Some Design Considerations for High-Frequency Transistor Amplifiers

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The major problem in the design of high-frequency transistor amplifiers is the interaction between the output and the input of the amplifier caused by the internal feedback of the transistor. This problem is illustrated and the two common design approaches to a solution of the problem are discussed. Nyquist's criterion of stability and Bode's feedback theory are then used to obtain an engineering evaluation of the relative merits of these two design approaches from a stability standpoint. The positive nature of the internal transistor feedback is established in this stability evaluation. Finally, Bode's feedback theory is used to consider the relative merits of some of the broad banding techniques used in transistor video amplifier design. The over-all analysis shows that many of the most practical and stable linear transistor amplifiers are very simple and can be built with a minimum of design effort.

I. INTRODUCTION

A survey of the mass of available literature on high-frequency transistor amplifier design discloses the constantly present problem of amplifier sensitivity and even instability, especially when so-called maximum available gain amplifier designs are attempted. This problem is the result of the internal positive feedback inherent in all known transistors. This paper is particularly directed toward a better understanding of transistor internal feedback and its relationship to transistor amplifier design and performance. A fresh and practical engineering approach to the problem of transistor amplifier sensitivity and stability evaluation is presented. The presentation is largely concerned with basic design principles.* Specific amplifier design discussion is limited to that needed

* The material in the paper covers the basic design principles presented in a talk on "The Design of RF and Video Amplifiers" given by the author as one of a series of six lectures on *Transistors — Their Circuits and Applications*, sponsored by the Dallas, Texas, Section of the Institute of Radio Engineers.

to provide engineering illustrations of these principles. References are then made to published material where more complete details on specific amplifiers can be found.

II. HIGH-FREQUENCY TRANSISTOR CHARACTERIZATION

Before considering amplifier design techniques, we must have some means of characterizing the transistor in terms of its performance as an electrical circuit element.* This paper will rely largely on a small-signal characterization in which the transistor is represented by the generalized equivalent T of Fig. 1 with a single internal generator in the branch corresponding to the collector of the transistor. The details of the impedances, each one of which can be written in terms of lumped constants

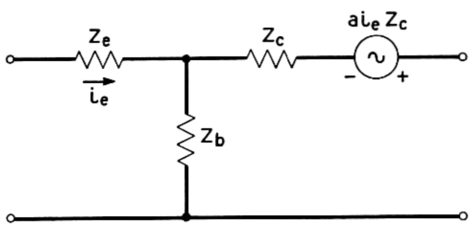

Fig. 1 — Transistor equivalent T.

that are directly relatable to the physical structure of the transistor, will be presented only when needed.

For simplicity in writing circuit equations, the transistor collector current generator, ai_e, which would normally appear across the collector impedance Z_c, has been replaced by the voltage generator, ai_eZ_c, in series with Z_c, in accordance with Thevenin's theorem. (a is frequency-dependent.) Except for a small phase error, a is closely approximated at frequencies below f_a by the expression

$$a = \frac{a_0}{1 + j\frac{f}{f_a}}, \tag{1}$$

where f_a is the frequency at which the amplitude of a is 3 db below its low frequency value, a_0.[2] For simplicity α, the short-circuit common base current gain will be used interchangeably with a in the discussion to follow.

No parasitic capacitances are shown in Fig. 1, since these will be con-

* For a resume of transistor equivalent circuits, see Pritchard.[1]

sidered as part of the terminating networks, except for the capacitance between output and input — that is, collector-to-emitter capacitance in the common base configuration and collector-to-base capacitance in the common emitter configuration. And, in order to simplify the consideration of the internal feedback effects, these latter capacitances will be neglected except in the discussion of the common emitter neutralized amplifier. However, input-to-output capacitance can be very troublesome, especially in the common base configuration at very high frequencies.

The equivalent T circuit representation illustrated in Fig. 1 is particularly useful in three respects. First, it represents the transistor with sufficient accuracy to be used in generalized circuit evaluation. Secondly, it can be used in any of the three possible transistor connections without change. Finally, since the various components of the circuit are directly relatable to the physical structure of the transistor, the effect of the transistor structure on circuit performance can be better understood, and effects that might otherwise be obscured may be uncovered.

When a precise amplifier circuit design in a particular frequency region is undertaken, a four-pole parameter circuit equivalence may be more accurate and more convenient.* However, this paper will make only limited use of this type of characterization for two reasons. First, the examination of amplifier stability, which is one of the major objectives of the paper, is more easily accomplished with the equivalent T of Fig. 1. In fact, the positive nature of the internal feedback of the transistor is not apparent in the hybrid parameter four-pole analysis of the common emitter transistor configuration. This is because the positive feedback is concealed in the forward transfer current ratio, h_{21e}. Secondly, many electronic circuit engineers are more accustomed to the two-terminal design techniques of vacuum tube circuitry below the UHF region. And this paper shows that those amplifiers that can be built on a two-terminal basis with limited impedance measurements and slide rule computations are often the better transistor amplifiers.

III. RADIO FREQUENCY AMPLIFIER DESIGN

In designing a radio frequency transistor amplifier, the immediate problem is to determine the proper choice of terminal networks for the transistor to obtain the greatest possible gain consistent with the other requirements on the amplifier. The first approach to a solution is given by linear network theory. A conjugate-matched-impedance generator

* For a presentation of the more common four-pole parameter equivalences, see Linvill and Schimpf.[3]

should be connected to the input and a conjugate-matched-impedance load should be connected to the output. However, since the transistor itself is a network of complex impedances ("complex" is used here and hereafter in the sense of having real and imaginary parts) and contains an internal generator that is a complex function of frequency, and also has built-in internal feedback, the required generator and load impedances are themselves complex. To say that the determination of these required impedances is difficult is a gross understatement. Even the computation of the transistor gain between known complex impedances becomes unduly complicated.

An alternate approach to the determination of suitable generator and load impedances is therefore used. The power delivered to the load with either a constant-current or a constant-voltage input generator is determined, and the input-matching problem is then considered separately. This approach will be illustrated by considering an elementary design of a 4-mc wide, 30-mc center-band frequency common base IF amplifier, using a 30-mc alpha-cutoff-frequency germanium transistor. The simplified equivalent T circuit of the transistor is shown in Fig. 2(a). The load impedance should be conjugately matched to the output impedance of the transistor with an open circuit input, since a current generator is being assumed at the input. This impedance is closely approximated by the reactance of the collector junction capacitance, C_c. A positive reactance equal to the negative reactance of C_c at the center-band frequency of 30 mc is therefore chosen as the reactance portion of the load. This is the 14-microhenry inductance of Fig. 2(b). Since the resistance component of the transistor output impedance is very small, bandwidth considerations rather than matching determine the resistance component of the load. A 4-mc bandwidth centered at 30 mc calls for a 19.9K

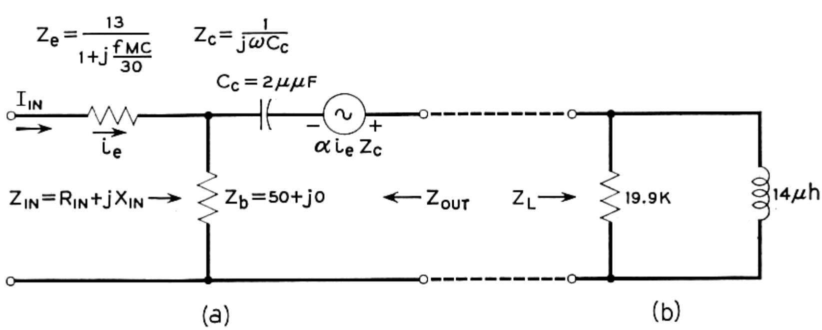

Fig. 2 — Transistor 30-mc single-stage IF amplifier: (a) transistor equivalent circuit; (b) amplifier load impedance.

shunt resistance, as shown in Fig. 2(b). Now, in accordance with the design plan, an input generator conjugately matched to the transistor input impedance with the output terminated in the selected load should be provided to complete the single-stage amplifier design. This input impedance, Z_{IN}, was computed, and its resistance and reactance components are plotted in Fig. 3(a) and 3(b) respectively. The resistance component is seen to vary by a factor of 5 to 1 throughout the desired band, and actually becomes negative at frequencies just below the bottom of the band. The reactance component likewise varies widely throughout the band, going from approximately 200 ohms at the bottom to zero at the top of the band. Anything but a conjugately matched generator at the input would distort the bandpass characteristic designed into the load impedance. Since this generator would have to incorporate the output impedance of the preceding transistor in a multistage amplifier, plus a suitable impedance transformation to obtain gain, its design would be at best very complicated. The design is therefore in serious trouble. An understanding of the source of the trouble is essential to a solution to the problem.

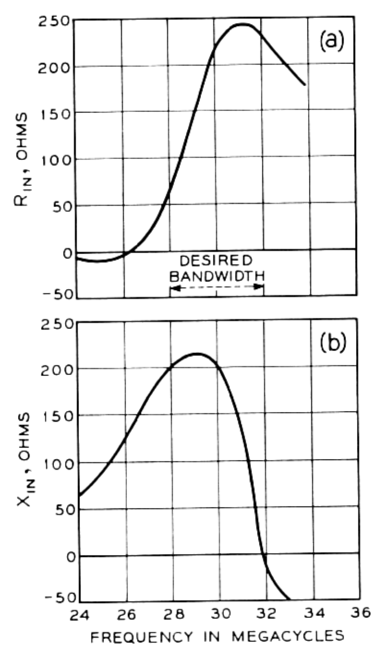

Fig. 3 — Input impedance for amplifier of Fig. 2: (a) resistance component; (b) reactance component.

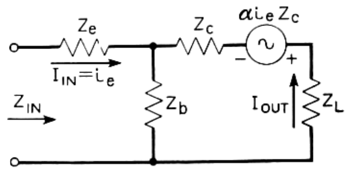

Fig. 4 — Equivalent T for common base connection with load Z_L .

The difficult nature of this input impedance is a direct result of the internal feedback in the transistor. Fig. 4 shows the equivalent circuit of Fig. 1 in the common base connection terminated with a constant-current generator input, I_{IN} , and a load impedance, Z_L . The input impedance Z_{IN} is given by

$$Z_{IN} \doteq Z_e + Z_b \left(1 - \frac{\alpha}{1 + \frac{Z_L}{Z_C}} \right). \tag{2}$$

In the common emitter connection, the other of the two more commonly used transistor configurations, the corresponding input impedance is given by

$$Z_{IN} \doteq Z_b + Z_e \left(1 + \frac{\alpha}{(1 - \alpha)\left[1 + \frac{Z_L}{Z_C\,(1 - \alpha)} \right]} \right). \tag{3}$$

Equations (2) and (3) show that, regardless of the common connection, the input impedance to the transistor is a function of the load impedance and the common base short-circuit current gain α, both of which are, as a rule, complex. And so the complicated complex input impedance that was uncovered in the amplifier example above follows naturally. Even if the design problem were not so complicated, and if physically realizable impedances with the proper impedance transformation for interstages of multistage amplifiers could be built, the alignment problem of the multistage amplifier would be an extremely difficult one. This is verified in the large mass of technical literature discussing interstage alignment and band-skewing problems as a result of adjacent interstage interaction. It is therefore apparent that, before practical high-frequency transistor amplifiers can be built, it is necessary to reduce the effect of the load impedance on the input impedance to a point where it is no longer a serious problem. This can be done either by "neutralization"* or by output-to-load-impedance mismatch.

* Neutralization is placed in quotation marks to call attention to the fact that it is quite different from neutralization as we know it in vacuum tubes. The characteristics of transistor neutralization will be discussed in more detail later.

IV. NEUTRALIZED AMPLIFIER DESIGN

The neutralized solution to the input-output impedance interaction will be considered first. The common emitter connection will be used, since this is the more common neutralized configuration. This is because more gain is obtainable in this connection at frequencies below the common base cutoff frequency of the transistor. Fig. 5 shows the transistor equivalent circuit of Fig. 1 in the common emitter connection, with a neutralizing impedance, Z_N, connected between the collector output and the base input and with the base input open. An external generator, V_g, is connected between the collector and the common emitter terminals. The following equations define the voltage and current relations of the circuit of Fig. 5:

$$\begin{aligned} i_1 (Z_e + Z_c - \alpha Z_c) + i_2 Z_c \qquad\qquad &= V_g , \\ i_1 (Z_c - \alpha Z_c) \qquad + i_2(Z_b + Z_c + Z_N) &= 0, \end{aligned} \tag{4}$$

$$i_1 = \frac{V_g \Delta_{11}}{\Delta} \qquad i_2 = \frac{V_g \Delta_{12}}{\Delta}, \tag{5}$$

where Δ is the circuit determinant of (4). Then

$$E_1 = i_2 Z_b - i_1 Z_e = V_g \left[\frac{Z_b \Delta_{12} - Z_e \Delta_{11}}{\Delta} \right], \tag{6}$$

which gives the input voltage, E_1, in terms of the output generator, V_g. If E_1 is made zero regardless of the value of V_g, the input impedance is then independent of the load voltage and therefore of the load impedance when the amplifier is terminated at its output. Solving for the value of Z_N required to make $E_1 = 0$ gives

$$-Z_N = Z_c + \frac{Z_b}{Z_e} Z_c(1 - \alpha) + Z_b . \tag{7}$$

The required neutralization impedance, Z_N, turns out to be negative, which indicates that a phase reversal is needed in the neutralization

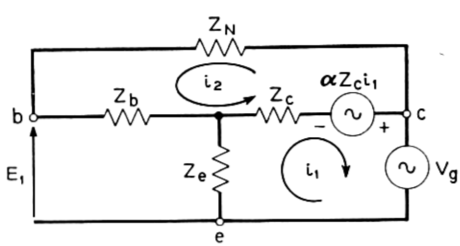

Fig. 5 — Equivalent T for common emitter connection with neutralization.

current feedback path in order to produce a positive neutralization impedance, Z'_n, which is the negative of Z_n. A phase-reversing transformer is therefore used, as shown in Fig. 6, which gives a generalized schematic diagram of a common emitter neutralized amplifier. A step-down is used between the collector and neutralizing windings of the phase-reversing transformer in order to distribute the effect of the loading of the neutralization impedance between the collector and the emitter. The transformer is tuned to the desired center-band frequency, and the load is the input impedance to the following identical stage in a multistage amplifier. If the approximation of (1) for α and $-j/\omega C_c$ for Z_c are substituted in (7) for Z_n, and if Z_b and Z_e are assumed to be real and constant, then (7) can be solved for Z'_n in terms of its real and imaginary parts:

$$Z'_n = -Z_n = Z_b \left\{ \frac{1}{\omega_\alpha C_c Z_e} \left[\frac{\alpha_0}{1 + \left(\frac{f}{f_\alpha}\right)^2} \right] + 1 \right\} - j \frac{1}{\omega C_c} \left\{ 1 + \frac{Z_b}{Z_e} \left[1 - \frac{\alpha_0}{1 + \left(\frac{f}{f_\alpha}\right)^2} \right] \right\} \tag{8}$$

As given by (8), Z'_n can be approximated by two resistances and a capacitance throughout a reasonably broad band of frequency, as shown in the network for Z'_n in Fig. 6. The dotted capacitances, C_{cb} and nC_{cb}, of Fig. 6, show how the input-to-output capacitance in the neutralized common emitter amplifier can be compensated for by a corresponding capacitance in the neutralizing impedance. Only this portion of the neutralization corresponds to the neutralization of the output-to-input capacitance feedback in vacuum tubes. The load impedance, Z_L, of Fig. 6 is given in terms of the input admittance to the following transistor

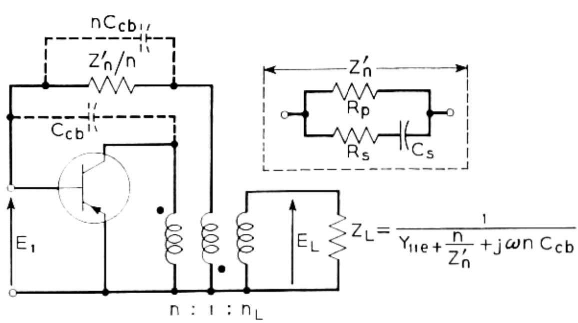

Fig. 6 — Schematic of single-stage common emitter neutralized amplifier.

with its output short-circuited designated as Y_{11e} and the neutralization impedance of the following transistor.

Although (7) gives the neutralization impedance in terms of the transistor equivalent T of Fig. 1 and is useful for qualitative understanding of the neutralization problem, it is not suffiicently accurate to determine the neutralization impedance required for an actual amplifier. A more accurate determination can be made from a four-pole parameter solution to the impedance Z'_n necessary to make Y_{12e}, the reverse transfer admittance of the circuit of Fig. 6, equal to zero. This was the technique used by Webster[4] in determining an expression for Z'_n in connection with the design of one of the best examples of a maximum gain neutralized common emitter amplifier to be found in the literature. However, even the four-pole approach fails to give a satisfactory determination of Z'_n for practical use, and so Z'_n is usually obtained experimentally by adjusting Z'_n until there is no appreciable change in the input impedance to the transistor across the bandwidth of the amplifier when the load is alternately normal and shorted. The input admittance is then given by

$$Y_{IN} = \frac{1}{Z_b + \dfrac{Z_e}{1 - \alpha}} + \frac{1}{\dfrac{Z'_n}{n}} + j\omega n C_{cb}, \tag{9}$$

which is the common emitter input admittance of the transistor with the collector shorted to the emitter plus the admittance added across the input by the neutralization impedance. The input admittance given by (9) is seen to be independent of the load impedance, and therefore the objective of having input impedance independent of output impedance is achieved.

The load impedance is then conjugately matched to the output impedance of the transistor with the input shorted. The generator is likewise conjugately matched to the input impedance given by (9). The power gain of the transistor can then be easily shown to be given by[4]

$$\text{power gain} = \frac{|\,Y_{21}\,|^2}{4G_{22}G_{11e}}, \tag{10}$$

where Y_{21} is the forward transfer admittance of the transistor with the output short circuited, G_{22} is the real part of the output admittance with the input short circuited, and G_{11e} is the real part of the input admittance common emitter with the output short circuited. This is a straightforward computation, since all the parameters are simple functions of the active device only and can be measured on a suitable impedance bridge as discussed by Webster.[4]

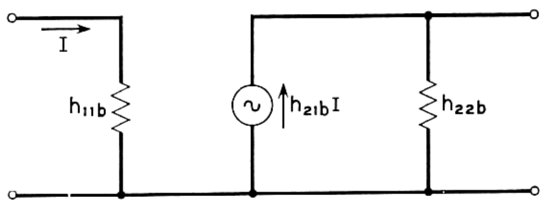

Fig. 7 — Transistor circuit, approximate.

Webster has built a five-stage, 75-db gain, 25-mc center-band frequency IF amplifier using the neutralization technique just discussed. The computed and measured amplifier checked to within 0.5 db in 75 db in gain and to 0.07 mc in 1.6 mc in bandwidth. This is an excellent example of the accuracy of the neutralization amplifier design technique in the prediction of available gain and bandwidth. However, a maximum-gain neutralized amplifier is far from easy to design, since the amplifier represents a delicate balance of feedback effects, which make it difficult to adjust and even more difficult to maintain stable. This will be discussed in greater detail later.

V. MISMATCHED AMPLIFIER DESIGN

The mismatch approach to making the input impedance independent of the output impedance will next be considered. Here we will use the common base connection for our discussion, since this connection has been most frequently used for mismatched RF amplifiers. However, mismatched common emitter RF amplifiers are becoming more frequent, due to the high-cutoff-frequency diffusion transistors currently available. The same principles apply to both types. A reexamination of (2) shows that the common base input impedance can be made substantially independent of load impedance if the load impedance, Z_L, is made small compared to the collector impedance, Z_c. This, of course, involves a loss of gain, but a considerable mismatch can be taken with a relatively small loss of gain. For instance, a 5-to-1 mismatch results in a gain reduction of less than 3 db, and a 10-to-1 mismatch results in a reduction of only 5 db.

With sufficient mismatch to make the input impedance essentially independent of the output impedance, the common base equivalent circuit of the transistor with a constant current generator is given in Fig. 7.* In this circuit, h_{11b} is the impedance looking into the input of the com-

* The equivalent circuit of Fig. 7 has been referred to by Linvill[3] as a "circuit approximate".

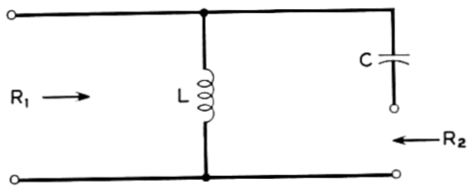

Fig. 8 — Single-tuned reactance-network transformer.

mon base transistor with output short circuited, h_{21b} is the ratio of the common base short-circuit output current to the input current and h_{22b} is the admittance looking into the common base output with the input open.

Even if the degree of mismatch is not great, the circuit of Fig. 7 gives a fair approximation to the true circuit. The approximation is sufficiently good to rough out interstage coupling networks, which can then be adjusted on the bench in the laboratory. Since there is little interaction of the output load circuit on the input impedance, the gain for a given load and generator impedance can be easily computed, in the same manner in which the gain is computed for a neutralized amplifier. It will be somewhat simpler, since no neutralization impedance is present. It must be remembered, however, that the output is no longer matched when the output power is computed.

Since a reversing transformer is not needed for neutralization purposes, simple impedance transformation between the high-impedance collector output of one stage and the low-impedance emitter input of the following stage can be used. The simplest type of impedance transformation corresponding to a single-tuned transformer is shown in Fig. 8. If R_2 is the resistance component of the impedance looking into the emitter of the following stage, and if the reactance component of the impedance is combined with the reactance of C, the load impedance R_1 facing the collector of the preceding stage at center frequency is increased to Q^2R_2, where Q is the ratio of the band center frequency reactance of either C or L to R_2. This simple circuit has the disadvantage that the circuit Q, and consequently the transmission bandwidth, and the transformation ratio are not independent. Therefore, a double-tuned reactance transformation network equivalent to a double-tuned transformer is usually used in the interstage.

Fig. 9 gives the schematic circuit of a single stage of a 70-mc germanium tetrode mismatched amplifier designed by Schimpf[3] using a double-tuned reactance transformation network. The short-circuited input impedance to the transistor was of the order of 75 ohms. This impedance also had a reactance component which varied somewhat throughout the transmitted band, but the variation was not sufficient to seriously complicate the band-width adjustment of the interstage network. The high-side impedance looking into the coupling network was approximately 1500 ohms, which gave sufficient impedance transformation ratio to provide substantial stage gain (approximately 9 db per stage) but presented sufficient mismatch to the collector of the preceding transistor to meet the requirements of making the input impedance substantially equal to the short-circuited input impedance — or, in other words, independent of the load. At the time this amplifier was designed, 70-mc impedance-measuring equipment was not available to the designer. Therefore, judicious extrapolations were made from measurements on a radio frequency bridge at frequencies of 30 mc and below. The interstage transformation network was then designed and built with the adjustable elements shown in Fig. 9. The circuit was then bench-adjusted in the laboratory with a sweep-frequency signal generator and a high-frequency oscilloscope across the load. This is the technique that was referred to earlier when it was stated that excellent amplifiers can be built without complicated impedance measurements and a minimum of slide rule computations.

The relative independence of this circuit design technique on transistor parameters was dramatically demonstrated when Schimpf placed one of the first research models of the germanium diffused-base transistor in a circuit that, except for the omission of the second base of the tetrode, was substantially identical with the circuit of Fig. 9. In spite of the wide difference in electrical characteristics of the diffused-base and tetrode

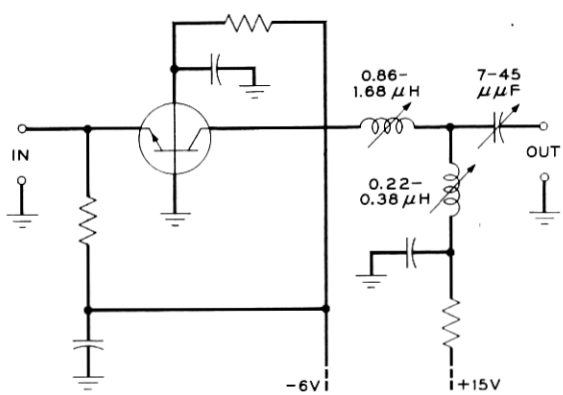

Fig. 9 — Single-stage 70-mc mismatched IF amplifier.

transistors, the circuit was alignable to give a 20-mc IF band centered at 70 mc. Because of the superior high-frequency performance of the diffused-base transistor, the stage gain was 14 db, as compared with 9 db for the tetrode, and the gain was flat to ±0.1 db across the 20-mc band.

VI. TRANSISTOR AMPLIFIER STABILITY

Up to now nothing has been said about the stability of the two types of amplifiers which have been described. And since stability of broadband transistor amplifiers is one of the most important design considerations, the relative stability of the two types of amplifiers that have just been described will now be considered.

Bode[5] has pointed out that the stability of any active network can be determined in terms of the positions of its poles and zeros in the complex frequency plane. However, if we have a known structure whose gain characteristics are satisfactory, it is a long and tedious process in general to determine whether the roots of the structure meet the stability requirement. Furthermore, if the structure is not stable this approach does not necessarily tell us what to do to make it stable. What is needed, therefore, is some means of transferring the restrictions on the poles and zeros into equivalent restrictions on the behavior of the circuit at real frequencies. This we have in the Nyquist criterion of stability,[6] which is used so effectively in the design of negative feedback amplifiers and which it is proposed that we use in the evaluation of the stability of our transistor amplifiers.

The Nyquist criterion of stability is simply stated. The open-feedback loop gain of a feedback amplifier — usually referred to as $\mu\beta$ — is determined in magnitude and phase across a frequency band broad enough to include all frequencies at which the gain is greater than 0.1 in magnitude. The individual values of magnitude and phase are then plotted in polar coordinates and connected to form a closed loop terminating close to the origin. If this loop encloses the point (1,0), the amplifier is unstable; if it does not, it is stable.[6] However, the external gain of the amplifier may be extremely sensitive to changes in amplifier components at frequencies where $\mu\beta$ is in the close vicinity of the (1,0) point, normally called the Nyquist point. In soundly designed negative feedback amplifiers the Nyquist plot approaches the Nyquist point only at frequencies well outside the useful frequency band of the amplifier.

Suppose now that the Nyquist criterion of stability is used to examine critically the stability of our transistor amplifiers. It is generally known that the transistor has built-in feedback, due to its internal base resist-

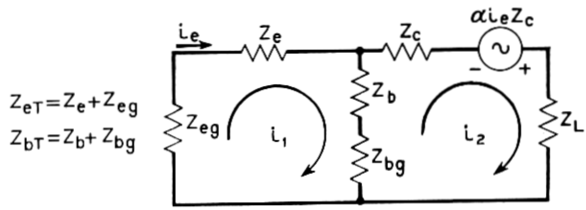

Fig. 10 — Generalized transistor amplifier equivalent circuit.

ance. Furthermore, the behavior of the common emitter transistor amplifier is analogous to that of the common cathode vacuum tube amplifier, since there is a reversal of phase of signal from input to output. Therefore, a reduction in forward gain occurs when a portion of the output signal is fed back to the input through a nonphase-reversing external circuit. As a result, the common emitter configuration is usually considered to be a negative feedback connection.* However, although the incremental feedback due to the external feedback path is negative, the residual or net feedback of the transistor considered as a single-stage amplifier is still positive. In fact, unless external feedback is applied through a suitably phased impedance-matching transformer or other active amplifying devices, a single-stage transistor amplifier—or a single stage of a multistage transistor amplifier—is always of itself a positive feedback amplifier, regardless of which of its elements is made the common connection of the stage.

The positive nature of the feedback is demonstrated in Fig. 10. Here the generalized equivalent circuit of a transistor is shown with generator impedances, Z_{eg} and Z_{bg} , in the emitter and base circuits respectively, and a load impedance, Z_L , in the collector circuit. This load impedance could just as well have been made a generator impedance, thereby making the circuit completely general for any transistor connection. The equations relating the voltages and currents of Fig. 10 are

$$\begin{aligned} i_1(Z_{eT} + Z_{bT}) - i_2 Z_{bT} &= 0, \\ -i_1(\alpha Z_c + Z_{bT}) + i_2(Z_{bT} + Z_c + Z_L) &= 0, \end{aligned} \tag{11}$$

where $Z_{eT} = Z_e + Z_{eg}$, the total impedance in the emitter, and $Z_{bT} = Z_b + Z_{bg}$, the total impedance in the base. The feedback loop gain $\mu\beta$ for this circuit is given by

$$\mu\beta = 1 - \frac{\Delta}{\Delta_0} \doteq \alpha \frac{Z_{bT}}{Z_{eT} + Z_{bT}} \frac{Z_c}{Z_c + Z_L}, \tag{12}$$

* This misconception has been strengthened by hybrid four-pole analysis of the common emitter transistor, since the positive feedback is concealed in the forward transfer parameter or common emitter short-circuit current gain.

where Δ is the circuit determinant of (11) and Δ_0 is the circuit determinant when the active generator, α, is 0. If all the impedances except Z_c are resistive and Z_L is somewhat smaller than Z_c, the feedback loop gain below the common base cutoff frequency will always fall in the right half of the Nyquist polar plot, indicating that the feedback is positive. Note that the choice of the common transistor connection does not influence this result.* In the common emitter iterative amplifier, where the total impedance in the base is much greater than the total impedance in the emitter and the load impedance is small compared to the collector impedance, the feedback loop gain is given by

$$\mu\beta \doteq \alpha = \frac{\alpha_0}{1 + j\dfrac{f}{f_\alpha}}, \tag{13}$$

or the common base short-circuit current gain of the transistor.† Since α has a frequency characteristic which for purposes of discussion can be approximated by an RC cutoff as shown in (13), the Nyquist diagram for this common emitter amplifier becomes a semicircle of diameter α_0, with its center at $\alpha_0/2$ on the zero phase axis and situated below the zero phase axis as shown in the Nyquist plot (a) of Fig. 11.‡ As α_0 approaches unity, a desirable characteristic in a common emitter amplifier, the Nyquist diagram approaches the Nyquist point, (1,0).

The high- and low-frequency cutoff portions of the Nyquist diagram for a soundly designed negative feedback amplifier are also shown in the Nyquist plot (b) of Fig. 11. Note that the negative feedback amplifier stays out of the shaded area bounded by the $\pm 30°$ axes and gain magnitude greater than 0.5. This shaded area represents the stability margins usually maintained for well-designed negative feedback amplifiers, and corresponds to a loop gain of less than 0.5, or -6 db when the loop phase is between $\pm 30°$. This requirement is strictly for stability margins against oscillation in the frequency regions where positive feedback occurs and, these regions are well above or well below the operating amplification band of the amplifier. In contrast, the useful amplification band of our common emitter amplifier falls on that portion of its $\mu\beta$ diagram

* The positive nature of the internal transistor feedback regardless of the common terminal of the transistor has been confirmed by R. B. Blackman of the mathematical research department of Bell Telephone Laboratories.

† Equation (13) neglects a passive component of the feedback loop gain or return ratio which is negligibly small.

‡ The Nyquist plot should include $\mu\beta$ plotted with its imaginary part negative of normal as well as normal. This returns the loop to zero for amplifiers whose gain is not zero at dc as in the present case. However, since this type of plot merely gives a mirror image across its 0–180° axis with the imaginary part of $\mu\beta$ having its normal sign, this half of the plot is not usually shown.

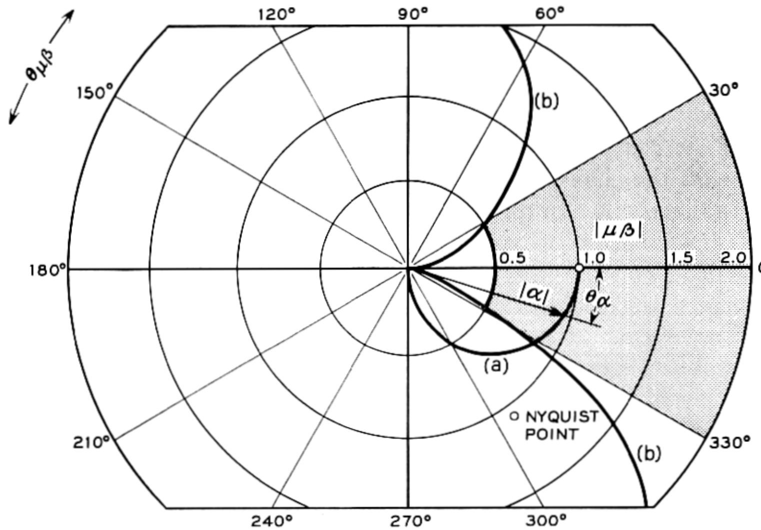

Fig. 11 — Nyquist diagram for amplifier of Fig. 10 with $Z_{bT} \gg Z_{et}$ and $Z_L \ll Z_c$

that is in closest proximity to the Nyquist point. The only reason why this is tolerable is because α_0 is a function of the physical structure of the device and because a well-designed and well-behaved junction transistor can be counted upon to stay reasonably constant. In any event, if the dc biases are held reasonably constant, α_0 can be expected to remain less than unity where circuit oscillation will not occur. However, even though there is little danger of oscillation, the positive nature of the inband feedback of the transistor is the basis of the stability problem in high-frequency amplifier design, as will be shown.

Fig. 12(a) shows the schematic diagram of a single-stage neutralized amplifier having a 4-mc bandwidth centered at 25 mc. This amplifier is similar to the amplifier designed by Webster.[4] The feedback loop gain of the amplifier has been computed using a mathematical trick for opening the feedback loop suggested by Blackman.[7] This trick consists of inserting a generator current, i_e, in the emitter and computing the current returned to the emitter, $\hat{i}_e$, through the two feedback paths — the internal feedback of the transistor and the feedback through the neutralization impedance, Z_N/n. The significance of the "^" is that the current so designated is not reamplified by the current gain α of the transistor or, in other words, that the loop transmission is mathematically stopped at a single round trip. The feedback loop gain is then given by $\hat{i}_e/i_e$, and can be written by inspection from the schematic circuit of

Fig. 12(a). This feedback loop gain is given to a close approximation by

$$\mu\beta = \frac{\hat{i}_e}{i_e} = |\,\mu\beta\,|\,\theta_{\mu\beta}$$

$$\doteq \frac{\alpha Z_c}{Z_c + Z_L}\,\frac{Z_b + Z_s}{Z_b + Z_s + Z_e} + \frac{\alpha Z_c}{Z_c + Z_L}\,\frac{Z_L}{Z_N}\,\frac{Z_s}{Z_b + Z_s + Z_e} \quad (14)$$

$$\doteq \frac{\alpha Z_c}{Z_c + Z_L}\left[1 + \frac{\dfrac{Z_L Z_s}{Z_N} - Z_e}{Z_b + Z_s + Z_e}\right].$$

The loop gain given by (14) was computed across a band of frequencies extending well above and well below the pass band of the amplifer. These

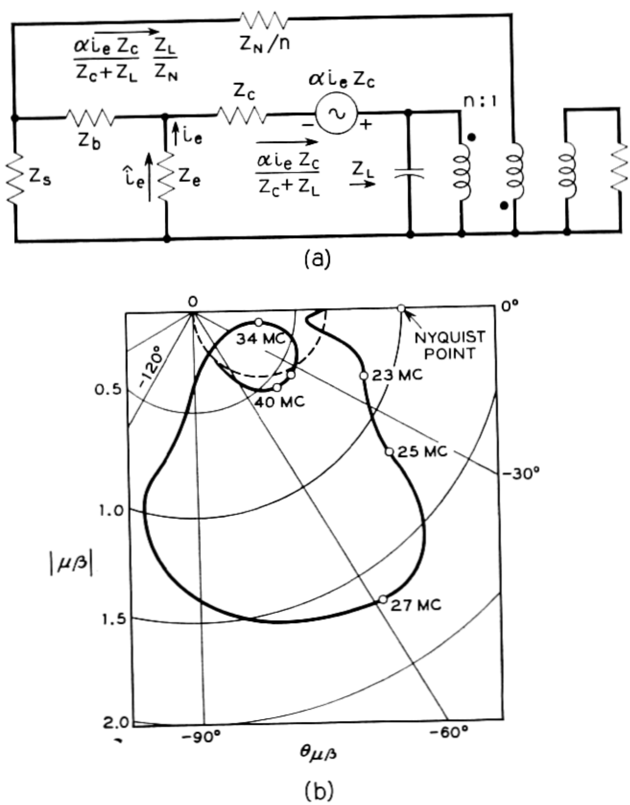

Fig. 12 — Stability evaluation of neutralized amplifier: (a) circuit schematic; (b) Nyquist diagram.

gains were plotted on a polar diagram to give the Nyquist diagram of Fig. 12(b). It will be noted that the Nyquist diagram does not include the Nyquist point (1,0), and therefore the amplifier would be expected to be stable in the sense of being free from oscillation. However, if the sensitivity of the amplifier to changes in transistor parameters caused by normal drift, battery changes or ambient conditions were examined, large changes in gain would be expected, in view of the fact that the feedback loop gain has real parts in excess of unity at angles of the order of 30° within the transmission band. A complete study of sensitivity can be made in accordance with the techniques described (Ref. 5, Chapters 4 through 6). The fact that the transmission gain of this amplifier is sensitive to environmental and transistor parameter changes due to bias shifts has been confirmed experimentally.

At this point, it is well to stop and consider the nature of the feedback through the neutralization impedance, Z_N. The common emitter connection owes its high current gain to the internal positive feedback in the transistor, which was discussed above. The open input–short-circuited output common emitter amplifier has a Nyquist diagram falling very close to the Nyquist critical point (1,0). [See (12) and (13) and Nyquist diagram (a) of Fig. 11.] When finite impedances are placed in the collector and base circuits, the real part of the positive feedback is reduced, or there is an increment of negative feedback introduced. This moves the Nyquist diagram away from the critical positive feedback area, or in the direction of greater amplifier stability. However, the internal feedback residue is still positive. When the neutralization circuit is added for the purpose of removing the dependence of input impedance on load impedance, the direction of the current through the neutralization circuit is such as to cancel the negative increment of feedback mentioned above. It therefore adds a positive increment to the internal positive feedback residue, and moves the total feedback in the direction of the Nyquist critical point or in the direction of lesser amplifier stability. In the iterative common emitter amplifier, the Nyquist diagram was held within the stability requirement by α_0 holding less than unity. In the neutralized amplifier, a small shift in the critical balance between the positive feedback neutralization current and the negative increment of internal positive feedback caused by the finite impedance terminations can move the Nyquist diagram beyond the (1,0) critical point, and the amplifier will oscillate. Anyone who has built a "maximum available gain" neutralized amplifier is aware of the criticalness of this balance and the tendency toward oscillation.[8] At best, the critical feedback balance results in a gain-sensitive amplifier, as pointed out above in the discussion of the Nyquist diagram of Fig. 12(b).

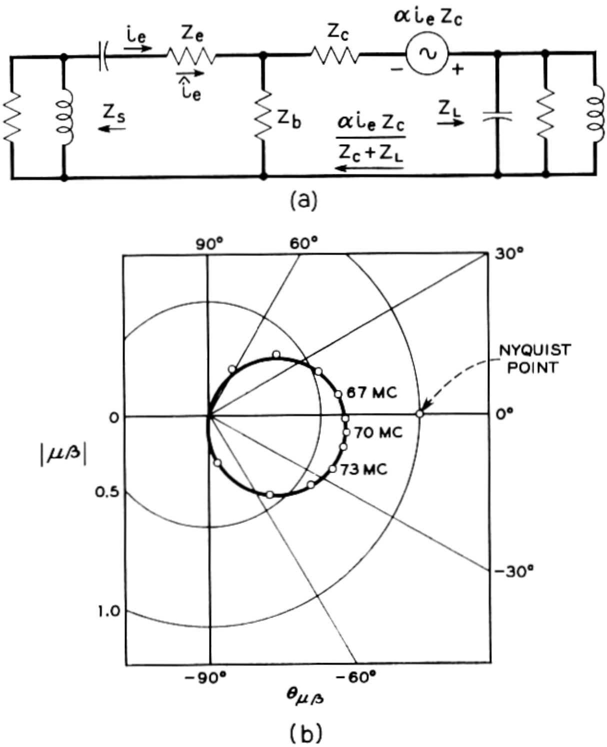

Fig. 13 — Stability evaluation of mismatched amplifier: (a) equivalent circuit; (b) Nyquist diagram.

Now consider the stability of the mismatched type of transistor RF amplifier. Fig. 13(a) shows the schematic circuit of a common base mismatched amplifier using single-tuned reactance coupling of the type referred to above. This amplifier has a transmission bandwidth of 6 mc centered at 70 mc, and a stage gain of approximately 10 db. Looking from the collector into the emitter of the following stage, one sees an effective moderate-Q parallel-resonant circuit whereas, looking from the emitter back toward the collector, one sees a rather high-Q series-resonant tuned circuit, as is shown in the schematic. The feedback loop gain of this amplifier was obtained by the technique used for the amplifier of Fig. 12(a). This gain is given by

$$\mu\beta = \frac{\hat{i}_e}{i_e} = |\,\mu\beta\,|\,\theta$$

$$\doteq \frac{\alpha Z_c}{Z_c + Z_L}\,\frac{Z_b}{Z_b + Z_e + Z_s}$$

or

$$\mu\beta \doteq \alpha \frac{Z_b}{Z_b + Z_e + Z_s}, \tag{15}$$

when Z_L is appreciably less than Z_c, the condition for mismatch design. Using (15), the loop gain was computed at a sufficient number of points to give the Nyquist diagram shown in Fig. 13(b). The center-band and band-edge frequency points are indicated on the Nyquist diagram. Note the inevitable positive feedback in the transmission band. However, the feedback though positive is fractional (i.e., there is a net loss around the feedback loop) and well below the critical unity value. Furthermore, the loop feedback gain decreases and rapidly goes to zero both above and below the center-band frequencies. An examination of the circuit and the feedback loop gain given by (15) shows the reason for this. Since the load impedance, Z_L, is a moderately high-Q parallel-resonant circuit, it rapidly approaches zero at frequencies outside the transmission band, thereby increasing the degree of mismatch away from the center-band frequency. At the same time, Z_s, the source or generator impedance, is an even greater-Q series-resonant circuit, so that it reaches a high impedance very rapidly away from the center-band frequency. Since Z_s appears only in the denominator of the expression of (15) for feedback loop gain, this means that $\mu\beta$ rapidly goes to zero, due to the high outband impedance of the generator, Z_s. The same behavior would be experienced with a double-tuned interstage circuit, except that the $\mu\beta$ diagram would consist of two loops, due to the added pole and zero in the reactance interstage. These two loops would both pull away from the Nyquist point area towards the origin in the same manner as does the loop gain of Fig. 13(b). Because of the avoidance of positive feedbacks having real parts approaching unity, it would be expected that the mismatch amplifiers would be not only more stable, but also less sensitive to changes than are the neutralized amplifiers, and this is confirmed by experimental results. The price paid for this improved stability and reduction in gain sensitivity is lower stage gain. In return for the gain sacrifice, we also obtain greater ease of design, greater ease of interstage alignment and less complicated circuitry.

How then does one decide on the choice of the neutralized or mismatched techniques? The answer to this question is largely dependent upon economic and system requirement considerations. If a consumer product is being designed where competition demands maximum gain to keep down cost factors and where the failure of an amplifier means only an occasional service call, then the maximum-gain neutralized amplifier might be selected. However, if a system is being designed where amplifier failure would cause malfunctioning of a large and costly system, reliability considerations would favor the more conservative mismatch approach, in spite of the lower stage gains obtained. Intermediate situations might suggest a combination of neutralization and mismatch, with higher gains than could be obtained with the straight mismatched amplifier and with feedback loop diagrams midway between the extremes of Figs. 12 and 13. It is interesting to note that, with the great reduction in the collector capacitances of VHF transistors, the mismatch that automatically occurs from the impracticability of simultaneously obtaining output matching and very broadband interstages results in a compromise mismatch-neutralized circuit of the type just mentioned. Actually, the experience with these circuits has shown that the neutralization is not critical when the degree of mismatch is fairly high, and may be omitted.

VII. VIDEO AMPLIFIER DESIGN

In the design of video amplifiers, the mismatch approach is practically dictated by the broadband requirements and the limitation on the maximum impedance available with a given irreducible circuit capacitance, in accordance with the Bode resistance integral theorem (Ref. 5, Chapters 4 through 6). And so we can use the high common emitter current gain without danger of circuit oscillation. However, the gain sensitivity problem still exists, as will be shown.

With the new high-frequency-cutoff diffusion transistors, common emitter short-circuit current gains of 12 db and higher at 100 mc are now commercially available. These make possible common emitter iterative amplifiers with the collector of one transistor coupled directly into the base of the following transistor — except for a blocking condenser when simple bias circuits are required. Such an amplifier was built by C. E. Paul of Bell Telephone Laboratories with early models of the germanium diffused-base transistor. A picture of this amplifier is shown in Fig. 14. The amplifier has three common emitter iterative stages, a gain of 70 db and a bandwidth of close to 10 mc using the simplest possible resistance-capacitance coupled interstages. The amplifier requires a total power of less than 100 milliwatts and occupies a volume of less than 2 cubic inches. This amplifier demonstrates the great potential of the common emitter transistor connection in video circuits.

With the simple iterative common emitter amplifier, the single-stage bandwidth is determined by $(1 - \alpha_0)f_\alpha$, where f_α is the common base cutoff frequency. This bandwidth will vary widely from transistor to transistor, due to variations in α_0 and f_α. If bandwidths narrower than $(1 - \alpha_0)f_\alpha$ are needed, they can be obtained most easily by choosing transistors with higher α_0 or lower f_α. However, for today's broadband video and baseband amplifiers, bandwidths greater than the normal common emitter bandwidths are frequently required, and some means of trading gain for bandwidth is needed. This can be accomplished by feeding back a portion of the output signal to the input, in accordance with the technique illustrated in Fig. 15. Fig. 15(a) shows a single-stage common emitter amplifier in which the load impedance is small compared to the collector impedance, a situation which exists in the iterative common emitter amplifier. The current gain of the amplifier is given by

$$\frac{i_2}{i_1} \doteq \frac{\alpha}{1 - \alpha}. \tag{16}$$

The current gain given by (16) is plotted as curve A of Fig. 15(b) for a transistor having an α_0 of 0.97 and a common emitter cutoff frequency

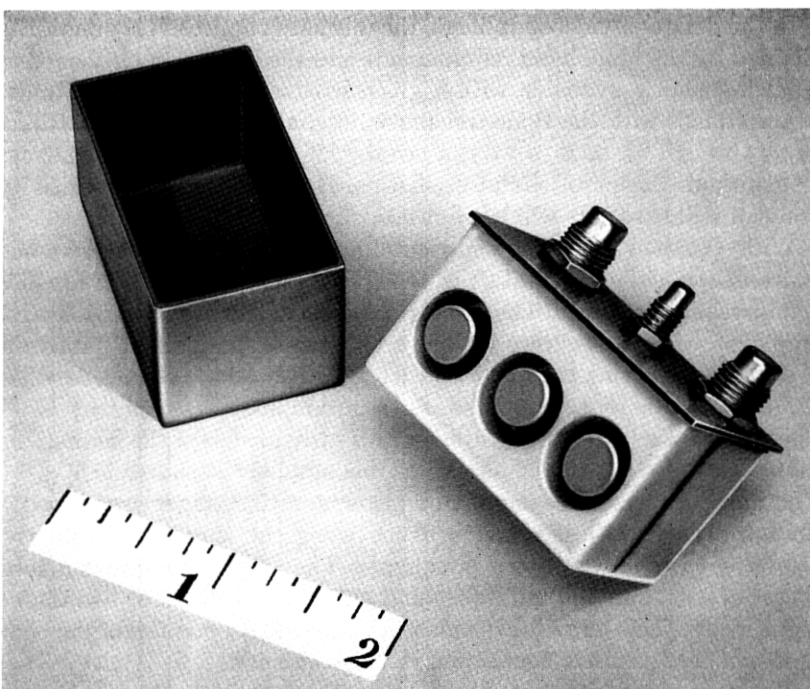

Fig. 14 — Three-stage common emitter video amplifier.

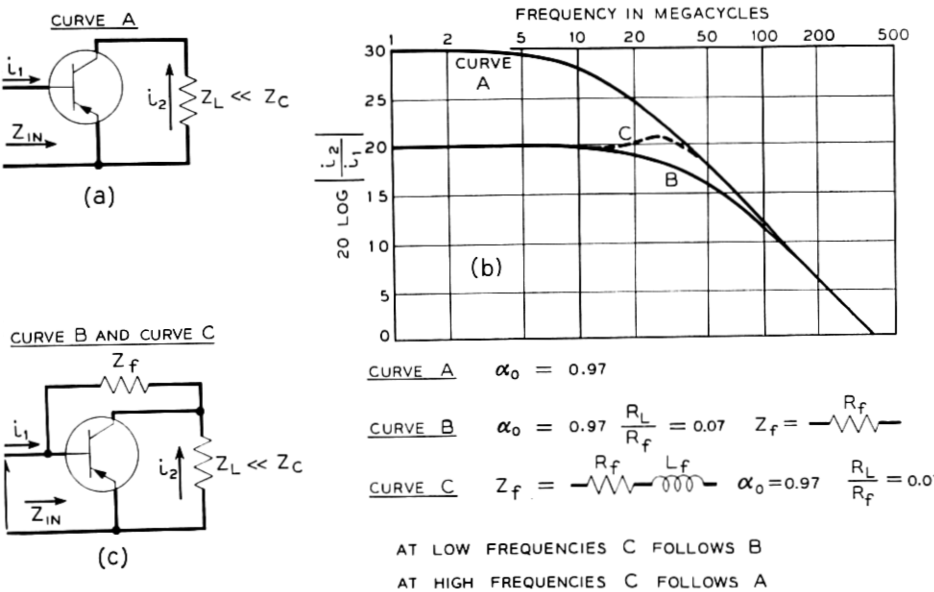

Fig. 15 — Single-stage common emitter amplifier with and without shunt feedback: (a) amplifier without feedback; (b) current gain of amplifier; (c) amplifier with shunt feedback.

of 14 mc — the frequency at which the common emitter current gain is 3 db below its low frequency value of $\alpha_0/(1 - \alpha_0)$. If a broader bandwidth is desired, this can be obtained by feeding back a portion of the output to the input through a feedback impedance, Z_f, connected between the collector and base, as shown in Fig. 15(c). The current gain of this transistor is given by

$$\frac{i_2}{i_1} \doteq \frac{\alpha}{1 - \alpha + \dfrac{Z_L}{Z_f}}. \tag{17}$$

If the simplified expression of (1) for a or α is placed in (17), the current gain as a function of frequency for Z_L/Z_f real is given by

$$\frac{i_2}{i_1} = \frac{\alpha_0}{(1 - \alpha_0)\left[1 + \dfrac{Z_L}{Z_f(1 - \alpha_0)}\right]\left\{1 + j\dfrac{f}{f_\alpha(1 - \alpha_0)\left[1 + \dfrac{\alpha_0 Z_L}{Z_f(1 - \alpha_0)}\right]}\right\}}, \tag{18}$$

so that, except for the ratio

$$\frac{1 + \dfrac{Z_L}{Z_f\,(1 - \alpha_0)}}{1 + \alpha_0 \dfrac{Z_L}{Z_f\,(1 - \alpha_0)}},$$

which is normally close to unity, the low frequency gain is decreased and the cutoff frequency is increased by the same amount, namely

$$1 + \frac{Z_L}{Z_f(1 - \alpha_0)}.$$

This is shown in curve B of Fig. 15(b), where the common emitter current gain is plotted for the transistor assumed for curve A, with a resistance, R_f, connected between its collector and base such that

$$R_L/R_f = 0.07.$$

Note that the low-frequency gain of curve B is down 10 db, or a factor of about one to three in magnitude, from that of curve A, and that the cutoff frequency has been increased by about the same factor. The asymptotic current gains of curves A and B at very high frequencies differ only slightly in magnitude, so they are shown identical in Fig. 15(b). By opening up the feedback path between the collector and base at high frequencies, curve B can be made to move into curve A before the asymptotic region is reached as is illustrated in the dotted curve, curve C. This can be accomplished most simply by making Z_f a resistance and inductance in series. In this way, approximately an extra octave of bandwidth can be obtained with no additional in-band gain sacrifice. However, there will be somewhat greater delay distortion when the amplifier is used for the amplification of narrow pulses than there would be if the cutoff were allowed to proceed in normal RC fashion, as in curve B of Fig. 15(b).

The simplicity of the above technique of trading gain for bandwidth is illustrated in the two-stage diffused-base common emitter video amplifier shown schematically in Fig. 16.* The transistors used in this amplifier have a normal common emitter short-circuit gain given by curve A of Fig. 15, and R_L/R_f is made 0.07 to make $(1 - \alpha_0) + R_L/R_f$ equal to 0.1 and give a low-frequency current gain of magnitude 10 or a

* This amplifier was devised by the author and presented at the June 1955 Semiconductor Device conference in Philadelphia, Pa., to demonstrate the broadband capabilities of the original research models of diffused base germanium transistors. For a description of these transistors see Lee.[9] For more complete information on this type of video amplifier see Ballentine and Blecher.[10]

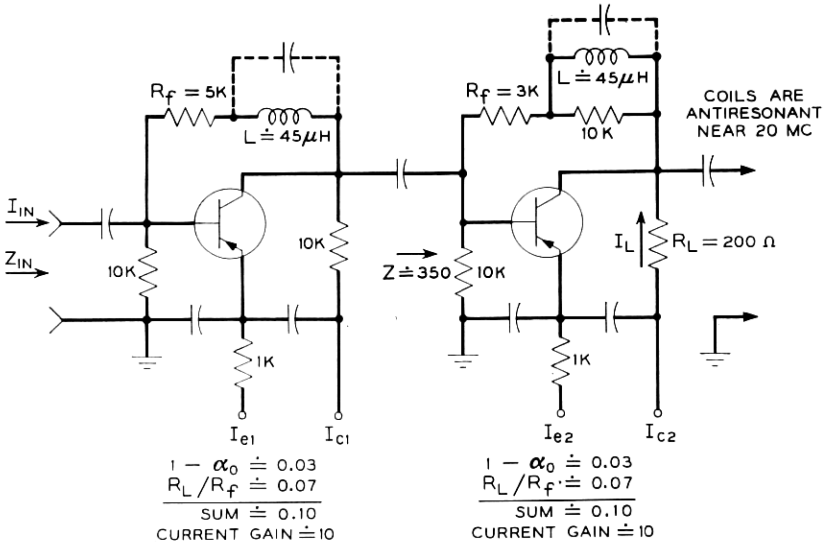

Fig. 16 — Two-stage common emitter amplifier with shunt feedback on each stage.

current gain of 20 db. The feedback path is opened at the high-frequency end of the band by the 45-microhenry coil in series with the feedback resistance of each stage. The dotted capacitances are the distributed capacitances of the coils, which produce a parallel resonance and essentially open-circuit impedance at the top end of the band. Therefore, the feedback path is effectively opened, and the normal common emitter current gain without feedback is obtained.

The current gain of the two-stage amplifier of Fig. 16 is plotted as a function of frequency in Fig. 17. The amplifier has a two-stage gain of 40 db flat to ±0.5 db up to 20 mc.

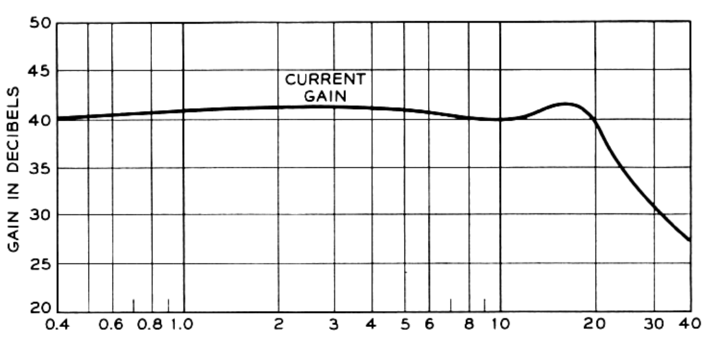

Fig. 17 — Current gain for amplifier of Fig. 16.

Although the technique described above is an easy way to trade gain for bandwidth, it is also an inefficient way. This is a result of the fact that, although a single stage of the amplifier of Fig. 16 behaves like a negative feedback amplifier, in that the forward gain is reduced as the feedback is increased, it is in fact still a positive feedback amplifier in accordance with our earlier analysis of transistor internal feedback. This is shown from Fig. 18, where the generalized T schematic of the transistor is given in the common emitter connection with a load resistance, R_L, and a feedback resistance, R_f. The various currents resulting from an injected emitter current, i_e, are also shown. Using Blackman's technique for determining feedback loop gain, $\mu\beta$, or the return ratio of the amplifier of Fig. 18, can be written by inspection as follows:

$$\hat{i}_e \doteq \alpha i_e - \alpha i_e \frac{R_L}{R_f},$$
$$\mu\beta = \frac{\hat{i}_e}{i_e} \doteq \alpha \left(1 - \frac{R_L}{R_f}\right), \tag{19}$$

or, for the amplifier of Fig. 16,

$$\mu\beta \doteq \alpha\,(1 - 0.07) = 0.93\,\alpha. \tag{20}$$

Equation (19) shows that, even though the magnitude of the feedback has been reduced by the factor $(1 - R_L/R_f)$, $\mu\beta$ is still positive and close to unity. In other words, even though the incremental feedback through the feedback resistance, R_f, is negative, the residual or net feedback is

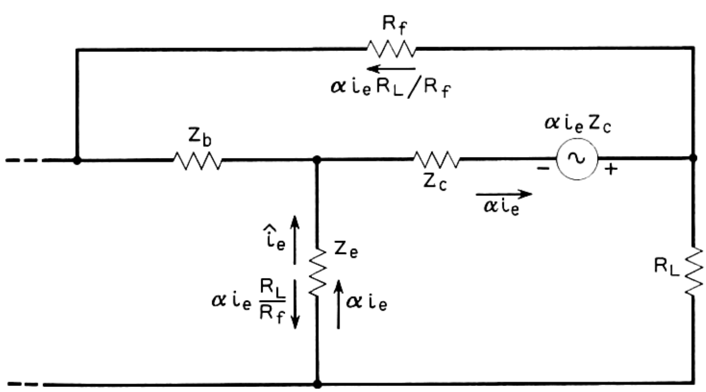

Fig. 18 — Equivalent circuit for common emitter amplifier with shunt feedback.

still positive. This can be practically verified by examining the variation in external gain with change in α.

The current gain of the circuit of Fig. 18 can be obtained from (17) and is given by:

$$\frac{i_2}{i_1} \doteq \frac{\alpha}{(1 - \alpha) + \frac{R_L}{R_f}}. \tag{21}$$

If we compute the variation in current gain expressed as a fraction of the initial current gain in terms of the variation in α expressed as a fraction of the original value of α we get,

$$\frac{d\frac{i_2}{i_1}}{\frac{i_2}{i_1}} = \frac{d_\alpha}{\alpha}\left[\frac{1}{1 - \alpha\left(1 - \frac{R_L}{R_f}\right)}\right], \tag{22}$$

which, from (19), gives

$$\frac{d\frac{i_2}{i_1}}{\frac{i_2}{i_1}} = \frac{d_\alpha}{\alpha}\left(\frac{1}{1 - \mu\beta}\right). \tag{23}$$

Since $\mu\beta$ from (20) is positive and only slightly less than unity, (23) shows that the terminal current gain changes much more rapidly than does the current generator gain, α, of the active device. This is the reverse of a negative feedback effect and is characteristic of the residual positive feedback which has been shown to exist. The variation in current gain due to a given change in α is less than it was before the R_f feedback path was added, which is in accordance with our statement that R_f represents a negative increment of feedback, but that the feedback loop gain residue is still positive.

In many instances, the decrease in external gain change for a given change in active-element gain obtained by the simple circuit of Fig. 16 is sufficient. However, where lower external gain change is required — and somewhat greater circuit complexity is therefore justified — gain can be more effectively traded for bandwidth by feedback around a minimum of two common emitter stages, as shown in the schematic of Fig. 19. The first stage only is shown in generalized equivalent T form, since it is here that the feedback path is mathematically broken to compute the main feedback gain (i.e. the feedback gain through the R_f feedback path).

Again using the Blackman technique, the feedback loop gain $\mu\beta$ can be obtained by inspection of Fig. 19 as follows:

$$\hat{i}_{e1} = \alpha_1 i_e \frac{Z_b}{Z_b + Z_e + R_{ef}} < 0.5\,\alpha_1 i_e, \tag{24}$$

$$\begin{aligned} \hat{i}_{e2} &\doteq \frac{\alpha_1\alpha_2}{1 - \alpha_2} \frac{R_L}{R_L + R_f} \frac{R_{ef}}{Z_b + Z_e + R_{ef}} i_e \\ &\doteq \frac{\alpha_1\alpha_2}{1 - \alpha_2} \frac{R_L}{R_L + R_f} i_e, \end{aligned} \tag{25}$$

$$\hat{i}_e = \hat{i}_{e1} - \hat{i}_{e2} \doteq -\frac{\alpha_1\alpha_2}{1 - \alpha_2} \frac{R_L}{R_L + R_f} i_e, \tag{26}$$

$$\mu\beta = \frac{\hat{i}_e}{i_e} \doteq -\frac{\alpha_1\alpha_2}{1 - \alpha_2} \frac{R_L}{R_L + R_f}. \tag{27}$$

If $R_L \geqq R_f$ and α_1 and α_2 are close to unity, then $\mu\beta \gg 1$. When the loop gain is much larger than unity, the feedback voltage, V_f, will be approximately equal to the applied generator voltage, V_g, when a steady state signal is applied to the input. Therefore, since

$$V_f = \frac{R_{ef}}{R_f} V_{\text{out}} \qquad \text{and} \qquad V_f \doteq V_g, \tag{28}$$

$$\text{amplifier voltage gain} = \frac{V_{\text{out}}}{V_g} \doteq \frac{R_f}{R_{ef}}, \tag{29}$$

and the voltage gain of the amplifier is substantially independent of change in gain of the active elements — in this case, the transistors. This

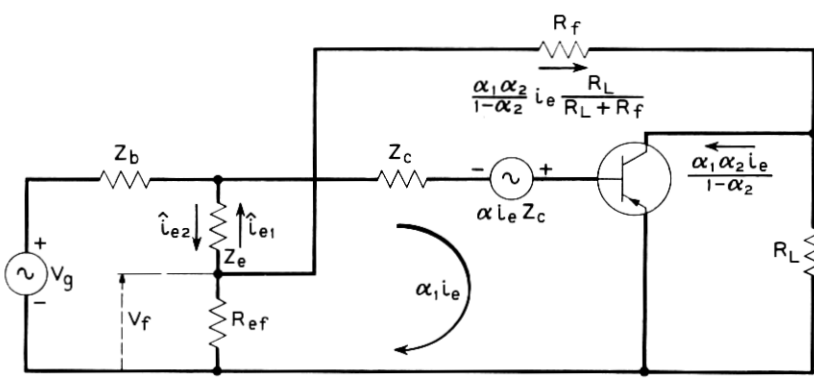

Fig. 19 — Equivalent circuit for two-stage common emitter amplifier with shunt-to-series feedback.

is the anticipated result for a true negative feedback amplifier with feedback loop gain much greater than unity. It is important to notice that the approximation of (29) to the feedback loop gain is good only for $R_L \geqq R_f$, so that this circuit is essentially a voltage feedback amplifier. Since the circuit of Fig. 19 contains basically only two 6-db-per-octave asymptotic cutoffs, it is an intrinsically stable circuit requiring only simple if any feedback loop equalization. The amplifier of Fig. 19 is a voltage amplifier with a high input impedance and a low output impedance, since it has series feedback at the input and shunt feedback at the output. It can be made a current amplifier with low input impedance and high output impedance by feeding back from a resistance in the emitter circuit of the second transistor through a feedback resistance R_f to the base of the first transistor. The approximate design formulae for this configuration can be obtained in the same manner as were those for the voltage amplifier shown schematically in Fig. 19.

If high linearity as well as high stability, or if unusually high stability is required in an amplifier, either of the broadband video or relatively narrowband linear type, then the two-stage amplifier of Fig. 19 is still inefficient from the standpoint of trading gain for bandwidth. In this case, the most efficient circuit is a three-common-emitter-stage single-loop feedback amplifier.[11,12] This, of course, involves the complexity of interstage and feedback network design inherent to the stabilization of a three-stage negative feedback amplifier. This is a consequence of the potential instability associated with the minimum asymptotic cutoff of 18 db per octave associated with three active stages.

In conclusion, it may be stated that the requirements of a large percentage of the radio frequency and video or baseband transistor amplifiers can be met by the circuits of Figs. 9, 16 and 19. These circuits demonstrate the simplicity with which basically sound and stable transistor amplifiers can be built, providing that the basic nature of the internal feedback of the transistor is understood, and the fatal mistake of attempting to obtain so called "maximum available gain" is not made.

Additional material which may be of interest to designers of RF and video amplifiers: neutralization — Cheng;[13] stability — Stern;[14] video amplifiers — Brunn;[15] alignable receivers — Gibbons.[16]

VIII. ACKNOWLEDGMENT

The author wishes to express his appreciation to L. G. Schimpf and C. E. Paul for making available information on the circuits and performance of their amplifiers for use as examples of the various types of amplifiers discussed in this paper. He also wishes to thank R. B. Blackman for checking the positive feedback theory of single-stage transistor amplification and for his suggestion of the simplified technique for mathematically breaking the feedback loop to compute return ratio.

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