EE223 Analog Integrated Circuits
Fall 2018

Lecture 18: Frequency Response 2

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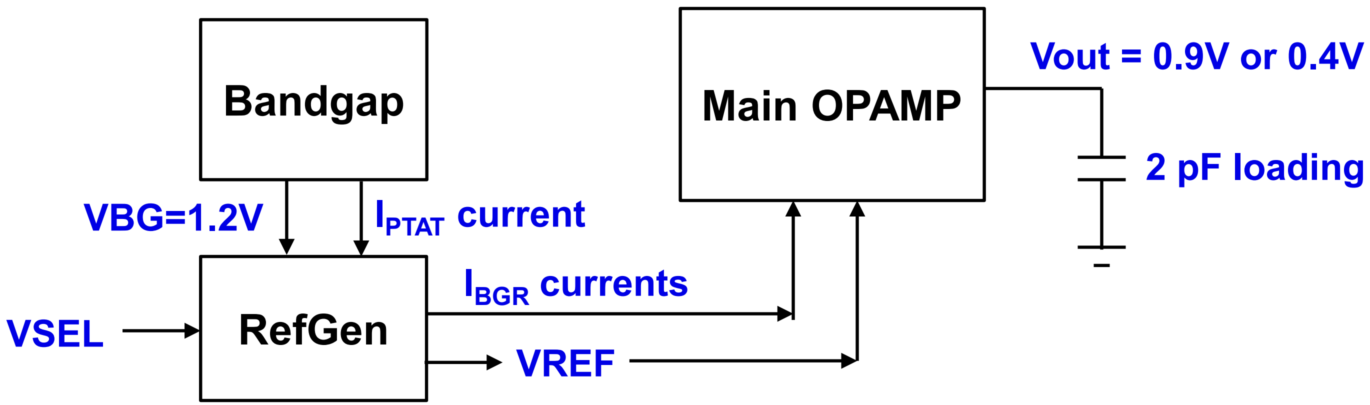
ENG-259

Project Description

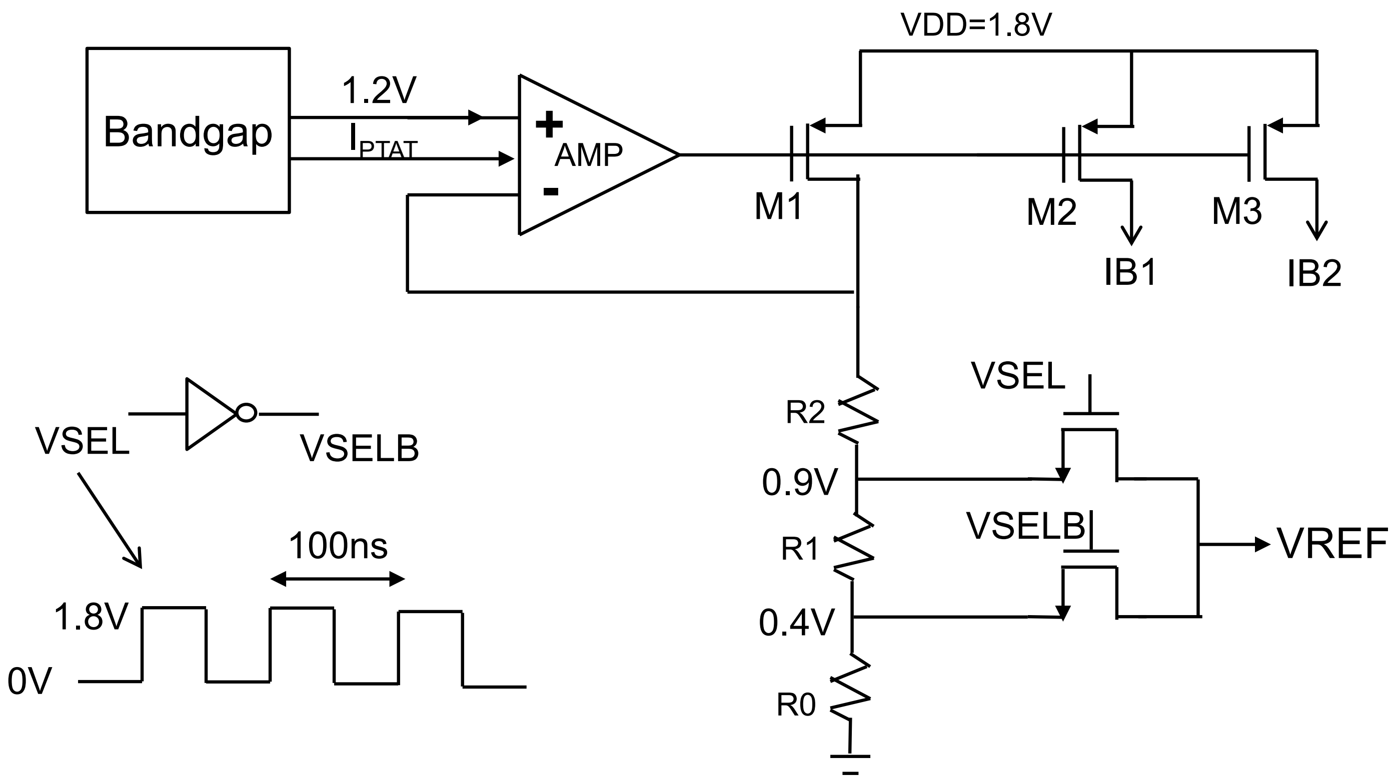


VSEL pulse is used to create either VREF = 0.9V or 0.4V.
See page 6 for the circuit diagram to implement this.

Main Circuit – High Gain OPAMP
Auxiliary Circuits – Bandgap, RefGen

- VDD=1.8V
- Device types available for the design
  → nmos2v, pmos2v, nmos2v_nat, vpnp5, resnsppoly, Ideal cap
- Results should meet the requirement over the following PVT corners
  - TT, 1.8V, 27C
  - SS, FF
  - 1.7V, 1.9V
  - -40C, 125C

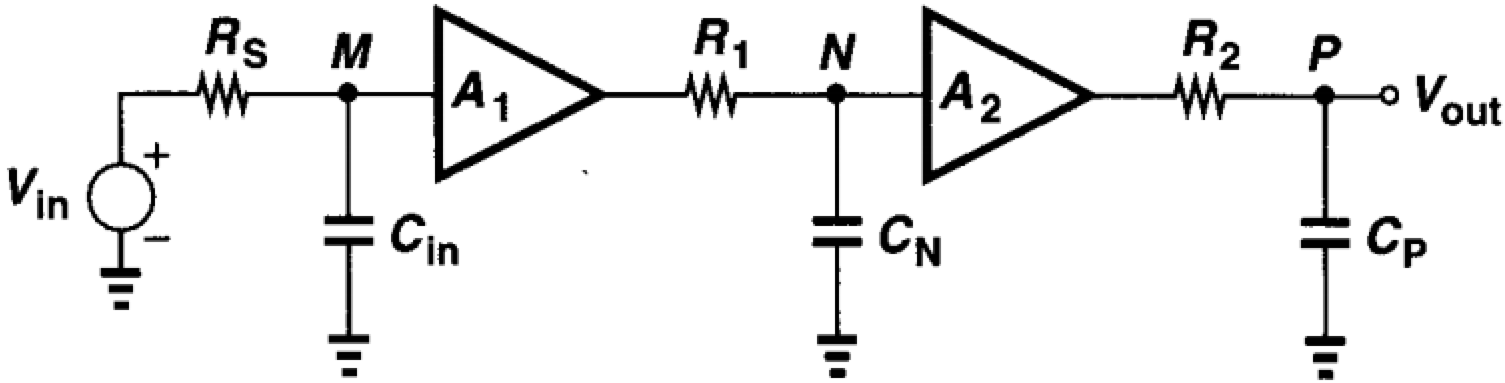
Reference Generator



Design Points

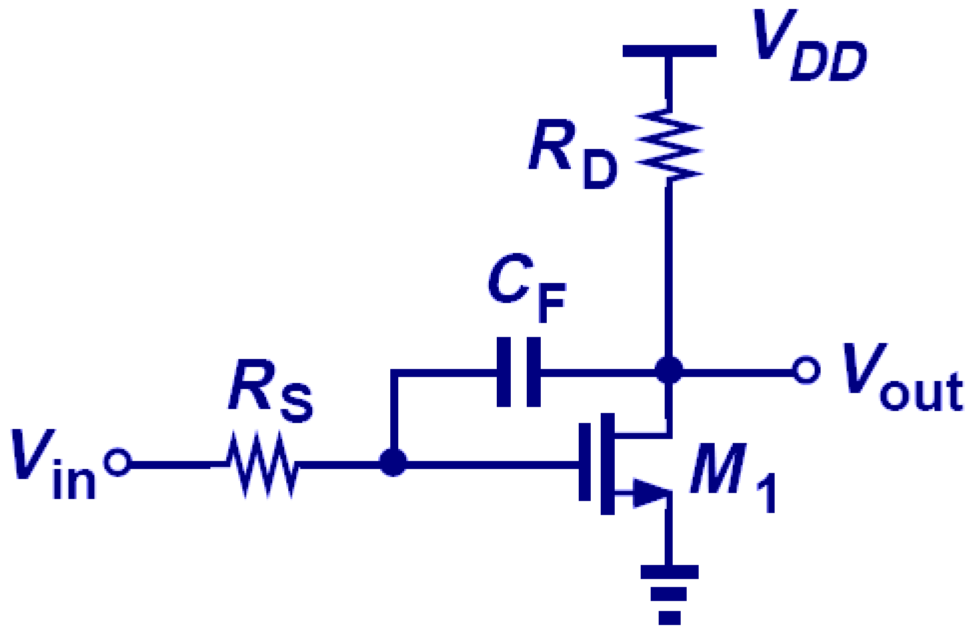
1. AMP and M1 will constitute a 2-stage RC compensated OPAMP.
2. AMP needs to be stable in unity gain configuration.
3. VREF is the input voltage to the main OPAMP.
4. VSEL is the input pulse you have to apply to measure the output of the main OPAMP.
5. IB1 and IB2 will supply the currents required in the Main Amplifier.

Association of Poles with Nodes



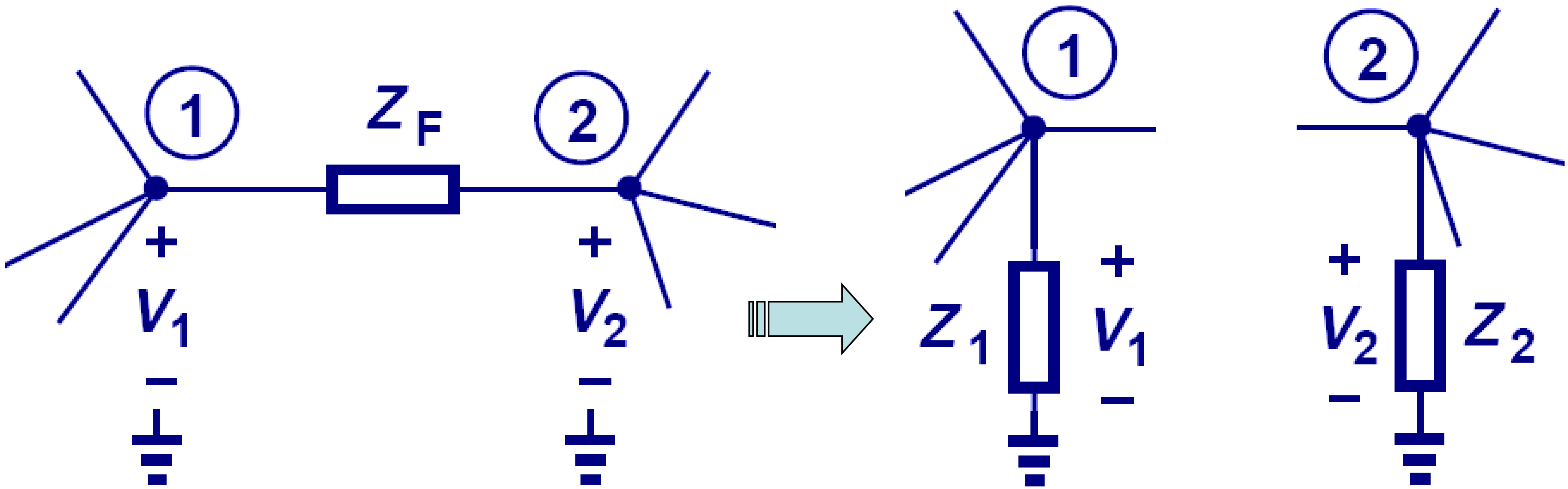
$$\frac{V_{out}}{V_{in}}(s) = \frac{A_1}{1 + R_S C_{in} s} \cdot \frac{A_2}{1 + R_1 C_N s} \cdot \frac{1}{1 + R_2 C_P s}$$

Circuit with Floating Capacitor



- The pole of a circuit is computed by finding the effective resistance and capacitance from a node to GROUND.
- The circuit above creates a problem since neither terminal of C_F is grounded.

Miller's Theorem

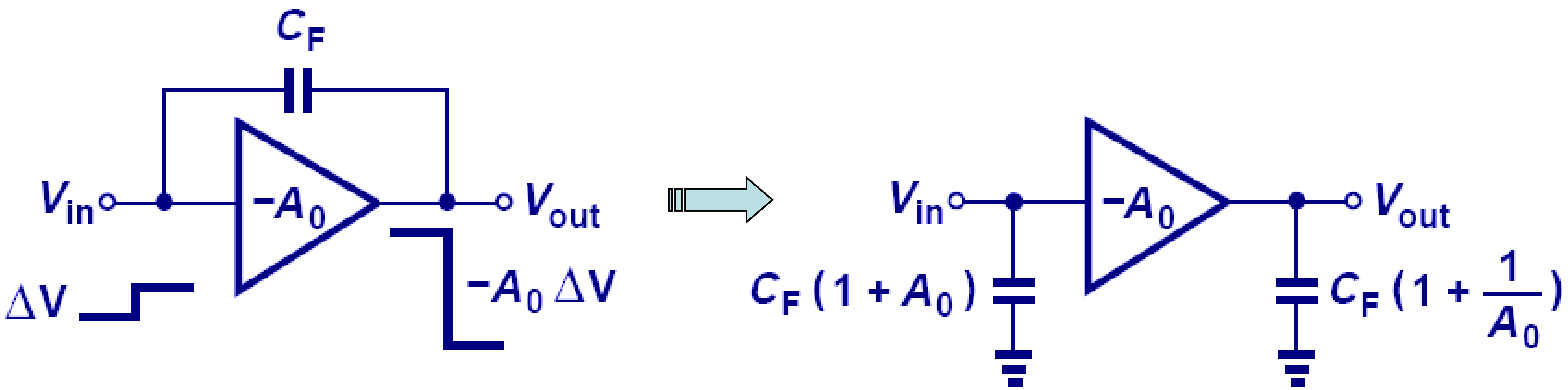


$$Z_1 = \frac{Z_F}{1 - A_v}$$

$$Z_2 = \frac{Z_F}{1 - 1/A_v}$$

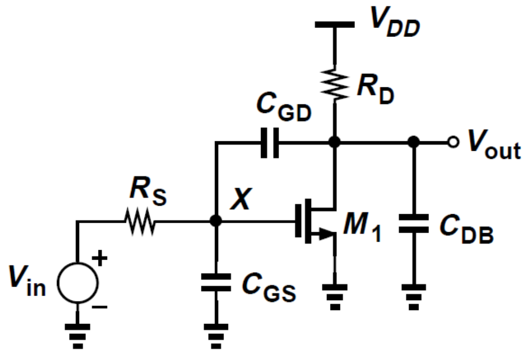
- If A_v is the gain from node 1 to 2, then a floating impedance Z_F can be converted to two grounded impedances Z_1 and Z_2.

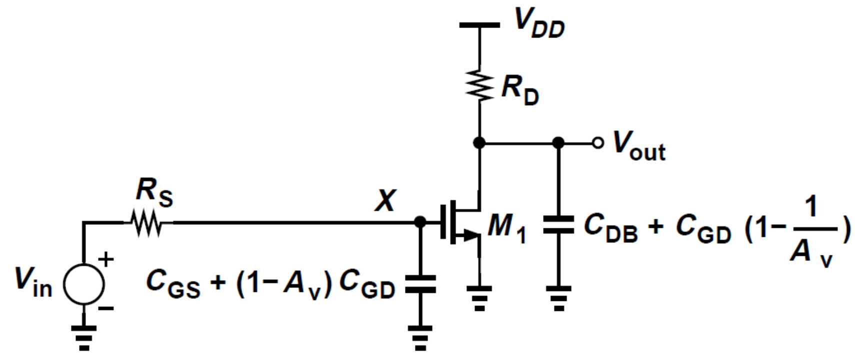
Miller Multiplication



- With Miller's theorem, we can separate the floating capacitor. However, the input capacitor is larger than the original floating capacitor. We call this **Miller multiplication**.

CS Frequency Response using Miller's Theorem





- **The magnitude of the "input" pole**

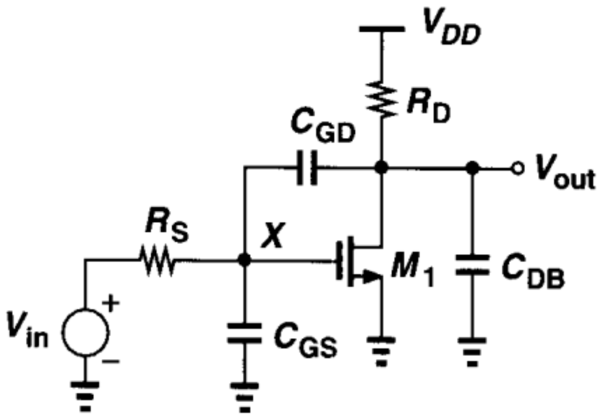
$$\omega_{in} = \frac{1}{R_S[C_{GS} + (1 + g_m R_D)C_{GD}]}$$

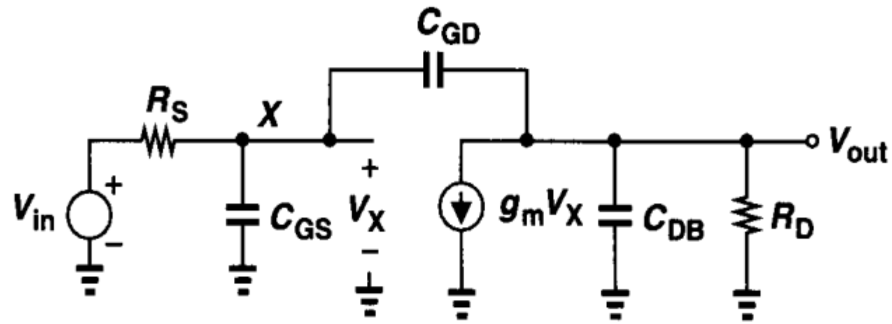
- **At the output node**

$$\omega_{out} = \frac{1}{R_D(C_{DB} + C_{GD})}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{-g_m R_D}{\left(1 + \frac{s}{\omega_{in}}\right)\left(1 + \frac{s}{\omega_{out}}\right)}$$

Exact Analysis of Common Source Frequency Response





$$\frac{V_X - V_{in}}{R_S} + V_X C_{GS} s + (V_X - V_{out}) C_{GD} s = 0$$

$$(V_{out} - V_X) C_{GD} s + g_m V_X + V_{out}\left(\frac{1}{R_D} + C_{DB} s\right) = 0.$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1 + g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD} + C_{DB})]s + 1}$$

$$\xi = C_{GS}C_{GD} + C_{GS}C_{DB} + C_{GD}C_{DB}$$

Dominant Pole Approximation

$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1 + g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD} + C_{DB})]s + 1}$$

$$D = \left(\frac{s}{\omega_{p1}} + 1\right)\left(\frac{s}{\omega_{p2}} + 1\right)$$

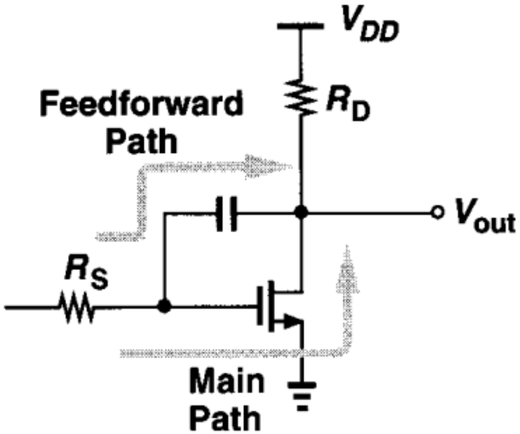
$$= \frac{s^2}{\omega_{p1}\omega_{p2}} + \left(\frac{1}{\omega_{p1}} + \frac{1}{\omega_{p2}}\right)s + 1 \quad = as^2 + bs + 1$$

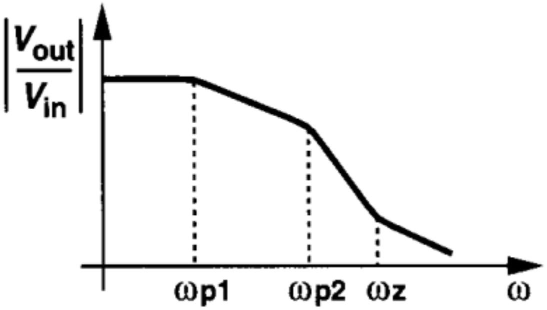
If we assume $|\omega_{p1}| \ll |\omega_{p2}|$ $\quad \rightarrow \quad b = \frac{1}{\omega_{p1}}$

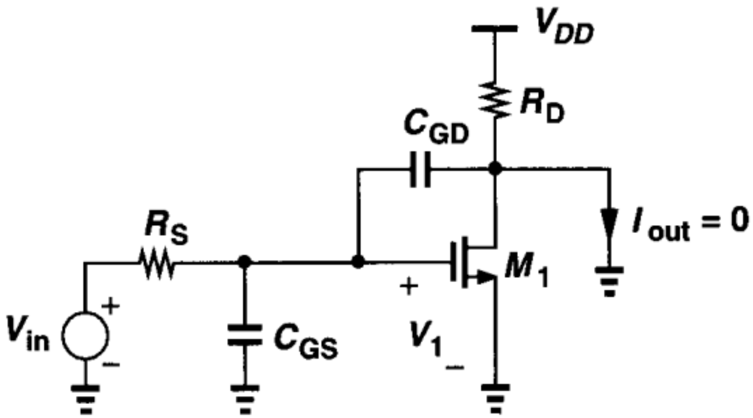
$$\omega_{p1} = \frac{1}{R_S(1 + g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD} + C_{DB})} \approx \omega_{in} = \frac{1}{R_S[C_{GS} + (1 + g_m R_D)C_{GD}]}$$

Zero in the Transfer Function

$$\frac{V_{out}}{V_{in}}(s) = \frac{(C_{GD}s - g_m)R_D}{R_S R_D \xi s^2 + [R_S(1+g_m R_D)C_{GD} + R_S C_{GS} + R_D(C_{GD}+C_{DB})]s + 1}$$





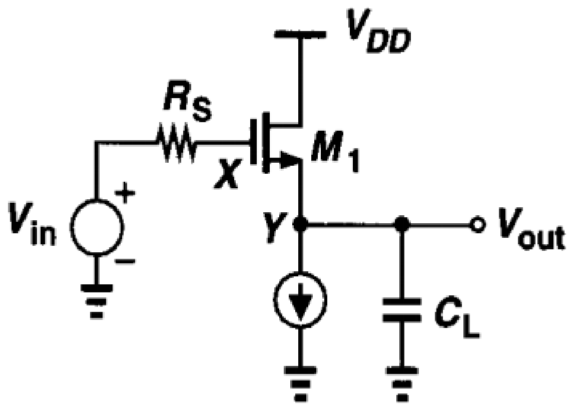


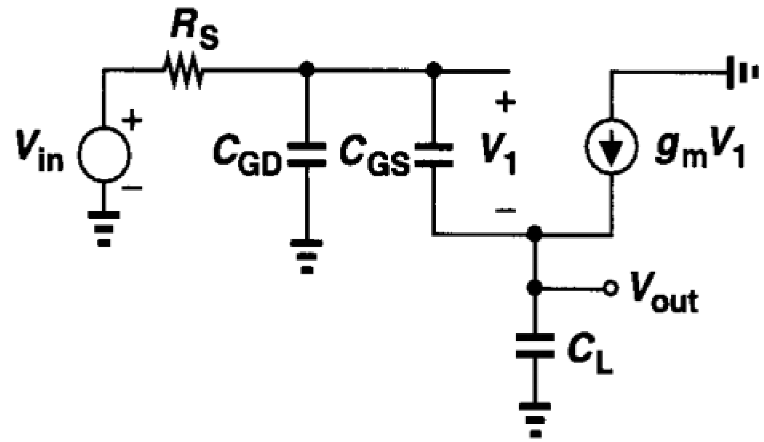
$$V_1 C_{GD} s_z = g_m V_1$$

$$s_z = +g_m / C_{GD}$$

RHP Zero :
Capacitance between input and output of inverting stage

Source Follower Frequency Response





LHP Zero :
Capacitance between input and output of non-inverting stage

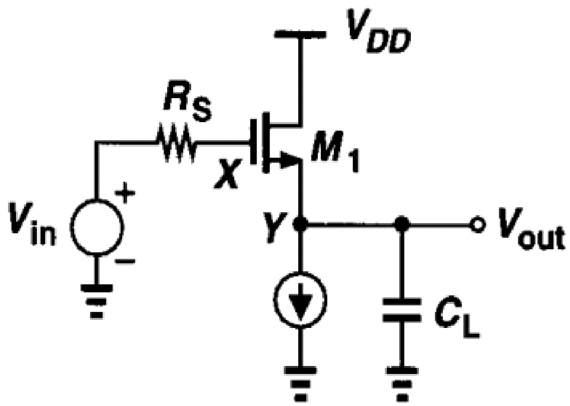
$$V_1 C_{GS} s + g_m V_1 = V_{out} C_L s$$

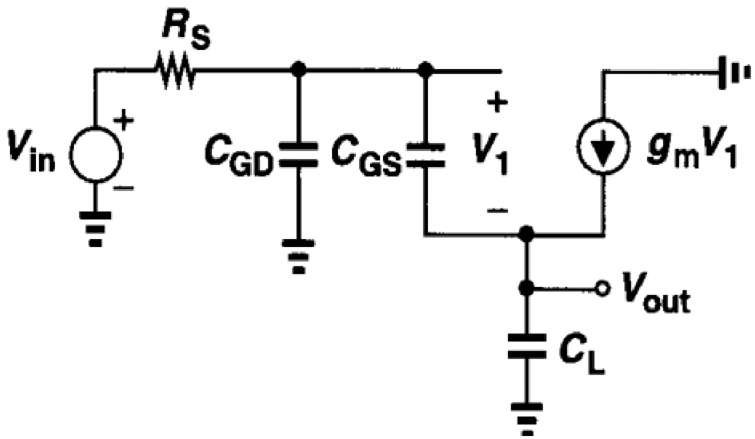
$$V_1 = \frac{C_L s}{g_m + C_{GS} s} V_{out}$$

$$V_{in} = R_S[V_1 C_{GS} s + (V_1 + V_{out}) C_{GD} s] + V_1 + V_{out}$$

$$\frac{V_{out}}{V_{in}}(s) = \frac{g_m + C_{GS} s}{R_S(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L)s^2 + (g_m R_S C_{GD} + C_L + C_{GS})s + g_m}$$

Source Follower Frequency Response



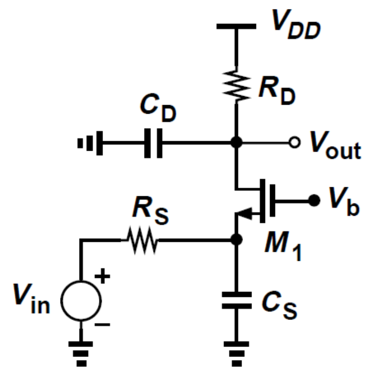


$$\frac{V_{out}}{V_{in}}(s) = \frac{g_m + C_{GS}s}{R_S(C_{GS}C_L + C_{GS}C_{GD} + C_{GD}C_L)s^2 + (g_m R_S C_{GD} + C_L + C_{GS})s + g_m}$$

$$\omega_{p1} \approx \frac{g_m}{g_m R_S C_{GD} + C_L + C_{GS}} = \frac{1}{R_S C_{GD} + \dfrac{C_L + C_{GS}}{g_m}}.$$

If $R_S = 0$, then $\omega_{p1} = g_m/(C_L + C_{GS})$

Common Gate Stage



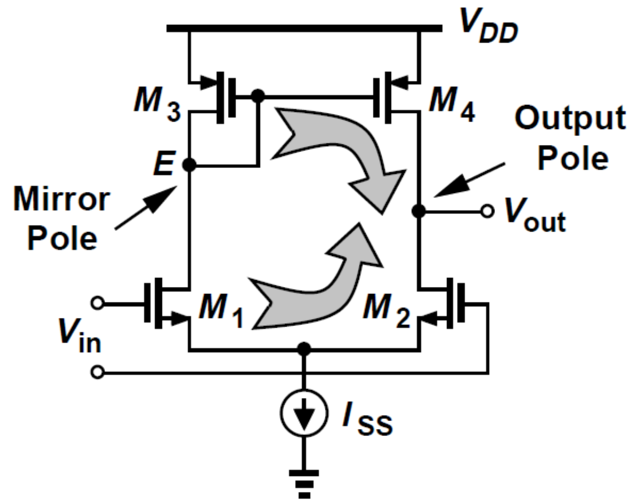
Common-gate stage at high frequencies

- A transfer function

$$\frac{V_{out}}{V_{in}}(s) = \frac{(g_m + g_{mb})R_D}{1 + (g_m + g_{mb})R_S} \frac{1}{\left(1 + \frac{C_S}{g_m + g_{mb} + R_S^{-1}} s\right)(1 + R_D C_D s)}$$

- No Miller multiplication of capacitances.
- Use when a low input impedance is required
- In cascode stages

Differential Pair with Active Loads - OTA

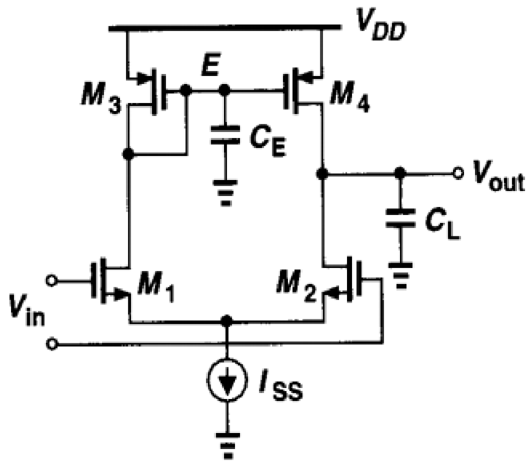


- How many poles does this circuit have?
- The pole associated with node E is called a “mirror pole.”

$$\frac{V_{out}}{V_{in}} = \frac{g_{mN} r_{ON}(2g_{mP} + C_E s) r_{OP}}{2r_{OP} r_{ON} C_E C_L s^2 + [(2r_{ON} + r_{OP})C_E + r_{OP}(1 + 2g_{mP} r_{ON})C_L]s + 2g_{mP}(r_{ON} + r_{OP})}$$

$$\omega_{p1} \approx \frac{1}{(r_{ON} \| r_{OP})C_L} \qquad \omega_{p2} \approx \frac{g_{mP}}{C_E}$$

Differential Pair with Active Loads

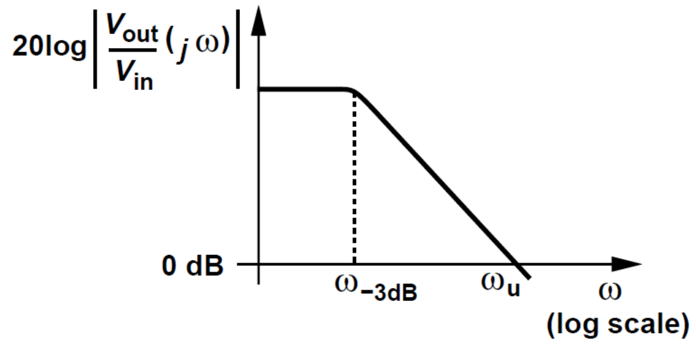


- A zero with a magnitude of $2g_{mP}/C_E$ in the left half plane.
- The appearance of such a zero is due to
  - "slow path" (M1,M3 and M4 to Vout) $A_0/[(1+s/\omega_{p1})(1+s/\omega_{p2})]$
  - "fast path" (M2 to Vout) $A_0/(1+s/\omega_{p1})$

$$\frac{V_{out}}{V_{in}} = \frac{A_0}{1+s/\omega_{p1}}\left(\frac{1}{1+s/\omega_{p2}}+1\right)$$

$$= \frac{A_0(2+s/\omega_{p2})}{(1+s/\omega_{p1})(1+s/\omega_{p2})}$$

Amplifier Gain-Bandwidth Trade-Offs



- We wish to maximize both the gain and the bandwidth of amplifiers.
- we are interested in both the 3-dB bandwidth, ω_{-3dB} , and the "unity-gain" bandwidth, ω_u.