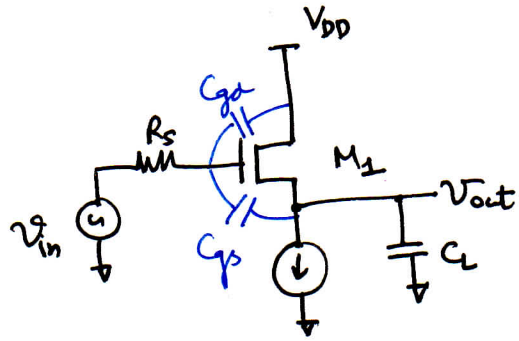
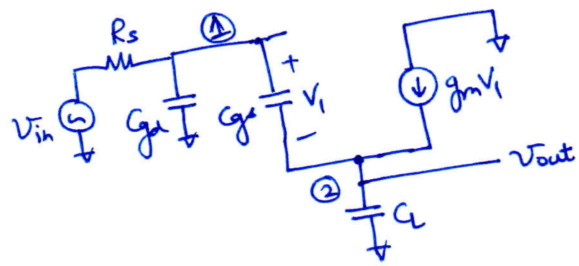
Source follower frequency Response





* strong interaction between nodes ① & ② through C_{gs}.

↳ can't associate a pole with each node

node ② $V_1 C_{gs} s + g_m V_1 = V_{out} C_L s$

$$\Rightarrow V_1 = \frac{C_L s}{g_m + C_{gs} \cdot s} V_{out}$$

* KVL beginning from V_{in}

$$V_{in} = R_S \left[V_1 C_{gs} s + (V_1 + V_{out}) C_{gd} s \right] + V_1 + V_{out}$$

Substitute V_1

$$\Rightarrow \frac{V_{out}}{V_{in}}(s) = \frac{(g_m + C_{gs} s)}{R_S \left[C_{gs} C_L + C_{gs} C_{gd} + C_{gd} C_L \right] s^2 + \left[g_m R_S C_{gd} + C_L + C_{gs} \right] s + g_m}$$

$\Rightarrow$ LHP zero

$\omega_z = -\frac{g_m}{C_{gs}}$ $\Rightarrow$ signal conducted by C_{gs}, adds to the signal carried by the main transistor with the same polarity.

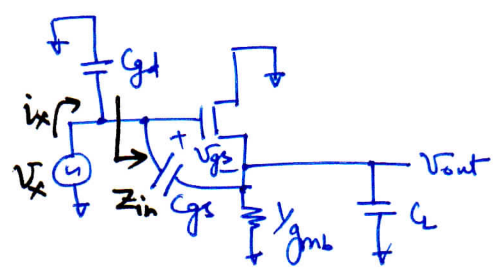
Assuming $|\omega_{p1}| \ll |\omega_{p2}|$

Dominant pole is at

$$\omega_{p1} \cong \frac{g_m}{g_m R_s C_{gd} + C_L + C_{gs}}$$

$$= \frac{1}{R_s C_{gd} + \frac{C_L + C_{gs}}{g_m}} = \frac{g_m}{C_L + C_{gs}} \text{ iff } R_s = 0.$$

Input impedance of SF:



* C_{gd} shunts the input

$$v_{gs} = \frac{i_x}{sC_{gs}} \Rightarrow i_d = g_m v_{gs} = \frac{g_m i_x}{sC_{gs}}$$

$$\Rightarrow v_x = \frac{i_x}{C_{gs}} + \left(i_x + \frac{g_m i_x}{sC_{gs}}\right)\left(\frac{1}{g_{mb}} \| \frac{1}{sC_L}\right)$$

$$\Rightarrow Z_{in_1} = \frac{1}{sC_{gs}} + \left(1 + \frac{g_m}{sC_{gs}}\right)\frac{1}{g_{mb} + sC_L}$$

@ low-frequencies

$$g_{mb} \gg |sC_L|$$

$$\Rightarrow Z_{in_1} \cong \frac{1}{sC_{gs}}\left(1 + \frac{g_m}{g_{mb}}\right) + \frac{1}{g_{mb}} = \frac{1}{\frac{sC_{gs}}{\left(1 + \frac{g_m}{g_{mb}}\right)}} + \frac{1}{g_{mb}}$$

$\Rightarrow$ Input cap is a fraction of $C_{gs} + C_{gd}$.

At higher frequencies

$$g_{mb} << |sC_L|$$

$$\Rightarrow Z_{in_1} \simeq \frac{1}{sC_{gs}} + \frac{1}{sC_L} + \frac{g_m}{C_{gs} C_L s^2}$$

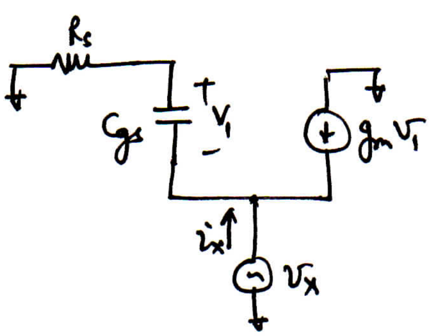
$\Rightarrow$ for same $s = j\omega$

$$Z_{in_1} = \underbrace{\frac{1}{j\omega}\left[\frac{1}{C_{gs}} + \frac{1}{C_L}\right]}_{\text{Capacitance}} - \underbrace{\frac{g_m}{C_{gs} C_L \omega^2}}_{\text{negative resistance}}$$

$$Re(Z_{in}) = R_{in_1} = \frac{-g_m}{C_{gs} C_L \omega^2} \leftarrow \text{useful in oscillators.}$$

Output impedance:

neglecting g_{mb} & C_{gd}



$$V_1 C_{gs} s + g_m V_1 = -i_x \longrightarrow (1)$$

&

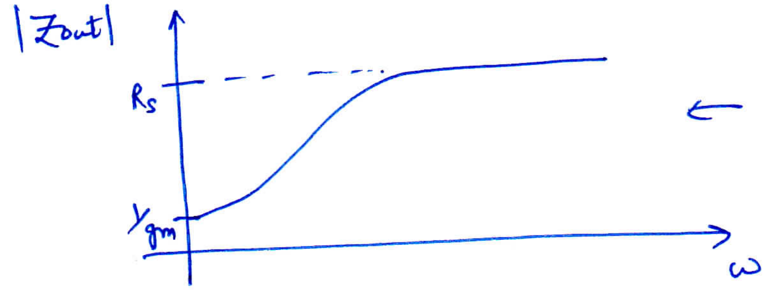
$$V_1 C_{gs} s R_s + V_1 = -V_x \longrightarrow (2)$$

$\frac{(2)}{(1)} \Rightarrow$

$$Z_{out} = \frac{V_x}{i_x} = \frac{R_s C_{gs} s + 1}{g_m + sC_{gs}}$$

@ low frequencies ⇒ $Z_{out} \approx 1/g_m$ ⟶ Expected from a SF

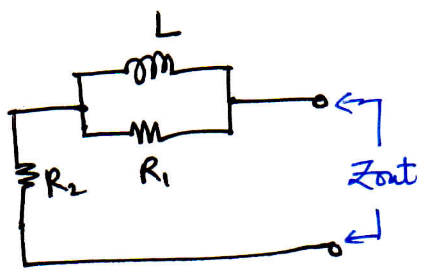
@ very high frequencies ⇒ $Z_{out} \simeq R_s$ ($\because C_{gs}$ shorts gate & source)



← realistic scenario as in a buffer
$1/g_m \ll R_s$

* Since the Z_{out} ↑ with frequency
↳ should contain an inductive component!
($j\omega L$)

Equivalent impedance can be derived to be:



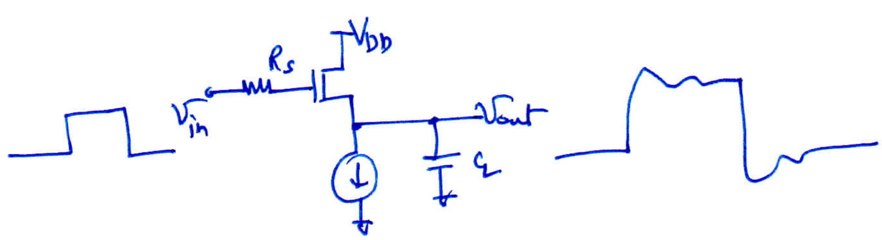
$R_2 = 1/g_m$

$R_1 = R_s - 1/g_m$

$L = \frac{C_{gs}}{g_m}\left(R_s - 1/g_m\right)$

$\because L = \frac{C_{gs}}{g_m}\left(R_s - 1/g_m\right)$

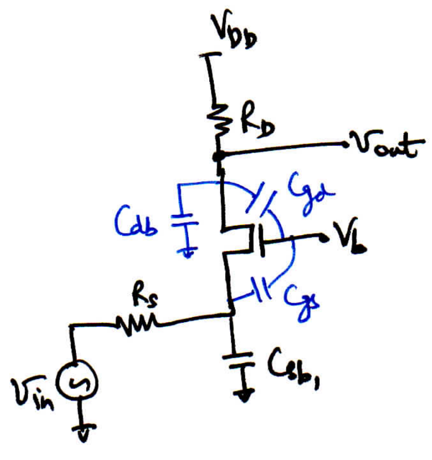
⇒ If the SF is driven by a large R_s
⇒ then it exhibits substantial inductive behavior
↳ manifests as ringing



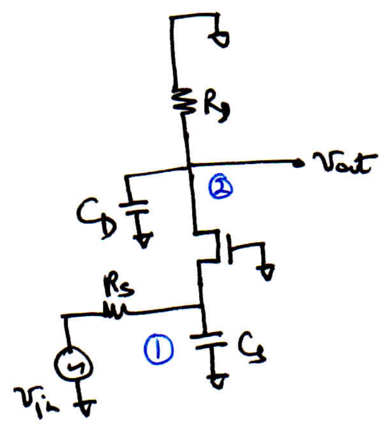
* Useful as gain peaking loads!

Common Gate Frequency Response

In a CG, input and output nodes are 'isolated' if $\lambda = 0$.



$\Rightarrow$



@ node ①

$$C_S = C_{gs_1} + C_{sb_1}$$

\+ Since the nodes are isolated $\Rightarrow$ no Miller cap

$\Rightarrow$ no zeros in the frequency response

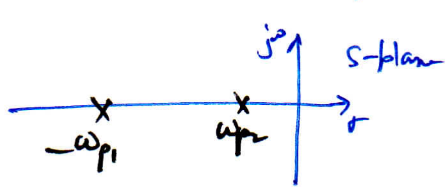
$$\Rightarrow \omega_{p_1} = \omega_{in} = \frac{1}{\left(R_S \,\|\, \frac{1}{g_m + g_{mb}}\right) \cdot C_S}$$

@ node ②

$$C_D = C_{gd_1} + C_{db_1}$$

$$\Rightarrow \omega_{p_2} = \omega_{out} = \frac{1}{R_D C_D}$$

$$\Rightarrow \frac{V_{out}}{V_{in}}(s) = \frac{(g_m + g_{mb}) R_D}{1 + (g_m + g_{mb}) R_S} \cdot \frac{1}{(1 + s/\omega_{p_1})(1 + s/\omega_{p_2})}$$



⇒ No Miller multiplication of Capacitor

↳ potentially wide-band

↳ low impedance input may load the driving stage

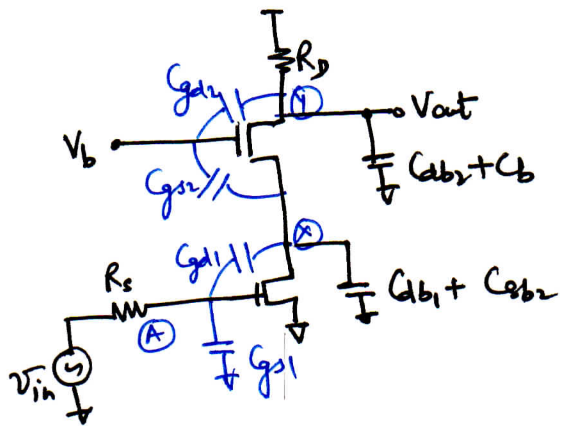
↳ good for current input (i.e. transimpedance) amplifiers

As derived earlier

$$Z_{in} \simeq \frac{Z_L}{(g_m + g_{mb}) r_o} + \frac{1}{(g_m + g_{mb})}$$

where $Z_L = \left(R_D \,||\, \frac{1}{sC_D} \right)$

Cascode Amplifier frequency Response:



* Miller effect of C_{gd1} is determined by the gain from A to X

↳ Assuming low value of R_D, this gain is $-g_{m1} \times \frac{1}{g_{m2}+g_{mb2}}$ (assuming $\gamma=0$)

$$\simeq \frac{-g_{m1}}{g_{m2}} \simeq -1$$

* Assume M_1 & M_2 to be identical

⇒ C_{gd1} is multiplied by 2x, instead of $|A_v|$

⇒ Miller effect is less significant in cascode amplifiers than in a CS stage. (normal CS)

$$\Rightarrow \omega_{pA} = \frac{1}{R_S\left[C_{gs1} + \left(1+\frac{g_{m1}}{g_{m2}+g_{mb2}}\right)C_{gd1}\right]}$$

* pole due to node X

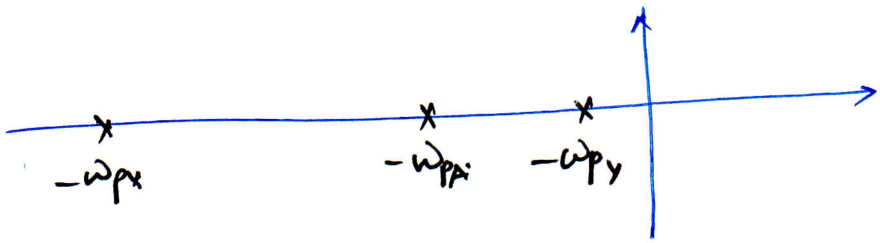
$$C_X = 2C_{gd1} + C_{db1} + C_{sb2} + C_{gs2}$$

$$\Rightarrow \omega_{pX} = \frac{g_{m2}+g_{mb2}}{2C_{gd1} + C_{db1} + C_{sb2} + C_{gs2}}$$

* Output pole: → third pole

$$\omega_{py} = \frac{1}{R_D(C_{db2} + C_L + C_{gd2})}$$

→ Must choose ω_{px} to be farther away from the other two
↳ important for stability



* What if R_D is replaced by a cascode current source load?
↳ higher DC gain → $A_{DC} = (g_{mn} r_{on}^2) \| (g_{mp} r_{op}^2)$.

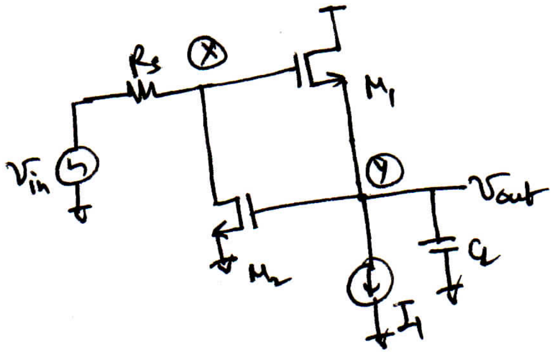
⇒ Impedance at node ⓧ reaches higher value (Derive this yourself)
↳ pole at node ⓧ can be quite lower than $\frac{(g_{m2} + g_{m2b})}{C_X}$

↳ But in practice this pole doesn't degrade stability much.

⇒ But always be mindful of such scenarios!!

Exercise:

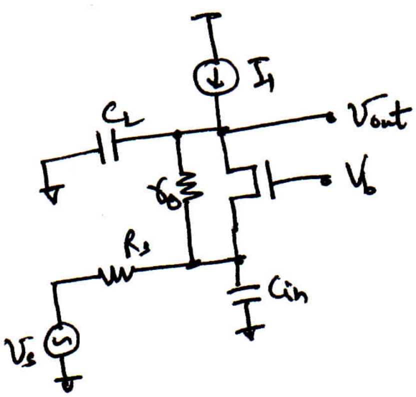
1. Find the transfer function of the circuit (Assume $\lambda = 0$, $\gamma = 0$)



Ans: $$\frac{V_{out}}{V_{in}}(s) = \frac{g_{m1} + C_{xy}s}{R_S \xi s^2 + [C_y + g_{m1} R_S C_x + (1+g_{m2}R_S) C_{xy}]s + g_{m1}(1+g_{m2}R_S)}$$

$$C_{xy} = C_{gs1} + C_{gd2}$$

2. Find $\frac{V_{out}}{V_{in}}(s)$ & Z_{in} for the circuit ($\lambda \neq 0$)



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

$$Z_{in} = \frac{1}{g_m + g_{mb}} + \frac{1}{sC_L}\,\frac{1}{(g_m + g_{mb}) r_o}$$