Contents

Part I

A. 2N2222A h-parameters

For V = 10V, I = 1 mA, f = 1kHz, and T = 25° C, the following h-parameters were obtained in **Table 1.1**:

Table 1.1: Table showing the values of parameters for the 2N2222A transistor from online datasheets[1]

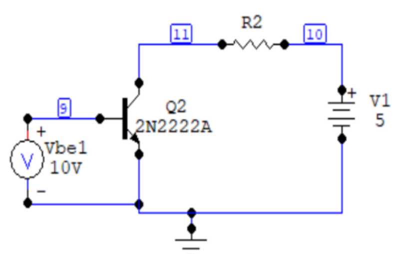
Hybrid-π parameter	h-parameter	Min. Value	Max. Value	Unit
β	h_{fe}	50	300	–
r_π	h_{ie}	2	8	$k\Omega$
r_o	h_{oe}	5	35	μS

B. Comparing “Measured” Values to Datasheet Values

This section covers how the Hybrid-π parameters for the 2N2222A are measured using a simulation software.

I. I_B vs V_{BE}

Circuit 1.1 was used to plot the graph of I_B vs V_{BE} (**Figure 1.1**).



Circuit 1.1: Circuit used to plot I_B vs

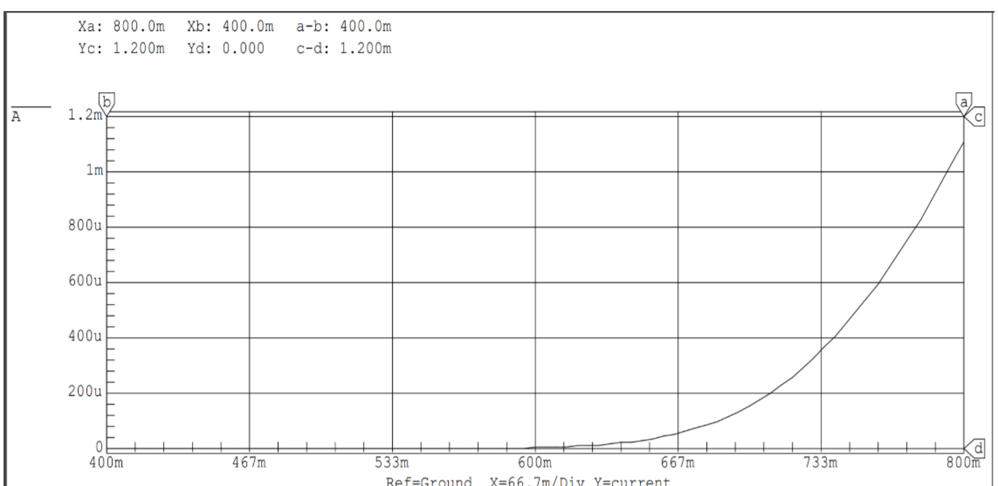
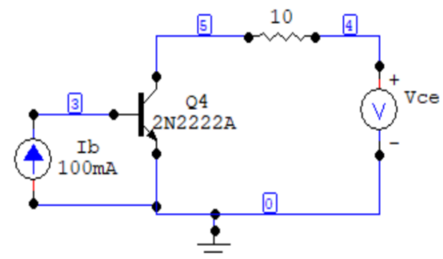


Figure 1.1: I_B vs V_{BE}

[1] https://www.onsemi.com/pub/Collateral/P2N2222A-D.PDF

II. I_C vs V_{CE} varying I_B



Circuit 1.2: Circuit used to plot

Circuit 1.2 was used to plot the graph of I_C vs V_{CE} with I_B as a variable parameter (**Figure 1.2**). By performing a DC Sweep, V_{CE} was varied from 0V to 6V and I_B was varied from 1μA to 10μA with 1μA step value. Ic was then measured to produce **Figure 1.2**. Using the cursors, the graph corresponding to 1 mA and 5V was located, as indicated on the graph by cursor a and c, and it corresponds to the $I_B = 6\mu A$ graph, which is the value of I_B that will be used.

Given that $I_C = 1\ mA$, we can calculate β as follows:

$$\beta = \frac{I_C}{I_B} = \frac{0.001}{0.000006} = 166.67$$

For $V_T = 25\ mV$, we can calculate the value of r_π using the value of β we calculated above:

$$g_m = \frac{I_C}{V_T} = \frac{0.001}{0.025} = 0.04\ V = 40\ mS$$

$$r_\pi = \frac{\beta}{g_m} = \frac{166.67}{0.04} = 4.167\ k\Omega$$

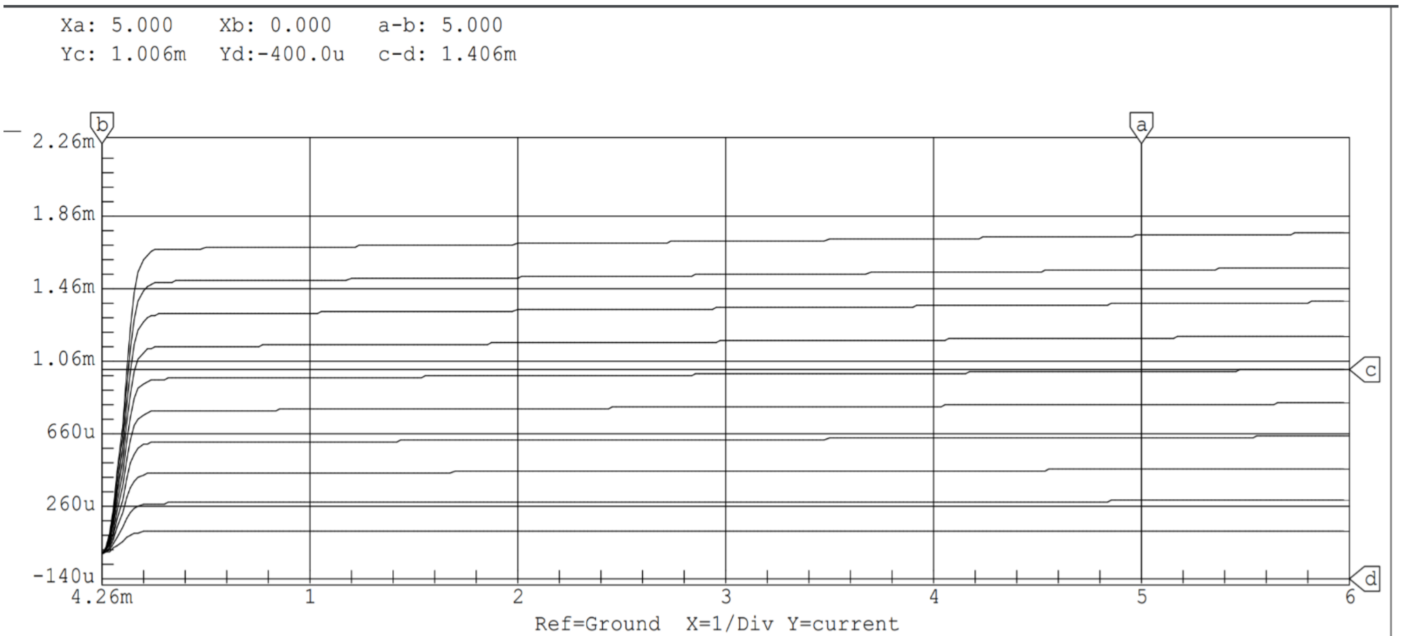


Figure 1.2: I_C vs V_{CE} Plot with I_B as a variable parameter

III. Using the I_C vs V_{CE} varying V_{BE} graph to estimate Early Voltage V_A

Circuit 1.1 was used again to generate the I_C vs V_{CE} Plot with V_{BE} as a variable parameter (**Figure 1.3**), however this time, V_{BE} is varied instead of I_B, from 0 to 800mV. Using the cursors, the slope of the graph was measured using two data sets

$$(x_1, y_1) = (67.5V, 0.1931\ A) \text{ and } (x_2, y_2) = (67.2V, 0.1928\ A)$$

The slope, m, was calculated as follows:

$$m = \frac{y_2 - y_1}{x_2 - x_1}$$

$$\frac{0.1928 - 0.1931}{67.2 - 67.5} = 0.0009545 \approx 0.00095\ \frac{V}{A}$$

Now the equation of the graph is in the form $y = mx + b$, where b is the y-intercept of the graph, which we need to find in order to solve for the early Voltage. Using data point (x_1, y_1) and the slope m, the y-intercept was found to be 0.1287, giving us the equation of the graph $y = 0.00095x + 0.1287$. In order to find the Early Voltage, we need to extrapolate the graph to find the x-intercept (at y=0). We can use the equation of the line found earlier to calculate the x-intercept as follows:

$$0 = m(-V_A) + b$$

$$V_A = \frac{b}{m} = \frac{0.1287}{0.00095} = 134.80\ V$$

The Early Voltage was then used to find r_o, given $I_C = 1\ mA$:

$$r_o = \frac{V_A}{I_C} = \frac{134.8}{0.001} = 134.8\ k\Omega = 7.418\ \mu S$$

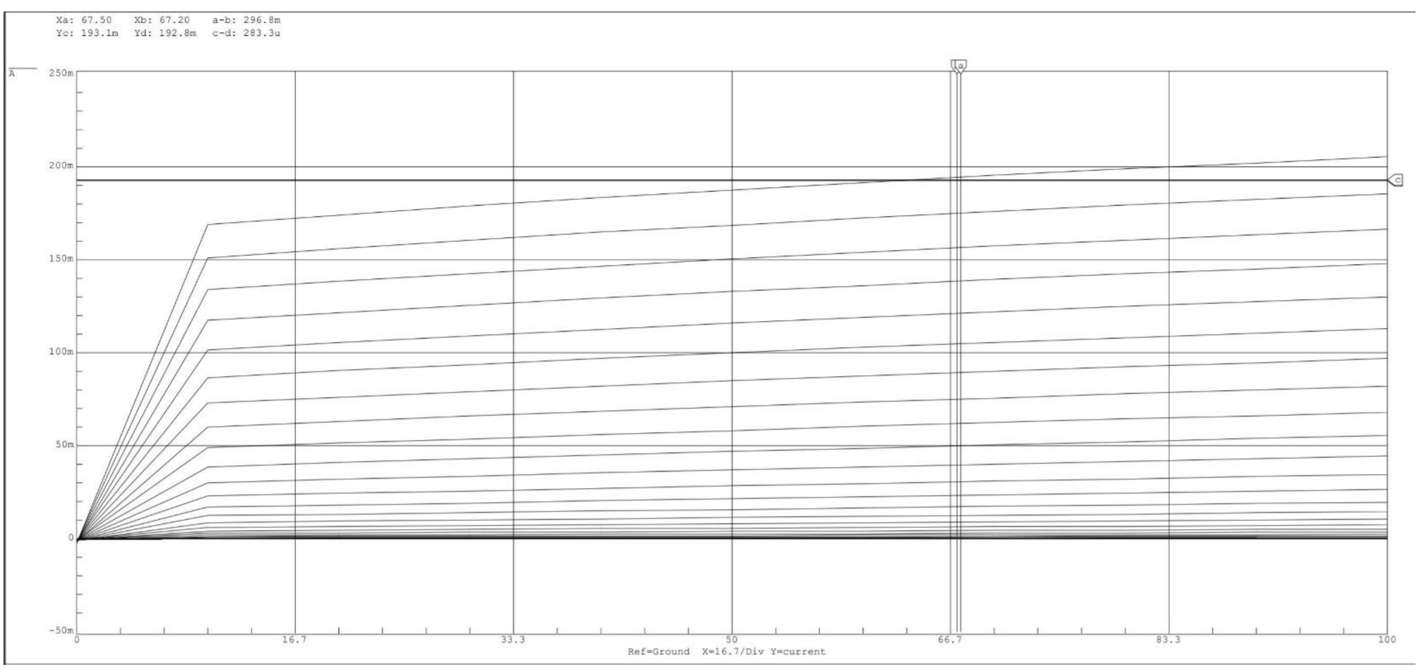


Figure 1.3: I_C vs V_{CE} Plot with V_{BE} as a variable parameter

Comparing "measured" values with datasheet values

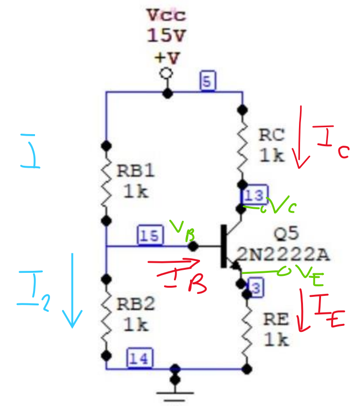
As shown in **Table 1.2**, the values "measured" above fall within the range of values obtained from the datasheet in part a. This makes the approximations of the measured values adequate, since they fall within the acceptable ranges for the 2N2222A transistor as specified by its datasheet.

Table 1.2: Table comparing the "measured" values and datasheet values of the Hybrid-π parameters for the 2N2222A

Hybrid-π parameter	h-parameter	Measured value	Datasheet Range	Unit
β	h_{fe}	166.7	50 – 300	–
r_π	h_{ie}	4.167	2 – 8	$k\Omega$
r_o	h_{oe}	7.418	5 – 35	μS

C. Simple Bias Network for NPN Transistor

I. Biasing the Circuit and Measuring D.C Operating Point



Circuit1.3: Bias Circuit for 2N2222A

To bias **Circuit 1.3**, we need the following values:

$V_{CC} = 15$ $\quad V_{CE} = 4\ V$ $\quad I_C = 1\ mA$ $\quad I_B = \frac{I_C}{\beta} = \frac{1\ \text{mA}}{166.67} = 6\ uA$

$I_E = I_C + I_B = 1.006\ mA$ $\quad V_{BE} = 0.7\ V$ *(an assumption we always make)*

Using mesh analysis,

$$V_{CC} = I_C R_C + I_E R_E + V_{CE}$$

Knowing that $R_E = \frac{R_C}{2}$

$$15\ V = 1\ \text{mA}\ \times R_C + 1.006\ \text{mA}\ \times \frac{R_C}{2} + 4\ V$$

We can then solve for R_C and R_E:

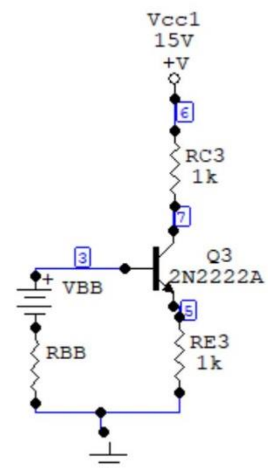
$$R_C = 7.318\ k\Omega \qquad R_E = 3.659\ k\Omega$$

Now, we can proceed to find the voltages V_E, V_B and V_C:

$$V_E = I_E R_E = 3659\ V \times 1.006\ \text{mA} = 3.618\ \text{V}$$

$$V_B = V_E + V_{BE} = 3.618 + 0.7 = 4.318\ V$$

$$V_C = V_E + V_{CE} = 3.618 + 4 = 7.618\ V$$



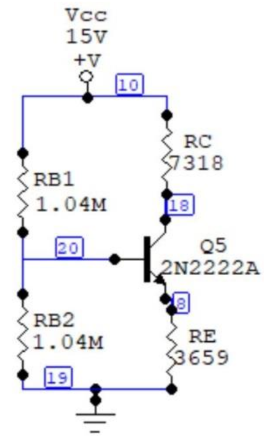
Circuit 1.4: Thevenin Equivalent of the Bias Circuit

Now we need to find $R_{B1}$and R_{B2}, so we perform a mesh analysis on the Thevenin equivalent circuit of the bias circuit (**Circuit 1.4**) as follows:

$$V_{BB} = I_B R_{BB} + V_{BE} + I_E R_E = I_B R_{BB} + V_B \qquad \textbf{e.q. (1)}$$

$$V_{BB} = \frac{V_{CC} \times R_{B2}}{R_{B1}+R_{B2}} \qquad \text{(Using Voltage Divider Method)}$$

$$R_{BB} = \frac{R_{B1} \times R_{B2}}{R_{B1}+R_{B2}}$$



Circuit 1.5: Biased Circuit with Values

Note: Here and in all subsequent bias circuit calculations, the assumption that $R_{B1} = R_{B2}$ was made to achieve maximum power transfer (and for ease of calculation).

And from **Circuit 1.3** we have

$$I_1 = I_2 + I_B \qquad \textbf{e.q. (2)}$$

where

$$I_1 = \frac{V_{CC} - V_B}{R_{B1}} = \frac{15-4.318}{R_{B1}} \qquad I_2 = \frac{V_B}{R_{B2}} = \frac{4.318}{R_{B2}}$$

Solving **e.q. (1)** and **e.q. (2)**, we get:

$$R_{B1} = R_{B2} = 1.04\ M\Omega$$

The bias circuit was then simulated using the values calculated above, and the D.C. operating point of the circuit was measured:

$$\boxed{V_B = 4.325\ V \quad I_B = 6.107\ \mu A \quad V_C = 7.599\ V \quad I_C = 1.011\ mA \quad V_E = 3.723\ V \quad I_E = 1.017\ mA}$$

II. Measuring D.C. Operating Point Using the 1/3rd Rule

Now we will the use the first version of the 1/3rd Rule to Bias the Circuit. The Rule is:

$$V_B = \frac{1}{3}\ V_{CC} \qquad V_C = \frac{2}{3}\ V_{CC} \qquad I_1 = \frac{I_E}{\sqrt{\beta}}$$

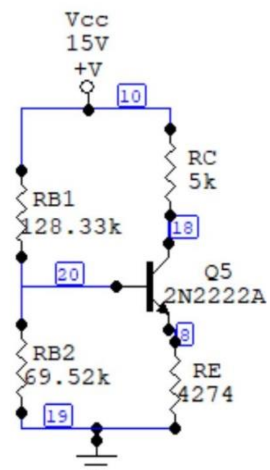
Thus,

$$V_B = \frac{15}{3} = 5\ V \qquad V_C = \frac{30}{3} = 10V \qquad V_E = V_B - 0.7\ V = 4.3\ V$$

ow we can proceed to find the currents:

$$I_C = 1\ mA \qquad I_B = \frac{I_C}{\beta} = \frac{1\ mA}{166.67} = 6\ \mu\text{A} \qquad I_E = I_C + I_B = 1.006\ m$$

$$I_1 = \frac{I_E}{\sqrt{\beta}} = \frac{1.006\ mA}{\sqrt{166.67}} = 77.79\ \mu\text{A} \qquad I_2 = I_1 - I_B = 71.92\ \mu\text{A}$$



Circuit 1.6: Biased Circuit with Values obtained from the 1/3rd Rule

And finally, we can calculate the resistor values using $R = \frac{V}{I}$:

$$R_E = \frac{V_E}{I_E} = \frac{4.3\ V}{1.006\ mA} = 4.274\ k\Omega \qquad R_C = \frac{V_{CC} - V_C}{I_C} = \frac{15-10}{1\ mA} = 5\ k\Omega$$

$$R_{B1} = \frac{V_{CC} - V_B}{I_1} = \frac{15-5V}{77.79\ \mu\text{A}} = 128.33\ k\Omega \qquad R_{B2} = \frac{V_B}{I_2} = \frac{5V}{71.92\ \mu\text{A}} = 69.52\ k\Omega$$

Now, the 1/3rd Rule Bias circuit has been modelled on CircuitMaker (**Circuit 1.**6) to measure its D.C. Operating Point:

$$\boxed{V_B = 4.966\ V \quad I_B = 6.087\ \mu A \quad V_C = 9.889\ V \quad I_C = 1.022\ mA \quad V_E = 4.395\ V \quad I_E = 1.028\ mA}$$

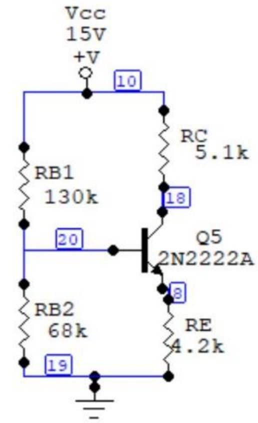
III. Measuring D.C. Operating Point Using the Closest Most Commonly Available Resistor Values

The closest commonly available resistor values to the calculated values in part II are as follows:

$$R_E = 4.3\ k\Omega \qquad R_C = 5.1\ k\Omega$$

$$R_{B1} = 130\ k\Omega \qquad R_{B2} = 68\ k\Omega$$

These values were then used to model the circuit and measure the D.C. Operating Point as in **Circuit 1.7**:



Circuit 1.7: Biased Circuit with Closest Commonly Available Values

$V_B = 4.888\ V \quad I_B = 57.71\ \mu A \quad V_C = 9.946\ V \quad I_C = 1.580\ mA \quad V_E = 4.287\ V \quad I_E = 1.638\ mA$

IV. Comparing operating Points

Table 1.2: Table comparing the operating points from parts I, II, and II for the 2N2222A:

D.C. Operating Point	Bias Circuit	1/3rd Rule	Closest Common Resistor Values
V_B	$4.325\ V$	$4.966\ V$	$4.888\ V$
I_B	$6.107\ \mu A$	$6.087\ \mu A$	$5.912\ \mu A$
V_C	$7.599\ V$	$9.889\ V$	$9.946\ V$
I_C	$1.011\ mA$	$1.022\ mA$	$0.9911\ mA$
V_E	3.723	$4.395\ V$	$4.287V$
I_E	$1.017\ mA$	$1.028\ mA$	$0.997\ mA$

D. Other Transistors

Note: For conciseness, the steps that have already been explained in part b and c were not shown again.

2N3904

Using **Circuit 1.2** with the 2N3904 transistor a DC sweep was performed (as done in part c, ii) to locate the I_B value corresponding to $I_C = 1\ mA$ and $V_{CE} = 5\ V$, which was found to be $I_B = 8.5\ \mu A$.

The following values were then determined similarly:

$\beta = 117.7$ $\quad g_m = \frac{I_C}{V_T} = \frac{0.001}{0.025} = 0.04\ V = 40\ mS$ $\quad r_\pi = \frac{\beta}{g_m} = \frac{117.7}{0.04} = 2.943\ k\Omega$

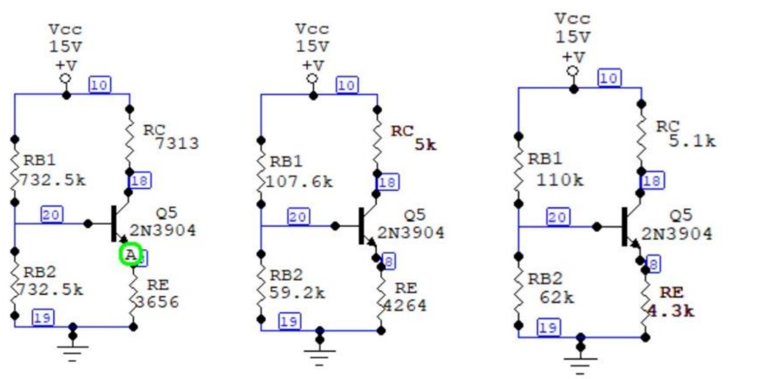


Figure 1.4: (a) Bias Circuit (b) 1/3rd Rule (c) Common Resistor Values

Table 1.3: Table comparing the operating points from parts I, II, and II for the 2N3904:

D.C. Operating Point	Bias Circuit	1/3rd Rule	Closest Common Resistor Values
V_B	$4.910\ V$	$5.086\ V$	$5.158\ V$
I_B	$7.072\ \mu A$	$6.222\ \mu A$	$6.275\ \mu A$
V_C	$6.442\ V$	$9.773\ V$	$9.629\ V$
I_C	$1.170\ mA$	$1.045\ mA$	$1.053\ mA$
V_E	$4.304\ V$	$4.484\ V$	$4.556\ V$
I_E	$1.177\ mA$	$1.052\ mA$	$1.059\ mA$

2N4401

Using **Circuit 1.2** with the 2N4401transistor a DC sweep was performed (as done in part c, ii) to locate the I_B value corresponding to $I_C = 1\ mA$ and $V_{CE} = 5\ V$, which was found to be $I_B = 7\ \mu A$. The following values were then determined similarly:

$\beta = 142.9$ $g_m = \frac{I_C}{V_T} = \frac{0.001}{0.025} = 0.04\ V = 40\ mS$ $r_\pi = \frac{\beta}{g_m} = \frac{142.9}{0.04} = 3.573\ k\Omega$

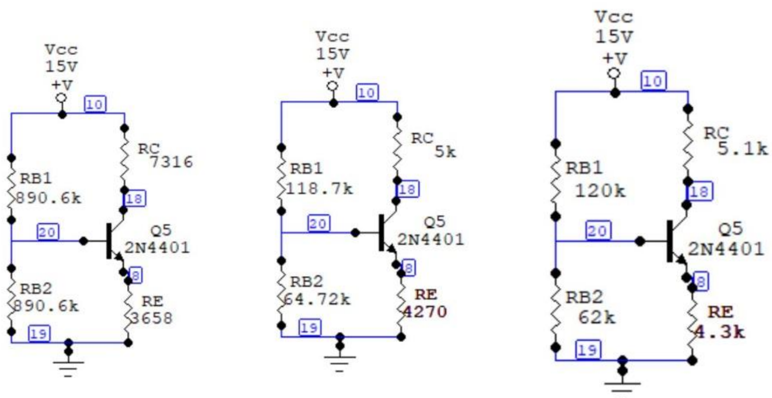


Figure 1.4: (a) Bias Circuit (b) 1/3rd Rule (c) Common Resistor Values

<u>**Table 1.4:** Table comparing the operating points from parts I, II, and II for the 2N4401:</u>

D.C. Operating Point	Bias Circuit	1/3rd Rule	Closest Common Resistor Values
V_B	$4.589\ V$	$5.035\ V$	$4.869\ V$
I_B	$6.538\ \mu A$	$6.150\ \mu A$	$5.887\ \mu A$
V_C	$7.078V$	$9.736\ V$	$9.967\ V$
I_C	$1.083\ mA$	$1.032\ mA$	$0.9868\ mA$
V_E	$3.985\ V$	$4.443\ V$	$4.269\ V$
I_E	$1.089\ mA$	$1.038mA$	$0.9927\ mA$

As shown in the tables above for each transistor, the D.C Operating Points for the circuit obtained through the 1/3rd Rule are closer to those obtained through using the most commonly available resistors than those obtained through normal bias of the circuit using Thevenin equivalent. However, such a comparison is not enough to make that generalisation, since the values were measured using Circuitmaker, a very old software with a lot of glitches. Moreover, when the normal bias circuit was calculated using Thevenin equivalent, the assumption that $R_{B1} = R_{B2}$ was made in order to achieve maximum power transfer, which resulted in resistance values that are very different from those found using the 1/3rd rule. Using the most commonly available resistor values for the 1/3rd rule resistances is an appropriate approximation to make since the D.C. operating points are very similar, and the difference can be neglected.

Part II

Note: In this section, calcualtions relating to poles and zeros shown in this report are kept concise as they were covered in Mini-project 1.

Figure 2.1 shows the Common-Emitter Amplifier circuit configuration which we will simulate in this section.

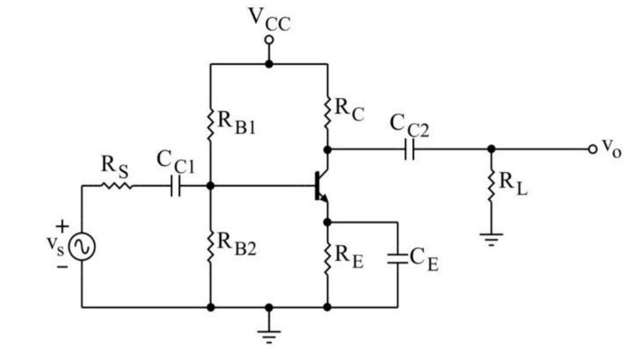


Figure 2.1: The Common-Emitter Amplifier Configuration

The Values found in part 1, c, iii will be used, along with the given values in problem:

$R_E = 4.3\ k\Omega$ $\quad R_C = R_L = 5.1\ k\Omega$ $\quad R_{B1} = 130\ k\Omega$ $\quad R_{B2} = 68\ k\Omega$ $\quad \beta = 166.67$

$R_S = 50\ \Omega$ $\quad C_{C1} = C_{C2} = C_E = 10\ \mu F$ $\quad R_{BB} = \frac{R_{B1} \times R_{B2}}{R_{B1} + R_{B2}} = 44.65\ k\Omega$ $\quad r_\pi = 4.3\ k\Omega$

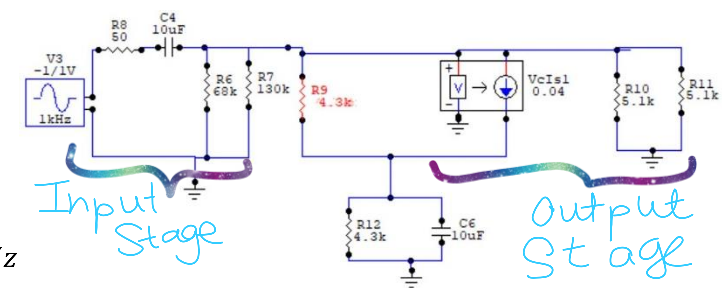
R_L was chosen to be equal to R_C to achieve maximum power transfer.

A. Identifying Poles and Zeros

First, we start by finding the low frequency poles and zeros using the Low-Frequency Small-Signal model of the circuit (**Circuit 2.1**), as follows:

$$\omega_{Lp1} = \frac{1}{\tau_{C_{C1}}} = \frac{1}{(R_S + R_{BB}||(r_\pi + R_E(1+\beta)) \times C_{C1}}$$

$$= 2.375 \frac{rad}{sec} = 0.3779 Hz$$



Circuit 2.1: Low Frequency Small Signal Model

$$\omega_{Lp2} = \frac{1}{\tau_{C_E}} = \frac{1}{(R_C + R_L) \times C_{C2}} = 9.804 \frac{rad}{sec} = 1.560\ Hz$$

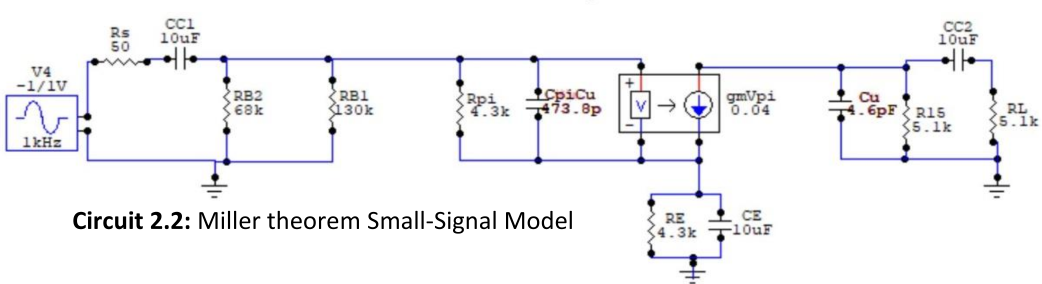
For C_{C2}, the output stage is decoupled from the input stage:

$$\omega_{Lp3} = \frac{1}{\tau_{C_{C2}}} = \frac{1}{\frac{(R_S||R_{BB} + r_\pi)}{1+\beta}\ ||R_E \times C_E} = 3.878\ k\frac{rad}{sec} = 617.2\ Hz$$

The low frequency zeros due to C_{C1} and C_{C2} are 0, and the third zero can be calculated as follows:

$$\omega_{Lz3} = \frac{1}{R_E \times C_E} = 23.26 \frac{rad}{sec} = 3.701 Hz$$

Firstly, we will start by applying Miller's theorem to our Small-Signal Model of the circuit (**Circuit 2.2**).



Circuit 2.2: Miller theorem Small-Signal Model

The value of C_μ and C_π, were calculated using the graphs provided in the datasheet, using the operating point previously found in part I:

$$V_{BE} = 0.7\,V \qquad\qquad V_{CB} = 5\,V$$

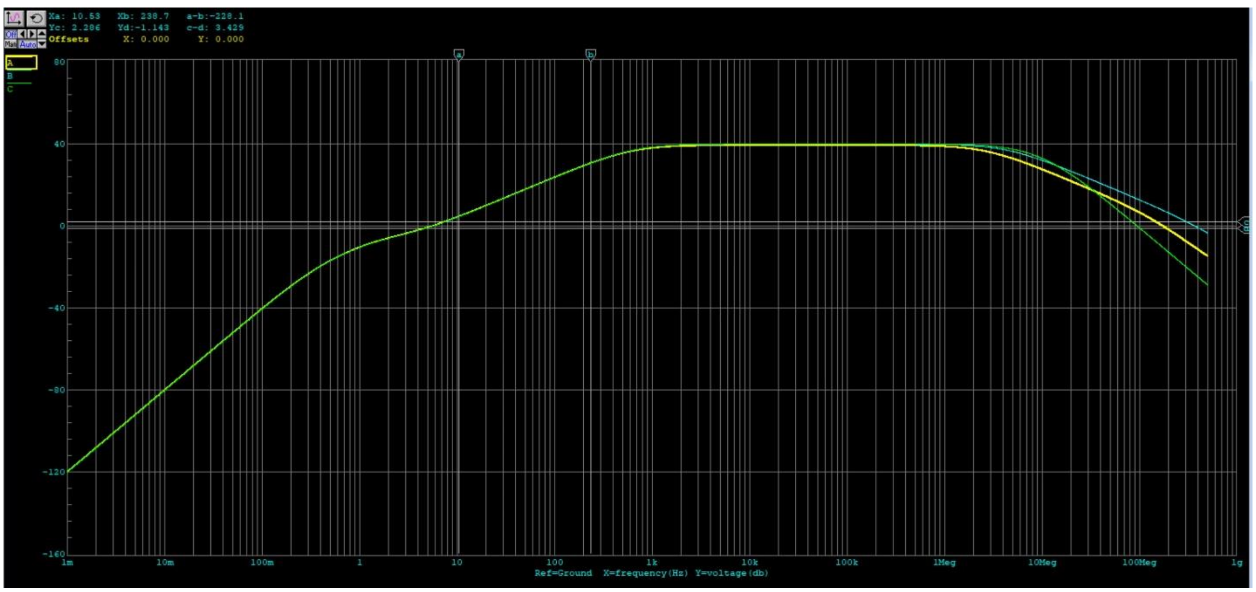
These voltages were traced on the graph to be $C_\mu = 4.7\,pF$ and $C_\pi = 19\,pF$. These were then used to calculate the Miller capacitances as shown in **Circuit 2.2**.

To find the High-Frequency poles, Miller's theorem was used to find applied on the input and output of **Circuit 2.2.**

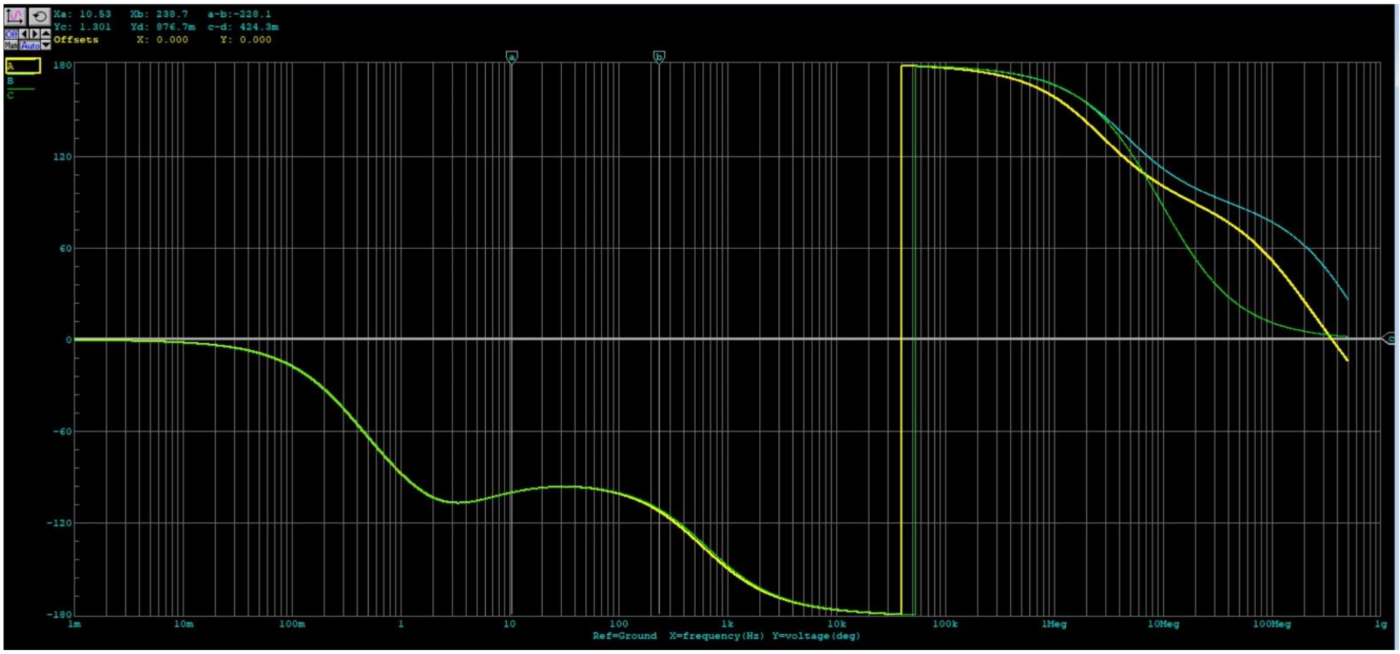
$$\omega_{Hp1} = \frac{1}{\tau_{C_{473.8+4.7}}} = \frac{1}{(R_S||R_{BB}||r_\pi)\ \times (473.8 + 4.7)pF} = 42.33M\ \frac{rad}{sec} = 6.737\ MHz$$

$$\omega_{Hp2} = \frac{1}{\tau_{C_\mu}} = \frac{1}{(R_C||R_L)\ \times 4.7pF} = 83.44M\ \frac{rad}{sec} = 13.28\ MHz$$

The high frequency zeros of this circuit are at infinity.



Graph 2.1: Bode Magnitude plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green)



Graph 2.2: Bode Phase Plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green)

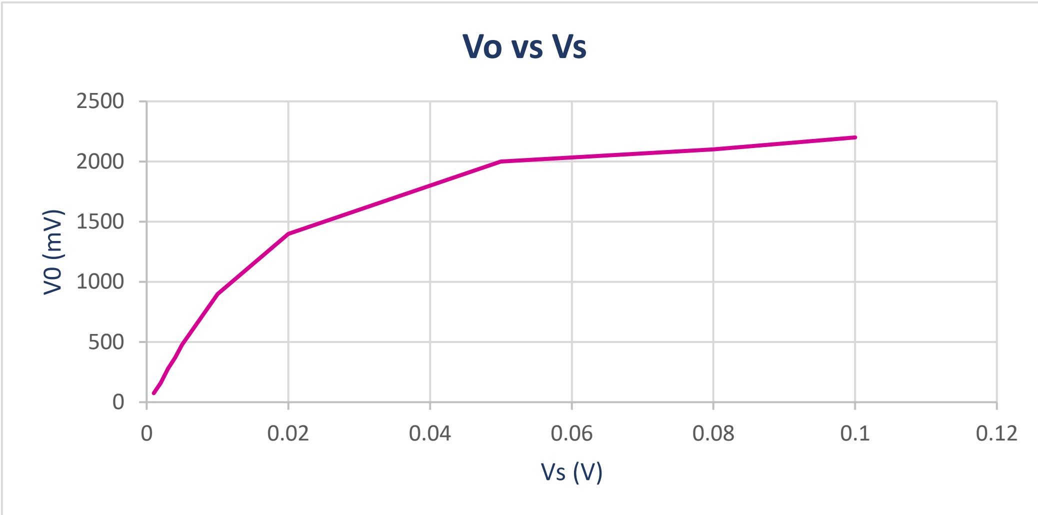
<u>**Table 2.1:** Table comparing the values of the poles and zeros (Hz) obtained graphically and by calculation:</u>

Pole/Zero	ω_{Lp1}	ω_{Lp2}	ω_{Lp3}	ω_{Hp1}	ω_{Hp2}	ω_{Lz1}	ω_{Lz2}	ω_{Lz3}	ω_{Hz1}	ω_{Hz2}
Calculated Value	0.3779	1.560	617.2	6.373M	13.28M	0	0	3.701	∞	∞
Measured Value	0.3594	1.668	655.6	6.391M	13.43M	0	0	6.556	∞	∞
% Error	5.15	6.47	5.86	0.28	1.12	–	–	2.86	–	–

As shown in the table, the calculated values and measured values are very close, and lie within a very small percentage error of one another, which makes the small signal model an accurate method to approximate the poles and zeros of the transistor.

B. Mid-band Frequency

From **Graph 2.1**, the point at 20 kHz on the curve is about exactly the middle of the band, which is why I chose a mid-band frequency of 20 kHz. I then proceeded to vary the source peak voltage from 0.1 to 1, which provided a reasonable range of data points, giving the desired shape of the Voltage Transfer Curve.



Graph 2.3: Voltage Transfer Curve for the 2N3904

From the graph, it appears that the graph stops being Linear at Vs = 0.01V.

C. Measuring Input Impedance

Now to calculate Input Impedance, I used the Input stage from **Circuit 2.1** to calculate the Input impedance as follows:

$$R_{in} = R_{BB}||r_\pi = 44.65k||4300 = 3.922\ k\Omega$$

Then, I measured the Input Impedance by simulating the circuit, which turned out to be:

$$R_{in} = \frac{V}{I} = \frac{4.487V}{910.5\mu A} = 4.928\ k\Omega$$

The calculated Input Impedance is smaller than the measure input impedance by about 13%, which is not a negligible error, however, both values are in the same order of magnitude. This is, however, not surprising, since Miller's theorem only provides an approximation to the circuit.

D. Measuring Output Impedance

Now to calculate Output Impedance, I used the output stage from **Circuit 2.1** to calculate the Output impedance as follows:

$$R_{out} = R_C = 5.1\ k\Omega$$

Then, I measured the Output Impedance by simulating the circuit , which turned out to be:

$$R_{out} = \frac{V}{I} = \frac{3.57\ mV}{700\ nA} = 5.1\ k\Omega$$

The calculated Output Impedance is the exact same value as the measured Output Impedance, which is what was expected.

E. 2N3904

The Values found in part 1, c, iii will be used, along with the given values in problem:

$R_E = 4.3\ k\Omega$ $\quad R_C = R_L = 5.1\ k\Omega$ $\quad R_{B1} = 110\ k\Omega$ $\quad R_{B2} = 62\ k\Omega$ $\quad \beta = 117.65$

$R_S = 50\ \Omega$ $\quad C_{C1} = C_{C2} = C_E = 10\ \mu F$ $\quad R_{BB} = \frac{R_{B1} \times R_{B2}}{R_{B1} + R_{B2}} = 39.65\ k\Omega$ $\quad r_\pi = 3\ k\Omega$

R_L was chosen to be equal to R_C to achieve maximum power transfer.

A. Identifying Poles and Zeros

First, we start by finding the low frequency poles and zeros using the Low-Frequency Small-Signal model of the circuit (**Circuit 2.1**), as follows:

$$\omega_{Lp1} = \frac{1}{\tau_{C_{C1}}} = \frac{1}{(R_S + R_{BB}||(r_\pi + R_E(1+\beta))\ \times C_{C1}}$$

$$= 2.713 \frac{rad}{sec} = 0.4318 Hz$$

$$\omega_{Lp2} = \frac{1}{\tau_{C_E}} = \frac{1}{(R_C + R_L) \times C_{C2}} = 9.804 \frac{rad}{sec} = 1.560\ Hz$$

For C_{C2}, the output stage is decoupled from the input stage:

$$\omega_{Lp3} = \frac{1}{\tau_{C_{C2}}} = \frac{1}{\frac{(R_S||R_{BB} + r_\pi)}{1+\beta}\ ||R_E \times C_E} = 3.914\ k \frac{rad}{sec} = 622.9\ Hz$$

The low frequency zeros due to C_{C1} and C_{C2} are 0, and the third zero can be calculated as follows:

$$\omega_{Lz3} = \frac{1}{R_E \times C_E} = 23.26 \frac{rad}{sec} = 3.701 Hz$$

Firstly, we will start by applying Miller's theorem to our Small-Signal Model of the circuit (**Circuit 2.2**).

The value of C_μ and C_π, were calculated using the graphs provided in the datasheet, using the operating point previously found in part I:

$$V_{BE} = 0.7\ V \qquad V_{CB} = 5\ V$$

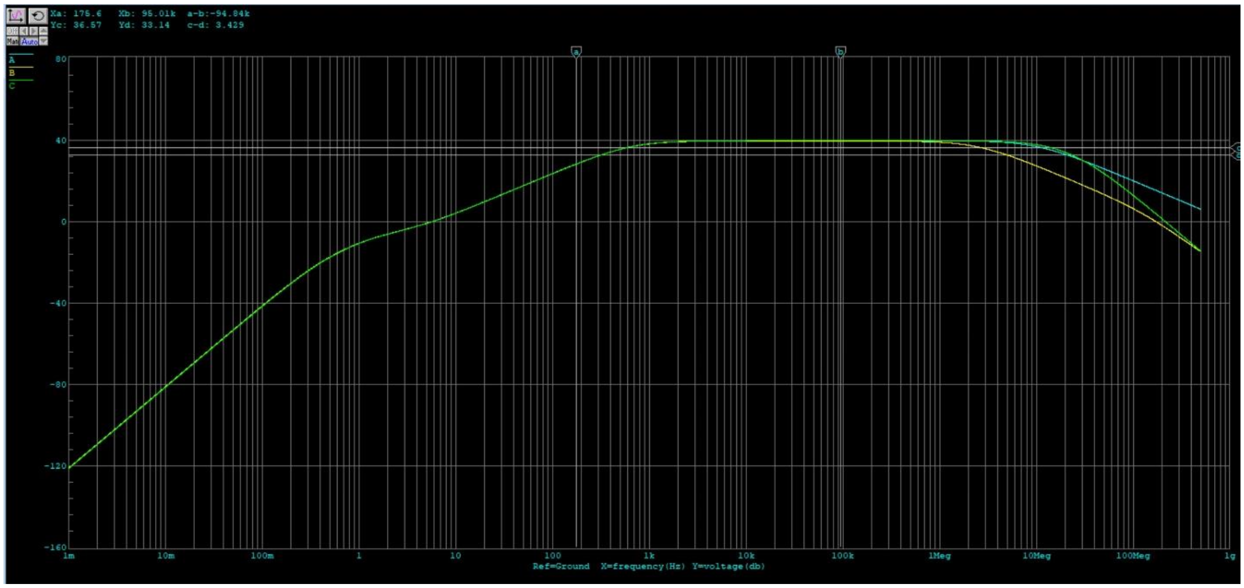
These voltages were traced on the graph to be $C_\mu = 2\ pF$ and $C_\pi = 3.5\ pF$. These were then used to calculate the Miller capacitances as shown in **Circuit 2.2**.

To find the High-Frequency poles, Miller's theorem was used to find applied on the input and output of **Circuit 2.2.**

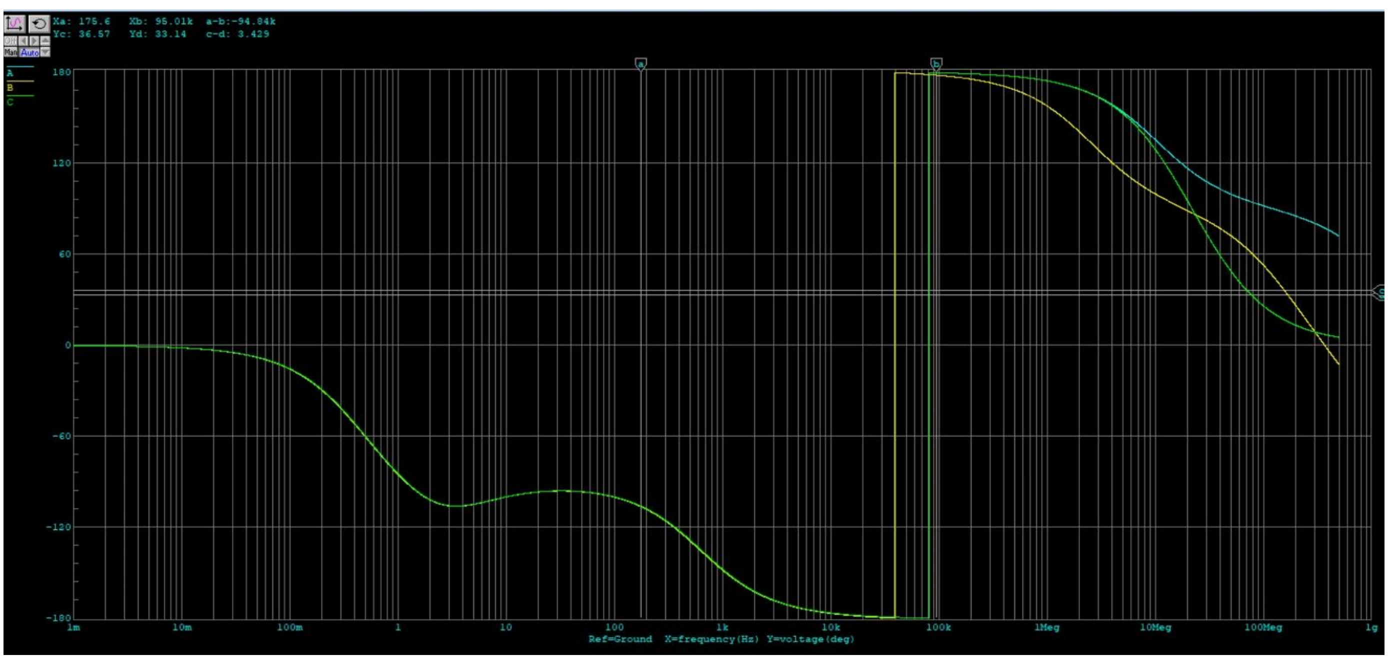
$$\omega_{Hp1} = \frac{1}{\tau_{C_{473.8+4.7}}} = \frac{1}{(R_S\big||R_{BB}\big||r_\pi)\ \times (209.5) pF} = 97.18\ M\ \frac{rad}{sec} = 15.45\ MHz$$

$$\omega_{Hp2} = \frac{1}{\tau_{C_\mu}} = \frac{1}{(R_C||R_L)\ \times 2\ pF} = 196.1M\ \frac{rad}{sec} = 31.21\ MHz$$

The high frequency zeros of this circuit are at infinity.



Graph 2.4: Bode Magnitude plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green) for the 2N3904



Graph 2.5: Bode Phase Plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green)

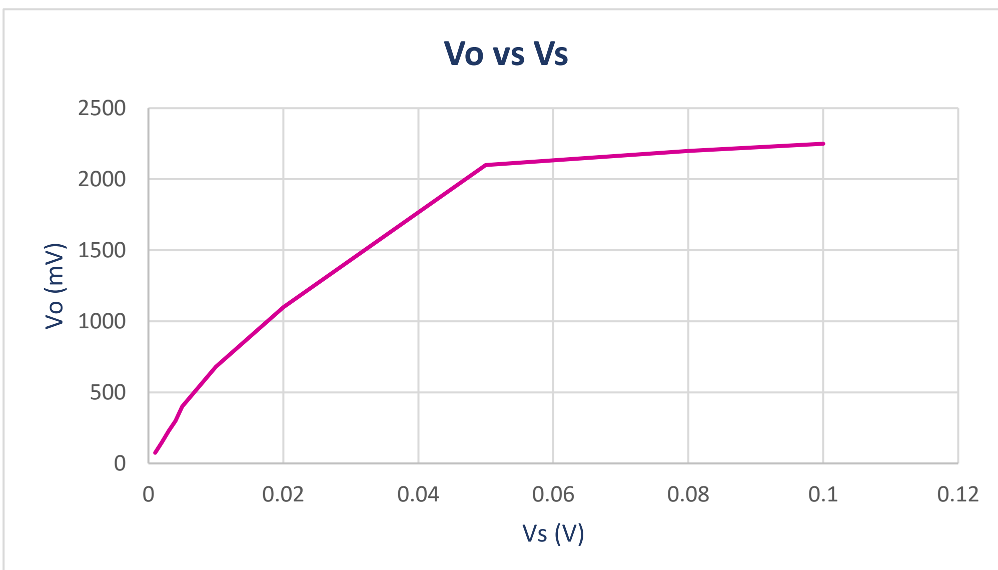
Table 2.2: Table comparing the values of the poles and zeros (Hz) obtained graphically and by calculation:

Pole/Zero	ω_{Lp1}	ω_{Lp2}	ω_{Lp3}	ω_{Hp1}	ω_{Hp2}	ω_{Lz1}	ω_{Lz2}	ω_{Lz3}	ω_{Hz1}	ω_{Hz2}
Calculated Value	0.4318	1.560	622.9	15.45M	31.21M	0	0	3.701	∞	∞
Measured Value	0.4512	1.668	633.7	15.93M	30.74M	0	0	6.556	∞	∞
% Error	4.30	6.47	1.70	3.01	1.53	–	–	2.86	–	–

As with the previous transistor, the calculated values and measured values are very close, lying within a very small percentage error of one another. There is not a significant difference between the percentage error found in the 2N2222A values and the 2N3904 values.

B. Mid-band Frequency

From **Graph 2.4**, the point at 20 kHz on the curve is about exactly the middle of the band, which is why, again, I chose a mid-band frequency of 20 kHz. Similar to the procedure followed for the 2N2222A, I obtained a Voltage Transfer Curve for the 2N3904:



Graph 2.6: Voltage Transfer Curve

From the graph, it seems that the curve stops being linear about Vs=0.05V.

E. 2N4401

The Values found in part 1, c, iii will be used, along with the given values in problem:

$R_E = 4.3\ k\Omega$ $R_C = R_L = 5.1\ k\Omega$ $R_{B1} = 120\ k\Omega$ $R_{B2} = 62\ k\Omega$ $\beta = 142.86$

$R_S = 50\ \Omega$ $C_{C1} = C_{C2} = C_E = 10\ \mu F$ $R_{BB} = \frac{R_{B1}\times R_{B2}}{R_{B1}+R_{B2}} = 40.88\ k\Omega$ $r_\pi = 3.572\ k\Omega$

R_L was chosen to be equal to R_C to achieve maximum power transfer.

A. Identifying Poles and Zeros

First, we start by finding the low frequency poles and zeros using the Low-Frequency Small-Signal model of the circuit (**Circuit 2.1**), as follows:

$$\omega_{Lp1} = \frac{1}{\tau_{C_{C1}}} = \frac{1}{(R_S + R_{BB}||(r_\pi + R_E(1+\beta))\ \times C_{C1}}$$

$$= 2.604\frac{rad}{sec} = 0.4144 Hz$$

$$\omega_{Lp2} = \frac{1}{\tau_{C_E}} = \frac{1}{(R_C + R_L) \times C_{C2}} = 9.804\frac{rad}{sec} = 1.560\ Hz$$

For C_{C2}, the output stage is decoupled from the input stage:

$$\omega_{Lp3} = \frac{1}{\tau_{C_{C2}}} = \frac{1}{\frac{(R_S||R_{BB} + r_\pi)}{1+\beta}\ ||R_E \times C_E} = 3.995\ k\frac{rad}{sec} = 635.8\ Hz$$

The low frequency zeros due to C_{C1} and C_{C2} are 0, and the third zero can be calculated as follows:

$$\omega_{Lz3} = \frac{1}{R_E \times C_E} = 23.26\frac{rad}{sec} = 3.701 Hz$$

Firstly, we will start by applying Miller's theorem to our Small-Signal Model of the circuit (**Circuit 2.2**).

The value of C_μ and C_π, were calculated using the graphs provided in the datasheet, using the operating point previously found in part I:

$$V_{BE} = 0.7\ V \qquad V_{CB} = 5\ V$$

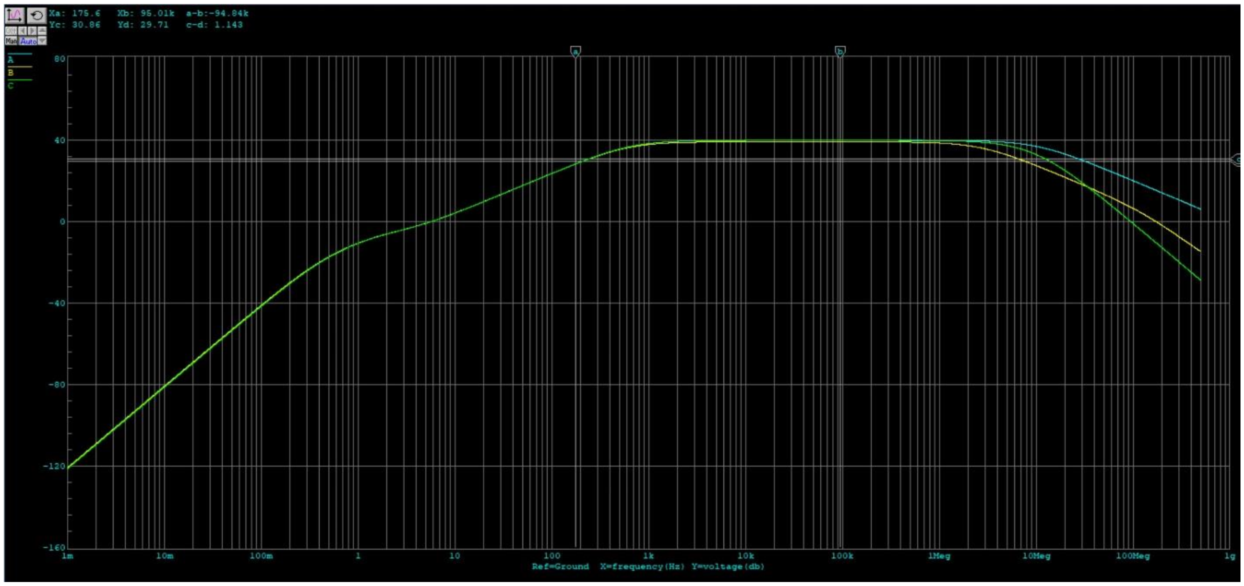
These voltages were traced on the graph to be $C_\mu = 4.5\ pF$ and $C_\pi = 19\ pF$. These were then used to calculate the Miller capacitances as shown in **Circuit 2.2**.

To find the High-Frequency poles, Miller's theorem was used to find applied on the input and output of **Circuit 2.2.**

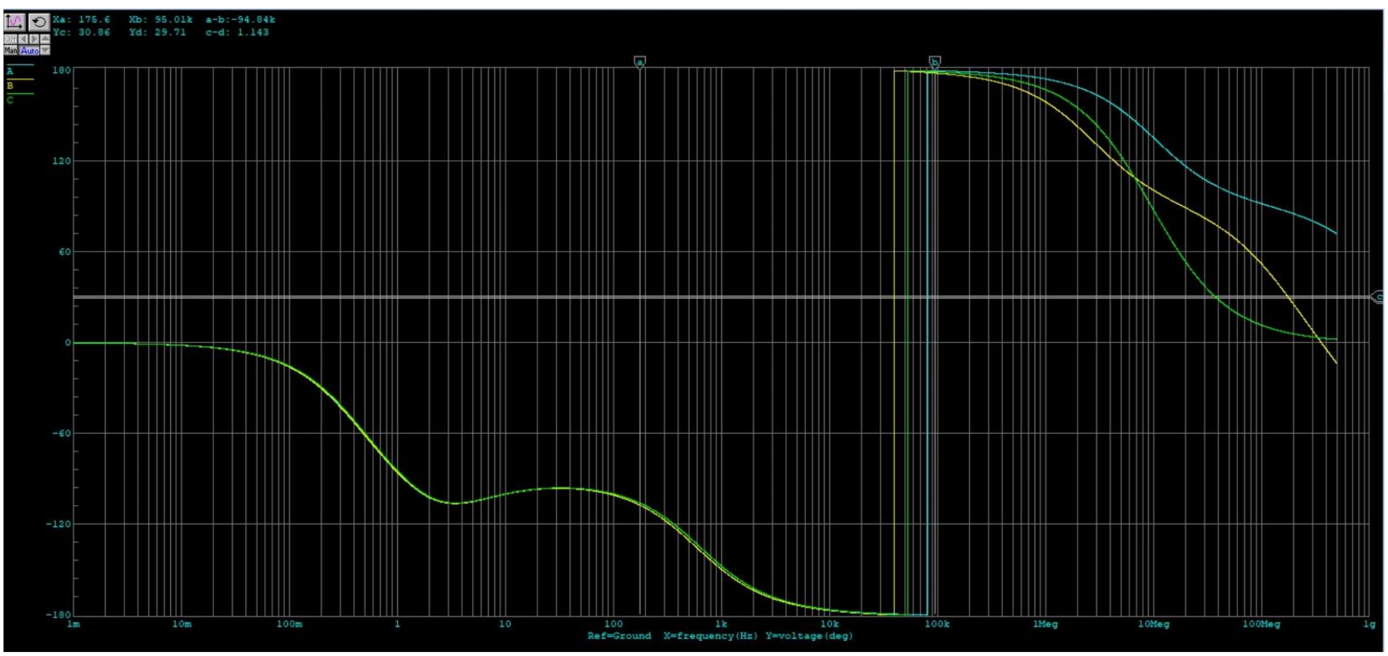
$$\omega_{Hp1} = \frac{1}{\tau_{C_{473.8+4.7}}} = \frac{1}{(R_S\big|\big|R_{BB}\big|\big|r_\pi)\ \times (482.5)pF} = 42.08\ M\ \frac{rad}{sec} = 6.698\ MHz$$

$$\omega_{Hp2} = \frac{1}{\tau_{C_\mu}} = \frac{1}{(R_C||R_L)\ \times 4.5\ pF} = 87.15M\ \frac{rad}{sec} = 13.87\ MHz$$

The high frequency zeros of this circuit are at infinity.



Graph 2.7: Bode Magnitude plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green) for the 2N4401



Graph 2.8: Bode Phase Plot comparing the Original Circuit (Yellow), Small Signal Model (Blue) & Miller's Approximation of the circuit (Green)

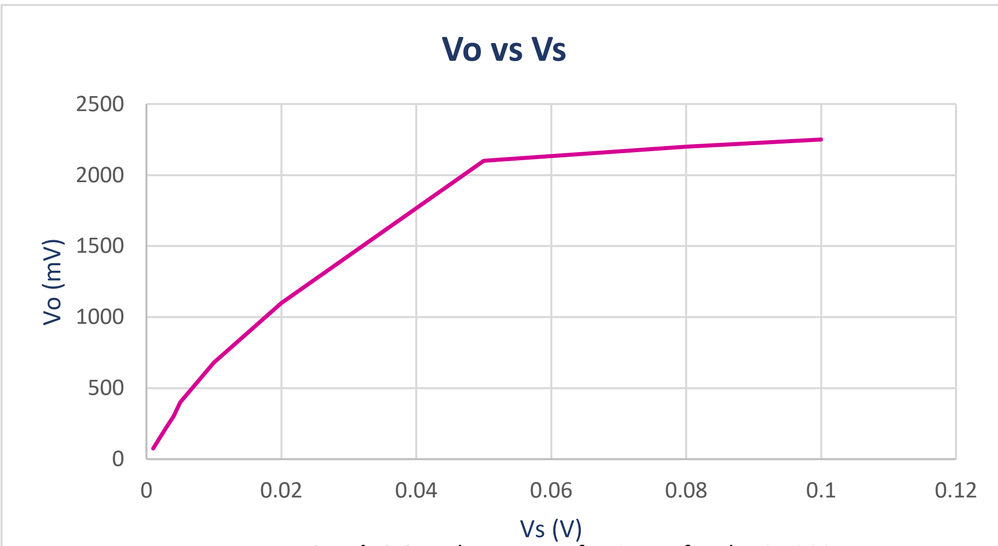
Table 2.3: Table comparing the values of the poles and zeros (Hz) obtained graphically and by calculation:

Pole/Zero	ω_{Lp1}	ω_{Lp2}	ω_{Lp3}	ω_{Hp1}	ω_{Hp2}	ω_{Lz1}	ω_{Lz2}	ω_{Lz3}	ω_{Hz1}	ω_{Hz2}
Calculated Value	0.4144	1.560	622.9	6.698M	13.87M	0	0	3.701	∞	∞
Measured Value	0.4213	1.593	631.2	6.557M	13.88M	0	0	3.512	∞	∞
% Error	1.64	2.07	1.31	2.15	0.07	–	–	5.38	–	–

The trend continues, and the calculated values and measured values are very close for this transistor too, as demonstrated by the percentage error.

B. Mid-band Frequency

From **Graph 2.8**, the point at 20 kHz on the curve is about exactly the middle of the band, which is why, again, I chose a mid-band frequency of 20 kHz. Like the procedure followed for the 2N2222A and the 2N3904, I obtained a Voltage Transfer Curve for the 2N3904:



Graph 2.9: Voltage Transfer Curve for the 2N3904

From the graph, it seems that the curve stops being linear about Vs=0.05V.

Selecting the Best Performing Transistor: 2N3904

Judging from the Saturation Voltages found above, the 2N3904 and the 2N4401 have a higher saturation voltage and are therefore much better in performance than the 2N2222A. This is because the 2N2222A transistor will reach its saturation point the fastest out of all 3. After observing the datasheets for the remaining two transistors, both transistors are more or less the same, except that the 2N4401 has a higher current rating, and so the 2N3904 is a better transistor as it can operate more efficiently at low currents than the 2N4401. It is also widely used and commonly available.

Part III

Note: In this section, calcualtions relating to poles and zeros shown in this report are kept concise as they were covered in Mini-project 1.

Figure 3.1 shows the Common-Base Amplifier circuit configuration which we will simulate in this section.

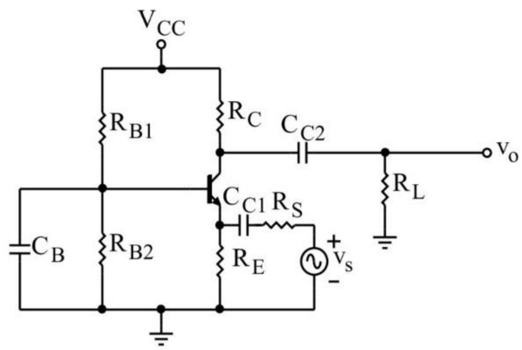


Figure 3.1: The Common-Base Amplifier Configuration

The Values found in part 1, c, iii will be used, along with the given values in problem:

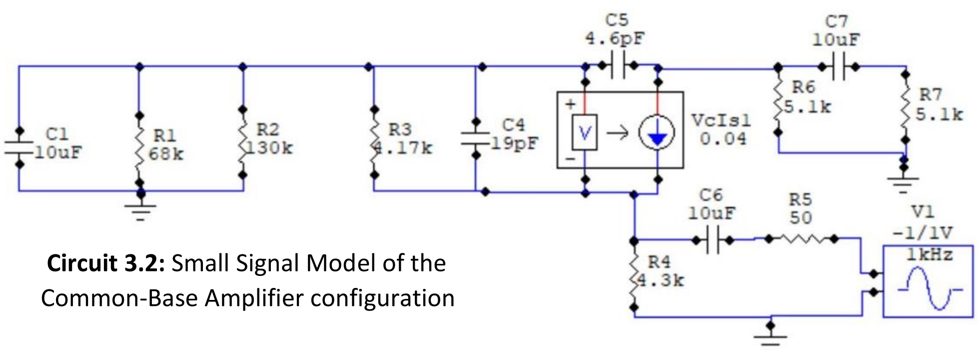
$R_E = 4.3\ k\Omega$ $\quad R_C = R_L = 5.1\ k\Omega$ $\quad R_{B1} = 130\ k\Omega$ $\quad R_{B2} = 68\ k\Omega$ $\quad \beta = 166.67$

$R_S = 50\ \Omega$ $\quad C_{C1} = C_{C2} = C_E = 10\ \mu F$ $\quad R_{BB} = \frac{R_{B1} \times R_{B2}}{R_{B1} + R_{B2}} = 44.65\ k\Omega$ $\quad r_\pi = 4.3\ k\Omega$

R_L was chosen to be equal to R_C to achieve maximum power transfer.

A. Identifying Poles and Zeros

First, we start by finding the low frequency poles and zeros using the Low-Frequency Small-Signal model of the circuit (**Circuit 2.1**), as follows:



Circuit 3.2: Small Signal Model of the Common-Base Amplifier configuration

$$\omega_{Lp1} = \frac{1}{\tau_{C_B}} = \frac{1}{(R_{BB}||(r_\pi + (1+\beta)R_E) \times C_B} = 2.378 \frac{rad}{sec} = 0.3784\ Hz$$

$$\omega_{Lp2} = \frac{1}{\tau_{C_{C2}}} = \frac{1}{(R_C + R_L) \times C_{C2}} = 9.804 \frac{rad}{sec} = 1.560\ Hz$$

$$\omega_{Lp3} = \frac{1}{\tau_{C_{C1}}} = \frac{1}{(R_S + (r_\pi || \frac{R_E}{1+\beta}))\ \times C_{C1}} = 1.324\ k \frac{rad}{sec} = 210.8\ Hz$$

$$\omega_{Lz1} = \omega_{Lz2} = 0$$

$$\omega_{Lz3} = \frac{1}{R_{BB} \times C_B} = 2.240 \frac{rad}{sec} = 0.3565\ Hz$$

The value of C_μ and C_π, were calculated using the graphs provided in the datasheet, using the operating point previously found in part I:

$$V_{BE} = 0.7\ V \qquad V_{CB} = 5\ V$$

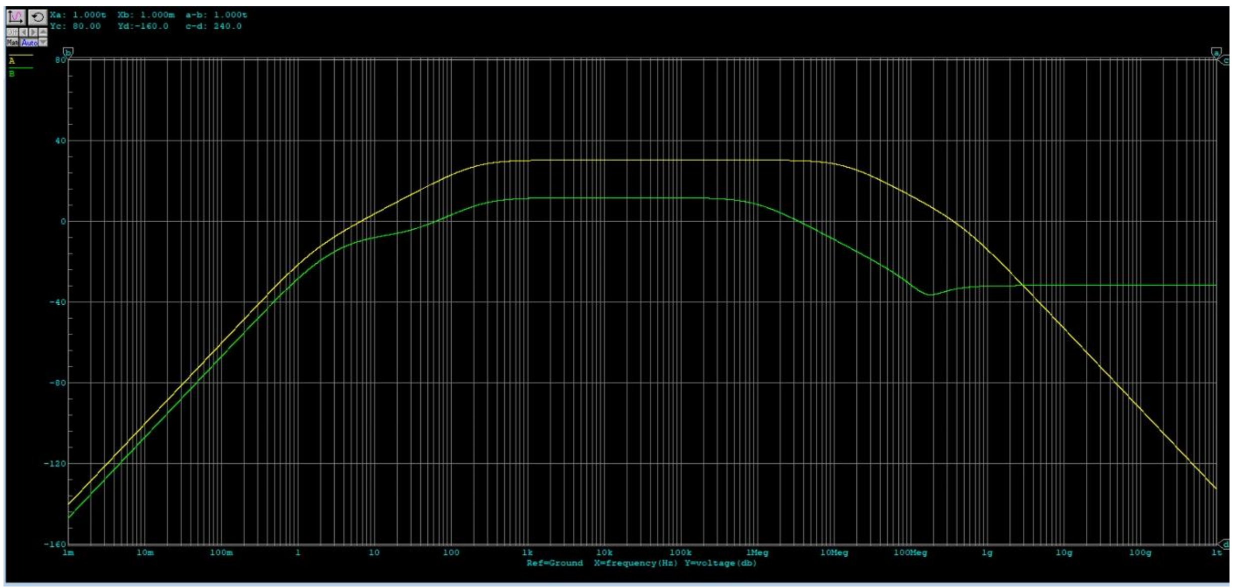
These voltages were traced on the graph to be $C_{\mu} = 4.7\ pF$ and $C_{\pi} = 19\ pF$. These were then used to calculate the Miller capacitances as shown in **Circuit 2.2**.

To find the High-Frequency poles, Miller's theorem was used to find applied on the input and output of **Circuit 2.2.**

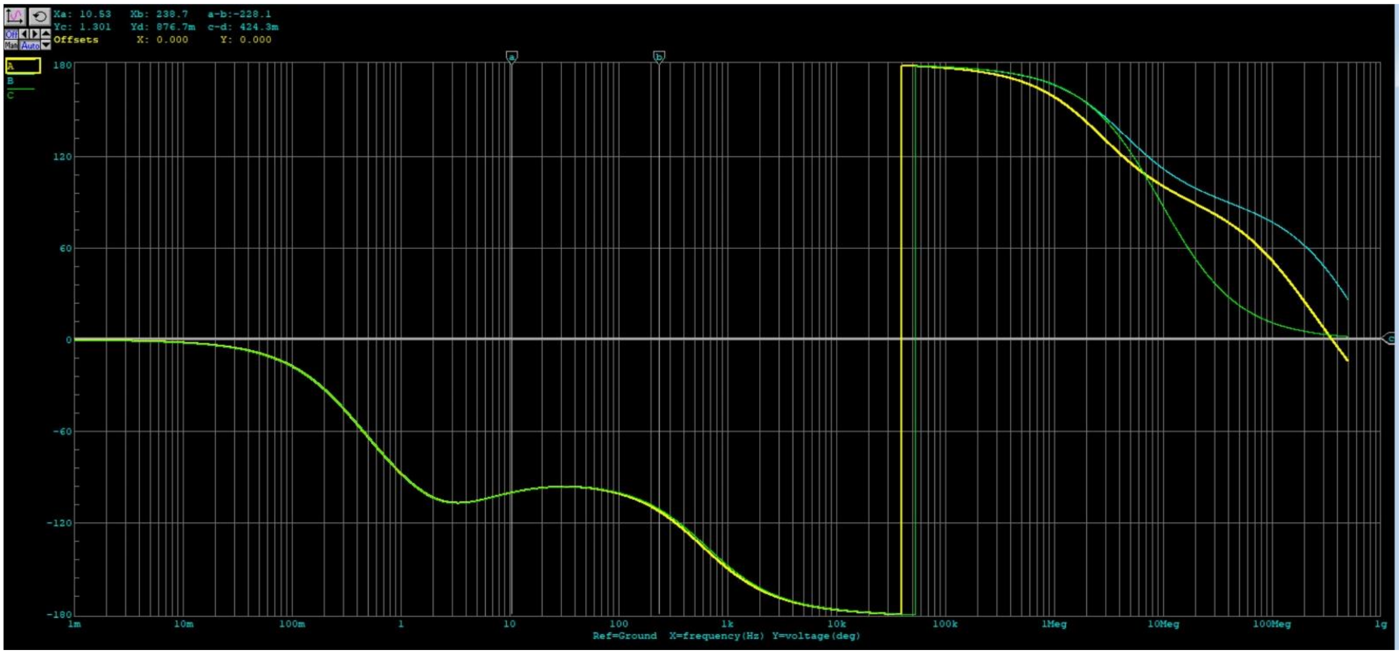
$$\omega_{Hp1} = \frac{1}{\tau_{C_{\mu}}} = \frac{1}{C_{\mu}\ \times R_{BB}} = 4.765M\ \frac{rad}{sec} = \ 758.4\ kHz$$

$$\omega_{Hp2} = \frac{1}{\tau_{C_{473.8+4.7}}} = \frac{1}{(R_S||r_{\pi}||\frac{R_E}{1+\beta})\ \times 19pF} = 3.117\ G\ \frac{rad}{sec} = 496.1\ MHz$$

The high frequency zeros of this circuit are at infinity.



Graph 3.1: Bode Magnitude plot comparing the Original Circuit (Green), Small Signal Model (yellow)



Graph 3.2: Bode Phase plot comparing the Original Circuit (Green), Small Signal Model (yellow)

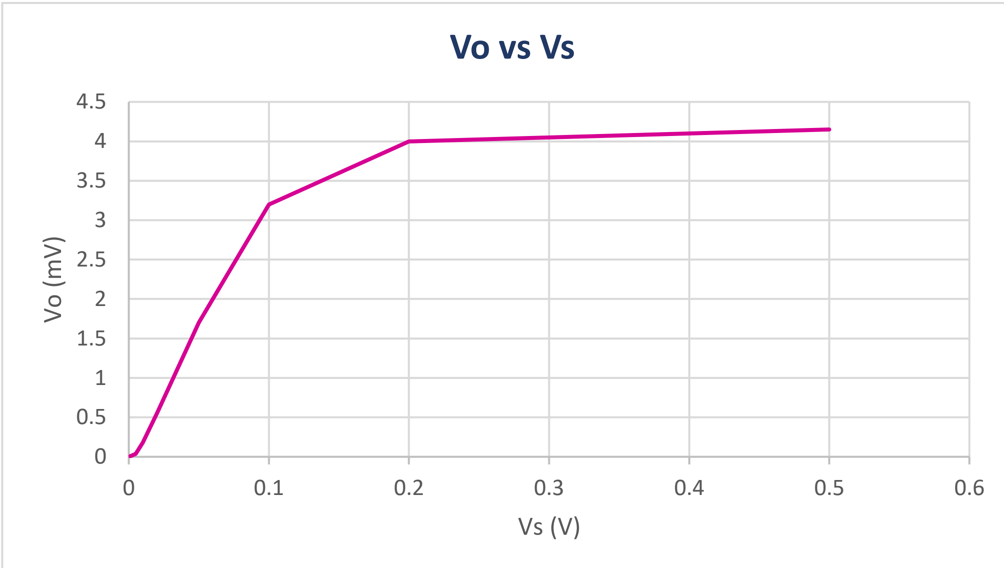
<u>**Table 3.1:** Table comparing the values of the poles and zeros (Hz) obtained graphically and by calculation:</u>

Pole/Zero	ω_{Lp1}	ω_{Lp2}	ω_{Lp3}	ω_{Hp1}	ω_{Hp2}	ω_{Lz1}	ω_{Lz2}	ω_{Lz3}	ω_{Hz1}	ω_{Hz2}
Calculated Value	0.3784	1.560	210.8	758.4k	496.1M	0	0	0.3565	∞	∞
Measured Value	0.3671	1.670	213.4	732.1k	494.4M	0	0	0.3332	∞	∞
% Error	3.08	6.59	1.22	3.59	0.34	–	–	6.99	–	–

As shown in the table, the calculated values and measured values are very close, and lie within a very small percentage error of one another, which makes the small signal model an accurate method to approximate the poles and zeros of the transistor.

B. Mid-band Frequency

From **Graph 3.1**, the point at 20 kHz on the curve is about exactly the middle of the band, which is why I, again chose a mid-band frequency of 20 kHz. I then proceeded to vary the source peak voltage from 0.1 to 1, which provided a reasonable range of data points, giving the desired shape of the Voltage Transfer Curve.



Graph 3.3: Voltage Transfer Curve

From the graph, it appears that the graph stops being Linear at Vs = 0.5V.

C. Measuring Input Impedance

I used the Input stage from **Circuit 3.2** to calculate the Input impedance as follows:

$$R_{in} = R_E || \frac{r_\pi}{1+\beta} = 25.49\ \Omega$$

Then, I measured the Input Impedance by simulating the circuit, giving me the value:

$$R_{in} = \frac{V}{I} = \frac{245.2\ mV}{9.674\ mA} = 25.34\ \Omega$$

The calculated Impedance matches the measured impedance, with a very negligible difference. This indicates that the approximation we made in our calculation is accurate.

D. Measuring Output Impedance

I used the output stage from **Circuit 3.2** to calculate the Output impedance as follows:

$$R_{out} = R_C = 5.1\ k\Omega$$

Then, I measured the Output Impedance by simulating the circuit:

$$R_{out} = \frac{V}{I} = \frac{2.485\ V}{487.3\ \mu A} = 5.1\ k\Omega$$

The calculated Output Impedance is the exact same value as the measured Output Impedance, which is what was expected.

Conclusion

The goal of this assignment is to explore and become familiar with hybrid-pi models and biasing of two important transistor configurations; the common-emitter and the common-base configurations. Throughout this assignment it became clearer why some transistors are better than others for specific applications.

For example, as mentioned earlier, the 2N3904 is a better performing transistor for the purpose of an amplifier, as it has a lower current rating than the 2N4401. Nevertheless, if we were to use a transistor as a switch, the 2N4401 may be a better option, since it achieves saturation mode faster.

Another important observation made was that the Common-Base Amplifier is a much a better Amplifier than the Common-Emitter since it has a much higher voltage gain. Moreover, the estimations made using the hybrid-pi model proved to be accurate, as they fall within a small range of error from the actual measured value, which can be neglected. Nonetheless, the accuracy of the measurements made graphically throughout the project could be improved by using a more accurate and up-to-date simulation software.

Finally, determining the best performing transistor depends on a lot of factors, most importantly the application it is intended for. Although I stated earlier that the 2N3904 is the best performing out of all three transistors in this project, the other two transistors could be better performing for other applications.

Aside: This project was quite long, and it is currently 4:58 am.