16 Step Analog/Digital Step Sequencer and Synthesizer

By

Justin Sconza

Kyle Vondrak

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TA: Eric Clark

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Abstract

This report chronicles the ideation, prototyping, development and refinement of a 16-step analog musical sequencer with digital tempo input. Sixteen steps in time consisting of sixteen voltages on slider potentiometers are applied in sequence to a voltage controlled oscillator which performs linear voltage to exponential frequency conversion. The voltage controlled oscillator, or VCO for short, is capable of three signal waveform types: triangle, sawtooth and square. Additionally, beneath each of the sixteen slider potentiometers governing the steps, there are three pushbutton switches which enable the user to 1) hold on a given step in order to properly tune that particular note in the sweeping melody, 2) to rest on a given step as in skip that particular note as the melody is playing out, 3) to start the sequence over at a given step, thereby ignoring the remaining steps. The pace at which the sequence is swept through, as in the tempo of the melody, is digitally programmable from 0 to 9999 beats-per-minute, or BPM as it is more commonly known.

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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.

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.

1.2 High Level Requirements

The main high level requirements for our system are described below

- 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

The project can be divided into six main sub-sections: the synthesizer stage, the TTL stage, the MCU stage, the output stage, the user interface and the power sections. The TTL logic section counts out the musical steps and performs the button features described above. The MCU allows for digital tempo input. The output stage serves to modify the output signals in order to be able to skip steps. The user interface stage allows the user to adjust the offset and range of the voltage levels, and to display the current step on a set of LEDs. As far as the power goes, our main objective was to have the entire product rely on one source so that it can be "plug and play".

By the end of the semester, everything we intended to accomplish was done. There are a few broken buttons and a broken LED on the user interface board, and there are many little things we would do differently next time, however, all in all, everything works exactly as we planned it should. Figure 1 shows the overall block diagram and how the stages connect.

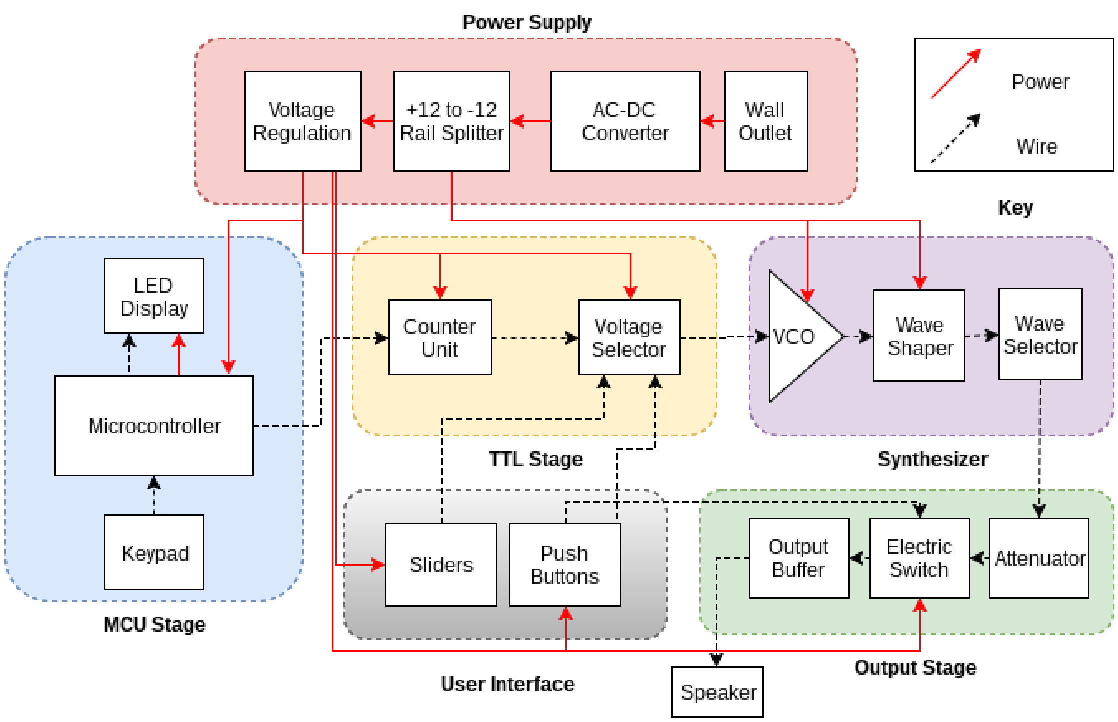


Figure 1: Block Diagram for Overall System

2.1 Design Details

2.1.1 Synthesizer Stage

The VCO is a linear voltage to exponential frequency oscillator tuned for an octave per volt. It maps the manageable voltage range of 0 to 5 V to the exponential frequency range of 60 Hz to (60 Hz) $* 2^5 = 1.92$ kHz. Our system has two inputs to the VCO, one from the analog MUX sweeping through each of the steps in the sequencer and a second from an additional slider potentiometer set to sweep 0 to 5 V. This additional slider is for transposing the sequencer melody by applying an offset voltage.

The VCO consists of two sub-circuits, the actual linear voltage to exponential frequency oscillator, which we shall refer to as the VCO, and then a second stage for shaping the sawtooth signal into triangle and square waves as well, which we shall call the waveshaper section. Figure 2 shows the first sub-circuit, the VCO.

Chamberlain [3] and Schmitz [4] provide great explanations of how an octave per volt synthesizer circuit works.

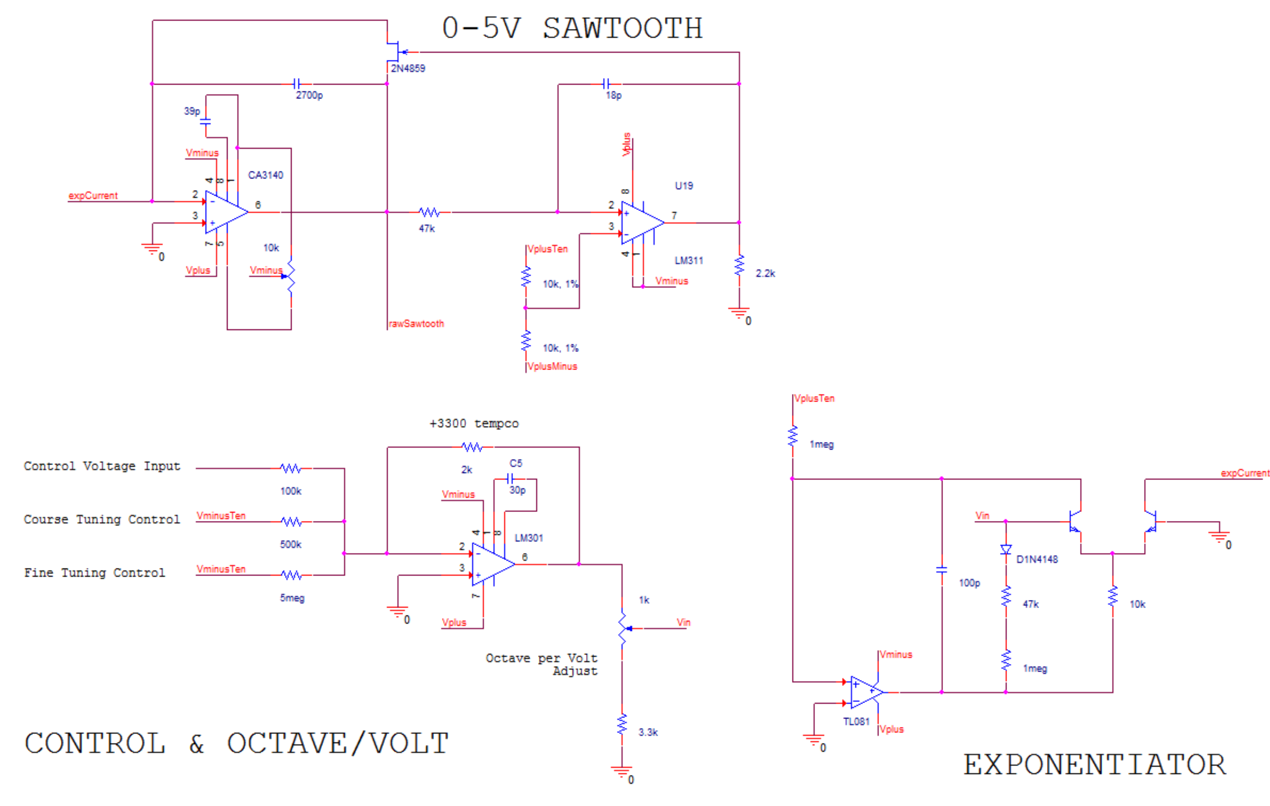


Figure 2: Schematics for VCO sub-circuit

The waveshaper section, illustrated in Figure 3, take the 0 to 5 V sawtooth signal output from the VCO and converts it into three separate ±10 V waveform signals, a sawtooth, triangle and square. The reason for different shapes is because each has a different musical timbre owing to differing amounts of harmonics, the triangle having the least and therefore sounding the gentlest, and the sawtooth having the most, therefore sounding the harshest. Waveshapes from these three different generators can be seen in Appendix B.

The sawtooth generator takes our 0 to 5 V sawtooth, and amplifies it to a peak to peak voltage of 20 V.

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.

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.

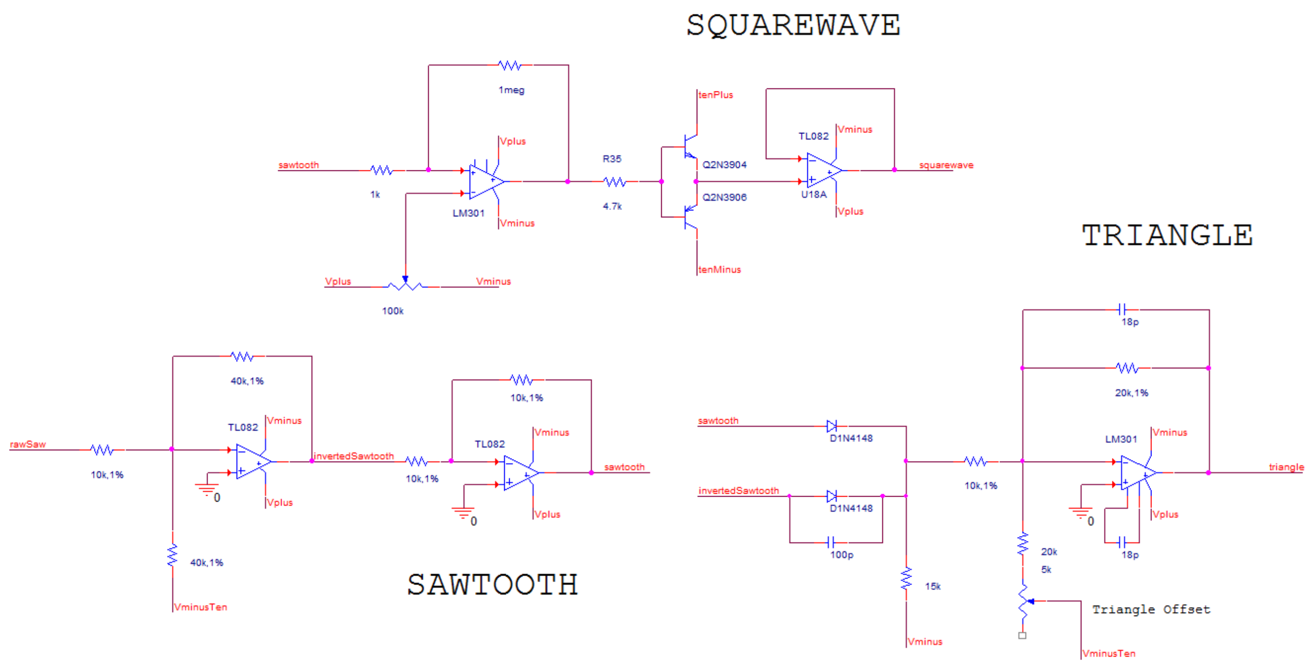


Figure 3: Waveshaper Schematics

2.1.2 MCU Stage

The MCU stage consists of an ARM microcontroller, a numeric keypad and an LED display. The microcontroller scans the keypad for input over I^2C, converts that input to a PWM signal for driving the TTL counter, and finally instructs the LED display over I^2C to show the current tempo. A detailed description of the software can be found in Section 2.1.7.

Figure 4 shows the full MCU schematics

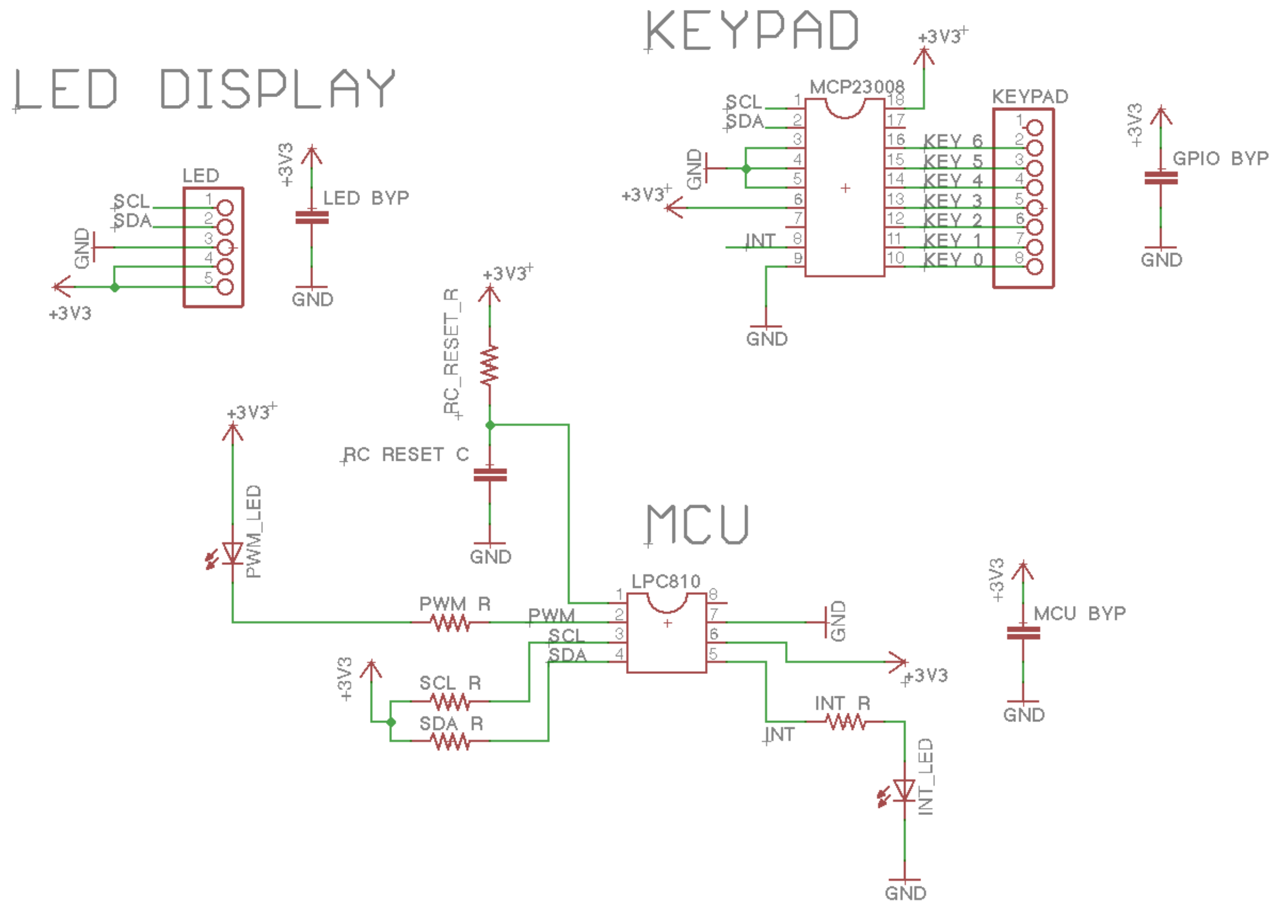


Figure 4: MCU, Keypad and LED Schematics

2.1.3 TTL Stage

Technically speaking, HCMOS logic ICs were used, although in line with the general convention, we are referring to this stage as transistor-transistor logic. A great application note by Texas Instruments on HCMOS design considerations [5], was found extremely helpful throughout the design of the project.

The TTL stage consists of multiple logic sections: 1) a priority encoder combined with a comparator for starting the sequence over at a given step, 2) another priority encoder used for the hold functionality, 3) a decoder for sweeping through the LEDs as well as driving the diode logic behind the step skipping functionality, 4) the counter driven by the MCU and driving the step-sequencer 5) and finally, the analog MUXes connected to each of the voltages set by the user on the slider potentiometers and swept through by the counter.

Figure 5 details the TTL counter, LED decoders and analog MUX schematics.

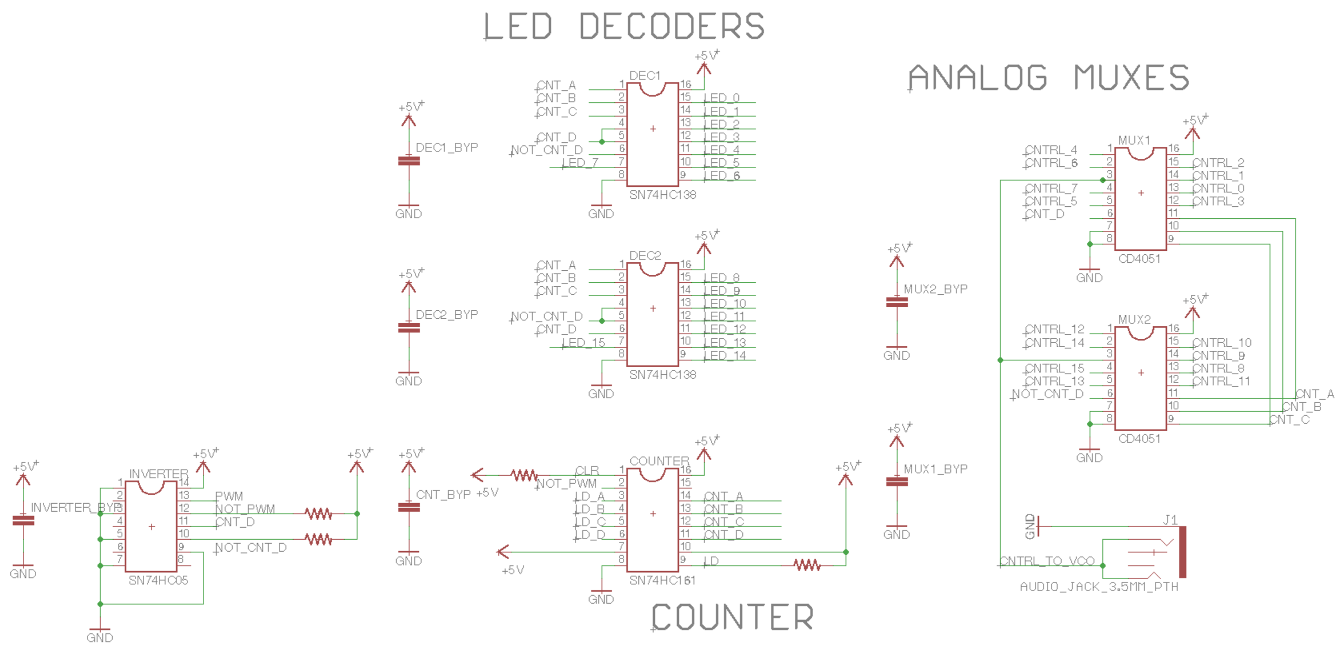


Figure 5: TTL Counter, LED Decoder and Analog MUX Schematics

Figure 6 shows the priority encoders used for the start-over and hold features of the user interface.

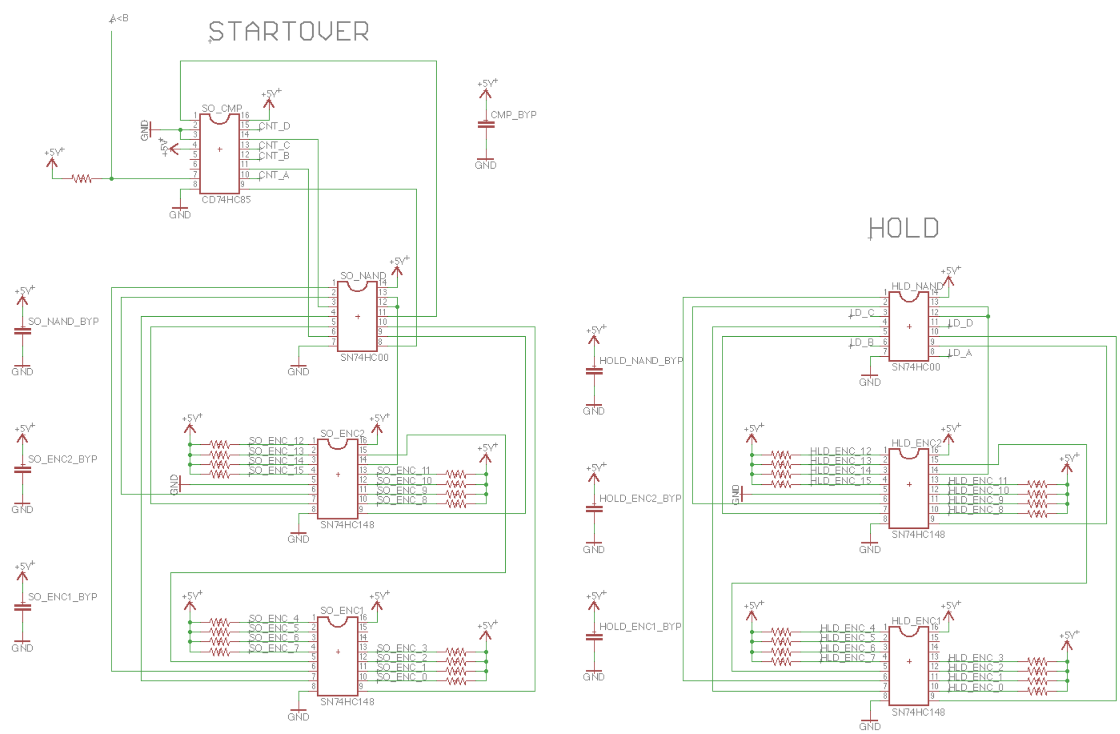


Figure 6: Priority Encoder Circuits for the Start-over and Hold Features of the User Interface

2.1.4 Power Supply

The main supply is a 24 V wall-wart adapter, chosen for the maximum current without being too expensive, which comes out to be 1.8 A, and about $20. After the wall-wart input, the power stage can be further divided into separate analog and digital sections.

For the analog power section, everything is accomplished using linear regulators. The +24 V is split into a ±12 V line using a rail-splitter IC. The +12 V is then regulated down to +10 V using an adjustable linear regulator and the -12 V is likewise regulated up to -10 V using an adjustable negative linear regulator.

Figure 7 shows the analog power schematics. Looking at the figure, the left most circuit is the rail-splitter used to create ±12 V from the +24 V wall wart adapter. In the middle of the figure, we have the linear regulator circuit, which creates ±10 V from ±12 V. And finally, the right-most circuit is for creating a two-sided voltage that the CD4016 electric switch can handle.

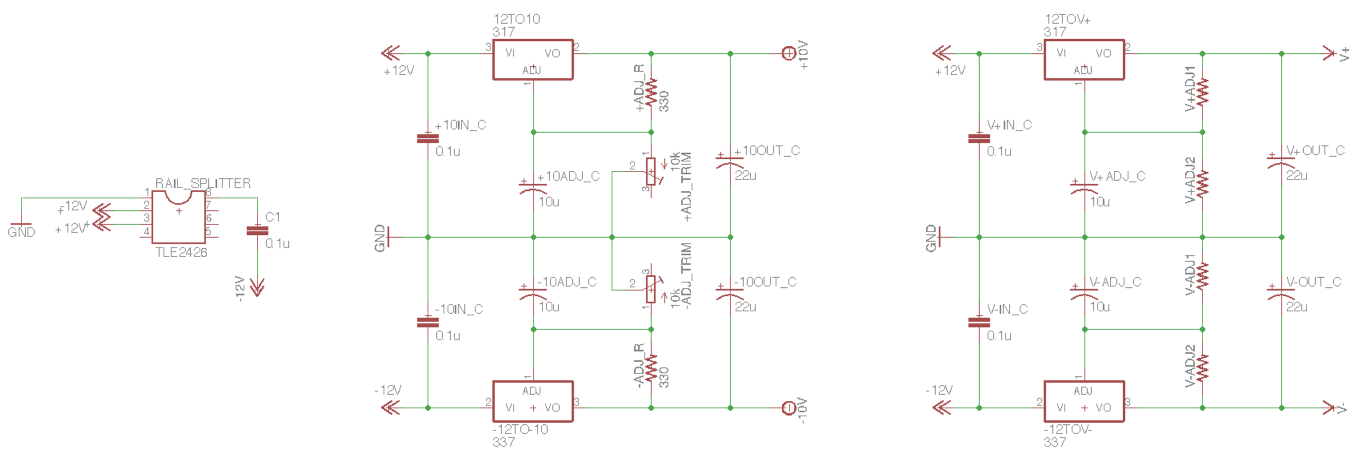


Figure 7: Analog Power Schematics

The digital power section consists of two DC-DC converters for creating the +5 V and +3.3 V lines. The reason for using DC-DC converters is because they offer much better isolation between sections. The first converter steps +12 V down to +5 V and the second converter steps +5 V down to +3.3 V.

In Figure 8, the digital power circuit is shown. As stated earlier, the digital power schematic consists of two MC34063A step-down DC-DC converters. The components were selected by carefully pouring over the datasheet [6] as well as a great application note written by ON Semiconductor [7].

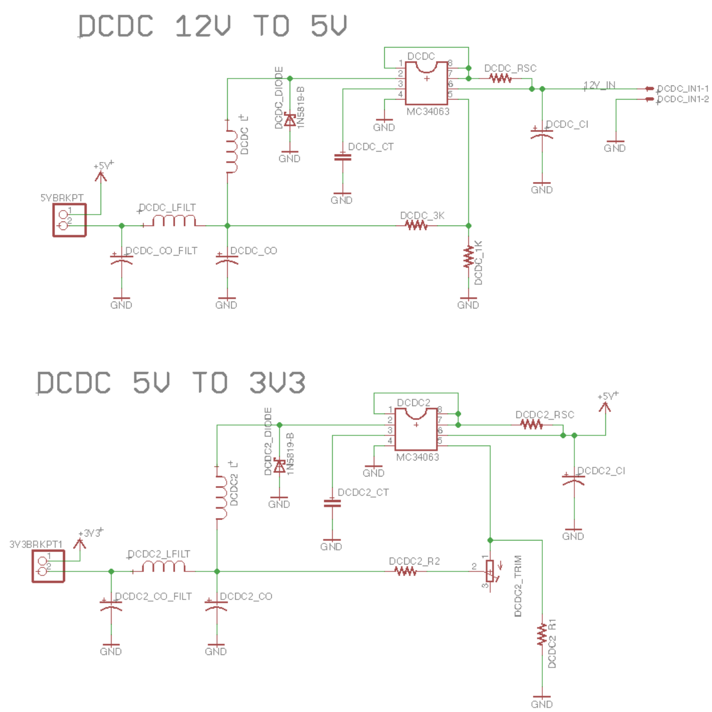


Figure 8: DC-DC Converter Schematics for the Digital Section

2.1.5 Output Stage

Figure 9 shows the three sub-circuits which make up the output stage. Looking at the figure below, on the top we have the attenuator which serves to reduce the ±10 V signal from the VCO stage down to ±2 V, a much more manageable level for the CD4016 electric switch as well as for the output speakers.

In the middle of the figure, we have the signal-to-skip circuit, followed by the CD4016 analog MUX. The first part of this middle section of the figure serves to scale up the TTL level signal coming from the LED decoders. This signal needs to be scaled up in order to reliably drive the control input on the CD4016. A simple RC filter before the signal is applied to the control input of the CD4016 in order to reduce the switching related noise on the output signal whenever a step was skipped. The RC filter takes advantage of the fact that the CD4016 is actually a "continuous" switch that allows for VCA like control. By rolling off the sharp edge of the signal, the CD4016 rolls off its control which eliminates the switching noise on the output. The time constant of the RC filter was determined in real-time by trial and error.

Finally, the CD4016 output is sent through a simple unity gain inverting amplifier, through a blocking capacitor and to the speakers for one's listening enjoyment.

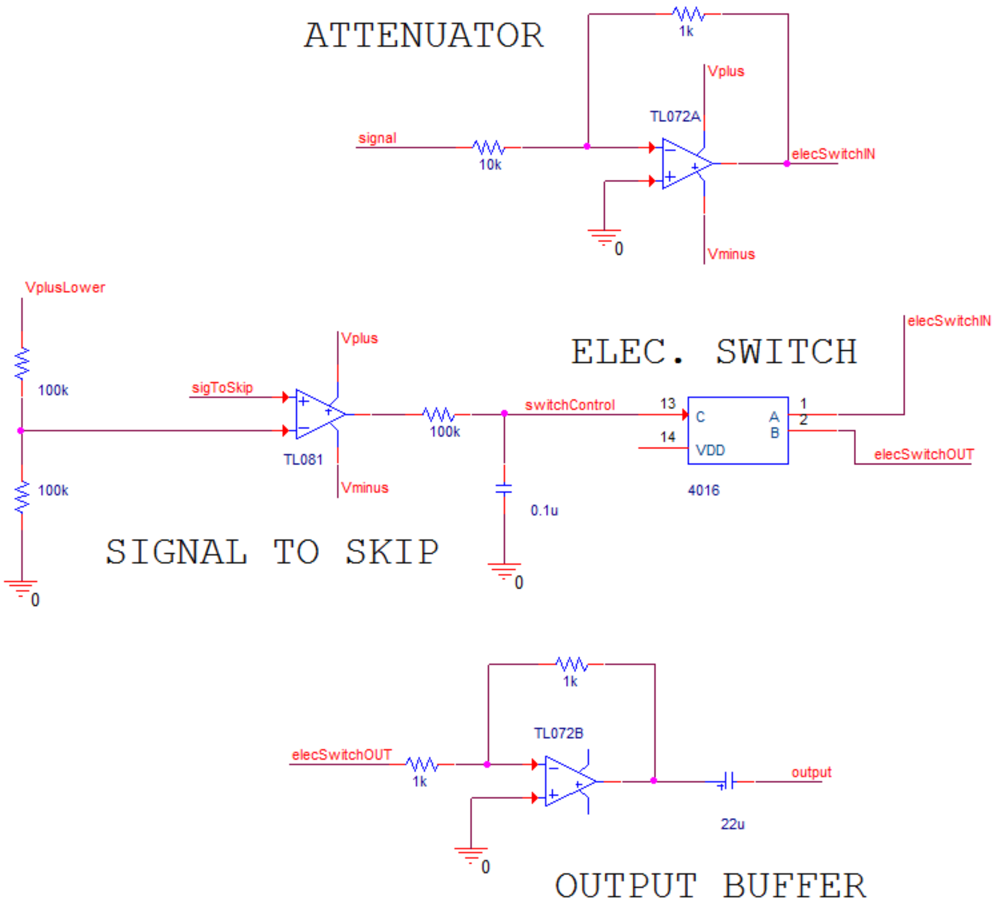


Figure 9: Output Stage Schematics

2.1.6 User Interface Stage

The following two figures are of the user interface schematic. Figure 10 illustrates the range and offset slider potentiometers as well as two example step circuits. The offset slider is discussed above and the range slider simply determines the range of voltage that each of the step sliders are allowed to sweep through, anywhere from 0 to 5 V. When they sweep through 5 V, the full five octave range is available, and when they sweep through 1 V, one octave is available, etc.

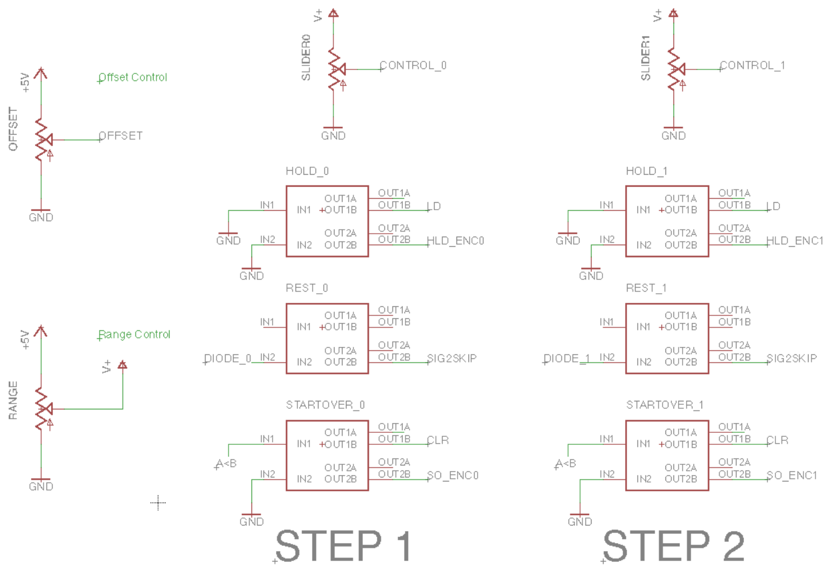


Figure 10: Offset and Range Control and Two Example UI Steps

Figure 11 illustrates the schematic layout of the step LEDs and diode logic circuit. The right is the diode logic which governs the user functionality for skipping notes.

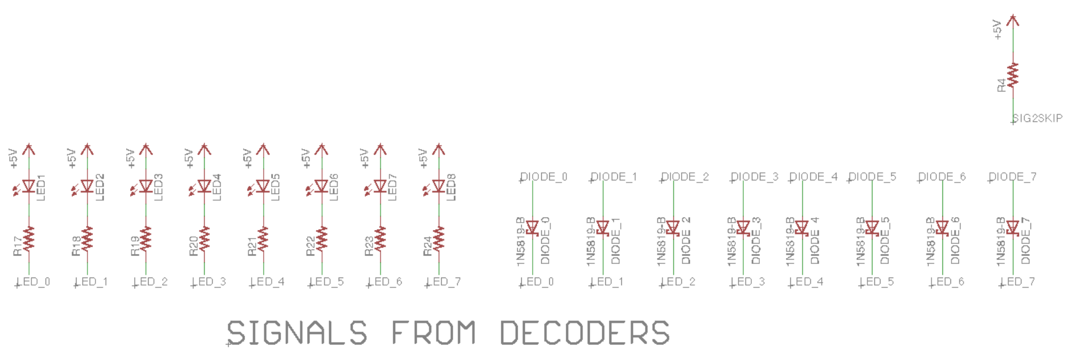


Figure 11: Step LEDs and Diode Logic Schematic

2.1.7 Software

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

$$[BPM] \xRightarrow{\text{Divide by 60}} [Hz] \tag{1}$$

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.

The system clock of the LPC810 is 30 MHz. During the design phase of the project, we were under the impression that the system clock of the chip was 12 MHz, however it is actually 30 MHz. We figured this out by simply printing the macro defined SYSTEM CLOCK. Furthermore, it was determined by trial and error, using an oscilloscope, that a more precise value of the system clock is actually 29925187.03 Hz. So when the user enters a BPM, in code, we simply perform the the following formula using 29925187 for System Clock.

$$\frac{\text{System Clock} * 60}{\text{BPM}} = \text{PWM Count} \tag{2}$$

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.

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; However, the LPC810, having only six pins that are not 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.

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 shown in Figure 12.

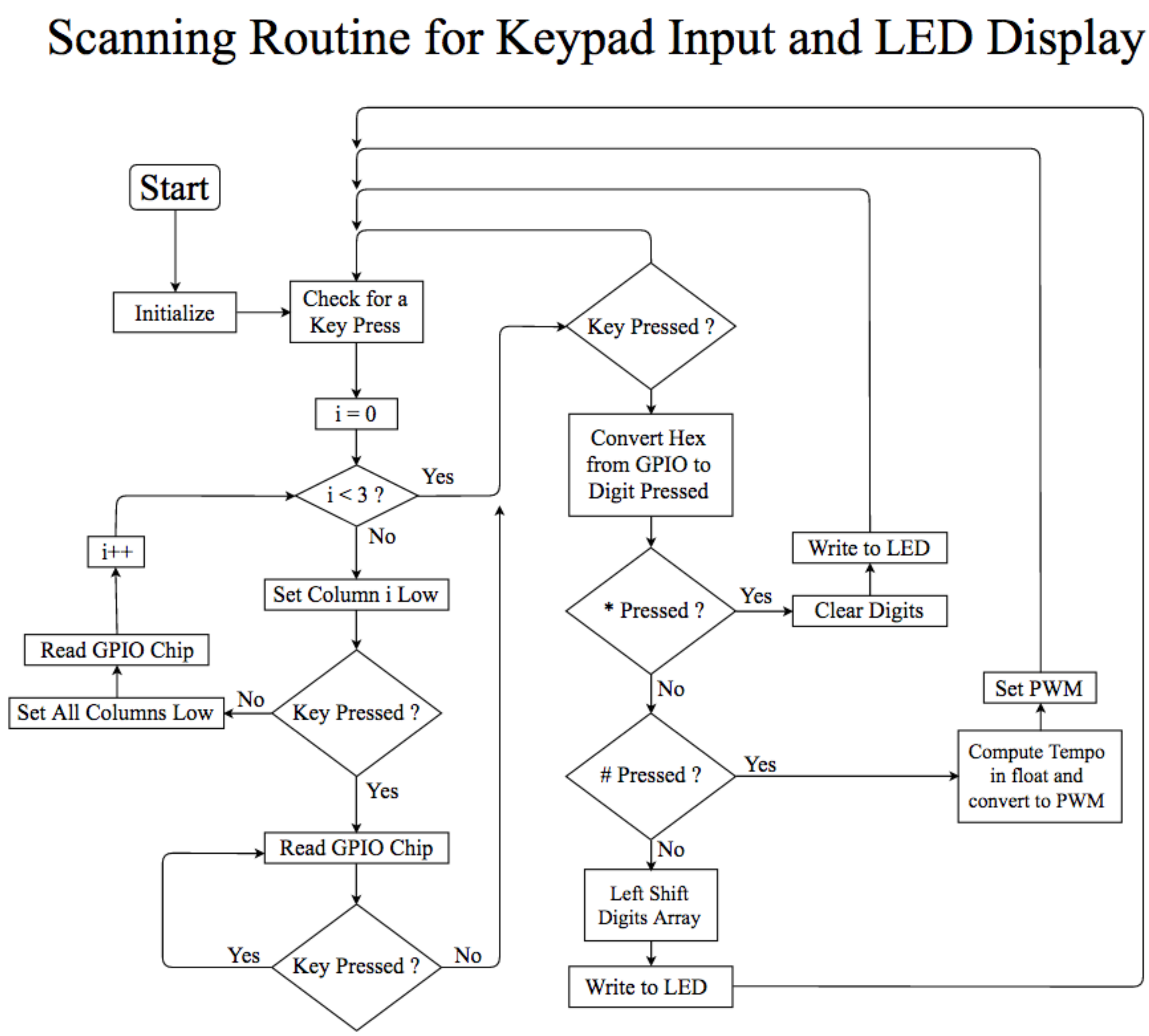


Figure 12: Software Flowchart for MCU Stage

3 Design Verification

3.1 Synthesizer Stage

We verified the synthesizer stage by connecting the audio output of the system to an oscilloscope. Waveshapes showing the verification of our requirements can be seen in Appendix B. They also highlight the fact that our target voltage to frequency conversion is working as expected. The cents difference across the control voltage range is shown in Table 1.

Table 1: Cents Difference in Synthesizer

Control Volts (V)	Exp. Freq. f_1 (Hz)	Meas. Freq. f_2 (Hz)	Cents
0	60	60.025	0.7212
1	120	119.9	-1.4433
2	240	240.1	0.7212
3	480	479.8	-0.7215
4	960	959.7	-0.5411
5	1920	1921	0.9014

Table 2 shows the measured sawtooth reset time, and the percent error compared to the period of the waveform. Table 3 shows the trapezoidal component compared to the period of the waveform.

Table 2: Sawtooth Reset Time and Percent Error

Control Volts (V)	Meas. Period (ms)	Reset Time (ms)	% Error
0	16.7	0.132	0.77
1	8.36	0.141	1.68
2	4.16	0.137	3.29
3	2.09	0.135	6.46
4	1.04	0.135	12.98
5	0.52	0.139	26.73

Table 3: Trapezoidal Component Time and Percent Error for Squarewave

Control Volts (V)	Meas. Period (ms)	Reset Time (ms)	% Error
0	16.7	0.164	0.99
1	8.36	0.164	1.97
2	4.16	0.16	3.95
3	2.09	0.17	8.17
4	1.04	0.164	15.7
5	0.52	0.164	31.4

It is important to note that the sawtooth reset time and trapezoidal time stay constant across all values of control voltage. This is because it is a value specified by our hardware choice, and cannot be changed. This causes our system to meet the system requirements at low control voltage when the output frequency is low, and the error increases as the frequency increases.

The peak to peak value of the output waveforms also stays within 10% of 20 V.

3.2 MCU Stage

The MCU stage was tested on the final build of the system. The MCU stage was able to achieve all the requirements. The keypad was successful in serving as an input to the microcontroller, and the LED display was effective in displaying the current BPM of the system. The microcontroller was able convert the BPM to a PWM wave in a reasonable amount of time, less than 200 ms and fast enough to be unnoticeable to the user. Figure 13 shows the percent error of the BPM across the range of 60 to 243.

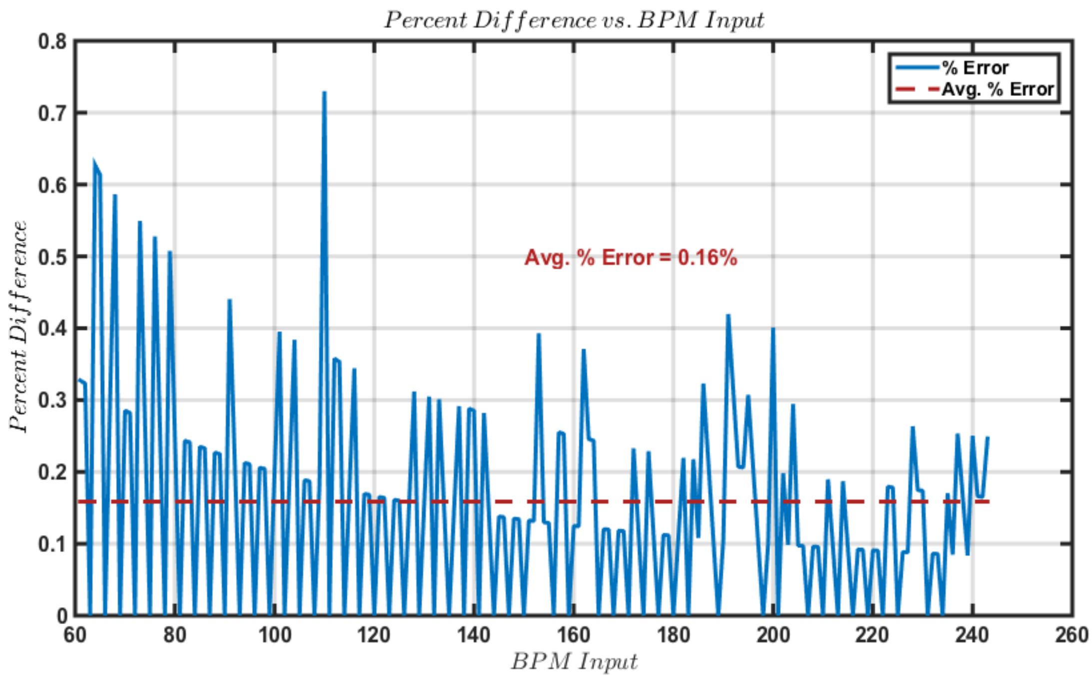


Figure 13: Percent Error across BPM range from 60 to 242

Across the BPM range of 60 to 243 the system, the average BPM error was 0.16%, which is much less than the 1% imposed by our requirements. We should also be checking the higher BPM ranges. However, from about 240 to 1000 BPM, the tempo is so much faster than any human could ever play along with, that accuracy becomes less of a concern. And above 1000 BPM, the tempo is no longer within the audible range. What happens in this region is a sort of smearing of the notes into one continuous yet still audibly discrete single note and therefore tempo accuracy has little meaning.

3.3 TTL Stage

The buttons were able to perform their intended tasks within 50 ms. Their operation happens almost instantaneously, which was our goal.

The voltage range of the sliders was measured to be 0.005 to 5.01 V. This is versus a desired range of 0.00 to 5.00 V. Because the lower limit is zero, to speak of a percentage difference or percent error does not make much sense because any number, no matter how infinitesimal is very different than zero. Perhaps a better metric of accuracy would be the difference between the desired and measured endpoints from the full range of 5V. This would give a 0.1% figure on the lower limit and a 0.2% figure on the upper limit.

3.4 Power Supply

The power supply was tested when the system was running, so that the components would be operating at their intended current draw. The regulated positive voltage was measured at +10.01 V, which gives a 0.1% error. The regulated negative voltage was measured at -10.00 V, which gives a zero percent error. The DC-DC step-down voltage for the MCU was measured at 3.300 V, which gives a zero percent error. The DC-DC step-down voltage for the TTL section was measured at 5.01 V, which makes for a 0.2% error.

4 Cost

4.1 Parts

The full parts list can be seen in Appendix E. Table 4 breakdown of cost per block. The total cost of parts is \$186.05

Table 4: Cost per Block Breakdown

Block	Cost (\$)
MCU Stage	24.81
TTL Stage	60.30
Synthesizer Stage	78.43
Power Supply	22.51
Total	186.05

4.2 Labor

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

$$2 * \frac{\$35}{\text{hour}} * \frac{10 \text{ hours}}{\text{week}} * 16 \text{ weeks} * 2.5 = \$28000 \tag{3}$$

This brings the total cost of parts and labor to \$28186.05

5 Conclusion

5.1 Accomplishments

As stated above, we accomplished everything we set out to do. There were a few broken buttons on the user interface board, but otherwise everything works as intended.

5.2 Uncertainties

5.2.1 Fixing the Current Demand Problem

The biggest mid-stream change that we needed to make was to the power section of the project. Basically, we thought linear regulators could handle all of the voltage level shifting involved in combining the VCO, TTL and MCU sections. But we were wrong.

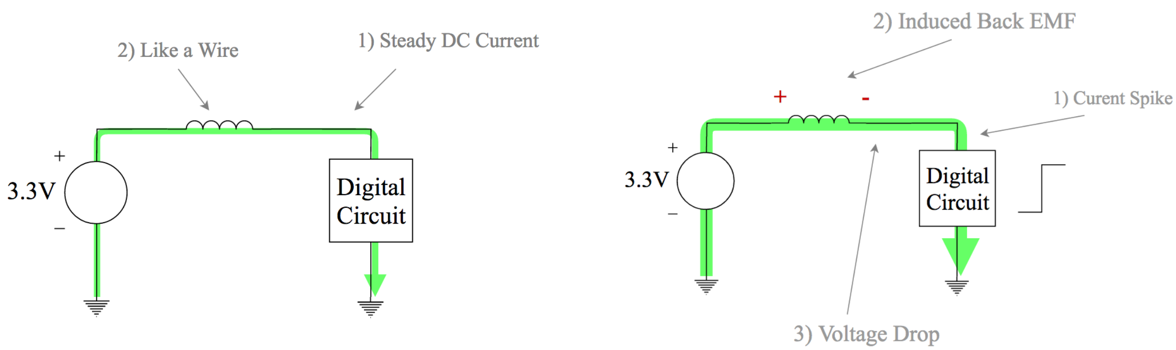
Our initial plan for the +5V and +3.3V lines was to adjust the +12V down to +5V and then the +5V down to +3.3V using linear regulators.

In terms of currents, roughly speaking, at around 10mA or even less, the VCO doesnt draw that much. The TTL stage draws around 60 mA. The real culprit as far as current is concerned is the LED display, which is controlled over I2C by the microcontroller.

Using the linear regulator approach for the +5 V and +3.3 V lines, when displaying 8888 (the combination of numbers that uses the most LED's on the display), the DC current draw was roughly 70 mA and when no numbers were displayed, the DC current draw was about 60 mA. So, the DC current draw was not really the problem. The problem comes from the transient moments when the display is actually switching its LED's from off to on.

Again, using the linear regulator approach, whenever a new tempo was entered by the user, the current demanded by the display to turn on a new combination of LED's was significant. By looking on a scope, we saw shark-tooth-shaped dips on the MCU's voltage line whenever the current demand spiked. These voltage sags weren't even worth measuring or looking further into because clearly, this approach wasn't going to work. At any random one of the voltage drops, the MCU would inevitably turn off and then back on right away, resetting the tempo and starting the sequencer over, which wasn't acceptable.

A visual explanation of what was happening is provided here. Basically, the self inductance of the wires connecting a given subsection to the larger system will set up a back EMF whenever a huge rush of current is demanded by subsection.



(a) DC Current, Inductance Acts Like a Wire

(b) Current Spike, Induced Back EMF

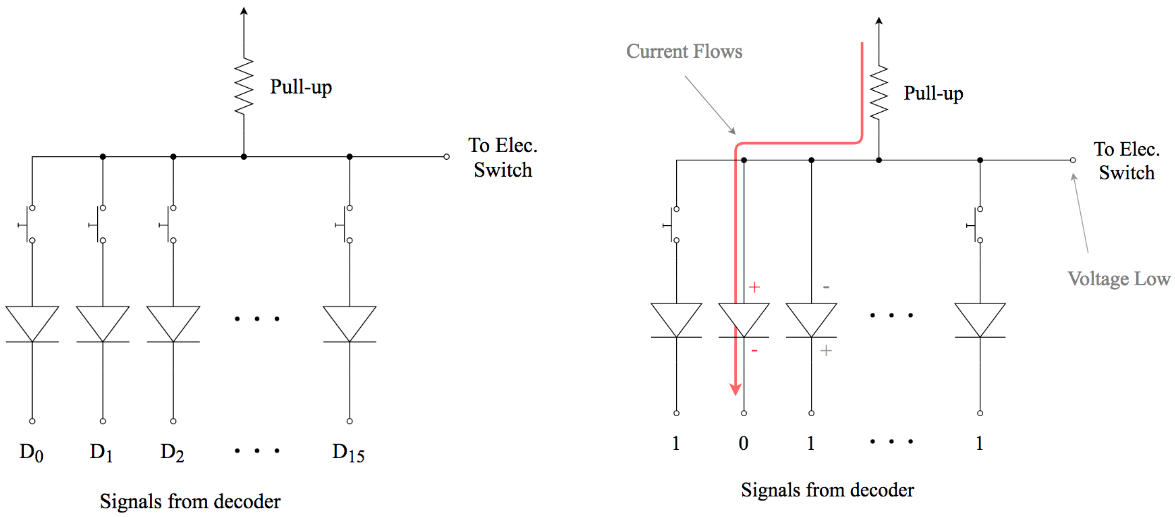
Figure 14: Current Spike Issue in MCU Stage

So, we decided to still use linear regulators for creating the 12 down to 10 V lines but instead use two DC-DC converters for the +5 V and +3.3 V lines. The reason for using DC-DC converters is because they offer much better isolation between sections. The first converter steps 12 V down to 5 V and the second converter steps 5 V down to 3.3 V. This made a huge difference. The shark-tooth-shaped dips in the power line on the MCU were completely eliminated.

5.2.2 Fixing the Design for Skipping Steps in the Melody Sequence

The next big change we had to make was regarding the circuit logic allowing for the user to skip steps in the sequence with the push of a button. Essentially, we just had it wrong. Our design for this feature was never going to work and after one iteration of prototyping this was made obvious. Fortunately, it wasn't too late.

Our solution for skipping steps in the sequence with the push of a button was to use a diode logic circuit. It works like this: each physical step in the sequence has an associated LED above its slider potentiometer. A given LED lights up when the counter driving the decoders force the output connected to that LED low, allowing for current to flow through the LED and turn it on. For the step-skipping circuit, we basically tapped the outputs of the decoders driving the LED's above each step in the sequence. Each of these signals is connected to the cathode of a diode which is connected to an ON-ON SPST switch which is then connected through a shared pull-up resistor to logic high. This is perhaps easier explained using the following figures.



(a) Diode Logic Circuit

(b) Diode Logic Circuit in Action

Figure 15: Diode Logic

The basic idea behind our diode logic circuit is presented in Figure 15a. As explained above, switches are used in series with diodes that have their cathodes connected to the outputs of the decoder IC being driven by TTL counter.

Looking at Figure 15b, we see that the two buttons corresponding to the second and third steps of the melody happen to be pressed, indicating that the user desires to pass over those notes as the melody is swept through. We also see that the note in the melody currently being played corresponds to the second step. Following the explanations indicated in Figure 15b, we see that indeed, the signal routed to the electric switch will go from high to low. This will cause the electric switch to turn off the output driving the speakers thereby skipping the second note in the melody. Still looking at Figure 15b, we also see that although the third button is pushed, the diode connected to this third branch is reverse biased and therefore can be considered an open circuit. This last bit is critical in order to prevent undefined outputs on the decoder.

This solution proved to work exactly as intended.

5.3 Ethical considerations

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].

5.4 Future work

For the next iteration of this project, we would make many small changes and a few big changes.

As far as the small changes go, we would reduce the use of trimpots as much as possible. Having variable resistors makes each build slightly more complicated. It would be better to get the values of those resistors and hard wire those values into the design rather than leave them to be tuned for each build.

Another small change we would make is to use ribbon cables for every inter-connection between sub-circuits. We made four separate PCBs and connecting them ended up taking hours when it could have taken minutes if we had built ribbon cables with proper ports into the design in the first place.

As far as big changes go, the first thing we would do is implement the TTL stage on a programmable logic device like an FPGA chip or perhaps a CPLD chip. This type of implementation would retain the application specific digital logic functionality however, programming it in software would not only streamline the design process, but it would make modifications much easier to do later on, should they be required or desired.

Another big change we would make is to add two additional modules. The first would be a voltage controlled filter (VCF) and the second would be a voltage controlled amplifier (VCA). The VCA would replace the CD4016 electric switch currently used for turning off individual steps in the sequence. Instead, the button to turn off the switch would trigger the VCA to instantly put the output attenuation at a maximum thereby turning off that note. With the VCA instead of a simple electronic switch, the user would have the ability to shape the contour of the sound in time by applying an envelope to the notes in the melody. For example, the volume of each step may be set to rise and/or decay linearly throughout its duration rather than simply turn on and off. As for the VCF, this would provide the user a means to further shape the harmonic spectrum of the waveform in order to create a more musical sound.

Appendix A Requirement and Verification Table

Table 5: System Requirements and Verifications

Requirement	Verification	Verification status (Y or N)
MCU Stage 1. Must be able to convert a user input BPM ranging from 1 to 9999 into a PWM signal measured in Hz within 200 ms. 2. Must output the user defined tempo to within 1%. 3. MCU must be able to successfully read keypad interrupts across I^2C bus communication line. 4. MCU must be able to successfully write to the LED driver across I^2C bus communication line.	1. (a) Use a precision stop-watch to verify that once the pound key is pressed, a PWM signal begins in under 200 ms. 2. (a) Input a tempo into the MCU unit. (b) Connect the PWM output pin to an oscilloscope. (c) Observe the output wave and verify the frequency within 1% of the input tempo. 3. (a) Loop read statements in a while loop using software on the MCU. (b) Use an I^2C decoder on an oscilloscope to verify the read is executed correctly. 4. (a) Loop write statements in a while loop using software on the MCU. (b) Use an I^2C decoder on an oscilloscope to verify the write is executed correctly.	
TTL Stage 1. Button press must perform its task within 50 ms. 2. Potentiometers on the analog MUX inputs are able to create a voltage range from 0 to 5 V.	1. (a) Use a precision stop-watch to ensure the desired outcome based the button pushed is met within 50 ms. 2. (a) Connect multimeter across the potentiometers. (b) Sweep the potentiometers resistance and ensure the multimeter displays voltages from 0 to 5 V.	

Continued on next page

Table 5 – continued from previous page

Requirement	Verification	Verification status (Y or N)
Synthesizer Stage 1. Must achieve one octave of frequency change within 20 cents of accuracy for every one volt of control input. 2. The sawtooth generator within the oscillator must have a reset time of less than 1% the duration of its period. 3. The peak-to-peak value of each of the three waveforms must be 20 V to within 10%. 4. Any trapezoidal component to the squarewave time-domain signal must be less than 1% of the total duration of the signal period. 5. The crossover portion of the rectified sawtooth which forms the triangle waveform must be triangular to within 20%.	1. (a) Start from 0 volts, measure the frequency on the oscilloscope. (b) Change the input to 1 volt and measure the new frequency to ensure it has doubled its previous value to within 20 cents. (c) Iterate this procedure up to 5 volts control input. 2. (a) Hook up the sawtooth generator output to the oscilloscope. Start with the control voltage at 0 V. (b) Measure the reset time using an oscilloscope to verify that it is less than 1% of the duration of the period for that frequency. (c) 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. 3. (a) Set control voltage to 1 V. (b) Connect the oscilloscope to the output of one of the waveshapers. (c) Measure the peak-to-peak voltage and ensure it is within 10% of 20 V. (d) Repeat b. and c. for the other two waveshapers. 4. (a) Set control voltage to 0 V. (b) Connect the oscilloscope to the squarewave waveshaper. (c) Measure the time-axis underneath the reset slopes to verify that it is less than 1% of the total time. (d) Repeat b. and c. for two frequencies per octave covering the VCOs range, up to 5 V in 1 V increments. 5. (a) Set control voltage to 0 V. (b) Connect the oscilloscope to the output of the triangle waveshaper. (c) 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.	

Continued on next page

Table 5 – continued from previous page

Requirement	Verification	Verification status (Y or N)
Power Supply 1. Positive voltages of +12 V and + 10 V should be supplied by LM317T chips to within 5%. 2. Negative voltages of -12 V and -10 V should be supplied by LM337T chips to within 5%. 3. 3.3 V DC-DC converter should supply 3.3 V ±5%. 4. 5 V DC-DC converter should supply 5 V ±5%.	1. (a) Connect multimeter to the output pin of the LM317T set to +12 V. (b) Use a constant current electronic load to apply a current load of 500 mA. (c) Ensure the mulitmeter rating stays within 5% of +12 V. (d) Repeat Step a to c for the LM317T chip set to -10 V. 2. (a) Connect multimeter to the output pin of the LM337T set to -12 V. (b) Use a constant current electronic load to apply a current load of 500 mA. (c) Ensure the mulitmeter rating stays within 5% of -12 V. (d) Repeat Step 1 to 3 for the LM337T chip set to -10V. 3. (a) Connect the output of the multimeter to the output pin of the 3.3 V IC. (b) Use a constant current electronic load to apply a current load of 500 mA. (c) Ensure the multimeter reading stays within 5% of 3.3 V. 4. (a) Connect the output of the multimeter to the output pin of the 5 V IC. (b) Use a constant current electronic load to apply a current load of 500 mA. (c) Ensure the multimeter reading stays within 5	

Appendix B Measured Waveforms

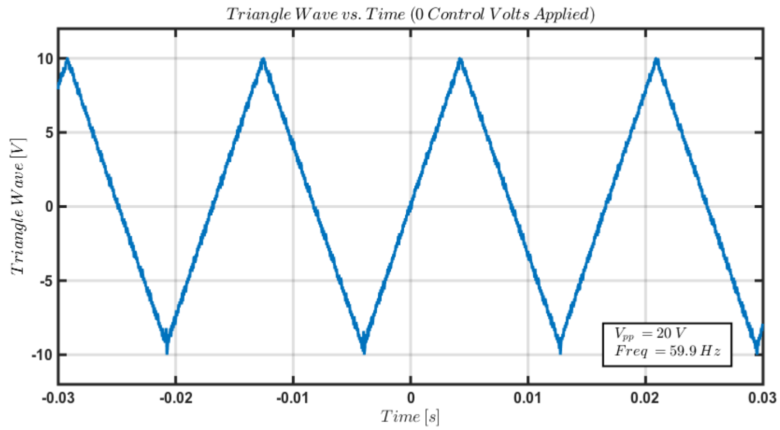


Figure 16: Triangle Wave with 0 Control Volts Applied, Freq. = 59.9 Hz

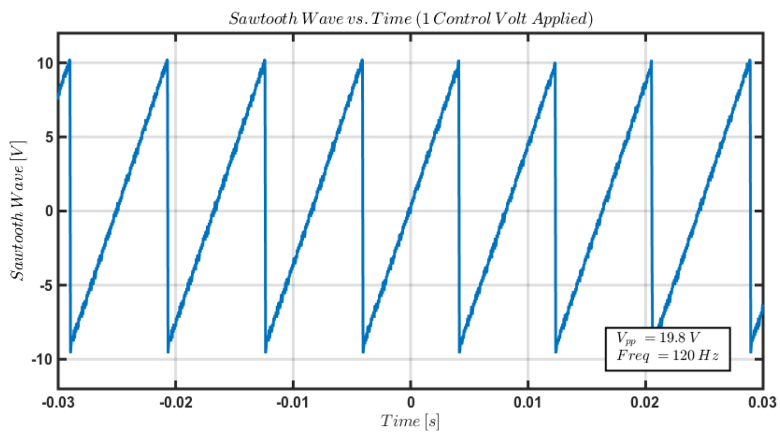


Figure 17: Sawtooth Wave with 1 Control Volt Applied, Freq. = 120 Hz

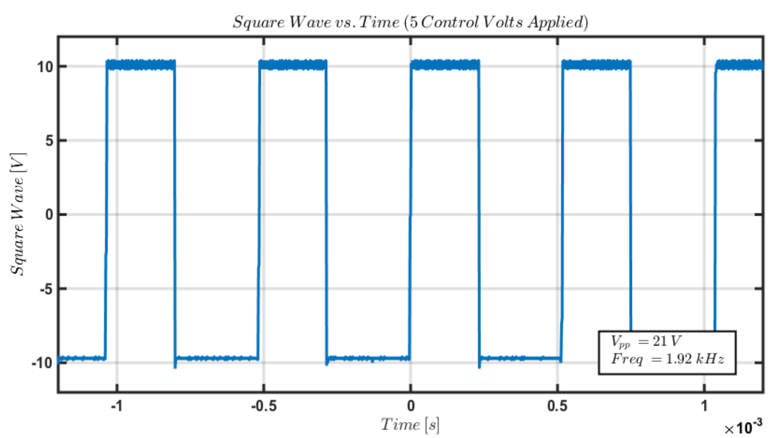


Figure 18: SquareWave with 5 Control Volts Applied, Freq. = 1.92 kHz

Appendix C 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.

C.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

$$i_c = I_S e^{\frac{v_{BE}}{V_T}} \tag{4}$$

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.

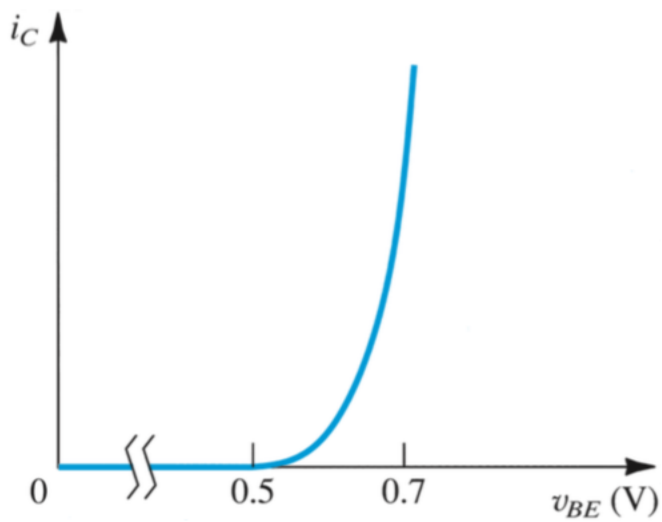


Figure 19: Exponential Collector Current for Linear Base-Emitter Voltage [10]

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 4, 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.

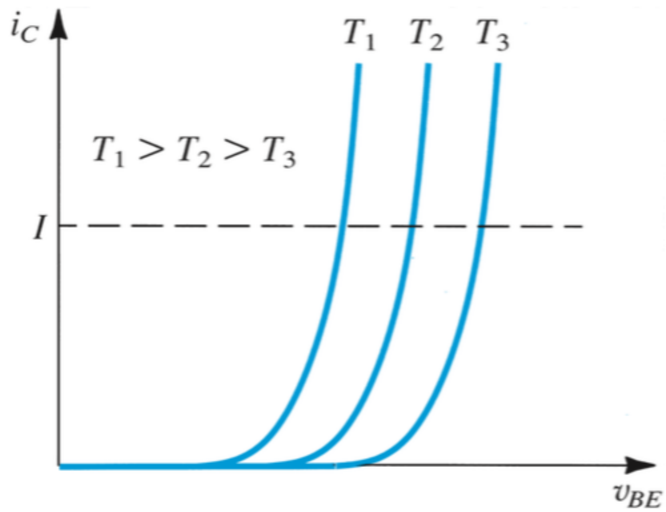


Figure 20: Effects of Temperature on Base-Emitter Voltage [10]

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.

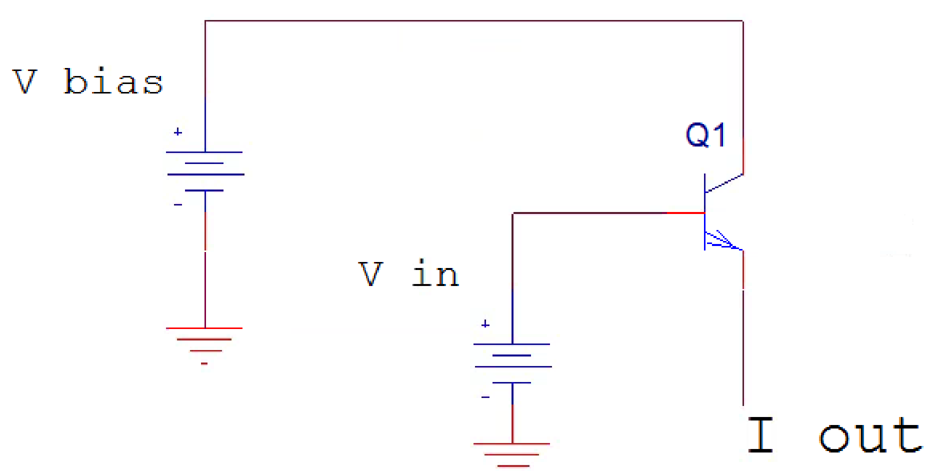


Figure 21: Simple Common-Emitter Based Linear Voltage to Exponential Current

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.

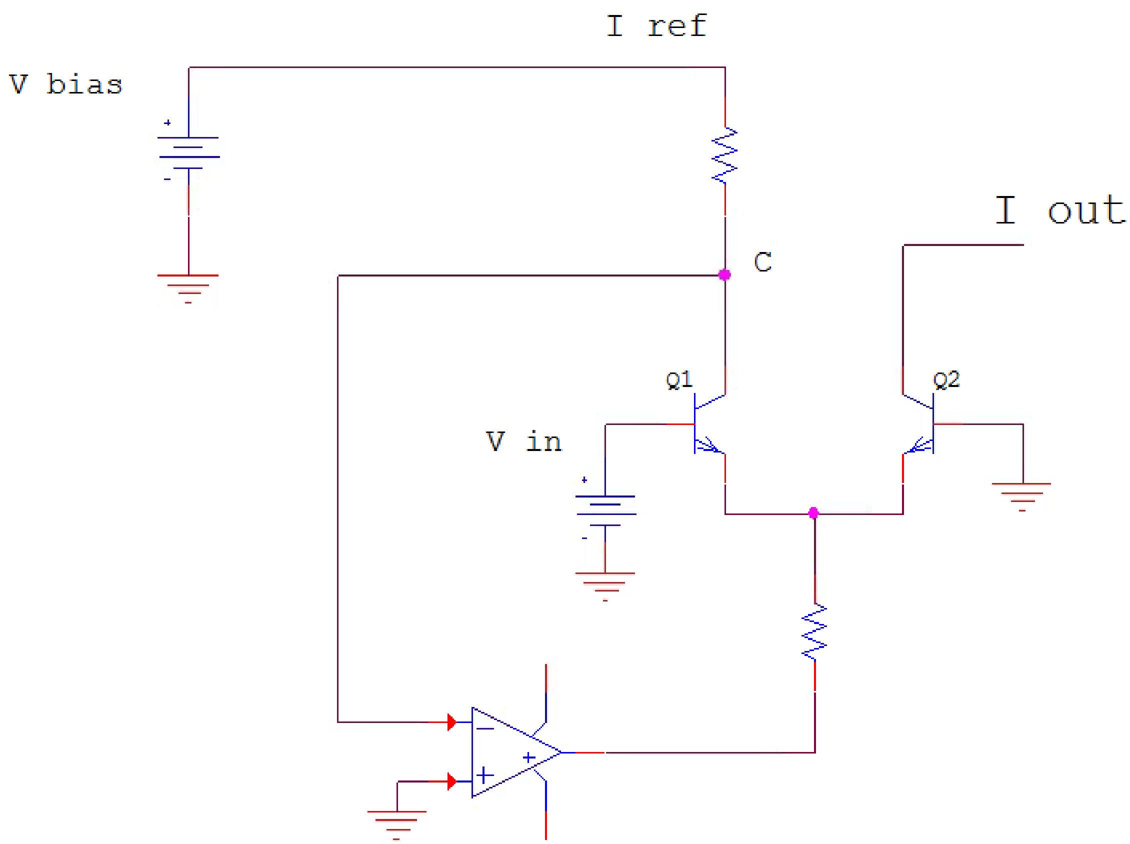


Figure 22: Solution to the Temperature Problem

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

$$V_{IN} - V_{BE1} + V_{BE2} = 0 \tag{5}$$

This can be rearranged to give

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

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

$$I_{ref} = I_S e^{\frac{v_{BE1}}{V_T}} \tag{7}$$

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

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 22. 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.

C.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.

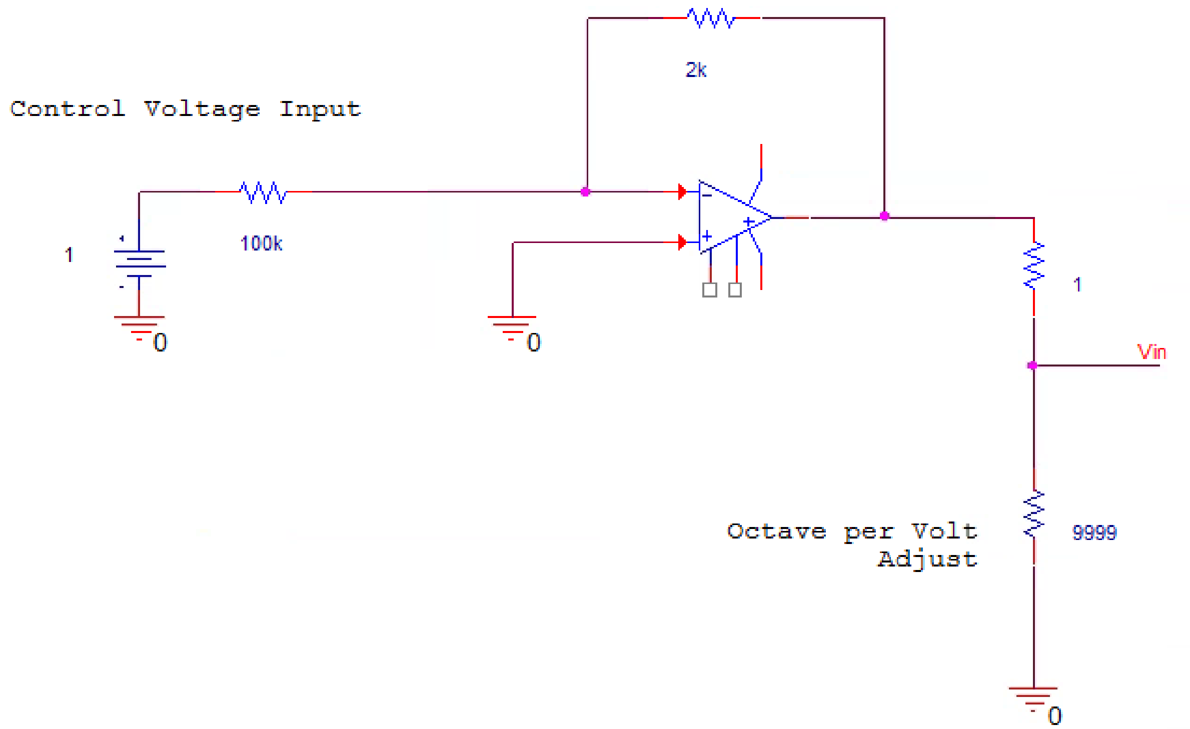


Figure 23: Scaling One Volt Down to Roughly $18mV$

Looking at Figure 23, 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 23. We have the following.

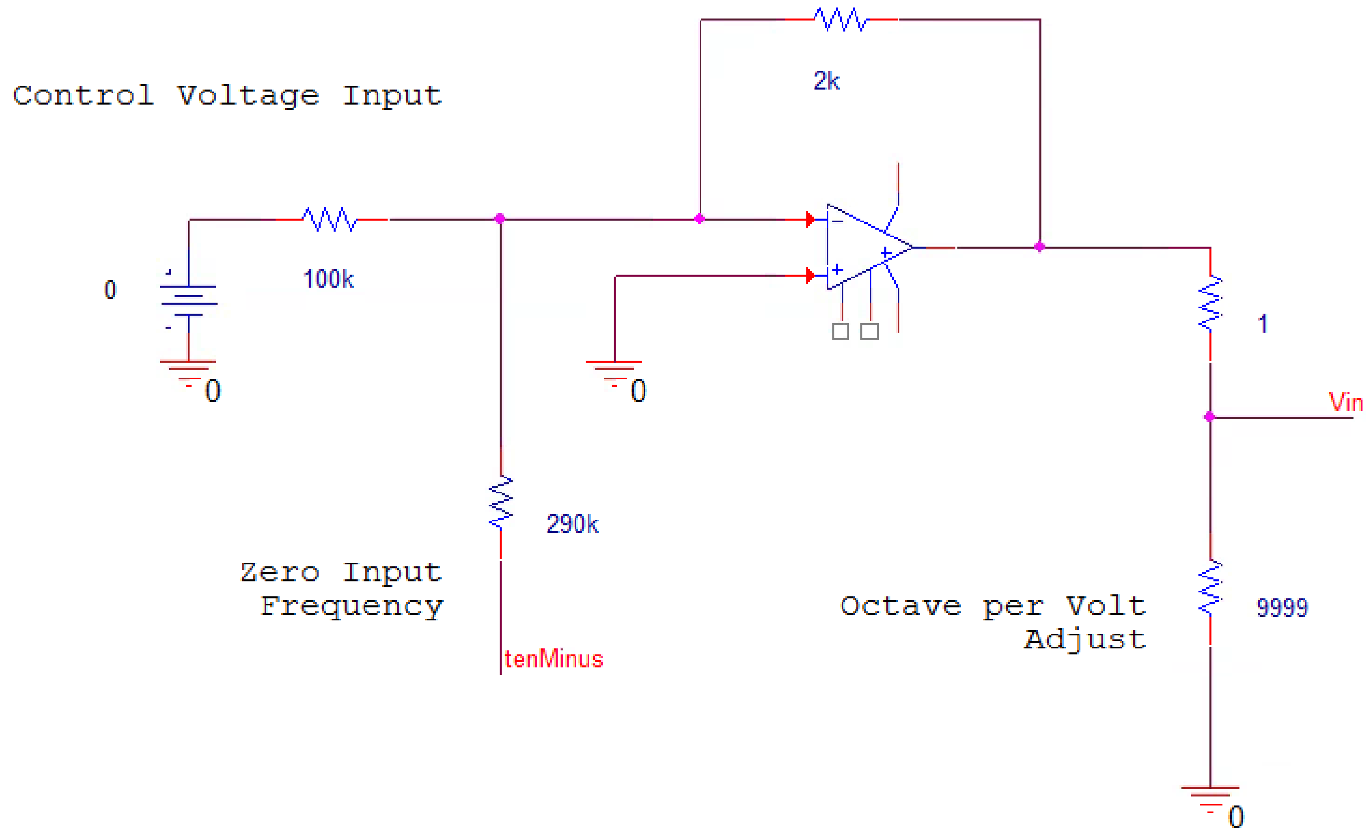


Figure 24: Setting the Base Frequency

The only difference between Figures 23 and 24 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.

C.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.

C.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.

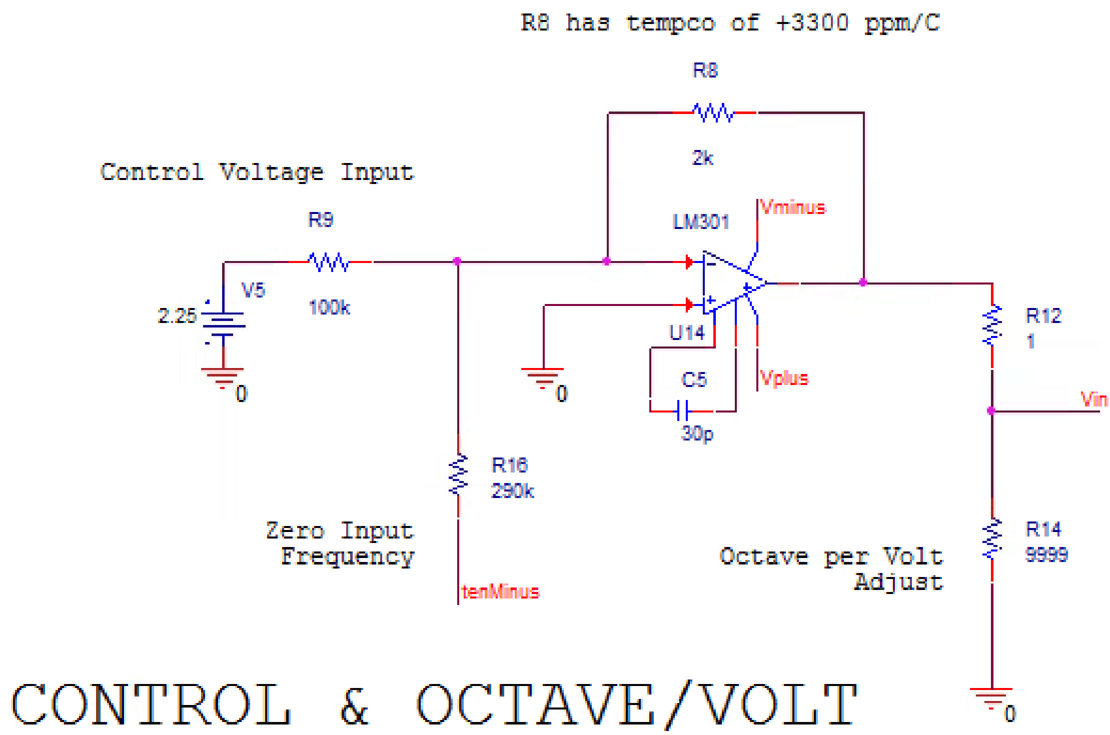


Figure 25: Input Sub-Circuit of VCO

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 $^\circ$ 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 22. 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.

C.5 The Exponentiator Sub-Circuit of the VCO

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

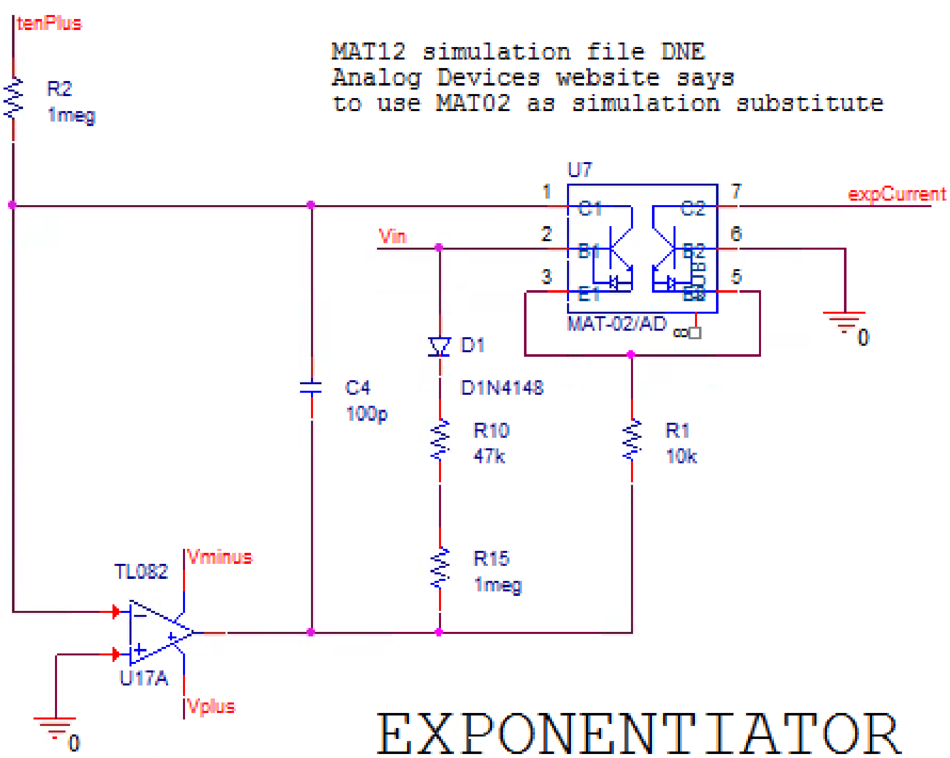


Figure 26: Exponentiator Sub-Circuit of VCO

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 26.

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 4. 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 26. 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 26 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.

C.6 The Oscillator Sub-Circuit of the VCO

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

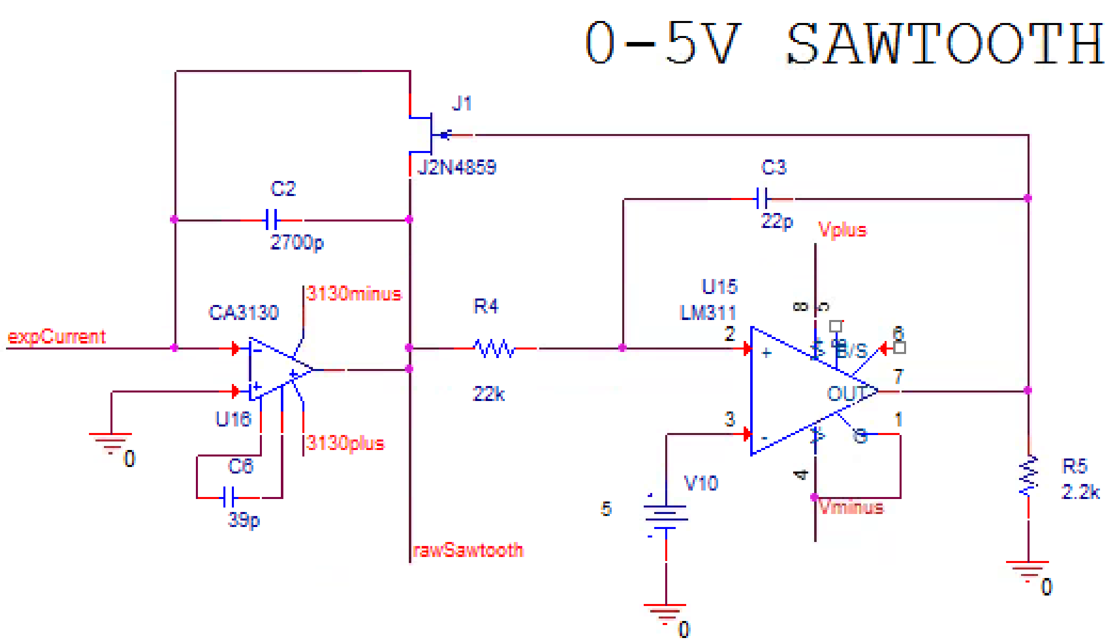


Figure 27: Oscillator Sub-Circuit of VCO

From a high-level perspective, looking at Figure 27, 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.

Appendix D Systematic Procedure Developed for Getting an Octave per Volt from the VCO

The following steps provide a procedure for tuning the VCO. In this example, the base frequency is 60 Hz.

1. Start with a high frequency on the oscillator. For example, put 4 control volts on the input and turn the zero-input trimpot until the frequency is $60Hz * 2^4 = 960Hz$.
2. Set zero control volts on the input and tune the octave per volt trimpot for a frequency of 60 Hz. It is important to have allowed a good amount of wiggle room on the oct/volt trimpot before doing step 1.
3. Check 1 V for 120 Hz, 2 V for 240 Hz, 3 V for 480 Hz and 4 V for 960 Hz.

Appendix E Complete 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%.

Figure 28: Complete Parts List

Appendix F PCB Board Layouts

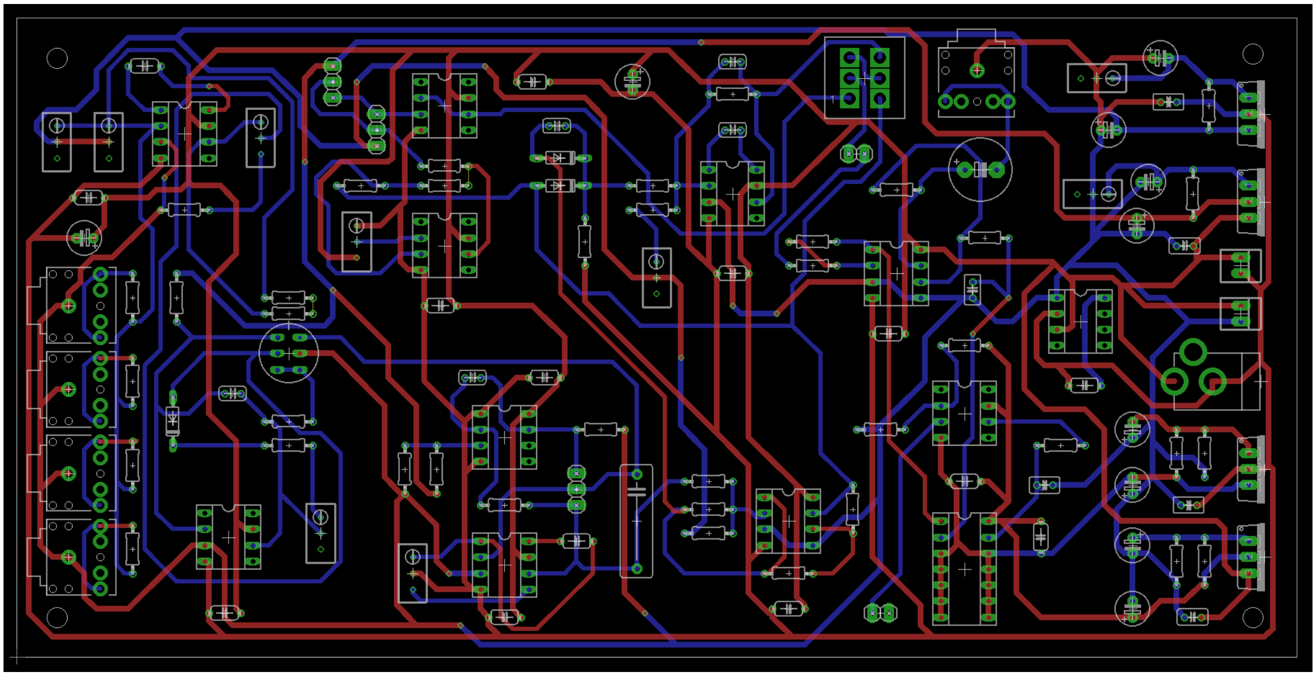


Figure 29: PCB Layout for Analog Components

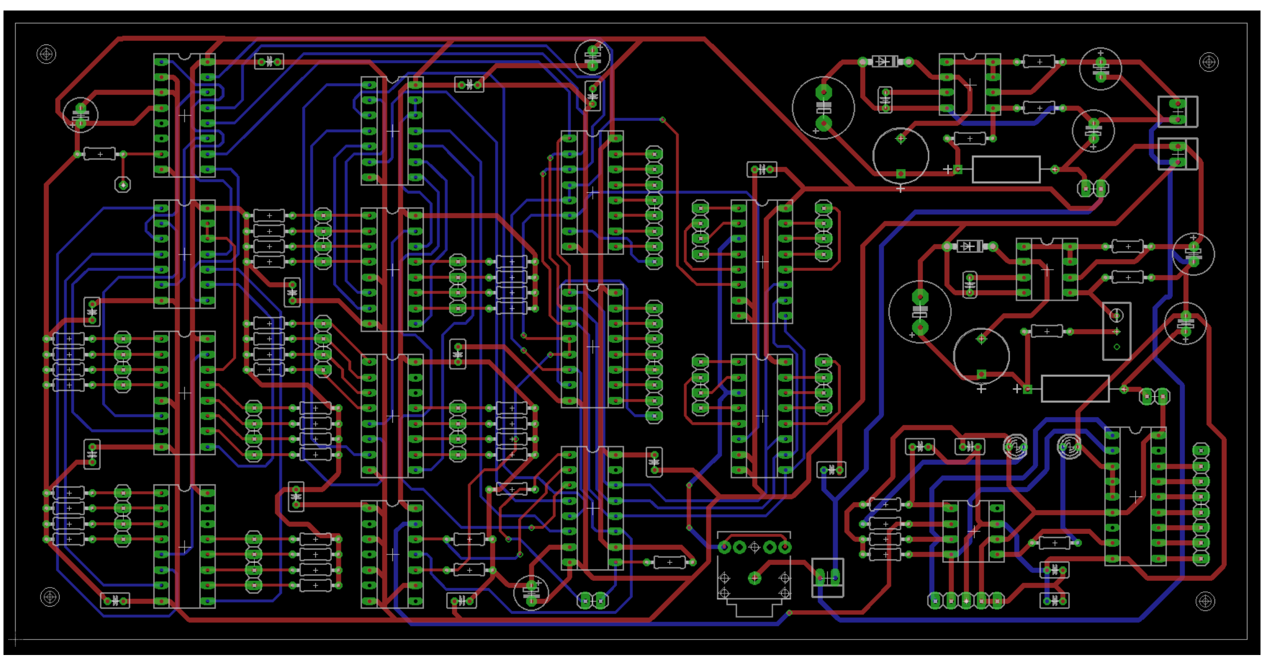


Figure 30: PCB Layout for Digital Components

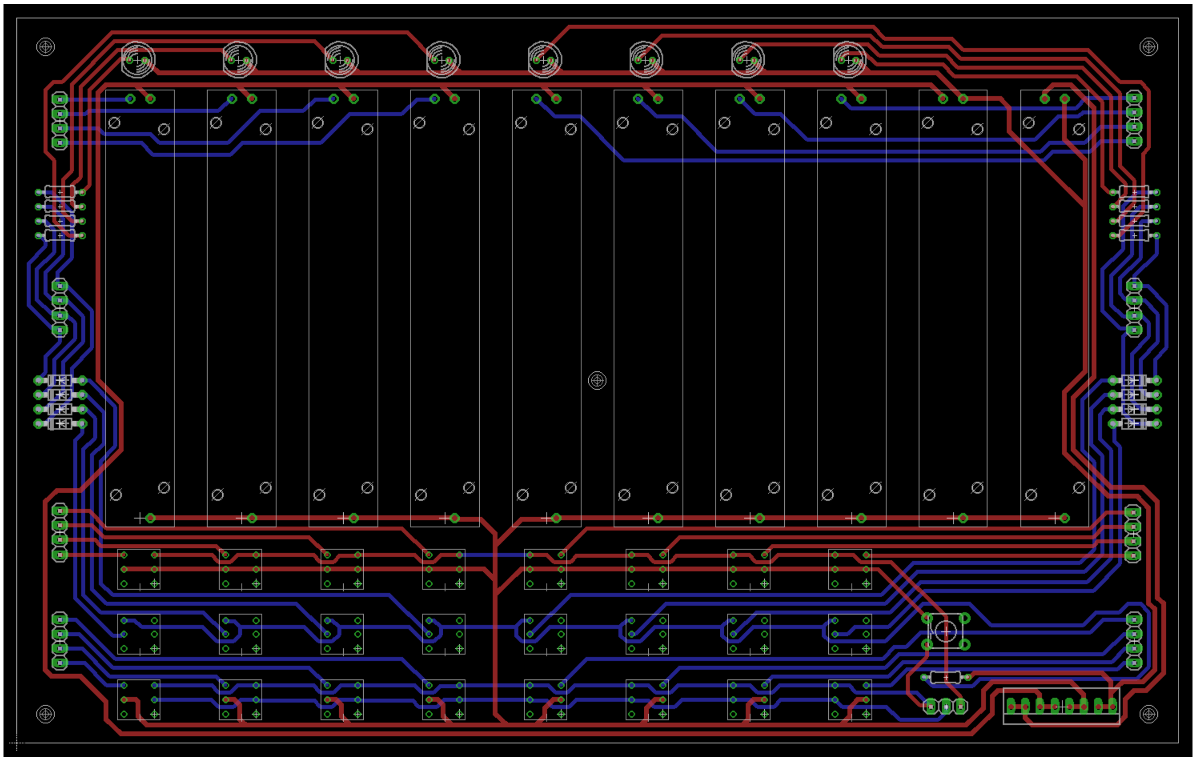


Figure 31: PCB Layout for User Interface Components

References

[1] "Doepfer dark time - step sequencer with blue leds," 2017, Sweetwater. [Online]. Available: https://www.sweetwater.com/store/detail/DarkTimeBL?adpos=1t1&creative=175866237996&device=c&matchtype=b&network=g&gclid=COKLt92u-9ECFQMMaQod_ZIC_A

[2] "Aluminium minimoog voyager." [Online]. Available: https://www.moogmusic.com/products/minimoog-voyagers/aluminium-minimoog-voyager

[3] H. Chamberlin, *Musical Applications of Microprocessors.* United Kingdom: Hayden Bk.Co.,U.S., 07 1981.

[4] R. Schmitz, "A tutorial on exponential convertors and temperature compensation," Apr 2000. [Online]. Available: http://www.schmitzbits.de/expo_tutorial/index.html

[5] T. Instruments, "Hcmos design considerations," sep 2002. [Online]. Available: http://www.ti.com/lit/an/scla007a/scla007a.pdf

[6] O. Semiconductor, "Mc34063a - buck / boost / inverting regulator, switching, 1.5 a," dec 2016. [Online]. Available: https://www.onsemi.com/pub/Collateral/MC34063A-D.PDF

[7] O. Semiconductor, "Theory and applications of the mc34063 and a78s40 switching regulator control circuits," dec 2013. [Online]. Available: https://www.onsemi.com/pub/Collateral/AN920-D.PDF

[8] S. Electronics, "Power and thermal dissipation," 2010. [Online]. Available: https://www.sparkfun.com/tutorials/217

[9] "Ieee code of ethics," 2017. [Online]. Available: http://www.ieee.org/about/corporate/governance/p7-8.html

[10] A. Sedra, *Microelectronic circuits.* New York, NY: Oxford University Press, 2015.