Analog Electronic Circuits - Assignment 4

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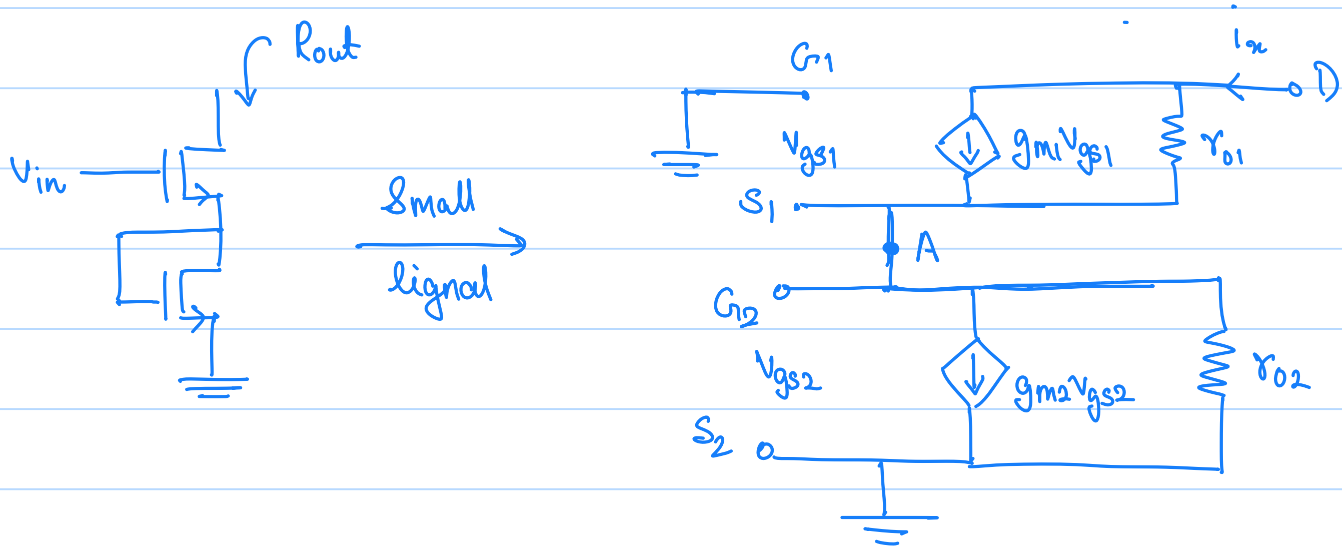
I)

1.



$$V_{gs1} = V_{G1} - V_{S1} = 0 - V_A$$

$$V_{gs1} = -V_A$$

$$i_x = g_{m1} V_{gs1} + \frac{V_x - V_A}{r_{o1}} \quad \text{(By KCL)}$$

$$\Rightarrow i_x = -g_{m1} V_A - \frac{V_x - V_A}{r_{o1}}$$

$$i_x + g_{m1} V_A = \frac{V_A - V_x}{r_{o1}}$$

$$\Rightarrow V_x = V_A - r_{o1} i_x - g_{m1} V_A = V_A (1 - g_{m1}) - r_{o1} i_x$$

In M2,

$$i_A = i_x \quad (\text{By KCL})$$

$$i_A = g_{m2} V_{gs2} + \frac{V_A}{r_{o2}}$$

$$V_{gs2} = V_{G2} - V_{S2} = V_A$$

$$\Rightarrow \quad i_x = g_{m2} V_A + \frac{V_A}{r_{o2}}$$

$$i_x = V_A \left(g_{m2} + \frac{1}{r_{o2}} \right)$$

$$V_A = \frac{i_x}{g_{m2} + \frac{1}{r_{o2}}} \quad \text{———} \quad (2)$$

Substituting (2) in (1),

$$V_x = \frac{i_x}{g_{m2} + \frac{1}{r_{o2}}} (1 - g_{m1}) - r_{o1} i_x$$

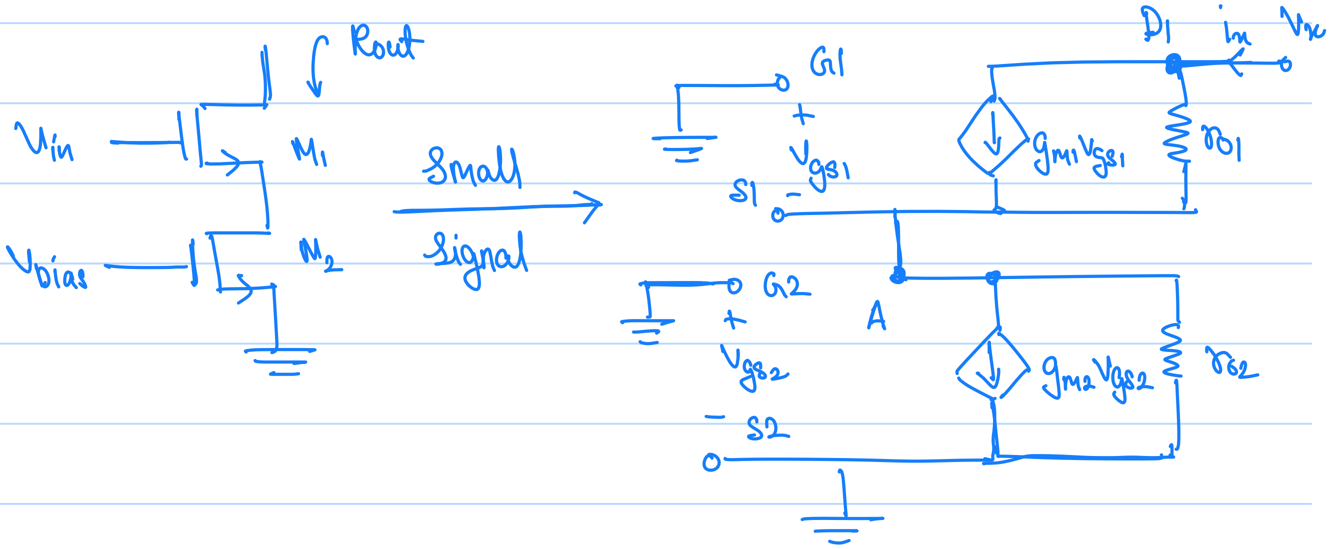
$$\frac{V_x}{i_x} = \frac{(1 - g_{m1}) r_{o2}}{g_{m2} r_{o2} + 1} - r_{o1}$$

$$\Rightarrow \quad Z_{out} = \frac{r_{o2} - g_{m1} r_{o2}}{1 + g_{m2} r_{o2}} - r_{o1}$$

With Body Effect,

$$Z_{out} = \frac{r_{o2} - (g_{m1} + g_{mb1}) r_{o2}}{1 + (g_{m2} + g_{mb2}) r_{o2}} - r_{o1}$$

2.



$$V_{gs2} = V_{G2} - V_{S2} = \underline{0}$$

$$\Rightarrow g_m V_{gs2} = \underline{0}$$

KCL at A, $\quad i_A = g_{m2}V_{gs2} + \dfrac{V_A}{r_{o2}}, \quad i_A = i_x$

$$\Rightarrow i_x = \frac{V_A}{r_{o2}} \Rightarrow V_A = i_x r_{o2} \quad \text{—— (1)}$$

$$V_{gs1} = V_{G1} - V_{S1} = -V_A$$

KCL at D_1, $\Rightarrow i_x = g_{m1}V_{gs1} + \dfrac{V_x - V_A}{r_{o1}}$

$$\Rightarrow i_x = -g_{m1}V_A + \frac{V_x - V_A}{r_{o1}}$$

Substituting (1),

$$i_x = -g_{m1}(i_x r_{o2}) + \frac{V_x - i_x r_{o2}}{r_{o1}}$$

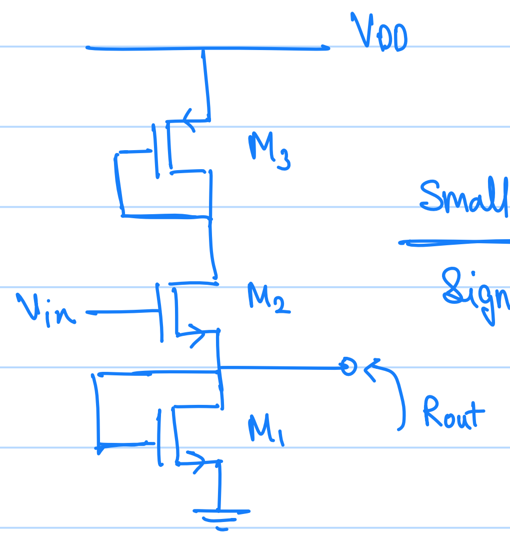
$$\Rightarrow i_x + g_{m1} i_x r_{o2} = \frac{V_x}{r_{o1}} - \frac{i_x r_{o2}}{r_{o1}}$$

$$= i_x r_{o1} + g_{m1} i_x r_{o1} r_{o2} + i_x r_{o2} = V_x$$

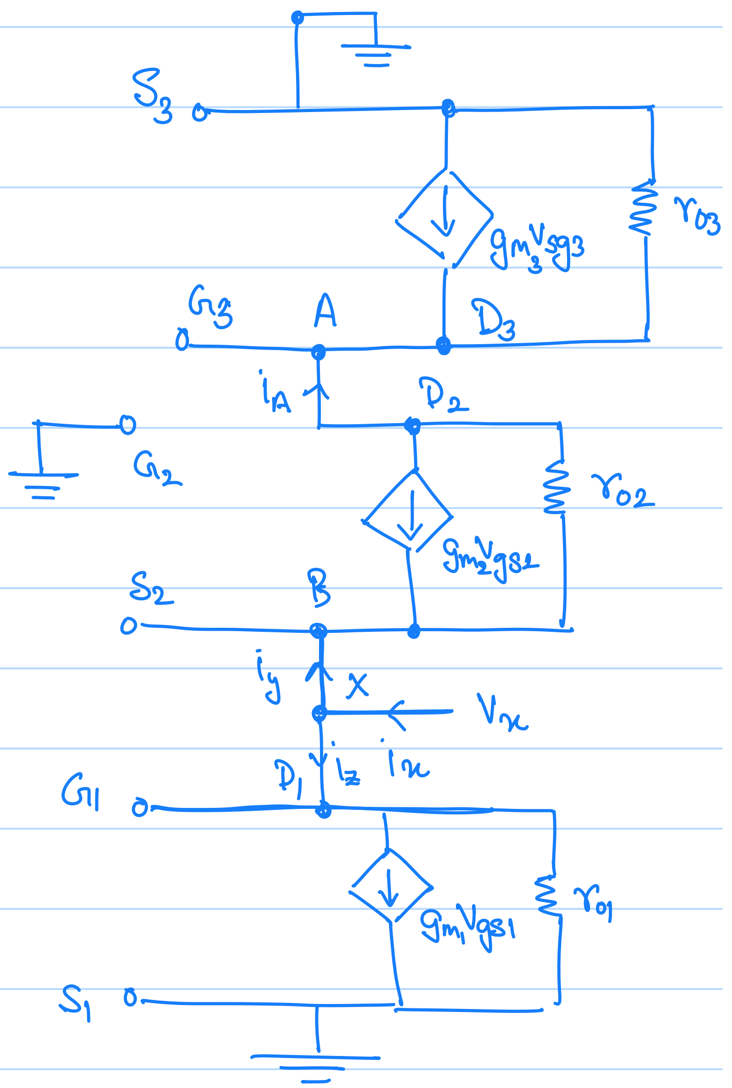
$$\Rightarrow \frac{V_x}{i_x} = r_{o1} + r_{o2} + g_{m1} r_{o1} r_{o2}$$

$$\Rightarrow Z_{out} = r_{o1} + r_{o2} + g_{m1} r_{o1} r_{o2} \xrightarrow[\text{Effect}]{\text{Body}} \boxed{Z_{out} = r_{o1} + r_{o2} + r_{o1} r_{o2} (g_{m1} + g_{mb1})}$$

3.



$\xrightarrow[\text{Signal}]{\text{Small}}$



KCL at X,

$$i_x = i_y + i_z$$

KCL at D_1,

$$i_z = g_{m1} V_{gs1} + \frac{V_x}{r_{o1}}$$

$$V_{gs1} = V_{G1} - V_{S1} = V_x$$

$$\Rightarrow i_z = g_{m1} V_x + \frac{V_x}{r_{o1}} = V_x \left(g_{m1} + \frac{1}{r_{o1}} \right) \quad \text{——— } (1)$$

KCL at B,

$$-i_y = g_{m2} V_{gs2} + \frac{V_A - V_x}{r_{o2}}$$

$$V_{gs2} = V_{G2} - V_{S2} = -V_x$$

$$\Rightarrow -i_y = -g_{m2} V_x + \frac{V_A - V_x}{r_{o2}}$$

$$= \quad i_y = g_{m2} V_x - \frac{V_A - V_x}{r_{o2}}$$

$$= \quad i_y = V_x \left(g_{m2} + \frac{1}{r_{o2}} \right) - \frac{V_A}{r_{o2}} \quad \text{———} \quad (2)$$

KCL at A,

$$i_A = i_y$$

$$-i_A = g_{m3} V_{sg3} - \frac{V_A}{r_{o3}}$$

$$V_{sg3} = V_{S3} - V_{G3} = 0 - V_A = -V_A$$

$$\Rightarrow -i_y = -g_{m3} V_A - \frac{V_A}{r_{o3}}$$

$$i_y = V_A \left(g_{m3} + \frac{1}{r_{o3}} \right)$$

$$V_A = \frac{i_y r_{o3}}{g_{m3} r_{o3} + 1} \quad \text{———} \quad (3)$$

Substituting (3) in (2),

$$i_y = V_x \left(g_{m2} + \frac{1}{r_{o2}} \right) - \left(\frac{1}{r_{o2}} \right) \left(\frac{i_y r_{o3}}{g_{m3} r_{o3} + 1} \right)$$

$$i_y + i_y\left(\frac{r_{o3}}{(\cancel{r_{o2}})(g_{m3}r_{o3}+1)}\right) = V_x\left(\frac{g_{m2}r_{o2}+1}{\cancel{r_{o2}}}\right)$$

$$i_y\left(1+\frac{r_{o3}}{g_{m3}r_{o3}+1}\right) = V_x\left(g_{m2}r_{o2}+1\right)$$

$$i_y = \frac{V_x(g_{m2}r_{o2}+1)(g_{m3}r_{o3}+1)}{g_{m3}r_{o3}+r_{o3}+1} \qquad (4)$$

(1) + (4),

$$i_y + i_z = V_x\left(\frac{(g_{m2}r_{o2}+1)(g_{m3}r_{o3}+1)}{g_{m3}r_{o3}+r_{o3}+1}\right) + V_x\left(g_{m1}+\frac{1}{r_{o1}}\right)$$

$$i_x = V_x\left(\frac{(g_{m2}r_{o2}+1)(g_{m3}r_{o3}+1)}{g_{m3}r_{o3}+r_{o3}+1} + \frac{g_{m1}r_{o1}+1}{r_{o1}}\right)$$

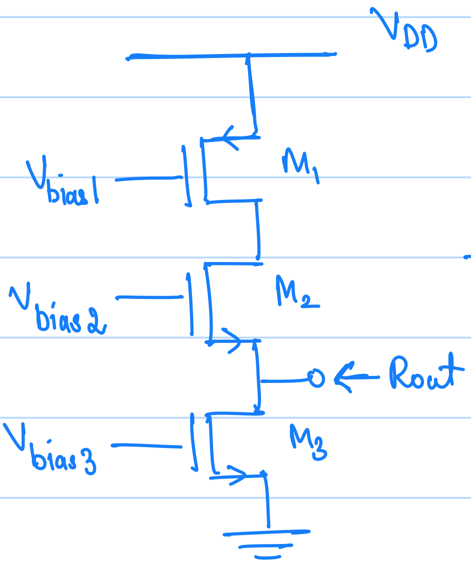
$$= V_x\left(\frac{(g_{m2}r_{o2}+1)(g_{m3}r_{o3}+1)r_{o1} + (g_{m1}r_{o1}+1)(g_{m3}r_{o3}+r_{o3}+1)}{(r_{o1})(g_{m3}r_{o3}+r_{o3}+1)}\right)$$

$$\Rightarrow \quad \frac{V_x}{i_x} = Z_{out} = \frac{(r_{o1})(g_{m3}r_{o3}+r_{o3}+1)}{r_{o1}(g_{m2}r_{o2}+1)(g_{m3}r_{o3}+1) + (g_{m1}r_{o1}+1)(g_{m3}r_{o3}+r_{o3}+1)}$$

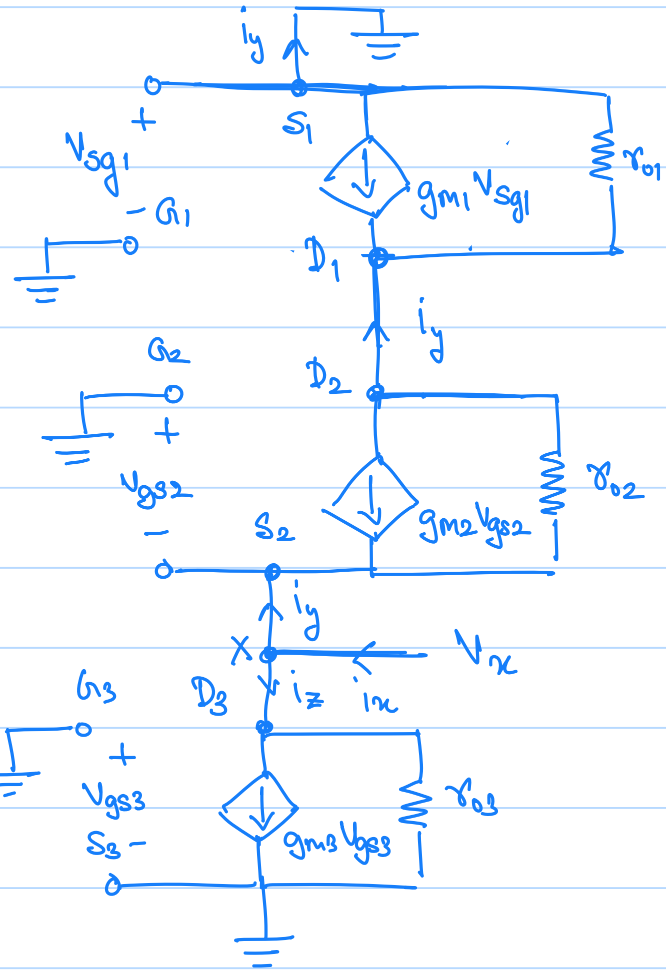
With Body Effect,

$$\boxed{Z_{out} = \frac{r_{o1}\left((g_{m3}+g_{mb3})r_{o3}+r_{o3}+1\right)}{r_{o1}\left(r_{o2}(g_{m2}+g_{mb2})+1\right)\left(r_{o3}(g_{m3}+g_{mb3})+1 + \left(r_{o1}(g_{m1}+g_{mb1})+1\right)\left(r_{o3}(g_{m3}+g_{mb3}+1)+1\right)}}$$

4.



Small signal



KCL on X,

$$i_x = i_y + i_z \quad \text{——} \quad (1)$$

KCL on D_3

$$i_z = g_{m3} V_{gs3} + \frac{V_x}{r_{o3}}$$

$$V_{gs3} = V_{G3} - V_{S3} = \underline{0}$$

$$\Rightarrow \quad i_z = \frac{V_x}{r_{o3}} \quad \text{——} \quad (2)$$

KCL on S_2,

$$-i_y = g_{m2} V_{gs2} + \frac{V_{D2} - V_x}{r_{o2}}$$

$$V_{gs2} = V_{G2} - V_{S2} = -V_x$$

$$-i_y = -g_{m2} V_x + \frac{V_{D2}}{r_{o2}} - \frac{V_x}{r_{o2}}$$

$$i_y = v_x\left(g_{m2} + \frac{1}{r_{o2}}\right) - \frac{v_{D2}}{r_{o2}} \quad \text{———} \quad (3)$$

KCL on D_2,

$$-i_y = g_{m1}v_{sg1} - \frac{v_{D2}}{r_{o1}}$$

$$v_{sg1} = v_{s1} - v_{a1} = \underline{\underline{0}}$$

$$\Rightarrow i_y = \frac{v_{D2}}{r_{o1}}$$

$$\Rightarrow v_{D2} = i_y r_{o1} \quad \text{———} \quad (4)$$

Substituting (4) in (3),

$$i_y = v_x\left(g_{m2} + \frac{1}{r_{o2}}\right) - i_y\frac{r_{o1}}{r_{o2}}$$

$$i_y\left(1 + \frac{r_{o1}}{r_{o2}}\right) = v_x\left(g_{m2} + \frac{1}{r_{o2}}\right)$$

$$i_y\left(\frac{r_{o2} + r_{o1}}{\cancel{r_{o2}}}\right) = v_x\left(\frac{g_{m2}r_{o2} + 1}{\cancel{r_{o2}}}\right)$$

$$i_y = v_x\left(\frac{g_{m2}r_{o2} + 1}{r_{o2} + r_{o1}}\right) \quad \text{———} \quad (5)$$

(5) + (2)

$$i_z + i_y = \frac{v_x}{r_{o3}} + v_x\left(\frac{g_{m2}r_{o2} + 1}{r_{o2} + r_{o1}}\right)$$

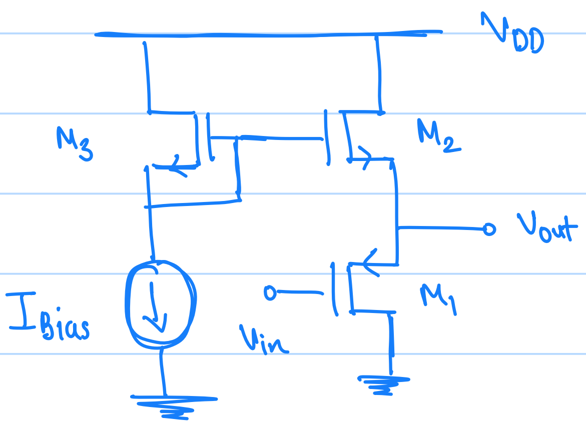
$$i_x = V_x\left(\frac{1}{r_{o3}} + \frac{g_{m2}r_{o2}+1}{r_{o2}+r_{o1}}\right)$$

$$\frac{V_x}{i_x} = \frac{1}{\frac{1}{r_{o3}} + \frac{g_{m2}r_{o2}+1}{r_{o2}+r_{o1}}}$$

$$Z_{out} = \frac{(r_{o2}+r_{o1})(r_{o3})}{r_{o2}+r_{o1}+r_{o3}+g_{m2}r_{o2}r_{o3}} \xrightarrow[\text{Effect}]{\text{Body}} \boxed{Z_{out} = \frac{(r_{o2}+r_{o1})(r_{o3})}{r_{o1}+r_{o2}+r_{o3}+(g_{m2}+g_{mb2})r_{o2}r_{o3}}}$$

II)

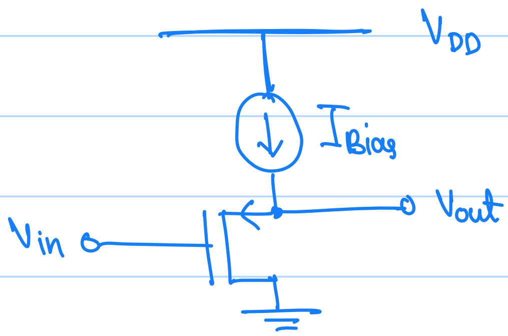
A PMOS CS Amplifier w/ NMOS current mirror as load :



In a current mirror circuit, the current across M_2 is equal to the current across M_3 (assuming M_2 and M_3 are identical and ideal)
They have a high output impedance and therefore work as good current sources.

Due to this, when a current mirror is used as a load, it provides high load impedance, which greatly increases the gain of a CS amplifier (III to using a PMOS as an active load), as the gain of a CS amplifier increases with increasing load impedance.

An equivalent circuit for the above amplifier is,



Comparing a PMOS CS w/ NMOS Current Mirror, with a NMOS CS w/ PMOS Current Mirror,

- NMOS MOSFETs usually have a higher carrier mobility than PMOS, due to material properties.

- Therefore, NMOS will have higher g_m ($g_m = \mu_n C_{ox} \frac{W}{L} V_{DSAT}$ $\Rightarrow$ $g_m \propto \mu_n$) than PMOS, which leads to a better gain.

- Since PMOS has a lower g_m, their output impedance will be higher. This makes them a better load due to higher load impedance.

$\Rightarrow$ NMOS is better suited for the gain stage and PMOS is better suited for the load stage.

$\therefore$ A NMOS CS Amplifier w/ PMOS Current Mirror would have better gain.

III) Amplifier Design 1

(a) Given I_D vs V_{GS} simulations and Aspect Ratio $\frac{W}{L} = \frac{1\mu}{1\mu}$, we need to find V_T and $\mu_n C_{ox}$

An example I_D vs V_{GS} simulation for a TSMC 180nm type MOSFET, with $V_{DD} = 1V$ and $R_L = 5k\Omega$ is given below

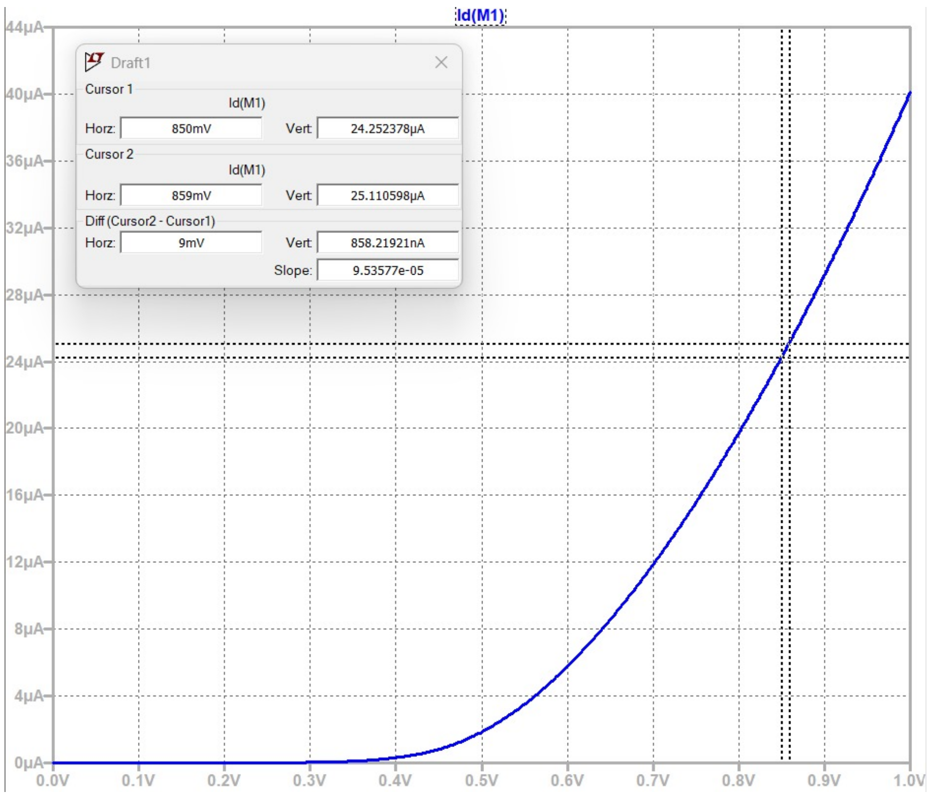


Figure 1: LTSpice Plot of I_D vs V_{GS}

We can use the Transconductance method to find $V_{TH}, \mu_n C_{ox}$.

(a) Find a point where the MOSFET is in Saturation Region. (Assume)

(b) Find I_D and V_{GS} at the point. Verify that the point is in Saturation by checking if $V_{DS} > V_{GS}$. ($V_{DS} > V_{GS} \implies V_{DS} > V_{GS} - V_{TH}$ which implies Saturation, $V_{DS} = V_{DD} - I_D R_L$)

(c) Find the Slope at the point. This gives us Transconductance g_m.

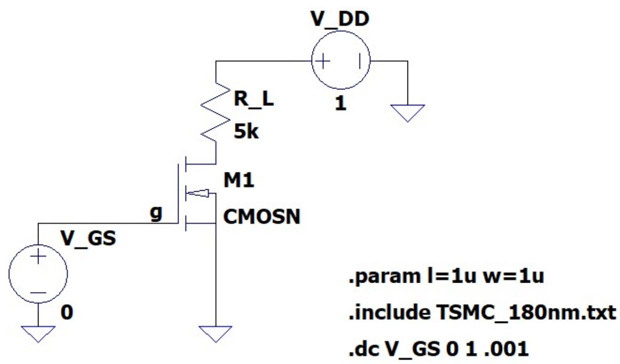


Figure 2: LTSpice Schematic

(d) We know that,

$$g_m = \sqrt{2K_n I_D} \tag{1}$$

$$g_m = \frac{2I_D}{V_{GS} - V_{TH}} \tag{2}$$

The values of I_D, V_{GS} and g_m are known. So using Equations (1) and (2), we can find K_n and V_{TH}.

For the given plot,

$$\begin{aligned}
g_m &= Slope = 9.53 \times 10^{-5} \\
g_m &= \sqrt{2K_n I_D} \\
\implies K_n &= \frac{g_m^2}{2I_D} \\
&= \frac{(9.53 \times 10^{-5})^2}{2(24.25 \times 10^{-6})} \\
\therefore K_n &= 1.872 \times 10^{-4} A/V^2
\end{aligned}$$

$$\begin{aligned} g_m &= \frac{2I_D}{V_{GS} - V_{TH}} \\ \implies V_{TH} &= V_{GS} - \frac{2I_D}{g_m} \\ &= 0.85 - \frac{2 \times 24.25 \times 10^{-6}}{9.53 \times 10^{-6}} = 0.85 - 0.508 \\ \therefore V_{TH} &= 0.342\ V \end{aligned}$$

(b) Using the parameters given in the question, the Schematic and the obtained plots are as follows

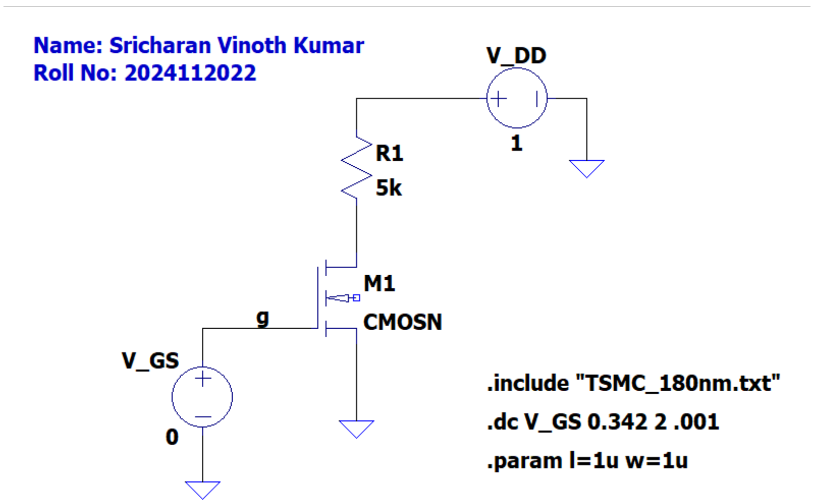


Figure 3: LTSpice Schematic

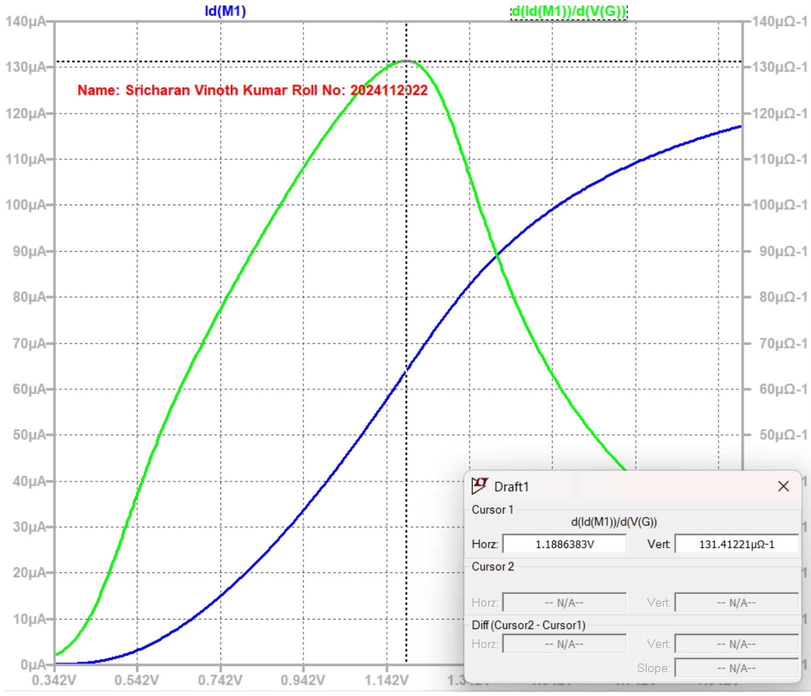


Figure 4: Plot of g_m(Green) and I_D(Blue) vs V_{GS}

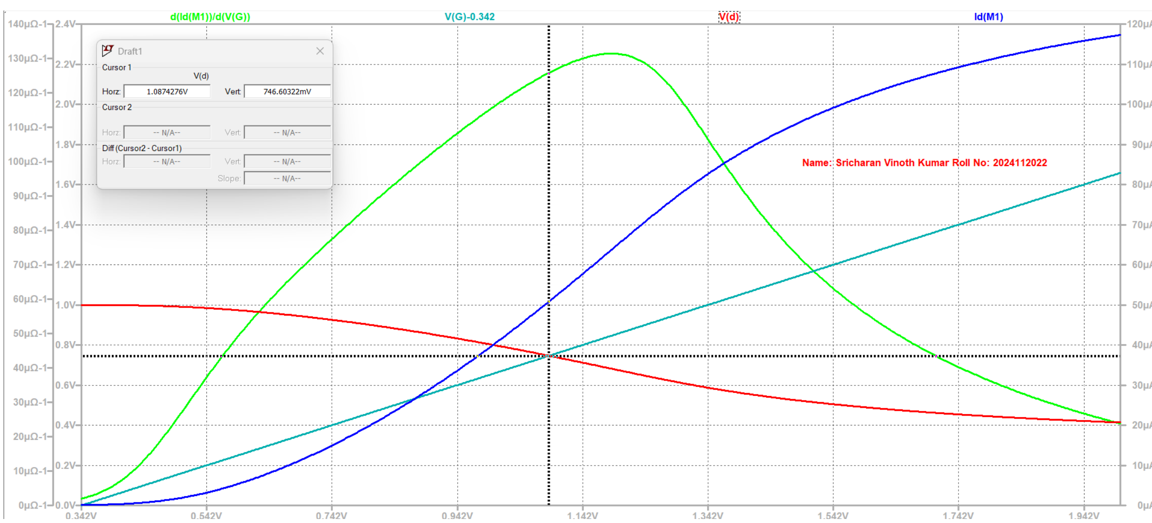


Figure 5: Plot of V_{OV}(Dark Green), V_{DS}(Red), g_m(Green) and I_D(Blue) vs V_{GS}

From the plot in Figure 4, we get Maximum g_m as $131.4122\Omega^{-1}$. In Figure 5, the transisiton from Saturation to Linear region of operation is highlighted by the cursor, at $V_{GS} \approx 1V$, which is the crossover point for the plots of V_{OV} and V_{DS}. This is close to the maximum value of g_m, implying that the maximum

Transconductance is achieved as we leave Saturation. However Transconductance also decreases as we go deeper into Linear.

(c) For $\frac{W}{L} = \frac{5\mu}{1\mu}$, the obtained plots are,

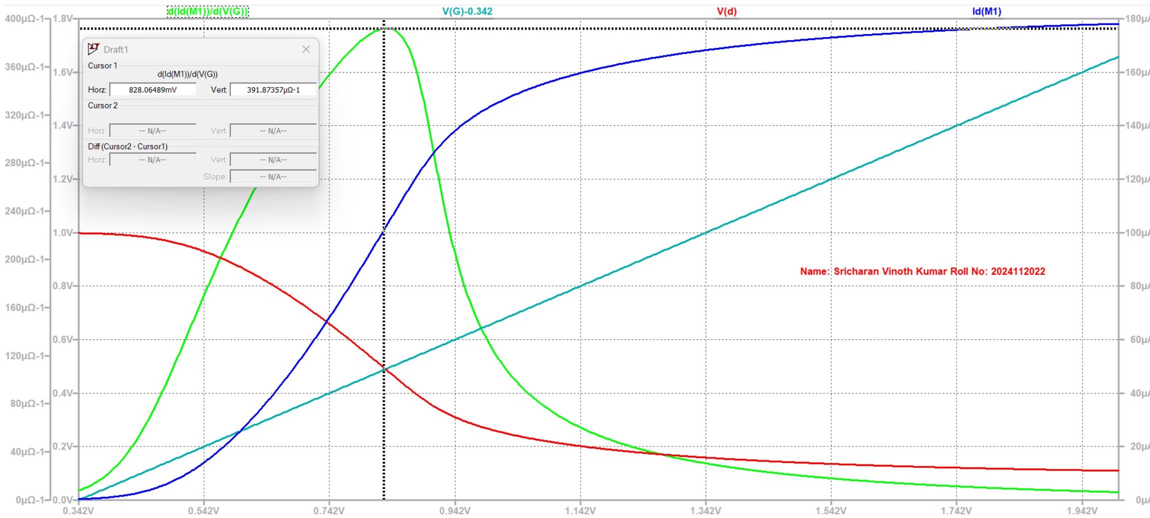


Figure 6: Plot of V_{OV}(Dark Green), V_{DS}(Red), g_m(Green) and I_D(Blue) vs V_{GS}

We see that with increase in W/L, our maximum Transconductance has increased to $\approx 391.873\Omega^{-1}$. Also our previous result, that maximum Transconductance is achieved at the edge of Saturation, is much more prominent in this graph, as the maximum g_m point coincides with the crossover point between V_{OV} and V_{DS}.
We know that the gain of the Amplifier is given by,

$$A_v = -g_m R_L$$

Since we see an increase in Transconductance, we can conclude that **the gain will increase.**
The increase in the Aspect Ratio, will also cause the size of the MOSFET to increase, which increases the parasitic capacitances, as,

$$C = \epsilon_o A/d \implies C \propto A \propto Width$$

Due to an increase in the capacitances, **the bandwidth will decrease.**
Also, the drain current of the Amplifier increases, which increases Power Con-

sumption. The increase in drain current is due to,

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS} - V_{TH})^2 \quad \text{,(assuming Saturation)}$$
$$\implies I_D \propto W/L$$

The output swing of the Amplifier will be increased due to the Harmonic Component, which is,

$$V_{omaxHD} = \frac{I_D R_L HD}{12.5} \implies V_{omaxHD} \propto I_D \propto W/L$$

However this increase is not definite, since the output swing also depends on the rail voltages and the bias voltage, since,

$$Output\ Swing = min\{V_{omaxHD}, V_{omax1}, V_{omax2}\}$$
$$V_{omax1} = V_{DD} - V_{DSQ}$$
$$V_{omax2} = V_{DSQ} - V_{SS}$$

Therefore it is possible that this increase is negated.
In conclusion, the trade-offs for increasing gain using W/L is a decreased bandwidth and higher power consumption.

IV) Amplifier Design 2

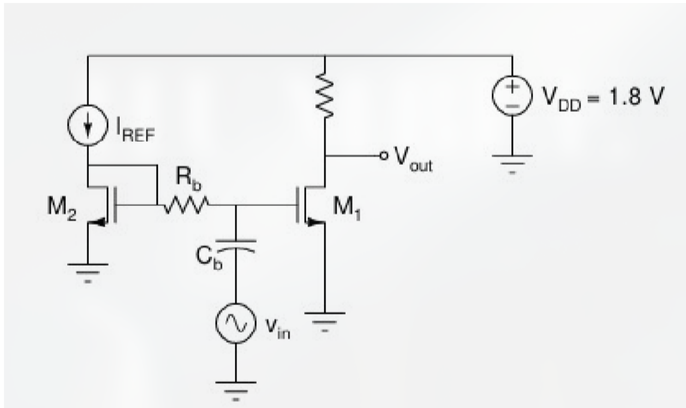
(a) Given Parameters:

- $V_{DD} = 1.8V$
- Resistive Load $= 15k\Omega$
- Overdrive Voltage of Input MOSFET $= 200mV$
- Minimum input frequency $= 100Hz$

Assuming M1 and M2 are the same type of MOSFET as the one used in III, (TSMC 180nm), we will use the following parameters,

- $\mu_n C_{ox} = 1.872 \times 10^{-4}\ A/V^2$
- $V_{TH} = 0.342V$

Schematic:



We know that $V_{TH} = 0.342V$, so

$$V_{ov} = V_{GS} - V_{TH} = 0.2V \implies V_{GS} = 0.542V \tag{3}$$

From the Schematic, we can infer the following:

1. The Amplifier is in Common Source Topology.
2. The Gain stage MOSFET (M_1) is part of a Current Mirror Circuit with M_2

Using Result 2, we can state that,

$$I_{DS1} = d \times I_{REF} \tag{4}$$

Where $d = \frac{(W/L)_1}{(W/L)_2}$. To minimize Power consumption, we need to minimize both I_{REF} and I_{DS1}. Ideally, we can have $d > 1$, since I_{DS1} depends on our gain so is independent of this equation, and Power $= V_{DD}(I_{REF} + _{DS1})$, so having $d > 1$, reduces I_{REF}, reducing power consumption. But practically this is not feasible since this decreases the accuracy of the current mirrors and complicates our design and layout. Therefore it is preferred to have $d = 1$,ie, Mirroring. So, we will take M_1 and M_2 to have identical sizes.
For a Common Source Amplifier, we know that the Gain is,

$$\begin{aligned} A_v &= -g_{m1} \times (Load\ Impedance) \\ \implies A_v &= -g_{m1} \times (R_L \parallel r_o) \\ \therefore\ A_v &\approx -g_{m1} R_L \end{aligned} \tag{5}$$

We know that the minimum gain must be 10. To minimize power consumption, we take gain as 10. Using that equation, we get,

$$g_{m1} = \frac{A_v}{R_L} = \frac{10}{R_L} = \frac{10}{15000} = 6.67 \times 10^{-4}\ \Omega^{-1} \tag{6}$$

We know that $g_{m1} = \frac{2I_{DS1}}{V_{ov}}$. Substituting that in Equation 5, we get,

$$\begin{aligned} \frac{2I_D}{V_{ov}} &= \frac{10}{R_L} \\ \implies I_{DS1} &= \frac{5V_{ov}}{R_L} \\ \implies I_{DS1} &= \frac{5 \times 0.2}{15000} \\ \implies I_{DS1} &\approx 6.67 \times 10^{-5} A \approx 66.7 \mu A \\ \therefore I_{REF} &\approx 67 \mu A \end{aligned} \tag{7}$$

We also know that $g_{m1} = \sqrt{2K_n I_{DS1}}$. Using this, we get,

$$K_n = \frac{g_{m1}^2}{2I_{DS1}} = \frac{(6.67 \times 10^{-4})^2}{2 \times (6.67 \times 10^{-5})} \approx 3.33 \times 10^{-3} \tag{8}$$

From III we know that $\mu_n C_{ox} = 1.872 \times 10^{-4} A/V^2$. Since $K_n = \mu_n C_{ox} \frac{W}{L}$,

$$\implies \frac{W}{L} = \frac{K_n}{\mu_n C_{ox}} = \frac{3.33 \times 10^{-3}}{1.872 \times 10^{-4}} \approx 17.58 \tag{9}$$

It is given that the minimum frequency that can be applied at the input is $100Hz$. Since the R_B, C_B network acts like a High Pass filter, the cutoff frequency of the filter must be 2-3 times lower (conventional) than $100Hz$ to allow it to pass through with minimal attenuation.

Take the cutoff frequency to be $100/3 \approx 66Hz$. Also we take R_B to be an arbitarily high value, say $100k\Omega$, since this makes the impedance across the filter less dominated by the capacitance ($z = R_B \parallel \frac{1}{wC} \approx R_B$), since that component is frequency dependent, resulting in an frequency independent impedance across the filter (for any frequency above the cutoff frequency that is). To find corresponding C_B,

$$\begin{aligned} f_{-3db} &= \frac{1}{2\pi R_B C_B} \\ \implies C_B &= \frac{1}{2\pi R_B f_{-3db}} = \frac{1}{2\pi \times 10^5 \times 66} \approx 24nF \end{aligned} \tag{10}$$

Therefore, we get the parameters for our Amplifier design as,

- $\frac{(W/L)_1}{(W/L)_2} = 1 \implies M_1$ and M_2 have identical dimensions.
- $I_{REF} = 66.7\mu A$
- $\frac{W}{L} = 17.58$
- $R_B = 100k\Omega$
- $C_B = 24nF$

The Power Consumption will be,

$$P = V_{DD}(I_{REF} + I_{DS}) = 1.8V \times 2 \times 66.7\mu A \approx 0.24\ mW \tag{11}$$

To get the effects of increasing each of these factors by 10%:
Increasing I_{REF} by 10%,

$$\begin{aligned}
I_{REF} &\rightarrow (1.1)I_{REF} \implies (1.1)I_{DS1} \\
g_m &= \sqrt{2K_n I_{DS1}} \\
g_{m'} &= \sqrt{2K_n I_{DS1}(1.1)} \\
\implies \Delta g_m &= \sqrt{2K_n I_{DS1}}(\sqrt{1.1} - 1) = g_m(\sqrt{1.1} - 1) \\
\implies \frac{\Delta g_m}{g_m} &= \sqrt{1.1} - 1 \approx 0.0488 = 4.88\% \\
A_v &= -g_m R_L \\
\Delta A_v &= -\Delta g_m R_L \\
\implies \frac{\Delta A_v}{A_v} &= \frac{\Delta g_m}{g_m} = 4.88\% \\
g_m &= K_n V_{ov} \\
\implies \frac{\Delta g_m}{g_m} &= \frac{\Delta V_{GS}}{V_{ov}} = 0.0488 \\
\implies \Delta V_{GS} &= 0.0488 \times 0.2 = 0.00976 \\
\implies \frac{\Delta V_{GS}}{V_{GS}} &= \frac{0.00976}{0.542} \approx 0.018 = 1.8\% \\
P &= 2V_{DD}I_{REF} \\
\frac{\Delta P}{P} &= \frac{\Delta I_{REF}}{I_{REF}} = 10\%
\end{aligned}$$

Summary of Changes,

- g_m increases by 4.88%
- Gain increases by 4.88%
- V_{GS} increases by 1.8%
- Power consumption increases by 10%

Increasing $\frac{W}{L}$ by 10%,

$$\begin{aligned}
\frac{W}{L} &\rightarrow (1.1)\frac{W}{L} \\
K_n &= \mu_n C_{ox}\frac{W}{L} \\
\implies \frac{\Delta K_n}{K_n} &= \frac{\Delta(W/L)}{(W/L)} = 10\% \\
g_m &= \sqrt{2K_n I_{DS1}} \\
\Delta g_m &= \sqrt{2K_n I_{DS1}}(\sqrt{1.1} - 1) = 0.048 g_m \\
\frac{\Delta g_m}{g_m} &= 0.0488 = 4.88\% \\
\implies \frac{\Delta A_v}{A_v} &= 4.88\% \\
I_{DS1} &= \frac{1}{2}K_n V_{ov}^2 \implies V_{ov} = \sqrt{\frac{2I_{DS1}}{K_n}} \\
\implies V_{ov}\prime &= V_{ov}\sqrt{\frac{1}{1.1}} \approx 0.953 \times 0.2 \approx 0.19V \\
\implies V_{GS}\prime &= V_{TH} + V_{ov}\prime = 0.532 \implies \Delta V_{GS} = 0.01 \\
\implies \frac{\Delta V_{GS}}{V_{GS}} &= \frac{-0.01}{0.532} \approx -1.88\%
\end{aligned}$$

Summary of Changes

- g_m increases by 4.88%
- Gain increases by 4.88%
- V_{GS} decreases by 1.88%
- Power Consumption is unaffected

Since the values of R_B and C_B matter only for the high pass filter as described in the design procedure, only the effective bandwidth of the Amplifier will be affected. All the other parameters will remain unchanged. We know that,

$$f_{-3db} = \frac{1}{2\pi R_B C_B} \tag{12}$$

By this equation, we can state that the effects of changes in R_B and C_B will be identical.

Increasing R_B by 10%,

$$\begin{aligned} f_{-3db}\prime &= \frac{1}{2\pi R_B(1.1)C_B} = \frac{1}{1.1} f_{-3db} = \frac{1}{1.1} 66 = 60Hz \\ \frac{\Delta f_{-3db}}{f_{-3db}} &= (60 - 66)/(66) = \frac{-1}{11} \approx -0.091 = -9.1\% \end{aligned} \tag{13}$$

Increasing R_B, decreases the cutoff frequency by 9.1%. We can state that increasing C_B by 10% will also have the same effect.

(b) Simulations:

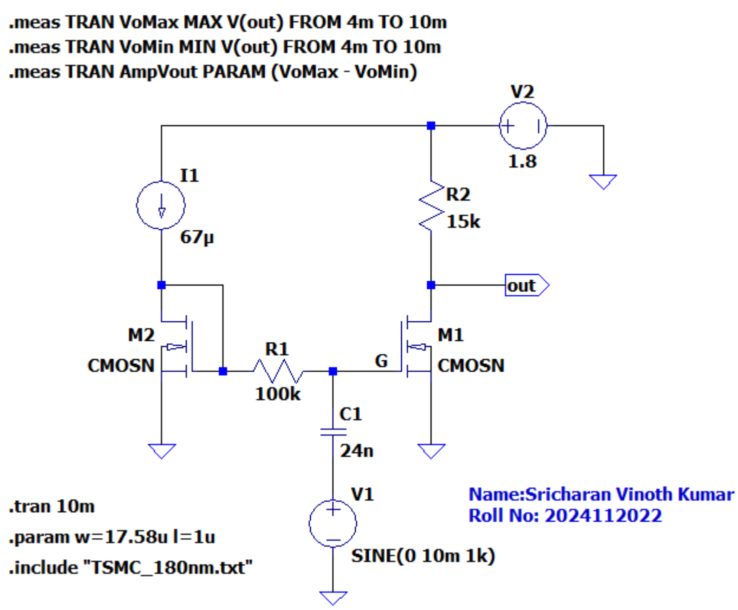


Figure 7: LTSpice Schematic with the designed parameters

Input signal: $10\sin(2\pi(1000)t)mV$

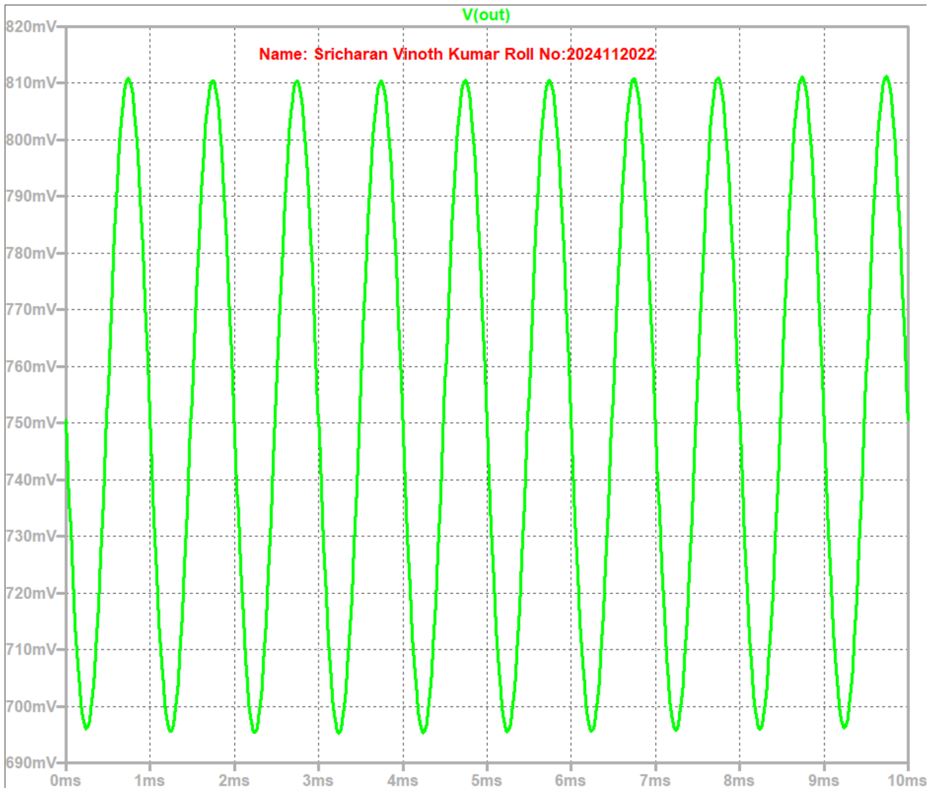


Figure 8: Plot of output signal

```
VoMax: MAX(V(out))=0.811338067055 FROM 0.004 TO 0.01
VoMin: MIN(V(out))=0.695251047611 FROM 0.004 TO 0.01
AmpVout: (VoMax - VoMin)=0.116087019444
```

Figure 9: Measurement of Amplitude of the Output Signal.

We get gain as, $A_v = 0.116087/0.01 = 11.6087$.

Increasing I_{REF} by 10%

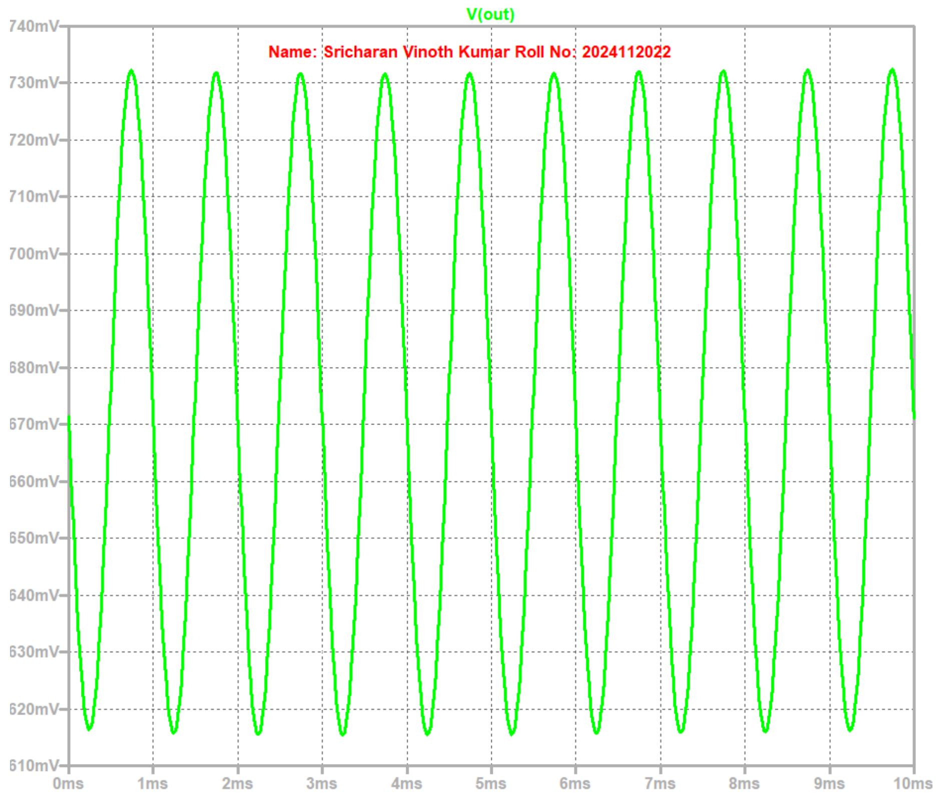


Figure 10: Plot of output signal

```
VoMax: MAX(V(out))=0.732568562031 FROM 0.004 TO 0.01
VoMin: MIN(V(out))=0.615359187126 FROM 0.004 TO 0.01
AmpVout: (VoMax - VoMin)=0.117209374905
```

Figure 11: Measurement of Amplitude of the Output Signal.

Gain has increased to $A_v = 0.117209/0.01 = 11.7209$, which is expected since we are expecting an increase in Gain as calculated in (a).
Increasing $\frac{W}{L}$ by 10%,

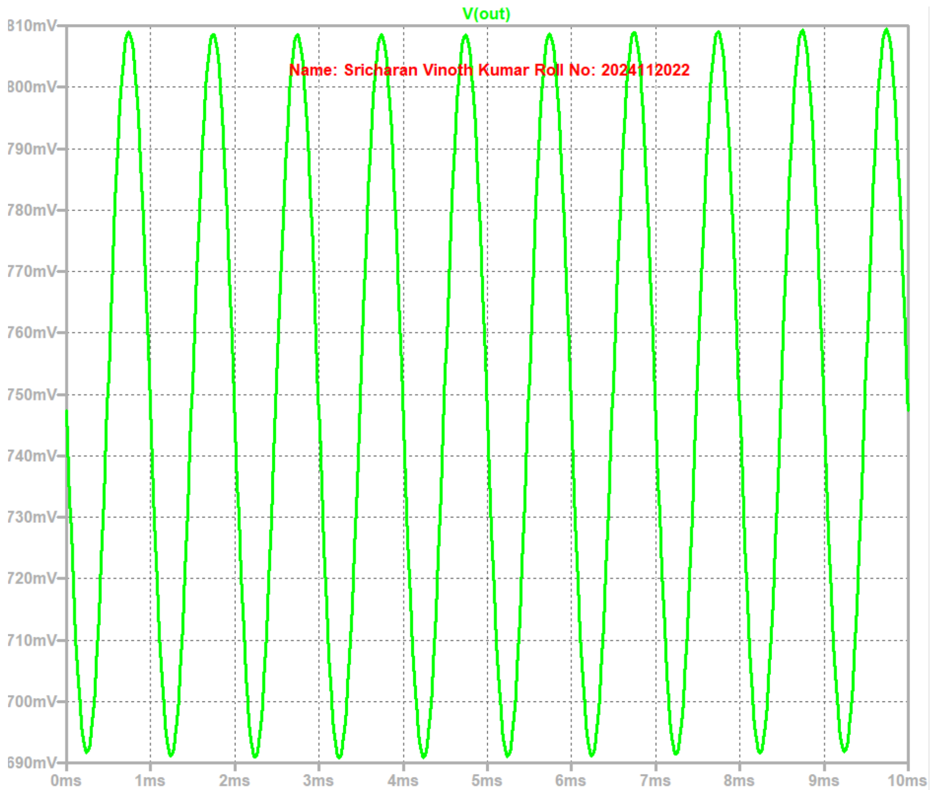


Figure 12: Plot of output signal

```
VoMax: MAX(V(out))=0.809469103813 FROM 0.004 TO 0.01
VoMin: MIN(V(out))=0.690835952759 FROM 0.004 TO 0.01
AmpVout: (VoMax - VoMin)=0.118633151054
```

Figure 13: Measurement of Amplitude of the Output Signal.

Gain has increased to $A_v = 0.118633/0.01 = 11.8633$, which is expected since we are expecting an increase in Gain as calculated in (a).

(c) Frequency Response Analysis of the Designed Amplifier:

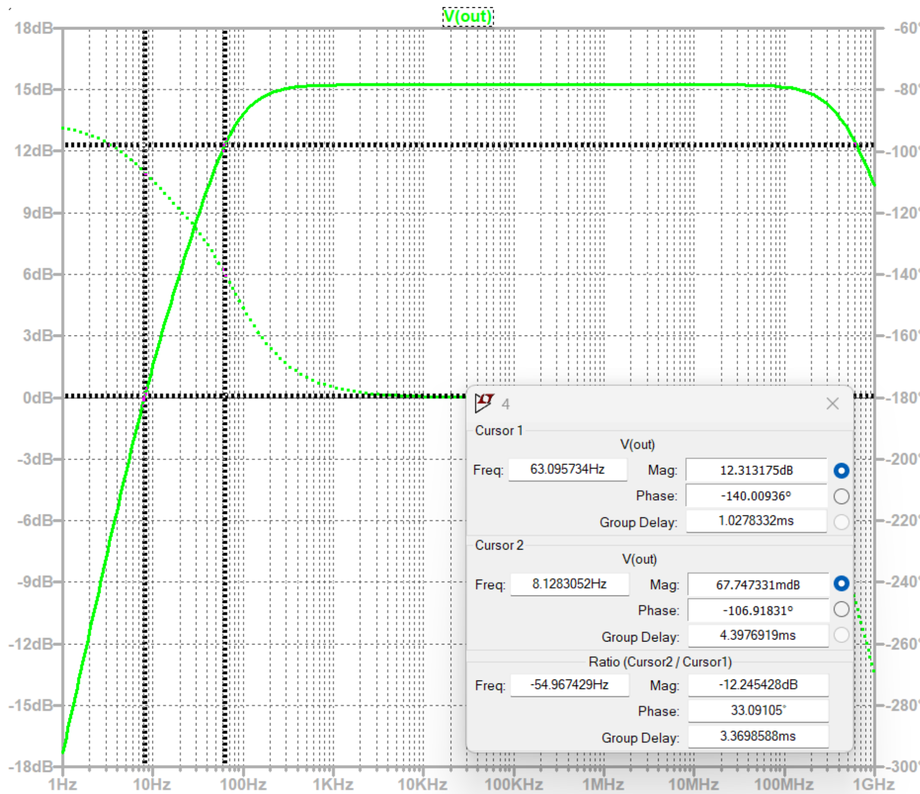


Figure 14: Result of Frequency Response Analysis

Cursor 1 shows the -3db frequency as ≈ 63.095Hz, and Cursor 2 shows the Unity Gain Frequency as 8.128Hz Increasing R_B by 10%, we get

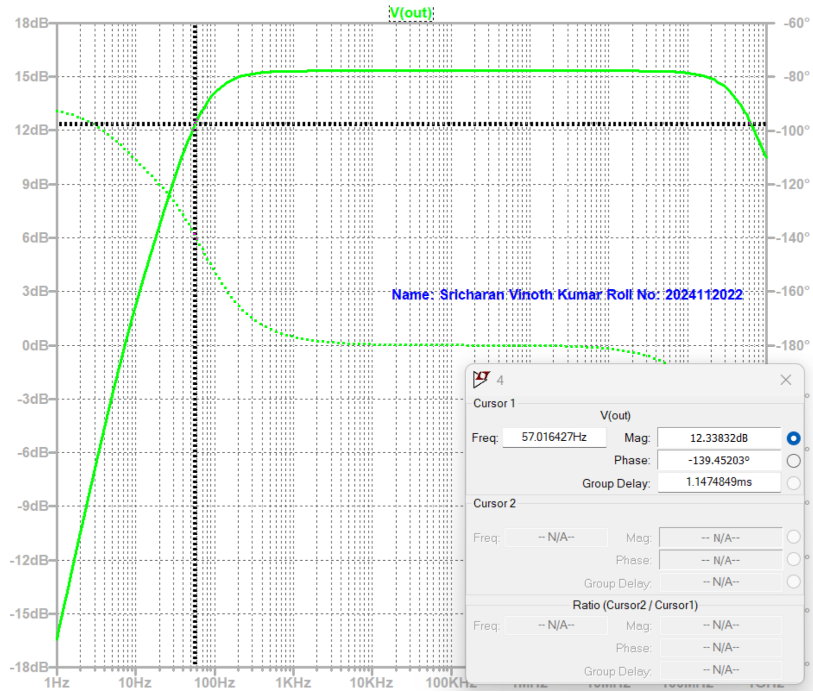


Figure 15: Result of Frequency Response Analysis

The -3db frequency has reduced to 57Hz, which is expected as we have calculated a decrease in f_{-3db} in (a)

Increasing C_B by 10%, we get

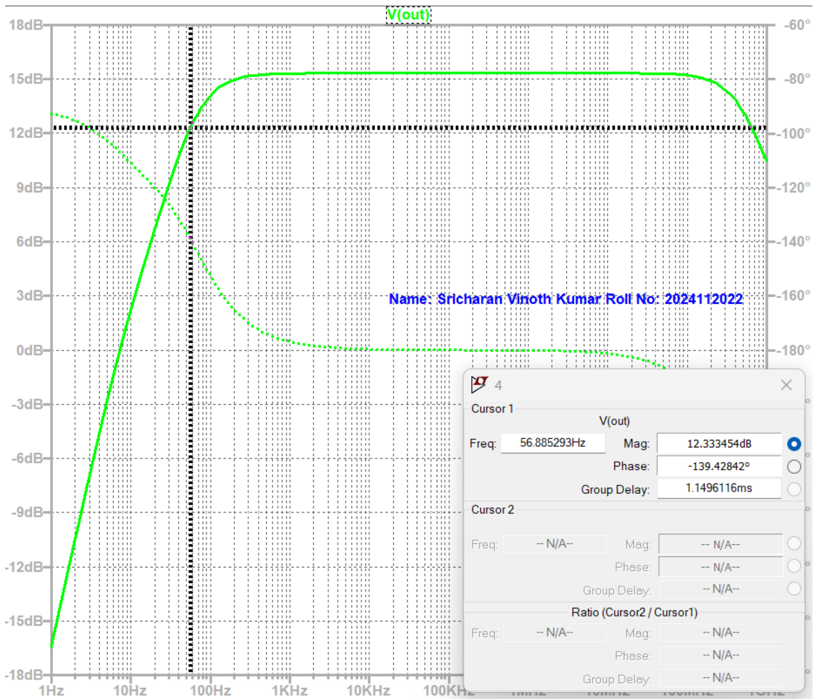


Figure 16: Result of Frequency Response Analysis

The -3db frequency has reduced to 56.8Hz, which is expected as we have calculated a decrease in f_{-3db} in (a)

This also shows that the effect of reducing R_B and C_B are similar.