Bias Circuits
Section 4

Constant Current Sources

IC Bias Circuit Challenges Bias circuits in integrated circuits are a special challenge because of the following:

- Poor tolerances in manufacturing capability
- Poor temperature coefficients
- Limitations on practical component values
- Desire not to have coupling capacitors
- Limited choice of active devices

The basic current mirror consists of two transistors where the base and collector of the reference transistor are shorted together and the emitters of both transistors are tied to ground. The reference transistor, Q1, therefore has a collector voltage of 0.7 volts above ground. The reference current is then determined by the resistor size between the power supply and the collector. Since the bases of the two transistors are shorted together and the emitters of both transistors are grounded, they both have the same base-emitter voltage, $V_{BE}$. It is known from the Shockley diode equation that emitter current and hence approximately the collector current is related together.

$$V_{BE} = \frac{kT}{q} \ln \frac{I_c}{A J_s}$$

$J_s$ is the saturation current for the transistor base-emitter junction. It depends on intrinsic semiconductor parameters and is the same for all devices on the chip. Hence, when two devices have the same $V_{BE}$, the collector currents of the two devices will be proportional to the junction area $A$. Consequently,

$$\frac{I_{C1}}{I_{C2}} = \frac{A_1}{A_2}.$$ 

In a the standard current mirror with both transistor emitters shorted to ground, $V_{BE1} = V_{BE2}$ so that the collector currents in both Q1 and Q2 should be equal. Actually, because of finite transistor $\beta$ and differing emitter areas in the two transistors, there could be some difference between the reference current coming from the power supply of Q1 and
the collector current of Q2. Because of the processing of the two transistors is almost exactly the same, it is assumed the \( \beta \)'s of the two transistors are the same. Kirchoff’s current law says that

\[
I_{ref} = I_{C1} + I_{B1} + I_{B2}
\]

Using this and the area ratio above, gives the output collector current.

\[
I_{C2} = \frac{A_2/A_1 I_{ref}\beta}{1 + \beta + A_2/A_1}
\]

When \( A_1 = A_2 \),

\[
I_{C2} = \frac{I_{ref}}{1 + 2/\beta} \approx I_{ref} \left( 1 - \frac{2}{\beta} \right)
\]

Typically, \( I_{C2} \approx I_{ref} \) to within 5%.

Several currents can be derived from the reference current, in which case

\[
I_{C1} = I_{C2} = I_{C3} \ldots I_{CN} \approx I_{ref} \left( 1 - \frac{N + 1}{\beta} \right)
\]

The above expressions assume that the Early voltage is \( \infty \), which may not be necessarily a good approximation. In the same elementary two transistor current mirror used above, the Early voltage, \( V_A \) is assumed finite while the areas are equal. From the forward biased Schottky diode equation (modified),

\[
I_{C1} = I_S e^{V_{BE}/V_T} \left( 1 + \frac{V_{CE1}}{V_A} \right)
\]

\[
I_{C2} = I_S e^{V_{BE}/V_T} \left( 1 + \frac{V_{CE2}}{V_A} \right)
\]

The ratio of these two currents is

\[
\frac{I_{C2}}{I_{C1}} = \frac{1 + \frac{V_{CE2}}{V_A}}{1 + \frac{V_{CE1}}{V_A}}
\]

From Kirchoff’s current law,

\[
I_{ref} = I_{C1} + 2I_B \approx I_{C1} \left( 1 + \frac{2}{\beta} \right).
\]

Thus for finite Early voltage, the output collector current of Q2 is given in terms of the reference current as follows.

\[
I_{C2} \approx \frac{1 + \frac{V_{CE2}}{V_A}}{1 + \frac{V_{CE1}}{V_A}} \left( 1 - \frac{2}{\beta} \right) I_{ref}
\]
**Base Compensation Current Mirror** In the previous circuit, the reference current is diminished by the base current which in turn is governed by the values of the transistor β’s. Adding a third transistor reduces the loading on the reference current needed to provide the required base currents for Q1 and Q2. If the β for this third transistor is β₃ then

\[
I_{C2} \approx I_{ref} \left(1 - \frac{2}{\beta \beta_3}\right)
\]

The effect on the current mirror is that the effective β’s of Q1 and Q2 are enhanced by β₃. It is however not safe to assume that \(\beta = \beta_3\) since the collector current of Q1 and Q2 is much larger than the collector current of Q3. These results can be summarized in the following chart.

<table>
<thead>
<tr>
<th>Comment</th>
<th>UN compensated</th>
<th>Base Compensation</th>
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</thead>
<tbody>
<tr>
<td>(I_{C2})</td>
<td>(I_{ref} \frac{1 + 2}{\beta})</td>
<td>(I_{ref} \left(1 + \frac{2}{\beta (1 + \beta_3)}\right))</td>
</tr>
<tr>
<td>(I_{C_2})</td>
<td>(I_{ref} \left(1 - \frac{2}{\beta}\right))</td>
<td>(I_{ref} \left(1 - \frac{2}{\beta \beta_3}\right))</td>
</tr>
<tr>
<td>(I_{C_1} \ldots I_{C_n})</td>
<td>(I_{ref} \left(1 - \frac{N + 1}{\beta}\right))</td>
<td>(I_{ref} \left(1 - \frac{N + 1}{\beta \beta_3}\right))</td>
</tr>
<tr>
<td>(A_1 \neq A_2 \cdot I_{C_2})</td>
<td>(I_{ref} \frac{A_2}{A_1} \left(1 - \frac{1 + A_2/A_1}{\beta}\right))</td>
<td>(I_{ref} \frac{A_2}{A_1} \left(1 - \frac{1 + A_2/A_1}{\beta \beta_3}\right))</td>
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**Resistor Ratioed Current Mirror** The standard current mirror can be modified by putting a resistance \(R_1\) in the emitter of Q1 and \(R_2\) in the emitter of Q2. Consequently, the voltage between the base of each transistor to ground is the same, but \(V_{BE1} \neq V_{BE2}\). Since this is typically a small but critical difference, the \(V_{BE}\) voltages can no longer be assumed to be merely 0.6 Volts. The Shockley diode equation must be used to relate current to \(V_{BE}\). After using Kirchoff’s voltage between the base and ground of each transistor, ie.

\[
I_{ref}R_1 + V_{BE1} = I_{C_1}R_2 + V_{BE2}.
\]

As a result, the output current relative to the reference current is given as follows.

\[
I_{C_2} = I_{ref} \frac{R_1}{R_2} \left[1 - \frac{V_T \ln \frac{I_{C_2}}{I_{ref}}}{I_{ref} R_1}\right]
\]

To get low output currents without using large resistances (that are costly in real estate) the value of \(R_1\) can be set to 0 in the above equation. This is the so called Widlar current source.

\[
I_{C_2} = \frac{V_T}{R_2} \ln \frac{I_{ref}}{I_{C_2}}
\]

To achieve a 100:1 current ratio, ie. \(I_{ref} = 1\ mA, C_2 = 10\ \mu A\), then a resistance value of \(R_2 = 12k\Omega\).
Wilson Current Mirror The Wilson current mirror like the base compensation circuit has three transistors and has the effect of reducing the difference between the reference current and the output current, $I_{C2}$. The output current is given below.

$$I_{C2} = I_{ref} \left[ 1 - \frac{2}{\beta^2 + 2\beta + 2} \right]$$

The error is proportional to $2/\beta^2$. This circuit has a very high output resistance (that could be obtained using the hybrid $\pi$ circuit).

$$R_{out} \approx \frac{\beta V_A}{2}$$

The Thevenin voltage is also high,

$$V_{Thev} = \frac{\beta V_A}{2}$$

so that the Wilson mirror has qualities for being a good current source. In general, the Widlar current source is useful for obtaining small currents, and the Wilson current source is noted for having a high output resistance and low sensitivity to base currents.

Supply Independent Bias There is still some dependence of the output current on the value of the power supply voltage, $V_{CC}$. If the bias current is made to depend ultimately on some transistor $V_{BE}$ rather than on $V_{CC}$, then the output current would be independent of the supply voltage. When a resistor, $R_2$ is put into the emitter of the output transistor, Q2, then the reference current is given as follows.

$$I_{ref} = \frac{V_{CC} - 2V_{BE}}{R_2} = I_1 \approx \frac{V_{CC}}{R_1}$$

The output current that travels through the collector of Q2 is

$$I_{out} = \frac{V_{BE}}{R_2}$$

and

$$V_{BE} = V_t \ln \left( \frac{I_1}{I_s} \right).$$

Consequently, the output current given here,

$$I_{out} = \frac{V_t \ln(I_1/I_s)}{R_2} \approx \frac{V_t}{R_2} \ln \left( \frac{V_{CC}}{R_1 I_s} \right)$$

depends only on the $\ln(V_{CC})$ rather than directly on $V_{CC}$. While the voltage sensitivity has been improved, it is still dependent on temperature through $V_t$. 

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Bootstrap or Self Bias Circuit This circuit consists of 6 transistors, and also has improved current sensitivity with respect to changes in the supply voltage. In this circuit, assume that $\beta = \infty$ and the Early voltage $V_A = \infty$. As a result, the base currents are zero. The $V_{BE1}$ sets the current $I_2$.

$$I_2 = \frac{V_{BE1}}{R_2} = \frac{V_t}{R_2} \ln \left( \frac{I_1}{I_2} \right)$$

The transistors Q4 and Q5 act as a current mirror with reference current being the $I_2$ just obtained. This sets the value of $I_1 \approx I_2$. There are now two equations that relate $I_1$ to $I_2$. When both are plotted on a curve of $I_2$ vs. $I_1$, the intersection of the two curves gives the operating point of Q2. Once $I_2$ is known, the current mirror sets the output current.

$$I_{out} \approx I_1$$

Since $I_2$ is also a possible stable solution it is necessary to insure that the current is $\neq 0$ by making sure some current is fed to the collector of Q1.

An estimate of the temperature stability, $S_T$, is given by

$$S_T = \frac{1}{I_{out}} \frac{\partial I_{out}}{\partial T}$$

Since,

$$I_{out} \approx I_1 \approx I_2 \approx \frac{V_{BE1}}{R_2}$$

then

$$\frac{\partial I_{out}}{\partial T} = \frac{1}{R_2} \frac{\partial V_{BE1}}{\partial T} - \frac{V_{BE1}}{R_2^2} \frac{\partial R_2}{\partial T} = \frac{I_{out}}{V_{BE1}} \frac{\partial V_{BE1}}{\partial T} - \frac{I_{out}}{R_2} \frac{\partial R_2}{\partial T}.$$ 

The temperature sensitivity

$$S_T = \frac{1}{V_{BE1}} \frac{\partial V_{BE1}}{\partial T} - \frac{1}{R_2} \frac{\partial R_2}{\partial T}$$

has two derivatives with respect to temperature change. The first derivative is negative (as described later) in spite of the $kT/q$ in front of the expression for $V_{BE1}$. The resistance nearly always increases with temperature so the second derivative is positive. The two temperature sensitivities tend to add thereby producing poor temperature sensitivity.

Low Impedance Voltage Source An ideal current source has a high output impedance and an ideal voltage source has a low output impedance. Low output impedance is associated with the emitter follower circuit.

DC Level Shift Circuits In discrete amplifiers, coupling capacitors are used to provide a DC block between stages. That means the same bias levels can be used for each stage. Without the blocking capacitors, the voltage levels in succeeding stages would have to
keep rising in order to maintain appropriate bias levels for each stage. However, blocking capacitors are not economical in an integrated circuit, so other approaches are needed for level shifting. Furthermore, to prevent loading the output of a previous stage and provide a good voltage source to the succeeding stage, the circuit needs a high input impedance and a low output impedance. The emitter follower circuit fits this criterion. There are four emitter follower type circuits that have these characteristics.

a.) The first one is merely a voltage divider in the emitter circuit. The output voltage is

\[ V_2 = \frac{(V_1 - V_{BE})}{R_1 + R_2}. \]

In this case the AC voltage gain is \(< 1\).

b.) In this second circuit, the resistor closest to the emitter \( (R_1) \) is replaced by a Zener diode. In this circuit,

\[ V_2 = V_1 - V_{BE} - V_Z \]

where \( V_Z \) is the Zener voltage which typically is in the 6 - 9 volt range. The Zener diode is based on avalanche under high reverse bias, so is noisy.

c.) The Zener diode can be replaced with \( n \) forward biased diodes, each diode having a forward voltage drop of \( V_{BE} \). The output voltage is

\[ V_2 = V_1 - (n+1)V_{BE}. \]

As in the emitter follower, the output resistance is

\[ R_{out} = \frac{1}{g_m} + \frac{R_{bb} + r_b}{1 + \beta} \approx (n+1) \frac{V_t}{I_c}. \]

Each diode needs an isolation pocket and therefore takes up additional room. Also, the output voltage, \( V_2 \) is temperature dependent.

d.) The \( V_{BE} \) multiplier replaces the series of forward biased diodes with a transistor that is used to control the current of the emitter follower. The collector-emitter voltage of this added transistor is the same as the voltage drop through the series \( R_1, R_2 \) used to bias this transistor.

\[ V_{CE} = I_2(R_1 + R_2) \]

The value \( I_2 \) is the current going through the two resistors. Assuming the transistor \( \beta \) is high, \( I_2 \) is also given by the following.

\[ I_2 = \frac{V_{BE}}{R_2} \]
Eliminating $I_2$ between these two equations gives an expression for $V_{CE}$.

$$V_{CE} = V_{BE} \left( 1 + \frac{R_1}{R_2} \right)$$

The output voltage is given as follows.

$$V_2 = V_1 - V_{BE} - V_{CE}$$

$$= V_1 - V_{BE} \left( 2 + \frac{R_1}{R_2} \right)$$

**Temperature Independent Biasing** A voltage source and a temperature independent source are two different types of circuits with different goals. A voltage source is a low impedance source. A temperature independent source seeks to provide a voltage that does not drift with temperature. Monolithic circuit components typically drift a few thousand ppm /°C (parts per million). In particular, resistors, except thin film types, have bad temperature drift characteristics. However, it is possible to achieve temperature stability with 100 to 20 ppm/°C in a monolithic circuit with appropriate circuit design.

One circuit that has two forward biased diodes and two resistors in the emitter of the transistor develops a reference voltage across $R_2$ and a diode above ground. This voltage is

$$V_{ref} = \frac{R_2 V_Z + V_{BE}(R_1 - 2R_2)}{R_1 + R_2}$$

Assuming the major portion of the temperature stability is associated with the active devices, the change in the reference voltage should ideally be zero.

$$\frac{\partial V_{ref}}{\partial T} = \frac{1}{R_1 + R_2} \left[ R_2 \frac{\partial V_Z}{\partial T} + (R_1 - 2R_2) \frac{\partial V_{BE}}{\partial T} \right] = 0$$

To achieve temperature stability,

$$\frac{R_1 - 2R_2}{R_2} = - \frac{\partial V_Z}{\partial T} \frac{\partial V_{BE}}{\partial T}$$

Typically, $\frac{\partial V_Z}{\partial T} = +2.5 \text{ mV/°C}$ and $\frac{\partial V_{BE}}{\partial T} = -2 \text{ mV/°C}$. In this latter expression, it must be remembered that $n_i$ which is buried in the saturation current term for $V_{BE}$ is also highly temperature dependent so that in the end it has a negative temperature coefficient. In choosing resistor values, $R_1 > 2R_2$, are chosen to minimize temperature instability as suggested. The main problem in using Zener diodes for temperature stability is their intrinsic noise and their requirement of typically at least 10 V to operate properly.

**Band-gap Reference Circuits** There are a variety of circuits that produce a reference voltage which is of the form

$$V_{ref} = V_{BE} + kV_t$$
where $V_{BE}$ has a negative temperature coefficient and $V_t$ has a positive temperature coefficient. By appropriate choice of $k$ (typically dependent of a resistance ratio), low temperature sensitivity can be achieved.

**Active Load** The typical common emitter amplifiers have a voltage gain of $-g_m R_L$. For high gain, a large load resistance is needed, but these are difficult to make in an integrated circuit format. A current mirror source can be used as a load, and these are easier to make than actual resistors. If the input transistor is Q1 and the current mirror is made up of two PNP transistors Q2 and Q3. When the Early voltage for the NPN transistor is $V_{AN}$ and for the PNP transistors $V_{AP}$. After some analysis, the output voltage across the collector-emitter terminals of the input transistor Q1 is

$$V_o = (V_{CC} - V_{BE3}) \frac{V_{AN}}{V_{AN} + V_{AP}}.$$ 

A small signal analysis shows that the output resistance and voltage gain are given below.

$$R_{out} = r_o1 || r_o2$$

$$A_v = -g_m (r_o1 || r_o2 || R_L)$$

This circuit can be unstable, but this can be overcome by using a differential amplifier.

**Offset Voltage for Active Loads** The offset voltage is

$$V_{os} = V_{BE1} - V_{BE2} = V_t \ln \frac{I_{C1}}{I_{C2}}$$

As was found for case with resistor loads,

$$\Delta I_C = I_{C1} - I_{C2}$$

$$I_C = \frac{1}{2} (I_{C1} + I_{C2}).$$

The offset voltage is then found.

$$V_{os} = V_t \left[ \frac{2}{\beta} + \frac{I_L}{I_C} \right]$$

A typical offset voltage is about 2 mV while that for one with resistor loads is 1.5 mV. If base compensation is used in this circuit by adding another PNP transistor, Q5. The offset voltage with this circuit is

$$V_{os} = \frac{V_t2}{1 + \beta_3 \beta_5}.$$ 

For $\beta_3 = 25$ and $\beta_5 = 5$ the offset voltage is only $416 \mu V$. 

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