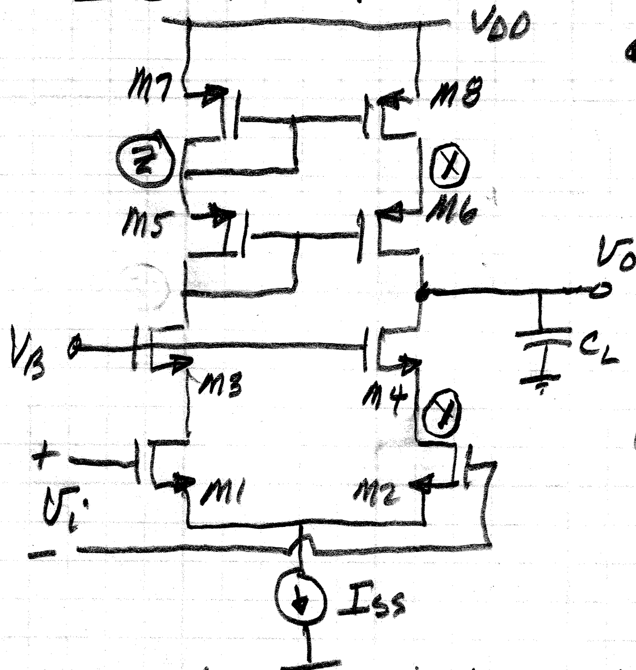
• Telescopic Opamp (cont.):



• Problems with this opamp:

(i) limited output swing

$V_O(max) = V_X - |V_{OV6}|$

$= \boxed{V_{DD} - |V_{T7}| - |V_{OV7}| - |V_{OV6}|}$

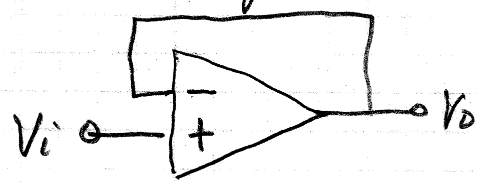
with $|V_{T5}| = |V_{T6}|$ & $|V_{OV5}| = |V_{OV6}|$

(ii) $V_O(min) = V_Y + V_{OV4}$

$= \boxed{V_B - V_{T4}}$

with V_B fixed from generator.

• Very limited input Common-mode Range:



Unity-gain follower

$$V_{ic} = \frac{v^+ + v^-}{2} = v_i$$

(i) $V_{ic}(min) = \boxed{V_B - V_{T4}}$

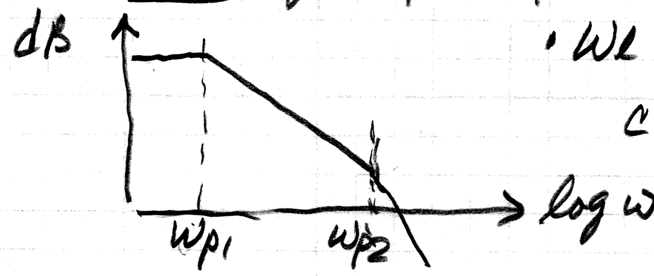
(ii) $V_{ic}(max) = \boxed{V_B - V_{ov4}}$

with $V_{T2} = V_{T4}$.

$CMR_i \equiv$ input CM range $= V_{ic}(max) - V_{ic}(min) = \boxed{V_{T4} - V_{ov4}}$

Very low CMR_i ➡ Telescopic opamp not suitable unity-gain buffers or for high dynamic range.

• Poor frequency response due to non-dominant pole:



• We want $\frac{\omega_{p2}}{\omega_{p1}}$ large for closed-loop stability

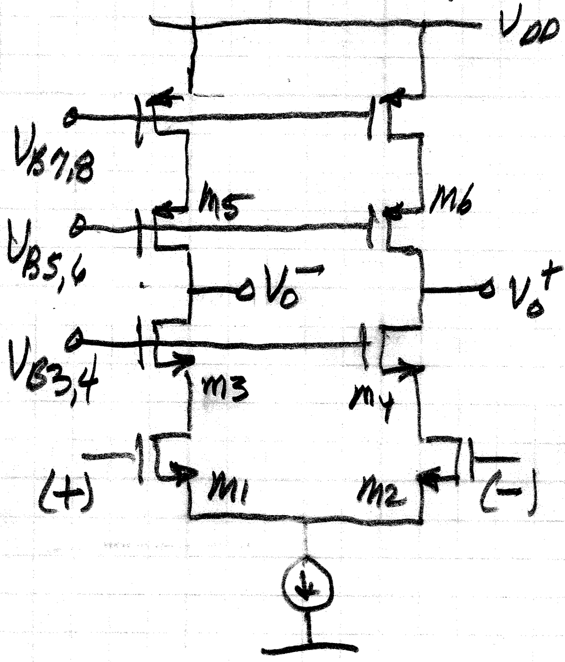
Dominant pole $= \omega_{p1} \simeq \dfrac{1}{R_o C_o} \approx \dfrac{1}{g_m r_o^2 C_L}$ (Equal g_m, r_o, etc.)

First non-dominant pole $= \omega_{p2}$ (Probably at $\boxed{Z}$ because of high capacitance there)

$$\omega_{p2} \simeq \frac{g_{m7}}{C_{ox}\left[(W/L)_7 + (W/L)_8\right]}$$

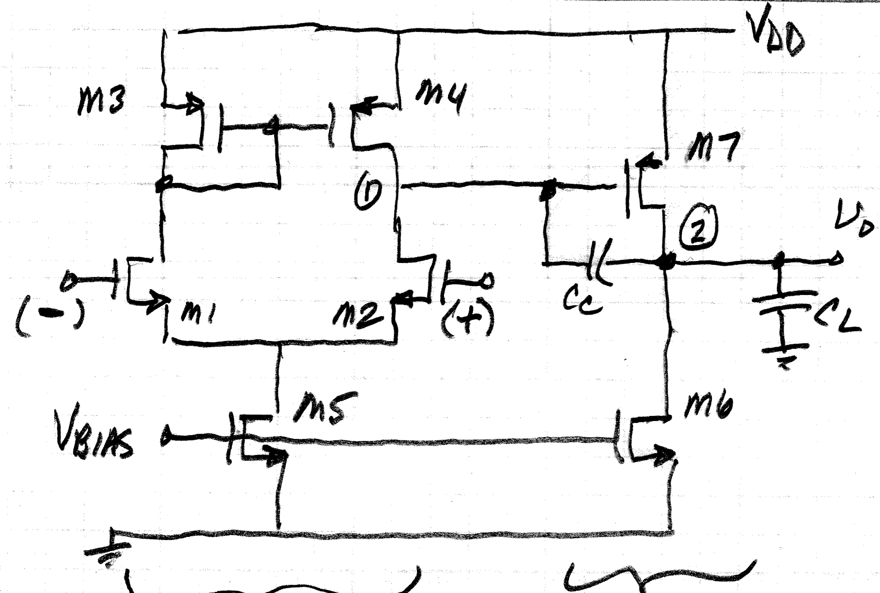
$\therefore$ Low-frequency non-dominant pole, ω_{p2}, due to current mirror node.

<u>One Solution:</u> <u>Fully-differential</u> (at both input and output):



- 6 dB increase in Dynamic Range
- Improved CMRR and PSRR
- ω_{p2} at higher frequency (i.e., no mirror node). Thus, better closed-loop stability.
- But, CM Feedback circuit needed to set DC Voltages at the output nodes.

- <u>Another solution</u>: Two-stage CMOS Opamp



Input Differential Stage with NMOS tail current source

PMOS common-source second stage with NMOS current source load.

• Two stages connected in cascade.

• $C_c \equiv$ Miller-multiplied frequency compensation capacitor

• Determine (+) and (−) input terminals

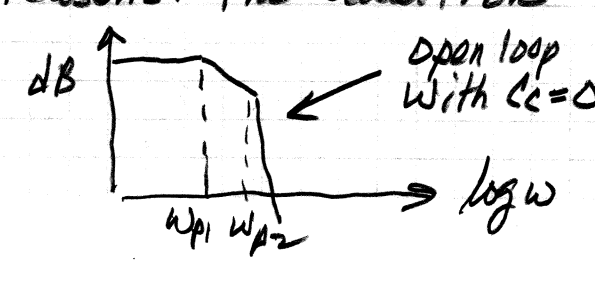
• Determine DC bias conditions (Assume $V_o(DC) = \frac{V_{DD}}{2}$)
(Feedback will set DC voltage at V_o)

Gain: Stage 1: $a_{v1} = \frac{v_{①}}{v_i} = \frac{-g_{m1}}{g_{o2}+g_{o4}}$

Stage 2: $a_{v2} = \frac{v_o}{v_{①}} = \frac{-g_{m7}}{g_{o6}+g_{o7}}$

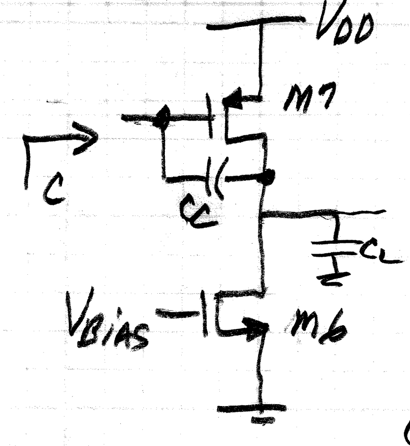
Overall gain = $\boxed{a_v = a_{v1} \cdot a_{v2} = \frac{g_{m1}}{g_{o2}+g_{o4}} \cdot \frac{g_{m7}}{g_{o6}+g_{o7}}}$

Note: $R_{o①} \simeq R_o$ so $\omega_{p1} \simeq \omega_{p2}$ without C_c. We want an approximate one-pole response out to ω_T for stability reasons. The addition of C_c helps achieve this.



Now, add C_c compensation capacitor:

Miller Effect:



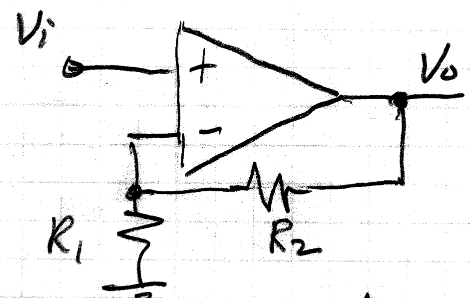
$$C = C_c(1 - a_{v2})$$
$$= C_c\left(1 + \frac{g_{m7}}{g_{o6}+g_{o7}}\right)$$
$$C \simeq \frac{g_{m7}}{g_{o6}+g_{o7}} C_c$$
$$(\text{e.g. } C = 100\, C_c)$$

- Huge savings in area of C_c using Miller Effect.

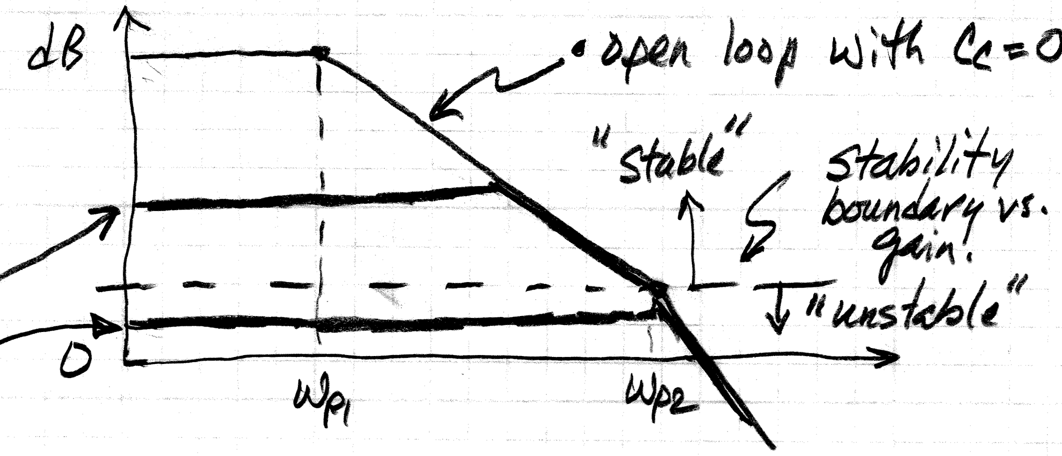
$\therefore$ Dominant pole $= \omega_{p1}$

$$\boxed{\omega_{p1} = \frac{1}{g_{o2}+g_{o4}} \cdot \frac{g_{m7} C_c}{g_{o6}+g_{o7}}}$$

- Typical closed-loop configuration: (negative FB)



$$\boxed{A_v = \frac{V_o}{V_i} = 1 + \frac{R_2}{R_1} = \text{closed-loop gain}}$$

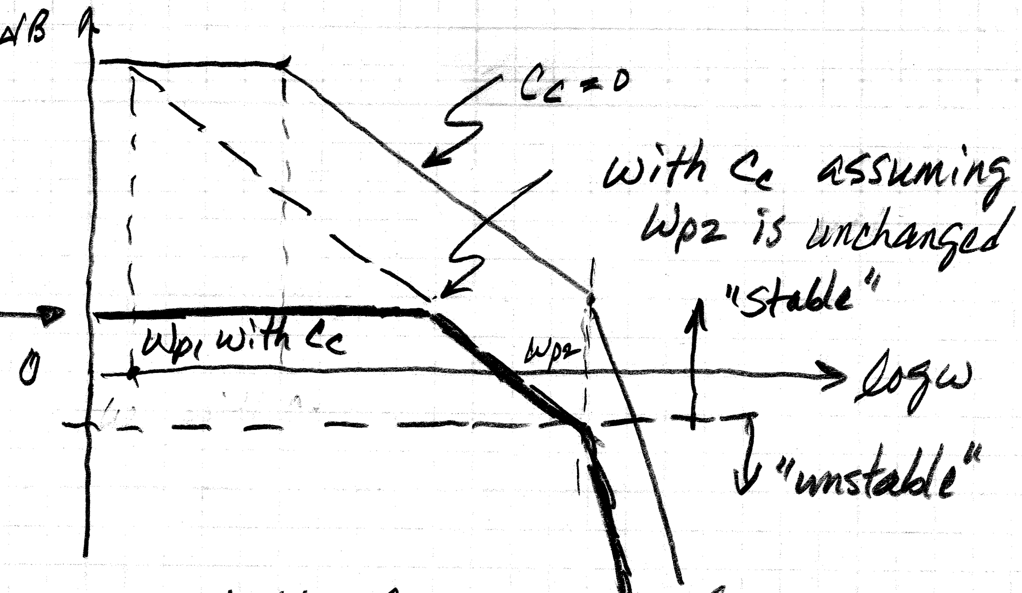


Case 1: $1 + R_2/R_1$ large (Easily stable)

Case 2: $1 + R_2/R_1$ small ("unstable")

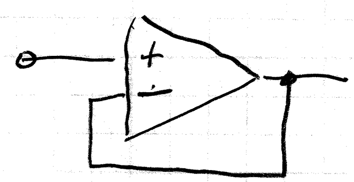
Now, add C_c:

Case 2: (previous page)



- Opamp is now stable for values of $A_v \geq 1$

From this, we see that unity-gain is worst-case for closed-loop stability.

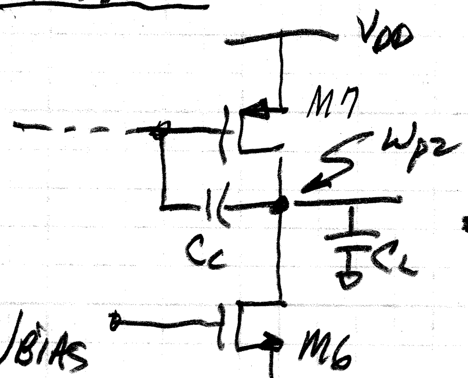


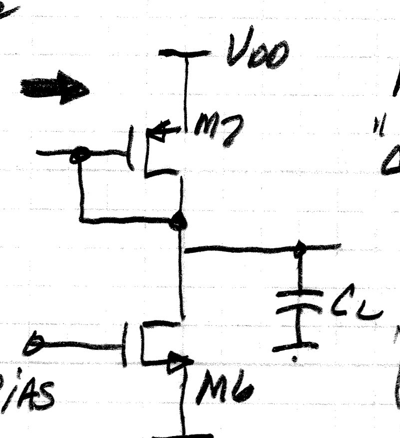
- Worst-case FB circuit for stability

In reality, ω_{p2} moves to a higher frequency!

Intuition:

- At high frequencies, assume C_c becomes a short circuit:





M_7 is now "diode connected"

$$\omega_{p2} \simeq \frac{g_{m7} + g_{o6}}{C_L}$$

$$\boxed{\omega_{p2} \simeq \frac{g_{m7}}{C_L}}$$

∴ Very good closed-loop stability with pole-splitting Miller compensation!!