Some Circuit Design Techniques for Linear Integrated Circuits

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Abstract—**This paper describes some novel circuit design techniques for linear integrated circuits. These techniques make use of elements which can be constructed easily in monolithic form to avoid processing difficulties or substantial yield losses in manufacturing. Methods are shown which eliminate the need for the wide variety of components and tight component tolerances usually required with discrete designs, which substitute parts which can be made simply, and which make use of special characteristics obtainable with monolithic construction.**

Introduction

THE COMPONENTS available in integrated circuit processes which are in production today are limited both in type and in range of values. This limitation has been more troublesome with linear circuits than with digital, since many more different kinds of parts are used with conventional designs. Therefore, serious problems have arisen in adapting discrete-component designs to monolithic construction.

In many cases an attempt has been made to develop new technology in order to integrate existing designs. This approach has met with varying degrees of success, but has generally resulted in circuits which are difficult to produce in large volume. To realize a practical microcircuit with any certainty and within a reasonable period of time, it becomes necessary to use existing production technology. Fortunately, with some specialized circuit design techniques, it is possible in certain cases to achieve performance with present technology which is equal to or better than that obtainable with discrete-component circuits. These techniques make it possible to avoid the restrictions imposed by limited types of components, poor tolerances, and restricted range of component values. They make use of the inherent advantages of integrated circuits: close matching of active and passive devices over a wide temperature range, excellent thermal coupling throughout the circuit, the economy of using a large number of active devices, the freedom of selection of active device geometries, and the availability of devices which have no exact discrete-element counterpart.

Some of these techniques will be discussed; an attempt will be made to indicate in a broad sense what can be done rather than going into great detail on particulars. Examples of practical circuits will be given to illustrate important points.

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Biasing Circuits

One of the most basic problems encountered in integrated circuits is bias stabilization of a common emitter amplifier. Conventional methods [1] usually require substantial dc degeneration and a bypass capacitor to reduce the degeneration at the frequencies to be amplified. With integrated circuits, the required bypass capacitors are much too large to be practical. In the past, this problem has been overcome using some sort of differential [2] or emitter-coupled [3] amplifier connection. These solutions have been adequate in many instances, but they suffer from a lack of versatility.

The close matching of components and tight thermal coupling obtained in integrated circuits permit much more radical solutions. An example is given in Fig. 1(a). A current source can be implemented by imposing the emitter-base voltage of a diode-connected transistor operating at one collector current across the emitter-base junction of a second transistor. If the two transistors are identical, their collector currents will be equal; hence, the operating current of the current source can be determined from the resistor (R_1) and the supply voltage (V^+). Experiment has shown that this biasing scheme is stable over a wide temperature range, giving collector current matches between the biasing and operating devices typically better than five percent, even for power dissipations in Q_2 above 100 mW.

An extension of this idea is shown in Fig. 1(b). A transformer with a low resistance secondary can be inserted between the biasing transistor (Q_1) and the second transistor (Q_2). Then Q_2 is stably biased as an amplifier without requiring any bypass elements, and the transformer secondary is coupled to the amplifier without disturbing the bias conditions.

A third and more subtle variation is given in Fig. 1(c). If R_3 and R_4 as well as Q_1 and Q_2 are identical, the collector currents of the two transistors will be equal, since their bases are driven from a common voltage point through equal resistances. The collector current of Q_1 will be given by

$$I_{C1} = \frac{V^+ - V_{BE}}{R_1} - \left(2 + \frac{R_3}{R_1}\right) I_B \tag{1}$$

where a single V_{BE} and I_B term is used since both transistors are identical. For $V_{BE} \ll V^+$ and $I_B \ll I_C$,

$$I_{C1} = I_{C2} \cong \frac{V^+}{R_1}. \tag{2}$$

If $R_2 = \frac{1}{2}R_1$, then

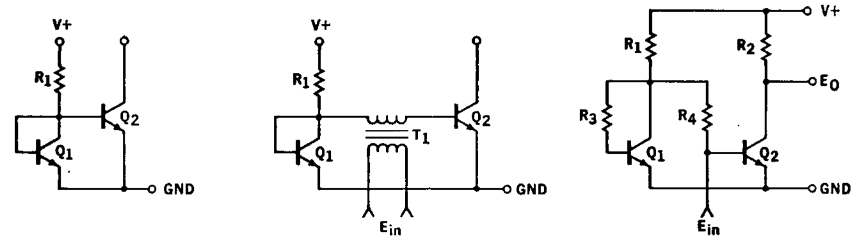


Fig. 1. Biasing techniques applicable to integrated circuitry which take advantage of precise matching and close thermal coupling. (a) Current source. (b) Transformer coupled amplifier. (c) RC-coupled amplifier.

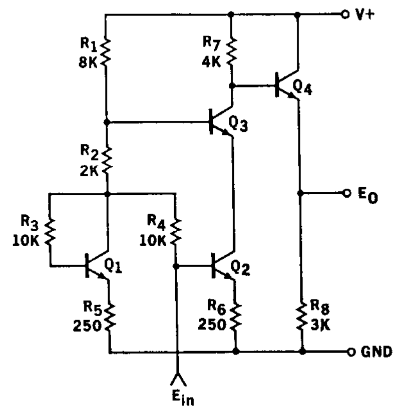


Fig. 2. General purpose wideband amplifier using balanced biasing techniques.

$$E_0 \cong \frac{V^+}{2} \tag{3}$$

which means that the amplifier will be biased at its optimum operating point, at one half the supply voltage, independent of the supply voltage as well as temperature and dependent only on how well the parts within the integrated circuit match.

An evaluation of the effects of mismatch on bias stability is given in the Appendix.

A simple amplifier using the biasing of Fig. 1(c) is illustrated in Fig. 2. An emitter degeneration resistor (R_6) is employed in conjunction with R_5 to control gain and raise input impedance without disturbing the balanced biasing. A cascode connection of Q_2 with Q_3 reduces input capacitance while the emitter follower (Q_4) gives a low output impedance.

It can be seen from the foregoing that the matching characteristics of a monolithic circuit have made possible biasing methods which are superior to those practically attainable with discrete designs. Excellent biasing stability is achieved over a wide temperature range without wasting any of the supply voltage across bias stabilization networks. In addition, no bypass capacitors are required.

The use of a circuit similar to that in Fig. 1(c) as the second stage of differential input, single-ended output amplifiers is described in [4]–[6].

Constant Current Source

The formation of current sources in the microampere current range can be difficult with integrated circuits because of the relatively large resistance values usually required. A circuit is shown in Fig. 3 which makes possible a current source with outputs in the tens of microamperes using resistances of only a few kilohms. It makes use of the predictable difference of the emitter-base voltage of two transistors operating at different collector currents. Its operation can be described as follows: the collector current of a transistor is given as a function of emitter-base voltage by

$$I_C = I_S \exp\cdot\left(\frac{qV_{BE}}{kT}\right), \tag{4}$$

where I_S is the so-called saturation current, q is the charge of an electron, k is Boltzmann's constant and T is the absolute temperature. This expression holds up to high currents where emitter contact and base spreading resistances become important and down to low currents where collector leakage currents cause inaccuracy. It has been shown to be valid, within a percent or so, for operation over at least six decades of collector current with well-made silicon transistors [7]–[9]. This contrasts with similar expressions for diode current and the emitter current of transistors which show substantial error over three decades of current operation.

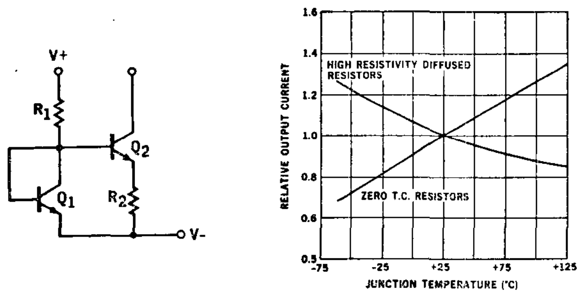


Fig. 3. A current source for generating very small currents using moderate value resistors.

Solving (4) for V_{BE} gives

$$V_{BE} = \frac{kT}{q} \log_e \frac{I_C}{I_S}. \tag{5}$$

This expression can be used to find the emitter-base voltage difference between two transistors:

$$\begin{aligned} \Delta V_{BE} &= V_{BE1} - V_{BE2} \\ &= \frac{kT}{q} \log_e \frac{I_{C1}}{I_{S1}} - \frac{kT}{q} \log_e \frac{I_{C2}}{I_{S2}} \\ &= \frac{kT}{q} \log_e \frac{I_{C1}}{I_{C2}} + \frac{kT}{q} \log_e \frac{I_{S2}}{I_{S1}}. \end{aligned} \tag{6}$$

For equal collector currents, this becomes

$$\Delta V_{BE} = \frac{kT}{q} \log_e \frac{I_{S2}}{I_{S1}}. \tag{7}$$

Considerable testing has shown that for adjacent, identical integrated circuit transistors this term is typically less than 0.5 mV. It is also relatively independent of the current level, as might be expected since I_S should be a constant. Hence, the emitter-base voltage differential between adjacent integrated circuit transistors operating at different collector currents is given by

$$\Delta V_{BE} = \frac{kT}{q} \log_e \frac{I_{C1}}{I_{C2}} \tag{8}$$

within a fraction of a millivolt.

With the circuit in Fig. 3, a collector current which is large by comparison to the desired current-source current is passed through the diode-connected baising transistor Q_1. Its emitter-base voltage is used to bias the current-source transistor Q_2. If the base currents of the transistors are neglected for simplicity, the resistance required to determine the current-source current is given by

$$R_2 = \frac{\Delta V_{BE}}{I_{C2}} = \frac{kT}{qI_{C2}} \log_e \frac{I_{C1}}{I_{C2}}. \tag{9}$$

Or, for the circuit in Fig. 3,

$$R_2 = \frac{kT}{qI_{C2}} \log_e \left[\frac{V^+ - V_{BE}}{R_1 I_{C2}}\right]. \tag{10}$$

The effect of nonzero base currents can be easily determined in that they both subtract directly from I_{C1}, and I_{B2} subtracts from I_{C2}.

One interesting feature of this circuit is that for $V^+ \gg V_{BE}$ and $I_{C1} \gg I_{C2}$ the output current will vary roughly as the logarithm of the supply voltage (V^+). Therefore, if the current source is used in such an application as the input stage of an operational amplifier, the operating collector current and voltage gain of the input stage will vary little over an extremely wide range of supply voltages.

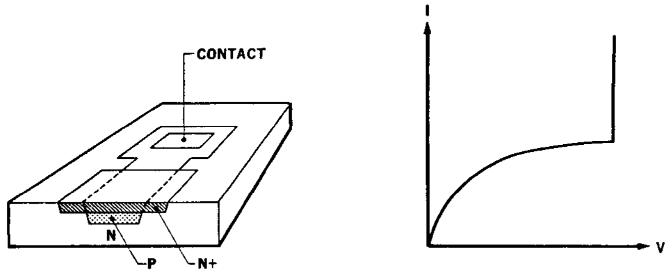
From (8) it can be seen that the emitter-base voltage differential is a linear function of absolute temperature. Therefore, it might be expected that the output current of the current source would vary in a similar manner. Such is the case, as illustrated in Fig. 3. The plot is for $I_{C1} \cong 50 I_{C2}$ with both zero temperature coefficient resistors and high-resistivity diffused resistors (bulk impurity concentration less than 10^{17} atoms/cm^3). It is notable that diffused resistors provide overcompensation for this characteristic.

Pinch Resistors

A potentially useful element in integrated circuits which has received much mention but little actual application is the pinch resistor [10]. It is an ordinary diffused (base) resistor, the cross-sectional area of which has been effectively reduced by making an emitter diffusion on top of it [see Fig. 4(a)]. The emitter diffusion raises the sheet resistivity from the usual 100 or 200 Ω/sq to 10 KΩ/sq or higher. This permits rather large resistors to be made in a relatively small area. The pinch resistor, however, has several limiting characteristics. As can be seen from Fig. 4(b), it is linear only for small voltage drops and it has a low breakdown voltage (5 to 10 volts). Neither the linear nor the nonlinear portions of the characteristics can be controlled well and the resistance at the origin can easily vary over a 4-to-1 range in a normal production situation. In addition, the resistor has a very strong positive temperature coefficient, changing by about 3 to 1 over the -55°C to $+125$°C temperature range.

On the other hand, there is a strong correlation between the pinch resistor values and transistor current gains obtained in manufacture. Within a given process, the sheet resitivity is roughly proportional to the current gains near the current-gain peak (where surface effects have little influence on the current gain). Further, the resistors tend to track with the current gains over temperature. The matching of identical pinch resistors is also nearly as good as base resistors and substantially better than transistor current gains, since the current gains are affected by unpredictable surface phenomena whereas the pinch resistors are not.

Figure 2 provides an example of where pinch resistors can be used effectively. Both R_3 and R_4 have small voltage drops across them, and it would be advantageous to have these resistor values proportional to the transistor current gain to obtain the highest possible input im-



(a) (b)

Fig. 4. Pinch resistors. (a) Structure. (b) Characteristics.

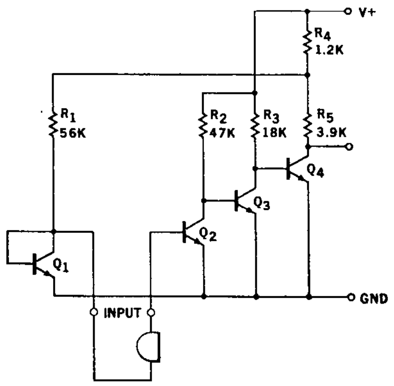


Fig. 5. A low-voltage high-gain microphone preamplifier illustrating the use of pinch resistors.

pedance consistant with satisfactory bias stability.

Another application is the preamplifier shown in Fig. 5 which was designed as part of a hearing aid amplifier. With hearing aids, the maximum supply voltage is 1.55 volts, so the voltage sensitivity and low breakdown of pinch resistors is of little concern. However, power drain is a problem, so large resistances are needed. In this circuit only matching of the pinch resistors (R_1, R_2, and R_3) is required for proper operation. The fact that the pinch resistors correlate with current gain makes the circuit far less sensitive to current gain variations: the pinch resistors and current gains can be varied simultaneously over a range greater than 7 to 1 without any noticeable degradation of performance.

The preceeding examples show that even though pinch resistors have extremely poor characteristics by discrete component standards, certain characteristics, namely good matching, a correlation between resistor values and current gain and high sheet resistivities make them extremely useful elements in circuit design. The pinch resistors can actually function better than precision resistors in certain applications in that they can be used to help compensate for production variations in current gain and the change in gain with temperature.

Conclusions

Biasing techniques have been demonstrated which eliminate the need for the large bypass capacitors required with conventional designs and yield superior performance. A microampere current source using resistance values at least an order of magnitude less than those normally required was also shown. Finally, pinch resistors, devices which can give large resistances in microcircuits, were discussed with respect to both their limitations and their unique properties; examples of how they could be used effectively in circuit design were introduced.

It has been suggested that there are far better approaches available than directly adapting discrete component designs to microcircuits. Many of the restrictions imposed by monolithic construction can be overcome on a circuit design level. This is of particular practical significance because circuit design is a nonrecurring cost in a particular microcircuit while restrictive component tolerances or extra processing steps represent a coutinuing expense in manufacture.

Appendix

The Effect of Mismatches on Balanced Biasing

The assumption that the two transistors in Fig. 1(c) were identical is, of course, not entirely true in a practical microcircuit. This deviation from reality will now be considered.

In (1) of the text, the mismatch in emitter-base voltage and base currents are certainly third-order effects since the absolute values of V_{BE} and I_B are second-order terms. Therefore, (1) will be assumed to hold in this analysis. Equation (2) is, however, strongly affected by the match between Q_1 and Q_2. This influence will be reflected by the equation

$$I_{C2} = I_{C1} + \Delta I_{C2}, \tag{1a}$$

where ΔI_{C2} is the change in collector current of Q_2 due to unequal emitter-base voltages and base currents in Q_1 and Q_2.

Both of the preceding mismatch terms can be combined in

$$\Delta V_{IN} = \Delta V_{BE} + R_4 \Delta I_B \tag{2a}$$

where ΔV_{BE} and ΔI_B are the emitter-base voltage difference and base current difference of the two devices operating at equal collector currents. (These terms are chosen because they are important in the performance of dc amplifiers and because data on the parameters is common, both as transistor pairs and as complete integrated amplifiers.) This ΔV_{IN} will have the same effect on bias stability as a dc input voltage, equal in magnitude, in series with R_4 on an ideally balanced amplifier.

In order to determine the relationship between ΔV_{IN} and ΔI_{C2}, which is the objective of this Appendix, it becomes necessary to introduce the two expressions found in [7],

$$I_C \propto \exp\cdot\left(\frac{qV_{BE}}{kT}\right) \tag{3a}$$

and

$$I_B \propto \exp\cdot\left(\frac{qV_{BE}}{mkT}\right), \tag{4a}$$

which have the same range of validity stated for (4) in the text.

Differentiation of (3a) gives

$$\frac{\Delta I_C}{\Delta V_{BE}} = \frac{qI_C}{kT}. \tag{5a}$$

Similarly, for (4a)

$$\frac{\Delta I_B}{\Delta V_{BE}} = \frac{qI_B}{mkT} \tag{6a}$$

whereas

$$R_{IN} = \frac{mkT}{qI_B}. \tag{7a}$$

From this,

$$\Delta I_{C2} = \left(\frac{R_{IN}\Delta V_{IN}}{R_{IN} + R_4}\right)\left(\frac{qI_{C2}}{kT}\right)$$

$$= \frac{m(\Delta V_{BE} + R_4 \Delta I_B)}{\dfrac{mkT}{q} + R_4 I_{B2}} I_{C2} \tag{8a}$$

where m is electrically significant as the ratio of the ac current gain to the dc current gain at the operating current level (typically 1.6 for low-current operation and 1.2 for operation approaching the current gain peak).

Hence,

$$\Delta I_{C2} \cong \frac{m(\Delta V_{BE} + R_4 \Delta I_B)}{\dfrac{mkT}{q} + R_4 I_B} I_{C1}. \tag{9a}$$

For $R_{IN} \ll R_4$ this becomes

$$\frac{\Delta I_{C2}}{I_{C1}} = \frac{m\Delta I_B}{I_B} = \Delta h_{fe} \tag{10a}$$

where

$$\Delta h_{fe} = \frac{h_{fe2}}{h_{fe1}}.$$

For $R_{IN} \gg R_4$,

$$\frac{\Delta I_{C2}}{\Delta I_{C1}} = \frac{q\Delta V_{BE}}{kT}. \tag{11a}$$

Since R_{IN} varies drastically over the $-55°C$ to $+125°C$ temperature range commonly expected for integrated circuits (10a) can be taken to represent the low-temperature extreme and (11a) the high-temperature extreme. Therefore, a more-than-worst-case solution for full temperature range operation is

$$\frac{\Delta I_{C2}}{I_{C1}} \leq \frac{q\Delta V_{BE}}{kT} + \Delta h_{fe}. \tag{12a}$$

To give some feeling for the results obtainable, the first term in (12a) is typically less than 0.02 while the second term is typically less than 0.07 for full temperature range operation.

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