Review of CMOS amplifiers

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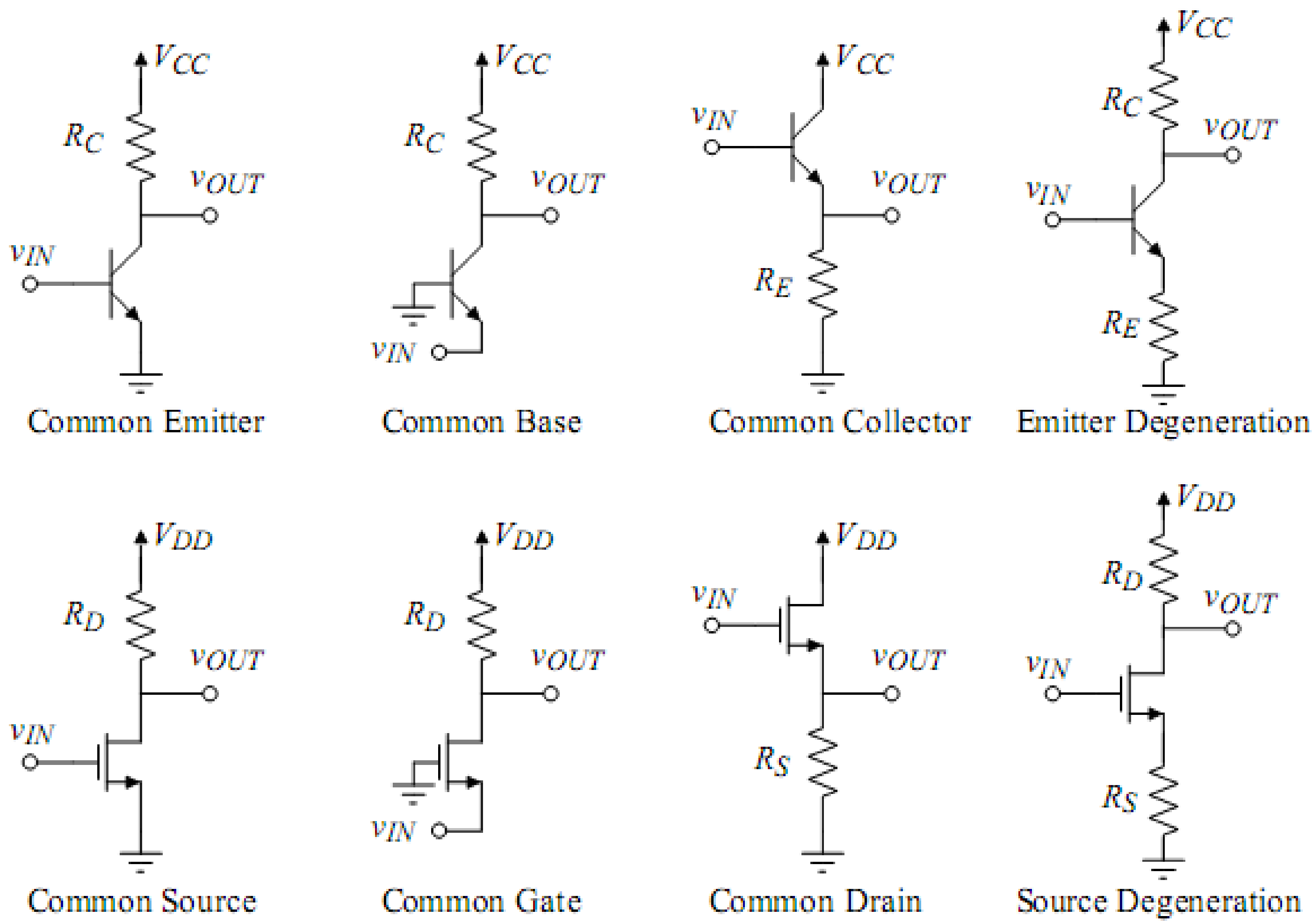
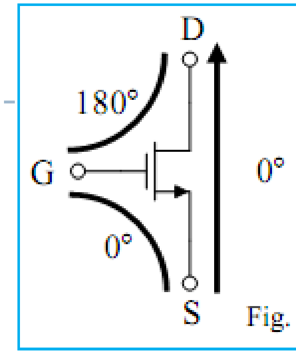
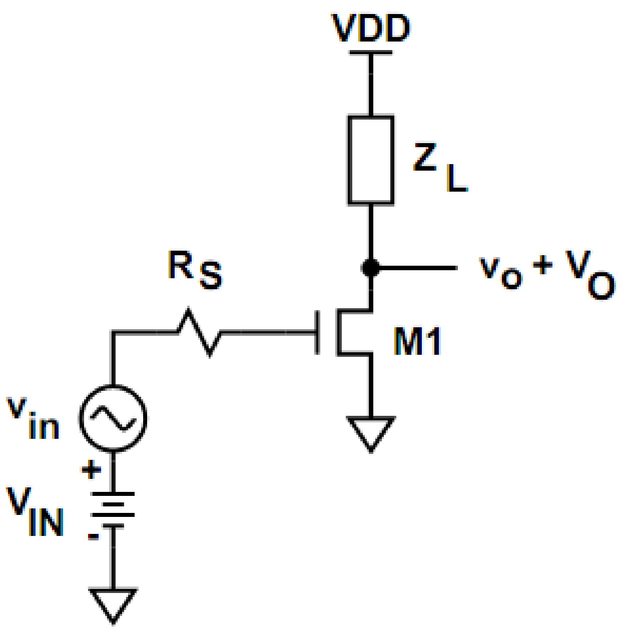
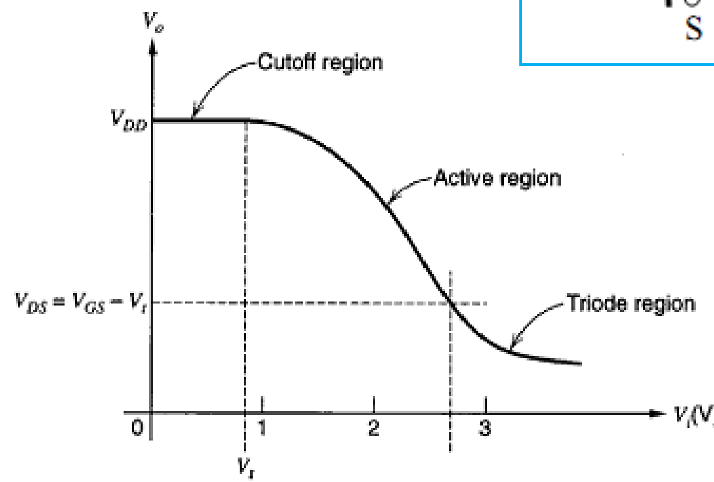
Basic CMOS amplifiers



Common-source amplifier



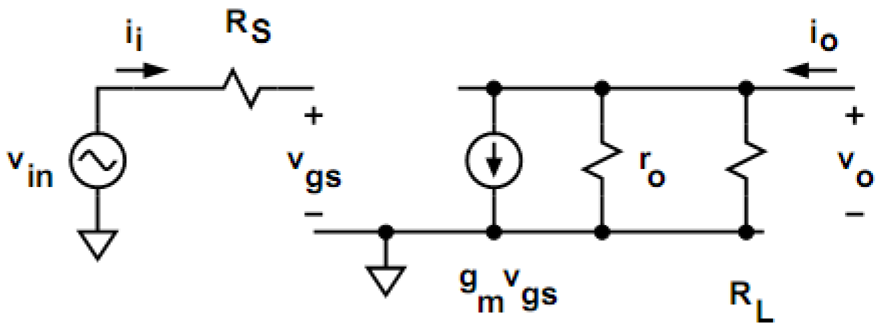


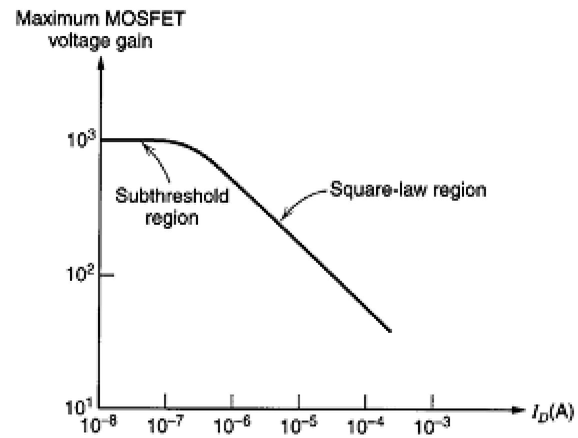


$$V_o = V_{DD} - I_d \cdot R_L = V_{DD} - \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_i - V_{tn})^2 \cdot R_L$$

- DC voltage V_I is chosen to bias M1 so that M1 is in active (saturation) region and its drain voltage is near the midpoint of the output swing ($V_O \approx V_{DD}/2$).

Common-source amplifier – small-signal analysis



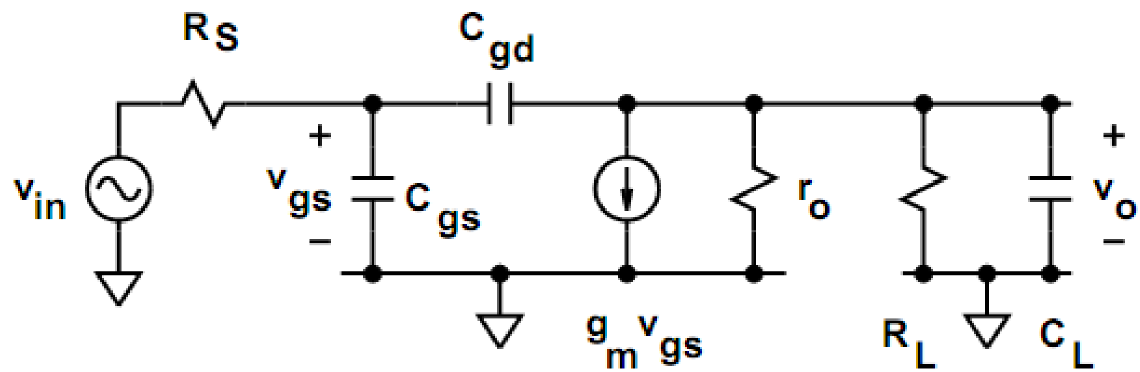


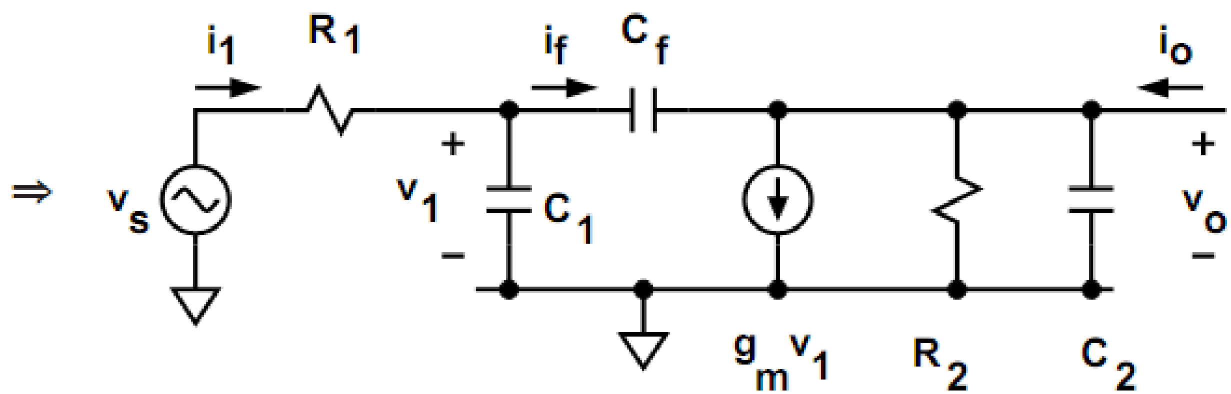
$$A_v(0) = -g_m(r_o \parallel R_L) = -g_m R'_L \qquad R_i = \infty \qquad R_o = r_o \parallel R_L = R'_L$$

- Note that for $R_L \to \infty$, $R'_L \to r_o$, and

$$|A_v(0)|_{max} = g_m r_o = \frac{I_D}{V_{ov}/2} \times \frac{L_{eff}}{I_D} \left| \frac{\partial L_{eff}}{\partial V_{DS}} \right|^{-1} = \frac{2L_{eff}}{V_{ov}} \left| \frac{\partial L_{eff}}{\partial V_{DS}} \right|^{-1}$$

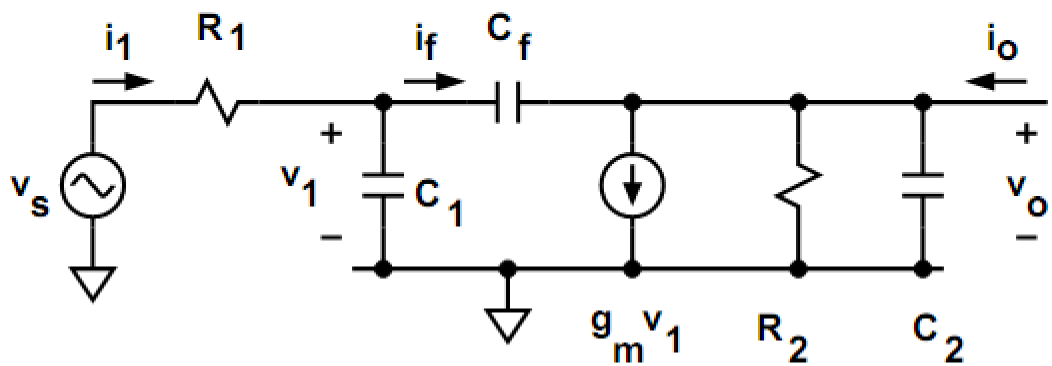
Common-source amplifier – small-signal analysis





$v_s = v_i \quad R_1 = R_S \quad R_2 = r_o \| R_L \quad C_1 = C_{gs} \quad C_2 = C_L \quad C_f = C_{gd}$

Miller approximation



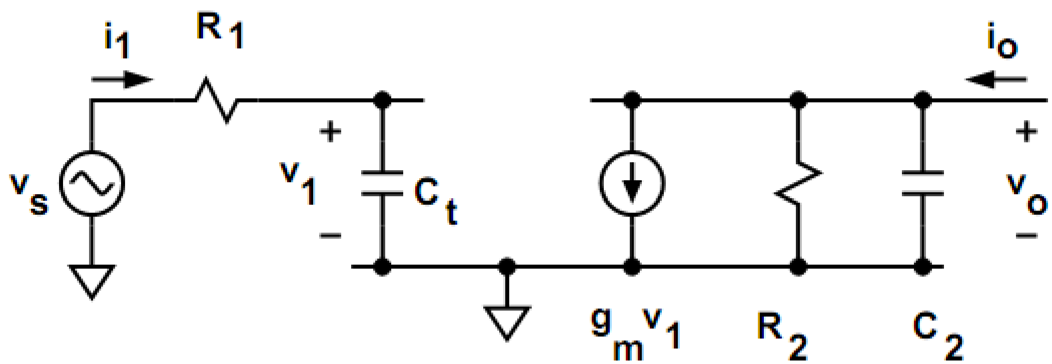
$$v_o = (-g_m v_1 + i_f)\left(R_2 \parallel \frac{1}{sC_2}\right) \qquad i_f = (v_1 - v_o)sC_f$$

If R_2-C_2 is a non-dominant pole, then, at the frequencies of interest

$$v_o \approx -g_m R_2 v_1 \qquad i_f \approx (v_1 + g_m R_2 v_1)sC_f \qquad \Rightarrow \qquad \frac{i_f}{v_i} = s(1 + g_m R_2)C_f = sC_M$$

$$C_M = (1 + g_m R_2)\cdot C_f = (1 + a_{v0})\cdot C_f = \text{Miller Capacitance}$$

Miller approximation

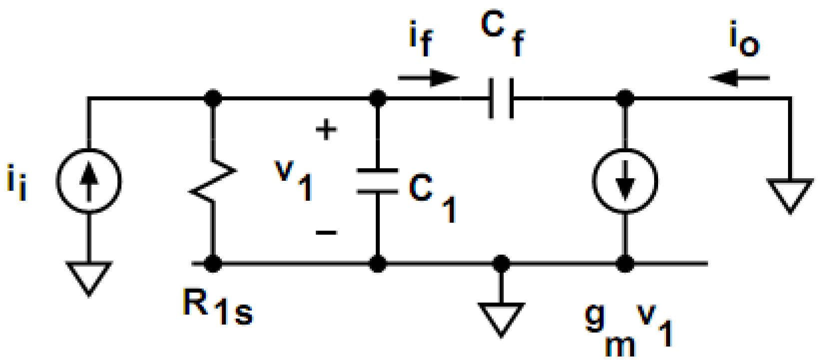


$$C_t = C_1 + C_M = C_1 + (1 + g_m R_2)C_f$$

$$A_v(s) = \frac{v_o}{v_s} = A_v(0)\frac{1}{(1 - s/p_1)(1 - s/p_2)}$$

$$A_v(0) = -g_m R_2 \qquad p_1 = \frac{1}{R_1 C_t} \qquad p_2 = \frac{1}{R_2 C_2}$$

MOSFET's Transition frequency



$$i_o \approx -g_m \times v_1 = -g_m \times i_1 \frac{R_{1s}}{1 + R_{1s}(C_1 + C_f)s}$$

$$\text{Short-Circuit Current Gain} = \beta(s) = -\frac{i_o}{i_i} = \frac{g_m R_{1s}}{1 + R_{1s}(C_1 + C_f)s}$$

$$\text{Transition Frequency} = \omega_T \approx \frac{g_m}{C_1 + C_f}$$

$$R_{1s} = \infty \qquad C_1 = C_{gs} = \frac{2}{3} C_{ox} W L \qquad C_f = C_{gd}$$

$$\omega_T = 2\pi f_T = \frac{g_m}{C_{gs} + C_{gd}}$$

MOSFET's Transition frequency

To calculate intrinsic device speed, let $\omega_T \approx g_m / C_{gs}$.

- For square-law device,

$$g_m = \mu C_{ox} \frac{W}{L} V_{ov} \quad \Rightarrow \quad \omega_T = \frac{3}{2} \cdot \frac{\mu}{L^2} \cdot V_{ov}$$

- For device with carrier velocity saturation,

$$g_m = W C_{ox} v_{scl} \quad \Rightarrow \quad \omega_T = \frac{3}{2} \cdot \frac{v_{scl}}{L}$$

MOSFET's Transition frequency

For MOSTs in the weak inversion region,

$$\omega_T = \frac{g_m}{C_{gb}} \qquad g_m = \frac{I_D}{U_T}\frac{C_{ox}}{C_{ox}+C_{depl}} \qquad C_{gb} = WL \times \frac{C_{ox}C_{depl}}{C_{ox}+C_{depl}}$$

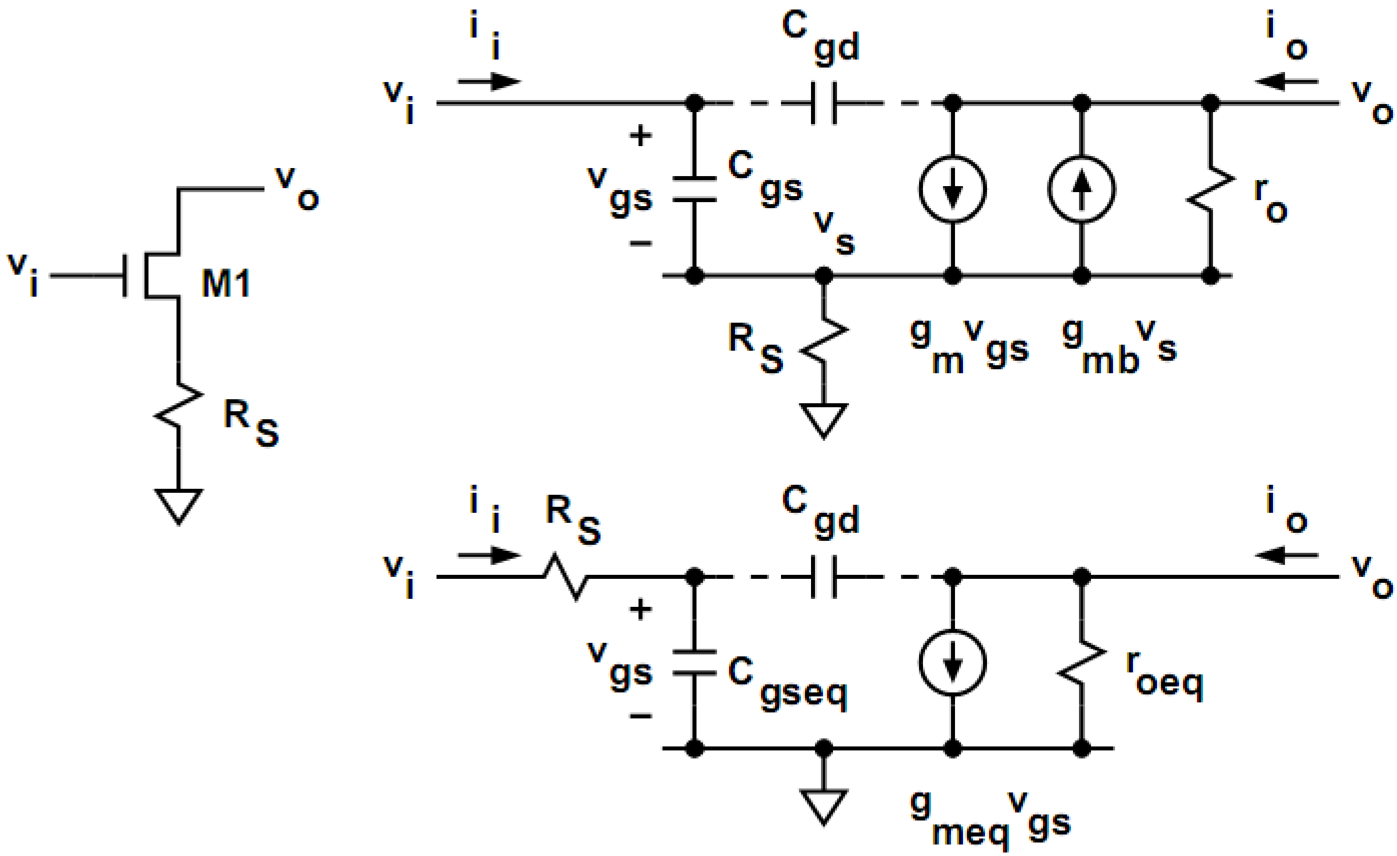
$$\omega_T = \frac{I_D}{U_T}\cdot\frac{1}{WLC_{depl}} = \frac{I_t}{U_T}\cdot\frac{1}{C_{depl}}\cdot\frac{1}{L^2}\cdot\frac{I_D}{I_M}$$

- $I_M = I_t \cdot W/L$ is the maximum I_D for device in weak inversion.

Since $I_t \propto D_n$ and $D_n = \mu U_T$, we have

$$\omega_T \simeq \frac{D_n}{L^2}\cdot\frac{I_D}{I_M} \simeq \frac{\mu}{L^2}\cdot U_T\cdot\frac{I_D}{I_M}$$

Common-source amplifier with source degeneration



Common-source amplifier with source degeneration

To find C_{gseq} and g_{meq}, let $v_o = 0$, then

$$(g_m + sC_{gs})(v_i - v_s) = (G_S + g_{mb} + g_o)v_s$$

At frequencies where $\omega \ll \omega_T = g_m/C_{gs}$,

$$\frac{v_s}{v_i} = \frac{g_m + sC_{gs}}{g_m + g_{mb} + G_S + g_o + sC_{gs}} \approx \frac{g_m}{g_m + g_{mb} + G_S + g_o}$$

$$g_{meq} = \frac{-i_o}{v_i} = g_m\left(1 - \frac{v_s}{v_i}\right) - (g_{mb} + g_o)\frac{v_s}{v_i} = \frac{g_m G_S}{g_m + g_{mb} + G_S + g_o} = \frac{g_m}{1 + (g_m + g_{mb})R_S + \frac{R_S}{r_o}}$$

$$\frac{i_i}{v_i} = sC_{gs}\left(1 - \frac{v_s}{v_i}\right) = sC_{gs} \cdot \frac{1 + g_{mb}R_S + \frac{R_S}{r_o}}{1 + (g_m + g_{mb})R_S + \frac{R_S}{r_o}}$$

Common-source amplifier with source degeneration

- If $r_o \gg R_S$,

$$g_{meq} \approx \frac{g_m}{1+(g_m+g_{mb})R_S} = \frac{g_m}{1+(1+\chi)g_m R_S}$$

$$C_{gseq} = C_{gs} \cdot \frac{1+g_{mb}R_S}{1+(g_m+g_{mb})R_S} = C_{gs} \cdot \frac{1+\chi g_m R_S}{1+(1+\chi)g_m R_S}$$

To find r_{oeq}, let $v_i = 0$, then

$$(g_m + g_{mb} + sC_{gs} + G_S)v_s = g_o(v_o - v_s)$$

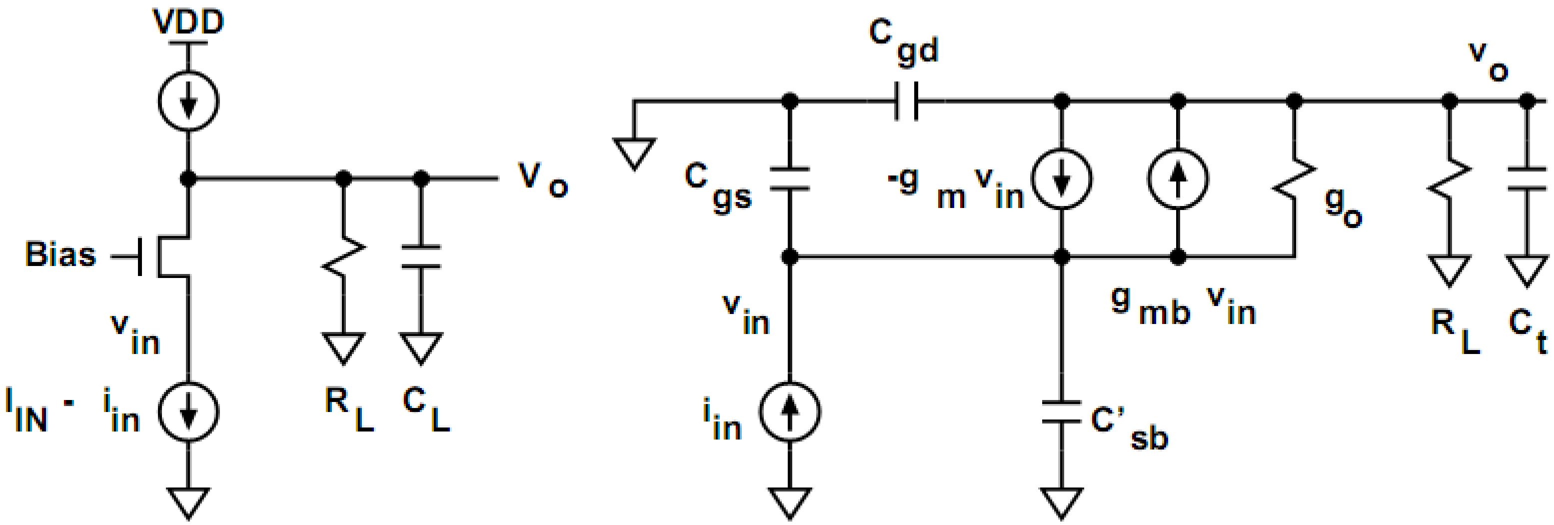
$$\frac{v_s}{v_o} = \frac{g_o}{g_m+g_{mb}+G_S+g_o+sC_{gs}} \approx \frac{g_o}{g_m+g_{mb}+G_S+g_o}$$

$$g_{oeq} = \frac{i_o}{v_o} = g_o\left(1-\frac{v_e}{v_o}\right) - (g_m+g_{mb})\frac{v_e}{v_o} = \frac{g_o G_S}{g_m+g_{mb}+G_S+g_o}$$

$$r_{oeq} = R_S + r_o[1+(g_m+g_{mb})R_S]$$

- r_{oeq} can be made arbitrarily large by increasing R_S.

Common-gate amplifier



$$g'_m = g_m + g_{mb} \qquad C'_L = C_t + C_{gd} = C_L + C_{db} + C_{gd} \qquad C_{in} = C_{gs} + C'_{sb}$$

The nodal equations are

$$i_{in} = (g'_m + sC_{in})v_{in} - g_o(v_o - v_{in}) \qquad g'_m v_{in} = (G_L + sC'_L)v_o + g_o(v_o - v_{in})$$

Common-gate amplifier

If the $g_o(v_o - v_{in})$ terms are neglected, then

$$\text{Transimpedance} = Z_t(s) = \frac{v_o}{i_{in}} = \frac{R_L}{\left(1 - s/p_1\right)\left(1 - s/p_2\right)}$$

$$p_1 = -\frac{g'_m}{C_{in}} = -\frac{g'_m}{C_{gs} + C'_{sb}} \quad p_2 = -\frac{1}{R_L C'_L}$$

$$\text{Input Impedance} = Z_{in}(s) = \frac{v_{in}(s)}{i_{in}(s)} = \frac{1/g'_m}{1 - s/p_1}$$

$$\text{Current Gain} = \frac{i_o(s)}{i_{in}(s)} = \frac{g'_m v_{in}}{i_{in}} = g'_m Z_{in}(s) = \frac{1}{1 - s/p_1}$$

- Note that $p_1 \simeq \omega_T = g_m/(C_{gs} + C_{gd})$.

Common-gate amplifier

If g_o is considered,

$$\text{Voltage Gain} = A_v(s) = \frac{v_o}{v_{in}} = \frac{g'_m + g_o}{g_o + G_L + sC'_L}$$

$$\text{Input Impedance} = Y_{in}(s) = \frac{i_{in}}{v_{in}} = g'_m + sC_{in} - g_o(A_v - 1)$$

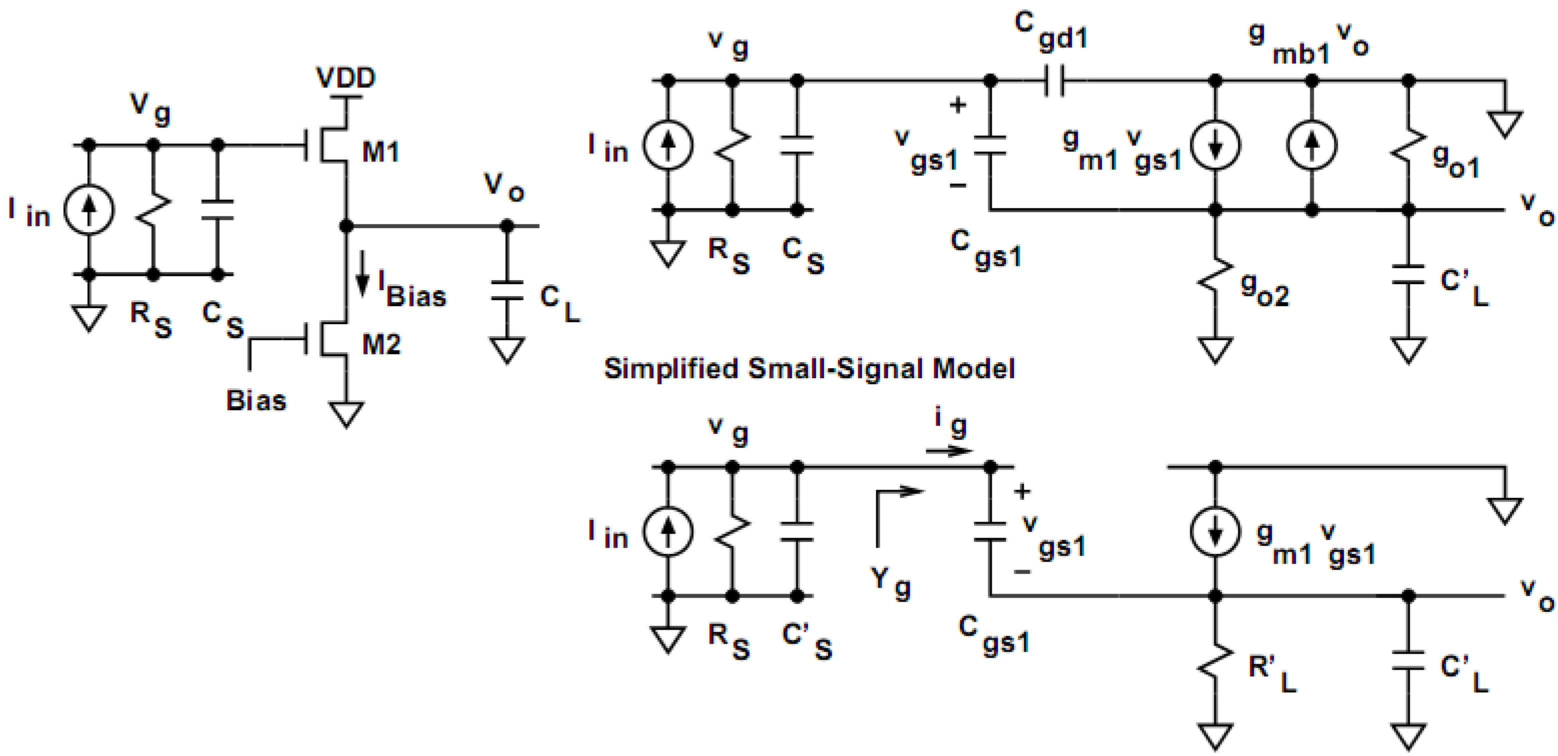
$$Z_t(s) = \frac{v_o}{i_{in}} = \frac{A_v(s)}{Y_{in}(s)}$$

- At low frequencies where $\omega \to 0$, assuming $g'_m \gg g_o$,

$$A_v = \frac{g'_m + g_o}{g_o + G_L} \approx \frac{g'_m}{g_o + G_L}$$

$$\Rightarrow \quad Y_{in} = g'_m - \frac{g'_m}{1 + \frac{G_L}{g_o}} + g_o \approx \frac{g'_m}{1 + \frac{R_L}{r_o}} \qquad Z_t = \frac{A_v}{Y_{in}} \approx R_L$$

Common-drain amplifier



$$C'_S = C_S + C_{gd1} \qquad C'_L = C_L + C'_{sb1} + C_{db2} + C_{gd2} \qquad G'_L = g_{o1} + g_{o2} + g_{mb1} = \frac{1}{R'_L}$$

Common-drain amplifier

Summing the currents at the output node, we have

$$(g_{m1} + sC_{gs1})(v_g - v_o) - v_o(sC'_L + G'_L) = 0$$

The voltage gain from gate to output is

$$A_{vg}(s) \equiv \frac{v_o(s)}{v_g(s)} = \frac{g_{m1} + sC_{gs1}}{g_{m1} + G'_L + s(C_{gs1} + C'_L)} = A_{vg}(0)\frac{1 - s/z_1}{1 - s/p_1}$$

$$A_{vg}(0) = \frac{g_{m1}}{g_{m1} + G'_L} = \frac{g_{m1}}{g_{m1} + g_{mb1} + g_{o1} + g_{o2}} \qquad A_{vg}(\infty) = \frac{C_{gs1}}{C_{gs1} + C'_L}$$

$$z_1 = -\frac{g_{m1}}{C_{gs1}} \approx -\omega_T \qquad p_1 = -\frac{g_{m1} + G'_L}{C_{gs1} + C'_L}$$

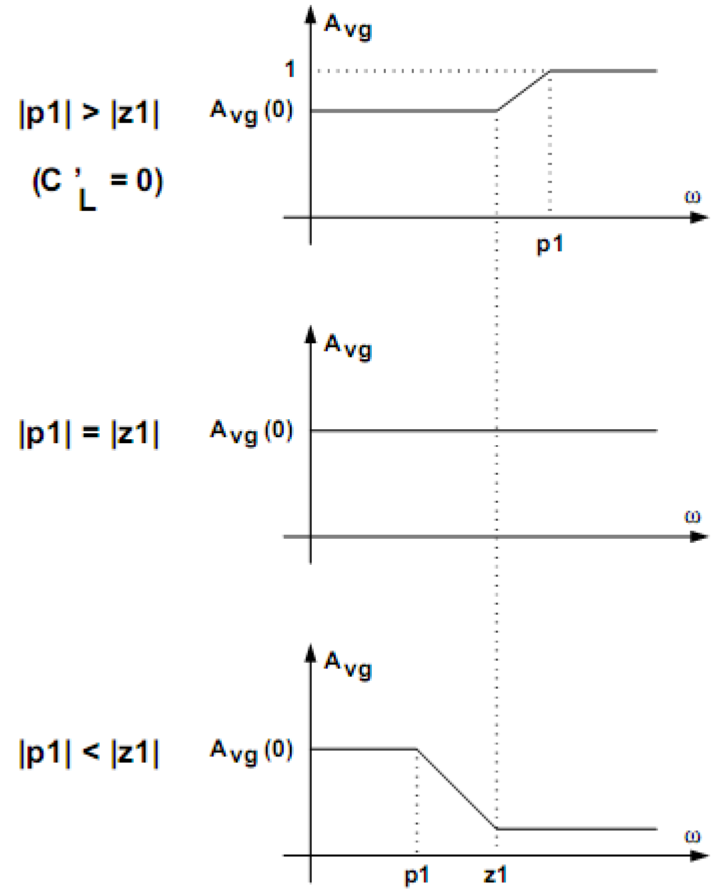
Common-drain amplifier

For most practical cases

$$g_{o1} + g_{o2} \ll g_{m1} + g_{mb1} = g_{m1}(1 + \chi)$$

$$A_{vg}(0) \approx \frac{g_{m1}}{g_{m1} + g_{mb1}} \approx \frac{1}{1 + \chi_1}$$

$$p_1 \approx -\frac{g_{m1}(1 + \chi_1)}{C_{gs1} + C'_L} \approx (1 + \chi_1)\left(\frac{1}{1 + \frac{C'_L}{C_{gs1}}}\right) z_1$$



Common-drain amplifier

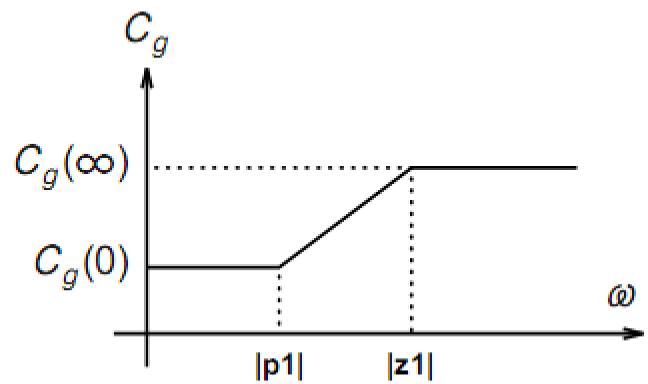
The input admittance looking into the gate is

$$Y_g(s) = \frac{i_g}{v_g} = sC_{gs1}[1 - A_{vg}(s)] = \frac{sC_{gs1}(G'_L + sC'_L)}{g_{m1} + G'_L + s(C_{gs1} + C'_L)}$$

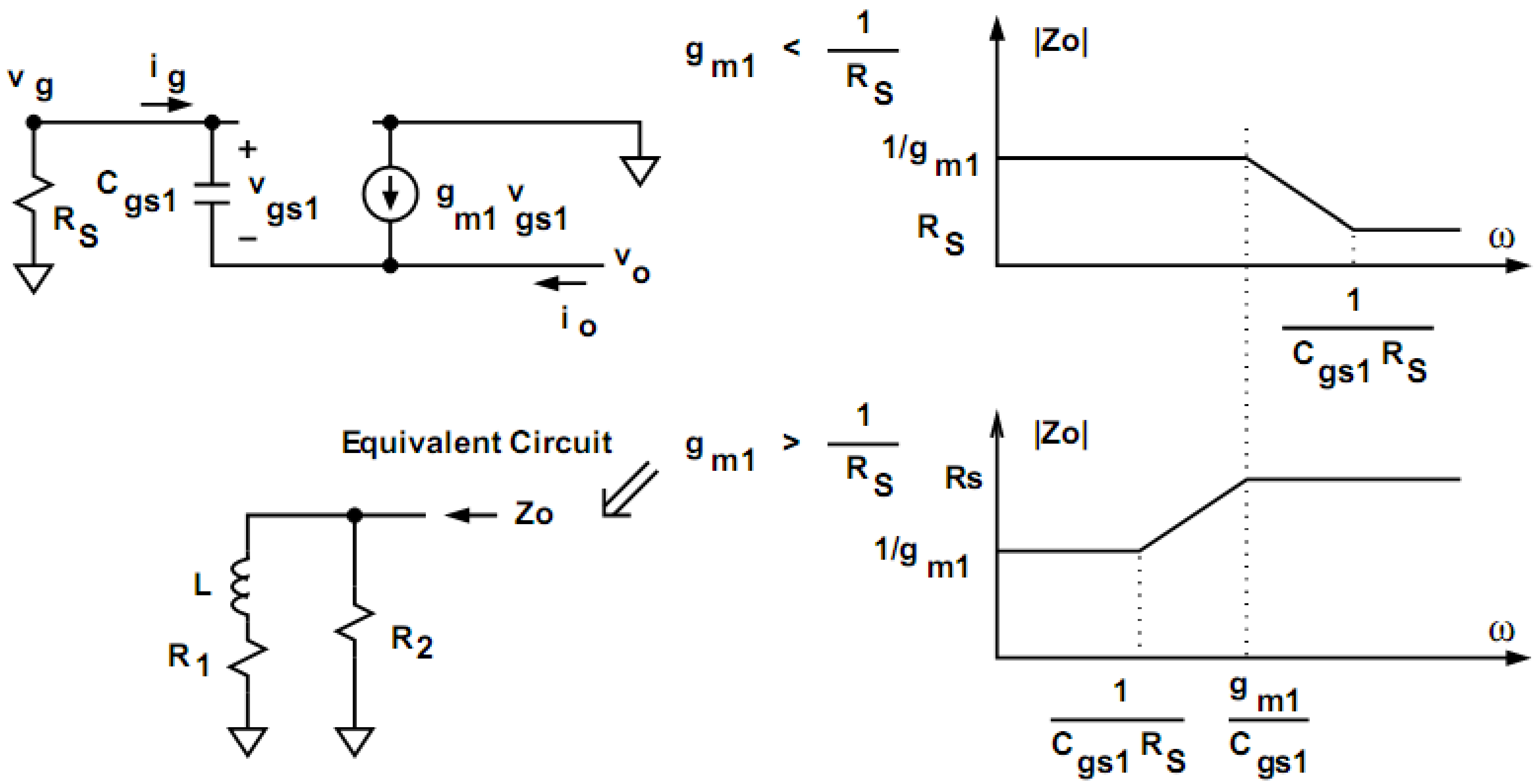
Define the capacitance looking into the gate as

$$Y_g(s) = sC_g(s)$$

$$\begin{aligned} C_g(j\omega) &= C_{gs1}[1 - A_{vg}(j\omega)] \\ C_g(0) &= C_{gs1}[1 - A_{vg}(0)] \\ C_g(\infty) &= C_{gs1}[1 - A_{vg}(\infty)] = \frac{C_{gs1}C'_L}{C_{gs1} + C'_L} \end{aligned}$$



Common-drain amplifier



$$i_o = -(g_{m1} + sC_{gs1})(v_g - v_o) \qquad G_S v_g + sC_{gs1}(v_g - v_o) = 0$$

Common-drain amplifier

The output admittance is

$$Y_o(s) = \frac{1}{Z_o(s)} \equiv \frac{i_o}{v_o} = \frac{G_S(g_{m1} + sC_{gs1})}{G_S + sC_{gs1}} = G_S + \frac{G_S(g_{m1} - G_S)}{G_S + sC_{gs1}} = G_S + \frac{1}{\frac{1}{g_{m1}-G_S} + \frac{sC_{gs1}R_S}{g_{m1}-G_S}}$$

- Note that

$$Z_o(0) = \frac{1}{g_{m1}} \qquad Z_o(\infty) = R_S$$

- The equivalent circuit is

$$R_1 = \frac{1}{g_{m1} - G_S} \qquad R_2 = R_S \qquad L = \frac{R_S C_{gs1}}{g_{m1} - G_S}$$

Dominant pole approximation

The response of an amplifier has the form of

$$A(s) = A(0)\frac{N(s)}{D(s)} = A(0)\frac{1 + a_1 s + a_2 s^2 + \cdots + a_m s^m}{1 + b_1 s + b_2 s^2 + \cdots + b_n s^n} \approx \frac{A(0)}{\left(1 - \frac{s}{p_1}\right)\left(1 - \frac{s}{p_2}\right)\cdots\left(1 - \frac{s}{p_n}\right)}$$

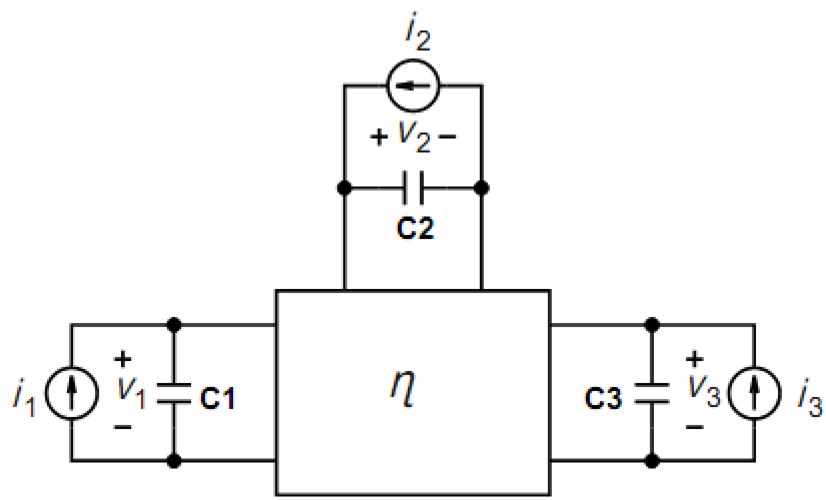
If $|p_1| \ll |p_2|, |p_3|, \cdots, |p_n|$, then p_1 is a dominant pole. We have

$$b_1 = -\frac{1}{p_1} - \frac{1}{p_2} - \cdots - \frac{1}{p_n} \approx -\frac{1}{p_1} = \left|\frac{1}{p_1}\right|$$

$$|A(j\omega)| = \frac{A(0)}{\sqrt{\left[1 + \left(\frac{\omega}{p_1}\right)^2\right]\left[1 + \left(\frac{\omega}{p_2}\right)^2\right]\cdots\left[1 + \left(\frac{\omega}{p_n}\right)^2\right]}} \approx \frac{A(0)}{\sqrt{1 + \left(\frac{\omega}{p_1}\right)^2}}$$

$$-3\text{ dB Bandwidth} = \omega_{-3\text{dB}} \approx |p_1| \approx \frac{1}{b_1}$$

Zero-value time constants

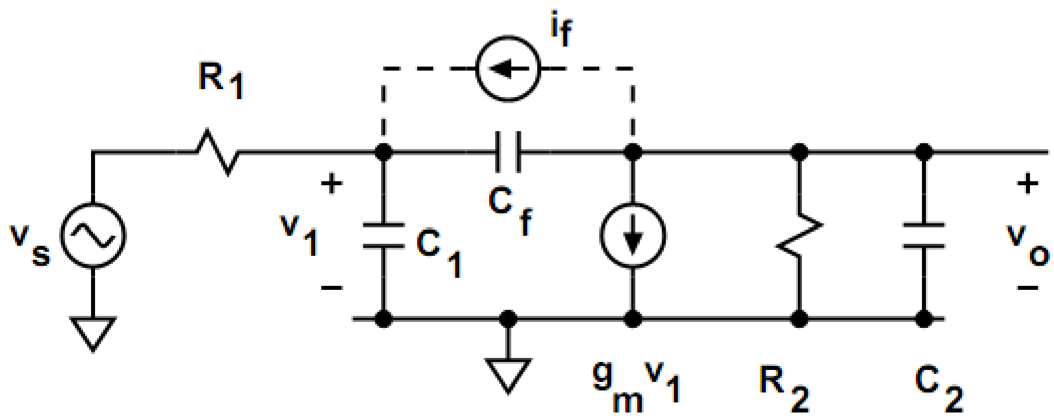


- η is a linear active network without energy storage.
- The b_1 in the denominator of the system function can be expressed as

$$b_1 = \sum T_0 = R_{10}C_1 + R_{20}C_2 + R_{30}C_3 + \cdots$$

R_{i0} is the driving point resistance seen by C_i with all capacitors equal to zero.

Zero-value time constants



$$R_{10} = R_1$$
$$R_{20} = R_2$$

To determine R_{f0}, replace C_f with a current source i_f, then

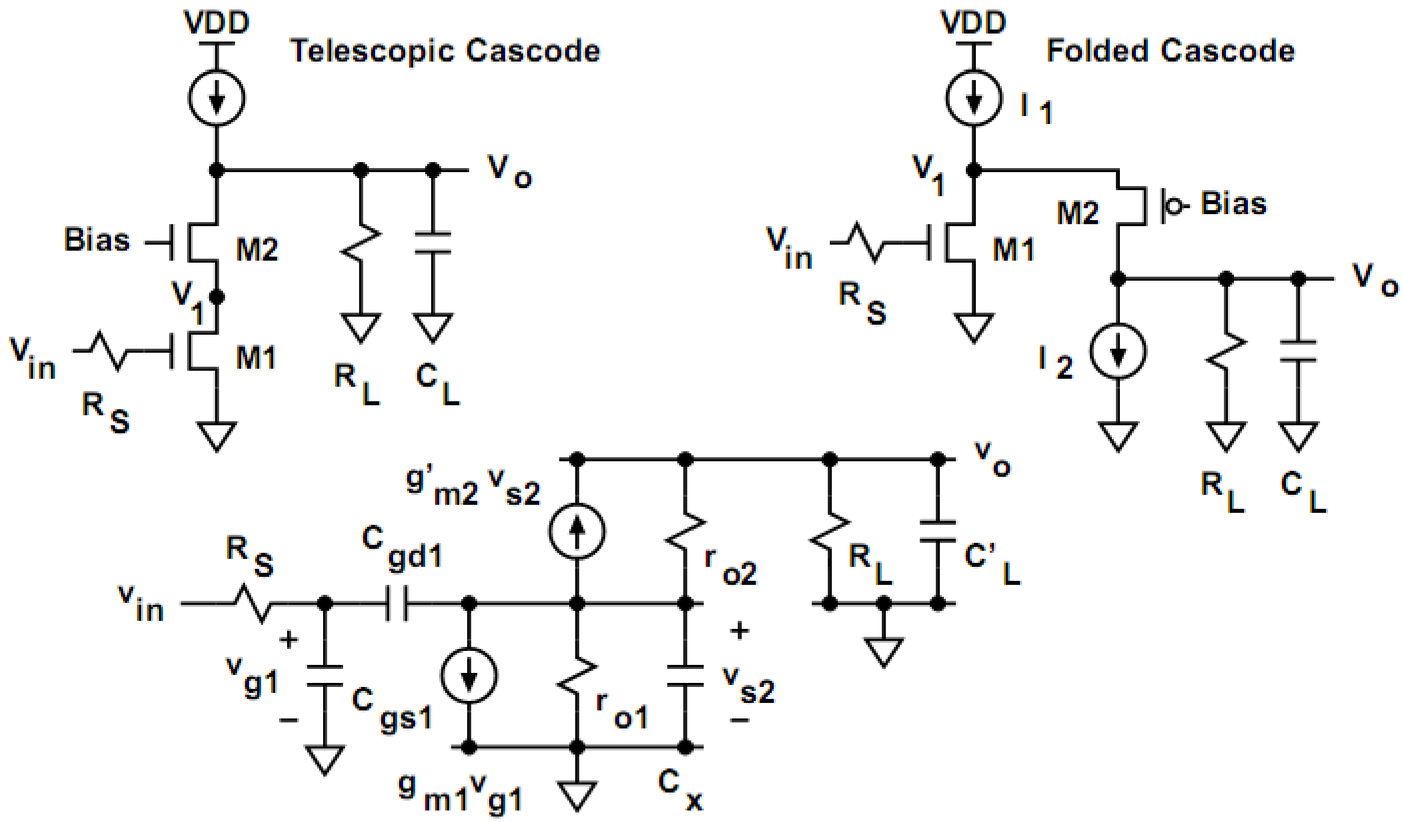
$$v_1 = i_f R_1 \qquad v_o = -(i_f + g_m v_1)R_2$$

$$R_{f0} = \frac{v_1 - v_o}{i_f} = R_1 + R_2 + g_m R_1 R_2 = R_1\left(1 + g_m R_2 + \frac{R_2}{R_1}\right)$$

We have

$$b_1 = \sum T_0 = R_1 C_1 + R_2 C_2 + (R_1 + R_2 + g_m R_1 R_2)C_f$$

Cascode amplifiers



$g'_{m2} = g_{m2} + g_{mb2}$ $\quad C_x = C_{db1} + C'_{sb2} + C_{gs2}$ $\quad C'_L = C_L + C_{db2} + C_{gd2}$

Cascode amplifiers

The output impedance looking into M2's drain is

$$R_{ot2} = r_{o1} + (g'_{m2} r_{o1} + 1) r_{o2} \approx g'_{m2} r_{o1} r_{o2}$$

The input admittance looking into M2's source is

$$G_{in2} \approx \frac{g'_{m2}}{g_{o2}/G_L + 1}$$

The overall voltage gain is

$$A_v = \frac{v_o}{v_{in}} \approx -\frac{g_{m1}}{g_{o1} + G_{in2}} \times \frac{g'_{m2}}{g_{o2} + G_L} = g_{m1} \times \frac{g'_{m2} r_{o1} r_{o2} R_L}{r_{o2} + g'_{m2} r_{o1} r_{o2} + R_L} \approx g_{m1} \times (R_{ot2} \parallel R_L)$$

- Let $g_m = g_{m1} = g'_{m2}$, $r_o = r_{o1} = r_{o2}$, and $g_m \gg g_o$. If $R_L = R_{ot2} = g_m r_o^2$, then

$$G_{in2} \approx \frac{g_m}{g_o R_L + 1} \approx \frac{g_m}{g_m r_o + 1} \approx g_o \qquad A_v \approx -\frac{g_m}{2g_o} \frac{g_m}{g_o + G_L} \approx -\frac{1}{2}\left(\frac{g_m}{g_o}\right)^2$$

Cascode amplifiers

Using the zero-value time constant method, we have

$$R_{gs10} = R_S \qquad R_{gd10} = R_S + R_{x0} + g_{m1} R_S R_{x0}$$

$$R_{x0} = r_{o1} \parallel R_{in2} \approx r_{o1} \parallel \left[(1/g'_{m2})(g_{o2}/G_L + 1)\right] \qquad R_{L0} = R_L \parallel R_{ot2} \approx R_L \parallel (g'_{m2} r_{o1} r_{o2})$$

$$\sum T_0 = R_{gs10} C_{gs1} + R_{gd10} C_{gd1} + R_{x0} C_x + R_{L0} C_L \qquad \omega_{-3dB} = 1/\left(\sum T_0\right)$$

- Let $g_m = g_{m1} = g'_{m2}$, $r_o = r_{o1} = r_{o2}$, $g_m \gg g_o$. If $R_L = R_{ot2} = g_m r_o^2$ and $R_S = r_o$, then

$$R_{in2} \approx r_o \quad R_{x0} = \frac{r_o}{2} \quad R_{L0} \approx \frac{g_m r_o^2}{2} \quad R_{gd10} \approx R_S + \frac{r_o}{2} + \frac{g_m r_o R_S}{2} \approx \frac{g_m r_o^2}{2}$$

$$\Rightarrow \quad \sum T_0 = R_S C_{gs1} + \frac{g_m r_o^2}{2} C_{gd1} + \frac{r_o}{2} C_x + \frac{g_m r_o^2}{2} C_L$$

- $R_{L0} C_L$ usually is the dominant term, unless R_S is very large.

Cascode amplifiers

Let $R_L = R_{ot2} = g'_{m2} r_{o1} r_{o2}$, then

$$G_{in2} \approx \frac{g'_{m2}}{g_{o2} R_L + 1} \approx \frac{g'_{m2}}{g'_{m2} r_{o1} + 1} \approx g_{o1}$$

The dc gain is

$$A_v(0) \approx -\frac{g_{m1}}{g_{o1} + G_{in2}} \times \frac{g'_{m2}}{g_{o2} + G_L} \approx -\frac{1}{2} \cdot \frac{g_{m1}}{g_{o1}} \cdot \frac{g'_{m2}}{g_{o2}}$$

The dominant pole is

$$p_1 = -\frac{1}{R_{L0} C_L} \approx -\frac{2}{g'_{m2} r_{o1} r_{o2} C_L}$$

At frequencies where $|p_1| \ll \omega \ll |p_2|$,

$$A_v(s) = \frac{A_v(0)}{1 - \frac{s}{p_1}} \approx \frac{A_v(0)}{-\frac{s}{p_1}} \approx -\frac{g_{m1}}{sC_L} = -\frac{\omega_u}{s} \qquad \omega_u = A_v(0) \cdot p_1 = \frac{g_{m1}}{C_L}$$

Cascode amplifiers

The second pole is approximately at

$$p_2 = -\frac{g_{o1} + Y_{in2}}{C_x}$$

Y_{in2} is the resistance looking into the M2's source at high frequencies.

$$Y_{in2} = g'_{m2} - g_{o2}\left(\frac{v_o}{v_{s2}} - 1\right)$$

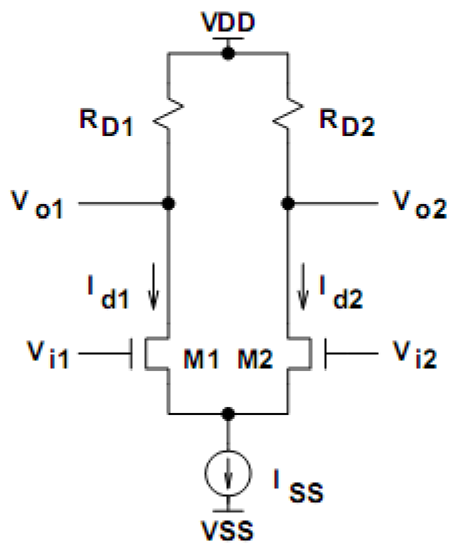
- At frequencies $\omega \gg (g_{o2} + G_L)/C_L$,

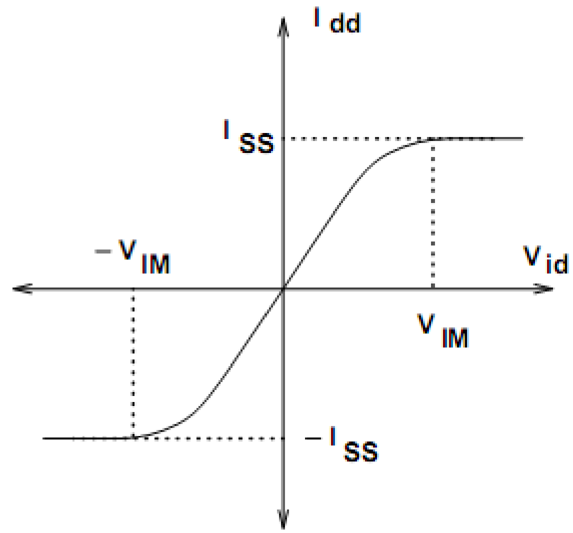
$$\frac{v_o}{v_{s2}} = \frac{g'_{m2} + g_{o2}}{g_{o2} + G_L + sC_L} \approx \frac{g'_{m2}}{sC_L} \quad \Rightarrow \quad Y_{in2} \approx g'_{m2}\left(1 - \frac{g_{o2}}{sC_L}\right) + g_{o2} \approx g'_{m2}$$

$$p_2 \approx -\frac{g'_{m2}}{C_x} \approx -\frac{g_{m2}}{KC_{gs2}} \approx -\frac{\omega_T}{K}$$

where K is between 1 and 2 (usually closer to 1).

Differential amplifier





Assume

- M1=M2.
- $R_{D1} = R_{D2} \equiv R_D$.
- Neglect r_o.
- $R_{SS} \to \infty$.

$$V_{id} \equiv V_{i1} - V_{i2} \qquad I_{dd} \equiv I_{d1} - I_{d2} \qquad V_{od} \equiv V_{o1} - V_{o2} = -I_{dd}R_D$$

Assume M1 and M2 are in the saturation region,

$$I_{d1} = \frac{1}{2}k(V_{gs1} - V_t)^2 \qquad I_{d2} = \frac{1}{2}k(V_{gs2} - V_t)^2 \qquad k = \mu C_{ox}\frac{W}{L}$$

Differential amplifier – large-signal analysis

Summing currents at the common source node, we have

$$I_{d1} + I_{d2} = I_{SS} \quad \Rightarrow \quad I_{d1} = \frac{I_{SS}}{2} + \frac{I_{dd}}{2} \qquad I_{d2} = \frac{I_{SS}}{2} - \frac{I_{dd}}{2}$$

The gate voltages can be written as

$$V_{gs1} = V_t + \sqrt{\frac{2I_{d1}}{K}} \qquad V_{gs2} = V_t + \sqrt{\frac{2I_{d2}}{K}}$$

The differential input voltage is

$$V_{id} = V_{gs1} - V_{gs2} = \sqrt{\frac{2I_{d1}}{k}} - \sqrt{\frac{2I_{d2}}{k}} = \sqrt{\frac{2}{k}}\left(\sqrt{I_{d1}} - \sqrt{I_{d2}}\right)$$

Squaring

$$V_{id}^2 = \frac{2}{k}\left[I_{d1} + I_{d2} - 2\sqrt{I_{d1}I_{d2}}\right] = \frac{2}{k}\left[I_{SS} - \sqrt{I_{SS}^2 - I_{dd}^2}\right]$$

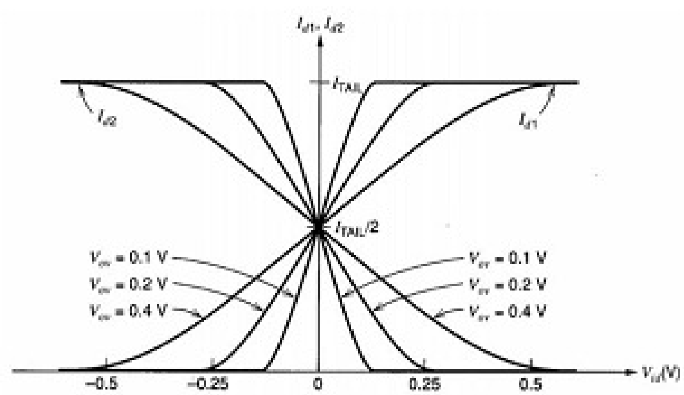
Differential amplifier – large-signal analysis

Rearrange, then we have

$$I_{dd} = \frac{k}{2} V_{id} \sqrt{\frac{4I_{SS}}{k} - V_{id}^2} \quad \text{and} \quad I_{d1} = \frac{I_{SS}}{2} + \frac{I_{dd}}{2} \qquad I_{d2} = \frac{I_{SS}}{2} - \frac{I_{dd}}{2}$$

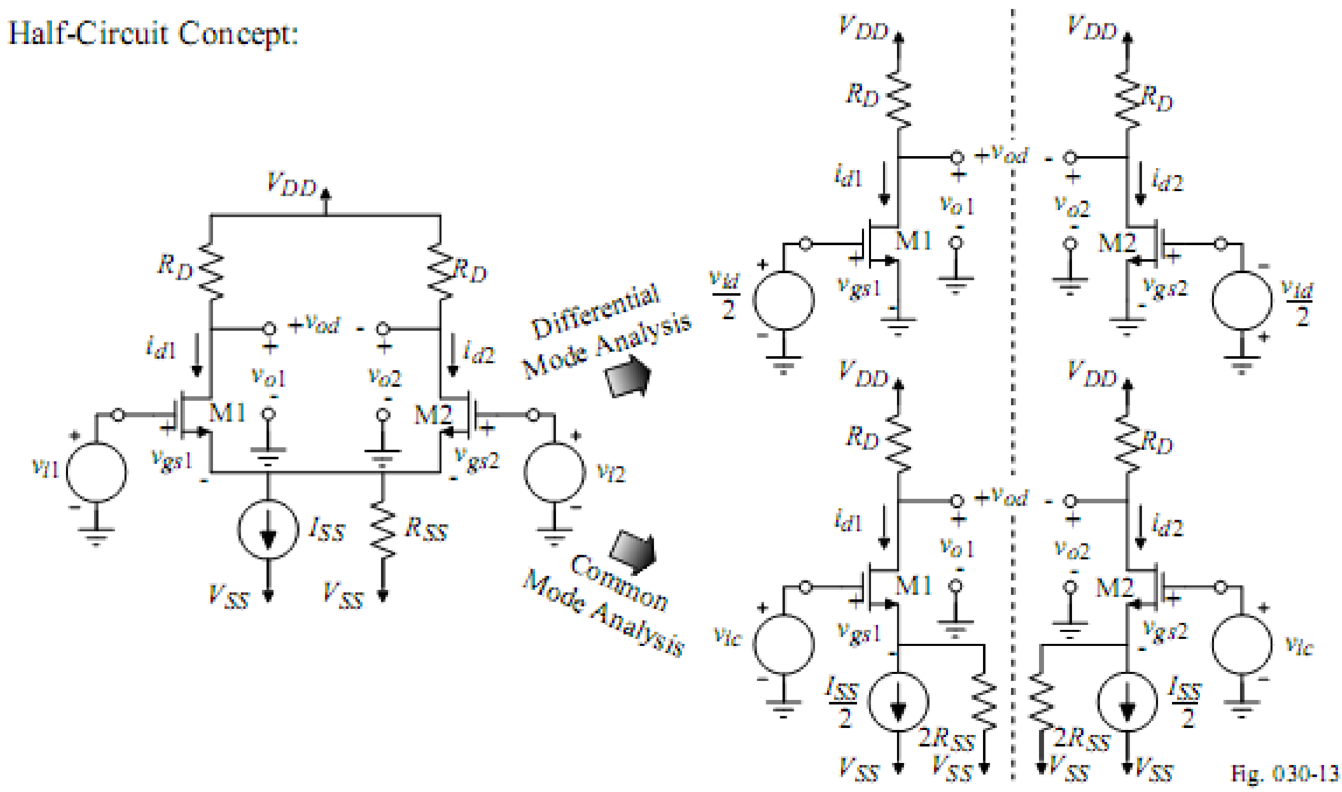
Define V_{IM} as the differential input voltage at which one of the MOST is turned off, i.e.,

$$I_{SS} = \frac{k}{2} V_{IM} \sqrt{\frac{4I_{SS}}{k} - V_{IM}^2} \Rightarrow V_{IM} = \sqrt{\frac{2I_{SS}}{k}} = \sqrt{2}\,(V_{ov1})\big|_{V_{id}=0} = \sqrt{2}\,(V_{ov2})\big|_{V_{id}=0}$$



Differential- and common-mode small-signal analysis

The small-signal performance of a differential amplifier can be separated into a differential mode and common mode analysis. This separation allows us to take advantage of the following simplifications.



Note: The half-circuit concept is valid as long as the resistance seen looking into each source is approximately the same.

Differential-mode analysis

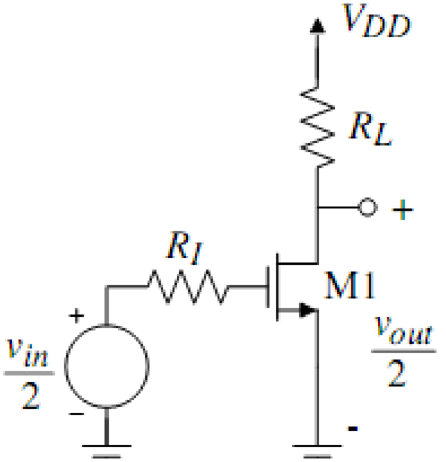


Fig. 070

Small Signal Model:

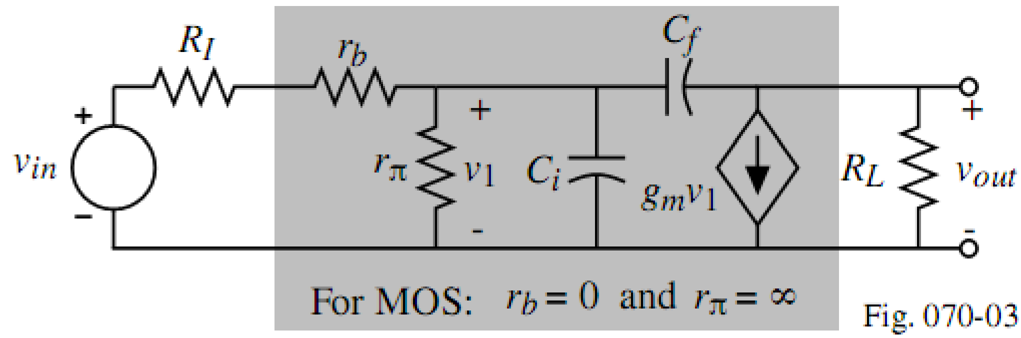


Fig. 070-03

Miller Approach:

Assume that $R_L < (1/\omega C_f)$, then $v_{out} \approx -g_m R_L v_1$

Therefore, the small-signal model can be approximated as,

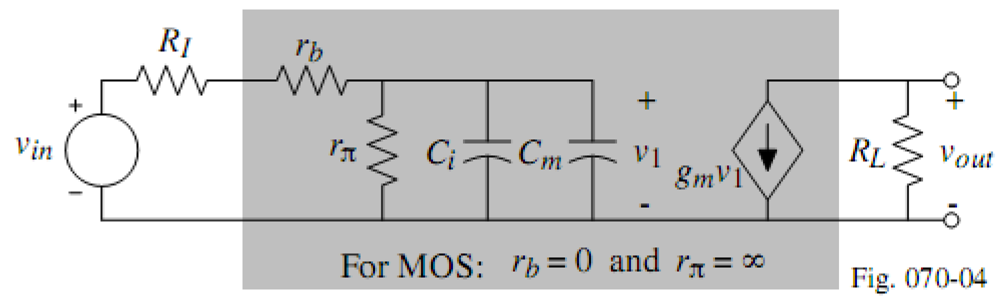


Fig. 070-04

where

$$C_m = C_f(1+g_m R_L)$$

Differential-mode analysis

The small-signal analysis of the previous circuit defining $C_t = C_i + C_m$ is,

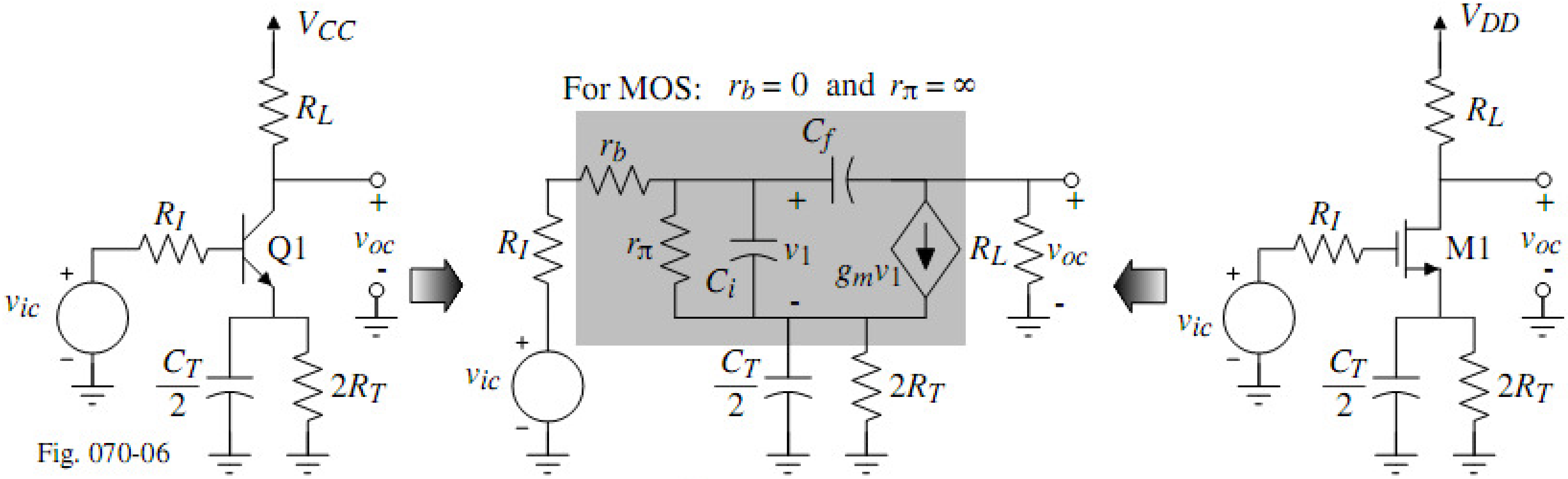
$$\frac{v_{out}}{v_{in}} = \left(\frac{v_{out}}{v_1}\right)\left(\frac{v_1}{v_{in}}\right) = (-g_m R_L)\left(\frac{\dfrac{r_\pi}{1+r_\pi C_t s}}{\dfrac{r_\pi}{1+r_\pi C_t s} + R_I + r_b}\right) = -g_m R_L\left(\frac{r_\pi}{r_\pi + R_I + r_b}\right)\left(\frac{1}{1+\dfrac{s r_\pi C_t (R_I + r_b)}{r_\pi + R_I + r_b}}\right)$$

Therefore we see that the gain (K), pole (p_1), and -3dB frequency (ω_{-3dB}) is given as,

	K	p_1	ω_{-3dB}
BJT	$-g_m R_L\left(\dfrac{r_\pi}{r_\pi + R_I + r_b}\right)$	$-\dfrac{r_\pi + R_I + r_b}{r_\pi C_t (R_I + r_b)}$	$\dfrac{r_\pi + R_I + r_b}{r_\pi C_t (R_I + r_b)}$
MOS	$-g_m R_L$	$-\dfrac{1}{C_t R_I}$	$\dfrac{1}{C_t R_I}$

Common-mode analysis

Assumptions: Tail capacitance is dominant and self-resistance is negligible.



$$\therefore \quad A_{cm} = \frac{v_{oc}}{v_{ic}} \approx -\frac{R_L}{Z_T} \quad \text{where} \quad Z_T = \frac{2R_T}{1+sR_TC_T} \quad \Rightarrow \quad A_{cm} = \frac{v_{oc}}{v_{ic}} \approx -\frac{R_L}{2R_T}(1+sR_TC_T)$$

This zero at $\omega = 1/R_TC_T$ causes the CM gain to increase, resulting in a *CMRR* decrease.

$$CMRR = \left|\frac{A_{dm}}{A_{cm}}\right| = \frac{\left(\dfrac{g_m 2R_T r_\pi}{r_\pi + R_I + r_b}\right)\left(\dfrac{1}{1+\dfrac{sr_\pi C_t(R_I+r_b)}{r_\pi + R_I + r_b}}\right)}{(1+sR_TC_T)} = \left(\frac{g_m 2R_T r_\pi}{r_\pi + R_I + r_b}\right)\left(\frac{1}{(1+s/\omega_T)(1+s/p_1)}\right)$$

CMRR frequency response

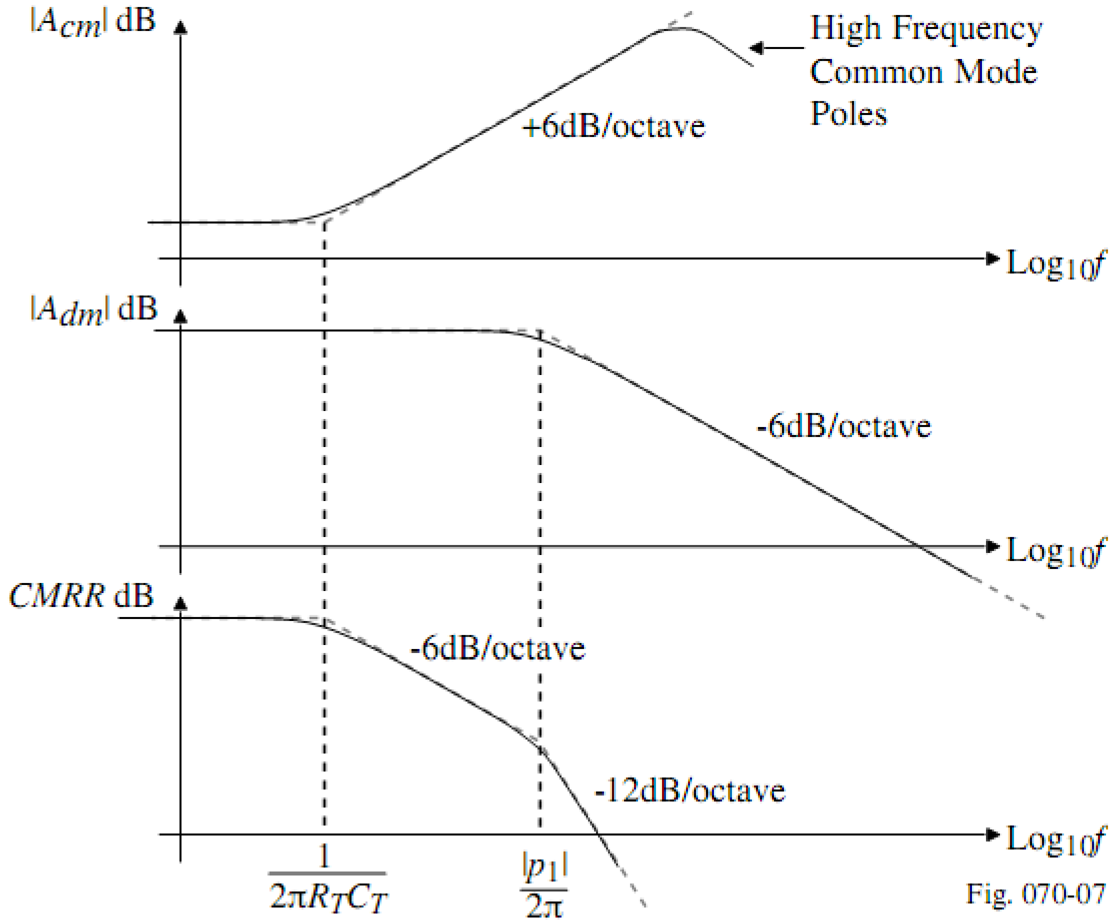
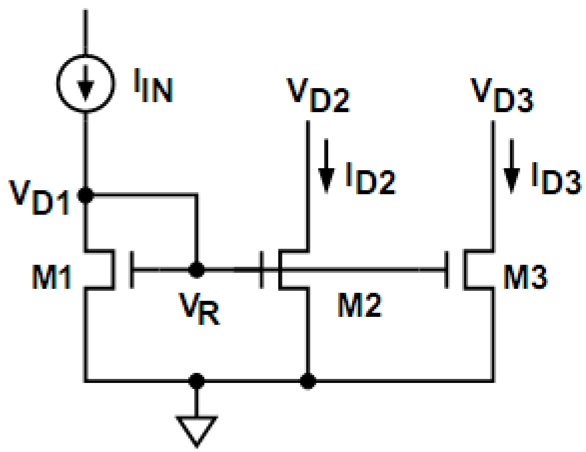


Fig. 070-07

Current-mirrors



$$k' = \mu_n C_{ox}$$

$$I_{IN} = \frac{1}{2}k'\left(\frac{W}{L}\right)_1 (V_R - V_{t1})^2(1 + \lambda_1 V_{D1})$$

$$I_{D2} = \frac{1}{2}k'\left(\frac{W}{L}\right)_2 (V_R - V_{t2})^2(1 + \lambda_2 V_{D2})$$

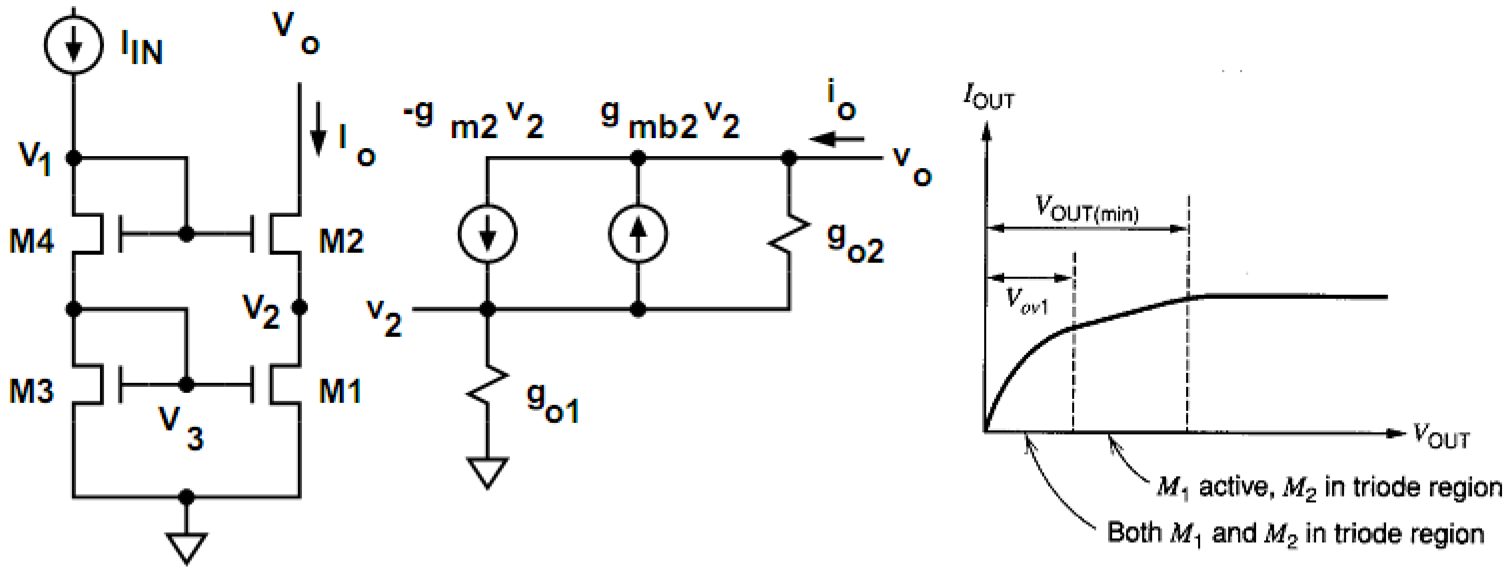
$$I_{D3} = \frac{1}{2}k'\left(\frac{W}{L}\right)_3 (V_R - V_{t3})^2(1 + \lambda_3 V_{D3})$$

Let $V_{t1} = V_{t2} = V_t$ and $\lambda_1 = \lambda_2 = \lambda$, then

$$I_{D2} = I_{IN} \cdot \frac{(W/L)_2}{(W/L)_1} \cdot \frac{1 + \lambda V_{D2}}{1 + \lambda V_{D1}} = I_{IN} \cdot \frac{(W/L)_2}{(W/L)_1} \cdot (1 + \epsilon) \qquad \epsilon \approx \lambda(V_{D2} - V_{D1}) = \frac{V_{D2} - V_{D1}}{V_A}$$

$$R_{o2} = r_{o2} = \frac{1}{\lambda_2 I_{D2}} \qquad V_{o2(min)} = V_{ov2} \approx V_{ov1} \approx \sqrt{\frac{2I_{IN}}{k'(W/L)_1}} \qquad V_{DD(min)} = V_{GS1} = V_t + V_{ov1}$$

Cascode current mirror



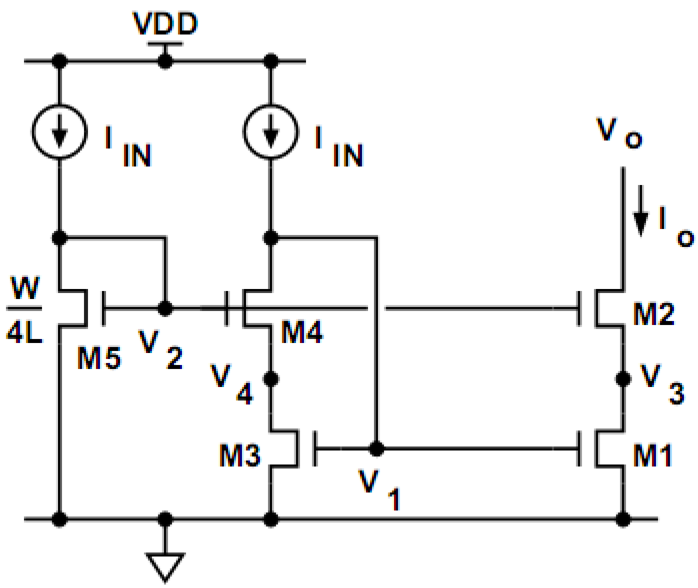
$$R_o = r_{o1}r_{o2}(g_{m2} + g_{mb2} + g_{o1} + g_{o2}) \approx r_{o1}r_{o2}g_{m2}(1 + \chi_2)$$

$$V_{o(min)} = V_{DS1} + V_{DSAT2} = V_{t3} + V_{ov3} + V_{ov2}$$

$$V_{DD(min)} = V_{GS3} + V_{GS4} = V_{t3} + V_{t4} + V_{ov3} + V_{ov4}$$

$$\epsilon \simeq 0$$

High-swing cascode current mirror



$$V_1 = V_t + V_{ov}$$

$$V_2 = V_t + 2V_{ov}$$

$$V_3 = V_{ov}$$

$$V_4 = V_{ov}$$

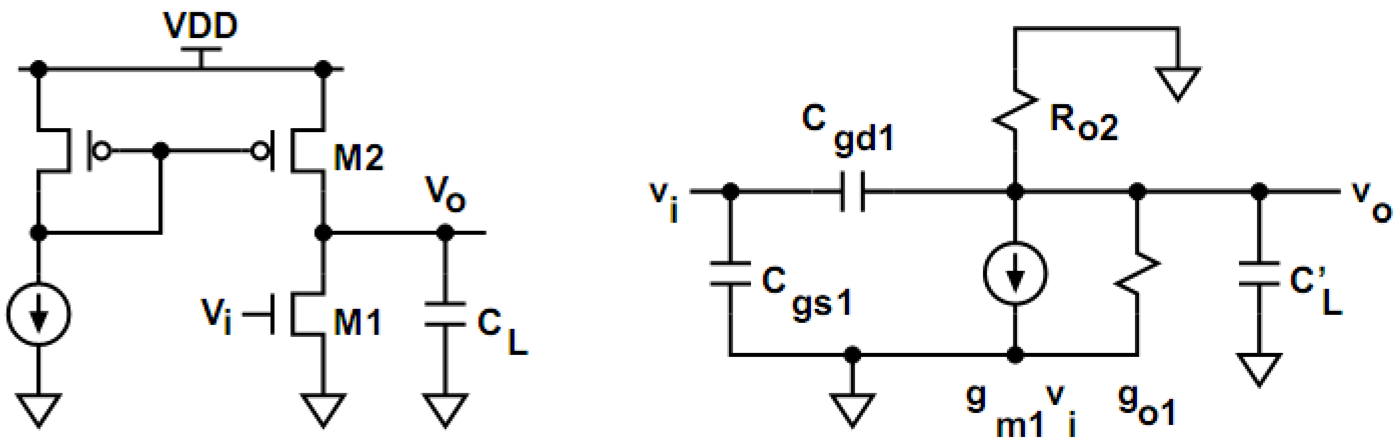
$$V_{o(min)} = V_{DS1} + V_{DSAT2} = 2V_{ov}$$

$$V_{DD(min)} = V_2 = V_t + 2V_{ov}$$

$$\epsilon = 0$$

- In practice, select $(W/L)_5 < (1/4)(W/L)$ due to body effect and design margin.
- To bias M2 and M4 in the active region, want $V_2 - V_1 < V_t \quad \Rightarrow \quad V_{ov} < V_t$.

Common-source with current source load



$$G'_L = g_{o1} + G_{o2} \qquad C'_L = C_L + C_{o2}$$

- The V_o range in normal operation is between V_{DSAT1} and $V_{DD} - V_{o2(min)}$.

Common-source amplifier with active loads