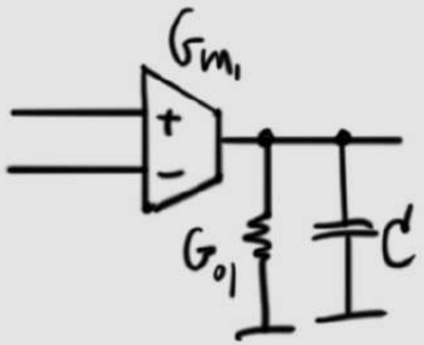
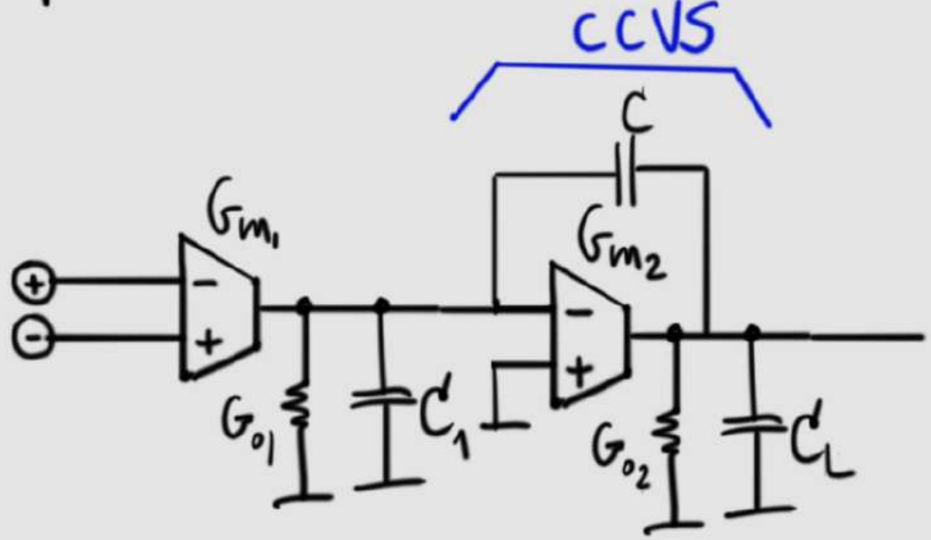
Single stage op amp is a transconductor driving a capacitive load:



$$A_0 = \frac{G_{m1}}{G_{o1}} = g_{m_1}(r_{ds_1} \| r_{ds_3})$$

$$\omega_u = \frac{G_{m_1}}{C}$$

2-stage op amp is the combination of a transconductor + a "CCVS"



$$A_0 = \frac{G_{m1}}{G_{o1}} \cdot \frac{G_{m2}}{G_{o2}}$$

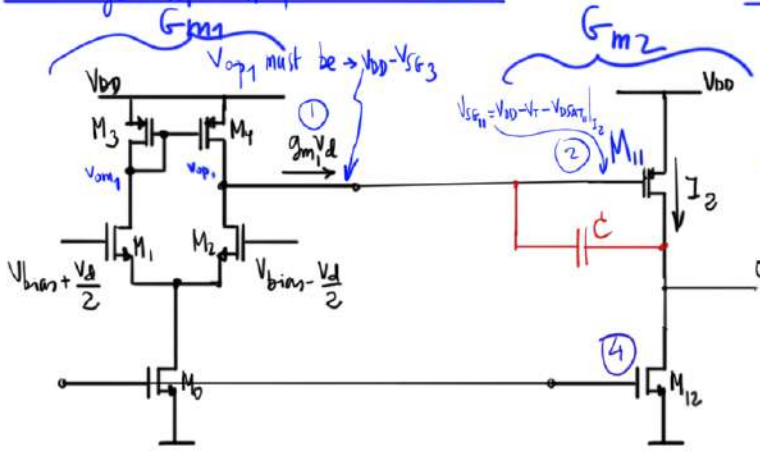
$$\omega_u = \frac{G_{m1}}{C}, \quad p_2 = -\frac{G_{m2}}{C_L + C_1}$$

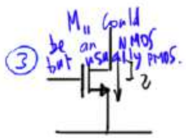
Same as the single stage.

Implementation:

1st stage: G_{m_1}: Diff pair transconductor

2nd stage: Single transistor = Another transconductor.





① From the output of the 1st transconductor a current $g_{m_1} v_d$ flows IF WE HAVE THE PROPER TERMINATION ON V_{op1}.
Quiescent voltage on V_{op_1} must be equal to V_{om_1} which is $V_{DD} - V_{SG_3}$.

② If we add a PMOS as 2nd stage, this needs to be biased with a specific I_2 that produces its $V_{SG_{11}}$ to be equal to V_{SG_3}.

$$V_{op_1} = V_{DD} - V_{SG_3} = V_{DD} - V_{SG_{11}} = V_{DD} - V_{T_p} - V_{DSAT_{11}}\Big|_{I_2}$$

Remember:

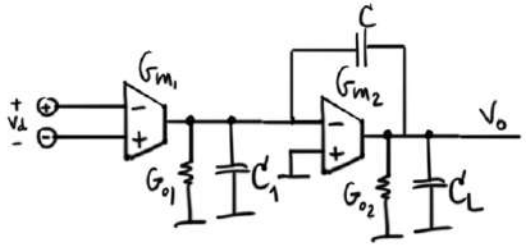
$V_{DSAT} = V_{OV} = V_{GS} - V_T \Big|$ at limit of saturation

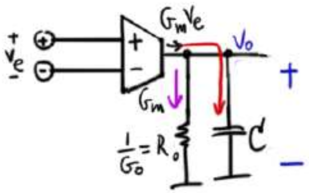
$$\boxed{V_{DSAT_{11}} = \sqrt{\frac{2 I_2}{\mu C_{ox} \frac{W_{11}}{L_{11}}}}}$$

③ We could use an NMOS 2nd stage but we would need to ensure its $V_{GS} = V_{Tn} + V_{DSAT}|_{I_2}$ was equal to $V_{DD} - V_{SG3}$ which is more inconvenient to match.

④ Load of 2nd stage is a current source, to get high DC gain on 2nd stage.

$|||$ **Miller-compensated op-amp.**





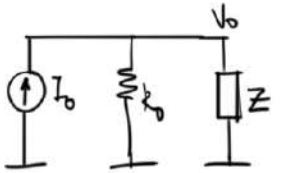
① Our intention is to get a voltage amplifier/integrator.

② We want to 'convert' the current $G_m V_e$ to a voltage V_o by pushing the current into the cap C.

③ * BUT some current flows into R_o hence we don't get all current into cap C. So we don't get full $V_o = \frac{G_m V_e}{sC}$ but a divided version $V_o = \frac{G_m V_e}{sC + G_o}$

⇓

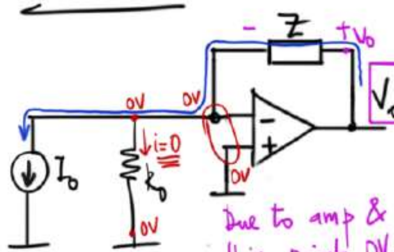
It's like this:



Due to R_o we don't get $V_o = I_o \cdot Z$ but we get $V = I_o \cdot (Z // R_o)$ (which is lower).

⇒

Solution = CCVS



$\boxed{V_o = I_o \cdot Z}$

Due to amp & feedback this point 0V ⇒ No current flows through R_o ⇒ All current flows through target Z ⇒

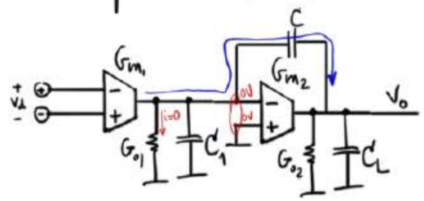
⇒ We achieve $V_o = I_o \cdot Z$ as desired.

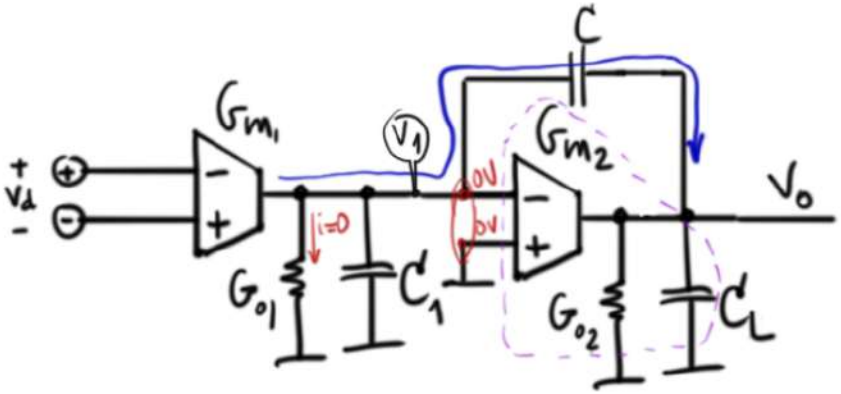
(Z can be a CAP)

⇒

This is why we stack the 2 stages like this:

(Miller-compensated op amp)





* It's 2^{nd} order. 2 poles.

(even though we have 3 caps, but they are in a loop, so only 2 cap voltages are independent).

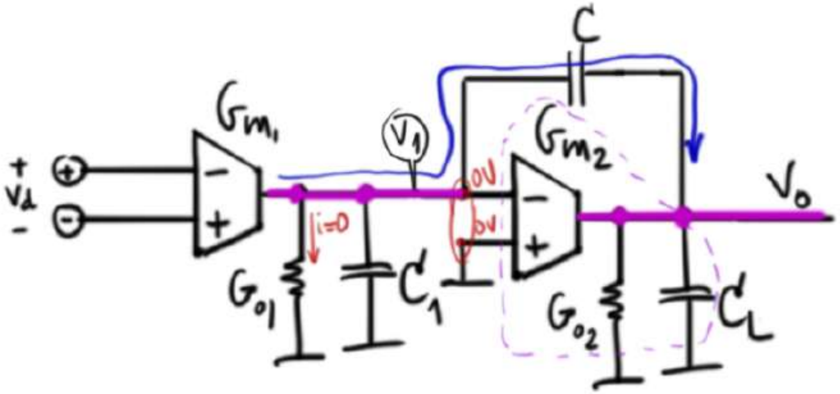
* DC gain:

at DC, caps are open circuit.

$I_{OUT_1} = G_{m_1} v_d$; $V_{OUT_1} = \frac{G_{m_1} v_d}{G_{o1}}$

$I_{OUT_2} = G_{m_2} V_{OUT_1} = \frac{G_{m_1} G_{m_2}}{G_{o1}} v_d$; $v_o = \frac{I_{OUT_2}}{G_{o2}} = \frac{G_{m_1}}{G_{o1}} \frac{G_{m_2}}{G_{o2}} v_d$; $\Rightarrow$

$$\Rightarrow \boxed{A_o = \left.\frac{v_o}{v_d}\right|_{DC} = \frac{G_{m_1}}{G_{o1}} \frac{G_{m_2}}{G_{o2}} = G_{m_1} R_{o1} \cdot G_{m_2} R_{o2}}$$



We have 2 nodes (V_1 & V_0).

$(Y \cdot V = I)$

Admitance matrix $\times$ Vector of voltages $=$ Vector of currents

$2\times 2 \quad \times \quad 2\times 1 \quad = \quad 2\times 1$

Current flowing into V_1 due to V_0

$$\begin{bmatrix} s(C_1+C)+G_{01} & -sC \\ G_{m_2}-sC & s(C_L+C)+G_{o2} \end{bmatrix} \times \begin{bmatrix} V_1 \\ V_0 \end{bmatrix} = \begin{bmatrix} -G_{m_1}V_e \\ 0 \end{bmatrix}$$

Total admittance hanging off node V_1

Current flowing into node V_0 (from G_{m2} & from C cap) due to V_1

Total admittance hanging off node V_0

We solve V_0 with Cramer's rule:

$$V_0 = \frac{\begin{vmatrix} s(C_1+C)+G_{01} & -G_{m1}v_e \\ G_{m2}-sC & 0 \end{vmatrix}}{\begin{vmatrix} s(C_1+C)+G_{01} & -sC \\ G_{m2}-sC & s(C_L+C)+G_{02} \end{vmatrix}} =$$

vector of currents substituted in place of 2nd column so we get V_0.

$$\overbrace{s^2(C_1+C)(C_L+C)}^{C_1C_L+\cancel{C^2}+C_1C+C_LC}+G_{01}G_{02}+$$

$$sG_{02}(C_1+C)+sG_{01}(C_L+C)$$

$$+sCG_{m2}-\cancel{s^2C^2}$$

$$= \frac{G_{m1}(G_{m2}-sC)v_e}{s^2(C_1C+CC_L+C_1C_L)+s\left[C(G_{m2}+G_{01}+G_{02})+C_LG_{01}+C_1G_{02}\right]+G_{01}G_{02}} \Rightarrow$$

$$\Rightarrow \boxed{\frac{V_o}{V_e} = \frac{G_{m1}(G_{m2} - sC)}{s^2(C_1C + CC_L + C_1C_L) + s\left[C(G_{m2} + G_{o1} + G_{o2}) + C_L G_{o1} + C_1 G_{o2}\right] + G_{o1}G_{o2}}}$$

2^{nd} order (as expected)

THE FULL TRANSFER FUNCTION OF THE 2-STAGE MILLER-COMPENSATED OP-AMP.

DC gain: $(s=0)$

$$\left.\frac{V_o}{V_e}\right|_{s=0} = \frac{G_{m1}G_{m2}}{G_{o1}G_{o2}}$$

(as expected)

2 poles:

* To get these values we do an APPROXIMATION.
Too complicated to solve $as^2+bs+c=0$
We assume $|s_1| \ll |s_2|$ (the 2 roots)
so $s_1 \simeq -\frac{c}{b}$ & $s_2 \simeq -\frac{b}{a}$

1 zero: $z_1 = +\frac{G_{m2}}{C}$

Using the previous approximation, the poles are at:

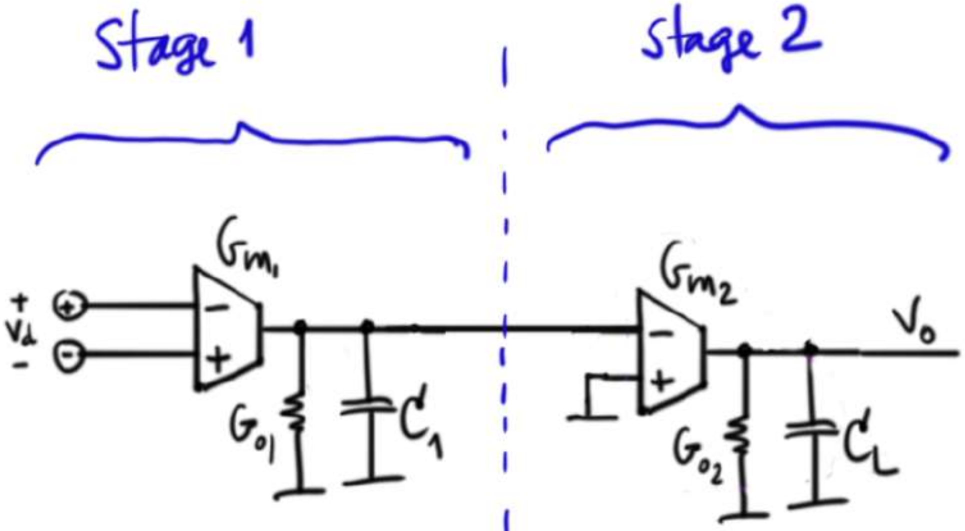
$$p_1 \simeq -\frac{c}{b} = \frac{G_{o1}G_{o2}}{C\left(G_{m2}+G_{o2}+G_{o1}\right)+C_1G_{o2}+C_LG_{o1}} = \boxed{\frac{G_{o1}}{C\left(\frac{G_{m2}}{G_{o2}}+1+\frac{G_{o1}}{G_{o2}}\right)+C_1+C_L\frac{G_{o1}}{G_{o2}}}}$$

p_1

$$p_2 \simeq -\frac{b}{a} = -\frac{C\left(G_{m2}+G_{o2}+G_{o1}\right)+C_1G_{o2}+C_LG_{o1}}{CC_1+C_1C_L+C_LC_1} \underset{\times\,\&\,\div\,(C+C_1)}{=} \boxed{-\frac{\frac{C}{C+C_1}G_{m2}+G_{o2}+G_{o1}\frac{C+C_L}{C+C_1}}{C_L+\frac{C_1C}{C+C_1}}}$$

p_2

Poles by inspection:

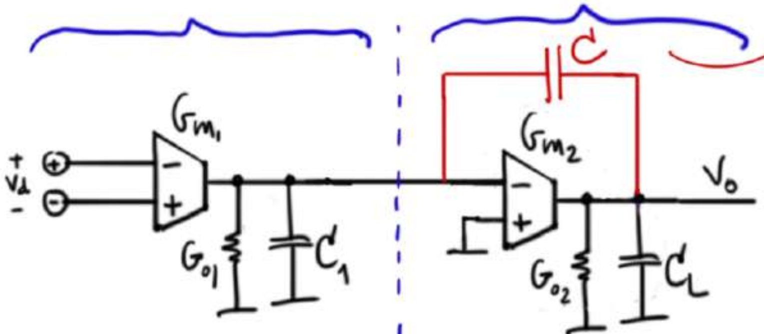


$$p_1 = -\frac{G_{m1}}{C_1}$$

$$p_2 = -\frac{G_{m2}}{C_L}$$

* If 2 stages concatenated, the 2 transfer functions multiply.
* So the 2 poles are just the 2 individual poles of the single stages.

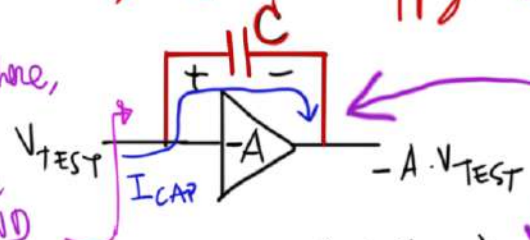
Poles by inspection:



$P_1 = -\frac{G_{o1}}{C_1}$ $\qquad$ $P_2 = -\frac{G_{o2}}{C_L}$

→ If cap around stage 2, we can no longer look at poles by inspection like before. ⇒ So we apply Miller's theorem.

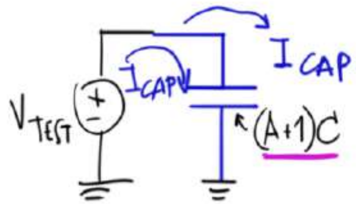
Seen from here, the cap C looks like a cap to GND of value $(A+1)C$.



MILLER'S THEOREM

$$I_{CAP} = \frac{V_{TEST} - (-A \cdot V_{TEST})}{1/sC} = (A+1)V_{TEST} \cdot sC$$

the same



$$I_{CAP} = \frac{V_{TEST}}{1/s(A+1)C} = V_{TEST} \cdot s(A+1)C$$

$C_{MM} = (A+1)C = \left(\frac{G_{m2}}{G_{o2}} + 1\right)C$

C_{MM} = "Miller-multiplied cap" seen from the INPUT side.

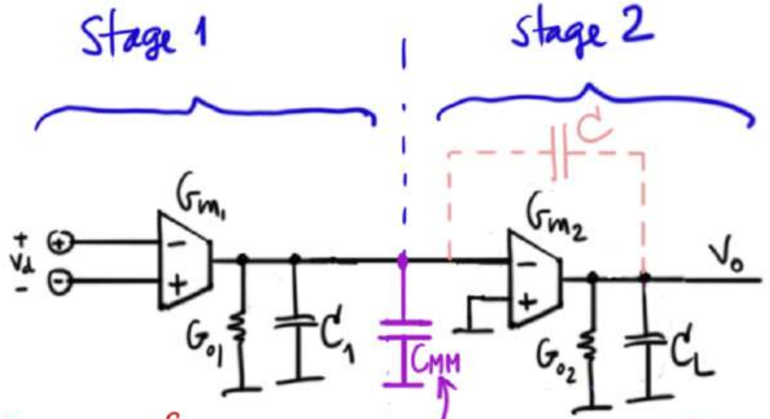
Let's call this "C_{MM}"

$C_{MM} = (A+1)C$

Where A is the voltage gain of the 2^{nd} stage $= A_{o2} = G_{m_2} \cdot R_{o2} = \frac{G_{m2}}{G_{o2}}$

So ⇒ $\boxed{C_{MM} = \left(1 + \frac{G_{m2}}{G_{o2}}\right)C}$

Poles by inspection:



$$P_1 \simeq -\frac{G_{o1}}{C_1 + C_{MM}} \simeq$$

$$P_2 = -\frac{G_{o2}}{C_L}$$

$$\simeq \frac{G_{o1}}{C_1 + \left(\frac{G_{m2}}{G_{o2}} + 1\right) C}$$

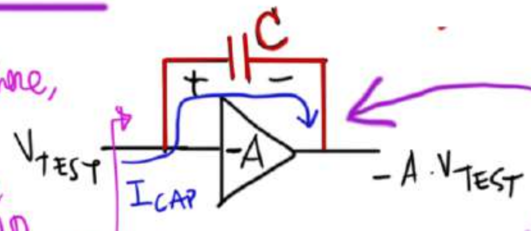
Approximation since 2nd stage is not an ideal opamp.

- New expression for the dominant pole p_1.
- Notice how it has shifted down in frequency (pole splitting)

Miller's theorem.

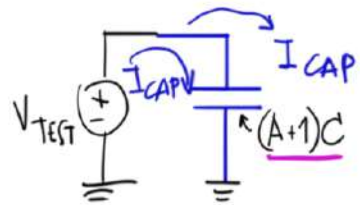
MILLER'S THEOREM

Seen from here, the cap C looks like a cap to GND of value $(A+1)C$.



$$I_{CAP} = \frac{V_{TEST} - (-A \cdot V_{TEST})}{1/sC} = (A+1) V_{TEST} \cdot sC$$

the same



$$I_{CAP} = \frac{V_{TEST}}{1/s(A+1)C} = V_{TEST} \cdot s\boxed{(A+1)C}$$

$$\boxed{C_{MM} = (A+1)C = \left(\frac{G_{m2}}{G_{o2}} + 1\right) C}$$

C_{MM} = "Miller-multiplied cap" seen from the INPUT side.

Without Miller cap C

$$p_1 = -\frac{G_{o1}}{C_1}$$

p_1 moved down to lower frequency

"POLE SPLITTING"

p_2 moved to higher frequency

$$p_2 = -\frac{G_{o2}}{C_L}$$

With Miller cap C

$$-\frac{G_{o1}}{C\left(\frac{G_{m2}}{G_{o2}} + 1 + \boxed{\frac{G_{o1}}{G_{o2}}}\right) + C_1 + \boxed{C_L \frac{G_{o1}}{G_{o2}}}} \simeq \frac{G_{o1}}{C_1 + \underbrace{\left(\frac{G_{m2}}{G_{o2}} + 1\right)C}_{C_{MM}}}$$

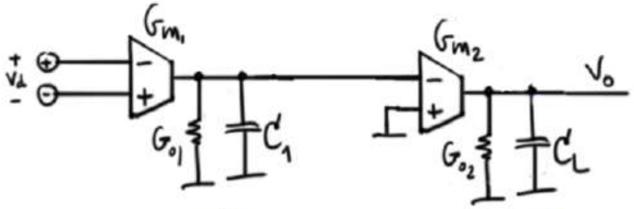
This is the expression for p_1 that we will always use

Approximation: this term $C_L \cdot \frac{G_{o1}}{G_{o2}}$ happens to be of very low significance when compared to C_1 and to $\left(\frac{G_{m2}}{G_{o2}} + 1\right)C$.

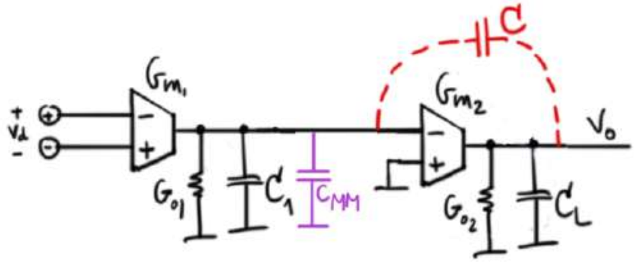
Approximation: This term $\frac{G_{o1}}{G_{o2}}$ is small in comparison with $\frac{G_{m2}}{G_{o2}} + 1$ so we can neglect it

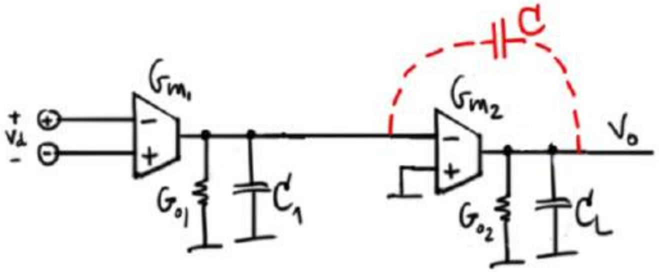
$$-\frac{\frac{C}{C+C_1} G_{m2} + G_{o2} + \boxed{G_{o1} \frac{C+C_L}{C+C_1}}}{C_L + \frac{C_1 C}{C+C_1}} \simeq \frac{G_{m2} \frac{C}{C+C_1} + G_{o2}}{C_L + \frac{C \cdot C_1}{C+C_1}}$$

Approximation: this term we also usually neglect it.



$$p_1 = -\frac{G_{o1}}{C_1} \qquad p_2 = -\frac{G_{o2}}{C_L}$$





2-STAGE OPAMP(*) At transconductor level

DC gain: $A_0 = \frac{G_{m1}}{G_{o1}} \frac{G_{m2}}{G_{o2}}$ ← Higher DC gain on 2-stage opamp compared to single stage. This is one key advantage.

$$p_1 \simeq -\frac{G_{o1}}{\underbrace{C\left(\frac{G_{m2}}{G_{o2}}+1\right)}_{C_{MM}}+C_1}$$

$$\boxed{\omega_u} = A_0 \cdot p_1 = \frac{G_{m1} G_{m2}}{\cancel{G_{o1}} G_{o2}} \cdot \frac{\cancel{G_{o1}}}{C\left(\frac{G_{m2}}{G_{o2}}+1\right)+C_1} =$$

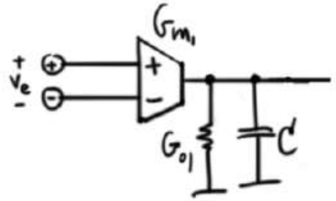
$$= \frac{G_{m1} G_{m2}}{\underbrace{C \cdot G_{m2}}_{\text{largest term}} + C\, G_{o2} + C_1} \simeq$$

$$\simeq \frac{G_{m1} \cancel{G_{m2}}}{C \cdot \cancel{G_{m2}}} \simeq \boxed{\frac{G_{m1}}{C}}$$

Same ω_u as the single stage one.

$$p_2 \simeq -\frac{G_{m2} \cdot \frac{C}{C+C_1} + G_{o2}}{C_L + \frac{C \cdot C_1}{C+C_1}}$$

$$z = +\frac{G_{m2}}{C}$$



SINGLE STAGE OPAMP(*)

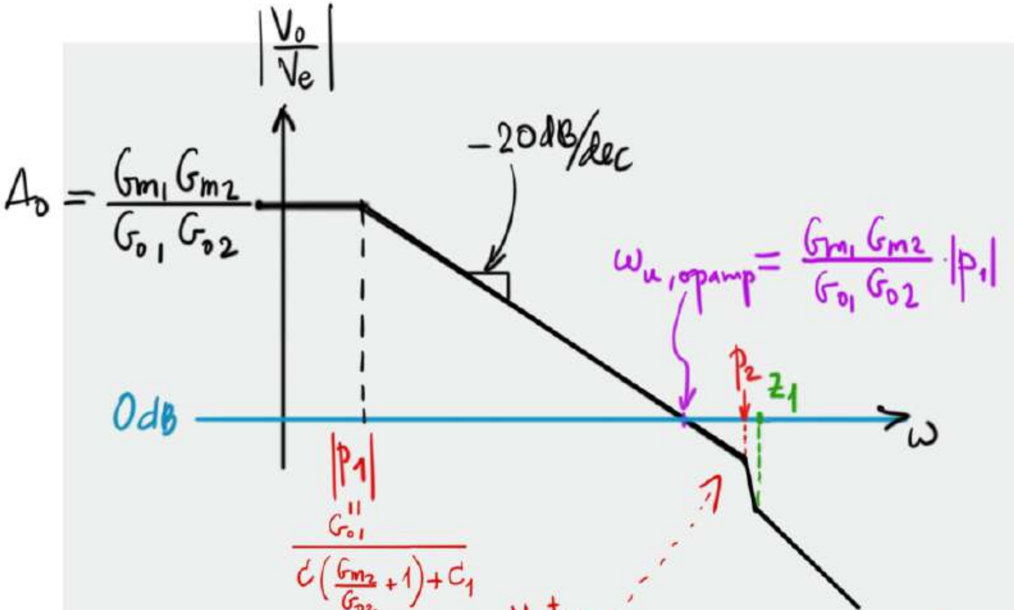
(*) At transconductor level, not at transistor level (no parasitic caps considered, hence just single pole system)

DC gain: $A_0 = G_{m1} \cdot R_{OUT,1} = \frac{G_{m1}}{G_{o1}}$

$$p_1 = \frac{G_{o1}}{C}$$

$$\omega_u = A_0 \cdot p_1 = \frac{G_{m1}}{C}$$

2-STAGE OPAMP(*) At transconductor level



Note:
Here we're drawn p_2 after ω_u which would be good for stability. However it's not like p_2 value will always be higher than ω_u, (as was the case in the single stage opamp), so we need to ensure p_2 is at high enough freq.

We call $\omega_{u,opamp}$ the "open loop" transfer function of the circuit, WITHOUT FEEDBACK

$$\Rightarrow \omega_{u,opamp} = \underbrace{\frac{G_{m1} G_{m2}}{\cancel{G_{o1}} G_{o2}}}_{A_0} \cdot \underbrace{\frac{\cancel{G_{o1}}}{C\left(\frac{G_{m2}}{G_{o2}}+1\right)+C_1}}_{p_1} \simeq \frac{G_{m1}}{C}$$

$$\omega_{u,opamp} = A_0 \cdot p_1$$

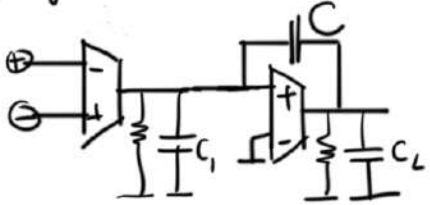
This matches the ω_u of the single stage opamp.

Is the 2-stage op-amp "unconditionally" stable?

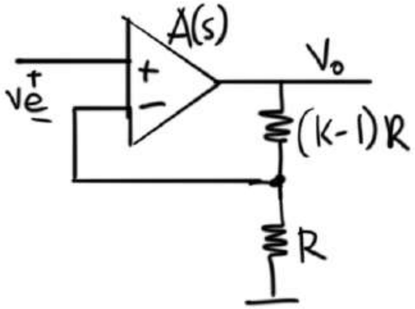
No, not unconditionally.

The transfer function of the opamp (WITHOUT FEEDBACK)

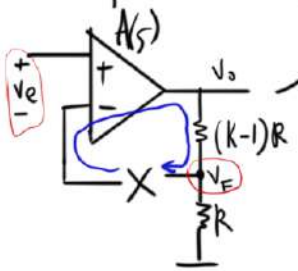
$$A(s) = \frac{V_o}{V_e} = \frac{A_o\left(1 - \frac{s}{z_1}\right)}{\left(1 + \frac{s}{-p_1}\right)\left(1 + \frac{s}{-p_2}\right)}$$



Now if we use this in a feedback configuration:



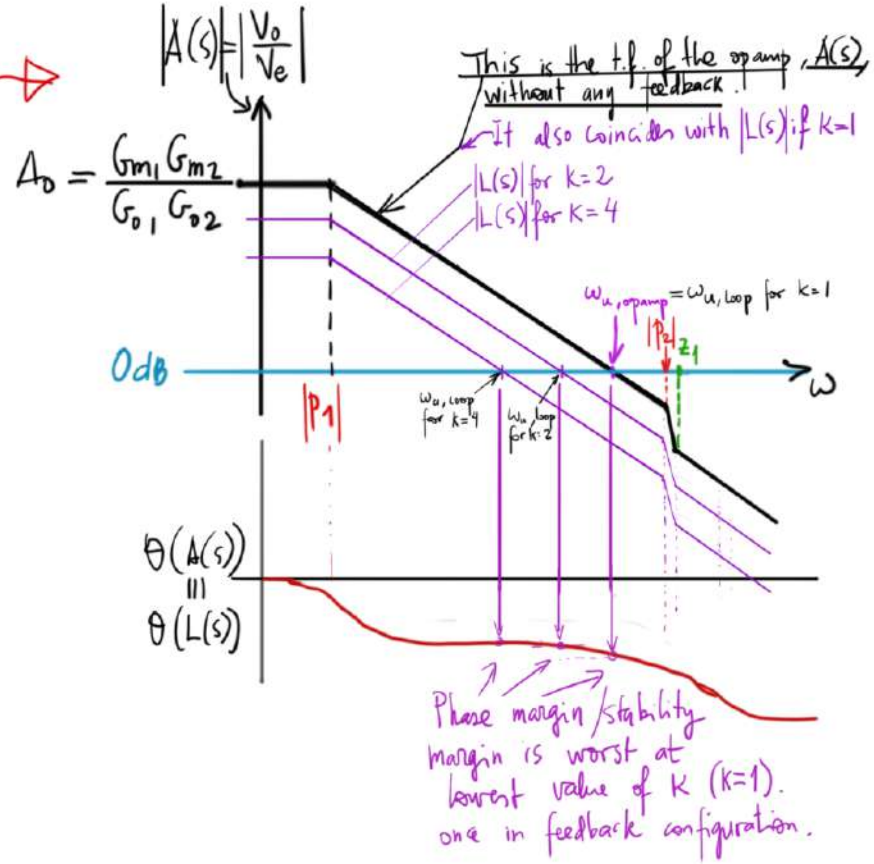
The LOOP GAIN $L(s)$ of this is (breaking the loop & going around it)



$$V_o = A(s)\underbrace{(V_p - V_m)}_{V_e} \;;\; V_e = \frac{V_o}{A(s)}$$

$$V_F = \frac{R}{(K-1)R + R} V_o = \frac{R}{KR} V_o = \frac{V_o}{K}$$

$$\boxed{L(s) = \frac{V_F}{V_e} = \frac{\cancel{V_o}/K}{\cancel{V_o}/A(s)} = \frac{A(s)}{K}}$$

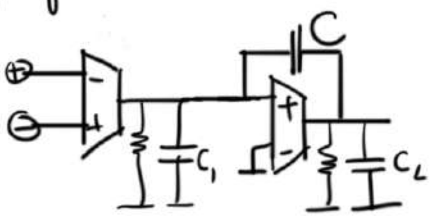


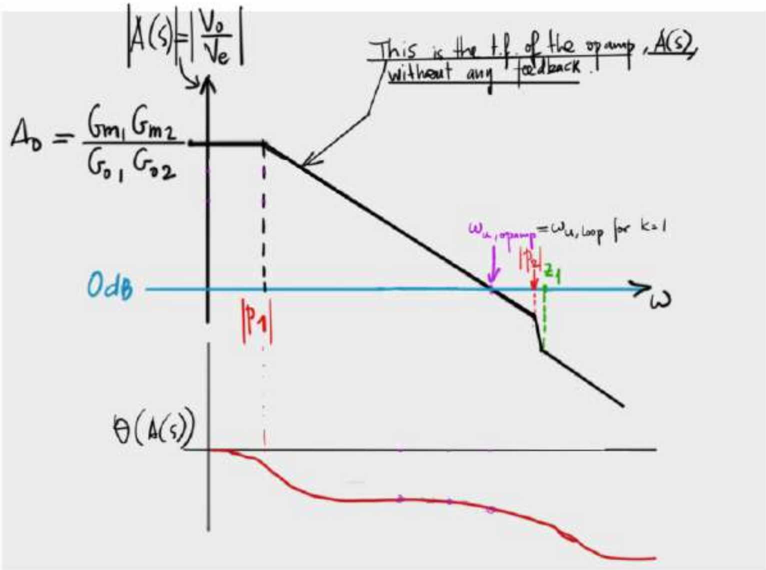
Is the 2-stage op-amp "unconditionally" stable?

No, not unconditionally.

The transfer function of the opamp (WITHOUT FEEDBACK)

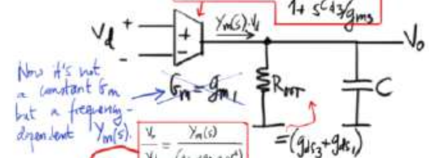
$$A(s)=\frac{V_o}{V_e}=\frac{A_o\left(1-\frac{s}{z_1}\right)}{\left(1+\frac{s}{-p_1}\right)\left(1+\frac{s}{-p_2}\right)}$$





REMINDER, the single stage op amp (5T) with para' caps was UNCONDITIONALLY STABLE but the 2-stage is not.

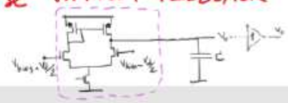
$$\boxed{Y_m(s)=g_{m1}\cdot\frac{1+sC_{d3}/2g_{m3}}{1+sC_{d3}/g_{m3}}}$$

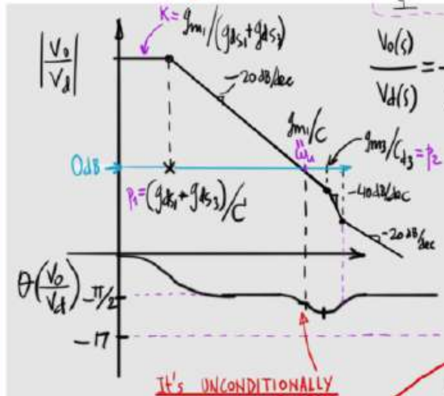


Now it's not a constant G_m → ~~$G_m=g_{m1}$~~ but a frequency-dependent $Y_m(s)$.

$$\boxed{\frac{V_o}{V_d}=\frac{Y_m(s)}{(g_{ds3}+g_{ds1}+sC)}}$$

This is the freq response WITHOUT FEEDBACK





$$\frac{V_o(s)}{V_d(s)}=\frac{g_{m1}}{g_{ds1}+g_{ds3}+sC}\cdot\frac{1+s\frac{C_{d3}}{2g_{m3}}}{1+s\frac{C_{d3}}{g_{m3}}}$$

Here we have "assumed" that p_2 comes after the 0dB crossing point.

i.e. $p_2>\omega_u$

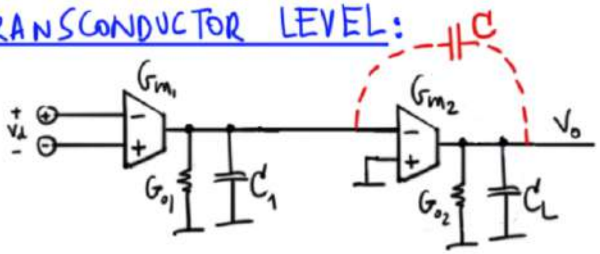
$$\frac{g_{m3}}{C_{d3}}>\frac{g_{m1}}{C}$$

Hence it crosses 0dB with -20dB/dec slope and it's stable.

It's UNCONDITIONALLY STABLE

If C (load) >>> C_{d3} (parasitics) then p_2 is after ω_u.

TRANSCONDUCTOR LEVEL:



2-STAGE OPAMP (*) At transconductor level

DC gain: $A_0 = \frac{G_{m1}}{G_{o1}}\frac{G_{m2}}{G_{o2}}$ ← Higher DC gain on 2-stage opamp compared to single stage. This is one key advantage.

$$p_1 \simeq -\frac{G_{o1}}{\underbrace{C\left(\frac{G_{m2}}{G_{o2}}+1\right)}_{C_{MM}}+C_1}$$

$$\boxed{\omega_u} = A_0 \cdot p_1 = \frac{G_{m1}G_{m2}}{\cancel{G_{o1}}G_{o2}} \cdot \frac{\cancel{G_{o1}}}{C\left(\frac{G_{m2}}{G_{o2}}+1\right)+C_1} = \frac{G_{m1}G_{m2}}{\underbrace{C\,G_{m2}}_{\text{longest term}}+C\,G_{o2}+C_1} \simeq \frac{G_{m1}\cancel{G_{m2}}}{C\cdot\cancel{G_{m2}}} \simeq \boxed{\frac{G_{m1}}{C}}$$

Same ω_u as the single stage one.

$$p_2 \simeq -\frac{G_{m2}\cdot\frac{C}{C+C_1}+G_{o2}}{C_L+\frac{C\cdot C_1}{C+C_1}}$$

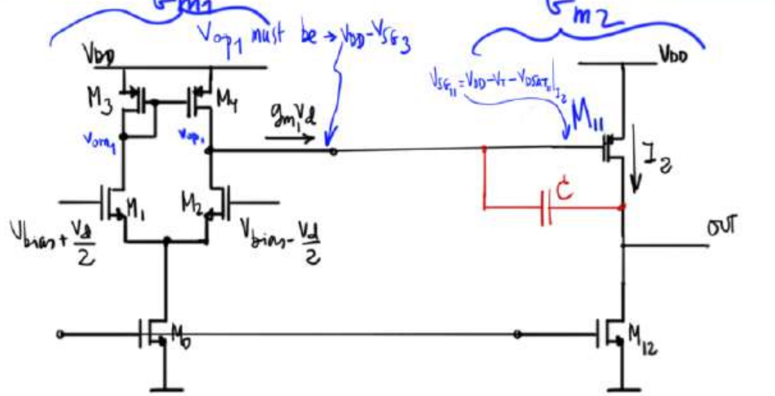
$$z = +\frac{G_{m2}}{C}$$

$$\text{TF: } \frac{V_o}{V_e} = A_0 \cdot \frac{1-S/z_1}{\left(1+\frac{S}{-p_1}\right)\left(1+\frac{S}{-p_2}\right)}$$

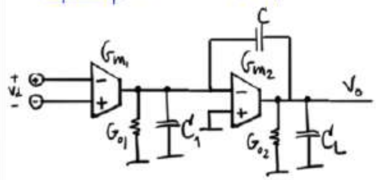
TRANSISTOR LEVEL:

1st stage: G_{m1}: Diff pair transconductor

2nd stage: Single transistor



Miller-compensated op-amp. |||



DC gain: $A_0 = \frac{G_{m1}}{G_{o1}}\frac{G_{m2}}{G_{o2}}$

$$\omega_u \simeq \frac{G_{m1}}{C}$$

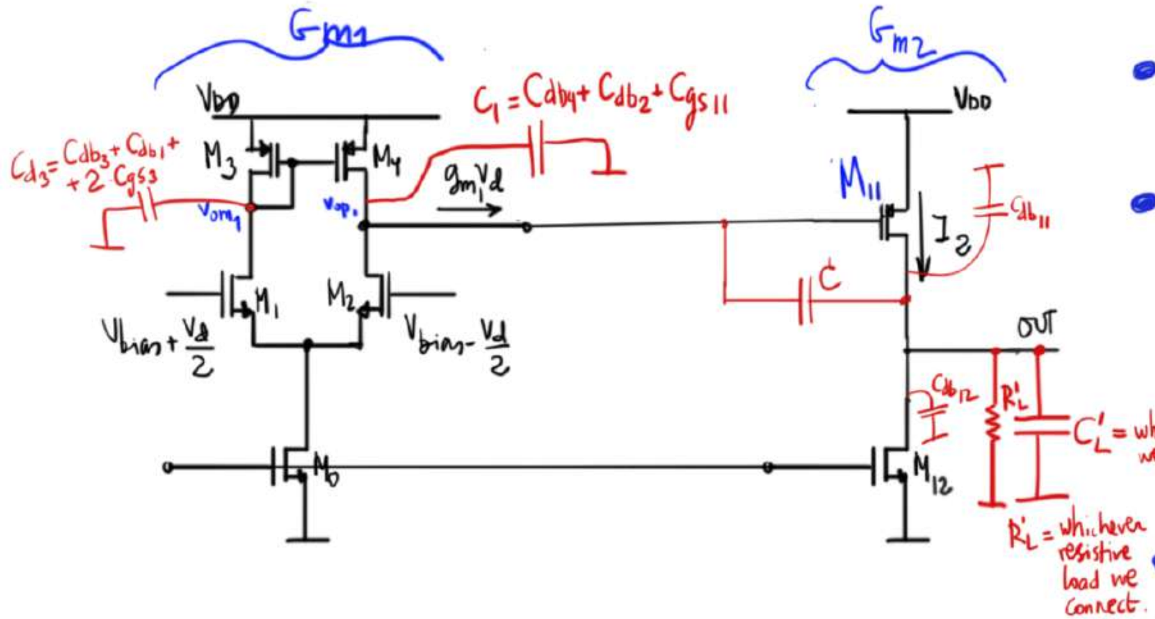
$$p_1 \simeq -\frac{G_{o1}}{\underbrace{C\left(\frac{G_{m2}}{G_{o2}}+1\right)}_{C_{MM}}+C_1}$$

$$p_2 \simeq -\frac{G_{m2}\cdot\frac{C}{C+C_1}+G_{o2}}{C_L+\frac{C\cdot C_1}{C+C_1}}$$

$$z = +\frac{G_{m2}}{C}$$

$$\text{TF: } \frac{V_o}{V_e} = A_0 \cdot \frac{1-S/z_1}{\left(1+\frac{S}{-p_1}\right)\left(1+\frac{S}{-p_2}\right)}$$

LET'S IDENTIFY COMPONENT VALUES:

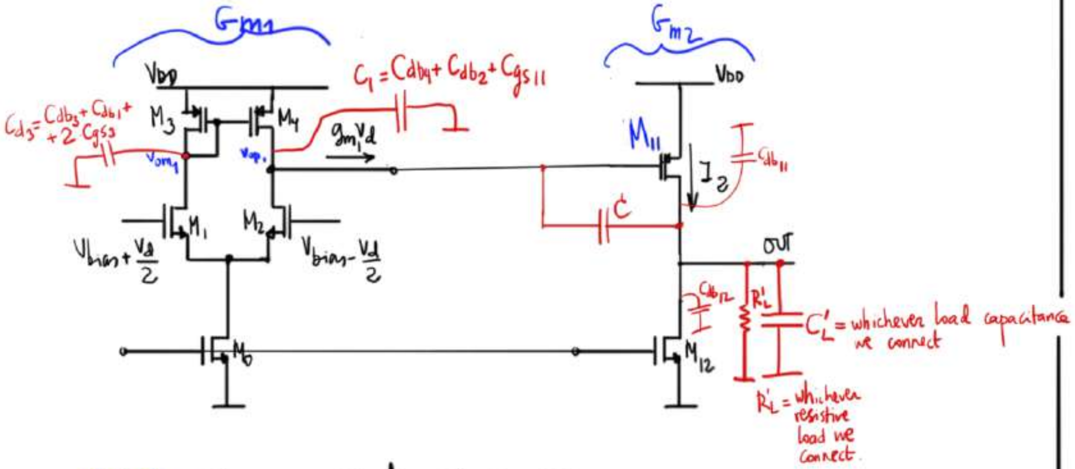


- $C_1 = C_{db4} + C_{db2} + C_{gs11}$
- $C_L = C_L' + \underbrace{C_{db12}} + \underbrace{C_{db11}}$

  Actual values of these we usually just get the values from the simulator.

- $G_{01} = g_{ds1} + g_{ds3} \quad \left(= \frac{1}{r_{ds1} /\!/ r_{ds3}} \right)$
- $G_{02} = g_{ds11} + g_{ds12} + G_L'$
- $Y_m(s) = g_{m1} \cdot \dfrac{1 + s\frac{C_{d3}}{2 \cdot g_{m3}}}{1 + s\frac{C_{d3}}{g_{m3}}}$ ← $Y_m(s)$ is what we use in place of G_{m1}
- $G_{m2} = g_{m11}$

AT TRANSISTOR LEVEL:



NOISE: Same as single stage one.

$$S_{vin} = \frac{16}{3}\frac{KT}{g_{m1}}\left(1+\frac{g_{m3}}{g_{m1}}\right)$$

OFFSET:
- Random: Same as single stage one.
$$\sigma_{V_{os}}^2 = \sigma_{V_{T12}}^2 + \sigma_{V_{T34}}^2$$
- Systematic: Minimize it by equalizing current densities M_4 & M_{11}.
Otherwise $\Delta V = \dfrac{V_{SG_3}\big|_{I_0/2} - V_{SG_{11}}\big|_{I_{12}}}{\dfrac{g_{m1}}{g_{ds1}+g_{ds3}}}$

TF of 2-stage opamp:

$$\text{TF}: \quad \frac{V_o}{V_e} = A_0 \cdot \underbrace{\frac{1+s\frac{C_{d3}}{2g_{m3}}}{1+s\frac{C_{d3}}{g_{m3}}}}_{\text{from } Y_m(s)} \cdot \frac{1-s/z_1}{\left(1+\frac{s}{-p_1}\right)\left(1+\frac{s}{-p_2}\right)}$$

Inserting values:

Note: Here it's better to insert values into the "sub-components" of the TF (DC gain, $p_1, p_2, p_3 \dots$) rather than trying to get a long expression for TF.

$$\left\{\begin{aligned} &\text{DC gain}: \; A_0 = \frac{G_{m1}}{G_{o1}}\frac{G_{m2}}{G_{o2}} \\ &p_1 \simeq -\frac{G_{o1}}{C\underbrace{\left(\frac{G_{m2}}{G_{o2}}+1\right)}_{C_{MM}}+C_1} \\ &p_2 \simeq -\frac{G_{m2}\cdot\frac{C}{C+C_1}+G_{o2}}{C_L+\frac{C\cdot C_1}{C+C_1}} \\ &z_1 = +\frac{G_{m2}}{C} \\ &\omega_u \simeq \frac{G_{m1}}{C} \end{aligned}\right.$$

Specific values at transistor Level:

- $A_0 = \dfrac{g_{m1}}{g_{ds1}+g_{ds3}} \cdot \dfrac{g_{m11}}{g_{ds11}+g_{ds12}+G_L'}$

- $\omega_u = \dfrac{g_{m1}}{C}$; • $p_1 = -\dfrac{g_{ds1}+g_{ds3}}{C\left(\frac{g_{m11}}{G_{o2}}+1\right)+C_1}$ ← whatever, just substitute

- $p_2 = -\dfrac{g_{m11}\cdot\frac{C}{C+C_1}+G_{o2}}{\frac{C\cdot C_1}{C+C_1}+C_L}$; • $p_3 = -\dfrac{g_{m3}}{C_{d3}}$ ← Coming from $Y_m(s)$

- $z_1 = +\dfrac{g_{m11}}{C}$; • $z_2 = -\dfrac{2g_{m3}}{C_{d3}}$ ← Coming from $Y_m(s)$

$\left.\begin{aligned} g_{m1} &\sim g_{m3} \\ C &\gg C_{d3}\end{aligned}\right\}$ z_1 lower freq than z_2

EXAMPLE:

Let's say:

We want $A_0 = 1000$

with an $R_L = 1\,K\Omega$
$(G_L = 1\,mS)$

With a single stage opamp:

$$A_0 = g_{m_1}\underbrace{(r_{ds_1} \| r_{ds_3} \| R_L)}_{\simeq R_L} \simeq g_{m_1} R_L \Rightarrow$$

$$\Rightarrow g_{m_1} = \frac{A_0}{R_L} = \frac{1000}{1000\,\Omega} = 1\,S = 1000\,mS$$

To get a g_{m_1} of $1000\,mS$ we would need a huge current so big power dissipation.

With a 2-stage opamp:

$$A_0 = \frac{g_{m_1}}{g_{ds_1} + g_{ds_3}} \cdot \frac{g_{m_{11}}}{g_{ds_{11}} + g_{ds_{12}} + \underbrace{G_L'}_{1K\Omega}} \simeq \underbrace{\frac{g_{m_1}}{g_{ds_1} + g_{ds_3}}}_{\sim 100 \text{ from } 1^{st} \text{ stage}} \cdot \underbrace{\frac{g_{m_{11}}}{G_L'}}_{10 \text{ from } 2^{nd} \text{ stage}} = 100 \times 10 = 1000$$

From 2^{nd} stage we need to get only a gain of ~ 10:

$$\frac{g_{m_{11}}}{G_L'} = 10 \Rightarrow g_{m_{11}} = 10 \cdot G_L' = \frac{10}{R_L} = \frac{10}{1K\Omega} = \frac{1}{100} = 10\,mS$$

From 1^{st} stage we need to get a gain of ~ 100:

$\dfrac{g_{m_1}}{g_{ds_1} + g_{ds_3}} = 100 \rightarrow$ From stability considerations we know that G_{m_1} has to be significantly less than G_{m_2}. So if $G_{m_2} = g_{m_{11}} = 10\,mS$, we make $g_{m_1} \simeq 2\,mS$ and we get $g_{ds_1} + g_{ds_3} \simeq$

$$g_{ds_1} + g_{ds_3} = \frac{1}{(r_{ds_1} \| r_{ds_3})} =$$

$$\frac{g_{m_1}}{g_{ds_1} + g_{ds_3}} = 100; \quad = \frac{g_{m_1}}{100} = \frac{2\,mS}{100} = 0.02\,S \Rightarrow$$

$$\Rightarrow r_{ds_1} \| r_{ds_3} = \frac{1}{0.02\,S} = 50\,K\Omega$$

These values are more sensible & easy to achieve