Ch2: Transistors

INTRODUCTION

The transistor is our most important example of an "active" component, a device that can amplify, producing an output signal with more power in it than the input signal. The additional power comes from an external source of power (the power supply, to be exact). Note that voltage amplification isn't what matters, since, for example, a step-up transformer, a "passive" component just like a resistor or capacitor, has voltage gain but no power gain. Devices with power gain are distinguishable by their ability to make oscillators, by feeding some output signal back into the input.

It is interesting to note that the property of power amplification seemed very important to the inventors of the transistor. Almost the 'first thing they did to convince themselves that they had really invented something was to power a loudspeaker from a transistor, observing that the output signal sounded louder than the input signal.

The transistor is the essential ingredient of every electronic circuit, from the simplest amplifier or oscillator to the most elaborate digital computer. Integrated circuits (ICs), which have largely replaced circuits constructed from discrete transistors, are themselves merely arrays of transistors and other components built from a single chip of semiconductor material.

A good understanding of transistors is very important, even if most of your circuits are made from ICs, because you need to understand the input and output properties of the IC in order to connect it to the rest of your circuit and to the outside world. In addition, the transistor is the single most powerful resource for interfacing, whether between ICs and other circuitry or between one subcircuit and another. Finally, there are frequent (some might say too frequent) situations where the right IC just doesn't exist, and you have to rely on discrete transistor circuitry to do the job. As you will see, transistors have an excitement all their own. Learning how they work can be great fun.

Our treatment of transistors is going to be quite different from that of many other books. It is common practice to use the h-parameter model and equivalent circuit. In our opinion that is unnecessarily complicated and unintuitive. Not only does circuit behavior tend to be revealed to you as something that drops out of elaborate equations, rather than deriving from a clear understanding in your own mind as to how the circuit functions; you also have the tendency to lose sight of which parameters of transistor behavior you can count on and, more important, which ones can vary over large ranges.

In this chapter we will build up instead a very simple introductory transistor model and immediately work out some circuits with it. Soon its limitations will become apparent; then we will expand the model to include the respected Ebers-Moll conventions. With the Ebers-Moll equations and a simple 3-terminal model, you will have a good understanding of transistors; you won't need to do a lot of calculations, and your designs will be first-rate. In particular, they will be largely independent of the poorly controlled transistor parameters such as current gain.

Some important engineering notation should be mentioned. Voltage at a transistor terminal (relative to ground) is indicated by a single subscript (C, B, or E): V_C is the collector voltage, for instance. Voltage between two terminals is indicated by a double subscript: V_{BE} is the base-to-emitter voltage drop, for instance. If the same letter is repeated, that means a power-supply voltage: V_{CC} is the (positive) power-supply voltage associated with the collector, and V_{EE} is the (negative) supply voltage associated with the emitter.

2.01 First transistor model: current amplifier

Let's begin. A transistor is a 3-terminal device (Fig. 2.1) available in 2 flavors (npn and pnp), with properties that meet the following rules for npn transistors (for pnp simply reverse all polarities):

1. The collector must be more positive than the emitter.

2. The base-emitter and base-collector circuits behave like diodes (Fig. 2.2). Normally the base-emitter diode is conducting and the base-collector diode is reverse-biased, i.e., the applied voltage is in the opposite direction to easy current flow.

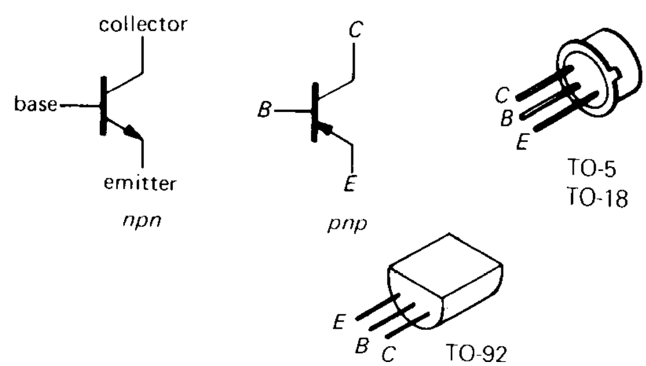


Figure 2.1. Transistor symbols, and small transistor packages.

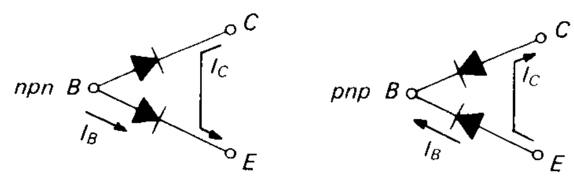


Figure 2.2. An ohmmeter's view of a transistor's terminals.

3. Any given transistor has maximum values of I_C, I_B, and V_{CE} that cannot be exceeded without costing the exceeder the price of a new transistor (for typical values, see Table 2.1). There are also other limits, such as power dissipation ($I_C V_{CE}$), temperature, V_{BE}, etc., that you must keep in mind.

4. When rules 1–3 are obeyed, I_C is roughly proportional to I_B and can be written as

$$I_C = h_{FE} I_B = \beta I_B$$

where h_{FE}, the current gain (also called beta), is typically about 100. Both I_C and I_E flow to the emitter. Note: The collector current is not due to forward conduction of the base-collector diode;

that diode is reverse-biased. Just think of it as "transistor action."

Property 4 gives the transistor its usefulness: A small current flowing into the base controls a much larger current flowing into the collector.

Warning: h_{FE} is not a "good" transistor parameter; for instance, its value can vary from 50 to 250 for different specimens of a given transistor type. It also depends upon the collector current, collector-to-emitter voltage, and temperature. *A circuit that depends on a particular value for h_{FE} is a bad circuit.*

Note particularly the effect of property 2. This means you can't go sticking a voltage across the base-emitter terminals, because an enormous current will flow if the base is more positive than the emitter by more than about 0.6 to 0.8 volt (forward diode drop). This rule also implies that an operating transistor has $V_B \approx V_E + 0.6$ volt ($V_B = V_E + V_{BE}$). Again, polarities are normally given for *npn* transistors; reverse them for *pnp.*

Let us emphasize again that you should not try to think of the collector current as diode conduction. It isn't, because the collector-base diode normally has voltages applied across it in the reverse direction. Furthermore, collector current varies very little with collector voltage (it behaves like a not-too-great current source), unlike forward diode conduction, where the current rises very rapidly with applied voltage.

SOME BASIC TRANSISTOR CIRCUITS

2.02 Transistor switch

Look at the circuit in Figure 2.3. This application, in which a small control current enables a much larger current to flow in another circuit, is called a transistor switch. From the preceding rules it is easy to understand. When the mechanical switch is open, there is no base current. So, from

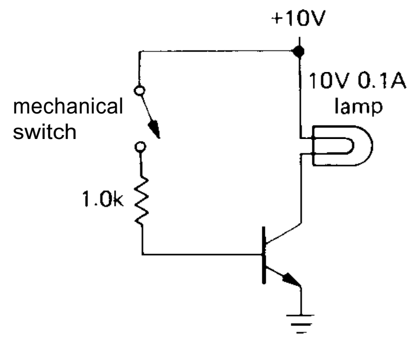


Figure 2.3. Transistor switch example.

rule 4, there is no collector current. The lamp is off.

When the switch is closed, the base rises to 0.6 volt (base-emitter diode is in forward conduction). The drop across the base resistor is 9.4 volts, so the base current is 9.4mA. Blind application of rule 4 gives I_C = 940mA (for a typical beta of 100). That is wrong. Why? Because rule 4 holds only if rule 1 is obeyed; at a collector current of 100mA the lamp has 10 volts across it. To get a higher current you would have to pull the collector below ground. A transistor can't do this, and the result is what's called saturation – the collector goes as close to ground as it can (typical saturation voltages are about 0.05–0.2V, see Appendix G) and stays there. In this case, the lamp goes on, with its rated 10 volts across it.

Overdriving the base (we used 9.4mA when 1.0mA would have barely sufficed) makes the circuit conservative; in this particular case it is a good idea, since a lamp draws more current when cold (the resistance of a lamp when cold is 5 to 10 times lower than its resistance at operating current). Also transistor beta drops at low collector-to-base voltages, so some extra base current is necessary to bring a transistor into full saturation (see Appendix G). Incidentally, in a real circuit you would probably put a resistor from base to ground (perhaps 10k in this case) to make sure the base is at ground with the switch open. It wouldn't affect the "on" operation, because it would sink only 0.06mA from the base circuit.

There are certain cautions to be observed when designing transistor switches: *1.* Choose the base resistor conservatively to get plenty of excess base current, especially when driving lamps, because of the reduced beta at low V_{CE}. This is also a good idea for high-speed switching, because of capacitive effects and reduced beta at very high frequencies (many megahertz). A small "speedup" capacitor is often connected across the base resistor to improve high-speed performance.
2. If the load swings below ground for some reason (e.g., it is driven from ac, or it is inductive), use a diode in series with the collector (or a diode in the reverse direction to ground) to prevent collector-base conduction on negative swings.
3. For inductive loads, protect the transistor with a diode across the load, as shown in Figure 2.4. Without the diode the inductor will swing the collector to a large positive voltage when the switch is opened, most likely exceeding the collector-emitter breakdown voltage, as the inductor tries to maintain its "on" current from V_{CC} to the collector (see the discussion of inductors in Section 1.31).

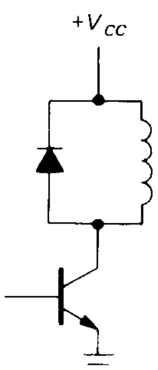


Figure 2.4. Always use a suppression diode when switching an inductive load.

Transistor switches enable you to switch very rapidly, typically in a small fraction of a microsecond. Also, you can switch many different circuits with a single control signal. One further advantage is the possibility of remote cold switching, in which only dc control voltages snake around through cables to reach front-panel switches, rather than the electronically inferior approach of having the signals themselves traveling through cables and switches (if you run lots of signals through cables, you're likely to get capacitive pickup as well as some signal degradation).

"Transistor man"

Figure 2.5 presents a cartoon that will help you understand some limits of transistor

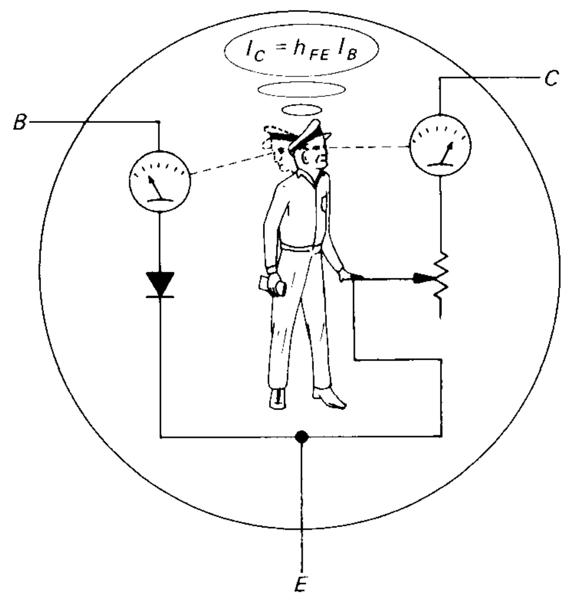


Figure 2.5. "Transistor man" observes the base current, and adjusts the output rheostat in an attempt to maintain the output current h_{FE} times larger.

behavior. The little man's perpetual task in life is to try to keep $I_C = h_{FE}I_B$; however, he is only allowed to turn the knob on the variable resistor. Thus he can go from a short circuit (saturation) to an open circuit (transistor in the "off" state), or anything in between, but he isn't allowed to use batteries, current sources, etc. One warning is in order here: Don't think that the collector of a transistor looks like a resistor. It doesn't. Rather, it looks approximately like a poor-quality constant-current sink (the value of current depending on the signal applied to the base), primarily because of this little man's efforts.

Another thing to keep in mind is that, at any given time, a transistor may be (a) cut off (no collector current), (b) in the active region (some collector current, and collector voltage more than a few tenths of a volt above the emitter), or (c) in saturation (collector within a few tenths of a volt of the emitter). See Appendix G on transistor saturation for more details.

2.03 Emitter follower

Figure 2.6 shows an example of an *emitter follower*. It is called that because the output terminal is the emitter, which follows the input (the base), less one diode drop:

$V_E \approx V_B - 0.6$ volt

The output is a replica of the input, but 0.6 to 0.7 volt less positive. For this circuit, V_{in} must stay at +0.6 volt or more, or else the output will sit at ground. By returning the emitter resistor to a negative supply voltage, you can permit negative voltage swings as well. Note that there is no collector resistor in an emitter follower.

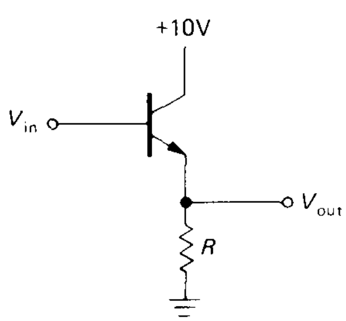


Figure 2.6. Emitter follower.

At first glance this circuit may appear useless, until you realize that the input impedance is much larger than the output impedance, as will be demonstrated shortly. This means that the circuit requires less power from the signal source to drive a given load than would be the case if the signal source were to drive the load directly. Or a signal of some internal impedance (in the Thévenin sense) can now drive a load of comparable or even lower impedance without loss of amplitude (from the usual voltage-divider effect). In other words, an emitter follower has current gain, even though it has no voltage gain. It has power gain. Voltage gain isn't everything!

Impedances of sources and loads

This last point is very important and is worth some more discussion before we calculate in detail the beneficial effects of emitter followers. In electronic circuits, you're always hooking the output of something to the input of something else, as suggested in Figure 2.7. The signal source might be the output of an amplifier stage (with Thevenin equivalent series impedance $\mathbf{Z}_{out}$), driving the next stage or perhaps a load (of some input impedance $\mathbf{Z}_{in}$). In general, the loading effect of the following stage causes a reduction of signal, as we discussed earlier in Section 1.05. For this reason it is usually best to keep $Z_{out} \ll Z_{in}$ (a factor of 10 is a comfortable rule of thumb).

In some situations it is OK to forgo this general goal of making the source stiff compared with the load. In particular, if the load is always connected (e.g., within a circuit) and if it presents a known and constant $\mathbf{Z}_{in}$, it is not too serious if it "loads" the source. However, it is always nicer if signal levels don't change when a load is connected. Also, if $\mathbf{Z}_{in}$ varies with signal level, then having a stiff source ($Z_{out} \ll Z_{in}$) assures linearity, where otherwise the level-dependent voltage divider would cause distortion.

Finally, there are two situations where $Z_{out} \ll Z_{in}$ is actually the wrong thing to

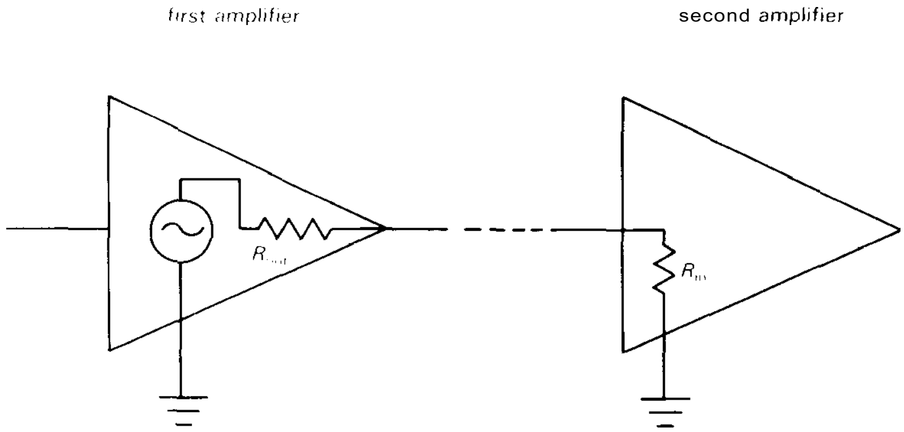


Figure 2.7. Illustrating circuit "loading" as a voltage divider.

do: In radiofrequency circuits we usually *match* impedances ($Z_{\text{out}} = Z_{\text{in}}$), for reasons we'll describe in Chapter 14. A second exception applies if the signal being coupled is a *current* rather than a voltage. In that case the situation is reversed, and one strives to make $Z_{\text{in}} \ll Z_{\text{out}}$ ($Z_{\text{out}} = \infty$, for a current source).

Input and output impedances of emitter followers

As you have just seen, the emitter follower is useful for changing impedances of signals or loads. To put it bluntly, that's the whole point of an emitter follower.

Let's calculate the input and output impedances of the emitter follower. In the preceding circuit we will consider R to be the load (in practice it sometimes is the load; otherwise the load is in parallel with R, but with R dominating the parallel resistance anyway). Make a voltage change ΔV_B at the base; the corresponding change at the emitter is $\Delta V_E = \Delta V_B$. Then the change in emitter current is

$$\Delta I_E = \Delta V_B / R$$

so

$$\Delta I_B = \frac{1}{h_{fe}+1}\Delta I_E = \frac{\Delta V_B}{R(h_{fe}+1)}$$

(using $I_E = I_C + I_B$). The input resistance is $\Delta V_B/\Delta I_B$. Therefore

$$r_{\text{in}} = (h_{fe}+1)R$$

The transistor beta (h_{fe}) is typically about 100, so a low-impedance load looks like a much higher impedance at the base; it is easier to drive.

In the preceding calculation, as in Chapter 1, we have used lower-case symbols such as h_{fe} to signify small-signal (incremental) quantities. Frequently one concentrates on the *changes* in voltages (or currents) in a circuit, rather than the steady (dc) values of those voltages (or currents). This is most common when these "small-signal" variations represent a possible signal, as in an audio amplifier, riding on a steady dc "bias" (see Section 2.05). The distinction between dc current gain (h_{FE}) and small-signal current gain (h_{fe}) isn't always made clear, and the term beta is used for both. That's alright, since $h_{fe} \approx h_{FE}$ (except at very high frequencies), and you never assume you know them accurately, anyway.

Although we used resistances in the preceding derivation, we could generalize to complex impedances by allowing ΔV_B, ΔI_B, etc., to become complex numbers. We would find that the same transformation rule applies for impedances: $Z_{in} = (h_{fe} + 1)Z_{load}$.

We could do a similar calculation to find that the output impedance Z_{out} of an emitter follower (the impedance looking into the emitter) driven from a source of internal impedance Z_{source} is given by

$$Z_{out} = \frac{Z_{source}}{h_{fe} + 1}$$

Strictly speaking, the output impedance of the circuit should also include the parallel resistance of R, but in practice Z_{out} (the impedance looking into the emitter) dominates.

EXERCISE 2.1

Show that the preceding relationship is correct. Hint: Hold the source voltage fixed, and find the change in output current for a given change in output voltage. Remember that the source voltage is connected to the base through a series resistor.

Because of these nice properties, emitter followers find application in many situations, e.g., making low-impedance signal sources within a circuit (or at outputs), making stiff voltage references from higher-impedance references (formed from voltage dividers, say), and generally isolating signal sources from the loading effects of subsequent stages.

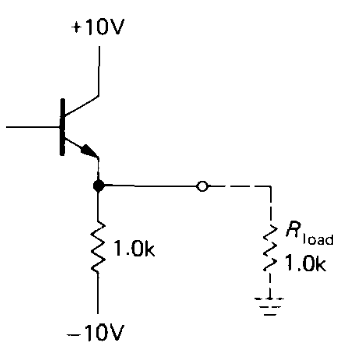


Figure 2.8. An npn emitter follower can source plenty of current through the transistor, but can sink limited current only through its emitter resistor.

EXERCISE 2.2

Use a follower with base driven from a voltage divider to provide a stiff source of +5 volts from an available regulated +15 volt supply. Load current (max) = 25mA. Choose your resistor values so that the output voltage doesn't drop more than 5% under full load.

Important points about followers

1. Notice (Section 2.01, rule **4**) that in an emitter follower the npn transistor can only "source" current. For instance, in the loaded circuit shown in Figure 2.8 the output can swing to within a transistor saturation voltage drop of V_{CC} (about +9.9V), but it cannot go more negative than −5 volts. That is because on the extreme negative swing, the transistor can do no more than turn off, which it does at **−4.4** volts input (−5V output). Further negative swing at the input results in backbiasing of the base-emitter junction, but no further change in output. The output, for a 10 volt amplitude sine-wave input, looks as shown in Figure 2.9.

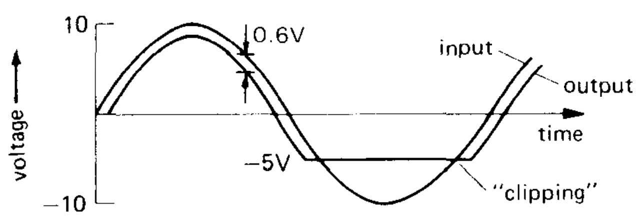


Figure 2.9. Illustrating the asymmetrical current drive capability of the npn emitter follower.

Another way to view the problem is to say that the emitter follower has low small-signal output impedance. Its large-signal output impedance is much larger (as large as R_E). The output impedance changes over from its small-signal value to its large-signal value at the point where the transistor goes out of the active region (in this case at an output voltage of -5V). To put this point another way, a low value of small-signal output impedance doesn't necessarily mean that the circuit can generate large signal swings into a low-resistance load. Low small-signal output impedance doesn't imply large output current capability.

Possible solutions to this problem involve either decreasing the value of the emitter resistor (with greater power dissipation in resistor and transistor), using a pnp transistor (if all signals are negative only), or using a "push-pull" configuration, in which two complementary transistors (one npn, one pnp), are used (Section 2.15). This sort of problem can also come up when the load of an emitter follower contains voltage or current sources of its own. This happens most often with regulated power supplies (the output is usually an emitter follower) driving a circuit that has other power supplies.

2. Always remember that the base-emitter reverse breakdown voltage for silicon transistors is small, quite often as little as *6* volts. Input swings large enough to take the transistor out of conduction can easily result in breakdown (with consequent degradation of h_{FE}) unless a protective diode is added (Fig. 2.10).

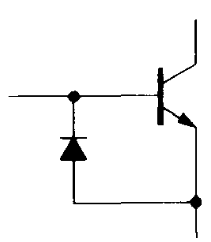


Figure 2.10. A diode prevents base-emitter reverse voltage breakdown.

3. The voltage gain of an emitter follower is actually slightly less than 1.0, because the base-emitter voltage drop is not really constant, but depends slightly on collector current. You will see how to handle that later in the chapter, when we have the Ebers-Moll equation.

2.04 Emitter followers as voltage regulators

The simplest regulated supply of voltage is simply a zener (Fig. 2.11). Some current must flow through the zener, so you choose

$$\frac{V_{in} - V_{out}}{R} > I_{out}\ (\max)$$

Because V_{in} isn't regulated, you use the lowest value of V_{in} that might occur for this formula. This is called worst-case design. In practice, you would also worry about component tolerances, line-voltage limits, etc., designing to accommodate the worst possible combination that would ever occur.

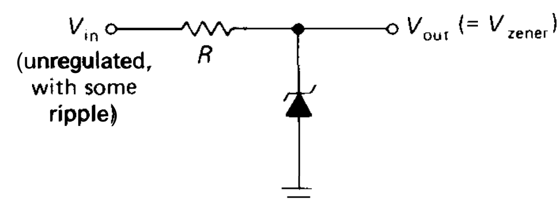


Figure 2.11. Simple zener voltage regulator.

The zener must be able to dissipate

$$P_{zener} = \left(\frac{V_{in} - V_{out}}{R} - I_{out}\right) V_{zener}$$

Again, for worst-case design, you would use V_{in} (max), R_{min}, and I_{out} (min).

EXERCISE 2.3

Design a +10 volt regulated supply for load currents from 0 to 100mA; the input voltage is +20 to +25 volts. Allow at least 10mA zener current under all (worst-case) conditions. What power rating must the zener have?

This simple zener-regulated supply is sometimes used for noncritical circuits, or circuits using little supply current. However, it has limited usefulness, for several reasons:

1. V_{out} isn't adjustable, or settable to a precise value.

2. Zener diodes give only moderate ripple rejection and regulation against changes of input or load, owing to their finite dynamic impedance.

3. For widely varying load currents a high-power zener is often necessary to handle the dissipation at low load current.

By using an emitter follower to isolate the zener, you get the improved circuit shown in Figure 2.12. Now the situation is much better. Zener current can be made relatively independent of load current, since the transistor base current is small, and far lower zener power dissipation is possible (reduced by as much as $1/h_{FE}$). The collector resistor R_C can be added to protect the transistor from momentary output short circuits by limiting the current, even though it is not essential to the emitter follower function. Choose R_C so that the voltage drop across it is less than the drop across R for the highest normal load current.

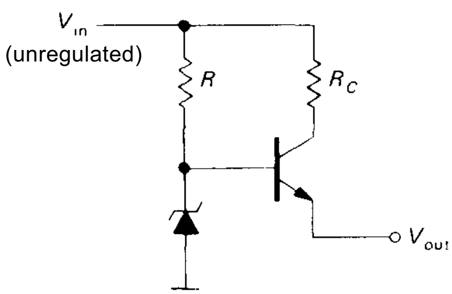


Figure 2.12. Zener regulator with follower, for increased output current. R_C protects the transistor by limiting maximum output current.

EXERCISE **2.4**

Design a +10 volt supply with the same specifications as in Exercise 2.3. Use a zener and emitter follower. Calculate worst-case dissipation in transistor and zener. What is the percentage change in zener current from the no-load condition to full load? Compare with your previous circuit.

A nice variation of this circuit aims to eliminate the effect of ripple current (through R) on the zener voltage by supplying the zener current from a current source, which is the subject of Section 2.06. An alternative method uses a low-pass filter in the zener bias circuit (Fig. 2.13). R is chosen to provide sufficient zener current. Then C is chosen large enough so that $RC \gg 1/f_{\text{ripple}}$. (In a variation of this circuit, the upper resistor is replaced by a diode.)

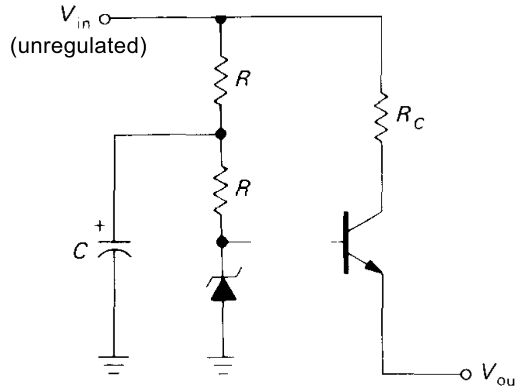


Figure 2.13. Reducing ripple in the zener regulator.

Later you will see better voltage regulators, ones in which you can vary the output easily and continuously, using feedback. They are also better voltage sources, with output impedances measured in milliohms, temperature coefficients of a few parts per million per degree centigrade, etc.

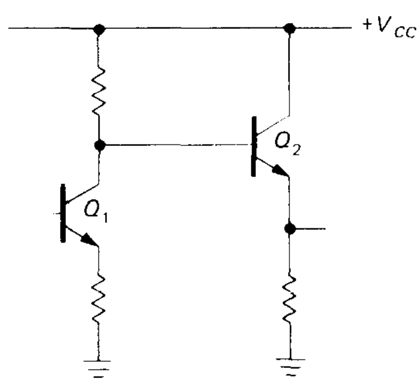


Figure 2.14

2.05 Emitter follower biasing

When an emitter follower is driven from a preceding stage in a circuit, it is usually OK to connect its base directly to the previous stage's output, as shown in Figure 2.14.

Because the signal on Q_1's collector is always within the range of the power supplies, Q_2's base will be between V_{CC} and ground, and therefore Q_2 is in the active region (neither cut off nor saturated), with its base-emitter diode in conduction and its collector at least a few tenths of a volt more positive than its emitter. Sometimes, though, the input to a follower may not be so conveniently situated with respect to the supply voltages. A typical example is a capacitively coupled (or ac-coupled) signal from some external source (e.g., an audio signal input to a high-fidelity amplifier). In that case the signal's average voltage is zero, and direct coupling to an emitter follower will give an output like that in Figure 2.15.

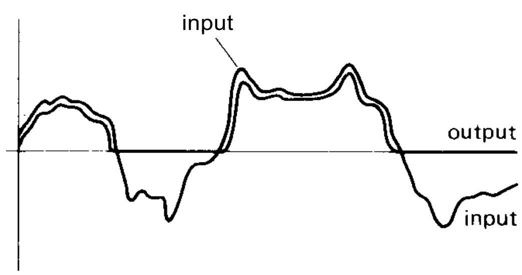


Figure 2.15. A transistor amplifier powered from a single positive supply cannot generate negative voltage swings at the transistor output terminal.

It is necessary to *bias* the follower (in fact, any transistor amplifier) so that collector current flows during the entire signal swing. In this case a voltage divider is the simplest way (Fig. 2.16). R_1 and R_2 are chosen to put the base halfway between ground and V_{CC} with no input signal, i.e., R_1 and R_2 are approximately equal. The process of selecting the operating voltages in a circuit, in the absence of applied signals, is known as setticg the *quiescent point.* In this case, as in most cases, the quiescent point is chosen to allow maximum symmetrical signal swing of the output waveform without *clipping* (flattening of the top or bottom of the waveform). What values should R_1 and R_2 have? Applying our general principle (Section 1.05), we make the impedance of the dc bias source (the impedance looking into the voltage divider) small compared with the load it drives (the dc impedance looking into the base of the follower). In this case,

$$R_1 \| R_2 \ll h_{FE} R_E$$

This is approximately equivalent to saying that the current flowing in the voltage divider should be large compared with the current drawn by the base.

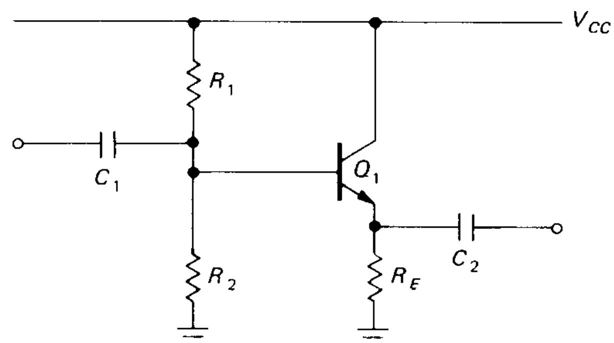


Figure 2.16. An ac-coupled emitter follower. Note base bias voltage divider.

Emitter follower design example

As an actual design example, let's make an emitter follower for audio signals (20Hz to 20kHz). V_{CC} is +15 volts, and quiescent current is to be 1mA.

Step 1. Choose V_E. For the largest possible symmetrical swing without clipping, $V_E = 0.5V_{CC}$, or +7.5 volts.

Step 2. Choose RE. For a quiescent current of 1mA, $R_E = 7.5$k.

Step 3. Choose R_1 and R_2. V_B is V_E+ 0.6, or 8.1 volts. This determines the ratio of R_1 to R_2 as 1:1.17. The preceding loading criterion requires that the parallel resistance of R_1 and R_2 be about 75k or less (one-tenth of 7.5k times h_{FE}).

Suitable standard values are $R_1 = 130\text{k}$, $R_2 = 150\text{k}$.

***Step* 4.** Choose C_1. C_1 forms a high-pass filter with the impedance it sees as a load, namely the impedance looking into the base in parallel with the impedance looking into the base voltage divider. If we assume that the load this circuit will drive is large compared with the emitter resistor, then the impedance looking into the base is $h_{FE}R_E$, about 750k. The divider looks like 70k. So the capacitor sees a load of about 63k, and it should have a value of at least $0.15\mu\text{F}$ so that the 3dB point will be below the lowest frequency of interest, 20Hz.

***Step* 5.** Choose C_2. C_2 forms a high-pass filter in combination with the load impedance, which is unknown. However, it is safe to assume that the load impedance won't be smaller than R_E, which gives a value for C_2 of at least $1.0\mu\text{F}$ to put the 3dB point below 20Hz. Because there are now two cascaded high-pass filter sections, the capacitor values should be increased somewhat to prevent large attenuation (reduction of signal amplitude, in this case 6dB) at the lowest frequency of interest. $C_1 = 0.5\mu\text{F}$ and $C_2 = 3.3\mu\text{F}$ might be good choices.

Followers with split supplies

Because signals often are "near ground," it is convenient to use symmetrical positive and negative supplies. This simplifies biasing and eliminates coupling capacitors (Fig. 2.17).

Warning: You must always provide a dc path for base bias current, even if it goes only to ground. In the preceding circuit it is assumed that the signal source has a dc path to ground. If not (e.g., if the signal is capacitively coupled), you must provide a resistor to ground (Fig. 2.18). R_B could be about one-tenth of $h_{FE}R_E$, as before.

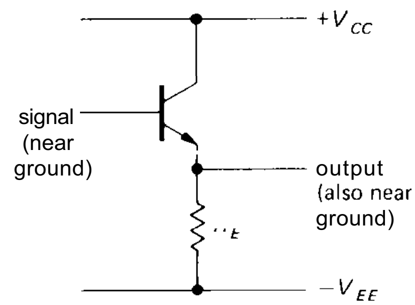


Figure 2.17. A dc-coupled emitter follower with split supply.

EXERCISE 2.5

Design an emitter follower with ±15 volt supplies to operate over the audio range (20Hz–20kHz). Use 5mA quiescent current and capacitive input coupling.

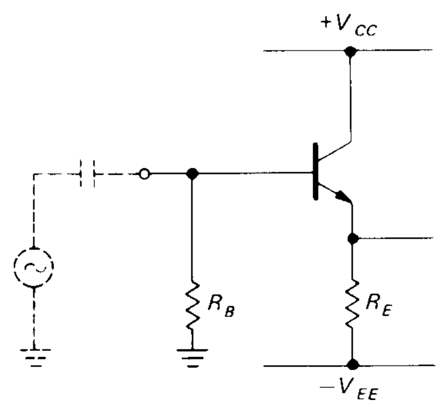


Figure 2.18

Bad biasing

Unfortunately, you sometimes see circuits like the disaster shown in Figure 2.19. R_B was chosen by assuming a particular value for h_{FE} (100), estimating the base current, and then hoping for a 7 volt drop across R_B. This is a bad design; h_{FE} is not a good parameter and will vary considerably. By using voltage biasing with a stiff voltage divider, as in the detailed example presented earlier, the quiescent point is insensitive to variations in transistor beta. For instance, in the previous design example the emitter voltage will increase by only 0.35 volt (5%) for a transistor with $h_{FE} = 200$ instead of the nominal

$h_{FE} = 100$. As with this emitter follower example, it is just as easy to fall into this trap and design bad transistor circuits in the other transistor configurations (e.g., the common-emitter amplifier, which we will treat later in this chapter).

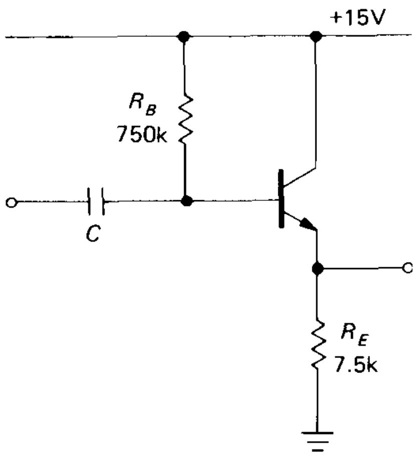


Figure 2.19. Don't do this!

2.06 Transistor current source

Current sources, although often neglected, are as important and as useful as voltage sources. They often provide an excellent way to bias transistors, and they are unequaled as "active loads" for super-gain amplifier stages and as emitter sources for differential amplifiers. Integrators, sawtooth generators, and ramp generators need current sources. They provide wide-voltage-range pull-ups within amplifier and regulator circuits. And, finally, there are applications in the outside world that require constant current sources, e.g., electrophoresis or electrochemistry.

Resistor plus voltage source

The simplest approximation to a current source is shown in Figure 2.20. As long as $R_{\text{load}} \ll R$ (in other words, $V_{\text{load}} \ll$ V), the current is nearly constant and is approximately

$$I = V/R$$

The load doesn't have to be resistive. A capacitor will charge at a constant rate, as long as $V_{\text{capacitor}} \ll$ V; this is just the first part of the exponential charging curve of an RC.

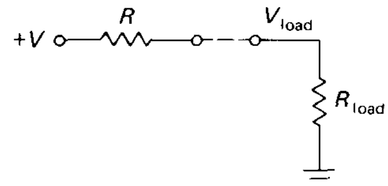


Figure 2.20

There are several drawbacks to a simple resistor current source. In order to make a good approximation to a current source, you must use large voltages, with lots of power dissipation in the resistor. In addition, the current isn't easily *programmable*, i.e., controllable over a large range via a voltage somewhere else in the circuit.

EXERCISE 2.6

If you want a current source constant to 1% over a load voltage range of 0 to +10 volts, how large a voltage source must you use in series with a single resistor?

EXERCISE 2.7

Suppose you want a 10mA current in the preceding problem. How much power is dissipated in the series resistor? How much gets to the load?

Transistor current source

Fortunately, it is possible to make a very good current source with a transistor (Fig. 2.21). It works like this : Applying V_B to the base, with $V_B > 0.6$ volt, ensures that the emitter is always conducting:

$$V_E = V_B - 0.6 \text{ volt}$$

So

$$I_E = V_E/R_E = (V_B - 0.6 \text{ volt})/R_E$$

But, since $I_E \approx I_C$ for large h_{FE},

$$I_C \approx (V_B - 0.6 \text{ volt})/R_E$$

independent of V_C, as long as the transistor is not saturated ($V_C > V_E$+ 0.2 volt).

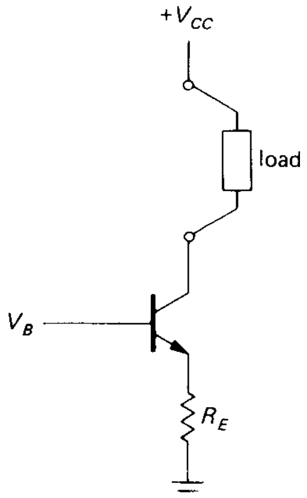


Figure 2.21. Transistor current source: basic concept.

Current-source biasing

The base voltage can be provided in a number of ways. A voltage divider is OK, as long as it is stiff enough. As before, the criterion is that its impedance should be much less than the dc impedance looking into the base ($h_{FE}R_E$). Or you can use a zener diode, biased from V_{CC}, or even a few forward-biased diodes in series from base to the corresponding emitter supply. Figure 2.22 shows some examples. In the last example (Fig. 2.22C), a *pnp* transistor *sources* current to a load returned to ground. The other examples (using *npn* transistors) should properly be called current *sinks,* but the usual practice is to call all of them current sources. ["Sink" and "source" simply refer to the direction of current flow: If a circuit *supplies* (positive) current to a point, it is a *source,* and vice versa.] In the first circuit, the voltage-divider impedance of ~1.3k is very stiff compared with the impedance looking into the base of about 100k (for h_{FE} = 100), so any changes in beta with collector voltage will not much affect the output current by causing the base voltage to change. In the other two circuits the biasing resistors are chosen to provide several milliamps to bring the diodes into conduction.

Compliance

A current source can provide constant current to the load only over some finite range of load voltage. To do otherwise would be equivalent to providing infinite power. The output voltage range over which a current source behaves well is called its output *compliance.* For the preceding transistor current sources, the compliance is set by the requirement that

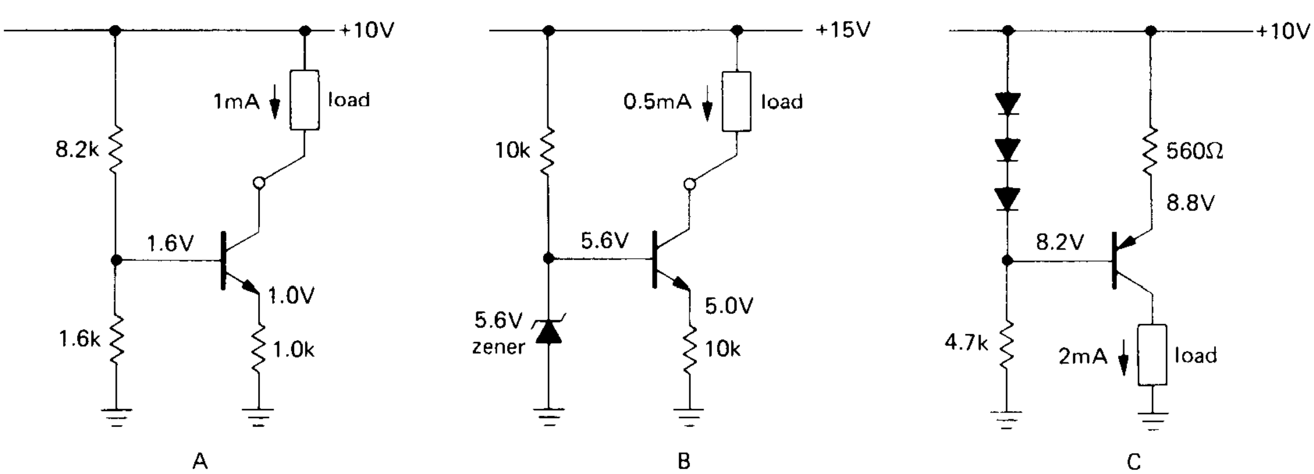


Figure 2.22. Transistor–current–source circuits, illustrating three methods of base biasing; ***npn*** transistors **sink** current, whereas *pnp* transistors ***source*** current. The circuit in C illustrates a load returned to ground.

the transistors stay in the active region. Thus in the first circuit the voltage at the collector can go down until the transistor is almost in saturation, perhaps +1.2 volts at the collector. The second circuit, with its higher emitter voltage, can sink current down to a collector voltage of about +5.2 volts.

In all cases the collector voltage can range from a value near saturation all the way up to the supply voltage. For example, the last circuit can source current to the load for any voltage between zero and about +8.6 volts across the load. In fact, the load might even contain batteries or power supplies of its own, carrying the collector beyond the supply voltage. That's OK, but you must watch out for transistor breakdown (V_{CE} must not exceed BV_{CEO}, the specified collector-emitter breakdown voltage) and also for excessive power dissipation (set by $I_C V_{CE}$). As you will see in Section 6.07, there is an additional safe-operating-area constraint on power transistors.

EXERCISE 2.8

You have +5 and +15 volt regulated supplies available in a circuit. Design a 5mA npn current source (sink) using the +5 volts on the base. What is the output compliance?

A current source doesn't have to have a fixed voltage at the base. By varying V_B you get a voltage-programmable current source. The input signal swing v_{in} (remember, lower-case symbols mean variations) must stay small enough so that the emitter voltage never drops to zero, if the output current is to reflect input voltage variations smoothly. The result will be a current source with variations in output current proportional to the variations in input voltage, $i_{\text{out}} = v_{\text{in}}/R_E$.

□ *Deficiencies of current sources*

To what extent does this kind of current source depart from the ideal? In other words, does the load current vary with voltage, i.e., have a finite ($R_{Th} < \infty$) Thévenin equivalent resistance, and if so why? There are two kinds of effects:

1. Both V_{BE} (Early effect) and h_{FE} vary slightly with collector-to-emitter voltage at a given collector current. The changes in V_{BE} produced by voltage swings across the load cause the output current to change, because the emitter voltage (and therefore the emitter current) changes, even with a fixed applied base voltage. Changes in h_{FE} produce small changes in output (collector) current for fixed emitter current, since $I_C = I_E - I_B$; in addition, there are small changes in applied base voltage produced by the variable loading of the nonzero bias source impedance as h_{FE} (and therefore the base current) changes. These effects are small. For instance, the current from the circuit in Figure 2.22A varied about 0.5% in actual measurements with a 2N3565 transistor. In particular, for load voltages varying from zero to 8 volts, the Early effect contributed 0.5%, and transistor heating effects contributed 0.2%. In addition, variations in h_{FE} contributed 0.05% (note the stiff divider). Thus these variations result in a less-than-perfect current source: The output current depends slightly on voltage and therefore has less than infinite impedance. Later you will see methods that get around this difficulty.

2. V_{BE} and also h_{FE} depend on temperature. This causes drifts in output current with changes in ambient temperature; in addition, the transistor junction temperature varies as the load voltage is changed (because of variation in transistor dissipation), resulting in departure from ideal current source behavior. The change of V_{BE} with ambient temperature can be compensated with a circuit like that shown in Figure 2.23, in which Q_2's base-emitter drop is compensated by the drop in emitter follower Q_1, with similar temperature dependence. R_3, incidentally, is a

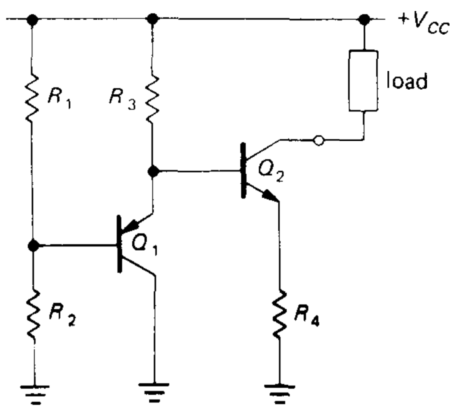


Figure 2.23. One method of **temperature-**compensating a current source.

pull-up resistor for Q_1, since Q_2's base sinks current, which Q_1 cannot source.

□ *Improving current-source performance*

In general, the effects of variability in V_{BE}, whether caused by temperature dependence (approximately $-2\text{mV/}^\circ\text{C}$) or by dependence on V_{CE} (the Early effect, given roughly by $\Delta V_{BE} \approx -0.0001\,\Delta V_{CE}$), can be minimized by choosing the emitter voltage to be large enough (at least 1V, say) so that changes in V_{BE} of tens of millivolts will not result in large fractional changes in the voltage across the emitter resistor (remember that the *base* voltage is what is held constant by your circuit). For instance, choosing $V_E = 0.1$ volt (i.e., applying about 0.7V to the base) would cause 10% variations in output current for 10mV changes in V_{BE}, whereas the choice $V_E = 1.0$ volt would result in 1% current variations for the same V_{BE} changes. Don't get carried away, though. Remember that the lower limit of output compliance is set by the emitter voltage. Using a 5 volt emitter voltage for a current source running from a +10 volt supply limits the output compliance to slightly less than 5 volts (the collector can go from about V_E+ 0.2V to V_{CC}, i.e., from 5.2V to 10V).

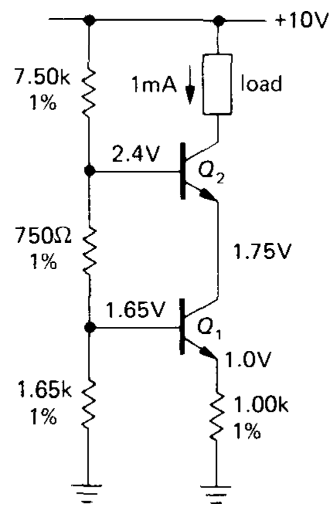


Figure 2.24. **Cascode** current source for improved current stability with load voltage variations.

Figure 2.24 shows a circuit modification that improves current-source performance significantly. Current source Q_1 functions as before, but with collector voltage held fixed by Q_2's emitter. The load sees the same current as before, since Q_2's collector and emitter currents are nearly equal (large h_{FE}). But with this circuit the V_{CE} of Q_1 doesn't change with load voltage, thus eliminating the small changes in V_{BE} from Early effect and **dissipation-**induced temperature changes. Measurements with 2N3565s gave 0.1% current variation for load voltages from 0 to 8 volts; to obtain performance of this accuracy it is important to use stable 1% resistors, as shown. (Incidentally, this circuit connection also finds use in high-frequency amplifiers, where it is known as the "cascode.") Later you will see current source techniques using op-amps and feedback that circumvent the problem of V_{BE} variation altogether.

The effects of variability of h_{FE} can be minimized by choosing transistors with large h_{FE}, so that the base current contribution to the emitter current is relatively small.

Figure 2.25 shows one last current source, whose output current doesn't depend on supply voltage. In this circuit, Q_1's V_{BE} across R_2 sets the output current, independent of V_{CC}:

$$I_{\text{out}} = V_{BE}/R_2$$

R_1 biases Q_2 and holds Q_1's collector at two diode drops below V_{CC}, eliminating Early effect as in the previous circuit. This circuit is not temperature-compensated; the voltage across R_2 decreases approximately 2.1mV/°C, causing the output current to decrease approximately 0.3%/°C.

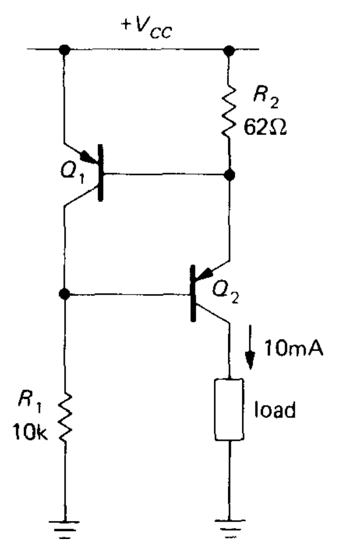


Figure 2.25. Transistor V_{BE}-referenced current source.

2.07 Common-emitter amplifier

Consider a current source with a resistor as load (Fig. 2.26). The collector voltage is

$$V_C = V_{CC} - I_C R_C$$

We could capacitively couple a signal to the base to cause the collector voltage to vary. Consider the example in Figure 2.27. C is chosen so that all frequencies of interest are passed by the high-pass filter it forms in combination with the parallel resistance of the base biasing resistors (the

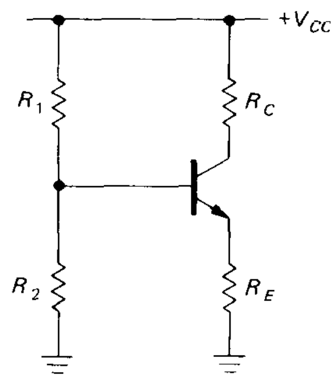


Figure 2.26

impedance looking into the base itself will usually be much larger because of the way the base resistors are chosen, and it can be ignored); that is,

$$C \geq \frac{1}{2\pi f(R_1 \| R_2)}$$

The quiescent collector current is 1.0mA because of the applied base bias and the 1.0k emitter resistor. That current puts the collector at +10 volts (+20V, minus 1.0mA through 10k). Now imagine an applied wiggle in base voltage v_B. The emitter follows with $v_E = v_B$, which causes a wiggle in emitter current

$$i_E = v_E/R_E = v_B/R_E$$

and nearly the same change in collector current (h_{fe} is large). So the initial wiggle in base voltage finally causes a collector voltage wiggle

$$v_C = -i_C R_C = -v_B(R_C/R_E)$$

Aha! It's a voltage *amplifier*, with a voltage amplification (or "gain") given by

$$\text{gain} = v_{\text{out}}/v_{\text{in}} = -R_C/R_E$$

In this case the gain is −10,000/1000, or −10. The minus sign means that a positive wiggle at the input gets turned into a negative wiggle (10 times as large) at the output. This is called a common-emitter *amplifier* with emitter degeneration.

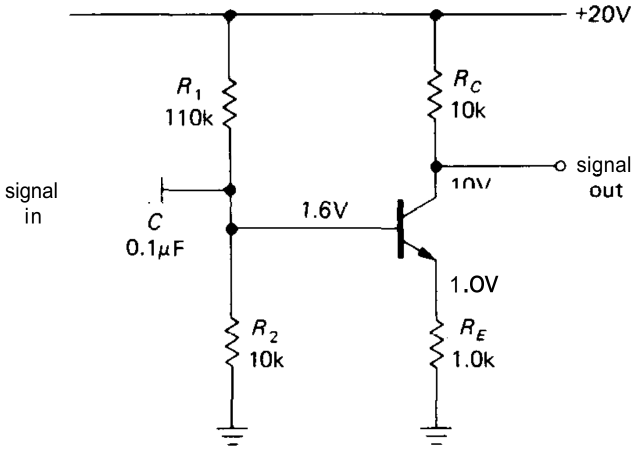


Figure 2.27. An ac common-emitter amplifier with emitter degeneration. Note that the output terminal is the collector, rather than the emitter.

Input and output impedance of the common-emitter amplifier

We can easily determine the input and output impedances of the amplifier. The input signal sees, in parallel, 110k, 10k, and the impedance looking into the base. The latter is about 100k (h_{fe} times R_E), so the input impedance (dominated by the 10k) is about 8k. The input coupling capacitor thus forms a high-pass filter, with the 3dB point at 200Hz. The signal driving the amplifier sees $0.1\mu F$ in series with 8k, which to signals of normal frequencies (well above the 3dB point) just looks like 8k.

The output impedance is 10k in parallel with the impedance looking into the collector. What is that? Well, remember that if you snip off the collector resistor, you're simply looking into a current source. The collector impedance is very large (measured in megohms), and so the output impedance is just the value of the collector resistor, 10k. It is worth remembering that the impedance looking into a transistor's collector is high, whereas the impedance looking into the emitter is low (as in the emitter follower). Although the output impedance of a common-emitter amplifier will be dominated by the collector load resistor, the output impedance of an emitter follower will not be dominated by the emitter load resistor, but rather by the impedance looking into the emitter.

2.08 Unity-gain phase splitter

Sometimes it is useful to generate a signal and its inverse, i.e., two signals 180° out of phase. That's easy to do – just use an emitter-degenerated amplifier with a gain of -1 (Fig. 2.28). The quiescent collector voltage is set to $0.75V_{CC}$, rather than the usual $0.5V_{CC}$, in order to achieve the same result – maximum symmetrical output swing without clipping at either output. The collector can swing from $0.5V_{CC}$ to V_{CC}, whereas the emitter can swing from ground to $0.5V_{CC}$.

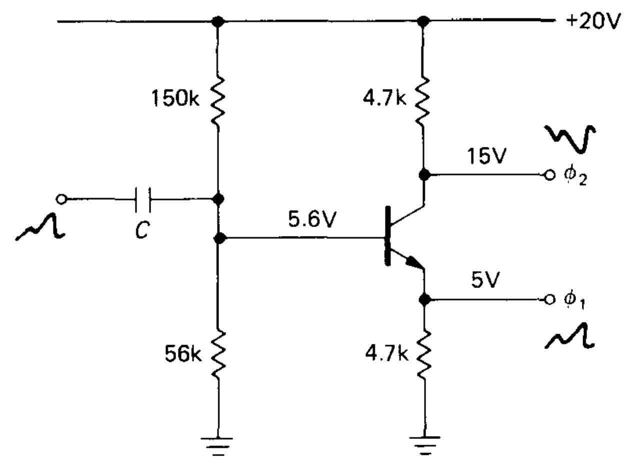


Figure 2.28. Unity-gain phase splitter.

Note that the phase-splitter outputs must be loaded with equal (or very high) impedances at the two outputs in order to maintain gain symmetry.

Phase *shifter*

A nice use of the phase splitter is shown in Figure 2.29. This circuit gives (for a sine wave input) an output sine wave of adjustable phase (from zero to 180°), but with constant amplitude. It can be best understood with a phasor diagram of voltages (see Chapter 1); representing the input signal by a unit vector along the real axis, the signals look as shown in Figure 2.30.

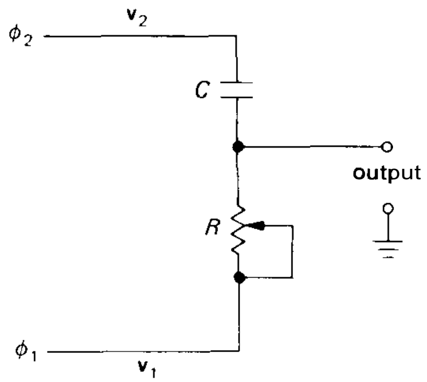


Figure 2.29. Constant-amplitude phase shifter.

Signal vectors $\mathbf{v}_R$ and $\mathbf{v}_C$ must be at right angles, and they must add to form a vector of constant length along the real axis. There is a theorem from geometry that says that the locus of such points is a circle. So the resultant vector (the output voltage) always has unit length, i.e., the same amplitude as the input, and its phase can vary from nearly zero to nearly 180° relative to the input wave as R is varied from nearly zero to a value much larger than Z_C at the operating frequency. However, note that the phase shift also depends on the frequency of the input signal for a given setting of the potentiometer R. It is worth noting that a simple RC high-pass (or low-pass) network could also be used as an adjustable phase shifter. However, its output amplitude would vary over an enormous range as the phase shift was adjusted.

An additional concern here is the ability of the phase-splitter circuit to drive the RC phase shifter as a load. Ideally, the load should present an impedance that is large compared with the collector and emitter resistors. As a result, this circuit is of limited utility where a wide range of phase shifts is required. You will see improved phase-splitter techniques in Chapter 4.

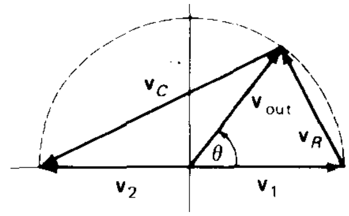


Figure 2.30. Phasor diagram for phase shifter.

2.09 Transconductance

In the preceding section we figured out the operation of the emitter-degenerated amplifier by (a) imagining an applied base voltage swing and seeing that the emitter voltage had the same swing, then (b) calculating the emitter current swing; then, ignoring the small base current contribution, we got the collector current swing and thus (c) the collector voltage swing. The voltage gain was then simply the ratio of collector (output) voltage swing to base (input) voltage swing.

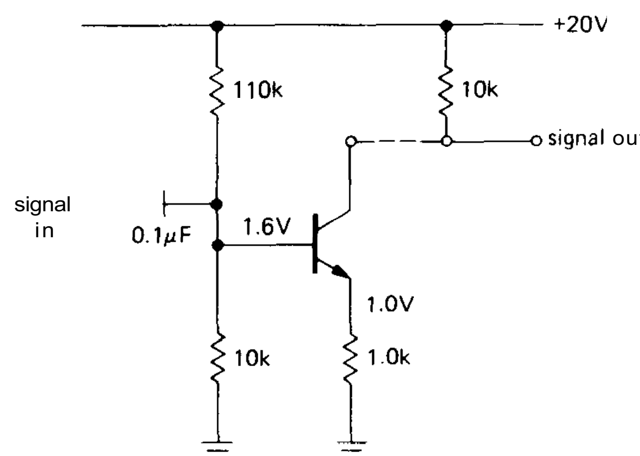


Figure 2.31. The common-emitter amplifier is a transconductance stage driving a (resistive) load.

There's another way to think about this kind of amplifier. Imagine breaking it apart, as in Figure 2.31. The first part is a voltage-controlled current source, with quiescent current of 1.0mA and gain of -1mA/V. Gain means the ratio output/input; in this case the gain has units of current/voltage, or 1/resistance. The inverse of resistance is called conductance (the inverse of reactance is susceptance, and the inverse of impedance is admittance) and has a special unit, the siemens, which used to be called the mho (ohm spelled backward). An amplifier whose gain has units of conductance is called a transconductance amplifier; the ratio I_{out}/V_{in} is called the transconductance, g_m.

Think of the first part of the circuit as a transconductance amplifier, i.e., a voltage-to-current amplifier with transconductance g_m (gain) of 1mA/V (1000μS, or 1mS, which is just $1/R_E$). The second part of the circuit is the load resistor, an "amplifier" that converts current to voltage. This resistor could be called a transresistance amplifier, and its gain (r_m) has units of voltage/current, or resistance. In this case its quiescent voltage is V_{CC}, and its gain (transresistance) is 10kV/A (10kΩ), which is just R_C. Connecting the two parts together gives you a voltage amplifier. You get the overall gain by multiplying the two gains. In this case G $= g_m R_C = R_C/R_E$, or -10, a unitless number equal to the ratio (output voltage)/(input voltage).

This is a useful way to think about an amplifier, because you can analyze performance of the sections independently. For example, you can analyze the transconductance part of the amplifier by evaluating g_m for different circuit configurations or even different devices, such as field-effect transistors (FETs). Then you can analyze the transresistance (or load) part by considering gain versus voltage swing trade-offs. If you are interested in the overall voltage gain, it is given by $G_V = g_m r_m$, where r_m is the transresistance of the load. Ultimately the substitution of an active load (current source), with its extremely high transresistance, can yield one-stage voltage gains of 10,000 or more. The *cascode* configuration, which we will discuss later, is another example easily understood with this approach.

In Chapter 4, which deals with operational amplifiers, you will see further examples of amplifiers with voltages or currents as inputs or outputs; voltage amplifiers (voltage to voltage), current amplifiers (current to current), and transresistance amplifiers (current to voltage).

Turning up the gain: limitations of the *simple* model

The voltage gain of the emitter-degenerated amplifier is $-R_C/R_E$, according to our model. What happens as R_E is reduced toward zero? The equation predicts that the gain will rise without limit. But if we made actual measurements of the preceding circuit, keeping the quiescent current constant at 1mA, we would find that the gain would level off at about 400 when R_E is zero, i.e., with the emitter grounded. We would also find that the amplifier would become significantly nonlinear (the output would not be a faithful replica of the input), the input impedance would become small and nonlinear, and the biasing would become critical and unstable with temperature. Clearly our transistor model is incomplete and needs to be modified in order to handle this circuit situation, as well as others we will talk about shortly. Our fixed-up model, which we will call the transconductance model, will be accurate enough for the remainder of the book.

EBERS-MOLL MODEL APPLIED TO BASIC TRANSISTOR CIRCUITS

2.10 Improved transistor model: transconductance amplifier

The important change is in property 4 (Section 2.01), where we said earlier that $I_C = h_{FE} I_B$. We thought of the transistor as a current amplifier whose input circuit behaved like a diode. That's roughly correct, and for some applications it's good enough. But to understand differential amplifiers, logarithmic converters, temperature compensation, and other important applications, you must think of the transistor as a *transconductance* device – collector current is determined by base-to-emitter *voltage.*

Here's the modified property 4:

4. When rules 1–3 (Section 2.01) are obeyed, I_C is related to V_{BE} by

$$I_C = I_S \left[\exp\left(\frac{V_{BE}}{V_T}\right) - 1\right]$$

where $V_T = kT/q = 25.3\text{mV}$ at room temperature (68°F, 20°C), q is the electron charge (1.60×10^{-19} coulombs), k is Boltzmann's constant (1.38×10^{-23} joules/°K), T is the absolute temperature in degrees Kelvin (°K = °C + 273.16), and I_S is the saturation current of the particular transistor (depends on T). Then the base current, which also depends on V_{BE}, can be approximated by

$$I_B = I_C/h_{FE}$$

where the "constant" h_{FE} is typically in the range 20 to 1000, but depends on transistor type, I_C, V_{CE}, and temperature. I_S represents the reverse leakage current. In the active region $I_C \gg I_S$, and therefore the −1 term can be neglected in comparison with the exponential.

The equation for I_C is known as the Ebers-Moll equation. It also approximately describes the current versus voltage for a diode, if V_T is multiplied by a correction factor m between 1 and 2. For transistors it is important to realize that the collector current is accurately determined by the base-emitter voltage, rather than by the base current (the base current is then roughly determined by h_{FE}), and that this exponential law is accurate over an enormous range of currents, typically from nanoamps to milliamps. Figure 2.32 makes the point graphically. If you measure the base current at various collector currents, you will get a graph of h_{FE} versus I_C like that in Figure 2.33.

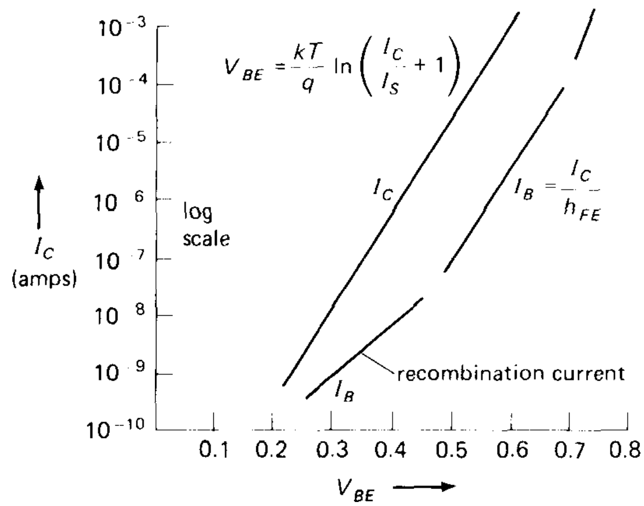


Figure 2.32. Transistor base and collector currents as functions of base-to-emitter voltage VBE.

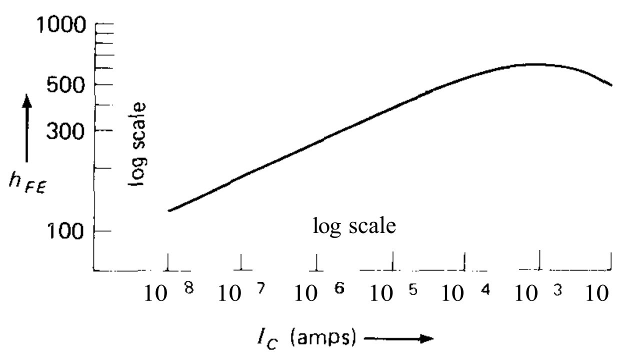


Figure 2.33. Typical transistor current gain (h_{FE}) versus collector current.

Although the Ebers-Moll equation tells us that the base-emitter voltage "programs" the collector current, this property may not be directly usable in practice (biasing a transistor by applying a base voltage) because of the large temperature coefficient of base-emitter voltage. You will see later how the Ebers-Moll equation provides insight and solutions to this problem.

Rules of thumb for transistor design

From the Ebers-Moll equation we can get

several important quantities we will be using often in circuit design:

1. The steepness of the diode curve. How much do we need to increase V_{BE} to increase I_C by a factor of 10? From the Ebers-Moll equation, that's just V_T log, 10, or 60mV at room temperature. *Base voltage increases 60mV per decade of collector current.* Equivalently, $I_C = I_{C0}e^{\Delta V/25}$, where A V is in millivolts.

2. The small-signal impedance looking into the emitter, for the base held at a fixed voltage. Taking the derivative of V_{BE} with respect to I_C, you get

$$r_e = V_T/I_C = 25/I_C \text{ ohms}$$

where I_C is in milliamps. The numerical value $25/I_C$ is for room temperature. This *intrinsic* emitter resistance, r_e, acts as if it is in series with the emitter in all transistor circuits. It limits the gain of a grounded emitter amplifier, causes an emitter follower to have a voltage gain of slightly less than unity, and prevents the output impedance of an emitter follower from reaching zero. Note that the transconductance of a grounded emitter amplifier is $g_m = 1/r_e$.

3. The temperature dependence of V_{BE}. A glance at the Ebers-Moll equation suggests that V_{BE} has a positive temperature coefficient. However, because of the temperature dependence of I_S, V_{BE} *decreases* about 2.1mV/°C. It is roughly proportional to $1/T_{abs}$, where T_{abs} is the absolute temperature.

There is one additional quantity we will need on occasion, although it is not derivable from the Ebers-Moll equation. It is the Early effect we described in Section 2.06, and it sets important limits on current-source and amplifier performance, for example:

4. Early effect. V_{BE} varies slightly with changing V_{CE} at constant I_C. This effect is caused by changing effective base width, and it is given, approximately, by

$$\Delta V_{BE} = -\alpha \Delta V_{CE}$$

where $\alpha \approx 0.0001$.

These are the essential quantities we need. With them we will be able to handle most problems of transistor circuit design, and we will have little need to refer to the Ebers-Moll equation itself.

2.11 The emitter follower revisited

Before looking again at the common-emitter amplifier with the benefit of our new transistor model, let's take a quick look at the humble emitter follower. The Ebers-Moll model predicts that an emitter follower should have nonzero output impedance, even when driven by a voltage source, because of finite r_e (item 2, above). The same effect also produces a voltage gain slightly less than unity, because r_e forms a voltage divider with the load resistor.

These effects are easy to calculate. With fixed base voltage, the impedance looking back into the emitter is just $R_{out} = dV_{BE}/dI_E$; but $I_E \approx I_C$, so $R_{out} \approx r_e$, the intrinsic emitter resistance [$r_e = 25/I_C$(mA)]. For example, in Figure 2.34A, the load sees a driving impedance of $r_e = 25$ ohms, since $I_C = 1$mA. (This is paralleled by the emitter resistor R_E, if used; but in practice R_E will always be much larger than r_e.) Figure 2.34B shows a more typical situation, with finite source resistance R_S (for simplicity we've omitted the obligatory biasing components – base divider and blocking capacitor – which are shown in Fig. 2.34C). In this case the emitter follower's output impedance is just r_e in series with $R_s/(h_{fe}+1)$ (again paralleled by an unimportant R_E, if present). For example, if R, = 1k and $I_C = 1$mA, $R_{out} = 35$ ohms (assuming $h_{fe} = 100$). It is easy to show that the intrinsic emitter r_e also figures into an emitter follower's *input* impedance, just as if it were in series with the load (actually, parallel combination of load resistor and

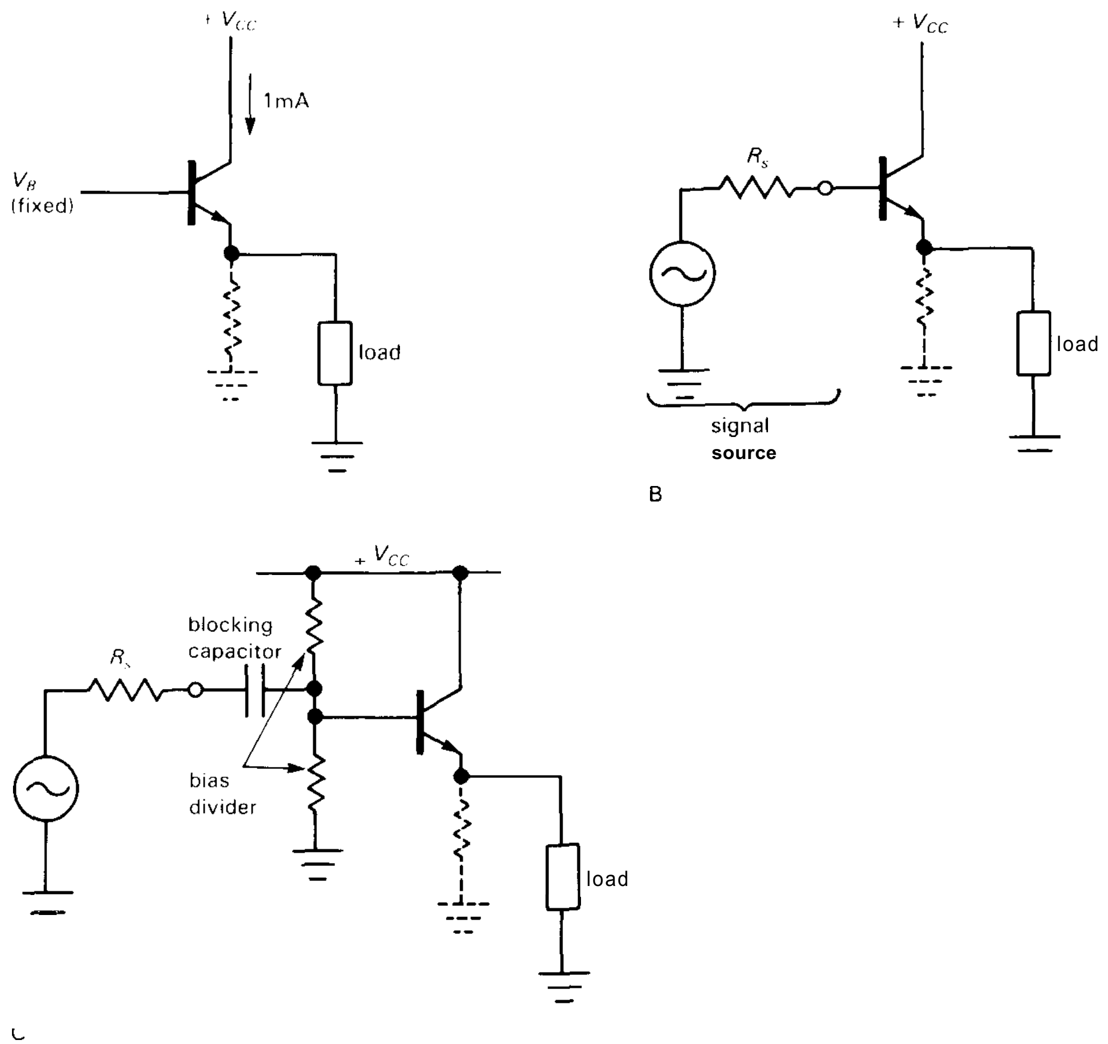


Figure 2.34

emitter resistor). In other words, for the emitter follower circuit the effect of the Ebers–Moll model is simply to add a series emitter resistance r_e to our earlier results.

The voltage gain of an emitter follower is slightly less than unity, owing to the voltage divider produced by r_e and the load. It is simple to calculate, because the output is at the junction of r_e and R_{load}: $G_V = v_{\text{out}}/v_{\text{in}} = R_L/(r_e + R_L)$. Thus, for example, a follower running at 1mA quiescent current, with 1k load, has a voltage gain of 0.976. Engineers sometimes like to write the gain in terms of the transconductance, to put it in a form that holds for FETs also (see Section 3.07); in that case (using g $= 1/r_e$) you get $G_V = R_L g_m/(1 + R_L g_m)$.

2.12 The common-emitter amplifier revisited

Previously we got wrong answers for the voltage gain of the common-emitter amplifier with emitter resistor (sometimes called emitter degeneration) when we set the emitter resistor equal to zero. The problem is that the transistor has $25/I_C$(mA) ohms of built-in (intrinsic) emitter resistance r_e that must be added to the actual external emitter resistor. This resistance is significant only when small emitter resistors (or none at all) are used. So, for instance, the amplifier we considered previously will have a voltage gain of $-10\text{k}/r_e$, or -400, when the external emitter resistor is zero. The input impedance is not zero, as we would have predicted earlier ($h_{fe}R_E$); it is approximately $h_{fe}r_e$, or in this case (1mA quiescent current) about 2.5k.

The terms "grounded emitter" and "common emitter" are sometimes used interchangeably, and they can be confusing. We will use the phrase "grounded emitter amplifier" to mean a common-emitter amplifier with $R_E = 0$. A common-emitter amplifier stage may have an emitter resistor; what matters is that the emitter circuit is common to the input circuit and the output circuit.

Shortcomings of the single-stage grounded emitter amplifier

The extra voltage gain you get by using $R_E = 0$ comes at the expense of other properties of the amplifier. In fact, the grounded emitter amplifier, in spite of its popularity in textbooks, should be avoided except in circuits with overall negative feedback. In order to see why, consider Figure 2.35.

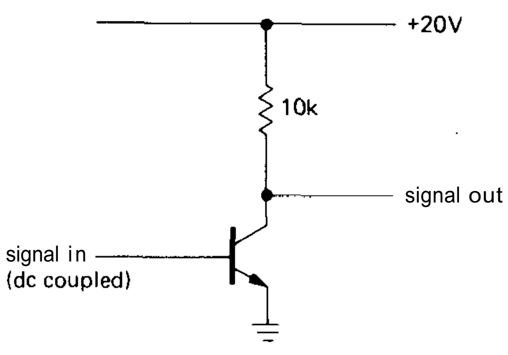


Figure 2.35. Common-emitter amplifier without emitter degeneration.

1. Nonlinearity. The gain is $G = -g_m R_C = -R_C/r_e = -R_C I_C(\text{mA})/25$, so for a quiescent current of 1mA, the gain is -400. But I_C varies as the output signal varies. For this example, the gain will vary from -800 ($V_{out} = 0$, $I_C = 2\text{mA}$) down to zero ($V_{out} = V_{CC}$, $I_C = 0$). For a triangle-wave input, the output will look like that in Figure 2.36. The amplifier has high distortion, or poor linearity. The grounded emitter amplifier without feedback is useful only for small signal swings about the quiescent point. By contrast, the emitter-degenerated amplifier has gain almost entirely independent of collector current, as long as $R_E \gg r_e$, and can be used for undistorted amplification even with large signal swings.

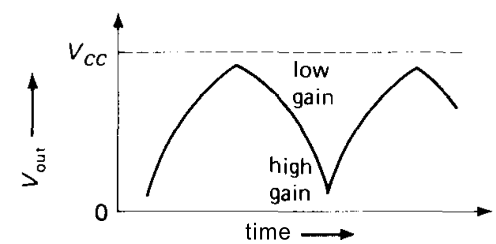


Figure 2.36. Nonlinear output waveform from grounded emitter amplifier.

2. Input impedance. The input impedance is roughly $Z_{in} = h_{fe}r_e = 25\ h_{fe}/I_C(\text{mA})$ ohms. Once again, I_C varies over the signal swing, giving a varying input impedance. Unless the signal source driving the base has low impedance, you will wind up with nonlinearity due to the nonlinear variable voltage divider formed from the signal source and the amplifier's input impedance. By contrast, the input impedance of an emitter-degenerated amplifier is constant and high.

3. *Biasing.* The grounded emitter amplifier is difficult to bias. It might be tempting just to apply a voltage (from a voltage divider) that gives the right quiescent current according to the Ebers-Moll equation. That won't work, because of the temperature dependence of V_{BE} (at fixed I_C), which varies about 2.1mV/°C (it actually decreases with increasing T because of the variation of I_S with T; as a result, V_{BE} is roughly proportional to $1/T$, the absolute temperature). This means that the collector current (for fixed V_{BE}) will increase by a factor of 10 for a 30°C rise

in temperature. Such unstable biasing is useless, because even rather small changes in temperature will cause the amplifier to saturate. For example, a grounded emitter stage biased with the collector at half the supply voltage will go into saturation if the temperature rises by 8°C.

EXERCISE 2.9

Verify that an 8°C rise in ambient temperature will cause a base-voltage-biased grounded emitter stage to saturate, assuming that it was initially biased for $V_C = 0.5V_{CC}$.

Some solutions to the biasing problem will be discussed in the following sections. By contrast, the emitter-degenerated amplifier achieves stable biasing by applying a voltage to the base, most of which appears across the emitter resistor, thus determining the quiescent current.

Emitter resistor as feedback

Adding an external series resistor to the intrinsic emitter resistance r_e (emitter degeneration) improves many properties of the common-emitter amplifier, at the expense of gain. You will see the same thing happening in Chapters 4 and 5 , when we discuss *negative feedback,* an important technique for improving amplifier characteristics by feeding back some of the output signal to reduce the effective input signal. The similarity here is no coincidence; the emitter-degenerated amplifier itself uses a form of negative feedback. Think of the transistor as a transconductance device, determining collector current (and therefore output voltage) according to the voltage applied between the base and emitter; but the input to the amplifier is the voltage from base to ground. So the voltage from base to emitter is the input voltage, *minus a sample of the output* ($I_E R_E$). That's negative feedback, and that's why emitter degeneration improves most properties of the amplifier (improved linearity and stability and increased input impedance; also the output impedance would be reduced if the feedback were taken directly from the collector). Great things to look forward to in Chapters 4 and 5!

2.13 Biasing the common-emitter amplifier

If you must have the highest possible gain (or if the amplifier stage is inside a feedback loop), it is possible to arrange successful biasing of a common-emitter amplifier. There are three solutions that can be applied, separately or in combination: bypassed emitter resistor, matched biasing transistor, and dc feedback.

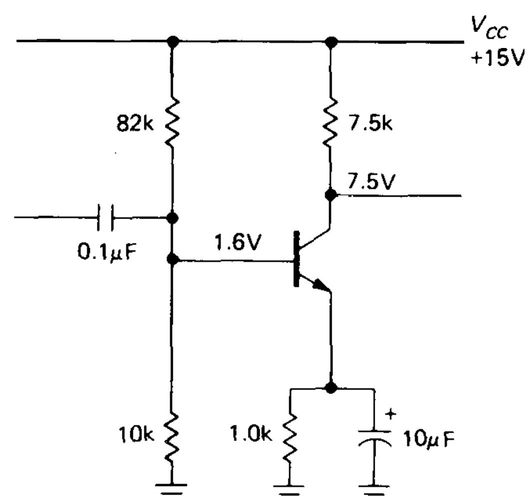


Figure 2.37. A bypassed emitter resistor can be used to improve the bias stability of a grounded emitter amplifier.

Bypassed emitter resistor

Use a bypassed emitter resistor, biasing as for the degenerated amplifier, as shown in Figure 2.37. In this case R_E has been chosen about $0.1R_C$, for ease of biasing; if R_E is too small, the emitter voltage will be much smaller than the base-emitter drop, leading to temperature instability of the quiescent point as V_{BE} varies with temperature. The emitter bypass capacitor is chosen by making its impedance small compared with r_e (not R_E) at the lowest frequency of interest. In this case its impedance is 25 ohms at 650Hz. At signal frequencies the input coupling capacitor sees an impedance of 10k in parallel with the base impedance, in this case h_{fe} times 25 ohms, or roughly 2.5k. At dc, the impedance looking into the base is much larger (h_{fe} times the emitter resistor, or about 100k), which is why stable biasing is possible.

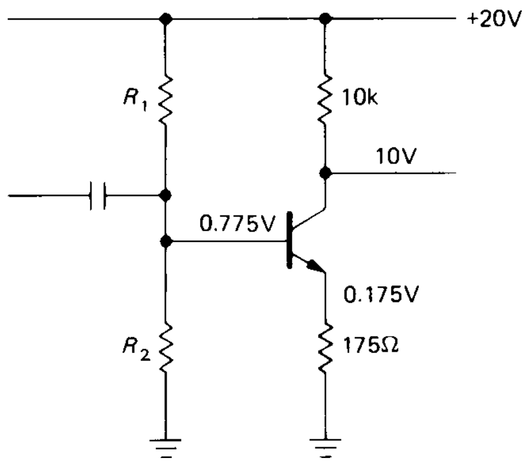


Figure 2.38

A variation on this circuit consists of using two emitter resistors in series, one of them bypassed. For instance, suppose you want an amplifier with a voltage gain of 50, quiescent current of 1mA, and V_{CC} of +20 volts, for signals from 20Hz to 20kHz. If you try to use the emitter-degenerated circuit, you will have the circuit shown in Figure 2.38. The collector resistor is chosen to put the quiescent collector voltage at $0.5V_{CC}$. Then the emitter resistor is chosen for the required gain, including the effects of the r_e of $25/I_C$(mA). The problem is that the emitter voltage of only 0.175 volt will vary significantly as the ~0.6 volt of base-emitter drop varies with temperature (−2.1mV/°C, approximately), since the base is held at constant voltage by R_1 and R_2; for instance, you can verify that an increase of 20°C will cause the collector current to increase by nearly 25%.

The solution here is to add some bypassed emitter resistance for stable biasing, with no change in gain at signal frequencies (Fig. 2.39). As before, the collector resistor is chosen to put the collector at 10 volts ($0.5V_{CC}$). Then the unbypassed emitter resistor is chosen to give a gain of 50, including the intrinsic emitter resistance r, = $25/I_C$(mA). Enough bypassed emitter resistance is added to make stable biasing possible (one-tenth of the collector resistance is a good rule). The base voltage is chosen to give 1mA of emitter current, with impedance about one-tenth the dc impedance looking into the base (in this case about 100k). The emitter bypass capacitor is chosen to have low impedance compared with 180+25 ohms at the lowest signal frequencies. Finally, the input coupling capacitor is chosen to have low impedancc compared with the signal-frequency input impedance of the amplifier, which is equal to the voltage divider impedance in parallel with $(180 + 25)h_{fe}$ ohms (the 820Ω is bypassed, and looks like a short at signal frequencies).

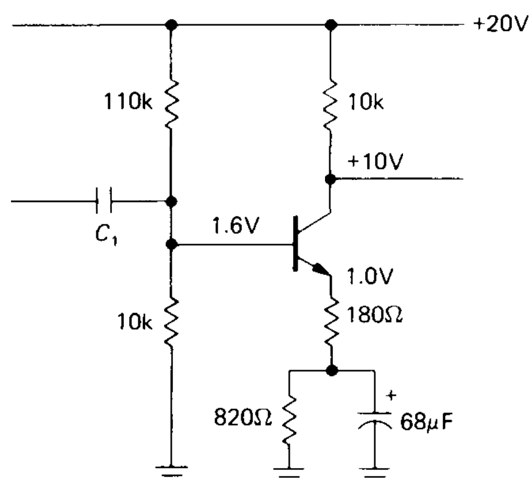


Figure 2.39. A common-emitter amplifier combining bias stability, linearity, and large voltage gain.

An alternative circuit splits the signal and dc paths (Fig. 2.40). This lets you vary the gain (by changing the 180Ω resistor) without bias change.

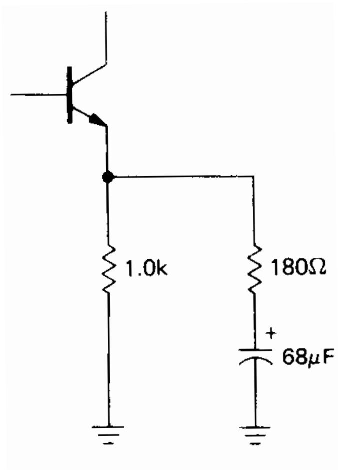


Figure **2.40.** Equivalent emitter circuit for Figure **2.39.**

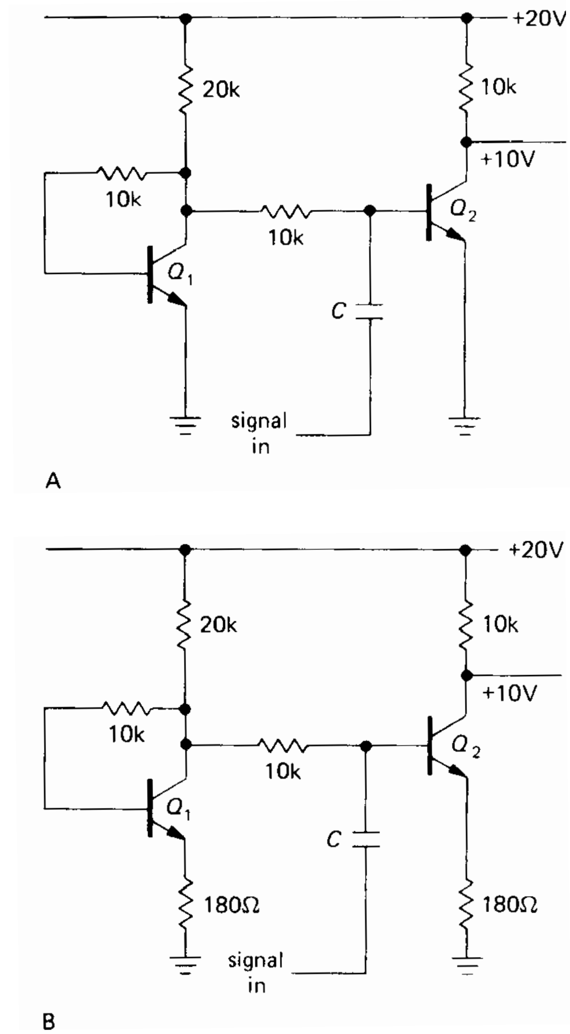


Figure **2.41.** Biasing scheme with compensated V_{BE} drop.

□ *Matched biasing transistor*

Use a matched transistor to generate the correct base voltage for the required collector current; this ensures automatic temperature compensation (Fig. 2.41). Q_1's collector is drawing 1mA, since it is guaranteed to be near ground (about one V_{BE} drop above ground, to be exact); if Q_1 and Q_2 are a matched pair (available as a single device, with the two transistors on one piece of silicon), then Q_2 will also be biased to draw 1mA, putting its collector at +10 volts and allowing a full ±10 volt symmetrical swing on its collector. Changes in temperature are of no importance, as long as both transistors are at the same temperature. This is a good reason for using a "monolithic" dual transistor.

Feedback at dc

Use dc feedback to stabilize the quiescent point. Figure 2.42 shows one method. By taking the bias voltage from the collector, rather than from V_{CC}, you get some measure of bias stability. The base sits one diode drop above ground; since its bias comes from a 10:1 divider, the collector is at 11 diode drops above ground, or about 7 volts. Any tendency for the transistor

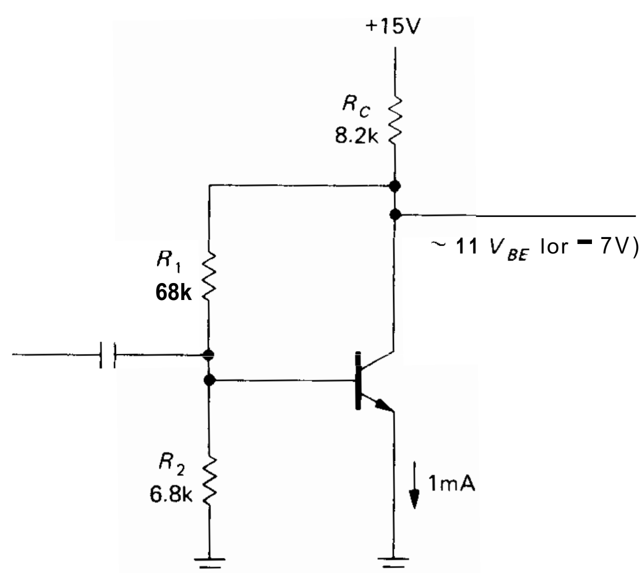


Figure **2.42.** Bias stability is improved by feedback.

to saturate (e.g., if it happens to have unusually high beta) is stabilized, since the dropping collector voltage will reduce the base bias. This scheme is acceptable if great stability is not required. The quiescent point is liable to drift a volt or so as the ambient (surrounding) temperature changes, since the base-emitter voltage has a significant temperature coefficient. Better stability is possible if several stages of amplification are included within the feedback loop. You will see examples later in connection with feedback.

A better understanding of feedback is really necessary to understand this circuit. For instance, feedback acts to reduce the input and output impedances. The input signal sees R_1's resistance effectively reduced by the voltage gain of the stage. In this case it is equivalent to a resistor of about 300 ohms to ground. In Chapter 4 we will treat feedback in enough detail so that you will be able to figure the voltage gain and terminal impedance of this circuit.

Note that the base bias resistor values could be increased in order to raise the input impedance, but you should then take into account the non-negligible base current. Suitable values might be $R_1 = 220\text{k}$ and $R_2 = 33\text{k}$. An alternative approach might be to bypass the feedback resistance in order to eliminate feedback (and therefore lowered input impedance) at signal frequencies (Fig. 2.43).

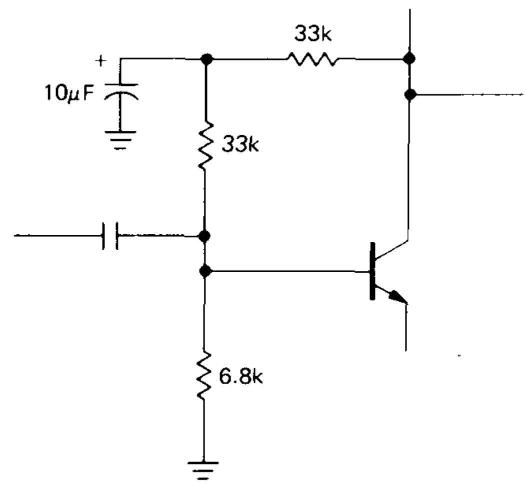


Figure 2.43. Eliminating feedback at signal frequencies.

Comments on biasing and gain

One important point about grounded emitter amplifier stages: You might think that the voltage gain can be raised by increasing the quiescent current, since the intrinsic emitter resistance r_e drops with rising current. Although r_e does go down with increasing collector current, the smaller collector resistor you need to obtain the same quiescent collector voltage just cancels the advantage. In fact, you can show that the small-signal voltage gain of a grounded emitter amplifier biased to $0.5V_{CC}$ is given by $G = 20V_{CC}$, independent of quiescent current.

EXERCISE 2.10
Show that the preceding statement is true.

If you need more voltage gain in one stage, one approach is to use a current source as an *active* load. Since its impedance is very high, single-stage voltage gains of 1000 or more are possible. Such an arrangement cannot be used with the biasing schemes we have discussed, but must be part of an overall dc feedback loop, a subject we will discuss in the next chapter. You should be sure such an amplifier looks into a high-impedance load; otherwise the gain obtained by high collector load impedance will be lost. Something like an emitter follower, a field-effect transistor (FET), or an op-amp presents a good load.

In radiofrequency amplifiers intended for use only over a narrow frequency range, it is common to use a parallel LC circuit as a collector load; in that case very high voltage gain is possible, since the LC circuit has high impedance (like a current source) at the signal frequency, with low impedance at dc. Since the LC

is "tuned," out-of-band interfering signals (and distortion) are effectively rejected. Additional bonuses are the possibility of peak-to-peak output swings of $2V_{CC}$ and the use of transformer coupling from the inductor.

EXERCISE 2.11

Design a tuned common-emitter amplifier stage to operate at 100kHz. Use a bypassed emitter resistor, and set the quiescent current at 1.0mA. Assume $V_{CC} = +15$ volts and L = 1.0mH, and put a 6.2k resistor across the LC to set Q = 10 (to get a 10% bandpass; see Section 1.22). Use capacitive input coupling.

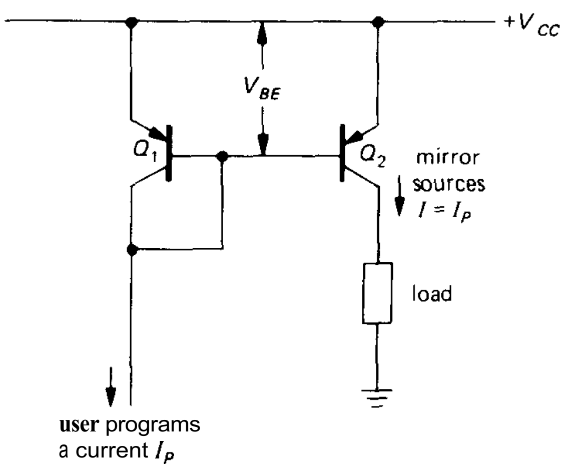


Figure 2.44. Classic bipolar-transistor matched-pair current mirror. Note the common convention of referring to the positive supply as V_{CC}, even when pnp transistors are used.

2.14 Current mirrors

The technique of matched base-emitter biasing can be used to make what is called a ***current mirror*** (Fig. *2.44)*. You "program" the mirror by sinking a current from Q_1's collector. That causes a V_{BE} for Q_1 appropriate to that current at the circuit temperature and for that transistor type. Q_2, matched to Q_1 (a monolithic dual transistor is ideal), is thereby programmed to source the same current to the load. The small base currents are unimportant.

One nice feature of this circuit is voltage compliance of the output transistor current source to within a few tenths of a volt of V_{CC}, since there is no emitter resistor drop to contend with. Also, in many applications it is handy to be able to program a current with a current. An easy way to generate the control current I_P is with a resistor (Fig. 2.45). Since the bases are a diode drop below V_{CC}, the 14.4k resistor produces a control current, and therefore an output current, of 1mA. Current mirrors can be used in transistor circuits whenever a current source is needed. They're very popular in integrated circuits, where (a) matched transistors abound and (b) the designer tries to make circuits that will work over a large range of supply voltages. There are even resistorless integrated circuit op-amps in which the operating current of the whole amplifier is set by one external resistor, with all the quiescent currents of the individual amplifier stages inside being determined by current mirrors.

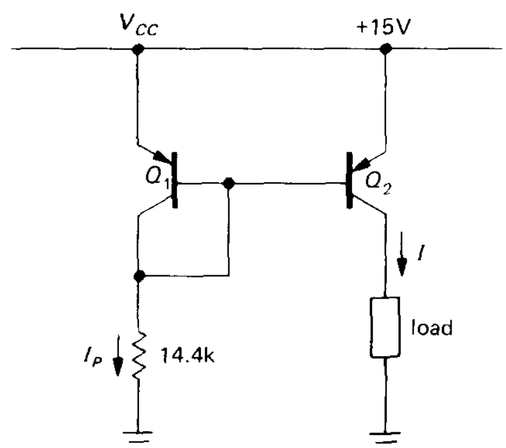


Figure 2.45

Current mirror limitations due to Early effect

One problem with the simple current mirror is that the output current varies a bit with changes in output voltage, i.e., the output impedance is not infinite. This is because of the slight variation of V_{BE} with collector voltage at a given current in Q_2 (due to Early effect); in other words, the curve of collector current versus collector-emitter voltage at a fixed base-emitter voltage is not flat (Fig. 2.46). In practice, the

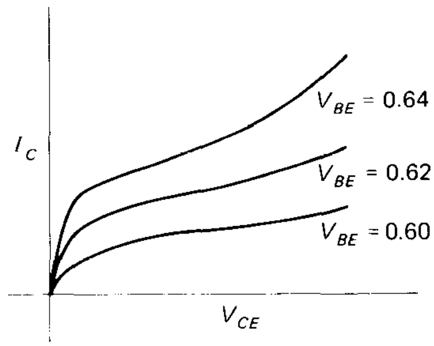


Figure 2.46

current might vary 25% or so over the output compliance range – much poorer performance than the current source with emitter resistor discussed earlier.

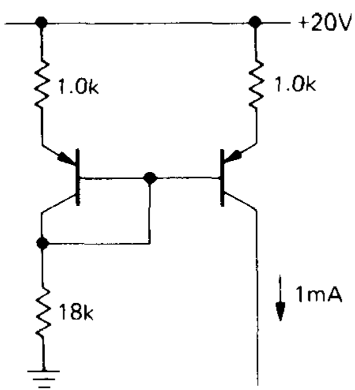


Figure 2.47. Improved current mirror.

One solution, if a better current source is needed (it often isn't), is the circuit shown in Figure 2.47. The emitter resistors are chosen to have at least a few tenths of a volt drop; this makes the circuit a far better current source, since the small variations of V_{BE} with V_{CE} are now negligible in determining the output current. Again, matched transistors should be used.

Wilson mirror

Another current mirror with very constant current is shown in the clever circuit of Figure 2.48. Q_1 and Q_2 are in the usual mirror configuration, but Q_3 now keeps Q_1's collector fixed at two diode drops

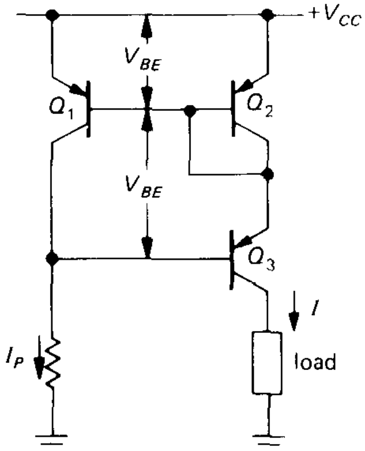


Figure 2.48. Wilson current mirror. Good stability with load variations is achieved through cascode transistor Q_3, which reduces voltage variations across Q_1.

below V_{CC}. That circumvents the Early effect in Q_1, whose collector is now the programming terminal, with Q_2 now sourcing the output current. Q_3 does not affect the balance of currents, since its base current is negligible; its only function is to pin Q_1's collector. The result is that both current-determining transistors (Q_1 and Q_2) have fixed collector-emitter drops; you can think of Q_3 as simply passing the output current through to a variable-voltage load (a similar trick is used in the cascode connection, which you will see later in the chapter). Q_3, by the way, does not have to be matched to Q_1 and Q_2.

Multiple outputs and current ratios

Current mirrors can be expanded to source (or sink, with *npn* transistors) current to several loads. Figure 2.49 shows the idea. Note that if one of the current source transistors saturates (e.g., if its load is disconnected), its base robs current from the shared base reference line, reducing the other output currents. The situation is

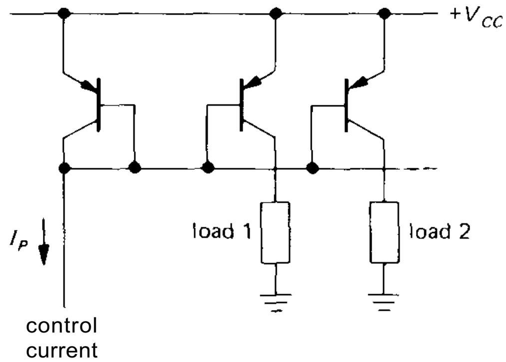


Figure 2.49. Current mirror with multiple outputs. This circuit is commonly used to obtain multiple programmable current sources.

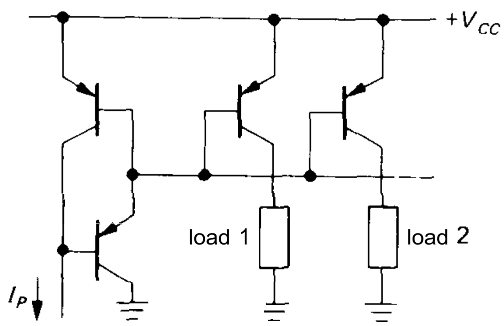


Figure 2.50

rescued by adding another transistor (Fig. 2.50).

Figure 2.51 shows two variations on the multiple-mirror idea. These circuits mirror twice (or half) the control current. In the design of integrated circuits, current mirrors with any desired current ratio can be made by adjusting the size of the emitter junctions appropriately.

Texas Instruments offers complete monolithic Wilson current mirrors in convenient TO-92 transistor packages. Their TL011 series includes 1:1, 1:2, 1:4, and 2:1 ratios, with output compliance from 1.2 to 40 volts. The Wilson configuration gives good current source performance – at constant programming current the output current increases by only 0.05% per volt – and they are very inexpensive (50 cents or less). Unfortunately, these useful devices are available in npn polarity only.

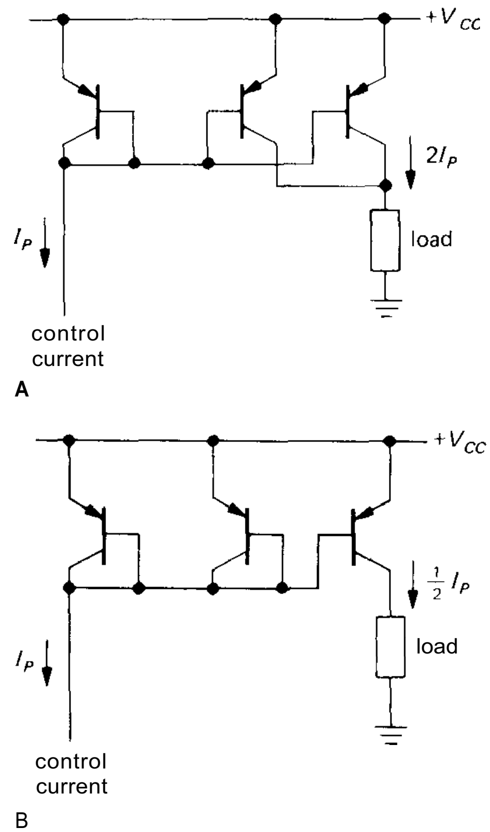


Figure 2.51. Current mirrors with current ratios other than 1:1.

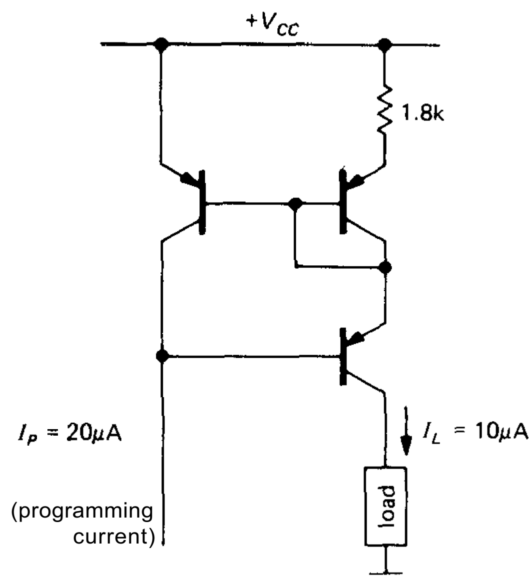


Figure 2.52. Modifying current-source output with an emitter resistor. Note that the output current is no longer a simple multiple of the progamming current.

Another way to generate an output current that is a fraction of the programming current is to add a resistor in the emitter circuit of the output transistor (Fig. 2.52). In any circuit where the transistors are operating at different current densities, the Ebers-Moll equation predicts that the difference in V_{BE} depends only on the ratio of the current densities. For matched transistors, the ratio of collector currents equals the ratio of current densities. The graph in Figure 2.53 is handy for determining the difference in base-emitter drops in such a situation. This makes it easy to design a "ratio mirror."

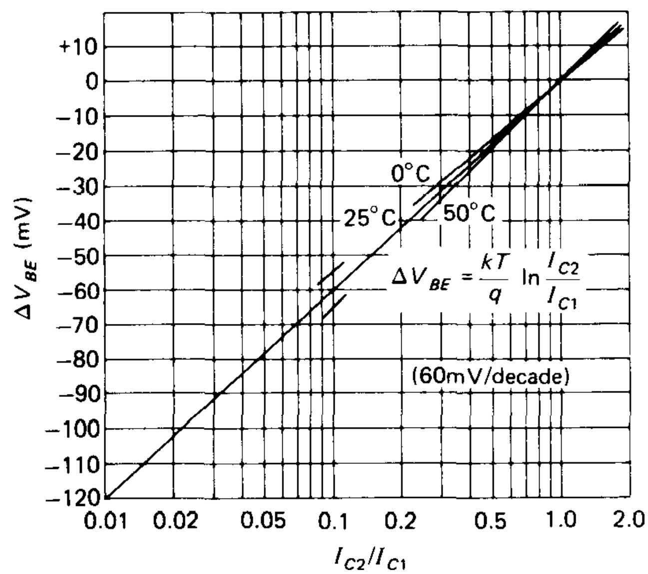


Figure 2.53. Collector current ratios for matched transistors as determined by the difference in applied base-emitter voltages.

EXERCISE 2.12
Show that the ratio mirror in Figure 2.52 works as advertised.

SOME AMPLIFIER BUILDING BLOCKS

□ 2.15 Push-pull output stages

As we mentioned earlier in the chapter, an *npn* emitter follower cannot sink current, and a *pnp* follower cannot source current. The result is that a single-ended follower operating between split supplies can drive a ground-returned load only if a high quiescent current is used (this is sometimes called a class A amplifier). The quiescent current must be at least as large as the maximum output current during peaks of the waveform, resulting in high quiescent power dissipation. For instance, Figure 2.54 shows a follower circuit to drive an 8 ohm load with up to 10 watts of audio. The *pnp* follower Q_1 is included to reduce drive requirements and to cancel Q_2's V_{BE} offset (zero volts input gives zero volts output). Q_1 could, of course, be omitted for simplicity. The hefty current source in Q_1's emitter load is used to ensure that there is sufficient base drive to Q_2 at the top of the signal swing. A resistor as emitter load would be inferior because it would have to be a rather low value (50Ω or less) in order to guarantee at least 50mA of base drive to Q_2 at the peak of the swing, when load current would be maximum and the drop across the resistor would be minimum; the resultant quiescent current in Q_1 would be excessive.

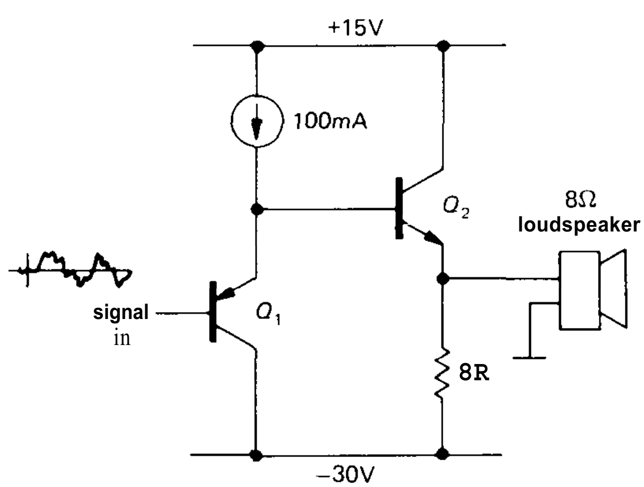


Figure 2.54. A 10 watt loudspeaker amplifier, built with a single-ended emitter follower, dissipates 165 watts of quiescent power!

The output of this example circuit can swing to nearly ±15 volts (peak) in both directions, giving the desired output power (9V rms across 8Ω). However, the output transistor dissipates 55 watts with no signal, and the emitter resistor dissipates another 110 watts. Quiescent power dissipation many times greater than the maximum output power is characteristic of this kind of class A circuit (transistor always in conduction); this obviously leaves a lot to be desired in applications where any significant amount of power is involved.

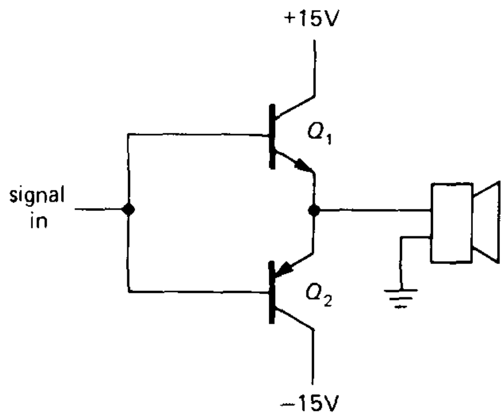


Figure 2.55. Push-pull emitter follower.

Figure 2.55 shows a push-pull follower to do the same job. Q_1 conducts on positive swings, Q_2 on negative swings. With zero input voltage, there is no collector current and no power dissipation. At 10 watts output power there is less than 10 watts dissipation in each transistor.

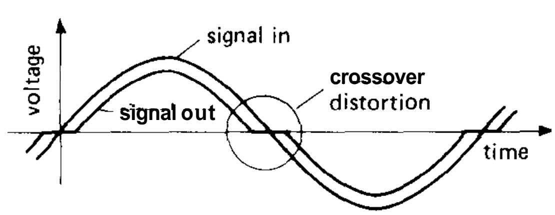


Figure 2.56. Crossover distortion in the push-pull follower.

□ *Crossover distortion in push-pull stages*

There is a problem with the preceding circuit as drawn. The output trails the input by a V_{BE} drop; on positive swings the output is about 0.6 volt less positive than the input, and the reverse for negative swings. For an input sine wave, the output would look as shown in Figure 2.56. In the language of the audio business, this is called crossover distortion. The best cure (feedback offers another method, although it is not entirely satisfactory) is to bias the push-pull stage into slight conduction, as in Figure 2.57.

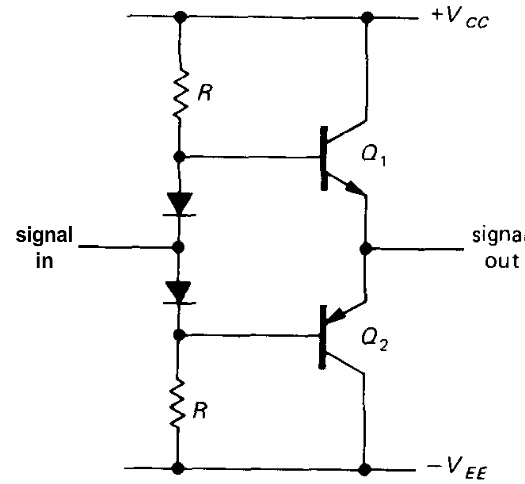


Figure 2.57. Biasing the push-pull follower to eliminate crossover distortion.

The bias resistors R bring the diodes into forward conduction, holding Q_1's base a diode drop above the input signal and Q_2's base a diode drop below the input signal. Now, as the input signal crosses through zero, conduction passes from Q_2 to Q_1; one of the output transistors is always on. R is chosen to provide enough base current for the output transistors at the peak output swing. For instance, with ±20 volt supplies and an 8 ohm load running up to 10 watts sine-wave power, the peak base voltage is about 13.5 volts, and the peak load current is about 1.6 amps. Assuming a transistor beta of 50 (power transistors generally have lower current gain than small-signal transistors), the 32mA of necessary base current will require base resistors of about 220 ohms (6.5V from V_{CC} to base at peak swing).

Thermal stability in class B push-pull amplifiers

The preceding amplifier (sometimes called a class B amplifier, meaning that each transistor conducts over half the cycle) has one bad feature: It is not thermally stable. As the output transistors warm up (and they will get hot, because they are dissipating power when signal is applied), their V_{BE} drops, and quiescent collector current begins to flow. The added heat this produces causes the situation to get worse, with the strong possibility of what is called *thermal runaway* (whether it runs away or not depends on a number of factors, including how large a "heat sink" is used, how well the diode temperature tracks the transistor, etc.). Even without runaway, better control over the circuit is needed, usually with the sort of arrangement shown in Figure 2.58.

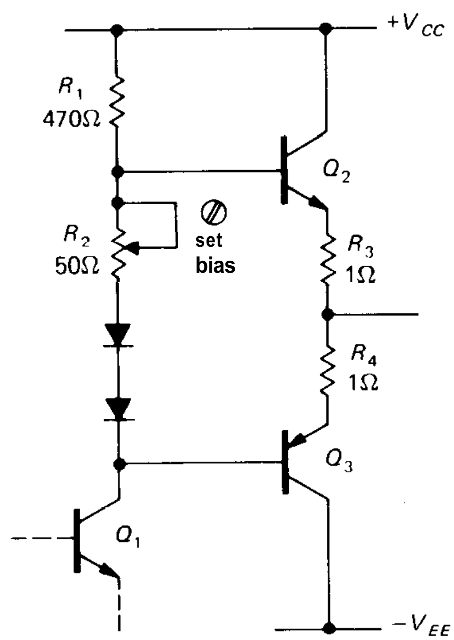


Figure 2.58. Small emitter resistors improve thermal stability in the push-pull follower.

For variety, the input is shown coming from the collector of the previous stage; R_1 now serves the **dual** purpose of being Q_1's collector resistor and providing current to bias the diodes and bias-setting resistor in the push-pull base circuit. Here R_3 and R_4, typically a few ohms or less, provide a "cushion" for the critical quiescent current biasing: The voltage between the bases of the output transistors must now be a bit greater than two diode drops, and you provide the extra with adjustable biasing resistor R_2 (often replaced by a third series diode). With a few tenths of a volt across R_3 and R_4, the temperature variation of V_{BE} doesn't cause the current to rise very rapidly (the larger the drop across R_3 and R_4, the less sensitive it is), and the circuit will be stable. Stability is improved by mounting the diodes in physical contact with the output transistors (or their heat sinks).

You can estimate the thermal stability of such a circuit by remembering that the base-emitter drop decreases by about 2.1mV for each 1°C rise and that the collector current increases by a factor of 10 for every 60mV increase in base-emitter voltage. For example, if R_2 were replaced by a diode, you would have three diode drops between the bases of Q_2 and Q_3, leaving about one diode drop across the series combination of R_3 and R_4. (The latter would then be chosen to give an appropriate quiescent current, perhaps 50mA for an audio power amplifier.) The worst case for thermal stability occurs if the biasing diodes are not thermally coupled to the output transistors.

Let us assume the worst and calculate the increase in output-stage quiescent current corresponding to a 30°C temperature rise in output transistor temperature. That's not a lot for a power amplifier, by the way. For that temperature rise, the V_{BE} of the output transistors will decrease by about 63mV at constant current, raising the voltage across R_3 and R_4 by about 20% (i.e., the quiescent current will rise by about 20%). The corresponding figure for the preceding amplifier circuit without emitter resistors (Fig. 2.57) will be a factor of 10 rise in quiescent current (recall that

I_C increases a decade per 60mV increase in V_{BE}), i.e., 1000%. The improved thermal stability of this biasing arrangement is evident.

This circuit has the additional advantage that by adjusting the quiescent current, you have some control over the amount of residual crossover distortion. A push-pull amplifier biased in this way to obtain substantial quiescent current at the crossover point is sometimes referred to as a class AB amplifier, meaning that both transistors conduct simultaneously during a portion of the cycle. In practice, you choose a quiescent current that is a good compromise between low distortion and excessive quiescent dissipation. Feedback, the subject of the next chapter, is almost always used to reduce distortion still further.

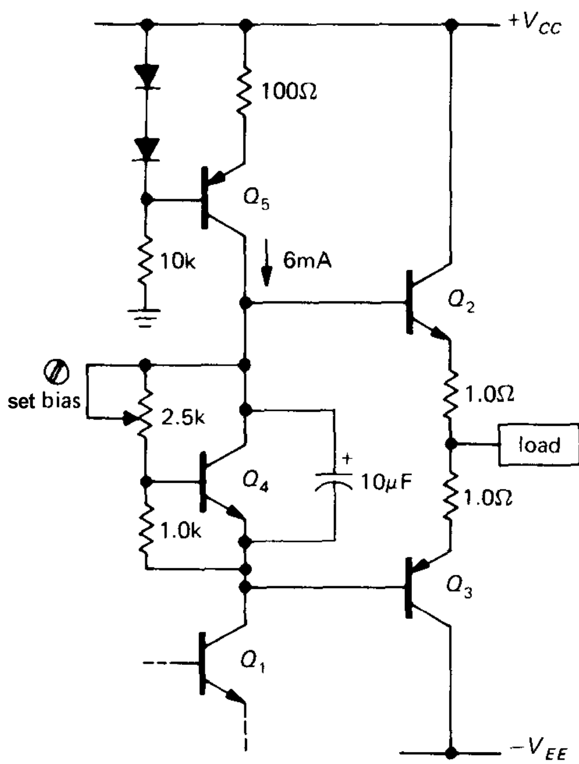


Figure 2.59. Biasing a push-pull output stage for low crossover distorion and good thermal stability.

An alternative method for biasing a push-pull follower is shown in Figure 2.59. Q_4 acts as an adjustable diode: The base resistors are a divider, and therefore Q_4's collector-emitter voltage will stabilize at a value that puts 1 diode drop from base to emitter, since any greater V_{CE} will bring it into heavy conduction. For instance, if both resistors were 1k, the transistor would turn on at 2 diode drops, collector to emitter. In this case, the bias adjustment lets you set the push-pull interbase voltage anywhere from 1 to 3.5 diode drops. The 10μF capacitor ensures that both output transistor bases see the same signal; such a bypass capacitor is a good idea for any biasing scheme you use. In this circuit, Q_1's collector resistor has been replaced by current source Q_5. That's a useful circuit variation, because with a resistor it is sometimes difficult to get enough base current to drive Q_2 near the top of the swing. A resistor small enough to drive Q_2 sufficiently results in high quiescent collector current in Q_1 (with high dissipation), and also reduced voltage gain (remember that $G = -R_{\text{collector}}/R_{\text{emitter}}$). Another solution to the problem of Q_2's base drive is the use of bootstrapping, a technique that will be discussed shortly.

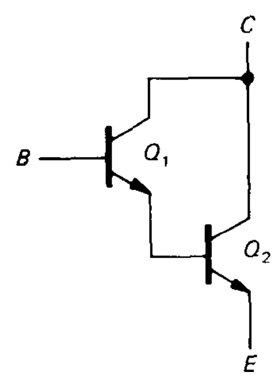


Figure 2.60. Darlington transistor configuration.

2.16 Darlington connection

If you hook two transistors together as in Figure 2.60, the result behaves like a single transistor with beta equal to the product of the two transistor betas. This can be very handy where high currents are involved (e.g., voltage regulators or power amplifier output stages), or for input stages of amplifiers where very high input impedance is necessary.

For a Darlington transistor the base-emitter drop is twice normal, and the saturation voltage is at least one diode drop (since Q_1's emitter must be a diode drop above Q_2's emitter). Also, the combination tends to act like a rather slow transistor because Q_1 cannot turn off Q_2 quickly. This problem is usually taken care of by including a resistor from base to emitter of Q_2 (Fig. 2.61). R also prevents leakage current through Q_1 from biasing

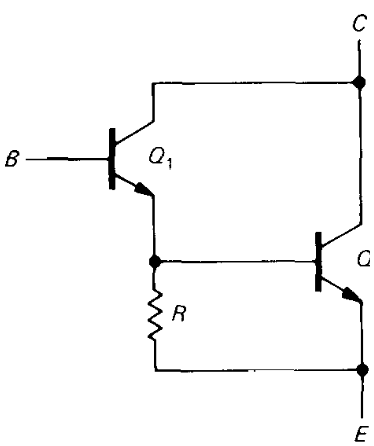


Figure 2.61. Improving turn-off speed in a Darlington pair.

Q_2 into conduction; its value is chosen so that Q_1's leakage current (nanoamps for small-signal transistors, as much as hundreds of microamps for power transistors) produces less than a diode drop across R and so that R doesn't sink a large proportion of Q_2's base current when it has a diode drop across it. Typically R might be a few hundred ohms in a power transistor Darlington, or a few thousand ohms for a small-signal Darlington.

Darlington transistors are available as single packages, usually with the base-emitter resistor included. A typical example is the *npn* power Darlington 2N6282, with current gain of 2400 (typically) at a collector current of 10 amps.

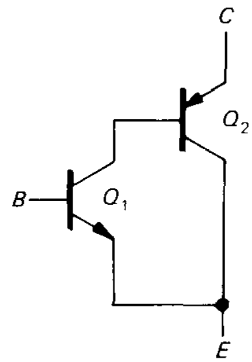


Figure 2.62. Sziklai connection ("complementary Darlington").

Sziklai connection

A similar beta-boosting configuration is the Sziklai connection, sometimes referred to as a complementary Darlington (Fig. 2.62). This combination behaves like an *npn* transistor, again with large beta. It has only a single base-emitter drop, but it also cannot saturate to less than a diode drop. A small resistor from base to emitter of Q_2 is advisable. This connection is common in push-pull power output stages where the designer wishes to use one polarity of output transistor only. Such a circuit is shown in Figure 2.63. As before, R_1 is Q_1's collector resistor. Darlington Q_2Q_3 behaves like a single *npn* transistor with high current gain. The Sziklai connected pair Q_4Q_5 behaves like a single high-gain *pnp* power transistor. As before, R_3 and R_4 are small. This circuit is sometimes called a pseudocomplementary push-pull follower. A true complementary stage would use a Darlington-connected *pnp* pair for Q_4Q_5.

Superbeta transistor

The Darlington connection and its near relatives should not be confused with the so-called superbeta transistor, a device with very high h_{FE} achieved through the manufacturing process. A typical superbeta transistor is the 2N5962, with a guaranteed minimum current gain of 450 at collector currents from 10μA to 10mA; it belongs to the 2N5961–2N5963 series; with a range of maximum V_{CE}s of 30 to 60 volts (if you need higher collector voltage, you have to settle for lower beta). Superbeta matched pairs are available for use in low-level amplifiers that require matched characteristics, a topic we will discuss in Section 2.18. Examples are the LM394 and MAT-01 series; these provide high-gain npn transistor pairs whose V_{BE}s are matched to a fraction of a millivolt (as little as 50μV in the best versions) and whose h_{FE}s are matched to about 1%. The MAT-03 is a pnp matched pair.

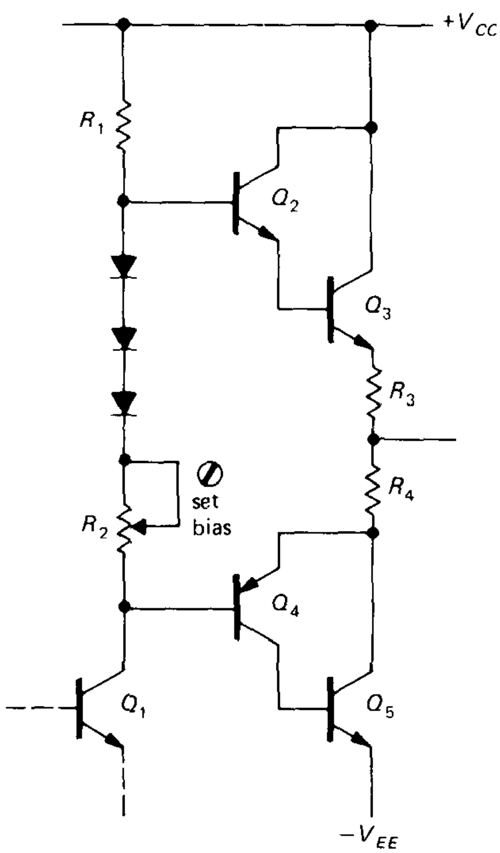


Figure 2.63. Push-pull power stage using only npn output transistors.

It is possible to combine superbeta transistors in a Darlington connection. Some commercial devices (e.g., the LM11 and LM316 op-amps) achieve base bias currents as low as 50 picoamps this way.

□ 2.17 Bootstrapping

When biasing an emitter follower, for instance, you choose the base voltage divider resistors so that the divider presents a stiff voltage source to the base, i.e., their parallel impedance is much less than the impedance looking into the base. For this reason the resulting circuit has an input impedance dominated by the voltage divider – the driving signal sees a much lower impedance than would otherwise be necessary. Figure 2.64 shows an example. The input

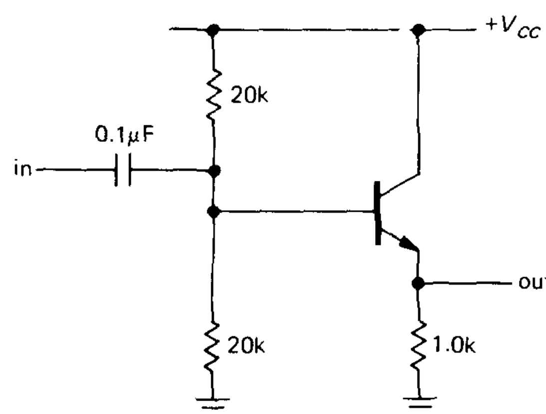


Figure 2.64

resistance of about 9k is mostly due to the voltage-divider impedance of 10k. It is always desirable to keep input impedances high, and anyway it's a shame to load the input with the divider, which, after all, is only there to bias the transistor. Bootstrapping is the colorful name given to a technique that circumvents this problem (Fig. 2.65). The transistor is biased by the divider R_1R_2 through series resistor R_3. C_2 is chosen to have low impedance at signal frequencies compared with the bias resistors. As always, bias is stable if the dc impedance seen from the base (in this case 9.7k) is much less than the dc impedance looking into the base (in this case approximately 100k). But now the signal-frequency input impedance is no longer the same as the dc impedance. Look at it this way: An input wiggle v_{in} results in an emitter wiggle $v_E \approx v_{\text{in}}$. So the change in current through bias resistor R_3 is $i = (v_{\text{in}} - v_E)/R_3 \approx 0$, i.e., Z_{in} (due to bias string) $= v_{\text{in}}/i_{\text{in}} \approx$ infinity. We've made the loading (shunt) impedance of the bias network very large at *signal frequencies.*

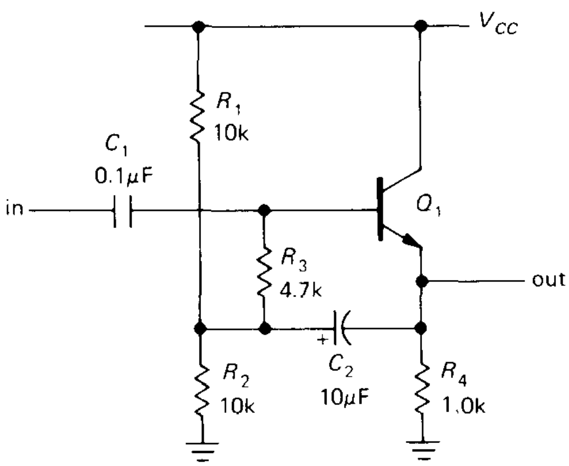


Figure 2.65. Raising the input impedance of an emitter follower at signal frequencies by bootstrapping the base bias divider.

Another way of seeing this is to notice that R_3 always has the same voltage across it at signal frequencies (since both ends of the resistor have the same voltage changes), i.e., it's a current source. But a current source has infinite impedance. Actually, the effective impedance is less than infinity because the gain of a follower is slightly less than 1. That is so because the base-emitter drop depends on collector current, which changes with the signal level. You could have predicted the same result from the voltage-dividing effect of the impedance looking into the emitter [$r_e = 25/I_C(\text{mA})$ ohms] combined with the emitter resistor. If the follower has voltage gain A ($A \approx 1$), the effective value of R_3 at signal frequencies is

$$R_3/(1 - A)$$

In practice the value of R_3 is effectively increased by a hundred or so, and the input impedance is then dominated by the transistor's base impedance. The emitter-degenerated amplifier can be bootstrapped in the same way, since the signal on the emitter follows the base. Note that the bias divider circuit is driven by the low-impedance emitter output at signal frequencies, thus isolating the input signal from this usual task.

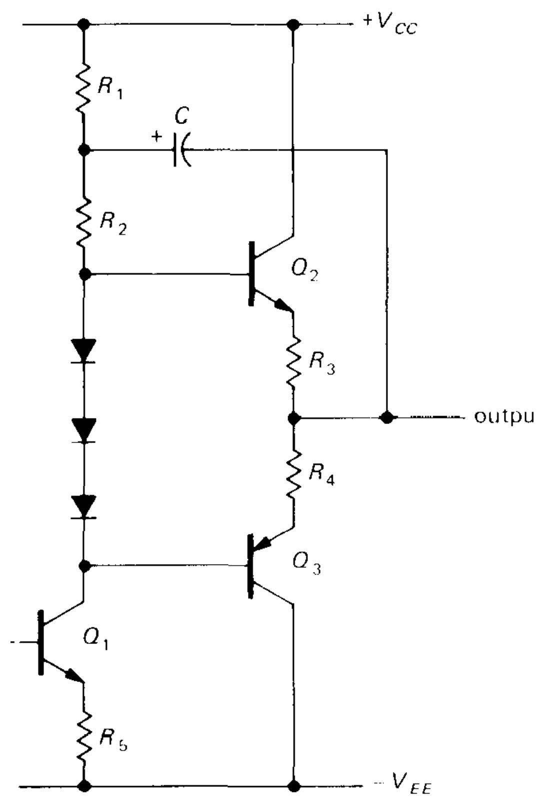


Figure 2.66. Bootstrapping driver-stage collector load resistor in a power amplifier.

□ *Bootstrapping collector load resistors*

The bootstrap principle can be used to increase the effective value of a transistor's collector load resistor, if that stage drives a follower. That can increase the voltage gain of the stage substantially [recall that $G_V = -g_m R_C$, with $g_m = 1/(R_E + r_e)$]. Figure 2.66 shows an example of a bootstrapped push-pull output stage similar to the push-pull follower circuit we saw earlier. Because the output follows Q_2's base signal, C bootstraps Q_1's collector load, keeping a constant voltage across R_2 as the signal varies (C must be chosen to have low impedance compared with R_1 and R_2 at all signal frequencies). That makes R_2 look like a current source, raising Q_1's voltage gain and maintaining good base drive to Q_2, even at the peaks of the signal swing. When the signal gets near V_{CC}, the junction of R_1 and R_2 actually rises above V_{CC} because of the stored charge in C. In this case, if $R_1 = R_2$ (not a bad choice) the junction between them rises to 1.5 times V_{CC} when the output reaches V_{CC}. This circuit has enjoyed considerable popularity in commercial audio amplifier design, although a simple current source in place of the bootstrap is superior, since it maintains the improvement at low frequencies and eliminates the undesirable electrolytic capacitor.

2.18 Differential amplifiers

The differential amplifier is a very common configuration used to amplify the difference voltage between two input signals. In the ideal case the output is entirely independent of the individual signal levels – only the difference matters. When both inputs change levels together, that's a *common-mode* input change. *A* differential change is called *normal mode.* A good differential amplifier has a high *common-mode rejection ratio* (CMRR), the ratio of response for a normal-mode signal to the response for a common-mode signal of the same amplitude. CMRR is usually specified in decibels. The common-mode input range is the voltage level over which the inputs may vary.

Differential amplifiers are important in applications where weak signals are contaminated by "pickup" and other miscellaneous noise. Examples include digital signals transferred over long cables (usually twisted pairs of wires), audio signals (the term "balanced" means differential, usually 600Ω impedance, in the audio business), radiofrequency signals (twin-lead cable is differential), electrocardiogram voltages, magnetic-core memory readout signals, and numerous other applications. A differential amplifier at the receiving end restores the original signal if the common-mode signals are not too large. Differential amplifiers are universally used in operational amplifiers, which we will come to soon. They're very important in dc amplifier design (amplifiers that amplify clear down to dc, i.e., have no coupling capacitors) because their symmetrical design is inherently compensated against thermal drifts.

Figure 2.67 shows the basic circuit. The output is taken off one collector with respect to ground; that is called a *single-ended output* and is the most common configuration. You can think of this amplifier as a device that amplifies a difference signal and converts it to a single-ended signal so that ordinary subcircuits (followers, current sources, etc.) can make use of the output. (If, instead, a differential output is desired, it is taken between the collectors.)

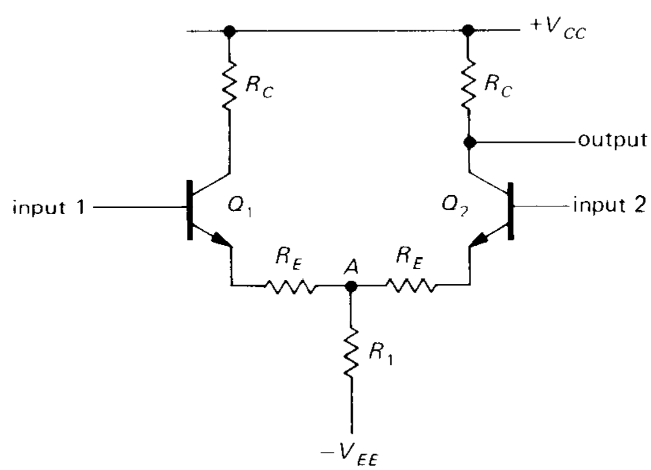


Figure 2.67. Classic transistor differential amplifier.

What is the gain? That's easy enough to calculate: Imagine a symmetrical input signal wiggle, in which input 1 rises by

v_{in} (a small-signal variation) and input 2 drops by the same amount. As long as both transistors stay in the active region, point A remains fixed. The gain is then determined as with the single transistor amplifier, remembering that the input change is actually twice the wiggle on either base: $G_{diff} = R_C/2(r_e + R_E)$. Typically R_E is small, 100 ohms or less, or it may be omitted entirely. Differential voltage gains of a few hundred are typical.

The common-mode gain can be determined by putting identical signals v_{in} on both inputs. If you think about it correctly (remembering that R_1 carries both emitter currents), you'll find $G_{CM} = -R_C/(2R_1 + R_E)$. Here we've ignored the small r_e, because R_1 is typically large, at least a few thousand ohms. We really could have ignored R_E as well. The CMRR is roughly $R_1/(r_e + R_E)$. Let's look at a typical example (Fig. 2.68) to get some familiarity with differential amplifiers.

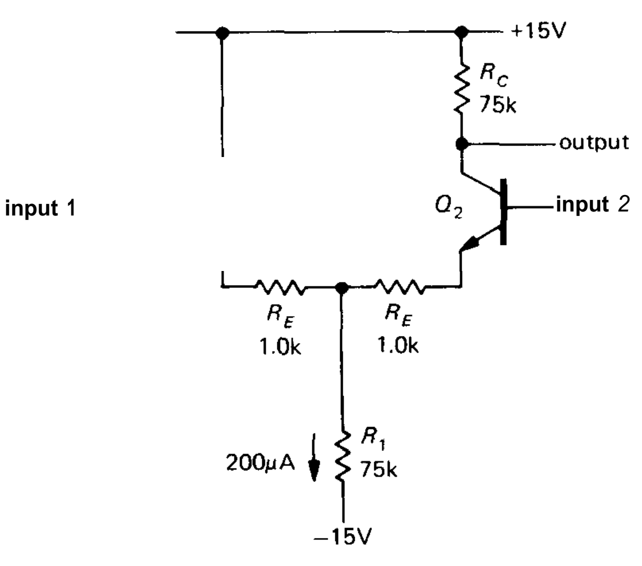
R_C is chosen for a quiescent current of $100\mu A$. As usual, we put the collector at $0.5V_{CC}$ for large dynamic range. Q_1's collector resistor can be omitted, since no output is taken there. R_1 is chosen to give total emitter current of $200\mu A$, split equally between the two sides when the (differential) input is zero. From the formulas just derived, this amplifier has a differential gain of 30 and a common-mode gain of 0.5. Omitting the 1.0k resistors raises the differential gain to 150, but drops the (differential) input impedance from about 250k to about 50k (you can substitute Darlington transistors in the input stage to raise the impedance into the megohm range, if necessary).

Remember that the maximum gain of a single-ended grounded emitter amplifier biased to $0.5V_{CC}$ is $20V_{CC}$. In the case of a differential amplifier the maximum differential gain ($R_E = 0$) is half that figure, or (for arbitrary quiescent point) 20 times the voltage across the collector resistor. The corresponding maximum CMRR (again with $R_E = 0$) is equal to 20 times the voltage across R_1.

EXERCISE 2.13

Verify that these expressions are correct. Then design a differential amplifier to your own specifications.

The differential amplifier is sometimes called a "long-tailed pair," because if the length of a resistor symbol indicated its magnitude, the circuit would look like Figure 2.69. The long tail determines the



$$G_{diff} = \frac{v_{out}}{v_1 - v_2} = \frac{R_C}{2(R_E + r_e)}$$

$$G_{CM} = -\frac{R_C}{2R_1 + R_E + r_e}$$

$$\text{CMRR} \approx \frac{R_1}{R_E + r_e}$$

Figure 2.68. Calculating differential amplifier performance.

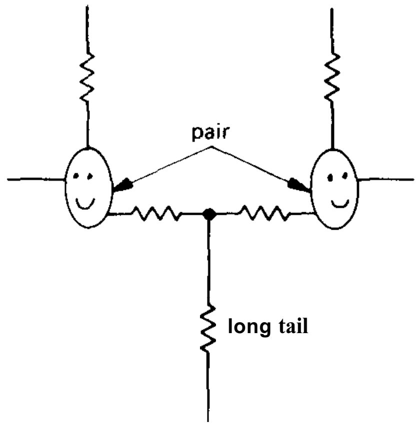


Figure 2.69

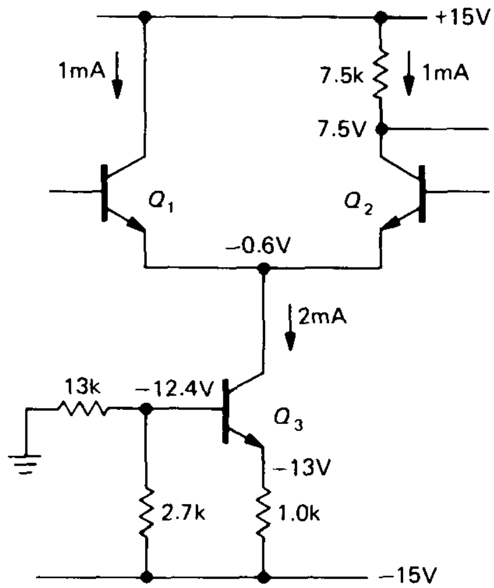


Figure 2.70. Improving CMRR of the differential amplifier with a current source.

common-mode gain, and the small inter-emitter resistance (including intrinsic emitter resistance r_e) determines the differential gain.

Current-source biasing

The common-mode gain of the differential amplifier can be reduced enormously by substituting a current source for R_1. Then R_1 effectively becomes very large, and the common-mode gain is nearly zero. If you prefer, just imagine a common-mode input swing; the emitter current source maintains a constant total emitter current, shared equally by the two collector circuits, by symmetry. The output is therefore unchanged. Figure 2.70 shows an example. The CMRR of this circuit, using an LM394 monolithic transistor pair for Q_1 and Q_2 and a 2N5963 current source is 100,000:1 (100dB). The common-mode input range for this circuit goes from −12 volts to +7 volts; it is limited at the low end by the compliance of the emitter current source and at the high end by the collector's quiescent voltage.

Be sure to remember that this amplifier, like all transistor amplifiers, must have a dc bias path to the bases. If the input is capacitively coupled, for instance, you must have base resistors to ground. An additional caution for differential amplifiers, particularly those without inter-emitter resistors: Bipolar transistors can tolerate only 6 volts of base-emitter reverse bias before breakdown; thus, applying a differential input voltage larger than this will destroy the input stage (if there is no inter-emitter resistor). An inter-emitter resistor limits the breakdown current and prevents destruction, but the transistors may be degraded (in h_{fe}, noise, etc.). In either case the input impedance drops drastically during reverse conduction.

Use in single-ended dc amplifiers

A differential amplifier makes an excellent dc amplifier, even for single-ended inputs. You just ground one of the inputs and connect the signal to the other (Fig. 2.71). You might think that the "unused" transistor could be eliminated. Not so! The differential configuration is inherently compensated for temperature drifts, and even when one input is at ground that transistor is still doing something: A temperature change causes both V_{BE}s to change the same amount, with no change in balance or output. That is, changes in V_{BE} are not amplified by G_{diff} (only by G_{CM}, which can be made essentially zero). Furthermore, the cancellation of V_{BE}s means that there are no 0.6 volt drops at the input to worry about. The quality of a dc amplifier constructed this way is limited only by mismatching of input V_{BE}s or their temperature coefficients. Commercial monolithic transistor pairs and commercial differential amplifier ICs are available with extremely good matching (e.g., the MAT-01 npn monolithic matched pair has a typical drift of V_{BE} between the two transistors of 0.15μV/°C and 0.2μV per month).

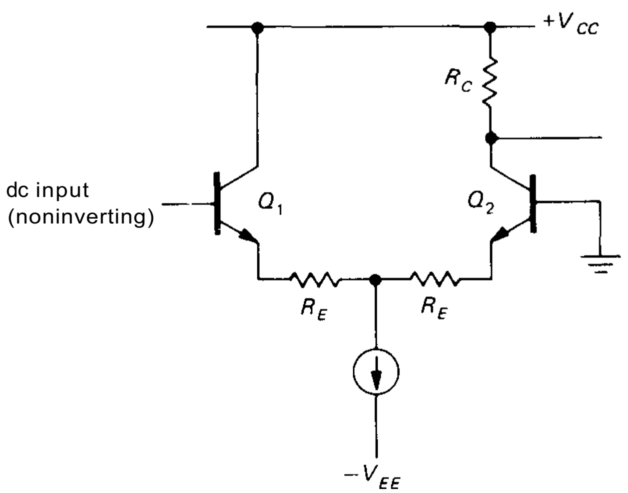


Figure 2.71. A differential amplifier can be used as a precision single-ended dc amplifier.

Either input could have been grounded in the preceding circuit example. The choice depends on whether or not the amplifier is supposed to invert the signal. (The configuration shown is preferable at high frequencies, however, because of *Miller effect*; see Section *2.19.)* The connection shown is noninverting, and so the inverting input has been grounded. This terminology carries over to op-amps, which are simply high-gain differential amplifiers.

Current mirror active load

As with the simple grounded emitter amplifier, it is sometimes desirable to have a single-stage differential amplifier with very high gain. An elegant solution is a current mirror active load (Fig. 2.72). Q_1Q_2 is the differential pair with emitter current source. Q_3 and Q_4, a current mirror, form the collector load. The high effective collector load impedance provided by the mirror yields voltage gains of 5000 or more, assuming no load at the amplifier's output. Such an amplifier is usually used only within a feedback loop, or as a comparator (discussed in the next section). Be sure to load such an amplifier with a high impedance, or the gain will drop enormously.

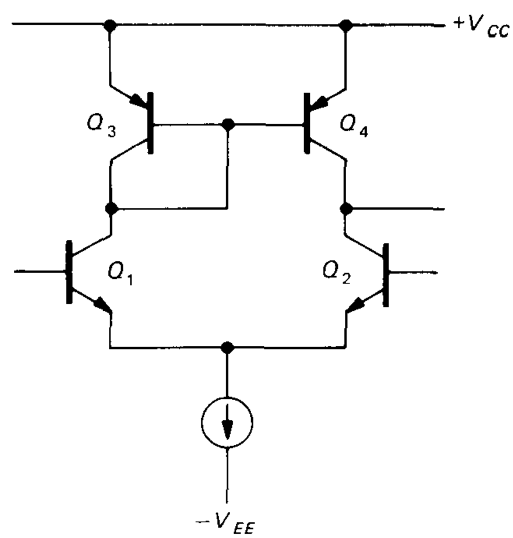


Figure 2.72. Differential amplifier with active current mirror load.

Differential amplifiers as phase splitters

The collectors of a symmetrical differential amplifier generate equal signal swings of opposite phase. By taking outputs from both collectors, you've got a phase splitter. Of course, you could also use a differential amplifier with both differential inputs and differential outputs. This differential output signal could then be used to drive an additional differential amplifier stage, with greatly improved overall common-mode rejection.

Differential amplifiers as comparators

Because of its high gain and stable characteristics, the differential amplifier is the main building block of the comparator, a circuit that tells which of two inputs is larger. They are used for all sorts of applications: switching on lights and heaters, generating square waves from triangles, detecting when a level in a circuit exceeds some particular threshold, class D amplifiers and pulse-code modulation, switching power supplies, etc. The basic idea is to connect a differential amplifier so that it turns a transistor switch on or off, depending on the relative levels of the input signals. The linear region of amplification is ignored, with one or the other of the two input transistors cut off at any time. A typical hookup is illustrated in the next section by a temperature-controlling circuit that uses a resistive temperature sensor (thermistor).

2.19 Capacitance and Miller effect

In our discussion so far we have used what amounts to a dc, or low-frequency, model of the transistor. Our simple current amplifier model and the more sophisticated Ebers-Moll transconductance model both deal with voltages, currents, and resistances seen at the various terminals. With these models alone we have managed to go quite far, and in fact these simple models contain nearly everything you will ever need to know to design transistor circuits. However, one important aspect that has serious impact on high-speed and high-frequency circuits has been neglected: the existence of capacitance in the external circuit and in the transistor junctions themselves. Indeed, at high frequencies the effects of capacitance often dominate circuit behavior; at 100 MHz a typical junction capacitance of 5pF has an impedance of 320 ohms!

We will deal with this important subject in detail in Chapter 13. At this point we would merely like to state the problem, illustrate some of its circuit incarnations, and suggest some methods of circumventing the problem. It would be a mistake to leave this chapter without realizing the nature of this problem. In the course of this brief discussion we will encounter the famous Miller *effect* and the use of configurations such as the cascode to overcome it.

Junction and circuit capacitance

Capacitance limits the speed at which the voltages within a circuit can swing ("slew rate"), owing to finite driving impedance or current. When a capacitance is driven by a finite source resistance, you see *RC* exponential charging behavior, whereas a capacitance driven by a current source leads to slew-rate-limited waveforms (ramps). As general guidance, reducing the source impedances and load capacitances and increasing the drive currents within a circuit will speed things up. However, there are some subtleties connected with feedback capacitance and input capacitance. Let's take a brief look.

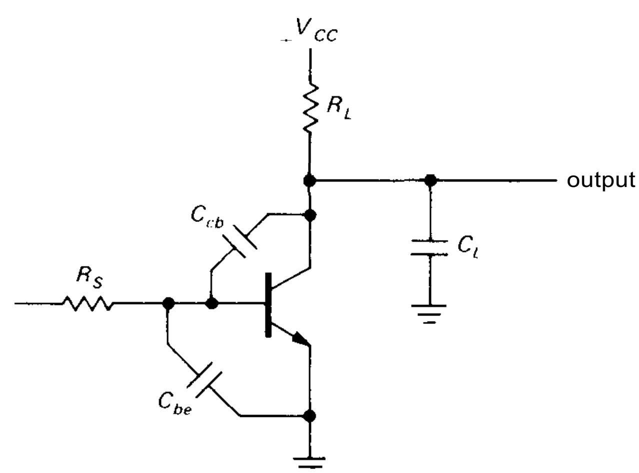


Figure 2.73. Junction and load capacitances in a transistor amplifier.

The circuit in Figure 2.73 illustrates most of the problems of junction capacitance. The output capacitance forms a time constant with the output resistance R_L (R_L includes both the collector and load resistances, and C_L includes both junction and load capacitances), giving a rolloff starting at some frequency $f = 1/2\pi R_L C_L$. The same is true for the input capacitance in combination with the source impedance R_S.

Miller effect

C_{cb} is another matter. The amplifier has some overall voltage gain G_V, so a small voltage wiggle at the input results in a wiggle G_V times larger (and inverted) at the collector. This means that the signal source sees a current through C_{cb} that is $G_V + 1$ times as large as if C_{cb} were connected from base to ground; i.e., for the purpose of input rolloff frequency calculations, the feedback capacitance behaves like a capacitor of value $C_{cb}(G_V + 1)$ from input to ground. This effective increase of C_{cb} is known as the Miller effect. It often dominates the rolloff characteristics of amplifiers, since a typical feedback capacitance of 4pF can look like several hundred picofarads to ground.

There are several methods available to beat the Miller effect. It is absent altogether in a grounded base stage. You can decrease the source impedance driving a grounded emitter stage by using an emitter follower. Figure 2.74 shows two other possibilities. The differential amplifier circuit (with no collector resistor in Q_1) has no Miller effect; you can think of it as an emitter follower driving a grounded base amplifier. The second circuit is the famous cascode configuration. Q_1 is a grounded emitter amplifier with R_L as its collector resistor. Q_2 is interposed in the collector path to prevent Q_1's collector from swinging (thereby eliminating the Miller effect) while passing the collector current through to the load resistor unchanged. V_+ is a fixed bias voltage, usually set a few volts above Q_1's emitter voltage to pin Q_1's collector and keep it in the active region. This fragment is incomplete as shown; you could either include a bypassed emitter resistor and base divider for biasing (as we did earlier in the chapter) or include it within an overall loop with feedback at dc. V_+ might be provided from a divider or zener, with bypassing to keep it stiff at signal frequencies.

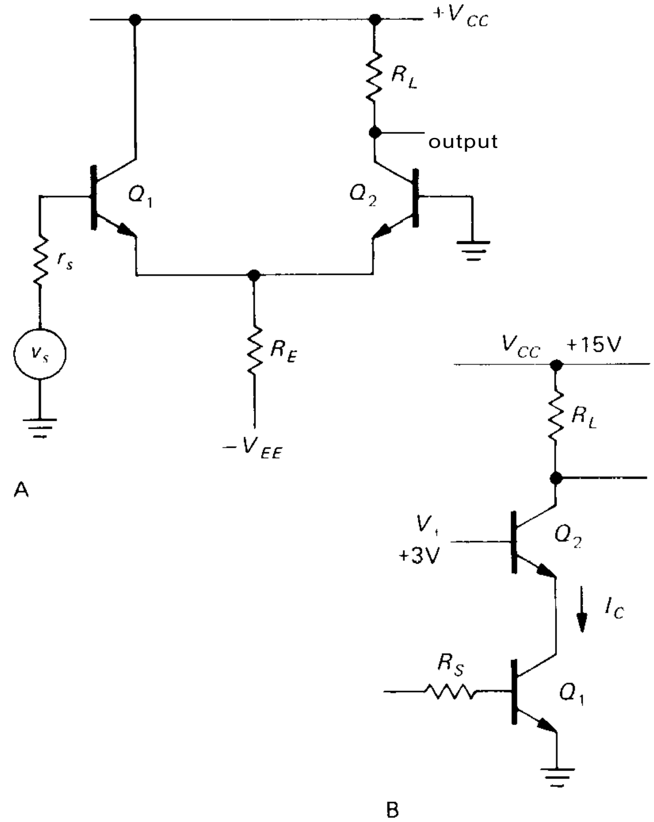


Figure 2.74. Two circuit configurations that avoid Miller effect. Circuit B is the cascode.

EXERCISE 2.14

Explain in detail why there is no Miller effect in either transistor in the preceding differential amplifier and cascode circuits.

Capacitive effects can be somewhat more complicated than this brief introduction might indicate. In particular: (a) The rolloffs due to feedback and output capacitances are not entirely independent; in the terminology of the trade there is pole splitting, an effect we will explain in the next chapter. (b) The input capacitance still has an effect, even with a stiff input signal source. In particular, current that flows through C_{be} is not amplified by the transistor. This base current "robbing" by the input capacitance causes the transistor's small-signal current gain h_{fe} to drop at high frequencies, eventually reaching unity at a frequency known as f_T. (c) To complicate matters, the junction capacitances depend on voltage. C_{be} changes so rapidly with base current that it is not even specified on transistor data sheets; f_T is given instead. (d) When a transistor is operated as a switch, effects associated with charge stored in the base region of a saturated transistor cause an additional loss of speed. We will take up these and other topics having to do with high-speed circuits in Chapter 13.

2.20 Field-effect transistors

In this chapter we have dealt exclusively with bipolar junction transistors (BJTs), characterized by the Ebers-Moll equation. BJTs were the original transistors, and they still dominate analog circuit design. However, it would be a mistake to continue without a few words of explanation about the other kind of transistor, the field-effect transistor (FET), which we will take up in detail in the next chapter.

The FET behaves in many ways like an ordinary bipolar transistor. It is a 3-terminal amplifying device, available in both polarities, with a terminal (the gate) that controls the current flow between the other two terminals (source and drain). It has a unique property, though: The gate draws no current, except for leakage. This means that extremely high input impedances are possible, limited only by capacitance and leakage effects. With FETs you don't have to worry about providing substantial base current, as was necessary with the BJT circuit design of this chapter. Input currents measured in picoamperes are commonplace. Yet the FET is a rugged and capable device, with voltage and current ratings comparable to those of bipolar transistors.

Most of the available devices fabricated with transistors (matched pairs, differential and operational amplifiers, comparators, high-current switches and amplifiers, radiofrequency amplifiers, and digital logic) are also available with FET construction, often with superior performance. Furthermore, microprocessors and memory (and other large-scale digital electronics) are built almost exclusively with FETs. Finally, the area of micropower design is dominated by FE'T circuits.

FETs are so important in electronic design that we will devote the next chapter to them, before treating operational amplifiers and feedback in Chapter 4. We urge the reader to be patient with us as we lay the groundwork in these first three difficult chapters; that patience will be rewarded many times over in the succeeding chapters, as we explore the enjoyable topics of circuit design with operational amplifiers and digital integrated circuits.

SOME TYPICAL TRANSISTOR CIRCUITS

To illustrate some of the ideas of this chapter, let's look at a few examples of circuits with transistors. The range of circuits we can cover is necessarily limited, since real-world circuits often use negative feedback, a subject we will cover in Chapter 4.

2.21 Regulated power supply

Figure 2.75 shows a very common configuration. R_1 normally holds Q_1 on; when the output reaches 10 volts, Q_2 goes into conduction (base at 5V), preventing further rise of output voltage by shunting base current from Q_1's base. The supply can be made adjustable by replacing R_2 and R_3

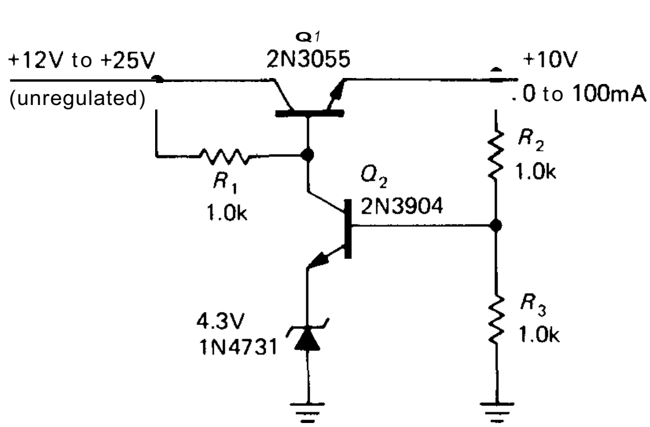


Figure 2.75. Feedback voltage regulator.

by a potentiometer. This is actually an example of negative feedback: Q_2 "looks at" the output and does something about it if the output isn't at the right voltage.

2.22 Temperature controller

The schematic diagram in Figure 2.76 shows a temperature controller based on a *thermistor* sensing element, a device that changes resistance with temperature. Differential Darlington $Q_1 - Q_4$ compares the voltage of the adjustable reference divider R_4–R_6 with the divider formed from the thermistor and R_2. (By comparing *ratios* from the same supply, the comparison becomes insensitive to supply variations; this particular configuration is called a **Wheatstone bridge**.) Current mirror Q_5Q_6 provides an active load to raise the gain, and mirror Q_7Q_8 provides emitter current. Q_9 compares the differential amplifier output with a fixed voltage, saturating Darlington $Q_{10}Q_{11}$, which supplies power to the heater, if the thermistor is too cold. R_9 is a current-sensing resistor that turns on protection transistor Q_{12} if the output current exceeds about 6 amps; that removes base drive from $Q_{10}Q_{11}$, preventing damage.

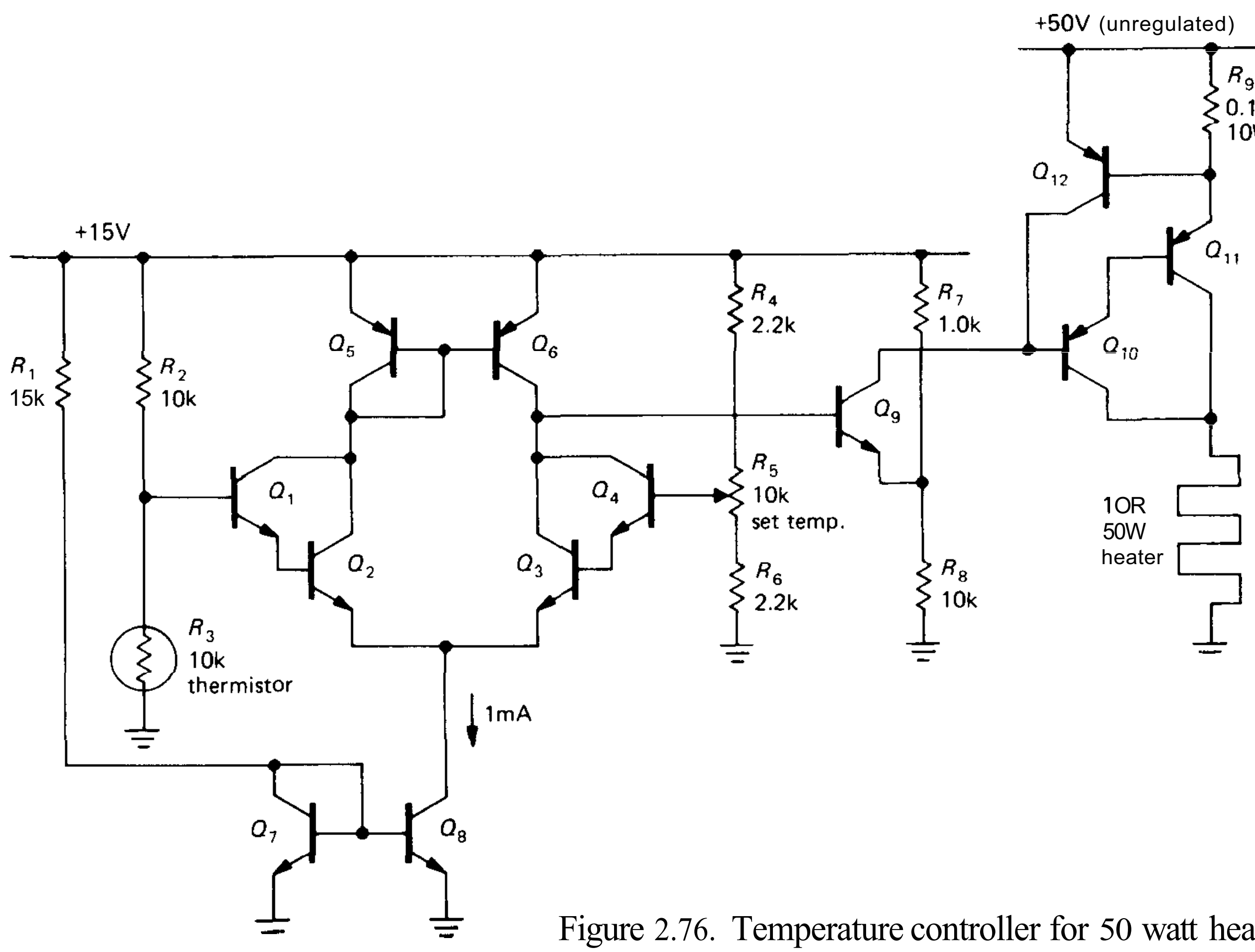


Figure 2.76. Temperature controller for 50 watt heater.

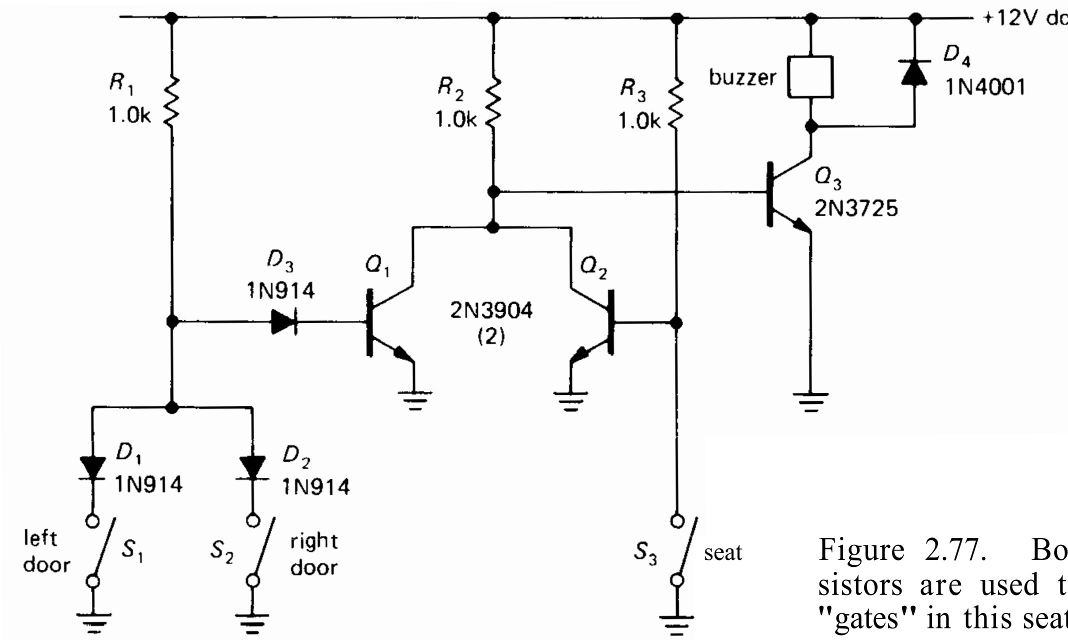


Figure 2.77. Both diodes and transistors are used to make digital logic "gates" in this seat-belt buzzer circuit.

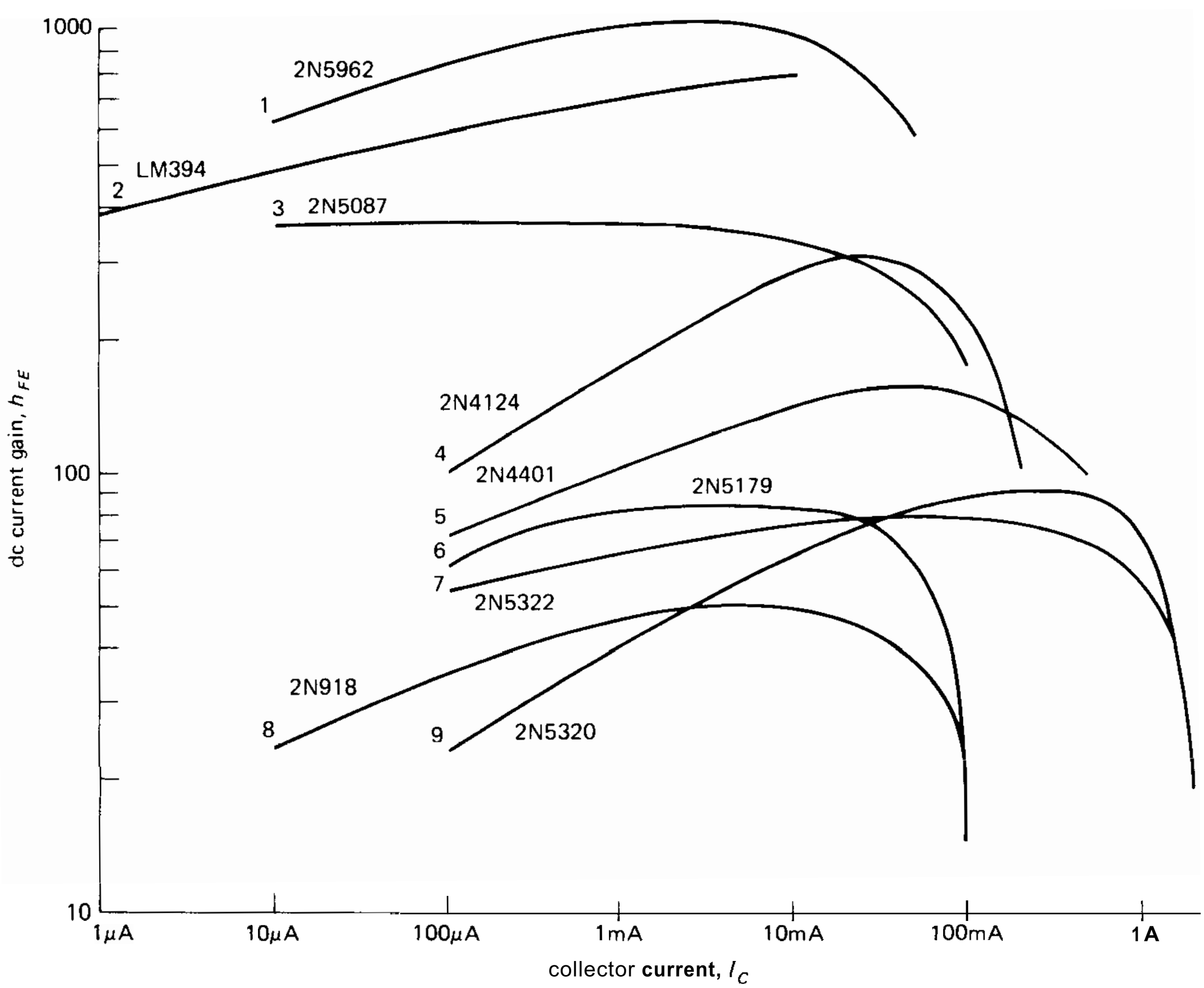


Figure 2.78. Curves of typical transistor current gain, h_{FE}, for a selection of transistors from Table 2.1. These curves are taken from manufacturers' literature. You can expect production spreads of +100%, −50% from the "typical" values graphed.

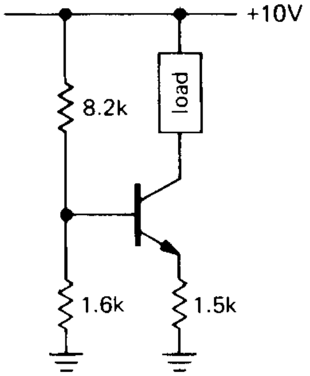


Figure 2.79

2.23 Simple logic with transistors and diodes

Figure 2.77 shows a circuit that performs a task we illustrated in Section 1.32: sounding a buzzer if either car door is open and the driver is seated. In this circuit the transistors all operate as switches (either off or saturated). Diodes D_1 and D_2 form what is called an OR gate, turning off Q_1 if either door is open (switch closed). However, the collector of Q_1 stays near ground, preventing the buzzer from sounding unless switch S_3 is also closed (driver seated); in that case R_2 turns Q_3 on, putting 12 volts across the buzzer. D_3 provides a diode drop so that Q_1 is off with S_1 or S_2 closed, and D_4 protects Q_3 from the buzzer's inductive turn-off transient. In Chapter 8 we will discuss logic circuitry in detail.

Table 2.1 presents a selection of useful and popular small-signal transistors; Figure 2.78 shows corresponding curves of current gain. See also Appendix K.

SELF-EXPLANATORY CIRCUITS

2.24 Good circuits

Figure 2.80 shows a couple of circuit ideas that use transistors.

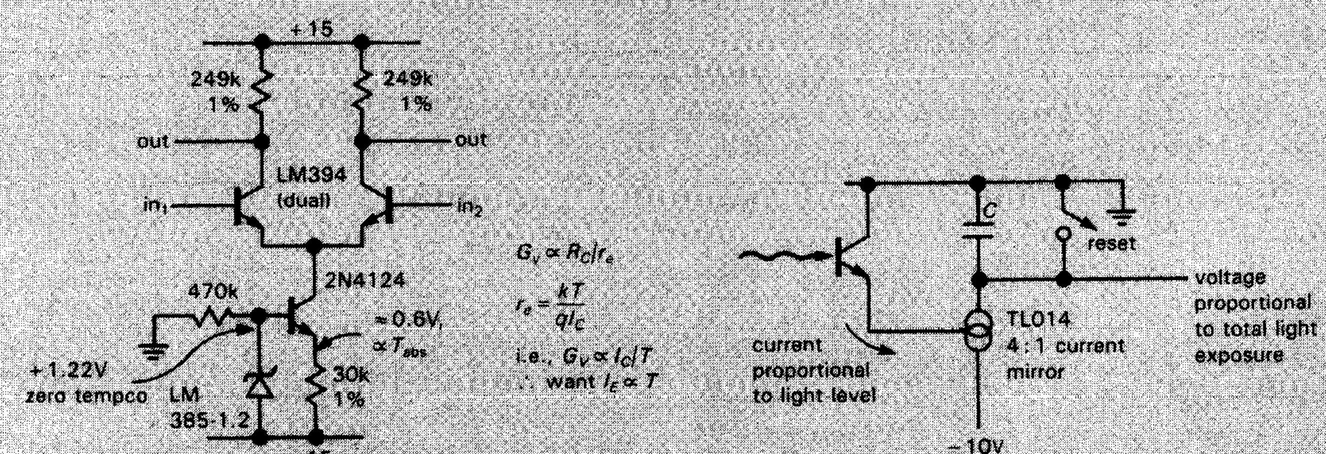
2.25 Bad circuits

A lot can be learned from your own mistakes or someone else's mistakes. In this section we present a gallery of blunders (Fig. 2.81). You can amuse yourself by thinking of variations on these bad circuits, and then avoiding them!

ADDITIONAL EXERCISES

(1) Design a transistor switch circuit that allows you to switch two loads to ground via saturated npn transistors. Closing switch *A* should cause both loads to be powered, whereas closing switch B should power only one load. Hint: Use diodes.

Circuit ideas

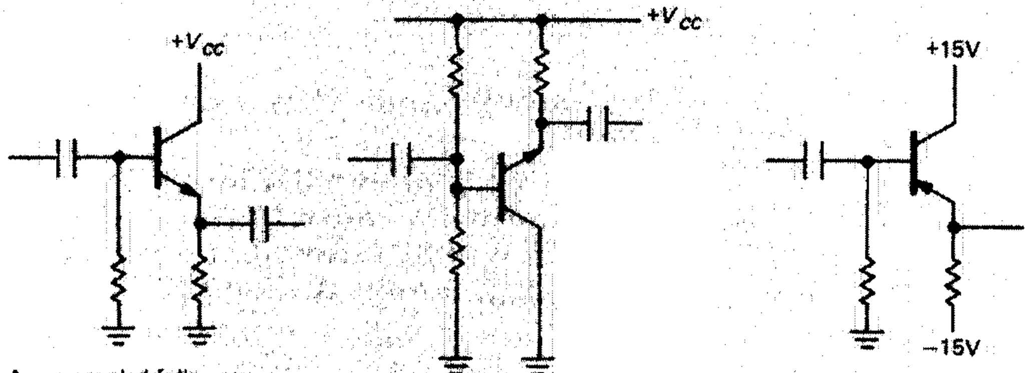


A. differential amplifier biasing for zero tempco of gain

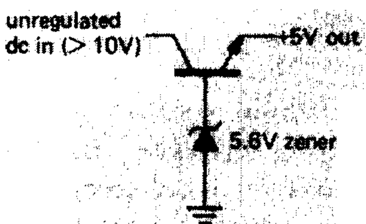
B. light integrator

Figure 2.80

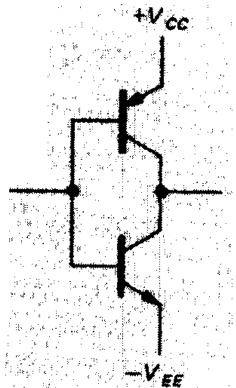
Bad circuits



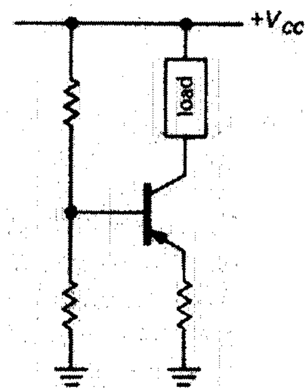
A ac coupled followers



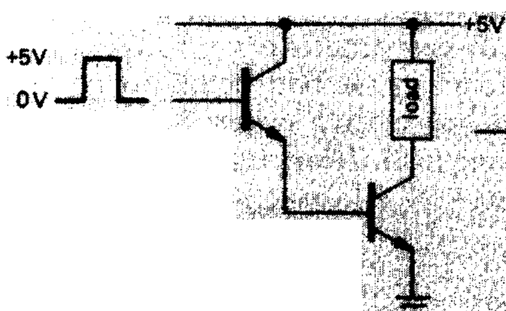
B +5V regulator



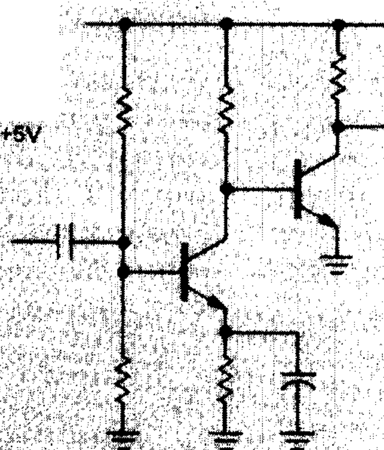
C push-pull follower



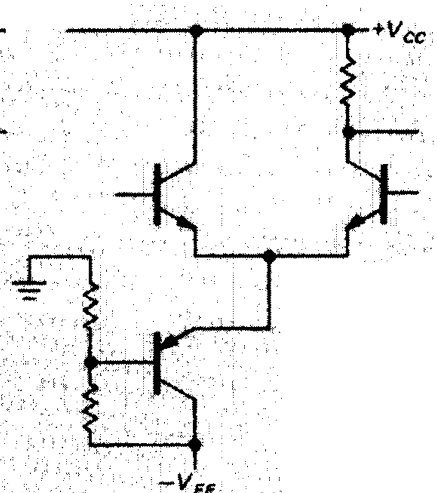
D current source



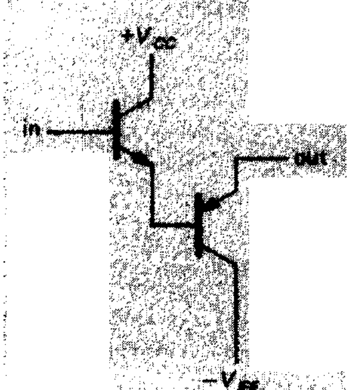
E high-current switch



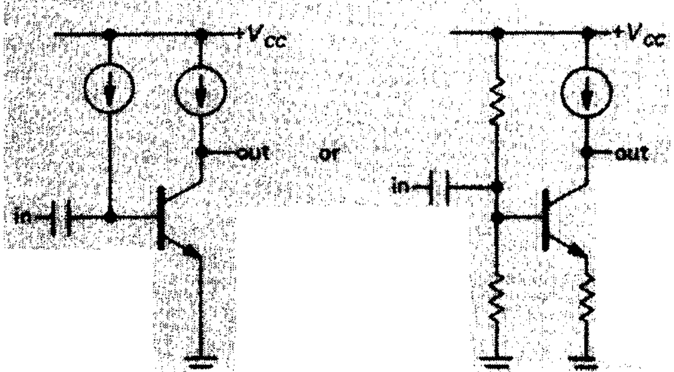
two-stage amplifier



G differential amplifier



H zero-offset follower



I high-gain ac amplifier

TABLE 2.1. SELECTED SMALL-SIGNAL TRANSISTORS[a]

	V_{CEO} (V)	I_C max (mA)	h_{FE} typ[b]	I_C (mA)	C_{cb} typ[c] (pF)	f_T typ[d] (MHz)	Gain curve	Metal TO-5[e] npn	Metal TO-5[e] pnp	Metal TO-18[f] npn	Metal TO-18[f] pnp	Plastic TO-92[h] npn	Plastic TO-92[h] pnp
General purpose	20	500	100	150	16	200		–	–	–	–	–	–
	25	200	200	2	1.8-2.8	300	4	–	–	–	–	4124	4126
	40	200	200	10	1.8-2.8	300		–	–	3947	3251	3904	3906
High gain, low noise	25	50	300	10	2-7	150		–	–	–	–	3391A[h],3707[h]	4058[h]
	25	300	250	50	4	300		–	–	–	–	6008[h]	6009[h]
	25	50	500	5	1.5-4	500		–	–	–	–	5089	–
	40	20	700	1	14	200	2	LM394	–	–	–	–	–
	45	50	1000	10	1.5	300	1	–	–	–	–	5962	–
	50	50	350	5	1.8	400	3	–	–	2848	3965	4967,5210	4965,5087
High current	30-60	600	150	150	5	300	5	2219	2905	2222	2907,3251	4401	4403
	50	1000	100	200	7	450		3725	5022	4014	–	–	–
	60	1000	70	80	15	100		2102,3107	4036	–	–	–	–
	75	2000	70	500	20	60	7,9	5320	5322	–	–	–	–
High voltage	150	600	100	10	3-6	250		–	4929	–	–	5550	5401
	300	1000	50	50	10	50		3439	5416	–	–	–	–
High speed	12	50	80	3	0.7	1500	6	–	–	5179	–	3662[h]	–
	12	100	50	8	1.5	900	8	–	–	918	4208	5770	–
	12	200	75	25	3	500		–	–	2369	2894	5769	5771

(a) all transistors are 2Nxxx numbers, except for the LM394 dual transistor. Devices listed on a single row are similar in characteristics and in some cases are electrically identical. (b) see figure 2.76. (c) at V_{CB}=10V. (d) see figure 13.4. (e) or TO-39. (f) or TO-72, TO-46. (h) TO-92 and its variants have two basic pinouts: EBC and ECB. Transistors with superscript h are ECB; all others are EBC.

(2) Consider the current source in Figure 2.79. ***(a)*** What is I_{load}? What is the output compliance? Assume V_{BE} is 0.6 volt. ***(b)*** If h_{FE} varies from 50 to 100 for collector voltages within the output compliance range, how much will the output current vary? (There are two effects here.) ***(c)*** If V_{BE} varies according to $\Delta V_{BE} = -0.0001\,\Delta V_{CE}$ (Early effect), how much will the load current vary over the compliance range? ***(d)*** What is the temperature coefficient of output current assuming that h_{FE} does not vary with temperature? What is the temperature coefficient of output current assuming that h_{FE} increases from its nominal value of 100 by 0.4%/°C?

(3) Design a common-emitter npn amplifier with voltage gain of 15, V_{CC} of +15 volts, and I_C of 0.5mA. Bias the collector at $0.5V_{CC}$, and put the low-frequency 3dB point at 100Hz.

(4) Bootstrap the circuit in the preceding problem in order to raise the input impedance. Choose the rolloff of the bootstrap appropriately.

(5) Design a dc-coupled differential amplifier with voltage gain of 50 (to a single-ended output) for input signals near ground, supply voltages of ±15 volts, and quiescent currents of 0.1mA in each transistor. Use a current source in the emitter and an emitter follower output stage.

(6) In this problem you will ultimately design an amplifier whose gain is controlled by an externally applied voltage (in Chapter 3 you will see how to do the same thing with FETs). (a) Begin by designing a long-tailed pair differential amplifier with emitter current source and no emitter resistors (undegenerated). Use ±15 volt supplies. Set I_C (each transistor) at 1mA, and use R_C = 1.0k. Calculate the voltage gain from a single-ended input (other input grounded) to a single-ended output. ***(b)*** Now modify the circuit so that an externally applied voltage controls the emitter current source. Give an approximate formula for the gain as a function of controlling voltage. (In a real circuit you might arrange a second set of voltage-controlled current sources to cancel the quiescent-point shift that gain changes produce in this circuit, or a differential-input second stage could be added to your circuit.)

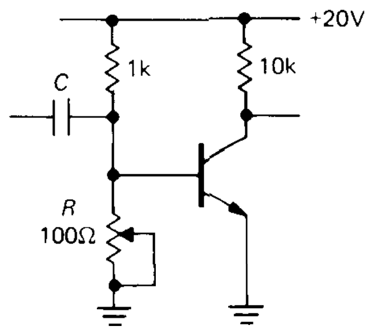


Figure 2.82

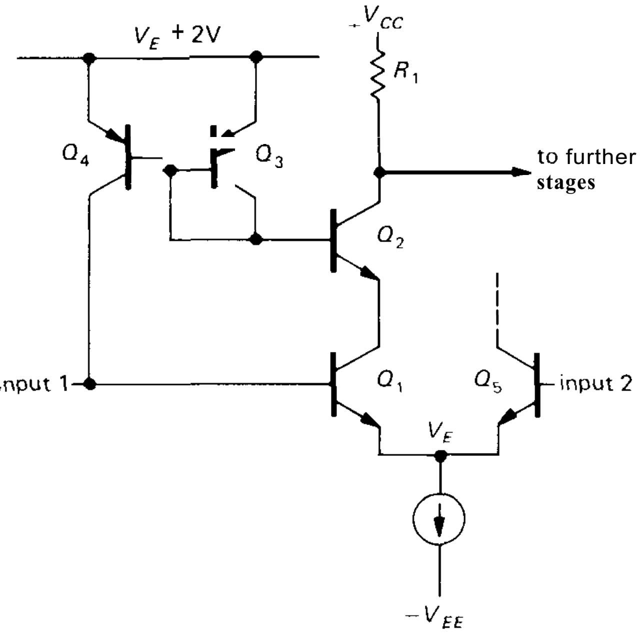


Figure 2.83. Base-current cancellation scheme, commonly used in high-quality operational amplifiers.

(7) Disregarding the lessons of this chapter, a disgruntled student builds the amplifier shown in Figure 2.82. He adjusts R until the quiescent point is $0.5V_{CC}$. ***(a)*** What is Z_{in} (at high frequencies where $Z_C \approx 0$)? ***(b)*** What is the small-signal voltage gain? ***(c)*** What rise in ambient temperature (roughly) will cause the transistor to saturate?

(8) Several commercially available precision op-amps (e.g., the venerable OP-07 and the recent LT1012) use the circuit in Figure **2.83** to cancel input bias current (only half of the symmetrical-input differential amplifier is shown in detail; the other half works the same way). Explain how the circuit works. Note: Q_1 and Q_2 are a beta-matched pair. Hint: It's all done with mirrors.