CHAPTER 6

Frequency Response of Amplifiers

Outline

1. **Miller Effect / Poles**
2. Common-Source (CS) Stage
3. Source Follower (SF)
4. Common-Gate (CG) Stage
5. Cascode Stage
6. Differential Pair (DP)

Why Frequency Response

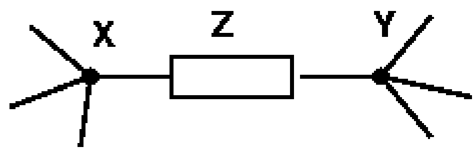
- In most analog circuits, the speed trades with many other parameters such as gain, power dissipation, and noise.

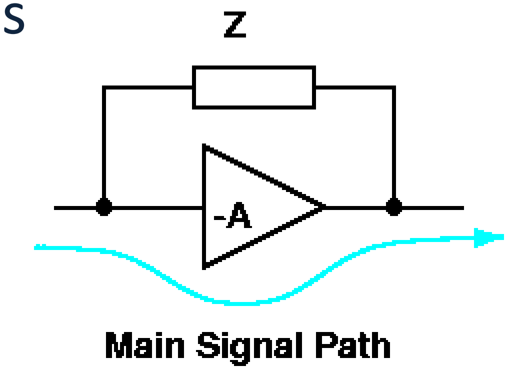
- Transient signal varied with time need to consider the capacitive and inductive effect inherent in circuit.

Miller Effect

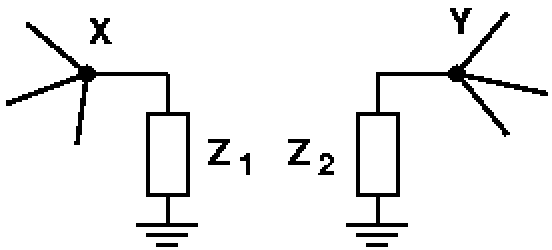
- For the current flowing through *Z* from *X* to *Y* is

$$\frac{V_X - V_Y}{Z} = \frac{V_X}{Z_1}$$



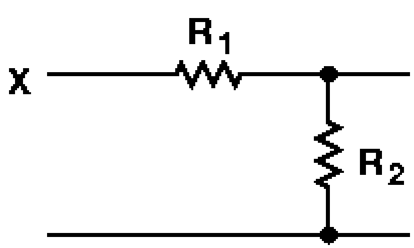


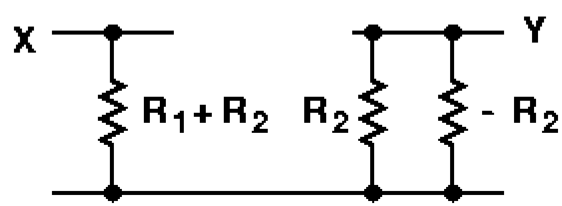
- That is



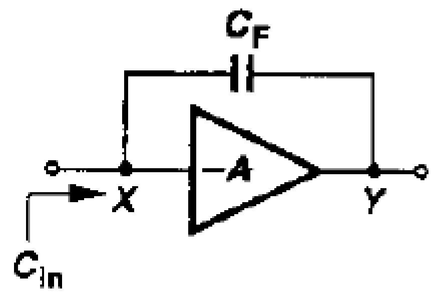
$$Z_1 = \frac{Z}{1 - \frac{V_Y}{V_X}} \qquad Z_2 = \frac{Z}{1 - \frac{V_X}{V_Y}}$$

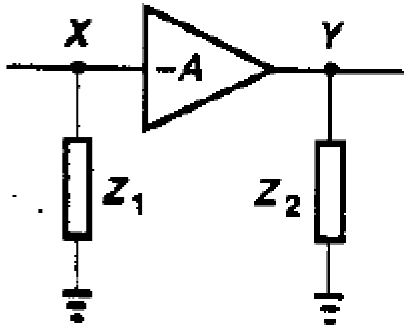
- If the impedance *Z* forms the only signal path between *X* and *Y* , then the conversion is often invalid.





Miller Effect of Feedback Capacitor



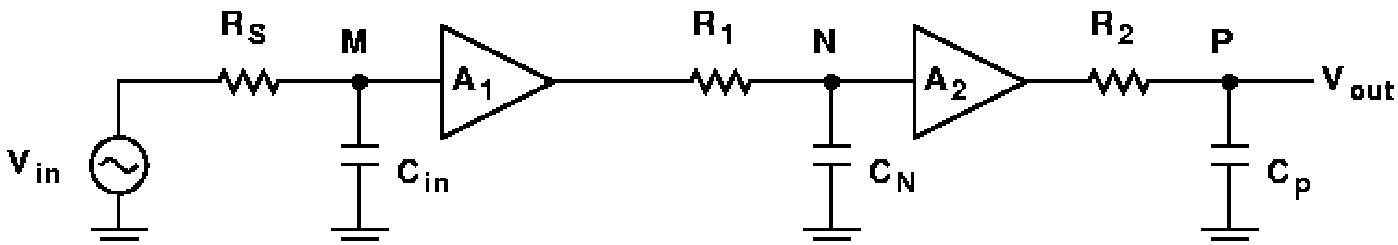


$$Z_1 = \frac{Z}{1-A_v} = \frac{Z}{1+A} = \frac{1}{sC_F(1+A)}, \quad C_1 = C_F(1+A)$$

$$Z_2 = \frac{Z}{1-(1/A_v)} = \frac{Z}{1+(1/A)} = \frac{1}{sC_F(1+A^{-1})}, \quad C_2 = C_F(1+A^{-1})$$

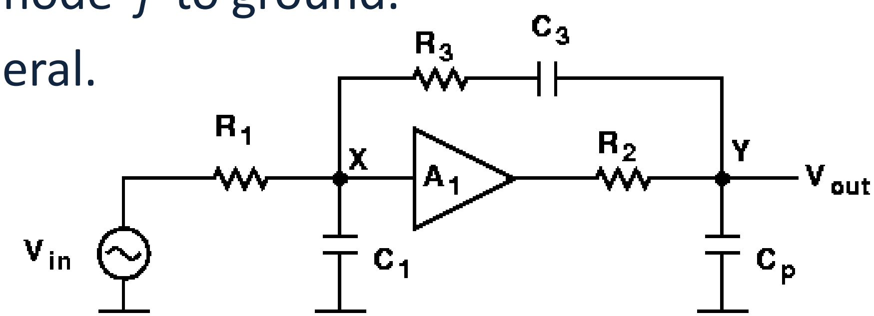
Association of Poles with Nodes

- Assume A_1 and A_2 are ideal amplifiers, R_1 and R_2 model the output resistance of each stage, C_{in} and C_N represent the input capacitance of each stage, and C_p denotes the load capacitance.



$$\frac{V_{out}}{V_{in}}(s) = \frac{A_1}{1+R_S C_{in} s} \cdot \frac{A_2}{1+R_1 C_N s} \cdot \frac{1}{1+R_2 C_p s}$$

- Each pole with one node of the circuit, i.e., $\omega_j = \tau_j^{-1}$, where τ_j is the product of the capacitance and resistance seen at node j to ground.
- The above statement is not valid in general.
  - The location of the pole is difficult to calculate.



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5. Cascode Stage
6. Differential Pair (DP)

Common Source Stage

- Provides voltage gain and high input impedance.

Vout/Vx = -gm(RD//ro) = -A
→ A = gm(RD//ro)
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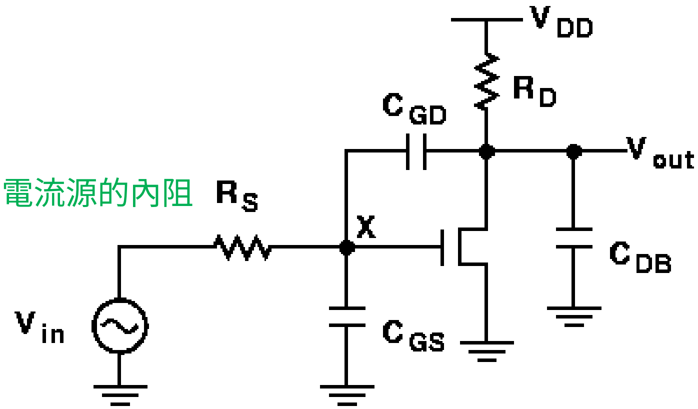


$$\omega_{in} = \frac{1}{R_S[C_{GS} + (1+g_m R_D)C_{GD}]}$$

Cin

沒有考慮ro

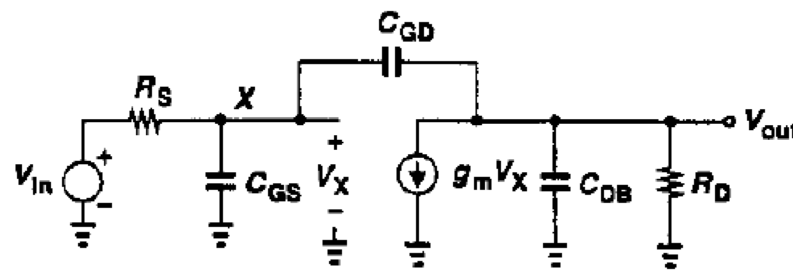
$$\omega_{out} = \frac{1}{R_D(C_{DB} + C_{GD})}$$

Cout

Cout = CGD(1+1/A)+CDB

Razavi認為A很大

- To obtain the exact transfer function



$$\frac{V_X - V_{in}}{R_S} + V_X C_{GS} s + (V_X - V_{out}) C_{GD} s = 0,$$

$$(V_{out} - V_X) C_{GD} s + g_m V_X + V_{out}\left(\frac{1}{R_D} + C_{DB} s\right) = 0,$$

$$\xi = C_{GS}C_{GD} + C_{GS}C_{DB} + C_{GD}C_{DB}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD} + C_{DB})]s + 1}$$

Dominant Pole Approximation

低頻的是Dominant pole

- Two independent capacitors (initial conditions) in the circuit yield a second order differential equation for the time response.

$$\frac{V_{out}}{V_{in}}(s)=\frac{(C_{GD}s-g_m)R_D}{R_SR_D\xi s^2+[R_S(1+g_mR_D)C_{GD}+R_SC_{GS}+R_D(C_{GD}+C_{DB})]s+1}$$

$$=-g_mR_D\frac{(1-C_{GD}s/g_m)}{R_SR_D\xi s^2+[R_S(1+g_mR_D)C_{GD}+R_SC_{GS}+R_D(C_{GD}+C_{DB})]s+1}$$

$$=-g_mR_D\frac{(1-s/z_1)}{\left(\frac{s}{\omega_{p1}}+1\right)\left(\frac{s}{\omega_{p2}}+1\right)}=-g_mR_D\frac{(1-s/z_1)}{\frac{s^2}{\omega_{p1}\omega_{p2}}+\left(\frac{1}{\omega_{p1}}+\frac{1}{\omega_{p2}}\right)s+1}$$

- If $|\omega_{p1}| << |\omega_{p2}|$

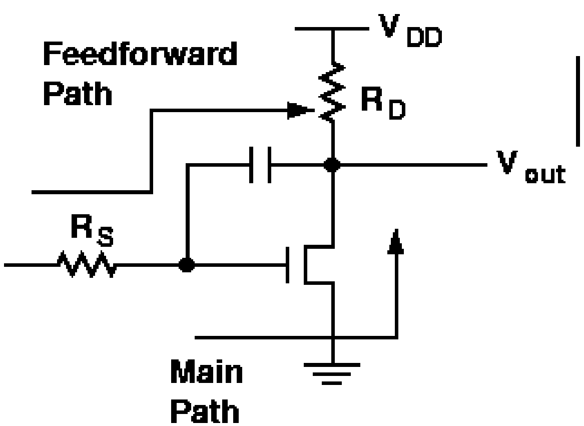
$$z_1=\frac{g_m}{C_{GD}},\quad \omega_{p1}=\frac{1}{R_S(1+g_mR_D)C_{GD}+R_SC_{GS}+R_D(C_{GD}+C_{DB})}$$

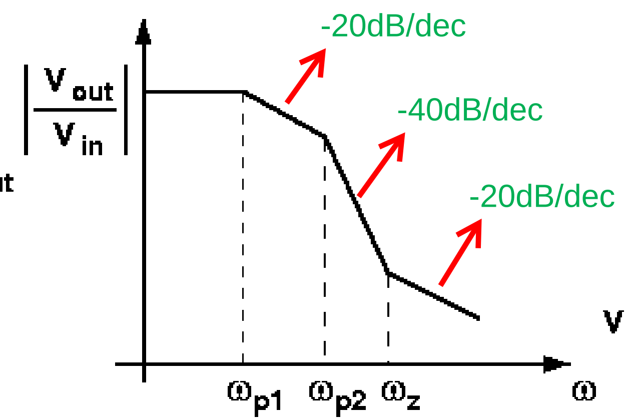
$$\omega_{p2}=\frac{1}{\omega_{p1}}\cdot\frac{1}{R_SR_D(C_{GS}C_{GD}+C_{GS}C_{DB}+C_{GD}C_{DB})}\cong\frac{R_SC_{GS}}{R_SR_D(C_{GS}C_{GD}+C_{GS}C_{DB})}=\frac{1}{R_D(C_{GD}+C_{DB})}$$

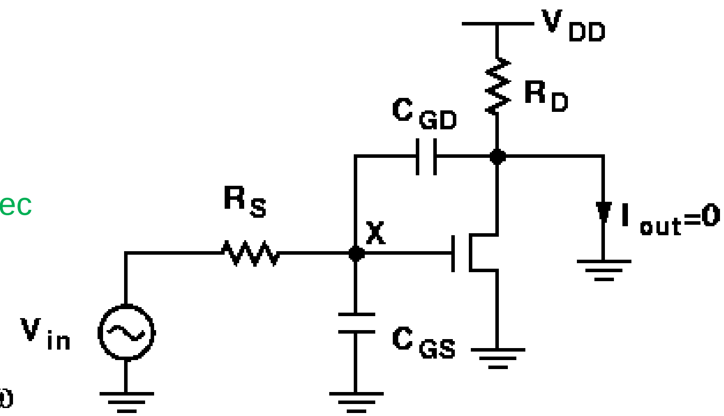
$$If\ \ C_{GS}>>(1+g_mR_D)C_{GD}+R_D(C_{GD}+C_{DB})/R_S$$

Right-Half Plane Zero

- C_{GD} provides a feed forward path that conducts the input signal to the output at very high frequencies.
- A zero in the right half plane causes stability issue in feedback Amp.







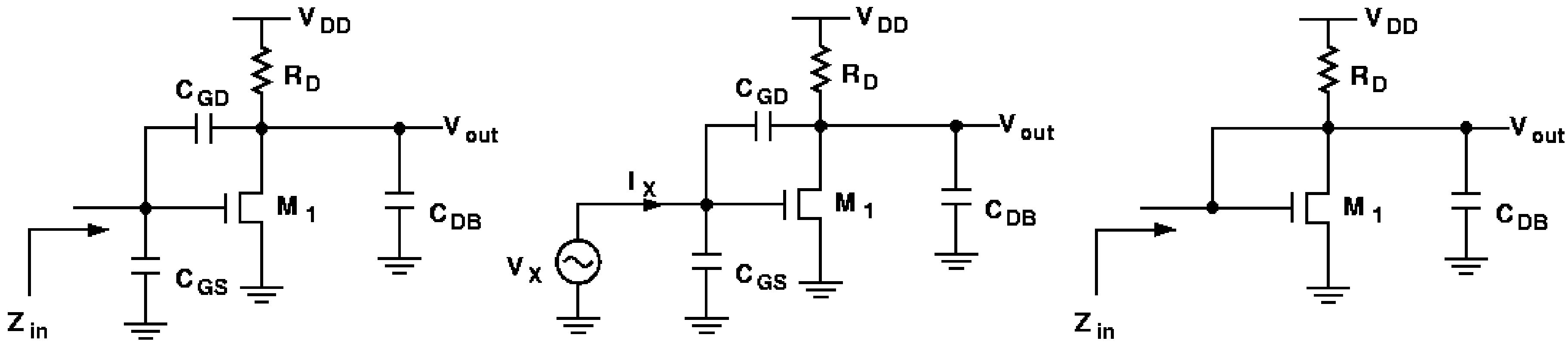
- The zero, s_z , can be computed by noting that the $V_{out}(s)/V_{in}(s)$ must drop to zero for $s = s_z$.
- The currents through C_{GD} and M_1 are equal and opposite.

$$V_x C_{GD} s_z = g_m V_x$$

$$s_z = \frac{g_m}{C_{GD}}$$

Input Impedance of a CS Stage

- As a first-order approximation $Z_{in} = \left\{\left[C_{GS} + (1+g_m R_D)\right] C_{GD}\right\}^{-1}$



- At high frequencies, take the effect of the output node into account

$$\text{Ignore } C_{GS}, \; (I_X - g_m V_X)\frac{R_D}{1+R_D C_{DB} s} + \frac{I_X}{C_{GD} s} = V_X \Rightarrow \frac{V_X}{I_X} = \frac{1+R_D(C_{GD}+C_{DB})s}{C_{GD}s(1+g_m R_D + R_D C_{DB} s)}$$

- At low frequencies $\quad Z_{in} \approx \dfrac{1}{sC_{GS}} \parallel \dfrac{1}{sC_{GD}(1+g_m R_D)}$

- At high frequencies $\quad Z(C_{GD}) = 0, \quad Z_{in} \approx \dfrac{1}{sC_{GS}} \parallel \dfrac{1}{g_m} \parallel R_D$ // 1/sCDB

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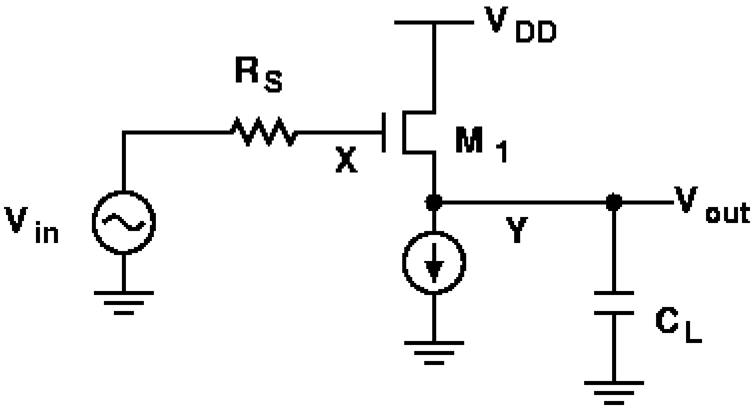
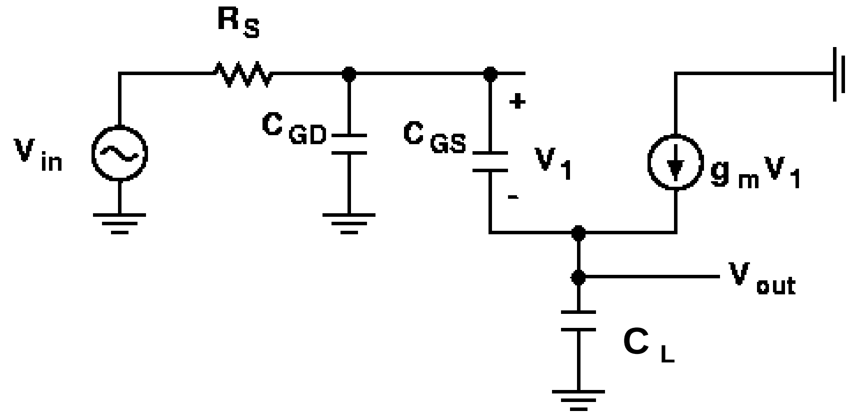
Source Follower

- Source followers are occasionally used as level shifters or buffers.





$$V_1 C_{GS} s + g_m V_1 = V_{out} s C_L \qquad V_1 = \frac{sC_L}{g_m + sC_{GS}} V_{out} \qquad V_{in} = R_S[V_1 C_{GS} s + (V_1 + V_{out}) C_{GD} s] + V_1 + V_{out}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{g_m + C_{GS} s}{R_S\left(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L\right)s^2 + \left(g_m R_S C_{GD} + C_L + C_{GS}\right)s + g_m}$$

$$= \frac{(1 + sC_{GS}/g_m)}{s^2 R_S\left(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L\right)/g_m + s\left(g_m R_S C_{GD} + C_L + C_{GS}\right)/g_m + 1}$$

➢ The signal conducted by C_{GS} at high frequencies adds with the same polarity to the signal produced by the intrinsic transistor. (*left half plane* zero)

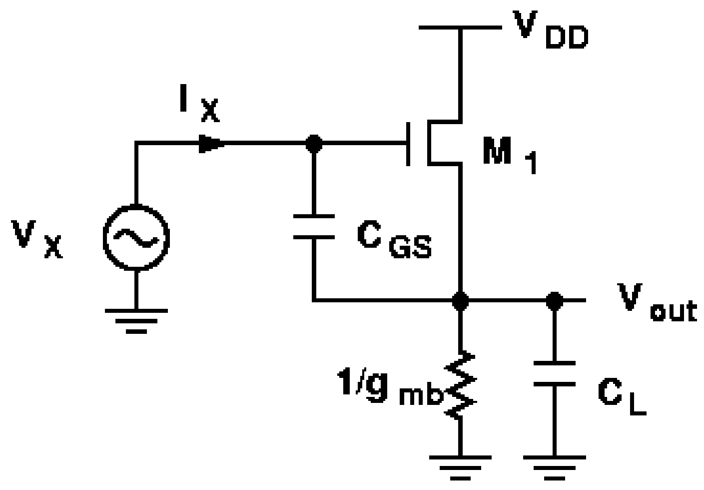
Input Impedance of Source Follower

$$\frac{V_{out}}{V_{in}}(s) = \frac{(1+sC_{GS}/g_m)}{s^2R_S\left(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L\right)/g_m + s\left(g_mR_SC_{GD} + C_L + C_{GS}\right)/g_m + 1}$$

- If the two poles are far apart

$$\omega_{p1} \approx \frac{g_m}{g_mR_SC_{GD} + C_L + C_{GS}} = \frac{1}{R_SC_{GD} + \dfrac{C_L + C_{GS}}{g_m}} \qquad If\ R_S = 0,\ \text{then}\ \omega_{p1} = \frac{g_m}{C_L + C_{GS}}$$

- Input impedance (ignore C_{GD})



$$V_X = \frac{I_X}{C_{GS}s} + \left(I_X + \frac{g_mI_X}{C_{GS}s}\right)\left(\frac{1}{g_{mb}} \parallel \frac{1}{sC_L}\right)$$

$$Z_{in} = \frac{1}{C_{GS}s} + \left(1 + \frac{g_m}{C_{GS}s}\right)\frac{1}{g_{mb} + C_Ls}$$

At low frequencies, $g_{mb} >> \ |C_Ls|$

$$Z_{in} = \frac{1}{C_{GS}s}\left(1 + \frac{g_m}{g_{mb}}\right) + \frac{1}{g_{mb}}$$

$$C_{in} = C_{GS}\frac{g_{mb}}{g_{mb} + g_m}$$

Input Impedance of Source Follower

- At relatively low frequencies, the equivalent input C_{in} = $\dfrac{C_{GS}g_{mb}}{g_m+g_{mb}}$
  - By Miller effect

$$A_v=\frac{g_m}{g_m+g_{mb}} \qquad C_{eq}=C_{GS}\left[1-\frac{g_m}{g_m+g_{mb}}\right]=\frac{C_{GS}g_{mb}}{g_m+g_{mb}}$$

  - The overall input capacitance is equal to C_{GD} plus a fraction of C_{GS}.
- At high frequencies,

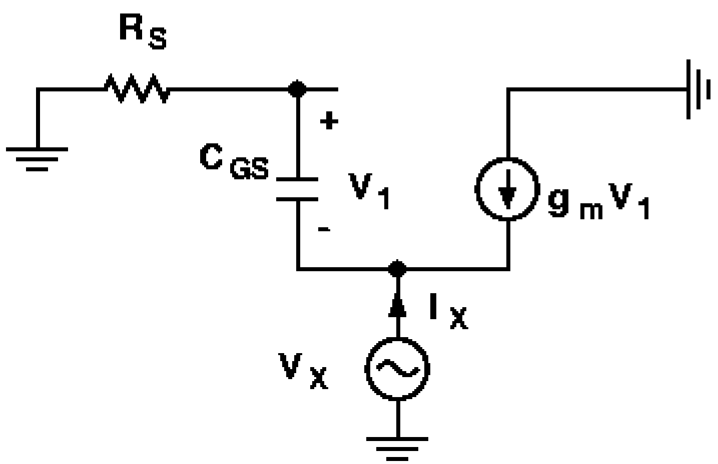
$$Z_{in}=\frac{1}{C_{GS}s}+\left(1+\frac{g_m}{C_{GS}s}\right)\frac{1}{g_{mb}+C_Ls}$$

$$g_{mb}<<\left|sC_L\right| \text{ and } Z_{in}\approx\frac{1}{sC_{GS}}+\frac{1}{sC_L}+\frac{g_m}{s^2C_{GS}C_L}=\frac{1}{j\omega C_{GS}}+\frac{1}{j\omega C_L}-\frac{g_m}{\omega^2C_{GS}C_L}$$

  - The input impedance consists of the series combination of capacitors C_{GS} and C_L and a negative resistance.
  - The negative resistance property can be utilized in oscillators.

Output Impedance of Source Follower

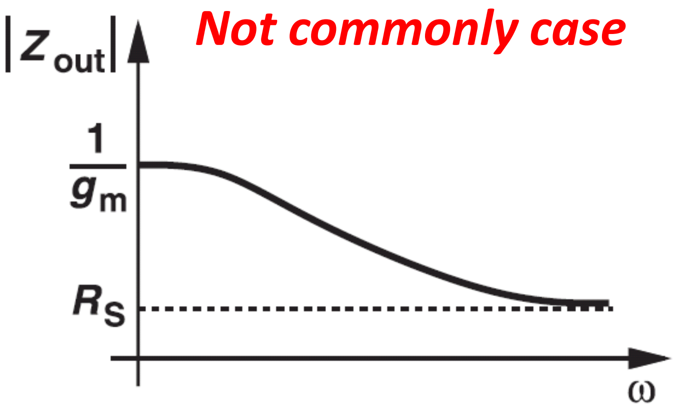
- Neglecting g_{mb} , C_{SB} & C_{GD}

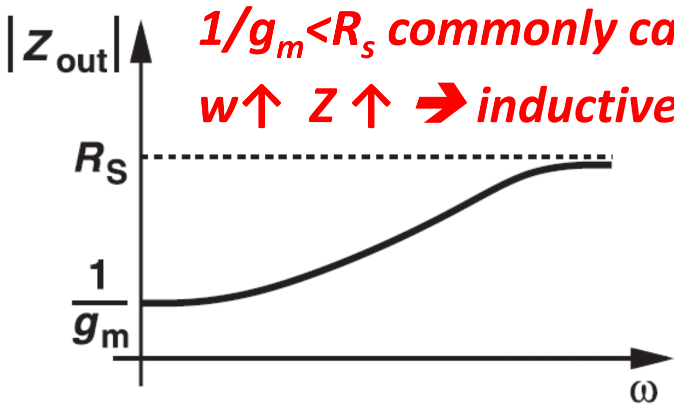


$$V_1 C_{GS} s + g_m V_1 = -I_X , \quad V_1 C_{GS} s R_S + V_1 = -V_X$$

$$Z_{out} = \frac{V_X}{I_X} = \frac{s R_S C_{GS} + 1}{g_m + s C_{GS}}$$

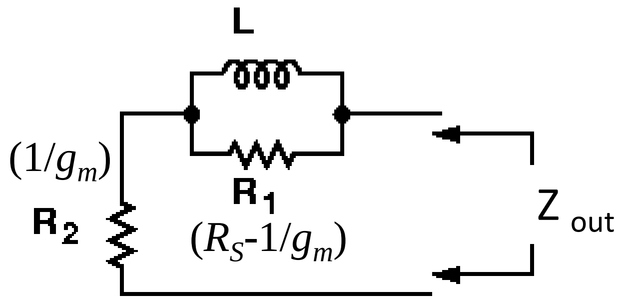
- At low frequencies, $Z_{out} \approx 1/g_m$
- At high frequencies, $Z_{out} \approx R_S$ with $Z(C_{GS}) = 0$
- If $1/g_m < R_S$, the output impedance contains an inductive component.





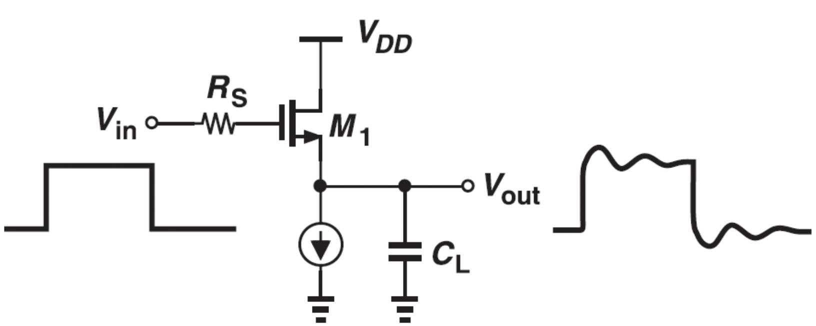
Output Impedance of SF

- Equivalent network of output impedance



- ➢ At low frequencies, $Z_{out} = 1/g_m$ when $\omega = 0$
- ➢ At high frequencies, $Z_{out} = R_S$ when $\omega = \infty$
- ➢ Therefore, $R_2 = 1/g_m \quad R_1 = R_S - 1/g_m$

$$Z_{out} - \frac{1}{g_m} = \frac{sC_{GS}(R_S - 1/g_m)}{g_m + sC_{GS}}, \quad \frac{1}{Z_{out} - \frac{1}{g_m}} = \frac{1}{R_S - 1/g_m} + \frac{1}{\frac{sC_{GS}}{g_m}(R_S - 1/g_m)}, \quad L = \frac{C_{GS}}{g_m}(R_S - 1/g_m)$$



- ➢ If a source follower is driven by a large resistance, then it exhibits substantial inductive behavior.
- ➢ This effect manifests itself as ringing in the step response. If it drive a large C_L.

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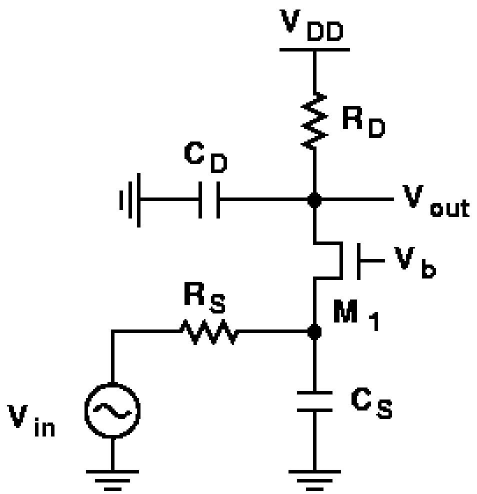
Common-Gate Stage

- Neglect channel length modulation effect



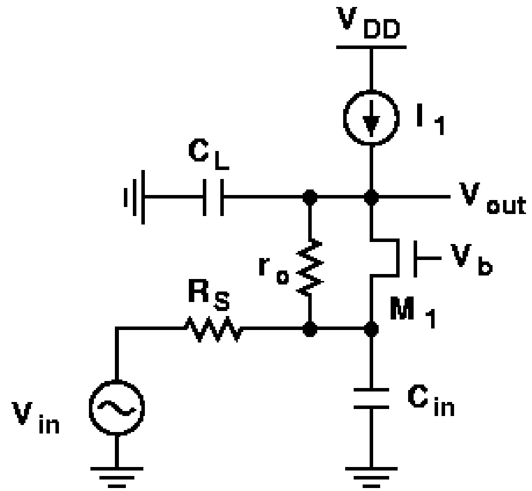
$$\frac{V_{out}}{V_{in}}(s)=\frac{(g_m+g_{mb})R_D}{1+(g_m+g_{mb})R_S}\frac{1}{\left(1+\frac{C_S}{g_m+g_{mb}+R_S^{-1}}s\right)(1+R_D C_D s)}$$

- It exhibits no Miller multiplication of capacitances, potentially achieving wide band.
- Low Input impedance may load the preceding stage.
- The DC level of the input signal must be quite low.

$$if\ \lambda\neq 0,\qquad Z_{in}\approx\frac{Z_L}{(g_m+g_{mb})r_O}+\frac{1}{g_m+g_{mb}},\qquad Z_L=R_D\|[1/sC_D]$$

- Since Z_{in} now depends on Z_L, it is difficult to associate a pole with the input node.

Input Impedance of Common Gate



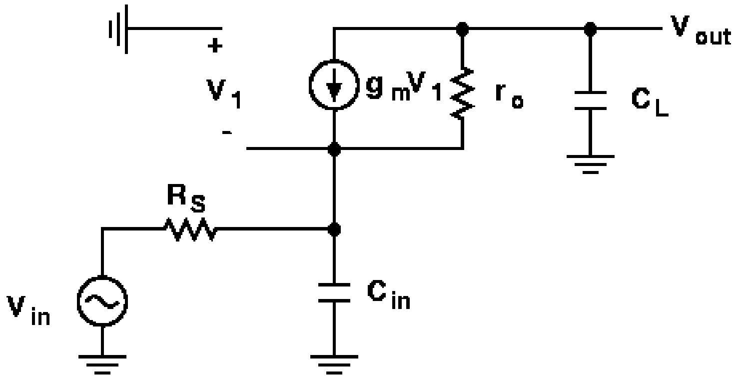
$$(-V_{out}C_L s + V_1 C_{in} s)R_S + V_{in} = -V_1$$

$$V_1 = -\frac{-V_{out}C_L s R_S + V_{in}}{1 + C_{in}R_S s}, \quad r_O(-V_{out}C_L s - g_m V_1) - V_1 = V_{out}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{1 + g_m r_O}{r_O C_L C_{in} R_S s^2 + [r_O C_L + C_{in} R_S + (1 + g_m r_O) C_L R_S]s + 1}$$

$$\frac{V_{out}}{V_{in}} = 1 + g_m r_O \,(\text{@ Low Freq}: s = jw = 0)$$

- The body effect can be included by simply replacing g_m with $g_m + g_{mb}$.



$$Z_{in} \approx \frac{Z_L}{(g_m + g_{mb})r_O} + \frac{1}{g_m + g_{mb}}$$

$$Z_L = \frac{1}{sC_L}, \quad Z_{in} = \frac{1}{g_m + g_{mb}} + \frac{1}{sC_L} \cdot \frac{1}{(g_m + g_{mb})r_O}$$

$$\text{As } C_L \text{ or } s \text{ increases, } Z_{in} = \frac{1}{g_m + g_{mb}}$$

$$\text{input pole } \omega_{p,in} = \frac{1}{\left(R_S \parallel \frac{1}{g_m + g_{mb}}\right)C_{in}}, \text{independent of } C_L$$

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Cascode Stage

- Common source + common gate gain stage.
- Increase the voltage gain of amplifiers and the output impedance of current source.

$$Z_{out} = (1 + g_{m2} r_{O2}) Z_X + r_{O2}$$

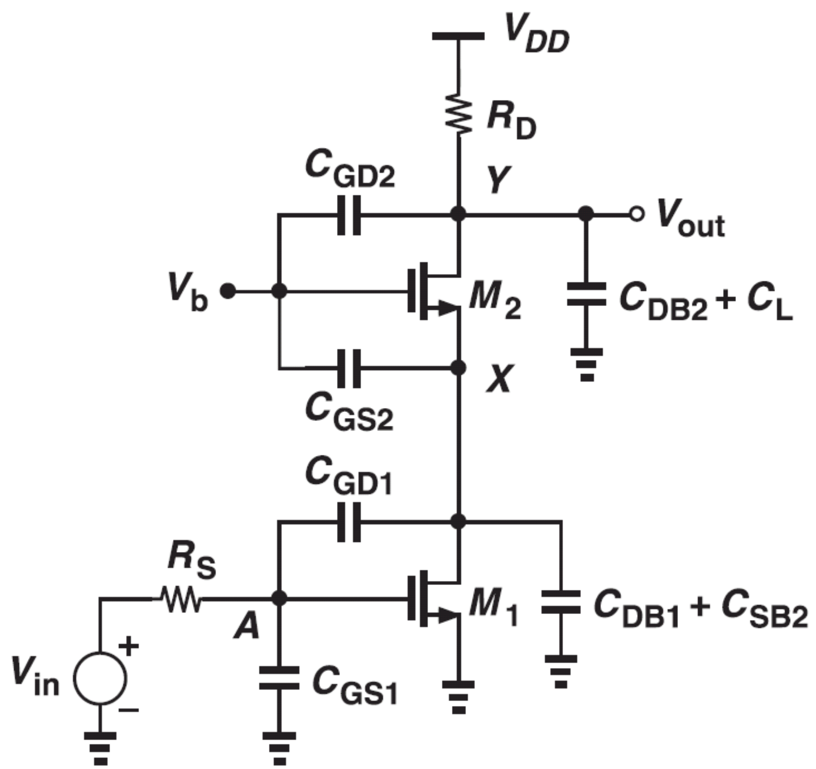
- Providing shielding – suppressing the Miller effect.
- The low frequency gain from M_1 gate to drain

$$A_v = -\frac{g_{m1}}{g_{m2} + g_{mb2}} \approx -1$$

- The input pole is at

$$\omega_{p,A} = \frac{1}{R_S \left[C_{GS1} + \left(1 + \frac{g_{m1}}{g_{m2} + g_{mb2}}\right) C_{GD1} \right]}$$

gain Av



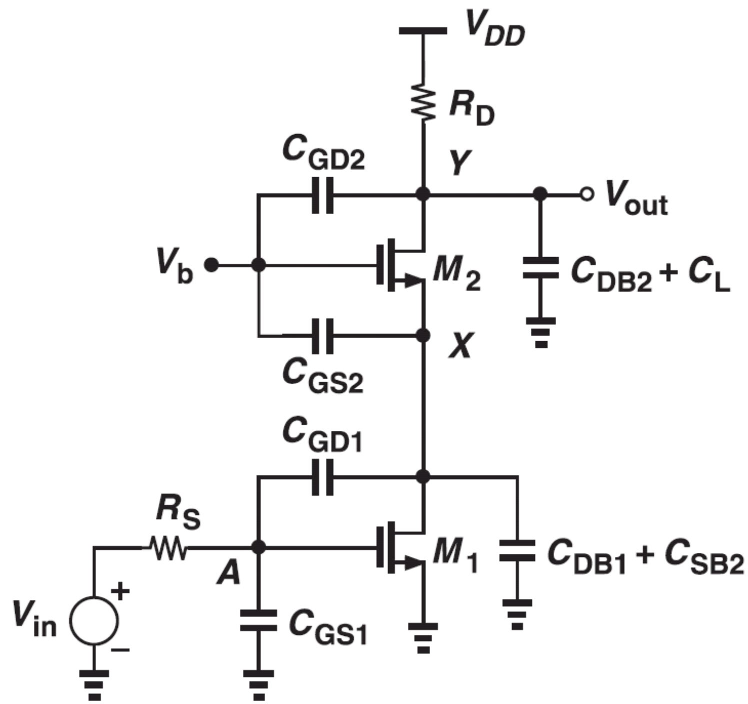
Miller effect被大幅降低

Cascode Stage

- The total capacitance at node X is (typically chosen farther from the origin than the other two)

$$C_X = 2C_{GD1} + C_{DB1} + C_{SB2} + C_{GS2}$$

$$\omega_{p,X} = \frac{1}{R_X C_X} = \frac{g_{m2} + g_{mb2}}{2C_{GD1} + C_{DB1} + C_{SB2} + C_{GS2}}$$



- The output pole at node Y is

$$\omega_{p,Y} = \frac{1}{R_D\left(C_{DB2} + C_L + C_{GD2}\right)}$$

- Output impedance of a cascode current source (neglect C_Y, C_{GD1})

$$Z_{out} = (1 + g_{m2} r_{O2}) Z_X + r_{O2}, Z_X = r_{O1} \| (C_X s)^{-1}$$

$$pole = (r_{O1} C_X)^{-1}$$

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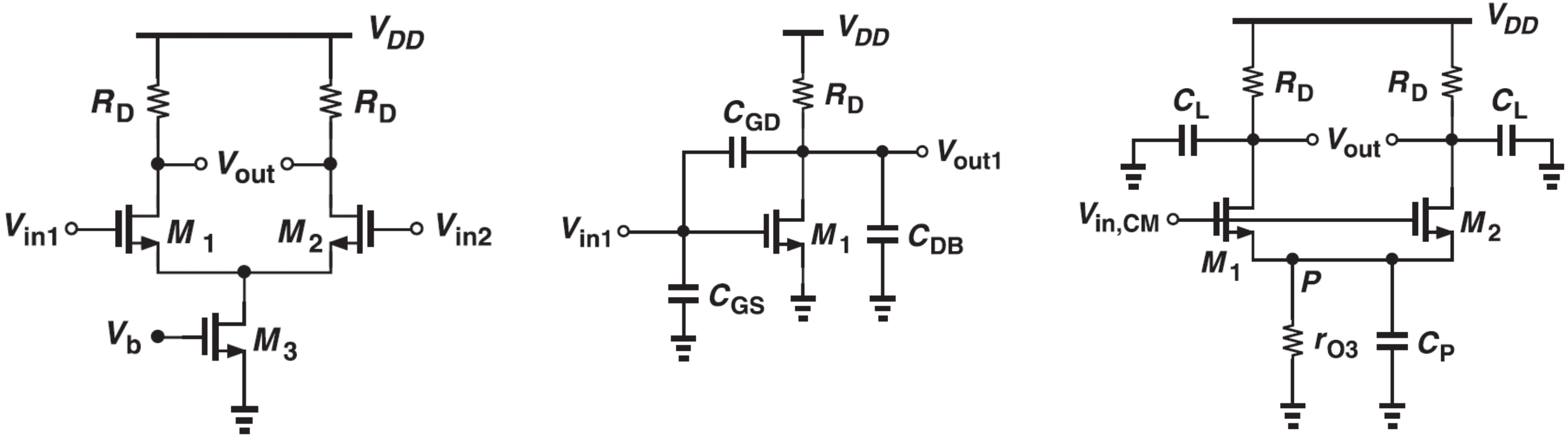
Differential Pair



- For differential signals, the response is identical to that of a CS stage.
- For CM gain, the total capacitance at node P determines the HF gain.

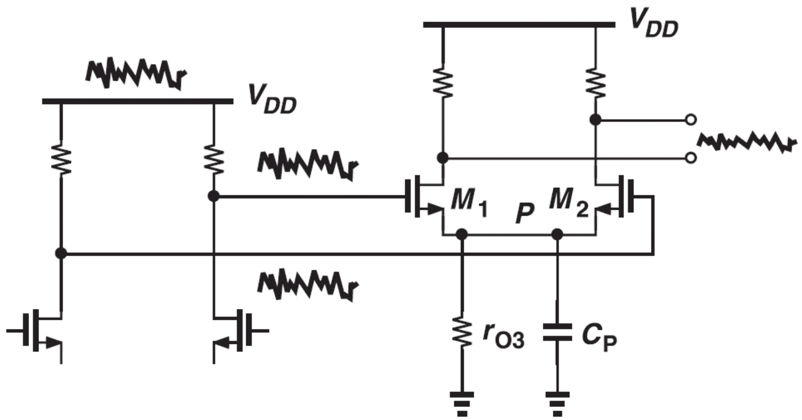
$$A_{v,CM} = -\Delta g_m \left[R_D \| \left(\frac{1}{C_L s} \right) \right] \Big/ \left\{ (g_{m1} + g_{m2}) \left[r_{O3} \| \left(\frac{1}{C_P s} \right) \right] + 1 \right\}$$

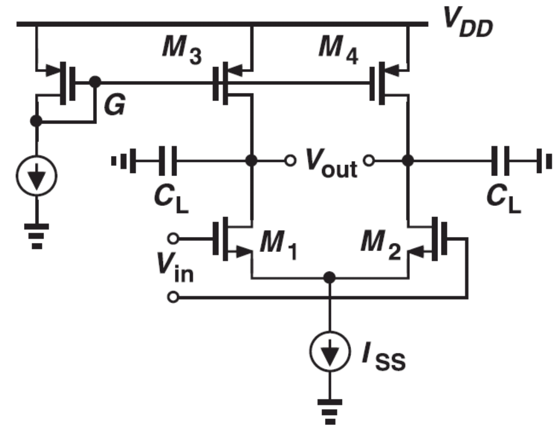
Av,cm_dm

CMRR會隨著頻率越高而越差

- If the output pole >> the pole at the tail node of current source, the CM rejection of the circuit degrades considerably at high frequencies (HF).
- If the V_{DD} contains HF noise and the circuit exhibits mismatches, the resulting CM disturbance at node P leads to a differential noise component at the output.

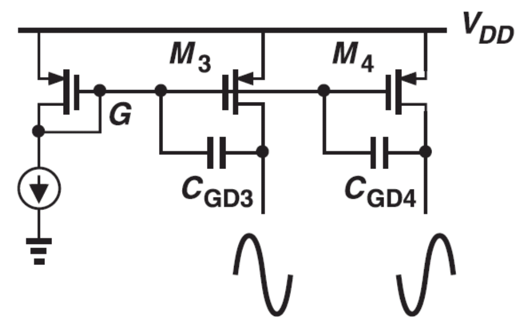
Effect of High-Frequency Supply Noise

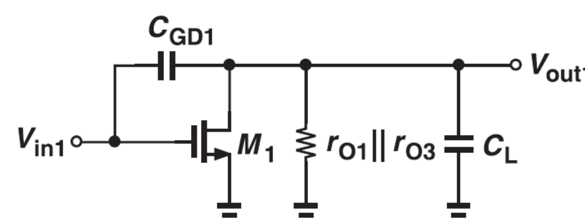




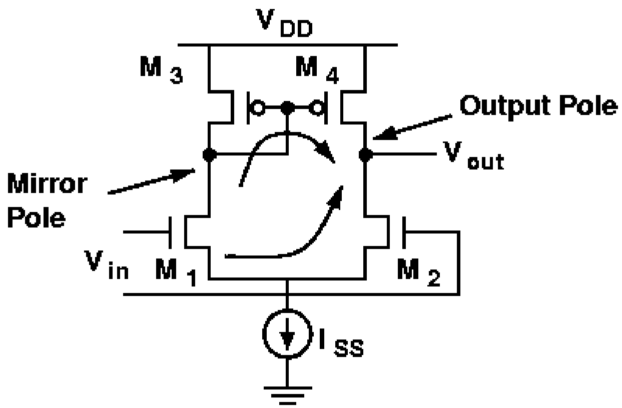
- Trade-off between voltage headroom and CMRR.
- To minimize voltage headroom, M_{tail} ↑, C_p ↑,high frequency CMRR ↓.
- For differential output signals, C_{GD3} and C_{GD4} conduct equal and opposite currents to G, making this node an AC ground.
- The dominant pole is at

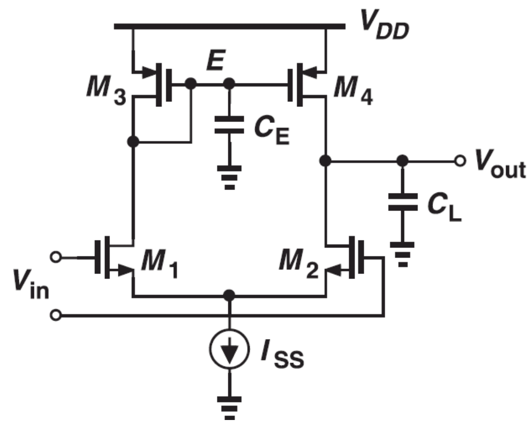
$$\omega_p = \frac{1}{[r_{O1} \| r_{O3}]C_L}$$

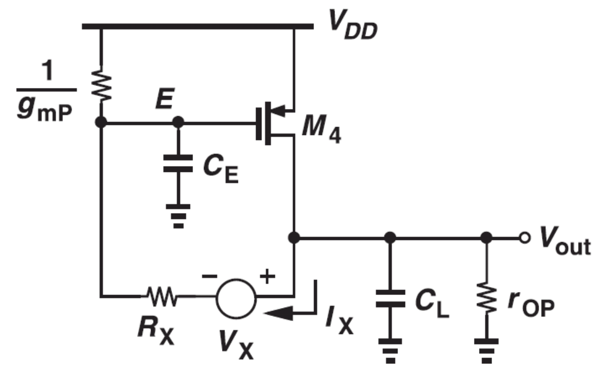




DP with Active Current Mirror







- This topology contains two signal paths with different transfer functions.
- Pole at node *E (Mirror pole)*

$$\omega_{pE}=\frac{g_{m3}}{C_E},\quad C_E \approx C_{GS3}+C_{GS4}+C_{DB3}+C_{DB1}+2C_{GD1}+g_{m4}\left(r_{O4}\parallel r_{O2}\right)C_{GD4}$$

$$V_X=g_{mN}r_{ON}V_{in},\quad R_X=2r_{ON},\quad V_E=\left(V_{out}-V_X\right)\frac{\left(C_E s+g_{mP}\right)^{-1}}{\left(C_E s+g_{mP}\right)^{-1}+R_X},\quad -g_{m4}V_E-I_X=V_{out}\left(sC_L+r_{OP}^{-1}\right)$$

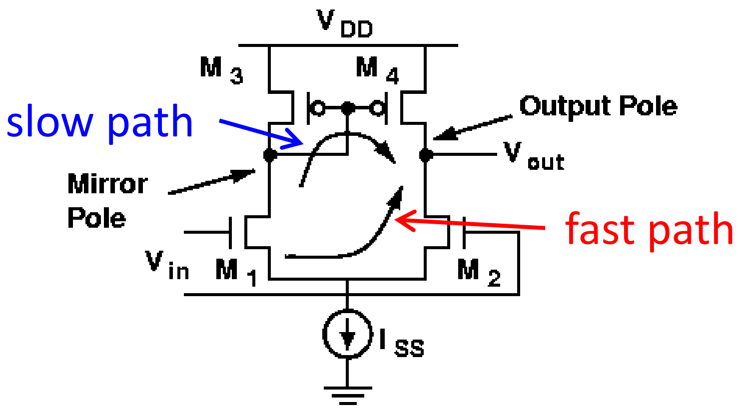
$$\frac{V_{out}}{V_{in}}=\frac{g_{mN}r_{ON}r_{OP}(2g_{mP}+sC_E)}{2r_{OP}r_{ON}C_E C_L s^2+\left[\left(2r_{ON}+r_{OP}\right)C_E+r_{OP}\left(1+2g_{mP}r_{ON}\right)C_L\right]s+2g_{mp}\left(r_{ON}+r_{OP}\right)}$$

$$\omega_{P1}\approx\frac{2g_{mP}\left(r_{ON}+r_{OP}\right)}{\left(2r_{ON}+r_{OP}\right)C_E+r_{OP}(1+2g_{mp}r_{ON})C_L}\approx\frac{1}{\left(r_{ON}\parallel r_{OP}\right)C_L}\quad \text{if}\quad 2g_{mP}r_{ON}>>1\qquad \omega_{p2}\approx\frac{g_{mP}}{C_E}$$

DP with Active Current Mirror

- To find zero
  - Consider that the circuit consists of a "slow path" (M_1, M_3, and M_4) in parallel with a "fast path" (M_1, and M_2)
  - Slow path

$$A_{v1} = \frac{A_0}{\left(1+\frac{s}{p_1}\right)\left(1+\frac{s}{p_2}\right)}$$



  - Fast path

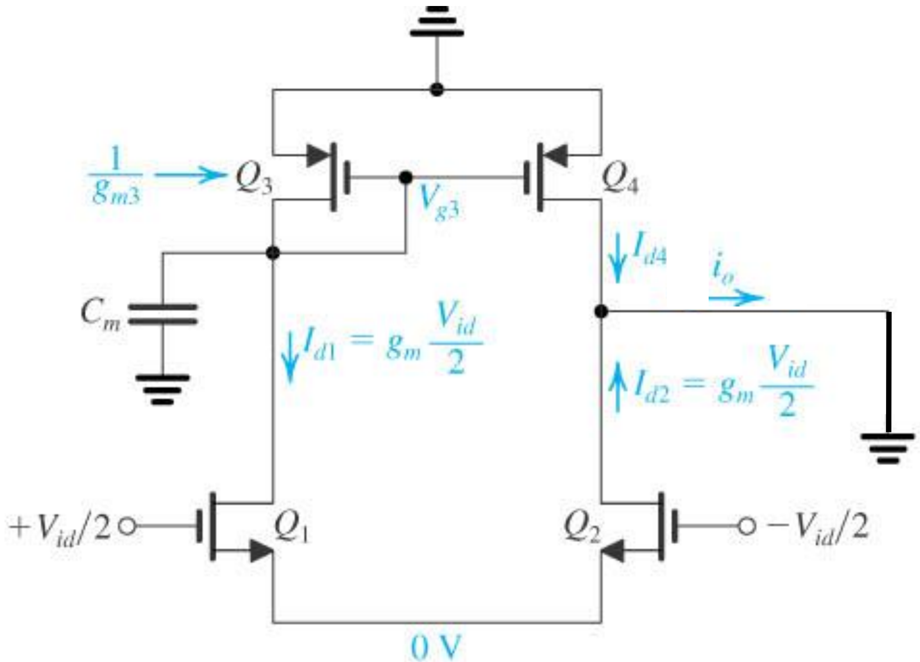
$$A_{v2} = \frac{A_0}{\left(1+\frac{s}{p_1}\right)}$$

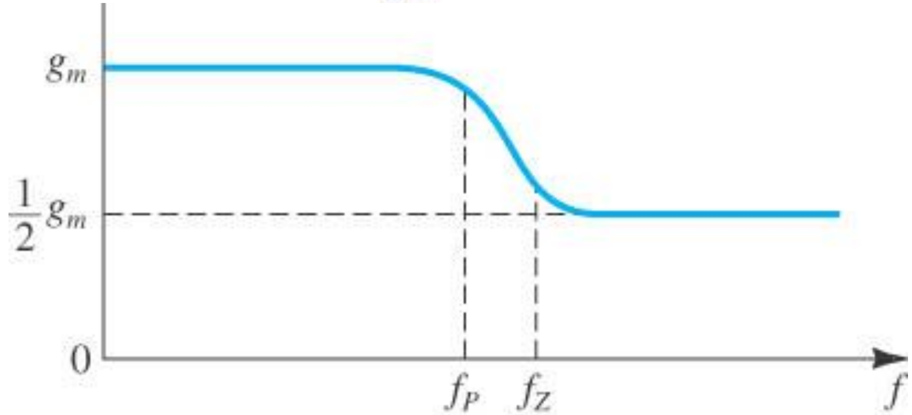
  - We have

$$\frac{V_{out}}{V_{in}} = A_{v1} + A_{v2} = \frac{A_0}{1+\frac{s}{p_1}}\left(1+\frac{1}{1+\frac{s}{p_2}}\right) = \frac{A_0\left(2+\frac{s}{p_2}\right)}{\left(1+\frac{s}{p_1}\right)\left(1+\frac{s}{p_2}\right)} \quad \Rightarrow \quad z = 2p_2$$

  - Fully differential pair has no mirror pole, which is better than differential-to-single ended amp.

G_m Frequency Response





At low frequency: C_m is OPEN

$$i_{d4} = i_{d1} = g_m \frac{V_{id}}{2}, i_{d2} = g_m \frac{V_{id}}{2}$$

$$i_o = i_{d4} + i_{d2} = g_m V_{id}$$

$$G_m = i_o / V_{id} = g_m$$

At high frequency: C_m is SHORT

$$V_{g3} = 0, i_{d4} = 0, i_o = i_2 = g_m \frac{V_{id}}{2}$$

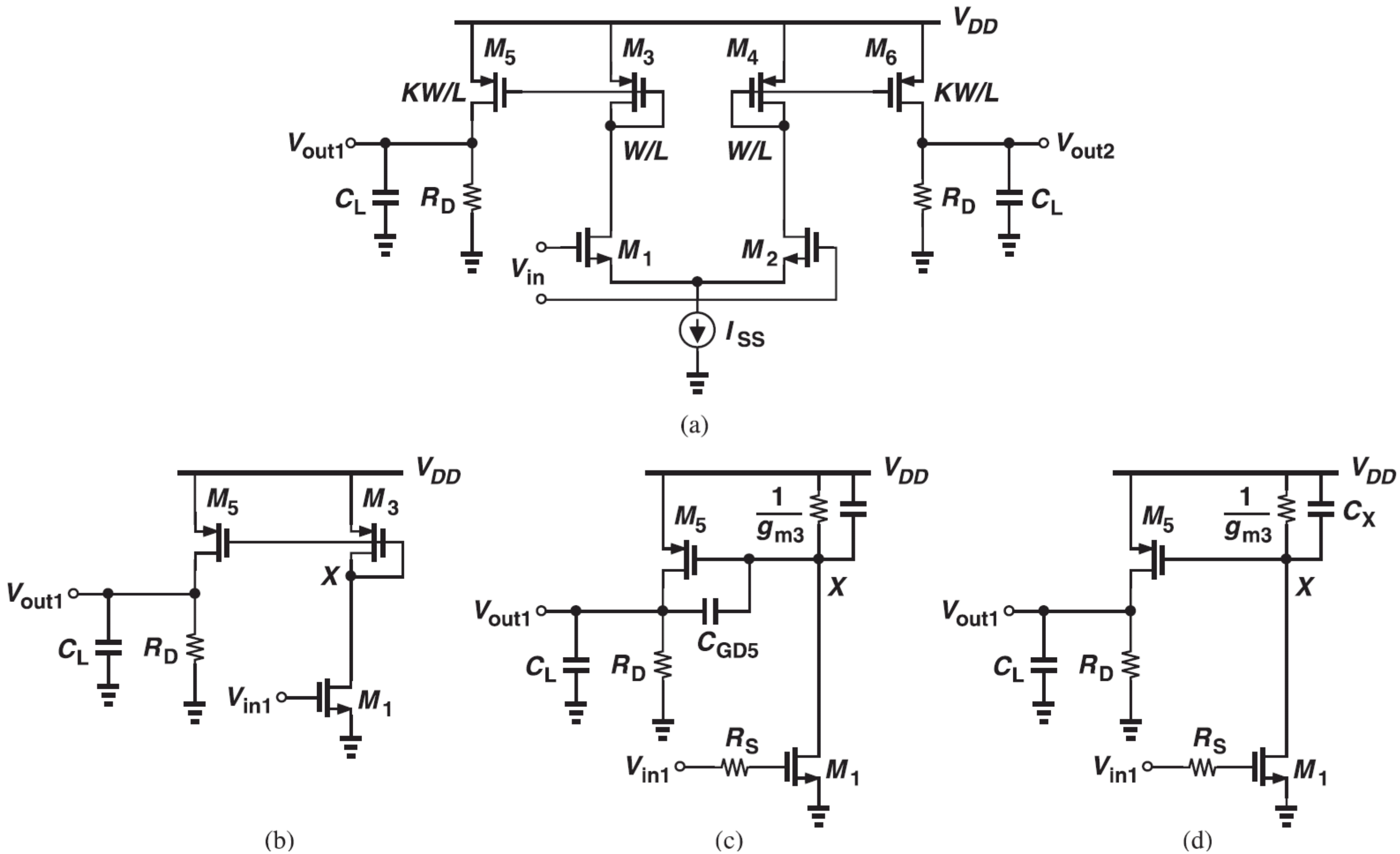
$$G_m = i_o / V_{id} = g_m / 2$$

G_m transfer function

$$G_m \equiv \frac{i_o}{V_{id}} = g_m \left(\frac{1 + sC_m / 2g_{m3}}{1 + sC_m / g_{m3}}\right)$$

$$f_P = \frac{g_{m3}}{2\pi C_m}, f_Z = \frac{2g_{m3}}{2\pi C_m}$$

Fully Differential Amplifier



$$\frac{V_{out1}}{V_X}(s) = -g_{m5}R_D\frac{1}{1+sR_DC_L}$$

$$C_X \approx C_{GS3} + C_{GS5} + C_{DB3} + C_{GD5}(1+g_{m5}R_D) + C_{DB1}$$