EE 240B – Fall 2019

Advanced Analog Integrated Circuits

Lecture 8: Operational Transconductance Amplifiers (II)

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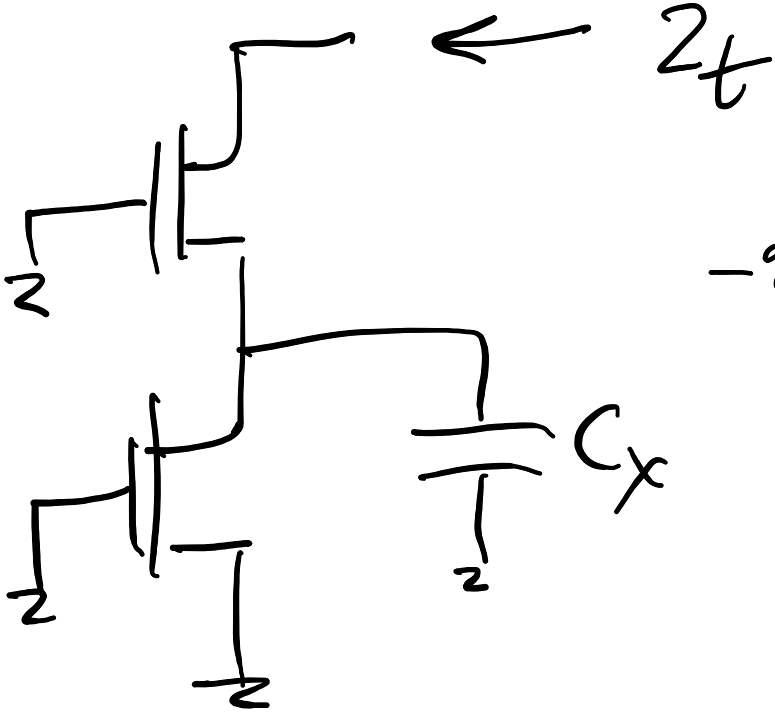
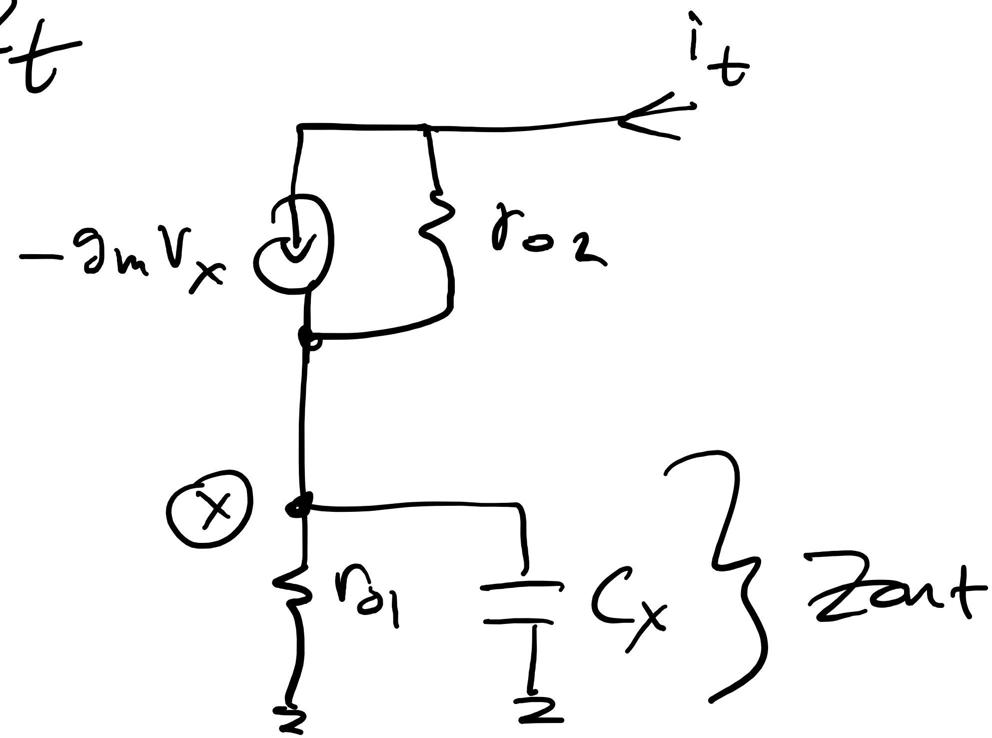
Closer Look at Cascode Dynamics





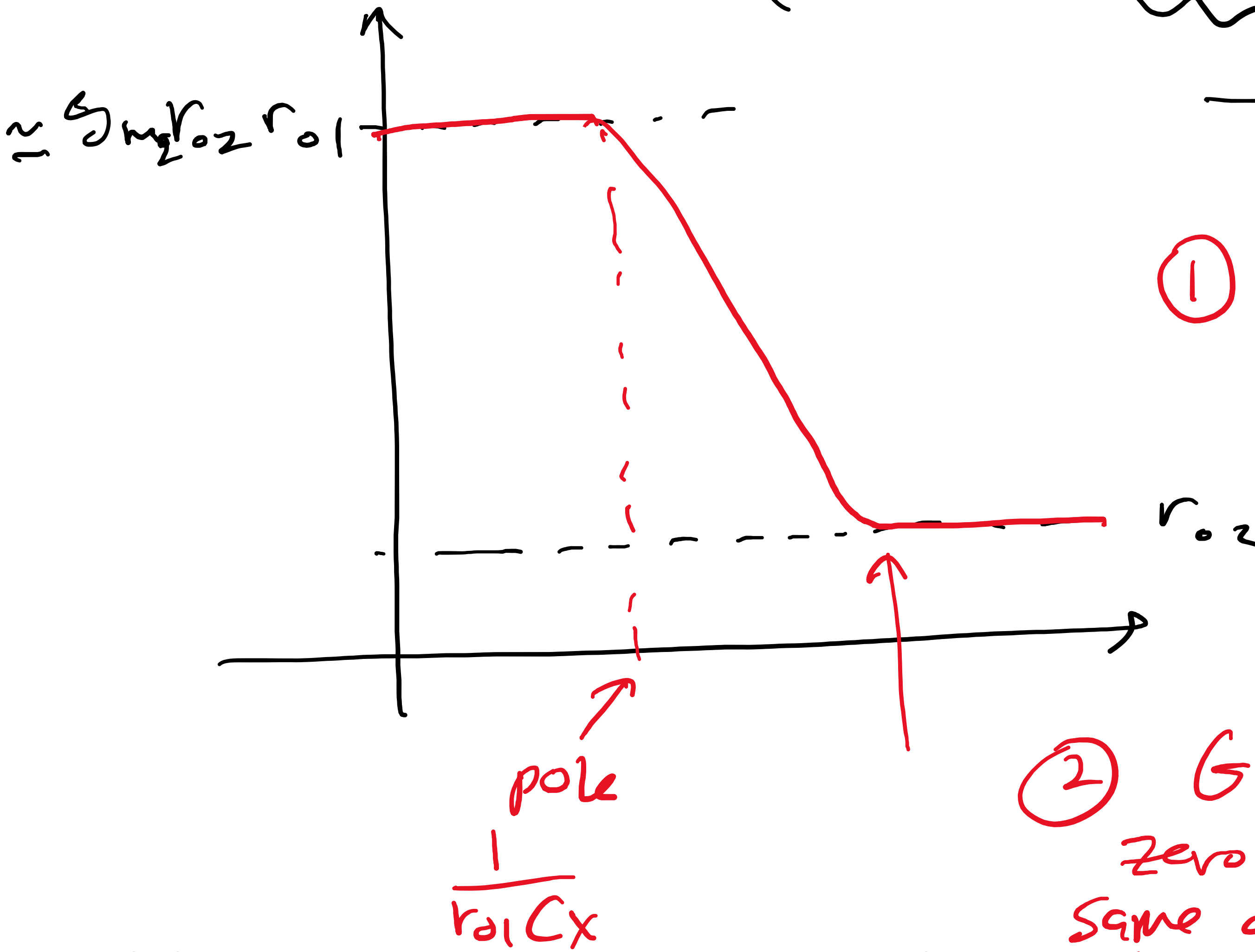
$$V_t = i_t \cdot Z_{out} + (i_t + g_m V_x)\, r_{o2}$$

$$i_t Z_{out}$$

$$V_t = i_t \left(Z_{out} + r_{o2} + g_m Z_{out} \cdot r_{o2} \right)$$

Cascode Z_{out} vs. f

$$Z_t = r_{o2} + (g_{m2} r_{o2} + 1)\,\underbrace{Z_{out}}_{\frac{r_{o1}}{1 + s\, r_{o1} C_x}}$$

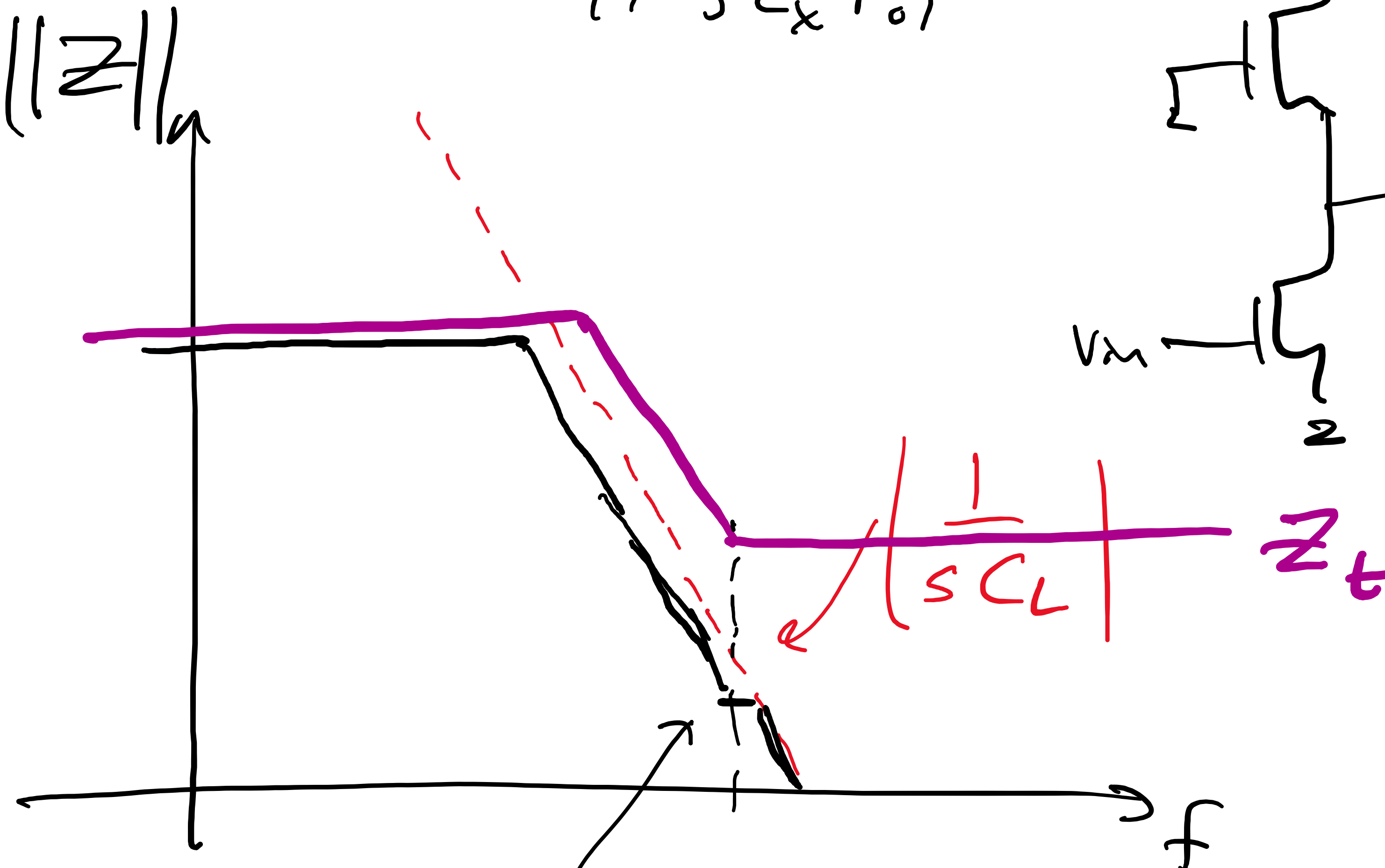


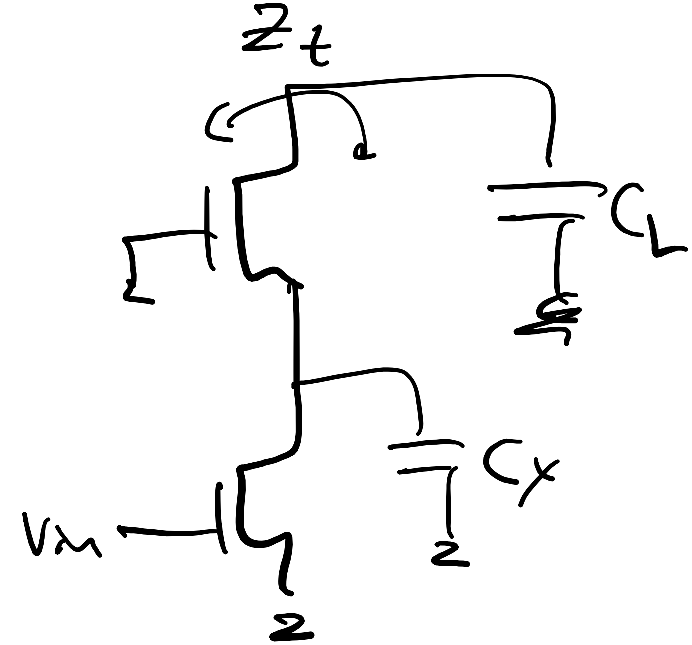
① $k + H_{LPF}$: has zero, & a pole at that of H_{LPF}

② $G = H_1 \| H_2$
zeros of G same as zeros of H_1 and H_2

Load + Cascode Z_{out} vs. f

$$Z_t = r_{o2} + \frac{r_{o1}(1 + g_m r_{o2})}{1 + s C_x r_{o1}}$$





Property

Pole-Zero Doublets

Discussion

- **Doublet generally not important in “simple” cascode since it shows up at high frequencies**

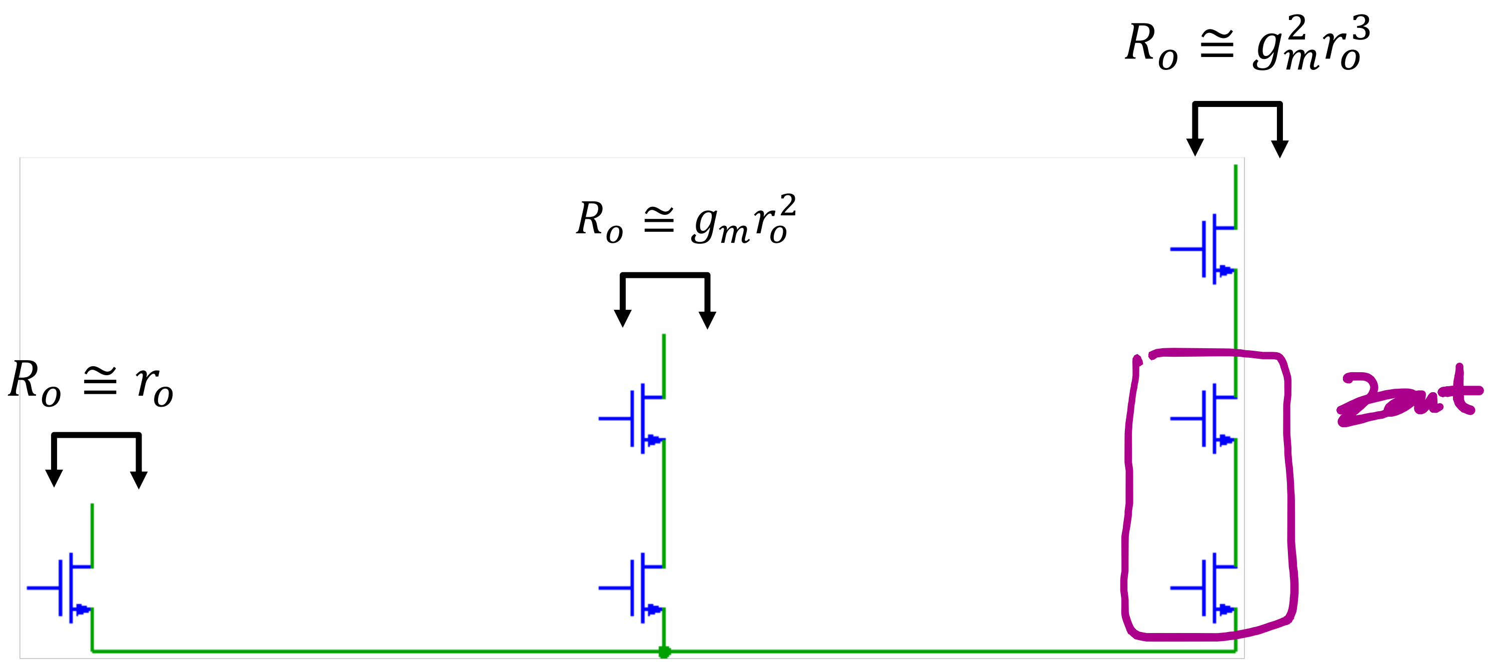
- **But, doublets can show up in similar circuits**
  - In particular, when you try and increase the gain beyond what a simple cascode can support

Openloop Gain : Price is swing!

$$a_{vo} = g_m R_o$$

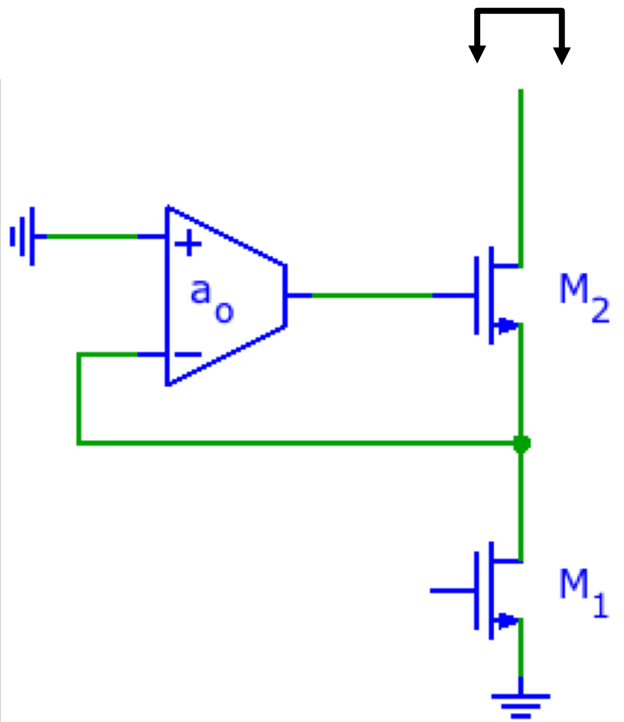
costs power
→ set by dynamics

"boost" R_o to get
sufficient gain



Gain Boosting

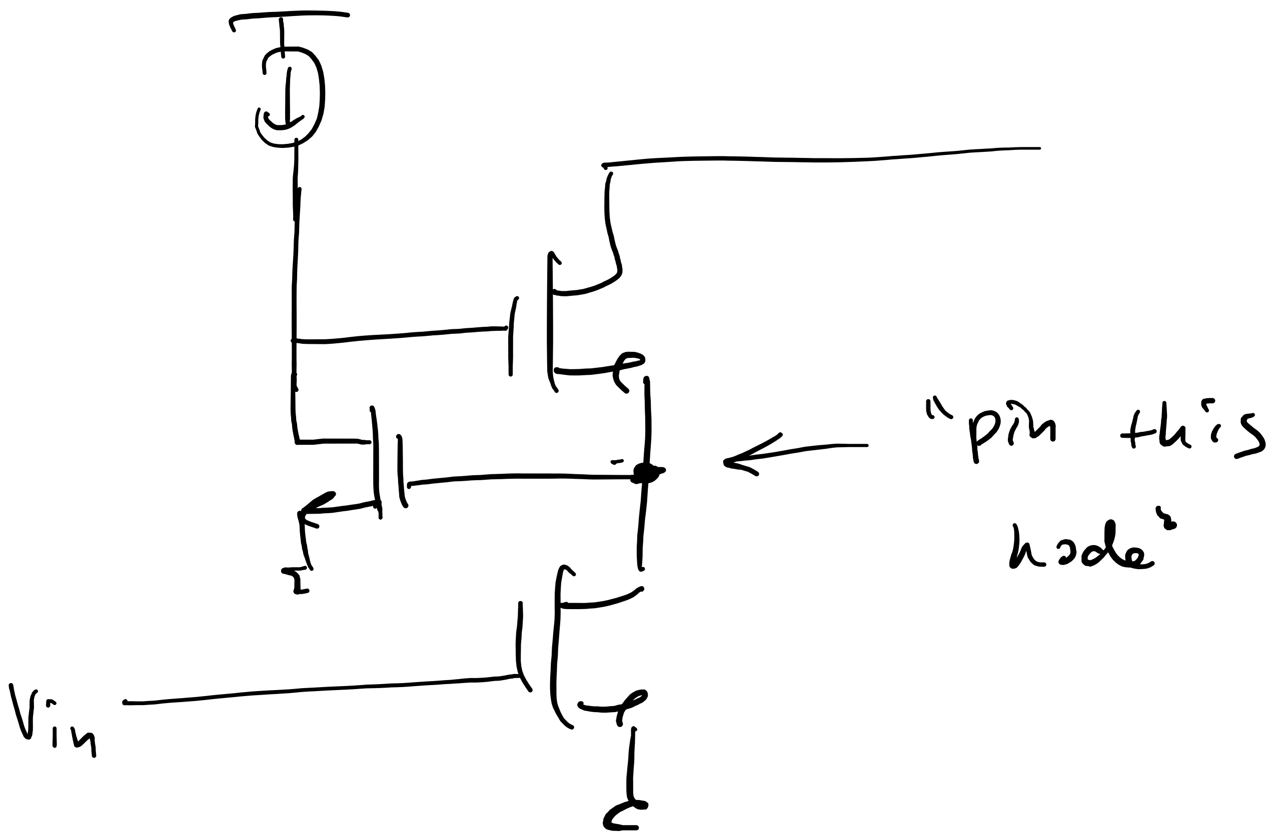
$R_o \cong r_{o1} g_{m2} r_{o2} a_o$



- **Use feedback to further increase R_{out}**
  - No increase of V_{min} (unlike double cascode)

- **References:**
  - B. J. Hosticka, “Improvement of the gain of MOS amplifiers,” JSSC, Dec. 1979 , pp. 1111-4.
  - E. Sackinger and W. Guggenbuhl, “A high-swing high-impedance MOS cascode circuit”, JSSC, Feb. 1990, pp. 289-298.
  - K. Bult, G. Geelen, “A fast-settling CMOS op-amp for SC circuits with 90-dB DC gain,” JSSC, Dec. 1990 , pp. 1379-84.

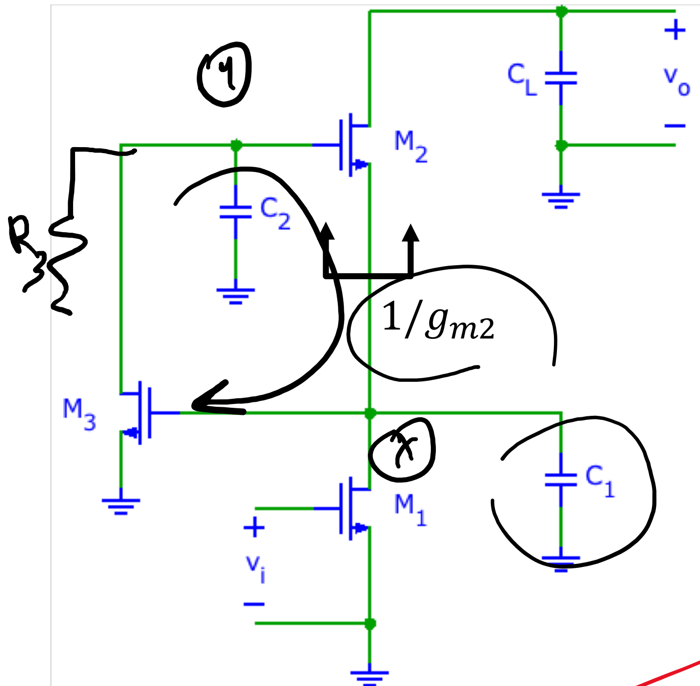
$R_o = (G_m r_{o2})\, r_{o1} = a_o g_{m2} \cdot r_{o1} \cdot r_{o2}$


"pin this node"
Vin

High Frequency Analysis

- Loop gain

$|T| = 1 \text{ @ } \omega \approx \frac{g_m}{C_2}$



$$T(s) = \frac{g_{m3}}{sC_2} \frac{1}{1 + s\frac{C_1}{g_{m2}}} = \frac{\omega_{u3}}{s} \frac{1}{1 + s\frac{C_1}{g_{m2}}}$$

- For acceptable phase margin

$$\frac{g_{m2}}{C_1} = \omega_{p2} > k\omega_{u3}$$

x $\quad p_1 \sim \frac{g_{m2}}{C_1}$

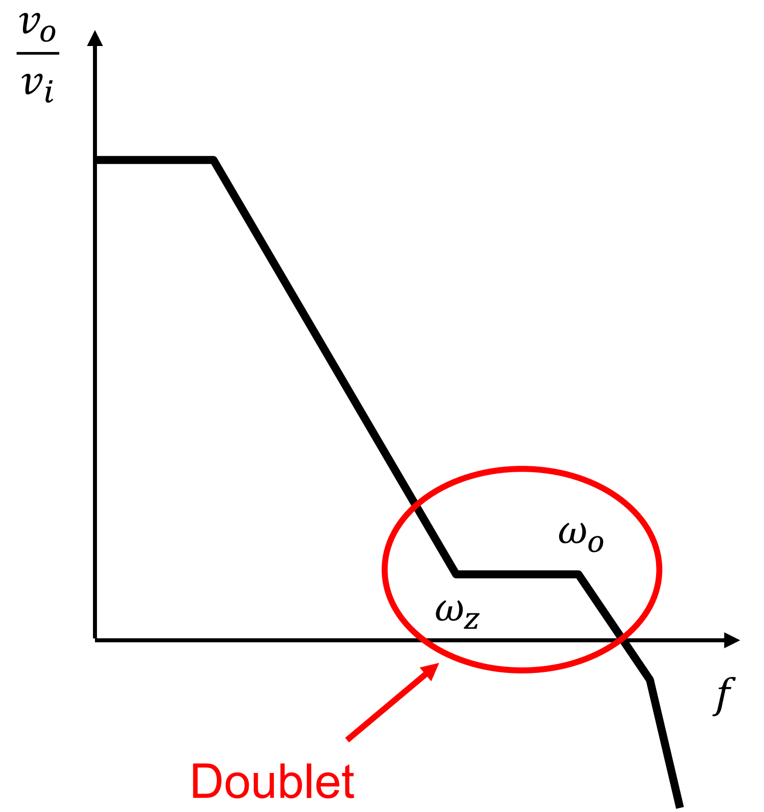
y $\quad p_2 = \frac{1}{C_2 R_3}$ } dominant pole

Overall Amplifier Response

$$\frac{v_o}{v_i} = \frac{g_{m1}}{sC_L} \frac{1 + \frac{s}{\omega_z}}{1 + \frac{s}{\omega_o Q} + \frac{s^2}{\omega_o^2}}$$

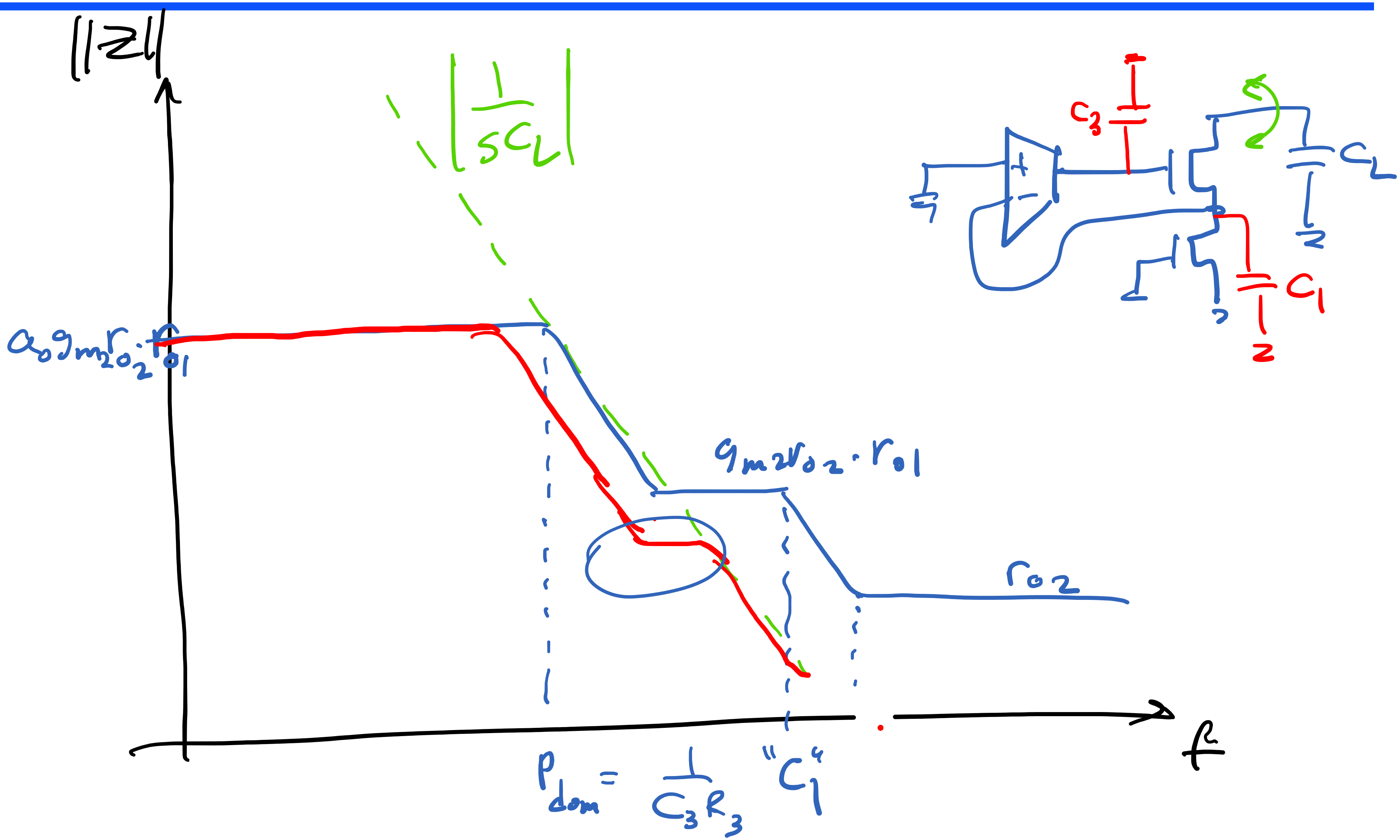
$$\omega_z = \omega_{u3} \qquad \omega_o = \sqrt{r}\,\omega_{u3} \qquad Q = \frac{1}{\sqrt{r}}$$

$$\text{for} \quad Q < 0.5 \quad \Longrightarrow \quad r > 4$$



Ref: M. Das, "Improved design criteria of gain boosted CMOS OTA with high-speed optimizations," IEEE CAS II, March 2002, pp. 204-7.

Graphical Derivation



Pole-Zero Doublets

- Often ”closely spaced”
  - May miss in Bode plot!
- Origins
  - Feedforward (e.g. Miller capacitance, C_{gd}, …)
  - Frequency dependent degeneration (series feedback, e.g. cascode)
- Concerns
  - Ringing (if high Q)
  - Slow settling

Ref: Y. Kamath, R. G. Meyer, and P. R. Gray, “Relationship between frequency response and settling time of operational amplifiers,” *IEEE J. Solid-State Circuits*, pp. 347–352, Dec. 1974.

Doublet Settling

- Amplifier model: replace G_{mo} with

$$G_m(s) = G_{mo}\frac{1+\dfrac{s}{\omega_z}}{1+\dfrac{s}{\omega_p}} \qquad \text{with} \qquad \begin{aligned} &\omega_p = \beta\omega_{-3dB}, \quad \omega_{-3dB} \text{ is bandwidth of } T(s) \\ &\omega_z = \frac{\omega_p}{\alpha} \\ &\alpha = 1+\varepsilon \quad \text{with} \quad |\varepsilon| << 1 \end{aligned}$$

- Closed-loop response

$$\frac{V_o}{V_{in}} = -c\frac{1}{1+s\dfrac{C_{Leff}}{FG_m(s)}} \cong -\frac{c}{1+\dfrac{s}{\omega_{-3dB}}}\left(\frac{1+\dfrac{s}{\omega_z}}{1+\dfrac{s}{\omega_{pp}}}\right) \qquad \text{with} \qquad \begin{aligned} &\omega_{-3dB} = \frac{FG_{mo}}{C_{Leff}} \\ &\omega_{pp} \cong \omega_p \end{aligned}$$

<u>Notation from</u> Y. Kamath, R. G. Meyer, and P. R. Gray, “Relationship between frequency response and settling time of operational amplifiers,” *IEEE J. Solid-State Circuits*, pp. 347–352, Dec. 1974.

Doublet Step Response

$$v_{o,step}(t) = -cV_{step}\left(1 + Ae^{-t\omega_{-3dB}} + Be^{-t\omega_{pp}}\right)$$

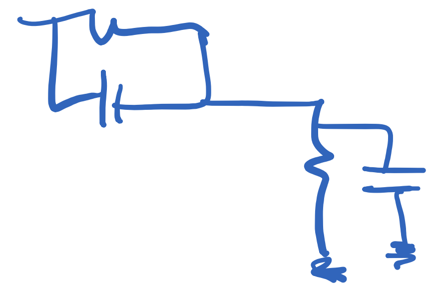
with

$$A \cong -1$$

$$B \cong \varepsilon \frac{\beta}{1-\beta^2}$$

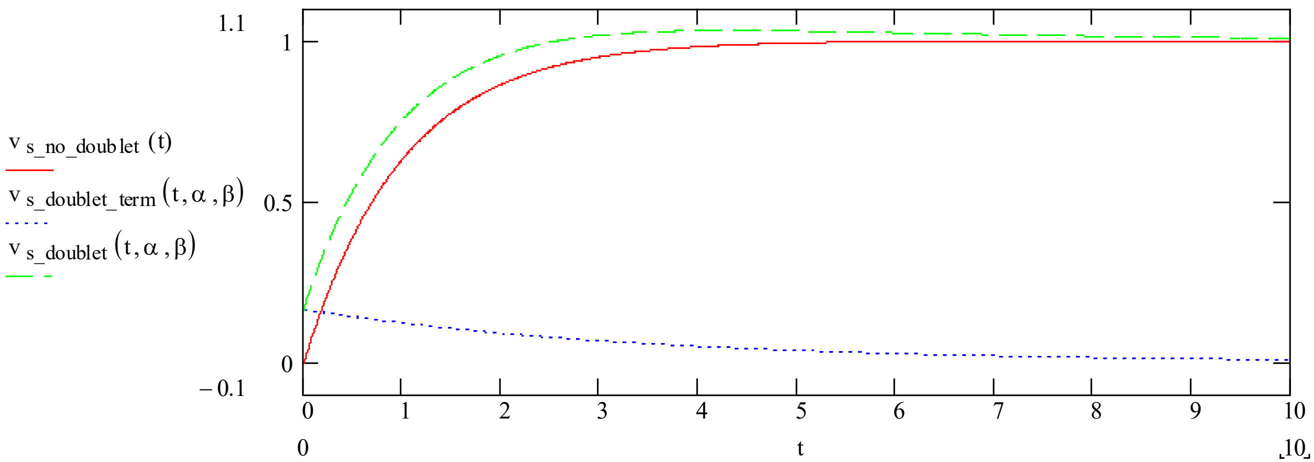
doublet

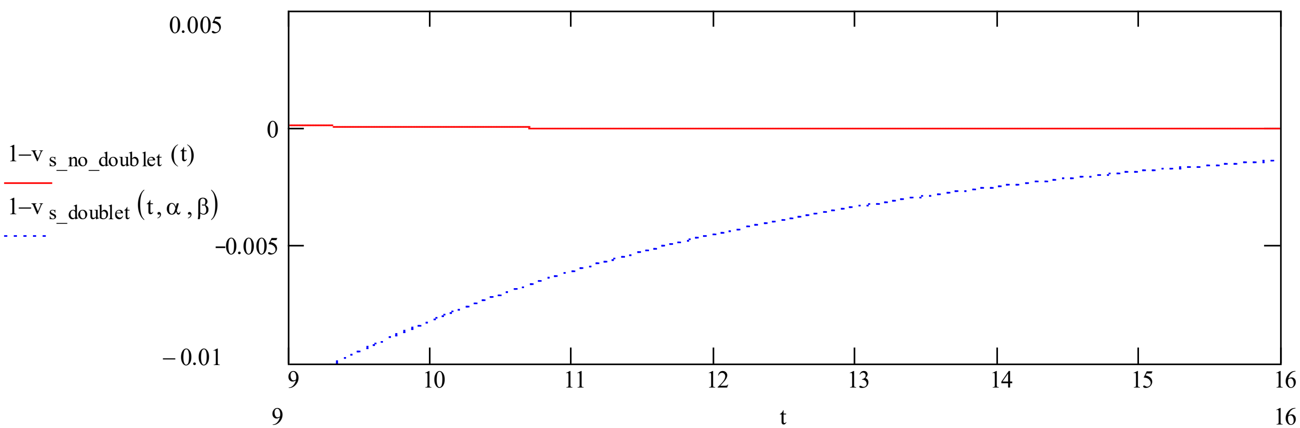
main amplifier



- Case A: $\omega_{pp} > \omega_{-3dB}$
  - Doublet settles faster than main amplifier
  - Has no impact on overall settling time (unless *B*>>*A*)
- Case B: $\omega_{pp} < \omega_{-3dB}$
  - Doublet settles more slowly than main amplifier
  - Determines settling time unless *B*<<*A*
  - Acceptable in “low precision” applications (e.g. oscilloscope probes)
  - Avoid when precise settling is required

Doublet Example





α = 1.5
β = 0.3

Gain Boosting – Doublets

zero in Z ⇒ leads to doublet

- Doublet is near ω_{u3}

$$\omega_{u3} > \beta\left(\frac{g_{m1}}{C_{Ltot}}\right) \text{ for fast settling}$$

amplifier gain/bandwidth

- Stability:

$$\omega_{u3} < \frac{g_{m2}}{C_1} \quad \text{(large } r\text{)}$$

parasitic pole of feedback loop

- Practical designs show 20-30% power penalty for booster
  - Since usually $C_2 < C_L$

If it works, do it again!

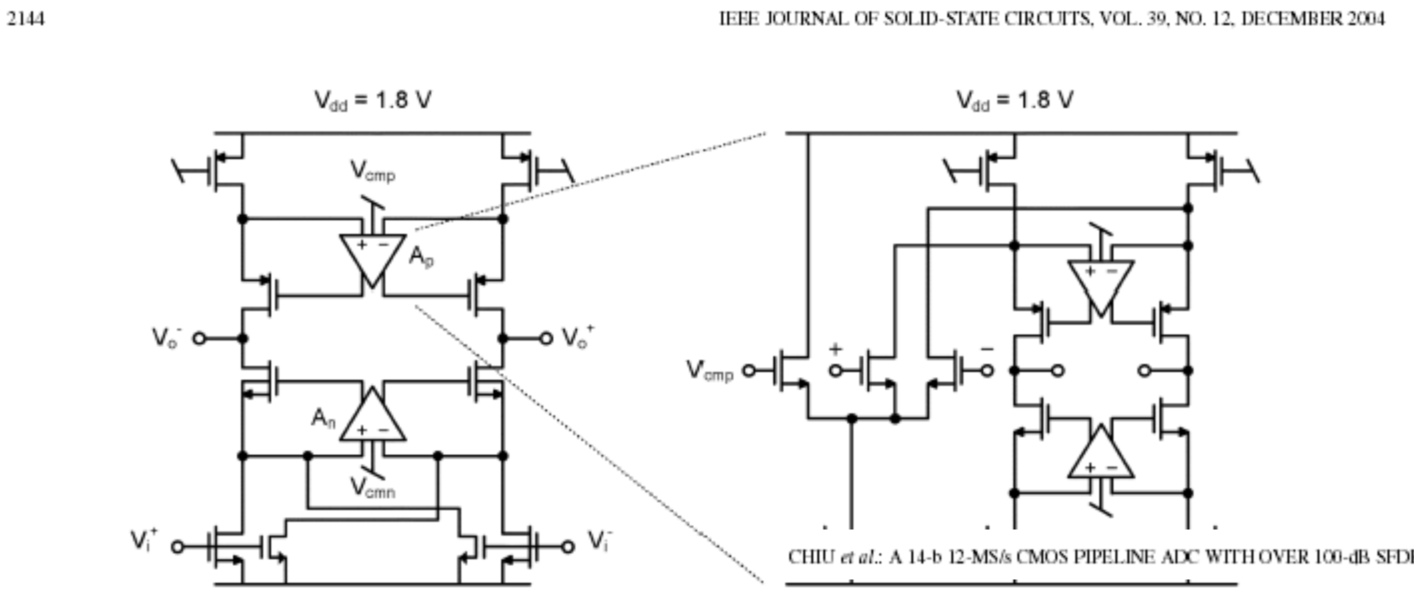
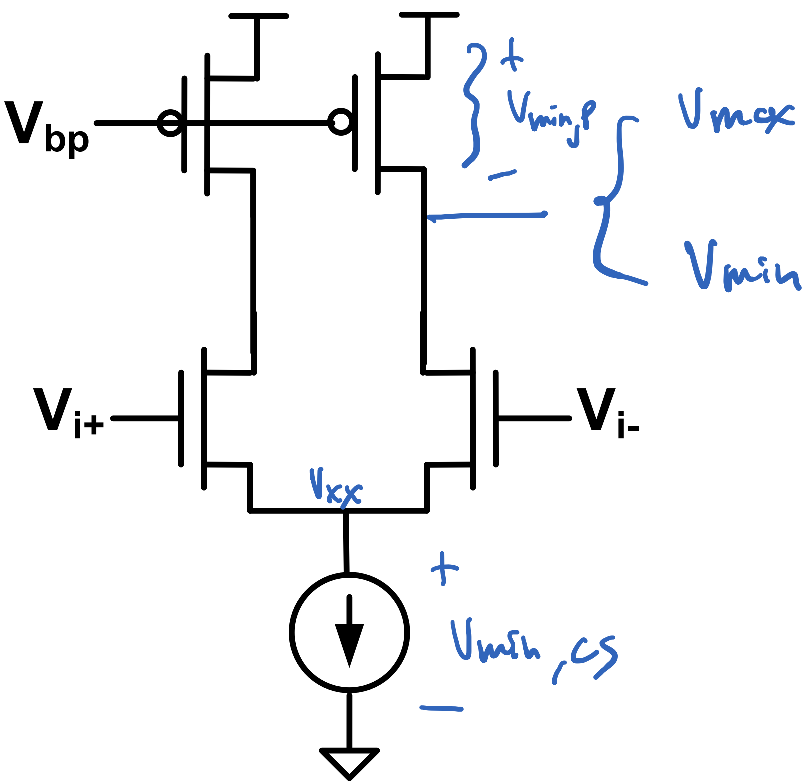


Fig. 8. Nested CMOS gain-boosting technique.

- Since in advanced scaled CMOS $g_m r_o$ is small, we can use nested gain boosting for higher output impedance.
- Watch out for pole-zero doublets!

Telescopic OTA: Common Mode vs. Swing



$$V_{max} = V_{DD} - V_{min,p}$$

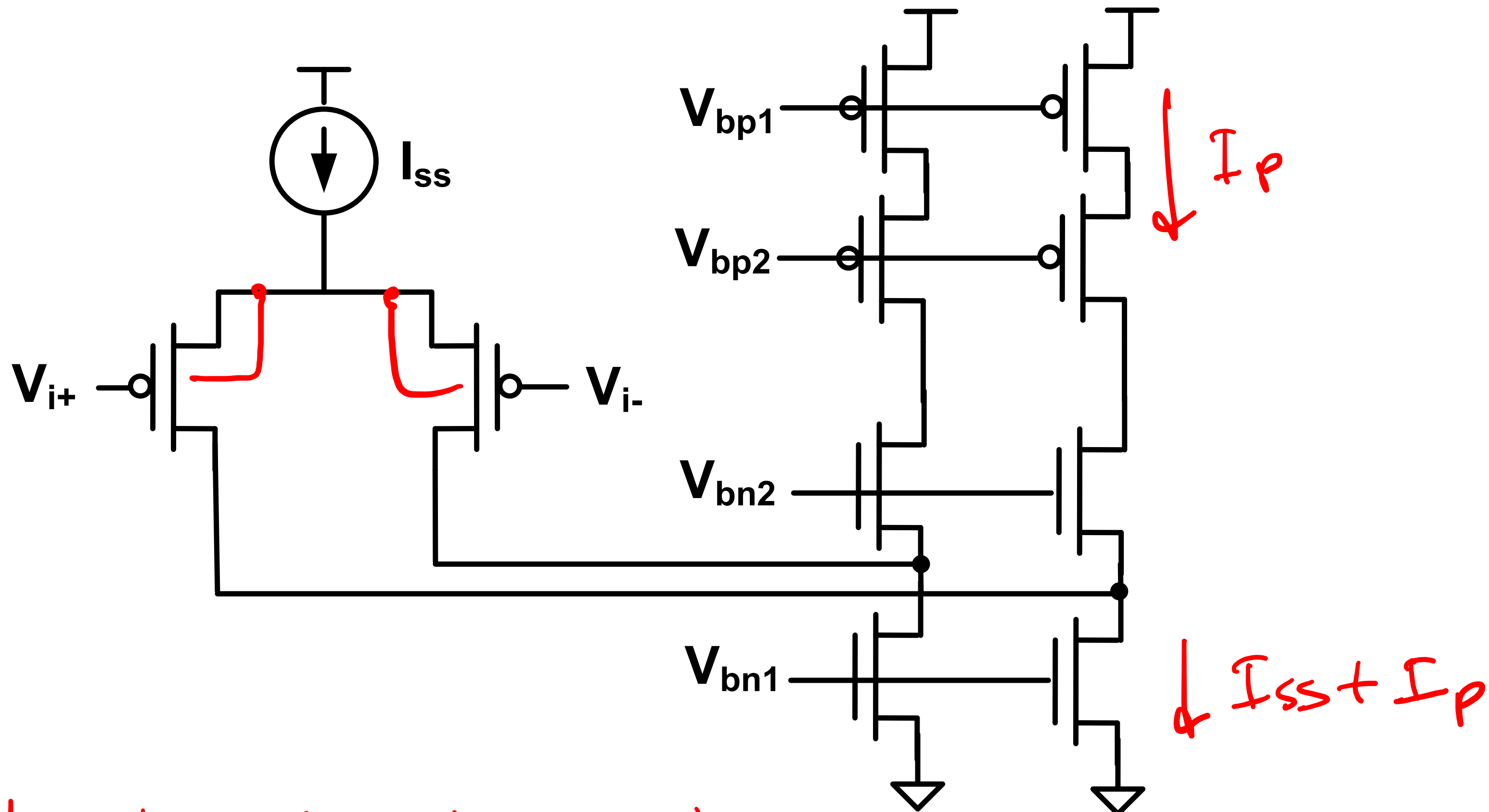
$$V_{min} = \text{set by CM}$$

$$V_{xx} = V_{CMi} - V_{th} - V_n^*$$

$$V_{min} = V_{min,n} + V_{xx}$$

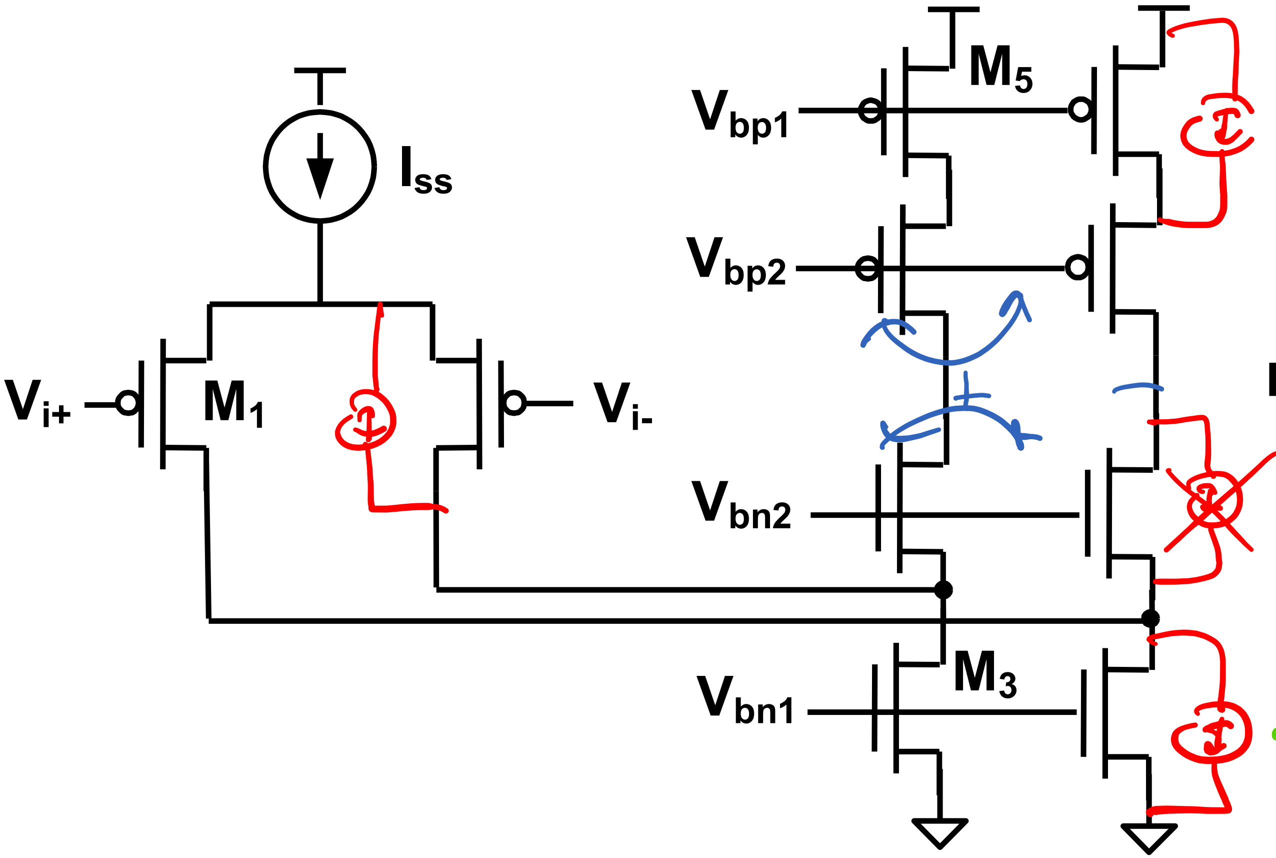
$$V_{sw} = V_{max} - V_{min}$$

Folded-Cascode Schematic



Input CM does not affect output swing

Folded-Cascode Noise



DC I_{ss}

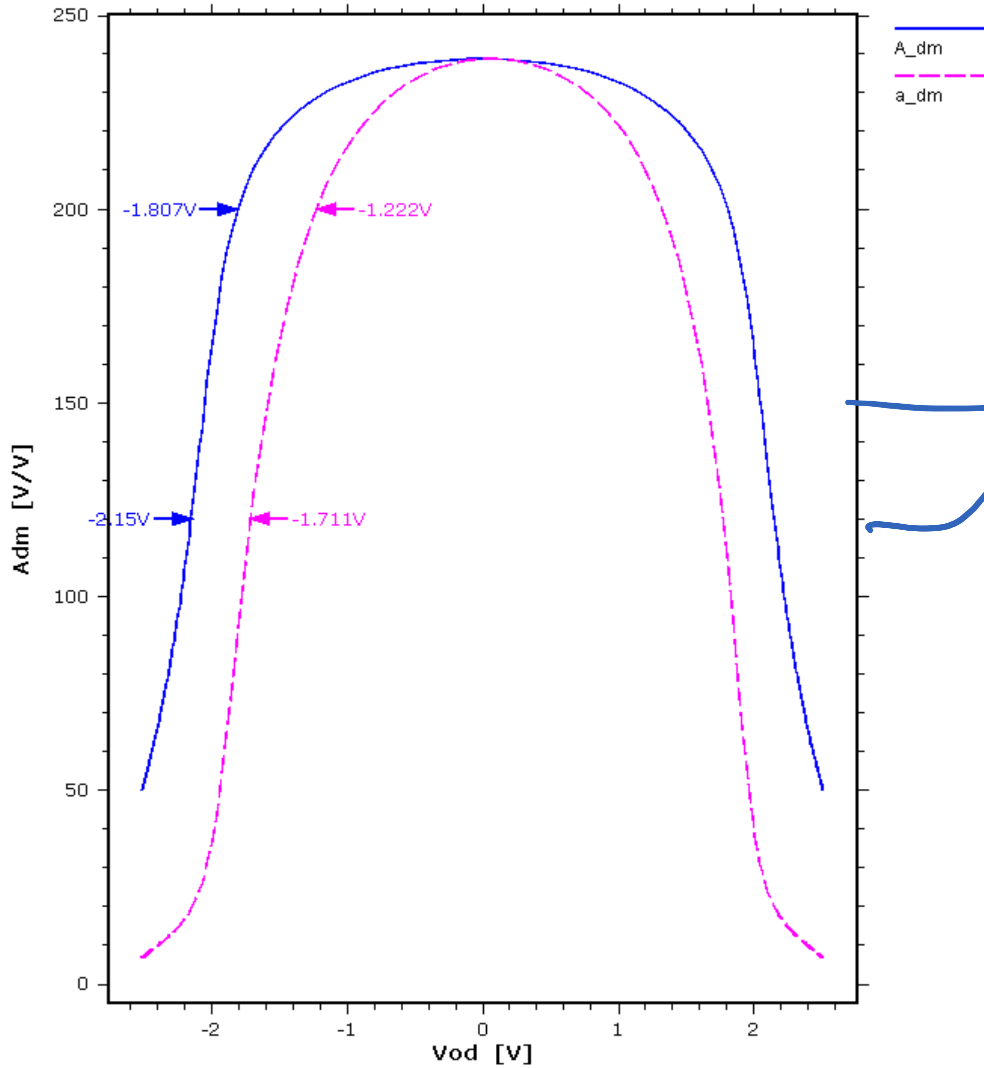
$$nf = 1 + \frac{V_1^*}{V_3^*} \times 2$$

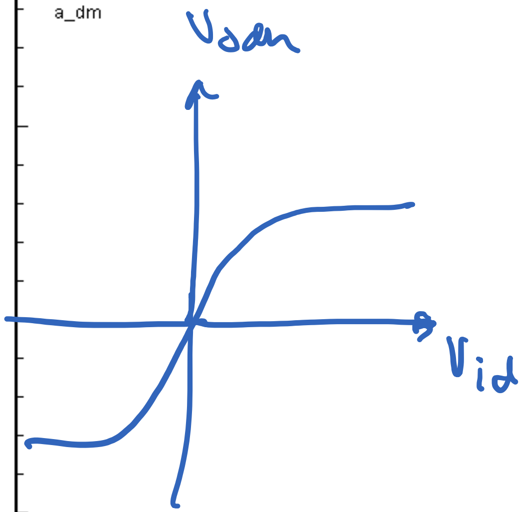
$2I_{ss}$

$$\frac{V_1^*}{V_5^*} \times 1$$

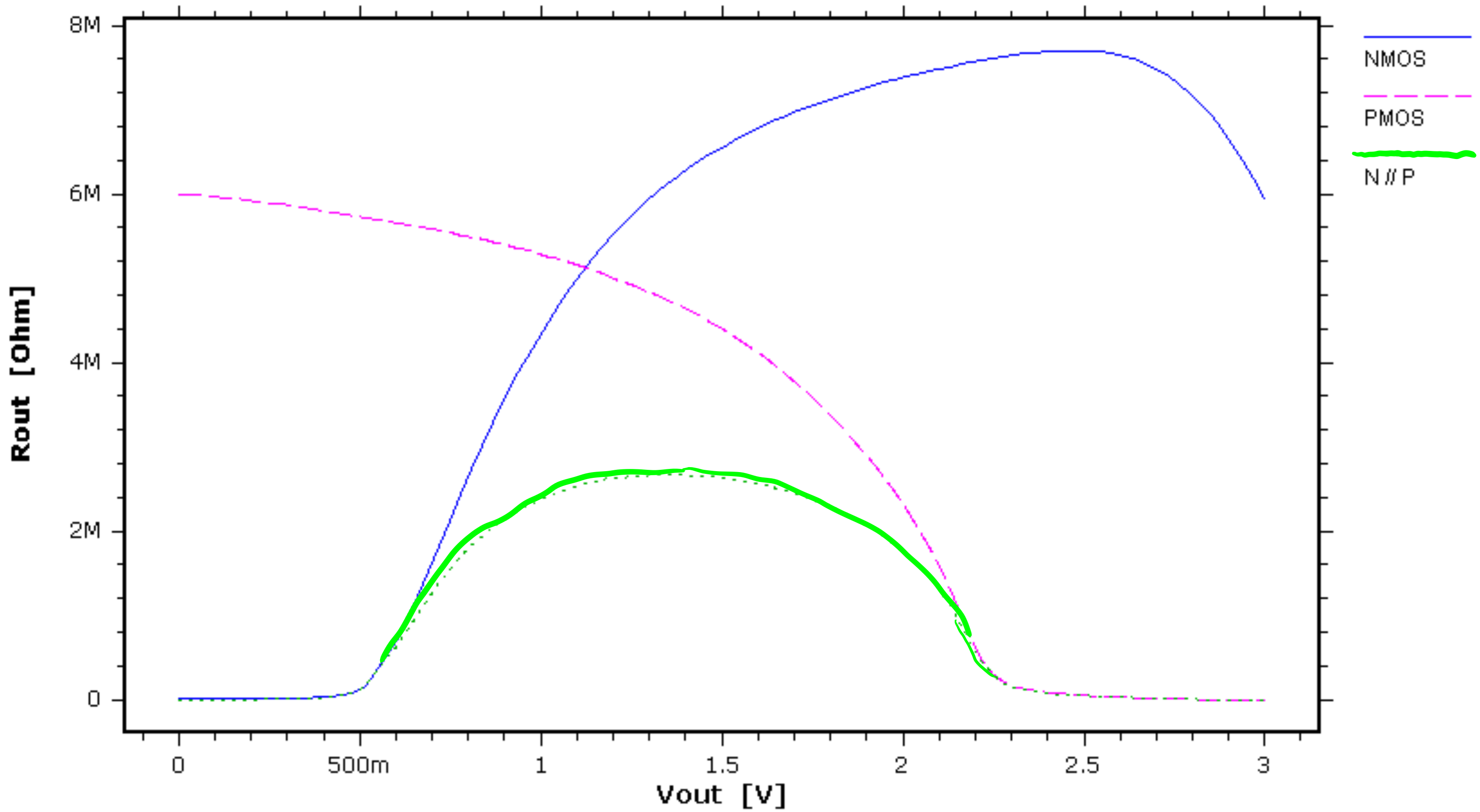
Assume DC currents all equal

Low Frequency Gain: a_v and A_v



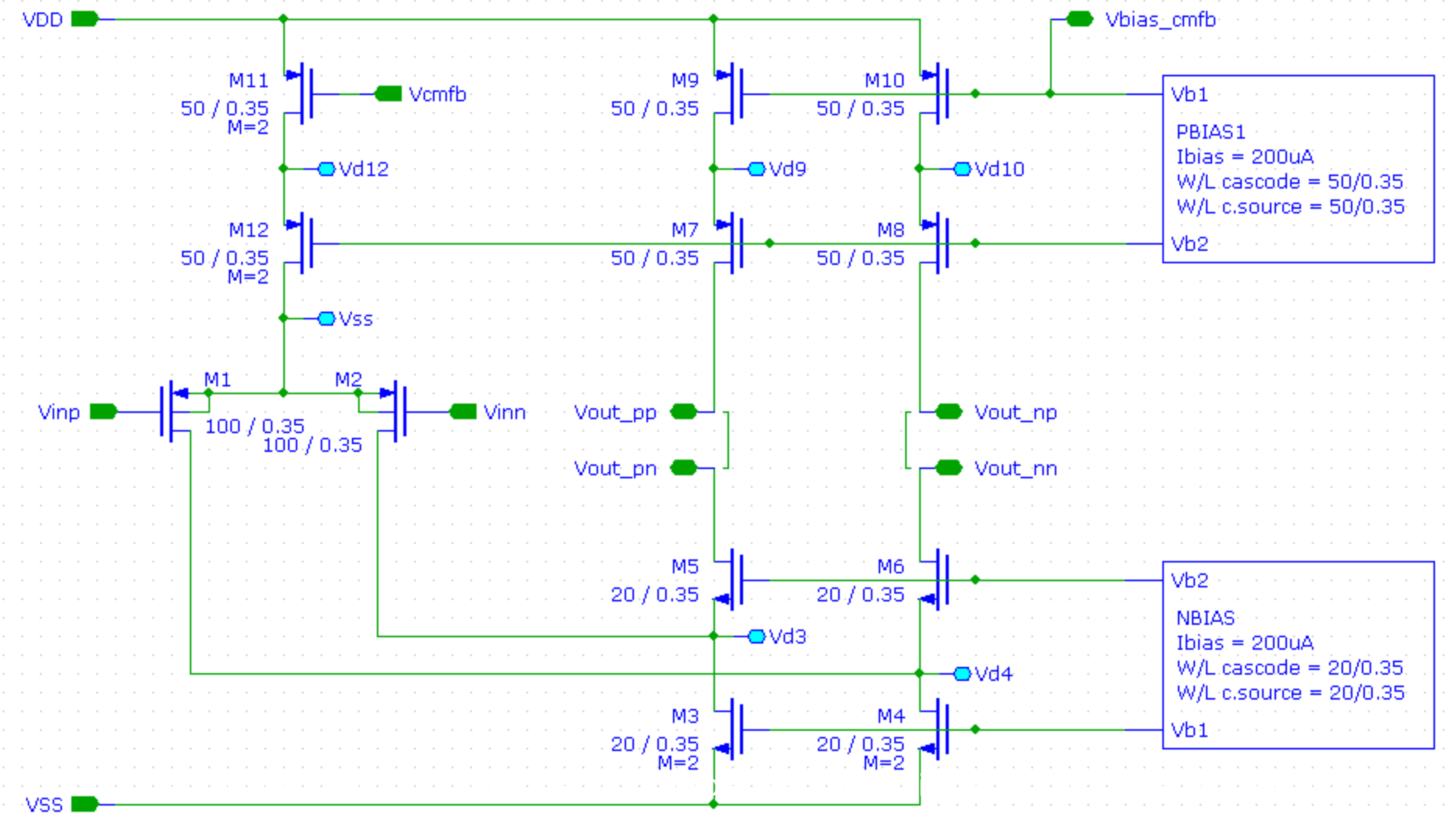


Output Resistance

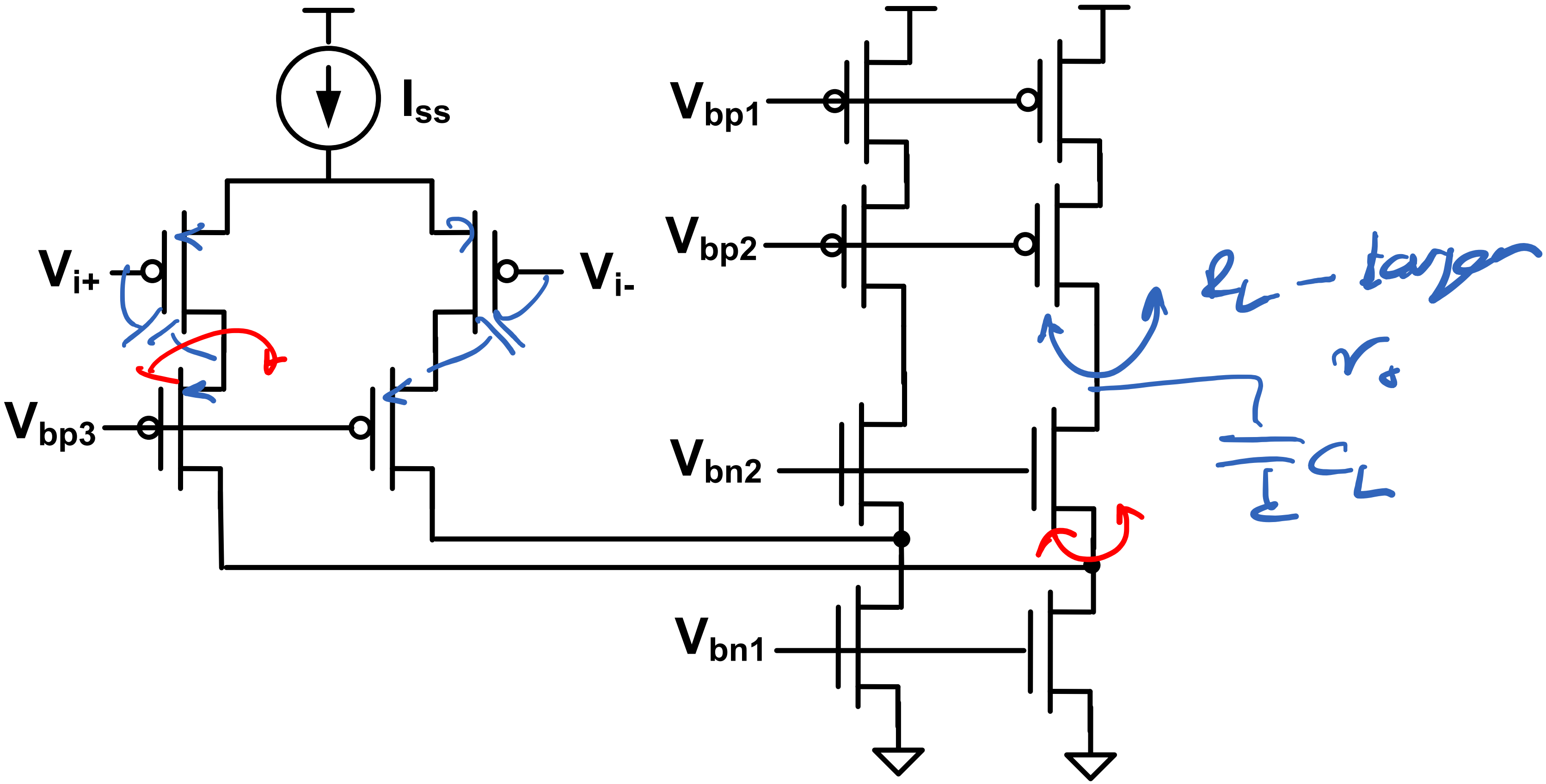


Beware of r_o imbalance between NMOS and PMOS current sources

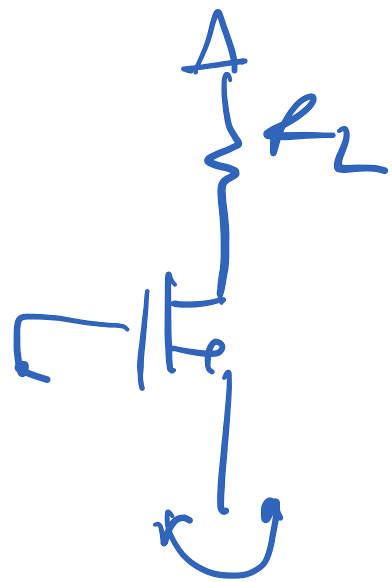
Simulation Schematic

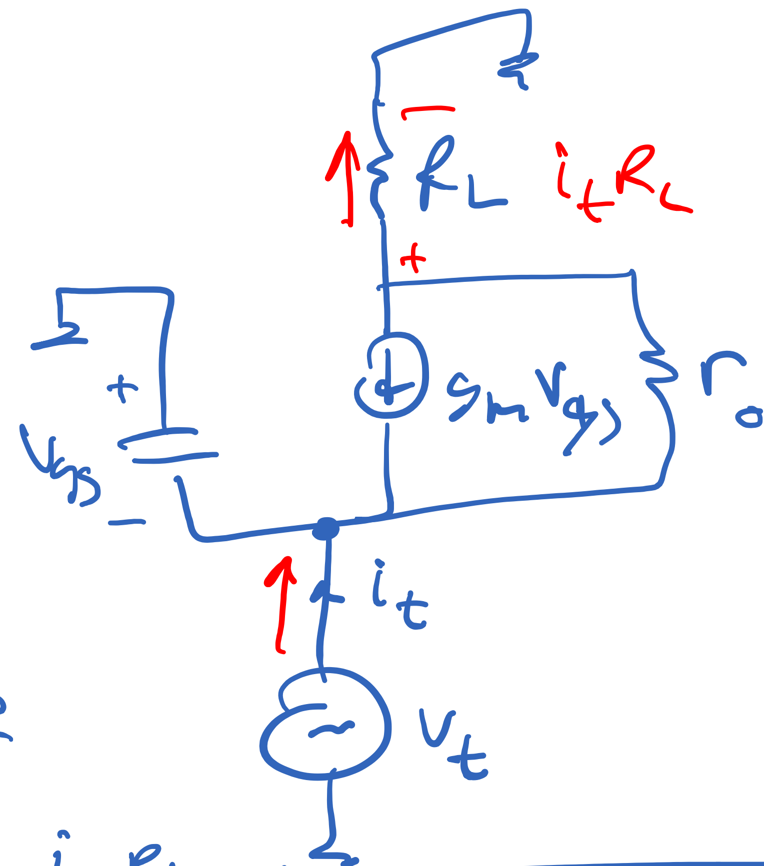


Input Cascode



Reminder: Cascode Zin





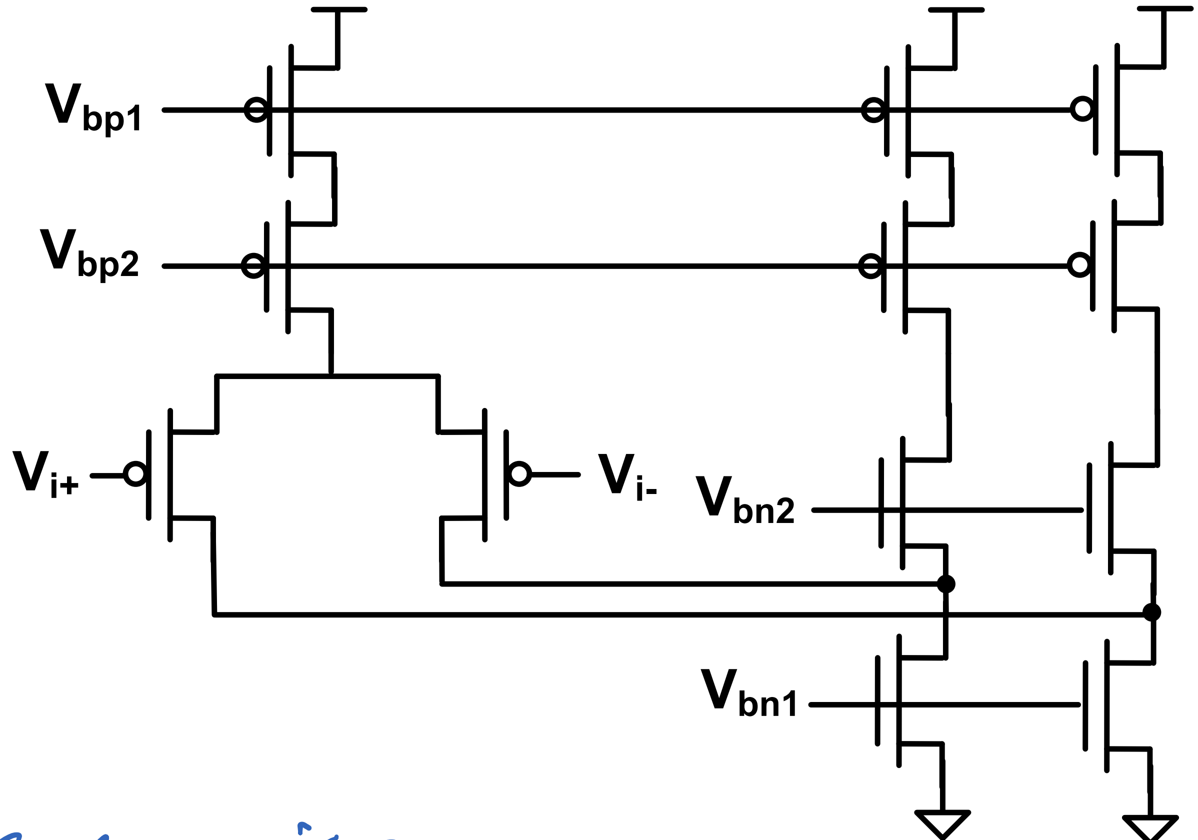
$$i_t = -g_m v_{gs} + \frac{v_t - v_o}{r_o}$$

$$= g_m v_t + \frac{v_t}{r_o} - i_t \frac{R_L}{r_o}$$

$$i_t\left(1 + \frac{R_L}{r_o}\right) = \left(g_m + \frac{1}{r_o}\right) v_t$$

$$\boxed{G_{in} \approx \frac{g_m}{1 + \frac{R_L}{r_o}}}$$

Biasing and Parasitic Feedback



CM noise