A Wideband Transistor IF Amplifier for Space and Terrestrial Repeaters Using Grounded-Base Transformer-Coupled Stages

By W. F. BODTMANN and C. L. RUTHROFF

(Manuscript received April 27, 1962)

A wideband transistor IF amplifier is described. A measured model has a gain of 41 db and is flat to ±0.1 db from 50 to 100 mc. Data on noise figure, effect of temperature changes and power output are presented. An electronically variable attenuator is described which is suitable for use in such an amplifier as part of the AGC circuit.

I. INTRODUCTION

The general objectives of the work discussed in this paper are as follows:

(i) To design a wideband IF amplifier suitable for use on heavy-route, long-haul microwave radio relay systems.[1,2]

(ii) To design a wideband IF amplifier with low power consumption and good temperature stability for use in microwave radio relay systems[3,4] which may be battery powered and which must operate at outdoor ambient temperatures.

In accordance with these objectives it was decided to design an amplifier to be substantially flat over a 40-mc band centered at 70 mc. Transistors were chosen as the active elements because of the low bias power required, and the common-base circuit configuration was selected because it appeared likely to result in a stable band shape with respect to temperature.

A part of the initial philosophy was to start with a stable wideband amplifier. The desired frequency selective circuits could then be added as linear lumped circuits whose gain and phase properties are in good control. Such circuits could include any desired gain and phase equalization. This approach results in an IF amplifier with an excellent delay characteristic.

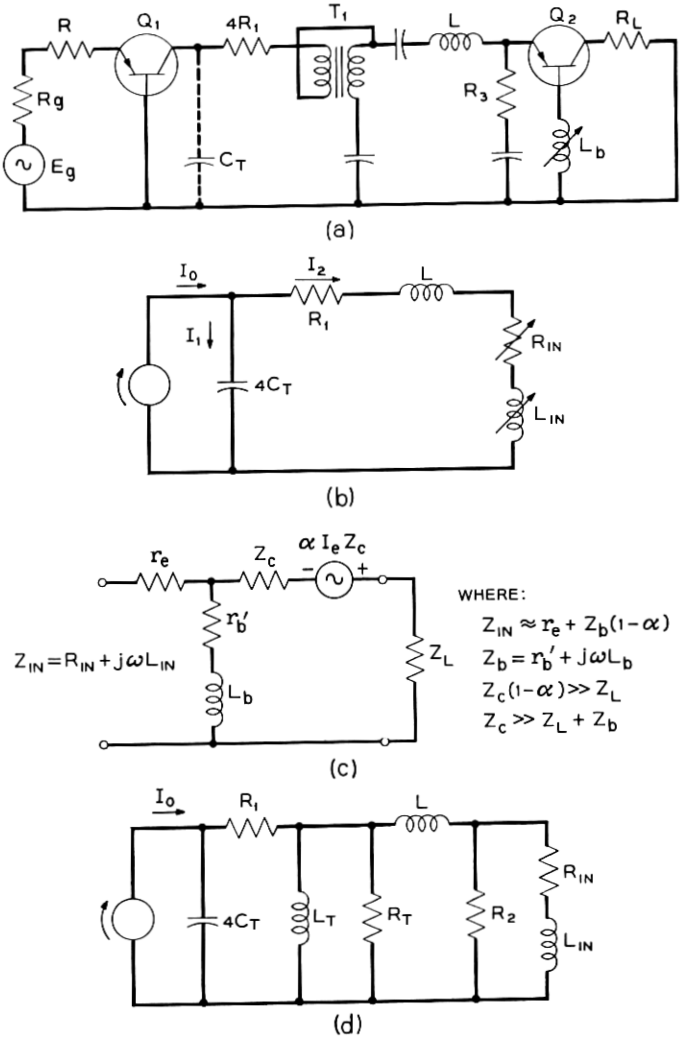


Fig. 1 — (a) Typical interstage circuit, (b) equivalent circuit for the interstage, (c) equivalent circuit for computation of input impedance, (d) equivalent interstage circuit including losses.

II. THE IF CIRCUIT[5,6]

The common-base configuration was chosen for several reasons. The transistors available have alpha cutoff frequencies in the region 500–700 mc. If these transistors were used in the common-emitter configuration in this application, the IF band would extend well beyond the beta cutoff frequency and considerable equalization would be necessary to shape the IF characteristic. Furthermore, changes in ambient temperature result in a shift in both alpha and beta cutoff frequencies. The alpha cutoff is well beyond the IF band but the beta cutoff is not. Hence, for operation over a reasonably large temperature range, the common-base stage has better stability with respect to band shape and midband gain.

Simple broadband transformers[7] are available which can be used in the common-base configuration and result in an interstage with at most one tuning element per stage.

The interstage circuit is shown in Fig. 1(a). L is the stray inductance, C_T the total interstage capacitance, L_b is an inductance added in the base of transistor Q_2, T_1 is a wideband transformer and R_1 has been added to help shape the interstage characteristic. This circuit has been redrawn in Fig. 1(*b*), where the assumptions have been made that the output of transistor Q_1 is a constant current and T_1 is an ideal transformer with an impedance ratio of 4:1. Transistor Q_2 has been replaced by its input impedance, which is considered to be $R_{IN} + j\omega L_{IN}$ where both R_{IN} and L_{IN} are functions of frequency. The blocking capacitors and bias circuits have been omitted. From Fig. 1(*b*) we see that the output capacitance of transistor Q_1, the input impedance of transistor Q_2, stray inductance L, and resistance R_1 form a peaking circuit. The equivalent circuit used to compute the input impedance of Q_2 is shown in Fig. 1(c). It has been assumed that $Z_c(1 - \alpha) \gg Z_L$ and $Z_c \gg Z_L + Z_b$. In the Appendix it is shown that the frequency response of Fig. 1(b) is

$$\left| \frac{I_0}{I_2} \right|^2 \approx \frac{(R_T - \mu)^2 + (\mu\gamma + X_L - X_C)^2}{X_C^2} \tag{1}$$

where $R_T = R_1 + r_e + r_b'$

$\mu = (r_b' + \gamma\omega L_b)/(1 + \gamma^2)$

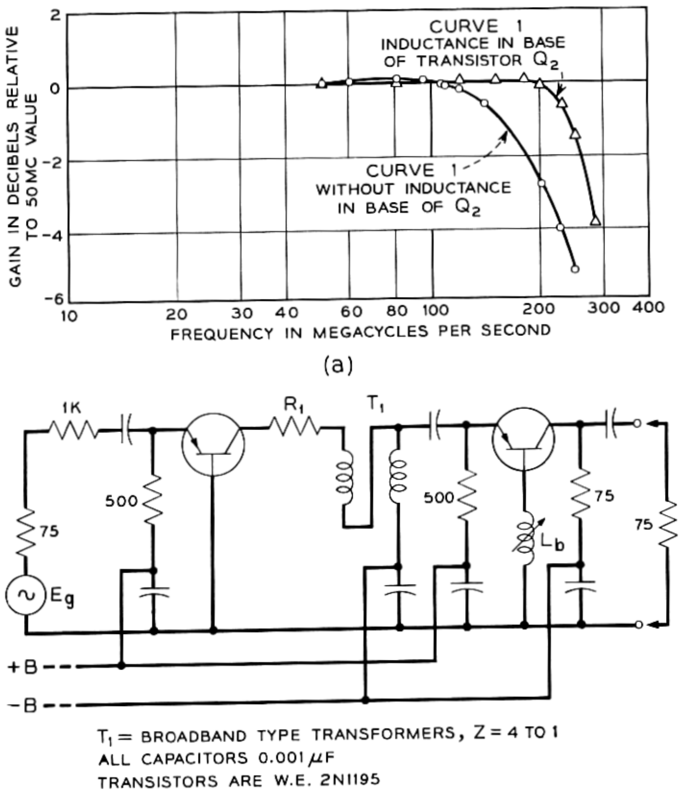
$\gamma = f/f_\alpha$

$r_e =$ emitter resistance

$r_b' =$ high-frequency base resistance.

The high-frequency response of this circuit can be controlled by adjusting R_1 or L_b. Amplifiers have been built using both schemes. The effect of changing R_1 is to change the Q of the circuit. The effect of adding a suitable inductance, L_b, in the base is a little more complicated but results in a better frequency response.

An inductance in series with the base of a transistor changes the transistor input impedance by decreasing the resistive part and increasing the reactive part. (See Appendix.) Thus, instead of adjusting R_1, a variable inductance in series with the base can be used to tune the interstage. Even though this method of tuning adds inductance to the emitter circuit it does not decrease the bandwidth of that portion of the response which is flat to a few tenths of a decibel. In fact if R_1 and L_b are properly chosen, the amplifier will be compensated over a wide frequency range. This is demonstrated in Figs. 2, 3, 4. Fig. 2(a) shows the measured responses of the peaking circuit with and without added



(a)

T_1 = BROADBAND TYPE TRANSFORMERS, Z = 4 TO 1
ALL CAPACITORS 0.001 μF
TRANSISTORS ARE W.E. 2N1195

(b)

Fig. 2 — (a) Measured responses of series peaking circuit, (b) circuit used to measure interstage series peaking response.

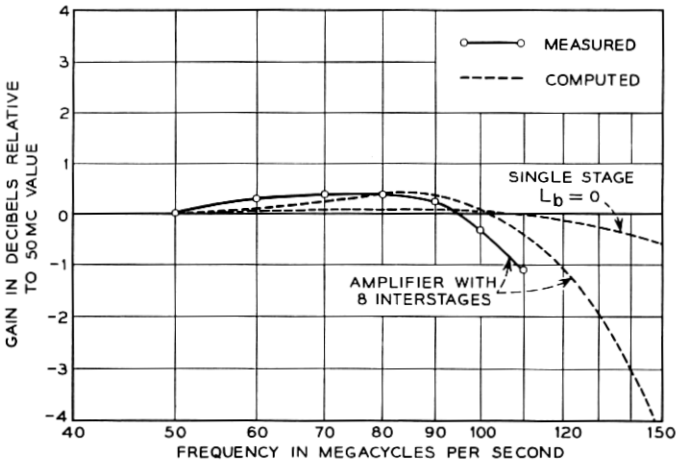


Fig. 3 — Amplitude-frequency response with no added inductance in base of transistor.

inductance in the base circuit. The circuit measured is shown in Fig. 2(b). The computed curves are identical to the measured ones when the following parameters are used in (1):

$$f_\alpha = 600 \text{ mc} \qquad C_T = 4\mu\mu f \qquad V_{CB} = -10\text{v}$$

$$L = 0.02\mu\text{h} \qquad L_b = 0.05\mu\text{h} \qquad I_E = 10 \text{ ma}$$

$$r_b' = 50 \text{ ohms} \qquad r_e = 3 \text{ ohms}$$

$$R_T = 115 \text{ ohms when } L_b = 0.05\mu\text{h}$$

$$R_T = 110 \text{ ohms when } L_b = 0.$$

Fig. 3 shows the computed response of a single stage and computed and measured responses for eight identical interstages, with no added inductance in the base lead. The transistor parameters are as listed above. $L_b = 0$ and $R_1 = 54$ ohms. For the computed responses α has been assumed equal to one and the transformers ideal. Since these assumptions are not exactly true, the measured response can be expected to be poorer than the computed response. The midband insertion gain between 75-ohm impedances is 41 db.

Fig. 4 shows the computed responses of a single interstage with two bias conditions and the measured response of eight interstages with inductance added in the base lead. The bias conditions for curve 1 are $V_{CB} = -10$v and $I_E = 10$ ma; for curve 2, $V_{CB} = -6$v and $I_E = 6$ ma. The bias conditions for the measured response are the same as for curve

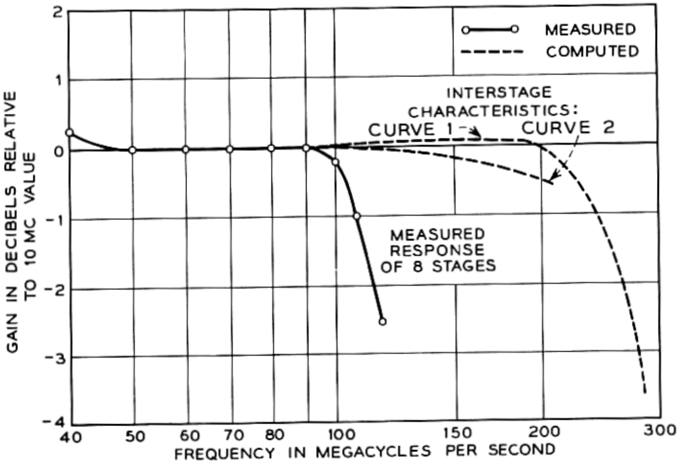


Fig. 4 — Amplitude-frequency response with inductance added in base of transistor.

2 in all except the last two stages which are discussed later. As suggested by Fig. 4, reducing the bias power results in some sacrifice in bandwidth. The final selection of bias conditions is a compromise between a flat response over the desired bandwidth and low power consumption for the amplifier. For the computation of curve 2 the circuit parameters are:

$$f_\alpha = 400 \text{ mc} \qquad C_T = 4.5\mu\mu\text{f} \qquad R_1 = 59 \text{ ohms}$$

$$r_b' = 50 \text{ ohms} \qquad L_b = 0.05\mu\text{h}$$

$$r_e = 4.3 \text{ ohms} \qquad L = 0.02\mu\text{h}$$

The measured midband insertion gain between 75-ohm impedances is 41 db for eight interstages. The dc power consumption is 0.7 watt.

III. INPUT AND OUTPUT CIRCUITS

A complete schematic diagram of the amplifier with a matched input circuit and two output circuits, is shown in Fig. 5. The input circuit shown is similar to the interstages in that a resistor is added in series with the emitter to provide the proper input resistance of 75 ohms, and a capacitor is shunted across the input to peak the circuit at higher frequencies. The return loss of the amplifier input as a function of frequency is shown in Fig. 6.

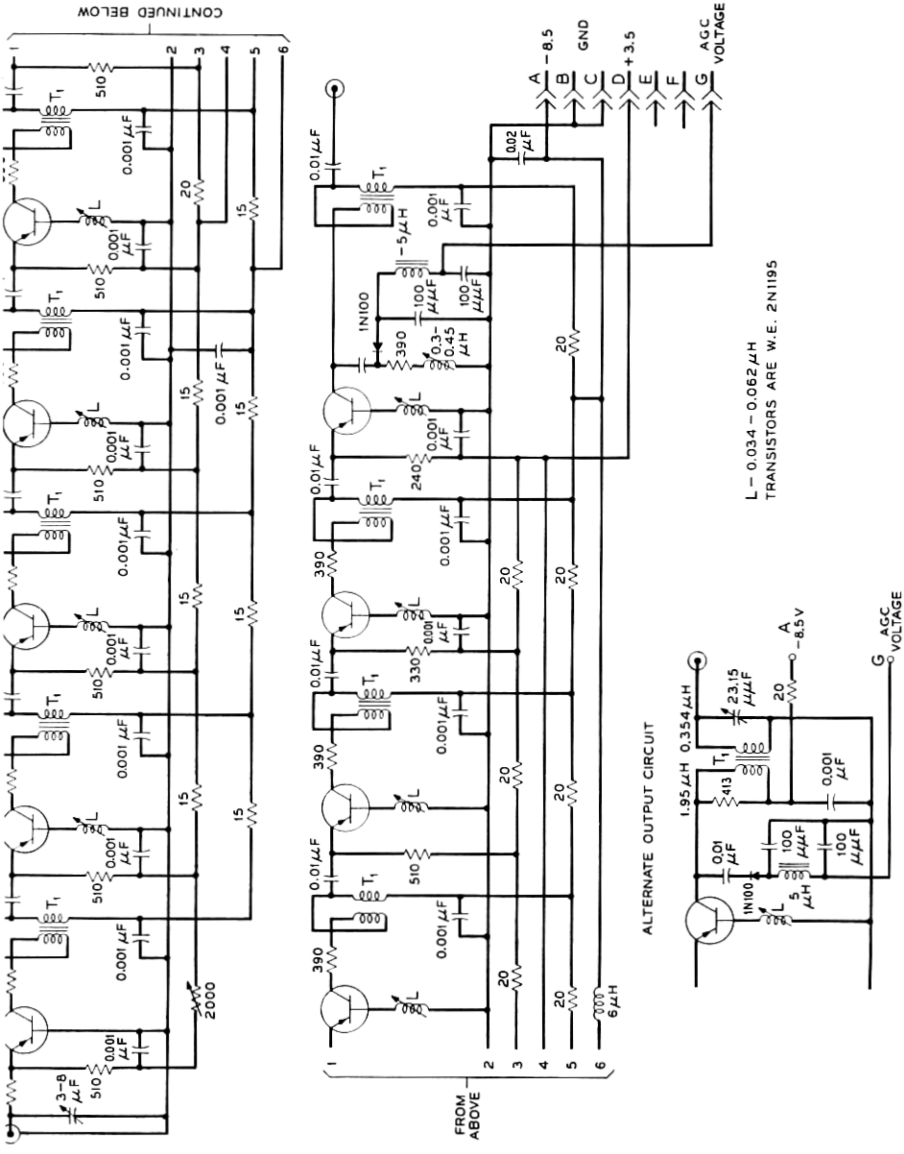


Fig. 5 — IF amplifier schematic.

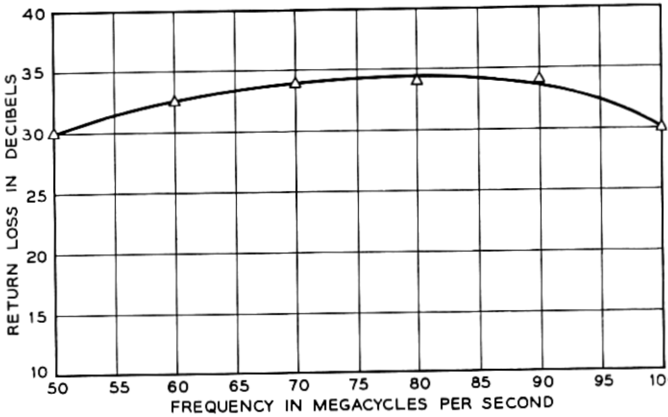


Fig. 6 — Return loss vs frequency for input circuit of IF amplifier of Fig. 5 measured at 75 ohms.

The input conditions for operation with a low noise figure are discussed in Section VIII.

Two output circuits are shown in Fig. 5. One output circuit consists of a broadband transformer and a shunt peaking circuit. The other has a double-tuned transformer and is shown as the alternate output circuit in Fig. 5. The return losses of the two output circuits, measured at 75 ohms, are given in Fig. 7. The double-tuned transformer adds a measured variation of about one-tenth of a db to the frequency response. All data discussed in this paper were taken with the output circuit having the broadband transformer.

The AGC detectors shown in Fig. 5 have been designed so as to have no effect upon the frequency response. AGC problems are discussed in Section IX.

IV. POWER OUTPUT AND FREQUENCY RESPONSE

The power output of the amplifier depends upon the bias conditions in the output transistors, which in turn depend upon the maximum ambient temperature expected. The last two transistors in Fig. 5 have increased bias currents to allow for increased power output. The bias is arranged so that the rated junction temperature is not exceeded at the high end of the ambient temperature range, 130°F.

The compression characteristic of this amplifier for a single frequency sine wave input at 70 mc is given in Fig. 8. In addition to this information it is desirable to know the frequency response as a function of output

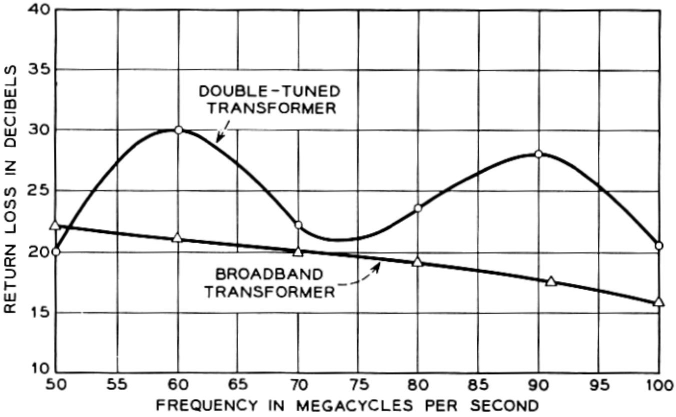


Fig. 7 — Return loss vs frequency for output circuits of IF amplifier of Fig. 5 measured at 75 ohms.

level. This is shown in Fig. 9. Fig. 9(a) shows the response for an output power of +6 dbm and a negligible compression. Fig. 9(b) shows responses for several other levels where it is to be noted that while the details of the responses have been preserved, the level differences between the curves have not.

V. TEMPERATURE-FREQUENCY RESPONSE

The frequency response as a function of temperature has been obtained for the amplifier that does not have added inductance in the base circuit. The data are presented in Fig. 10, plotted relative to the gain at 50 mc. The 50-mc gain decreases at high temperatures and increases at low temperatures. The total change in 50-mc gain in the range −40°F to +130°F is under 2 db.

VI. TRANSFORMERS

The interstage transformers used in this amplifier have been described elsewhere[7] and only a brief description is given here. These transformers are of a transmission line type and consist of a short bifilar winding on a small nickel-zinc ferrite toroidal core. The winding is formed by twisting a pair of wires and then winding the pair on the core. Fig. 11 is a schematic of the transformer and a plot of the measured frequency response into resistive terminations. The turns ratio is 2:1.

This type of transformer may be regarded as an ideal transformer in

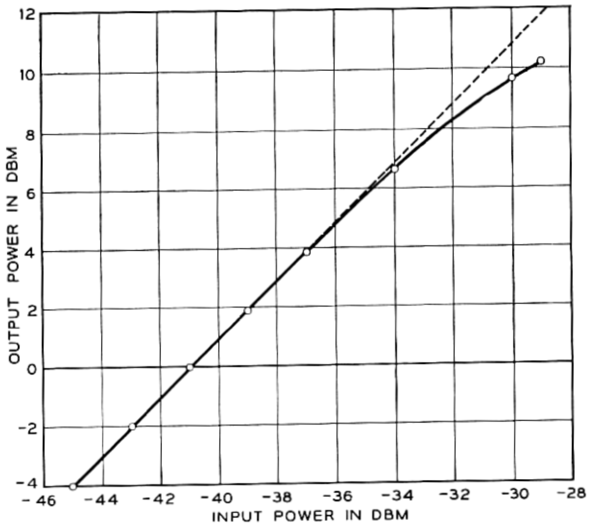


Fig. 8 — Compression characteristic for amplifier of Fig. 5.

series with a transmission line whose electrical length equals that of the bifilar winding. The analysis in this paper has assumed only the ideal transformer, which is reasonable as long as the winding is short compared to a wavelength at the highest frequency of interest.

The permeability and Q of this ferrite has been optimized in this frequency range to give minimum loss. With this ferrite and a winding 1.6 inches long, the transformer has the performance shown in Fig. 11.

VII. GAIN PER STAGE

All of the current gain is obtained from the interstage transformer so the maximum current gain per stage using a transformer with an impedance ratio of 4:1 would be 6 db. There are three sources of current loss in each interstage which reduce this gain. As seen in Fig. 1(d), they are:

(i) The transformer inductance L_T and resistance R_T which shunts the emitter-base circuit of Q_2.

(ii) The emitter bias resistor R_2 which shunts the input circuit of Q_2. This loss can be reduced at the expense of greater bias power.

(iii) The current gain of $Q_2 = \alpha < 1$.

The total loss from these sources is 0.8 db so that the actual measured gain per stage is 5.2 db.

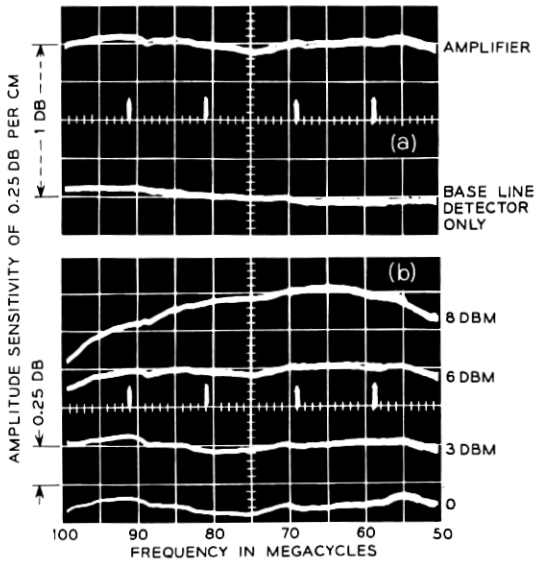


Fig. 9 — (a) Frequency response of amplifier and base line at output power level of +6 dbm and insertion gain of 41 db, (b) frequency response of IF amplifier as a function of the output level in dbm.

VIII. NOISE FIGURES

The amplifier of Fig. 5 has its input impedance adjusted to 75 ohms as indicated in Fig. 6. With this input circuit it is suitable for use as a main IF amplifier. For use as a preamplifier, however, the input circuit can be modified to improve the noise figure. Such a circuit is shown in Fig. 12. Note that the turns ratio of the interstage transformer is 8:3. The measurements reported below were made on the amplifier of Fig. 5 with the input circuit of Fig. 12 and using a diode noise generator. The bias on the first stage was $I_E = 3$ ma, $V_{CB} = -7$v.

Some amplifier noise figure measurements with W.E. 2N1195 transistors in the input circuit are shown in Fig. 13. The curve is the noise figure for the best of 5 transistors, and the amplifier noise figure is plotted versus generator resistance. Also shown are the amplifier noise figures for the 5 transistors with a generator resistance of 70 ohms. These measurements were made in a 0.7-mc band centered at 71 mc. At 50 mc the noise figure is typically 0.5 db lower and at 100 mc, 0.75 db higher.

It should be emphasized that these are amplifier noise figures and not transistor noise figures. In such a broadband amplifier, care must be taken to minimize the noise contribution of the second transistor. This

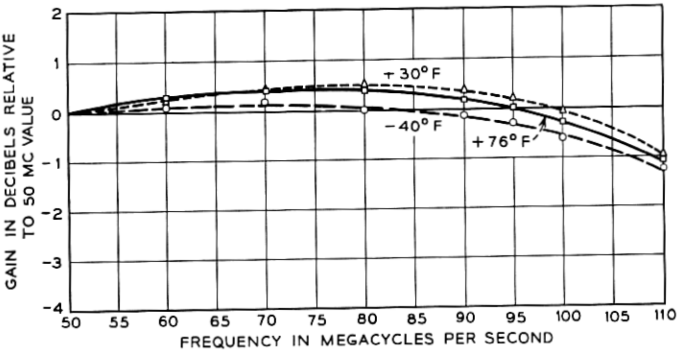


Fig. 10 — Frequency response vs temperature of IF amplifier.

accounts for the higher turns ratio of the interstage transformer. Use of the transformer described in Section VI results in an increase of 0.5 db in noise figure.

The noise contribution of the second transistor depends upon the available gain of the preceding circuit,[8] and the available gain depends in part upon the generator impedance and the transformer turns ratio. It is to be expected, then, that for narrower bandwidths the noise contribution of the second transistor can be further reduced, since higher turns ratios can be used with the corresponding increase in available gain of the first stage.

For the broadband common-base amplifier the best noise figures are obtained with transistors having low optimum source resistances. This comes about because the available gain increases as the optimum source resistance decreases, and the greater the available gain, the lower the noise contribution of the second transistor. It follows that for best results a low-noise transistor having a low optimum source resistance should be used.

The effect of the input circuit of Fig. 12 on the frequency response of the amplifier is quite small. With this input circuit the response was measured using a 75-ohm source impedance and the response equivalent to that of Fig. 9 was obtained by slight readjustment of the base lead inductance in the second stage.

IX. AUTOMATIC GAIN CONTROL CIRCUIT

Wideband transistor IF amplifiers present special problems with respect to AGC in that it is not always feasible to merely adjust the bias to obtain gain control. The method presented here consists of an elec-

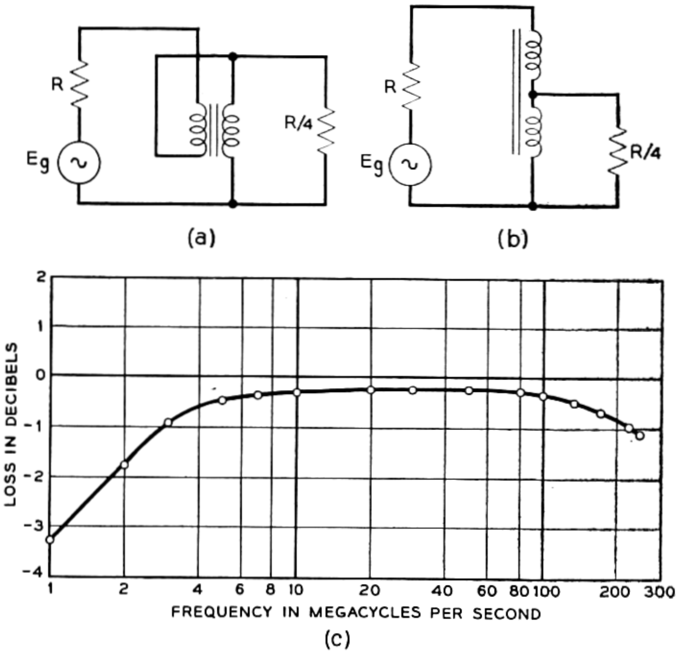


Fig. 11 — Transmission-line type transformer with impedance transformation of 4 to 1. (a) Transmission line form, (b) low-frequency equivalent, (c) amplitude vs frequency of transformer measured between 75 ohms and 18.75 ohms.

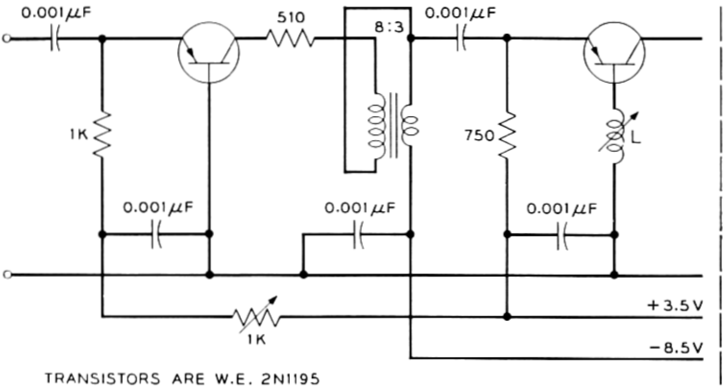


Fig. 12 — Input circuit for low noise figure.

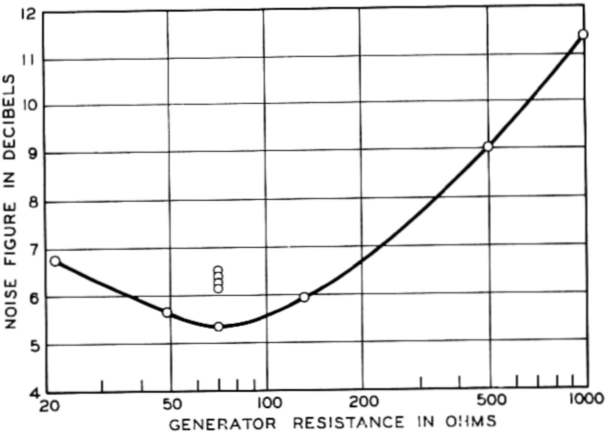


Fig. 13 — Noise figure measurements.

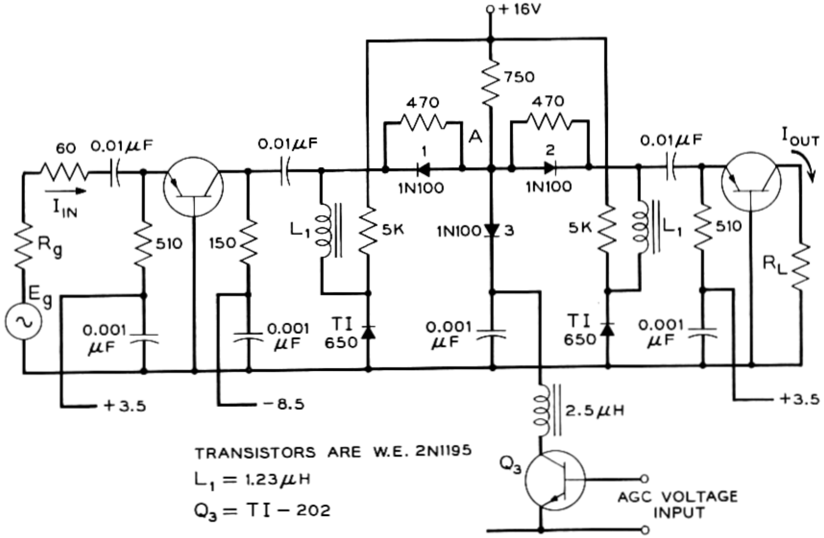


Fig. 14 — AGC attenuator.

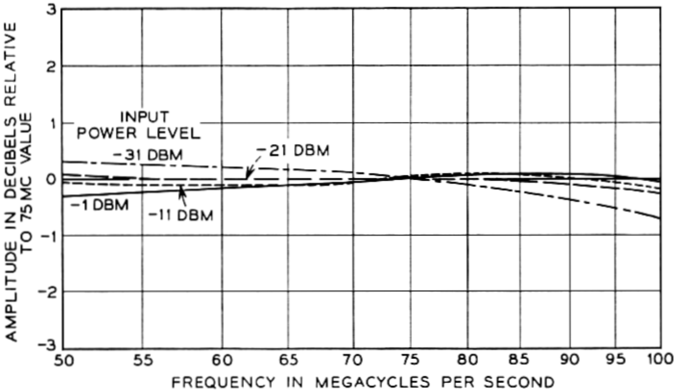


Fig. 15 — Frequency response of variable IF attenuator of Fig. 14. Output level held constant at −37 dbm.

tronically variable attenuator which can be inserted in the amplifier at any suitable point. The attenuator is described schematically in Fig. 14 and is basically a T-pad formed by three diodes. The three diodes are supplied with current at point A through the 750-ohm resistance. Transistor Q_3 controls the current through diode 3 which in turn controls the current through diodes 1 and 2. The resistance of the diodes is a function of the current through them so the attenuation of the T-pad can be controlled by adjusting the current in diode 3. This is done by applying the AGC detector output to the emitter of transistor Q_3.

If Q_3 is cut off, the current at point A splits between diodes 1 and 2, making them low resistances; diode 3 becomes a high resistance. This is the minimum-loss condition. When Q_3 conducts heavily, all of the current

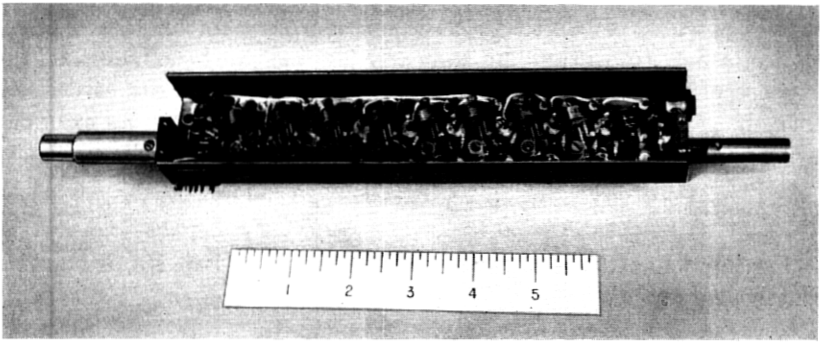


Fig. 16 — Wideband IF amplifier.

at point A goes to diode 3, lowering its resistance and cutting off diodes 1 and 2. This is the maximum-loss condition. The reference diodes (TI-650) provide a 4-volt bias that assures that diodes 1 and 2 can be cut off and also provides collector voltage for Q_3.

The frequency response of the attenuator is given in Fig. 15. These curves show the variation in response relative to the 75-mc value as the power input to transistor Q_1 (Fig. 14) is varied from -1 dbm to -31 dbm. For inputs greater than $+3$ dbm the response changes shape rather than slope.

The minimum measured insertion loss of the circuit of Fig. 14 is 6 db. The power consumption of the attenuator is 0.35 watt.

X. MECHANICAL ARRANGEMENT

The photograph of Fig. 16 shows the mechanical arrangement of one section of the amplifier which has 41 db of gain. The transistors are mounted inside the chassis to minimize stray inductances and to prevent accidental grounding of the collector, which is connected to the transistor header.

XI. DISCUSSION

It has been demonstrated that high quality wideband IF amplifiers can be made using transistors. Such amplifiers are capable of operation over large temperature ranges and small input power. For instance, an amplifier with a flat bandwidth of 50 mc, 90 db of gain, $+12$ dbm of output power, 34 db AGC range, a noise figure of 6 db, and capable of operation over a temperature range of $-40°$F to $+130°$F is feasible. Such an amplifier would require about 2 watts of bias power and would contain 23 transistors and 6 diodes.

It is possible, of course, to trade one attribute for another. For instance, a flat band extending to nearly 200 mc can be achieved at the expense of increased bias power. If a narrower bandwidth is desired, then the gain per stage can be increased. This can be done, for example, by using two transformers per stage or using another type of transformer.

APPENDIX

The input impedance of the equivalent circuit of Fig. 1(c) is approximately

$$Z_{IN} \approx r_e + Z_b(1 - \alpha) \tag{1}$$

where

$$Z_c(1 - \alpha) \gg Z_L$$

$$Z_c \gg Z_L + Z_b .$$

Now let

$$\alpha = \frac{\alpha_0}{1 + j\gamma} \quad \text{and} \quad \alpha_0 \approx 1$$

$$\gamma = f/f_\alpha$$

$$Z_b = r_b' + jX_b$$

$$X_b = \omega L_b = \gamma \omega_\alpha L_b = \gamma X_{b\alpha}$$

After these substitutions (1) becomes

$$Z_{IN} \approx \frac{(r_e + r_b')(1 + \gamma^2) - (r_b' + \gamma^2 X_{b\alpha}) + j(\gamma r_b' + \gamma^3 X_{b\alpha})}{1 + \gamma^2}. \quad (2)$$

Thus, by adding inductance L_b in the base circuit the input impedance is changed by the addition of the real term

$$[-\gamma^2 X_{b\alpha}/(1 + \gamma^2)] \text{ and the reactive term } j[X_{b\alpha}\gamma^3/(1 + \gamma^2)].$$

When the input impedance of the transistor becomes part of the peaking circuit of Fig. 1(b), it becomes part of the resonant circuit formed by C_T, R_1, L, and Z_{IN}. The total inductance of this circuit will increase and the total resistance will decrease with increasing frequency due to the variation of Z_{IN}, thus changing the shape of the resonance curve. The greatest effects will be observed at the higher frequencies.

The peaking circuit to be analyzed is shown in Fig. 1(b). The total interstage capacitance is C_T, stray inductance L, and inductance added at base L_b. R_1 is used to adjust circuit Q. The output current of transistor Q_1 is I_0. The transformer has been assumed ideal.

I_2 is the emitter current of Q_2. Solving this circuit we get

$$\frac{I_0}{I_2} = \frac{R_1 + R_{IN} + j(X_{L_{IN}} - X_C)}{-jX_C} \quad (3)$$

and substituting from (1) and (2)

$$\frac{I_0}{I_2} = \frac{R_1 + r_e + r_b' - \dfrac{r_b' + \gamma X_b}{1 + \gamma^2} + j\left[\dfrac{(r_b' + \gamma X_b)\gamma}{1 + \gamma^2} + X_L - X_C\right]}{-jX_C}$$

$$= \frac{R_T - \mu + j(\mu\gamma + X_L - X_C)}{-jX_C}. \quad (4)$$

$$\left|\frac{I_0}{I_2}\right|^2 = \frac{(R_T - \mu)^2 + (\mu\gamma + X_L - X_C)^2}{X^2{}_C}, \tag{5}$$

where

$$R_T = R_1 + r_e + r_b'$$

$$\mu = \frac{r_b' + \gamma\omega L_b}{1 + \gamma^2}$$

$$\gamma = f/f_\alpha .$$

REFERENCES

1. McDavitt, M. B., 6000-Megacycle-Per-Second Radio Relay System for Broad-Band Long-Haul Service in the Bell System, A.I.E.E. Transactions, Part I, No. 34, January, 1958.
2. Curtis, H. E., Collins, T. R. D., and Jamison, B. C., Interstitial Channels for Doubling TD-2 Radio System Capacity, B.S.T.J., **39,** November, 1960, p. 1505.
3. Ruthroff, C. L., and Tillotson, L. C., An Experimental 'Short Hop' Microwave System, Bell Laboratories Record, June, 1960, p. 202
4. Gammie, J., and Hathaway, S. D., The TJ Radio System, B.S.T.J., **39,** July, 1960, p. 821.
5. Linvill, J. G., and Schimpf, L. G., The Design of Tetrode Transistor Amplifiers, B.S.T.J., **35,** July, 1956, p. 813.
6. Saari, V. R., Transistor 70-mc IF Amplifier, 1958 Transistor and Solid State Circuits Conference, Philadelphia, Pennsylvania.
7. Ruthroff, C. L., Some Broadband Transformers, Proc. I.R.E., **47,** August, 1959, p. 1337.
8. Friis, H. T., Noise Figures of Radio Receivers, Proc. I.R.E., **32,** July, 1944, p. 419.