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DESIGN CONSIDERATIONS FOR DIRECT-COUPLED TRANSISTOR AMPLIFIERS

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DESIGN CONSIDERATIONS FOR DIRECT-COUPLED TRANSISTOR AMPLIFIERS

Approved

Stuart Wm. Seeley

An outstanding source of difficulty in d-c amplifier design is drift. With drift as with noise the use of degenerative feedback does not improve the ratio of the signal to the undesired disturbance. By analogy to an approach used with the noise problem the procedure to be described is to seek a relationship between the driving source and the amplifier which will yield optimum performance with respect to signal-to-drift ratio.

It is shown that:

1. The maximum obtainable signal-to-drift ratio of an amplifier is achieved by choosing the optimum source resistance.
2. The signal-to-drift capability of a transistor and the optimum source resistance nearly the same for the three configurations; common emitter, common collector and common base.
3. Negative feedback can change the optimum source resistance and generally degrades the signal-to-drift ratio.
4. Improved signal-to-drift ratio can be obtained at reduced emitter currents.

Introduction

Probably the outstanding source of difficulty in d-c amplifier design is drift. With transistors the changes caused by temperature variation are particularly troublesome in this respect[1]. On this account much effort has been expended to realize transistor circuits which achieve a stable operating point over a wide range of temperature. In many cases the use of d-c feedback accomplishes this result very successfully and may appear to offer a remedy for drift in the d-c amplifier. However, while degenerative feedback does indeed reduce drift, the result is deceptive. The desired signal is also reduced.

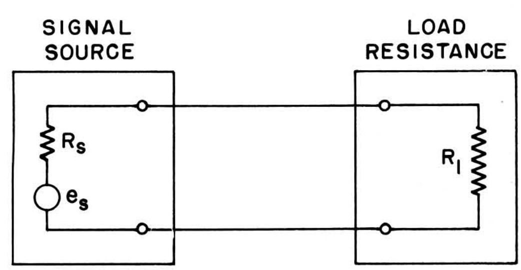
This aspect of the problem of drift is much like that of noise in a-c amplifiers. In both cases the signal and the undesired disturbance are inseparable and feedback does not distinguish the two. By analogy to the approach used with the noise problem[2,3,4] the procedure described here is to seek a relationship between the driving source and the amplifier which will lead to optimum performance with respect to signal-to-drift ratio.

That such an approach might prove rewarding can be inferred by considering the chief sources of drift in the transistor. These are the changes in saturation current and d-c input conductance due to temperature variation. If the change of saturation current, I_{CO}, were the only source of drift it would be desirable to drive a d-c transistor amplifier from a constant-voltage source. Increasing saturation current would then have no effect at the input and drift would be minimized. If, on the other hand, only changes in the input conductance were associated with drift it would be desirable to drive the amplifier from a constant-current source.

These ideas are now set forth in detail. Proceeding from fundamental definitions the concepts of optimum source resistance and obtainable signal-to-drift ratio for a direct-coupled amplifier are developed. These principles are then applied to certain aspects of the problem of drift of transistors operated over an extended temperature range.

Available Signal Power[5]

The signal source of Fig. 1 has an internal resistance, R_S, and supplies an electromotive force of e_S volts. It delivers $e_S^2 R_1/(R_1 + R_S)^2$ watts to a load resistance of R_S



$$\text{Output power} = \frac{e_s^2 R_l}{(R_l + R_s)^2}$$

$$\text{Available power} = \frac{e_s^2}{4 R_s}$$

Fig. 1 – Available signal power.

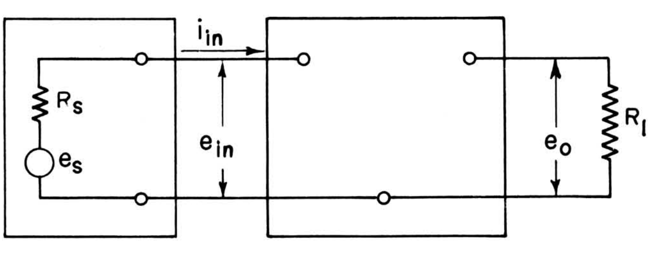
ohms. This power is maximum and equal to $e_s^2/4R_s$ when the load resistance is matched to the source resistance; i.e., when $R_l = R_s$. The quantity $e_s^2/4R_s$ is called the available signal power of the source, since it represents the maximum power the source is capable of delivering.

This available signal power is to be regarded as a defined quantity which provides another useful means of specifying the source. As such the available signal power is characteristic of the source alone independent of the external circuit. This is the case even though available signal power is defined by reference to a matched load.

Gain[5,6]

The amplifier of Fig. 2 delivers output signal to a load resistance, R_l. A signal generator of source resistance, R_s, drives the amplifier. The power gain of the amplifier is the ratio of the signal power delivered to the load to the signal power supplied to the input terminals; e.g., $G_p = P_{out}/P_{in}$. As such, power gain is a property of the amplifier and the load alone in that it can be specified apart from the source. However, to determine either the power output of the amplifier or the power input to the amplifier, it is necessary to specify in addition both the source parameters, e_s and R_s, and the input resistance of the amplifier, R_i. The formula for the power input to the amplifier is

$$P_{in} = \frac{e_s^2 R_i}{(R_s + R_i)^2}$$



$$G_p = \frac{P_{out}}{P_{in}} = \frac{e_o^2/R_l}{e_{in}\, i_{in}}$$

$$G_T = \frac{P_{out}}{P_{av}} = \frac{e_o^2 \;/ R_l}{e_s^2/4R_s}$$

Fig. 2 – Amplifier gain.

The expression for the power output, then, is

$$P_{out} = G_p P_{in} = G_p \left[\frac{e_s^2 R_i}{(R_s + R_i)} \right]$$

The gain of the circuit of Fig. 2 called transducer gain is defined as the ratio of the power delivered to the load, e_o^2/R_l, to the power available from the signal source, $e_s^2/4R_s$; that is, $G_T = P_{out}/P_{av}$. It is significant that this available power is not, generally, the power being delivered to the input terminals of the amplifier. Instead this definition of gain can be regarded as the ratio of the power actually obtained using the amplifier to the power which could be obtained from the source alone. Such a ratio is meaningful to the designer as a measure of what the use of the amplifier accomplishes.

Transducer gain, G_T, is a property of the amplifier and the source and as such must be specified for a given source. The relationships involved are

$$G_p = \frac{P_{out}}{P_{in}} \qquad P_{in} = \frac{e_s^2 R_i}{(R_s + R_i)^2}$$

$$G_T = \frac{P_{out}}{P_{av}} \qquad P_{av} = \frac{e_s^2}{4R_s}$$

from which

$$G_T = G_p \frac{P_{in}}{P_{av}} = G_p \left[\frac{4R_s R_i}{(R_s + R_i)^2} \right]$$

Thus, if transducer gain is specified, only the available source power is needed to determine the power output. This permits a comparison of amplifiers driven by the same source.

The utility of the definition of transducer gain is illustrated by the following example. In Fig. 3 amplifier A has an input resistance of one megohm. The power gain (not the transducer gain) of this amplifier is 10^5 or 50 db. Amplifier A is driven by a source with internal resistance of one kilohm. If the open-circuit source voltage is one volt the power delivered to the input terminals of the amplifier is calculated to be:

$$P_{in} = \left[\frac{e_s}{R_l + R_s} \right]^2 R_l = \frac{1000 \times 10^3}{(1001 \times 10^3)^2} \approx 10^{-6} \text{ watt}$$

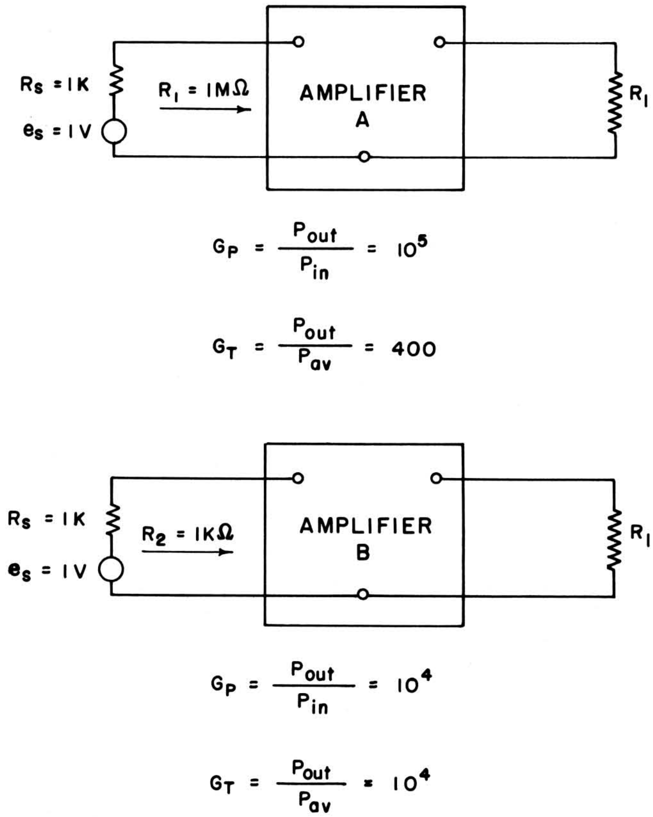


Fig. 3 – Comparison of amplifiers.

The power output is then 1/10 watt.

Amplifier B, on the other hand, has an input resistance of one kilohm. The power gain, G_p, of this amplifier is 10^4 or 40 db. If amplifier B is driven by the same source as amplifier A, the power delivered to the input terminals is:

$$P_{in} = \left[\frac{e_S}{R_2 + R_S}\right]^2 R_2 = \frac{1000}{(2000)^2} = 2.5 \times 10^{-4} \text{ watt}$$

The resulting power output is 2.5 watts. Thus, for the specified source the signal power delivered by Amplifier B exceeds that of A even though its power gain is lower.

Specifying available source power as input power takes into account the relationship between the amplifier and the driving source. Defined in this way the transducer gain of amplifier A driven by the given source is 400, or 26 db; the transducer gain of amplifier B driven by the given source is 10^4, or 40 db. Thus the definition of transducer gain provides a basis for comparison of amplifiers driven by the same source.

The transducer gain of the amplifier of Fig. 2 depends upon the load resistance and upon the source resistance. If power gain is the sole consideration it is desirable to match the source resistance to the input of the amplifier. This is the condition required to utilize fully the available power of which the source is capable. For this matched condition the power input, defined as available power from the source, is actually driving the amplifier and power gain and transducer gain are the same.

Likewise for constant input the amplifier delivers maximum power to a matched load resistance. The ratio of this power available to the load to power available from the generator is called available gain. If both the source and the load are matched to the amplifier the maximum available gain is obtained. Under these conditions of matched input and matched output power gain, transducer gain, available gain and maximum available gain are identical.

A given signal source of e_S volts with an internal resistance, R_S can be matched to the input of an amplifier by means of an ideal transformer of turns ratio n. The combination is equivalent to a signal source of ne_S volts with an internal resistance, n^2R_S, where n is the turns ratio of the transformer. The available power of the source is unchanged by the transformer, and is delivered to the amplifier by setting the equivalent resistance n^2R_S, equal to the input resistance of the amplifier. This is also the condition for obtaining at the input terminals of the amplifier the maximum current and the maximum voltage from the given source. Thus, whether drive for the amplifier is considered to be current, voltage or power the condition for optimum drive from the given source is the same.

Drift

For a direct-coupled amplifier both gain and the problem of drift are important considerations. Drift appears as an error which is not distinguishable from the desired signal in the output of the direct-coupled amplifier. The drift magnitude alone does not adequately specify the significance of drift in an amplifier. A given drift magnitude in a high-gain amplifier may represent less relative error than a much lower value where the gain is low. The signal-to-drift ratio is used to express the relative error represented by drift.

In order to specify the significance of drift as a relative error it is convenient to refer the drift to the input of the amplifier. For the amplifier and signal source of Fig. 2 this is accomplished by determining that input power (available power from the source) which is sufficient to correct the drift. This gives the error due to drift in terms of equivalent input power and independently of the amplifier gain.

Now by definition $S = G\, S_{av}$ where S is the signal in the output, S_{av} is input signal power from the source and G is the gain of the amplifier.

Likewise, $D = G\,D_{av}$ where D is the magnitude of drift in the output and D_{av} is the equivalent input power sufficient to correct for drift. It is thus possible to express the ratio of signal to drift in terms of values at the input of the amplifier, independent of the gain since:

$$\frac{S}{D} = \frac{S_{av}}{D_{av}}$$

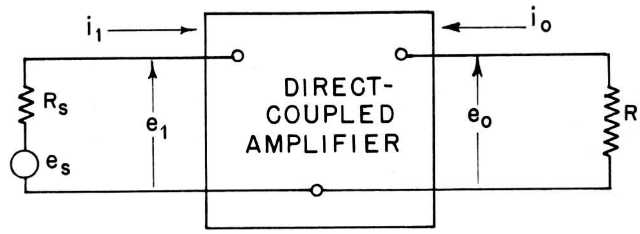
This value of the equivalent input power sufficient to correct for drift is an important performance characteristic of a direct-coupled amplifier with its driving source. This value alone is a criterion of the performance obtained. It can be used as a basis for comparison among amplifiers. It follows, therefore, that a prime objective of direct-coupled amplifier design is to minimize the available power from the source needed to correct for drift.

Optimum Source Resistance And Obtainable Signal-To-Drift Ratio

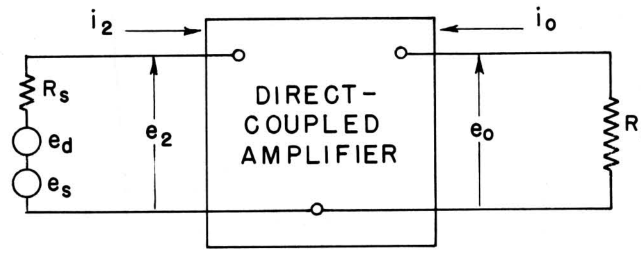
The problem of drift in direct-coupled amplifiers has been attacked in a number of ways. The methods range from chopper techniques used with complex high gain amplifiers[7] to carefully designed compensation schemes for transistors operated over an extended temperature range.[8,9] In each case it may be possible to realize improved performance up to a maximum obtainable signal-to-drift ratio for a given amplifier. It is possible to define this capability in terms of measurable values and to prescribe precisely the conditions for achieving the limit. The relationships involved are developed below.

Matching, which has been discussed in connection with available power and maximum available gain, is the design procedure for achieving maximum power transfer. The idea of matching for best noise figure is also a well established concept.[4] It turns out that the same sort of approach yields a useful result for the signal-to-drift ratio of a direct-coupled amplifier. For a given amplifier there is an optimum source resistance for which the equivalent input power sufficient to correct for drift is a minimum. With this optimum source resistance the obtainable signal-to-drift ratio for the amplifier is achieved.

An expression for the optimum source resistance can be developed from Fig. 4. Fig. 4a represents a driftless condition of the amplifier and source. Fig. 4b represents a condition of the amplifier wherein the source contributes input drive to correct the drift.



DRIFTLESS AMPLIFIER AND SOURCE



AMPLIFIER WITH THE SOURCE PROVIDING DRIFT CORRECTION

Fig. 4 – Drift referred to the input of an amplifier.

From Fig. 4:

$$e_S = i_1 R_S + e_1 \quad and \quad e_d + e_S = i_2 R_S + e_2$$

where:

R_S = source resistance

e_S = source signal voltage

e_d = drift correction voltage

i_1, e_1 = input current and voltage for the driftless condition of the amplifier

i_2, e_2 = input current and voltage with drift in the output corrected.

then:

$$e_d = (i_2 - i_1) R_S + (e_2 - e_1)$$

Now it is desired to select a value of source resistance, R_{SO}, which yields a minimum for the available power, $D_{av} = (e_d)^2/4R_S$, sufficient to correct for drift. It follows that:

$$D_{av} = \frac{(i_2 - i_1)^2 R_S^2 + 2(i_2 - i_1) R_S (e_2 - e_1) + (e_2 - e_1)^2}{4R_S}$$

and that the minimum is obtained for

$$\frac{\partial D_{av}}{\partial R_S} = 0 = \frac{(i_2 - i_1)^2}{4} - \frac{(e_2 - e_1)^2}{4R_S^2}$$

This gives for the optimum source resistance, R_{SO}, simply

$$R_{SO} = \frac{e_2 - e_1}{i_2 - i_1}$$

where $(e_2 - e_1)$ and $(i_2 - i_1)$, respectively, represent that change of voltage and current at the input terminals of the amplifier which corrects for drift. The corresponding value of the equivalent input power to correct for drift is:

$$D_{av} = (e_2 - e_1)(i_2 - i_1)$$

These simple relations specify both the obtainable signal-to-drift ratio of a given amplifier and the condition for achieving the result.

Considerations In The Design Of Direct-Coupled Transistor Amplifiers

General Design Considerations

These considerations have been applied to the problem of the drift of transistor direct-coupled amplifiers due to temperature change. The measurements are described in the next section of this bulletin. The results are summarized below and then treated in detail individually. These are:

1. The optimum source resistance and the obtainable signal-to-drift ratio as measured by the power needed to correct for drift are closely the same for the three configurations; common emitter, common base and common collector.

2. Feedback does not improve the signal-to-drift ratio but generally increases somewhat the available power sufficient to correct the drift. Feedback may increase or decrease the optimum source resistance.

3. Silicon transistors of otherwise comparable performance capabilities, as contrasted to germanium, can be expected to provide 10 db or more improvement of obtainable signal-to-drift ratio for the same temperature range at elevated temperatures. The optimum source resistance for silicon transistors is an order of magnitude greater than for germanium transistors operated comparably.

4. The obtainable signal-to-drift ratio can be improved by operating at lower emitter currents. The results realized at I_e = 0.1 ma as contrasted to those at I_e = 1.0 ma represent 3 to 5 db improvement.

5. Compensation is useful in improving the signal-to-drift ratio. If applied at the output the optimum source resistance is unchanged. By using compensation at the input it is possible to alter the optimum source so as to improve the match for obtainable signal-to-drift ratio.

The Three Configurations

The optimum source resistance and the obtainable signal-to-drift ratio are closely the same for the three configurations; common emitter, common base and common collector. Experimental results obtained for an RCA 2N79 transistor and for an RCA 2N139 transistor are shown in Fig. 5. In each case values for the available power sufficient to correct for drift are plotted against values of the source resistance used. In each case optimum values for the three configurations are nearly the same.

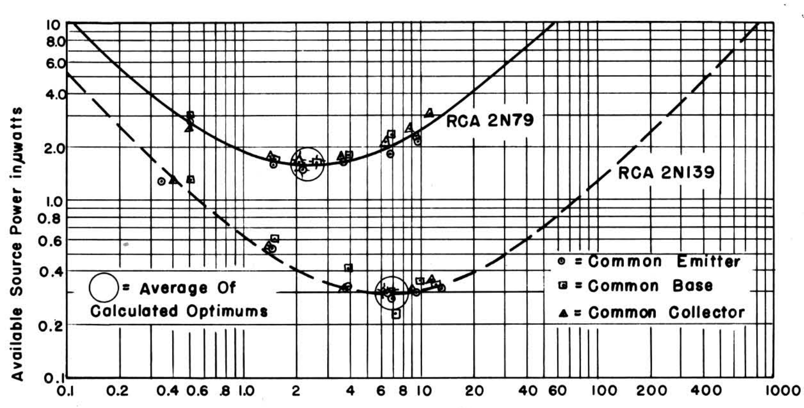
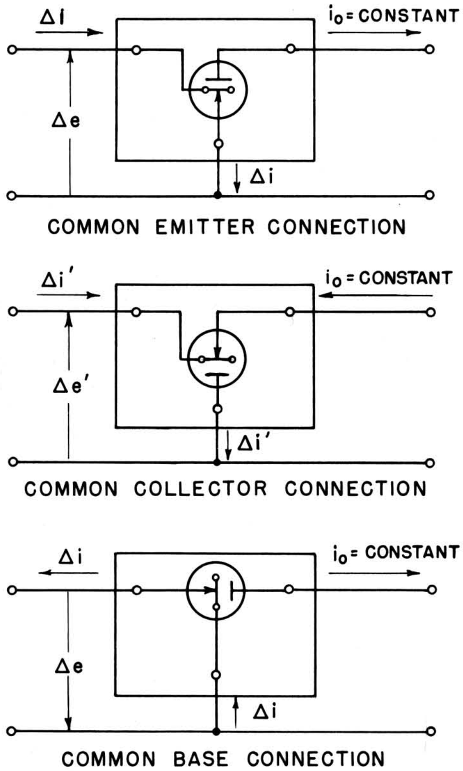


Fig. 5 – Available power to correct for drift due to temperature change plotted as a function of source resistance for the three configurations; common emitter, common base, common collector. The temperature range is 30° C to 50° C.

That this result is the valid one is demonstrated by Fig. 6. Comparison of the common base and common emitter configurations shows that the same output terminal is involved and that the input terminals are simply reversed. If the output currents are held constant and identical for the two cases and if the device remains unchanged for measurements at a given temperature then the changes at the input terminals must be the same.



$\left.\begin{matrix}\Delta e\\ \Delta i\end{matrix}\right\}$ = INPUT CHANGES TO CORRECT FOR DRIFT

$$R_{SO} = \frac{\Delta e}{\Delta i}$$

$$D_{AV} = \Delta e \cdot \Delta i$$

Fig. 6 – The optimum source resistance and obtainable signal-to-drift ratio are closely the same for the three connections.

Slightly less current and voltage changes are required to hold emitter current constant, so the obtainable signal-to-drift ratio is slightly greater for the common collector connection.

Feedback

Negative feedback does not improve the signal-to-drift ratio for uncompensated transistors but rather increases somewhat the available power sufficient to correct for drift. Results obtained using emitter feedback are shown with those for the common emitter connection in Fig. 7 for an RCA 2N79 transistor. Both the optimum source resistance and the power to correct for drift have increased.

Results obtained using collector feedback are shown with those for the common emitter connection in Fig. 8 for an RCA 2N139 transistor. In this case the power to correct for drift has increased but the optimum source resistance has decreased.

That these are results to be expected can be seen from Fig. 9. In the case of emitter feedback the correction current flowing in the feedback resistor results in a voltage drop which appears in the input loop. This increases the input power required and the optimum source resistance. In a like manner for collector feedback the input correction voltage results in a current flow in the feedback resistor which drains additional current and power from the source. In this case the optimum source resistance decreases.

This increase of power to correct for drift is a general result of feedback. Fig. 10 is a basic circuit of four fundamental methods of feedback applied simultaneously. In Fig. 10 the resistor R_1 provides current feedback proportional to the output voltage; R_2 provides current feedback proportional to the output current; R_3 provides voltage feedback proportional to the output current; and R_4 provides voltage feedback proportional to the output voltage. In each case the added elements

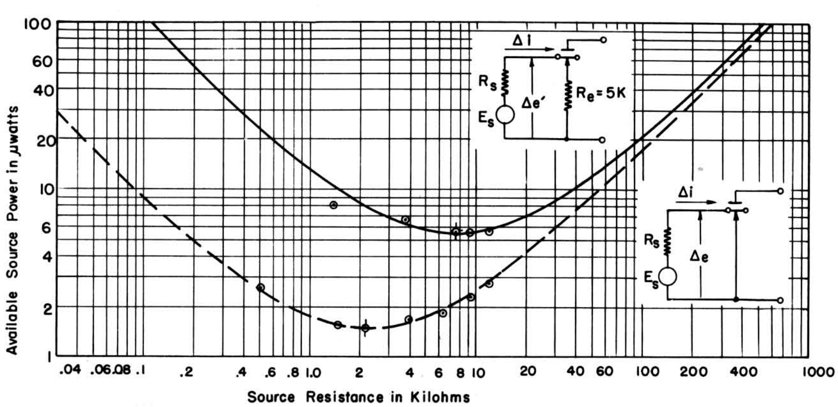


Fig. 7 – Available power to correct for drift due to temperature change plotted as a function of source resistance for the common emitter connection with and without emitter feedback. The temperature range is 30° C to 50° C. The transistor is an RCA 2N79.

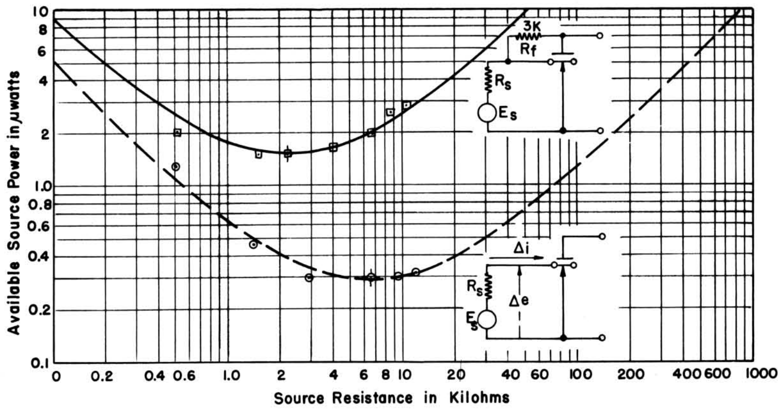


Fig. 8 – Available power to correct for drift due to temperature change plotted as a function of source resistance for the common emitter connection with and without collector feedback. The temperature range is 30° C to 50° C.

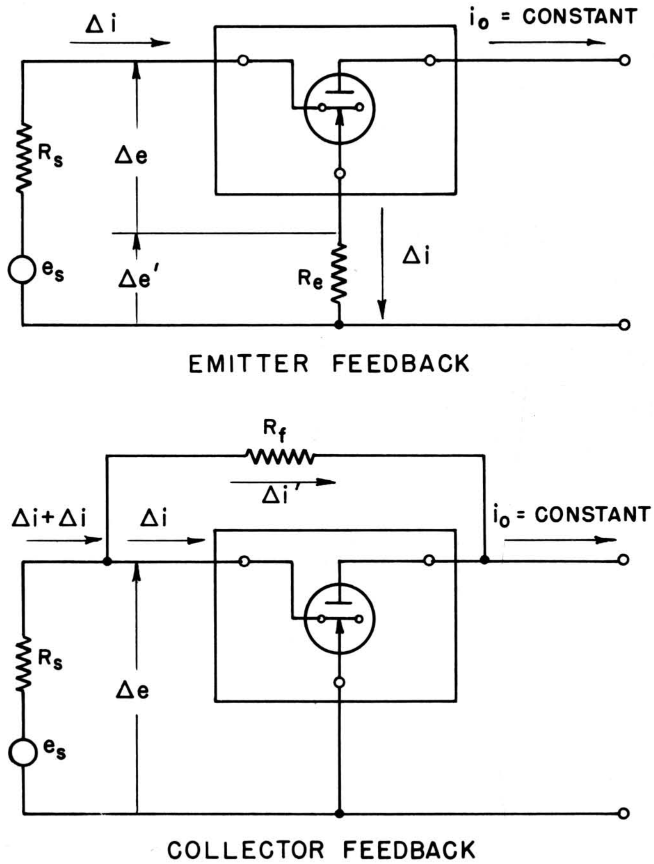
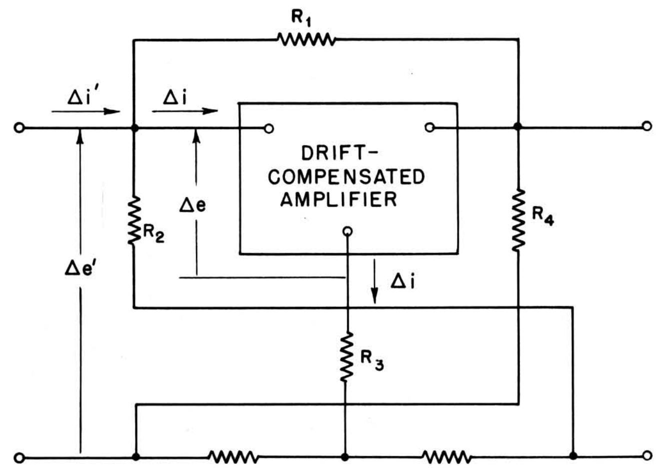


Fig. 9 – Feedback increases the power to correct for drift reducing the signal-to-drift ratio.



Δe , Δi = INPUT CHANGES WHICH CORRECT DRIFT OF AMPLIFIER ALONE

Δe', Δi' = INPUT CHANGES WHICH CORRECT DRIFT OF AMPLIFIER WITH FEEDBACK

Fig. 10 – Basic circuit of four fundamental methods of feedback applied simultaneously.

between the input terminals have the effect of increasing the input power required to correct for drift.

Silicon and Germanium with Extended Temperature Ranges

Results obtained with a germanium transistor (RCA 2N109) and with a silicon transistor (TI 904-A) for different temperature ranges are plotted in Figs. 11 and 12, respectively. The power to correct for drift increases rapidly as the temperature range is extended. The optimum source resistance decreases slightly. The results for the silicon transistor represent considerable improvement over that obtainable with germanium.

In Fig. 13, also, the uncorrected collector current is plotted against the source resistance used. The magnitude of drift current decreases with increasing source resistance as previously reported[1]. The drift current magnitude is not a minimum for the optimum source resistance in this case. It is the available power needed to correct for drift that reaches a minimum for the optimum source resistance and this achieves the maximum obtainable signal-to-drift ratio. This is not a contradiction. Even though the drift magnitude is minimized by

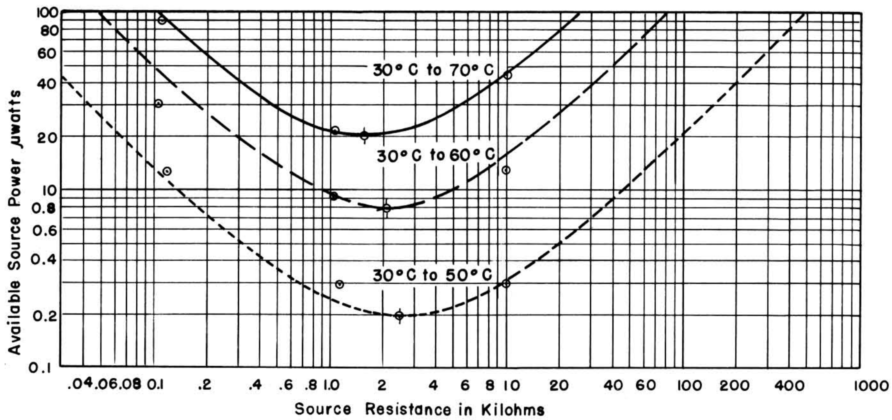


Fig. 11 – Available power to correct for drift due to temperature change plotted as a function of source resistance for a germanium transistor (RCA 2N109). The temperature ranges are 30° C to 50° C, 30° C to 60° C, 30° C to 70° C.

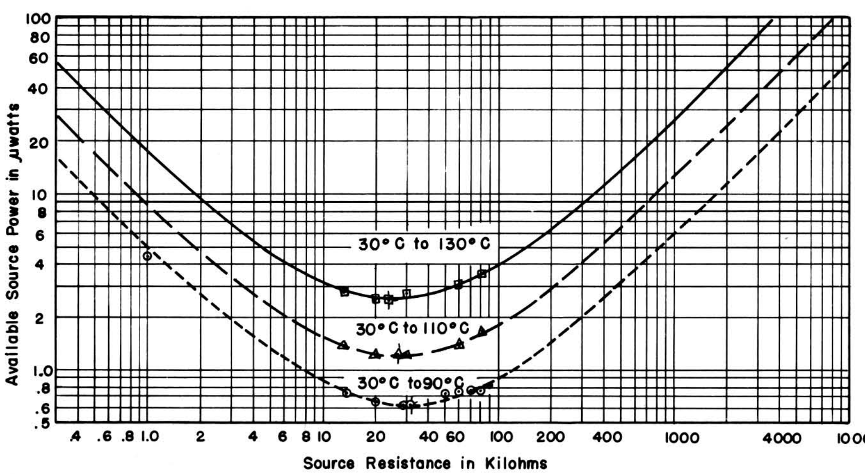


Fig. 12 – Available power to correct for drift due to temperature change plotted as a function of source resistance for a silicon transistor (TI 904-A). The temperature ranges are 30° C to 90° C, 30° C to 110° C, 30° C to 130° C.

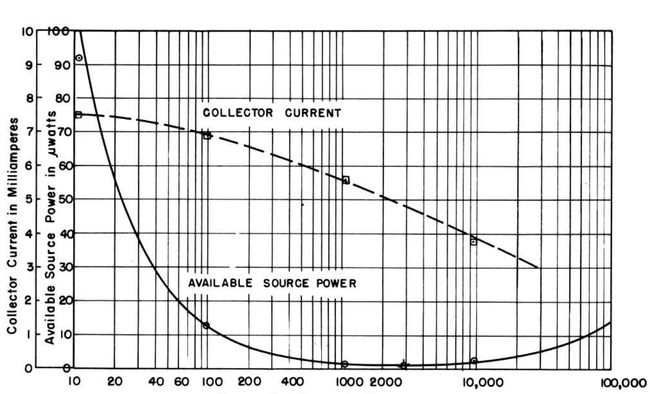


Fig. 13 – Available power to correct for drift due to temperature change and collector current plotted as a function of source resistance. The transistor is an RCA 2N109. The temperature range is 30° C to 50° C.

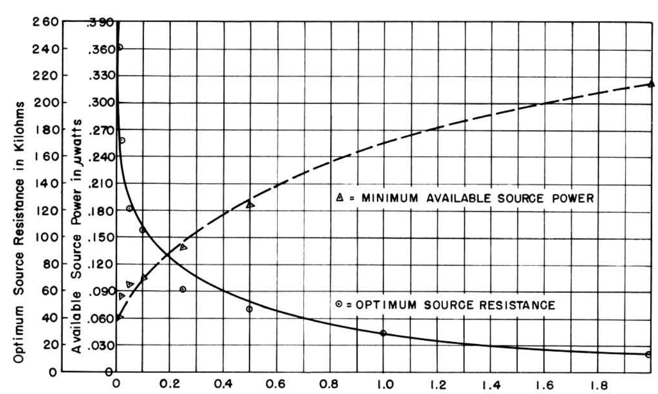


Fig. 14 – The optimum source resistance and the minimum available power to correct drift plotted as a function of emitter current for a silicon transistor (TI 904-A). Collector-emitter voltage is 6 volts. The temperature range is 30° C to 70° C.

using a very large source resistance, at the same time the mismatch between the source and the amplifier becomes so great that the available power sufficient to correct for drift is greatly increased. The drift represents the least error in terms of available power from the source when the optimum source resistance is used.

Operating Point

The obtainable signal-to-drift ratio can be improved by operating at lower emitter currents. Fig. 14 is a plot of optimum source resistance and drift correction power at different emitter currents for a silicon transistor (TI-904-A). The drift correction power decreases with lowered emitter currents while the optimum source resistance increases. Here the limit of usefulness is set by the decrease of transistor gain at low emitter currents

Compensation

Compensation is useful for improving the signal-to-drift ratio of transistor direct-coupled amplifiers. Compensation can be used at the input of an amplifier so as to improve the signal-to-drift ratio and also to alter the optimum source resistance. This is important because it admits the possibility of designing the amplifier to match a fixed source. A simple scheme is to supply compensation current at the input of the amplifier by means of a reversed biased diode which is chosen to follow the temperature-produced changes of the transistor as closely as possible. Results obtained for an RCA 2N109 with and without such diode compensation are plotted in Fig. 15. The optimum source resistance has been increased and the obtainable signal-to-drift ratio improved.

It is possible, by supplying an excess of reverse current at the base of the transistor, to overcompensate so that the direction of drift due to temperature change becomes reversed. This yields the interesting result that drift in the output is eliminated if the optimum source resistance is used with the amplifier. In Fig. 16 collector current is plotted against source resistance for such a case. The drift component of the collector current passes through zero and reverses at the optimum value of the source resistance.

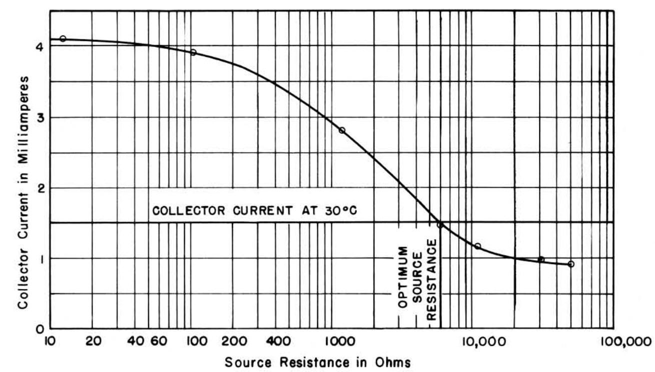


Fig. 16 – Collector current at 50°C plotted as a function of source resistance for a germanium transistor with excess reverse current compensation at the input. The drift component of collector current is eliminated at the optimum source resistance.

The means by which this eliminates drift is illustrated by Fig. 17. The reverse saturation current from the diode flows to the device and to the source. If the source resistance is so chosen that the resulting voltage drop is that required for compensation, drift is eliminated. Such a result is possible if the voltage and current change needed for compensation are opposite in sign.

Certain practical limits exist for this case. The optimum source resistance for an amplifier will be sub-

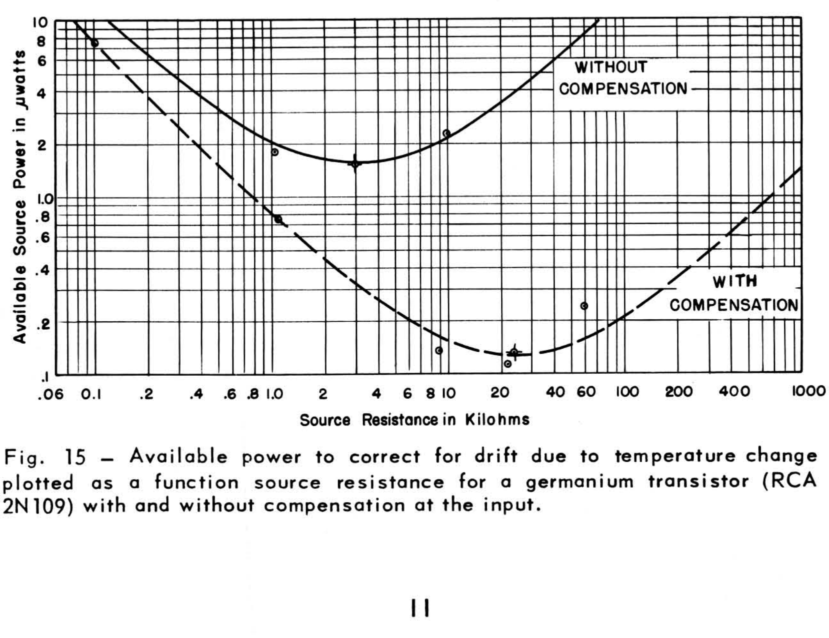


Fig. 15 – Available power to correct for drift due to temperature change plotted as a function source resistance for a germanium transistor (RCA 2N109) with and without compensation at the input.

ject to some spread of tolerance limits; a like situation exists for the source resistance so that an exact match is improbable. Also the optimum source resistance changes somewhat with the temperature increment involved. Such factors combine to insure that drift will not be completely eliminated.

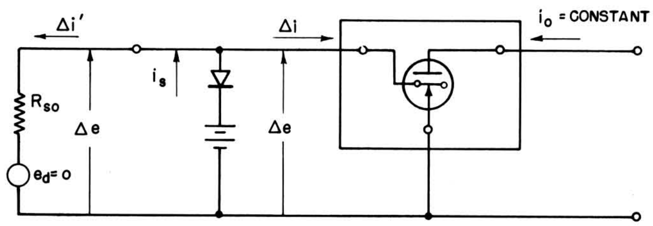


Fig. 17 – Compensation reverses drift correction current.

Application

The correct application of these principles to the design of a direct-coupled amplifier requires careful appraisal of design objectives. For example, the distinction between drift magnitude and signal-to-drift ratio is an important one. Minimum drift does not necessarily imply best signal-to-drift ratio. Negative feedback reduces drift but also decreases the signal-to-drift ratio. A common collector stage drifts much less than common emitter but the signal-to-drift capabilities are nearly equal. Performance requirements with respect to drift, gain, and signal to drift are therefore separate factors to take into account.

The signal-to-drift ratio of an amplifier is improved by reducing available source power which will correct drift. With a particular amplifier this value is a minimum for the optimum source resistance. However, in many cases the source resistance is fixed and not equal to the optimum source resistance of the amplifier. This means only that if the source resistance could be altered without losing available power, then the obtainable signal-to-drift ratio for that amplifier could be achieved.

No increase of signal-to-drift ratio is realized by attempting to improve the match with added resistance between the signal source and the input of an amplifier. If such added resistance is regarded as part of the signal source the resulting available signal power is lower. If the added resistance is regarded as part of the amplifier it absorbs power from the source, reducing the power delivered as effective drive to the amplifier. In any case, that amplifier which provides the best signal-to-drift ratio requires the least available power from the given source to correct drift.

There are widespread applications of direct-coupled amplifiers for which the source resistance can be altered without loss of available source power. An example is the use of a direct-coupled amplifier in the AFC loop of a controlled oscillator as shown in Fig. 18. In this application the direct-coupled amplifier is used to drive a frequency control element which varies the frequency of the Variable Frequency Oscillator. The d-c source for this amplifier is a phase discriminator which provides an output if there is a frequency shift of the VFO. The output of the phase discriminator represents a comparison between the VFO and the crystal-controlled oscillator, which provides a signal for automatic frequency control.

In this application, drift of the d-c amplifier results in an associated drift of the VFO. Consequently, the frequency stability of the VFO depends upon the signal-to-drift ratio of the amplifier. In such a case the obtainable signal-to-drift ratio of the amplifier can be achieved if the source resistance represented by the phase discriminator is set equal to the optimum source resistance of the amplifiers.

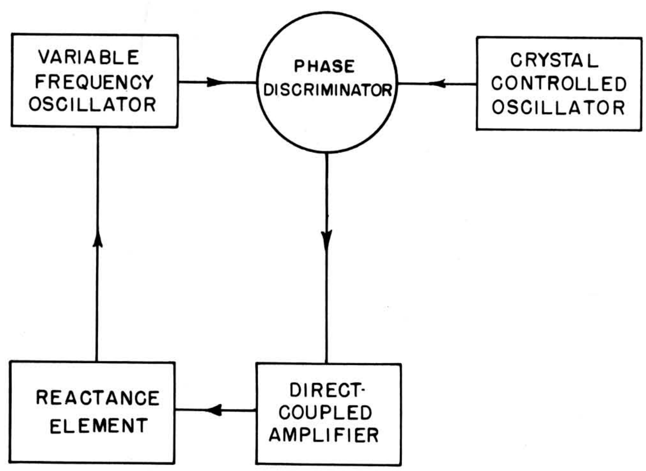


Fig. 18 – Direct-coupled amplifier for automatic frequency control.

Measurement Of Available Power To Correct For Drift

The circuit of Fig. 19 was used to determine the available source power to correct for drift due to temperature change. In this circuit E_{st} is a standard voltage source; E_v is a variable voltage source. The important metered quantities are E, the voltage difference between the sources, I_b, the base current and I_c, the collector current. The procedure is to set the variable voltage equal to the standard at the initial temperature and then to adjust R_2 and R_1 for the desired source resistance and operating point. Then the variable voltage source, E_v, is used to correct for drift due to the temperature change of interest; i.e., the collector current is maintained constant. The available source power required to do this

$$\frac{\left(\dfrac{E\,R_1}{R_1 + R_2}\right)^2}{4\left(\dfrac{E_1\,R_2}{R_1 + R_2}\right)} = \frac{E^2\,R_1}{4\,R_2\,(R_2 + R_1)}$$

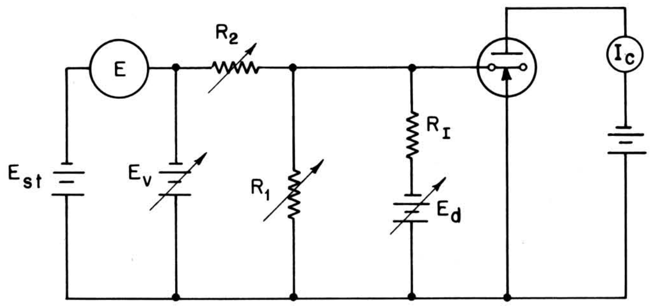


Fig. 19 – Measurement of available source power to correct for drift.

Values of this available power obtained over a range of source resistance show a minimum for the optimum source resistance.

It is possible to determine the optimum source resistance directly from one such measurement since $R_{SO} = \Delta e/\Delta i$.

The current change at the input, Δi, is the difference between base currents at the initial and final temperatures for constant collector current. The corresponding voltage change at the input, Δe, can be determined since the equivalent source voltage and resistance are known. The relationship in terms of measured quantities is:

$$\Delta e = \frac{E\,R_1}{R_1 + R_2} - \frac{R_1\,R_2}{R_1 + R_2}\,(I_{bf} - I_{bi})$$

where I_{bi} and I_{bf} are the initial and final base current readings respectively. This permits a check of each determination and fixes the optimum value more precisely than would be possible with plotted curve alone.

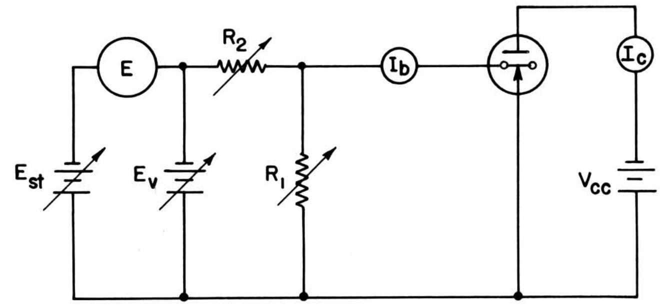


Fig. 20 – Measurement of small currents to correct for drift.

If the input current changes are very small some difficulty is encountered with this procedure of direct measurement. Very small currents can be introduced readily by the method shown in Fig. 20. The operating point and source resistance are established as before by manipulating R_1 and R_2. Small input current changes are introduced by means of the large resistor R_1 and voltage source E_d. For example, if R_I is 10 megohms and E_d is one volt an input current of 1/10 μa is obtained. Two measurements are required to fix the values of current and voltage change at the input of the transistor. Thus:

$$\Delta e = (i_1 - \Delta i)\,R_{S1}$$

$$\Delta e = (i_2 - \Delta i)\,R_{S2}$$

where i_1 and i_2 are the small current increments introduced and R_{S1} and R_{S2} are the different equivalent source resistances chosen for the determination. Then:

$$\Delta i = \frac{R_{S1}\,i_1 - R_{S2}\,i_2}{R_{S1} - R_{S1}}$$

$$\Delta e = \frac{R_{S1}\,R_{S2}\,(i_1 - i_2)}{R_{S2} - R_{S1}}$$

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References

1. LB-979 *Temperature Effects in Circuits Using Junction Transistors.* H. C. Lin and A. A. Barco.
2. H. T. Friis, 'Noise Figures of Radio Receivers,' *Proc. IRE,* July, 1944.
3. H. Goldberg, 'Some Notes on Noise Figures,' *Proc. IRE,* October, 1948.
4. E. W. Herold, 'An Analysis of the Signal-to-Noise Ratio of Ultra High Frequency Receivers,' *RCA Review,* Vol. 6, January 1942. pp 302-332.
5. Standards on Transducers; Definitions of Terms, 1951, *Proc. IRE,* August, 1951, pp 897-900.
6. Wallace & Pietenpol, 'Properties and Applications of n-p-n Transistors,' *Proc. IRE.*
7. F. H. Blecher, 'Transistor Circuits for Analog and Digital Systems,' *B.S.T.J.,* Vol. 35, March, 1956. pp 661-670.
8. E. Keonjian, 'Temperature-Compensated DC Transistor Amplifier,' *Proc. IRE,* Vol. 42, April, 1954, pp 661-670.
9. J. W. Stanton, 'A Transistorized D-C Amplifier,' *IRE Transactions On Circuit Theory,* pp 65-66.