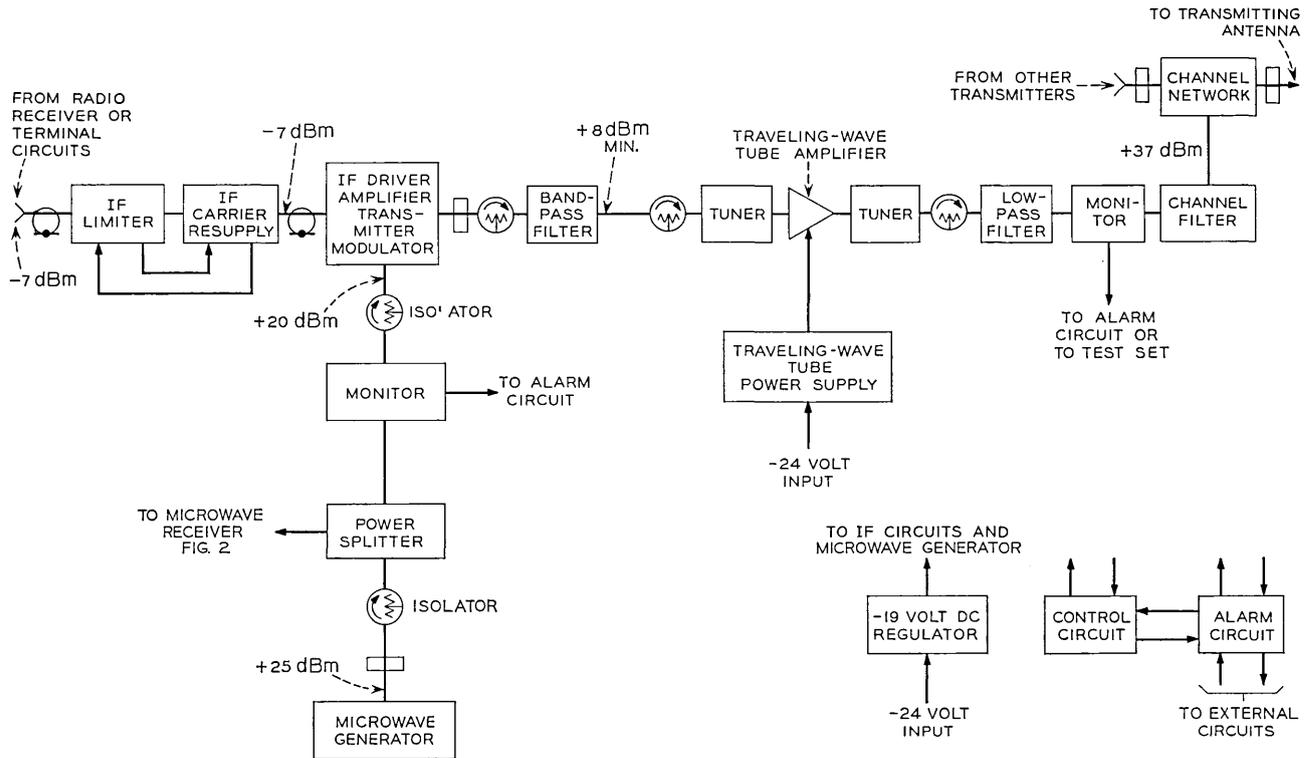


Fig. 1 — Block diagram of TD-3 microwave transmitter.

IF amplifiers to full gain and limiters from spreading high-level noise. Noise spreading would cause adjacent working channels to become unsatisfactory because of excessive noise spillover. The tone modulation on the resupplied carrier provides information to the automatic switching system to indicate that the channel is defective.

The IF driver amplifier-transmitter modulator<sup>8</sup> converts the 70 MHz IF signal to a microwave signal at the transmitter output frequency. The driver amplifier provides approximately 7 dB of gain. The modulator uses a waveguide hybrid junction and a pair of varactor diodes to achieve an upconversion gain of about 8 dB. The beat oscillator power, +20 dBm, is supplied by the microwave generator. The upper and lower sidebands and the beat oscillator leakage signal appear at the output port.

For some channels the modulator is used as an upper sideband upconverter; for the others, as a lower sideband upconverter. The sideband to be transmitted is selected by a band-pass filter. The undesired sideband, which serves as an idler frequency for the upconverter, is reflected by the filter and is dissipated in the reverse loss of an isolator between the upconverter and the filter. Allowing idler frequency currents to flow in this manner results in stable operation, as discussed in Ref. 8. The sideband selecting filter provides about 15 dB of loss to the beat oscillator frequency, and about 30 dB to the unwanted sideband. These losses assure that the power of the 2A-B product generated in the TWT by the beat oscillator and the unwanted sideband is acceptably low.

The desired input to the TWT is a frequency-modulated signal consisting of a carrier and FM sidebands. However, if the signal is transmitted through a circuit that has amplitude or delay distortion, AM components are introduced into the signal. These components are transformed by the relatively high AM-PM conversion coefficient of the TWT into PM components, and they appear as cross-modulation noise. By using low selectivity in the sideband selecting filter,<sup>9</sup> the "in-band" amplitude and delay distortion is small and AM-PM cross-modulation noise from this source is negligible.

The traveling wave tube amplifier uses permanent magnet periodic focusing.<sup>10</sup> It provides approximately 34 dB of gain with an output power of +37 dBm and an amplitude flatness of 0.02 dB over the 12 MHz radio channel. The low pass filter following the amplifier provides 50 dB of loss to second and third harmonics of the carrier frequency.

A monitor, consisting of a directional coupler and a diode detector, is used during transmitter alignment and for monitoring the output power.

The channel filter provides selectivity for the transmitter to assure that energy spillover into adjacent channels is within tolerable levels.<sup>9</sup>

The channel network provides additional selectivity and permits the tandem connection of up to six bays for combining transmitter signals for application to the common transmitting waveguide and antenna.<sup>9</sup>

Isolators, tuners, and attenuators are used in the signal path as required to improve return loss and transmission, and to adjust signal powers.

The microwave generator provides an output of +25 dBm at the appropriate 4 GHz frequency for use as a beat oscillator signal for the radio transmitter and receiver.<sup>11</sup> It uses a crystal oscillator-amplifier unit producing about +37 dBm at  $\frac{1}{3}$  of the output frequency. Varactor diodes in tandem connected quadrupler, doubler, and quadrupler circuits provide the multiplication of 32.

The power splitter, used in repeater bays, is a directional coupler which divides the microwave generator power between the transmitter and receiver loads. Resonance isolators are used to control the reverse transmission characteristics in the generator power distribution circuit. This insures that signals in the transmitter do not couple into the receiver (or vice versa) and produce interfering tones.

A directional coupler and diode detector at the output of the microwave generator are used for testing and monitoring the output power.

An inverter regulator unit supplies electrode voltages to the TWT amplifier. Other circuits are powered from a -19 volt regulator. Control and alarm circuits for alignment and surveillance are provided.

### 3.1.2 *Microwave Receiver*

Figure 2 shows the microwave receiver with typical signal powers. The receiver accepts an input signal on one of 24 radio channels located in the 3710-4170 MHz frequency range. The carrier input power depends on path length and has a nominal value of -28 dBm. The output power is -7.0 dBm at 70 MHz IF.

A channel separation network selects the proper channel and applies it to the channel band-pass filter which provides additional selectivity to reduce adjacent channel interference. An isolator

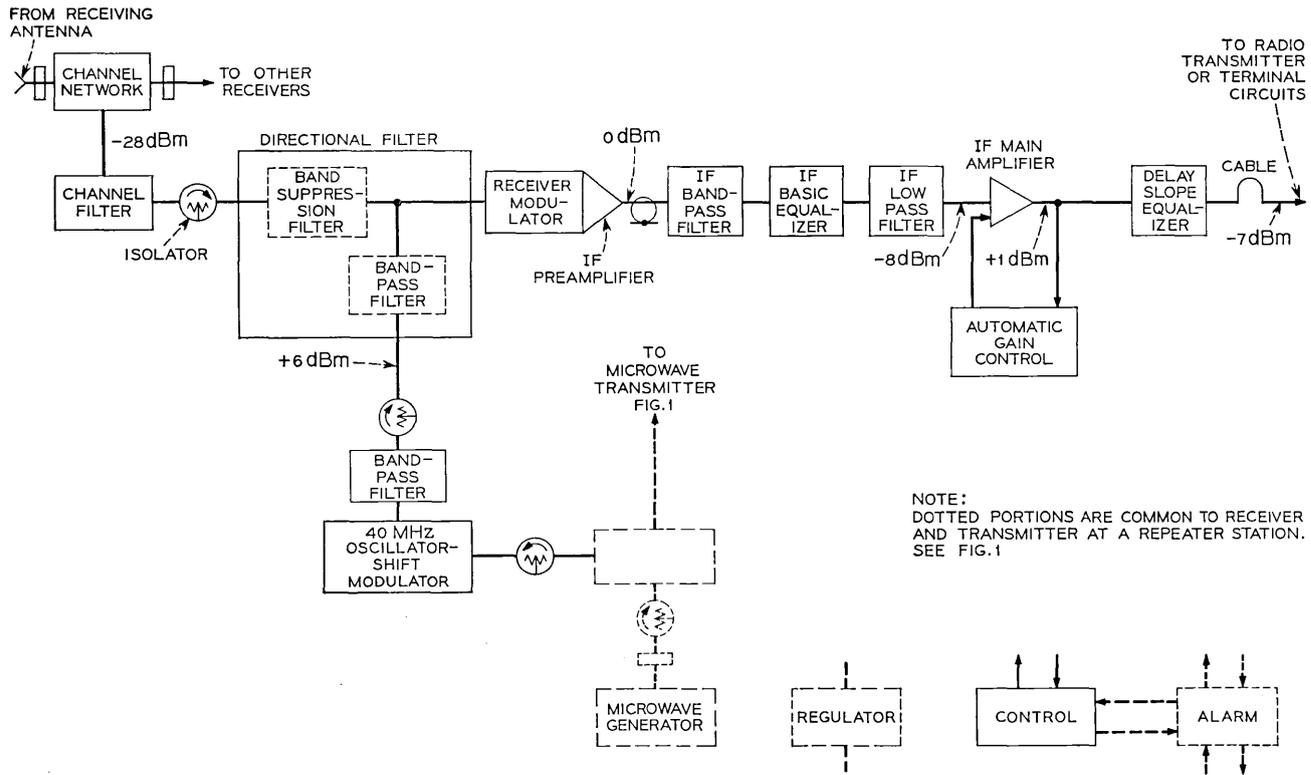


Fig. 2 — Block diagram of TD-3 microwave receiver.

terminates the image frequency generated in the receiver modulator.

The modulator preamplifier consists of an unbalanced Schottky barrier diode modulator and a low noise IF preamplifier.<sup>12</sup> The incoming microwave and beat oscillator signals are combined and applied to the input port of the modulator by a directional filter. The modulator translates the microwave input signal to the 70 MHz IF. The IF output of the modulator is applied to a five-stage preamplifier, which provides about 33 dB of gain. The gain can be adjusted to provide an output of 0 dBm for input signals ranging from -22 to -30 dBm. The noise figure of the modulator-preamplifier is about 6.8 dB.

The IF band-pass filter provides additional receiver selectivity.<sup>9</sup> Its amplitude and delay are internally equalized over the 64-76 MHz range. The basic equalizer corrects the amplitude and delay distortion of the transmitter and receiver microwave filters and networks over the 64-76 MHz range. The IF low-pass filter attenuates harmonics of the 70 MHz carrier frequency. It also is internally equalized over the 64-76 MHz range.

The IF main amplifier has a gain of 9 dB under normal received signal conditions and produces an output of +1 dBm.<sup>7</sup> During receiver input signal fading, the amplifier variolosses circuits, controlled by the AGC circuit, maintain the output power constant for "down" fades as great as 35 dB and "up" fades as large as 10 dB.

A linear (in dB) meter indication of received signal power is provided by a linearizing circuit which monitors the total variolosses control current.

Delay slope equalizers are used as required following the main amplifier to compensate for residual delay slope distortion. A length of IF coaxial cable, as required, is used to build all channels in a switching section to the same electrical length. This insures that large baseband signal phase changes will not occur during a switch to a protection channel. An attenuator is used to build out the loss between the IF main amplifier and the receiver output to maintain an output of -7 dBm.

In a repeater station bay, the beat oscillator signal for the receiver is supplied by a 40 MHz oscillator-shift modulator which shifts the microwave generator frequency by 40 MHz.<sup>8</sup> The shift modulator uses a waveguide hybrid junction and a pair of silicon diodes. The 40 MHz oscillator delivers +17 dBm to the diodes. Approximately +17 dBm of power at the microwave generator frequency also is

applied to the diodes. Either the sum or difference frequency output signal, at a power of about +6 dBm, is selected by a band-pass filter for application to the receiver modulator. The filter reflects the undesired sideband into the modulator, thus improving modulator efficiency.

In a main station radio bay, the shift oscillator-modulator is not used because there are separate microwave generators for the microwave transmitter and receiver. Power, control, and alarm circuits are similar to those used in the transmitter.

### 3.2 *Thermal Noise*

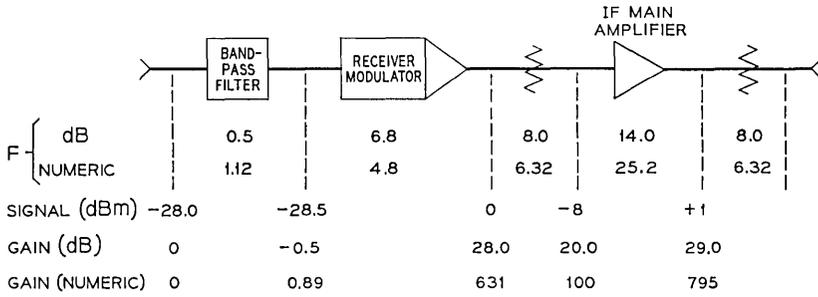
#### 3.2.1 *Noise Figure of a Radio Receiver*

In the TD-3 microwave receiver, the input to the channel band pass filter is the reference point at which the received signal and the receiver noise figure are measured. Typical values indicating the noise contributions for the several parts of the receiver are shown in Fig. 3a. The nominal over-all noise figure of the receiver is 7.5 dB.

With the normal received carrier power and receiver noise figure known, the rms frequency deviation caused by receiver noise as a function of baseband frequency may be calculated. If the system signal deviation is also known, the thermal noise contributed by the receiver may be determined. The rms signal deviation per message band as a function of baseband frequency is shown in Fig. 4. Using the value for a message circuit at 5.77 MHz, and a 7.5 dB receiver noise figure, a top channel circuit noise of 13.5 dBmC0 is obtained. This meets the receiver thermal noise objective imposed by fade margin, but is slightly larger (by 0.1 dB) than the noise allocated to the complete repeater (see Section 2.2.1). Also, so far only receiver signal path noise contributions have been considered. Thermal noise contributed by the microwave generator and the microwave transmitter also must be considered.

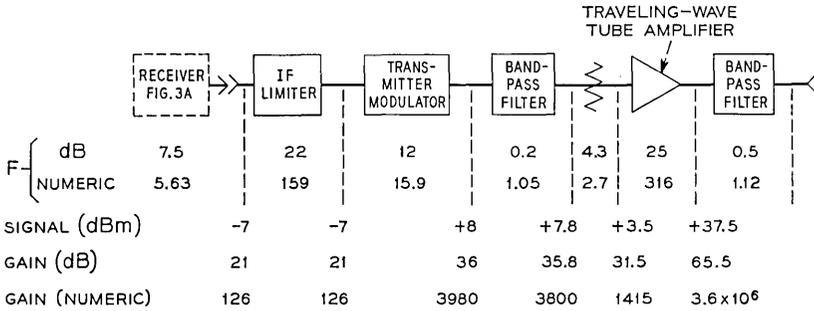
#### 3.2.2 *Radio Repeater Noise*

Under normal received signal conditions, the microwave transmitter (excluding the microwave generator) makes a small but significant contribution to repeater noise. Figure 3b shows that the noise figure of the complete repeater is about 8.5 dB. Curve A of Figure 5 presents the computed thermal noise for a 4000 mile radio line with 140 repeaters, each with 8.5 dB noise figure and a -28 dBm received signal.



$$\begin{aligned}
 F_{\text{RECEIVER}} &= 1.12 + \frac{4.8-1}{0.89} + \frac{6.32-1}{631} + \frac{25.2-1}{100} + \frac{6.32-1}{795} \\
 &= 1.12 + 4.25 + 0.008 + 0.242 + 0.007 \\
 &= 5.63 \text{ OR } \sim 7.5 \text{ dB}
 \end{aligned}$$

(a)



$$\begin{aligned}
 F_{\text{REPEATER}} &= 5.63 + \frac{159-1}{126} + \frac{15.9-1}{126} + \frac{1.05-1}{3980} + \frac{2.7-1}{3800} + \frac{316-1}{1415} + \frac{1.12-1}{3.6 \times 10^6} \\
 &= 5.63 + 1.25 + 0.12 + 0 + 0 + 0.22 + 0 \\
 &= 7.22 \text{ OR } \sim 8.5 \text{ dB}
 \end{aligned}$$

(b)

Fig. 3 — Receiver (a) and repeater (b) noise figure.

### 3.2.3 System Thermal Noise from Microwave Generators

Noise sidebands surrounding the beat oscillator signal contribute to system noise. The sidebands appear as a result of noise within the local oscillator producing phase modulation of the output carrier. The magnitude of the modulation is directly proportional to the carrier-to-noise ratio. Since the output signal of a modulator used as a frequency shifter contains the phase modulation of both input signals, a given carrier-to-noise ratio at the beat oscillator input

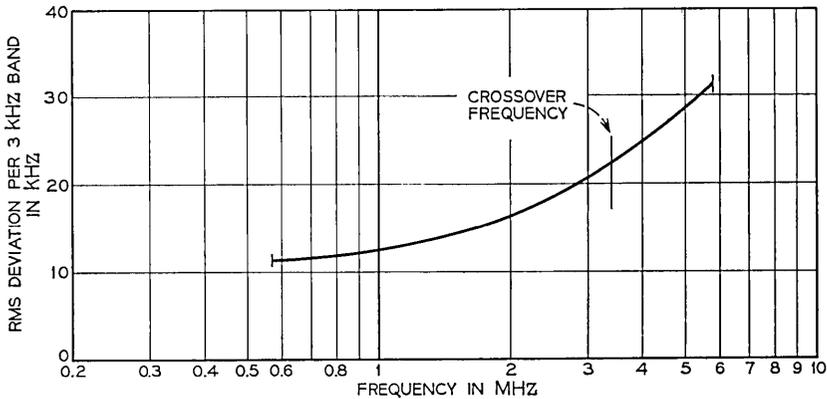


Fig. 4 — Frequency deviation characteristic for 1200 telephone channels. Total RMS deviation: 780 KHz, baseband range: 0.564 to 5.772 MHz.

will have the same effect as an equal carrier-to-noise ratio at the signal input to the receiver modulator.

Figure 6 shows the carrier-to-noise ratio (for a 1 Hz noise band) for a typical TD-3 microwave generator. If separate generators were used for receivers and for transmitters at all stations, the noise

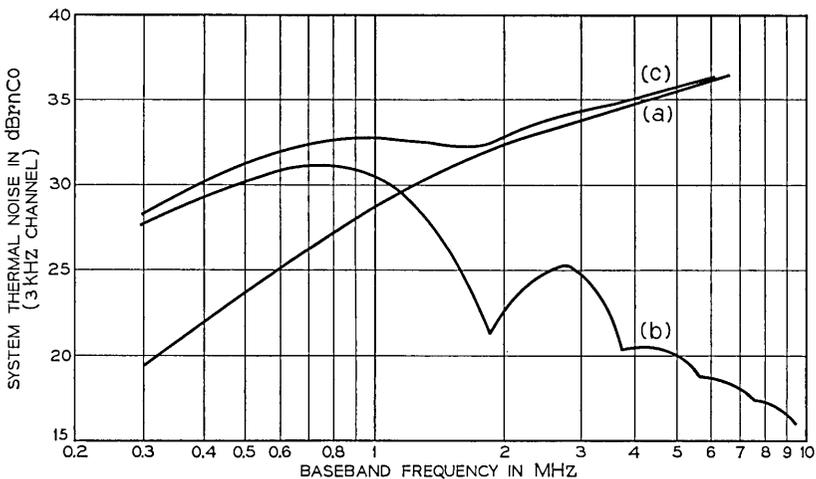


Fig. 5 — System thermal noise. (a) 140 repeaters, each with 8.5 dB noise figure and  $-28.0$  dBm received signal level. (b) 182 generators, 98 at repeater stations and 84 at main stations. (c) Total of a and b.

contributed by the several generators could be added on a power basis to obtain the total system noise from this source. At repeater stations, however, a common generator is used for the receiver and transmitter modulators, and the generator noise introduced at the two modulators will add for some baseband frequencies, and will cancel for others.\* The result is a periodic scalloping of generator noise. The scallop period is determined by the difference between the electrical lengths of two paths within the radio repeater. The first path is the direct one from the microwave generator to the trans-

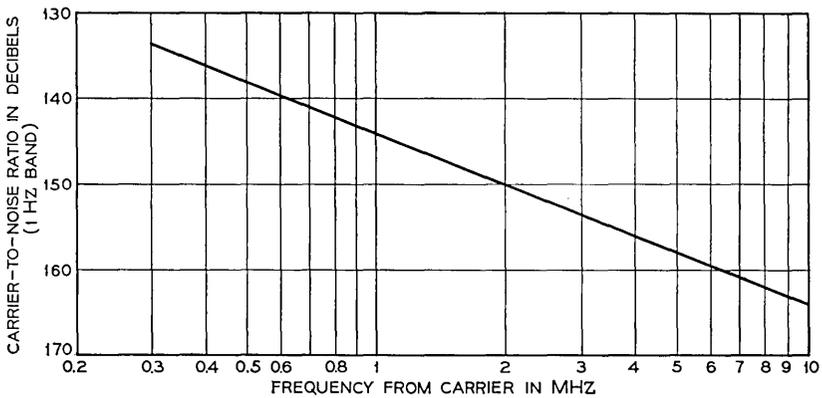


Fig. 6—Typical carrier-to-noise performance of microwave generator.

mitter modulator, and the second is from the generator to the transmitter modulator via the receiver modulator. The expression for calculating the noise addition at a repeater station is given in Appendix A.

Using the data of Fig. 6, and a TD-3 typical delay difference,  $D_3 - D_1 - D_2$  (see Appendix A), of 555 nanoseconds, curve b of Fig. 5 shows the total noise contributed by the two generators at each of 42 main stations, and the single generators at the 98 repeater stations for a 4000-mile system. The gradual decrease of the envelope amplitude from low to high baseband frequencies results from the attenuation characteristic of the band pass filter at the output of the shift modulator used at repeater stations.

\* The noise contribution of the shift modulator-oscillator is small and may be neglected.

### 3.2.4 *Total Thermal Noise of the Radio Line*

Curve c of Fig. 5 results from power addition of curves a and b, and represents the total thermal noise of the radio line model. Notice that generator noise is controlling at low baseband frequencies, and that the total thermal noise is less than 33 dBrnC0 for frequencies below 2 MHz. Above this frequency, noise from sources other than the generators generally dominates, and the total thermal noise gradually increases to 36 dBrnC0 in the top message channel.

The top message circuit thermal noise objective was established at 34.9 dBrnC0, as discussed in Ref. 1. Figure 5 shows that this objective will be missed by about 1 dB. To meet total noise objectives, therefore, the cross modulation noise must be kept below the 36.3 dBrnC0 assigned to it in Section 2.2.2.

## 3.3 *Selectivity Considerations*

### 3.3.1 *Introduction*

The TD-3 microwave transmitter and receiver provide required selectivity in passive networks. The active circuits, designed for broadband response, contribute a negligible amount to the total amplitude and delay distortion. The equalization compensates for the amplitude and delay distortion introduced by the passive selective networks.

The selectivity chosen was influenced greatly by the need to use the frequency plan and antenna arrangement in current use for the TD-2 radio system.<sup>13, 14</sup> This plan uses interspersed transmitter and receiver frequencies, with as little as 20 MHz spacing between carrier frequencies. The antenna side-to-side and back-to-back coupling losses, and the cross-polarization discrimination provide some isolation between the interfering and the disturbed radio channels. These losses, however, are usually insufficient to meet performance objectives, and selective networks must be provided in the transmitter and receiver. In addition to control of intrasystem interferences, selectivity must be provided to meet FCC regulations.

### 3.3.2 *Selectivity in the Microwave Transmitter*

Selectivity is required in the transmitter to suppress the unwanted sideband and beat oscillator signals at the output of the transmitting modulator, and to suppress carrier harmonics generated in the modulator and the TWT amplifier.

The selectivity required falls into two categories: that which must

be placed between the transmitting modulator and the TWT amplifier, and that which must be placed after the TWT amplifier. If the TWT amplifier were perfectly linear, all of the selectivity could be located either ahead of or after it, or it could be arbitrarily divided. Since the amplifier is nonlinear, two considerations apply. First, enough selectivity must be placed ahead of the amplifier to keep sufficiently low the third order modulation products generated in the amplifier by the transmitter modulator products (beat oscillator and upper and lower sideband signals). Second, low selectivity must be used ahead of the amplifier to minimize in-band distortion. This reduces the amount of amplitude modulation added to the FM signal and thus restricts the AM to PM conversion occurring in the TWT amplifier to a tolerable amount.

The  $2A-B$  product generated in the TWT amplifier by the beat oscillator and undesired sideband signals is an exact replica of the desired signal, both in spectrum (no frequency inversion) and index of modulation. However, this product generated in the tube will be delayed relative to the desired sideband; this will be a source of echo-type cross modulation noise. To keep this noise within tolerable limits, it is necessary to attenuate the beat oscillator and undesired sideband signals before they are applied to the traveling wave tube amplifier.

A derivation of the requirements for selectivity at this location is given in Appendix B, which is an example of the analytical methods used to derive selectivity requirements. The requirements in Appendix B were based on measured intermodulation coefficients of typical traveling wave tube amplifier models, and 25 dB suppression of the beat oscillator signal at the transmitter modulator output. In practice, the 25 dB suppression may not always be maintained, and some TWTs may have higher modulation coefficients. Therefore, the sideband selecting filter ahead of the traveling wave tube amplifier has losses (about 15 and 30 dB at 70 and 140 MHz, respectively, from the desired sideband) somewhat greater than those derived in Appendix B. The in-band distortion of this filter is sufficiently low so that cross modulation noise from the AM-PM conversion in the amplifier is negligible.

This filter does not attenuate sufficiently the beat oscillator and unwanted sideband signals to meet requirements at the transmitter output. Therefore, additional selectivity is provided by the channel band-pass filter and by the channel separation network. Actually,

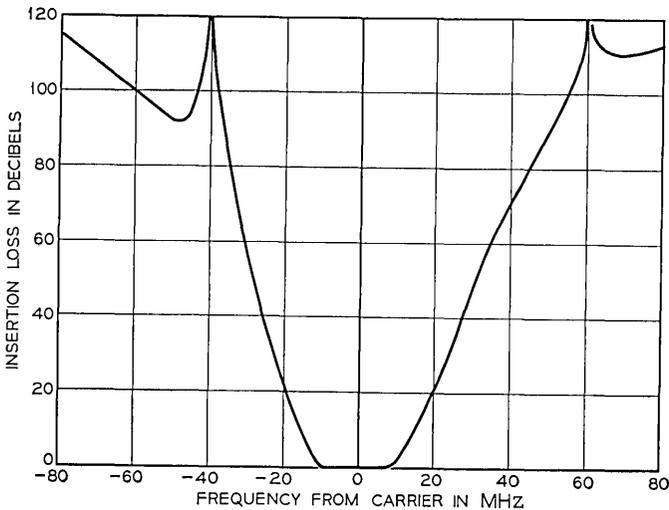


Fig. 7— Microwave selectivity characteristic for microwave receiver. (This characteristic also applies to the microwave transmitter because the same network-filter combination is used.)

the channel band pass filter has somewhat more selectivity than needed, but to reduce the number of filter designs, a filter identical to the one used in the microwave receiver was adopted. The combined selectivity characteristic for the channel network and the channel filter is shown in Fig. 7.

The channel separation network and the channel band pass filter cannot be relied on to provide sufficient suppression of carrier harmonics generated in the TWT amplifier since these waveguide designs can propagate modes other than the dominant one at harmonic frequencies. Therefore, a coaxial low-pass filter is used. It provides more than 50 dB of loss at the second and third harmonic frequencies of the output signal.

### 3.3.3 Microwave Receiver Selectivity Requirements

Selectivity is needed in the receiver to:

- (i) Sufficiently reduce the levels of interfering tones at the receiver output to permit the message circuit tone objectives to be met.
- (ii) Prevent interference from signals at the receiver image frequency.

(iii) Prevent the receiver automatic gain control circuit operation from being affected by intrasystem interference.

(iv) Prevent the "turn-on" threshold of the carrier resupply circuit of the transmitter from being affected by interferences.

(v) Reduce the noise from direct adjacent channel interference to acceptable values.

As with the transmitter, selectivity can be placed at several points in the circuit. Considerable latitude is possible in choosing between microwave and IF selectivity. The plan followed was this: to provide only sufficient selectivity at microwave frequencies (ahead of the receiver modulator) to control third order modulation products generated in the receiver modulator, and to provide the remainder of the selectivity at the intermediate frequency.

In determining the selectivity required, it was assumed that system antenna side-to-side coupling losses of 80 dB or more, and antenna cross-polarization discriminations of 25 dB or more, would be realized. As we already explained, a parametric amplifier was originally used ahead of the receiver modulator. The microwave selectivity provided was based on the modulation performance of this combination. This resulted in more selectivity than that required without the parametric amplifier, but is of little importance because the total selectivity requirements remain about the same. If the microwave selectivity were to be reduced, additional IF selectivity would be needed.

*3.3.3.1 Interfering Tones.* The considerations governing interfering tones are treated in Ref. 1. The required carrier-to-interfering tone ratio, at the output of the receiver, is given by Fig. 6 of that article. For example, a single tone at a frequency corresponding to a baseband frequency of 5.77 MHz (75.77 or 64.23 MHz at the receiver output) must be 94 dB below the 70 MHz IF carrier at the receiver output. An IF tone of this power will produce a tone of  $-70$  dBm0 in a message circuit located at a baseband frequency of 5.77 MHz.

Tones will result from third order modulation products formed in the receiver modulator by unwanted carriers spaced at 20 MHz intervals and transmitted to a receiver through antenna coupling losses, cross-polarization discrimination, and leaks. Products of the  $A+B-C$  and  $2A-B$  type produced by unwanted carriers at  $\pm 20$  and  $\pm 40$  MHz from the desired carrier are the most serious. If all carriers were precisely on assigned frequency, the modulation products would fall at the desired carrier frequency, and unless the products were unreasonably large, no system degradation would result. However, be-

cause of frequency errors in the system, the product may appear as a tone in the message portion of the baseband of the disturbed channel.

Carriers spaced at 80 MHz which are fed to the common waveguide connected to the microwave receivers can produce tones by a similar process. However, when sufficient microwave selectivity is provided to control the effects of the  $\pm 20$  and  $\pm 40$  MHz carriers, it will provide adequate suppression of the carriers at 80 MHz spacing. The selectivity provided is shown in Fig. 7.

Tones can be generated in the receiver modulator by the unwanted sideband signal of the shift modulator. A filter with 112 dB loss at the unwanted sideband frequency (80 MHz from desired output) is used at the shift modulator output to assure that such tones are within tolerable limits.

**3.3.3.2 Image Interference.** Signals at the receiver image frequency (140 MHz from the signal frequency and on the same side as the beat oscillator frequency) will produce a 70 MHz IF output signal. The selectivity provided to control third order modulation products is more than adequate to suppress interference at the image frequency.

Radiation of the beat oscillator frequency must be controlled to prevent interference within the system and with other systems. Such protection is provided by the networks just treated, and by the signal combining directional filter (Ref. 12) immediately ahead of the modulator.

**3.3.3.3 AGC Circuit Protection.** Both the IF main amplifier and the AGC detector circuits are broad band. Therefore, selectivity must be provided ahead of the AGC detector to prevent interfering tones from affecting its output. Tones at 50 and 90 MHz, caused by adjacent channel carriers, are the most troublesome. These tones may remain constant in amplitude while the desired signal is fading, and the AGC operation may be locked up for deep fades. The selectivity provided by the microwave networks is inadequate to prevent such lock-up. However, an IF band-pass filter must be used to control other effects of adjacent channel interference. By locating this filter ahead of the IF main amplifier, the AGC circuit is also protected. The filter used is equalized in amplitude and delay over a 16 MHz band centered at 70 MHz. Its attenuation characteristic is shown in Ref. 9.

**3.3.3.4 Carrier Resupply Operation.\*** Normally, the carrier resupply

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\*This might have been included in Section 3.3.2, but because its performance is influenced primarily by the receiver selectivity, it is included here.

is set to turn on when the receiver input signal fades about 50 dB. A nonfading interfering carrier will therefore affect this unit before it affects the AGC circuit; hence, additional filtering is included in the carrier resupply itself.

*3.3.3.5 Adjacent Channel Interference.* The modulation on an adjacent channel can be a serious source of interference.<sup>1</sup> The usual worst cause of this interference is an adjacent channel ( $\pm 20$  MHz) transmitter coupling into the receiver of the disturbed channel via the antenna side-to-side coupling. The interference is most severe during fades of the disturbed channel, and it affects both the message circuit portion of the baseband and the 9 MHz noise slot used to determine the need for a switch to a protection channel. The amount and distribution of selectivity required to provide a specified degree of protection from this type of interference cannot be easily determined. It is fairly certain that the total receiver selectivity, and probably the transmitter selectivity, is important in controlling this type of interference. More work is planned to obtain understanding of the mechanisms involved. In any event, the field tests on the TD-3 system indicate that adequate protection is provided against adjacent channel interference.

### 3.4 *Equalization*

#### 3.4.1 *Amplitude and Delay*

All active circuits are designed with broadband response, and IF selective networks are self-equalized. The remaining equalization needed is that necessary to compensate for the amplitude and delay distortion of the microwave networks in the transmitter and receiver. At the time TD-3 was developed, suitable microwave equalizers could not be designed. Therefore an IF equalizer, located in the microwave receiver, is used to compensate for the distortion of the microwave networks of a transmitter at one station, and for the distortion of the microwave networks of the associated receiver at the following station. Thus, the equalization is applied on each hop and a corrected signal is made available for the next hop.

To determine the shape needed for the basic equalizer, a large number of simulated radio hops were measured in the laboratory. The amplitude response of the hop was measured to an accuracy of about 0.01 dB, and the delay response to an accuracy of about 0.1 nanosecond. Measurements were made from the IF input of the microwave transmitter to the IF output of the associated micro-

wave receiver. For the delay measurements, the accuracy available using swept frequency measurements was inadequate, and it was necessary to design and construct a precise point-by-point IF delay measuring set. With this set and a broadband upconverter and down-converter it was possible to measure individual microwave networks.

The measurements showed that one group of distortion patterns involve receivers and transmitters that use beat oscillator frequencies below the signal frequencies, and another group with frequencies above the signal frequencies. Another finding, and a very welcome one, was that within a group, the distortion characteristics to be equalized were virtually independent of channel frequency. This meant that just two types of IF basic equalizer were needed. Figure 8 shows the delay distortion characteristic for a hop.

Notice that the curve is not symmetrical about the carrier frequency. Therefore, an IF equalizer designed to compensate for this distortion will need to have this same shape for a channel with the beat oscillator frequency located below the signal frequency. It will need to have a reversed frequency shape for a channel that has the beat oscillator frequency located above the signal frequency.

Using the data collected in the measurement program, a sixth order least squares approximation fit was developed for each type of shape to be equalized. This same procedure was followed for both delay and amplitude shapes. The resulting delay shape to be equalized is shown in Fig. 8, the amplitude shape in Fig. 9. As described

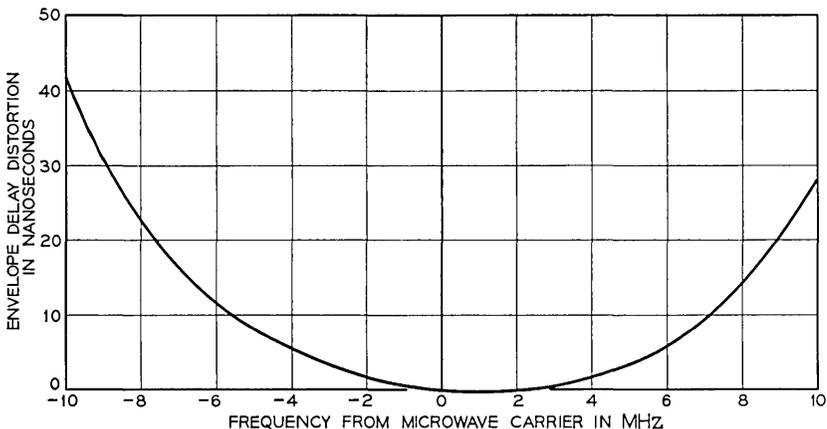


Fig. 8—Envelope delay characteristic of unequalized radio hop. Tandem connection of two channel networks and two channel filters.

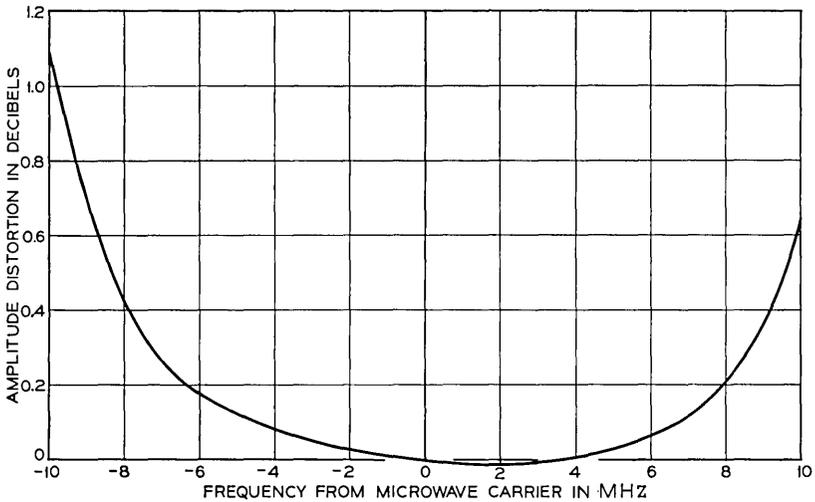


Fig. 9—Amplitude characteristic of unequaled radio hop. Tandem connection of two channel networks and two channel filters.

in Ref. 9, IF equalizers were designed to compensate for these shapes. Adding this equalizer to the hop results in the delay characteristic of Fig. 10, and the amplitude characteristic of Fig. 11.

Over the signal first order sideband spectrum of  $70 \pm 6$  MHz, the residual delay slope is approximately 0.025 nanoseconds per MHz compared with an objective of 0.1 nanoseconds per MHz. The residual

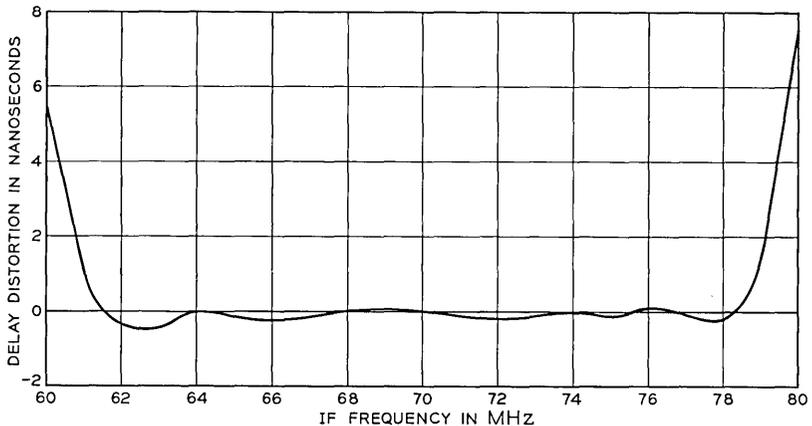


Fig. 10—Envelope delay characteristic of equalized radio hop.

delay ripple is 0.2 nanoseconds peak with a period of about 4 MHz; the objective for a delay ripple of this period is 0.9 nanoseconds peak. Over the  $70 \pm 6$  MHz band, the amplitude distortion is less than 0.03 dB. This meets the objective of 0.05 dB. These data illustrate the close fit between the distortion introduced by the microwave networks and the equalization provided.

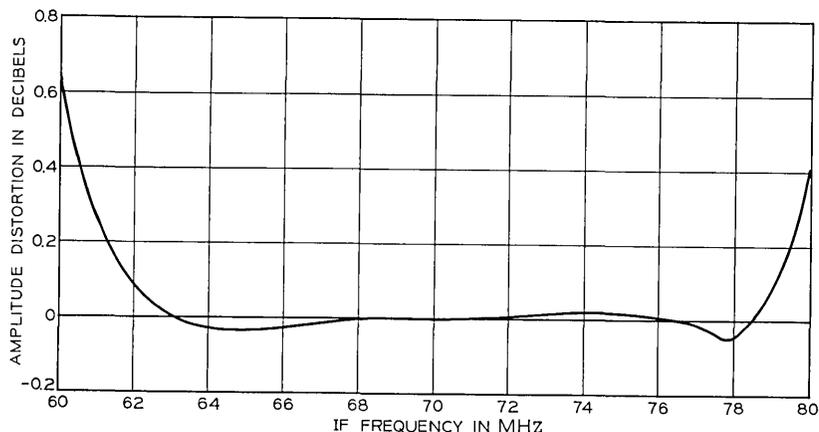


Fig. 11 — Amplitude characteristic of equalized radio hop.

### 3.4.2 Mop-up

Although the match between the characteristics of the microwave networks and the IF equalizer is excellent, some residual distortion will accumulate as a number of hops are connected; therefore, mop-up equalization is provided.

Laboratory and field measurements have shown that the residual amplitude distortion is negligible and that the residual delay distortion is predominantly slope contributed mainly by the adjacent channel network.\* Therefore, a series of fixed delay slope equalizers are available to provide delay slope values of +0.5, +0.25, 0, -0.25, and -0.5 nanoseconds per MHz.

\*In a TD-3 radio station, up to six bays may be connected in tandem, with the channel networks connected in ascending order by frequency, and the lowest frequency network closest to the antenna. For a full bay lineup the networks are spaced in frequency by 80 MHz. Thus the incoming signal to the first bay in the line-up passes through no previous network, the signal for the second bay passes through one previous network, and so on. Only the previous network (80 MHz lower in frequency) adds appreciable delay slope to the desired signal.

The type and number of slope equalizers needed is determined by measuring the delay distortion of a switching section. The slope correction then is distributed among the receivers of the switching section, with not more than one slope equalizer per receiver. This plan is less detrimental to baseband response than equalizing at one location.

### 3.4.3 *Effects of Temperature and Humidity*

It was recognized early in TD-3 development that changes in ambient temperature or humidity would shift the resonant frequencies of the cavities in the microwave networks and filters, but the IF basic equalizer, operating at a much lower frequency, would remain relatively stable. Therefore, as ambient conditions changed, mismatch would occur. To minimize this mismatch, the microwave networks considered in Section 3.3 are constructed of invar, and the radio stations are air conditioned.

As a result of incomplete control, however, some mismatch still will occur. The residual delay distortion from this mismatch will be essentially delay slope whose effect on system cross modulation noise performance can be estimated if some assumptions are made.

To help make such an estimate, a model of a 4000-mile route was hypothesized. The route extended from Boston to San Francisco via Atlanta and Dallas, and was comprised of 140 repeaters. It was assumed that the route was divided into 16 sections with FM terminals at the ends of each section, and with complete baseband circuit frogging at each terminal. Study of three years' weather records for four months in summer and early fall revealed that extremes of temperature and humidity were quite likely to occur simultaneously at all stations in a section of the model. It was logical to assume that cross modulation noise caused by miscalibration from weather would add systematically for stations within a section. Because of circuit frogging, noise of the 16 sections might be expected to add randomly. Therefore, using the 33.3 dBrnC0 allocation from Section 2.2.2, the per-repeater noise allocation becomes:

$$\begin{aligned} N_{REP} &= 33.3 - 10 \log_{10} 16 - 20 \log_{10} \frac{140}{16} \\ &= +2.5 \text{ dBrnC0} \end{aligned}$$

Figure 12 shows the center frequency shift versus temperature for invar networks. Figure 13 shows the shift versus relative humidity.

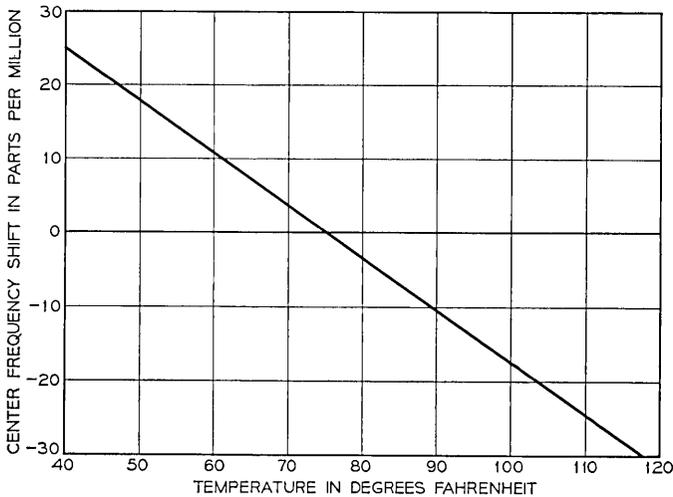


Fig. 12—Frequency shift of tuned cavities versus ambient temperature with relative humidity at 0 per cent.

Values of frequency shift, determined from Figs. 12 or 13, may be used with Fig. 14 to determine the repeater cross-modulation noise resulting from temperature or humidity. For instance, if relative humidity is held at 0 per cent, the repeater cross modulation noise caused by the temperature changing from 75 to 120°F will be +3.5 dBrnC0.\* Similarly, if a fixed temperature of 75°F is assumed, the repeater cross modulation noise caused by the relative humidity changing from 40 to 100 per cent will be in excess of 10 dBrnC0. Thus control of temperature and relative humidity is required to meet the cross modulation noise objective of +2.5 dBrnC0.

To reduce the cost of air-conditioning equipment, a decision was made to control the temperature within stations only to  $75 \pm 20^\circ\text{F}$ . This temperature range requires the application of dry air (less than 5 per cent relative humidity) to the waveguide networks. The cross modulation noise caused by misalignment between the equalizer and the repeater then will meet the +2.5 dBrnC0 objective.

#### 3.4.4 Cross Modulation Noise

No allowance was made for noise caused by transmission distortion followed by an AM-PM converting device when cross modulation

\* Had the microwave networks been constructed of copper rather than invar, this would have been +23.5 dBrnC0.

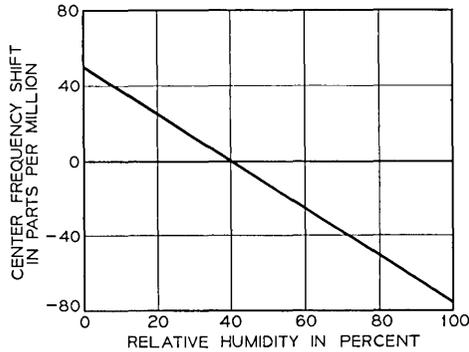


Fig. 13—Frequency shift of tuned cavities versus relative humidity with temperature at 75°F. (Adjusted at 40 per cent relative humidity.)

noise was allocated. It was recognized that circuits would have to be designed to minimize the AM-PM conversion factor. An objective of 0.25 degree per dB maximum was established for this factor for each of the IF active circuits. The objective for the traveling wave tube amplifier was established at 4 degrees per dB maximum.

Tests on the initial installation indicated that excessive cross modu-

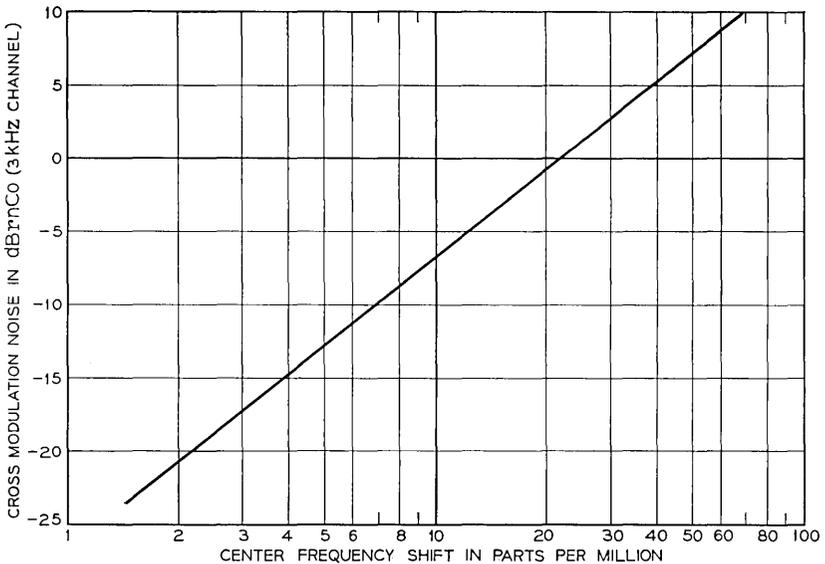


Fig. 14—Cross modulation noise in one repeater caused by frequency shift of tuned cavities.

lation noise was occurring in the IF main amplifier and in the traveling wave tube amplifier. This noise was made negligible by relocating the basic equalizer ahead of the IF main amplifier, by moving the highly selective microwave channel band pass filter to the output side of the TWT amplifier, and by providing a relatively broad (low distortion) sideband-selecting filter ahead of the TWT.

Analytical techniques serve to confirm the validity of these choices.<sup>6</sup> For instance, using typical values of 0.4 and 3.5 degrees per dB for the IF amplifier\* and the traveling wave tube amplifier, respectively, Table II gives maximum distortions which may be permitted ahead

TABLE II—DISTORTION AHEAD OF AN AM-PM CONVERTER†

Distortion type	Magnitude of distortion at $\pm 6$ MHz	
	K = 0.4°/dB (nanoseconds)	K = 3.5°/dB
Linear delay	19.2	6.0
Parabolic delay	4.9	0.6
Cubic delay	16.2	1.9
	(decibels)	
Linear gain	3.8	1.2
Parabolic gain	0.8	0.1
Cubic gain	5.1	0.4
Quartic gain	0.5	0.04

† Distortion which will produce +1.8 dBmC0 of cross modulation noise in the top message channel.

of either of these converting devices to produce cross modulation noise of 1.8 dBmC0.

If the cross modulation noise from either of these sources is assumed to add on a power basis for 140 repeaters, the system noise would be 1.8 + 21.5 or 23.3 dBmC0. This value is 13 dB lower than the total cross modulation noise allocation of 36.3 dBmC0, and therefore would increase the system cross modulation noise by about 0.2 dB.

Figure 8 shows that the delay distortion of the microwave networks and filters of a hop is approximately 12 and 6 nanoseconds at 6 MHz removed from the carrier frequency. If the basic equalizer is located after the IF amplifier, this distortion will appear in the IF signal applied

\*The objective of 0.25 degree per dB maximum for the IF amplifier was not met; a typical number is 0.4 degree per dB. However, this was a tradeoff with the noise figure requirement. Since the basic equalizer was repositioned ahead of the IF amplifier thereby reducing the noise caused by AM-PM conversion, it was judged more important to keep the noise figure low.

to the amplifier. The amplitude distortion of the microwave networks and filters is less than 0.2 dB at  $\pm 6$  MHz: this appears to be no problem.

If the highly selective channel filter was located ahead of the traveling wave tube, it would produce delay distortion of about 11 and 13 nanoseconds at  $\pm 6$  MHz removed from the carrier frequency. By moving the channel filter to the output of the traveling wave tube, and using a broad sideband selecting filter, the delay distortion preceding the tube is reduced to 0.3 nanoseconds at  $\pm 6$  MHz. This will contribute substantially less than +1.8 dB of cross modulation noise. Similarly, the 0.2 dB amplitude distortion in the channel filter has been reduced to 0.03 dB by using the broad side-band-selecting filter ahead of the tube. As in the case of the IF amplifier, the delay distortion ahead of the AM-PM converter is more serious than the amplitude distortion.

#### IV. EQUIPMENT DESCRIPTION

At a repeater station, the microwave receiver and transmitter in a bay serve one direction of transmission only, and the IF output of the receiver is connected directly to the IF input of the associated transmitter. This combination is called a repeater station bay, and is illustrated in Fig. 15. The bay uses a single microwave generator and a single -19 volt regulator. The equipment described here is manufactured by the Western Electric Company for Bell System use only.

At a main station, the microwave transmitter and receiver serve opposite directions of transmission. The IF output of the receiver, and the IF input to the transmitter, are connected to IF switching, patching, and distribution circuits in the station. This main station bay uses one microwave generator and one regulator for the transmitter, and a second one of each for the receiver to provide independent operation for the two directions of transmission. This improves reliability and facilitates maintenance.

The transmitter, receiver, and associated control units are packaged on a 9-foot, 19-inch, unequal flange, duct-type bay framework about 22½ inches wide and 15½ inches deep.

A clear plastic cover (top of bay in Fig. 15) protects the precisely tuned channel networks from mechanical damage which would change electrical characteristics.

The bays can be mounted back-to-back or against a wall; all mounted equipment is accessible and removable from the front.

The lower portion of the bay contains one regulator and one micro-

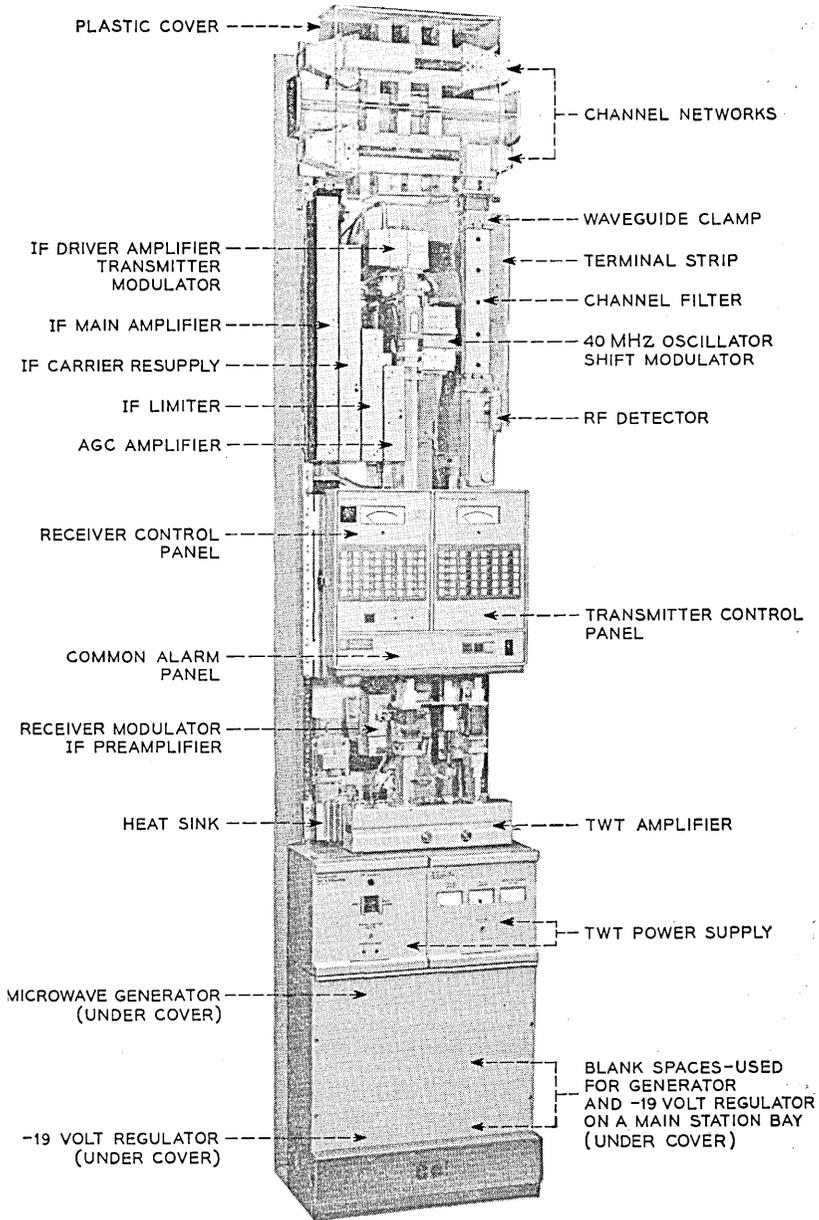


Fig. 15 — TD-3 transmitter-receiver bay for repeater station.

wave generator for a repeater station bay, and two regulators and two generators for a main station bay. These components plug in to permit easy removal. They are supported by nylon slides and protected by a front cover with quick release fasteners.

The traveling wave tube amplifier power supply mounts just above the upper microwave generator. It consists of two plug-in units: the low-voltage inverter unit on the left must be removed before the high-voltage rectifier unit on the right can be removed. Removing the low-voltage unit de-energizes the high-voltage unit making it safe for personnel. The power supply voltages are applied to the TWT amplifier by a cable that passes through an opening in the top of the power supply framework. The cable terminates in a plug that connects to a receptacle on the tube.

The TWT amplifier is on top of the power supply, with its power receptacle facing the supply. This arrangement encloses the power cord and its connector for greatest personnel safety. A dummy test load may be mounted in two slots on top of the supply framework without removing the TWT, and the power connection may be made to it. The dummy load is used to determine if the TWT amplifier or the power supply is faulty in case of transmitter output power failure.

The collector electrode of the TWT amplifier is connected to a cooling block on the TWT package. To achieve long life for the tube, it is necessary to restrict the cooling block temperature to 150°F maximum. To accomplish this, the TWT cooling block is connected to a heat sink made of cooling fins bolted to the bay uprights.

The control and alarm circuits are housed in a doorlike type framework mounted near the middle of the bay. The receiver control panel is at the left, the transmitter panel is at the right, and the common alarm panel is at the bottom. The door latches on the left side and swings open for access to apparatus behind the door, to read meter-relays, or to remove control units easily.

The receiver control units for repeater and main station bays are identical, and include a circuit breaker to prevent overloading of the -24 volt source, pushbuttons for connecting metered functions to a panel meter, a switch to select manual or automatic gain control, and a meter-relay to monitor the microwave generator output power. When the microwave generator output drops by a predetermined amount, the meter-relay provides an alarm. A pushbutton switch permits checking the voltage regulator output.

The transmitter control unit for a repeater station bay includes

pushbuttons for connecting metered functions to a panel meter, and a meter-relay to monitor the transmitter output power and to provide a low output alarm. The transmitter control circuit for a main station is similar except that a circuit breaker and a pushbutton switch for checking the voltage regulator output is provided.

The common alarm unit receives transmitter and receiver alarms as grounds on the input leads and translates them into closed contacts on the output leads for operation of external audible and visible alarms. It also has an alarm cutoff key for disabling an external audible alarm, and an alarm reset switch for resetting the receiver and transmitter meter-relays.

The receiver, shift, and transmitter modulators are physically and electrically associated with the IF preamplifier, 40 MHz shift oscillator, and IF driver amplifier, respectively.

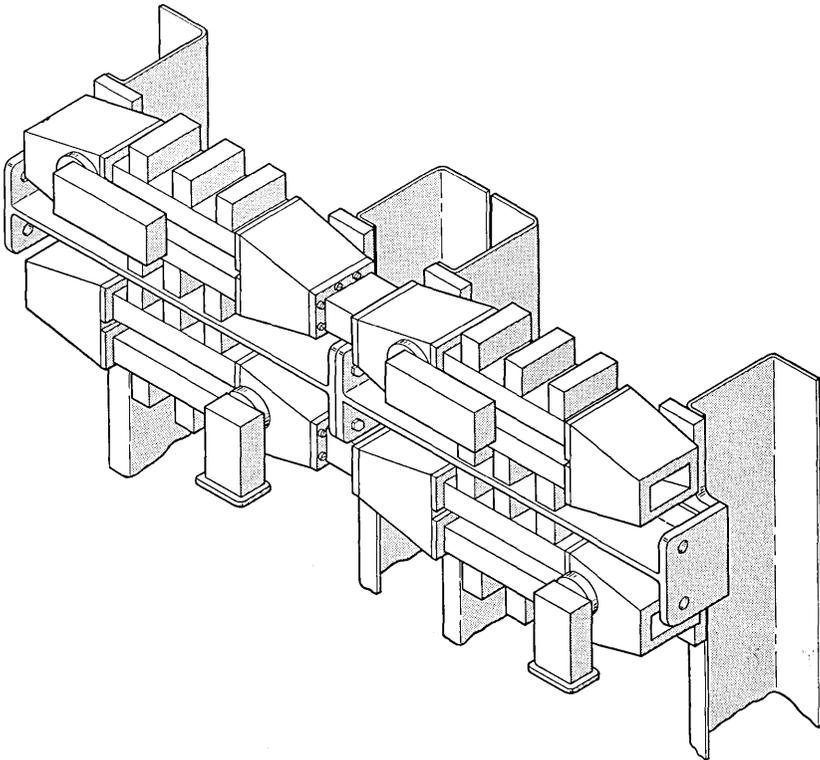


Fig. 16— Channel separation network casting assembly.

The four active IF units are mounted on a doorlike framework at the left of the bay directly above the control door assembly. The units are from left to right: main amplifier, carrier resupply, limiter and AGC amplifier. The door is hinged on the left side and swings open to expose the connectors at the rear and the waveguide apparatus behind the door.

The passive IF units are mounted to the bay framework and are located so as to keep the interconnecting cable to a minimum. The units are: band-pass filter, low-pass filter, basic equalizer, and delay slope equalizer.

The channel separating networks are attached to a rugged casting at the top of the bay. The casting is designed to connect to castings of adjacent bays as illustrated by Fig. 16. This arrangement aligns adjacent network ports within a fraction of an inch to permit them to be connected by short sections of flexible waveguide; it eliminates mounting and positioning the networks of the bays separately in a line since the connection of the castings provides sufficient alignment.

Several sections of waveguide within the bay are removable to permit access for routine testing. These sections are connected by waveguide clamps to permit removal.

#### APPENDIX A

##### *Baseband Noise*

It can be shown that the output signal of a modulator used as a frequency shifter contains the phase modulation of both the input signal and beat oscillator signal. In a repeater station, one microwave generator serves as the beat oscillator for both the receiver and transmitter modulators. Therefore, generator noise will appear as phase modulation at the output of both of these modulators. The addition of these two noise outputs is dependent upon their phase relationships. This appendix derives the equation for calculating the baseband noise due to the addition of noise at the outputs of the two modulators.

let  $\phi_m(t)$  = message modulation on signal

$\phi_n(t)$  = noise modulation on beat oscillator signal

$D_1$  = absolute delay in the RF path between the microwave generator and the receiver modulator

$D_2$  = absolute delay in the IF path between receiver and transmitter modulators

$D_3$  = absolute delay in the RF path between the microwave generator and the transmitter modulator.

Receiver Modulator

$$\text{Input Signal} = \cos [\omega_1(t) + \phi_m(t)]$$

Receiver Modulator

$$\text{Beat Oscillator Signal} = \cos [\omega_2(t + D_1) + \phi_n(t + D_1)]$$

Transmitter Modulator

$$\text{Beat Oscillator Signal} = \cos [\omega_3(t + D_3) + \phi_n(t + D_3)]$$

Then the receiver modulator output signal is:

$$K_1 \cos [(\omega_1 - \omega_2)t - \omega_2 D_1 + \phi_m(t) - \phi_n(t + D_1)]$$

The transmitter modulator input signal is:

$$K_2 \cos [(\omega_1 - \omega_2)(t + D_2) - \omega_2 D_1 + \phi_m(t + D_2) - \phi_n(t + D_1 + D_2)]$$

The transmitter modulator output signal is:

$$K_3 \cos [(\omega_1 - \omega_2)(t + D_2) - \omega_2 D_1 + \omega_3(t + D_3) + \phi_n(t + D_3) + \phi_m(t + D_2) - \phi_n(t + D_1 + D_2)]$$

Collecting the phase modulation terms and shifting the time axis by  $D_1 + D_2$ :

$$\phi_n(t + D_3 - D_1 - D_2) - \phi_n(t) + \phi_m(t - D_1)$$

The output of the discriminator in the FM receiver is proportional to the time derivative of phase modulation terms at its input. Therefore  $\phi_n(t)$  becomes  $\phi'_n(t)$  at the discriminator output. The transfer function which operates on  $\phi'_n(t)$  may be found as follows:

$$\begin{aligned} \phi'_n(t + D_3 - D_1 - D_2) - \phi'_n(t) &= \phi'_n(t) \exp [-j(D_3 - D_1 - D_2)\omega] - 1 \\ &= -\phi'_n(t) \exp \frac{-j(D_3 - D_1 - D_2)\omega}{2} \\ &\quad \cdot \left[ \exp \frac{j(D_3 - D_1 - D_2)\omega}{2} - \exp \frac{-j(D_3 - D_1 - D_2)\omega}{2} \right] \end{aligned}$$

where

$\omega$  = baseband frequency

The magnitude of the above function

$$= 2\phi'_n(t) \sin \frac{(D_3 - D_1 - D_2)\omega}{2}$$

or in dB,

$$= 6 + 20 \log_{10} \phi'_n(t) + 20 \log_{10} \sin \frac{(D_3 - D_1 - D_2)\omega}{2}$$

This transfer function may be used to calculate the addition of noise at the output of a receiver modulator and transmitter modulator at a repeater station. It is seen that the transfer function has a maximum value (in-phase addition) of

$$6 + 20 \log_{10} \phi'_n(t) \quad \text{when} \quad \frac{(D_3 - D_1 - D_2)\omega}{2} = \frac{\pi}{2}, \frac{3\pi}{2}, \frac{5\pi}{2}, \text{ etc.}$$

The transfer function has a minimum value of  $-\infty$  (cancellation) when  $(D_3 - D_1 - D_2)\omega/2 = 0, \pi, 2\pi, \text{ etc.}$

It should be noted that  $D_3 - D_1 - D_2 = D_3 - (D_1 + D_2)$ .

Therefore the difference in delay between the two paths from the generator to the transmitter modulator determines the period of this function.

## APPENDIX B

### *Selectivity*

This appendix shows the derivation of selectivity requirements at the traveling wave tube amplifier input in the microwave transmitter. The signal applied to the input from the transmitter modulator has three components which are assumed to have the following powers:

(i) The desired sideband, +8.5 dBm.

(ii) The beat oscillator frequency, displaced 70 MHz with respect to the desired sideband, -5 dBm. (The transmitter modulator is assumed to provide 25 dB of suppression to the +20 dBm beat oscillator input.)

(iii) The undesired sideband, displaced 140 MHz with respect to the desired sideband (in same direction as the best oscillator frequency), +8.5 dBm.

The beat oscillator and undesired sideband signals will produce a  $2A-B$  product which is an exact replica of the desired signal. However, because of delay differences, this product will be delayed relative to the desired sideband and will be a source of echo-type cross modulation noise.

To establish filter selectivity requirements, a delay difference of 35 nanoseconds between the desired and undesired sidebands, and the

beat oscillator frequency has been assumed. Available contour curves showed that the ratio between desired sideband power and the echo power can be 24 dB lower at RF than the required signal-to-noise ratio at baseband.<sup>4</sup>

Now if the noise resulting from the  $2A-B$  product is permitted to increase system intermodulation noise by 0.1 dB, the echo noise should be  $36.3 - 16 = 20.3$  dBmC0. If systematic addition is assumed for 140 repeaters, the per-repeater requirement for echo noise is  $20.3 - 20 \log_{10} 140 = -22.7$  dBmC0.

The signal-to-noise requirement at baseband may be derived as follows:

The signal at baseband can be derived using the system constants listed in Ref. 1; that is, a  $-10$  dBm0 power for an average talker and a 25 per cent activity factor. This yields a talker power of  $-16$  dBm0 in each of the 1200 voice channels. Reference 5 shows that a power of 0 dBm0 corresponds to  $+88$  dBmC0. Therefore, the signal at baseband is  $-16 + 88 = +72$  dBmC0.

The noise at baseband is  $-22.7$  dBmC0 as derived above.

Therefore, the signal-to-noise ratio at baseband becomes  $+72 - (-22.7) = 94.7$  dB.

When the 24 dB factor from above is included, the ratio between desired sideband power and echo power at the tube input must be  $94.7 - 24 \cong 71$  dB.

Thus, for a desired sideband power of  $+8.5$  dBm, the  $2A-B$  product power referred to the input to the tube must be less than  $-62.5$  dBm. The intermodulation performance of the traveling wave tube amplifier can be described by the following equation:

$$P_{2A-B} = 2P_A + P_B + K_{2A-B} \text{ dBm.} \quad (1)$$

where all powers are referred to the tube input and where

$P_{2A-B}$  = Power of  $2A - B$  product in dBm

$P_A$  = Power of beat oscillator frequency in dBm

$P_B$  = Power of undesired sideband in dBm

$K_{2A-B}$  =  $-22$  dB =  $2A - B$  intermodulation coefficient measured on traveling wave tube amplifier.

From above,

$$\begin{aligned} -62.5 &= 2P_A + P_B - 22 \\ \therefore 2P_A + P_B &= -40.5 \text{ dBm.} \end{aligned} \quad (2)$$

For no filtering between the upconverter and the traveling-wave-tube amplifier, the values for  $P_A$  and  $P_B$  are equal to  $-5$  and  $+8.5$  dBm, respectively, and obviously will not satisfy equation (2).

The problem at this point is to select filter loss values at 70 MHz and 140 MHz removed from the desired carrier in order to reduce  $P_A$  and  $P_B$  at the amplifier input so that the conditions of equation (2) may be satisfied.

By substituting the following values into equation (2),

$$P_A = (-5 - FL_{70}) \text{ dBm}$$

$$P_B = (+8.5 - FL_{140}) \text{ dBm}$$

where  $FL_{70}$  and  $FL_{140}$  represent the filter loss at 70 and 140 MHz, respectively, from the desired sideband,

$$2(-5 - FL_{70}) + (+8.5 - FL_{140}) = -40.5 \text{ dBm.}$$

$$\therefore 2FL_{70} + FL_{140} = 39 \text{ dB.} \quad (3)$$

The following tabulation shows the range of losses that may be provided to satisfy equation (3):

Loss in dB		dBm		
$FL_{70}$	$FL_{140}$	$P_A$	$P_B$	$2P_A + P_B$
5	29	-10	-20.5	-40.5
10	19	-15	-10.5	-40.5

The loss of the filter, therefore, should be in the range of 5 to 10 dB at 70 MHz from the desired carrier, and in the range of 29 to 19 dB at 140 MHz from the desired carrier.

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# Active IF Units for the Transmitter and Receiver

By G. L. FENDERSON, J. J. JANSEN, and S. H. LEE

(Manuscript received November 27, 1967)

*TD-3 IF circuits are designed to operate with a carrier frequency of 70 MHz with extremely flat transmission characteristics over the 60–80 MHz band. The automatic gain control of the IF main amplifier and the excess gain of the limiter together provide an effective AGC range of more than 40 dB. The units are designed to provide low noise figures and low AM-PM conversion.*

## I. INTRODUCTION

The TD-3 microwave radio system uses heterodyne repeaters in which the IF frequency band is centered at 70 MHz. System equalization, and part of the selectivity is provided at 70 MHz along with over half of the maximum available gain. The IF gain is controlled by an automatic gain control (AGC) circuit to compensate for variations in received signal power at the input of the microwave receiver.

The repeater also includes an IF limiter and an IF carrier resupply unit. The limiter suppresses incidental amplitude modulation which may be added to the frequency modulation signal by transmission deviations. The resupply circuit automatically inserts a substitute carrier in the event that the transmitted carrier fails or suffers a very deep fade. Loss of the transmitted carrier would cause IF amplifiers in succeeding repeaters to go to maximum gain, causing a noise build-up in the failed channel that would spread into the adjacent radio channels.

The design and performance characteristics of the IF preamplifier, the IF main amplifier and AGC circuit, the limiter, and the carrier resupply unit are described in this paper. The design of the IF driver amplifier for the transmitter modulator is covered in a companion paper.<sup>1</sup> Equipment described in this article is manufactured by the Western Electric Co. for Bell System use only.

## II. DESIGN CONSIDERATIONS

The major design problems were to achieve:

- (i) Satisfactory transmission characteristics
- (ii) The required AGC range without appreciably degrading the repeater noise figure or the transmission characteristic of the main IF amplifier
- (iii) Low noise figures
- (iv) Adequate power handling capability with low AM to PM conversion.

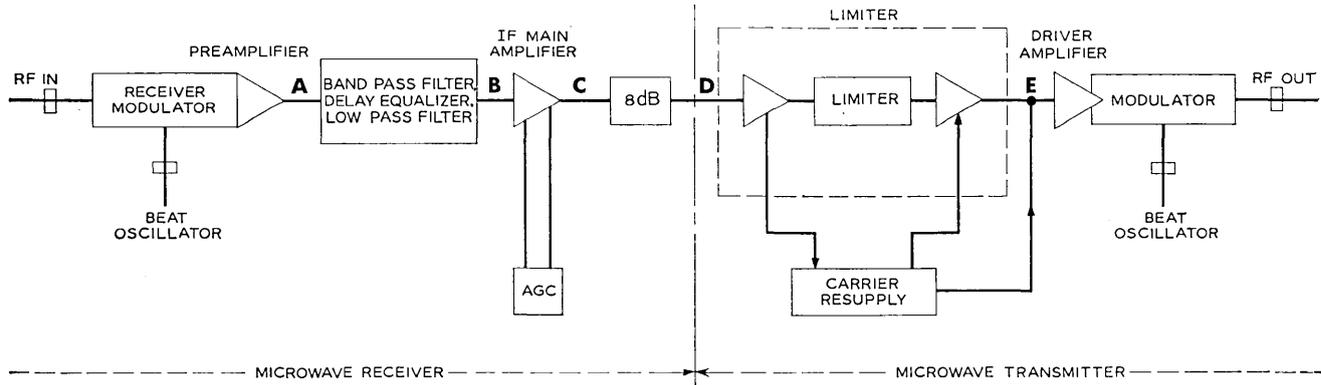
The design philosophy adopted was to make all the active circuits broad band to achieve good control over amplitude and delay characteristics. The 3 dB points of individual IF amplifiers fall typically below 5 MHz and above 130 MHz; and the transmission characteristics can be adjusted to within 0.01 dB over a range of  $70 \pm 6$  MHz (for the first order sidebands) and to within 0.03 dB over the remainder of the  $\pm 10$  MHz channel width. Band shaping is accomplished in the passive networks.<sup>2</sup>

The broadband approach, however, means that harmonics of the IF signal generated in the amplifier stages are not attenuated significantly. Filters must be relied on where harmonics are troublesome. For example, if second harmonics generated in the preamplifier were allowed to reach the IF main amplifier, they would modulate with the fundamental in the IF main amplifier to produce products at fundamental frequency. Since the delay through the IF filter and equalizer is greater at 70 MHz than at 140 MHz, these products may be viewed as leading echoes that contribute to modulation noise. This problem is most severe with an up-fade or a high signal level.

To remove this noise source, a delay equalized low-pass filter is used between the preamplifier and the IF main amplifier. The interfering effect of an echo is greatly reduced, however, as the delay difference between the fundamental and the harmonic is minimized. If the delay difference is sufficiently small, a low-pass filter may not be required. Thus, harmonics produced in the IF main amplifier do not produce echoes that contribute appreciable modulation noise in the limiter, because networks with large delays are not used between the IF main amplifier and the limiter.

The AM to PM conversion objective set for these amplifiers is  $\frac{1}{4}$  degree per dB at normal transmission level.

Figure 1 illustrates the types and positions of both the active and



Operating condition	A	B	C	D	E
Normal received power -28.5 dBm	0dBm	-8dBm	+1dBm	-7dBm	-7dBm
6 dB upfade	+ 6	- 2	+ 1	- 7	- 7
35 dB fade	-35	-43	+ 1	- 7	- 7
41 dB fade (100A operation)	-41	-49	- 5	-13	- 7.4
49 dB fade	-49	-57	-13	-21	-12
50 dB fade Carr resupply operates	-50	-58	-14	-22	- 7

Fig. 1 — IF portion of a radio repeater.

passive IF units in a radio repeater. The chart shows the signal power at various points for several operating conditions. The characteristics of the IF networks, and the considerations that led to their assigned positions in the repeater are covered in a companion paper.<sup>3</sup> At main stations the loss of the attenuator following the main IF amplifier is less than 8 dB to allow for the loss of mop-up equalizers. Enough loss has been provided to allow for further mop-up equalization if operating experience shows this to be desirable. For example, parabolic delay correction could be introduced readily.

The first data line in the Fig. 1 chart describes the average normal case for which the signal at the input to the receiver modulator is  $-28.5$  dBm. The gain of the IF preamplifier can be adjusted to achieve an output of 0 dBm for a microwave input signal falling between  $-22$  dBm and  $-30$  dBm. The IF main amplifier is operated at  $+1$  dBm output while the limiter has an over-all gain of unity and is designed to operate at  $-7$  dBm.

A short radio path is not the only situation that results in a received signal higher than  $-28.5$  dBm. Reflection in the path between two microwave radio stations can enhance as well as fade the received signal. In practice, a 6 dB up-fade occurs relatively frequently for the longer hops and may last for an appreciable length of time. Higher up-fades are experienced occasionally. Hence the load handling ability of the preamplifier and the IF main amplifier must be adequate to permit at least a 6 dB up-fade without an appreciable increase in system cross modulation.

The IF main amplifier for TD-3 has a 35 dB AGC range. Fades deeper than 35 dB will result in a decreasing output of the IF main amplifier; however, the limiter provides added fade range. At the point where the noise of a system reaches the limit of acceptability (55 dBmCO), the 100A switching system operates to substitute the protection radio channel for the working radio channel.<sup>4</sup> Assuming 14 dBmCO of thermal noise per hop, the 100A system will operate when a single hop fades about 41 dB.

The Fig. 1 chart also shows signal powers for deep fades. For a 49 dB fade, for example, the driver-amplifier receives a signal 5 dB down from normal. (Because of the compression characteristics of the transmitting modulator and the traveling wave tube, however, the transmitted output power drops by only about 2 dB.) When the fade reaches about 50 dB, noise spreading into adjacent radio channels

may become serious. The carrier resupply unit is set to operate at this fade depth. Through transmission is interrupted and a substitute carrier is applied to the microwave transmitter.

Since the 100A switching system must be able to recognize a failed channel, the carrier resupply delivers an FM signal to the driver amplifier. This FM signal has 9 MHz sidebands for a failed working channel and 7 MHz sidebands for a failed protection channel which the initiator of the switching system interprets as noise build-up in a 9 MHz slot or the absence of a head end bridge. The carrier resupply unit also initiates an alarm that operates 45 seconds after the resupply itself has operated. The delay insures that the alarm system does not respond to short duration atmospheric fades.

III. IF MAIN AMPLIFIER

The IF main amplifier is the output amplifier for the microwave receiver. It provides an output of +1 dBm and normally operates at a gain of 9 dB. Under control of the associated AGC unit, this gain can be increased to 44 dB or decreased to less than -1 dB. The amplifier uses 18 silicon transistors (WE 45B) in 10 gain stages and 7 variolossor stages. A monitor stage has an output to drive the AGC circuit.

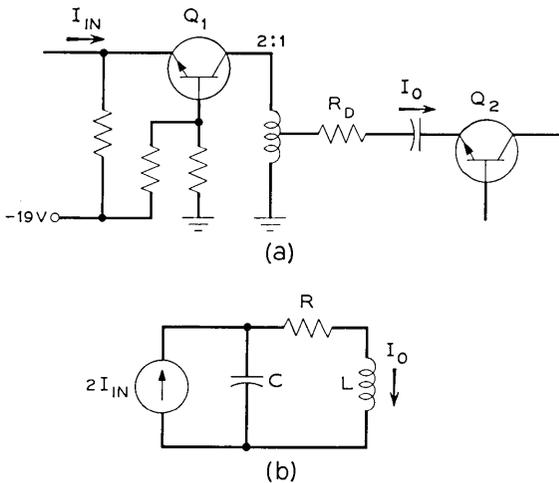


Fig. 2—Schematic and equivalent circuit of common base transformer coupled stage.

The basic gain building block of the amplifier, shown in Fig. 2a, is the common-base, transformer-coupled gain stage.<sup>5</sup> Figure 2b is a simplified equivalent circuit where  $C$  represents the sum of the collector and transformer capacitances,  $R$  is the sum of the damping resistor ( $R_D$ ) and the emitter diffusion resistance, while  $L$  is the sum of the transformer leakage and emitter inductance. These elements as shown have been referred to the low impedance side of the transformer. The transmission expression is given by:

$$20 \log | I_o/I_{IN} | = 6 - 10 \log [(f/f_p)^4 + (4\delta^2 - 2)(f/f_p)^2 + 1] \quad (1)$$

where

$$f_p = \frac{1}{2\pi(LC)^{1/2}}$$

$$\delta = \pi RCf_p .$$

Using a transformer with a 2:1 turns ratio, the theoretical maximum gain per stage is 6 dB, but owing to transformer core losses, the actual gain which can be achieved is approximately 5 dB. The frequency response is determined by the second term of equation (1). The important parameters are  $\delta$  (the damping factor) and  $f_p$  (the peak frequency).

The optimum value of  $\delta$  (hence  $R_D$ ) may be found by setting the derivative of equation (1) equal to zero at the center of the IF band. The resulting transmission expression becomes:

$$20 \log | I_o/I_{IN} | = 6 - 10 \log [(f/f_p)^4 - 2(f_c/f_p)^2(f/f_p)^2 + 1], \quad (2)$$

with

$$R = \left\{ [1 - (f_c/f_p)^2] \frac{L}{C} \right\}^{1/2}, \quad \text{and} \quad R_D = R - R_E . \quad (3)$$

$f_c$  is the center frequency of the IF band and  $R_E$  is the emitter resistance.

Under these conditions the remaining transmission distortion is primarily parabolic, as shown in Fig. 3a, with the magnitude of the distortion determined by the peak frequency ( $f_p$ ) as shown in Fig. 3b. The peak frequency is therefore a convenient figure of merit for the stage and can be measured by noticing the frequency at which the response peaks when the damping resistor value is zero.

Consider next two stages in tandem, where each stage has a transmission characteristic corresponding to Fig. 3a. Stagger tuning may be accomplished by adjusting the first stage damping so that  $f_{c1}$  is below  $f_c$  (equation 3), and the second stage damping so that  $f_{c2}$  is above  $f_c$ . Optimum choices for  $f_{c1}$  and  $f_{c2}$  will yield an over-all transmission characteristic with virtually no linear or parabolic gain distortion. It is not necessary to provide each stage in a multistage amplifier with adjustable damping because the damping of two stages can be adjusted to introduce a parabolic gain distortion which is opposite to that shown in Fig. 3a, and which therefore can be used to compensate for the distortion of other stages. If two stages are made adjustable and the remaining stages are damped corresponding to a nominal optimum in accord with equation (3), the over-all transmission characteristic may be adjusted to have virtually no linear or parabolic transmission distortion. The IF main amplifier was designed in this manner; two stages have adjustable damping and the remaining eight have fixed damping.

Stagger tuning eases the peak frequency requirement considerably. If all ten gain stages were made alike, Fig. 3b shows that an  $f_p$  greater than 300 MHz would be required to obtain an over-all transmission characteristic that was flat to 0.01 dB over the 20 MHz band. However, a high value of  $f_p$  is still desirable for stability and easy adjustment.

The two components that contribute significantly to determining peak frequency are the transistor and the transformer. The Western Electric 45B transistor was selected as a compromise between peak frequency and power handling capability.<sup>6</sup> The transformer, designed

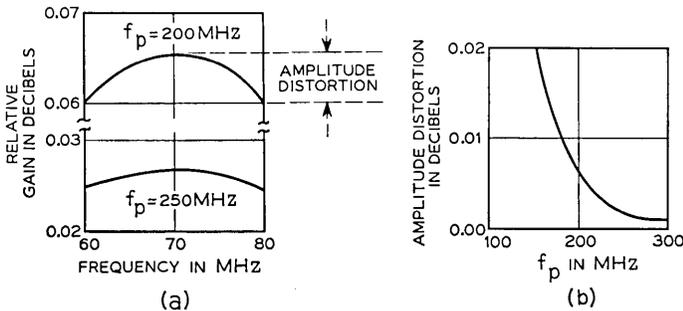


Fig. 3 — Amplitude distortion for a single stage.

specifically for this application, has the transmission line design described by C. L. Ruthroff.<sup>7</sup> The resulting peak frequency obtained for the circuit, including the devices and the parasitic inductance and capacitance of the external circuitry, is typically 240 MHz.

The automatic gain control is accomplished by variable loss stages located between gain stages. Figure 4 shows the schematic diagram and equivalent circuit of the variolossor. It is desirable that the transmission shape and delay be invariant with loss. This can be accomplished by placing the diode at a constant resistance point in the circuit, and can be achieved if

$$R_1 = R_2 + R_E = (L_{IN}/C_{OB})^{\frac{1}{2}} \tag{4}$$

The transmission expression under these conditions becomes

$$I_o/I_{in} = \frac{R_D}{[R_D + (L_{IN}/C_{OB})^{\frac{1}{2}}]} \cdot \frac{1/L_{IN}C_{OB}}{[\omega + (1/L_{IN}C_{OB})^{\frac{1}{2}}]^2} \tag{5}$$

where

$C_{OB}$  is the transistor output capacitance

$L_{IN}$  is the emitter input inductance

$R_E$  is the emitter input resistance.

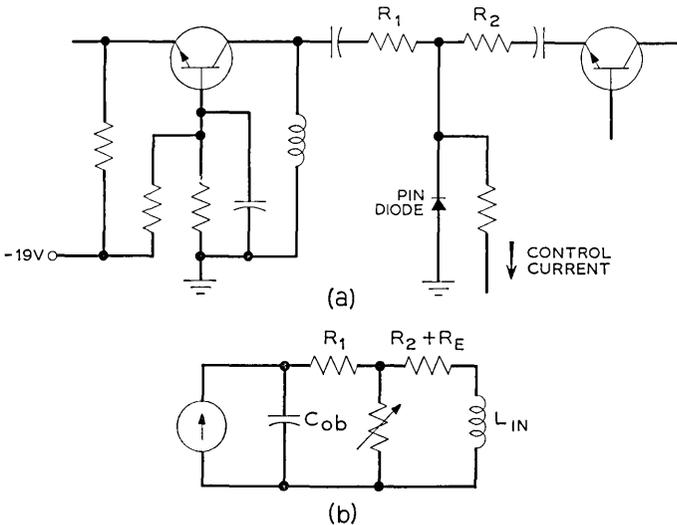


Fig. 4 — Schematic and equivalent circuit of variolossor.

This equation represents a low-pass filter with a 6 dB cutoff frequency of

$$\frac{1}{2\pi(L_{IN}C_{OB})^{\frac{1}{2}}}$$

For 45B transistors used in the amplifier, this frequency is approximately 1400 MHz, and the frequency response of the variolosses is, therefore, virtually flat over the 60 to 80 MHz band. The variable resistance element of the variolosses is a 474A PIN diode.<sup>6</sup> Above the frequency at which the diffusion capacitances bypass the junction resistances of the diode, the resistance of the diode is dependent on the resistance of the intrinsic region alone. The resistance of the intrinsic layer depends on the mechanism of conductivity modulation. And to permit the mechanism to operate, the intrinsic layer must be relatively wide, and the silicon must be sufficiently pure to result in a long lifetime and long transmit time for the minority carriers. Under these conditions the diode cannot follow the instantaneous variations of the high frequency signal; it acts like a linear resistance whose value is controlled by the direct current.

The locations of the variolosses in the complete amplifier circuit are critical because they affect the noise figure and the linearity of the amplifier. Because of noise, the signal power at every point in the amplifier should be greater than the signal power at the amplifier input. From linearity considerations it follows that the output of the amplifier should be the highest signal power point. These considerations, and the AGC range to be provided, determine the locations and the number of variolosses stages. The relative positions adopted for the gain and variolosses stages is shown in Fig. 5 along with a level diagram for conditions of normal and maximum gain. Each gain-variolosses stage combination has a gain ranging from less than -1 dB to 5 dB, depending on the control current supplied by the AGC unit. Thus at least 45 dB total gain range is provided by the seven stages to accommodate 35 dB down-fades and up-fades of at least 10 dB.

There are two transistor stages following the last variolosses with the output of Q17 matched to a 75-ohm impedance. These stages contribute a gain of about 4 dB. The monitor stage, Q18, is similar to the gain stages and isolates the main amplifier from the AGC amplifier.

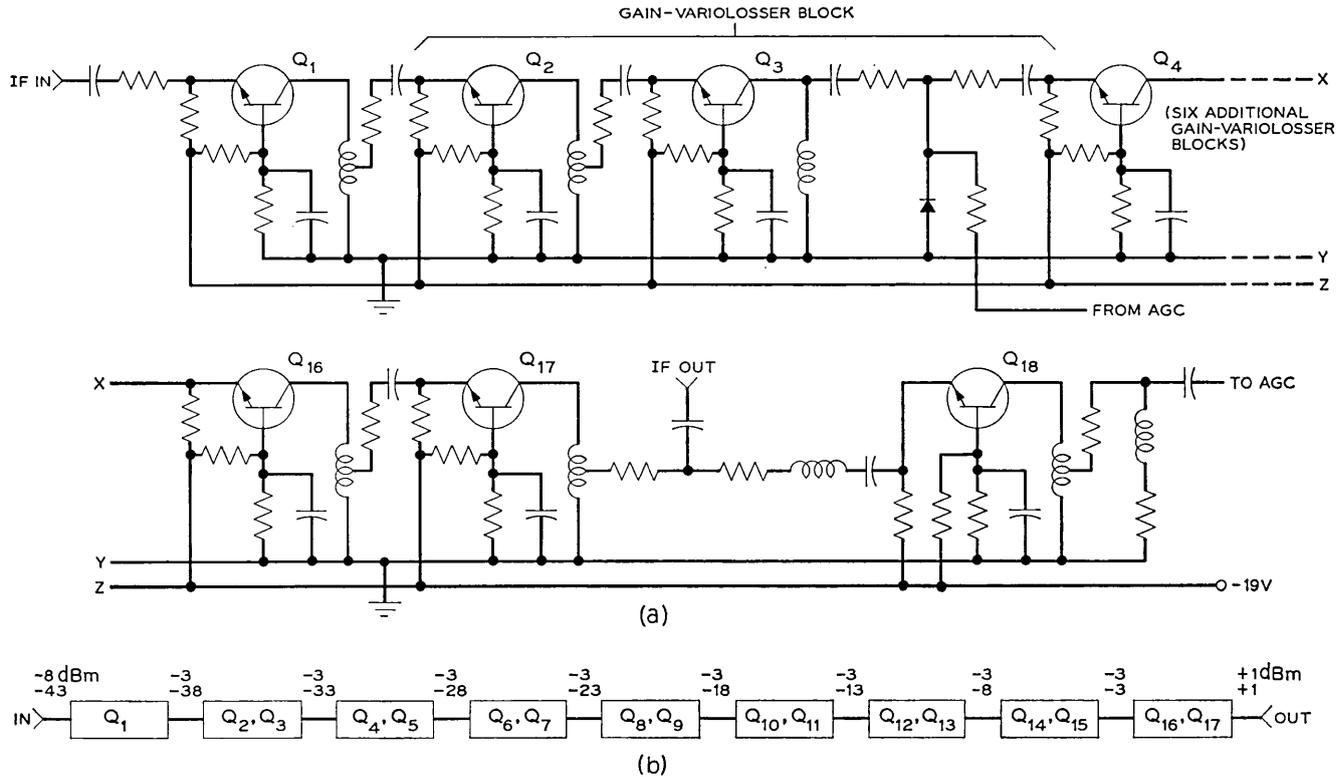


Fig. 5 — IF main amplifier schematic (a) and level diagram (b).

Figure 6 shows the transmission characteristic of the amplifier under normal and maximum gain. The increasing gain at low frequencies for the normal gain results from the PIN diode. At low frequencies the junction capacitance does not completely bypass the junction resistance with the result that the effective diode resistance

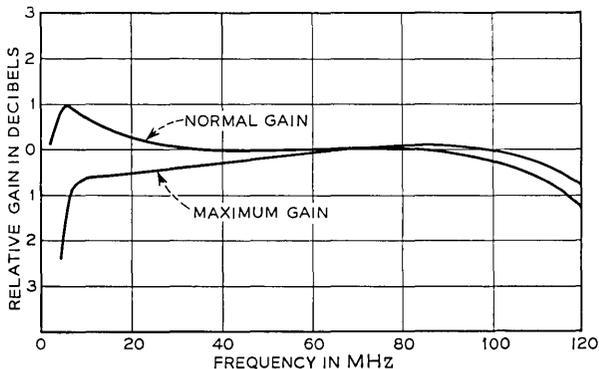


Fig. 6 — Transmission characteristic of IF main amplifier.

increases at low frequencies. Figure 7 shows the noise figure as a function of gain. At the normal gain of 9 dB, the noise figure of 14 dB will contribute about 0.2 dB to the over-all receiver noise. Figures 8 and 9 show the amplitude to phase conversion\* and harmonic distortion for the amplifier as functions of gain.

The bias for each transistor stage is set at 15 mA and 8 volts, resulting in a power dissipation of 120 milliwatts per transistor. The complete amplifier draws 300 mA at 19 volts resulting in a total power consumption of approximately 6 watts. The 15 mA bias current is the center of the transistor operating current range with respect to  $\alpha$  and  $f_t$ .

\* AM-PM conversion measurements reported in this paper were made using a test instrument developed by L. J. Sisti of Bell Telephone Laboratories, and J. M. Hancsarik, formerly of BTL. The set makes a direct dynamic measurement that has proven to be particularly useful in measuring small AM-PM coefficients. The set uses an amplitude modulator that is substantially free of incidental FM, and a balanced detector that is substantially insensitive to AM. AM-PM measurements can be made using envelope frequencies of 100 kHz, 1 MHz, or 6 MHz. The 100 kHz envelope frequency was used most frequently for which the resolution is 0.01° per dB and the absolute accuracy better than 0.1° per dB.

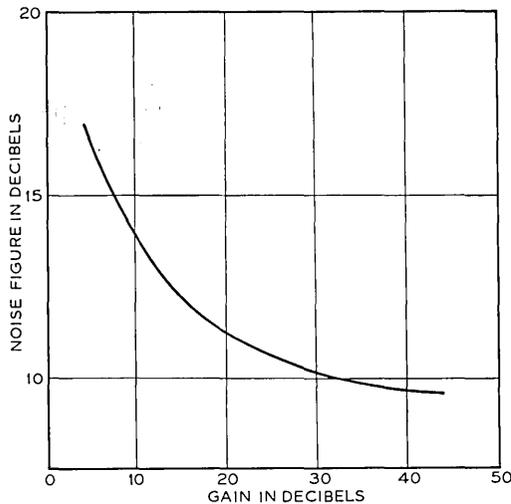


Fig. 7 — Noise figure vs gain of the IF main amplifier.

Figure 10 is a photograph of the amplifier housed in a  $2 \times 4 \times 24$  inch aluminum casting. The printed circuit layout was designed to minimize parasitic reactances in order to obtain the highest possible peak frequency for the basic gain stage.

#### IV. AGC CIRCUIT

The AGC circuit amplifies and rectifies a portion of the IF output of the main amplifier.\* The rectified signal is compared with a reference voltage in a differential amplifier. The net error signal changes current flow through the amplifier variolossor diodes in a direction to minimize the error signal. The AGC unit also provides a current that is directly proportional to received microwave signal strength.

Figure 11 gives block and schematic diagrams of the AGC amplifier. The input common base stage provides impedance matching to the 75-ohm cable and isolation between the AGC filter and the IF main amplifier. The filter prevents interfering tones from affecting the performance of the amplifier but is not sharp enough to disturb sweep delay measurements over the 64–76 MHz range made with the main amplifier operating on AGC.

\* The AGC circuit was developed by R. D. Thomas.

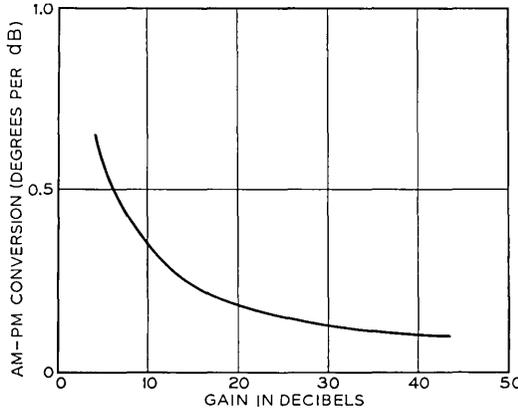


Fig. 8 — AM-PM conversion vs gain of IF main amplifier.

The filter characteristic is shown in Fig. 12. The output of Q2 is detected by CR1 and applied to the input of a three-stage dc amplifier. The output of the dc amplifier is the control current for the variolossor diodes which completes the feedback loop. The loop gain is such that the output power of the main amplifier decreases 0.2 dB for a 35 dB change in input power. The value of capacitor *C* is selected to provide a sufficient gain and phase margin and yet allow the AGC circuit to follow fade rates up to 100 dB per second. Since the AGC circuit is fast enough to follow a 30 Hz sweep, the circuit may be disabled so that swept transmission measurements may be made. Swept delay measurements are unaffected, however, because small variations in amplitude do not affect the response of the phase

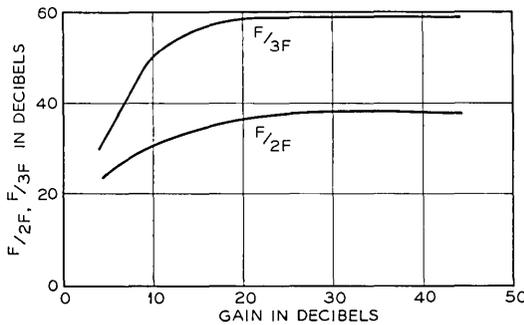


Fig. 9 — Harmonic distortion vs gain of IF main amplifier.

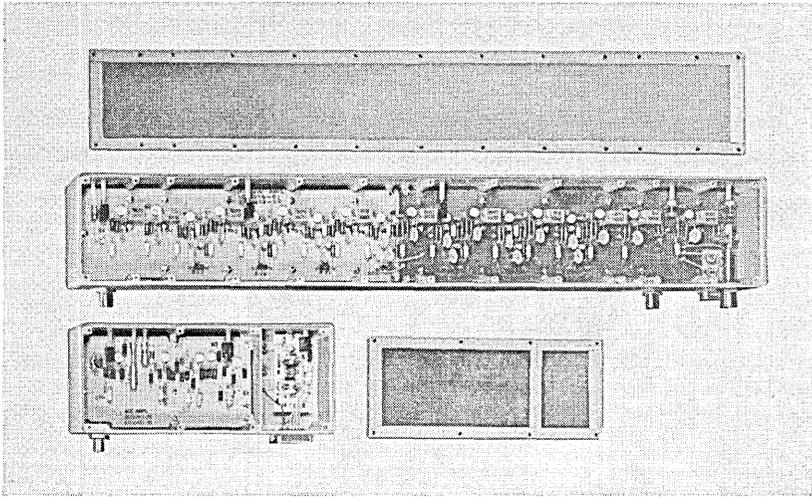


Fig. 10 — IF main amplifier and AGC unit.

detector of the delay measuring set. The operating point of the differential amplifier is set by adjusting the base voltage of Q4. This adjustment is used to set the output power of the main amplifier at +1 dBm. The metering network at the output of the dc amplifier is used to indicate received signal level. Figure 13 shows the relationship between the reading of the TD-3 panel meter and repeater input power.

Temperature compensation is built into the AGC circuit so that the output of the main amplifier does not change by more than 0.5 dB between 40 and 140°F. Figure 10 also includes a photograph of the AGC unit housed in an aluminum casting 2 × 4 × 9 inches.

#### V. IF PREAMPLIFIER

The noise figure of the entire receiver is very sensitive to the noise figure of the IF preamplifier because its input is at the lowest signal power point in the receiver. Shielding the preamplifier against extraneous interference is also important. Obtaining a low noise figure consistent with adequate load handling capability and flat transmission and delay characteristics was emphasized in the design. The receiver modulator is described elsewhere.<sup>8</sup>

A low noise amplifier requires a low noise transistor for the first stage and the stage must have sufficient gain to minimize the noise

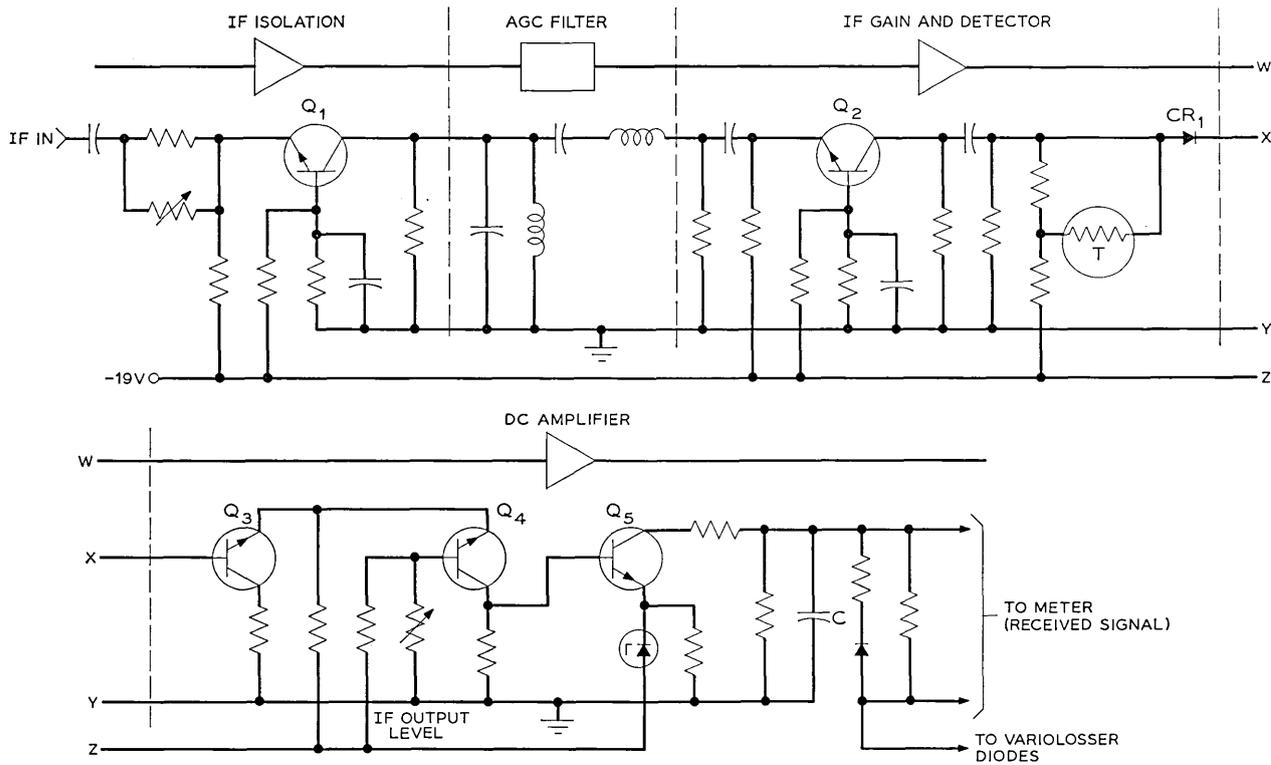


Fig. 11 — Block diagram and schematic of AGC amplifier.

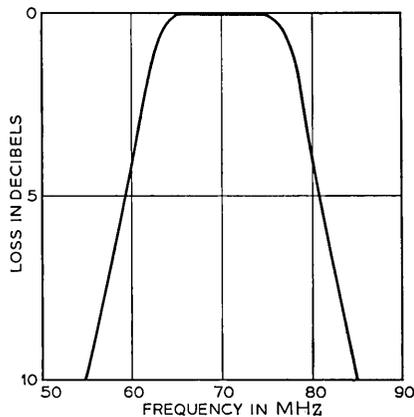


Fig. 12 — AGC filter characteristic.

contributions of succeeding stages. The common emitter configuration is most effective for this, but for a transmission band centered on 70 MHz, this configuration results in a transmission characteristic which falls off at 6 dB per octave owing to transistor  $\beta$  cutoff. To compensate, a second stage can be used which is also a common emitter but which includes feedback from collector to base to provide a transmission characteristic that rises 6 dB per octave over the desired band. This arrangement is used and the transmission characteristic of the combination is virtually flat.

The input transistor for the preamplifier is a Western Electric type 45J.<sup>6</sup> The transistor noise figure is 2.5 dB or less when measured with a 50-ohm source impedance and a collector current of 12 mA. With an  $f_T$  between 800 and 1200 MHz, it is possible to realize a gain of about 17 dB at 70 MHz for the first stage. Since the nominal source impedance (receiver modulator) is 50 ohms, no impedance correcting network is used at the amplifier input. While an imperfect impedance match at this point does introduce transmission shape in some cases, sufficient range of transmission adjustment is provided elsewhere in the preamplifier to achieve a flat over-all characteristic.

The second stage transistor is a Western Electric type 45G biased at about 38 mA and 4.5 volts. Operated in the common emitter configuration with the shunt feedback network shown in Fig. 14, the transmission characteristic is controlled by adjusting the load impedance  $R_L$ , the potentiometer designated SHAPE. In this manner the

transmission characteristic of the first two stages combined can be adjusted as shown in Fig. 15.

The three remaining stages are transformer-coupled common base circuits as described in Section III. The damping resistor associated with Q4 is a potentiometer for further controlling the transmission characteristic. In normal operation the preamplifier SLOPE and SHAPE controls are adjusted to obtain a flat characteristic from the receiver modulator input to the preamplifier output. A variable loss T network is located between Q4 and Q5 to permit the preamplifier output power to be set to 0 dBm for at least an 8 dB range of receiver RF input power. The output of the preamplifier is matched to a 75-ohm line with a network which provides more than 35 dB return loss over the 60–80 MHz band. The preamplifier also includes a control to permit bias adjustment for the Schottky-barrier modulator diode.

Since unequalized microwave networks with considerable amplitude and delay distortion precede the preamplifier, the applied signal has a significant index of amplitude modulation. Hence, the AM-PM conversion characteristic of the preamplifier is important. For a nominal received signal of  $-28.5$  dBm, the AM-PM coefficient of the preamplifier is less than  $0.02^\circ$  per dB. For a received signal of  $-20$  dBm a coefficient of about  $0.25^\circ$  per dB may be expected. This is considered satisfactory because cross-modulation noise from AM-PM conversion for a single repeater would be only about 5 dBmCO, and because several hops are not likely to up-fade simultaneously. Typical

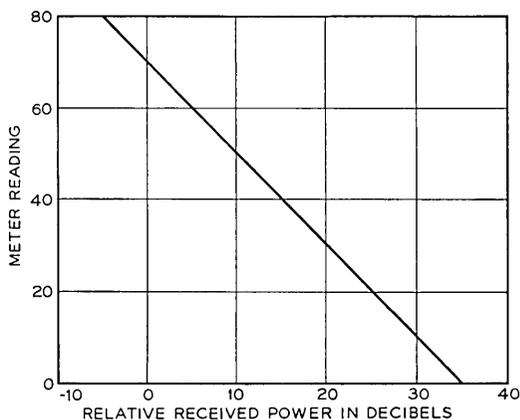


Fig. 13 — Received signal level meter reading vs relative received signal power.

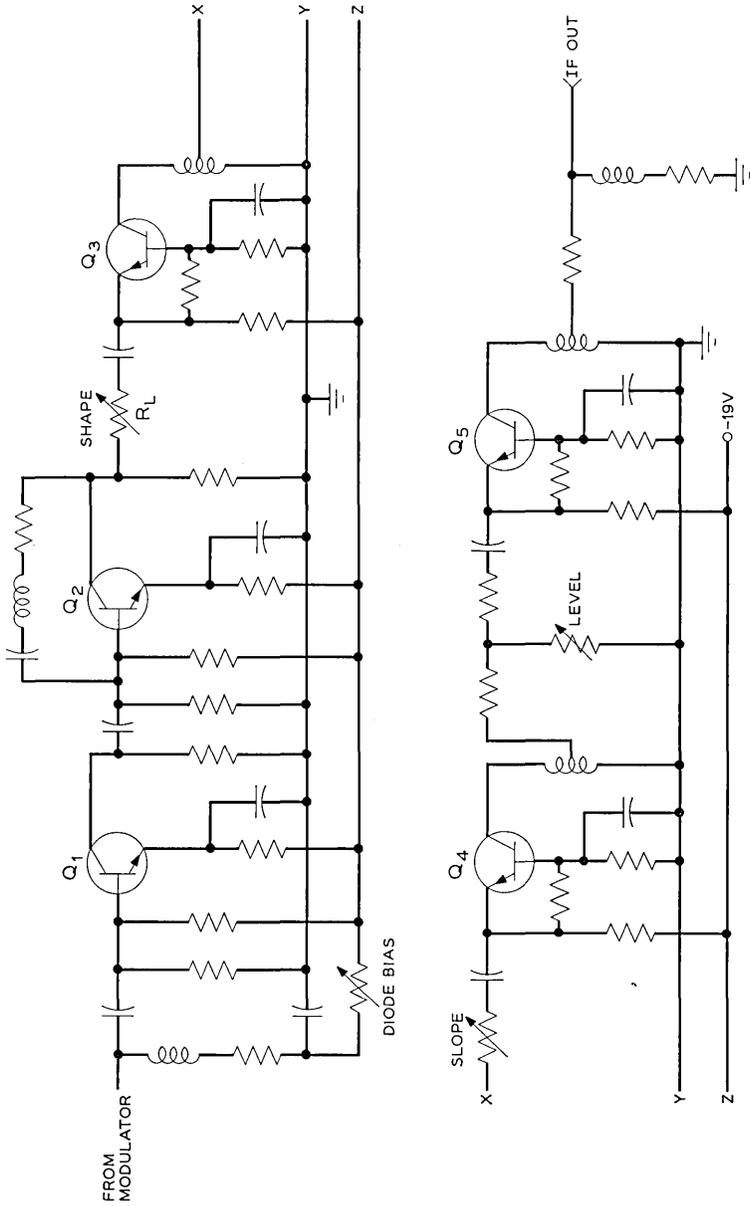


Fig. 14 — Preamplifier schematic.

compression and harmonic output characteristics versus input signal are shown in Fig. 16.

The over-all noise figure of the preamplifier is essentially the noise figure of the first stage plus about 0.5 dB, and for most transistors this total falls between 2.5 and 3.0 dB. While some improvement in noise figure may be obtained by reducing the collector currents in Q1 and Q2, the load handling ability of the amplifier is reduced thereby resulting in higher cross-modulation noise, particularly during up-fades.

## VI. THE IF LIMITER

The first circuit in the microwave transmitter is an IF limiter. It prevents the accumulation of AM resulting from residual transmission distortions. This is particularly important at main stations where considerable switching equipment and cabling is located between a microwave receiver and a microwave transmitter. The limiter also increases the effective AGC range of the repeater, and provides an output for the carrier resupply unit. The nominal input and output signals of the limiter are  $-7$  dBm.

The limiter consists of an input amplifier, the limiter proper, and an output amplifier. Both amplifiers use the common base, transformer-coupled stages described in Section III and shown in Fig. 17. The input amplifier provides a current gain of about 15 dB to the limiter section. The input signal for the carrier resupply unit is also derived from this amplifier.

The limiter section of the unit is shown in some detail in Fig. 17. It is a series clipper limiter, a design well suited to the low impedance

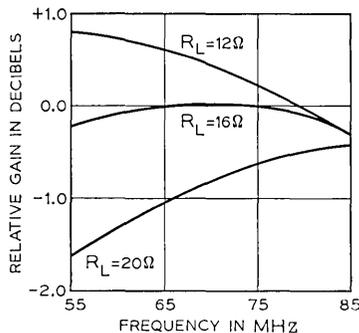


Fig. 15—Transmission characteristic of first two stages of IF preamplifier.

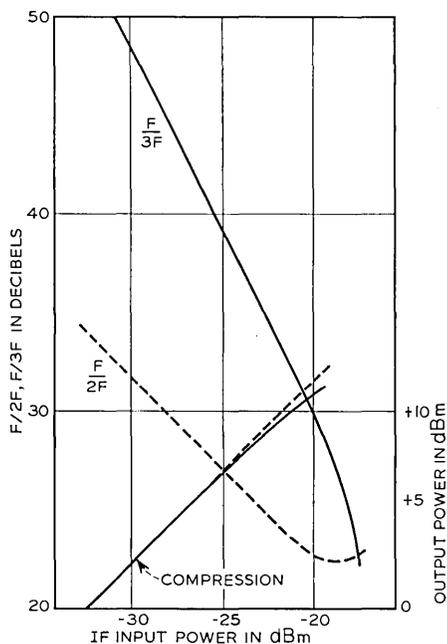


Fig. 16 — Compression and harmonic distortion characteristics of the preamplifier.

environment of transistor circuitry. The diodes CR1 and CR2 are 479A epitaxial silicon Schottky barrier diodes.<sup>6</sup> The diodes feature low capacitance, fast reverse recovery time, and a high back-to-forward resistance ratio.

The function of a limiter in an FM system is to reduce the index of amplitude modulation of the input signal and to accomplish this in such a way as to introduce a minimum of amplitude to phase conversion in the process. The objectives for this design were to suppress AM modulation by at least 30 dB and to incur AM to PM conversion of less than 0.2° per dB.

Laboratory experience, supported by analog computer simulation work, has shown that the performance of a limiter is critically dependent upon the presence of capacitance across the diodes and upon the source and load impedances. The over-all performance of the limiter was optimized experimentally. Although some improvement in AM-PM conversion for high envelope frequencies could be obtained by building out the source impedance to a constant value, this was not deemed necessary in this application.

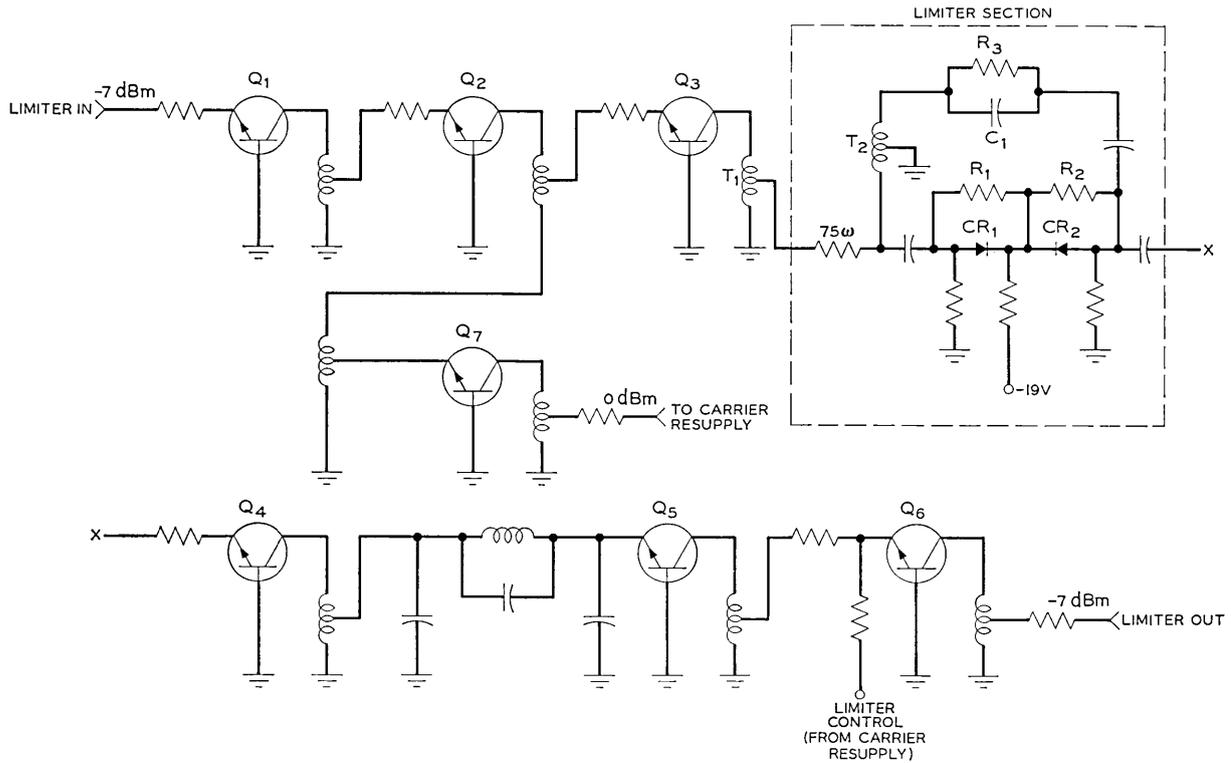


Fig. 17 — Limiter circuit—simplifier schematic.

The AM-PM contribution from the shunt capacitance of the diodes is balanced by the compensating circuit consisting of T2, R3, and C1 shown in Fig. 17. Transformer T2 is used to derive a signal having the same amplitude but of opposite phase from the signal applied to the diodes. R3 and C1 were selected experimentally to minimize AM-PM conversion. Resistors R1 and R2, low in value compared to the reverse resistances of the diodes, were added to make the operation of the limiter less dependent upon the reverse characteristics of the diodes.

To truly optimize the limiter, R3 and C1 should be made adjustable. This was not found to be necessary. The element values needed to achieve balance were worked out for average limiter diodes, and the capacitance of the diodes are held to such relatively close limits that the AM-PM is acceptable for all diodes. Circuit parasitics are held sufficiently close by virtue of the printed wiring circuit layout.

The efficiency of the limiter balancing technique is illustrated in Fig. 18 which shows both AM-PM conversion and AM suppression

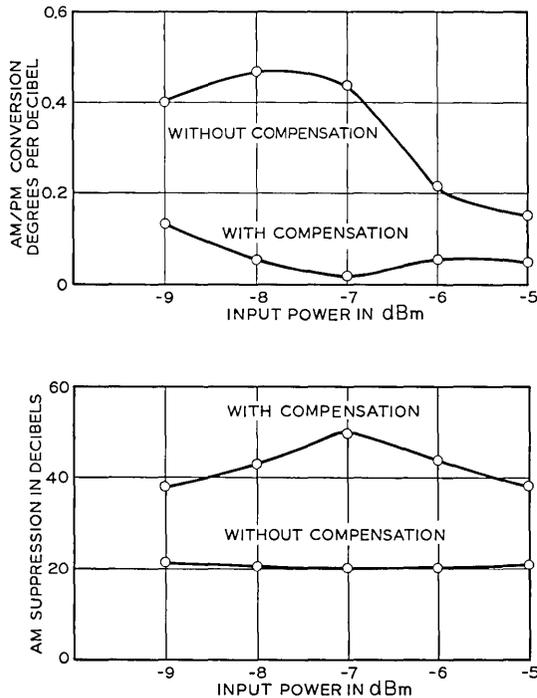


Fig. 18 — Limiter AM-PM conversion and AM suppression vs input power.  $f_c = 70$  MHz,  $f_m = 100$  kHz.

versus the limiter unit input power. The compensation makes a substantial improvement in both characteristics. Figure 19 presents a curve of AM suppression versus envelope frequency for a  $-7$  dBm input to the limiter unit, while Fig. 20 shows a compression characteristic.

The output amplifier consists of three additional common base transformer coupled stages, and includes a bandpass filter to attenuate harmonics, particularly the third harmonic, generated by the limiting process. Controls permit adjustment of the transmission characteristic

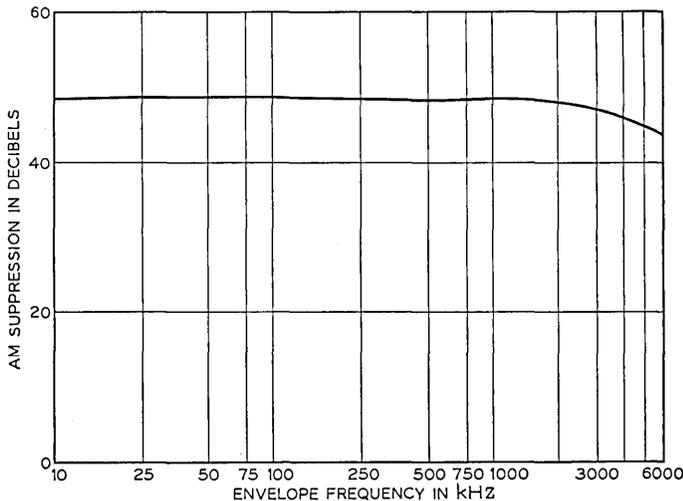


Fig. 19—Limiter AM suppression vs frequency.  $f_c = 70$  MHz, input =  $-7$  dBm.

and the output power. Operation of the carrier resupply unit disables the limiter output amplifier by cutting off transistor Q6. Thus, when the substitute carrier is applied to the driver amplifier, the high noise from the failed or faded receiver is blocked.

The limiter unit is housed in a die cast box about  $2 \times 4 \times 12$  inches. Figure 21 shows the unit with one cover removed.

## VII. CARRIER RESUPPLY

The carrier resupply circuit applies a 70 MHz frequency modulated signal to the transmitter when the normal signal fails, thereby preventing successive hops from going to full noise.\* The substitute

\* The carrier resupply unit was developed by B. F. Matas.

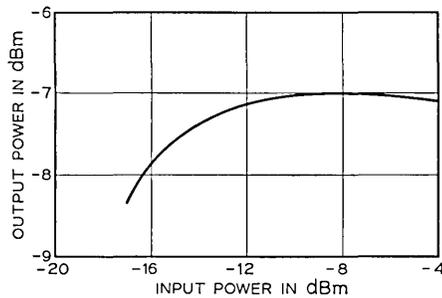


Fig. 20 — Limiter compression vs input power.  $f_c = 70$  MHz.

signal is modulated to insure that the 100A protection switching system will recognize this condition as a channel failure.

The desired output of the carrier resupply unit is an FM signal with a modulating frequency of 7 MHz or 9 MHz. The FM signal is produced by generating a 70 MHz signal and a 61 MHz signal (or 63 MHz for the protection channels). These signals are combined and applied to a limiter as shown by Fig. 22 for conversion to an FM signal.

The IF carrier nominally supplied to the microwave transmitter is monitored at the limiter as shown in Fig. 17. This monitored signal is amplified and rectified by the amplifier-detector in the resupply unit, and the resulting dc signal operates a Schmidt trigger circuit to control the state of the logic circuits. These circuits control the diode gate which determines whether the limiter output or the carrier resupply output is used to drive the microwave repeater.

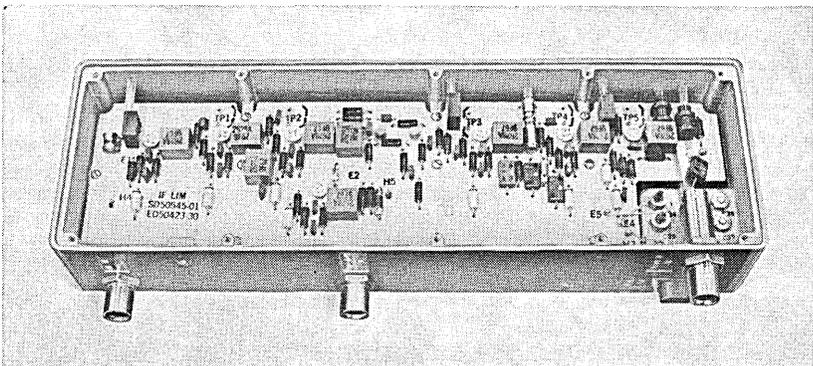
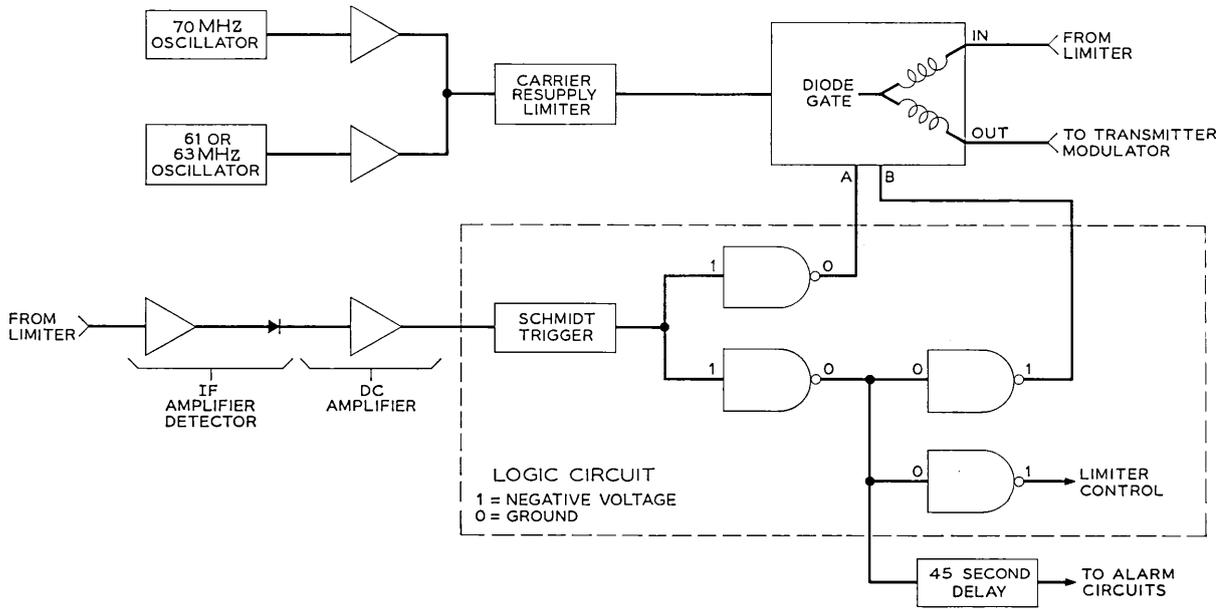


Fig. 21 — Limiter unit.



INTERMEDIATE FREQUENCY UNITS

Fig. 22 — Simplified block diagram IF carrier resupply.

When the 70 MHz carrier at the input to the microwave transmitter drops by 15 dB (corresponding to a 50 dB fade) the logic changes state with the following results:

- (i) The diode gate is turned on to place the resupply signal on the channel in less than 100 microseconds. (The restore time is less than the operate time.)
- (ii) The system limiter is biased to attenuate its output by more than 20 dB.
- (iii) A timing circuit is energized. This initiates an alarm after 45 seconds.

When the normal carrier reappears and reaches a level corresponding to a 48 dB fade, the Schmidt trigger restores and all functions are normalized. The operate and restore points of the resupply circuit are adjustable over wide ranges.

Figure 23 shows the 70 MHz carrier oscillator and amplifier. The oscillator is a Clapp common emitter circuit in which the collector is tuned to the fifth overtone of the crystal. Two stages of amplification are used to achieve +20 dBm of carrier power for the resupply limiter. A second oscillator uses a similar circuit but operates at either 61 or 63 MHz in another Clapp circuit. Since the sideband power requirement is less than the carrier power requirement, only a single gain stage is associated with this second oscillator and its gain is made adjustable to permit setting the deviation to the desired value. Both oscillators are temperature compensated to hold frequencies to within  $\pm 4$  kHz over a 30 to 140°F ambient temperature range.

As shown in Fig. 22, the outputs of both oscillators are combined and applied to a limiter very similar to the circuit described in Section VI. The output of the limiter is an FM signal. The deviation of the resupplied carrier is adjusted to accommodate the 100A switching system and is about 80 kHz peak for the 9 MHz modulated carrier and about 0.55 MHz peak for the 7 MHz modulated carrier.

Figure 24 shows the amplifier-detector-trigger circuit. The amplifier consists of two common emitter stages with a gain control between stages to set the trip point. A band-pass filter is located immediately ahead of the detector to introduce about 13 dB of loss  $\pm 10$  MHz from the carrier. This filter must be wide enough to prevent the resupply from operating when a channel is swept, but it must be narrow enough to prevent interfering tones which may be located approximately 10 MHz on either side of the carrier from interfering significantly with

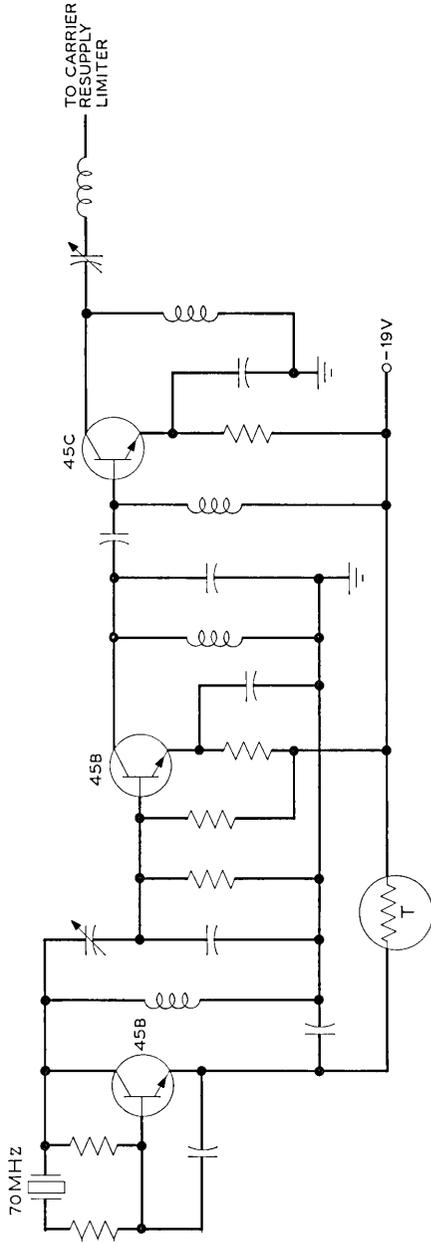


Fig. 23 — Crystal oscillator and amplifier.

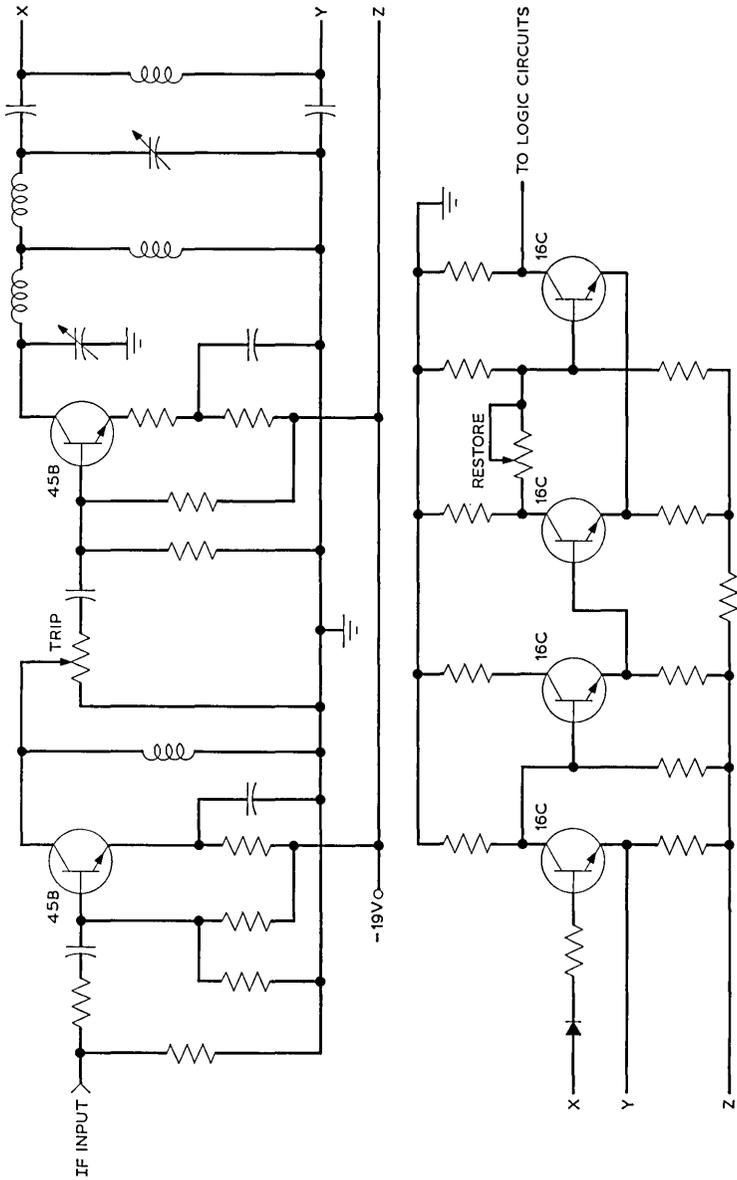


Fig. 24 — IF monitor and control circuits.

the desired operation of the resupply. The restore adjustment in the trigger circuit is set about 2 dB higher than the trip point to assure stable operation,

The Schmidt trigger operates logic circuitry involving six PNP transistors and three diodes to accomplish the functions just mentioned. One of these functions is to operate the diode gate, shown in Fig. 25. For normal operation of the system, the series diodes are reverse biased while the shunt diode is forward biased. In this state the gate introduces a loss of more than 90 dB to the resupply signal. The high loss is controlled by the high ratio of reverse to forward impedance of the 480A diodes and by extensive shielding. Fig. 25 also shows how the transmission path is completed from the system limiter to the transmitting modulator. The low-pass filter section associated with the gate presents return losses of better than 35 dB over the 60–80 MHz band.

Figure 26 shows the carrier resupply unit. The cast framework,  $19\frac{1}{4} \times 4\frac{1}{2} \times 2$  inches, is divided into eight compartments to isolate the circuits from each other. Both covers are completely gasketed, part of the extensive shielding required because the oscillators operate continuously. Meeting the 100 microseconds transfer requirement in ways other than continuous operation of the oscillators would have been more difficult and less reliable.

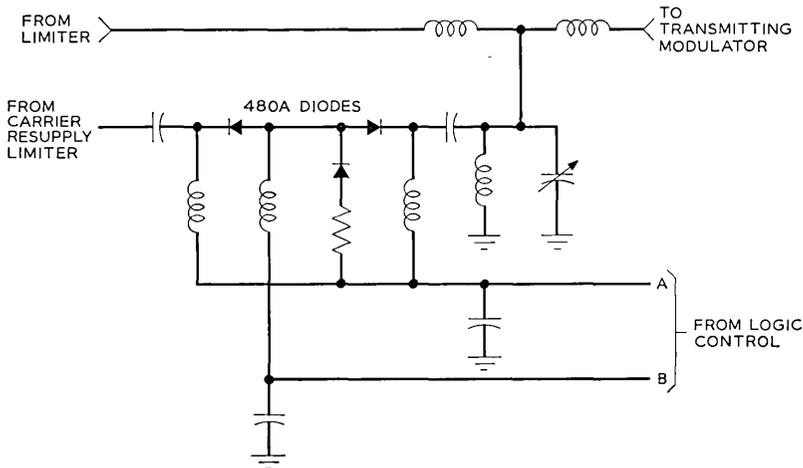


Fig. 25 — Gate circuit.

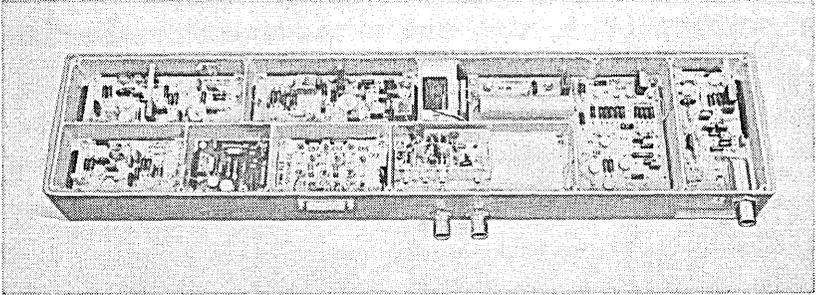


Fig. 26 — IF carrier resupply.

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# Schottky Barrier Receiver Modulator

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*This paper describes the receiver modulator of the TD-3 repeater. This modulator is a single-diode downconverter with an integral IF preamplifier. Its significant features are the use of a Schottky barrier diode, a waveguide directional filter, two lowpass filters for harmonic suppression, and image frequency absorption. The amplitude transmission characteristic of the modulator is flat to  $\pm 0.01$  dB for  $\pm 6$  MHz, and the average noise figure is 6.7 dB.*

## I. INTRODUCTION

The receiver modulator of the TD-3 repeater is the first unit in the signal path which uses active devices. Because it is in the signal path, its transmission characteristics must be well behaved, and because it is at the front end of the repeater, its low thermal noise is of considerable importance to the performance of the system.

## II. PERFORMANCE OBJECTIVES AND BASIC CONFIGURATION

### 2.1 Objectives

The objective of the development was to obtain a solid state downconverter-preamplifier for the microwave receiver of the TD-3 system. The unit had to be reliable, inexpensive, and in a shape to ease the transmitter-receiver bay designing. It should be easy to manufacture and maintain, and should have a minimum of adjustments. It should be designed to last at least 20 years and be rugged enough to withstand the shock and temperature extremes of shipment. In addition, it should conform to the following electrical specifications:

Signal midband frequency:  $f_{SI} = 3710, 3730, \dots 4190$  MHz, according to the frequency plan discussed in Ref. 1.

Local oscillator frequency:  $f_{LO} = 3780, 3800, \dots 4120$  MHz, according to the frequency plan discussed in Ref. 1.

Midband intermediate frequency:  $f_{IF} = 70$  MHz.

Signal power:  $P_{SI} = -28$  dBm nominally; for short hops, upfades and in anticipation of a possible increase of the transmitter output power, the design should be capable of operating satisfactorily with input signals as high as  $P_{SI} = -19$  dBm.

Local oscillator power:  $P_{LO} \leq 6$  dBm.

IF output power:  $P_{IF} = 0$  dBm.

Signal to IF transmission: *Amplitude*—as flat as practicable; if possible within  $\pm 0.005$  dB for  $f_{SI} \pm 6$  MHz,  $\pm 0.015$  dB for  $f_{SI} \pm 10$  MHz; *Delay*—less than 0.1 ns distortion for  $f_{SI} \pm 6$  MHz.

Noise figure: as low as possible, but less than 8 dB.

Signal input port: WR-229 waveguide.

Signal return loss:  $RL_{SI} \geq 30$  dB for  $f_{SI} \pm 10$  MHz.

Local oscillator input port: WR-229 waveguide.

Local oscillator return loss:  $RL_{LO} \geq 10$  dB for  $f_{LO}$ .

IF output port:  $75 \Omega$

IF output return loss:  $RL_{IF} \geq 35$  dB for  $f_{IF} \pm 10$  MHz.

Local oscillator suppression: local oscillator port (*i*) to signal port  $\geq 35$  dB; (*ii*) to IF output port  $\geq 50$  dB.

Local oscillator harmonics: The second and third harmonic of  $f_{LO}$  shall each be at least 40 dB below  $P_{LO}$  at the signal port and at the local oscillator port.

RF leakage: comparable to a good waveguide flange joint.

Temperature range:  $75^\circ\text{F} \pm 10^\circ\text{F}$  nominally, but operative from  $40^\circ$  to  $140^\circ\text{F}$ .

Bias voltage:  $-19$  V to ground.

Diode bias current monitoring meter:  $20\mu\text{A}$  ( $5900\Omega$ ).

All of these requirements were met by the final design. However, the amplitude transmission characteristic from signal to IF is slightly more sensitive to temperature within the  $75^\circ\text{F} \pm 10^\circ\text{F}$  temperature range than ideally desirable, but this is not expected to noticeably impair the performance of the system.

## 2.2 Basic Design Considerations

Four basic considerations and resulting decisions significantly influenced the design of the downconverter-preamplifier unit.

(*i*) A Schottky barrier diode was chosen because it appeared able to yield the required low noise figure.

(*ii*) A single diode was used, that is, an unbalanced downconverter was developed. The major advantages of such a converter are that

the cost of diodes for the converter is lowered by at least 50 per cent since a matched pair of diodes is not required; the diode can be biased externally, regardless of the polarity of the supply voltage, for example, only a negative voltage to ground is available in TD-3; and the connection to the IF preamplifier is simple and does not require a balanced transformer.

The noise performance of this downconverter should be comparable with its balanced counterpart, even though the unbalanced downconverter does not cancel the noise bands centered around  $f = 0$  generated by local oscillator noise beating with the local oscillator carrier. However, this lack of noise suppression is of no consequence for most radio systems because the local oscillator power generally passes through a narrow bandpass filter so that no significant noise energy is contained in frequency regions  $\pm 70$  MHz away from the local oscillator carrier. Appendix A gives a more detailed analysis of the noise performance of a balanced and an unbalanced downconverter using a simple model for the diode.

(iii) The power generated in the downconverter at the image frequency band is absorbed rather than reflected to the diode with a suitable phase shift. As is well known, reflecting the power would raise the efficiency of the conversion process and hence lower the noise figure. It was estimated from preliminary investigations that the penalty in noise figure for absorbing the image power is approximately 0.7 dB. This appeared to be a fair price for the advantages gained. Since in absorbing the image power, for example, in an isolator, no narrowband reactance has to be placed into the path of the signal, there is a better chance that the very stringent requirements pertaining to the amplitude and delay response of the downconverter can be met. Furthermore, no adjustment of an image reflecting reactive circuit is required.

(iv) Since difficulties often have been encountered from harmonics and harmonic sidebands of the local oscillator frequency originating in the diode, the diode was placed between two lowpass filters. This not only assures that the level of harmonics emanating from the downconverter is drastically reduced; it is very unlikely that the impedances of the connecting circuits at the harmonic frequencies will influence the performance of the downconverter appreciably. This is quite important, since these impedances are usually not under control.

These four basic decisions lead to the block diagram shown in Fig.

1. The signal power and the local oscillator power are combined in the adding network and pass to the diode through the isolator and the first lowpass filter. The cutoff frequency of this lowpass filter is at approximately  $1.5 \times$  (average local oscillator frequency) so that it passes the local oscillator frequency and the signal and image band of all channels without any appreciable attenuation, and that it stops all undesirable harmonics and harmonic sidebands of the local oscillator frequency present at the diode.

The isolator has broadband characteristics so that it passes the signal band and the local oscillator frequency in the forward direction and absorbs these frequencies as well as the image band in the reverse direction. It also makes the return loss at the signal and local oscillator port virtually independent of the diode properties.

The intermediate frequency generated in the diode passes through the second lowpass filter, is amplified in the preamplifier and appears at the IF output port. The dc-bias is derived from the  $-19$  V available in the preamplifier and is fed to the diode from the preamplifier through the second lowpass filter. The cutoff frequency of this lowpass filter is at approximately  $0.5 \times$  (average local oscillator frequency) so that it passes the dc-bias and the IF band without any appreciable attenuation, and that it attenuates all undesirable microwave frequencies present at the diode.

It was decided to realize the combination of the lowpass filter, the diode mount, and the second lowpass filter as a waveguide-coaxial structure. It appeared then that a very suitable realization for the adding network would be a waveguide directional filter of the type

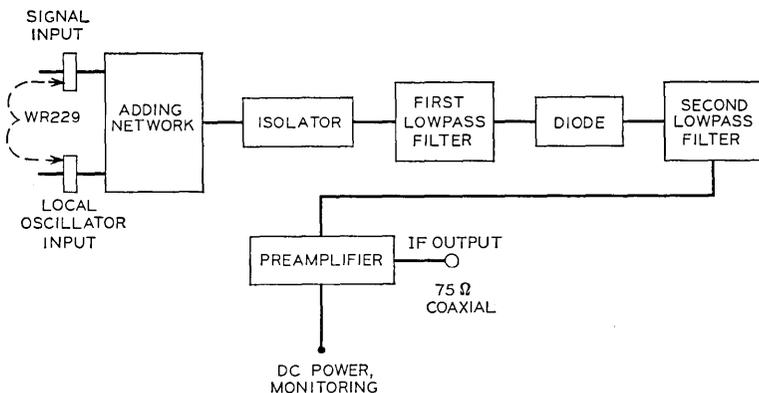


Fig. 1 — Basic downconverter-preamplifier block diagram.

described in Ref. 2. Since any filter is most sensitive to manufacturing deviations and environmental conditions around the frequency where its resonators resonate, it was decided to pass the local oscillator frequency through the bandpass section of the directional filter and the signal band through the bandstop section. Thus, all resonators of the directional filter resonate at the local oscillator frequency, and the response of the filter at the signal band is rather insensitive to manufacturing deviations and environmental conditions. The resulting block diagram is shown in Fig. 2.

Although the location of the isolator in Fig. 2 is ideal from the viewpoint of the downconverter design, it was necessary to move the isolator to the signal input port to avoid multiple reflections between the bandstop section of the directional filter and the channel bandpass filter which precedes the downconverter.<sup>1</sup> Therefore, the configuration shown in Fig. 3 was finally adopted.

A block not in Fig. 2 is the step transducer between the directional filter and the first lowpass filter. This transducer is necessary because the first lowpass filter requires reduced height waveguide as the connecting waveguide. This lowpass filter must be a multimode lowpass filter, that is, it must stop harmonics and harmonic sidebands of the local oscillator frequency generated in the diode irrespective of the mode in which they occur. The only known solution for this requirement is the "waffle iron filter,"<sup>3</sup> which must be operated between reduced height guides. No transducer is required between the first lowpass filter and the diode because initial estimates indicated that

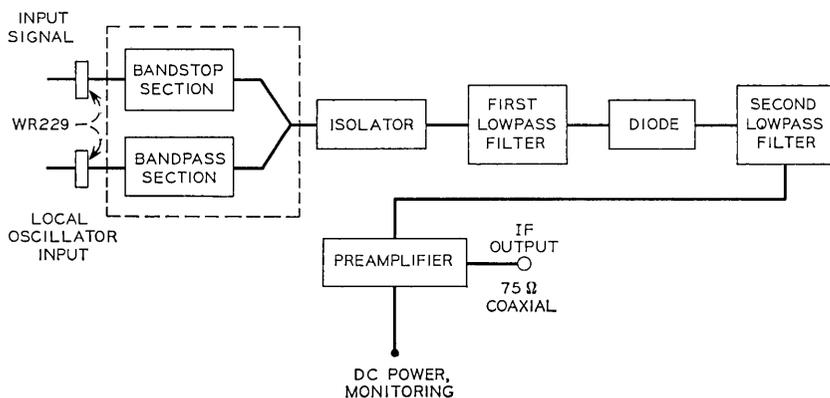


Fig. 2 — Detailed downconverter-preamplifier block diagram.

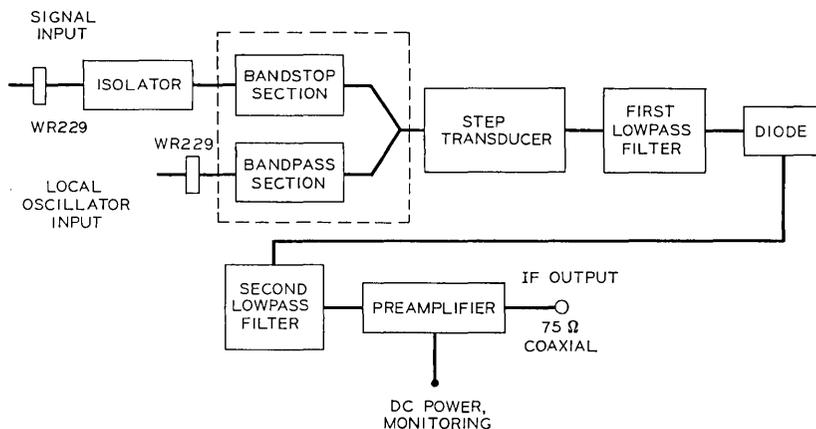


Fig. 3—Final downconverter-preamplifier block diagram.

the Schottky barrier diode could be operated quite well in a reduced height waveguide.

### III. DETAILED ELECTRICAL DESIGN

#### 3.1 Directional Filter

The structure, the properties and the design of the kind of waveguide directional filter used for this downconverter are described in Ref. 2. Therefore this discussion is confined to establishing the requirements, deciding the values of  $n$  (number of bandpass and band-stop cavities) and  $Q_T$  (selectivity factor) of the filter, and presenting the actual results obtained.

Fig. 4 is a sketch of the directional filter for  $n=2$ . As a result of the requirements discussed in Section 2, the following requirements were imposed on the directional filter:

Midband frequency:  $f_{L0}$ .

Transmission from port 3 to port 1: *Amplitude*—Flat to within  $\pm 0.005$  dB for the two 12 MHz bands centered at  $f_{L0} \pm 70$  MHz and flat to within  $\pm 0.015$  dB for the two 20 MHz bands centered at  $f_{L0} \pm 70$  MHz; *Delay*—Less than 0.1 ns distortion for the two 12 MHz bands centered at  $f_{L0} \pm 70$  MHz.

Insertion loss at  $f_{L0}$  from port 2 to port 3:  $\geq 5$  dB. (The isolator provides another 30 dB.)

Return loss at port 1:  $\geq 30$  dB for the two 20 MHz bands centered at  $f_{L0} \pm 70$  MHz.

Return loss at port 2:  $\geq 20$  dB at  $f_{L0}$ .

Return loss at port 3:  $\geq 30$  dB for the two 20 MHz bands centered at  $f_{L0} \pm 70$  MHz.

Ports: WR-229 waveguide.

Temperature range:  $75^\circ\text{F} \pm 10^\circ\text{F}$  but operative from  $40^\circ$  to  $140^\circ\text{F}$ .

It is obvious from these requirements that the  $Q_T$  of the filter should be made as high as possible in order to meet the specifications with the smallest possible  $n$ .<sup>2</sup> The limiting factor in this connection is the dissipation loss in the filter, which reduces the insertion loss from port 2 to port 3 and raises the insertion loss from port 2 to port 1, as  $Q_T$  is increased for a given  $n$ . For this reason  $n = 1$  could not be used since it would require too high a  $Q_T$ . Consequently:

$$n = 2 \quad (1)$$

$$Q_T = 140 \quad (2)$$

were chosen since  $Q_T \approx 150$  is known to be a reasonable upper limit.

The results obtained from a typical filter designed to these specifications are as follows:

Return loss at ports 1 and 2 as Fig. 5 shows.

Return loss at port 3:  $\geq 40$  dB for the two 60 MHz bands centered at  $f_{L0} \pm 70$  MHz.

Insertion loss from port 2 to port 3 as Fig. 6 shows.

Transmission from port 3 to port 1: *Amplitude*—At  $f_{L0} \pm 70$  MHz, the insertion loss is 0.005 dB; for the two 20 MHz bands centered at  $f_{L0} \pm 70$  MHz, the insertion loss is flat to within less than  $\pm 0.01$  dB;

*Delay*—It can be computed that the distortion will be a slope of 0.084 ns per 12 MHz for the two 12 MHz bands centered at  $f_{L0} \pm 70$  MHz.

The insertion loss from port 2 to port 1 at  $f_{L0}$  was measured to be typically 0.3 dB. In contrast with the other microwave filters of the TD-3 system,<sup>4</sup> it was decided to fabricate this filter out of copper instead

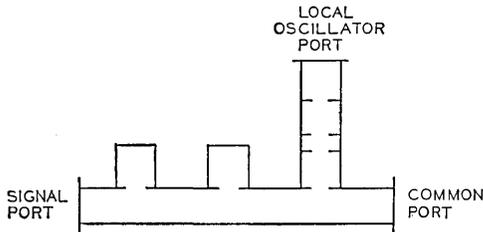


Fig. 4 — Directional filter. (Common: port 1, local oscillator: port 2, signal: port 3.)

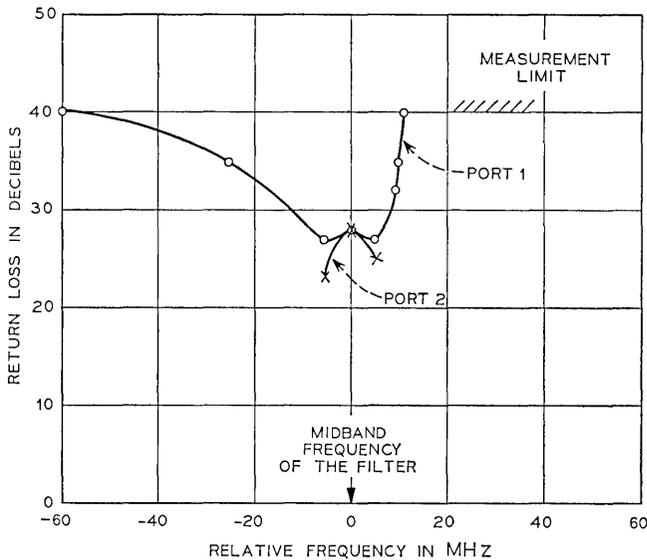


Fig. 5 — Return loss of ports 1 and 2 of the directional filter.

of copper-clad Invar, since it is apparent from the requirements that the frequency shift caused by changes in temperature and relative humidity can be tolerated.

### 3.2 Converter Block

The following general approach was used in designing the converter block consisting of the step transducer, the first and second lowpass filter, and the diode section:

(i) The dimensions of the converter block were experimentally determined in such a way that a *broadband* match was obtained at the microwave input port to the WR-229 waveguide (Fig. 7). This broadband match was obtained using local oscillator power levels of  $P_{LO} = 0, 3, \text{ and } 6$  dBm (maximum level available) and reasonable forward diode bias currents from 4 to 12 mA. No mechanical adjustment was required for any fixed  $P_{LO}$  to obtain this match for all diodes and around all local oscillator frequencies. However, some adjustment of the bias current for different diodes and different local oscillator frequencies was required. It was assumed that the converter block designed by this procedure will behave approximately like a

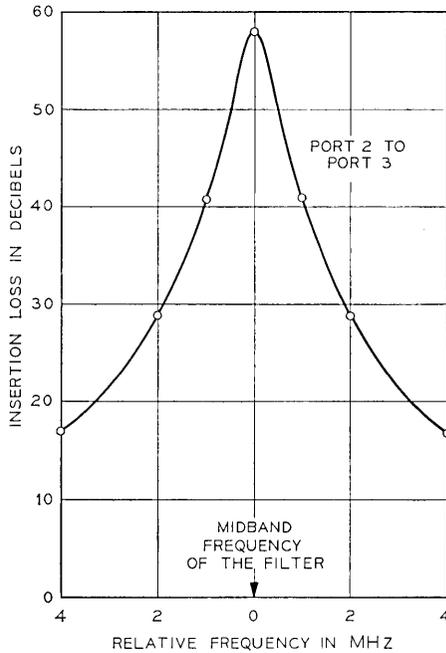


Fig. 6 — Insertion loss between ports 2 and 3 of the directional filter.

nonlinear conductance (that is, real admittance), when operated as a downconverter with the pertinent  $f_{LO}$  and at an intermediate frequency of 70 MHz, for two reasons: (1) The broadband match indicates that the diode and diode mounting reactances have been successfully tuned out around the pertinent  $f_{LO}$ . (2) These reactances do not play any significant role at frequencies around 70 MHz and, similarly, the second lowpass filter does not provide any additional reactances because its cutoff frequency is well above 70 MHz, and its electrical length is short compared with the wavelength.

(ii) As a consequence of this, it can be expected that the converter block, after being adjusted with the bias current for an optimum

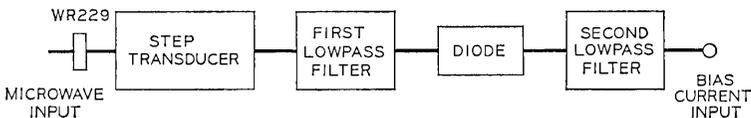


Fig. 7 — Block diagram of the converter block.

match at the pertinent  $f_{LO}$  and when connected to form a downconverter as shown in Fig. 8, exhibits an IF output impedance  $Z_{IFout}$ , which is approximately real and constant over the band from 60 to 80 MHz.

(iii) For the same reason, this downconverter, when operated as shown in Fig. 9, can be expected to exhibit a signal to IF transmission characteristic with virtually no amplitude or delay distortion for the two 20 MHz bands centered around  $f_{LO} \pm 70$  MHz and to yield a conversion loss, defined as

$$CL/dB = 10 \lg P_{SI}/P_{IF}, \tag{3}$$

between 3 dB and 5 dB.<sup>5</sup>  $P_{SI}$  is the incident signal power at the signal input port, and  $P_{IF}$  is the IF power absorbed by the IF load  $R_{IF}$  in Fig. 9. This load shall be a constant real resistance approximately equal to  $Z_{IFout}$  in order to extract the available IF power from the converter.

(iv) Assuming that the input impedance of the IF preamplifier is made approximately equal to  $R_{IF}$  and that the noise figure in dB of the IF preamplifier when measured from  $R_{IF}$  is equal to  $NF_{IF}$ , the total single sideband noise figure  $NF_{Total}$  in dB of the complete downconverter-preamplifier unit (Fig. 3) referred to a signal generator matched to the WR-229 waveguide can then be estimated to be:<sup>5</sup>

$$NF_{Total} = CL + NF_{IF} + 10 \lg \left( 1 + \frac{1 - \frac{1}{2}10^{CL/10 \text{ dB}}}{10^{(CL+NF_{IF})/10 \text{ dB}}} \right) \leq CL + NF_{IF} \tag{4}$$

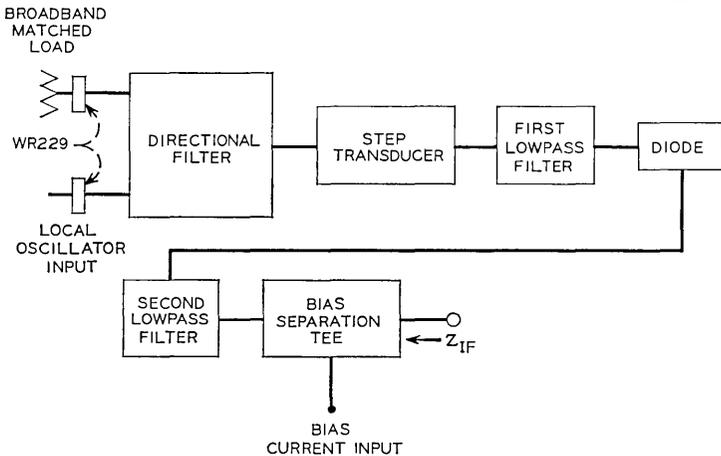


Fig. 8 — IF output impedance of the downconverter.

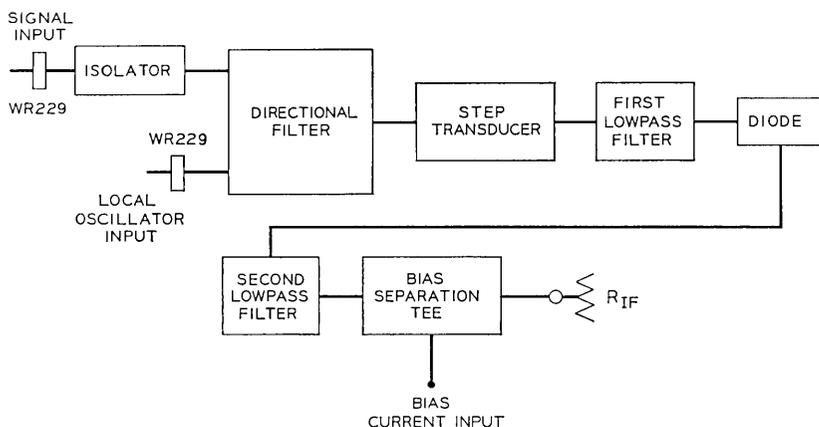


Fig. 9 — Block diagram of the downconverter.

Since  $NF_{IF}$  is about 3 dB, the performance objective of  $NF_{Total} < 8$  dB should be attainable.

The physical configuration chosen for the converter block is shown in cross section in Fig. 10.

The step transducer is a conventional<sup>6</sup> quarter wavelength transducer connecting from a height of 1.145 to 0.100 inch at a constant width of 2.29 in. It has three steps and provides a return loss of better than 22 dB from 3700 to 4200 MHz, which is entirely satisfactory for this application. The height of 0.100 in was chosen, since it is a convenient connecting height for the first lowpass filter, and was estimated to be a suitable height to obtain the desired broadband match of the diode to the local oscillator power.

The first lowpass filter is designed as a waffle iron filter, as explained in Section 2.2, based on a modification of a published design.<sup>7</sup> The number of sections was reduced to four and the design scaled according to the ratio of the waveguide widths. Judging from the published data, this should result in an insertion loss of at least 27 dB for a  $TE_{10}$  mode from 6 to 16 GHz. The return loss from 3700 to 4200 MHz was measured between  $2.29 \times 0.1$ -inch rectangular waveguides to be better than 14 dB, which is satisfactory in this application.

No attempt was made to measure the multimode stopband insertion loss because of the formidable problems of such a measurement. It was established later, however, by measurements of the level of

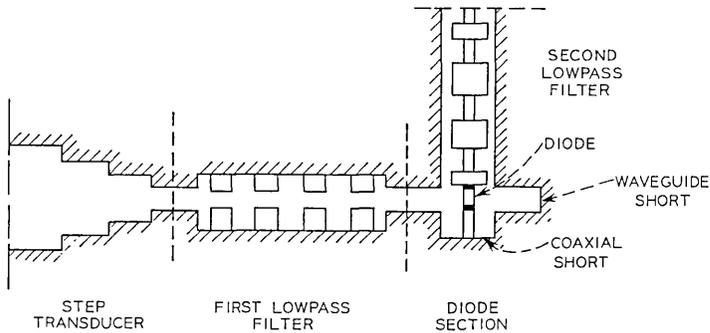


Fig. 10 — Cross section of the converter block.

harmonics of the local oscillator frequency emanating from the converter, that the rejection at these frequencies is adequate. Section 5.7 gives the results.

The second lowpass filter is a conventional stepped-impedance coaxial lowpass filter designed according to image parameter theory.<sup>8</sup> It has three sections, its outer conductor has a 0.288 inch diameter, and it starts with a low impedance line. The image cutoff frequency is 2460 MHz, and the filter is designed to furnish better than 40 dB of image attenuation from 3.7 GHz to at least 8.4 GHz. The image impedance at  $f=0$  is  $40\Omega$ , resulting in virtually the same image impedance across the entire 60 to 80 MHz band. The value of  $40\Omega$  represents a compromise between manufacturing cost of the center conductor and electrical requirements. Since  $Z_{IFout}$  turned out to be approximately  $50\Omega$ , an image impedance of  $50\Omega$  at  $f=0$  would have been ideal. This, however, would have resulted in a rather thin center conductor in the high impedance lines of the filter. The electrical performance of the filter was not measured independently because the design procedure is sufficiently accurate and the filter in the down-converter performs satisfactorily (see Section 5.7).

The diode section is a  $2.29 \times 0.1$ -inch rectangular waveguide shorted at one end ("waveguide short" in Fig. 10). The diode is located in the center of the waveguide across the narrow dimension between the second lowpass filter and a shorted section ("coaxial short" in Fig. 10) of  $30\Omega$  coaxial line (which has an outer diameter of 0.288 inch).

The three available degrees of freedom—the distance between the diode and the waveguide short, the distance between the diode and

the first lowpass filter, and the length of the coaxial short—were determined empirically to yield the desired broadband match of the converter block to the local oscillator power. The dimensions arrived at are 0.250, 0.350, and 0.060 inch, respectively. These dimensions are valid for  $P_{LO} = 6$  dBm (maximum power available), which was found to yield the lowest value for  $CL$ . The diode used is the Western Electric Company 497A gallium arsenide Schottky barrier diode.<sup>9</sup> (The equipment described in this article is manufactured by the Western Electric Company for Bell System use only.)

### 3.3 Remaining Components

The remaining components shown in Fig. 3 are the isolator and the IF preamplifier. Ref. 10 describes the preamplifier. The Western Electric Company 8A isolator was chosen because it has adequate broadband characteristics so that a single code covers all 24 signal midband frequencies with their associated local oscillator frequencies and image bands.

## IV. MECHANICAL DESIGN

The downconverter was designed very carefully to achieve a satisfactory compromise between the electrical requirements of high performance and reliability and the economic demands of low cost and easy manufacture. Starting with the basic arrangement of Fig. 11, it was decided to integrate the step transducer, waffle iron filter, and diode cavity into a single die-cast structure; the bandpass and band-stop filters were combined into the directional filter; the mounting on one side of the diode was integrated with the coaxial lowpass filter

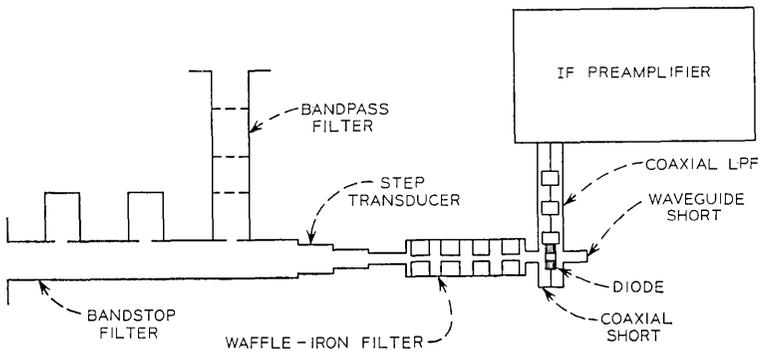


Fig. 11 — Basic arrangement of the downconverter-preamplifier assembly.

and on the other side with the coaxial short. It was recognized that particular effort would have to be devoted to the diode mounting in the final design. N-type connectors between the coaxial LPF and the IF preamplifier were helpful during early development but unnecessary in production.

The unit thus consists of six basic mechanical modules: directional filter, die-cast housing, collet assembly, diode, coaxial LPF, and IF preamplifier. Figure 12 shows the two halves of the die-cast housing, the collet assembly, the diode, and the coaxial LPF, and Fig. 13 shows a preproduction model of the entire unit. During final assembly, the modules are joined by screws, and two  $\frac{1}{4}$ -inch-long wire straps are added to connect the outer and inner conductor of the LPF to the IF preamplifier.

#### 4.1 *The Directional Filter*

Figure 4 is a sketch of the directional filter. It shows the location of the two-cavity bandstop filter and the two-cavity bandpass filter. Both filters are channel-frequency dependent, and their resonance cavities require high-conductivity walls to reduce losses. These properties separate the directional filter sufficiently from the other components to make it a separate module. Either type of filter is traditionally made from sections of standard copper waveguide.

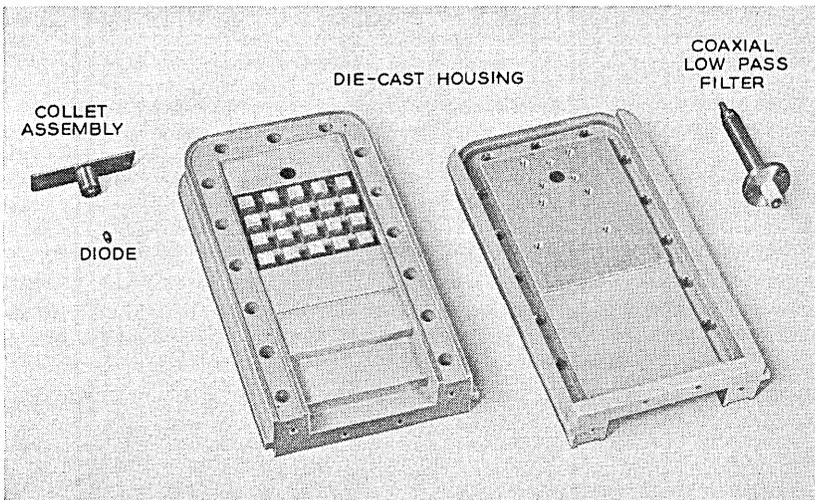


Fig. 12 — Four of the six mechanical modules.

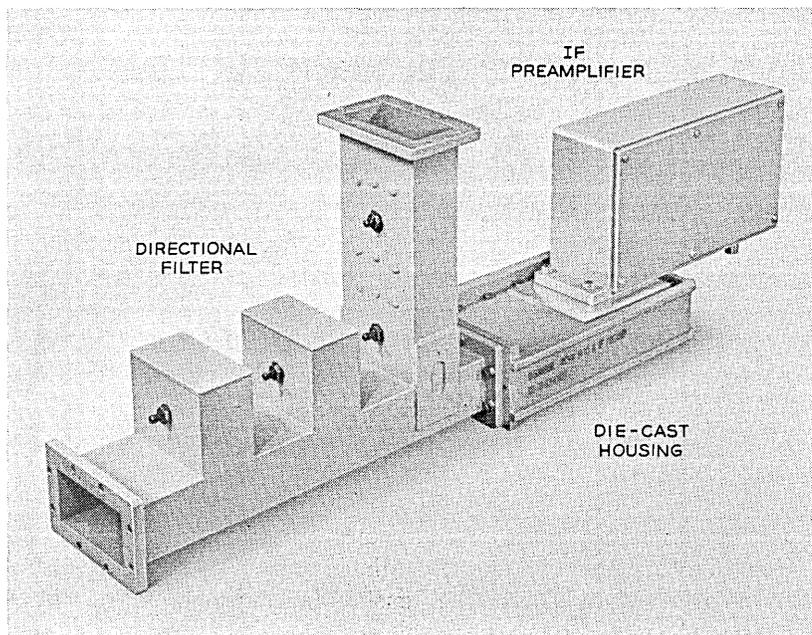


Fig. 13 — Preproduction model of the downconverter-preamplifier assembly.

Combining the two filters into a single unit was within existing manufacturing and testing technology so that no mechanical development was needed. The two shorted waveguide sections on the horizontal arm of the filter in Fig. 13 are the resonance cavities of the bandstop filter. They are coupled to the horizontal waveguide by circular holes in the waveguide wall. A similar hole couples the lower of the two bandpass cavities in the vertical arm to the common port. The upper end of this cavity and both ends of the upper bandpass cavity are defined by triple-post obstacles. In manufacture, precise jiggling is combined with differential soldering to achieve connections that are both electrically and mechanically reliable. Copper tuning screws permit compensation of manufacturing tolerances.

#### 4.2 Die-cast Housing

The housing is assembled from two machined aluminum castings. The two halves, which are identical as raw castings, differ slightly in their final form because one half is machined to hold the collet assembly, while the other is machined to hold the coaxial LP filter and

the IF preamplifier. The housing was designed to be die-cast, although all preproduction models of the casting were made in plaster molds, which produce castings of quality similar to die casting at very reasonable cost for small runs or prototypes.

In order to use two identical castings, the housing had to be divided perpendicularly to the RF currents in the waveguide sidewalls. Particular care was taken in designing the mating surfaces in order to prevent RF leakage into or out of the assembly. The walls of the casting are 0.64 inch thick, but electrical contact is made on two 0.10-inch wide lands only. Clearance holes for the clamping screws are located between the two lands so that, on assembly, uniform and high contact pressure is assured on the inner land area. The cast-in clearance holes are spaced one inch apart around the periphery of the casting. With properly tightened 0.164 UNC screws, no RF interference was encountered even when the castings were slightly warped. (Some warping of the castings occurs during ejection from the die, even though generous drafts are provided on all noncritical surfaces.)

Good die castings have adequate dimensional accuracy and surface texture so that the inside of the housing and the peripheral lands do not have to be machined.

The assembly of the housing is straightforward. A cylindrical mandrel aligns the hole of the coaxial short with that of the coaxial LP filter. A rectangular mandrel aligns the WR-229 port. Fifteen hexagonal socket head screws are inserted through the cast-in holes into hexagonal nuts on the opposite side to clamp the assembly together. (A trough-like recess along the periphery of the castings prevents the nuts from turning during assembly.) At this point, the waveguide flange is drilled and finished in accordance with the Bell System standards for WR-229 waveguide flanges.

Die-casting alloy A360 was selected for the housing because it combines good corrosion resistance, castability, and strength with adequate electrical conductivity. Screws and nuts are made of zinc-plated carbon steel to reduce corrosion in humid climates.

#### 4.3 *The Collet Assembly*

To be able to remove the diode without dismantling the modulator was an early design goal. Some sort of device was needed which would tightly grip one end of the diode and, when inserted into the modulator housing, plug the other end of the diode into a receptacle. Since the coaxial lowpass filter did not lend itself readily as the gripping

device, and since the IF preamplifier would have to be removed to gain access to the coaxial LPF, it was logical that the gripping device should be on the grounded end of the diode. This evolved into the collet assembly shown in Fig. 14.

The depth of collet insertion is limited by a 0.060-inch-thick shoulder in the housing. The inside diameter of the shoulder is the outside diameter of the previously-mentioned coaxial short, while the collet forms the inside diameter and the shorting plane. The shorting "plane" is actually a truncated cone. This not only assures that RF contact between the collet and the housing takes place at a well-defined diameter, that is, at the outside diameter of the short coaxial line, but it also reduces the likelihood of RF leakage into the receiver modulator from extraneous sources.

The collet is gold plated to reduce the dc contact resistance between diode and collet, as well as between collet and housing. The collet is designed to make cylindrical contact with the diode (rather than a circumferential line contact which, in the presence of friction, can lead to poor axial alignment of the diode in the collet).

A beam spring limits the axial force exerted by the collet assembly on the 0.060-inch thick shoulder in the housing. Experiments on a number of preproduction modulators indicated that a seating force of about 100 pounds is needed to prevent RF leakage.

The bending moment acting on the two screws which fasten the

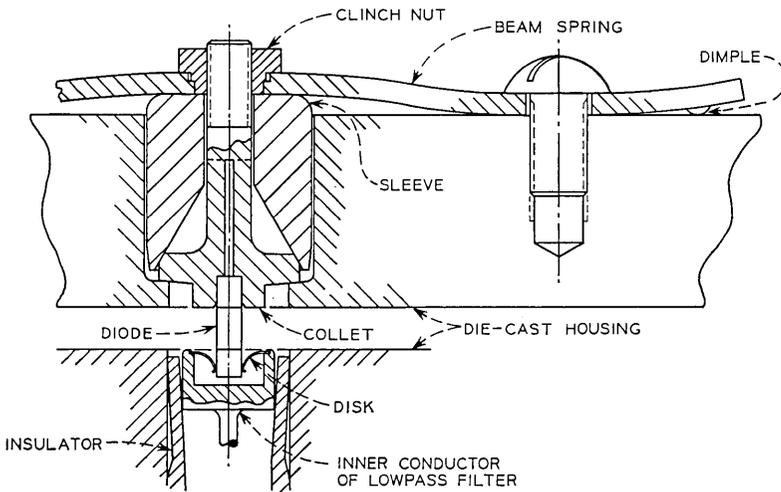


Fig. 14 — Diode mounting.

collet assembly to the housing is greatly reduced by dimpling the beam spring as shown in Fig. 14. Although the tensile force acting on the screws is thereby approximately doubled, the maximum stress in the screws is significantly reduced. Hard yellow brass was chosen for the beam spring because it combines adequate energy storage without heat treatment with good corrosion resistance, creep resistance, and sufficiently low hardness for pressing-in a commercial clinch nut. A nut that is integral with the beam spring guards against accidental release of the diode before the collet assembly is removed from the modulator.

#### 4.4 *The Coaxial Lowpass Filter*

The coaxial lowpass filter consists of three parts, the outer conductor, insulator, and inner conductor (see Fig. 15.) As mentioned in Section 4.3, one end of the inner conductor has to serve as the diode receptacle. The major problem was designing this receptacle: how to make repeatable electrical contact between the diode and the filter without unduly stressing the relatively fragile diode. A more conventional problem concerned the positioning of the inner conductor in relation to the outer conductor (which is fastened to the housing) so that the diode receptacle would be flush with the waveguide wall and centered in its hole in the housing.

The second problem was solved by mechanically interlocking the injection-molded insulator and the inner conductor and then staking the combination in place through the wall of the outer conductor. Radial play between the diode receptacle and the waveguide wall is eliminated by a circumferential ridge (on the insulator), the diameter of which is somewhat larger than the hole in the modulator housing.

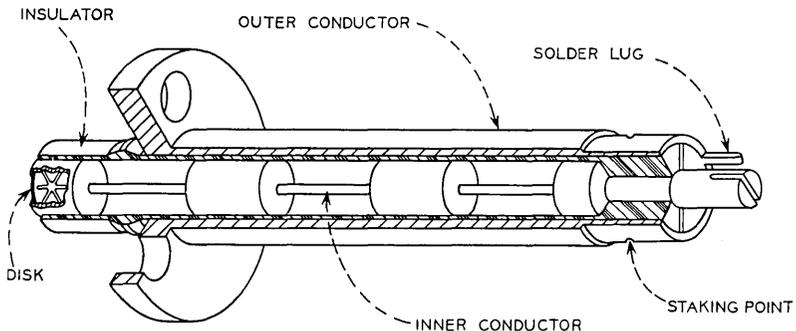


Fig. 15 — Coaxial lowpass filter.

A diametrical slot at each end of the insulator provides adequate radial compliance for both mechanical functions.

But designing a reliable diode receptacle was further complicated by the possibility of considerable diode misalignment from tolerance build-up in the collet assembly and from crookedness of the diode itself. After considering various alternatives, it was decided to attach a thin Beryllium-copper disk with six approximately triangular radial fingers (see Fig. 16) to the hollowed end of the inner conductor.

The constant-thickness finger is designed as a beam of uniform strength, that is, after a diode has been inserted, the maximum bending stress is the same in every cross section of the finger and hence its elastic deflection is at its greatest. The elastic deflection capability of the fingers is further increased by forming them after precipitation hardening rather than in the annealed state. It can be shown that the favorable residual stress distribution resulting from such a severe plastic deformation after hardening raises the elastic deformability by almost 50 per cent. The tips of the fingers are sharply bent to prevent them from digging into the diode during diode extraction.

It can be shown that the optimum number of radial fingers depends on the ratio of the width to the length of the radial slots separating the fingers. (The derivation is quite simple because of the constant-stress distribution in the fingers. Since we are interested in maximum elastic deformation of the fingers under a given load, we have to maximize the total potential energy stored in the disk. Since this is approximately proportional to the volume of the fingers, the optimum number of fingers is that which maximizes the volume of the fingers.) In general, the smaller the ratio of slot width to slot length, the greater are the optimum number of fingers and the elastic deforma-

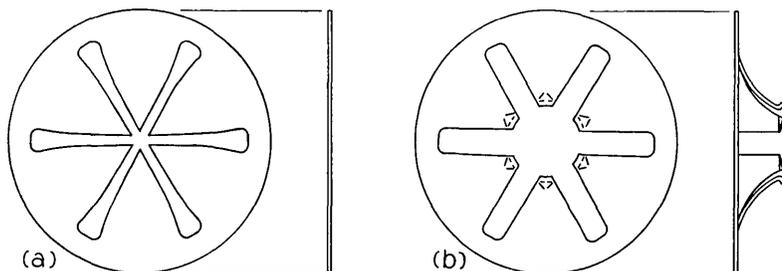


Fig. 16—(a) The disk blank after photo etching. (b) The disk ready for installation after hardening and cold forming.

tion under a given load. We found a slot width of 0.010 inch necessary to assure separation of the fingers during manufacture. For this width and a 0.085-inch slot length, the theoretical optimum number of fingers is between 6 and 7.

The disk is photo-etched from 0.002-inch beryllium-copper foil. Its rim is clamped between an annular shoulder and a spun-over thin lip, both of which are part of the inner conductor (see Fig. 15). Both the inner and outer conductors are machined from free-cutting brass. The disk and the inner conductor are gold plated to minimize dc resistance.

The diode insertion force levels off at about  $\frac{1}{2}$  pound after many insertions and extractions. The contact resistance between simulated diodes and inner conductors was found to be about 1 milliohm. This is negligible in comparison with the series resistance  $R_s$  of the diode which is typically 1 ohm.

## V. PERFORMANCE

### 5.1 Local Oscillator Return Loss

Figure 17 shows the local oscillator return loss of the converter block as discussed in item (i) of Section 3.2 (see Fig. 7). The measurement was made for three different diodes and in each case for three different bias currents, which were chosen to yield an optimum local oscillator match at  $f_{LO} = 3780, 3940,$  and  $4100$  MHz. The maximum local oscillator power of  $P_{LO} = 6$  dBm was used, since it yielded the lowest  $CL$ . It is evident that a rather good broadband match can be achieved for each diode around any local oscillator frequency between 3780 and 4120 MHz.

It was found that the total capacitance  $C_{TO}$  of the diode at zero bias is virtually the only characteristic of the diode which affects the bias current required for an optimum match at a given local oscillator frequency; as a consequence, diodes with equal  $C_{TO}$  require practically identical bias currents.

The diode specification<sup>9</sup> calls for  $0.30 \text{ pF} \leq C_{TO} \leq 0.60 \text{ pF}$ . Therefore, Fig. 17 with  $C_{TO} = 0.31, 0.45,$  and  $0.61 \text{ pF}$  is quite representative of the range of  $C_{TO}$  that will be available. It can, therefore, be concluded from Fig. 17 that the local oscillator return loss of the converter block is above 15 dB for any diode and for any local oscillator frequency between 3780 and 4120 MHz after proper adjustment of the bias current. Since the return loss at port 2 of the directional filter (Figs. 4 and 5) is well above 20 dB, the return loss  $RL_{LO}$  at  $f_{LO}$  of the downconverter-

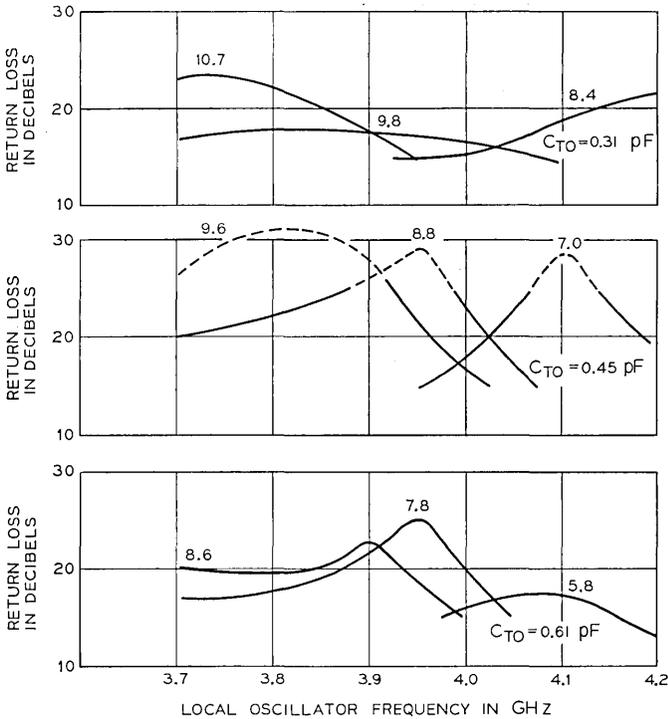


Fig. 17 — Measured local oscillator return loss of the converter block. (Diode bias current in mA.)

preamplifier unit at its local oscillator input port (Fig. 3) is well above the specified value of 10 dB, typically  $> 15$  dB.

It was also found that the incremental change in bias current needed to shift the optimum local oscillator return loss point from one local oscillator frequency to another is practically independent of the diode and approximately linearly related to the frequency difference. To a good approximation, the bias current,  $I$ , required for an optimum match at  $f_{LO}$  can be obtained from:

$$I(f_{LO}) = I(3940 \text{ MHz}) - 1.2 \text{ mA} \frac{f_{LO}/\text{MHz} - 3940}{340}. \quad (5)$$

Finally, it was determined that, in order to remain at the point of optimum local oscillator match with varying local oscillator power, the bias current should be kept constant. In practice this was realized

to a good approximation by biasing the diode from the available voltage of  $-19\text{ V}$  through a series resistor.

5.2 *IF Output Impedance*

Figure 18 shows the IF output impedance  $Z_{IFout}$  discussed in (ii) of Section 3.2 for the three diodes mentioned in Section 5.1.  $Z_{IFout}$  is fairly independent of the particular diode and local oscillator frequency, and it is approximately real and constant from 60 to 80 MHz. The mean value is:

$$Z_{IFout} = 58\Omega - j16\Omega. \tag{6}$$

5.3 *Transmission Characteristics*

Figure 19 shows a typical signal-to-IF amplitude transmission characteristic of the downconverter as discussed in (iii) of Section 3.2.

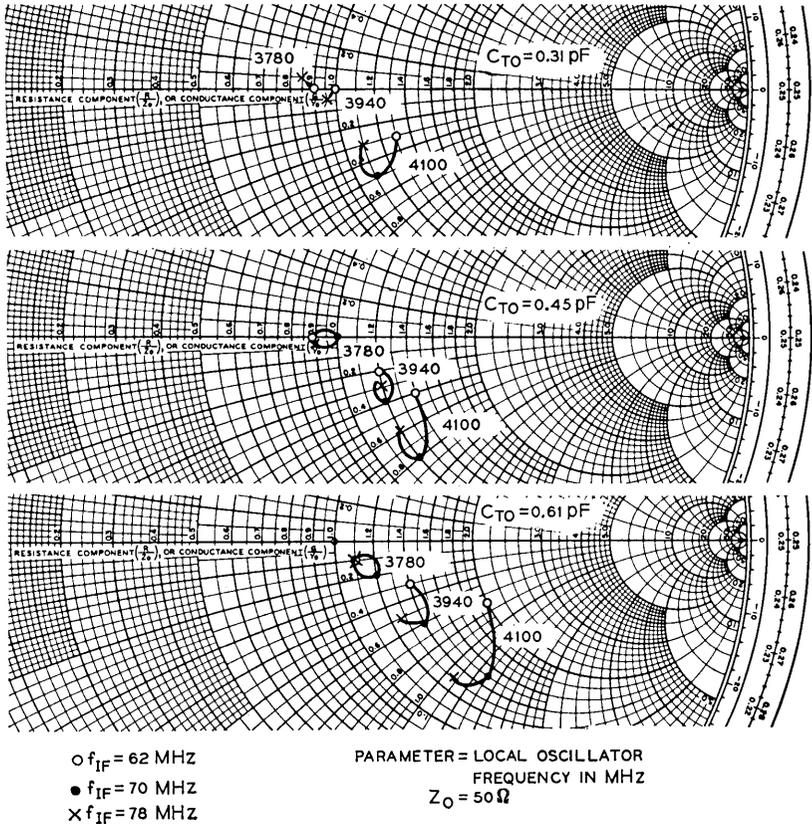


Fig. 18 — Measured IF output impedance of the downconverter.

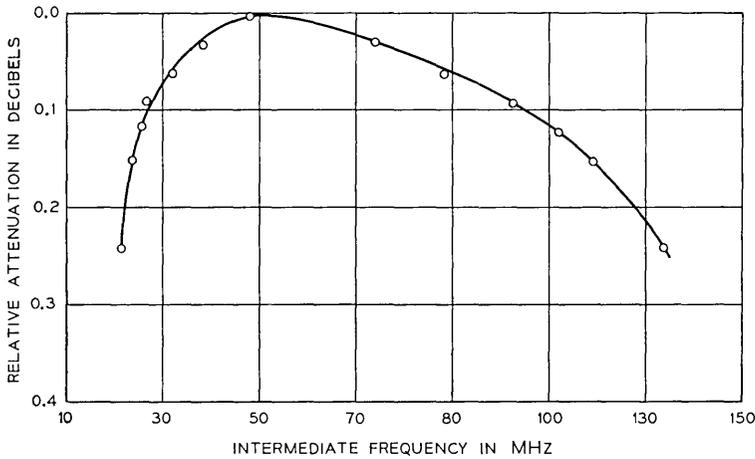


Fig. 19—Measured signal-to-IF amplifier transmission characteristic of the downconverter.

Because of the availability of test equipment, the IF load  $R_{IF}$ , which shall approximately equal  $Z_{IFout}$  as explained in Section 3.2, was chosen to be:

$$R_{IF} = 50\Omega. \quad (7)$$

A typical conversion loss  $CL$  at 70 MHz is:

$$CL = 4.2 \text{ dB}. \quad (8)$$

Figure 19 shows that the signal-to-IF amplitude transmission characteristic of the downconverter typically exhibits a slope of 0.03 dB per 20 MHz from 60 to 80 MHz, which is definitely small enough to be corrected in the IF preamplifier.

Figure 20 shows a typical signal-to-IF amplitude transmission characteristic of the downconverter-preamplifier unit (Fig. 3) after properly adjusting the two transmission amplitude controls and the output level control of the preamplifier. The specifications for transmission amplitude flatness (indicated tolerance field) are met with ample margin. However, notice that although the response of Fig. 20 is quite flat, it is considerably narrower than that of Fig. 19 because of the bandpass characteristic of the IF preamplifier.

#### 5.4 Noise Figure

For various signal and local oscillator frequencies, Fig. 21 shows the median total noise figure  $NF_{Total}$  of a downconverter-preamplifier

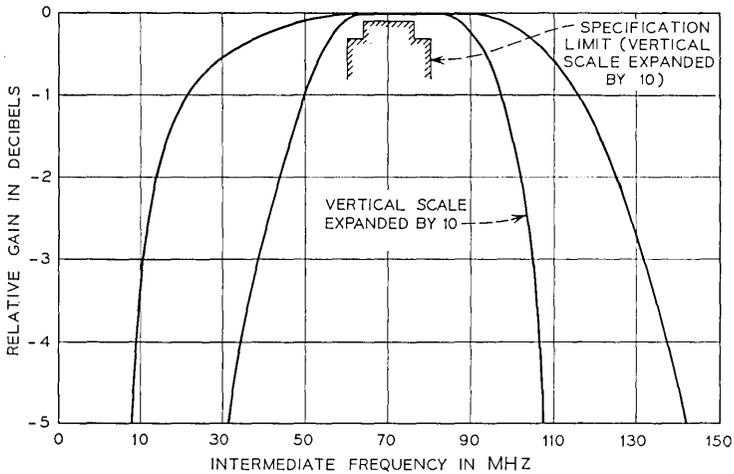


Fig. 20—Measured signal-to-IF amplitude transmission characteristic of the downconverter-preamplifier unit.

unit (Fig. 3) obtained from a sample of 97 diodes. The noise figure is lowest near the center of the band, and rises by 0.1 and 0.3 dB at the low and the high end of the band, respectively. The noise figure of the IF preamplifier,  $NF_{IF}$  as defined in Section 3.2, was in this case:

$$NF_{IF} = 2.8 \text{ dB} \quad (9)$$

measured from  $R_{IF} = 50\Omega$ .

It must be pointed out here, that as mentioned in Ref. 10, the input impedance of the IF preamplifier is *not* equal to  $R_{IF} = 50\Omega$ , contrary to the assumption made in (iv) of Section 3.2. Hence, equation (4) cannot be expected to give the relationship between a typical  $NF_{Total}$  of Fig. 21 and  $CL$  and  $NF_{IF}$  of equations (8) and (9), respectively, since  $CL$  is measured with  $R_{IF} = 50\Omega$  as load impedance.

The diode specification<sup>9</sup> calls for a maximum total noise of 7.3 dB for the downconverter-preamplifier unit with  $NF_{IF} = 2.8$  dB and at  $f_{SI} = 3950$  MHz. (The isolator with an insertion loss of approximately 0.2 dB is excluded). The IF preamplifier specification limits  $NF_{IF}$  to 3.0 dB maximum. Thus the total noise figure is limited to 8.0 dB maximum at the limits of the band (7.5 dB + 0.3 dB degradation at the high end of the band + 0.2 db from the isolator), which meets the requirement of < 8.0 dB. However, the average result for 49 pre-production models was  $NF_{Total} = 6.7$  dB (isolator excluded).

### 5.5 Compression Characteristics

Figure 22 shows the compression characteristic of the downconverter-preamplifier unit. It is seen that the break point of the downconverter is remarkably high ( $P_{st} = -4$  dBm) considering that  $P_{LO} = 6$  dBm. The compression characteristic of the downconverter-preamplifier unit is clearly determined by the preamplifier.

### 5.6 Level of Unwanted Signals

The results of a test made on a downconverter-preamplifier unit are:

Local oscillator suppression from the local oscillator port to:

- (i) the signal port: 89 dB.
- (ii) the IF preamplifier input: 65 dB.

Level of the second harmonic of the local oscillator at:

- (iii) the signal port: more than 76 dB below  $P_{LO}$ .
- (iv) the IF preamplifier input: more than 76 dB below  $P_{LO}$ .

Level of the third harmonic of the local oscillator at:

- (v) the signal port: more than 51 dB below  $P_{LO}$ .
- (vi) the IF preamplifier input: more than 38 dB below  $P_{LO}$ .

RF leakage:

- (vii) satisfactorily low.

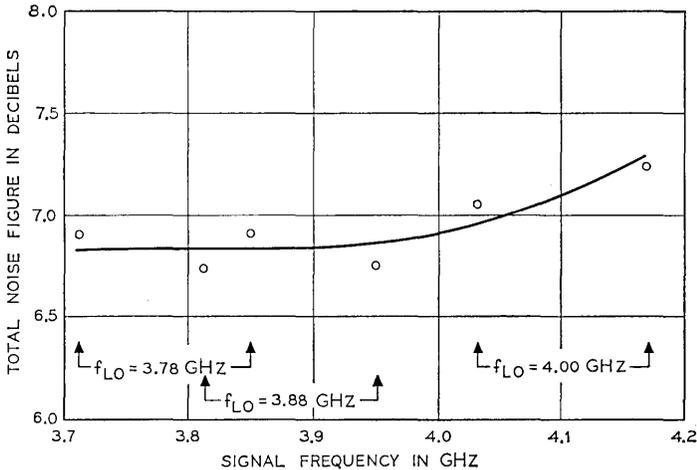


Fig. 21 — Median measured total noise figure of a downconverter-preamplifier unit for 97 diodes.

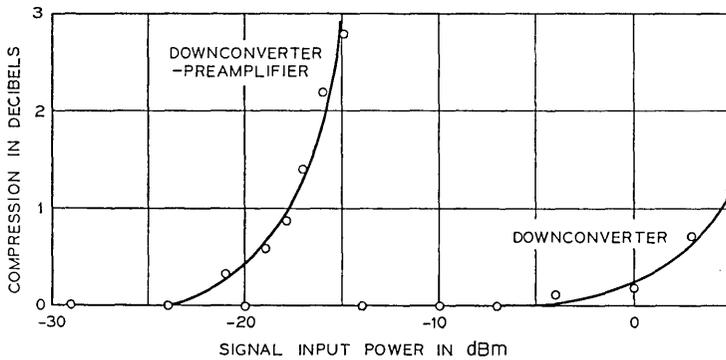


Fig. 22 — Measured compression characteristics of the downconverter and the downconverter-preamplifier unit.

Thus all requirements have been met with ample margin, except possibly for (vi). However, it can be expected that there will be at least 2 dB of attenuation between the input and output of the IF preamplifier for the third harmonic of the local oscillator frequency.

### 5.7 Temperature Behavior

A temperature test of a preproduction downconverter-preamplifier unit (Fig. 3) gave the following results:

Temperature (°F):	40	70	80	90	140
Local oscillator suppression from local oscillator port to signal port without isolator (dB):	29	—	51	—	28
Slope of the transmission characteristic from signal to IF (dB per 12 MHz):	0.04	0.01	0	-0.02	-0.16

These results show that the local oscillator suppression from the local oscillator port to the signal port stays well above the required 5 dB. The results also show that the amplitude transmission characteristic from signal to IF (0.01 dB per 12 MHz) is slightly more sensitive to temperature within the  $\pm 10^\circ\text{F}$  range, than would be desired. This, however, is not expected to impair the performance of the system noticeably. Performance at the extreme temperatures is such that the system will still be able to render a good service, as required.

## APPENDIX

It is the purpose of the Appendix to show that the noise performance of a single-diode downconverter is identical to that of a balanced downconverter, if the noise of the local oscillator is properly band limited.

The diode is modeled as a memoryless, frequency independent, quadratic scatterer with a built-in noise source:

$$b(t) = A_1 a(t) + A_2 a^2(t) + b_N(t).$$

$a(t)$  is the incident wave,  $b(t)$  is the reflected wave,  $b_N(t)$  is that portion of  $b(t)$  which results from the built-in noise source. All waves are normalized to a reference impedance  $R$ .  $A_1$  and  $A_2$  are real constants. For a linear network,  $A_1$  would be called the reflection coefficient.

To study the performance of this diode model in a single-diode downconverter, the converter is modeled as shown in Fig. 23. The "ideal multiplexer" is a lossless network with the following scattering matrix.

$$\begin{pmatrix} b_1 \\ b_2 \\ b_3 \\ b_4 \\ b_5 \\ b_6 \end{pmatrix} = \begin{pmatrix} S_{11} & 0 & 0 & 0 & 0 & S_{16} \\ 0 & S_{22} & 0 & 0 & 0 & S_{26} \\ 0 & 0 & S_{33} & 0 & 0 & S_{36} \\ 0 & 0 & 0 & S_{44} & 0 & S_{46} \\ 0 & 0 & 0 & 0 & S_{55} & S_{56} \\ S_{16} & S_{26} & S_{36} & S_{46} & S_{56} & 0 \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \\ a_3 \\ a_4 \\ a_5 \\ a_6 \end{pmatrix}$$

The nonzero elements of this matrix have the following values as a function of frequency:

- (i)  $S_{ii} = 0$  for the frequency range  $B_i$ , otherwise  $|S_{ii}| = 1$ .
- (ii)  $S_{i6} = 1$  for the frequency range  $B_i$ , otherwise  $S_{i6} = 0$ .
- (iii) The five  $B_i$  ( $i=1, 2, 3, 4$  and  $5$ ) are defined as:

$$B_1: 0 \leq \omega \leq 2\alpha$$

$$B_2: \omega_2 - 2\alpha \leq \omega \leq \omega_2 + 2\alpha$$

$$B_3: \omega_3 - 2\alpha \leq \omega \leq \omega_3 + 2\alpha$$

$$B_4: \omega_3 - \omega_2 - 2\alpha \leq \omega \leq \omega_3 - \omega_2 + 2\alpha$$

$$\omega_3 + \omega_2 - 2\alpha \leq \omega \leq \omega_3 + \omega_2 + 2\alpha$$

$$B_5: \text{All other } \omega.$$

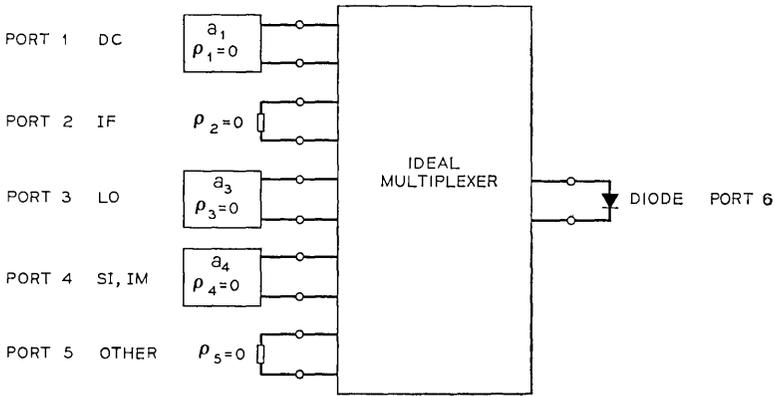


Fig. 23 — Model of a single-diode downconverter.

Here  $\omega_3$  denotes the frequency of the local oscillator carrier,  $(\omega_3 \pm \omega_2)$  the frequency of the signal carrier and the image carrier, respectively, and  $\omega_2$  the frequency of the IF carrier. This also explains the nomenclature used at ports 1 through 5 (dc, IF, and so on). The frequency  $\alpha$  is restricted to  $\alpha < \omega_2/4$ , otherwise arbitrary.

As shown in Fig. 23, ports 1 through 5 are terminated in sources or loads. All are matched ( $\rho_i=0$ ) and the source waves are:

$$a_1(t) = a_1$$

$$a_3(t) = a_3[1 + m_3(t)] \cos [\omega_3 t + \varphi_3(t) + \psi_3]$$

$$a_4(t) = a_4[1 + m_4(t)] \cos [\omega_3 t + \omega_2 t + \varphi_4(t) + \psi_4]$$

$a_1, a_3, a_4, \psi_3, \psi_4$  are real constants,  $m_3(t), \varphi_3(t)$  describe AM and PM noise impressed on the local oscillator carrier, and  $m_4(t), \varphi_4(t)$  describe AM and PM noise or modulation impressed on the signal carrier. It is assumed that the spectra of  $m_3(t)$  and  $m_4(t)$  are limited to the frequency range  $\pm\alpha/2$ , and that the spectra of  $\cos [\omega_3 t + \varphi_3(t) + \psi_3]$  and  $\cos [\omega_3 t + \omega_2 t + \varphi_4(t) + \psi_4]$  are limited to the frequency ranges  $\omega_3 \pm \alpha/2$  and  $\omega_3 + \omega_2 \pm \alpha/2$ , respectively.

Hence, the spectra of  $a_3(t)$  and  $a_4(t)$  are limited to the frequency ranges  $\omega_3 \pm \alpha$  and  $\omega_3 + \omega_2 \pm \alpha$ , respectively. It should be remembered that, in addition to  $a_1(t), a_3(t)$  and  $a_4(t)$ , white thermal noise is delivered to the ideal multiplexer at ports 1 through 5 from each of the matches presented to the multiplexer at these ports. These incident noise waves are denoted by  $a_{N1}(t), a_{N2}(t)$ , and so on.

The wave incident to the diode at port 6 can then be calculated:

$$a(t) = a_1(t) + a_3(t) + a_4(t) + a_N(t)$$

where  $a_N(t)$  represents the white thermal noise wave incident to the diode from the match which the multiplexer presents to the diode. Obviously it is:

$$a_N(t) = a_{N1}(t)_{B_1} + a_{N2}(t)_{B_2} + a_{N3}(t)_{B_3} + a_{N4}(t)_{B_4} + a_{N5}(t)_{B_5}$$

where the subscripts indicate that only the components falling within the respective frequency range are to be taken. If it is now assumed that:

$$a_4, a_N(t) \ll a_1, a_3$$

the wave reflected from the diode can readily be calculated. In particular, the wave  $b_2(t)$  incident to the load at port 2 (IF) is:

$$\begin{aligned} b_2(t) &= A_2 a_3 a_4 [1 + m_3(t)][1 + m_4(t)] \cos [\omega_2 t + \varphi_4(t) - \varphi_3(t) + \psi_4 - \psi_3] \\ &\quad + \langle A_1 a_N(t) + b_N(t) + 2A_2 a_1 a_N(t) + 2A_2 a_3 [1 + m_3(t)] \\ &\quad \cdot a_N(t) \cos [\omega_3 t + \varphi_3(t) + \psi_3] \rangle_{B_2} \\ &= A_2 a_3 a_4 [1 + m_3(t)][1 + m_4(t)] \cos [\omega_2 t + \varphi_4(t) - \varphi_3(t) + \psi_4 - \psi_3] \\ &\quad + (A_1 + 2A_2 a_1) a_{N2}(t)_{B_2} + b_N(t)_{B_2} \\ &\quad + \langle 2A_2 a_3 [1 + m_3(t)] \cos [\omega_3 t + \varphi_3(t) + \psi_3] a_{N4}(t)_{B_4} \rangle_{B_2} . \end{aligned}$$

To study the performance of the same diode model when used in a balanced downconverter, the converter is modeled as shown in Fig. 24.

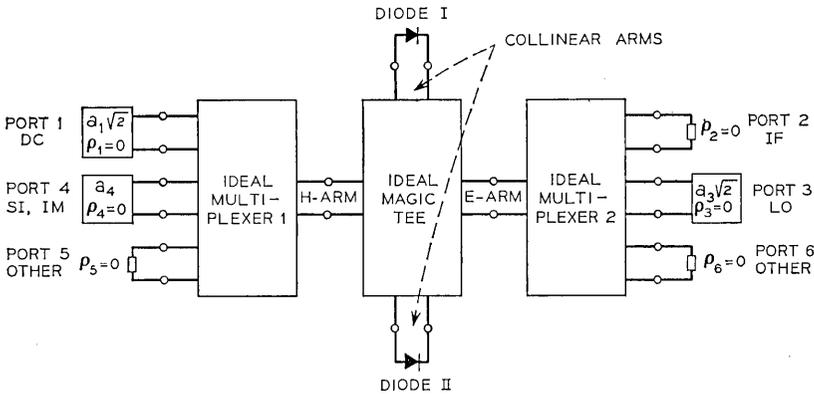


Fig. 24 — Model of a balanced downconverter.

The ideal tee hybrid is a lossless hybrid matched at all four ports. It has infinite isolation, and the transmission coefficients are  $2^{-\frac{1}{2}}$  from the H-arm to both collinear arms and  $\pm 2^{-\frac{1}{2}}$  from the E-arms to the top (diode I) and bottom (diode II) collinear arm, respectively. The two ideal multiplexers have properties analogous to those outlined for the ideal multiplexer of Fig. 23.

Notice that the signal source at port 4 remained unaltered in strength (since this can commonly not be changed), but that the power of both the DC and LO source have been increased by a factor of 2 (wave amplitudes increased  $2^{\frac{1}{2}}$ ) in order to provide the same DC and LO power to each diode as before for the single diode.

The wave incident to diode I is then:

$$a_I(t) = a_1(t) + a_3(t) + 2^{-\frac{1}{2}}a_4(t) + 2^{-\frac{1}{2}} \\ \cdot [a_{N1}(t)_{B_1} + a_{N4}(t)_{B_4} + a_{N5}(t)_{B_5} + a_{N2}(t)_{B_2} + a_{N3}(t)_{B_3} + a_{N6}(t)_{B_6}]$$

and the wave incident to diode II is:

$$A_{II}(t) = a_1(t) - a_3(t) + 2^{-\frac{1}{2}}a_4(t) + 2^{-\frac{1}{2}} \\ \cdot [a_{N1}(t)_{B_1} + a_{N4}(t)_{B_4} + a_{N5}(t)_{B_5} - a_{N2}(t)_{B_2} - a_{N3}(t)_{B_3} - a_{N6}(t)_{B_6}].$$

The resulting wave  $b_2(t)$  incident to the load at port 2 (IF) is now:

$$b_2(t) = A_2a_3a_4[1 + m_3(t)][1 + m_4(t)] \cos [\omega_2t + \varphi_4(t) - \varphi_3(t) + \psi_4 - \psi_3] \\ + \langle A_1[a_{N2}(t)_{B_2} + a_{N3}(t)_{B_3} + a_{N6}(t)_{B_6}] + 2^{-\frac{1}{2}}[b_{NI}(t) - b_{NII}(t)] \\ + 2A_2a_1[a_{N2}(t)_{B_2} + a_{N3}(t)_{B_3} + a_{N6}(t)_{B_6}] + 2A_2a_3[1 + m_3(t)] \\ \cdot [a_{N1}(t)_{B_1} + a_{N4}(t)_{B_4} + a_{N5}(t)_{B_5}] \cos [\omega_3t + \varphi_3(t) + \psi_3] \rangle_{B_2} \\ = A_2a_3a_4[1 + m_3(t)][1 + m_4(t)] \cos [\omega_2t + \varphi_4(t) - \varphi_3(t) + \psi_4 - \psi_3] \\ + (A_1 + 2A_2a_1)a_{N2}(t)_{B_2} + 2^{-\frac{1}{2}}[b_{NI}(t) - b_{NII}(t)]_{B_2} \\ + \langle 2A_2a_3[1 + m_3(t)] \cos [\omega_3t + \varphi_3(t) + \psi_3]a_{N4}(t)_{B_4} \rangle_{B_2}.$$

$b_{NI}(t)$  and  $b_{NII}(t)$  are the built-in noise waves of diode I and diode II, respectively.

Since  $b_{NI}(t)$  and  $b_{NII}(t)$  are uncorrelated and represent the same noise as  $b_N(t)$  for the unbalanced downconverter, it is seen that  $b_2(t)$  is equal for the balanced and the single-diode downconverter. As can be seen by going through the detailed analysis, this is a consequence of the fact that the spectra of  $m_3(t)$  and  $a_3(t)$  are limited to the frequency ranges  $\pm \alpha/2$  and  $\omega_3 \pm \alpha$ , respectively, and that  $\alpha < \omega_2/4$ .

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# Transmitter Modulator and Receiver Shift Modulator

By ANDRAS HAMORI and PAUL L. PENNEY

(Manuscript received November 14, 1967)

*The frequency upconverter provides one of the basic building blocks of heterodyne radio systems. In the TD-3 repeater, two waveguide modulators based on a balanced tee hybrid structure provide the required upconverter functions. The transmitter modulator uses high efficiency varactor diodes to convert the IF signal into an RF output. The second modulator, called the shift modulator, provides a 40 MHz shift in frequency between the transmitter local oscillator signal and the receiver local oscillator signal. This circuit uses varistor diodes to obtain simple broadband performance.*

## I. INTRODUCTION

In the TD-3 radio repeater, two upconverters are used: the transmitter modulator and the receiver shift modulator.<sup>1</sup> Although the electrical requirements for these modulators are quite different, the microwave circuit is similar for both.

In a TD-3 microwave transmitter, an FM modulated 70 MHz IF signal is converted by the transmitter modulator to a microwave signal before amplification and transmission. Because the transmitter modulator is a true upconverter, the output contains two sidebands either of which may be used depending on the channel frequency of the particular repeater.

The TD-3 frequency plan requires a 40 MHz difference between the received and transmitted signals in a repeater bay.<sup>2</sup> Consequently, separate local oscillator signals are required for the receiver and the transmitter. Because of system frequency stability and economy, however, it is very desirable to use the same microwave generator for both applications.\* Therefore, the second local oscillator signal is provided by shifting the common generator frequency by 40 MHz

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\* For reasons explained in Ref. 1, separate microwave generators are used for the receiver and the transmitter of a main station bay.

with the shift modulator. Since this RF to RF conversion is lossy, the shift modulator is in the receiver portion of the TD-3 repeater where the lower local oscillator power is required. Like the transmitter modulator, the shift modulator operates as either a lower sideband upconverter or an upper sideband upconverter, depending on the channel frequency.

II. BASIC STRUCTURE

The transmitter and shift modulators use the same basic microwave structure consisting of a waveguide tee hybrid with coaxial circuits connected to the collinear arms. Having common parts for both modulator designs yields economies in manufacture, maintenance, and repair.

Figure 1 is a block diagram of the common circuit configuration and Fig. 2 is a cross-section of the actual modulator structure, including a connecting IF driver amplifier or 40 MHz oscillator circuit. Figures 3 and 4 are photographs of the equipment. The tee hybrid

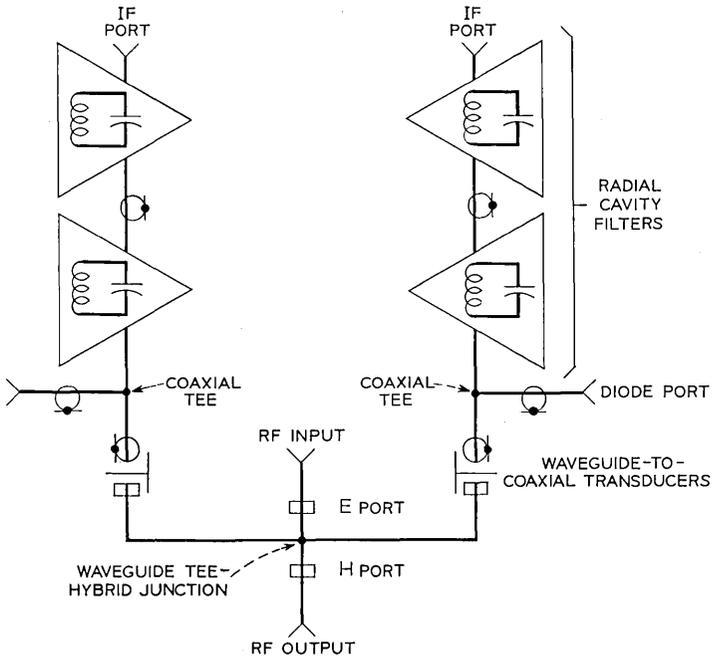


Fig. 1—Block diagram of common modulator structure.

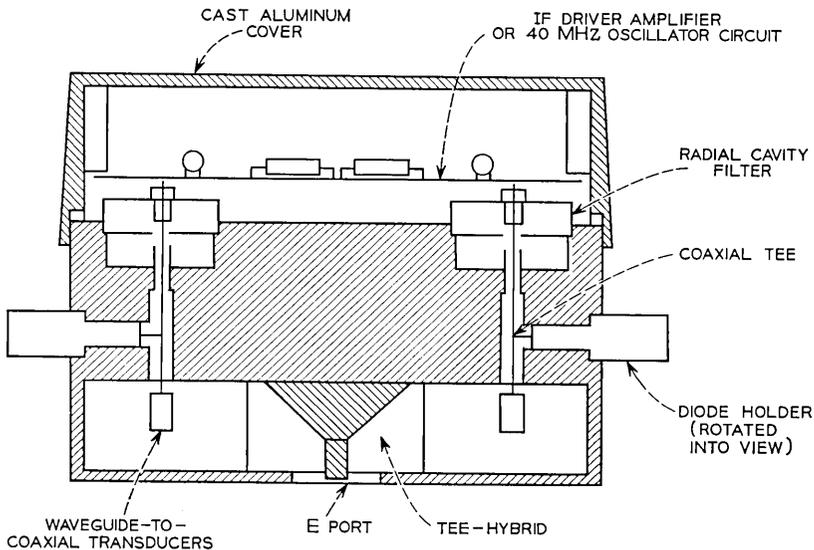


Fig. 2—Common modulator structure.

provides good isolation between the waveguide input and output ports. The magnitude of this isolation depends only on the physical symmetry of the entire modulator structure and the equality of the impedances of the diodes. Impedance at the input and output ports of the hybrid is controlled by three reactive elements: a post and an inductive iris for matching the source impedance to the E port, and a tuned cone for matching at the H port.

A waveguide-to-coaxial transducer provides the transition to a  $50\Omega$  coaxial line at each collinear arm of the hybrid junction. The transducer probe is offset in the guide to provide a good match over the entire TD-3 band. A coaxial-tee circuit provides interconnections among the transducer, the diode, and the low frequency (40 MHz or 70 MHz) port. To obtain best modulator efficiency, microwave energy is prevented from entering the low frequency port by use of a two-section radial cavity filter. This filter provides about 45 dB rejection to frequencies in the TD-3 band, but has little effect at 40 and 70 MHz. The open circuit is obtained at the junction of the coaxial lines by spacing the effective open circuit of the radial cavities one-half wavelength from the junction. This one-half wavelength section of coaxial line is composed of two parts: a quarter wavelength of 50-ohm line and a quarter wavelength of 20-ohm line. The 20-ohm

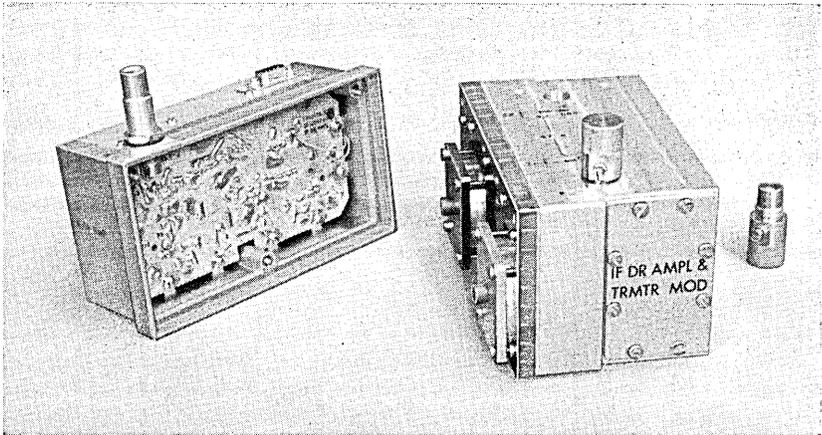


Fig. 3 — Transmitter modulator and IF driver amplifier.

section is used as a transformer and enhances the high impedance provided by the cavities.

The transistor circuits for both modulators are contained in cast aluminum covers that attach directly to the modulator block. In the transmitter modulator, the IF driver amplifier is housed in this cover, while in the 40 MHz shift modulator the cover contains the 40 MHz oscillator circuitry. This construction combines the shortest possible connections between the transistor and modulator circuits with the convenience of easily removable active circuits.

### III. TRANSMITTER MODULATOR AND IF DRIVER AMPLIFIER

The transmitter modulator is a balanced parametric upconverter using a matched pair of silicon epitaxial varactor diodes, Western Electric Co.\* type 471A. (The equipment described in this article is manufactured for Bell System use only.) The varactor diodes for this modulator circuit were developed especially for this application.<sup>3</sup> The use of a balanced structure based on a tee hybrid junction provides 3 dB more output power than could be obtained with a single diode, and provides 25 dB or more suppression of local oscillator power in the output circuit. The latter feature reduces the selectivity required in the sideband selecting filter. For a local oscillator input power of +20 dBm and an IF (70 MHz) input power of

\* Manufacturing and supply unit of the Bell System.

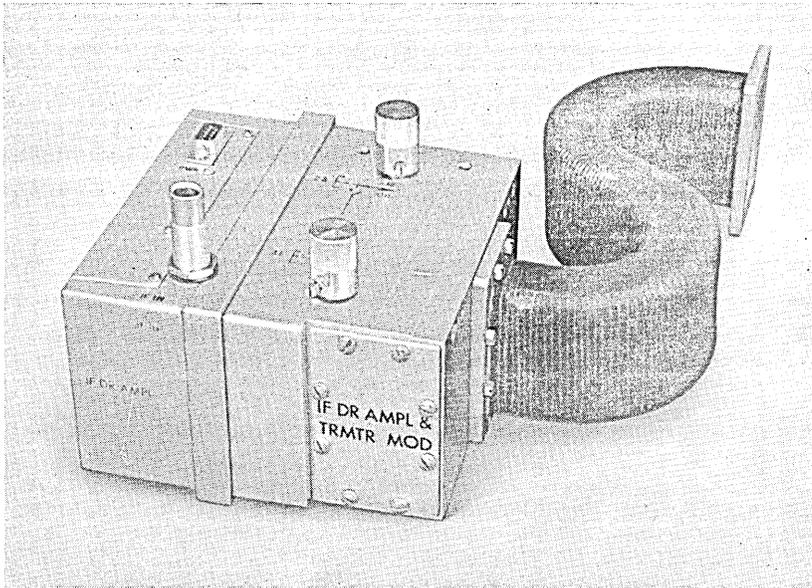


Fig. 4 — Assembled transmitter modulator with connecting waveguide.

about +3 dBm, the microwave output power of the transmitter modulator is at least +8 dBm and some production units reach +12 dBm.

The 471A diodes are mounted in coaxial holders which match the diodes to the 50Ω coaxial lines. Figure 5 shows details of the diode mounting arrangement. The diode matching circuit consists of a fixed

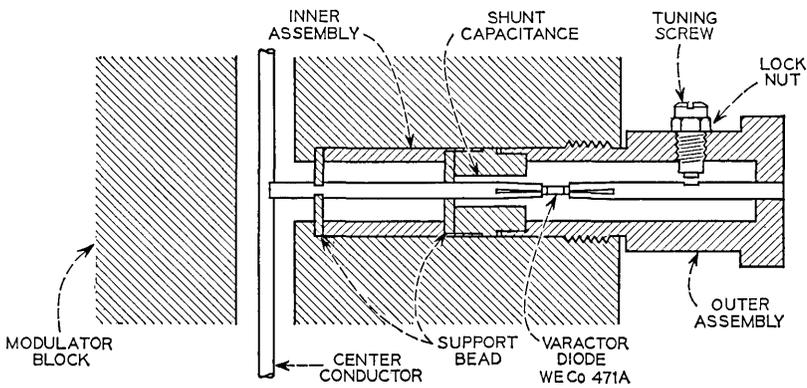


Fig. 5 — Tunable diode mount.

shunt capacitance located ahead of the diode, an adjustable capacitance, and a short-circuited section of coaxial line located behind the diode.

Upconverter modulators using varactor diodes often exhibit parametric oscillations at frequencies where high  $Q$  resonances exist. These oscillations can occur at frequencies at which the diode is not terminated. Since the waveguide-to-coaxial transducer provides a satisfactory match only at frequencies in the TD-3 frequency band, two additional kinds of terminations are incorporated in the modulator design to eliminate spurious oscillations. To provide reasonable terminations at frequencies up to 1000 MHz, a 50-ohm resistor is connected at each IF input port as part of the IF driver amplifier. For frequencies above 1000 MHz, a coaxial microwave termination is used on the IF side of each radial cavity as shown in Fig. 6. These microwave terminations have no effect at frequencies in the TD-3 band because of the isolation provided by the radial cavities, and have less than 0.1 dB loss at IF frequencies.

At the output port of the modulator, an isolator is used followed by a waveguide band-pass filter.<sup>1</sup> The band-pass filter selects and passes the desired sideband and reflects the undesired sideband into the isolator where it is absorbed. By absorbing the undesired sideband rather than reflecting it into the modulator for reconversion, a flatter modulator transmission characteristic is obtained.

In the radio transmitter, the IF driver amplifier provides an interface

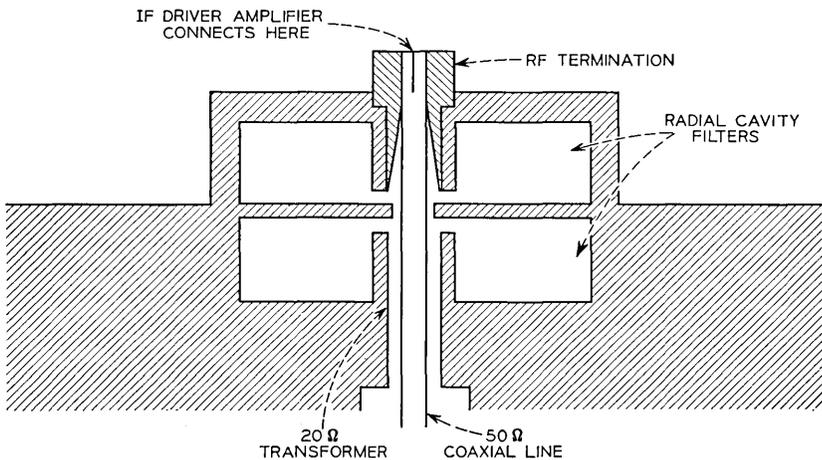


Fig. 6 — RF termination.

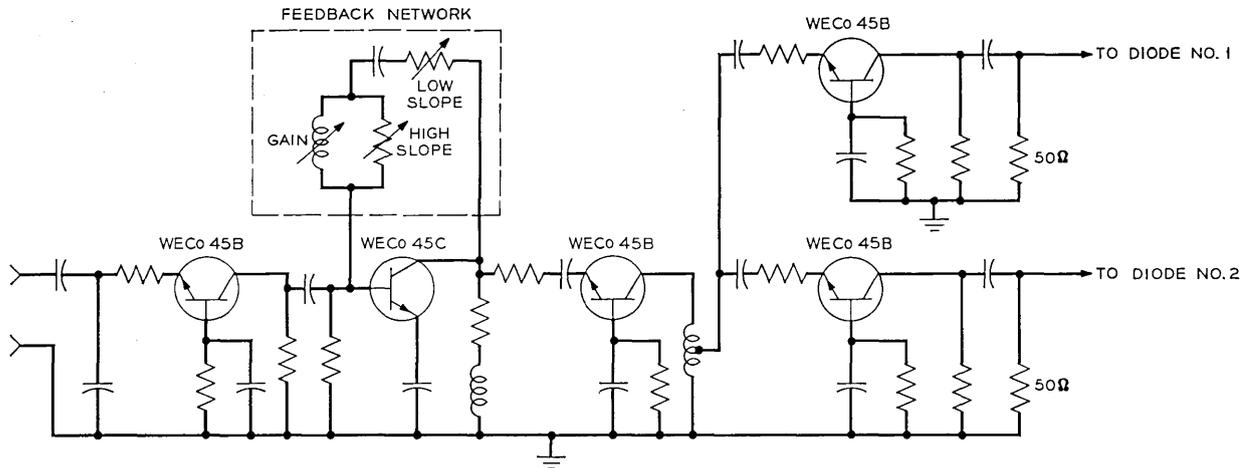


Fig. 7 — IF driver amplifier.

between the limiter and the transmitter modulator.<sup>1</sup> The amplifier provides about 7 dB gain from the 75-ohm input to each of two 50-ohm outputs with transmission flatness better than 0.03 dB over the  $70 \pm 10$  MHz band. Figure 7 is a simplified diagram of the circuit used. It consists of five stages with a common base input stage for impedance matching. The input return loss is better than 35 dB. The second stage is a common emitter amplifier with shunt feedback. Adjustment for over-all circuit gain and transmission shape are provided in the feedback network. The output stages use common base configurations. Separate stages are used for each 50-ohm output to provide isolation between the modulator varactor diodes.

The amplifier operates from a regulated  $-19$  volt supply and draws about 120 mA of current. The amplifier is constructed on a printed circuit board and is mounted in a cast aluminum housing which mounts directly on the transmitter modulator structure.

#### IV. SHIFT MODULATOR AND 40 MHz OSCILLATOR

The TD-3 shift modulator uses a matched pair of silicon point contact diodes, WECO. type 416C, in a balanced circuit based on a tee hybrid junction. See Fig. 8. Use of a balanced circuit provides 3 dB higher output level and more suppression of the microwave beat frequency input signal than could be obtained with a single diode in an unbalanced circuit. Varistor rather than varactor diodes were selected for this circuit because conversion gain was not required and only a relatively low output was needed.

The 416C diodes are mounted in coaxial holders which provide matching to the 50-ohm line as shown in Fig. 9. A four section low-pass filter is used in front of each diode to suppress by at least 40 dB the microwave second harmonics generated by the diodes. This amount of suppression is required to ensure an adequately low 80 MHz signal at the output of the receiver modulator.<sup>2</sup> This 80 MHz signal is produced in the receiver modulator as a difference product of the second harmonic of the microwave input and the second harmonic of the microwave output signals of the shift modulator.

At the output of the modulator, the desired sideband is selected by a waveguide band-pass filter while the undesired sideband is reflected into the modulator. Because the modulator has a single frequency output, the phase of the reflected sideband is easily optimized to obtain about 1 dB increase in output.

The modulator requires a  $+17$  dBm microwave input and a  $+17.5$

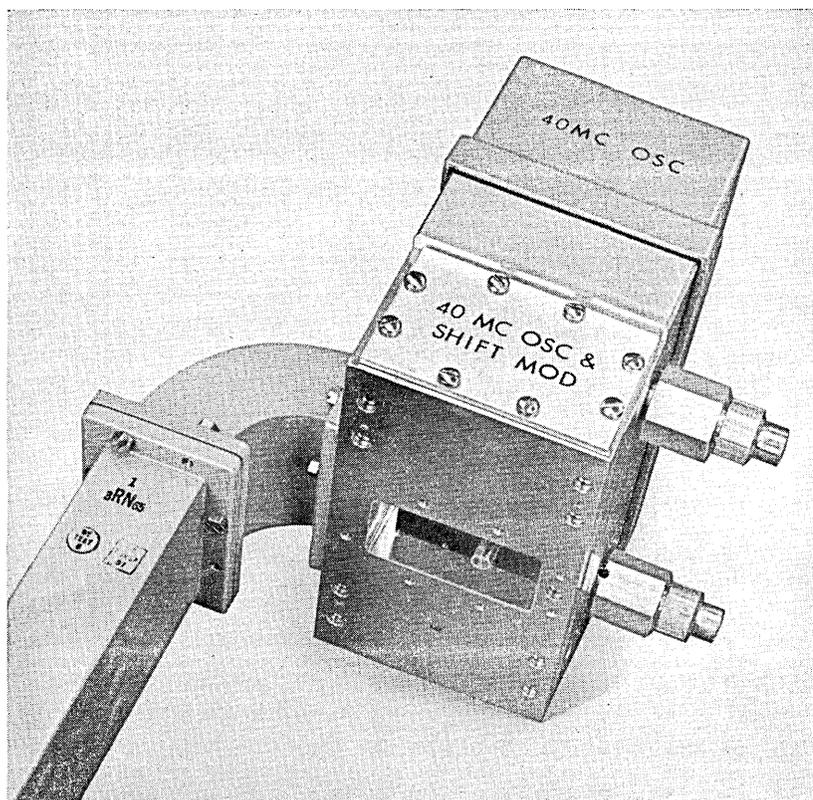


Fig. 8—Shift modulator and 40 MHz oscillator.

dBm 40 MHz input to provide an output signal of at least +6.5 dBm; some production units reach +8.5 dBm.

The 40 MHz signal is generated in the shift oscillator (see the simplified schematic diagram, Fig. 10.) The circuit consists of three stages: a crystal-controlled oscillator, a tuned common emitter buffer amplifier, and a tuned common emitter output amplifier.

The oscillator stage uses a third overtone crystal which operates in series resonance. A small variable capacitor,  $C_1$ , in series with this crystal, provides a frequency adjustment range of about  $\pm 500$  Hz. A thermistor in the emitter circuit of  $Q_1$  reduces over-all circuit output level variations resulting from temperature changes. The frequency of the oscillator is stable to  $\pm 400$  Hz for a  $\pm 10^\circ\text{F}$  temperature change.

An output power control with about 6 dB range is provided in the

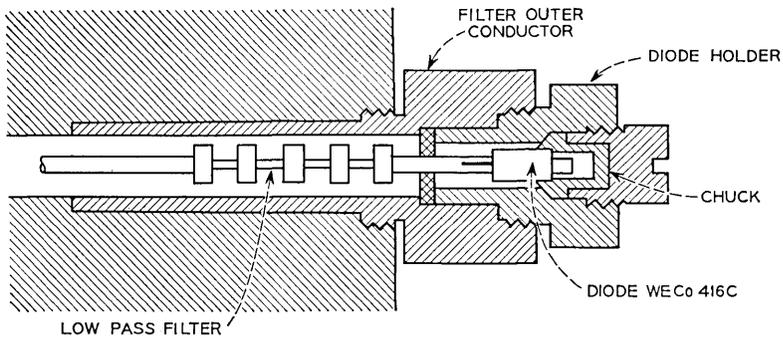


Fig. 9—Shift modulator diode holder.

circuit by a simple resistive voltage divider ( $R_1$  and  $R_2$ ) between the two amplifier stages. In order to measure output power and frequency in service, a monitor jack is provided. This access port is decoupled by 20 dB when terminated in 75 ohms and produces negligible effect on the output circuit. When the monitoring tap is not used, it is terminated with a 75-ohm load. A small portion of the 40 MHz output signal is rectified by  $D_1$  and used for external metering.

Like the IF driver amplifier, the circuit is housed in a cast aluminum cover which attaches directly to the shift modulator structure.

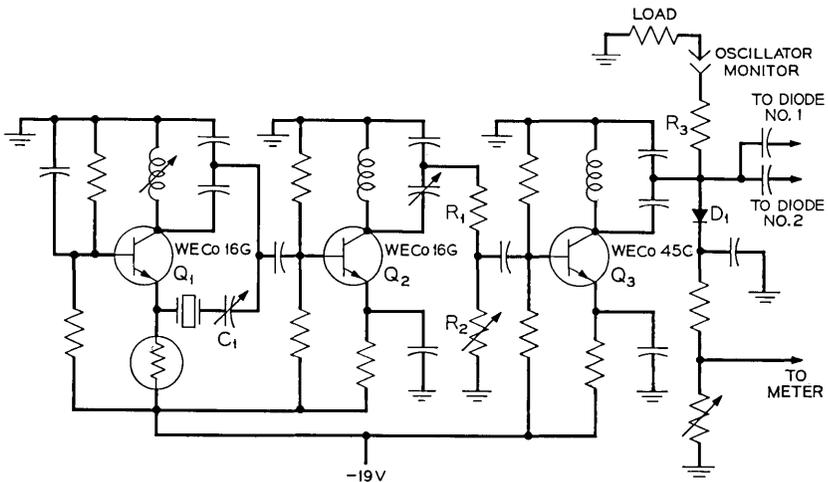


Fig. 10—40 MHz shift oscillator.

The total current drain for the circuit is about 55 mA from a  $-19V$  regulated supply.

#### V. ACKNOWLEDGMENTS

The development of the TD-3 transmitter and shift modulators was the combined accomplishment of many individuals. Specifically, the authors wish to acknowledge the contributions of P. I. Sandsmark and R. W. Judkins for the basic modulator circuit design, of S. H. Lee for the IF driver amplified design, and of O. Giust for the 40 MHz oscillator design.

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# Microwave Generator

By T. A. ABELE and D. J. LEONARD

(Manuscript received January 19, 1968)

*The TD-3 microwave generator furnishes all of the local oscillator power used in a TD-3 transmitter-receiver bay. It delivers an output power of 0.5W at a frequency which can be adjusted to any one of 17 frequencies. In designing the generator special emphasis was placed on reliability, ease of tuning, and on obtaining a carrier with high spectral purity and frequency stability. This paper describes the generator's circuit configuration and operation, and discusses important design considerations and techniques.*

## I. INTRODUCTION

The microwave generator is a highly frequency-stable, low-noise source of microwave power which is the local oscillator supply for the TD-3 microwave transmitters and receivers. It delivers 0.5W output power and is adjustable to any one of 17 different frequencies: 3780, 3800 . . . 4100 MHz.\*

To achieve great reliability, minimum maintenance, and low power consumption, it was decided to use only solid state devices. Various devices were considered: varactors, step recovery diodes, and transistors; however, at design time, varactors were the only solid state devices that appeared able to operate at the required frequencies and provide the power levels and reliability demanded.

Once varactors were chosen it was possible to determine the over-all circuit configuration. A crystal oscillator operating in the range of 125 MHz† was needed to obtain the required frequency stability. Then a chain, consisting of a cascade of several varactor multipliers, was used to obtain efficient frequency multiplication up to 4000 MHz. A power amplifier had to be inserted between the oscillator and the

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\* This equipment is manufactured for Bell System use only.

† In general 125, 500, 1000, and 4000 MHz are used to denote any frequency between 118.125 and 128.125 MHz and their fourth, eighth, or thirty-second multiple, respectively.

multiplier chain to drive the passive, lossy multipliers with sufficient input power to obtain the desired output power at 4000 MHz.

A significant aspect in the development of this generator was the need for a compromise between the usual need for efficiently obtaining a high level carrier, and the need to generate a carrier of high spectral purity; that is, with extremely low noise sidebands. These two requirements led to a rather extensive use of experimental design techniques, because available theory precluded exact design procedures, particularly with the multiplier chain. Instead, the design was based on simple principles which were experimentally refined. For example, even the expected efficiencies of the multipliers could not easily be calculated because the required power levels necessitate driving the varactors into the forward charge storage region, the voltage-charge relationships of the varactors do not follow a simple law, and the series resistance of the varactors is a complicated nonlinear function of charge.<sup>1</sup> In addition, a noise model was not obtainable. Therefore, although it would have been possible to extend Burckhardt's<sup>2</sup> work regarding efficiencies to this situation, an experimental approach was deemed the most satisfactory solution.

## II. DESIGN OBJECTIVES

Since the microwave generator furnishes all of the local oscillator microwave power to operate a TD-3 repeater, and since failure would cause loss of a radio channel, high reliability is a prime objective. It is attained by using solid state devices which should insure a generator failure rate of less than  $8.7 \times 10^{-3}$  failures per repeater year.

All performance objectives discussed in this section should be met over the normal operating temperature range of  $75 \pm 10^\circ\text{F}$ . However, the generator must continue to operate (possibly with degraded performance) from 40 to  $140^\circ\text{F}$  to assure system operation in case the air-conditioning malfunctions.

The other performance objectives are:

*Output Power:* 26.5 dBm  $\pm 1$  dB.

*Frequency:* The generator should be able to be tuned to each of the 17 frequencies (3780, 3800  $\cdots$  4100 MHz) by choosing the proper crystal and by tuning the oscillator, the amplifier, and the multiplier chain.

*Frequency Stability:* The inherent stability should be  $\pm 1$  part in  $10^6$  between maintenance intervals.

*Noise:* The noise objective was set to provide less than 1 dB con-

tribution from the microwave generator to the repeater noise figure. This objective, expressed in carrier-to-noise-per-cycle ratio, is shown on Fig. 1 as a function of frequency increment  $\Delta f$  away from  $f_c$ .\*

*DC Requirements:* The generator should be designed to operate from the  $-19\text{V}$  repeater power supply.

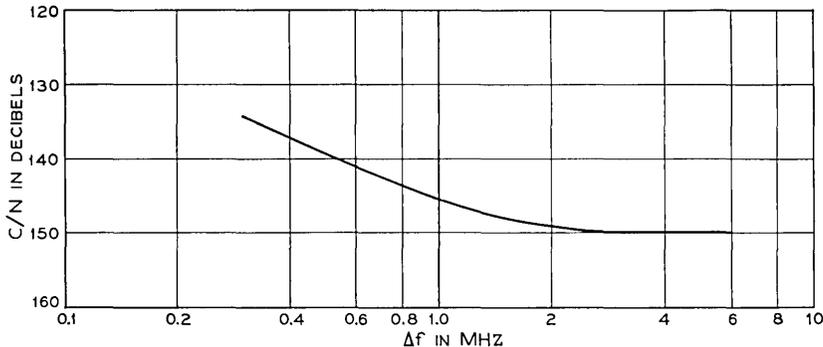


Fig. 1—C/N objective for the output carrier.  $C$  = carrier output power in dBm;  $N$  = total noise power in a 1 Hz band in dBm;  $C/N$  in dB =  $C - N$ ;  $\Delta f$  = frequency increment away from  $f_c$ ;  $f_c$  = any one of the 17 output frequencies (3780, 3800 . . . 4100 MHz).

*Miscellaneous: Instant start:* The generator should operate on the application of dc power; that is, with a properly tuned generator, removal and restoral of dc power should not require retuning of any circuits to obtain the original output. **Monitor points:** Test points should connect to the main metering panel ( $20 \mu\text{A}$ ,  $5900\Omega$  meter resistance) in order to monitor the condition of the generator without removing it from service. There should be a frequency monitoring point. **Output port:** The output port should be a coaxial  $N$ -type connector.

All of these objectives were met by the final design. The performance characteristics are discussed in detail in Section V.

### III. GENERAL DESIGN

#### 3.1 Block Diagram

Figure 2 is a block diagram of the microwave generator. It consists of two major parts: an oscillator-amplifier and a frequency multiplier chain. The oscillator-amplifier consists of a crystal oscillator and a power amplifier, and the multiplier chain consists of a first quadrupler,

\*  $f_c$  is any one of the 17 carrier frequencies just mentioned.

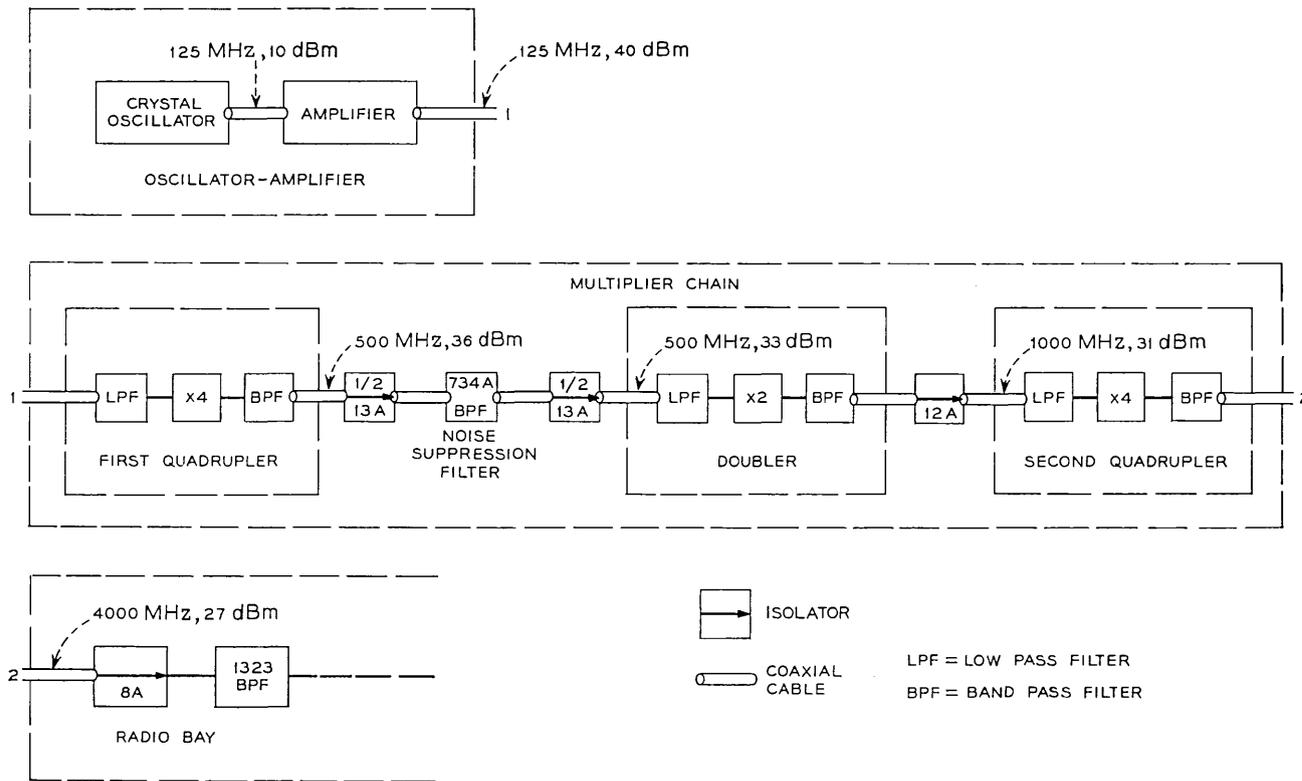


Fig. 2 — Block diagram of the microwave generator.

a noise suppression filter, a doubler and a second quadrupler. To facilitate development and testing, all of the units are designed to operate between  $50\Omega$  impedances.

The oscillator is a crystal-controlled transistor oscillator which operates between 118.125 and 128.125 MHz by choice of crystal and proper tuning. It delivers at least 10 dBm of power to the amplifier whose input impedance is  $50\Omega$  ( $\geq 20$  dB return loss). There is a point for monitoring the transistor's emitter current.

The power amplifier has four transistor stages: two class A stages and two class C stages. It is tunable over the same frequency range as the oscillator and can deliver at least 40 dBm of power to the first quadrupler. The output power can be varied from 38.5 to 40 dBm by adjusting an interstage network. This range of output levels is required since the generator must have an output power of about 26.5 dBm and the multiplier chain exhibits a loss ranging from 12 to 13.5 dB. There are points for monitoring frequency, output power level, and individual transistor currents.

The first quadrupler and the doubler are lumped element varactor multipliers; the second quadrupler is a distributed element varactor multiplier. All multipliers are self-biased and are shunt multipliers in order to facilitate better heat sinking. The output level of the chain is 26.5 dBm with input levels ranging from 38.5 to 40 dBm.

A waveguide bandpass filter is used at the output of the chain to eliminate spurious tones at frequencies  $\pm 125$ ,  $\pm 250$ ,  $\pm 375 \dots$  MHz removed from the desired 4000 MHz output. This filter is located in the radio bay (see Fig. 2).

### 3.2 *Physical Characteristics*

Figure 3 shows the generator, which measures 21 by 11 by 7 inches. Figure 4 shows the oscillator-amplifier part, and Fig. 5 shows the three multipliers. As Fig. 2 indicates, the components are connected by coaxial cable. The use of this flexible  $50\Omega$  cable (RG214/U or RG98/U), simplifies equipment layout and permits the easy disconnection needed for tuning the chain.

### 3.3 *Multiplier Chain Design*

#### 3.3.1 *General*

For the general layout of the multiplier chain, three empirical observations were most important. They concerned interaction, noise, and impedance.

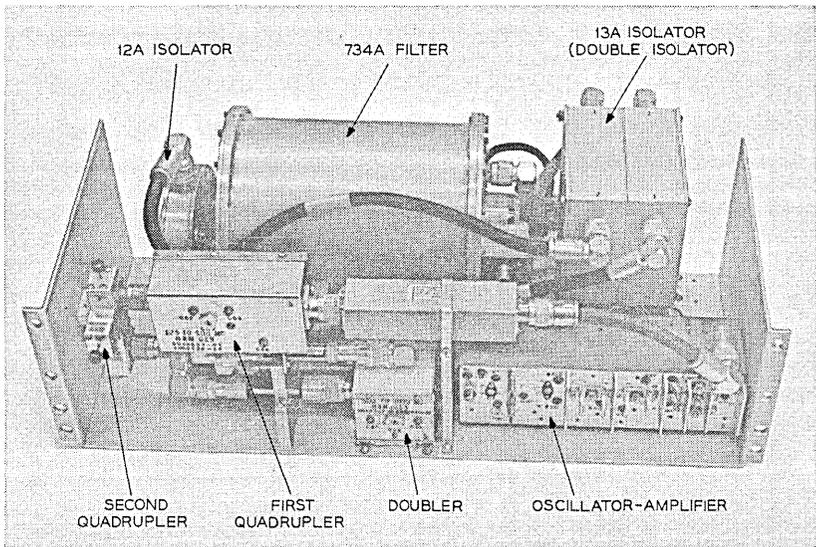


Fig. 3— Complete generator.

It was found that in cascading multipliers it is important to prevent interaction among the multipliers at undesired harmonic frequencies generated in the chain. Therefore, each multiplier was imbedded between filters which pass only the desired frequency at the input and the output.

It was verified that all multipliers behave like ideal multipliers of a multiplication factor  $m$ . Therefore, they amplify the FM correlated part of the noise power spectrum of the input carrier by a factor, expressed in dB, of  $20 \log m$ .<sup>3</sup> However, an additional noise component which existed even when the input carrier was essentially free of noise was observed in the noise power spectrum of the output carrier. It was concluded that this additional noise component originates in the multiplier.

When incorporating a properly operating multiplier in a chain it was found that, in order to keep the FM correlated noise originating in the multiplier consistently low and to ensure, when tuning the multiplier in the chain, that this condition remains coincident with that of maximum carrier output power, the multiplier must be imbedded between operating impedances which are not reactive at the carrier frequency or at frequencies close enough so that the band-

widths of the impedances facing the multiplier diode are appreciably reduced.

Therefore, a multiplier is not suitable as a load impedance for another multiplier, since it is reactive at the carrier frequency as long as it does not operate and could, therefore, cause starting and tuning problems. A narrow band filter is also not suitable either at the input or output of a multiplier. When placed at the output the effective bandwidth of the multiplier output circuit becomes much narrower than that of its input circuit. As Ref. 4 leads one to expect, this was found to produce instabilities even under perfectly tuned conditions. When placed at the input of a multiplier, a narrow band filter causes instabilities when the multiplier or the filter are being tuned, or when temperature and humidity changes cause the filter to become slightly detuned during operation.

The observation of multiplier interaction led to the inclusion of a low pass filter at the input of each multiplier and a band pass filter at the output of each multiplier (see Fig. 2). The filters are directly connected to the multipliers to avoid long line effects at the rejected

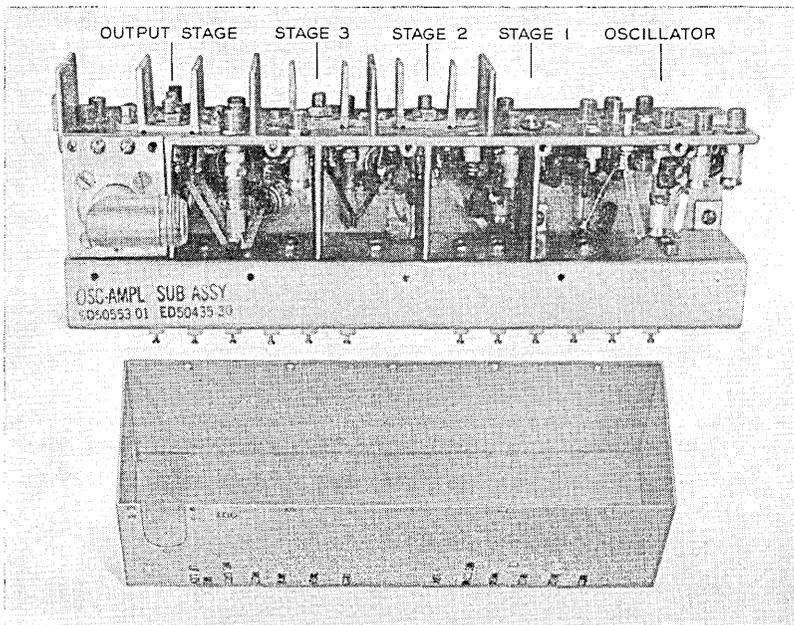


Fig. 4 — Oscillator-amplifier.

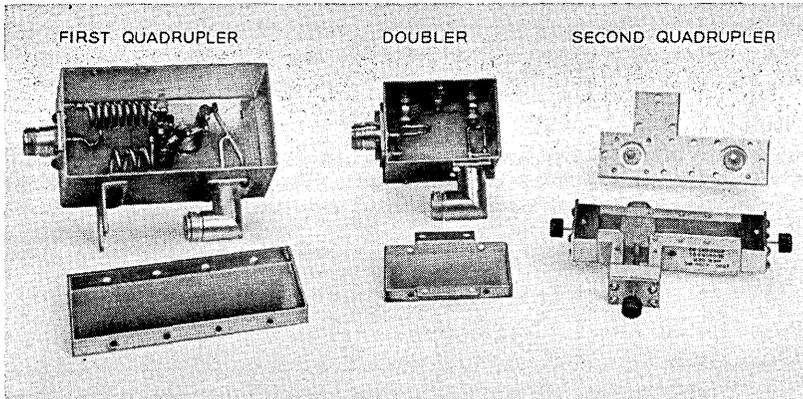


Fig. 5—The three frequency multipliers.

frequencies. All are commercially available filters designed to operate between  $50\Omega$  impedances.

The internally generated noise determined the optimum placement of a noise suppression filter within the chain. This filter is a narrow bandpass filter which, as Section 3.3.2 shows, is required to sufficiently increase the ratio between the carrier level at the output of the chain and the level of the FM correlated part of the noise power spectrum around the carrier at the output of the chain. If the multiplier chain consisted of ideal multipliers which only amplified the input noise, the filter could be placed at 125, 500, 1000, or 4000 MHz, provided that in all cases it furnished the same discrimination at a frequency increment  $\Delta f$  away from the respective carrier frequency  $f_c$ . Therefore  $Q_L$ , the loaded  $Q$  of the filter, would have to be proportional to  $f_c$ . Hence, in the case of ideal multipliers, the filter should be placed at 125 MHz because: (i) assuming that  $Q_I$ , the intrinsic  $Q$  of the resonator, is proportional to  $1/(f)^{1/2}$ , the midband insertion loss of the filter would be lowest, and (ii) the variation in insertion loss at  $f_c$  resulting from a given temperature variation would be smallest.

In reality, however, the multiplier chain does not consist of ideal multipliers; some noise originates in the multipliers themselves. It is obvious that if this contribution were dominating, the filter should be placed at 4000 MHz. After a number of measurements (described Section 3.3.2) it was decided that placing the filter at 500 MHz was the best compromise for this chain.

The impedance phenomena led to the inclusion of three isolators (realized by terminated circulators). One each was put at the input

and the output of the noise suppression filter, and one between the doubler and the second quadrupler. An additional isolator is actually connected to the chain at its output in the radio bay to guarantee the proper load impedance for the second quadrupler (see Fig. 2). It was found that an isolator is not required between the amplifier and the first quadrupler (in accordance with Ref. 4).

3.3.2 Noise Suppression Filter

Starting with the  $C/N$  objective (Fig. 1) at the generator output, the maximum permissible noise levels  $N_{IN}$  at the various multiplier inputs and at the input to the amplifier can be calculated under the assumption that in each case all transmission components between the respective input and the output of the generator contain no noise sources. This has been done in Fig. 6 for  $\Delta f = 2$  MHz using measured values for the carrier levels and the  $C/N$  degradations of the multipliers and amplifier.

Notice that the measured  $C/N$  degradations are 13 dB for the quadruplers and 7 dB for the doubler; that is, 1 dB higher than the theoretical value of 6 dB per frequency doubling. The increment frequency  $\Delta f = 2$  MHz was chosen for the calculations in Fig. 6 because (i) the  $C/N$  at the output of the oscillator, as pointed out in Section 5.2.3, is essentially constant in the band from 0.3 to 10 MHz, and (ii) the  $C/N$  objective (Fig. 1) decreases at approximately 6 dB per octave for frequencies below 2 MHz. Consequently, a chain containing a one-cavity noise suppression filter which meets the  $C/N$  objective at  $\Delta f = 2$  MHz will meet the  $C/N$  objective for all  $\Delta f$ .

It follows from Fig. 6 that very severe requirements would have to be placed on the individual units. For example, the oscillator must

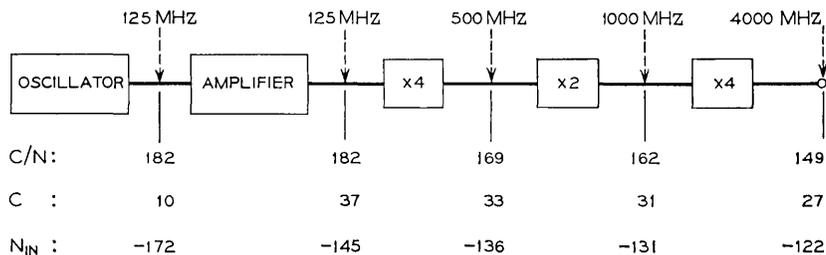


Fig. 6—Maximum permissible  $N$  at the input of the various multipliers and the amplifier (without noise suppression filter).  $C/N$  = maximum permissible  $C/N$  in dB;  $C$  = carrier level in dBm;  $N_{IN}$  = maximum permissible  $N$  in dBm. Carrier levels and  $C/N$  degradations are based on measurements.  $\Delta f = 2$  MHz.

have noise sidebands which are only 2 dB above thermal noise (-174 dBm) and the amplifier and multipliers must have no internal noise sources. It was therefore concluded that a noise suppression filter is required. Although it is generally most desirable to place the noise suppression filter at the lowest frequency point, that is, 125 MHz, it is not possible because of the internal multiplier noise.

The noise contributed by the multipliers ranges from 25 to 35 dB above thermal noise when referred to the input of the respective multiplier. These noise contributions were calculated from measurements made on the multipliers with calibrated noise sources and filters of known characteristics at the input of each multiplier stage (125, 500, and 1000 MHz). Also, the noise contributed by the amplifier is 11 dB above thermal noise.

Figure 7 shows the effect of these internal noise sources separately and combined. It can be seen that filtering the noise at the 125 MHz point would not improve the output  $C/N$  because the -139 dBm of internal noise contribution of the first quadrupler alone already results in -130 dBm at the input of the doubler which is 6 dB more than the -136 dBm requirement at this point. It can also be seen that the internal noise contributed by the doubler is sufficiently below that required at the doubler input to meet the output objective, and filtering would be effective at this point. Although the filtering could be done at higher frequencies, it was decided to place the filter at the 500 MHz point because of the arguments presented in Section 3.3.1.

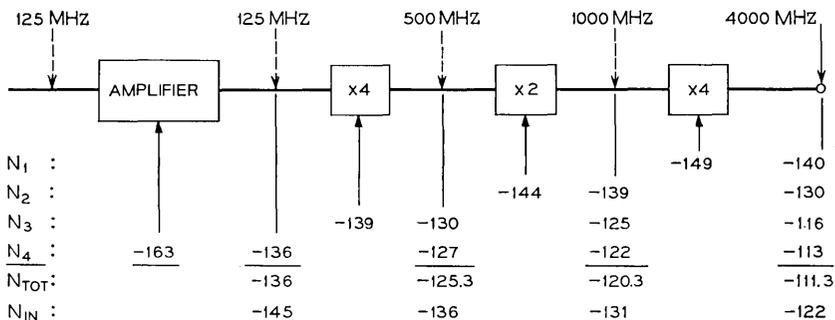


Fig. 7—Effect of internal noise contributions.  $N$  in dBm contributed by the internal noise of:  $N_1$ , the second quadrupler;  $N_2$ , the doubler;  $N_3$ , the first quadrupler;  $N_4$ , the amplifier.  $N_{TOT}$  is the total  $N$  contributed by the internal noise sources, and  $N_{IN}$  is the maximum permissible noise in dBm.  $C/N$  degradations are based on measurements.  $\Delta f = 2$  MHz.

Regarding the filter design, it was decided that a temperature compensated coaxial resonator could be developed which would have a 3 dB bandwidth of 300 kHz. This filter would provide about 20 dB of loss at  $\Delta f = 2$  MHz.

Figure 8 shows the maximum permissible noise levels at the input of each of the units with the noise suppression filter located at 500 MHz. It can be seen that the  $N_{IN}$  for the amplifier and first quadrupler inputs are now much more reasonable than those in Fig. 6.

### 3.4 Generator Tuning

The entire generator can be tuned to any one of the 17 different frequencies by selecting the appropriate crystal and tuning the various resonant circuits. The major criterion for correct tuning is maximum power output. That is, the oscillator resonant circuits are tuned for maximum power into a 50Ω load.

For tuning the amplifier, swept frequency techniques have proven to be very useful. Therefore maximum power output coupled with a smooth resonant response is the criterion for proper tuning. The input return loss of the amplifier is maximized at the frequency of interest and the interstage tuning controls are then adjusted for smooth response and maximum power at the desired frequency.

Swept tuning techniques were also used in the development of the multiplier chain. However, once the final circuit configuration was obtained, the adjustment for maximum power resulted in the proper response. The multiplier chain is tuned successively for the first quadrupler, the noise suppression filter, the doubler, and the last quadrupler.

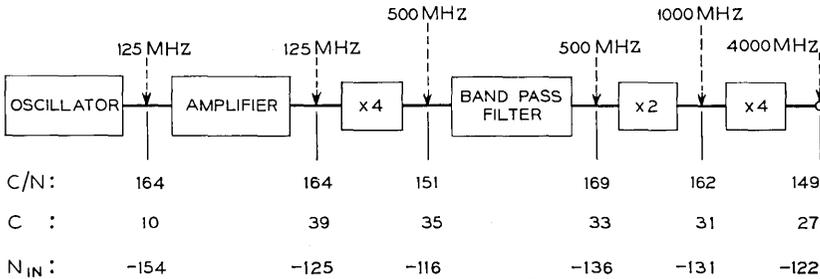


Fig. 8—Maximum permissible  $N$  at the input of the various multipliers and the amplifier (with noise suppression filter).  $C/N$  = maximum permissible  $C/N$  in dB;  $C$  = carrier level in dBm;  $N_{in}$  = maximum permissible  $N$  in dBm. Carrier levels and  $C/N$  degradations are based on measurements.  $\Delta f = 2$  MHz.

## IV. DETAILED ELECTRICAL DESIGN

4.1 *Oscillator*

The oscillator shown schematically in Fig. 9 is a modified Butler circuit<sup>5</sup> which was suggested by W. L. Smith of the Allentown Laboratory. The transistor is a Western Electric Company 45A transistor which is described elsewhere in the issue.<sup>6</sup> The transistor operates as an amplifier which drives a tuned collector load comprised of L3 in parallel with the total capacitive reactance seen by the collector. Part of this reactance, that from C2 and C3, is used to derive the feedback path which permits energy in the collector circuit to be fed back to the emitter through a series resonant crystal. The crystal is a Western Electric Co. 108 AB crystal which operates in the thickness mode at its fifth overtone. Another part of the collector reactance, from C4 and C5, is used to match the collector impedance to 50 ohms at the output. The resonant circuit (L1, C1) connected to the base is used as a fine tuning adjustment to pull the crystal frequency slightly. Inductor L2 is used to tune out the parasitic capacitance of the crystal. The necessary biasing resistors are not shown.

The oscillator delivers 10 dBm at its output. In order to achieve the  $\pm 1$  ppm frequency stability required by the generator only 2 mW of power can be dissipated in the crystal; hence the maximum output power was limited to 10 mW (10 dBm). If it were not for this requirement, the output power of the oscillator could have been increased, which, as experiments indicated, would have resulted in a better carrier to noise ratio.

4.2 *Amplifier*

The amplifier, shown schematically in Fig. 10, consists of two class A stages followed by two class C stages. The amplifier gain is

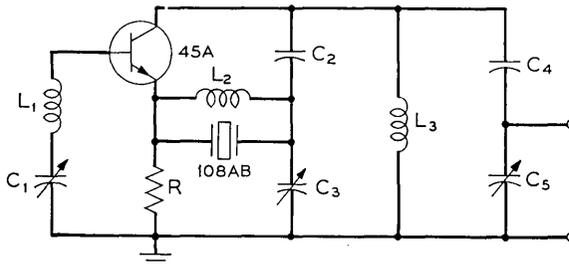


Fig. 9 — Oscillator circuit.

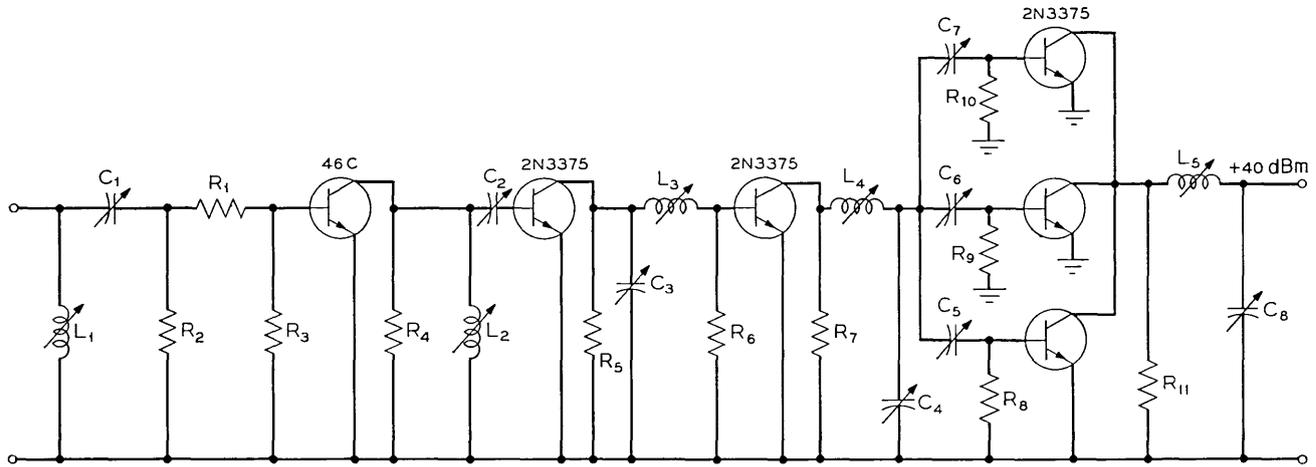


Fig. 10 — Amplifier circuit.

30 dB; that is, it can deliver 40 dBm into a  $50\Omega$  load when driven with a power of 10 dBm. It has an input return loss greater than 20 dB referred to  $50\Omega$ .

The first stage of the amplifier has an inherent gain of about 13 dB, but because of a lossy input network the net gain is about 10 dB and the output power is approximately 20 dBm. The second stage provides about 8 dB of gain and provides a level of 28 dBm to the input of the first class C stage. The first class C stage provides 6 dB of gain, thus delivering 34 dBm to the composite output stage. The output stage provides another 6 dB of gain which permits the amplifier to deliver 40 dBm to the multiplier chain.

The combined amplifier and oscillator operates at approximately 35 to 40 per cent efficiency; that is, it requires 25 to 30 W dc input power to obtain 10 W of 125 MHz power at the output.

The oscillator requires the amplifier to have a constant resistive input impedance in order to avoid interaction effects which generate noise, produce instabilities, and result in starting problems. It was assumed, and experimentally verified, that an input return loss of the amplifier of greater than 20 dB would be sufficient to provide satisfactory operation.

Since noise performance is very important, it was decided to operate the first stage at as low a noise figure and as high a gain as practicable. Common emitter biasing of a Western Electric Co. 46C transistor was chosen. The input impedance problem then consisted of matching the  $50\Omega$  oscillator output into the very low impedance of the first amplifier stage.

Because the input impedance of this stage depends considerably on transistor parameters, second stage input impedance, and tuning frequency, it was decided to buffer the input of the stage with a resistive pad (R1, R2, and R3). The transistor impedance of approximately  $Z = (10 + j50)\Omega$  was thereby transformed to an impedance in the range of  $Z = (25 + j10)\Omega$  at the pad input. It was then a relatively easy task to design a reactive transformer (L1 and C1) to transform from the pad impedance to  $50\Omega$  at the amplifier input.

Although more complicated lossless networks were investigated, this network gave a good compromise among noise performance, stability, sensitivity to transistor variations, ease of tuning, and ability to tune the oscillator and amplifier independently.

The interstage network (L2, C2) is designed to match the collector impedance of the first stage to the base impedance of the second stage. In order to simplify this network and to avoid stability prob-

lems as a function of tuning, the interstage has a resistor,  $R_4$ , which, in effect, lowers the interstage  $Q$  and places an upper bound on the collector load impedance at high frequencies; that is, at frequencies above the band of interest but below the  $f_t$  of the transistors. The networks ( $L_3$ - $C_3$ ,  $L_4$ - $C_4$ ) and resistors ( $R_5$ ,  $R_7$ ,  $R_{11}$ ) used between the remaining stages were designed in a similar manner.

Class A operation was chosen for the second stage. This arrangement minimizes the impedance variations which are reflected back to the input of the first stage, thereby minimizing the oscillator matching problem discussed previously. This stage uses an RCA 2N3375 npn transistor.

The third and fourth stages operate in the class C mode and use RCA 2N3375 npn transistors. In order to realize the desired power output with a power source of only  $-19$  volts, it was necessary to parallel three transistors in the output stage. It is desirable for each of the output transistors to operate at approximately the same level. The input impedance to each transistor is very low; even small variations in wiring inductance among the three inputs would be sufficient to result in unequal input levels. Feedback techniques to increase the input impedances would improve this but would be too wasteful of output power. Therefore, it was decided to independently feed each of the bases through tunable capacitors ( $C_5$ ,  $C_6$ , and  $C_7$ ).

The output network ( $L_5$ ,  $C_8$ ) is designed to match the collector impedance of the output stage to  $50\Omega$ . In addition, this network provides some rejection to the harmonics of the output frequency.

#### 4.3 Multiplier Chain

Figure 11 shows the detailed circuit of the first quadrupler. It consists of the varactor diode, a shunt resistor for self-bias, an input resonant circuit ( $L_1$ ,  $C_1$ ) tuned to 125 MHz, an output resonant circuit ( $L_3$ ,  $C_3$ ) tuned to 500 MHz, and an idler resonator ( $L_2$ ,  $C_2$ ) tuned to 250 MHz. The input is connected to a tap on  $L_1$ , and the output is connected to a tap on  $L_3$ , to provide  $50\Omega$  input and output impedances.

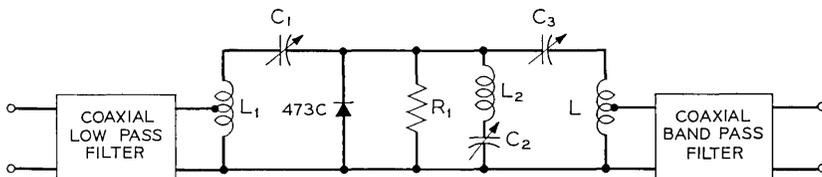


Fig. 11 — Circuit of the first quadrupler.

A low pass filter (LPF) is placed at the input to pass 125 MHz and to reject all other harmonics. A band pass filter (BPF) is placed at the output to pass 500 MHz and to reject all other harmonics. With an input power level of about 39 dBm from the oscillator amplifier the output power level is approximately 35 dBm. The varactor diode is a Western Electric Co. 473C diode, described in Ref. 1. With a typical breakdown voltage of 85V, a typical series resistance  $0.25\Omega$ , and a typical capacity at zero bias of 53 picofarads, this graded junction diode is driven into the forward charge storage region by the input power of approximately 8W.

The first quadrupler is followed by a coaxial isolator realized as a terminated circulator which provides approximately 20 dB of isolation (see Fig. 2). The forward loss is typically 0.25 dB. The next component in the chain is a Western Electric Co. 734A noise suppression filter (Fig. 2). It is a tunable temperature-compensated single cavity coaxial band pass filter with approximately 2 dB of loss at midband frequency and a bandwidth between the 3 dB points of approximately 300 kHz. The filter is followed by a second coaxial isolator realized as a terminated circulator. Both of these circulators, the one at the input and the one at the output of the 734A filter, are physically integrated into one package coded as a Western Electric Co. 13A circulator (see Fig. 2).

Figure 12 shows the detailed circuit of the doubler. It consists of the varactor diode, a shunt resistor for self-bias, an input resonant circuit (C4, L4) tuned to 500 MHz, and an output resonant circuit (L5, C5) tuned to 1000 MHz. The input is connected to a tap on L4 and the output is connected to a tap on L5 to provide 50 $\Omega$  input and output impedances.

A low pass filter is placed at the input to pass 500 MHz and to reject all other harmonics. A band pass filter is placed at the output to pass 1000 MHz and to reject all other harmonics. With an input power level of about 33 dBm, the output power level is approximately 31 dBm. The varactor diode is a Western Electric Co. 473B diode, described in Ref. 1. Also, this diode is driven into the forward charge storage region with an input power of approximately 2 W.

As shown in Fig. 2, the doubler is followed by another coaxial isolator realized as a terminated Western Electric Co. 12A circulator providing about 20 dB of isolation. The forward loss of this isolator is approximately 0.2 dB.

Figure 13 shows the detailed circuit of the second quadrupler. This quadrupler, which uses transmission line elements, consists of the

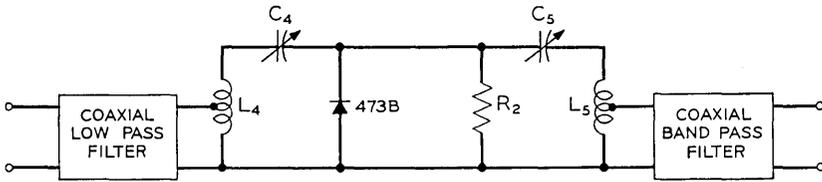


Fig. 12 — Doubler circuit.

varactor diode, a shunt resistor for self-bias, an input resonant circuit consisting of  $C_6$  and a shorted piece of coaxial transmission line ( $L_6$ ) tuned to 1000 MHz, an output resonant circuit consisting of  $C_8$  and a shorted piece of coaxial transmission line ( $L_8$ ) tuned to 4000 MHz, and an idler resonant circuit consisting of  $C_7$  and a shorted piece of coaxial transmission line ( $L_7$ ) tuned to 2000 MHz.

The input is tapped along  $L_6$ , and the output is tapped along  $L_8$  for impedance matching into  $50\Omega$ . A low pass filter (LPF 1) is placed at the input to pass 1000 MHz and to reject all other harmonics. In order to insure the proper phase of the 2000 MHz idler signal reflected by this filter, a suitable length of  $50\Omega$  line ( $L$ ) is inserted between the filter and the multiplier input.

A band pass filter, comprised of a low pass filter (LPF 2) and a high pass filter is connected to the output to pass 4000 MHz and to reject all other harmonics. With an input power level of about 31 dBm the output power level is approximately 26.5 dBm. The varactor diode is a Western Electric Co. 473A diode described in Ref. 1. This diode, too, is driven into the forward charge storage region with an input power of approximately 1.3 W.

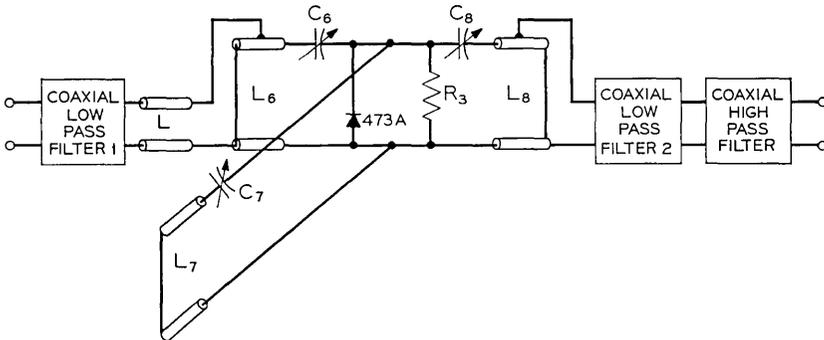


Fig. 13 — Circuit of the second quadrupler.

Since it was recognized that in all three multipliers the varactor diodes were driven into the forward charge storage region and therefore available design theory<sup>7</sup> could not be applied, the multipliers were designed by using simple circuit principles and making improvements experimentally. The filters used at the input and output of each multiplier are commercial units designed to operate between 50 $\Omega$  impedances.

## V. PERFORMANCE

### 5.1 *Reliability*

In the basic TD-3 System design it was established that a TD-3 microwave repeater could be permitted to have a maximum of 0.12 failure per year for the active devices based on the over-all TD-3 reliability requirements. About 20 per cent of this figure was allocated for the solid state devices and about 80 per cent for the traveling wave tube. Since there are approximately 200 solid state devices in TD-3 this corresponds to about 13.7 fits\* per device (2740 fits per bay).

Because this was considered too few for the power devices in the generator, the generator power devices were allocated 100 fit performance objectives and all others were allocated 10. Since only seven devices fall into the high power category the total number of fits per bay is virtually unchanged ( $7 \times 100 + 193 \times 10 = 2630$  fits per bay). The seven power devices and the three other devices result in 730 fits permitted for the generator's active devices.

The other passive devices will contribute an additional 270 fits since there are 135 components which should have 2 fit performance or better. This results in 1000 fit performance predicted for the generator which corresponds to 0.0087 failure per generator year.

Up to now there is nothing quantitative to report on the generator failure rate. There have been approximately 30 generator years logged in the laboratory and about 100 generator years logged in the field with no device failures. However, there has been one generator failure because of a defective capacitor. Nevertheless, based on the device reliability described in Refs. 1 and 6, and the reliability objectives specified for the passive components, it appears that the objectives will be met.

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\* A fit is a failure per 10<sup>9</sup> hours.

### 5.2 Noise Performance

As indicated in Section II, one of the main objectives is to obtain satisfactory noise performance between 0.3 and 6 MHz from  $f_c$  where  $f_c$  can be any one of the 17 frequencies between 3780 and 4100 MHz. Therefore, the noise very close to the carrier was of no interest and was not investigated.\*

The carrier to noise ratio was measured at various points in the generator. The direct measurements were made at the oscillator output (125 MHz), the amplifier output (125 MHz) and the multiplier chain output (4000 MHz).

#### 5.2.1 Measurement Method

The basic technique used to measure the carrier to noise ratio consists of measuring the carrier power with a standard power meter and then using the setup shown in Fig. 14 to remove the carrier and measure the noise on a calibrated spectrum analyzer.

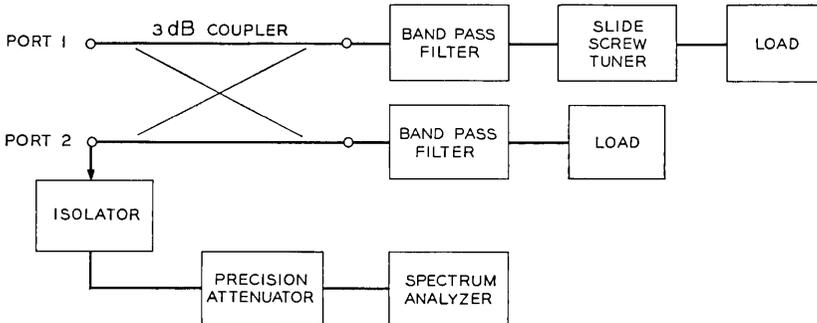


Fig. 14 — Noise measurement setup.

The equipment in Fig. 14 between ports 1 and 2 is a 3 dB coupler, two identical very narrow bandpass filters, and one fixed and one variable termination (fixed termination and slide screw tuner). By tuning both filters to  $f_c$  and adjusting the variable load, the insertion loss between port 1 and 2 can be made to behave like a narrow bandstop filter with infinite loss at  $f_c$ . The 3 dB points were  $\pm 0.1$  MHz for the 125 MHz bridge and  $\pm 0.25$  MHz for the 4000 MHz bridge.

The remainder of the setup, consisting of isolator, calibrated at-

\* For TV transmission, this statement is inaccurate. However, systems tests involving transmission of TV signals disclosed no impairment from microwave generator noise.

tenuator and spectrum analyzer, merely serves as an indicator. The purpose of the narrow bandstop filter (port 1 to port 2) is to prevent overloading the spectrum analyzer with the carrier power while passing the noise sidebands with virtually no attenuation. By using a calibrated attenuator, the noise at port 1 at any frequency  $\Delta f$  away from  $f_c$  can be compared with a known white noise at port 1 (noise lamp). Then by measuring the carrier level at port 1, the carrier to noise ratio at port 1 can be computed.

### 5.2.2 Over-all Noise Performance

Figure 15 shows the typical carrier to noise ratio for a complete microwave generator as a function of  $\Delta f$  about any 4000 MHz  $f_c$  as measured with the 4000 MHz bridge described above. The noise performance was measured for several generators at 40°, 75°, and 140°F, and it was found that temperature has virtually no influence.

### 5.2.3 Oscillator-Amplifier Noise Performance

The noise performance of the oscillator and amplifier was evaluated with the 125 MHz noise measurement bridge described in Section 5.2.1. The C/N at the oscillator output was 166 dB and was flat in the band between  $\pm 0.3$  and  $\pm 6$  MHz from  $f_c$ . It was observed that the C/N could be improved, if necessary, by increasing the carrier power output of the oscillator. This was not done, however, mainly because it was not necessary (a C/N of 164 dB would be adequate according to Fig. 8), but also because it was desirable to limit the power dissipated in the crystal to preserve the desired frequency stability.

The C/N at the amplifier output is shown in Fig. 16. The noise

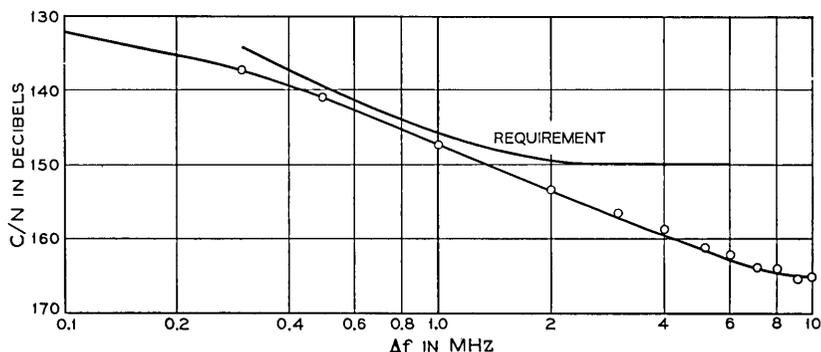


Fig. 15—Measured C/N—Performance at the output of the generator (4000 MHz).

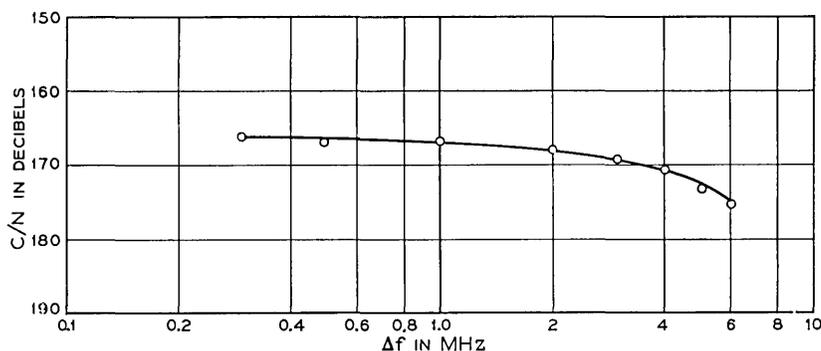


Fig. 16—Measured C/N—Performance at the output of the amplifier (125 MHz).

figure at the input to the amplifier's first transistor is approximately 8 dB, but, because the input network is lossy, the noise level at the amplifier input is approximately 11 dB. This corresponds to  $N = -163$  dBm. Based on the oscillator output C/N, the noise contributed from the oscillator is  $-156$  dBm. Therefore, the addition of the lossy input network did not contribute appreciably (0.6 dB) to the output C/N of the amplifier.

The improvement of C/N at frequencies between 1 and 6 MHz from  $f_c$  results from the band shaping in the amplifier. Again, this is not significant at this point because further improvements of the amplifier C/N would not necessarily realize an improved C/N at the generator output, since the internal noise of the multipliers would soon become controlling.

### 5.3 Output Power and Frequency Stability

Figure 17 shows the variation of the generator output power and frequency as a function of temperature. The design objectives were met over the  $75 \pm 10^\circ\text{F}$  temperature range. In fact, the output power level stays within the requirement over the extended range of 40 to  $140^\circ\text{F}$ . However, the frequency changes by about 5 ppm over the extended range. Both of these variations are acceptable from a system standpoint.

## VI. ACKNOWLEDGMENTS

The development of this generator was successful because of the work and cooperation of many of our colleagues at Bell Telephone Laboratories. The early development was done by H. W. Andrews,

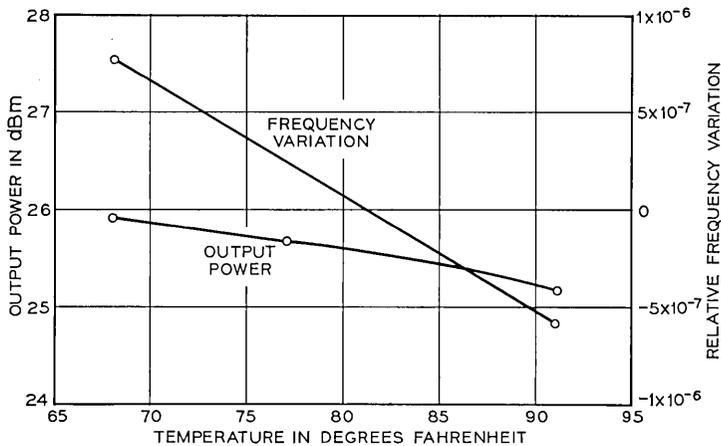


Fig. 17—Measured variation of the output power and output frequency of the generator as a function of temperature.

L. L. Gutman, and E. T. Harkless. We are indebted to W. D. Baker, L. L. Dale, J. A. Flynn, R. A. Svenson, and C. N. Tanga for their contributions. In addition, the Reading and Allentown Laboratories' help with crystals, transistors, and varactors is gratefully acknowledged. And we wish to thank J. T. Bangert for his encouragement and support.

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# Active Solid-State Devices

By H. E. ELDER

(Manuscript received October 5, 1967)

*This paper describes the electrical and physical characteristics of 20 new solid-state devices developed for TD-3. Among them are new high-frequency planar, epitaxial, NPN silicon transistors which are used in intermediate-frequency amplifiers and in the FM deviator. The microwave power for the transmitter and receiver is supplied from a three-stage varactor multiplier chain which uses three new epitaxial silicon varactor diodes. A high-quality epitaxial silicon varactor diode pair is the up-converter in the transmitter modulator. A new epitaxial gallium arsenide Schottky barrier mixer diode assures a low noise figure for the receiver modulator. Two point-contact silicon IF and RF monitor diodes were developed using the same miniature package in which the transmitter modulator and receiver modulator diodes are encapsulated. Two very stable diffused silicon diodes provide a stable output from the FM deviator; in-process preaging at accelerated conditions is used to assure satisfactory stability. Two new epitaxial silicon Schottky barrier diodes are used in the IF amplifier-limiter and discriminator circuits, which require a nonlinear resistance with negligible capacitance and recovery time. Stringent voltage breakdown, corona and mounting requirements for the high voltage power supply rectifiers were satisfied by molding two high-voltage diode rectifier assemblies in a high-dielectric plastic. The need for a pure variable resistance in the IF variolosses was satisfied by the development of a new silicon PIN diode.*

## INTRODUCTION

Twenty new solid-state devices were developed to meet the TD-3 requirements. The FM terminal transmitter needed a variable capacitor diode (457A) as the active element in the deviator, a very stable voltage regulator diode (446AC), and a transistor oscillator (44A). A family of medium-power varactors (473A, B, and C) was required for the microwave carrier supply. A high-quality silicon varactor diode pair (471A) was developed for the transmitter modulator. Two new

rectifiers (463A and 464A) were developed for the high voltage power supply for the traveling wave tube.

For the receiver, a low-noise gallium arsenide Schottky-barrier mixer diode (497A) and a low-noise IF preamplifier transistor (45J) were developed to assure the low-noise figure required of the TD-3 repeater. The second stage of the preamplifier required a transistor that could handle a large signal without harmful distortion (45G). The FM receiver also needed a transistor with controlled input impedance (45B) and a silicon Schottky-barrier diode (479A) in the IF amplifier-limiter, and a similar diode (479B) in the discriminator. The 45C transistor was developed for locations where the signal level exceeds the level at which the 45B transistor can maintain sufficiently low distortion. A silicon p-i-n diode (474A) was developed for the variolossler. Two silicon point contact diodes (488A and 493A) were developed for use in IF and RF monitor circuits.

#### I. TRANSISTORS By N. J. Chaplin and R. M. Jacobs

Concurrent with the conception and design of the TD-3 system, major advances in transistor technology occurred which have been exploited to reduce system noise, distortion, and power consumption, and to improve system reliability.

These advances resulted in the development of a family of high-frequency, planar, epitaxial, npn silicon transistors for use in the various intermediate-frequency amplifiers of the repeaters and terminals, and the FM terminal deviator. Typical values of important electrical characteristics of these transistors are given in Table I.

##### 1.1 General Description

The family of transistors developed for the TD-3 system are characterized by a 1 Ghz gain-bandwidth product and have been designed to yield optimum performance in the various applications. The 45A transistor, illustrated in Fig. 1, is the basic transistor of the family and is designed for operation at collector currents in the range of

TABLE I—TRANSISTORS FOR THE TD-3 SYSTEM

Parameter	Transistor					
	44A	45A	45B	45C	45G	45J
	Typical values					
Gain bandwidth product (mHz)	1000	1000	1000	1000	1050	1000
$NF$ (dB)	—	—	5.5	4.5	3.5	2.0
$C_{ob}$ (pF at 5 V)	1.2	0.9	1.2	1.9	3.5	4.0
$C_{ib}$ (pF at 1.5 V)	0.8	0.8	1.6	3.2	6.0	4.5

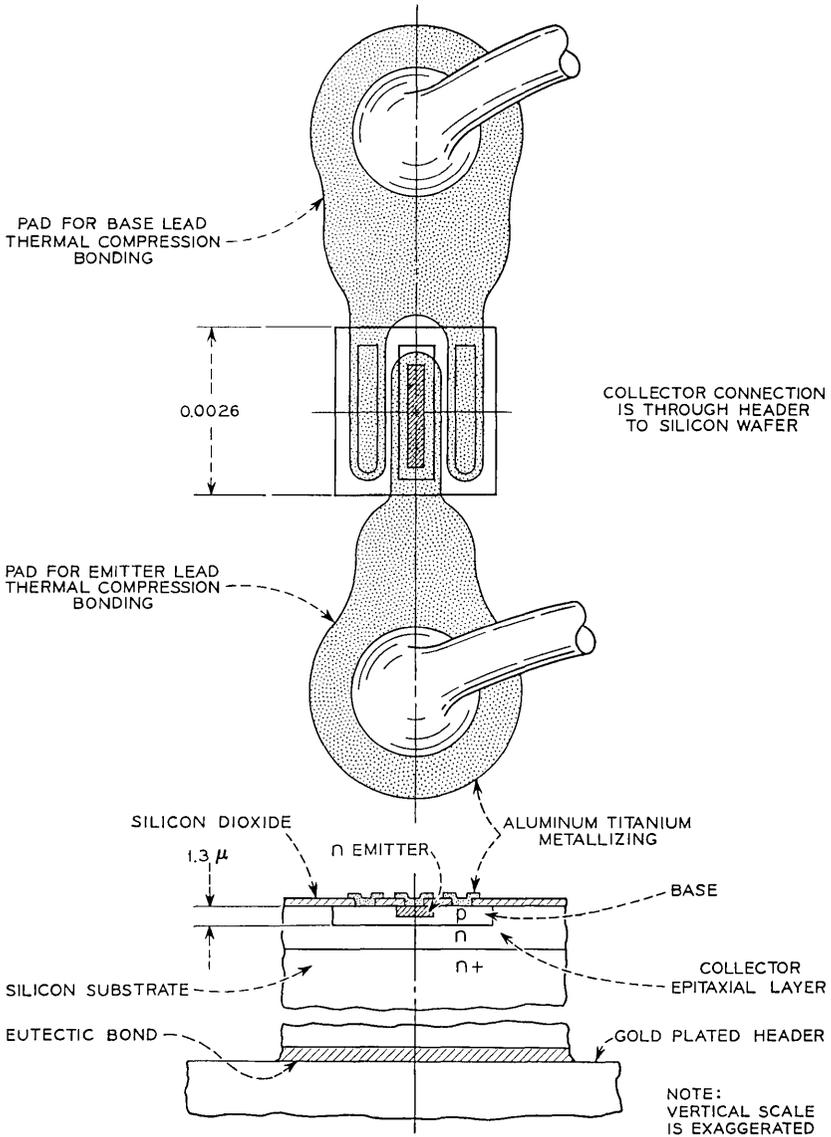


Fig. 1 — 45A transistor.

5 to 10 milliamperes. Other transistors of the family are fabricated by increasing the number of emitter stripes in order to increase the current handling ability and decrease the base resistance. The 45B, 45C and 45G transistors have twice, four and eight times the collector current rating and one half, one fourth and one eighth the base resistance of the 45A.

The transistor wafers are encapsulated in the metal-ceramic package illustrated in Fig. 2. This package was designed specifically for this transistor family to minimize parasitic capacitance, lead inductances, and to provide a relatively low thermal impedance. In addition, the package insulates the transistor collector from the heat sink. This package has been registered with the Joint Electron Device engineering Council and is designated TO-112. A transistor electrically similar to the 45A type has been encapsulated in a conventional TO-18 package for use in the FM deviator cavity and is designated the 44A. The collector of the 44A is connected directly to the deviator cavity through the metallic package. This method of connecting the collector to the cavity minimizes collector inductance and so maximizes the deviation range.

### 1.2 Main and Limiter IF Amplifiers

The stringent requirements of the TD-3 system IF amplifiers are discussed in Ref. 1. These requirements determine the critical tran-

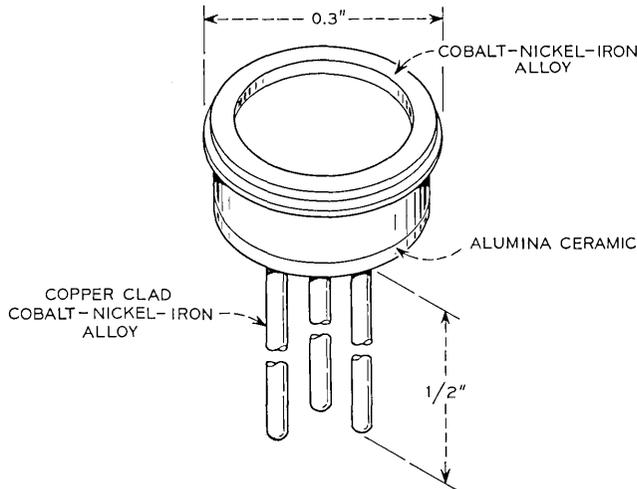


Fig. 2 — JEDEC TO-112 package.

sistor parameters, such as low noise figure, low output capacitance, controlled input impedance, high gain-bandwidth product, and the allowable distortion level.

To illustrate these parameter values, the 45B must have an output capacitance of less than 1.4 picofarads, a gain-band product of  $1000 \pm 200$  MHz, and common base input impedance at the operating frequency of not more than  $5 \pm j5$  ohms. The 45C transistor is used where the signal exceeds the level at which the 45B transistor can maintain sufficiently low distortion.

### 1.3 IF Preamplifier

As discussed elsewhere in this issue the low noise first stage of the preamplifier together with the low noise performance of the 497A Schottky-barrier diode have made it possible to meet the TD-3 system specifications for noise figure without radio frequency or parametric amplification.<sup>1</sup>

The needs of the preamplifier first stage were met by the development of a completely new transistor that has noise figures considerably lower than those previously attainable at the relatively high bias currents required for satisfactory TD-3 system performance. A bias current of 12 milliamperes is required to limit the distortion during the large signal condition of "up-fade" in short hop operation.

The 45J transistor was designed for this application. A typical noise figure of 2 dB at a 12 milliamperes bias current was achieved by decreasing the base resistance of the transistor to five ohms. This reduction in base resistance was achieved by reducing the emitter width and increasing the total emitter length.

This lowering of transistor base resistance results in a considerably lower noise figure as evident from Fig. 3.\* The new design also uses low resistivity diffused (five ohms per square) p type conductors in place of the usual metal base stripes. This structure permits covering the entire base-emitter inter-digitated structure with emitter metalizing. The metallized emitter contact is sufficiently large for wire bonding without an extension over the collector silicon and hence creates no parasitic collector-emitter capacity.

Figure 4 shows the structure of this transistor, designated 45J.

\* Figure 3 has been plotted using Nielson's equation and the assumptions shown. (See Ref. 2.) The equation is:

$$\text{N.F.} = 1 + \frac{r_e}{2R_g} + \frac{r_b'}{R_g} + \frac{(R_g + r_e + r_b')^2}{2\alpha_0 \cdot R_g \cdot r_e} \left[ 1/h_{FE} + \frac{I_{co}}{I_e} + \left( \frac{f}{f_\alpha} \right)^2 \right]$$

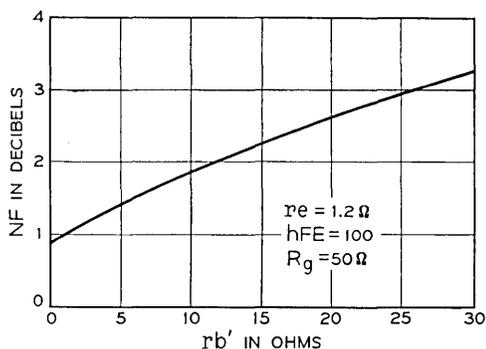


Fig. 3—Calculated noise variation ( $r_e = 1.2$  ohms,  $h_{FE} = 100$ ,  $R_G = 50$  ohms).

Table II compares characteristics of the 45J and 45G transistors. The lower noise and base resistance of the 45J are evident.

While the noise figure of the preamplifier comes largely from the first stage, a smaller contribution occurs from stage two. Because of the 17 decibel gain of the first stage, the 3.5 dB noise figure of the 45G is adequate for the second stage, but the bias current is increased to 38 milliamperes in order to handle the larger signal level without harmful distortion.

#### 1.4 Terminals

Transistor requirements for the intermediate-frequency amplifiers in the TD-3 terminals are similar to those discussed in Sections 1.2 and 1.3. It has been necessary to use an additional transistor, the 44A, in the oscillator cavities of the FM deviator.

The 44A transistor uses the wafer of the 45A transistor mounted in a standard TO-18 package. This package is conveniently mounted in the resonant cavity since the collector is electrically connected to the metal package as described in Section 1.1.

The FM deviator requires devices stable with respect to capacity variation for the lifetime of the device.<sup>3</sup> Transistor samples were subjected to accelerated aging tests which predicted typical output capacitance variations of 0.01 pfd or less over life. Since the deviator diode capacitances are many times larger than those of the transistors. The contribution of capacitance variation caused by the transistor is negligibly small.

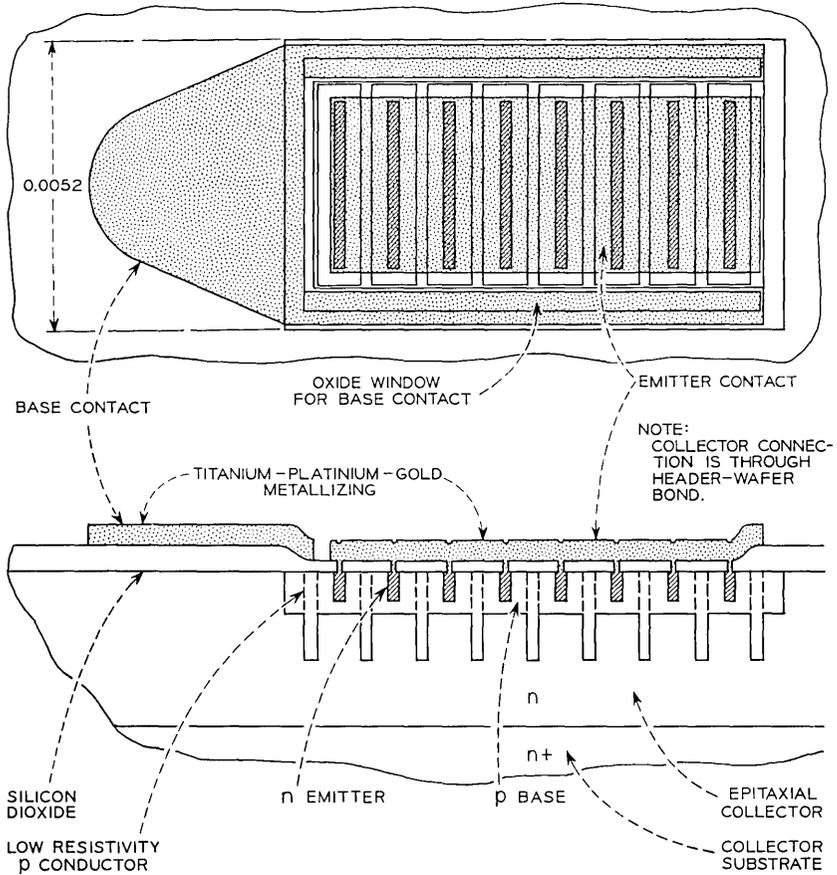


Fig. 4 — 45J transistor.

TABLE II—COMPARISON OF 45G AND 45J TRANSISTORS

Transistor	Typical values	
	45G	45J
Noise figure (dB)	3.5	2.0
Base resistance (ohms)	20	5

### 1.5 *Transistor Reliability*

During development of the 44 and 45 type transistors, several thousand models were subjected to accelerated aging under conditions of both power and temperature stress. Extrapolation of these data justifies a predicted transistor failure rate of less than ten failures per  $10^9$  transistor operating hours over a twenty-year period at junction temperatures of  $125^\circ\text{C}$  for all transistors except the 45J.

Similar data for the 45J transistor, which is fabricated with platinum-titanium-gold metallizing<sup>4</sup> predicts ten failures per  $10^9$  device-hours at  $250^\circ\text{C}$ .<sup>5</sup> Using this contact, junction operating temperatures are no longer limited by system reliability considerations, rather they are limited by the values of transistor parameters at such elevated temperatures.

The reliability information obtained through system use at this early date agrees with the extrapolated accelerated aging results.

### 1.6 *Conclusion*

44A, 45A, 45B, 45C and 45G devices have been developed and placed in manufacture to meet stringent requirements of the TD-3 system.

A vastly improved design in noise figure and reliability (45J) has been developed which, in combination with the Schottky diode, has permitted elimination of an expensive parametric amplifier or microwave preamplifier.

## II. HARMONIC GENERATOR VARACTOR DIODES FOR THE MICROWAVE CARRIER SUPPLY By F. W. Crigler and D. R. Decker

The microwave beat oscillator power for the microwave transmitter and microwave receiver of the TD-3 radio-relay system<sup>6</sup> is supplied by an all-solid-state circuit incorporating a crystal-controlled oscillator, a transistor amplifier, and a three-stage varactor-multiplier chain. Three codes of medium power, high efficiency varactor diodes were needed for this circuitry since there were no suitable diodes at the beginning of the TD-3 project. The 7-watt output\* of the transistor amplifier at 125 MHz is converted to about one-half-watt at 4 GHz by the varactor multipliers. The design requirements for the

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\* As described in Ref. 7, the output of the transistor amplifier is adjustable over the range of 7 to 10 watts.

beat oscillator supply (microwave generator) include adequate frequency stability and power capability, and very low noise.

### 2.1 *Circuit Requirements*

The three multiplier stages each contain a single varactor diode. The differences in operating frequency and power level between multiplier stages call for differences in diode characteristics. However, all three diodes must combine high power handling capability with efficiency and reliability.

A 473C diode in the first multiplier stage converts 7.0 watts from the transistor amplifier at 125 MHz to 2.8 watts at 500 MHz. The output frequency of the first stage is doubled to a nominal power of 1.2 watts at 1 GHz by a 473B diode multiplier circuit. The multiplier stages are decoupled from each other with isolators and frequency selective networks to minimize interaction during tuning and to eliminate undesired harmonics. A narrow band filter is used between the first and second multiplier stages to attenuate noise sidebands. Finally, the 1 GHz output is quadrupled to a minimum power of 400 mW at 4 GHz with a 473A diode circuit. The over-all efficiency of conversion from 125 MHz to 4 GHz is about 6 per cent. The efficiency includes all the losses of the interstage isolation and frequency selective circuitry that provide stable tunability and low noise performance.

### 2.2 *Diode Requirements*

The diode parameters that must be controlled to specified limits in order to assure desired multiplier performance include junction capacitance  $C_j$ , series resistance  $R_s$ , breakdown voltage  $BV$ , and thermal resistance  $\theta_{J-C}$ . An equivalent circuit of the 473 diode is shown in Fig. 5a. Since the construction of the wafer and package is such that the fringing capacitance  $C_f$  is small enough to be neglected, a simplified circuit may be used as shown in Fig. 5b. The total capacitance  $C_t$  includes the package capacitance  $C_c$  and the junction capacitance.

The diode junction capacitance is determined by two considerations. First, the capacitance must be large enough to allow the diode to handle the desired power without being driven into avalanche breakdown. Second, a choice of circuit characteristics determines what range of diode capacitance will provide high efficiency multiplier operation. Typically, the junction capacitance is chosen on the basis

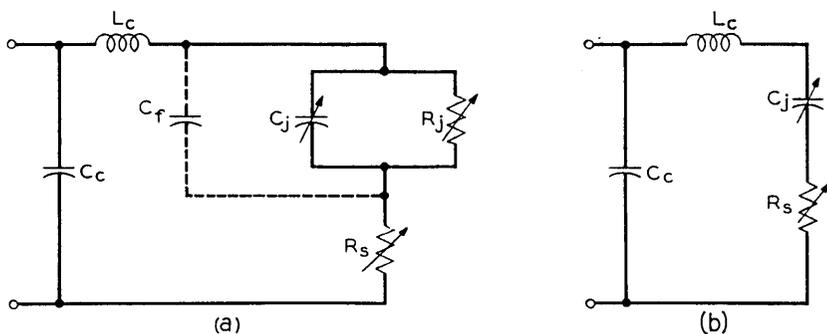


Fig. 5—(a) Diode equivalent circuit. (b) Simplified equivalent circuit.  $C_c$ , package capacitance;  $L_c$ , package inductance;  $C_f$ , distributed fringing capacitance;  $C_j$ , junction capacitance;  $R_j$ , junction resistance;  $R_s$ , diode series resistance.

of circuit design considerations, and the breakdown voltage is chosen to provide adequate power handling capability and good efficiency. Noisy performance may be encountered if the diodes are allowed to operate into the breakdown region because of the inherently noisy avalanche process.

The diode series resistance is the only lossy element of the varactor. For frequency multipliers it is desirable to make the series resistance as low as possible. For prescribed capacitance, the lowest usable breakdown voltage gives the minimum obtainable series resistance.

The power dissipation capabilities required for each of the 473A, B, and C diodes can be derived from the maximum input power and maximum conversion loss for each stage of multiplication as shown in Table III.

### 2.3 Diode Design and Construction

The 473A, B, and C harmonic-generator diodes are diffused, planar, epitaxial, silicon diodes in the V package. (Section 2.4 describes the V package.) All three device codes are manufactured by the same

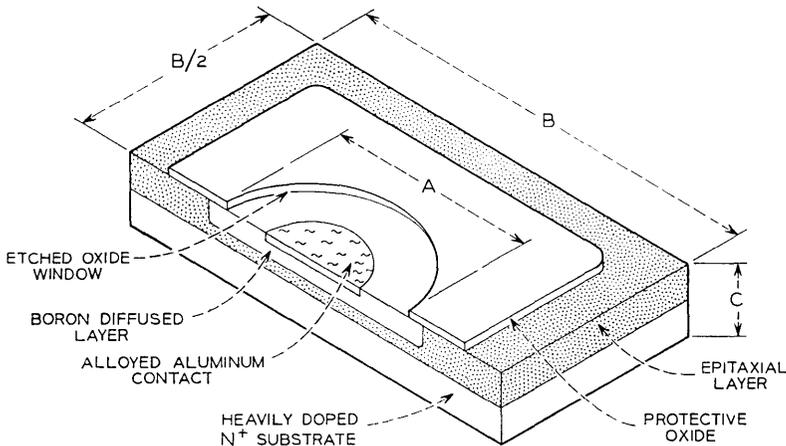
TABLE III—MULTIPLIER CONVERSION LOSS AND POWER DISSIPATION

Stage multiplier	Diode	Maximum diode conversion loss (dB)	Maximum power (W)	
			Input	Dissipated
125–500 MHz	473C	2.5	7.1	3.1
0.5–1.0 GHz	473B	2.0	2.0	0.8
1.0–4.0 GHz	473A	4.0	1.1	0.7

process. The differences in final device characteristics are controlled by choice of epitaxial starting material and by a difference in junction area. Figure 6 shows a cross-section of the planar 473 wafer.

The use of epitaxial silicon material for these devices is dictated by the need for a high breakdown voltage and a low series resistance. The epitaxial layer furnishes the lightly-doped region needed for high breakdown voltage while the heavily-doped substrate provides support and electrical contact for the lightly-doped layer without adding appreciably to the series resistance of the diode.

The intrinsic series resistance of an epitaxial diode is determined by the required epitaxial layer thickness and resistivity to sustain the breakdown voltage.<sup>8</sup> The parasitic resistance contributed by the substrate and package amounts to only 0.1 to 0.3 ohms. The breakdown voltage of an epitaxial diode is determined mainly by the epitaxial layer resistivity and the distance between the junction and the region of rapidly rising impurity concentration near the substrate. There is thus a compromise that must be made between a wide region



473	Dimension (inches)		
	A	B	C
	Window size	Wafer size	Wafer thickness
A	0.0072	0.020	0.0050-0.0070
B	0.0150	0.030	0.0050-0.0070
C	0.0284	0.040	0.0050-0.0070

Fig. 6 — Planar wafer cross section.

of high-resistivity material to sustain a high breakdown voltage and a narrow region of low-resistivity material to minimize the series resistance. The 473 design is an engineering compromise in this respect for the specified breakdown voltage.

The diode capacitance is governed mainly by the epitaxial layer resistivity and the junction area.<sup>9</sup> The 473A, B, and C diodes are designed by first tailoring the epitaxial material characteristics and the diffusion process to achieve the desired breakdown voltage with the least series resistance, and then adjusting the junction area to obtain the proper junction capacitance.

These diodes are fabricated by standard planar-photolithography.<sup>10</sup> An oxide film 8000 to 10,000 Å thick is first grown on the epitaxial silicon slice. Windows are etched in the oxide through a mask of photoresist, and boron is diffused into the slice to the desired junction depth. After diffusion, the windows are reopened and an alloyed aluminum contact is applied. The slice is scribed and broken into wafers, and the individual wafers are gold-silicon-eutectic bonded into a metal-ceramic V package. A gold tape is bonded by thermal compression to the wafer contact and to the package rim. The subassembly is vacuum baked before sealing the package by brazing on the end cap.

#### 2.4 *Device Characteristics*

The V package is one of several standard microwave diode packages used in the Bell System. (The "V" is simply a sequential designation assigned by Bell Laboratories.) Figure 7 shows a cutaway view. The primary characteristics of this package are low shunt capacitance, low series inductance, good power dissipation, and low cost. One figure of merit of a microwave diode package is the self-resonant frequency computed from the capacitance and inductance of the package. For this package, the self-resonant frequency is 15.9 GHz using 0.25 pF and 0.4 nH for the shunt-capacitance and series-inductance, respectively. The package contribution to power dissipation capability (thermal resistance) is dependent mainly on wafer bond area. For the 473A, B, and C, the package contribution is less than 30 per cent of the total thermal resistance.

The primary characteristics of the 473 type varactors are listed in Table IV. Figure 8 shows distributions of breakdown voltage for diodes meeting these characteristics. Since there is no upper limit on breakdown voltage, the spread of the distribution is mainly deter-

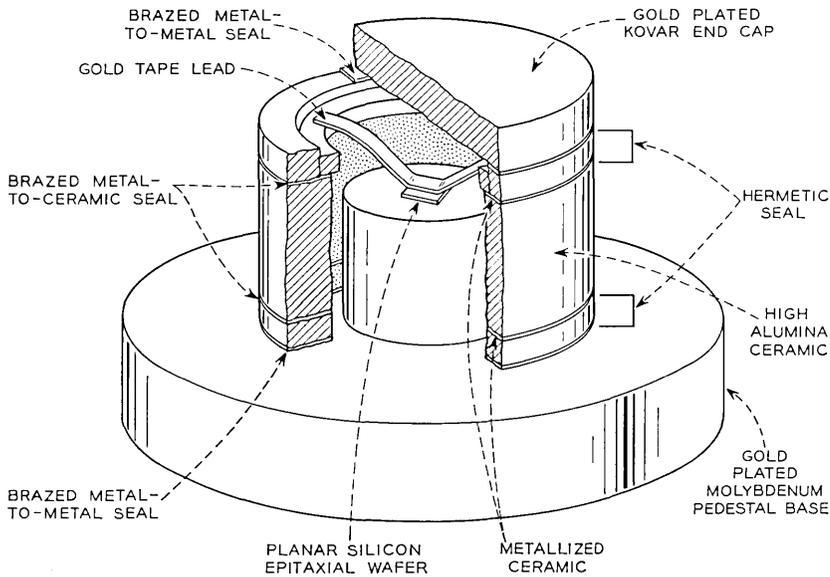


Fig. 7 — Cutaway view of V package.

mined by the capacitance and resistance limits for each code. As mentioned previously, these parameters are controlled by the characteristics of the epitaxial material and the diffusion profile. Typical dc forward and reverse-voltage characteristics are shown in Fig. 9a. High dc impedance is maintained in the operating region between breakdown voltage and forward conduction with a rapid rise in current occurring in the avalanche region for a small change in voltage. The forward voltage is exponential over many decades of current, producing a straight line plot on the log-linear scale shown.

In Fig. 9b are 473A junction capacitance and series resistance as functions of bias. The 473B and C diodes exhibit similar characteristics. The nonlinear capacitance-voltage characteristic shown is typical for a reverse-biased varactor. However, in the TD-3 genera-

TABLE IV—473 CHARACTERISTICS

	473A	473B	473C
$BV$ (V) at $I_r = 10 \mu A$	>60	>80	>70
$C_T$ (pF) at $V = 0, f = 1$ MHz	3.0 to 6.0	11.0 to 19.0	46.0 to 68.0
$R_s$ (ohms) at $V = 0, f = 1.3$ GHz	<1.5	<0.9	<0.5
$\theta_{J-C}$ ( $^{\circ}C/W$ )	<38	<22	<18

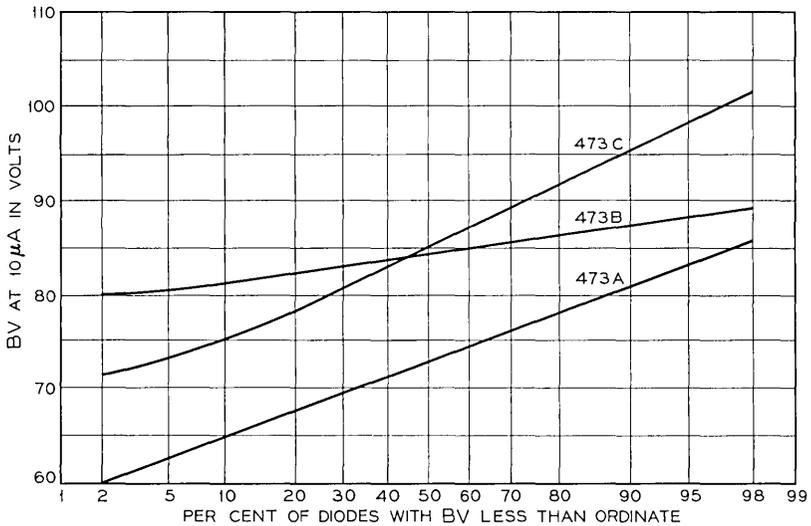


Fig. 8—Breakdown voltage.

tor, the devices are driven into forward bias as well, and minority carriers are injected across the junction. These carriers are stored during the forward cycle and contribute to a large injection capacitance. The relative importance of charge storage effects on circuit performance is being investigated.

Total capacitance distributions are shown in Fig. 10. The capacitance of the 473A diodes falls in the upper part of the allowable range as a result of making the junction area as large as possible to assure a low series resistance. Capacitance measurements at 1.3 GHz or higher would require inductive reactance corrections for the B and C diodes. This problem has been avoided by using a capacitance bridge at 1 MHz.

The series resistance is obtained from measurements of voltage standing-wave ratio at 1.3 GHz. Reflections from the diode, terminating a coaxial line, are measured with a slotted line. A solid coin-silver slug in the shape of the package (short) is used to determine line losses, and a package without the silicon wafer (open) is used to establish the reference plane of the diode. The diode series resistance is calculated by computer using the following equation:<sup>11</sup>

$$R_s = Z_0 \frac{S(1 + \cot^2 \beta d)}{S^2 + \cot^2 \beta d}$$

where:

$Z_0$  is the characteristic impedance of the transmission line.

$S$  is the voltage standing-wave ratio corrected for line and holder losses using the solid package short

$\beta$ , the phase constant of the line, is  $2\pi/\lambda$ , where  $\lambda$  is the wavelength  
 $d$  is the distance between the voltage minimum of the diode and a reference plane established at the voltage minimum for the open package.

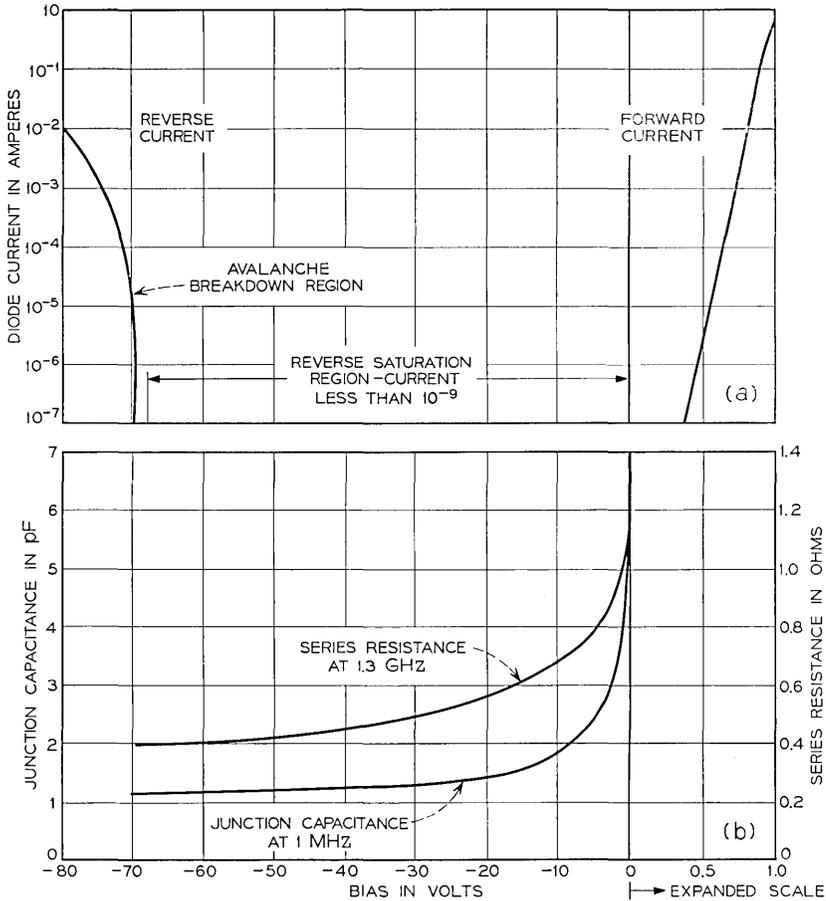


Fig. 9—(a) 473A varactor forward and reverse current as a function of bias voltage. (b) Varactor junction capacitance and series resistance as a function of bias voltage.

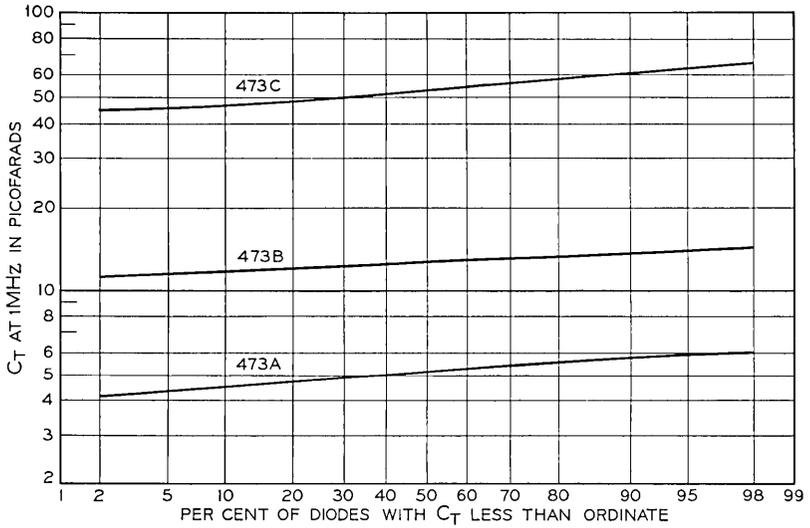


Fig. 10 — Diode total capacitance.

Figure 11 shows distributions of series resistance at zero bias. Typical values of resistance which are obtained at 100 mA forward current, on devices meeting the above specifications, are 0.25, 0.20, and 0.15 ohms for the 473A, B, and C diodes, respectively. These low resistances assure low diode losses during multiplier operation.

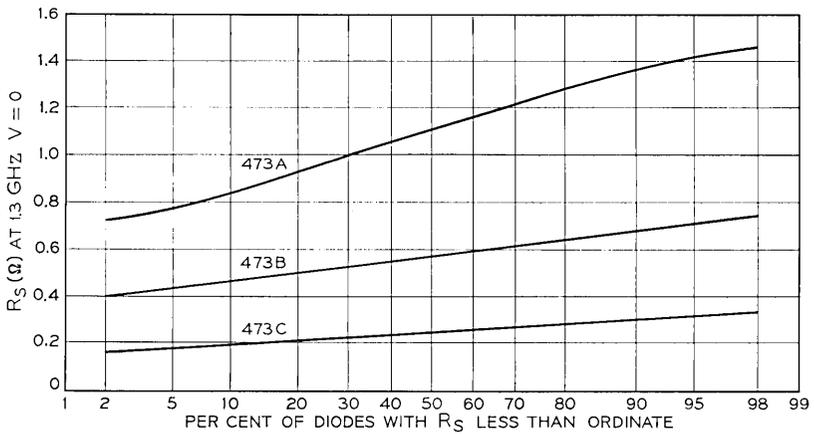


Fig. 11 — Diode series resistance.

Since relatively large power is dissipated by the 473 diodes, it is necessary to limit the diode thermal resistance to avoid degraded performance which would result from excessive device temperature during operation. Typical distributions of thermal resistance are shown in Fig. 12.

### 2.5 Reliability

Both temperature and power-aging experiments have been used throughout device development to evaluate designs. Analysis of temperature-step-stress aging<sup>12</sup> data on early developmental models led to the introduction of a more reliable wafer contact. High-temperature aging under reverse bias to evaluate the effectiveness of the oxide in stabilizing the surface of the wafer is a standard reliability test. Improved oxides have been obtained by using information from these tests and from metal oxide semiconductor measurements.

Manufacture of highly reliable devices requires constant attention to cleanliness in fabrication to assure a clean environment for the wafer. All devices are given a process screening at 220°C for 72 hours to weed out unreliable products. Temperature cycling is also used as a screening test. Encapsulations are tested for leaks in two steps. One uses a radioactive-source-tracer to detect very small leaks; the other

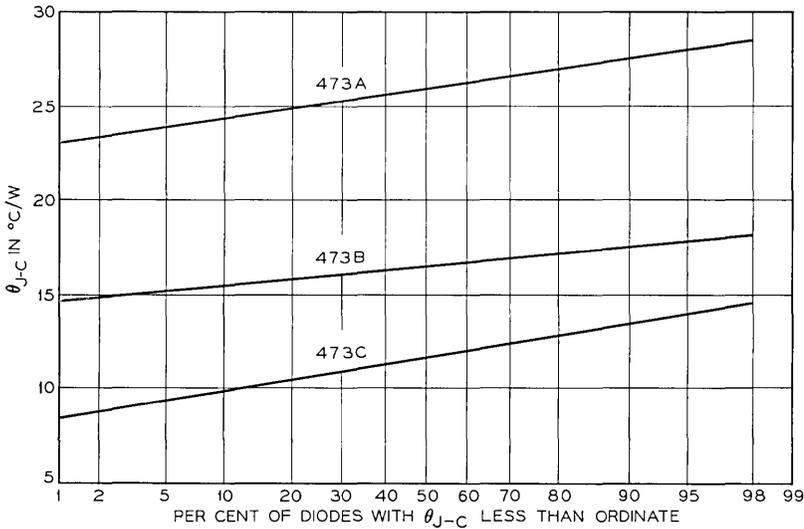


Fig. 12 — Thermal resistance.

TABLE V—100-HOUR POWER AGING REQUIREMENTS

Diode	Power (W)		$V_R$ (V)
	Rated	Aging	
473A	2.0	3.0	50
473B	3.0	4.5	70
473C	4.0	6.0	60

uses pressurized dye to indicate larger leaks; combined they assure a hermetically sealed package. All devices undergo a 16-hour test at 175 per cent of rated power to assure reliable operation in the multiplier circuit. A centrifuge test at 20,000G and a shock test at 1,500G complete the screening.

A sample lot of devices is required to pass 100-hour power aging tests as indicated in Table V. Diodes are switched between forward conduction and reverse-bias operation ( $V_R = BV - 10V$ ) at 60 Hz during aging. The forward current is adjusted to obtain the correct aging power. About 1 per cent of the diodes in a sample from a properly manufactured lot fail under the conditions of Table V.

### 2.6 Summary

The 473A, B, and C are medium power varactor diodes that are useful as harmonic generators in the range 0.1 to 10 GHz. The diodes are planar, passivated structures encapsulated in a miniature pill-type ceramic package. Process screening tests assure high quality. In the TD-3 microwave generator circuitry, these diodes generate the desired beat oscillator power while meeting additional requirements of low noise and tunability.

### III. U-PACKAGE MICROWAVE DIODES

By H. E. Elder, T. P. Lee, F. H. Levien, D. C. Redline and N. C. Vanderwal

The 471A transmitter modulator diode pair, the 488A IF monitor diode, the 493A RF monitor diode, and the 497A receiver modulator diode are encapsulated in the "U package." The "U" is a sequential designation assigned by Bell Laboratories.

The U package was designed to fill a need for a hermetic microwave package with low parasitic inductance and capacitance. Figure 13 shows a U-package silicon varactor. The other diodes which use the U package differ in semiconductor wafer material, processing, and internal contacts. The 488A and 493A diodes have internal tungsten

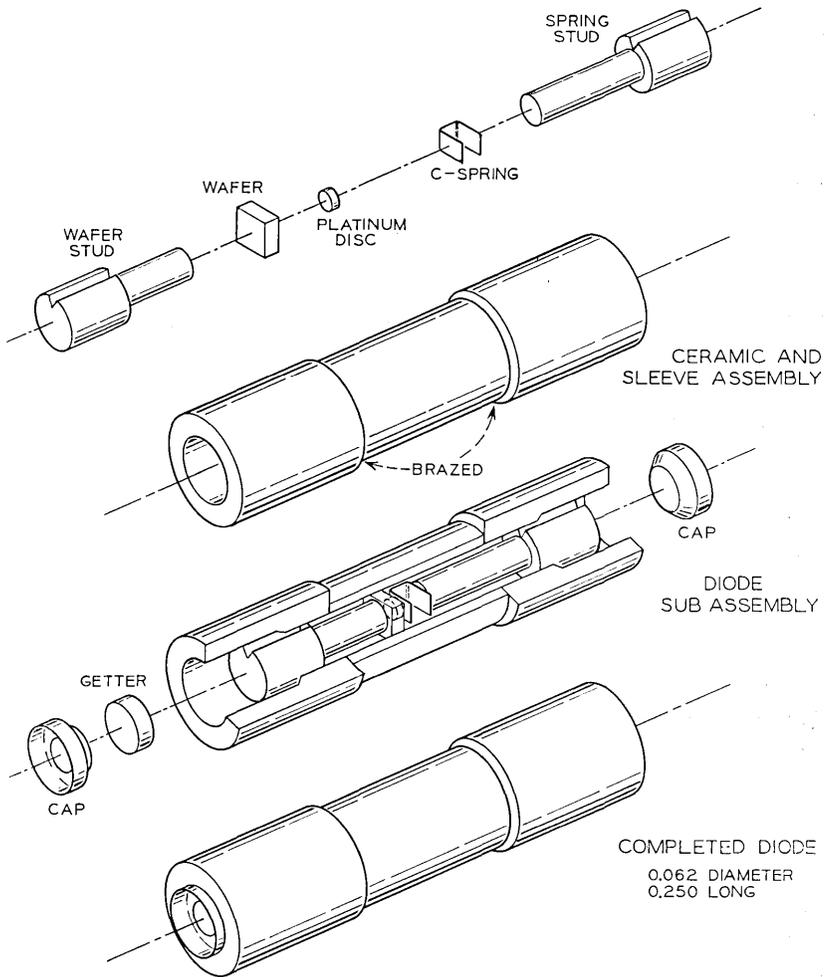


Fig. 13 — 471A diode in a U package.

wire S-spring point contacts on the top of the silicon wafer, and the 497A gallium arsenide diode uses the basic gold-plated platinum disk contact and C-spring shown for the 471A in Fig. 13. The 488A diode has external nickel wire leads welded to the endcaps of the U package.

### 3.1 Silicon Varactor

The 471A diffused epitaxial silicon mesa varactor was developed to satisfy the need for an upconverter diode that would give about

8.5 dB conversion gain in the TD-3 transmitter modulator.<sup>13</sup> A satisfactory Western Electric diode did not exist before the 471A was coded. In order to provide the desired transmitter modulator performance, the 471A diode was developed with the following characteristics:

Semiconductor wafer			
$C_{i(0V)}$ (pF)	$R_{s(0V)}$ ( $\Omega$ )	$BV$ (V)	$f_{c(0V)}$ (GHz)
0.68 to 1.08	max. 3.0	min. 15.	min. 75.
Package parasitics		Pairing requirements	
$C_c$ & $C_f$ (pF)	$L_{s(nh)}$	$\Delta C_{T(0V)}$	$\Delta R_{s(0V)}$ ( $\Omega$ )
0.12	1.4	max. 0.1	max. 0.5

$C_{j(0V)}$  is the zero-bias junction capacitance,  $R_{s(0V)}$  is the zero-bias microwave series resistance,  $BV$  is the dc breakdown voltage,  $f_{c(0V)}$  is the zero-bias cutoff frequency,  $C_c$  is the case capacitance (about 0.07 pF) and  $C_f$  is the fringing capacitance (about 0.05 pF) between the contact spring and the silicon wafer.

The series inductance,  $L_s$ , is about 1.4 nh in the 4 GHz, 0.090-inch high waveguide circuit used with the transmission resonance method for measuring inductance and microwave resistance of diodes as described by B. C. DeLoach.<sup>14</sup> The pairing requirements, needed in the balanced hybrid transmitter modulator circuit to suppress the carrier, are that the two diodes may differ by no more than 0.1 pF in total zero-bias capacitance and no more than 0.5 ohm in zero-bias series resistance.

Varactor action is assured by the wafer design and the manufacturing processes which lead to the following  $C$ - $V$  relationship

$$C_i = \frac{C_{j(0V)}}{\left(1 - \frac{V}{0.6}\right)^{1/3}}$$

where

$C_i$  is the junction capacitance,

$C_{j(0V)}$  is the junction capacitance at zero-bias

and

$V$  is the applied voltage.

Studies showed that the transmitter modulator output power correlated well with diode resistance and capacitance, as Fig. 14 shows. In both cases, the plotted diode parameter is the average value of the two diodes involved. For a minimum total output power of 6 dBm from the transmitter modulator, the maximum series resistance for each diode was set at 3.0 ohms. The corresponding total capacitance ranged between 0.8 and 1.2 pF.

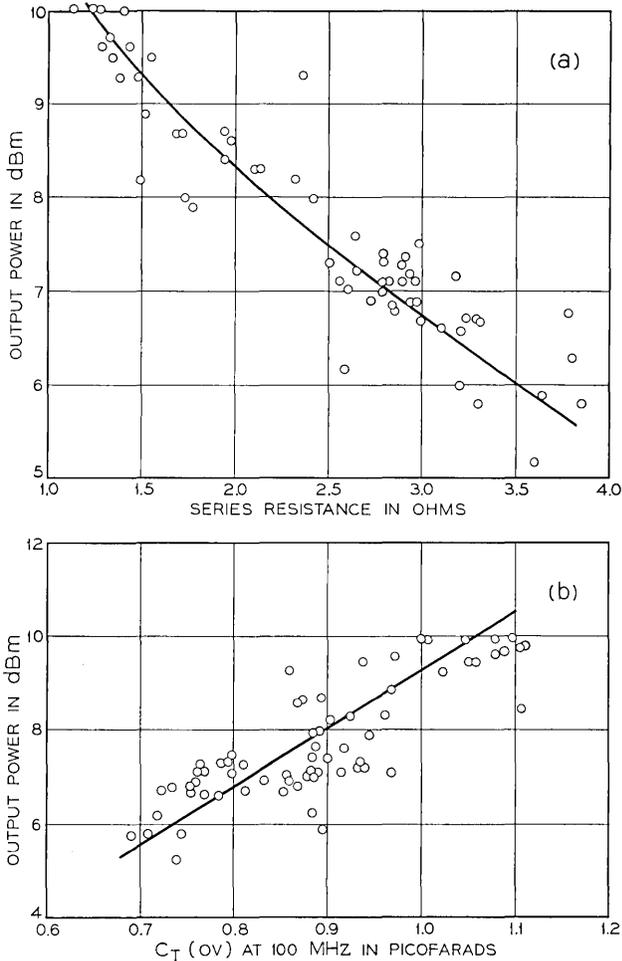


Fig. 14—Transmitter modulator output power with (a) diode series resistance and (b) diode capacitance.

### 3.1.1 Wafer Design

The TD-3 transmitter modulator circuit requires diodes with zero-bias capacitance of about 0.9 pF, and series resistance less than 3 ohms. The difficulty in fabricating such diodes arises from keeping this low resistance in a diode with such low capacitance and consequent small junction area. In addition, this application requires that the diodes have a 15 volt minimum breakdown.

The breakdown voltage and series resistance tend to be mutually incompatible and are normally met only by compromising. In conventional diodes, the resistivity of the silicon wafer is chosen to lie between a lower limit set by the  $BV$  and an upper limit set by  $R_s$ . However, in the 471A design only a lower limit on resistivity is necessary. As Fig. 15 shows, cross-diffusion between a p-type doping impurity deposited on the surface and an n-type doping impurity from the substrate is produced within the epitaxial layer. This cross diffusion provides a low  $R_s$  which is independent of the original resistivity of the epitaxial layer.

A typical slice has a 7 micron layer of 0.3 ohm-cm silicon grown on a 5 mil thick phosphorous-doped substrate of about 0.001 ohm-cm resistivity. The slice is boron diffused to form a cross-diffused p-n junction midway in the epitaxial layer as shown in Fig. 15. The incoming boron atoms produce a diffusion gradient which meets the phosphorous doping impurity diffusing out from the substrate. Since both phosphorous and boron have about equal diffusion constants, the profiles meet about halfway through the epitaxial layer. The p-n junction thus formed has a small enough doping gradient to sustain the  $BV$  desired, yet the gradient provides a heavily doped region on either side of the junction regions to reduce diode resistance.

Another stringent requirement for this diode is the tight tolerance on diode junction capacitance,  $C_{j(0V)}$ . To meet the  $\pm 0.2$  pF allowable variation, a method of chemical etching was developed that allowed for precise control of diode junction area. The initial step is to prepare 0.001-inch thick platinum disks with diameters from 0.0006 to 0.0021 inch, and a thin evaporated layer of gold and palladium.

After the capacitance per unit area of a particular lot of wafers has been determined, the diameter is chosen and a disk is alloy bonded to the surface of the diffused epitaxial silicon wafer while the wafer itself is simultaneously alloy bonded to the gold plated stud. The wafer-stud assembly is then immersed in an acid mixture, which attacks the bare silicon wafer. That portion of the wafer under

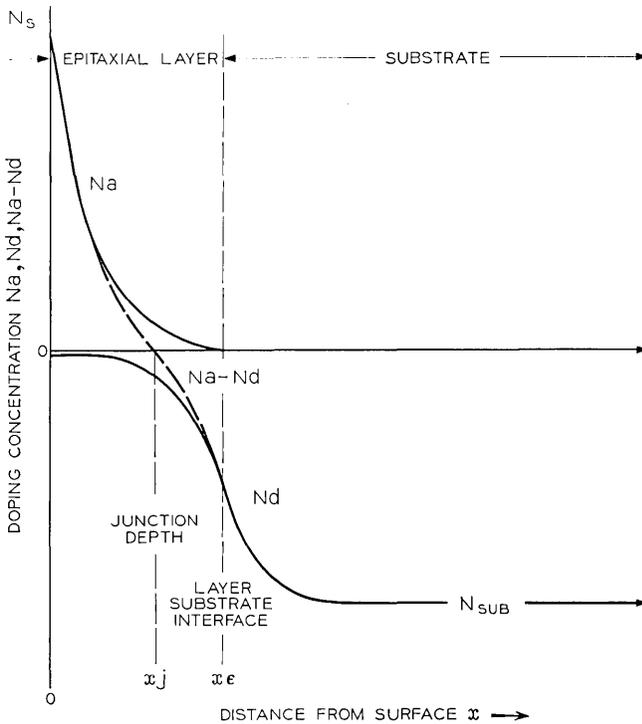


Fig. 15 — Schematic doping distribution for cross diffused junction.

the disk is protected and remains intact. The result is a mesa, the diameter of which controls the area of the junction and therefore the junction capacitance.

Now the wafer junction capacitance is carefully measured. If it is above the desired value, another brief acid immersion is sufficient to bring the capacitance within specifications. Thus, the ability to control the disk size and etching time, permits control of junction capacitance.

### 3.1.2 Diode Reliability

Device reliability is achieved first by careful attention to cleanliness in manufacturing, and second, by screening finished devices to eliminate weak units.

To insure that all volatile contaminants are driven off the interior surfaces of the diode, they are heated in a vacuum at 300°C for two

hours, then pure, dry nitrogen is allowed to enter the diodes which are then hermetically sealed.

To prevent deleterious effects of unwanted gases evolved during the closure weld and during subsequent aging, a glass frit gas-absorber is placed in one end of the unit before final welding.

To identify completed diodes that may, however, still be unstable because of some chemical or ionic contaminants, all diodes are "process aged" by subjecting them for three days to a 220°C temperature after which they are electrically tested. All diodes which shift out of the electrical limits are discarded.

To eliminate mechanically weak diodes, all are subjected to the following tests:

- (i) Centrifuge at 27,000g for 1 minute in each of 3 directions.
- (ii) Temperature cycle from -65°C to +150°C, 5 times.
- (iii) Mechanical shock at 1,500g, 5 times in each of 3 directions.

Each diode is then leak tested to insure that a diode is still hermetically sealed. Only those devices which survive all of these screening tests are used in TD-3. Step stress power aging data indicate the failure rate to be less than 100 failures per  $10^9$  device-hours at 50 mW per diode in the TD-3 transmitter modulator. Field data in a military parametric amplifier at slightly lower power levels (about 20 mW) indicate that five diodes out of 4000 apparently changed in total capacitance enough to necessitate their removal in the first  $10^8$  device hours. This indicates that a rate of 100 failures per  $10^9$  device-hours is a reasonable upper limit (with a 95 per cent confidence limit) for the 471A diode.

### 3.2 *Silicon Point Contact Detectors*

The 488A and 493A silicon point contact diodes were needed for IF and RF detector circuits because there were no hermetically sealed diodes available with the required detector characteristics. The gallium arsenide diode which ultimately became the 497A diode could have been used; however, it will probably always cost significantly more than silicon point contact diodes.

The 488A and 493A diodes are very similar. Both are in the U package and they use the same basic aluminum doped silicon material, but the wafers are processed differently. The 488A wafer is oxidized at 975°C and the 493A wafer is oxidized at 927°C. The ultimate desired results, caused by different out-diffusion profiles, are a higher

reverse impedance for the 488A diode and a lower forward resistance for 493A. A more obvious difference is that the 488A diode has 0.020-inch nickel leads welded on each end of the package and the 493A diode has no leads.

The diode specifications, based on applications studies are:

	$I_F$ at 1.0 Vdc	$I_R$
	(mA)	( $\mu$ A)
488A	20 min.	100 max. at $-2$ Vdc
493A*	20 min.	100 max. at $-1$ Vdc

As with the 471A varactor diode, the 488A and 493A diodes are vacuum-baked at  $300^\circ\text{C}$  for two hours. The final end cap weld is made in dry nitrogen. The assembled diodes are given a 72 hour process aging at  $220^\circ\text{C}$ . Leads are welded axially to the 488A end caps after aging. Both diodes are then given shock, temperature cycle, and acceleration tests before their final electrical tests.

Reliability tests indicate that the failure rate of the 488A and 493A diodes will be negligible. The diodes are rated at 30 mW; however, the various TD-3 applications are at significantly lower power levels (less than 10 mW). Temperature step stress aging indicates a failure rate of less than one failure per  $10^9$  device-hours. Long term power aging at 250 mW and 310 mW on 14 and 15 devices, respectively, has resulted in no failures after 6000 hours of aging. At 390 mW, eight out of 17 diodes failed after 6000 hours. There are no significant trends in diode parameter changes at the two lower powers, therefore the tests must continue for some time to get a good estimate of the failure rate.

### 3.3 Gallium Arsenide Schottky-Barrier Diode

The 497A gallium arsenide Schottky-barrier diode has been designed to replace the 416C silicon point-contact diode as the nonlinear resistor element in the receiver modulator or downconverter of TD-3. The 497A in an improved downconverter circuit, followed by a new low-noise transistor IF amplifier, has so improved receiver noise that

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\*The 493A specification also requires application tests in a 61A detector with minimum values of rectified current in a 1200 ohm load resistor at two different TD-3 power levels and two different TD-3 frequencies. Also, an input return loss of 20 dB minimum is required at both extremes of the TD-3 frequency band.

no microwave preamplifier is necessary to achieve TD-3 thermal noise objectives.

### 3.3.1 Required Electrical Characteristics

The present theory of downconverter diode operation uses approximations appropriate to small-signal analysis;<sup>15</sup> the 497A diode is driven in a large-signal mode by the TD-3 local oscillator and dc bias currents. Therefore, a new detailed characterization of the semiconductor wafer interaction with the circuit is needed. Further, this knowledge should be used to develop a more refined and appropriate general theory of downconverter operation. Such work is in progress in many laboratories, including our own, but is not complete. The only adequate measure now of downconverter diode suitability is a test of its performance in the microwave circuit.

This empirical approach was used to design a suitable diode for TD-3. First, an evaporated-metal Schottky-barrier diode was chosen because stored-charge effects which would degrade the desired varistor action are negligible,<sup>16</sup> and because there are well understood and controlled processes for fabricating such devices. Second, a readily tunable, low-loss, narrow-band downconverter circuit was built and used to search for the optimum semiconductor doping level and junction diameter for wafers packaged in the U encapsulation. Third, a suitable fixed-tuned, low-loss, broad-band downconverter circuit was designed using the apparent optimum device from this study.<sup>17</sup> Finally, using the broad-band circuit, the acceptable range of diode parameters was determined. They are shown in Table VI.

The forward series resistance,  $R_{sf}$ , is the resistance appearing in the diode current-voltage relationship  $I = I_S \exp [(q/nkT) (V_A - IR_{sf})] - I_S$  where  $n$  is an experimental constant near 1.

The graphical determination of  $R_{sf}$  for a typical 497A diode is illustrated in Fig. 16. Unlike the  $R_{s(OR)}$  common in varactor charac-

TABLE VI—ELECTRICAL REQUIREMENTS FOR THE TD-3 RECEIVER MODULATOR DIODE, TYPE 497A

Noise figure ( $f = 3.95$ GHz NF <sub>IF</sub> = 2.4 dB, $I_D = 8$ mA dc)	NF 6.9 dB maximum
Capacitance ( $V_A = 0$ volts)	C 0.2 pF minimum 0.6 pF maximum
Reverse current ( $V_A = -6$ volts)	$I_R$ $10^{-4}$ A maximum
Forward series resistance	$R_{sf}$ 2 $\Omega$ maximum
Power dissipation	P 0.05 W maximum

terization,<sup>9</sup>  $R_{sf}$  includes the resistance of most of the zero-bias depletion region, since the depletion region collapses almost completely under forward bias.\*  $R_{sf}$  is discussed further in Sections 3.3.4 and 3.3.5.

### 3.3.2 Structure

The U encapsulation was chosen for the 497A diode, resulting in assembly processes represented by Fig. 13. A gold-tin solder pre-form is used for low resistance contact to the back of the GaAs wafer,

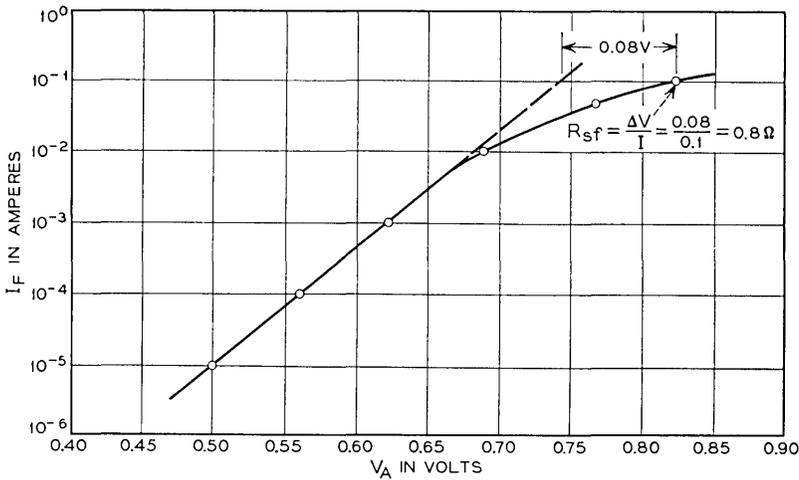


Fig. 16—Forward current-voltage relationship for a typical 497A diode illustrating the graphical determination of  $R_{sf}$ .

and a 0.0018-inch diameter by 0.001-inch thick gold-coated platinum disk is used to reduce fringing capacitance between the front of the wafer and the C-spring contact.† Both contacts are effected during a single heating cycle under about 50 gm-wt compressive force. Use of the platinum-disk raised contact requires that an overlay contact on a protective insulator be used between the disk and the smaller-diameter metal-semiconductor Schottky-barrier junction.

Figure 17 is a cross-sectional drawing of the junction designed to

\* Measurement of  $R_{sf}$  in  $p$ - $n$  junction diodes at low frequencies measures only the resistance of the wafer substrate and contacts, since injected charge modulates the conductivity of the epitaxial layer in this case. This modulation is not significant in Schottky-barrier diodes at normally-encountered current densities.

† The platinum disk is used only for contacts. The disk size in the 471A silicon mesa diode described in Section 3.1 is chosen as an etching mask to control junction capacitance.

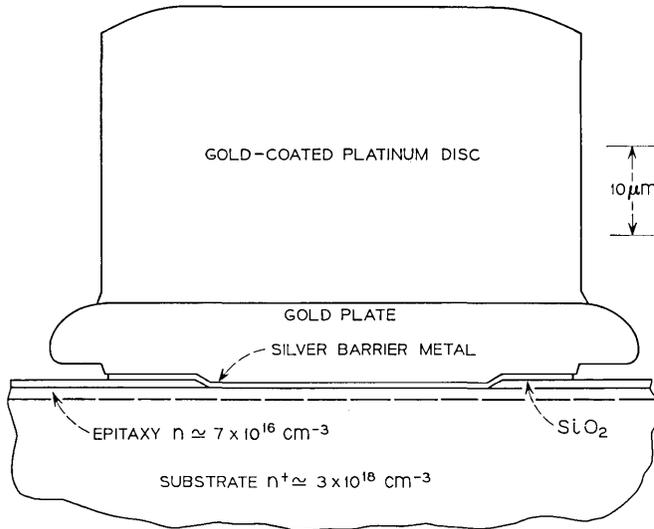


Fig. 17 — Cross section of gallium arsenide Schottky-barrier diode junction.

meet both the requirements of Table VI and the mechanical requirements imposed by the encapsulation.\* A deposited layer of  $\text{SiO}_2$  serves as the overlay insulator. Junction windows are etched through the  $\text{SiO}_2$  photolithographically, after which a thin layer of titanium is evaporated onto the slice to improve the adherence of the silver barrier metal which is next evaporated onto the slice. A second photolithographic process defines apertures through which the gold contacts are electroplated. These contacts serve as etching masks during removal of silver and titanium from areas where they are not needed. This etching is the last operation before wafering and contact bonding.

### 3.3.3 *Advances in Technology*

Significant advances in control and characterization of thin n-on-n<sup>+</sup> GaAs epitaxial layers were required to meet the 497A device requirements. A silicon-doped, water-vapor-transport system<sup>18</sup> and tin, tellurium, or selenium-doped halide-transport systems<sup>19, 20</sup> have all achieved the required steep doping gradient between substrate and epitaxial layer doping levels. A sample chip from each slice is evaluated for doping profile and for effective electron mobility using large and small-area evaporated Schottky-barrier diodes, respectively.

\* Prototype Schottky-barrier junctions on gallium arsenide developed by J. C. Irvin contributed to this junction design.

A computerized  $dC/dV$  technique<sup>21</sup> has been adapted for use on GaAs to obtain doping profile plots such as the typical one shown in Fig. 18. The  $R_{sf}$  of a sample of 0.001-inch diameter diodes is measured, using a pulse-formed alloyed-tin substrate contact, to insure adequate electron mobility.

In the complex, interdependent series of heating cycles required during processing, reverse-current degradation resulting from oxide contamination, and  $R_{sf}$  increases caused by contamination of the epitaxial layer with an acceptor (probably copper), are problems which have been brought under control during diode development.

To achieve adequate adherence of the photolithographic medium to the silver barrier metal during gold electroplating, the photolithographic process is carefully controlled and the time in the plating bath reduced to a minimum.

### 3.3.4 Wafer Design

Because devices commonly used as downconverters have barrier heights ( $\Phi$ ) varying from 0.1 to 0.9 volts, and are all forward-biased to similar current levels in downconverter operation, zero-bias cutoff

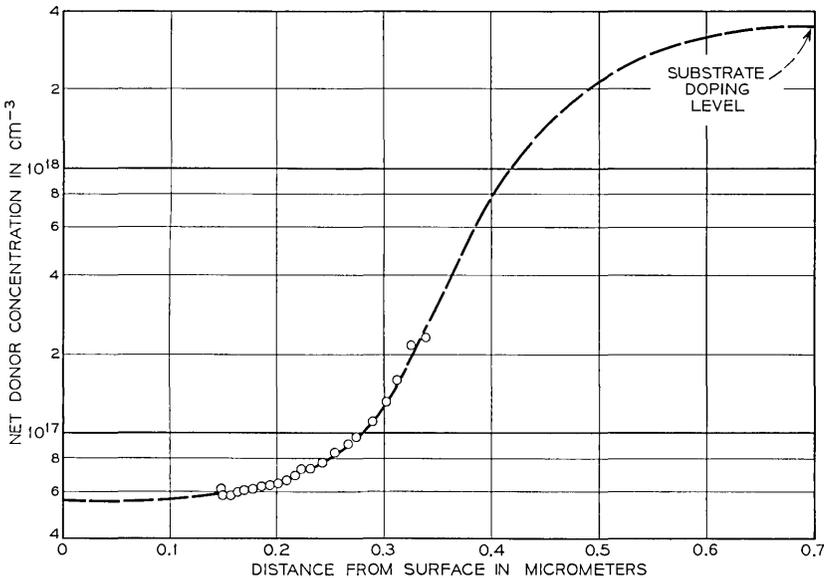


Fig. 18—A typical doping vs distance profile for the epitaxial gallium-arsenide layers used for the 497A diode.

frequency is not a useful characterization parameter. A more meaningful parameter is the forward-bias cutoff frequency

$$f_{cf} = \frac{1}{2\pi C_{jf} R_{sf}}$$

$C_{jf}$  is defined at  $V_A = \Phi - 0.1$  Vdc and can be calculated from the more common  $C_{j(ov)}$  as  $C_{jf} = C_{j0} (10 \Phi)^{1/2}$  by making the reasonable approximation of uniform doping within the zero-bias depletion layer. The bias point  $V_A = \Phi - 0.1$  Vdc is chosen since for many Schottky-barrier diodes, approximately this voltage produces the forward current at which the junction resistance  $R_j = R_{sf}$ , where  $R_j = nkT/qI$ .

Figure 19 shows the calculated  $f_{cf}$  as a function of epitaxial doping density for n-type gallium arsenide Schottky-barrier diodes. Curves for several junction diameters are shown. A curve for p-type silicon is also included, since this is used in the 416C point-contact diode. A contact resistance of  $0.1 \Omega$  and a  $0.5 \mu\text{m}$ -thick epitaxial layer are assumed. For GaAs, layers having practical values of electron mobility on substrates of  $0.002 \Omega\text{-cm}$  resistivity are assumed. In practice, the lightly-doped layer

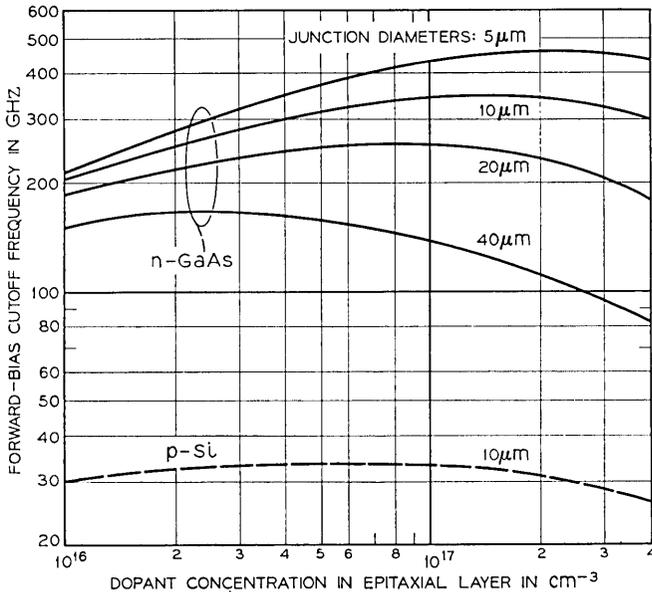


Fig. 19 — Theoretical forward-bias cutoff frequencies of Schottky-barrier varistors in n-type GaAs and p-type Si, as a function of doping impurity density in a  $0.5 \mu\text{-thick}$  epitaxial layer.

in the 416C is formed by out-diffusion of impurities and by "tapping," and is  $\approx 0.2 \mu\text{m}$  thick, resulting in a nominal  $f_{cf}$  of 50 GHz.

If there were no fringing capacitance or burnout constraints, a 5  $\mu\text{m}$ -diameter junction on a GaAs epitaxial layer doped near  $2 \times 10^{17} \text{ cm}^{-3}$  would be near optimum. However, the thermal resistivity of GaAs is triple that of silicon,<sup>22, 23</sup> and the outside diameter of the overlay is fixed by the choice of encapsulation and contacting technique. These considerations, coupled with the case capacitance of 0.12 pF, change the optimum to a junction diameter near 25  $\mu\text{m}$ , a doping level near  $5 \times 10^{16} \text{ cm}^{-3}$ , and an epitaxial layer thickness of 0.2  $\mu\text{m}$ . Manufacturing tolerances force a range of epitaxial thicknesses, doping levels, and junction diameters varying from these optima toward thicker layers, higher doping, and smaller junctions.

### 3.3.5 Results

Figure 20 shows the experimental relationship obtained between  $f_{cf}$  and over-all noise figure for a tunable 6 GHz downconverter with reflected image and  $NF_{IF} = 1.1 \text{ dB}$ , and for fixed-tuned downconverters with absorbed image and  $NF_{IF} = 2.4 \text{ dB}$ , suitable for use in TD-3. The left end of each curve is based on measurements of

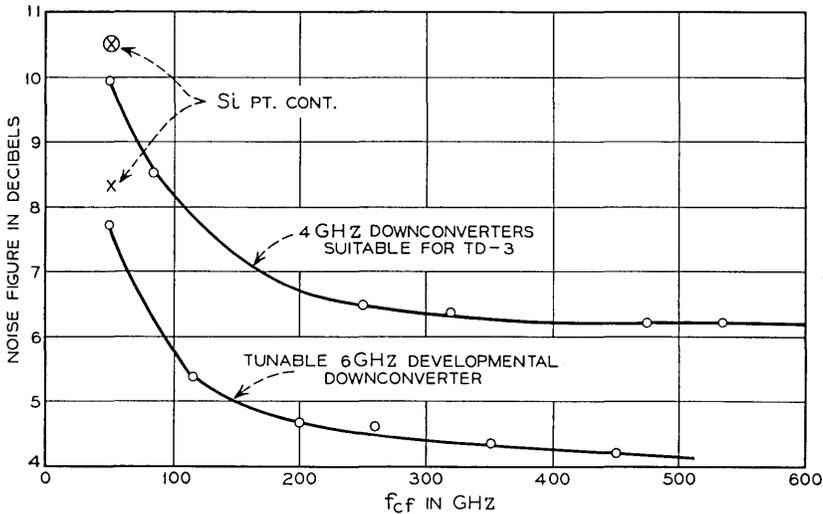


Fig. 20 — Overall noise figure vs  $f_{cf}$  for a 6 GHz reflected-image downconverter with  $NF_{IF} = 1.1 \text{ dB}$ , and for a 4 GHz absorbed-image downconverter with  $NF_{IF} = 2.4 \text{ dB}$ .

silicon point-contact diodes with junctions like the 416C. The excess noise characteristics of such junctions causes these points to be about 0.6 dB higher than the curve, which is drawn for Schottky-barrier junctions. All the other experimental points are median values for groups of diodes fabricated during the development of the 497A.

Based on these experimental curves, the design of the downconverter diode for TD-3 must provide a value of  $f_{c,f}$  somewhat above 175 GHz to assure NF below 6.9 dB. For a typical total capacitance of 0.45 pF, this requires  $R_{s,f} \leq 1.1 \Omega$  (overlay capacitance  $\simeq 0.05$  pF,  $\Phi \simeq 0.8$  V). Lack of control of electron mobility in thin epitaxial layers of GaAs prevents calculation of the resistance expected from the measured doping-versus distance profiles. However, assuming contact resistances of  $0.1 \Omega$ , and calculating the spreading resistance from a  $25 \mu\text{m}$  diameter junction in the substrate<sup>24</sup> to be  $0.2 \Omega$ , we find that for a  $0.5 \mu\text{m}$  thick epitaxial layer, uniformly doped at  $5 \times 10^{16} \text{ cm}^{-3}$ , the average electron mobility must be at least  $1600 \text{ cm}^2 \text{ V}^{-1} \text{ sec}^{-1}$ .

Referring to the requirements for the 497A diode listed in Table VI, typical values of parameters for the device being manufactured are:

$NF$	6.5 dB
$C$	0.45 pF
$I_R$	$10^{-9}$ A
$R_{s,f}$	$0.8 \Omega$
$BV(I_R = 10 \mu\text{A})$	12 V

The mechanical and thermal process stresses used to screen 471A diodes are also used for the 497A. Figure 21 shows the step-stress aging curves for temperature and 60 cycle, 6 volt peak-reverse-voltage power stressing. For use in TD-3, a rate of less than 10 failures per  $10^9$  device-hours is predicted. These data are being supplemented by long-term aging tests.

#### IV. HIGHLY STABLE DIODES FOR THE IF DEVIATOR

By S. M. Forst, G. F. Foxhall, and G. A. Kelley, Jr.

This paper discusses the development of two semiconductor diodes to be used in the IF deviator of the 3A FM terminal transmitter. These devices have been coded 446AC, an 8.2 volt regulator diode, and 457A, a variable capacitance diode. While the two may be thought of as dissimilar device types, the system requires a like feature in each: great stability of electrical characteristics.

The heterodyne deviator contains two voltage controlled Colpitts

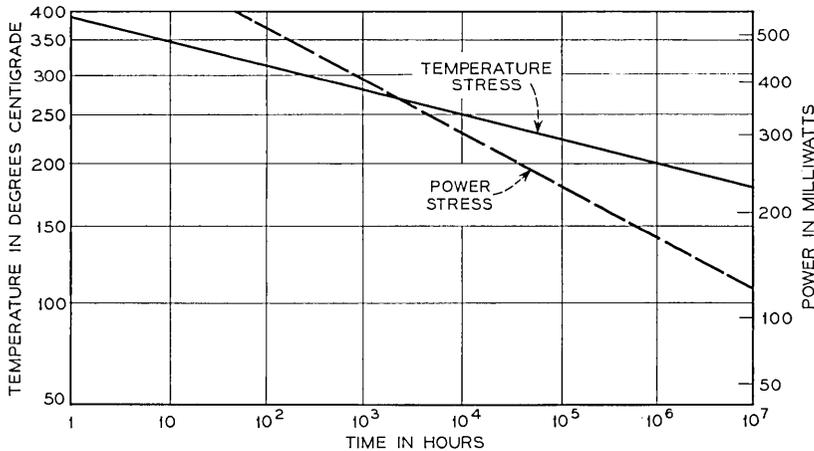


Fig. 21 — 497A stress aging: median-failure stress vs time.

oscillators operating at 186.67 and 256.67 MHz. The output frequencies are mixed, and the difference frequency of 70 MHz is extracted. The latter must have a drift instability of  $\pm 100$  kHz maximum for a three-month interval.

An ultrastable 5 volt regulated supply using the 446AC diode provides bias for the 457A diodes in the oscillator tank circuits. (See Fig. 22). The baseband signal superimposed on this bias modulates the capacitance of each 457A diode, and consequently the output frequencies of the oscillators as well.

Since the stability of every component in the deviator system will affect the eventual stability of the output, an allowable range of frequency drift was assigned to each code.<sup>3</sup> For the 446AC diode, the allotted drift of 7 kHz means that the breakdown voltage must be stable to within 7 millivolts. The allowable drift of 43 kHz by the 457A diode places requirements on two characteristics. The reverse leakage current must be stable to within 2.3 nanoamperes to prevent excessive loading of the bias supply, and the capacitance must be stable to within 0.0078 pF.

#### 4.1 446AC Diode

##### 4.1.1 Desired Characteristics

The 446AC diode was the simpler of the two devices to develop since it is an extension of the existing 446T diode design. The 446T

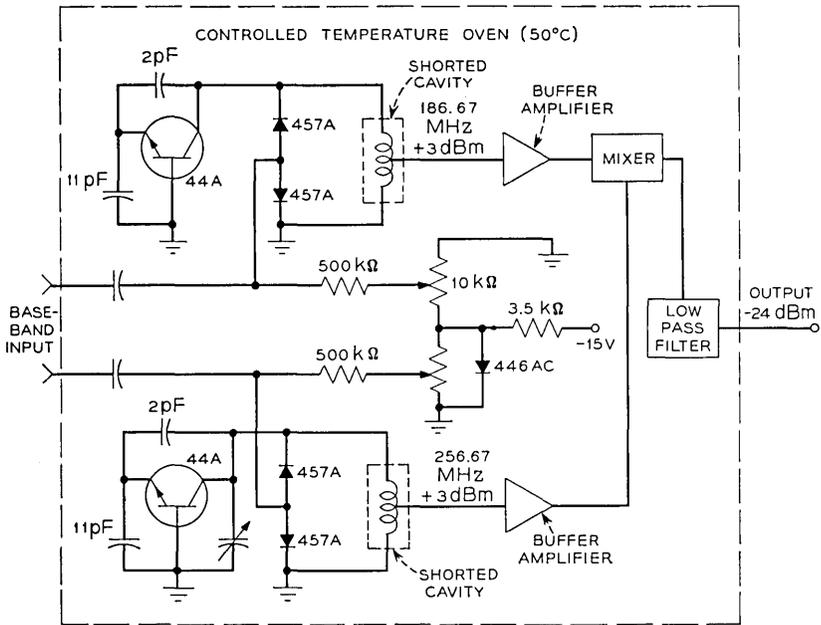


Fig. 22 — Simplified diagram of the 3A FM terminal transmitter deviator.

is an 8.2 volt regulator diode with a 5 per cent range on breakdown voltage. It was the additional requirement of 7 mV stability which necessitated the new code.

#### 4.1.2 Description and Fabrication

The 446AC is a member of a family of K package devices. This implies a hermetic metal encapsulation as shown in Fig. 23. The body of the device is coated with grey paint; the leads are solder coated.

Manufacturing begins with the diffusion of an n-type layer (phosphorus is used as the doping impurity) into a slice of p-type silicon. An undesired n-layer is removed from the back side of the slice when the latter is lapped to give a final thickness of 0.007 inch. Metallic contacts are applied to the slice through plating and sintering; the outer coating is gold to facilitate subsequent wafer and wire bonding.

Individual diode wafers are formed by a stream of airborne abrasive. Round metal discs are held against the slice during this operation to protect those areas on the slice which will eventually become



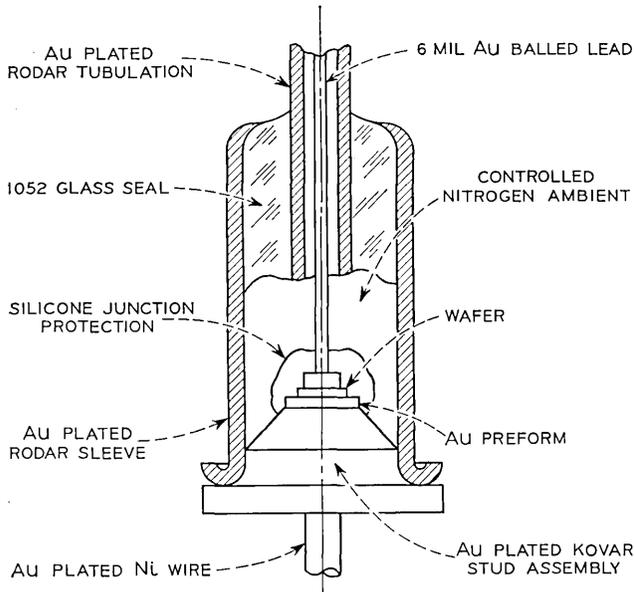


Fig. 24 — Internal structure of the K package assembly.

able. The relatively large tolerance on drift used for these accelerated test conditions is sufficient to provide a 7 mV stability at 16 mW use.

The expected failure rate of K package diodes is normally considered to be about one failure per  $10^9$  hours at a junction temperature of  $60^\circ\text{C}$ . The results of initial attempts to measure failure rate under an end point requirement of 7 mV stability were unfortunately obscured by variations in ambient temperature while readings were being taken. The experiment did show, however, that the device was indeed capable of meeting the requirement. This is verified by 10

TABLE VII—ELECTRICAL CHARACTERISTICS OF THE 446AC DIODE

		Min.	Max.	Typical	Units
Breakdown voltage ( $I_R = 10$ mA dc)	$BV$	7.8	8.6	8.2	Vdc
Forward voltage ( $I_R = 400$ mA dc)	$V_F$	—	1.0	0.9	Vdc
Saturation current ( $V_R = 6.5$ Vdc)	$I_S$	—	2.0	0.3	$\mu\text{A dc}$
Breakdown impedance ( $I_R = 10$ mA dc)	$bz$	—	7	5	ohms
Stability, breakdown voltage ( $I_R = 1.3$ mA dc)					
Accelerated tests	$\Delta BV$		50		mVdc
Use conditions	$\Delta BV$		7		mVdc

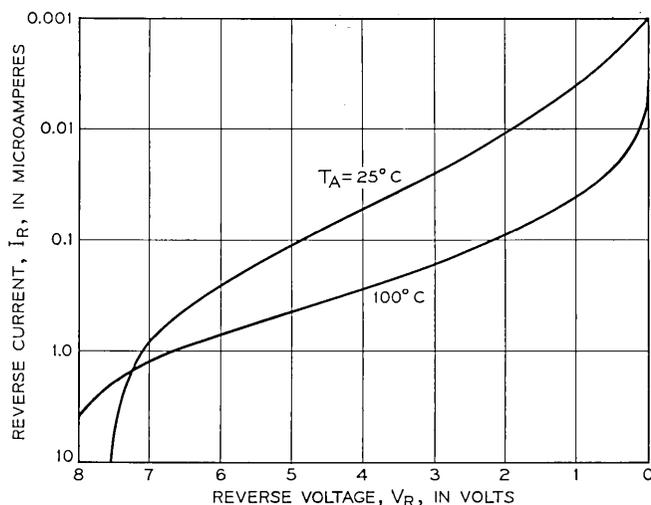


Fig. 25 — Reverse characteristics of the 446AC diode.

months of field experience during which no system failures have been recorded. A redesigned experiment to give more accurate values is under way.

## 4.2 457A Diode

### 4.2.1 Desired Characteristics

It was immediately apparent that many of the objective requirements for the 457A diode would make necessary the development of an entirely new code. Table VIII lists these requirements, of which

TABLE VIII — PROPOSED CHARACTERISTICS FOR THE 457A DIODE

		Min.	Max.	Units
Capacitance ( $V_R = 8$ Vdc)	$C$	18	22	pF
Sensitivity ( $C = V^{-n}$ )	$n$	1/3		
Inductance	$L$		1	nh
Series resistance	$R_S$		0.5	ohms
Breakdown voltage ( $I_R = 5 \mu\text{A}$ dc)	$BV$	25		Vdc
Reverse current ( $V_R = 11$ Vdc)	$I_R$		10	nAdc
Stability		as stable as possible		
Capacitance	$\Delta C$			
Reverse current	$\Delta I_R$			

capacitance, series resistance, inductance, and stability were considered to be controlling.

#### 4.2.2 Description and Fabrication

An S-package design was selected primarily because of its low inductance. This incorporates two stud lead assemblies sandwiching the wafer, with a glass sleeve sealed along the major portion of the length of the studs. (See Fig. 26.) Because the body is so small, colored coding bands replace the usual alphanumeric identification.

Manufacturing is similar to that for the 446AC. An n-type slice is simultaneously diffused with both p and n layers on opposite sides of the slice. This leaves a very thin region in the middle of the slice where the silicon resistivity is high. This technique provides the low series resistance necessary to obtain high Q. Plating and sintering apply a metal contact to the slice.

The diode wafers are formed by an abrasive stream, but the stream is focused to cut a restricted kerf, and the resulting wafers are square.

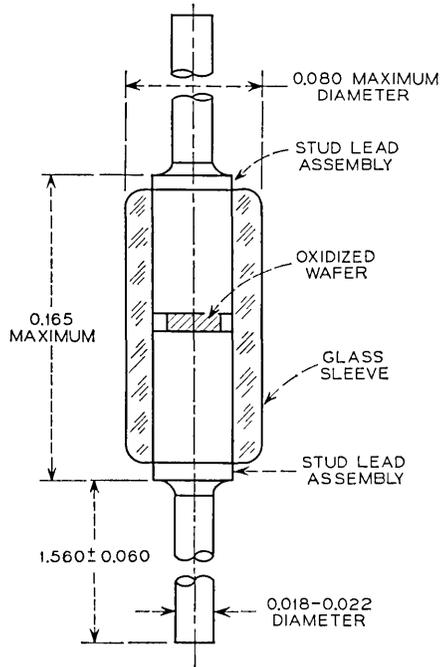


Fig. 26 — S package assembly.

The damaged edges are etched, and then an oxide is grown on the exposed silicon for passivation and protection. (See Fig. 27.) It is this oxide which prevents damage to the wafer during sealing, which is done at 830°C, and provides the required stability of electrical characteristics.

#### 4.2.3 *Electrical Characteristics and Reliability*

As the development program progressed, it became evident that some of the requirements would have to be modified slightly to be compatible with manufacturing. For example, Fig. 28 shows distributions of series resistance achieved over successive runs. Based on these curves, a value of 0.7 ohm was specified as a maximum for series resistance, slightly greater than the 0.5 ohm requested. Table IX shows the final electrical characteristics. Figures 29 and 30 show the dependence of capacitance on voltage and temperature, respectively.

As with the 446AC diode, preaging assures that the 457A meets the stability requirements. All devices are subjected to 1000 hours of stress at a reverse bias of 5 volts and an ambient temperature of 150°C. Characteristics are measured at 250 hour intervals, and those devices drifting more than 0.1 pF or 2.3 nanoamperes are rejected.

A measuring system and associated facilities to detect capacitance changes as small as 0.0001 pF has been constructed. An experiment is under way to determine the expected failure rate at 5 volts reverse bias use condition and 50°C ambient temperature. At the time of this writing, no system failures caused by the 457A have been re-

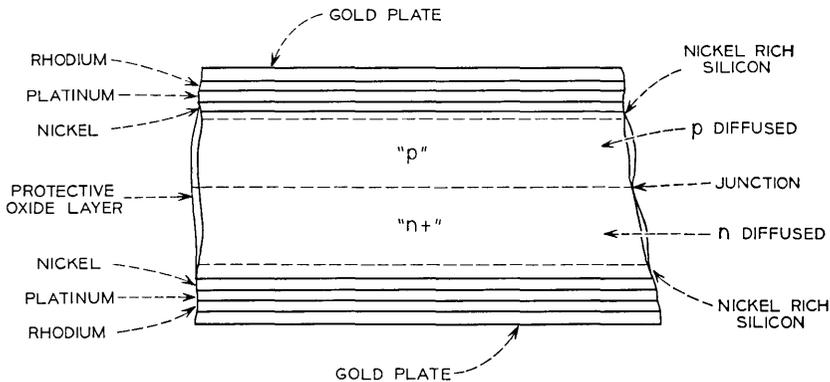


Fig. 27 — Completed 457A wafer.

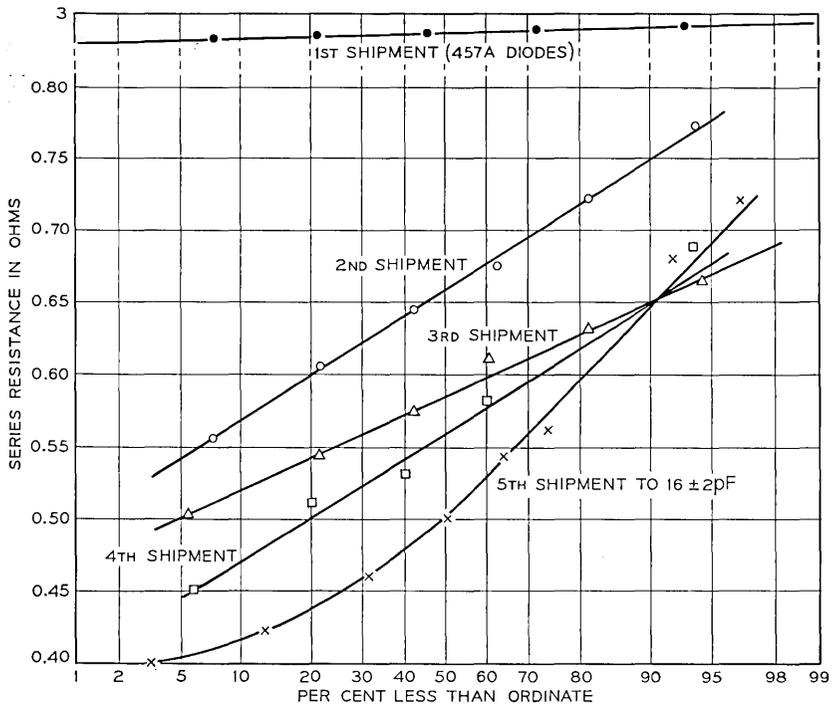


Fig. 28—Series resistance (ohms) per cent less than ordinate.  $f = 100$  MHz,  $V_R = 5$  volts,  $R_s = 1/(2\pi f c Q)$ .

TABLE IX—ELECTRICAL CHARACTERISTICS OF THE 457A DIODE

		Min.	Max.	Typical	Units
Breakdown voltage ( $I_R = 5 \mu\text{Adc}$ )	$BV$	50	—	60	Vdc
Forward voltage ( $I_F = 100 \text{ mAdc}$ )	$V_F$	—	1.0	0.82	Vdc
Reverse current ( $V_R = 5$ volts)	$I_R$	—	0.010	0.0005	$\mu\text{Adc}$
Capacitance ( $V_R = 5$ volts, $f = 100$ KHz)	$C$	15	17	—	pF
Quality factor ( $V_R = 5$ volts, $f = 100$ MHz)	$Q$	130	—	175	—
Series resistance ( $V_R = 5$ volts, $f = 100$ MHz)	$R_S$	—	0.7	0.6	ohms
Inductance	$L$	—	—	3	nh
Stability					
Capacitance ( $V_R = 5$ volts, $f = 100$ KHz)					
Accelerated tests	$\Delta C$	—	$\pm 0.10$	—	pF
Use conditions	$\Delta C$	—	$\pm 0.0078$	—	pF
Reverse current ( $V_R = 5$ volts)	$\Delta I_R$	—	$\pm 2.3$	—	nA

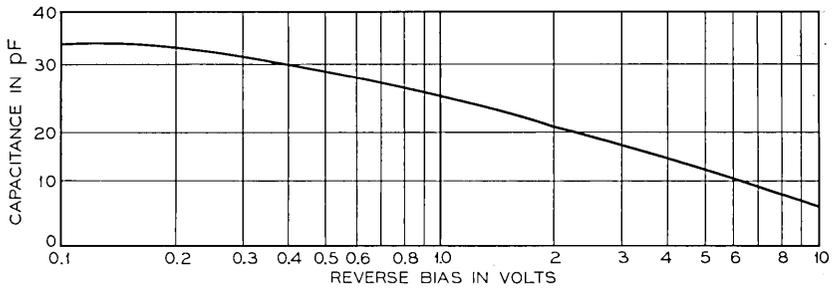


Fig. 29 — Dependence of the 457A diode's capacitance on reverse voltage.

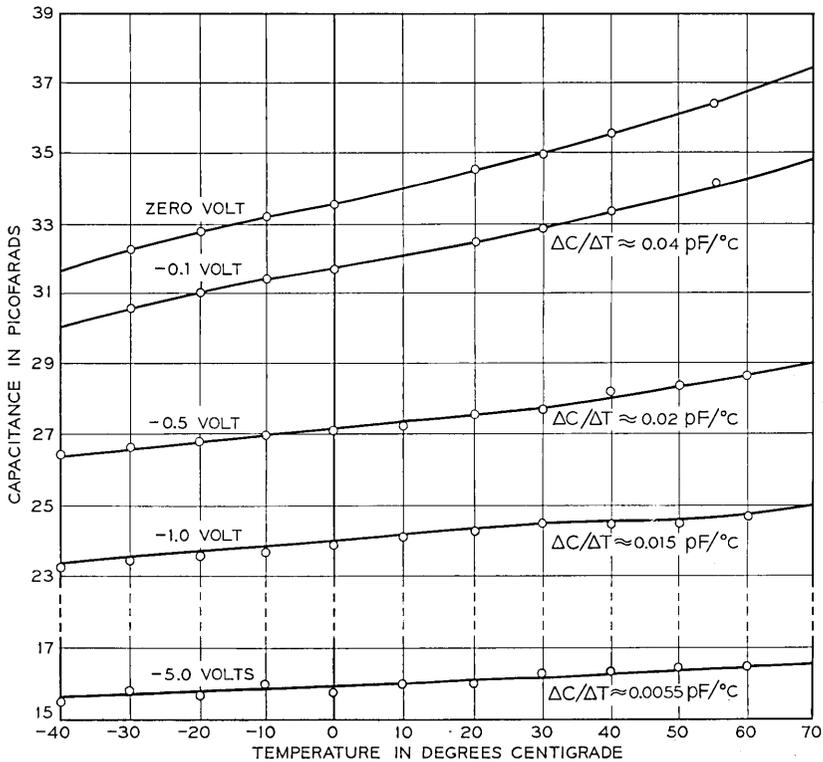


Fig. 30 — Dependence of the 457A diode's capacitance on temperature.

corded, verifying that the screening procedure is passing only acceptable devices.

#### 4.3 Summary

Two devices have been developed for the 3A FM deviator. They are the 446AC, an 8.2 volt regulator, and the 457A, a variable capacitance diode.

Over the 3-month system maintenance interval, the breakdown voltage of the 446AC must be stable to within 7 millivolts; and the capacitance and reverse current of the 457A must be stable to within 0.0078 pF and 2.3 nanoamperes, respectively. Accelerated preaging assures such compliance with these requirements that to date there have been no system failures because of drift in either diode.

### V. SILICON SCHOTTKY BARRIER DIODES By H. J. Lory

#### 5.1 Applications and System Requirements

The 479A and 479B diodes are planer epitaxial silicon Schottky barrier diodes developed for the TD-3 system.

The 479A is used in the IF limiter of the TD-3 microwave transmitter<sup>1</sup> and in the TD-3 FM terminal receiver.<sup>3</sup> To provide the required limiting, a diode was needed with low capacitance (less than 0.6 pF), low recovery time (less than 500 ps), and moderate breakdown voltages (greater than 10 volts). Since a diode with all these characteristics was not available, the 479A was developed.

The 479B is used in discriminator circuits of the FM terminal receiver and transmitter.<sup>3</sup> Here, a high rectification efficiency is required at 70 MHz; this condition also calls for low capacitance and reverse recovery time. The bias on the diode is normally less than ten volts, but may sometimes go as high as 15 volts during certain testing procedures. Hence, the 479B has a higher reverse bias voltage specification than the 479A (20 volts vs 10 volts, respectively). Point contact diodes have suitably low capacitances and reverse recovery times, but their reverse breakdown is low. Hence, a Schottky barrier device is used.

In this class of device, conduction is almost entirely by majority carriers, so that there is negligible reverse recovery time. The desired low forward impedance can be achieved simultaneously with low capacitance by using epitaxial silicon with a closely controlled thickness and a doping level consistent with the breakdown voltage requirement.

## 5.2 Configuration

Initially, an evaporated gold dot on bare silicon was used. While this diode performed system functions satisfactorily, it was sensitive to ambient moisture and high temperature during fabrication. Moreover, the fragility of the gold-silicon bond led to low shock and centrifuge endurance compared with other configurations.

The configuration finally used is a modification of that developed by Kahng and Lepselter.<sup>25</sup> Figure 31 shows a cross section of the 479A diode wafer. (The 479B is fabricated similarly to the 479A, but with different epitaxial film parameters.) The principal features are:

(i) A Schottky barrier junction, 0.001-inch in diameter, between silicon and an "alloy" formed by a solid-solid reaction of silicon and palladium at 475°C.

(ii) A protective steam-grown  $\text{SiO}_2$  passivated layer.

(iii) A protective overlay of platinum and palladium, 0.002 inch in diameter, sealed to the oxide by a thin ( $\sim 300 \text{ \AA}$ ) layer of chromium.

This yields a die which has little sensitivity to ambient moisture and which may be eutectically bonded to the gold plated header without deteriorating electrical characteristics.

## 5.3 Choice of Package

In order to reduce capacitance shunting the diode, the die was isolated from the metal platform of the TO-18 package, which was then grounded with a third lead. Figure 32 illustrates the package

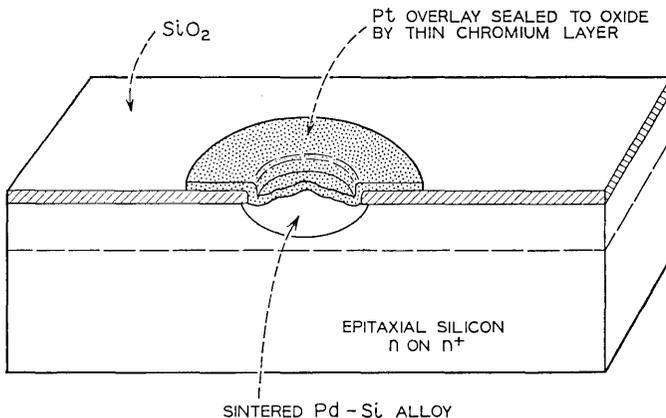


Fig. 31 — Cross section of Cr-Pd-Pt Schottky barrier diode wafer (not to scale).

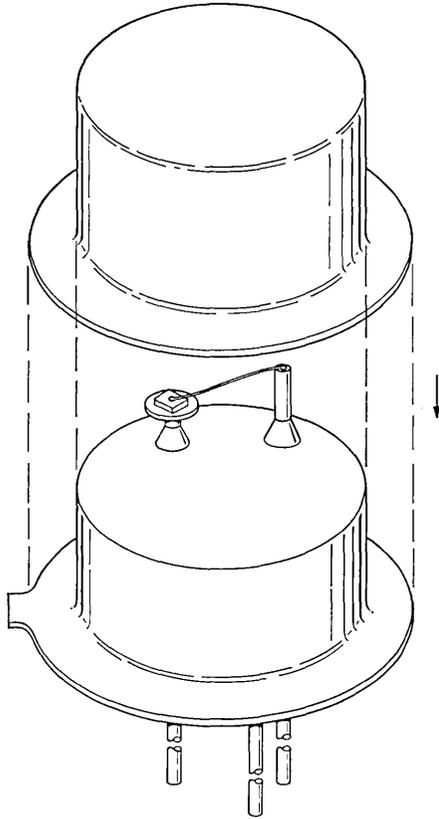


Fig. 32—Packaged 479A diode.

used. The wafer die is mounted on a “nail head” lead, and a 0.7 mil wire connects to the second lead. This configuration leads to a typical diode zero bias capacitance of 0.35 pF, of which 0.05 pF is case capacitance.

#### 5.4 *Electrical Characteristics*

Figure 33 illustrates the total zero bias capacitance distribution for a sample of 476 randomly selected units. The  $\pm 1 \sigma$  points cover a range of 0.075 pF.

The reverse recovery times of most 479A diodes are too low (less than 200 ns) to measure with available techniques. Testing to the 500 ns test specification limit is performed by switching from a for-

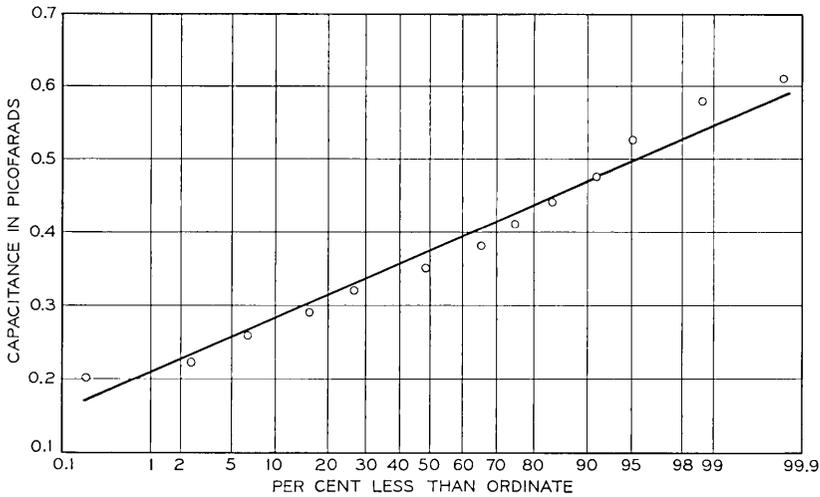


Fig. 33 — Distribution of capacitance for 479A diode.

ward bias of 10 mA to a reverse current of 10 mA while monitoring the waveform on a sampling oscilloscope. Figure 34 shows this waveform. The reverse peak is caused by the shunt capacitance; the ringing is associated with the lead inductance (approximately 2 nH) and the shunt capacitance. The reverse recovery time associated with minority carrier injection is not detectable here; it is less than 200 picoseconds.

Figure 35 shows the forward current-voltage characteristic of the

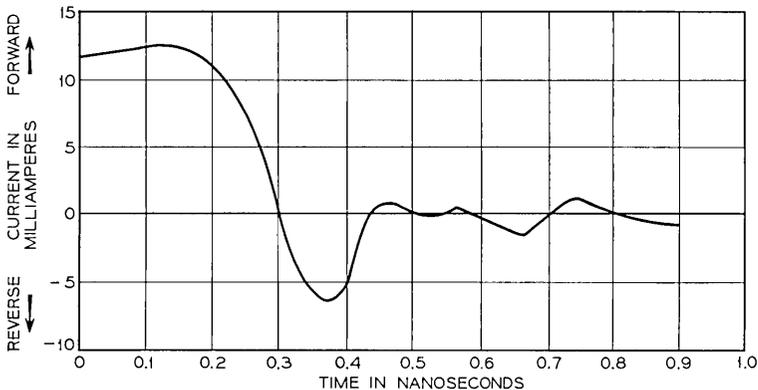


Fig. 34 — Reverse recovery characteristic of 479A diode.

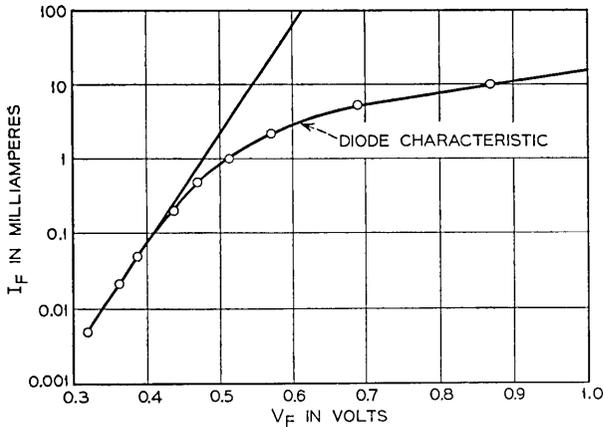


Fig. 35 — Forward characteristic of a typical 479A diode.

479A diode. At lower current levels, the characteristic fits the curve  $I = I_o [\exp(qV/nkT) - 1]$ . The number  $n$  varies from 1.01 to 1.30; the theoretical value<sup>25</sup> is 1.03. At higher current levels there is a deviation from the exponential relationship which is only partly explained by resistive drop in the epitaxial film. The problem of identifying the mechanisms affecting Schottky barrier current-voltage characteristics is complex and has been considered by many authors.<sup>26-30</sup> A quantitatively accurate model for the 479A diode has not been derived.

Reverse currents for the 479A and 479B are typically  $\sim 0.002$  microamperes. Breakdown voltage is a function of the material and of the minimum curvature radius of the metal-semiconductor junction; it ranges typically from 15 to 25 volts.

### 5.5 Aging and Reliability

In order to test the basic susceptibility of the wafer to environmental influences, 29 diodes were aged in steam at 250 and 300°C on unencapsulated headers for 65 to 161 hours. Table X shows the results. No failures occurred at 250°C, and only three failed from environmental causes at 300°. In some cases, a high  $V_F$  reading occurred because of deterioration of the ohmic back contact; in every such case, rebonding brought the unit back within specifications.

Representative temperature aging data on encapsulated diodes is shown in Fig. 36. Four groups of 40 or more units were subjected to

TABLE X—AGING 479A AND 479B DIODES\*

Temperature (°C)	Time (cumulative hours)	Number tested	Number failures	$V_F$	$I_s$	$BV$	Open	Short	Damaged in testing
250	120	29	0						
250	160	29	0						
300	65	29	4		1			1	2
300	161	25	3		1				2

\* Steam, unencapsulated, complete overlap.

aging for fixed times at increasing temperatures from 150 to 450°C. The time intervals chosen were 2, 16, 168 and 1000 hours.

Failures at the lower temperatures most often resulted from breakdown voltage deterioration, while failures at higher temperatures were caused by high  $V_F$ , open circuits, or a combination of these with low  $BV$ . The median failure temperatures generally lie at or above the silicon-gold eutectic temperature. In many cases, diodes with low breakdown voltages tended to heal; that is, a diode would go below the  $BV$  specification on one temperature and then return to specification at a higher temperature. In all cases, however, regression curves were plotted using the criterion that, once a diode failed, it was con-

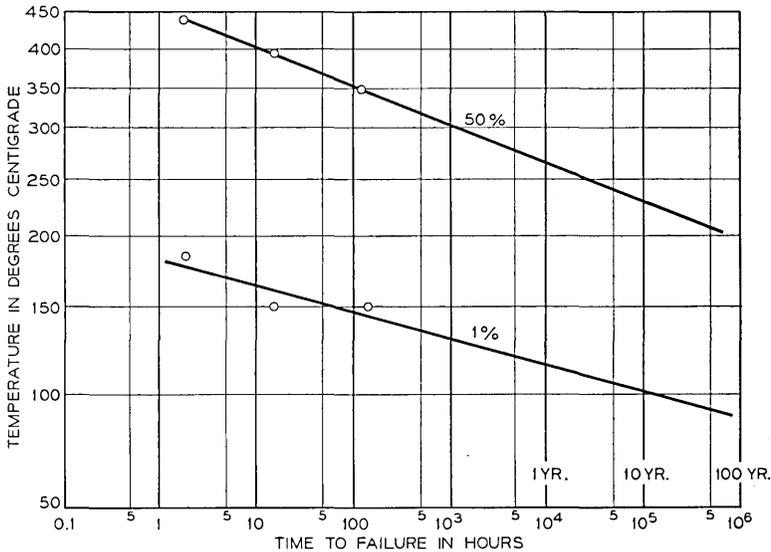


Fig. 36 — 479A aging curve, no in-process aging.

sidered defective for all subsequent measurements. Subsequent forward bias and high-temperature reverse bias aging revealed no bias-dependent modes of failure. The thermal impedance of the diode is about 1000°C per watt, and the maximum power to which the diode is subjected in use is 10 mW. This leads to a predicted failure rate of less than 10 failures per 10<sup>9</sup> device-hours for all TD-3 applications.

#### VI. HIGH-VOLTAGE RECTIFIER DIODES By I. C. Savadelis

Two new high voltage diode rectifier assemblies were required for the traveling wave tube power supply. High altitude operation imposed stringent voltage breakdown and corona requirements on these diode rectifiers. Multiple diodes were necessary to attain the high voltage requirements. Molding the multiple diodes in a single high dielectric constant plastic encapsulation assembly enhanced high altitude operation and simplified the mechanical mounting problem.

Type 426J one watt diodes were selected for the higher voltage diode requirements. Since these diodes were to be used in a voltage doubling circuit and in a full-wave bridge configuration, it was advantageous to mold four of them into a single package. The voltage doubling circuit required a breakdown voltage of 4800 volts in each leg. Therefore, two pairs of diodes were used in each leg with a third solder terminal between each leg. This allows the diode pairs to be used also in a full-wave bridge. The maximum forward voltage drop required across each pair of the 426J diodes was 7.5 volts at 300mA. The power rating of the assembly is 5.0 watts. This assembly was coded the 463A diode (multiple semiconductor type).

The molding concept was carried over to the other diode rectifier in the power supply. A bridge rectifier was needed with a minimum breakdown of 1200 volts between the terminals of each leg. Four 426G diodes molded into a four-terminal bridge configuration fulfilled this requirement. The power rating of the assembly is 5 watts and has a maximum forward voltage drop of 2.1 volts at 600mA. This assembly was coded 464A.

It was necessary that the molded assemblies mount easily and have a thermal impedance to the mounting surface lower than the individual lead mounted diodes. The molded assemblies are rectangular, with two mounting ears diagonally opposite each other. This gives a package that is approximately 2.75 inches long, 1.5 inches wide and 0.8 inch deep, as shown in Fig. 37.

The assemblies are fabricated using the diodes, a high dielectric

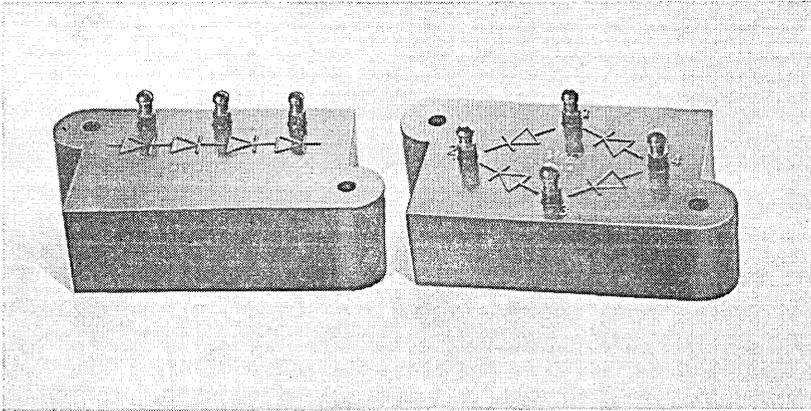


Fig. 37 — The 463A and 464A diodes.

constant silicone resin, an alumina filled epoxy, and a premolded silicone shell with terminals. Figure 38 shows the general assembly structure. The silicone resin serves as a high dielectric insulator across the glass seal area of the diode and mechanically decouples the diode from the epoxy resin. The epoxy supports the diode and wiring structure, and seals the diode from moisture. An alumina filling agent in the epoxy lowers the thermal impedance of the package, resulting in lower diode junction temperatures. The shell, although not required for the design of a molded assembly, was used to reduce fabrication cost. Because the 463A and 464A are high voltage diodes, care was

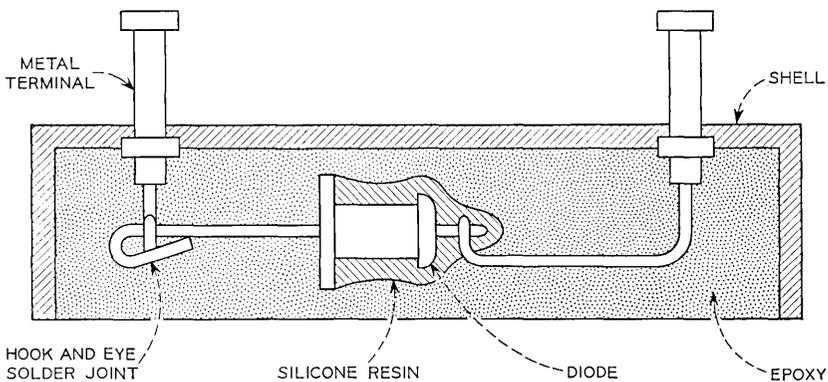


Fig. 38 — Cross section of 463A and 464A diode assembly.

exercised during fabrication to avoid pinholes and voids in the molding materials and sharp projections on the metal parts.

The individual diodes were coated with the silicone resin and were evacuated to a vacuum of 5 to 10 mm of Hg for 15 minutes to remove all air bubbles. The resin was cured in a 150°C oven for 15 minutes. The operation was repeated 3 times resulting in a 0.012-inch thick coating.

The coated diodes are then assembled into the desired circuitry and placed in the silicone shell. Where the diode leads join, a hook and eye solder joint is used and care taken to avoid sharp points and voids in the solder joint.

Filling the shell with the alumina loaded epoxy resin is the next step. The epoxy resin is mixed with the catalyst or hardening agent, thoroughly blended, and subjected to a vacuum of 5 to 10 mm of Hg to remove any gas. Then the epoxy resin is poured into the shell and the assembly degassed at 200 microns of Hg. The assembly is cured for 6 hours at 85°C, which is sufficient to harden the epoxy; and then at 150°C for 16 hours. Both curing steps are performed in an air atmosphere. This two-step cure is necessary to keep epoxy shrinkage to a minimum.

The completed assemblies are ac corona tested. For this test, all the terminals are connected and the corona is measured between the shorted terminals and a grounded plane adjacent to the assembly mounting surface. The observed corona voltage, for a charge transfer of 20 picocoulombs, is in excess of 6000 volts rms.

The standard 1 watt (426 type) diode used in fabricating the molded assemblies has a junction-to-ambient thermal impedance of approximately 50°C per watt. However, when molded into an assembly this thermal impedance is lowered to 32°C per watt. This improved heat conductance has allowed the molded assemblies to be conservatively rated at 5 watts.

Environmental tests were performed on the molded diode assemblies. These included temperature cycling from -65 to 125°C for five 90-minute cycles, thermal shock from 100 to 0°C, ten day moisture, accelerated moisture for 3 cycles and storage at 150°C for 1000 hours. The molded assemblies adequately met these tests, which are in excess of the environmental conditions encountered in TD-3 operation. Based on aging data of the 426G and 426J diodes, the reliability estimate of the 463A and 464A diodes for the TD-3 operating conditions is better than 100 failures per 10<sup>9</sup> device-hours of operation.

## VII. PIN DIODE By R. J. McClure

An ultraflat IF amplifier using a variolossor stage is required for the TD-3 system.<sup>1</sup> In the variolossor stage a diode whose impedance is a pure resistance independent of frequency is needed. The diode must be capable of carrying a peak signal current of 3 ma (for 5 dB loss) without affecting the diode resistance. Any nonlinear resistance will generate harmonics. Available diodes exhibited too much frequency dependence of the forward impedance, so development of the 474A pin diode (p type-intrinsic-n) was required.

The pin diode is a semiconductor device which exhibits a resistance that varies with the forward bias current. The resistance under forward bias is roughly inversely proportional to the bias current (see Fig. 39). The resistance is nearly independent of frequency and can be controlled by very low currents.

To achieve 5 dB loss at the 3 ma peak current, the diode resistance must be 126  $\Omega$ . From Fig. 39 the dc bias must be 0.35 mAdc. The ratio of the peak ac to the required dc is 8.57. Even at this high ratio, the ac does not appreciably change the resistance of the pin diode.

The 474A pin diode is a diffused silicon, mesa-etched device mounted in a TO-18 package. The resistance of the i-layer cannot follow the instantaneous variations in the IF signal because of the

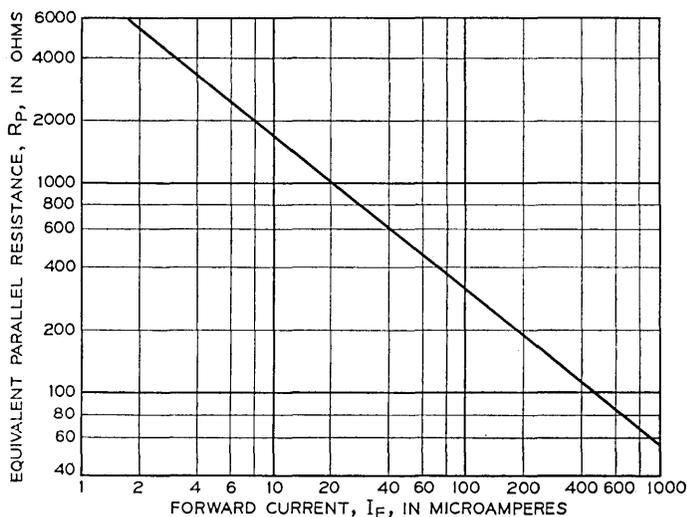


Fig. 39—Typical resistance vs forward current characteristic for the 474A pin diode.

long transit time of the i-layer and the lifetime achieved after processing. At frequencies above which the junction impedance (which is frequency dependent) is small compared to the i-layer impedance, the pin diode acts as a linear resistor whose value is controlled by the direct current.

Design of the 474A has been optimized to achieve small frequency dependence of the total diode impedance above 5 MHz. Figure 40 shows a comparison of the normalized insertion loss versus frequency for the 474A diode and for the 1N100 germanium point contact diode which had been used before the pin diode was developed. The 474A diode with its superior frequency response provides improved transmission characteristics for the IF amplifier.

The equivalent circuit for the 474A diode is shown in Fig. 41. The circuit consists of the impedance of the p-i and n-i junctions, the impedance of the i-layer, and stray reactances associated with the package and the diode wafer itself. Stray reactances and the junction impedances should be as small as possible in order to minimize the frequency dependence of the IF amplifier transmission as the loss is varied.

The frequency dependence of the total diode impedance stems from the junction impedance  $Z_j$ , which varies inversely as  $I_o(\omega\tau)^{\frac{1}{2}}$ , where  $I_o$  is the diode dc,  $\omega$  is the angular frequency, and  $\tau$  is the lifetime in the i-layer. For small frequency dependence of the total diode impedance, the frequency dependent  $Z_j$  should be small compared with the i-layer impedance. Clearly, for a given  $I_o$  and  $\omega$ ,  $Z_j$  will be as small as possible if the lifetime in the i-layer is as high as possible. The lifetime in the i-layer is 1 to 2 microseconds after processing is completed.

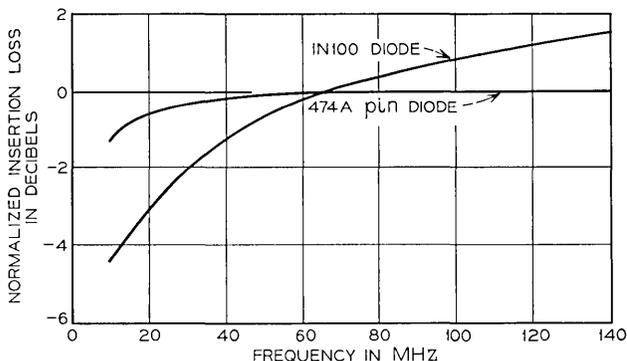


Fig. 40 — Frequency response of a commonly used diode compared with that of a specially designed pin diode.

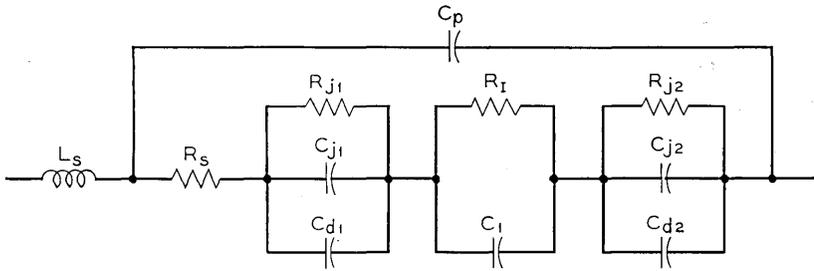


Fig. 41—Equivalent circuit of a pin diode.  $L_s$  = series inductance,  $C_p$  = package capacitance,  $R_I$  = I layer resistance,  $C_I$  = I layer capacitance,  $C_{j1}$  &  $C_{j2}$  = two junction capacitances,  $R_{j1}$  &  $R_{j2}$  = two junction resistances,  $C_{d1}$  &  $C_{d2}$  = two diffusion capacitances, and  $R_s$  = diode series resistance.

The other alternatives available to minimize  $Z_j$  are to increase  $I_o$  and to decrease the junction diffusion voltages. The diode current may be increased while keeping the total impedance constant by making the i-layer width close to  $2L_o$  ( $L_o$  is the dc diffusion length for current carriers in the i-layer). The decrease in junction impedance is compensated for by an increase in i-layer resistance. The i-layer width cannot be increased much beyond  $2L_o$  without losing conductivity modulation and thus impairing the diode variolossor action.

As a result of these considerations, the i-layer width is designed for 4.0 mils for the 474A diode. The junction diffusion voltages are decreased by decreasing the junction doping gradients. The gradient at the p-i junction cannot be decreased too much since, below some value of the gradient, a depletion layer will cease to exist and injection at the junction will be lost. The optimum gradient for the p-i junction is achieved by a complementary error function boron diffusion to a depth of 0.4 mils. The n-i junction depth is about 1.2 mils.

Care is required in the fabrication of the 474A diode to obtain the required i-layer width of 4.0 mils. Phosphorus is first diffused into p-type silicon ( $\rho > 5000 \Omega\text{-cm}$ ) to the required junction depth. Then the back side of the slice is lapped to a final thickness such that, after a boron diffusion, the final i-layer thickness is 4.0 mils. After these two diffusion steps the i-layer resistivity has decreased from greater than  $5000 \Omega\text{-cm}$  to about  $2000 \Omega\text{-cm}$ . After the boron diffusion, aluminum contacts are evaporated and alloyed to the phosphorus side of the slice. Finally, 12 mil circular mesas are etched into the slice, and the slice is diamond-scribed into wafers. The wafers are packaged using conventional wafer and wire bonding techniques in gettered TO-18 cans.

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# The Traveling-Wave Tube Amplifier for the Microwave Transmitter

By C. E. BRADFORD

(Manuscript received January 18, 1968)

*The traveling-wave tube is the final microwave power amplifier in the transmitter of the TD-3 radio relay system. The important requirements of this amplifier are to furnish 6 watts power output at 35 dB gain with less than 0.02 dB variation in gain over any 12 MHz channel, to be free of noise and distortion, and to have a long life.*

*The tube contains a Pierce-type electron gun, a helix slow-wave structure with a sputtered tantalum lossy section, and a conduction-cooled uni-potential collector. This new radio-relay tube is packaged with a periodic permanent magnetic circuit as an integral unit, which eliminates the need for beam focusing adjustments in the field and reduces the size of the amplifier.*

## I. INTRODUCTION

Because a traveling-wave tube can provide substantial, stable, microwave power output at high levels of gain with very little noise or distortion, it is used as the transmitting amplifier for the TD-3 radio-relay repeater. The 461A traveling-wave tube, manufactured by the Western Electric Co. (for Bell System use only), has been developed to furnish 6 watts power output at 4 GHz with 35 dB of gain.\* The same tube is used for all channels in the TD-3 frequency band. Adequate flatness of the gain-frequency characteristic over a given channel is achieved at the time of tube installation by a tuning adjustment in the waveguide adjoining the tube.

The tube operates at 2 to 3 dB below maximum power output to keep intermodulation effects resulting from gain saturation low. At this level, the amplifier is essentially linear in operation with distortion products down 15 dB or more.

It is desirable that a traveling-wave tube for a radio relay system

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\* Losses after the TWT may cause the repeater output to be as low as 5 watts.

be a completely packaged device, so that it can be given the optimum focusing-field adjustment at manufacture with no further adjustments required in the field. This avoids the need for special test equipment in the field and duplication of tube adjustment, thus lowering costs. The package should be returnable at the end of tube life so that the magnetic structure can be reused. It was therefore decided to design a device that is packaged during manufacturing.

This device uses the periodic permanent magnet technique for focusing the electron beam.<sup>1-3</sup> Figure 1 shows the entire amplifier

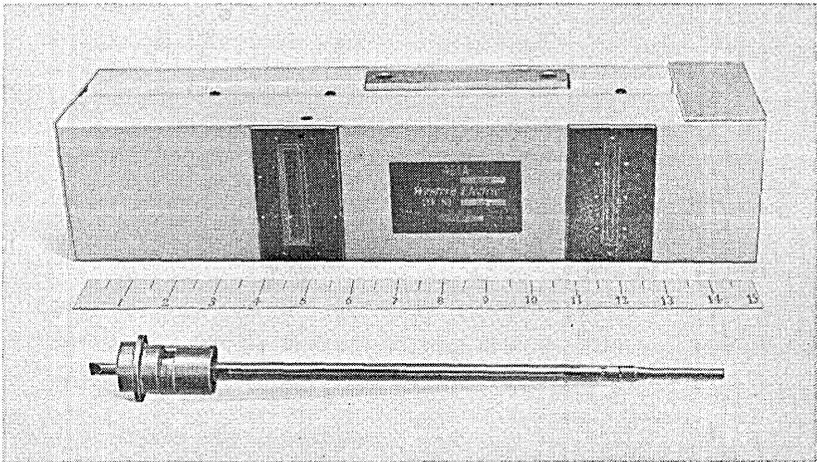


Fig. 1 — The 461A packaged traveling-wave tube.

which measures  $3\frac{1}{2} \times 4 \times 16$  inches and weighs 16 pounds. Below the ruler is the vacuum tube\* itself which contains the active elements of the amplifier.

System reliability is obtained through a combination of long tube life and provisions to monitor anode voltage for obtaining advance warning of tube failure. This allows scheduling tube replacement at a time when the channel is not in service. Improved cleaning and degassing procedures are used during manufacturing to obtain long life; whereas careful monitoring of shifts in tube characteristics during operation yield accurate failure warning.

RF input and output connections to the tube are made by non-standard reduced-height waveguide. Tuning-screw impedance-match-

\* By standard nomenclature, the entire device is an "electron tube."

ing devices, needed to obtain 0.02 dB (maximum) transmission flatness over the channel bandwidth, are part of the microwave transmitter rather than of the tube.

Power to run the tube is provided by a solid-state regulated high-voltage supply,<sup>4</sup> which in turn is driven by the nominal -24 volt dc input to the TD-3 bay.

Heat dissipated from the electron beam impinging on the collector of the tube is conducted to a heat sink which is an integral part of the TD-3 bay.

Table I summarizes the system requirements upon the 461A tube and typical device performance.

TABLE I—461A TWT REQUIREMENTS AND PERFORMANCE

	System requirement	Typical performance
Operating frequency, GHz	3.7-4.2	3.7-4.2
Operating power output, watts	{5.9	6.2
Saturation power output, watts	—	11
Gain at operating power output, dB	32-38	35
Gain flatness in any 12-MHz band from 3.7 to 4.2 GHz, dB	0.02 max	0.02 max
Input return loss (with tuner), dB	—	20-25
Amplitude-to-phase-modulation conversion at operating power output, degrees per dB	4 max	3.1
Product of gain and noise figure, dB	62 max	59.4

The high-level spurious noise is required not to exceed by more than 10 dB the thermal noise (with the noise bandwidth fixed by the FM detector) in the output of the tube at low-level drive. This is typically the case.

## II. ELECTRICAL DESIGN CONSIDERATIONS

Although the 461A tube's overall performance is superior to the 444A traveling-wave tube used as the transmitter power amplifier in the original TH radio relay system,<sup>5</sup> there are similarities in the electrical requirements for the two tubes.\* Because considerable experience has been gained with the design of the 444A and with variations of this design, certain of the 461A design parameters have been chosen to be similar to corresponding 444A parameters. The beam current, the cathode current density and the beam voltage (helix-to-cathode potential) were chosen to be approximately the same as for the 444A.

\* A prototype of the 444A, the M1789, is described in Ref. 6.

### 2.1 *Electrode Voltages and Beam Current*

It is desirable to depress the collector potential below that of the helix to reduce the beam power dissipated on the collector. Furthermore, since the collector must be cooled, it is convenient to operate it at ground potential. Because a depressed collector operates typically midway between cathode and helix potentials, the depression and grounding of the collector also simplifies power supply design by keeping the helix-to-ground and cathode-to-ground voltages low. Corona problems in power supplies become troublesome as voltages exceed about 1500 volts. For the 461A it was decided that the cathode should be operated at  $-1400$  volts, the helix at  $+1200$  volts and the anode at  $+1400$  volts. With the beam voltage of 2600 volts, a beam current of 40 mA is sufficient for the required power output.

### 2.2 *Other Parameters*

With the beam voltage and beam current established, the other principal design parameters are:

- Cathode current density  $\approx 200$  mA/cm<sup>2</sup>
- Beam-to-helix diameter ratio,  $b/a \approx 0.5$
- Cathode temperature = 740°C (true)
- Helix diameter,  $a = 0.1023$  inch.

From these parameters, one may calculate values of helix<sup>7</sup> and gun<sup>8</sup> design constants. For the desired beam voltage, the helix must have 30 turns per inch and is calculated to give 9.8 dB gain per inch. This sets the required active length\* at 5.33 inches. Adding the attenuation section and transitional turn makes the total helix length 6.86 inches.

With these values of helix and beam diameter, the gun should be designed to produce a minimum beam diameter (for 95 per cent of the beam current) of 0.0410 inches. This leads to gun design values of 12° 25' for the convergence half-angle  $\theta$  and 0.256 inches for the cathode-anode spacing,  $d$ .

### 2.3 *Magnetic Focusing*

In his work on periodic permanent magnet focusing, K. J. Harker demonstrated that the magnetic flux required for minimum beam ripple is related to the space-charge density and the magnetic shielding of the cathode.<sup>9</sup> Using his curves, minimum ripple in the 461A

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\* It is necessary also to allow for losses, particularly in establishing the circuit wave initially.

should occur when the peak value of magnetic flux is about 325 gauss. This theoretically optimum value of peak flux density calls for perfect beam launching and a perfectly straight magnetic structure. This is impractical, and little deterioration results in practice from the use of higher-than-optimum flux densities.

An extensive study of magnet structures giving a variety of fields has resulted in the adoption of an empirical value of 1000 gauss for the peak flux density in the 461A.\* This value limits the helix intercept current to less than  $\frac{1}{4}$  per cent of the beam current over a wide range of beam currents and voltages. The magnetic alloy Alnico 8 is used. This alloy has the advantage of not requiring compensation for temperature changes. The focusing field has a period of 0.550 inch with a polepiece inside diameter of 0.312 inch.

### III. CONSTRUCTION

New techniques were developed for more economical and accurate helix construction, improved cooling, and to permit the use of reduced-height waveguide.

#### 3.1 *The Helix*

The helix consists of 0.010-inch molybdenum wire, supported by three ceramic rods glazed to each helix turn. Helices are constructed by winding the helix on a mandrel, releasing the helix from the mandrel by partially annealing the wire and allowing it to "spring back" from the mandrel.

However, if the helix is not allowed to spring back, three advantages occur. First, it is cheaper to make, because fewer hand operations are involved. Second, higher electronic efficiency results because the exact pitch is maintained over the entire helix length (comparatively large dimensional changes accompany springback).

The input waveguide and output waveguide are a fixed distance apart. Best match is obtained when the ends of the helix are located the same distance apart and then positioned to be located at the respective waveguide. Therefore, a third advantage for a no-spring-back helix would be that an exact length could be more easily maintained, thereby simplifying the problem of obtaining simultaneously RF matches at both the input and output waveguides. This is very important with the 461A tube because of the increased difficulty of matching brought on by use of reduced-height waveguide.

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\* Corresponding, in Mr. Harker's nomenclature, to  $\alpha/\beta > 3$  and  $\beta < 0.06$ .

Therefore, a no-springback helix has been incorporated into the 461A design. After being wound, the helix is prevented from springing back by a weighted fixture. The mandrel on which the helix is wound is coated with aquadag, which burns off during the glazing of the wire to the rods. With the aquadag coating gone, the mandrel can be removed even though springback has not been allowed.

### 3.2 *Applied Helix-Loss*

In order to keep the amplifier from oscillating, it is necessary to deliberately apply a lossy material to a section of the helix. Historically, a favorite material has been carbon, often in the form of sprayed-on aquadag. Usually, the input and output sections of helix are isolated by as much as 80 dB of loss.

Experience in the TH radio relay system has shown that many of the 444A traveling-wave tubes began to oscillate after 10,000 hours of operating life. This was traced to a decrease in the helix loss, presumably because of oxidation of the aquadag.

To avoid this problem in the TD-3 and TH systems, the lossy material used in the 461A and 444A is tantalum applied by sputtering. The required loss is easily obtained and all evidence to date shows that the helix loss does not deteriorate with use.

### 3.3 *The Periodic Permanent Magnet*

The advantages of a periodic permanent magnet structure are elimination of the need to adjust beam focus in the field and reduction of amplifier size. The periodic permanent magnet is assembled by stacking rings of Alnico 8 magnetized with alternating polarity between polepiece rings so that the field reverses at each polepiece. The input and output waveguides, which must carry RF energy through the stack, are made with reduced heights of 0.200 and 0.100 inch, respectively, to disrupt the stack geometry as little as possible.\* Each waveguide is introduced between two polepieces. The polepieces used in the vicinity of the waveguide have offset flanges (see Fig. 2) to allow a maximum magnet thickness at the waveguide, without a change in the magnetic period. Compensation for leakage flux is acquired by adjusting the degree of magnetization of individual magnets.

The polepieces are aligned concentrically with a precise mandrel. The stack is potted by applying epoxy adhesive externally and is

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\* Standard waveguide for this frequency has a height of 1.145 inches.

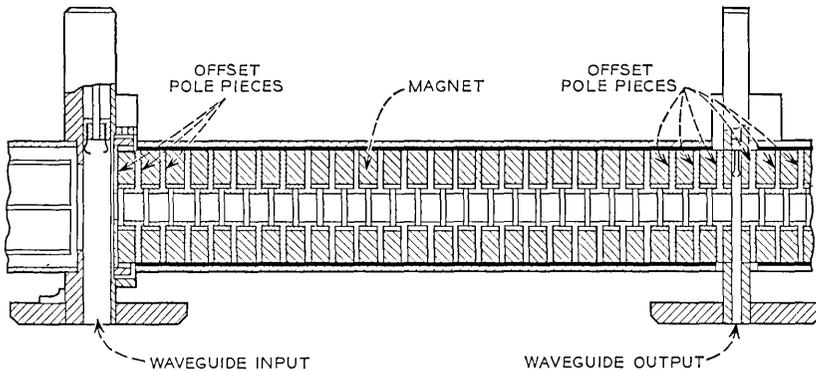


Fig. 2 — A cross section of the periodic permanent magnet assembly.

bonded with epoxy into monel supporting covers. No fasteners are used.

A special gun-end magnet is required to reduce the leakage field strength to 18 gauss at the cathode position. This was found to give best noise performance. In Fig. 3, typical curves for noise, helix intercept current, power and gain are shown as a function of peak flux density.

### 3.4 *Electron Gun*

The final 461A design configuration includes a spherical surfaced cathode in the electron gun. Previous devices were made with a "coned cathode." The effect of the cone at the center of the cathode was to produce a semihollow beam and thereby reduce the magnitude of ion oscillations which could modulate the beam. The cone was used in tubes using a uniform focusing field, but it was found in the course of this work that the cone is incompatible with periodic permanent magnet focusing, and increases ion noise, thermal noise, and helix intercept current.

Figure 4 shows the electron gun structure and a choke used to prevent R.F. power from leaking out of the tube through the gun.

### 3.5 *Cooling Block*

The TWT electron collector intercepts the 0.040 ampere beam at about 1400 electron volts so that some 56 watts must be dissipated. In the interest of preventing spurious ion oscillations and for the

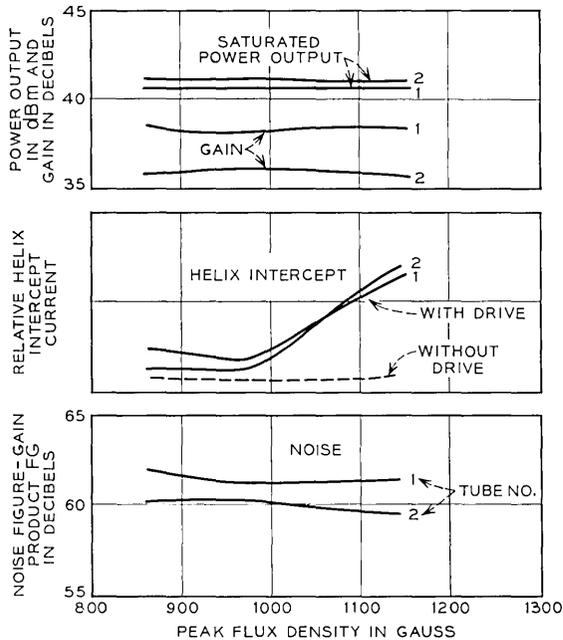


Fig. 3—The effects of the magnitude of the peak flux density on several characteristics.

sake of long life, the collector must be maintained at a reasonably low temperature (below  $150^{\circ}\text{C}$ ).

It was decided that rather than to use forced air, as is done in the TH system, the 461A collector should be cooled by conduction to fins mounted on the TD-3 bay.

In order to allow alignment of the electron beam with the magnetic field, the hole in a cooling block is made 3 mils larger than the collector. The small annular space between the collector and cooling block is filled with a silicon heat-conducting paste which provides excellent heat conduction from the collector to the cooling block.

The collector is held radially by a reference surface located between the magnet stack and the cooling block as shown in Fig. 5. The aluminum cooling block is itself conductively coupled to cooling fins which are part of the TD-3 bay.

### 3.6 Enclosure

The package is enclosed in a sheet steel container for RF shielding, strength, and mechanical protection.

## IV. PROCESSING

The tube parts are kept in dustproof super-clean areas. All parts are water-washed with ultrasonic agitation. How long an unbroken film of pure water remains over the surface of the part after it is withdrawn from the bath is used as an indication of cleanliness. All parts are further cleaned by heating in wet hydrogen. The subassemblies of the gun and the helix are degassed by induction heating in a vacuum furnace. Normally the temperatures used for induction heating are limited by the softening points of glaze joints, braze joints, or ceramics.

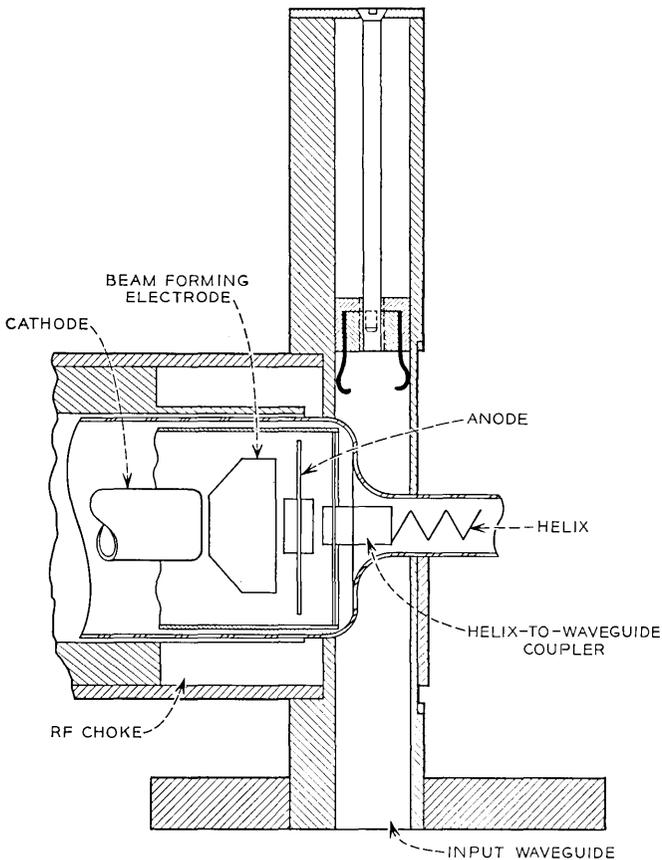


Fig. 4—A cross section of the electron gun region of the 461A tube.

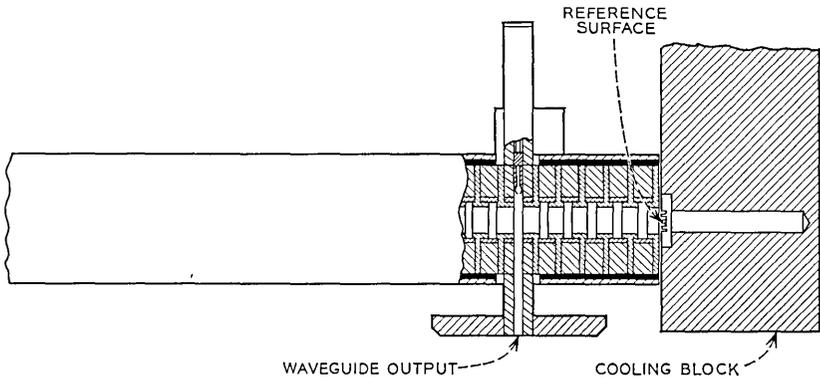


Fig. 5 — Cross section of the magnet stack, the cooling block assembly.

## V. TUBE PERFORMANCE

### 5.1 Power Output and Gain

Although high efficiency is not a primary design objective for a radio relay tube, (especially since tube cost tends to be increased by designing for greater efficiency) the efficiency of the 461A for saturated operation is a quite reasonable 20 per cent. The saturated efficiency of the 444A tube used in the TH system is about 21 per cent. The traveling-wave tube in the *Telstar*<sup>®</sup> communications satellite was specifically designed for high efficiency and gives in excess of 30 per cent saturated efficiency.<sup>10</sup> The desired gain has been obtained with the typical tube-to-tube variation illustrated in Fig. 6. Through the use of waveguide tuners at the input and output ports of the tube, the transmission characteristic over any 15 MHz band can be flattened to less than 0.02 dB variation peak-to-peak.

The input and output matches must be such that reflections do not upset the characteristics of the bandpass filters facing the tube. For this reason, isolators are used at the input and output of the tube. However, the need to avoid gain ripple makes it important to achieve a good match at the input. The input return loss in the 461A is typically greater than 25 dB when the waveguide tuner is used.

### 5.2 Thermal Noise

Since TD-3 is a frequency-modulated system, limiters are used in every repeater to remove the AM component of noise modulation. Therefore, only FM noise is important.

Thermal noise has been determined by the standard technique of

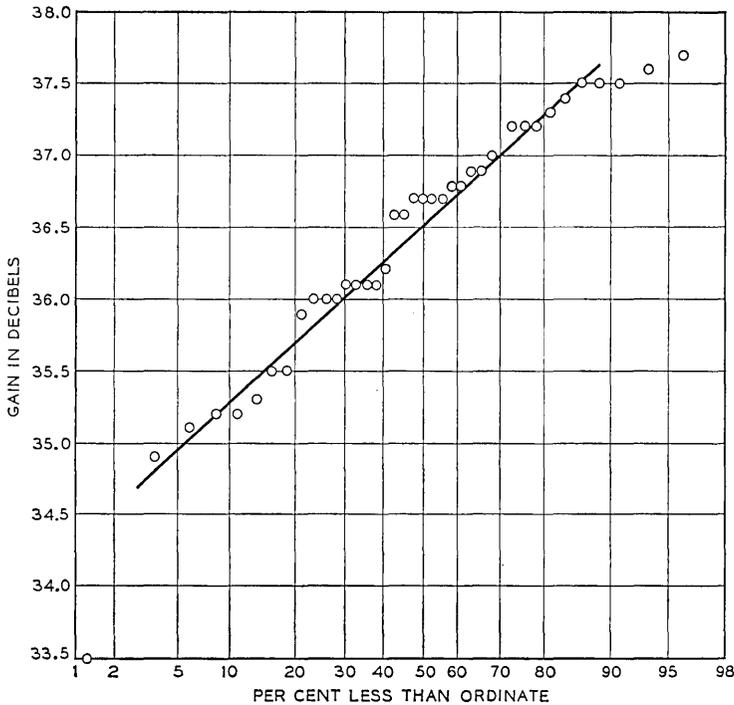


Fig. 6 — Distribution of 461A gain at midband.

measuring the noise power output through a 10 MHz bandpass filter. However, it is necessary to know the noise figure in the presence of a carrier at high level drive. The maximum carrier-to-noise ratio of which the 461A tube is capable is about 156 dB per hertz bandwidth. It is apparent from this that care must be used to filter out the carrier when measuring noise in the presence of a carrier. To do this, one sets the filter frequency away from carrier frequency making certain to obtain at least 100 dB of isolation at the carrier frequency.

The gain at the frequency ( $f_o$ ), where noise power is measured, will be compressed because of the presence of the carrier at ( $f_c$ ). If the ratio of low-level gain ( $= G_o$ ) to compressed gain is designated as  $C$ , and the high-level noise figure is designated as  $F_{38}$ ,\* then

$$\frac{N_{oc}}{kTB} = \frac{F_{38}G_o}{C}$$

\*The designation  $F_{38}$  is used because the nominal output power is 38 dBm (6-1/3 watts) at the output port.

where the output noise power is  $N_{oc}$ , measured through a filter of bandwidth  $B$  ( $k$  is Boltzmann's constant and  $T$  is absolute temperature).

The noise figure has also been determined by using the output of an FM detector and has been shown to be the same as that obtained using the method first described. Figure 7 shows the effect of changes in helix voltage on noise figure  $F$ , gain  $G$  and gain-noise figure product  $FG$ .

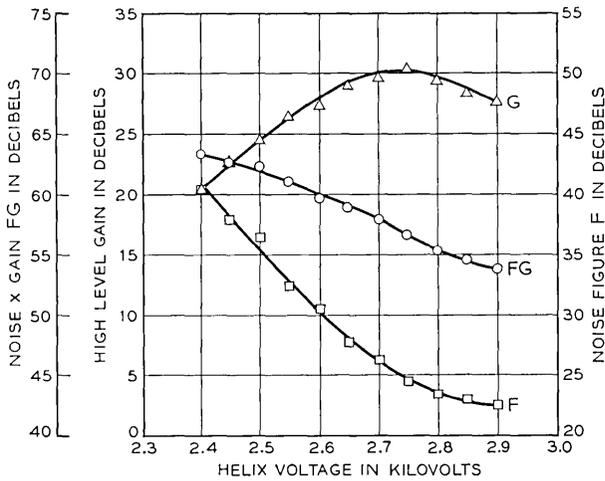


Fig. 7 — Typical variation of noise and gain with helix voltage.

The cathode must be immersed in a small magnetic field of about 18 gauss in order to inhibit the growth of a space-charge wave of noise. The nature of the variation of low-level noise and helix intercept current with cathode field is shown in Fig. 8. The measurements were taken using a coil over the tube in the cathode region. Increasing the cathode field coil current beyond 250 mA, which corresponds to a cathode field of 10 to 15 gauss, has no further effect on the gain-noise figure product  $FG$ ,\* even though noise figure continue to drop, because of the simultaneous increase in gain.

### 5.3 Spurious Modulation

The ion noise power in the output of the TWT is measured with respect to the thermal noise output with an FM detector. Thermal

\* More important to system design than noise figure.

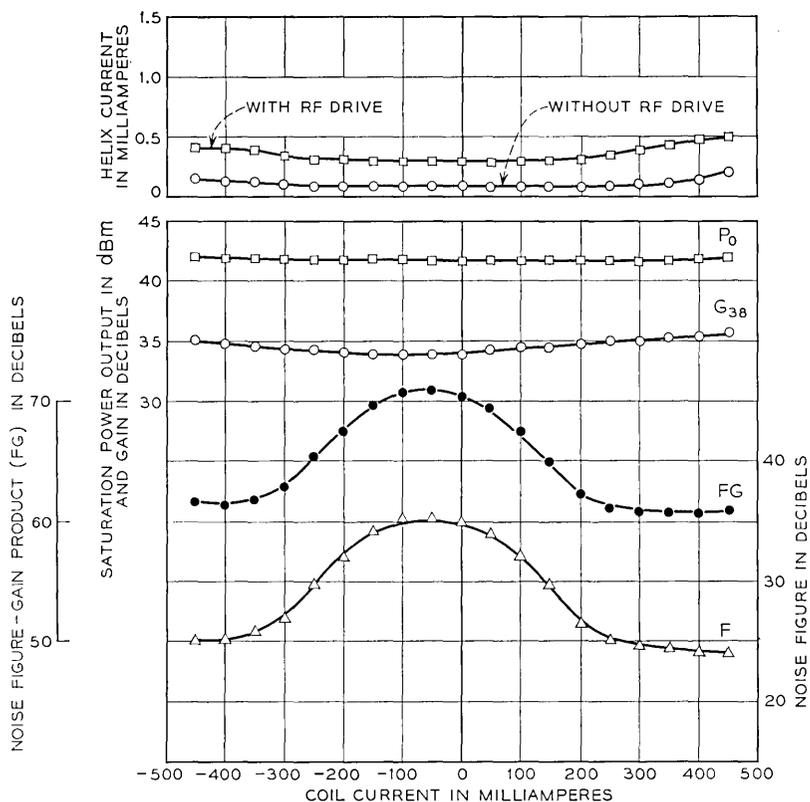


Fig. 8—Noise and power versus cathode field.

noise power measured at baseband at the output of an FM detector varies in magnitude as the square of the baseband frequency. Ion noise will generally appear as a series of spikes riding above thermal noise.

Figure 9 is a plot of thermal noise vs tuned frequency of the noise analyzer. An ion-noise spike is evident at a baseband frequency of 7 MHz and a power output of 38 dBm. Since the response of an FM detector to a thermal noise input is proportional to the ratio of noise to carrier power, the magnitude of ion noise can be fully specified as being so many dB above thermal noise, for the detector bandwidth to be used, and for the nominal output power.

Ion noise appears in the output of most 461A tubes and is reduced to an acceptably low level by aging. The outgassing of helix parts by

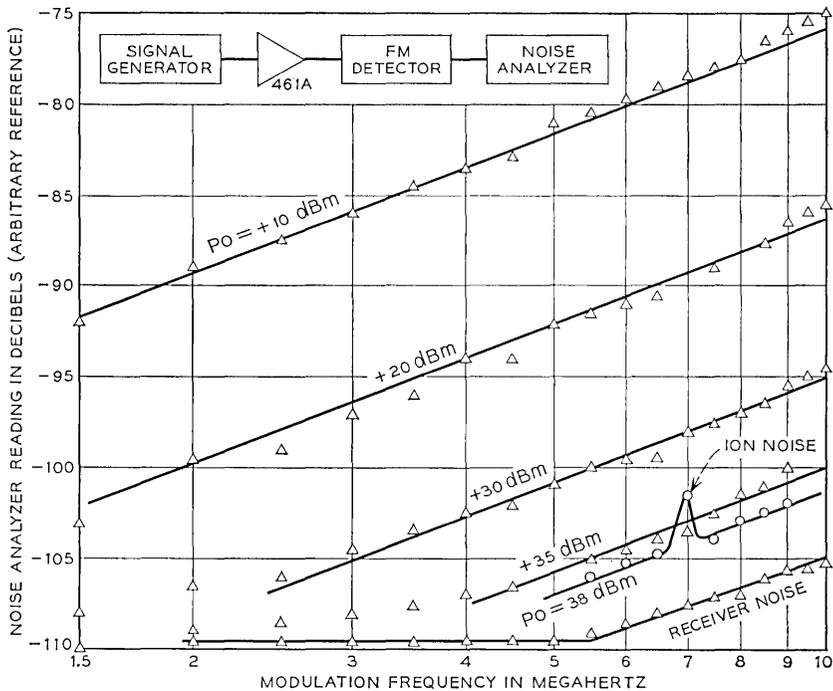


Fig. 9—Thermal noise characteristics of a 461A TWT showing an ion noise spike at 7 MHz.

electron bombardment is the principal purpose of this aging process, which typically takes 100 to 200 hours.

Intermodulation in general refers to the fact that in active devices, it is normal to find extraneous frequencies in the output signal. For instance, if two signals enter the device at  $\omega_A$  and  $\omega_B$ , one finds in the output signals at  $\omega_A$ ,  $\omega_B$ ,  $2\omega_A$ ,  $2\omega_B$  and intermodulation products at  $2\omega_A - \omega_B$ ,  $2\omega_B - \omega_A$ , and so on. Signals at  $\omega = 2\omega_A - \omega_B$  are known as third order distortion products and those at  $3\omega_A - 2\omega_B$  as fifth order distortion products, and so on. The harmonics at  $2\omega_A$  and  $2\omega_B$  can be removed from the output by using a low-pass filter.

If one assumes that only distortion products of the  $2\omega_A - \omega_B$  type are obtained, by measuring them one can estimate the degree of AM-to-PM conversion as a measure of intermodulation.\* This has been done using two input signals, one at  $\omega_A$  and the other at  $\omega_B$ .

\* Although the basic device phenomenon is intermodulation, in systems work the parameter of concern is AM/PM conversion. Also see Ref. 6.

For output power levels ranging from 36 to 40 dBm, the AM-to-PM conversion in 461A tubes was found to be between 2 and 5 degrees per dB amplitude modulation.

#### 5.4 *Life and Prediction of Failures*

In order to obtain good reliability for the system, sudden unpredictable failures should be infrequent. The traveling-wave tube is expected to have a field lifetime between 30,000 and 50,000 hours. Since other system components have expected lifetimes upwards of 200,000 hours, the end-of-life of the traveling-wave tube will be the most frequent cause of a system channel going off the air. System reliability can therefore be substantially improved if the failure of the traveling-wave tube can be predicted. A soon-to-fail tube can then be replaced by a new one while the channel is not being used.

In the TH radio relay system, a dip test has been used to predict imminent failure. This consists of determining the dip in beam current after suddenly reducing the heater voltage to zero, then restoring it to its original value after a specified number of seconds. However, studies of the TH system have shown that the approach of cathode failures can be predicted better on the basis of a record of the increase and rate of increase of anode voltage needed to give a 40 mA beam current during life. Therefore, an accurate anode voltage meter has been provided in the TD-3 TWT power supply, to permit a voltage record to be kept for each tube.

#### 5.5 *The Power Supply*

The power supply<sup>4</sup> provides a well-regulated heater and helix voltage. The dc heater voltage is regulated to  $\pm 2$  per cent, which allows the cathode to operate at lower temperatures for long life without risking the instability which can occur if the cathode temperature becomes too low. The heater ripple is kept below 0.1 volt to prevent heater ripple from appearing as gain ripple. The anode and helix regulation is made small enough ( $\pm 45$  volts,  $\pm 12$  volts, respectively) to permit the gain setting to be held to within  $\approx 0.3$  dB. However, the ripple requirement (1.8 volts, 0.4 volts) is governed by the requirement that the phase deviation resulting from the ripple be 67 dB below the signal deviation, as required for television. Collector voltage requirements are predicated mainly on an allowable increase in helix current over that at the optimum collector voltage.

Proper start-up contributes to the long life of the tube. To accom-

plish this, a built-in timer delays the application of high voltage and provides an overvoltage for the heater, so that when the high voltages are automatically applied, there will be sufficient cathode emission for good beam focus. This is especially important for older tubes.

#### VI. CONCLUSION

The desired TD-3 repeater power output of 38 dBm is obtained with a noise output of about 59 dB (24 dB noise figure) at a gain of 35 dB. The intermodulation distortion at this power level is equivalent to about 3 degrees per dB.

To achieve long life, the device has been conservatively designed at a moderate cathode loading. It is expected, from our experience with field performance of the 444A tube, and our laboratory life tests of prototype tubes, to achieve a field life of about 40,000 operating hours. In order to obtain optimum convenience in service, the device has been packaged for easy installation; periodic permanent magnet beam focusing is used to reduce weight and ease handling.

#### VII. ACKNOWLEDGMENT

A project of this complexity represents the work of many development engineers of the Reading Laboratory, each responsible for a vital contribution. We give a special word of thanks for the efforts of Messrs. L. K. S. Haas and H. P. Ross. These gentlemen were responsible for transforming the prototype device into the final design for manufacture, and for many valuable features of that final form.

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# Networks

By E. J. DRAZY, R. C. MacLEAN and R. E. SHEEHEY

(Manuscript received March 7, 1968.)

*This article describes the design and development of the microwave and intermediate-frequency filters, which provide the frequency selectivity required in the TD-3 microwave transmitter and receiver, and of the equalizers, which compensate for amplitude and delay distortions introduced by the filters.*

## I. INTRODUCTION

As described in the first article of this issue, all selectivity required by the microwave transmitter and microwave receiver of the TD-3 radio system is provided by passive networks.<sup>1</sup>

The general plan followed in the design and development of these networks was to:

(i) Determine, from the system requirements, the preliminary in band and out of band loss requirements for the microwave networks.

(ii) Design microwave networks meeting the loss requirements with a minimum of amplitude distortion and with only broad-shape types of amplitude variation over the passband. The return loss requirement of the networks was generally specified as 30 dB minimum.

(iii) Calculate and measure the in-band loss distortions of the microwave networks to set preliminary requirements for the IF equalization networks.

(iv) Develop a preliminary design of IF equalizer meeting these preliminary requirements in order to perform system tests to establish that the network characteristics were satisfactory.

(v) Construct and measure characteristics of a substantial number of models of the microwave networks which introduce distortions. These results were used to specify the actual requirements of the IF equalizing networks.

(vi) Redesign the IF equalizing networks to compensate for the nominal distortions introduced by the microwave networks.

(vii) Establish the IF selective network requirements and design the network to meet the systems requirements. Equalization sections were included to compensate for the in-band distortions of the filter sections.

## II. MICROWAVE TRANSMISSION NETWORKS

Frequency selective microwave transmission\* networks are required in the TD-3 microwave transmitter to select the desired sideband output of the transmitting modulator, to suppress unwanted modulation products which might fall within the passbands of microwave receivers operating on other channels, to provide a means for efficiently coupling as many as six microwave transmitters to a common waveguide output while providing sufficient isolation to prevent interaction, and to prevent the emission of harmonics and other modulation products which might interfere with other radio services.

In the microwave receiver, selective microwave networks are required to efficiently couple as many as six receivers to a common waveguide, to prevent adjacent channel interference<sup>2</sup> and to suppress interfering frequencies which would degrade the system's performance.

### 2.1 *Transmitter Microwave Transmission Path*

The location and nature of each selective microwave component in the transmission path of the TD-3 microwave transmitter is shown in Fig. 1a.† In Fig. 1a, the 1336 type bandpass filter transmits the desired sideband from the transmitting modulator to the TWT amplifier, but suppresses the carrier leak, undesired sidebands, and other out-of-band modulation products and noise which would otherwise degrade the performance of the system. The 1336 type bandpass filter following the TWT provides most of the adjacent- and adjoining-channel attenuation,‡ the 1326A low-pass filter prevents transmission of harmonics of the modulated carrier, which might otherwise interfere with services operating on harmonically related frequencies or, by modulation with corresponding harmonics from other bays, produce interfering tones within the TD-3 system itself. Finally, the 1418 type channel combining network§ provides additional adjacent- and ad-

\* We define transmission networks as those which pass the modulated signal.

† Fig. 1 includes only those components essential to this discussion. For complete block diagrams, see Ref. 3.

‡ Adjoining channels are considered to be those within the 3.7 to 4.2 GHz common-carrier band. At frequencies appreciably higher than those within this band, the performance of narrow-band waveguide filters is unpredictable because of the possibility of multimode propagation.

§ For reasons discussed in Section 2.3, the channel-combining and channel-separation networks are identical.

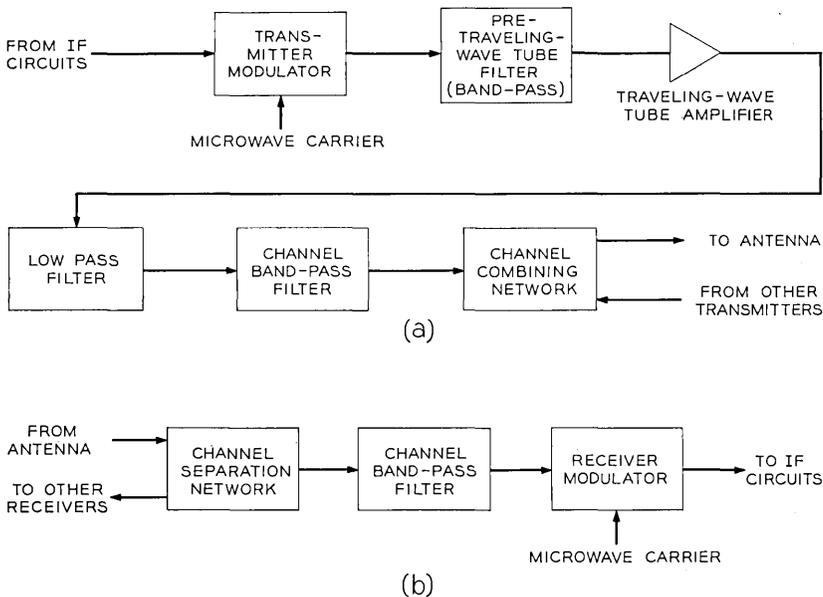


Fig. 1—Location of selective microwave transmission components in TD-3 microwave (a) transmitter and (b) receiver.

joining-channel selectivity and couples the transmitter to the common antenna-feed waveguide without substantial loss.

### 2.2 Receiver Microwave Transmission Path

In the microwave receiver, as shown in Fig. 1b, the 1418 type channel-separation network selects the desired channel, while passing the remaining ones on to the appropriate receivers via the common waveguide. The 1322 type bandpass filter provides supplementary adjacent- and adjoining-channel selectivity.

### 2.3 Requirements, Constraints and Practical Considerations

Identical in-band transmission objectives were established for each of the selective transmission networks, namely, a ripple-free insertion loss characteristic with a value no greater than 0.5 dB at midband, flat to  $\pm 0.01$  dB within the range  $f_0$  (the midband frequency)  $\pm 6$  MHz, and to  $\pm 0.10$  dB within the range  $f_0 \pm 10$  MHz.

Similarly, common objectives governing permissible variations of network performance because of changes in ambient conditions were specified for all microwave networks as follows in response to any temperature change within  $+40^\circ$  to  $+140^\circ\text{F}$ :

(i) The change in envelope delay distortion of any filter characteristic shall be less than 0.1 nanosecond within the range  $f_0 \pm 6$  MHz.

(ii) The in-band insertion loss versus frequency characteristic of any selective network shall shift by no more than  $\pm 1.0$  MHz.

(iii) At frequencies of  $f_0 \pm 70$  and  $\pm 80$  MHz, the insertion losses of the selective networks shall decrease by no more than 10 dB.

The first two of these requirements are necessary to ensure that the amplitude and delay characteristics of the microwave networks are at all times complementary to those of the correcting amplitude and delay equalizers, which operate at the intermediate frequency, and so exhibit relatively insignificant variation of characteristics with changes in ambient conditions. The third ensures that adjoining-channel interferences will not become intolerable. The first two requirements proved to be governing, and about equally restrictive, in all cases.

The out-of-band selectivity requirements for both the microwave transmitter and receiver were specified in terms of the total discrimination of the channel separation network and channel bandpass filters together; the distribution of these losses among the networks was left to the discretion of the networks' designer. In addition to those stated in the preceding paragraph these requirements were, for the transmitter, discrimination of at least 20 dB at frequencies  $f_0 \pm 20$  MHz, and at least 80 dB at frequencies  $f_0 \pm 70$  MHz; and for the receiver, discrimination of at least 20 dB at frequencies  $f_0 \pm 20$  MHz, and of at least 100 dB at frequencies  $f_0 \pm 80$  MHz.

Notice that the requirements for the transmitter and receiver can be fulfilled by very nearly identical sets of filters; the economics resulting from using actually identical sets proved to outweigh the disadvantages.

To provide an easily equalizable attenuation and delay characteristic, the maximally-flat-amplitude shape was chosen as the design basis for all selective microwave networks. The allocation of out-of-band insertion loss between the channel-separation and combining networks and the bandpass filters was dictated largely by the space available for the former, which, to permit side-by-side disposition and connection of transmitter-receiver bays, are best located horizontally and transversely in the 19-inch wide bays.

#### 2.4 Channel-Combining and Separation Networks

The basic configuration selected for the channel-combining and separation networks is the constant-resistance one originally described

by Lewis and Tillotson,<sup>4</sup> and used successfully in the TD-2, TH and TJ radio systems. However, instead of the rod-and-disc resonators used in Lewis and Tillotson's filters, the resonators of the band rejection filters which form the selective elements of the networks for the TD-3 system consist of rectangular waveguide cavities, spaced along the main waveguide at three-quarter wavelength intervals, and coupled to it by circular irises.\* The cavities form filters with lower dissipative loss, improved stability, and higher passband return loss.

Asymmetry of the filter characteristic, caused by the series inductance inherent in the coupling irises, is corrected by capacitive compensating studs, centered on the broad wall of the main waveguide opposite each iris. Each resonator cavity is designed to resonate at a frequency slightly above the design center frequency of the filter and has a capacitive tuning screw, which is adjusted to resonate the cavity at exactly the design frequency during factory tuning.

To ensure minimum shifting of the transmission characteristics of these networks with temperature changes, the component filters are (except for the connecting flanges) completely fabricated from drawn WR229 waveguide tubing of copper-clad low-temperature coefficient nickel alloy (Invar). After fabrication, the filter is copper plated inside and out to 0.0002 inches thick to prevent corrosion on portions of the nickel alloy which have been exposed during machining, and to ensure low surface resistivity. A band rejection filter is shown in Fig. 2.

Available space limited the number of resonators per band-reject filter to three. This, in turn, limited the obtainable insertion loss to the separating port to about 18 dB at  $f_0 \pm 80$  MHz, while satisfying the in-band requirements. These two parameters (the number of resonators, and the nondissipative insertion loss at one frequency) are sufficient to determine the electrical design of a maximally-flat amplitude filter.

Typical insertion-loss and delay characteristics are shown in Fig. 3. A detailed theoretical and experimental investigation has established that the asymmetries largely result from errors in the lengths of the waveguide sections connecting the resonators. A more refined design procedure which greatly improves the symmetry has resulted.<sup>6</sup> The ohmic losses cause the in-band transmission to not conform with the stated requirements. Intermediate frequency amplitude equalizers compensate for this deviation (see Section II).

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\* Rectangular resonators are also used in the analogous band-rejection filters for the TJ, TH, and TM radio systems. See Ref. 5.

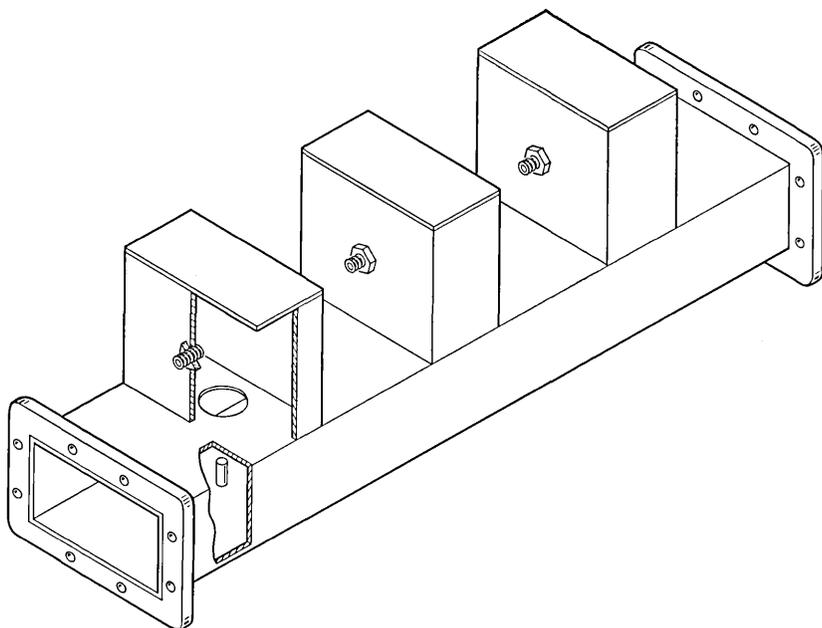


Fig. 2—Band reject filter used in channel combining and separating network.

A total of 24 codes of networks is required, one for each microwave channel. The input (4B) and output (5B) hybrid junctions are common to all codes; the band-reject filters, however, differ among codes dimensionally, as well as in center frequency, to provide the most uniform possible performance from channel to channel. Figure 4 shows a 1418 type channel separating or combining network.

### 2.5 Channel Bandpass Filters

The allocation of 18 dB of discrimination at  $f_0 \pm 80$  MHz to the channel separating and combining networks leaves a residue of 82 dB at this frequency to be provided by the channel bandpass filters. This can be provided by a five-resonator maximally flat amplitude filter, which will also satisfy the in-band transmission requirements. Such a filter becomes somewhat unwieldy when composed of the usual assembly of approximately half-wavelength resonators, bounded by single cylindrical-post susceptances and coupled by three-quarter wavelength lines (necessary to avoid spurious coupling from evanescent modes created by the susceptible posts).

Effort was therefore directed toward reducing its size without sacrificing performance. A possible alternative, the direct-coupled filter, was rejected because of the extreme mechanical precision which would be required in forming and positioning the high-susceptance coupling elements needed in such narrow bandwidth filters. A second alternative, to select a configuration of susceptible elements which would reduce coupling caused by evanescent modes and so permit closer spacing of the resonators, appeared to be more desirable.

An array of three cylindrical posts, equal in diameter and uniformly spaced across the waveguide (see Refs. 7 through 10), was found to permit reducing the coupling lines to one-quarter wavelength long. The posts have reasonable dimensions for manufacturing. Cylindrical posts were favored because they are the simplest to form within close tolerances and to position accurately within the waveguide. Data were obtained empirically relating post array dimensions to susceptance. This information, gathered with the most accurate measuring methods available, was used in the filter designs, following Mumford's<sup>11</sup> method.

After several prototype filters performed satisfactorily, an IBM 7094

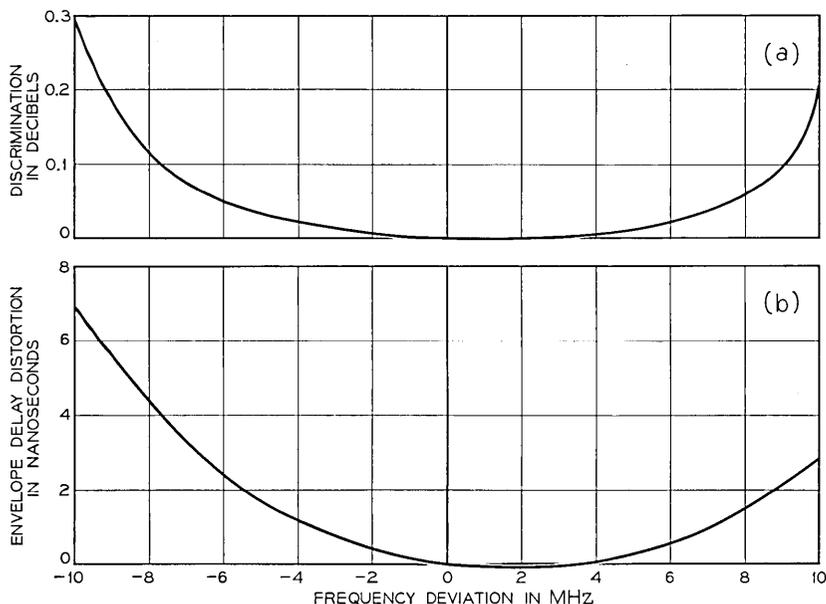


Fig. 3—Discrimination and delay characteristics of channel combining and separating networks.

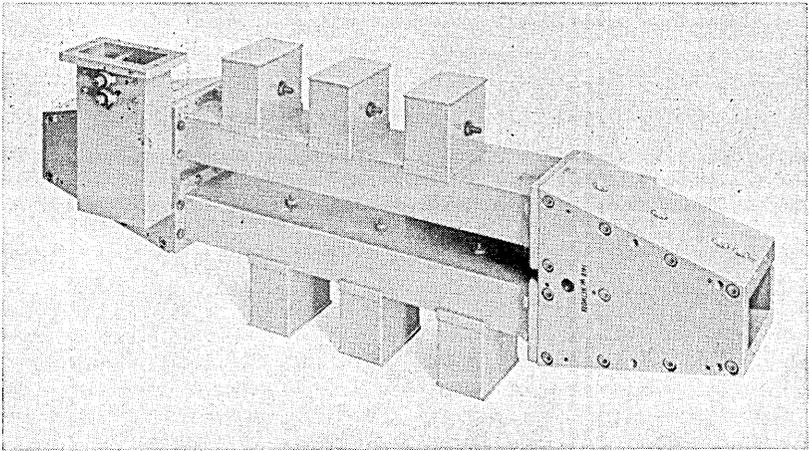


Fig. 4 — Channel combining and separating networks.

computer was programmed to mechanically and electrically design filters for the remaining channels. As with the band rejection filters, each of the 24 filter codes entails a distinct physical arrangement of elements for the greatest possible uniformity of performance from channel to channel.

Each filter consists of a section of drawn WR229 waveguide of copper-clad low-temperature coefficient nickel alloy. The posts, composed of the same alloy, are cylindrically ground to a tolerance of  $\pm 0.0002$  inch, then copper plated and inserted in holes formed in the broad walls of the waveguide by electrical discharge machining (to avoid burrs), and soft-soldered in place. After fabrication, the entire filter is copper-plated to a thickness of 0.0002 inch to prevent corrosion and to ensure a low-resistivity surface. Again, each resonator is designed to resonate slightly above the nominal center frequency of the filter and has a capacitive tuning screw to permit exact adjustment to the specified frequency.

In Fig. 5, a 1322 channel bandpass filter is cut to show the internal arrangement of susceptive posts. Figure 6 shows typical measured amplitude and delay distortion characteristics, as well as the asymmetry which is the subject of investigation. Ohmic losses also contribute to the departure from the ideal "maximally flat" characteristic, which is compensated for in the IF amplitude equalizer. The channel bandpass filters, as well as the channel separating and combining networks, have a frequency-temperature coefficient of  $-5$  kHz per degree Fahrenheit.

### 2.6 *Filter Before TWT*

The first article in this issue mentioned that excessive delay distortion in the transmitting channel bandpass filter (which preceded the TWT), and TWT amplitude to phase conversion were producing excessive modulation noise in the initial TD-3 installation.<sup>1</sup> To alleviate this the bandpass filter was moved to follow the TWT. To prevent overloading the TWT by the carrier leak and the unused sideband produced by the transmitting modulator, a very broad bandpass filter was designed and put between the transmitting modulator and the TWT. It was to be only selective enough to acceptably reduce signals while introducing negligible amplitude and delay distortion in the passband. Because its amplitude and delay distortions are very small, this filter can be made of all copper construction in a WR229 waveguide. This filter, number 1336, consists of three triple-post resonators. It provides 15 dB of discrimination at  $f_m \pm 70$  MHz, less than 0.1 ns of delay distortion, and immeasurably small amplitude distortion in the band.

### 2.7 *Low-pass Filter*

The attenuation characteristics of the bandpass filters are unpredictable at frequencies for which the waveguide containing them can propagate modes besides the dominant mode. Furthermore, even without high-order modes, the distributed element resonators of the filters will exhibit resonances at frequencies which are approximately integral multiples of the operating frequencies, and so introduce spurious passbands. Therefore, to ensure adequate attenuation of harmonic frequencies of the microwave carrier, it is necessary to provide supplementary high-frequency attenuation. A low-pass filter

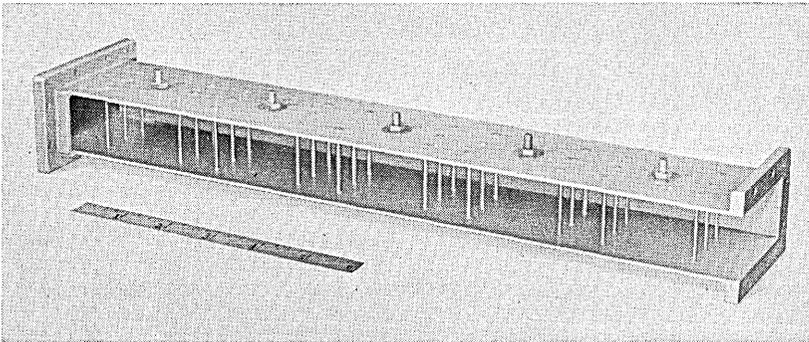


Fig. 5 — Channel bandpass filter cut to show internal construction.

which follows the highly nonlinear TWT of the microwave transmitter does this.

The requirements placed on the low-pass filter were: insertion loss less than 0.2 dB at 3.7 to 4.2 GHz, and greater than 50 dB at 7.4 to 12.64 GHz; and return loss greater than 30 dB from 3.7 to 4.2 GHz.

A coaxial structure of alternating sections of low- and high-impedance coaxial line provides the required stop-band loss. In-line coaxial-to-waveguide transducers couple the coaxial element to a WR229 waveguide, as shown in Fig. 7.

The coaxial attenuating element has seven sections, mid-shunt termination, constant K, and a nominal image impedance of 50 ohms. The inside diameter of the outer conductor is small (0.288 inches) to minimize spurious passbands from propagation in modes other than the TEM in the normally attenuating frequency range. To ensure minimum loss in the passband, only two thin discs were used as dielectric supports, one at either end of the coaxial structure. Shock and vibration tests have proven this design mechanically adequate.

The cutaway in Fig. 7 shows details of the coaxial-to-waveguide transducer. The tuneable stub and axial probe form the inductive

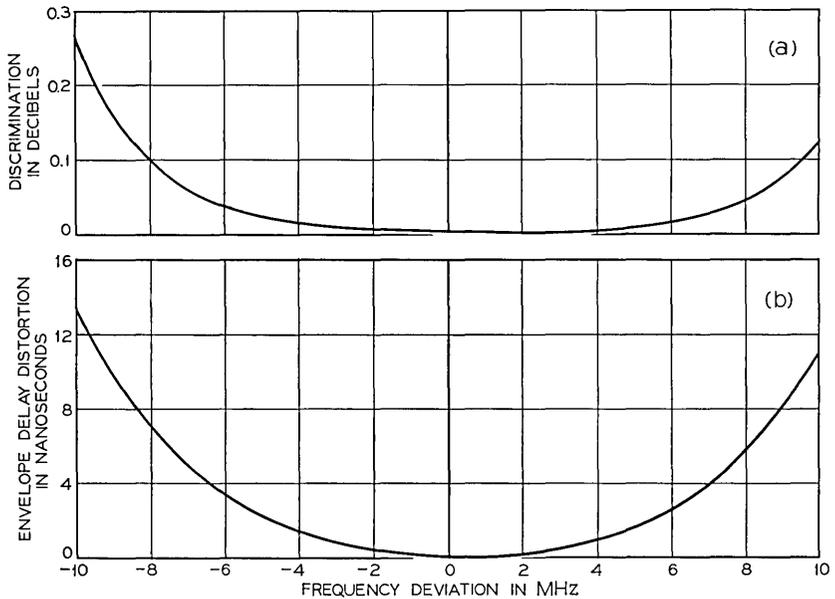


Fig. 6—Discrimination and delay characteristics of channel bandpass filter.

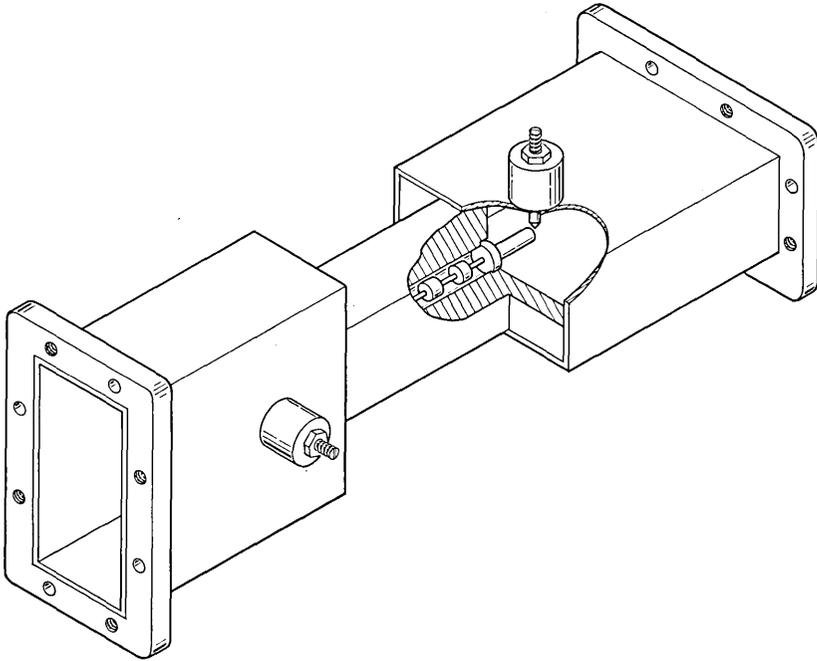


Fig. 7 — Microwave low pass filter.

elements and the air gap between them forms the capacitive element of a series-resonant coupling loop with reactances small enough to fulfill the inband return loss requirement. Besides being a convenient in-line arrangement of waveguide parts, these transducers permit the waveguide sections to rotate 90 degrees as shown in Fig. 7, thus eliminating the need for a waveguide twist, which would otherwise be required.

The insertion loss and return loss characteristics of the low-pass 1326A filter, are shown in Fig. 8.

### 2.8 *Miscellaneous Microwave Networks*

The TD-3 transmitters and receivers contain conventional attenuators, directional couplers, terminators, and a single-resonator filter. There is no need to describe them here.

## III. IF BANDPASS FILTER

An equalized IF bandpass filter, with extremely good passband performance, was required to increase the selectivity of the TD-3

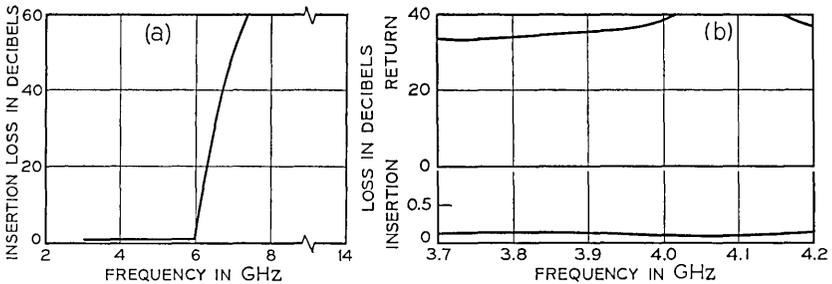


Fig. 8. (a) Insertion loss (3.0 to 12.6 GHz) and (b) insertion and return loss characteristics of low pass filter in TD-3 frequency band.

receiver. This filter, with a passband from 62-78 MHz, selects against out-of-band interference which could cause improper operation of the AGC or carrier resupply circuits. The filter allowable passband distortion has to be kept to a minimum. Therefore, the following stringent objectives were specified: a maximum peak delay distortion of  $\pm 0.3$  nanoseconds, a maximum peak amplitude distortion of  $\pm .05$  dB, and a minimum return loss of 30 dB. This filter was called 745A; its stopband objectives are shown in Fig. 9.

### 3.1 Electrical Design

Insertion loss and image parameter methods of designing the filter were explored. The insertion loss technique is a more exact method of design in that the insertion loss and return loss characteristics can be specified, and a circuit configuration realizing these characteristics is then synthesized. By the image parameter method a circuit configuration is first chosen, then the circuit parameters are adjusted until the required characteristics are obtained. The image parameter design was chosen for this filter because a circuit that better compensates for parasitic elements was more readily obtained.

The filter consists of two double M-derived, asymmetrical, transformed half-sections. The transformation was performed to obtain capacitors to ground that can be adjusted to compensate for the parasitic capacitances. Figure 10 is a schematic diagram of the filter. Figure 9 shows typical measured characteristics of the unequaled filter. The delay distortion is 12 nanoseconds, and the loss distortion is 0.3 dB over the 62-78 MHz frequency range.

A slight change in the frequency of the filter attenuation peaks by temperature and aging will cause a large change in the delay at the band edge. Therefore greater emphasis was placed on equalizing the

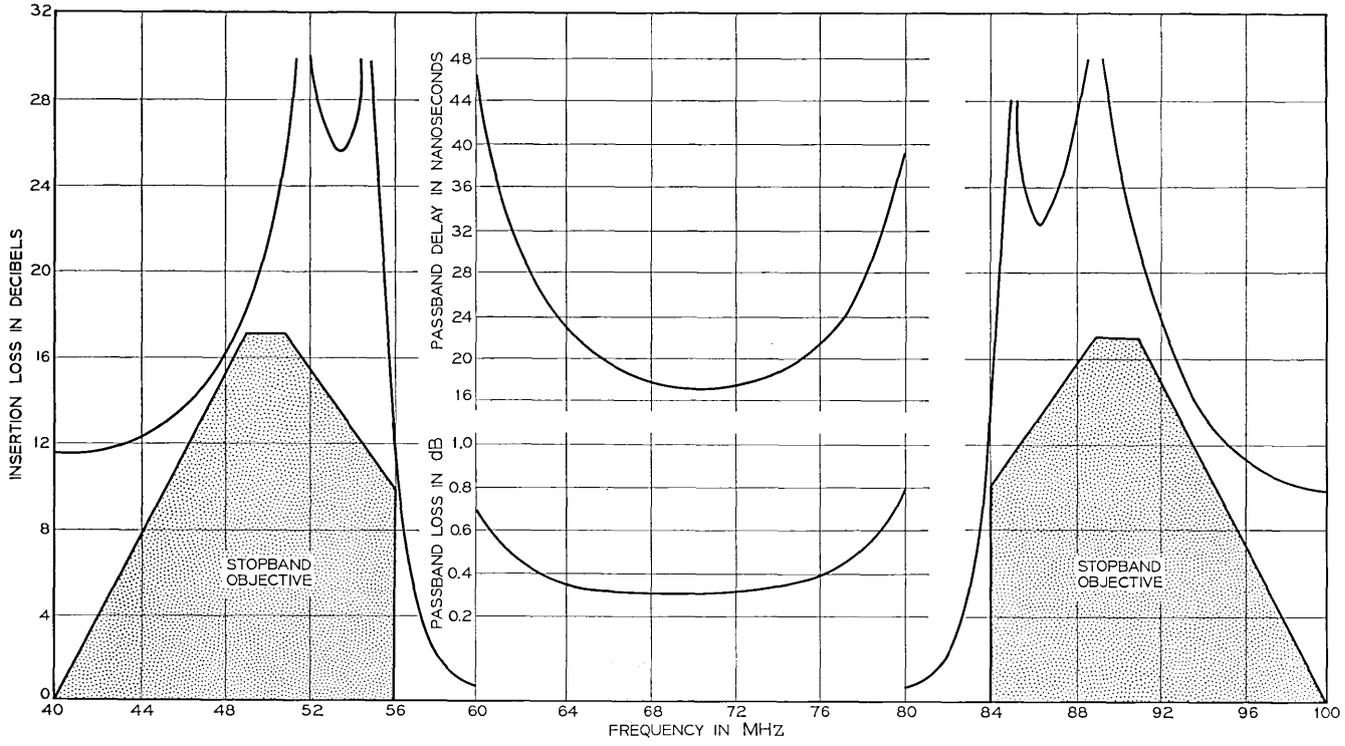


Fig. 9 — Characteristics of the unequalized 745A filter.

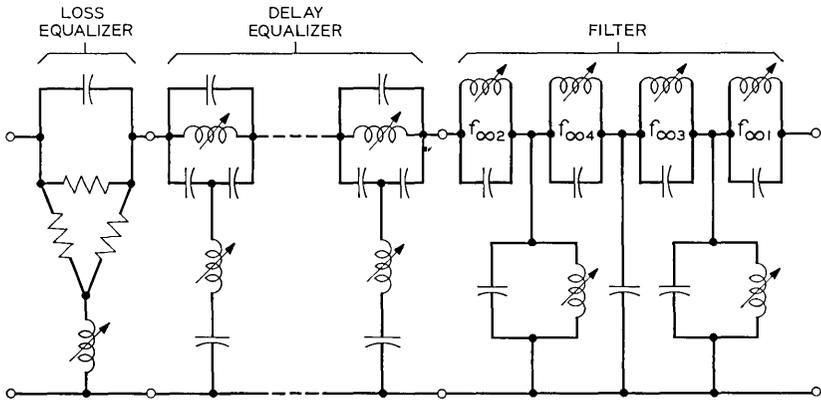


Fig. 10 — Schematic diagram of the 745A filter.

filter over the 64-76 MHz frequency range than over the 62-78 MHz frequency range. Four bridged-T  $360^\circ$  all-pass sections are used to equalize the delay distortion of the filter to within the specified objectives. One bridged-T amplitude equalizer section is used to compensate the combined amplitude distortions of the filter and the four delay equalizing sections. The filter and its equalizers are connected in tandem as shown in Fig. 10.

The typical measured characteristics of the equalized filter are shown in Fig. 11. The delay distortion is  $\pm 0.25$  nanoseconds over the 62-78 MHz frequency range. The peak-to-peak loss distortion is 0.05 dB over the same frequency range. The return loss (compared with 75 ohms) is greater than 30 dB over the 62-78 MHz frequency range.

### 3.2 Mechanical Design

The 745A filter is housed in a sectionalized, cast-aluminum base with a folded sheet aluminum cover (see Fig. 12). Each inductor is enclosed in an individual compartment in the channel to minimize the magnetic coupling between inductors. The closely controlled mechanical tolerances achieved with this type of construction help to keep constant any parasitics caused by proximity. Printed wiring paths are used where possible to eliminate varying lead lengths of hand wiring.

### 3.3 Summary

A filter with extremely good performance has been developed by using a digital computer program to get the best electrical design and

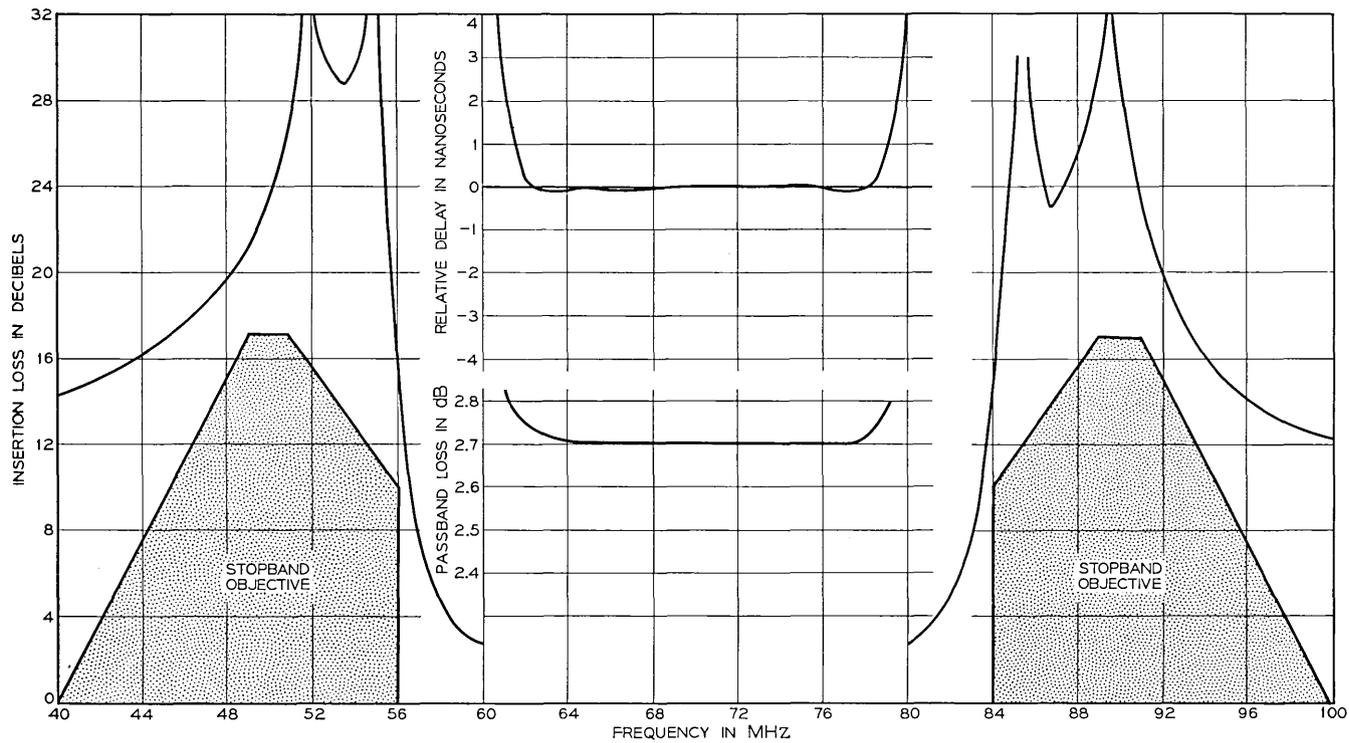


Fig. 11 — Characteristics of the equalized 745A filter.

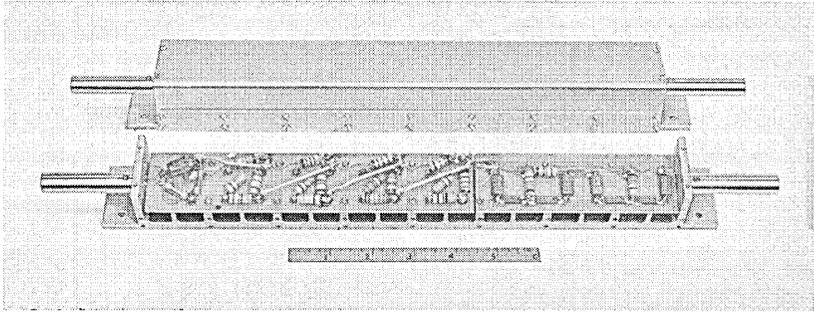


Fig. 12 — The 745A filter.

parasitic-corrected element values, and by using a mechanical configuration with closely controlled tolerances to keep constant parasitics caused by proximity.

The 745A filter represents a great improvement in performance over previous IF filters. For example, the manufacturing limit on the delay distortion is reduced to one third that of the TD-2 intermediate frequency filter (574A), the limit on loss distortion has been cut in half, and the minimum return loss has been increased by three decibels.

#### IV. BASIC IF DELAY AND AMPLITUDE EQUALIZERS

##### 4.1 *Function*

Most circuits in the TD-3 repeater contain active devices which are relatively broadband and therefore introduce little transmission distortion. The IF filters used in the TD-3 system are self equalized. Therefore almost all of the delay and amplitude distortion is introduced by the No. 1418 microwave channel separation networks and No. 1322 channel bandpass filters.

There are 24 No. 1418 microwave channel separation networks and 24 No. 1322 channel bandpass filters which operate in the 24 microwave channels. Many of these networks and filters were measured to determine similarities in passband characteristics about their respective center frequencies.

These passband characteristics are in two groups, those where the beat oscillator is above the signal frequency and those where the beat oscillator is below the signal frequency. Consequently, only two basic IF equalizers are needed. The 794A equalizer is used with the first

group of channels, and the 793A equalizer is used with the second. The delay and amplitude distortions of the first group is the mirror image (about 70 MHz) of the second group.

#### 4.2 Requirements

The average characteristics of two No. 1322 and two No. 1418 networks in tandem was determined from the measurements. To reduce the influence of measurement inaccuracies, the average delay and amplitude characteristics were approximated by a sixth order polynomial (using the least squares method). These smooth delay and amplitude characteristics are shown in Figs. 13 and 14.

TD-3 was designed for 1200 message circuits, which requires the transmission characteristic of the basic equalizers to be carefully controlled over a 12 MHz band. However, in order to provide some transmission performance margin for possible future, but presently unknown, system loads, the transmission characteristic is controlled over a 16 MHz band. Therefore the basic equalizers were designed to meet stringent inband tolerances between 62 and 78 MHz.

The objective of the basic equalizers is to equalize the one-hop

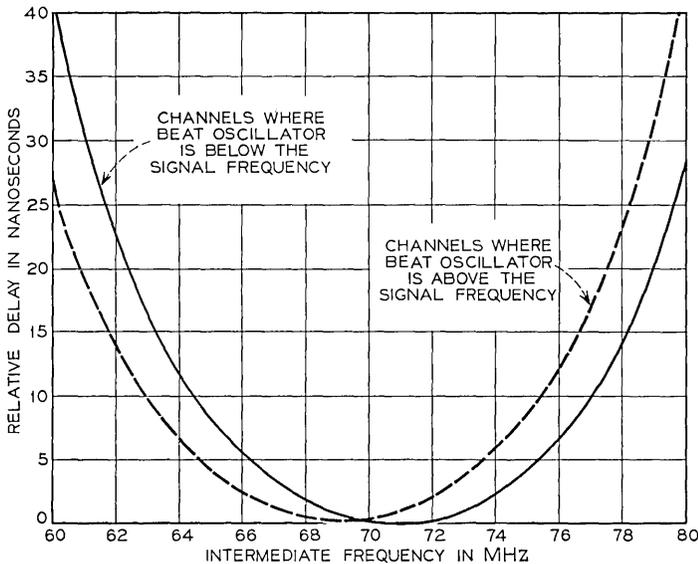


Fig. 13 — Average one hop delay characteristic.

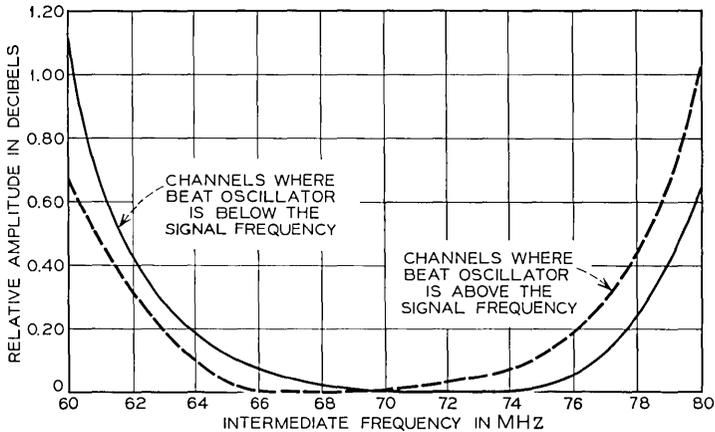


Fig. 14 — Average one hop amplitude characteristic.

characteristics, shown in Figs. 13 and 14, to within the following tolerances:

Frequency (MHz)	Delay		Insertion loss (dB)		Minimum return loss (dB)
	Slope	Ripple	Slope	Parabolic	
62-78	—	$\pm 0.2$ ns	—	$\pm 0.05$	30
60-80	$\leq 0.1$ ns/MHz	$\pm 2.0$ ns	0.20 max.	$\pm 0.50$	25

The input-output impedance level of the basic equalizers is 75 ohm-unbalanced. The constant resistance, unbalanced bridged-T type of delay and insertion loss sections is used to achieve a high return loss.

#### 4.2.1 Theoretical Design

Since the design procedure for both of the basic equalizers is the same, the design of only one equalizer, the 794A, is presented, along with the results of both equalizers.

The requirements on the basic equalizers are very stringent, and a great deal of work was needed to achieve a balance between performance and cost. Hand calculations indicated that the required delay match could be achieved using six 360 degree all-pass delay sections. In order to optimize these results, the  $b$  (stiffness parameter) and  $f_c$  (resonant frequency parameter) for each delay section were used as input to a computer program. This program matched the sixth order polynomial requirements with equalizers of six, seven, and eight delay sections. A final design using seven delay sections

was chosen as a compromise between the number of delay sections and practical considerations such as element values and the values of  $b$ . Experience shows values of  $b$  between 2.5 and 10.0 to be the most desirable to build in the IF range.

The computer delay match over the 62-78 and 60-80 MHz bands was  $\pm 0.02$  and  $\pm 0.17$  ns, respectively.

Since there is a finite  $Q$  associated with the inductors used in the delay sections, the delay sections are lossy. The total delay of the delay equalizer is concave down, and it is a characteristic of this type of delay equalizer to have a similar loss shape. Therefore, the loss of the delay sections tend to self-equalize the loss shape of the microwave networks.

There is also some delay shaping associated with amplitude equalizers. One amplitude section meets the design objectives for this equalizer; however, the single section is complex and would be difficult to tune in the IF range. Therefore, two simpler sections in tandem were designed which are simpler to tune and can be constructed in the standard mechanical assembly. The computed amplitude match with the two sections was  $\pm 0.03$  and  $\pm 0.08$  dB over the 62-78 and 60-80 MHz ranges, respectively. Figure 15 shows the complete schematic for the 794A equalizer. The schematic for the 793A equalizer is the same except that it has one additional delay section.

4.2.2 Laboratory Measured Performance

The measured performance of the first laboratory models of 793A and 794A equalizers is:

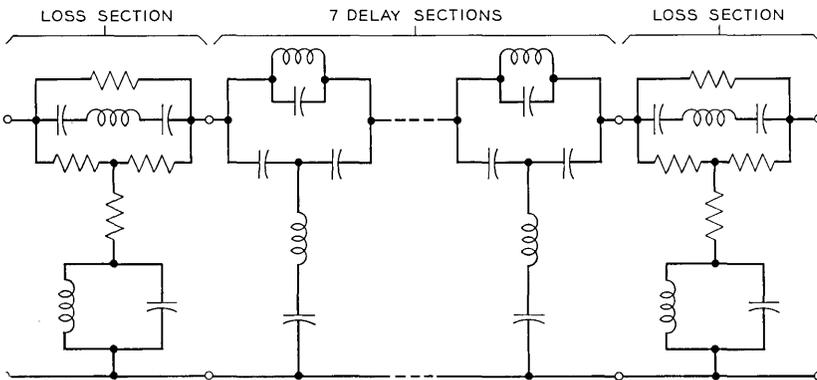


Fig. 15 — Schematic for the TD-3 basic delay and amplitude equalizer, 794A.

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Insertion match	Return
62-78	$\pm 0.15$	$\pm 0.04$	35 min
60-80	$\pm 0.50$	$\pm 0.20$	33 min

The insertion loss at 70 MHz is 5.3 dB, and the absolute delay at 70 MHz is 180 ns.

Additional models of both equalizers were constructed, using capacitors and resistors that were within  $\pm 1$  percent of those used in the first laboratory models. Using the  $\pm 1$  percent variations in element values, which is the closest tolerance available, the same degree of delay match and high return loss was not achieved.

#### 4.2.3 *Manufacturing Procedure and Requirements*

The manufacturing requirements for the basic equalizers are:

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Insertion match	Return
62-78	$\pm 0.3$	$\pm 0.05$	28 min
60-80	$\pm 2.0$	$\pm 0.50$	25 min

The delay and amplitude tolerances of  $\pm 0.3$  ns and  $\pm 0.05$  dB is the best accuracy that can be achieved on the equipment available for testing the basic equalizers. In order to obtain this accuracy, the equalizers are measured by comparison; that is, an unknown equalizer is measured against a standard 793A and 794A equalizer. This adjusting and testing is done on a test set which visually displays the input and output return loss of the equalizer being tested as well as the delay and insertion loss differences between the equalizer being tested and standard equalizer.

Each delay and loss section is tuned individually before it is assembled into a complete equalizer. Once the equalizer is completely assembled, it is returned by the comparison method.

#### 4.2.4 *Mechanical Design*

The basic equalizers are stacked in assembly. Each section, delay and loss, is assembled on a printed wiring card and inserted in cans mounted on a frame. The cards are then wired together and each can is sealed with a separate cover, as the bottom of Fig. 16 shows. After the equalizer has been tested, the completed assembly is enclosed by

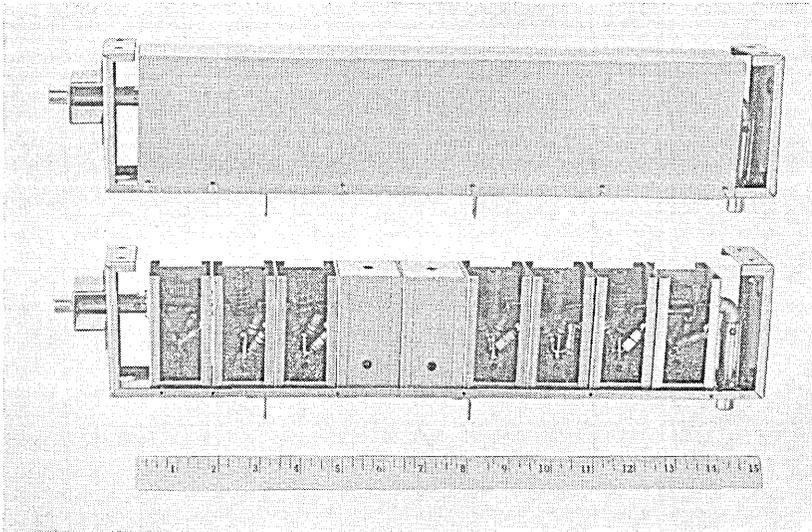


Fig. 16 — 794A equalizer.

a single cover, as shown at the top of Fig. 16, to increase mechanical strength.

Each section is in a separate can to reduce the interaction between sections; Printed wiring boards reduce the differences that could occur with hand wiring. The two inductors in the 360-degree all-pass bridged-T delay sections are mounted perpendicular to each other to reduce the magnetic coupling between them, and to enable the inductors to be adjusted with the covers over the individual cans. The bridging capacitors are hand wired between the inductors; the distance between inductors is about the same as the bridging capacitor length, so there is little room for extra lead. Figure 17 shows the printed wiring card and its assembly. This assembly has the good mechanical symmetry needed for a high return loss.

## V. "MOP-UP" EQUALIZATION

### 5.1 *Function*

Since each basic equalizer is used with different sets of RF networks, it was not possible with only one fixed IF equalizer in each repeater to achieve the nearly perfect equalization required by the TD-3 radio system. Additional "mop-up" equalization is required on

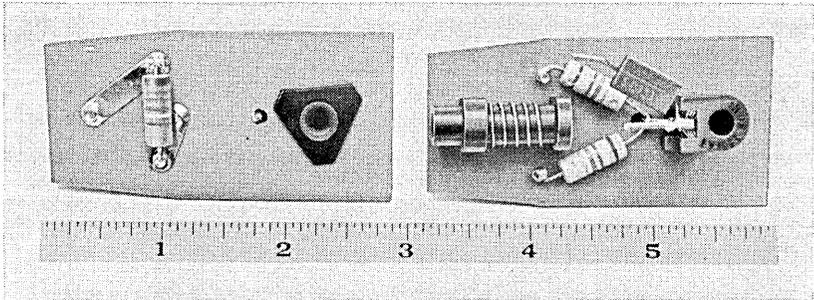


Fig. 17 — Assembly of printed wiring cards.

each hop. Delay slope is the most prevalent and potentially serious component of the residual characteristic of each hop.

### 5.2 Fixed Delay Equalizers

#### 5.2.1 Requirements

The amount of delay slope required to mop-up a radio channel is subject to change. Thus, slopers must be added, removed or replaced from time to time. In order that the absolute delay at 70 MHz can be retained with minimum effort, it is required that all mop-up equalizers have the same absolute delay at 70 MHz.

Based on the TD-3 field trial experience and the per-repeater linear envelope delay distortion requirements, it was concluded that four codes of slope equalizers were needed.

Nominal slopes of +0.25, -0.25, +0.50 and -0.50 ns per MHz are required, along with an equalizer which has the same absolute delay at 70 MHz as the slopers, but which is nominally without envelope delay distortion (zero delay slope) and amplitude distortion.

Requirements for these equalizers are:

Frequency (MHz)	Delay match (ns)	Loss (dB)	
		Amplitude	Return
62-78	$\pm .3$	$\pm 0.02$	33.0 min
60-80	$\pm .5$	$\pm 0.05$	30.0 min
70		1.0 max	

In addition, the absolute delay at 70 MHz for all five equalizers is within  $\pm 1.0$  ns.

### 5.2.2 Design and Performance

The electrical design and adjustment procedures for all five mop-up equalizers are the same as those for the IF basic equalizers.

The codes and manufacturing requirements for each equalizer are:

Code	Delay slope (ns per MHz)	Insertion loss, 70 MHz (dB)
918A	- .50	0.35 ± 0.1
918B	- .25	0.35 ± 0.1
919A	+ .25	0.60 ± 0.1
919B	0.00	0.50 ± 0.1
920A	+ .50	0.80 ± 0.1

Additional manufacturing requirements common to all five equalizers are:

Frequency (MHz)	Delay match (ns)	Amplitude distortion (dB)	Return loss (dB min)
62-78	± .3	± .03	33
60-80	± .5	± .05	30

Absolute delay at 70 MHz =  $23.4 \pm 1.0$  ns.

## VI. 747A LOW PASS FILTER

### 6.1 Function

The IF preamplifier used in the TD-3 repeater generates harmonics of the 60-80 MHz fundamental signal. This amplifier is followed by an IF main amplifier. When the IF basic equalizer is placed between these units, a substantial delay ripple occurs. This ripple occurs because the absolute delay of the basic IF equalizers is considerably greater at the fundamental frequencies than at the harmonics. Thus, the fundamental frequency is substantially delayed in relation to its harmonics.

Because of small nonlinearities in the main amplifier, the delayed fundamental and its harmonics produce modulation products at the fundamental frequency which have a different phase from the incident fundamental frequency. This causes the substantial delay distortion. When a low pass filter, which attenuates the second and third harmonics of the 60-80 MHz range, is inserted before the repeater equalizer, the ripple is greatly reduced. The low pass filter prevents the harmonics from reaching the IF main amplifier.

6.2 *Electrical Design*

6.2.1 *Requirements*

It is not practical to incorporate the low pass filter as part of the basic equalizer circuit. Further, separation of the low pass filter from the equalizer is consistent with the original philosophy of keeping the basic repeater equalization separate from other IF networks.<sup>3</sup>

For reasons described in Section IV, the equalized passband of the low pass filter is 62-78 MHz. However, the tightest inband requirements are placed on the 64-76 MHz band. The requirements for the low pass filter are:

Frequency (MHz)	EDD (ns)	Loss (dB)	
		Insertion	Return
64-76	±0.05	±0.01	30.0 min
60-80	±0.10	±0.05	27.0 min
70		1.0 max	
120-240		21.0 min	

The input and output impedances are 75 ohm, unbalanced.

6.2.2 *Theoretical Design*

The 747A is a fifth order Cauer elliptic low pass filter which is designed with an equal ripple passband and equal minima stopband. The computed passband loss distortion of this filter with a finite Q of 150 is 0.1 dB. The minimum return loss in the passband is 34 dB. The delay equalization is achieved with two 360-degree all-pass bridged-T sections, and the loss equalization with one bridged-T amplitude section. Figure 18 is a schematic diagram of the 747A filter.

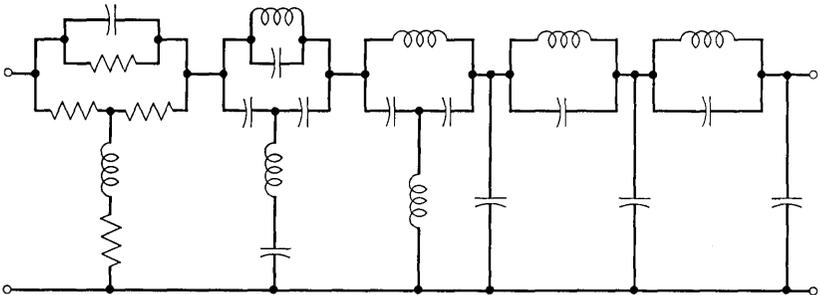


Fig. 18 — Schematic for the 747A filter.

6.2.3 *Measured Performance*

The measured performance and manufacturing requirements for the 747A filter are:

Frequency (MHz)	EDD (ns)	Loss (dB)	
		Insertion	Return
64-76	$\pm 0.30$	$\pm 0.02$	30.0 min
60-80	$\pm 0.30$	$\pm 0.05$	27.0 min
70		0.60 max	
120-240		21.0 min	

Figure 19 shows its typical characteristics.

6.2.4 *Mechanical Design*

The 747A filter is assembled on a printed wiring board which is mounted on an eight-section cast aluminum base. The inductors are mounted on the component side of the printed wiring board and shielded from each other by walls in the base. The capacitors and resistors are mounted on the conductor side of the board. Figure 20 shows the assembly.

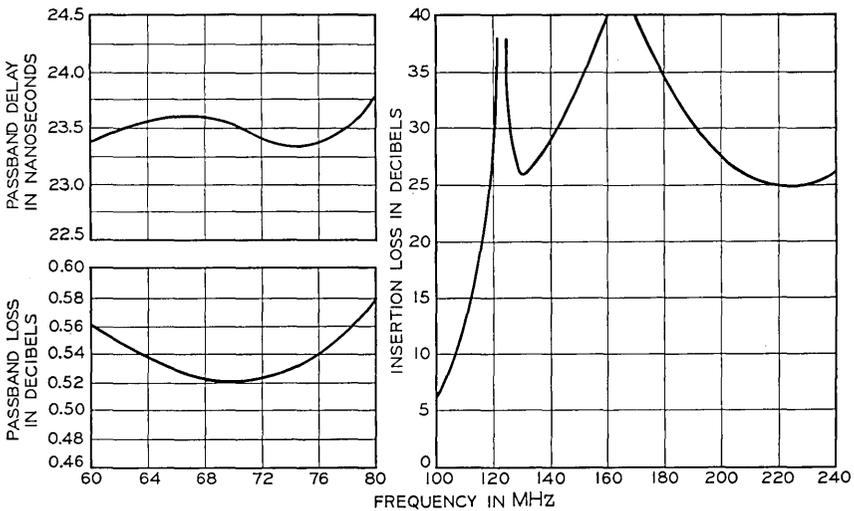


Fig. 19 — Characteristics of the 747A filter.

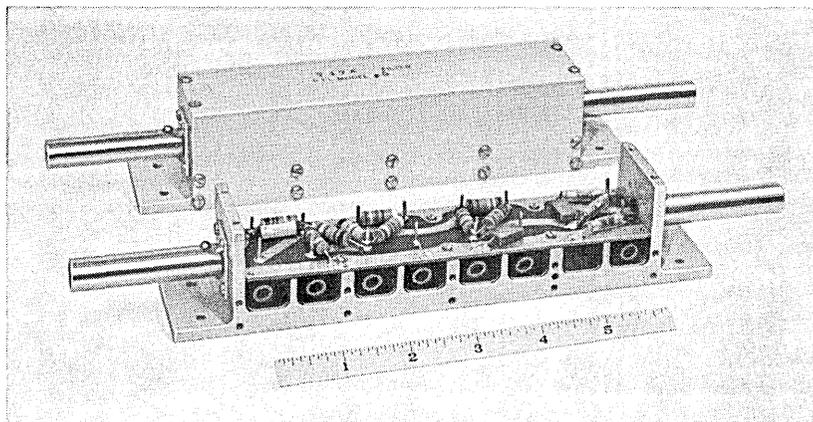


Fig. 20 — 747A filter.

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## 3A FM Terminal Transmitter and Receiver

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*FM terminals form an important subsystem of long-haul microwave radio systems as the link between the baseband signal and the 70 MHz FM signal. Designed primarily for use on the TD-3 system, 3A FM terminals are also compatible with the TD-2 system on which they are finding wide application. Development objectives included the use of solid-state circuitry throughout, emphasis on reliability, and performance consistent with TD-3 objectives with up to 16 terminal pairs in tandem in 4000 miles.*

*The FM transmitter uses two voltage controlled oscillators at 186 and 256 MHz which are frequency modulated by the baseband signal and heterodyned down to the 70 MHz IF. Transmitter frequency stability is ensured by controlling the oscillator's environmental temperature and stringent pre-aging of critical circuit elements. The FM receiver uses a balanced, parallel resonant type discriminator preceded by two limiters which ensure good AM suppression and wide dynamic range. The gain of a terminal pair is 16 dB with 124 ohm balanced baseband input and output impedances.*

### I. INTRODUCTION

As indicated in a companion paper, FM terminals perform the initial and final modulation steps in the TD-3 microwave radio system.<sup>1</sup> The FM transmitter converts the baseband signal to a frequency modulated signal centered at 70 MHz; the FM receiver performs the reverse function of recovering the baseband modulation from the FM signal. Although primarily intended for use in the TD-3 system, the 3A FM terminal equipment\* also is used for improved performance in the TD-2 system.<sup>2</sup>

Design emphasis was placed on: (i) reliability and minimum maintenance through the use of solid-state circuitry throughout, (ii) im-

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\* Manufactured by the Western Electric Company for Bell System use only.

proved performance consistent with the more stringent TD-3 objectives, and (iii) reduced cost and size.

The types of baseband signals to be transmitted over the terminals are discussed by S. D. Hathaway and others.<sup>1</sup> In general terms they consist of 1200 message circuits comprising two multiplex master-groups or a single NTSC\* color television signal. FM terminals are required at each end of a radio route and at intermediate points where the baseband signal or some portion of it must be added or dropped. The 3A design is based on a maximum of 16 terminal pairs in tandem in 4000 miles.

## II. DESIGN OBJECTIVES

With appropriate assumptions about the law of impairment addition, the system allocation to terminals can be converted to the performance objectives for a single terminal pair given in Table I.

These objectives apply with the appropriate pre-emphasis shapes which are chosen to optimize over-all system performance. Since terminal noise and distortion are not controlling, preemphasis shapes were not available as a design parameter for optimizing terminal performance. The pre-emphasis shapes chosen (see Fig. 1), were based on an extension of the shapes that were found to be optimum for the TD-2 system.

The message and television signals carried by TD-3 have an upper frequency limit of approximately 6 MHz. However, to minimize the influence of terminals on high-end frequency response and to allow for possible future applications, the upper frequency limit was set at 10 MHz. The lower frequency limit is set by the 60 Hz component of a television signal. To minimize phase distortion of the low frequency components of the TV signal, it is necessary to keep all low frequency cutoffs well below 60 Hz. This is satisfactorily controlled by maintaining the transmission response essentially flat down to 6 Hz.

Cross modulation objectives can be stated in terms of nonlinearity or harmonic performance.<sup>3</sup> These characterizations, however, are less appropriate than a dBnC0 objective with simulated message loading since situations can arise where optimum noise loading and optimum linearity are not coincident. Nevertheless, since good linearity is in a broad sense a necessary condition for satisfactory performance with

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\* Based on standards of the National Television System Committee.

TABLE I—DESIGN OBJECTIVES FOR ONE FM TERMINAL PAIR

Baseband transmission Bandwidth ( $\pm 0.1$ dB point) Gain stability	6 Hz to 10 MHz $\pm 0.25$ dB per six-month maintenance interval
Total noise First order nonlinearity ( $\pm 4$ MHz) Second order nonlinearity ( $\pm 4$ MHz)	21 dBrc0 <1.3% <2.0%
Differential gain  Differential phase Television weighted high frequency signal-to-noise ratio	Satisfactory differential gain is ensured by the telephone cross-modula- tion objectives $\pm 0.3^\circ$ >73 dB
Center frequency of FM transmitter Peak frequency deviation Deviation sense	70 MHz $\pm 100$ kHz 4 MHz Positive going signal on the tip of the input plug produces a decrease in transmitter output fre- quency
Transmitter longitudinal suppression	>53 dB up to 1 MHz
Change in receiver demodulation sensitivity for a 10 dB reduction in IF input Microphonics	<0.25 dB  Negligible
Change in transmitter carrier fre- quency between zero and full modulation Operating temperature range	<20 kHz  0 to 50°C

a live message load, it is useful to have a linearity objective for design purposes.

The requirement relating to carrier frequency stability as a function of modulation level is associated with the desire to use the carrier null or Crosby technique to set transmitter deviation accurately.<sup>4</sup> Since the detection of a carrier null in the presence of adjacent sidebands requires a narrow band receiver, it is essential that the application or removal of modulation does not shift the carrier outside the receiver passband.

### III. FM TRANSMITTER

The FM transmitter provides a +10 dBm output signal centered at 70 MHz with frequency modulation linearly related to the baseband

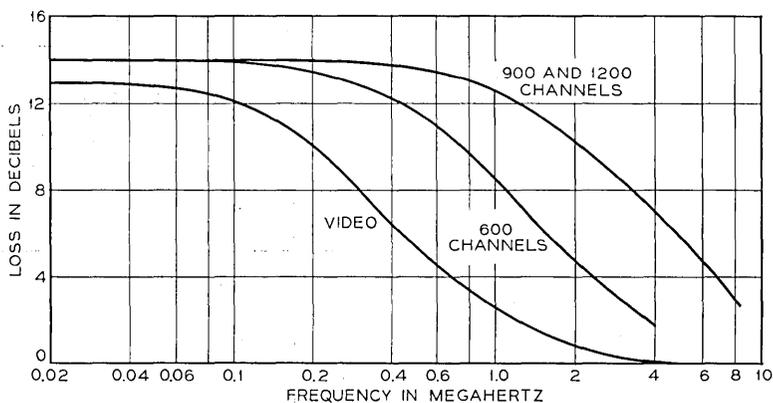


Fig. 1 — Pre-emphasis characteristics.

(input) signal. A  $-12$  dBV input signal produces 8 MHz peak-to-peak deviation.

The development of the transmitter involved the examination of various alternative solid-state modulators. These fall into two broad categories: (i) direct modulators which operate and are deviated at the 70 MHz intermediate frequency; (ii) heterodyne type modulators in which the outputs of two higher frequency voltage-controlled oscillators are combined in a mixer to generate a difference frequency of 70 MHz. For reasons related to the status of device development, reproducibility, and high frequency baseband response, the heterodyne approach was used in the 3A development.

### 3.1 General Description

The broad features of the FM terminal transmitter are shown in the block diagram of Fig. 2.

The modulating signal is applied to the 124 ohm balanced input of the baseband amplifier. Two outputs from this amplifier, in anti-phase, drive two voltage controlled oscillators with frequencies centered at 186.67 and 256.67 MHz. The deviated outputs from these oscillators are applied through buffer amplifiers to a mixer circuit where the 70 MHz difference frequency is generated. A low pass filter in the mixer output rejects unwanted products and passes the desired intermediate frequency to the succeeding filter-equalizer unit. The filter portion of this unit provides additional suppression to out-of-band products generated by the mixer. The equalizer provides amplitude equalization for the low pass filter as well as delay equali-

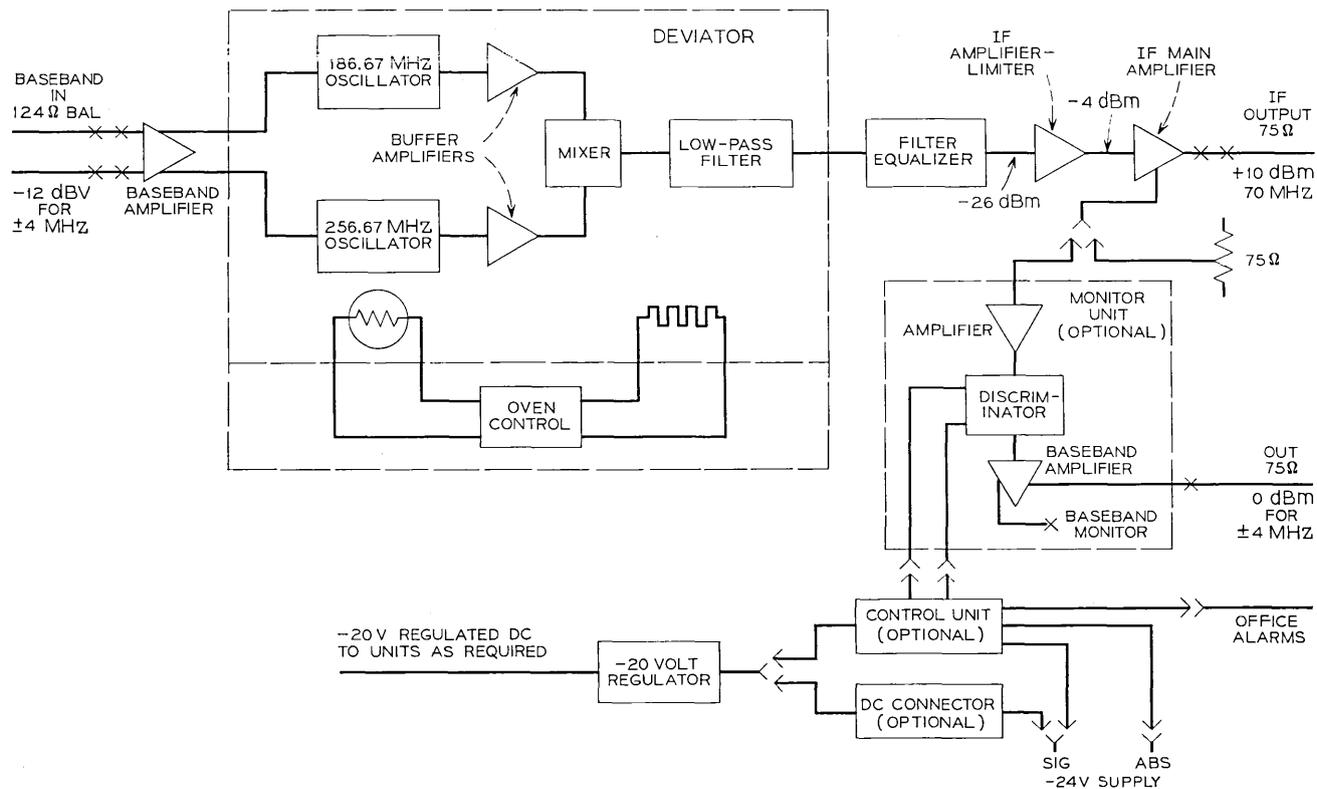


Fig. 2 — Block diagram of FM transmitter.

zation for the complete transmitter. The signal then passes to an amplifier-limiter and finally to an output IF amplifier which delivers +10 dBm.

Part of the IF output from the main amplifier can be connected to a monitor circuit which is essentially an FM receiver without limiting. The discriminator output from this circuit is used to provide alarm indications when the transmitter IF output power or average frequency departs by a prescribed amount from nominal. The alarm outputs are not provided directly from the monitor but through alarm relays in the control unit. Both the monitor unit and control unit are optional and are generally not used when the terminal is included within an automatic terminal protection switching system.

The frequency of the transmitter is maintained within 100 kHz of nominal by closely regulating the temperature of the entire deviator assembly. This is accompanied by special pre-aging of critical components within the deviator to ensure that the stability objectives are met for extended periods. The use of this form of frequency stabilization as opposed to automatic frequency control has the merit of placing no limitation on the low-end frequency response of the transmitter.

Figure 3 is a photograph of the FM transmitter.

### 3.2 Baseband Amplifier

The baseband amplifier provides a nominal 19.5 dB of voltage gain flat to  $\pm 0.05$  dB in the frequency band 6 Hz to 10 MHz between the balanced 124 ohm input and the balanced varactor diode loads presented by the deviator. It must also suppress low frequency common mode tones by more than 53 dB.

Figure 4 is a simplified schematic diagram of the amplifier. The amplifier is balanced in all stages and operates in push-pull class A manner with identical stages for "tip" and "ring" sides of the signal.

The input stages  $Q_1$  and  $Q_{101}$  provide no signal gain, but together with transistor  $Q_7$  provide more than 65 dB of voltage loss to common mode tones applied to the input. In addition, transistor  $Q_7$  provides a canceling voltage for low frequency power supply noise suppression. This is accomplished by applying the power supply noise to the emitter of  $Q_7$  through resistor  $R_{16}$  which is adjusted to provide a noise voltage at the bases of  $Q_2$  and  $Q_{102}$  equal to the power supply noise present at their emitters. In this manner the noise volt-

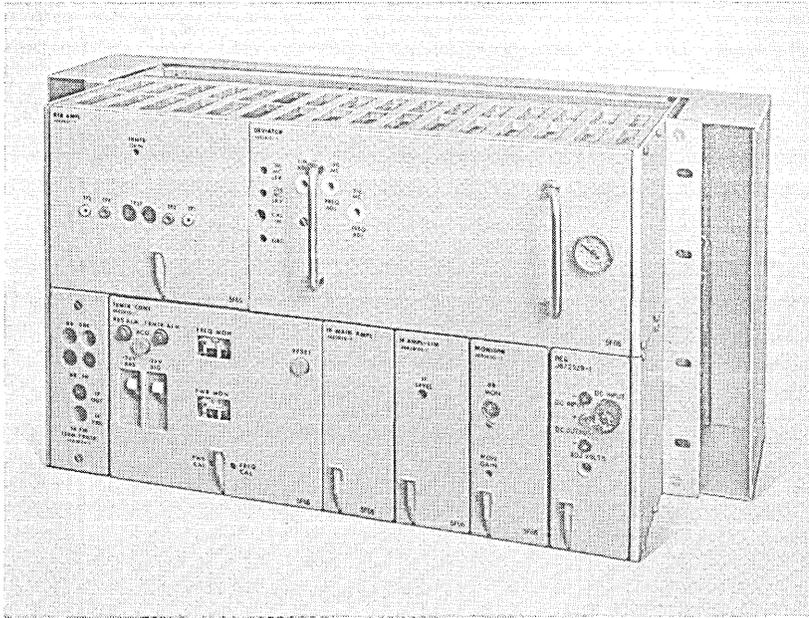


Fig. 3—3A FM transmitter.

age at either side of the amplifier output is reduced by more than 20 dB.

The amplifier gain is provided by two 2-stage series input, shunt output balanced feedback amplifiers. Bias stability is obtained by dc feedback. The high input impedance of the series input feedback circuit allows frequency response down to 0.6 Hz with coupling capacitors of reasonable size (500  $\mu$ f) between stages and without the use of low frequency loop gain compensation.

Gain adjustment is accomplished in the first balanced feedback amplifier by a single potentiometer common to the "tip" and "ring" sides. The gain potentiometer is connected to equal dc voltage points in the two amplifiers which eliminates bothersome transients during gain adjustment.

The voltage gain of the first balanced feedback amplifier is adjustable from 4 to 12 dB and the gain of the second amplifier is fixed at 11 dB. As shown in Fig. 5, the output amplifier has 38° of phase margin at 50 MHz with no phase crossover below 500 MHz. The use of ferrite beads in place of resistors in the phase shaping network  $C_5$  and  $CM_3$  provides smooth loop transmission where the inductance

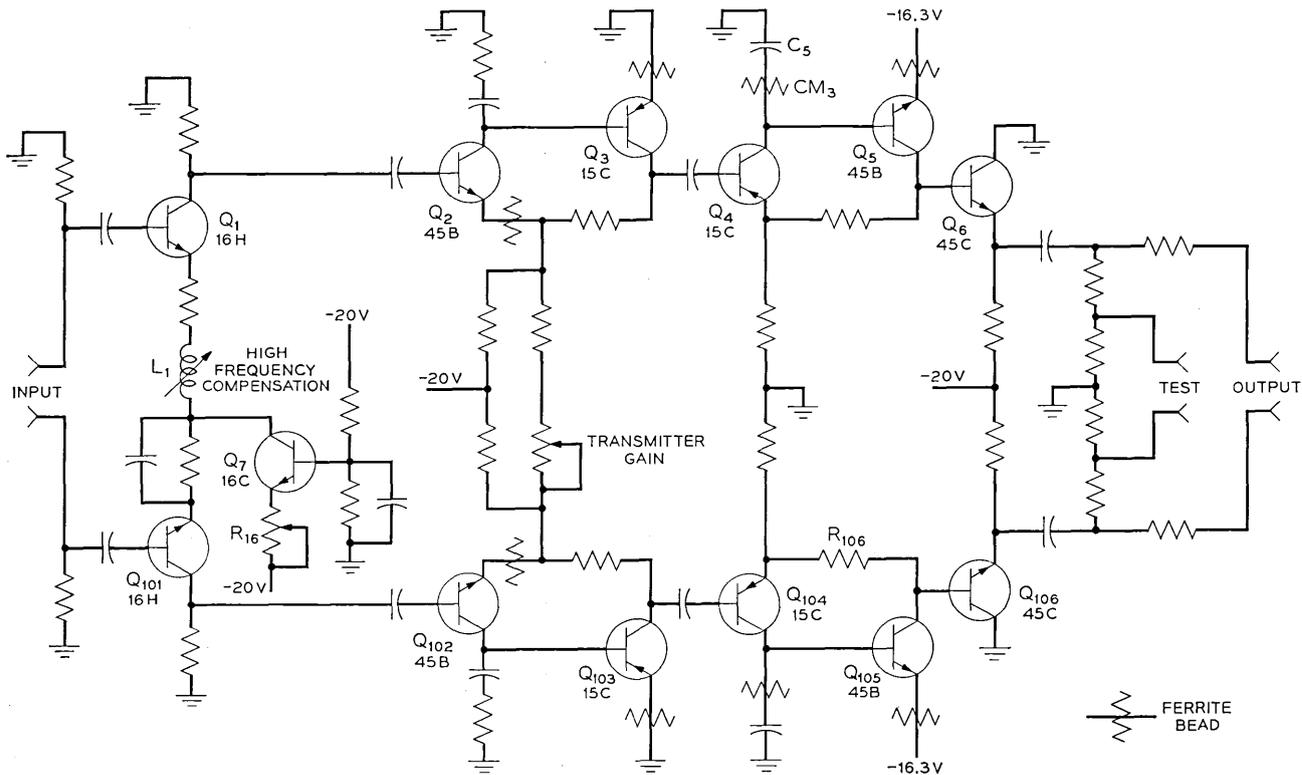


Fig. 4—Simplified schematic diagram of transmitter baseband amplifier.

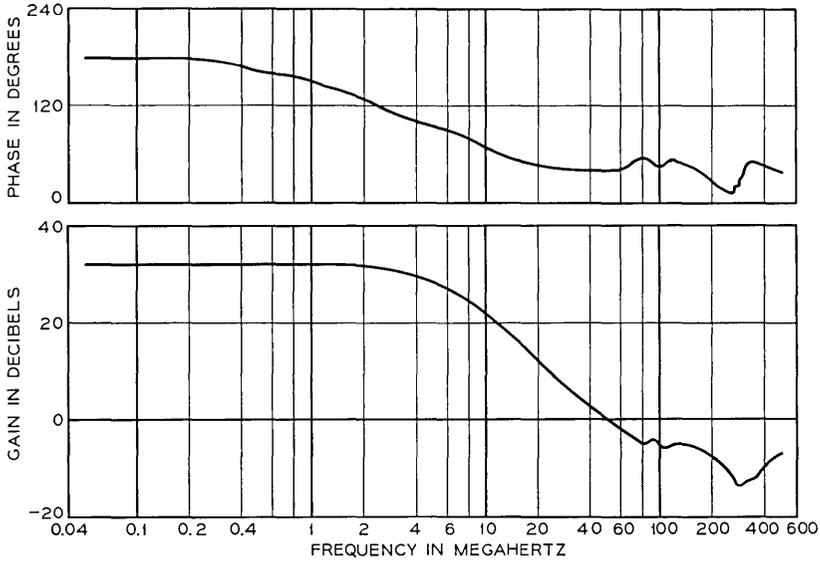


Fig. 5 — Output amplifier open-loop gain and phase.

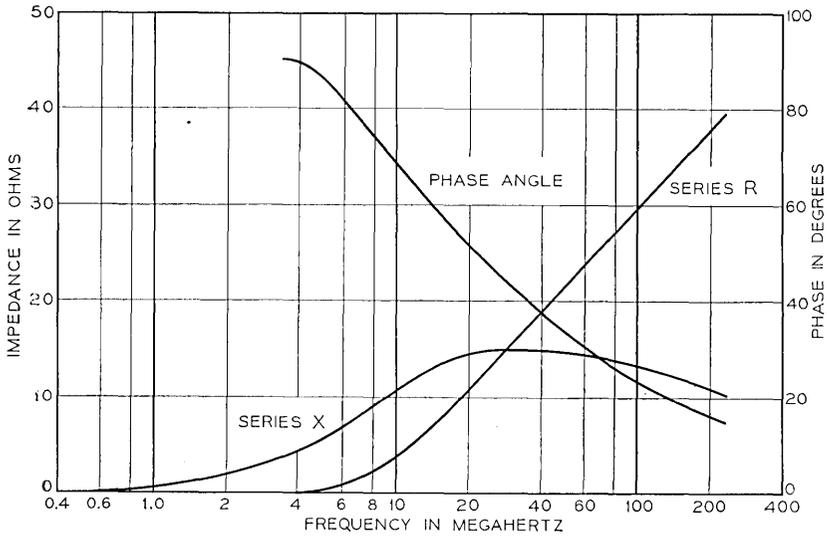


Fig. 6 — Ferrite bead impedance characteristics.

of an equivalent resistor would have produced a resonant ripple at 50 MHz.

Figure 6 shows the resistance and inductive reactance of these beads versus frequency. The phase angle rapidly decreases from about  $90^\circ$  at 5 MHz to  $34^\circ$  at 50 MHz. The increase in resistance with frequency and the decreasing phase angle give these devices ideal characteristics for reducing loop gain at high frequencies with minimum phase shifts. These beads are also used for emitter feedback in the second stage of the feedback amplifiers.

Emitter followers are used as output stages to isolate the feedback loop from the load impedance. The emitter followers work into the high impedance of varactor diodes through an equalizing network for flat baseband transmission.

The high frequency response of the over-all amplifier is adjusted by the common feedback inductor  $L_1$  to achieve a response flat to  $\pm 0.05$  dB from 6 Hz to 10 MHz. Figure 7 shows the frequency response of a baseband amplifier. The 3 dB down points are 0.6 Hz and 45 MHz.

The nonlinearity of this amplifier is less than 0.15 percent, which adds virtually no degradation to the terminal.

### 3.3 Deviator

Frequency modulation in the FM transmitter is performed by the deviator. A heterodyne deviator was chosen since it requires higher oscillator frequencies which: (i) simplify the separation of oscillator

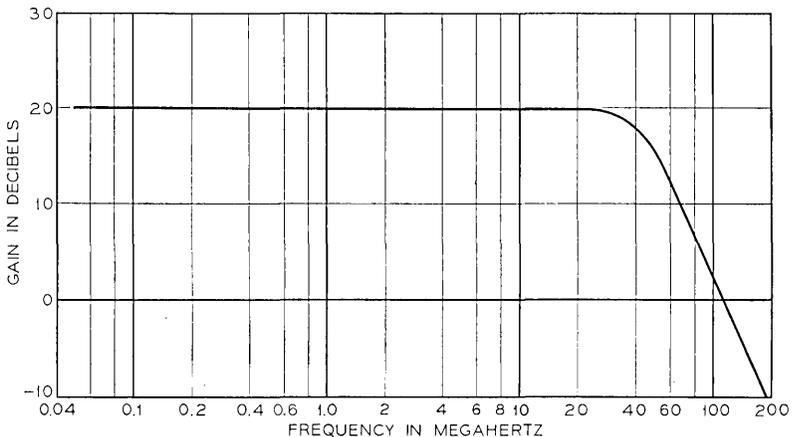


Fig. 7 — Baseband amplifier frequency response.

and baseband signals, (ii) have an inherently higher modulating frequency capability, and (iii) have lower nonlinearity because of their smaller percentage deviation. The principal disadvantage of the heterodyne method is the generation of undesired products in the mixer which in practice necessitates lower deviator output powers.

The choice of oscillator frequencies for the heterodyne deviator should, on the one hand, be as high as possible to minimize percentage deviation and hence improve linearity, and on other hand, should be kept as low as possible to avoid device limitations and reduce the importance of strays. In the 3A deviator, the percentage deviation on each oscillator is halved by deviating both oscillators in antiphase. As discussed in Appendix A, this push-pull type situation leads to the cancellation of first-order nonlinearities so that the first significant nonlinearity is the smaller second-order term.

The results of Appendix A indicate that to keep this second-order term below two per cent, using graded junction varactor diodes as the voltage sensitive tuning elements, the lowest heterodyne oscillator frequency should lie above 150 MHz. An examination of spurious mixer products as a function of oscillator frequencies shows that the nearest satisfactory location above 150 MHz is 186 MHz with the high frequency oscillator 70 MHz above this at 256 MHz. Designating these frequencies as  $V$  and  $C$ , respectively, the lowest order mixer product falling within the 60 to 80 MHz band is the 10V-7C term. This product will be least interfering if it falls exactly at 70 MHz leading to the final choice of rest frequencies at 186.667 MHz and 256.667 MHz. However, in arriving at requirements for the 10V-7C product level, no advantage was taken of its optimum location at 70 MHz on the grounds that it would not be accurately enough controlled in frequency to assure its continued location at the chosen design frequency.

To insure IF frequency stability, temperature control was chosen in preference to automatic frequency control. This was based on economic advantage, the elimination of low frequency response limitations because of an automatic frequency control loop, and the general advantages which accrue from temperature stabilization of the semiconductor circuit environment. However, the lack of an automatic frequency control circuit imposes stringent requirements on oscillator frequency stability.

The voltage-controlled oscillators are Colpitts-type circuits such as in Fig. 8. The varactor diodes which have the voltage capacitance

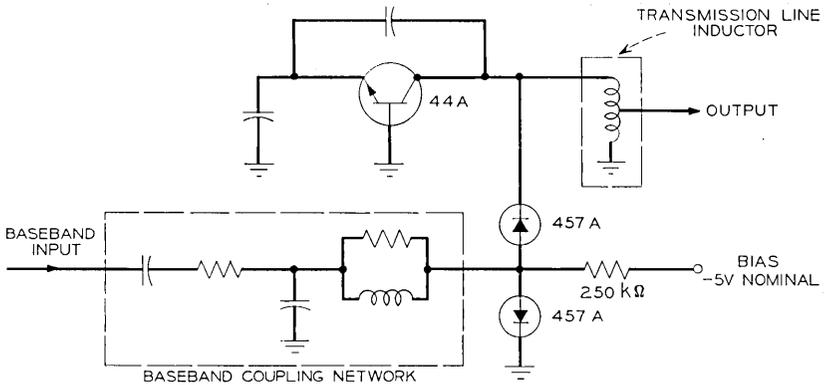


Fig. 8—Simplified voltage controlled oscillator circuit.

law illustrated in Fig. 9 are connected in opposite directions across the oscillator tank circuit with the modulating signal applied at the center point. This reduces modulation of the tank circuit capacitance by the oscillator output signal. An R-L-C coupling network, terminated by the two varactor diodes in parallel, ensures that the modulating voltage applied to the diodes is virtually independent of frequency over the entire baseband range.

The capacitance of the two varactor diodes in series at the nominal -5 volts bias is 8 pF. This value was chosen as a compromise between providing a reasonably high ratio of diode capacitance to fixed capacitance and at the same time ensuring that the tank circuit inductance did not become unduly small in relation to circuit strays.

The tank circuit inductance consists of a 100-ohm short-circuited coaxial transmission line of approximately 27 degrees electrical length.

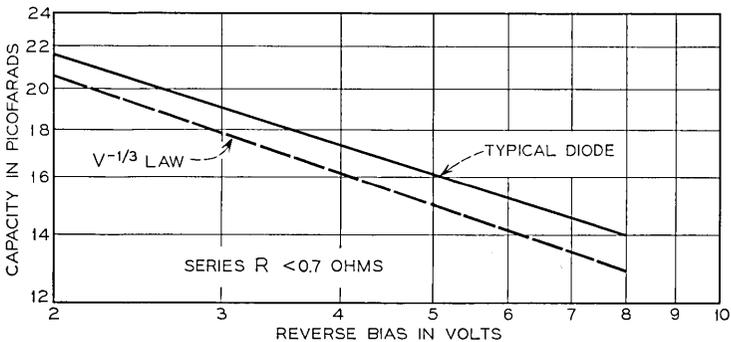


Fig. 9—Typical capacitance characteristics of 457A varactor diode.

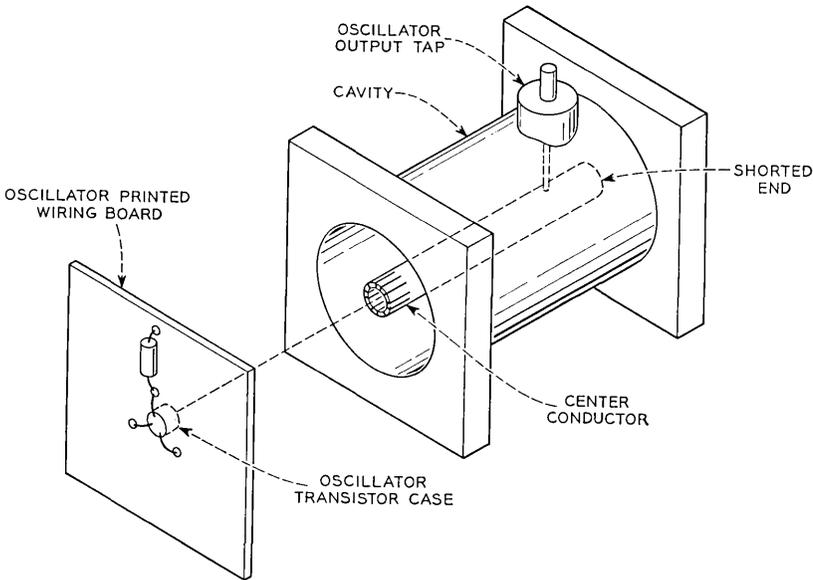


Fig. 10 — Transmission line inductor assembly.

As illustrated in Fig. 10, a chuck on the open end of the transmission line center conductor accepts the transistor case which is connected internally to the collector. This gives an intimate and well-controlled connection between the collector and tank circuit inductance.

The transmission line inductor is very stable, highly reproducible through control of the mechanical dimensions, and quite insensitive to mechanical shock. The center conductor also provides a good heat sink for thermal stabilization of the oscillator transistor. Output power is derived at a convenient 50-ohm impedance level by tapping the center conductor near the short-circuited end of the transmission line. To eliminate hysteresis effects and frequency discontinuities during warmup, the same material is used on inner and outer conductors. One disadvantage of the transmission line inductor is that its first-order equivalent circuit is an inductor in parallel with a small capacitor which increases the total effective stray capacitance.

The approximate effective equivalent circuit of the 186 MHz oscillator tank circuit is shown in Fig. 11. The ratio of stray to total tank circuit capacitance is 0.5 and, using the results of the Appendix, the computed second-order deviator nonlinearity is 1.8 percent for  $\pm 4$  MHz deviation. This is in good agreement with typical measured performance.

The 256 MHz oscillator is similar to its lower frequency counter-

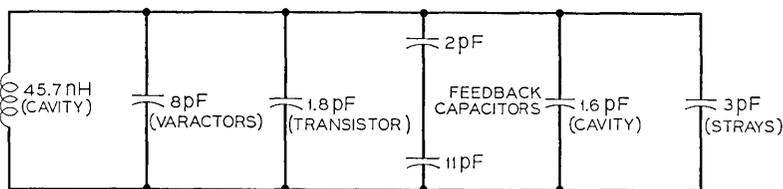


Fig. 11 — Equivalent tank circuit of 186 MHz oscillator.

part except for an additional trimmer capacitor across the tank circuit. Assuming equal deviation sensitivities, one would expect the 256 MHz oscillator to be more linear than the 186 MHz unit as a result of the smaller percentage deviation. Since it is intended that first-order nonlinearities in the two oscillators should cancel, the linearity of the 256 MHz oscillator must be diminished by providing additional “strays” in the form of the trimmer.

This trimmer is used to optimize linearity by adjusting for zero first-order deviator nonlinearity. The need for the trimmer can also be demonstrated from equation 15 of the Appendix. For  $n = \frac{1}{3}$ , this equation requires that the high frequency oscillator stray capacitance must be greater than the stray capacitance of the low frequency oscillator for zero first-order nonlinearity.

To ensure a stable bias voltage for the oscillator varactor diodes, the  $-20$  volt output from the dc regulator is first reduced to  $-15$  volts by a voltage regulating diode circuit. A second voltage regulating circuit using a 446AC diode further reduces this to  $-8.2$  volts which is applied to two adjustable voltage divider networks which supply the bias for the individual oscillators. Since the  $-8$  volt bias supply is common to the two oscillators, frequency changes resulting from residual voltage variations tend to cancel.

Special care has been used in the detailed design of the oscillators to ensure good frequency stability. The principal factors influencing long-term frequency stability are breakdown-voltage aging on the 446AC diode, capacitance aging on the 457A varactor diodes, and leakage current changes on the varactor diodes. Device aging objectives were determined starting from an intermediate frequency stability objective of 100 kHz over a six-month maintenance interval. Of this total, 50 kHz was allocated to semiconductor aging and ambient temperature effects while the remaining 50 kHz was allocated to all other causes of frequency drift. The 50 kHz allowance for semi-

conductor and temperature effects was further divided in the manner shown by Fig. 12 to provide objectives for individual devices.

To achieve these aging objectives, the 446AC diode and 457A varactor diode are subjected to stringent pre-aging as described by S. M. Forst and others.<sup>5</sup>

The major means for controlling frequency is the proportionally controlled oven which maintains the environmental temperature of the oscillator, buffer-amplifier, and mixer assembly within 1° of 50°C. The control characteristic for the oven is shown in Fig. 13. The loss of control at 45°C results from power dissipation in the circuitry within the oven. Since the design operating temperature range for the terminal is 0 to 50°C, the loss of oven control at 45°C results in a degradation in frequency stability over the top 5° of the range. This degradation is small, however, since the individual oscillators have temperature coefficients from 15 to 18 kHz per degree Centigrade with a net temperature coefficient of approximately 3 kHz per degree Centigrade for the intermediate frequency output.

Field measurements indicate that the long-term frequency stability objectives are being met. Figure 14 shows a typical frequency-time record for a seven-month monitoring period.

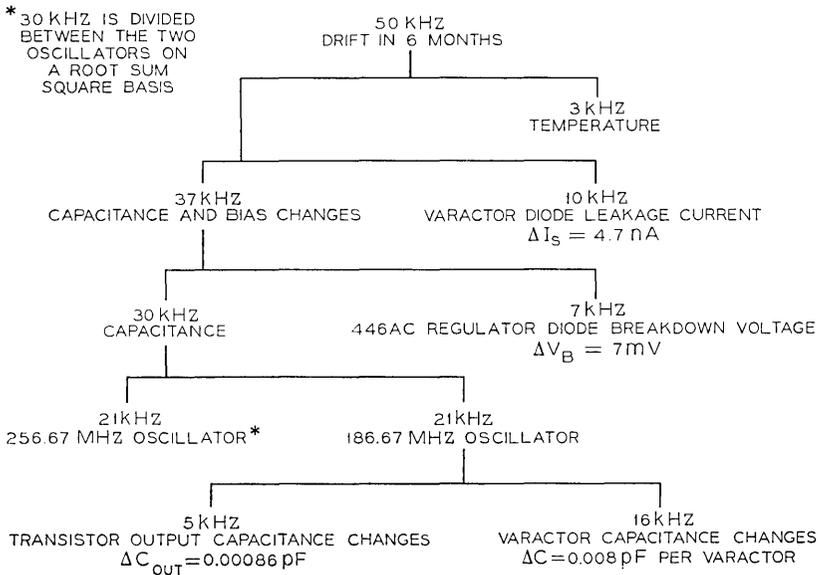


Fig. 12 — Frequency tolerance allocations.

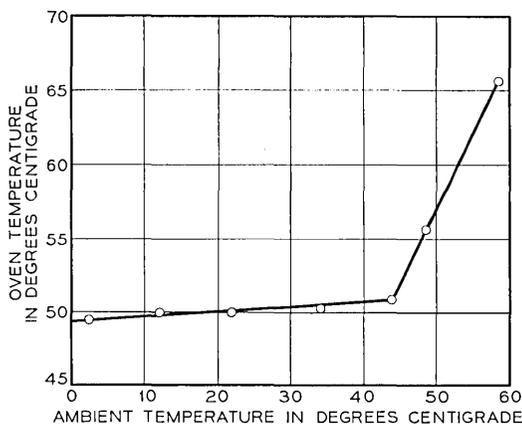


Fig. 13 — Oven temperature control characteristic.

Diode detector circuits are bridged across the oscillator outputs to provide a dc indication of oscillator output level. In the case of the 256 MHz oscillator, provision is also made for applying an external signal in parallel with the oscillator output and its detector circuit. For alignment, this external signal is the output of a crystal-controlled 256.667 MHz oscillator to which the frequency of the internal 256 MHz oscillator can be locked.

The lock-in is indicated by a dip in the dc output of the detector diode. Under lock-in conditions, the 186 MHz oscillator can be adjusted to obtain a precise 70 MHz output from the deviator. Thereafter, with the external reference removed, the internal 256 MHz oscillator can be accurately set by re-establishing a precise 70 MHz intermediate frequency. This procedure is used so that only one reference frequency oscillator need be provided.

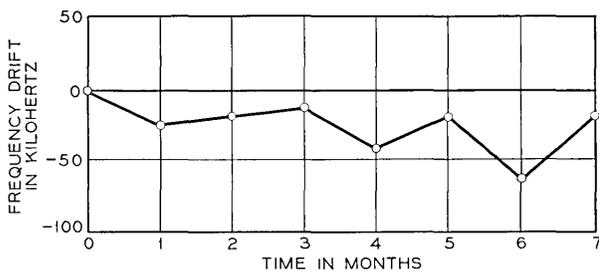


Fig. 14 — Typical frequency stability record for a production transmitter.

The outputs of the deviated oscillators at approximately +2 dBm are connected through single-stage, common-base amplifiers to the transistor mixer circuit as shown in Fig. 15.

The emitter circuit of the mixer transistor is biased to minimize undesired products and provides the desired difference output at -24 dBm. The transistor mixer has the desirable feature that the impedance presented by the output circuit has very little influence on the modulation performance of the emitter circuit. The relatively high 186 and 256 MHz oscillator signals in the mixer output are suppressed by the 697A low pass filter. The loss characteristic of the 697A filter is shown in Fig. 16.

Out-of-band mixer products must be limited in magnitude to minimize adjacent channel interference and in-band interference resulting from modulation sidebands. Undesired in-band products must be at least 96 dB below the carrier. The only significant in-band tone is the high order 10V-7C product and it is the 96 dB objective for this tone that limits the maximum output of the mixer to -24 dBm.

The modulation sensitivity of the deviator ranges from 2.9 to 4.2 MHz per volt, depending upon the particular 457A varactor diodes used. Since all transmitters must have the same sensitivity, this variation is compensated for by providing adequate gain range in the baseband amplifier. The sensitivity is virtually constant over the whole baseband range and in combination with the 3A receiver meets the  $\pm 0.1$  dB baseband response objective.

### 3.4 IF Amplifier-Limiter

Residual amplitude modulation in the output of the deviator is removed by the amplifier-limiter which also provides 22 dB of gain. The design is an adaptation of the amplifier-limiter used in the 100A protection switching system for pilot stabilization.<sup>6</sup> The limiter is very similar to the TD-3 repeater limiter described in Ref. 7 except that the level-to-phase neutralizing network has been eliminated.

The amplifier-limiter provides 20 dB of AM suppression and has

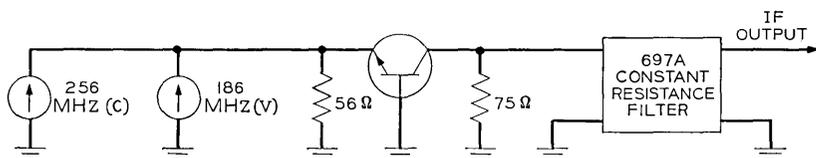


Fig. 15 — Simplified mixer circuit.

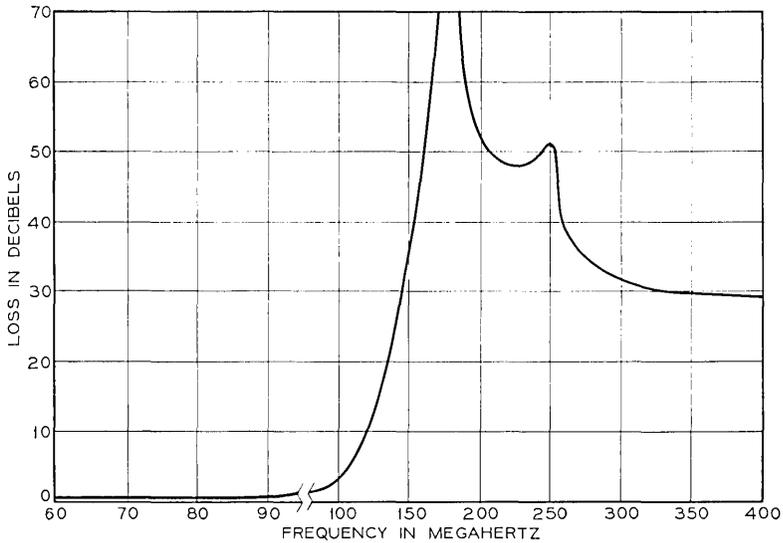


Fig. 16—697A filter loss characteristic.

input and output return losses in excess of 32 dB. The passband frequency response is flat within 0.03 dB from 60 to 80 MHz.

### 3.5 IF Main Amplifier

The IF main amplifier raises the  $-4$  dBm output of the amplifier-limiter to the FM transmitter output level of  $+10$  dBm. The individual stages use the same basic design as the repeater IF amplifier.<sup>7</sup>

A second output, 12.2 dB below the main output, is available for driving the optional monitor circuit. Input and output return losses on the main amplifier exceed 35 dB. The frequency response is flat within 0.03 dB in the 60 to 80 MHz band.

### 3.6 Frequency and Level Monitor

The frequency and level monitoring circuit is essentially an FM receiver without limiting. It provides output indications proportional to the error in average intermediate frequency and the error in IF output power relative to its nominal  $+10$  dBm. The error signals are applied to meter relays in the transmitter control unit and the relay-operate points are set to provide alarm outputs when the intermediate frequency error exceeds  $\pm 200$  kHz or the change in IF output power exceeds  $\pm 2$  dB. The monitor unit also provides two independent 75-

ohm baseband outputs that can supply a 0 dBm signal for 8 MHz peak-to-peak sine wave deviation.

The frequency alarm point at an error of  $\pm 200$  kHz is beyond the  $\pm 100$  kHz objective for transmitter frequency stability. Since a properly functioning transmitter will stay within the 100 kHz objective, the primary function of the 200 kHz alarm indication is to call attention to a trouble condition when it reaches a point where it will begin to have an adverse effect on system performance.

The discriminator-baseband amplifier portion of the monitor circuit is very similar to the corresponding circuitry in the 3A FM receiver which is described in Section 4.3. Power and frequency alarm indications are derived from the sum and difference outputs, respectively, of the two balanced discriminator detector diodes.

### 3.7 Transmitter Control Unit

The transmitter control unit provides individual alarm outputs, an alarm cutoff feature, and alarm circuit breaker protection, self-contained within the terminal. This unit is optional and is not provided when the monitor unit is omitted.

Off-frequency and IF level error signals from the monitor unit operate individual meter relays with high and low latching type contacts. The meter relays provide external alarm indications and a local visual indication of trouble conditions.

## IV. FM RECEIVER

The FM receiver accepts the 70 MHz FM signal and delivers a balanced baseband output of +4 dBV in a 124 ohm circuit for 8 MHz peak-to-peak deviation.

Performance requirements with respect to linearity, baseband response, and stability are similar to (and must be compatible with) the over-all terminal pair objectives.

The design is fairly conventional and similar to the approaches used on corresponding equipment in the TH and TD-2 systems.<sup>2, 3</sup>

### 4.1 General Description

Figure 17 is a block diagram of the FM receiver. The IF input to the receiver is applied to the 75-ohm input of the limiter-amplifier at a power of -7 dBm (+3 dBm on an earlier version of the receiver). Here, the first of two limiters is the major source of AM suppression in the receiver and has low amplitude-to-phase conversion. Follow-

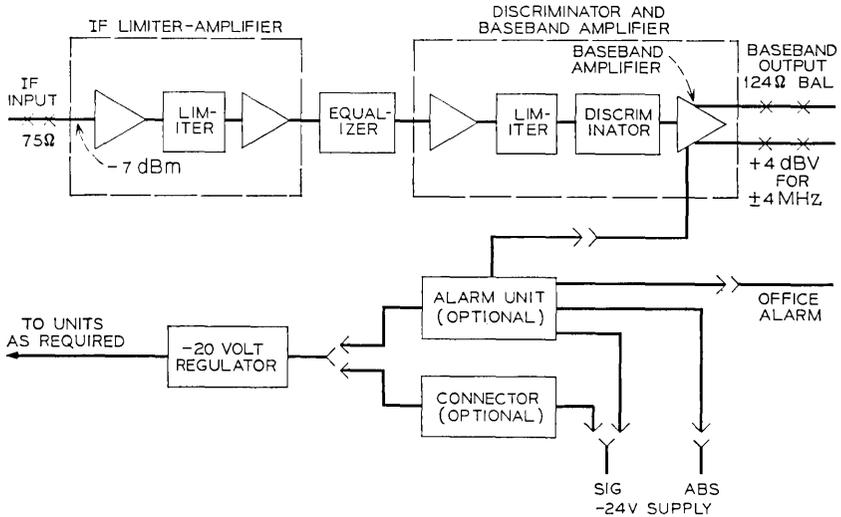


Fig. 17 — Block diagram of FM receiver.

ing the limiter are several stages of gain and an equalizer which corrects for delay distortion in the over-all receiver. The second limiter extends the dynamic range of the receiver such that performance is essentially unaffected by reductions up to 10 dB in IF input power.

The output of the second limiter is applied to the discriminator where it is demodulated. The recovered signal is then amplified and an output of +4 dBV is provided from a balanced 124-ohm impedance for a peak-to-peak deviation of 8 MHz.

An optional alarm unit in the receiver provides an indication of low or high IF level at the discriminator and an indication of changes in the dc conditions on the baseband amplifier transistors. Special care was taken in the design of the receiver to ensure reproducible adjustment of the discriminator networks which are the major source of envelope delay distortion in the terminal. Field adjustment controls were also carefully selected to minimize the possibility of degrading the delay equalization.

Figure 18 is a photograph of the receiver.

#### 4.2 IF Limiter-Amplifier

Early models of the limiter-amplifier operated at an IF input power of +3 dBm. This was later changed to -7 dBm to compensate for office cabling losses.

A common emitter configuration is used for the input stage while all other gain stages have the common base configuration used in the repeater IF amplifier described by G. L. Fenderson and others.<sup>7</sup> The series diode limiter is of the compensated type described in the same paper.

The limiter-amplifier suppresses amplitude modulation by at least 40 dB at nominal drive. It provides an output of +8 dBm, flat within 0.03 dB between 60 and 80 MHz.

#### 4.3 Discriminator-Baseband Amplifier

The discriminator-baseband amplifier performs the demodulation function in the FM receiver. The first stages of the unit comprise a limiter and a low pass filter. This limiter reduces amplitude modulation by 25 dB and stabilizes the IF input to the discriminator. Because of the combined action of the two receiver limiters, a 10 dB drop in IF input (from nominal) produces less than a 0.25 dB reduction in baseband output. This is illustrated in Fig. 19 where receiver sensitivity is plotted as a function of IF input power.

The limiter output is temperature stabilized by thermistor control of the limiter diode bias current. The limiter circuit is very similar to corresponding stages of the amplifier-limiter described in Section 3.4. A low pass filter following the limiter rejects harmonics of the 70 MHz IF to prevent distortion products being generated in the discriminator detectors.

To make the baseband amplifier noise contribution negligible, it

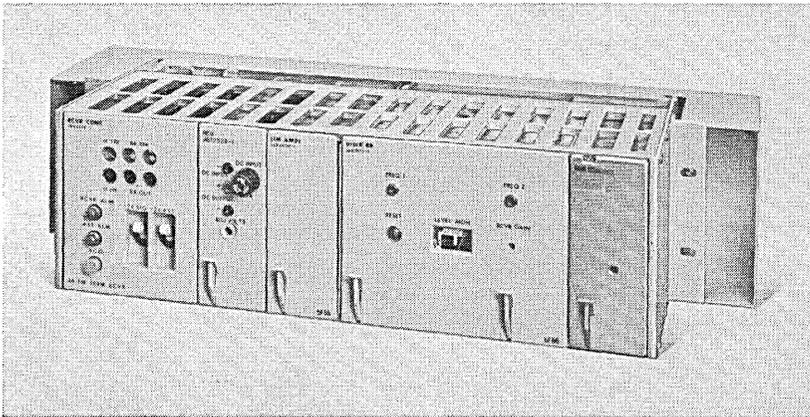


Fig. 18 — 3A FM receiver.

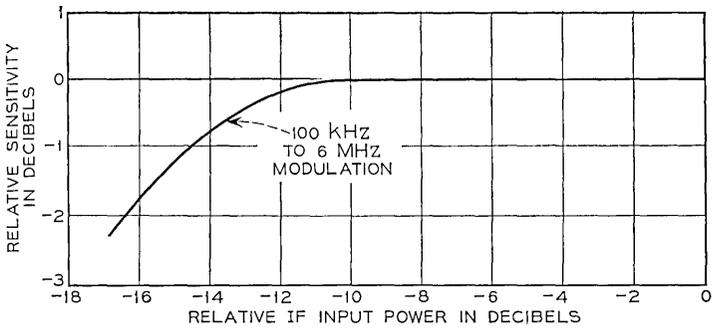


Fig. 19 — Relative receiver demodulation sensitivity as a function of IF input power.

was necessary to provide a discriminator sensitivity of at least 0.25 volt per MHz. For a given input power and degree of complexity, there is in general a trade-off between discriminator sensitivity and linearity. Using the ratio of sensitivity to nonlinearity as a figure of merit, several circuit configurations were compared before selecting the modified balanced resonant discriminator similar to that used in the TH radio system.<sup>3</sup>

To determine the relationship between sensitivity and nonlinearity in the balanced resonant discriminator, the idealized circuit of Fig. 20 was optimized on the computer. The computer was programmed to adjust the circuit elements to obtain the best match between  $|E_2| - |E_3|$  as a function of frequency and a straight line of specified slope

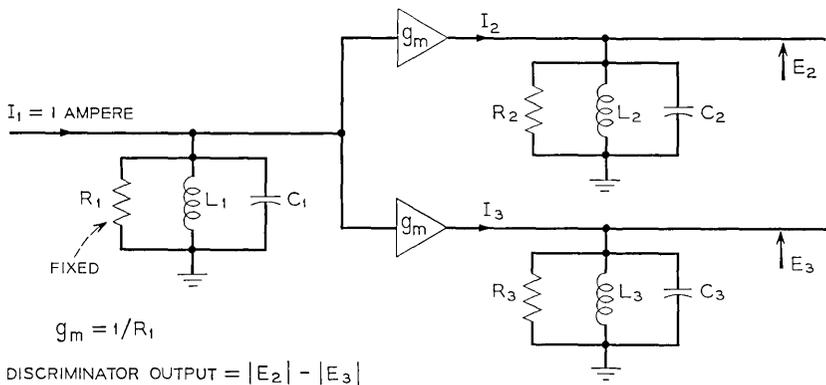


Fig. 20 — Idealized balanced resonant discriminator circuit.

passing through zero at 70 MHz. By repeating this process for several different slope objectives corresponding to different sensitivities, the relationship between sensitivity, linearity, and element values was determined.

The amplifiers in Fig. 20 have infinite input and output impedances whereas in the actual circuit, common-base transistor stages with low input impedances are used. If the damping resistor  $R_1$  is in series with the emitters, the emitter currents are proportional to the voltage across  $R_1$  so that the idealized circuit becomes a satisfactory representation of the actual circuit. Good agreement was obtained between measured and computed sensitivity and linearity. In Fig. 21, the high and low frequency tank circuits in the collectors of  $Q_6$  and  $Q_4$  provide impedance peaks at 89 and 51 MHz, respectively. The highly damped tank in the collector circuit of  $Q_3$  provides shaping of the common IF signal around 70 MHz to attain an over-all discriminator linearity of better than 0.5 per cent for 8 MHz peak-to-peak deviation.

Printed wiring gives close control of most of the stray inductance and, to a considerable extent, the stray capacitance. By using close tolerance, tension wound inductors, and by adjusting the circuit capacitance with trimmers to correct for differences in transistor and diode capacitances, excellent reproducibility of the discriminator delay characteristics was obtained. The high and low frequency tank circuits of the discriminator need no field adjustments. In the common or center frequency tank circuit, both the inductor and capacitor are variable to facilitate alignment.

An important and useful technique to reduce and control stray inductance and couplings in high frequency ground paths was the use of a ground plane as illustrated in Fig. 22. This ground plane is placed about one half inch below the printed wiring board with threaded studs connecting wherever ground reinforcement is needed on the printed wiring board.

A high-speed epitaxial silicon Schottky-barrier diode, coded 479B, with low forward voltage and high reverse breakdown voltage was developed for the 70 MHz detector. High reverse breakdown voltage allows relatively high discriminator drive with correspondingly improved discriminator sensitivity. A combination of high detector load impedance and high detection efficiency results in the detector circuits contributing less than 10 per cent to discriminator tank circuit loading.

The high sensitivity of the discriminator provides such high signal

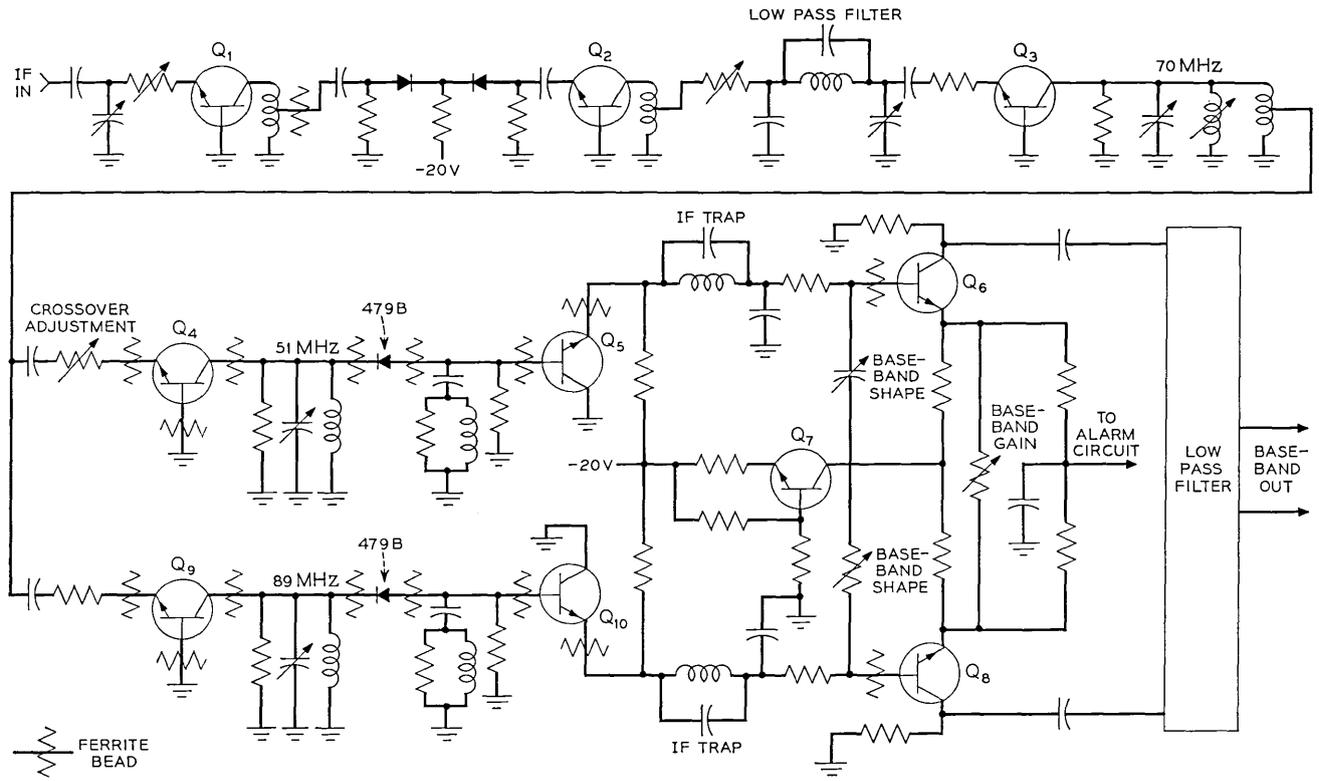


Fig. 21 — Simplified schematic diagram of discriminator-baseband amplifier unit.

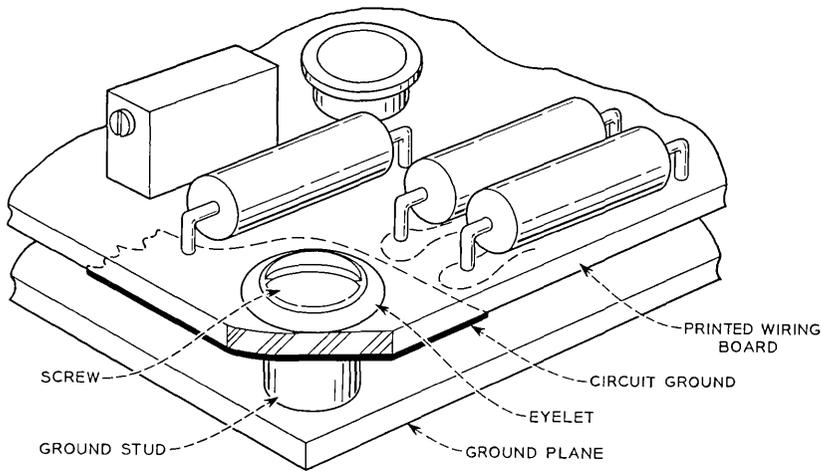


Fig. 22 — Grounding technique.

input to the baseband amplifier that little voltage gain is needed and the amplifier noise is relatively unimportant. The controlling design requirements for this amplifier were impedance transformation to the 124-ohm output, differential balance, and frequency response. The low-end frequency response of 3 dB down at 0.6 Hz led to the choice of dc coupling between stages and a circuit configuration permitting the use of low-voltage output coupling capacitors to minimize space requirements.

Emitter followers were chosen for the input stages of the amplifier to provide a high impedance load for the discriminator. Resonant 70 MHz traps in the emitter follower outputs attenuate the IF signal and its harmonics. These IF signals are further attenuated by the low pass filter in the amplifier output to be less than 3 millivolts from tip or ring to ground. The high end frequency response is controlled by an adjustable RC filter which is connected between the inputs of  $Q_6$  and  $Q_8$  to adjust the over-all transmitter and receiver frequency response flat to  $\pm 0.1$  dB from 6 Hz to 10 MHz.

The baseband signals from the two halves of the discriminator must be subtracted to cancel their first-order nonlinearities. The differential amplifier,  $Q_6$ ,  $Q_7$ , and  $Q_8$ , performs this combining function.

An optional IF level alarm is used to detect the loss of IF signal in the receiver. The discriminator diode detector outputs are dc coupled through the baseband amplifier to the emitters of  $Q_6$  and  $Q_8$ .

The IF level on both halves of the discriminator as well as baseband amplifier bias conditions are therefore monitored by summing these emitter voltages. The monitored voltage is stored in a capacitor and compared to a reference voltage by a zero center, latching-type, meter relay. The meter relay will operate with either an increase or decrease in monitored voltages equivalent to 3 dB changes in IF level at the discriminator. The capacitor smooths rapid voltage changes to prevent alarms on hit type failures. It also stores sufficient energy to latch the meter relay when circuit power is removed. The meter relay contacts energize alarm relays in the control unit to provide external alarm indications, and a local visual indication of trouble.

#### 4.4 Receiver control Unit

The receiver control unit serves the same purpose and is very similar to the corresponding transmitter unit. In this case, however, the meter relay which responds to loss of IF signal at the discriminator is located on the discriminator-baseband amplifier unit.

### V. POWER REQUIREMENTS

The FM terminals operate from a  $-24$  volt office battery. The transmitter requires a maximum of 3.0 amperes from the  $-24$  volt supply (signal battery) with a typical requirement of 1.5 amperes at ambient temperatures around  $20^{\circ}\text{C}$ . The dependence on temperature results primarily from variations in power requirements on the deviator oven. The  $-24$  volt receiver drain is 0.75 ampere. When equipped with alarm and control units, the transmitter and receiver each require a maximum of 0.2 ampere from the  $-24$  volt alarm battery supply.

In both the transmitter and receiver, the signal battery supply is fed through a voltage regulator providing an output of  $-20$  volts  $\pm 1$  percent.

### VI. TANDEM PERFORMANCE

Based on laboratory and production experience, the tandem performance of the FM transmitter and receiver has shown a very satisfactory degree of reproducibility so that the quoted performance can be regarded as typical. Compared with similar vacuum tube equipment, performance variations with time are relatively small and in most instances there is not yet enough experience to quantitatively characterize the stability.

6.1 *Noise Load Performance*

The most satisfactory measure of terminal performance from the standpoint of message applications is based on the use of band-limited fluctuation noise to simulate the multi-channel speech load. Noise and cross-modulation performance measured in this manner is presented in Figs. 23, 24, and 25 corresponding to 600, 900, and 1200 channel message loading, respectively. The performance at reference drive with 1200 channel loading is well within the TD-3 objective of 21 dBrnC0.

6.2 *Fluctuation Noise*

Fluctuation noise, generated by a tandem transmitter and receiver, and measured at the receiver output, is shown in Fig. 26. This total noise is made up of several distinguishable contributors as indicated in the figure. Fluctuation noise at high baseband frequencies is controlled by the noise figure of the transmitter amplifier-limiter which has a relatively low input signal power of  $-26$  dBm. In the baseband region below about 1 MHz, the main fluctuation noise sources in

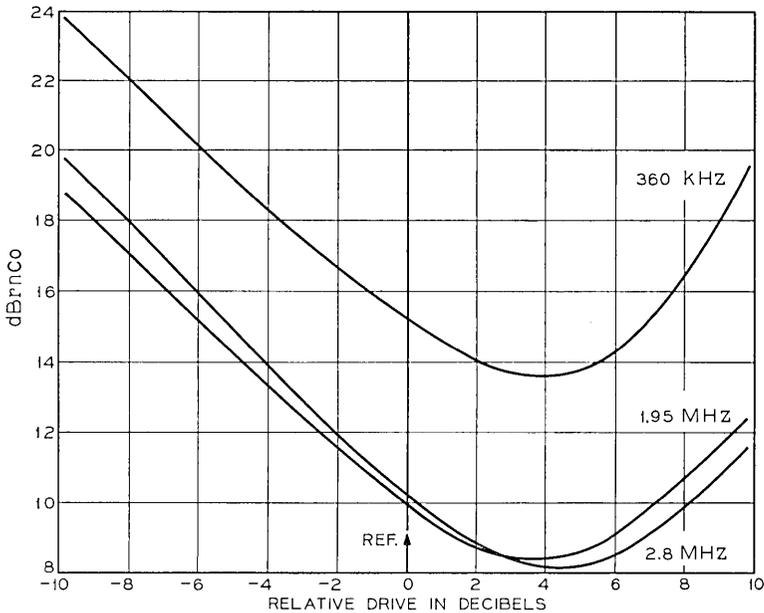


Fig. 23 — 600 channel noise load performance of a terminal pair; reference drive at  $-23.8$  dBm.

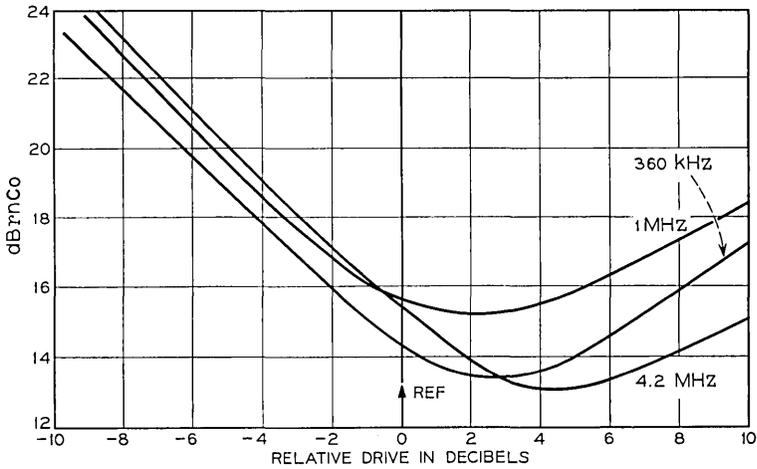


Fig. 24—900 channel noise load performance of a terminal pair; reference drive at  $-23.5$  dBm.

order of importance are the transmitter baseband amplifier, the deviator, and the receiver.

For television, in addition to the spectral nature of the noise, the ratio of signal to total noise is of interest. This is usually expressed as the ratio of peak-to-peak video signal to weighted (rms) noise

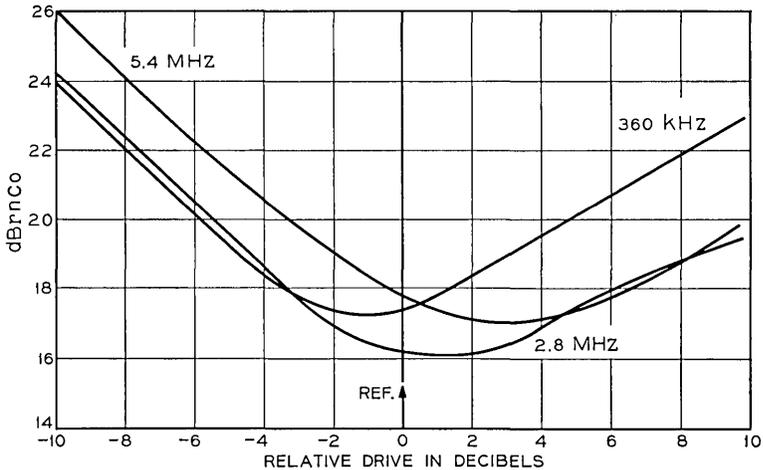


Fig. 25—1200 channel noise load performance of a terminal pair; reference drive at  $-23.2$  dBm.

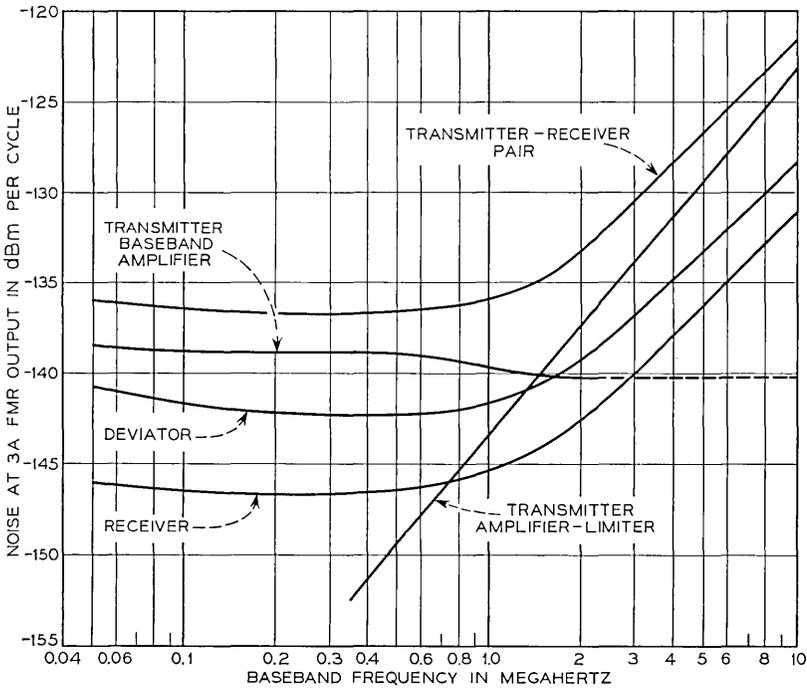


Fig. 26—Composition of FM terminal fluctuation noise.

where the color weighting curve is as shown in Fig. 27. For the noise measurement, the weighted baseband noise is generally separated into two regions, one encompassing the low frequency end of the band up to 4 kHz and the other encompassing the range from 4 kHz up to approximately 4.5 MHz. The noise is measured at the FM receiver output following the video de-emphasis network. Typical values for signal-to-weighted low frequency noise and signal-to-weighted (color) high frequency noise are 91 and 84 dB, respectively, meeting the corresponding 40 and 73 dB objectives with a comfortable margin.

6.3 Baseband Amplitude Response

Figure 28 shows a typical baseband response characteristic.

For television purposes the low frequency response is frequently characterized by the response to a 60 Hz square wave. Slope on the square wave output, expressed as a percentage of the peak-to-peak

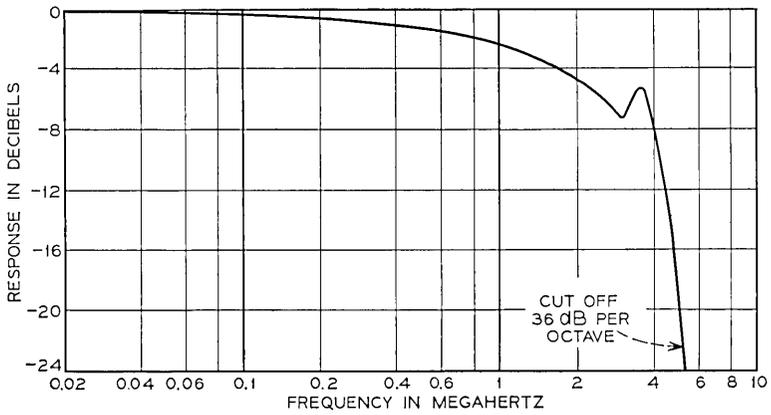


Fig. 27 — Color TV noise weighting curve.

signal is a measure of the phase fidelity at the fundamental frequency. Typical slope, measured at the output of a 3A terminal pair, is 1.5 per cent, which is consistent with a lower 3 dB cutoff frequency of approximately 0.5 Hz.

6.4 *Differential Gain, Differential Phase and Linearity*

With 8 MHz peak-to-peak deviation, the differential gain is 0.1 dB and the differential phase is 0.15 degree; both are predominantly linear in shape. Since pre-emphasis reduces the low frequency video components by 12 dB, the differential gain and phase experienced by the TV signal will be a factor of four less and is therefore well within the design objectives. Over-all nonlinearity is primarily determined by the transmitter which has a residual parabolic component of typically 1.5 per cent for 8 MHz peak-to-peak deviation.

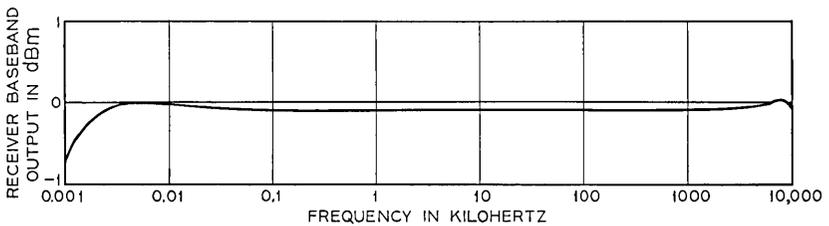


Fig. 28 — Baseband frequency response of a terminal pair.

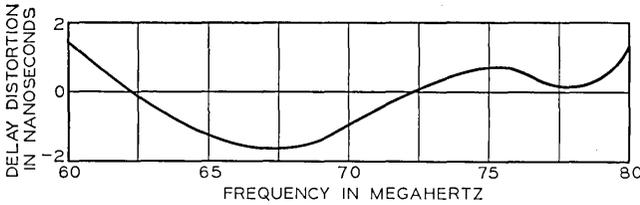


Fig. 29 — Envelope delay distortion of a terminal pair.

6.5 Envelope Delay Distortion

For all practical purposes the FM transmitter and receiver are separately delay equalized. In the transmitter there is residual envelope delay distortion of approximately 2 nanoseconds parabolic at  $\pm 10$  MHz. The residual delay distortion of the over-all terminal pair is shown in Fig. 29.

6.6 Spurious Transmitter Output Tones

Spurious tones present in the deviator output are band-limited and amplitude-limited in succeeding portions of the transmitter. The resulting relative amplitudes of these spurious tones are listed in Table II.

VII. TEST FACILITIES

Because of the expected frequent application of 3A FM terminals in existing TD-2 stations, it was decided to make maximum use of existing terminal test facilities. For the maintenance of 3A terminals, the TD-2 FM terminal test console is supplemented by a new test panel which replaces a blank panel in the test console. The composi-

TABLE II — OUT-OF-BAND TONES AT THE FM TRANSMITTER OUTPUT

Tone frequency (MHz)	Product	Power relative to 70 MHz IF output (dB)
23.33	4V-3C	< -87
46.67	3V-2C	-76
93.33	5V-4C	-79
116.67	2V-C	-69
140.00	2C-2V	-23
186.67	V	-86
256.67	C	< -87

tion and functions of the test panel are described fully in Ref. 8.

The general approach to maintenance testing has been to confine routine tests to input-output measurements. Trouble on individual units is isolated by substituting a spare unit for the suspected defective unit.

#### VIII. EQUIPMENT FEATURES

All major components in the 3A transmitter and receiver are plug-in units for easy replacement and maintenance. They are located in an open box framework suitable for mounting on a 19-inch duct bay. The transmitter has two rows of plug-in units totaling 10½ inches high. The receiver has one row of units totaling 5¼ inches high. The transmitter and receiver weigh about 45 and 18 pounds, respectively.

The terminals also are available in portable carrying cases. A companion 117 volt, 60 Hz rectifier unit is available with an output capacity of 4 amperes at -24 volts. This unit can power a transmitter or receiver individually or both units simultaneously. Since the FM receiver may be used more frequently in portable form and because its power requirements are modest, an optional 117 volt rectifier unit is available, mounted inside the rear cover of the receiver case. In all instances the portable terminals are equipped for optional connection direct to the -24 volt office battery. Figure 30 shows the portable terminals with front covers removed and stacked for use.

#### APPENDIX

##### A.1 *Linearity of Varactor Diode Voltage Controlled Oscillator*

In the region where a quasistationary analysis is appropriate, the relationship between instantaneous frequency and modulating voltage may be expressed as a power series of the form

$$f(v) = f_0 + \sum_i a_i v^i \quad (1)$$

where  $f(v)$  = instantaneous oscillator frequency

$f_0$  = oscillator rest frequency

$v$  = applied signal voltage.

Using an approach similar to that discussed in Appendix A of Ref. 3, the following relationships can be defined in terms of the coefficients in equation (1).

$$\text{Deviation sensitivity} = a_1 \tag{2}$$

$$\text{First order nonlinearity} = \frac{4a_2 \Delta f}{a_1^2} \tag{3}$$

$$\text{Second order nonlinearity} = \frac{3a_3 \Delta f^2}{a_1^3} \tag{4}$$

where  $\Delta f$  = peak frequency deviation.

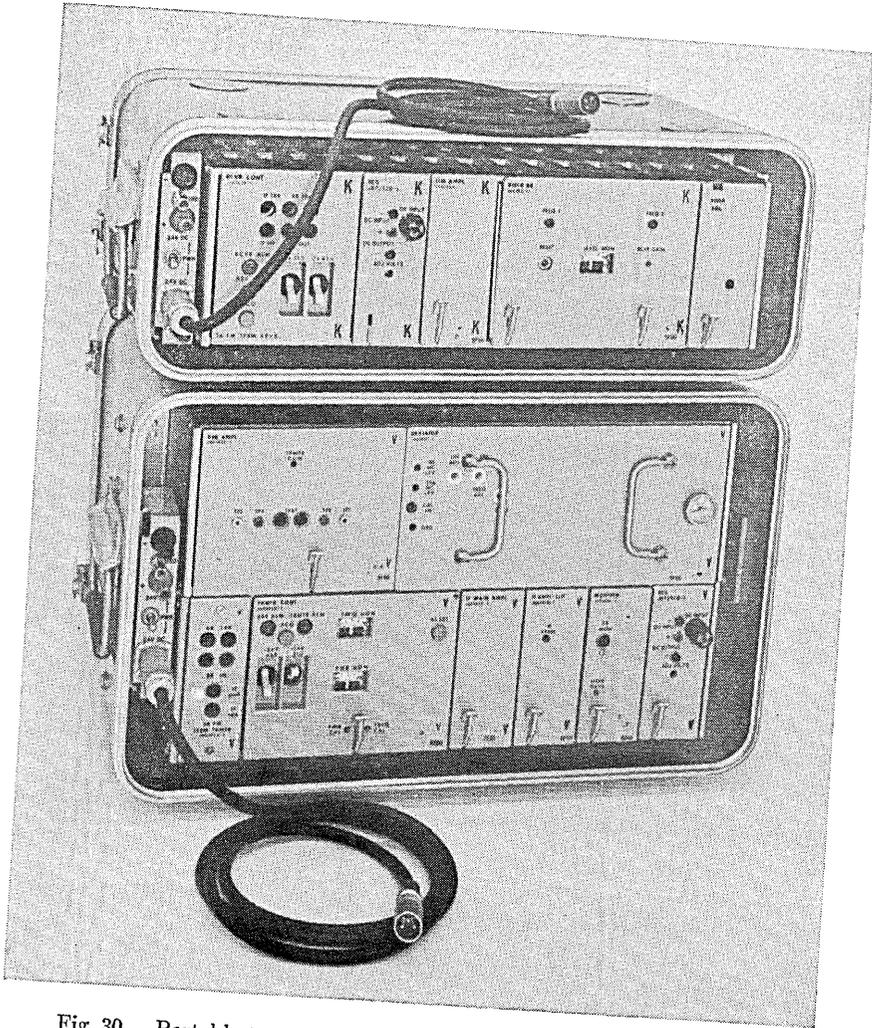


Fig. 30 — Portable transmitter and receiver equipped for dc operation.

Notice that equation 3 defines peak-to-peak nonlinearity in contrast with Ref. 3, which defines the peak value. The choice is somewhat arbitrary except that the peak-to-peak definition is more in keeping with values commonly measured on linearity test equipment.

Consider the parallel resonant tank circuit of a voltage controlled oscillator as shown in Fig. 31.

- $L$  = effective shunt inductance
- $C_s$  = total stray capacitance
- $C_v(v)$  = varactor diode capacitance.

The resonant frequency of this circuit is

$$f(v) = \frac{1}{2\pi} [L(C_v(v) + C_s)]^{-1/2}. \quad (5)$$

The diode capacitance  $C_v(v)$  can be written in the form

$$C_v(v) = \frac{C_0}{(V_0 + v)^n} \quad (6)$$

where  $C_0$  is constant,  $V_0$  is the fixed bias voltage on the diode junction and  $n$  is an exponent which depends on the doping profile at the diode junction. For abrupt junction diodes  $n$  is approximately  $1/2$  while the exponent for graded junction devices is close to  $1/3$ .

By substituting equation 6 in equation 5 a closed form expression for  $f(v)$  in terms of the modulating voltage  $v$  is obtained. This expression can be expanded in a MacLaurin series about  $v = 0$  and the resulting coefficients can be equated with those in equation 1 to give

$$a_1 = \frac{nf_0}{2V_0} \left(1 - \frac{C_s}{C_B}\right) \quad (7)$$

$$a_2 = \frac{1}{2} \frac{nf_0}{(2V_0)^2} \left(1 - \frac{C_s}{C_B}\right) \left(n - 3n \frac{C_s}{C_B} - 2\right) \quad (8)$$

$$a_3 = \frac{1}{6} \frac{nf_0}{(2V_0)^3} \left(1 - \frac{C_s}{C_B}\right) \cdot \left[ n^2 - 6n + 8 - 6n(2n - 3) \frac{C_s}{C_B} + 15n^2 \left(\frac{C_s}{C_B}\right)^2 \right]. \quad (9)$$

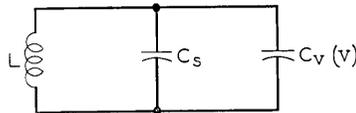


Fig. 31 — Voltage controlled oscillator tank circuit.

In the above expressions,  $C_B$  is the total tank circuit capacitance with  $v = 0$  or

$$C_B = C_s + \frac{C_0}{V_0^n} \tag{10}$$

Equations 7, 8 and 9 may now be substituted in equations 2, 3 and 4 to yield the following results.

$$\text{Deviation sensitivity} = \frac{nf_0}{2V_0} \left(1 - \frac{C_s}{C_B}\right) \tag{11}$$

$$\text{First order nonlinearity} = \frac{\Delta f}{f_0} \left[ \frac{2\left(n - 3n \frac{C_s}{C_B} - 2\right)}{n\left(1 - \frac{C_s}{C_B}\right)} \right] \tag{12}$$

Second order non-linearity

$$= \left(\frac{\Delta f}{f_0}\right)^2 \left[ \frac{n^2 - 6n + 8 - 6n(2n - 3) \frac{C_s}{C_B} + 15n^2 \left(\frac{C_s}{C_B}\right)^2}{2n^2 \left(1 - \frac{C_s}{C_B}\right)^2} \right] \tag{13}$$

Notice that the deviation sensitivity is inversely proportional to the bias voltage and that the nonlinearities depend on the percentage deviation.

### A.2 Linearity of the Heterodyne Deviator

The heterodyne deviator in the 3A transmitter uses two voltage controlled oscillators whose rest frequencies differ by 70 MHz and which are deviated by baseband signals in antiphase. The desired difference frequency at the mixer output may therefore be expressed as

$$f(v) = f_h(v/2) - f_\ell(-v/2) = f_{0h} - f_{0\ell} + \sum_i (1/2)^i [a_{ih}v^i - a_{i\ell}(-v)^i] \tag{14}$$

where the subscripts  $h$  and  $\ell$  refer to the high and low frequency oscillators, respectively. Also notice that  $v$  in equation 14 is the total balanced input voltage to the deviator which accounts for the  $(\frac{1}{2})^i$  term in the power series.

With  $V_{0\ell}$  and  $C_{s\ell}/C_{B\ell}$  defined, there are two degrees of freedom in the high frequency oscillator, namely,  $V_{0h}$  and  $C_{sh}/C_{Bh}$ . This permits satis-

fyng two arbitrary conditions by appropriate choice of these quantities. In the 3A deviator it was chosen to design the two oscillators with equal sensitivities and equal  $a_2$  terms. To satisfy these requirements,  $C_{sh}/C_{Bh}$  must be related to  $C_{st}/C_{Bt}$  in the following way.

$$\frac{C_{sh}}{C_{Bh}} = 1 - \frac{2(n+1)\left(1 - \frac{C_{st}}{C_{Bt}}\right)}{3n\left(1 - \frac{C_{st}}{C_{Bt}}\right) - \frac{f_{0h}}{f_{0t}} - \frac{f_{0h}}{f_{0t}}\left[n - 3n\frac{C_{st}}{C_{Bt}} - 2\right]}. \quad (15)$$

When the preceding conditions have been satisfied, the first order nonlinearity is zero and the dominant second order nonlinearity is given by

$$\frac{3(a_{3h} + a_{3t}) \Delta f^2}{(a_{1h} + a_{1t})^3}. \quad (16)$$

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## Test Equipment

By R. T. COONEY, F. M. KLISCH, and C. P. SUSEN

(Manuscript received January 4, 1968)

*This paper describes the transmitter-receiver test set and the FM terminal test panel, designed for maintaining the TD-3 radio equipment, and discusses the instrumentation and measuring techniques used to achieve accuracy consistent with the high performance of the TD-3 system. Circuits of special interest are covered in some detail.*

### I. INTRODUCTION

The TD-3 test equipment for routine maintenance and troubleshooting procedures is designed to match the high performance of the TD-3 system and is simple and convenient to operate. The units have good long term stability and may be readily interconnected for testing or self-calibration. This is desirable because much of the TD-3 equipment is located in remote, unmanned sites where testing time must be kept to a minimum.

The equipment consists of a transmitter-receiver test set mounted in a rolling console and an FM terminal test panel mounted in (and used with) the existing TD-2 FM terminal test set.

The major functions of the test set are to measure amplitude and return loss in the 3700 to 4200 MHz (RF) and the 50 to 100 MHz (IF) frequency ranges, AM noise figure, frequency, and RF and IF power. These are similar to the functions of other test consoles,<sup>1,2</sup> but significant advances have been made in implementation. Solid state devices and RF coaxial instrumentation made such advances possible. Accurate IF frequency markers, low level power measurements, and direct reading frequency measurements are used extensively.

The FM terminal test panel is designed to enable the TD-2 FM terminal test set to test TD-3's 3A FM terminals. A major influence to this approach was that the 3A terminals<sup>3</sup> are now standard equipment in TD-2 production and are replacing TD-2 terminals in the TD-2 improvement program. Because many 3A terminals are being

installed in TD-2 stations, it was decided to use existing terminal test equipment wherever feasible. The test panel has the greater accuracy required for 3A terminals.

The new test panel, which replaces a blank panel in the TD-2 FM terminal test console,<sup>2</sup> measures FM transmitter deviation sensitivity, deviator oscillator frequencies, baseband transmission, and sensitivity of the frequency alarm and level alarm circuits.

## II. TRANSMITTER-RECEIVER TEST SET

The transmitter-receiver test set is used to measure IF and RF amplitude characteristics, IF and RF return loss, IF and RF power, frequency, and AM noise figure. Its major instruments are the IF and RF sweep oscillators, RF and IF detectors, oscilloscope, power meter, counter, IF amplifier, and noise lamp. They are integrated into a single console with a common control circuit, their power supplies, and peripheral equipment. The common control circuit is a centralized point for interconnecting many of the instruments.

Figure 1 shows the console, which also houses cables, coaxial to waveguide transducers, attenuators, and similar items.

### 2.1 *Design Considerations*

In recent years high quality test instruments have become increasingly available from many sources. Such instruments have been incorporated in the test set, usually with minor modifications. New circuits were developed to perform specific functions for which instruments were not available.

One of the most stringent functions that the test set must perform is to measure amplitude deviations of approximately one hundredth of a dB on IF to IF, IF to RF, RF to IF, and RF to RF measurements. This usually is achieved in laboratory and factory test apparatus by the comparison technique; that is, comparing the amplitude characteristic of the unit being tested with the characteristic of a known standard. But this is not practical for field tests.

The TD-3 transmitter-receiver test set uses the simpler and more economical direct method of amplitude measurement, which consists of inserting the unit to be tested between a signal source which has constant amplitude vs frequency and a detector which gives constant dc output vs frequency. This method trades the complexity of the comparison method for more stringent requirements on the swept signal source and the detector.

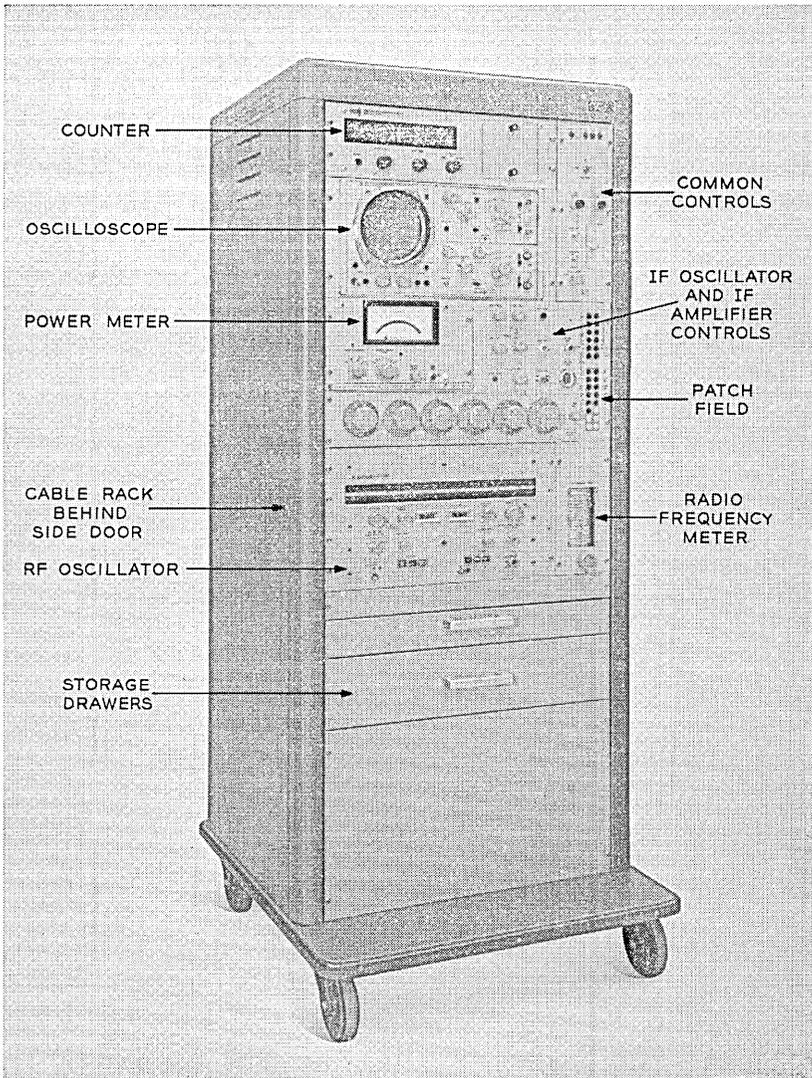


Fig. 1—The transmitter-receiver test set.

Measuring amplitude deviations of approximately one hundredth of a dB by direct techniques ideally would require test apparatus with deviations much less than 0.01 dB. This is beyond present technology. The over-all deviations in the test set IF amplitude measuring apparatus are less than 0.15 dB from 60 to 80 MHz and in the

RF apparatus the deviations are less than 0.02 dB over any 20 MHz portion of the band. This is considered sufficient for routine TD-3 maintenance and adjustment.

The test set components must have excellent level stability and low power supply "hum" to display amplitude characteristics on the oscilloscope at high sensitivities. The sensitivity may be as high as 0.05 dB per centimeter of vertical oscilloscope deflection. This corresponds to an oscilloscope sensitivity of approximately 1 mv per centimeter. The level stability of the IF and RF units must be such that the oscilloscope trace changes considerably less than 0.05 dB while the measurement is being made and the power supply hum must be kept much lower than 1 mv.

The required stability was achieved by using stable power supplies,<sup>4</sup> temperature-compensated circuitry, and feedback techniques. Power supply hum is kept satisfactorily low by using low ripple power supplies,<sup>4</sup> by eliminating ground loops, and by shielding dc leads from transformers, motors, and other sources of power supply hum. Ground loop problems have been severe in other test sets,<sup>1</sup> particularly when connected with radio bays whose ground potential is slightly different from the test set ground. The TD-3 test set eliminates the ground loop by using an IF detector mounted in the test set and by using a dc block with the RF detector. The total power supply hum on the test set oscilloscope trace is less than 0.001 db, or 0.2 mv.

Coaxial cables rather than waveguides are used through the RF portions of the test set for compactness and the convenience of the operator. Impedance interactions from the cable connecting the input of the unit being tested to the test set are eliminated by using a leveling detector. This detector is located at the end of the coaxial cable at the input to the unit being tested and provides a dc control voltage to regulate the output power of the RF oscillator. This insures a constant input level to the unit being tested regardless of any distortions introduced by the connecting cable. The output power of the IF oscillator is regulated within the test set since impedance interactions in the IF connecting cables are small.

## 2.2 *Transmission Measurements*

Figure 2 illustrates the method of making transmission measurements. Either the RF oscillator or the IF oscillator may be the signal source. Similarly, either the RF or IF detector may be used.

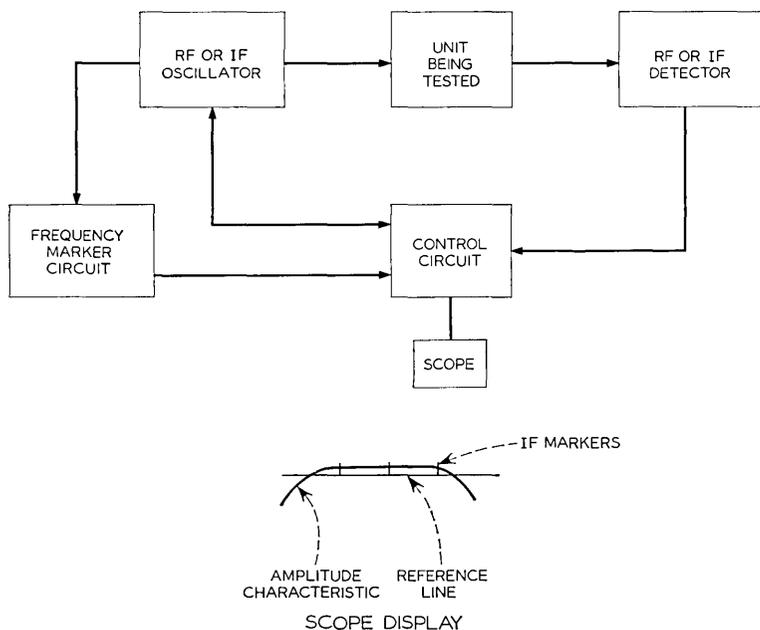


Fig. 2 — Transmission measurement.

This permits measurement of any IF or RF unit, modulator, or combination of units. The oscillators give constant output power and the detectors have flat frequency responses over their respective bands. This permits a direct display on the oscilloscope of the amplitude characteristic without the necessity of any corrections for test set characteristics.

The amount of patching required is kept to a minimum consistent with test set flexibility. The desired mode of operation is selected by a control switch which connects the output of the appropriate detector and frequency marker through the control circuit to the oscilloscope. The control circuit alternately connects the test signal and a dc reference voltage to the oscilloscope at a 31 Hz rate.

The oscilloscope display consists of a plot of the amplitude characteristic vs frequency of the unit being tested with appropriate frequency markers and a dc reference line. The vertical sensitivity is variable to 0.05 dB per cm maximum sensitivity. The sweep width is variable from 0 to 50 MHz and from 0 to 500 MHz for the IF and RF oscillators, respectively.

### 2.3 IF Return Loss

The IF return loss bridge shown in Fig. 3 measures IF return loss. The bridge separates the wave reflected from the unit being tested from the incident wave, and returns it to the test set to be amplified, detected, and presented on the oscilloscope. The oscilloscope displays return loss vs frequency. The dc reference line provides a calibrated reference which is set with the test set IF attenuator. IF frequency markers indicate frequency. The accuracy of the measurement is  $\pm 1.5$  dB when measuring 40 dB return losses. Lower return losses are measured with correspondingly better accuracy.

The bridge technique overcomes the disadvantages of directional coupler and reflectometer methods sometimes used. A directional coupler has a 6 dB per octave frequency characteristic which must be compensated for when displaying the return loss characteristic. This is generally difficult to achieve with sufficient accuracy. The

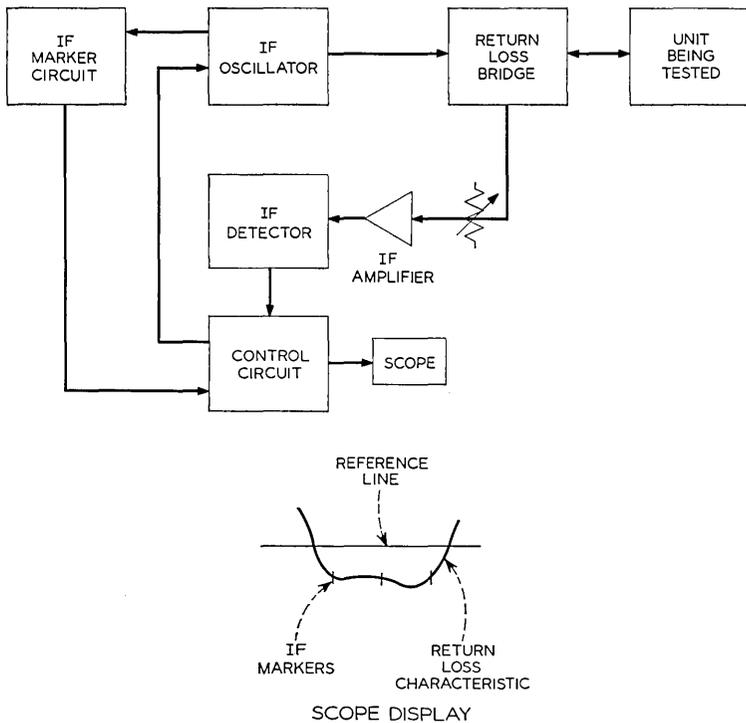


Fig. 3 — IF return loss measurement.

return loss bridge has a flat frequency response and requires no compensation. Reflectometer measurements using amplitude modulation of the test signal are affected by nonlinear elements in the test circuit, thus resulting in poor accuracy. No modulation is required with the bridge method.

#### 2.4 RF Return Loss

With this test set, RF return loss measurements can be made only on circuits having the reflecting discontinuity located at a considerable electrical distance from the test equipment. This restriction was imposed because it satisfies the TD-3 system requirements and allows a simple measurement technique to be used. The technique consists of combining the incident and reflected waves in a detector by means of a power divider as Fig. 4 shows. As the frequency of the oscillator is varied, the phase difference between the incident and the reflected wave changes. For the normal 20 MHz sweep width and waveguide lengths of greater than 10 feet, several in phase and out of phase conditions exist over the swept frequency range. The output of the detector changes as the phase changes, thus producing a ripple on the oscilloscope. The amplitude of the ripple is a function of the return loss, and the frequency spacing of the ripples is a function

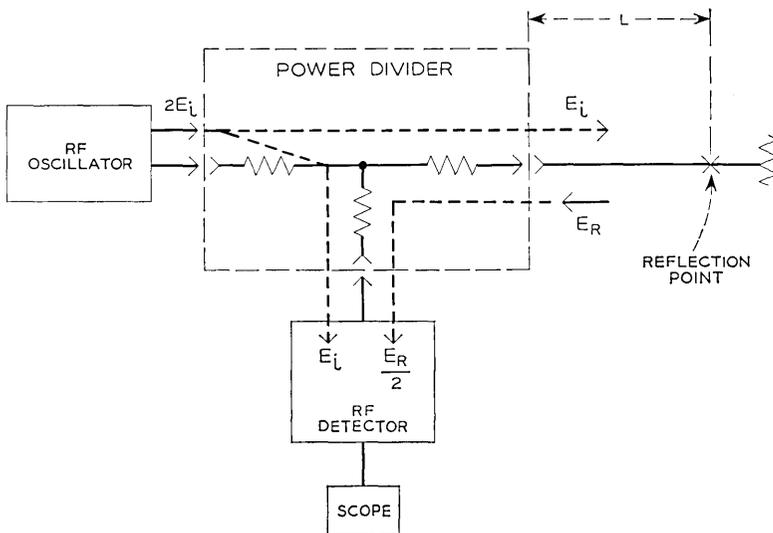


Fig. 4 — RF return loss measurement.

of the electrical distance between the power divider and the point of reflection. Thus, information on the magnitude and location of the reflection is obtained from this measurement.

The magnitude of the peak to peak ripple in dB may be determined from the calibrated oscilloscope display. The ripple amplitude may easily be converted to return loss. As shown in Fig. 4, the magnitude of the incident wave at the detector input is  $E_i$  and the magnitude of the reflected wave is  $E_r/2$ . The factor of  $1/2$  takes the 6 dB loss of the power divider into account. When the two signals are in phase the input is  $E_i + E_r/2$  and when the signals are out of phase the input is  $E_i - E_r/2$ . The magnitude of the ripple ( $R$ ) in dB is:

$$R = 20 \log \left[ \frac{E_i + E_r/2}{E_i - E_r/2} \right] = 20 \log \left[ \frac{1 + |\Gamma|/2}{1 - |\Gamma|/2} \right]. \quad (1)$$

$$\Gamma = \text{reflection coefficient} = E_r/E_i$$

Solving (1) for  $\Gamma$

$$|\Gamma| = 2 \left[ \frac{10^{R/20} - 1}{10^{R/20} + 1} \right].$$

But, return loss ( $RL$ ) =  $-20 \log 1/|\Gamma|$

$$RL = 20 \log \left[ \frac{10^{R/20} + 1}{2(10^{R/20} - 1)} \right]. \quad (2)$$

Figure 5 is a plot of equation (2).

### 2.5 Frequency Markers

Both IF and RF markers are provided to indicate frequency on swept measurements. Three IF markers are provided; two use crystal-controlled tuned circuits and one uses a variable oscillator. The mark-

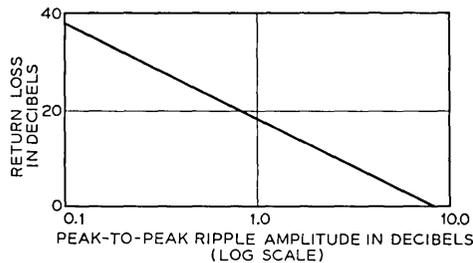


Fig. 5 — Conversion of ripple amplitude to return loss.

ers are driven by a sample of the IF swept signal and appear as "birdies" on the oscilloscope presentation. The variable marker can be shifted from 55 to 95 MHz; the frequency is read directly on the counter, which minimizes marker frequency error. The "birdie" width is such that the frequency of any point on an IF characteristic can be determined within 0.1 MHz.

Figure 6 is a schematic diagram of the IF marker circuit. The circuit consists of a 55 to 95 MHz oscillator associated with transistor Q1 and two tuned circuits consisting of a 64 MHz and a 76 MHz crystal. When the IF swept signal coincides with the frequency of the crystals or the oscillator, a "birdie" is produced at the collector

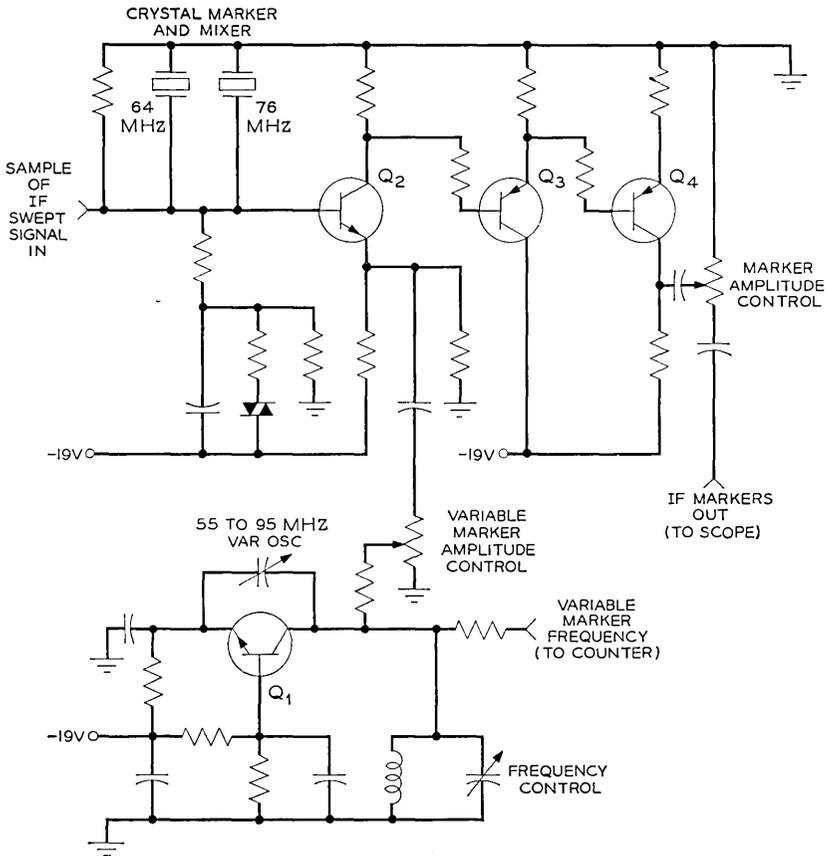


Fig. 6 — IF marker circuit.

of Q2. The "birdie" is amplified by transistors Q3 and Q4 and applied directly to the oscilloscope display. Amplitude controls are provided to adjust the amplitude of the markers.

The IF markers are used for IF to IF, IF to RF, and RF to IF measurements. When the IF oscillator is used in the measurement, the sample of the IF swept signal is obtained from the IF oscillator. When the IF oscillator is not used, the sample of the IF swept signal is obtained from the IF detector. In each case, the output of the marker circuit is added directly to the oscilloscope display and does not pass through the unit being tested.

The RF marker is provided by a resonant-cavity frequency meter which is lightly coupled to the output of the RF oscillator. The frequency meter has a diode rectifier which provides a dc pulse when the frequency of the RF oscillator is at the resonant frequency of the cavity. The pulse is added directly to the scope presentation in the same manner as the IF markers. The RF frequency can be measured to about  $\pm 0.6$  MHz accuracy.

The RF marker is used for RF to RF measurements and is available for RF to IF measurements. Its use for RF to IF measurements is usually limited to calibration because of the more accurate and more convenient IF markers available.

### 2.6 Power Measurements

The thermocouple type power meter measures IF and RF power from +10 to -30 dBm, to an accuracy of 0.3 dB or better at room temperature. It consists of an indicating unit and three separate power heads, each of which houses a thermocouple. The head used for IF measurements has an impedance of 75 ohms. The heads used for RF measurements each have an impedance of 50 ohms. The IF head and one of the RF heads are used for measuring power outside the test set and are accessible by front panel connectors. The other RF power head is connected to a monitor point on the RF oscillator and provides a continuous measurement of the RF output level. This allows the output of the oscillator to be changed a known amount without the use of a precision variable attenuator.

### 2.7 Frequency Measurement

The frequency counter is used to measure the frequencies of the crystal oscillators in the microwave generator, carrier resupply, and shift oscillator of the TD-3 radio equipment as well as the frequency

of the IF marker. The counter is a direct reading unit and measures frequencies as high as 135 MHz. Its stability is such that an accuracy of 5 parts in  $10^7 \pm 1$  digit is maintained for a year. Using a gating time of 0.1 seconds, a 125 MHz oscillator may be conveniently adjusted to a frequency within  $\pm 2$  parts in  $10^7$ .

A special technique is used to measure the frequency of the carrier resupply.<sup>5</sup> The resupply provides a 70 MHz carrier and a low level signal of either 61 or 63 MHz at its output. To measure the frequency of the 61 or 63 MHz signal, it is first separated from the 70 MHz carrier by a low pass filter. The amplitude of the signal is then increased by the test set IF amplifier to a power sufficient to drive the counter. However, the amplifier adds noise to the signal and this noise impairs the accuracy of the measurement. A band pass filter between the amplifier output and the counter reduces the effect of the noise.

### 2.8 Noise Figure

The AM noise figure of the TD-3 radio receiver is measured by comparing its noise with the noise of a gaseous discharge tube using the "Y factor" technique. The AM noise is measured rather than the more meaningful FM noise because of the simplicity and economy of the test apparatus. Despite its shortcomings, AM noise measurement is a useful maintenance tool.

The gaseous discharge tube is an argon lamp mounted in a coaxial structure. The noise output of the lamp is available on a type N connector located on the front panel of the test set. A noise measurement using the Y factor technique is made in the following manner. The output of the noise lamp is connected to the input of the receiver being tested. The output of the receiver is connected through the test set IF amplifier and the band pass filter to the power meter. Power meter readings are noted with the noise lamp ON and with it OFF. The Y factor (in dB) is the difference in the two power meter readings.

The noise figure is determined from the Y factor by the following relationship:<sup>6</sup>

$$F = 10 \log \left[ \frac{T_{\text{ex}}}{Y - 1} \right] \text{ dB} \quad (3)$$

where

$F$  = noise figure in dB

$T_{ex}$  = excess noise temperature of noise generator and connecting cables

$Y$  = measured  $Y$  factor ratio.

The excess noise temperature ratio of the test set generator and cables is  $13.9 \pm 0.5$  dB. Figure 7 is a plot of equation (3) with  $F$ ,  $Y$ , and  $T_{ex}$  given in dB. The dotted curves show the accuracy limitations caused by the variation of the excess noise temperature.

### 2.9 Control and Common Circuits

The control circuit merges the several units of the test set into a single functional unit. A single switch in the control circuit makes many of the interconnections required for the different measurements. The control circuit also provides the oscilloscope drive circuit, and generates the 31 Hz IF sweep drive signal.

The four-position switch makes the proper connections for IF to IF, IF to RF, RF to IF, and RF to RF measurements. It selects either the IF or RF detector, the IF or RF markers, and connects the appropriate sweep control voltage to the oscilloscope. The remainder of the connections are made manually on the front panel patch field.

The oscilloscope drive circuit alternately connects the detector output and a dc reference voltage to the vertical input of the oscilloscope. The reference line is applied to the oscilloscope during the retrace interval of the sweep signal. The circuit is synchronized to either the 31 Hz sinusoidal IF sweep signal or the 31 Hz triangular RF sweep signal. There are separate synchronizing controls. The reference line is connected to the scope during the negative going portion of the sweep signal and the test signal is connected during the positive going portion of the sweep signal. Coarse and fine controls position the traces on the oscilloscope.

The dc reference line serves two functions. First, it provides a reference line on the oscilloscope whose relationship to the test signal is independent of any oscilloscope adjustments. This permits use of the oscilloscope controls to set up calibration and position without upsetting the reference position. Second, use of the sweep signal retrace interval for applying the reference line allows the RF and IF oscillators to be on continuously rather than being blanked during the retrace. This permits power measurements to be made while the RF and IF oscillators are being swept, which could not be done without error if the oscillators were blanked during the retrace.

The sweep drive signal for the IF oscillator is provided by a modi-

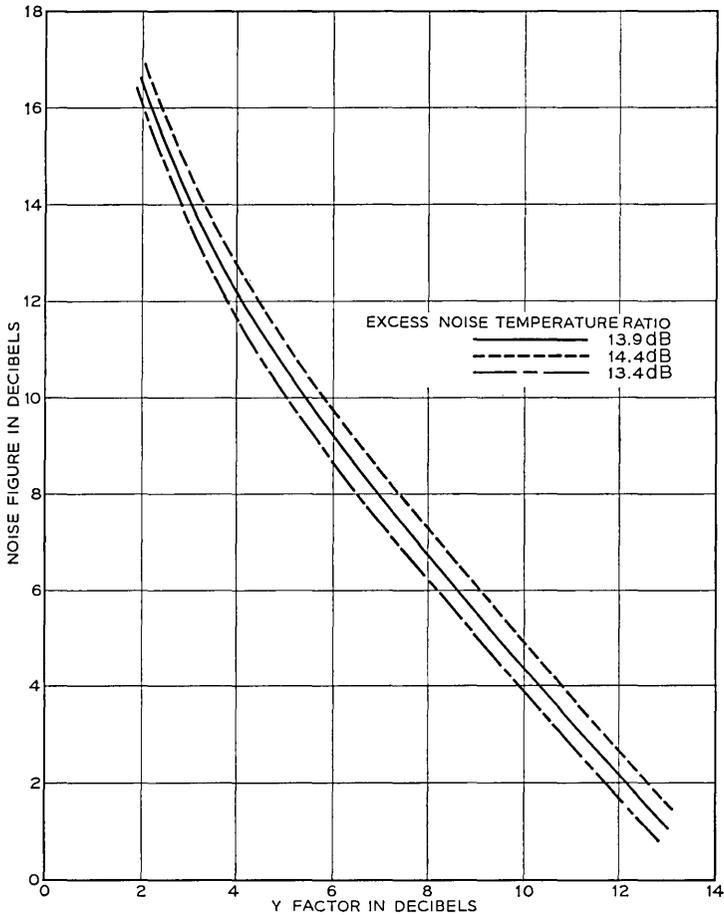


Fig. 7 — Conversion of Y factor to noise figure.

fied Wein Bridge oscillator as shown in Fig. 8. It consists of a 4-stage RC amplifier, operating as an oscillator, followed by two amplifier stages. The amplifier stages provide an output level of 26 volts peak-to-peak which is required to drive the IF oscillator to its maximum sweep width of 50 MHz. In order to achieve this voltage output, the amplifier stages are powered by a  $-40$ -volt power supply instead of the  $-19$ -volt supply.

#### 2.10 IF Detector

The IF detector consists of two IF amplifier stages followed by a diode detector as shown in Fig. 9. The use of the amplifier stages has

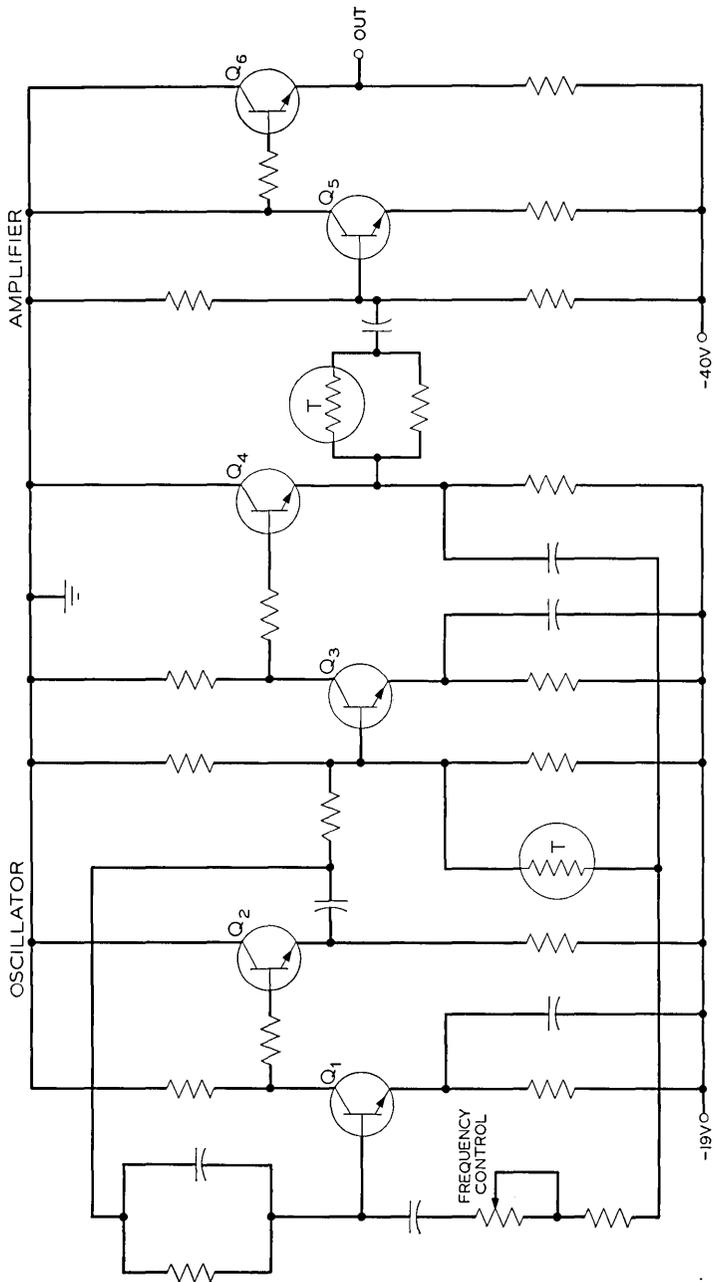


Fig. 8 — 31 Hz oscillator circuit.

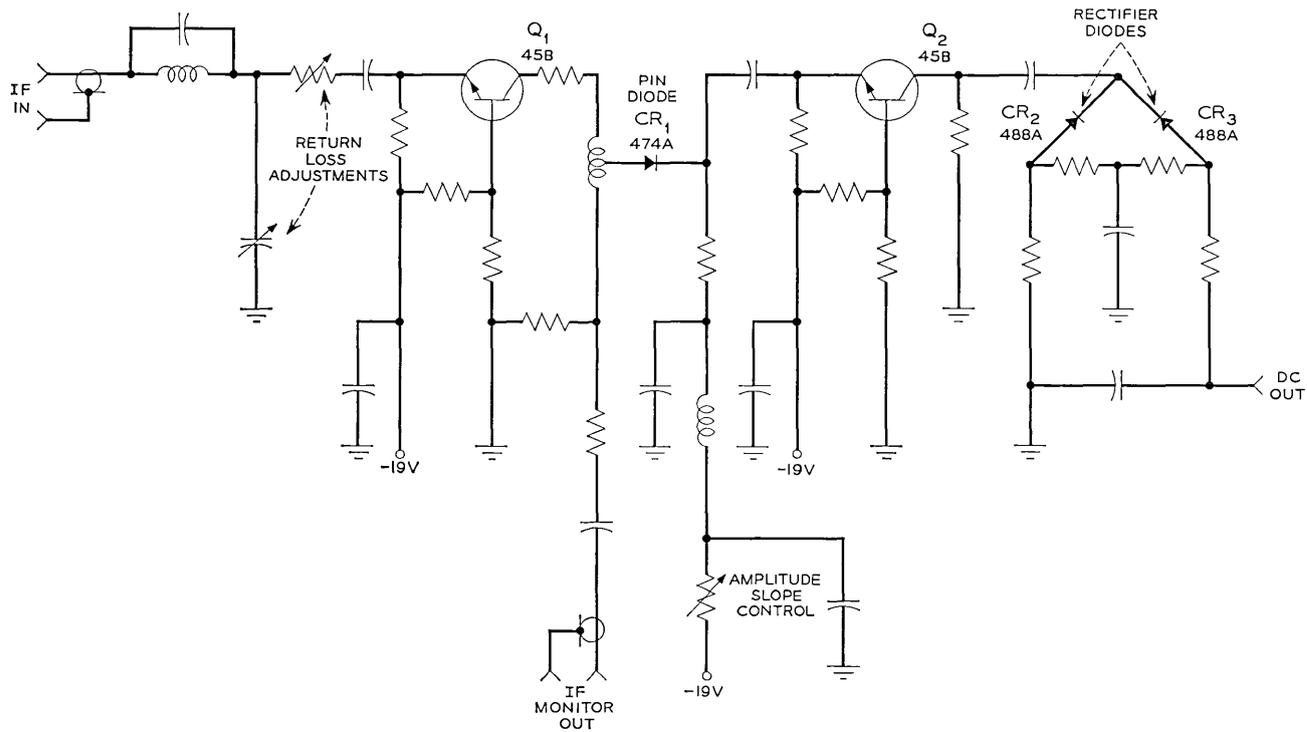


Fig. 9 — IF detector circuit.

great impact on the detector performance. First, the sensitivity is improved by the gain of the amplifier stages. Second, the input return loss is good since the input to the amplifier stages can be well matched to 75 ohms. Third, the amplifier provides an amplitude slope control which may be located outside the detector. Fourth, the nonlinear detector diodes are isolated from the connecting circuit by the amplifier stages. These improved detector characteristics allow the IF detector to achieve high performance and allow the connection to the detector to be made with a long IF cable.

Detectors used in earlier test sets had to be connected directly to the unit being tested. The return loss of the earlier detector sometimes was poor; in addition, the diode generated harmonics of the signal which appeared at the detector input. If a long cable were used to connect the detector to the unit being tested, interferences caused by reflected and phase shifted fundamental and harmonic signals caused a distorted signal envelope (for swept input signals) to appear at the detector output. The good return loss and harmonic isolation provided in the TD-3 test set detector virtually eliminate these problems and permit a cable to be used to connect the detector, mounted on the test set, to the unit being tested. With the detector mounted on the test set, ground loop disturbances, mentioned earlier, are less troublesome.

The IF power available in TD-3 for detection is  $-7$  dBm. This low power, in combination with the high sensitivity required, makes IF amplification in the detector necessary. The amplification consists of two transistor transformer-coupled stages similar to those used in the TD-3 IF main amplifier circuits.<sup>5</sup> A harmonic filter at the input reduces the detector response to harmonics present in the output of the unit undergoing the test. A signal approximately 30 dB lower than the input signal is provided by a monitor circuit to drive the IF marker circuit.

An amplitude slope control in the detector compensates for slope introduced by the connecting IF cabling and test set variations. The control is a screwdriver adjustment on the front panel. As shown in Fig. 9, the slope control circuit consists of a PIN diode between the two amplifier stages. The amplitude slope of the amplifier is controlled by the resistance of the diode which is varied by changing the direct current through the diode.

The detector uses two diodes in a full wave rectification circuit with dc loading to give an output proportional to the average value

of the rectified input signal. This configuration was chosen to provide high fundamental sensitivity and low harmonic sensitivity.

In contrast, the RF detector has higher power available for detection (0 dBm) and the RF harmonics are far removed from the TD-3 frequency range. RF power is detected with a high sensitivity diode detector preceded by a low pass filter. Both these units have coaxial structures with type N connectors and are obtained commercially.

### 2.11 *Test Set IF Amplifier*

The test set IF amplifier provides increased sensitivity or power level for IF return loss, noise figure, and carrier resupply frequency measurements. The amplifier is the IF main amplifier described in Ref. 5 with the AGC function of the main amplifier removed, and with additional gain. The gain of the test set IF amplifier is manually adjustable from 45 to 60 dB.

### 2.12 *Power Supplies*

Most units in the test set operate directly from commercial 117 volt ac power. Some use a power supply and a 19 volt regulator.<sup>4</sup>

The power supply provides outputs of +300, -150, -40, and -24 V dc, and 6.3 V ac. The -24 V dc output drives the -19 volt regulator. The IF oscillator uses the +300 and -150 V dc, and the 6.3 V ac. The control and common circuits use the +300 and -40 V dc in addition to the -19 V dc output of the regulator.

### 2.13 *Equipment Considerations*

Figure 1 identifies various parts of the transmitter-receiver test set. Power supply controls are available at the bottom rear of the console. The left side is a door, hinged full length at the rear, to allow access to a cable storage area. The console is slightly wider than similar consoles to concentrate useful functions at a comfortable working height. The base extends forward for greater mechanical stability.

The test bay and equipment undergoing tests are connected by 75-ohm coaxial cables for IF, and by 50-ohm coaxial cable for RF measurements. Special stainless steel type N RF connectors are used in critical locations to maintain proper impedance characteristics. The patch field provides a convenient means of establishing various test connections by use of standard patch plugs and cables.

Most front panel units and the power supplies are supported by

their front mounting screws and by horizontal aluminum angle rails running from front to rear. The storage drawer contains molded plastic inserts to hold small loose pieces of test equipment.

### III. FM TERMINAL TEST PANEL

The new 3A FM terminal test panel is used to measure and adjust:  
 FM transmitter deviation sensitivity

Deviator oscillator frequencies (186 and 256 MHz)

Baseband transmission (0.1 to 10 MHz)

Transmitter frequency and level alarms.

Such measurements as FM linearity and the 60 Hz square wave test of low frequency transmission response are made with equipment originally designed for the TD-2 radio system.

Since the 3A FM terminals have unprecedented stability and reliability, the FM test panel was designed only to perform routine adjustments and locate defective terminal plug-in units. When a defective unit has been located, it is simply replaced with a spare unit.

#### 3.1 General Equipment Features

The 3A FM terminal test panel, shown in Fig. 10, was designed to simply and easily make the precise measurements needed to insure that the terminals meet their stringent performance requirements. The panel has plug-in circuit packages for flexibility and to aid in trouble shooting. All of the circuits are transistorized and designed for maximum stability and long life. Two microammeters in a sepa-

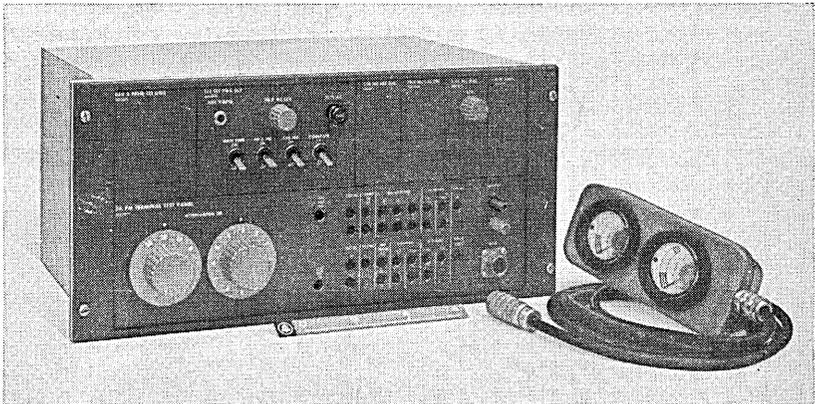


Fig. 10 — 3A FM terminal test panel.

rate case are connected to the panel by a detachable ten foot flexible cord. This separate case may be held in the hand or hung on the terminal bay for convenience when making adjustments.

Connections to the terminal equipment being tested are made with flexible cables to the test panel patch field, which is laid out to connect the various test panel units by standard coaxial patch plugs. Figure 11 shows the test panel mounted in the TD-2 FM terminal test set, where previously there had been a blank panel.

The test panel plug-in units are powered by a self-contained 20 volt regulated dc power supply which operates from a 117 V ac supply. A pair of connectors on the front of the test panel makes the 20 volt supply available for powering individual terminal plug-in units at a maintenance bench.

### 3.2 FM Transmitter Deviation Sensitivity

FM deviation sensitivity is defined as the ratio of the peak frequency deviation to the amplitude of the input baseband signal. Deviation sensitivity is controlled by adjusting the baseband gain of the FM transmitter. The method used to accurately adjust the transmitter deviation sensitivity is known as the Bessel null, or Crosby, technique.<sup>7</sup> This technique uses the principle that the carrier component of a sinusoidally modulated FM signal vanishes for certain values of modulation index. Consider the FM signal  $M(t)$  which arises when the carrier  $C(t) = A_c \cos \omega_c t$  is frequency modulated by the sinusoid  $V(t) = -A_v \sin \omega_v t$ :

$$\begin{aligned} M(t) &= A_c \cos \left[ \omega_c t + k \int V(t) dt \right] = A_c \cos \left[ \omega_c t + \frac{kA_v}{\omega_v} \cos \omega_v t \right] \\ &= A_c \cos [\omega_c t + X \cos \omega_v t]. \end{aligned} \quad (4)$$

Here, the index of modulation  $X$ , or peak phase deviation, is expressed as the ratio of the peak frequency deviation  $kA_v$ , divided by the modulating frequency  $\omega_v$ :

$$X = \frac{kA_v}{\omega_v} = \frac{\Delta\omega_c}{\omega_v} = \frac{\Delta f_c}{f_v}. \quad (5)$$

The FM signal  $M(t)$  may be expanded by use of the Bessel function identity to separate the various frequency components:

$$M(t) = A_c \sum_{n=-\infty}^{\infty} J_n(X) \cos \left( \omega_c t + n\omega_v t + \frac{n\pi}{2} \right). \quad (6)$$

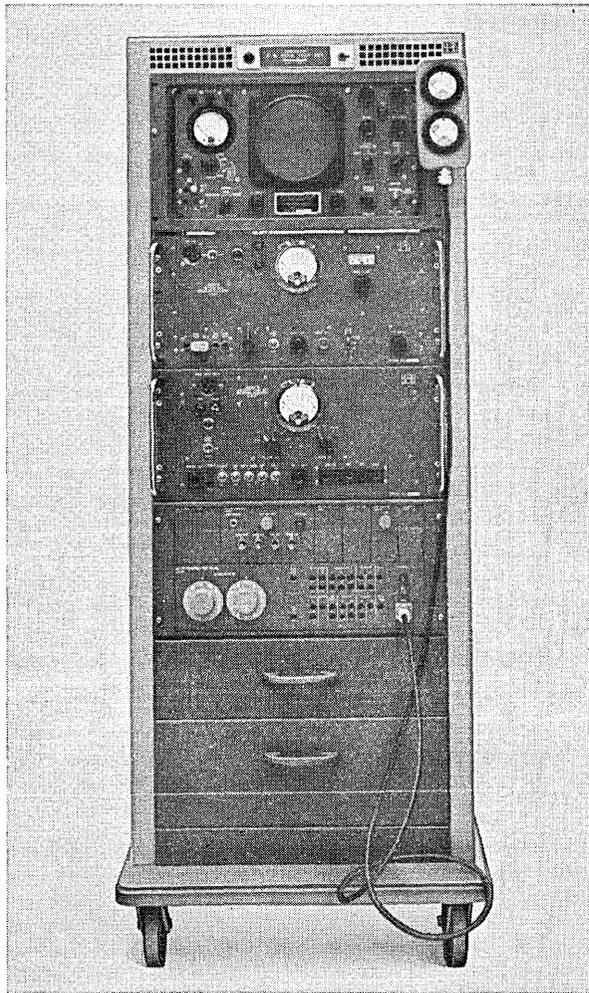


Fig. 11 — TD-2 FM terminal test set with 3A FM terminal test panel in place.

Here  $J_n(X)$  is the Bessel function of the first kind of the  $n$ th order and of argument  $X$ . It is seen from equation (6) that the magnitude of the carrier component of the FM signal is given by  $A_c J_0(X)$ . For certain values of the argument  $X$ ,  $J_0(X)$  goes to zero, the carrier component of the FM signal vanishes, and all of the signal energy appears in the sidebands. The lowest such value of modulation index for which  $J_0(X) = 0$  is  $X = 2.405$ .

For TD-3 use, it is desired to establish the transmitter deviation sensitivity so that a  $-12$  dBm sinusoidal baseband input signal will produce a peak deviation of 4 MHz. In actually adjusting deviation sensitivity, however, it is not important to use a  $-12$  dBm baseband signal, and in fact, the actual signal level chosen was  $-16$  dBm. The corresponding peak deviation is 4 MHz reduced by 4 dB, or 2.53 MHz. The corresponding baseband frequency which results in a carrier null may be calculated from equation (5):

$$f_s = \frac{\Delta f_c}{X} = \frac{2.53 \text{ MHz}}{2.405} = 1.05 \text{ MHz.} \quad (7)$$

Thus, to obtain the desired deviation sensitivity, a 1.05 MHz signal is applied to the baseband input at  $-16$  dBm, and the gain of the baseband amplifier is adjusted until a carrier null is observed. The technique for measuring the carrier null, shown in Figure 12, pro-

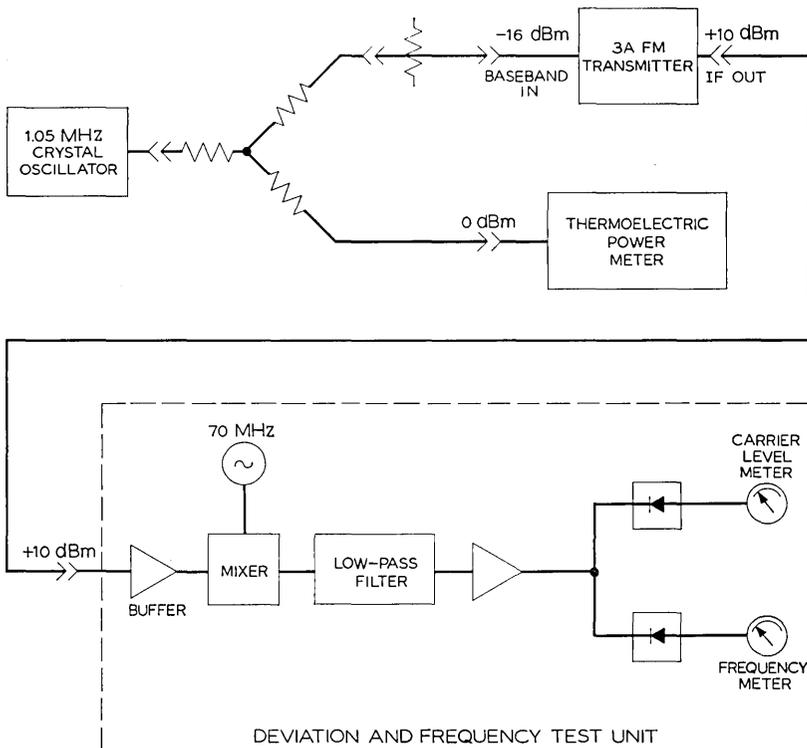


Fig. 12 — Measurement of 3A FM transmitter deviation sensitivity.

vides sufficient sensitivity and frequency separation to detect a carrier null of greater than 60 dB.

The accuracy to which the deviation sensitivity can be adjusted is limited by the accuracy to which the frequency and power of the baseband input signal are known. Therefore, a 1.05 MHz crystal-controlled oscillator is provided in the test panel, and the oscillator output is monitored by means of an external thermocouple power meter. The two stage oscillator provides an output power variable from +2 to +10 dBm into a balanced 124 ohm load, with a frequency accuracy of  $\pm .005$  percent.

The test panel deviation and frequency test unit provides a sensitive indication of the carrier null. To accomplish this, it must first separate the carrier from its sidebands, which are spaced at multiples of 1.05 MHz from the 70 MHz carrier. To achieve this separation without using a costly narrow bandpass filter at the carrier frequency, the input FM signal is heterodyned against a 70 MHz crystal-controlled local oscillator. Following this frequency down-conversion, the sideband energy at 1.05 MHz and above is filtered, leaving only the carrier beat component which is amplified and detected. The detected level of the carrier beat component is displayed on the carrier level microammeter.

To establish the proper transmitter deviation sensitivity, the transmitter baseband gain is adjusted until a dip is observed on the carrier level meter, indicating a carrier null. The gain range of the transmitter baseband amplifier is limited to prevent any carrier null except that corresponding to the first zero of the Bessel function  $J_0(X)$ . Use of the Bessel null technique results in an accuracy of better than  $\pm 0.05$  dB in the adjustment of the transmitter baseband gain or deviation sensitivity. Figure 13 illustrates the accuracy to which the transmitter deviation sensitivity is adjusted as a function of the null achieved on the test panel carrier level meter when using the Bessel null technique.

### 3.3 *FM Transmitter Rest Frequencies*

An additional feature of the deviation and frequency test unit is that it may be used to accurately adjust the frequency of an unmodulated input signal to 70 MHz. This is accomplished by a frequency counting detector which provides a dc voltage proportional to the frequency of the carrier beat signal. This voltage is displayed on the hand-held frequency microammeter and is proportional to the frequency difference between the input signal and the local oscillator.

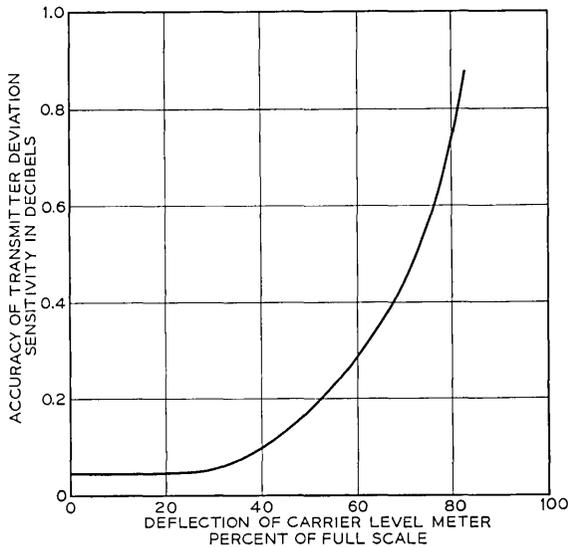


Fig. 13—Accuracy achieved when using Bessel null technique to adjust FM deviation sensitivity. (The meter is fully deflected when the transmitter being tested is undeviated.)

Thus, if the frequency of the input signal is adjusted until a null shows on the frequency microammeter, the input signal will be identical in frequency to the 70 MHz local oscillator.

In addition to establishing an accurate 70 MHz transmitter carrier frequency, it is necessary that the rest frequencies of the 256 and 186 MHz voltage-controlled oscillators in the transmitter deviator be accurately established. This is important because the rest frequencies have been carefully selected to minimize the interference created by high order modulation products which fall in the baseband spectrum. Therefore, there is a 256 MHz crystal controlled oscillator in the test panel as a stable reference, composed of oscillator, buffer, frequency doubling, and output amplifier stages which will deliver +13 dBm into a 50 ohm load with a frequency accuracy of  $\pm 0.006$  percent.

Figure 14 shows the procedure for adjusting the rest frequencies of the varactor diode controlled oscillators in the FM deviator. The reference signal is injected into the 256 MHz deviator oscillator and the varactor bias is adjusted until the 256 MHz oscillator becomes phase locked to the reference oscillator, as indicated by the detected output voltage of the deviator oscillator. When the 256 MHz oscillator is

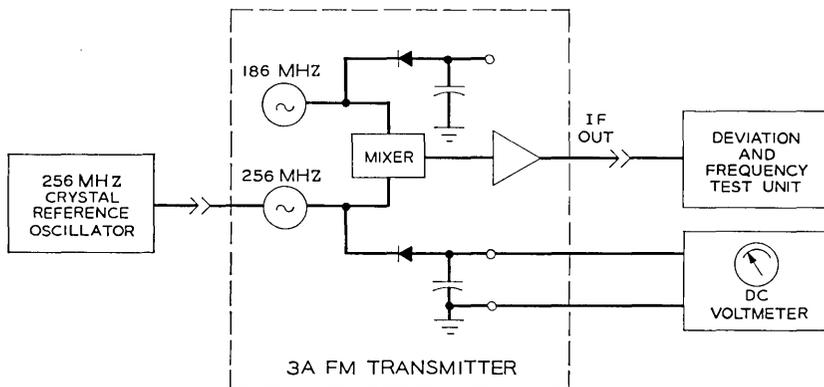


Fig. 14—Adjustment of varactor oscillator rest frequencies in the 3A FM transmitter.

locked, the 186 MHz deviator oscillator is brought to its proper rest frequency by adjusting for a 70 MHz carrier frequency as measured by a null on the frequency meter of the deviation and frequency test unit. The 256 MHz reference oscillator is then removed, and the 256 MHz deviator oscillator is adjusted for a 70 MHz carrier frequency. With this method, the oscillator rest frequencies can easily be set to within  $\pm 5$  kHz of their nominal values.

The frequency microammeter has a high sensitivity range with a full scale deflection of 20 kHz and a low sensitivity range with a full scale deflection of 100 kHz. The desired range is selected with a push-button on the hand-held case. Figure 15 illustrates the accuracy of the transmitter 70 MHz IF output frequency as a function of the deflection of the frequency meter.

### 3.4 Baseband Transmission

The baseband frequency response of an FM terminal pair must be maintained flat within  $\pm 0.1$  dB from 6 Hz to 10 MHz. It is interesting that this allows less transmission shaping than the frequency roll-off of only twelve feet of cable. To achieve such accuracy, a comparison test circuit with an external signal generator and a thermocouple power meter was chosen. This test circuit can be used to measure the baseband transmission of either an FM terminal pair or a transmitter baseband amplifier plug-in unit at a maintenance bench. A typical arrangement of this test circuit, illustrating an FM transmitter baseband amplifier under test, is shown in Fig. 16.

Since the balanced to unbalanced transformer and the test set baseband amplifier are employed in the common transmission path, their frequency response is not critical with respect to the over-all transmission accuracy. The test set baseband amplifier is a three stage, direct-coupled feedback amplifier which provides approximately 12 dB of gain between 100 kHz and 10 MHz. Cables for patching between the terminals and the test panel are provided with the test panel, primarily to increase the accuracy of transmission measurements by controlling the cable length. A length of cable is placed in the reference arm of the comparison test circuit inside the test set in order to maintain equal lengths in the reference and test paths.

### 3.5 Frequency and Level Alarms

The 3A FM transmitter activates station alarms whenever the carrier frequency drifts by more than 200 kHz from its nominal center frequency of 70 MHz, or whenever the transmitter IF output power changes by more than 3 dB from its nominal value of +10 dBm. A 70.2 MHz crystal-controlled oscillator in the test panel is a standard for adjusting the sensitivity of the transmitter control circuits. This three stage circuit provides a variable output power of +7 to +13 dBm into a 75 ohm load, with a frequency accuracy of  $\pm 0.006$  percent.

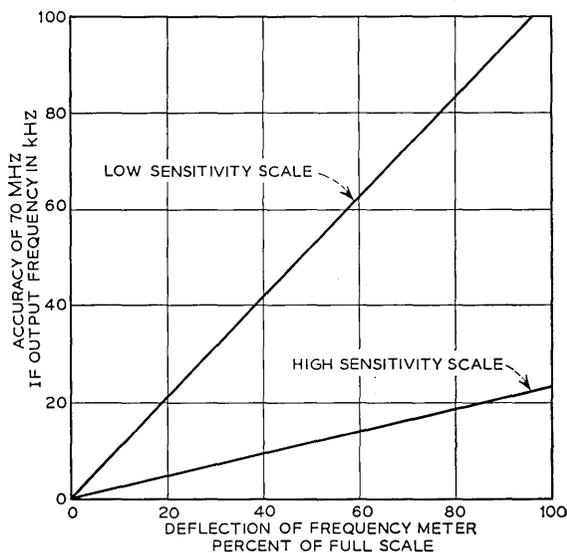


Fig. 15 — Accuracy achieved in adjusting FM transmitter output frequency.

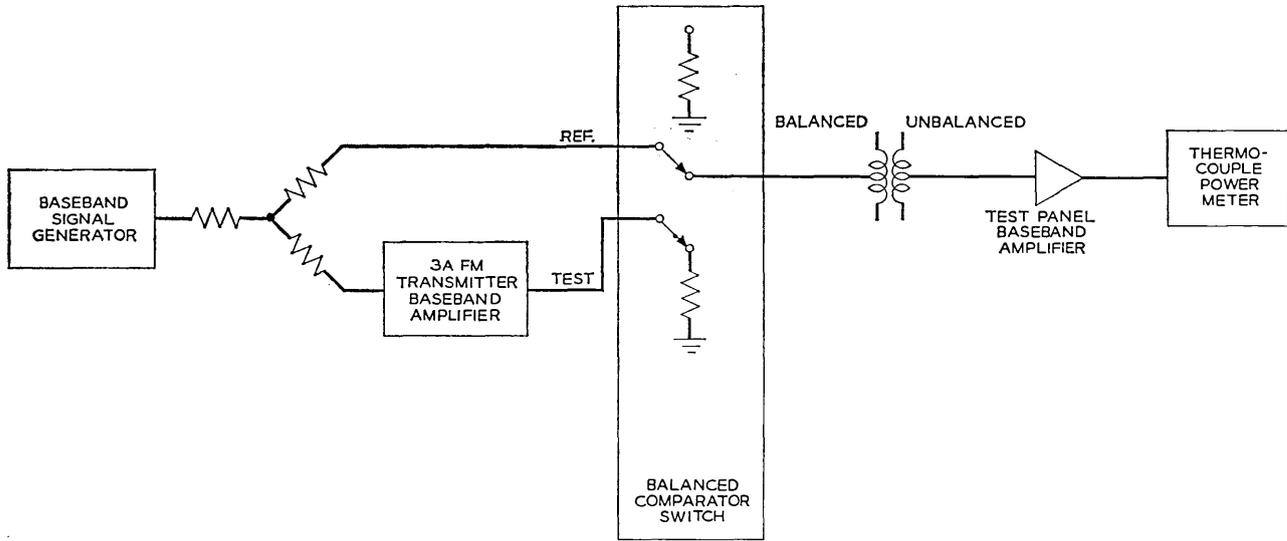


Fig. 16— Measurement of 3A FM transmitter baseband amplifier transmission.

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# Power System

By W. E. JEWETT and S. MOTTEL

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*This paper describes the power required by the TD-3 radio relay system. It discusses the battery plant which serves as the basic source of power, the traveling-wave tube power supply, the regulated -19-volt power supply, the transmitter-receiver test set power supply, the dc distribution arrangement, and the radio grounding circuit.*

## I. INTRODUCTION

The transmission equipment in the TD-3 radio system is designed to be powered from common, -24-volt battery plants. Where voltages other than -24 volts are required, for the traveling-wave tube, the IF circuits, the microwave generator, and the like, they are furnished by power supplies which are powered from the common -24-volt battery plant. An exception is the transmitter-receiver test set power supply which is powered from commercial 60 hertz ac.

The stringent noise requirements of the TD-3 radio system dictated a precisely-controlled approach to dc power distribution. DC power feeders from the common -24-volt battery plant through a battery distribution circuit breaker board are segregated into noisy, quiet, and undesignated groups. The noisy group, those feeders on which there is appreciable noise, feed primarily converter-type power supplies. The quiet group consists of feeders whose loads are susceptible to noise, such as the regulated -19-volt power supply. The undesignated group primarily feeds alarm battery supply loads which generally are not noisy, but will on occasion, such as when a relay drops out, produce noise.

With rare exceptions, a separate ground return feeder of equal cross sectional area is run with each hot (off ground) feeder from the circuit breaker board to the equipment bays. This minimizes current flowing in the equipment bay frameworks and thus reduces system noise.

In the TD-3 main and repeater stations of the initial installation,

the tower is an integral part of the building. Therefore, it was essential that the radio grounding circuit be designed with extreme care to protect persons and equipment from lightning. In addition to the usual internal and external ring grounds, another ring ground is embedded in the roof of the building. The four tower legs are connected to the internal and external rings as well as to the roof ring. Inside the building, all equipment bays are directly connected to the internal ring ground; and when mechanical bonding of adjacent bays does not result in a good electrical connection, these bays are interconnected electrically.

## II. BATTERY PLANT

From a power point of view, one of the objectives in the design of the TD-3 radio system was that it be compatible with existing, common, -24-volt battery plants. This includes 11-cell plants with emergency cells, and 12-cell plants with or without emergency cells. This design objective was met. However, a 12-cell plant with emergency cells is preferred because there is greater reserve time during a commercial power failure, and less dc distribution copper is needed because of the increase in the allowable loop feeder drop from the battery to the equipment bays with such a plant.

The float voltage of 11-cell plants is 23.9 volts  $\pm 1$  percent, while that of 12-cell plants with or without emergency cells is 26.0 volts  $\pm 1$  percent. The TD-3 radio system will perform satisfactorily with battery plant voltage from 22 to 28 volts.

## III. POWER SUPPLIES

### 3.1 *Traveling-Wave Tube*

The traveling-wave tube power supply is a solid state, dc to dc converter which furnishes power to all of the TWT electrodes. The electrical requirements for the supply are shown in Table I.

A dc to dc converter was obviously required to convert the battery voltage to the relatively high dc voltages required by the TWT electrodes. The converter was designed to have an oscillating frequency of 2 kHz. This frequency was selected because it offered the best compromise between power supply size and cost. At 2 kHz, reasonably small transformers and filters are required, yet relatively slow switching, inexpensive transistors can be tolerated.

Figure 1 is a simplified block diagram of the power supply. Notice

TABLE I—TWT SUPPLY REQUIREMENTS

Electrode	Voltage	Voltage stability	Current	Ripple (rms)
Anode	Adjustable, +60 to +500 V with respect to the helix voltage	$\pm 60$ V with respect to cathode	0-1 mA	0.3 V
Helix	Adjustable, +2500 to +2900 V with respect to the cathode voltage	$\pm 12$ V with respect to cathode	0-4 mA	0.14 V
Cathode	Fixed, -1420 V with respect to the collector voltage	1185 to 1615 V with respect to collector	35-40 mA	1.0 V
Collector	Connected to ground	—	—	—
Heater*	7.5 V below the cathode	$\pm 0.15$ V (70°-80°F) $\pm 0.375$ V (40°-140°F)	0.8-0.95 A	0.075 V

\* Heater voltage is 9.1 V upon initial turn-on of the power supply and automatically drops to 7.5 V after an adjustable 120 to 300 seconds.

that complete input-to-output dc isolation is obtained from the transformers in the supply. The description which follows relies on Fig. 1. Input power from the -24-volt battery plant, by way of an input filter, flows into the 2 kHz inverter which serves as the source of power for all of the rectifiers in the supply. Power for the cathode is derived from the 2 kHz inverter through the cathode transformer and the cathode rectifier and filter; one side of the cathode rectifier and filter output is connected to the collector which is grounded. The other side is connected to the cathode. Anode and helix power is furnished from their respective rectifiers and filters, by way of their transformers, through a common ac series regulator. Power for the heater flows through an electronic time delay relay to the heater transformer and then to the heater rectifier and filter.

The helix potential and, to a lesser extent, the anode potential, both relative to the cathode, are regulated by sampling the helix-to-cathode voltage with a voltage divider and comparing this sampled voltage to a reference voltage. The difference or error voltage is amplified, chopped at a 2 kHz rate, passed through a dc isolating transformer, and used to adjust the bias of the ac series regulator so as to vary the voltages applied to the primary windings of the helix and anode transformers

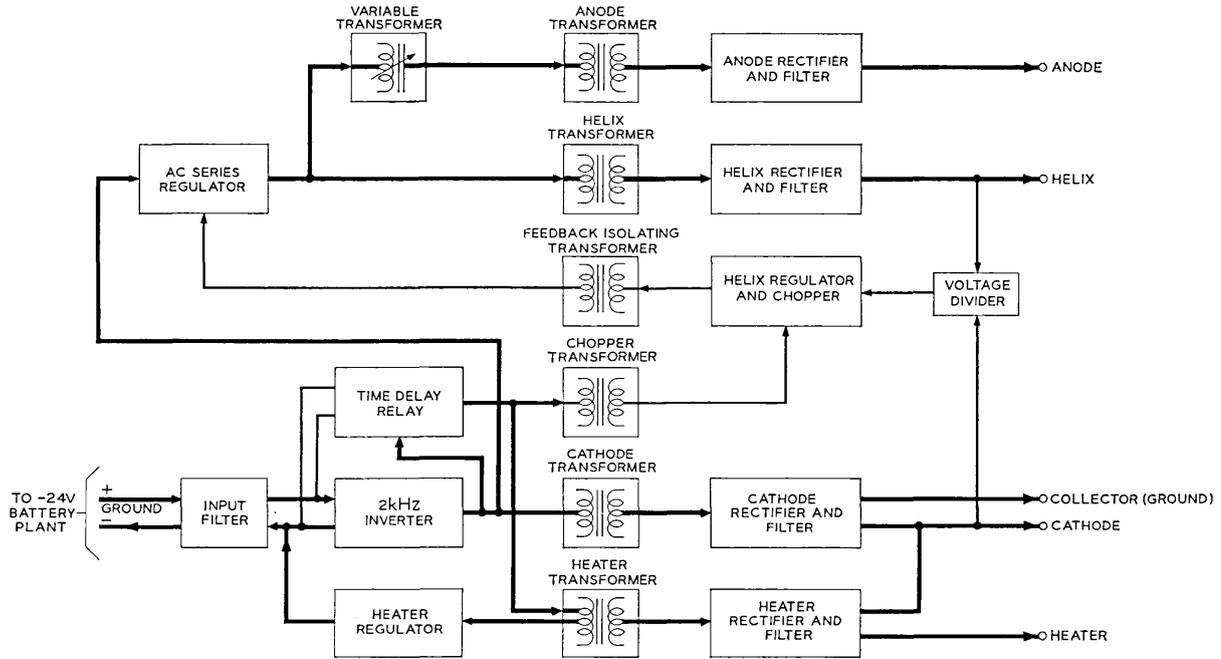


Fig. 1 — Simplified block diagram of TWT supply.

according to the dictates of the error signal. Because the helix-to-cathode voltage is sampled, it is regulated to better than  $\pm 12$  volts for any helix voltage between 2500 and 2900 volts. The anode voltage, because of the open-loop output impedance of the helix circuit and the output impedance of the anode circuit, will be regulated to a lesser degree,  $\pm 60$  volts for any anode output voltage between 2560 and 3400 volts. The heater voltage is regulated by the heater regulator which maintains the primary voltage to the heater transformer to better than  $\pm 2$  percent during normal operation.

Figure 2 illustrates, in simplified form, the high voltage circuits, excluding the heater, and the ac series regulator. Variable transformer  $T_1$  permits the anode voltage to be level-set from 60 to 500 volts above the helix. The regulated rectifier consisting of diodes  $CR_1$ ,  $CR_2$ ,  $CR_3$ ,  $CR_4$ , and resistor  $R_1$  provides a 44-volt bias which insures that the anode will always be positive with respect to the helix.

The helix-to-cathode voltage divider consisting of resistors  $R_2$ ,  $R_3$ , and  $R_4$  is designed so that the input voltage level to the error detector circuit is near ground potential. The output voltage of the feedback isolation transformer is rectified by diodes  $CR_5$  and  $CR_6$  and controls the collector-emitter voltage drop of the transistor in the ac series regulator. Because this transistor is in series with the parallel combination of variable transformer  $T_1$  and transformer  $T_3$ , it controls the amplitude of the square-wave voltage applied to these transformers according to the dictates of the helix voltage regulator feedback loop.

Figure 3 is a simplified schematic diagram of the heater regulator circuit. Heater regulation is accomplished by maintaining voltage  $V$  of Fig. 3 constant. This voltage plus the drop across either diode  $CR_3$  or  $CR_4$ , (which conduct on alternate half-cycles), is the voltage applied to the primary of heater transformer  $T_2$ . The power supply has an adjustable electronic time delay relay which permits the heater voltage at the tube to be high initially for a period which is adjustable from 120 to 300 seconds. When the TWT supply is shipped from the factory, the time delay is set to 180 seconds. During this initial time the voltage at the heater is  $-9.1$  volts. After the relay has timed out, the heater voltage is reduced to  $-7.5$  volts.

When the polarity of the primary winding on cathode transformer  $T_1$  is positive at terminal 1 and negative at terminal 2, diode  $CR_2$  will conduct and diode  $CR_1$  will block. For this condition, current will flow from terminal 1 of  $T_1$  through  $CR_2$ , through the time delay

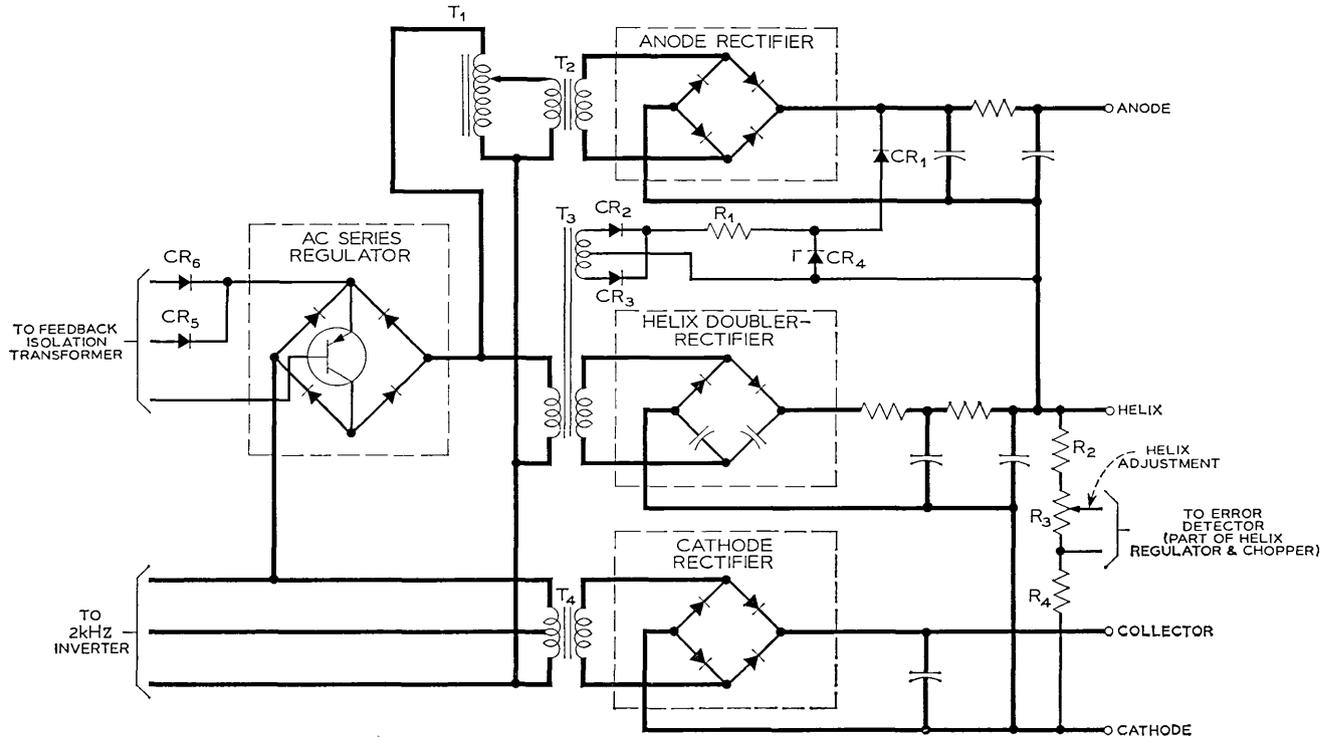


Fig. 2—Simplified high voltage circuits, excluding the heater, and the helix ac series regulator arrangement.

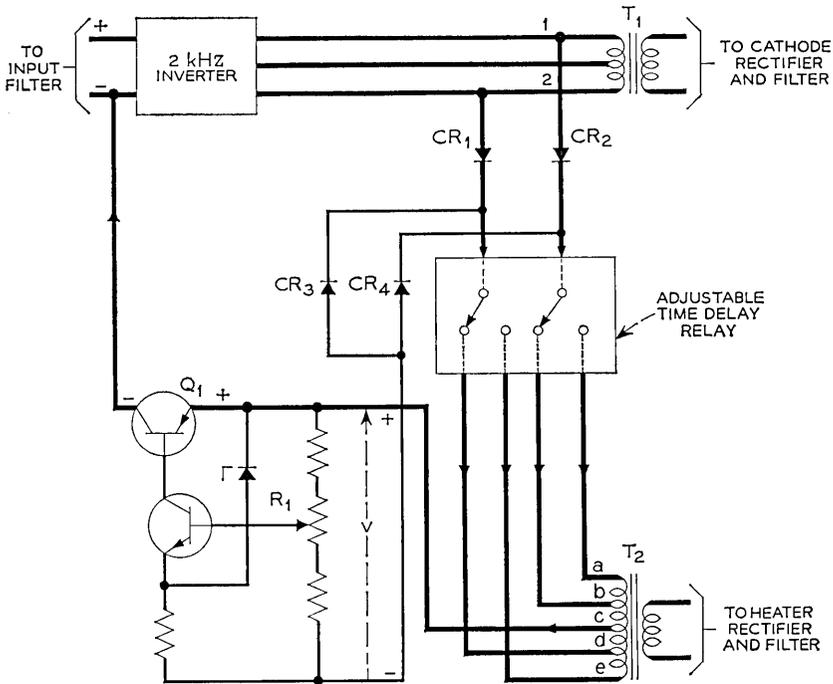


Fig. 3 — Simplified heater regulator circuit.

relay, and into terminal *b* of heater transformer  $T_2$ . This current leaves the primary of  $T_2$  by way of center tap terminal *c*, through heater regulator series transistor  $Q_1$ , and back to the negative side of the input filter. On the alternate half-cycle, primary current will flow through  $CR_1$ , through the time delay and into terminal *d* of  $T_2$ . This current also leaves by way of terminal *c* of  $T_2$ . During the initial overvoltage period, the time delay relay connects terminals *b* and *d* of  $T_2$  to the cathodes of  $CR_2$  and  $CR_1$ , respectively. After the overvoltage period has passed, terminals *a* and *e* are connected to the cathodes of  $CR_2$  and  $CR_1$ . The heater is regulated by the automatic variation of the dc voltage across transistor  $Q_1$  according to the dictates of voltage  $V$ . Potentiometer  $R_1$  level-sets the heater voltage.

The TWT power supply will be made in considerable quantity; its mechanical design, testing, and assembly, and their associated costs, are of considerable importance. In addition, it should have

a pleasing appearance. Figure 4 shows the general mechanical features of the power supply, which consists of two units in a metal case. The inverter and heater regulator unit (left) and the rectifiers and helix regulator unit are both inserted into an over-all metal housing. The two are joined through connectors at the rear of the housing and at the rear of the plug-in units. A plug-equipped cord from the rectifiers and helix regulator unit connects to the traveling wave tube which is mounted above the power supply. The same cord connects to the test load, described later in this section.

Because of the electrical potentials in this unit, personnel safety had to be considered. The entire power supply is mounted in the metal housing in such a way that the inverter and heater regulator unit on the left must be removed before the high voltage right-hand unit. The right unit actually extends behind the low voltage left-hand unit, and an attempt to remove the high voltage unit first will automatically eject the low voltage unit. Removing the left-hand unit disables the circuits to the high voltage unit so that high voltages are no longer present. There is further electrical interlocking to the traveling-wave tube by the internal wiring through the associated connectors of the power supply and the traveling-wave tube.

Figures 5 and 6 show a number of design details. The low voltage

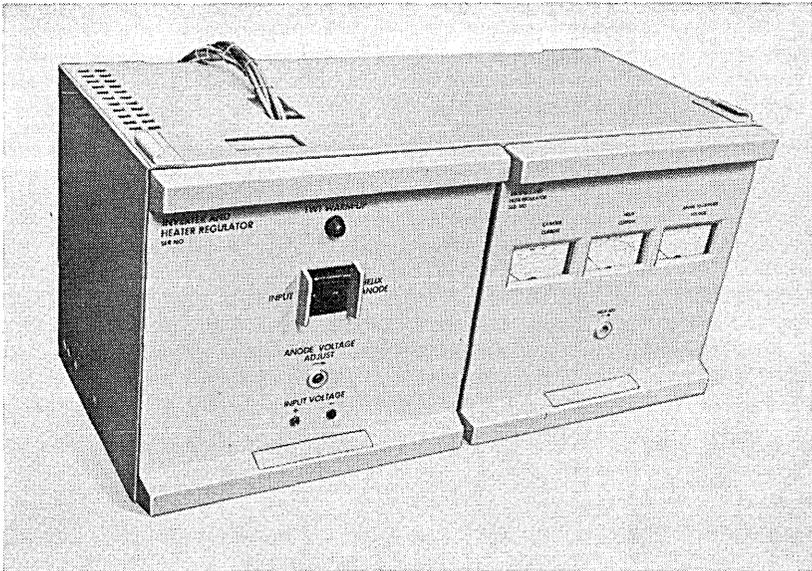


Fig. 4 — TWT power supply.

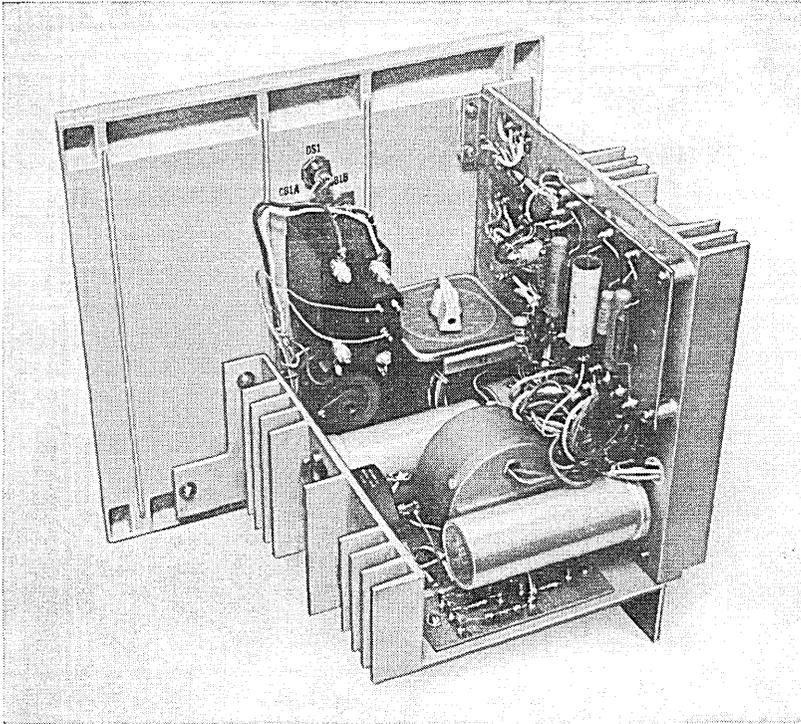


Fig. 5 — Inverter and heater regulator unit of the TWT power supply.

unit in Fig. 5 uses three castings for mounting components, transistors, diodes, and connectors. This permitted the use of an integral heat sink and a mechanical structure to dissipate the heat generated within the low voltage package. The unit may be opened for assembly and repair by dropping the hinged front panel and dropping the vertical casting on the left (viewed from the front). The castings on the left unit include fins for dissipating the heat from the transistors and diodes mounted on it with insulating washers.

The rectifier and regulator unit shown in Fig. 6 has two die castings; the panel casting with beveled edges for the meter openings and a chassis casting which holds the transformers, capacitors and associated high voltage components (mounted on insulating panels because of the high voltages). A printed wiring board in the rectifier regulator unit holds the helix regulator circuit which operates at high voltages. When the hinged front panel and a hinged plastic panel

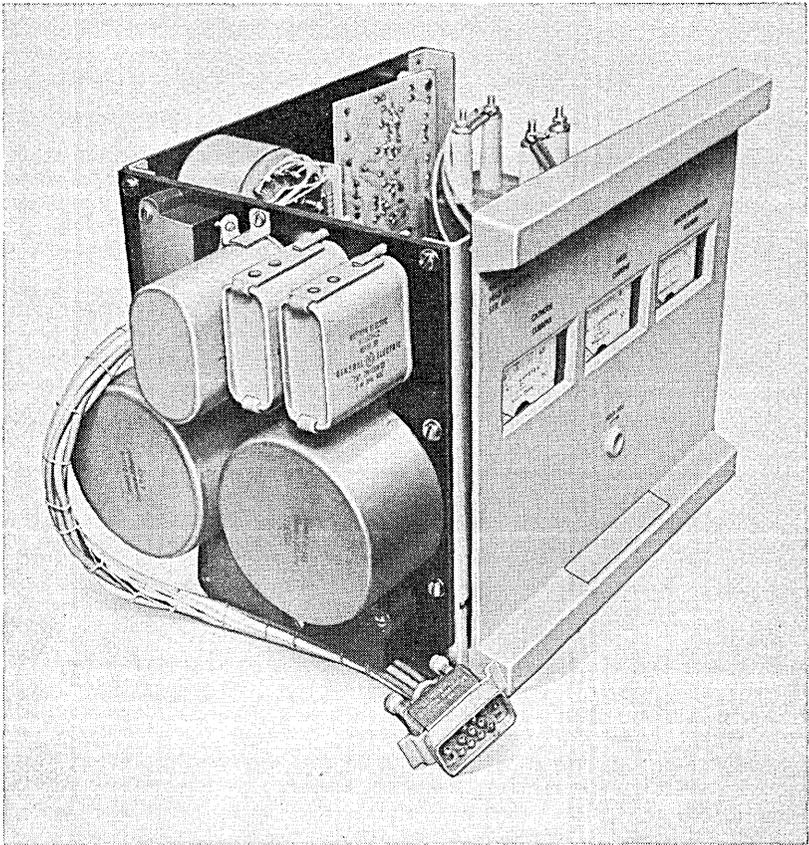


Fig. 6 — Rectifiers and helix regulator section of the TWT power supply.

at the rear are opened the apparatus is accessible for assembly, repair, and troubleshooting. The panels can be dropped only when the power unit has been removed from its housing. Insulating sheets and high voltage wire have been used extensively because of the high potentials.

The coils in the power unit are given standard dielectric and corona tests. The high voltage coils are epoxy encapsulated with metallic shields to reduce noise which could couple into other circuits in the repeater bay and finally appear in the baseband of the TD-3 system.

Figure 7 shows two molded high voltage diode assemblies developed for the TD-3 power supply.<sup>1</sup> With these molded assemblies, the diodes occupy less space compared with conventional individual mounting, and there are fewer exposed solder connections (with their

corona generating possibilities). These packages permit the use of the enclosed diodes to the full limit of their voltage rating without derating for altitude, as has been necessary for discrete diodes because of the very small tubulation to body spacing.

If a faulty TWT power supply is connected to a good TWT, it is possible that the power supply will damage the tube. For this reason, and for one other, a test load for the TWT power supply was designed and is provided at each radio station. When the TWT power supply is first installed on a transmitter-receiver bay it is checked with the test load connected to its output terminals. Once it has been determined that the TWT power supply is functioning properly, the test load is removed and the TWT connected. The second reason for a test load is to help maintenance people determine, in the event of a TWT amplifier failure (TWT and its power supply combined,) whether the TWT or the power supply is faulty.

The test load is shown in Fig. 8. It consists of resistive loads which simulate several quiescent states of TWT electrode voltages and currents, a multipurpose voltmeter, a cathode current milliammeter, a helix current milliammeter, and a heater voltmeter. The multipurpose voltmeter can be switched to measure the anode-to-cathode, helix-to-cathode, and collector-to-cathode voltages. The test load attaches to the same plug-in connector as the TWT.

When used to check the TWT power supply, the test load is mounted on top of the power supply in front of the TWT, as Fig. 9 shows. Connections pass through a cutout in the top of power supply case and the bottom of the test load. The test load has a handle at the

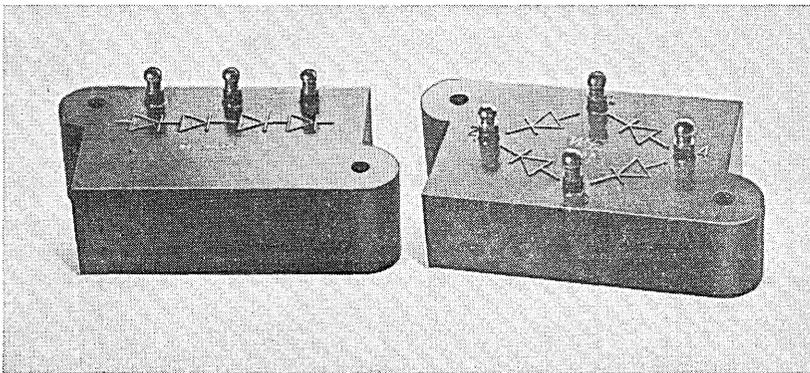


Fig. 7—Molded high voltage diode assemblies used in the TWT power supply.

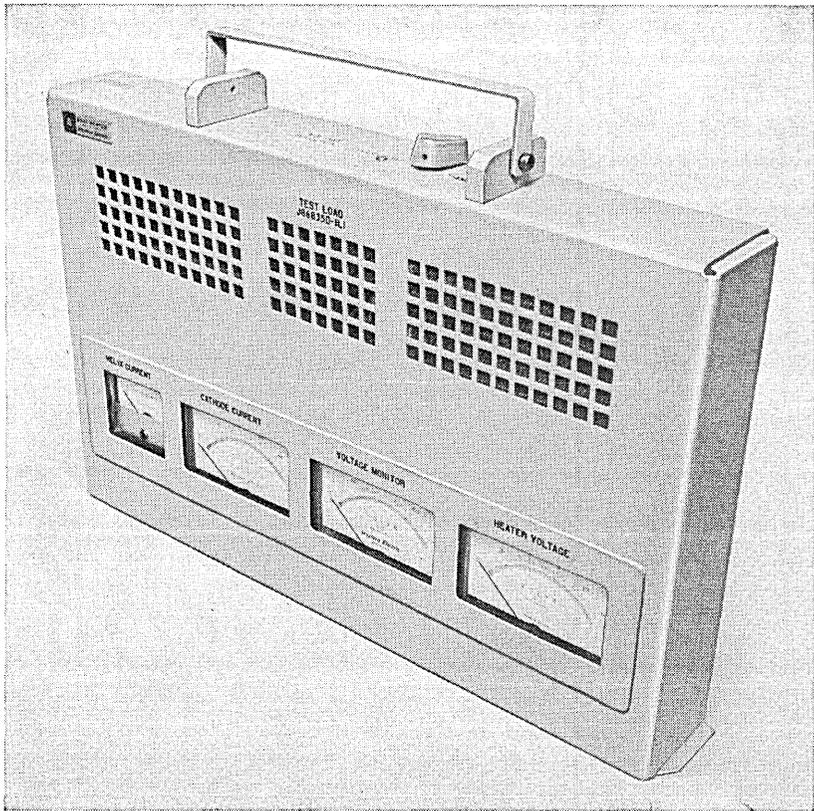


Fig. 8 — Test load for the TWT power supply.

top for easy carrying. When not in use it will normally be stored in a spare parts cabinet. One test load per station is sufficient.

### 3.2 Regulated $-19$ -Volt Supply

This supply which powers the IF circuits, microwave generator, and so on, in the transmitter-receiver bay is a solid state, active, series line regulator which drops the battery plant voltage to  $-19 \pm 0.2$  volts. Two of these supplies are used in each bay in a switching main station and one is used in each bay in a repeater station. One is also used in the transmitter-receiver test set. The requirements for this power supply are:

Input range	21–27 V dc
Output	$-19$ V dc

Output stability ( $75 \pm 5^\circ\text{F}$ )	$\pm 0.2 \text{ V}$
Output stability ( $40\text{--}140^\circ\text{F}$ )	$\pm 0.4 \text{ V}$
Output current	0–4 A
Output ripple voltage	1 mv rms
Output high voltage alarm at	–20 V
Output low voltage alarm at	–18 V.

Fig. 10 depicts a simplified circuit of the regulated, –19-volt power supply. Output voltage is regulated by varying the emitter-collector dc voltage drop of transistor  $Q_1$ , which is in series with the

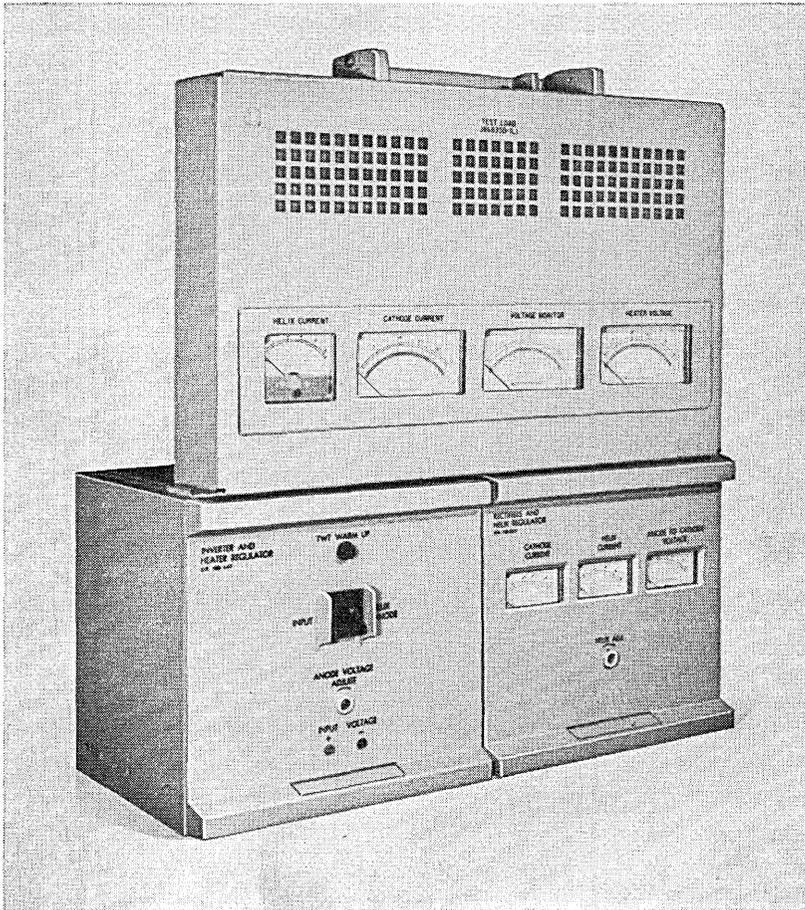


Fig. 9 — Test load atop TWT power supply.

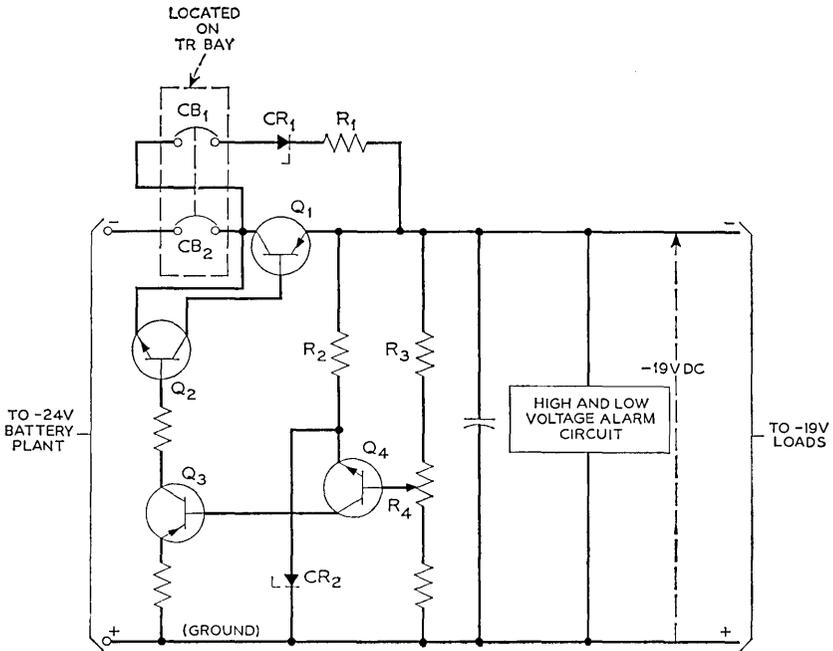


Fig. 10 — Simplified  $-19$ -volt regulator power supply circuit.

battery and the load, according to the dictates of the feedback loop.

Should the output voltage of the regulator increase by an amount  $\Delta v$ , (this could result from a decrease in load current), the voltage across resistor  $R_2$  would increase by  $\Delta v$  because  $CR_2$  is a voltage reference diode operating in the breakdown region. The increase in voltage across the series combination of  $R_3$  and the upper part of  $R_4$  would be equal to  $k\Delta v$  where  $k$  is a real number less than unity. The base to emitter voltage of transistor  $Q_4$  would decrease by an amount equal to  $(1-k)\Delta v$ . The net result would be a decrease in the base current of  $Q_4$  and thus a decrease in the collector current of this transistor. Because the base current of  $Q_3$  is the same current flowing in the collector of  $Q_4$ , it is also reduced, and so is the collector current of  $Q_3$ . This chain of events will result in a decrease in both the collector current of  $Q_2$  and the base current of  $Q_1$ . The decrease in the base current of  $Q_1$  will result in an increase in the emitter-collector dc voltage drop of this transistor and this, in turn, will return the output voltage of the regulator to  $-19$  volts.

Similar reasoning can be applied if the output voltage of the regulator decreases by  $\Delta v$ . The result will, of course, be a decrease in the emitter-collector dc voltage drop of transistor  $Q_1$ . The  $-19$ -volt output is level-set with potentiometer,  $R_4$ .

Circuit breaker  $CB_2$ , rated at 5 amperes, is the overload protection device for the power supply when the battery voltage is in the normal range of 24 to 26 volts.  $CB_1$ , a companion circuit breaker to  $CB_2$ , is rated at 2 amperes and serves as the overload protection device when the battery voltage is low. The series combination of  $CR_1$  and  $R_1$  provides a starting current path for the power supply during turn-on.

The regulated  $-19$ -volt power supply is shown in Fig. 11. It is a plug-in unit which consists of a combined one-piece, die-cast frame, heat sink, and handle. The volume allotted for this unit at the bottom of the radio bay, coupled with the need for a heat sink that could dissipate 38 watts, indicated that the most efficient and economical

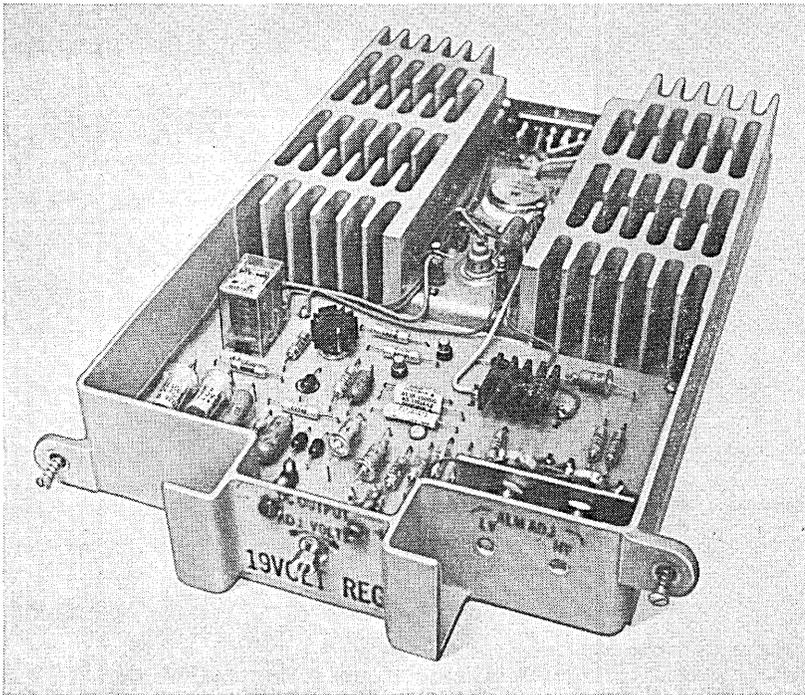


Fig. 11 — The regulated  $-19$ -volt power supply.

approach would be an aluminum die-cast one-piece plug-in unit. Overall, it is about 10 inches wide, 2 inches high, and 13½ inches deep. The epoxy glass printed board which contains most of the components is mounted within the frame. The heat sink area accommodates a power transistor and several other heat dissipating components. The face of the unit has such maintenance components as a pair of pin jacks for external connection to a dc voltmeter, and a potentiometer for adjusting the output voltage. A plug which makes external connections is mounted on the rear of the frame. The unit is fastened in place by two quick-release fasteners attached to the frame assembly. It is designed so that two power supplies may be placed beside each other in the bottom space of a TD-3 radio bay.

### 3.3 Transmitter-Receiver Test Set

The power supply for the transmitter-receiver test set<sup>2</sup> is a magnetically and electronically regulated unit which furnishes power at five different voltage levels. The requirements for this power supply are given in Table II.

Shown in Fig. 12 is a simplified block diagram of the power supply. Two ferroresonant regulators designated No. 1 and No. 2 receive 60 Hz, 117 V ac power through the 4-ampere circuit breaker, CB. Two ferroresonant regulators were needed to prevent excessive cross-regulation and because two different short circuit droop characteristics were required. The droop characteristic of ferroresonant regulator No. 2 protects the +300-volt active series regulator against overloads. All other outputs are protected by fuses.

The 6.4 V ac output is derived from an unregulated winding on ferroresonant regulator No. 1. Therefore this voltage follows the 60

TABLE II—TRANSMITTER-RECEIVER TEST SET POWER SUPPLY REQUIREMENTS

Output voltage	Voltage stability	Output current	Ripple (rms)
+300 V dc	±0.3 V	80-130 mA	1 mV
-150 V dc	±0.8 V	7-10 mA	1 mV
-40 V dc	±0.3 V	5-10 mA	1 mV
-24 V dc	-22.5 to -26 V	250-950 mA	300 mV
6.4 V ac	±0.83 V	3.3-3.9 A	—

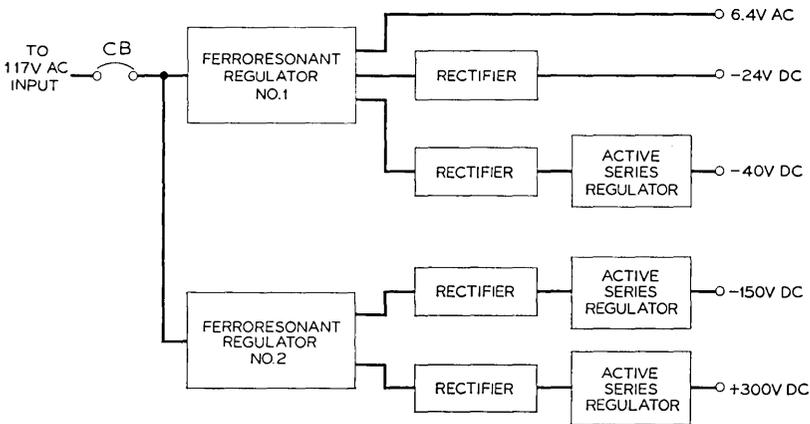


Fig. 12 — Simplified block diagram of the test set power supply.

Hz input voltage. The ac input voltages to the four rectifiers are regulated to minimize the voltage variations at the inputs of the three active series regulators. All of the active series regulators are conventional in design.

Figure 13 shows the power supply for the test set. It mounts at the rear bottom of the test set. The power supply consists of an open aluminum structure with most of the apparatus mounted on top, and the wiring on the bottom of the chassis. The transformers, resonating capacitors and  $-24$ -volt filter capacitors are on the chassis. The front panel of the power supply has fuse holders, fuses, circuit breaker, and access to the four plug-in units. Die-cast frames and guides are used to mount the four plug-in cards. One card contains all the rectifiers and protective diodes for the entire power supply. The other three cards are the voltage regulators for the  $+300$ -,  $-150$ -, and  $-40$ -volt outputs.

The input to the power supply is protected by a circuit breaker on the front panel. The input and output connections are made through quick-disconnect plug and jack assemblies located on the rear of the chassis. Various voltage adjustments, test points, and controls are mounted on the front panel. An extender assembly is used for servicing the power supply. The extender has a protective grid so that the plug-in units, when inserted, may be checked with safety. In order for the extender and the plug-in to be engaged the grid must be in a position to protect the users against the potentials present.

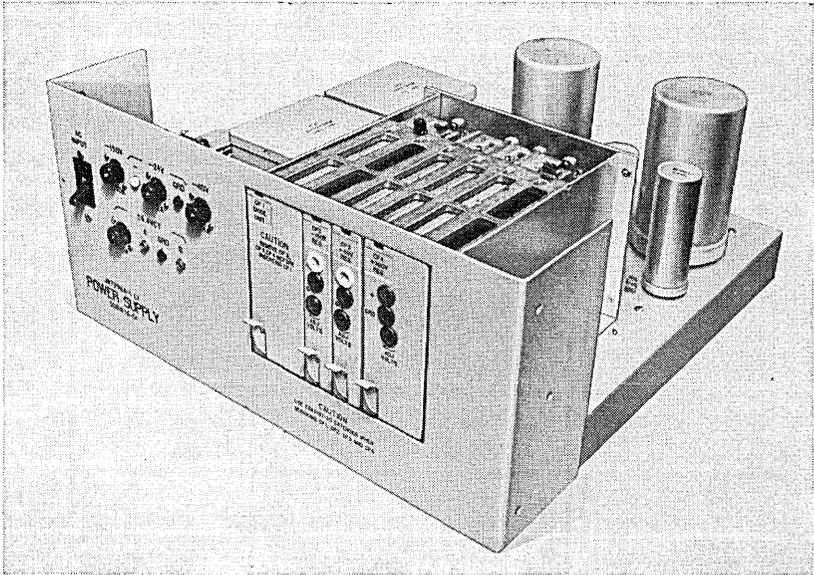


Fig. 13— Test set power supply.

#### IV. THE DC DISTRIBUTION CIRCUIT

The two paramount considerations which influenced the design of the dc distribution circuit were system electrical noise, and reliability. To minimize electrical noise, dc power feeders from the  $-24$ -volt battery plant fuse panel via the battery distribution circuit breaker bay are segregated into three groups: the noisy group which primarily feeds converter power supply loads, the quiet group which feeds noise-sensitive loads, and the undesignated group which feeds loads which can, on occasion, produce noise on their dc power feeder leads. The reliability criterion is that no more than half of the channels in one transmitting direction (one end) shall be lost as the result of a battery distribution fuse failure.

The allowable loop voltage drop from the  $-24$ -volt battery is determined by the type of battery plant used. For an 11-cell plant with emergency cells, 1.25 volts are permitted; for 12-cell plants with emergency cells, 1.75 volts are permitted; and for 12-cell plants without emergency cells, 1.0 volt is permitted. These allowable drops include the voltage drop across the circuit breaker in the circuit breaker bay, and agree with the recommended discharge loop voltage

drops which appear in common systems -24-volt battery plant drawings.

Figure 14 illustrates a simplified schematic of the -24-volt dc distribution circuit for one direction of transmission. Half of the channels transmitting in one direction receive their -24-volt distribution from quiet Q bus 1 and noisy N bus 1; the other half receive theirs from Q bus 2 and N bus 2. Not only is dc power distributed to the transmitter-receiver bays in such a manner as to preserve the reliability objective of the system, but both plus and minus 24-volt dc power also is distributed to all equipment associated with each transmitter-receiver bay in such a manner as to preserve the reliability objective. The other equipment includes 100A protection switching and 3A FM terminals.

In other words, it is not possible for the loss of one fuse at a battery plant fuse panel to cause a station to lose, in one transmitting direction, the transmitter-receiver bays associated with channels 1 through 6 (or any other designations they may have) and channels 7 through 12 of the 100A protection switching or other channel associated equipment. Four circuit breakers at the circuit breaker bay are required for each transmitter-receiver bay in a main station, and three for each in a repeater station.<sup>3</sup> Figure 14 shows that at least five -24-volt dc distribution fuses at the battery plant fuse panel are required for each direction of transmission.

Overcurrent tripping or the inadvertent turn-off of any circuit breaker in the breaker bay will set off a major office alarm. An alarm cutoff button on the circuit breaker bay can be pushed to silence the bell, turn off the red aisle guard lamps and turn on the white. Lamps on the alarm panel indicate which bus is affected.

The battery distribution circuit breaker bay is available as a 7-foot high unit in the same kind of cabinet as the newer -24-volt battery plant, or as a 9-foot duct type bay typical of radio equipment. The 7-foot cabinet is typically positioned adjacent to the battery power plant. The duct bay, which is compatible with 100A protection switching and 3A FM terminal equipment, is placed close to the load center of the equipment. In most instances it is expected that the 7-foot cabinet will be adequate for the TD-3 radio bays in main or repeater stations, and the 9-foot duct bay will be used to distribute to the various IF terminal and associated equipment. Both arrangements may be furnished with groups of circuit breaker panels.

The 7-foot cabinet includes ground bus-bar details so that a metallic

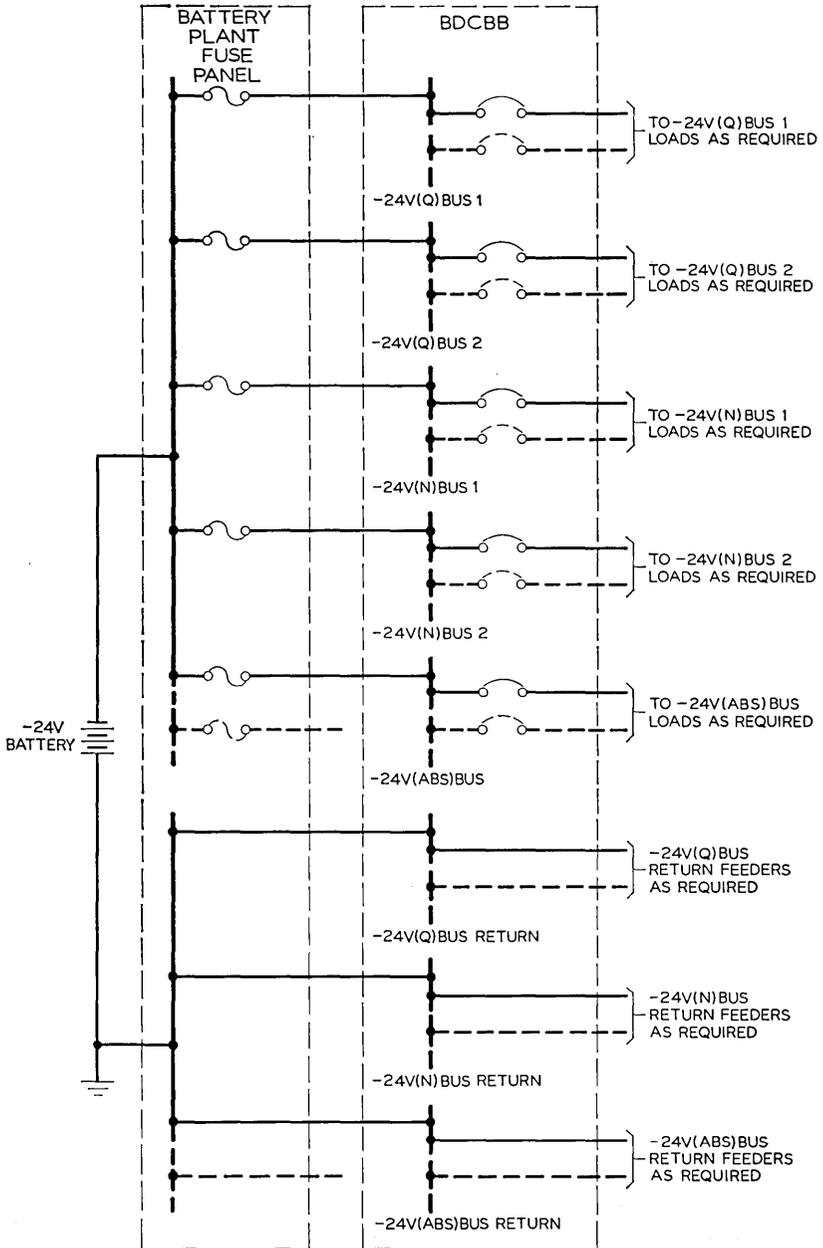


Fig. 14 — Simplified schematic diagram of the -24V dc distribution circuit for one direction (one end) of transmission.

pair is available for each circuit to be served by the cabinet. The 9-foot bay has two removable rear covers, for the upper and lower sections. Covers are for access where circuits are being added.

#### V. RADIO GROUNDING CIRCUIT

Tower-integrated buildings, as might be expected, are extremely vulnerable to lightning strokes. It is very important, therefore, that the radio grounding circuit for such a building be extremely well designed. The circuit design objective is to minimize the impedance to ground seen by the lightning-produced current. This, in turn, minimizes the potential differences within the building produced by this current.

Figure 15 illustrates, in simplified form, the radio grounding circuit for the tower-integrated, TD-3 radio buildings. Outside the building the external ring ground, which consists of No. 2, bare, tinned, copper wire, connects 5-foot,  $\frac{1}{2}$  inch diameter ferrous ground rods a maximum of 10 feet apart. The external ring ground is buried not less than  $1\frac{1}{2}$  feet below the final grade level. The internal ring ground consists of bare, No. 2 copper wire and is on the walls inside the building high enough to eliminate possible mechanical damage, yet low enough for visual inspection. The roof ring ground, which is also bare, No. 2 copper wire, is embedded in the concrete roof slab. The three ring grounds are electrically connected at many well spaced points, and are connected to the reinforcing steel of the building. Structural building steel, including the tower, is connected to the three ring grounds in many places. The four waveguide hatch plates are connected directly to the roof ring ground and all waveguides are connected to their respective hatch plates as well as to the internal ring ground.

Inside the building, two bare No. 2 copper wires run the length of the building, one on each side of the main aisle. These two wires are connected at both ends to the internal ring ground and, thus, form part of the internal ring. Each equipment bay is directly connected with a bare No. 6 copper wire to a bare No. 2 copper wire which runs along the cable rack of each bay line-up. This arrangement can be seen in Fig. 15. When the station is completely filled, the internal ring ground will be a mesh suspended above and electrically connected to all equipment with leads as short as practicable. When a station is grounded in this manner, large differences of potential between adjacent equipment during a lightning stroke are impossible.

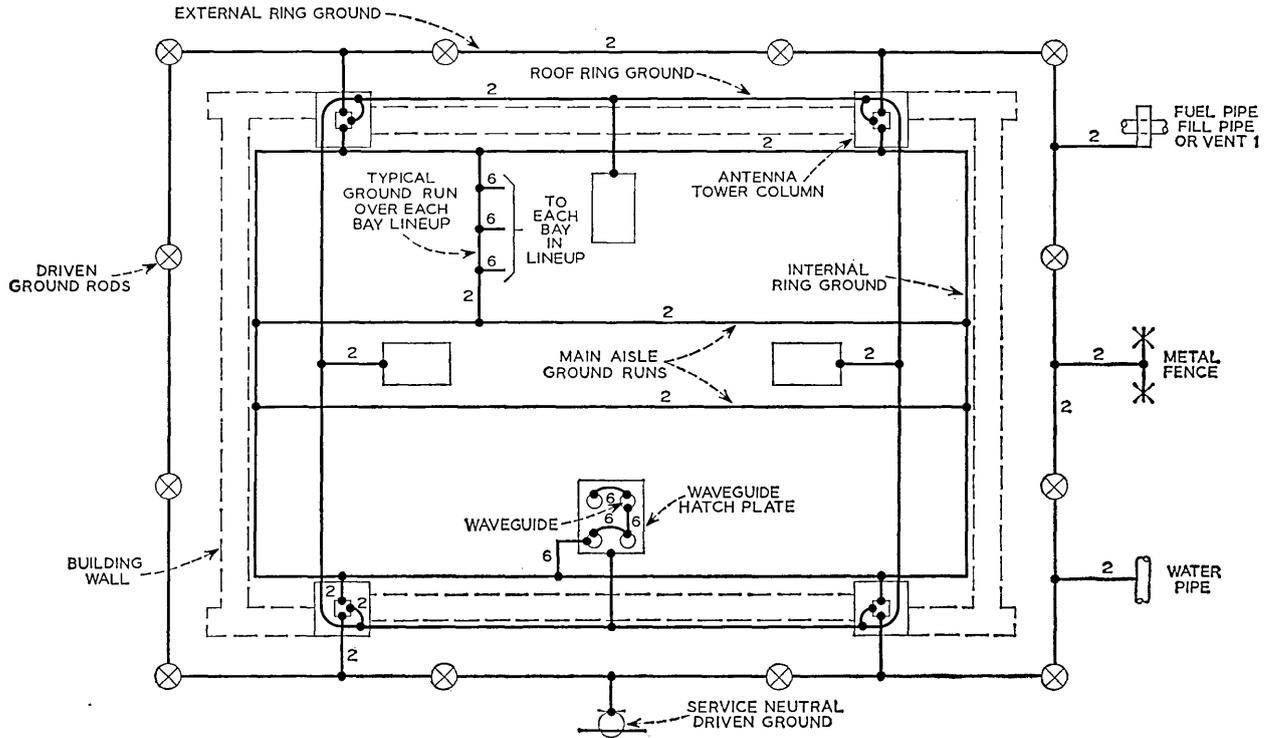


Fig. 15—Simplified radio grounding circuit for the tower-integrated buildings.

## VI. ACKNOWLEDGMENTS

Many members of the Electronic Power Systems Laboratory have contributed to the development of the power system for TD-3. Among them the authors wish to mention T. G. Blanchard, S. J. Brolin, W. H. Macfie, G. W. Meszaros, R. P. O'Connell, E. W. Paessler, M. Parente, and F. J. Salvo, as well as former members P. C. Chen and E. Koczukur. Special mention is made of E. A. Hake whose knowledge of the TD-3 system and recommendations contributed more than substantially to power system work.

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# Microwave Radio Equipment and Building Considerations

By R. J. SKRABAL and J. A. WORD

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*This paper points out salient equipment and building considerations peculiar to the TD-3 radio system and in particular emphasizes the important differences in station design that exist for TD-3 compared with TD-2 radio.*

## I. INTRODUCTION

Microwave radio relay stations have existed in the Bell System in several forms since 1950.<sup>1</sup> In this time, the equipment and the layouts within the buildings have changed to accommodate the new designs that have been introduced into these systems.

The introduction of the TD-3 repeater bay offers further opportunities for improvements particularly in the simplification of power requirements. The basic power for a TD-3 station is  $-24$  volts dc. Those power plant elements peculiar to TD-2, namely the  $+250$ ,  $+130$ , and  $-12$ V dc plants have been eliminated. Voltages other than  $-24$ V dc are furnished by dc-to-dc converters, except for  $+24$ V dc which may be furnished by a separate plant in some stations.

Except for the power plant, the typical TD-3 station will not differ greatly from recently constructed TD-2 stations, since the 100A IF protection switching system,<sup>2</sup> the 3A FM terminals, and the 3A solid-state wire line entrance links have been previously introduced into the TD-2 system. The 200A protection switching system for use with the 3A FM terminals and wire line entrance links has been developed for long haul microwave radio systems.

## II. TYPES OF STATIONS

There are two general types of TD radio stations: main radio stations and radio repeater stations. Main stations may be end points or intermediate points in the system where signals are accepted from

and delivered to terminal facilities. In addition, main stations provide points of flexibility in the system for facilities such as IF patching and monitoring, automatic protection switching, dropping or picking up local or spur radio signals and maintenance switching from regular to protection channels. Such stations also are convenient locations for performing system tests and may include baseband multiplexing equipment. Main stations may be partially or fully attended.

The radio repeater stations provide transmission gain and maintain line-of-sight paths. They comprise the majority of the stations in any large system and are normally unattended. Unlike main stations, the repeater stations are quite uniform in their makeup and are more adaptable to standardized floor plans and building construction.

### III. MICROWAVE EQUIPMENT

The general design concept for all TD-3 equipment is to provide, insofar as practical, for assembly, wiring and testing of complete bays in the manufacturing shop. This results in a reduction of job engineering, installation, and field testing effort. For convenience and speed in restoration of service in the event of failure and, to reduce routine maintenance effort, extensive use is made of plug-in units which are carried as spare parts in the stations and at centralized maintenance repair centers. The major equipment required in a radio station is listed in Table I.

The microwave transmitters, receivers, and associated equipment in the radio room are mounted on 9-foot unequal flange, duct-type frameworks (see Fig. 1). This applies in general to all items except power plants which are mounted on 7-foot frames. The transmitter-receiver bays require only front access for maintenance and may be mounted against a wall or back-to-back with other bays. In general, the remaining equipment requires front and rear access. The color scheme in a TD-3 station is light gray with border items such as guard rails and end guards in a contrasting darker blue gray color. The superstructure with cable rack is just above the tops of the frames. Indoor waveguide arrangements are similar to TD-2 and, as in the latest versions of TD-2, use close tolerance flanges to assure a return loss of at least 40 dB.

Use has been made of a short piece of flexible waveguide to connect between the dropping networks of adjacent bays thus insuring a better connection and less possibility of R.F leakage. This also simplifies to a considerable extent installer effort in accurate alignment

TABLE I—MAJOR EQUIPMENT IN A RADIO STATION

Unit	Station	
	Main	Repeater
1. Transmitter-receiver bay	Maximum of 12 for each route direction	Maximum of 24 for a through route
2. Transmitter-receiver auxiliary channel*	One for each route direction	Maximum of 2 for a through route
3. Multiplex equipment to derive voice circuits on auxiliary channels	X	X
4. C1 alarm and order circuit equipment	Maximum 7 bays (21 scans—882 alarms and or indications and 490 orders)	Maximum 2 bays (3 scans) for most stations
5. 100A protection switching equipment	3 bays for each route direction	Maximum 2 bays if required to protect dropped channels
6. IF patch and access	One bay provides for a maximum of 12 2-way radio channels in each of four directions	—
7. 3A FM terminals and FM terminal patching	1 bay for each 4 pairs of transmitters and receivers	As required for drop or pickup
8. 200A FM terminal and wire line entrance link protection switching	Approximately 3 bays in the radio room and 2 bays in the multiplex area for each 1 × 12 switching system	—
9. IF restoration	X	—
10. IF and baseband monitoring	1 bay	—
11. Test equipment and spare parts	X	X
12. Dehydrator for dry air supply to antenna system and to the transmitter-receiver bays	1 or more as required	1 for average station

\* This channel is used to provide order wire service, and to carry control tones for IF protection switching and alarm systems. Alternatively, wire lines may be used instead of the auxiliary radio channel.

Unit	Station	
	Main	Repeater
13. Tower obstruction light controls	X	X
14. AC power distribution	X	X
15. Engine alternator plant	X	X
16. 111A -24 volt power plant	X	X
17. -24 volt battery	X	X
18. 111A +24 volt power plant and associated battery	X	If required
19. 24 volt battery distribution circuit breaker board	Two	One
20. Heating and air conditioning equipment	X	X
21. Antennas and associated waveguide	X	X

X = as required.

of bays which was required in initial TD-2 systems when hard waveguide fittings were used.

Attention is given to isolation of particularly noisy leads on the cable racks to reduce noise interference problems. This is accomplished by segregating such leads in metallic duct, thin walled conduit, or assigning such leads to dedicated cable racks or runs. Power leads to high-transient producing loads such as dc-to-dc converters come under this classification.

In addition to the above, leads carrying control signals between radio bays and units, such as alarm bays, are paired or looped insofar as practical to minimize noise problems. This also applies to power leads.

#### IV. FLOOR PLANS

Figure 2 is a block schematic diagram of a main station complex with two-way radio channels in each of four directions. This figure

illustrates the most efficient arrangement of equipment for a main station with the 70 MHz IF patch and access bay, shown as an octagon, as the focal point of the office. If the equipment could be laid out in this manner, the IF cable lengths would be minimized. Reducing cable length reduces the cross modulation noise resulting from echoes and facilitates the program to standardize IF levels in the radio station.

The typical floor plan (Fig. 3) shows a similarity to the ideal ar-

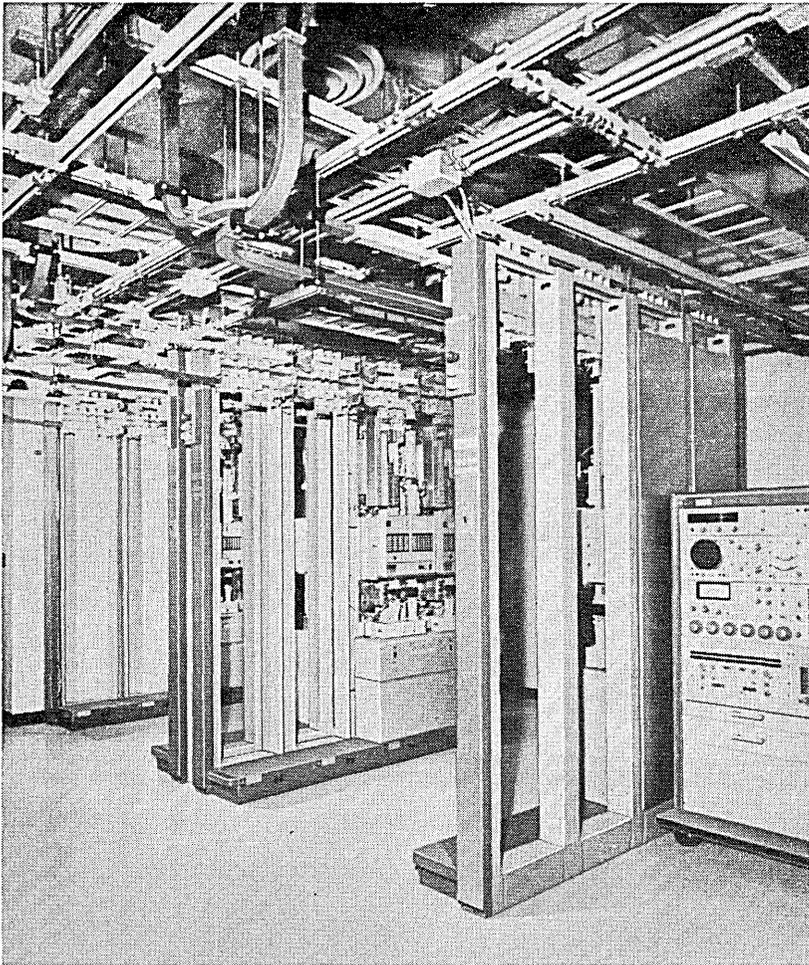


Fig. 1 — Microwave radio equipment bays in a station lineup.

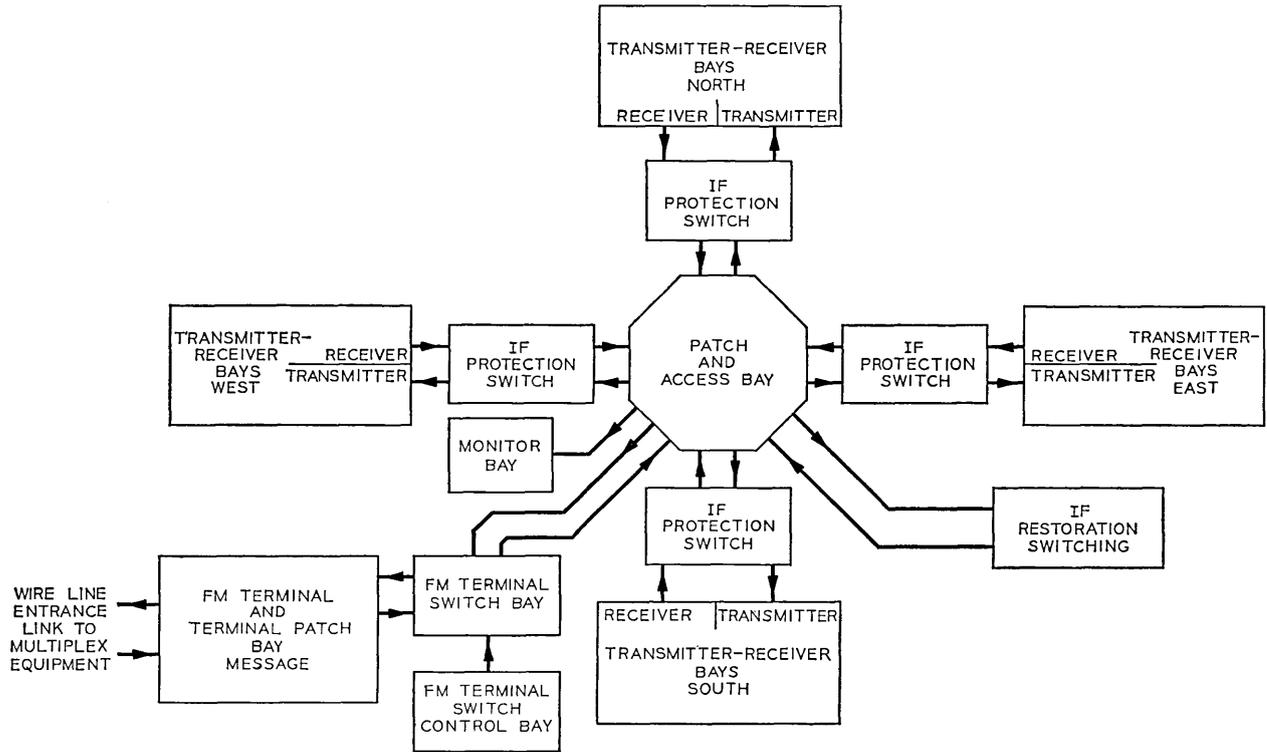


Fig. 2 — Typical block diagram of a TD-3 main station.

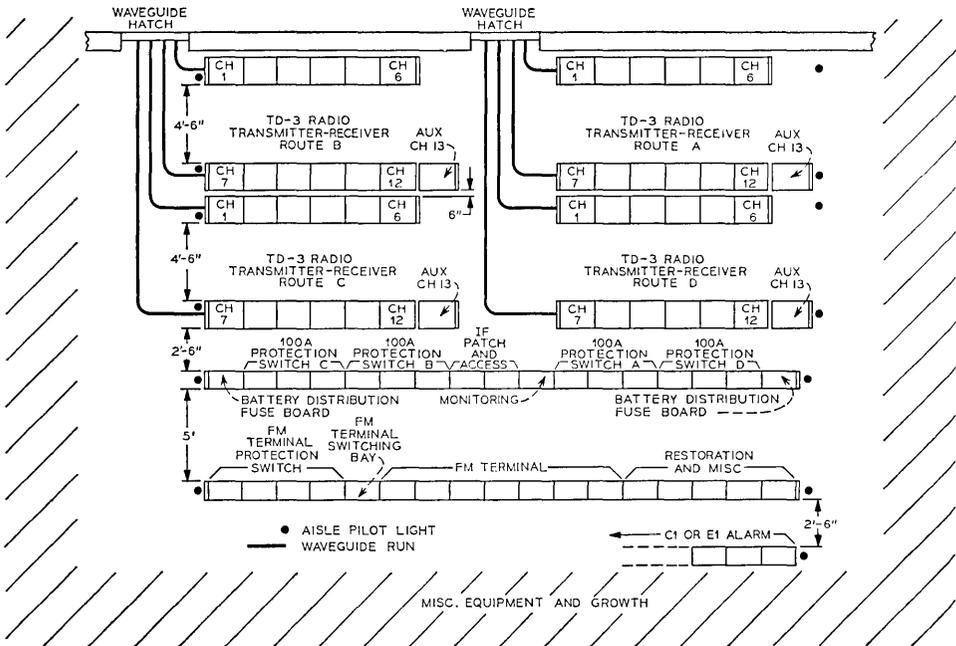


Fig. 3 — Typical main station equipment layout.

rangement of Fig. 2. The IF patch and access bay is centrally located and the other equipment is located so as to minimize the lengths of IF cables and thus approach the ideal as far as practical. The location of the transmitter-receiver bays and the use of two waveguide hatches, as depicted in Fig. 3, minimizes the length of the inside antenna waveguide. Figure 4 shows the layout for a repeater station. In this case there are no IF cabling problems since each transmitter-receiver bay is internally connected as a through circuit. The floor plans in Figs. 3 and 4 are typical of new stations now being engineered. Figures 5 and 6 are floor plans for main and repeater stations, respectively, used on the Arkansas-Mississippi trial. The primary consideration in these layouts is the placement of the transmitter-receiver bays in the strategic area directly under the tower in order to minimize the length of the waveguide runs.

#### V. PATCH AND ACCESS BAY

The IF patch and access bay (see Figs. 7 and 8) has jack fields for terminating the inputs and outputs of radio channels (both regular

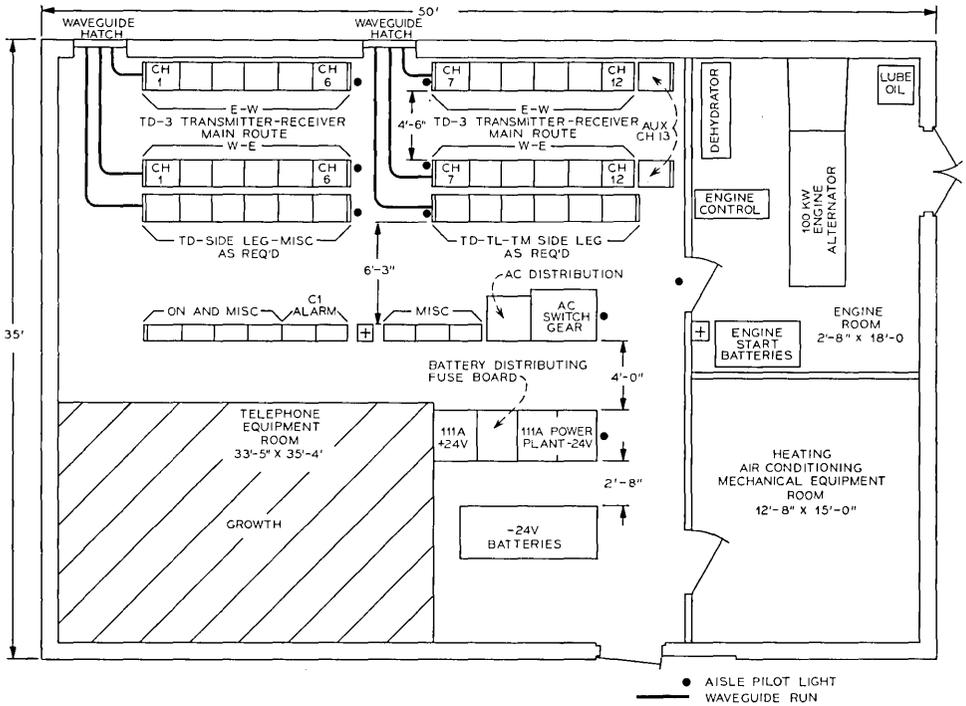


Fig. 4 — Typical repeater station equipment layout.

and protection) for flexibility and for maintenance and testing. This bay can be considered as a hub which interconnects the IF cabling from the protection switching bays, FM terminal patch bays, monitoring bays, restoration bays, and so on. The IF patch and access bays contain jack mountings for twelve incoming and twelve outgoing channels with their associated monitoring taps for four directions of transmission. The channels appearing in this bay may also be connected to reed type coaxial switches and coaxial type directional couplers for connecting through restoration equipment associated with FM terminal facilities or they may be connected through to trunks directly associated with FM terminal equipment. Since not all incoming circuits require switches and directional couplers, this bay has been designed to have a maximum capacity for 36 switches and couplers.

A revision of the monitoring bay is being developed to be associated with the patch and access bay. This monitoring bay will in-

clude a 3A FM terminal receiver to reduce IF signals to baseband, a de-emphasis network to compensate for the preemphasis of the signal at the originating terminal, an IF amplifier to compensate for the inherent loss of a monitoring tap, and a video monitoring amplifier and an "A" scope for visual observation of the signal. Facilities for local and express order circuits will also be included.

## VI. ALARMS AND MAINTENANCE ORDER CIRCUITS

The individual audible and visible alarm equipment in TD-3 offices are decentralized with the individual alarm relay circuits located

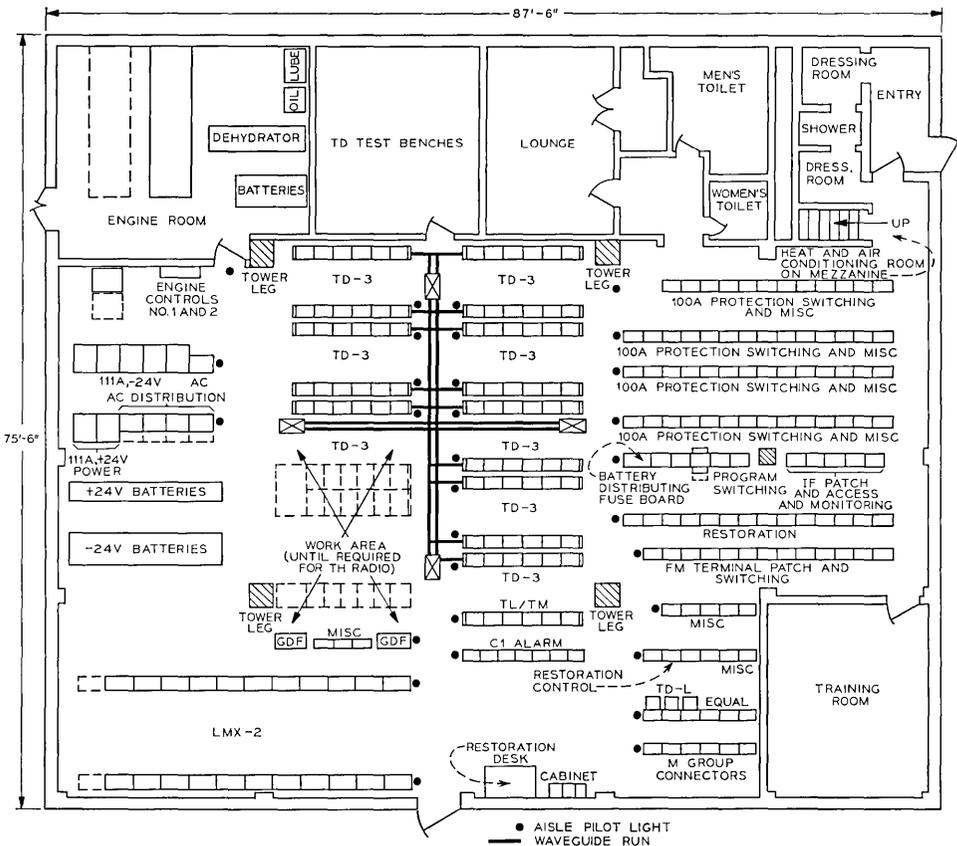


Fig. 5—Main station floor plan at Arkabutla, Mississippi, on the initial TD-3 route.

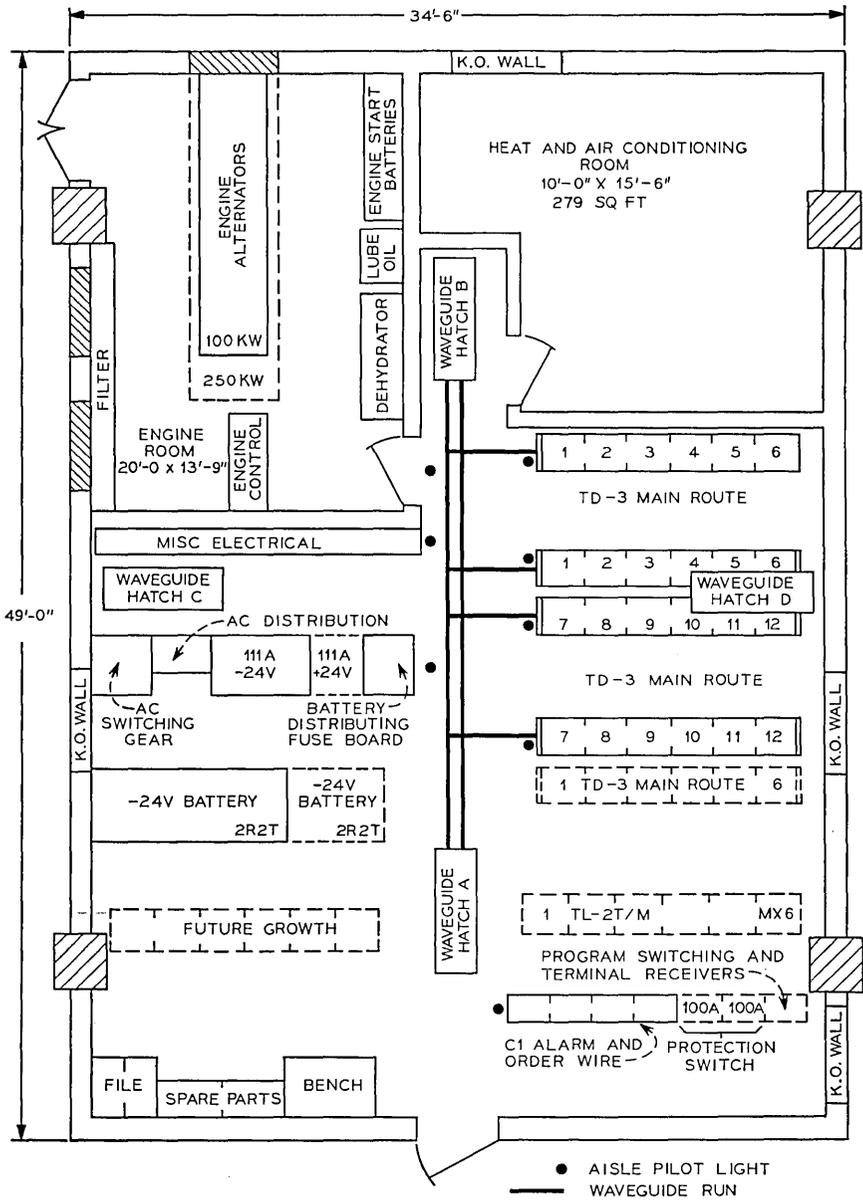


Fig. 6 — Floor plan of repeater stations on the initial TD-3 route.

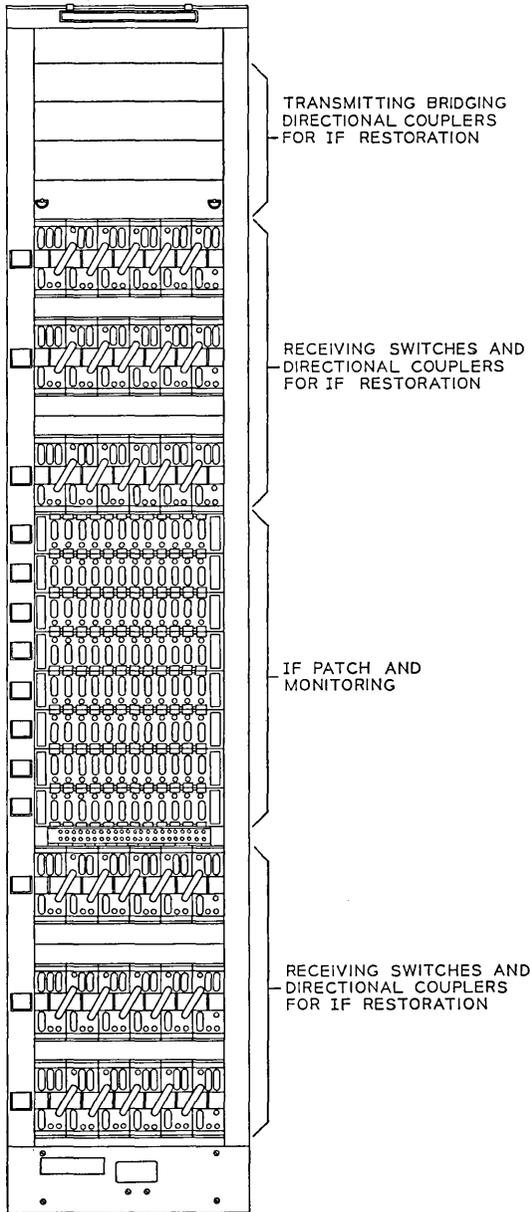


Fig. 7 — IF patch and access bay.

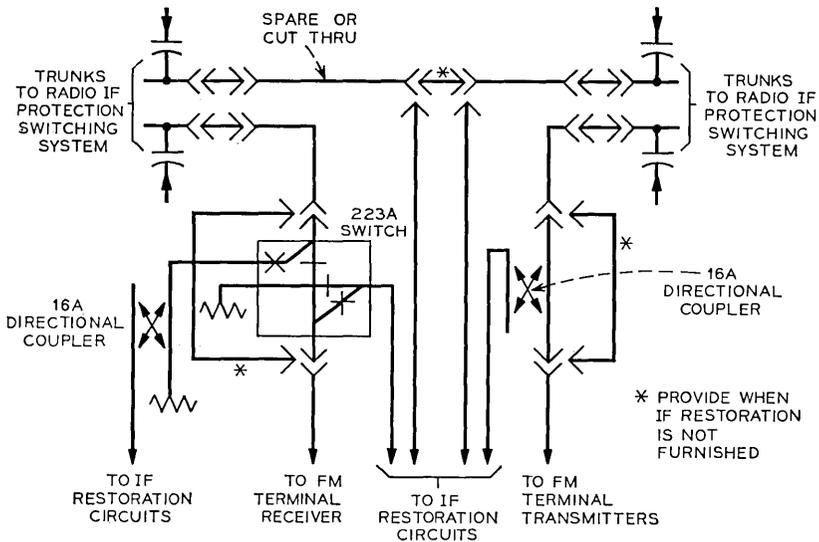


Fig. 8—Typical application circuit for IF patch and access bay with monitoring taps.

in the equipment units where the alarms originate. The alarm system, which directs maintenance personnel by means of pilot lights on the aisles to the point of the alarm, is adapted for both main and repeater stations. Relays for common office alarm circuits such as open door and tower navigation lights, are mounted together at one location. These circuits also connect to the aisle pilot system. The existing C1 alarm and order system<sup>3</sup> or the new E1 system may be used as the connecting link to transmit alarm and status information from unattended stations to the attended control point and to transmit orders outward from the control point. The C1 alarm and order system is in wide use in the field. It uses relay type circuitry and has a maximum combined capacity of 882 alarm and station indications and 490 orders for each of as many as 12 stations per system. The new E1 system, using a combination of solid state and relay circuitry, has expandable capacity and will fulfill any foreseeable requirements for TD-3 radio with respect to transmission of information and the number of stations served. It has also been drastically reduced in size; that is, a maximum C1 system requires 7 frames of equipment whereas a comparable E1 system requires only a single frame.

New maintenance order circuits also are being introduced which include these basic improvements over the existing systems: (i) solid

state circuitry is used, (ii) the order circuits are independent of the E1 transmission facility (iii) conventional dialing arrangements are available for signalling between stations, and (iv) the express and local order circuits may be interconnected when required.

#### VII. DEHYDRATORS

A dehydrator is available for TD-3 offices to provide dry air at a positive pressure to the outdoor antenna waveguide system to prevent water accumulation thru condensation or leakage. Dry air is also used to charge the transmitter-receiver critical waveguide filter elements. These Invar filters require regulation of temperature and humidity to maintain frequency stability. By direct introduction of dry air into the filter cavities the station temperature may be allowed to vary within reasonable limits namely,  $70 \pm 20^\circ\text{F}$  and no requirements need be placed on humidity. The dehydrator operates on a combined refrigerating and desiccant principle and will provide 100 cubic feet per hour of  $-40^\circ$  dew point air at a pressure of approximately 7 inches of water. It is estimated that a leakage of about 0.8 cubic foot per hour per bay, or approximately 20 cubic feet per hour for a repeater station with 24 bays, will occur. Available information shows that an average antenna system may leak about 8 cubic feet per hour, or a station equipped with 4 antennas may leak at a rate of about 32 cubic feet per hour. However, corrective maintenance is not undertaken until antenna system leakage approaches 15 cubic feet per hour. This leakage rate, therefore, determines the ultimate air usage and results in a total approximate requirement of 80 cubic feet per hour for a repeater station. This amount can be furnished by a single dehydrator. At larger repeater stations and at main stations additional dehydrators are needed. The dry air for the antenna system is delivered to the waveguides at the pressure windows by a manifolding arrangement. This dry air to the channel dropping filters is introduced through the inside waveguide using a flange-type fitting located adjacent to the waveguide pressure window. When two separate dehydrators are furnished, separate manifolding runs are used. A manual transfer valve is provided so that in the event of failure of one dehydrator, the remaining unit can provide the total supply at some reduction in pressure.

#### VIII. ANTENNA SYSTEM

The horn reflector antenna with circular waveguide and systems combining (or separation) networks, used in the earlier TD-2 and TH systems, is also used with TD-3. This antenna system, capable

of simultaneously carrying orthogonally-polarized signals in the common carrier frequency bands of 3.7 to 4.2, 5.925 to 6.425 and 10.7 to 11.7 GHz, has been described in detail in previous papers.<sup>4-6</sup>

A vertical run of circular copper waveguide (2.81 inch inside diameter) connects the antenna to appropriate systems combining networks at the base of the antenna supporting structure. Since two polarizations, vertical and horizontal, are available for each of the three frequency bands, as many as six rectangular waveguides may connect from a single run of circular waveguide to the transmitter-receiver bays in the microwave radio equipment room. The antennas, networks, and all outdoor waveguides are pressurized with dry air. An RF window in the waveguide at the building entrance separates the outside and inside waveguide sections.

Generally, a repeater station is equipped with four antennas, one transmitting and one receiving in each direction, while stations located at route intersections may have six or more antenna systems. Therefore, plans for fully-equipped stations must provide for as many as six rectangular waveguides per antenna—at least 24 waveguides at a repeater station and more at main stations.

To avoid undue signal attenuation and to control noise contributions of the antenna system, the waveguide runs should be kept as short as possible and, therefore, must be carefully engineered and installed. Standardization or pre-engineering of waveguide runs is highly desirable but difficult to achieve with rigid waveguide. The arrangements for the tower-mounted circular waveguides and the indoor portions of the rectangular waveguides have been standardized. The outdoor rectangular waveguide runs are generally specially engineered for separate building and tower structures—each situated according to its own requirements with respect to radio path directions and local site conditions. Nevertheless, they usually follow typical patterns.

## IX. BUILDINGS AND TOWERS

The building and tower shown in Fig. 9 are typical of those being used for TD-3, TD-2, and TH. The building is a reinforced concrete structure with 1625 square feet of equipment floor space.

The tower will accommodate four horn reflector antennas and a pair of 8-foot-diameter parabolic antennas. At the station shown in Fig. 9, the shrouded antennas, mounted approximately 50 feet below



Fig. 9— This installation at Dumas, Mississippi, is typical of current TD-3 repeater stations.

the horn-reflectors, are used in a narrow-band space diversity system, operating at 4190 or 4198 MHz, which provides a high-reliability auxiliary channel for order and alarm circuits.

The towers used at main stations have the capacity for mounting

eight horn reflectors on two top platforms. Both "four and eight antenna" towers of this tapered truss design are available in heights up to 350 feet. Galvanized ASTM A 36 steel is used, although for economy some leg sections are made of high strength steel (ASTM A 441). Pier-and-pad footings, independent of the building foundations, are generally used so that the tower is 15 to 20 feet from the building. The separation is sometimes greater, although in recent construction it has been held to a minimum. For a four-antenna tower, the tower base width varies from about 18 to 53 feet, and is even larger for an eight antenna tower.

The systems combining networks are mounted on the outside face of the tower. Waveguide support trusses carry the rectangular waveguides from these networks to the entrance panels in the rear wall of the building. Both the networks and the horizontal waveguide runs are covered to prevent damage from falling objects or ice.

It is difficult to standardize outdoor waveguide arrangements with separate buildings and towers because of (i) the large number of variations in tower base dimensions and building-tower separations, (ii) the occasional positioning of the tower at an odd angle with respect to the building plan, and (iii) the need to connect antennas located in various positions on the tower with transmitter-receiver bays which are uniformly placed within the building. Furthermore, the building-tower separation and the relatively large dimensions of the tower base result in long rectangular waveguide runs. To eliminate many of the problems associated with outdoor waveguide arrangements, a new approach was used on the TD-3 system installed between Alexander, Arkansas, and Arkabutla, Mississippi. A "unitized" building-tower arrangement with relatively short pre-engineered waveguide runs was used. In this arrangement, the tower is mounted directly on a reinforced concrete building. The tower legs are anchored to the building columns and the tower footings are integral with the building foundation.

"Unitized" designs were developed for both repeater and main stations. The unattended repeater building provides approximately 1800 square feet of equipment floor space, including separate engine and mechanical equipment rooms. Figure 10 shows the "unitized" type at Stuttgart, Arkansas, where the tower is about 287 feet high. Anticipating future growth, this tower is designed to support as many as six horn-reflector antennas. The availability of eight possible antenna mounting positions on a large platform offers a wide range of orienta-

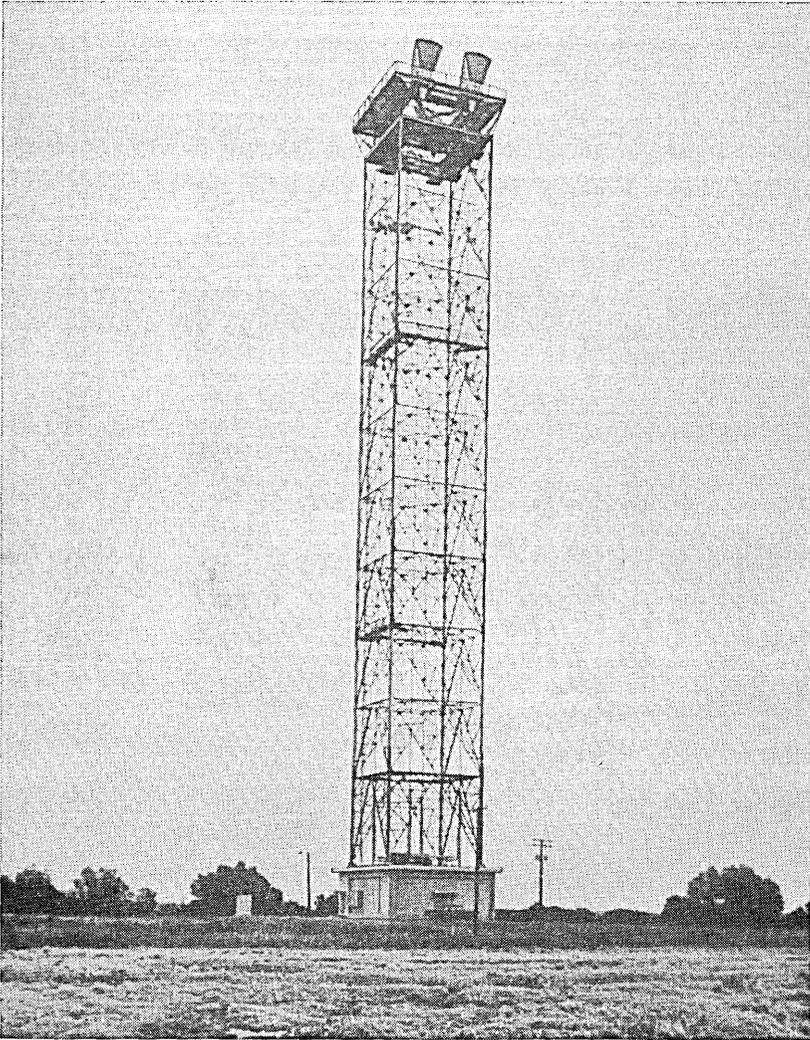


Fig. 10 — The repeater station at Stuttgart, Arkansas, on the initial TD-3 route, uses a “unitized” building-tower arrangement.

tion possibilities, so that it rarely will be necessary to orient the tower with respect to the microwave paths as must frequently be done with conventional towers for four antennas.

Corresponding main junction buildings, such as those provided at

Alexander, Arkansas, and Arkabutla, Mississippi, are several times as large as those used for the repeaters. They provide additional space for the protection switching, terminal, order wire and alarm, patching, and test equipment, as covered in Table I, as well as multiplex or other locally required equipment. The building at Alexander, with a 137 foot tower (Fig. 11), is designed to support 12 horn-reflector antennas—four on a second platform. A second floor on one side of the building (right in Fig. 11) houses air conditioning and heating equipment.

The towers in the "unitized" construction are galvanized steel 33 feet on a side, up to 350 feet high. The legs are wide flange H sections of high strength steel (ASTM A 441), ranging from six inches deep and weighing 15.5 pounds per foot near the top, to 14 inches deep and weighing 150 pounds per foot at the base of the heaviest tower. Each baseplate is secured by anchor bolts to the building column on which it rests. For the heaviest tower, eight 3-inch diameter bolts are required at each leg. The towers are designed to withstand loads corresponding to approximately 100 mph winds. In a 70 mph wind, the antenna platforms will deflect less than  $\frac{1}{4}$  degree.

The large and uniform tower cross-section permits the running of the circular waveguide straight and inside the tower, rather than in the sweeping arc (500-foot radius) which brings it tangent to the outside face of the usual tapered tower. Short rectangular waveguide runs connect from the systems combining networks, mounted inside the tower face at roof level, to one of four entrance hatches on the roof. Networks and rectangular waveguide are protected from falling objects or ice by steel grating supported at the top of the tower base section which may be at a level 18 or 31 feet above the roof. In a departure from usual practice, the rectangular waveguide is positioned with its long dimension vertical so that small amounts of water or other foreign matter which may enter the guide can collect on the narrow side, where it has less effect on transmission.

Both the conventional and "unitized" buildings have exterior wall panels designed to facilitate their removal in the event the building must be enlarged. They also provide facilities for 100 kW engine alternators with room to expand the emergency power service to 225 kW if necessary.

Air conditioning in the trial installation was designed to maintain the radio equipment space at  $75 \pm 10^\circ\text{F}$ , with relative humidity below 50 percent. As the Invar waveguide filter elements are now being pres-

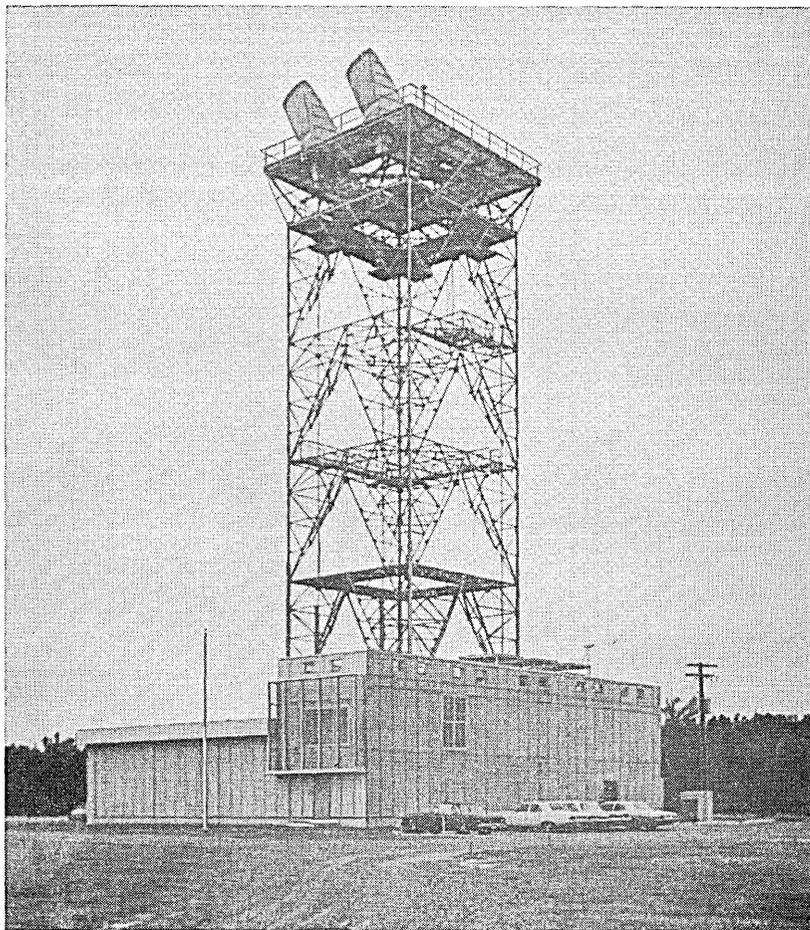


Fig. 11 — The main station at Alexander, Arkansas, on the initial TD-3 route.

surized with dry air, the station temperature requirement for future construction is  $70 \pm 20^{\circ}\text{F}$  with no limits on relative humidity.

In accordance with Bell System plans for continuity of communications in the event of nuclear attack, all structures are designed to withstand 2 psi peak overpressure and associated blast effects. Attended main stations are equipped with decontamination and emergency living facilities. Adequate shielding is provided to protect working personnel from fallout radiation.

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ALFRED J. ALBERTS, B.S.E.E., 1957, Newark College of Engineering; Bell Telephone Laboratories, 1947—. Mr. Alberts helped design microwave filters and networks for early experimental radio relay systems and for the TD-2 microwave system. He has been concerned with designing and developing microwave and coaxial apparatus for subsequent radio relay systems, and is engaged in the mechanical design of microwave transmission components for the TH-3 and TD-3 radio systems.

JOHN F. BARRY, B.S., 1961, and M.S., 1963, both from Northeastern University; Bell Telephone Laboratories, 1961—. Mr. Barry's work has included circuit development for TH Radio, *Bellboy*<sup>®</sup> personal signaling set, and FM terminals. He now is designing IF switching circuits for broadband restoration. Member, Eta Kappa Nu, Tau Beta Pi.

C. E. BRADFORD, B.S. (E.E.), 1942, Worcester Polytechnic Institute; M.S. and Ph.D. (physics), 1954, Illinois Institute of Technology; Bell Telephone Laboratories 1954—. Mr. Bradford has investigated the effect of manufacturing tolerances on the operation of electron guns, and has been involved in designing and developing Bell System and military traveling-wave tubes, Bell System masers and a military magnetron. Also, he designed and developed the traveling-wave tube for TH-3 radio system. He is responsible for design engineering for traveling-wave tubes in production by the Western Electric Company. Member, IEEE, A.I.P., Sigma Xi.

RICHARD T. COONEY, B.S.E.E., 1959, University of Rhode Island; M.S.E.E., 1962, Northeastern University; Bell Telephone Laboratories, 1959—. Mr. Cooney's work has included designing and developing analog portions of the 100A protection switching system, developing the TD-3 transmitter-receiver test set, a noise reduction program on the TH radio system and designing new protection switching systems for microwave radio. Member, IEEE, Eta Kappa Nu, Tau Beta Pi, Phi Kappa Phi.

ELBERT J. DRAZY, B.S.E.E., 1942, Purdue University; Bell Telephone Laboratories 1942—. Mr. Drazy has been concerned with developing test equipment for microwave radar and for microwave radio relay and video transmission systems, as well as developing carrier supplies for the L-type multiplex equipment, and FM terminals for long-haul microwave radio relay systems. Since 1964 he has supervised a group developing microwave networks for radio systems. Member, Tau Beta Pi, Eta Kappa Nu, Sigma Xi.

HARRY E. ELDER, who coordinated the writing of the 14 other men who wrote sections for the article, "Active Solid-State Devices," received his B.S. in E.E. from the University of Illinois in 1948, his M.S. in E.E. from Newark College of Engineering in 1954, and graduated from Bell Telephone Laboratories Communications Development Training Program in 1956. He joined Bell Laboratories in 1951 where he has been concerned with developing electron tubes and solid-state devices. Mr. Elder has been supervisor of microwave diode final development, klystron development, and microwave masers. He is now supervisor of microwave diode final development. He has one traveling wave tube patent and helped write *Microwave Semiconductor Devices and Their Circuit Applications*, McGraw-Hill, 1968. Member, Eta Kappa Nu, Tau Beta Pi, Sigma Tau, Phi Kappa Phi, Phi Lambda Upsilon.

G. L. FENDERSON, B.S.E.E., 1960, University of Maine; M.S.E.E., 1963, Northeastern University; Bell Telephone Laboratories, 1960—. Mr. Fenderson's early work at Bell Laboratories dealt with the development of central office noncommon control switching apparatus. Since 1962 he has been designing and developing broadband solid state amplifiers for microwave radio systems.

JAMES GAMMIE, B.Sc. (E.E.), 1944, University of Aberdeen, Scotland; B.Sc. (Mathematics), 1951, Birkbeck College, University of

London; Bell Telephone Laboratories, 1952—. Mr. Gammie has been concerned with developing transmission systems, including the L3 coaxial system and the TJ and TL short-haul microwave radio systems. He is now developing FM terminals and wire line entrance links for radio relay systems. Member, IEEE.

ANDRAS HAMORI, Dipl. Eng., EE, 1955, Technical University of Budapest, Hungary; Bell Telephone Laboratories, 1957—. Mr. Hamori was engaged in developing the decoder and the line repeater for the T1 carrier system. Later he worked on developing the exerciser and the test set for the 100A switching system. He also was associated with developing the transmitter modulator and the microwave generator for the TD-3 radio system. He now supervises a group developing the TH-3 radio repeater.

S. D. HATHAWAY, B.E.E., 1947, University of Virginia; M.S.E.E., 1950, Virginia Polytechnic Institute; M.S.E.E., 1952, University of Illinois Bell Telephone Laboratories 1952—. Mr. Hathaway has been engaged in systems engineering on microwave radio relay systems. He is Head of Radio Systems Engineering Department. Member, IEEE, Eta Kappa Nu, Tau Beta Pi.

W. G. HENSEL, B.S.E.E., 1929, Ohio Northern University; Bell Telephone Laboratories, 1929—. Mr. Hensel is Head, 4 GHz Long Haul Microwave Department. Member, IEEE.

J. JAN JANSEN, S.B. and S.M., 1939, Massachusetts Institute of Technology; Bell Telephone Laboratories 1939—. His early assignments included television terminal circuit design for the L1 coaxial system and the A1 video transmission system. During World War II he worked on several military airborne radar systems. He participated in designing the A2A video system and the television terminals for the L3 coaxial system. Since 1957 he has headed groups in developing the L3, T1, and microwave radio systems. Member, Sigma Xi, Eta Kappa Nu, IEEE.

R. M. JENSEN, B.S. in E.E., 1937, Purdue University; Bell Telephone Laboratories, 1937—. He first was engaged in developing audio equipment, and then worked in quality assurance. In 1938 he began developing L-type multiplex networks, and during World War II he was concerned with tuned amplifiers, coaxial filters, and crystal-

controlled oscillators. In 1956 Mr. Jensen was assigned to develop microwave networks, and in 1959 he became a supervisor. In 1964 he began supervising a group designing equipment for TL-2/TM-1 radio relay, FM terminals, wire line entrance link, and TD-3 radio relay, and now his group is designing TH-3 radio relay equipment. Member, Sigma Pi Sigma.

WILLIAM E. JEWETT, B.S.E.E., 1953, University of Kentucky; M.S.E.E., 1965, Stevens Institute of Technology; Bell Telephone Laboratories, 1953—. Mr. Jewett has worked on military and telephone apparatus, and regulated power supplies for telephone systems. In 1962 he was made supervisor of a military and telephone apparatus group, and in 1963 he became supervisor of a power conversion group. Member, IEEE, Eta Kappa Nu, Tau Beta Pi.

DOUGLAS R. JORDAN, B.S.E.E., 1952, University of Buffalo; Bell Telephone Laboratories, 1952—. He has worked primarily on the TH and TD-3 microwave radio relay systems. On TH, his assignments included work on the traveling wave tube amplifier, the auxiliary radio channel, the broadband channel equalization, and the antenna system. He supervises a TD-3 system analysis group responsible for establishing the requirements for the microwave transmitter and receiver, and evaluating the radio system performance. Member, IEEE.

FRANCIS M. KLISCH, B.E.E., 1963, George Washington University; M.S.E.E., 1965, Northeastern University; Bell Telephone Laboratories, 1963—. Mr. Klisch has been engaged in designing FM terminals and FM terminal test equipment for the TD-3 radio system and developing the new TH-3 6 GHz long haul radio system. Member, Tau Beta Pi.

SHEE H. LEE, B.S.E.E., 1961, Tufts University; M.S.E.E., 1963, Northeastern University; Bell Telephone Laboratories, 1961—. Since joining Bell Laboratories, Mr. Lee has been involved in designing and developing IF circuits for microwave radio systems.

N. E. LENTZ, B.S.E.E., 1952, Washington State University; Bell Telephone Laboratories 1952—. He completed Bell Laboratories' Communications Development Training Program in 1955, and worked on video amplifiers and test equipment for the TD-2 radio system. Mr. Lentz holds several patents for the timing circuits he designed for the

T-carrier PCM system. Since 1961 he has been developing circuits for FM terminals. Member, Tau Beta Pi, Sigma Tau.

D. J. LEONARD, B.E.E., 1956, and M.E.E., 1960, New York University. Bell Telephone Laboratories 1956—. Mr. Leonard has been concerned with the circuit design and analysis of amplifiers, networks, and logic circuits for various transmission and signaling systems. He has led groups responsible for transmission system circuit development as well as component design, including microwave circuits and networks. He is Director of the Transmission Technology Laboratory. Member, Tau Beta Pi, Eta Kappa Nu.

R. C. MACLEAN, Assoc. E. E., 1958, Franklin Technical Institute; Bell Telephone Laboratories, 1958—. Mr. MacLean has been involved in the design of filters and equalizers for various communication systems and as programmer in the analog computer simulation of various circuits and systems. He is now designing and evaluating active circuits and networks for communication systems.

SAMUEL MOTTEL, B.S.M.E., 1950, City College of New York; M.S.M.E., 1968, Newark College of Engineering; Bell Telephone Laboratories, 1952—. Mr. Mottel has been concerned with power equipment and systems development since joining the Laboratories. He has worked on power for carrier systems, microwave systems, Bell System and military submarine cable systems, data systems and various military communications systems. Since 1963 he has supervised a group working in systems power equipment development areas. Member, A.S.M.E.

PAUL L. PENNEY, B.S.E.E., 1963 and M.S.E.E., 1965, Northeastern University; Bell Telephone Laboratories, 1963—. Mr. Penney had worked on developing microwave upconverter and downconverter circuits for the TD-3 radio system. He is now working on broadband IF transistor amplifier circuits. Member, Tau Beta Pi, Eta Kappa Nu.

RONALD C. PRIME, B.Sc. in E.E., 1958, University of Southampton (England); M.S.E., 1961, Princeton; Bell Telephone Laboratories 1961—. His first assignment with Bell Laboratories was to design logic circuits for the 100A switching system which is used to protect TD-2 and TD-3 radio. Later he was engaged in TD-3 repeater bay,

and system measurements. He supervises the systems analysis group responsible for TH-3 radio. Graduate member, IEE.

CHUNG-LI REN, B.S.E.E., 1953, Taiwan College of Engineering; M.S.E.E., 1957, University of Notre Dame; Ph.D. (Electro-physics), 1964, Polytechnic Institute of Brooklyn; Bell Telephone Laboratories, 1965—. At Bell Laboratories, Mr. Ren has done modal analysis of waveguides with arbitrary cross sections, development of solid state down converters, analysis of coupled TEM line filters with non-adjacent coupling, and the design of waveguide filters. He is now developing millimeter wave transmission components. Member, Sigma Xi.

ROBERT E. ROWE, B.S.E.E., University of Maine, June 1960; M.S.E.E., Northeastern University, June 1962; Bell Telephone Laboratories 1960—. Mr. Rowe worked on designing and developing the TL radio test set and the 100A protection switching system test set. Since 1965 he has worked on developing the TD-3 radio microwave transmitter and receiver. Member, IEEE, Tau Beta Pi, Phi Kappa Phi.

ROBERT C. SALVAGE, Associate degree in EE, 1955, Nework College of Engineering Night School; Western Electric Co., 1953-1955; Bell Telephone Laboratories, 1955—. Mr. Salvage started work in the TD-2 microwave system and has since contributed to designs in the TJ, TL, and TM short haul microwave systems. He is now doing development work on FM terminals.

R. E. SHEEHY, E.E., Assoc., 1960, Wentworth Institute; Bell Telephone Laboratories, 1960—. Mr. Sheehy has been involved in developing various types of electrical networks, and is now designing and developing networks for radio systems.

ROBERT E. SHERMAN, B.S.E.E., 1944, Michigan Technological University; Bell Telephone Laboratories, 1946—. Mr. Sherman's early work consisted of designing high-frequency transmission transformers, including those employed in the original transatlantic cable system. He has supervised groups designing resistors, inductors, and transformers. Since 1963 he has been supervising a group responsible for designing microwave radio repeaters, modulators, and micro-

wave generators for the new TD-3 system. Senior Member IEEE, member, Tau Beta Pi, Eta Kappa Nu.

ROBERT J. SKRABAL, B.S.M.E., 1949, Columbia University; Bell Telephone Laboratories, 1953—. Since 1959, he has been supervising a group concerned with the mechanical and structural design of microwave antennas, towers and other structures for radio relay systems. He had been involved in the design of a 30-foot tropospheric scatter antenna for the DEW Line and in the development of special air conditioning systems for military electronic equipment. Member, A.S.M.E.; registered professional engineer in the State of New York.

CHARLES P. SUSEN, B.E.E., 1953, Rensselaer Polytechnic Institute; Bell Telephone Laboratories, 1953—. Mr. Susen's work has included designing the microwave carrier supply and the broadband receiver for the TH radio system, work on the 100A protection switching system, and designing the TD-3 radio transmitter-receiver test console. He was responsible for TD-2 equipment improvement, and now is designing mastergroup branching and combining units for use with MMX2 on radio.

GERD A. TUCHEN, Vordiplom, M.E., Technische Hochschule Dresden, 1954; M.S.M.E., California Institute of Technology, 1957; M.S.E.E., New York University, 1959; Bell Telephone Laboratories, 1957—. Mr. Tuchen was first engaged in the mechanical design of the SD ocean cable repeater and later in the mechanical design and systems integration of the *Telstar*<sup>®</sup> satellite and other communications satellites. He is supervisor of the Mechanical Group in the Transmission Components Department. Member, A.S.M.E.

JOHN A. WORD, B.S., 1930, University of California (Berkeley); Bell Telephone Laboratories, 1930—. Prior to World War II he worked on the design of toll terminal room equipment. During World War II he worked on sonar, communications counter measures and microwave radio. At present, he supervises a group in the equipment design of short- and long-haul microwave radio systems. Member, Tau Beta Pi and Eta Kappa Nu; associate member, Sigma Xi; senior member, IEEE; registered professional engineer in the State of New York.

