THANKS TO THE LOADING EFFECTS $a_{12} \ll a_1^{CE}$ AS THE MILLER MULTIPLIER IS STRONGLY REDUCED.

AS A RESULT $\omega_{H_2} \gg \omega_H^{CE} \Rightarrow \omega_H > \omega_H^{CE}$!!

THE BUFFER STAGE ALLOWS TO MAKE THE AMPLIFIER BANDWIDTH LARGER.

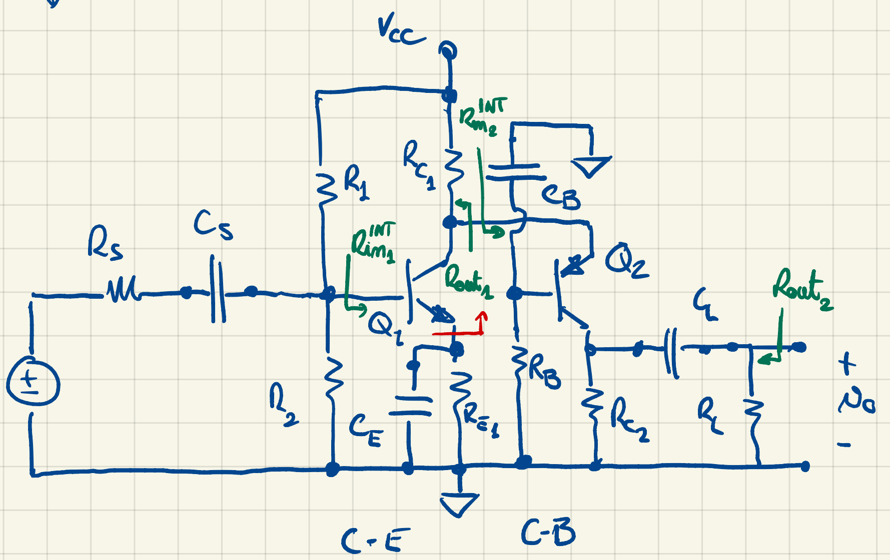
N.B. DO NOT FORGET THE POTENTIAL RESONANCE BETWEEN Z_{out}^{CC} AND Z_{in}^{CE}

ADDITIONAL DETAILS: $m = 3$ EVEN IF $N = 4$ AT HIGH FREQUENCY, BECAUSE WE HAVE A LOOP INVOLVING $C_{\pi_1}, C_{\mu_1}, C_{\pi_2}$.

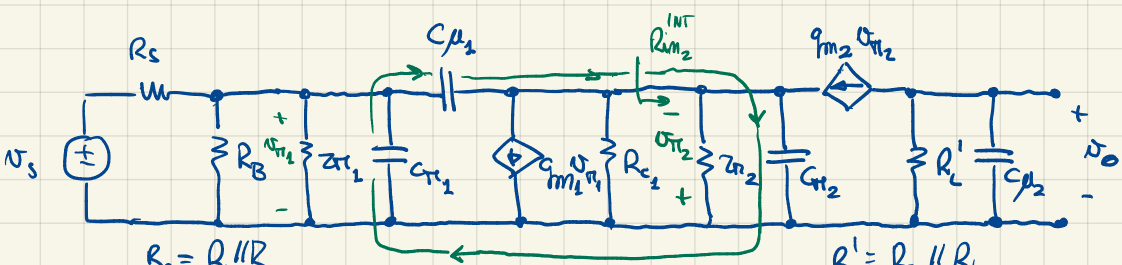
- BESIDES, v_0 GOES TO ZERO LIKE $\frac{1}{s}$ AS $s \to +\infty$ WHICH MEANS $m = 2$.
  NOTICE THE CAPACITIVE VOLTAGE DIVIDER MADE UP BY C_{π_1} AND C_{π_2}!

- THE TWO ZEROS CAN BE FOUND AS IN THE SINGLE STAGE CE AND CC AMPLIFIER.

◇ FREQUENCY RESPONSE OF THE CASCODE (CE-CB) AMPLIFIER



- Q_1 (I_{C_1}, V_{CE_1}), Q_2 (I_{C_2}, V_{CE_2})
- CHECK F.A.R. BIAS
- DETERMINE SMALL SIGNAL PARAMETERS $g_{m_{1,2}}$, $r_{\pi_{1,2}}$, $C_{\mu_{1,2}}$, $C_{\pi_{1,2}}$ ...
- DRAW THE SMALL SIGNAL EQUIVALENT CIRCUIT



$R_B = R_1 // R_2$ $\quad$ $R_L' = R_{C_2} // R_L$

$N = 4$ BUT WE HAVE A LOOP $\Rightarrow$ $m = 3$

$m = 1$ AS $v_0 \to 0$ AS FAST AS $1/s^2$ DUE TO C_{μ_2} AND C_{π_1} (INDEPENDENTLY).

$$R_{in_1}^{INT} = r_{\pi_1}$$

$$R_{out_1} = R_{C_1} // r_{o_1} \cong R_{C_1}$$

$$R_{in_2}^{INT} = r_{\pi_2} // \frac{1}{g_{m_2}} = \frac{r_{\pi_2}}{\beta_{o_2}+1} \cong \frac{1}{g_{m_2}}$$

$$R_{out_2} \cong R_L'$$

.1

$$A_v^{MB} = \underbrace{-g_{m_2} R_L'}_{\text{CB GAIN}} \cdot \underbrace{g_{m_1} \cdot (R_{C_1} // R_{in_2}^{INT})}_{\text{CE GAIN WITH LOADING EFFEC}} \cdot \underbrace{\frac{R_{in_1}^{INT} // R_B}{R_{in_1}^{INT} // R_B + R_S}}_{\alpha_i} \cong -\alpha_i \cdot g_{m_1} R_L'$$

$\frac{1}{g_{m2}}$

THE CASCODE STAGE HAS ALMOST THE SAME GAIN OF A CE STAGE WITH THE SAME LOAD.

. 2 SCTC METHOD AND DOMINANT POLE ASSUMPTION

$$\omega_L \cong \frac{1}{R_S^{sc} C_S} + \frac{1}{R_E^{sc} C_E} + \frac{1}{R_B^{sc} C_B} + \frac{1}{R_L^{sc} C_L}$$

$$R_S^{sc} = R_S + R_B // R_{in_1}^{INT}$$

$$R_E^{sc} = R_{E_1} // \frac{r_{\pi_1} + R_S // R_B}{\beta_{o_1}+1} \cong R_{E_1} // \frac{1}{g_{m_1}} \cong \frac{1}{g_{m_1}}$$

← MOST IMPORTANT FOR THE DESIGN OF ω_L

$$R_B^{sc} = R_B // \left[r_{\pi_2} + (\beta_{o_2}+1) R_{out_1}\right]$$

$$R_L^{sc} = R_{C_2} + R_L$$

.3 OCTC METHOD AND DOMINANT POLE ASSUMPTION

$$\omega_H \cong \frac{1}{\underbrace{R_{\pi_1}^{o} C_{\pi_1} + R_{\mu_1}^{o} C_{\mu_1}}_{\frac{1}{a_{11}}} + \underbrace{R_{\pi_2}^{o} C_{\pi_2} + R_{\mu_2}^{o} C_{\mu_2}}_{\frac{1}{a_{12}}}} = \frac{1}{\frac{1}{\omega_{H_1}} + \frac{1}{\omega_{H_2}}}$$

THE LOADING EFFECT STRONGLY REDUCES THE MILLER'S MULTIPLIER

$$R_{\pi_1}^{o} = r_{\pi_1} // R_B // R_S$$

$$R_{\mu_1}^{o} = R_{\pi_1}^{o} + R_{C_1} // R_{in_2}^{INT}\left(1 + g_{m_1} R_{\pi_1}^{o}\right) = R_{C_1} // R_{in_2}^{INT} + R_{\pi_1}^{o}\left(1 + g_{m_1} R_{C_1} // R_{in_2}^{INT}\right)$$

$$R_{\pi_2}^{o} = R_{out_1} // R_{in_2}^{INT} \cong R_{in_2}^{INT}$$

$$R_{\mu_2}^{o} = R_L'$$

CB IS NOT AFFECTED (ALMOST) BY THE CE STAGE

CONCLUSION: BY REDUCING THE MILLER EFFECT ON THE CE STAGE THE CASCODE AMPLIFIER HAS A **MUCH WIDER BANDWIDTH** THAN A SIMPLE CE OF EQUAL GAIN.

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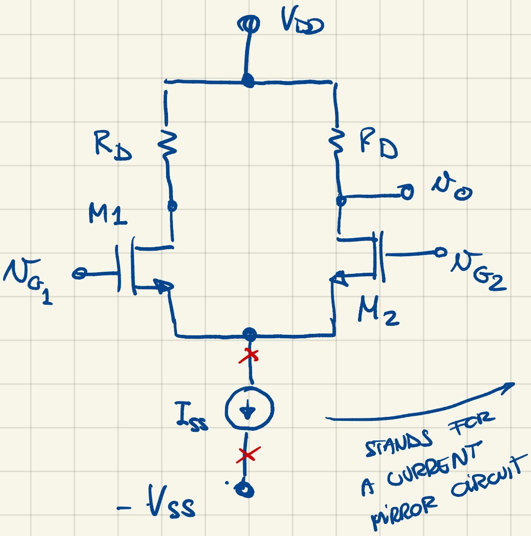
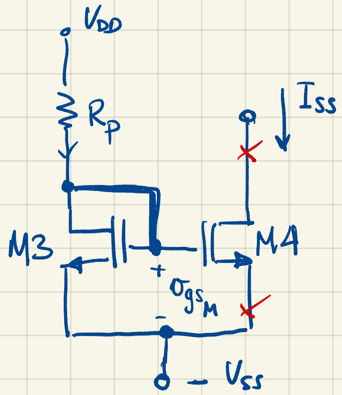
FREQUENCY RESPONSE OF A MOSFET BASED DIFFERENTIAL STAGE





THE CURRENT MIRROR DEVICES ARE MUCH LARGER (TYPICALLY) THAN THE AMPLIFIER'S

THE PURPOSE OF THE CURRENT MIRROR IS TO BIAS THE DIFFERENTIAL STAGE SO THAT WHEN $V_{G_1} = V_{G_2} = 0$ THEN

$$I_{D_1} = I_{D_2} = \frac{I_{SS}}{2}$$

Hyp: THE CIRCUIT IS PERFECTLY SYMMETRICAL

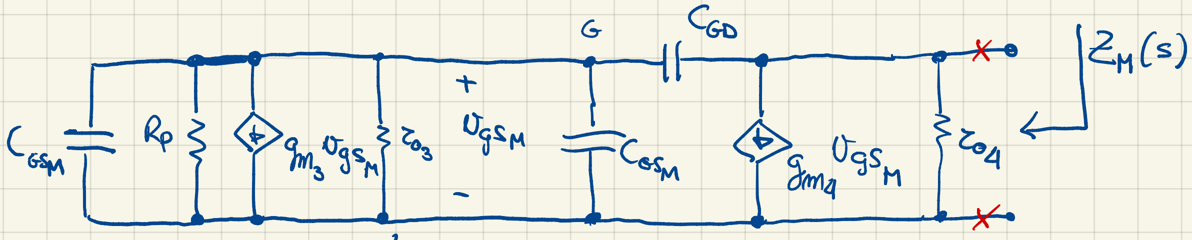
SMALL SIGNAL ANALYSIS IS **TWOFOLD**

- **DM ANALYSIS** $\quad v_{g_1} = \frac{v_{dm}}{2} \qquad v_{g_2} = -\frac{v_{dm}}{2}$
- **CM ANALYSIS** $\quad v_{g_1} = v_{g_2} = v_{cm}$

WHERE

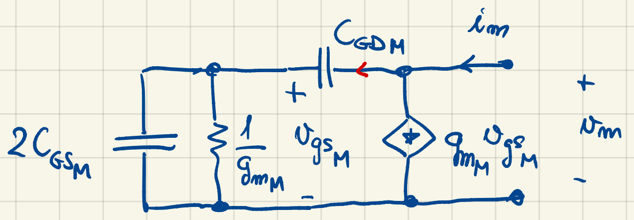
$$\begin{cases} v_{dm} = v_{g_1} - v_{g_2} \\ v_{cm} = \dfrac{v_{g_1} + v_{g_2}}{2} \end{cases} \iff \begin{cases} v_{g_1} = v_{cm} + \dfrac{v_{dm}}{2} \\ v_{g_2} = v_{cm} - \dfrac{v_{dm}}{2} \end{cases}$$

LET'S ANALYZE THE CURRENT MIRROR FIRST, ASSUMING $g_{m_3} = g_{m_4} = g_{m_M}$



$$R_p \,//\, r_{o_3} \,//\, \frac{1}{g_{m_M}} \simeq \frac{1}{g_{m_M}}$$

$$Z_M(s) = r_{o4} \,//\, Z_y(s)$$



$$C_{GD_M} \gg C_{GD}$$

$$C_{GS_M} \gg C_{GS}$$

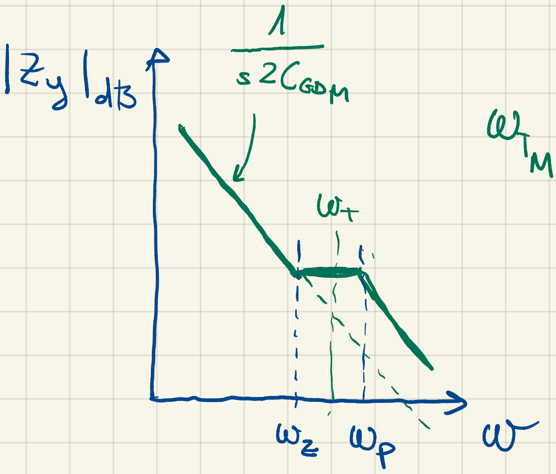
$$v_{gs_M} = v_m \cdot \frac{\dfrac{1}{g_{m_M} + 2sC_{GS_M}}}{\dfrac{1}{g_{m_M} + 2sC_{GS_M}} + \dfrac{1}{sC_{GD_M}}} = v_m \cdot \frac{sC_{GD_M}}{g_{m_M} + 2sC_{GS_M} + sC_{GD_M}}$$

$$i_m = g_{m_M} \cdot v_{gs_M} + \left(v_m - v_{gs_M}\right) sC_{GD_M} =$$

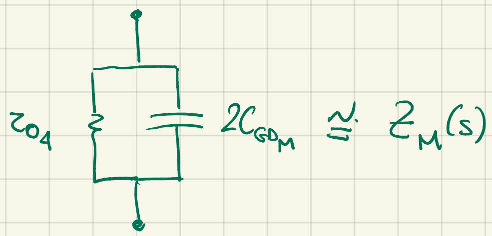
$$= v_m \cdot \left[\frac{\left(g_{m_M} - sC_{GD_M}\right) sC_{GD_M}}{g_{m_M} + s\left(C_{GD_M} + 2C_{GS_M}\right)} + sC_{GD_M} \right] =$$

$$= v_m \cdot \frac{sC_{GD_M}\left(g_{m_M} - \cancel{sC_{GD_M}} + g_{m_M} + \cancel{sC_{GD_M}} + 2sC_{GS_M}\right)}{g_{m_M} + s\left(C_{GD_M} + 2C_{GS_M}\right)}$$

$$\frac{v_m}{i_m} \triangleq Z_y(s) = \frac{1 + s\dfrac{C_{GD_M} + 2C_{GS_M}}{g_{m_M}}}{2sC_{GD_M}\left(1 + \dfrac{sC_{GS_M}}{g_{m_M}}\right)} = \frac{1 + \dfrac{s}{\omega_z}}{s2C_{GD_M} \cdot \left(1 + \dfrac{s}{\omega_p}\right)}$$



$$\omega_{T_M} = \frac{g_{m_M}}{C_{GS_M} + C_{GD_M}}$$



$$Z_M(s) \cong r_{o4} \,//\, \frac{1}{s2C_{GD_M}} = \frac{r_{o4}}{1 + s2r_{o4}C_{GD_M}}$$

BECAUSE $\omega_z < \omega_T < \omega_p$, ω_z IS CLOSE TO ω_T AND SO IS ω_p, WE CAN NEGLECT THE ZERO-POLE COUPLE AND APPROXIMATE Z_M AS IN HERE