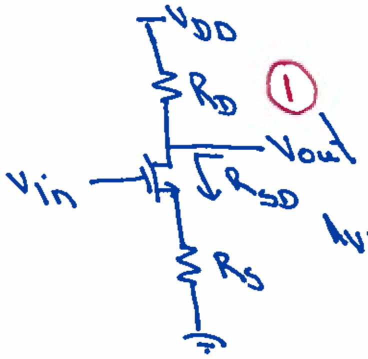
Analogue Electronics 2 (E2.2) 2014 — SOLUTIONS

Q1 – BOOKWORK/CALCULATION FOR NEW EXAMPLE

1 (a) Source degeneration is when a circuit element (eg. resistor) is connected between the source terminal (in a MOSFET) and the common node (i.e. GND for NMOS and VDD for PMOS). (1)



(1)

$$A_v = -\frac{R_D}{\frac{1}{g_m}+R_S}$$ (1)

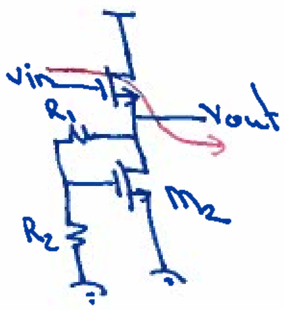
Advantages (compared to CS)

1. output resistance R_{SD} is increased from r_{o1} to $(g_{m1}r_{o1}R_S + R_S + r_{o1})$ (1/2)
2. Large signal operation is linearised (i.e. larger small signal region). (1/2)

Disadvantages (compared to CS)

1. reduced voltage (o/p) headroom. (1/2)
2. reduced voltage gain. (1/2)

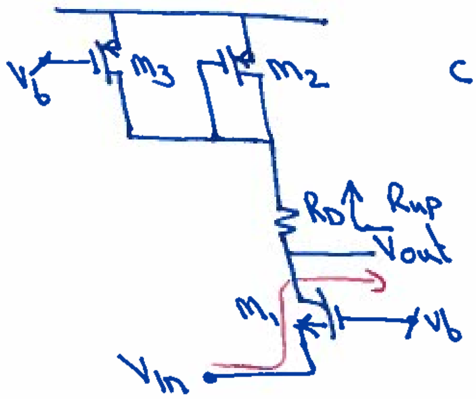
(b) (i)



Source follower (1) $\therefore A_v = \dfrac{R_S \| r_o}{\frac{1}{g_m} + R_S\| r_o}$ (1)

$R_S = r_{o2} \| (R_1+R_2)$ $\therefore A_v = \dfrac{r_{o1}\|r_{o2}\|(R_1+R_2)}{\frac{1}{g_{m1}} + r_{o1}\|r_{o2}\|(R_1+R_2)}$ (3)

(ii)



C.G. amplifier (1) $\therefore A_v = +g_m(r_o \| r_D)$ (1)

$A_v = +g_{m1}(r_{o1} \| R_{up})$ (1)

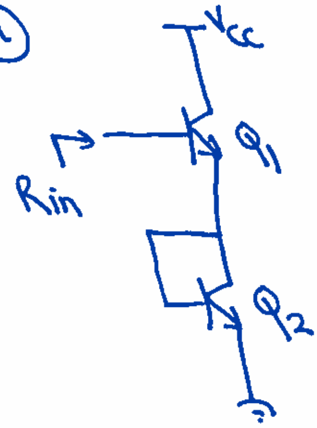
$R_{up} = R_D + \frac{1}{g_{m2}} \| r_{o2} \| r_{o3}$

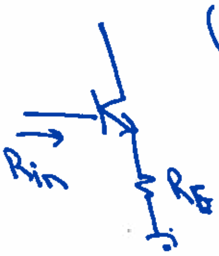
$\approx R_D + \frac{1}{g_{m2}} \| r_{o3}$

(assuming $\frac{1}{g_{m2}} \ll r_{o2}$) (1)

$\therefore A_v = +g_{m1}\left(r_{o1} \| \left(R_D + \frac{1}{g_{m2}} \| r_{o3}\right)\right)$ (1)

(c) (i)





(Use $R_{in} = r_\pi + (\beta+1)R_E$) (1)

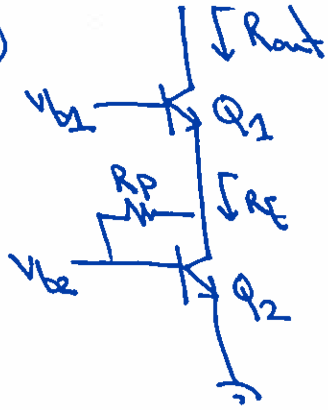
where $R_E = \frac{1}{g_{m2}} \| r_{\pi 2} \| r_{o2}$ (1)

$$\therefore R_{in} = r_{\pi 1} + (\beta+1)\left(\frac{1}{g_{m2}} \| r_{\pi 2} \| r_{o2}\right) \quad (1)$$

$$\approx r_{\pi 1} + (\beta+1)\left(\frac{1}{g_{m2}}\right) \quad (1)$$

(assuming $\frac{1}{g_{m2}} \ll r_{\pi 2} < r_{o2}$) (1)

(ii)



emitter degeneration. (1)

$$\therefore R_{out} = g_{m1} r_{o1}(R_E \| r_{\pi 1}) + (R_E \| r_{\pi 1}) + r_{o1} \quad (1)$$

where $R_E = R_p \| r_{o2}$ (1)

$$\therefore R_{out} = g_{m1} r_{o1}(R_p \| r_{o2} \| r_{\pi 1}) + (R_p \| r_{o2} \| r_{\pi 1}) + r_{o1}$$

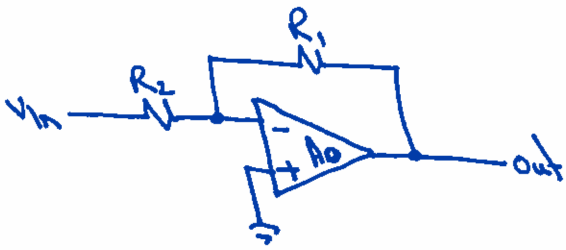
(2)

(d) (i) Output voltage range of a differential amplifier depends on the number of devices that are "stacked" which may limit the headroom. This is because all devices must remain in saturation (i.e. $V_{GS} - V_{TH} \leq V_{DS}$). For example, using cascode devices will limit the o/p range. (3)

(ii) The slew rate of an amplifier gives a measure on how fast the output can change. As this is fundamentally due to the output current charging and discharging the load, the SR is limited by the bias current (or "drivability"). (2)

(e) $A_V(\text{o.L}) = 106\,dB = 10^{\left(\frac{106}{20}\right)} = 200{,}000$ (1)

$A_V(\text{c.L}) = 200$

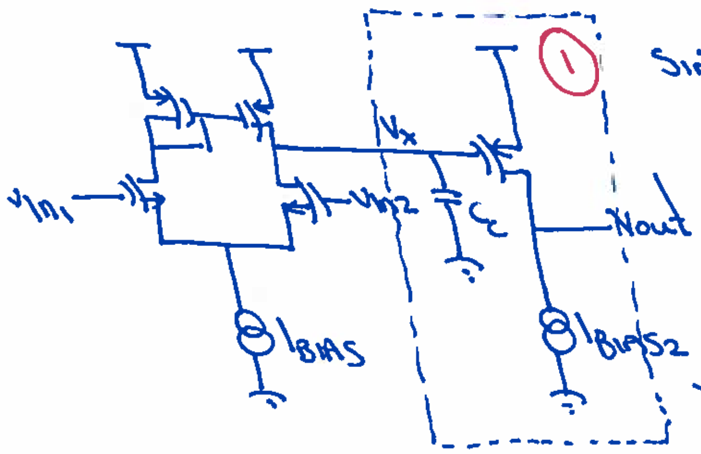


$$A_V(cL) = -\frac{R_1}{R_2}\left(1 - \frac{1}{A_0}\left(1+\frac{R_1}{R_2}\right)\right) \quad (2)$$

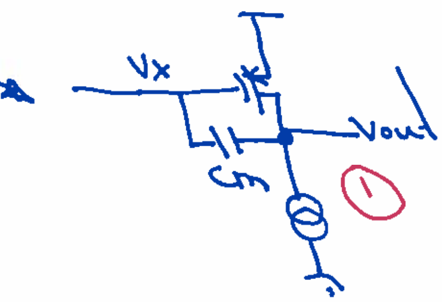
$$= -200\left(1 - \underbrace{\frac{1}{200000}(1+200)}_{\varepsilon}\right) \quad (1)$$

$\therefore\ \varepsilon = 0.1005\%$ (1)

(f) The typical procedure to improve phase margin in op-amp design is to add a compensation capacitance such as to limit the bandwidth so sufficient phase margin is achieved. By exploiting the Miller effect this capacitor value (2)

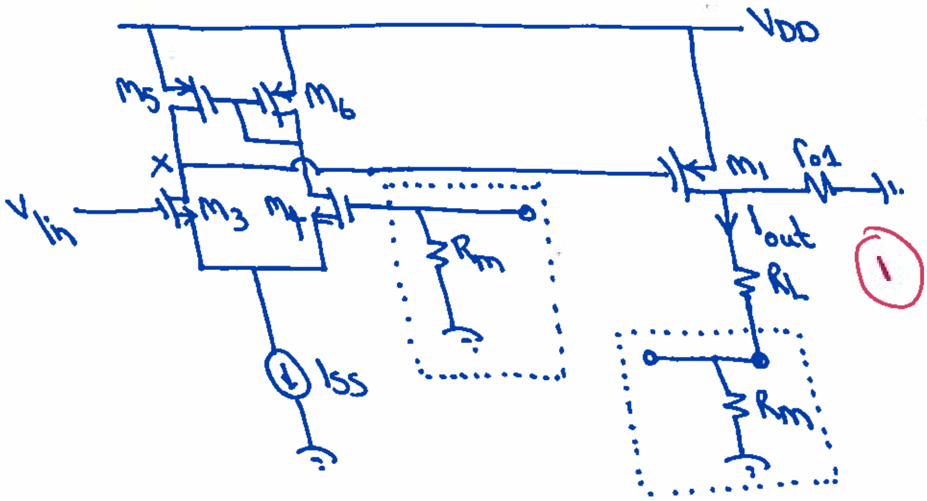


Since $\frac{V_{out}}{V_x} = A_{v2}$ (-ve amplification stage) a floating capacitor placed across this will appear as: $C_c = C_m(1+|A_{v2}|)$ at the input (1)



Q2- A NEW THEORETICAL APPLICATION

2 a (i) open loop gain → need to "break" the loop.

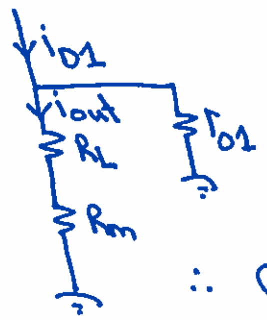


1

$$G_m(\text{open loop gain}) = \frac{I_{out}}{V_{in}} = \frac{V_x}{V_{in}}\cdot\frac{I_{out}}{V_x}$$ ½

$$\frac{V_x}{V_{in}} = -g_{m3}(r_{o3}\|r_{o5})$$ 1

$$\frac{I_{out}}{V_x} = \frac{I_{out}}{I_{D1}}\times\underbrace{\frac{I_{D1}}{V_x}}_{=g_{m1}}$$ ½

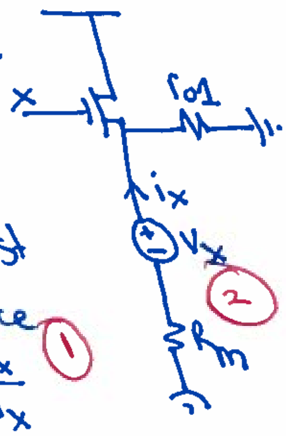


$$\frac{I_{out}}{I_{D1}} = \frac{r_{o1}}{r_{o1}+(R_L+R_m)}$$ 1

$$\therefore G_m = \frac{+g_{m1}g_{m3}r_{o1}(r_{o3}\|r_{o5})}{r_{o1}+R_L+R_m}$$ 1

(ii) R_{out} (OL)

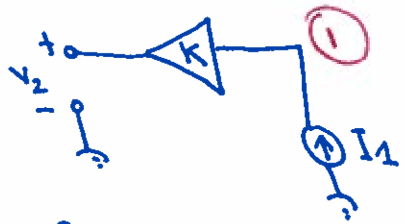
Need to replace R_L with a test voltage source and $\therefore R_{out} = \frac{V_x}{i_x}$ 1



2

$$R_{out}(OL) = r_{o1} + R_m$$ 2

(b) feedback factor K



$K = \frac{V_2}{I_1}$

$\therefore V_2 = I_{out} \cdot R_m$

$\therefore K = \frac{V_2}{I_{out}} = R_m$ 1

Error Signal = i/p − feedback $= V_{in} - R_m I_{out} = \varepsilon$ 1

$\therefore \varepsilon = \frac{V_{in}}{1 + G_m \cdot K}$ 1 $= \frac{V_{in}}{1 + R_m G_m}$ 1

(c) Transconductance amplifier ∴ voltage in, current out

∴ Ideally $R_{in} = \infty$

$R_{out} = \infty$

∴ −ve feedback will increase both R_{in} and R_{out}.

$\therefore R_{out} = (1 + \text{loopgain}) R_{out}(OL)$

$= (1 + R_m G_m)(R_m + r_{o1})$

(d) Loop gain $= (1 + R_m G_m)$ 1 so either R_m can be increased or $G_m(OL)$ 1

Ideally R_m should be kept small ∴ need to increase $G_m(OL)$ 1

$$G_m(OL) = \frac{g_{m1} g_{m3} r_{o1} (r_{o3} \| r_{o5})}{r_{o1} + R_L + R_m}$$

If bias current is increased eg. I_{SS}, this would increase g_{m1} but would also decrease r_{o3}, r_{o5}. 1

However can increase g_{m1} (and/or g_{m3}) by increasing its (W/L). 1

(e) Rm should ideally be kept small so majority of voltage drop is across the load. (2)

→ Given $V_{DD} = 10V$ and $I_{out} = 5mA$

However Rm also needs to provide sufficient bias voltage to sustain device M4 in saturation. (1)

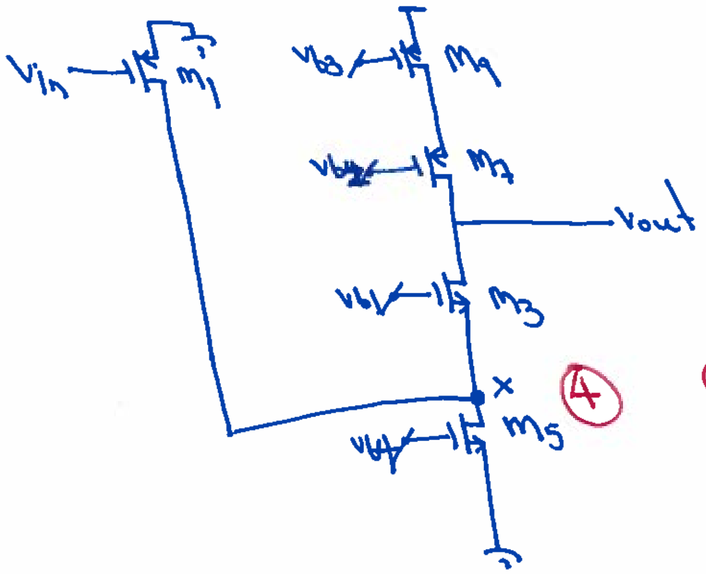
$\therefore V(R_m) > V_{GS}(sat)$

→ lets design for $V(R_m) = 0.5V$ (1)

$$\therefore R_m = \frac{0.5V}{5mA} = 100\,\Omega$$ (1)

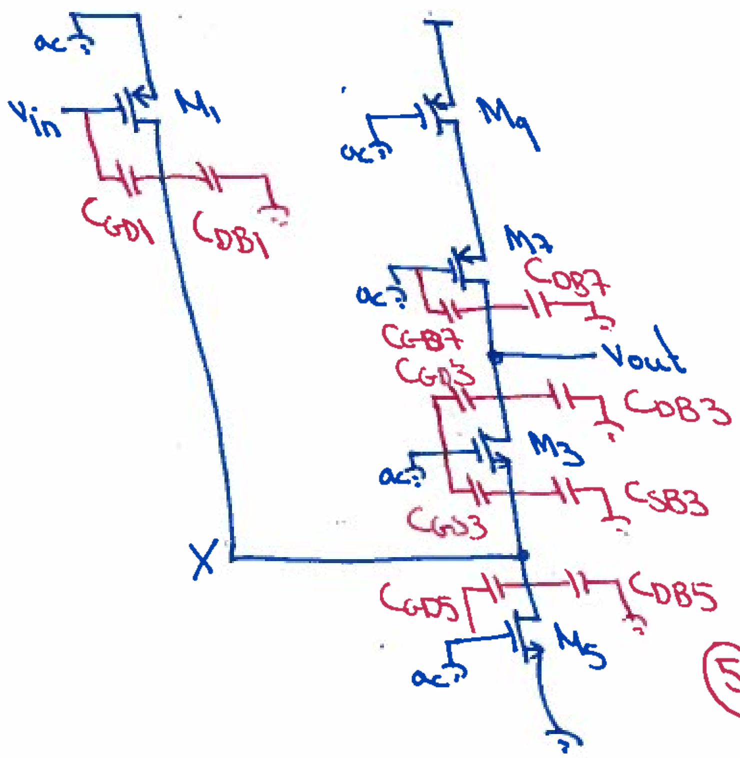
Q3—CALCULATION FOR A NEW EXAMPLE

③ ⓐ (i) Assuming perfect symmetry, sources of diff. pair can be connected to AC. ground. ①



④ (−½ mark each mistake)

(ii) with parasitics associated with nodes x and vout.



⑤ (−½ mark each mistake)

for (i) and (ii) OK as single diagram

(b) $A_v = -G_m R_{out}$

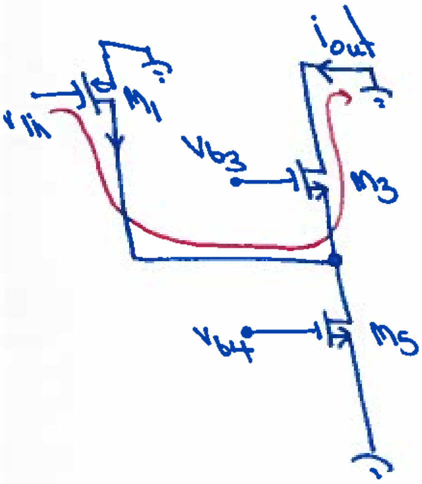
$R_{out} = R_{up} \| R_{down}$

$R_{up} = g_{m7} r_{o7} r_{o9} + r_{o7} + r_{o9}$

$R_{down} = g_{m3} r_{o3}(r_{o5} \| r_{o1}) + r_{o3} + r_{o5} \| r_{o1}$

$\therefore$ assuming $g_m r_o \gg 1$ 1

$R_{out} \approx g_{m7} r_{o7} r_{o9} + g_{m3} r_{o3}(r_{o5} \| r_{o1})$ 1



$i_{out} = I_{d3}$

$i_{d3} + i_{d5} = i_{d1}$ (assuming $\frac{1}{g_{m3}} \ll r_{o5}$, then $i_{d3} \approx i_{d1}$) 1

$i_{d1} = v_{in} g_{m1}$

$\therefore \frac{i_{out}}{v_{in}} = G_m = g_{m1}$ 1

$\therefore A_v = -g_{m1}\left[g_{m7} r_{o7} r_{o9} + g_{m3} r_{o3}(r_{o1} \| r_{o5})\right]$ 1

(c) $g_m = \sqrt{2\mu C_{ox} \frac{W}{L} I_D}$

$r_o = \frac{1}{\lambda I_D}$ 1

$g_{m1} = \sqrt{2(200\mu)(200)\,200\mu} = 4mS$

$g_{m3} = \sqrt{2(200\mu)(1)\,250\mu} = 316\mu S$

$g_{m7} = \sqrt{2(100\mu)\left(\frac{5}{2}\right)250\mu} = 353\mu S$

$r_{o7} = \frac{1}{0.2(250\mu)} = 20k\Omega$

$r_{o1} = \frac{1}{0.2(200\mu)} = 25k\Omega$ 3

$r_{o3} = \frac{1}{0.1(250\mu)} = 40k\Omega$

$r_{o5} = \frac{1}{0.1(50\mu)} = 200k\Omega$

$r_{o9} = \frac{1}{0.2(250\mu)} = 20k\Omega$

$\therefore A_v = -4m\big(353\mu(20k)(20k) + 316\mu(40k)(25k \| 200k)\big) = -1688 = 64.5\,dB$ 1

(d) $f_{px} = \dfrac{1}{2\pi R_x C_x}$ (1)

$f_{pout} = \dfrac{1}{2\pi R_{out} C_{out}}$ (1)

Since $C_{DB} = C_{SB} = \emptyset$

$C_x = C_{GD1}\left(1 + \dfrac{g_{m1}}{g_{m3}}\right) + \cancel{C_{DB1}} + C_{GD5} + \cancel{C_{DB5}} + C_{GS3} + \cancel{C_{SB3}}$ (1)

$R_x = r_{o1} \| r_{o5} \| \dfrac{1}{g_{m3}} \| r_{o3} \simeq \dfrac{1}{g_{m3}}$ (assuming $g_m \ll r_o$) (1)

$C_{out} = C_{GD7} + \cancel{C_{DB7}} + C_{GD3} + \cancel{C_{DB3}}$ (1)

R_{out} = as in (b).

(e) $C_x = \left[(0.2)100\left(1 + \dfrac{40}{3}\right) + (0.2)\left(\dfrac{4}{2}\right) + \dfrac{2}{3}(12)(2)(2)\right] fF$

$= 319fF$ (1)

$R_x = \dfrac{1}{g_{m3}} = \dfrac{1}{316\mu S} = 3.16k\Omega$ (1)

$R_{out} = 422k\Omega$ (from (c))

$C_{out} = 5(0.2) + 2(0.2) = 1.4fF$ (1)

$\therefore f_{px} = \dfrac{1}{2\pi(3.16k)319f} = 157.9MHz$ (1)

$f_{pout} = \dfrac{1}{2\pi(1.4fF)(422k)} = 269.4MHz$ (1)