ELEC 301: Electronic Circuits

Mini Project 2: Biasing and the Common Emitter Amplifier

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1. Introduction

This project report outlines the modeling of bipolar junction transistors modeled for small signal operation using a hybrid-π model. Two basic single transistor amplifier circuits will be examined, including the common emitter and common base amplifiers. Under the assumption of small signals, transistors are assumed to work in the active region, as large signals may cause the transistors to enter saturation or cut-off modes, in which the hybrid-π model no longer describes the circuit's operation. The transistors used in this project are 2N2222A, 2N3904, and 2N4401. The datasheet for these transistors are obtained from manufacturer's websites.

2. Objective

The objective of this mini project is to strengthen our understanding of transistor's hybrid-π model and issues surrounding the biassing of transistors. Three commonly available transistors will be analyzed and measured in the context of the characteristics of an important transistor amplifier.

3. Project

3.1 Part 1: Biasing Transistors

A. Finding the Values of the Small Signal Parameters

The h-parameters of the 2N2222A transistor are first found on its datasheet, retrieved from STMicroelectronics' website. The values for h_{fe}, h_{ie} and h_{oe} at V_{CE}= 10 V, I_C= 1 mA, f = 1 kHz, and T = 25℃ are shown in Table 1.1.

Parameter	Name	Min Value	Max Value	Mean
h_{fe}	Small Signal Current Gain	50	300	175
h_{ie}	Input Impedance	2 kΩ	8 kΩ	5 kΩ
h_{oe}	Output Impedance	5 µS	35 µS	20 µS

Table 1.1: Small Signal Parameters from 2N2222A Datasheet

As the 2N2222A Datasheet contained both minimum and maximum values, the mean of the values will be used in future calculations.

B. Plotting Parameters and Estimating Early Voltage

Using an LTSpice DC sweep simulation on V_{BE}, we plot the relationship between I_B vs V_{BE} using the circuit shown in Figure 1.1, and probing the base of the 2N2222A transistor to obtain a current reading. The relationship between I_B and V_{BE} is shown as a plot in Figure 1.2.

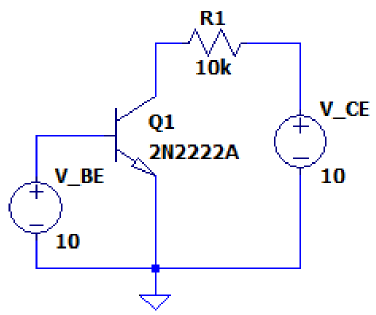


Figure 1.1: I_B vs V_{BE} Circuit

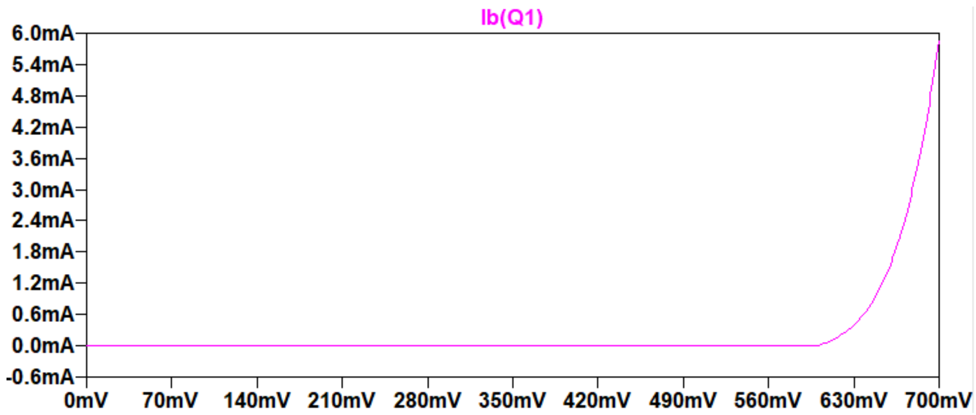


Figure 1.2: I_B vs V_{BE} Plot

Next we look at the circuit shown in Figure 1.3 to observe the relationship between I_C and V_{CE} with I_B being a variable parameter. a DC sweep on the current source I_B and the voltage source V_{CE}. The resulting relationship plot is shown in Figure 1.4. At the lowest curve, $I_B = 1\ \mu A$, and at the highest curve, $I_B = 10\ \mu A$.

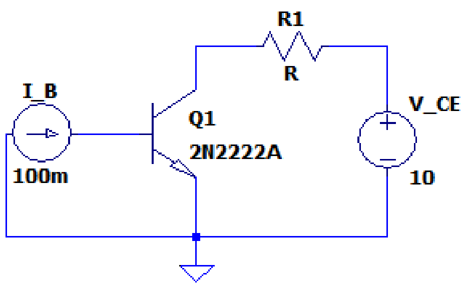


Figure 1.3: I_C vs V_{CE} Circuit

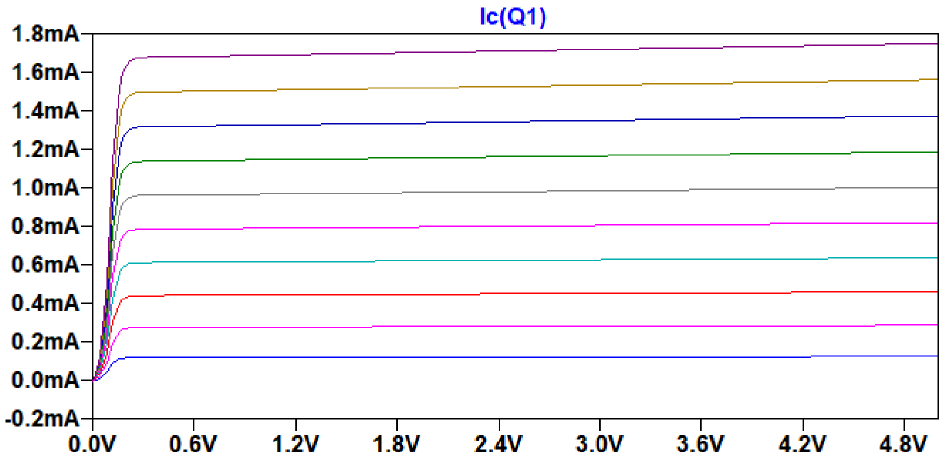


Figure 1.4: I_C vs V_{CE} Plot With Varying I_B Parameter

Finally, we once again look at the circuit shown in Figure 1.1, but this time to observe the relationship between I_C and V_{CE} with V_{BE} acting as a variable parameter. A DC sweep on the voltage source V_{BE} and the voltage source V_{CE} is simulated, and we probe the resistor at I_C to plot. The resulting relationship plot is shown in Figure 1.5.

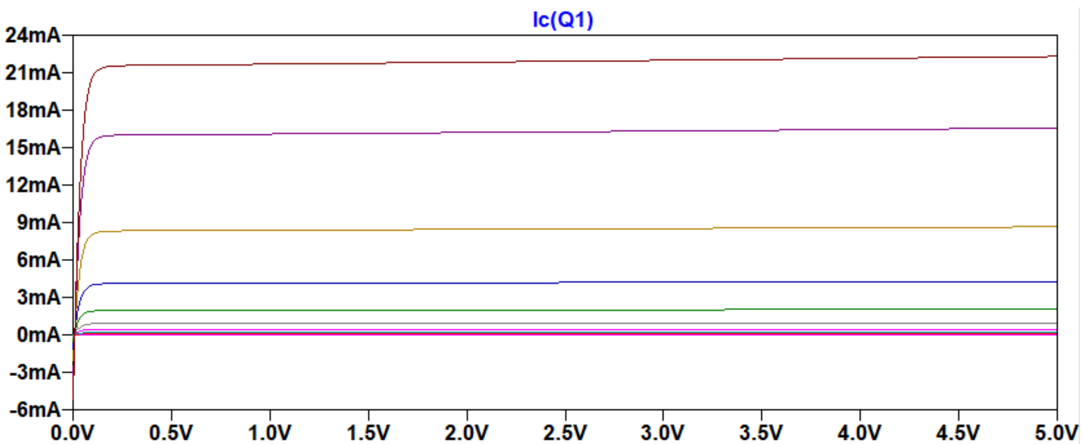


Figure 1.5: I_C vs V_{CE} Plot With Varying V_{BE} Parameter

As varying I_B provides a more straight-forward linear relation, we use Figure 1.4 to find I_B graphically at $V_{CE} = 5\ V$ and $I_C = 1\ mA$. As a DC Sweep was simulated from $1\ \mu A\ to\ 10\ \mu A$ for I_B in $1\ \mu A$ increments, we see that the I_B value corresponding to V_{CE} and I_C is the 6th curve from the bottom, therefore $I_B = 6\ \mu A$.

The forward current gain β, being the ratio of I_C to I_B, can be solved using $I_C = \beta I_B$.

Therefore, $\beta = 166.667$.

Next, we solve for the transconductance gain g_m, where $V_T = 25\ mV$ at T = 25℃:

$g_m = \frac{I_C}{V_T} = \frac{1}{25} = 0.04\ S$. Then, $r_\pi = \frac{\beta}{g_m} = \frac{166.667}{0.04} = 4.166\ k\Omega$

Next, we estimate the early voltage V_A by finding the x-intercept of the extended I_B slopes as shown in Figure 1.6. Using the slope of the active region curve in the $I_C\ vs\ V_{CE}$ plot in Figure 1.4, the slope equation y = mx + b can be determined and solved at $I_C = 0$ to find the x-intercept $- V_A$.

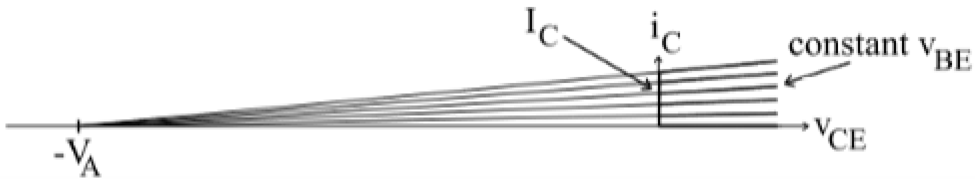


Figure 1.6: Finding V_A

We take the slope of I_B at $10\ \mu A$ at the arbitrary points of V_{CE} at 0.6 V and 6V:
$m = \frac{1.76m - 1.68m}{6 - 0.6} = 1.4815 * 10^{-5}\ S$. By extending I_B's slope to the y axis, we find the y-intercept to be approximately b = 1.669 mA. Setting $I_C = 0$, the slope equation is $0 = 1.4815 * 10^{-5}x + 1.669 * 10^{-3}$. Solving for the x-intercept, $- V_A = - 112.657\ V$.
r_o is given as the inverse of the change in i_c as a function of v_{CE}, and can be approximated in terms of V_A as $r_o = \frac{V_A}{I_C} = \frac{112.657\ V}{1\ mA} = 112.657\ k\Omega$.
Our measured values are compared to those of the 2N2222A datasheet, calculated with the same formulas as above, and are shown in Table 1.2. The small signal current gain β is shown to be in the range of [50, 300]. Using these datasheet values, we approximate the other parameters and find that β, r_π, and g_m were measured adequately. Looking at the I_C current plots in the datasheet further shows our r_o and V_A measurements to be good approximations.

	β	r_π	g_m
Measured	~167	$4.166\ k\Omega$	0.04 S
Datasheet	~175	~4.375 $k\Omega$	~0.04 S

Table 1.2: Measured Values vs Datasheet Calculated Values

C. Biasing the Transistor in its Active Region

In this section, we bias the transistor shown in Figure 1.7 in its active region using different methods, at $V_{CC} = 15\ V$ and $I_C = 1\ mA$ to find its DC operating point.

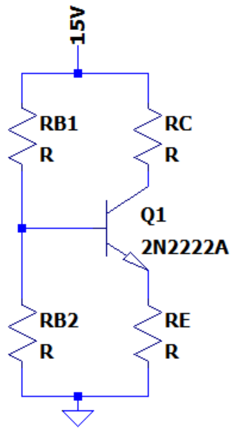


Figure 1.8: 2N2222A Biasing Transistor Circuit

i) DC Operating Point Using Measured Parameters

From Part B., $\beta = 167$. At DC operating point, $V_{BE} = 0.7\ V$ approximately. Using an arbitrary $V_{CE} = 4V$, and $R_E = \frac{R_C}{2}$, we can bias this circuit and measure the DC operating point. Then, $I_B = \frac{I_C}{\beta} = 6\ \mu A$, and $I_E = I_B + I_C = 1.006\ mA$. R_C can be found by KVL analysis on the collector and emitter: $V_{CC} = R_C I_C + V_{CE} + R_E I_E$.

Plugging in the values above and $R_E = \frac{R_C}{2}$, we can solve for $R_C = 7.318\ k\Omega$ and $R_E = 3.659\ k\Omega$.

Using the found currents and resistance, the transistor voltages can now be calculated:

- Emitter Voltage: $V_E = R_E I_E = 3.681\ V$
- Collector Voltage: $V_C = V_E + V_{CE} = 7.681\ V$
- Base Voltage: $V_B = V_E + V_{BE} = 4.381\ V$

The resistors R_{B1} and R_{B2}should be designed such that the base current I_B flowing in the circuit corresponds to that of the operating point I_B. To minimize power dissipation, we choose arbitrarily large and standard common values for BJT transistors:
$R_{B1} = R_{B2} = 1.2\ M\Omega$. Using these resistor values allow us to obtain an operating point near that of the current and voltage values calculated above. Using Thevenin, we know that

$R_{BB} = R_{B1}||R_{B2}$ and $V_{BB} = I_B[(1 + \beta)R_E + R_{BB}] + V_{BE} = \frac{R_{B2}}{R_{B1}+R_{B2}}V_{CC} = 7.5\ V$.

R_C	R_E	R_{B1}	R_{B1}	R_{BB}
$7.318\ k\Omega$	$3.659\ k\Omega$	$1.2\ M\Omega$	$1.2\ M\Omega$	$0.6\ M\Omega$

Table 1.3: Designed Transistor Resistance

Using previous formulas above and the resistances in Table 1.3, we calculate the new DC operating point to be as follows in Table 1.4.

I_C	I_B	I_E	V_C	V_B	V_E
0.9339 mA	$5.6\ \mu A$	0.9395 mA	7.417 V	4.117 V	3.417 V

Table 1.4 Operating Point Using Measured Values at $R_{B1} = R_{B2} = 1.2\ M\Omega$

ii) DC Operating Point Using ⅓ Rule

Using the First Version of the ⅓ Rule with $V_{CC} = 15\ V$, $I_C = 1\ mA$, $V_{BE} = 0.7\ V$ and $\beta = 167$, the transistor voltages are found as such:

- $V_B = \frac{1}{3}V_{CC} = 5\ V$
- $V_C = \frac{2}{3}V_{CC} = 10\ V$
- $V_E = V_B - V_{BE} = 4.3\ V$

Similarly, the transistor currents are found using the following current equations:

- $I_B = \frac{I_C}{\beta} = 6\ \mu A$
- $I_E = I_B + I_C = 1.006\ mA$

Finally, calculating the transistor resistance yields:

- $R_C = \frac{V_{CC} - V_C}{I_C} = 5\ k\Omega$
- $R_E = \frac{V_E}{I_E} = 4.274\ k\Omega$
- $R_{B1} = \frac{V_C}{I_1} = \frac{V_C}{\frac{I_E}{\sqrt{\beta}}} = 128.457\ k\Omega$
- $R_{B2} = \frac{V_B}{I_1 - I_B} = \frac{V_B}{\frac{I_E}{\sqrt{\beta}} - I_B} = 69.592\ k\Omega$

The final DC operating point using the resistor values used in ⅓ Rule is shown in Table 1.5

I_C	I_B	I_E	V_C	V_B	V_E
1 mA	$5.9\ \mu A$	$1.006\ mA$	9.99 V	5 V	4.5 V

Table 1.5 Operating Point Using ⅓ Rule

iii) DC Operating Point Using Closest Common Resistors

Choosing the closest commonly available resistors to those calculated in ii), we obtain Table 1.6. Calculating the new DC operating point using these resistor values yields Table 1.7.

R_C	R_E	R_{B1}	R_{B2}
$5.1\ k\Omega$	$4.3\ k\Omega$	$130\ k\Omega$	$68\ k\Omega$

Table 1.6: ⅓ Rule Using Standard Resistor Values

I_C	I_B	I_E	V_C	V_B	V_E
0.99 mA	$5.9\ \mu A$	$0.99\ mA$	9.9 V	4.89 V	5.49 V

Table 1.7 Operating Point Using ⅓ Rule and Standard Resistor Values

iv) Comparison

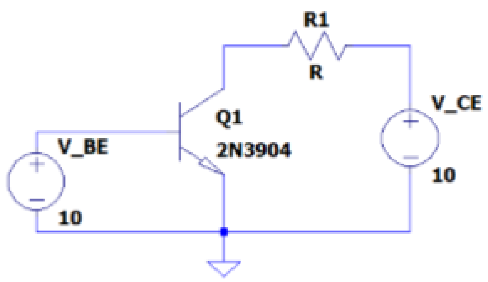
The DC operating point values obtained in Tables 1.4, 1.5, and 1.7 are all decently similar, with the largest difference being between the measured values in Table 1.4 and the values obtained using the ⅓ Rule in Table 1.5 and 1.7. Comparison Tables 1.5 and 1.7, the values are once again very similar, however replacing the resistance values by their standard values introduce some inaccuracies, resulting in a larger error margin.

D. DC Operating Points for 2N3904 and 2N4401

We now replace the 2N2222A transistor with a 2N3904 and 2N4401 transistor. We once again measure its parameters and bias the transistor in its active region to find the DC operating point, using the same methods as seen in previous sections.

2N3904

Similarly to section B, we perform a DC Sweep on V_{BE} and I_B using the circuits shown in Figure 1.9 a) and 1.9 b).



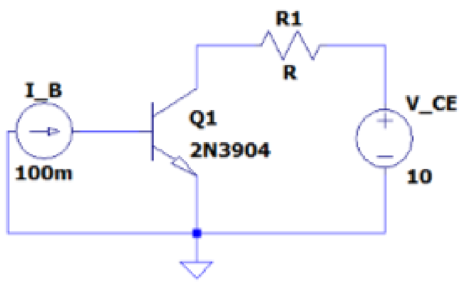


Figure 1.9 a) 2N3904 $I_B\ vs\ V_{BE}$ Circuit

Figure 1.9 b) 2N3904 $I_C\ vs\ V_{CE}$ Circuit

Plotting the relationship curves of $I_B\ vs\ V_{BE}$ and $I_C\ vs\ V_{CE}$, we generate a plot similar to Figure 1.4 to find I_B at $V_{CE} = 5\ V$ and $I_C = 1\ mA$. We see that $I_B = 8.5\ \mu A$. Then using $I_C = \beta I_B$, $\beta = 118$. The transconductance gain $g_m = \frac{I_C}{V_T} = \frac{1}{25} = 0.04\ S$. Then, $r_\pi = \frac{\beta}{g_m} = \frac{118}{0.04} = 2.950\ k\Omega$.

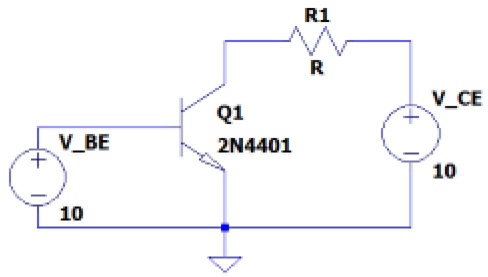
Using the resistor values in Table 1.4 and replacing the transistor in Figure 1.8 by a 2N3904 transistor, we bias our circuit with and without the ⅓ Rule to compare the resulting operating points as shown in Table 1.8.

	I_C	I_B	I_E	V_C	V_B	V_E
Measured	0.89 mA	$6.99\ \mu A$	$0.81\ mA$	9.7 V	3.32 V	2.7 V
⅓ Rule	0.97 mA	$8.01\ \mu A$	$0.96\ mA$	10.2 V	4.9 V	4.26 V

Table 1.8 2N3904 Operating Point Using Measured Values and ⅓ Rule

2N4401

We now use a 2N4401 transistor to simulate a DC sweep on V_{BE} and I_B using the circuits shown in Figure 1.10 a) and 1.9 b).



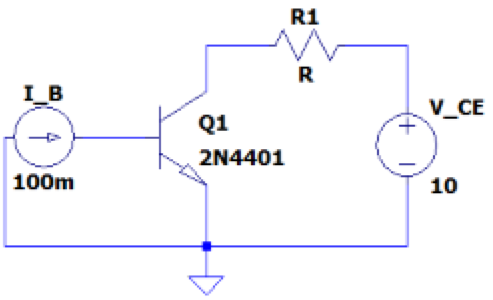


Figure 1.10 a) 2N4401 $I_B\ vs\ V_{BE}$ Circuit   Figure 1.10 b) 2N4401 $I_C\ vs\ V_{CE}$ Circuit

Plotting the relationship curves of $I_B\ vs\ V_{BE}$ and $I_C\ vs\ V_{CE}$, we generate a plot similar to Figure 1.4 to find I_B at $V_{CE} = 5\ V$ and $I_C = 1\ mA$. We see that $I_B = 6.8\ \mu A$, $\beta = 147$ and $g_m = 0.04\ S$. Then, $r_\pi = \frac{\beta}{g_m} = \frac{147}{0.04} = 3.675\ k\Omega$.

Using the resistor values in Table 1.4 and replacing the transistor in Figure 1.8 by a 2N4401 transistor, we bias our circuit with and without the ⅓ Rule to compare the resulting operating points as shown in Table 1.9.

	I_C	I_B	I_E	V_C	V_B	V_E
Measured	0.91 mA	$6.4\ \mu A$	$0.95\ mA$	8.9 V	3.7 V	3.07 V
⅓ Rule	0.97 mA	$6.66\ \mu A$	$0.98\ mA$	10.1 V	4.8 V	4.33 V

Table 1.9 2N4401 Operating Point Using Measured Values and ⅓ Rule

Observations

While the ⅓ Rule resulted in a very accurate DC operating point for the 2N2222A transistor, it did not perform as well for the 2N3904 and 2N4401 transistors, although the results were nonetheless acceptable, with the largest disparity being in the voltage values. It can be concluded that using the ⅓ Rule to bias a BJT transistor is an adequate method to use for transistors with larger β values.

3.2 Part 2: Common-Emitter Amplifier

A. Transistor Poles and Zeros

Using the biased transistor circuit shown in Figure 1.8, and given $C_{C1} = C_{C2} = C_E = 10\ \mu F$, $R_S = 50\ \Omega$, and $R_L = R_C$, the common-emitter amplifier in Figure 2.1 is created.

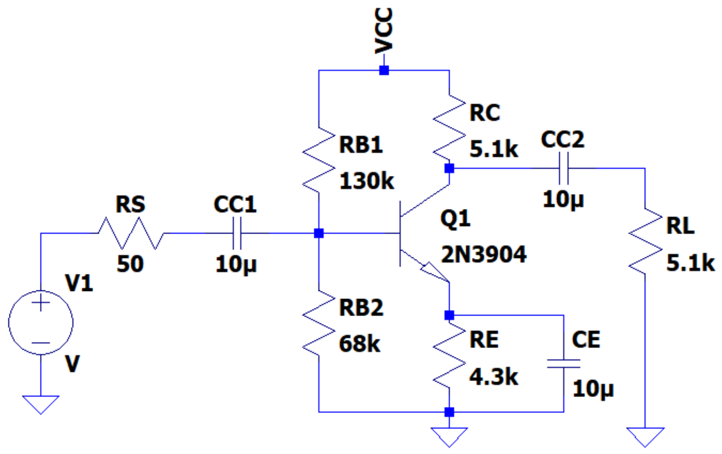


Figure 2.1: Common-Emitter Amplifier

Magnitude and phase Bode plots are plotted for this circuit as shown in Figure 2.2. The poles and low frequency zeros can be identified on the plot by linear approximation and fitting the slopes to the magnitude plot such as in Figure 2.3.

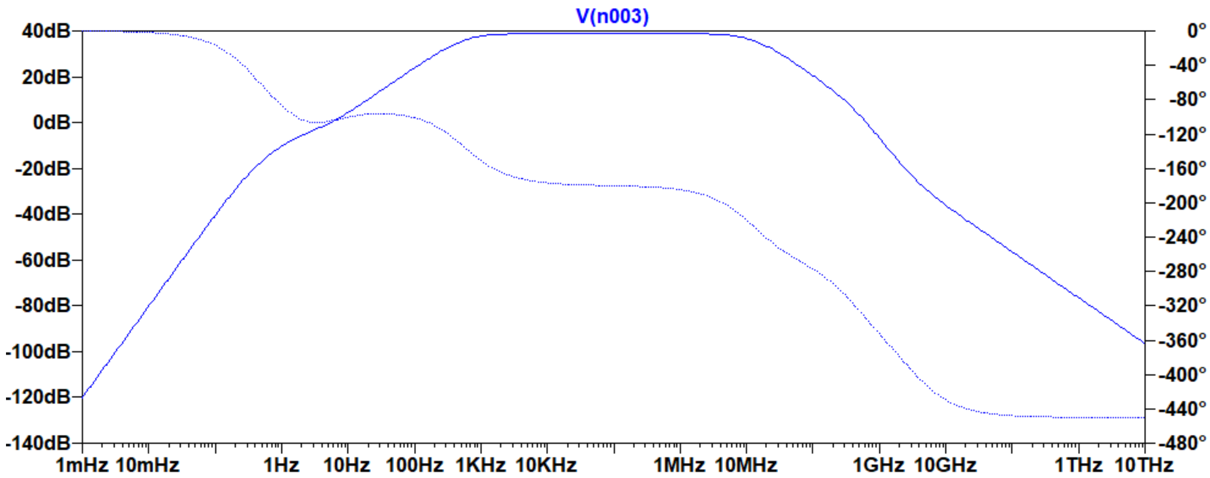


Figure 2.2: 2N3904 Magnitude and Phase Bode Plots

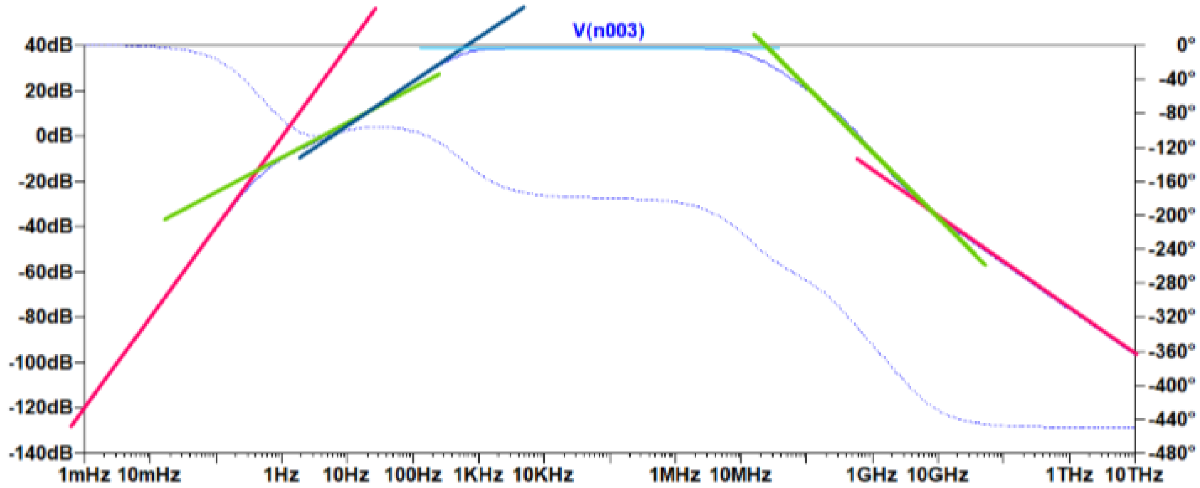


Figure 2.3: 2N3904 Magnitude Bode Plot Linear Approximation

The measured poles and zeroes locations are shown in Table 2.1.

ω_{LZ1}	ω_{LZ2}	ω_{LZ2}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
0	0	5.5	0.34	1.5	605	N/A	N/A	16M	68M

Table 2.1: 2N3904 Poles and Zeros From Measured Bode Plot

Next, we obtain the poles and zeroes locations by calculation. The high-frequency capacitors C_π and C_μ can be approximated using the relationship curves found in the 2N3904 transistor datasheet. From Part 1, it is known that $V_{BE} = 0.7\,V$ and $V_{CB} = 5\,V$. From the 2N3904 datasheet plot, $C_\pi = 3.5\,pF$ and $C_\mu = 2\,pF$. From Part 1, $g_m = 0.04\,S$ and $r_\pi = 2.950\,k\Omega$. To calculate the zeros and poles locations, a small signal model of the transistor is created as shown in Figure 2.4.

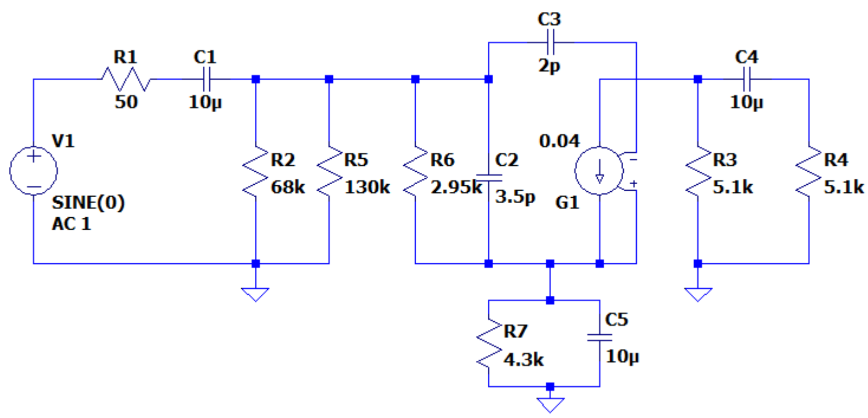


Figure 2.4: 2N3904 Small Signal Model

Applying Miller's theorem, $k = -g_m(5.1k \,||\, 5.1k) = -102$, $C_{\mu1} = C_\mu(1 - k) = 206\,pF$, $C_{\mu2} = C_\mu = 2p$, yielding the simplified circuit in Figure 2.5.

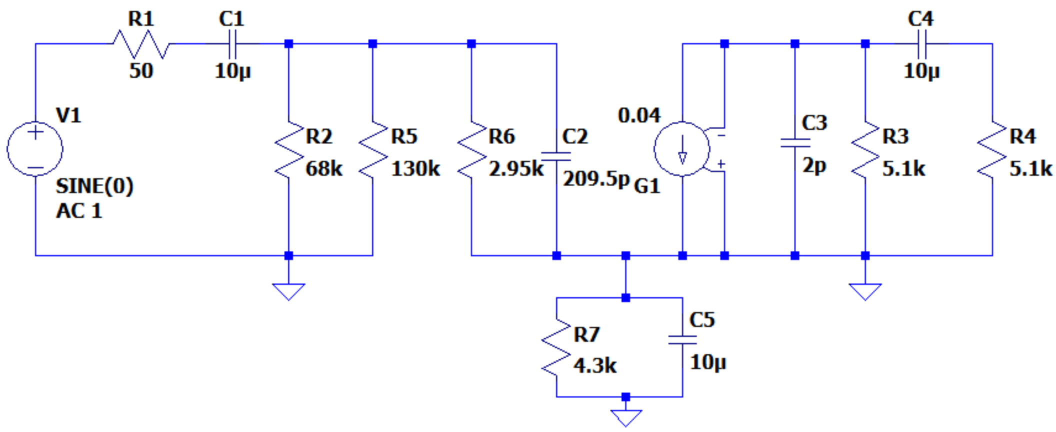


Figure 2.5: 2N3904 Miller's Theorem Applied

We use the short circuit time constant (SCTC) and open circuit time constants (OCTC) to find the poles and zeros:

At high frequency, low frequency 10 μF capacitors are shorted. The high frequency poles are calculated as follows:

$\omega_{HP1} = 209.5p(r_\pi \,||\, R_{BB} \,||\, R_S)^{-1} = 15.4\ MHz$

$\omega_{HP2} = (C_{\mu 2}(R_C \,||\, R_L))^{-1} = 31.2\ MHz$

At low frequency, we have two zeroes $\omega_{LZ1} = \omega_{LZ2} = 0\ Hz$ due to the coupling capacitance. the emitter capacitance introduces a third zero at $\omega_{LZ3} = \frac{1}{R_E C_E} = 3.7\ Hz$. The output pole due to the coupling capacitor is calculated as

$\omega_{LP2} = (C_{C2}(R_C + R_L))^{-1} = 1.5\ Hz$. Then using SCTC and OCTC,

$\omega_{LP3} = (C_{C2}(R_E \,||\, \frac{(R_{BB} \,||\, R_S) + r_\pi}{1+\beta}))^{-1} = 635\ Hz$, and

$\omega_{LP1} = (C_{C1}(R_S + (R_E * (1 + \beta) + r_\pi) \,||\, R_{BB})))^{-1} = 0.38\ Hz$

The calculated poles and zeroes locations are summarized in Table 2.2.

ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
0	0	3.7	0.38	1.5	635	N/A	N/A	15.4M	31.2M

Table 2.2: 2N3904 Calculated Poles and Zeros

Next, we repeat this process for a 2N4401 transistor, and obtain the poles and zeroes location by plotting and taking the linear approximation, then by calculations. The same methods are used as for 2N3904 above, these findings are shown in Table 2.3.

	ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
Meas	0	0	5.3	0.48	5.7	3.8k	N/A	N/A	106M	178M
Calc	0	0	23.3	2.39	9.8	4k	N/A	N/A	96.9M	196M

Table 2.3: 2N4401 Measured and Calculated Poles and Zeros

Comparing our measured poles and zeros frequencies to our calculated values, the values tend to be near that of the calculated values for low frequency poles. However the high frequency poles were slightly shifted due to the application of the Miller's theorem, resulting in a few inaccuracies. Furthermore, it becomes gradually difficult to discern pole values the steeper the slope and the closer to other poles they are, resulting in a larger margin of error.

B. Midband Frequency

From Figure 2.2, it is observed that the midband ranges from approximately 1 kHz to 5 MHz. We choose a low midband frequency of 20kHz and adjust the input voltage amplitude until the output domain becomes non-linear.

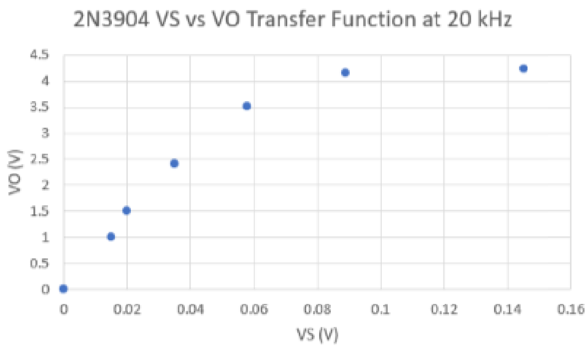


Figure 2.6 a) 2N3904 Transfer Curve 20kHz

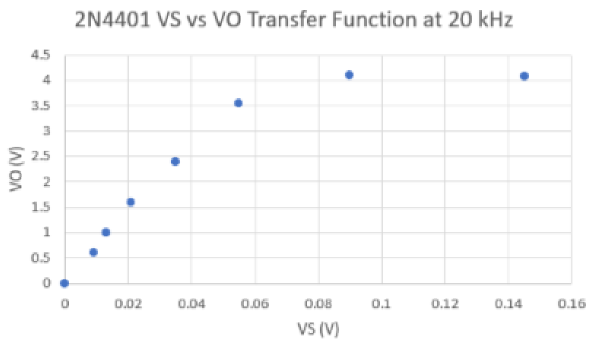


Figure 2.6 b) 2N4401 Transfer Curve 20kHz

C. Input Impedance

The input impedance of 2N3904 at a midband frequency of 20 kHz can be measured using $R_{in} = \frac{V_B}{I_{in}}$. By probing the input current in the transistor, we find that

$R_{in} = \frac{5.79\,V}{1.74\,mA} = 3.327\,k\Omega$. When calculating the input resistance at midband, the capacitors short, therefore $R_{in} = R_{B1} \,||\, R_{B2} \,||\, r_\pi = 2.867\,k\Omega$.

The input impedance of 2N4401 at a midband frequency of 20 kHz can be measured in the same way. When measured, $R_{in} = \frac{5.13\,V}{1.82\,mA} = 2.818\,k\Omega$, and when calculated, $R_{in} = R_{B1} \,||\, R_{B2} \,||\, r_\pi = 2.395\,k\Omega$. For both transistors, these values are approximately close, but the measured values tend to be larger than when calculated due to some inconsistencies.

D. Output Impedance

The output impedance of 2N3904 at a midband frequency of 20 kHz can be measured using $R_{out} = \frac{V_C}{I_{out}}$. By probing the output current in the transistor, we find that

$R_{out} = \frac{6.42\,V}{2.37\,mA} = 2.708\,k\Omega$. When calculating the input resistance at midband, the capacitors short, therefore $R_{in} = R_C \,||\, R_L = 2.55\,k\Omega$. Similarly, the measured output impedance of 2N4401 is $R_{out} = \frac{6.12\,V}{2.02\,mA} = 3.029\,k\Omega$ and calculated is $R_{in} = R_C \,||\, R_L = 2.55\,k\Omega$. For both transistors, these values are approximately close, but the measured values tend to be larger than when calculated due to some inconsistencies.

E. Overall Comparison

Both 2N3904 and 2N4401 perform well, however while the 2N3904 operates better at low currents, the 2N4401 can handle higher currents. For most common uses, the 2N3904 is a common choice as it is versatile and will provide a good performance in most cases.

3.3 Part 3: Common-Base Amplifier

A. Transistor Poles and Zeros

Given $C_{C1} = C_{C2} = C_E = 10\,\mu F$, $R_S = 50\,\Omega$, and $R_L = R_C$, the common-base amplifier in Figure 3.1 is created.

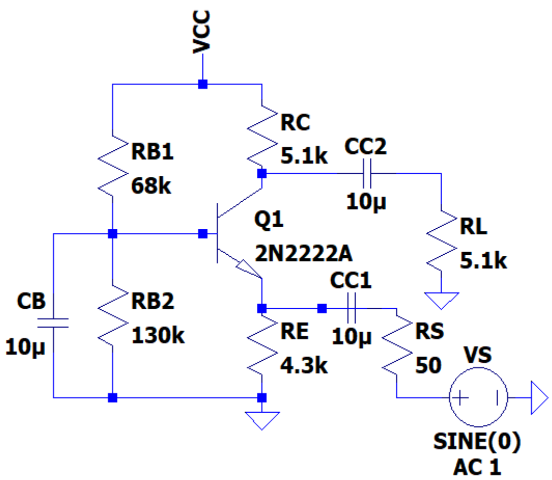


Figure 3.1: Common-Base Amplifier

Magnitude and phase Bode plots are plotted for this circuit as shown in Figure 3.2. The poles and low frequency zeros can be identified on the plot by linear approximation and fitting the slopes to the magnitude plot similarly to Figure 2.3 in Part 2.

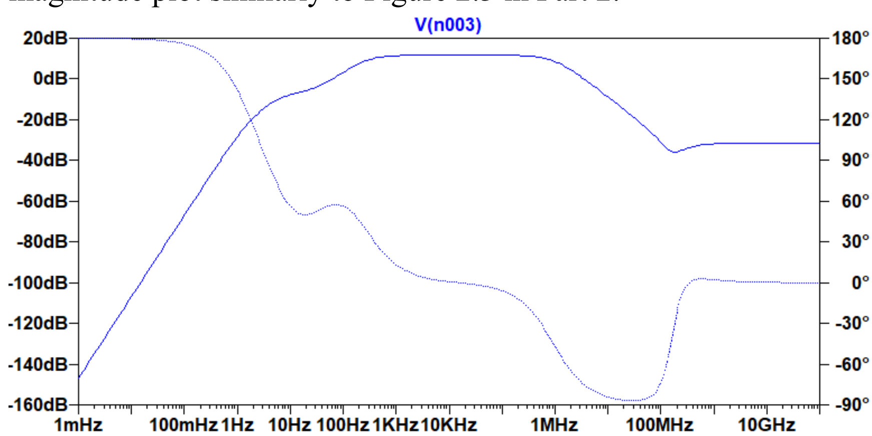


Figure 3.2: 2N2222A Magnitude and Phase Bode Plots

Next, we obtain the poles and zeroes locations by calculation using SCTC and OCTC, similarly to Part 2, using a small signal model shown in Figure 3.3.

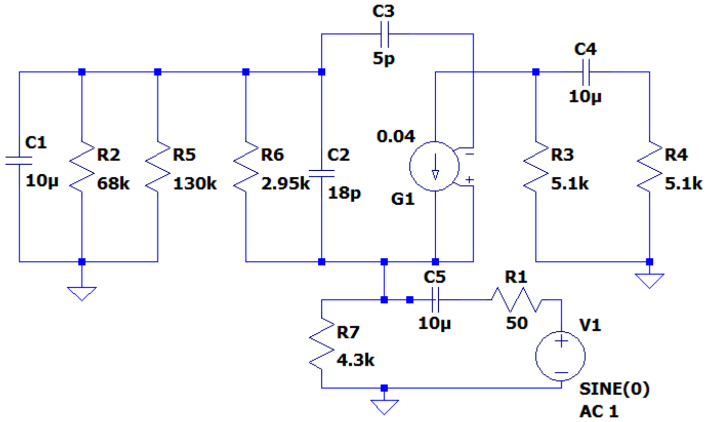


Figure 3.3 2N2222A Common-Base Amplifier Small Signal Model

From Part 1 and 2, it is known that $V_{BE} = 0.7\,V$, $V_{CB} = 5\,V$, $\beta = 167$, $g_m = 0.04\,S$ and $r_\pi = 2.950\,k\Omega$. From the 2N2222A datasheet curves, $C_\pi = 18\,pF$ and $C_\mu = 5\,pF$.
At low frequency, high-frequency capacitors are open. Then $\omega_{LZ1} = \omega_{LZ2} = 0$ and

$\omega_{LZ3} = \frac{1}{10\mu(68k\,||\,130k)} = 0.356\,Hz$. Next we use OCTC on C_B and SCTC on C_{C1} to find:

$\omega_{LP1} = (10\mu(68k\,||\,130k\,||\,(2.95k + 4.3k(1+\beta)))^{-1} = 0.378\,Hz$

$\omega_{LP3} = 10\mu((2.95k(\frac{1}{1+\beta})\,||\,4.3k) + 50) = 235.83\,Hz$

$\omega_{LP2} = (10\mu(5.1k + 5.1k))^{-1} = 1.56\,Hz$

Next, the low frequency capacitors are shorted to compute high-frequency poles:

$\omega_{HP1} = (18p(2.95k(\frac{1}{1+\beta})\,||\,(4.3k\,||\,50)))^{-1} = 682\,MHz$

$\omega_{HP2} = (5p(5.1k\,||\,5.1k))^{-1} = 12.5\,MHz$

The measured and calculated poles and zeroes locations are shown in Table 3.1. While the values are similar for the most part, zeros and poles that are near each other are difficult to see on a plot and may lead to some inaccuracies.

	ω_{LZ1}	ω_{LZ2}	ω_{LZ3}	ω_{LP1}	ω_{LP2}	ω_{LP3}	ω_{HZ1}	ω_{HZ2}	ω_{HP1}	ω_{HP2}
Meas	0	0	0.98	0.42	3.1	243	N/A	N/A	600M	15M
Calc	0	0	0.35	0.38	1.56	235	inf	inf	682M	12.5M

Table 3.1: 2N2222A Poles and Zeros From Measured Bode Plot

B. Midband Frequency

From Figure 3.2, it is observed that the midband ranges from approximately 1 kHz to 1 MHz. We choose a low midband frequency of 10kHz and adjust the input voltage amplitude until the output domain becomes non-linear as shown in Figure 3.4.

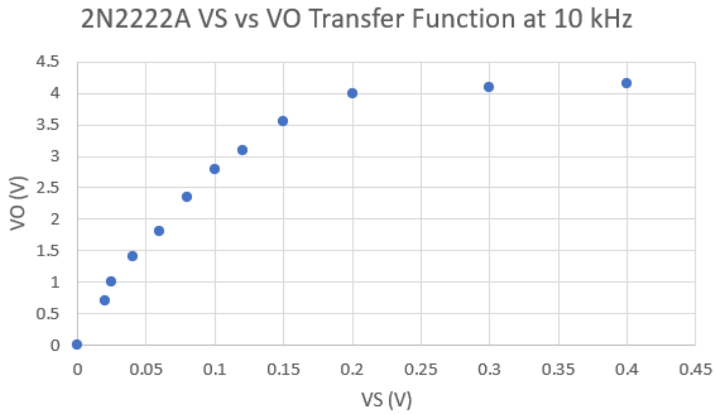


Figure 3.4 2N2222A Transfer Curve 10kHz

C. Input Impedance

The input impedance of 2N2222A at a midband frequency of 10 kHz can be measured by probing the voltage before R_s and probing the current after R_s: $R_{in} = \frac{V_{in}}{I_{in}} = 29\ \Omega$. When calculating the input resistance at midband without R_S, the capacitors short, therefore $R_{in} = R_E || \frac{r_\pi}{1+\beta} = 24\ \Omega$. These values are approximately close, but the measured values tend to be larger than when calculated due to some inconsistencies.

D. Output Impedance

The output impedance of 2N2222A at a midband frequency of 10 kHz can be measured by probing the output current in the transistor, we find that $R_{out} = \frac{5.13\ V}{1.77\ mA} = 2.7898\ k\Omega$. When calculating the input resistance at midband, the capacitors short, therefore $R_{in} = R_C || R_L = 2.55\ k\Omega$. These values are approximately close, but the measured values tend to be larger than when calculated due to some inconsistencies.

4. Discussion and Conclusion

This project consisted in modeling, biasing, and testing different types of transistor amplifiers, such as Common-base and Common-emitter amplifiers. First the characteristics of transistors at given voltages and currents were analyzed then and compared with our calculated values and ones from the transistor's datasheet. Next, the transistors were biased to their DC operating point using various methods, such as by measurement, using the ⅓ rule, then using the ⅓ rule with standard resistors. Finally we identified the poles and zeroes locations by analyzing the Bode plot and observed the difference between calculated and measured values. The effects of large signals were observed on small signal models, and we received insight on the performance of the approximate models for transistors.

6. References

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