The TH Broadband Radio Transmitter and Receiver

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A general description of the electrical and mechanical features of the TH radio system is given in the two preceding articles.[1,2] *This paper describes the broadband radio receivers and transmitters in detail. Special attention is given to the new features: the receiver modulator with germanium diodes and its associated preamplifier with a noise figure of 10 db; the IF main amplifier in which nearly all adjustments have been eliminated; the amplifier-limiter with unusually low AM/PM conversion; the high-power transmitter modulator using a varactor diode to avoid conversion loss; and finally the 5-watt traveling-wave tube amplifier.*

I. BROADBAND RECEIVER

1.1 *General*

Fig. 1 is a block diagram of the TH radio receiver. The incoming signal from the antenna waveguide system is selected by the channel separation network. After further filtering by the channel bandpass filter, the signal is applied to the receiver modulator. Here it is mixed with a beat oscillator (BO) frequency derived from the microwave carrier supply, to provide an intermediate frequency centered at 74.1 mc.

The IF signal is amplified first by a low noise preamplifier of the cascode type and then by the IF main amplifier. Under normal (no-fade) conditions the IF main amplifier provides about one-half of the gain of the radio repeater. Sufficient additional gain is available, as determined by the automatic gain control (AGC) circuit, to keep the receiver output level constant for repeater input signal levels ranging from 5 db above to 25 db below the normal. For short sections, of low path loss, a pad of suitable value (not shown in Fig. 1) is used between the preamplifier and main amplifier to keep the AGC circuit properly centered. Equalization follows the IF main amplifier. All receivers have a basic equalizer which compensates for gain and delay departures from

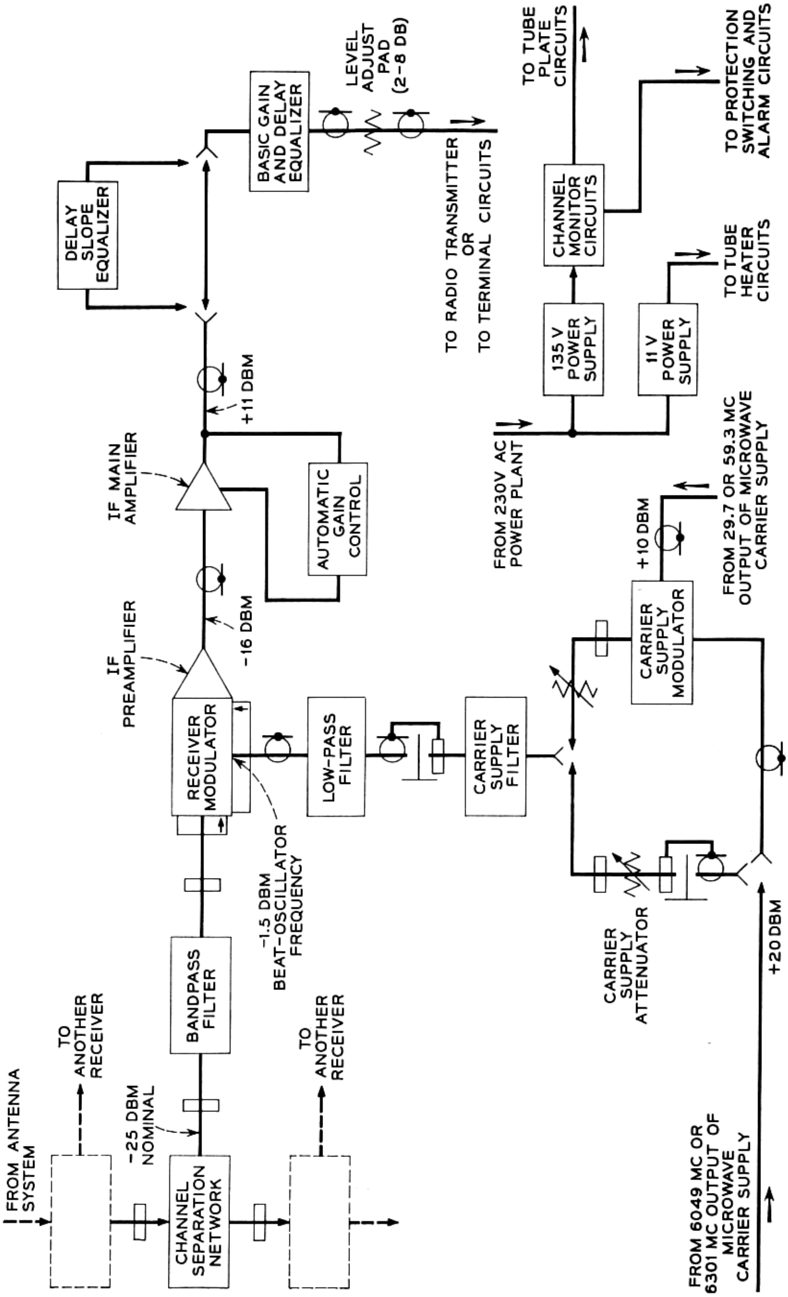

Fig. 1 — Block diagram of TH radio receiver.

ideal transmission for both the transmitter and receiver of the section. A delay slope equalizer is inserted between the IF main amplifier and the basic equalizer when required to correct for delay slope distortion which has accumulated in several repeaters. Not shown on Fig. 1 is a 93.8-mc rejection (trap) filter, connected between preamplifier and main amplifier and used only on channels 11 or 21 and 18 or 28; this filter blocks spurious transmission of the adjacent auxiliary radio channel through the broadband channel.

The nominal microwave signal power at the input to the channel bandpass filter is -25 dbm* and the output of the IF main amplifier is set for a constant level of approximately $+11$ dbm. The pad at the receiver output is chosen to give the proper power input at the connecting equipment.

The BO frequency of 6049 mc (or 6301 mc) is used directly for the receiver modulators of channels 12 and 17 (or 22 and 27). For the other channels, a carrier supply modulator is required as part of the radio receiver. This modulator shifts the 6049 mc (or 6301 mc) by either 29.7 mc or 59.3 mc. The carrier supply filter at the output of the modulator selects the desired BO frequency. However, this filter does not provide attenuation to the second harmonic output at 12 kmc from the modulator, which was found to be a source of interference. The low pass filter, following the carrier supply filter, attenuates these undesired frequencies. An attenuator is used to set the BO input power to the receiver modulator to the correct value for optimum noise figure of the crystals.

Indications of lack of input signal or malfunction of equipment are provided by the channel monitor circuit. This circuit informs the protection switching and alarm circuits when the input signal falls to approximately -52 dbm.

Fig. 2 is a photograph of the radio receiver. The multiposition switch and the connector immediately below it are used with an external meter to check the condition of the receiver's electron tubes and diodes while the receiver is in use.

1.2 *Receiver Modulator and IF Preamplifier*

A photograph of the receiver modulator and IF preamplifier is shown in Fig. 3. Since the signal level is at a minimum at this point in the receiver, the best possible signal-to-noise ratio is required, consistent with good transmission. The receiver modulator is built in a dual waveguide

* This value was chosen early in the TH development for design purposes. As discussed in Ref. 1, the currently accepted value is -27 dbm, based on an rms section loss of 64 db.

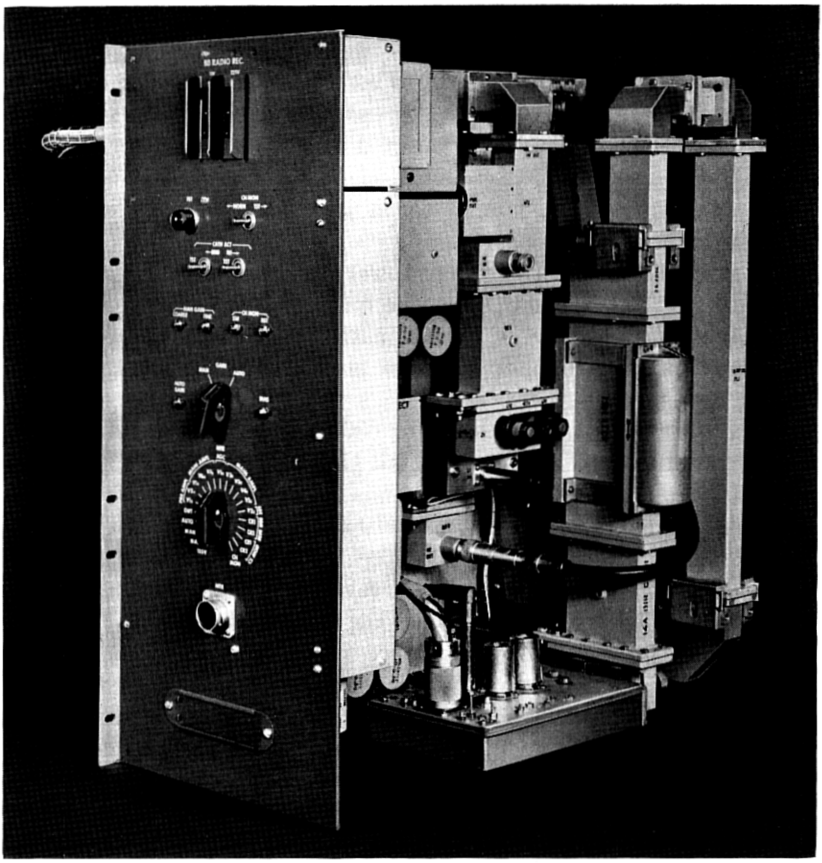

Fig. 2 — The broadband radio receiver.

configuration with four elements, the diode mount, the short slot hybrid junction, the dual isolator and the input transducer. It uses two germanium crystals in a balanced modulator circuit designed for optimum noise performance. The balanced circuit suppresses any amplitude modulation noise present on the BO input. The dual isolators absorb spurious frequencies generated in the modulator and provide a good input impedance. The IF preamplifier is a two-stage low noise amplifier of the cascode type using triode electron tubes. The noise figure for a typical receiver modulator and IF preamplifier with new crystals and tubes is about 10 db. The transmission is essentially flat over the band from 64 mc to 84 mc, dropping off by about 0.3 db at 58 mc and 90 mc.

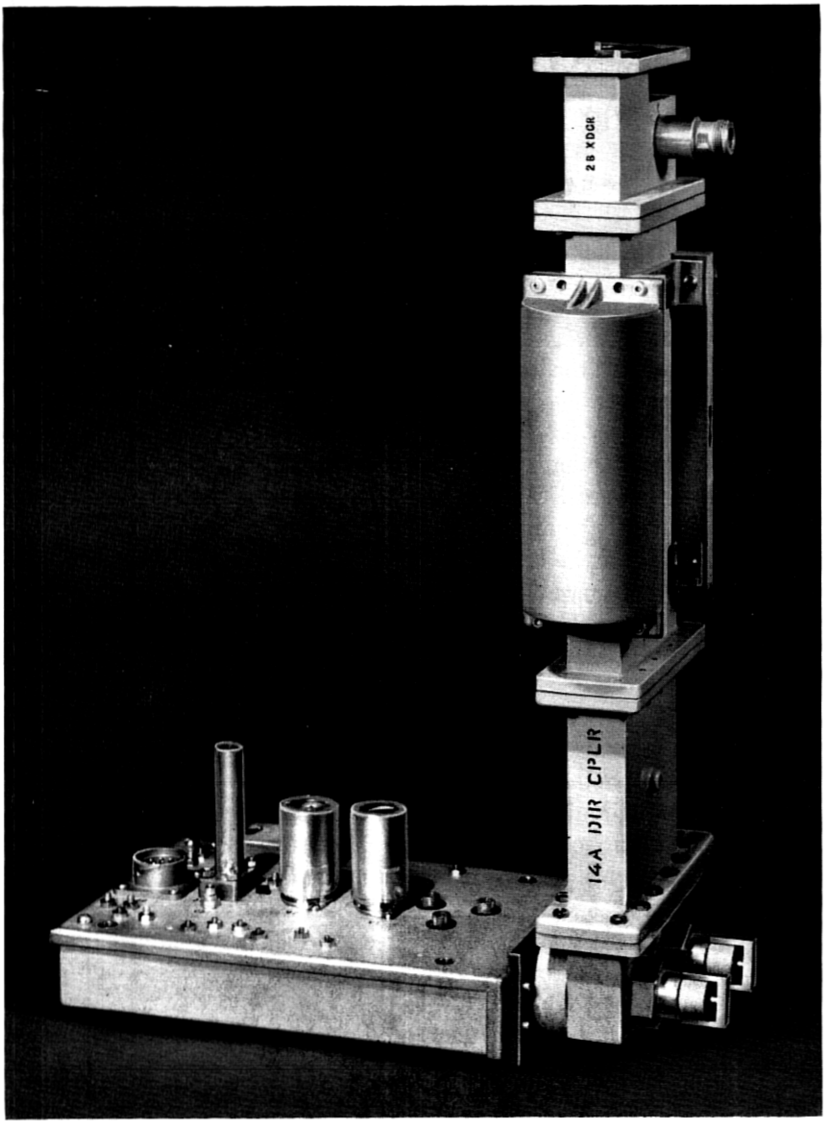

Fig. 3 — The receiver modulator (right) and IF preamplifier (left).

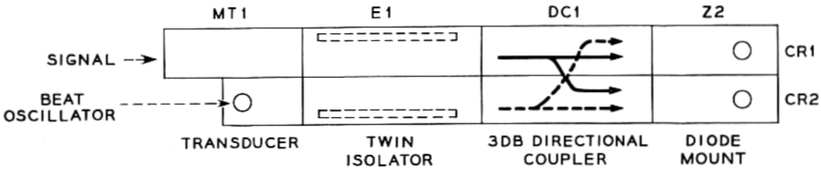

Fig. 4 — Simplified block schematic of the receiver modulator.

1.3 *Receiver Modulator*

A simplified block schematic of the receiver modulator is given in Fig. 4. The received signal is coupled through the transducer and one side of the twin isolator into the short slot junction which forms a 3-db directional coupler. The BO frequency is coupled to the directional coupler through the coaxial-to-waveguide portion of the input transducer and the other half of the twin isolator. The directional coupler divides the signal and BO powers equally and applies them to the crystals CR1 and CR2 of the diode mount.

A line drawing of the diode mount is shown in Fig. 5. The configuration is that of two coaxial-to-waveguide transducers placed side by side and sharing a common E plane wall, with the diodes in the transition region, i.e., partly in the waveguide and partly in the coaxial cavity. The depth that the coaxial sleeve penetrates into the waveguide controls the coupling between the coaxial line and the waveguide, thus matching the resistive portion of the diode impedance to that of the waveguide. The reactance is tuned out by the properly positioned fixed waveguide short circuit behind the junction and by adjustment of the length of the coaxial cavity, which also serves to compensate for differences between diodes. These latter adjustments, called RF TUNER 1 and RF TUNER 2 on Fig. 5, are the only mechanical adjustments.

The modulator uses germanium diodes, which are matched pairs of 1N263-type diodes. Optimum noise figure is obtained with a BO power of about −4.5 dbm and a fixed bias of about 0.3 volt from a 200-ohm source applied to each diode. The modulator has a conversion loss of approximately 6 db.

The IF circuit is designed to provide adequate decoupling to microwave frequencies and at the same time minimum capacitance. The decoupling is obtained through the use of an RF choke in the IF output lead. The choke consists of two one-quarter wavelength radial lines. One of these lines is resonant in each half of the 5925-mc to 6425-mc band. This gives a minimum insertion loss of 30 db over the band. The position of the choke along the IF output line is so chosen that a low impedance is presented at the waveguide wall. Through the use of this construction, the total capacitance of the IF output terminals in parallel has been kept to 12 mmfd. Minimum capacitance is necessary because of the broad IF bandwidth to be transmitted.

The short-slot hybrid junction consists of two waveguides sharing a

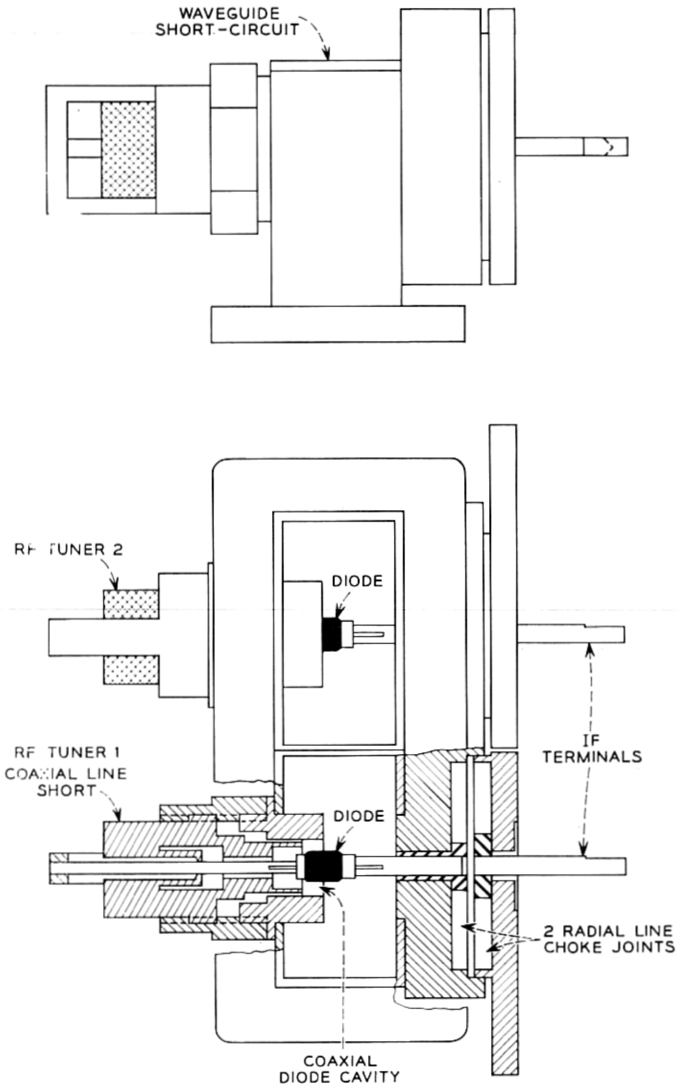

Fig. 5 — The diode mount.

common E plane wall in which there is a coupling slot to form a hybrid junction.[3] Power applied to one input arm divides equally between the two opposite output arms with no coupling to the third arm. Instead of the 0° or 180° phase difference between output arms as found in the hybrid T, the short slot coupler has a 90° phase difference. This causes the reflected signal power from the diode modulator to appear only in the BO arm rather than in the signal arm. Typical couplers have return losses greater than 24 db, a power split equal within 0.2 db, and a directivity of 23 db or better over the 500-mc band.

The twin isolator consists of two similar isolators mounted side by side with the waveguides having one common E plane wall. The isolators present a low loss to signals proceeding toward the modulator, and high loss to reflected energy.

One of the products generated by the modulator falls at twice the BO frequency minus the signal frequency. If this product is reflected back to the modulator in the correct phase, an improvement in the noise figure of up to one-half db can be obtained. However, the disadvantages of using this image frequency energy are more important. First, a different mechanical spacing of the channel bandpass filter would be required for each channel, and second, the variation in phase would cause the IF output impedance to vary over the band, which would adversely affect the transmission performance. An isolator, therefore, is provided in the BO arm to absorb the image as well as any signal energy reflected from the modulator.

1.4 *IF Preamplifier*

Fig. 6 is a schematic of the IF preamplifier circuit. It uses two 417A triodes connected in a cascode circuit between the IF terminals of the balanced modulator and the 75-ohm output coaxial line. The gain of the preamplifier is typically 15 db. As shown on Fig. 3, it is assembled in a shielded chassis about five inches wide, eight inches long and $1\frac{1}{2}$ inches deep.

The preamplifier is designed to provide a transmission characteristic which is stable over the life of the tubes, along with optimum noise performance. Factory and field adjustments have been minimized by the use of compensation circuits and by control of wiring in the transmission path. The compensation circuits are in the cathode circuit of the grounded-cathode first stage and in the grid circuit of the grounded-grid second stage of the cascode.

For a stable transmission characteristic, the input impedance of a tube, which is the termination of an interstage network, must be inde-

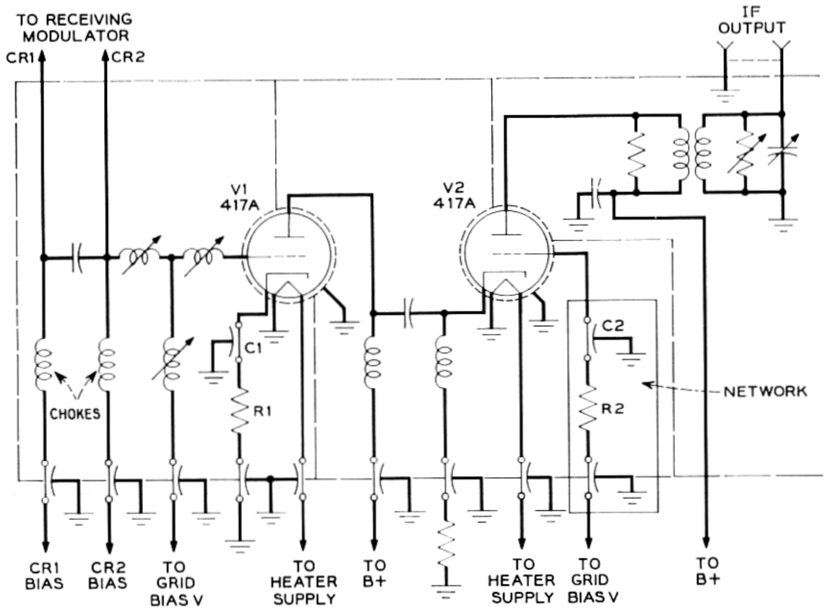

Fig. 6 — Simplified schematic of the IF preamplifier.

pendent of the voltage gain of the stage. For a triode with resistive load, the input capacitance C_{in} is given by

$$C_{\text{in}} = C_{\text{gk}} + (1 + A)C_{\text{gp}}$$

where C_{gk} is the hot grid-to-cathode capacitance, and A is the voltage amplification of the stage; C_{gp} is the grid-to-plate capacitance. Thus the input capacitance is a function of the voltage gain of the stage. This change in capacitance is commonly called the Miller effect. In many circuits this effect is cancelled by neutralizing techniques, one of which is to parallel the grid-to-plate capacitance with a suitable coil to form a parallel resonant circuit. Neutralization of this type could not be used because of the wide frequency band of 58 mc to 90 mc. However, resistance in the cathode circuit produces negative input capacitance, which also changes with the gain of the stage. By proper choice of resistor value (R1 of Fig. 6) the input capacitance is reduced and made, for practical purposes, independent of tube gain.

The lower curve of Fig. 7 gives the input resistance of a 417A in a grounded-cathode stage in which the connection to ground of the cathode pin of the tube socket is made as short as possible. This input resistive loading is due to cathode lead inductance and varies inversely with both

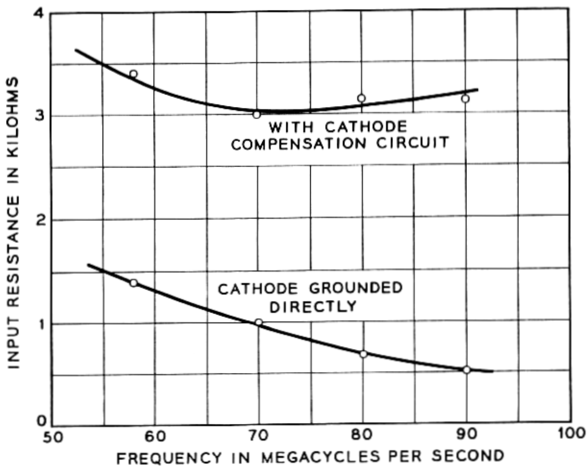

Fig. 7 — Input resistance of a W.E. 417A triode: cathode grounded directly and with cathode compensation circuit.

the square of the frequency and the operating gain. These variations in the input resistance make it a far from ideal termination for an interstage network. The inductance of the series resistor added for capacitance compensation makes the input resistance even lower. However, a capacitor in parallel with this resistor (C1 of Fig. 6) forms a broad series resonant circuit with the cathode lead inductance, and reduces the net inductance. The resultant input resistance, with the compensation circuit added, is plotted in the upper curve of Fig. 7.

Stability of the input impedance to the grounded-grid stage is obtained in a similar way by placing a small resistor and capacitor (R2 and C2 of Fig. 6) in parallel in the grid lead.

The coupling network between the receiver modulator and the grounded-cathode stage is an adjustable three-coil network in a T configuration. It is mismatched to obtain optimum noise performance, and is designed to operate with the IF impedance of the modulator as one termination and with the input impedance to the tube as the other. The only important resistive loading is applied by the real component of the IF impedance of the crystals.

The interstage network between the tubes appears as an extremely simple one on paper. However, stray impedances play an important role. The lead inductance of the connection between plate and cathode, the series blocking capacitor, the output capacitance of V1, and the input capacitance of V2 combine to form a filter with adequate bandpass. This network is also misterminated, to obtain an optimum noise figure from the grounded-grid stage. Plate voltage for V1 and cathode voltage for V2 are applied through high impedance chokes. No adjustments are required since the lead inductance is carefully controlled during manufacture.

The output circuit uses a double-tuned transformer of the type discussed more fully in the next section of this paper. One capacitor adjustment is required to compensate for circuit variations. The output impedance is 75 ohms, to better than 20-db return loss.

1.5 *IF Main Amplifier*

The IF main amplifier operates at a gain of 27 db for the nominal receiver input signal level of −25 dbm, and has a maximum gain with new tubes of about 60 db. The gain-frequency characteristic for a typical IF amplifier at nominal gain setting is essentially flat over the 64-mc to 84-mc band, dropping at 58 mc and 90 mc by about 0.6 db. Changes in the transmission characteristic with change in operating gain of 25 db are typically within 0.5 db.

A photograph of the amplifier is shown in Fig. 8. The amplifier is assembled in a shielded chassis approximately $4\frac{1}{2}$ inches wide, $1\frac{1}{2}$ inches deep and 20 inches long. The electron tubes are covered by a common electrostatic shield. Cooling air from the bay is supplied to the tubes by means of a conduit mounted on the shield. Circuit elements are

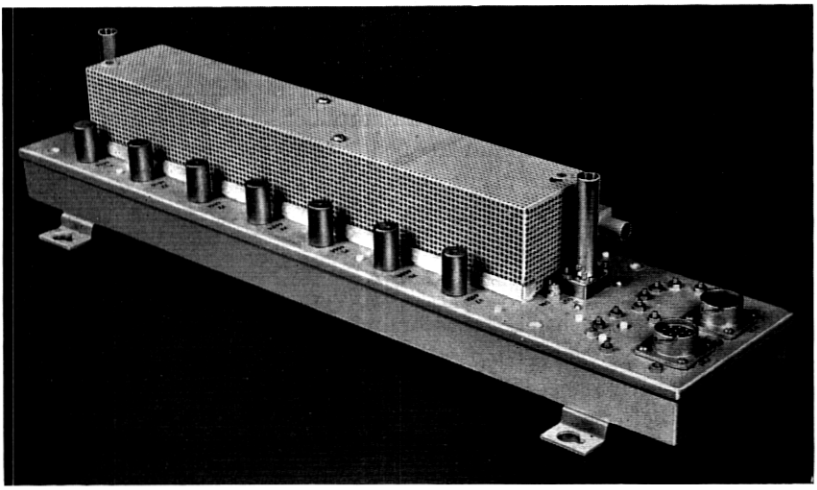

Fig. 8 — The IF main amplifier.

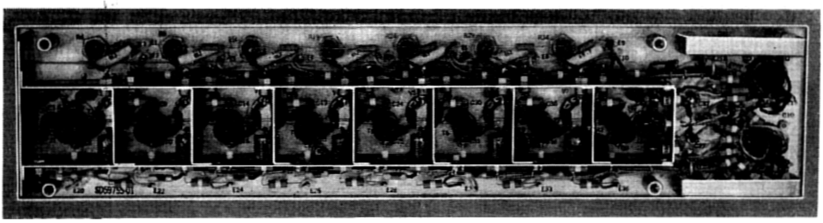

Fig. 9 — IF main amplifier, rear view with cover removed.

mounted on the underside of the chassis with each stage individually shielded, as shown by Fig. 9. Multi-pin connectors, which carry the necessary plate and grid supply voltages, also connect the individual cathode voltages to the control panel of the receiver for test purposes.

A simplified schematic of the amplifier is shown in Fig. 10. It uses a new high-performance tetrode, the Western Electric 448A, specially designed for this application.[4] Seven stages of amplification are used, the first, sixth, and seventh being fixed gain stages, the remaining four being variable gain stages. Fixed gain is used on the first stage to obtain the most stable input impedance, on the six and seventh to maintain the required power output over the range of operating gain. Connected across the output of the amplifier is a monitoring circuit which provides a dc voltage, dependent upon the signal level, to the automatic gain control (AGC) circuit. The latter controls the grid bias of the variable gain stages to maintain the output power constant as the input to the receiver changes due to fading, etc. By careful control of the tube parameters and of wiring during manufacture, by the use of compensation circuits, and by the use of an interstage transformer of a new type with a high coefficient of coupling, an amplifier has been achieved which requires no adjustment of gain-frequency characteristic when tubes are replaced. The input and output circuits each have two variable capacitors to permit optimum adjustment of impedance.

1.6 *Electron Tube Circuit*

The 448A tetrode has a transconductance of approximately 31,000 micromhos. It has very close spacings between the control grid and cathode and between the control grid and screen grid, and relatively large spacing between the screen grid and plate. This wide spacing makes possible a low in-circuit output capacitance, which averages 5 mmfd. The close control grid-to-cathode spacing makes negligible the effect of transit time on input impedance for frequencies up to at least 100 mc.

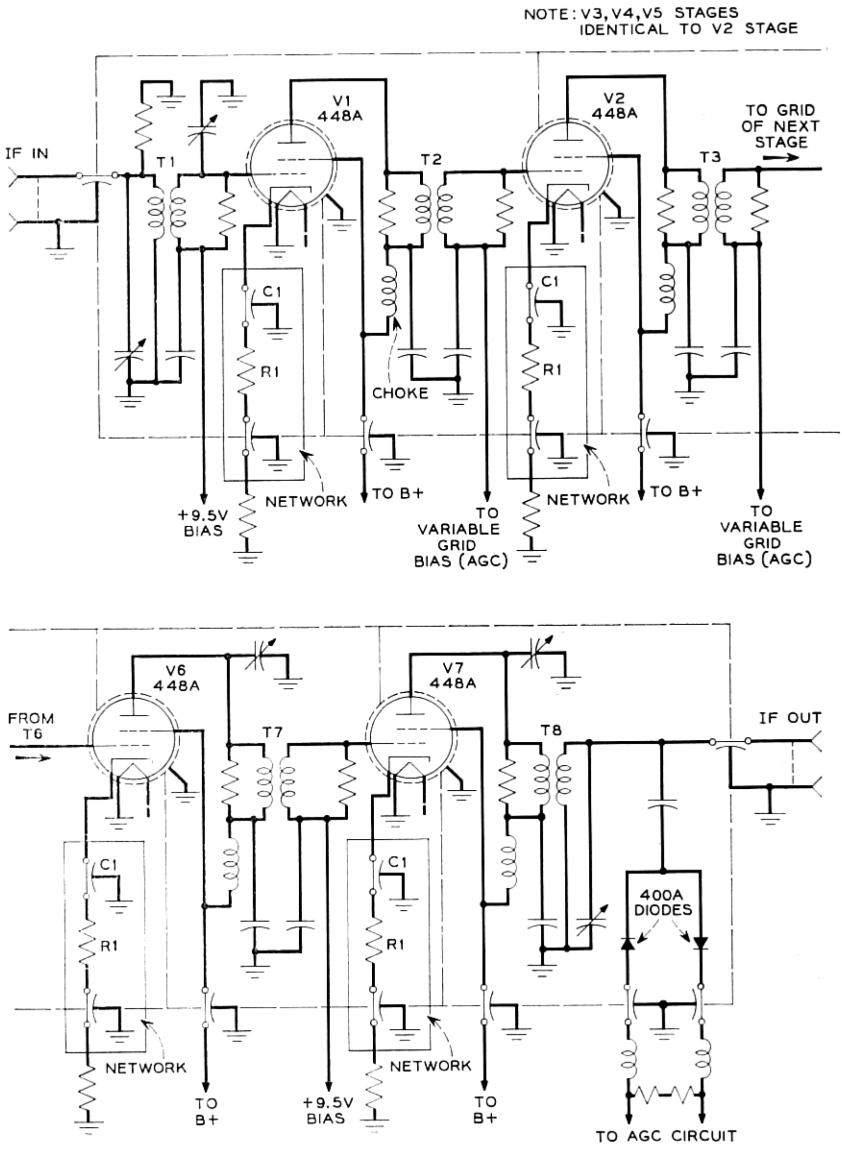

Fig. 10 — Simplified schematic of the IF main amplifier.

The input capacitance as measured in the circuit averages 23 mmfd. This tube therefore has a high gain-bandwidth factor, not all of which is realized in the amplifier since negative feedback is introduced to improve transmission stability.

The termination of an input or interstage network is the input impedance of the tube. For high transconductance tubes operating in the 50-mc to 100-mc region, this impedance is far from ideal. Cathode lead inductance gives rise to a positive input resistive loading, while screen lead inductance results in a negative input resistive loading. Both vary inversely with the square of the frequency and the operating gain. Furthermore, the input capacitance is a function of the total space current flowing past the control grid. Since the gain of several of the stages must be changed to provide AGC, the effect of these non-constant impedances must be minimized so that transmission through the stages will not be a function of the operating gain.

The required transmission stability is obtained by the use of compensation circuits in the cathode circuits of the tubes as described above for the preamplifier. The input capacitance of the 448A, as determined by measurement, is very nearly a linear function of transconductance over the operating current range for the gain-controlled stages. Resistance in the cathode circuit, while decreasing the stage gain, introduces a negative capacitance in the grid circuit, which also varies linearly with transconductance. With a suitable resistor in the cathode circuit, the input capacitance is stabilized at 23 mmfd over the operating gain range.

The more important contributor to the input resistive loading is the cathode lead inductance. To minimize this inductance, three cathode leads connected to widely separated base pins are provided in the tube. These are connected externally by a low inductance path consisting of a brass plate which joins together the appropriate tube socket pins. This results in an in-circuit cathode lead inductance of less than 0.01 microhenry. However, even this low inductance gives rise to an input resistance of a few hundred ohms at 90 mc at maximum gain. Fig. 11 shows measured values of the input resistance with the shorting plate directly grounded, as a function of frequency and cathode current. This input loading is comparable to the terminating resistance of the interstage on the grid side (approximately 110 ohms).

The resistor which provides compensation for input capacitance changes adds inductance to the cathode lead. However, by proper choice of a capacitor in parallel with the resistor, a compensation circuit was designed which reduces the effective cathode-to-ground inductance to about 0.005 microhenry over the 50-mc to 90-mc band.

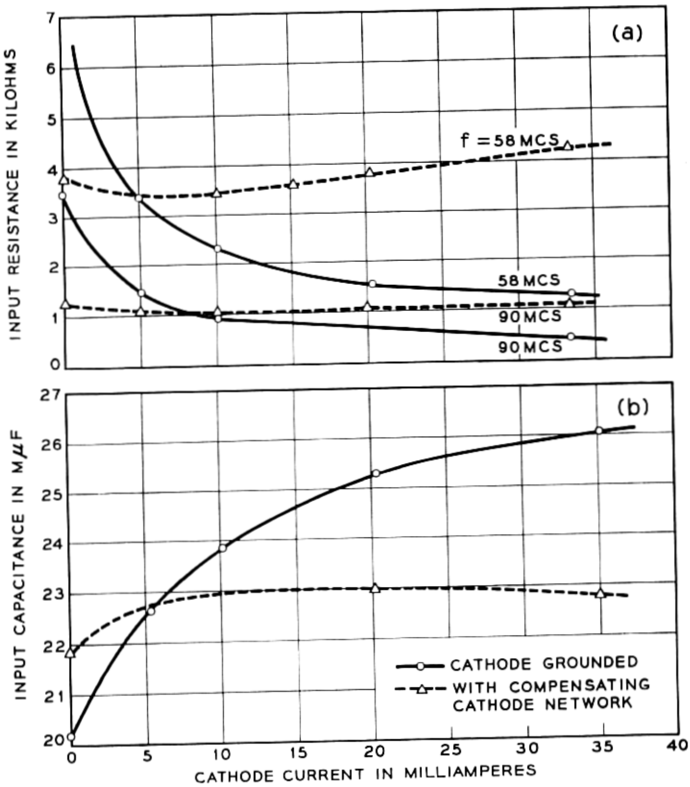

Fig. 11 — Input impedance of the W.E. 448A tetrode.

To ensure adequate control of leads in this very critical cathode circuit, the combination of the resistor and parallel capacitor along with the bypass capacitor was designed into a closely controlled network assembly. This network (designated 491A) is visible in the photograph of the wiring side of the chassis, Fig. 9.

Screen grid lead inductance contributes a negative resistance across the input terminals of the tube. The 448A uses a single internal screen lead. The external lead to the bypassing point is made intentionally long. The resulting negative resistance cancels to a large extent the loading due to the remaining cathode lead inductance. Fig. 11 shows the components of input impedance with the final compensation. The net effect is that of 23 mmfd shunted by approximately 1000 ohms. For all practical purposes, the resistance is independent of the value of cathode current, and for all frequencies is at least ten times that of the interstage terminating resistance.

All stages of the amplifier use dc feedback provided by a combination of self bias supplied by the cathode resistors and positive grid voltage obtained from a stabilized source. This minimizes gain variations from tube to tube and due to tube aging. The variable gain stages have less dc feedback than the fixed gain stages because of limitations imposed by the AGC circuit.

1.7 *Coupling Networks*

Three different networks are used in the amplifier: a terminated input network which couples from the 75-ohm input coaxial to the grid of the first tube, an interstage network, and a terminated output network which couples the plate of the last tube to the 75-ohm output coaxial line. Studies of various types of networks indicated that to meet the exacting requirements of minimum adjustment and permissible transmission deviations of a few hundredths of a db per stage, terminated networks would have to be used throughout. Of those studied, a design described by Rideout[5] which gives a slightly over-coupled characteristic and has the advantage of simplicity in design, is used. For the bandwidth required and the ratio of the terminating capacitances (5 mmfd for the tube output, 23 mmfd for the tube input) the network cannot be realized with a simple T or π configuration. However, it is realizable with either a tapped coil or a two-winding transformer. The disadvantage of the tapped coil is that a series blocking capacitor is required. For this reason the two-winding type was chosen.

The transformer is encased in an epoxy resin cylindrical block approximately $\frac{5}{8}$ inch in diameter by $1\frac{1}{8}$ inches high. The external view and stages in assembly showing the internal construction of a typical interstage transformer are shown in Fig. 12. Connections to the windings are made through four radial wires spaced at 90° intervals. The two windings are in separate grooves of a double thread cut in a plastic tube. The small mechanical separation between windings achieved in this way makes possible the required high coefficient of coupling without the use of a ferrite core. Because of the different impedance ratios required, the input, output and interstage transformers, although of the same general design, differ in a number of ways, such as the length-to-width ratio of the plastic tube, the number of turns and pitch of the windings, and the size of wire used. Externally, however, they are all similar in appearance. To achieve adequate uniformity in electrical characteristics, the individual inductances are held to less than ± 3 per cent of the nominal design value. Reproducibility of the transformers therefore depends upon very close control of all dimensions. The basic

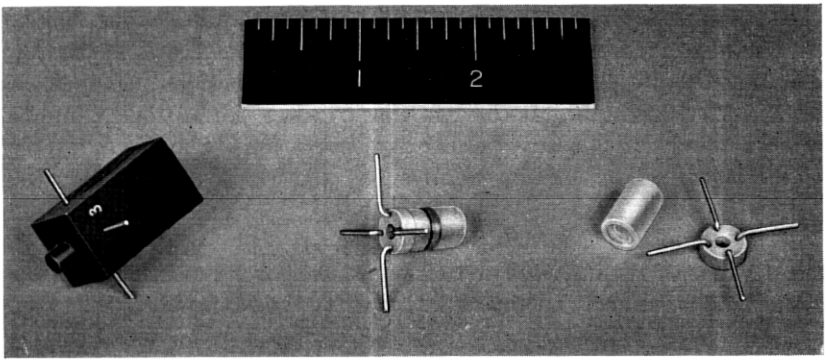

Fig. 12 — Typical interstage transformer.

transformer design has been so successful that it is used not only in the IF amplifier but throughout TH in broadband IF circuits.

To meet the objective of a minimum of factory and maintenance adjustments, a high degree of control over each of the components is required. Equally important is that their relationship to one another when installed in the amplifier remain the same from unit to unit. For this reason, the transformers are located in position by means of a keyed hole, and the interstage loading resistors are located uniformly. All components not directly in the IF transmission path are located outside the shielded area.

To minimize reflections (echoes) a good impedance match to the 75-ohm coaxial line is required at input and output. Typically the amplifiers have a minimum of 27 db return loss over the 64-mc to 84-mc band. To provide this performance consistently, two pairs of variable capacitors are added to correct for circuit variations. One additional variable capacitor is used in the sixth interstage. This capacitor is adjusted at the factory for optimum gain-frequency characteristic.

1.8 *Automatic Gain Control and Channel Monitor Circuits*

The AGC circuit is straightforward. A voltage doubler rectifier circuit at the output of the IF main amplifier monitors the signal power. Its output voltage, together with an adjustable reference voltage, is applied to the grid of a stabilized dc amplifier.[6] The reference voltage effectively sets the IF power at the main amplifier output. The amplified dc signal is applied to a cathode follower which translates it to the correct voltage level for control of the grid bias of the tubes in the IF amplifier. Thus an increase in IF power at the amplifier output results in an increase in negative grid bias to the gain-controlled stages. A series resistance and shunt capacitance in the bias lead set the time constant and hence the response of the circuit to sudden IF signal changes. Diodes in the bias load limit the voltage rise when the IF main amplifier input is removed. The use of grid bias gain control is feasible as a result of the transmission stability obtained by the use of the compensating circuits in the cathode circuits of the tubes.

If the input signal is suddenly reduced by 27 db, approximately 30 milliseconds is required for the amplifier output to return to normal. This response time is adequate to follow the most rapid signal fading expected.

The channel monitor circuit initiates a local alarm and originates an order to the protection switching system when a deep fade of the incoming signal occurs. A fade results in an increase in the current drawn by the IF amplifier tubes. The increase in the voltage drop that this causes across a resistor in series with the plate supply is used to trigger a bistable transistor circuit. This in turn controls a relay to operate the alarm and call for a protective switch.

II. BROADBAND TRANSMITTER

2.1 *General*

The TH radio transmitter is shown in block schematic form in Fig. 13. The IF input signal from the radio receiver or terminal equipment is first applied to the amplifier-limiter. The limiter effectively removes any amplitude modulation that might be present in the input signal and which could be converted into phase modulation by the traveling-wave tube (TWT) amplifier. Such modulation would appear in the system as unequalizable distortion. Associated with the amplifier-limiter is the carrier resupply circuit, which provides a local carrier source if a deep fade or an equipment failure occurs.

The IF signal from the limiter output passes with negligible loss through the carrier resupply to the transmitter modulator. There, after passing through a buffer amplifier, the signal is mixed with the BO frequency in a low-loss balanced modulator. After selection of the appropriate sideband by the channel bandpass filter, the signal is applied to the transmitter amplifier. This TWT amplifier provides about 32-db gain and an output power of +37 dbm (5 watts). A filter following the amplifier attenuates its second harmonic output an additional 25 db. Next is an isolator to provide an adequate termination for the channel separation network. After passing through the power output monitor,

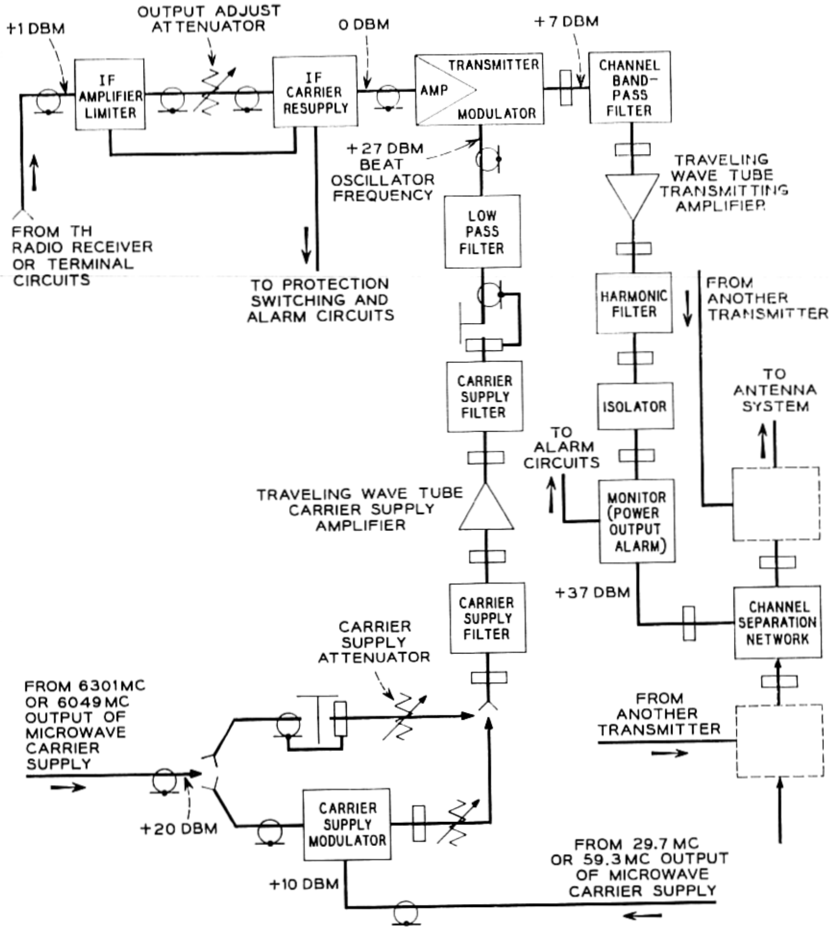

Fig. 13 — Block schematic of the TH radio transmitter.

the signal is fed to the channel separation network, which combines it with the outputs of other transmitters for transmission to the antenna system.

The power monitor circuit gives an indication to the repeater alarm circuits if the power drops more than 4.5 db. It consists of a 23-db directional coupler in the signal path feeding a crystal detector whose output current controls a meter-type relay. The meter also provides a visual indication of the output power.

The BO frequency for the transmitter modulator is obtained from the microwave carrier supply, as already described for the receiver.

However, here a relatively high power +27 dbm (0.5 watt) is required. As shown in Fig. 13, a second TWT amplifier, of the same type as used in the transmission path, is used to provide this power. Although this single-frequency, low-power application of the transmitting amplifier appears inefficient, it avoids introducing a second type of microwave amplifier into the repeater and permits the use of a common power supply for both amplifiers.

The carrier supply filter is divided into two sections, one ahead and one after the TWT. The combination provides the same amount of filtering as in the receiver. The section at the input of the TWT attenuates the unwanted sideband sufficiently to prevent cross-modulation in the TWT. The section at the output attenuates noise at signal frequency generated by the TWT amplifier. The low-pass filter attenuates second harmonic energy, as in the receiver.

Fig. 14 is a photograph of the radio transmitter. The main items visible on the front panel are the access doors to the traveling-wave tubes,

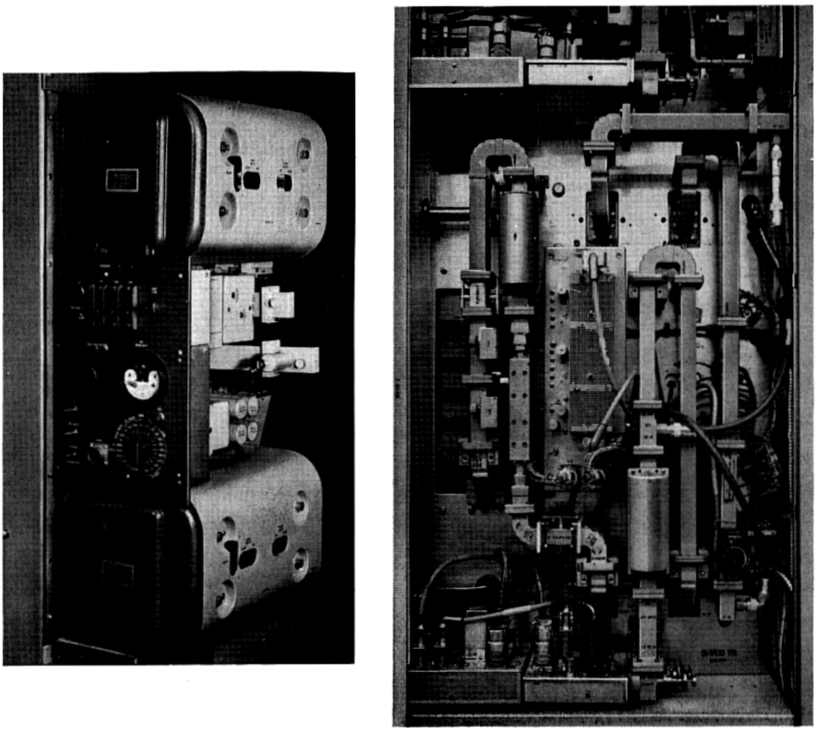

Fig. 14 — The broadband radio transmitter, and left side view.

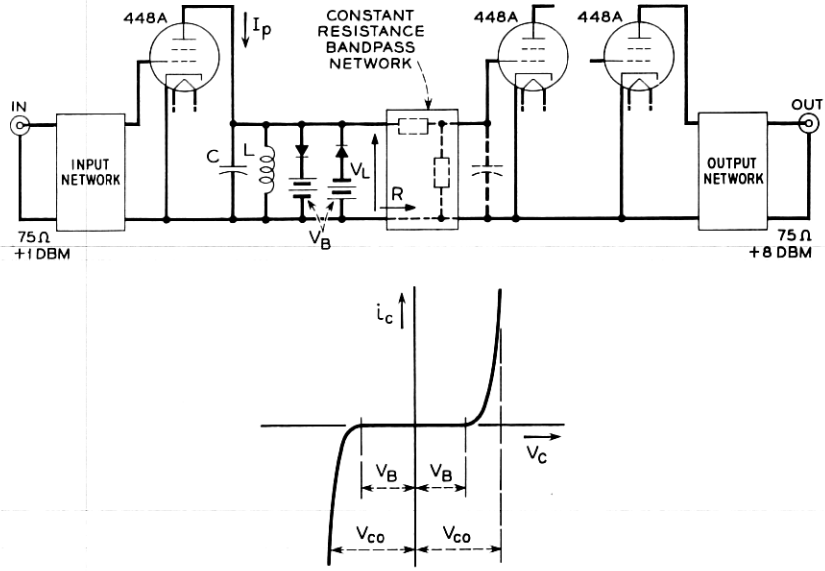

Fig. 15 — Simplified schematic of the IF amplifier-limiter, with typical voltage-current dc clipping characteristic.

the attenuator used to set the output level to +37 dbm, the output monitor meter, and the multi-position meter switch.

2.2 *Amplifier-Limiter*

A simplified schematic of the amplifier-limiter is shown in Fig. 15 and a photograph in Fig. 16. It uses three 448A high-transconductance tubes with compensation circuits identical to those in the IF main amplifier. To provide the required limiting, two 431A gold-bonded germanium diodes are used in each of the two identical interstages. The interstages are designed so as to provide maximum compression with a minimum of amplitude modulation to phase modulation (AM/PM) conversion. For this reason, the diodes are arranged in a symmetrical voltage clipping circuit in a parallel RLC circuit resonant at approximately midband. The diodes work into a constant resistance bandpass network. This network transforms the capacitance at the input to the following tube to a resistance of about 150 ohms in shunt with a small capacitance across the diodes. Harmonic frequencies generated by the diodes lie outside the bandpass of the network and are absorbed. The input and output networks are similar to those used in the IF main amplifier. Diode monitors (not shown) which produce a voltage proportional to

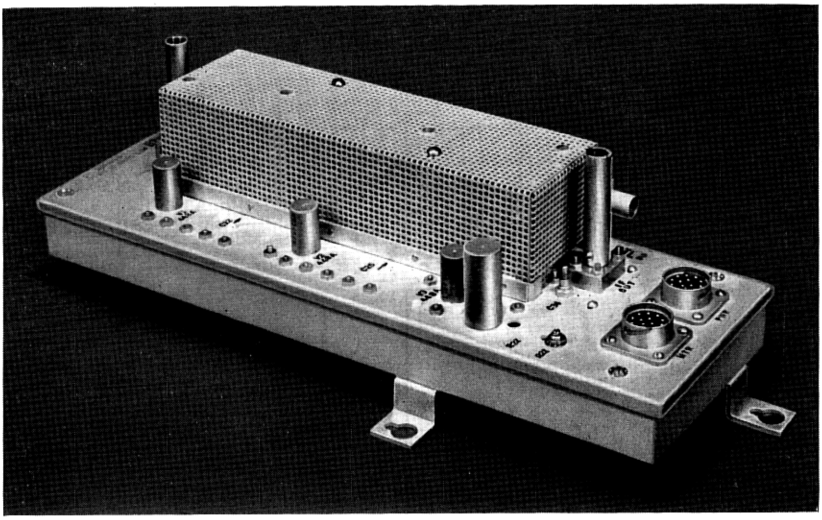

Fig. 16 — The IF amplifier-limiter.

the signal level are used at both the input and output. The voltage from the input monitor is used to operate the carrier resupply circuit, which is switched on when the signal power at this point drops 3 db. Both monitors are connected to the metering circuits to provide a convenient means of level checking.

The gain through the amplifier-limiter is normally about 7 db. When the incoming carrier is lost, the limiter acts as a linear amplifier with about 22 db of gain. Operation of the carrier resupply circuit provides a bias to the three tubes which reduces this gain to a loss of 2 db.

2.3 *Interstage Networks*

The design of the interstage networks is a compromise among maximum compression, minimum AM/PM conversion and a suitable gain-frequency characteristic.

The compression, C, is defined as

$$C = \frac{dV_i/V_i}{dV_0/V_0} \tag{1}$$

where V_i is the input voltage and V_0 is the output voltage. Compression is expressed in db by taking 20 log C. Fig. 15 shows a typical voltage-current characteristic of a symmetrical voltage clipping circuit. The factors affecting the compression of such a circuit are the voltage-current characteristic of the diodes, the peak current supplied by the source, and the clipping level, which is dependent upon the bias supplied to the diodes.

The 431A gold-bonded germanium diode has a low forward resistance and good high-frequency characteristics. At 10 ma forward current, the ac resistance is between 4 and 5 ohms. A nearly ideal characteristic can be obtained with a low bias voltage of approximately 0.7 volt.

The dc clipping characteristic shown in Fig. 15 can be approximated by a power series containing odd powers only and for practical purposes, in the clipping region, can be reduced to one term, namely $i_c \sim V_c^n$, where n is an odd integer. The exponent n increases with diode bias, i.e. as the characteristic becomes more nearly ideal. The maximum compression, C, of the double-diode circuit, as calculated by nonlinear circuit analysis, is given by

$$C = n. \tag{2}$$

A value of $n = 15$ gives the best fit for the characteristic using 431A diodes. This means that 23.5 db of compression should be available from this stage. However, actual measurements show only 16 db. The main reason for this discrepancy lies in the over-simplified model for the diodes. Minority carrier storage in the diodes prevents current and voltage from following the dc diode characteristic at high speeds. Lifetime of the carrier in the 431A diode is relatively high, between one and two microseconds. Other types of diodes, such as the microwave diodes used in down converters, have negligible storage effects, but the amount of limiting attainable is small because of the rather high forward resistance of these diodes. High-speed computer diodes also exhibit very small storage effects (lifetimes of one to two millimicroseconds). Experiments with such diodes showed that higher compression could be obtained, but the capacitance of the diodes was rather high and AM/PM conversion, as explained later, would have been adversely affected.

In the Appendix, a simplified analysis of the circuit of Fig. 15 shows that the AM/PM conversion, P, is given by

$$P = 106 \frac{\Delta f c V_{co}}{I_p} \text{ degrees/db} \tag{3}$$

where

Δf = frequency deviation of the signal from center frequency,
V_{co} = voltage at which clipping occurs,
c = shunt capacitance across the diodes, and
I_p = peak current into the diodes.

This expression shows that the capacitance and the clipping voltage should be small while the peak current delivered should be large, to keep the AM/PM conversion small. Exact tuning of the limiter circuit to the incoming carrier ($\Delta f = 0$) would seem to eliminate AM/PM conversion. However, a more exact solution, taking into account higher harmonics, indicates that there exists no frequency at which P is zero for all values of I_p. Measurements taken on the complete amplifier show this. (It should be noted that most of the AM/PM conversion is produced in the first stage, since the compression of the first stage reduces the AM input to the second stage.) The clipping voltage is kept small by the use of the low forward resistance 431A diode, and a high peak current is obtained by using a high transconductance tube, the 448A, as the driver. The shunt capacitance across the diodes is kept low by the network between the diodes and the following tube. On the tube side, the network is terminated by the input capacitance of the tube, namely 23 mmfd. To the diodes the network presents a resistance of 150 ohms, constant within 10 per cent, and a parallel capacitance of about 1 mmfd over a frequency range of 40 mc to 200 mc. Since the plate capacitance of the 448A is only about 2 mmfd, the use of a con-

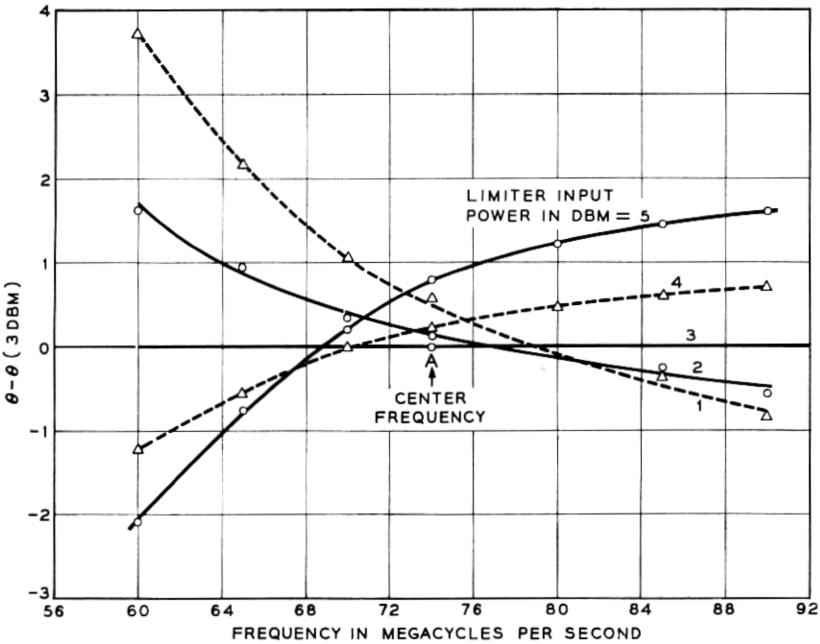

Fig. 17 — AM to PM conversion as a function of frequency with limiter input power as a parameter.

stant resistance network to reduce this capacitance is not justified. The capacitance of the two 431A diodes is only 0.9 mmfd when measured with small signals at 1 volt reverse bias. However, when measured in the circuit under heavy clipping, it is as high as 7 mmfd. This effect is again caused by minority carrier storage effects. The total capacitance across the circuit at high levels is approximately 10 mmfd. With $V_{co} = 1.1$ volts, $I_p = 15$ ma, substitution in (3) gives

$$P = 0.078\ \Delta f \text{ deg/db for } f \text{ in mc.} \tag{4}$$

For $\Delta f = \pm 15$ mc, the expected P would therefore be $0.078 \times 30 = 2.3$ deg/db. This can be compared with the measured results shown in Fig. 17. The curves show a total phase change from 60 mc to 90 mc of 1.9° at +4 dbm input power and 2.1° at +2 dbm input power, compared to a +3 dbm reference power. The very simple limiter model, therefore, gives results which agree well with measurements.

Fig. 17 shows the phase curves for the different input powers intersecting each other at certain frequencies. The intersection can be shifted in frequency by changing the parallel inductance L in the diode circuit. In this way it is possible to minimize AM/PM conversion for the carrier frequency of 74.1 mc at the nominal input power to the limiter. Fig. 18 shows the phase shift at center frequency as a function of input power. The AM/PM conversion is zero at +3-dbm input power but increases rather rapidly for higher and lower powers. It stays below 0.2 deg/db (or −30 db) over a range of more than 1.5 db. The input power to the limiter is controlled by the AGC of the receiver to less than ±0.5 db variation about the nominal power, so that the AM/PM objective is always met. (The measurements of Figs. 17 and 18 were based on a

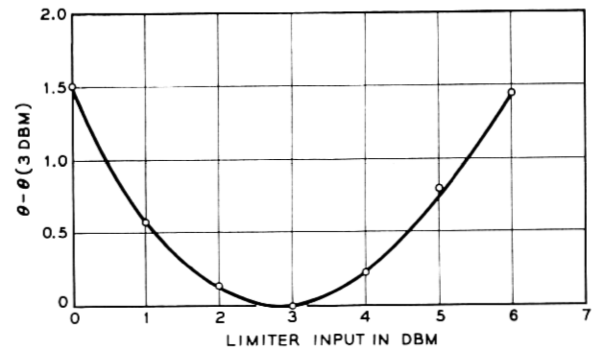

Fig. 18 — AM to PM conversion at 74 mc as a function of limiter input power.

nominal input power to the limiter of +3 dbm. This power was later lowered to the present value of +1 dbm. The inductance L was changed in value to optimize the circuit at the new power).

The limiter has to operate over a frequency range from 58 mc to 90 mc with only a few tenths of a decibel drop in transmission at the edges of the band. The basic diode network of Fig. 15 is inherently very broadband due to the limiting process. Its bandwidth is $B = f_0/Q$, and using (7), (8), and (9) we obtain:

$$B = \frac{I_p}{8cV_{co}} \tag{5}$$

which gives 131 mc if the values used for (9) above are inserted. The constant resistance networks have a bandpass characteristic which is flat to 0.01 db over the whole band. They also serve as harmonic filters, especially for the fairly strong third harmonic generated by the diodes. The over-all transmission characteristic of the limiter is flat to ±0.05 db from 64 mc to 84 mc and drops not more than 0.2 db at 58 mc and 90 mc.

2.4 *Dynamic Characteristics*

The preceding discussion considers only static amplitude changes. In practice the amplitude changes may have rates of up to 10 mc or even higher. Because of the large bandwidth of the diode circuit, the dynamic situation is little different. An approximate nonlinear analysis of the dynamic case indicates that the equations for compression and AM/PM conversion have to be multiplied by

$$\left[1 + \left(\frac{2f_m}{B}\right)^2\right]^{-\frac{1}{2}}$$

where f_m is the modulating frequency and B is given by (5). For f_m = 10 mc, this factor has a value of 0.988 and can therefore be neglected.

Fig. 19 shows dynamic compression in db for the complete two-stage limiter as a function of the modulating frequency. The curve is very flat with frequency, although the drop is somewhat greater than indicated by the theory. This compression curve can be affected by the impedance of the diode bias supply, through detection and remodulation. Such effects are avoided by the use of a very low resistive impedance bias source. A temperature compensation circuit changes the bias for the diodes in the second limiting stage to counteract temperature effect on the forward voltage drop of the diodes. In this way the output power of the limiter is constant to 0.05 db over a temperature range from 10° to 40°C.

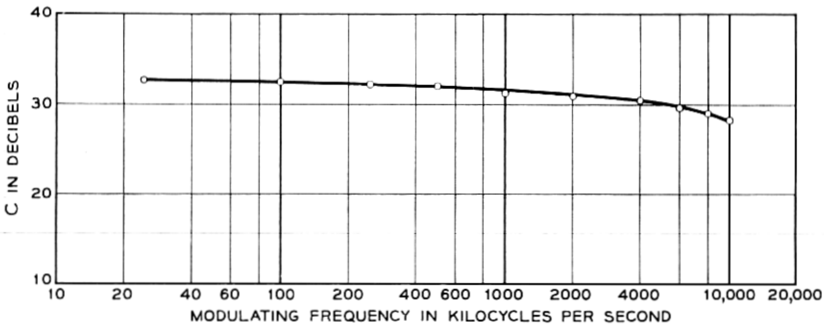

Fig. 19 — Dynamic compression for two-stage limiter as a function of modulating frequency.

The dynamic AM/PM conversion of the complete limiter was also measured. The accuracy of such measurements is affected by the difficulty of generating a 74-mc AM test signal free from PM. The results will include the AM/PM conversion generated by the unsymmetrical delay characteristic of the passive networks in the limiter circuits as well as AM/PM conversion due to limiter action. The measurements indicate the over-all dynamic AM/PM conversion is better than −30 db up to a modulating frequency of 10 mc.

2.5 *IF Carrier Resupply*

Without remedial measures, loss of the carrier allows the limiter and main IF amplifier to go to maximum gain. Within two repeater sections, the TWT amplifiers are saturated at full power with noise spread over a wide band. In the adjacent channels, this noise power is intolerably large. Protection against this is the function of the carrier resupply. When a carrier is lost, the gain of the limiter will reach maximum in a few millimicroseconds and the gain of the main IF amplifier in a few milliseconds. The carrier resupply replaces the lost carrier rapidly (within less than 0.1 millisec) and prevents subsequent repeaters from going to maximum gain. In addition, the limiter gain is reduced to attenuate the incoming noise and prevent it from modulating the new carrier.

The carrier resupply unit is a fully transistorized circuit containing a dc amplifier, a trigger circuit, a crystal-controlled oscillator, an amplifier, and circuits to reduce the limiter gain and operate a relay. The relay actuates the alarm and the automatic protection switching circuits. The ac equivalent circuit of the oscillator and amplifier is shown in Fig. 20; the other circuits, being conventional, are not shown. In the

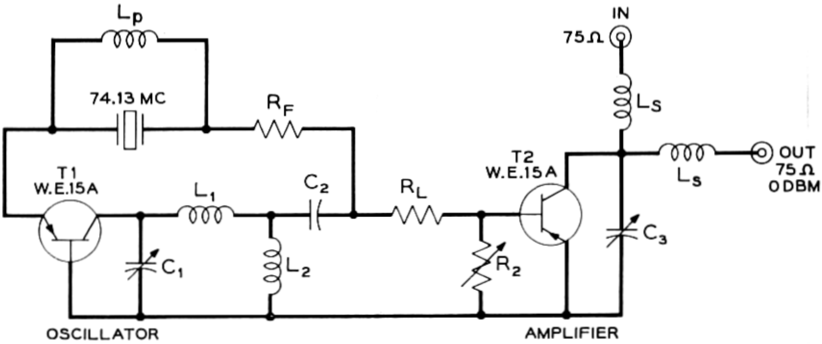

Fig. 20 — The ac equivalent of carrier resupply IF oscillator and amplifier.

oscillator and amplifier stages, high-frequency transistors having an alpha cutoff frequency of 750 mc are used. When a carrier has to be supplied, the transistors T1 and T2 are switched from cutoff to the operating point by the trigger circuit. The crystal oscillator starts oscillating in 60 microseconds. This rapid start time is obtained by resistor R_F in series with the crystal which lowers its Q. The crystal operates in the series-resonant mode at 74.1 mc, and the inductor L_p tunes out the parallel crystal capacitance.

The crystal is in the feedback loop of a grounded base transistor. The current gain necessary to produce oscillations and feed power into the load R_L is provided by the circuit C_1, L_1, C_2, L_2, which acts as an ideal transformer at the oscillating frequency of 74.1 mc. The second transistor T2 is connected as a grounded emitter amplifier. The current gain of this stage is approximately 15 db at 74 mc. The collector of T2 is bridged across a 75-ohm coaxial line by means of an extremely wide band low-pass filter consisting of L_S, C_3, L_S. R_2 adjusts the carrier power delivered to the 75-ohm output.

Measurements show the frequency of the oscillator to remain within ± 10 kc over a wide temperature range. This is well within the required limit of ± 50 kc. The starting time of the oscillator is dependent on temperature, but the variation is only a few microseconds over the range expected in the temperature-controlled radio repeater rooms.

2.6 *Transmitter and Carrier Supply Modulators*

The conversion gain available from the use of variable capacitance diodes as up-converters is important in the performance of the high-power microwave modulators used in the TH system. The transmitter modulator is used in the radio transmitter to convert the frequency

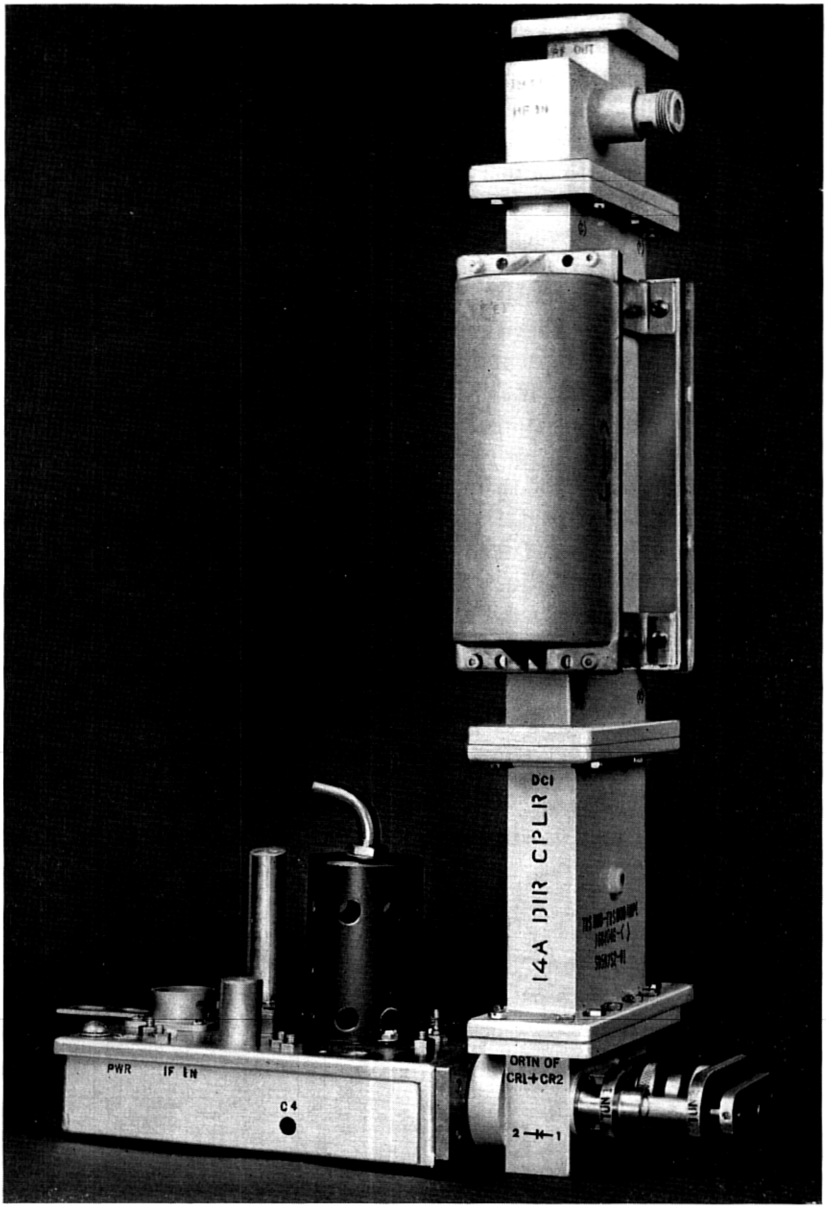

Fig. 21 — Complete transmitter modulator.

modulated IF signal of 74.1 mc to the required microwave frequency. Transmission performance is important in this modulator, and the diodes are biased to give approximately unity gain in order to obtain wide bandwidth. A photograph of a complete transmitter modulator is shown in Fig. 21. The carrier supply modulators are used in the microwave carrier supply to shift a microwave carrier by 252 mc, and in the radio transmitter and receiver to shift a microwave carrier by 29.7 mc or 59.3 mc. These are single-frequency devices, and conversion efficiency has been optimized.

The two types of modulators use a similar double-waveguide type of diode mount, and are shown schematically in Fig. 22. The diode mount provides a means of matching the 427A diode to WR159 waveguide, and is shown in Fig. 23. It is the same in principle as, but differs in detail from, the receiver modulator diode mount of Fig. 5. To match the resistive component of the diode impedance to the characteristic impedance of the waveguide, the coaxial line in which the diode is mounted is placed an appropriate distance off center in the waveguide. To compensate for differences between diodes, the coupling between the diode and the waveguide is adjusted by controlling the distance that the outer conductor of the coaxial line extends into the waveguide (RF TUNER 1 of Fig. 23). Reactance at the junction is partially cancelled by a waveguide short circuit fixed in position behind the coaxial extension. The reactance that remains is tuned out by means of a short circuit placed in the coaxial line. In the case of the transmitter modulator, this short circuit is made adjustable to compensate for differences between diodes (RF TUNER 2 of Fig. 23) and is set for optimum transmission over the band. In the carrier supply modulators, the position of the short circuit is fixed.

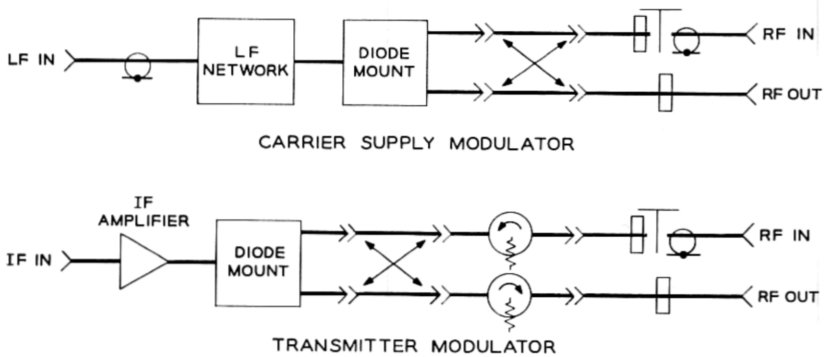

Fig. 22 — Schematics of the two types of modulators.

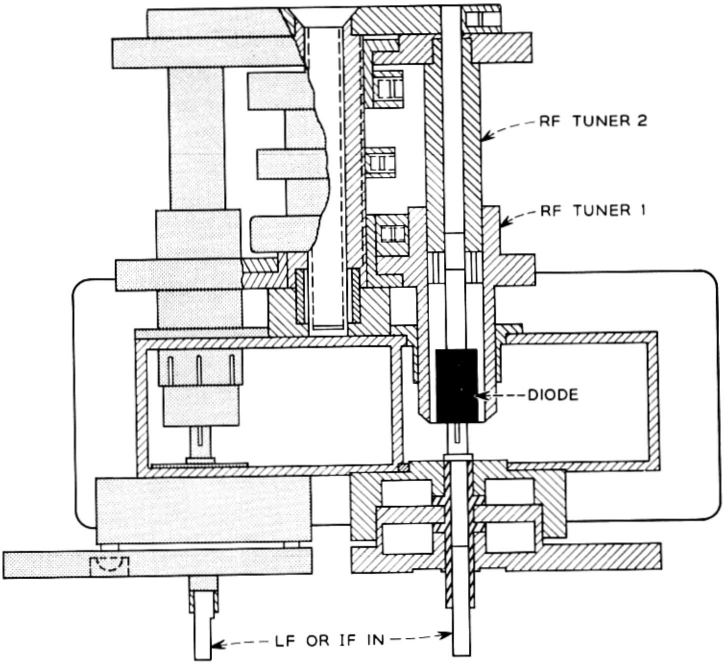

Fig. 23 — Double diode mount for transmitter and carrier supply modulator.

The use of radial line RF chokes in the input leads, a short-slot hybrid junction, and a double isolator in the transmitter modulator parallels the receiver modulator design. The dual isolator is omitted in the single-frequency carrier supply modulators.

In the carrier supply modulators, the low frequency is fed to the two diodes in parallel from a 75-ohm cable. Simple networks are used to match the cable impedance to the impedance of the two diodes. For the transmitter modulator, however, the IF signal is supplied to the diodes through a one-stage amplifier. The purpose of the amplifier is to isolate the 75-ohm cable from the diode IF impedance, which changes considerably over the 30-mc wide IF band. This amplifier employs a 448A which operates in the same manner as in the IF main amplifier. An input transformer matches the 75-ohm cable to the grid of the tube. A variable series inductor is in the plate circuit to correct for a slope in the RF output over each channel, and a relatively low plate load is used to minimize the gain-frequency variations due to variation of diode IF impedance.

With a microwave carrier (BO) input of +27 dbm and an IF input of 0 dbm to the one-stage amplifier, the microwave output power of the transmitter modulator is +7 dbm. The output is flat to within 0.05 db over a 20-mc band and flat to within 0.25 db over a 32-mc band. The BO energy appearing at the output is at least 20 db below its power at the input to the modulator.

The 427A diode is capable of giving conversion gain if sufficient negative bias is applied to it, the gain increasing as the negative bias is increased. This is due to the fact that the 427A diode actually is a variable capacitance p-n junction, or varactor, and, therefore is low loss. This phenomenon was demonstrated first on the laboratory version of this diode by Uhlir.[7] However, the impedance of the 427A diode becomes more frequency sensitive as the gain is increased, and for this reason the diodes are operated at relatively low conversion gain in the TH equipment. A bias of −1 volt gives approximately 0-db gain, which meets system requirements for all high-level modulators except the 252-mc carrier supply modulator. The latter is operated at −4 volts bias to offset the loss of gain inherent in a higher frequency input. In all cases the required bias is obtained from a resistor in series with the diode.

III. TRAVELING-WAVE TUBE AMPLIFIER

3.1 *General*

The TWT microwave amplifier has a power output of five watts and uses the Western Electric Company 444A tube, which is the production version of a 6000-mc TWT described in an earlier paper.[8] A photograph of the amplifier is shown in Fig. 24, and a line drawing of a cross section through it in Fig. 25. The amplifier uses a permanent magnet focusing structure which consists of two permanent magnets, two pole pieces, a field straightener, and a movable gun shield. The magnets and the associated pole pieces provide a uniform magnetic field along the axis of the electron beam of the traveling-wave tube. The field straightener unit is placed concentrically with the axis of the electron beam between the input and output waveguide to reduce any transverse field present. The gun shield which is near the electron gun region of the tube controls the field at the cathode as well as providing a means for magnetic focusing of the electron beam through the use of mechanical controls. Waveguide circuits are used to couple signals into and out of the tube. The impedance match between the waveguides and the helix of the tube is adjusted by mechanical control of the positions of the waveguide along the axis of the tube and of a shorting plunger in the waveguide.

Fig. 24 — The TH traveling-wave tube amplifier.

The collector and the tube envelope are air cooled. The entire structure is enclosed in a magnetic shield to eliminate interaction with adjacent TWT amplifiers and other circuits. In addition, the shield and the asso ciated mechanical and electrical interlocks protect operating personnel from the high voltages required for the tube.

The TWT amplifier has two applications in the transmitter; in the signal path (transmitter amplifier) and in the carrier supply path for the transmitter modulator (carrier supply amplifier). Requirements for the transmitter amplifier will be considered first.

The transmitter amplifier power output requirement is +37 dbm (5 watts). The allowable gain is dependent on the noise figure of the traveling-wave tube. The 444A has a noise figure of 27 db to 30 db, which is sufficiently high to make some contribution to the over-all repeater noise figure. The nominal gain of the repeater is 62 db (+37 dbm output, −25 dbm input), and the noise figure of the average-age radio receiver is about 11 db. If we allow the transmitter amplifier to make an additional contribution of 0.3 db to the repeater noise figure, then the average gain allowable for the amplifier is found to be about 33 db. The gain of the amplifier is therefore held to the range of 30 db to 35 db. In those cases where the amplifier gain exceeds 35 db at +37

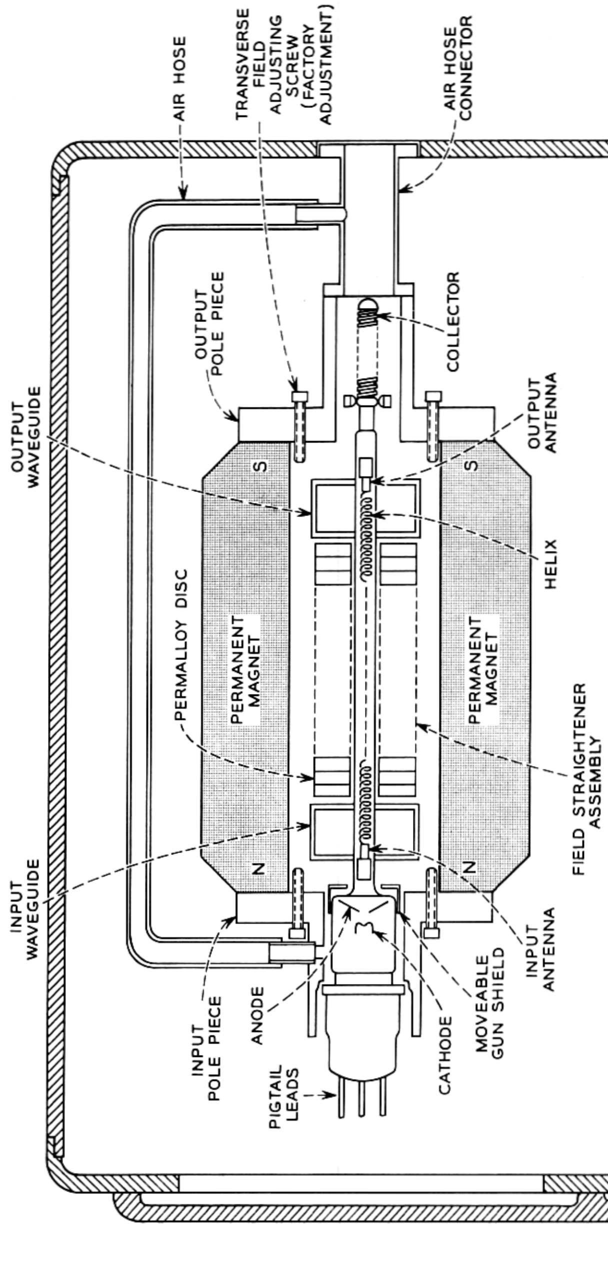

Fig. 25 — Cross section of traveling-wave tube amplifier.

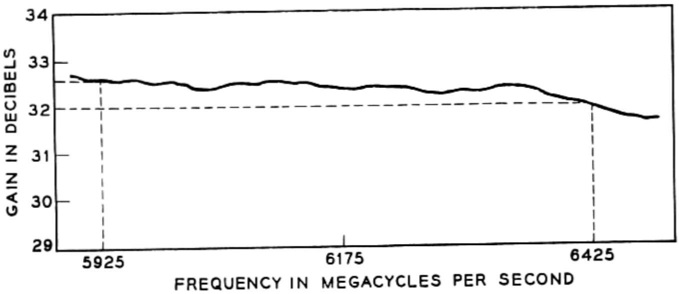

Fig. 26 — Typical gain-frequency response of the traveling-wave tube amplifier.

dbm output, the helix voltage is lowered until the gain is reduced to 35 db.

The gain-frequency response of the amplifier is substantially flat over the 5925 to 6425-mc band, and a typical example is shown in Fig. 26. The input return loss is 25 db minimum to provide a proper termination for the channel bandpass filter. The active or hot output return loss is about 10 db and is controlled principally by internal reflections in the tube. At least 25 db return loss is needed for terminating the channel separation network, and an isolator is used to meet this requirement.

Spurious radiation requirements call for second harmonic (12 kmc) output at 50 db below the carrier. As the amplifier at +37 dbm output may have the second harmonic down only 25 db, a low-pass filter providing a minimum of 25 db attenuation to the second harmonic is inserted immediately following the amplifier. Since 12-kmc energy can be propagated in WR 159 waveguide in several modes, it is not attenuated appreciably by the channel bandpass and channel separation networks.

The carrier supply amplifier output power requirement is +27 dbm (0.5 watt). Here, also, the allowable gain is determined by the noise figure of the 444A TWT. Noise on the carrier supply at BO frequency causes phase modulation, which appears as noise on the signal. To minimize this noise contribution, an input power of about +6 dbm is used, and the TWT gain is reduced by lowering the helix voltage. This provides a high carrier-to-noise ratio.

The structure associated with the TWT must focus the electron beam, couple energy in and out of the slow-wave structure, and dissipate the heat generated by the expended electron beam. Other design requirements arise from over-all system considerations. The economics of continuous operation in unattended remote repeater stations dictates that the power required to focus the beam and to cool the tube be minimized. The compact arrangement of the radio equipment demands that the external magnetic fields of the focusing structure be negligible, or they will adversely affect the operation of nearby equipment. Furthermore, as the electron structure is the only part of the amplifier that deteriorates with age, its replacement in the field with a minimum number of simple adjustments is required for desirable operating and maintenance practices.

3.2 *Magnetic Circuit*

The description of the magnetic circuit divides conveniently into three parts. These correspond to the regions surrounding the electron gun, the helix, and the collector of the TWT. To a greater or lesser extent, the distribution of magnetic field in all three regions affects beam focusing, noise performance, and collector efficiency.

A minimum longitudinal focusing field of 580 oersteds is provided for the proper operation of the 444A tube. To achieve long life, the intercept current to the helix is kept to less than 2 per cent of the beam current of 40 ma. This requires that the maximum value of the transverse field not exceed about two oersteds.

The focusing field could be generated either by a solenoid or by permanent magnets in either a periodic or straight field configuration. The cost of providing electric power to operate the solenoid, however, is prohibitive. Although the periodic focusing method has the advantage of a size and weight reduction when compared to the straight-field method, this is not of primary importance in a ground installation such as the TH system. The straight-field structure is inherently simple in form and lends itself more easily than the periodic structure to the introduction of the broadband waveguide coupling circuits required to provide the necessary transmission performance. Shielding of the periodic structure, to control external fields, is relatively easy compared to the straight-field structure. However, at the time of this development, the high coercive force magnetic materials available for periodic structures were not sufficiently temperature insensitive to be practical. The structure chosen was therefore of the straight-field type.

The design of the magnetic circuit, shown in Fig. 25, can be achieved from the procedures described by M. S. Glass.[9] A prototype structure, having no external magnetic shield, required $19\frac{3}{4}$ pounds of Alnico VI to produce the required 580 oersteds in the 7.1-inch gap between the pole pieces. In the final design the addition of the external shield increased the magnet loading appreciably. As a result, it was not only

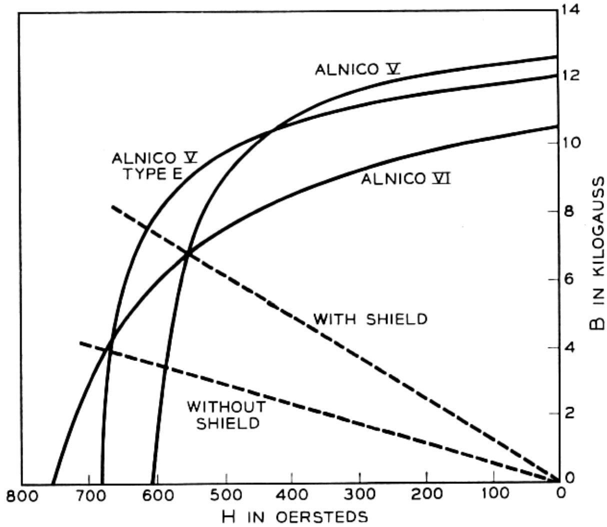

Fig. 27 — B-H curves showing effect of shielding on permanent magnet loading.

necessary to increase the amount of magnetic material to 27 pounds but to use Alnico V, type E, in place of Alnico VI. Fig. 27 shows the B-H curves for Alnico V, Alnico V type E, and Alnico VI, together with the load lines for the structure with and without shielding. It can be seen that the high coercive force of Alnico VI could not be utilized due to the loading of the shield. A plot of a typical longitudinal field distribution is shown in Fig. 28.

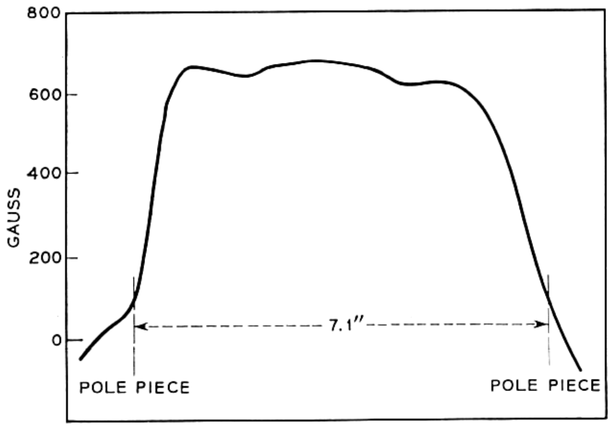

Fig. 28 — Typical longitudinal field distribution along axis of tube.

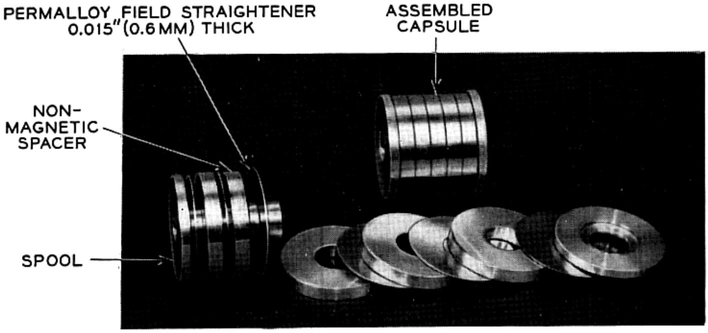

Fig. 29 — Field-straightener assembly.

To realize the transverse field requirements, it is necessary first to align the axis of the helix accurately with that of the magnetic circuit. If the net transverse component is to be less than two oersteds, the maximum mechanical misalignment must be less than 2/600 radian or about 0.2 degree. Secondly, it is necessary to provide as uniform a magnetic field as possible. Since the two magnets in the circuit are poled alike, minor differences between the magnets generate transverse fields. In the space between the waveguides these transverse fields are controlled by the field-straightener assembly, which is shown in Fig. 29. This consists of 18 thin Permalloy discs spaced $\frac{3}{16}$-inch apart. The high permeability of the discs normal to the axis of the tube shorts out the transverse fields. The discs are assembled in three identical cartridges, six discs to a cartridge, and are encased in a precision-machined aluminum casting. As shown in Fig. 30, the complete field-straightener assembly, in turn, is suspended between the waveguides by the two nonmagnetic stainless steel rods that connect the pole pieces. In the region occupied by the waveguides, field straighteners cannot be used, and the transverse fields in these regions are controlled by the adjustment of four iron screws inserted through each pole piece. Fig. 31 shows a typical transverse field plot.

Special attention was given to the shape of the field buildup at the ends of the structure, particularly at the cathode or input end. The 444A uses a converging Pierce-type electron gun, which is normally operated with the cathode completely shielded from the magnetic field.[8] Best focus results if the magnetic field is introduced in the region between the accelerating anode and the point where the electrostatic fields in the gun cause the beam to reach its minimum diameter in the absence of the magnetic field. However, the noise figure is importantly affected

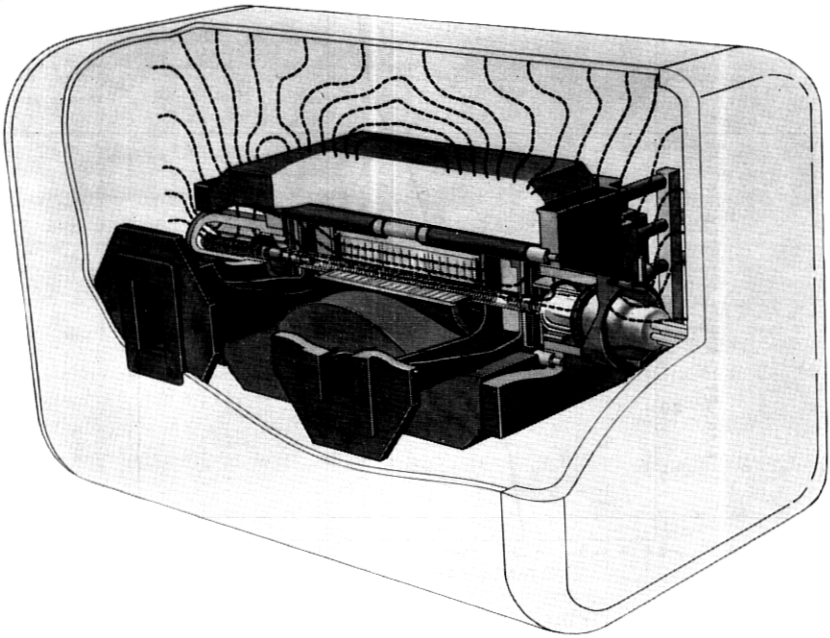

Fig. 30 — Magnetic assembly of the traveling-wave tube amplifier.

by the amount of focusing field in the gun region. Measurements made with a minimum field at the cathode showed that at the +37 dbm rated output power, the noise figure of the tube increased from about a low-power output value of 30 db to 40 db or higher. This is attributed to a growing noise wave on the electron stream, which depends on, among other things, the drive on the tube.[10] This growing noise wave can be

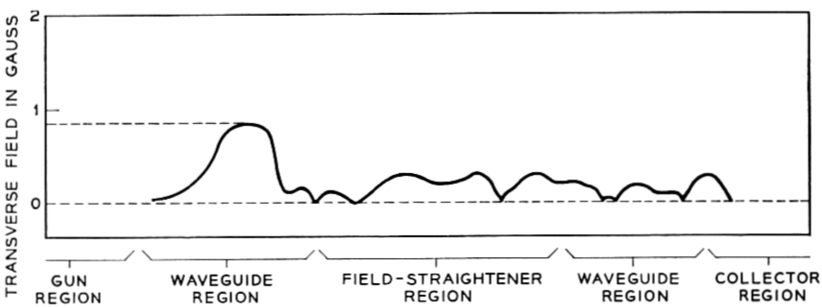

Fig. 31 — Typical plot of transverse field along axis of the tube.

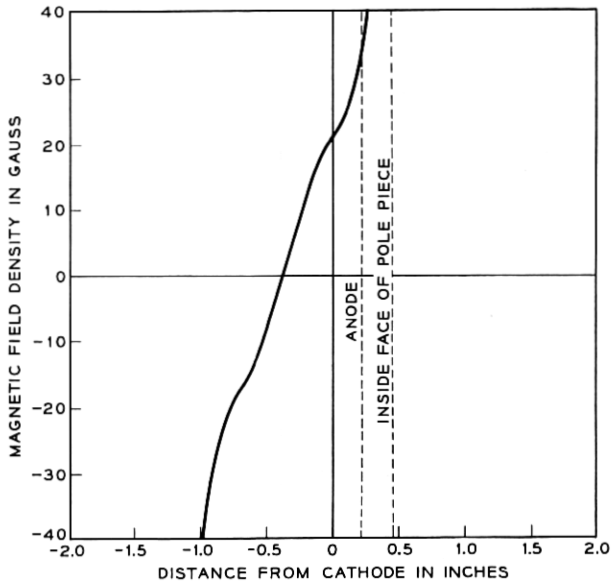

Fig. 32 — Longitudinal field in gun region of the traveling-wave tube amplifier.

reduced by increasing the magnetic field at the cathode. A field of 15 gauss is adequate to maintain the noise figure relatively constant throughout the tube operating range. The gun region magnetic structure is designed accordingly. In production, the cathode field is adjusted by selection of a suitable size soft-iron ring slipped over the gun-end shield.

As shown in Fig. 25, the magnetic field configuration in the gun region is controlled by the end shield and the gun shield, and by the magnetic material (Kovar) details used in the base of the tube itself. The most effective parameter for control of the field is the design of gun shield. The magnitude and shape of the field depend on the length of the gun shield, the size of the hole at the helix end of the shield, the type material and thickness of the gun shield face and walls, and the longitudinal positioning of the shield with respect to the tube and the rest of the magnetic circuit. Optimization of the dimensions provides the desired conditions as shown in Fig. 32.

Besides controlling the field conditions in the gun region, the gun shield is also used as a fine adjustment of the focus, by two controls which position it in a plane perpendicular to the tube axis. This movement produces a small transverse magnetic field in the vicinity of the hole in the gun shield but causes essentially no change in the field conditions elsewhere. This small transverse field is sufficient to provide a successful focusing adjustment with the close mechanical tolerances imposed on the 444A and the magnetic circuit.

The magnetic field conditions at the collector end of the circuit are not particularly critical. Excessive helix intercept current results unless the focusing field is maintained close to a value of 600 oersteds out to the end of the helix, but the rate of field decay beyond the end of the helix is not important.

The design of the external shield is controlled by the tolerance of adjacent equipment to magnetic fields. To keep the size reasonable, the shield must be located in a region of high field intensity. Under these conditions, the magnetic properties of low carbon steel are superior to those of most alloys normally used for shielding. The sides of the shield are fabricated from $\frac{1}{8}$-inch low carbon steel. The thickness is determined by the allowable external field. Magnetic iron castings are used at the ends to complete the shield. The external field is less than one oersted at one-half inch from the center of the sides of the complete assembly.

A large dc-powered solenoid, having an inside diameter of $12\frac{1}{4}$ inches and a length of 24 inches with a magnetic field intensity of 4000 oersteds, is used to magnetize the complete assembly.

Specialized flux-measuring equipment is required for determining the longitudinal and transverse components of the focusing field. The helix of the 444A has an inside diameter of 0.080 inch. To measure the field in the region of the electron beam, longitudinal and transverse search coils of approximately the same diameter as the helix are used. These coils mount in an accurately machined rod which is supported from the surfaces used to locate the TWT in the magnetic structure.

Two types of search coils and associated equipment are used. The first type, used primarily for longitudinal field measurement, consists of an air-core coil and an integrator of a type similar to that described by Cioffi.[11] The second type, primarily for transverse-field measurement, uses a permalloy-core coil as a magnettor with its associated equipment.[12] In both instruments the field is plotted on an X-Y recorder as a function of search coil position.

3.3 *Helix-to-Waveguide Transducer*

The microwave signal is coupled into the TWT by surrounding an antenna attached to the helix with a section of standard rectangular waveguide, to form a transducer. As shown in Fig. 33, a radial-line choke is incorporated to form an effective ground plane at the end of the helix. Dimensions A and B are both a quarter-wavelength long, and a microwave short circuit appears across the opening CC. The performance of the choke is broadbanded by making the characteristic impedance of section A high relative to that of section B. Ordinarily, many turns of the helix are exposed to the field in the guide, which results in weak coupling. In this transducer the coupling is increased by stretching out the last turn of the helix to aid in matching it to an antenna or post which can be coupled to the waveguide. In Fig. 33 the height of the post is denoted by D and length of the last turn on the helix by E.

To a first approximation, the equivalent circuit of the helix-waveguide junction, as shown in Fig. 33, may be represented by the parallel combination of a conductance G and a susceptance jB connected across a transmission line; Y_0 is the characteristic admittance of the waveguide. For values of G near Y_0, the conductance is determined mainly by the height of the post and the susceptance by the pitch of the last helix turn. Increasing the post height increases G; increasing the length of the last helix turn decreases jB. By proper choice of dimensions, therefore, it is possible to obtain a match between helix and waveguide.

For small values of jB, G is almost constant with frequency, while jB itself decreases somewhat as the frequency is increased. If the waveguide is terminated by a microwave short circuit somewhat less than a quarter wavelength beyond the helix, this termination will look like a small inductive susceptance at the helix. This susceptance also decreases as the frequency increases. By dimensioning the post so that the value of G equals Y_0, and the last helix turn so that its capacitive susceptance, jB, cancels the inductive susceptance due to the short circuit, it is possible to produce a broadband match between helix and waveguide. Fig. 34 shows the return loss of the typical transducer over the 6-kmc band.

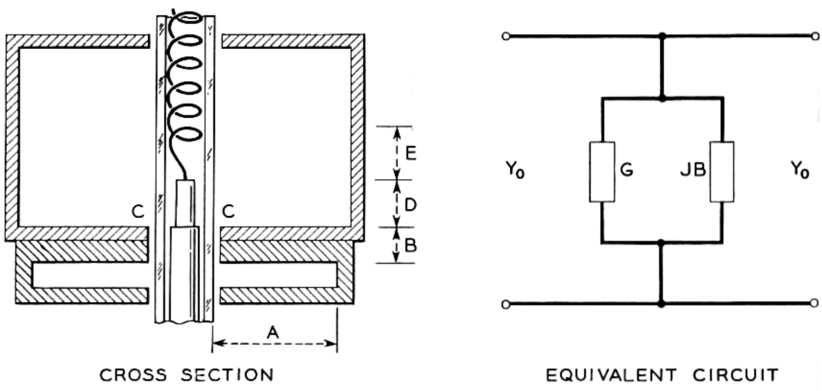

Fig. 33 — Cross section of helix-to-waveguide transducer and equivalent circuit.

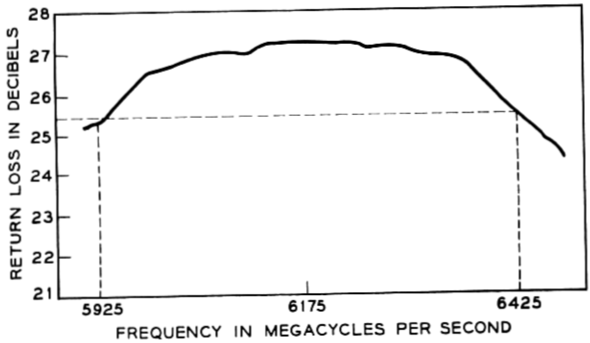

Fig. 34 — Return loss of typical transducer.

The same transducer arrangement is used at both ends of the helix. However, while the input match of the amplifier is that of the helix-waveguide transducer, the output match is controlled by minute periodic imperfections in the helix.[8] Any reflection of the output signal from the circuits beyond travels backward along the helix, essentially undiminished, to the helix attenuator. Impedance variations along the helix resulting from mechanical imperfections or at the helix attenuator cause a portion of the reflected signal to be re-reflected toward the output and be amplified by the gain existing between the point of reflection and the output. The output transducer connects to a microwave harmonic filter whose out-of-band impedance is reactive. Over a wide range of frequencies, therefore, the helix is mismatched either by the filter or the transducer. As a result, the traveling-wave tube itself must be short-circuit stable if the amplifier is not to oscillate.

In addition, it is necessary to control any external coupling between output and input. The radial line chokes do not provide perfect ground planes at all frequencies; some energy escapes along the extension of the helix post, which can be coupled to the other end of the structure through a crude cavity formed by the magnetic shield. To reduce the leakage, beryllium copper fingers between the outside of the choke and the pole piece are provided. The high-voltage connection from the terminal board at the input end to the collector is a second source of coupling. Lossy beads are put on the collector lead to provide attenuation in this path.

3.4 *Collector Cooling*

In normal operation, the beam current of the 444A is 40 ma and the collector voltage is 1250 volts, providing 50 watts to be dissipated in

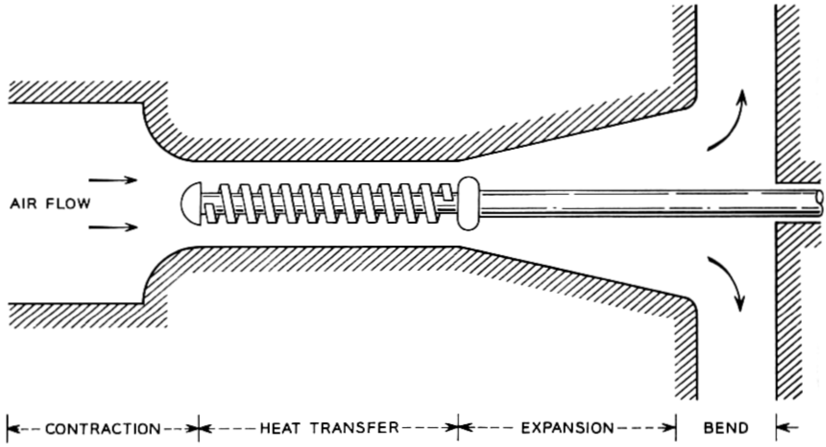

Fig. 35 — Cross section of the collector cooling chamber.

the collector. Since the collector is only about $1\frac{3}{4}$ inches long and $\frac{5}{16}$ inch in diameter, the heat dissipation is approximately 30 watts/in^2. To prevent serious outgassing, the temperature of the collector must not exceed 300°F, and forced air cooling is employed.

To minimize the pumping power required for forced air cooling, it is necessary to minimize the product of pressure drop and volume rate of flow. Fig. 35 shows a cross section of the collector cooling chamber. For a given rate of flow, the energy losses in each section are directly proportional to the corresponding pressure drops, but only the energy loss in the heat transfer section contributes to cooling the collector. The expansion and compression sections were designed for minimum pressure drops.

The pressure drop across the heat transfer section and the heat transfer to the cooling air are both a function of the air velocity and the roughness of the collector surface. The collector surface conditions that would minimize the pumping power for a given heat transfer were determined experimentally. With an initial air temperature of 70°F, the required static pressure drop and volume rate of flow are $4\frac{1}{4}$ inches of water and four cubic feet per minute. This corresponds to a pumping power of approximately two watts.

A small quantity of air is piped from the collector inlet to the electron gun envelope to keep its temperature from exceeding 250°F. The final pressure and volume requirements for the amplifier, which allow for a higher ambient temperature, are eight inches of water and six cubic feet per minute. The corresponding pump power is approximately $5\frac{1}{2}$ watts.

3.5 *Thermal and Ion Oscillation Noise*

The contribution of the TWT amplifier noise figure to the repeater performance has been discussed. To measure the TWT frequency modulation noise, a microwave signal is applied to the amplifier from a klystron. The carrier and noise sidebands appearing at the amplifier output are amplified and demodulated by a skeletonized radio receiver and FM receiver. The recovered video noise spectrum, of the familiar triangular shape, can be measured with an ordinary selective analyzer or AM detector. The measuring circuit is calibrated by using a known noise source.

A second cause of noise impairment in the system due to the TWT results from ion oscillation in the electron stream, which gives rise to spurious modulation of the carrier. In the 444A this modulation appears as a relatively narrow band of energy in excess of normal tube noise, usually at about 2.8 mc each side of the signal carrier. By using a relatively short period of aging under full RF drive conditions during manufacture, it has been possible to hold the ion oscillation noise at baseband to 10 db or less above normal tube noise. As the ion density within the tube decreases with operation, the excess noise disappears after a few hundred hours.

The allowance of 10 db above thermal noise for ion oscillation assumes that only a few new tubes would be installed in a 4000-mile system at any time within the few hundred hours required for tube ion cleanup. Since the thermal noise contribution of one repeater is 21 db below that of the system, and the TWT contribution 12 db below that, the increase in noise due to one tube with ion noise 10 db above thermal in one repeater is negligible. In fact, if 10 tubes were changed at once, the increase in noise for a 4000-mile circuit would be only about 0.3 db, which would appear in only a few dozen telephone channels for a few hundred hours.

3.6 *Mechanical Design*

The over-all dimensions of the amplifier are approximately $9\frac{3}{4}$ by 10 by $17\frac{1}{4}$ inches, and the complete unit weighs 86 pounds, of which the permanent magnets account for 27 pounds and the external shield accounts for 36 pounds. The front and rear of the shield are cast iron and engage interlocking tabs on the ends of the side covers. Large aluminum castings which attach to each pole piece suspend the magnetic circuit inside the shield. Four screws, accessible only through the front door, run the full length of the amplifier and secure the front and rear castings. The amplifier mounts on a vertical panel on four large brass pins that protrude from the panel. These four pins engage the aluminum castings through holes on the side cover. The ends of the mounting pins are tapped to accept captive screws which come through from the opposite side cover. A locked end door gives access for replacement of the electron tube. To obtain the key to unlock the door, it is necessary to lock the high voltage power supply to the OFF condition.

IV. EQUALIZATION

Equalization is an area which is still under development, as discussed in Ref. 1. The description here is confined to the non-adjustable equalizers mounted on the broadband receivers. One type is the basic gain and delay equalizer used in every receiver. It is intended to compensate for departures from ideal of the transmission characteristics of one complete radio section. Its characteristic will eventually be determined by the average transmission characteristics of a large number of radio sections. Fig. 36 shows the average characteristic of the first 30 manufactured transmitter-receiver pairs, as determined from factory point-by-point measurements. The envelope delay distortion (EDD) curve includes the limiter, but this was necessarily excluded in determining the loss characteristic.

The present basic equalizer is made up of five delay sections in tandem with a gain-correcting section. The delay sections are 360° all-pass networks in a bridged-T, constant-R configuration. The gain sec-

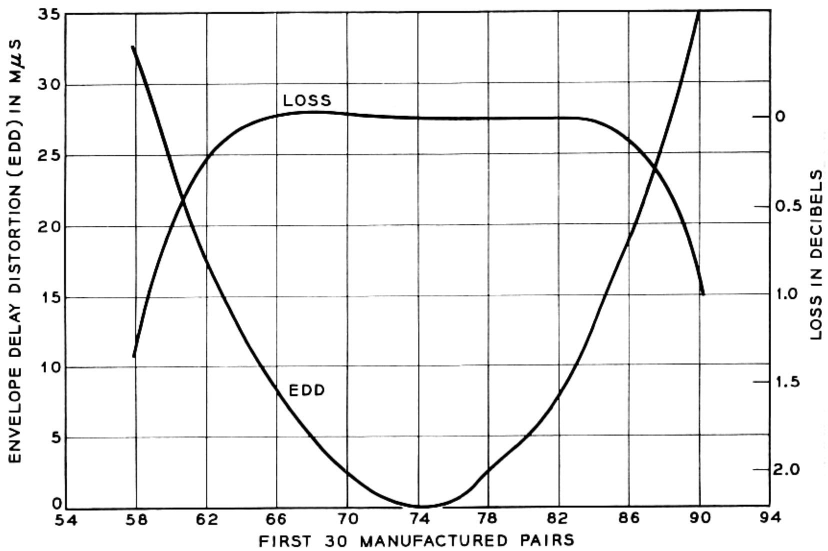

Fig. 36 — Average characteristic of the first 30 manufactured transmitter-receiver pairs.

tion is similar. Each section is individually shielded in cans which are attached to a common U-shaped chassis. The equalizer matches the repeater to about ± 1 mμs over 60 mc to 88 mc.

The transmission characteristics vary from channel to channel, for several reasons, although much effort is taken to minimize this. Part of the variation is due to the differences in frequency, even though the microwave filters are specifically designed for constant bandwidth. Part is due to unavoidable manufacturing variations. Part arises from the varying number of channel separation networks which are traversed on the path to the antenna. The most important part of this channel-to-channel variation is linear envelope delay distortion, or delay slope. Accordingly, two sizes of delay slope equalizer are made available: $+1$ mμs/mc and -1 mμs/mc. Over-all EDD measurements of a switching section determine the amount of slope correction needed for each channel. The necessary number of slope equalizers are then distributed among the repeaters of the channel. For the worst channel, one-third of the repeaters may need delay slopers. They are mounted on the receiver, as indicated in Fig. 1. Their design is very similar to the basic equalizer: two delay and one loss sections are used. The delay shapes are linear to about ± 0.2 mμs over 60 mc to 88 mc.

V. ACKNOWLEDGMENTS

A project of this magnitude represents the work of many competent development engineers, and this article could not have been written without their assistance. Special mention should be made of D. R. Jordan, P. R. Wickliffe and the late F. W. Koller for their work on the traveling-wave tube amplifier. H. W. Andrews, M. G. Davis, and P. I. Sandsmark contributed to the modulator designs, and A. J. Giger was largely responsible for the amplifier-limiter and carrier resupply designs. Many others are to be thanked for their contributions.

APPENDIX

To show the dependence of AM/PM conversion on circuit parameters.

The phase shift, θ, between the driving voltage V_L and the current I_p across a parallel RLC circuit for small frequency deviations Δf from center frequency is

$$\theta = -2Q \frac{\Delta f}{f_0}, \tag{6}$$

where f_0 is the resonant frequency, and

$$Q = 2\pi f_0 RC. \tag{7}$$

In a clipping network, like Fig. 15, but with perfect diodes, the fundamental component V_L of the output voltage is constant in amplitude. At f_0 the apparent parallel resistance R_A is then given by

$$R_A = V_L/I_p. \tag{8}$$

Now if we assume that (at $\pm V_{co}$) the diodes are clipping the incident sine waves symmetrically and so close to the base line that the voltage across the diodes appears essentially as a square wave, then the fundamental component of this square wave is given by

$$V_L = \frac{4}{\pi} V_{co}. \tag{9}$$

Combining (6) to (9) we obtain

$$\theta = -16 \frac{\Delta f c V_{co}}{I_p}. \tag{10}$$

Now AM/PM conversion is defined by

$$P = \frac{d\theta}{dV_i/V_i} \tag{11}$$

where $d\theta$ is the infinitesimal change of phase in radians caused by a fractional change dV_i/V_i of input voltage. Since I_p is proportional to V_i, (11) may be written as

$$P = \frac{d\theta}{dI_p/I_p}. \tag{12}$$

P may be expressed in degrees per db by multiplying by the factor 6.6. Differentiating θ with respect to I_p in (10) and substituting in (12) we obtain

$$P = 16 \frac{\Delta f c V_{co}}{I_p}, \tag{13}$$

or if P is expressed in degrees per db,

$$P = 105.6 \frac{\Delta f c V_{co}}{I_p}. \tag{14}$$

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