The Solid-State Receiver in the TL Radio System

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The availability of reliable high-frequency solid-state devices and the application of new circuit concepts have made possible the development of completely solid-state IF and baseband circuits for the TL microwave radio system. These include (1) a 70-mc IF amplifier with 20-mc 3-db bandwidth, 105 db of gain, and 60 db of automatic gain control, (2) a remodulation-type limiter, (3) discriminator, automatic frequency control, and squelch circuits, and (4) two 6-mc baseband feedback amplifiers. All circuits have been designed to operate over a temperature range of at least −20°C to +60°C. It has been demonstrated that electrical performance comparable to or better than that obtained with electron tube circuits may be achieved while gaining considerably in power drain and reliability. The new circuit techniques and the design considerations which led to their development are presented.

I. INTRODUCTION

When the junction transistor was first announced, it was apparent to many that it would eventually replace the electron tube as an active element in many communications systems. Its small size, low power drain, ruggedness, reliability, and potentially low cost all contribute to its widespread usefulness in the development of new electronic circuits and in the redesign of existing apparatus. The growth in diversity of applications is directly related to the properties of the devices which become available or can be made available in production quantities. This article reports on another step in this expansion — solid-state circuits for a wideband microwave communications system. The development of this new system became both technically feasible and economically practical when diffused-base transistors with excellent high-frequency performance and reliability became available in large quantities at low cost.

Although improvements in device capabilities were very important, the success of the development described herein is due also to the application of new circuit design concepts which differ considerably from conventional electron tube circuit design practice. These innovations and their supporting philosophy will be discussed.

The TL radio relay system as a whole will be described in another article.[1] That article should be consulted for an over-all system description and for the results of early field applications. The present article is restricted to a description of the IF and baseband circuits.

II. TL IF AND BASEBAND CIRCUITS

2.1 *General Description*

A simplified block diagram of the TL receiver is shown in Fig. 1. Let it be mentioned that, in the entire radio system, the only nonsolid-state components are the beating oscillator and transmitting klystrons. Attention in this article is directed to the solid-state circuits which are enclosed by a dashed line in Fig. 1. The IF signal is amplified by the IF

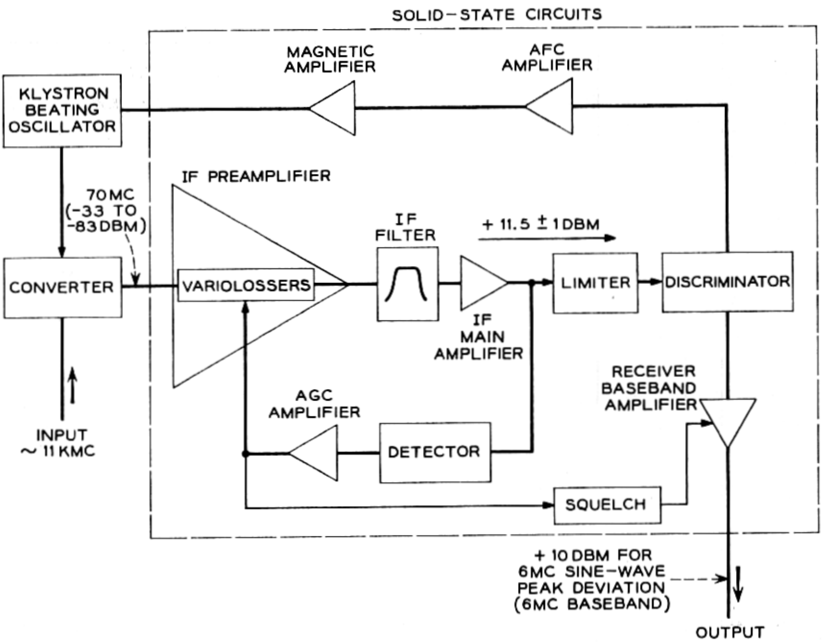

Fig. 1 — Simplified block diagram of wideband FM receiver.

preamplifier and IF main amplifier. A passive filter located between these amplifiers limits the bandwidth. AM noise and an AM component arising from FM-to-AM conversion in the IF amplifiers are suppressed by the limiter. Baseband intelligence is detected in the discriminator. The resulting signal is amplified in the receiver baseband amplifier and is then applied to the transmitter at a repeater or delivered to appropriate terminal equipment. In the transmitter, baseband signals are amplified by the transmitter baseband amplifier (a transistor circuit not shown in Fig. 1). This amplifier drives the repeller electrode of the transmitting klystron.

Three other circuits are included in the receiver: (1) an automatic gain control (AGC) circuit which adjusts the gain of the IF preamplifier to compensate for variations in received signal level; (2) an automatic frequency control (AFC) circuit which controls the voltage on the repeller electrode of the beating oscillator klystron, adjusting its oscillating frequency in such a way as to keep the IF carrier frequency centered in the IF passband; and (3) a squelch circuit which prevents noise from feeding through the receiver during abnormal fades or periods of absence of the incoming carrier.

Fig. 2 is a photograph of the IF and baseband circuits, except for the transmitter baseband amplifier. The four compartments, from bottom to top, contain (1) the IF preamplifier; (2) the passive filter; (3) the IF main and AGC amplifiers; and (4) the squelch circuit, transistor AFC amplifier, limiter, discriminator, and receiver baseband amplifier.

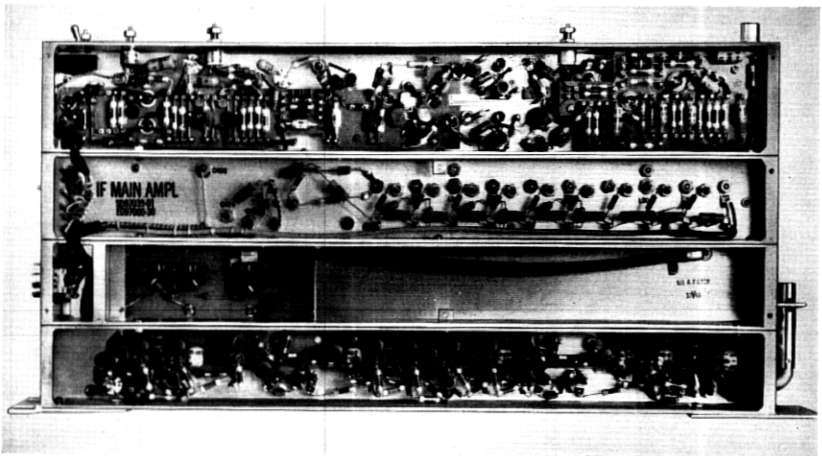

Fig. 2 — IF and baseband circuits.

The transmitter baseband amplifier is packaged separately and is shown in Fig. 3.

2.2 *Some Departures from Conventional Design Philosophy*

2.2.1 *IF Amplifiers and Automatic Gain Control*

In order to utilize the microwave medium efficiently, it is necessary to have a wide IF band which is precisely positioned and defined. In the usual wideband electron tube IF amplifier, the passive interstage coupling networks define the IF band. Variable dc bias can be applied to the tubes to change the gain electronically for AGC without changing the normalized frequency response. This convenient property is due to the fact that the principal band-limiting mechanisms in electron tube IF amplifier circuits — namely, the input and output capacitances — are not strongly bias-dependent. Lossless interstage networks are designed to include these capacitances as elements, and gain-bandwidth product is preserved.

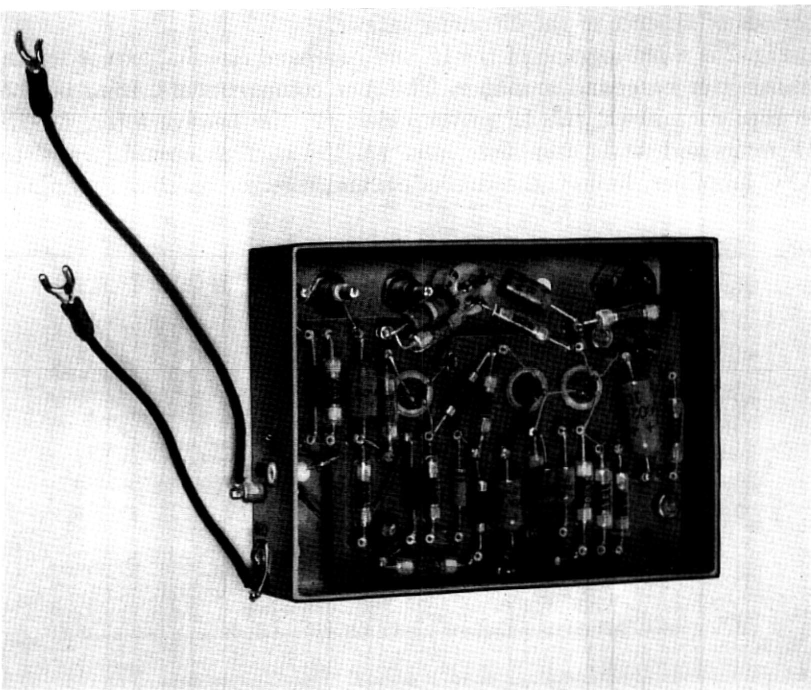

Fig. 3 — Transmitter baseband amplifier.

2.2.1.1 *Inherent Transistor Characteristics.* Certain inherent characteristics of transistors dictate a departure from the conventional electron tube approach.* These characteristics are listed below:

A. The principal mechanism which causes the gain of a transistor amplifier to fall off at high frequencies is transit time — i.e., the finite time it takes minority carriers in the base region to complete their journey from emitter to collector. (This effect occurs in electron tubes, but it has only secondary importance up to 100 mc for electron tubes used in IF service.)

B. Not only are the input and output impedances quite frequency-dependent, but they also depend on the dc bias and on the gain of the transistor. Hence, it is generally not possible to design interstage band-limiting coupling networks which would be satisfactory for a wide range of transistors and dc operating points.

C. A transistor stage exhibits both forward and reverse transmission; hence, the input impedance depends on the load impedance, and the output impedance depends on the generator impedance.

D. A transistor IF amplifier using presently available diffused-base transistors has a wider inherent bandwidth capability than a conventional electron tube IF amplifier.

2.2.1.2 *Design Considerations.* Some design considerations for transistor wideband IF amplifiers are stated below. These follow rather naturally from the characteristics listed above.

A. For wideband applications, the transistor amplifier configuration should be a low-pass rather than a bandpass structure.† Useful low-pass configurations are discussed in the Appendix. The cutoff frequency of the low-pass amplifier is above the upper edge of the IF band, and the over-all IF characteristic is determined by imbedding a passive bandpass filter in the cascade of IF stages. The transistor amplifier stage has a sufficiently large inherent bandwidth to make this technique practical. Because transit-time effect is the primary cause of gain rolloff at high frequencies, extending the bandwidth on the low side does not reduce the gain obtained in the ultimate band.

By using a low-pass configuration, envelope delay distortion, a primary limitation in FM systems, is minimized; and the small amount present is of such a nature that it is easily equalized. A low-pass tran-

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* It is assumed in the following discussion that the IF band falls between the beta and alpha cutoff frequencies of the transistor. This condition is satisfied for the TL radio IF amplifiers.

† This does not mean low-pass in the strictest sense; i.e., the amplifier chain need not pass dc. By low-pass is meant that the frequency rolloff above the IF band is determined by a low-pass structure.

sistor wideband IF amplifier has proven to be far superior to an electron tube IF amplifier from the standpoint of envelope delay distortion.

A further advantage of the use of the low-pass configuration is that adjustment is greatly simplified, because the IF band is a small part of the passband of the active circuits. Also, a wide range of transistor parameters can be tolerated.

The passive bandpass filter, which may include a simple envelope delay equalizer, can be adjusted prior to installation in the IF system if the IF amplifier is designed to present controlled terminating impedances for the filter. The position of the passive bandpass filter should be near enough to the output of the amplifier to prevent out-of-band noise, originating in the IF stages following the filter, from contributing appreciable noise power at the IF output. On the other hand, it must be near enough to the IF amplifier input to prevent intermodulation of out-of-band noise from occurring during weak signal conditions.

B. To obtain a low receiver noise figure, the IF amplifier input transistor stages should be designed to utilize as much available power gain as possible while, of course, taking into account such factors as stability and input impedance. This technique will minimize the effect of the noise figures of the second and third stages on the over-all noise figure. The undesirable departure from the ideal "flat" transmission characteristic, which is inherent in obtaining high gain, can be compensated with an equalizer network following the second or third stage. See the Appendix for a discussion of the "doublet circuit," which uses this principle.

C. The gain variation of the IF amplifier for AGC purposes should be achieved by using separate wideband variolossers. As in the TL system, these may employ semiconductor diodes. This is a departure from the standard technique of varying the gain by changing the dc bias on the amplifier stages. Generally, an intolerable amount of change in the normalized IF transmission characteristic will result if the latter method of gain control is used. The wideband variolossers are passive attenuator networks whose IF transconductances are controlled by the direct current flowing through them. Variolossers can be designed so that the normalized IF transmission characteristic changes only slightly over a wide range of loss settings. The required loss range (60 db for the TL system) must generally be split up between two or more variolossers. The number of variolossers used and their position in the IF chain must be chosen carefully in accordance with the system requirements and the limitations of the variolossers. The maximum IF input level to the variolossers must be restricted so that the diodes remain reasonably linear; otherwise, undesirable IF transmission characteristic changes and excessive AM-to-PM conversion will result. However, the loss must not be allowed to accumulate too close to the input end of the amplifier; otherwise, the system noise figure under strong signal conditions will be unnecessarily high.

As illustration, consider what happens as the input carrier level increases: The automatic gain control system will increase the loss in the variolossers in order to hold the IF amplifier output level constant. Because of this increased variolosser loss, and because of the noise generated in the IF amplifier stages which follow the variolossers, the receiver noise figure will increase. Of course, the IF output signal-to-noise ratio is also increased and a better output signal is obtained. However, the receivers spend most of the time operating under strong signal conditions, so the strong signal S/N ratio (which varies inversely with noise figure) must be kept considerably better than that allowed during localized deep signal fades.* It is important, therefore, to use enough variolossers and to locate them properly in the IF amplifier chain.

It is good design practice to have the passive bandpass filter located between the last variolosser and the output of the IF amplifier to prevent any spurious out-of-band distortion products that might be generated in the variolossers during strong signal conditions from being remodulated into the desired band in a later part of the amplifier.

2.2.2 *Limiter*

The conventional technique for suppressing the AM component of modulation on the IF amplifier output signal is to pass the signal through one or more amplitude limiters. In recent years, these limiters have taken the form of clippers containing semiconductor diodes. When more than one limiter is used, it is necessary to provide buffering amplifiers between them in order to achieve adequate limiting action. Ruthroff[2] has pointed out that this process is inefficient, and he has proposed an improved circuit which has been called *the remodulation limiter*. This circuit derives its efficiency from the fact that it senses the AM present on the incoming IF waveform and then amplitude modulates this IF signal in such a way that the original AM is canceled. A version of a particular form of the remodulation limiter is described in more

* The requirement on strong signal noise figure is based on the cumulative effects of noise in a multihop system, and it is therefore related to the number of hops. On the other hand, the weak signal noise figure requirement is relatively independent of the number of hops because of the localized nature of deep fades.

detail later in this paper. Whereas the clipper-type of circuit typically requires four diodes and two transistor-amplifying stages for 25 db of AM index suppression, a remodulation limiter with equivalent performance contains just two diodes.

2.2.3 *Automatic Frequency Control*

The receiver AFC circuit* in the TL radio system is conventional insofar as it provides a negative feedback loop which centers the IF signal in the IF passband. The klystron-beating oscillator frequency is controlled by an error signal which is sensed at the discriminator, amplified, and impressed on the klystron repeller. The part of the feedback loop between the discriminator and the klystron repeller must necessarily be direct-coupled; other considerations dictate that the repeller voltage be a high negative voltage, about −500 volts. The problem of direct-coupling the transistor circuits, which operate near zero volts, to the klystron repeller has been overcome by the use of a magnetic amplifier (magamp). The magamp serves to completely isolate the low- and high-voltage circuits while effectively maintaining direct coupling and adding to the AFC feedback loop gain.

2.3 *Circuit Description*

2.3.1 *IF Amplifiers*

2.3.1.1 *IF Preamplifier*. The input circuit of the IF preamplifier (Fig. 4) consists of a pair of direct-coupled common-emitter stages followed by a high-pass filter. The principal advantage of this combination, which is called a "doublet," is that it yields the best noise figure of the various circuits investigated† as well as an acceptable input return loss. The remaining stages of IF amplification consist of wideband common-base, transformer-coupled circuits. To obtain good transistor interchangeability and insensitivity to temperature change, the stages are padded and mismatched, resulting in a power gain for each common-base stage of about 6 db. Two diode attenuator networks, the variolossers, are included in the preamplifier, dividing it into three approximately equal gain segments. Their placement is such that they have a negligible (<0.1

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* Transmitter AFC is not required in the TL radio system because of the specially designed klystron and klystron cooling system.[3]

† Circuits investigated included the common-base stage with a wideband transformer interstage network, the doublet circuit, and the common-emitter stage with frequency-dependent shunt feedback. These configurations are compared in the Appendix.

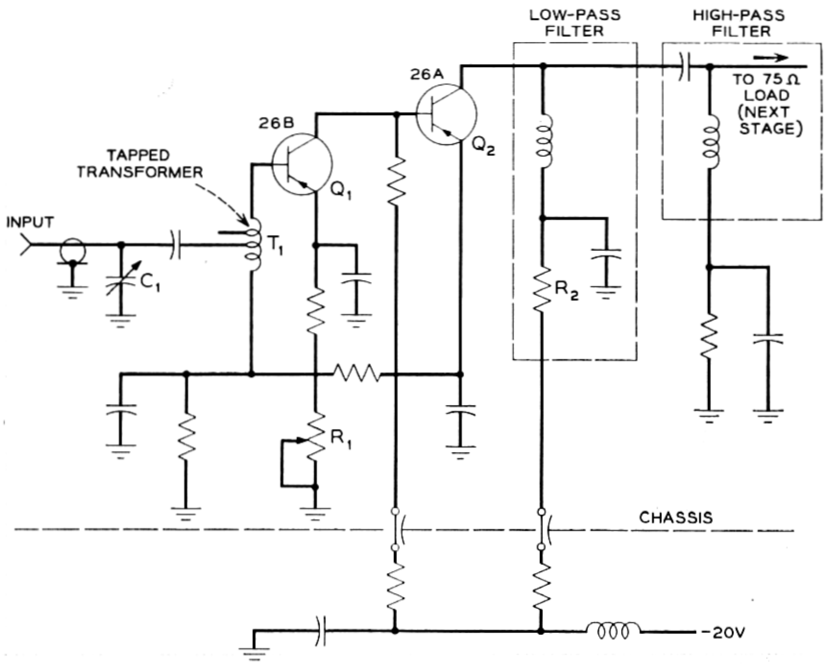


Fig. 4 — Low-noise input section.

db) effect on the noise figure under weak-signal conditions and remain sufficiently linear under strong-signal conditions. The preamplifier maximum gain is adjusted at room temperature to 57 db, flat between 64 and 76 mc within about ±0.2 db.

Input return loss adjustments are applied in the form of potentiometer R_1 (which varies the bias current in transistor Q_1), taps on transformer T_1, and variable capacitor C_1 (Fig. 4).

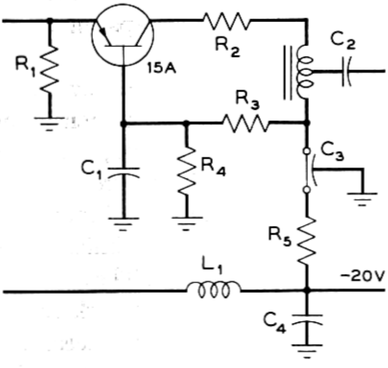
The two common-emitter stages of the doublet exhibit a downward transmission slope approaching 12 db per octave in the IF band; the high-pass filter, which has a cutoff at approximately 100 mc, equalizes this rolloff. This combination yields a very good noise figure because the available power gain provided between the preamplifier input jack and the input of transistor Q_2 is relatively large, thus minimizing the influence of the second transistor on the over-all noise figure. The constant resistive load for transistor Q_2 consists of a high-pass network and a low-pass network, which have complementary input impedances. Resistor R_2 is the termination for the low-pass network, and the third transistor stage terminates the high-pass network.

A diagram of a typical common-base stage is shown in Fig. 5. In this configuration, the transistor has a current gain slightly less than unity in the IF band. The interstage transformer, which is a ferrite-core distributed autotransformer,[4,5,6] provides a current step-up ratio of approximately 1:2, making the over-all power gain slightly less than 6 db per stage. Since a transistor of the type used is capable of providing a maximum unilateral gain of about 15 db at 80 mc, there is evidently a considerable sacrifice of gain in order to obtain a high degree of stability and transistor interchangeability. The slope of the gain-frequency characteristic in the IF band can be adjusted by changing the damping resistor R_2, which is introduced to control gain peaking at higher frequencies. (Also, a variable inductor is added in the base lead of some of the stages to provide a small, continuous, additional slope adjustment.)

Fig. 6 indicates the make-up of the complete preamplifier. Potentiometer R_4 is a gain control having 16 db range. Used to adjust the over-all IF amplifier gain to 105 db, it does not unduly affect the good output return loss of the preamplifier. (A good termination is needed for proper filter operation.)

2.3.1.2 *Variolossers.* Two variable-loss pi networks of germanium point-contact diodes are included in the preamplifier (Fig. 6) to maintain a constant level out of the main amplifier. The input levels from the converter may vary from about -33 dbm to less than -83 dbm. Each variolosser is able to insert from 1 to 30 db of flat loss over the IF band. This loss range, greater than that required to correct for IF input level variations, allows for temperature and aging effects on the gain of the amplifier.

The impedance of each variolosser diode to signal frequencies is



TYPICAL VALUES:

$R_1 = 825$, $R_2 = 464$, $R_3 = 2150$

$R_4 = 1780$, $R_5 = 10$, $L_1 = 7$

$C_1 = 0.0005$, $C_2 = 0.005$, $C_3 = 0.001$, $C_4 = 1$

RESISTANCE IN OHMS

INDUCTANCE IN μH

CAPACITANCE IN μF

Fig. 5 — Typical common-base stage.

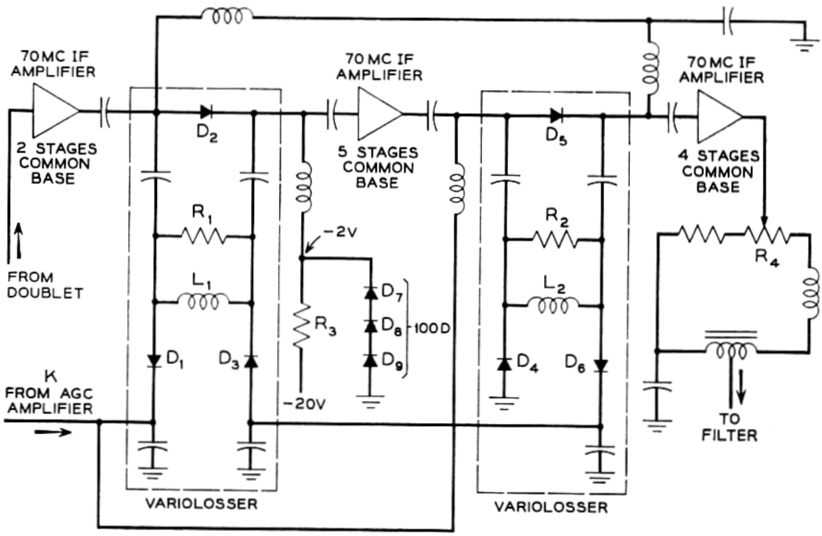


Fig. 6 — Gain control circuits in the preamplifier.

controlled by the direct current passing through it. Direct current is caused to flow in such a manner that as the current increases in the shunt diodes, it decreases in the series diodes (and vice versa). Since the diodes appear in pi networks at IF, variable loss is obtained without producing much variation of the input and output impedances of the variolossers. The currents are controlled by the output of the AGC amplifier (see Fig. 8 below), which responds to amplitude-modulation frequencies ranging from dc up to about 50 cps. Diodes D_7, D_8 and D_9 are used as a -2.0-volt dc supply and are forward-biased through resistor R_3. (These diodes become starved of current under fade conditions, and the reduced voltage ensures a low minimum IF loss in the variolossers.) The series and shunt diode sets are connected in parallel for incremental currents supplied from the K lead; but the -2.0-volt source is placed within a loop passing through all six attenuator diodes, thus providing a condition wherein their dynamic resistances can simultaneously equal about 130 ohms. This condition occurs when the input dc lead from the AGC amplifier carries no current, and it corresponds to a medium loss condition in the variolossers. The two variolossers are so interconnected that they conduct the same dc currents, thereby forcing their loss values to track together. Inductors L_1 and L_2 carry control current and also counteract the effect of the series-diode capacitance, which is important when the diode resistances reach their highest level. Resistors R_1 and R_2 provide an upper limit for the series path impedance, thereby forcing the shunt diodes to carry more of the loss burden at higher IF input signal levels.

A network containing a thermistor applies a temperature-dependent bias voltage in series with the shunt diodes of the variolossers. The effect is to equalize the drift which tends to occur in the AGC amplifier output current, and thus to prevent drift in the squelch firing level. The thermistor network, not shown in Fig. 6, is inserted between diode D_4 and ground.

2.3.1.3 *IF Filter* (Fig. 1). The main functions of the IF filter are to delimit the IF bandwidth precisely with a minimum of ripple or slope within the band and to equalize the delay of the over-all IF amplifier. Besides limiting thermal noise, it also prevents out-of-band interfering signals and harmonics generated in the variolossers from entering the main amplifier. Systems considerations[1] dictate that the 3-db frequencies be 60 and 80 mc, and that the loss be flat to within ± 0.1 db from 64 to 76 mc. It is designed to work between precise 75-ohm terminations.

2.3.1.4 *IF Main Amplifier* (Fig. 7). The IF main amplifier is a cascade of nine common-base stages, each developing slightly less than 6 db of gain, and a parallel common-base output stage capable of delivering a maximum power of $+13$ dbm into the limiter. The nominal power gain of this amplifier is 48 db, flat to within ± 0.2 db over the 64- to 76-mc IF band.

The driver stage Q_9 is similar to the earlier stages, except that it is followed by an additional transformer to give an over-all 4:1 current step-up at the last interstage (for driver linearity). Resistors R_2 and R_3 ensure equal driving currents for the parallel transistors. Resistor R_4 helps to provide a good output return loss for a 200-ohm load (the limiter).

2.3.2 *Automatic Gain Control Circuit* (Fig. 8)

For proper operation of the limiter, it is necessary that the output level of the IF main amplifier be held nearly constant regardless of changes in RF signal level and temperature. Furthermore, the IF amplifier stages preceding the output stage must not be overdriven; otherwise, too much spurious phase modulation will result. These desired conditions are achieved through the use of an automatic gain control circuit consisting of an IF detector, a direct-coupled amplifier called the AGC amplifier, and the variolossers described in Section 2.3.1.2. The AGC amplifier, excited by the detector, feeds back a current (K lead) to control the loss due to the variolossers and thereby

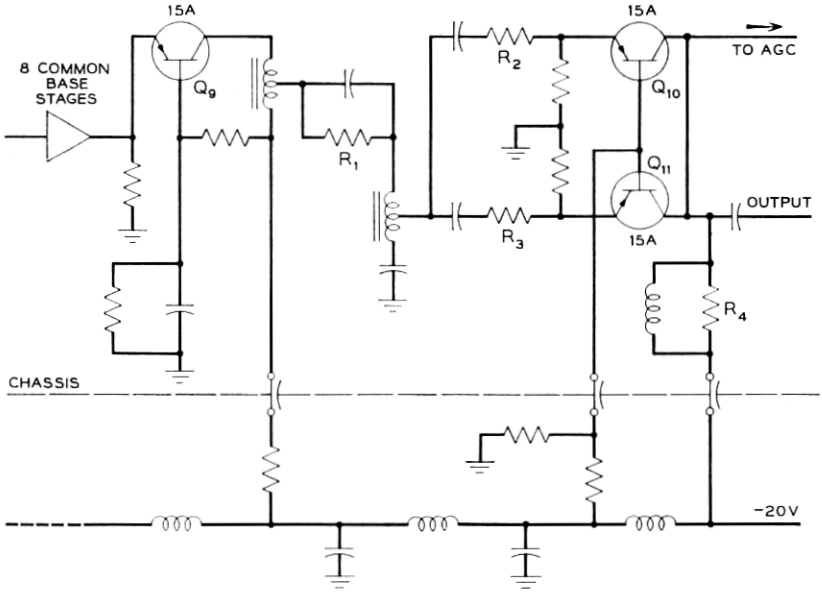


Fig. 7 — Main IF amplifier.

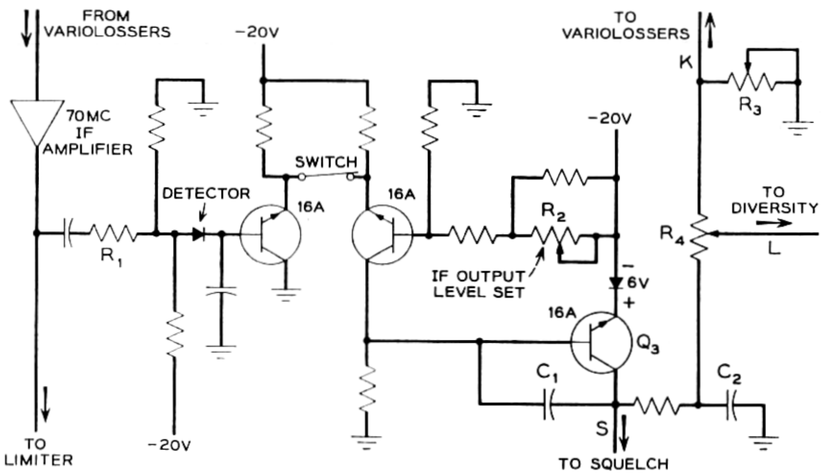


Fig. 8 — AGC amplifier.

compensates any tendency for the IF output level to change. The dc amplifier has sufficient gain to multiply the small current increment drawn from the detector to the relatively large current needed to drive the variolossers. This circuit holds the limiter input level to +11.5 dbm ±1 db for IF input levels ranging from −83 dbm to −33 dbm over the temperature range from −20°C to +60°C.

A masking resistor R_1 is inserted to partially isolate the output of the IF main amplifier from the nonlinear input impedance of the AGC detector. This reduces the AM-to-PM conversion of the receiver. The detector-diode voltage varies almost linearly with the IF output voltage and is therefore used as a limiter input level monitor.

The dc amplifier consists of a differential stage followed by a common-emitter stage. The differential stage is used to minimize the drift in IF output level due to changes in temperature. The amplifier has one net phase reversal, providing negative feedback around the AGC loop. The closed-loop bandwidth of the AGC system is limited to about 50 cps by capacitors C_1 and C_2, which produce the only significant cutoffs occurring in the feedback-vs-frequency characteristic of the AGC system.

A switch is incorporated in the differential stage to permit the AGC loop to be opened for test purposes. The potentiometer R_2 serves a dual purpose. With the AGC loop closed (switch on), it sets the dc reference to which the detector output level is compared; and since the variolossers are automatically adjusted to make the difference between these levels zero, this potentiometer sets the IF output level. When the loop is open, the potentiometer is used to manually adjust the loss of the variolossers.

The collector voltage of transistor Q_3 is a monotonic function of the received carrier level. This voltage is used to trigger a squelch circuit (lead S), to excite a diversity switching circuit (lead L) and to drive a signal strength meter (also lead L). Potentiometer R_3 serves as a calibration control for this voltage, and potentiometer R_4 provides an additional adjustment for the voltage on lead L.

2.3.3 *Limiter* (Fig. 9)

The output signal of the IF main amplifier will contain both AM and FM components of noise and baseband signals.* Since the discriminator will respond to AM signals as well as to FM signals, a limiter is used to greatly reduce the AM component of the signal entering the discriminator.

* The AM baseband signal component is due to the action of a non-flat system transmission characteristic on the FM signal.

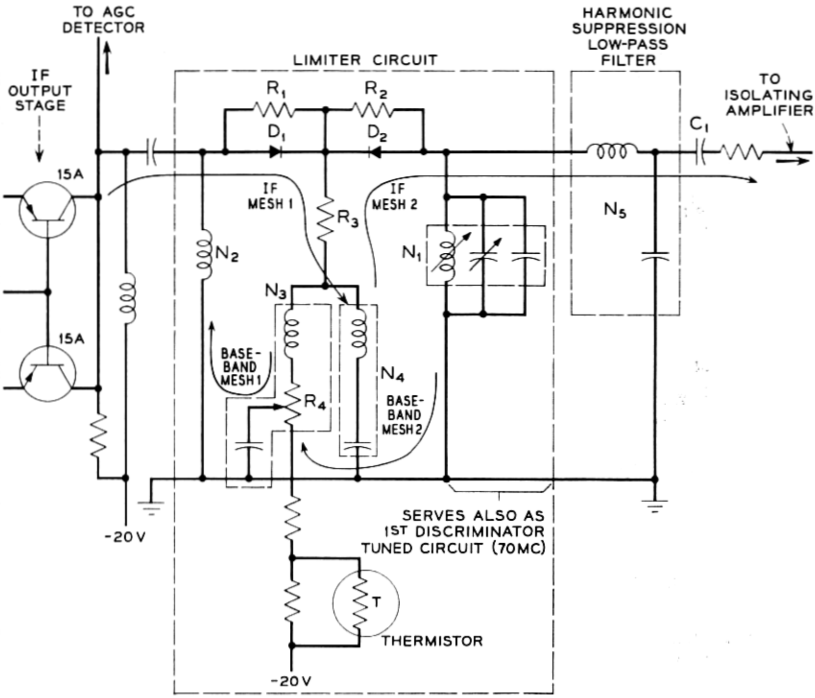


Fig. 9 — Remodulation-type limiter circuit.

The operation of the limiter can be understood from a consideration of Fig. 9.[2] The limiter resembles a simple series diode clipper in which the diodes are used to open up the transmission path after a certain level threshold is reached. One can think of the first and second diodes as an amplitude detector and an amplitude modulator, respectively. This process of detection and modulation is so performed that the net AM on the limiter output signal is minimized. Some of the incoming AM sideband energy is coupled from the IF mesh 1 to IF mesh 2 through the shunt path consisting of resistor R_3 and bandpass network N_4. The envelope of the IF signal is detected by diode D_1 in baseband mesh 1 and this baseband signal is coupled through the shunt path consisting of resistor R_3 and low-pass network N_3 to baseband mesh 2. There the baseband signal is impressed on the IF carrier by diode D_2 and appears across the load at sideband frequencies in phase opposition to the energy coupled from IF mesh 1. Note that networks N_1 and N_5 are required to cause the baseband and sideband signals to flow in the proper meshes and to attenuate undesirable modulation products. (Also, N_1 is a 70-mc antiresonant tank circuit which serves as a corrective circuit for linearizing the discriminator characteristic.) Potentiometer R_4 is adjusted to achieve cancellation of the AM. The dc source in the baseband shunt path is used to bias the diodes D_1 and D_2 and hence set the limiter output level. The thermistor compensates for temperature drift in the gain (or loss) of the limiter, discriminator and baseband amplifiers. The FM signal behaves in the same manner as the carrier, suffering only loss in passing through the limiter. Insertion loss of the limiter is about 10 db, and the circuit has been designed for acceptable limiting for input levels ranging from +10 to +13 dbm.

Resistors R_1 and R_2 shunt the diodes, thereby permitting the use of diodes with a wide range of reverse impedances. These resistors, being in parallel with the low forward impedances, do not affect diode performance in the forward bias state. The load driven by the limiter is the input network of the discriminator. Capacitor C_1 tunes out the reactive part of the input impedance of the limiter-discriminator isolating transistor at the IF center frequency.

2.3.4 *Triple-Tuned Balanced Discriminator* (Fig. 10)

The discriminator extracts the baseband information from the input FM signal. After passing through the first common-base isolating amplifier, the FM signal traverses a wide-band transformer and then drives two separate branches. Each of these paths contains a common-base isolating amplifier, a parallel resonant circuit, an amplitude detector, a low-pass filter, and two terminating resistors. The discriminator output baseband signal is the sum of the output currents of the two paths. Two outputs are provided: an ac-coupled output to the receiver baseband amplifier and a direct-coupled output to the AFC amplifier.

Circuit N_1 (Fig. 9) is tuned approximately to the 70-mc carrier frequency, circuit N_2 (Fig. 10) to 85 mc, and circuit N_3 to 55 mc. The use of circuits N_2 and N_3 alone yields the familiar "S" curve; however, the low-Q tuned circuit N_1 significantly improves the linearity of the discriminator.[7] Since the attainment of adequate linearity requires precise adjustment of both the Q and resonant frequency of the tuned circuits, both the inductor and capacitor of each tank are adjustable. Proper phasing of the outputs of the two paths is accomplished by connecting the discriminator diodes as indicated. (This connection avoids the use of a costly and large-size broadband transformer.)

Negative-coefficient capacitors are used to maintain the proper resonant frequencies of the two tuned circuits of Fig. 10 as the temperature changes. Equal forward-bias voltages of approximately 0.3 volt are

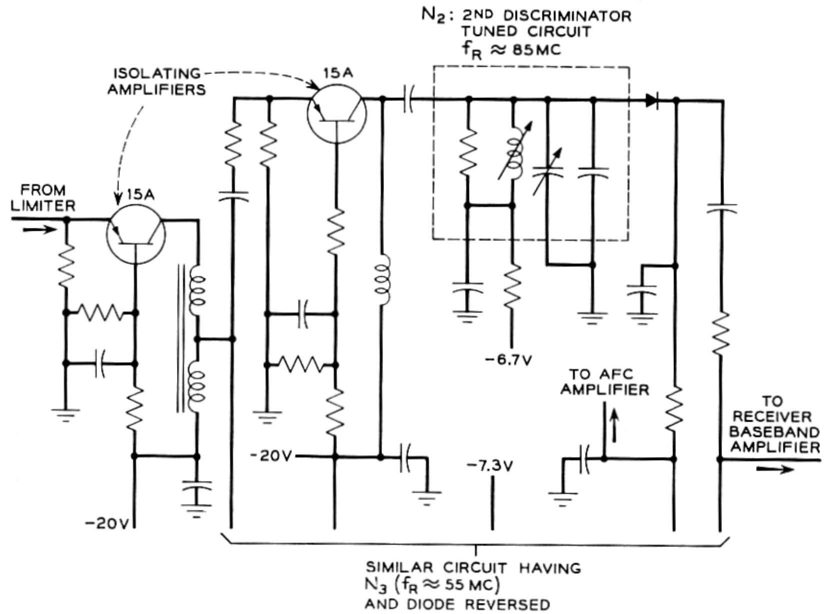


Fig. 10 — Balanced discriminator.

maintained across each of the diodes to improve the linearity of the detectors.

2.3.5 *Automatic Frequency Control Amplifier* (Fig. 11)

The AFC loop contains two cascaded amplifier circuits, a two-transistor amplifier followed by a magnetic amplifier. The primary purpose of the transistor AFC amplifier is to isolate the discriminator and baseband amplifier from the magnetic amplifier, since the latter amplifier produces a substantial 1800-cycle signal rich in harmonics which must be attenuated to avoid spurious baseband amplifier output tones.

The transistor AFC amplifier consists of a common-emitter stage direct-coupled to a common-collector stage. Shunt negative feedback provides bias and gain stabilization and low input and output impedances. (The low input impedance helps isolate the two paths of the balanced discriminator, and it simultaneously stabilizes the bias applied to the discriminator diodes.) Attenuation of undesirable tones from the magnetic amplifier is provided by the low output impedance of the transistor AFC amplifier and by an RC filter between the discriminator and the transistor AFC amplifier. The input current from the discrimi-

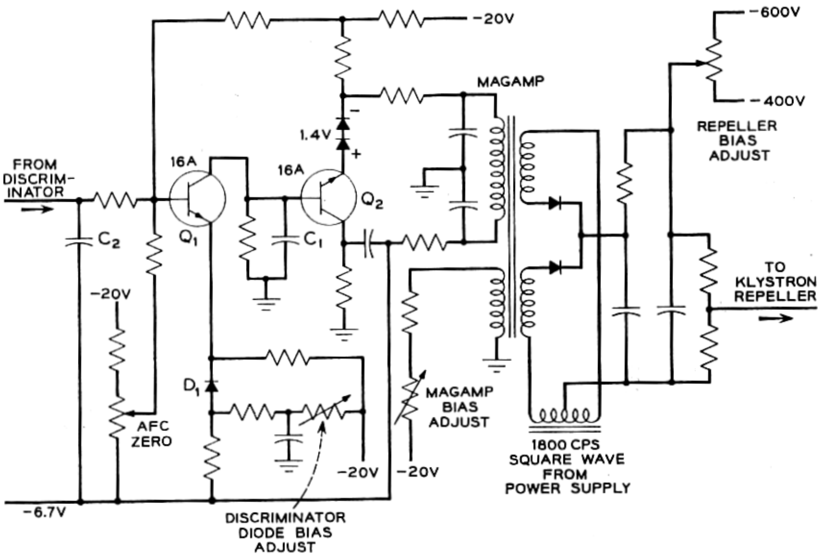


Fig. 11 — AFC amplifier.

nator is very nearly zero when the IF carrier frequency is 70 mc because the discriminator diodes are oppositely poled.

Since the transistor AFC amplifier output voltage must be bipolar and since the dc power supply has only one polarity (−20 volts) with respect to dc ground, a −6.7 volt dc voltage source is used as the transistor AFC amplifier "dc ground." The dc output of the AFC amplifier is relatively insensitive to variations in both the −20-volt power supply and the −6.7-volt source. The transistor AFC amplifier output is approximately 40 mv/mc of carrier deviation.

Temperature stability of the transistor AFC amplifier is achieved by canceling Q_1 base-emitter voltage-drop changes with diode D_1 and by achieving beta drift compensation by making the collector current of Q_1 and the base current of Q_2 approximately equal in magnitude. Capacitor C_1 is used to shape the feedback versus frequency characteristic of the transistor AFC amplifier and assures stability. Capacitor C_2 reduces 1800-cycle energy originating in the magnetic amplifier.

The main gain-producing element of the AFC loop is the magamp, which provides a voltage gain of about 950. Its cutoff frequency, which controls the response time of the AFC loop, is about one cycle per second. The 1800-cycle square wave required for operation of the magamp is obtained from a winding on the dc-to-dc converter which supplies high voltage for the klystrons.

2.3.6 *Receiver Baseband Amplifier* (Fig. 12)

The receiver baseband amplifier follows the discriminator. It is a direct-coupled cascade of three common-emitter transistor stages with shunt feedback through a "T" network. Two virtually identical input currents are applied to the amplifier from the two branches of the discriminator. The external current gain of the amplifier is given, to a good approximation, by the ratio of the feedback network short-circuit transfer impedance (13 to 25 kilohms) to the output-lead resistance (145 ohms). The gain may be varied over a 5-db range by adjusting potentiometer R_5.

The output stage is a parallel combination of a pnp germanium unit

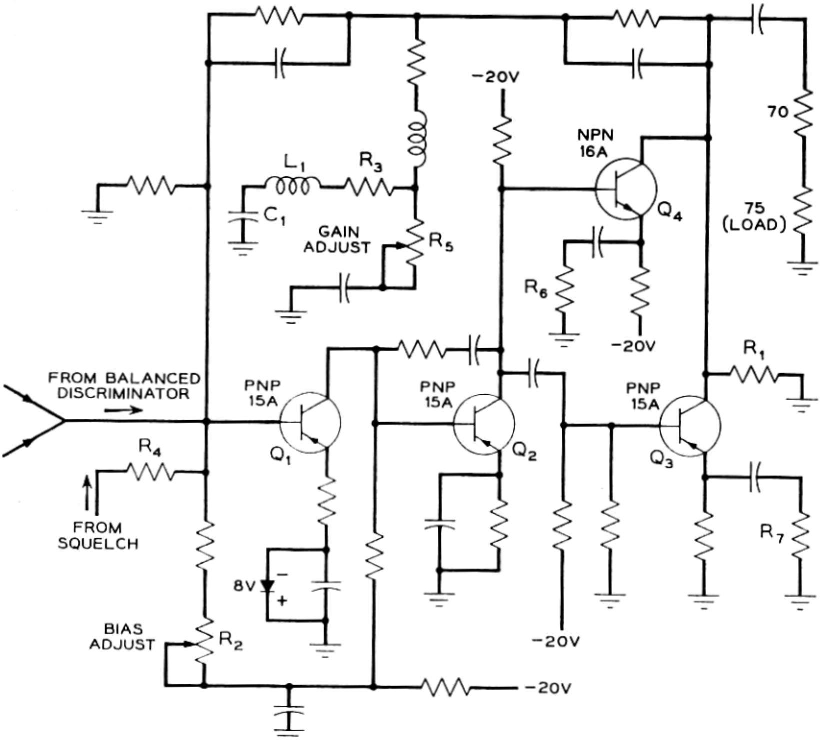


Fig. 12 — Receiver baseband amplifier.

having a common-emitter current-gain cutoff at about 15 mc and an npn silicon unit with a cutoff at about 3 mc. Both transistors contribute substantially to the ac load current out to 6 mc (the highest baseband frequency), with the result that more current can be driven into the load under given distortion limits than can be provided by one transistor alone. The dc current passed by transistor Q_3 into the collector node is drawn out of that node by transistor Q_4, allowing a reduction of the output admittance shunting the load. The available output power is thus increased. Resistor R_1 sets the amount by which the bias current in the silicon transistor Q_4 exceeds that in the germanium transistor Q_3. The small emitter resistors R_6 and R_7 ensure that the two transistors share the load equally by making their input impedances proportional to their respective incremental current gains.

The local shunt feedback in the second stage allows the circuit to accommodate high-gain units in the output stage without becoming unstable; and it also compensates, to some degree, for the drop in incremental gain of the output stage during part of the signal cycle, maintaining the over-all feedback and thereby reducing distortion.

Because of the large amount of over-all shunt feedback, the output impedance of the amplifier is roughly 5 ohms augmented by the 70-ohm padding resistance, which yields a good output return loss. Potentiometer R_2 adjusts the dc collector voltage of transistors Q_3 and Q_4. The network formed by resistor R_3, inductor L_1 and capacitor C_1 improves the stability margins when potentiometer R_5 is set for maximum resistance.

The amplifier is switched off during deep carrier fades by an input from a squelch circuit sufficient to drive it into saturation. This input current is passed through resistor R_4, which is large enough to mask out the output capacitance of the squelch circuit.

2.3.7 *Transmitter Baseband Amplifier* (Fig. 13)

A baseband amplifier immediately precedes the transmitter klystron in the TL system. This amplifier provides an adjustable voltage gain of 27 $\pm$ 4 db between a 75-ohm source and the klystron repeller load (which, including wiring capacitance, behaves like a 35-$\mu\mu$f capacitor).

The amplifier uses three transistors. The first stage uses the common-base configuration; the second stage is common-collector; and the third stage is common-emitter, providing the net phase reversal needed for negative shunt feedback. Because of the very low input impedance provided by the common-base stage and the over-all negative feedback, a 75-ohm resistor is added to give excellent input return loss.

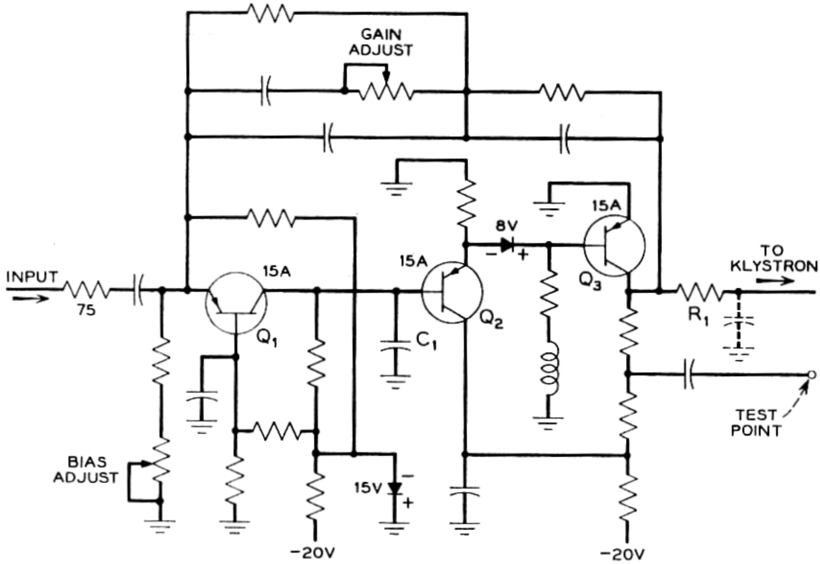


Fig. 13 — Transmitter baseband amplifier.

Capacitor C_1 provides the initial feedback cutoff (at about 5 mc). This cutoff slides to compensate for changes in the current gains of Q_2 and Q_3 since it depends on the input impedance of the common-collector stage. Resistor R_1 improves the phase margin of the feedback ($A\beta$) by halting the falling asymptote due to the 35-$\mu\mu$f load capacitance at about 47 mc.

2.3.8 *Squelch Circuit* (Fig. 14)

The receiver contains a bistable trigger circuit which renders the baseband amplifier inoperative during deep radio path fades. This

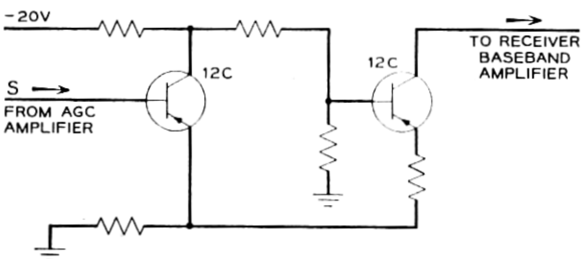


Fig. 14 — Squelch circuit.

circuit is a Schmitt trigger operating from the variolosser-control voltage produced by the AGC amplifier. This voltage, which appears at the collector of transistor Q_3 in Fig. 8, becomes less negative as the IF input carrier fades, and the circuit is designed to trigger in the vicinity of -2.8 volts. This voltage level is caused to correspond to a carrier level of from -80 to -95 dbm by means of potentiometer R_3 in Fig. 8. When the Schmitt trigger circuit fires, it injects about 1.5 ma of current into the summing node of the receiver baseband amplifier, driving this amplifier into saturation since as little as 0.8 ma will saturate it.

2.4 *Circuit Performance*

2.4.1 *IF Amplifier*

The most important electrical performance features of the IF amplifier in the TL radio receiver may be described by graphs showing the over-all gain-frequency characteristic at different incoming signal levels and at extremes of temperature.

All of the curves in Fig. 15 are normalized with respect to the gain at 70 mc and 25°C with the passive filter in. The data show that both the in-band and wideband spectral distortion is small as the ambient temperature is changed from -20°C to $+60$°C. Furthermore, the gain level at 70 mc without AGC varies a maximum of 2 db over the same temperature range. Between 50 and 90 mc the "bowed" shape of the curves is primarily due to the output circuit of the IF amplifier (Fig. 7); the in-band effect is negligible.

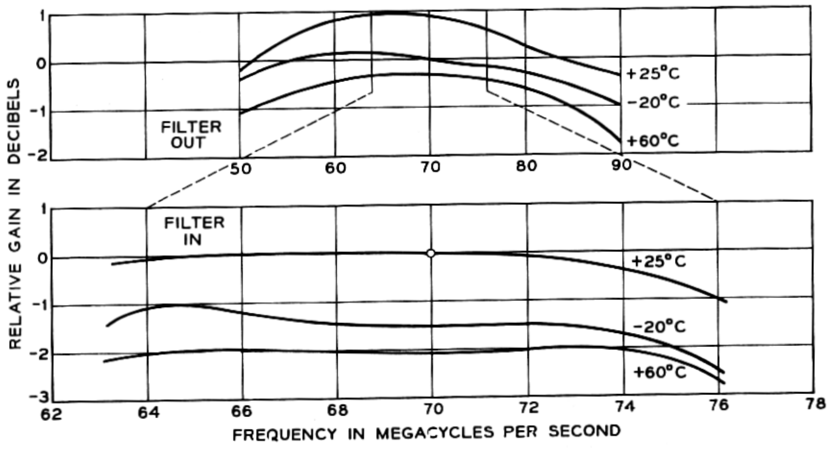


Fig. 15 — Gain-frequency characteristic of IF amplifier.

The IF amplifier output level vs frequency at different input levels and temperature extremes is shown in Fig. 16. These curves demonstrate the tightness of the AGC.* The output power is set to +11.5 dbm at 70 mc, 25°C with an input level of −58 dbm. The maximum level shift encountered at 70 mc as temperature and input level vary is ±0.6 db, and the maximum tilt is 0.13 db/mc across the band. The IF main amplifier provides +11.5 dbm of power with less than 0.5 db compression working into the 200-ohm limiter input impedance.

In addition to transmission-shape and output-level stability, another important requirement is that the envelope delay distortion be small. Fig. 17 shows the envelope delay in nanoseconds at two input levels and at temperature extremes. The maximum change in envelope delay over the 64- to 76-mc band is observed to be 12 nanoseconds. Most of the change with frequency occurs during deep fades where noise, rather than delay distortion, is controlling. The envelope delay characteristic of the IF filter is presented in Fig. 18 along with the insertion and return loss characteristics. The 70-mc loss is about one db, and the envelope delay shift across the 64- to 76-mc band is 6 nanoseconds.

The noise figure of the IF amplifier during fades, measured from a 75-ohm source impedance, is typically about 6 db. Fig. 19 is a plot of noise figure as a function of input carrier level in dbm at 25°C. The

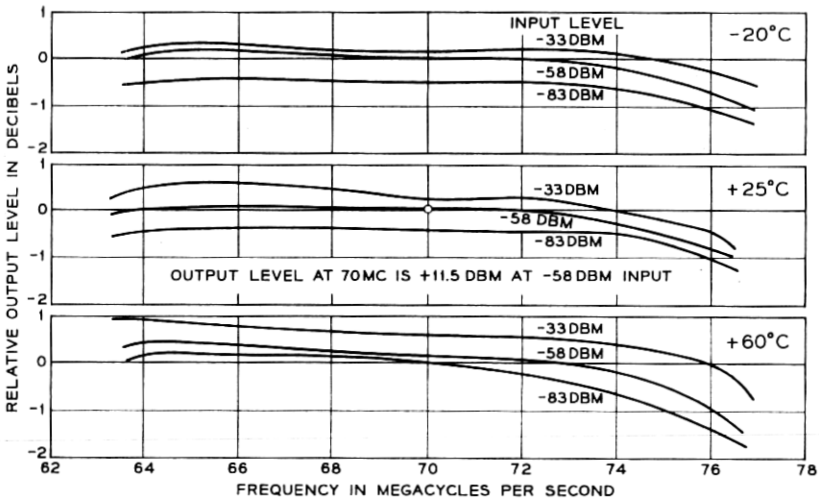


Fig. 16 — IF output level versus frequency (with filter and AGC).

* The frequency response of the AGC system, for the purpose of this measurement, is made much lower than the sweep frequency of the signal source.

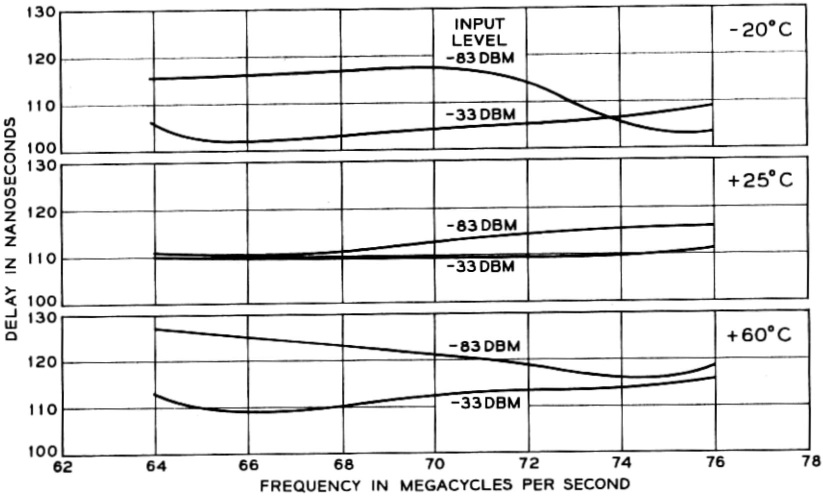


Fig. 17 — Delay-frequency characteristic of IF amplifier (with filter and AGC).

input circuit is adjusted so that a compromise is reached between low noise figure and return loss, typically 26 db from 64 to 76 mc.

2.4.2 *AGC Circuit*

The stability of the closed AGC feedback loop may be measured by applying a step function from a current source at the K lead of the AGC amplifier (Fig. 8) and observing the output voltage on the collector of Q_3 as a function of time. Fig. 20 shows a sketch of this waveform for three input level conditions representing different amounts of AGC loop gain. An input level of -58 dbm yields the worst transient response. The amount of overshoot is a measure of the stability margins, and the rise time and ripple frequency relate to the response time. The AGC system cuts off at approximately 50 cps, and the maximum overshoot of 6 db suggests a minimum phase margin of about 20 degrees.

2.4.3 *Limiter*

Measurements of AM suppression of the limiter for three temperatures and three drive levels into the IF preamplifier are given as functions of baseband frequency in Fig. 21. The limiter has been optimized for maximum AM suppression at 100 kc, room temperature, and a $+11.5$-dbm drive level and still provides at least 18-db suppression over the

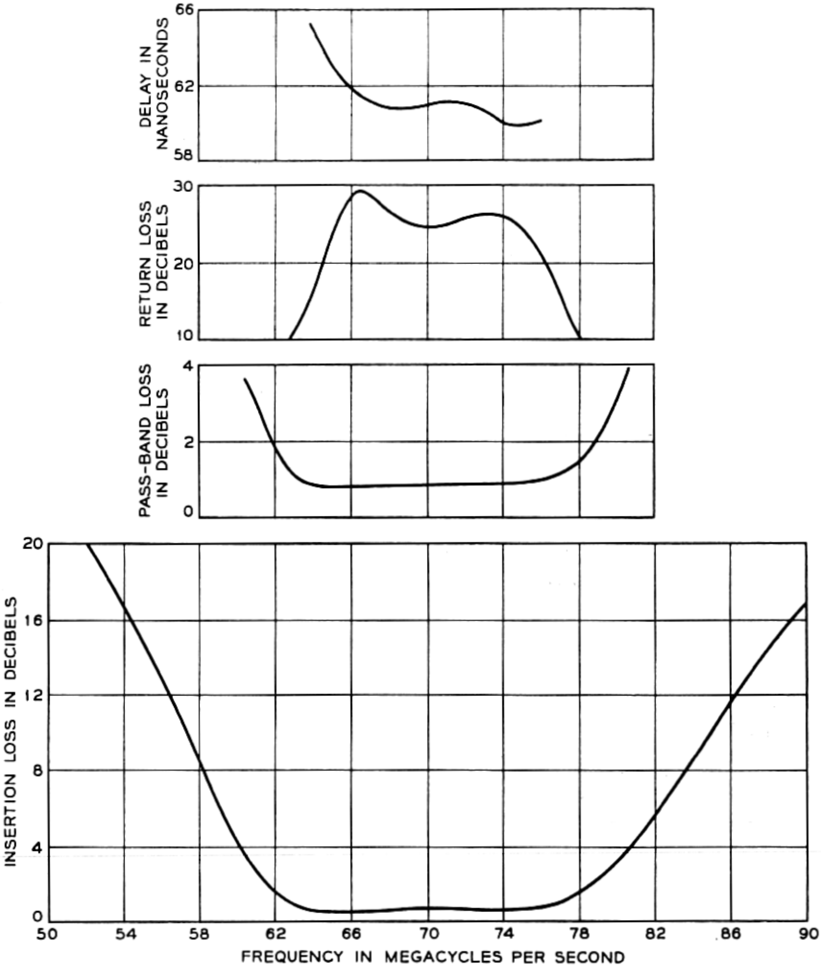


Fig. 18 — Loss, delay and return loss of IF filter.

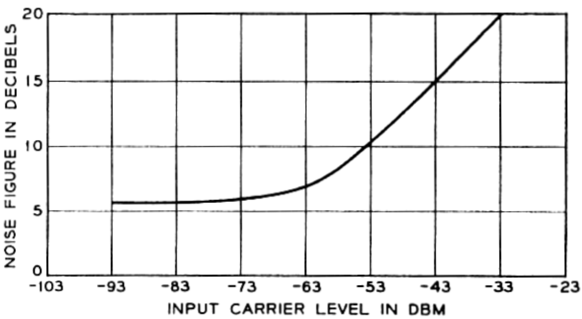


Fig. 19 — Noise figure versus input carrier level, 25°C.

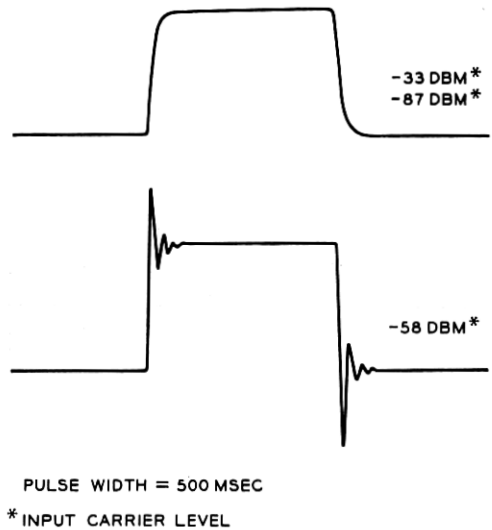


Fig. 20 — Step response of AGC system.

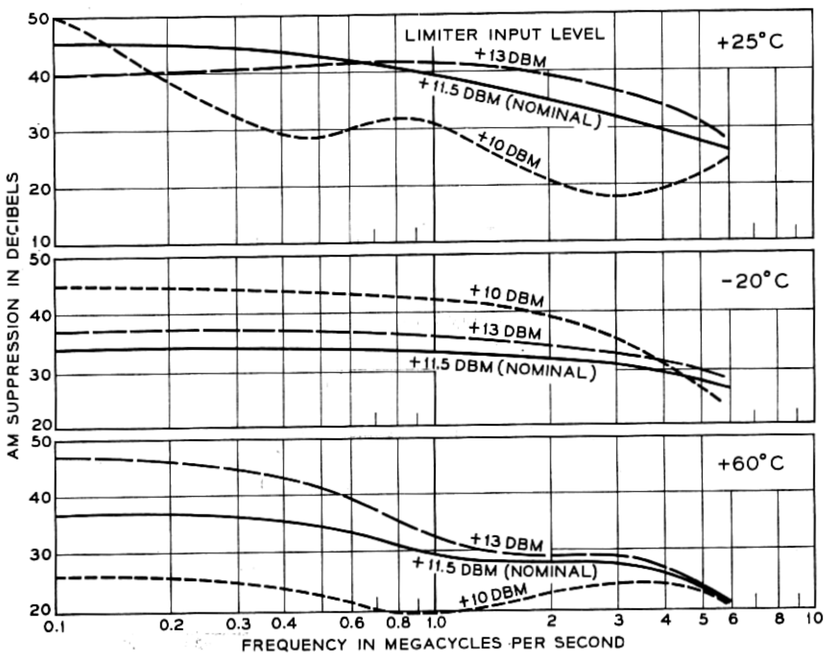


Fig. 21 — Limiter AM suppression

6-mc baseband under the worst conditions of temperature and of drive level permitted by the AGC.

2.4.4 *Discriminator and Transistor AFC Amplifier*

A discriminator is judged primarily by its linearity, stability of its zero crossover frequency and sensitivity.[7] The over-all receiver nonlinearity is typically measured to be better than 4, 7 and 4 per cent at temperatures of +25, −20 and +60°C, respectively, for a peak deviation of 6 megacycles; and this is approximately the nonlinearity of the discriminator.

Measurements on the composite discriminator-transistor AFC amplifier indicate that the shift in zero crossover frequency is ±200 kc, and the nominal sensitivity of 40 mv per megacycle changes ±2 mv per megacycle over this temperature range. The rejection of the 1800-cycle signal and harmonics of this frequency generated by the magnetic amplifier is sufficient to keep the level of this signal at the receiver-baseband output down to the order of −50 dbm.

2.4.5 *AFC Loop*

The AFC loop performance, aside from drift which was discussed in the previous section, is relatively independent of temperature. A typical circuit has a loop gain of 30 db, 10-db gain margin and 60 degrees phase margin.

2.4.6 *Receiver Baseband Amplifier*

The open-loop gain and phase vs frequency characteristics of a typical receiver baseband amplifier are shown in Fig. 22 for a medium gain setting and room temperature. The closed-loop gain curves for −20°C, +25°C and +60°C are also shown. It is clear that the external gain from 200 cps to 2 mc stays constant within ±0.08 db over the temperature range, and between 2 mc and 6 mc it is constant within ±0.23 db. The open-loop characteristic exhibits a 6-db per octave rolloff, which is desirable when flatness of external gain near the unity feedback crossover frequency is important. The feedback amounts to 31 db at midband and 13 db at 6 mc. The phase and gain margins are 47° and 6 db, respectively. (A small gain margin such as this can be tolerated when a large part of the phase shift tracks well with the f_T's of the transistors* and, hence, with the crossover frequency.) Since feedback is maintained down to dc, there is no question of low-frequency oscillation.

* f_T is defined as the frequency at which the common-emitter short-circuit current gain is unity. It is a convenient figure of merit for the bandwidth of the transistor.

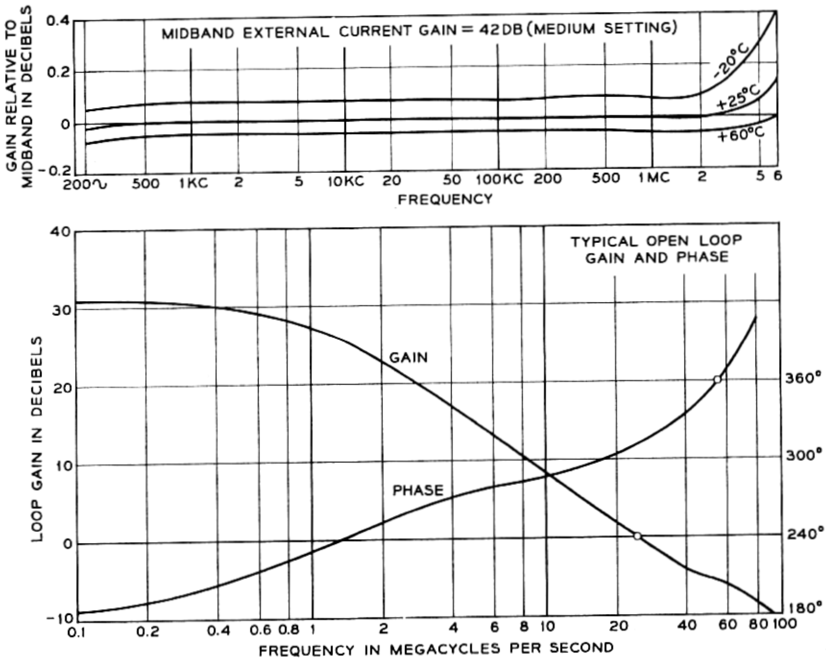


Fig. 22 — Gain-frequency curves — typical receiver baseband amplifier.

The second- and third-harmonic distortion products are down at least 37 db and 46 db, respectively, for maximum external gain (minimum feedback condition) and full output level of +10 dbm at 6 mc. These figures become better rapidly as frequency decreases, reaching −60 dbm and −70 dbm, respectively, at 500 kc. The output noise power of the amplifier in a 500-cycle band for frequencies above 100 kc is approximately −92 dbm.

2.4.7 *Transmitter Baseband Amplifier*

Measurements similar to those made on the receiver baseband amplifier were made on the transmitter amplifier. Fig. 23 displays the gain-frequency characteristics of a typical transmitter baseband amplifier at a medium gain setting and room temperature. This particular model contains transistors with average f_T's; thus, it represents a typical case. The phase and gain margins are 53° and 8 db, respectively.

Distortion measurements taken on the transmitter baseband amplifier indicate that the second and third harmonics are, respectively, at

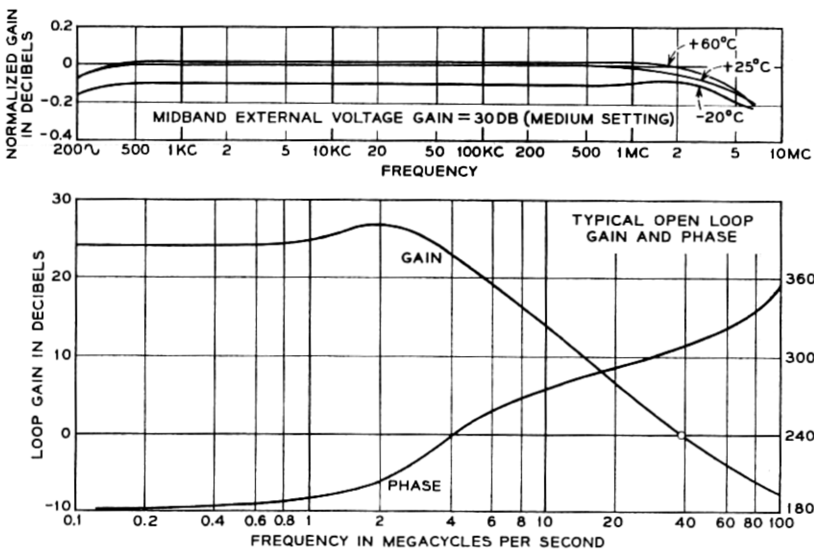


Fig. 23 — Gain-frequency curves — typical transmitter baseband amplifier.

least 54 db and 70 db below the fundamental at 6 mc and full output voltage (4 volts peak).

2.4.8 *Squelch Circuit*

The squelch threshold level for a typical receiver was measured as a function of temperature. The firing-level setting of −87 dbm drifted to −84 dbm and −85 dbm at −20°C and +60°C, respectively. The hysteresis between turn-on and turn-off was less than 1 db.

2.4.9 *AM-to-PM Conversion*

The over-all AM-to-PM conversion of a receiver was measured for temperatures ranging from −20°C to +60°C and for IF input levels ranging from −33 to −83 dbm. The worst AM-to-PM conversion was 0.017 radian for a 10 per cent AM index.

III. CONCLUSION AND REMARKS

Certain apparent limitations in the use of solid-state components in a wideband microwave radio relay system have been overcome. The fresh approaches which have been applied have, in some cases, resulted in performance which is superior to that achieved with comparable electron tube designs. This improved performance is over and above that achieved through inherent advantages of solid-state devices over conventional electron tubes, which include reduced power drain, small size, reliability, and potentially low cost.

It has recently come to our attention that others working in the field of microwave radio systems have arrived through parallel efforts at some of the same design techniques described herein.[8] It is hoped that the publication of this paper will help expand the field of application of solid-state devices and will encourage others to design around apparent shortcomings of solid-state devices by refusing to have their thinking completely channeled by what already exists.

IV. ACKNOWLEDGMENTS

The solid-state FM receiver circuits described in this paper are the result of the efforts of many members of Bell Telephone Laboratories. Some of the design concepts were born in the Radio Research Department, which triggered the development of the TL system; and Messrs. L. C. Tillotson, C. L. Ruthroff, N. E. Chasek and W. F. Bodtmann were particularly influential there. The members of the Transmission Division who contributed to the project are numerous, but the efforts of Mr. L. F. Willey on the development of the limiter and discriminator, Mr. J. Gammie's work on the AFC magamp, Mr. G. H. Klemm's equipment design effort, the many helpful suggestions of Mr. F. H. Blecher and the accumulation of several notebooks full of important data by Mr. H. A. Hageman are particularly noteworthy.

APPENDIX

A.1 *Transistor "Low-Pass" IF Amplifiers*

At least three basic configurations are useful in transistor wideband IF amplifiers. These are the common-base stage with a wideband transformer interstage network, the doublet circuit, and the common-emitter stage with frequency-dependent shunt feedback.

A.1.1 *The Common-Base Stage with a Wide-Band Transformer Interstage Network*

Because the current gain of a common-base stage is less than unity, this configuration requires the use of a current step-up interstage network to achieve power gain when a cascade of similar stages is used as a wideband amplifier. Transformers are available which provide a constant current step-up over a wide band of frequencies[4,5,6] between terminating impedances typical of solid-state circuits. The basic common-base configuration is shown in Fig. 5. It can be shown that to achieve "flat" amplitude response extending to within an order of magnitude of f_T, damping resistor R_2 is required. Use of a transformer having a 1:2 current step-up results in a power gain slightly less than 6 db per stage.

In the TL IF amplifier stages, the transformer turns ratio is much lower than that which would result in maximum gain. This makes the amplifiers quite insensitive to changes in transistor parameters. This implies a very stable transmission characteristic over a wide temperature range and large variations in power supply voltage. Another result is that the circuits accept an extremely wide range of transistor parameters, either due to statistical distribution or to aging; and adjustment of the transmission characteristic is relatively simple.

The input and output impedances are relatively stable with temperature and dc bias and can be represented by simple passive networks. Hence, the common-base stage can be compensated to yield a stable resistive input or output termination over a wide frequency range. This feature makes it particularly useful for connection to passive bandpass filters or other equalizer networks. Since the common-base static characteristics are much more linear than those of other configurations, this configuration also gives superior performance in high-level stages. However, one must look to other configurations, such as the doublet, for the best noise figures in input stages.

A.1.2 *The Doublet Circuit*

This circuit, shown in Fig. 4, takes advantage of the fact that certain transistor pairs, or doublets, without interstage transforming networks, can provide an over-all gain stage which is more immune to temperature and power supply variations than each transistor considered separately. The reason for this improvement is the fortunate circumstance that undesirable variations on the individual stages tend to cancel for certain configurations. Two combinations which have been shown to exhibit this desirable property are the common-emitter — common-emitter doublet and the common-emitter — common-collector doublet. The gain characteristic, though stable, rolls off smoothly at high frequencies; hence, it is necessary to follow the transistors with an equalizer network, which usually takes the form of a constant-resistance, high-pass filter. The high-pass filter is designed to have a cutoff frequency above the

IF band, and the rolloff of the filter compensates for the rolloff of the transistors.

The gain level of a doublet is a fairly strong function of the f_T of the transistors used; however, the over-all normalized gain-frequency characteristic is stable with temperature and bias variations. Exclusive use of the doublet in a high-gain amplifier would probably dictate either a tight control of the average f_T of the transistors used or a larger dynamic range of automatic gain control. The power gain of the doublet used in the TL radio receiver ranges from 15 to 20 db.

The common-emitter — common-emitter doublet configuration has been found to be the one best suited for use as the input circuit for wideband IF amplifiers. It provides the lowest noise figure of all of the configurations evaluated and, through the use of a simple impedance-matching network, can yield a good input return loss over a wide IF band. The low noise figure is due to the facts that the input of the amplifier is separated by a block of relatively high gain from the remainder of the amplifier and that the natural input impedance of the doublet amplifier needs only minor correction to achieve good input return loss at the desired impedance levels (usually 50 to 125 ohms).

A.1.3 *The Common-Emitter Configuration with Frequency-Dependent Shunt Feedback*

Of all forms of wideband amplifiers studied, the configuration which most easily yields the broadest bandwidths is the common-emitter stage with frequency-dependent shunt local feedback.[9,10] This circuit, shown in Fig. 24, has a two-terminal RL network connected between base and collector. The effect of the RL network is to reduce the gain of the stage by application of negative feedback. Resistor R_1 determines the low-frequency gain, and inductor L effectively removes the feedback

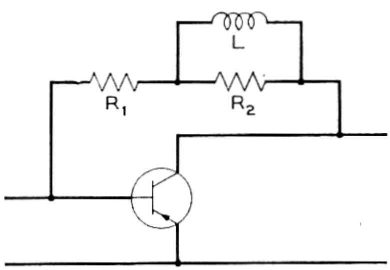


Fig. 24 — Common-emitter stage with shunt local feedback.

at high frequencies and hence is a broadbanding element. Resistor R_2 damps the resonance which occurs between the inductor L and the capacitive reactance presented by the transistor. It has been demonstrated that this configuration is very flexible in that gain and bandwidth can easily be exchanged, and the gain-bandwidth product is approximately given by f_T.[9]

A cascade of common-emitter stages with shunt feedback exhibits only very slight changes in the normalized amplitude response for wide variations in temperature and power supply voltage. However, because the gain level of each stage is highly dependent on f_T and r_b' and on the input impedance of the following stage, the over-all gain level varies considerably. By specifying tight limits on the average values of the parameters for a given set of transistors (usually only the average f_T need be specified), the over-all absolute gain can be kept within close limits. Of course, a moderate amount of gain adjustment may be had by changing the value of R_1 in the feedback network. This configuration cannot compete with the doublet circuit as a low-noise input stage or with the common-base configuration as a low-distortion, high-level stage; however, it appears to be the best configuration to use when extremely large gain-bandwidth products are required.

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