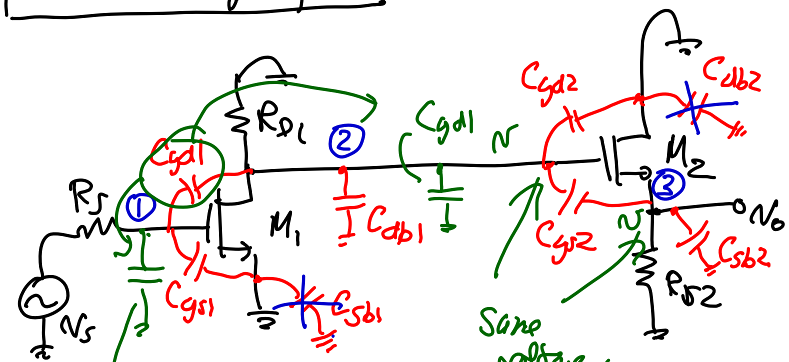
EE 140/240a: Analog Integrated Circuits
Lecture 7w: Active Loads I

CTN 9/17/15

Lecture 7: Active Loads I

- Announcements:
- Lab 1 next week – report to your lab section
- HW#3 online; due next Wednesday at 8 a.m.
- Passed out some extra computer account sheets
  - Come to my office if you still don't have one
- Lecture Topics:
  - Short Ckt Time Constant (SCTC) Analysis
  - Example Low Freq. Response Determination
  - Active Loads
    - —Why active loads?
    - —Examples of actively loaded amplifiers
- -----------------------------------
- Last Time:
- OCTC analysis to get dominant high frequency pole
- Ended with two-stage amplifier example

Ex. MOS Two-Stage Amplifier



Same voltage (w/ unity gain)

$C_M = C_{gd1}(1+g_{m1}R_{D1})$

$\tau_{①} = \left[C_{gd1}(1+g_{m1}R_{D1}) + C_{gs1}\right] R_S$

$\tau_{②} = \left[C_{db1} + C_{gd1} + C_{gd2}\right](\cancel{r_{o1}} \| R_{D1})$

$\tau_{③} = C_{sb2}\left(R_{S2} \| \frac{1}{g_{m2}+g_{mb2}}\right)$

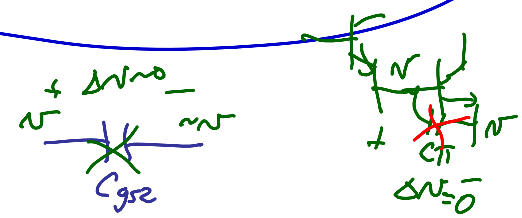
$\cancel{\tau_{gs2} = C_{gs2}\left(\frac{R_{D1}+R_{S2}}{1+(g_{m2}+g_{mb2})R_{S2}}\right)}$

C_{gs2}

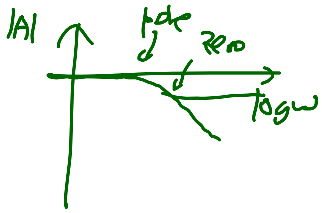
$+ \Delta v = 0V -$

No voltage drop across C_{gs2} ∴ it's effectively not there!

$$\boxed{\omega_H = \frac{1}{\tau_{①} + \tau_{②} + \tau_{③} + \cancel{\tau_{gs2}}}}$$

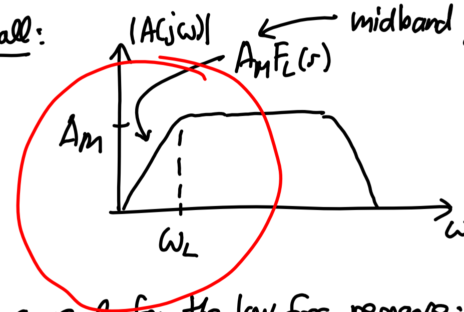


For a bipolar follower:



Low Freq. Amplifier Response Using Short Circuit Time Constant Analysis (SCTC)

Recall:



In general, for the low freq. response:

$$F_L(s) = \frac{s^{n_L} + d_1 s^{(n_L-1)} + \cdots}{s^{n_L} + e_1 s^{(n_L-1)} + \cdots}, \quad n_L = \#\text{poles} = \#\text{zeros}$$

We can express the coefficient e_1 by:

$$e_1 = \omega_{p1} + \omega_{p2} + \cdots + \omega_{pn_L}$$

Similar analysis to that used for OCTC...

For the case of a dominant pole:
↳ i.e., the highest freq. pole

$$F_L(s) \cong \frac{s}{s+\omega_L} = \frac{s}{s+e_1} \rightarrow e_1 \cong \omega_{p1} = \omega_L$$

$$\therefore \boxed{\omega_L \cong e_1 = \sum_j \omega_{pj} = \sum_j \frac{1}{C_j R_{js}} = \sum_j \frac{1}{\tau_{js}}}$$

where $C_j \triangleq$ various large (>10nF) capacitors in the ckt. (e.g., the bypass caps.)

$R_{js} \triangleq$ driving point resistance seen between the terminals of C_j determined with:

For reading, can go to Sedra & Smith

1. all large capacitors short-circuited, except C_j, which is replaced by the test voltage source for R determination

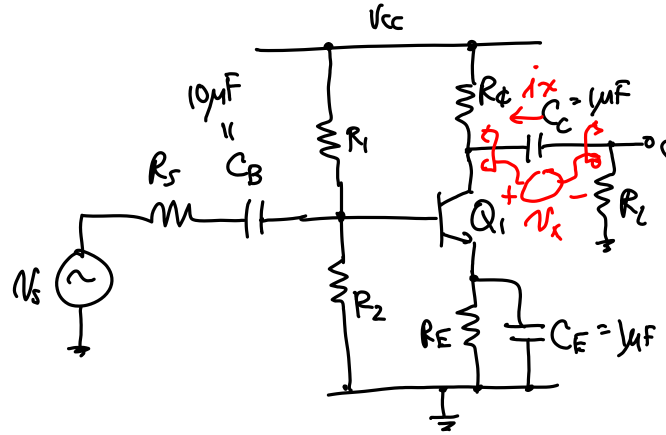
② all independent sources eliminated (i.e., short voltage sources, open current sources)

③ open all H.F. capacitors (i.e., small caps in the pF range, or < 1nF)

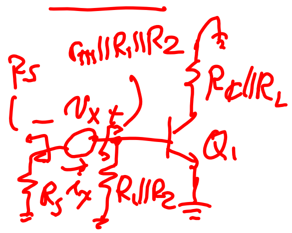
Again, for the case where there are no dominant poles, a reasonable approximation is:

$$\omega_L \cong \sqrt{\omega_{P1}^2 + \omega_{P2}^2 - 2\omega_{Z1}^2 - 2\omega_{Z2}^2}$$

Ex: Determine the L.F. response of the C.E. Amplifier:



(a) τ due C_B: short ckt. C_C & C_E



$R_{BS} = \frac{v_x}{i_x}$ = resistance in series w/ v_x

$R_{BS} = R_s + r_\pi || R_1 || R_2$

$\tau_B = R_{BS} C_B = (R_s + r_\pi || R_1 || R_2) C_B$

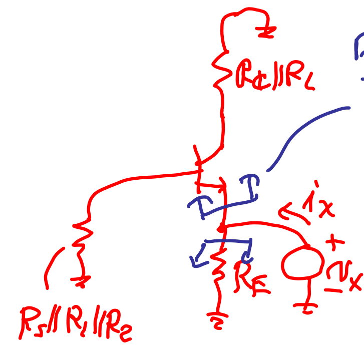
$$\omega_{pB} = \frac{1}{\tau_B} = \frac{1}{(R_s + r_\pi || R_1 || R_2) C_B}$$

(b) τ due to C_C: short ckt. C_B & C_E

⇒ again: R_{CS} = resistance in series w C_C seen on both sides

$$\tau_C = (R_L + \underbrace{R_C || r_o}_{\sim R_C}) C_C \rightarrow \omega_{pC} = \frac{1}{\tau_C} = \frac{1}{(R_C + R_L) C_C}$$

(c) τ due to C_E: short C_B & C_C



$$\frac{r_\pi + R_s || R_1 || R_2}{\beta + 1}$$

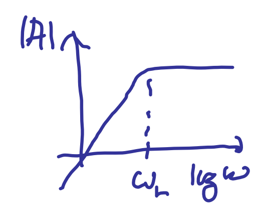
$$R_{ES} = R_E || \frac{r_\pi + R_s || R_1 || R_2}{\beta + 1}$$

$\tau_E = R_{ES} C_E$

$$\omega_{pE} = \frac{1}{\tau_E} = \frac{1}{R_{ES} C_E}$$

and finally:

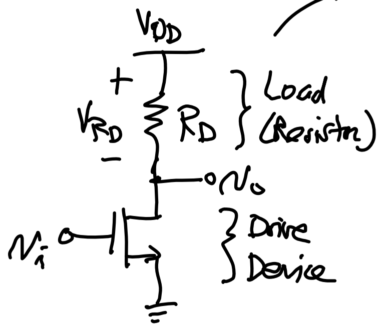
$$\boxed{\omega_L = \omega_{pB} + \omega_{pC} + \omega_{pE}}$$



Active Loads

⇒ why use them?

Gain: $\frac{v_o}{v_i} = -g_m R_D$



For $\frac{v_o}{v_i}\uparrow$, must:

① Raise g_m → raise I_D

Problem: $V_{RD} = I_D R_D \uparrow$

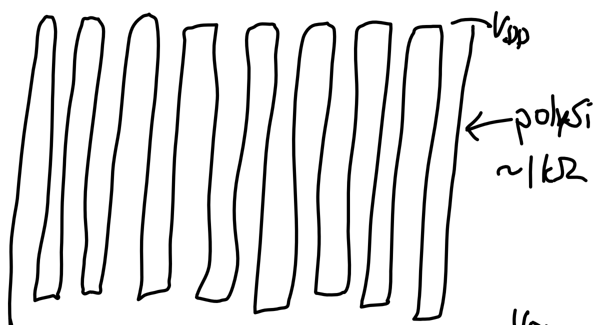
Limited by supply V_{DD}

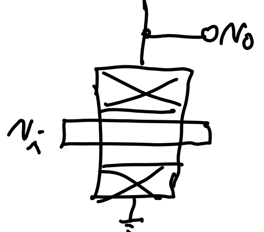
② raise R_D → again, $V_{RD}\uparrow$ → Limited by supply V_{DD}

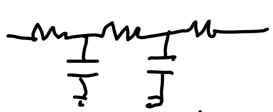
another problem: area consumption by R_D

high cost!

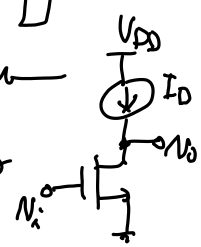
Layout:





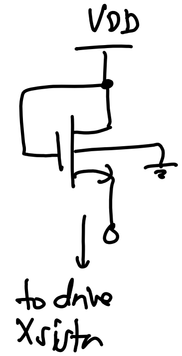


Better Solution:

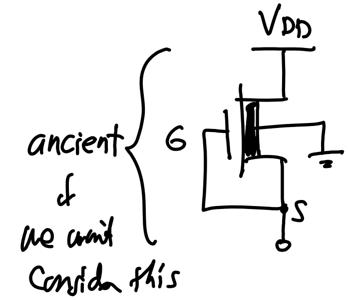


Types of Active Loads → want a current source

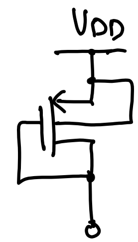
Diode-Connected Enhancement Load:



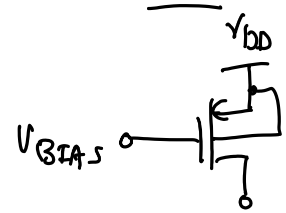
Depletion Load!



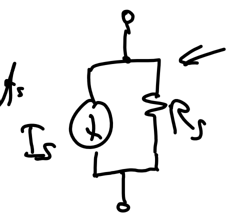
Diode-Connected PMOS Load:



PMOS Current Source Load:

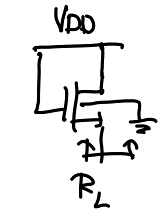


Norton Equivalent:

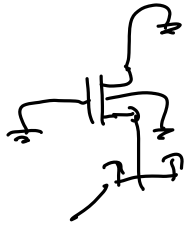


⇐ ideal when $R_s = \infty$

Diode-Connected Enhancement Load



s.s. Ckt. $\longrightarrow$ (do by inspection)

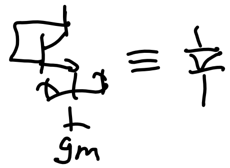


$R_L = \frac{1}{g_m} \| \frac{1}{g_{mb}} = \frac{1}{g_m + g_{mb}}$

$\sim 1\text{–}10\,k\Omega$

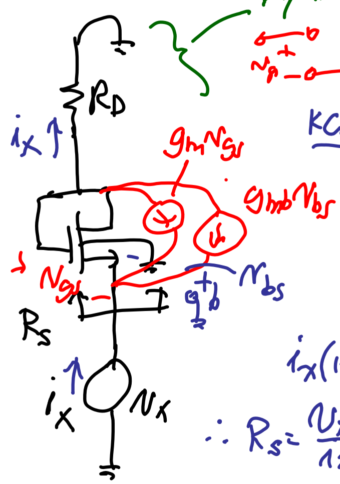
↓

"Diode-Connected"

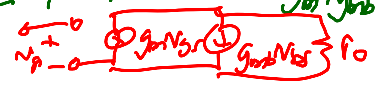


$\frac{1}{g_m + g_{mb}} \approx \frac{1}{g_m}$

How about this?



by inspection: $R_S \approx \frac{1}{g_m + g_{mb}} + R_D$



KCL: $i_x = -g_m v_{gs} - g_{mb} v_{bs}$

$= -g_m(-(v_x - i_x R_D)) - g_{mb}(-v_x)$

$= g_m v_x - g_m R_D i_x + g_{mb} v_x$

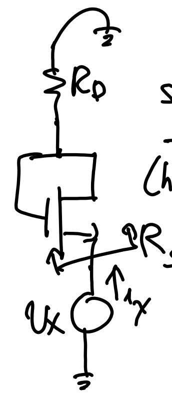
$i_x(1 + g_m R_D) = (g_m + g_{mb}) v_x$

$\therefore R_S = \frac{v_x}{i_x} = \frac{1 + g_m R_D}{g_m + g_{mb}} = \boxed{\frac{\frac{1}{g_m} + R_D}{1+\eta} = R_S}$

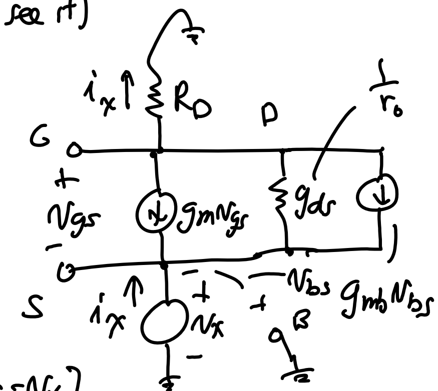
$\boxed{R_S = \frac{1}{g_m + g_{mb}} + \frac{R_D}{1+\eta}}$

Full hybrid-π analysis:

(in case you want to see it)



s.s.ckt. $\longrightarrow$ (hybrid-π)



$[v_{bs} = -v_x]$

$i_x = -g_m v_{gs} + \underbrace{g_{ds}(-v_{gs})}_{\text{neg.}} - g_{mb} v_{bs}$

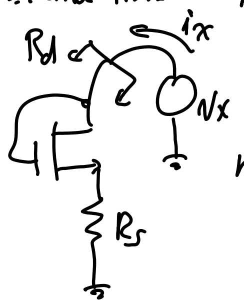
$= g_m(v_x - i_x R_D) + g_{mb} v_x$

$R_S = \frac{v_x}{i_x} = \frac{1 + g_m R_D}{g_m + g_{mb}} = \frac{1}{g_m + g_{mb}} + \frac{R_D}{1+\eta}$

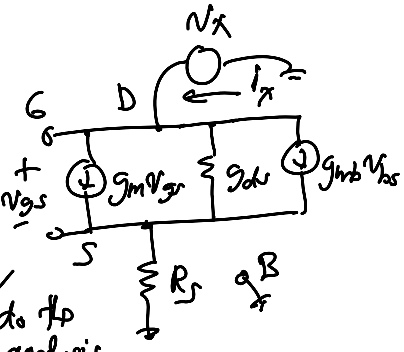
$\approx \frac{1}{g_m + g_{mb}} + R_D$

$$\boxed{R_S \cong \frac{\frac{1}{g_m} + R_D}{1+\eta}}$$

... and from the top:

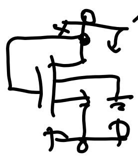


S.S. Ckt. ⟹ hybrid-π



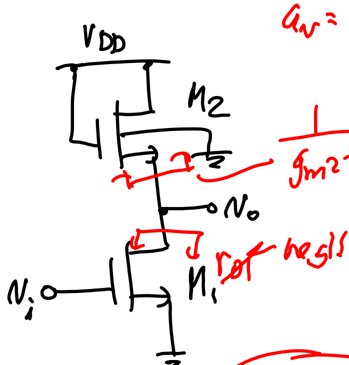
do the analysis

$$R_d = \frac{v_x}{i_x} = \frac{1}{g_m} + (1+\eta)R_s \approx \frac{1}{g_m} + R_s$$



$\cong \quad \frac{1}{g_m} \qquad \frac{1}{g_m + g_{mb}} \approx \frac{1}{g_m}$

Apply to a C.S. Ckt.



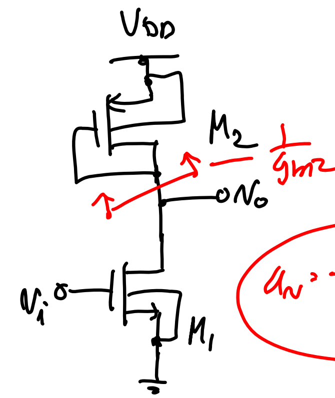
$\frac{1}{g_{m2}+g_{mb2}}$

$$a_v = \frac{v_o}{v_i} = -\frac{g_{m1}}{g_{m2}+g_{mb2}} = -\frac{1}{(1+\eta)}\frac{g_{m1}}{g_{m2}}$$

$$a_v = -\frac{1}{1+\eta}\frac{\sqrt{2\mu_n C_{ox}(W/L)_1 I_D}}{\sqrt{2\mu_n C_{ox}(W/L)_2 I_D}}$$

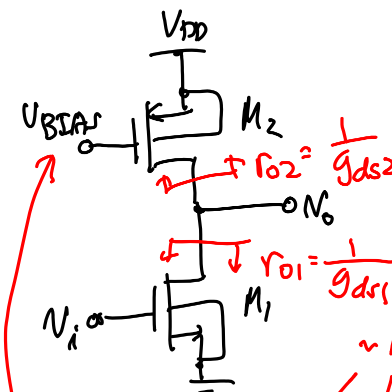
$$a_v = -\frac{1}{1+\eta}\sqrt{\frac{(W/L)_1}{(W/L)_2}}$$

Diode-Connected PMOS Load



$$a_v = -\frac{g_{m1}}{g_{m2}} = -\sqrt{\frac{\mu_n (W/L)_1}{\mu_p (W/L)_2}}$$

PMOS Current Source Load

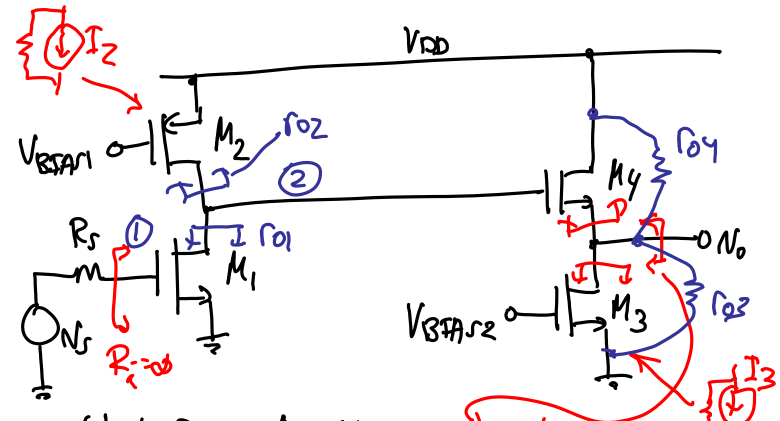


~MΩ's

$$a_v = \frac{v_o}{v_i} = -g_{m1}(r_{o1} \| r_{o2}) = -\frac{g_{m1}}{g_{ds1}+g_{ds2}}$$

Gain is huge! → but... requires V_{BIAS}

Ex. Multi-Stage Actively-Loaded MOS Ckt.



Want R_i, R_o, A_v, ω_H.

neg!

$$R_o = \cancel{r_{o3}} \,\|\, \frac{1}{g_{m4}+g_{mb4}}$$

$$a_v = \frac{v_1}{v_s}\cdot\frac{v_2}{v_1}\cdot\frac{v_o}{v_2}$$

$$= (1)\left(-g_{m1}(r_{o1}\|r_{o2})\right)\cdot\frac{g_{m4}(r_{o3}\cancel{\|r_{o4}})}{1+(g_{m4}+g_{mb4})(\cancel{r_{o3}\|r_{o4}})}$$

$$a_v = -g_{m1}(r_{o1}\|r_{o2})\left(\frac{g_{m4}}{g_{m4}+g_{mb4}}\right)$$

Aside:

$$\frac{v_o}{v_i} = \frac{g_m R_E}{1+g_m R_E}$$

Now, tackle freq. response: