ECE 100: Linear Electronic Systems

Professor: Drew Hall

Lab 3:
Voltage Follower Circuit

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Abstract

This project aims to study the non-inverting voltage follower circuit when implemented with real op-amps.Ideal op-amps have infinite gain, infinite input impedance, and zero output impedance. In contrast, real op-amps are very complex, usually consisting of 10s of transistors and many passive elements. We experimented with an LF411 op-amp using two different Voltage-Follower circuits in order to examine the non-idealities. One configuration was without a load, and another with a capacitive load.

Experimental Procedure

Tools:

- Oscilloscope
- Function Generator
- Breadboard
- Electrical Components (LF411, Capacitors, Resistors)
- Digital Multimeter (Recommended)
- DC Dual-Channel Power Supply
- PSpice Simulation Software

A first-order Voltage-Controlled Voltage Source (VCVS) model captures much of the important behavior of a real op-amp. In this model, we define the gain as $A(s) = \frac{2\pi G}{S}$ where G is the unity-gain bandwidth. It is usually greater than 1 MHz, sometimes as high as 1 GHz. The input impedance is $R_i < 1M\Omega$, and the output impedance is $R_0 < 100\Omega$.

Section 1: Analysis

Part 1: Voltage Follower - Unloaded

The normal follower-with-gain circuit is shown below. Here, we assume that $R_0 \ll (R1 + R2)$ and $R_i \gg R1 \| R2$, where R_0 and R_i are the output and input impedances of the op-amp, and R1 and R2 are the resistors in the feedback network. The assumptions will be valid if R2 and R1 are of the order of 10kΩ.

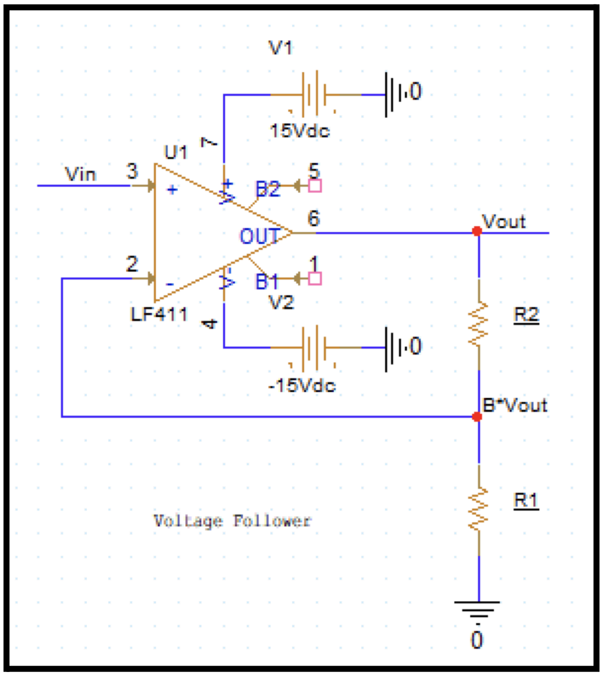


Figure 1.1.1: Schematic of unloaded op-amp

(a)

Derive rationalized polynomial expression for $A_{CL}(s)$, $Z_{IN}(s)$, $Z_{OUT}(s)$ in terms of $G(Hz)$, B, R_i, R_0. Hint: Use $A(s) = \frac{2\pi G}{s}$.

(b)

Show that $A_{CL}(s)$ is the transfer function of a low-pass filter with a dc gain of $\frac{1}{B}$ and a –3dB bandwidth of $G \times B$. Therefore, the product of the dc gain and the bandwidth is 1

$\frac{1}{B} \times G \times B = G$. This is the origin of the useful rule "the gain-bandwidth product of a feedback amplifier is constant." One must remember that this "rule" only applies to the first-order model of the op-amp.

(c)

Show that $Z_{IN}(s)$ can be modeled as a capacitor in series with R_i. So even when A is large, Z_{IN} is not resistive; it is capacitive.

(d)

Show that $Z_{OUT}(s)$ can be modeled as an inductor in parallel with Ro. So even when A is large, Z_{OUT} is not resistive but inductive.

Part 2: Voltage Follower with Capacitive Load

These results do not preclude the voltage follower from being used as a buffer but suggest that it may behave unexpectedly. For example, when loaded with a large capacitance, the buffer may behave like a resonant circuit rather than the ideal voltage source that one might have expected.

(a)

Write the circuit's transfer function (closed-loop gain) when a capacitor C_L is connected from the output to ground. Here $R_0$and $C_L$act like a low-pass filter. Put your expression in the general form. You will find that the denominator of the transfer function $\left(\frac{V_{out}(s)}{V_{in}(s)}\right)$ looks exactly like an RLC resonant circuit. Find ω_0 and ζ in terms of G, B, and$R_0 C_L$.

(b)

If the C_L is large, for example, if the buffer is driving a long coaxial cable run, the damping factor can be quite small, and the step response will show a great deal of "ringing." This is a common problem in both analog and digital integrated circuits. In large-scale digital systems, it can occur in the buffers which drive the system clock to the various logic components. It can be reduced or eliminated by placing a "compensation" resistor R_C in series with C_L. The output will still be taken across C_L. Find the transfer function $\left(\frac{V_{out}(s)}{V_{in}(s)}\right)$ of the modified circuit. Furthermore, show that ζ can be adjusted by choosing R_C without significantly changing the bandwidth ω_0.

Section 2: Simulations

Part 1: Unloaded Gain-Bandwidth Product

(a)

Simulate a voltage follower with a dc gain of 1, 3, 10, 30, and 100. Instead of an op-amp, use a VCVS with an open-loop gain $A(s) = \frac{2\pi G}{S}$, where $G = 10^6 Hz$. Measure the –3dB bandwidth in each case. Confirm that the Gain-Bandwidth product is constant. You can do all the simulations simultaneously using the Parameter property and a Parameter Sweep. Add a Parameter box to your schematic. Enter a parameter called "r" and give it a value of 1k. Then set $R2 = 10k$ and $R1 = \{r\}$. In the analysis setup menu, add a Parameter Sweep with a value list. Enter the values of R1 required to give the gains listed above. PSpice will then do an AC Sweep for each value of r in the list and overlay them all. Make a copy of the plot and your schematic. You can measure the –3dB frequency manually with the cursor.

(b)

Repeat the simulation for a voltage follower with a gain of 1, 3, 10, 30, and 100 using an LF411 op-amp. Remember to connect the power supplies. Note that the –3dB bandwidth is defined with respect to the dc gain, not the maximum gain. Make a copy of the plot. You will see that the gain-bandwidth product is constant for the higher gains but increases near unity gain. In fact, the LF411 op-amp has several poles and a right half-plane zero. The phase lag due to these high-frequency "features" causes the gain-bandwidth product to change. They also have other interesting effects - they can make the follower oscillate under certain conditions.

Part 2: Effect of Capacitive Load

(a)

Use the LF411 op-amp in the unity gain voltage follower configuration to simulate the step response for load capacitances of 100pF, 300pF, 1nF, 2nF, 4nF, 6nF, and 10nF. Use the PARAM feature to get PSpice to run the entire list simultaneously and overlay the plots. A time range of 0 to 2μs should be sufficient. Make a copy of the plot. You will see that loads larger than 100pF can cause a lot of ringing. The typical capacitance of a coaxial cable is $30\frac{pF}{ft}$. So, it is easy to get a lot of ringing while driving a cable incautiously.

Note: Refer to the image below of the pulse waveform to understand the parameters.

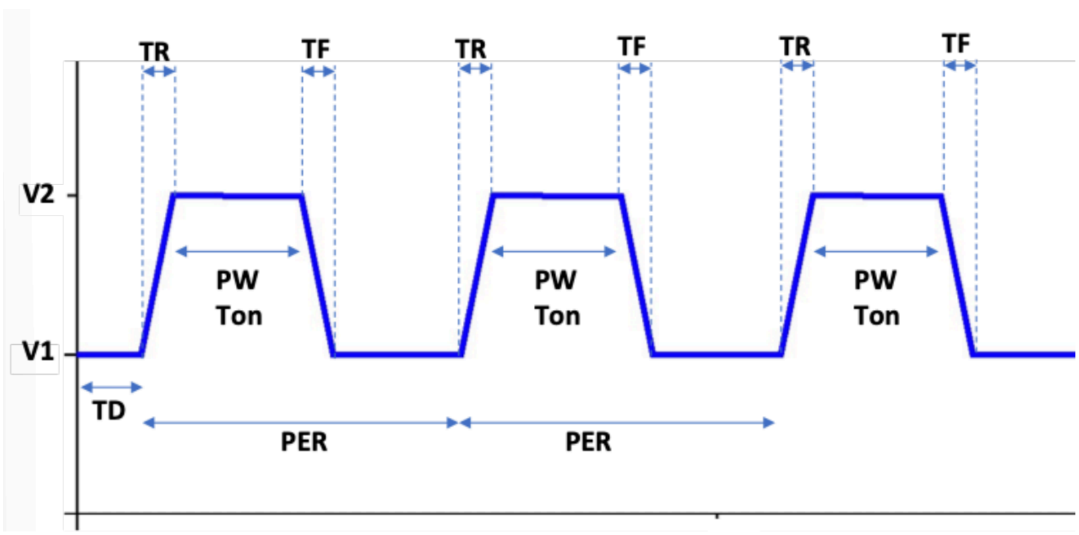


Figure 2.1.1: Parameters for Pulse-Width response

(b)

Put a 25Ω resistor in series with the load capacitance and rerun the simulation. Make a copy of this plot. You will see that it clearly mitigates the overshoot. In practice, you would adjust the compensation resistor to be as small as possible, consistent with the desired overshoot, because larger resistors will cause a slower rise time. Generally, you would like the shortest feasible rise time. What is the smallest resistor to keep the overshoot < 11% for all loads?

(c)

In a practical situation, you would have a certain load capacitance, say 3 nF, and you would want to compensate for it to get the shortest rise-time consistent with some maximum overshoot, say 15%. Connect a 3nF load and optimize the compensation resistor. What resistor works best? What was the optimal 10% to 90% rise time? Make a plot of the optimized step response.

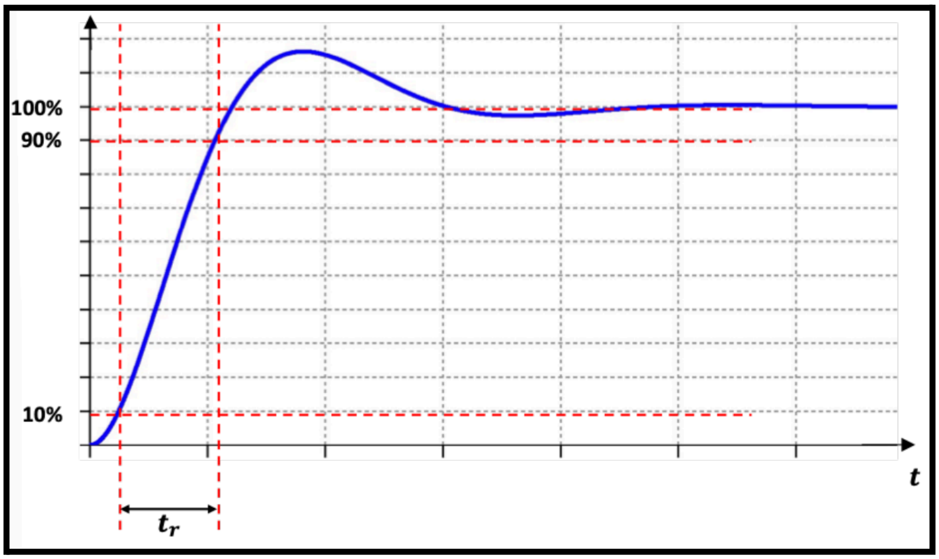


Figure 2.1.2: Overshoot Parameters with Capacitive Load

Section 3: Measurements

Part 1: Unloaded Gain-Bandwidth Product

In this section, you will measure the gain bandwidth product of the follower under the same conditions as the simulation in Part 1-(b). The measurements will require some care, and the circuit must be laid out neatly because the bandwidth is large. Bypass both dc supplies to the ground with a 0.1µF capacitor close to the chip. Special care is needed for the feedback when $B = 1$. You will not be able to use $R2 = 10K$ and $R1 = Some\ large\ value$ because the capacitance on the breadboard is about 6pF to ground and will cause enough phase shift to greatly increase the ringing (or even make it oscillate). Set **R2 = short circuit** and **R1 = open circuit** instead. The exact values of the gain are unimportant – use the closest convenient resistors.

First, check the calibration of your scope probes and adjust them if necessary. The 10X probe setting must match the oscilloscope setting. It will be necessary to respect the "slew-rate limit". For the LF411 this is about $10\frac{v}{\mu s}$. This means that $\left|\frac{dV_0}{dt}\right| < 10\frac{v}{\mu s}$. If $V_0(t) = Asin(2\pi ft)$ then $\frac{dV_0}{dt} = 2\pi fAcos(2\pi ft)$. So we need $A < \frac{10^7}{2\pi f}$ or slew rate limiting will occur and mess up our measurements, e.g. at $f = 1MHz$ the peak-to-peak $V_0 < 3.18V$. If the gain = 100, the input voltage will be only 31.8mV . This is low to measure or trigger from, and you may need to trigger from the output and use averaging to reduce the noise on the input. Using the 10X probe setting on the output will be necessary because the 1X probe has considerable capacitance. You can use the 1X probe setting on the input, though. Measure the gain-bandwidth product for each case you simulated in Simulation 1-(b). Remember that –3dB is with respect to the dc gain, not the maximum gain.

Part 2: Capacitive Load

In this section, you will measure the effect of a capacitive load on a unity-gain follower, as simulated in Simulation Part 2. The same precautions against slew rate limiting apply. The input step must be $\ll$ 1V, and you must use the 10X probe setting on the output. You might as well use it on both because both signals will be about the same amplitude. Remember to use $R2\ =$ short circuit and $R1\ =$ open circuit or you will see too much overshoot caused by the breadboard.

(a)

Measure the overshoot for (roughly) the same capacitances that you used in the simulation. Here the exact values are not important. You may find that the follower oscillates with some values of C_L. If it does oscillate, the amplitude will be limited by the slew rate, so you won't necessarily see a large signal. What you might see is a ringing that never dies away. If this happens, make a hard copy for your report.

(b)

In a practical situation, you would have a certain load capacitance, say 3nF, and you would want to compensate for it to get the shortest rise-time consistent with some maximum overshoot, say 11%. Connect a 3nF load and optimize the compensation resistor. What resistor works best? What was the optimal 10% to 90% rise time?

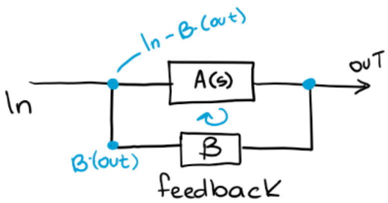
Results

Section 1: Analysis

Part 1: Voltage Follower - Unloaded

(a)

Derive rationalized polynomial expression for $A_{CL}(s)$, $Z_{IN}(s)$, $Z_{OUT}(s)$ in terms of $G(Hz)$, B, R_i, R_0.

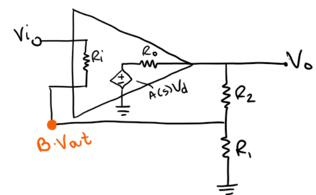


$$OUT = a(s)\left[IN - B\cdot(OUT)\right]$$

$$OUT = a(s)\cdot IN - B\cdot(OUT)\cdot a(s)$$

$$OUT\left(1 + a(s)\cdot B\right) = a(s)\cdot IN$$

$$\boxed{A_{cl} = \frac{OUT}{IN} = \frac{a(s)}{1 + a(s)\cdot B}}$$



$$I_{IN} = \frac{V_{IN} - BV_o}{R_i} = \frac{V_{in}}{R_i}(1 - BH)$$

$$* \quad H = \frac{V_o}{V_{IN}} = \frac{aB}{1 + aB} \quad *$$

$$I_{IN} = \frac{V_{IN}}{R_i}\left(1 - \frac{aB}{1 + aB}\right)$$

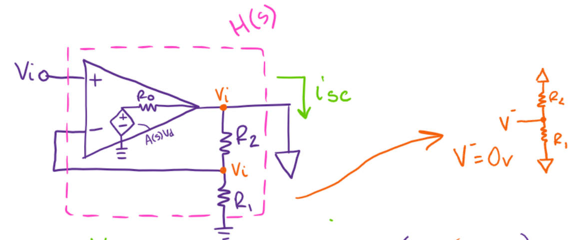
$$= \frac{V_{IN}}{R_i}\left(\frac{1}{1 + aB}\right)$$

$$= \frac{V_{IN}}{R_i}\left(\frac{1}{1 + \left(\frac{2\pi G}{s}\right)B}\right)$$

$$I_{IN} = \frac{V_{IN}}{R_i}\left(\frac{1}{1 + \frac{2\pi GB}{s}}\right)$$

$$Z_{IN} = \frac{V_{IN}}{I_{IN}} = \frac{\cancel{V_{IN}}}{\left(\frac{\cancel{V_{IN}}}{R_i}\right)\frac{1}{1 + \frac{2\pi GB}{s}}}$$

$$\boxed{Z_{IN} = R_i\left(1 + \frac{2\pi GB}{s}\right)}$$



V_{oc}

$$V_o = H(s)\cdot V_i$$

$$V_o = \left[\frac{a(s)}{1 + a(s)B}\right]V_i$$

I_{sc}

$$V^- = 0v$$

$$* \quad R_1 \,\&\, R_2 >> R_o \quad *$$

$$I_{sc} = \left(\frac{V_i - 0}{R_o}\right)a(s)$$

$$I_{sc} = \frac{V_i\cdot a(s)}{R_o}$$

$$Z_{OUT} = \frac{V_{oc}}{I_{sc}} = \frac{\left(\frac{\cancel{a(s)}}{1 + a(s)B}\cdot \cancel{V_i}\right)}{\left(\frac{\cancel{a(s)}}{R_o}\cdot \cancel{V_i}\right)}$$

$$Z_{OUT} = \frac{R_o}{1 + a(s)\cdot B}$$

$$= \frac{R_o}{1 + \left(\frac{2\pi G}{s}\right)B}$$

$$\boxed{Z_{OUT} = R_o\left(\frac{1}{1 + \frac{2\pi GB}{s}}\right)}$$

(b)

Show that $A_{CL}(s)$ is the transfer function of a low-pass filter with a dc gain of $\frac{1}{B}$

$$\boxed{A_{cl} = \frac{OUT}{IN} = \frac{a(s)}{1 + a(s)\cdot B}}$$

$$A_{cl} = \frac{\left(\frac{2\pi G}{s}\right)}{1 + \left(\frac{2\pi G}{s}\right)B}$$

$$= \frac{\cancel{\left(\frac{2\pi G}{s}\right)}}{\cancel{\frac{2\pi G}{s}}\left[\frac{1}{\left(\frac{2\pi G}{s}\right)} + B\right]}$$

$$A_{cl} = \frac{1}{\frac{s}{2\pi G} + B}$$

$$\boxed{A_{cl}(s) = \underbrace{\frac{1}{B}}_{\text{DC Gain}}\frac{1}{\left(\frac{s}{2\pi GB} + 1\right)}}$$

$A_{CL}(s)$ is the transfer function of a low pass filter with a dc gain of $\frac{1}{B}$ and bandwidth GB . A real pole indicated that the system will attenuate frequencies.

$$Gain\ bandwidth\ product\ =\ \frac{1}{\beta}\cdot G\beta\ =\ G$$

(c)

Show that $Z_{IN}(s)$ can be modeled as a capacitor in series with R_i Show that $Z_{OUT}(s)$ can be modeled as an inductor in parallel with Ro.

$$Z_{IN} = \frac{V_{IN}}{I_{IN}} = \frac{\cancel{V_{IN}}}{\left(\frac{\cancel{V_{IN}}}{R_i}\,\frac{1}{1+\frac{2\pi GB}{s}}\right)}$$

$$\boxed{Z_{IN} = R_i\left(1+\frac{2\pi GB}{s}\right)}$$

$$Z_{IN} = R_i + C_{eff}$$

$$* \; C_{eff} = \frac{1}{2\pi GBR_i} \; *$$

$$\boxed{Z_{IN} = R_i + \frac{1}{sC_{eff}}}$$

(d)

Show that $Z_{OUT}(s)$ can be modeled as an inductor in parallel with Ro.

$$\boxed{Z_{OUT} = R_o\left(\frac{1}{1+\frac{2\pi GB}{s}}\right)}$$

$$\star \; Z_{OUT} = R_o \| L_{eff} \; \star$$

$$\boxed{L_{eff} = \frac{R_o}{2\pi GB}}$$

Solving for R_1 Values

Calculating Gain

$$\text{Gain} = \frac{1}{B} \qquad B = \frac{R_1}{R_1 + R_2}$$

$$\text{Gain} = \frac{R_1 + R_2}{R_1}$$

$$R_1 G = R_1 + R_2$$

$$R_1 (G - 1) = R_2$$

$$R_1 = \frac{R_2}{G - 1}$$

when $G = 100$: $R_2 = 10k$

$$R_1 = \frac{10k}{99} = 101.01\ \Omega$$

when $G = 30$

$$R_1 = \frac{10k}{29} = 344.82\ \Omega$$

when $G = 10$

$$R_1 = \frac{10k}{9} = 1111\ \Omega$$

when $G = 3$

$$R_1 = \frac{10k}{2} = 5k\Omega$$

when $G = 1$

$$R_1 = \frac{10k}{.01} = 1\,\text{Meg}\ \Omega$$

$$H(0) = \frac{1}{B} \qquad \cancel{\frac{1}{1}}$$

$$H(0) = \frac{1}{B} \qquad \beta = \frac{R_1}{R_1 + R_2}$$

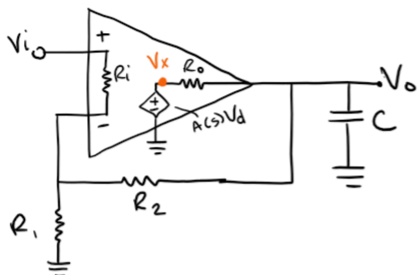
$H(j\omega)$

We calculated the resistor values we needed in order to achieve the desired gains to test the circuit.

Part 2: Voltage Follower with Capacitive Load

(a)

Write the circuit's transfer function (closed-loop gain) when a capacitor C_L is connected from the output to ground.



$$V^- = B\cdot V_o = H(s)V_o$$

$$V_x = a(s)\left[V^+ - V^-\right]$$

$$V_x = a(s)\left[V_i - BV_o\right]$$

$$\star \quad B = \frac{R_1}{R_1+R_2} \quad \star$$

Assuming R_1 & $R_2 >> R_o$

$$V_o = \frac{V_x}{1+sR_oC}$$

$$V_x = a(s)\left[V_{IN} - B\left(\frac{V_x}{1+sR_oC}\right)\right]$$

$$V_x + \frac{a(s)\cdot B\cdot V_x}{1+sR_oC} = a(s)V_{IN}$$

$$V_x\left(1+\frac{a(s)\cdot B}{1+sR_oC}\right) = a(s)V_{IN}$$

$$\frac{V_x}{V_{IN}} = \frac{a(s)}{\left(1+\frac{a(s)\,B}{1+sR_oC}\right)}$$

$$= \frac{a(s)}{\left(\frac{(1+sR_oC)+a(s)B}{1+sR_oC}\right)}$$

$$\frac{V_x}{V_{IN}} = \frac{(1+sR_oC)\cdot a(s)}{1+sR_oC+a(s)\cdot B} \quad (1)$$

Using (1)

$$\frac{V_{OUT}}{V_{IN}} = \frac{V_{OUT}}{V_x}\cdot\frac{V_x}{V_{IN}}$$

$$= \left(\frac{1}{\cancel{1+sR_oC}}\right)\cdot\left[\frac{(\cancel{1+sR_oC})\cdot a(s)}{1+sR_oC+a(s)\cdot B}\right]$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{a(s)}{1+sR_oC+a(s)\cdot B}$$

$$* \quad \tau = R_oC \quad *$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{a(s)}{1+s\tau+a(s)\cdot B} = \frac{\left(\frac{2\pi G}{s}\right)}{1+s\tau+\left(\frac{2\pi GB}{s}\right)}$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{\left(\frac{2\pi G}{\cancel{s}}\right)}{\left(\frac{s+s^2\tau+2\pi GB}{\cancel{s}}\right)} = \frac{2\pi G}{2\pi GB+s+s^2\tau}$$

$$= \frac{2\cancel{\pi}G}{2\cancel{\pi}GB\left(1+\frac{s}{2\pi GB}+\frac{s^2\tau}{2\pi GB}\right)}$$

$$\boxed{\frac{V_{OUT}}{V_{IN}} = \frac{1}{B}\left(\frac{1}{1+\frac{s}{2\pi GB}+\frac{s^2\tau}{2\pi GB}}\right)}$$

General Form:

$$H(s) = \frac{1}{1+2\zeta\frac{s}{\omega_o}+\frac{s^2}{\omega_o^2}}$$

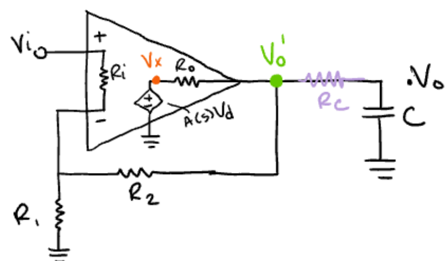
$$\frac{2\zeta}{\omega_o} = \frac{1}{2\pi GB} \qquad \frac{1}{\omega_o^2} = \frac{\tau}{2\pi GB}$$

$$\zeta = \frac{\omega_o}{4\pi GB} \qquad \boxed{\omega_o = \sqrt{\frac{2\pi GB}{\tau}}}$$

$$\boxed{\zeta = \frac{1}{4\pi GB}\sqrt{\frac{2\pi GB}{\tau}}}$$

(b)

Find ω_0 and ζ in terms of G, B, and $R_0 C_L$.



$$V_o' = a(s)[V_{in} - BV_o]\left(\frac{R_c + \frac{1}{sC}}{R_o + R_c + \frac{1}{sC}}\right)\cdot\left(\frac{sC}{sC}\right)$$

$$= a(s)[V_{in} - BV_o]\left(\frac{sCR_c + 1}{sCR_o + sCR_c + 1}\right)$$

assuming $R_1 \& R_2 \gg R_o$

$\tau_L = R_o C$, $\tau_c = R_c C$

$$V_o' = a(s)[V_{in} - BV_o]\left(\frac{s\tau_c + 1}{s(\tau_L + \tau_c) + 1}\right)$$

$$= a(s)\left[V_{IN} - B\left(\frac{V_o'}{1 + s\tau_c}\right)\right]\left(\frac{s\tau_c + 1}{s(\tau_L + \tau_c) + 1}\right)$$

$$= \left(a(s)V_{IN} - \frac{a(s)B\cdot V_o'}{\cancel{1 + s\tau_c}}\right)\left(\frac{\cancel{s\tau_c + 1}}{s(\tau_L + \tau_c) + 1}\right)$$

$$V_o' = a(s)V_{IN}\left(\frac{s\tau_c + 1}{s(\tau_L + \tau_c) + 1}\right) - \left(\frac{a(s)\cdot BV_o'}{s(\tau_L + \tau_c) + 1}\right)$$

$$V_o'\left(1 + \frac{a(s)\cdot B}{s(\tau_L + \tau_c) + 1}\right) = a(s)V_{in}\left(\frac{s\tau_c + 1}{s(\tau_L + \tau_c) + 1}\right)$$

$$\frac{V_o'}{V_{in}} = \frac{\left[\frac{a(s)\cdot(s\tau_c + 1)}{s(\tau_c + \tau_L) + 1}\right]}{\left[1 + \frac{a(s)\cdot B}{s(\tau_L + \tau_c) + 1}\right]}$$

$$= \left[\frac{a(s)\cdot(s\tau_c + 1)}{\cancel{s(\tau_c + \tau_L) + 1}}\right]\cdot\left[\frac{s(\cancel{\tau_L + \tau_c}) + 1}{s(\tau_L + \tau_c) + 1 + a(s)\cdot B}\right]$$

$$\frac{V_o'}{V_{in}} = \frac{a(s)\cdot(s\tau_c + 1)}{s(\tau_L + \tau_c) + 1 + a(s)B} \quad (1)$$

$*\ a(s) = \frac{2\pi G}{s}\ *$

after simplification

$$\frac{V_o'}{V_{IN}} = \frac{1}{B}\,\frac{1}{1 + s\frac{1}{2\pi GB} + \frac{1 + s(\tau_c + \tau_L)}{1 + s\tau_L}}$$

using (1)

$$\frac{V_o}{V_o'} = \frac{1}{1 + s\tau_L}$$

$$\frac{V_{OUT}}{V_{in}} = \frac{V_{OUT}}{\cancel{V_o'}}\cdot\frac{\cancel{V_o'}}{V_{in}} = \left(\frac{1}{1 + s\tau_L}\right)\left[\frac{1}{B}\cdot\frac{1}{1 + s\frac{1}{2\pi GB} + \frac{1 + s(\tau_c + \tau_L)}{1 + s\tau_L}}\right]$$

$$\boxed{\frac{V_{OUT}}{V_{IN}} = \frac{1}{B}\,\frac{1}{1 + s\left(\tau_c + \frac{1}{2\pi GB}\right) + s^2\frac{(\tau_L + \tau_c)}{2\pi GB}}}$$

New	OLD (w/o R_c)
$\omega_0 = \sqrt{\frac{2\pi GB}{\tau_L + \tau_c}}$	$\omega_0 = \sqrt{\frac{2\pi GB}{\tau_L}}$
$\zeta = \frac{1}{2}\left(\tau_c + \frac{1}{2\pi GB}\right)\sqrt{\frac{2\pi GB}{\tau_L + \tau_c}}$	$\zeta = \frac{1}{4\pi GB}\sqrt{\frac{2\pi GB}{\tau}}$

We see that ξ is more sensitive to changes and ω_0 is less sensitive to changes.

Section 2: Simulation

Part 1: Unloaded Gain-Bandwidth Product

(a)

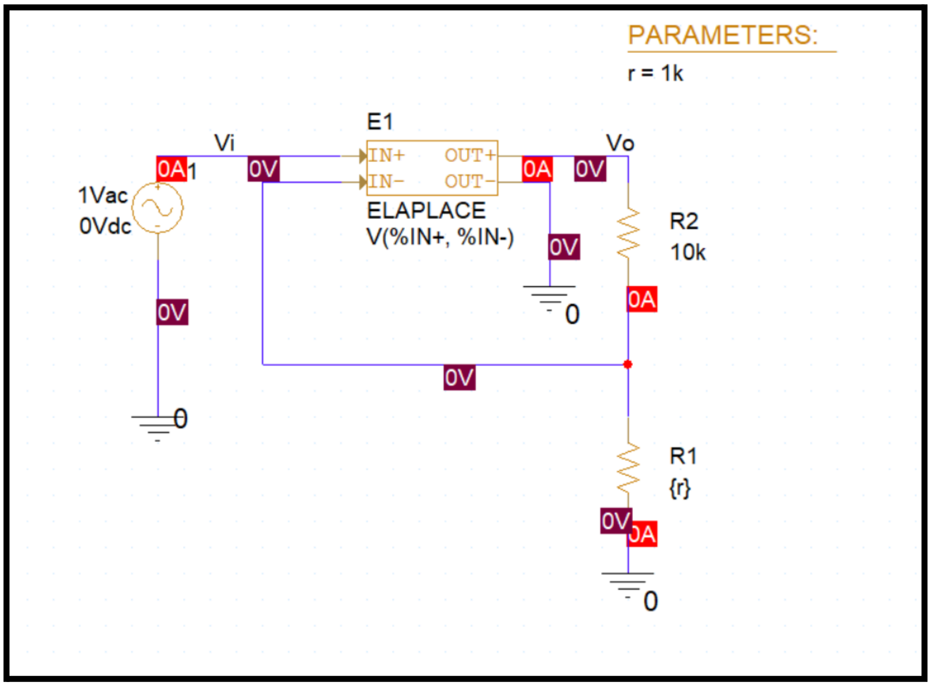


Figure 2.1.3: PSpice schematic of unloaded op-amp

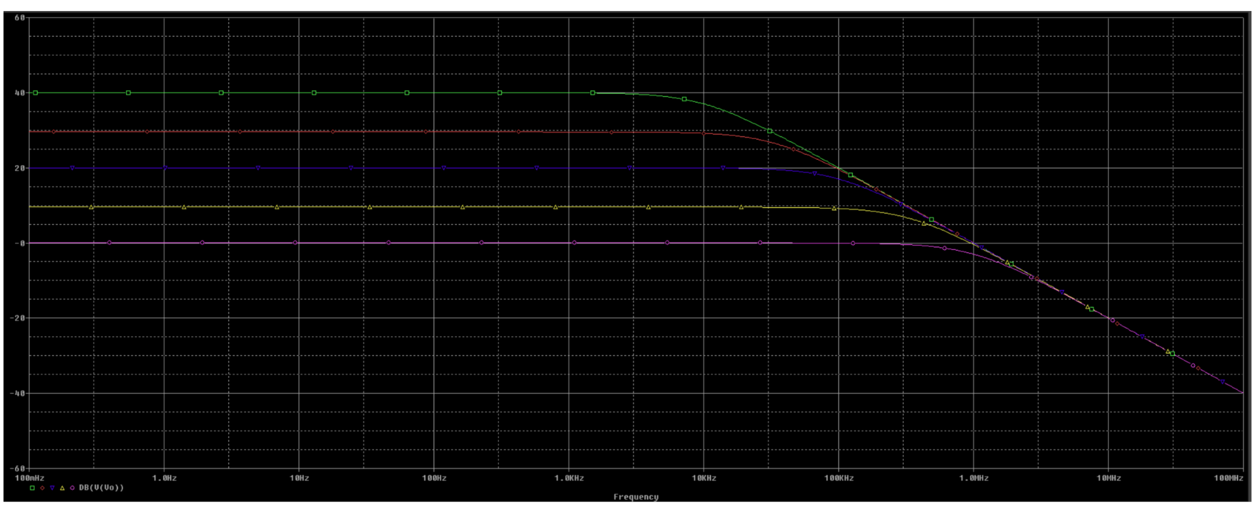


Figure 2.1.4: Parametric Sweep for all Gains

(100→Green, 30 → Red, 10 → Blue, 3→ Yellow, 1 → Purple)

$$Gain\ bandwidth\ product\ =\ a_0\omega_p$$

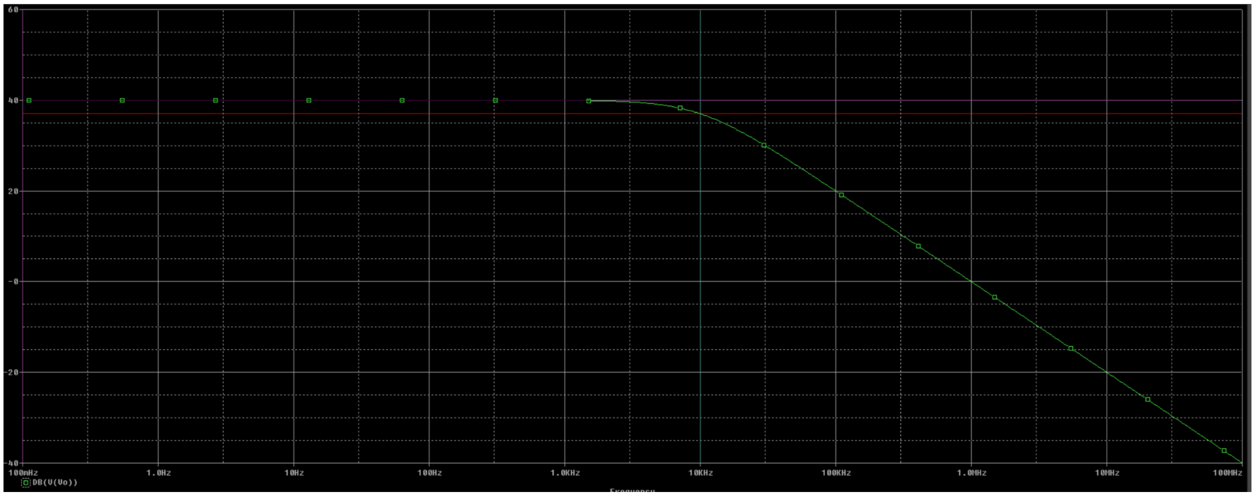


Figure 2.1.5: DC gain of 100 with Resistor Value 101Ω

For a_0= 100: Gain: 40dB at 36.990 dB , ω_p= 10.000K

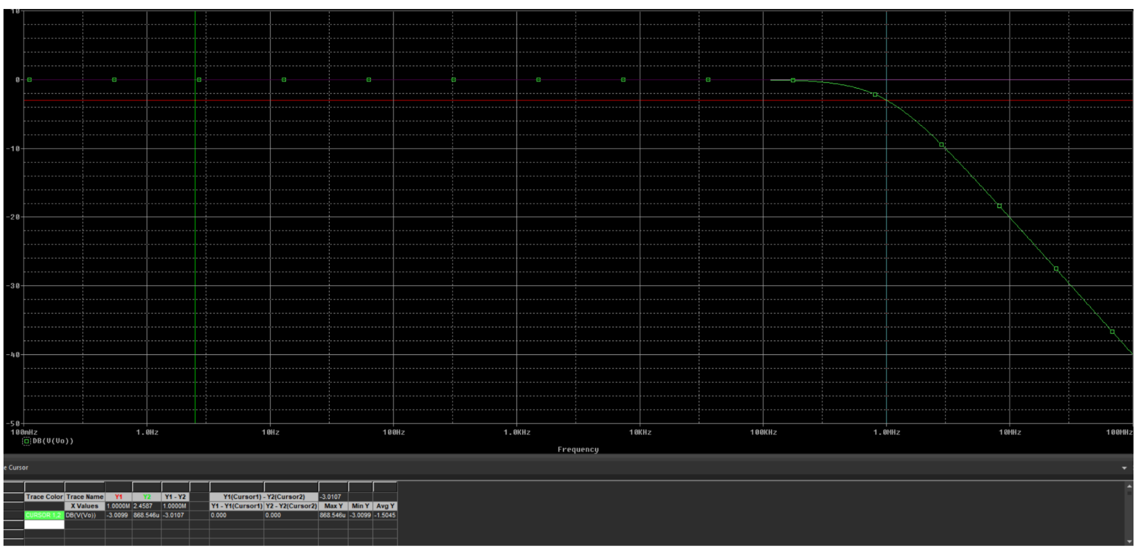


Figure 2.1.6: DC gain of 1 with Resistor Value 1MegΩ

For a_0= 1: Gain:868.546u dB at-3.0099 dB , ω_p=1.0000MHz

$$Gain\ bandwidth\ product = 1 * 1.0000M = 1M$$

Parameter	Gain	Desired Gain	**(Simulated)** Frequency at A_D	Gain Bandwidth Product
Units	A (dB)	$A_D = (A - 3dB)$	ω_p	$= a_0\omega_p$
Data	A = 1 (0dB)	-3 dB	1.000 MHz	1MHz
	A = 3 (9.542 dB)	+6.4951 dB	331.841kHz	0.995MHz
	A = 10 (20 dB)	+16.9 dB	100.000 KHz	$1MHz$
	A = 30 (29.538 dB)	+26.526 dB	33.184 KHz	0.995MHz
	A = 100 (40 dB)	+36.990 dB	10.000 KHz	1M Hz

Figure 2.1.7: Table of all Simulated ω_p

(b)

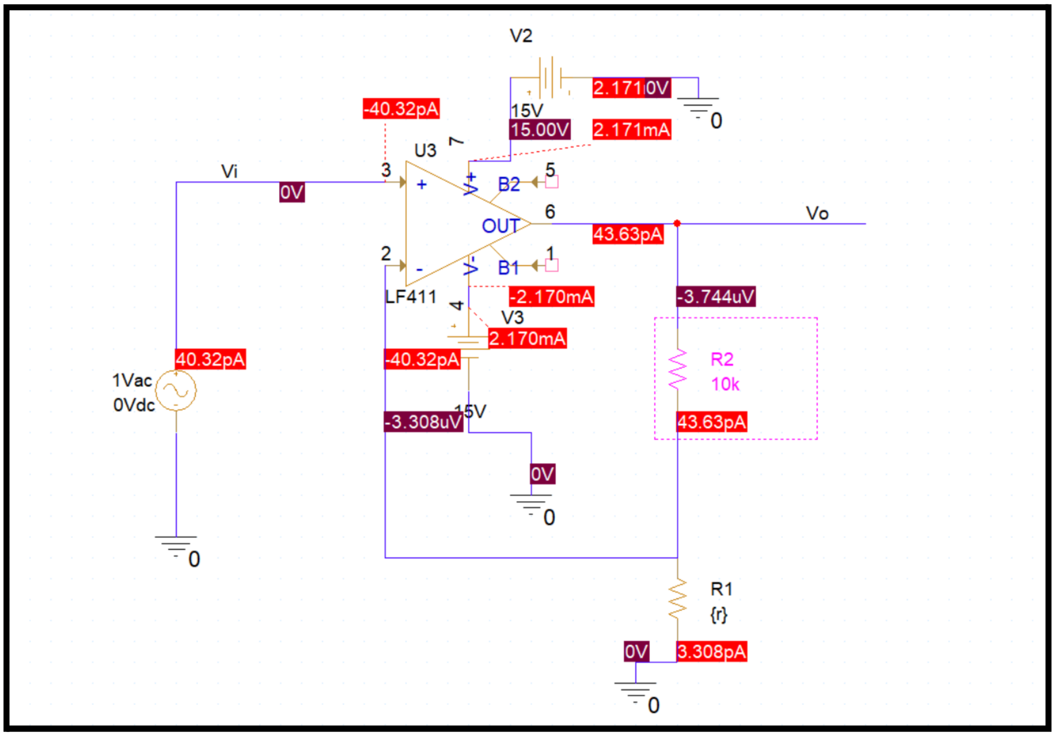


Figure 2.1.8 : Schematic with Resistor Sweep on R_1 and R_2 = 10k Ω

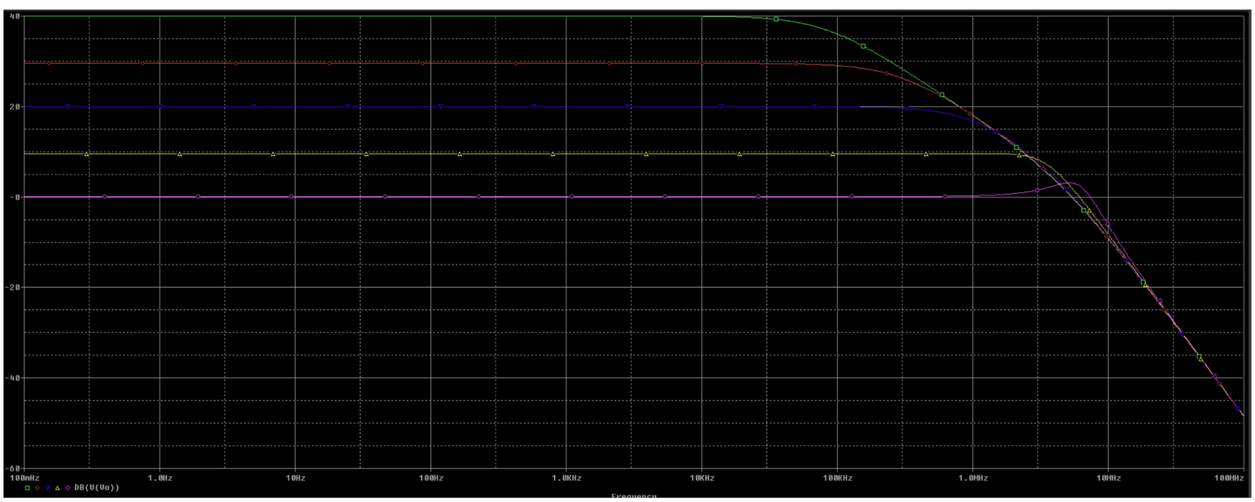


Figure 2.1.9: Parametric Sweep for all gains with calculated R_1

Gain (100→Green, 30 → Red, 10 → Blue, 3→ Yellow, 1 → Purple) for calculated R_1

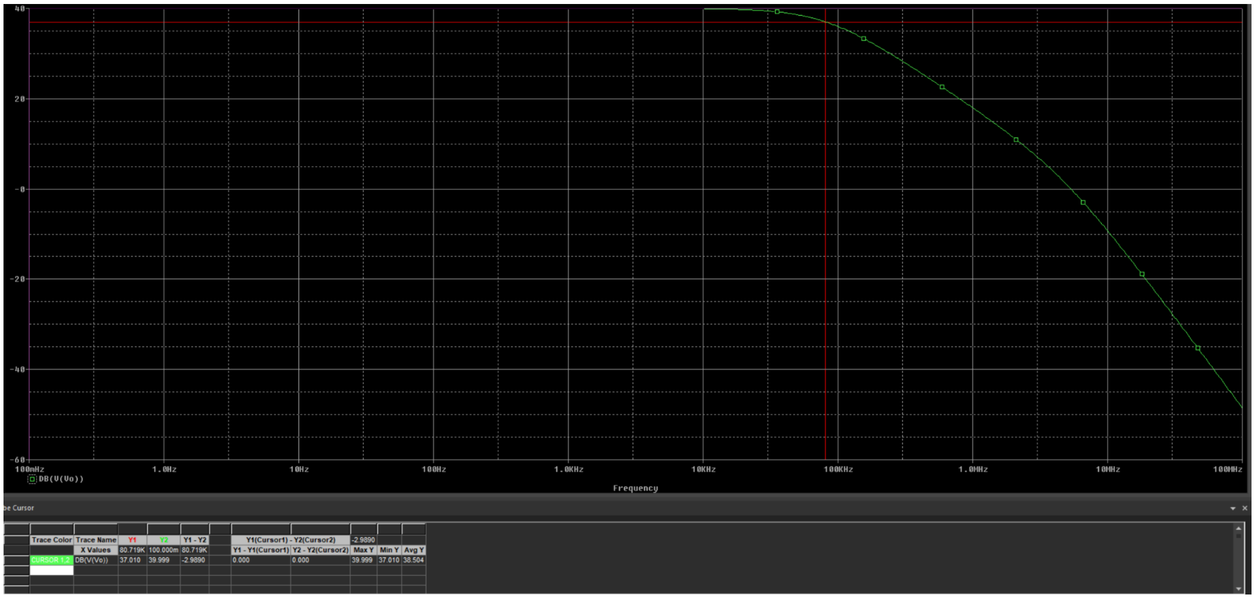


Figure 2.1.10: DC gain of 100 with Resistor Value 101Ω

For a_0= 100: Gain:39.999dB At 37.010 dB , ω_p=80.719KHz

$$Gain\ bandwidth\ product\ =\ 100\ *\ 80.719K\ =\ 8.072M$$

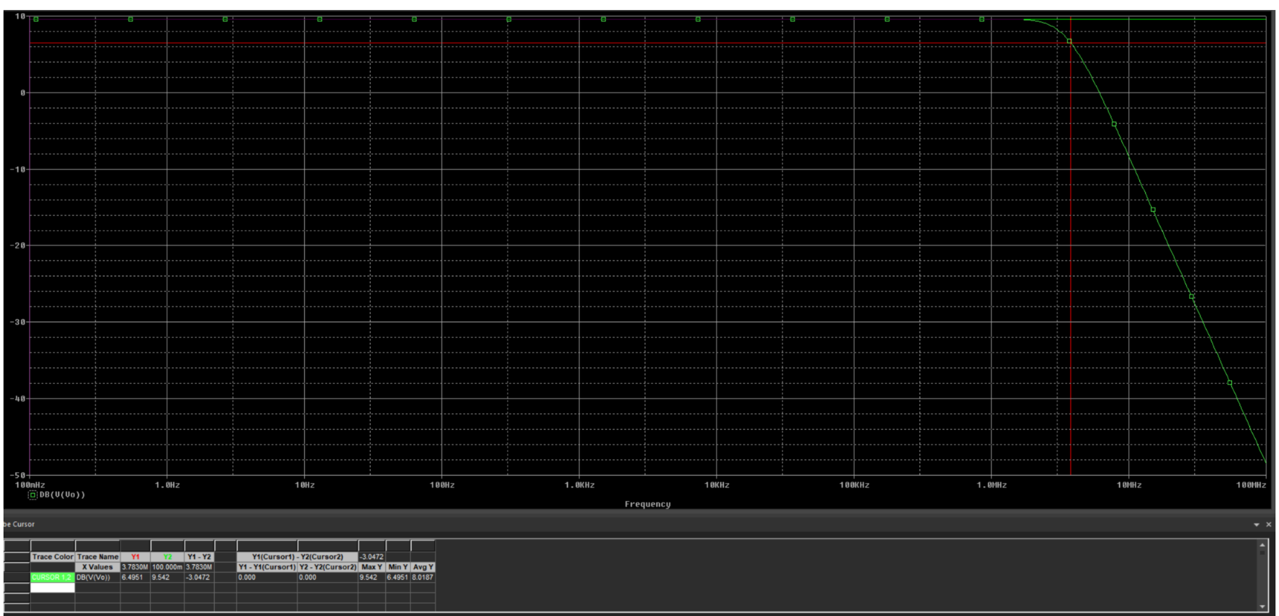


Figure 2.1.11: DC gain of 3 with Resistor Value 5kΩ

For a_0= 3: Gain: 9.542dB At 6.4951dB , ω_p=3.7830MHz

$$Gain\ bandwidth\ product\ =\ 3\ *\ 3.7830M\ =\ 11.349M$$

Parameter	Gain	Desired Gain	**(Simulated)** Frequency at A_D	Gain Bandwidth Product
Units	A (dB)	$A_D = (A - 3dB)$	ω_p	$= a_0\omega_p$
Data	A = 1 (0dB)	-3 dB	8.5114 MHz	8.512MHz
	A = 3 (9.542 dB)	+6.4951 dB	3.7830 MHz	11.35Mhz
	A = 10 (20 dB)	+17 dB	947.860 KHz	$9.479 MHz$
	A = 30 (29.538 dB)	+26.526 dB	282.594 KHz	8.478MHz
	A = 100 (40 dB)	+36.990 dB	80.719 KHz	8.072M Hz

Figure 2.1.12: Table of Simulated ω_p and Gain Bandwidth Product for Desired Gain

Part 2: Voltage Follower with Capacitive Load

(a)

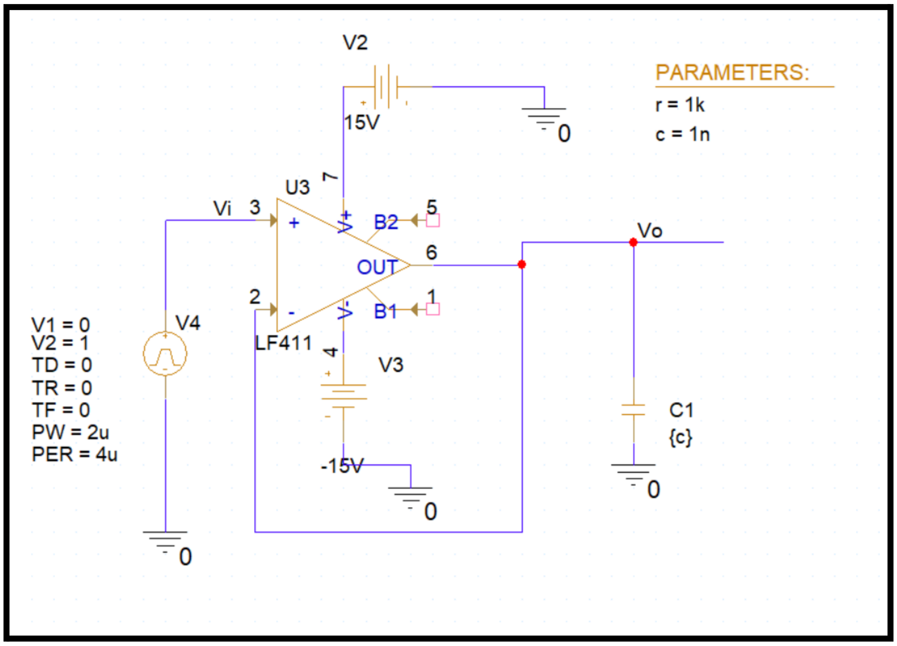


Figure 2.2.1 : Schematic of Unity-Gain Follower for load Capacitances

C = 100pF, 300pF, 1nF, 4nF, 6nF, and 10nF

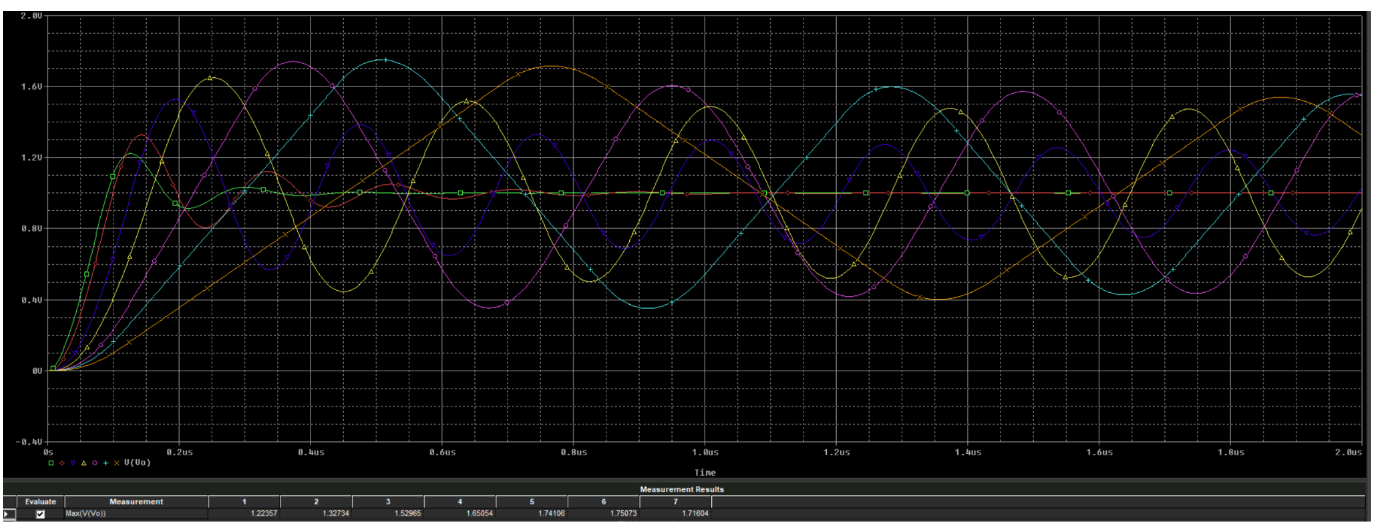


Figure 2.2.2: Plots for C = 100pF, 300pF, 1nF, 4nF, 6nF, and 10nF

We see a lot of overshooting here. 22% for green, 32% for red, 52% for purple, 65% for yellow, 74% for pink , 75% for blue , 71% for orange.

As Capacitance goes upOvershoot goes up and saturates

b)

Put a 25 Ω resistor. Adjust compensation resistors to be as small as possible, consistent with the desired overshoot, because larger resistors will cause a slower rise time.

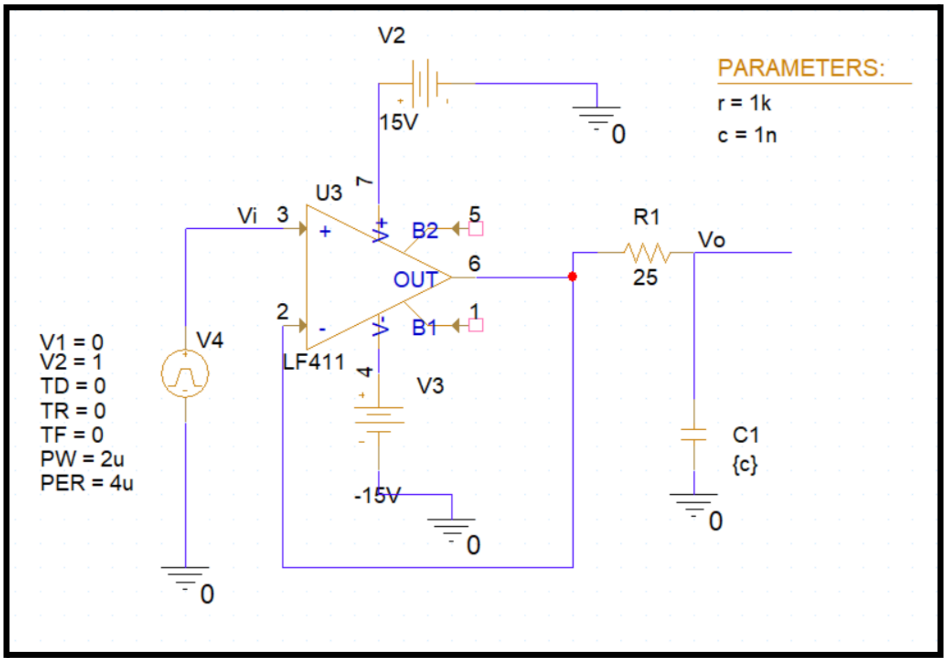


Figure 2.2.3: Schematic of Unity-Gain Follower with Compensation Resistor (25Ω)

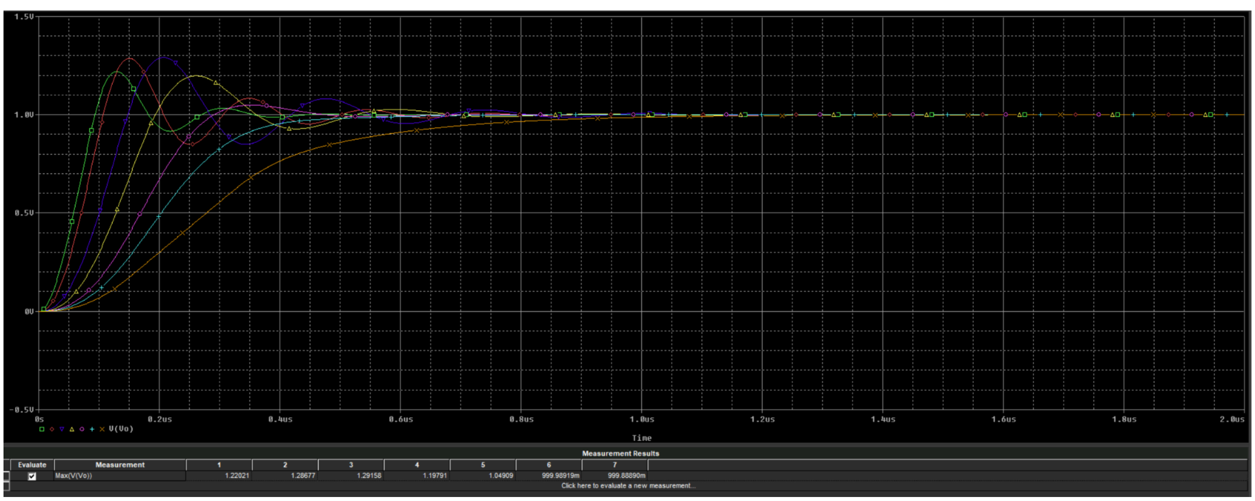


Figure 2.2.4: Plot of Unity-Gain Follower with $R_C = 25\Omega$ for all C

22% for Green, 26% for Red, 29% for Purple, 19% for Yellow, 4% for Pink , 0% for Blue , 0% for Orange.

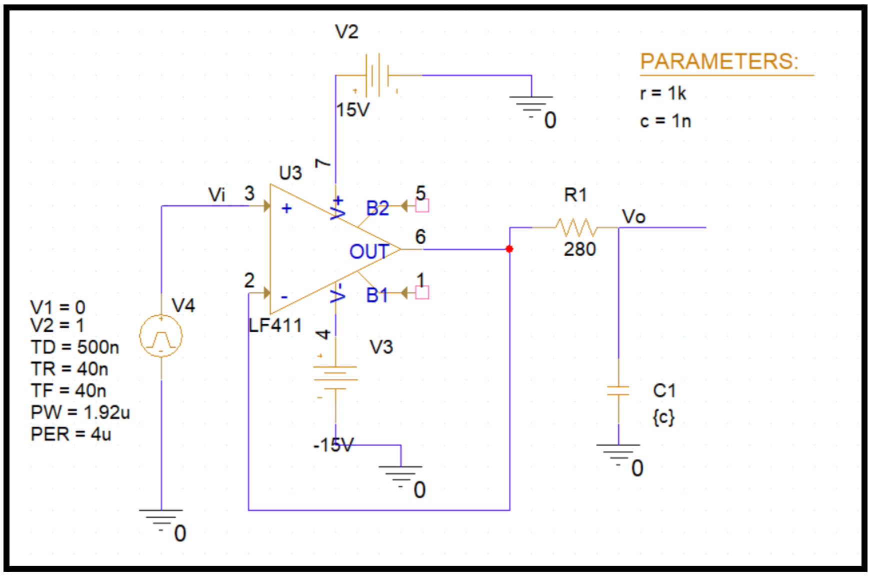


Figure 2.2.5: Schematic of Unity-Gain Follower with $R_C = 280\Omega$

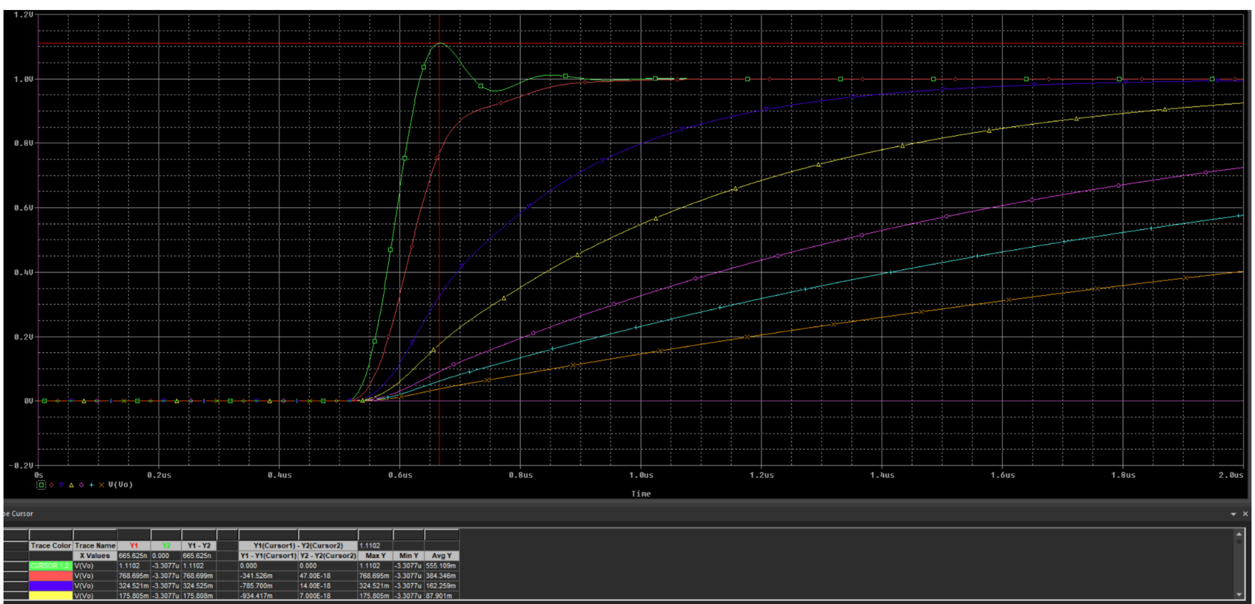


Figure 2.2.6: Plot of Unity-Gain Follower with $R_C = 280\Omega$ for all C

This compensation resistor value of 280 Ω gave us an overshoot of 11.02% greater than 11% with a capacitance of 100pF. Therefore, we have to increase resistance to lower overshoot %.

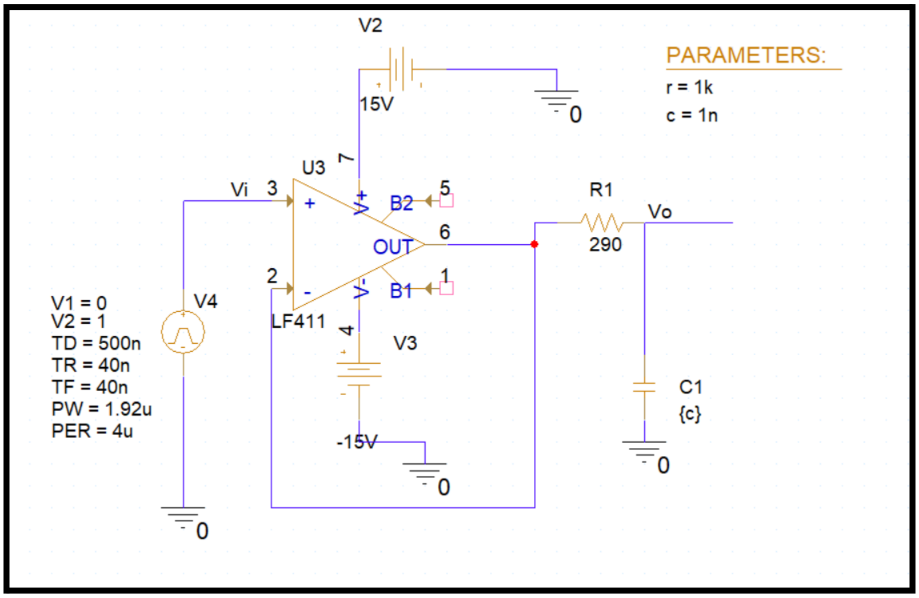


Figure 2.2.7: Schematic of Unity-Gain Follower with $R_C = 290\Omega$

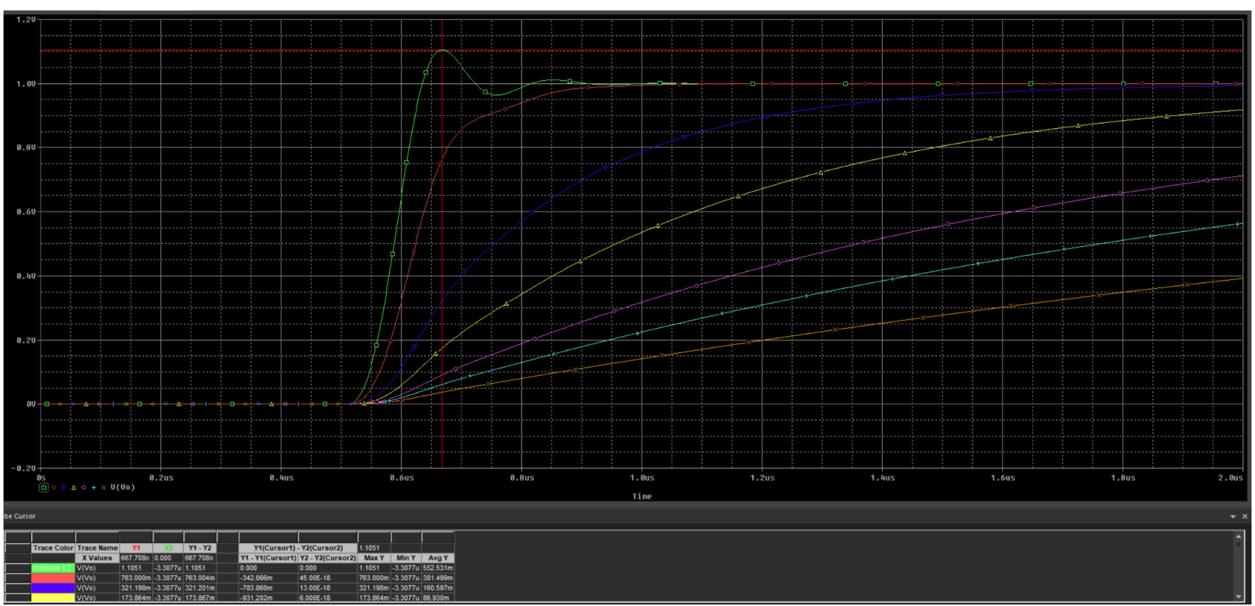


Figure 2.2.8: Plot of Unity-Gain Follower with $R_C = 290\Omega$ for all C

This gave us a max overshoot <11% for all loads.

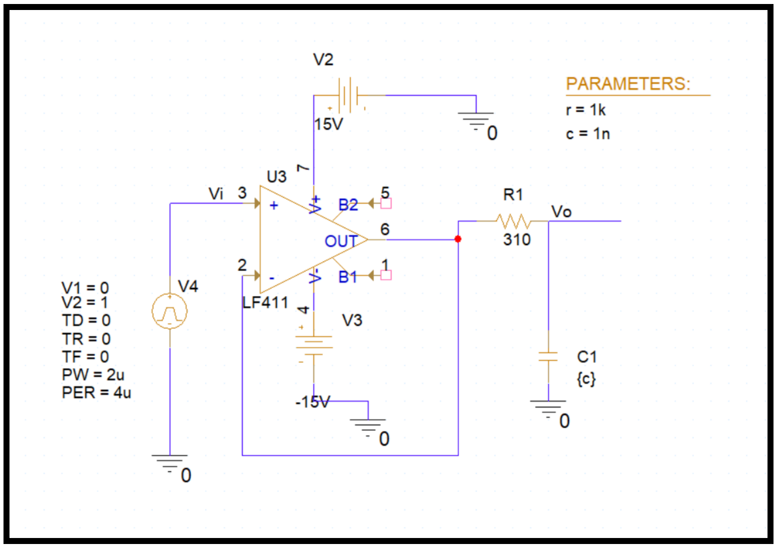


Figure 2.2.9: Schematic of Unity-Gain Follower with $R_C = 310\Omega$

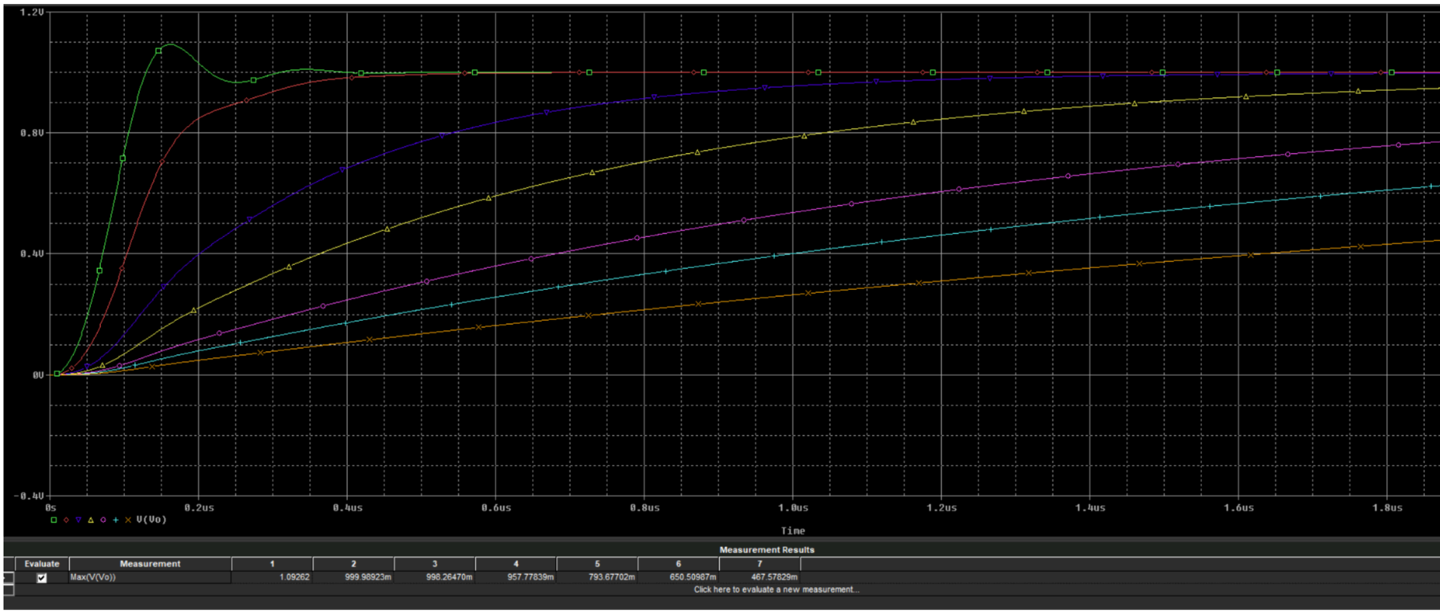


Figure 2.2.10: Plot of Unity-Gain Follower with $R_C = 310\Omega$ for all C

Overshoot : 9% for Green, None for Red, None for Purple, None for Yellow, None for Pink , None for Blue ,None for Orange.

We decided to change the rise time for the pulse voltage input so we can see it instantly.We see Overshoot% is less than 11 for all loads. However, we can get the rise time to be faster because our max overshoot is only 9 % and we can make some changes. We decided to reduce the resistance because this would reduce the rise time which would be faster.

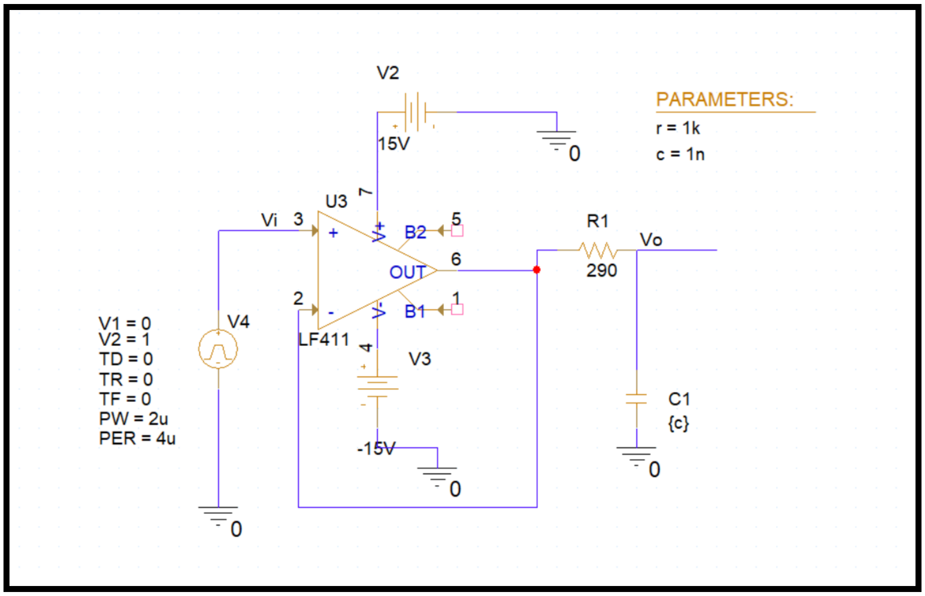


Figure 2.2.11 Schematic of Unity-Gain Follower with R_C =290Ω

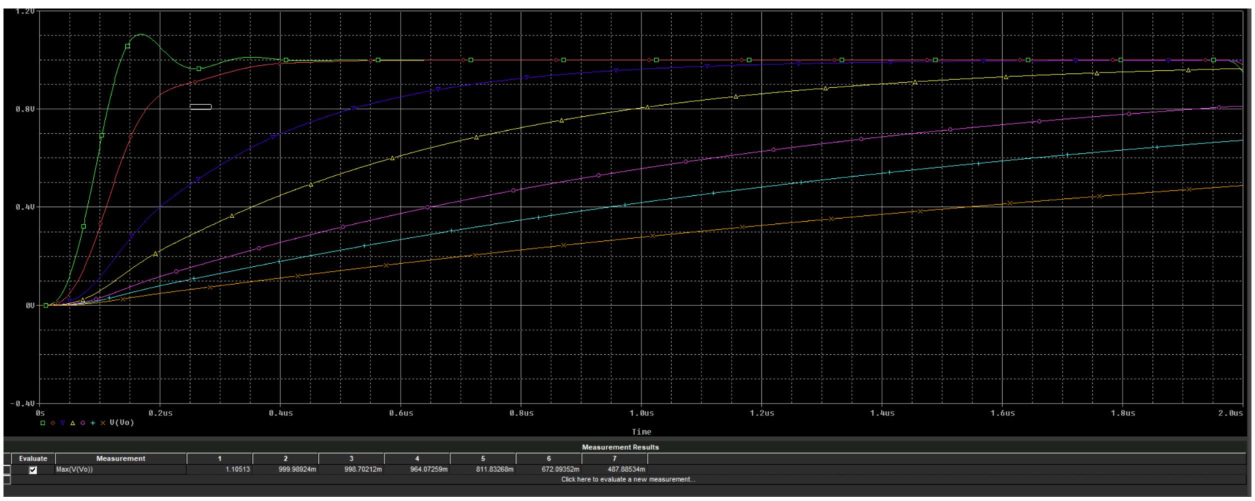


Figure 2.2.12: Plot of Unity-Gain Follower with R_C = 290Ω for all C

Resistor Value Ω	Max Overshoot of all loads
25 Ω	29.152 %
280 Ω	11.02 %
290 Ω	10.5 %

C)

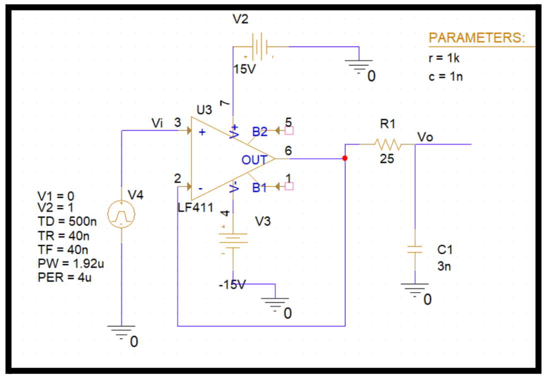


Figure 2.2.13 : Schematic of Unity-Gain Follower with R_C = 25Ω and load C = 3nF

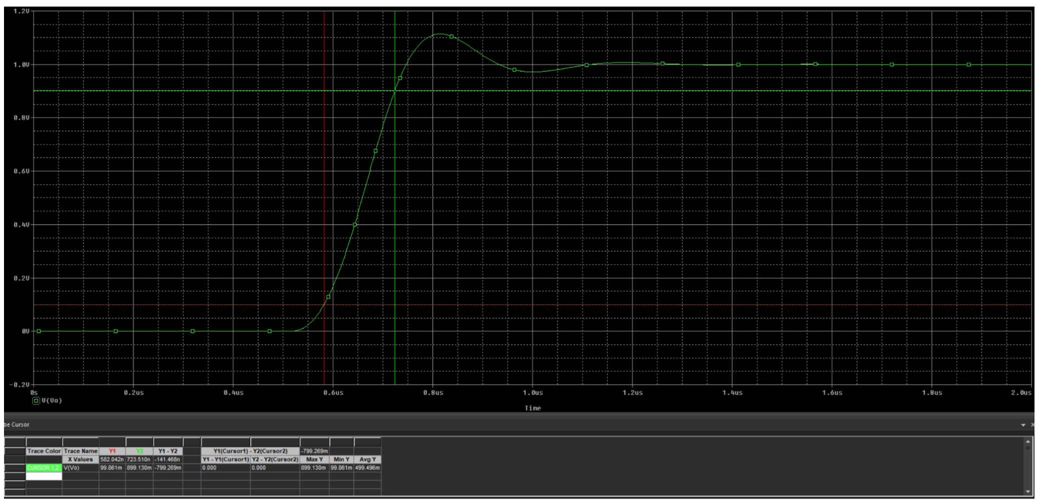


Figure 2.2.14: Plot of Unity-Gain Follower with R_C = 25Ω for load C = 3nF

Rise time: 141.468ns , Overshoot: 11.4 %

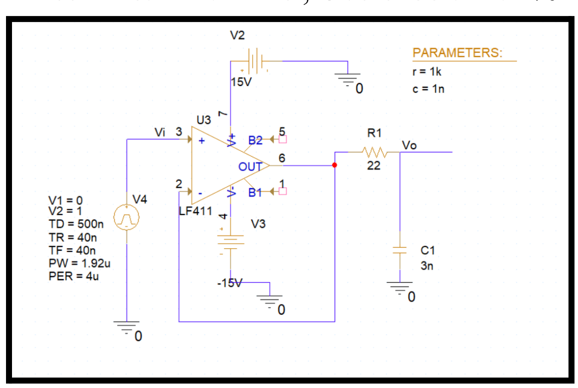


Figure 2.2.15: Schematic of Unity-Gain Follower with R_C =22Ω and load C = 3nF

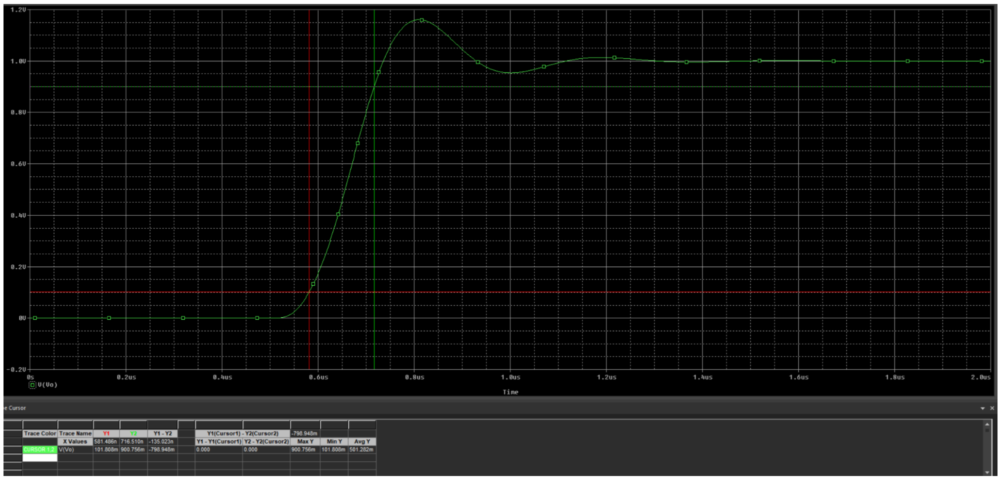


Figure 2.2.16: Plot of Unity-Gain Follower with $R_C = 22\Omega$ for load C = 3nF

Overshoot:16.02 % (not what we want because too much overshoot) , Rise time: 135.023ns

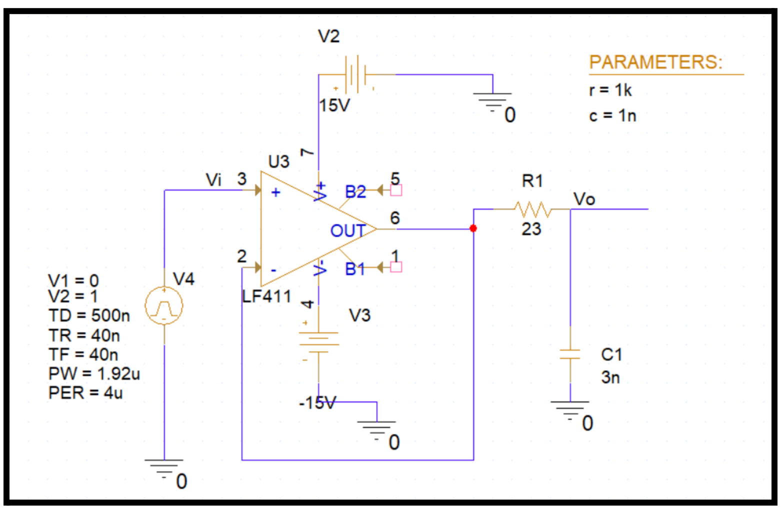


Figure 2.2.17: Schematic of Unity-Gain Follower with $R_C = 23\Omega$ and load C = 3nF

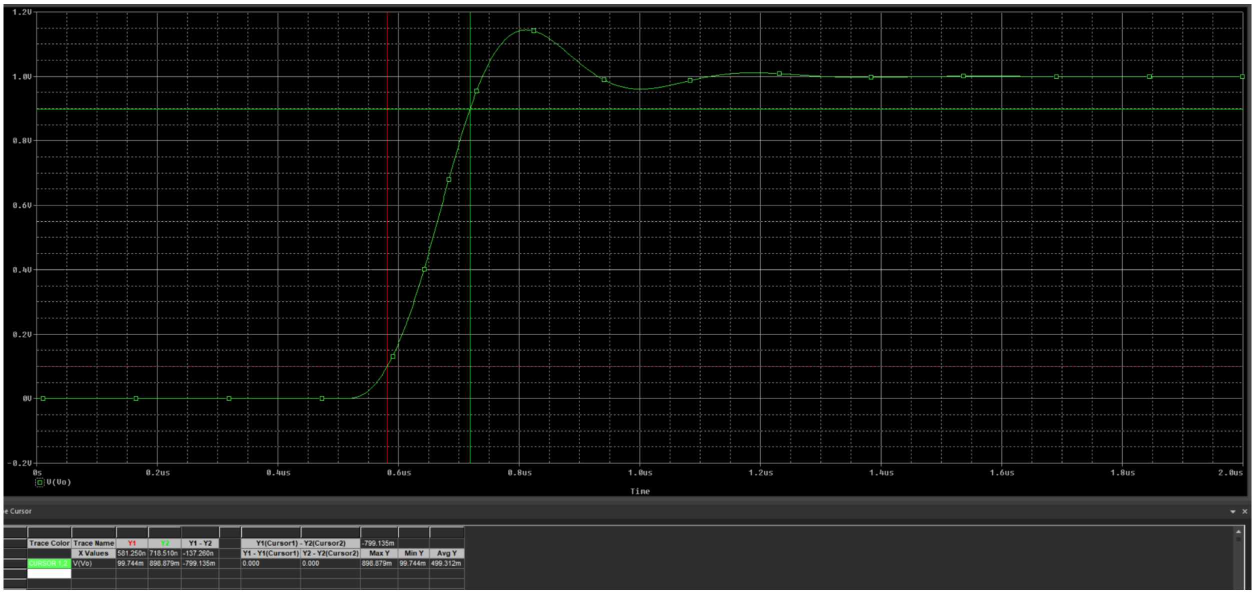


Figure 2.2.18: Plot of Unity-Gain Follower with R_C = 23Ω)for load C = 3nF

Rise Time: 137.260ns
Overshoot max = 14.3 %
This is what we want because it gives us the fastest rise time and overshoot is under 15%.

Resistor Value Ω	Overshoot %	Rise Time
25 Ω	11.4%	141.468 ns
22 Ω	16.02%	135.023 ns
23 Ω	14.3%	137.260 ns

The data shows that as resistor values decrease, overshoot increases and rise time tends to decrease. The time constant (τ=RC) gives us a representation of how the circuit responds to change in voltage. A smaller τ would mean a quicker response and larger overshoot.

Section 3: Experiment

Part 1) Un-loaded

Parameter	Gain	Desired Gain	**(Measured)** Frequency at A_D	**(Measured)** Gain Bandwidth Product	**(Simulated)** Frequency at A_D	$\left(\frac{V_{out}}{V_{in}}\right)$	Measured Gain	R1	R2
Units	A (dB)	$A_D = (A - 3dB)$	ω_p	$a_0\omega_p$	ω_p	$\left(\frac{mV_{out}}{mV_{in}}\right)$	dB	Ω	Ω
Data	**A = 1** (0dB)	-3 dB	7.950 MHz	7.950 MHz	8.5114 MHz	$\left(\frac{31.2}{44}\right)$	-2.99 dB	Shorted	Open
	A = 3 (9.542 dB)	+6.4951 dB	1.7 MHz	5.1 MHz	3.7830 MHz	$\left(\frac{100}{46}\right)$	6.745 dB	5k	10k
	A = 10 (20 dB)	+17 dB	365 kHz	3.65 MHz	947.860 KHz	$\left(\frac{328}{46}\right)$	17.062 dB	1.1k	10k
	A = 30 (29.538 dB)	+26.526 dB	115 kHz	3.450 MHz	282.594 KHz	$\left(\frac{800}{37.6}\right)$	26.558 dB	345	10k
	A = 100 (40 dB)	+36.990 dB	30 kHz	3.0 MHz	80.719 KHz	$\left(\frac{880}{12.4}\right)$	37.021 dB	100	10k

Table 3.1.1: Measured vs. Simulated Gain Results

As gain increases, we see frequency at which the desired gain is measured/simulated decreases. β is the feedback factor which also controls the DC gain at low frequencies. Our measured and simulated ω_p was off about 40% for each gain. This shows how components can affect the behavior of circuits.

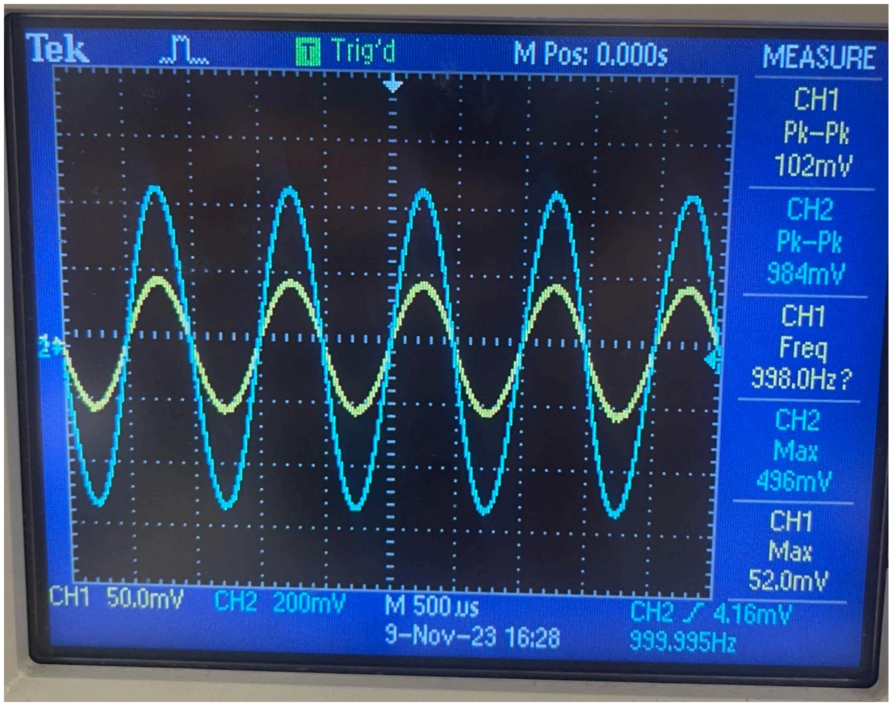


Figure 3.1.2: Oscilloscope Reading - DC gain of 10 at 1k Hz

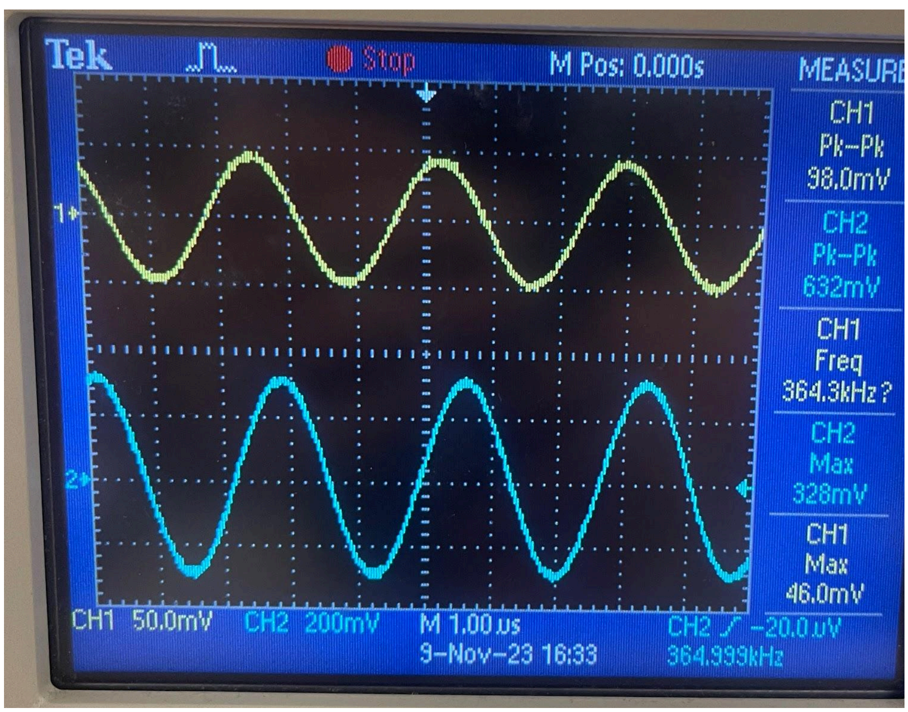


Figure 3.1.3: Oscilloscope Reading - DC Gain of 10 at 17 dB

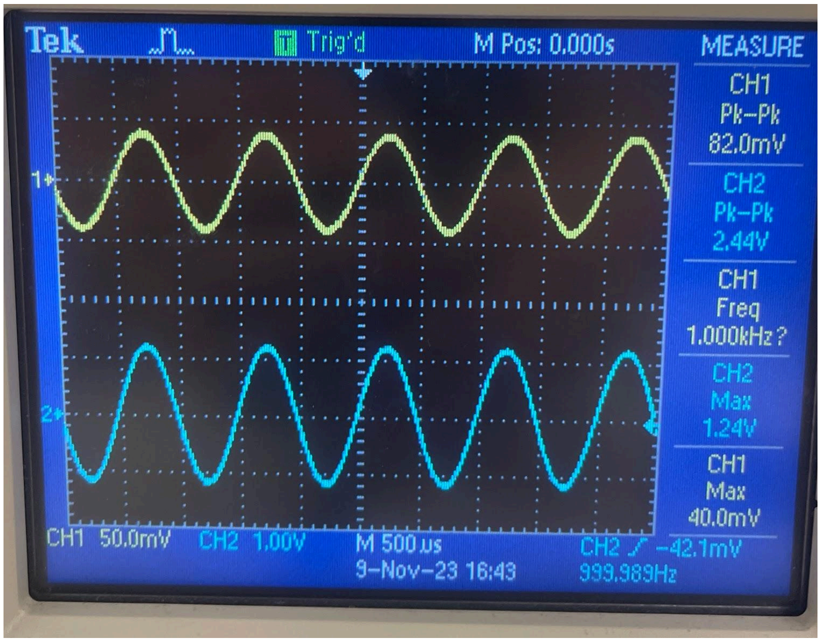


Figure 3.1.4: Oscilloscope Reading - DC gain of 30

Part 2: Capacitor Load

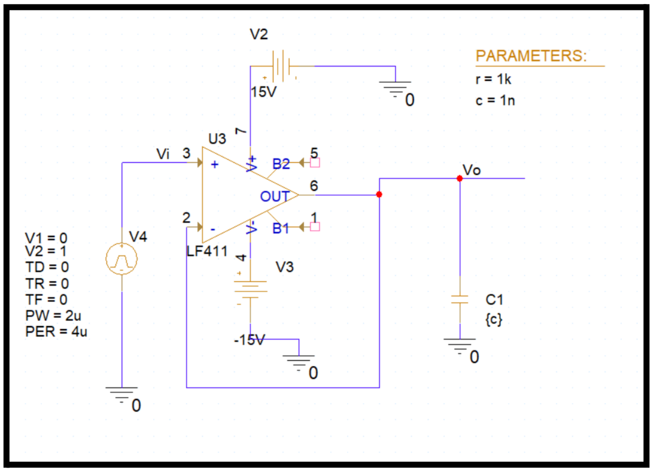


Figure 3.2.1: Schematic of Voltage Follower with Capacitive Load

(a)

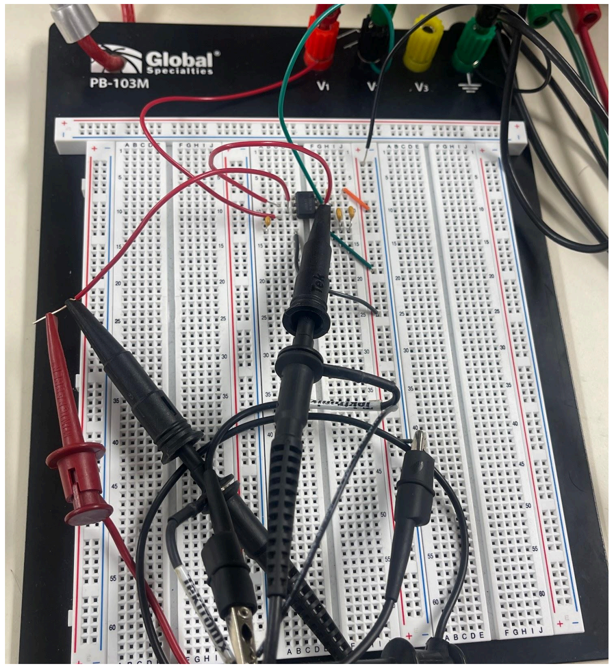


Figure 3.2.2: Loaded Voltage-Follower Circuit

Capacitor Value	Delta output	Delta Input	Overshoot %
100 pF	1.22 V	.464 V	32.6 %
300 pF	1.4 V	.464 V	52.14 %
1 nF	.76 V	.464 V	63.79 %
2 nF	.800 V	.464 V	72.414 %
4.2 nF	.848 V	.464 V	82.76 %
6 nF	.880 V	.464 V	89.66 %
10 nF	.864 V	.464 V	86.207 %

Figure 3.2.2: Table of Overshoot Values with Varying Capacitors

We see a similar trend; Capacitance goes up, Overshoot goes up and hits peak at 6nF and then slightly decreases at 10nF. We see that with higher capacitance, it causes higher overshoots in the response to a step input. There is a slight decrease of shoot from 6nF to 10 nF and that might be because of the circuit's ability to dampen oscillations. Another thing that might affect this is interaction of resistance and inductive elements.

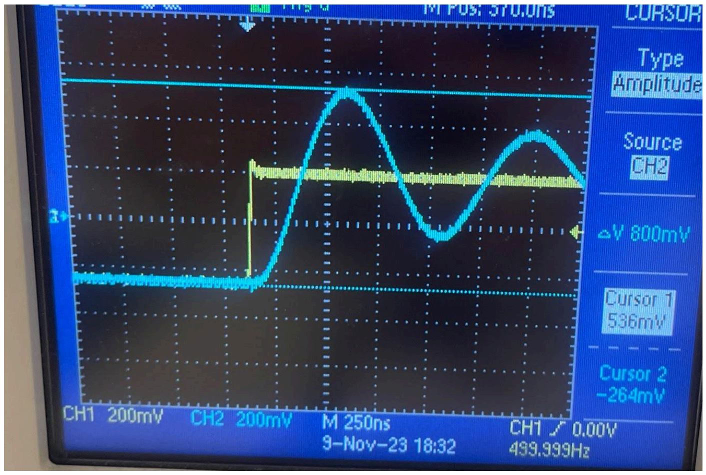


Figure 3.2.3: Oscilloscope Plot - V_o and V_i $vs.$ $Time$ (C = 2nF)

V_o $(Blue)$ and $V_i(Yellow)$

Rise time: 200 ns

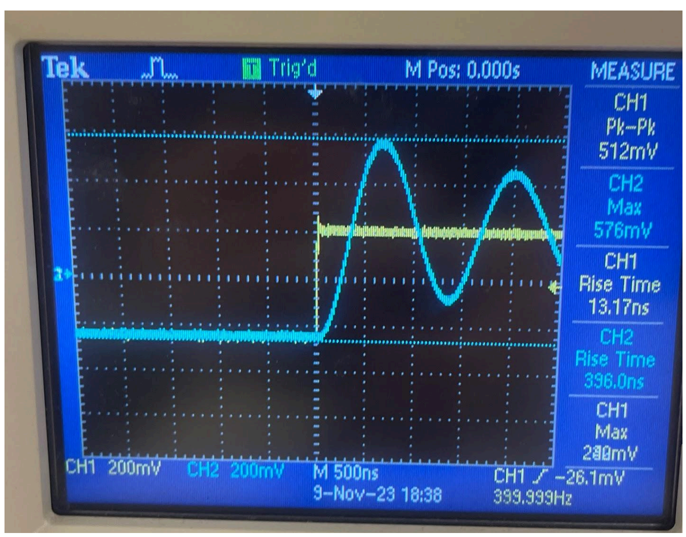


Figure 3.2.4: Oscilloscope Plot - V_o and V_i $vs.$ $Time$ (C = 4nF)

V_o $(Blue)$ and $V_i(Yellow)$

Rise Time: 396 ns

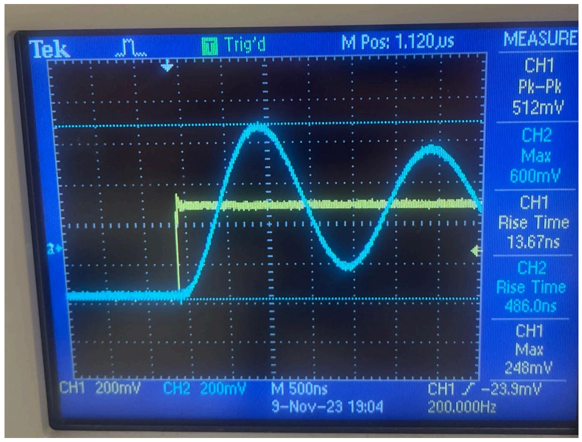


Figure 3.2.5: Oscilloscope Plot - V_o and V_i vs. Time (C = 10nF)

V_o (Blue) and V_i (Yellow)

Rise Time: 486 ns

(b)

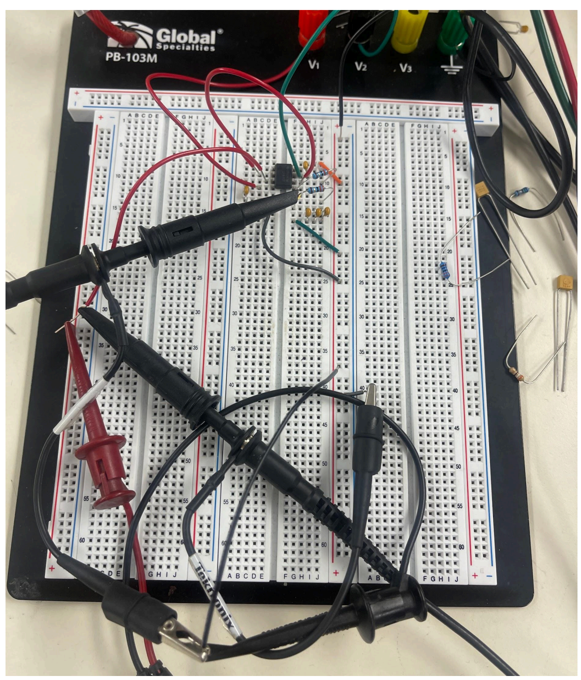


Figure 3.2.6: Circuit after capacitor replacement (3nF)

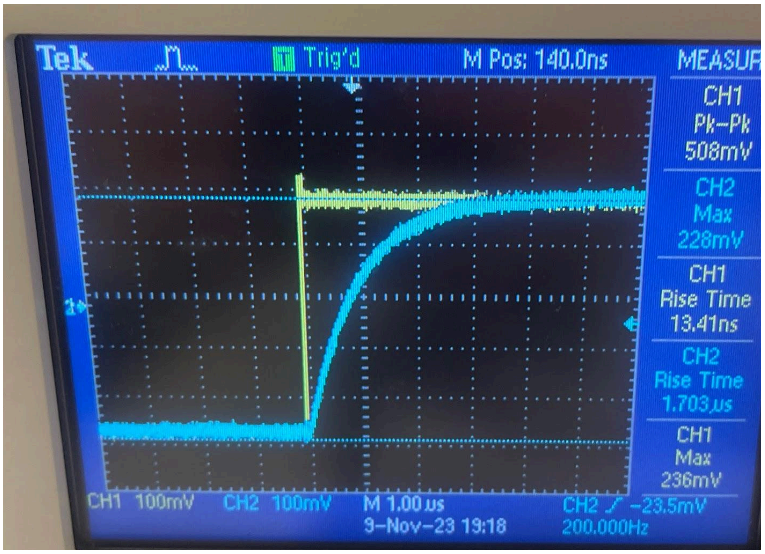


Figure 3.2.7: Oscilloscope Plot - V_o and V_i vs. Time (C = 10nF, $R_C = 330\Omega$)

V_o (Blue) and V_i(Yellow)

No overshoot ,Rise Time 1.7 us

Figure 3.2.8: Oscilloscope Plot - V_o and V_i vs. Time (C = 10nF, $R_C = 75\Omega$)

V_o (Blue) and V_i(Yellow)

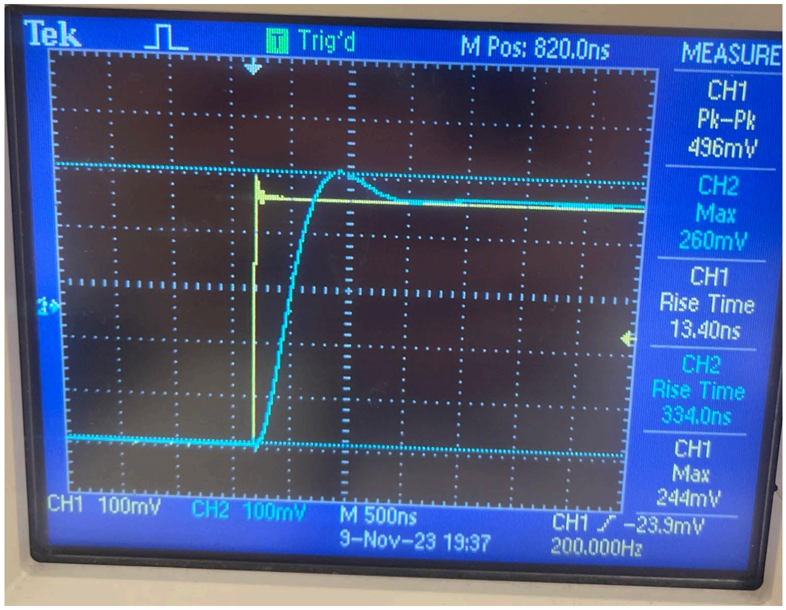


Figure 3.2.9: Oscilloscope Plot - V_o and V_i $vs.$ $Time$ (C = 10nF, $R_C = 68\Omega$)

$V_o\,(Blue)\ and\ V_i(Yellow)$

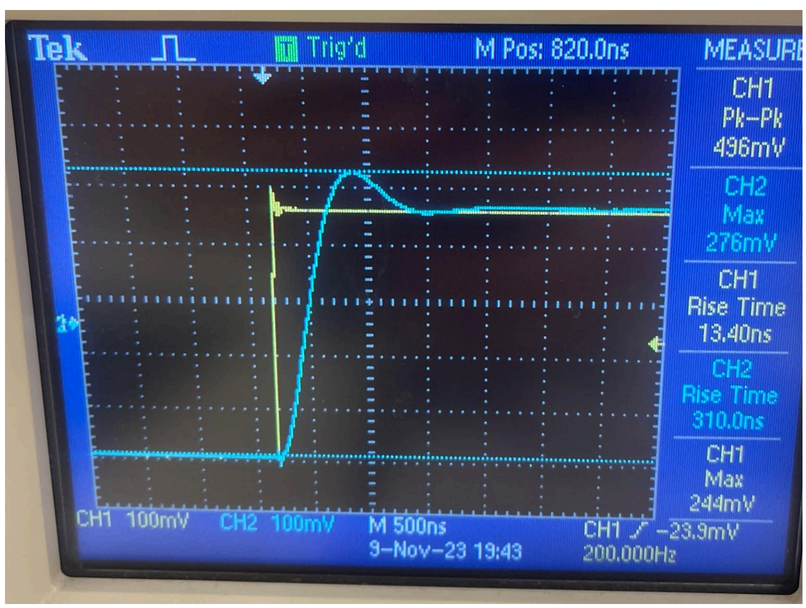


Figure 3.2.10: Oscilloscope Plot - V_o and V_i $vs.$ $Time$ (C = 10nF, $R_C = 60\Omega$)

$V_o\,(Blue)\ and\ V_i(Yellow)$

Resistor Value Ω	ΔOutput	ΔInput	Overshoot %	Rise Time
27 Ω	.64 V	.452 V	41%	232 ns
39 Ω	.572 V	.452 V	26.55%	254 ns
75 Ω	.488 V	.452 V	7.965%	356 ns
68 Ω	.496 V	.452 V	9.735%	334 ns
60 Ω	.516 V	.452 V	14%	310 ns
66 Ω	.500V	.452 V	10.6%	324 ns

Figure 3.2.11: Table of Overshoot Values with Varying Resistors

66 Ω met the required conditions →Max overshoot is under 11% and the most optimal rise time of 324 ns.

Conclusion

In our study of a basic operational amplifier (op-amp) voltage amplifier, we initially focused on formulating a general transfer function and determining the input/output impedances. We then confirmed the unity gain bandwidth and DC gain through the closed-loop gain analysis. Our observations revealed capacitive elements in the input impedance and inductive elements in the output impedance. We developed new transfer functions for scenarios involving a capacitive load and a capacitive load with resistive compensation. We noted that capacitive loads introduced oscillations in the step response, which were effectively dampened by incorporating a compensatory resistor. We found that the natural frequency and damping coefficient ξ affects the output because the adjustment of resistor and capacitance will have a big affect on ξ while having minimal impact on the bandwidth.

For practical testing, we used a simple voltage controlled voltage source (ideal op-amp) and a Texas Instruments LF411C operational amplifier. We experimented with gains of 1, 3, 10, and 100. The unity gain bandwidths were approximately 1MHz for the VCVS and 8MHz for the LF411C. Further testing on the LF411C, particularly in unity gain mode with varying capacitive loads (ranging from 100pF to 10nF), displayed a trend of more pronounced ringing in the output as the capacitance increased. Adding . The overshoot % generally decreases as the resistor value increases, reaching a minimum at 23 ohms. The rise time increases with higher resistor values, suggesting that a larger resistance slows down the response of the circuit. Optimizing the compensatory resistor for a 3nF load, we found that a 66 Ω resistor was ideal, limiting overshoot to 11% and achieving a rise time of approximately 324 nanoseconds. In circuits with resistors and capacitors, the RC time constant is a key factor in determining how quickly a circuit responds to a change in voltage. A higher RC constant($\boldsymbol{\tau}$) means slowing charging and discharging of the capacitor, which affects the overshoot.

In the final phase of our experiment, we constructed LF411-based voltage amplifiers with gains of 1, 3, 10, 30, and 100. We observed that the -3dB points were around 7.95 MHz, 1.7 MHz, 365 kHz, 115 kHz, and 30 kHz respectively for these gain levels. It was also noted that higher gains necessitated smaller input signals to prevent distortion due to slew rate or current saturation limitations. Our observed and calculated values of the cutoff frequency ω_p showed a discrepancy of 40% across various gain values. This shows how components can affect the behavior of the circuit.