Integrated Circuit Biasing and Active Loads

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Two Transistor Current Source

- The two-transistor current source is also called a current mirror, which consists of two matched (or identical) transistors, Q1 and Q2, operating at the same temperature.

$$I_{REF} = I_{C1} + I_{B1} + I_{B2} = I_{C1} + 2I_{B1}$$
$$= I_{C2} + 2I_{B2} = I_{C2}\left(1 + \frac{2}{\beta}\right)$$

$$I_O = I_{C2} = I_{REF} \Big/ \left(1 + \frac{2}{\beta}\right)$$

$$I_{REF} = \frac{V^+ - V_{BE} - V^-}{R_1}$$

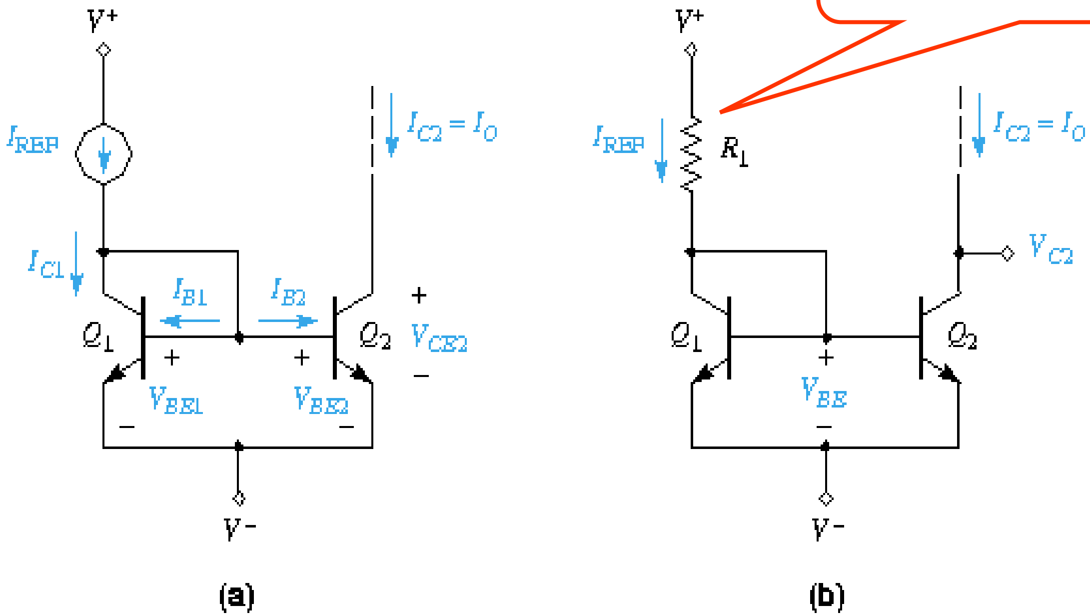


Figure 10.2 (a) Basic two-transistor current source; (b) two-transistor current source with reference resistor R_1

Design Example 10.1 **Objective:** Design a two-transistor current source to provide a specific output current.

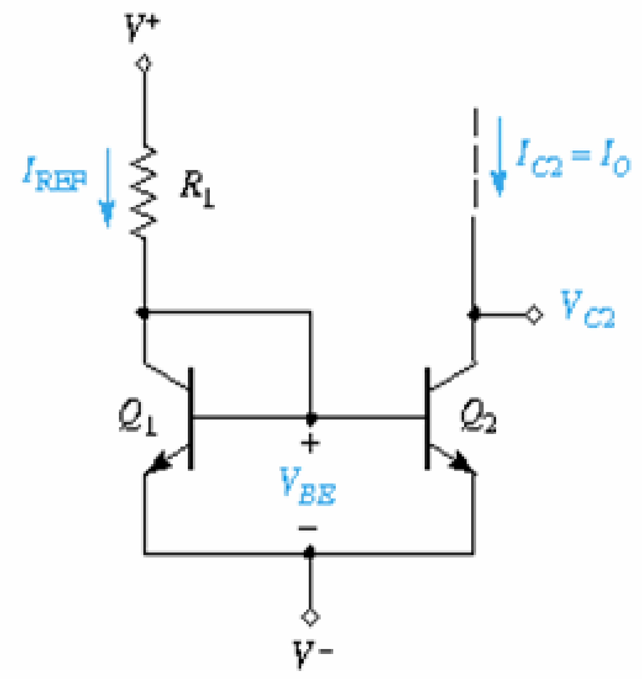
Consider the circuit shown in Figure 10.2(b). The transistor parameters are: $V_{BE}(\text{on}) = 0.6\,\text{V}$, $\beta = 100$, and $V_A = \infty$. The output current is to be $200\,\mu\text{A}$ with $V^+ = 5\,\text{V}$ and $V^- = 0$.

Solution: The reference current can be written as

$$I_{\text{REF}} = I_O\left(1 + \frac{2}{\beta}\right) = (200)\left(1 + \frac{2}{100}\right) = 204\,\mu\text{A}$$

From Equation (10.1), resistor R_1 is found to be

$$R_1 = \frac{V^+ - V_{BE}}{I_{\text{REF}}} = \frac{5 - 0.6}{0.204} = 21.6\,\text{k}\Omega$$



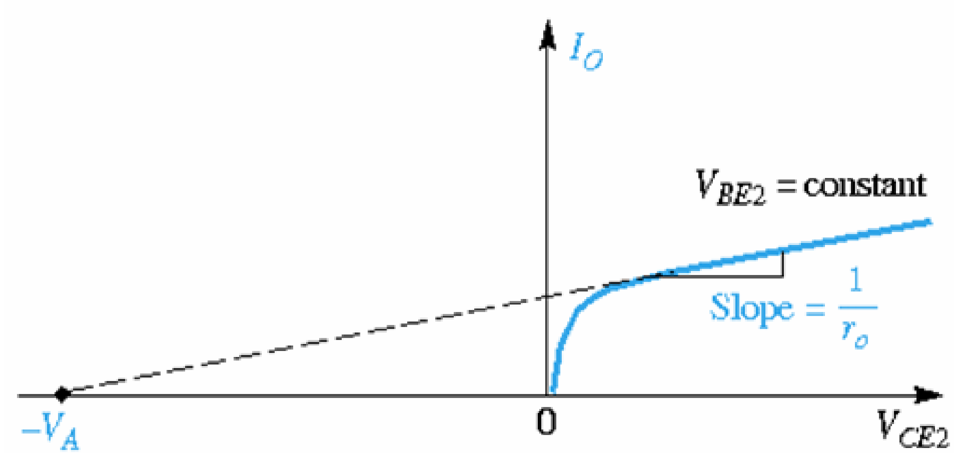
Output Resistance with Early Effect

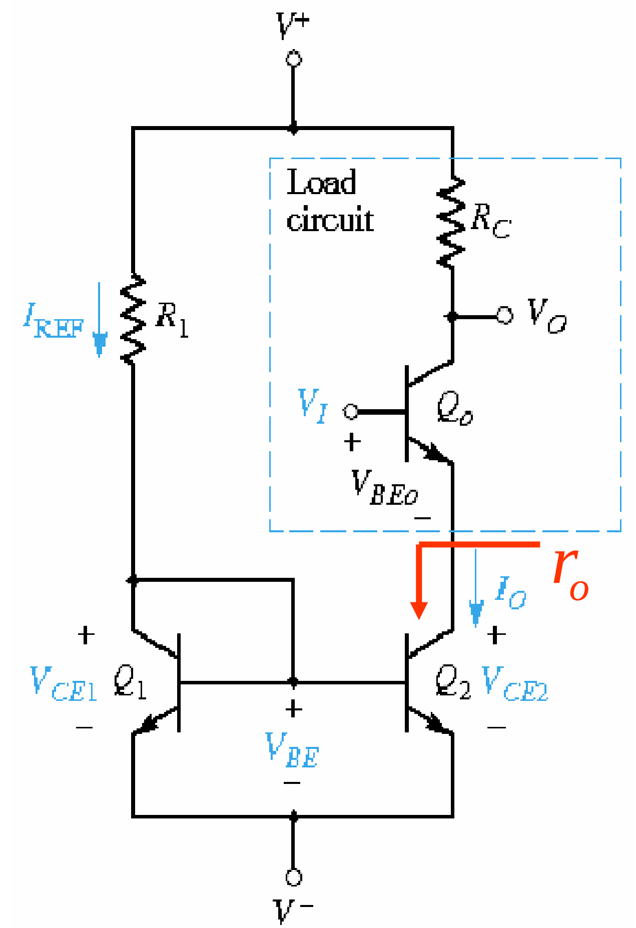
- Taking the Early effect into account

$$\frac{I_O}{I_{REF}} = \frac{1}{1+2/\beta} \times \frac{1+V_{CE2}/V_A}{1+V_{CE1}/V_A}$$

$$V_{CE2} = V_I - V_{BEo} - V^-$$

For finite Early voltage, a change in the V_I dc bias condition in the load circuit affects the collector-emitter voltage of Q_2.





$$\frac{dI_O}{dV_{CE2}} = \frac{I_{REF}}{1+2/\beta} \times \frac{1}{V_A} \times \frac{1}{1+V_{BE}/V_A}$$

$$\approx \frac{I_O}{V_A} = \frac{1}{r_O} \quad (V_{BE} << V_A)$$

Example 10.2 **Objective:** Determine the change in load current produced by a change in collector–emitter voltage in a two-transistor current source.

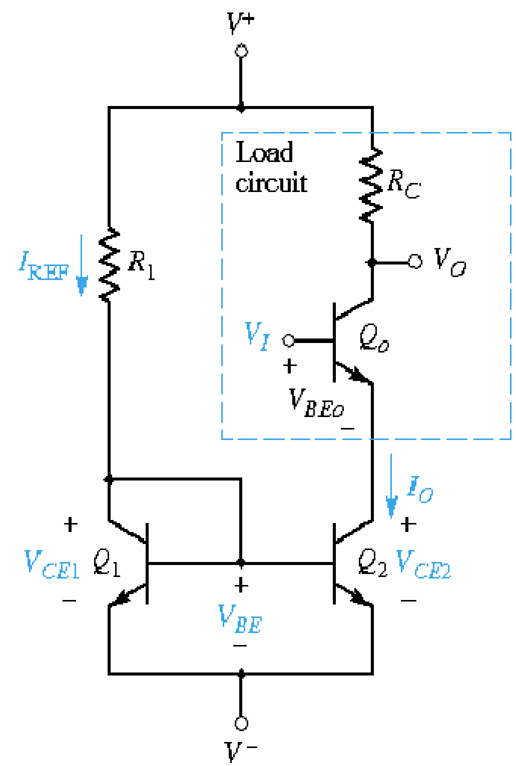
Consider the circuit shown in Figure 10.3. The circuit parameters are: $V^+ = 5\,\text{V}$, $V^- = -5\,\text{V}$, and $R_1 = 9.3\,\text{k}\Omega$. Assume the transistor parameters are: $\beta = 50$, $V_{BE}(\text{on}) = 0.7\,\text{V}$, and $V_A = 80\,\text{V}$. Determine the change in I_O as V_{CE2} changes from 0.7 V to 5 V.

Solution: The reference current is

$$I_{\text{REF}} = \frac{V^+ - V_{BE}(\text{on}) - V^-}{R_1} = \frac{5 - 0.7 - (-5)}{9.3} = 1.0\,\text{mA}$$

For $V_{CE2} = 0.7\,\text{V}$, transistors Q_1 and Q_2 are identically biased. From Equation (10.5), we then have

$$I_O = \frac{I_{\text{REF}}}{1 + \frac{2}{\beta}} = \frac{1.0}{1 + \frac{2}{50}} = 0.962\,\text{mA}$$



From Equation (10.8), the small-signal output resistance is

$$r_o = \frac{V_A}{I_O} = \frac{80}{0.962} = 83.2\,\text{k}\Omega$$

The change in load current is determined from

$$\frac{dI_O}{dV_{CE2}} = \frac{1}{r_o}$$

or

$$dI_O = \frac{1}{r_o}dV_{CE2} = \frac{1}{83.2}(5 - 0.7) = 0.052\,\text{mA}$$

The percent change in output current is therefore

$$\frac{dI_O}{I_O} = \frac{0.052}{0.962} = 0.054 \Rightarrow 5.4\%$$

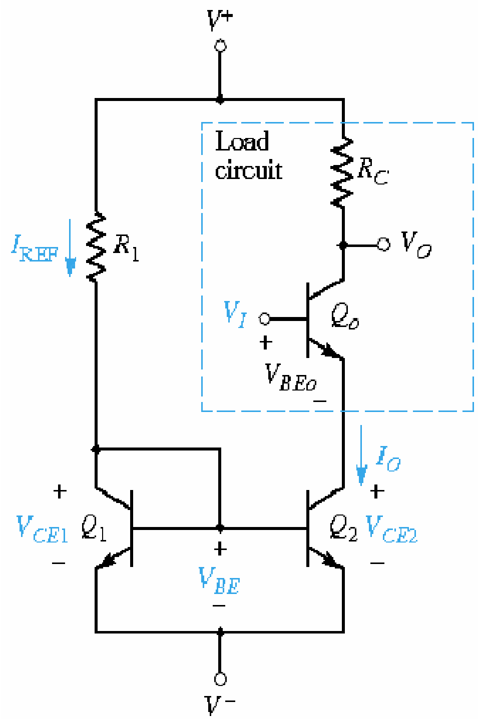
Mismatched Transistors

- In practice, transistor Q1 and Q2 may not be exactly identical.

$$I_{REF} \approx I_{C1} = I_{S1} e^{V_{BE}/V_T}$$

$$I_O = I_{C2} = I_{S2} e^{V_{BE}/V_T}$$

If Q1 and Q2 are not identical, then $I_{S1} \neq I_{S2}$.



- Relationship between the bias and reference currents

$$I_O = I_{REF} \left(\frac{I_{S2}}{I_{S1}} \right)$$

- I_S is a strong function of temperature. Thus Q1 and Q2 must be close to one another on the semiconductor for the similar operation situation (including temperature).

- By using different sizes of transistors, we can design the circuit such that $I_O \neq I_{REF}$.

Basic Three Transistor Current Source

Assume that all transistor are identical.

$$I_{REF} = I_{C1} + I_{B3}$$

$$I_{B3} = \frac{I_{E3}}{1+\beta_3} = \frac{2I_{B2}}{1+\beta_3} = \frac{2I_O}{\beta_2(1+\beta_3)}$$

$$I_{C1} = I_{C2} = I_O$$

$$I_{REF} = I_{C2} + \frac{2I_{C2}}{\beta_2(1+\beta_3)} = I_{C2}\left[1 + \frac{2}{\beta_2(1+\beta_3)}\right]$$

$$I_O = I_{REF} \Big/ \left[1 + \frac{2}{\beta_2(1+\beta_3)}\right]$$

$$I_{REF} = \frac{V^+ - V_{BE3} - V_{BE} - V^-}{R_1}$$

$$\approx \frac{V^+ - 2V_{BE} - V^-}{R_1}$$

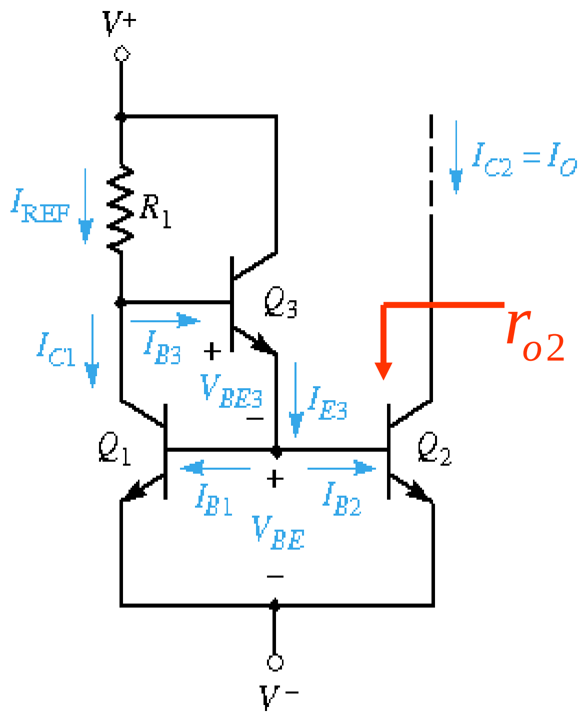


Figure 10.5 Basic three-transistor current source

- The approximation of $I_O \cong I_{REF}$ is better.
- The change in load current with a change in β is much smaller.

Cascode Current Source

- Current-source circuits can be designed such that the output resistance is much greater than that of the two-transistor circuit.
- For a constant reference current, the base voltages of Q2 and Q4 are constant, which implies these terminals are at signal ground.

$$V_{be4} = -I_x(r_{o2} // r_{\pi 4})$$

$$I_x = g_{m4}V_{be4} + \left(\frac{V_x - I_x(r_{o2} // r_{\pi 4})}{r_{o4}}\right)$$

$$= -g_{m4}I_x(r_{o2} // r_{\pi 4}) + \left(\frac{V_x - I_x(r_{o2} // r_{\pi 4})}{r_{o4}}\right)$$

$$R_o = \frac{V_x}{I_x} = r_{o4}(1+\beta) + r_{\pi 4}$$

$$\approx \beta r_{o4}$$

Assume that all transistor are identical.

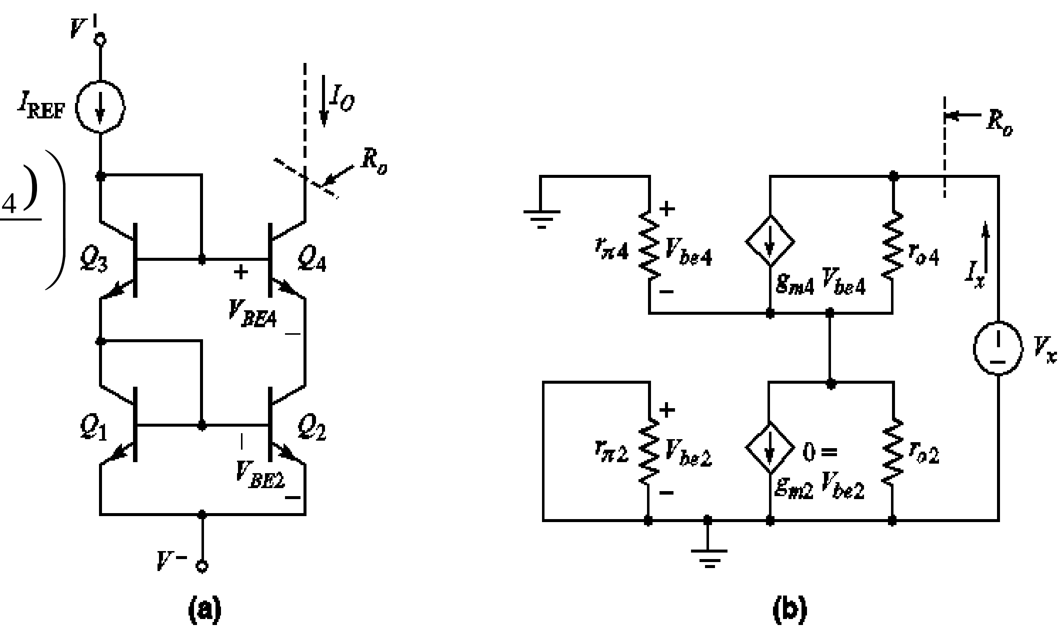
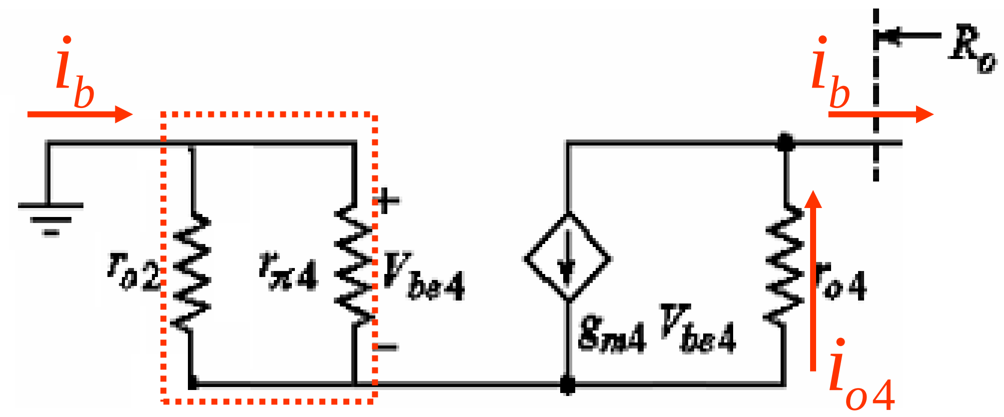


Figure 10.7 (a) Bipolar cascode current mirror; (b) small-signal equivalent circuit

Resistance Analysis of the Cascode Current Source



$$g_{m4}V_{be4} = g_{m4}(r_{o2} // r_{\pi 4})i_b$$

$$i_{o4} = [1 + g_{m4}(r_{o2} // r_{\pi 4})]i_b$$

$$\boxed{R_o = (r_{o2} // r_{\pi 4}) + r_{o4}[1 + g_{m4}(r_{o2} // r_{\pi 4})]}$$

$$\approx r_{\pi 4} + r_{o4}(1 + g_{m4}r_{\pi 4})$$

$$= r_{\pi 4} + r_{o4}(1 + \beta)$$

$$\approx \beta r_{o4}$$

Wilson Current Source

$$I_{REF} = I_{C1} + I_{B3} = I_{C2} + I_{B3}$$

$$I_{E3} = I_{C2} + 2I_{B2} = I_{C2}\left(1 + \frac{2}{\beta}\right)$$

$$I_{C2} = I_{E3} \Big/ \left(1 + \frac{2}{\beta}\right) = \frac{1}{1 + 2/\beta} \times \frac{1+\beta}{\beta} I_{C3}$$

$$= \frac{1+\beta}{2+\beta} I_{C3}$$

$$I_{REF} = \frac{1+\beta}{2+\beta} I_{C3} + \frac{I_{C3}}{\beta}$$

$$I_O = I_{REF} \times \frac{1}{1 + \dfrac{2}{\beta(2+\beta)}}$$

Assume that all transistor are identical.

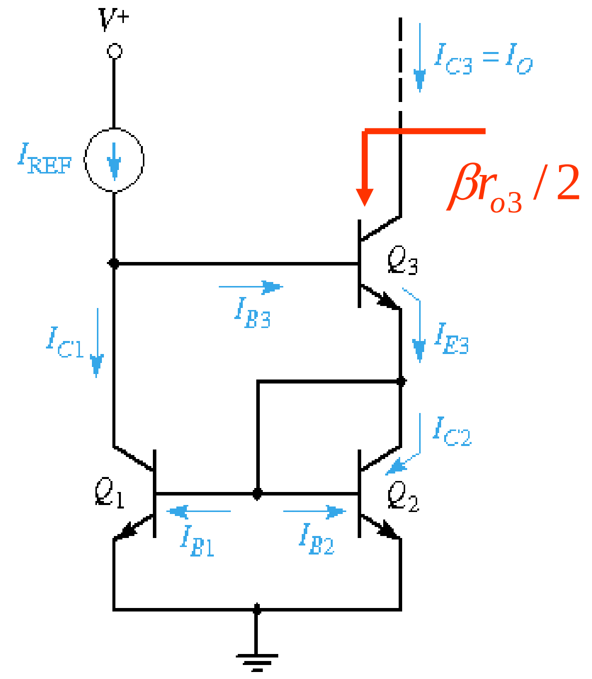


Figure 10.8 Wilson current source

Widlar Current Source

$$I_{REF} \approx I_{C1} = I_S e^{V_{BE1}/V_T} \quad (\beta >> 1)$$

$$I_O = I_{C2} = I_S e^{V_{BE2}/V_T}$$

$$V_{BE1} = V_T \ln\left(\frac{I_{REF}}{I_S}\right)$$

$$V_{BE2} = V_T \ln\left(\frac{I_O}{I_S}\right)$$

$$V_{BE1} - V_{BE2} = V_T \ln\left(\frac{I_{REF}}{I_O}\right)$$

$$= I_{E2} R_E \approx I_O R_E$$

$$I_O R_E = V_T \ln\left(\frac{I_{REF}}{I_O}\right)$$

Assume that all transistor are identical.

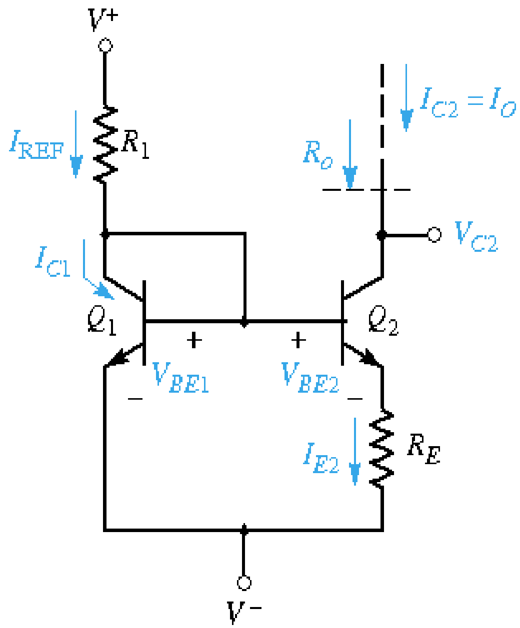


Figure 10.9 Widlar current source

$$V_{BE1} > V_{BE2} \Rightarrow I_O << I_{REF}$$

Design Example 10.4 Objective: Design a Widlar current source to achieve specified reference and load currents.

Design the Widlar current source to produce $I_{\text{REF}} = 1\,\text{mA}$ and $I_O = 12\,\mu\text{A}$. Let $V^+ = 5\,\text{V}$ and $V^- = -5\,\text{V}$. Assume $V_{BE1} = 0.7\,\text{V}$ at the reference current of 1 mA.

Solution: Resistance R_1 is

$$R_1 = \frac{V^+ - V_{BE1} - V^-}{I_{\text{REF}}} = \frac{5 - 0.7 - (-5)}{1} = 9.3\,\text{k}\Omega$$

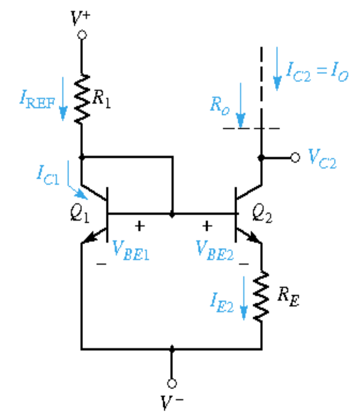


Figure 10.9 Widlar current source

Resistance R_E is, from Equation (10.30),

$$R_E = \frac{V_T}{I_O} \ln\left(\frac{I_{\text{REF}}}{I_O}\right) = \frac{0.026}{0.012} \ln\left(\frac{1}{0.012}\right) = 9.58\,\text{k}\Omega$$

From Equation (10.29), we can determine the difference between the two B–E voltages, as follows:

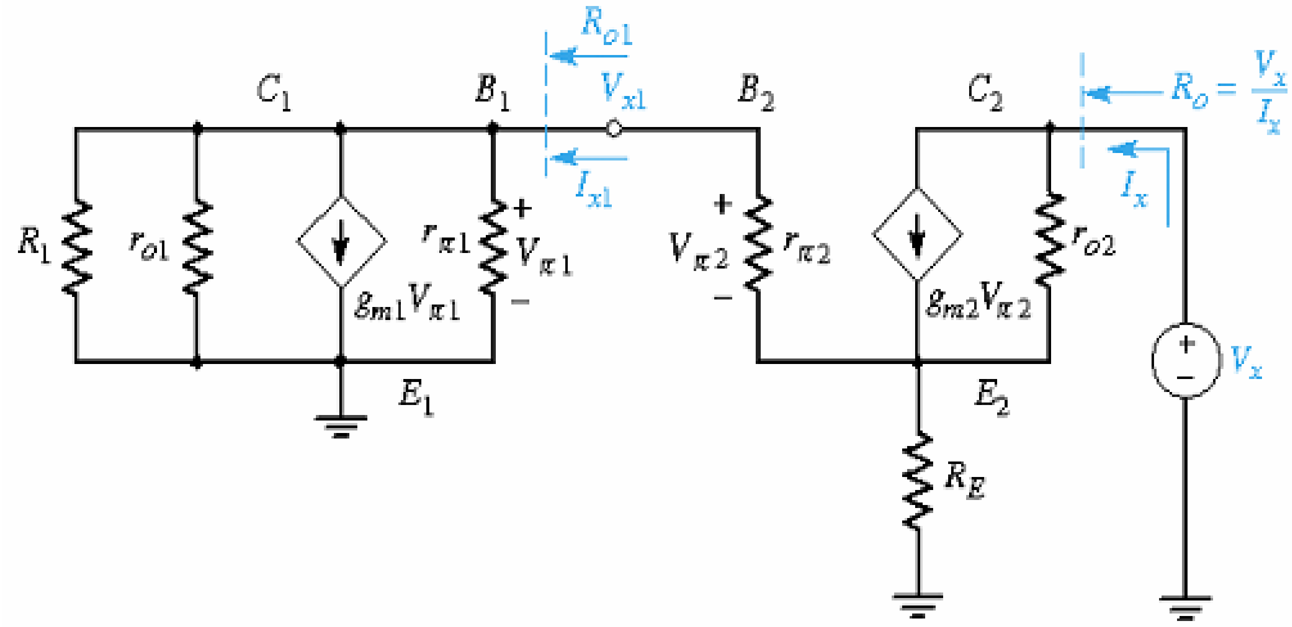
$$V_{BE1} - V_{BE2} = I_O R_E = (12 \times 10^{-6})(9.58 \times 10^3) = 0.115\,\text{V}$$

Output Resistance

$$R_{o1} = r_{\pi 1} \,//\, \frac{1}{g_{m1}} \,//\, r_{o1} \,//\, R_1$$

(For typical parameters, it is small.)

Next, we calculate the approximate value for R_{o1}. If $I_{\mathrm{REF}} = 1\,\mathrm{mA}$, then for $\beta = 100$, $r_{\pi 1} = 2.6\,\mathrm{k}\Omega$ and $g_{m1} = 38.5\,\mathrm{mA/V}$. Assume that $R_1 = 9.3\,\mathrm{k}\Omega$ and $r_{o1} = \infty$. For these conditions, $R_{o1} \cong 0.026\,\mathrm{k}\Omega = 26\,\Omega$. For a load current of $I_O = 12\,\mu\mathrm{A}$, we find $r_{\pi 2} = 217\,\mathrm{k}\Omega$. Resistance R_{o1} is in series with $r_{\pi 2}$, and since $R_{o1} \ll r_{\pi 2}$, we can neglect the effect of R_{o1}, which means that the base of Q_2 is essentially at signal ground.



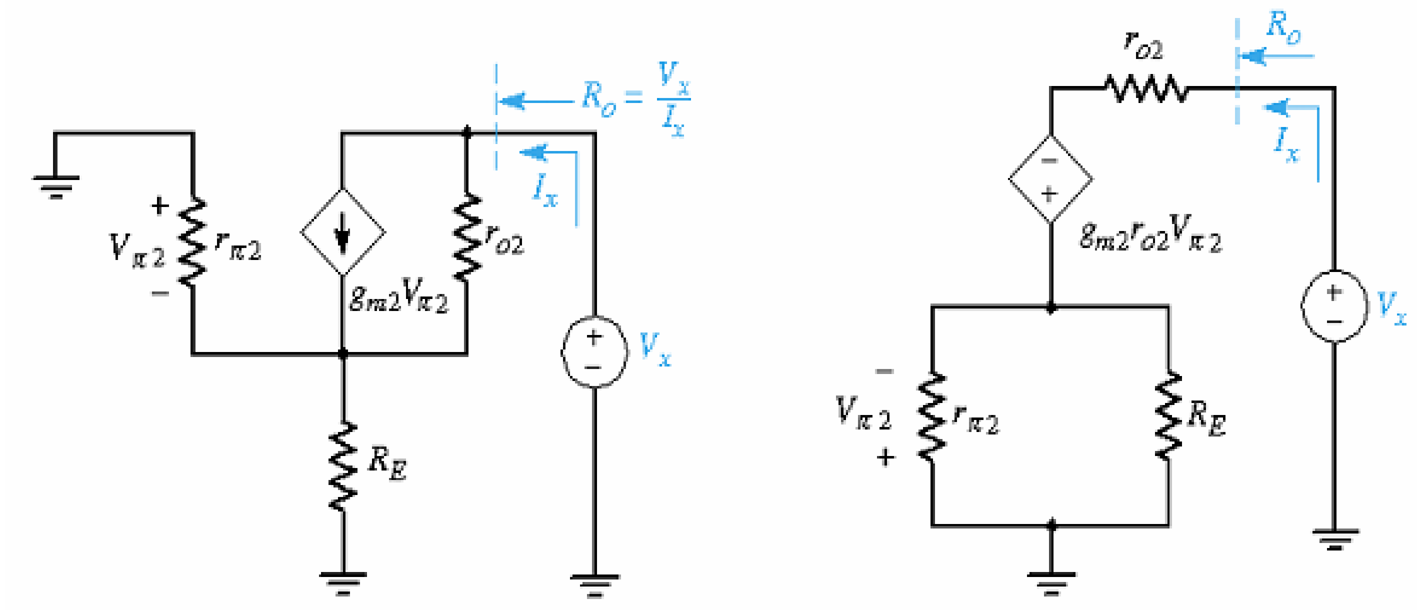
Output Resistance

$$V_{\pi 2} = -I_x(R_E // r_{\pi 2})$$

$$V_x = I_x r_{o2} - g_{m2} r_{o2} V_{\pi 2} + I_x(R_E // r_{\pi 2})$$

$$\frac{V_x}{I_x} = R_o = r_{o2}\left[1 + (R_E // r_{\pi 2})(g_{m2} + 1/r_{o2})\right]$$

$$\approx r_{o2}[1 + g_{m2}(R_E // r_{\pi 2})]$$

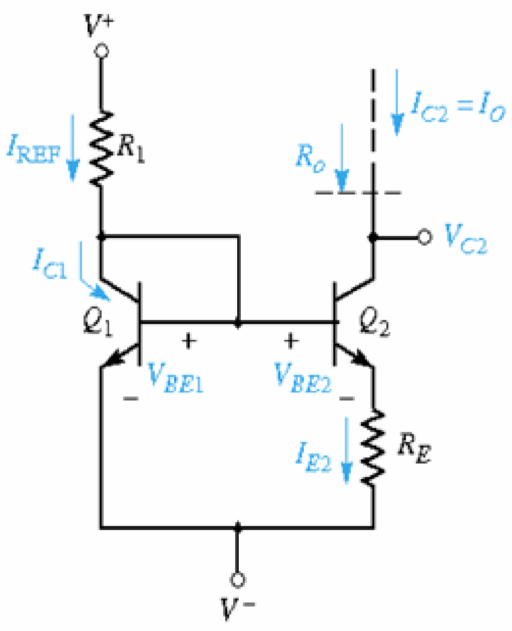


Example 10.5 **Objective:** Determine the change in load current with a change in collector voltage in a Widlar current source.

Consider the circuit in Figure 10.9. The parameters are: $V^+ = 5\,\text{V}$, $V^- = -5\,\text{V}$, $R_1 = 9.3\,\text{k}\Omega$, and $R_E = 9.58\,\text{k}\Omega$. Let $V_A = 80\,\text{V}$ and $\beta = 100$. Determine the change in I_O as V_{C2} changes by 4 V.

Solution: From Example 10.4, we have $I_O = 12\,\mu\text{A}$. The small-signal collector resistance is

$$r_{o2} = \frac{V_A}{I_O} = \frac{80}{0.012} \Rightarrow 6.67\,\text{M}\Omega$$



We can determine that

$$g_{m2} = \frac{I_O}{V_T} = \frac{0.012}{0.026} = 0.462\,\text{mA/V}$$

and

$$r_{\pi 2} = \frac{\beta V_T}{I_O} = \frac{(100)(0.026)}{0.012} = 217\,\text{k}\Omega$$

The output resistance of the circuit is

$$R_o = r_{o2}[1 + g_{m2}(R_E \| r_{\pi 2})] = (6.67) \cdot [1 + (0.462)(9.58 \| 217)] = 34.9\,\text{M}\Omega$$

From Equation (10.31), the change in load current is

$$dI_O = \frac{1}{R_o} dV_{C2} = \frac{1}{34.9 \times 10^6} \times 4 \Rightarrow 0.115\,\mu\text{A}$$

The percentage change in output current is then

$$\frac{dI_O}{I_O} = \frac{0.115}{12} = 0.0096 \Rightarrow 0.96\%$$

Multitransistor Current Mirrors

$$I_{O1} = I_{O2} = \ldots = I_{ON} = \frac{I_{REF}}{1 + \dfrac{1+N}{\beta}}$$

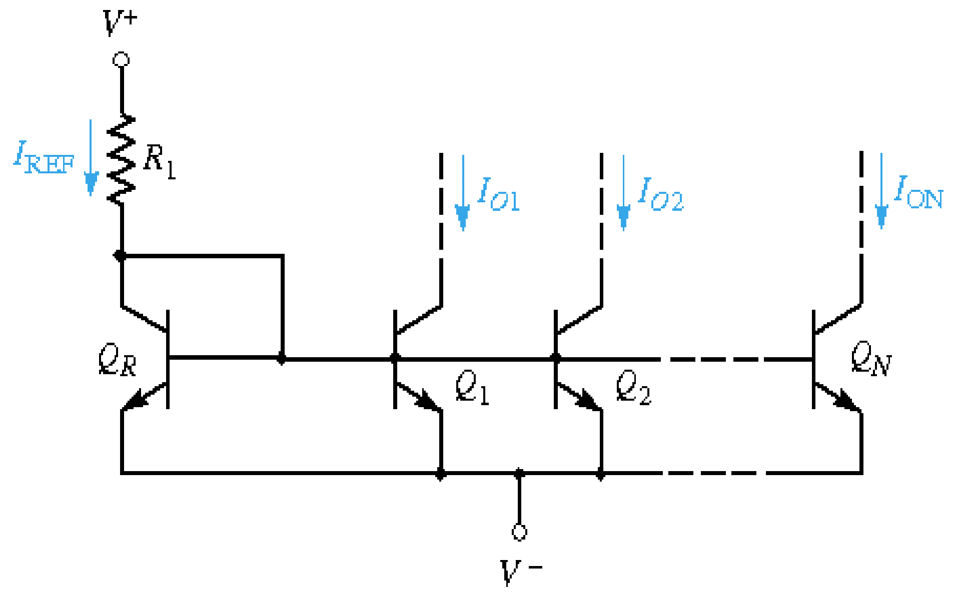
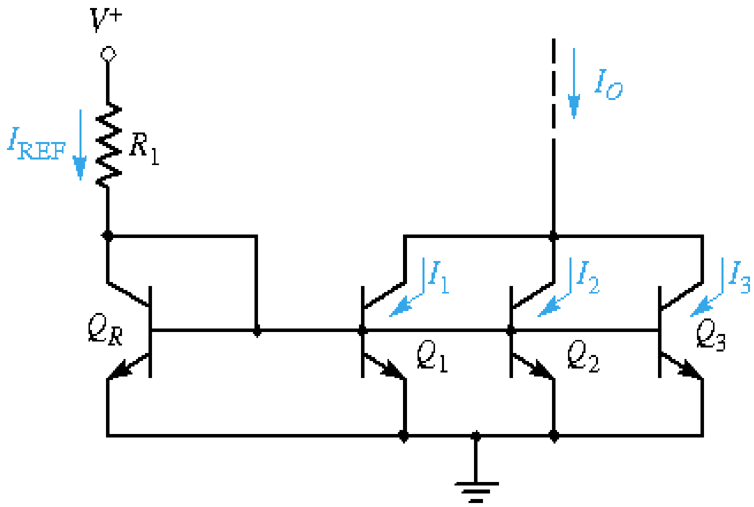


Figure 10.12 Multitransistor current mirror

Multioutput Transistor Current Source



$$I_1 = I_2 = I_3 \approx I_{REF}$$

$$I_O \approx 3I_{REF}$$

Figure 10.13 Multioutput transistor current source

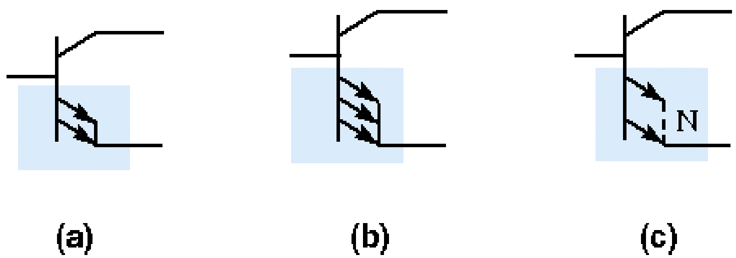


Figure 10.14 Equivalent circuit symbols (a) two transistors in parallel, (b) three transistors in parallel, and (c) *N* transistors in parallel

Generalized Current Mirrors

Assume that all transistor are identical and the base current effects are neglected.

$$I_{REF} = \frac{V^+ - V_{EB}(Q_{R1}) - V_{BE}(Q_{R2}) - V^-}{R_1}$$

$$I_{O1} = I_{REF}$$

$$I_{O2} = 2I_{REF}$$

$$I_{O3} = I_{REF}$$

$$I_{O4} = 3I_{REF}$$

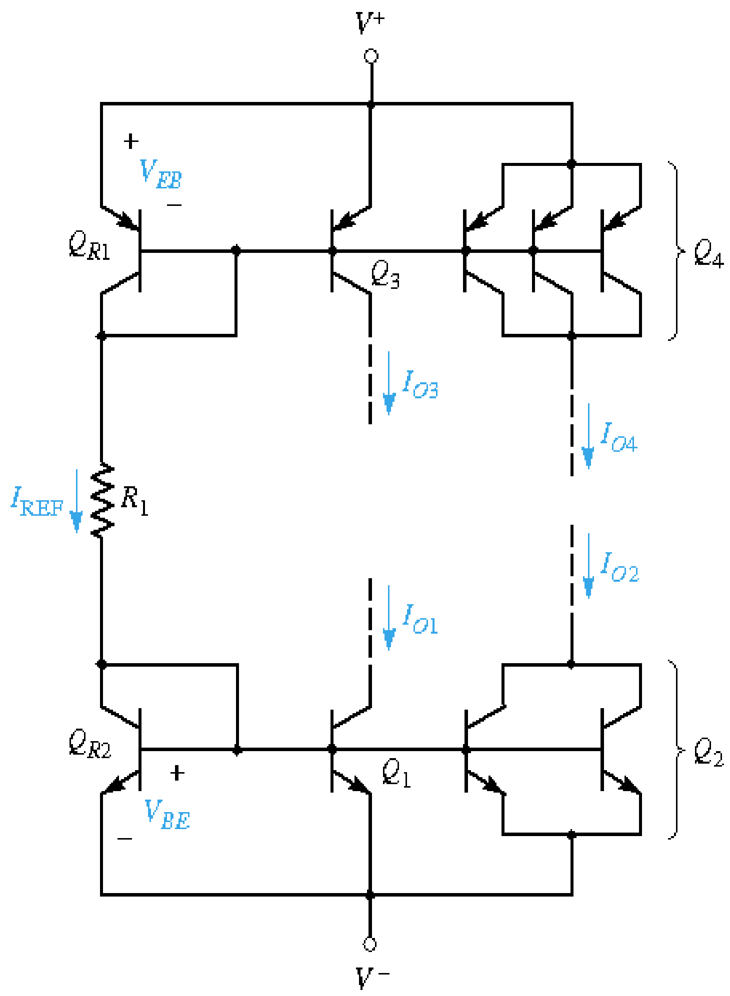


Figure 10.15 Generalized current mirror

Design Example 10.6 Objective: Design a generalized current mirror.

Consider the current mirror shown in Figure 10.15, with parameters $V^+ = 5\,\text{V}$ and $V^- = -5\,\text{V}$. Neglect base currents and assume $V_{BE} = V_{EB} = 0.6\,\text{V}$. Design the circuit such that $I_{O2} = 400\,\mu\text{A}$. Determine I_{REF}, I_{O1}, I_{O3}, I_{O4}, and R_1.

Solution: For $I_{O2} = 400\,\mu\text{A}$, we have

$$I_{\text{REF}} = I_{O1} = I_{O3} = 200\,\mu\text{A} \qquad \text{and} \qquad I_{O4} = 600\,\mu\text{A}$$

Resistor R_1 is

$$R_1 = \frac{V^+ - V_{EB}(\text{Q}_{\text{R1}}) - V_{BE}(\text{Q}_{\text{R2}}) - V^-}{I_{\text{REF}}} = \frac{5 - 0.6 - 0.6 - (-5)}{0.2}$$

or

$$R_1 = 44\,\text{k}\Omega$$

Comment: If the load and reference currents are to be within a factor of approximately four of each other, it is more efficient, from an IC point of view, to adjust the B–E areas of the transistors to achieve the specified currents rather than use the Widlar current source with its additional resistors.

Two Transistor MOSFET Current Source

$$I_{REF} = K_{n1}(V_{GS} - V_{TN1})^2$$

$$V_{GS} = V_{TN1} + \sqrt{\frac{I_{REF}}{K_{n1}}}$$

$$I_O = K_{n2}(V_{GS} - V_{TN2})^2$$

$$= K_{n2}\left[\sqrt{\frac{I_{REF}}{K_{n1}}} + V_{TN1} - V_{TN2}\right]^2$$

If M1 and M2 are identical transistors, then $I_O = I_{REF}$.

The output current can be controlled by aspect ratios,

$$I_O = \frac{K_{n2}}{K_{n1}} \cdot I_{REF} = \frac{(W/L)_2}{(W/L)_1} \cdot I_{REF}$$

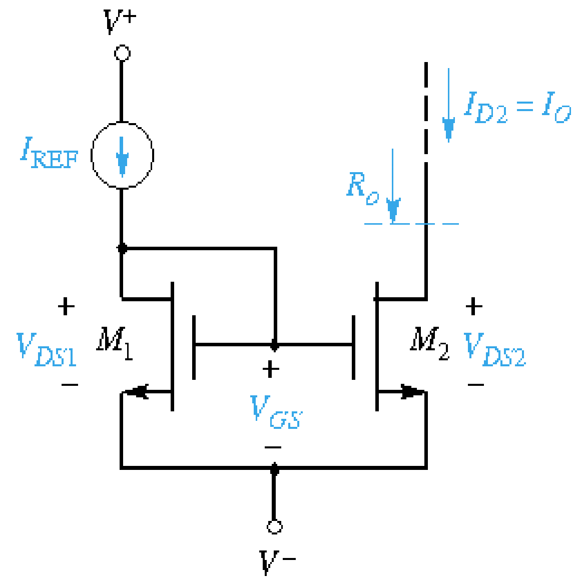


Figure 10.16 Basic two-transistor MOSFET current source

Output Resistance

- Taking into account the finite output resistance of transistors,

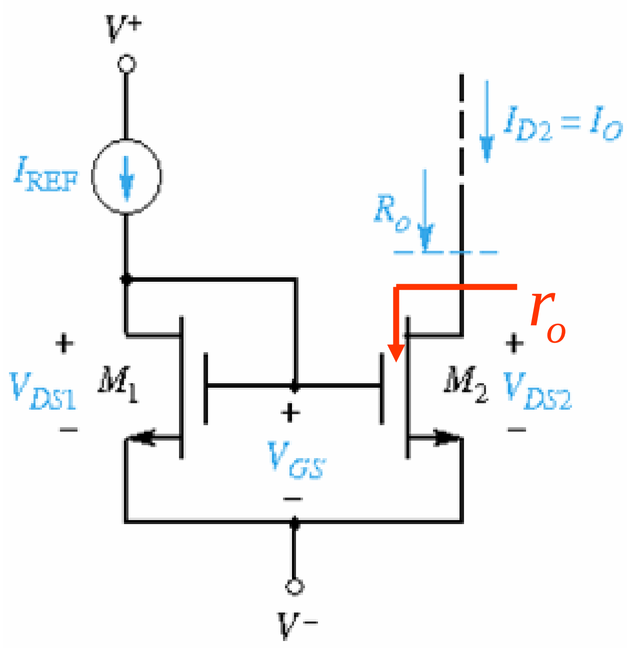
$$I_{REF} = K_{n1}(V_{GS} - V_{TN1})^2(1+\lambda_1 V_{DS1})$$
$$I_O = K_{n2}(V_{GS} - V_{TN2})^2(1+\lambda_2 V_{DS2})$$

Assume that all physical parameters are identical for both devices.

$$\frac{I_O}{I_{REF}} = \frac{(W/L)_2}{(W/L)_1} \cdot \frac{(1+\lambda V_{DS2})}{(1+\lambda V_{DS1})}$$

Let $(W/L)_2 = (W/L)_1$.

$$\frac{1}{R_O} = \frac{dI_O}{dV_{DS2}} \approx \lambda I_{REF} = \frac{1}{r_o}$$



Reference Current

- The M3 transistor is used as a resistor.

$$K_{n1}(V_{GS1}-V_{TN1})^2 = K_{n3}(V_{GS3}-V_{TN3})^2$$

$$\begin{cases} V_{GS1} = \sqrt{\dfrac{(W/L)_3}{(W/L)_1}} \cdot V_{GS3} + \left(1-\sqrt{\dfrac{(W/L)_3}{(W/L)_1}}\right) \cdot V_{TN} \\ V^+ = V_{GS1} + V_{GS3} + V^- \end{cases}$$

$$V_{GS1} = \frac{k}{1+k} \cdot (V^+ - V^-) + \frac{1-k}{1+k} \cdot V_{TN}$$
$$= V_{GS2}$$

$$k = \sqrt{\frac{(W/L)_3}{(W/L)_1}}$$

$$I_O = \left(\frac{W}{L}\right)_2 \left(\frac{1}{2}\mu_n C_{ox}\right)(V_{GS2} - V_{TN})^2$$

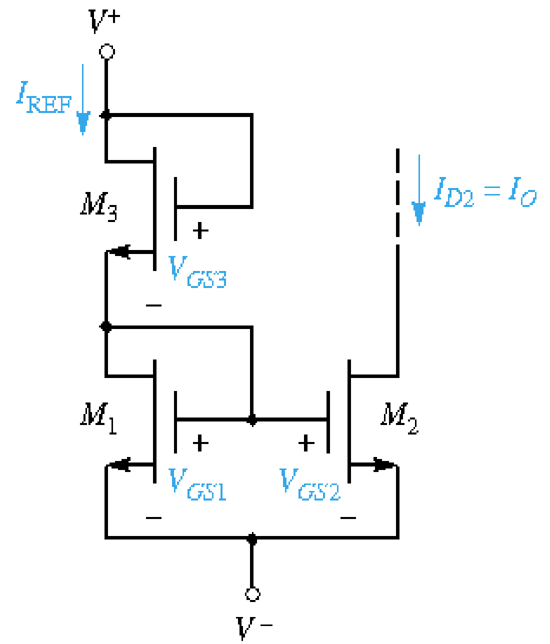


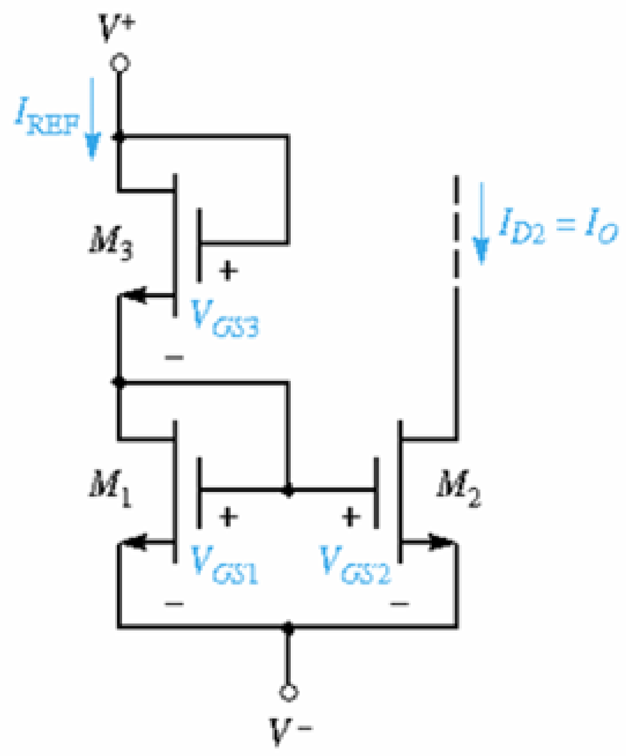
Figure 10.17 MOSFET current source

Design Example 10.7 **Objective:** Design a MOSFET current source to meet specified current values.

Consider the current source in Figure 10.17, with transistor parameters $\frac{1}{2}\mu_n C_{ox} = 20\,\mu\text{A/V}^2$, $V_{TN} = 1\,\text{V}$, and $\lambda = 0$. Let $V^+ = 5\,\text{V}$ and $V^- = 0$. Design the circuit such that $I_{\text{REF}} = 0.25\,\text{mA}$ and $I_O = 0.10\,\text{mA}$.

Solution: If we choose V_{GS2} to be fairly small, yet greater than V_{TN}, then M_2 will remain biased in the saturation region over a fairly large range of V_{DS2} values. Let $V_{GS2} = 1.85\,\text{V}$. Then, from Equation (10.53), we can write

$$\left(\frac{W}{L}\right)_2 = \frac{I_O}{\left(\frac{1}{2}\mu_n C_{ox}\right)(V_{GS2} - V_{TN})^2} = \frac{0.10}{(0.02)(1.85 - 1)^2} = 6.92$$



The reference current is

$$I_{\text{REF}} = \left(\frac{W}{L}\right)_1 \left(\frac{1}{2}\mu_n C_{ox}\right)(V_{GS1} - V_{TN})^2$$

Since $V_{GS1} = V_{GS2}$, we have

$$\left(\frac{W}{L}\right)_1 = \frac{I_{\text{REF}}}{\left(\frac{1}{2}\mu_n C_{ox}\right)(V_{GS2} - V_{TN})^2} = \frac{0.25}{(0.02)(1.85 - 1)^2} = 17.3$$

The value of V_{GS3} is

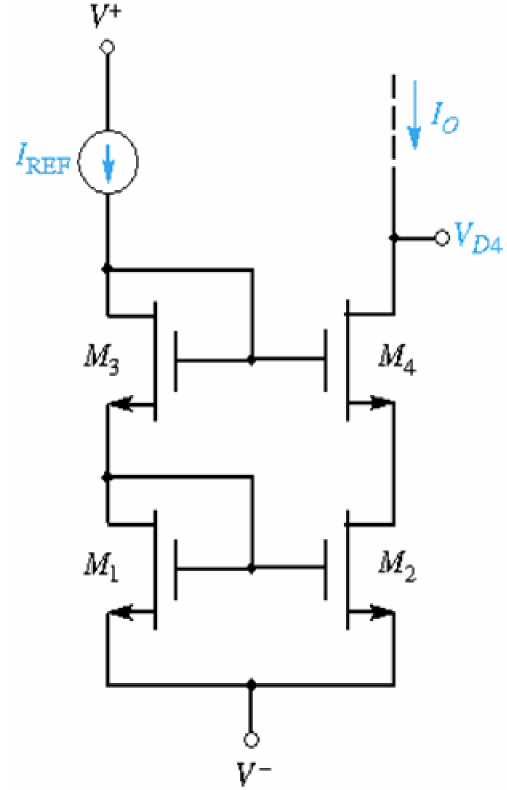
$$V_{GS3} = (V^+ - V^-) - V_{GS1} = 5 - 1.85 = 3.15\,\text{V}$$

Then, since $I_{\text{REF}} = K_{n3}(V_{GS3} - V_{TN})^2$, we have

$$\left(\frac{W}{L}\right)_3 = \frac{I_{\text{REF}}}{\left(\frac{1}{2}\mu_n C_{ox}\right)(V_{GS3} - V_{TN})^2} = \frac{0.25}{(0.02)(3.15 - 1)^2} = 2.70$$

Cascode Current Mirror

- The output resistance is designed to be much greater than that of the two-transistor circuit.

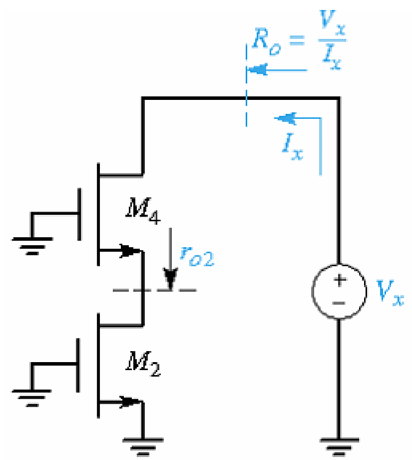


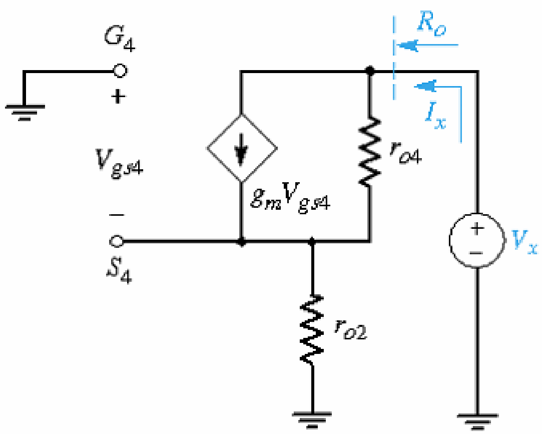
$$
\begin{cases}
I_x = g_m V_{gs4} + \dfrac{V_x - (-V_{gs4})}{r_{o4}} \\
V_{gs4} = -I_x r_{o2}
\end{cases}
$$

$$I_x + \frac{r_{o2}}{r_{o4}} I_x + g_m r_{o2} I_x = \frac{V_x}{r_{o4}}$$

$$R_o = \frac{V_x}{I_x} = r_{o4} + r_{o2}(1 + g_m r_{o4})$$

$$\approx r_{o4} + g_m r_{o2} r_{o4}$$





Example 10.8 **Objective:** Compare the output resistance of the cascode MOSFET current source to that of the two-transistor current source.

Consider the two-transistor current source in Figure 10.17 and the cascode current source in Figure 10.19. Assume $I_{\mathrm{REF}} = I_O = 100\,\mu\mathrm{A}$ in both circuits, $\lambda = 0.01\,\mathrm{V}^{-1}$ for all transistors, and $g_m = 0.5\,\mathrm{mA/V}$.

Solution: The output resistance of the two-transistor current source is, from Equation (10.48)

$$r_o = \frac{1}{\lambda I_{\mathrm{REF}}} = \frac{1}{(0.01)(0.10)} \Rightarrow 1\,\mathrm{M\Omega}$$

For the cascode circuit, we have $r_{o2} = r_{o4} = 1\,\mathrm{M\Omega}$. Therefore, the output resistance of the cascode circuit is, from Equation (10.57),

$$R_o = r_{o4} + r_{o2}(1 + g_m r_{o4}) = 1 + (1)[1 + (0.5 \times 10^{-3})(10^6)]$$

or

$$R_o = 502\,\mathrm{M\Omega}$$

Comment: The output resistance of the cascode current source is substantially larger than that of the basic two-transistor circuit. Since $dI_O \propto 1/R_o$, the load current in the cascode circuit is more stable against variations in output voltage.

Design Pointer: Achieving the output resistance of $502\,\mathrm{M\Omega}$ assumes the transistors are ideal. In fact, small leakage currents will begin to be a factor in actual output resistance values, so a value of $502\,\mathrm{M\Omega}$ may not be achieved in reality.

Wilson Current Mirror

- For modified MOSFET Wilson current source circuit, the drain-to-source voltages of M1, M2, and M4 are held constant.
- The primary advantage of these circuits is the increase in output resistance, which further stabilizes the load current.

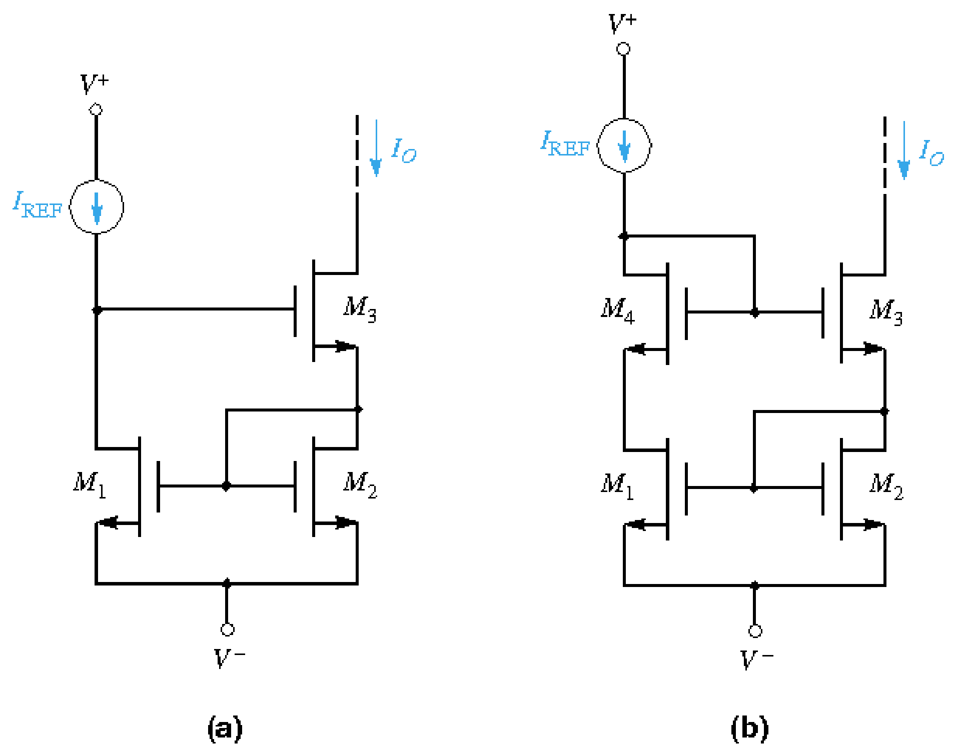


Figure 10.21 (a) MOSFET Wilson current source and (b) modified MOSFET Wilson current source

Bias-Independent Current Source

- PMOS devices are matched, the current I_{D1} and I_{D2} are equal.

$$I_{D1} = \frac{k_n'}{2}\left(\frac{W}{L}\right)_1 (V_{GS1} - V_{TN})^2$$

$$= I_{D2} = \frac{k_n'}{2}\left(\frac{W}{L}\right)_2 (V_{GS2} - V_{TN})^2$$

$$\sqrt{\frac{(W/L)_1}{(W/L)_2}} \cdot (V_{GS1} - V_{TN}) = V_{GS2} - V_{TN}$$

$$V_{GS2} = V_{GS1} - I_{D2}R = V_{GS1} - I_{D1}R$$

$$\because I_{D1} = K_{n1}(V_{GS1} - V_{TN})^2$$

$$\Rightarrow V_{GS1} - V_{TN} = \sqrt{\frac{I_{D1}}{K_{n1}}}$$

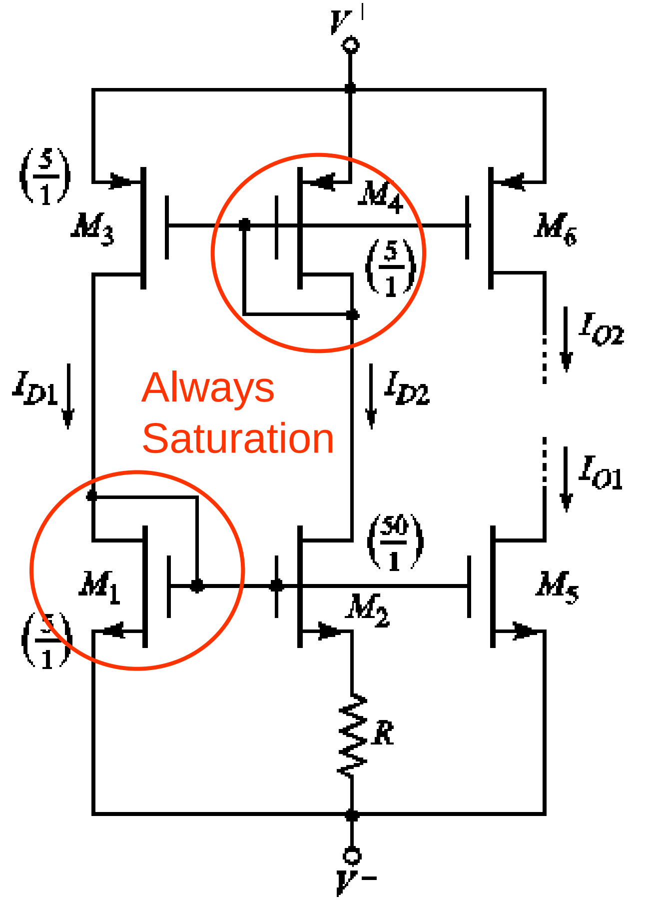


Figure 10.23 Bias-independent MOSFET current mirror

Bias-Independent Current Source

$$\sqrt{\frac{(W/L)_1}{(W/L)_2}}\cdot(V_{GS1}-V_{TN})=V_{GS2}-V_{TN}$$

$$\Downarrow$$

$$\sqrt{\frac{(W/L)_1}{(W/L)_2}}\cdot\sqrt{\frac{I_{D1}}{K_{n1}}}=V_{GS1}-V_{TN}-I_{D1}R$$

$$=\sqrt{\frac{I_{D1}}{K_{n1}}}-I_{D1}R$$

$$R=\frac{1}{\sqrt{K_{n1}I_{D1}}}\left(1-\sqrt{\frac{(W/L)_1}{(W/L)_2}}\right)$$

(For a given current I_{D1}, one can find the value of R.)

The currents I_{D1} and I_{D2} are independent of the supplied voltages as long as M2 and M3 are biased in the saturation region.

JFET Current Sources

- The device remains biased in the saturation region.

$$v_{DS} \geq v_{DS}(\text{sat}) = v_{GS} - V_P = |V_P| \qquad (V_P \text{ is negative})$$

In the saturation region, the current is

$$i_D = I_{DSS}\left(1 - \frac{v_{GS}}{V_P}\right)^2 (1 + \lambda v_{DS}) = I_{DSS}(1 + \lambda v_{DS})$$

$$\frac{1}{r_o} = \frac{di_D}{dv_{DS}} = \lambda I_{DSS}$$

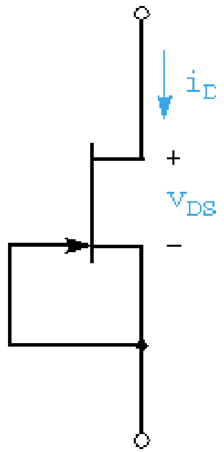


Figure 10.24 Depletion-mode JFET connected as a current source

Example 10.9 **Objective:** Determine the currents and voltages in a simple JFET circuit biased with a constant-current source.

Consider the circuit shown in Figure 10.25. The transistor parameters are: $I_{DSS1} = 2\,\text{mA}$, $I_{DSS2} = 1\,\text{mA}$, $V_{P1} = V_{P2} = -1.5\,\text{V}$, and $\lambda_1 = \lambda_2 = 0.05\,\text{V}^{-1}$. Determine the minimum values of V_S and V_I such that Q_2 is biased in the saturation region. What is the value of I_O?

Solution: In order for Q_2 to remain biased in the saturation region, we must have $v_{DS} \geq |V_P| = 1.5\,\text{V}$, from Equation (10.69). The minimum value of V_S is then

$$V_S(\text{min}) - V^- = v_{DS}(\text{min}) = 1.5\,\text{V}$$

or

$$V_S(\text{min}) = 1.5 + V^- = 1.5 + (-5) = -3.5\,\text{V}$$

From Equation (10.70), the output current is

$$i_D = I_O = I_{DSS2}(1 + \lambda v_{DS}) = (1)[1 + (0.05)(1.5)] = 1.08\,\text{mA}$$

As a first approximation in calculating the minimum value of V_I, we neglect the effect of λ in transistor Q_1. Then, assuming Q_1 is biased in the saturation region, we have

$$i_D = I_{DSS1}\left(1 - \frac{v_{GS1}}{V_{P1}}\right)^2$$

or

$$1.08 = 2\left(1 - \frac{v_{GS1}}{(-1.5)}\right)^2$$

which yields

$$v_{GS1} = -0.40\,\text{V}$$

We see that

$$v_{GS1} = -0.40\,\text{V} = V_I - V_S = V_I - (-3.5)$$

or

$$V_I = -3.90\,\text{V}$$

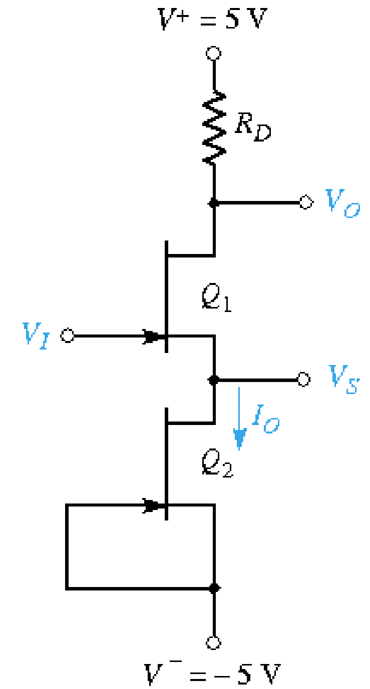


Figure 10.25 The dc equivalent circuit of simple JFET amplifier biased with JFET current source

Cascode JFET Current Source

- Increase the output resistance of a JFET current source

Assuming Q1 and Q2 are identical,

$$i_D = I_{DSS}(1+\lambda v_{DS1}) = I_{DSS}\left(1-\frac{v_{GS2}}{V_P}\right)^2 (1+\lambda v_{DS2})$$

$$v_{GS2} = -v_{DS1},\ \ v_{DS2} = V_{DS} - v_{DS1}$$

For a given V_{DS} , v_{DS1} (and then i_D) can be determined by

$$(1+\lambda v_{DS1}) = \left(1+\frac{v_{DS1}}{V_P}\right)^2 [1+\lambda(V_{DS} - v_{DS1})]$$

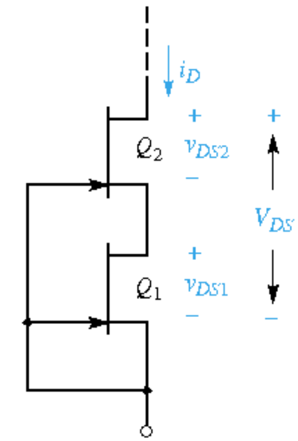


Figure 10.26 JFET cascode current source

- Output Resistance

$$I_x = g_m V_{gs2} + [V_x - (-V_{gs2})]/r_{o2}$$
$$= g_m(-I_x r_{o1}) \ + [V_x - (-V_{gs2})]/r_{o2}$$

$$R_o = \frac{V_x}{I_x} = r_{o2} + r_{o1} + g_m r_{o1} r_{o2}$$
$$= r_{o2} + r_{o1}(1+g_m r_{o2})$$

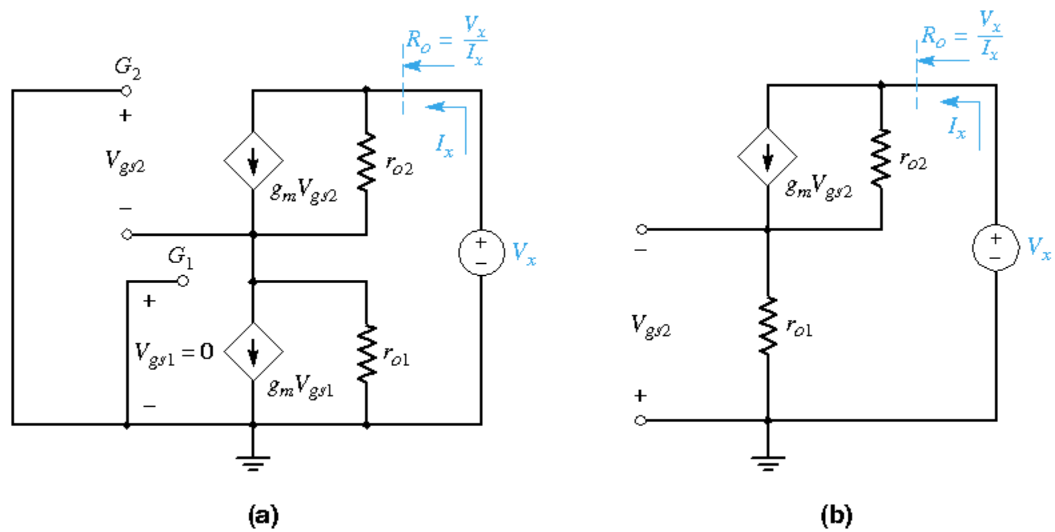


Figure 10.27 (a) Equivalent circuit, using phasor notation, of the JFET cascode current source for determining output resistance and (b) final configuration

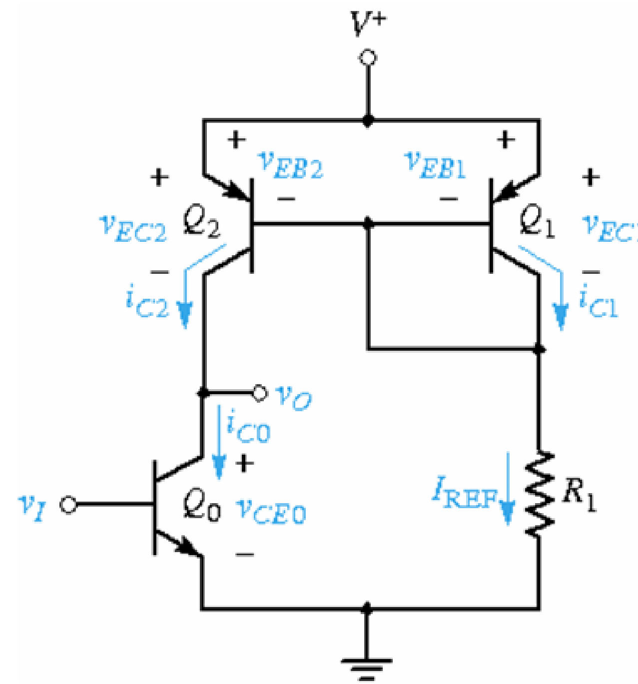
DC Analysis of BJT Active Load Circuits

- Q2 is referred to as the active load device for driver transistor Q0.

$$I_{C0} = I_{S0}[e^{V_I/V_T}]\left(1+\frac{V_{CE0}}{V_{AN}}\right)$$

$$I_{C2} = I_{S2}[e^{V_{EB2}/V_T}]\left(1+\frac{V_{EC2}}{V_{AP}}\right)$$

$$I_{REF} \approx I_{C1} = I_{S1}[e^{V_{EB1}/V_T}]\left(1+\frac{V_{EC1}}{V_{AP}}\right)$$



Assuming Q1 and Q2 are identical,

$$I_{S1} = I_{S2} \text{ and } V_{EC1} = V_{EB1} = V_{EB2}.$$

$$V_O = V_{CE0} = \frac{V_{AN}V_{AP}}{V_{AN}+V_{AP}}\left[1-\frac{I_{S0}e^{V_I/V_T}}{I_{REF}}\right]+\frac{V_{AN}}{V_{AN}+V_{AP}}(V^+ - V_{EB2})$$

Voltage Gain of BJT Active Load Circuits

- Voltage Transfer Function and Load Curve

$$V_{CE0}(\text{sat}) < V_O < V^+ - V_{EC2}(\text{sat})$$

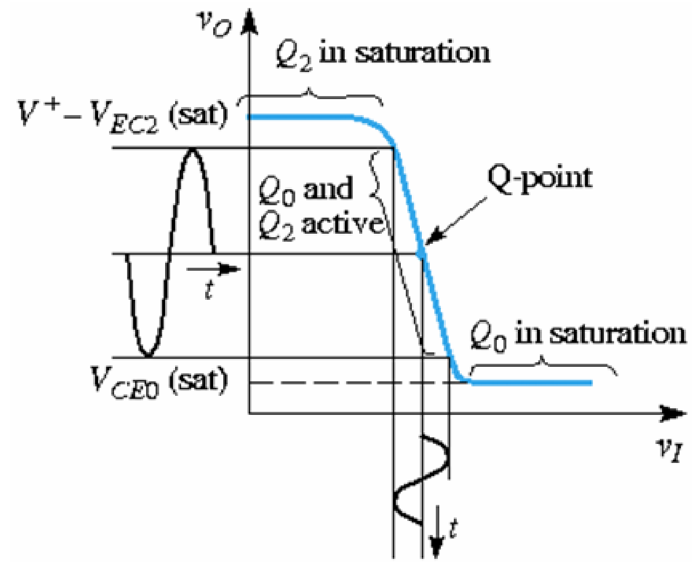
- Voltage Gain

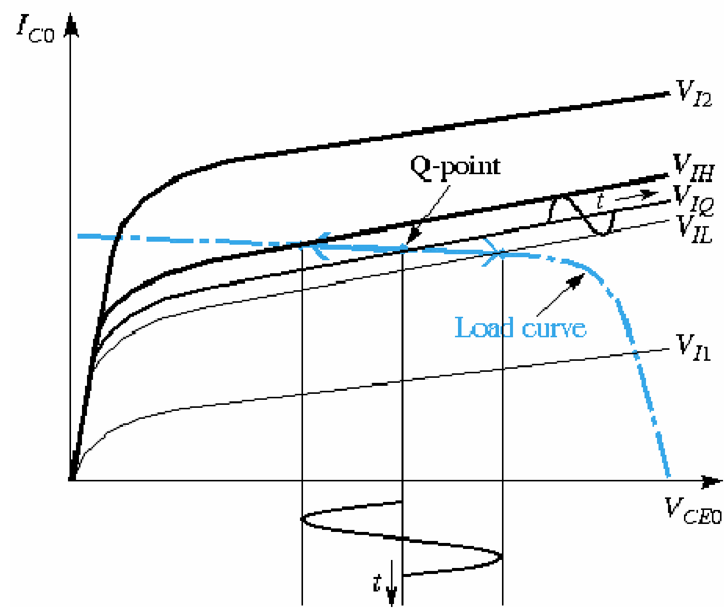
$$A_v = \frac{dV_O}{dV_I} = -\left(\frac{V_{AN}V_{AP}}{V_{AN}+V_{AP}}\right)\left(\frac{I_{S0}}{I_{REF}}\right)\left(\frac{1}{V_T}\right)e^{V_I/V_T}$$

$$I_{REF} \approx I_{S0}e^{V_I/V_T}$$

$$A_v = \frac{dV_O}{dV_I} = -\left(\frac{V_{AN}V_{AP}}{V_{AN}+V_{AP}}\right)\left(\frac{1}{V_T}\right)$$

$$= \frac{-(1/V_T)}{\dfrac{1}{V_{AN}}+\dfrac{1}{V_{AP}}}$$





DC Analysis of MOSFET Active Load Circuit

$$I_{REF} = K_{p1}(V_{SG} - |V_{TP1}|)^2(1+\lambda_1 V_{SD1})$$

$$I_2 = K_{p2}(V_{SG} - |V_{TP2}|)^2(1+\lambda_2 V_{SD2})$$

Assuming M1 and M2 are identical,

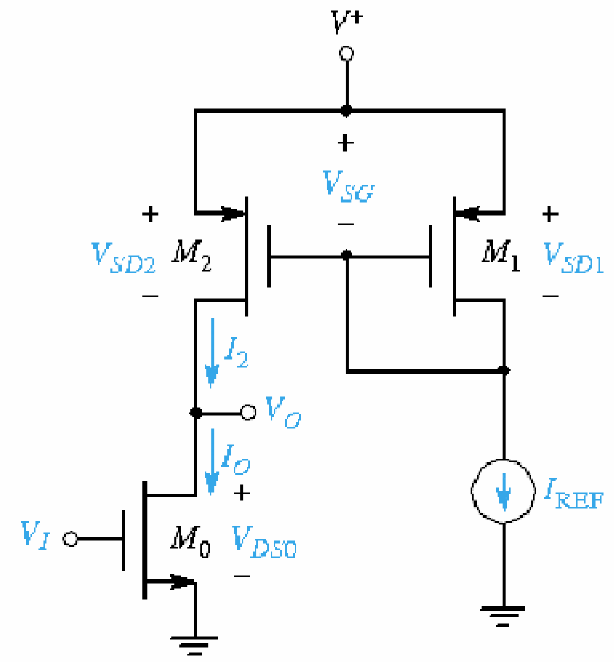
$$\lambda_1 = \lambda_2 \equiv \lambda_p, V_{TP1} = V_{TP2} \equiv V_{TP}, \text{ and } K_{p1} = K_{p2} \equiv K_p.$$

$$V_{SD1} = V_{SG},\ V_O = V^+ - V_{SD2}$$

$$\frac{I_{REF}}{I_2} = \frac{(1+\lambda_p V_{SD1})}{(1+\lambda_p V_{SD2})} = \frac{(1+\lambda_p V_{SG})}{(1+\lambda_p (V^+ - V_O))}$$

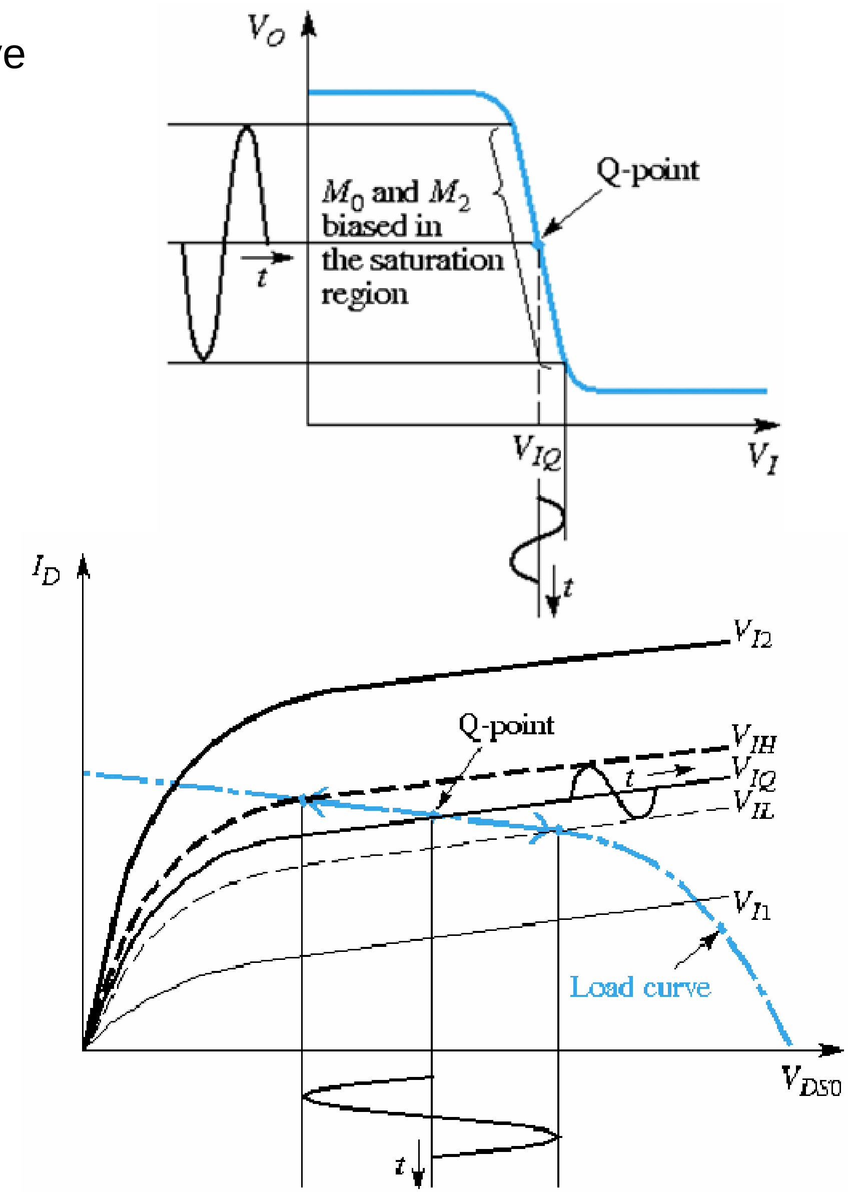
$$= \frac{I_{REF}}{K_n(V_I - V_{TN})^2(1+\lambda_n V_O)}$$

$$V_O = \frac{[1+\lambda_p(V^+ - V_{SG})]}{\lambda_n + \lambda_p} - \frac{K_n(V_I - V_{TN})^2}{I_{REF}(\lambda_n + \lambda_p)}$$



Voltage Gain of MOSFET Active Load Circuits

- Voltage Transfer Function and Load Curve



- Voltage Gain

$$A_v = \frac{dV_O}{dV_I} = \frac{-2K_n(V_I - V_{TN})}{I_{REF}(\lambda_n + \lambda_p)}$$

$$g_m = 2K_n(V_I - V_{TN})$$

$$r_{on} = 1/\lambda_n I_{REF} \qquad r_{op} = 1/\lambda_p I_{REF}$$

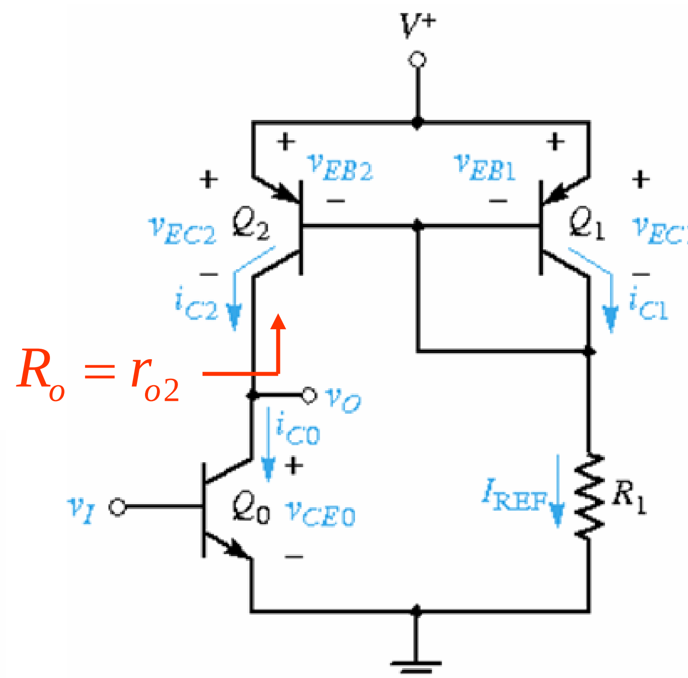
$$A_v = \frac{-g_m}{\left(\frac{1}{r_{on}} + \frac{1}{r_{op}}\right)} = -g_m(r_{on} // r_{op})$$

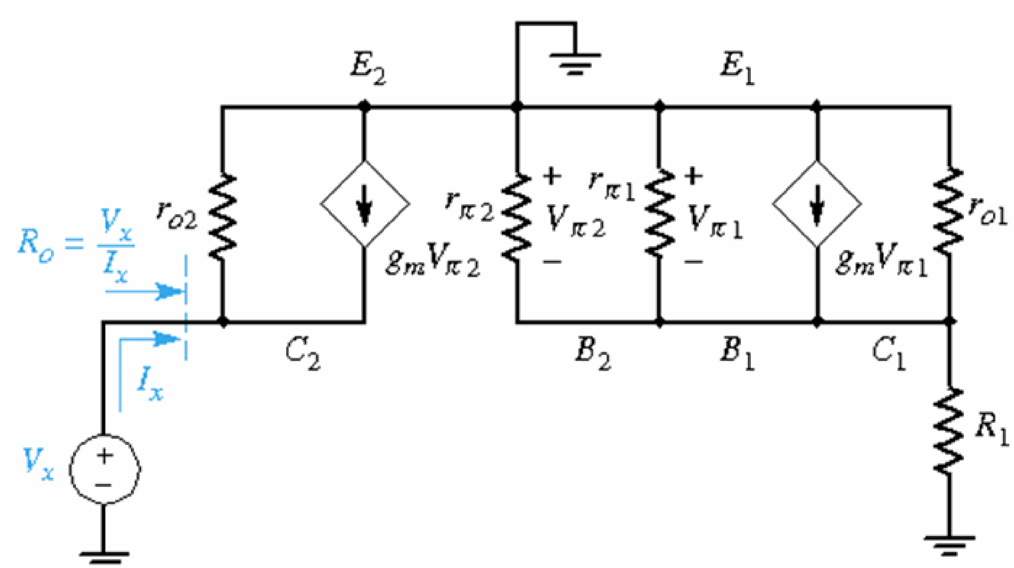
Small-Signal Analysis of BJT Active Load Circuit

- In the Q1 portion of the equivalent circuit, there are no independent AC sources to excite any current or voltages.

$$V_{\pi 1} = V_{\pi 2} = 0$$

$$R_o = r_{o2}$$





Small-Signal Voltage Gain

$$A_v = \frac{V_o}{V_i} = -g_m(r_o // R_L // r_{o2})$$

$$= \frac{-g_m}{\left(\frac{1}{r_o} + \frac{1}{R_L} + \frac{1}{r_{o2}}\right)} = \frac{-g_m}{g_0 + g_L + g_{o2}}$$

$$g_m = I_{Co} / V_T$$

$$g_o = I_{Co} / V_{AN}$$

$$g_{o2} = I_{Co} / V_{AP}$$

$$g_L = 1 / R_L$$

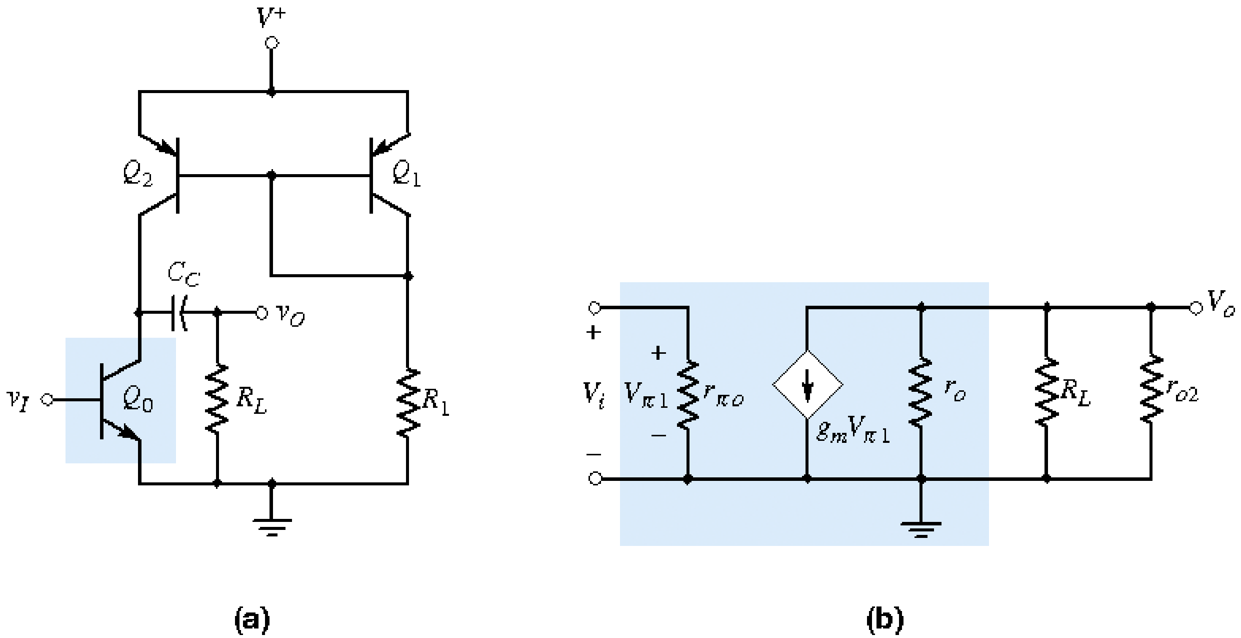


Figure 10.37 (a) Simple BJT amplifier with active load and load resistance and (b) small-signal equivalent circuit

Example 10.11 **Objective:** Calculate the small-signal voltage gain of an amplifier with an active load and a load resistance R_L.

For the circuit in Figure 10.37(a), the transistor parameters are $V_{AN} = 120\,\text{V}$ and $V_{AP} = 80\,\text{V}$. Let $V_T = 0.026\,\text{V}$ and $I_{Co} = 1\,\text{mA}$. Determine the small-signal voltage gain for load resistances of $R_L = \infty$, $100\,\text{k}\Omega$, and $10\,\text{k}\Omega$.

Solution: For $R_L = \infty$, Equation (10.96) reduces to

$$A_v = \frac{-\left(\frac{1}{V_T}\right)}{\left(\frac{1}{V_{AN}} + \frac{1}{V_{AP}}\right)} = \frac{-\left(\frac{1}{0.026}\right)}{\left(\frac{1}{120} + \frac{1}{80}\right)} = -1846$$

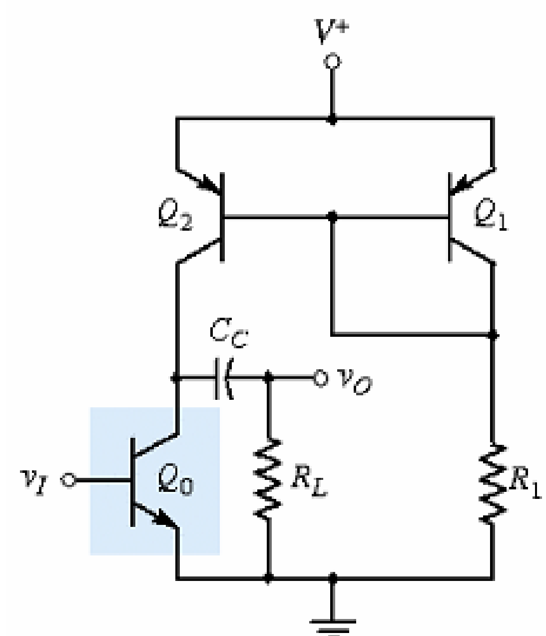
which is the same as that determined for the open-circuit configuration in Example 10.10.

For $R_L = 100\,\text{k}\Omega$, the small-signal voltage gain is

$$A_v = \frac{-\left(\frac{1}{0.026}\right)}{\left(\frac{1}{120} + \frac{1}{100} + \frac{1}{80}\right)} = \frac{-38.46}{0.00833 + 0.010 + 0.0125} = -1247$$

and for $R_L = 10\,\text{k}\Omega$, the voltage gain is

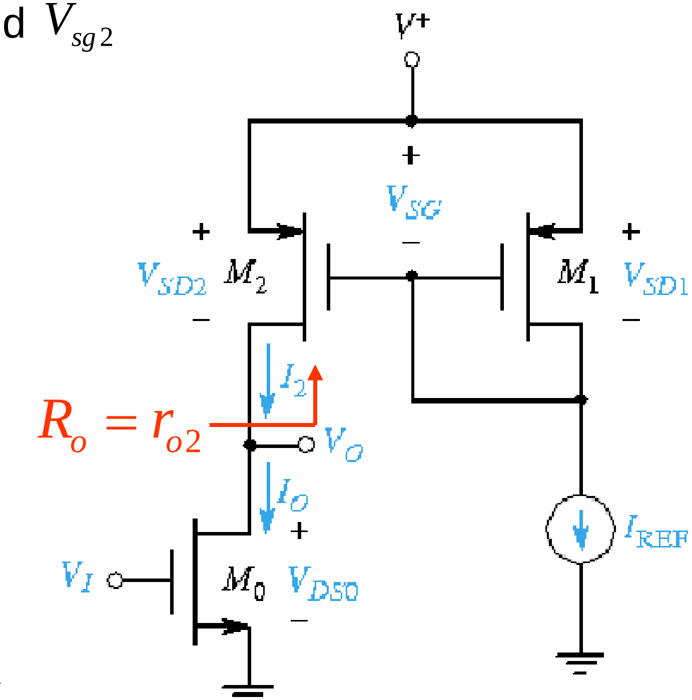
$$A_v = \frac{-\left(\frac{1}{0.026}\right)}{\left(\frac{1}{120} + \frac{1}{10} + \frac{1}{80}\right)} = \frac{-38.46}{0.00833 + 0.10 + 0.0125} = -318$$



Small-Signal Analysis of MOSFET Active Load circuit

- There is no AC excitation, the signal voltage V_{sg1} and V_{sg2} are zero.

$$R_o = r_{o2}$$



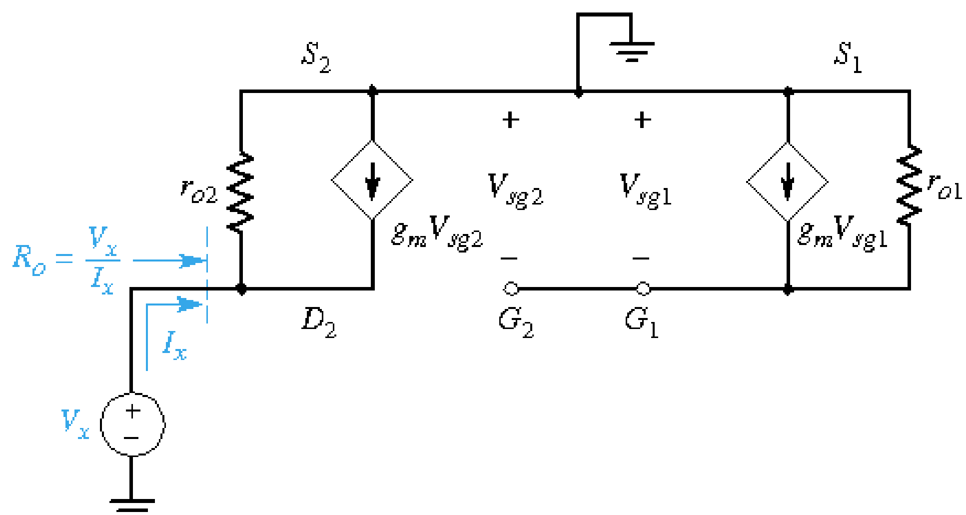


Figure 10.38 Small-signal equivalent circuit of the MOSFET active load circuit

Small-Signal Voltage Gain

$$A_v = \frac{V_o}{V_i} = -g_m(r_o \,//\, R_L \,//\, r_{o2})$$

$$= \frac{-g_m}{g_0 + g_L + g_{o2}}$$

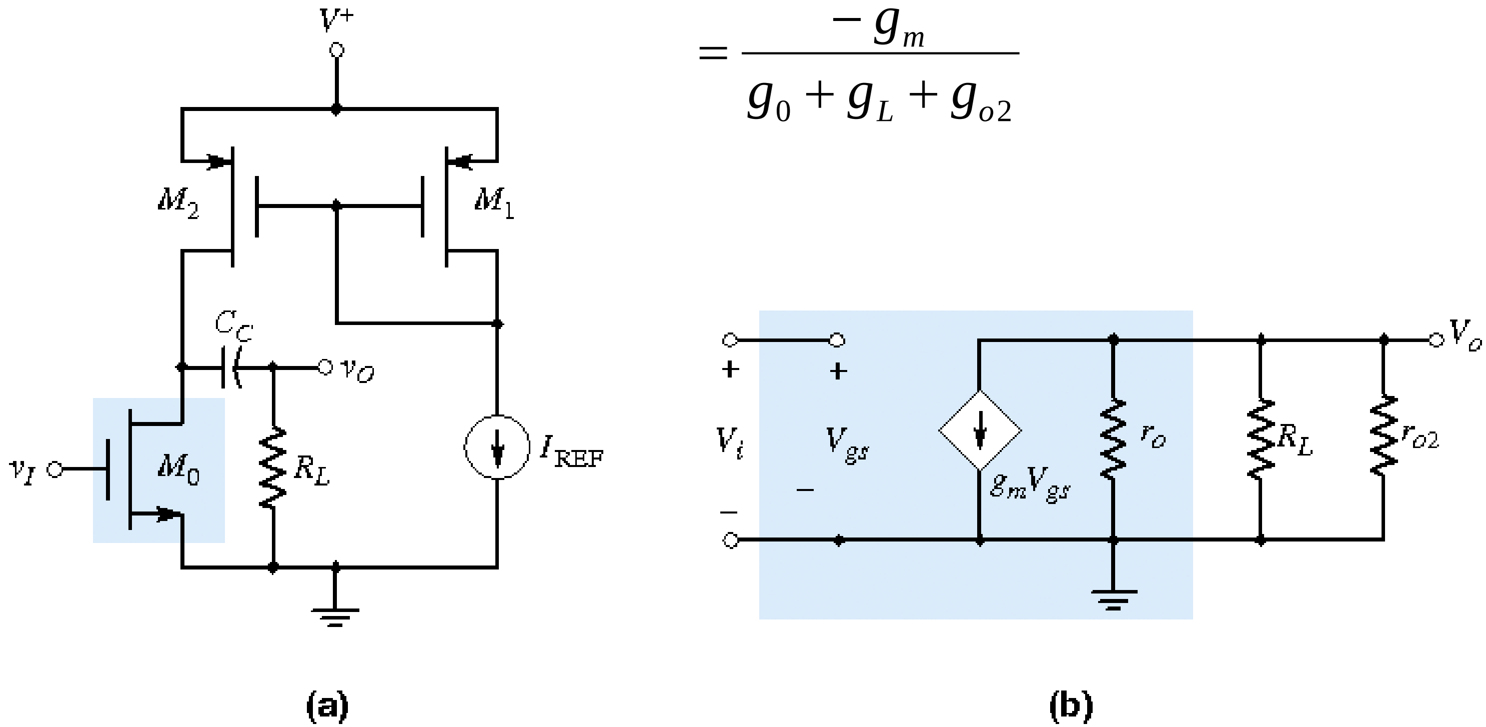


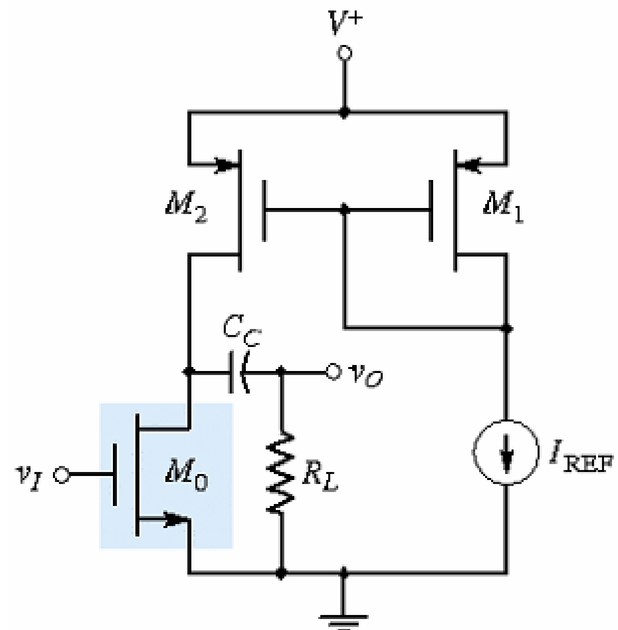
Figure 10.39 (a) Simple MOSFET amplifier with active load and load resistance and (b) small-signal equivalent circuit

Example 10.12 **Objective:** Calculate the small-signal voltage gain of an NMOS amplifier with an active load.

For the amplifier shown in Figure 10.39(a) the transistor parameters are: $\lambda_n = \lambda_p = 0.01\,\mathrm{V}^{-1}$, $V_{TN} = 1\,\mathrm{V}$, and $K_n = 1\,\mathrm{mA/V}^2$. Assume M_1 and M_2 are matched and $I_{\mathrm{REF}} = 0.5\,\mathrm{mA}$. Calculate the small-signal voltage gain for load resistances of $R_L = \infty$ and $100\,\mathrm{k\Omega}$.

Solution: Since M_1 and M_2 are matched, then $I_O = I_{\mathrm{REF}}$, and the transconductance is

$$g_m = 2\sqrt{K_n I_{\mathrm{REF}}} = 2\sqrt{(1)(0.5)} = 1.41\,\mathrm{mA/V}$$



The small-signal transistor conductances are

$$g_o = g_{o2} = \lambda I_{\mathrm{REF}} = (0.01)(0.5) = 0.005\,\mathrm{mA/V}$$

For $R_L = \infty$, Equation (10.99) reduces to

$$A_v = \frac{-g_m}{g_o + g_{o2}} = \frac{-1.41}{0.005 + 0.005} = -141$$

For $R_L = 100\,\mathrm{k\Omega}$ ($g_L = 0.01\,\mathrm{mA/V}$), the voltage gain is

$$A_v = \frac{-g_m}{g_o + g_L + g_{o2}} = \frac{-1.41}{0.005 + 0.01 + 0.005} = -70.5$$

Comment: The magnitude of the small-signal voltage gain of MOSFET amplifiers with active loads is substantially larger than for those with resistive loads, but it is still smaller than equivalent bipolar circuits, because of the smaller transconductance for the MOSFET.

Advanced MOSFET Active Load

$$\begin{cases} g_m V_{gs1} + \dfrac{-V_{gs2}}{r_{o1}} = g_m V_{gs2} + \dfrac{V_o - (-V_{gs2})}{r_{o2}} \\ \dfrac{V_o}{R_{o3}} + \dfrac{V_o - (-V_{gs2})}{r_{o2}} + g_m V_{gs2} = 0 \end{cases}$$

$$A_v = \frac{V_o}{V_i} = \frac{-g_m^2}{\dfrac{g_m}{R_{o3}} + \dfrac{1}{r_{o1} r_{o2}}} \approx \frac{-g_m^2}{\dfrac{1}{r_{o3} r_{o4}} + \dfrac{1}{r_{o1} r_{o2}}}$$

$$R_{o3} = r_{o3} + r_{o4}(1 + g_m r_{o3})$$

(See Cascode Current Mirror, p.24)

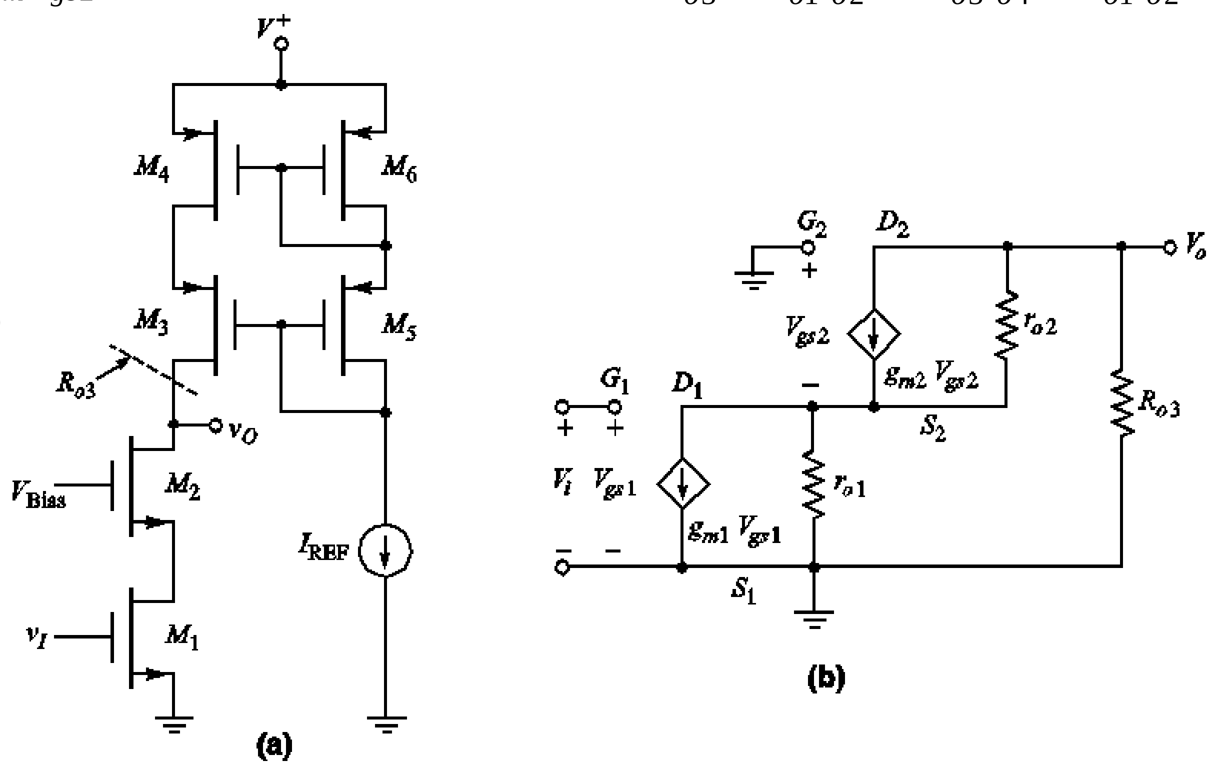


Figure 10.40 (a) MOSFET cascode amplifying stage with cascode active load; (b) small-signal equivalent circuit