Labs VIII-IX: Creation of an Analog Voice Scrambler Using a Balanced Modulator and Higher Order Filters

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ECE 4446: Audio Engineering Lab

November 2nd, 2017

Purpose:

The purpose of this lab was to create a device, called an analog voice scrambler, which takes an input signal (voice or music) and outputs a scrambled version of that signal by reflecting and shifting the input frequency content. For example, if the experimenter were to speak the word “hello” into the system, the output would sound like “Helay”. If the experimenter were to say “Helay” into the system, the output would be “Hello”. An analog voice scrambler has many applications such as a very basic analog message encryption device. The creation of the circuit introduced the experimenter to the application of complex filters such as the 2nd Order Chebyshev filter and the 5th order Elliptical filter. It also introduced the experimenter to the balanced modulator, a circuit which multiplies an input signal by a square wave.

System Theory:

The foundation of this voice scrambler stems from the well-known trigonometric identity: $Cos(a) * Cos(b) = \frac{1}{2}[Cos(a+b) + Cos(a-b)]$. This states that multiplying two sinusoids with differing frequencies will create two sinusoids which are spaced evenly about a center frequency. Therefore, if you multiply the human voice by a sinusoid, the output will be similar to that found in Figure 1. The higher frequency output section is identical to the input voice signal but shifted upwards. The lower frequency component is the reflected input voice signal shifted downwards. If the upper frequency content is filtered out, the remaining frequency content would contain only the reflected input signal. The reflection in the frequency domain effectively “scrambles” the input. Multiplying an input signal by a sinusoid is very difficult in analog electronics. However, multiplying an input signal by a square wave is much easier via a balanced modulator circuit. The extra frequency components created by the square wave’s harmonics can be filtered out via low pass filter. The following section details how a balanced modulator was used to create an analog voice scrambler.

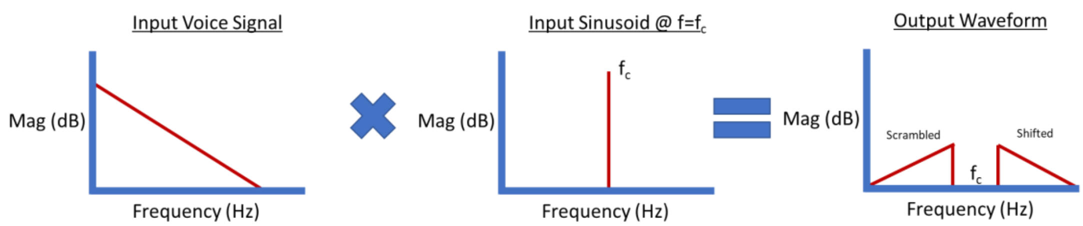


Figure 1. Overview of the effect of multiplying an input voice signal by a sinusoid.

The overall setup follows the block diagram shown below in Figure 2. The input signal is fed into a 2nd order band-pass filter which limits the incoming frequencies to the telephone band (between 300 Hz and 3000 Hz). This filter is realized by cascading a second order high pass Chebyshev filter with a second order low pass Chebyshev filter. The filtered signal is then placed into the balanced modulator. The balanced modulator multiplies the input signal by a square wave and requires that a square wave, oscillating at $f=f_c$, be provided at the V_C input. After the balanced modulator, the signal consists of pairs of frequency content which are symmetric about the harmonic frequencies of the square wave at V_c. However, only one of these pieces of frequency content is desired (the scrambled one detailed in Figure 1 above). Therefore, another filter must be used to remove all other undesired frequencies. One issue with filtering out this signal is the fact that the symmetric pairs are very close to one another. Thus, a high order low pass filter is required so that only the scrambled content below the fundamental frequency remains. Another issue is that the f_c frequency will permeate throughout the system and needs to be removed. A 5th order elliptical low pass filter can satisfy both of these requirements. This filter has an extremely steep response and contains a resonant null in the stop band. This resonant null can be designed to occur at f_c. The output now only contains the scrambled content and can be played through a speaker so that the experimenter can hear the scrambled content.

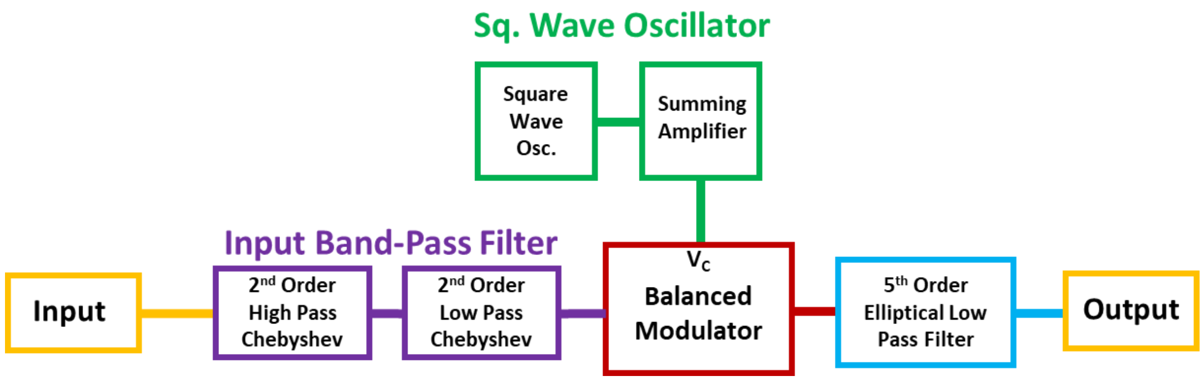


Figure 2. System overview of the analog voice scrambler.

Specifications:

The major design specifications for the experiment are detailed below along with techniques used by the experimenter to account for them. The methods to realize these techniques are explained in later sections.

- **Input band-pass filter should have upper and lower target -3dB frequencies of 300Hz and 3kHz:** Created by setting the -3dB of the cascaded high pass and low pass filters to 300Hz and 3kHz respectively.
- **Input band-pass filter should have a midband gain of 1 and a dB ripple of 0.5dB:** Set via the passive components of one low-pass Chebyshev filter and one high-pass Chebyshev filter realized by Sallen-Key filter topologies.
- **The square wave oscillator is to have the target frequency of 3.3Khz:** Set via the passive components on the square wave oscillator.
- **Output low-pass filter is to have a response as flat as possible to 3000 Hz and roll off as fast as possible above that frequency:** Realized by using a 5th order elliptical low pass filter.
- **Output low-pass filter maximum ω_s / ω_c value should be 1.5 and Output low-pass filter should have a ripple of 0.5 dB in the passband:** Elliptical Alignments Table in Dr. Leach's Filter Potpourri shows the required transfer function coefficients required to create such a filter. Set via the passive components of two biquadratic filters and an active low-pass filter.

Designing the Chebyshev Band-Pass Filter:

Most of the vowels and consonants required to comprehend audible speech lie within the telephone band, which lies between 300Hz and 3kHz. Using a 2nd order low-pass Chebyshev filter with a cutoff at 3kHz cascaded with and a 2nd order high-pass filter with a cutoff at 300Hz, a bandpass filter with a steeper roll off was made. Chebyshev filters have steeper roller offs than Butterworth filters, but have ripples in the passband. The specified required specifications were a 0.5dB ripple and a unity midband gain. Figure 3 and Figure 4 show the desired low-pass and high-pass filters respectively. f_c is the frequency at which the ripple stops, and f_3 is the -3dB frequency. They differ by a factor of the parameter x. Using the equations below, found in Dr. Leach's Filter Potpourri, x was calculated, and ultimately, the coefficients of the second order transfer function were calculated.

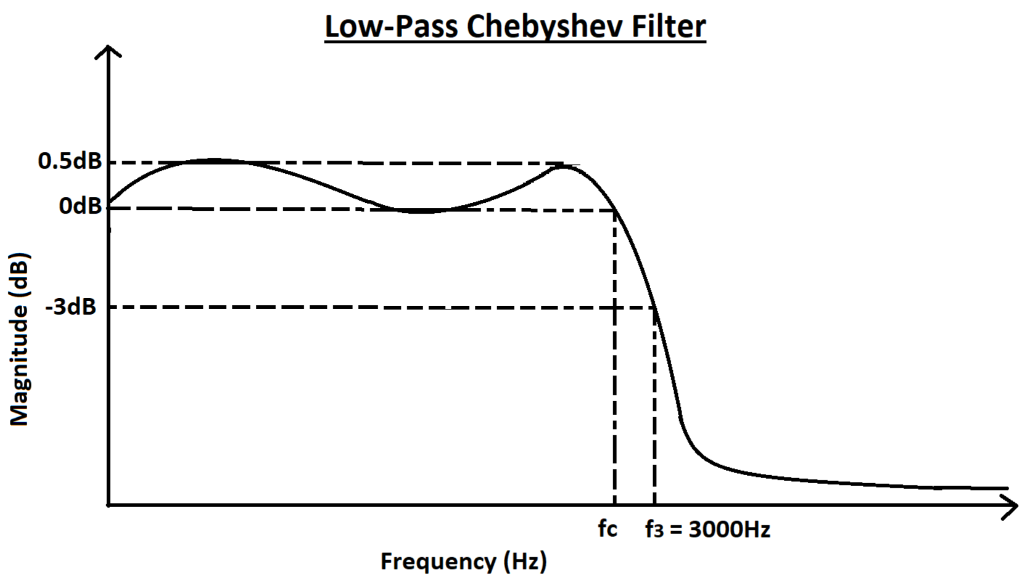


Figure 3. Low-pass Chebyshev Filter cutoff frequency at 3000Hz.

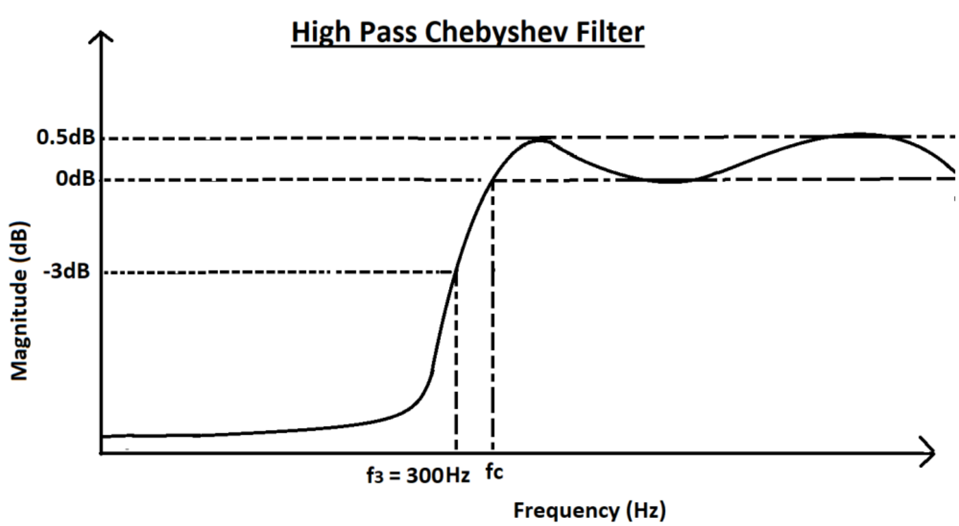


Figure 4. High-pass Chebyshev Filter with cutoff frequency at 300Hz.

The second order Chebyshev polynomial is $t_2(x) = 2x^2$-1. ϵ is a parameter determined by the ripple.

$$\epsilon = \sqrt{10^{\frac{ripple(dB)}{10}} - 1} = 0.3493114002$$

Using the Chebyshev polynomial and ϵ, the variable x was be calculated.

$$t_2(x) - \sqrt{\frac{1}{\epsilon^2} + 2t_2^2(0)} = 0$$

$$2x^2 - 1 - \sqrt{\frac{1}{\epsilon^2} + (-1)^2} = 0$$

$$x = \sqrt{\frac{1 + \sqrt{\frac{1}{\epsilon^2} + 2}}{2}} = 1.447935846$$

Using x, the cutoff frequency was calculated.

Low-Pass Filter:

$$\omega_c = \frac{\omega_3}{x} = \frac{2\pi * 3000}{1.447935846} = 13018\, rad/s$$

High-Pass Filter:

$$\omega_c = \omega_3 * x = 2\pi * 300 * 1.447935846 = 2729\, rad/s$$

The desired transfer function of the low pass filter is in the form

$$T_{LP}(s) = K \frac{1}{\left(\frac{s}{a\omega_c}\right)^2 + \left(\frac{s}{a\omega_c}\right) * \left(\frac{1}{b}\right) + 1}$$

To find these coefficients for the lowpass filter, the parameter h and θ are required. The order of the filter, n, is 2.

$$\theta = \frac{2-1}{n} * 90° = 45°$$

$$h = \tanh\left(\frac{1}{2}\sinh^{-1}\left(\frac{1}{\epsilon}\right)\right) = 0.7099423250$$

$$a = \sqrt{\frac{1}{1-h^2} - \sin^2(\theta)} = 1.231341799$$

$$b = \frac{\sqrt{1 + \frac{1}{h^2\tan^2(\theta)}}}{2} = 0.8637209781$$

Filter Transformation:

Using filter transformation, by substituting $p = \frac{s}{\omega_c}$, the high-pass of the desired equivalent were calculated.

$$T_{LP}(p) = K\frac{1}{\left(\frac{p}{a}\right)^2 + \left(\frac{p}{a}\right) * \left(\frac{1}{b}\right) + 1}$$

$$T_{HP}(p) = T_{LP}\left(\frac{1}{p}\right) = K\frac{1}{\left(\frac{1}{ap}\right)^2 + \left(\frac{1}{ap}\right) * \left(\frac{1}{b}\right) + 1} = K\frac{1}{\left(\frac{1}{ap}\right)^2 + \left(\frac{1}{ap}\right) * \left(\frac{1}{b}\right) + 1} * \frac{(ap)^2}{(ap)^2}$$
$$= K\frac{1}{\left(\frac{1}{ap}\right)^2 + \left(\frac{1}{ap}\right) * \left(\frac{1}{b}\right) + 1} * \frac{(ap)^2}{(ap)^2} = K\frac{(ap)^2}{(ap)^2 + (ap) * \left(\frac{1}{b}\right) + 1}$$

$$T_{HP}(s) = K\frac{\left(\frac{as}{\omega_c}\right)^2}{\left(\frac{as}{\omega_c}\right)^2 + \left(\frac{as}{\omega_c}\right) * \left(\frac{1}{b}\right) + 1}$$

Thus, using the coefficients for the low-pass filter, the coefficients of the high-pass filter were also calculated. Note that ω_c was not the same for the two filters.

Component of Sallen-Key Filter calculation:

The low-pass and high-pass second order Chebyshev filters were realized by a Sallen-Key topology. The desired transfer function needed to be in the form

$$T(s) = K\frac{One\ of\ the\ demoninator\ terms}{\left(\frac{s}{\omega_c}\right)^2 + \left(\frac{s}{\omega_c}\right) * \left(\frac{1}{Q}\right) + 1}$$

Since K is 1, the low-pass and high-pass transfer functions were written in this form.

$$T_{LP}(s) = K\frac{1}{\left(\frac{s}{a\omega_c}\right)^2 + \left(\frac{s}{a\omega_c}\right) * \left(\frac{1}{b}\right) + 1} = \frac{1}{\left(\frac{s}{16030}\right)^2 + \left(\frac{s}{16030}\right) * \left(\frac{1}{0.8637}\right) + 1}$$

$$T_{HP}(s) = K\frac{\left(\frac{as}{\omega_c}\right)^2}{\left(\frac{as}{\omega_c}\right)^2 + \left(\frac{as}{\omega_c}\right) * \left(\frac{1}{b}\right) + 1} = \frac{\frac{s}{2217}}{\left(\frac{s}{2217}\right)^2 + \left(\frac{s}{2217}\right) * \left(\frac{1}{0.8637}\right) + 1}$$

With the Sallen-Key topology given in Figure 5, the components of the low-pass filter were calculated using the equations given in Dr. Leach's Filter Potpourri. $C_1$and C_2 were arbitrarily chosen to be 0.1µF and 0.01µF respectively, which was acceptable so long as $4Q^2C_2/\ C_1 < 1$. R_1 and R_2 were calculated. They were interchangeable.

$$R_{1,2} = \frac{1}{2Q\omega_c C_2} \pm \sqrt{1 - 4Q^2 * \frac{C_2}{C_1}} = 6636\Omega, 586\Omega$$

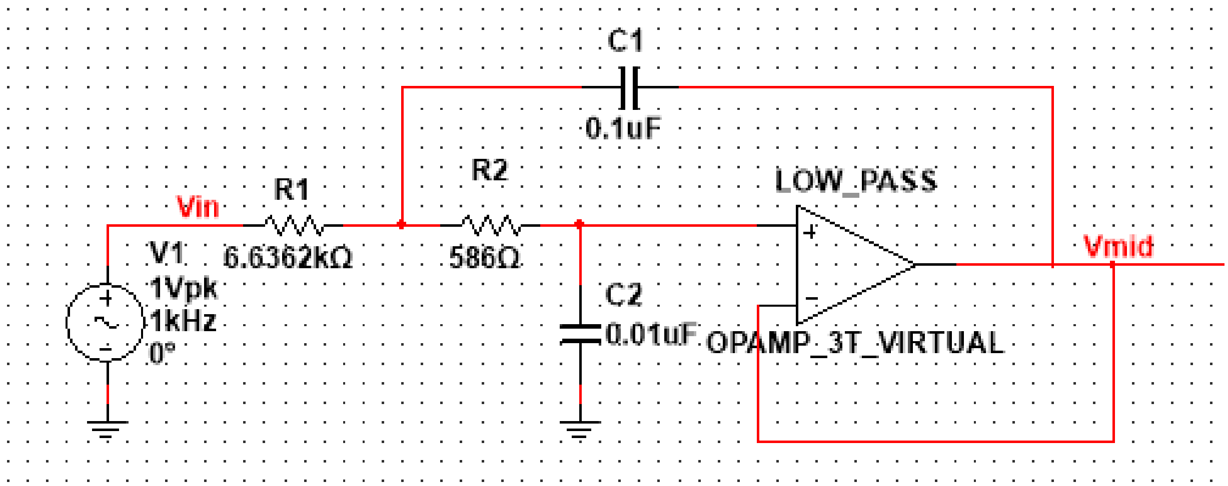


Figure 5. Sallen-Key filter implementing a low-pass Chebyshev Filter with cutoff frequency at 3000Hz.

With the Sallen-Key topology given in Figure 6, the components of the high-pass filter were calculated using the equations given in Dr. Leach's Filter Potpourri. $C_1$and C_2 were arbitrarily both chosen to be 0.1µF. R_1 and R_2 were calculated.

$$R_1 = \frac{1}{2Q\omega_c C_2} = 5.475k\Omega$$

$$R_2 = \frac{2Q}{\omega_c C} = 16.339k\Omega$$

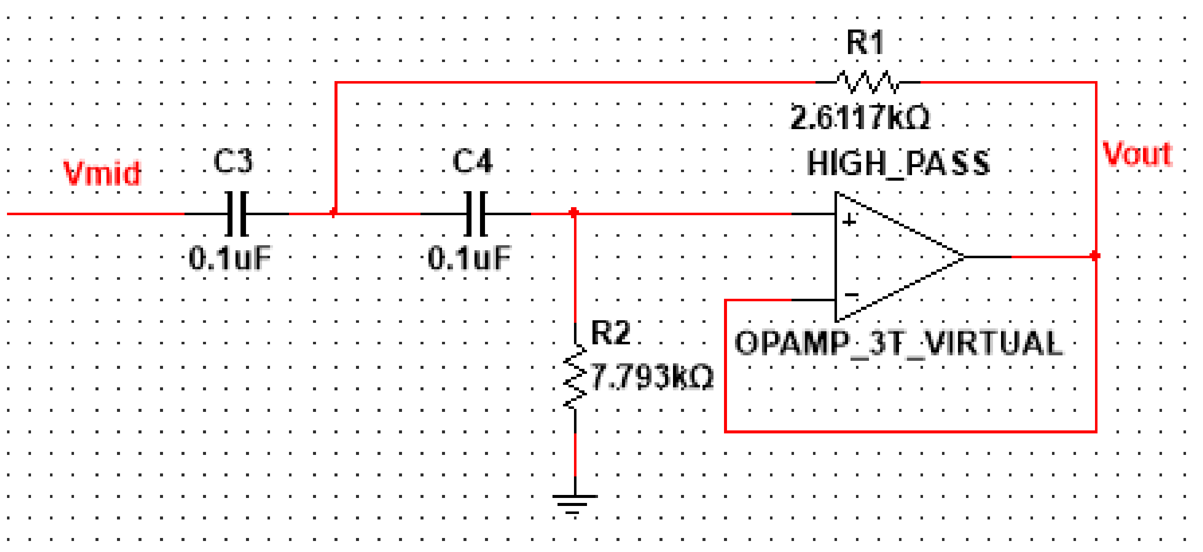


Figure 6. Sallen-Key filter implementing a high-pass Chebyshev Filter with cutoff frequency at 300Hz.

Cascading the filters shown above resulted in a bandpass filter from 300Hz to 3kHz. The actual results and simulated results are shown in Figure 14 and Figure 15.

Designing the Balanced Modulator:

The balanced modulator is a circuit which multiplies an input signal by a square wave and is shown below in Figure 7. V_C is the input square wave signal and V_I is the input voice signal provided by the output of the band-pass filter. It is easiest to understand the operation of the balanced modulator circuit when it is analyzed in two separate conditions: when V_C is high, and when V_C is low. When V_C is low, the JFET acts like an open circuit. Thus, the operational amplifier acts like an inverting amplifier with $A_V = -\frac{R_F}{R_5}$. When V_C is high, the JFET acts like a short circuit and the amplifier has a gain of $A_v = \left(\frac{1}{R3+R4}\right)\left(R4 - \frac{RF*R3}{R5}\right)$. The derivation of these gains can be found in Appendix 1. This circuit should be designed such that the gains of each condition are equal and opposite. The component values were given in the design guide as the following values: R1=1MΩ, R2=10kΩ, R3=100kΩ, R4=200kΩ, R5=200kΩ, and RF=100kΩ. These values create a gain of 0.5 and -0.5 for the two previously mentioned conditions.

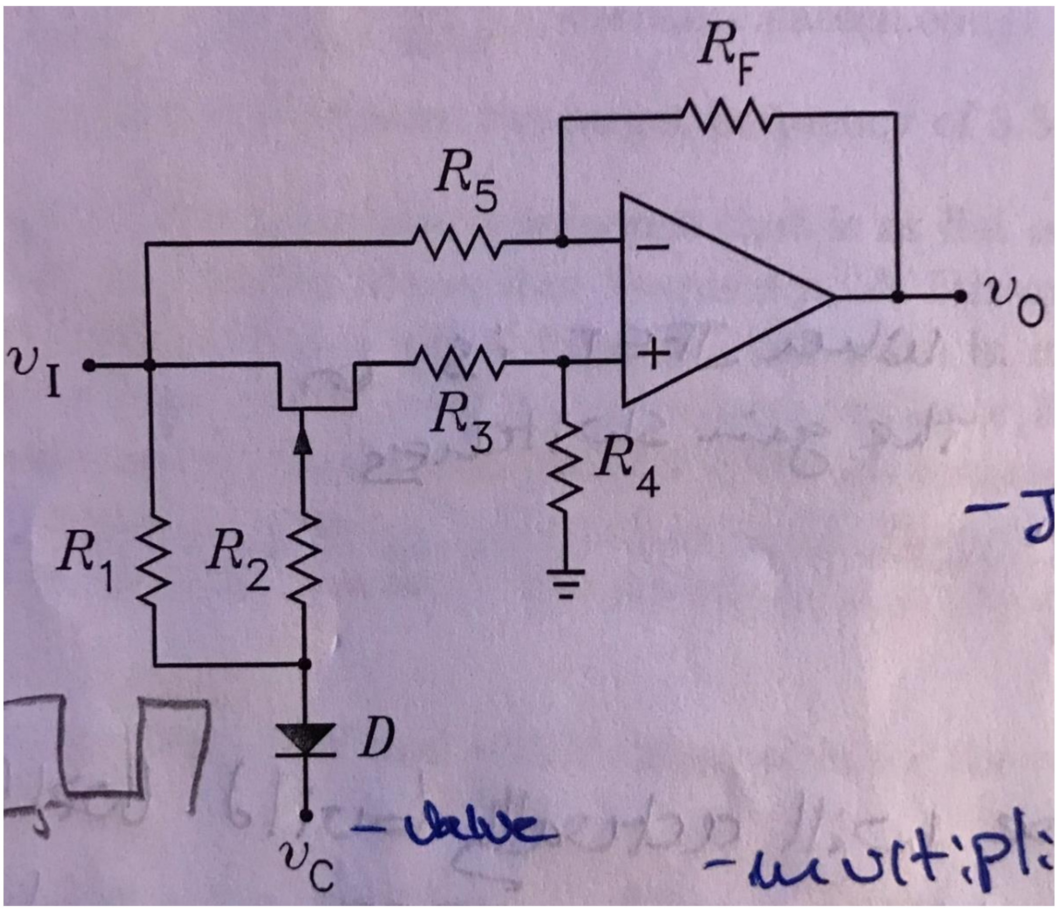


Figure 7. Balanced modulator circuit used to multiply incoming audio signals by a square wave.

After creating the balanced modulator, the control signal for V_C was designed. The input needed to be a 3.3 kHz square wave varying between V_{MAX}=1V and V_{MIN}=-5V (Vpp of 6V). The JFET acts like a short circuit when V_{GS}=0V and like an open circuit when $V_{GS}=V_{TO}$. The values for maximum and minimum voltages were chosen to assure that the square wave swept through the entire operational range of the JFET. Therefore, the overall circuit design needed to be a square wave oscillator cascaded with an inverting summer circuit to fix both the amplitude and the DC offset. The overall oscillator circuit is shown below in Figure 8. The period of oscillation can be found through: $T = 2RC1 * \ln\left[\frac{1+\lambda}{1-\lambda}\right]$ where $\lambda = \frac{R1}{R1+R2}$. R1 and R2 were arbitrarily chosen as 100kΩ in order to set λ equal to 0.5. C1 was also arbitrarily chosen as 0.1µF since it is easily found in the lab. Using these values in the above equation, R was found to be 1379Ω. The output of this square wave was then cascaded with an inverting summing amplifier to fix both the amplitude and the DC offset. The output voltage of this circuit is provided through the equation: $V_o = -\mathrm{Rf}(\frac{\mathrm{Vsqpp}}{\mathrm{R3}} + \frac{V2}{R4})$ where Vsqpp is the output peak to peak value of the square wave oscillator and V2 is tied to the 15V supply rail. Rf was arbitrarily selected as 75kΩ. The Vsq term sets the amplitude while the V2 term sets the DC offset. Thus, the equation $-Rf * \left(\frac{Vsqpp}{R3}\right) = -6$ was solved to provide a 6Vpp output signal. This required an R3 value of 300kΩ. R4 was then solved to provide a DC offset of -2V to shift the maximum and minimum values to 1V and -5V respectively. The equation $-2 = -Rf * (\frac{15}{R4})$, provided that R4=562500Ω. This circuit was breadboarded and tested to show an output waveform of 3.1 kHz varying between -4.93 V and 1.48 V (shown in Figure 9). The frequency was lower than expected. Therefore, the value of R was reduced to bring the square wave frequency to a more acceptable value of 3.35 kHz.

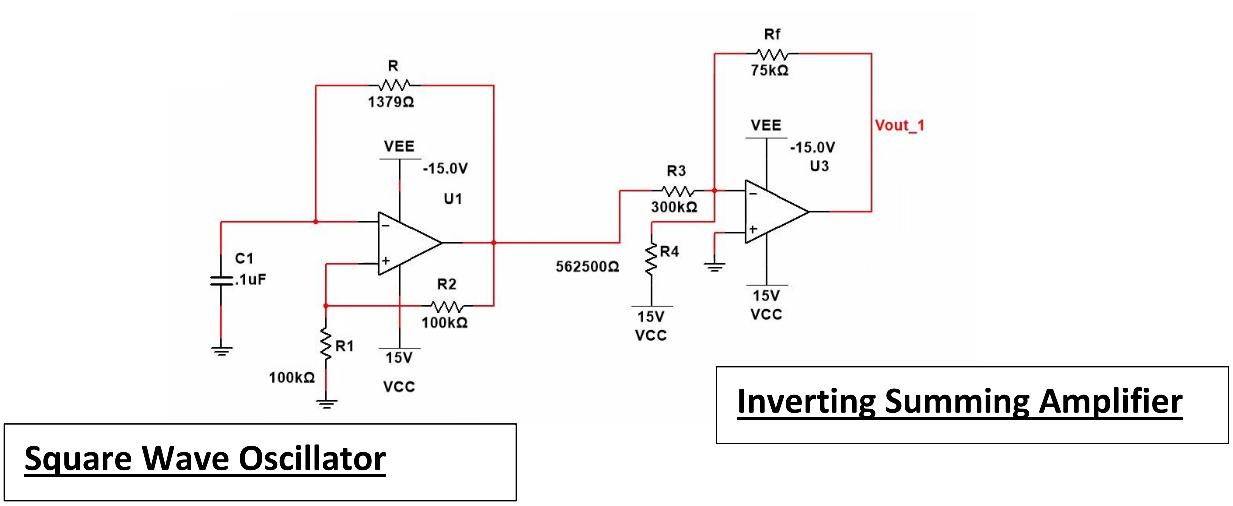


Figure 8. Square wave oscillator circuit cascaded with inverting summing amplifier to create V_c waveform.

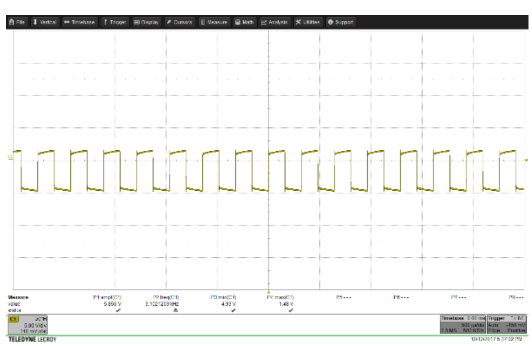


Figure 9. Square wave oscillator output with fixed amplitude and adjusted DC offset.

Designing the Elliptical Filter:

The signal from the balanced modulator contained the desired frequencies, which ranged from 300Hz to 3kHz. However, when the signal was modulated with the square wave, reflections of the desired frequency were also produced, which centered around the odd harmonics of the square wave. A filter is thus needed to remove every reflection except for the desired frequency. Since the 3.3kHz square wave needed to be filtered out, preserving up to 3kHz as much as possible, a steep roll-off was required. This was done by using a 5th order low-pass elliptical filter shown in Figure 10. The required specifications were to have a 0.5dB ripple, and a ratio of the start of the stopband frequency (f_s shown in red) to the cutoff frequency (f_c shown in green) of 1.5. Since it was an odd ordered filter, the ripples caused negative dB ripple, and the passband ends at f_c where the gain is -0.5dB for the last time.

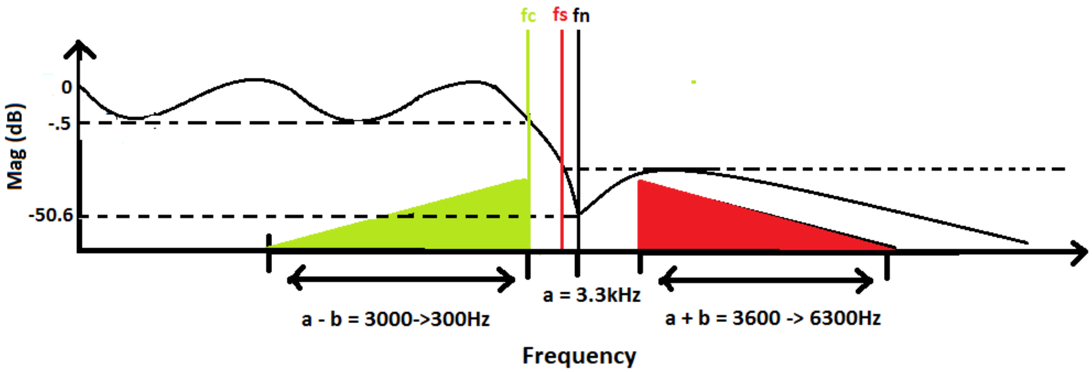


Figure 10. Transfer function graph of the desired filter which filters out unwanted frequencies (red triangle) and retains the desired frequencies (green triangle).

Looking at the Elliptical Alignments Table of -0.5dB ripple in Dr. Leach's Filter Potpourri (Page 18) in https://leachlegacy.ece.gatech.edu/ece4435/filtrpot.pdf, with n = 5 and $\frac{\omega_s}{\omega_c} = 1.5$, the desired coefficients of the required transfer function were found, and the gain at the null frequency was found to be -50.6dB. The desired coefficients are shown in the transfer function below.

$$\mathrm{T(s)} = \frac{1}{\frac{s}{0.42597\omega_c}+1} + \frac{\left(\frac{s}{1.5574\omega_c}\right)^2+1}{\left(\frac{s}{1.0158\omega_c}\right)^2+\left(\frac{s}{1.0158\omega_c}\right)*\left(\frac{1}{6.2145}\right)+1} + \frac{\left(\frac{s}{2.3319\omega_c}\right)^2+1}{\left(\frac{s}{0.70455\omega_c}\right)^2+\left(\frac{s}{0.70455\omega_c}\right)*\left(\frac{1}{1.3310}\right)+1}$$

$$\text{Where } \omega_c = \frac{\omega_n}{c_1} = \frac{f_n*2\pi}{c_1} = \frac{3300*2\pi}{1.5574} = 13313\, rad/s$$

The 5th order transfer function equation shown above can be broken up into one 1st order, and two 2nd order transfer functions. This was achieved by cascading an active 1st order low-pass filter with two 2nd order biquadratic filters.

Active 1st Order Low-Pass RC Filter:

$$T_{LP}(\mathrm{s}) = \frac{1}{\frac{s}{0.42597\omega_c}+1} = \frac{1}{\frac{s}{5671}+1}$$

Using an arbitrary an arbitrary capacitor value of 0.1μF and $R = \frac{1}{\omega * C}$ at corner frequency, the resistance was calculated to be 1.76kΩ. The circuit topology is shown below in Figure 11.

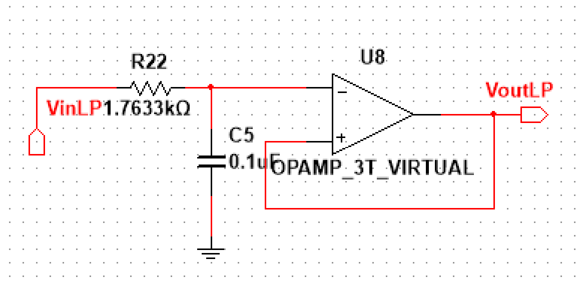


Figure 11. 1st Order low-pass RC filter.

Designing Biquadratic Filters:

Given the cutoff frequencies ω_N, ω_D gain K and quality factor Q of a 2nd order transfer function, the component values of a biquadratic filter were calculated below, with schematic shown in Figure 12 and Figure 13, and the actual and simulated results shown in Figure 20 and Figure 21 respectively. The two 2nd order transfer functions needed to be in the form

$$T(s) = K \frac{\left(\frac{s}{\omega_N}\right)^2 + 1}{\left(\frac{s}{\omega_D}\right)^2 + \left(\frac{s}{\omega_D}\right) * \left(\frac{1}{Q}\right) + 1}$$

First Biquad:

$$T_{B1}(s) = \frac{\left(\frac{s}{1.5574\omega_c}\right)^2 + 1}{\left(\frac{s}{1.0158\omega_c}\right)^2 + \left(\frac{s}{1.0158\omega_c}\right) * \left(\frac{1}{6.2145}\right) + 1} = \frac{\left(\frac{s}{20735}\right)^2 + 1}{\left(\frac{s}{13524}\right)^2 + \left(\frac{s}{13524}\right) * \left(\frac{1}{6.2145}\right) + 1}$$

Second Biquad:

$$T_{B2}(s) = \frac{\left(\frac{s}{2.3319\omega_c}\right)^2 + 1}{\left(\frac{s}{0.70455\omega_c}\right)^2 + \left(\frac{s}{0.70455\omega_c}\right) * \left(\frac{1}{1.3310}\right) + 1} = \frac{\left(\frac{s}{31046}\right)^2 + 1}{\left(\frac{s}{10104}\right)^2 + \left(\frac{s}{10104}\right) * \left(\frac{1}{1.3310}\right) + 1}$$

Using arbitrary capacitor values of 0.1μF, and gain K = 1, the rest of the resistor values were calculated.

$$R = \frac{1}{\omega_D C}$$

$$R_1 = \frac{R}{K} = R$$

$$R_4 = Q * R$$

$$R_2 = R_1 * \left(\frac{\omega_N}{\omega_D}\right)^2$$

$$R_3 = Q * R_2$$

Using these equations, the resistor values were calculated, and their values are shown in the schematic in Figure 12 and Figure 13.

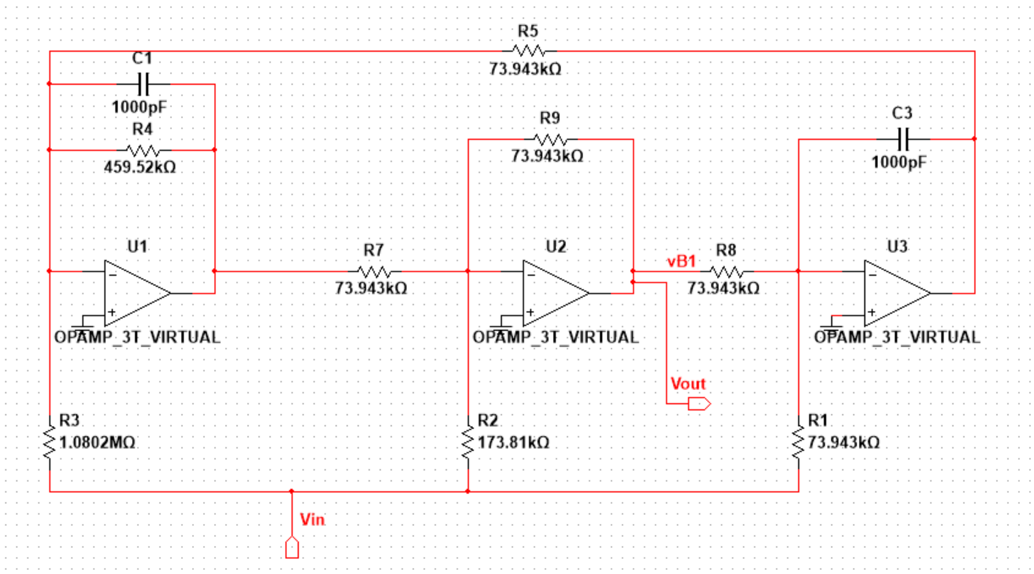


Figure 12. Schematic of the first 2^{nd} order biquadratic filter.

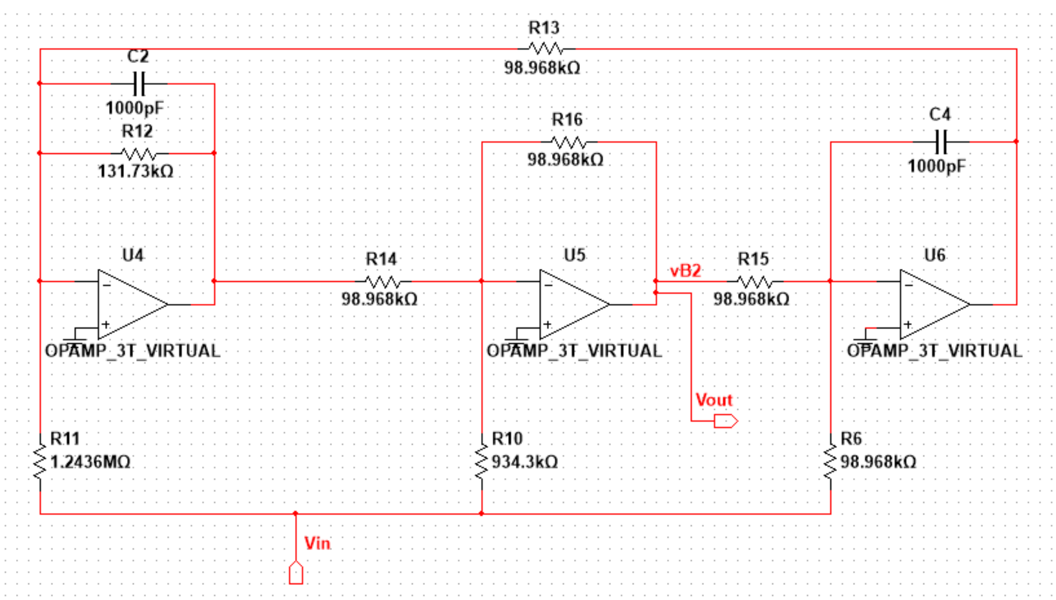


Figure 13. Schematic of the second 2^{nd} order biquadratic filter.

Cascading the 1 first order, and the 2 second order low-pass filters resulted in the 5^{th} order low-pass elliptical filter, with a null frequency at 3.3kHz. The actual and simulated results are shown in Figure 20 and Figure 21 respectively.

Analog Voice Scrambler Analysis:

Several quantitative and qualitative tests were conducted on the voice scrambler to assess its functionality and quality. The following sections explain these tests as well as provide the resulting waveform captures and frequency spectra.

Testing the Input Chebyshev Band-Pass Filter:

The specifications noted that the input Chebyshev band-pass filter was to have -3dB frequencies at 300 Hz and 3 kHz. This was tested by connecting the filter to the impedance analyzer to check the frequency response. Channel A was connected to the input of the filter and Channel B was connected to the output of the filter. The obtained frequency response is shown below in Figure 14. The cursors show that the lower -3dB frequency was 298.35 Hz (0.6% error) and the upper -3dB frequency was 3 kHz (~0% error). The specifications stated that the midband gain was to be unity and the passband ripple was to be 0.5dB. This frequency response shows that a midband gain of 0.709dB or 1.085 (linear) was achieved. This was only 8.5% away from the target value. A ripple of 0.709dB is shown in the frequency response. In linear scale, this equates to a gain of 1.085 and is 2% away from the target value. This frequency response shows that all specifications and requirements were met by the input Chebyshev band-pass filter meaning that the proper range of frequencies were supplied to the balanced modulator. The theoretical frequency response for the Chebyshev filter is shown in Figure 15. This shows that the -3 dB frequencies match up accordingly and that the midband gain is slightly above 0 as well.

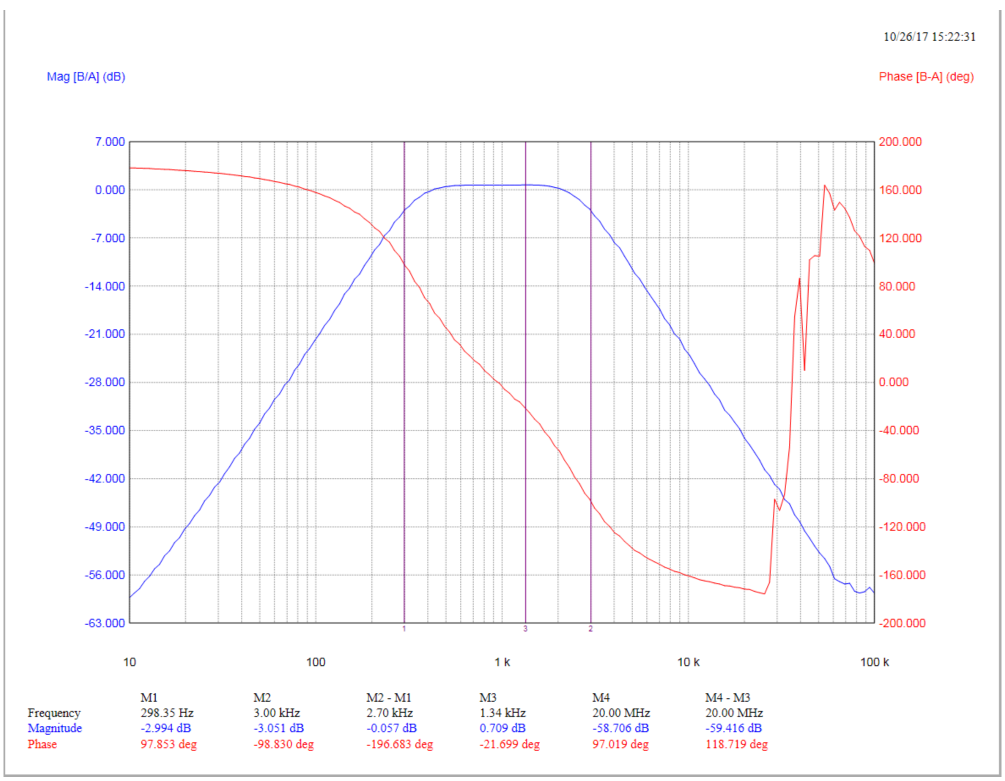


Figure 14. Frequency response of the input Chebyshev band-pass filter showing -3dB frequencies of 298.35 Hz and 3 kHz.

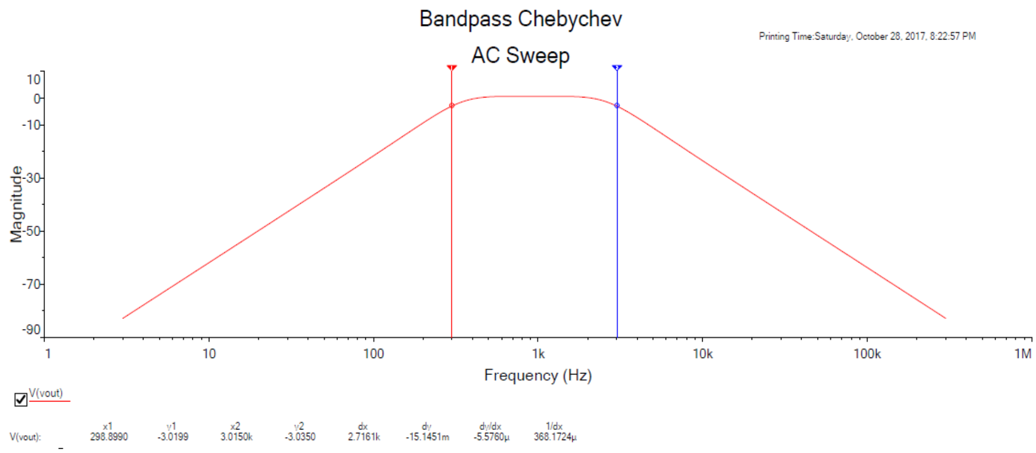


Figure 15. Theoretical input bandpass Chebyshev filter frequency response.

Testing the Balanced Modulator Circuit:

The balanced modulator circuit was tested next. Both time domain and frequency domain data were gathered for varying input frequencies. Figure 16 shows the time domain capture for a 400 Hz 1Vpp sinusoidal input voltage (yellow trace). The blue trace shows the 3.3 kHz square wave at the V_C input and the red trace shows the balanced modulator's output. This waveform capture clearly shows that the balanced modulator is functioning properly. First, the square wave is 3.35 kHz which is only 1.5% off from the target frequency of 3.3 kHz. When the blue square wave goes low, the input yellow sinusoid is multiplied by -0.5 (shown by cursors). When the blue square wave goes high, the input yellow sinusoid is multiplied by +0.5 (shown by cursors). The red trace clearly shows that this is occurring and that the balanced modulator is working properly.

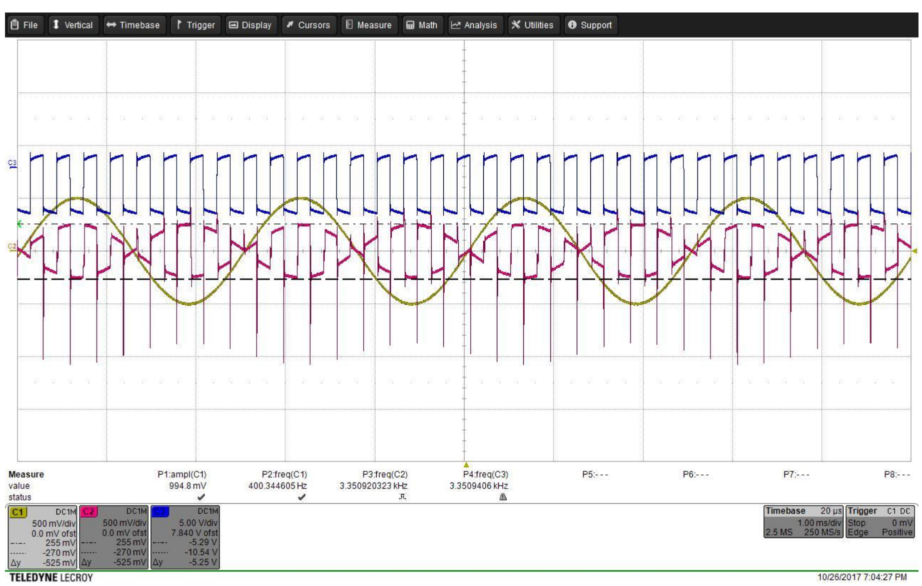


Figure 16. Time domain capture for 400 Hz sine wave input. (Yellow: 400 Hz sine wave input; Blue: 3.3kHz square wave input present at V_C; Red: Balanced modulator output).

Figure 17 shows the frequency spectrum of the balanced modulator output for the same 1 Vpp 400 Hz input sinusoid. Peaks in the frequency response were expected to occur at 2.9 kHz (3.3 kHz - 400 Hz), 3.7 kHz (3.3 kHz + 400 Hz) and 9.5 kHz (9.9 kHz - 400 Hz). These peaks were expected due to the trig identity detailed in the introduction. A square wave is composed of single frequency components at the odd harmonic values from the fundamental frequency of 3.3 kHz. Therefore, while the input 400 Hz sine wave was multiplied by a 3.3 kHz sine wave, it was also multiplied by a 9.9kHz sine wave. The trigonometric identity states that this should create sinusoids at 3.3 kHz +/- 400 Hz, and 9.9 kHz +/- 400Hz. Since the span of the spectrum was only 100 Hz to 10 kHz, only the 2.9 kHz, 3.7 kHz, and 9.5 kHz components were visible. The actual measured values were 2.952kHz at -6.53dB, 3.752kHz at -6.71dB, and 9.645kHz at -16.82 dB. All of these values were less than 5% away from the target values and show that the balanced modulator was working properly. Most of the error was caused by the square wave oscillator being 3.35 kHz instead of 3.3 kHz. The 2.952kHz and 3.752 kHz signal showed an attenuation by ~-6dB. This is expected because the balanced modulator has a gain of 0.5 which equates to ~-6dB. There are several spurious components in the frequency spectrum. The largest of these components was measured with the cursors and was -24.6dB. These components are most likely are caused by the large voltage spikes that can be seen in the balanced modulator output in Figure 16.

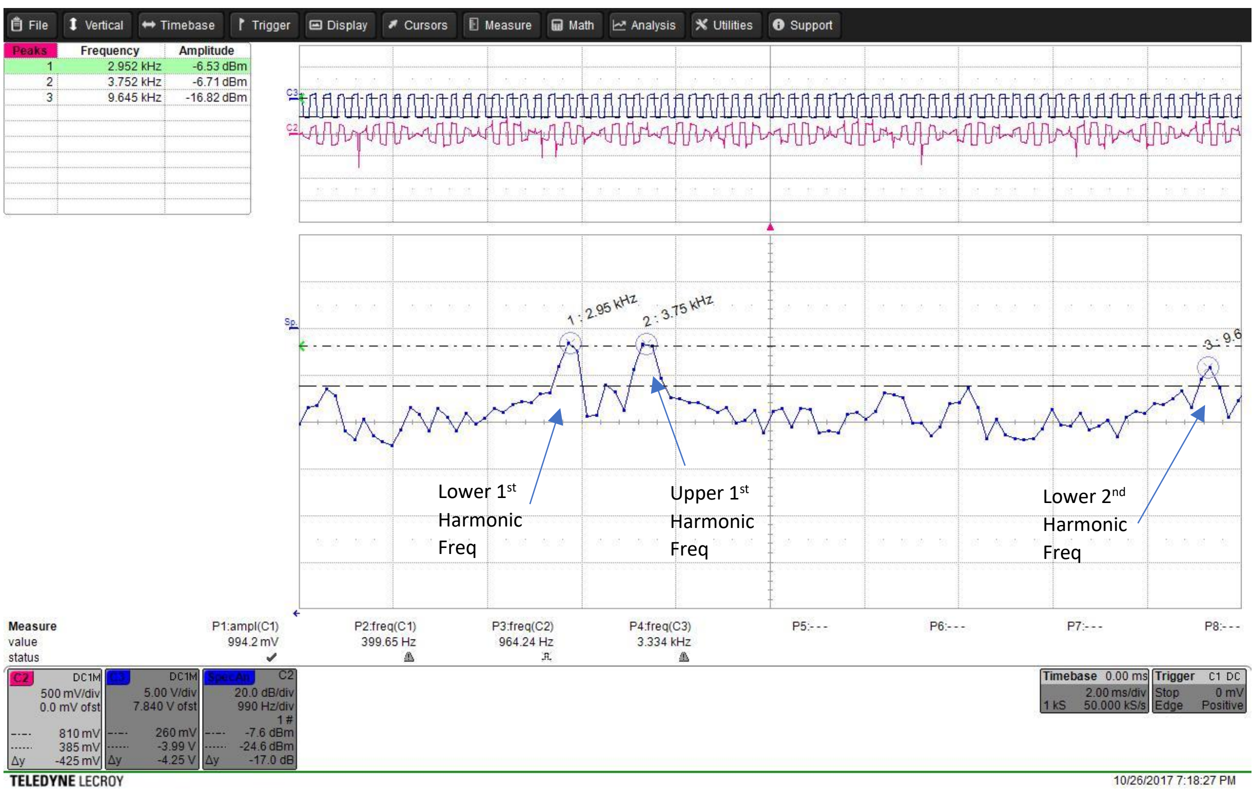


Figure 17. Frequency spectrum capture of the balanced modulator for a 400 Hz sine wave input.

Figure 18 shows the time domain capture for a 2000 Hz 1Vpp sinusoidal input voltage (yellow trace). The blue trace shows the 3.3 kHz square wave at the V_C input and the red trace shows the balanced modulator's output. This waveform capture clearly shows that the balanced modulator is functioning properly. First, the square wave is 3.35 kHz which is only 1.5% off from the target frequency of 3.3 kHz. When the blue square wave goes low, the input yellow sinusoid is multiplied by -0.5 (shown by cursors). When the blue square wave goes high, the input yellow sinusoid is multiplied by +0.5 (shown by cursors). The red trace clearly shows that this is occurring and that the balanced modulator is working properly.

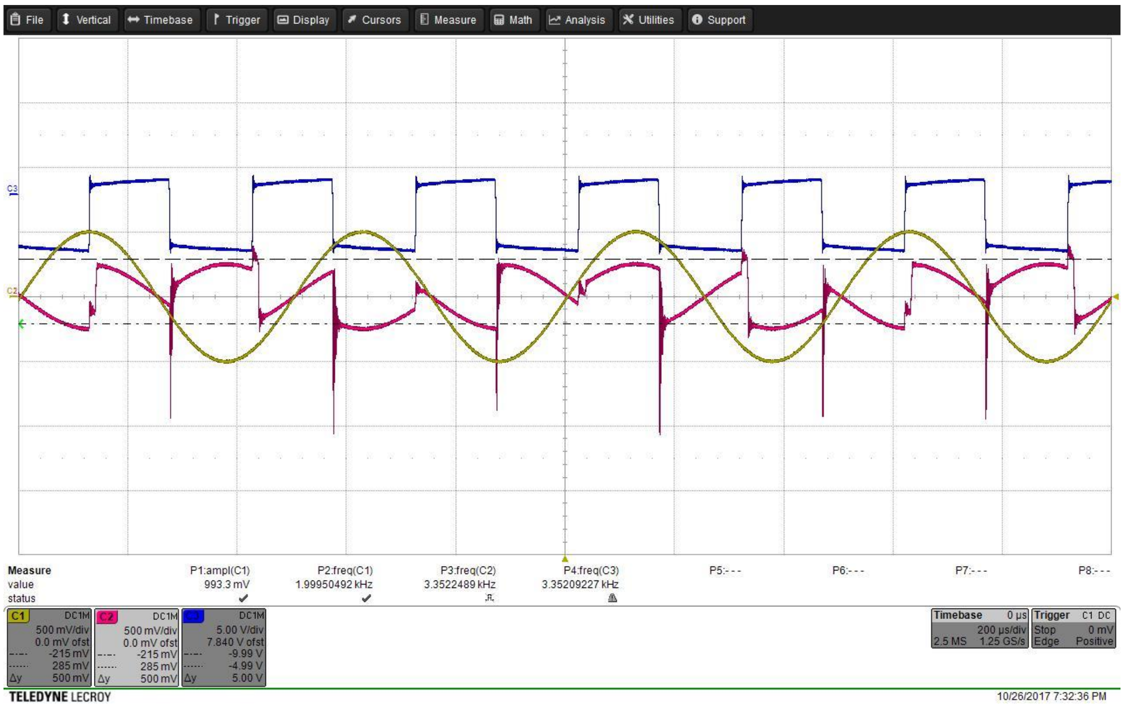


Figure 18. Time domain capture for 2000 Hz sine wave input. (Yellow: 2000 Hz sine wave input; Blue: 3.3kHz square wave input present at V_C; Red: Balanced modulator output).

Figure 19 shows the frequency spectrum of the balanced modulator output for the same 1 Vpp 2000 Hz input sinusoid. Peaks in the frequency response were expected to occur at 1.3 kHz (3.3 kHz - 2 kHz), 5.3 kHz (3.3 kHz + 2 kHz) and 7.9 kHz (9.9 kHz - 2000 kHz). These peaks were expected due to the trig identity detailed in the introduction. Since the span of the spectrum was only 100 Hz to 10 kHz, only the 1.3 kHz, 5.3 kHz, and 7.9 kHz components were visible. The actual measured values were 1.351 kHz at -6.11dB, 5.351 kHz at -6.3dB, and 8.052 kHz at -16.81 dB. All of these values were less than 5% away from the target values and show that the balanced modulator was working properly. Most of the error was caused by the square wave oscillator being 3.35 kHz instead of 3.3 kHz. The 2.952kHz and 3.752 kHz signal showed an attenuation by ~-6dB. This is expected because the balanced modulator has a gain of 0.5 which equates to ~-6dB. There are several spurious components in the frequency spectrum. The largest of these components was measured with the cursors and was -24.6dB. These components are most likely are caused by the large voltage spikes that can be seen in the balanced modulator output in Figure 18.

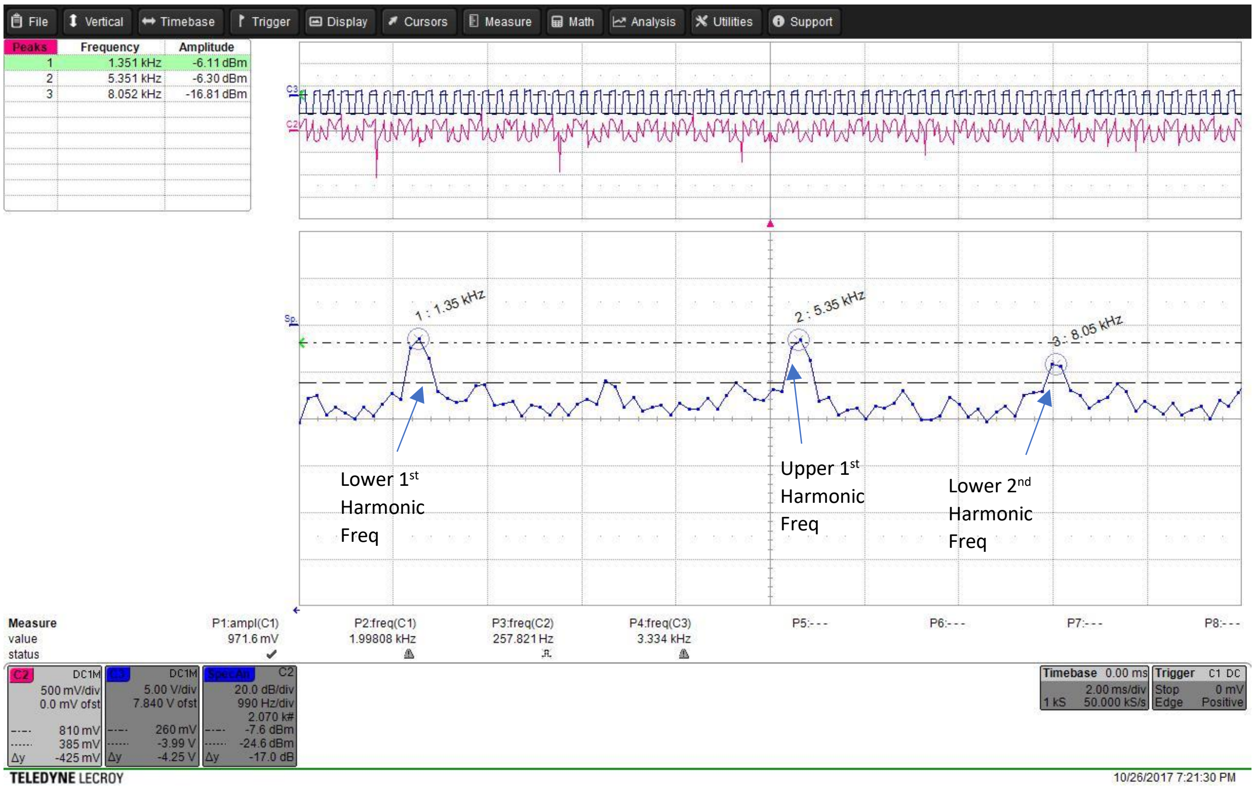


Figure 19. Frequency spectrum capture of the balanced modulator for a 2000 Hz sine wave input.

As the input frequency was varied from the original 400 Hz to 2000 Hz, the real-time spectrum showed the peaks spreading out evenly from their respective resonant frequencies. The lowest frequency component acted similar to how the scrambler will act on an incoming voice signal. The higher the input frequency, the lower the output frequency due to the previously explained frequency reflection. The elliptical filter was used to filter out the higher, undesired frequency components to leave only the reflected "scrambled" frequency content.

Testing the Elliptical Filter:

The elliptical filter was tested next. The goal for this filter was to have a null at 3.3 kHz, a ripple of 0.5dB in the passband, and have a maximum ω_s/ω_c value of 1.5. The frequency response of the filter was captured using the impedance analyzer and is shown in Figure 20. The null frequency was 3.47 kHz which was an acceptable 5% from the target value of 3.3 kHz. The cutoff frequency was f_c=2.26 kHz and the stop frequency was 3.33 kHz which provided an ω_s/ω_c value of 1.47 which is suitably below the maximum value of 1.5. The second zero occurred at 5.11 kHz. This showed that the elliptical filter was working as desired. The theoretical elliptical filter frequency response is shown in Figure 21 for comparison. The null frequency was 3.29 kHz which is acceptably close to the obtained result. The ω_s/ω_c value was 1.5 which was also within the specifications.

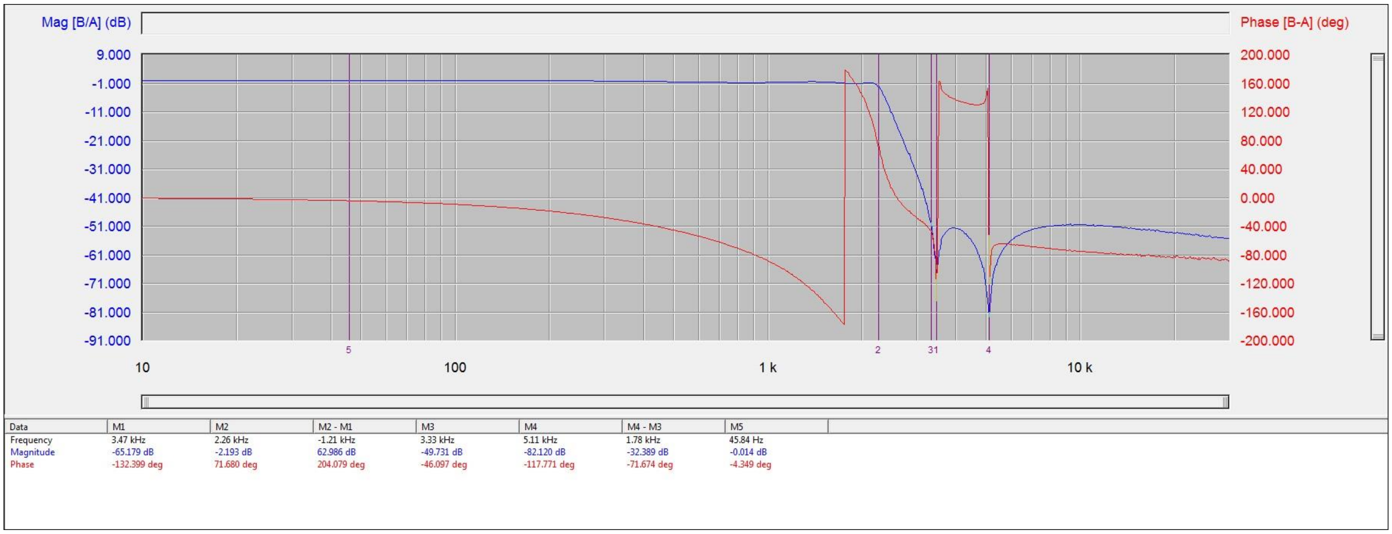


Figure 20. Frequency response of the 5th order elliptical filter. (Null frequency=3.47kHz, f_c=2.26kHz, f_s=3.33kHz)

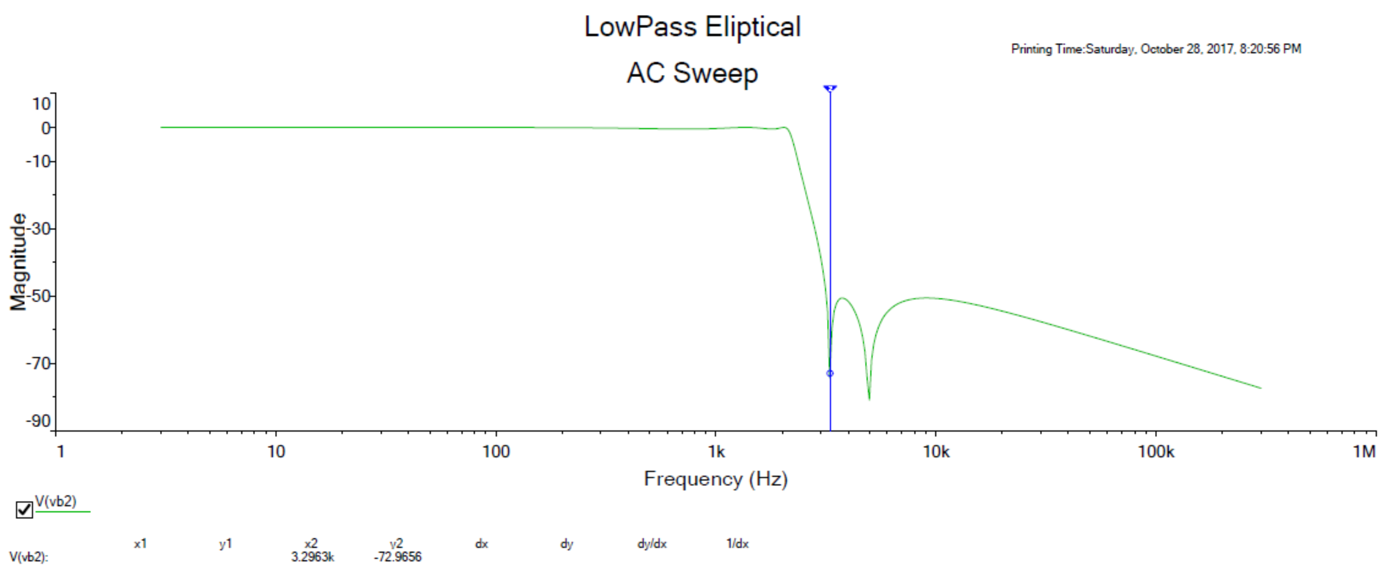


Figure 21. Theoretical elliptical frequency response from Multisim.

The last quantitative test was of the overall system. An input sinusoid was placed into the Chebyshev band pass which was connected to the balanced modulator and the elliptical filter. The output was taken from the elliptical filter. The Lecroy oscilloscope was used to measure the frequency spectrum of the output signal. This test was to confirm that the elliptical filter was properly filtering out all frequencies except the scrambled frequency content (below the square wave's fundamental frequency of 3.3 kHz). The first frequency spectrum (shown in Figure 22) was taken when the input signal was a 1 Vpp 400 Hz sinusoid. It is clear from the spectrum that the only remaining signal is one at 2.95 kHz (other than the two input frequencies). The 9.65 kHz component (visible in the unfiltered spectrum in Figure 17) has been successfully removed by the elliptical filter. The 2.95 kHz signal is very low (-37dB) because it is so close to the null frequency of 3.3 kHz. Therefore, it is still attenuated greatly by the elliptical filter. A 400 Hz and 3.35 kHz signal can be seen because the input signals permeate through the system.

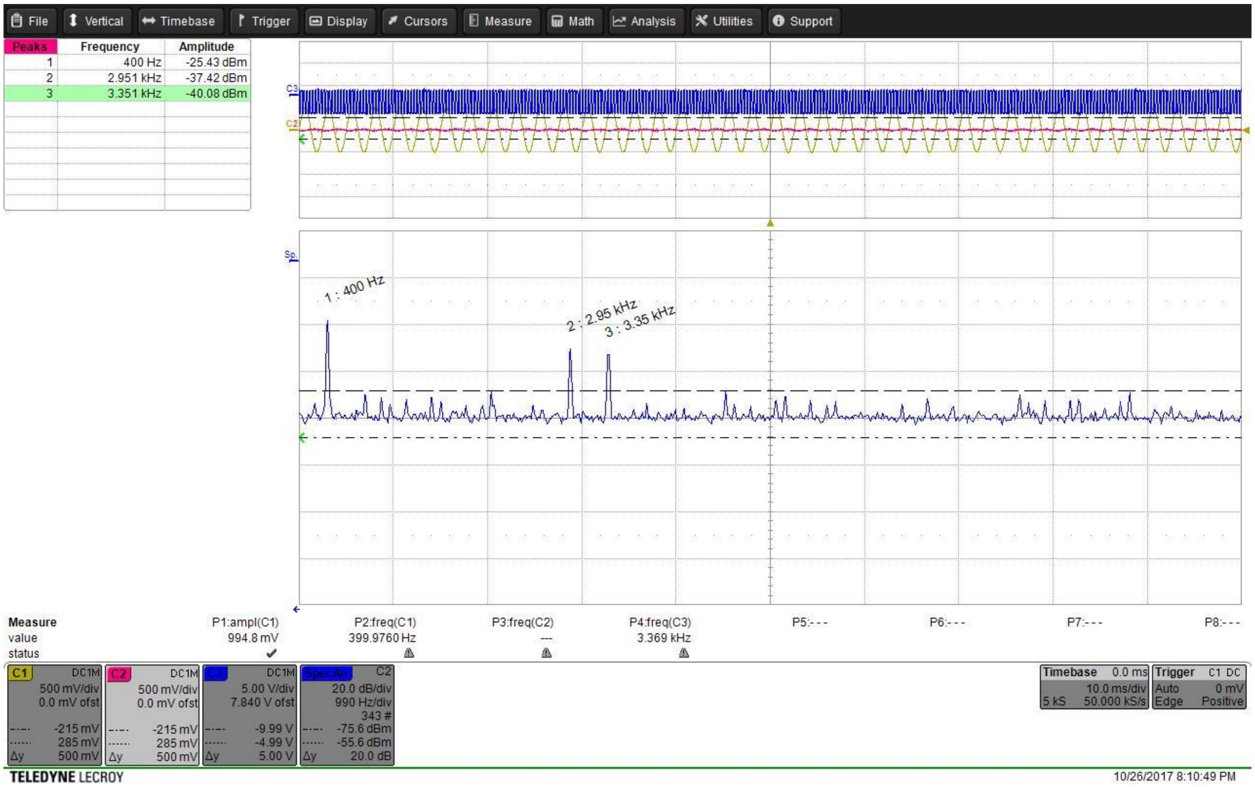


Figure 22. Frequency spectrum after the elliptical filter for a 400 Hz 1 Vpp input sinusoid.

Figure 23 shows the elliptical filter output frequency spectrum for when the input is a 2 kHz sinusoid. It is clear that the higher harmonic components that were previously visible (in Figure 19) have now been removed. The only signals that remain are the 1.35 kHz signal, a 2 kHz signal, and a 3.35 kHz signal. The 1.35 kHz signal is only attenuated by -7 dB which is mostly due to the balanced modulator's gain of 0.5 (-6dB). The 2 kHz signal and 3.35 kHz signal are due to the input signals permeating the system. While these signals still remain, they are at -25 dB and -40 dB respectively and should not be very noticeable in the output signal. The 1.35 kHz is attenuated far less than the other signals and will be the dominate output frequency from the device. Overall this test showed that the elliptical filter was working properly by filtering out all frequency content above the square wave's fundamental frequency.

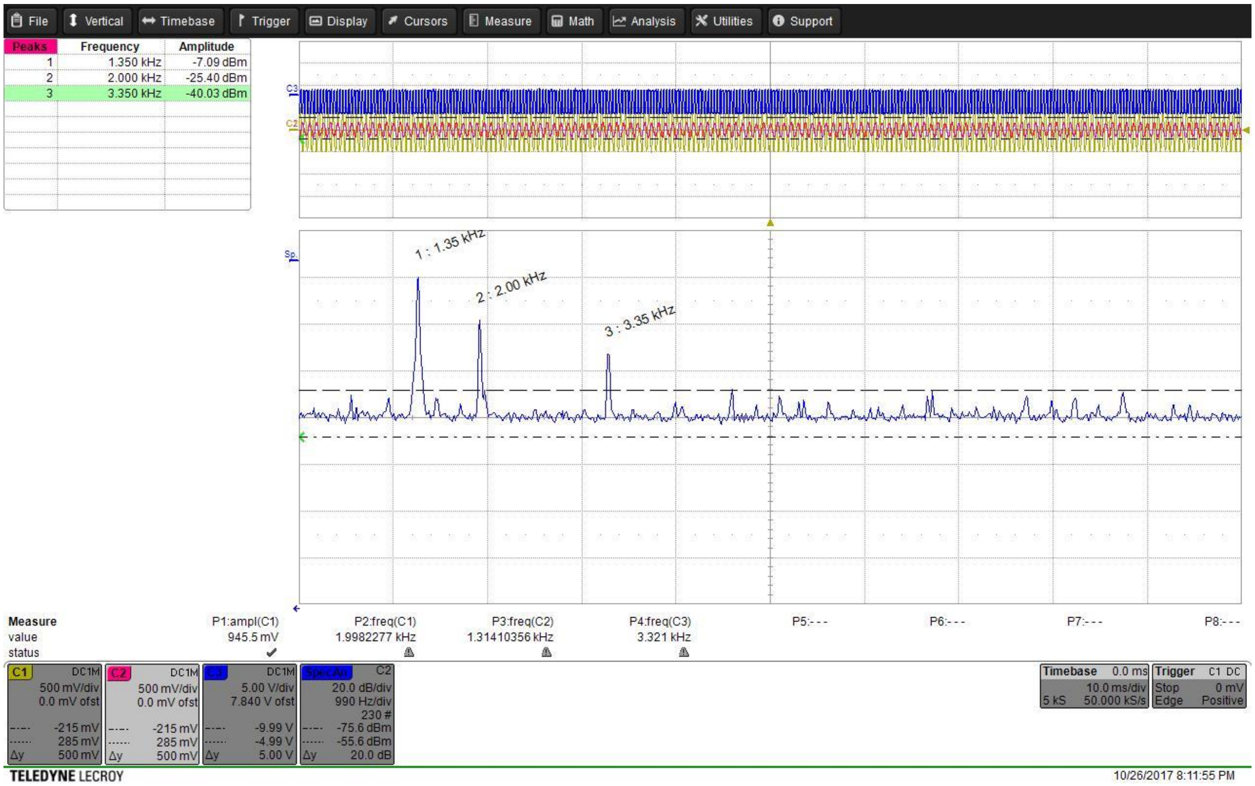


Figure 23. Frequency spectrum after the elliptical filter for a 2000 Hz 1 Vpp input sinusoid.

Qualitative System Test:

Now that the system has been quantitatively checked for proper functionality, the last test is to supply it with sound and qualitatively test the output. This test was conducted by connecting a cell phone to the circuit's input via 3.5mm audio jack and connecting the output signal to a speaker. A song was then played into the input and the output "scrambled" sound from the speaker was recorded by the experimenter. This "scrambled" recording was then played back through the system to result in the original audio signal from the speakers. This showed that the circuit was qualitatively performing as planned. The most noticeable problem with the system was that a soft high-pitched tone could be heard at the output speaker. This is most likely due to the 3.3 kHz tone making it through to the speakers. This issue could be rectified by placing a notch filter (with null at 3.35 kHz) at the output to make sure this is removed. While the output was audible, the 0.5 gain from the modulator did make the output signal harder to hear. Placing an amplifier on the output would fix this issue. A picture of the breadboarded circuit can be seen in Figure 24.

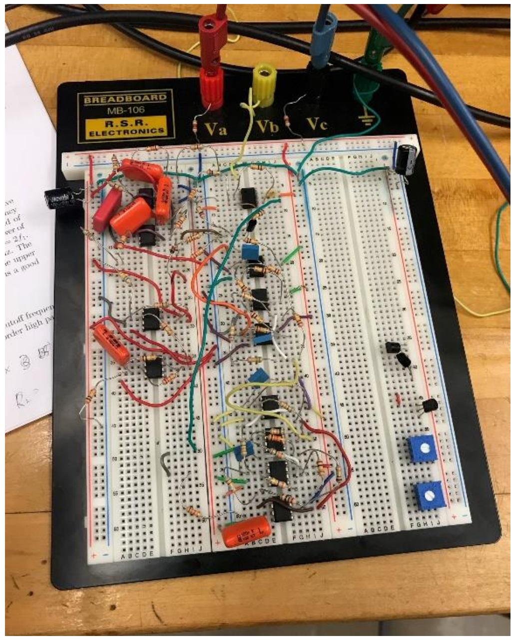


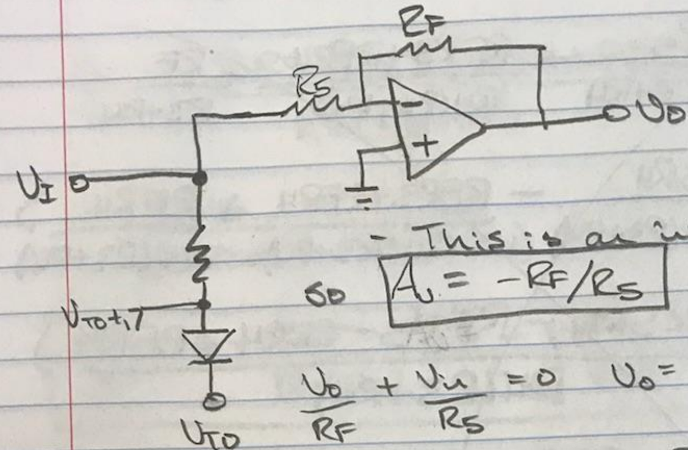
Figure 24. Breadboarded analog voice scrambler circuit.

Conclusions:

The goal of this lab was to create an analog voice scrambler device using 2nd order Chebyshev filters, a balanced modulator, and an elliptical filter. This lab introduced the experimenter to the Chebyshev and Elliptical filters which required in depth design work in order to meet the required specifications. Overall, all systems proved to meet the specified requirements and a fully functional analog voice scrambler was displayed via quantitative and qualitative testing. This circuit could be improved by re-designing the elliptical filter to do a better job at nulling out the 3.35 kHz frequency components due to the V_C square wave. It was also found that using a 3rd order elliptical filter instead of a 5th order elliptical filter had minimal effects on the output sound quality. Build cost and component count could be reduced by only using a 3rd order elliptical filter. Overall, the desired analog scrambler effect was proven when playing a “scrambled” audio file into the input of the circuit produced the original audio file at the output.

Appendix 1: Derivation of balanced modulator gain

When $V_{gs} = V_{TO}$ Equivalent Circuit:
JFET is open



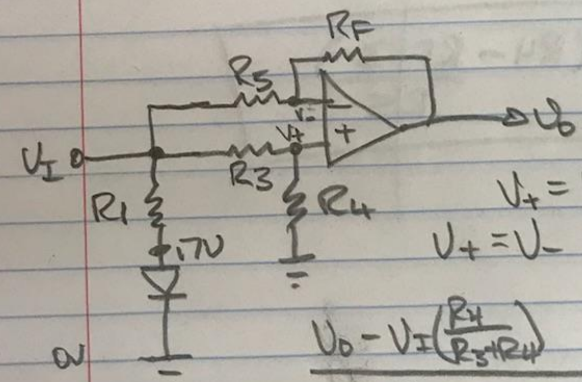
- This is an inverting amplifier

so $\boxed{A_v = -R_F/R_S}$

$$\frac{V_o}{R_F} + \frac{V_{in}}{R_S} = 0 \qquad V_o = -\frac{V_{in}}{R_S} R_F$$

$$\frac{V_o}{V_{in}} = -R_F/R_S$$

When $V_{gs} = 0V$ Equivalent Circuit
JFET is short



$$V_+ = V_I\left(\frac{R_4}{R_3+R_4}\right)$$

$$V_+ = V_-$$

$$\frac{V_o - V_I\left(\frac{R_4}{R_3+R_4}\right)}{R_F} + \frac{V_I - V_I(R_4/R_3+R_4)}{R_S} = 0$$

$$\frac{V_o}{V_I R_F} - \frac{R_4}{R_F(R_3+R_4)} + \frac{1}{R_S} - \frac{R_4}{R_S(R_3+R_4)} = 0$$

$$\frac{V_o}{V_I} = \frac{R_4}{R_3+R_4} - \frac{R_F}{R_S} + \frac{R_4 R_F}{R_S(R_3+R_4)} \rightarrow \frac{V_o}{V_I} = \frac{R_4}{R_3+R_4} - R_F(R_3+R_4$$

$$= \frac{R_4}{R_3+R_4} - \frac{R_F(R_3+R_4)}{R_S(R_3+R_4)} + \frac{R_4 R_F}{R_S(R_3+R_4)}$$

$$\frac{V_o}{V_I} = \frac{R4R5}{R5(R3+R4)} - \frac{RF(R3+R4)}{R5(R3+R4)} + \frac{R4RF}{R5(R3+R4)}$$

$$\frac{V_o}{V_I} = \frac{R4R5 - RFR3 \cancel{- RFR4} \cancel{+ RFR4}}{R5(R3+R4)}$$

$$= \frac{R4\cancel{R5}}{\cancel{R5}(R3+R4)} - \frac{RFR3}{R5(R3+R4)}$$

$$\boxed{\frac{V_o}{V_{in}} = A_v = \left(\frac{1}{R3+R4}\right)\left(R4 - \frac{RFR3}{R5}\right)}$$