FUN2 Final Project Report

Braden Desman, Garrett Delaney, and Nick Flora

***Abstract*—In this report, we explain the design, simulation, and verification of an audio processing circuit that flashes lights along to the treble and bass sounds in music. We designed our circuit to work well with “Feel So Close” by Calvin Harris as an input, using its frequency content to guide our design decisions. In order to design our circuit, we decomposed our system into five stages: a summing amplifier, highpass filter, lowpass filter, peak detector, and LED driver. The design, simulation, and verification of each stage will be explained in detail in this report. Ultimately, we succeeded in designing and implementing a circuit that performs this audio processing task well and produces a satisfactory visual response when our chosen song is used as the input. Along the way, we learned about using Operational Amplifiers in several different configurations, modeling components with the small-signal model, and how to apply concepts from signals and systems theory to circuits.**

I. Background Information and Rationale

The goal of this project is to design a signal processing circuit that takes a song as input and flashes lights in sync with the music. For our project, we decided to use “Feel So Close” by Calvin Harris as the song to design our circuit around. With this song, our goal was for our circuit to have one LED flash along with the treble content of the song, or the high frequencies, in the song and another LED flash along to the bass content of the song, or low frequencies. Visually, this corresponds to one LED lighting up when there are loud bass noises and the other lighting up for loud treble noises.

Our approach to this design problem was to separate the input signal into two parts, one containing the high frequencies (treble) and another containing the low frequencies (bass). Our circuit detects when the treble signal is loud and then turns on an LED. Likewise, it will detect when a bass signal is loud and turn on a different LED to show this visually. In the body of this report, we will go into detail about the specific implementation of these high level functions.

II. Design and Analysis

We chose "Feel So Close" by Calvin Harris because it has a good range of highs and lows and an EDM-type beat. For the low pass filter we determined a cutoff value of 300 Hz and a high pass cut off value of 400 Hz. We chose these values based on how the frequency spectrum in Audacity looks. We also used the built-in Audacity tools for applying a low pass and high pass filter in software. When we applied the high pass filter, the bass was largely inaudible. And likewise when we applied the low pass filter, the treble became minimized. We also wanted the high pass cut off frequency to be greater than the low pass cut off frequency so that a loud note near the corner frequency does not light up both LEDs at once.

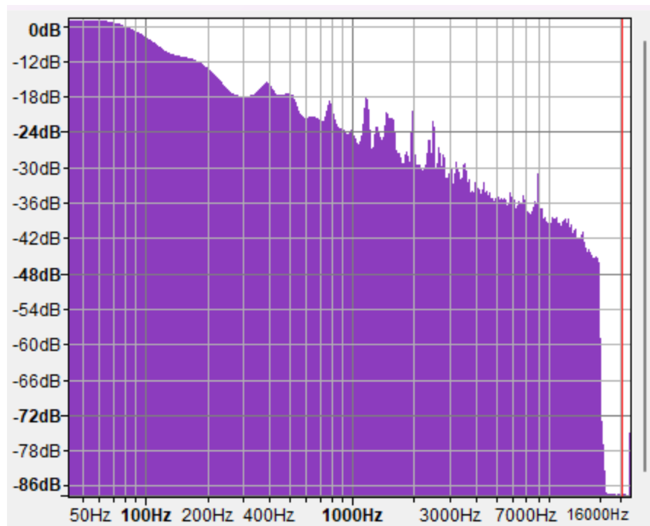

Fig. 1. Frequency spectrum of “Feel So Close” by Calvin Harris (entire song)

Before designing and implementing our signal processing circuit, we first decomposed the problem into five discrete chunks. Each chunk has its own stage in our circuit and will be described in detail in the following subsections. For now, we will provide an overview of the specification for each stage.

First, the left and right channel of the song input will go through our summing amplifier. We decided to make this the first stage so that stereo signals are no longer broken into their left and right channels. Because there may be some arbitrary DC offset in the input, this summing amplifier stage also contains a highpass filter before the left and right channels are added together. Because we only want to remove DC offsets with this filter, we decided to specify a very low corner frequency of 20 Hz so that the rest of the low frequency content of the song passes through with little attenuation. Once the DC offset is removed, we want this stage to add the left and right channels and simultaneously amplify them. Given the typical range of output magnitudes for a 3.5mm headphone jack and the supply rails of -4.5 to 4.5 V, we decided to specify a linear gain of -3 for summing amplifier. We chose this gain so that we could achieve good amplification of the input signal with little to no railing. We will describe the design of this stage

This is the final project report for ECE 2660 Fundamentals 2, Spring 2023 at the University of Virginia

in Section II-A including the architecture of the summer and how we chose specific component values to realize the specification described here.

The second stage of our circuit is a filtering stage to separate the input signal into its high frequency and low frequency content. After this stage, the high frequency and low frequency signals will be processed in parallel so that the LEDs can respond to one signal or the other. As mentioned before, we chose 300 Hz as the target lowpass cutoff frequency and 400 Hz as the target highpass frequency based on the frequency content of "Feel So Close". In Section II-B and Section II-C, we will explain how we implemented the Sallen-Key architecture and selected component values to realize the cutoff frequencies we described here. Because the signal was already amplified in the previous stage, we have specified the gain of these active filters to be 0 dB.

After the high frequency and low frequency channels pass through their respective filters, we elected to pass the signals through a peak detector. The peak detector gives the signals a 1.6V DC offset. This makes the inputted signals have a voltage very close to the MOSFET saturation region as the MOSFETs have a V_t of around 2 V, so the MOSFETs are easier to turn on. However, the offset is set low enough so that when no signal is inputted (for example, when the song isn't playing or the respective filter doesn't output a signal based on the frequency), then the LED is turned off. The output of the peak detector is then used as the gate voltage of an NMOS transistor which ultimately regulates the current through each LED, along with the resistor at the source of the MOSFET. The peak detector gives the signal a τ of 30Hz so that the LED is on long enough for the human eye to visually see it but not on too long so the LED doesn't overlap with other signals. The purpose of the peak detector is to ensure that the light flashes with the song but does not follow the peaks of the input signal exactly (i.e. flash too quickly to be perceived or vice versa). The design of the peak detector is explained in Section II-D, while the LED driver design is explained in Section II-E.

Before diving into the design of each stage, we would also like to make some comments about the power circuitry included in this system. While the circuit is powered by a 9 V battery, we wanted to include a stable ground reference voltage at the halfway point of the battery's supply range. This means that there is a rail splitter to supply a constant 0 V (ground) with V_{EE} at -4.5 V and V_{CC} at 4.5 V [1].

A. Summing Amplifier

The summing amplifier is the first stage of the project board. The design is slightly more involved than the standard op-amp summing amp. It is comprised of 2 main sub components: a high pass filter (HPF) and a summing amp. There is a T-shaped resistor feedback network to allow for separate designs the HPF cutoff frequency and the passband gain of summing amp. The design requirements were to select input capacitors C_3 and C_{14} and resistors R_{18} and R_{19} to create a corner frequency near $20Hz$ and resistor values that are at least $10k\Omega$. Humans cannot hear below $20Hz$ and this also removes any DC offset that the outputting audio device such as a laptop may have. We were also tasked with designing values for R_1, R_2, and R_5 such that there is a linear gain of 3 in the passband.

The first step is selecting components to create a cutoff frequency near $20Hz$. Let

$$C = C_3 = C_{14} \tag{1}$$

$$H(s) = \frac{V_{out}}{V_{in}} = \frac{-R_f}{Z_{in}} = \frac{-R_f}{R_{in} + \frac{1}{sC}} \tag{2}$$

$$H(s) = -R_f C \frac{s}{1 + sR_{in}C} = -R_f C \frac{s}{1 + \frac{s}{\omega_c}} \tag{3}$$

Where:

$$\omega_c = \frac{1}{R_{in}C} \tag{4}$$

$$f_c = \frac{\omega_c}{2\pi} = \frac{1}{2\pi R_{in} C} \tag{5}$$

Let us first arbitrarily choose a capacitor value since there are much less options in our kit compared to resistors.

$$C_3 = C_{14} = 0.1\mu F \tag{6}$$

Solve for R_{in}

$$R = \frac{1}{2\pi f C} = \frac{1}{2\pi(20Hz)(0.1\mu F)} = 79.6k\Omega \tag{7}$$

Choose the closest resistor in the lab kit:

$$R_{in} = R_{18} = R_{19} = 82k\Omega \tag{8}$$

Check the corner frequency is close enough:

$$f_c = \frac{1}{2\pi(82k\Omega)(0.1\mu F)} = 19.41Hz \approx 20Hz \tag{9}$$

The next step is to find the the transfer function of the circuit in the pass band. In the pass band (high frequencies), the capacitors C_3 and C_{14} behave like short circuits and thus can be ignored.

Let I_L be the left channel current through R_{18}, and I_R be the right channel current through R_{19}. Since the negative input terminal of the op amp is at virtual ground, by Ohms law:

$$I_L = \frac{V_L}{R_{in}} \tag{10}$$

$$I_R = \frac{V_R}{R_{in}} \tag{11}$$

Let:

$$V_{in} = V_L + V_R \tag{12}$$

Let I_2 be the current through R_2. By KCL, since the op-amp input terminals draw no current:

$$I_2 = I_L + I_R = \frac{V_L}{R_{in}} + \frac{V_R}{R_{in}} = \frac{V_{in}}{R_{in}} \tag{13}$$

By Ohm's law:

$$V_T = -I_2 R_2 = -V_{in}\frac{R_2}{R_{in}} \tag{14}$$

Let I_1 is the current through R_1 and I_5 be the current through R_5. By Ohm's Law they equal:

$$I_1 = \frac{V_T}{R_1} \tag{15}$$

$$I_5 = \frac{V_T - V_{out}}{R_5} \tag{16}$$

By KCL:

$$I_2 = I_1 + I_5 \tag{17}$$

Substituting in expressions for I_2, I_1, and I_5.

$$\frac{V_{in}}{R_{in}} = \frac{V_T}{R_1} + \frac{V_T - V_{out}}{R_5} \tag{18}$$

Following some algebra and plugging back in the expression for V_T (shown in Appendix B) this leads to a transfer function:

$$H(s) = \frac{V_{out}}{V_{in}} = -\frac{R_f}{R_{in}} \tag{19}$$

Where:

$$R_f = R_2 + R_5 + \frac{R_2 R_5}{R_1} \tag{20}$$

The next step was to solve for R_1, R_2, and R_5 such that there is a linear gain of 3 in the passband.

The gain magnitude is:

$$g = \frac{R_f}{R_{in}} \tag{21}$$

Suppose:

$$R = R_{in} = R_1 = R_2 = R_5 \tag{22}$$

We noticed that setting all the resistors equal to each other would conveniently lead to a gain of 3 in the passband. This seemed like the simplest design so we went with it.

$$g = \frac{R + R + \frac{R^2}{R}}{R} = 3 \tag{23}$$

Therefore:

$$R_1 = R_2 = R_5 = 82k\Omega \tag{24}$$

B. High Pass Filter (HPF)

We implemented our HPF using the Sallen-Key architecture depicted in Fig. 3. As mentioned before, we selected the high pass cutoff frequency to be 400 Hz (2513 rad/s) based on the frequency content of our song. We chose to make our filter critically damped so that the rolloff in the frequency domain would be 40 db/decade beyond the breakpoint frequency.

The equations for Q, the quality factor, and ω_0, the cutoff frequency, for a Sallen-Key filter are provided in Equation (26) and Equation (27) for a filter with a linear gain of 1.

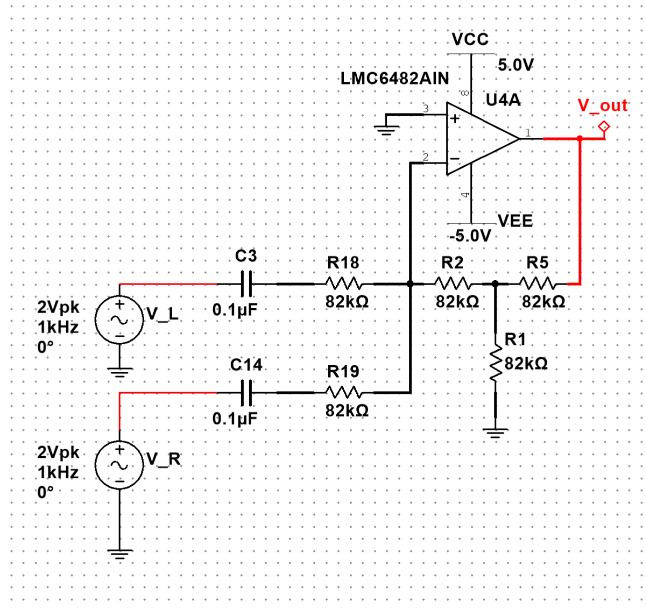

Fig. 2. Summing Amplifier Schematic

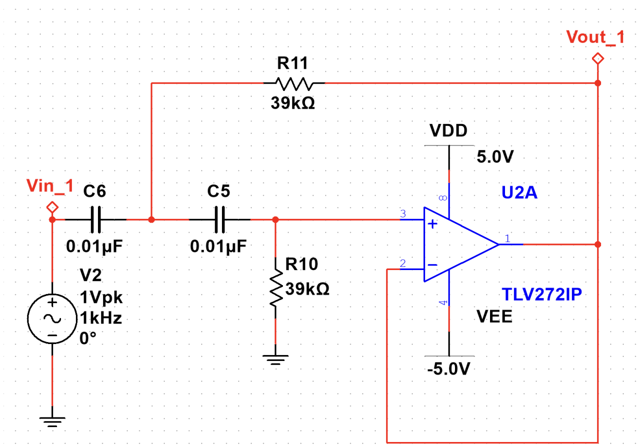

Fig. 3. Sallen-Key HPF Architecture Schematic

$$H(s) = \frac{s^2}{s^2 + s\frac{\omega_0}{Q} + \omega_0^2} \tag{25}$$

$$Q = \frac{\sqrt{R_{11}R_{10}C_5C_6}}{R_{11}C_6 + R_{11}C_5} \tag{26}$$

$$\omega_0 = \frac{1}{\sqrt{R_{10}R_{11}C_5C_6}} \tag{27}$$

We used a Python script to compare different set of values using these equations and the target Q and ω_0. Using this script we were able to select values for R_{10}, R_{11}, C_6, C_5. The code for this script is provided in Appendix A. With this code, we ended up with the following component values:

- $R_{10} = R_{11} = 39$ kΩ
- $C_5 = C_6 = 0.01$ μF.

While there were several sets of component values that were within the specified tolerance range, we chose to go with this specific combination because it yielded a cutoff frequency slightly greater than the target value (as opposed to the other solutions, which yielded cutoff frequencies below the target). Because the cutoff frequencies for the LPF and HPF are already fairly close, we didn't want to bring them any closer by choosing resistor and capacitor values that produced a corner frequency below the target value. Using the components listed above, we find that the cutoff frequency for our high pass filter is 408 Hz = 2564 rad/s. The linear gain of this active filter in the passband is 1, as intended.

C. Low Pass Filter (LPF)

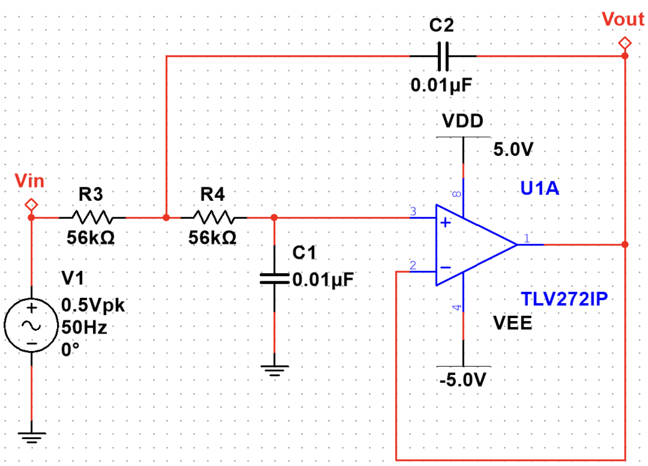

Fig. 4. Sallen-Key LPF Architecture Schematic

$$H(s) = \frac{\omega_0^2}{s^2 + s\frac{\omega_0}{Q} + \omega_0} \tag{28}$$

$$Q = \frac{\sqrt{R_3R_4C_1C_2}}{R_3C_1 + R_4C_1} \tag{29}$$

$$\omega_0 = \frac{1}{\sqrt{R_3R_4C_1C_2}} \tag{30}$$

We implemented our LPF using the Sallen-Key architecture depicted in Fig. 4. As mentioned before, we selected the low pass cutoff frequency to be 300 Hz (1885 rad/s) based on the frequency content of our song. We chose to make our filter critically damped so that the rolloff in the frequency domain would be 40 db/decade beyond the breakpoint frequency.

Since the quality factor Q, and cutoff frequency ω_0 for a lowpass Sallen-Key filter are governed by Equation (29) and Equation (30), we were again able to use a Python script in Appendix A to select values for R_3, R_4, C_1, C_2. Ultimately, we ended up with the following values:

- $R_3 = R_4 = 56$ kΩ
- $C_1 = C_2 = 0.01$ μF.

While there were several sets of component values that were within the specified tolerance range, we chose to go with this specific combination because it yielded a cutoff frequency slightly less than the target value (as opposed to the other solutions, which yielded cutoff frequencies above the target). Because the cutoff frequencies for the LPF and HPF are already fairly close, we didn't want to bring them any closer by choosing resistor and capacitor values that produced a corner frequency above the target value. Using the components listed above, we find that the cutoff frequency for our high pass filter is 284 Hz = 1786 rad/s.

D. Peak Detector

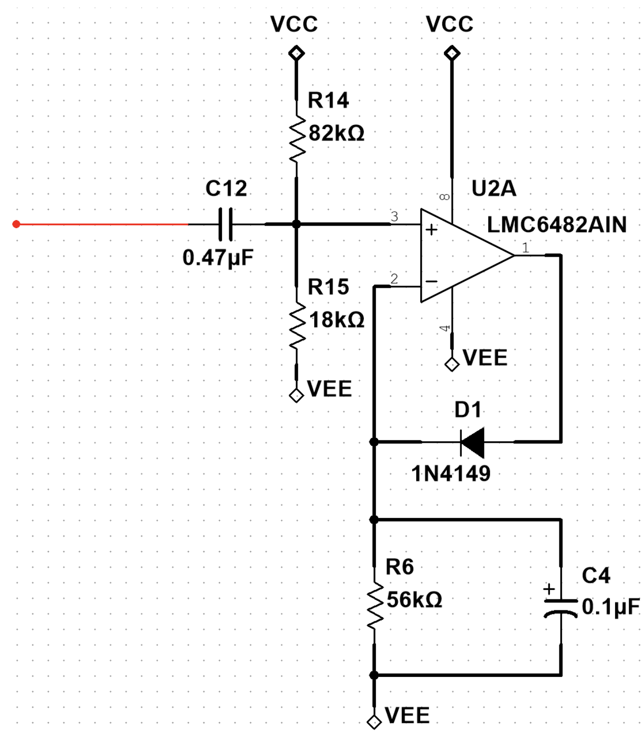

Fig. 5. Peak Detector Schematic

The peak detector circuit is provided in Fig. 5. Recall that the output of the peak detector is fed into the MOSFET. When there is no signal in the frequency range of interest, we want the MOSFET to be in cutoff. This means that the output of the peak detector must be less than or equal to the threshold voltage of the MOSFET when there is no input signal. Thus, in order to design the voltage divider immediately preceding the peak detector, we first needed to measure the MOSFET threshold voltage.

We obtained the threshold voltage of both of our MOSFETs by measuring the current in the saturation region for two different gate voltages. These measurements correspond to (V_{gs1}, I_{ds1}) and (V_{gs2}, I_{ds2}). We plugged these values into Equation (31) (from the lecture notes) to calculate our values for the threshold voltage.

$$V_t = \frac{-V_{gs2}\sqrt{\frac{I_{ds1}}{I_{ds2}}} + V_{gs1}}{1 - \sqrt{\frac{I_{ds1}}{I_{ds2}}}} \tag{31}$$

After collecting our experimental data, we calculated that both MOSFETs conveniently had the same threshold voltage of 2.06 V.

To ensure that the output of the peak detector was below this threshold voltage when there is no input signal, we determined that the voltage at the input of the Op Amp should be 2 V corresponding to the voltage divider formula in Equation (32).

$$2 = \frac{R_{17}}{R_{17}+R_{16}}9 \tag{32}$$

We were able to select the resistor values by inspection. Since both of our MOSFETs had the same threshold voltage, the resistor values for each peak detector are symmetrical: $R_{15} = R_{17} = 18$ kΩ and $R_{16} = R_{14} = 82$ kΩ. This corresponds to a DC offset at the noninverting terminal of our peak detector of 1.62 V, well below the threshold voltage of our MOSFETs, as desired.

Based on the frequencies at which humans can detect changes in light intensity (30-60 Hz) and sound volume (5 Hz), we decided to make our time constant 30 Hz. Equation (33) was used to calculate the value for R_6, and we chose $C_4 = 0.1$ μF since it was easily available in our kit. Again, we hypothesized that the component values should work for both the high pass channel and low pass channel, so we let $R_{12} = R_6 = 56$ kΩ and $C_7 = C_6 = 0.1$ μF.

$$R_6 = \frac{1}{2\pi f C} = \frac{1}{2\pi * (30Hz)(0.1\mu F)} \approx 56k\Omega \tag{33}$$

$$\tau = RC = (56k\Omega)(0.1\mu F) = 5.6ms \tag{34}$$

E. LED Driver

When designing our LED driver, we had two main concerns: ensuring that the current did not exceed the maximum rating for each LED and ensuring that the current was large enough so that the LEDs turned on brightly when the input signal is large.

We obtained the maximum current ratings for the LEDs from the data sheets. The maximum current is 30 mA and 7 mA for the green and red LEDs, respectively [2][3].

We measured the maximum voltage output by the computer for our song to be 3 V. Using Equation (35) from the lecture notes with $I_{ds} = 3$ mA and the K_n measurement from our MOSFET, we can calculate the resistance of the source resistor. $V_g = 3$ V, the maximum voltage output by the computer, and V_t =2.06 V (as measured previously). We selected $I_{ds} = 3$ mA since it was well beneath the max current rating for both LEDs.

$$I_{ds} = \frac{1}{2K_n R_s^2}\left(\sqrt{1 + 2K_n R_s (V_g - V_t)} - 1\right)^2 \tag{35}$$

Solving Equation (35) for R_s, we find $R_s = R_7 = 270$ Ω. Experimentally, we observed that this resistance value resulted a satisfactory visual response in the red LED. We also used this resistor value as a starting place for the green LED since it can handle higher current. However, experimentally we noticed that this resistor value only lit the green LED dimly, so we iteratively checked lower resistance values for the green LED until we found one that produced a satisfactory visual response. We ultimately selected $R_{13} = 150$ Ω for the high frequency channel (green LED) and $R_7 = 270$ Ω for the low frequency channel (red LED).

III. Simulations

A. Summing Amplifier

In Section II-A, we selected component values to ensure that our summing amplifier circuit has a linear gain of 3 (which corresponds to 9.54 dB) in the passband with a RC highpass filter preceding it. The breakpoint frequency of the highpass filter was designed to be 20 Hz. In Fig. 6, we show the simulated frequency response of our summing amplifier circuit from an AC Sweep in Multisim. This test was conducted on both input channels and produced identical results, as anticipated. Notice that in the passband (i.e., for frequencies higher than 20 Hz), the gain is 9.54 dB. This exactly matches the value it was designed to be. Since the RC highpass filter is a first order filter, we expect the gain at the breakpoint frequency to be 3 dB less than the gain in the passband. This is also confirmed in the figure as we see that $\left|H(f = 20\ Hz)\right| = 6.66$ dB.

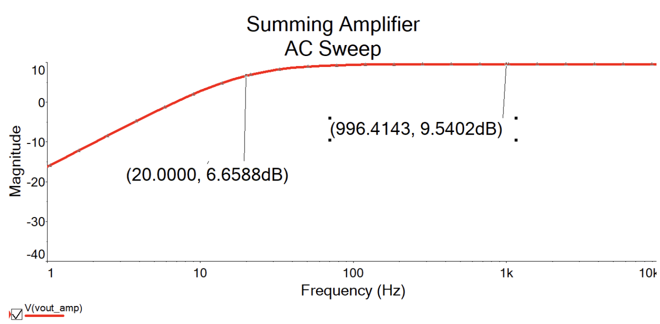

Fig. 6. Summing Amplifier Simulation Results

We also numerically verified our summing amplifier's functionality in the time domain. For this simulation, a 500 mV sinusoidal voltage source was input to both the left and the right channels. Given that our summing amplifier is inverting and is designed to have a linear gain of 3, we expect the output of the summing amplifier at the peak of the input signals to be $-(500mV + 500mV) * 3 = -3V$. Fig. 7 provides the results of our transient simulation of the summing amplifier circuit. Indeed, we see that the output of the amplifier when the input is two 500 mV sinusoids is -3 V at the peak of the input, as expected.

B. High Pass Filter

In Section II-B, we determined that with the component values we selected, our HPF cutoff frequency should be 408 Hz. Since the Sallen-Key HPF is a second order filter, this

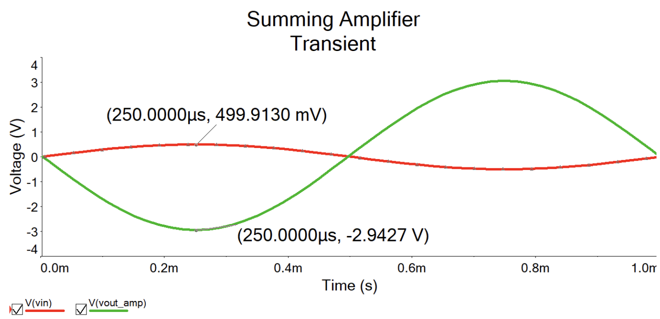

Fig. 7. Summing Amplifier Transient Simulation Results

corresponds to a drop of 6 dB from the gain in the passband at 408 Hz. Since the gain in the passband is 0 dB, we would expect to see that $|H(f = 408\ Hz)| = -6$ dB. We can see from the Multisim AC Sweep plot shown in Fig. 8 that this is exactly the behavior of our design. Moreover, we can clearly see that the passband - which corresponds to frequencies greater than 408 Hz - has a gain of approximately 0 dB as expected.

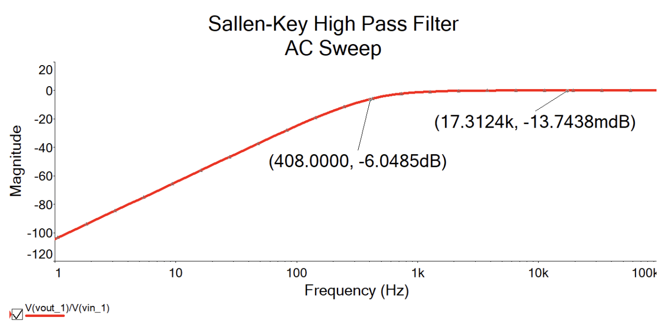

Fig. 8. Sallen-Key HPF Simulation Results

We also numerically verified the output of our filters in the time domain. First, we tested an input signal of 3 kHz, well above the cutoff frequency of 408 Hz. Given that this input is in the passband, we expect the output to be a phase shifted but not attenuated version of the input. In Fig. 9, we see that this is exactly what our simulation results show.

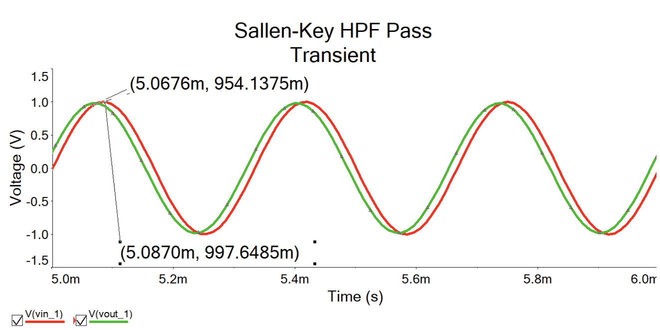

Fig. 9. HPF transient results for a signal in the passband

We also tested our design in the time domain for a 20 Hz signal, well below the corner frequency of 408 Hz. For this 20 Hz input, we performed a transient simulation and observed that the output was indeed attenuated as we would expect. The results of this simulation are provided in Fig. 10.

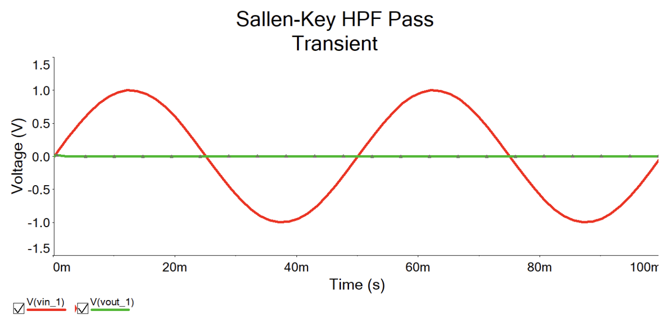

Fig. 10. HPF transient results for a signal below the passband

C. Low Pass Filter

In Section II-C, we determined that with the component values we selected, our LPF cutoff frequency should be 284 Hz. Since the Sallen-Key LPF is a second order filter, this corresponds to a drop of 6 dB from the gain in the passband at 284 Hz. Since the gain in the passband is 0 dB, we would expect to see that $|H(f = 284\ Hz)| = -6$ dB. We can see from the Multisim AC Sweep plot shown in Fig. 11 that this is exactly the behavior of our design. Moreover, we can clearly see that the passband - which corresponds to frequencies less than 284 Hz - has a gain of approximately 0 dB as expected.

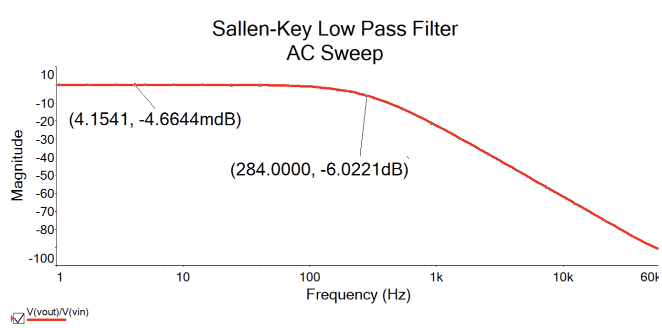

Fig. 11. Sallen-Key LPF Simulation Results

We also numerically verified the design of our LPF in the time domain. Fig. 12 shows the simulated transient output of our LPF when a 3 kHz sinusoidal signal is input. Since 3 kHz is well above our LPF corner frequency, we expect this signal to be attenuated strongly. Indeed, we can see that the green waveform which corresponds to the output of the filter is in fact strongly attenuated.

We also numerically verified the design of our LPF in the time domain for a signal in the passband. Fig. 13 shows the simulated transient output of our LPF when a 50 Hz sinusoidal signal is input. As this input is well below the corner frequency of 284 Hz, we expect this signal to be passed by the LPF. From Fig. 13, we can see that our LPF

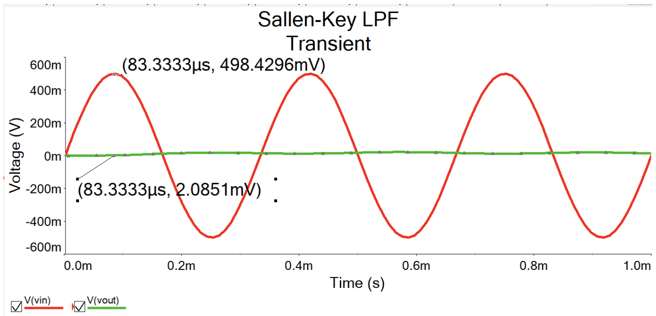

Fig. 12. LPF transient results for signal above the corner frequency.

does indeed pass this low frequency signal, albeit with a phase shift.

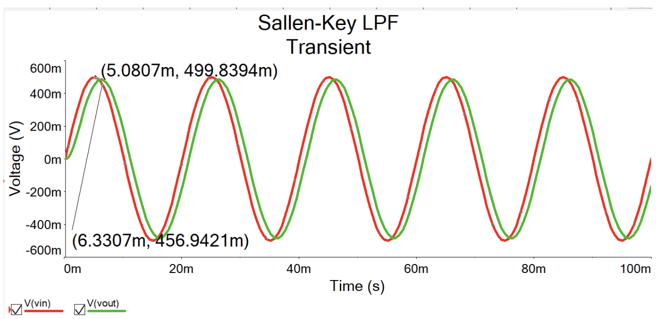

Fig. 13. LPF transient results for a signal in the passband

D. Peak Detector

To simulate the peak detector, we built it in Multisim, as shown in Fig. 14. That shows the peak detector for the red LED, but the design for the green LED was exactly the same. We inputted a $1V\ 1kHz$ sine wave as the input. The on-page connector $PDin$ shows the input to the op amp after passing through the big friendly capacitor (BFC) C_{12} and the added offset by the biasing resistors R_{14} and R_{15}. The transient results are shown in Fig. 15. The traces show $PDin - V_{EE}$ and $PDout - V_{EE}$ to reference them to V_{EE}, matching the experimental setup. The designed τ value for the peak detector was $5.6ms$. The cursors show that at time t_0 the voltage across the capacitor (green trace) is $2.5946V$, the peak as the capacitor is charging up. At time t_1, $500\mu s$ later, the voltage is at $2.3847V$. The equation for voltage across a discharging capacitor is $V_C(t) = V_0 e^{-t/\tau} = 2.5946Ve^{-500\mu s/5.6ms} = 2.3730V \approx 2.3847V$. Therefore the peak detector is working properly.

E. LED Driver

To simulate the LED driver, we added it to the peak detector simulation shown in Fig. 14. We also changed the input signal amplitude to 2V and left the frequency at 1 kHz. The only difference between red (low pass) LED driver and the green (high pass) LED driver designs was the source resistor. The red LED driver had a source resistor of

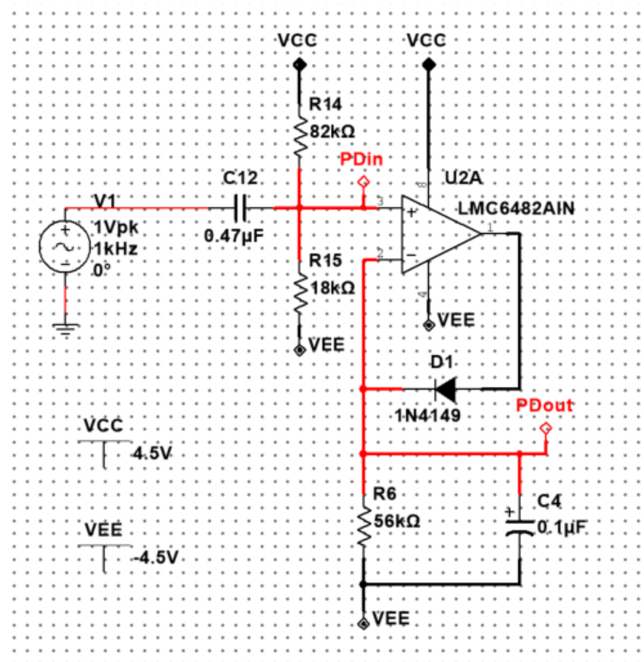

Fig. 14. Peak Detector Simulation Schematic

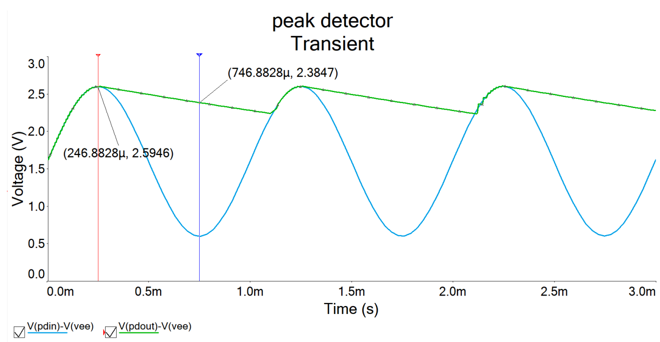

Fig. 15. Simulated Peak Detector Transient Results

$R_7 = 270\Omega$ and the green LED driver had a source resistor of $R_{13} = 150\Omega$. The maximum rating current for the red LED is 7 mA [3]. Notice in Fig. 16 that the current stays below 7 mA. The maximum rating current for the green LED is 30 mA [2]. Notice in Fig. 17 that due to a lower source resistor the green LED current goes higher, but is well below 30 mA.

F. System Simulations

Recall that the goal of our project is to have one of our LEDs turn on when there is substantial high frequency content and the other LED turn on when there is substantial low frequency content in the input. Although we are unable to use our chosen song as an input in Multisim, we were still able to verify the funcitonality of our entire system using two sinusoidal inputs of different frequencies.

Since we know that the brightness of an LED is proportional to the current moving through the LED, we used

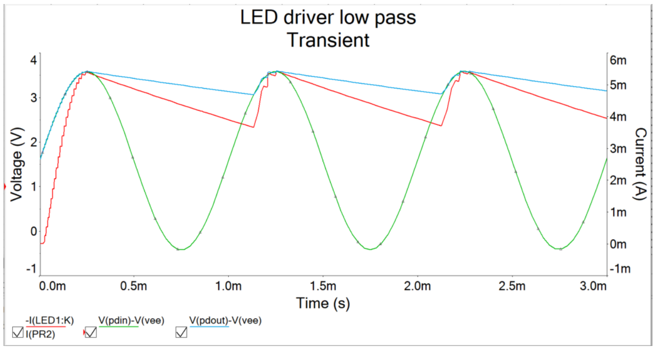

Fig. 16. Simulated Low Pass Red LED Driver Transient Results (current through LED shown in red trace)

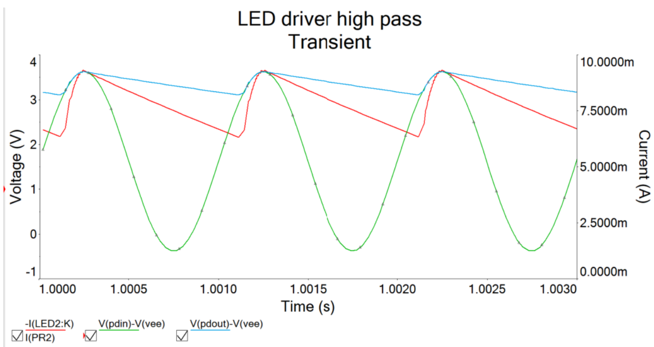

Fig. 17. Simulated High Pass Green LED Driver Transient Results (current through LED shown in red trace)

the current going through the resistor at the source of each MOSFET (which must be equal to the current going through each LED by Kichoff's current law) as a proxy for the brightness of the LED. In other words, we expect there to be lots of current (on the order of a few mA) going through the source resistor of the high-frequency channel when there are peaks in the high-frequency input signal. Likewise, we expect there to be lots of current going through the source resistor of the low-frequency channel when there are peaks in the low frequency input. These peaks in current correspond to the LEDs being on.

This behavior is exactly what we observe when we simulate the entire system in Multisim with a 1 kHz input signal on the left channel and a 100 Hz input signal on the right channel. Both signals had a magnitude of 500 mV. When there are (negative) peaks on the low frequency input, we see in Fig. 18 that there is substantial current through the resistor in series with the LED on the low-frequency channel (R_7). Likewise, we see substantial current going through the resistor in series with the LED on the high-frequency channel ($R13$) when there are (negative) peaks on the high-frequency input.

IV. Experimental Results

A. Summing Amplifier Experimental Results

These experimental results replicate the tests under Section III-A. To verify DC blocking and the passband gain, we

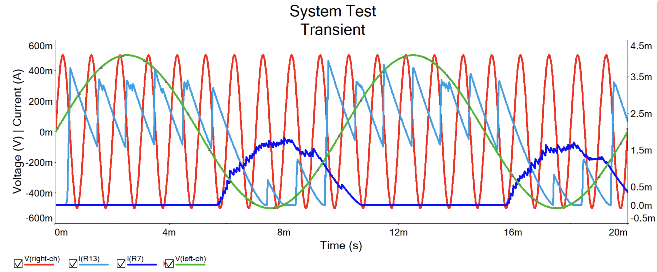

Fig. 18. System simulation results. The high frequency input is shown in red, and the current through the corresponding LED is shown in light blue. The low frequency input is shown in green, and the current through the corresponding LED is shown in dark blue. Note that current is shown on the right axis, in mA.

ran the WaveForms Network Analyzer (Bode plotter) for the AD2 on the left and right channels independently, with only one connected at a time. We ran the Bode plotter from 1 Hz to 100 kHz. Human hearing only ranges from 20 Hz to 20 kHz so I thought this would be an appropriate range for this song project. We left all the other inputs to the Network Analyzer at their default values such as the number of steps value at the 151.

Therefore the filter is supposed to have a gain of 6.53 dB at the corner frequency. At the corner frequency of 20Hz, the gain of left channel was observed to be 6.38 dB as shown in Fig. 19 and the gain of the right channel was 6.48 dB as shown in Fig. 20. Comparing the expected output to the observed output, we conclude the high pass filter is working properly with respect to the corner frequency. They are obviously not going to be exactly equal considering non-ideal components and resistors with tolerances.

The passband was designed to have a linear gain of 3, or $20log_{10}(3)dB = 9.54dB$. We arbitrarily looked at the gain at 1 kHz for the left and right channels to test the linear pass band gain. The left channel had a gain of 9.29dB as shown in Fig. 21 and the right channel had a gain of 9.27dB as shown in Fig. 22. Comparing the expected output to the observed output, we concluded the gain in the pass band is working properly. Again, they are not going to be exactly equal because of non-ideal aspects of the real world, but this is definitely close enough.

The testing conditions were a 4.5 V VCC and a -4.5 V VEE for the supplies, and a 0.3 V peak to peak 1 kHz sin wave signal input with 50 percent symmetry. The sin wave was inputted into Junction 12 to test the left channel, and Junction 15 to test the right channel for the time domain plots. The output was measured through Junction 4 for both domain plots.

To verify the summing feature, we used the superposition principle. We used the WaveForms Wavegen tool to input a 1 kHz 0.1V signal into the left channel only and the right channel only, one at a time. We chose 0.1V so that the op-amps would operate in their linear regions with a power supply of +/- 4.5V. We left the other values at their defaults such as offset of 0V, symmetry of 50%, and phase of 0°. We

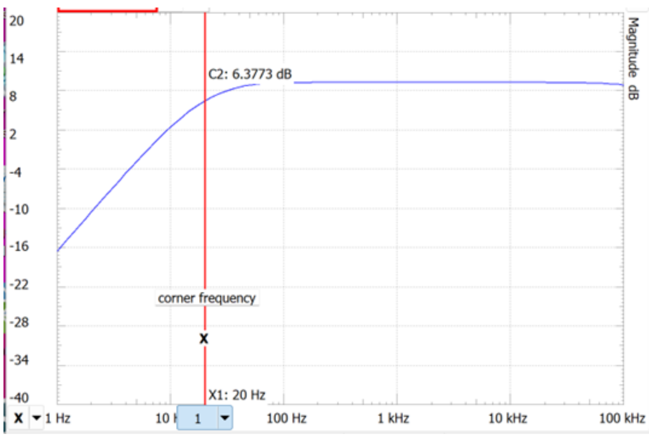

Fig. 19. Experimental Summing Amp Bode Plot for Left Channel Showing Corner Frequency

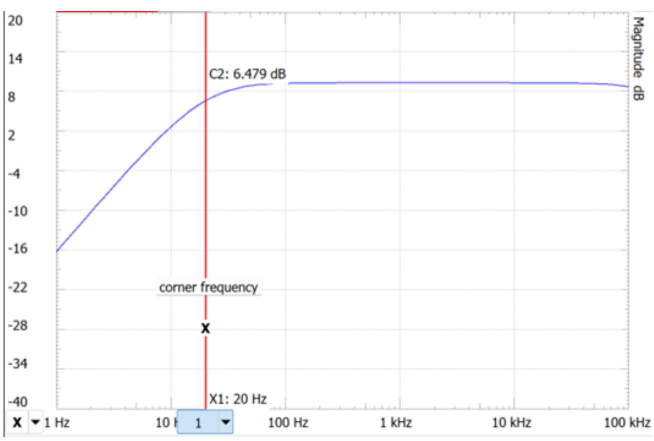

Fig. 20. Experimental Summing Amp Bode Plot for Right Channel Showing Corner Frequency

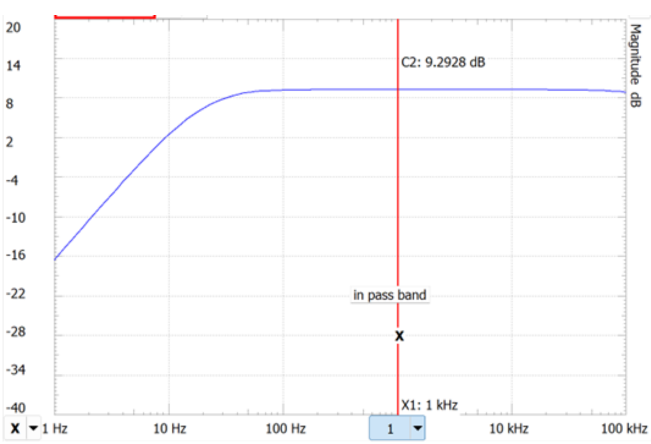

Fig. 21. Experimental Summing Amp Bode Plot for Left Channel Showing Passband Gain

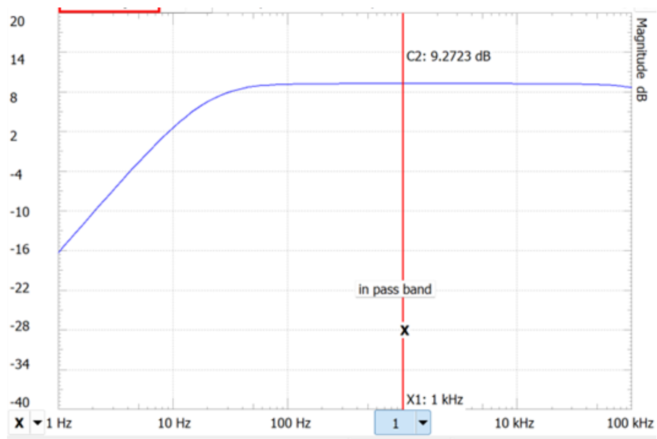

Fig. 22. Experimental Summing Amp Bode Plot for Right Channel Showing Passband Gain

then applied the same signal to both channels simultaneously and made sure to synchronized the input signals.

By the additivity principal of superposition, the system output of only the left channel added with the system output of only the right channel should equal the output if one added the left and right channel signals first and then ran the sum through the system.

To test this, we looked at the peak values of these sinusoidal waves. We observed the left and right channel time t=-1.28ms. The left channel had an input value of -0.103V and output of 0.300V as shown in Fig. 23 and the right channel had an input value of -0.102V and output of 0.295V as shown in Fig. 24. These outputs make sense considering the linear gain of 3 (or -3 since it's inverting nature) in the passband. We then observed the peak output of both channels at time t=-1.79ms. We probably should observed the same time, but it does not really matter because of the periodic nature of the signal. As shown in Fig. 25, the output of both channels was 0.595V, the sum of the two individual outputs, very close to the expected value of 0.6V. Thus, the additive nature works just as expected for an LTI system and the op-amp is operating in its linear region.

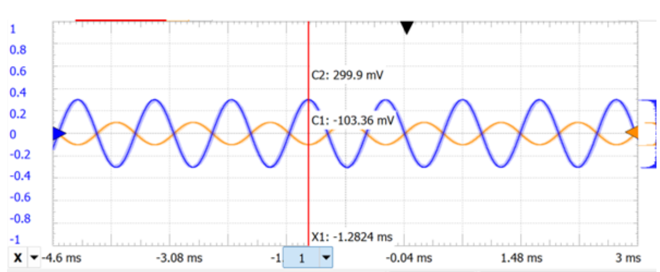

Fig. 23. Experimental Summing Amp Transient Plot for Left Channel

Fig. 26 shows the Summing amplifier with the song as the input, and the 9 V battery as the supplies with the output being measured through junction 4. The figure confirms that the summing amp adds both the left and right channels together and has a gain of -3. The orange line (Channel

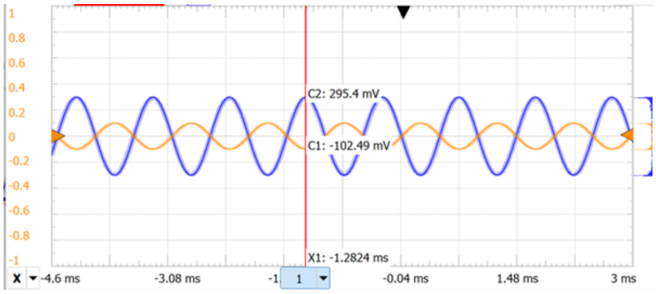

Fig. 24. Experimental Summing Amp Transient Plot for Right Channel

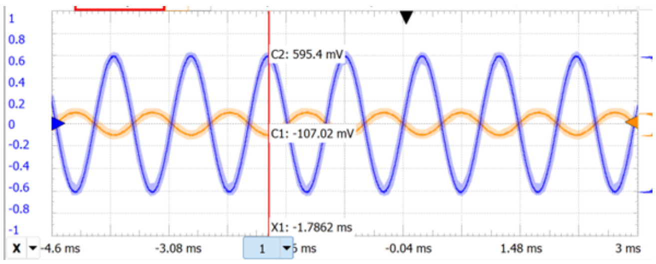

Fig. 25. Experimental Summing Amp Transient Plot for Both Channel

1) is the left channel of the input of the song, and the red line is the mathematical match up to the shape of the output (Channel 2). This figure verifies the summing amplifier for the song.

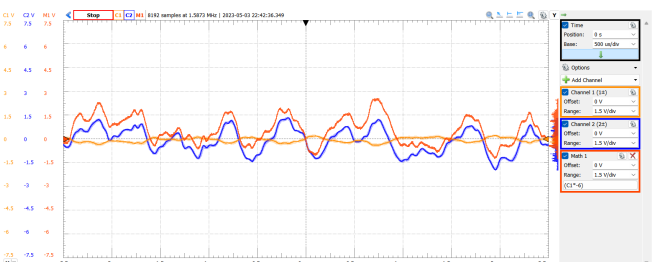

Fig. 26. Experimental result of the summing amplifier with the song as the input. Channel 1 is the input while the Red line is the mathematical manipulation of Channel 1 to match channel 2 to verify the summing amplifier.

B. *High Pass Filter Experimental Results*

In Section III-B, we numerically confirmed that our Sallen-Key HPF functions as we designed it to with a breakpoint frequency of 408 Hz. In this section, we will show that it performs well once built on the project board. In Fig. 27, we see that the experimental breakpoint frequency of our highpass filter is 442 Hz. We know that this is the actual breakpoint frequency because this is the frequency at which the gain is -6 dB. For the Sallen-Key filter, 6 dB below the passband gain (which is 0 dB in this case) is the location of the breakpoint frequency. While this cutoff frequency of 442 Hz does not exactly match the 408 Hz we designed it to be, this is certainly within the range of tolerance for the resistors and capacitors that determine the corner frequency. Additionally, from the perspective of the function we want our HPF to play, this slight increase in the breakpoint frequency from our ideal value should still allow the circuit to respond well to high frequencies in our chosen song. Indeed, looking at the frequency spectrum in Section II, there is still notable peaking in the frequency content of the song beyond 442 Hz, the experimental breakpoint frequency of our HPF. As a result, the deviation from our designed value should not have a negative impact on the functionality of our circuit. The testing conditions were a 4.5 V VCC and a -4.5 V VEE for the supplies, and the output was measured through Junction 7.

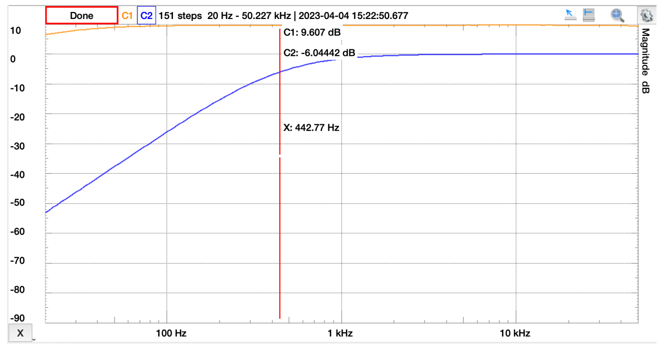

Fig. 27. Experimental Frequency Response of Sallen-Key HPF. The cursor indicates that the gain is -6 dB at 442 Hz.

Fig. 28 shows the high pass filter output with the song as the input and a 9 V battery as a supplies with the output being measured through Junction 7. This figure confirms the work ability of the HPF with the song as the input due to the higher peaks and shorter periods of the output than the LPF, verifying that higher frequencies are being filtered through.

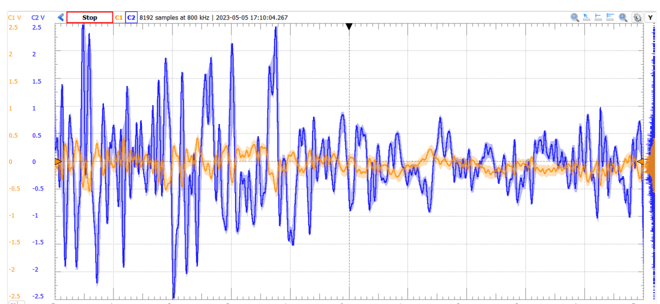

Fig. 28. Experimental Response of the HPF with the song as the input. Channel 1 is the input of the song and channel 2 is the output.

C. *Low Pass Filter Experimental Results*

In Section III-C, we numerically confirmed that our Sallen-Key LPF functions as we designed it to with a breakpoint frequency of 284 Hz. In this section, we will show that it performs well once built on the project board. In Fig. 29, we see that the experimental breakpoint frequency of our lowpass filter is 308 Hz. Again, we know that this is the cutoff frequency because it is the frequency at which the gain is 6 dB below the gain in the passband (the gain in the passband is 0 dB, and at $f = 308$ Hz the gain is approximately -6 dB. While this does cutoff frequency of 308 Hz does not exactly match the 284 Hz we designed it to be, this is certainly within the range of tolerance for the resistors and capacitors that determine the corner frequency. Additionally, considering that we initially specified a cutoff frequency of 300 Hz (and were unable to achieve it using the components available), this experimental value for the cutoff frequency is actually more than satisfactory as it is close to our ideal value. The testing conditions were a 4.5 V VCC and a -4.5 V VEE for the supplies, and the output was measured through Junction 6.

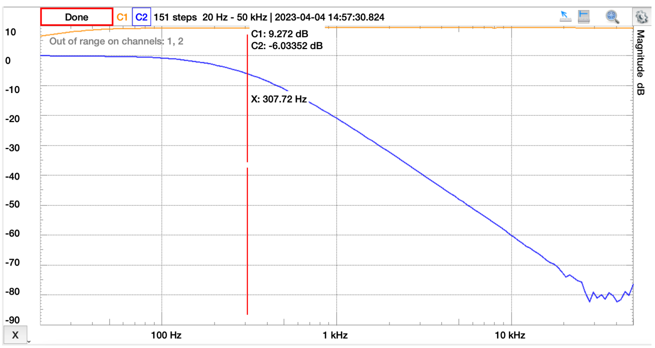

Fig. 29. Experimental Frequency Response of Sallen-Key LPF. The cursor indicates that the cutoff frequency occurs at 308 Hz (since the gain at this point is 6 dB less than the gain in the passband).

Fig. 30 shows the low pass filter output with the song as the input and a 9 V battery as a supplies with the output being measured through Junction 6. This figure confirms the work ability of the LPF with the song as the input due to the shorter peaks and longer periods of the output than the HPF, verifying that lower frequencies are being filtered through.

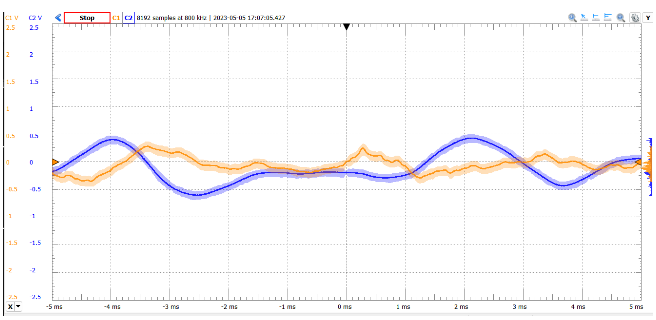

Fig. 30. Experimental Response of the LPF with the song as the input. Channel 1 is the input of the song and channel 2 is the output.

D. Peak Detector Experimental Results

The peak detector experimental results below show two things that verify the design of the peak detector on the board works. The peak detector was tested by putting an input into J6 from the AD2 Wavegen. The input was a sine wave with an amplitude of 1V and a frequency of 1KHz. The supplies for VCC and VEE were set at 4.5 V and -4.5 V, respectfully. The output (channel 2) was measured through J8. First, there was a DC offset of 1.6 V, that is show with the input (channel 1) being 1.6V lower than channel 2 (the output) in Fig. 31. This is also show when there is no sine input and the output is 1.6 V higher as seen in Fig. 32. This was important for the design as the peak detector makes the voltage at the gate of the MOSFET 1.6 V when the song is off (Battery input only). This makes the MOSFET in its cutoff region, as its V_t is around 2 V. However, having V_G at 1.6 V makes the MOSFET easier to move into saturation when the song is inputted and an additional voltage is added. The experimental results in Fig. 31 also show the τ being 5.6ms (around 30Hz). This is ideal as τ should be between 5 and 60 Hz. The cursors show that at time t_0 the voltage across the capacitor (blue trace, channel 2) is $2.613V$, the peak as the capacitor is charging up. At time t_1, $872.9\mu s$ later, the voltage is at $2.265V$. The equation for voltage across a discharging capacitor is $V_C(t) = V_0 e^{-t/\tau} = 2.613V e^{-872.9\mu s/5.6ms} = 2.236V \approx 2.265V$. Therefore the peak detector is working properly.

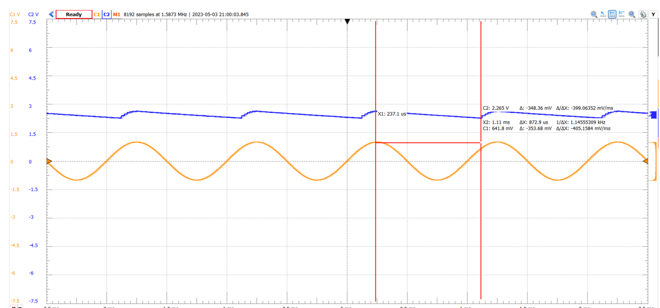

Fig. 31. Experimental Response of Peak detector with a sine wave input. The input to the peak detector is the yellow waveform, while the output of this stage is the blue waveform. Note that the sine wave input was measured before the DC offset was added. This results in the peak detector output having a DC offset when compared to the input.

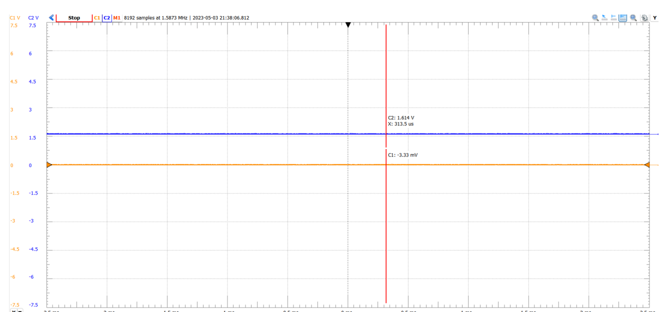

Fig. 32. Experimental Response of Peak detector with no sine wave input. When there is a DC input (the yellow waveform) we can clearly see that the output of the peak detector has the desired DC offset of 1.6 V to place the MOSFET in the cutoff region just below saturation.

E. LED Driver Experimental Results

The LED Driver experimental results below shows the current across Rs, which is the the current drop across each LED. The current drop across both LEDs was desired to be around 7 mA, with the high pass LED requiring more current than the low pass LED. This is shown below in the LPF and HPF LEDs figures. Fig. 33 shows the LPF LED having a max current of 4.5 mA flowing through it. Fig. 34 shows the HPF LED having a max current of 7.792 mA flowing through it. Additionally, this is ideal because the LPF LED (Green) has a max current of 7 mA, while the HPF LED (red) has a much higher current maximum of 30 mA, based on their respective data sheets.

The testing conditions were a 4.5 V VCC and a -4.5 V VEE for the supplies, and a 2 V peak to peak 1 kHz sin wave signal input with 50 percent symmetry. The sin waves were inputted into Junction 6 with the LPF, and the output was measured through Junction 13. For the HPF, the sin waves were inputted into Junction 7 and the output was measured through Junction 16.

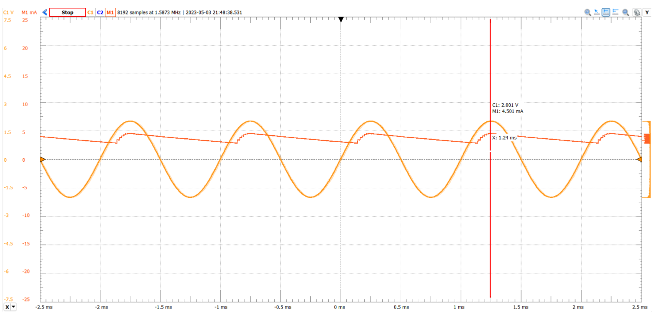

Fig. 33. Experimental Response of the Low pass LED driver. The yellow waveform shows the input to the circuit, while the red waveform shows the current through the LED.

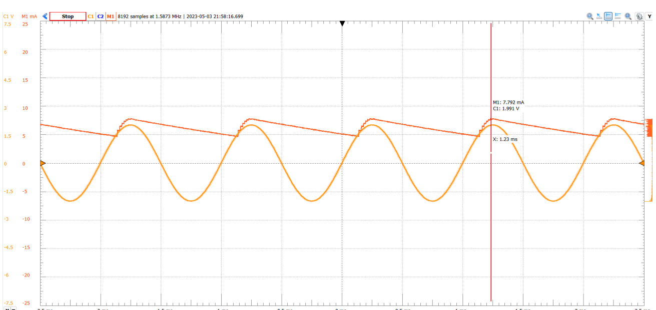

Fig. 34. Experimental Response of the High pass LED driver. Again, the yellow waveform shows the input to the circuit, while the red waveform shows the current through the LED.

F. System Level Experimental Results

The system experimental results below show the whole systems output with the song as its input. Fig. 35 shows the output of the system with the song inputted. The graph shows how the current is increased and decreased at different periods, based on the point in the song where there are higher frequencies (as this output was measured after the HPF). This graph shows the system working because of these different peaks in current, with the lower/flat peaks keeping the LED off (a point in the song with lower frequencies), and the higher peaks turning the LED on at that specific time. Not enough information is known to what the frequencies were at that specific point in the song, however seeing the change in the peaks shows that the output is responding in a proper manner to the input. Also, refer to the next paragraph to see visuals of the LEDs responding to the proper frequencies. We also saw it sufficient enough to show one of the MOSFET outputs, because the goal was to show the shifting currents based on their corresponding frequencies. Fig. 36 shows the output of the system at junction 16 with no song inputted. With just the DC inputted, the current shown running through the system is 0 mA, which indicates the LED is turned off. The testing conditions were using a 9 V battery for supply and the song as the input, and measuring the output at Junction 16.

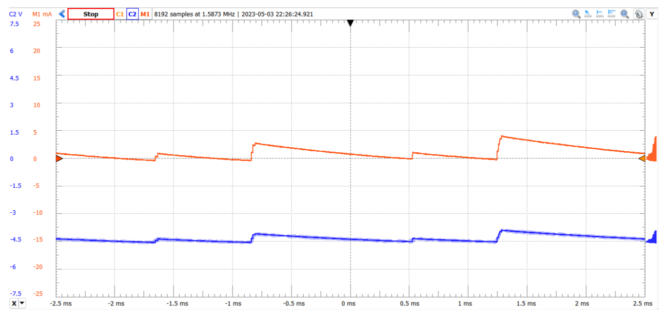

Fig. 35. Experimental Response of the system with the song, which the Red line properly shows different peaks of current based on the frequencies of the song at that particular time.

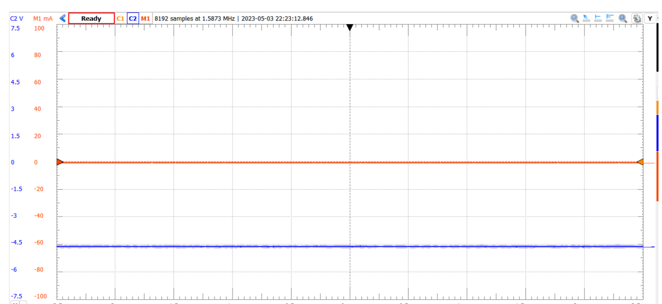

Fig. 36. Experimental Response of the system without the song, with the Red channel being the properly referenced output, showing that the LEDs are off even when the battery is plugged in.

Finally, we would like to provide evidence that our circuit produces the desired visual output. Before the beat drops in “Feel So Close”, there is significant frequency content in the treble range but little frequency content in the bass range. This corresponds to only the green LED being on, as shown in Fig. 37.

When the beat drops, there is substantial frequency content at both high and low frequencies, and we intend for both LEDs to be on at this part of the song. Indeed, this is what we observe as shown in Fig. 38.

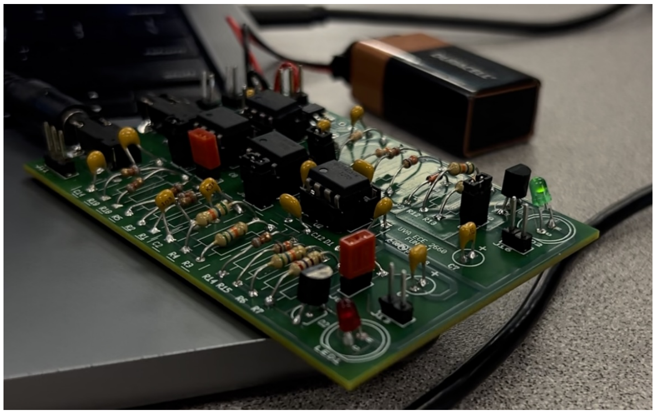

Fig. 37. When only high frequencies are present, the green LED is on and the red LED is off.

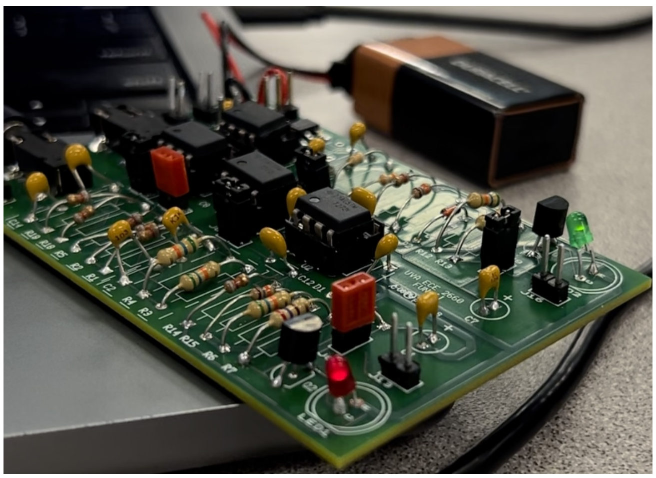

Fig. 38. When high and low frequencies are present, both LEDs are on.

V. Challenges and Workarounds

What challenges did you face in this project? How did you overcome them?

C1: One of the op amp holders was soldered in backwards. W1: The group kept note of the backwards U2, and instead of looking at the notch on the OPamp holder, the group instead looked at the "dot" on the board, which was visible. This made sure the OP amp was oriented correctly. C2: Misplacing an op amp in the rail splitter holder. W2: The group re looked at the schematic when the circuit was faulty, and realized the op amp was in the wrong spot. C3: Filters were not properly turning on. W3: The group misplaced/didn't use shunts in the proper junctions, which left parts of the circuit open. C4: Both the MOSFETs were found to be faulty, W4: The MOSFETs that were soldered onto the PCB board had extremely low Vt values, which would result in a faulty circuit. What the group did was we found two working MOSFETs, acquired their Vt and KN values, then replaced the old MOSFETs on the board with the working ones.

VI. Conclusions and Thoughts for Future Classes

The group learned how to utilize filters to have LEDs light up on either treble or bass. We accomplished this using Calvin Harris's song "Feel so close". The group learned this was primarily done through using a high and low pass filter, with the low pass filter outputting bass frequencies and the high pass filter outputting treble frequencies.

This project taught us a lot about using op amps in various configurations. Specifically, we learned how to use and apply op amps in the summing, Sallen-Key, and peak detector configurations. We also gained experience using the small-signal model when working with the MOSFETs. This project helped connect the theoretical work we did in class regarding the Fourier Transform to its applications in signal processing and audio.

Beyond the technical aspects of this project, we also learned a lot about the design process, especially in terms of making tradeoffs. One example of this can be seen in the LPF/HPF stages. Rather than using passive filters, which use fewer components and do not need an op amp, we chose to use active filters in the Sallen-Key architecture. Although this requires more hardware, it also results in a filter with a steeper rolloff of 40 db/decade versus the 20 db/decade rolloff obtained by using a passive filter.

We have a number of recommendations for future versions of this project if it is used again. First off, we think it would be better to begin working on the project later in the semester rather than right in the beginning. In the beginning of the semester, we didn't have a clear understanding of what each stage of the project does. In our opinion, it would be better to cover all of the material used in the project and then spend the last few weeks completing the entire design. If it were done this way, we would have a clear understanding of how the signal should move through the system and be processed while designing each stage.

We also think that in future semesters the MOSFETs should not be soldered onto the board until the LED driver has been designed and the transistors have been characterized. We had lots of issues with characterizing our MOSFETs as mentioned in Section V, and this was mainly due to the fact that they were soldered on too early and could possibly have been damaged by electrostatic discharge.

To future students working on this project - we strongly advise that you document your progress as you work on each stage rather than writing everything down after finishing the project. An easy way to do this is to simply write each section of the report as you complete that work, whether it be design, simulation, or experimental. It is also extremely helpful to be very organized with how you save your figures so that you don't lose them and can avoid repeating simulations or experiments.

VII. Collaboration Statement

An updated version of the team contract is shown in Fig. 39. It shows a detailed account of the work load distribution across the project. Overall, we worked as a cohesive team and the work was shared evenly.

Task	**Braden**	**Garrett**	**Nick**
Soldering non-design components	Group 1	Group 2	Group 3
Board beep check	x		
Input filter and summer: design/analysis	x	x	x
Input filter and summer: simulations	x	x	x
Input filter and summer: soldering			x
Input filter and summer: board testing	x		
LPF: design and analysis	x	x	x
LPF: simulations	x	x	x
LPF: soldering			x
LPF: board testing		x	
HPF: design and analysis	x	x	x
HPF: simulations	x	x	x
HPF: soldering			x
HPF: board testing	x		
Peak detector: design and analysis	x	x	x
Peak detector: simulations		x	
Peak detector: soldering			x
Peak detector: board testing	x		
MOSFET+LED: design and analysis	x	x	x
MOSFET+LED: simulations		x	
MOSFET+LED: soldering			x
MOSFET+LED: board testing	x		
Final project report tasks	Design, analysis, simulations, and experimental for HPF and Peak detector	Design, analysis, simulations, and experimental for LED driver and Summing amp	Design, analysis, simulations, and experimental for LPF and System level
Final project demo video tasks	Purpose and goals, interesting or challenging aspect of design	Explanation of filters in more detail, editing together the clips	Walk through of the high level functionality of each block, conlusion and takeaways

Fig. 39. Updated Team Contract

References

[1] *The "rail splitter" precision virtual ground*, TLE2426, Texas Instruments, 1998. [Online]. Available: https://www.digikey.com/en/products/detail/texas-instruments/TLE2426IP/371938.

[2] *High efficiency led in ϕ 3 mm tinted diffused package*, TLHR440, Vishay Semiconductors, 2022. [Online]. Available: https://www.vishay.com/docs/83006/tlhg440.pdf.

[3] *Low current led lamps*, HLMP-4700, Broadcom, 2021. [Online]. Available: https://media.digikey.com/pdf/Data%5C%20Sheets/Avago%5C%20PDFs/HLMP-47zz,HLMP-17zz_2021-08-09.pdf.

Appendix A
Code

Low pass filter code:

```python
from math import *

# resistors in lab kit that are at least 10 kOhm
resistors = [10e3, 12e3, 15e3, 18e3, 22e3, 27e3, 33e3, 39e3, 47e3, 56e3, 68e3, 82e3,
             100e3, 120e3, 150e3, 180e3, 220e3, 270e3, 330e3, 390e3, 470e3, 560e3, 680e3, 820e3,
             1e6]  # Ohms

capacitors = [0.01e-6, 0.1e-6, 0.47e-6, 1e-6, 4.7e-6, 10e-6, 100e-6]  # F

cutoff_desired = 300  # Hz
cutoff_tolerance = 50  # Hz

option = 1

for r3 in resistors:
    for r4 in resistors:
        for c1 in capacitors:
            for c2 in capacitors:
                q = sqrt(r3 * r4 * c1 * c2) / (r3 * c1 + r4 * c1)
                w_0 = 1 / sqrt(r3 * r4 * c1 * c2)
                f_0 = w_0 / 2 / pi
                if 0.5 <= q <= 0.707 and abs(f_0 - cutoff_desired) <= cutoff_tolerance:
                    print(f'design option: {option}')
                    print(f'{r3=}, {r4=} Ohms')
                    print(f'{c1=}, {c2=} F')
                    print(f'q={round(q, 4)}')
                    print(f'f_0={round(f_0, 4)} Hz\n')
                    option += 1
```

High pass filter code:

```python
from math import *

# resistors in lab kit that are at least 10 kOhm
resistors = [10e3, 12e3, 15e3, 18e3, 22e3, 27e3, 33e3, 39e3, 47e3, 56e3, 68e3, 82e3,
             100e3, 120e3, 150e3, 180e3, 220e3, 270e3, 330e3, 390e3, 470e3, 560e3, 680e3, 820e3,
             1e6]  # Ohms

capacitors = [0.01e-6, 0.1e-6, 0.47e-6, 1e-6, 4.7e-6, 10e-6, 100e-6]  # F

cutoff_desired = 400  # Hz
cutoff_tolerance = 5  # Hz

option = 1

for r10 in resistors:
    for r11 in resistors:
        for c5 in capacitors:
            for c6 in capacitors:
                q = sqrt(r10 * r11 * c6 * c5) / (r11*c5 + r11*c6)
                w_0 = 1 / sqrt(r10 * r11 * c6 * c5)
                f_0 = w_0 / 2 / pi
                if 0.5 <= q <= 0.707 and abs(f_0 - cutoff_desired) <= cutoff_tolerance:
                    print(f'design option: {option}')
                    print(f'{r10=}, {r11=} Ohms')
                    print(f'{c5=}, {c6=} F')
                    print(f'q={round(q, 4)}')
                    print(f'f_0={round(f_0, 4)} Hz\n')
                    option += 1
```

Appendix B
Derivations

Summing amp transfer function derivation, algebraic steps (no ECE analysis)

$$V_T = -(V_L + V_R)\frac{R_2}{R_{in}} = -V_{in}\frac{R_2}{R_{in}} \tag{36}$$

$$\frac{V_{in}}{R_{in}} = \frac{V_T}{R_1} + \frac{V_T - V_{out}}{R_5} \tag{37}$$

$$\frac{V_T - V_{out}}{R_5} = \frac{V_{in}}{R_{in}} - \frac{V_T}{R_1} \tag{38}$$

$$V_T - V_{out} = R_5(\frac{V_{in}}{R_{in}} - \frac{V_T}{R_1}) \tag{39}$$

$$V_{out} = V_T - R_5(\frac{V_{in}}{R_{in}} - \frac{V_T}{R_1}) \tag{40}$$

$$V_{out} = -V_{in}\frac{R_2}{R_{in}} - V_{in}\frac{R_5}{R_{in}} - V_{in}\frac{R_2 R_5}{R_{in} R_1} \tag{41}$$

$$H(s) = \frac{V_{out}}{V_{in}} = -\frac{R_f}{R_{in}} \tag{42}$$

$$R_f = R_2 + R_5 + \frac{R_2 R_5}{R_1} \tag{43}$$