CHAPTER 8

Building Blocks of Integrated-Circuit Amplifiers
Figure 8.1 Circuit for a basic MOSFET constant-current source. For proper operation, the output terminal, that is, the drain of $Q_2$, must be connected to a circuit that ensures that $Q_2$ operates in saturation.

Figure 8.2 Basic MOSFET current mirror. $Q_2$ should be in saturation, $V_{GD} < V_t$

$I_O \approx I_{REF}$ for matched MOSFETS
MOSFET current mirror

\[
\frac{I_O}{I_{REF}} = \frac{(W/L)_2}{(W/L)_1}
\]

Easy to change \(I_O\) by choosing \(W_2\)

Channel-length modulation (Early) effect:

1) finite output resistance

\[
r_{o2} = \frac{V_{A2}}{I_O} = \frac{1}{\lambda_2 I_O}
\]

2) (in other words) \(I_O\) slightly depends on \(V_O\)

\[
I_O = \frac{(W/L)_2}{(W/L)_1} I_{REF} \left(1 + \frac{V_O - V_{GS}}{V_{A2}}\right)
\]

Figure 8.2 Basic MOSFET current mirror.
Figure 8.4 A current-steering circuit.
Figure 8.7 The basic BJT current mirror.

Figure 8.9 A simple BJT current source.

\[ I_O \approx I_{REF} \] for matched BJTs
Figure 8.10 Generation of a number of constant currents of various magnitudes.
BJT current mirror

\[ I_{\text{REF}} = I_C + 2 \frac{I_C}{\beta} = I_C \left(1 + \frac{2}{\beta}\right) \]

\[ \frac{I_O}{I_{\text{REF}}} = \frac{1}{1 + 2/\beta} \approx 1 - \frac{2}{\beta} \]

With Early effect:

\[ I_O = \frac{I_{\text{REF}}}{1 + 2/\beta} \left(1 + \frac{V_O - V_{BE}}{V_{A2}}\right) \]

Output resistance \( R_o = r_{o2} = V_{A2}/I_O \)

Figure 8.8 Analysis of the current mirror taking into account the finite \( \beta \) of the BJTs.
BJT current mirror with base-current compensation

Instead of $I_O \approx (1 - 2/\beta) I_{REF}$ for simple mirror, it is now better:

$$I_{REF} = I_O \left[1 + \frac{2}{\beta(\beta + 1)}\right]$$

$$I_O \approx \left(1 - \frac{2}{\beta^2}\right) I_{REF}$$

However, output resistance is still not improved, $R_o = r_{o2}$

Figure 8.11 A current mirror with base-current compensation.
BJT Wilson mirror

Figure 8.40 The Wilson bipolar current mirror: circuit showing analysis to determine the current transfer ratio

\[ I_o = I_c \frac{(1 + 2/\beta)\beta}{\beta + 1} \]

Reduces inaccuracy of \( I_o/I_{REF} \) and also improves (increases) \( R_o \)

\[ I_o \approx \left(1 - \frac{2}{\beta^2}\right) I_{REF} \]

\[ R_o = \beta_3 r_{o3}/2 \]

(derivations are not trivial)

We will discuss some other improved current mirrors later
Basic IC design philosophy: resistors are expensive (especially large resistances), transistors are cheap. Try to avoid resistors (do as much as possible with transistors).

Idea of active load: replace load resistors with transistors or with transistor-based circuits (current mirrors).

\[ A_v = -g_m (R_D \parallel r_o) \]

Wish to increase \( R_D \)

Use transistor(s) to create current source \( I \).

Then \( A_v = -g_{m1} (r_{o2} \parallel r_{o1}) \).
Figure 8.15 (a) The CS amplifier with the current-source load implemented with a $p$-channel MOSFET $Q_2$; (b) the circuit with $Q_2$ replaced with its large-signal model; and (c) small-signal equivalent circuit of the amplifier.

\[ A_v = -g_{m1}(r_{o1} \parallel r_{o2}) \]

(dc value of the output voltage is not obvious)
CS amplifier with current mirror as active load

\[ A_v = -g_{m1} R_o \]

\[ R_o = r_{o1} \parallel r_{o2} \]

\[ R_i = \infty \]

\[ A_v = -g_{m1} (r_{o1} \parallel r_{o2}) \]

((dc value of the output voltage is not obvious)

Figure 8.16
CG amplifier with current mirror as active load

\[ A_v = \left( g_{m1} + \frac{1}{r_{o1}} \right) (r_{o1} \parallel r_{o2}) \]

\[ \approx g_{m1} (r_{o1} \parallel r_{o2}) \]

\[ R_o = r_{o1} \parallel r_{o2} \]

\[ R_i \approx \frac{1}{g_{m1}} \left( 1 + \frac{r_{o2}}{r_{o1}} \right) \]
Figure 8.45 (a) A source follower biased with a current mirror $Q_2$–$Q_3$ and with the body terminal indicated. Note that the source cannot be connected to the body and thus the body effect should be taken into account. (b) Equivalent circuit.

$$A_v = \frac{g_{m1}}{g_{m1} + 1/r_{o1} + 1/r_{o3}} \approx 1$$

$$R_i \approx \infty$$

$$R_o = \frac{1}{g_{m1} \parallel r_{o1} \parallel r_{o3}} \approx \frac{1}{g_{m1}}$$

Actually, the body effect is important, then

$$A_v \approx \frac{1}{1 + \chi} = \frac{1}{1 + g_{mb1}/g_{m1}}$$
Next subject: MOS cascode

MOS cascode: common-source stage loaded with common-gate stage (similar with BJT: CE loaded with CB)

Idea: 1) increase output resistance ⇒ increase voltage gain
2) fast circuit because $Q_1$ is loaded with a rather small $1/g_{m2}$

$$A_v = -G_m R_o \quad G_m \approx g_{m1}$$

$$R_o \approx g_{m2} r_{o1} r_{o2}$$

More accurately,

$$G_m = \frac{g_{m2} + 1/r_{o2}}{g_{m2} + 1/r_{o1} + 1/r_{o2}} \quad g_{m1}$$

$$R_o = r_{o1} + r_{o2} + g_{m2} r_{o1} r_{o2}$$

Figure 8.30 (a) A MOS cascode amplifier with an ideal current-source load
Derivation of output resistance $R_o$ for cascode

\[ v_{gs2} = -i_x r_{o1} \]

\[ v_x = i_x r_{o1} + r_{o2} (i_x - g_{m2} v_{gs2}) \]
\[ = i_x r_{o1} + r_{o2} i_x (1 + g_{m2} r_{o1}) \]
\[ = i_x [r_{o1} + r_{o2} (1 + g_{m2} r_{o1})] \]

\[ R_o = r_{o1} + r_{o2} + g_{m2} r_{o1} r_{o2} \approx g_{m2} r_{o1} r_{o2} \]

usual output resistance $r_{o2}$ is increased by $g_{m2} r_{o1}$

Similarly, for a BJT cascode (will need later)

\[ R_o = (r_{o1} \parallel r_{\pi2}) + r_{o2} + g_{m2} (r_{o1} \parallel r_{\pi2}) r_{o2} \]

\[ \approx g_{m2} r_{\pi2} r_{o2} = \beta_2 r_{o2} \]

$r_{o2}$ is increased by factor $\beta_2$
MOS cascode with ideal current source

\[ R_o \approx g_{m2} r_{o1} r_{o2} \]
\[ A_v \approx -g_{m1} R_o \approx -g_{m1} r_{o1} g_{m2} r_{o2} \]

As if two stages of amplification, but faster operation (first transistor loaded with small \(1/g_{m2}\)).

Actually, needs a very good current source (with output resistance comparable to \(R_o\)). Simple current mirror is not good enough (only \(r_o\)), \(\Rightarrow\) we need either an improved current mirror (discuss later) or another cascode.

Figure 8.30 (a) A MOS cascode amplifier with an ideal current-source load
Figure 8.31 (a) A MOS cascode amplifier
loaded in a simple PMOS current source $Q_3$.

Output resistance is $R_o \parallel r_{o3}$. Voltage gain is limited by $r_{o3}$.

Not quite good for voltage gain, but still fast.
Figure 8.33 A cascode amplifier with a cascode current-source load.

MOS cascode with cascode current source

\[ R_{op} = g_{m3}r_{o3}r_{o4} \]

\[ R_{on} = g_{m2}r_{o1}r_{o2} \]

\[ A_v = -g_{m1}(R_{on} \parallel R_{op}) \]

If all \( g_m \) are equal and all \( r_o \) are equal,

then \[ A_v = -\frac{1}{2}(g_mr_o)^2 \]
MOS double cascode

Each time increase output resistance

\[ r_{o2} \rightarrow (g_{m2} r_{o2}) r_{01} \]
\[ \rightarrow (g_{m3} r_{o3})(g_{m2} r_{o2}) r_{01} \]

\[ A_v \approx -(g_m r_o)^3 \quad \text{if ideal current source} \]

\[ \text{if double-cascode as the current source, then } \times \frac{1}{2} \]

Figure 8.35 Double cascoding.
MOS folded cascode

PMOS common gate load.

Equivalent to usual cascode. Avoids "stacking" (requiring too large voltage).

Figure 8.36 The folded cascode.
BJT cascode

\[ R_o = (r_{o1} \parallel r_{\pi2}) + r_{o2} + g_{m2} r_{o2} (r_{o1} \parallel r_{\pi2}) \]

\[ \approx g_{m2} r_{\pi2} r_{o2} = \beta_2 r_{o2} \]

(as derived earlier)

\[ A_v \approx -g_{m1} R_o \approx -g_{m1} \beta_2 r_{o2} \]

Impossible to double-cascode because \( R_o \) would still be the same

(though can double-cascode with MOSFET in BiCMOS technology)

Figure 8.37 (a) A BJT cascode amplifier with an ideal current-source load;

Needs “good” current source (otherwise significantly less \( A_v \))
BJT cascode with cascode current source

\[ A_v = -g_{m1} \left( \beta_2 r_{o2} \parallel \beta_3 r_{o3} \right) \]

**Figure 8.38** A BJT cascode amplifier with a cascode current source.
BiCMOS cascode

MOS loaded with BJT:
large input resistance of MOS and
large output resistance of BJT.
Also, faster (loaded with $r_e$).

$$A_v = -g_{m1} \beta_2 r_{o2}$$

Figure P8.81
Next subject: improved current mirrors

Basic BJT current mirror

\[ I_0 \approx (1 - 2/\beta)I_{REF} \]

\[ R_o = r_{o2} \text{ (not large)} \]

Wilson BJT current mirror

\[ I_0 \approx (1 - 2/\beta^2)I_{REF} \quad \text{(better)} \]

\[ R_o = \beta_3 r_{o3}/2 \quad \text{(larger)} \]
Wilson MOS mirror

Figure 8.41 The Wilson MOS mirror: (a) circuit; (c) modified circuit.

\[ R_o = g_{m3} r_{o3} r_{o2} \]

Derivation is not trivial.
Result looks similar to cascode, but with different transistor

Extra transistor to balance (so that the same voltages in both branches)
Figure 8.39 A cascode MOS current mirror.

\[ R_o = g_{m3} r_{o3} r_{o2} \]

Drawback: stacked transistors, “eats up” more voltage,

\[ V_o > V_t + 2 V_{OV} \]

(since \( V_{G3} = 2( V_t + 2V_{OV} ) \) from \( Q_1 \) and \( Q_4 \))

(same drawback for Wilson mirror)
Wildar current source

Resistor $R_E$ decreases $V_{BE2} \Rightarrow$ decreases $I_O$

Assume matched transistors

$$V_{BE1} - V_{BE2} = V_T \ln \frac{I_{REF}}{I_O}$$

$$I_O R_E = V_T \ln \frac{I_{REF}}{I_O}$$

$$R_o = r_0 \left[ 1 + g_m (R_E \parallel r_\pi) \right]$$

(increased output resistance)

Figure 8.42 The Widlar current source.
Next subject: Some useful transistor pairings

(a) CC–CE: increases $R_i$ (due to emitter follower), makes faster (not obvious)

(b) The same with MOS. Faster (no improvement of $R_i$)

(c) The same in BiCMOS: better $R_i$ than in (a), better $g_m$ than in (b)
Figure 8.47 (a) The Darlington configuration.

\[ \beta = \beta_1 \beta_2 \]
CC-CB (CD-CG) configuration

Figure 8.48 (a) A CC–CB amplifier. (b) Another version of the CC–CB circuit with $Q_2$ implemented using a pnp transistor. (c) The MOSFET version of the circuit in (a).

(a) CC-CB: fast because of CB, while large $R_i$ because of the follower
(b) The same with pnp BJT for CB
(c) MOSFET version of (a)