Active IF Units for the Transmitter and Receiver

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(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.[1] 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.

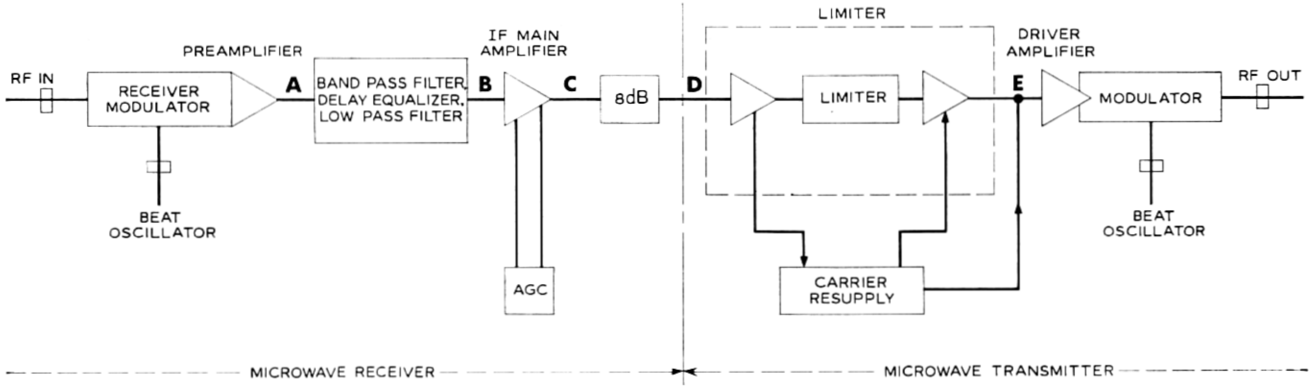
The design philosophy adopted was to make all the active circuits broad band to achieve good control over amplitude and delay characteristics. The 3 dB points of individual IF amplifiers fall typically below 5 MHz and above 130 MHz; and the transmission characteristics can be adjusted to within 0.01 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.[2]

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 ¼ 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.[3] 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 dBrnCO), the 100A switching system operates to substitute the protection radio channel for the working radio channel.[4] Assuming 14 dBrnCO 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 variolosser stages. A monitor stage has an output to drive the AGC circuit.

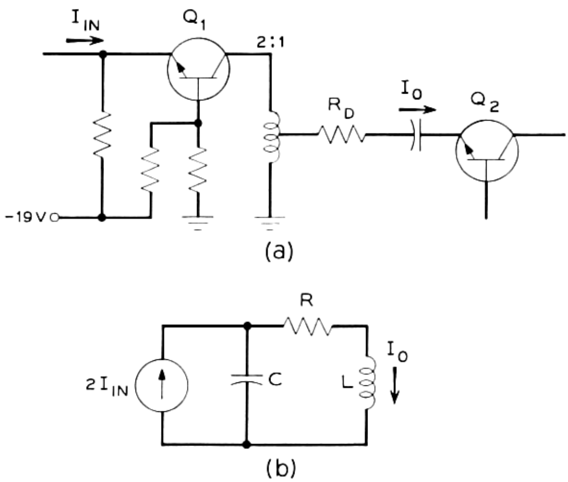


Fig. 2 — Schematic and equivalent circuit of common base transformer coupled stage.

The basic gain building block of the amplifier, shown in Fig. 2a, is the common-base, transformer-coupled gain stage.[5] 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:

$$20 \log | I_0/I_{IN} | = 6 - 10 \log [(f/f_p)^4 + (4\delta^2 - 2)(f/f_p)^2 + 1] \tag{1}$$

where

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

$$\delta = \pi RCf_p \ .$$

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 δ (the damping factor) and f_p (the peak frequency).

The optimum value of δ (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:

$$20 \log | I_0/I_{IN} | = 6 - 10 \log [(f/f_p)^4 - 2(f_c/f_p)^2(f/f_p)^2 + 1], \tag{2}$$

with

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

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.[6] The transformer, designed

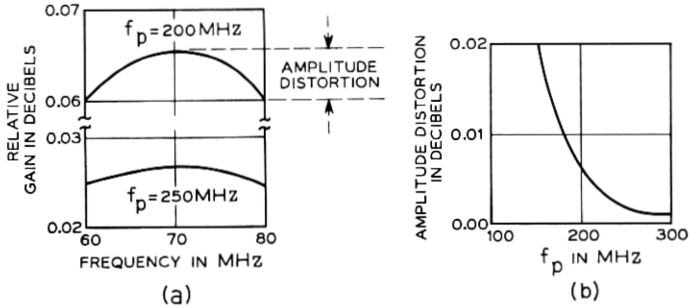


Fig. 3 — Amplitude distortion for a single stage.

specifically for this application, has the transmission line design described by C. L. Ruthroff.[7] 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 variolosser. 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

$$I_0/I_{in} = \frac{R_D}{[R_D + (L_{IN}/C_{OB})^{\frac{1}{2}}]} \cdot \frac{1/L_{IN}C_{OB}}{[\omega + (1/L_{IN}C_{OB})^{\frac{1}{2}}]^2}, \tag{5}$$

where

C_{OB} is the transistor output capacitance
L_{IN} is the emitter input inductance
R_E is the emitter input resistance.

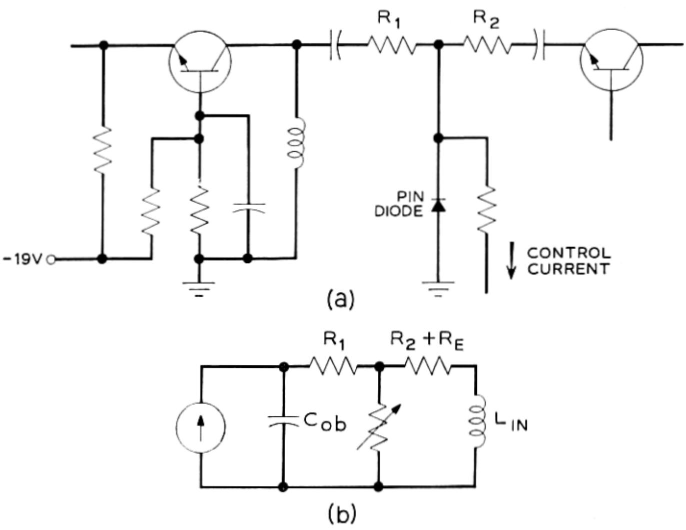


Fig. 4 — Schematic and equivalent circuit of variolosser.

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 variolosser is, therefore, virtually flat over the 60 to 80 MHz band. The variable resistance element of the variolosser is a 474A PIN diode.[6] 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 variolossers 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 variolosser stages. The relative positions adopted for the gain and variolosser stages is shown in Fig. 5 along with a level diagram for conditions of normal and maximum gain. Each gain-variolosser 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 variolosser 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.

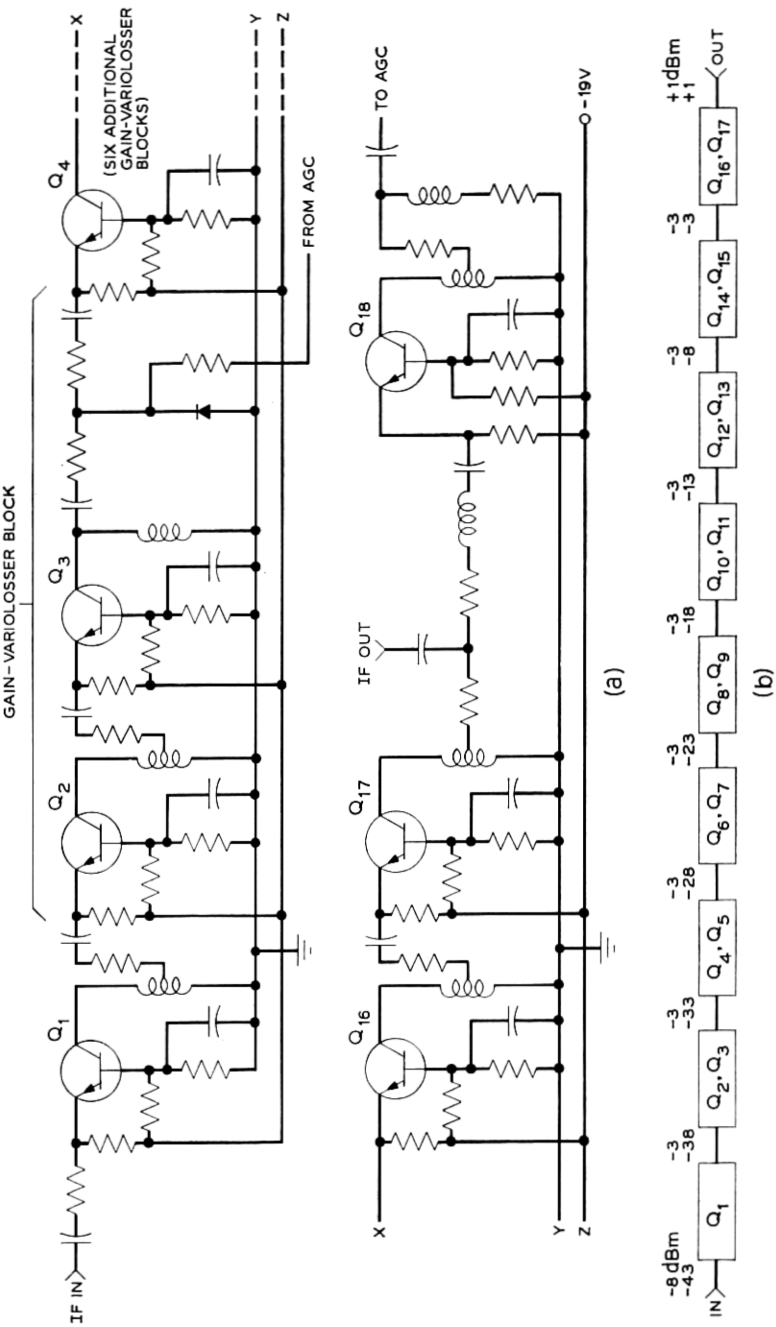


Fig. 5 — IF main amplifier schematic (a) and level diagram (b).

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

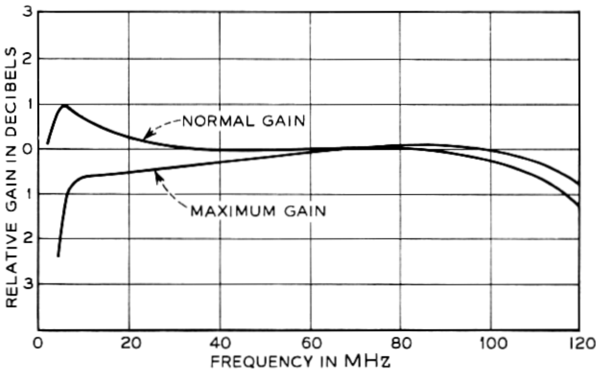


Fig. 6 — Transmission characteristic of IF main amplifier.

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 α and f_t.

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* 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.

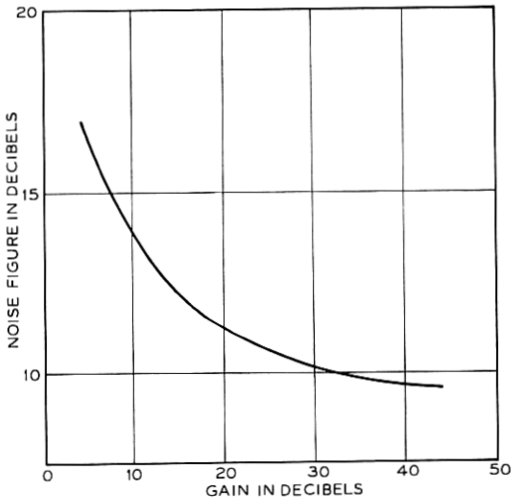

Fig. 7 — Noise figure vs gain of the IF main amplifier.

Figure 10 is a photograph of the amplifier housed in a 2 × 4 × 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 variolosser 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.

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

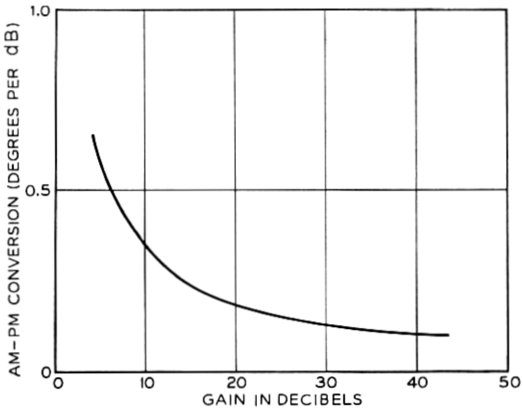

Fig. 8 — AM-PM conversion vs gain of IF main amplifier.

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 variolosser 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

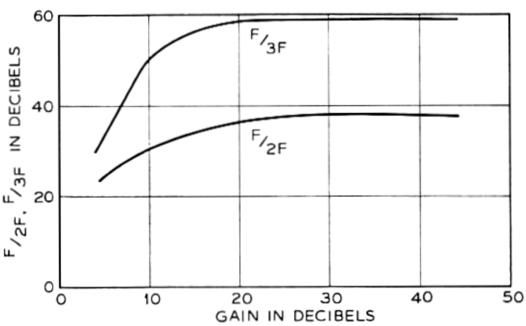

Fig. 9 — Harmonic distortion vs gain of IF main amplifier.

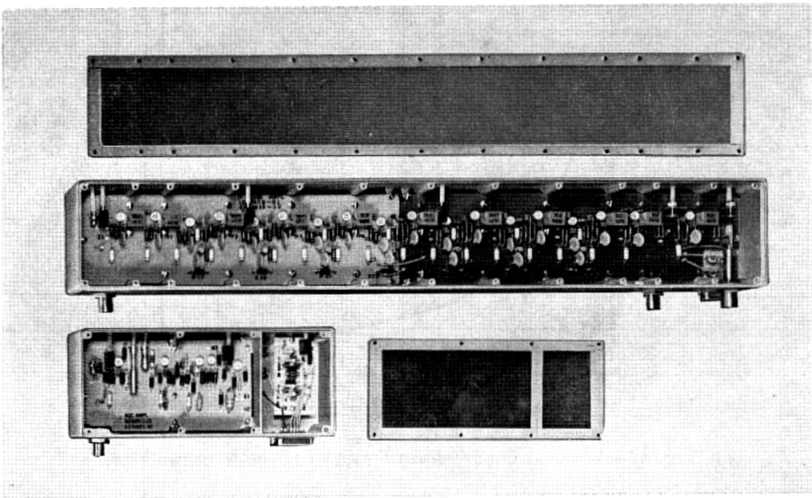

Fig. 10 — IF main amplifier and AGC unit.

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.[8]

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

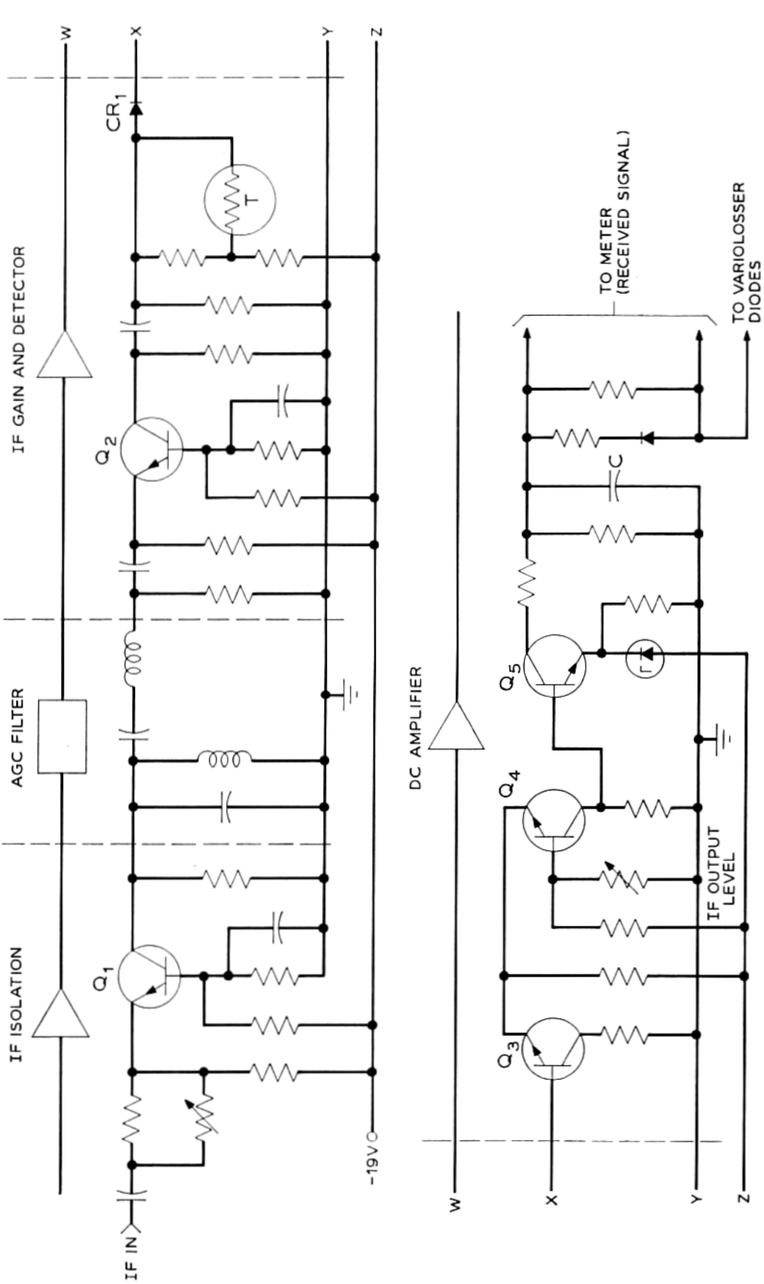

Fig. 11 — Block diagram and schematic of AGC amplifier.

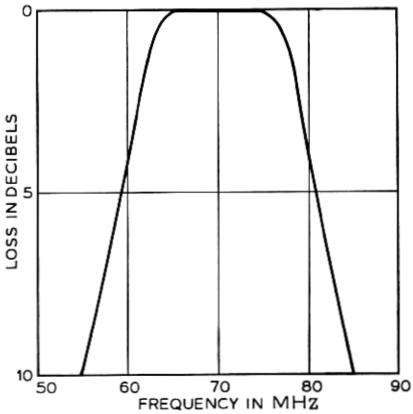

Fig. 12 — AGC filter characteristic.

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 β 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.[6] 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° per dB. For a received signal of −20 dBm a coefficient of about 0.25° 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 dBrnCO, and because several hops are not likely to up-fade simultaneously. Typical

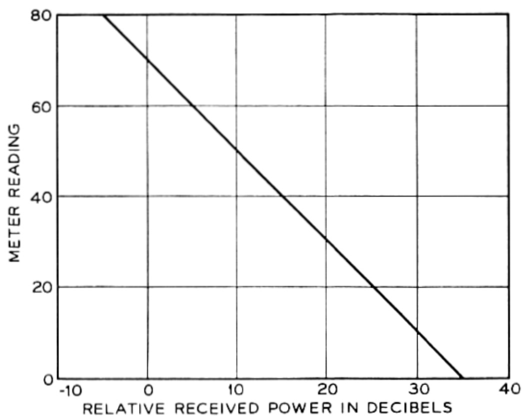

Fig. 13 — Received signal level meter reading vs relative received signal power.

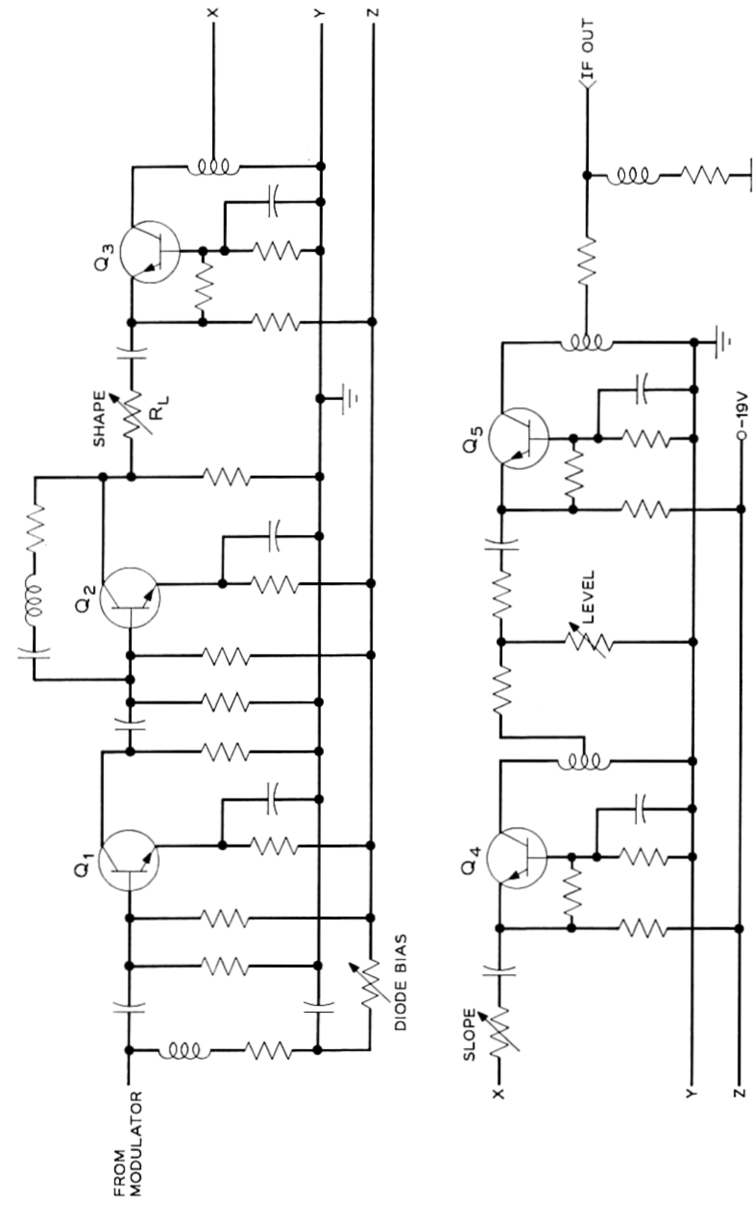

Fig. 14 — Preamplifier schematic.

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

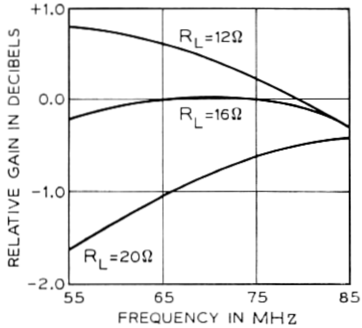

Fig. 15 — Transmission characteristic of first two stages of IF preamplifier.

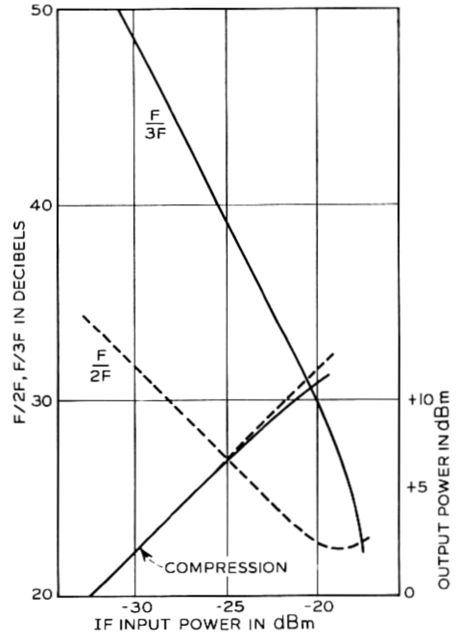

Fig. 16 — Compression and harmonic distortion characteristics of the preamplifier.

environment of transistor circuitry. The diodes CR1 and CR2 are 479A epitaxial silicon Schottky barrier diodes.[6] 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.

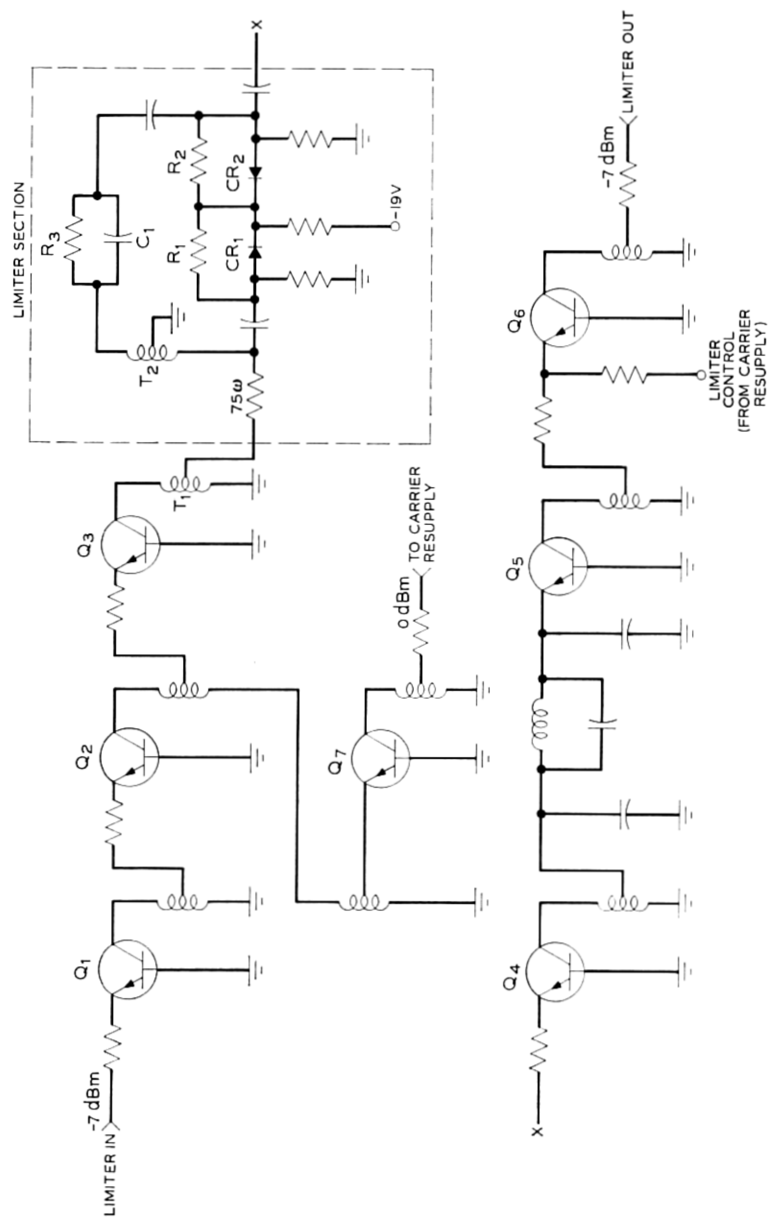

Fig. 17 — Limiter circuit—simplifier schematic.

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

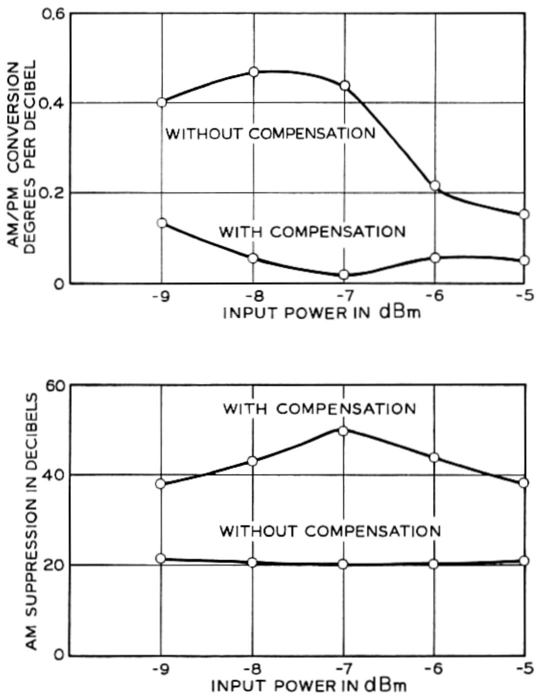

Fig. 18 — Limiter AM-PM conversion and AM suppression vs input power. $f_c = 70$ MHz, $f_m = 100$ kHz.

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

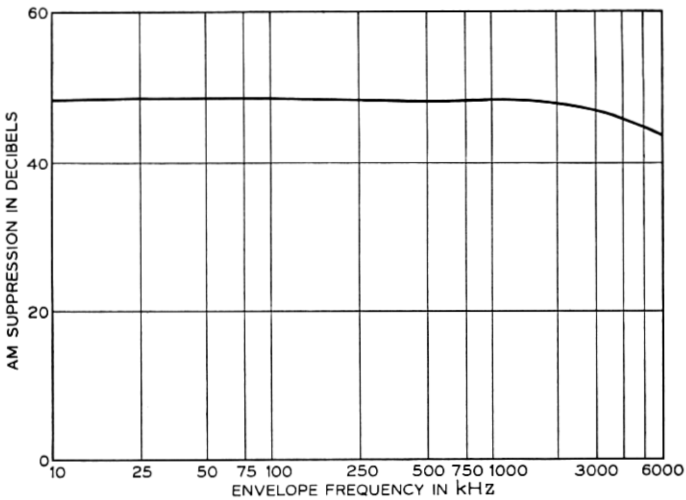

Fig. 19 — Limiter AM suppression vs frequency. f_c = 70 MHz, input = −7 dBm.

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 × 4 × 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.

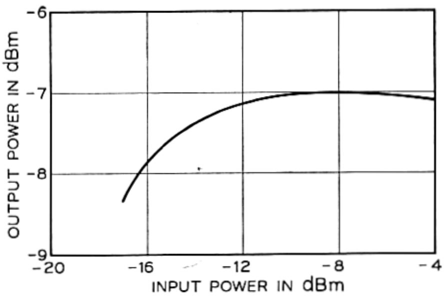

Fig. 20 — Limiter compression vs input power. $f_c = 70$ MHz.

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.

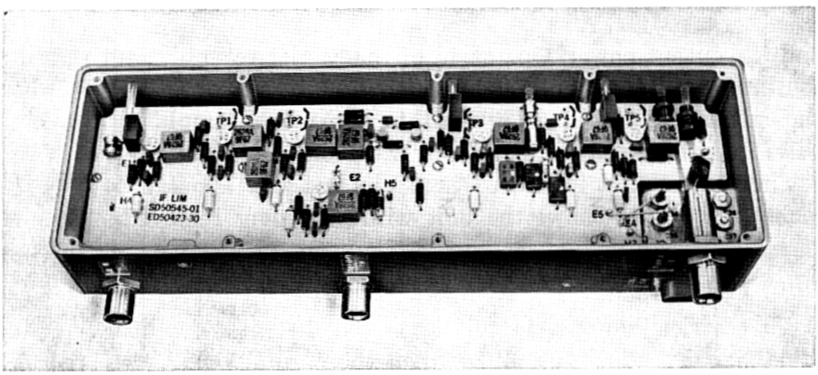

Fig. 21 — Limiter unit.

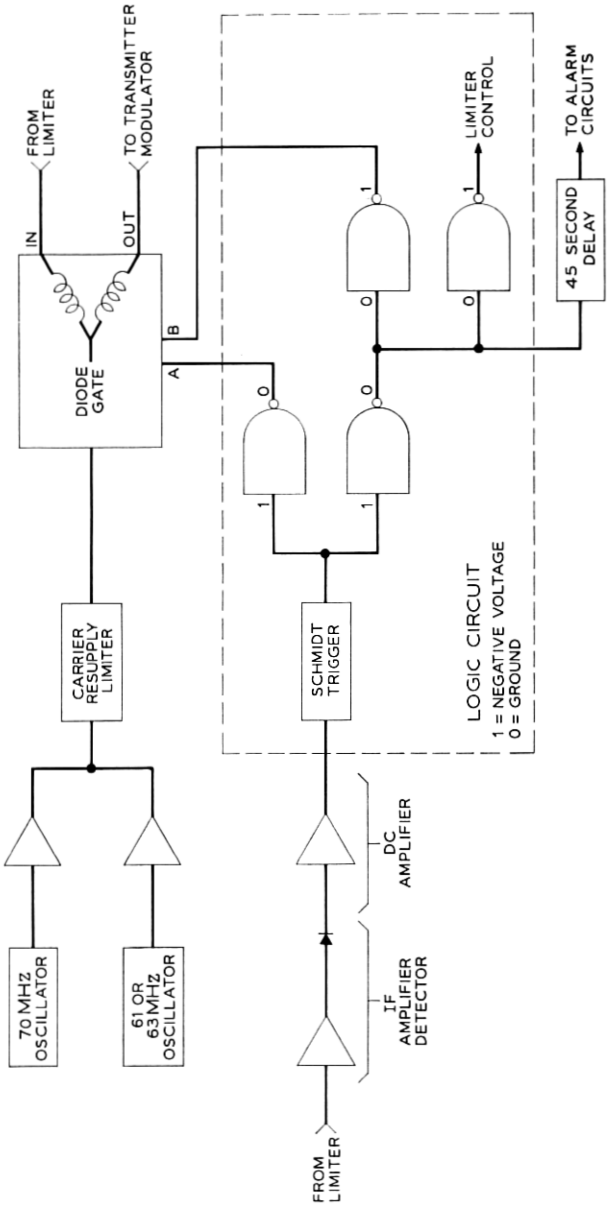

Fig. 22 — Simplified block diagram IF carrier resupply.

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 ±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 ±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

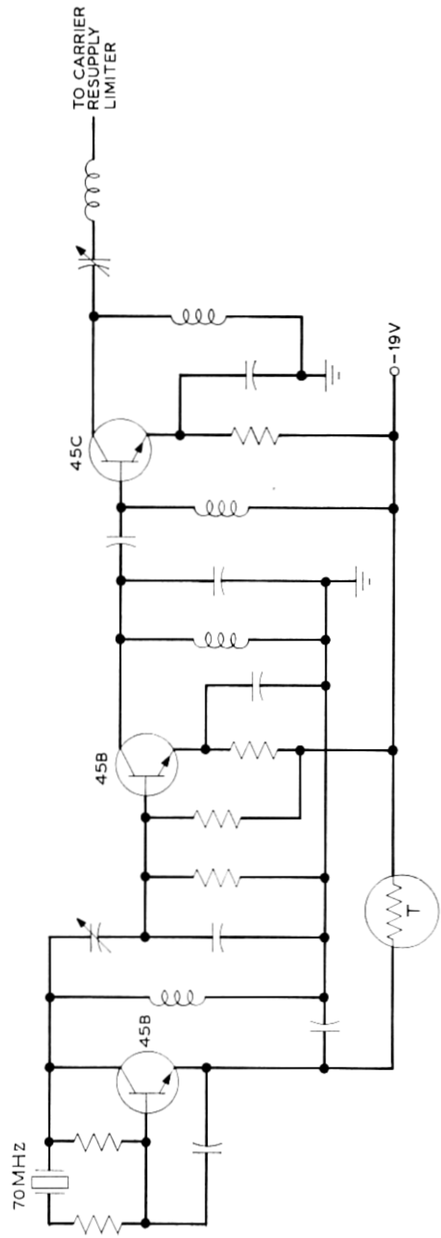

Fig. 23 — Crystal oscillator and amplifier.

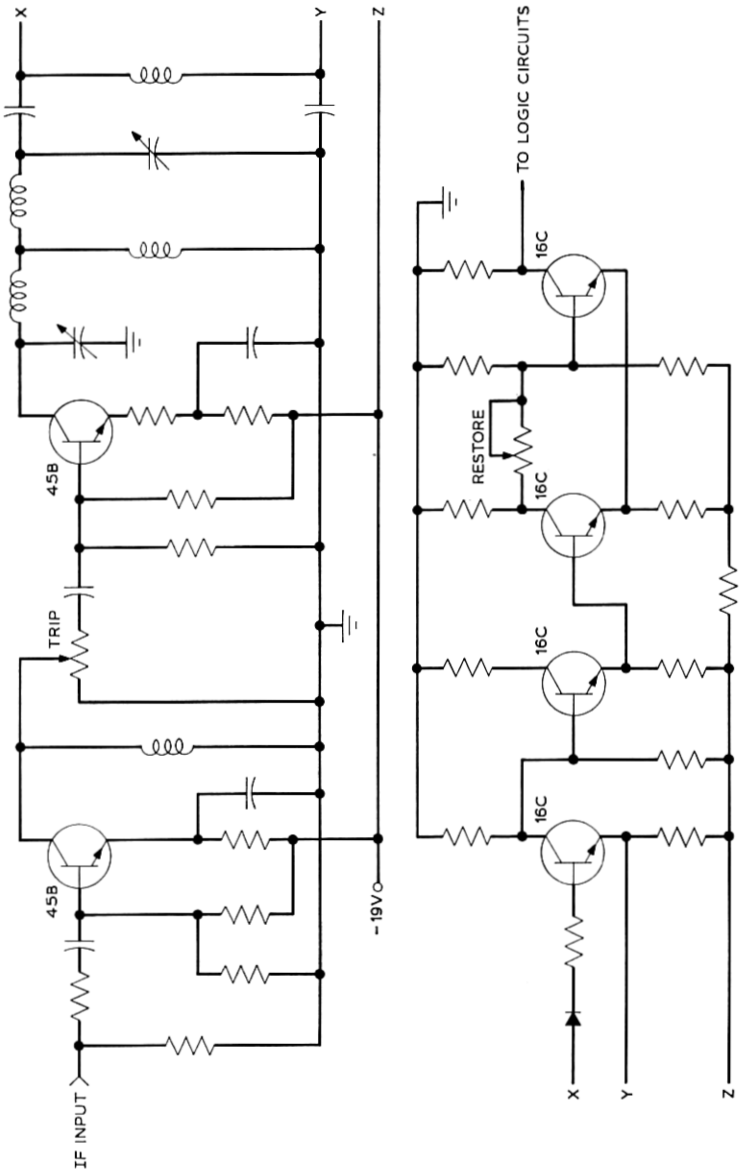

Fig. 24 — IF monitor and control circuits.

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, 19¼ × 4½ × 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.

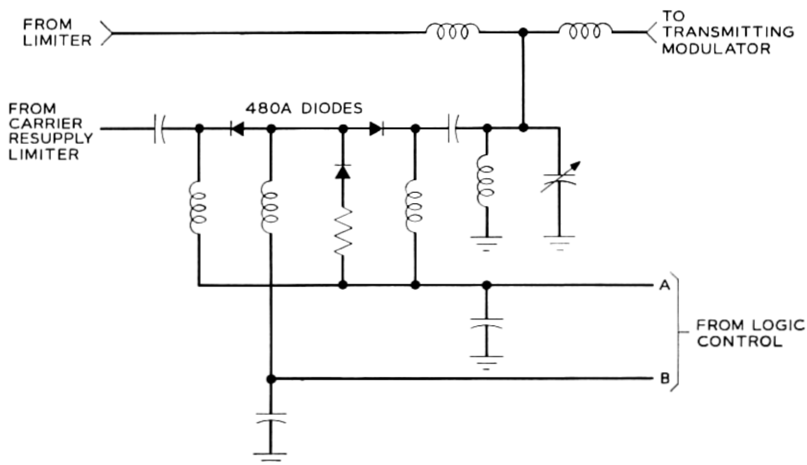

Fig. 25 — Gate circuit.

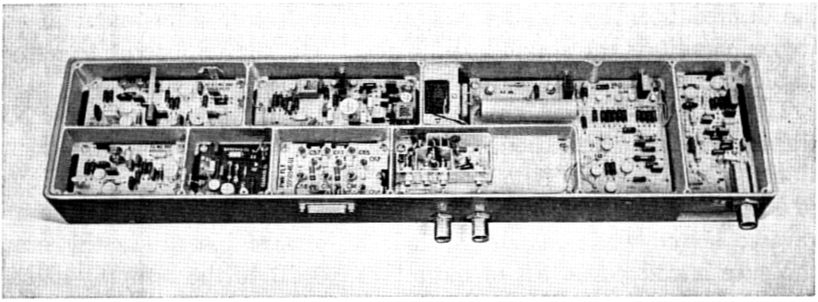

Fig. 26 — IF carrier resupply.

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