LECTURE 20 – LOW INPUT RESISTANCE AMPLIFIERS – THE COMMON GATE, CASCODE AND CURRENT AMPLIFIERS

LECTURE ORGANIZATION

Outline

- Voltage driven common gate amplifiers
- Voltage driven cascode amplifier
- Non-voltage driven cascode amplifier – the Miller effect
- Further considerations of cascode amplifiers
- Current amplifiers
- Summary

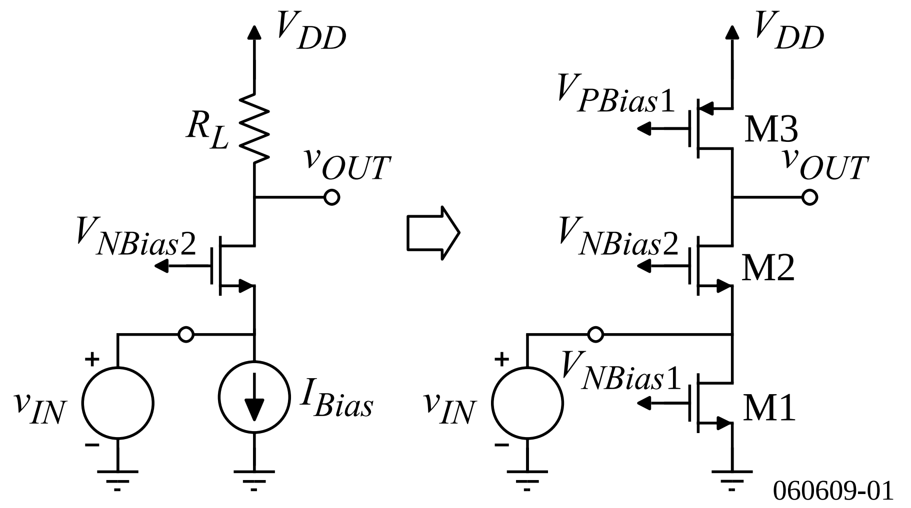
***CMOS Analog Circuit Design*, 3rd Edition Reference**

Pages 218-236

VOLTAGE-DRIVEN COMMON GATE AMPLIFIER

Common Gate Amplifier

Circuit:

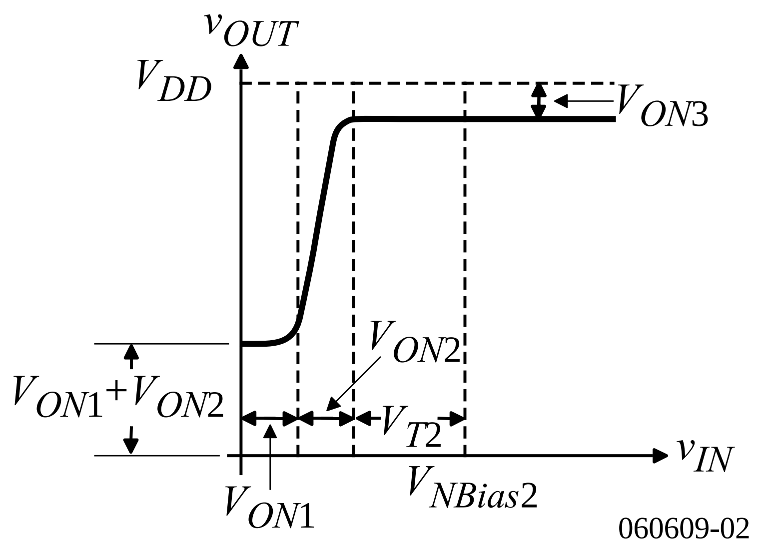


Large Signal Characteristics:

$V_{OUT}(\text{max}) \approx V_{DD} - V_{DS3}(\text{sat})$

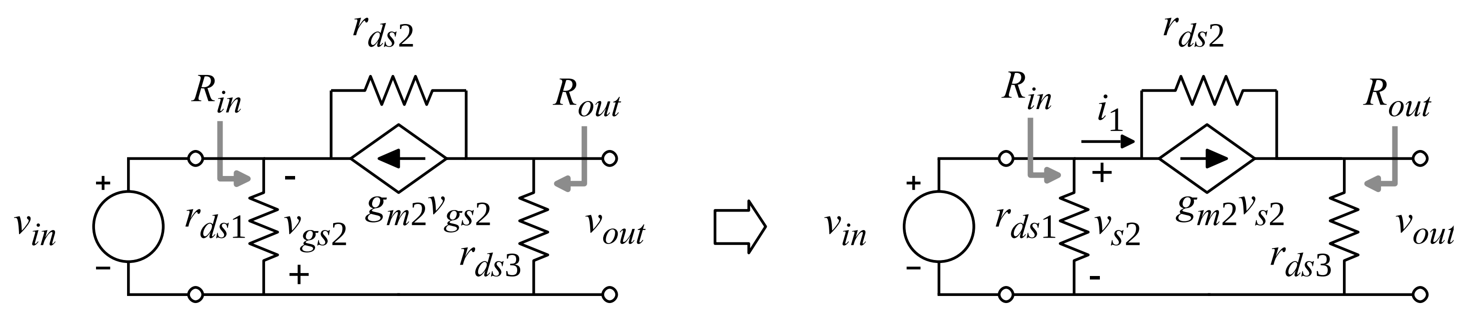
$V_{OUT}(\text{min}) \approx V_{DS1}(\text{sat}) + V_{DS2}(\text{sat})$

Note $V_{DS1}(\text{sat}) = V_{ON1}$



Small Signal Performance of the Common Gate Amplifier

Small signal model:



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$$v_{out} = g_{m2}v_{s2}\left(\frac{r_{ds2}}{r_{ds2}+r_{ds3}}\right)r_{ds3} = \left(\frac{g_{m2}r_{ds2}r_{ds3}}{r_{ds2}+r_{ds3}}\right)v_{in} \quad\rightarrow\quad \boxed{A_v = \frac{v_{out}}{v_{in}} = +\frac{g_{m2}r_{ds2}r_{ds3}}{r_{ds2}+r_{ds3}}}$$

$R_{in} = R_{in}'||r_{ds1}$, R_{in}' is found as follows

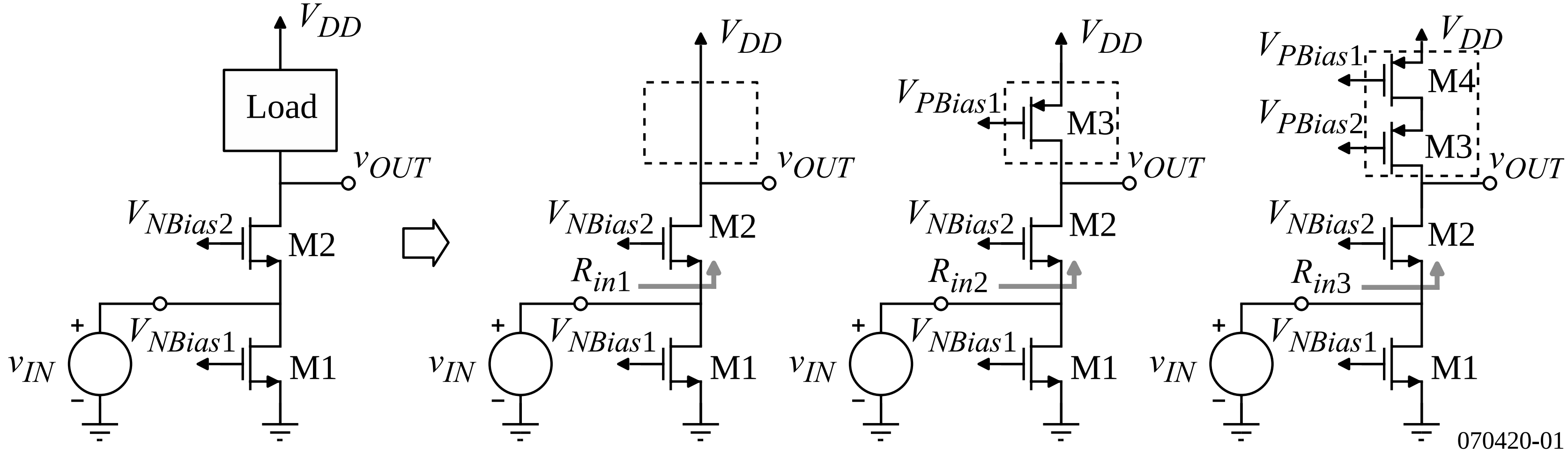
$$v_{s2} = (i_1 - g_{m2}v_{s2})r_{ds2} + i_1r_{ds3} = i_1(r_{ds2} + r_{ds3}) - g_{m2}\, r_{ds2}v_{s2}$$

$$R_{in}' = \frac{v_{s2}}{i_1} = \frac{r_{ds2} + r_{ds3}}{1 + g_{m2}r_{ds2}} \quad\rightarrow\quad \boxed{R_{in} = r_{ds1}||\frac{r_{ds2} + r_{ds3}}{1 + g_{m2}r_{ds2}}}$$

$$\boxed{R_{out} \approx r_{ds2}||r_{ds3}}$$

Influence of the Load on the Input Resistance of a Common Gate Amplifier

Consider a common gate amplifier with a general load:



From the previous page, the input resistance to the common gate configuration is,

$$R_{in} = \frac{r_{ds2} + R_{Load}}{1 + g_{m2}r_{ds2}}$$

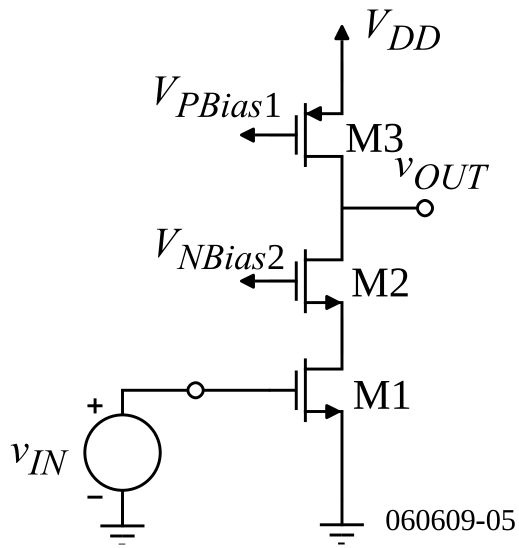
For the various loads shown, R_{in} becomes:

$$R_{in1} = \frac{r_{ds2}}{1+g_{m2}r_{ds2}} \approx \frac{1}{g_{m2}} \qquad R_{in2} = \frac{r_{ds2}+r_{ds3}}{1+ g_{m2}r_{ds2}} \approx \frac{2}{g_{m2}} \qquad R_{in3} = \frac{r_{ds2}+r_{ds4}g_{m3}r_{ds3}}{1+ g_{m2}r_{ds2}} \approx r_{ds}!!!$$

∴The input resistance of the common gate configuration depends on the load at the drain.

VOLTAGE-DRIVEN CASCODE AMPLIFIER

Cascode[†] Amplifier



Advantages of the cascode amplifier:

- Increases the output resistance and gain (if M3 is cascoded also)
- Eliminates the Miller effect when the input source resistance is large

[†] "Cascode" = "*Casc*aded tri*ode*" see H. Wallman, A.B. Macnee, and C.P. Gadsden, "A Low-Noise Amplifier, *Proc. IRE*, vol. 36, pp. 700-708, June 1948.

Large-Signal Characteristics of the Cascode Amplifier

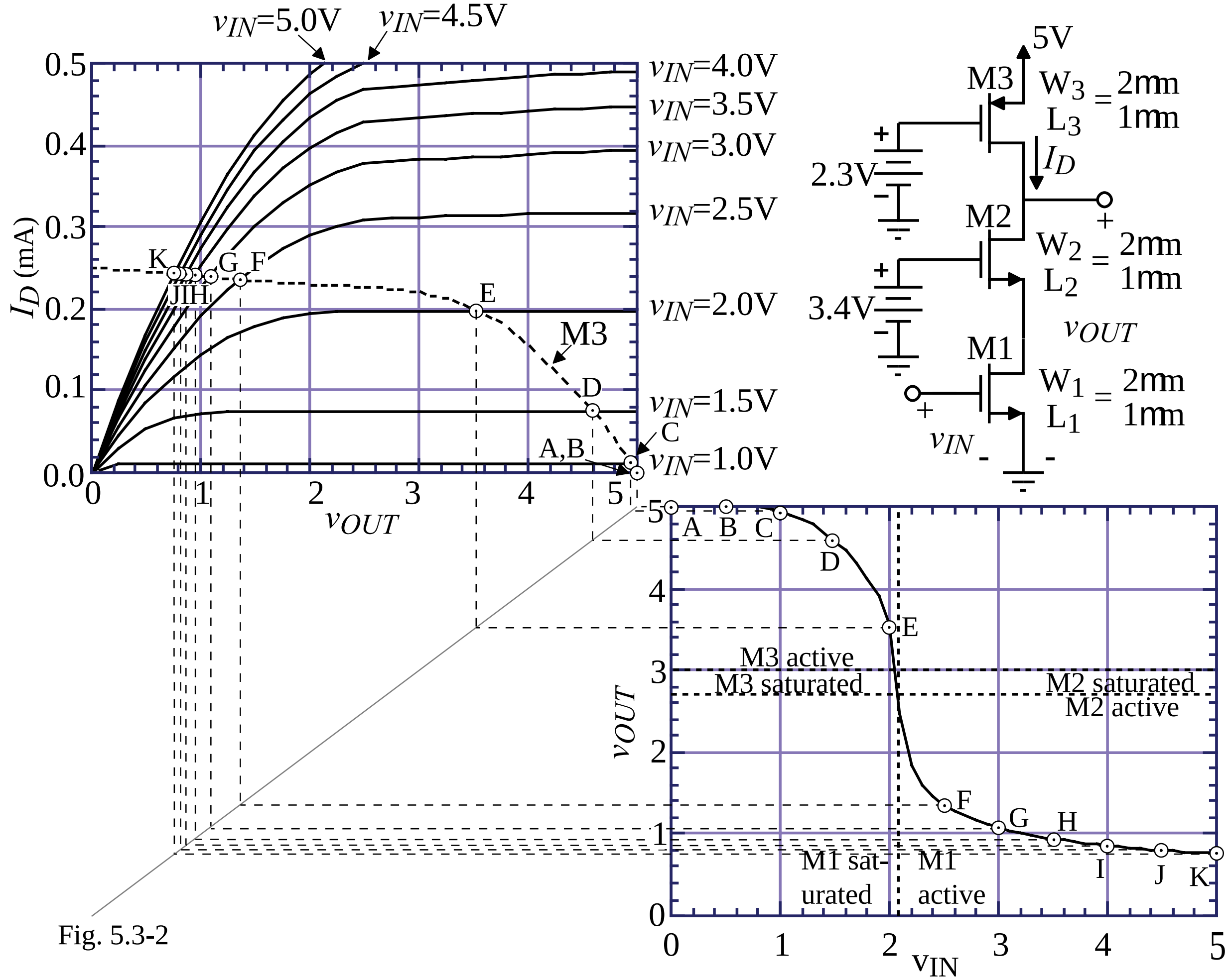


Fig. 5.3-2

M1 sat. when V_{GG2}-$V_{GS2} \geq V_{GS1}$-V_T $\rightarrow$ $v_{IN} \leq 0.5(V_{GG2}+V_{TN})$ where $V_{GS1}=V_{GS2}$

M2 sat. when $V_{DS2} \geq V_{GS2}-V_{TN} \rightarrow v_{OUT}-V_{DS1} \geq V_{GG2}-V_{DS1}-V_{TN} \rightarrow v_{OUT} \geq V_{GG2}-V_{TN}$

M3 is saturated when $V_{DD}-v_{OUT} \geq V_{DD} - V_{GG3} - |V_{TP}|$ $\rightarrow$ $v_{OUT} \leq V_{GG3} + |V_{TP}|$

Large-Signal Voltage Swing Limits of the Cascode Amplifier

Maximum output voltage, v_{OUT}(max):

$$v_{OUT}(\max) = V_{DD}$$

Minimum output voltage, v_{OUT}(min):

Referencing all potentials to the negative power supply (ground in this case), we may express the current through each of the devices, M1 through M3, as

$$i_{D1} = \beta_1 \left((V_{DD} - V_{T1})v_{DS1} - \frac{v_{DS1}{}^2}{2} \right) \approx \beta_1(V_{DD} - V_{T1})v_{DS1}$$

$$i_{D2} = \beta_2 \left((V_{GG2} - v_{DS1} - V_{T2})(v_{OUT} - v_{DS1}) - \frac{(v_{OUT} - v_{DS1})^2}{2} \right)$$

$$\cong \beta_2(V_{GG2} - v_{DS1} - V_{T2})(v_{OUT} - v_{DS1})$$

and

$$i_{D3} = \frac{\beta_3}{2}(V_{DD} - V_{GG3} - |V_{T3}|)^2$$

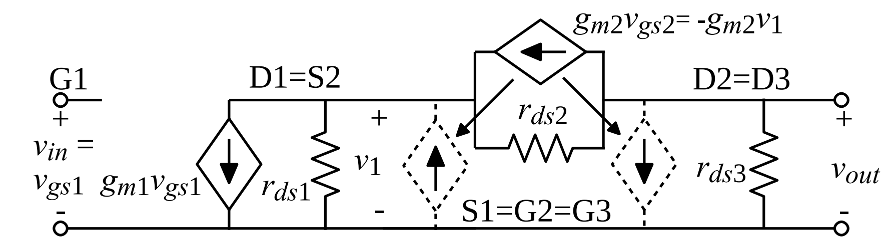
where we have also assumed that both v_{DS1} and v_{OUT} are small, and $v_{IN} = V_{DD}$.

Solving for v_{OUT} by realizing that $i_{D1} = i_{D2} = i_{D3}$ and $\beta_1 = \beta_2$ we get,

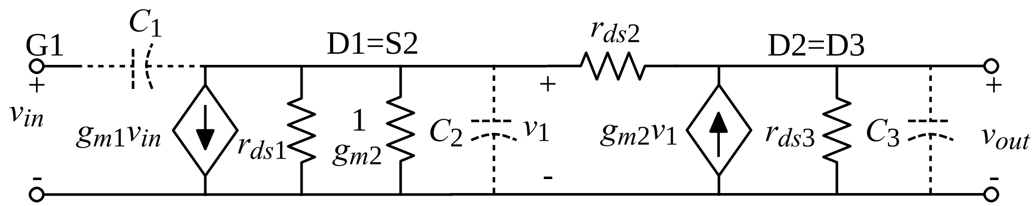
$$v_{OUT}(\min) = \frac{\beta_3}{2\beta_2}(V_{DD} - V_{GG3} - |V_{T3}|)^2 \left(\frac{1}{V_{GG2} - V_{T2}} + \frac{1}{V_{DD} - V_{T1}} \right)$$

Small-Signal Midband Performance of the Cascode Amplifier

Small-signal model:



Small-signal model of cascode amplifier neglecting the bulk ef fect on M2.



Simplified equivalent model of the above circuit. Fig. 5.3-3

Using nodal analysis, we can write,

$$[g_{ds1} + g_{ds2} + g_{m2}]v_1 - g_{ds2}v_{\text{out}} = -g_{m1}v_{\text{in}}$$

$$-[g_{ds2} + g_{m2}]v_1 + (g_{ds2} + g_{ds3})v_{\text{out}} = 0$$

Solving for $v_{\text{out}}/v_{\text{in}}$ yields

$$\frac{v_{\text{out}}}{v_{\text{in}}} = \frac{-g_{m1}(g_{ds2} + g_{m2})}{g_{ds1}g_{ds2} + g_{ds1}g_{ds3} + g_{ds2}g_{ds3} + g_{ds3}g_{m2}} \cong \frac{-g_{m1}}{g_{ds3}} = -\sqrt{\frac{2K'_1W_1}{L_1I_D\lambda^2{}_3}}$$

The small-signal output resistance is,

$$r_{out} = [r_{ds1} + r_{ds2} + g_{m2}r_{ds1}r_{ds2}]||r_{ds3} \cong r_{ds3}$$

Frequency Response of the Cascode Amplifier

Small-signal model ($R_S = 0$):

where

$C_1 = C_{gd1}$,

$C_2 = C_{bd1}+C_{bs2}+C_{gs2}$, and

$C_3 = C_{bd2}+C_{bd3}+C_{gd2}+C_{gd3}+C_L$

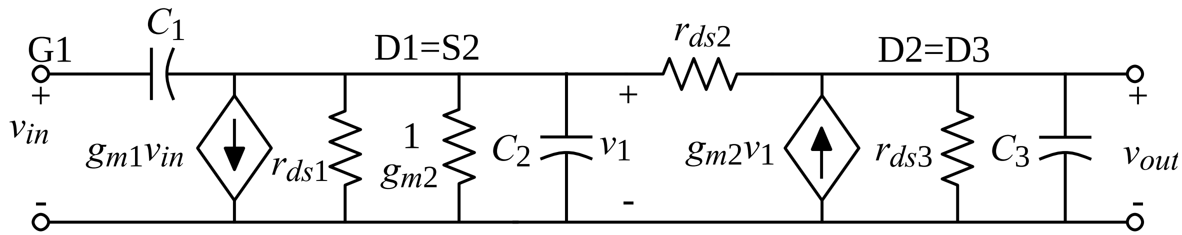


Fig. 5.3-4A

The nodal equations now become:

$$(g_{m2} + g_{ds1} + g_{ds2} + sC_1 + sC_2)v_1 - g_{ds2}v_{\text{out}} = -(g_{m1} - sC_1)v_{\text{in}}$$

and

$$-(g_{ds2} + g_{m2})v_1 + (g_{ds2} + g_{ds3} + sC_3)v_{\text{out}} = 0$$

Solving for $V_{\text{out}}(s)/V_{\text{in}}(s)$ gives,

$$\frac{V_{\text{out}}(s)}{V_{\text{in}}(s)} = \left(\frac{1}{1 + as + bs^2}\right)\left(\frac{-(g_{m1} - sC_1)(g_{ds2} + g_{m2})}{g_{ds1}g_{ds2} + g_{ds3}(g_{m2} + g_{ds1} + g_{ds2})}\right)$$

where

$$a = \frac{C_3(g_{ds1} + g_{ds2} + g_{m2}) + C_2(g_{ds2} + g_{ds3}) + C_1(g_{ds2} + g_{ds3})}{g_{ds1}g_{ds2} + g_{ds3}(g_{m2} + g_{ds1} + g_{ds2})}$$

and

$$b = \frac{C_3(C_1 + C_2)}{g_{ds1}g_{ds2} + g_{ds3}(g_{m2} + g_{ds1} + g_{ds2})}$$

A Simplified Method of Finding an Algebraic Expression for the Two Poles

Assume that a general second-order polynomial can be written as:

$$P(s) = 1 + as + bs^2 = \left(1 - \frac{s}{p_1}\right)\left(1 - \frac{s}{p_2}\right) \quad = 1 - s\left(\frac{1}{p_1} + \frac{1}{p_2}\right) + \frac{s^2}{p_1 p_2}$$

Now if $|p_2| >> |p_1|$, then $P(s)$ can be simplified as

$$P(s) \approx 1 - \frac{s}{p_1} + \frac{s^2}{p_1 p_2}$$

Therefore we may write p_1 and p_2 in terms of a and b as

$$p_1 = \frac{-1}{a} \text{ and } p_2 = \frac{-a}{b}$$

Applying this to the previous problem gives,

$$p_1 = \frac{-[g_{ds1}g_{ds2} + g_{ds3}(g_{m2} + g_{ds1} + g_{ds2})]}{C_3(g_{ds1} + g_{ds2} + g_{m2}) + C_2(g_{ds2} + g_{ds3}) + C_1(g_{ds2} + g_{ds3})} \approx \frac{-g_{ds3}}{C_3}$$

The nondominant root p_2 is given as

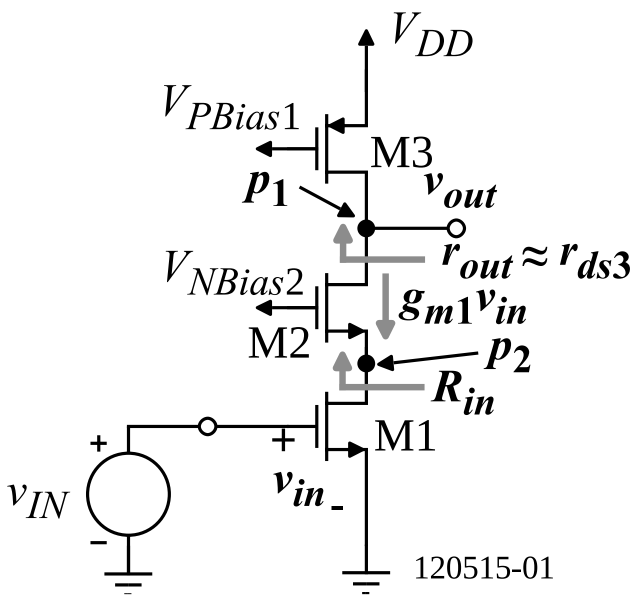
$$p_2 = \frac{-[C_3(g_{ds1} + g_{ds2} + g_{m2}) + C_2(g_{ds2} + g_{ds3}) + C_1(g_{ds2} + g_{ds3})]}{C_3(C_1 + C_2)} \approx \frac{-g_{m2}}{C_1 + C_2}$$

Assuming C_1, C_2, and C_3 are the same order of magnitude, and g_{m2} is greater than g_{ds3}, then $|p_1|$ is smaller than $|p_2|$. Therefore the approximation of $|p_2| >> |p_1|$ is valid.

Note that there is a right-half plane zero at $z_1 = g_{m1}/C_1$.

Repeating the Previous Example Using Intuitive Approach

Circuit:



Gain:

$$v_{out} \approx (-g_{m1}v_{in})\, r_{ds3} \quad \Rightarrow \quad \frac{v_{out}}{v_{in}} \approx -g_{m1}r_{ds3}$$

Poles:

1.) Dominant pole (one with the largest resistance to ground):

$$p_1 \approx \frac{-1}{r_{ds3}C_3}$$

2.) Next dominant pole is $p_2 \approx \dfrac{-1}{R_{in}(C_1+C_2)}$

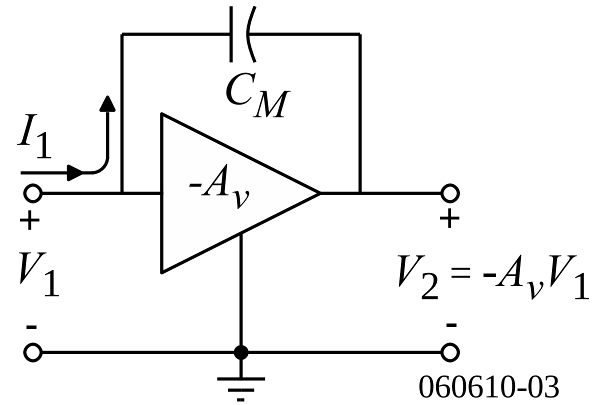
However, in this case, p_1 has already shorted the output to ground so that R_{in} is $\approx \dfrac{1}{g_{m2}}$ rather than $\approx \dfrac{2}{g_{m2}}$. Thus, $p_2 \approx \dfrac{-g_{m2}}{C_1+C_2}$.

Much easier!!!

NON-VOLTAGE DRIVEN CASCODE AMPLIFIER – THE MILLER EFFECT

Miller Effect

Consider the following inverting amplifier:



Solve for the input impedance:

$$Z_{in}(s) = \frac{V_1}{I_1}$$

$$I_1 = sC_M(V_1 - V_2) = sC_M(V_1 + A_vV_1) = sC_M(1 + A_v)V_1$$

Therefore,

$$Z_{in}(s) = \frac{V_1}{I_1} = \frac{V_1}{sC_M(1 + A_v)V_1} = \frac{1}{sC_M(1 + A_v)} = \frac{1}{sC_{eq}}$$

The Miller effect can take C_{gd} = 5fF and make it look like a 0.5pF capacitor in parallel with the input of the inverting amplifier ($A_v \approx$ -100).

If the source resistance is large, this creates a dominant pole at the input.

Simple Inverting Amplifier Driven with a High Source Resistance

Examine the frequency response of a current-source load inverter driven from a high resistance source:

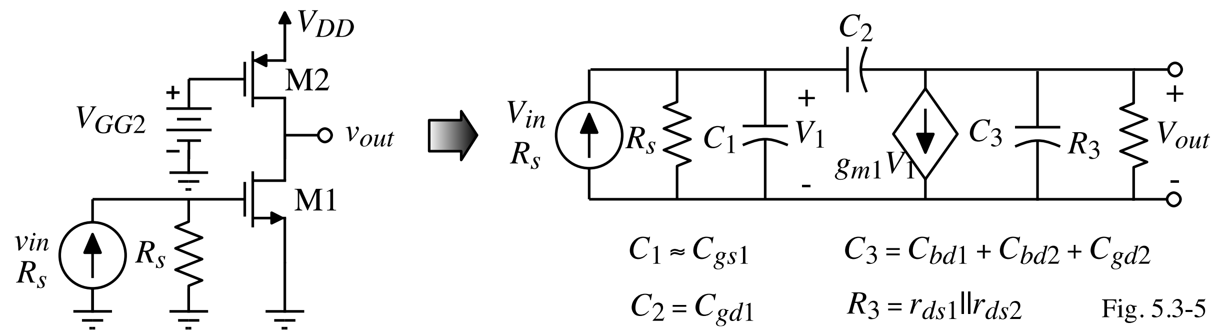


Fig. 5.3-5

Assuming the input is I_{in}, the nodal equations are,

$$[G_1 + s(C_1 + C_2)]V_1 - sC_2V_{out} = I_{in} \quad \text{and} \quad (g_{m1} - sC_2)V_1 + [G_3 + s(C_2 + C_3)]V_{out} = 0$$

where

$$G_1 = G_s\ (=1/R_s),\quad G_3 = g_{ds1} + g_{ds2},\quad C_1 = C_{gs1},\quad C_2 = C_{gd1} \quad \text{and } C_3 = C_{bd1} + C_{bd2} + C_{gd2}.$$

Solving for $V_{out}(s)/V_{in}(s)$ gives

$$\frac{V_{out}(s)}{V_{in}(s)} = \frac{(sC_2 - g_{m1})G_1}{G_1G_3 + s[G_3(C_1+C_2) + G_1(C_2+C_3) + g_{m1}C_2] + (C_1C_2 + C_1C_3 + C_2C_3)s^2} \quad \text{or,}$$

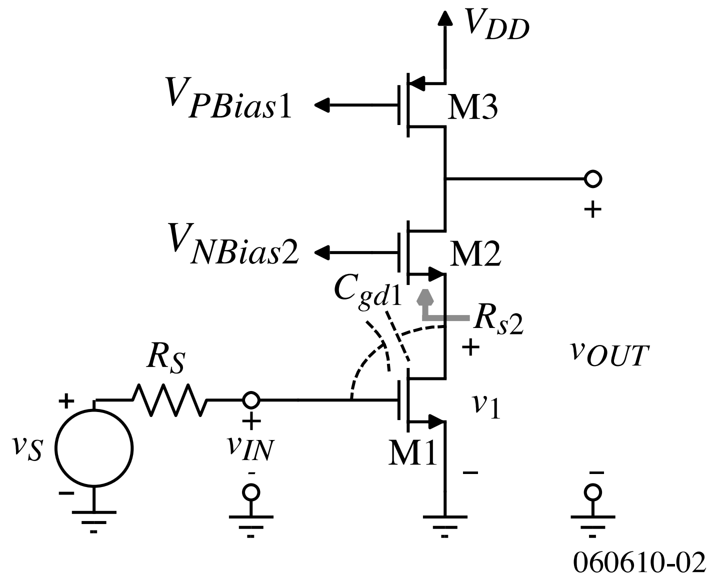
$$\frac{V_{out}(s)}{V_{in}(s)} = \left(\frac{-g_{m1}}{G_3}\right) \frac{[1 - s(C_2/g_{m1})]}{1 + [R_1(C_1+C_2) + R_3(C_2+C_3) + g_{m1}R_1R_3C_2]s + (C_1C_2 + C_1C_3 + C_2C_3)R_1R_3s^2}$$

Assuming that the poles are split allows the use of the previous technique to get,

$$p_1 = \frac{-1}{R_1(C_1+C_2) + R_3(C_2+C_3) + g_{m1}R_1R_3C_2} \cong \frac{-1}{g_{m1}R_1R_3C_2} \quad \text{and} \quad p_2 \cong \frac{-g_{m1}C_2}{C_1C_2 + C_1C_3 + C_2C_3}$$

How Does the Cascode Amplifier Solve the Miller Effect?

Cascode amplifier:



The Miller effect causes C_{gs1} to be increased by the value of $1 + (v_1/v_{in})$ and appear in parallel with the gate-source of M1 causing a dominant pole to occur.

The cascode amplifier eliminates this problem by keeping the value of v_1/v_{in} small by making the value of R_{s2} approximately $2/g_{m2}$.

Comparison of the Inverting and Cascode Non-Voltage Driven Amplifiers

The dominant pole of the inverting amplifier with a large source resistance was found to be

$$p_1(\text{inverter}) = \frac{-1}{R_1(C_1+C_2)+R_3(C_2+C_3)+g_{m1}R_1R_3C_2} \approx \frac{-1}{g_{m1}R_1R_3C_2}$$

Now if a cascode amplifier is used, R_3, can be approximated as $2/g_m$ of the cascoding transistor (assuming the drain sees an r_{ds} to ac ground).

$$\therefore \quad p_1(\text{cascode}) = \frac{-1}{R_1(C_1+C_2)+\left(\frac{2}{g_m}\right)(C_2+C_3)+g_{m1}R_1\left(\frac{2}{g_m}\right)C_2}$$

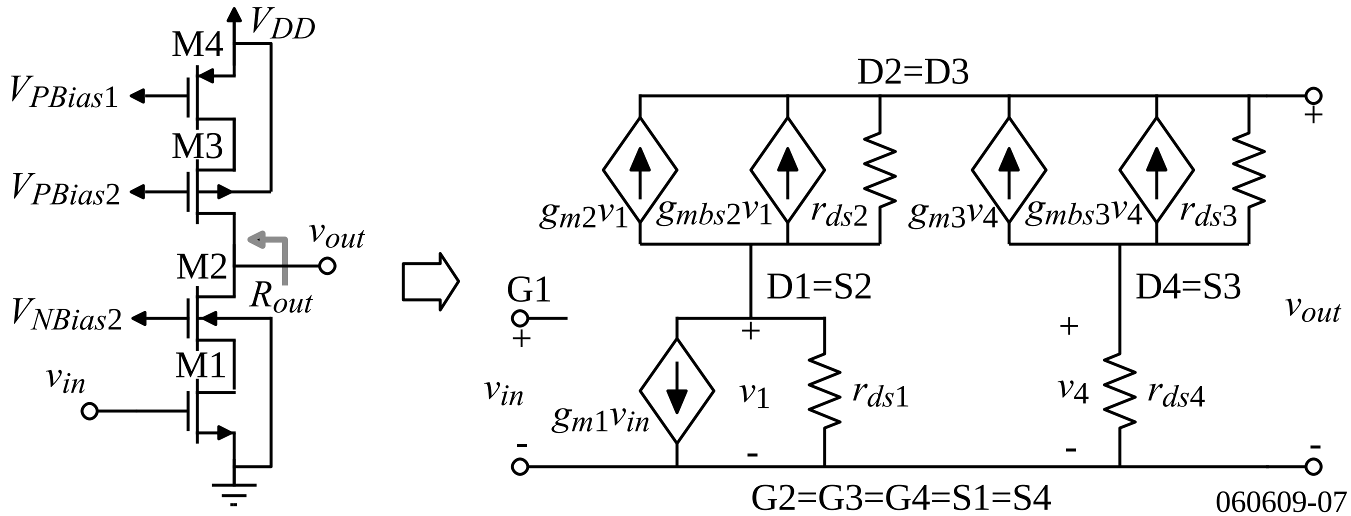
$$= \frac{-1}{R_1(C_1+C_2)+\left(\frac{2}{g_m}\right)(C_2+C_3)+2R_1C_2} \approx \frac{-1}{R_1(C_1+3C_2)}$$

Thus we see that p_1(cascode) >> p_1(inverter).

FURTHER CONSIDERATIONS OF CASCODE AMPLIFIERS

High Gain and High Output Resistance Cascode Amplifier

If the load of the cascode amplifier is a cascode current source, then both high output resistance and high voltage gain is achieved.



The output resistance is,

$$r_{out} \cong [g_{m2}r_{ds1}r_{ds2}] \,||\, [g_{m3}r_{ds3}r_{ds4}] = \frac{I_D^{-1.5}}{\dfrac{\lambda_1\lambda_2}{\sqrt{2K'_2(W/L)_2}} + \dfrac{\lambda_3\lambda_4}{\sqrt{2K'_3(W/L)_3}}}$$

Knowing r_{out}, the gain is simply

$$A_v = -g_{m1}r_{out} \cong -g_{m1}\{[g_{m2}r_{ds1}r_{ds2}] \,||\, [g_{m3}r_{ds3}r_{ds4}]\} \cong \frac{\sqrt{2K'_1(W/L)_1}\,I_D^{-1}}{\dfrac{\lambda_1\lambda_2}{\sqrt{2K'_2(W/L)_2}} + \dfrac{\lambda_3\lambda_4}{\sqrt{2K'_3(W/L)_3}}}$$

Example 20-1 - Comparison of the Cascode Amplifier Performance

Calculate the small-signal voltage gain, output resistance, the dominant pole, and the nondominant pole for the low-gain, cascode amplifier and the high-gain, cascode amplifier. Assume that I_D = 200 microamperes, that all W/L ratios are 2µm/1µm, and that the parameters of Table 3.1-2 are valid. The capacitors are assumed to be: C_{gd} = 3.5 fF, C_{gs} = 30 fF, $C_{bsn} = C_{bdn}$ = 24 fF, $C_{bsp} = C_{bdp}$ = 12 fF, and C_L = 1 pF.

Solution

The low-gain, cascode amplifier has the following small-signal performance (no upper cascode, just lower cascode):

$A_v = -37.1\text{V/V}$ $\qquad R_{out} = 125\text{k}\Omega$

$p_1 \approx -g_{ds3}/C_3 \rightarrow 1.22\text{ MHz}$ $\qquad p_2 \approx -g_{m2}/(C_1+C_2) \rightarrow 605\text{ MHz}$.

The high-gain, cascode amplifier has the following small-signal performance (with upper and lower cascode):

$A_v = -414\text{V/V}$ $\qquad R_{out} = 1.40\text{ M}\Omega$

$p_1 \approx -1/R_{out}C_3 \rightarrow 108\text{ kHz}$ $\qquad p_2 \approx -g_{m2}/(C_1+C_2) \rightarrow 579\text{ MHz}$

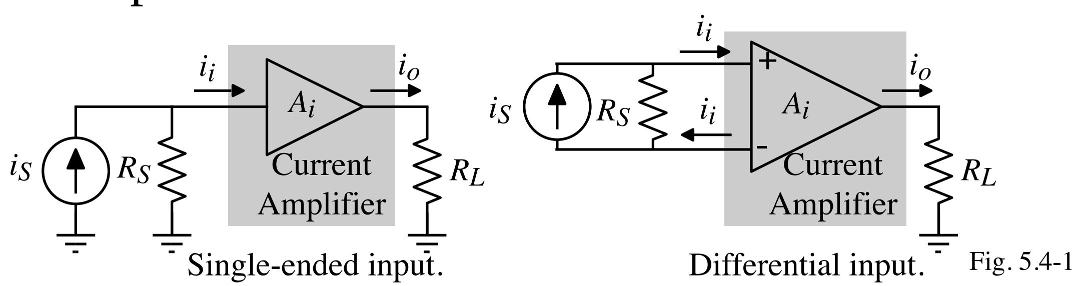
(Note at this frequency, the drain of M2 is shorted to ground by the load capacitance, C_L)

CURRENT AMPLIFIERS

What is a Current Amplifier?

- An amplifier that has a defined output-input current relationship
- Low input resistance
- High output resistance

Application of current amplifiers:



$R_S >> R_{in}$ and $R_{out} >> R_L$

Advantages of current amplifiers:

- Currents are not restricted by the power supply voltages so that wider dynamic ranges are possible with lower power supply voltages.
- -3dB bandwidth of a current amplifier using negative feedback is independent of the closed loop gain.

Frequency Response of a Current Amplifier with Current Feedback

Consider the following current amplifier with resistive negative feedback applied.

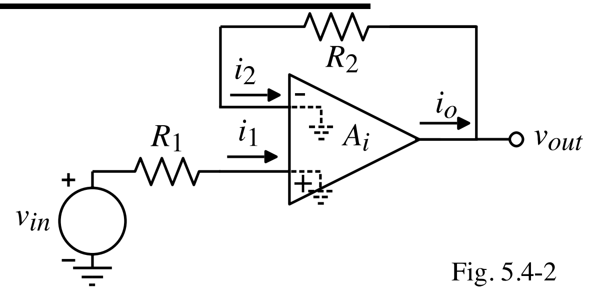


Fig. 5.4-2

Assuming that the small-signal resistance looking into the current amplifier is much less than R_1 or R_2,

$$i_o = A_i(i_1 - i_2) = A_i\left(\frac{v_{in}}{R_1} - i_o\right)$$

Solving for i_o gives

$$i_o = \left(\frac{A_i}{1+A_i}\right)\frac{v_{in}}{R_1} \quad \rightarrow \quad v_{out} = R_2 i_o = \frac{R_2}{R_1}\left(\frac{A_i}{1+A_i}\right)v_{in}$$

If $A_i(s) = \dfrac{A_o}{\dfrac{s}{\omega_A} + 1}$, then

$$\frac{v_{out}}{v_{in}} = \frac{R_2}{R_1}\left(\frac{1}{1+\dfrac{1}{A_i(s)}}\right) = \frac{R_2}{R_1}\left(\frac{A_o}{\dfrac{s}{\omega_A} + (1+A_o)}\right) = \frac{R_2}{R_1}\left(\frac{A_o}{1+A_o}\right)\left(\frac{1}{\dfrac{s}{\omega_A(1+A_o)} + 1}\right)$$

$$\therefore \quad \omega_{-3dB} = \omega_A(1+A_o)$$

Bandwidth Advantage of a Current Feedback Amplifier

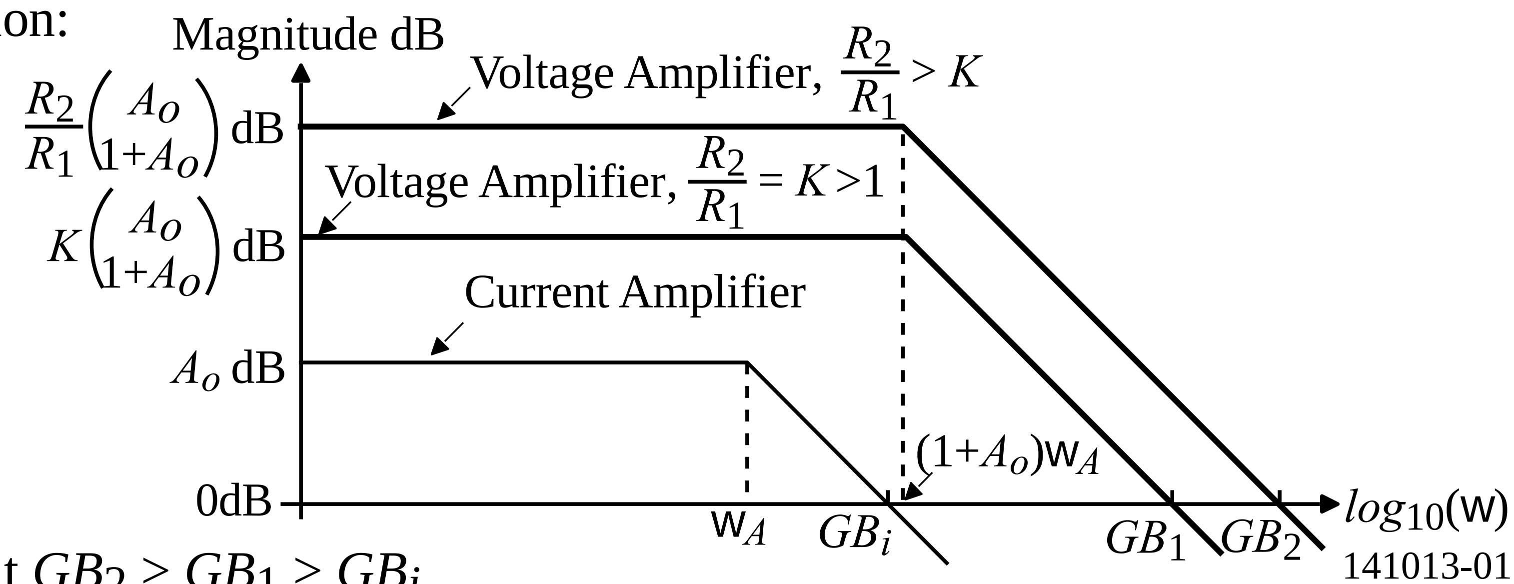
The unity-gainbandwidth is,

$$GB = |A_v(0)|\ \omega_{-3\text{dB}} = \frac{R_2 A_o}{R_1(1+A_o)} \cdot \omega_A(1+A_o) = \frac{R_2}{R_1} A_o \cdot \omega_A = \frac{R_2}{R_1} GB_i$$

where GB_i is the unity-gainbandwidth of the current amplifier.

Note that if GB_i is constant, then increasing R_2/R_1 (the voltage gain) increases GB.

Illustration:



Note that $GB_2 > GB_1 > GB_i$

The above illustration assumes that the *GB* of the voltage amplifier realizing the voltage buffer is greater than the *GB* achieved from the above method.

Current Amplifier using the Simple Current Mirror

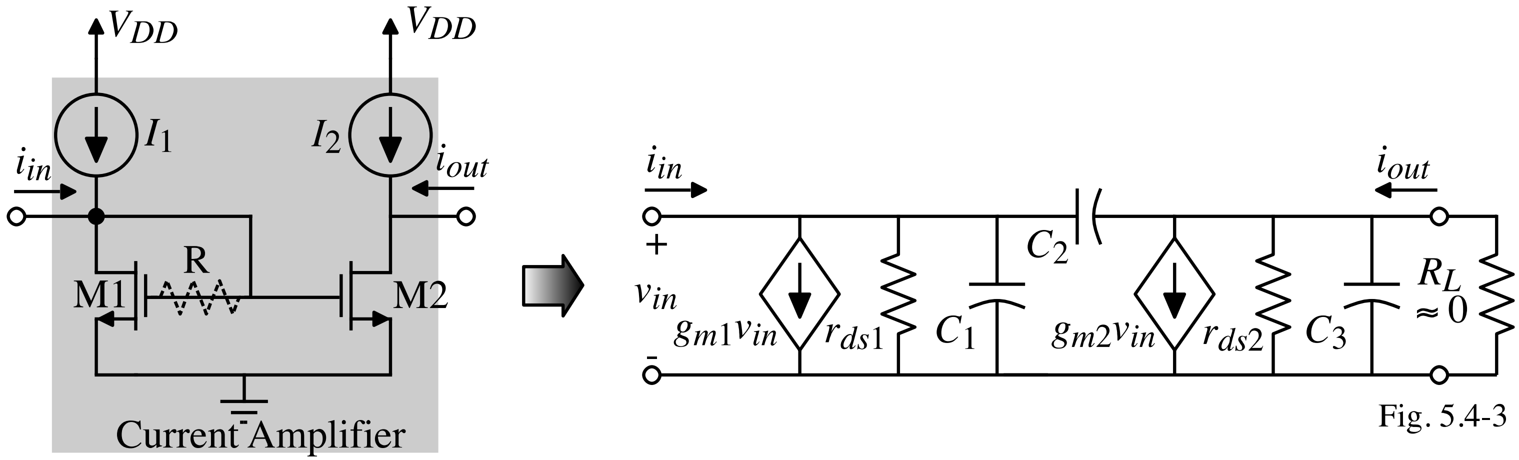


Fig. 5.4-3

$$R_{in} = \frac{1}{g_{m1}} \quad R_{out} = \frac{1}{\lambda_1 I_o} \quad \text{and} \quad A_i = \frac{W_2/L_2}{W_1/L_1} .$$

Frequency response:

$$p_1 = \frac{-(g_{m1}+g_{ds1})}{C_1+C_2} = \frac{-(g_{m1}+g_{ds1})}{C_{bd1}+C_{gs1}+C_{gs2}+C_{gd2}} \approx \frac{-g_{m1}}{C_{bd1}+C_{gs1}+C_{gs2}+C_{gd2}}$$

Note that the bandwidth can be almost doubled by including the resistor, R.

(R removes C_{gs1} from p_1)

Example 20-2 - Performance of a Simple Current Mirror as a Current Amplifier

Find the small-signal current gain, A_i, the input resistance, R_{in}, the output resistance, R_{out}, and the -3dB frequency in Hertz for the current amplifier of previous slide if $10I_1 = I_2 = 100\mu A$ and $W_2/L_2 = 10W_1/L_1 = 10\mu m/1\mu m$. Assume that $C_{bd1} = 10fF$, $C_{gs1} = C_{gs2} = 100fF$, and $C_{gd2} = 50fF$.

Solution

Ignoring channel modulation and mismatch effects, the small-signal current gain,

$$A_i = \frac{W_2/L_2}{W_1/L_1} \approx 10\text{A/A}.$$

The small-signal input resistance, R_{in}, is approximately $1/g_{m1}$ and is

$$R_{in} \approx \frac{1}{\sqrt{2K_N(1/1)10\mu A}} = \frac{1}{46.9\mu S} = 21.3k\Omega$$

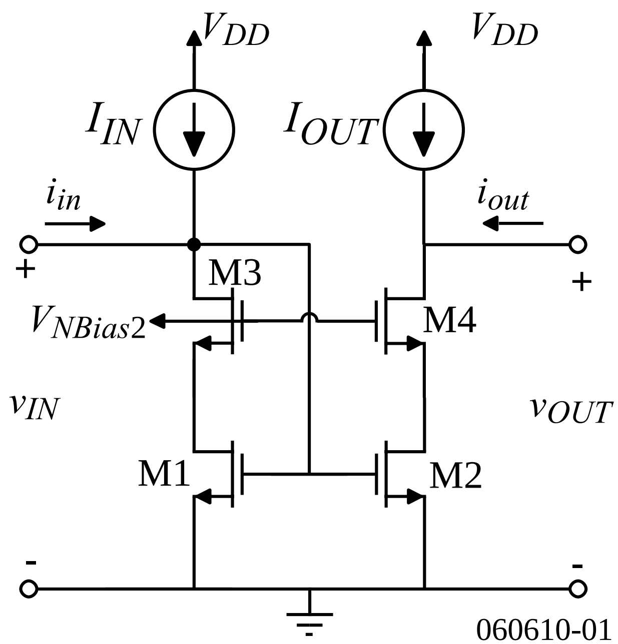
The small-signal output resistance is equal to

$$R_{out} = \frac{1}{\lambda_N I_2} = 250k\Omega.$$

The -3dB frequency is

$$\omega_{-3dB} = \frac{46.9\mu S}{260fF} = 180.4\text{x}10^6 \text{ radians/sec.} \quad \rightarrow \quad f_{-3dB} = 28.7 \text{ MHz}$$

Wide-Swing, Cascode Current Mirror Implementation of a Current Amplifier



$$R_{in} \approx \frac{1}{g_{m1}}, \quad R_{out} \approx r_{ds2} g_{m4} r_{ds4}, \quad \text{and} \quad A_i = \frac{W_2/L_2}{W_1/L_1}$$

Example 20-3 - Current Amplifier Implemented by the Wide-Swing, Cascode Current Mirror

Assume that I_{IN} and I_{OUT} of the wide-swing cascode current mirror are 100µA. Find the value of R_{in}, R_{out}, and A_i if the W/L ratios of all transistors are 182µm/1µm.

Solution

The input resistance requires g_{m1} which is $\sqrt{2\cdot 110\cdot 182\cdot 100} = 2\text{mS}$

$\therefore \quad R_{in} \approx 500\Omega$

From our knowledge of the cascode configuration, the small signal output resistance should be

$$R_{out} \approx g_{m4}r_{ds4}r_{ds2} = (2001\mu\text{S})(250\text{k}\Omega)(250\text{k}\Omega) = 125\text{M}\Omega$$

Because $V_{DS1} = V_{DS2}$, the small-signal current gain is

$$A_i = \frac{W_2/L_2}{W_1/L_1} = 1$$

Simulation results using the level 1 model for this example give

$R_{in} = 497\Omega$, $R_{out} = 164.7\text{M}\Omega$ and $A_i = 1.000$ A/A.

The value of V_{ON} for all transistors is

$$V_{ON} = \sqrt{\frac{2\cdot 100\mu\text{A}}{110\mu\text{A/V}^2\cdot 182}} = 0.1\text{V}$$

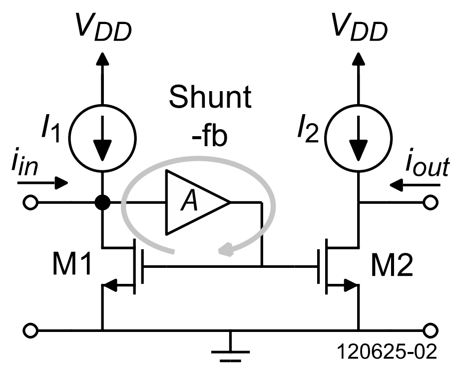
Low-Input Resistance Current Amplifier

To decrease R_{in} below $1/g_m$ requires feedback but what kind of feedback?

Consider Blackman's formulation for input resistance:

$$R_x = R_x(k{=}0)\left[\frac{1 + RR(\text{port shorted})}{1 + RR(\text{port opened})}\right]$$

Therefore, we want a configuration where the return ratio (RR) goes to zero when the port is shorted. We know that the shunt configuration shown below accomplishes this.



It is easy to see that the return ratio for the input shorted is zero and the return ratio for the input open is,

$$RR(\text{port opened}) = Ag_{m1}r_{ds1} \neq 0$$

Therefore based on these ideas, a low-input resistance realization is proposed on the next slide.

Low-Input Resistance Current Amplifier

Blackmann's formula:

Choosing g_{m1} as k, we see that,

$$R_x(k=0) = r_{ds1}$$

The circuits for calculating the shorted and open return-ratios are:

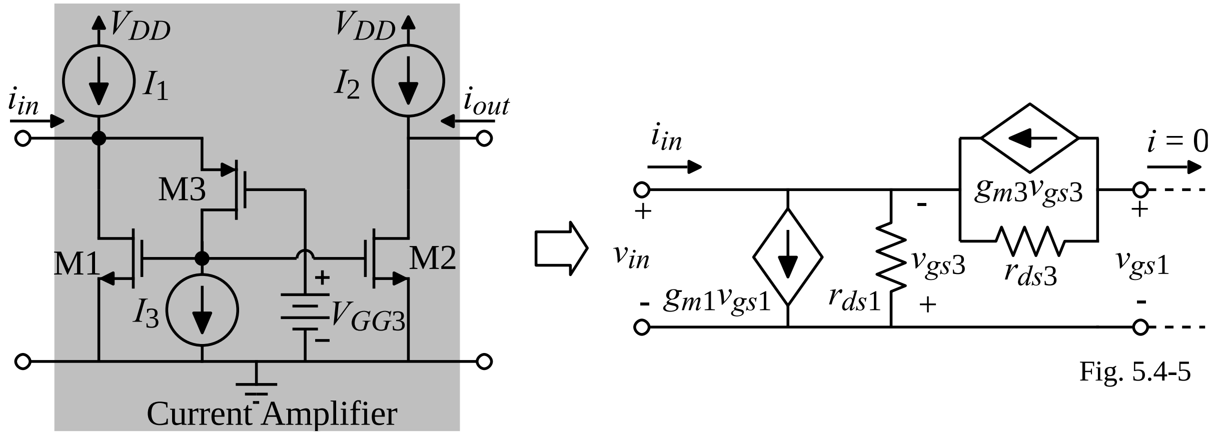
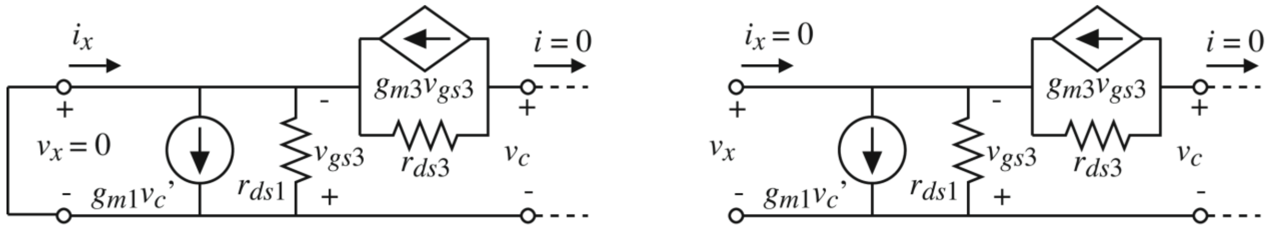


Fig. 5.4-5



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$$RR(v_x = 0):\ -\frac{v_c}{v_c'} = 0 \quad RR(i_x = 0):\ v_c = -\,v_{gs3}(1+ g_{m3}r_{ds3}) = -\,g_{m1}r_{ds1}\,(1+ g_{m3}r_{ds3})v_c'$$

$$\therefore \quad RR(i_x = 0) = -\frac{v_c}{v_c'} = g_{m1}r_{ds1}\,(1+ g_{m3}r_{ds3})$$

Finally, $R_x = R_{in} = r_{ds1}\dfrac{1+0}{1+g_{m1}r_{ds1}(1+ g_{m3}r_{ds3})} \approx \dfrac{1}{g_{m1}g_{m3}r_{ds3}}$

Small signal analysis gives the same result and is much easier to calculate.

Differential-Input, Current Amplifiers

Definitions for the differential-mode, i_{ID}, and common-mode, i_{IC}, input currents of the differential-input current amplifier.

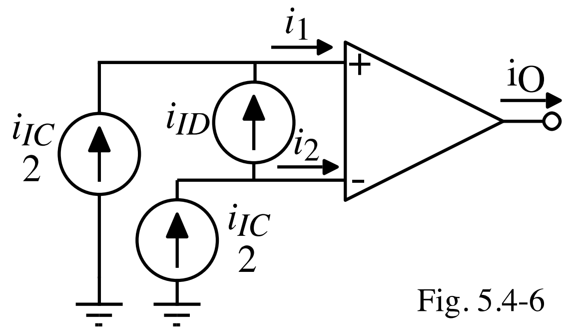
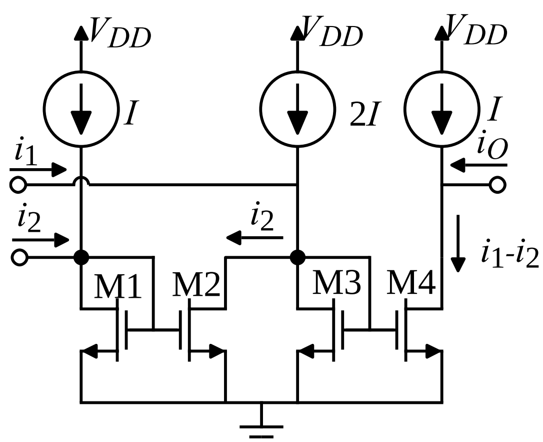


Fig. 5.4-6

$$i_o = A_{ID}i_{ID} \pm A_{IC}i_{IC} = A_{ID}(i_1 - i_2) \pm A_{IC}\left(\frac{i_1+i_2}{2}\right)$$

Implementations:



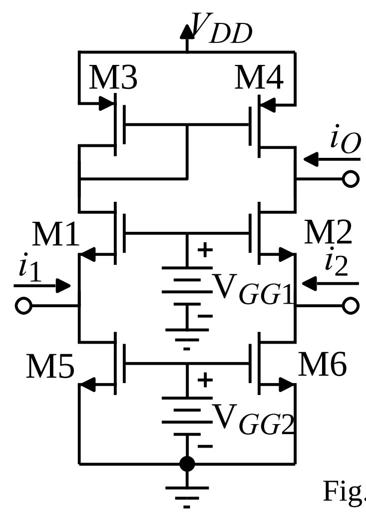


Fig. 5.4-7

SUMMARY

- Low input resistance amplifiers use the source as the input terminal with the gate generally on ground
- The input resistance to the common gate amplifier depends on what is connected to the drain
- The voltage driven common gate/common source amplifier has one dominant pole
- The current driven common gate/common source amplifier has two dominant poles
- The cascode amplifier eliminates the input dominant pole for the current driven common gate/common source amplifier
- Current amplifiers have a low input resistance, high output resistance, and a defined output-input current relationship
- Input resistances less than $1/g_m$ require feedback

  However, all feedback loops have internal poles that cause the benefits of negative feedback to vanish at high frequencies.

  In addition, feedback loops can have a slow time constant from a pole-zero pair.
- Voltage amplifiers using a current amplifier have high values of gain-bandwidth
- Current amplifiers are useful at low power supplies and for switched current applications

