CHAPTER 8

Building Blocks of Integrated-Circuit Amplifiers

Basic MOSFET current source (current mirror)

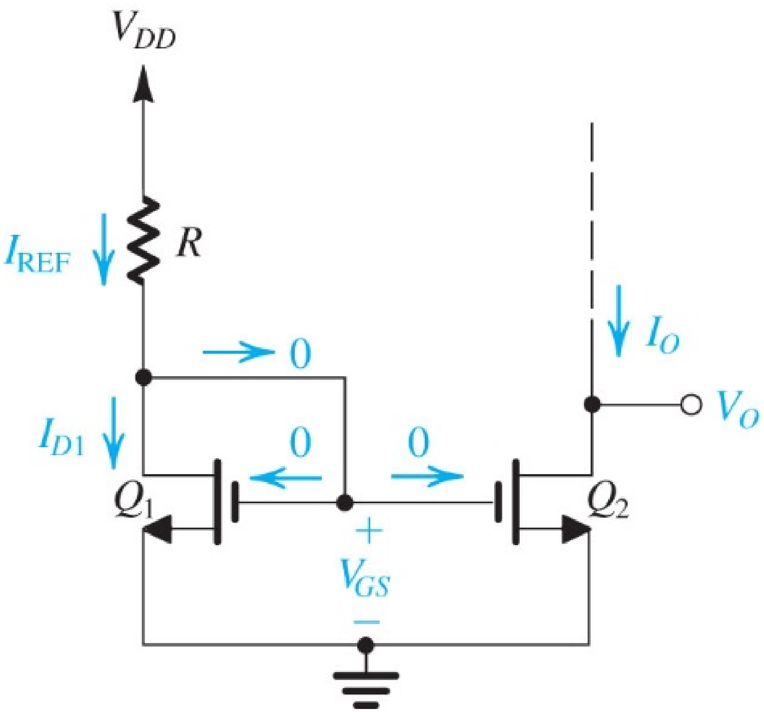


Figure 8.1 Circuit for a basic MOSFET constant-current source. For proper operation, the output terminal, that is, the drain of Q_2, must be connected to a circuit that ensures that Q_2 operates in saturation.

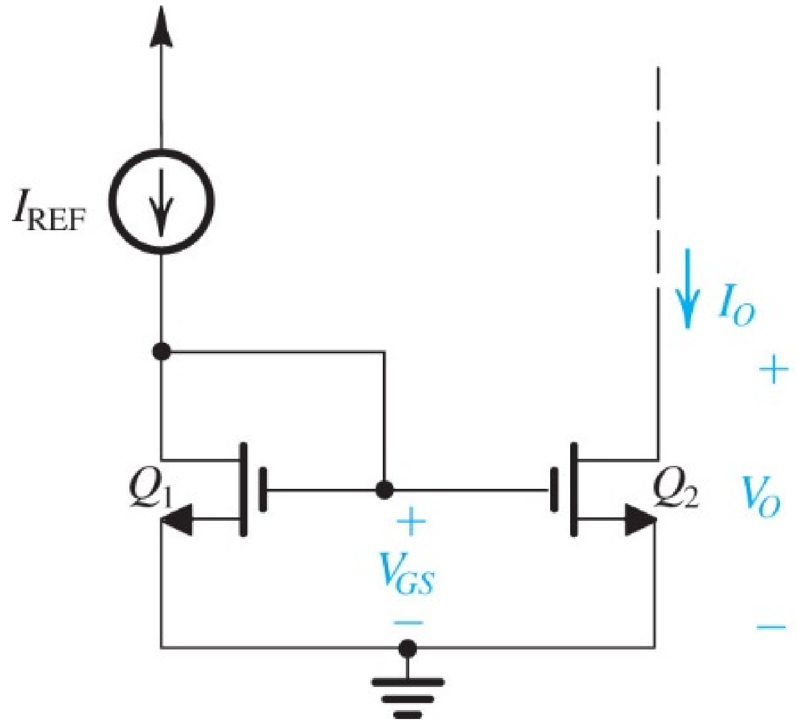


Figure 8.2 Basic MOSFET current mirror.

Q_2 should be in saturation,

$$V_{GD} < V_t$$

$I_O \approx I_{REF}$ for matched MOSFETS

MOSFET current mirror

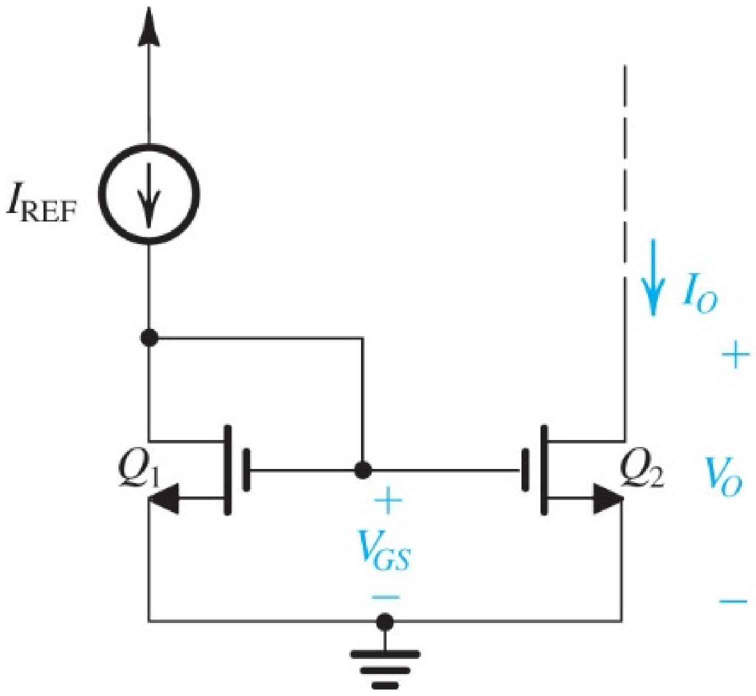


Figure 8.2 Basic MOSFET current mirror.

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

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

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

Current-steering circuit

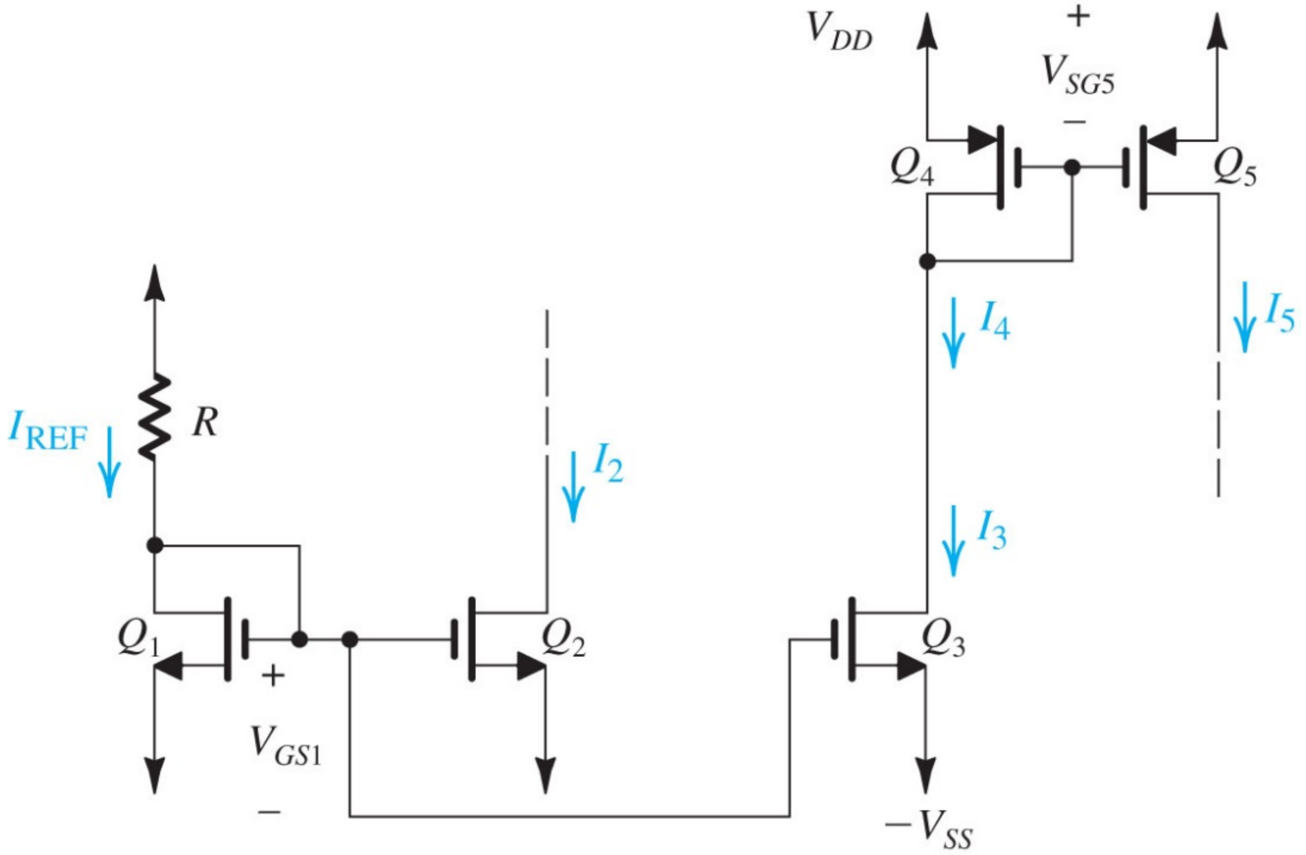


Figure 8.4 A current-steering circuit.

BJT current mirror

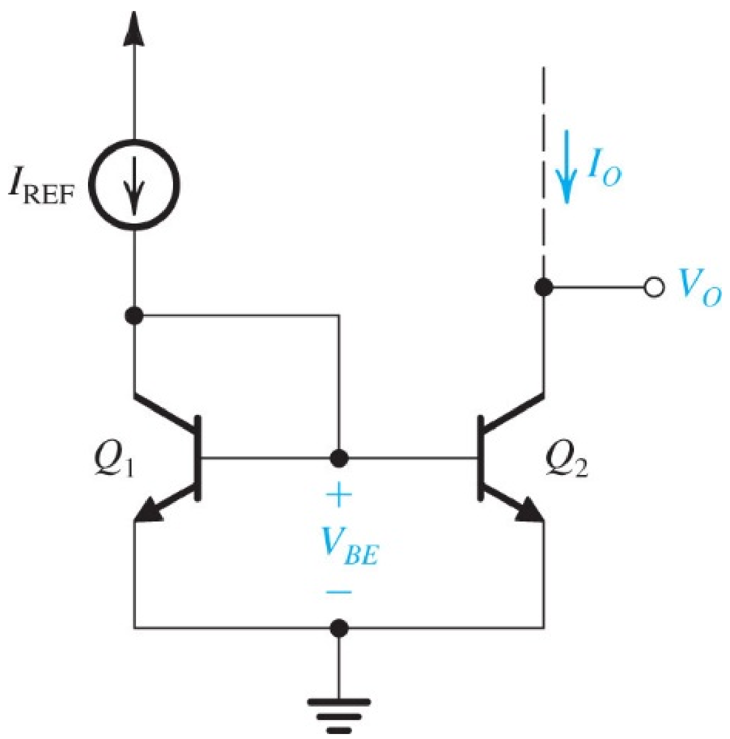


Figure 8.7 The basic BJT current mirror.

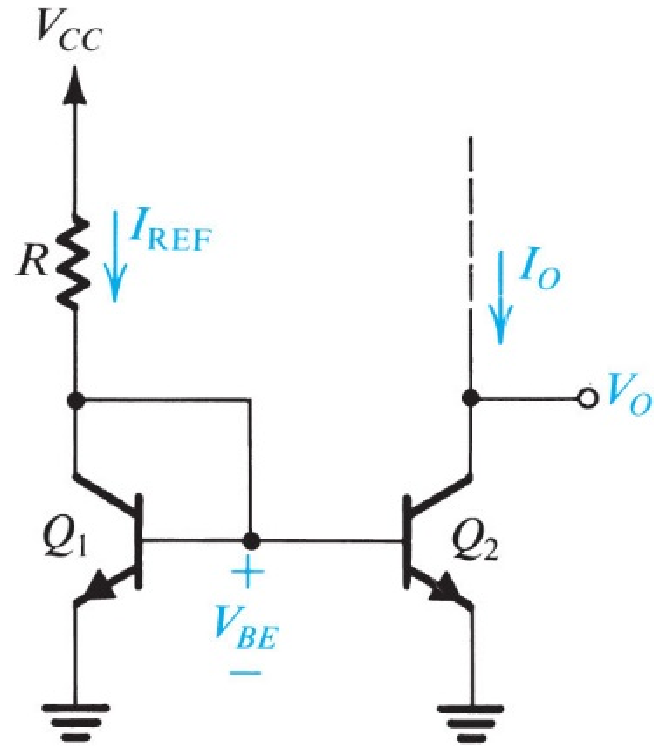


Figure 8.9 A simple BJT current source.

$I_O \approx I_{REF}$ for matched BJTs

Current steering with BJTs

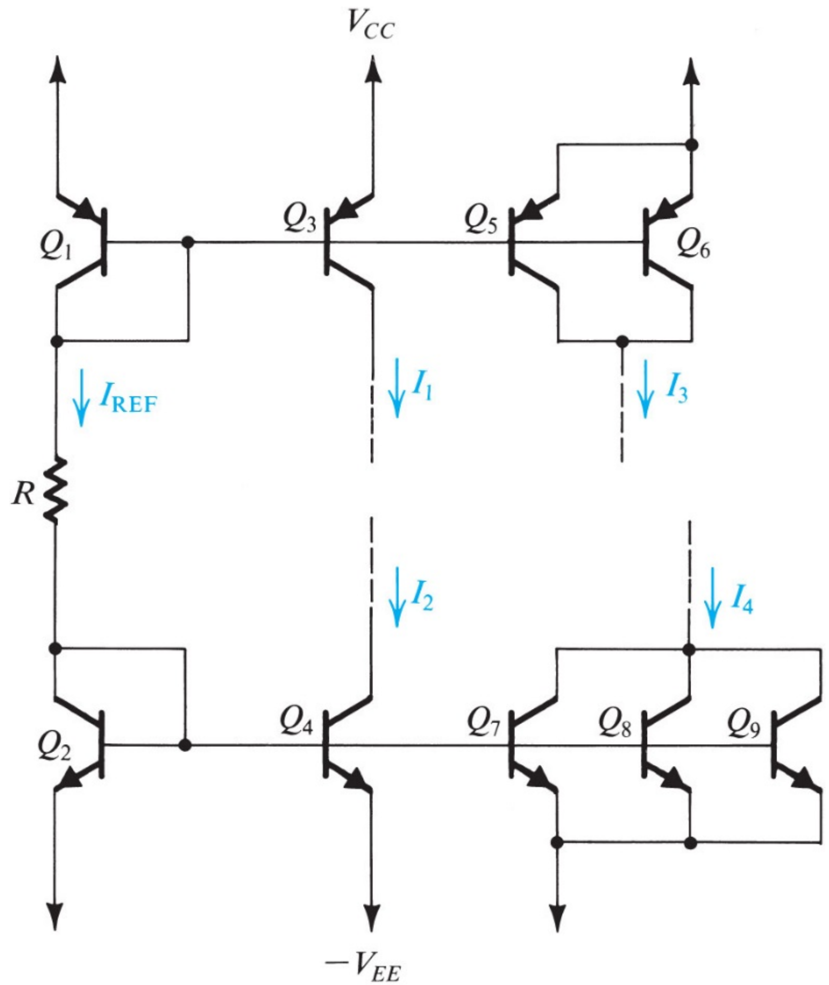


Figure 8.10 Generation of a number of constant currents of various magnitudes.

BJT current mirror

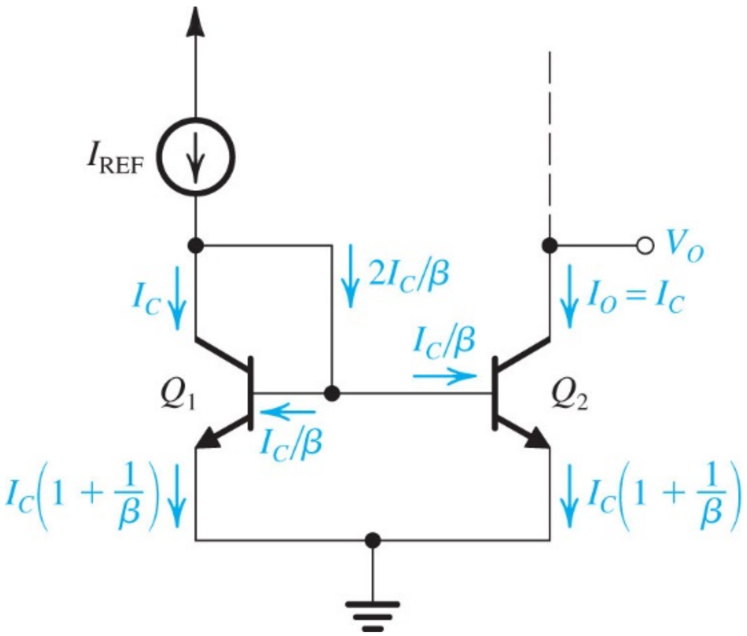


Figure 8.8 Analysis of the current mirror taking into account the finite β of the BJTs.

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

$$\frac{I_O}{I_{REF}} = \frac{1}{1 + 2/\beta} \approx 1 - \frac{2}{\beta}$$

With Early effect:

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

Output resistance $R_o = r_{o2} = V_{A2}/I_O$

BJT current mirror with base-current compensation

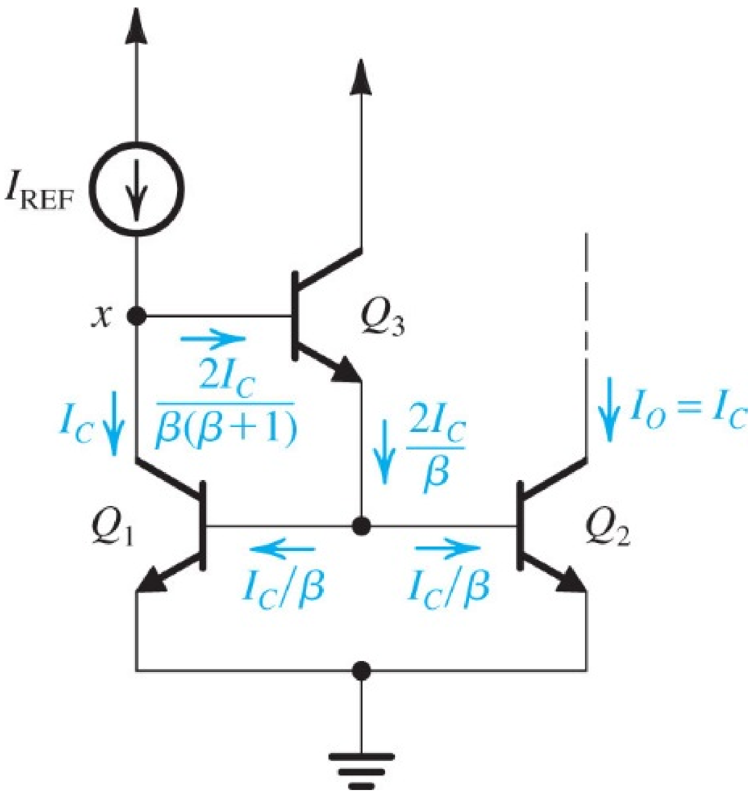


Figure 8.11 A current mirror with base-current compensation.

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, $R_o = r_{o2}$

BJT Wilson mirror

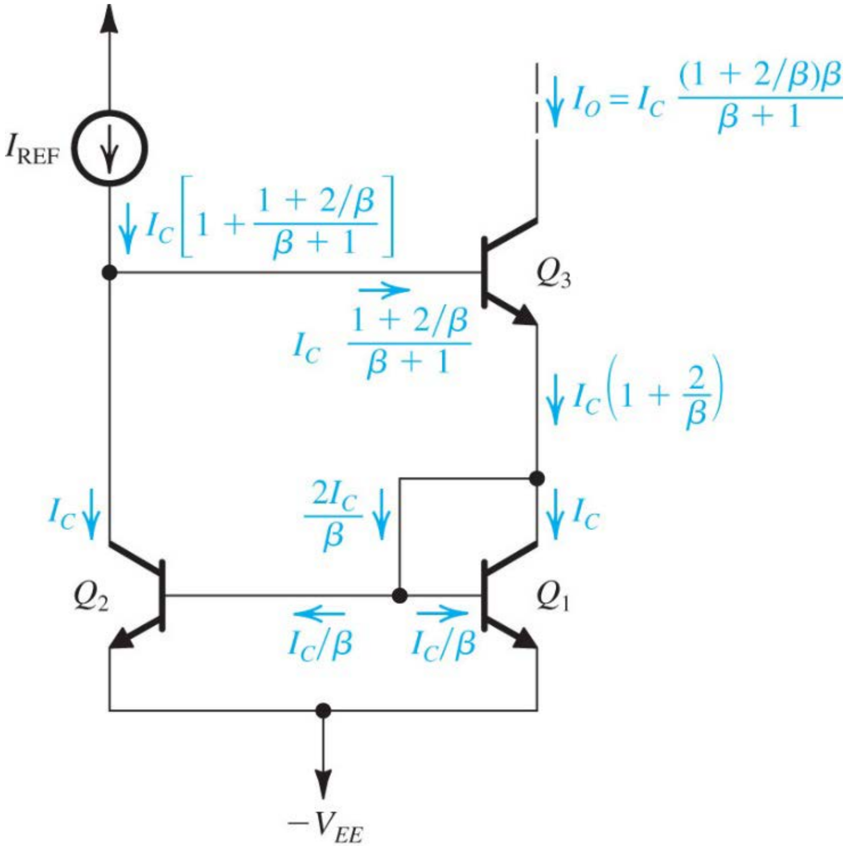


Figure 8.40 The Wilson bipolar current mirror: circuit showing analysis to determine the current transfer ratio

Reduces inaccuracy of I_O/I_{REF} and also improves (increases) R_O

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

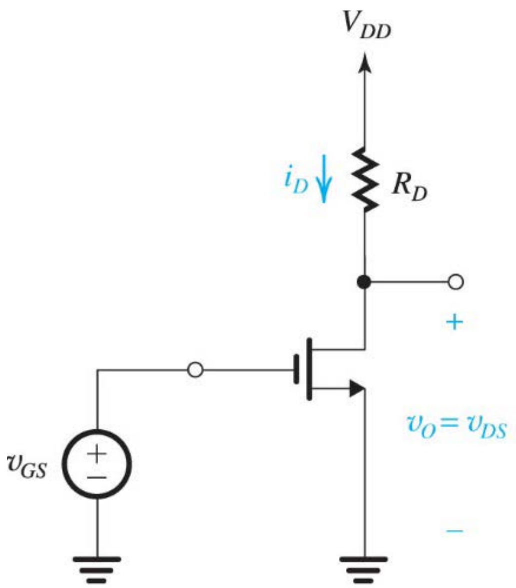
$$R_o = \beta_3 \, r_{o3}/2$$

(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).



$\Rightarrow$

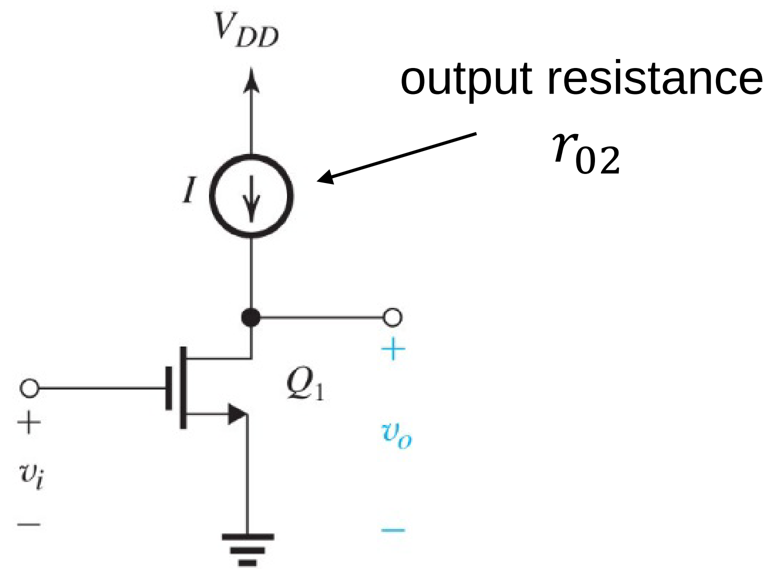


Figure 8.13

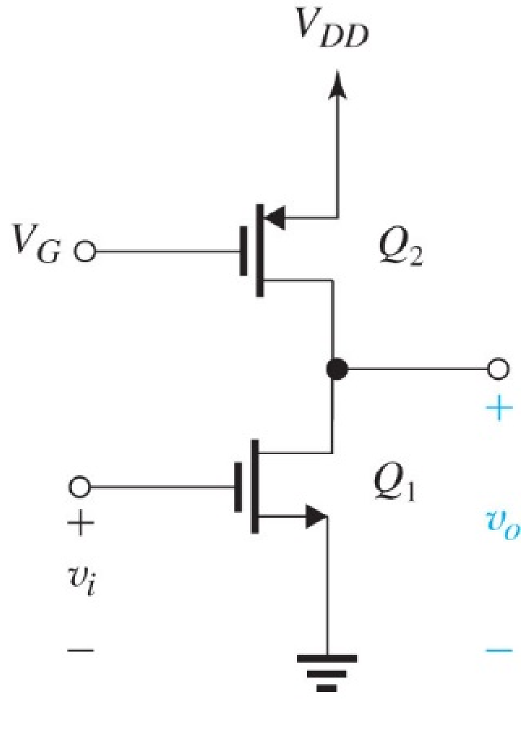
$$A_v = -g_m(R_D \parallel r_o)$$

Wish to increase R_D

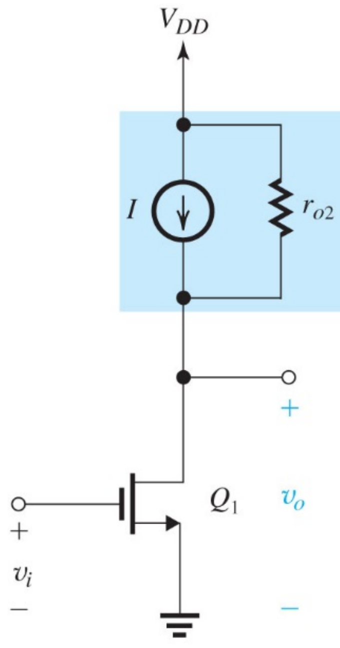
Use transistor(s) to create current source I.

Then $A_v = -g_{m1}(r_{o2} \parallel r_{o1})$.

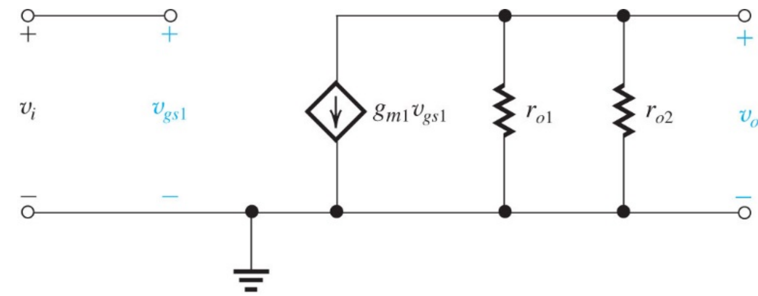
CS amplifier with PMOS active load



(a)



(b)



(c)

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

(dc value of the output voltage is not obvious)

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.

CS amplifier with current mirror as active load

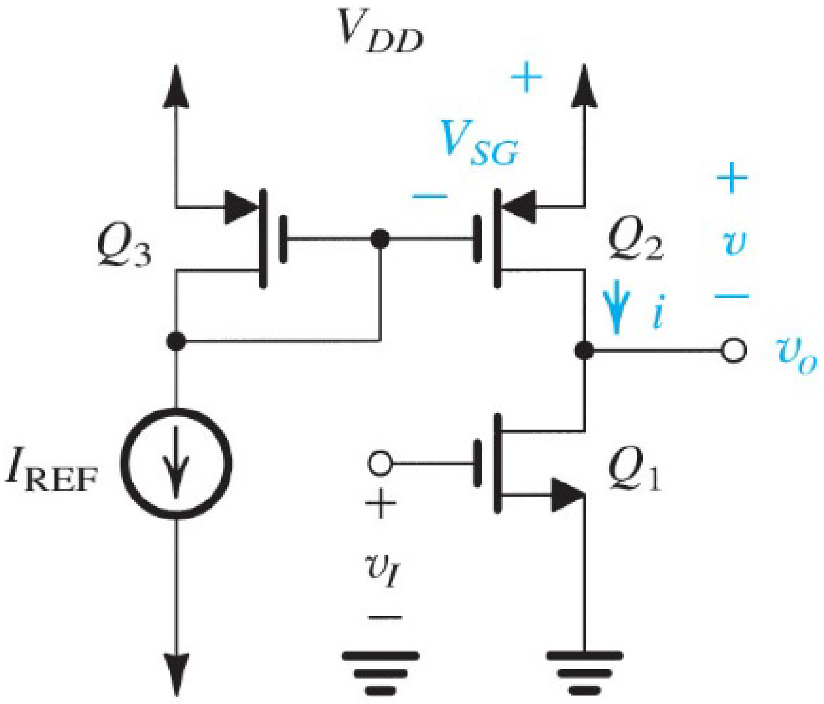


Figure 8.16

$$A_v = -g_{m1} R_o$$

$$R_o = r_{o1} \parallel r_{o2}$$

$$R_i = \infty$$

$$A_v = -g_{m1}\,(r_{o1} \parallel r_{02})$$

(dc value of the output voltage is not obvious)

CG amplifier with current mirror as active load

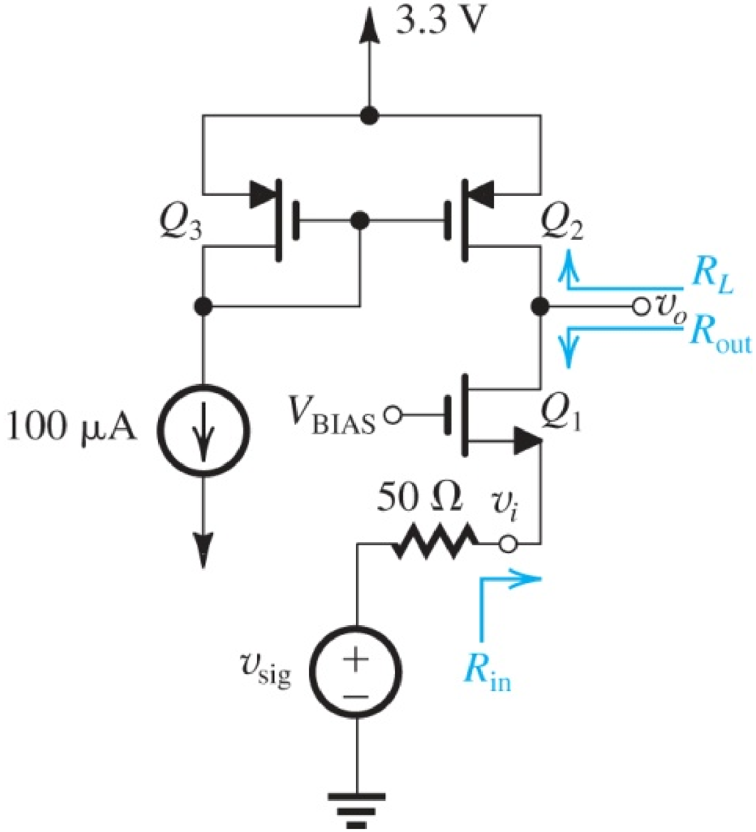


Figure P8.55

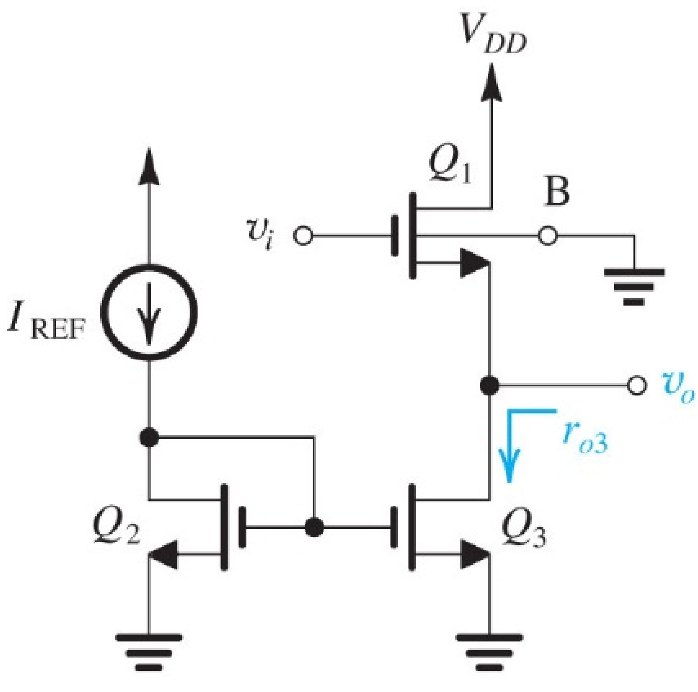
$$A_v = \left(g_{m1} + \frac{1}{r_{o1}}\right)(r_{o1} \parallel r_{o2})$$

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

$$R_o = r_{o1} \parallel r_{o2}$$

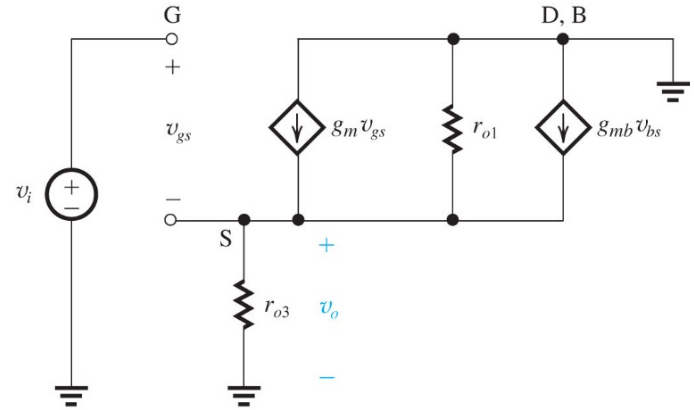
$$R_i \approx \frac{1}{g_{m1}}\left(1 + \frac{r_{o2}}{r_{o1}}\right)$$

Source follower with active load



(a)

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.



(b)

$$R_i \approx \infty$$

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

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

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

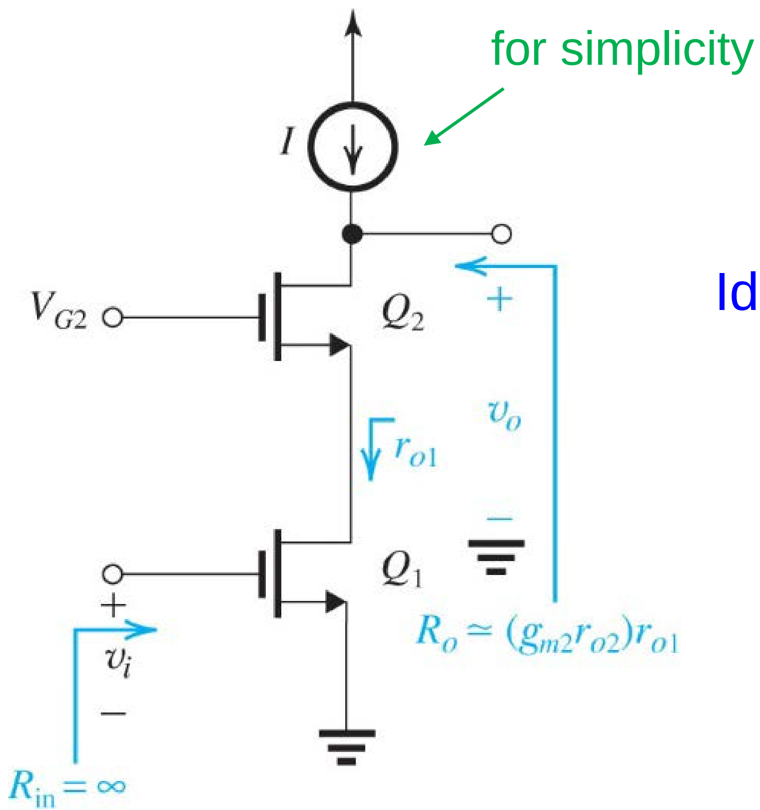


Figure 8.30 (**a**) A MOS cascode amplifier with an ideal current-source load

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 $1/g_{m2}$

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

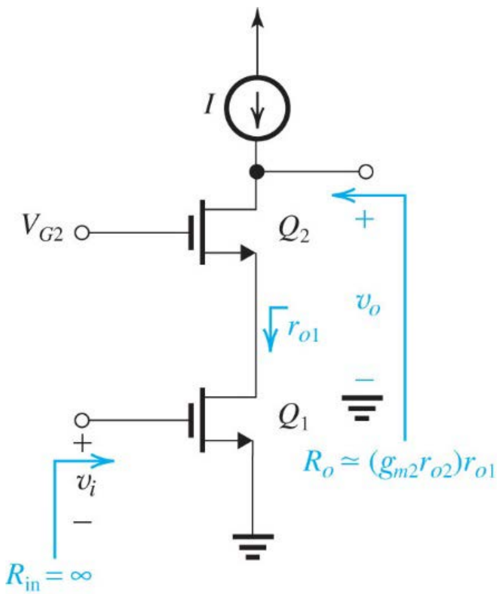
$$R_o \approx g_{m2}\, r_{o1}\, r_{o2}$$

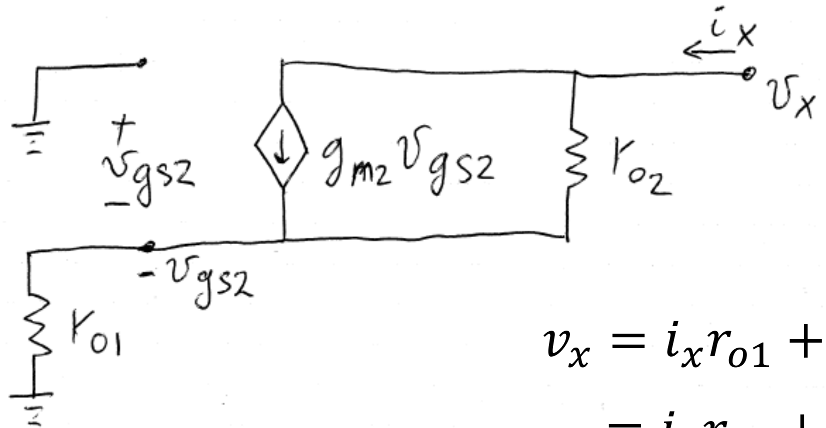
More accurately,

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

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

Derivation of output resistance R_o for cascode





$$v_{gs2} = -i_x r_{o1}$$

$$\begin{aligned} v_x &= i_x r_{o1} + r_{o2}\left(i_x - g_{m2} v_{gs2}\right) \\ &= i_x r_{o1} + r_{o2} i_x \left(1 + g_{m2} r_{o1}\right) \\ &= i_x\left[r_{o1} + r_{o2}\left(1 + g_{m2} r_{o1}\right)\right] \end{aligned}$$

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

usual output resistance r_{o2} is increased by $g_{m2} r_{01}$

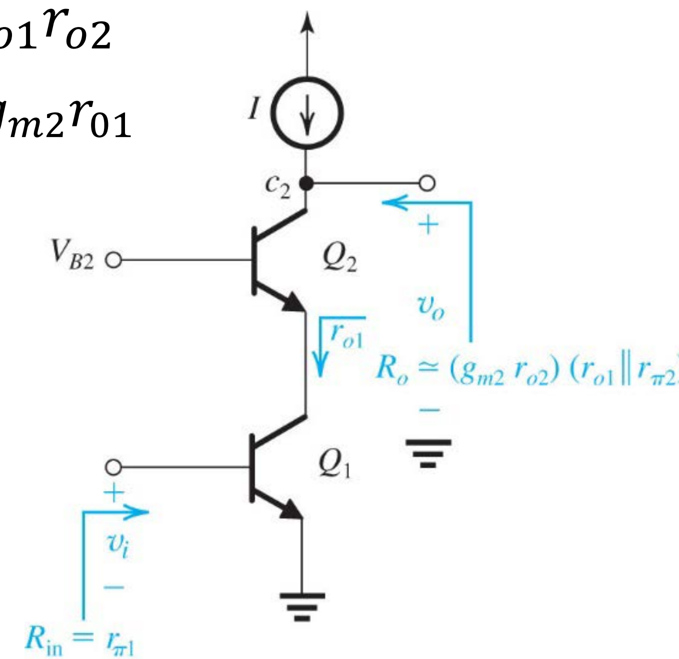
Similarly, for a BJT cascode (will need later)

$$R_o = \underline{(r_{o1} \| r_{\pi 2}) + r_{o2}} + g_{m2}(\underline{r_{o1}} \| r_{\pi 2})\, r_{o2}$$

neglect

$$\approx g_{m2} r_{\pi 2} r_{02} = \beta_2 r_{02}$$

r_{02} is increased by factor β_2



MOS cascode with ideal current source

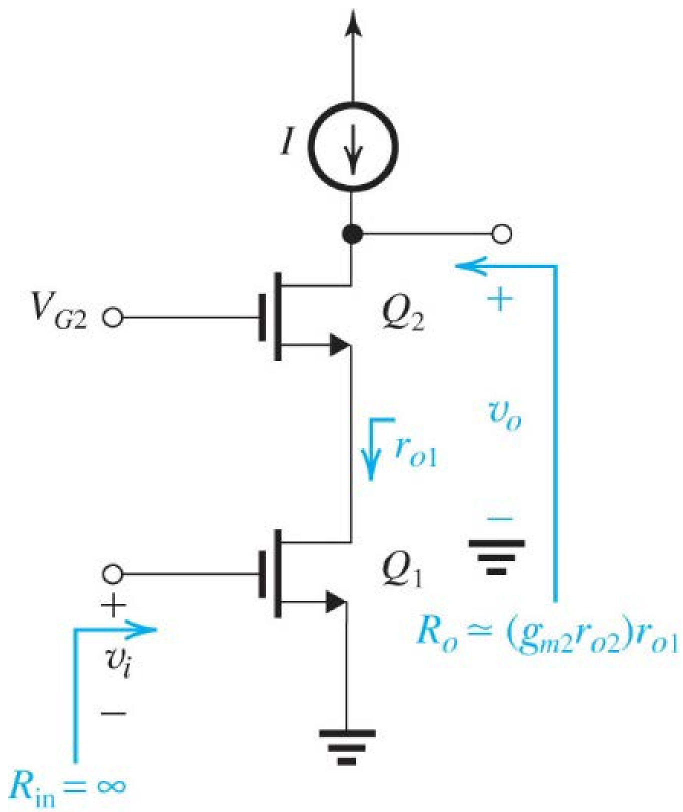


Figure 8.30 (**a**) A MOS cascode amplifier with an ideal current-source load

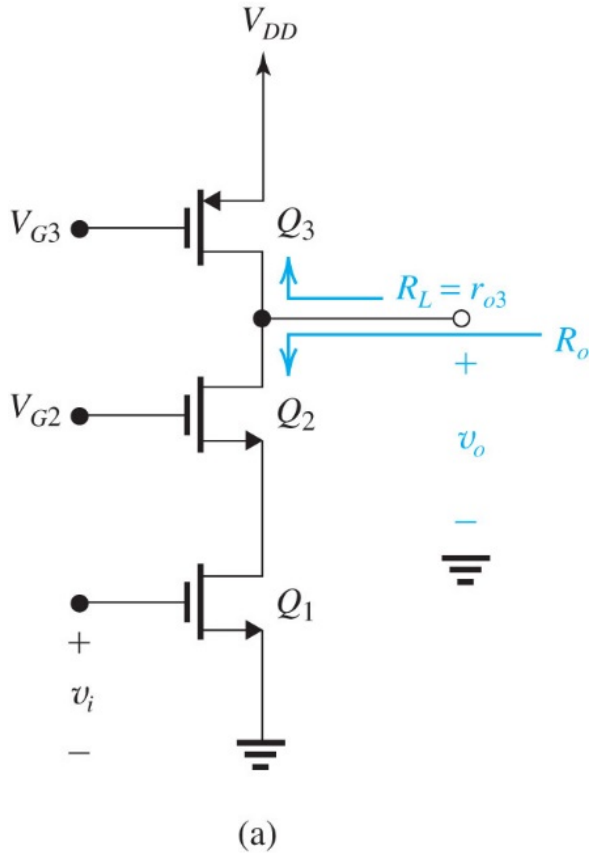
$$R_o \approx g_{m2}\, r_{o1}\, r_{o2}$$

$$A_v \approx -g_{m1} R_o \approx -g_{m1} r_{o1}\, g_{m2} r_{o2}$$

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.

MOS cascode with simple PMOS current source



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 8.31 (a) A MOS cascode amplifier loaded in a simple PMOS current source Q_3.

MOS cascode with cascode current source

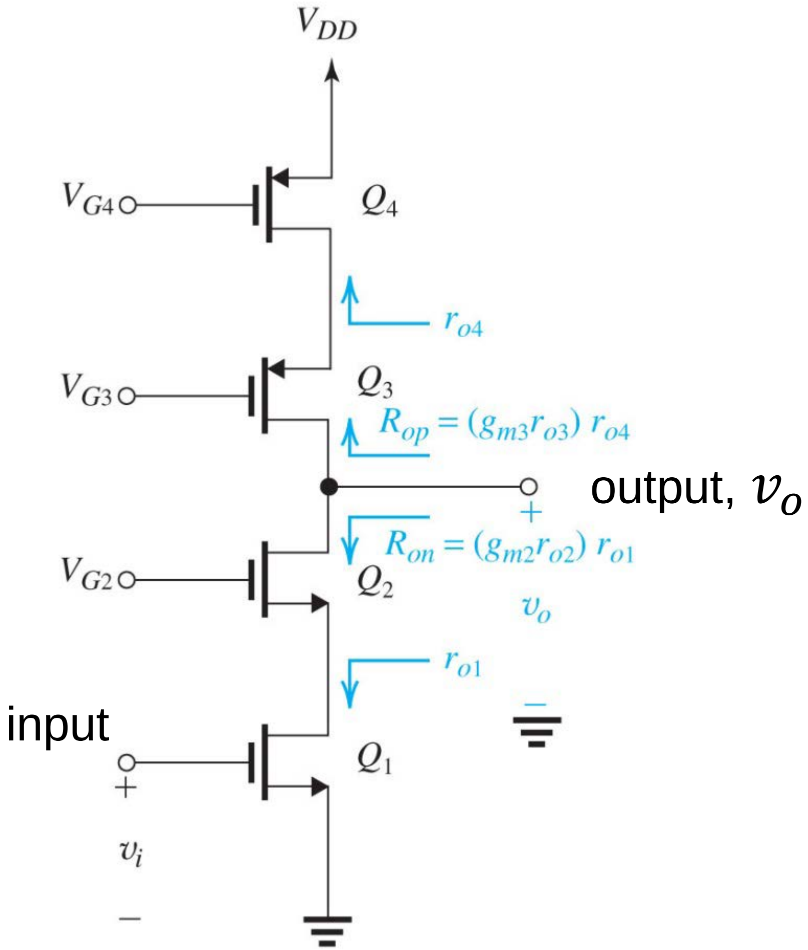


Figure 8.33 A cascode amplifier with
a cascode current-source load.

$$R_{op} = g_{m3}r_{o3}r_{o4}$$

$$R_{on} = g_{m2}r_{o1}r_{o2}$$

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

If all g_m are equal and all r_o are equal,

then $$A_v = -\frac{1}{2}(g_m r_o)^2$$

MOS double cascode

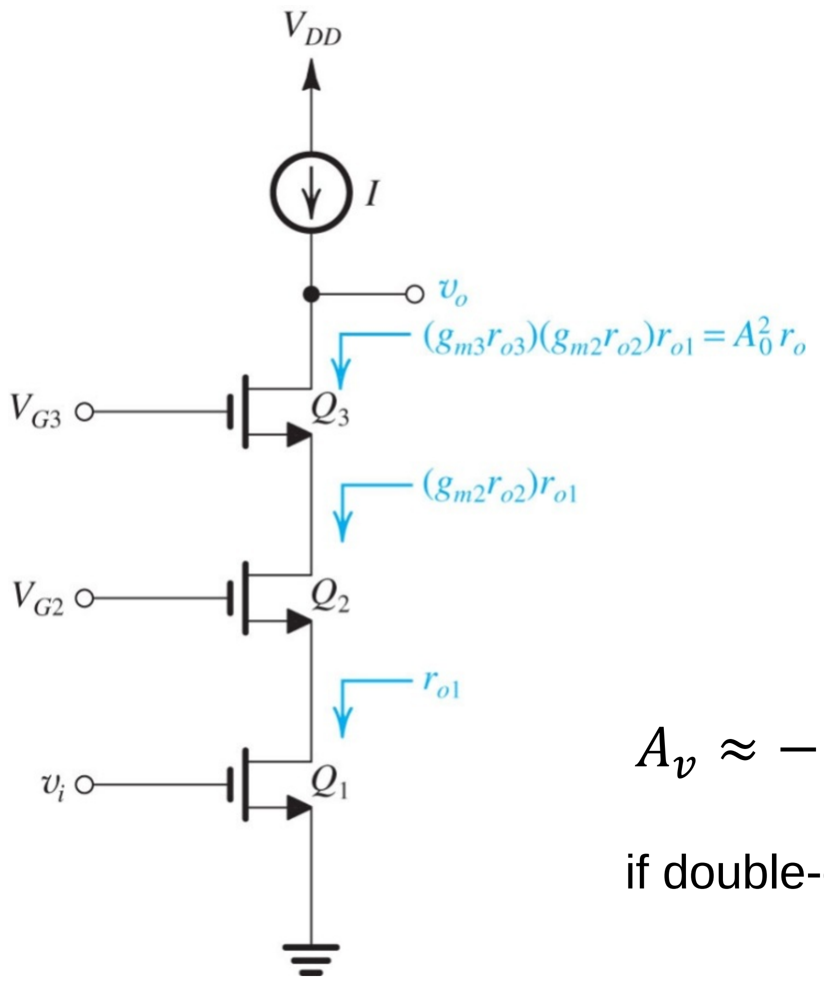


Figure 8.35 Double cascoding.

Each time increase output resistance

$$r_{o2} \rightarrow (g_{m2}\, r_{02})\, r_{01}$$

$$\rightarrow (g_{m3}\, r_{o3})(g_{m2}\, r_{o2})\, r_{01}$$

$A_v \approx -(g_m r_o)^3$ if ideal current source

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

MOS folded cascode

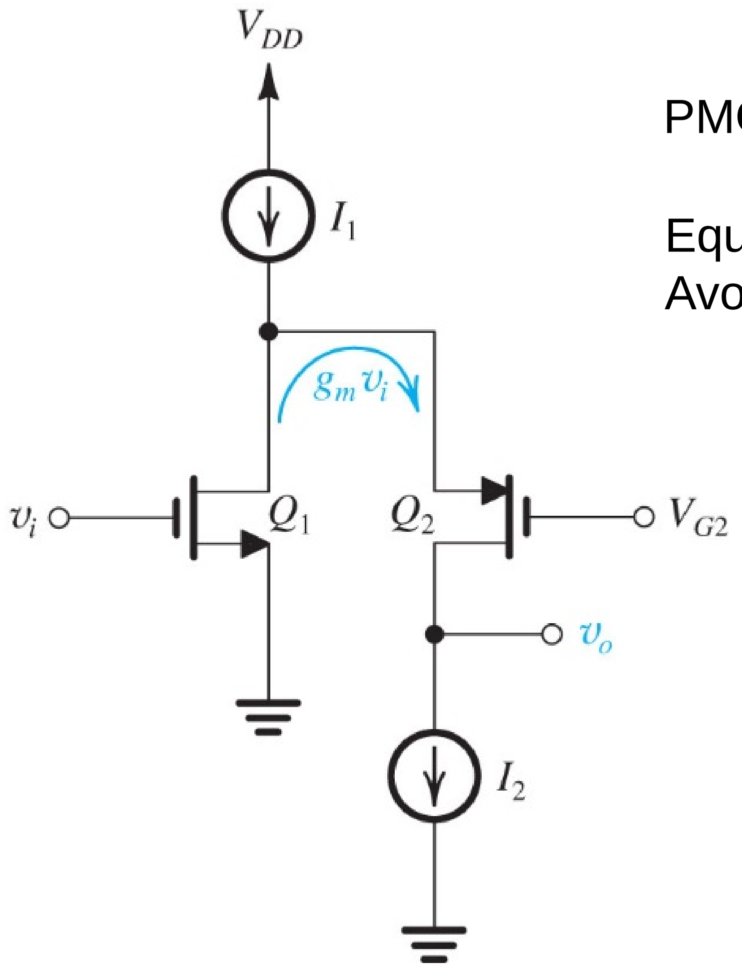


Figure 8.36 The folded cascode.

PMOS common gate load.

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

BJT cascode

Common-emitter loaded with common-base

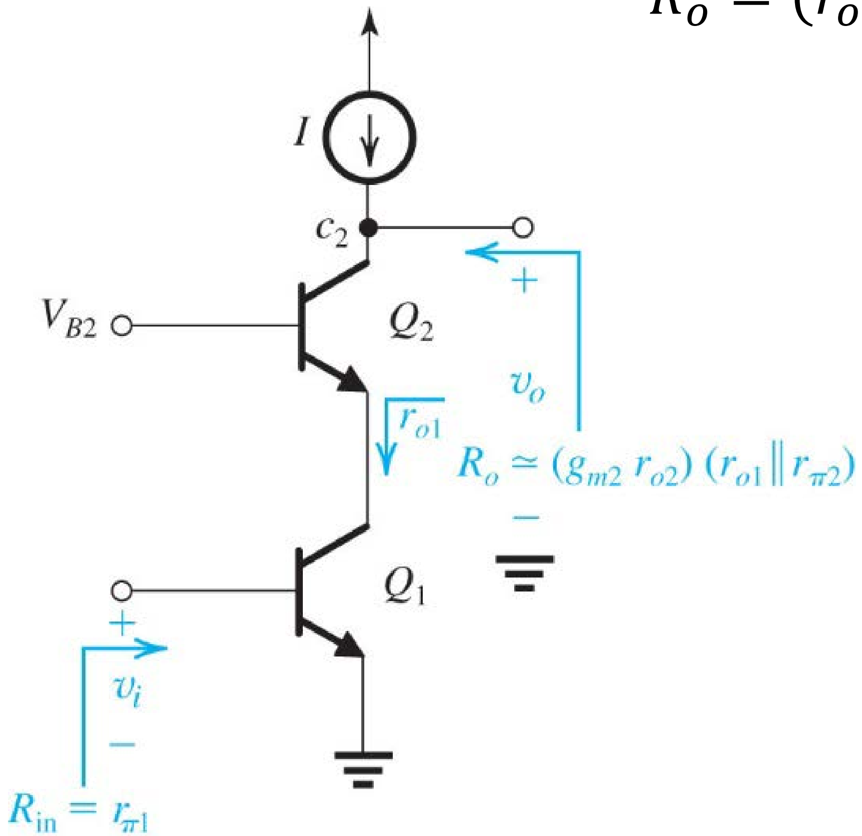


Figure 8.37 (**a**) A BJT cascode amplifier with an ideal current-source load;

$$R_o = (r_{o1} \| r_{\pi 2}) + r_{o2} + g_{m2}\, r_{o2}\, (r_{o1} \| r_{\pi 2})$$

$$\approx g_{m2} r_{\pi 2} r_{02} = \beta_2 r_{02}$$

(as derived earlier)

$$A_v \approx -g_{m1} R_o \approx -g_{m1} \beta_2 r_{02}$$

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

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

Needs “good” current source (otherwise significantly less A_v)

BJT cascode with cascode current source

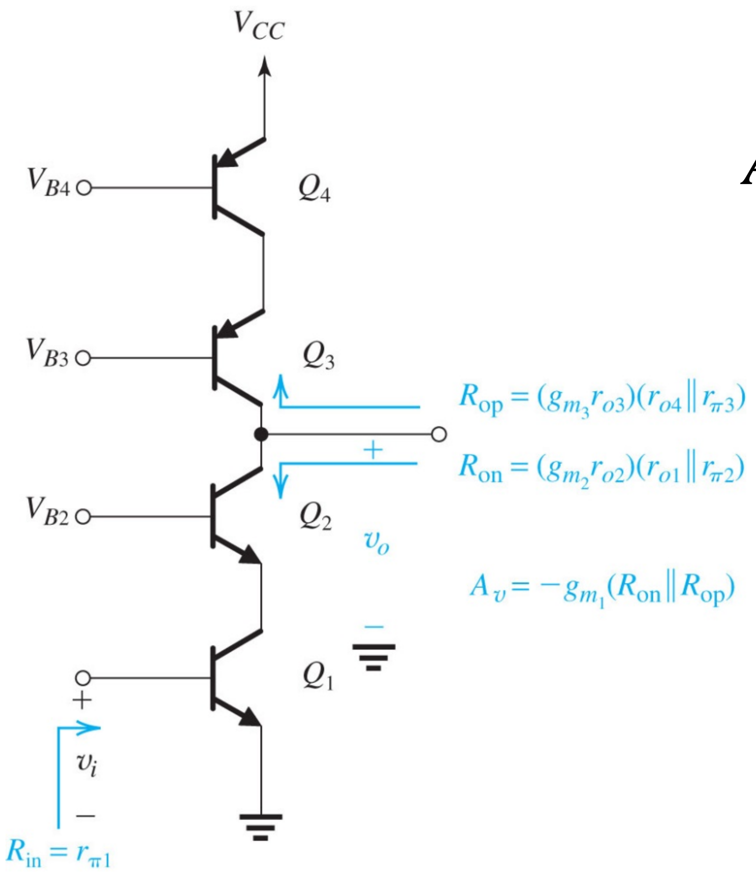


Figure 8.38 A BJT cascode amplifier with a cascode current source.

$$A_v = -g_{m1}\,(\beta_2 r_{o2} \parallel \beta_3 r_{o3})$$

BiCMOS cascode

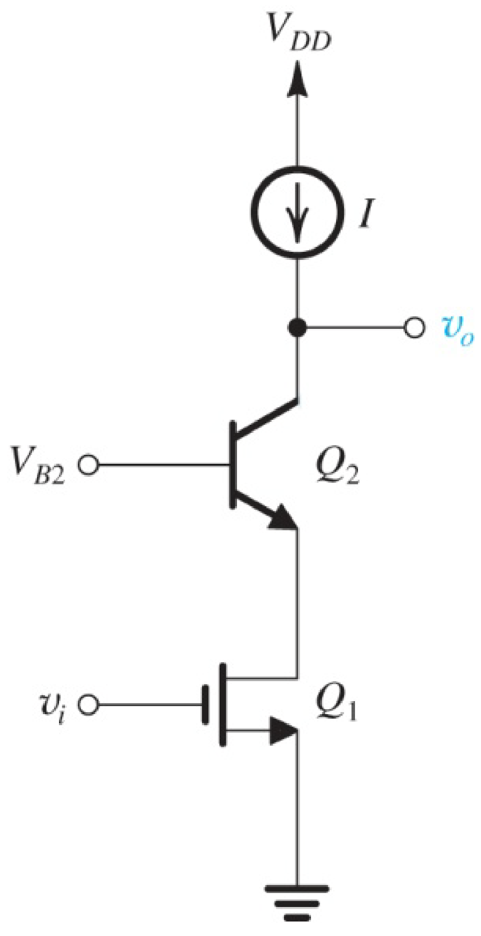


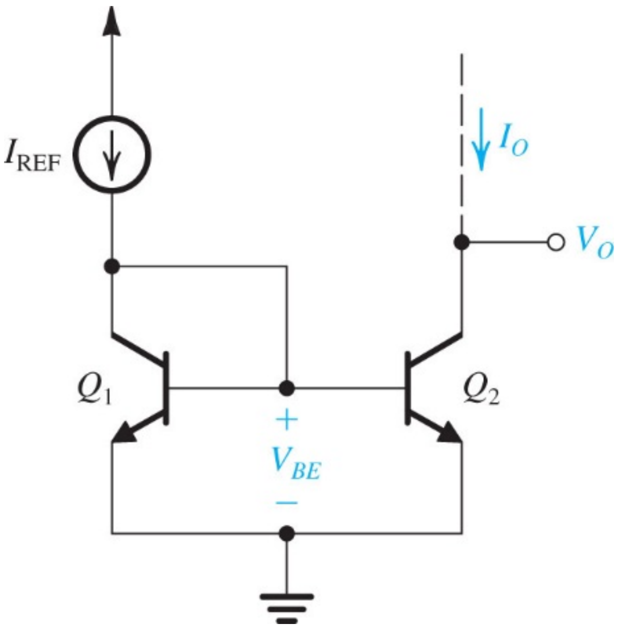
Figure P8.81

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

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

Next subject: improved current mirrors

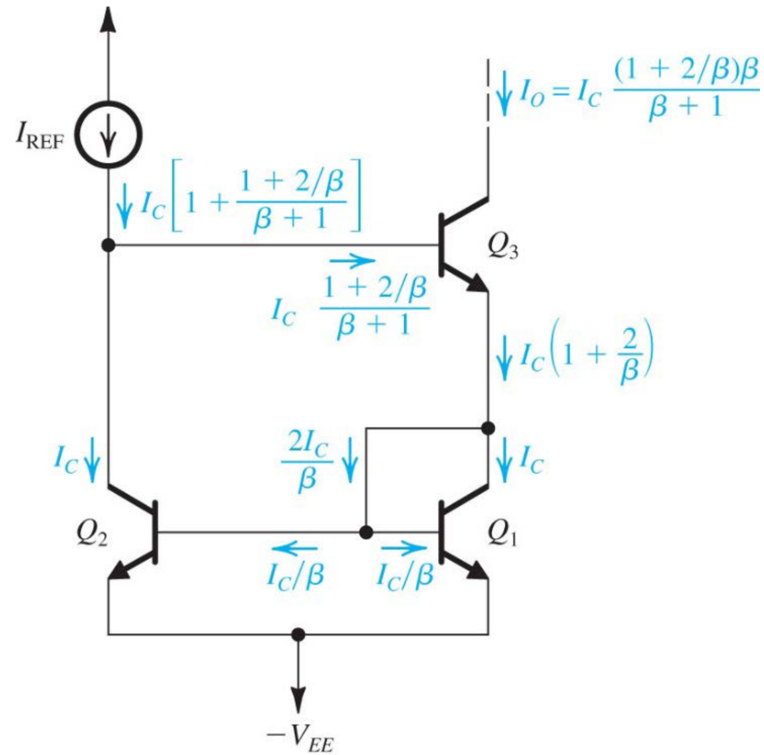
Basic BJT current mirror



$$I_O \approx (1 - 2/\beta) I_{REF}$$

$R_o = r_{o2}$ (not large)

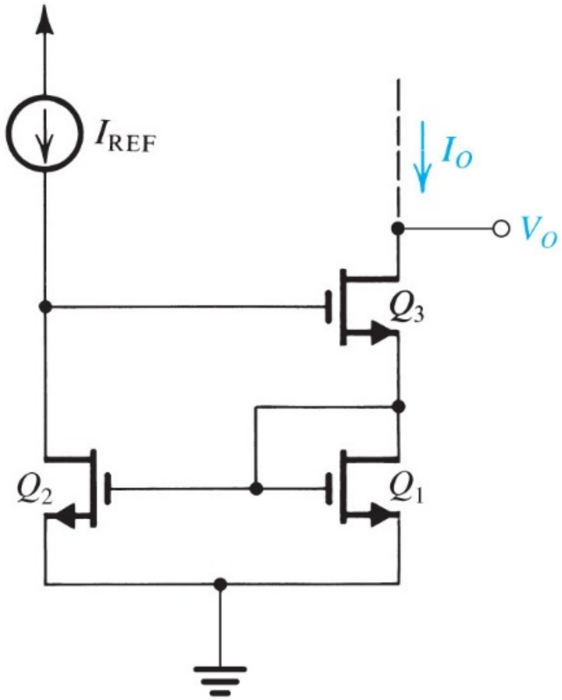
Wilson BJT current mirror



$I_O \approx (1 - 2/\beta^2)\, I_{REF}$ (better)

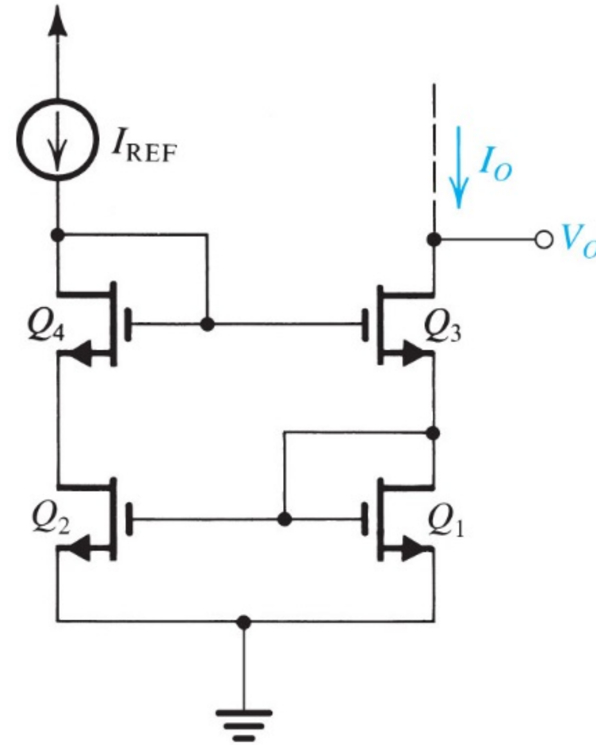
$R_o = \beta_3\, r_{o3}/2$ (larger)

Wilson MOS mirror



(a)

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



(c)

$$R_o = g_{m3}\, r_{o3}\, r_{o2}$$

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)

Cascode MOS mirror

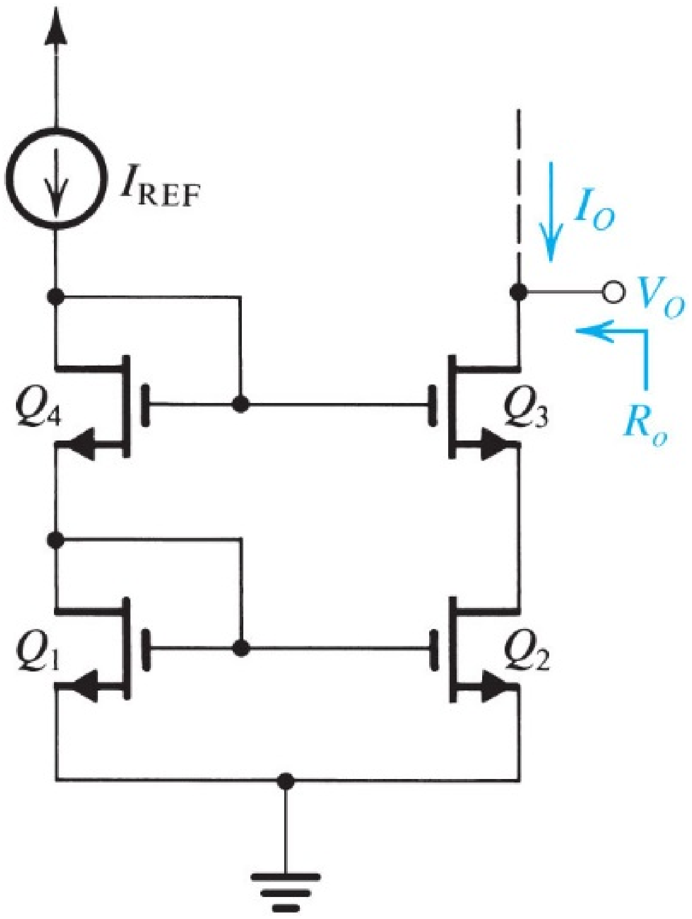


Figure 8.39 A cascode MOS current mirror.

$$R_o = g_{m3}\, r_{o3}\, r_{o2}$$

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

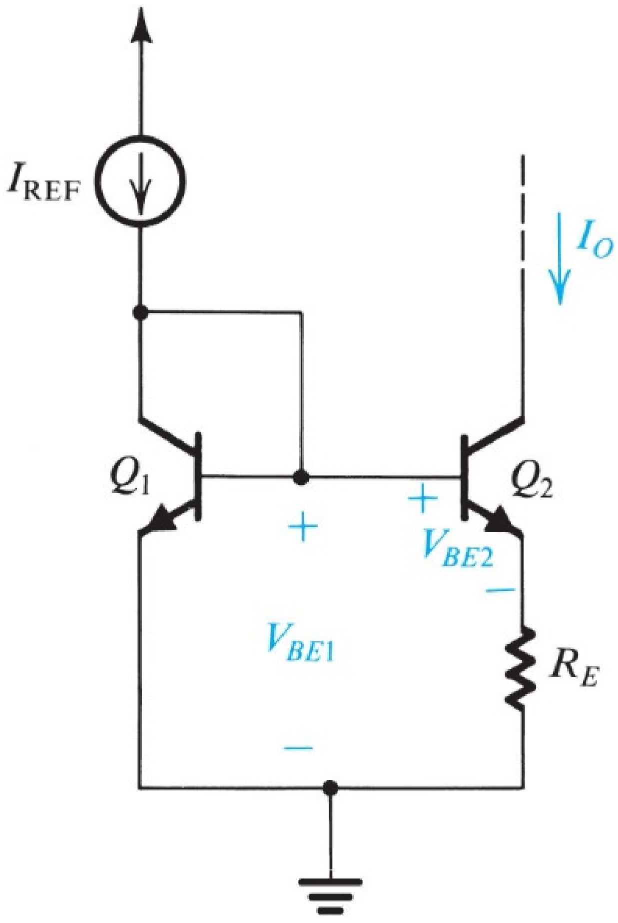


Figure 8.42 The Widlar current source.

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

Assume matched transistors

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

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

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

(increased output resistance)

Next subject: Some useful transistor pairings

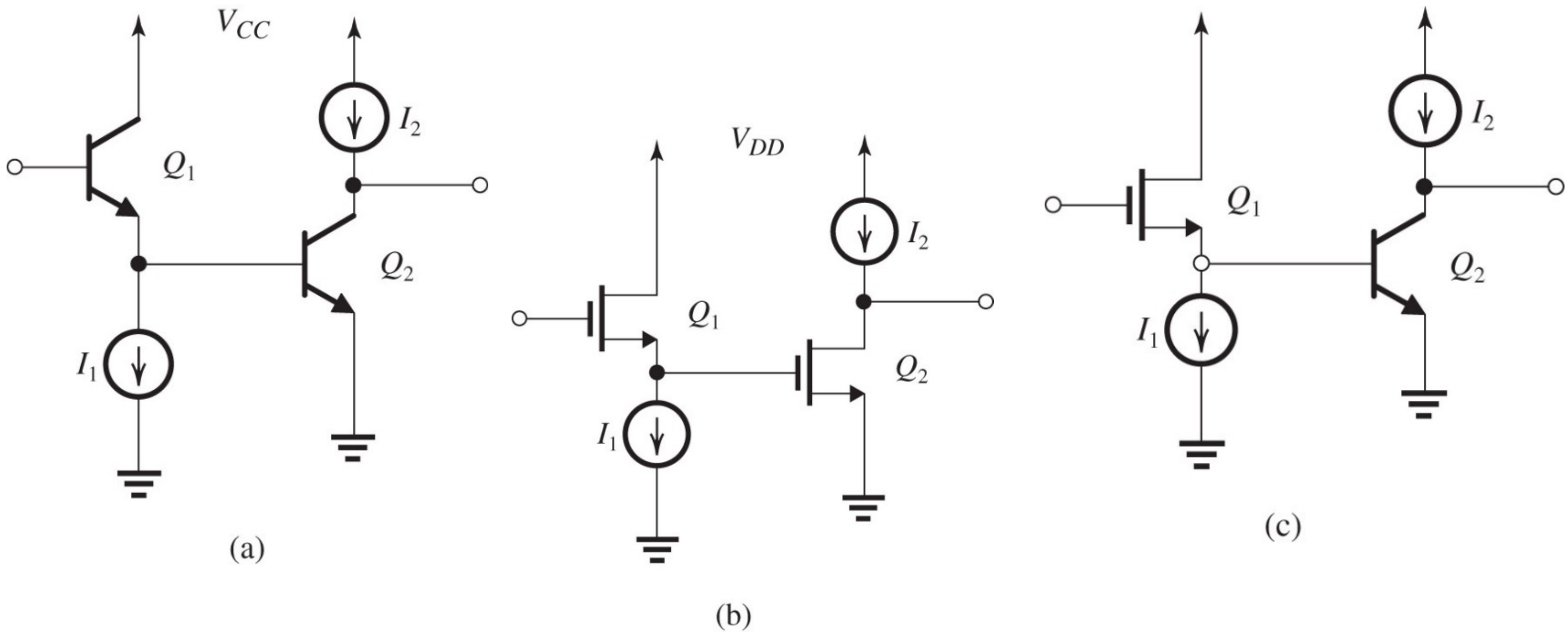


Figure 8.44 (**a**) CC–CE amplifier; (**b**) CD–CS amplifier; (**c**) CD–CE amplifier.

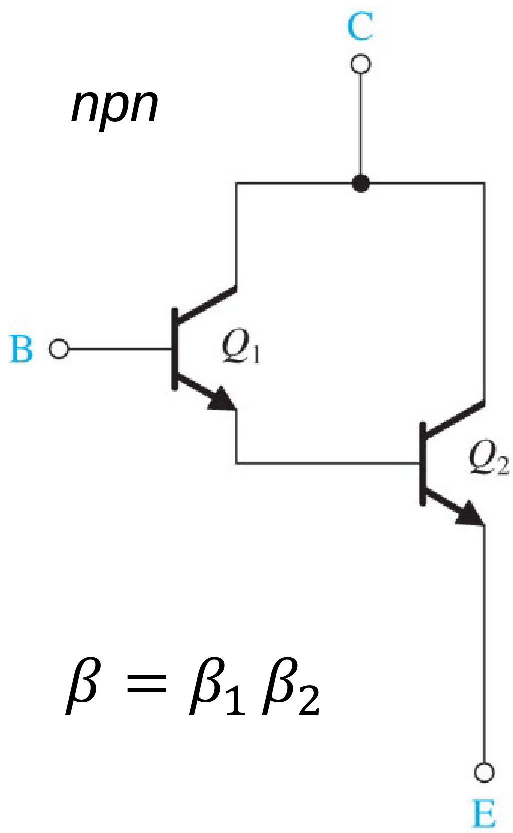
(a) CC-CE: increases R_i (due to emitter follower), makes faster (not obvious)

(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)

Darlington configuration

npn

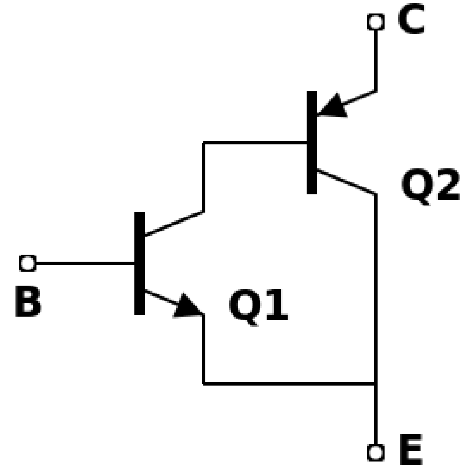


$$\beta = \beta_1 \beta_2$$

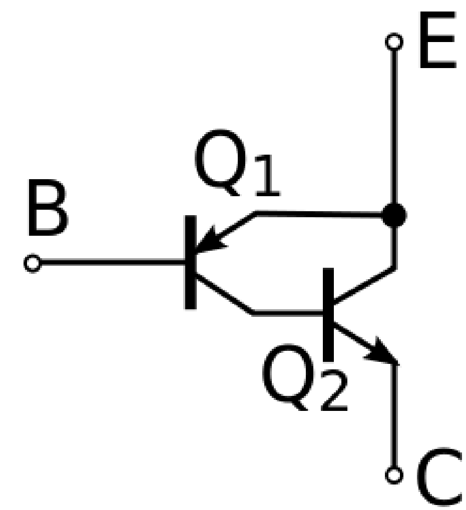
Figure 8.47 (**a**) The Darlington configuration.

Sziklai pair (compound, complementary Darlington)

npn



pnp



$$\beta = \beta_1 \beta_2$$

CC-CB (CD-CG) configuration

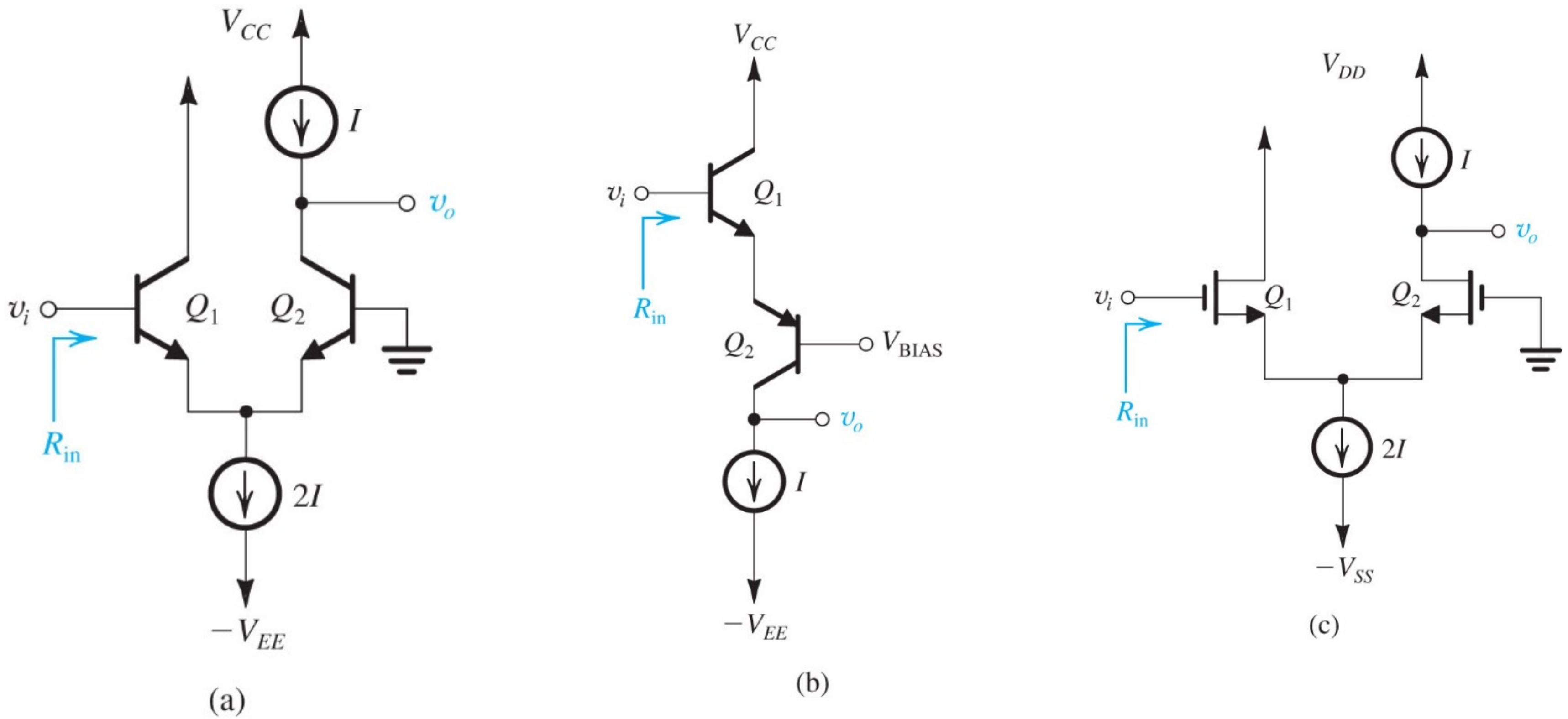


Figure 8.48 (**a**) A CC–CB amplifier. (**b**) Another version of the CC–CB circuit with Q_2 implemented using a *pnp* transistor. (**c**) The MOSFET version of the circuit in (**a**).

(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)