Mini-Project 2

UBC Elec 301

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Part 1:

Part A – Data Sheet lookup

Gathering data about the operation parameters of the 2N2222A transistor from its datasheet we obtained the following table of values

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

Table 1: h parameters of 2n2222A transistor

Part B – Simulated data gathering

In order to find the transistors parameters graphically I simulated the following circuit:

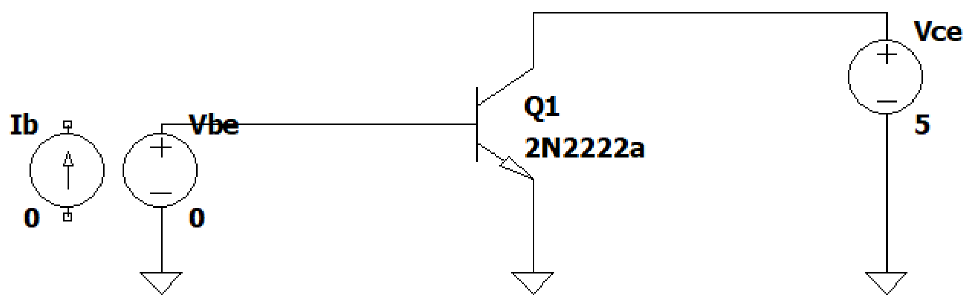


Figure 1-1: transistor circuit to find parameters

Throughout the data gathering process I swept through values of Vce, Vbe, and Ib. In the cases where we are looking at Ib I would replace the voltage source Vbe with the current source Ib.

We were also tasked to find our parameters at Ic = 1mA and Vce = 5V

First, I decided to use the plot of Ic vs Vce to determine the β of our transistor at operating point. We obtain this plot by probing the collector leg of transistor for current and setting our Vce and Ib sources to step through values from 0-6V and 100-700uA respectively.

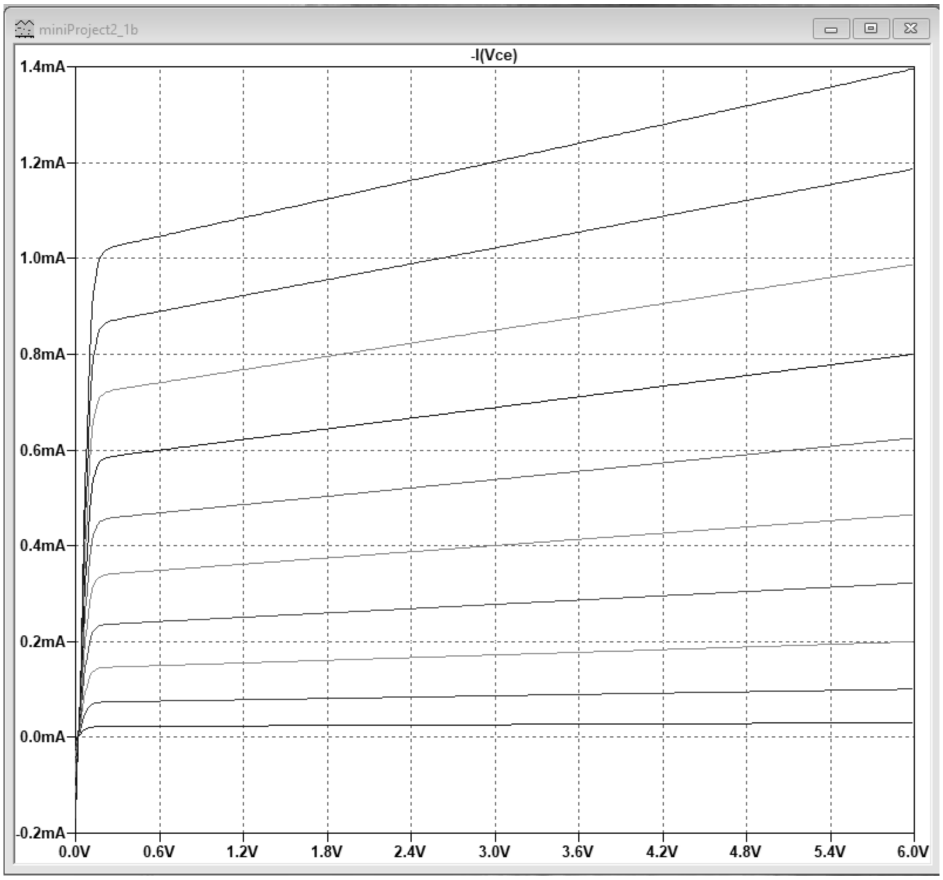


Figure 1-1 Plot of Ic vs Vce with variable Ib

In this graph we have Ic plotted on the y-axis and Vce plotted on the x-axis, each individual line represents a value of Ib from 100uA to approximately 700uA. I chose these values to show a legible representation of this graph. When finding my values, I sampled 100 plots at a time to close in on the value I wanted.

Using the plot of Ic vs Vce and sampling more data points for Ib we can find a plot corresponding to Vce = 5V and Ic = 1mA, the plot that best matches that corresponded to Ib = 6μA. Using the relationship Ic = βIb we can calculate β as 167

Second, we examine the relationship between Ib and Vbe by sweeping the voltage source Vbe from 0 to 1V and probing the base leg of our transistor to find current Ib. doing so gives us the following plot.

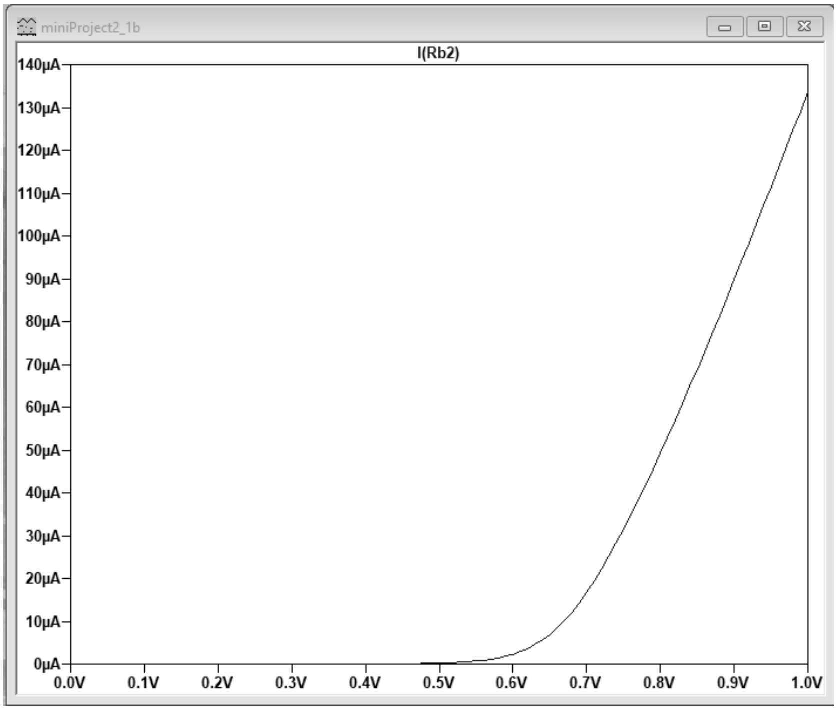


Figure 1-2 Plot of Ib vs Vbe

As shown in this plot Ib begins to increase shortly after Vbe reaches 0.5V, there is a noticeable increase after 0.6V, and Ib is even more noticeably increasing after 0.7V. Using this information, we can make assumptions about the Vbe at our operating point being somewhere in the 0.6-0.7V range.

Knowing β and that Ic = 1mA at our operating point we can calculate our expected Ib at operating point to be Ib = 6µA. By zooming in on this plot and analyzing the slope of the line around the point where Ib = 6µA we can obtain an estimate for $1/r_\pi$ giving us

$$r_\pi = 4.3k\Omega$$

We also see that Vbe is at 601mV for Ib = 6µA

$$r_\pi = \frac{v_{be}}{i_b} = \frac{\beta}{gm}$$

Using β and r_π we can calculate $gm = \frac{\beta}{r_\pi}$ = 0.0384

Finally, I used the plot of Ic vs Vce with variable Vbe to determine the remainder of our parameters at operating point, Va and r_o of our transistor at operating point. We obtain this plot by probing the collector leg of transistor for current and setting our Vce and Vbe sources to step through values from 0-6V and 100-700mV respectively.

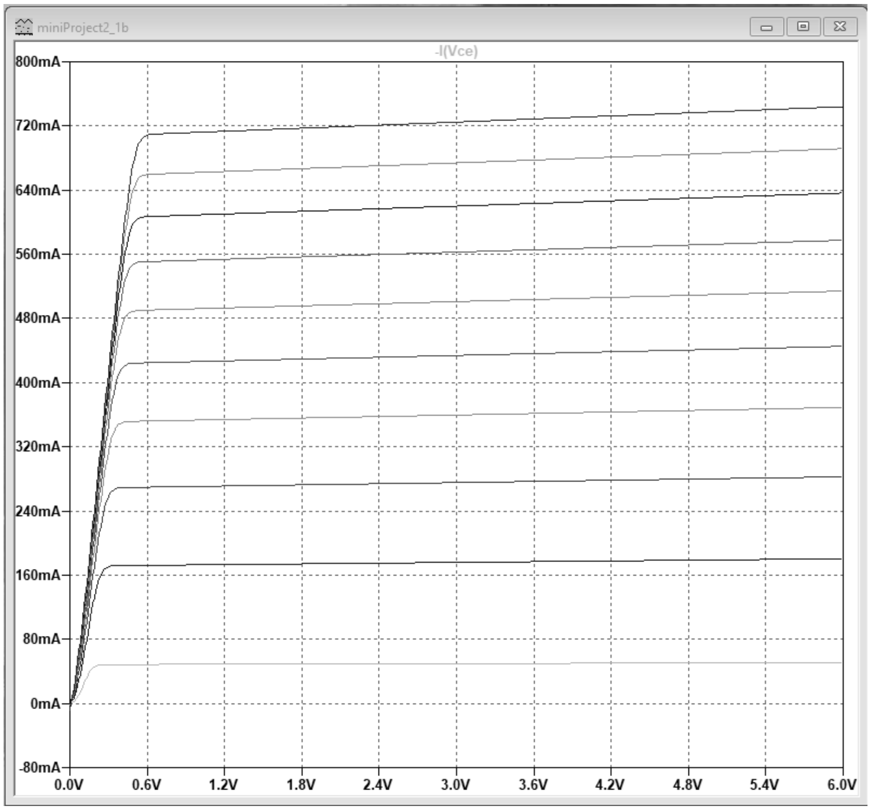


Figure 1-3 Plot of Ic vs Vce with variable Vbe

As with before I chose a graph that best indicated a good visual pattern of the data and experimentally, I took 100 or so plots at a time to narrow down my range to find the following values.

Using the plot of Ic vs Vce with variable Vbe to find the plot that passes through Vce = 5V and Ic = 1mA, we find the plot corresponding to Vbe = 601mV.

That slope of Ic/Vce corresponds to $1/r_o$ = 8.6μS as we learned in class

$$r_o = \left[\frac{\partial i_C}{\partial v_{CE}} \bigg|_{v_{BE} = constant} \right]^{-1}$$

Multiplying r_o = 115880 by Ic = 1mA we obtain Va = 115.8V

Or alternatively I estimated Va = 110.5V by extending my graph's x-axis into the negative and drawing lines over the slope of several of my lines and finding where the slopes would intersect the x-axis. Graphically finding this point I estimated it to be about 110.5 V thus confirming my other measurements.

Parameter	Found Value	Data Sheet Value
$1/r_o$	8.6 µS	5-35 µS
r_π	4.35 kΩ	2-8 kΩ
gm	0.0384	0.035
β	167	50 – 300
Va	115.88 V	200-28.5 V

Table 2: comparison of measured values to datasheet values

Comparing values, we can see that my measured values taken from the simulated transistor are all consistent with the values obtained from the transistor's datasheet.

Part C – Biasing

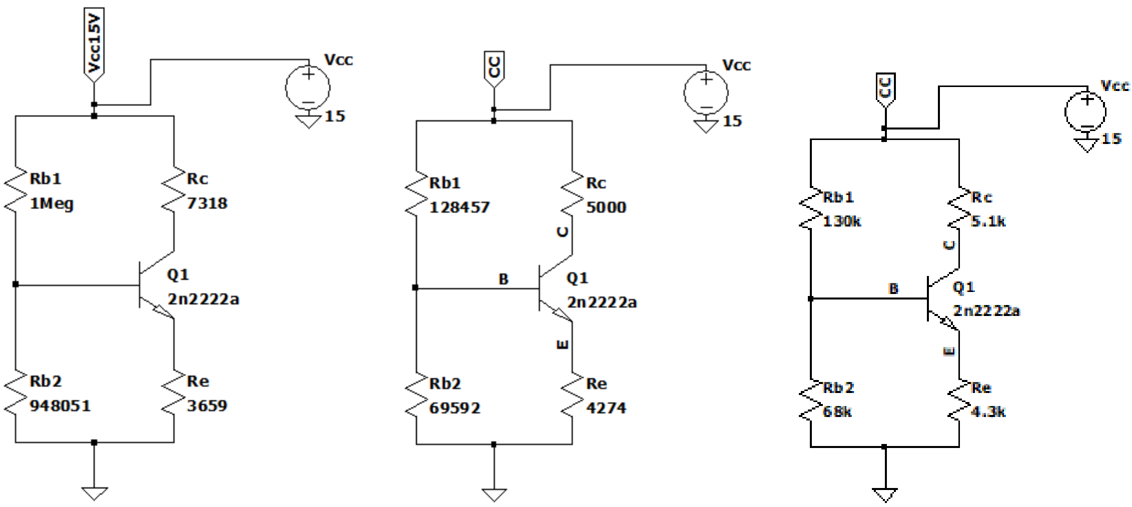


Figure 1-4: My bias circuit biased from simulated values (left), 1/3rd rule (middle), 1/3rd rule with standard resistors(right)

(i)

First, we take inventory of what we know:

I chose to make Vbe= 0.7V as it is an approximation that we use in class as well as a standard approximation made when biasing transistors. I don't believe that using 0.7 instead of 0.601 will make too much of a difference.

Vbe = 0.7V B=167, Ic = 1mA, Ib = 6µA, Ie = Ib+Ic = 1.006mA, Re= $\frac{Rc}{2}$

$$Vcc = IcRc + Vce + IeRe$$

$$15 = 1mA(Rc) + 4 + \frac{1.006mA(Rc)}{2}$$

$$Rc = 7.32k\Omega, \qquad Re = \frac{Rc}{2} = 3.66k\Omega$$

To find R1 and R2 we first transform our bias circuit using $R_{BB} = R_{B1}//R_{B2}$ and $V_{BB} = \frac{R_{B2}}{R_{B1}+R_{B2}} * V_{cc}$

Using nodal analysis, we get the equations: $V_{BB} = I_B R_{BB} + V_{BE} + I_E R_E$ and

$$\frac{V_{CC} - V_B}{R_{B1}} = \frac{V_B}{R_{B2}} + I_B$$

Solving the following system of equations (1) & (2), we find them to be unsolvable so instead, knowing that we need R_{B1} to be very large, so I set R_{B1} = 1MΩ

$$1)\ \frac{15 - 4.38}{R_{B1}} = \frac{4.38}{R_{B2}} + 6 * 10^{-6}$$

$$2)\ \frac{R_{B2}}{R_{B1} + R_{B2}}(15) = \left(\frac{R_{B1} * R_{B2}}{R_{B1} + R_{B2}}\right) 6 * 10^{-6} + 0.7 + 3.66 * 10^{3} + 1.006 * 10^{-3}$$

Calculating R_{B2} after setting R_{B1} to 1MΩ I calculated R_{B2} as 948kΩ

$$3)\ \frac{15 - 4.38}{1M} = \frac{4.38}{R_{B2}} + 6\mu$$

Simulating the circuit, I found the following to be the dc operating point:

V(C):	V(B):	V(E):	V(CC):	Ic(Q1):	Ib(Q1):	Ie(Q1):
7.594V	4.326V	3.7249V	15V	0.0010119A	6.10991µA	0.00101801A

(ii)

Given Ic = 1mA and Vcc = 15V the 1/3rd rule states that $V_B = \frac{Vcc}{3}$ = 5V and $V_C = Vcc * \frac{2}{3} = 10$V, using V_{BE} = 0.7V we get a V_E of 4.3V

We calculate $R_C = \frac{15-10}{1*10^{-3}} = 5k\Omega$ Using the relationship $I_C = \beta I_B$ and $I_E = I_C + I_B$ solving for I_B = 6µA and I_E = 1.006mA

We calculate $R_E = \frac{4.3}{1.006*10^{-3}} = 4274\Omega$

The 1/3rd rule also states that $I_{B1} = \frac{I_E}{\beta}$ = 77.847µA

Using I1 + Ib+I2 = 0 and $R = \frac{Vdrop}{I}$

We can solve for our remaining currents and resistors giving us R_{B1}=128.46kΩ and R_{B2}= 69.59kΩ for I_1 = 77.85µA and I_2=71.85µA

Giving us the dc operating point:

V(C):	V(B):	V(E):	V(CC):	Ic(Q1):	Ib(Q1):	Ie(Q1):
9.889 V	4.996 V	4.3945 V	15 V	0.001022 A	6.08633 µA	0.00102819 A

(iii)

Subbing in the common resistor values closest to our 1/3rd rule approximation we get a dc operating point of:

V(C):	V(B):	V(E):	V(CC):	Ic(Q1):	Ib(Q1):	Ie(Q1):
10.044 V	4.8877 V	4.28708 V	15 V	0.0009910 V	5.9067 µA	0.00099699 A

(iv)

Comparing the 3 different operating points we can see that I_C I_B I_E stay largely the same between all 3 models differing by much less than 10% despite the large difference in collector voltage between our first model and the 2 1/3rd rule models. This large difference in the collector voltage is most likely due to the value I arbitrarily assigned to R_{B1} assuming that it must be large. The 1/3rd rule approximated that resistor to be around 13% the size of my R_{B1}.

Part D – The other 2 transistors

By changing only, the transistor of my circuit biased using the 1/3rd rule with standard resistances and re-simulating I generated the following for the DC operating points of our other 2 transistors.

2n3904 dc operating point:

V(C):	V(B):	V(E):	V(CC):	Ic(Q1):	Ib(Q1):	Ie(Q1):
10.219 V	4.79045 V	4.14565 V	15 V	0.00095601 V	8.08712 µA	0.000964104 A

2N4401 dc operating point:

V(C):	V(B):	V(E):	V(CC):	Ic(Q1):	Ib(Q1):	Ie(Q1):
10.15 V	4.85615 V	4.19941 V	15 V	0.00096999 A	6.61557 µA	0.000976607 A

Comparison:

As seen biasing with the 1/3rd rule we were able to successfully bias 3 different transistors with the same circuit. Using the Data given by the dc operating point we can calculate parameters for each of our new transistors finding values for Vbe, Vcb, Vce, β, and r_π. We can see from the data that the 2n3904 has the largest Vcb at 5.428V, the smallest β of 118, the smallest r_π at 2.95kΩ and a Vbe of 0.65V.

These calculated values will be important later when comparing transistors and for making the decision of which transistor to use when building our amplifier.

Part 2:

Part A – Bode Plot Poles and Zeros

I chose a matched resistor for the load resistor to maximize power transfer by setting Rl = Rc.

In order to do a small signal model of our circuit we need values for the 2 internal junction capacitors Cπ and Cμ as well as a value of gm and r_π for all 3 of our transistors.

Rather than calculating gm for my other 2 transistors like I did in part 1 I opted to set gm = 0.04 for all 3 transistors, since gm = $\frac{Ic}{VT}$ where VT is the thermal equivalent of voltage = 0.025 V at room temperature and Ic = 1mA as we are biasing at Ic = 1mA. Measuring gm from part 1 I found it to be about 0.0384 therefore I believe my approximation to 0.04 is valid.

As for V_{CB} and β I will use values calculated from the dc operating point of each transistor I generated in the previous part. And then using the relationship of

$$r_\pi = \frac{v_{be}}{i_b} = \frac{\beta}{gm}$$

I will find my value of r_π.

Formulas were provided for Cμ and C_π:

$$C\pi = 2 * CJE + TF * gm$$

$$C\mu = \frac{CJC}{\left(1 + \frac{V_{CB}}{VJC}\right)^{MJC}}$$

Where CJC, MJC, TF, CJE are all values taken from our simulation model and V_{CB} is taken from my DC operating point of each transistor.

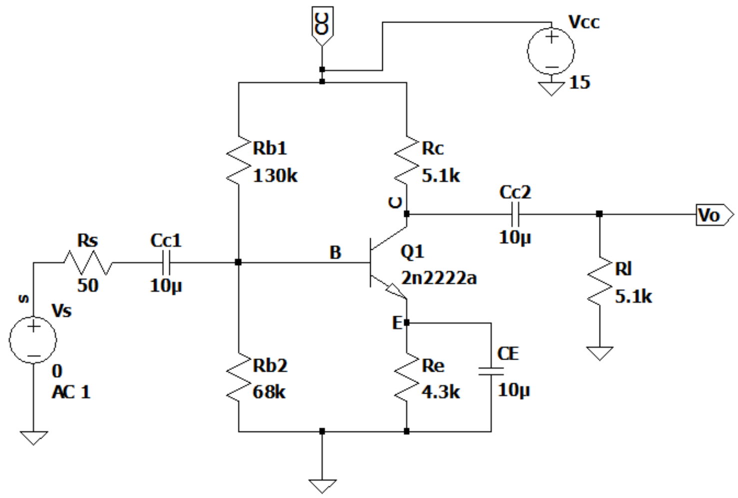


Figure 2-1 The Common emmited amplifier with matched load

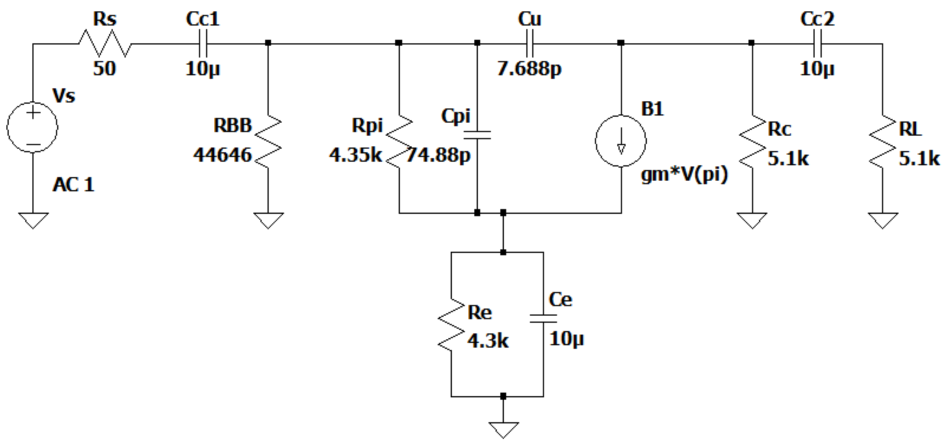


Figure 2-2 The small signal model

2N2222A transistor

For the 2n2222A transistor I found $r\pi$= 4.35k, Cµ = 7.688pF, Cπ = 74.88pF and gm = 0.04.

Calculating the location of the poles and zeros:

First, we split the small signal model of our circuit into the low frequency and high frequency response. For the sake of easier calculations, I decided to ignore the r_o resistance in the small signal model as we did in class. Since r_o is so much larger than the rest of our resistors its reasonable to say that the current passing through r_o is negligible and that it acts as an open circuit.

At high frequency all the 10µF capacitors are treated like short circuits and leaving the 2 internal junction capacitors untouched. Since Cµ is coupling the input and output we can apply miller's theorem to split the circuit and calculate our high frequency poles.

For the miller gain k, we calculate it as $-gm * (5.1k||5.1k) = -102$ allowing us to solve for our poles.

$$WHp1 = (866.7pF * (50||44.6k||4.35k))^{-1} = 23.4 * 10^6/sec = 3.7MHz$$

$$WHp2 = (7.76pF * (2.52k))^{-1} = 50.5 * 10^6/sec = 8.0MHz$$

Next, we calculate the low frequency poles and zeros from the low frequency response where all the high frequency capacitors are replaced by open circuits.

There are 2 zeros at 0 caused by the coupling capacitors. The 3rd zero is caused by the emitter capacitor and can be found by finding the frequency where the emitter admittance is 0. We find

$$Wlz = \frac{1}{ReCe} = (4.3k * 10\mu F)^{-1} = 23.2/sec = 3.6Hz$$

Since there is a dependent current source on the only branch connecting the output to the rest of our circuit we can treat the pole associated with the output coupling capacitor to be isolated and unaffected by the states of the other 2 low frequency capacitors, we can easily calculate:

$$Wlp2 = (10\mu F * 5.1k + 5.1k)^{-1} = 9.8/sec = 1.56Hz$$

Using the method of open circuit short circuit time constants, we can calculate the remaining 2 low frequency poles. However, one thing to be made aware of in the calculations would be the magnifying factor (1+β). The emitter impedances seen by the base are magnified by a factor of 1+ β and the base impedances as seen from the emitter are demagnified by a factor of 1+ β ie. Magnified by 1/1+ β

Calculating the time constants, we get:

$$t_{Cc1}^{OC} = 10\mu F * (50 + \left(44646\middle|\middle|\left(4.35k + 4.3k * (1 + 167)\right)\right) = 10\mu F * 42.09k = 0.4209/sec$$

$$Wlp1 = 2.37 = 378mHz$$

$$t_{CE}^{SC} = 10\mu F * (4.3k||\frac{4.35k + (50||44.6k)}{1 + 167}) = 10\mu F * 26.03 = 2.6 * 10^{-4}$$

$$Wlp3 = 3841.5 = 611Hz$$

Bode plot:

I then ran an AC sweep of my circuit (figure 2-1) measuring the output voltage at node Vo with the AC source set to a magnitude of 1V AC plotting a bode plot from 1mHz to 1THz in order to see the full plot beyond our mid-band.

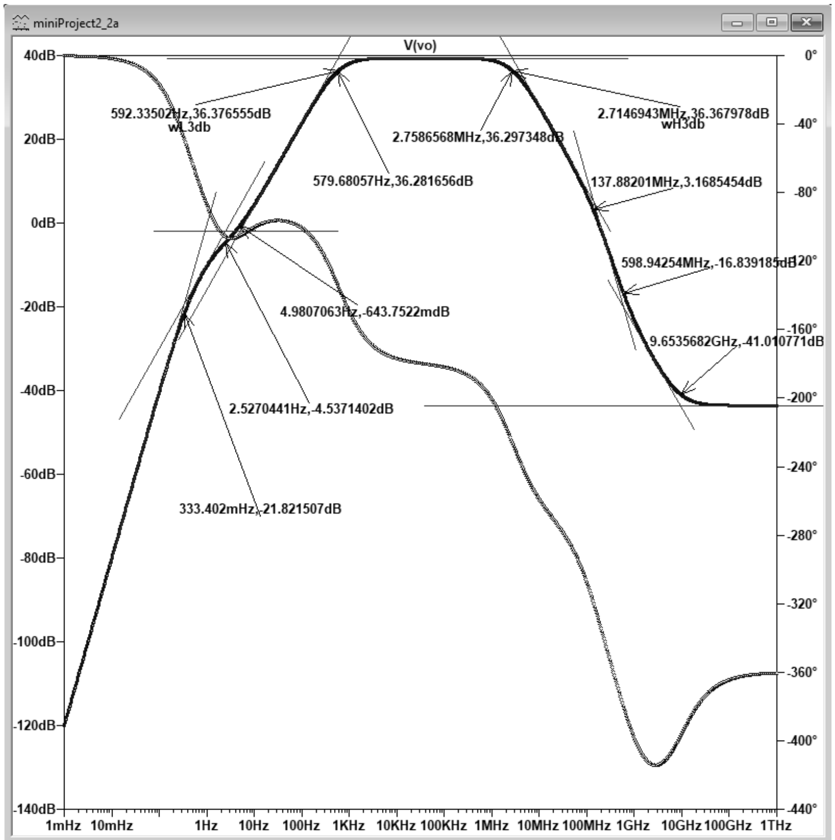


Figure 2-3 Bode plot of 2n2222a
Magnitude (Darker), phase (lighter)

As can be seen from my plot (figure 2-3) where I found the locations of all poles and zeros graphically, we obtain the following table of values

	WL_{Z1}	WL_{Z2}	WL_{Z3}	WL_{p1}	WL_{p2}	WL_{p3}	WH_{Z1}	WH_{Z2}	WH_{p1}	WH_{p2}
Graphical	0	0	4.98	333m	2.53	579.7	598.9M	9.65G	2.76M	137M
Calculated	0	0	3.7	378m	1.56	611	∞	∞	3.7M	8M

Table 3: 2n2222a poles and zero locations (Hz)

Comparison:

As shown our poles and zeros remain accurate to the actual poles remaining within a decade of difference with a few exceptions. Since we had to use miller's theorem to find the high poles, we weren't able to find the high zeros, however the high zeros happen after the mid band, so they are inconsequential for our purposes of operation. Its worth noting that the reason our WH_{p2} is so dramatically different from the graphical value is because that pole frequency is dependent on the resistance of r_o however, using the method we learned in class we decided to neglect r_o in our small signal model as it's inconsequential to our mid band and its large enough to treat it as an open circuit as it draws so little current when compared to the other resistors.

The other 2 transistors

For the remaining 2 transistors I used the same process to calculate the poles and zeros just using different values. The general equations that I used to calculate the poles and zeros can be found in the appendix along with the corresponding bode plots with marked pole and zero locations.

From my Dc operating point tables I find $V_{CB} = V_C - V_B$ and $\beta = \frac{I_C}{I_B}$

2N3904 transistor

For the 2n3904 transistor I set r_π= 2.95kΩ, Cμ =1pF, Cπ = 25pF and gm = 0.04.

For β and V_{CB} I got 118 and 5.428V respectively.

CJE, TF, CJC, VJC, MJC were 23.4p,512p,10.2p,750m, and 330m respectively which were plugged into the provided equations to give Cπ and Cμ.

	WL_{Z1}	WL_{Z2}	WL_{Z3}	WL_{p1}	WL_{p2}	WL_{p3}	WH_{Z1}	WH_{Z2}	WH_{p1}	WH_{p2}
Graphical	0	0	5.04	327m	2.415	598	2.89G	∞	11.1M	465.7M
Calculated	0	0	3.7	387m	1.56	635	∞	∞	15.82M	35.42M

Table 4: 2n3904 poles and zero locations (Hz)

Comparison:

Much like our original transistor we see the 2nd high pole is much higher than calculated and as explained previously this is caused by us neglecting r_o. Important things to note for this transistor would be the length of the mid band, as our bode plot is in mid band for more than twice the frequencies of our other transistors. The 2nd high frequency zero was also determined to be at such a high frequency that I decided it was equivalent to infinity.

2N4401 transistor

For the 2n4401 transistor I set r_π= 3675Ω, Cμ =5.12 pF, Cπ = 67.28pF and gm = 0.04.

For β and V_{CB} I got 118 and 5.428V respectively.

CJE, TF, CJC, VJC, MJC were 4.5p,400p,3.5p750m, and 330m respectively which were plugged into the provided equations to give Cπ and Cμ.

	WL_{Z1}	WL_{Z2}	WL_{Z3}	WL_{p1}	WL_{p2}	WL_{p3}	WH_{Z1}	WH_{Z2}	WH_{p1}	WH_{p2}
Graphical	0	0	7.25	311m	3.88	557	803M	11.4G	3.99M	155.7M
Calculated	0	0	3.7	381m	1.56	636	∞	∞	5.42M	12.07M

Table 5: 2n4401 poles and zero locations (Hz)

Comparison:

Much like our first transistor the poles mostly agree, we were unable to find the zero's at high frequency. It's worth noting for this transistor that the bandwidth of the mid band is about twice the length of our first transistor the 2N2222A.

Part B – Mid Band Frequency

For the 2N2222A transistor I found a mid band from approximately 600Hz to 2.7MHz keeping in line with that I decided to set my source to a 1Mhz frequency. Resulting in the following graph:

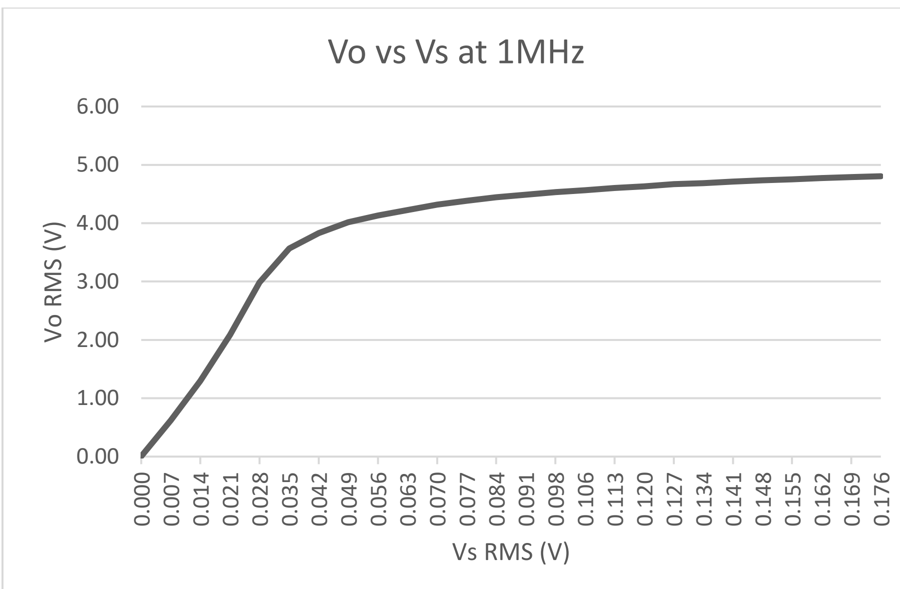


Figure 2-4: plot of Vo (RMS) vs Vs (RMS)

As shown by my graph the output voltage as a function of the input voltage we initially see the output voltages rising linearly with the input. However, at around 42mV RMS = 59.39mV PEAK we see that this linear behaviour ends, and the graph begins to flatten out. This behaviour is caused by the output signal being saturated and being unable to physically supply the expected linear voltage.

A plot of this behaviour in terms of pure input and output sinusoids can be found in the appendix.

Part C – Input Impedance

To find the input impedance of my amplifier I we look at the small signal AC model at mid band. At mid band frequency the emitter capacitor acts as a short and the Cμ and Cπ capacitors act as open circuits isolating the input and making the equivalent impedance = $r_\pi || R_{B1} || R_{B2}$ = 3.92kΩ

Simulating the circuit and performing an AC analysis after replacing our voltage source and source resistor with an AC current source set to 1A amplitude and then probing the voltage of our input node gives the equivalent resistance of the circuit at different frequencies. Since the input current is 1A we can say that the measured voltage is equivalent to the impedance according to ohm's law, V=IR.

I found the impedance at the frequency 6KHz to be 4.4kΩ which should correspond to mid band for my amplifier.

Using the same procedure, I found:

Transistor	Measured	Calculated
2N2222A	4.4k	3964
2N3904	3.7k	2767
2N4401	4.77k	3396

Table 6: Input impedance

Part D – Output Impedance

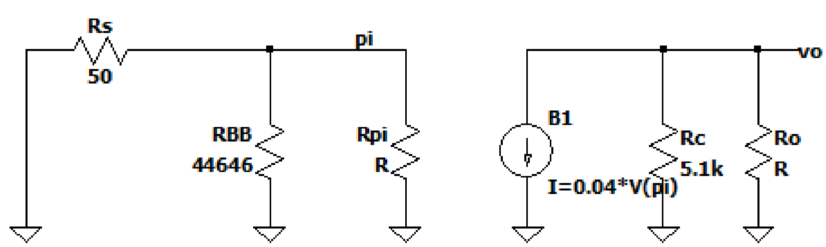


Figure 2-5: small signal model at Mid-band for calculating output resistance

To measure the output impedance, we look at the small signal model at mid band. In the small signal mid band model, the output of our amplifier is isolated from the rest of circuit and the only resistance seen should be from the collector resistor and r_o. If we apply Thevenin theorem we see that the only resistors that affect the output impedance is Rc and r_o

Output impedance is then given by $r_o || R_C$ for our other 2 transistors I chose values of r_o based on the data sheets for the respective transistors, I chose 27k and 60k as r_o for my 2n3904 and 2n4401 transistors respectively.

To measure output impedance, I connected a 6kHz AC source to my output much like I did with my input impedance and measured the voltage to give me the equivalent impedance.

πTransistor	Measured	Calculated
2N2222A	4.87k	4.89k
2N3904	4.22k	4.29k
2N4401	4.66k	4.7k

Table 7: Output impedance

Part E – Selecting the Best transistor

To make the best possible amplifier there's several things we want to look at, we want a high input impedance to minimize the effect of our amplifier on our source. We want low output impedance to minimize the power losses of our output signal. We want a bandwidth of our mid-band to be as large as possible in order to maximize the frequencies where it is amplifying. And finally, we want as high a saturation amplification as we can get.

Based on the above criteria I believe that the 2N3904 transistor is the best choice for building an amplifier. The bandwidth of the mid band is more than twice the size of the other 2 transistors. It also has the lowest output impedance of the 3 transistors. A close second choice would be the 2N4401 transistor as it has the 2nd biggest bandwidth as well as the 2nd smallest output impedance and the biggest saturation gain and input impedance. However, I chose the 2N3904 because of how much bigger the bandwidth is compared to the other 2 transistors. It has more than twice the operating range of the other 2 transistors because of it.

Conclusion:

In this mini project we learned to bias a transistor by hand and through simulations to turn it into a common emitter amplifier.

We discovered the convenience of the 1/3rd rule with biasing and how much of a time saver it can be when trying to bias a transistor. We saw that he 1/3rd rule provides a fairly accurate bias circuit with very little effort and that by biasing our transistor with the 1/3 we have enough leeway in our bias circuit that we can substitute in 2 different transistors and still a have a biased circuit.

We also got hands on experience calculating the small signal parameters of a transistor by plotting the Vbe, Vce, Ic, Ib data and finding the parameters graphically, and comparing the accuracy of our values to the ones found in the transistors data sheet.

We were also given the opportunity to apply the methods we learned in class to find the poles and zeros of a common emitter amplifier by calculating the short circuit time constant with the magnifying resistance.

We also saw the saturation point of our amplifier where our amplifier stops linearly amplifying our voltage and found the input and output impedances of our circuit to determine the best choice of transistor to use for our amplifier.

In the end I picked the 2N3904 as the best choice because of its outstanding mid band bandwidth.

Appendix:

General equations for calculating poles and zeros for our transistors:

$$k = -gm * (RC||RL)$$

$$WHp1 = (Cpi + Cu * (1-k) * (Rs||Rb1||Rb2||r_\pi))^{-1} = rad/sec * \frac{1}{2\pi} = Hz$$

$$WHp2 = \left(Cu * (1-\frac{1}{k}) * (RL||RC\right))^{-1} = rad/sec * \frac{1}{2\pi} = Hz$$

$$Wlp2 = (Cc1 * RL + Rc)^{-1} = rad/sec * \frac{1}{2\pi} = Hz$$

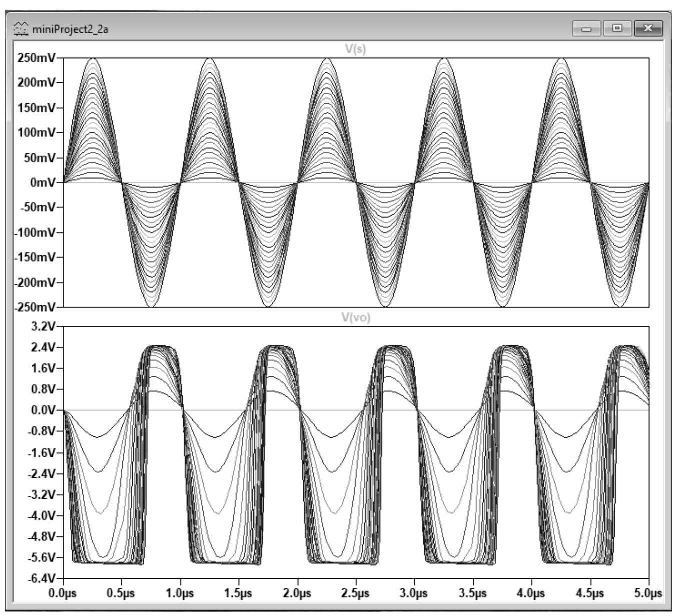
$$t_{Cc1}^{OC} = Cc1 * (50 + (Rb1||Rb2||(r_\pi + Re * (1+\beta))) = sec$$

$$Wlp1 = \frac{1}{t_{Cc1}^{OC}} = rad/sec * \frac{1}{2\pi} = Hz$$

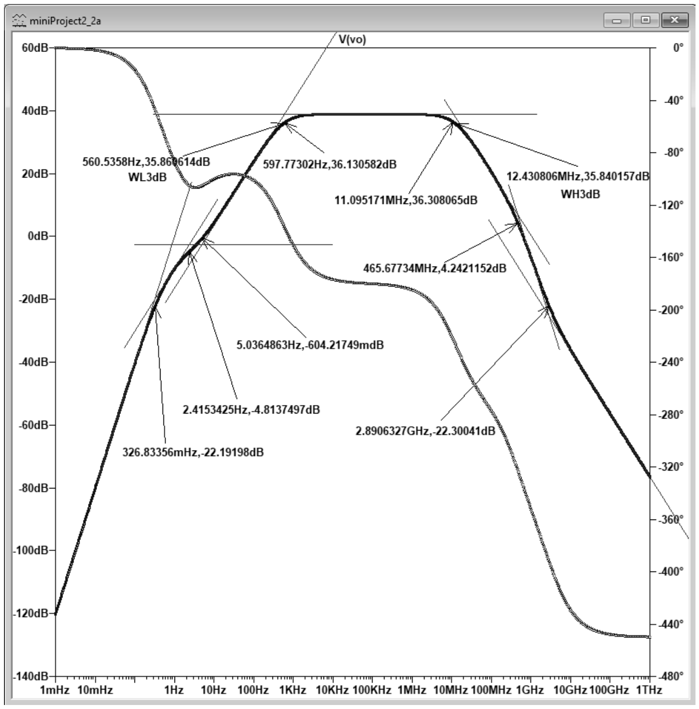
$$t_{CE}^{SC} = Ce * (Re||\left(\frac{(Rs||Rb1||Rb2) + r_\pi}{1+\beta}\right)) = sec$$

$$Wlp3 = \frac{1}{t_{CE}^{SC}} = \frac{rad}{sec} * \frac{1}{2\pi} = Hz$$

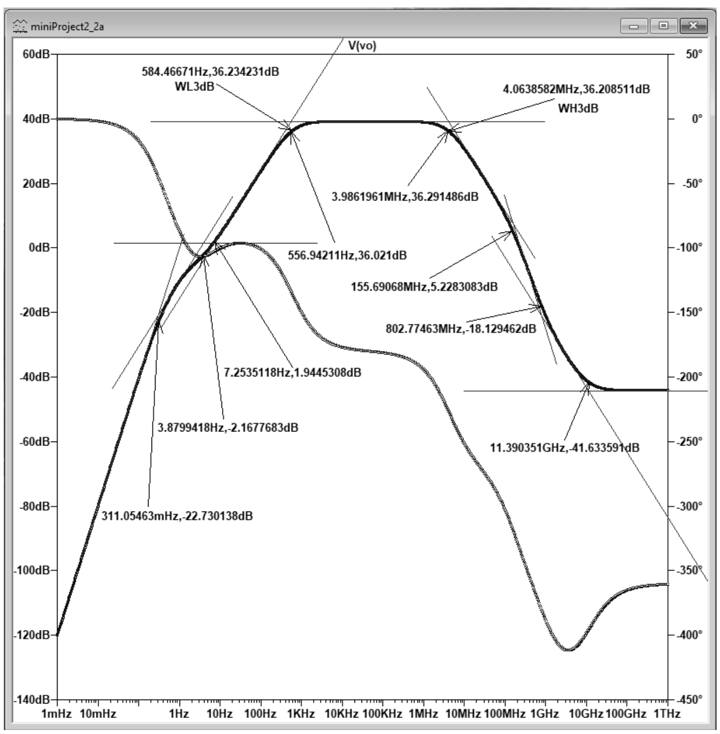
$$Wlz = \frac{1}{ReCe} = rad/sec * \frac{1}{2\pi} = Hz$$



Vo and Vs sinusoids



Bode plot of 2n3904(amplitude plot darkened)



Bode plot of 2n4401(amplitude plot darkened)