ANALOG SYNTHESIZER WITH AR ENVELOPING

EC412
MAY 5, 2013

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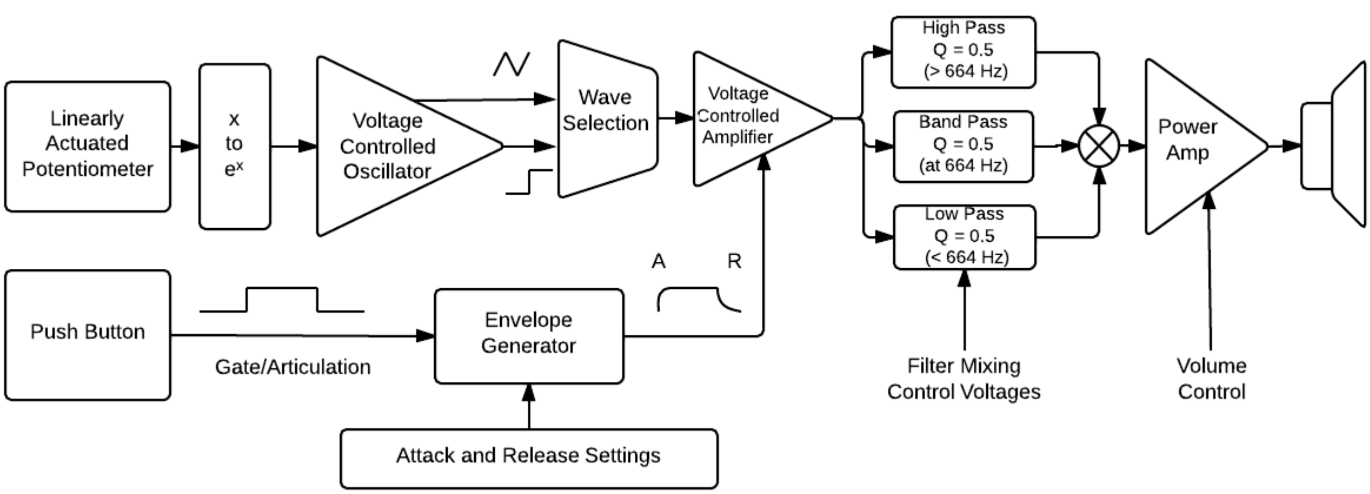
1. Introduction

Building an analog audio synthesizer is an exercise in low speed waveform generation and shaping with high quality needs. For our EC412 project we are creating an Analog Audio Synthesizer with AR (attack-release) enveloping. The goal of this project is to create a useable analog audio synthesizer inspired by a modular synthesizer design. Each component is designed so that with some work they could be used modularly. We made this synthesizer to be a good starting point for future adventures in analog synthesizer design.

Figure 1. Block Diagram.



The design of the synthesizer was broken down into they sub-systems seen in *Figure 1*. Each circuit was designed, simulated and then put onto a breadboard to be verified and connected together. The control voltage to generate the tone is provided by a linearly actuated potentiometer. This control voltage is converted from a linear curve to a $1\frac{V}{octave}$ exponential curve, which is industry standard for analog synthesizers. This control voltage feeds a VCO (Voltage Controlled Oscillator) that generates both a triangle and a square wave, for now only the square wave is used. The generated wave goes through a VCA (Voltage Controlled Amplifier) which receives its control voltage from the Attack-Release envelope generator, triggered by the gate button on the front panel. The resultant, enveloped waveform goes through a high pass, low pass and band pass filter, all with $\omega_0 = 664$ and $Q = 0.5$. The filtered signals are summed together and sent through a power amplifier to 8Ω speakers at about $1W$.

2. Circuits

Throughout the development of the synthesizer we used LTSpiceIV to simulate circuits. The final circuit layout was put together in KiCAD, and open source EDA program. All schematics are in *Appendix A* and will be referenced throughout this writeup by their page number within the appendix. Where appropriate the circuits used for the LTSpiceIV simulations will be provided. LTSpiceIV files are available on request

Figure 2. Picture of the input being tested.

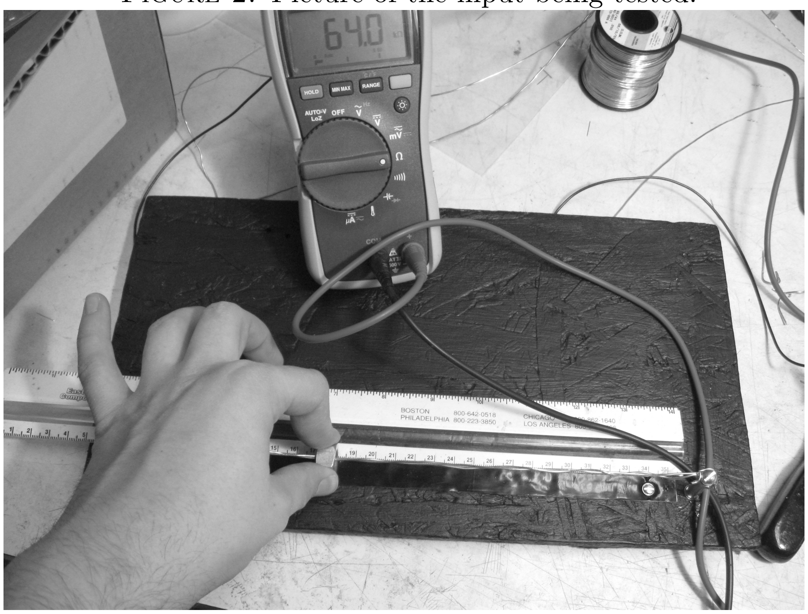
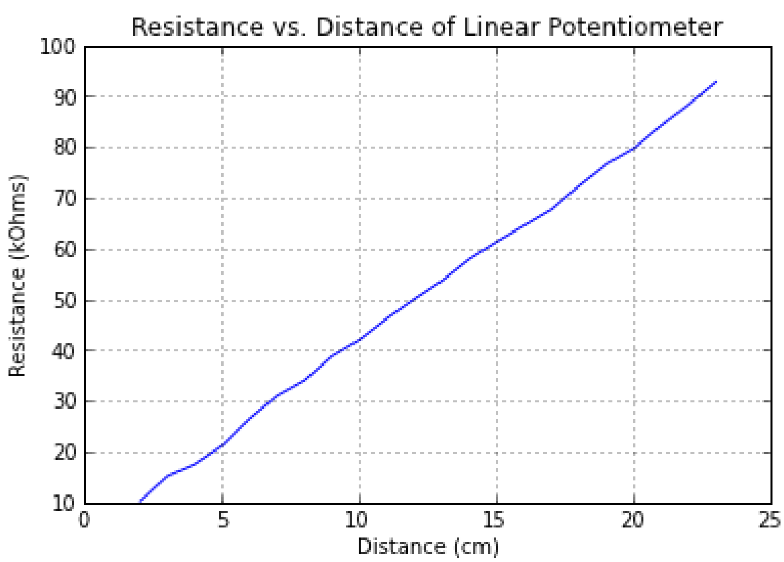


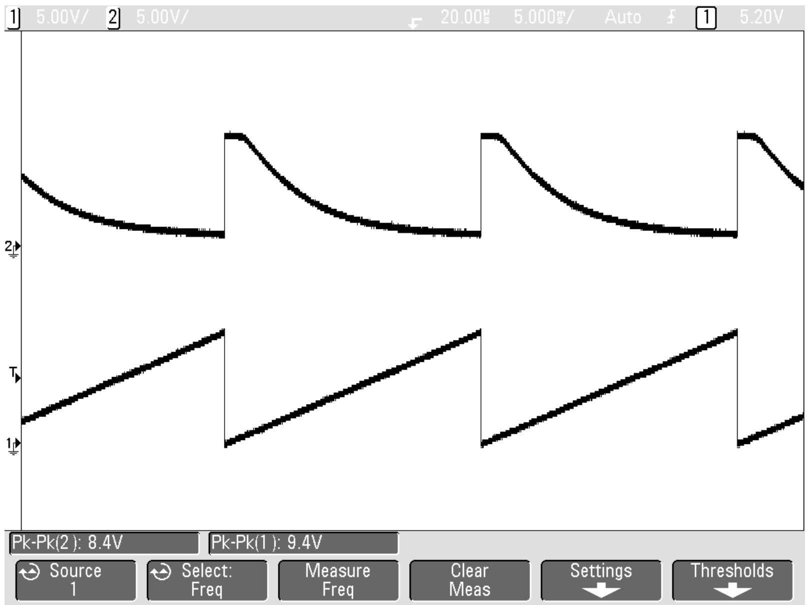
Figure 3. Input Stage Resistance versus Wiper Position.



2.1. **Input Stage.** The input stage of our synthesizer is a ribbon controller. This type of controller is essentially a linearly actuated potentiometer. We used a strip of conductive plastic, often found protecting electronics from ESD during shipping, mounted to a piece of wood and a wiper suspended above it to create a linearly actuated potentiometer. The conductive plastic has a resistance in the $k\Omega$ range. This input stage has been built (Fig. 2) and verified that it has a response which is linear (Fig. 3). The smaller nonlinearities in Figure 2 can be accounted for by low-accuracy wiper positioning measurement and slight variations in the width of the resistive element. This ribbon controller is simple treated as a potentiometer and can be seen in circuit on page *Appendix A-1*.

The gate is just a button hooked up as shown in page *Appendix A-1* such that its on voltage is $4.5V$. The gate is then buffered for the envelope generator.

FIGURE 4. Linear To Exponential Plot. Output on Top, Input on Bottom. Both at 5V/div and 5.0ms/div



2.2. **Linear to Exponential Converter.** In music each octave is 2x the previous octave's frequency, leading to an exponential increase in frequency for octaves. Our linear input stage is not sufficient so we need to convert the linear curve of the input stage to an exponential stage. Our goal was the $1V/Octave$ industry standard; however this goal was never reached due to difficulties in tuning. The circuit used is presented on page *Appendix A-2* and we will assume that $I_{S3} = I_{S4}$ and a DC model of the opamp U1A.

In this circuit we find that the op-amp U1A sources and sinks current at its output to keep Q3's collector current at a constant value, finding $I_{C1} = \frac{9-4.5}{R12} = 4.5\mu A$. The resistor network including R9, R11 and R10 is simply to tune the exponential. We can solve this by invoking Ebers-Moll large signal model with approximation that $I_{C1} >> I_S$. The math follows below:

$$V_{in} = V_{CV}\frac{R_{10}}{R9 + R10} \tag{1}$$

$$V_{in} - V_{BE3} = 4.5\ V - V_{BE4} \tag{2}$$

$$V_{in} - 4.5\ V = V_{BE3} - V_{BE4} = \Delta V_{BE} \tag{3}$$

$$V_{BE3} = V_T ln(\frac{I_{C3}}{I_S}) \tag{4}$$

$$V_{BE4} = V_T ln(\frac{I_{C4}}{I_S}) \tag{5}$$

$$\Delta V_{BE} = V_{in} - 4.5\ V = V_T ln(\frac{I_{C4}}{I_{C3}}) \tag{6}$$

$$I_{C4} = I_{C3} e^{\frac{\Delta V_{BE}}{V_T}} = I_{C3} e^{\frac{V_{in} - 4.5\ V}{V_T}} = I_{C3} e^{\frac{V_{CV} \frac{R_{10}}{R9+R10} - 4.5\ V}{V_T}} \tag{7}$$

The results of the above equations is that the current I_{C3} is exponentially related to tine input voltage V_{in} and its associated resistor network. We then use a PNP, Q5, based inverter circuit which adds some gain and is biased lifted two diode drops off of ground. Through simulation and verification we found that U1A oscillates and goes unstable, to reduce this a feedback capacitor was added between the output of U1A and the negative input of U1A. This experimentally stabilized the system with a C8 of $100pF$. The result can be viewed in Figure 4. For a tunable input stage R9 and R10 should be replaced with a potentiometer.

FIGURE 5. VCO Output Wave Forms (LTSpice IV).

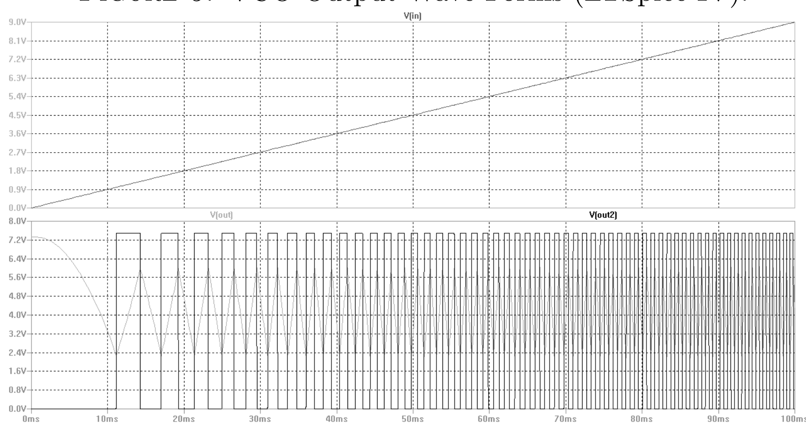
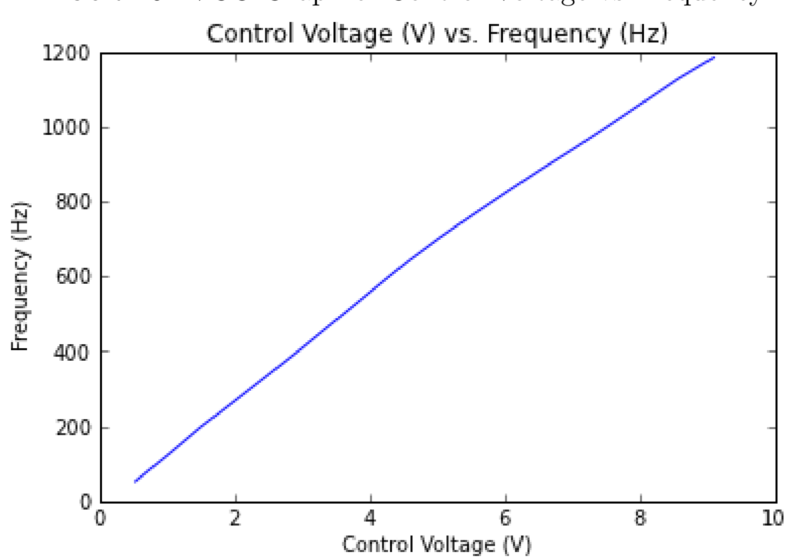


FIGURE 6. VCO Graph of Control Voltage vs Frequency.



2.3. **Voltage Controlled Oscillator.** The VCO schematic was inspired by an application in the LM358 datasheet [3] about using the LM358 as a VCO. The VCO presented on page *Appendix A-4* has two stages an integrator stage and a Schmitt Trigger stage. The integrator integrates the input signal creating a linearly decreasing output whose slope depends on the input voltage. The Schmitt Trigger will eventually trigger turning on the MOSFET Q7 which will force the integrators direction to reverse causing it to increase linearly with the same slope (because R8 is equal to R5/2). The frequency gets set by C11. The output of the VCO vs Control Voltage can be seen in Figure 5 and in Figure 6. In Figure 5 the output is from simulation with LTSpice and in Figure 6 the output is graphed from data taken using an oscilloscope. The results are rather linear and are suitable for our application. However only the square wave output was taken during demo.

Figure 7. VCA Output. Input (Top), Envelope (Middle) and Output (Bottom) (LTSpice IV)

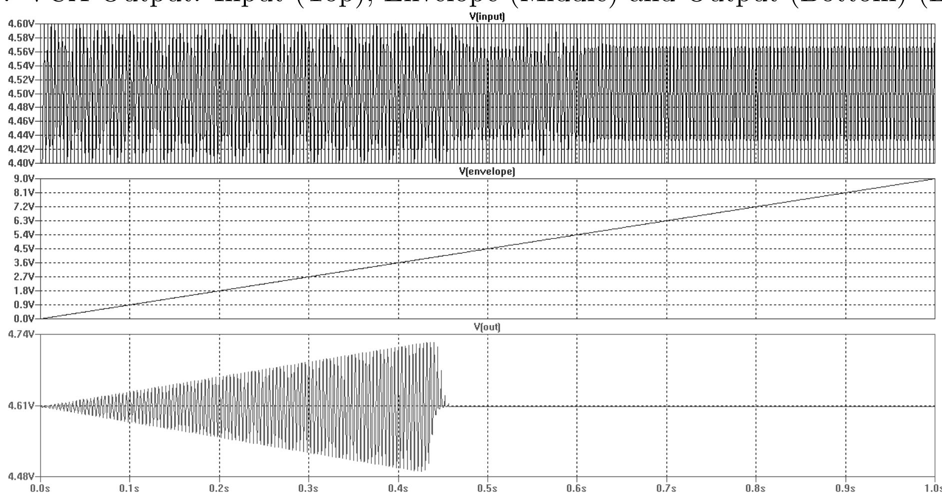
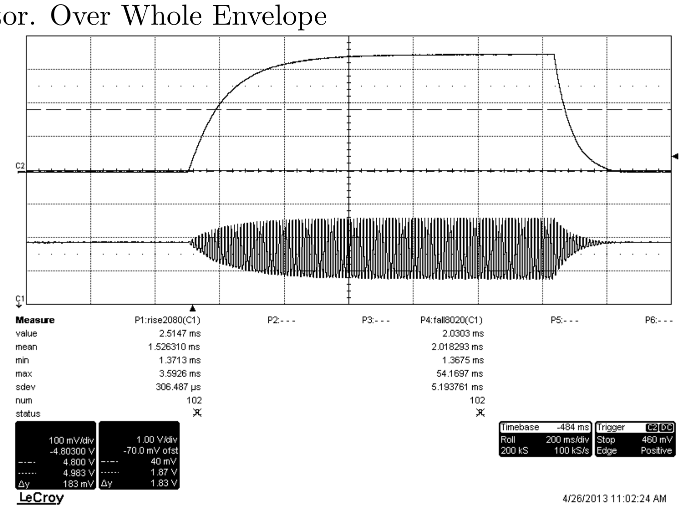
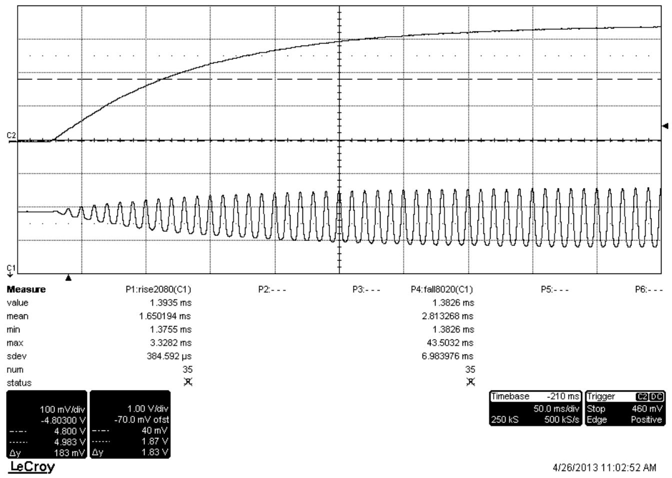


Figure 8. VCA Output w/ Sine Input (200mV peak to peak). Envelope provided by AR Envelope Generator. Over Whole Envelope



2.4. **Voltage Controlled Amplifier.** A voltage controlled amplifier is required to mix the envelope signal with the oscillation created by the VCO. A schematic for this amplifier is on page *Appendix A-5* and is a differential pair whose I_{bias} can be changed by some control voltage. It is important to note that this amplifier only functions within very limited ranges on both V_{in} and CV. As the control voltage CV increases the transconductance, g_m of the transistors increases and as the CV decreases the transconductance likewise decreases. The choice of BJTs is one of convenience. Since the BJT's small signal transconductance

FIGURE 9. VCA Output w/ Sine Input. Envelope provided by AR Envelope Generator. Zoomed In



relationship is $g_m = \frac{I_o}{V_T}$, the amplifier will have a linear gain relationship with CV. A MOSFET based circuit would have a square root based relationship, which could be compensated for by adjusting the voltage to current stage with U3, Q9 and R36.

However, there are some design tradeoffs. First of all if $R35$ and $R37$ are excluded then the circuit distorts very badly, but has a much more linear transfer function (compared to the control voltage). By putting in R35 and R37 we introduce a non-linearity because for small I_{bias} values the transconductance will be small and play a role, but as it grows larger it is removed from the equation and the gain only depends on R35 and R37. The final stage converts the differential output of the VCA to a single ended one. Ignoring R35 and R37 the resulting transfer function becomes:

$$v_{out-dm} = v_{in} * CV * \frac{R33}{R36 * V_T} \tag{8}$$

Once R35 and E37 are added in the resultant equation becomes:

$$v_{out-dm} = v_{in} * CV * \frac{R33}{R36 * V_T + R35 * CV} \tag{9}$$

To help develop the final circuit a combination of experimentation and simulation was used. The goal was to minimize the effect of R35, while still reducing distortion on the output waveform.The results of the simulations can be found in Figure 7, which has been optimized to reduce the distortion by tuning R35 and R37 to a high enough value that they help expand the effective amplification region of Q8 and Q10, while effecting the linearity of the CV-to-gain conversion minimally. If we assume R35 and R37 are small enough and I_bias is also rather small the governing equation becomes $|GAIN| = R33 * \frac{CV}{R36V_T}$. This is not the actual equation, but is close enough to use as an effective guiding equation. Using a 200mV-pp wave the graph in Figure 8 was plotted, providing proof that the VCA is working as expected. However, as we zoom in in Figure 9 the distortion is still visible on the sine wave. Future iterations will likely move to a CMOS VCA which has proven in simulation to have much better distortion characteristics.

2.5. **Envelope Generator.** For analog audio synthesizers envelope generators creates a wave form based on the gate input from the input controller, which is used to change the tone being generated in some way. The envelope generator used here is an A-R, or Attack-Release, envelope generator. It essentially adds an adjustable rise time and fall time to the the gate so you have control of how fast a note comes on. This

FIGURE 10. Measured Outputs from AR Envelope Generator to On-Off Gate ($V_{on} = 4.5V$), RV3 and RV4 set to maximum positions.

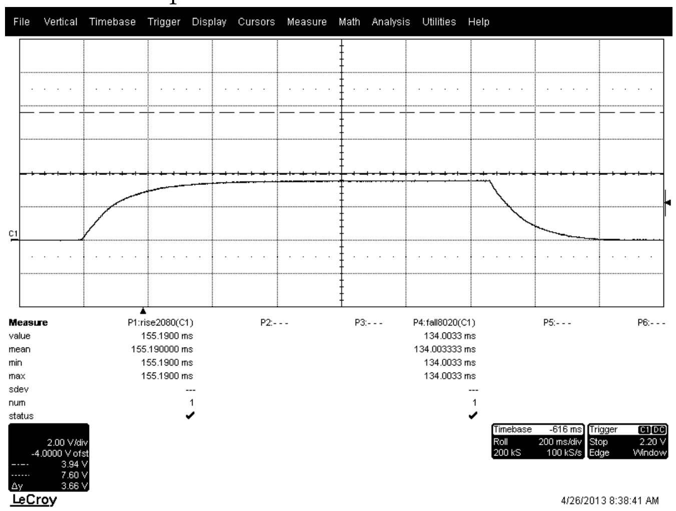
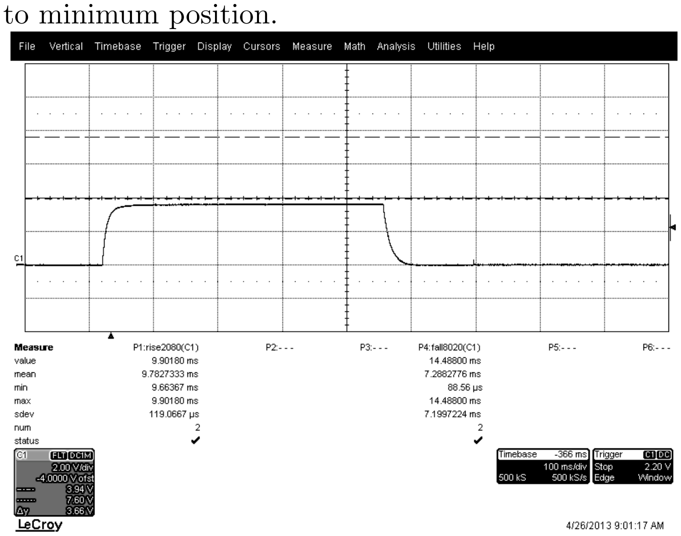


FIGURE 11. Measured Outputs from AR Envelope Generator to On-Off Gate ($V_{on} = 4.5V$), RV3 and RV4 set to minimum position.



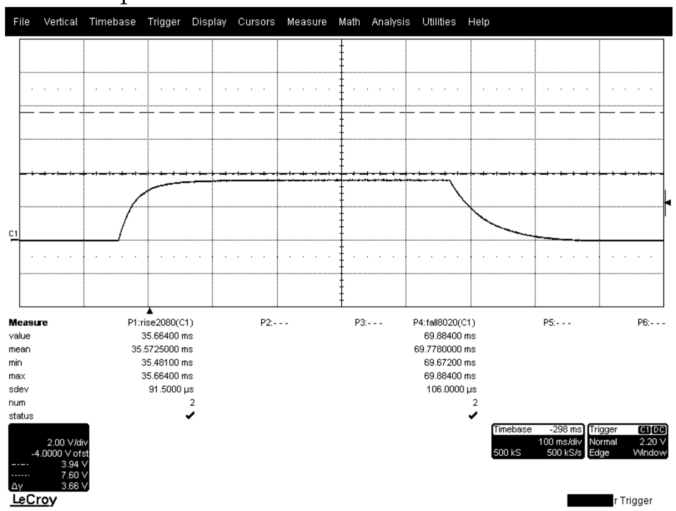
circuit is on page *Appendix A-3* and was developed by Christopher Woodall.

The circuit works by allowing a separate charge and discharge path for capacitor C10. When gate goes high C10 charges through D4, RV3 and R17. This time constant is approximately $\tau = (RV3 + R17) * 1uF$, meaning that the minimum time constant for 100k potentiometer is 1ms and the maximum time constant is 100ms. When gate goes low, the capacitor will discharge through D5, RV4 and R17, the time constant equation is the same with RV4 substituted for RV3 and the time constant limits, assuming a 100k potentiometer, are the same. Experimentally we have found a slight variation from this scheme, but it is close enough for our low precision purposes.

The results of this circuit when both RV3 and RV4 are set to minimums can be found in Figure 11. Similarly the results for the maximum time constants can be found in Figure 10. Since the two potentiometers are separate mixed rise and fall times can be achieved as see in 12.

2.6. **Filter Stage.** The output stage begins by buffering and inverting the output signal of the voltage controlled amplifier. ALl of the circuits for the output stage can be found on page *Appendix A-6*. The

FIGURE 12. Measured Outputs from AR Envelope Generator to On-Off Gate ($V_{on} = 4.5V$), RV3 and RV4 set to mixed positions.



initial design had us only buffering here, but due to noise and general attenuation too much of the original signal was lost by the time we got to the power amplifier. The amplification is done through a simple 6dB inverting amplifier. The output of the inverting amplifier gets fed into a modified Linkwitz-Riley crossover filter.

According to Microelectronics Circuits and Devices by Horenstein the quality factor of a generic second-order low-pass is: $Q = \sqrt{\frac{C_1}{C_2}} * \left(\frac{\sqrt{R_1*R_2}}{R_1+R_2}\right)$. The quality factors for a generic secon-order high-pass, and band-pass filters of Sallen-Key type is $Q = \sqrt{\frac{R_1}{R_2}} * \left(\frac{\sqrt{C_1*C_2}}{C_1+C_2}\right)$. The cutoff frequencies for all of these filters is: $\omega_0 = \frac{1}{\sqrt{R_1*R_2*C_1*C_2}}$.

The Linkwitz-Riley crossover filter is a popular configuration in mixing audio filters and is the basis for our Analog Synthesizer's filters. A Linkwitz-Riley crossover attempts to create a flat 0dB response from the sum of the low-pass and high-pass filters. It achieves this by setting the values for both filters to R1=R2=R and C1=C2=C. This results in equal crossover frequencies, as well as, Q factors in Eq. 10 for the low-pass filter and Eq. 11 for the high-pass and band-pass filters. The cutoff frequencies can now be solved with Eq. 12.

$$Q = \sqrt{\frac{C_1}{C_2}} * \left(\frac{\sqrt{R_1 * R_2}}{R_1 + R_2}\right) = \sqrt{\frac{C}{C}} * \left(\frac{\sqrt{R * R}}{R + R}\right) = 1 * \frac{R}{2R} = .5 \tag{10}$$

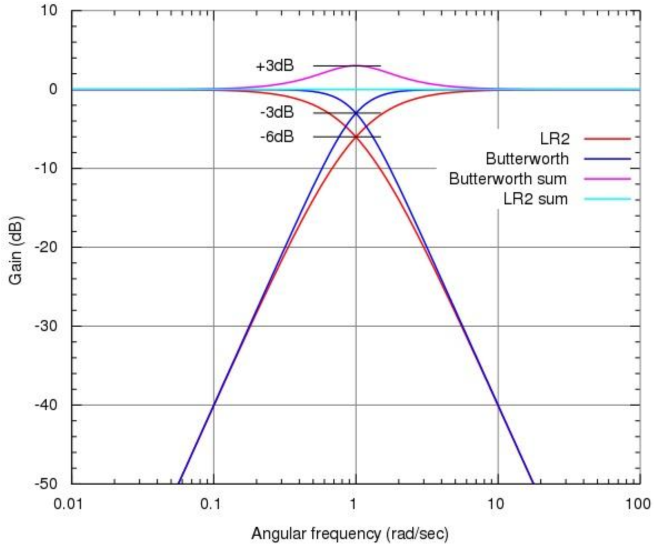
$$Q = \sqrt{\frac{R_1}{R_2}} * \left(\frac{\sqrt{C_1 * C_2}}{C_1 + C_2}\right) = \sqrt{\frac{R}{R}} * \left(\frac{\sqrt{C * C}}{C + C}\right) = 1 * \frac{C}{2C} = .5 \tag{11}$$

$$\omega_0 = \frac{1}{\sqrt{R_1 * R_2 * C_1 * C_2}} = \frac{1}{\sqrt{R * R * C * C}} = \frac{1}{\sqrt{R^2 * C^2}} = \frac{1}{R * C} \tag{12}$$

For the values we implemented: R=5.1k, C=47nF, we get ?0=664 Hz. A normal Linkwitz-Riley filter only sums low-pass and high-pass filters, creating a perfect 0dB crossover region, an example shown in Figure 13.

Since the main purpose of our filter network was not to create different volumes of audio at various frequencies but was instead to shape our input signal to different forms to create different sound effects. Because of this we included a band pass filter. This created a slight increase in the center frequency gain. At the center frequency, the gain of the bandpass filter is given in Eq 13.

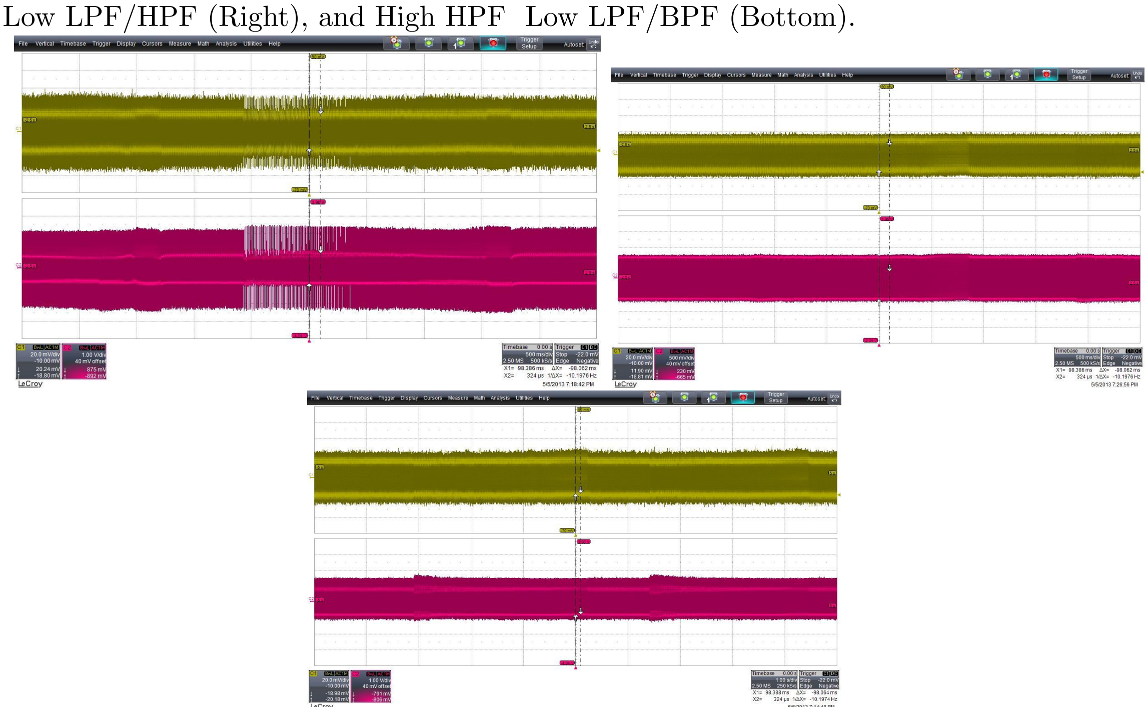
FIGURE 13. Linkwitz-Riley Crossover Graph [4]



$$\left|\frac{V_{out}}{V_{in}}\right|_{\omega=\omega_0} = \sqrt{\frac{R_2 * C_2}{R_1 * C_1}} * Q = \sqrt{\frac{RC}{RC}} * Q = Q = .5 \tag{13}$$

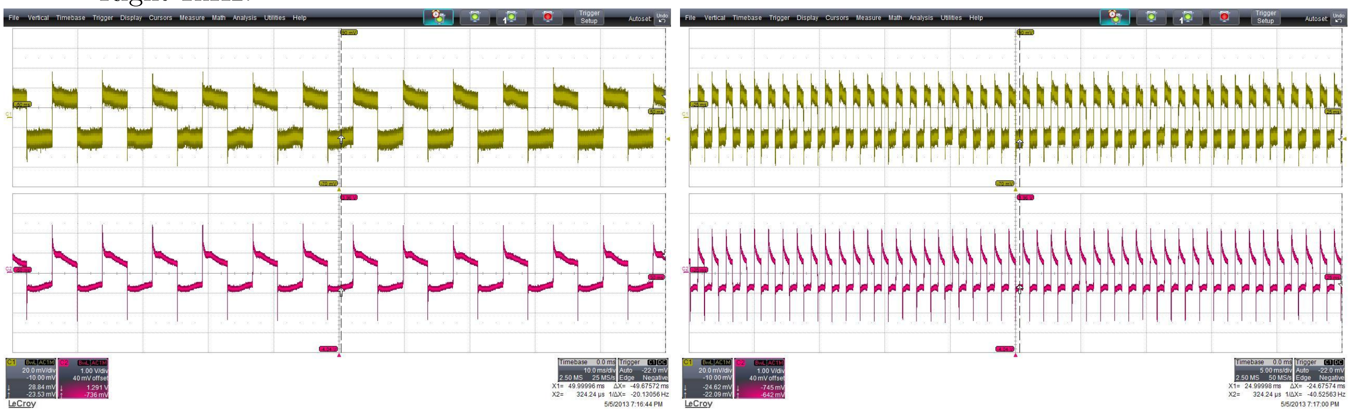
This creates an additional peak at -3dB from the band-pass filter. The sum of the three filters create a crossover peak at 3.5dB. Although this deviates from the ideal 0dB crossover behavior, it is unavoidable and considerably better then the response of the common alternatives, such as the Butterworth or Chebyshev configurations. Due to this there is a mostly flat response over the entire frequency range of our synth. Below are a frequency sweep for the sum of High LPF - Low BPF/HPF, High BPF Low LPF/HPF, and High HPF Low LPF/BPF, in that order in Figure 14

FIGURE 14. Frequency sweep for the sum of High LPF - Low BPF/HPF (Left), High BPF Low LPF/HPF (Right), and High HPF Low LPF/BPF (Bottom).



The waveform produced from high HP, medium BP, and medium LP is shown in Figure 15 at 180Hz and 1kHz.

FIGURE 15. Waveform produced from high HP, medium BP, and medium LP. Left 180 Hz, Right 1kHz.



The output of all three filters is fed into potentiometers which are then mixed in a summing amplifier. All three filter potentiometers are $10k\Omega$, while R_f is also a $10k\Omega$ resistor. R_f in this case controls the final volume of the synthesizer. The output of the amplifier is given by $V_{out} = -R_f * \left(\frac{V_{lpf}}{R_{lpf}} + \frac{V_{bpf}}{R_{bpf}} + \frac{V_{hpf}}{R_{hpf}}\right)$. Considering just the peak to peak voltages and not the actual waveform Vlpf and Vhpf can be written as .2V with their gain of 0dB. With a gain of -3dB, Vbpf can be written as .1V. This results in a Vout equation of: $V_{out} = -R_f * \left(\frac{.2V}{R_{lpf}} + \frac{.1V}{R_{bpf}} + \frac{.2V}{R_{hpf}}\right)$.

Seeing this it becomes obvious that all three signals get mixed together in such a way that the actual values for any given filter potentiometer doesn't affect the output volume by much (A notable exception occurring near the lowest resistor values) When you increase/decrease one of the pots the denominator in turn is also increased/decreased. Because of this R_f becomes the predominant controller of the output voltage while R_{lpf}, R_{bpf}, and R_{hpf} simply control how much of any given filter is in the output signal. Figure 16 shows various output signals for different levels of R_{lpf}, R_{bpf}, and R_{hpf}, with a constant R_f.

The current level of the R_f resistor determines the peak to peak voltage of the signal and therefore the output volume of the speakers. In Figure 17 is the results of a volume sweep up and down on R_f:

The output of the summing filter goes to the power amplifier, which is used to drive the speakers. A class AB amplifier was used, based on the minimally biased Class AB amplifier from Horenstein's book. For this configuration, $V_{out} \approx V_{in}$. The output of the of the summing amplifier is able to produce about 3V peak to peak.

With V_{out} going across the speaker, which is 8Ω and $max(V_{out}) \approx 3V$ we get $P = \frac{V^2}{R} = \frac{9V}{8\Omega} \approx 1.1W$. The final maximum output power is roughly 1.1W, which is plenty loud enough for our needs.

2.7. **Power.** For our project we decided to design the synthesizer to work on $9V$ single supply, as opposed to the standard $\pm 15V$ often used in Analog Synthesizer applications. To obtain the $9V$ supply we used a $12V$ wall-wart to go from AC to $12V$ DC and then a standard LM317 adjustable regulator circuit (found in *Appendix A-7*). The LM317 circuit, was pulled from the LM317 datasheet [1], and once the RV1 is adjusted will provide $9V$.

Using a single supply required us to derive a $4.5V$ voltage reference which could handle current draws up to 200mA. To do this we chose to use a basic class AB amplifier driven by a TLC08x BiCMOS op-amp [2]. This configuration allows for a high current draw by using a TIP31 and TIP32 to provide the actual

FIGURE 16. Output signals for various levels of R_{lpf}, R_{bpf}, and R_{hpf}. R_f remains constant.

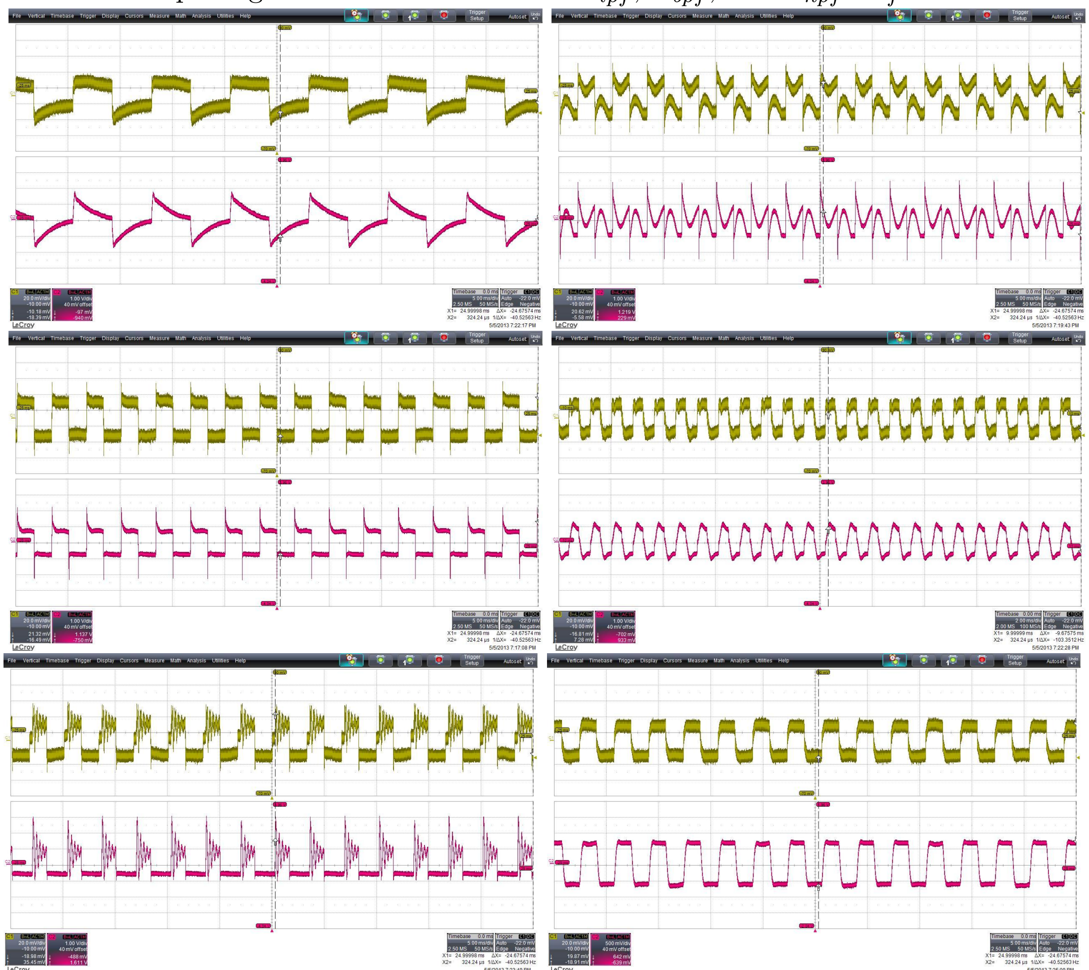
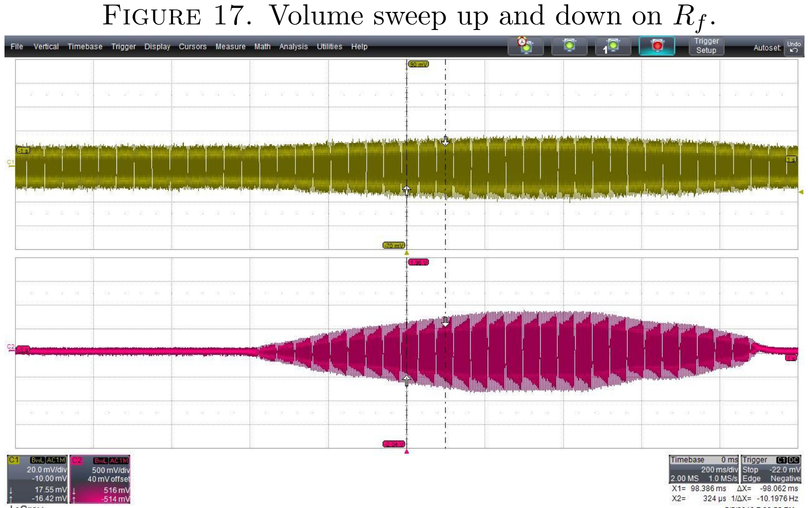


FIGURE 17. Volume sweep up and down on R_f.



current to whatever is attached to the virtual ground point. Furthermore capacitors C6 and C7 are added to help stabilize the 4.5V point and provide some additional current for short bursts. This circuit can be found at on page *Appendix A-7*. An on off LED and a switch are also present for convenience.

3. Conclusion

At the end of this project playable instrument was produced which could output filtered filtered square waves over a range from 100 Hz to 1000Hz, covering approximately 3 octaves of effective playing. Furthermore, effective and adjustable AR enveloping was produced, which could be mixed with the VCO generated waves to produce an output to send to the filter stage. Furthermore, the device was made independent from the lab supplies thanks to a power supply network created from a wall-wart and some additional supplies.

While the creation of a playable musical instrument was a success some improvements could be made. For example, the linear to exponential stage could use a better tuning network to achieve 1V/Octave, industry-standard, performance. This was not the only case of bad potentiometer choice: the filter summing stage could have used a better potentiometer setup to allow for easier use of the effective range of the summing amp. There were other features which got removed due to the reduction of our team from 3 team members to 2, namely, moving from an ADSR (Attack, Decay, Sustain and Release) envelope to a simple AR envelope generator. Other circuits, such as the VCA could have been improved with further iterations. Furthermore, a more modular and less error-prone design could have been implemented using PCB's, or protoboard.

Even with the problems listed above, this project was a decisive success considering the primary goal was to produce a base platform for future work in analog synthesizer design. With a VCO, VCA and output stage design present more effects and stages can be added to create more complex sounds. We plan on creating PCBs of the various sections over the summer and improving our design for our own entertainment.

References

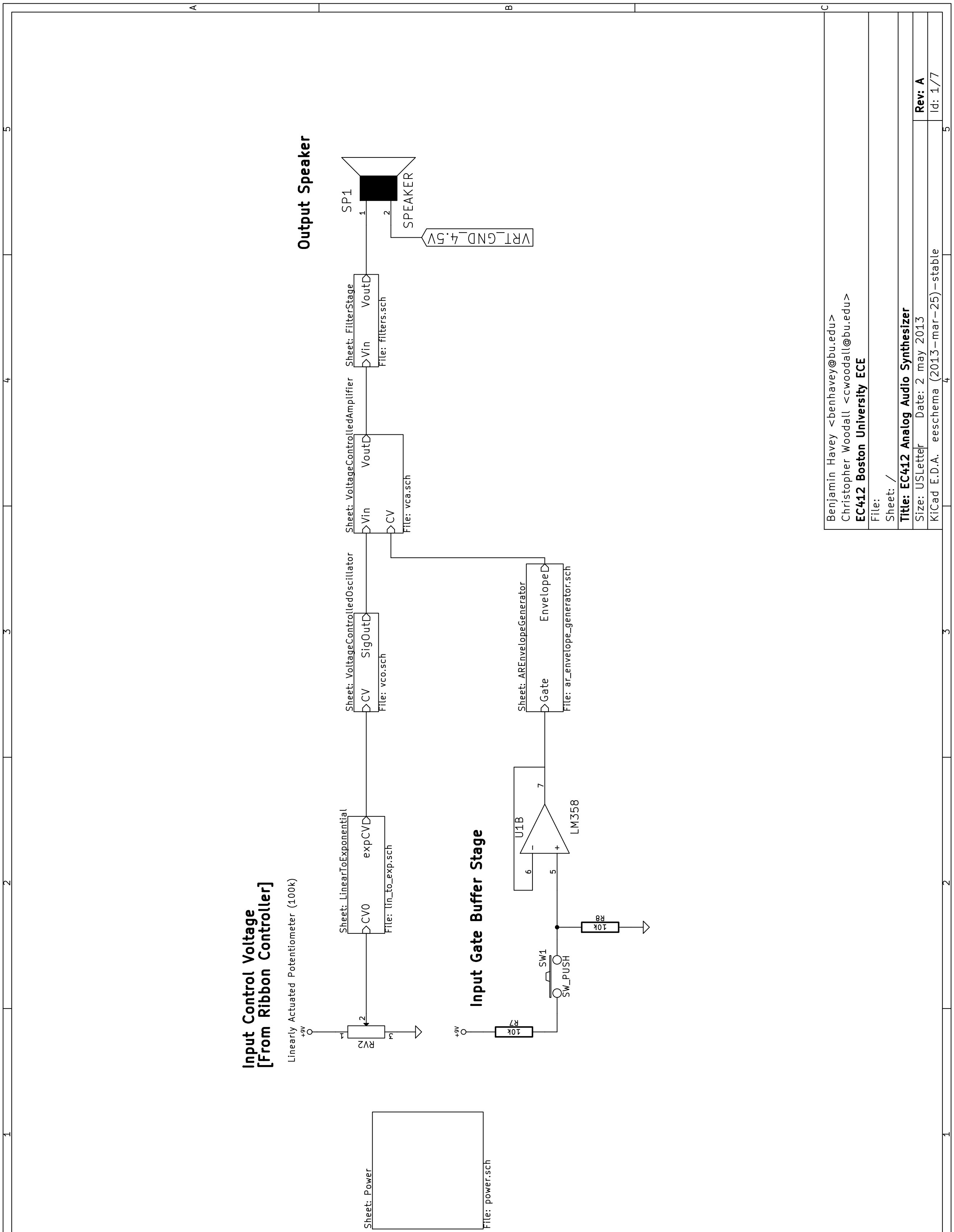
[1] Texas Instruments, "LM117/LM317A/LM317-N 3-Terminal Adjustable Regulator," LM317 datasheet, May 2004 [Revised Feb. 2011].

[2] Texas Instruments, "Family of Wide-Bandwidth High-Output-Drive Single Supply Operational Amplifiers," TLC08x datasheet, June 1999 [Revised Nov. 2000].

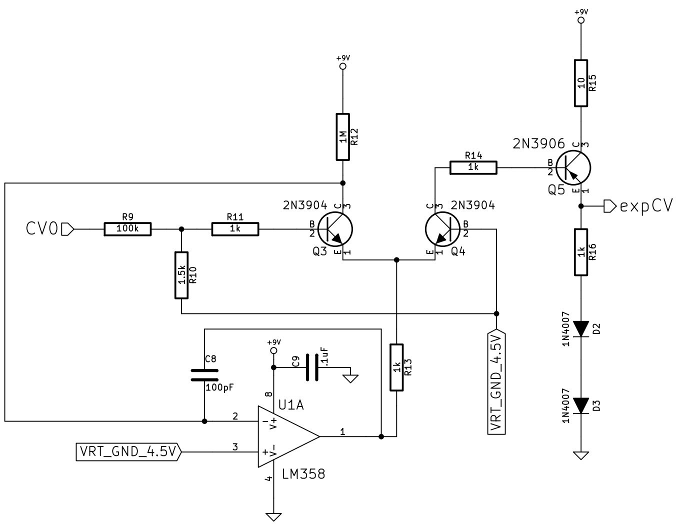
[3] Texas Instruments, "LM158/LM258/LM358/LM2904 Low Power Dual Operational Amplifiers," LM358 Datasheet, January 2000 [Revised Mar. 2013].

[4] Wikipedia, "LinkwitzRiley filter", May 5, 2013.
`http://en.wikipedia.org/wiki/Linkwitz-Riley_filter`

4. Appendix A: Schematics



Linear To Exponential Stage



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File:
Sheet: /LinearToExponential/

Title: EC412 Analog Audio Synthesizer

Size: USLetter Date: 2 may 2013 **Rev: A**

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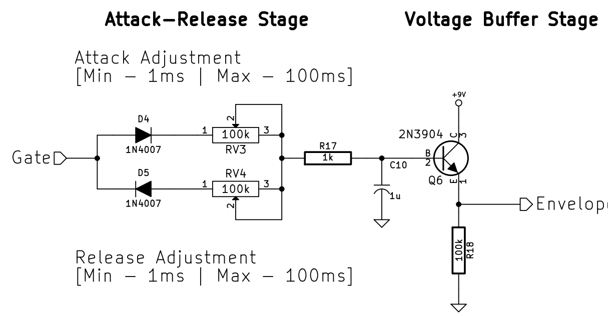
AR Envelope Generator Circuit



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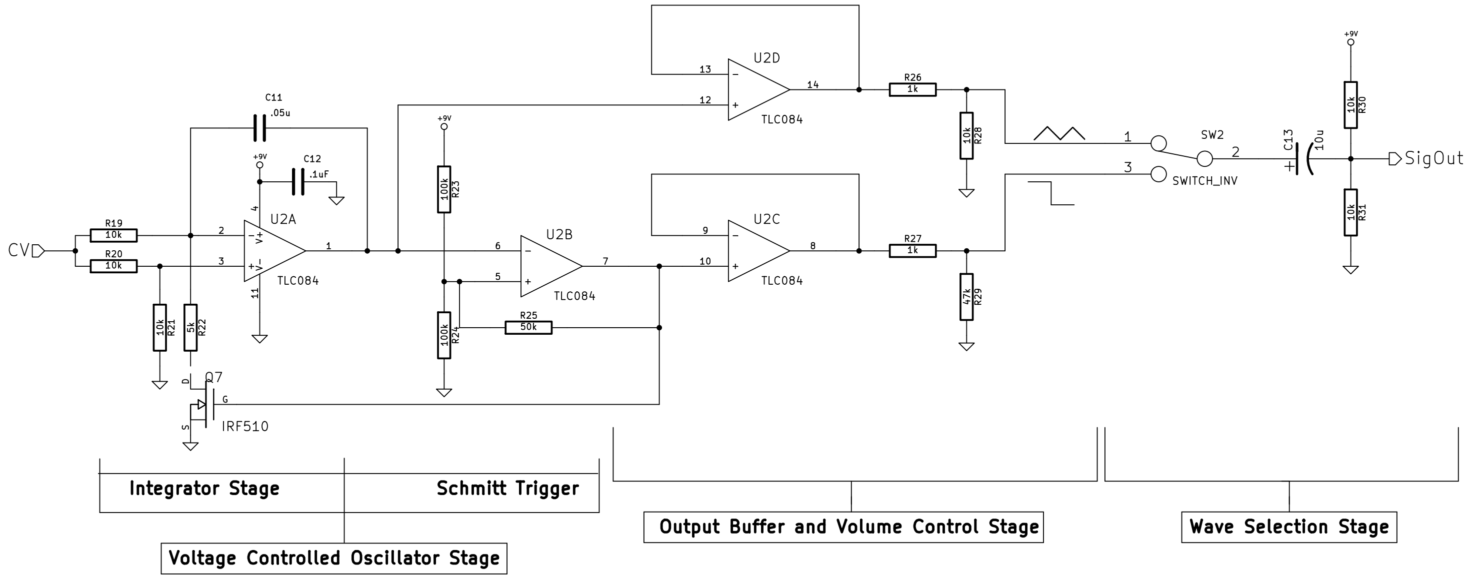
Voltage Controlled Oscillator



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Title: EC412 Analog Audio Synthesizer

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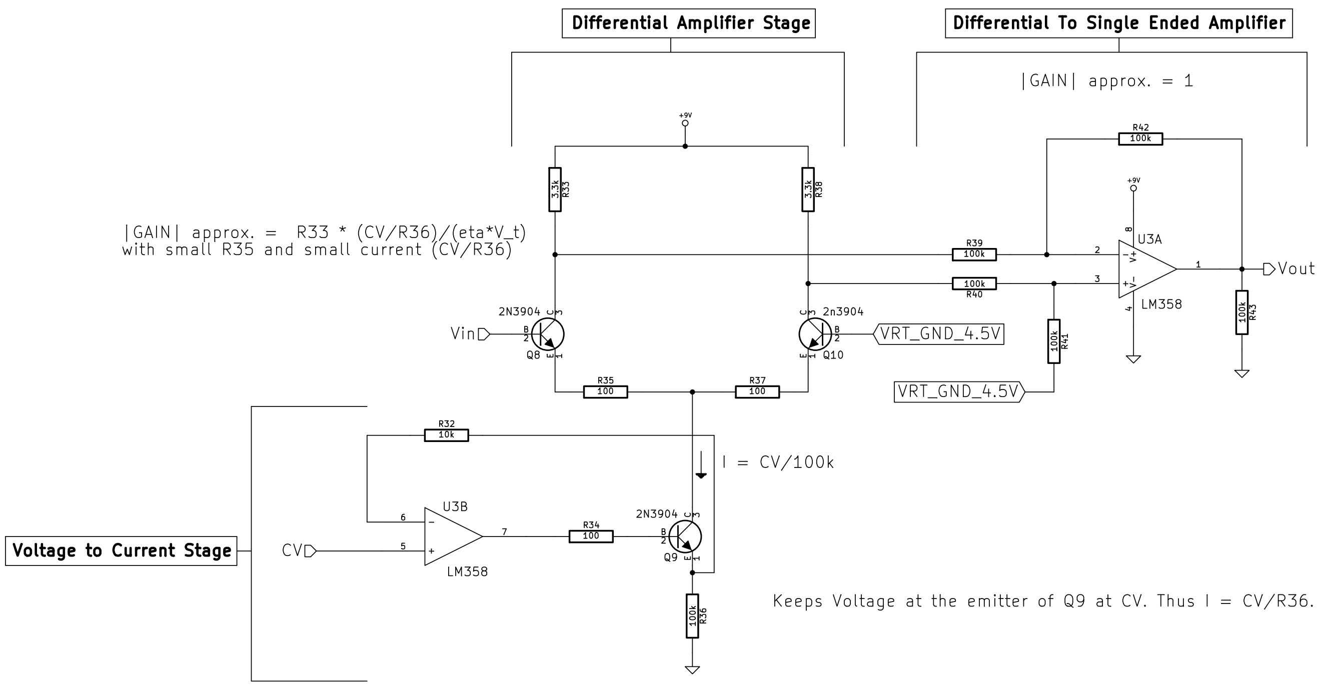
Voltage Controlled Amplifier (Variable Transconductance Amplifier)



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File:
Sheet: /VoltageControlledAmplifier/
Title: EC412 Analog Audio Synthesizer
Size: A4 Date: 2 may 2013 **Rev: A**
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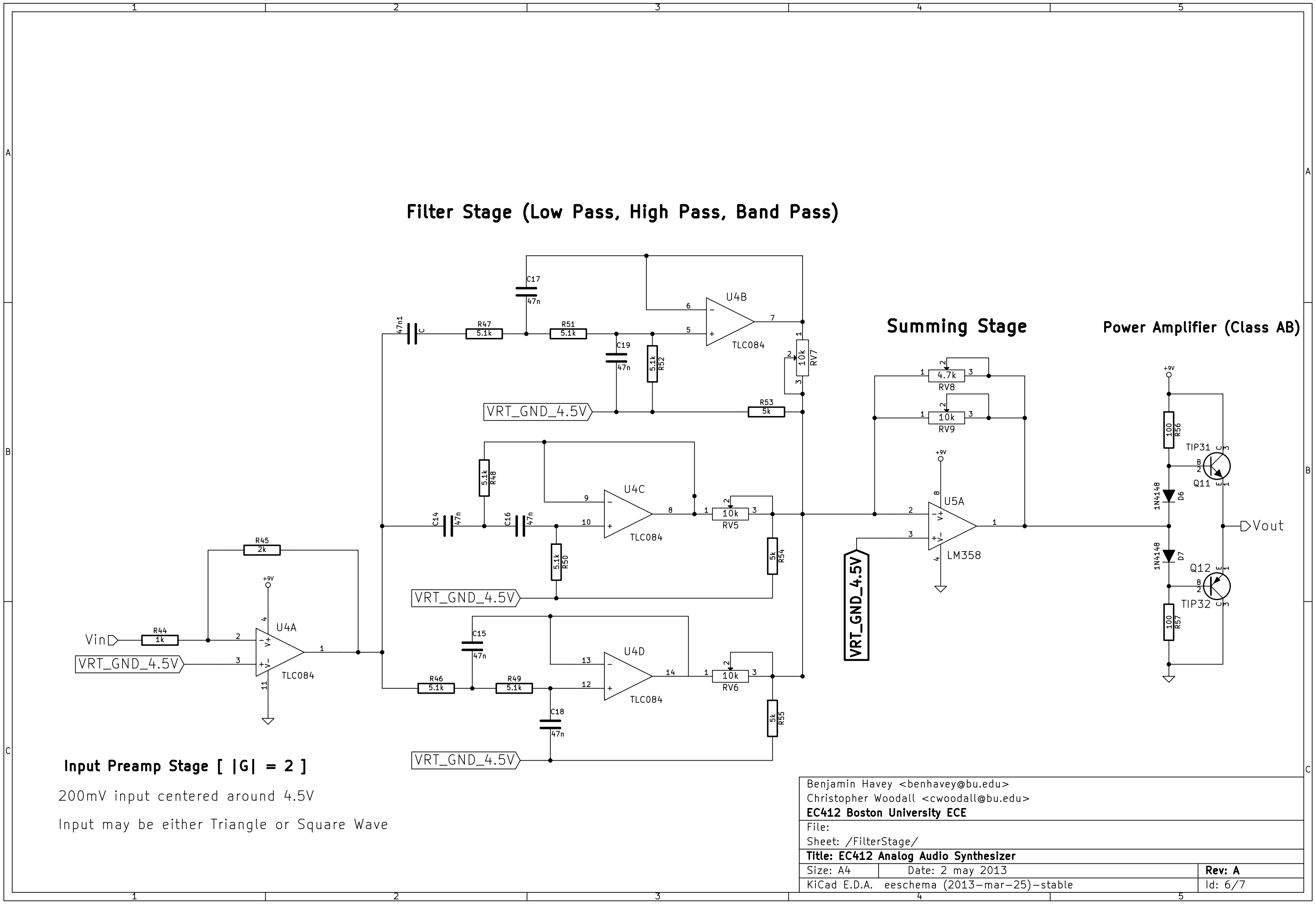
Filter Stage (Low Pass, High Pass, Band Pass)
Summing Stage
Power Amplifier (Class AB)
Input Preamp Stage [|G| = 2]
200mV input centered around 4.5V
Input may be either Triangle or Square Wave
Vin
Vout
VRT_GND_4.5V
U4A
U4B
U4C
U4D
TLC084
U5A
LM358
R44 1k
R45 2k
R46 5.1k
R47 5.1k
R48 5.1k
R49 5.1k
R50 5.1k
R51 5.1k
R52 5.1k
R53 5k
R54 5k
R55 5k
R56 100
R57 100
C14 47n
C15 47n
C16 47n
C17 47n
C18 47n
C19 47n
47n1
RV5 10k
RV6 10k
RV7 10k
RV8 4.7k
RV9 10k
D6 1N4148
D7 1N4148
Q11 TIP31
Q12 TIP32
+9V
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File:
Sheet: /FilterStage/
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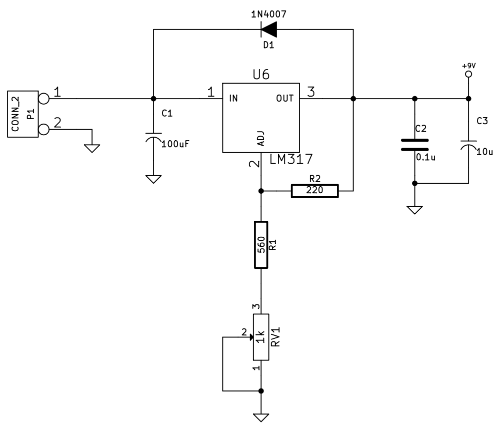
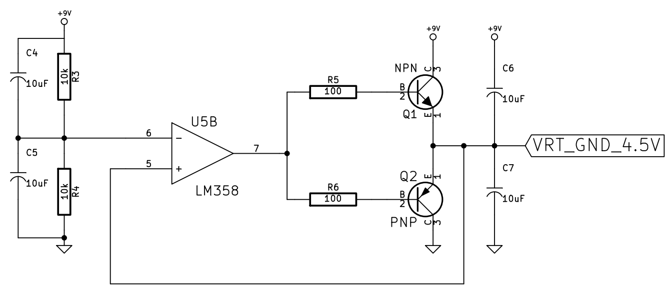
Date: 2 may 2013
Rev: A
KiCad E.D.A. eeschema (2013-mar-25)-stable
Id: 6/7

Voltage Regulator
[12V to 9V]



Virtual Ground Generator (4.5V with Adjustment)



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File:
Sheet: /Power/

Title: EC412 Analog Audio Synthesizer

Size: USLetter Date: 2 may 2013 **Rev: A**

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