BECE206L - Analog Circuits

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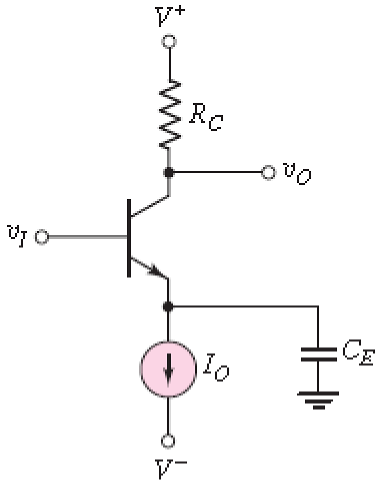
Module 3	MOSFET Active Biasing and Differential Amplifiers (6)
Introduction to Current Mirror – Basic, Wilson and Cascode Current Mirror, MOSFET Basic Differential Pair, Large Signal and Small Signal Analysis of Differential Amplifier, Differential Amplifier with active load.	

Introduction



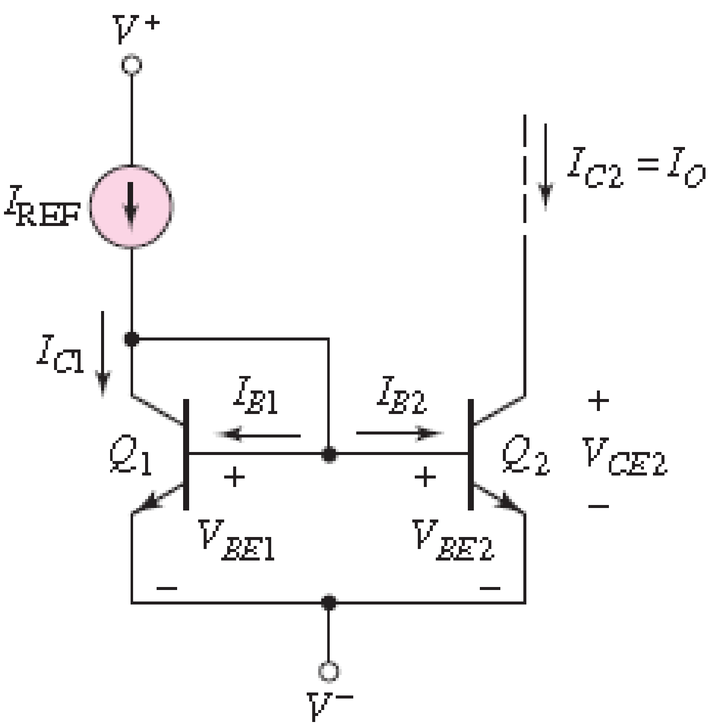
- Why is constant current source used for biasing in ICs?
  - resistor-intensive circuit would necessitate a large chip area
  - Hence, voltage divider biasing require relatively large area on IC
  - Also resistor biasing needs coupling capacitors in μF range, which is difficult to fabricate on an IC

Bipolar Transistor current sources

- Types of circuits to produce constant current I_o:
  - Two transistor current source
  - Three transistor current source
  - Widlar current source
  - Wilson Current source
  - Cascode current source

Two-Transistor current source

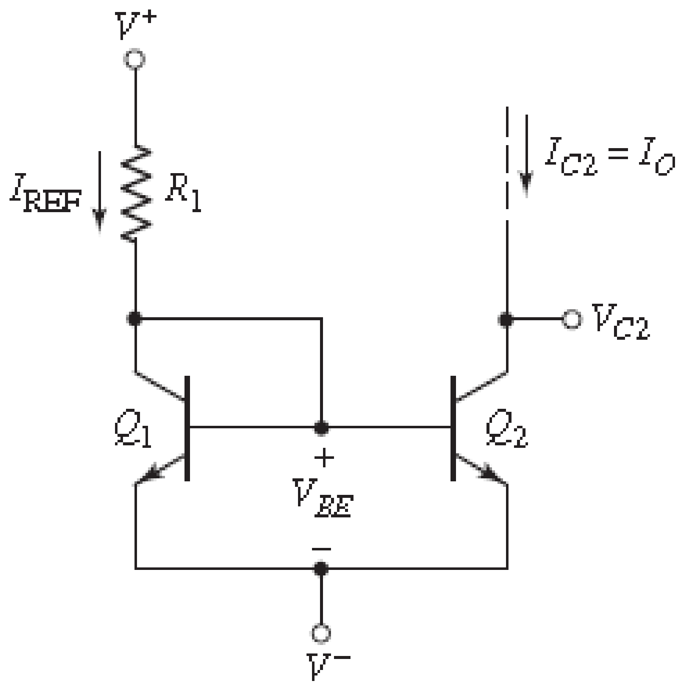
- Also called as **Current mirror**
- Use two *matched* or *identical* transistors-Q_1 & Q_2
  - Operating at same temperature
  - Base and emitter terminals are tied together
  - B-E voltage is same in both transistors
- Transistor Q_1 is used as diode
- When supply is ON,
  - V_{BE1} is established and I_{REF} flows
  - V_{BE2} is also same and turns ON Q_2
  - Generating load current I_o
  - I_o: used to bias a transistor circuit



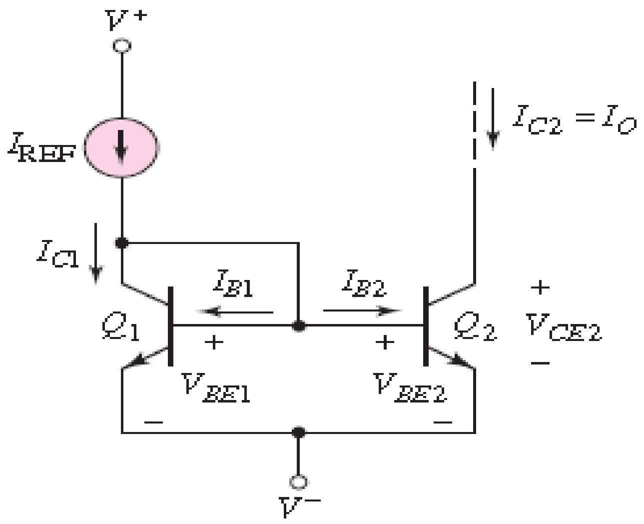
Two-transistor current source with resistor

- Current source can be replaced with resistor
- Then,

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



Current relationships



- Since V_{BE} is same in both the devices,

$$I_{B1} = I_{B2} \text{ and } I_{C1} = I_{C2}.$$

Transistor Q_2 is assumed to be biased in the forward-active region. If we sum the currents at the collector node of Q_1, we have

$$I_{\mathrm{REF}} = I_{C1} + I_{B1} + I_{B2} = I_{C1} + 2I_{B2}$$

Replacing I_{C1} by I_{C2} and noting that $I_{B2} = I_{C2}/\beta$,

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

The output current is then

$$I_{C2} = I_O = \frac{I_{\mathrm{REF}}}{1 + \frac{2}{\beta}}$$

assuming V_A is infinite

MOSFET Active Biasing

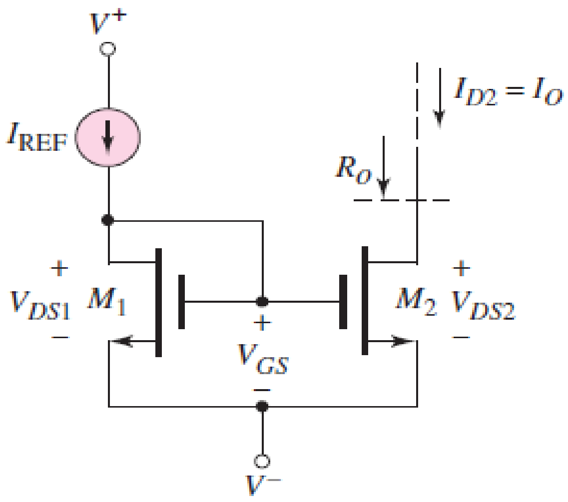
Outline

- Introduction to Current Mirror
- Basic Current Mirror
- Wilson Current Mirror
- Cascode Current Mirror

MOSFET Current Mirror

- Biasing in the integrated circuits is contant current sources by using
- The constant DC current generated at one location will be replicated in other locations for biasing various amplifier stages through a process known as “Current Steering”.
- The bias currents of various stages track eachother in case of changes in power supply voltage or in temperature.

Basic MOSFET Current Mirror



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

Solving for V_{GS} yields

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

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

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

If M_1 and M_2 are identical transistors, then $V_{TN1} = V_{TN2}$ and $K_{n1} = K_{n2}$, $\quad I_O = I_{REF}$

If the transistors are matched except for the aspect ratios, we find

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

Effect of Output Resistance

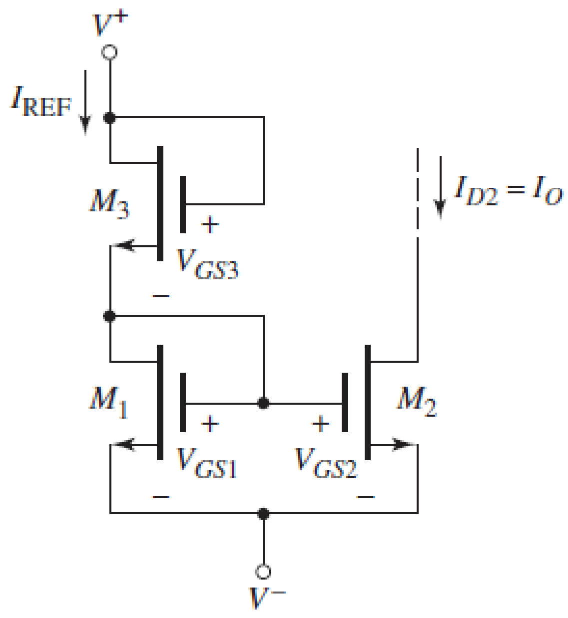
Taking into account the finite output resistance of the transistors, we can write the load and reference currents as

$$I_O = K_{n2}(V_{GS} - V_{TN2})^2(1 + \lambda_2 V_{DS2})$$

$$I_{\text{REF}} = K_{n1}(V_{GS} - V_{TN1})^2(1 + \lambda_1 V_{DS1})$$

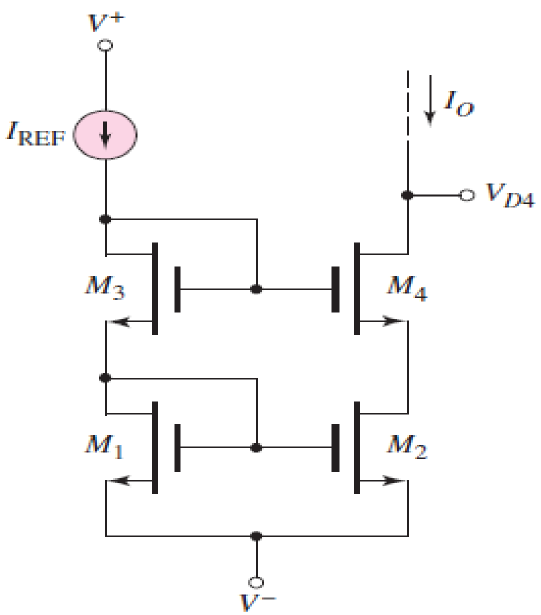
$$\frac{I_O}{I_{\text{REF}}} = \frac{(W/L)_2}{(W/L)_1} \cdot \frac{(1 + \lambda V_{DS2})}{(1 + \lambda V_{DS1})}$$

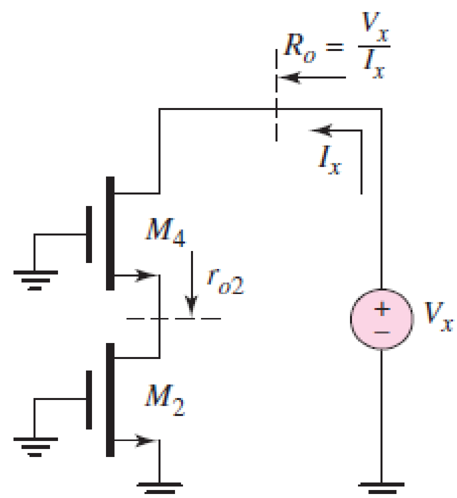
Replacing Resistor with M3

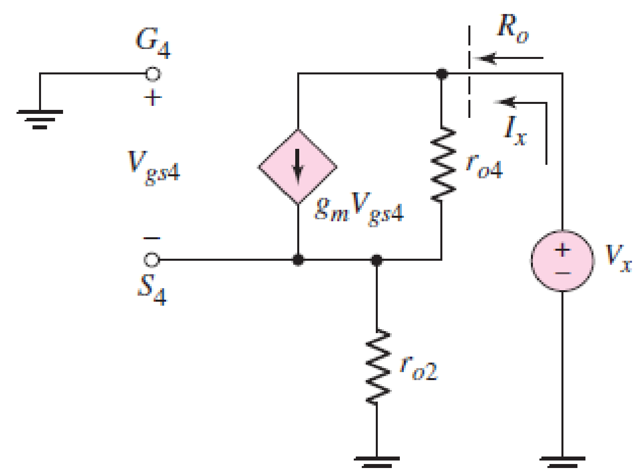


Cascode Current Mirror

In MOSFET current-source circuits, the output resistance is a measure of the stability with respect to changes in the output voltage. This output resistance can be increased by modifying the circuit, which is a **cascode current mirror.**

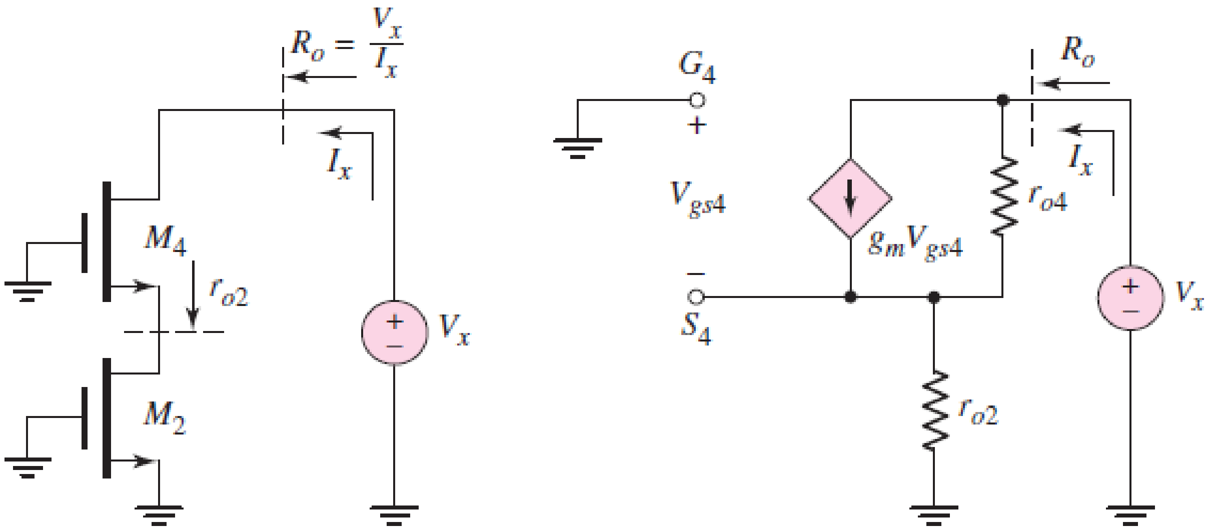






λ Assuming all transistors are identical, then $I_O = I_{REF}$

Since IREF is a constant, the gate voltages to M1 and M3 , and hence to M2 and M4, are constant. This is equivalent to an ac short circuit.



The small-signal resistance looking into the drain of M2 is rO2.

Writing a KCL equation, at the output node,

$$I_x = g_m V_{gs4} + \frac{V_x - (-V_{gs4})}{r_{o4}} \qquad V_{gs4} = -I_x r_{o2}$$

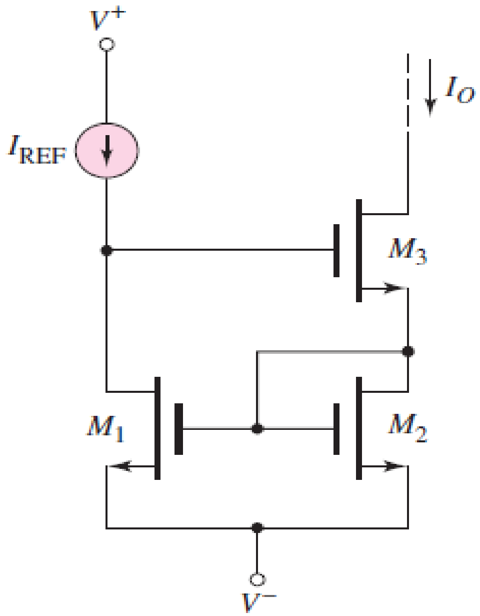
$$I_x + \frac{r_{o2}}{r_{o4}} I_x + g_m r_{o2} I_x = \frac{V_x}{r_{o4}}$$

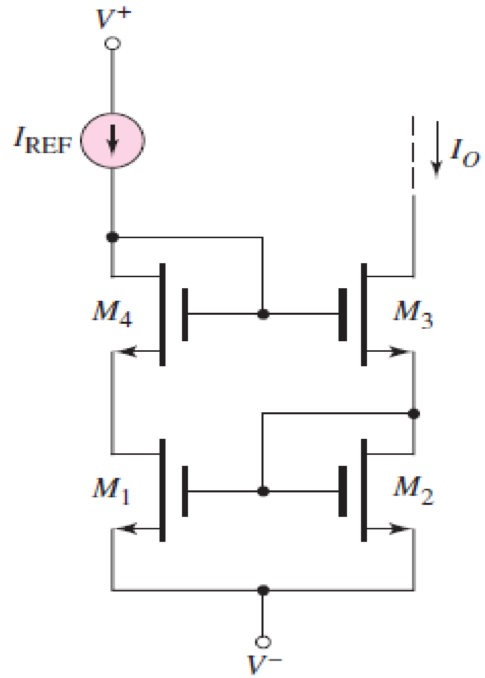
The output resistance is then

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

Normally, $g_m r_{o4} \gg 1$, which implies that the output resistance of this cascode configuration is much larger than that of the basic two-transistor current source.

Wilson Current Mirror





- Note that the VDS values of M1 and M2 are not equal. Since λ is not zero, the ratio IO/IREF is slightly different from the aspect ratios.
- This problem is solved in the modified Wilson Current Source

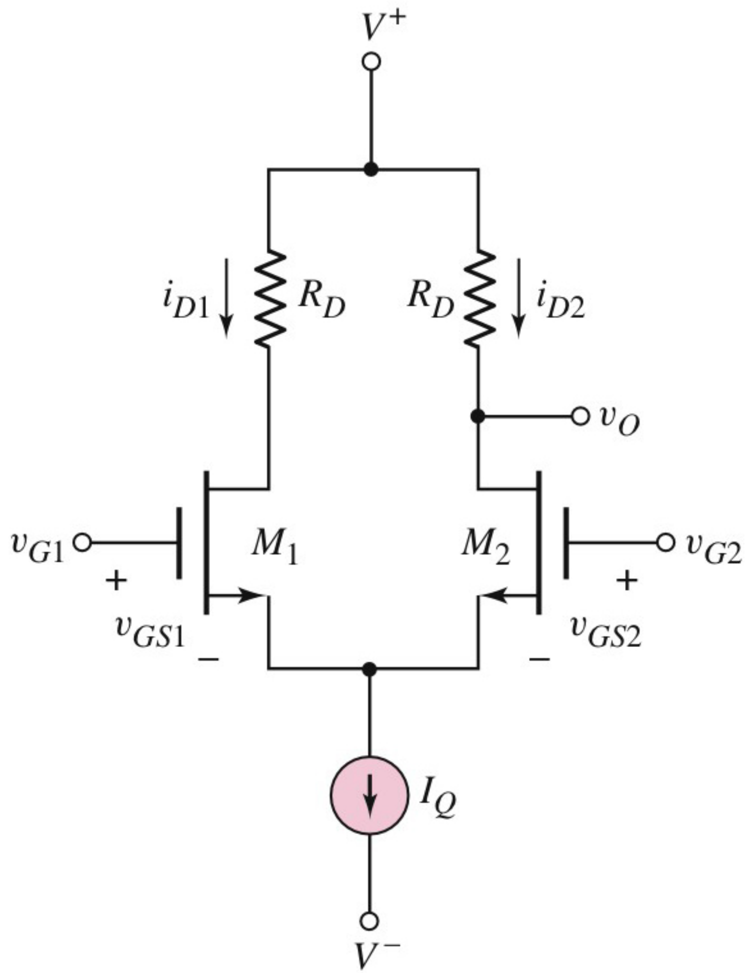
For a constant reference current, the drain-to-source voltages of M1 , M2 , and M4 are held constant.

The primary advantage of these circuits is the increase in output resistance, which further stabilizes the load current.

In this chapter, we will:

- Describe the characteristics and terminology of the ideal differential amplifier.
- Describe the characteristics of and analyze:
  - the basic FET differential amplifier.
  - BJT and FET differential amplifiers with active loads.
- Analyze the frequency response of the differential amplifier.

MOSFET Differential Pair



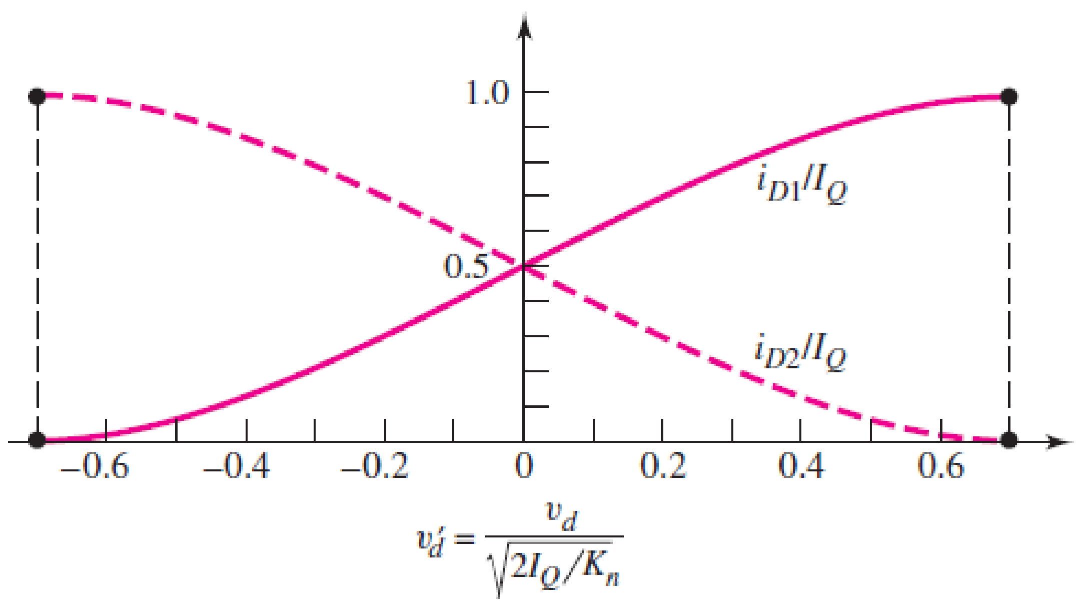
Refer class notes

The normalized drain currents are

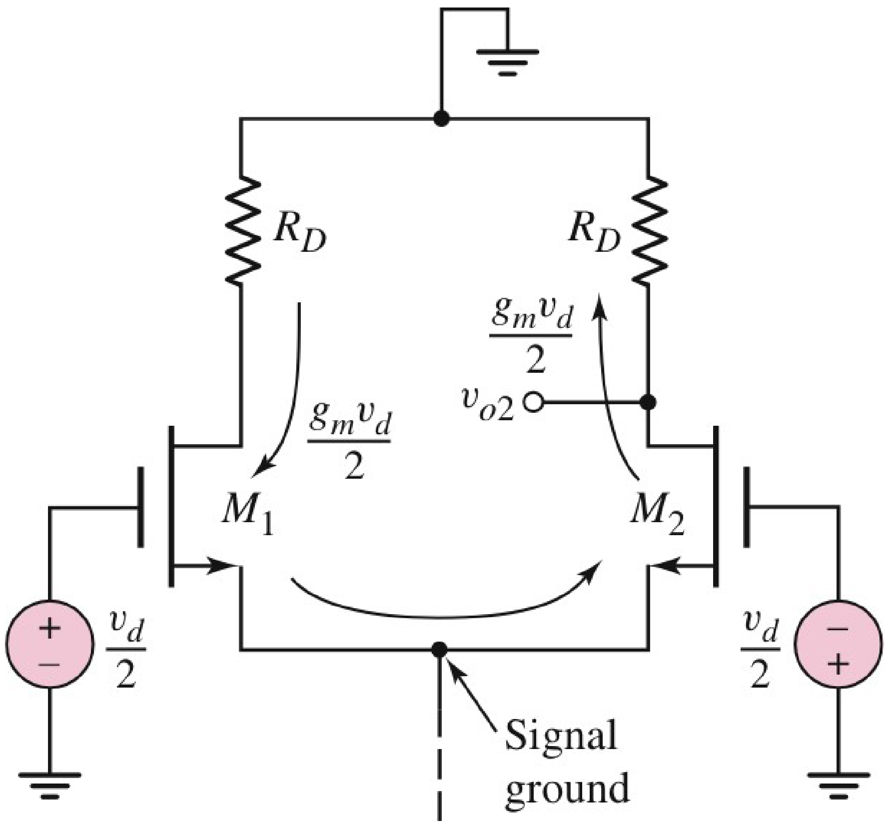
$$\frac{i_{D1}}{I_Q} = \frac{1}{2} + \sqrt{\frac{K_n}{2I_Q}} \cdot v_d \sqrt{1 - \left(\frac{K_n}{2I_Q}\right) v_d^2}$$

and

$$\frac{i_{D2}}{I_Q} = \frac{1}{2} - \sqrt{\frac{K_n}{2I_Q}} \cdot v_d \sqrt{1 - \left(\frac{K_n}{2I_Q}\right) v_d^2}$$



Differential mode gain



Assume that the output resistance looking into the current source is infinite.

Using this equivalent circuit, the one sided output voltage, at M_2 is given as,

$$v_{o2} \equiv v_o = +\left(\frac{g_m v_d}{2}\right) R_D$$

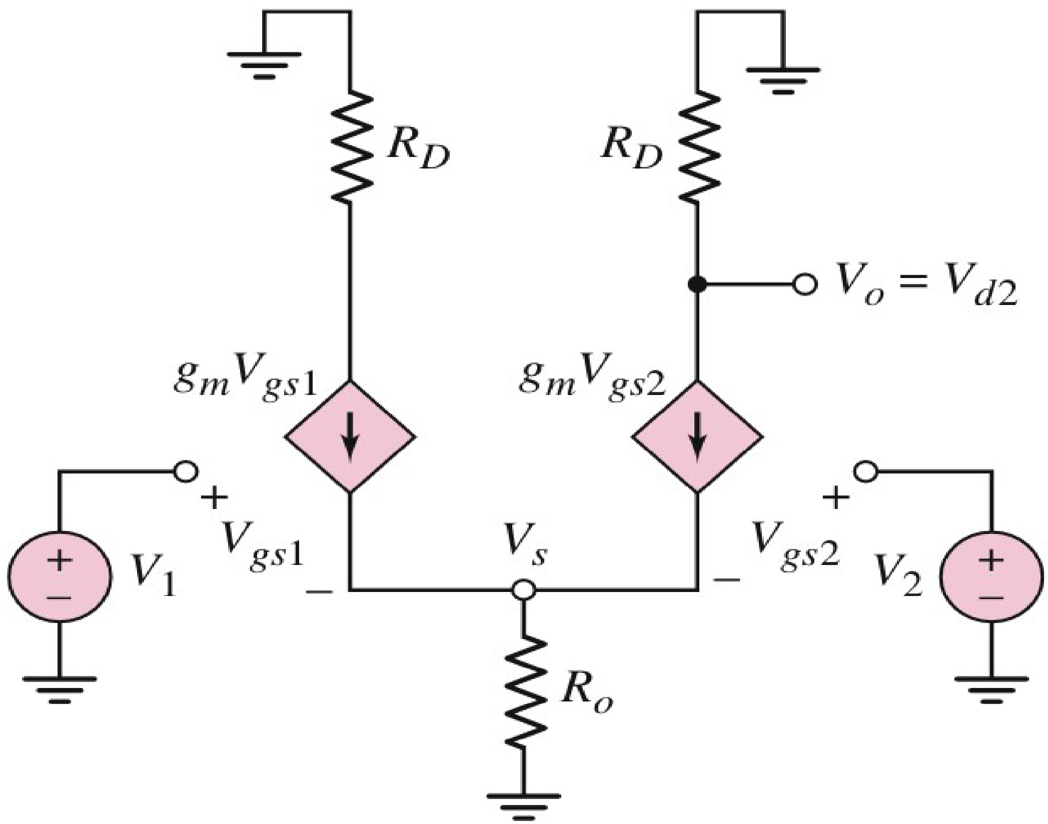
Differential mode gain and input impedances

- The differential voltage gain is

$$A_d = \frac{v_o}{v_d} = \frac{g_m R_D}{2} = \sqrt{\frac{K_n I_Q}{2}} \cdot R_D$$

- The differential and common-mode input impedances are:
  - We know that, at low frequencies, the input impedance of the MOSFET is infinite
  - Hence both the differential and common-mode input impedances are infinite too
  - As a design trade-off, sacrifice the differential mode gain.

Small-Signal Equivalent Circuit: MOSFET Differential Amplifier



Assume that the transistors are matched, with λ=0
The output resistance of constant current source is finite, R_o
The two transistors are biased with same quiescent current and

$$g_{m1} = g_{m2} \equiv g_m.$$

Refer Class notes

Writing a KCL equation at node V_s, we have

$$g_m V_{gs1} + g_m V_{gs2} = \frac{V_s}{R_o}$$

$V_{gs1} = V_1 - V_s$ and $V_{gs2} = V_2 - V_s$

$$g_m(V_1 + V_2 - 2V_s) = \frac{V_s}{R_o}$$

Solving for V_s we obtain

$$V_s = \frac{V_1 + V_2}{2 + \frac{1}{g_m R_o}}$$

For a one-sided output at the drain of M_2, we have

$$V_o = V_{d2} = -(g_m V_{gs2}) R_D = -(g_m R_D)(V_2 - V_s)$$

$$V_o = -g_m R_D \left[\frac{V_2 \left(1 + \frac{1}{g_m R_o}\right) - V_1}{2 + \frac{1}{g_m R_o}} \right]$$

$$V_o = \frac{g_m R_D}{2} V_d - \frac{g_m R_D}{1 + 2 g_m R_o} V_{cm}$$

The output voltage, in general form, is

$$V_o = A_d V_d + A_{cm} V_{cm}$$

The transconductance g_m of the MOSFET is

$$g_m = 2\sqrt{K_n I_{DQ}} = \sqrt{2K_n I_Q}$$

$$A_d = \frac{g_m R_D}{2} = \sqrt{2K_n I_Q}\left(\frac{R_D}{2}\right) = \sqrt{\frac{K_n I_Q}{2}} \cdot R_D$$

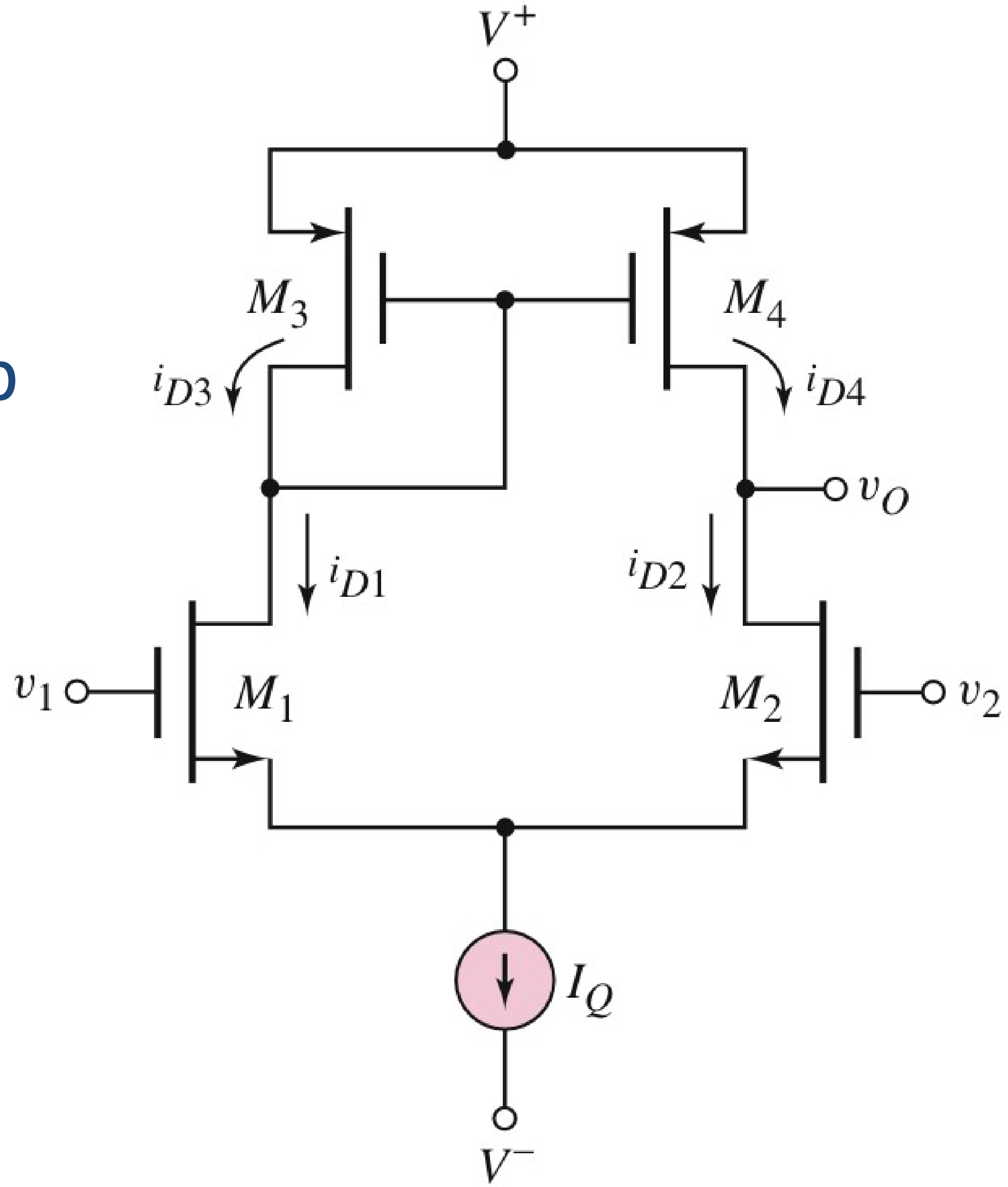
and the common-mode gain

$$A_{cm} = \frac{-g_m R_D}{1 + 2g_m R_o} = \frac{-\sqrt{2K_n I_Q} \cdot R_D}{1 + 2\sqrt{2K_n I_Q} \cdot R_o}$$

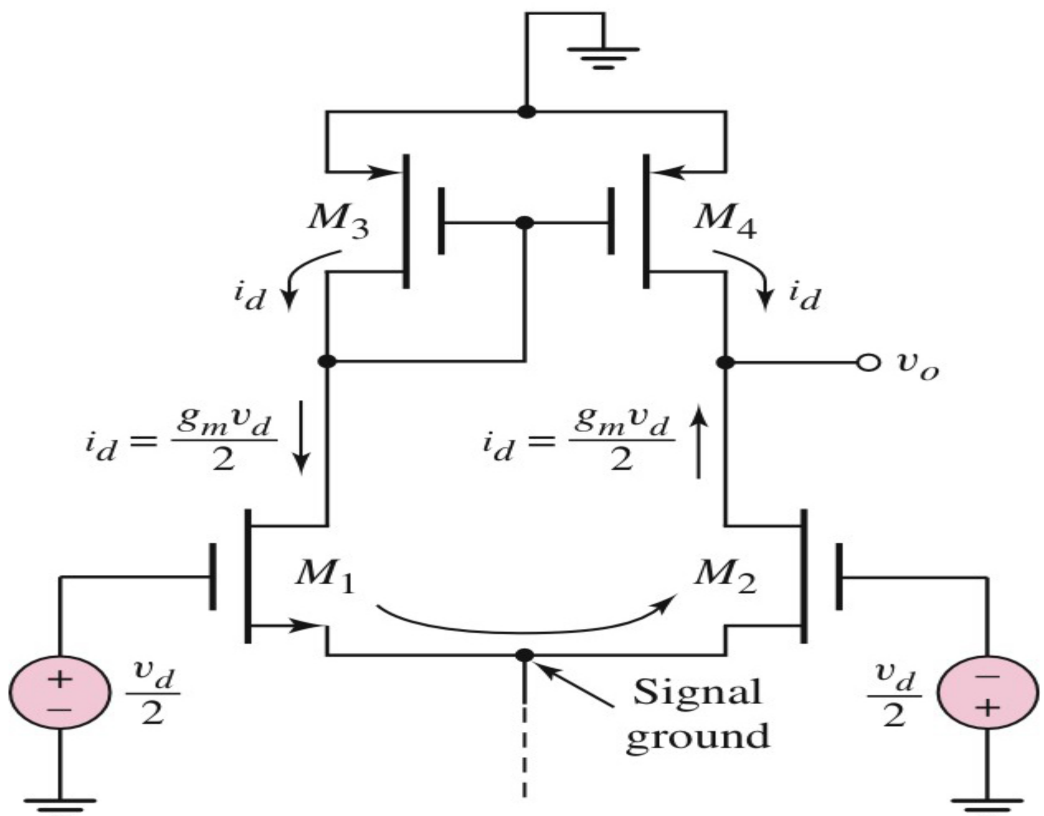
$$\text{CMRR} = \frac{1}{2}\left[1 + 2\sqrt{2K_n I_Q} \cdot R_o\right]$$

Refer class notes

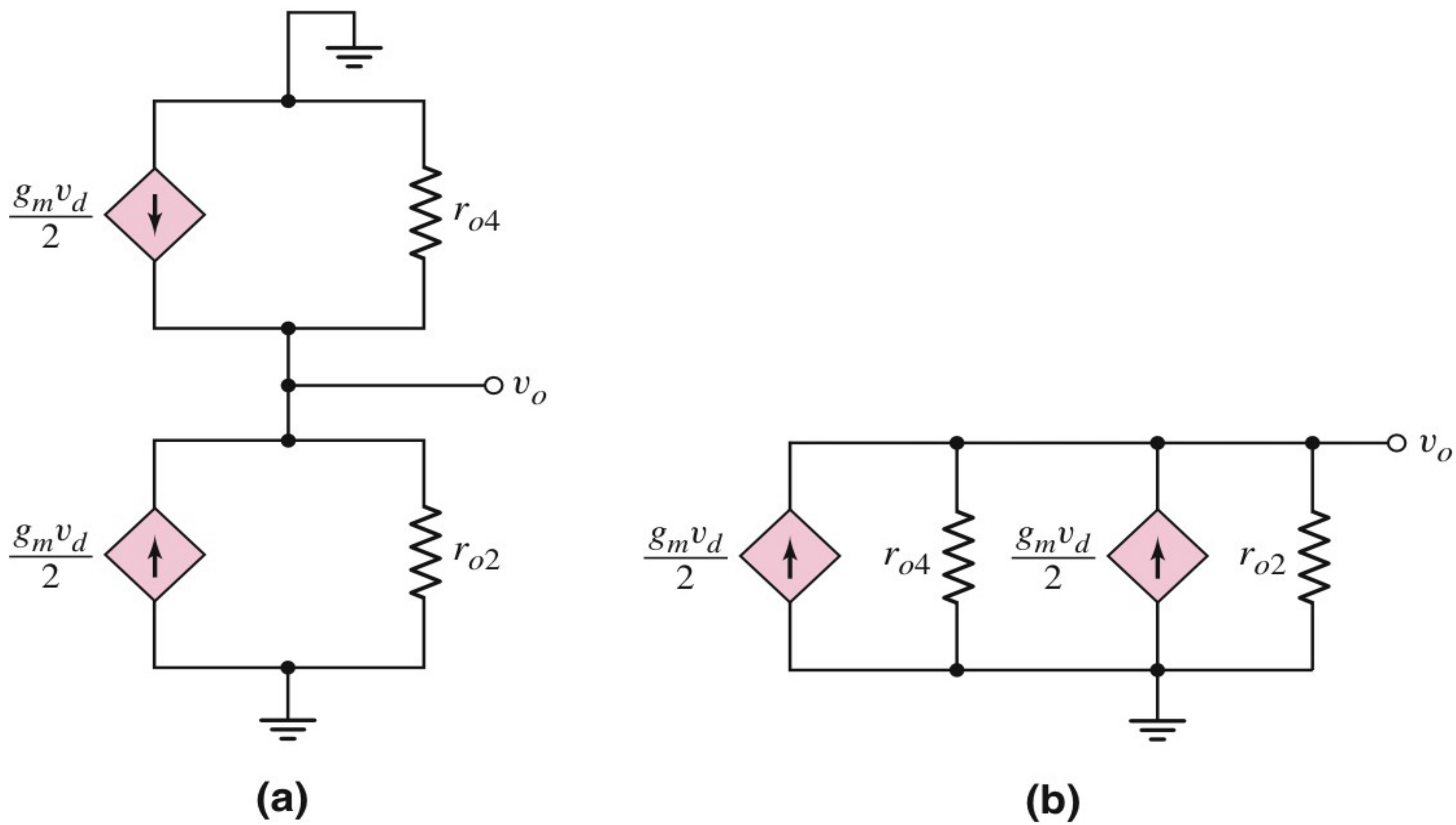
MOSFET Diff-Amp with Active Load



AC Equivalent Circuit: MOSFET Diff-Amp with Active Load



Small-Signal Equivalent Circuit: MOSFET Diff-Amplifier with Active Load



Differential Mode Gain

$$v_o = 2\left(\frac{g_m v_d}{2}\right)(r_{o2} \| r_{o4})$$

$$A_d = \frac{v_o}{v_d} = g_m(r_{o2} \| r_{o4})$$

Equation (11.106) can be rewritten in the form

$$A_d = \frac{g_m}{\dfrac{1}{r_{o2}} + \dfrac{1}{r_{o4}}} = \frac{g_m}{g_{o2} + g_{o4}}$$

$g_m = 2\sqrt{K_n I_D} = \sqrt{2K_n I_Q}$, $g_{o2} = \lambda_2 I_{DQ2} = (\lambda_2 I_Q)/2$, $g_{o4} = \lambda_4 I_{DQ4} = (\lambda_4 I_Q)/2$, then

$$A_d = \frac{2\sqrt{2K_n I_Q}}{I_Q(\lambda_2 + \lambda_4)} = 2\sqrt{\frac{2K_n}{I_Q}} \cdot \frac{1}{\lambda_2 + \lambda_4}$$