Mini Project 2

Single Transistor Amplifiers

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1 Introduction

The objective of this mini project is to investigate the methods used to model, analyize, and bias a transistor and or amplifier circuit. We will examine 3 commonly used NPN Bipolar junction transistors. Mainly: 2N2222A, 2N3904, 2N4401. All of which can be modelled in the circuit simulation software *CircuitMaker*.

2 Investigation

2.1 Part 1

2.1.1 A - Datasheet Lookup

First, we look up the necessary small-signal "H-parameters" of the 2N2222A transistor from its associated datasheet downloaded from the web.

We are interested in finding h_{fe}, h_{ie}, and h_{oe}. The following table contains the values for V_{CE}=10V, I_C=1mA, f=1kHz, and T=25°C.

Parameter	Description	Min	Max
h_{fe}	Small Signal Current Gain	50	300
h_{ie}	Input Impedence	2 kΩ	8 kΩ
h_{oe}	Output Impedence	5 μS	35 μS

Table 1: Datasheet values for small-signal value for 2N2222A

For the sake of consistency, we will use the mean of minimum and maximum values. So h_{fe}=175, h_{ie}=5kΩ, and h_{oe}=20μS or 50kΩ.

2.1.2 B - Characteristics

For this section, we obtain the "measured" values by simulating the circuit in *CircuitMaker* software. To obtain the characteristic curve of the relationship between I_B and V_{BE}, we draw the circuit in figure 1 and perform a DC sweep simulation. The simulation will sweep the V_{BE} source from 0V to 3V at a step of 20mV. The corresponding output graph is shown in figure 2.

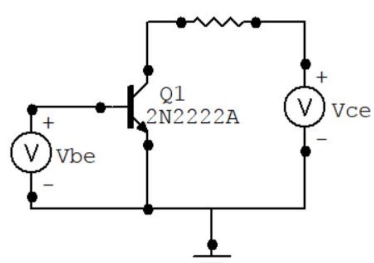


Figure 1: Circuit for finding I_B vs. V_{BE}

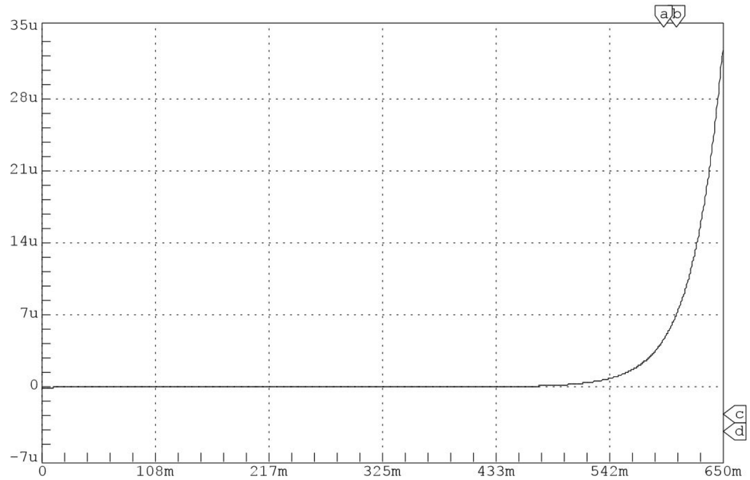


Figure 2: I_B vs. V_{BE} graph

Now, we observe the characteristics between I_C and V_{CE} with varying I_B. In this case, we build the circuit as shown in figure 3 and operate DC sweep on two sources. One is the current source I_B, from 1μA to 10μA with 1μA step. And second, the voltage source V_{CE} from 0V to 6V with a step of 20mV. The resulting graph is shown in figure 4.

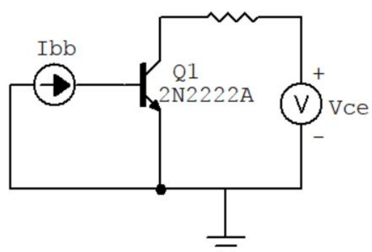


Figure 3: Circuit for finding I_C vs. V_{CE} with varying I_B

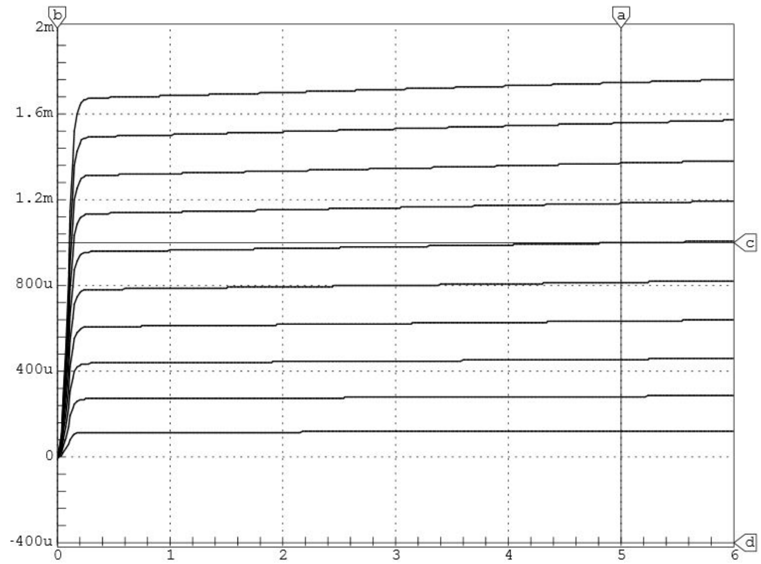


Figure 4: I_C vs. V_{CE} with varying I_B

In figure 4, the bottom most curve represents the characteristic where I_B=1μA/ The top most curve represents where I_b=10μA.

Now that we have obtained the characteristic curves, we can compute β, r_π, and r_o.

It's clear that at V_{CE}=5V and I_c=1mA, the collector current is I_B=6μA, as shown in figure 4. Using the fact that $I_C = \beta I_B$, we solve for β, and $\beta = \boxed{166.7}$.

Next, to find r_π, we must first compute the transconductance gain g_m with V_T=25mV for T=25°C.

$$g_m = \frac{I_C}{V_T} = \frac{1\text{mA}}{25\text{mV}} = 0.040\text{S}$$

Using the plot of I_C vs. V_{BE}, we can verify that our g_m is adequate since $g_m=\frac{I_C}{V_{BE}}$. As shown in figure 5, the slope at which I_C=1mA is approximately $\frac{450\mu\text{A}}{11.8\text{mV}} = 0.038$, which is close enough to 0.040 but we will use 0.038.

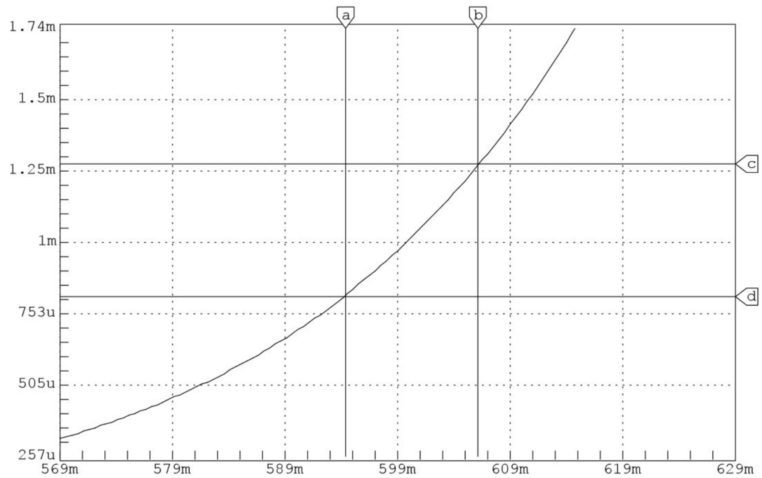


Figure 5: I_C vs. V_{BE}, where the slope is g_m

$$r_\pi = \frac{\beta}{g_m} = \frac{166.7}{0.040\text{mS}} = \boxed{4167\Omega}$$

Finally, to find r_o, we first need to find the early voltage V_A. We can compute this by obtaining the slope of the active region curves in the I_C vs. V_{CE} graph. Taking two sample points of the I_B=10μA curve at V_{CE}=1V and 6V. The difference in collector current is 76μA. Thus, the slope is $\frac{6-1\text{V}}{76\mu\text{A}} = 15.2\mu\text{S}$.

We find the line equation (in the form of $y = mx + b$) of the linear (active) component to be $y = (1.520 \times 10^{-5})x + 1.673 \times 10^{-3}$. Extrapolating the curve and solve for I_C=0, we find that $-V_A = -110\text{V}$. Thus, V_A= $\boxed{110\text{V}}$.

It follows that

$$r_o \approx \frac{V_A}{I_C} = \frac{110\text{V}}{1\text{mV}} = \boxed{110\text{k}\Omega}$$

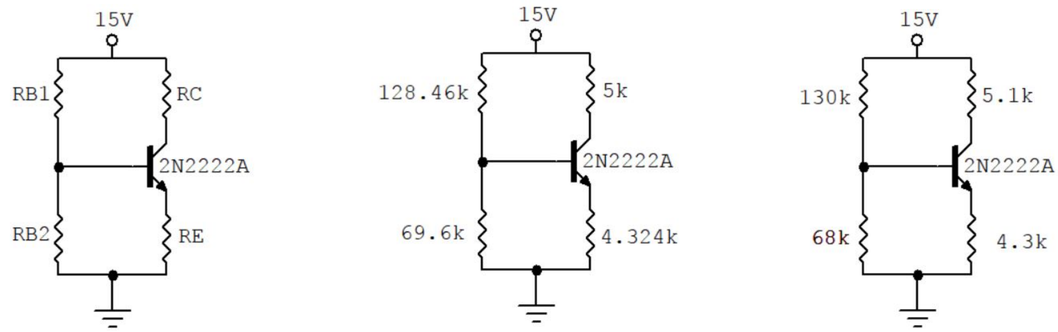
Note that V_{BE} is approximately 0.6V at DC operating point, we will use this value in future calculations.

Comparing Measured Values to Datasheet Values

The calculated (from measurements) small signal current gain is approximately 167. This is reasonable since the datasheet specifies the range of 50-300. Other parameters such as V_{BE} and g_m is adequate as well when compared to the curves in the datasheet.

2.1.3 C - Biasing

The following circuit in figure 6a is used to bias the BJT. Where V_{CC}=15V and I_C=1mA.



(a) Bias circuit (b) Using 1/3 rule (c) Using standard resistors

Figure 6: Bias circuits

Biasing from Measured Plot
Recall the parameters obtained from the plots:

$$\beta = 167, \quad \mathrm{V_{BE}} = 0.6\mathrm{V},$$

Also given that:

$$\mathrm{V_{CE}} = 5\mathrm{V}, \quad \mathrm{R_E} = \frac{1}{2}\mathrm{R_C}$$

We first compute all the currents.

$$I_C = 1\mathrm{mA}, \quad I_B = \frac{I_C}{\beta} = \frac{1\mathrm{mA}}{167} = 6\mu\mathrm{A}, \quad I_E = I_B + I_C = 1.006\mathrm{mA}$$

Next, doing mesh analysis from $\mathrm{V_{CC}}$ to emitter ground, we get the following equation. Plugging in $\mathrm{R_E}=\frac{1}{2}\mathrm{R_C}$ and all currents, we evaluate for $\mathrm{R_C}$ and $\mathrm{R_E}$.

$$\begin{aligned} 15 &= I_C R_C + 5 + I_E R_E \\ 10 &= I_C R_C + I_E \left(\frac{1}{2} R_C\right) \\ &= (1\mathrm{mA})(R_C)\left(1 + \frac{1}{2}\right) \\ \implies R_C &= \boxed{6.667\mathrm{k}\Omega} \end{aligned}$$

$$\begin{aligned} R_E &= \frac{1}{2} R_C \\ &= \frac{6.667\mathrm{k}\Omega}{2} \\ R_E &= \boxed{3.333\mathrm{k}\Omega} \end{aligned}$$

Then we calculate the voltage.

$$\begin{aligned} V_E &= R_E I_E = (1\mathrm{mA})(3.333\mathrm{k}\Omega) = \boxed{3.333\mathrm{V}} \\ V_C &= V_E + V_{CE} = 3.333\mathrm{V} + 5\mathrm{V} = \boxed{8.333V} \\ V_B &= V_E + V_{BE} = 3.333\mathrm{V} + 0.6\mathrm{V} = \boxed{3.933\mathrm{V}} \end{aligned}$$

To find R_{B_1} and R_{B_2}, we need two equations. The first equation comes from doing a mesh analysis from V_{BB} to ground.

$$V_{BB} = R_{BB} I_B + V_B$$

Where $V_{BB} = \frac{R_{B_2}}{R_{B_1}+R_{B_2}} V_{CC}$, $R_{BB} = R_{B_1} \| R_{B_2}$, and V_B=3.933V.

The second equation is from doing KCL at the node V_B.

$$I_1 = I_B + I_2$$

Where $I_1 = \frac{15\text{V}-3.933\text{V}}{R_{B_1}}$, $I_2 = \frac{3.933\text{V}}{R_{B_2}}$, and $I_B = 6\mu A$.

Thus,

$$\begin{cases} \frac{R_{B_2}}{R_{B_1}+R_{B_2}}(15\text{V}) = 6\mu\text{A}(R_{B_1} \parallel R_{B_2}) + 3.933\text{V} \\ \frac{11.067\text{V}}{R_{B_1}} = \frac{3.933\text{V}}{R_{B_2}} + 6\mu\text{A} \end{cases}$$

As it turns out, these two equations does not demonstrate a linear relationship. So the resistances cannot be solved using systems of equations. Thus, I picked R_{B_1} to be 1.2MΩ. Plugging into one of the equations above, we obtain R_{B_2} to be 1.22MΩ. The reason I chose this value is because the resistance is comparatively large, thus there is less power dissipated from these resistors. Also they are more economical as 1.2MΩ is a standard and common resistor value. Using two of the same resistor would allow the supplier to buy in bulk, thus is cheaper.

Table 2 shows the DC operating point for this design.

I_C	I_B	I_E	V_B	V_C	V_E
0.993mA	5.950μA	0.999mA	3.930V	8.381V	4.531V

Table 2: DC operating point with R_{B_1}=R_{B_2}=1.2MΩ resistors

Biasing using 1/3 Rule

In this case, we shall use the first version of the one-third rule. Recall that the first version of the one-third 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, we find the voltages:

$$V_B = \frac{15\text{V}}{3} = \boxed{5\text{V}}$$
$$V_C = \frac{2}{3}15\text{V} = \boxed{10\text{V}}$$
$$V_E = V_B - 0.6\text{V} = \boxed{4.4\text{V}}$$

Using $\beta = 167$ and V_{CE}=0.6V, we find the currents.

$$I_B = \frac{1}{\beta} I_C = \frac{1}{167} 1\text{mA} = \boxed{6\mu\text{A}}$$
$$I_E = I_C + I_B = 1\text{mA} + 6\mu\text{A} = \boxed{1.006\text{mA}}$$
$$I_1 = \frac{I_E}{\sqrt{\beta}} = \frac{1.006\text{mA}}{\sqrt{167}} = \boxed{77.85\mu\text{A}}$$
$$I_2 = I_1 - I_B = 77.85\mu\text{A} - 6\mu\text{A} = \boxed{71.85\mu\text{A}}$$

Finally, calculate resistance, which is just a matter of ohm's law.

$$R_C = \frac{V_{CC} - V_C}{I_C} = \frac{5\text{V}}{1\text{mA}} = \boxed{5\text{k}\Omega}$$
$$R_E = \frac{V_E}{I_E} = \frac{4.4\text{V}}{1.006\text{mA}} = \boxed{4.374\text{k}\Omega}$$
$$R_{B_1} = \frac{V_{CC} - V_B}{I_1} = \frac{10\text{V}}{77.85\mu\text{A}} = \boxed{128.45\text{k}\Omega}$$
$$R_{B_2} = \frac{V_B}{I_2} = \frac{5\text{V}}{71.85\mu\text{A}} = \boxed{69.6\text{k}\Omega}$$

The final bias circuit while using this method is as shown in table 6b. The DC operating point values are as follows in figure 3

I_C	I_B	I_E	V_B	V_C	V_E
1.000mA	5.927μA	1.006mA	5.001V	9.999V	5.602V

Table 3: DC operating point with 1/3 rule and calculated resistors

Using Standard Resistors
In this step, we replace the calculated resistor values with the nearest standard resistor value according to the *Standard Values List*[4]. The resulting circuit is as shown in figure 6c.

The DC operating point using the one third rule is very close to the desired specifications. However, the rounding to nearest standard value introduced some small error. The DC measured values is as shown in table 4.

I_C	I_B	I_E	V_B	V_C	V_E
0.991mA	5.923μA	0.997mA	4.888V	9.946V	5.489V

Table 4: DC operating point with 1/3 rule and standard resistors

Observations

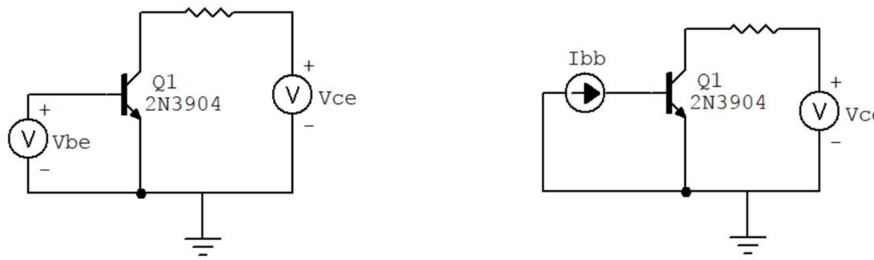
All methods provide an adequate bias circuit. The one third rule, however, is by far, much more versatile in terms of choice of components. It is also much faster to calculate the resistances and still end up with a decent circuit.

2.1.4 D - Different Transistors

Now we swap the 2N2222A with two other transistors, mainly the 2N3904 and the 2N4401 to see how they behave in the same bias circuit. But first, we need to "measure" the transistor parameters via simulation.

2N3094

The original testing circuit has been swapped for the 2N3094 transistor. The circuit setup for simulation is as seen in figure 7. Since the computation and analysis is identical to that of part (b) for 2N222A, I will only show results of calculations.



(a) I_B and I_C vs. V_{BE} sweep circuit (b) I_C vs. V_{CE} and I_B sweep circuit

Figure 7: Test circuits for 2N3904 transistors

For an operating collector current of 1mA, V_{BE}=0.645V. Using same method used for 2N2222A, we find the slope (g_m) of the I_C vs. V_{BE} at V_{BE}=0.645V to be g_m=0.037S.

Next, using the same conditions but using circuit in figure 7b, we find I_B=8.5μA using the characteristic curves similar to figure 4. Finally, given that $I_C=\beta I_B$, β=118 and $r_\pi=\frac{118}{0.037}=3.19\text{k}\Omega$.

Lastly, we plug the 2N3904 transistor into our original bias circuits. In particular, the bias circuit without 1/3 rule and the one with 1/3 rule (as shown in figure 8) and we analyze the DC operating currents and voltages. The results are in table 5.

(a) Bias circuit without 1/3 rule (b) Bias circuit using 1/3 rule

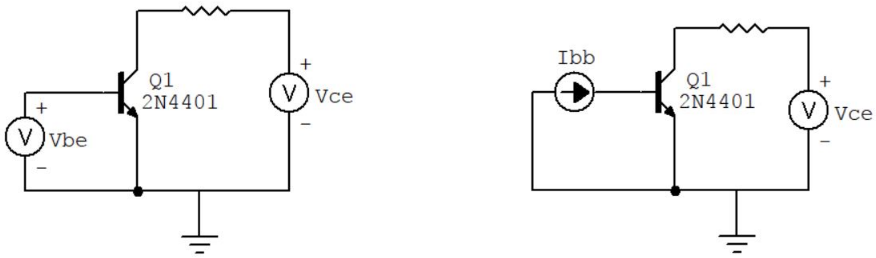
Figure 8: Test circuits for 2N3904 transistors

Bias circuit	I_C	I_B	I_C	V_B	V_C	V_E
no 1/3 rule	0.801mA	6.947μA	0.808mA	3.332V	9.662V	2.692V
with 1/3 rule	0.965mA	8.161μA	0.973mA	4.903V	10.170V	4.258V

Table 5: 2N3904 transistor DC operating points comparison

2N4401

The transistor in the circuit has once again been swapped, this time with the 2N4401 transistor. The circuit used for simulation and analysis (similar to 2N222A and 2N3904) is as follows (figure 9).



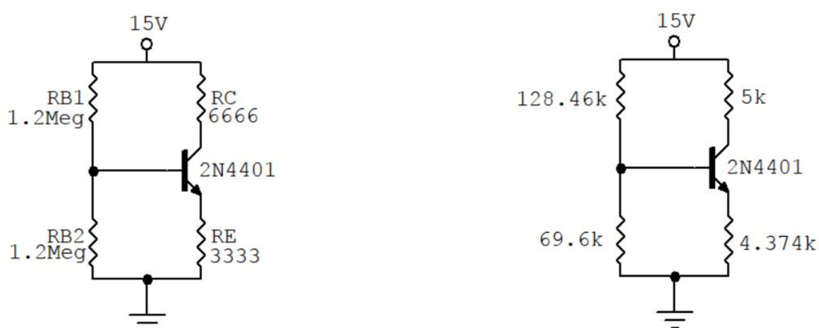
(a) I_B and I_C vs. V_{BE} sweep circuit (b) I_C vs. V_{CE} and I_B sweep circuit

Figure 9: Test circuits for 2N4401 transistors

First, we run the DC sweep simulation for the circuit in figure 9a At I_C=1mA, V_{BE}=0.66V, higher than the previous two transistors. We also find the slope of I_C vs. V_{BE} at I_C=1mA to be g_m=0.038S, as expected.

Moving on to the V_{CE} and I_B DC sweep simulation, using the circuit in figure 9b. At 1_C=1mA and V_{CE}=5V, using the characteristic curves, we find I_B=6.8μA. Thus β for the 2N4401 is 147 and $r_\pi = \frac{147}{0.038} = 3.87\text{k}\Omega$.

Again, pluggin the 2N4401 into the existing biasing circuit (such as in figure 10), we obtain the DC operating points as shown in table 6.



(a) Bias circuit without 1/3 rule (b) Bias circuit using 1/3 rule

Figure 10: Test circuits for 2N3904 transistors

Bias circuit	I_C	I_B	I_C	V_B	V_C	V_E
no 1/3 rule	0.911mA	6.315μA	0.917mA	3.711V	8.930V	3.056V
with 1/3 rule	0.979mA	6.668μA	0.986mA	4.970V	10.100V	4.313V

Table 6: 2N4401 transistor DC operating points comparison

Observation
Despite the bias circuit made without using the 1/3 rule working flawlessly for the 2N2222A transistor; as soon as we switch transistors(parameter change), we see that the bias circuit made using the 1/3 rule has much better tolerance than the former. Therefore, the 1/3 rule is a valid and accurate method of biasing a BJT within a relatively large β or h_{fe} values.

2.2 Part 2

2.2.1 A - 2N2222A Amplifier

We choose the bias circuit as shown in figure 6c and we configure it as an common-emitter amplifier by adding a source and some low frequency 10μF capacitors, as shown in figure 11. We choose R_L, the load resistance to be 5.1kΩ since the output impedance is 5.1kΩ. This allows maximum power transfer.

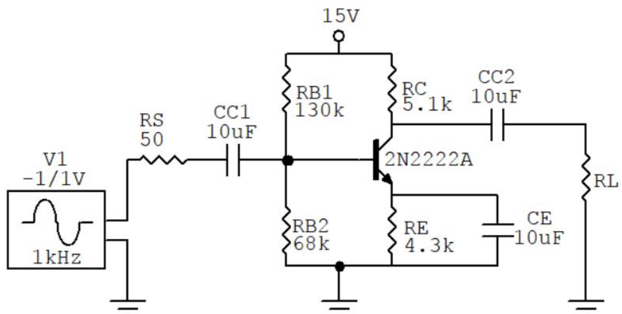
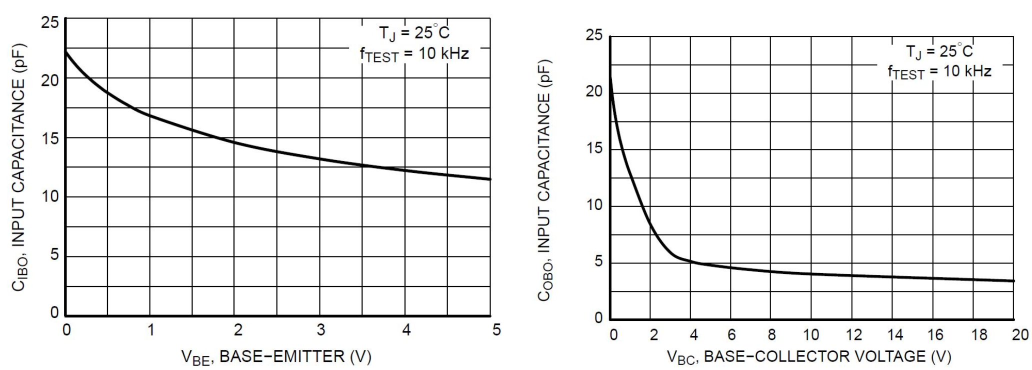


Figure 11: Common-emitter circuit for the 2N2222A

To find the high-frequency parasitic capacitances C_π and C_μ, we observe the curves provided in the 2N2222A datasheet. The said curves are in figure 12. Recall at DC operating point, V_{BE}=0.6V, and V_{CB}=5V, we find that C_π=18pF and C_μ=5pF.



(a) C_π vs. V_{BE} curve (b) C_μ vs. V_{CB} curve

Figure 12: Internal capacitances for the 2N2222A transistor from the datasheet[8]

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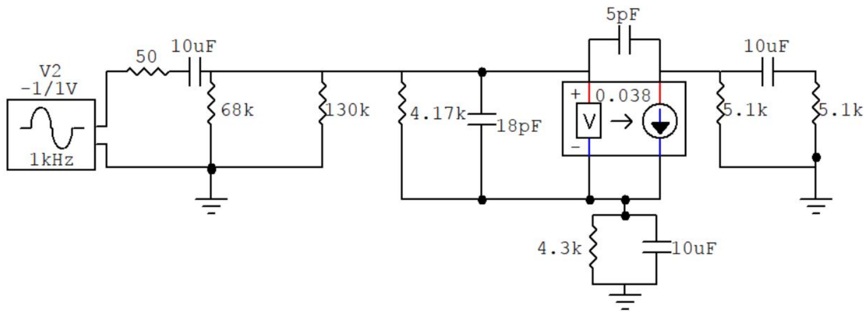
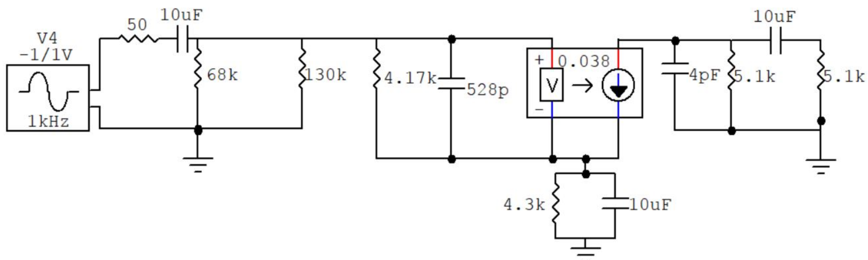
Once we have obtained the internal capacitances of the transistor, we can start modelling BJT.

A small-signal model (figure 13a) and small-signal-miller-equivalence model (figure 13b) is also drawn. Note that we used transconductance gain g_m of 0.038 and R_π of 4.17kΩ as calculated from previous parts.



(a) Circuit for the 2N2222A small-signal model



(b) Circuit for the small-signal model after miller transformation

Figure 13: 2N2222A equivalent circuits

Calculating the Location of the Zeroes and Poles

First, we shall split the small signal model into low frequency response and high frequency response. In low frequency response circuit, we can ignore the two high-frequency internal junction capacitances since they act as open circuit.

For high frequency, the low-frequency 10μF capacitors short out, leaving the two internal junction capacitors behind. Since C_μ is coupling the input and the output, we use Miller equivalence to decouple them. The resulting circuit is as shown in figure 13b. with all 10μF capacitors replaced with a short-circuit (Emitter is grounded).

For Miller equivalence, the coefficient k is $k = -g_m \times (5.1k\Omega \parallel 5.1k\Omega) = -102$. Thus, the two pole locations are calculated as follows.

$$\omega_{H_{P1}} = (528\text{pF} \times (50 \parallel 68\text{k} \parallel 130\text{k} \parallel 4.17\text{k}\Omega))^{-1} = \boxed{6.03\text{MHz}}$$
$$\omega_{H_{P2}} = (4\text{pF} \times (5.1\text{k}\Omega \parallel 5.1\text{k}\Omega))^{-1} = \boxed{15.6\text{MHz}}$$

Next, we calculate the low frequency zeroes and poles. For reference, the low frequency circuit looks like the one in figure 13a except with all the high-frequency capacitors replaced by an open circuit.

There are two zeroes ($\omega_{L_{Z1}}$, $\omega_{L_{Z2}}$) at zeroes due to the two coupling capacitors. The third zero is due to the emitter capacitor. Equating the admittance of the emitter network to 0, we rearrange and find that the location is $\omega_{L_{Z3}} = \frac{1}{R_E C_E} = \frac{1}{4.3\text{k}\Omega \times 10\mu\text{F}} = \boxed{3.7\text{Hz}}$.

The pole for the coupling capacitor at the output is

$$\omega_{L_{P2}} = \frac{1}{10\mu\text{F}(5.1\text{k} + 5.1\text{k})} = \boxed{1.56\text{Hz}}$$

For the other two poles, we shall use short-circuit time constant (SCTC) and open-circuit time constant (OCTC) tests.

Suppose that I want the emitter capacitor to have to conduct at a higher frequency compared to the input coupling capacitor. In other words, by the time the emitter capacitor starts to short, the input coupling capacitor should have already shorted. Thus we perform OCTC test on $\mathrm{C_{C1}}$ and SCTC test on $\mathrm{C_E}$.

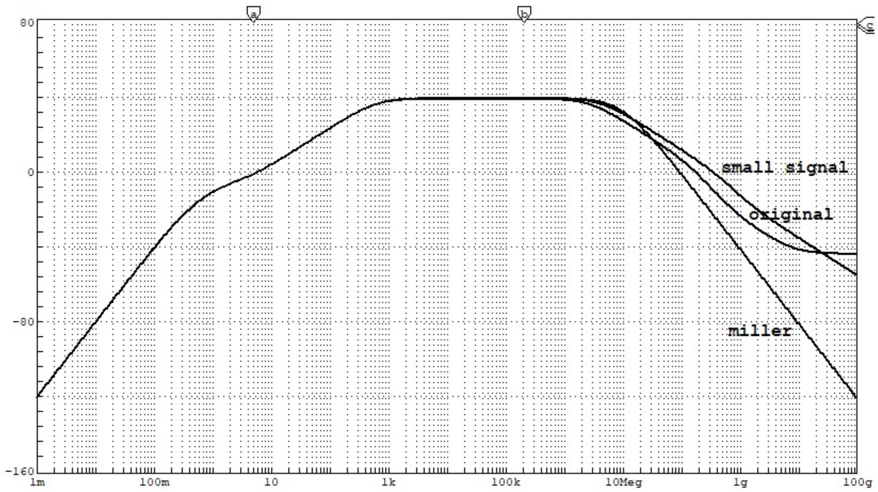
Note that the emitter impedances as seen by the base are magnified by a factor of $1+\beta$; and the base impedances seen by the emitter are demagnified by a factor of $\frac{1}{1+\beta}$. Where β=167. This can be proved by replacing the capacitor with a test source, and finding the test current. But for the sake of the report, we will assume this is true.

$$
\begin{aligned}
\tau_{OC}^{C_{C1}} &= 10\mu\mathrm{F} \times [(68\mathrm{k} \parallel 130k \parallel 4.17k + (1+\beta)4.3\mathrm{k}) + 50] = 0.42 \text{ seconds} \\
\omega_{L_{P1}} &= \boxed{380\mathrm{mHz}} \\
\tau_{SC}^{C_E} &= 10\mu\mathrm{F} \times \left[\left(\frac{1}{1+\beta}\right)(50 \parallel 68\mathrm{k} \parallel 130\mathrm{k} + 4.17\mathrm{k}) \parallel 4.3\mathrm{k}\right] = 0.25 \text{ milliseconds} \\
\omega_{L_{P3}} &= \boxed{637\mathrm{Hz}}
\end{aligned}
$$

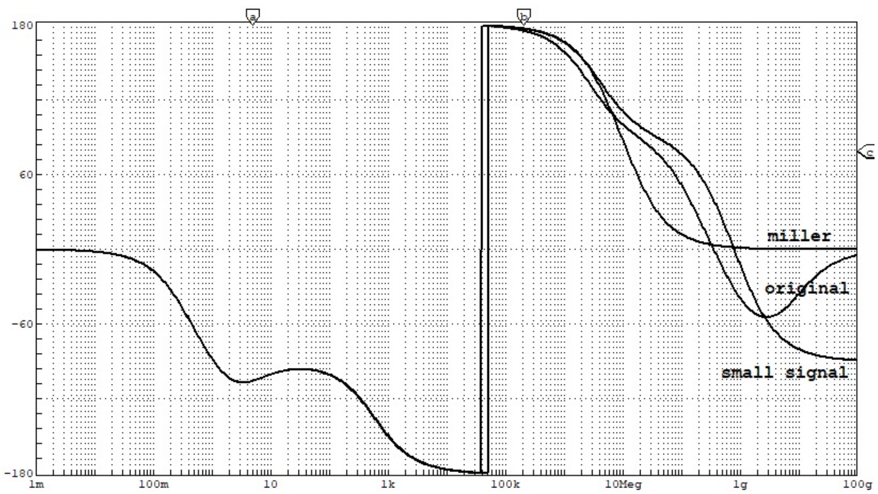
Indeed, the frequencies at which the capacitors transition make sense.

Bode Plots

Doing the AC sweep in the CircuitMaker simulation, we obtain the magnitude and phase bode plots as seen in figure 14. Notice that when we switched from the original circuit to the small-signal mode, the approximation lost a high-frequency pole. When we went from small-signal to miller equivalence, we lost another high-frequency pole.



(a) Magnitude bode plot



(b) Phase bode plot

Figure 14: 2N2222A amplifier bode plots

Comparing Poles and Zeroes to the Plot

Using the method similar to the one demonstrated in Mini-Project 1, we use linear approximation on the bode plot. Figure 15 depicts the process of approximating the poles and zeroes graphically. Regardless, the result of the poles and zeroes is as follows (Table 7). Refer above for calculations of the calculated locations of poles and zeroes.

	ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
Graphed	0	0	3.79	0.508	1.28	651	497M	8.93G	3.02M	134M
Calculated	0	0	3.7	0.380	1.56	637	∞	∞	6.03M	15.6M

Table 7: Poles and zeroes locations (Hz)

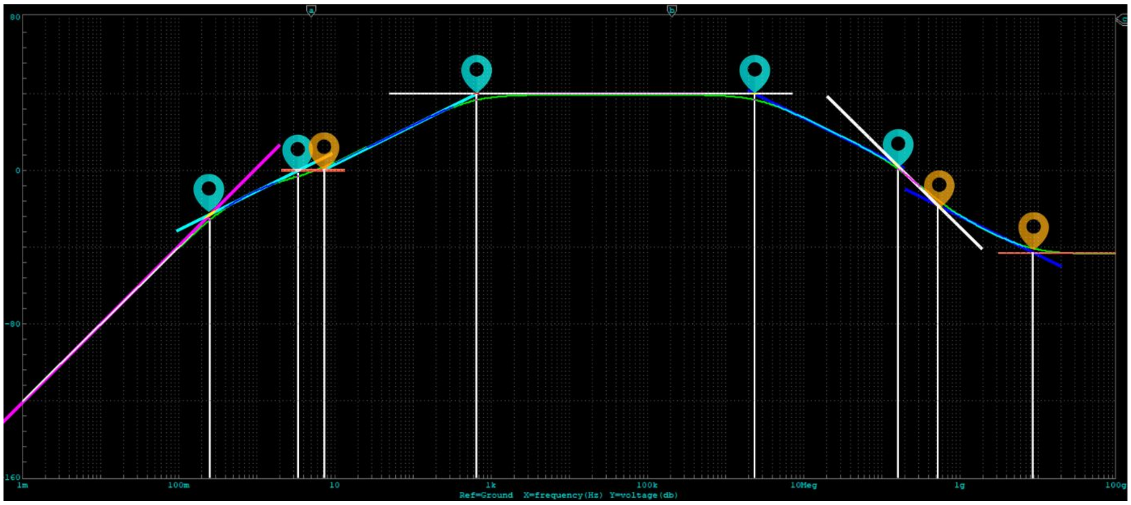


Figure 15: Finding locations of poles and zeroes graphically using linear approximation

Comparison

Because of the Miller transformation, the high-frequency poles got shifted right, thus increasing the bandwidth. It is noted before that the high-frequency zeroes disappear - or get moved to ∞, this is true as the calculated zeroes of the high-frequency capacitors tends to ∞. The two poles at low-frequency are slightly off because the poles are relative close to each other, thus the calculated value is becomes a rough estimate.

2.2.2 B - Varying Source Amplitude

Recall, from the bode plot from the previous part, the midband bandwidth consists from approximately 1kHz to 2MHz. For this part and future parts, We choose the midband frequency of 20kHz. Since it is near the low-end of the midband. This frequency is about the maximum frequency a human can hear.

By varying the source voltage amplitude (V_S) and measuring the amplitude of the output voltage (V_O). We get the following characteristic plot, shown in figure 16.

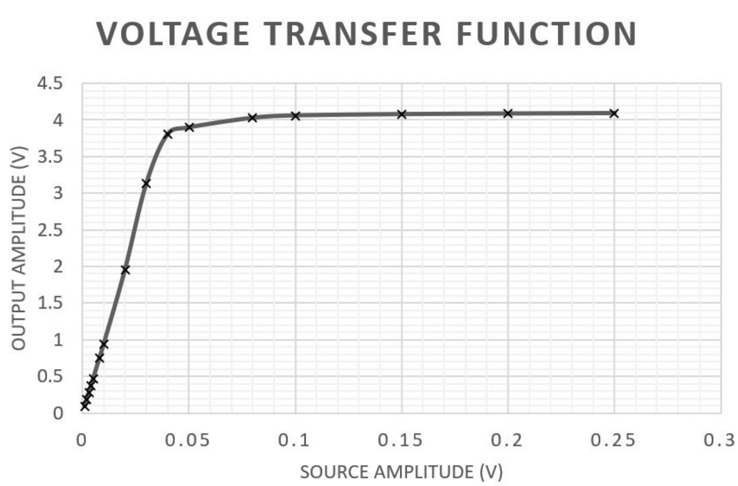


Figure 16: The voltage transfer (V_O vs. V_S)

We see the transfer function transitions from being linear to non-linear at around 0.01V. This is due to the output signal being saturated. Thus, we will use 0.01V for the remainder of the measurements.

2.2.3 C - Input Impedance

Calculated Input Impedance

As one could see the equivalent circuits for the 2N2222A amplifier in figure **??**, the input impedance is simply $\mathrm{R_{BB}} \parallel \mathrm{r}_\pi$. Where $\mathrm{R_{BB}} = \mathrm{R_{B1}} \parallel \mathrm{R_{B2}}$. This is because at midband, the emitter capacitor $\mathrm{C_E}$ shorts and thus the emitter is grounded.

$$R_{in} = R_{B1} \parallel R_{B2} \parallel r_\pi$$
$$R_{in} = \boxed{3.81\mathrm{k}\Omega}$$

Measured Input Impedance

At 10mV source amplitude and 20kHz source frequency. We can measure the current and the voltage since $\mathrm{R_{in}}=\frac{I_{in}}{V_B}$. Using the probe in *CircuitMaker*, we see that $\mathrm{I_{in}}=1.738\mu\mathrm{A}$ and $\mathrm{V_B}=6.992\mathrm{V}$. Therefore, the measured input impedance is:

$$R_{in} = \frac{6.992\mathrm{V}}{1.738\mu\mathrm{A}} = \boxed{4023\Omega}$$

Which is expected.

2.2.4 D - 2N3904 Amplifier

For the 2N3904, a similar circuit is set up for the transistor similar to that of figure 11, except with the transistor replaced.

Recall at the DC operating point using the same bias circuit (from part 1), $\mathrm{V_{BE}}=0.645\mathrm{V}$ and $\mathrm{V_{CB}}=5.27\mathrm{V}$. Thus, from the datasheet for 2N3904 (figure 17), we obtain the values for $\mathrm{C}_\pi=3.5\mathrm{pF}$, and $\mathrm{C}_\mu=2\mathrm{pF}$. Some other information from part 1 that are useful: $\mathrm{g_m}=0.037$, $\beta=118$, and $\mathrm{r}_\pi=3.19\mathrm{k}\Omega$.

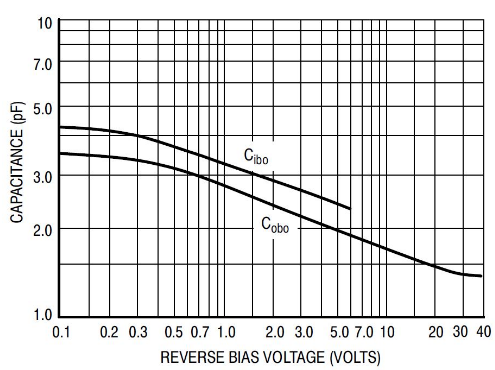


Figure 17: Internal capacitances for the 2N3904 [9]

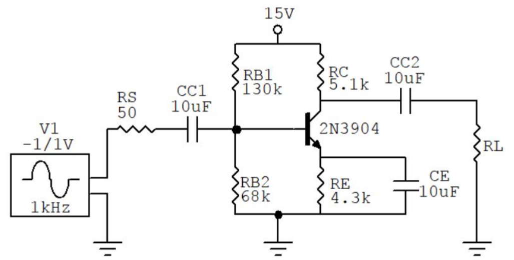
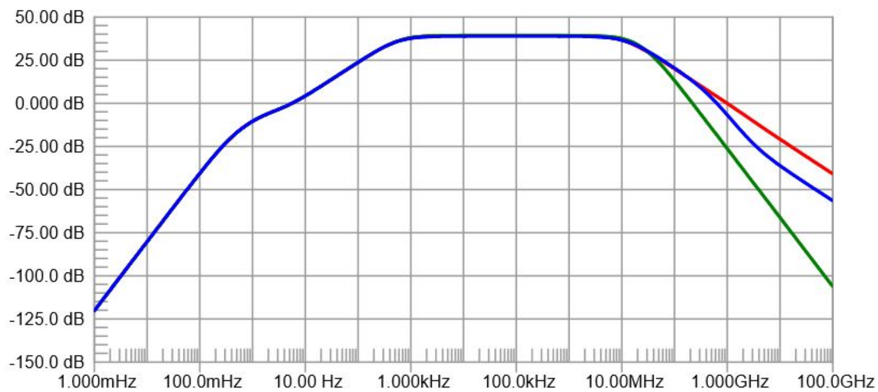
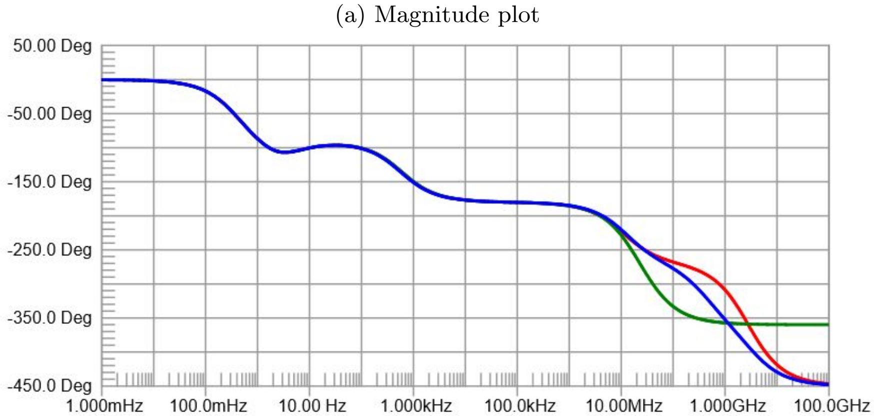


Figure 18: Amplifier circuit for the 2N3904

The circuit is drawn as shown in figure 18. The bode of the magnitude and phase is as shown in figure 19. The blue plot represents the unmodified amplifier circuit, the red represents the small signal model, and the green represents the miller-equivalence model.



(a) Magnitude plot



(b) Phase plot

Figure 19: Bode plots for the 2N3904 amplifier

.

The same algorithm or method is used to calculate the approximate pole locations. For more details, refer to the previous section where the poles and zeroes locations are calculated for the 2N2222A. The results are in table 8. The high-frequency poles are inaccurate possibly due to approximations such as small-signal and Miller transformation.

	ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
Graphed	0	0	5.6	0.35	1.4	650	-	-	14.5M	516M
Calculated	0	0	3.7	0.387	1.56	588	∞	∞	16,6M	31.2M

Table 8: Poles and zeroes locations of 2N3904 amplifier (Hz)

Again, we choose a midband frequency 20kHz, and vary the source amplitude V_S to find the voltage transfer curve. The voltage transfer curve is as shown in figure 20. Notice that the maximum source amplitude is 50mV before the output amplitude starts getting chopped (saturated).

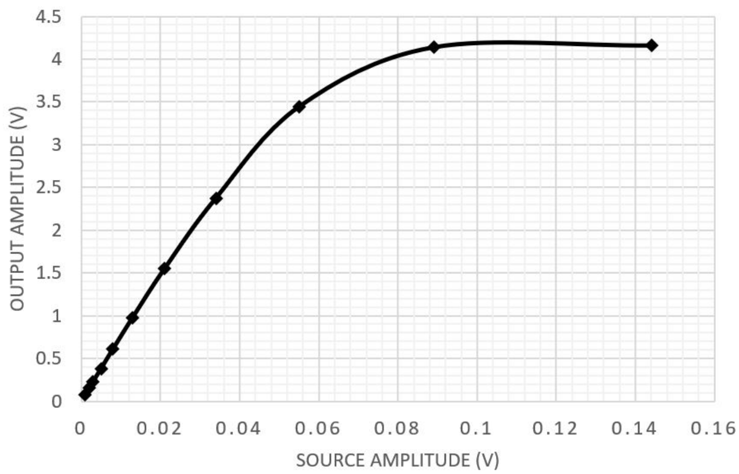


Figure 20: Voltage transfer curve for the 2N3904 amplifier

2.2.5 D - 2N4401 Amplifier

Same circuit is prepared for the 2N4401 transistor. Recall at DC operating point, V_{BE}=0.657V, V_{CB}=5.13V, g_m=0.038, β=147, r_π=3.87kΩ. Using the datasheet (figure 21) and DC operating point voltages, we find the internal capacitances: C_π=20pF and C_μ=4.5pF, same as the 2N3904 transistor.

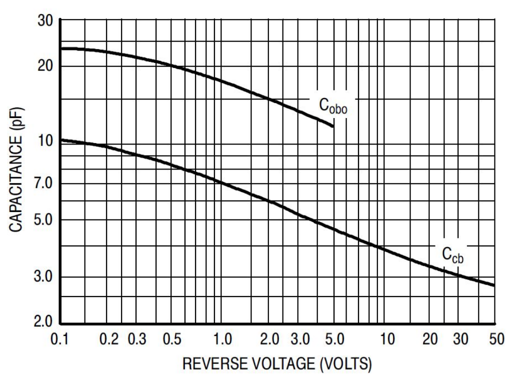
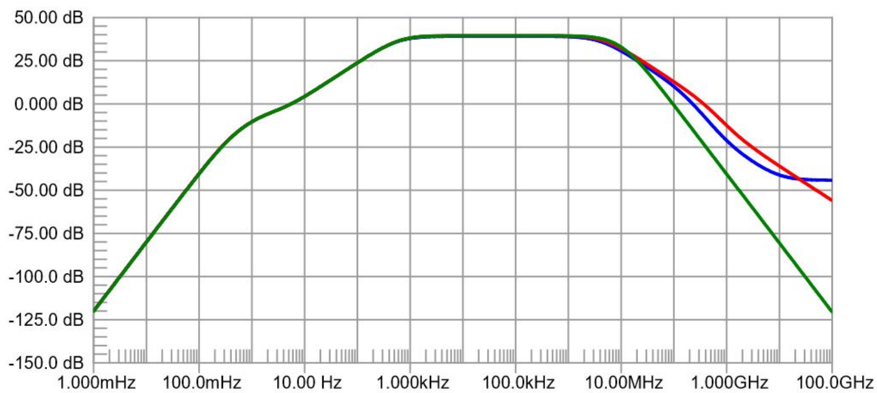
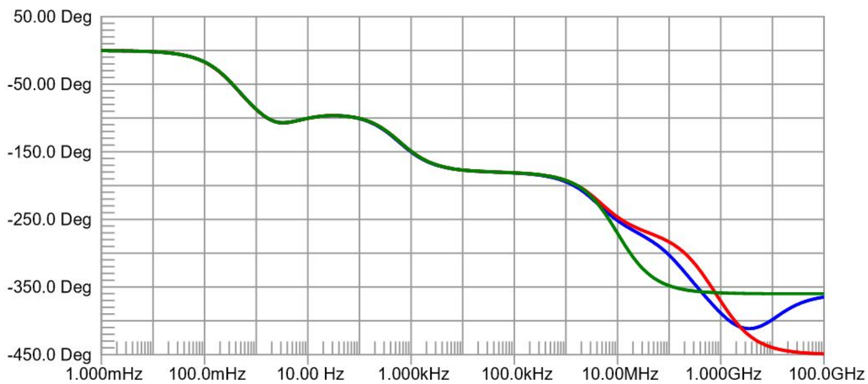


Figure 21: Internal capacitances for the 2N4401 transistor [10]

The bode plot of the amplifier is as follows in figure 22. The comparison of the calculated poles and graphed poles are in table 9. The details of the calculation and identifying poles and zeroes are identical to the one done for 2N2222A.



(a) Magnitude plot



(b) Phase plot

Figure 22: Bode plots for the 2N4401 amplifier

.

	$\omega_{L_{Z1}}$	$\omega_{L_{Z2}}$	$\omega_{L_{Z3}}$	$\omega_{L_{P1}}$	$\omega_{L_{P2}}$	$\omega_{L_{P3}}$	$\omega_{H_{Z1}}$	$\omega_{H_{Z2}}$	$\omega_{H_{P1}}$	$\omega_{H_{P2}}$
Graphed	0	0	6.11	0.41	1.33	647	1.6G	7.3G	6.22M	193M
Calculated	0	0	3.7	0.381	1.56	605	∞	∞	6.98M	13.9M

Table 9: Poles and zeroes locations of 2N4401 amplifier (Hz)

We can see that the high-frequency poles are inaccurate because of the approximation models we have made along the way. Also due to models and approximations such as small-signal and Miller transformation, the high-frequency zeroes are no longer visible. The low-frequency zero is slightly off because it is very close to other poles.

At the midband frequency of 20kHz, we vary the source amplitude. Once again, we found that at about 50mV is where the output signal starts to saturate, shown in figure 23

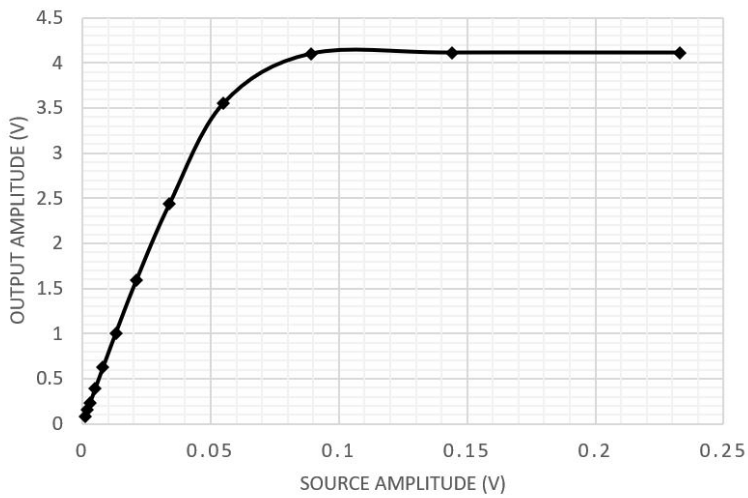


Figure 23: Voltage transfer curve for the 2N4401 amplifier

2.3 Part 3

In this section, we will use the 2N2222A transistor to build a common-base amplifier. First, we take the bias circuit as seen in figure 6c and add components until the circuit resembles a typical common-base amplifier. Figure 24 depicts the circuit. The small signal model is drawn as shown in figure 25 using transistor parameters of the 2N2222A calculated in part 1.

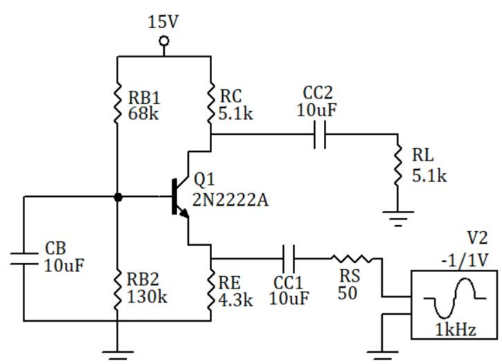


Figure 24: Common-base amplifier circuit for the 2N2222A transistor

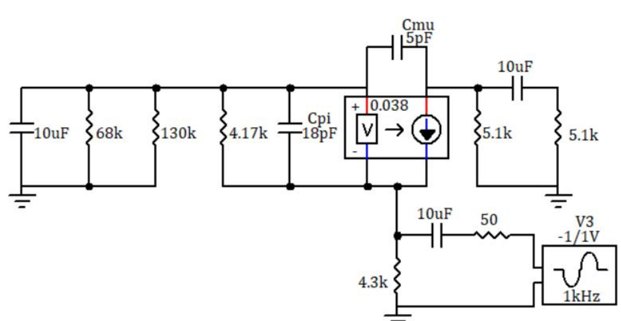
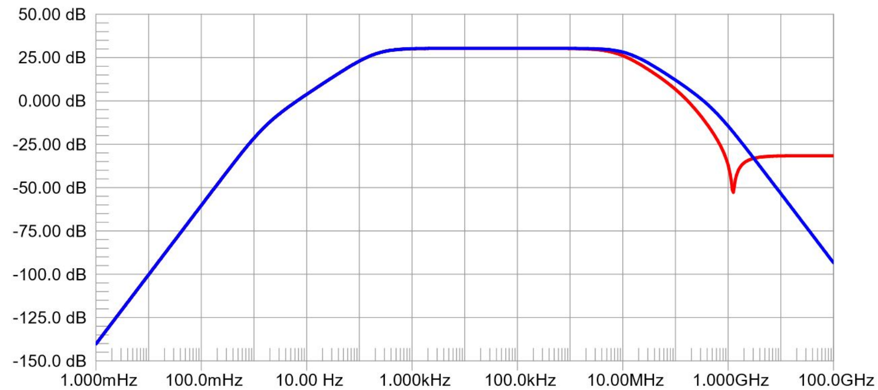


Figure 25: Small signal model of the common-base amplifier circuit

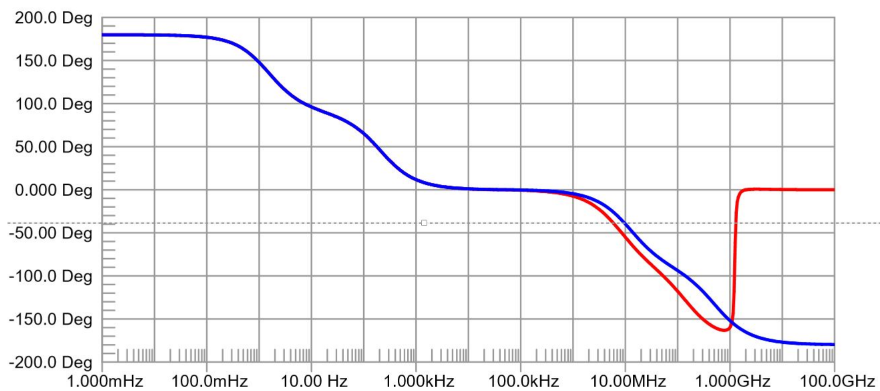
2.3.1 A - Bode Plots, Poles, and Zeroes

Bode Plots

The following is a bode plot of the amplifier circuit (figure 26).



(a) Magnitude plot



(b) Phase plot

Figure 26: Bode plots for the 2N2222A common-base amplifier

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The red plot is the original circuit, the blue plot is the small-signal model.

The sudden shift in phase from the phase bode plot and the cusp in the magnitude bode plot suggests that there a pair of complex zeros somewhere in the circuit. However, investigation into this subject is beyond the scope of this report.

Calculated Poles and Zeroes

First, we consider the low-frequency frequency response of the amplifier. The high-frequency capacitors act like open-circuits.

We first realize that there are two zeroes at zero, and one zero at the common base.

$$\begin{aligned}
\omega_{LZ1} &= 0 \\
\omega_{LZ2} &= 0 \\
\omega_{LZ3} &= \frac{1}{(68\text{k} \parallel 130\text{k})10\mu\text{F}} = 0.357\text{Hz}
\end{aligned}$$

Next, we use OCTC/SCTC tests to determine the low-frequency poles associated with the $\mathrm{C_B}$ and $\mathrm{C_{C1}}$. By inspection, we see that the OCTC for $\mathrm{C_B}$ is larger. So our strategy is to do OCTC test on $\mathrm{C_B}$ and SCTC test on $\mathrm{C_{C1}}$.

$$\begin{aligned}
\tau_{OC}^{C_B} &= 10\mu\text{F}(68\text{k} \parallel 130\text{k} \parallel (4.17\text{k} + (1+\beta)4.3\text{k})) = 0.421\text{s} \\
\omega_{LP1} &= \boxed{0.378\text{Hz}} \\
\tau_{SC}^{C_{C1}} &= 10\mu\text{F}\left(4.17\text{k}\left(\frac{1}{1+\beta}\right) \parallel 4.3\text{k} + 50\right) = 0.747\text{ms} \\
\omega_{LP3} &= \boxed{213\text{Hz}}
\end{aligned}$$

The last low-frequency pole is straight forward.

$$\omega_{LP2} = \frac{1}{10\mu\text{F}(10.2\text{k})} = \boxed{1.56\text{Hz}}$$

Finally, for the high-frequency poles, we short all the low frequency capacitors. Then the pole locations are computed as follows.

$$\begin{aligned}
\omega_{HP1} &= \left(18\text{pF}\left(4.17\text{k}\left(\frac{1}{1+\beta}\right) \parallel (4.3\text{k} \parallel 50)\right)\right)^{-1} \\
&= \boxed{535\text{MHz}} \\
\omega_{HP2} &= (5\text{pF} \times 2.55\text{k})^{-1} \\
&= \boxed{12.5\text{MHz}}
\end{aligned}$$

Comparison of Poles and Zeroes Locations

The poles and zero locations can be graphically determined using the same method done countless times before. Table 10 provides comparison between the calculated and plotted poles and zeroes. The values are not that accurate since it's difficult to find some low-frequency poles and zeroes as they are too close together. Also note that we think there is a complex pole at around 1.25GHz.

	ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
Graphed	0	0	1.2	0.86	3.76	237	1.25G*	1.25G*	606M	11.6M
Calculated	0	0	0.36	0.378	1.56	213	∞	∞	535M	12.5M

Table 10: Poles and zeroes locations of 2N2222A common base amplifier (Hz)

2.3.2 B - Varying Source Amplitude

For the remainder of the investigation, we use the midband frequency of 10kHz. We vary the amplitude of the source voltage similar to parts before. The result of the voltage transfer curve is as follows in figure **??**.

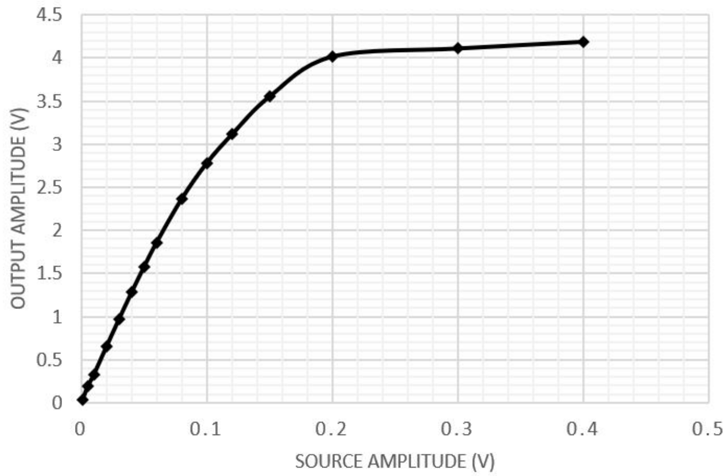

Figure 27: Voltage transfer curve

As we can observe, the common-base amplifier allows a much higher input amplitude as it can be up to 120mV before output starts distorting due to saturation.

2.3.3 C - Input Impedance

Calculated Input Impedance

The input impedance is the impedance seen by the source without the 50Ω source resistance. Because C_B, the low-frequency capacitor shorts out R_{B1} and R_{B2}, we are left with r_π. Thus the input impedance is:

$$R_{in} = R_E \parallel \frac{1}{1+\beta} r_\pi$$
$$R_{in} = \boxed{24.7\Omega}$$

Measured Input Impedance

As shown in figure 28, this is the circuit set up used to measure input impedance of the 2N2222A common base amplifier. We use 50mV for the source amplitude to ensure that there are no distortions in the output signal. Then we measure the current from the source at node A, and the voltage at node B. Then we can compute $R_{IN}=\frac{V}{I}=29.3\Omega$. Which is close enough.

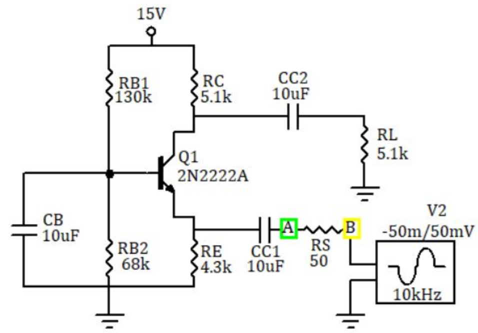

Figure 28: Circuit for measuring input impedance

3 Conclusion

In this investigation, we learned to apply the algorithm of modelling, analyzing, biassing, testing, and validating a single transistor amplifier.

First, we verified our understanding of the characteristics of transistors given a voltage or current. We compared our calculated values to the ones provided by the datasheet.

Next, we experimented with ways to bias a BJT for a DC operating point. We learned that 1/3 rule is very quick and versatile, and thus is a valid method of biasing a decent amplifier.

It follows that we applied the theory taught in lecture on poles and zeroes locations of the transfer function of common emitter and common base amplifiers to analyize the Bode plot. We learned that not all times our theoretical, approximate model will give an accurate insight, but often times it is close enough.

Lastly, we saw the effect of large signals on the BJT's small-signal model as the output is saturated or cutoff.

Finally, this mini-project is taking way too long and I think I need to end it here. It is 4:08 AM.

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