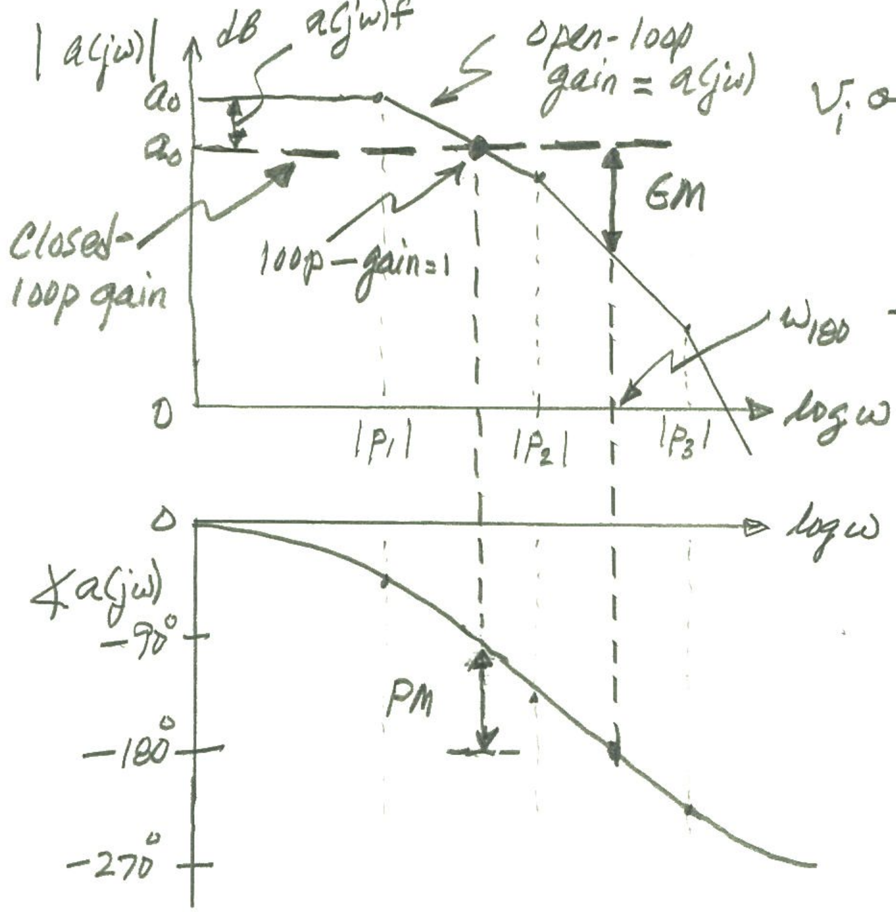
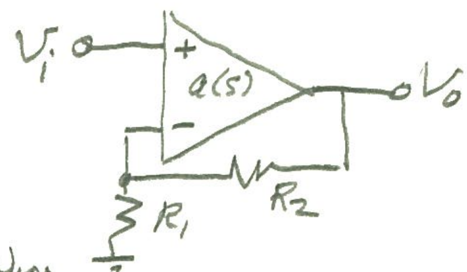


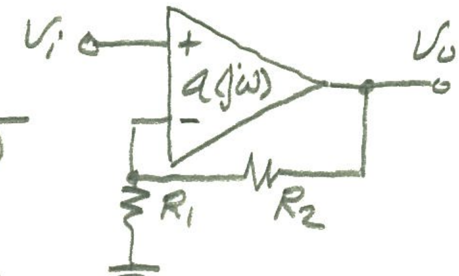


note: $\measuredangle a(j\omega)f = \measuredangle a(j\omega)$ because $f = \frac{R_1}{R_1+R_2}$ = constant

- PM ≡ phase margin = $180^\circ - \measuredangle a(j\omega)f$ where $|a(j\omega)f| = 1$ (PM ≃ 75° above)

GM ≡ gain margin = $|a(j\omega)f|$ at frequency, ω_{180}, where $\measuredangle a(j\omega)f = -180^\circ$

- Frequency Compensation:



$$A(s) = \frac{a(s)}{1+\underbrace{a(s)f}_{\text{loop transmission}}} = \frac{a(s)}{1+T(s)}$$

at DC, $a(s)f = T_o = a_o f \equiv$ loop gain

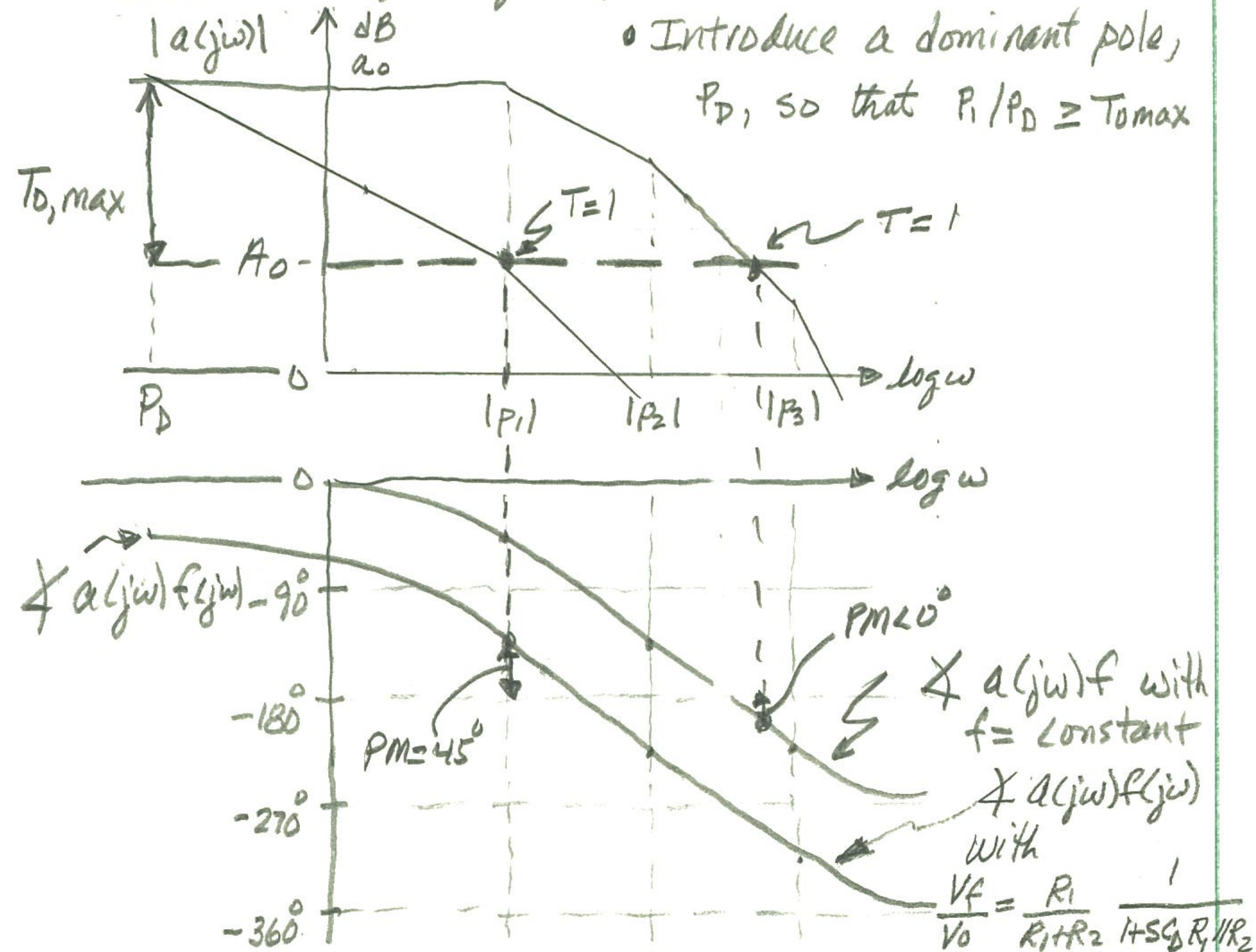
Instability occurs when $A(s) \rightarrow \infty$

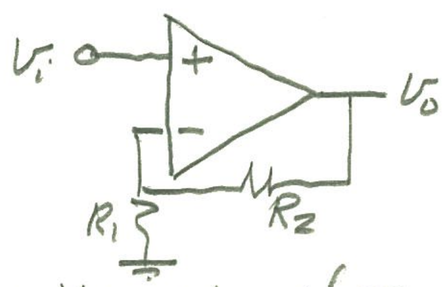
$$\Rightarrow A(s) = \frac{a(s)}{1+a(s)f} \rightarrow \frac{a(s)}{1-1} = \frac{a(s)}{0} \rightarrow \infty$$

Compensate for $T_{o,max}$ ($T_{o,max} = a_o \equiv$ worst case)

(A) Narrowbanding frequency compensation:

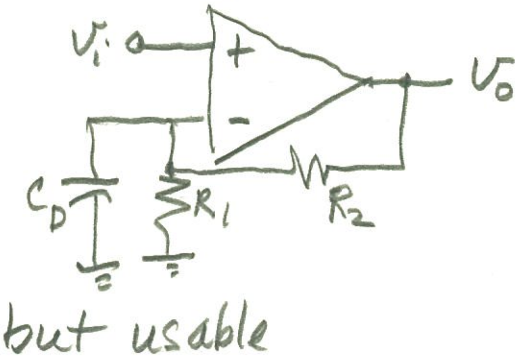
• Introduce a dominant pole, P_D, so that $P_1/P_D \geq T_{o\,max}$





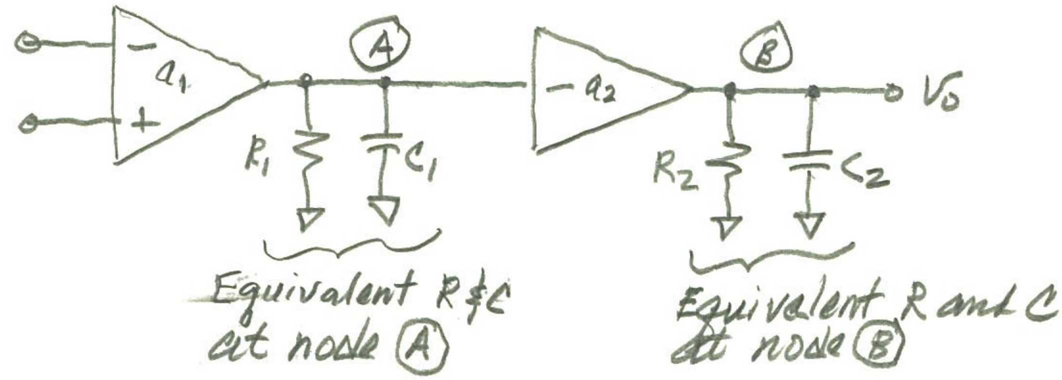
• for the value of T_o shown, the amplifier is unstable.

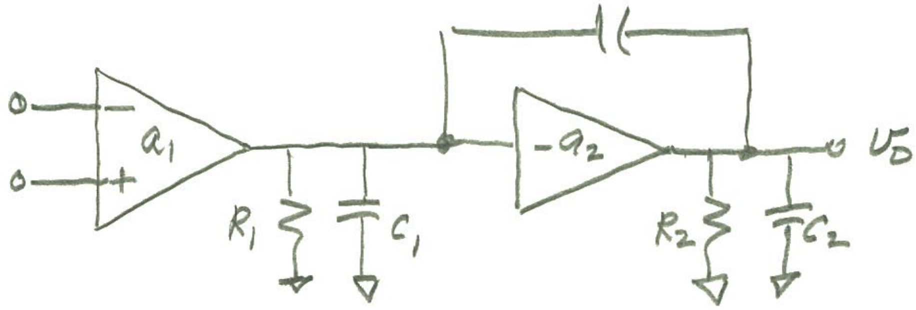
Add a dominant pole so that $P_1/P_D = T_{o\,max}$. This yields $PM = 45^\circ$. Put a pole in $f(j\omega)$ as shown:

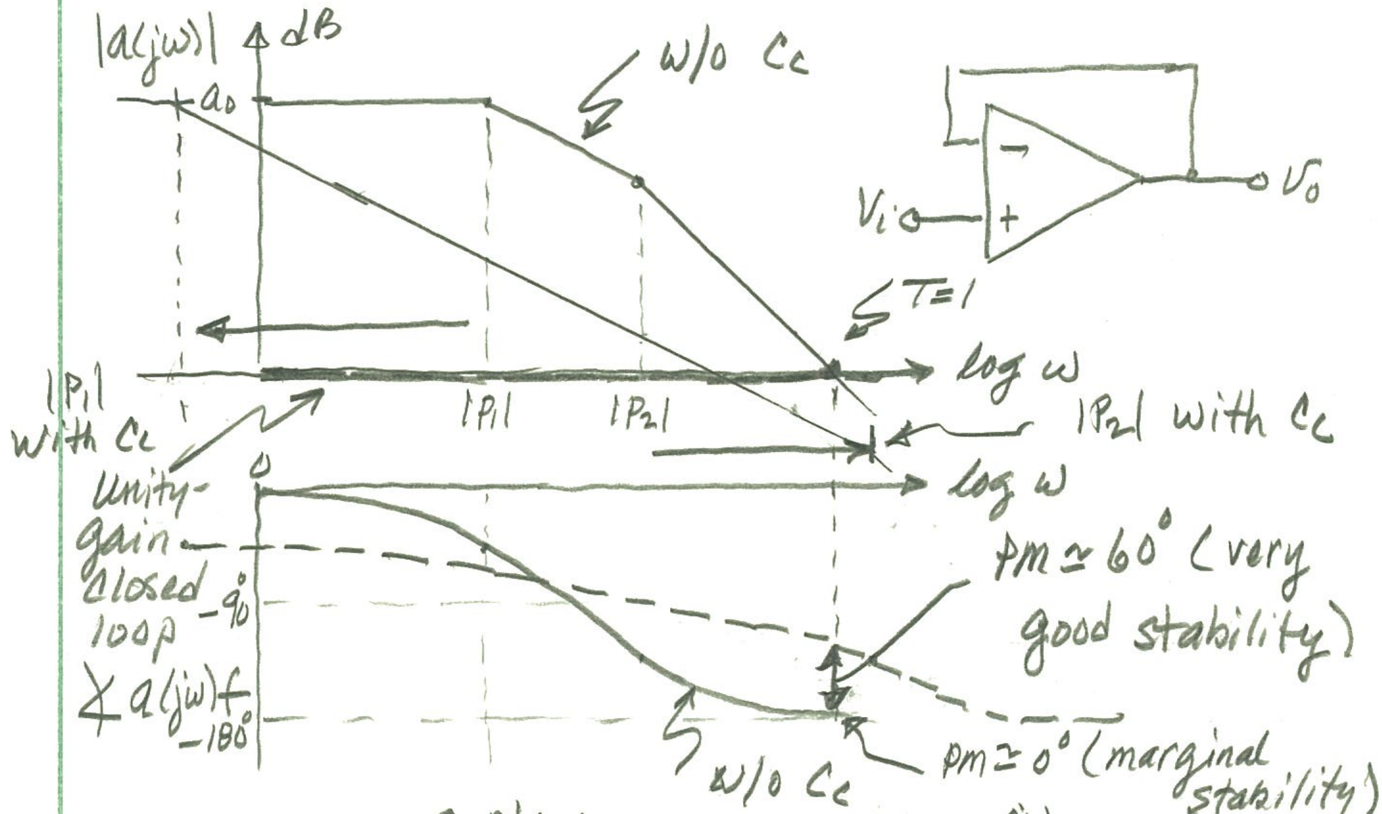


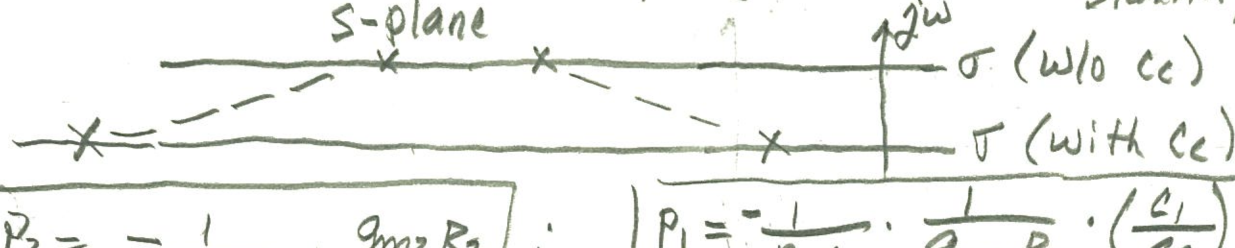
• Closed-loop amp is stable but usable bandwidth is greatly decreased.

(B) Pole-splitting Compensation (Miller Compensation):





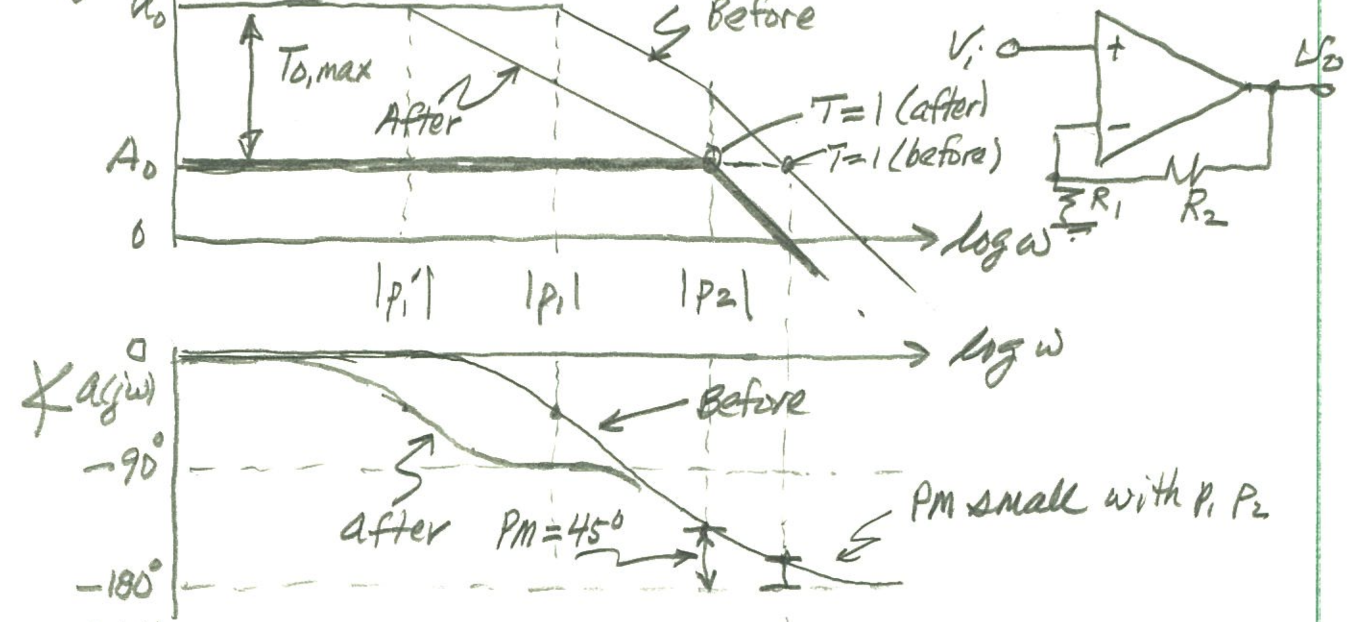




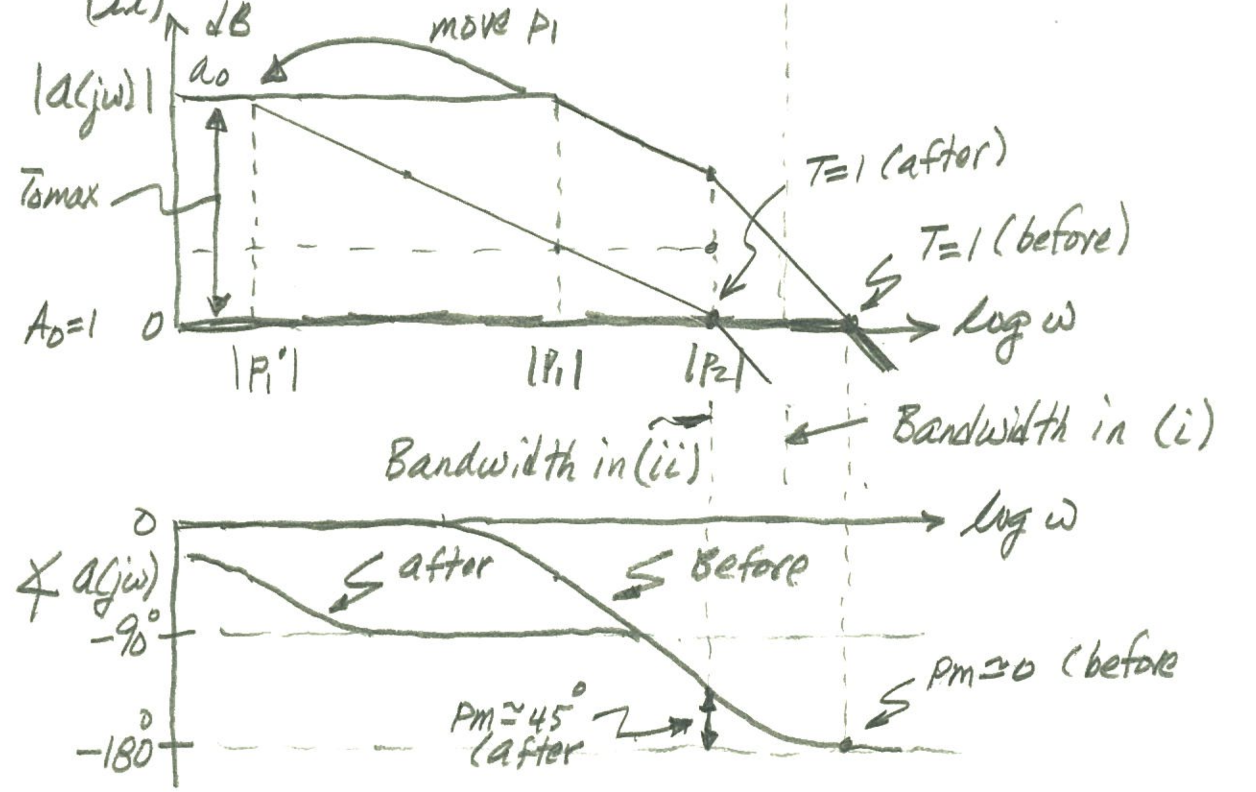
$$\boxed{P_2 = -\frac{1}{R_2 C_2} \cdot g_{m2} R_2} \; ; \quad \boxed{P_1 = -\frac{1}{R_1 C_1} \cdot \frac{1}{g_{m2} R_2} \cdot \left(\frac{C_1}{C_c}\right)}$$

- Choosing the value of C_c Based on closed-loop gain:

(i) closed-loop gain > 1: (Assume p_2 fixed and move p_1 for $PM = 45°$)



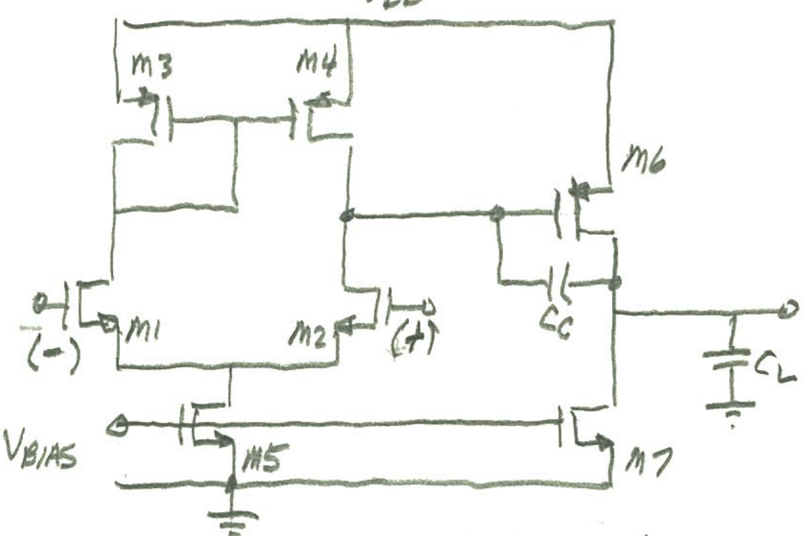
(ii)



If we compensate for unity-gain stability (i.e., the worst case), and then use with $A_0 > 1$, bandwidth is lost. Design guideline: <u>Do not over compensate</u>!

- <u>Choosing C_c</u>: (Assumptions for now) Approximate Method.
  - Assume no RHP zeros
  - $|p_3| >> |p_2|$
  - Impedance of $C_c = \frac{1}{sC_c}$ << surrounding impedances at high frequencies.

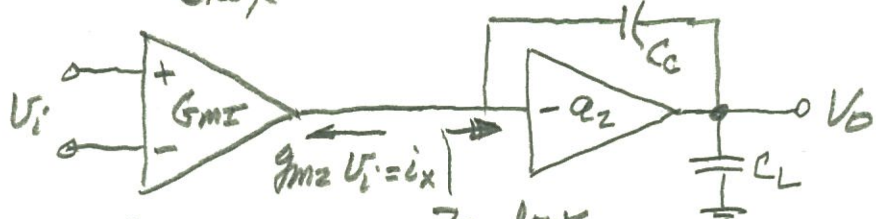
<u>case (i)</u>: Two-stage opamp with Miller Compensation



Differential transconductance stage

single-ended common-source amplifier

- consider the model of the two-stage miller compensated opamp below. We will use an approx. analysis at high frequencies where $T \to 1$ and PM is determined.

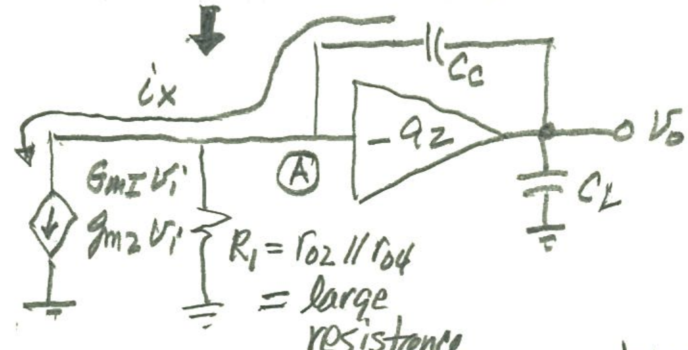


$G_{mI} = g_{m2}$

Z_c low at high freqs ;

<u>Assume:</u>

$$Z_c = \frac{1}{sC_c(1 + a_2(s))} << R_1$$

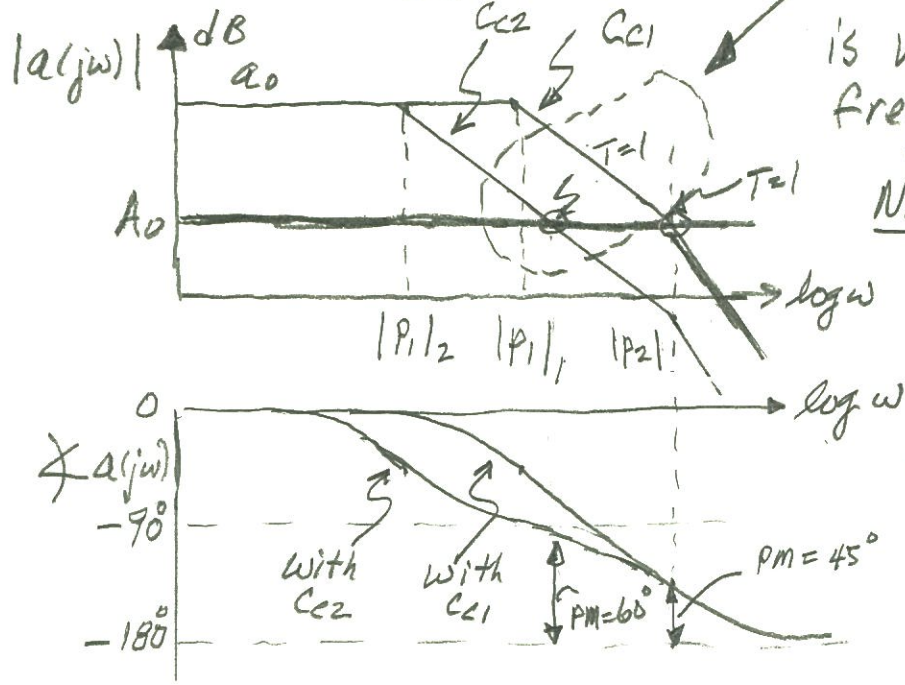


∴ Most of $g_{m2} v_i$ flows through C_c

treat Ⓐ as ground.

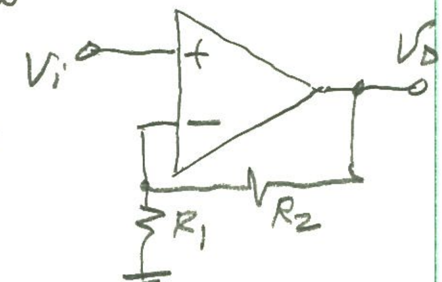
Hence, $i_x = G_{mI} V_i$, $V_o = \dfrac{i_x}{sC_c}$ $\Big\}$ $V_o = \dfrac{G_m V_i}{sC_c}$ $\Rightarrow$ $\boxed{\dfrac{V_o}{V_i}(s) = \dfrac{G_{mI}}{sC_c}}$

is valid for these frequencies.



Note: $C_{c2} > C_{c1}$

P_2 stays same.



<u>case (i)</u>: C_{c1} choosen with A_0 for $PM = 45°$.

Closed-loop gain = A_0

$$\left| \frac{V_o}{V_i}(j\omega) \right| = \frac{G_{mI}}{\omega C_c}$$

ω_{ULT} = "Unity loop gain" frequency

= ω where $|a(j\omega)f| = 1$

For $PM = 45°$, $\omega_{ULT} = \omega_2$ (second pole freq)

$$\therefore \left| \frac{V_o}{V_i}(j\omega_2) \right| = A_0 = \frac{G_{mI}}{\omega_2 C_{c1}} \Rightarrow \boxed{C_{c1} = \frac{G_{mI}}{\omega_2 A_0}}$$

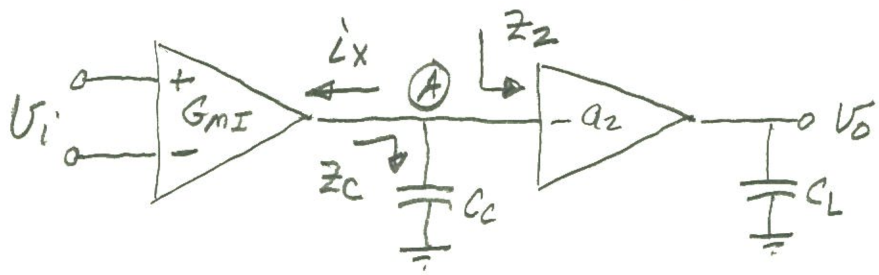
case (ii) C_{c2} choosen with A_0 for $PM = 60°$:

<u>Note</u>: $\omega_{ULT} < \omega_2$ in this case. It can be shown that $\omega_{ULT} = \dfrac{\omega_2}{1.73}$ for $PM = 60°$

$$\therefore \left| \frac{V_o}{V_i}\left(j\frac{\omega_2}{1.73}\right) \right| = A_0 = \frac{G_{mI}}{\frac{\omega_2}{1.73} C_{c2}} \Rightarrow \boxed{C_{c2} = \frac{1.73\, G_{mI}}{\omega_2 A_0}}$$

Remarks:
- Smaller A_o requires larger C_c
- Complex dependence on G_{mI} (more later)

Case (ii): Two-stage opamp with shunt compensation capacitance. (Not practical - just for comparison)



C_c is huge capacitor. Thus, $Z_c << Z_2$ at high frequencies.

Hence,
$$\left.\begin{aligned} v_{A} &= -\frac{G_{mI} v_i}{sC_c} \\ v_o &= -a_2 v_{A} \end{aligned}\right\}$$

$$v_o(s) = \frac{a_2 G_{mI} v_i(s)}{sC_c}$$

$$\frac{v_o}{v_i}(s) = \frac{a_2 G_{mI}}{sC_c}$$

For PM = 45°: Closed-loop gain must intersect the previous curves at ω_{ULT} for the desired PM

Hence, $$\left|\frac{v_o}{v_i}(j\omega_{ULT})\right| = A_o = \frac{a_2 G_{mI}}{\omega_2 C_{c1}}$$

$$\therefore \boxed{C_{c1} = \frac{a_2 G_{mI}}{\omega_2 A_o}}$$

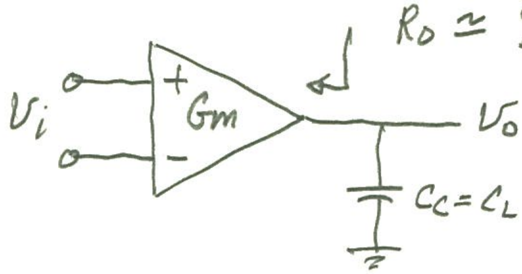
C_{c1} bigger by a_2 because we did not take advantage of Miller multiplication effect

For PM = 60°,

$$\boxed{C_{c2} = \frac{1.73\, a_2 G_{mI}}{\omega_2 A_o}}$$

Case (iii): Single-stage cascode opamp with shunt load capacitance (e.g., telescopic opamp)

$R_o \simeq \frac{g_m r_o^2}{2}$ (very high $\therefore$ transconductance amplifier)

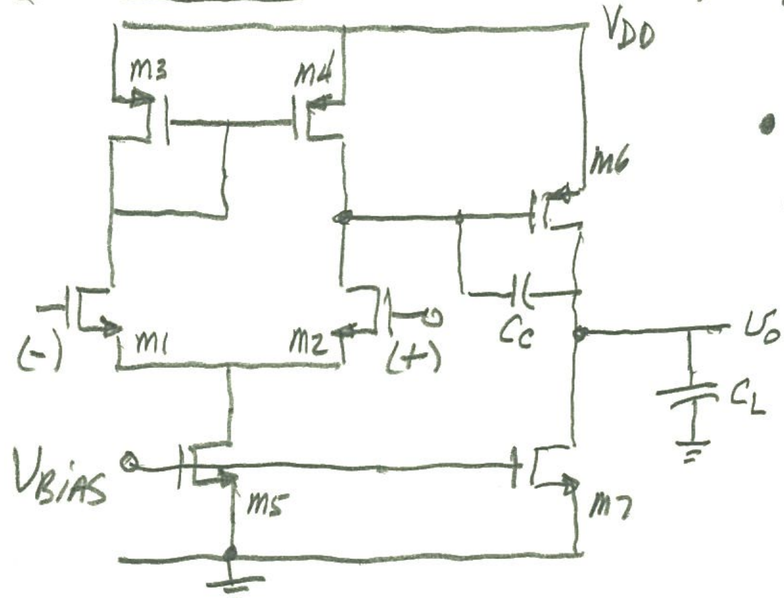


$\therefore$ $\underline{PM = 45^\circ}$: $\boxed{C_{C1} = C_{L1} = \frac{G_{m1}}{\omega_2 A_o}}$

$\underline{PM = 60^\circ}$: $\boxed{C_{C2} = C_{L2} = \frac{1.73\, G_{m1}}{\omega_2 A_o}}$

Note: One big point here: Might need to add capacitance in parallel with C_L for desired PM.

Actual Circuit: CMOS two-stage opamp compensation



- Relate the equations derived on page 147 to the circuit parameters of the ubiquitous two-stage opamp

First stage:

$G_{mI} = g_{m2}$

$R_I = r_{o2} \parallel r_{o4}$

Second-Stage:

$G_{mII} = g_{m6}$

$R_{II} = r_{o6} \parallel r_{o7}$