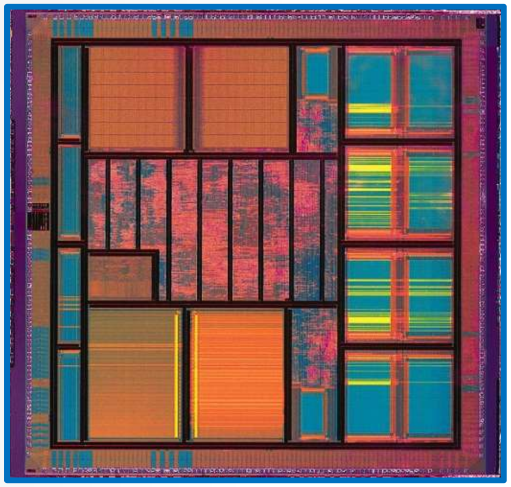
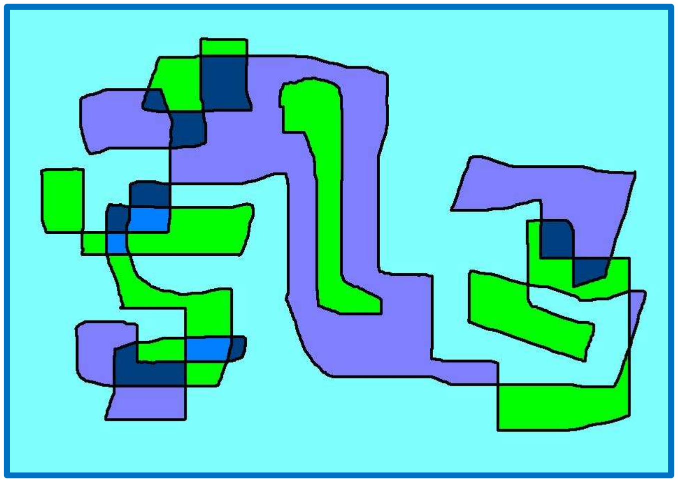
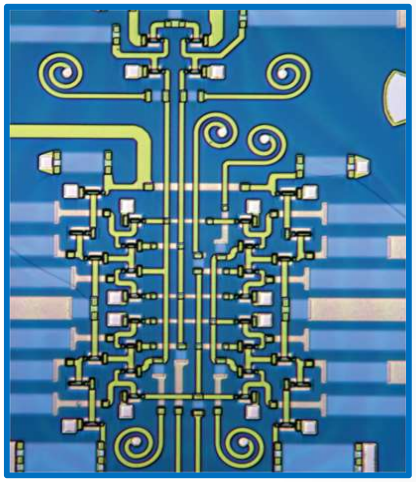
  

ENE/EIE 211 : Electronic Devices and Circuit Design II

Lecture 4: Current Mirrors and Single-stage IC Amps

Outline

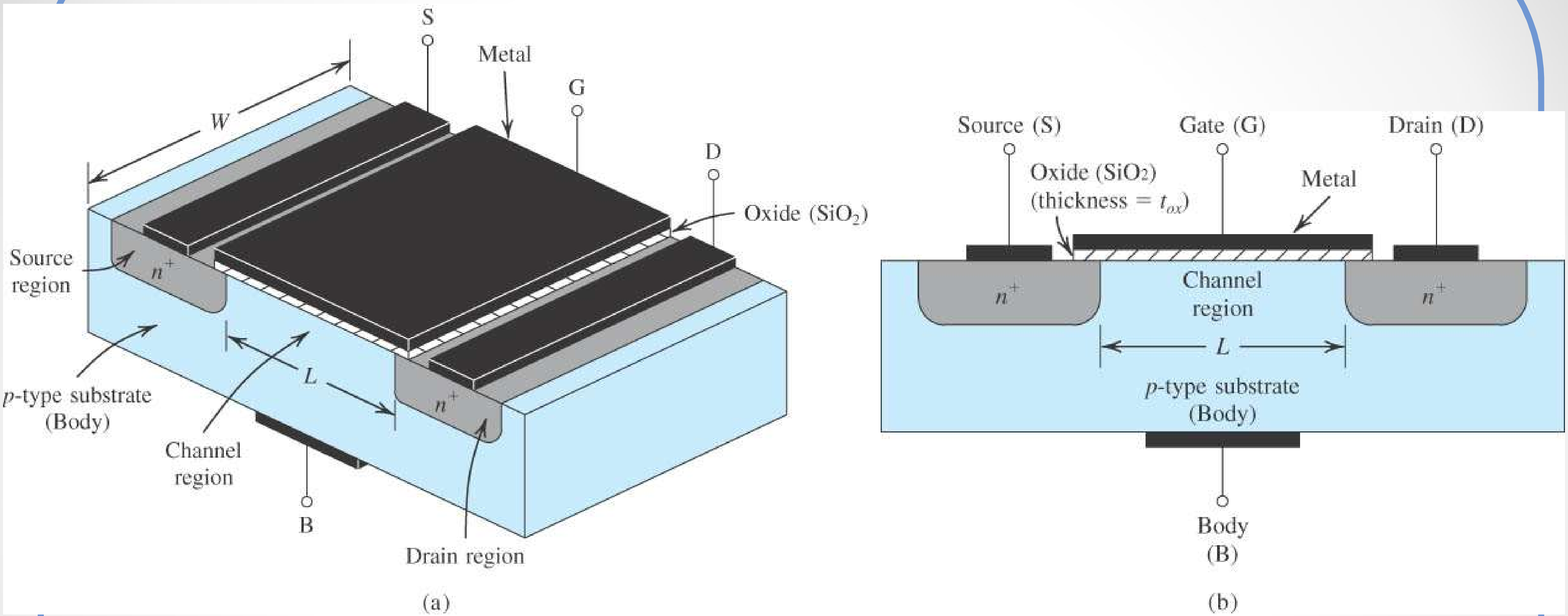
- Current Sources/ Current Mirrors
  - MOSFETs
  - BJTs
  - Wilson Current Mirror
  - Widlar Current Source
- High Frequency Response: Method for estimating the corner frequency (ω_{3dB})
- Single-staged Amplifier
  - Common Emitter Amplifier
  - Common Base Amplifier
- Cascode Amplifier
  - BJTs
  - BiCMOS

IC Biasing- Current Sources, Current Mirrors and Current-Steering Ckts

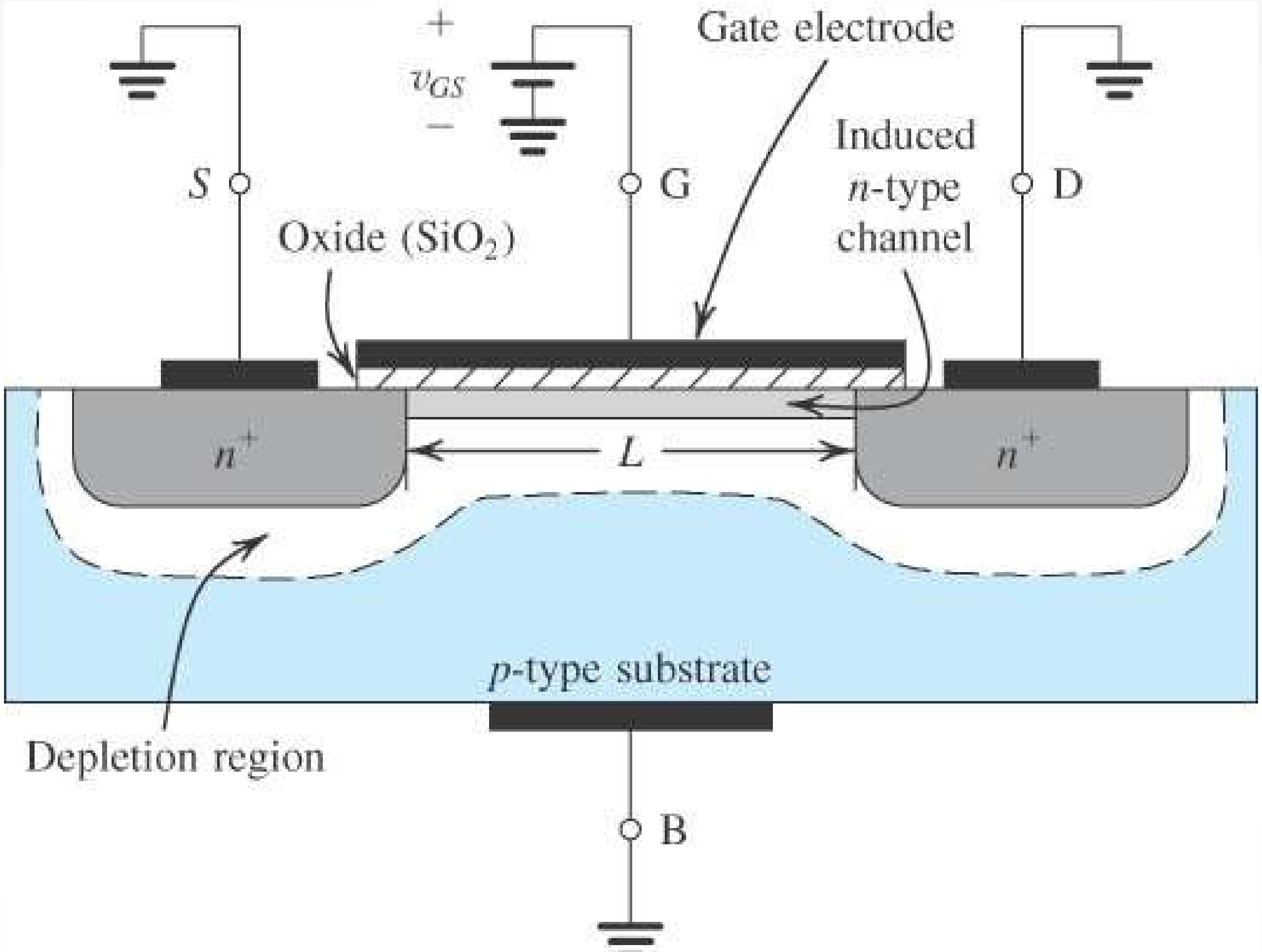
- Biasing in IC design is based on the use of constant- current sources.
- A constant dc I (called a reference I) is generated at one location and is then replicated at various other locations for biasing the various stages through a process known as current steering.
- This approach has the advantage that the effort expended on generating a predictable and stable reference current.
- Furthermore, the bias currents of the various stages track each other in case of changes in power-supply voltage or in temperature.


IMAX
THEATRE
Best Family Entertainment Value
Open Year Round
BUFFALO CROSSING
RV PARK
McDonald's

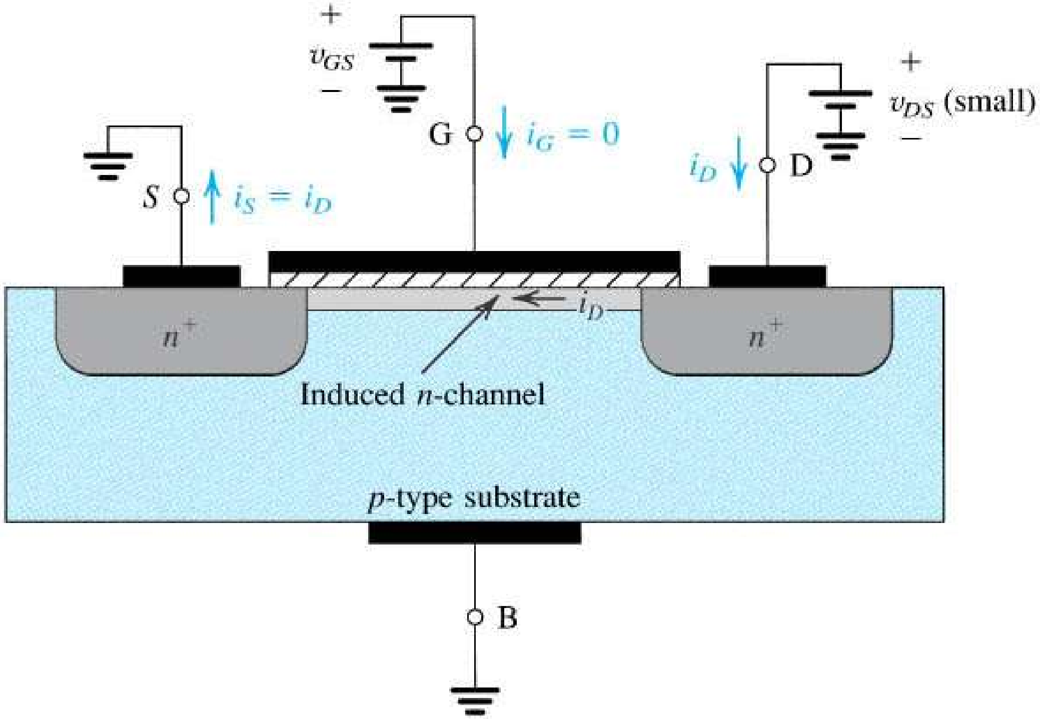
MOSFET Transistors: The Reviews

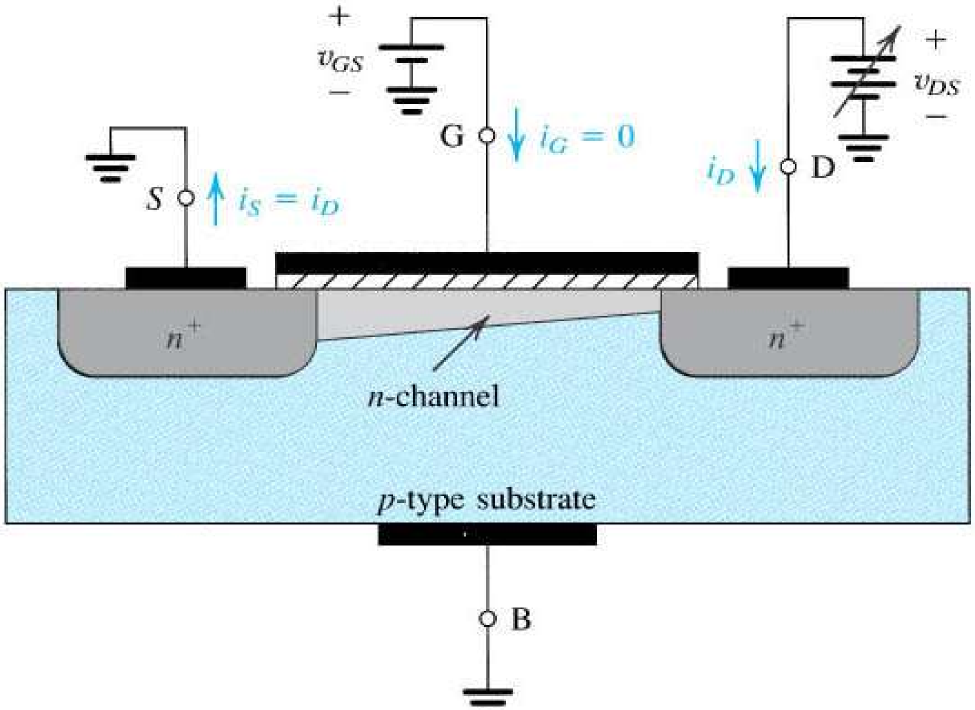


Physical structure of the enhancement-type NMOS transistor: **(a)** perspective view; **(b)** cross-section. Typically L = 0.1 to 3 mm, W = 0.2 to 100 mm, and the thickness of the oxide layer (t_{ox}) is in the range of 2 to 50 nm.



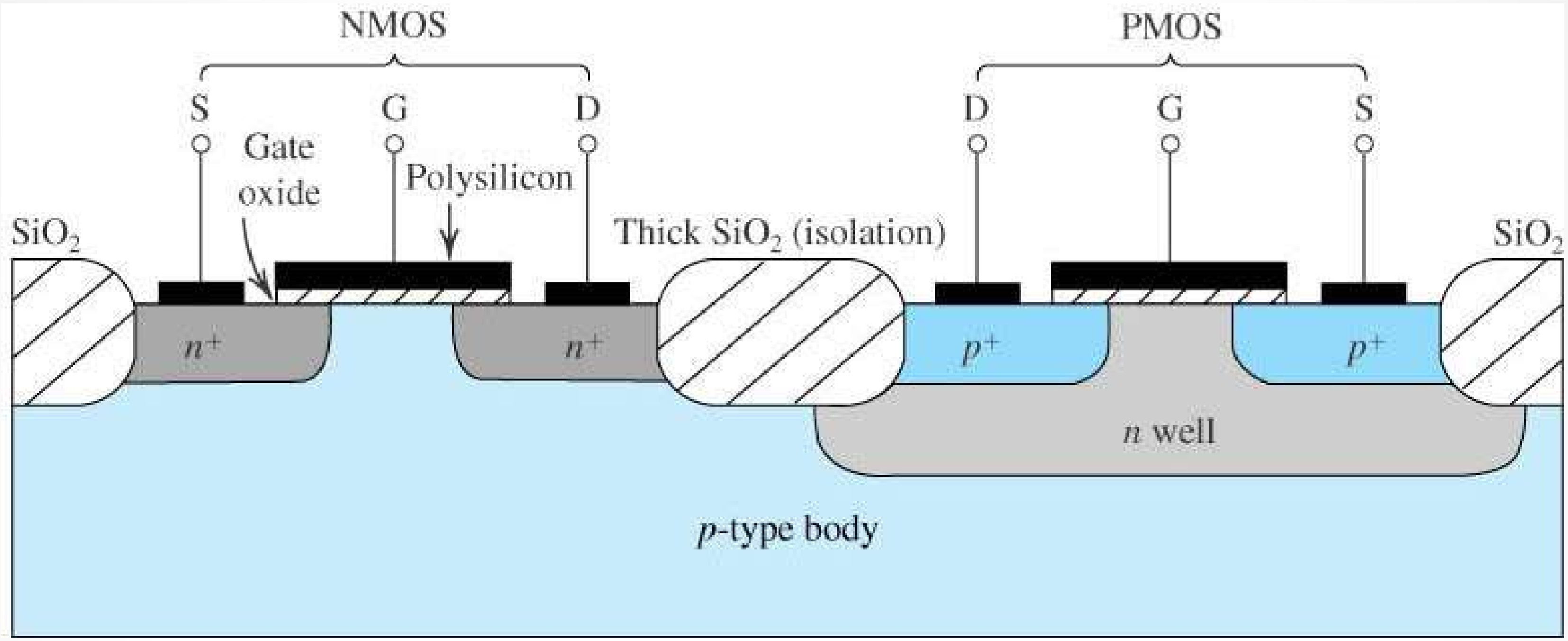
The enhancement-type NMOS transistor with a positive voltage applied to the gate. An *n* channel is induced at the top of the substrate beneath the gate.





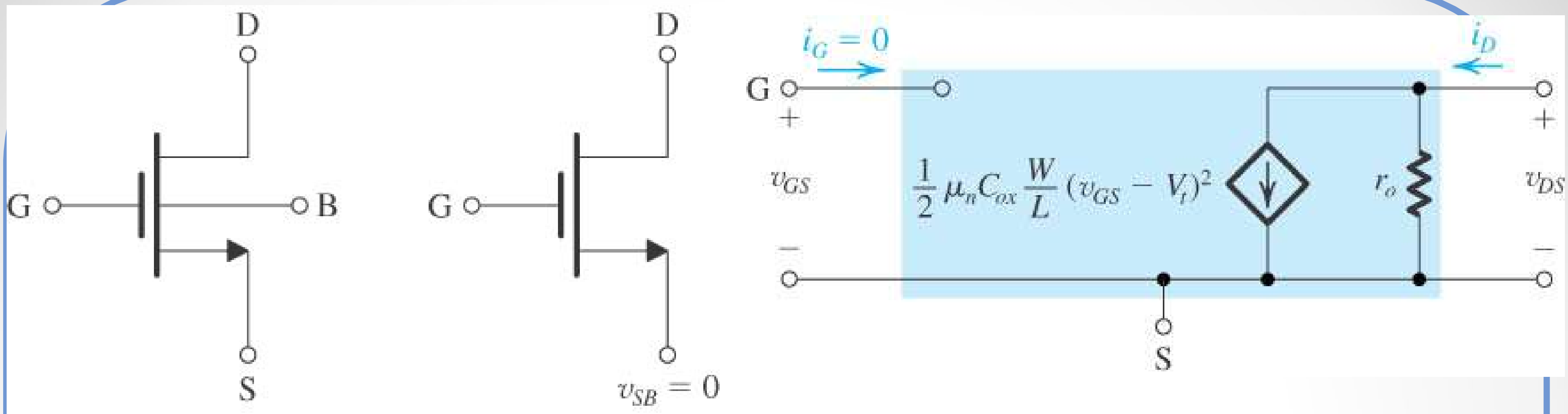
LEFT: An NMOS transistor with $v_{GS} > V_t$ and with a small v_{DS} applied. The device acts as a resistance whose value is determined by v_{GS}. Specifically, the channel conductance is proportional to $v_{GS} - V_t$, and thus i_D is proportional to $(v_{GS} - V_t)\, v_{DS}$. Note that the depletion region is not shown (for simplicity).

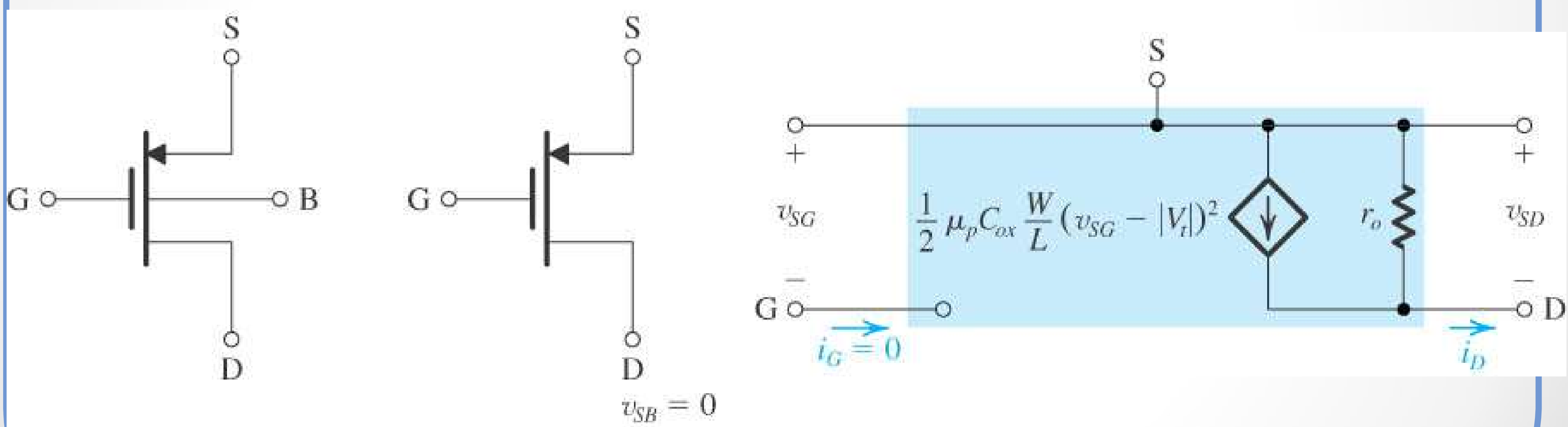
Right: Operation of the enhancement NMOS transistor as v_{DS} is increased. The induced channel acquires a tapered shape, and its resistance increases as v_{DS} is increased. Here, v_{GS} is kept constant at a value $> V_t$.



Cross-section of a CMOS integrated circuit. Note that the PMOS transistor is formed in a separate *n*-type region, known as an *n* well. Another arrangement is also possible in which an *n*-type body is used and the *n* device is formed in a *p* well.

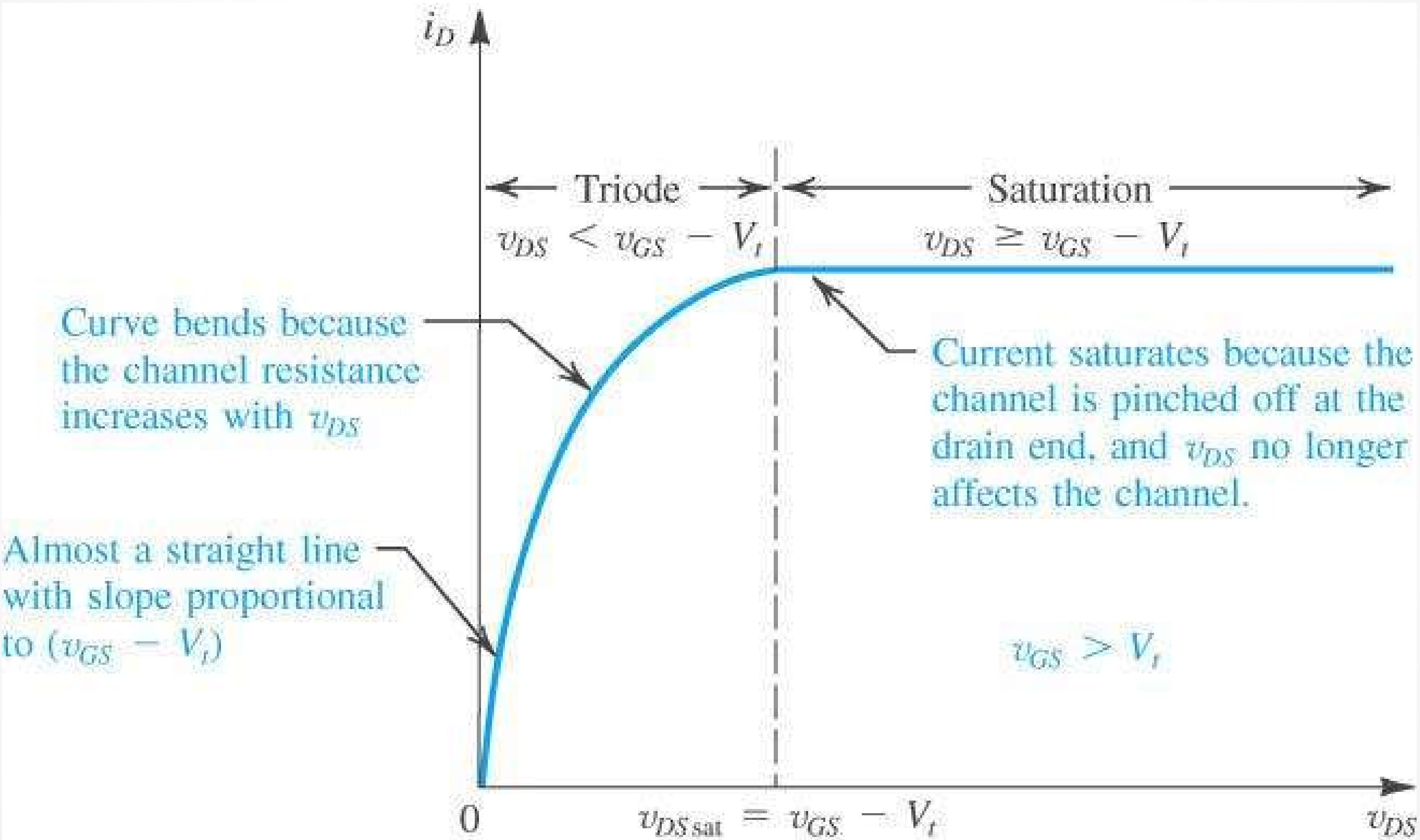
Not shown are the connections made to the *p*-type body and to the *n* well; the latter functions as the body terminal for the *p*-channel device.

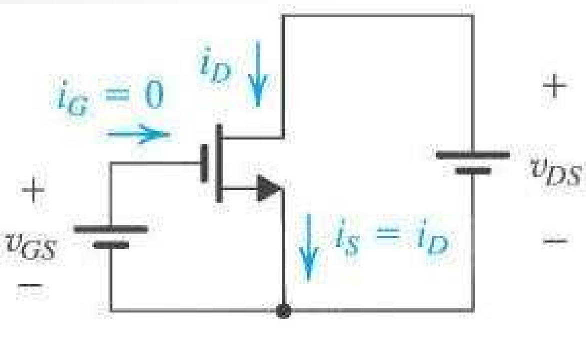




TOP: NMOS
BOTTOM: PMOS

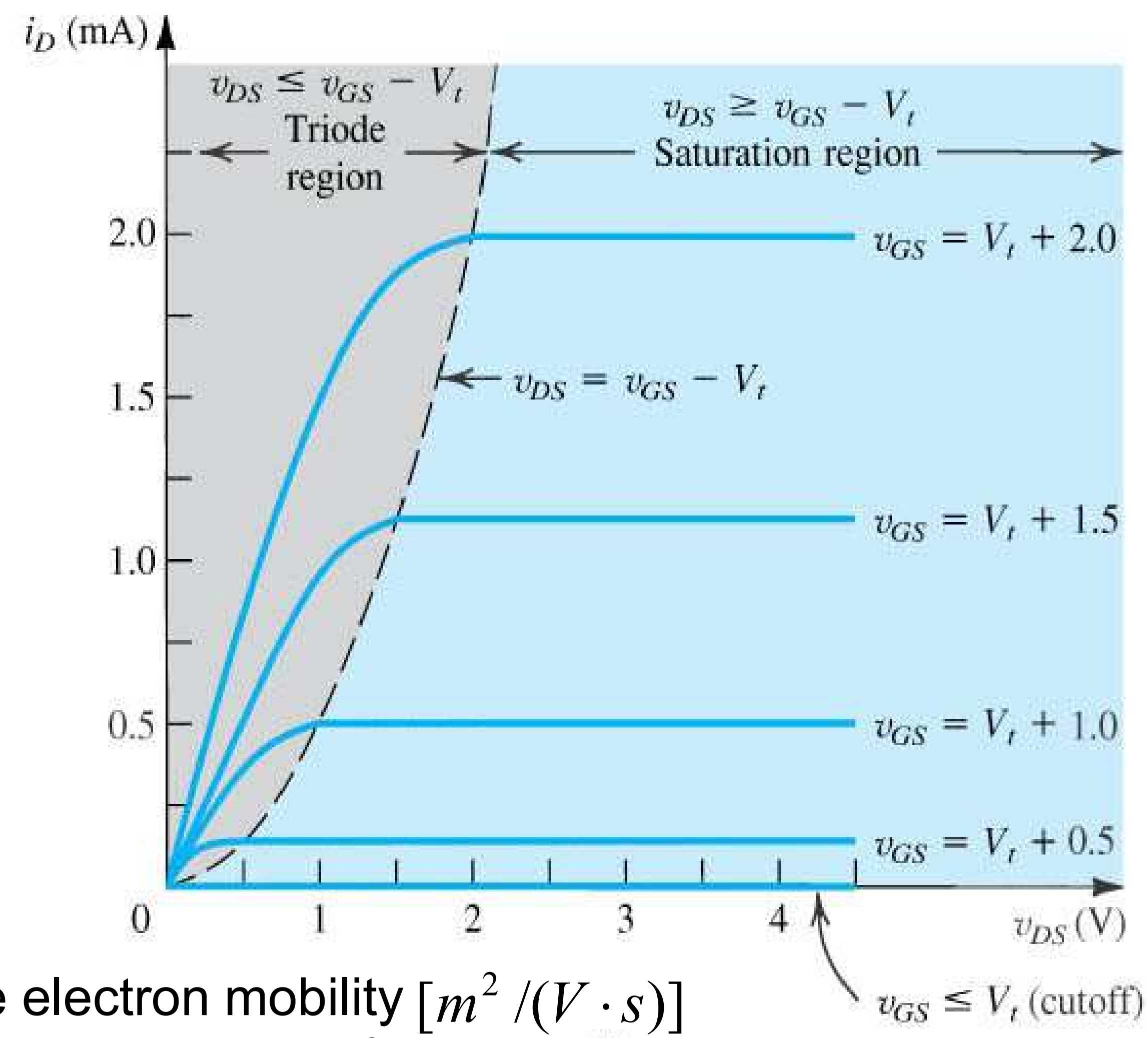
The Basic MOSFET Current Source





(a)

$$I_D = \frac{1}{2} k_n^{'} \left(\frac{W}{L} \right) (V_{GS} - V_{tn})^2$$

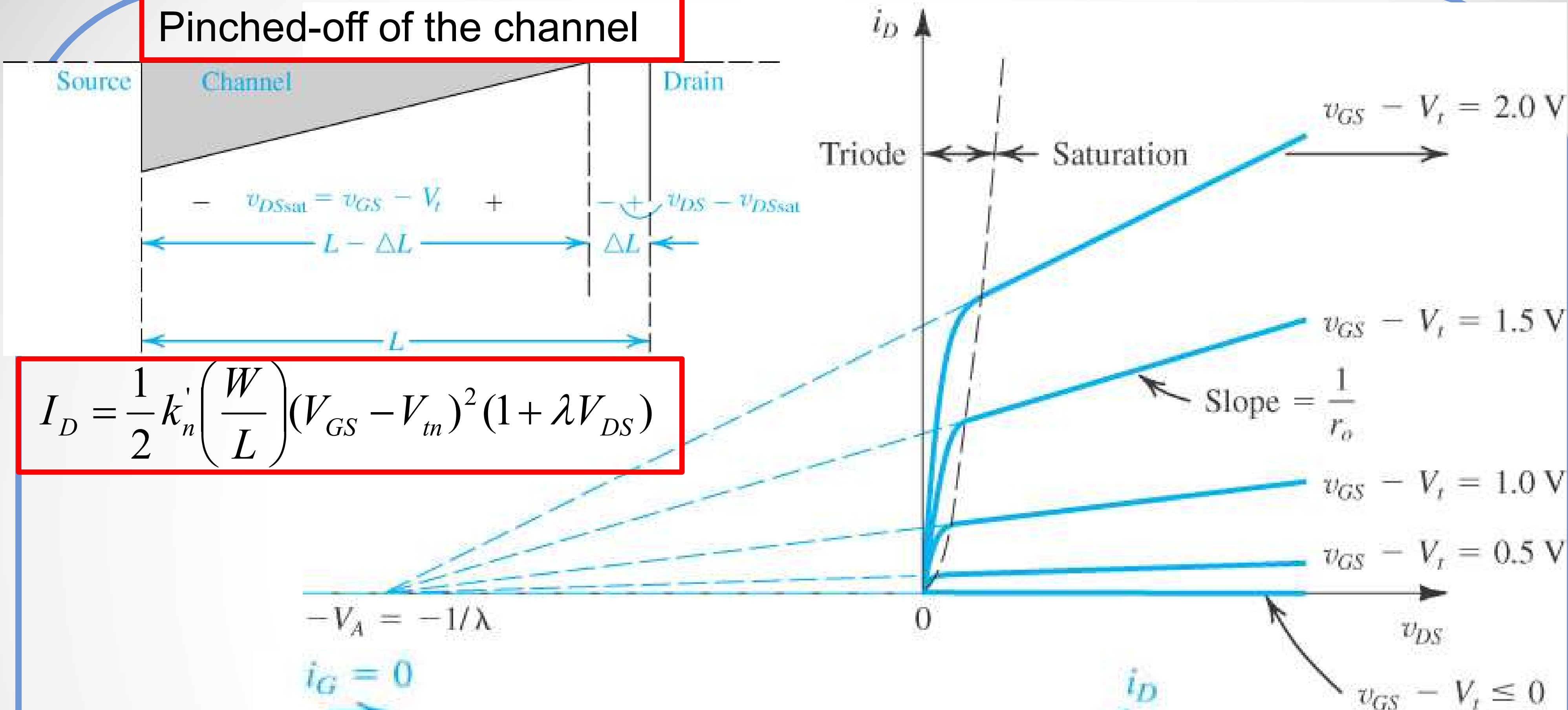


(b)

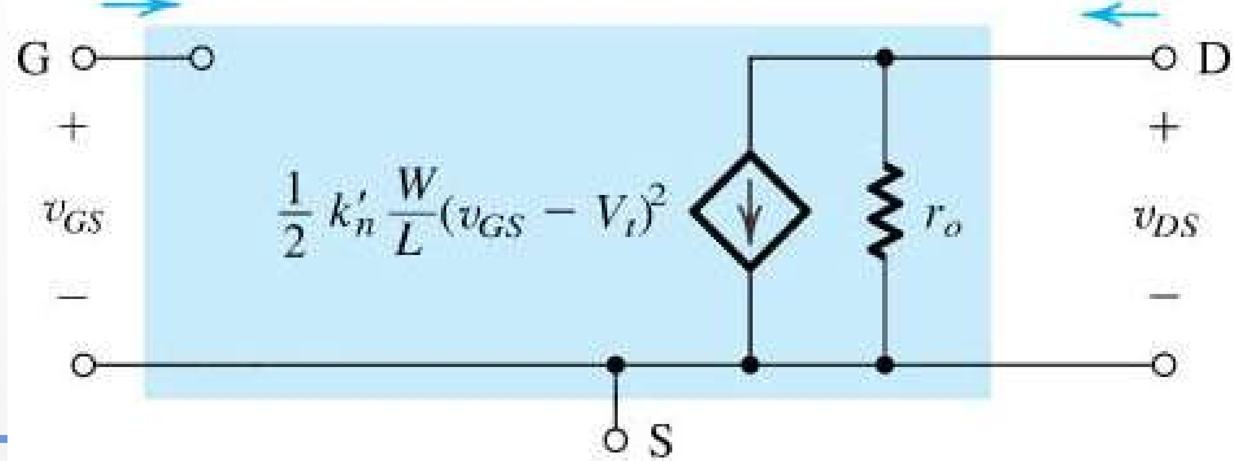
where $k_n = \mu_n C_{ox}$, with μ_n is the electron mobility $[m^2/(V \cdot s)]$ and C_{ox} is the oxide capacitance = ε_{ox}/t_{ox} [F/m^2].

(W/L) is called the aspect ratio. The current obeys the squared law theory.

Pinched-off of the channel

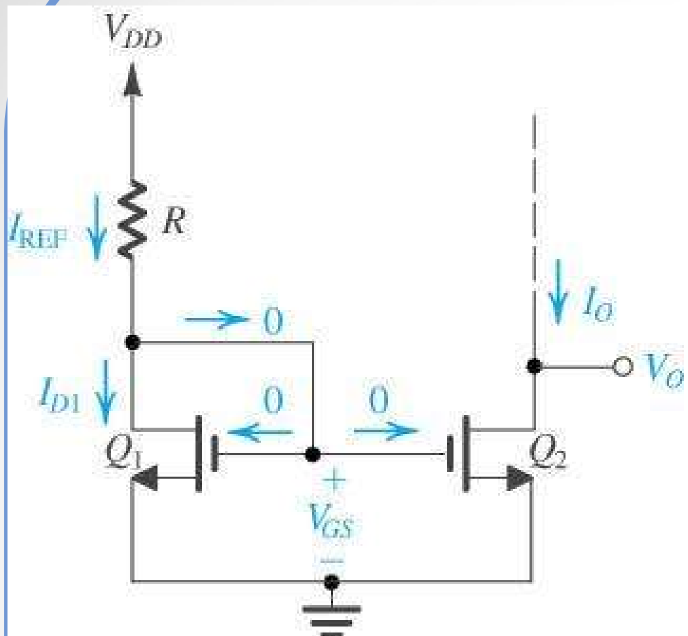


$$I_D = \frac{1}{2} k_n' \left(\frac{W}{L} \right) (V_{GS} - V_{tn})^2 (1 + \lambda V_{DS})$$



$$r_o = \frac{V_A}{I_D}$$

MOSFET current source



The heart of the ckt is transistor Q1, the drain of which is shorted to its gate, thereby forcing it to operate in the saturation mode with

$$I_{D1} = \frac{1}{2}k_n^{'}\left(\frac{W}{L}\right)_1 (V_{GS} - V_{tn})^2$$

Since the gate currents are zero, $\boxed{I_{D1} = I_{REF} = \frac{V_{DD} - V_{GS}}{R}}$

Transistor Q2 has the same V_{GS} as Q1, so

$$I_O = I_{D2} = \frac{1}{2}k_n^{'}\left(\frac{W}{L}\right)_2 (V_{GS} - V_{tn})^2$$

Take ratio of I_{REF} and I_O, ignore the channel-length modulation, we'll get

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

In the special case of identical transistors, $I_O = I_{REF}$, and the ckt simply replicates or mirrors the reference current in the output terminal. This has given the ckt composed of Q1 and Q2 the name current mirror, a name that is used irrespective of the ratio of device dimension.

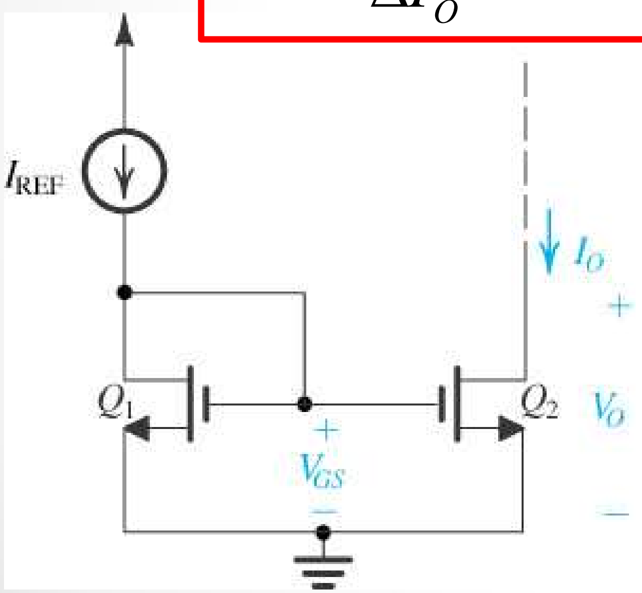
Effect of V_O on I_O

For the current source to operate well, Q2 must be in saturation: $V_O \geq V_{GS} - V_t$

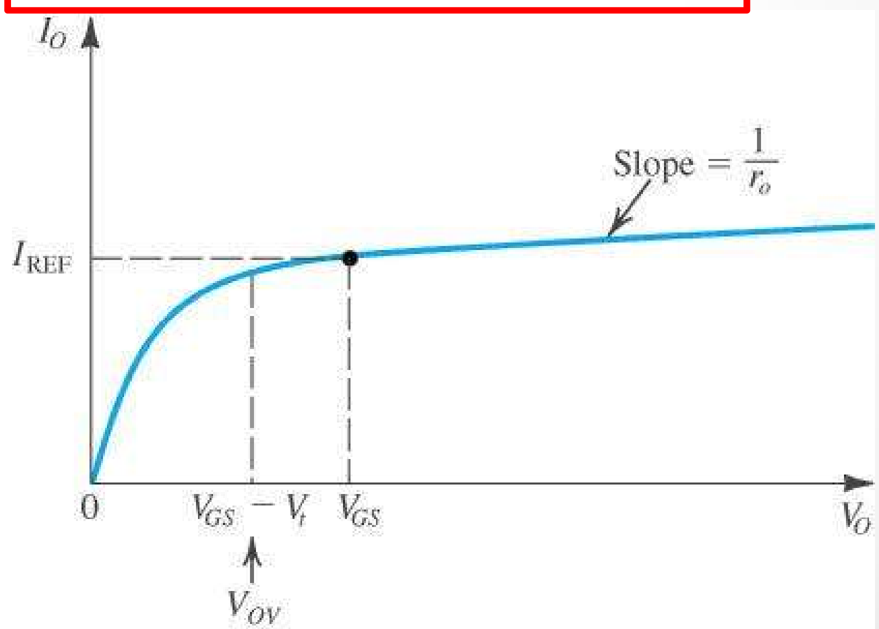
Or, equivalently, in terms of the "over drive" voltage V_{OV} : $V_O \geq V_{OV}$

The current source will operate properly with an output voltage V_O as low as V_{OV}, which is a few tenths of a volt. The current source and the current mirror have a finite output resistance R_O,

$$\boxed{R_O = \frac{\Delta V_O}{\Delta I_O} = r_{o2} = \frac{V_{A2}}{I_O}}$$ and $$\boxed{\frac{I_O}{I_{REF}} = \frac{(W/L)_2}{(W/L)_1}\left(1 + \frac{V_O - V_{GS}}{V_{A2}}\right)}$$

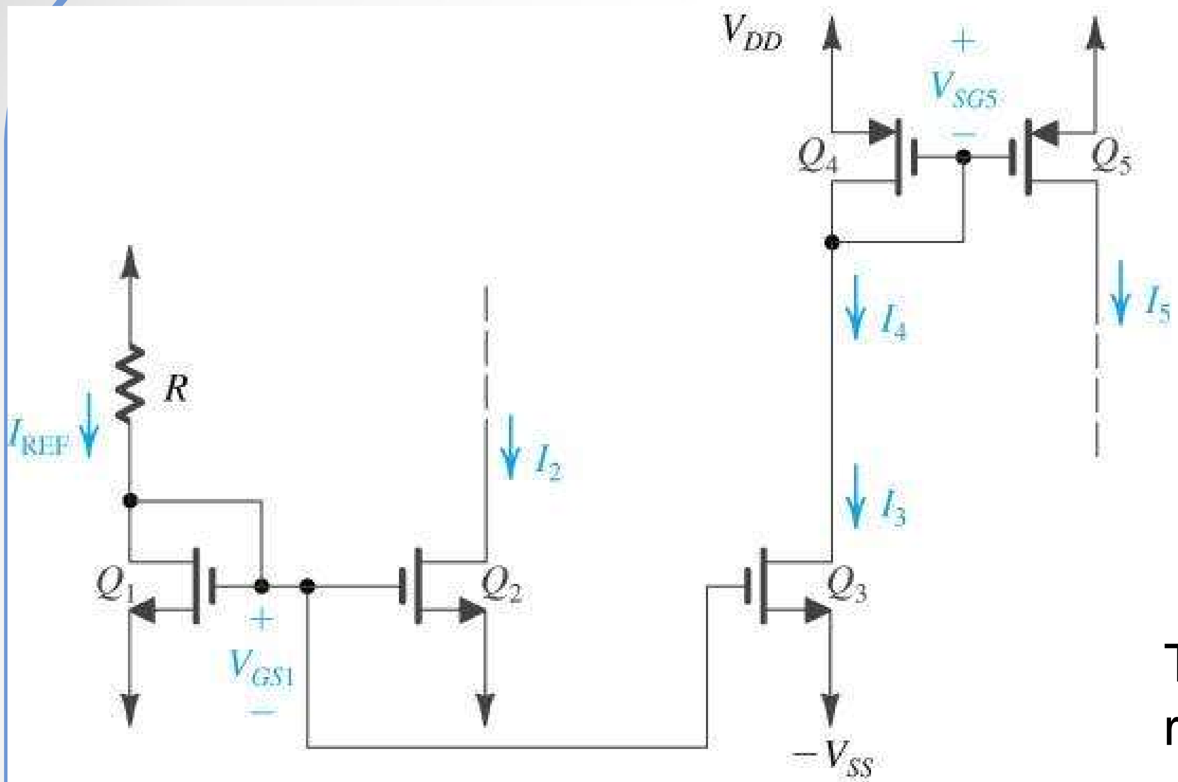


The current mirror



Example: Given V_{DD} = 3 V and using I_{REF} = 100 mA, it is required to design the current source ckt to obtain an output whose nominal value is 100 mA. Find R if Q_1 and Q_2 are matched and have L = 1 mm, W = 10 mm, Vt = 0.7V, and Kn'= 200mA/V^2. What is the lowest possible value of Vo? Assuming that for this process technology the Early voltage V_A'= 20 V/mm, find the output resistance of the current source.

MOS Steering Ckts



- Q_1 and R determine the reference current I_{REF}. Transistors Q_1, Q_2, Q_3 form a two-output current mirror,

$$\frac{I_2}{I_{REF}} = \frac{(W/L)_2}{(W/L)_1}$$

$$\frac{I_3}{I_{REF}} = \frac{(W/L)_3}{(W/L)_1}$$

To ensure operation in saturation region, $V_{D2}, V_{D3} \geq -V_{SS} + V_{GS1} - V_{tn}$

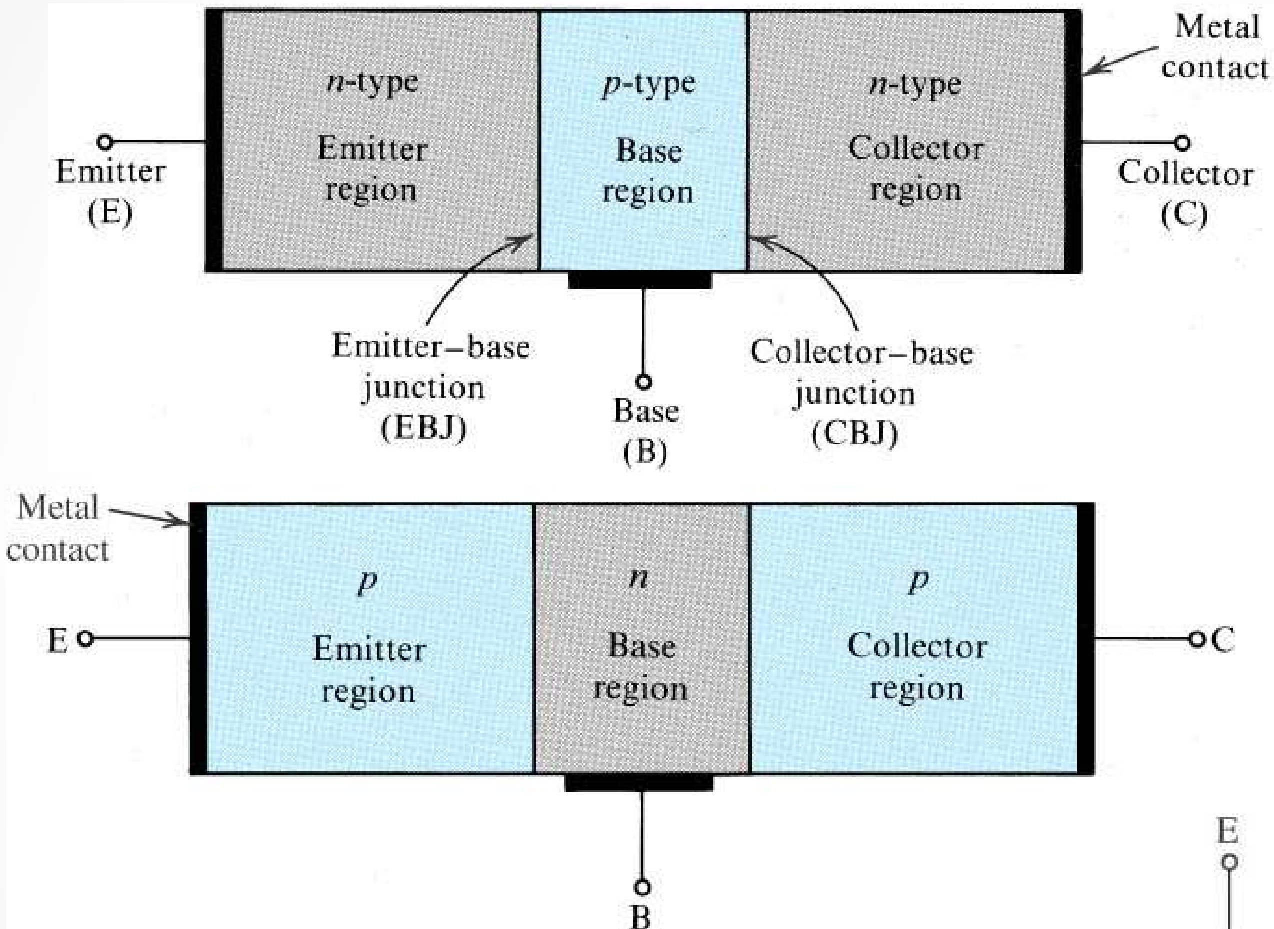
$$V_{D2}, V_{D3} \geq -V_{SS} + V_{OV1}$$

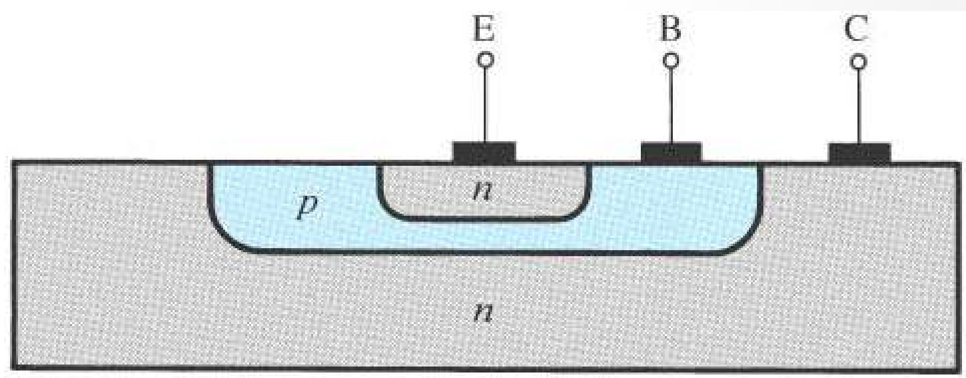
Also, $I_3 = I_4$, and $\frac{I_5}{I_4} = \frac{(W/L)_5}{(W/L)_4}$

To keep Q5 in saturation, its drain voltage should be $V_{D5} \leq V_{DD} - |V_{OV5}|$

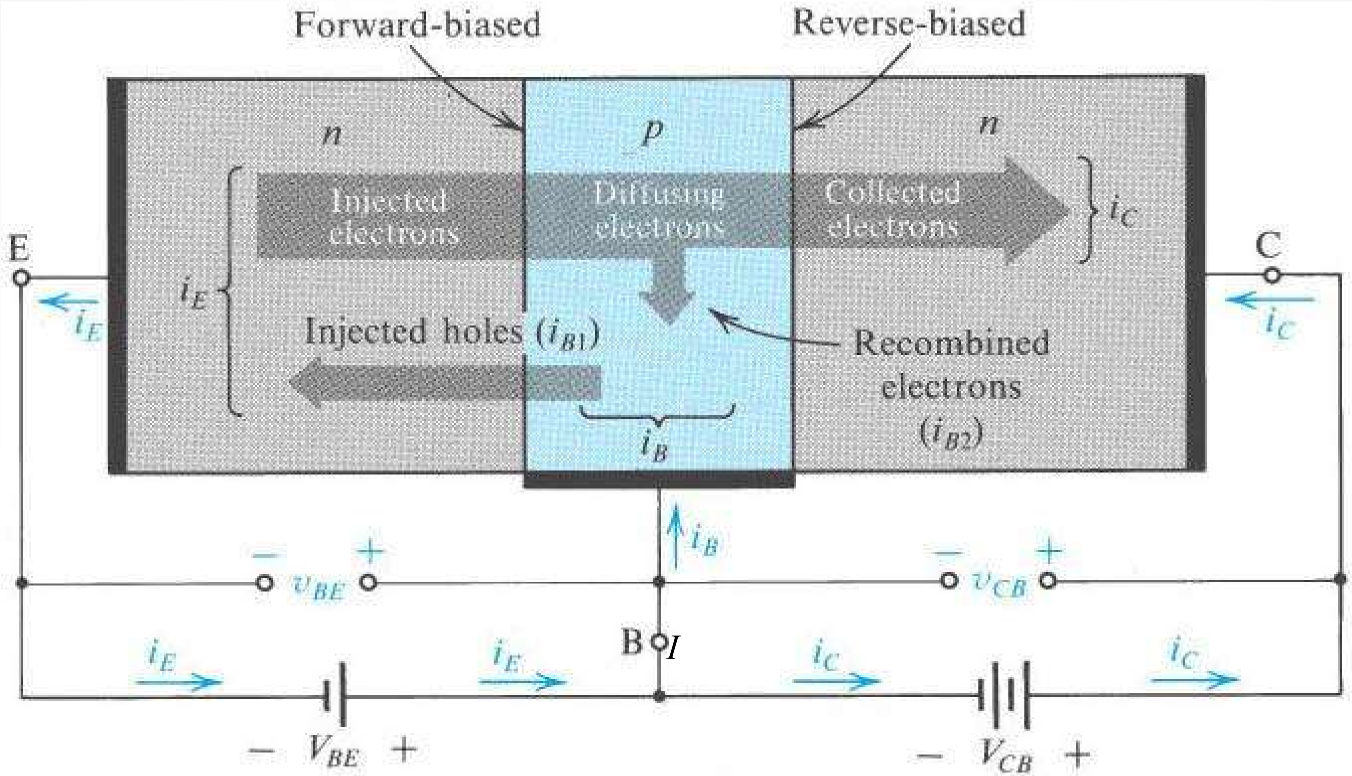
Note: Q_2 is called current sink since it pulls its current from a load (not shown) while Q_5 is called current source since it pushes its current into a load (not shown).

Bipolar Junction Transistors (BJTs): The Reviews





Cross-section of an *npn* BJT



Current flow in an *npn* transistor biased to operate in the active mode. (Reverse current components due to drift of thermally generated minority carriers are not shown.)

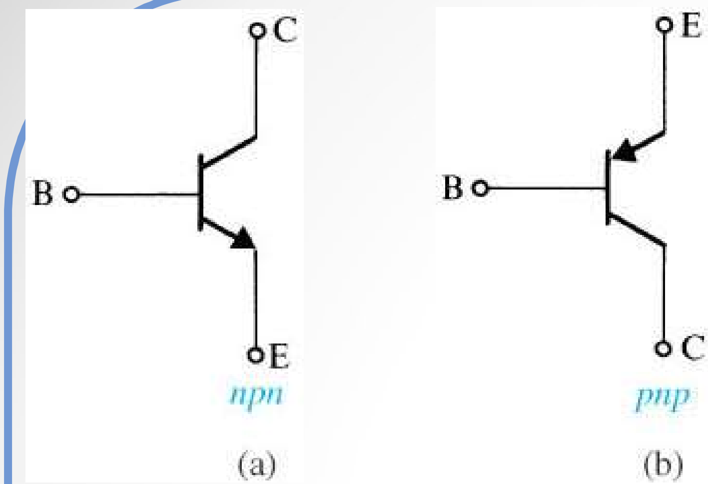
$$\boxed{I_E = I_C + I_B}$$

$$\boxed{I_E = (\beta + 1) I_B}$$

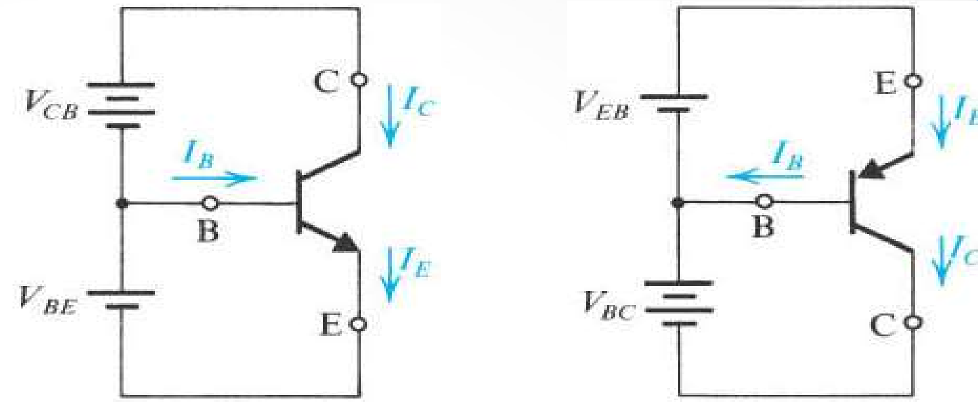
$$\boxed{\beta = \frac{\alpha}{1-\alpha}}$$

$$\boxed{\alpha = \frac{\beta}{\beta + 1} \approx 1}$$

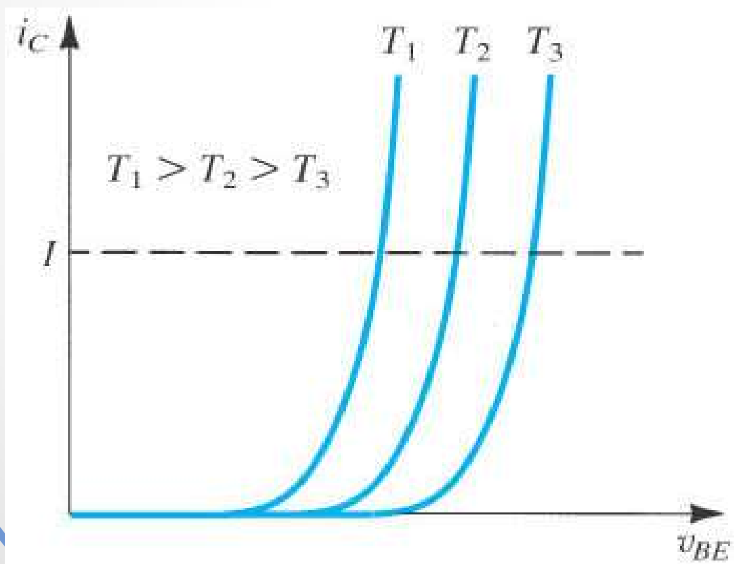
$$\boxed{I_C = \alpha I_E = \beta I_B}$$

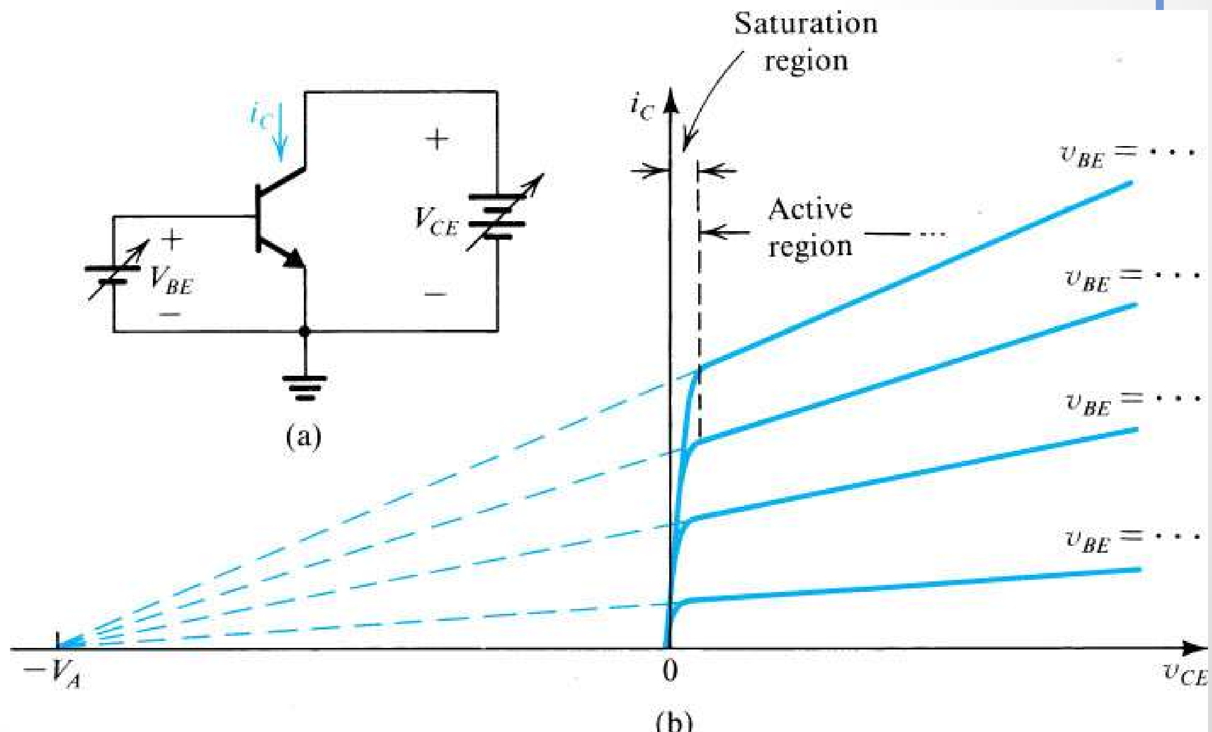


Circuit symbols for BJTs.

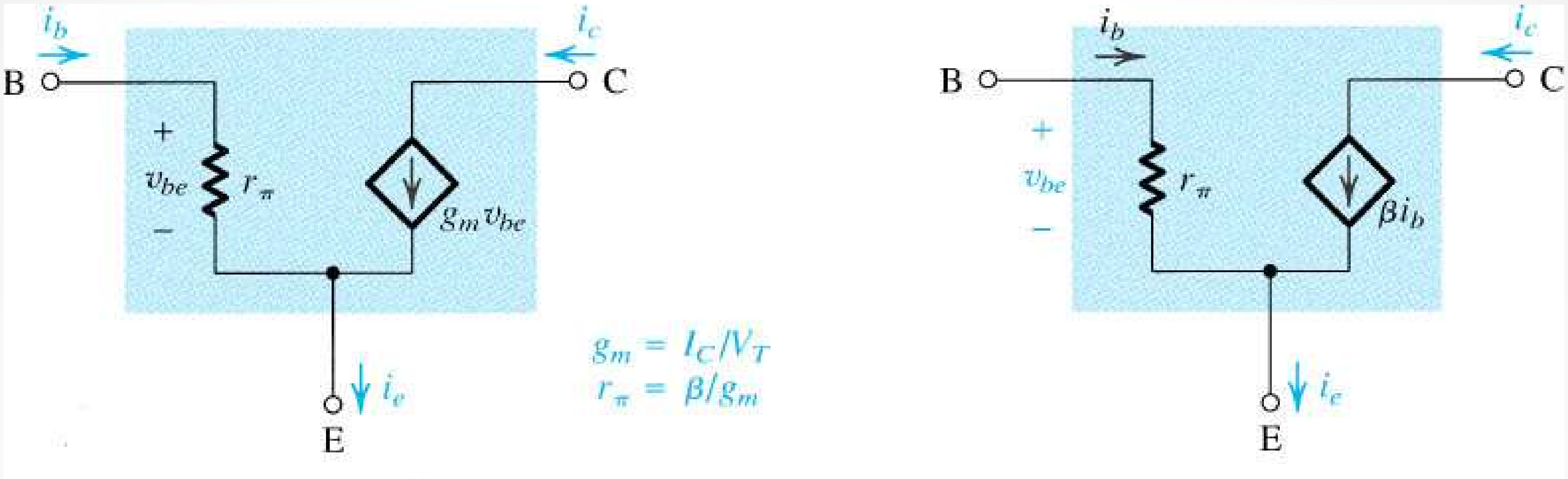


Voltage polarities and current flow in transistors biased in the active mode.

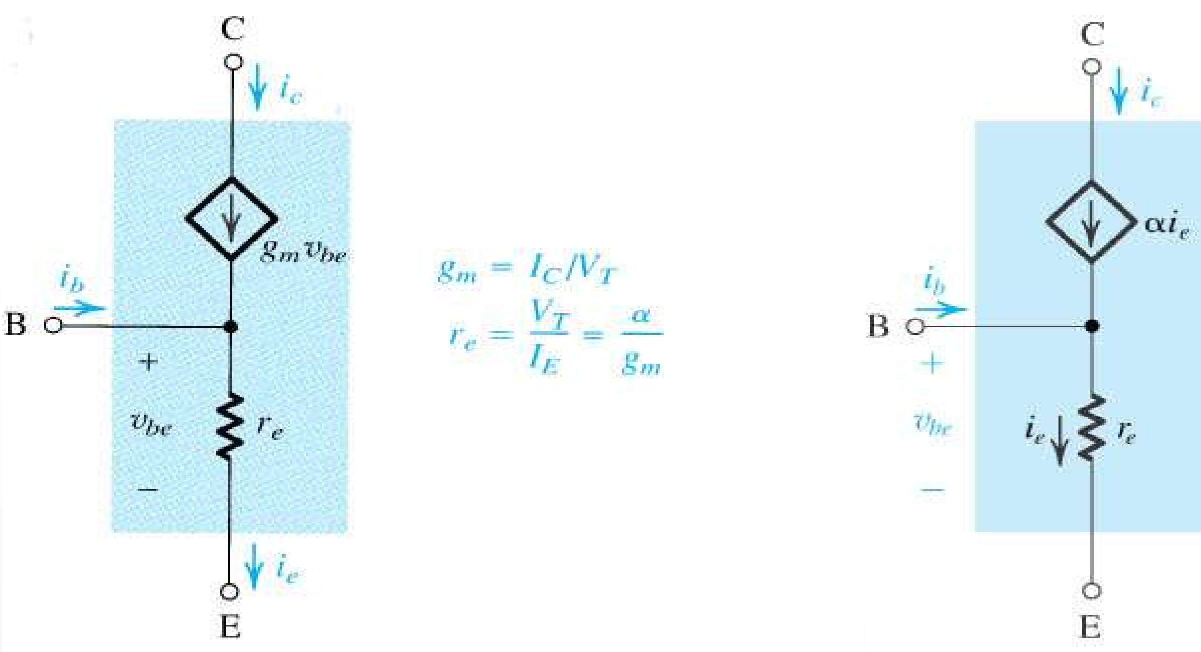




Small-signal Models: the hybrid-π model and the T-model



(a) (b)



(a) (b)

Model Parameters in Terms of DC Bias Currents

$$g_m = \frac{I_C}{V_T} \qquad r_e = \frac{V_T}{I_E} = \alpha\left(\frac{V_T}{I_C}\right) \qquad r_\pi = \frac{V_T}{I_B} = \beta\left(\frac{V_T}{I_C}\right) \qquad r_o = \frac{|V_A|}{I_C}$$

In Terms of g_m

$$r_e = \frac{\alpha}{g_m} \qquad r_\pi = \frac{\beta}{g_m}$$

In Terms of r_e

$$g_m = \frac{\alpha}{r_e} \qquad r_\pi = (\beta+1)r_e \qquad g_m + \frac{1}{r_\pi} = \frac{1}{r_e}$$

Relationships Between α and β

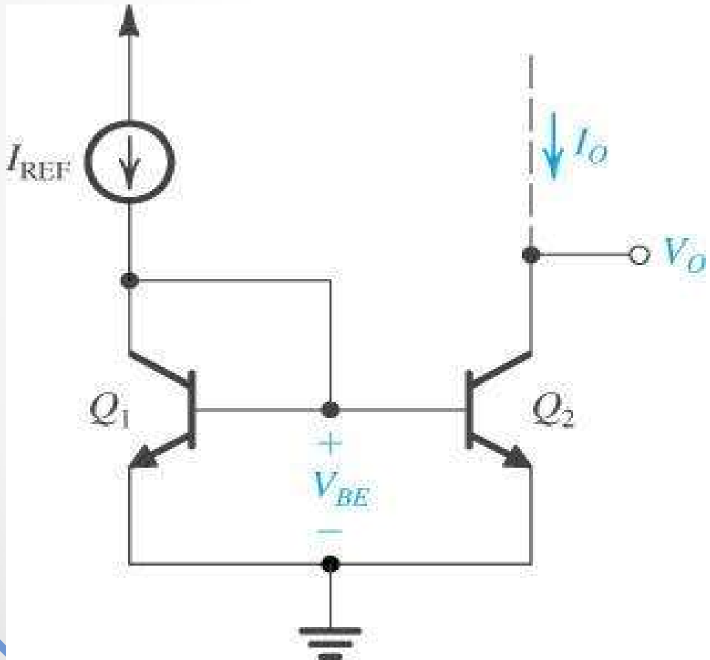
$$\beta = \frac{\alpha}{1-\alpha} \qquad \alpha = \frac{\beta}{\beta+1} \qquad \beta+1 = \frac{1}{1-\alpha}$$

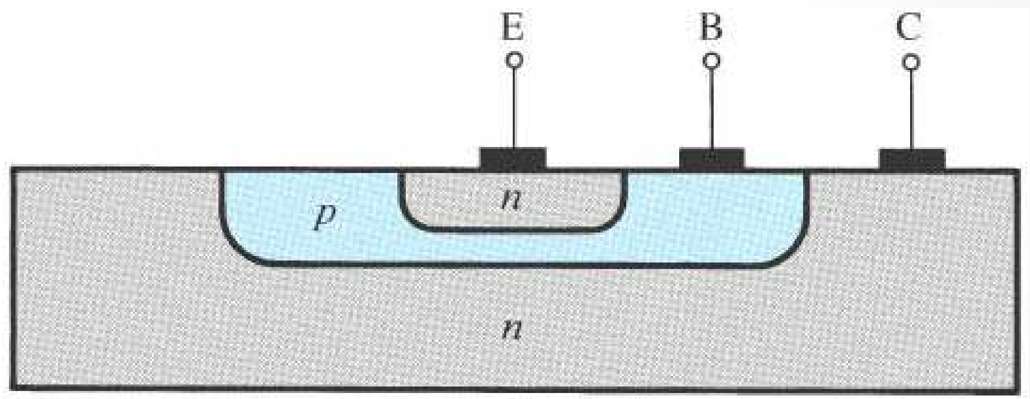
BJT current source and mirror

It works in a manner similar to that of the MOS mirror. However, there are 2 important differences: 1) the nonzero base current of BJT causes an error in the current transfer ratio of the bipolar mirror.

2) the current transfer ratio is determined by the relative areas of the emitter-based junctions (EBJ areas) of Q_1 and Q_2. In general,

$$\boxed{\frac{I_O}{I_{REF}} = \frac{I_{S2}}{I_{S1}} = \frac{\text{Area of EBJ of } Q_2}{\text{Area of EBJ of } Q_1} = m}$$





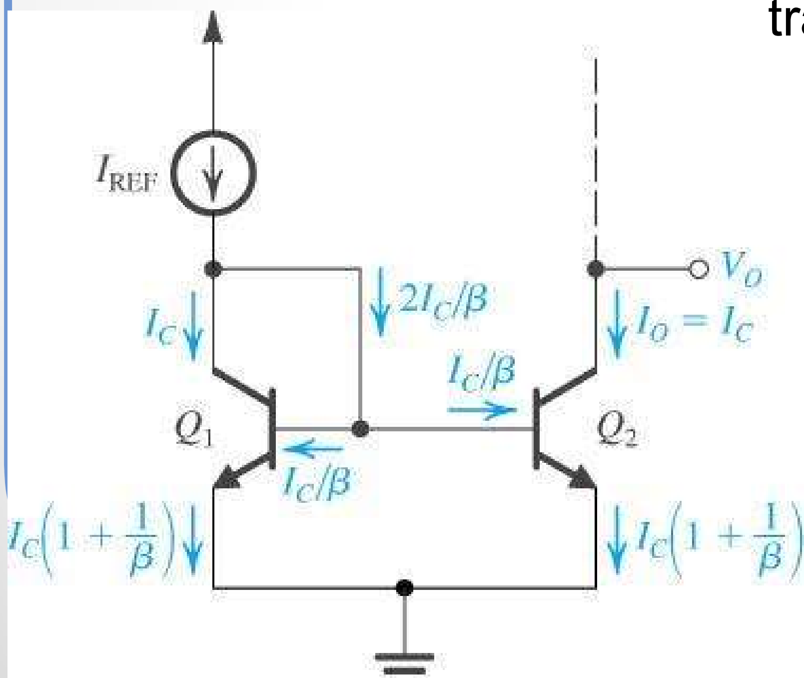
Since Q_1 and Q_2 are matched and have the same V_{BE}, their collector currents will be equal. A node equation at the collector of Q_1 yields

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

Since $I_O = I_C$, the current ratio would give $\dfrac{I_O}{I_{REF}} = \dfrac{I_C}{I_C(1+\frac{2}{\beta})} = \dfrac{1}{1+\frac{2}{\beta}}$

What happens when β = ∞ ?

For a mirror with current transfer ratio m: $\boxed{\dfrac{I_O}{I_{REF}} = \dfrac{m}{1+\frac{m+1}{\beta}}}$

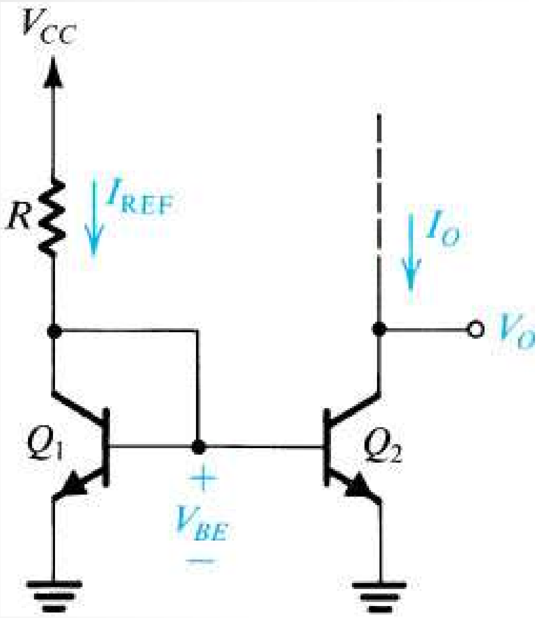


A general relation taken into an account of the Early effect:

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

$$\boxed{R_O = \frac{\Delta V_O}{\Delta I_O} = r_{o2} = \frac{V_{A2}}{I_O}}$$

A Simple Current Source



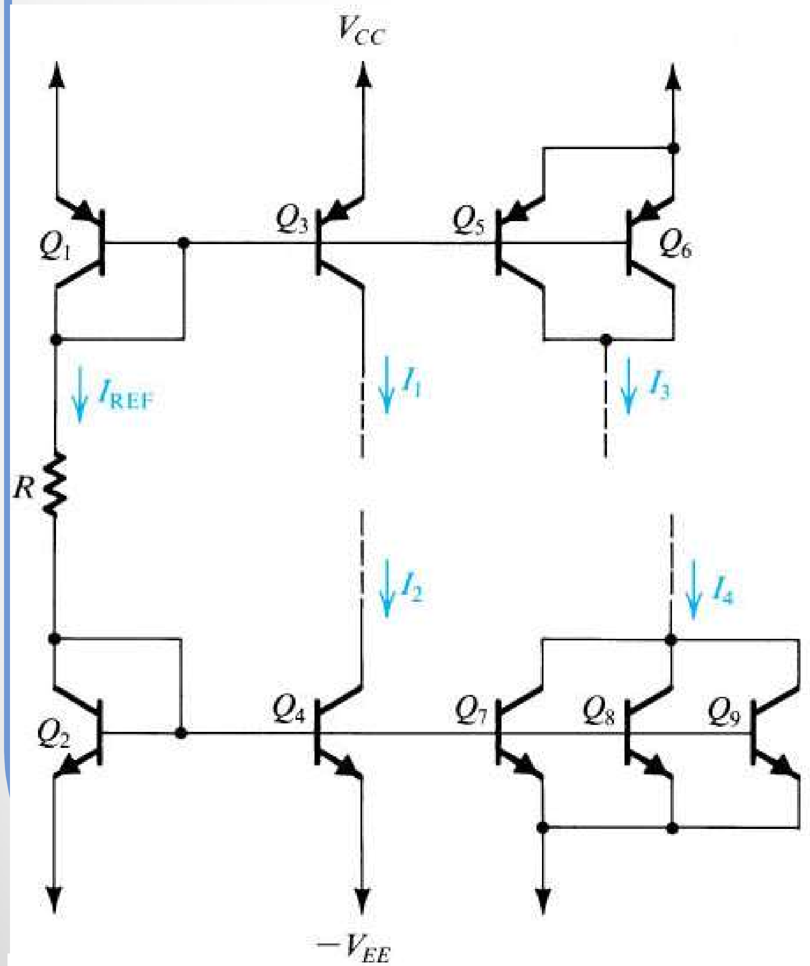
$$\boxed{I_{REF} = \frac{V_{CC} - V_{BE}}{R}}$$

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

$$R_O = \frac{\Delta V_O}{\Delta I_O} = r_{o2} = \frac{V_{A2}}{I_O} \approx \frac{V_{A2}}{I_{REF}}$$

Current Steering

To generate bias current for different amplifier stages in an IC, the current-steering can be used in the bipolar case. The I_{REF} is generated in the branch that consists of the diode-connected transistor Q_1, resistor and the diode-connected transistor Q_2:

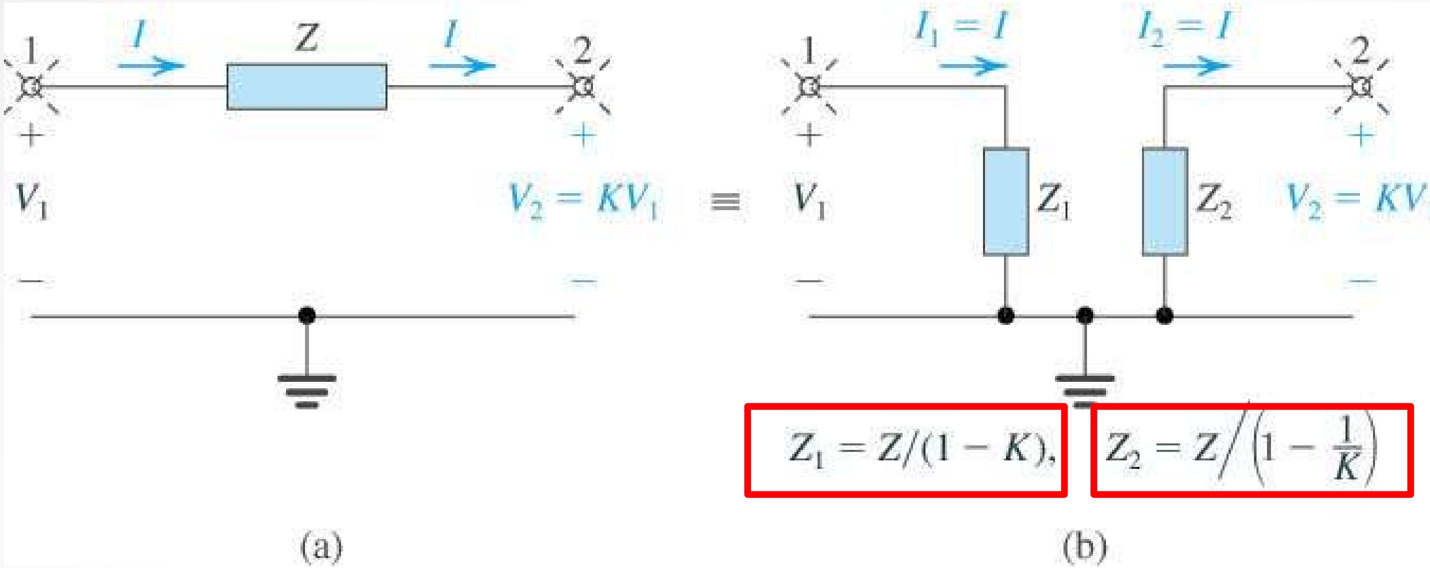


$$\boxed{I_{REF} = \frac{V_{CC} + V_{EE} - V_{EB1} - V_{BE2}}{R}}$$

(assume β = ∞ and ignore the Early effect)

- Q_1 forms current mirror with Q_3, so $I_1 = I_{REF}$.
- The voltage at the collector of Q_3 must not exceed Vcc – 0.3; otherwise Q_3 would enter the saturation region.
- If Q_5 and Q_6 are matched to Q_1, $I_3 = 2 \times I_{REF}$
- Q_4 forms a mirror with Q_2; thus $I_2 = I_{REF}$
- Q_4 sinks its current from parts of the ckt whose voltage should not decrease below $-V_{EE} + 0.3$
- *What is I_4 equal to?*

Miller's Theorem

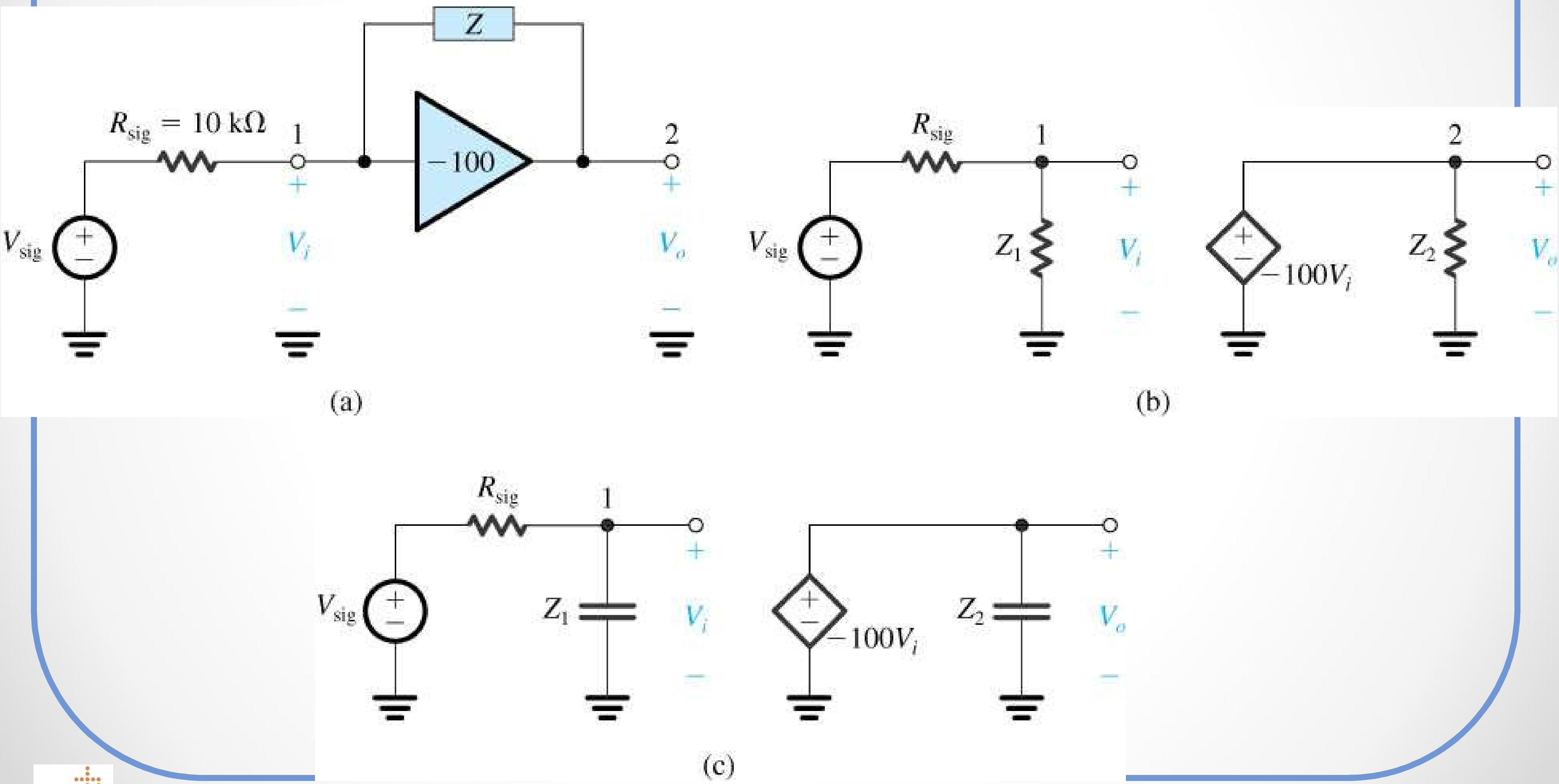


Miller's theorem states that *impedance Z can be replaced by two impedances: Z_1 connected between node 1 and ground and Z_2 connected between node 2 and ground.*

Proof: we must choose Z_1 and Z_2 so they draw an equivalent amount of current to and from their respective terminals.

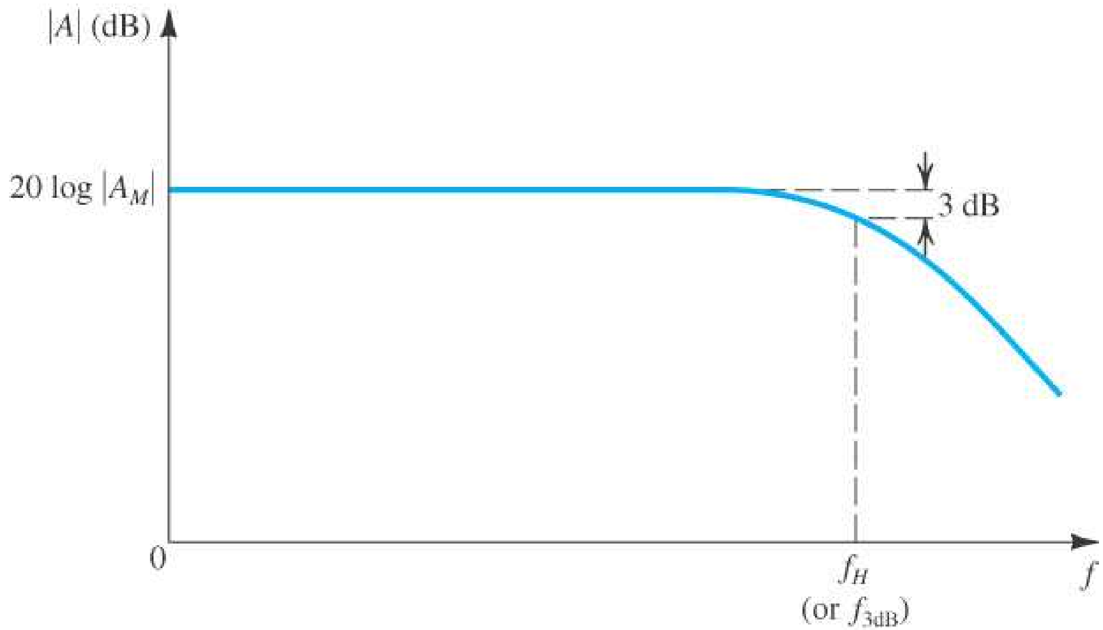
$$I_1 = \frac{V_1}{Z_1} = I = \left(\frac{V_1 - KV_1}{Z}\right) \qquad I_2 = \frac{0 - V_2}{Z_2} = \frac{0 - KV_1}{Z_2} = I = \frac{V_1 - KV_1}{Z}$$

Example: Figure below shows an ideal voltage amplifier having a gain of -100 V/V with an impedance Z connected between its output and input terminals. Find the Miller equivalent ckt when Z is (a) a 1-MΩ resistance and (b) a 1-pF capacitance. In each case, use the equivalent ckt to determine Vo/Vsig.



High Frequency Response

The amplifier ckts intended for fabrication using IC technology do not employ bypass capacitors or coupling capacitors. The frequency response of these directed-coupled or dc amplifiers takes the general form as shown. The gain remains constant at its midband value A_M down to zero frequency (dc).



The amplifier gain taken into account the internal transistor capacitances, can be expressed as a function of the complex-frequency variable s in the general form:

$$\boxed{A(s) = A_M F_H(s)}$$

where A_M is the midband gain whose value is found by analyzing the amp equivalent ckt while neglecting the internal capacitances.

$$F_H(s) = \frac{(1+s/\omega_{Z1})(1+s/\omega_{Z2})\cdots(1+s/\omega_{Zn})}{(1+s/\omega_{P1})(1+s/\omega_{P2})\cdots(1+s/\omega_{Pn})}$$

where ω_{P1}, $\omega_{P2,\ldots}$ ω_{Pn} are positive numbers representing the frequencies of the n real poles and $\omega_{z1,}$ $\omega_{z2}\ldots\omega_{zn}$ are positive, negative or infinite numbers representing the frequencies of the n real transmission zeros.

The amplifier designer is usually interested in the part of the high-frequency band that is close to the midband. This is because the designer needs to estimate or modify the value of the upper 3-dB frequency f_H.

Usually, the zeros are either at infinity of such high frequencies that it is insignificant.

If there's one pole, ω_{P1} that is much lower freq than any of other poles, then this pole will have the greatest effect on the value of the f_H, as it will dominate the high-freq response of the amplifier. In such case the response can be approximated as

$$\boxed{F_H(s) = \frac{1}{1+s/\omega_{P1}}}$$

If the dominant pole does not exist, the 3-dB freq can be determined from the plot of $|F_H(j\omega)|$. Alternatively, an approx. can be derived as follows: Consider the case of a ckt having 2 poles and 2 zeros in the high-freq band; that is,

$$F_H(s) = \frac{(1+s/\omega_{Z1})(1+s/\omega_{Z2})}{(1+s/\omega_{P1})(1+s/\omega_{P2})}$$

Substituting s = jω and taking the squared magnitude gives

$$|F_H(s)|^2 = \frac{(1+\omega^2/\omega_{Z1}^2)(1+\omega^2/\omega_{Z2}^2)}{(1+\omega^2/\omega_{P1}^2)(1+\omega^2/\omega_{P2}^2)}$$

By definition, at $\omega = \omega_H$, $|F_H|^2 = 0.5$, thus

$$\frac{1}{2} = \frac{(1+\omega_H^2/\omega_{Z1}^2)(1+\omega_H^2/\omega_{Z2}^2)}{(1+\omega_H^2/\omega_{P1}^2)(1+\omega_H^2/\omega_{P2}^2)}$$

$$= \frac{1+\omega_H^2\left(\dfrac{1}{\omega_{Z1}^2}+\dfrac{1}{\omega_{Z2}^2}\right)+\omega_H^4/\omega_{Z1}^2\omega_{Z2}^2}{1+\omega_H^2\left(\dfrac{1}{\omega_{P1}^2}+\dfrac{1}{\omega_{P2}^2}\right)+\omega_H^4/\omega_{P1}^2\omega_{P2}^2}$$

$$\omega_H \approx 1 / \sqrt{\frac{1}{\omega_{P1}^2} + \frac{1}{\omega_{P2}^2} - \frac{2}{\omega_{Z1}^2} - \frac{2}{\omega_{Z2}^2}}$$

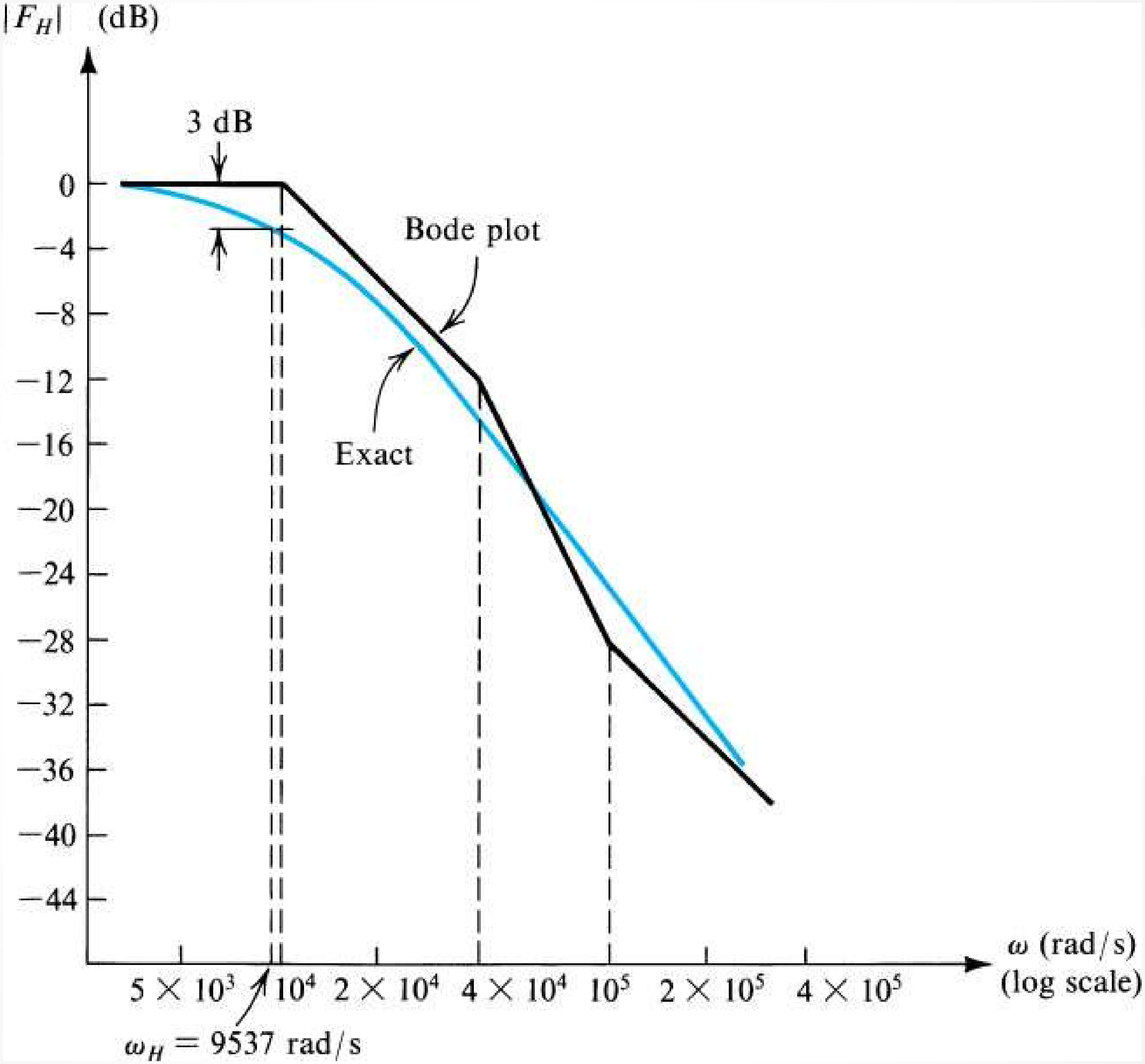
This relationship can be extended to any number of poles and zeros as

$$\boxed{\omega_H \approx 1 / \sqrt{\left(\frac{1}{\omega_{P1}^2} + \frac{1}{\omega_{P2}^2} + \cdots\right) - 2\left(\frac{1}{\omega_{Z1}^2} + \frac{1}{\omega_{Z2}^2} + \cdots\right)}}$$

Example: The high-frequency response of an amplifier is characterized by the transfer function

$$F_H(s) = \frac{1 - s/10^5}{(1 + s/10^4)(1 + s/4\times 10^4)}$$

Determine the 3-dB freq approximately.



$|F_H|$ (dB)
3 dB
0
−4
−8
−12
−16
−20
−24
−28
−32
−36
−40
−44
Bode plot
Exact
ω (rad/s)
(log scale)
5×10^3
10^4
2×10^4
4×10^4
10^5
2×10^5
4×10^5
$\omega_H = 9537$ rad/s

Using Open-Ckt Time Constants for the Approximate Determination of f_H

$$F_H(s) = \frac{1 + a_1 s + a_2 s^2 + \cdots + a_n s^n}{1 + b_1 s + b_2 s^2 + \cdots + b_n s^n} \quad \text{where} \quad b_1 = \frac{1}{\omega_{P1}} + \frac{1}{\omega_{P2}} + \cdots \frac{1}{\omega_{Pn}}$$

It can be shown that the value of b_1 can be obtained by considering the various capacitances in the high-freq equivalent ckt one at a time while reducing all other capacitors to zero (by replacing them with open ckts).

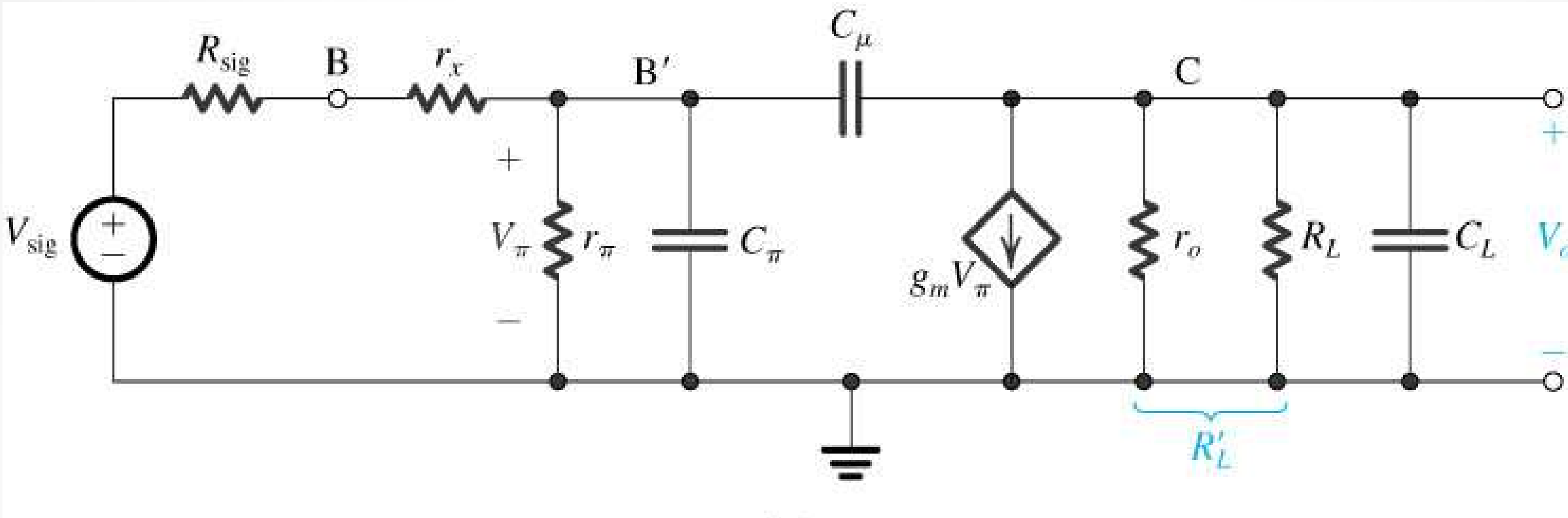
That is to obtain the contribution of capacitance Ci , we reduce all other capacitances to zero, reduce the input signal to zero, and determine the resistance R_{io} seen by C_i. This process is then repeated for all other capacitors in the circuit.

b_1 is computed by summing the individual time constants, called "open-ckt time const"

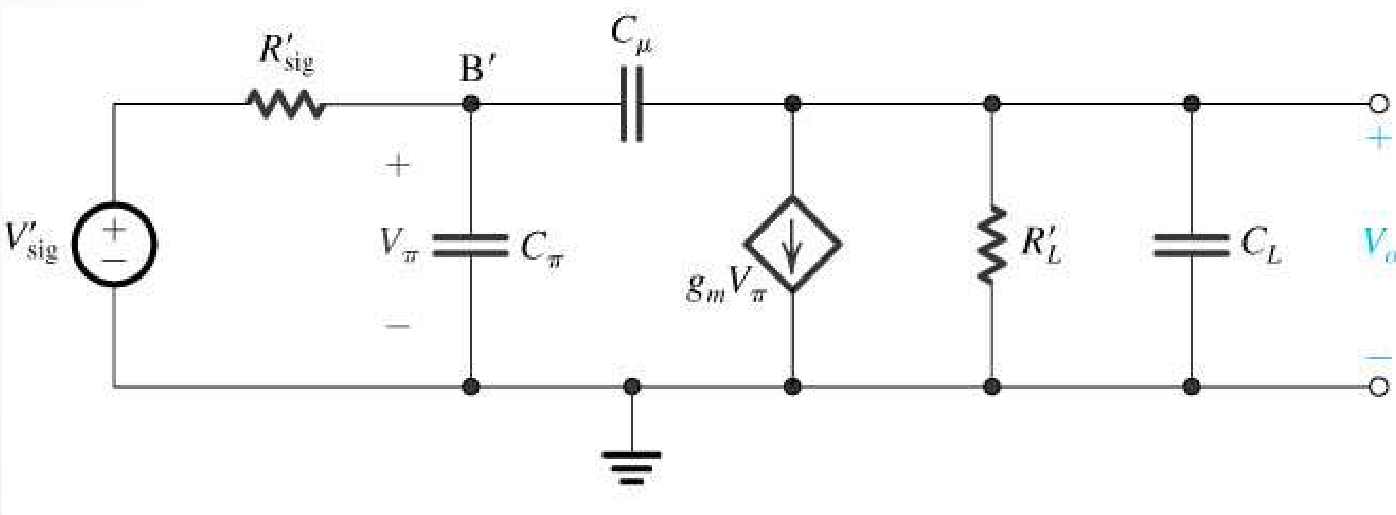
$$\boxed{b_1 = \sum_{i=1}^{n} C_i R_{io} = \tau_H}$$

$$\boxed{\omega_H = \frac{1}{2\pi\tau_H}}$$

Example: Determine the f_H of a CE Amplifier



(a)

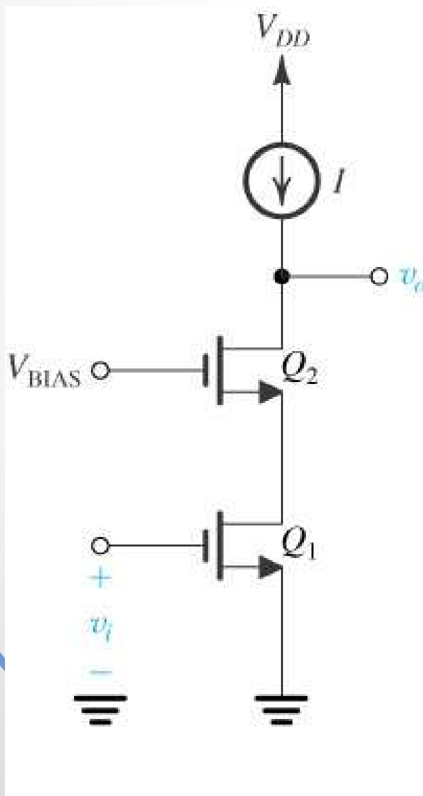


(b)

The Cascode Amplifier

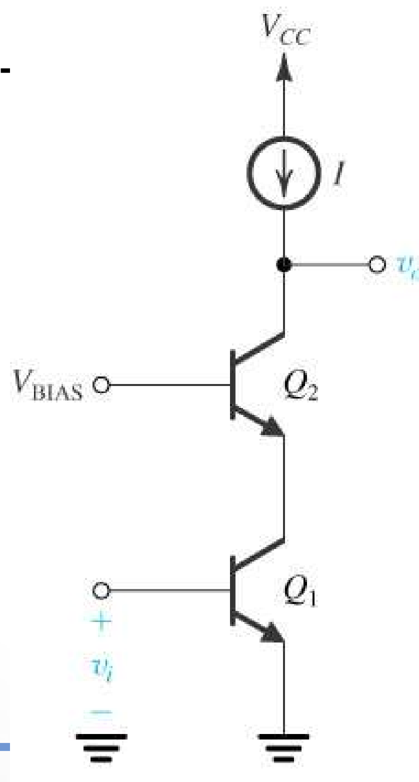
By placing a common-gate (common-base) amplifier stage in cascade with a common-source (common-emitter) amplifier stage, we have a **cascode** configuration.

The basic idea is to combine the **high R_{in}** and **large transconductance** (g_m) of a common-source (common-emitter) amplifier with the **current-buffering** property and **superior high-frequency response** of the common-gate (common-base) circuit.



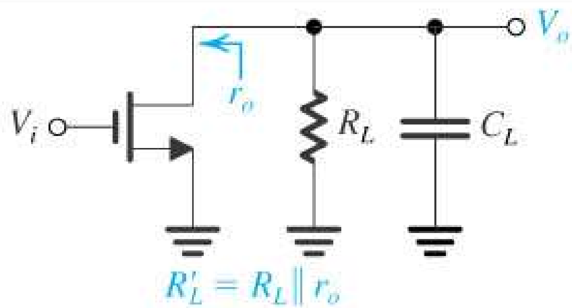
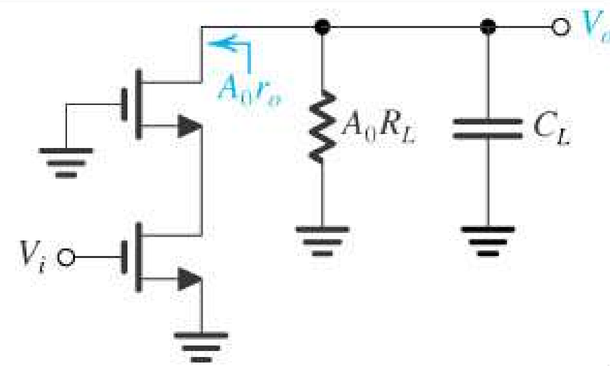
(a)

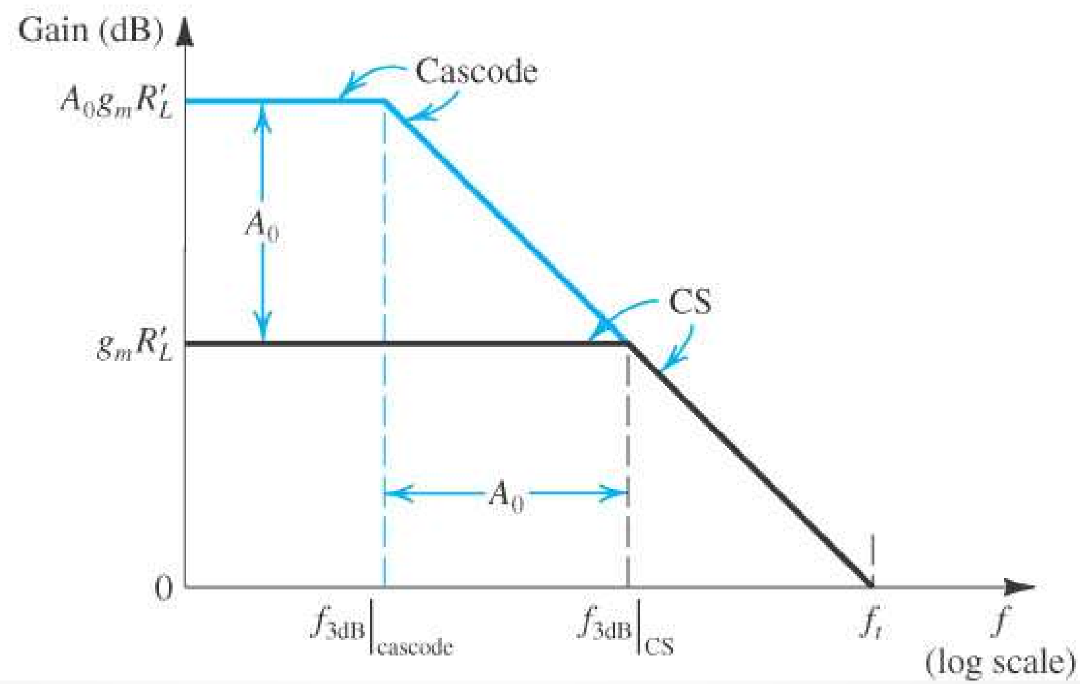
nd BJ



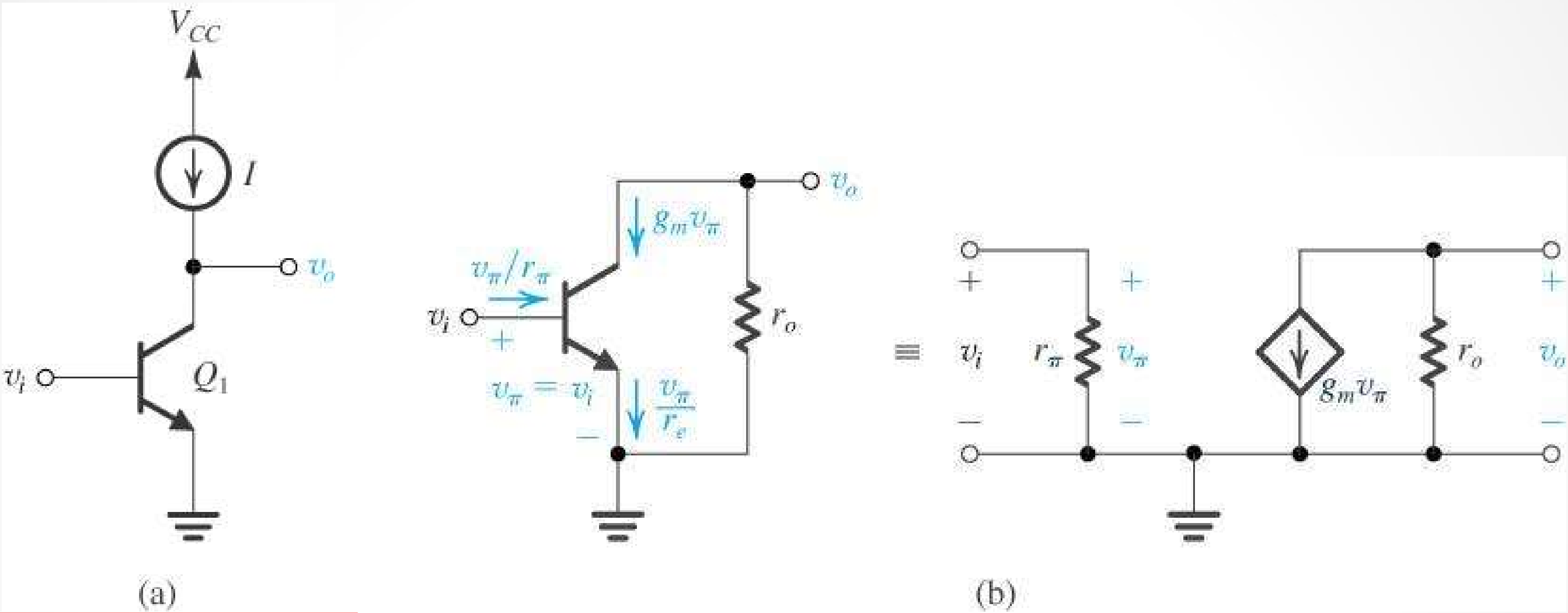
(a)

- In response to input signal voltage v_i, the CS transistor Q_1 conducts a current signal $gm_1 v_i$ in its drain terminal and feeds it to the source of the CG transistor Q_2 (cascode transistor).
- Q_2 passes signal to its drain and to the load R_L.
- Q_2 acts as a buffer, presenting low R_{in} to the drain of Q_1 and providing high R_{out} at output.

	Common Source	Cascode
Circuit		
DC Gain	$-g_m R'_L$	$-A_0 g_m R'_L$
f_{3dB}	$\dfrac{1}{2\pi(C_L + C_{gd})R'_L}$	$\dfrac{1}{2\pi(C_L + C_{gd})A_0 R'_L}$
f_t	$\dfrac{g_m}{2\pi(C_L + C_{gd})}$	$\dfrac{g_m}{2\pi(C_L + C_{gd})}$



Common-Emitter (CE) Amplifier: A Short Review

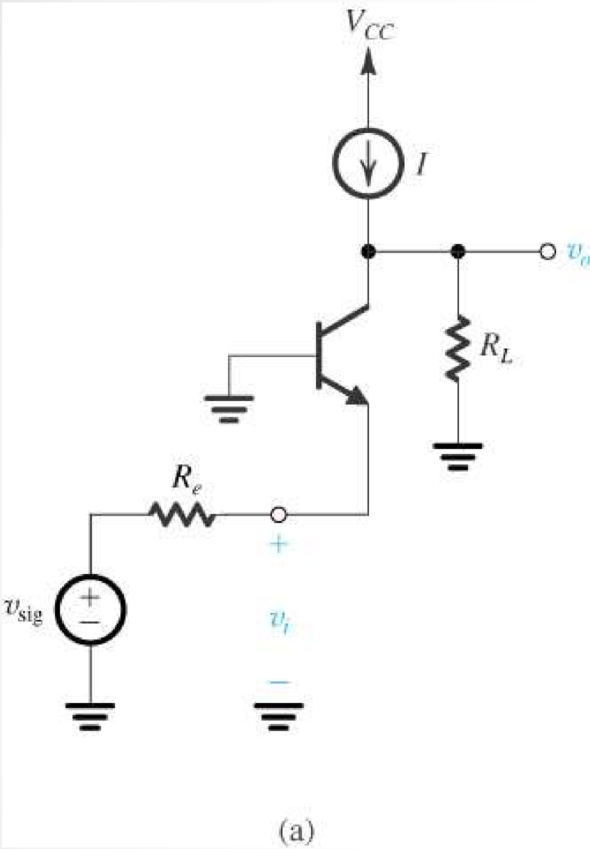


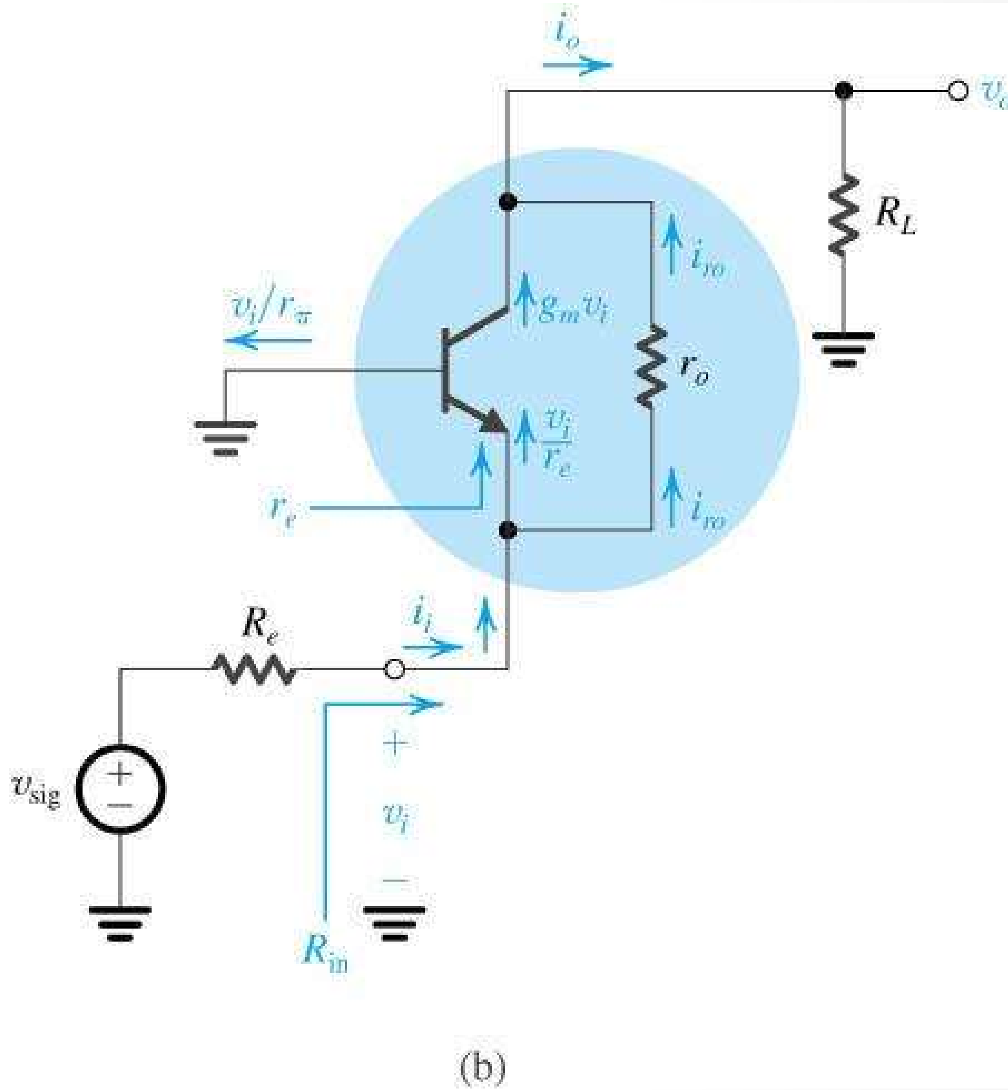
$$R_i = r_\pi$$

$$A_{vo} = -g_m r_o$$

$$R_o = r_o$$

High-Frequency Analysis of Common Base Amplifier





Note: the small-signal analysis performed directly on the ckt with the T model of the BJT used implicitly.

$$\boxed{i_o = i_i - v_i / r_\pi}$$

$$\boxed{R_{in} = \frac{r_o + R_L}{1 + \frac{r_o}{r_e} + \frac{R_L}{(\beta+1)r_e}} \approx r_e \frac{r_o + R_L}{r_o + R_L/(\beta+1)}}$$

$$\lim_{r_o \to \infty} R_{in} =$$

$$\lim_{R_L \to 0} R_{in} =$$

$$\lim_{R_L \to \infty} R_{in} =$$

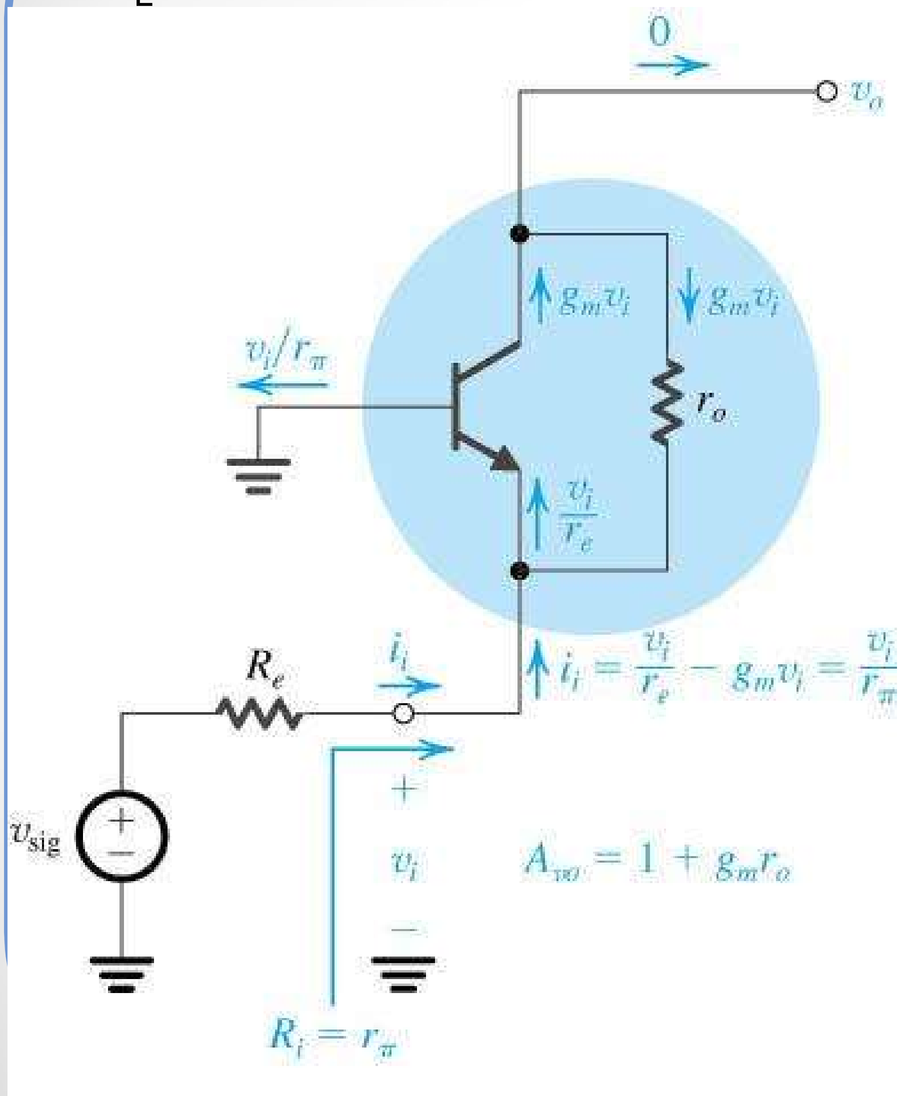
To find the open-ckt voltage gain and resistances, we'll use the ckt below with $R_L = \infty$

$$A_{vo} = 1 + g_m r_o = 1 + A_o$$

$$R_i = r_\pi \quad \text{and} \quad R_o = r_o$$

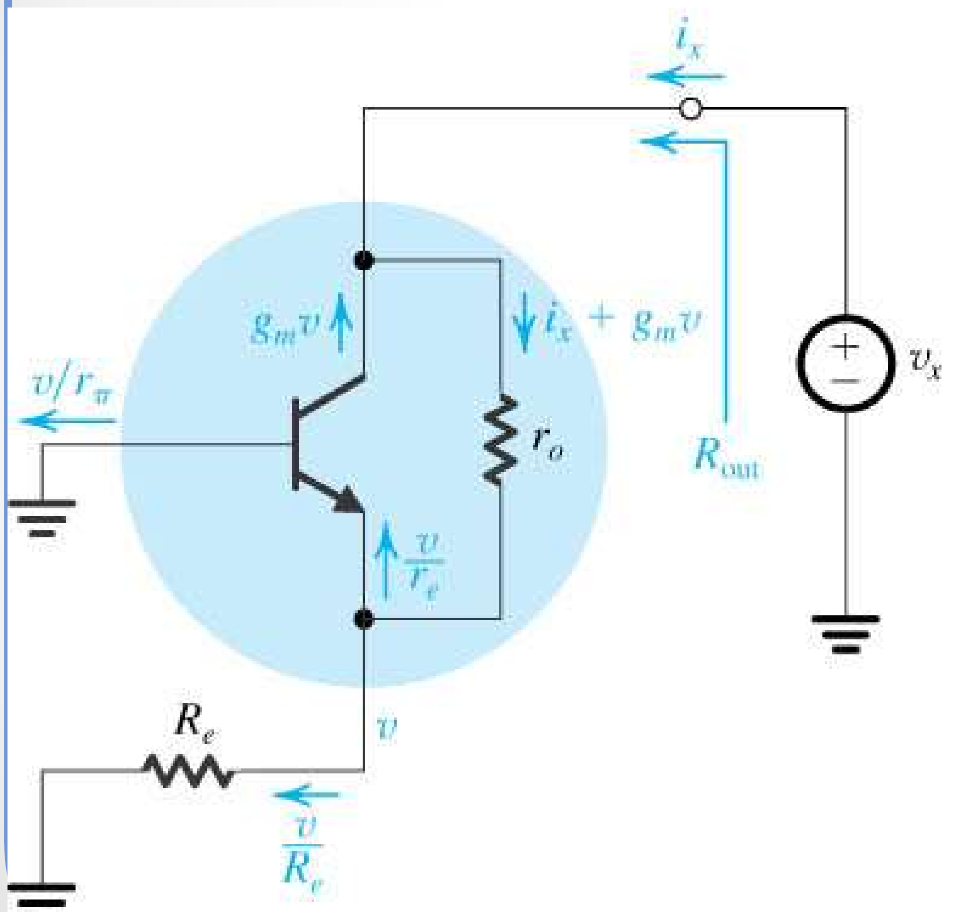
$$R_{out} = r_o + (1 + g_m r_o) R_e'$$

$$R_e' = R_e \| r_\pi$$



(c)

To find Rout, we put a test voltage Vx at the output and find R_{out} = Vx/Ix

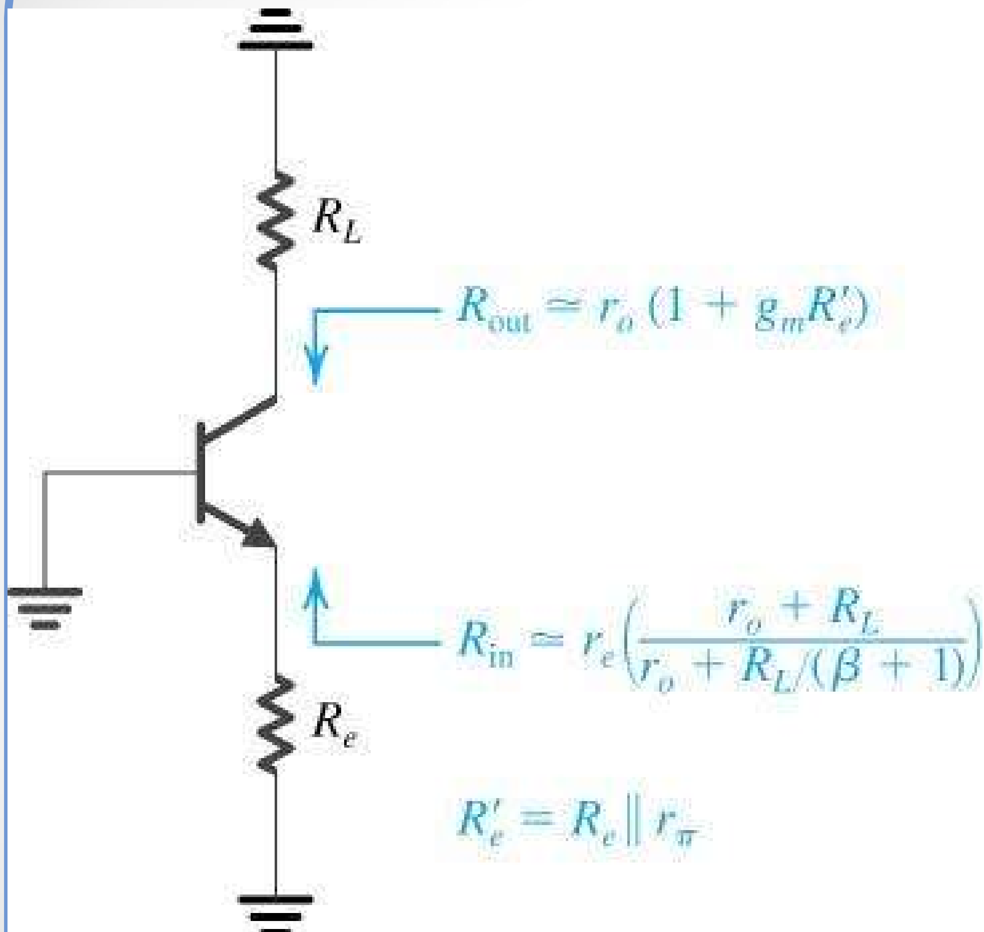


$$R_{out} = r_o + (1 + g_m r_o) R_e'$$

$$R_{out} = R_e' + (1 + g_m R_e') r_o$$

$$R_{out} \approx (1 + g_m R_e') r_o$$

Summary: input and output resistances of CB amplifier



The overall voltage gain G_v is given by

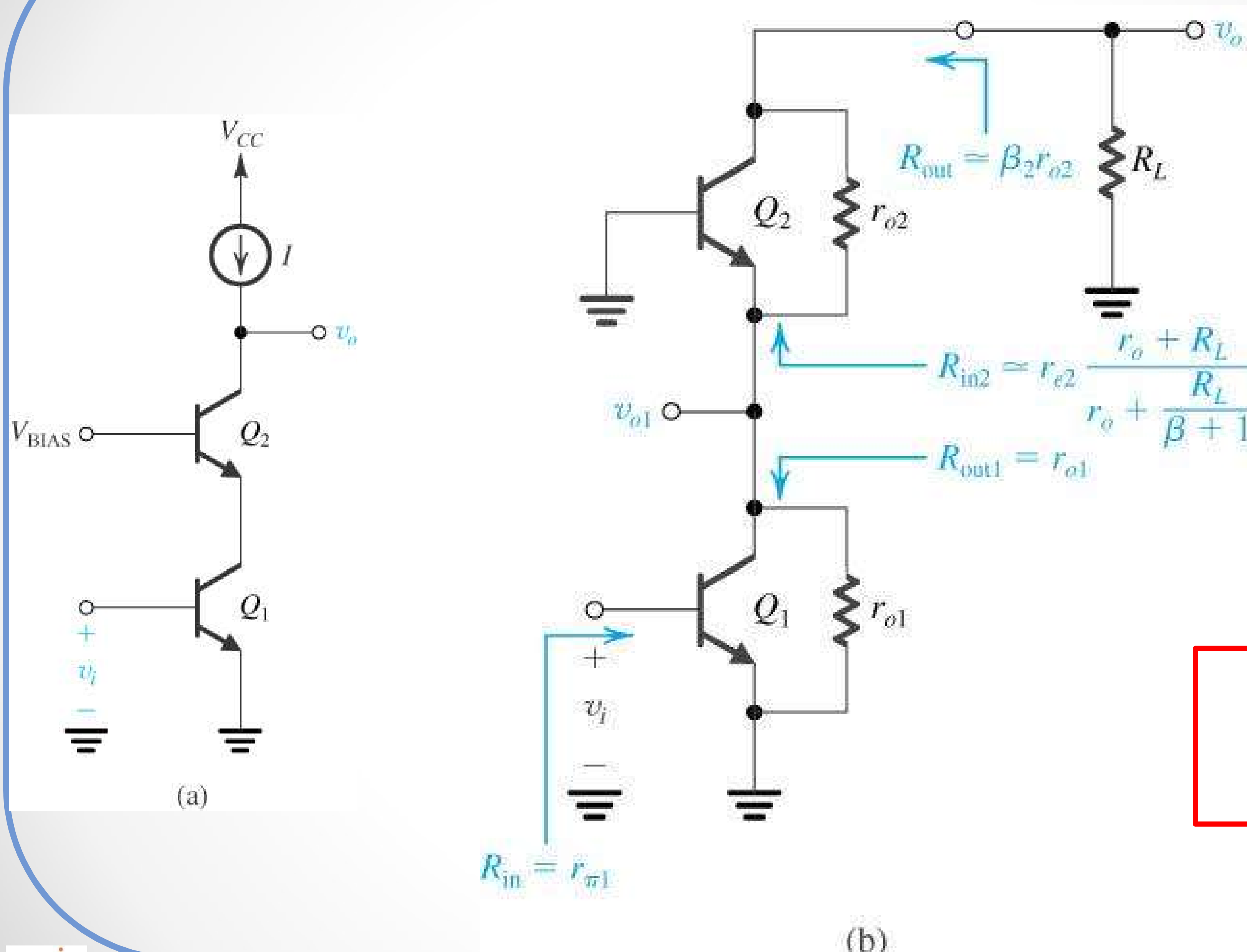
$$G_v = \frac{R_i}{R_i + R_e} A_{vo} \frac{R_L}{R_L + R_{out}}$$

$$G_v = \frac{r_\pi}{r_\pi + R_e} A_{vo} \frac{R_L}{R_L + R_{out}}$$

where

$$A_{vo} = 1 + g_m r_o = 1 + A_o$$

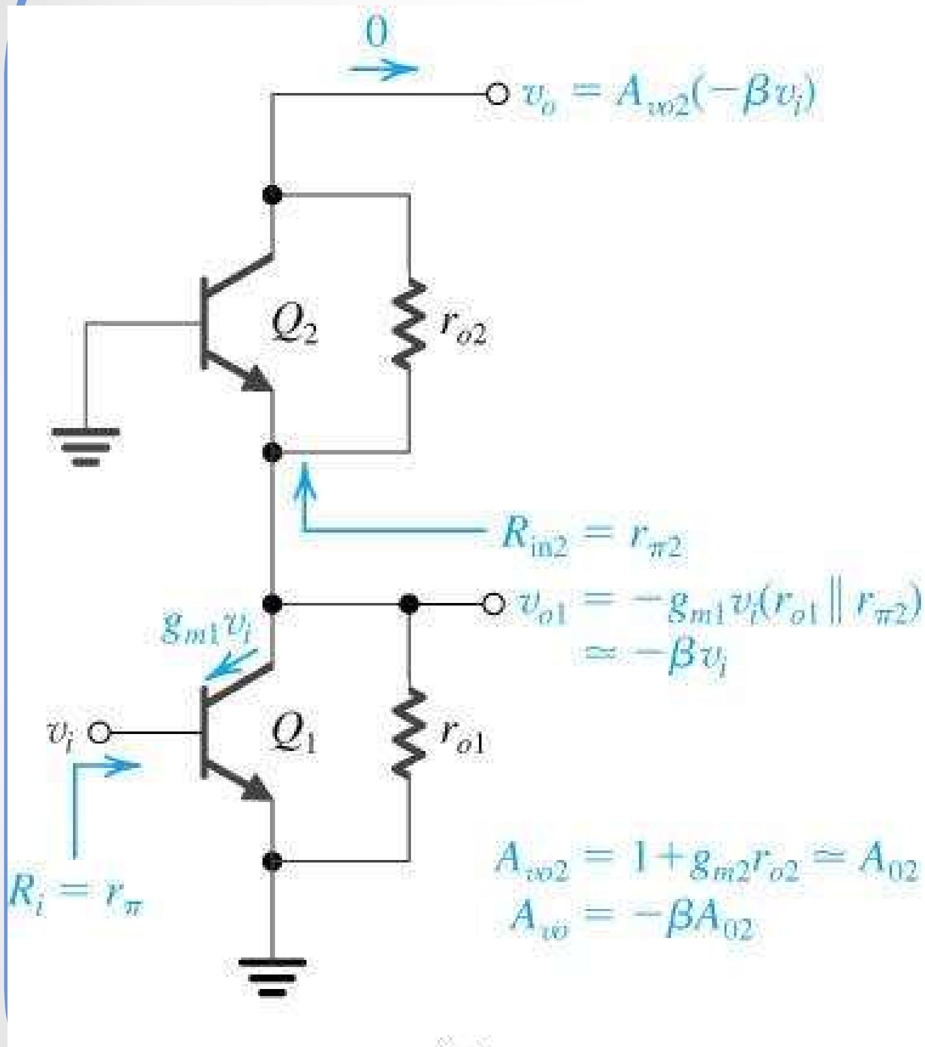
The Analysis of BJT Cascode



$$R_{in} = r_{\pi 1}$$

$$R_{out} \approx \beta_2 r_{o2}$$

To determine the open-ckt voltage gain,

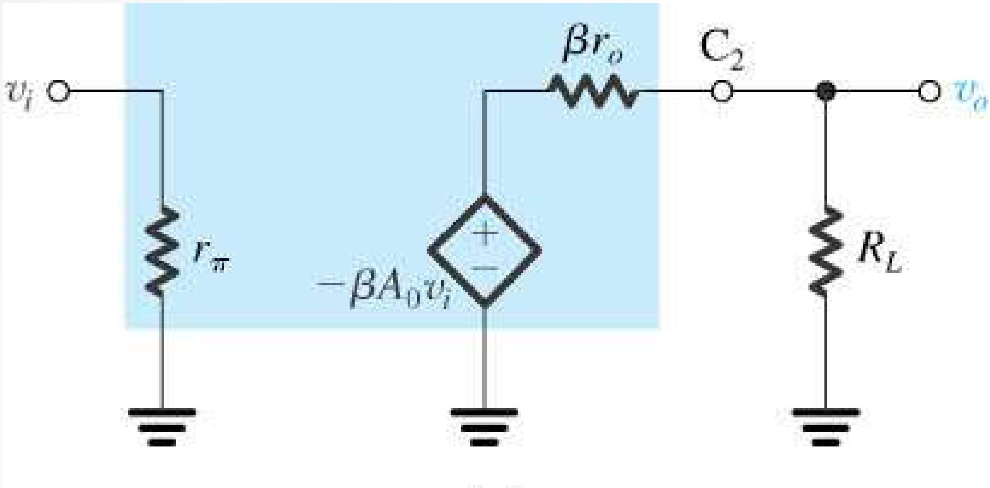


(c)

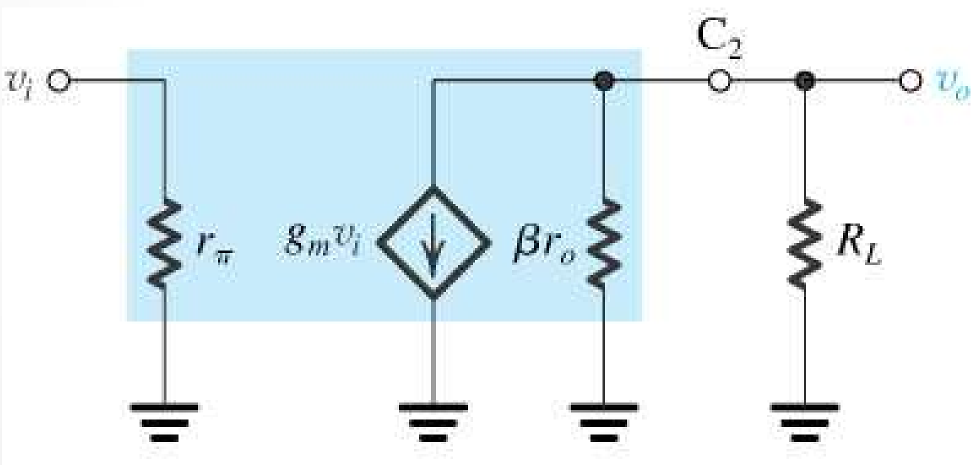
$$\boxed{A_{vo} = -\beta A_o}$$

- Compare to the CE amplifier, cascode increases both the open-ckt voltage gain (A_{vo}) and the output resistance (R_{out}) by a factor of β.
- In MOS cascode, the factor of increase in both parameters is A_o.

Equivalent ckt model of BJT cascode amplifier

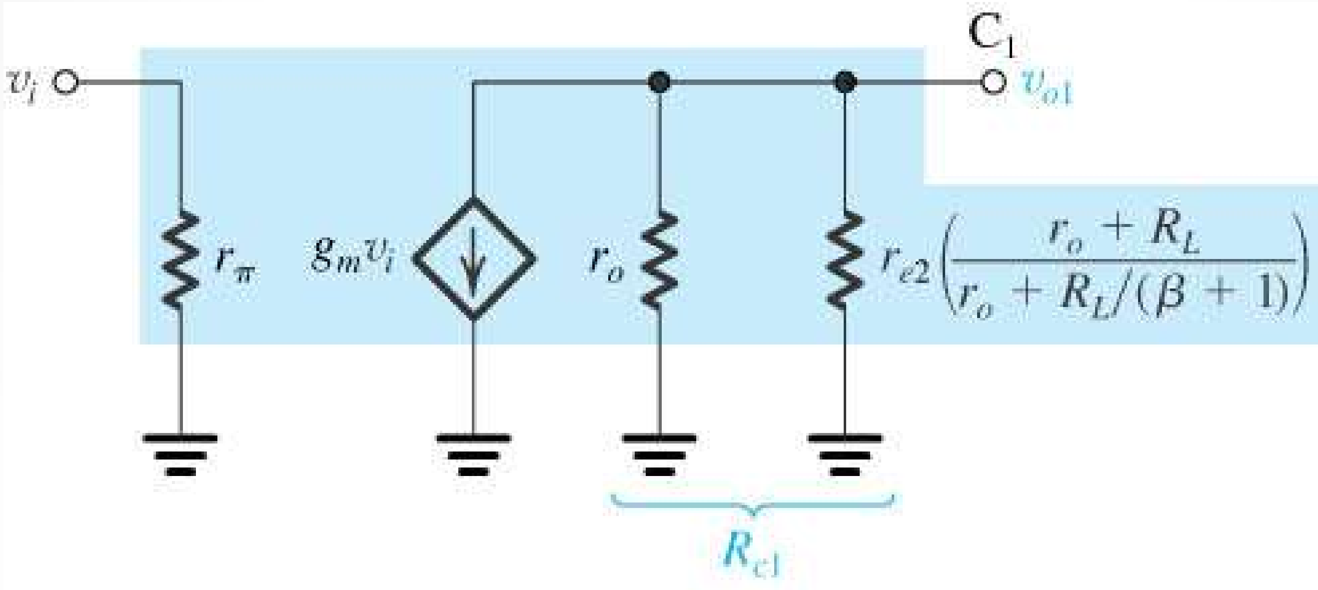


(a)



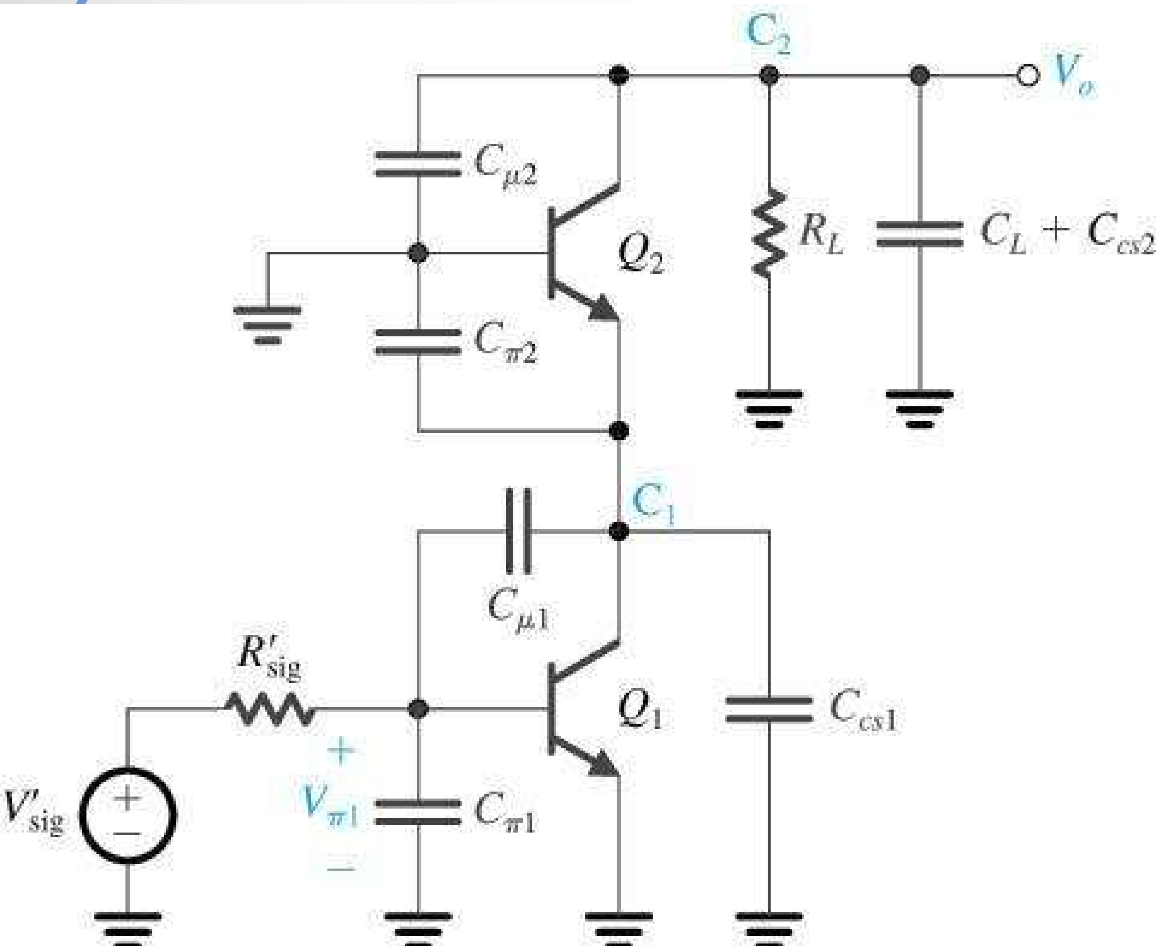
(b)

… Equivalenct ckt for determining gain of the CE stage, Q1.



(c)

High-Frequency Response of the BJT Cascode



$$R'_{sig} = r_{\pi 1} \| (r_{x1} + R_{sig})$$

$$R_{\pi 1} = R'_{sig}$$

$$R_{\mu 1} = R'_{sig}(1 + g_{m1}R_{c1}) + R_{c1}$$

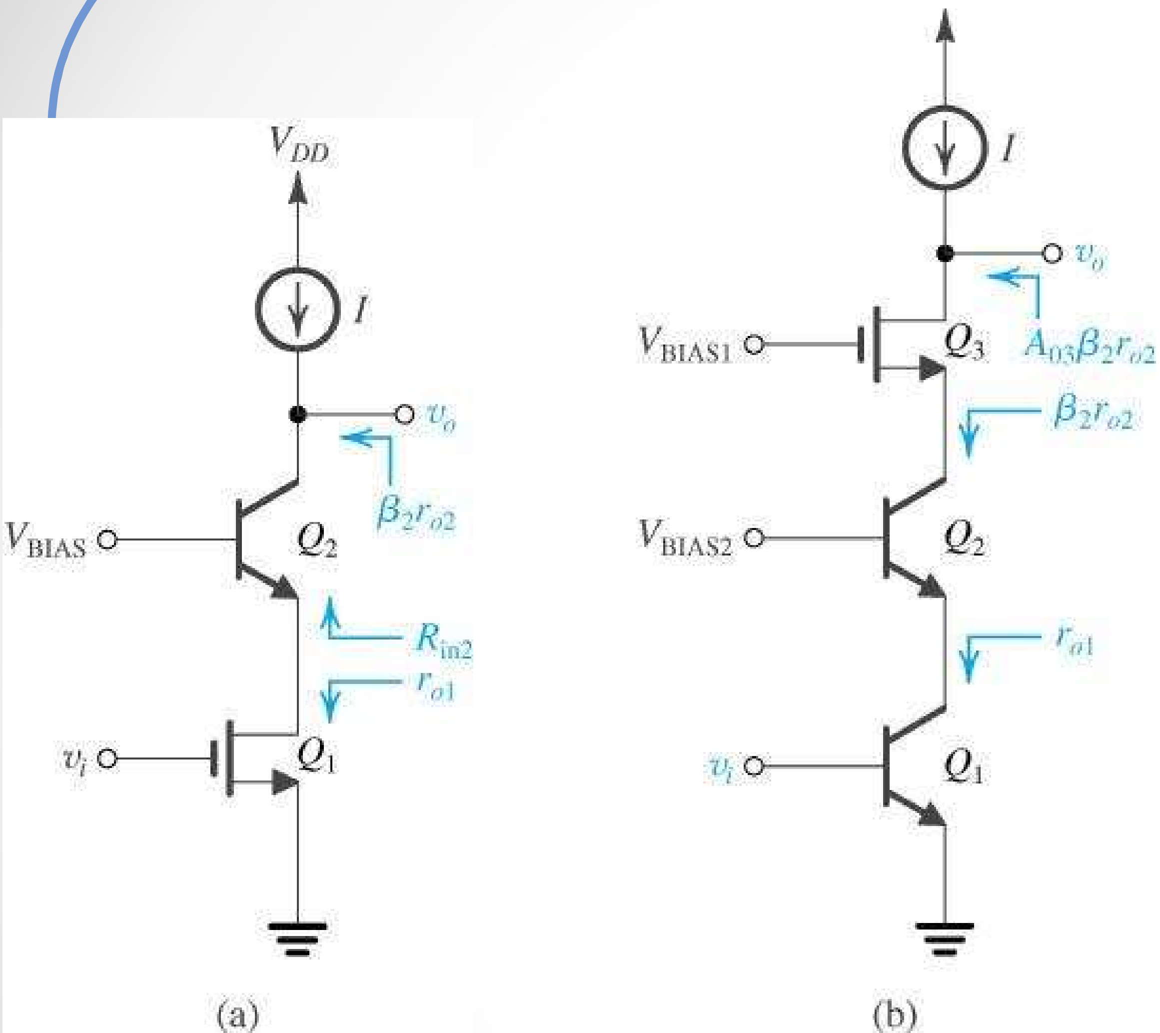
$$R_{c1} = r_{o1} \left\| \left[r_{e2}\left(\frac{r_{o2} + R_L}{r_{o2} + R_L/(\beta_2 + 1)} \right) \right] \right.$$

$$\tau_H = C_{\pi 1}R_{\pi 1} + C_{\mu 1}R_{\mu 1} + (C_{cs1} + C_{\pi 2})R_{c1} + (C_L + C_{cs2} + C_{\mu 2})(R_L \| R_{out})$$

$$f_H \simeq \frac{1}{2\pi\tau_H}$$

$$A_M = -\frac{r_\pi}{r_\pi + r_x + R_{sig}} g_m(\beta r_o \| R_L)$$

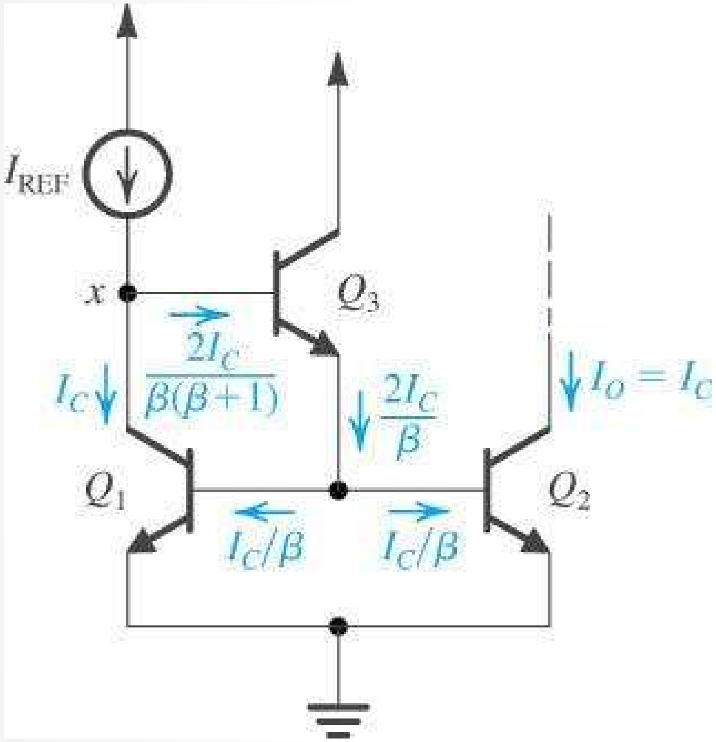
BiCMOS Cascodes



- MOSFET provides an infinite R_{in} while BJT cascode provide larger Rout since β of BJT is usually larger than Ao of the MOSFET. Also, r_o of BJT is higher than r_o of the MOSFET.
- BJT CB transistor provides lower R_{in2} than the CG transistor. Hence, reduce Miller effect in Q_1.
- The last ckt needs MOSFET to boost R_{out} since the max possible Rout provided by BJT is limited to βr_o. But there is no such limit in MOSFET.

Current-Mirror Ckts with Improved Performance

1. <u>A Bipolar with base current compensation</u>: the reduced dependency on β is achieved by adding transistor Q_3



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

$$I_O = I_C$$

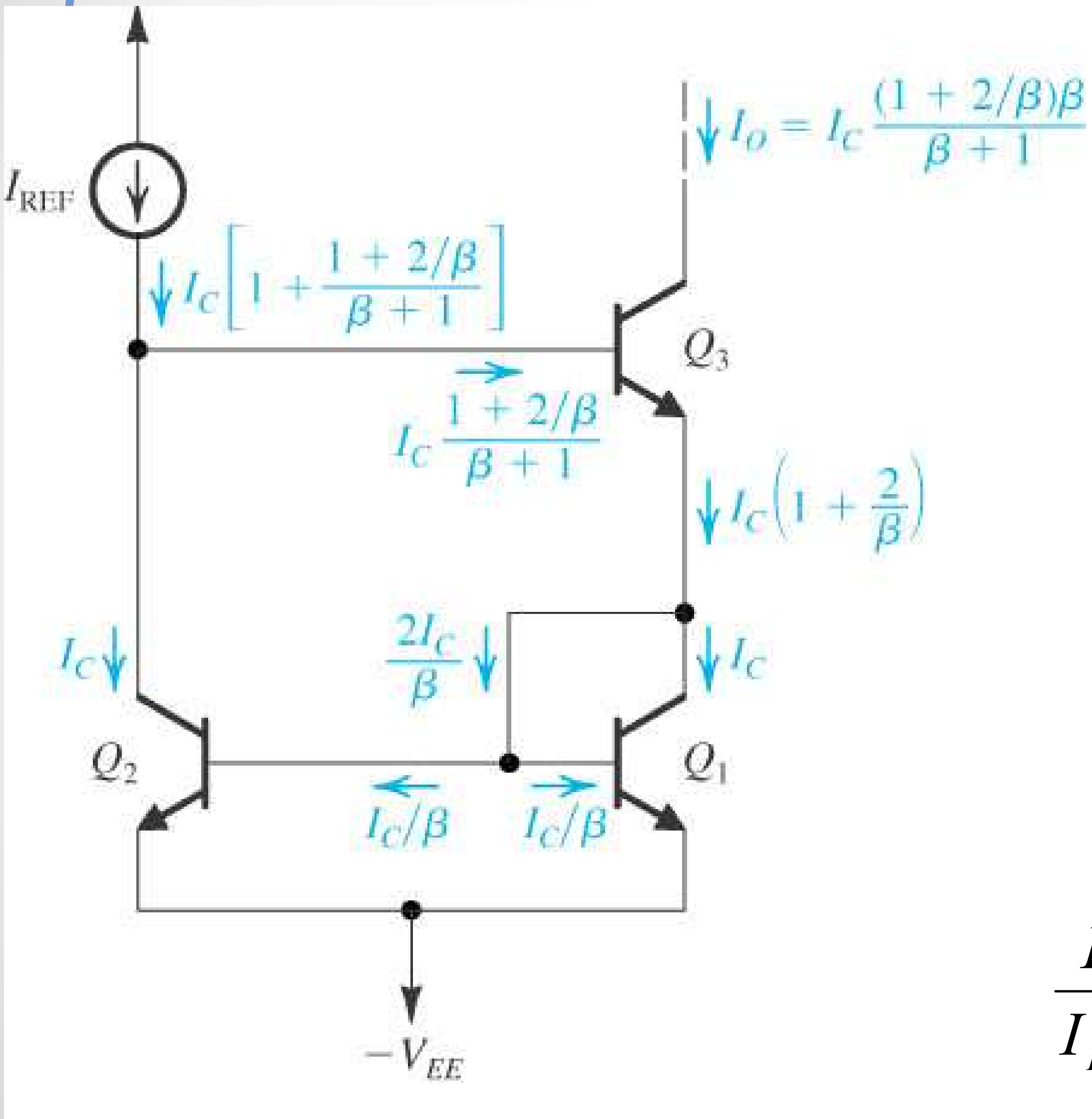
$$\boxed{\frac{I_O}{I_{REF}} = \frac{1}{1+2/(\beta^2+\beta)} \approx \frac{1}{1+2/\beta^2}}$$

If a reference current I_{REF} is not available, just connect V_x to V_{cc} through a resistor R.

$$I_{REF} = \frac{V_{CC} - V_{BE1} - V_{BE3}}{R}$$

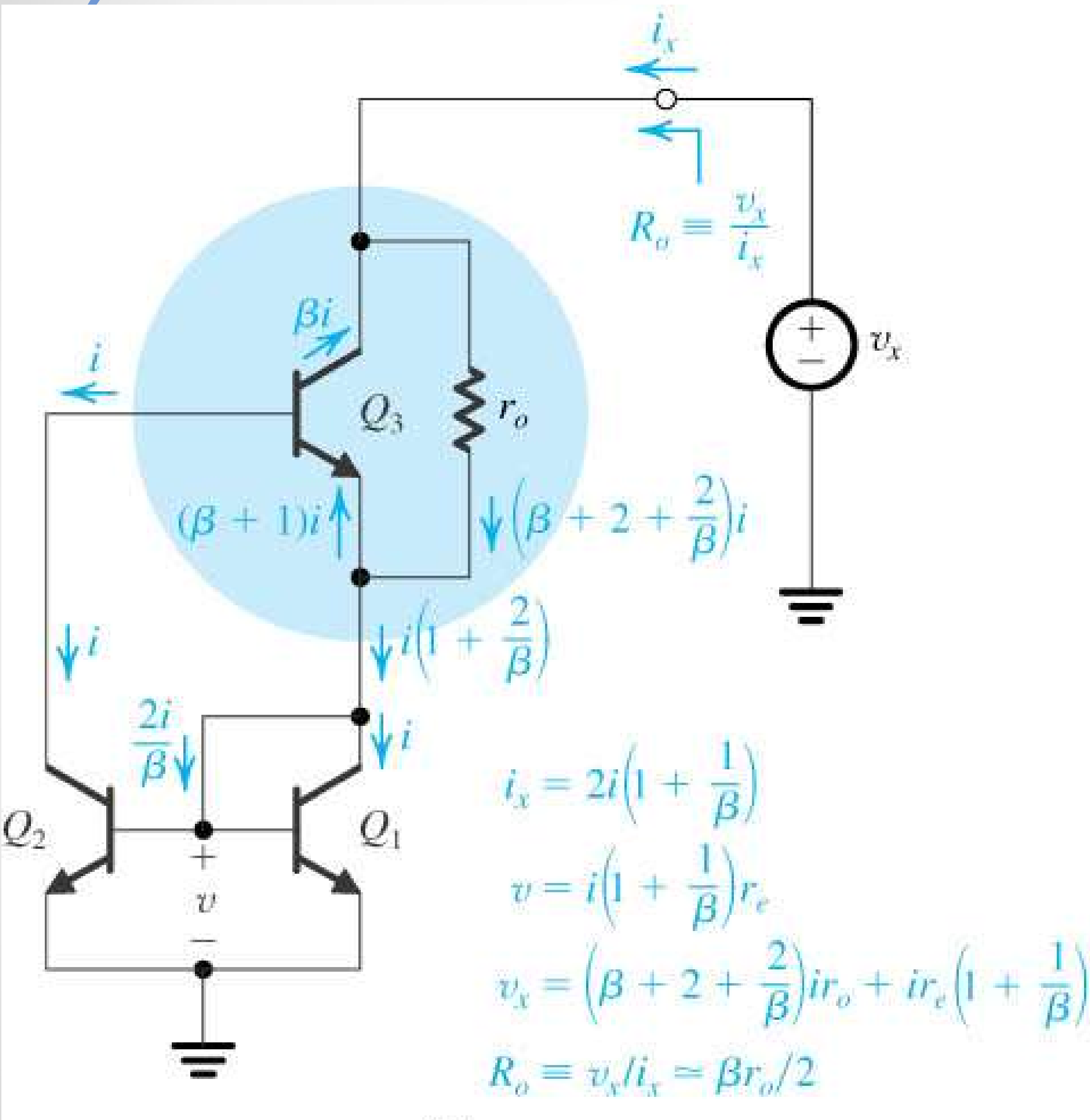
Current-Mirror Ckts with Improved Performance (cont.)

2. The Wilson Current Mirror: reduces the β dependence and increases R_{out}



(a)

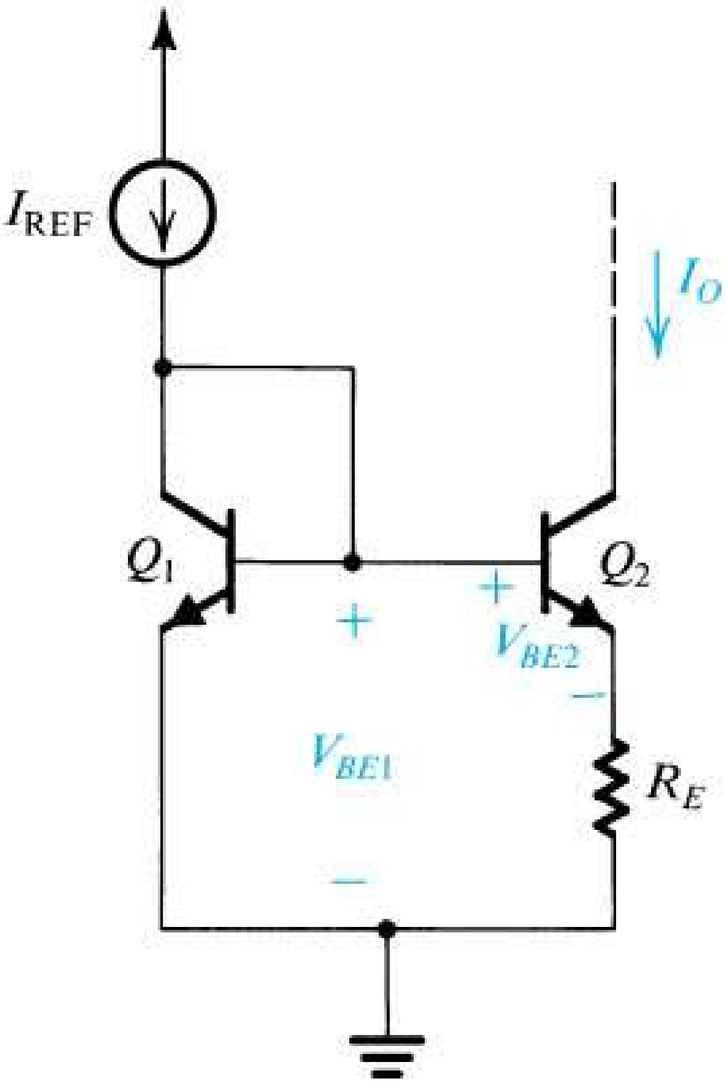
$$\frac{I_O}{I_{REF}} = \frac{1}{1+\dfrac{2}{\beta(\beta+2)}} \approx \frac{1}{1+2/\beta^2}$$



(b)

$$R_O \approx \beta r_o/2$$

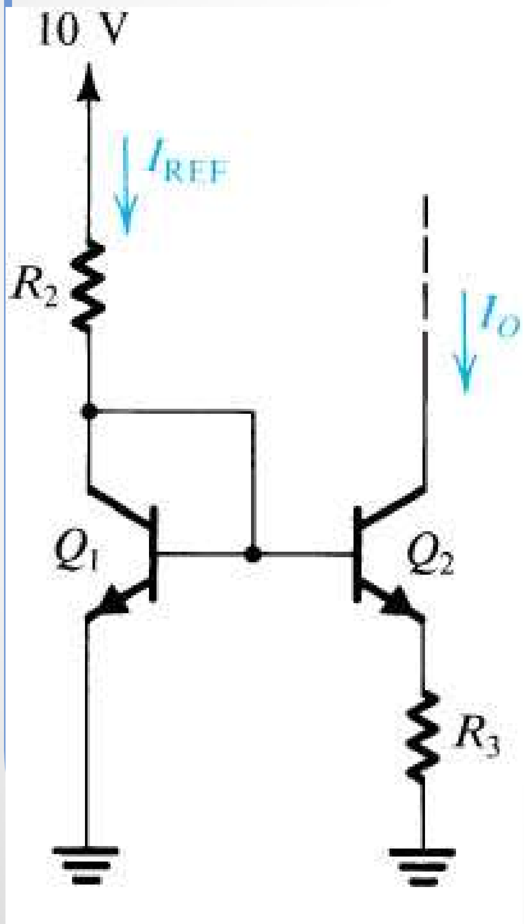
3. The Widlar Current Source: there is an R_E included in the emitter terminal



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

$$R_O = [1 + g_m (R_E \parallel r_\pi)] r_o$$

Example: From figure below, determine the values of the required resistors assuming that V_{BE} is 0.7 V at a current of 1 mA and neglecting the effect of β



(b)

Reference

Microelectronic Circuits by Adel S. Sedra & Kenneth C. Smith. Saunders College Publishing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