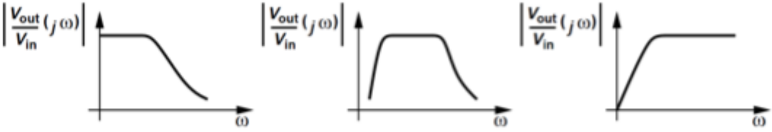
Chapter 6: Frequency Response of Amplifiers

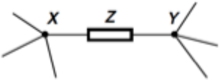
General Considerations



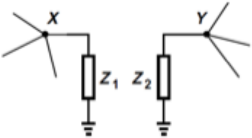
(a) (b) (c)

- In this chapter, we are primarily interested in the magnitude of the transfer function.
- The magnitude of a complex number a + jb is given by $\sqrt{a^2 + b^2}$
- Zeros and poles are respectively defined as the roots of the numerator and denominator of the transfer function.

Miller effect



(a)



(b)

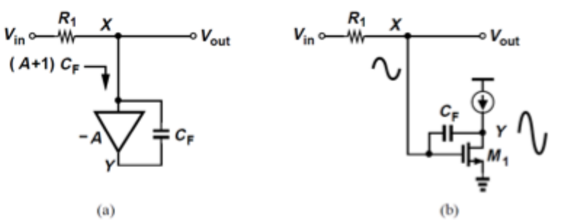
- If the circuit of Fig (a) can be converted to that of Fig (b), then

$$Z_1 = Z/(1 - A_v)$$

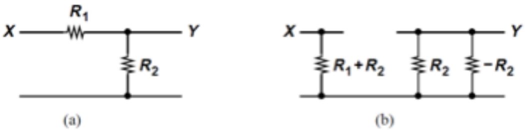
$$Z_2 = Z/(1 - A_v^{-1})$$

$$A_v = V_Y/V_X$$

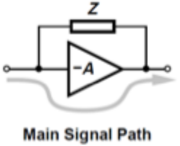
Example



- **A student needs a large Capacitor and decides to utilize the Miller multiplication**
- **What is the issues in this approach?**



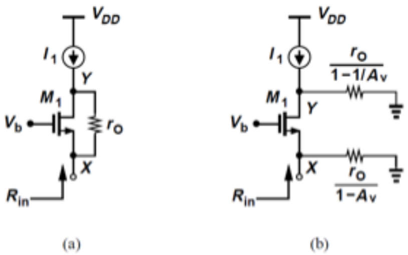
Improper application of Miller's theorem



Typical case for valid application of Miller's theorem.

- **Miller's theorem does not stipulate the conditions under which this conversion is valid.**
- **If the impedance Z forms the only signal path between X and Y , then the conversion is often invalid.**

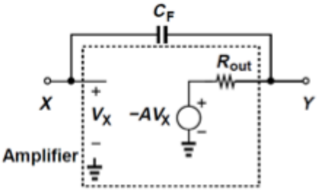
Example



- **Calculate the input resistance of the circuit shown.**
- **Since Av is usually greater than unity,** $r_O/(1-A_v)$ **is a negative resistance.**
- **Av =** $1+(g_m+g_{mb})r_O$

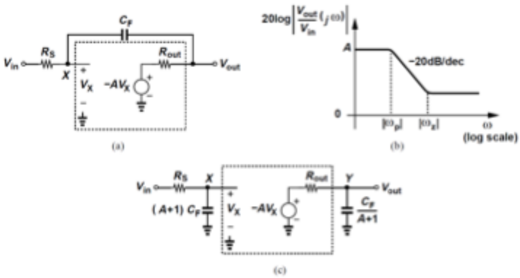
$$\begin{aligned}
R_{in} &= \frac{r_O}{1-[1+(g_m+g_{mb})r_O]} \Big\| \frac{1}{g_m+g_{mb}} \\
&= \frac{-1}{g_m+g_{mb}} \Big\| \frac{1}{g_m+g_{mb}} \\
&= \infty.
\end{aligned}$$

Example



- The value of Av = VY / VX must be calculated at the frequency of interest.
- In the figure, the equivalent circuit $V_Y \neq -AV_X$ that at high frequencies.
- In many cases we use the low-frequency value of VY / VX to gain insight.
- We call this approach "Miller's approximation."

Example



- **Direct Calculation:**

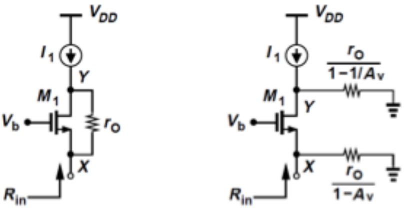
$$\frac{V_{out}}{V_{in}}(s) = \frac{R_{out}C_F s - A}{[(A+1)R_S + R_{out}]C_F s + 1}$$

- **Miller Aproximation:**

$$\frac{V_{out}}{V_{in}}(s) = \frac{-A}{[(1+A)R_S C_F s + 1]\left(\frac{1}{1+A^{-1}}C_F R_{out} s + 1\right)}$$

- **Miller's approximation has eliminated the zero and predicted two poles for the circuit!**

Example

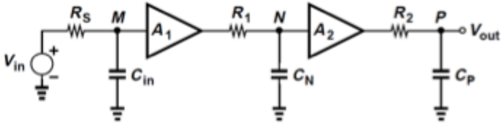


(a) (b)

$$R_{out} = \frac{r_O}{1 - 1/A_v}$$
$$= \frac{r_O}{1 - [1 + (g_m + g_{mb})r_O]^{-1}}$$
$$= \frac{1}{g_m + g_{mb}} + r_O,$$

- **Actual Rout =r0**
- **Miller's approximation:**
- **(1) it may eliminate zeros**
- **(2) it may predict additional poles**
- **(3) it does not correctly compute the "output" impedance.**

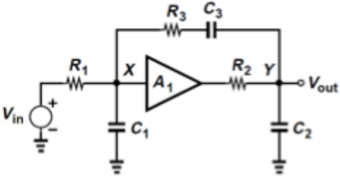
Association of Poles with Nodes

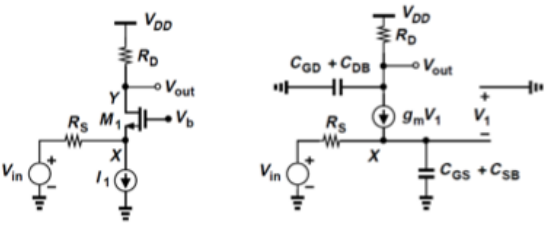


- The overall transfer function can be written as

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

- Each node in the circuit contributes one pole to the transfer function.
- Not valid in general. Example:





(a) (b)

- **At node X:**

$$\omega_{in} = \left[(C_{GS} + C_{SB}) \left(R_S \middle\| \frac{1}{g_m + g_{mb}} \right) \right]^{-1}$$

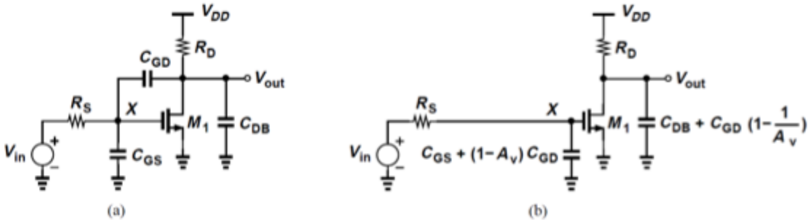
- **At node Y:**

$$\omega_{out} = [(C_{DG} + C_{DB}) R_D]^{-1}$$

- **The overall transfer function…**

$$\frac{V_{out}}{V_{in}}(s) = \frac{(g_m + g_{mb}) R_D}{1 + (g_m + g_{mb}) R_S} \cdot \frac{1}{\left(1 + \frac{s}{\omega_{in}}\right)\left(1 + \frac{s}{\omega_{out}}\right)}$$

Common-Source Stage



- **The magnitude of the “input” pole**

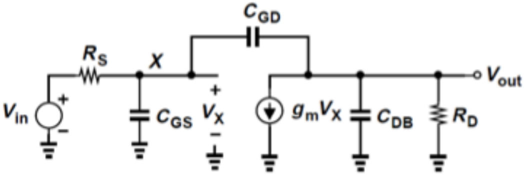
$$\omega_{in} = \frac{1}{R_S[C_{GS} + (1 + g_m R_D)C_{GD}]}$$

- **At the output node**

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

$$\frac{V_{out}}{V_{in}}(s) = \frac{-g_m R_D}{\left(1 + \frac{s}{\omega_{in}}\right)\left(1 + \frac{s}{\omega_{out}}\right)}$$

Direct Analysis



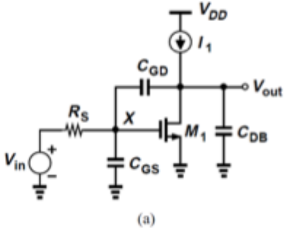
$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})]s + 1}$$

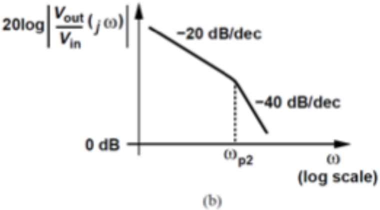
- While the denominator appears rather complicated, it can yield intuitive expressions for the two poles. $|\omega_{p1}| \ll |\omega_{p2}|$
- "Dominant pole" approximation.

$$\omega_{p1} = \frac{1}{R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})}$$

- The intuitive approach provides a rough estimate with much less effort.

Example





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

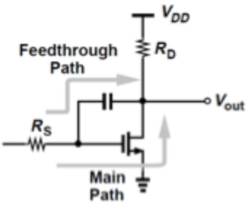
$$= \frac{C_{GD}s - g_m}{s[R_S(C_{GS}C_{GD} + C_{GS}C_{DB} + C_{GD}C_{DB})s + (g_m R_S + 1)C_{GD} + C_{DB}]}$$

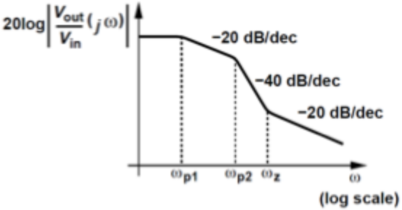
- **One pole is at the origin because the dc gain is infinity.**
- **For a large CDB or load capacitance**

$$\omega_{p2} \approx \frac{1}{R_S(C_{GS} + C_{GD})}$$

- **No miller multiplication. Why?**

Zero in Transfer Function





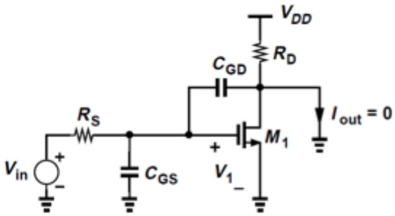
$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})]s + 1}$$

- **The transfer function of exhibits a zero given by**

$$\omega_z = +g_m/C_{GD}$$

- **C_{GD} provides a feedthrough path that conducts the input signal to the output at very high frequencies.**

Calculating zero in a CS stage



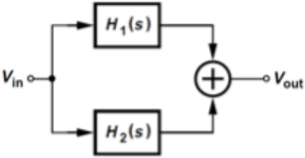
- The transfer function Vout(s)=Vin(s) must drop to zero for s = s_z.
- Therefore, the currents through C_{GD} and M1 are equal and opposite:

$$V_1 C_{GD} s_z = g_m V_1$$

- That is

$$s_z = +g_m / C_{GD}$$

Example

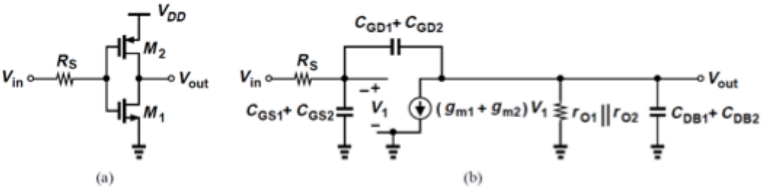


- **Can this (the zero) occur if H1(s) and H2(s) are first-order low-pass circuits?**
- **H1=** $A_1/(1+s/\omega_{p1})$ **and H2=** $A_2/(1+s/\omega_{p2})$

$$\frac{V_{out}}{V_{in}}(s) = \frac{(\frac{A_1}{\omega_{p2}} + \frac{A_2}{\omega_{p1}})s + A_1 + A_2}{(1+\frac{s}{\omega_{p1}})(1+\frac{s}{\omega_{p2}})}$$

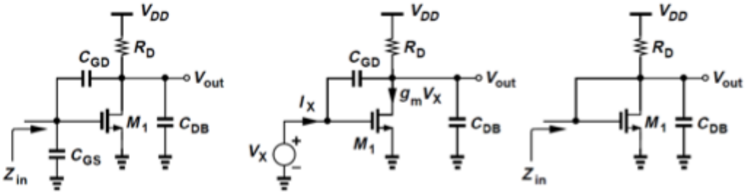
- **The overall transfer function contains a zero.**

Example



- **Since the corresponding terminals ofM1 and M2 are shorted to one another in the small-signal model, we merge the two transistors.**
- **The circuit thus has the same transfer function as the simple CS stage.**

Miller's Approximation



(a) (b) (c)

- **With the aid of Miller's approximation,**

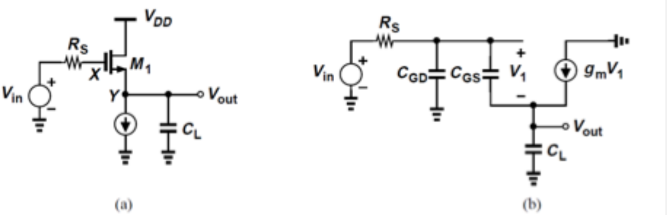
$$Z_{in} = \frac{1}{[C_{GS} + (1 + g_m R_D) C_{GD}]s}$$

- **But at high frequencies, the effect of the output node capacitance must be taken into account.**
- **Ignore C_{GS}**

$$\frac{V_X}{I_X} = \frac{1 + R_D(C_{GD} + C_{DB})s}{C_{GD}s(1 + g_m R_D + R_D C_{DB} s)}$$

- **if C_{GD} is large, it provides a low impedance path between the gate and drain of M1.**

Source Followers

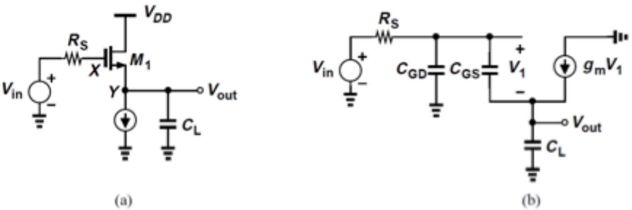


- The strong interaction between nodesX and Y through C_{GS} in makes it difficult to associate a pole with each node.

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

- Contains a zero in the left half plane. Why?

Example

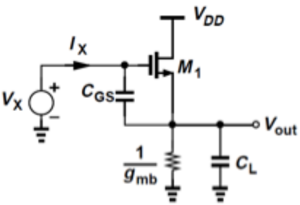


- **Transfer function if CL = 0?**

$$
\begin{aligned}
\frac{V_{out}}{V_{in}} &= \frac{g_m + C_{GS}s}{R_S C_{GS} C_{GD} s^2 + (g_m R_S C_{GD} + C_{GS})s + g_m} \\
&= \frac{g_m + C_{GS}s}{(1 + R_S C_{GD} s)(g_m + C_{GS}s)} \\
&= \frac{1}{1 + R_S C_{GD} s}.
\end{aligned}
$$

- **C_{GS} disappear**
- **In the absence of channel-length modulation and body effect, the voltage gain from the gate to the source is equal to unity.**

Input impedance



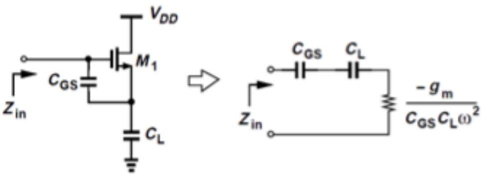
- C_{GD} simply shunts the input and can be ignored initially.

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

- If gmb = 0 and CL = 0, then Zin = ∞
- C_{GS} is entirely bootstrapped by the source follower and draws no current from the input.
- At Low frequency the overall input capacitance is equal to C_{GD} plus a fraction of C_{GS}.

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

Input Impedance

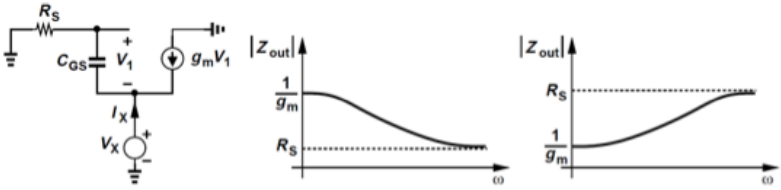


- **At high frequencies,** $g_{mb} \ll |C_L s|$

$$Z_{in} \approx \frac{1}{C_{GS}s} + \frac{1}{C_L s} + \frac{g_m}{C_{GS}C_L s^2}$$

- **A source follower driving a load capacitance exhibits a negative input resistance, possibly causing instability.**

Output Impedance

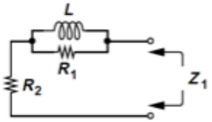


(a) (b) (c)

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

- **At low frequency:** $Z_{out} \approx 1/g_m$
- **At very high frequencies,** $Z_{out} \approx R_S$
- **Because CGS shorts the gate and the source.**
- **Which one of these variations is more realistic?**

Output Impedance



'. Equivalent output impedance of a source follower

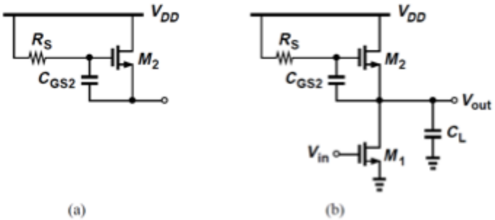
- **Since the output impedance increases with frequency, we postulate that it contains an inductive component.**

$$Z_{out} - \frac{1}{g_m} = \frac{C_{GS}s\left(R_S - \frac{1}{g_m}\right)}{g_m + C_{GS}s}$$

$$\frac{1}{Z_{out} - \frac{1}{g_m}} = \frac{1}{R_S - \frac{1}{g_m}} + \frac{1}{\frac{C_{GS}s}{g_m}\left(R_S - \frac{1}{g_m}\right)}$$

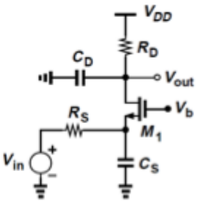
$$L = \frac{C_{GS}}{g_m}\left(R_S - \frac{1}{g_m}\right)$$

Example



- Can we construct a (two-terminal) inductor from a source follower?
- Yes, but non-ideal.
- It also incurs a parallel resistance and a series resistance.
- The inductance can partially cancel the load capacitance, CL, at high frequencies, thus extending the bandwidth.

Common-Gate Stage



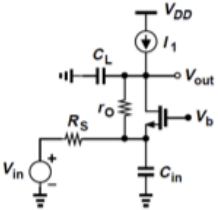
Common-gate stage at high frequencies

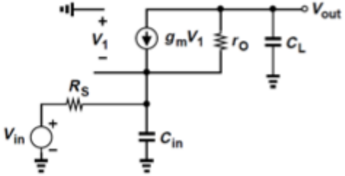
- **A transfer function**

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

- **No Miller multiplication of capacitances.**
- **R_D is typically maximized, so the dc level of the input signal must be quite low.**
- **As an amplifier in cases where a low input impedance is required**
- **In cascode stages.**

Example



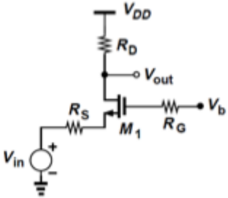


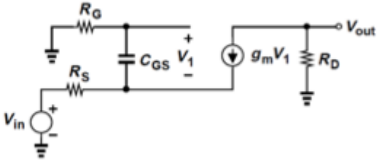
- Why Z_{in} becomes independent of C_L as this capacitance increases?

$$Z_{in} = \frac{1}{g_m + g_{mb}} + \frac{1}{C_L s} \cdot \frac{1}{(g_m + g_{mb})r_O}$$

- As C_L or s increases, Zin approaches 1/(gm + gmb)
- C_L lowers the voltage gain of the circuit, thereby suppressing the effect of the negative resistance introduced by Miller effect through rO.

CG Stage





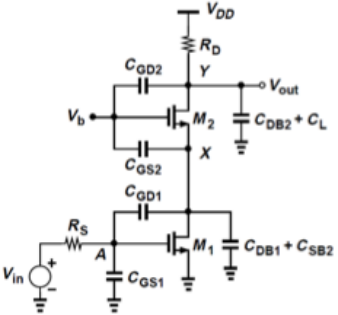
- The bias network providing the gate voltage exhibits a finite impedance.
- Consider only C_{GS} here.

$$\frac{V_{out}}{V_{in}} = \frac{g_m R_D}{(R_G + R_S)C_{GS}s + 1 + g_m R_S},$$

$$\omega_p = \frac{1 + g_m R_S}{(R_G + R_S)C_{GS}}.$$

- Lowering the pole magnitude.
- Output impedance of the circuit drops at high frequencies.

Cascode Stage



High-frequency model of a cascode stage.

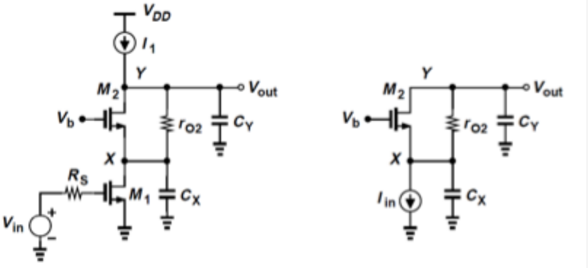
$$\omega_{p,A} = \frac{1}{R_S\left[C_{GS1} + \left(1 + \frac{g_{m1}}{g_{m2} + g_{mb2}}\right) C_{GD1}\right]}$$

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

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

- **Miller effect is less significant in cascode amplifiers than in common-source stages.**
- **But $\omega_{p,X}$ is typically quite higher than the other two.**
- **What if RD is replaced by a current source?**
- **Pole at node X may be quite lower, but transfer function will not affect much by this. See example.**

Example



(a) (b)

- **Compute the transfer function.**

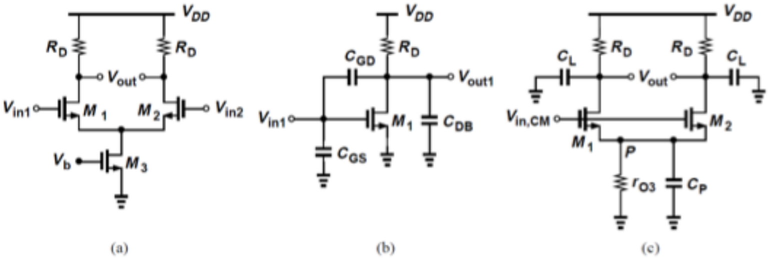
$$\frac{V_{out}}{I_{in}} = -\frac{g_{m2}r_{O2}+1}{C_X s} \cdot \frac{1}{1+(1+g_{m2}r_{O2})\frac{C_Y}{C_X}+C_Y r_{O2} s}$$

- **For** $g_{m2}r_{O2} \gg 1$

$$\frac{V_{out}}{V_{in}} = -\frac{g_{m1}g_{m2}}{C_Y C_X s}\frac{1}{g_{m2}/C_X + s}$$

- **The magnitude of the pole at node X is still given by gm2/Cx. Why?**

Differential Pair



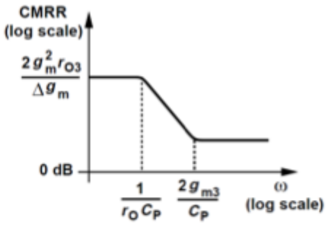
(a) Differential pair, (b) half-circuit equivalent, (c) equivalent circuit for common-mode inputs.

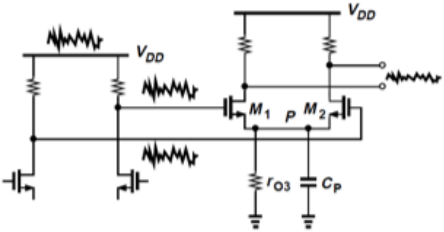
- **For differential signals, the response is identical to that of a common-source stage.**
- **the common-mode rejection of the circuit degrades considerably at high frequencies.**

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

- **Channel-length modulation, body effect, and other capacitances are neglected.**

Differential Pair

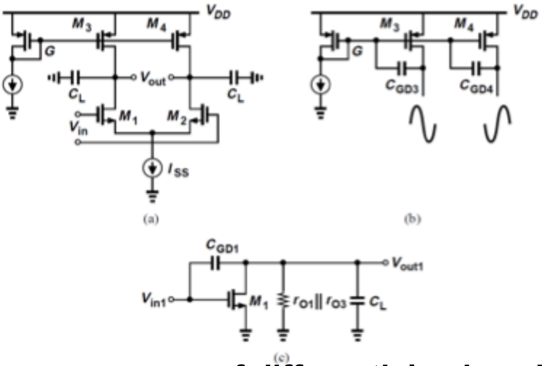




$$\text{CMRR} \approx \frac{g_m}{\Delta g_m} \frac{r_{O3} C_P s + 1 + 2 g_m r_{O3}}{r_{O3} C_P s + 1}$$

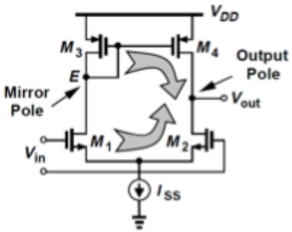
- This transfer function contains a zero and a pole.
- The magnitude of the zero is much greater than the pole.
- Common-mode disturbance at node P translates to a differential noise component at the output, if the supply voltage contains high-frequency noise and the circuit exhibits mismatches.
- Trade-off between voltage headroom and CMRR.

Diff Pair



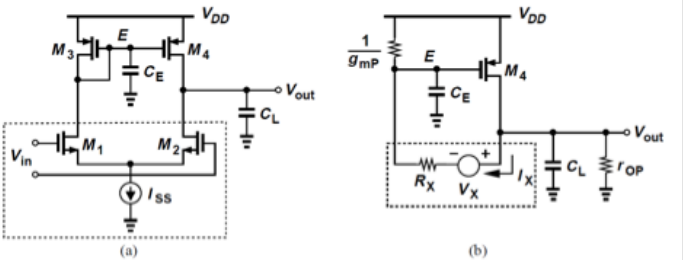
- **Frequency response of differential pairs with high-impedance loads.**
- **Fig (b) C_{GD3} and C_{GD4} conduct equal and opposite currents to node G, making this node an ac ground.**
- **The differential half circuit is depicted in Fig. (c).**
- **More on chapter 10**

Differential Pair with Active Load



- How many poles does this circuit have?
- The severe trade-off between gm and CGS of PMOS devices results in a pole that impacts the performance of the circuit.
- The pole associated with node E is called a "mirror pole."

Active Load



- **Replacing Vin, M1, and M2 by a Thevenin equivalent.**

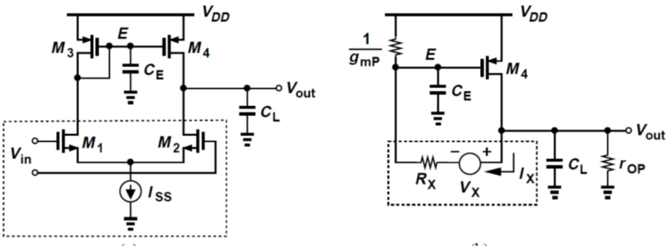
$$V_X = g_{mN} r_{ON} V_{in}$$

$$R_X = 2 r_{ON}$$

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

$$\omega_{p1} \approx \frac{1}{(r_{ON} \| r_{OP}) C_L} \qquad \omega_{p2} \approx \frac{g_{mP}}{C_E}$$

Active Load

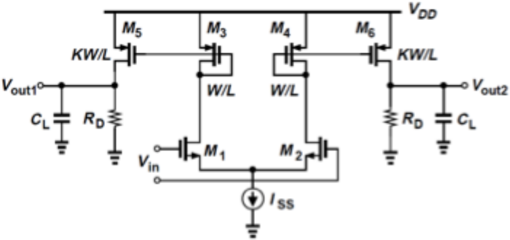


- A zero with a magnitude of $2g_{mP}/C_E$ in the left half plane.
- The appearance of such a zero can be understood by noting that the circuit consists of a "slow path" (M1,M3 and M4) in parallel with a "fast path" (M1 and M2) by $A_0/[(1+s/\omega_{p1})(1+s/\omega_{p2})]$

$A_0/(1+s/\omega_{p1})$

$$\frac{V_{out}}{V_{in}} = \frac{A_0}{1+s/\omega_{p1}}\left(\frac{1}{1+s/\omega_{p2}}+1\right)$$
$$= \frac{A_0(2+s/\omega_{p2})}{(1+s/\omega_{p1})(1+s/\omega_{p2})}.$$

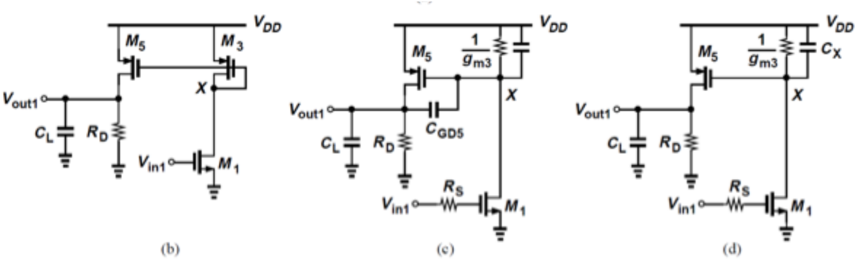
Example



(a)

- Not all fully differential circuits are free from mirror poles.
- Estimate the low-frequency gain and the transfer function of this circuit.

Example



- **M5 multiplies the drain current of M3 by K.**

$$A_v = g_{m1} K R_D$$

- **Assume $R_D C_L$ is relatively small so that the Miller multiplication of C_{GD5} can be approximated as**

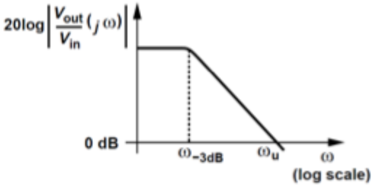
$$C_{GD5}(1 + g_{m5}R_D)$$

- **The overall tran**

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

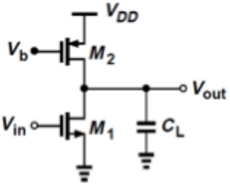
n equal to Vx/Vin1 multiplied by Vout1/Vx.

Gain-Bandwidth Trade-Offs



- **We wish to maximize both the gain and the bandwidth of amplifiers.**
- **we are interested in both the 3-dB bandwidth,** ω_{-3dB} **and the "unity-gain" bandwidth,** ω_u.

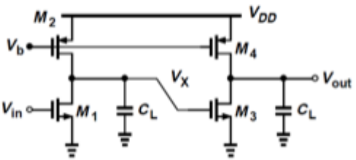
One pole circuit



$$
\begin{aligned}
GBW &= A_0\omega_p \\
&= g_{m1}(r_{O1}||r_{O2})\frac{1}{2\pi(r_{O1}||r_{O2})C_L} \\
&= \frac{g_{m1}}{2\pi C_L}.
\end{aligned}
$$

$$
\begin{aligned}
\omega_u &= \sqrt{A_0^2 - 1}\omega_p \\
&\approx A_0\omega_p
\end{aligned}
$$

Multi-Pole Circuits



- It is possible to increase the GBW product by cascading two or more gain stages.
- Assume the two stages are identical and neglect other capacitances.

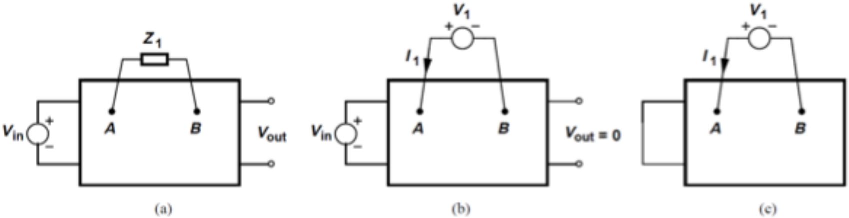
$$\frac{V_{out}}{V_{in}} = \frac{A_0^2}{(1+\frac{s}{\omega_p})^2}$$

$$\omega_{-3-dB} = \sqrt{\sqrt{2}-1}\omega_p \approx 0.64\omega_p.$$

$$GBW = \sqrt{\sqrt{2}-1}A_0^2\omega_p$$

- While raising the GBW product, cascading reduces the bandwidth.

Appendix A: Extra Element Theorem



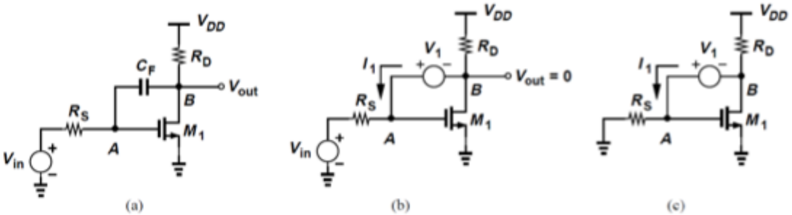
(a) Circuit with extra parallel element, Z_1, (b) $Z_{out,0}$ calculation, and (c) $Z_{in,0}$ calculation.

- **Suppose the transfer function of a circuit is known and denoted by H(s). Add an extra impedance Z1 between two nodes of the circuit.**
- **New transfer function:**

$$G(s) = H(s)\frac{1 + \frac{Z_{out,0}}{Z_1}}{1 + \frac{Z_{in,0}}{Z_1}}$$

- **Particularly useful for frequency response analysis.**

Example 1



- **Find the transfer function.**

$$H(s) = -g_m(R_D||r_O)$$

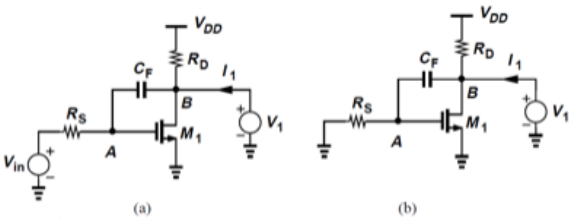
$$Z_{out,0} = -1/g_m$$

- **The negative sign of Zout,0 does not imply a negative impedance between A and B, since** $V_{in} \neq 0$

$$Z_{in,0} = (1+g_m R_S)R_D + R_S = (1+g_m R_D)R_S + R_D$$

$$G(s) = -g_m(R_D||r_O)\frac{1-\frac{1}{g_m}C_F s}{1+[(1+g_m R_D)R_S + R_D]C_F s}$$

Example 2



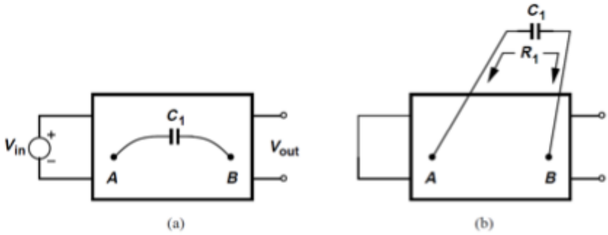
- **Include CB, from node B to ground.**

$$Z_{out,0} = 0$$

$$Z_{in,0} = \frac{R_D(R_S C_F s + 1)}{R_S(1 + g_m R_D) + R_D]C_F s + 1}$$

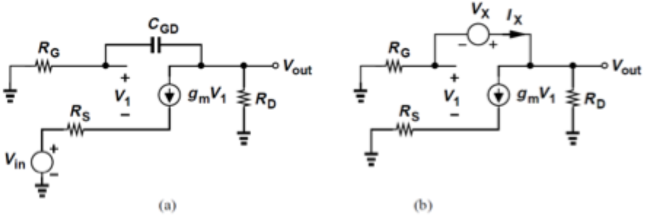
$$G(s) = -g_m(R_D||r_O)\frac{1 - \frac{C_F}{g_m}s}{1 + [(1 + g_m R_D)R_S + R_D]C_F s}\frac{1}{1 + \frac{R_D(R_S C_F s + 1)C_B s}{[R_S(1 + g_m R_D) + R_D]C_F s + 1}}$$

Appendix B: Zero-Value Time Constant



- Suppose a circuit contains one capacitor and no other storage elements.
- Wish to determine the pole of the system.
- Set the input to zero, compute the resistance, R1, seen by C1, and express the pole as 1/(R1C1).

Example



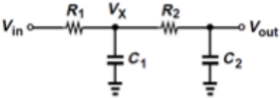
- **If only CGD is considered, determine the pole frequency.**

$$g_m V_1 R_S + V_1 = -I_X R_G$$

$$\frac{V_X}{I_X} = R_D + \left(\frac{g_m R_D}{1 + g_m R_S} + 1\right) R_G = R_{eq}$$

- **The pole is given by** $1/(R_{eq}C_{GD})$

Example



- Writing a KVL around Vin, R1, R2, and Vout gives

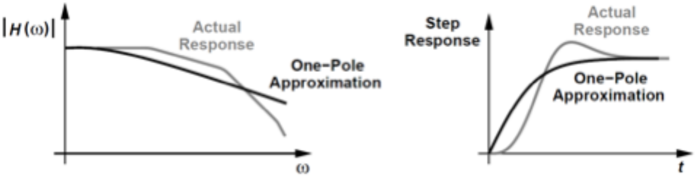
$$\frac{V_{out}}{V_{in}}(s) = \frac{1}{R_1R_2C_1C_2s^2 + [R_1C_1 + (R_1 + R_2)C_2]s + 1}$$

- The dominant pole is indeed equal to the inverse of the sum of the zero-value time constants. (need to prove)

$$B_s = \omega_{p1}^{-1} + \omega_{p2}^{-1} + \cdots + \omega_{p1}^{-n} = R_1C_1 + R_2C_2 + \cdots + R_nC_n$$

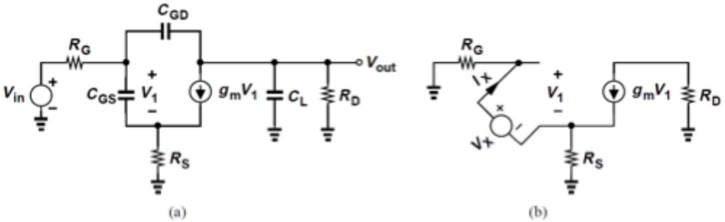
- The ZVTC method proves useful if we wish to estimate the 3-dB bandwidth of a circuit.

Example



- **Approximation of the frequency and time responses by one-pole counterparts.**

Example 1



- **Estimate the 3-dB bandwidth.**
- **Begin with the time constant associated with C_{GS} and set C_{GD} and C_L to zero.**

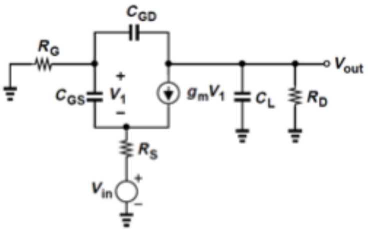
$$R_{CGS} = \frac{R_G + R_S}{1 + g_m R_S}$$

$$R_{CGD} = R_D + \left(\frac{g_m R_D}{1 + g_m R_S} + 1\right) R_G$$

- **The resistance seen by C_L is simply equal to R_D.**

$$\omega_{-3dB}^{-1} = \frac{R_G + R_S}{1 + g_m R_S} C_{GS} + \left[R_D + \left(\frac{g_m R_D}{1 + g_m R_S} + 1\right) R_G\right] C_{GD} + R_D C_L$$

Example2



- **Find 3-dB band width of a common-gate stage containing a gate resistance of RG and a source resistance of RS.**
- **The resulting equivalent circuits are identical for CS and CG stages, yielding the same time constants and hence the same bandwidth.**
- **Does this result contradict our earlier assertion that the CG stage is free from the Miller effect?**
- **No. Why?**