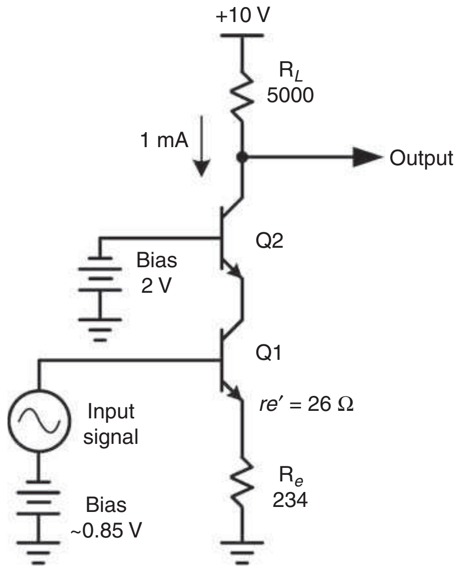

FIGURE 2.8 Cascode.

Cascode

A cascode stage is implemented by Q2 in Figure 2.8. The cascode stage is also called a *common base* stage because the base of its transistor is connected to AC ground. Here the cascode is being driven at its emitter by a CE stage comprising Q1. The most important function of a cascode stage is to provide isolation. It provides near-unity current gain, but can provide very substantial voltage gain. In some ways it is like the dual of an emitter follower.

A key benefit of the cascode stage is that it largely keeps the collector of the driving CE stage at a constant potential. It thus isolates the collector of the CE stage from the large swing of the output signal. This eliminates most of the effect of the collector-base capacitance of Q2, resulting in wider bandwidth due to suppression of the Miller effect. Similarly, it mitigates distortion caused by the nonlinear collector-base junction capacitance of the CE stage, since very little voltage swing now appears across the collector-base junction to modulate its capacitance.

The cascode connection also avoids most of the Early effect in the CE stage by nearly eliminating signal swing at its collector. A small amount of Early effect remains, however, because the signal swing at the base of the CE stage modulates the collector-base voltage slightly.

If the current gain of the cascode transistor is 100, then 99% of the signal current entering the emitter will flow in the collector. The input-output current gain is thus 0.99. This current transfer factor from emitter to collector is sometimes referred to as the *alpha* of the transistor.

The Early effect resistance *ro* still exists in the cascode transistor. It is represented as a resistance *ro* connected from collector to emitter. Suppose *ro* is only 10 kΩ. Is the output impedance of the collector of the cascode 10 kΩ? No, it is not.

Recall that 99% of the signal current entering the emitter of the cascode re-appears at the collector. This means that 99% of the current flowing in *ro* also returns to the collector. Only the lost 1% of the current in *ro* results in a change in the net collector current at the collector terminal. This means that the net effect of *ro* on the collector output impedance in the cascode is roughly like that of a 1-MΩ resistor to ground. This is why the output impedance of cascode stages is so high even though Early effect still is present in the cascode transistor.

$$R_{\text{out}} = \beta * ro$$

$$ro = \frac{VA + V_{\text{ce}}}{I_c}$$

$$ro \approx VA/I_{\text{c}} \qquad \text{at low } V_{\text{ce}}$$

$$R_{\text{out}} = \beta * VA/I_{\text{c}}$$

Notice that the product of β and *VA* is the Early effect figure of merit mentioned previously. The output resistance of a cascode is thus the FOM divided by the collector current.

$$R_{\text{out}} = \text{FOM}/I_{\text{c}}$$

Current Mirror

Figure 2.9a depicts a very useful circuit called a current mirror. If a given amount of current is sourced into Q1, that same amount of current will be sunk by Q2, assuming that the emitter degeneration resistors R1 and R2 are equal, that the transistor V_{be} drops are the same, and that losses through base currents can be ignored. The values of R1 and R2 will often be selected to drop about 100 mV to ensure decent matching in the face of unmatched transistor V_{be} drops, but this is not critical.

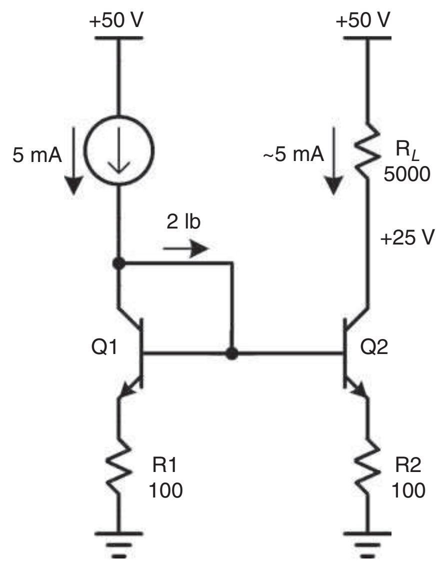

Figure 2.9a Simple current mirror.

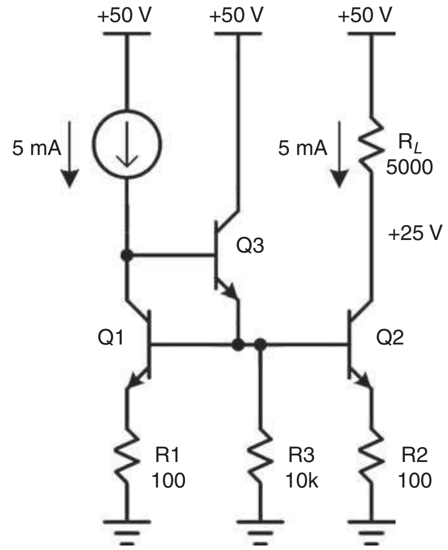

Figure 2.9b Improved current mirror.

If R1 and R2 are made different, a larger or smaller multiple if the input current can be made to flow in the collector of Q2. In practice, the base currents of Q1 and Q2 cause a small error in the output current with respect to the input current. In the example above, if transistor β is 100, the base current I_b of each transistor will be 50 μA, causing a total error of 100 μA, or 2% in the output current.

Figure 2.9b shows a variation of the current mirror that minimizes errors due to the finite current gain of the transistors. Here emitter follower Q3, often called a *helper* transistor, provides current gain to minimize that error. Resistor R3 assures that a small minimum amount of current flows in Q3 even if the current gains of Q1 and Q2 are very high. Note that the input node of the current mirror now sits one V_{be} higher above the supply rail than in Figure 2.9a.

Many other variations of current mirrors exist, such as the *Wilson* current mirror shown in Figure 2.9c. The Wilson current mirror includes transistors Q1, Q2, and Q3. Input current is applied to the base of Q3 and is largely balanced by current flowing in the collector of Q1. Input current that flows into the base of output transistor Q3 will turn Q3 on, with its emitter current flowing through Q2 and R2. Q1 and Q2 form a conventional current mirror. The emitter current of Q3 is mirrored and pulled from the source of input current by Q1.

Any difference between the current of Q1 and the input current is available to drive the base of Q3. If the input current exceeds the mirrored emitter current of Q3, the base voltage of Q3 will increase, causing the emitter current of Q3 to increase and self-correct the situation with feedback action. The equilibrium condition can be seen to be when the input current and the output current are the same, providing an overall 1:1 current mirror function.

Notice that in normal operation all three of the transistors operate at essentially the same current, namely the supplied input current. Ignoring the Early effect, all of the base currents will be the same if the betas are matched. Assume that each base current

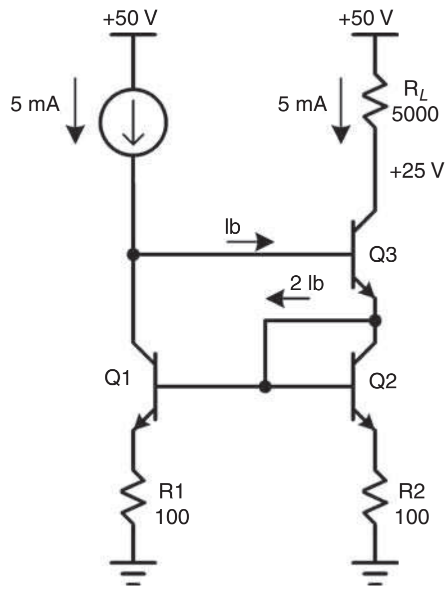

Figure 2.9c Wilson current mirror.

is I_b and that the collector current in Q1 is equal to I. It can be quickly seen that the input current must then be $I + I_b$ and that the emitter current of Q3 must be $I + 2I_b$. It is then evident that the collector current of Q3, which is the output current, will be $I + I_b$, which is the same as the input current. This illustrates the precision of the input-output relationship when the transistors are matched.

Transistor Q3 acts much like a cascode, and this helps the Wilson current mirror to achieve high output impedance. Transistors Q1 and Q2 operate at a low collector voltage, while output transistor Q3 will normally operate at a higher collector voltage. Thus, the Early effect will cause the base current of Q3 to be smaller, and this will result in a slightly higher voltage-dependent output current. This is reflected in the output resistance of the Wilson current mirror.

Current Sources

Current sources are used in many different ways in a power amplifier, and there are many different ways to make a current source. The distinguishing feature of a current source is that it is an element through which a current flows wherein that current is independent of the voltage across that element. The current source in the tail of the differential pair is a good example of its use.

Most current sources are based on the observation that if a known voltage is impressed across a resistor, a known current will flow. A simple current source is shown in Figure 2.10a. The voltage divider composed of R2 and R3 places 2.7 V at the base of Q1. After a V_{be} drop of 0.7 V, about 2 V is impressed across emitter resistor R1. If R1 is a 400-Ω resistor, 5 mA will flow in R1 and very nearly 5 mA will flow in the collector of Q1. The collector current of Q1 will be largely independent of the voltage at the collector of Q1, so the circuit behaves as a decent current source. The load resistance R_L is just shown for purposes of illustration. The output impedance of the current source itself (not including the shunting effect of R_L) will be determined largely by the Early effect in

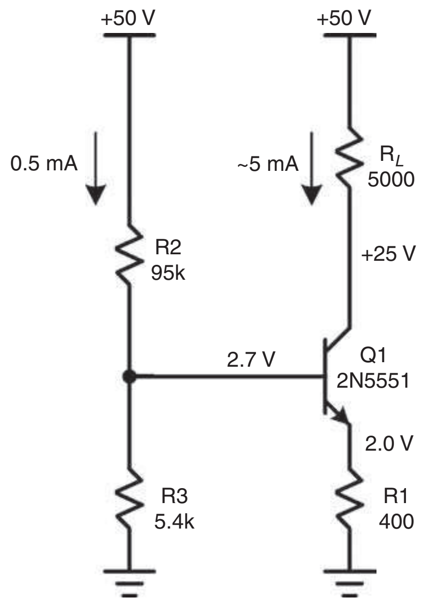

FIGURE 2.10a Simple current source.

the same way as for the CE stage. The output impedance for this current source is found by SPICE simulation to be about 290 kΩ.

In Figure 2.10b, R3 is replaced with a pair of silicon diodes. Here one diode drop is impressed across R1 to generate the desired current. The circuit employs 1N4148 diodes biased with the same 0.5 mA used in the voltage divider in the first example. Together

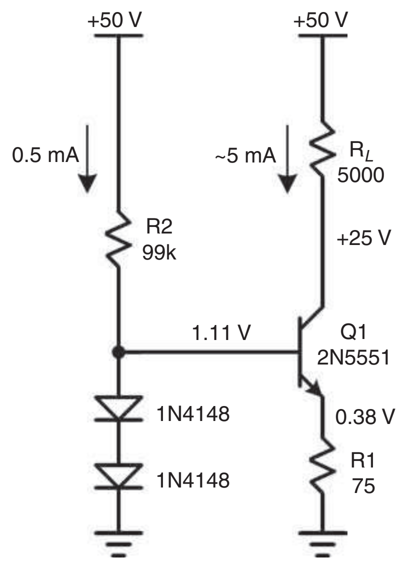

FIGURE 2.10b Current source using diodes.

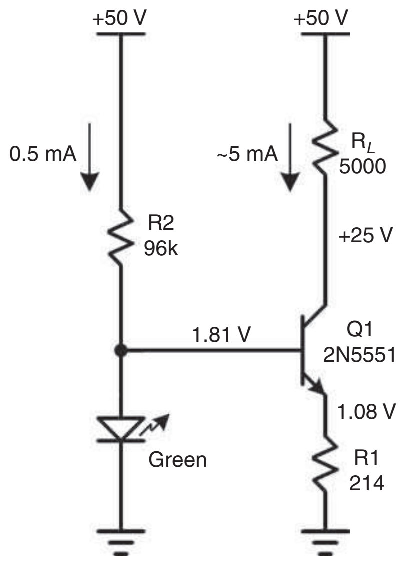

FIGURE 2.10c Current source using LED.

they drop only about 1.1 V, and about 0.38 V is impressed across the 75-Ω resistor R1. The output impedance of this current source is approximately 300 kΩ, about the same as the one above.

Turning to Figure 2.10c, R3 is replaced instead with a Green LED, providing a convenient voltage reference of about 1.8 V, putting about 1.1 V across R1. Once again, 0.5 mA is used to bias the LED. The output impedance of this current source is about 750 kΩ. It is higher than in the design of Figure 2.10b because there is effectively more emitter degeneration for Q1 with the larger value of R1.

R3 is replaced with a 6.2-V *Zener* diode in Figure 2.10d. This puts about 5.5 V across R1. The output impedance of this current source is about 2 MΩ, quite a bit higher than the earlier arrangements due to the larger emitter degeneration for Q1. The price paid here is that the base of the transistor is fully 6.2 V above the supply rail, reducing headroom in some applications.

In Figure 2.10e, a current mirror fed from a known supply voltage is used to implement a current source. Here a 1:1 current mirror is used and 5 mA is supplied from the known power rail. The output impedance of this current source is about 230 kΩ. Only 0.25 V is dropped across R1 (corresponding to 10:1 emitter degeneration), and the base is only 1 V above the rail.

Figure 2.10f illustrates a clever two-transistor feedback circuit that is used to force one V_{be} of voltage drop across R1. It does so by using transistor Q2 to effectively regulate the current of Q1. If the current of Q1 is too large, Q2 will be turned on harder and pull down on the base of Q1, adjusting its current downward appropriately. As in Figure 2.10a through 2.10d, a 0.5-mA current is supplied to bias the current source. This current flows through Q2. The output impedance of this current source is an impressive 3 MΩ. This circuit achieves higher output impedance than the Zener-based version and yet only requires the base of Q1 to be 1.4 V above the power rail. This circuit can also be used to place an overcurrent limit on a CE transistor stage implemented with Q1.

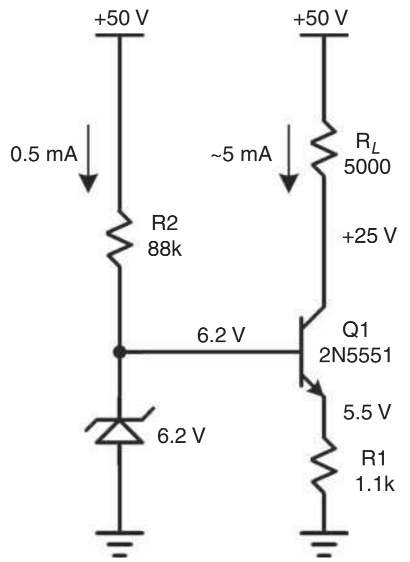

FIGURE 2.10d Current source using Zener diode.

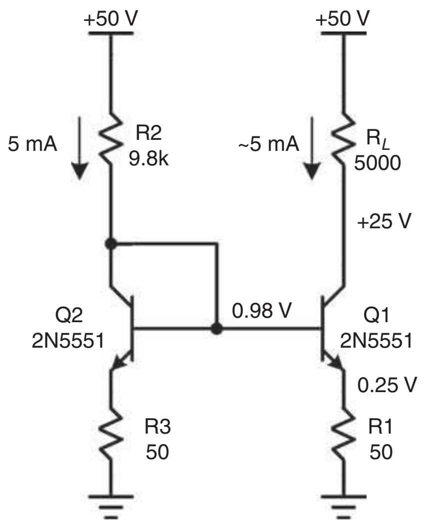

FIGURE 2.10e Current mirror current source.

This circuit will work satisfactorily even if less than 0.5 mA (one-tenth of the output current) is supplied as bias for Q2, but then the output impedance will fall to a lower value and the "quality" of the current source will suffer somewhat. This happens because at lower collector current, Q2 has less transconductance and its feedback control of the current variations in Q1 as a result of the Early effect is less strong. If the bias current is reduced to 0.1 mA, for example, the output impedance falls to about 1 MΩ.

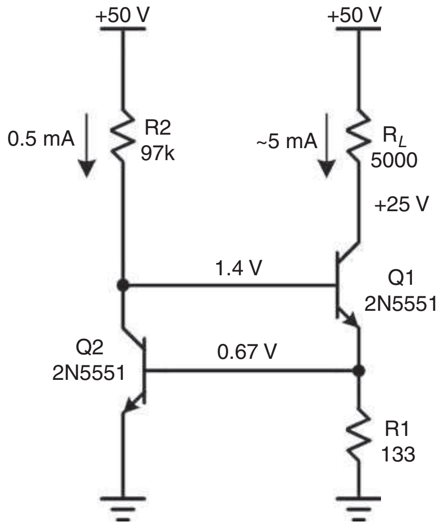

FIGURE 2.10f Feedback current source.

V_{be} Multiplier

Figure 2.11 shows what is called a V_{be} *multiplier*. This circuit is used when a voltage drop equal to some multiple of V_{be} drops is needed. This circuit is most often used as the bias spreader for power amplifier output stages, partly because its voltage is conveniently adjustable.

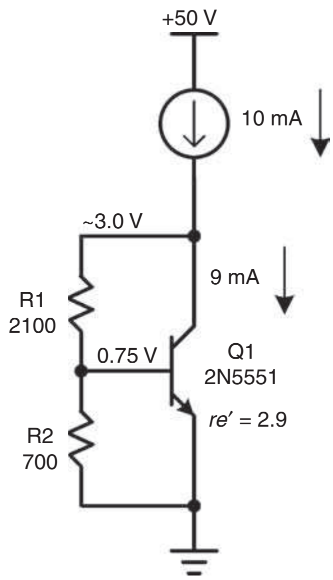

FIGURE 2.11 V_{be} multiplier.

In the circuit shown, the V_{be} of Q1 is multiplied by a factor of approximately 4. Notice that the voltage divider formed by R1 and R2 places about one-fourth of the collector voltage at the base of Q1. Thus, in equilibrium, when the voltage at the collector is at four V_{be}, one V_{be} will be at the base, just enough to turn on Q1 by the amount necessary to carry the current supplied. This is simply a shunt feedback circuit. In this arrangement, about 1 mA flows through the resistive divider while about 9 mA flows through Q1.

When the V_{be} multiplier is used as a bias spreader, R2 will be made adjustable with a trim pot. As R2 is made smaller the amount of bias voltage is increased. Notice that if for some reason R2 fails open, the voltage across the V_{be} multiplier falls to about one V_{be}, failing in the safe direction.

In practice the V_{be} multiplication ratio will be a bit greater than 4 due to the base current required by Q1 as a result of its finite current gain. The extra drop caused across R1 by the base current will slightly increase the collector voltage at equilibrium, making the apparent multiplier factor slightly larger than 4.

The impedance of the V_{be} multiplier is about 4 *re′* for Q1. At 9 mA, *re′* is 2.9 Ω, so ideally the impedance of the multiplier would be about 11.6 Ω. In practice, SPICE simulation shows it to be about 25 Ω. This larger value is mainly a result of the finite current gain of Q1.

The impedance of the V_{be} multiplier rises at high frequencies. This is a result of the fact that the impedance is established by a negative feedback process. The amount of feedback decreases at high frequencies and the impedance-reducing effect is lessened. The impedance of the V_{be} multiplier in Figure 2.11 is up by 3 dB at about 2.3 MHz and doubles for every octave increase in frequency from there. It is thus inductive. For this reason, the V_{be} multiplier is often shunted by a capacitor of 0.1 to 10 μF. A shunt capacitance of as little as 0.1 μF eliminates the increase in impedance at high frequencies.

2.3 Amplifier Design Analysis

Here we apply the understanding of transistors and circuit building blocks to analyze the basic power amplifier. Having accomplished this, we will be well armed to explore, evolve, and analyze the amplifier design steps that will be taken to achieve high performance in the next chapter.

Figure 2.12 is a schematic of a basic 50-W power amplifier that includes the three stages that appear in most solid-state power amplifiers.

- Differential input stage (IPS) comprising Q1–Q3
- Voltage Amplification Stage (VAS) comprising Q4, Q6, and Q7
- Output stage (OPS) comprising Q8–Q11

The design also includes a bias spreader implemented with Q5 connected as a V_{be} multiplier. Some details like coupling capacitors, input networks, and output networks have been left out for simplicity.

The amplifier of Figure 2.12 will be described in simple terms, so that those who are less familiar with circuit design will quickly come to understand its behavior. Those who are already familiar with these concepts can relax and skim through this section.