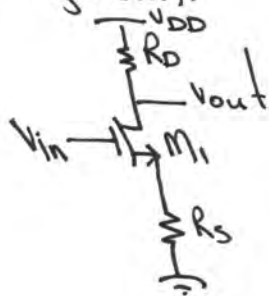


- ① (a) Source degeneration is when a circuit element (eg. resistor) is connected between the MOSFET source terminal and the common node (eg. V_{DD} or ground).



Advantage (over CS)

- Output impedance is increased (from $R_{D||}$ to $R_{D||}(r_{o1} + R_S + g_{m1}r_{o1}R_S)$).
- Large-signal operation is linearised (i.e. larger small-signal region).

Disadvantages (over CS)

- Reduced headroom
- Reduced voltage gain = $-\frac{R_D}{\frac{1}{g_m} + R_S}$ (from $-g_m R_D$)

(b) (i) $A_v = \frac{v_{out}}{v_{in}} = \frac{(R_E + \frac{1}{g_{m2}} || r_{o2}) || r_{o1}}{(R_E + \frac{1}{g_{m2}} || r_{o2}) || r_{o1} + \frac{1}{g_{m1}}}$ (if $\lambda > 0$)

or

$= \frac{R_E + \frac{1}{g_{m2}}}{R_E + \frac{1}{g_{m2}} + \frac{1}{g_{m1}}}$ (if $\lambda = 0$)

* question should state if $\lambda = 0$ or $\lambda > 0$

(ii) $A_v = \frac{v_{out}}{v_{in}} = \frac{v_{out}}{v_x} \times \frac{v_x}{v_{in}}$

$\frac{v_{out}}{v_x} = g_{m1} \cdot (g_{m2}r_{o2}R_E + R_E + r_{o2}) || r_{o1}$

$\frac{v_x}{v_{in}} = \frac{\frac{1}{g_{m1}}}{\frac{1}{g_{m1}} + R_S}$

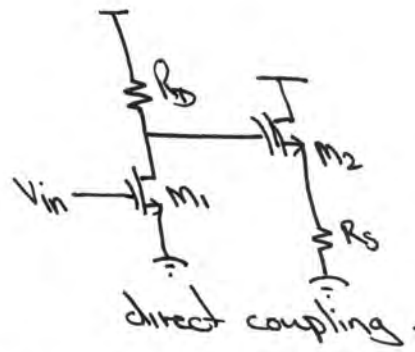
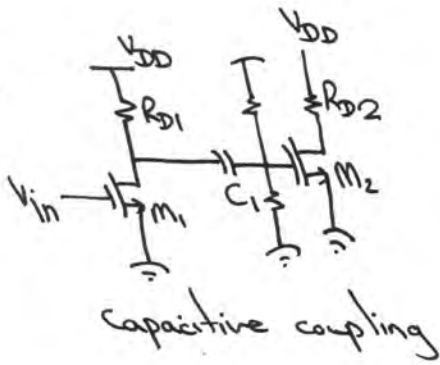
$A_v = \frac{(g_{m2}r_{o2}R_E + R_E + r_{o2}) || r_{o1}}{\frac{1}{g_{m1}} + R_S}$

* if $\lambda > 0$

or

$A_v = \frac{v_{out}}{v_{in}} = \frac{R_E}{\frac{1}{g_{m1}} + R_S}$ if $\lambda = 0$ *

(c)

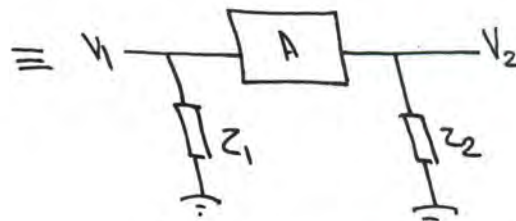
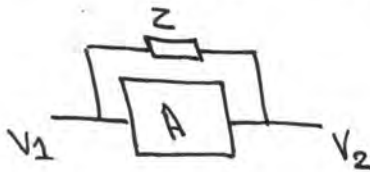


capacitive coupling is when amplifier stages are cascaded by using a capacitor for ac coupling, i.e. biasing each stage independently.

Direct coupling is when the amplifier stages are directly connected and ^{the} circuit must therefore be designed to be biased correctly as a whole.

In IC design, constraints in large capacitance realisation means that capacitive coupling is undesirable for low frequency signals.

(d)



$$\text{where } Z_1 = Z \left(\frac{1}{1+A} \right)$$

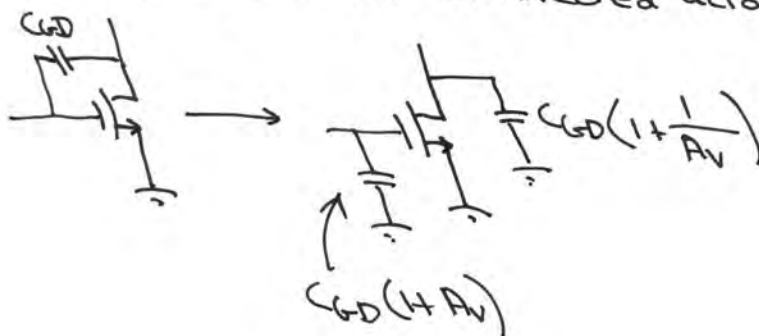
$$Z_2 = Z \left(\frac{1}{1+1/A} \right)$$

$$\text{if } A = \frac{V_2}{V_1}$$

if the impedance is a capacitor $\therefore C_1 = C(1+A)$

$$C_2 = C \left(1 + \frac{1}{A} \right)$$

The Miller's theorem is extremely useful in resolving floating capacitances, in particular connected across an amplifying device.



(e) ~~Gain~~ 4 advantages are:

1. gain desensitisation
2. extension of bandwidth
3. modification in input/output impedance
4. linearity improvement

for ① for example in op-amp \rightarrow open-loop gain may vary by $\pm 20\%$.
however when operated in closed loop (eg. non-inverting amplifier)

$$\frac{V_{out}}{V_{in}} = \frac{A_o}{1 + \frac{R_2}{R_1 + R_2} A_o} \text{ and if } A_o \gg 1 \Rightarrow \frac{V_{out}}{V_{in}} \approx 1 + \frac{R_2}{R_1}$$

for ② if we consider a one-pole amplifier:

$$A_1(s) = \frac{A_o}{1 + s/\omega_o}$$

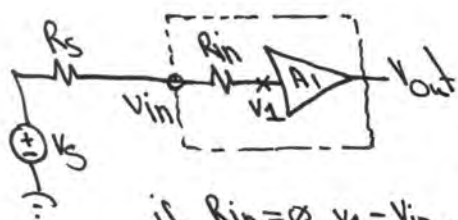
if we substitute into a closed-loop system:

$$\begin{aligned} \frac{Y}{X} &= \frac{\frac{A_o}{1 + s/\omega_o}}{1 + \frac{K A_o}{1 + s/\omega_o}} = \frac{A_o}{1 + K A_o + s/\omega_o} \\ &= \frac{A_o}{1 + K A_o} \cdot \frac{1}{1 + \frac{s}{(1 + K A_o)\omega_o}} \end{aligned}$$

$$\therefore \text{closed-loop BW} = (1 + K A_o)\omega_o$$

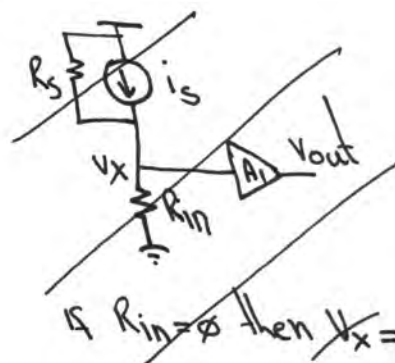
(f) voltage amplifier: $R_{in} = \infty$ ideally

transimpedance amplifier $R_{in} = \emptyset$ ideally



if $R_{in} = \emptyset$, $v_1 = V_{in}$, and $V_{out} = A_1 V_s$
otherwise:

$$V_{out} = \left(\frac{V_s R_{in}}{R_{in} + R_s} \right) A_1$$



~~If $R_{in} = \emptyset$ then $V_x =$~~

(g) The transit frequency (f_T) is defined as the frequency at which the current gain of a transistor becomes unity. This gives a measure of the maximum "speed" of the transistor.

$$f_T = \frac{1}{2\pi} \cdot \frac{g_m}{C_{BE}} \rightarrow \text{if } C_{BE} = 2.2 \text{ pF} \text{ and } I_C = 5 \text{ mA}$$

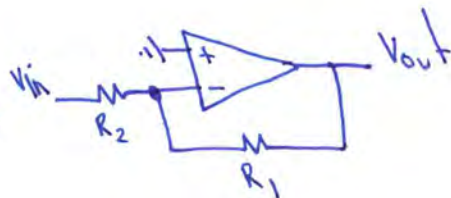
$$f_T = \frac{1}{2\pi} \frac{I_C}{V_T \cdot C_{BE}} = \frac{1 (5 \text{ mA})}{2\pi (26 \text{ mV}) (2.2 \text{ pF})} = 13.91 \text{ GHz}$$

(h) $A_V(\text{o.l.}) = 106 \text{ dB} = \frac{106}{20} = 200,000$

$A_V(\text{c.l.}) = 250$

As inverting amplifier

$\therefore \frac{R_1}{R_2} = 250$



$$\frac{V_{out}}{V_{in}} = -\frac{R_1}{R_2} \left(1 - \frac{1}{A_0} \left(1 + \frac{R_1}{R_2} \right) \right)$$

$$\frac{V_{out}}{V_{in}} = -250 \left(1 - \frac{1}{200000} \left(1 + 250 \right) \right)$$

ϵ

$\epsilon = 0.126\%$

(i) PVT Remember

P = Process variation

V = Power supply variation

T = Temperature

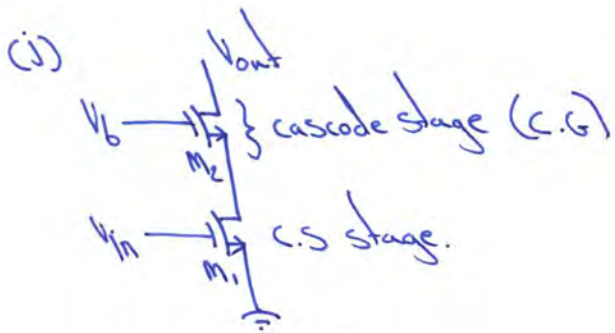
Three challenges of designing on chip reference (bias generator) circuits are the sensitivity (or dependence) on P, V and T.

To overcome apply design techniques as follows:

Process: make bias generator dependant on relative component ratios not absolute values.

Temperature: use combination of components with +ve and -ve temp. coefficient to cancel out

Power supply: Do not derive reference from P.S



A cascode stage can boost the output impedance of a C.S. stage from (r_{o1}) to $(g_{m2}r_{o1}r_{o2} + r_{o1} + r_{o2})$ and thus, if ~~$V_{out} = i_{out} \cdot R_{out}$~~

~~$\Rightarrow A_v = \frac{V_{out}}{V_{in}} = G_m R_{out}$~~

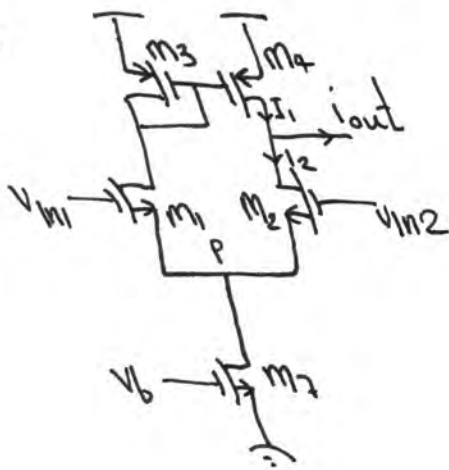
$V_{out} = i_{out} \times R_{out}$ where $i_{out} = G_m V_{in}$

$V_{out} = G_m V_{in} R_{out}$

$\frac{V_{out}}{V_{in}} = G_m R_{out} \therefore$ by increasing R_{out} , A_v is increased.

- Drawbacks:
- ① Reduced headroom therefore require higher supply voltage.
 - ② An additional device is required.

② (a)



$$i_1 = g_{m1} V_{in1} \text{ (assuming P is at gnd)}$$

$$i_2 = g_{m2} V_{in2}$$

$$i_{out} = g_{m1}(V_{in1} - V_{in2}) \text{ (assuming } M_1 = M_2)$$

$$R_{out} = r_{o4} \parallel r_{o2}$$

$$\therefore V_{out} = i_{out} R_{out} = g_{m1}(V_{in1} - V_{in2})[r_{o2} \parallel r_{o4}]$$

$$A_v = \frac{V_{out}}{V_{in1} - V_{in2}} = g_{m1}(r_{o2} \parallel r_{o4})$$

(b) $\frac{V_{out}}{V_y} = -g_{m5}(r_{o5} \parallel r_{o8})$ c.s stage.

$$A_v = \frac{V_{out}}{V_{in1} - V_{in2}} = -g_{m1} g_{m5}(r_{o2} \parallel r_{o4})(r_{o5} \parallel r_{o8})$$

(c) $\left(\frac{W}{L}\right)_6 = \frac{4}{2} \rightarrow \frac{V_{DD} - V_{GS}}{8K} = I_6, \quad I_6 = \frac{\mu_n C_{ox}}{2} \left(\frac{W}{L}\right)_6 (V_{GS} - V_{TH})^2$

$$\left(\frac{W}{L}\right)_7 = \frac{10}{2}$$

$$\left(\frac{W}{L}\right)_8 = \frac{20}{2}$$

$$\frac{V_{DD} - V_{GS}}{8K} = \frac{1}{2} \mu_n C_{ox} \left(\frac{W}{L}\right)_6 (V_{GS} - V_{TH})^2$$

$$1.8 - V_{GS} = 4000(200\mu)^2 (V_{GS} - 0.4)^2$$

$$1.8 - V_{GS} = 1.6(V_{GS}^2 + 0.16 - 0.8V_{GS})$$

$$\cancel{1.6V_{GS}^2 - 1.28V_{GS} + V_{GS} + 0.256 - 1.8 = 0}$$

$$\cancel{1.6V_{GS}^2 - 0.28V_{GS} - 1.544 = 0}$$

$$1.6V_{GS}^2 - 1.28V_{GS} + V_{GS} + 0.256 - 1.8 = 0$$

$$1.6V_{GS}^2 - 0.28V_{GS} - 1.544 = 0$$

$$V_{GS6} = 1.074V$$

$$\therefore I_D = \frac{1.8 - 1.074}{8K}$$

$$= 90.9\mu A$$

$$\therefore \text{Power} = I_{\text{total}} \times V_{DD}$$

$$I_{\text{total}} = I_6 + I_7 + I_8$$

$$= 90.9 \mu(1 + 2.5 + 5)$$

$$= 772 \mu\text{A}$$

$$\text{Power} = IV = 772 \mu(1.8) = \underline{1.4 \text{ mW}}$$

$$(d) \left. \begin{aligned} \left(\frac{W}{L}\right)_{1,2} &= \frac{250}{1} \\ \left(\frac{W}{L}\right)_{3,4} &= \frac{4}{4} \end{aligned} \right\}$$

$$A_V = g_{m1} g_{m5} (r_{o2} \parallel r_{o4}) (r_{o5} \parallel r_{o8})$$

$$\left(\frac{W}{L}\right)_5 = \frac{40}{4}$$

$$g_{m1} = \sqrt{2 I_D \frac{W}{L} \mu_n C_{ox}}$$

$$= \sqrt{2(2.5) 90.9 \mu 200 \mu (250)}$$

$$= 4.8 \text{ mS}$$

$$g_{m5} = \sqrt{2 I_D \frac{W}{L} \mu_n C_{ox}} = \sqrt{2(5) 90.9 \mu (100 \mu) 10}$$

$$= 0.953 \text{ mS}$$

$$r_{o2} = \frac{1}{\lambda_n I_{D2}} = \frac{1}{0.1(0.5)(90.9 \mu) 2.5} = 88 \text{ k}\Omega$$

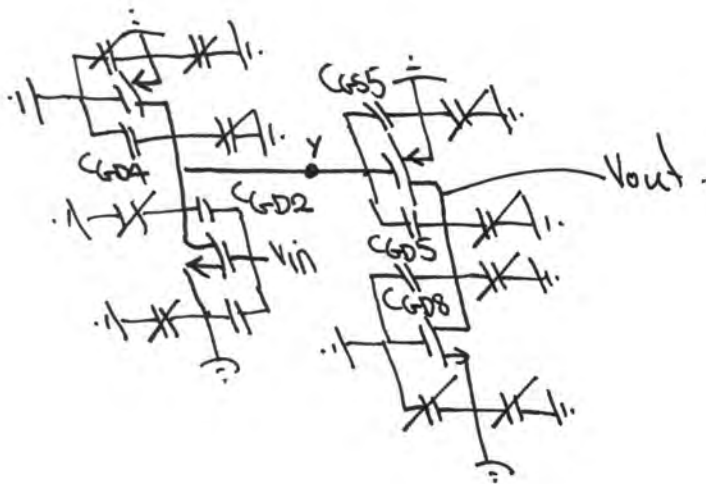
$$r_{o4} = \frac{1}{\lambda_p I_{D2}} = 44 \text{ k}\Omega$$

$$r_{o8} = \frac{1}{\lambda_n I_{D8}} = \frac{1}{0.1(0.5)(90.9 \mu)} = 22 \text{ k}\Omega$$

$$r_{o5} = \frac{1}{\lambda_p I_{D5}} = 11 \text{ k}\Omega$$

$$\Rightarrow A_V = (29.33 \text{ k})(7.3 \text{ k}) 0.953 \text{ m} \times 4.8 \text{ m} = 98.39$$

$$= 39.859 \text{ dB}$$



(e)

At node Y: $C_y = C_{GD4} + C_{GD2} \left(1 + \frac{1}{A_{v1}}\right) + C_{GS5} + C_{GD5} (1 + A_{v2})$

At node out: $C_{out} = C_{GD5} \left(1 + \frac{1}{A_{v2}}\right) + C_{GD8}$

$$\begin{aligned} \rightarrow C_y &= C_{GD4} + C_{GD2} + C_{GS5} + C_{GD5} (1 + g_{m5} (r_{o5} \parallel r_{o8})) \\ &= 4(0.2) + 250(0.2) + 40(4) \frac{2}{3}(12) + 40(0.2)(1 + 6.96) \\ &= 1.39 \text{ pF} \\ C_{out} &= C_{GD5} + C_{GD8} = 40(0.2) + 20(0.2) = 12 \text{ fF} \end{aligned}$$

$$R_y = r_{o2} \parallel r_{o4} = 29.33 \text{ k} \quad (\text{from (d)})$$

$$R_{out} = r_{o5} \parallel r_{o8} = 7.3 \text{ k} \quad (\text{from (d)})$$

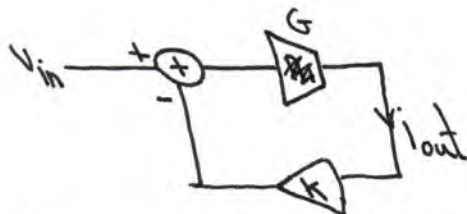
(f)

$$\therefore \omega f_{py} = \frac{1}{2\pi(29.33 \text{ k})(1.39 \text{ p})} = 3.9 \text{ MHz}$$

$$f_{pout} = \frac{1}{2\pi(7.3 \text{ k})(12 \text{ fF})} = 1.82 \text{ GHz}$$

f_{py} is the dominant pole

③ (a)



- (i) Gain of CE is negative and therefore the +ve terminal of amp. is used.
 (ii) Gain of EF is positive and therefore the -ve terminal of amp. is used.

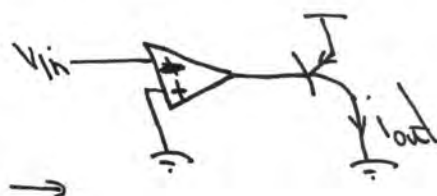
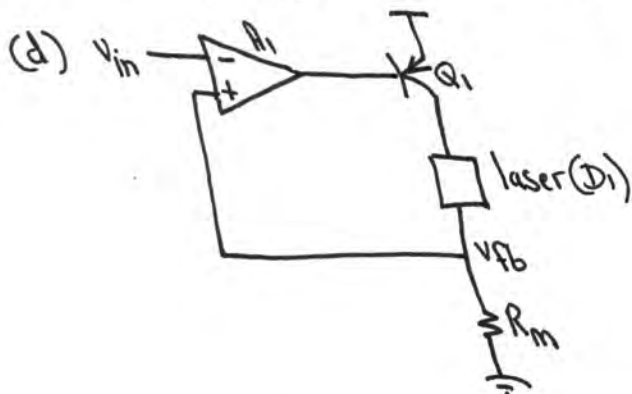
Both are -ve feedback.

(b) Voltage in, current out:

\therefore transconductance amplifier

(c) Ideal transconductance amplifier has high input impedance and high output impedance

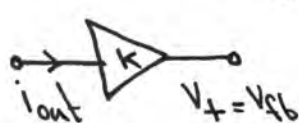
\therefore topology (i) common source is preferable as higher o/p impedance.



$$i_{out} = -V_{in}(A_1) \times -g_{m1}$$

$$\Rightarrow \frac{i_{out}}{V_{in}} = g_{m1} A_1 \text{ (open loop)}$$

(e) Feedback factor (K) takes i_{out} as its input and gives V_f as its o/p (V_{fb})



$$\therefore K = \frac{V_{fb}}{i_{out}} = \frac{i_{out} \times R_m}{i_{out}} = R_m.$$

$$\therefore \text{loop gain} = A_1 \times K = g_{m1} A_1 R_m$$

$$(f) \frac{i_{out}}{V_{in}} \text{ (closed-loop)} = \frac{G}{1+KG} = \frac{g_{m1} A_1}{1+g_{m1} A_1 R_m}$$

$$(g) \frac{V_{out}}{i_{out}} = R_{out} = R_m + r_{o1}$$

$$(h) R_{out} \text{ (closed-loop)} = (R_m + r_{o1})(1+g_{m1} A_1 R_m)$$

(i) * Note: This section cannot be solved as I_c and V_A are not given.

$$\rightarrow \text{if } I_c = 20\text{mA} \left\{ \begin{array}{l} g_m = \frac{I_c}{V_T} \text{ and } r_o = \frac{V_A}{I_c} \end{array} \right. \rightarrow \frac{i_{out}}{V_{in}} = \frac{1}{10} \text{ S}$$

$$\rightarrow R_{in}(CL) = 200 \text{ m}\Omega$$