16 Step Analog Sequencer with Digital Tempo Input

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ECE 445 Design Review - Spring 2017

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1 Introduction

1.1 Objective

Analog electronic musical instruments are generally very expensive. There are two main reasons for this. The first is simply because the market is small. The second reason is that certain analog, discrete-form electrical components, like matched BJT's for example, which are necessary for building analog electronic musical instruments, are expensive as well as not consistently manufactured.

Our objective is to build a modern take on a 70's style analog music synthesizer, which will be controlled by a digital step-sequencer. In the simplest possible terms, we aim to make a looping melody machine. The synthesizer will be driven by a voltage controlled oscillator implementing a linear voltage to exponential current circuit. A digitally controlled step-sequencer will continuously loop the oscillator through sixteen steps, defining sixteen notes in a user created melody.

Our implementation will be physically different in the sense that the schematic as well as the layout and design will be entirely our own. Additionally, these types of instruments are generally very over-priced and ours will cost much less than anything out there on the market currently. As an example of this last point, a good step-sequencer, like the Doepfer Dark Time, goes for around $725 [1]. A good analog VCO based synthesizer, like the Moog Voyager, goes for upwards of $3,000 [2]. And these two pieces must be bought separately in order to achieve what we aim to accomplish in one instrument.

1.2 Background

Musical electronic synthesizers with built-in with step-sequencers, as they are known today, first became available in the 1960's and 70's with the advent of commercially available transistors.

During this golden age of the analog audio synthesizer, these devices were extremely expensive and were consequently for professional use only. Two famous examples of musical step-sequencers from this era include the Moog 960 and the Buchla 100 [3]. Beginning in the early 1980's, as home computers and other consumer-level digital products became more widely available, step-sequencers became digitally controllable as well as affordable by the average enthusiast. Examples from this period include the Linn LM-1 and LM-2 machines [4].

Beginning in the early 1990's however, hardware step-sequencers were mostly replaced by software, GUI based instruments [5]. Today, there are no widely available analog synthesizers with built-in step sequencers. In fact, there are only a handful of boutique companies that make analog synthesizers and unfortunately, they do not function exclusively as step-sequencers. Likewise, there are only a few companies that make hardware step-sequencers and sadly, they do not sequence sounds synthesized internally.

Our instrument will be both self-contained (a step-sequencer combined with a synthesizer) as well as less expensive than anything on the market currently.

1.3 High-Level Requirements List

- Tempo of sequenced melody must be digitally user-controllable and re-settable to any integer value ranging from 1 to 9,999 beats per minute.
- Sequencer must be programmable, via physical buttons on the device, to rest for any step as well as start over at any step during its loop.
- The linear-to-exponential voltage controlled oscillator must be capable of saw tooth, square and triangle wave output.

2 Design

From a high-level perspective, our design is composed of four stages. The first is the MCU stage, where the user may input a desired tempo for the looping melody. The Transistor-Transistor Logic (TTL) stage is for the actual step-sequencing of the melody. The user will set sixteen voltages via potentiometers and these sixteen voltages will be looped through by a TTL counter in combination with an analog MUX. The synthesizer stage consists of the analog, voltage-controlled oscillator (VCO). The type of waveform output by the oscillator will be selectable by the user as saw tooth, square or triangle. The last stage is the power supply, which will supply and regulate the correct power requirements for each of the three previous stages.

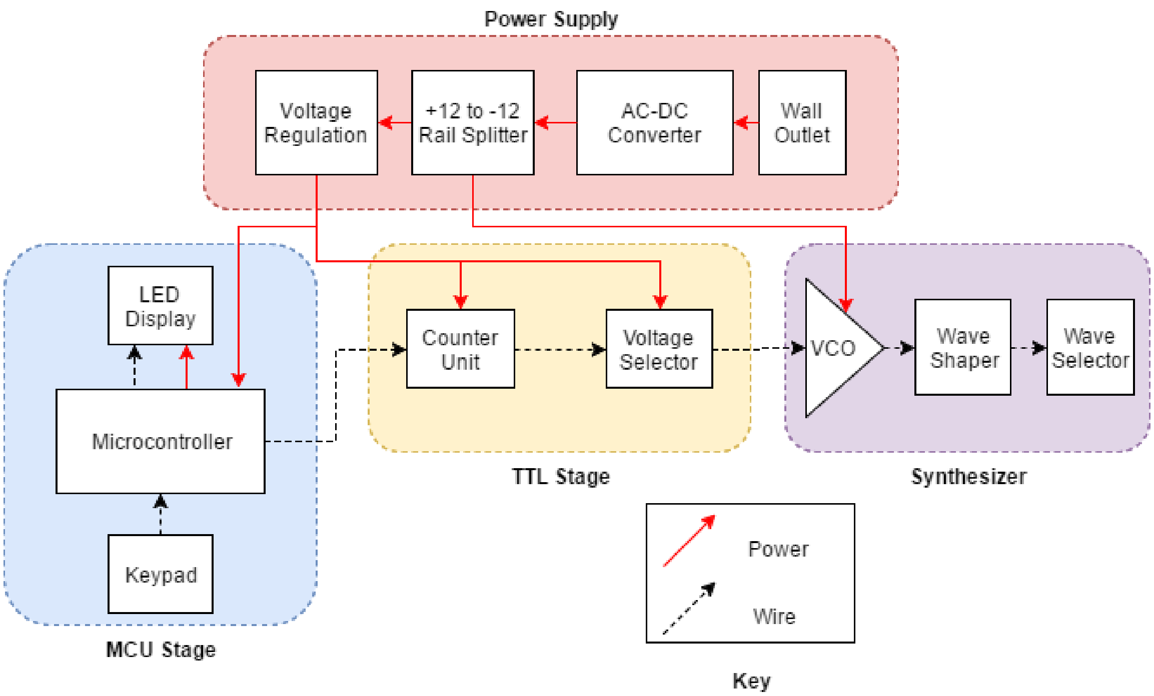

Figure 1: Block Diagram for Overall System

In Figure 2 below, we present a rough artist's rendition of what the physical device might look like upon completion. Because we do not have the sliders or buttons in hand, it does not make sense to design a physical layout with precise dimensions just yet. This figure may prove useful however in terms of getting a feel for what the instrument might look like.

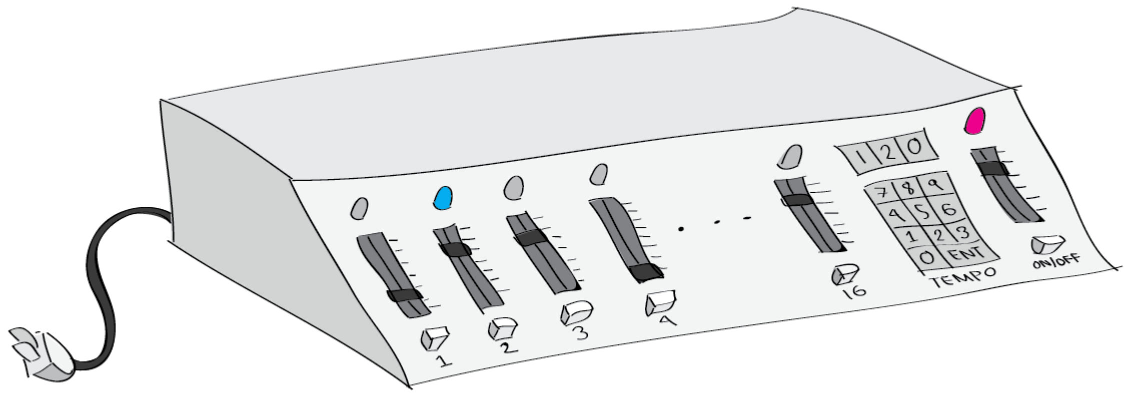

Figure 2: Artist Rendition

2.1 MCU Stage

The Microcontoller Unit (MCU) stage manages the tempo that will be output by the unit, and allows an interface for the user to input and see the beats-per-minute value that they are typing in. The MCU stage requires 3.3 V that will be supplied by the power supply.

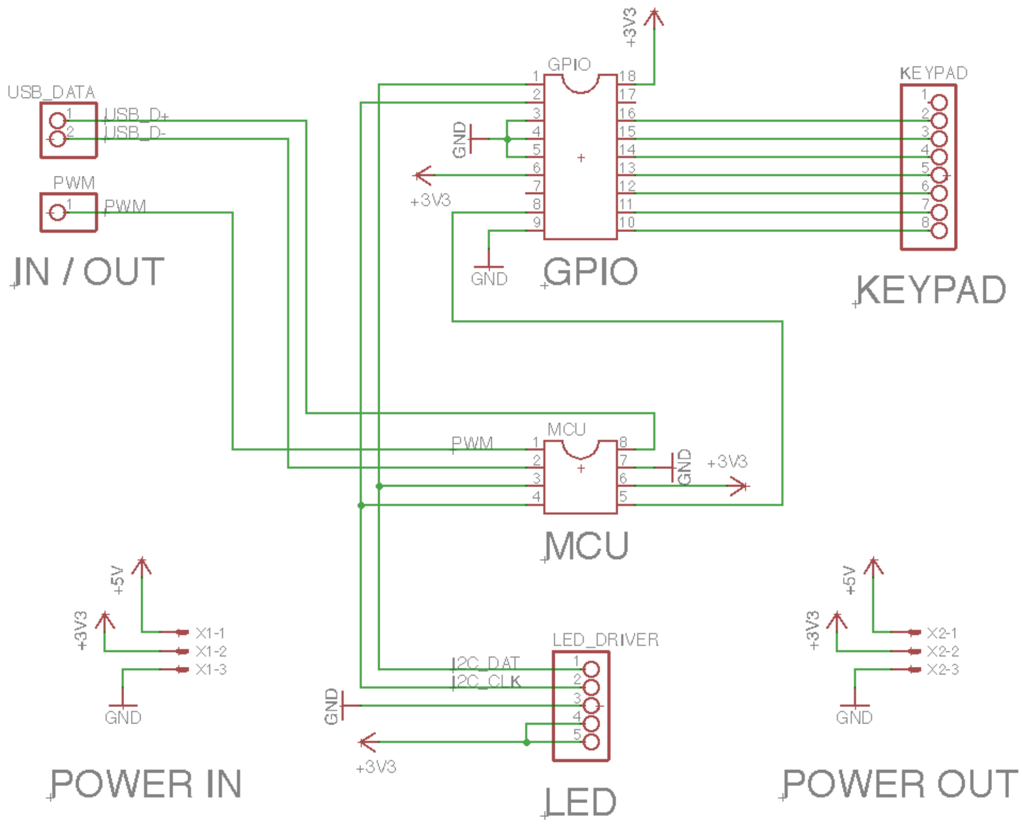

Figure 3: MCU Stage Eagle Schematic

2.1.1 Microcontroller

The microcontroller we have chosen is an LPC810 ARM MCU. It will handle converting the analog input from the keypad to a digital Pulse Width Modulation (PWM) wave that will then drive the clock for the TTL stage. It will also send analog output via Inter-Integrated Circuit (I^2C) to the 7-segment display so the user is able to see the current beats-per-minute and receive analog input via I^2C from the keypad for the button press.

Requirements	Verification
1. Must be able to convert a user input BPM ranging from 1 to 9999 into a PWM signal measured in Hz within 200ms.	1. Use a precision stop-watch to verify that once the pound key is pressed, a PWM signal begins in under 200ms.
2. Must output the user defined tempo to within ± 1%.	1. Input a tempo into the MCU unit. 2. Connect the PWM output pin to an oscilloscope. 3. Observe the output wave and verify the frequency within 1% of the input tempo.

2.1.2 Keypad

The keypad will allow the user to type in their desired beats-per-minute to control the ARM MCU. The $*$ and # keys correspond to clearing the tempo display and starting a new tempo. It is connected to the MCU via I^2C bus.

Requirements	Verification
1. MCU must be able to successfully read keypad interrupts across I^2C bus communication line.	1. Loop read statements in a while loop using software on the MCU. 2. Use an I^2C decoder on an oscilloscope to verify the read is executed correctly.

2.1.3 LED Display

The LED display will allow the user to know what the current beats-per-minute value of the overall system is. It should be updated every time the user changes the beats-per-minute value via the keypad. It is connected to the MCU via a Light Emitting Diode (LED) driver and I^2C bus.

Requirements	Verification
1. MCU must be able to successfully write to the LED driver across I^2C bus communication line.	1. Loop write statements in a while loop using software on the MCU. 2. Use an I^2C decoder on an oscilloscope to verify the write is executed correctly.

2.2 The TTL Stage

The TTL stage shown in Figure 4, and power requirements shown in Figure 5.

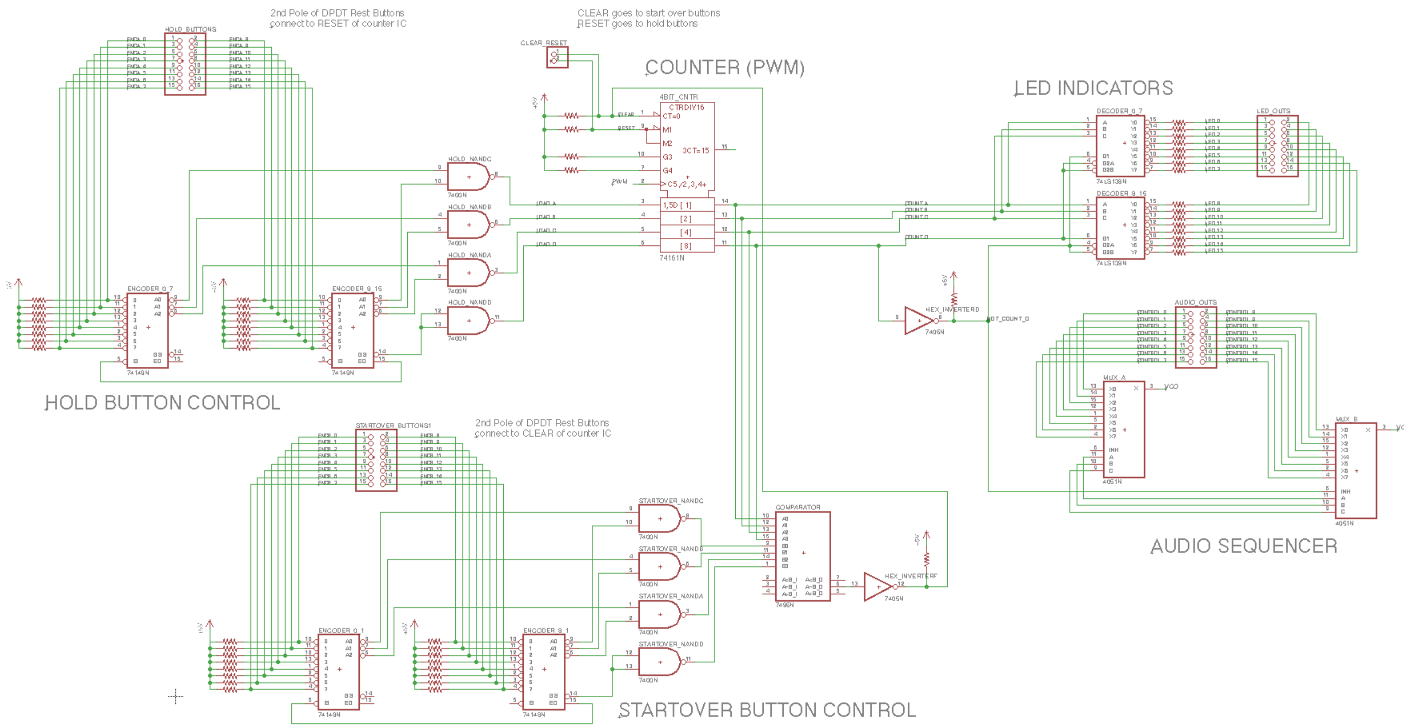

Figure 4: TTL Stage Schematic

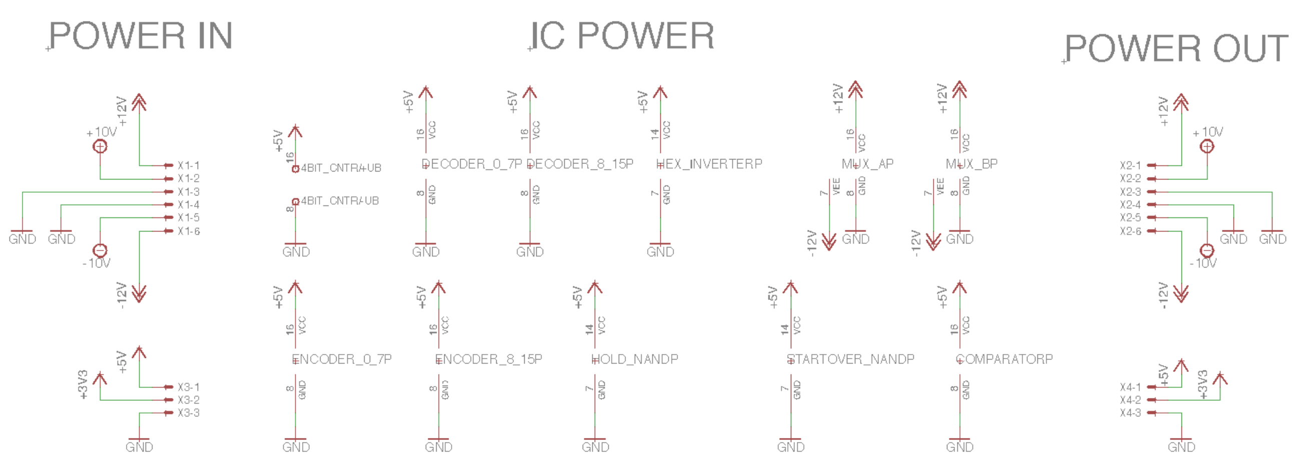

Figure 5: TTL Stage Power Schematic

The TTL stage consists of the counter, the LED display, the analog MUX, the hold and the startover sub-circuits, Figure 6.

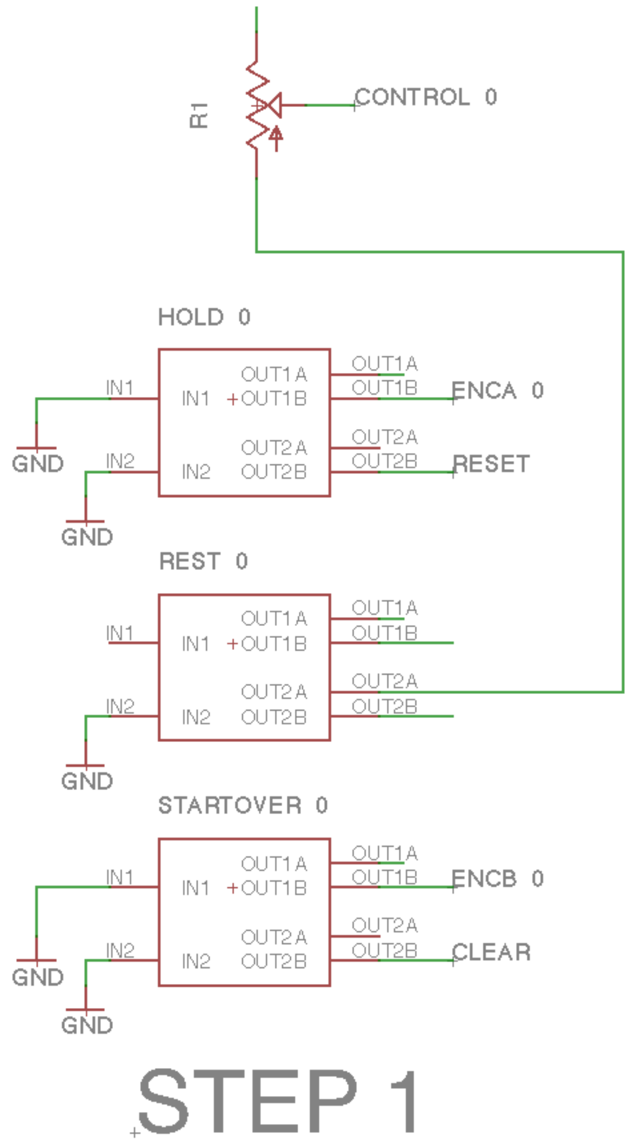

Figure 6: Single Step Schematic for Hold and Startover Sub-Circuit

The counter is a TTL counter Integrated Chip (IC) and it will be driven by the PWM signal generated by the MCU. The LED displays will be driven by decoder IC's which are fed by the counter. The analog MUX will be made to select from sixteen steps by the counter and pass through one potentiometer set voltage to the VCO. The hold sub-circuit will be driven by priority-line encoder IC's controlled by pushbutton switches. Similarly, the startover sub-circuit will be driven by a second priority-line encoder IC controlled by a second set of pushbutton switches.

However, since each sub-circuit is primarily one TTL chip, and because no verification can be done on a single chip, the TTL stage should be considered one overall unit as far as verification goes.

Requirements	Verification
1. Button press must perform its task within 50 ms.	1. Use a precision stop-watch to ensure the desired outcome based the button pushed is met within 50ms.
2. Analog MUX must pass through voltages from input to output to within 1%.	1. Use a DC Power Supply to connect a known, precise voltage to an analog MUX input. 2. Inhibit the MUX from switching channels. 3. Measure the output voltage using a multimeter to verify it is within 1% of set voltage.
3. Potentiometers on the analog mux inputs are able to create a voltage range from 0 to 5 V.	1. Connect multimeter across the potentiometers. 2. Sweep the potentiometers resistance and ensure the multimeter displays voltages from 0 to 5V.

2.3 Synthesizer

The synthesizer stage consists of the linear-to-exponential voltage controlled oscillator followed by a waveshaper section. The user will be able to select a sawtooth, square or triangle waveform for the output.

2.3.1 VCO

The VCO is a linear-to-exponential oscillator that allows us to map frequencies from 20Hz to 20kHz from a voltage that does not have to sweep through the same values. The VCO also will allow the user to feed additional smaller signal and lower frequencies oscillations on top of the voltage controlled note in order to provide frequency modulation, sometimes called vibrato in the musical world [6].

The VCO, Figure 7, consists of a control and octave per volt circuit, an exponentiator that allows us to take a linear voltage and create an exponential current, and then from that current we can generate a 0 to 5 V sawtooth with another circuit.

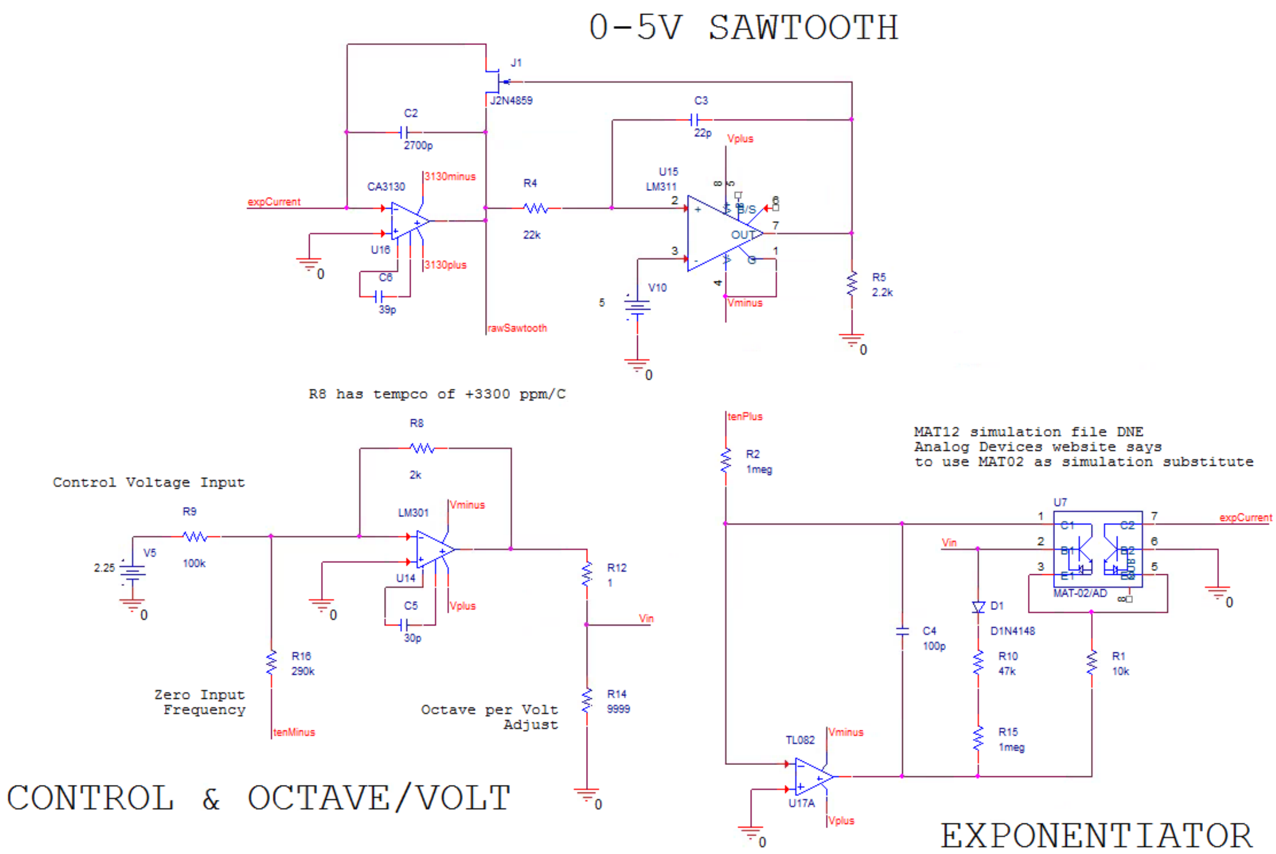

Figure 7: VCO Stage Schematic

Figure 8 shows an OrCad simulation of Figure 7. Voltage is plotted on the y axis, and time is along the x axis. As shown in Figure 8, the peak to peak voltage is 5 V.

Figure 8: OrCad Simulation of Voltage Controlled Oscillator Sawtooth Wave from 0 to 5 V

Requirements and verification for the VCO stage are shown below.

Requirements	Verification
1. Must achieve one octave of frequency change within 20 cents of accuracy for every one volt of control input.	1. Start from 0 volts, measure the frequency on the oscilloscope. 2. Change the input to 1 volt and measure the new frequency to ensure it has doubled its previous value to within 20 cents. 3. Iterate this procedure up to 5 volts control input.
2. The sawtooth generator within the oscillator must have a reset time of less than 1% the duration of its period.	1. Hook up the sawtooth generator output to the oscilloscope. Start with the control voltage at 0 V. 2. Measure the reset time using an oscilloscope to verify that it is less than 1% of the duration of the period for that frequency. 3. Iterate this procedure up to 5 V in 1 V increments and ensure the reset time is less than 1% of the duration of the period for that frequency.

2.3.2 Waveshaper

The waveshaper. Figure 9, takes the sawtooth output from the VCO and converts it into three different wave shapes, those being sawtooth, square and triangle. This will allow for different effects that the user will hear.

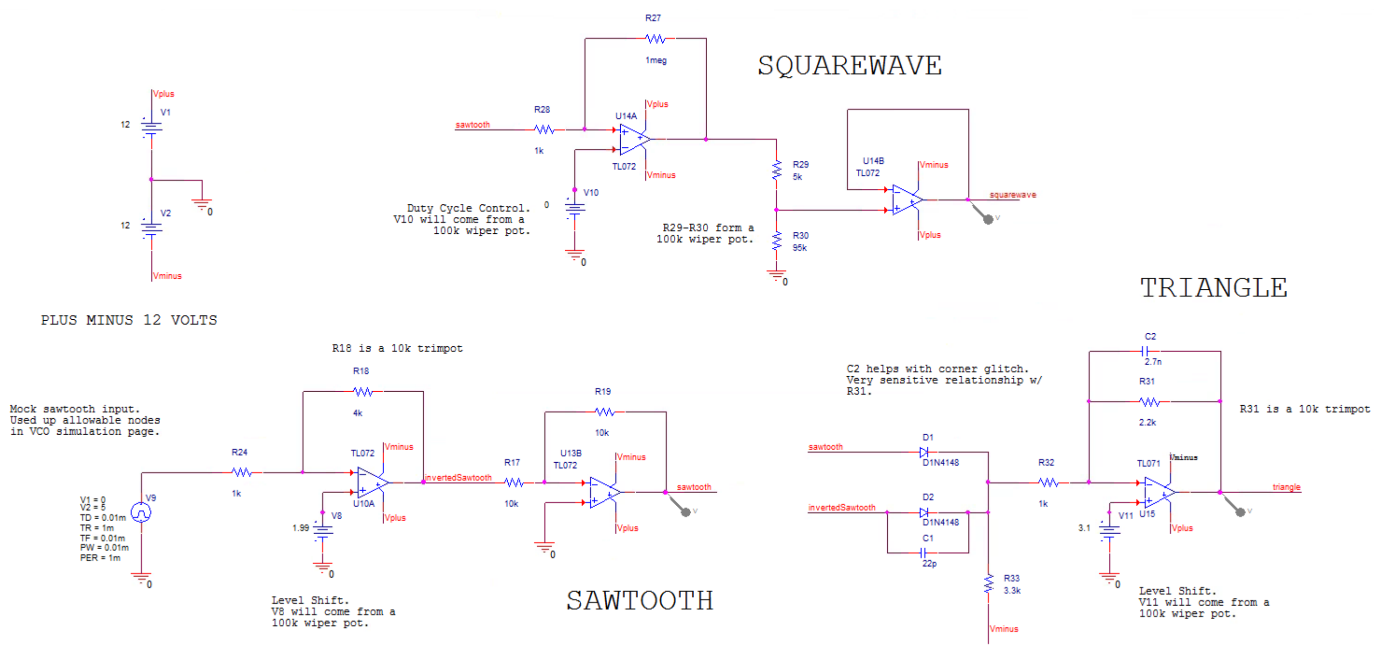

Figure 9: Waveshaper Schematic

The sawtooth generator takes our 0 to 5 V sawtooth, and amplifies it to a peak to peak voltage of 20 V. An OrCad simulation of this is shown in Figure 10. Voltage is plotted on the y axis, and time along the x axis. As can be seen in Figure 10, the peak to peak voltage is now 20 V.

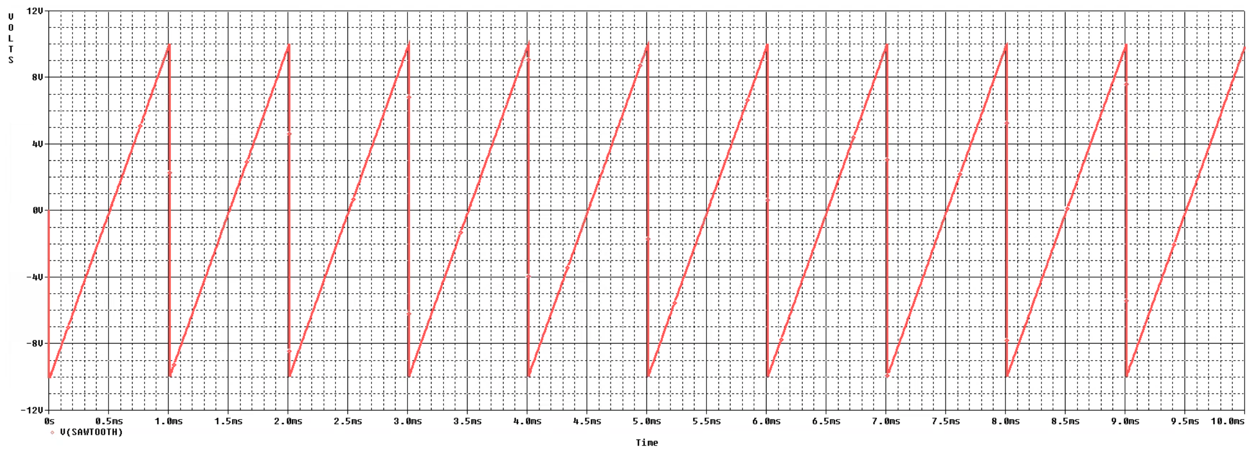

Figure 10: OrCad Simulation of Waveshaper Output for Sawtooth Wave

The square wave generator consists of a comparator and amplifier that will take our sawtooth wave, and compare it to the high and low peak voltage values. It will then force it to the high voltage, and then the low, creating a square wave. An OrCad simulation of this is shown in Figure 11. Voltage is plotted along the y axis and time along the x-axis. The peak to peak voltage is also 20 V.

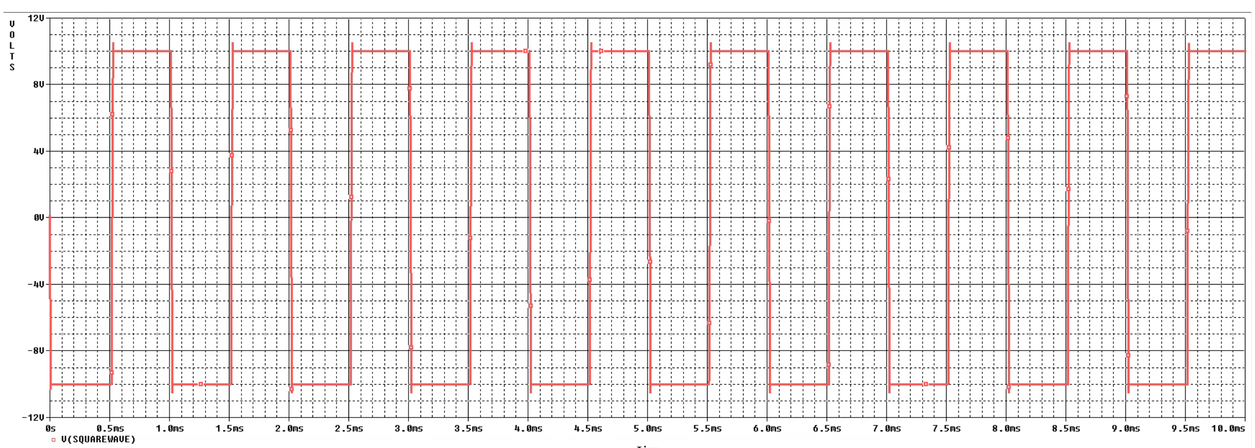

Figure 11: OrCad Simulation of Waveshaper Output for Square Wave

The triangle wave generator consists and inverter and a comparator. When the sawtooth input is positive voltage, it outputs it the same, but when it is negative, it inverts the signal to create a triangle shape. An OrCad simulation is shown in Figure 12. Voltage is plotted along the y axis and time along the x axis. The peak to peak voltage is 20 V.

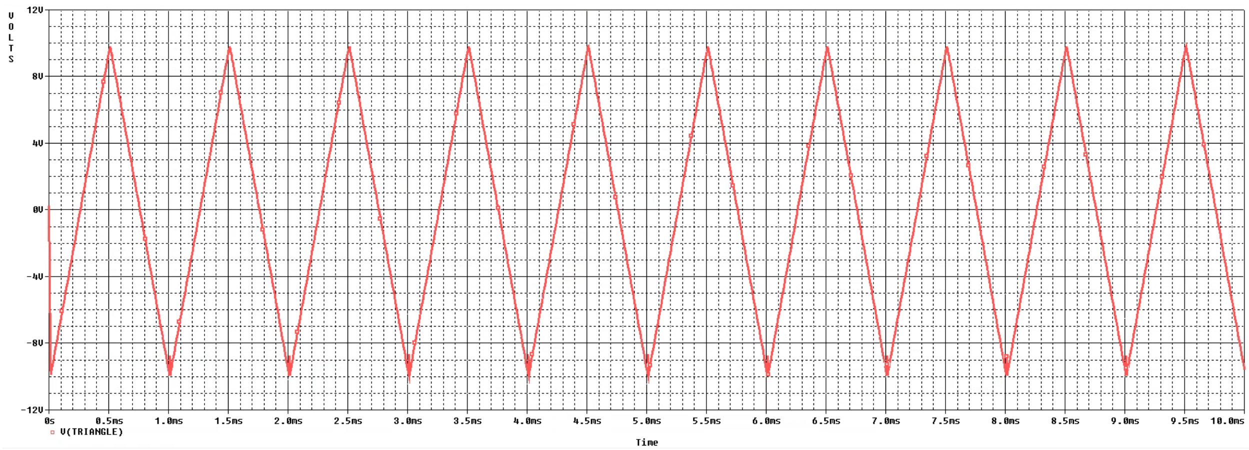

Figure 12: OrCad Simulation of Waveshaper Otuput for Triangle Wave

It is also important to note that in all the simulations that the frequency for each waveshape does not change. This is our goal, and allows us to create different types of sounds and tones form a single sawtooth generated by our VCO. This will give the user more freedom to play with sounds to achieve what they want.

Requirements for the wavehshaper section are shown below.

Requirements	Verification
1. The peak-to-peak value of each of the three waveforms must be 20 V to within 10%.	1. Set control voltage to 1 V. 2. Connect the oscilloscope to the output of one of the waveshapers. 3. Measure the peak-to-peak voltage and ensure it is within 10% of 20 V. 4. Repeat 2. and 3. for the other two waveshapers.
2. Any trapezoidal component to the squarewave time-domain signal must be less than 1% of the total duration of the signal period.	1. Set control voltage to 0 V. 2. Connect the oscilloscope to the squarewave waveshaper. 3. Measure the time-axis underneath the reset slopes to verify that it is less than 1% of the total time. 4. Repeat 2. and 3. for two frequencies per octave covering the VCO's range, up to 5 V in 1 V increments.
3. The crossover portion of the rectified sawtooth which forms the triangle waveform must be triangular to within 20%.	1. Set control voltage to 0 V. 2. Connect the oscilloscope to the output of the triangle waveshaper. 3. Take the derivative of the oscilloscope triangle waveform and measure the time duration of the cross-over portion of the signal to ensure that it is within 20% of the total time-duration of the period. 4. Repeat 3. covering the VCO's range, up to 5 V in 1 V increments.

2.4 Power Supply

The power supply, Figure 13 is required to keep the system running. Each of the previous three sections has different voltage requirements, so a power supply is needed to regulate the power to each stage. In particular, the VCO stage will require ± 12 V and ± 10 V, the TTL stage requires 5 V, and the MCU stage requires 3.3 V. We are designing this system to be plug and play, so we will be utilizing an off the shelf AC-DC converter to safely and effectively give us a 24 V DC signal that we can then regulate to the necessary voltages. Total power consumption and power consumption for each stage can be seen in Table 1.

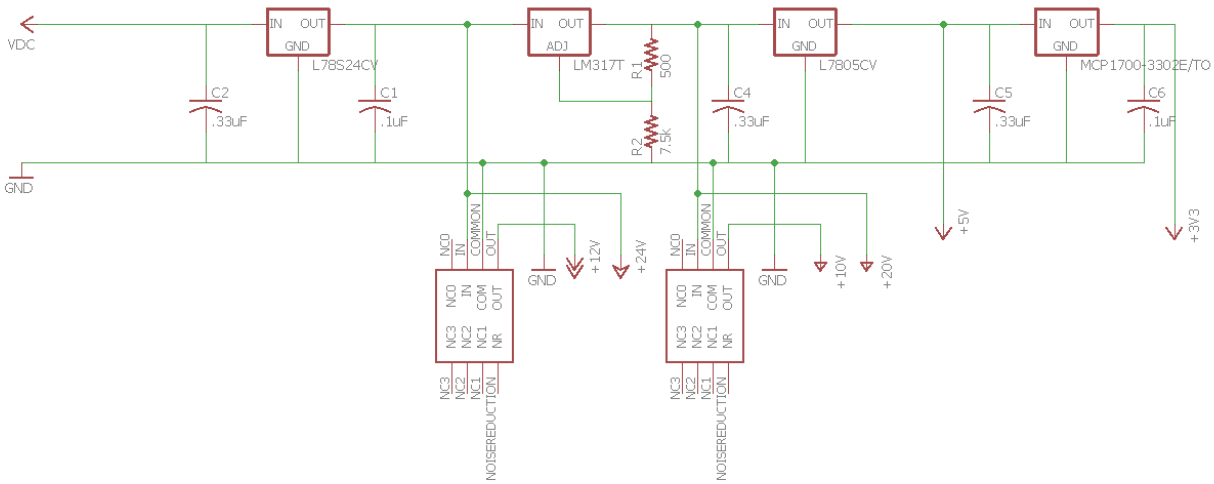

Figure 13: Power Supply Schematic

Table 1: Power Consumption for Each Stage

Part	Max Current Draw (mA)	Max Voltage (V)	Quantity	Max Power Consumption (mW)
MCU				
LPC810	3.3	3.3	1	10.89
LEDs	30	2	2	120
Total	33.3			130.89
TTL				
Counter	0.08	5	1	0.4
MUX	0.1	5	2	1
Decoder	0.008	5	2	320
Inverter	0.02	5	1	0.1
Encoder	0.08	5	4	1.6
NAND	0.02	5	2	0.2
Comparator	0.008	5	1	0.04
LEDs	30	2	2	120
Total	30.316			443.34
VCO				
Summing Op-Amp	1.8	10	1	500
Matched NPNs	20	40	1	800
Tail Op-Amp	1.4	36		50.4
JFET for Switching	50	15	1	1.8
Integrating Op-Amp	10	15	1	150
Comparator Op-Amp	7.5	30	1	225
Dual Op-Amp	1.4	36	1	50.4
Input Op-Amp	1.8	36	1	64.8
Emitter follower NPN	200	40	1	625
Emitter followre PNP	200	40	1	625
Buffer Op-Amp	1.4	36	1	50.4
Op-Amp	1.8	36	1	64.8
Total	497.1			3207.6
Power Supply				
3.3 V Regulator	33.3	1.7	1	0.644
5 V Regulator	30.316	7	1	212.212
20 V Regulator	497.1	4	1	1988.4
24 V Regulator	497.1	16	1	7953.6
Rail Splitter	20	40	2	1600
Total	1077.816			11.754856
Overall System				
Total	560.716			3793.584856

2.4.1 AC-DC Converter

This converter will take 120 VAC from a standard wall outlet and convert it to a nominal 24 VDC. We will then use an additional voltage regulator to regulate the voltage to a near constant 24 V. This whole system will allow the device to be plug and play, so it is usable in any space where standard wall outlets are available. According to the power calculation, the total current drawn by all three stages should be 1A, so a 24 V 1.5A AC-DC power supply should be more than enough to supply our system.

Requirements	Verification
1. Regulated output voltage must be 24 V ± 5%.	1. Connect multimeter to output pin of the 24 V Regulator 2. Use a constant current electronic load to apply a current load of 500mA. 3. Ensure the regulated voltage stay within 5% of 24 V.

2.4.2 Rail Splitter

The rail splitter section will allow us to take the regulated 24 V signal from the previous section, and from it create two different bipolar voltages for the VCO stage. A rail splitter IC will first take the 24 V signal and split it to ±12 V. We will also need another voltage regulator, a LM317T with the output adjusted to 20 V, which will be fed into another rail splitter that will give us our ±10 V. The LM317T will provide a constant 1.25 V reference between the output and adjustment terminal. We can use two resistor to create a voltage divider that will then allow us to adjust the output voltage of the LM317T [7]. Equation 1 shows the relationship between R_1, R_2 and V_{out}

$$V_{out} = 1.25(1 + \frac{R_2}{R_1}) \tag{1}$$

By choosing R_1 to be 500Ω and V_{out} to be 20 V and using Equation 1, we can see that R_2 should be $7.5k\Omega$. According to the power calculations, the current drawn by the VCO stage which will utilize the 24 V and 20 V regulators should draw about 500mA. Both of these regulators have a maximum current draw of 1.5A, so they will be able to handle the maximum current draw.

Requirements	Verification
1. 20 V Regulator should supply 20 V $\pm 5\%$	1. Connect multimeter to the output pin of the LM317T. 2. Use a constant current electronic load to apply a current load of 500 mA. 3. Ensure the mulitmeter rating stays within 5% of 20 V.
2. Rail splitter chips must split input voltage to one half $\pm 5\%$.	1. Connect multimeter to the output pin of the chip 2. Connect another multimeter to the input pin of the chip 3. Ensure the output voltage is within 5% of the input voltage.

2.4.3 Voltage Regulator

For the voltage regulation we will utilize a 3.3 V IC and a 5 V IC to supply the correct voltages to the MCU stage and the TTL stage, respectively. We will also utilize two Zener diodes in the VCO stage to regulate the +12 to +9.1 V and the -12 to -3.9 V as a certain op-amp will need this voltage requirement, while others will require the ± 12 V. The 5 V voltage regulator should be able to effectively regulate the +12 V down to 5 V. From the 5 V we will then utilize a 3.3 V regulator to supply the voltage to our MCU unit. According to the power calculations the current draw for the MCU stage is about 33mA, and the 3.3 V IC can handle maximum current draw of 250mA. The TTL stage has a maximum current draw of 30mA, and the 5 V IC can handle maximum current draw of 1.5A.

Requirements	Verification
1. 3.3 V Regulator should supply 3.3 V $\pm$ 5%.	1. Connect the output of the multimeter to the output pin of the 3.3 V IC. 2. Use a constant current electronic load to apply a current load of 35 mA. 3. Ensure the multimeter reading stays within 5% of 3.3 V.
2. 5 V Regulator should supply 5 V $\pm$ 5%	1. Connect the output of the multimeter to the output pin of the 5 V IC. 2. Use a constant current electronic load to apply a current load of 35 mA. 3. Ensure the multimeter reading stays within 5% of 3.3 V.
3. Zener diodes regulate voltage to 3.9 V $\pm$ 5% and 9.1 V $\pm$ 5%.	1. Connect one multimeter across the 3.9 V Zener, and another multimeter across the 9.1 V Zener. 2. Ensure that the first multimeter reads within 5% of 3.9 V, and the second multimeter reads within 5% of 9.1 V.

2.5 Software

2.5.1 BPM to Frequency Unit Conversion Process

Tempo in music is most often given in units of beats per minute, or BPM for short. To convert BPM to frequency in Hz, we perform the following dimensional analysis

$$1\frac{beat}{minute} * \frac{1\ minute}{60\ seconds} = 1\frac{beat}{second} = 1\frac{number}{second} = 1\ Hz \tag{2}$$

We can make the above formula even simpler and represent the conversion as follows

$$[BPM] \xRightarrow{\text{Divide by 60}} [Hz] \tag{3}$$

Putting it this way, 60 BPM corresponds to 1 Hz. Likewise, 120 BPM corresponds to 2 Hz and so and so forth.

Because microprocessor clock speeds work in units of Hz, while our system accepts input in units of BPM, we will need to convert from BPM to Hz in software before generating the PWM output driving the step sequencer.

2.5.2 PWM Calculation Based on a 12 MHz Clock

The microcontroller that we will be using is the LPC810, which has a clock speed of 12MHz. To understand how this works, if we wait 12 million "ticks" of LPC810 system clock, then we will have waited one second.

PWM on the LPC810 works in the following way. Two counters are created in software. Each counter has an associated event which occurs at the PWM output upon a match. In our case, the first event will be to set the output high and it will occur when the first counter has counted the amount of system "ticks" corresponding to the user desired tempo. The second event will be to the set the output low somewhere in the middle of the desired tempo. This will be our duty-cycle. Since the PWM signal is sent to a rising-edge triggered TTL counter IC, we don't care too much about when the output goes low within the overall period just as long as it isn't sooner than the amount of time it takes for the TTL counter to process a rising edge. So just to be safe, we'll set this second event to always occur at the midpoint of the PWM period.

For a visual understanding of how PWM is set up on the LPC810, we provide the following figure. In our case, we will simply set event 0 as the start-over point of our PWM signal and event 1 as the duty-cycle point of the signal.

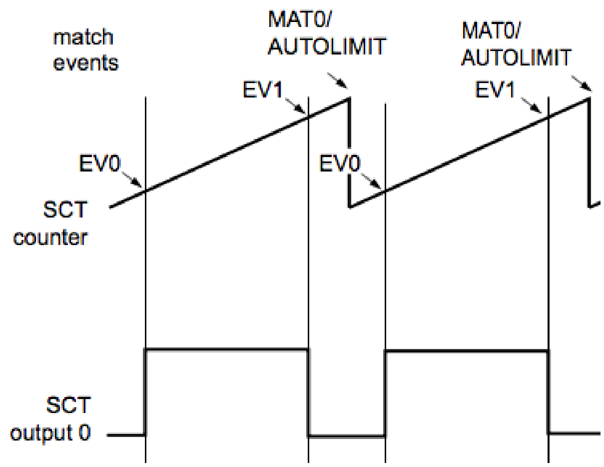

Figure 14: Visual Representation of Two Counters Driving a PWM Signal on the LPC810

If, for example, the user has input a desired step-sequencer tempo of 120 BPM, we will need to do the following. First, using the conversion process explained above, 120 BPM corresponds to 2 Hz. A frequency of 2 Hz is equal to a period of 0.5 seconds. Now if 12 million ticks of the 12MHz clock takes one second, then 6 million ticks will take 0.5 seconds. Therefore, for a user desired tempo of 120 BPM, we will need to set the pulse width modulation on the LPC810 to start over every 6 million ticks of its system clock. This process is generalized as follows.

$$\text{User Desired BPM} \xRightarrow{\text{divide by 60}} \text{Frequency} \xRightarrow{\text{system clock/frequency}} \text{PWM Period Count} \tag{4}$$

And for duty-cycle, we simply divide our calculated PWM period by two.

2.5.3 Scanning and I^2C

The keypad we are using for our user tempo input has 12 keys, just a like conventional phone keypad. Ideally, we would use 12 inputs on our microcontroller and call it a day. However, the LPC810, having only six pins that aren't power or ground, is severely limited in terms of input and output availability. In fact, with two pins taken up for USB programming and a third pin used for the PWM output, we only have two pins left. So, clearly the 12 keys on the number keypad will not fit.

The solution to this problem is two-fold. First, we boil the 12 inputs on the keypad down to 7 using a matrix technique which will be described shortly. Secondly, we connect these boiled-down, 7 keypad signals to a separate general purpose input/output (GPIO) expander chip which uses just two lines to communicate with the MCU over a protocol called inter-inter- chip, or I^2C as it is more commonly known.

Our numeric keypad can be described as a matrix consisting of four rows and three columns as shown below.

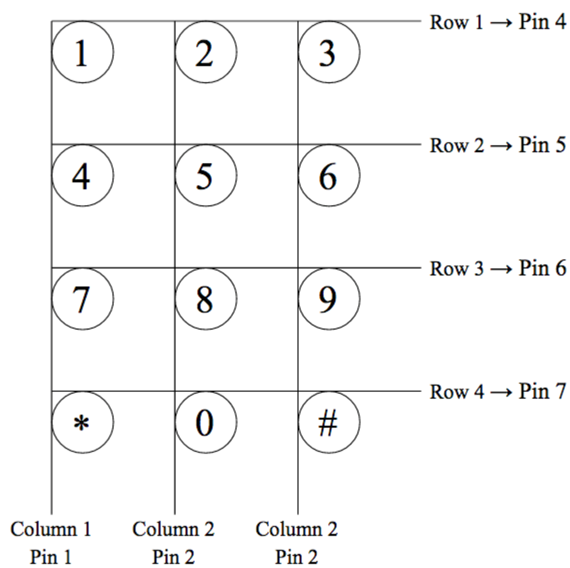

Figure 15: Matrix Representation of Numeric Keypad

Looking at Figure 15, we see that each column of the keypad connects to a unique pin on the GPIO chip. Likewise, each row connects to a unique pin on the GPIO chip.

Let's look at what the keypad is doing from a schematic perspective.

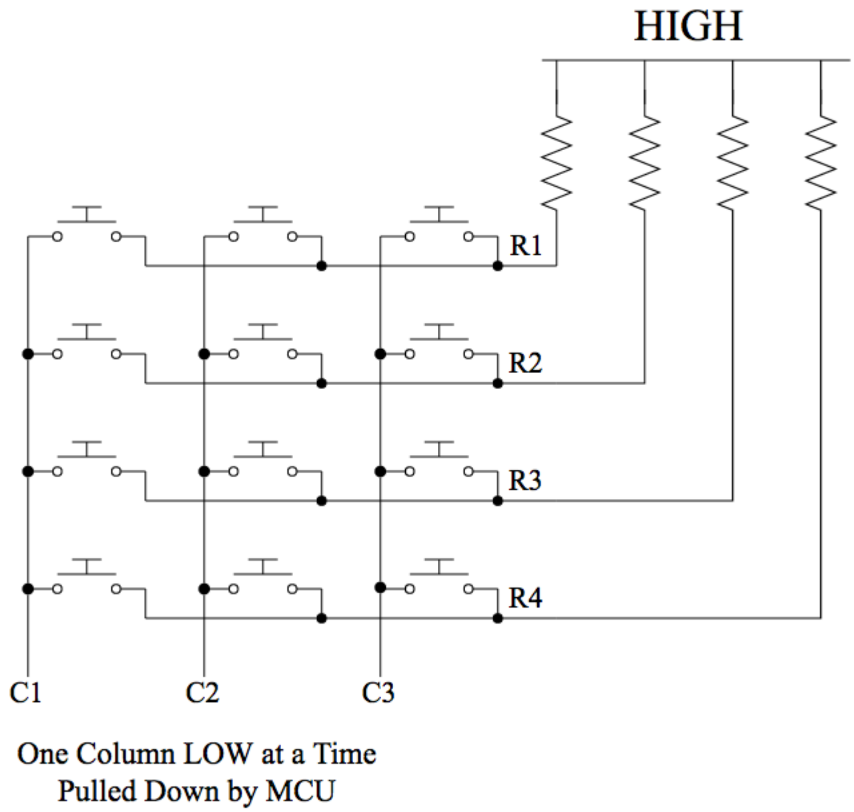

Figure 16: Matrix Schematic of Numeric Keypad

As we can see from Figure 16, each number corresponds to a pushbutton switch. The columns are connected to each other and likewise, so are the rows. If, for example, the number 8 switch is pushed, then row 3 will be electrically connected to column 2.

The GPIO expander has an interrupt feature which will trigger an interrupt any time an interrupt-enabled pin changes from a programmable value. For example, if pin 8 is set in software to be high then the GPIO chip will trigger an interrupt if pin 8 is detected to be low.

In software we will cycle through making one column low while leaving the other two high and then check whether any of the rows have changed from high to low. Looking at Figure 16, the rows are all tied high with pull-up resistors. So, if for example, column 2 is forced low and the user pushes key 8, then row 3 will connect to ground, forcing it low. The GPIO chip will then detect an interrupt. Once the interrupt is fired, we can read the pins on the GPIO chip, in order to find out that key 8 has been pushed, using the I^2C protocol, which will be described now.

I^2C is a form of serial communication rather than parallel communication. Essentially, it uses time rather than space to communicate. So if there are eight bits of information to send from one chip to another and there isn't enough space, i.e. enough pins on the two chips, to send the eight bits in parallel, then I^2C will handle the transfer of bits serially, in time, using just two lines of communication rather than eight.

The two lines that I^2C uses are for serial data, SDA, and the serial clock, SCL. If two chips are communicating, then there is one master and one slave. The master controls the bus in all cases except when reading from the slave, during which time the slave must send the requested information. The I^2C protocol is well documented and if the reader wishes to learn more about the details, information is widely available. The main takeaway in terms of our project is that we can boil down the 7 pins from the keypad to just 2 I^2C lines on our MCU chip.

2.5.4 MCU Flowchart

Now that we have established how to generate a PWM signal at the correct tempo, and how to take the user input from the keypad, we can construct an overall flowchart for the MCU stage.

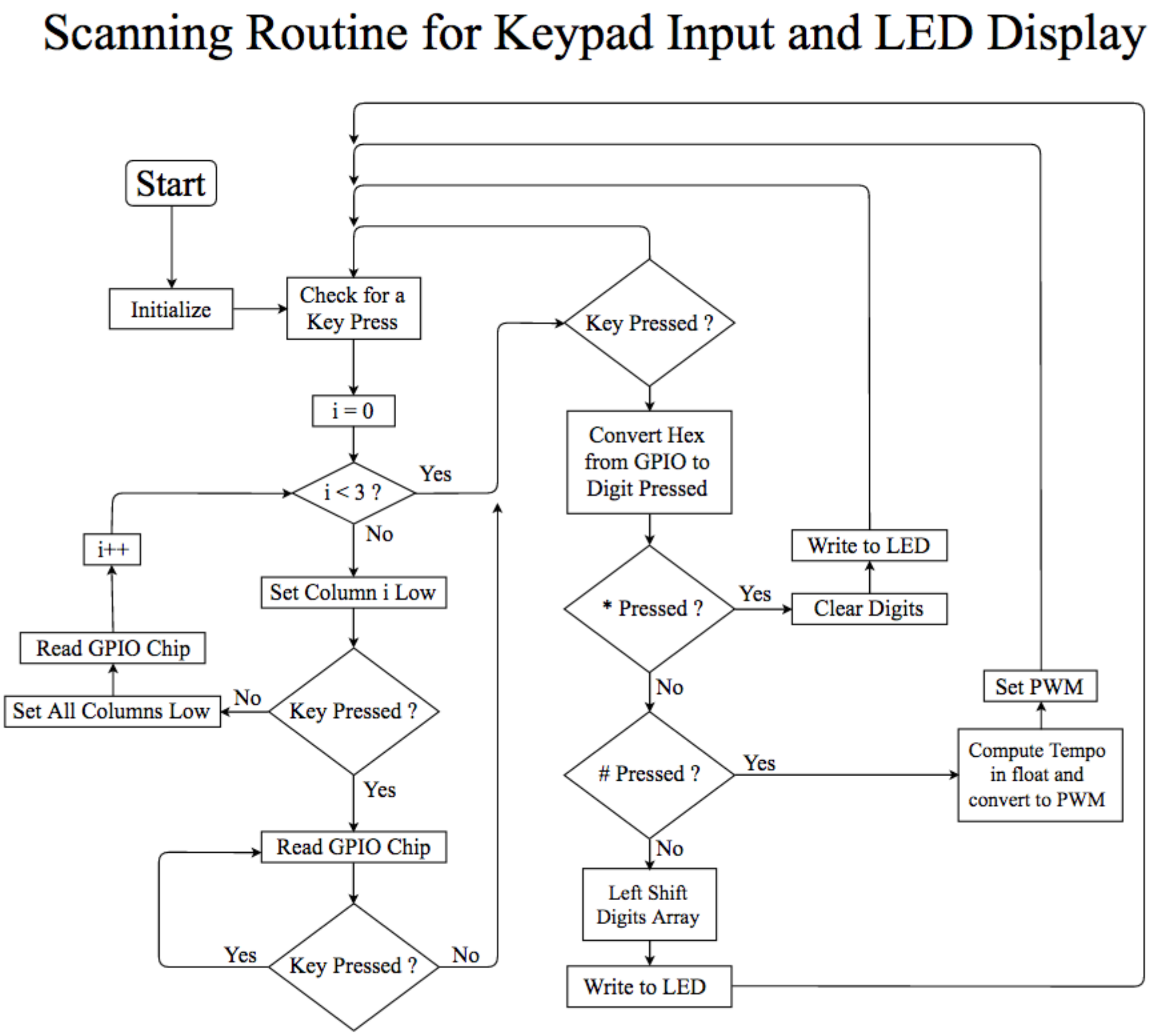

Figure 17: MCU Stage Software Flow Chart

2.6 Tolerance Analysis

This is our goal: to make a circuit that doubles the current every time the input voltage increases by one volt. A bipolar junction transistor converts linear voltage to exponential current, so naturally, we will start with the Bipolar Junction Transistor (BJT). As we will discover however, there are two obstacles to overcome. The first we shall call the temperature problem. The second hurdle is how to achieve an octave (or a doubling) of current for every one volt of input. Let's look at the temperature problem first.

2.6.1 The Analog VCO and its Temperature Problem

The collector current of an NPN tansistor as a function of its base-emitter voltage is given by

$$i_c = I_S e^{\frac{v_{BE}}{V_T}} \tag{5}$$

where I_S is a constant and V_T is proportional to the temperature T in kelvin. The graph for this equation is a simple exponential curve as shown below.

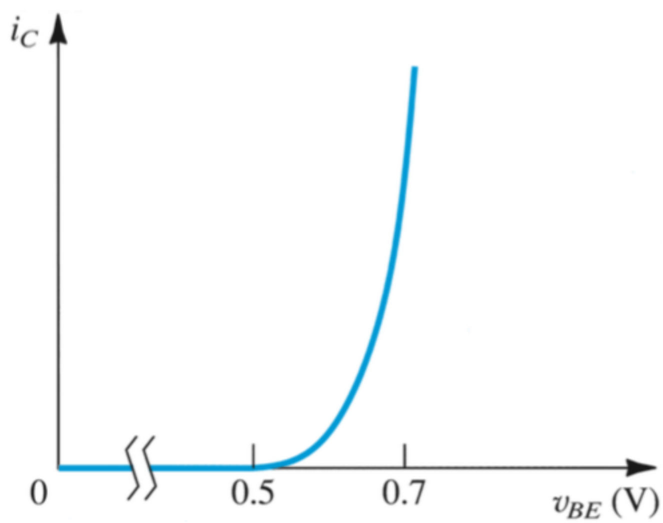

Figure 18: Exponential Collector Current for Linear Base-Emitter Voltage

For a fixed collector current, provided by some ideal current source, if the temperature changes and all other things remain constant, just by looking at Equation 5, we see that v_{BE} must drop. What this means graphically is that the curve given above will shift to the left with increasing temperature for a fixed collector current.

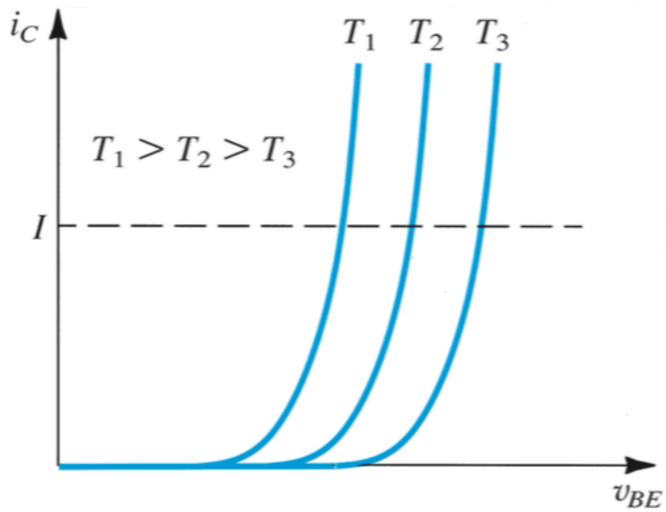

Figure 19: Effects of Temperature on Base-Emitter Voltage

Our goal with the VCO is to convert a linear voltage to an exponential current in order to drive a current-to-frequency oscillator. The reason we need linear to exponential conversion is because the audible range is from roughly 20 Hz to 20 kHz and so a direct linear mapping of voltage to frequency would not be practical.

If it weren't for this pesky temperature problem, a single NPN transistor would be the ideal choice for our linear-to-exponential converter. Our circuit would employ the simple common-emitter configuration as shown below.

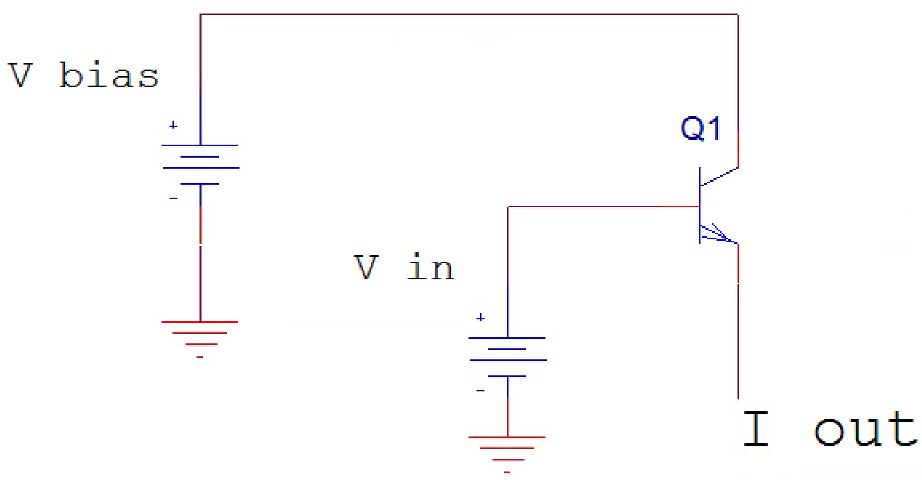

Figure 20: Simple Common-Emitter Based Linear Voltage to Exponential Current

Considering that there is negligible base current compared to the exponential collector current, we can safely say that the emitter current, I_{out}, is equal to the collector current. But alas, the temperature dependence of the base-emitter voltage prevents us from using this simple arrangement for our VCO circuit. To get around this problem, we can do the following.

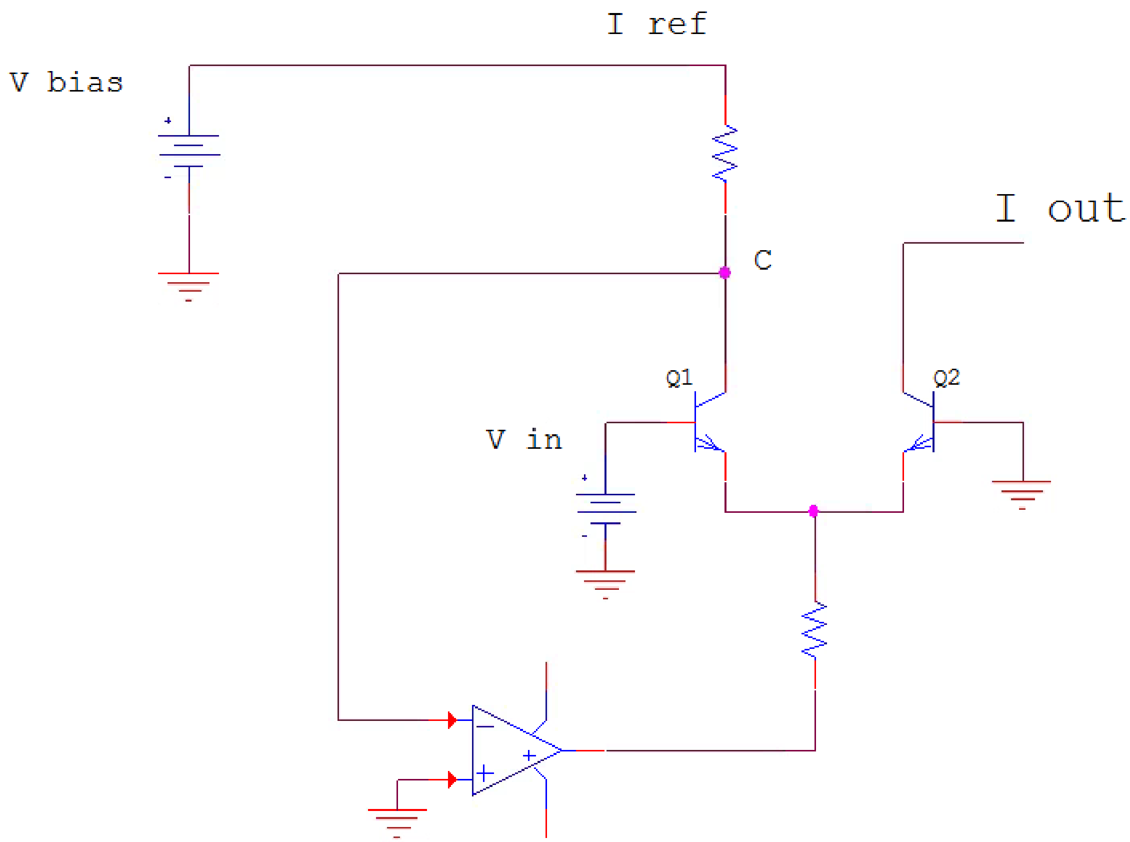

Figure 21: Solution to the Temperature Problem

At first glance, this second arrangement seems a lot more complicated than the simple common-emitter we had earlier. However, the analysis is straightforward. If we just look at the differential pair formed by Q1 and Q2, by KVL we have

$$V_{IN} - V_{BE1} + V_{BE2} = 0 \tag{6}$$

This can be rearranged to give

$$V_{BE1} - V_{BE2} = -V_{IN} \tag{7}$$

Using the exponential relationship between base-emitter voltage and collector current and the fact that the emitter current will essentially equal the collector current since there is very little base current, we have

$$I_{ref} = I_S e^{\frac{v_{BE1}}{V_T}} I_{out} = I_S e^{\frac{v_{BE2}}{V_T}} \tag{8}$$

Dividing these two equations and substituting for V_{IN}, we arrive at

$$I_{out} = I_{ref} e^{\frac{-V_{IN}}{V_T}} \tag{9}$$

So here we have the output current *linearly* related to the reference current and *exponentially* related to the input voltage. If we can get the reference current to remain constant, we will have a strictly exponential relationship between input voltage and output current.

To get the reference current constant, we use the op-amp as shown in Figure 21. This op-amp will do whatever it takes to make its inputs equal, which means that the voltage drop across Q1's collector resistor will remain constant and therefore so will the reference current.

Now looking at Equation 9, and with the ideal op-amp forcing I_{ref} to be constant, we have achieved a strictly exponential relationship between the input voltage and the output current.

Let's see how the temperature problem is taken care of. Now, if the temperature goes up for our fixed collector current through Q1, then as discussed earlier, its base-emitter voltage will decrease by the same amount. (The graph will shift left as shown above). Q2 however, doesn't have a nice, fixed reference current, and so with this rise in temperature, its collector current will increase. Now looking back at Q1, for its base-emitter voltage to drop, the tail voltage, V_{TAIL}, must rise if V_{IN} is fixed. This means that Q2's V_{BE} will rise, which will cause its current to decrease. This decrease in Q2's collector current will perfectly cancel the temperature related increase just described. In summary, Q2's collector current will increase due to a temperature rise and it will also decrease due to its base-emitter voltage decreasing and therefore the temperature problem is canceled out.

2.6.2 Achieving an Octave per Volt with the VCO

With the temperature problem out of the way, we can now focus on figuring out how to get an octave of current per one volt of input. An octave in dB is a doubling in linear terms. So what we need is a current that doubles itself for every one volt of input. With our updated schematic in mind, we can look at Equation 9 again.

For $I_{out} = I_{ref} * 2$, we need $e^{\frac{-V_{IN}}{V_T}} = 2$. Now, V_T is accepted to be around $26mV$, so we have $e^{\frac{-V_{IN}}{26}} = 2$. Ignoring the negative sign, which can be solved by simply inverting the input, and solving for V_{IN}, we have $V_{IN} = 26ln(2) = 18.02mV$. Therefore, we need the input voltage to increase by about $18mV$ for every doubling of the output current.

One volt of input is easily inverted and scaled down to $20mV$ with an op-amp based inverting amplifier having a gain of $\frac{2}{100}$. In other words $(1\ Volt\ Input * (\frac{2}{100}\ gain) = 20mV$. At the output of the inverting amplifier, we can then use a simple wiper potentiometer as a voltage divider to further fine-tune the $20mV$ down to the $18mV$ we need at the input of the linear-to-exponential converter circuit.

For the input schematic, we have the following.

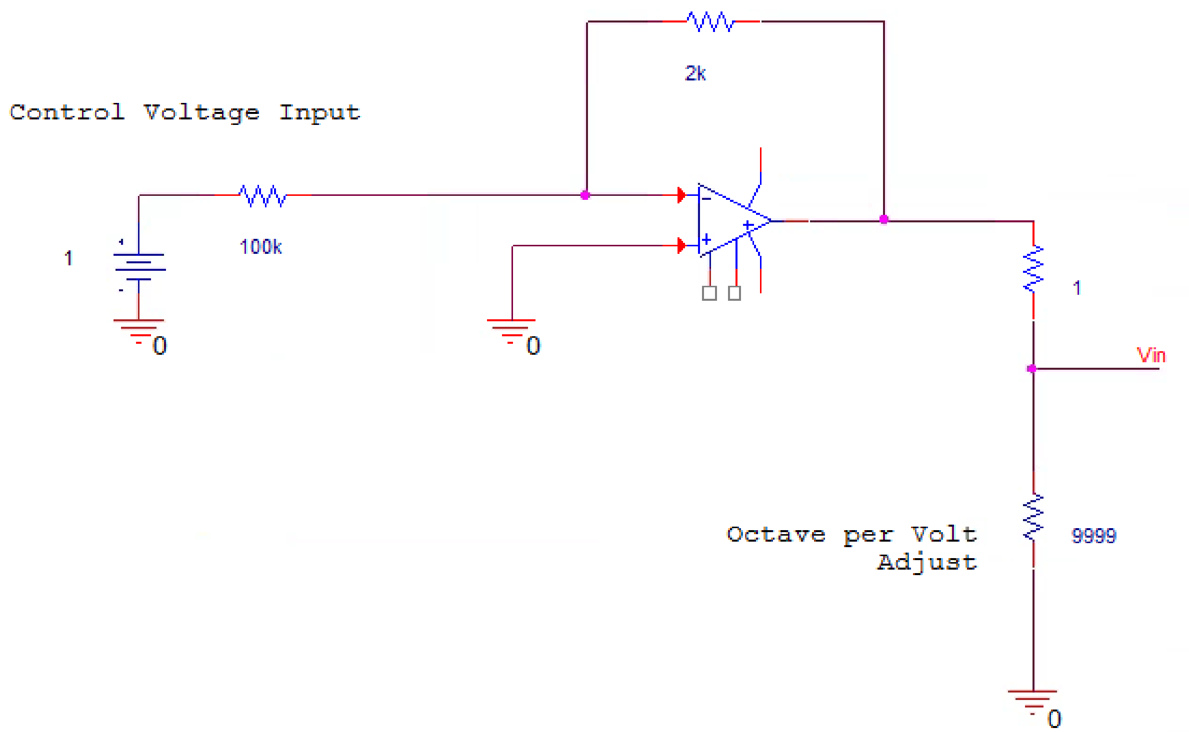

Figure 22: Scaling One Volt Down to Roughly $18mV$

Looking at Figure 22, we have a schematic representation of how we can scale one volt of input down to $18mV$ of output for the linear-to-exponential converter at the following stage (not shown).

There's just one more thing to take care of before diving into the actual VCO schematic. We need to make sure that with zero input voltage, we have a desired base frequency. Look at Equation 9 one more time, if $V_{IN} = 0$, then we have $I_{out} = I_{ref}$. A good base frequency is $60Hz$. So what we need to do is add a "zero-input" signal to the input stage of our converting amplifier in Figure 22. We have the following.

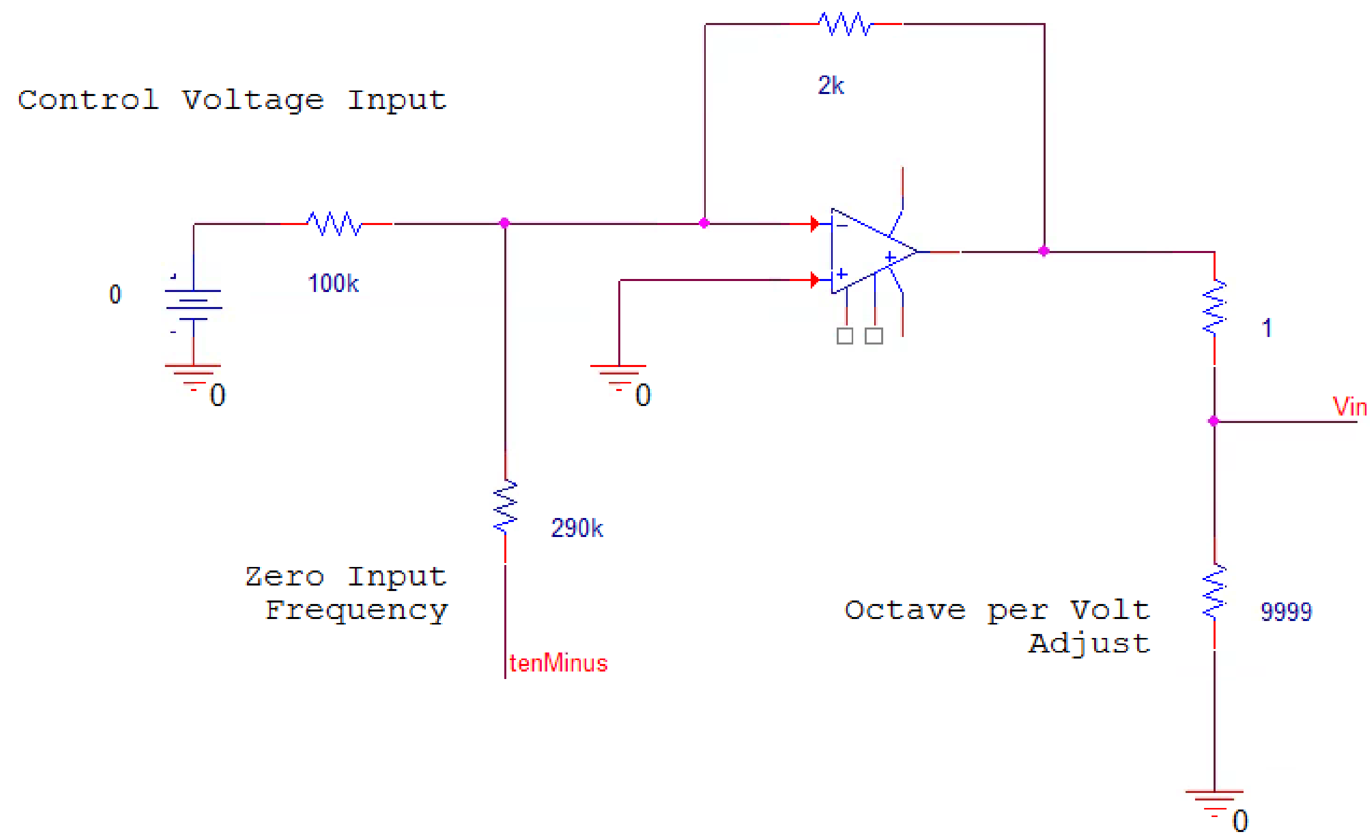

Figure 23: Setting the Base Frequency

The only difference between Figures 22 and 23 is an additional input voltage to the inverting amplifier. With more than one input, the inverting amplifier can now be officially called a summing amplifier. The resistor on the zero-input frequency line is a trimmer potentiometer. What we have to do is set zero volts at the control input and then set the zero-input frequency trimpot so that we have 60 Hz at the output of the oscillator circuit (not shown).

Having solved the temperature problem and having figured out how to achieve an octave of current (or a doubling) for every one volt of input, we are now ready to look at the three sub-circuits which make up the VCO.

2.6.3 Tolerance Analysis Regarding our VCO Schematic

There are three sub-circuit in our VCO. The first is the control input which handles the zero-frequency input as well as octave per volt adjustment. The second sub-circuit is the exponentiator which takes the voltage provided at the output of the control sub-circuit and converts it to an exponential current. The third sub-circuit is an integrator combined used in conjunction with a comparator to create a sawtooth signal from the current provided by the exponentiator sub-circuit.

Let's take a look at the control input sub-circuit.

2.6.4 The Control Input Sub-Circuit of the VCO

The following figure is of our actual input sub-circuit we are using in our design.

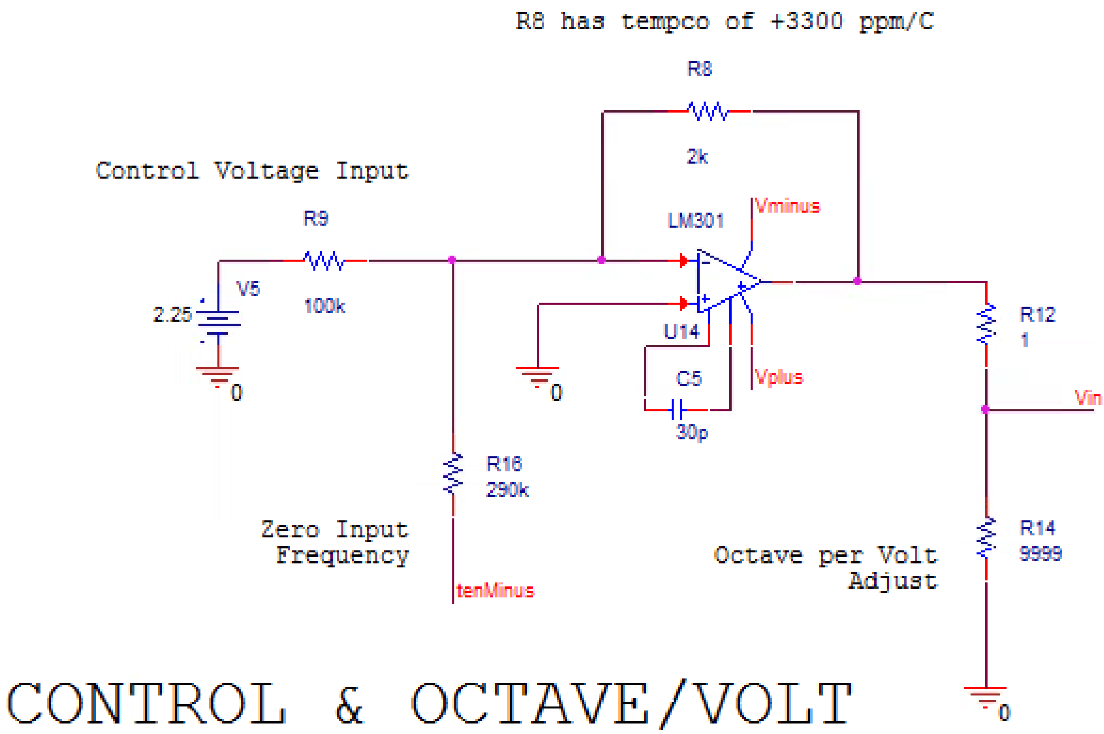

Figure 24: Input Sub-Circuit of VCO

First, there are no special requirements of this sub-circuit's op-amp, other than it behave ideally, and so the chosen LM301 is of the common, garden-variety type.

In terms of tolerance analysis, there are two components to look at here. The first is the $100k\Omega$ input resistor, which must have 1% tolerance. The reason for this is because the next stage, the exponentiator, is exponentially sensitive to control voltage input. And so we need to be sure that the gain of the LM301 inverting amplifier is precisely what it should be:

$$Gain = -\frac{2k}{100k} = 0.02 \tag{10}$$

The second consideration is the feedback resistor of the inverting amplifier, which has a temperature coefficient of +3,300 parts per million (ppm) per ° C. As discussed above, the collector current of the BJT, which performs the actual linear voltage to exponential current conversion, will increase with increasing temperature. We got around this problem by combining a differential pair with an op-amp as shown in Figure 21. However, it turns out that the exponentiator has a remaining temperature coefficient of -3,300 ppm. In order to cancel this out, a resistor with a +3,300 ppm temperature coefficient is required.

Conceptually, this is what's happening: let's say the temperature goes up. The remaining -3,300 ppm temperature coefficient of the exponentiator will cause more reference current to flow into it. Looking at Equation 9, more reference current means the base frequency will go up. This means the control voltage will not correspond to the frequency desired by the user. So the +3,300 ppm temperature coefficient (tempco) resistor in the feedback portion of the inverting amplifier will attenuate the control voltage by the exact amount required to bring the temperature related increase in reference current back down the required amount. This will only work if the tempco resistor increases in temperature by the same amount as the NPN transistors in the exponentiator. In order to ensure this, we will place the tempco resistor in thermal contact with the NPN transistors.

2.6.5 The Exponentiator Sub-Circuit of the VCO

The following figure is of our actual exponentiator sub-circuit we are using in our design.

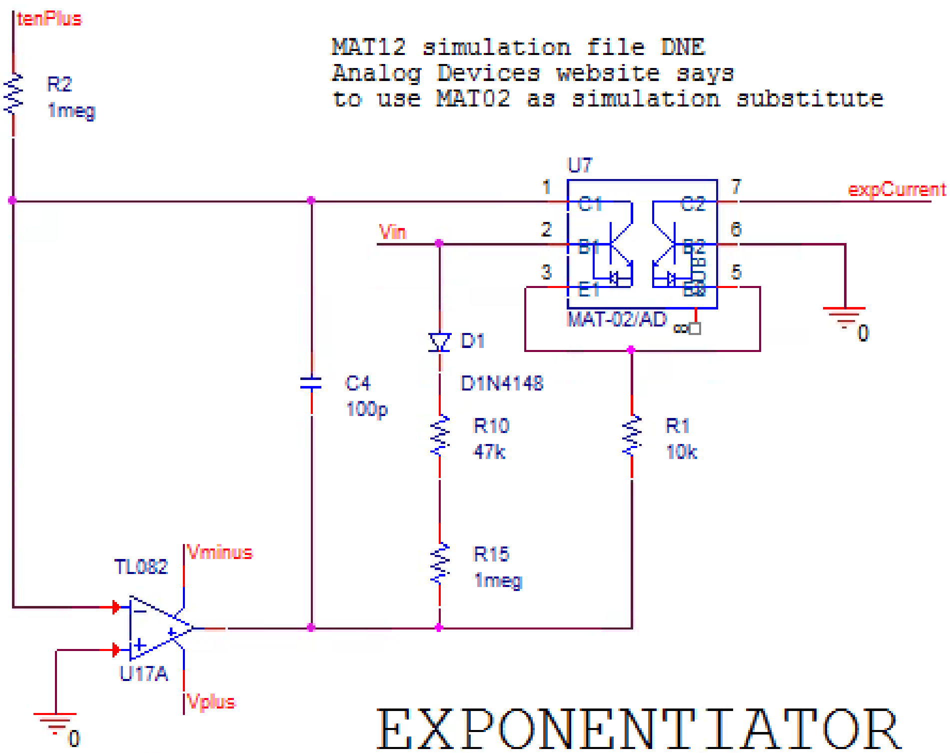

Figure 25: Exponentiator Sub-Circuit of VCO

The main considerations in our exponentiator circuit are the matched NPN transistors and the feedback path through from V_{in} to the output of the TL082 op-amp in Figure 25.

The NPN transistors must be perfectly matched. Looking at Equation 9 again, it's critical that we are able to perform the division of the two current equations at the step before arriving at Equation 5. The only way we can do this is if the I_S coefficient term as well as the $\frac{1}{V_T}$ term in the exponent are exactly the same. Otherwise, we would have to carry them around in the equation for output current, which would no longer have a reliable and simple relationship to the reference current as well as control voltage input. Analog Devices makes the perfect component for exactly this purpose, which we plan to use. It's a dual, matched NPN pair designed for audio applications called the MAT12.

Let's take a look at the feedback path from V_{in} to the output of the TL082 op-amp in Figure 25. The reason for this is to compensate for high frequencies being lower than they should be according to the control voltage input. As will be discussed in the next sub-section, the integrator portion of the oscillator has a finite discharge time. At high frequencies this discharge time may become significant when compared to the actual integration time, which is what we care about as it forms the period of the sawtooth signal. The discharge portion of the integrator signal *adds* time to the repeating sawtooth in the form of a negative slope when looking at the waveform. As just stated, at high frequencies, this can add significant time to the repeating signal, thereby lowering the frequency from what it should be according to the control voltage input.

The magnitude of this error is directly proportional to the control current from the converter. The addition of D1 and R15 in Figure 25 couples a voltage that is directly proportional to the control current back into the control input summer. This voltage is developed across the $10k\Omega$ protective resistor in series with the reference current regulator. The diode cancels a 0.6 V offset that exists at low values of control current.

Finally, let's look at the third sub-circuit within the VCO, which is an integrator in combination with a comparator, forming the actual oscillator signal.

2.6.6 The Oscillator Sub-Circuit of the VCO

The following figure is of our actual oscillator sub-circuit we are using in our design.

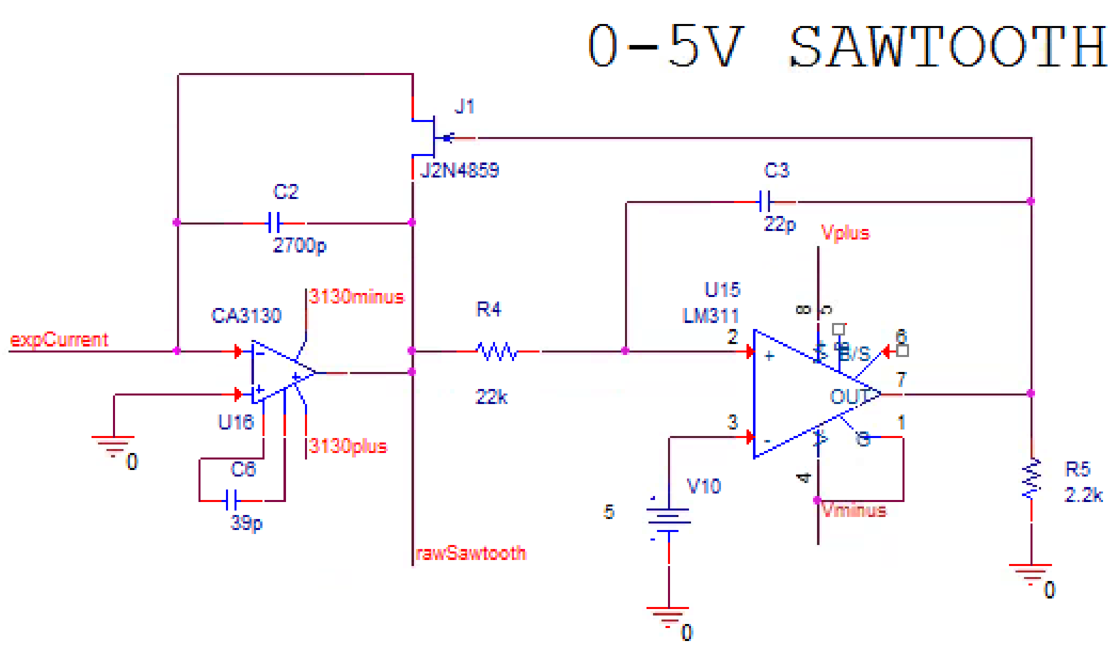

Figure 26: Oscillator Sub-Circuit of VCO

From a high-level perspective, looking at Figure 26, this is what's happening: a current from the exponentiator is being integrated into a voltage across the feedback capacitor of the first op-amp. This voltage is seen at the (+) input of the second op-amp, which functions as a comparator. The (−) input of the comparator is set to 5 V so that when the integrating voltage at the (+) input exceeds 5 V, the output of the comparator goes high. This causes the JFET to conduct, acting as a fast, momentary switch, which opens up a discharge path for the integrating capacitor. At this point, the capacitor has no voltage across it and so the comparator goes back low. Overall, we have just conceptually traced through one period of the sawtooth signal.

In terms of sensitivities and component choices, there are a few important considerations to make.

First, the CA3130 for the integrator is chosen for its ultra-high input impedance. The reason for this is because we don't want any of the current meant for integration across the capacitor to leak into the (−) input of the integrating op-amp. An ultra-high input impedance at the (−) input will prevent this from happening as much as is possible.

Second, the CA3130 that we just chose for its high input impedance cannot tolerate the ±12 V supply on the VCO and waveshaper circuits. So, we will need to adjust and level-shift its supply using Zener diodes.

The LM311 is designed to be a fast comparator, which is exactly what we need since we want to avoid adding time to the sawtooth signal at all costs.

The positive-feedback capacitor on the LM311 is to make its switch from low to high even faster, again to shave off as much of the unwanted discharge portion of the sawtooth signal. As soon as the integrator voltage reaches 5 V, the comparator output starts to go positive. As it does, the positive comparator input is forced even more positive through C3 giving positive feedback and causing the comparator to snap on.

C3 also serves to prevent the comparator from responding instantly to the drop in integrator voltage (due to the JFET switching on) by the charge it accumulated when the comparator was allowing for capacitor integration. In effect, C3 gives the JFET enough time to completely discharge the integrating capacitor.

The last consideration for the oscillator sub-circuit is the switching JFET. We choose the 2N4859 JFET since it has a high pinchoff and a low on-resistance. The high pinchoff is so we don't accidentally turn it on when we don't want to. And the low on-resistance is necessary for the capacitor to discharge as fast as possible.

3 Cost

Our development cost are estimated to be \$35/hour and 10 hours/week to complete our prototype. This gives us an estimated development cost of

$$2 * \frac{\$35}{hour} * \frac{10\ hours}{week} * 16\ weeks * 2.5 = \$28000 \tag{11}$$

The estimated cost of a prototype is \$186.05. The full parts list and cost per part can be seen below in Table 2. This brings our total cost of the project to be \$28186.05.

Table 2: Full Parts List

MCU Stage	Description	Value (if Applicable)	Manufacturer	Man. Part #	Supplier	Supplier Part #	Quantity	Cost	
	MCU- LPC810		NXP	LPC810M021FN8FP	Mouser	771-LPC810M021FN8FP	1	2.95	
	MCU Alternative - LPC812		NXP	LPC812M101JD20J	Mouser	771-LPC812M101JD20J	1	1.55	
	7-segment Display		Adafruit	1911	Adafruit	1911	1	9.95	
	Number Keypad		Adafruit	1824	Adafruit	1824	1	7.5	
	GPIO Expander		Microchip Technology	MCP23008-E/P	Mouser	579-MCP23008-E/P	1	1.08	
	Decoupling Capacitors	0.1 µF	Kemet	C320C104M5U5TA	Mouser	80-C320C104M5U	2	0.3	
	LED Indicators (Interrupts and PWM)		Lumex	SSL-LX5093IT	Mouser	696-SSL-LX5093IT	2	0.39	
	LED Resistors	470 Ω	Yageo	CFR-25JR-52-470R	Mouser	603-CFR-25JR-52470R	2	0.1	
	Pullup Resistors	10 kΩ	Yageo	CFR-25JR-52-10K	Mouser	603-CFR-25JR-5210K	2	0.1	
								24.81	MCU Total

TTL Stage	Description	Value (if Applicable)	Manufacturer	Man. Part #	Supplier	Supplier Part #	Quantity	Cost	
	Counter		Texas Instruments	SN74HC161N	Mouser	595-SN74HC161N	1	0.41	
	Analog Mux		Texas Instruments	CD4051BE	Mouser	595-CD4051BE	2	2	
	Decoders		Texas Instruments	SN74HC138N	Mouser	595-SN74HC138N	2	0.41	
	Hex Inverter		Texas Instruments	SN74HC05N	Mouser	595-SN74HC05N	1	0.5	
	Encoders		Texas Instruments	SN74HC148N	Mouser	595-SN74HC148N	4	0.41	
	Quad 2-input NAND		Texas Instruments	SN74HC00N	Mouser	595-SN74HC00N	2	0.46	
	Comparator		Texas Instruments	CD74HC85EE4	Mouser	595-CD74HC85EE4	1	1.15	
	LED Step Indicators		Lumex	SSL-LX5093IT	Mouser	696-SSL-LX5093IT	2	0.39	
	LED Resistors	470 Ω	Yageo	CFR-25JR-52-470R	Mouser	603-CFR-25JR-52470R	2	0.1	
	Pullup Resistors	1 kΩ	Yageo	CFR-25JR-52-1K	Mouser	603-CFR-25JR-521K	38	0.1	
	DPDT Pushbuttons		Apem	MHPS2283	Mouser	642-MHPS2283	32	0.64	
	Slider Potentiometers	10 kΩ	Bourns	PTA6043-2015DPB103	Mouser	652-PTA60432015DPB10	16	1.6	
								60.3	TTL Total

VCO Stage	Description	Value (if Applicable)	Manufacturer	Man. Part #	Supplier	Supplier Part #	Quantity	Cost	
Control Input Sub-Ckt	Tempco feedback resistor	1 kΩ , +3300 ppm	Akaneohm	LT16S102F33	Synth Rotek	LT16S102F33	1	3.5	
	Octave per volt trimpot	5 kΩ	Bourns	3296W-1-502LF	Mouser	652-3296W-1-502LF	1	2.41	
	Tune Series resistor	200 kΩ	Yageo	CFR-25JR-52-200K	Mouser		1	0.1	
	Tune trimpot	100 kΩ	Bourns	3296Y-1-104LF	Mouser	652-3296Y-1-104LF	1	2.41	
	Fine tune trimpot	5 MΩ	Bourns	3296Y-1-505LF	Mouser	652-3296Y-1-505LF	1	2.41	
	Course tune trimpot	500 kΩ	Bourns	3299W-1-504LF	Mouser	652-3299W-1-504LF	1	3.39	
	Input resistors (0.1% tolerance)	100 kΩ	Vishay / BC Components	UXB02070F1003BC100	Mouser	594-UXBB100K00B1A	2	0.77	
	Summing Op-amp	LM301	Texas Instruments	LM301AN/NOPB	Mouser	926-LM301AN/NOPB	1	0.81	
	Summing Op-amp Capacitor	30 pF	Kemet	C317C300J5G5TA	Mouser	80-C317C300J5G	1	0.82	
Exponentiator Sub-Ckt	Matched NPN's		Analog Devices	MAT12AHZ	Mouser	584-MAT12AHZ	1	29.6	
	Diff. pair tail resistor	10 kΩ	Yageo	CFR-25JR-52-10K	Mouser	603-CFR-25JR-5210K	1	0.1	
	High-freq. tracking diode	1N4148	Fairchild Semiconductor	1N4148	Mouser	512-1N4148	1	0.1	
	High-freq. tracking resistor	47 kΩ	Yageo	CFR-25JR-52-47K	Mouser	603-CFR-25JR-5247K	1	0.1	
	High-freq. tracking trimpot	1 MΩ	Bourns	3296W-1-105LF	Mouser	652-3296W-1-105LF	1	2.41	
	Op-amp feedback capacitor	100 pF	Kemet	C315C101K2R5TA	Mouser	80-C315C101K2R	1	0.34	
	Collector resistor	1 MΩ	Yageo	CFR-25JR-52-1M	Mouser	603-CFR-25JR-521M	1	0.1	
	Tail op-amp	TL081	Texas Instruments	TL081BCP	Mouser	595-TL081BCP	1	1.1	
Oscillator Sub-Ckt	Integrating Capacitor	2700 pF	Panasonic	ECQ-E10272KF	Mouser	667-ECQ-E10272KF	1	0.74	
	Cap for faster Switching	18 pF	Kemet	C315C180K2G5TA	Mouser	80-C315C180K2G	1	0.38	
	Alt. Cap for faster Switching	22 pF	Kemet	C315C220J2G5TA	Mouser	80-C315C220J2G	1	0.43	
	Chopper JFET for Switching	2N4859	InterFET	2N4859	Mouser	106-2N4859	1	7.4	
	Resistor between op-amps	15 kΩ	Yageo	CFR-25JR-52-15K	Mouser	603-CFR-25JR-5215K	1	0.1	
	Resistor at output of comparator	2.2 kΩ	Yageo	CFR-25JR-52-2K2	Mouser	603-CFR-25JR-522K2	1	0.1	
	Comparator VDR resistors	10 kΩ, 1%	KOA Speer	SPR1CT52R1002F	Mouser	660-SPR1CT52R1002F	2	0.25	
	Integrating Op-amp	CA3130	Intersil	CA3130EZ	Mouser	968-CA3130EZ	1	2.35	
	Integrating Op-amp Capacitor	39 pF	Kemet	C322C390J2G5TA	Mouser	80-C322C390J2G	1	0.43	
	Comparator op-amp	LM311	Texas Instruments	LM311P	Mouser	595-LM311P	1	0.6	
	Zener Diode for CA3130 Supply	9.1 V, 1%	Vishay Semiconductor	TZX9V1D-TR	Mouser	78-TZX9V1D	1	0.19	
	Zener Diode for CA3130 Supply	3.9 V, 2%	Vishay Semiconductor	TZX3V9C-TR	Mouser	78-TZX3V9C	1	0.19	
	Zener Circuit Capacitor	0.1 µF	Kemet	C320C104M5U5TA	Mouser	80-C320C104M5U	2	0.3	
	Zener Circuit Resistor	470 Ω	Yageo	CFR-25JR-52-470R	Mouser	603-CFR-25JR-52470R	1	0.1	
	Zener Circuit Resistor	1 kΩ	Yageo	CFR-25JR-52-1K	Mouser	603-CFR-25JR-521K	1	0.1	
Sawtooth Sub-Ckt	Dual op-amp	TL082	Texas Instruments	TL082BCP	Mouser	595-TL082BCP	1	1.42	
	Input resistor	10 kΩ, 1%	KOA Speer	SPR1CT52R1002F	Mouser	660-SPR1CT52R1002F	1	0.25	
	Feedback Resistor	39 kΩ, 2%	KOA Speer	CF1/4CT52R393G	Mouser	660-CF1/4CT52R393G	1	0.37	
	Offset Resistor	39 kΩ, 2%	KOA Speer	CF1/4CT52R393G	Mouser	660-CF1/4CT52R393G	1	0.37	
	Inverter input and feedback resistors	10 kΩ, 1%	KOA Speer	SPR1CT52R1002F	Mouser	660-SPR1CT52R1002F	2	0.25	
Squarewave Sub-Ckt	Input Op-amp	LM301	Texas Instruments	LM301AN/NOPB	Mouser	926-LM301AN/NOPB	1	0.81	
	Input resistor	1 kΩ	Yageo	CFR-25JR-52-1K	Mouser	603-CFR-25JR-521K	1	0.1	
	Feedback resistor	1 MΩ	Yageo	CFR-25JR-52-1M	Mouser	603-CFR-25JR-521M	1	0.1	
	Level-shift trimpot	100 kΩ	Bourns	3296Y-1-104LF	Mouser	652-3296Y-1-104LF	1	2.41	
	Base resistor	4.7 kΩ	Yageo	CFR-25JR-52-4K7	Mouser	603-CFR-25JR-524K7	1	0.1	
	Comp. emitter-follower NPN	NPN	Fairchild Semiconductor	2N3904BU	Mouser	512-2N3904BU	1	0.17	
	Comp. emitter-follower PNP	PNP	Fairchild Semiconductor	2N3906BU	Mouser	512-2N3906BU	1	0.19	
	Buffer Op-amp	TL081	Texas Instruments	TL081BCP	Mouser	595-TL081BCP	1	1.1	
Triangle Sub-Ckt	Op-amp	LM301	Texas Instruments	LM301AN/NOPB	Mouser	926-LM301AN/NOPB	1	0.81	
	Op-amp capacitor	18 pF	Kemet	C315C180K2G5TA	Mouser	80-C315C180K2G	1	0.38	
	Input resistor	10 kΩ, 1%	KOA Speer	SPR1CT52R1002F	Mouser	660-SPR1CT52R1002F	1	0.25	
	Feedback resistor (series makeshift)	2 @ 10 kΩ, 1%	KOA Speer	SPR1CT52R1002F	Mouser	660-SPR1CT52R1002F	2	0.25	
	Triangle offset resistor	10 kΩ	Yageo	CFR-25JR-52-10K	Mouser	603-CFR-25JR-5210K	2	0.1	
	Triangle offset trimpot	10 kΩ	Bourns	3296W-1-103LF	Mouser	652-3296W-1-103LF	1	2.41	
	Rectifying diodes	2 @ 1N4148	Fairchild Semiconductor	1N4148	Mouser	512-1N4148	2	0.1	
	Glitch reduction capacitor	100 pF	Kemet	C315C101K2R5TA	Mouser	80-C315C101K2R	1	0.34	
								78.43	VCO Total

Top-level Parts	Description	Value (if Applicable)	Manufacturer	Man. Part #	Supplier	Supplier Part #	Quantity	Cost	
	Rail splitter		Texas Instruments	TLE2426CP	Mouser	595-TLE2426CP	2	1.82	
	3.3 V Regulator	3.3 V 250mA	Microchip Technology	MCP1700-3302E/TO	Mouser	579-MCP1700-3302E/TO	1	0.4	
	5 V Regulator	5 V 1.5A	STMicroelectronics	L7805CV	Mouser	511-L7805CV	1	0.43	
	20 V Regulator	1.2 to 37 V 1.5A	STMicroelectronics	LM317T	Mouser	511-LM317T	1	0.52	
	24 V Regulator	24 V 1.5A	STMicroelectronics	L7824CV	Mouser	511-L7824CV	1	0.54	
	AC-DC Wall Power Supply	24 V 1.8A	Parts Express	120-054	Parts Express	120-054	1	16.98	
								22.51	Pow. Total
								186.05	Grand Total

Notes:
The TL08 series op-amps can all be used with a single TL084 instead.
The Sawtooth feedback and offset resistors are 39k and 2% but ideally, they should be 40k and 1%. So maybe it's better to do 40k and 5%.

4 Schedule

Week	Justin	Kyle	Both
2/27	Order Parts	Finish power supply calculations	Design review
3/6	Begin laying out VCO and waveshaper on breadboard	Work on Power Supply PCB	Soldering assignment
3/13	Begin connecting all module systems together on breadboard	Begin laying out power supply on breadboard	Breadboard prototype completed
3/20	Spring Break	Spring Break	
3/27	Revise and Submit PCB design to Machine Shop	Revise PCBs	Individual Progress Report
4/3	Debug and fine tune MCU Stage	Debug and fine tune TTL stage. Sign up for Mock Demo	
4/10	Debug and fine tune VCO stage	Debug and fine tune power supply. Sign up for mock presentation, demonstration and presentation	
4/17	Mock Demo	Mock Demo	All parts soldered together
4/24	Prepare for demonstration and mock presentation	Begin Final Paper	Demonstration
5/1	Prepare for Presentation	Prepare for Presentation	Turn in Final Paper

5 Ethics and Safety

There are a few safety concerns with out project. The concern would be making sure proper precaution is used when dealing with wall outlet voltage at 120V AC and converting it to 24V DC, which is for all intents and purposes, an absolutely manageable safety concern. We will utilize the one hand method and make sure our wall outlet contains a ground. Also, we will need to ensure that the AC-DC conversion is closed off from the user so they will never come into contact with high voltages.

Additionally, when dealing with high voltages, it can create large currents and dissipate heat [8]. We will need to ensure that precaution is taken on our part to ensure that this excess heat is handled correctly to avoid hazards in testing, but also in the final product so the user is never exposed to this.

We are responsible for all decisions made in the design of this product and it is our responsibility to disclose any issues that issues that might endanger the user per Section 1 of the IEEE Code of Ethics [9]. We believe that if properly designed, we will be able to mitigate these hazards to create a pleasant and enjoyable experience for the user.

Our system will be designed to be plug and play, being compatible with common wall outlets. Given the high voltage, and our lack of experience in designed high voltage converters and regulators, we have chosen to simply their design by utilizing ICs and off the shelf converters. This is in accordance with Section 7 of the IEEE Code of Ethics [9].

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