

XI. TRANSISTOR AND DIODE STUDIES*

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A. DESIGN OF TRANSFORMERLESS TRANSISTOR AUDIO AMPLIFIERS

Considerable effort by many organizations has been put into the development of high-fidelity, high-efficiency audio amplifiers that use power transistors. These amplifiers may be divided into two classes: those that use transformers, and those that employ complementary symmetry. Relatively little, if any, feedback is employed in the transformer-coupled amplifiers. This is due, in part, to the limitations that the transformers place on the frequency response. The use of complementary symmetry eliminates this limitation but, unfortunately, creates two new limitations. First, it is difficult to obtain commercially available complementary power transistors that are reasonably matched for gain of high-current levels. Second, a serious problem of bias stability exists. We have recently studied the design of single-ended, push-pull, transformerless output stages that use the same type of transistor and have built amplifiers that incorporate as much as 36-db over-all negative feedback. This design technique eliminates the problem of matched gain at high-current levels, but does not completely eliminate the stability problem.

Before describing this output stage, a few general points of interest in transformerless amplifier design will be stated. It is commonly known that the grounded base connection is not useful in transformerless design because the resulting power gain is less than unity. The advantages of the grounded collector over the grounded emitter from the standpoint of distortion are also well known (1). The advantage of the grounded collector from the standpoint of maximum phase shift at high frequencies is not as well known. It will be discussed here. An equivalent circuit for a grounded-emitter stage is shown in Fig. XI-1. With the assumption that the collector impedance z_c is infinite, the voltage gain is

$$\frac{e_o}{e_{in}} = \frac{a R_L}{(1-a)(R_g + r_b) + r_e} \quad (1)$$

Middlebrook's (2) approximation for the frequency dependence of the current gain a is

$$a = \frac{a_o(1 - m Ts)}{1 + Ts} \quad (2)$$

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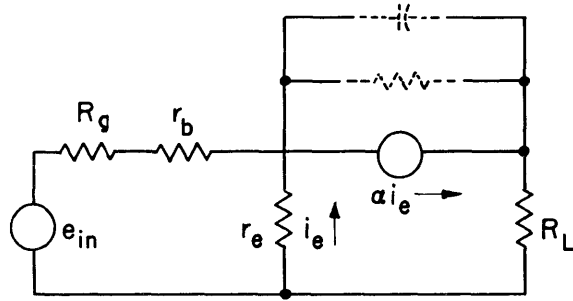


Fig. XI-1. Equivalent circuit of a grounded-emitter amplifier.

where $m = 0.204$, $T = 1.04/2\pi f_c$, and f_c is the frequency at which the current gain is 3 db below its low-frequency value. This approximation is very accurate for frequencies as high as f_c , and even reasonably good for frequencies as high as $4f_c$. Thus

$$\frac{e_o}{e_{in}} = \frac{R_L \alpha_o [1 - m Ts]}{R'_g(1 - \alpha_o) + r_e + Ts[R_g(1 + \alpha_o m) + r_e]} \quad (3)$$

where $R'_g = R_g + r_b$. The significant point in Eq. 3 is that the maximum phase shift is 180° . Introducing negative feedback by the simple means of increasing the ratio r_e/R'_g does not affect the maximum phase shift.

The equivalent circuit of a grounded-collector stage is shown in Fig. XI-2. The voltage gain of this stage is

$$\frac{e_o}{e_{in}} = \frac{R_L}{R_L + r_e + R'_g(1 - \alpha)} \quad (4)$$

Substituting for the frequency dependence of α , and letting $R'_L = R_L + r_e$, we obtain

$$\frac{e_o}{e_{in}} = \frac{R_L}{R'_L} \frac{1 + Ts}{1 + \frac{R_g(1 - \alpha_o)}{R'_L} + Ts \left[1 + \frac{R'_g}{R'_L} (1 + \alpha_o m) \right]} \quad (5)$$

$$\approx \frac{1 + Ts}{1 + Ts \left[1 + \frac{R'_g}{R'_L} (1 + m) \right]} \quad (6)$$

Thus, the voltage gain of a grounded-collector connection has the interesting property of a lead-lag network. The maximum phase shift is 90° or less. The frequency at

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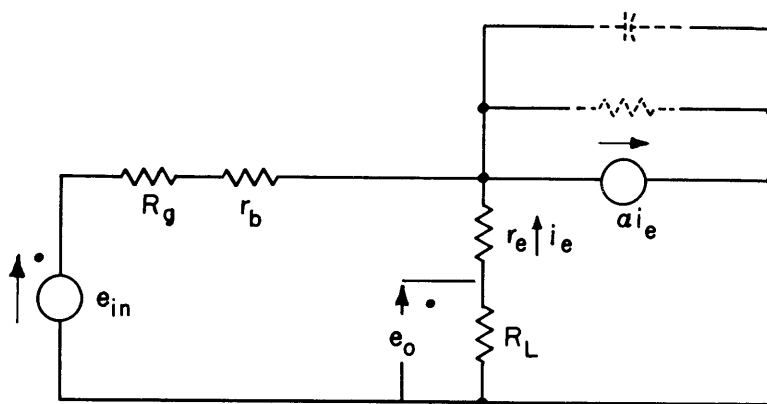


Fig. XI-2. Equivalent circuit of a grounded-collector amplifier.

which the maximum phase shift occurs is given by

$$f_{\theta_{\max}} \approx \frac{f_c}{\left[\frac{R'_g}{R'_L} (1+m) \right]^{1/2}} \quad (7)$$

The value of the maximum phase shift is

$$\theta_{\max} \approx \tan^{-1} \frac{\frac{R'_g}{R'_L} (1+m) - 1}{2 \left[\frac{R'_g}{R'_L} (1+m) \right]^{1/2}} \approx \tan^{-1} \frac{1.2 \frac{R'_g}{R'_L} - 1}{2.2 \left(\frac{R'_g}{R'_L} \right)^{1/2}} \quad (8)$$

As the ratio R'_g/R'_L approaches unity (the approximate condition for an available power gain of unity) the maximum phase shift approaches a value of a few degrees. For practical values of R'_g/R'_L between 5 and 10, the maximum phase shift varies from 45° to 58° .

The circuit diagram of a single-ended push-pull output stage together with a complementary driver stage is shown in Fig. XI-3. Class B operation is used in both the output and driver stages. For negative-going waveforms at the input, both T_1 and T_2 conduct; for positive-going waveforms, T_3 and T_4 conduct. It should be noted that T_3 and T_4 are both grounded-emitter connections. However, we shall show that the method of connection that includes a high degree of negative feedback causes the gain and the frequency response of two grounded-emitter stages to closely approximate those of the

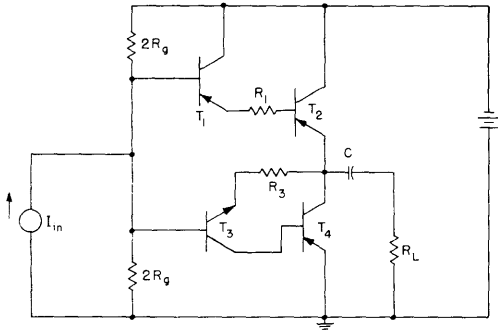


Fig. XI-3. Circuit diagram of a single-ended push-pull output stage and complementary driver stage.

grounded-collector stages. The use of complementary transistors in the driver stage provides the necessary phase inversion. Capacitance coupling to the load resistor allows the use of a single power source and slightly improves the bias stability of the circuit.

The mid-frequency gain of the two grounded-collector stages, if we assume that the collector impedances are infinite, is

$$\frac{e_o}{i_{in} R_g} = \frac{R_L}{R_L + r_{e_2} + \left(r_{b_2} + r_{e_1} + R_1 \right) (1 - a_1) + \left(R_g + r_{b_1} \right) (1 - a_1) (1 - a_2)} \quad (9)$$

$$\approx \frac{R_L}{R'_L} \frac{1}{1 + \frac{R'_g}{R'_L} (1 - a_1) (1 - a_2)} \quad (10)$$

If we include the frequency dependence of a_1 and a_2 , the high frequency gain becomes

$$\frac{e_o}{e_{in}} \approx \frac{(1 + T_1 s) (1 + T_2 s)}{1 + (T_1 + T_2) \left[1 + \frac{R'_g}{R'_L} (1 - a_o) (1+m) \right] s + T_1 T_2 \left[1 + \frac{R'_g}{R'_L} (1+m)^2 \right] s^2} \quad (11)$$

where $a_o = (a_1)_{s=0} = (a_2)_{s=0}$. If $T_2 > T_1$, then

$$\frac{e_o}{e_{in}} \approx \frac{(1 + T_1 s) (1 + T_2 s)}{1 + T_2 s \left[1 + 1.2 \frac{R'_g}{R'_L} (1-a) \right] + T_1 T_2 s^2 \left[1 + 1.4 \frac{R'_g}{R'_L} \right]} \quad (12)$$

The denominator has two poles. The location of these poles is a function of the ratio R'_g/R'_L . When this ratio is small the poles are located on the negative real axis; as the

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ratio increases the poles move toward each other; and after coincidence become complex.

The mid-frequency gain of the two grounded-emitter stages is

$$\frac{e_o}{e_{in}} = \frac{1}{1 + \frac{[R_g + r_{b3}(1 - a_3) + r_{e3} + R_3](1 - a_4)}{R_L[1 - a_3 + a_3 a_4]}} \quad (13)$$

If the magnitudes of a_3 and a_4 are equal, $T_4 > T_3$, and r_{e3} and R_3 are negligible. Then

$$\frac{e_o}{e_{in}} \approx \frac{1 + T_4 s + T_3 T_4 (1 + m^2) s^2}{1 + T_4 \left[1 - m + \frac{R'_g}{R_L} (1 - a) (1 + m) \right] s + T_3 T_4 \left[(1 + m)^2 \frac{R'_g}{R_L} + 1 + m^2 \right] s^2} \quad (14)$$

If T_4 is greater than T_3 , we have

$$\frac{e_o}{e_{in}} \approx \frac{(1 + T_4 s)(1 + T_3 s)}{1 + T_4 \left[0.8 + 1.2 \frac{R'_g}{R_L} (1 - a) \right] s + T_3 T_4 \left[1 + 1.4 \frac{R'_g}{R_L} \right] s^2} \quad (15)$$

Comparing Eqs. 12 and 13, we see that the high-frequency behaviors are quite similar. A comparison of Eqs. 10 and 13 for the condition that the current gain a is close to unity indicates that the mid-frequency gains are both very close to unity. It can also be shown that the input and output impedances of the two different circuits are approximately equal. Thus the low-frequency gains will also be equal.

The bias stability of the circuit shown in Fig. XI-3 has not been studied in great detail. Calculations show that the quiescent current has a term that is dependent on saturation currents of the form

$$\text{quiescent current term} \approx \frac{I_{co2} + I_{co4}}{1 - a'}$$

where a' is the approximate current gain of the output stage. This factor should indicate a relatively high sensitivity of quiescent current to the temperature of the collector junctions of the power transistors. However, in actual operation, the quiescent current, after long periods of operation at power levels as high as 5 watts, remains reasonably constant. This may well result from the inherent thermal stability of the power transistors rather than from properly stabilized circuitry.

With the use of commercially available transistors (such as Philco 1041 power transistors and General Electric 2N136 and 2N169 drivers), the maximum phase shift of the voltage gain of the output and driver stages can be of the order of 100° . Employing

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these design principles, we have constructed and tested 4-stage, 5-watt audio amplifiers which have proved stable, with 36-db over-all negative feedback from the speaker to the input stage. Interesting characteristics of such amplifiers are: (a) output impedance over the audio range of the order of 0.01 ohms; (b) frequency response that is flat within 1 db from 5 to 50,000 cps; and (c) square-wave response at 20 kc with less than 5 per cent overshoot.

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References

1. R. P. Crow and R. D. Mohler, Design of high fidelity 10 watt transistor amplifier, IRE Convention Record, Part 7, 1956, pp. 142-150.
2. R. D. Middlebrook, An Introduction to Junction Transistor Theory (John Wiley and Sons, Inc., New York, 1957), pp. 192-209.

B. DIELECTRIC MODULATORS

Temperature tests of the zero-signal balance of the RC bridge modulator (see the Quarterly Progress Report of July 15, 1957) that employs junction diodes have indicated nonretraceable characteristics. Since we now attribute this factor to the presence of water moisture rather than to a permanent change in the parameter values, we have constructed a hermetically sealed box containing the majority of the bridge components. The temperature tests will be repeated in the future.

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