Bass Guitar Effects Bank

6.101 Final Project Report - Spring 2016

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Abstract

Nowadays many musicians turn to digital signal processing, however software processing requires large digital overhead such as an operating system, incurs a large power cost, and reduces quality due to digitization. Therefore, in many circumstances analog audio processing is still preferred. In our project, we have created a customizable and modular analog effects bank for the bass guitar, which will focus on the freedom of the user to choose how to modulate their sound. Our system includes a variety of interconnectable and adjustable modules that the user can select and combine to achieve their desired overall effect. As a demonstration of our intrinsically modular design, we successfully integrated a bonus module into the system, further increasing user control.

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Overview (Mark and Israel)

In today's world of abundant computation, musicians have many options when they wish to produce effects or alter the sound of their instruments; however, digital signal processing requires large overhead and the audio signal must be sampled and converted into digital form which can degrade quality and add unwanted distortions. The features of analog signal processing, namely lower power consumption and cost, minimal perceptible delay, and preservation of fidelity, make it a clearly preferable alternative to the digital route for the highest quality.

The modules chosen for the effects bank showcase a wide breadth of analog engineering, with everything from sample and hold digitization circuits to amplification with minimal harmonic distortion. The breadth of audio processing achieved required knowledge of various oscillator implementations, passive and active filtering, rectification, transistor amplifier topologies, and a large variety of operational amplifier topologies. The effect designs drew inspiration from old design classics, but sometimes the implementation used modern parts and design paradigms. The result is a robust and effective product with a distinctly recognizable sound.

The effects bank consists of a pre-amplification stage containing a 3 channel equalizer, 6 audio effects modules (including the bonus fuzz face), and a commercial guitar amplifier depicted in the figure below. All of the effects stages are buffered and have a max output of 5Vp-p in order to ensure ample room for processing with ± 15V rails and high signal to noise ratio given the large amount of parasitics on the bread boards. With this design paradigm, the modules can be placed in any order and any combination the user desires without any consideration other than what sound will be produced.

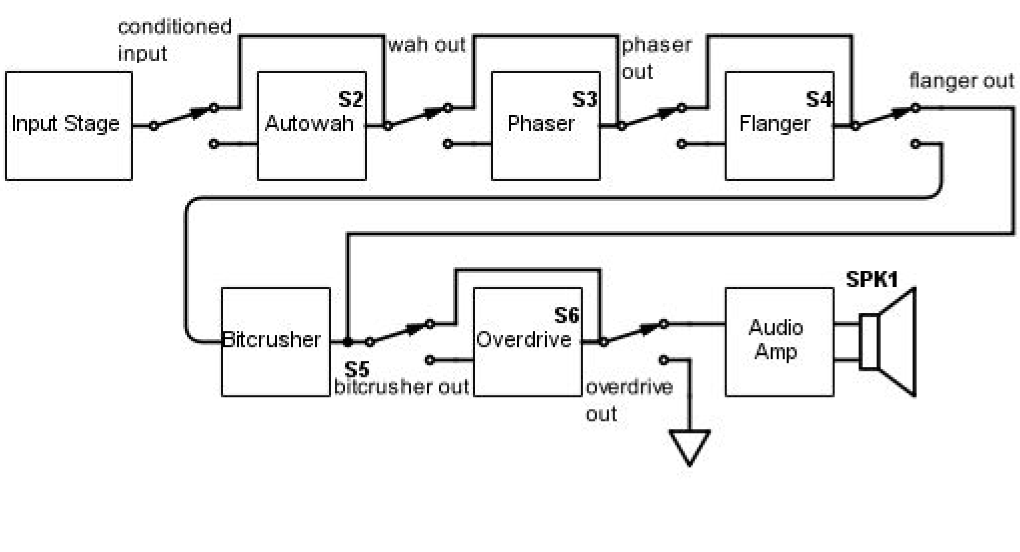
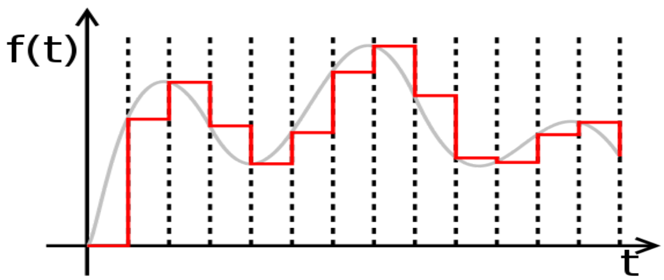


Figure 1. A block diagram showing a setup consisting of most of the modules.

Bitcrusher(Mark)

A relatively modern effect, the bitcrusher produces a distinctive sound that is iconic of early video game sound scapes. The bitcrusher is typically a low fidelity digital effect, but our module attempts to mimic its sound by using a circuit consisting of discrete components and operational amplifiers. The artificially low fidelity is achieved through sample rate reduction and resolution reduction, both of which are illustrated below.



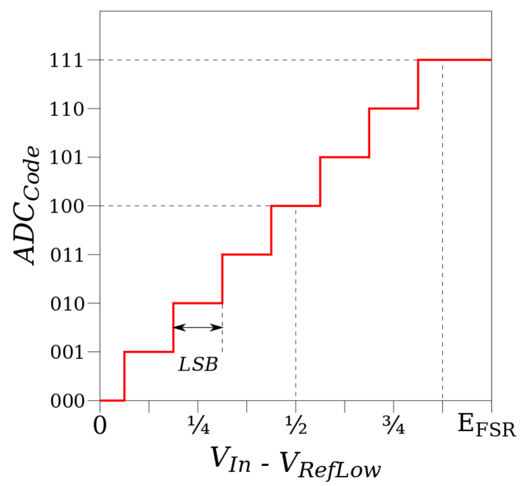


Figure 2. Demonstration of fidelity reduction
Left: an intuitive view on sample and hold, showing the continuous waveform turning into a discontinuous waveform with steps.
Right: a graph showing the input voltage corresponding to discrete output voltage levels on an ADC.

A continuous audio signal possessing an infinite sampling rate requires a sample and hold (SH) stage in order to reduce the sampling frequency. Then, the sampled waveform is passed through an analog to digital converter (ADC), which produces an output signal with a discrete set of amplitude values.

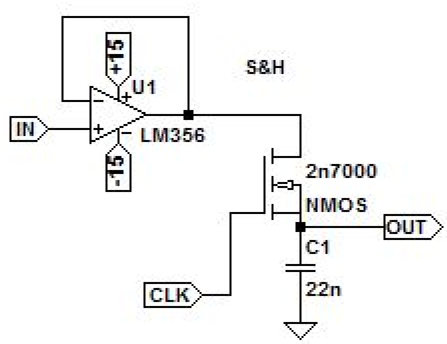


Figure 3. A standard sample and hold topology

Shown above is the SH stage. The input audio signal is buffered by U1. The 2N7000 NMOS is used as a switching MOSFET; when the CLK is high, the NMOS's source and drain are shorted, and the output of the buffer appears on the capacitor C1. Later, the CLK goes low, and the value in C1 remains the same until the next clock cycle. In this way the CLK frequency determines the sampling rate. Next, the timing circuit that produces the CLK signal is examined.

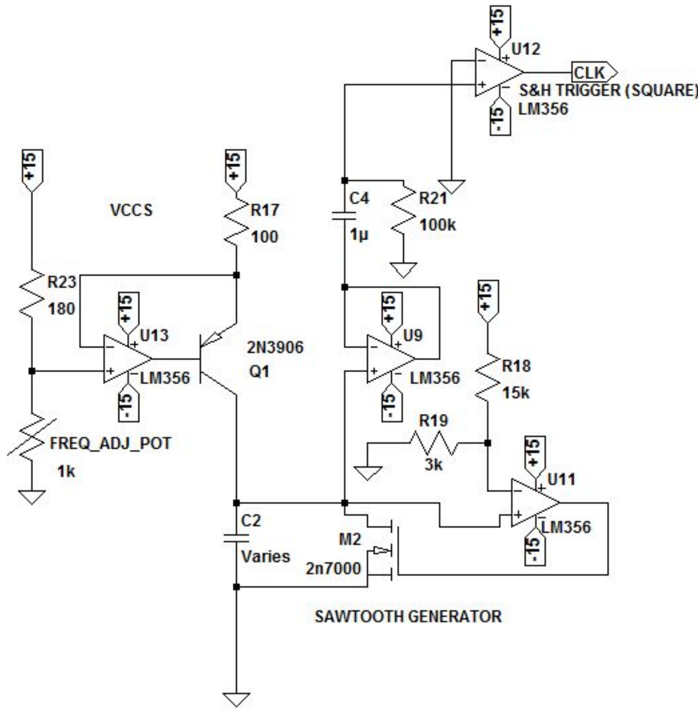
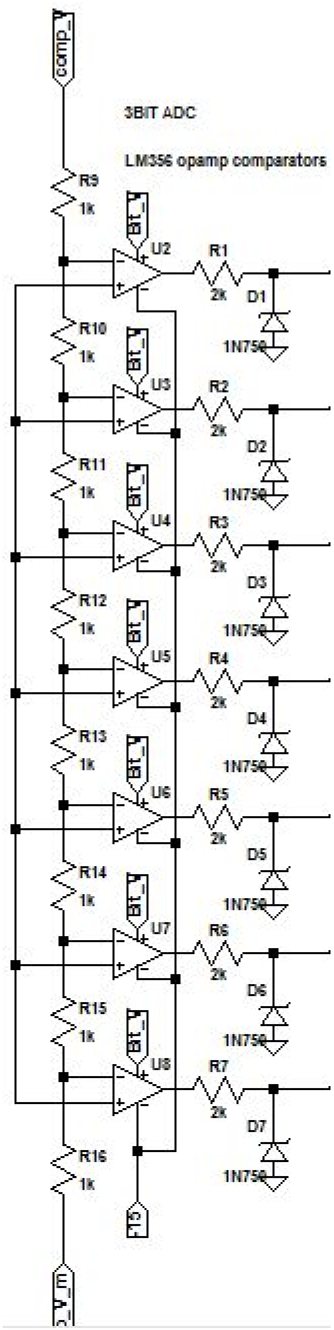


Figure 4. The timing circuitry for the SH stage, note the VCCS from lab 6

The left portion of the above schematic is a voltage controlled current source identical to that from lab 6. U13 has negative feedback, forcing the output of U13 to drive Q1 in a manner that ensures both of U13's input terminals have the same voltage. Due to U13's configuration, the emitter voltage of Q1 can be set by adjusting the 1k ohm potentiometer (FREQ_ADJ_POT). Consequently, the FREQ_ADJ_POT controls the current through R17 and thus the collector current as well as the rate at which C2 charges. If M2 is switched on and off at a fixed rate, C2 is repeatedly shorted and recharged which produces a sawtooth waveform across C2 with a frequency corresponding to the switching rate of M2. M2 is made to switch with a comparator composed of R18, R19, and U11. R18 and R19 also set the threshold voltage at which the U11 output will go from high to low, or vice versa. Note that this reference is not connected directly to ground, but instead draws a DC offset from a voltage divider. The reason for this is the signal being compared has a DC offset due to component restraints. C2 cannot drain completely during the short time M2 is conducting, thus the sawtooth waveform contains a DC offset. To remove the offset, the waveform is simply buffered by U9 and then high pass filtered. The resultant signal is then compared to ground by U12, which produces a low duty cycle squarewave - a suitable CLK signal. This clock circuit completes the SH stage, whose final output is sent to the ADC.

In order to simplify the circuitry and avoid using purpose built ICs, a direct conversion or “flash” ADC topology is used. The flash ADC topology, shown below in figure 5, utilizes a parallel bank of comparators each having a reference voltage from the same linear voltage ladder.



All of the non-inverting input terminals are connected to the input signal, in this case the sampled audio waveform. The resistor chain divides the maximum input amplitude of 5 volts evenly for each comparator; this sets a strict output voltage for each range of input voltages. The op amps are powered by +15V and - 15V bench voltage supply. As the input signal’s amplitude increases, the op amps’ outputs reach as close to +15V as they can, one by one. The opposite happens when the input signal decreases in amplitude; the topmost op amp, U2, will output as close to -15V as it can, and so on.

Since the rails of the op amps are not very uniform, reverse biased zener diodes are added to the output terminals in. The 1N750 zener diodes provide a relatively uniform 4.7V output when the op amp output is high, and about -0.6V output when the op amp’s output is low.

This discrete set of voltages are then summed together with an op amp adder, and then high pass and low pass filtered before the output buffer.

This module represents a truly analog alternative to digital signal processing, and it is able produces its own worthwhile effect. The main challenge for this module is the lack of online reference, since many bit crushers are digital in nature. It was difficult yet rewarding to complete a working and good sounding module starting with only a rough idea of how different stages should be built.

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Figure 5. The flash ADC ladder used in the bitcrusher, note the use of the zener diodes to achieve a uniform output voltage

Please see the appendix for the full schematic.

Flanger(Mark)

The flanger's iconic jet plane like "woosh" effect is instantly recognizable by hardcore rock fans. At the circuit level, the flanger functions effectively like a self sweeping comb filter. The filter is achieved by feeding back a varying time delayed version of the input signal and mixing it with the original signal. The flanger is a famous effect, and many schematics are available online; however, most designs are too complicated for the timeline and require extensive use of purpose built ICs such as the phase-locked-loop (PLL). Therefore, I chose the simplest flanger schematic available, and substituted the PLL in the clock circuit with a low frequency oscillator (LFO) controlling a voltage controlled oscillator (VCO).

It is important to explain how analog delay is achieved in this module. Without digital memory, the only practical choice is the bucket brigade IC (BBD).

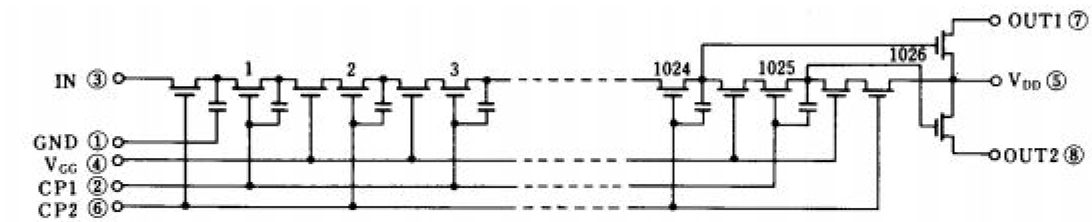


Figure 6. The MN3207 BBD device used to replace the MN3007 from the reference design

As shown above, the BBD can be thought of as a long chain of sample and hold stages - the sampled signal is passed down the delay line every clock cycle. The topology in figure 6 is the exact schematic of the MN3207 IC, which is the delay element in this module. This type of BBD is known as having "Complementary Output Topology", meaning the two source follower transistors, which connect to the last two consecutive capacitors, have two distinct outputs that can be summed with an op amp adder to produce a delayed version of the input with minimal distortions.

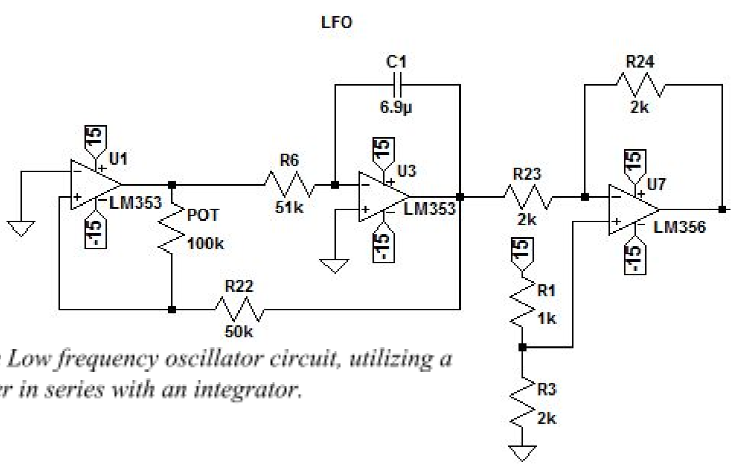


Figure 7. The Low frequency oscillator circuit, utilizing a schmitt trigger in series with an integrator.

CP1 and CP2 are the two clock inputs for the BBD, and they are opposite of one another. The clock frequency determines the length of the delay - a slower clock will result in a longer delay, and vice versa. Of course, since the flanger requires only 2-10 ms of delay, the clock frequency is kept between 100 kHz and 200 kHz as specified for this delay time.

The actual construction of the clock is a major challenge of this module. The low frequency oscillator (LFO) shown in figure 7 is the first component for the clock. U1 and U3 here form an astable oscillator. U1, R22, and the 100k potentiometer form a schmitt trigger with reference to ground, causing U1 to either output a positive or negative constant voltage close to the rails. The aforementioned output is integrated by R6, C1, and U3, transforming the square wave into a triangle wave needed to control the voltage controlled oscillator (VCO). The frequency of the output waveform can be adjusted by tuning the potentiometer, which dictates the switching voltage of the schmitt trigger. The triangle wave control waveform is given a DC offset by U7. The reason for adding a DC offset will be discussed along with the VCO.

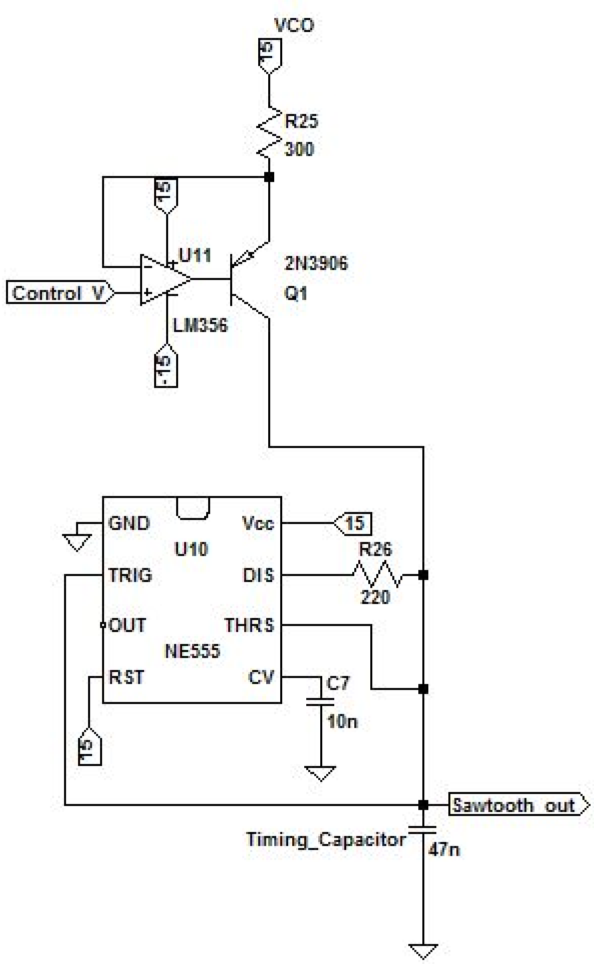


Figure 8. The VCO with a 555 timer

Shown on the left in figure 8 is the VCO. Note the top section is the same VCCS design used in the bit crusher, and it takes a control voltage from the output of the LFO. The LM356's output current is constrained such that if the control voltage is much lower than +15V, then the base current of Q1 required to produce the requested emitter current will exceed the output limit of the op amp, causing this VCCS topology to behave unexpectedly. The simple solution to avoid the VCCS failure scenario is to ensure the control voltage never decreases below the failure threshold by adding a DC offset as discussed above.

Due to the precise performance requirements of the BBD, the 555 was chosen over discrete components. The collector current of Q1 is slowly varying at the frequency of the LFO and this current charges the 555 sawtooth oscillator.

The discharge pin (DIS) will be connected to ground when the threshold pin (THRS) senses ⅔ Vcc, quickly draining the timing capacitor and then restarting the charging process. These actions of the 555 produces a sawtooth waveform at terminal of Timing_Capacitor. The sawtooth waveform is then fed into comparators U12 and U13 shown in figure 9 to produce a pair of opposite clock signals. Actual comparators are used here in order to safely meet the clock requirements of the BBD. The LM311 comparators are placed in parallel, with the sawtooth input connected to the inverting input on one comparator, and the non-inverting input on the other comparator. This ensures the fastest output of two opposite clock signals. The reference is made adjustable with POT1, so the duty cycle can be tuned as close to 50% as possible.

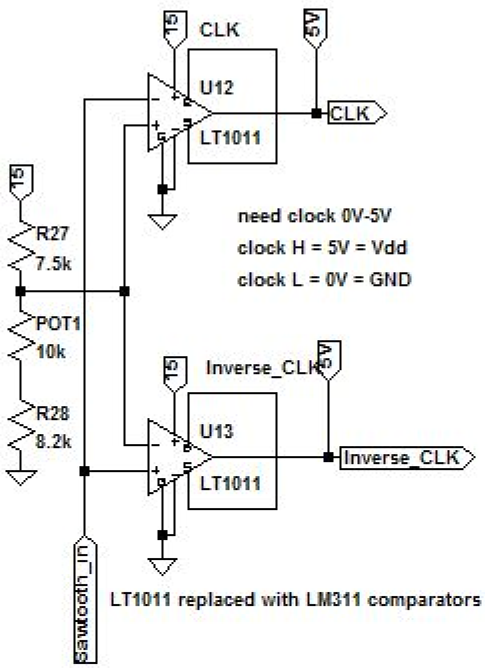


Figure 9. The parallel comparators

The timing circuitry is now complete, with outputs closely meeting the BBD specifications. The BBD is then biased according to the data sheet, figure 10 shows the arrangement for the BBD section.

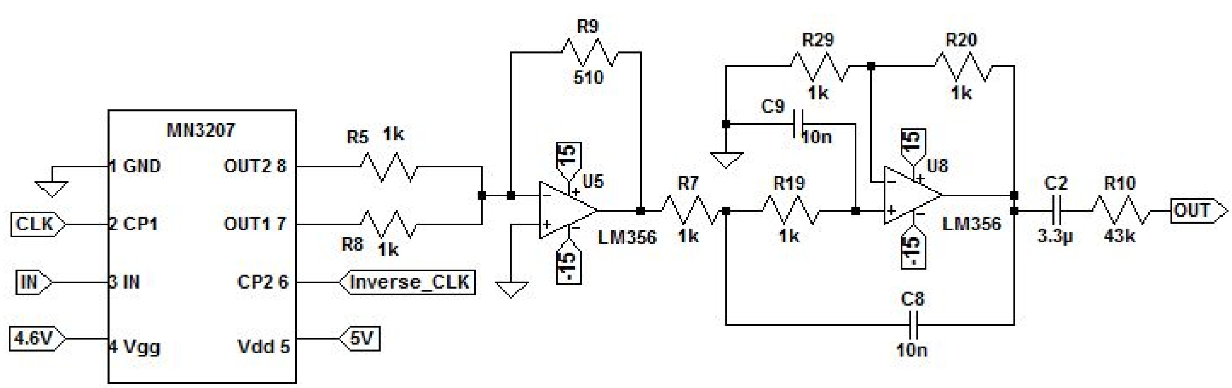


Fig 10. All connections to the BBD

The outputs of the BBD are summed by an op amp adder as mentioned before. Note that this adder has a gain of ½, as the feedback resistor R9 has half the resistance of that of the input resistors. R20 and R29 give the active low pass filter, U8, a gain of 2, since Gain = 1 + R20/R29. Setting the gain of the adder to ½ is a simple way to cancel out this gain. U8 is needed because the BBD output has a non-trivial amount of clock interference. The final HPF eliminates any DC offset before outputting to the next stage.

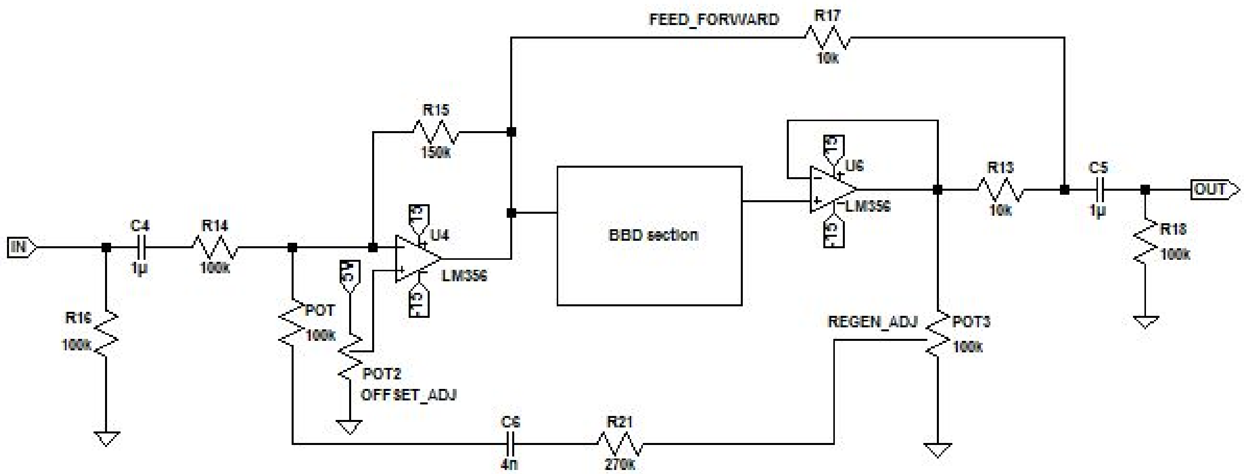


Figure 11. Input stage, output stage, and feedback paths added

With the BBD section complete, the input stage(left), the output stage(right), the feedforward path, and the regen feedback are connected as shown above in figure 11. The input stage is an inverting amplifier with an adjustable offset. This offset is needed for the BBD because this device requires the input to have an above zero rms voltage. Through experimentation, it is found that a DC offset of 2.5 V produces the cleanest output signal. The output stage is a simple buffer and a HPF to eliminate DC offset. The feedforward path adds the delayed and the original signal together. Lastly, the regen path splits a portion of the delayed output back into the input, which enhances the effect even more.

Please see the appendix for the full schematic.

Overdrive(Mark)

The overdrive effect was one of the first analog audio effects ever created, so old that it was originally built with vacuum tubes. Overdrive is a signature gain effect, meaning the distortion of the audio waveform is achieved through amplification. The working principle of the circuit is fairly simple: a sine wave passes through a high gain amplifier, causing the maximum amplitude to exceed the supply voltage and the tops of the sinusoid to flatten. This is known as hard clipping, a classic overdrive variant. In the extreme case, hard clipping turns a sine into a square wave, introducing new even and odd harmonics. Below is the full schematic of the overdrive module and the circuit level explanations.

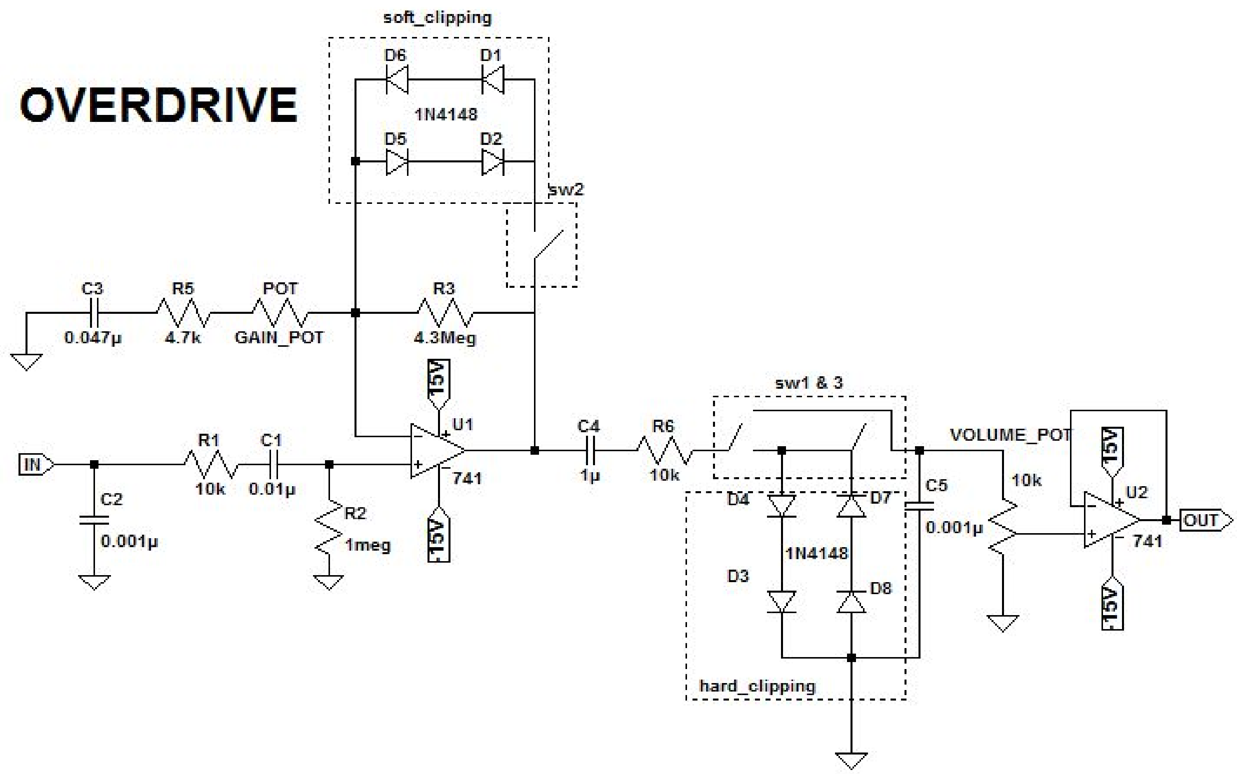


Figure 12. Inspired from the MXR Dist+ overdrive pedal for the electric guitar

The input signal is immediately low passed by C2, which also functions as a protection capacitor that eliminates any high voltage spikes that may harm the rest of the circuitry. The signal is then high pass filtered by C1 and R2 to eliminate any DC offset before amplification. This filtered signal is then fed into U1. If we open SW2 and disregard the soft clipping diodes, we can see that this is a noninverting amplifier. This amplifier has an adjustable gain controlled by a potentiometer (POT); furthermore, the gain also varies with the audio frequency as C3 introduces a frequency dependent impedance. An audio signal with higher frequency results in a lower impedance, this means higher notes will be amplified more, as the gain is inversely proportional to the sum of the impedances excluding the feedback resistor R3.

Now we have an amplified signal with amplitude large enough to be noticeably clipped by the two set of oppositely biased diodes. To achieve hard clipping, we turn on SW1 and SW3, while leaving SW2 open and excluding the soft clipping diodes. The amplified signal is DC blocked by C4, and then it sees the diodes. If a diode is connected between the signal path and the ground, the signal won't be shorted to ground until it exceeds the forward diode voltage. When the signal exceeds this forward voltage, the portion of the signal above that forward voltage will appear "chopped off." On the positive half of the waveform, D4 and D3 conducts and on the negative half, D7 and D8 conducts. Two diodes are connected in series in each direction because of preference, since the signal level is relatively large, a higher cut off threshold for the waveforms will preserve most of the waveform. This hard clipped signal is then divided by R6 and the volume control potentiometer (VOLUME_POT), which can be turned to adjust the output volume.

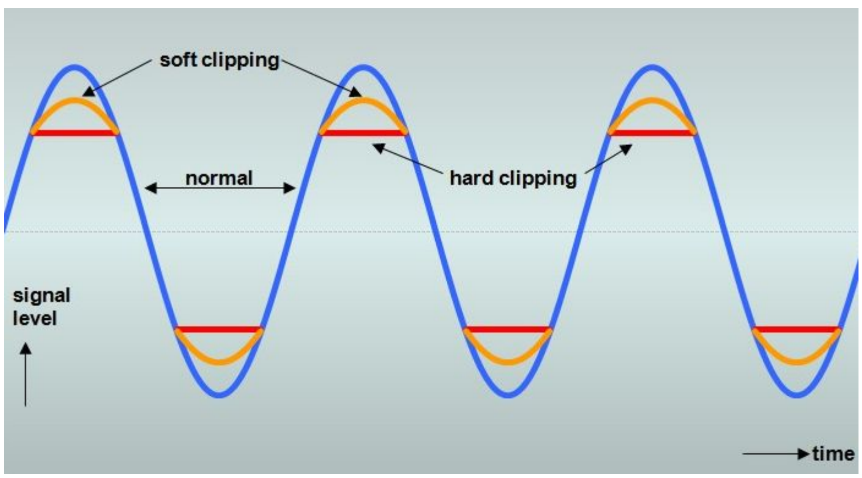


Figure 13. Hard clipping vs. soft clipping

Soft clipping is also achieved through manipulating the gain, but the output does not simply become clipped near the diode forward voltage. Instead, it appears the higher the amplitude, the lesser the gain. This is implemented by placing diodes in the previous manner, but in the feedback path. This negative feedback causes the gain to be inversely proportional to the signal amplitude, which is the big idea behind soft clipping. This placement achieves soft clipping again by utilizing the diode forward voltage. When the signal amplitude is low, the diodes are off, effectively disappearing from the circuit. So the amplifier functions as intended and a large gain is observed during this time. However, as the signal amplitude increases, the diodes will turn on, and effectively becoming shorts, shunting the feedback signal from R3, and turning the noninverting amplifier into a simple voltage buffer. This means the gain quickly drops to unity, and the flatter peaks are achieved.

Fuzz Face(Mark)

The fuzz face effect produces a highly distorted sound similar to that of the overdrive, and is usually referred to as "fuzz", hence the name. This is an old school analog audio effect and is very prevalent in rock n roll music due to its clean audio cut off; it is made popular by Jimmy Hendrix. The effect primarily depends on the selection of the transistors. The full Spice schematic is shown below, and explanations for each stage follow.

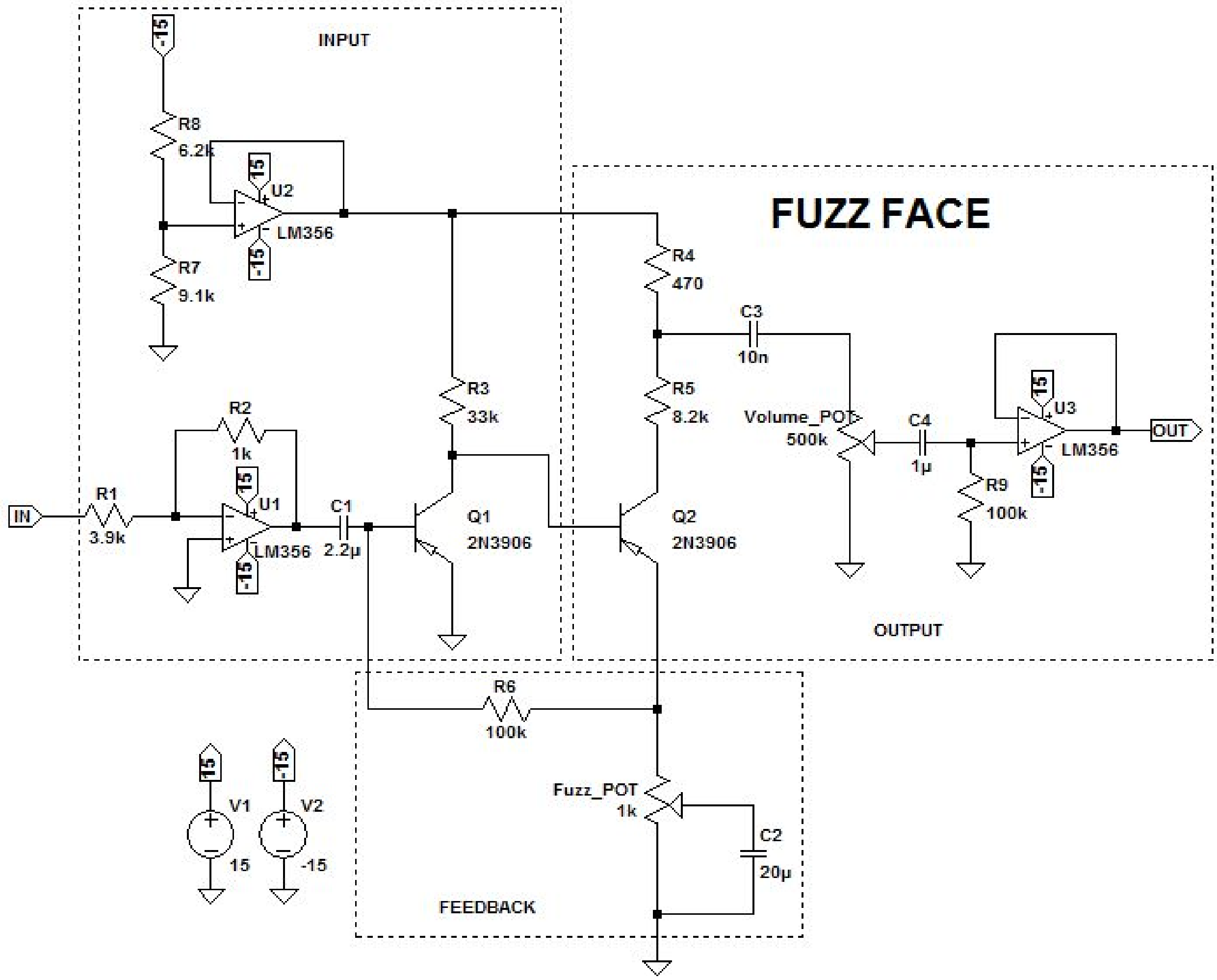


Figure 14. Fuzz Face full schematic. Thanks to Electrosmash.com for the inspiration

The circuit is designed using the common two stage BJT amplifier topology. Indeed, the distortion effect is inherently a "gain effect", which means the final sound is achieved through amplification of the input signal.

The input stage consists of an op amp amplifier with a fractional gain less than unity. This is added to the original design because our signal level of 5 V peak to peak (Vpp) is much higher than the standard audio processing signal level of 1 Vpp. The attenuated signal is then passed into the common emitter PNP amplifier consisted of Q1 and R3. Note that the circuit is designed to be powered by -9 V, which is provided by the voltage divider and the buffer, U2. The DC blocking capacitor C1 removes any DC offset from the attenuated signal. R3 sets the transistor Q1's bias point and collector current, which in turn determines the voltage gain. This gain is quickly calculated as below.

$$I_E = (V_{cc} - V_c) \div R1 = (9V - 1.6V) \div 33K = 0.22mA$$
$$g_m = I_E \div V_T = 0.22mA \div 25mV = 0.0088\,\mathrm{S}$$
$$Gain = -g_m \times R3 = -0.0088\,S \times 33K = 290$$

This amplified signal then goes to the output stage and the feedback loop. This is another common emitter amplifier, but it also has an adjustable emitter degeneration resistor (Fuzz_POT). This resistance lessens ß dependence of Q2, making the effect of temperature change less noticeable. It should be mentioned that having an effect that changes as the musicians plays (the transistor will heat up after power on) may actually be desirable, which is why this resistance is made adjustable. The addition of C2 also adds frequency dependency to the gain of the output transistor. The 20 microFarad value used in the lab demonstration is a relatively large capacitance, meaning all of the bass frequencies will be fully amplified without attenuation. R6 provides a negative feedback path from the Q2 emitter back to the Q1 base. This diverts currents away from the Q2 emitter and injects it into the Q1 base. Now the second function of the fuzz potentiometer becomes clear - setting this resistance high would cause more Q2 emitter current to be diverted, thus reducing the gain of the system.

The output waveform is extracted between R4 and R5, this waveform is DC blocked by C3 and goes through a potentiometer. The center tap is the output and the potentiometer controls the volume as it is a voltage divider. Of course, the output is buffered by U3 before exiting the module.

Pre-amplifier(Israel)

The output signal level of a bass guitar can vary widely from guitar to guitar as well as with playing dynamics. To Ensure proper processing in later stages, the signal must either be boosted to maintain good signal to noise ratio in later stages or compressed to prevent unwanted clipping. Additionally, pre-amplification stages should have a flat frequency response over the audible range to prevent unwanted harmonic distortion.

The input stage depicted below consists of a high pass filter with a cutoff frequency of 12.5 Hz in order to remove DC bias without affecting the audible range. The newly treated signal serves as the input to the first common source inverting JFET stage. R41 is a sense resistor for experimental measurements. The small signal gain of this stage is given by $A_v = \frac{-g_m R_L}{1+g_m R_S}$, where $g_m \approx 4$ mS and $A_v \approx -2.4$.

The output of the first stage is then fed into an RC network with optional Bass Cut and Ultra Low switches. When both the switches are open, the network functions as a high pass filter with a cutoff of approximately 110 Hz and attenuation set by POT1. When both switches are closed, R4 becomes shorted, making C2 a low pass filter with a cutoff of approximately 1 KHz. C4 and C1 are also put in parallel increasing their effective capacitance and thus pushing the cutoff of the high pass filter to approx. 300 mHz. Again, the attenuation of the signal is set by POT1. These effects are mostly separable; When bass cut is open, the cutoff of the highpass filter is around 100 Hz, effectively attenuating any note lower than the G string of a Bass guitar, but when it is closed it pushes the cutoff far below the audible range so that it removes DC offset but has no further operation on signals of interest. When Ultra Low is open it has no noticeable effect but when it is closed it attenuates high frequency overtones without affecting the principal pitch of any notes played.

It is worth noting that POT1 functions more as a volume control between stages to control the overall gain but cannot prevent the first stage from clipping the signal. The second amplifier stage has the same topology as the one above and the math is derived the exact same way; Its small signal gain is approximately -1.6.

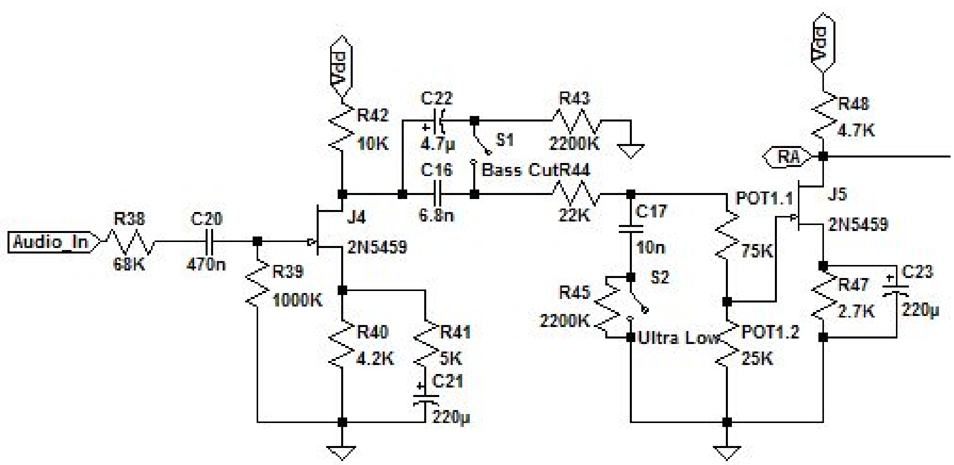


Figure 15: The first two stages of the pre-amplifier

The second amplifier output splits into two paths, one of which goes to a three band Equalizer (marked as RA in the schematic) and the other to a simple high pass filter to remove the DC offset of the amplifier bias. The Equalizer has a unity gain buffer stage U1 with a high pass filter to remove DC offset. This is followed by adjustable gain low pass filter with cutoff of 100 Hz, bandpass filter with center frequency of 1 KHz, and high pass filter with cutoff of 2 KHz. These filters are then summed by op-amp U2 and passed through a high pass filter to remove any DC Bias. The output RB then serves as an input to the an EQ Bypass switch shown in figure 17 that allows the user to choose between the unaltered and equalized audio signal.

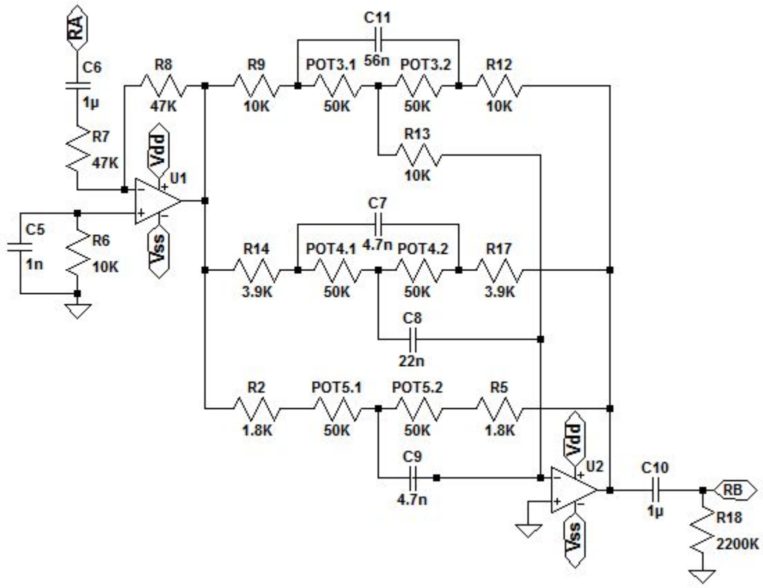


Figure 16: The 3 channel Equalizer stage of the pre-amplifier

After the EQ bypass switch the signal is passed through an RC network with a potentiometer volume control and an optional ultra high switch. With the switch open, the topology becomes a simple resistor divider volume control with a decoupling capacitor feeding into a common drain JFET amplifier. When the switch is closed, C18 gives high frequency signals an extra path to travel through, resulting in a magnitude and phase boost to overtones above 1 KHz. The common drain amplifier functions as a voltage buffer and the output is high pass filtered before being passed to the next module.

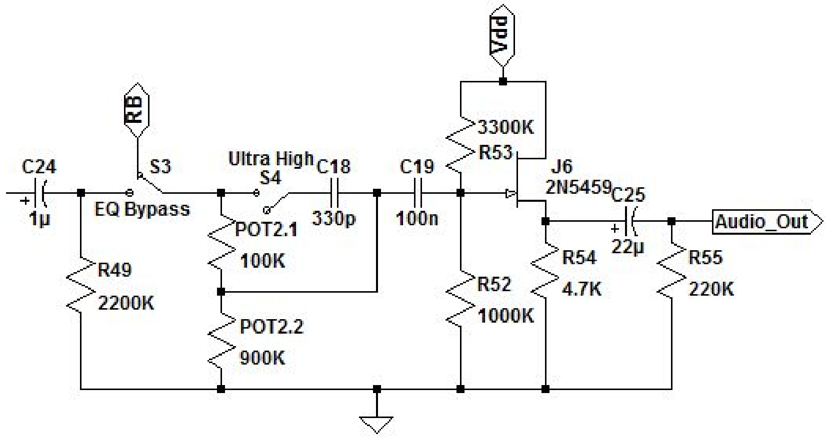


Figure 17: The final pre-amplification stage

Phaser (Israel)

The Phaser module produces a subtle shimmering effect that has a somewhat otherworldly feeling. The effect is achieved by splitting the signal into two feed-forward paths, one of which is treated with four all-pass filters, thereby altering the phase of the waveform, before being added back into the original. Adding the shifted signal to itself results in two notches, or disturbances in the original signal's frequency spectrum. Then, to get the characteristic sound, the phase shift of the all-pass filter stages is modulated by a low frequency oscillator to slowly move the notches up and down in frequency.

The Audio signal is first treated with a high pass filter and a unity gain buffer before being split into a clean feedforward and dirty signal path. The dirty path is passed through four all-pass filters with the topology shown in figure 17. First, without considering the effective resistance of the NFET J4, one can think of the resistor configuration below as an inverting amplifier with a virtual ground where the virtual ground is a high-pass filtered version of the input. The transfer function of this block is given by $\frac{V_{out}}{V_{in}} = \frac{sR3R4C2 - R5}{R3 + sR4R3C2}$. When R3 = R5 the pole and zero magnitudes of this equation cancel to give a gain of 1 but their phases add constructively giving a maximum phase shift of 180°. This equation also shows that the resistance R4 governs the location of the pole/zero when C2 is held constant; from here it is easy to show that J4 operating as a variable resistor can change the effective resistance of R4 and dynamically control how much phase shifting a given frequency receives.

The low frequency oscillator is an astable oscillator op-amp configuration that pulls up and pushes down the the voltage output by the potentiometer POT1. With POT2 the user can set how quickly the LFO oscillates (Thus how quickly the resistance of the NFETs change and consequently how quickly the notches shift up and down in frequency) and with POT1 one can set the initial resistance of the NFET variable resistors (thus what frequency the two filter notches oscillate around).

Finally the op-amp U7 sums the clean and dirty signals and high pass filters them to form the characteristic phaser sound.

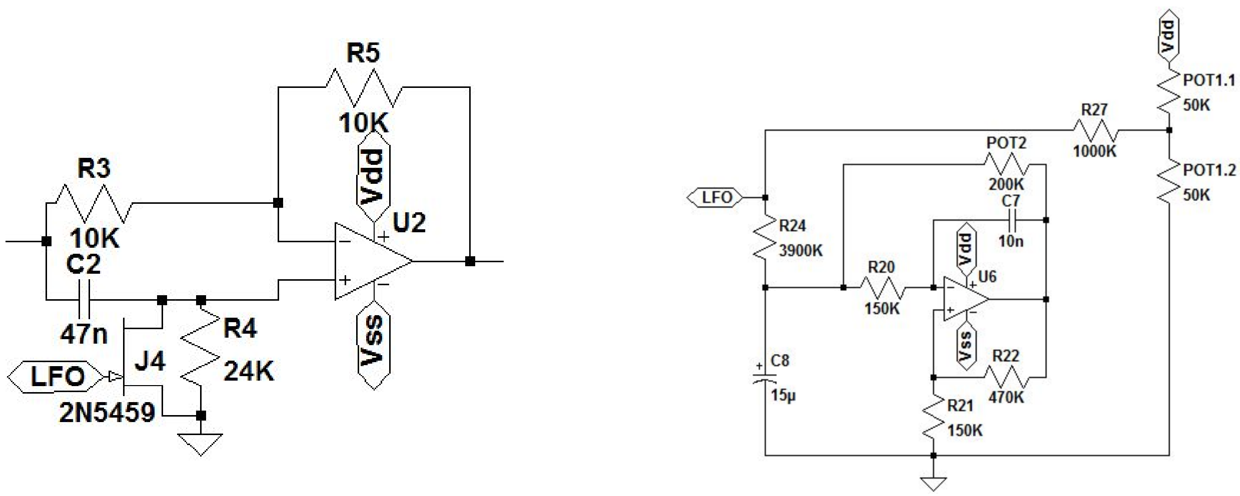


Figure 18: (Left) The basic all-pass filter block used in the phaser. (Right) The low frequency oscillator block.

Auto-Wah (Israel)

The auto wah effect produces a wavering sound that is controlled by the playing dynamic. The effect is made by shifting the cutoff frequency of a filter to accentuate certain frequencies being played based on voltage corresponding to playing dynamic.

An essential piece of a customizable auto-wah is the state variable filter because of its easily adjustable cutoff frequencies and resonance, as well as its ability to simultaneously output band-, high-, and low-pass filters.

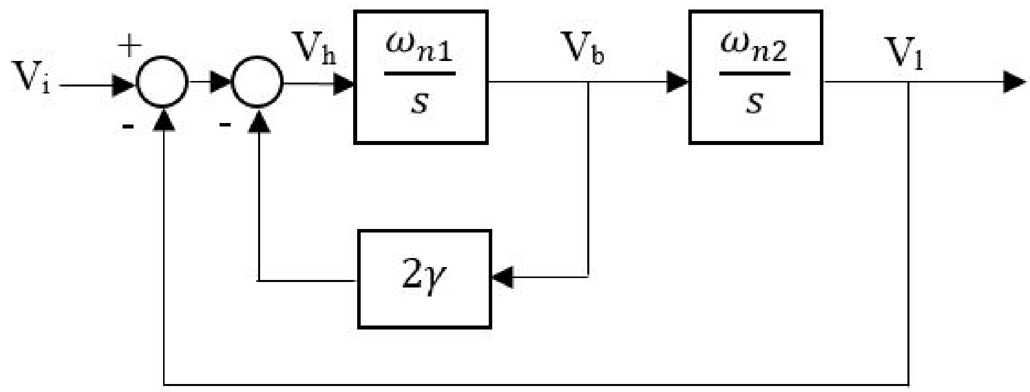


Figure 19: A Block diagram of a basic state variable filter topology.

The block diagram in Figure 19 depicts a conceptual overview of how the state variable filter operates in the laplace domain. It consists of two integrators with gains ω_{n1} and ω_{n2} respectively, as well as feedback of the bandpass output, V_b, with gain 2γ and of the low pass output, V_l, with unity gain into the input voltage, V_i, in order to produce the high pass output, V_h. Algebraic manipulation of the terms leads to the following transfer functions:

$$\frac{V_l}{V_i}(s) = \frac{\omega_{n1}\omega_{n2}}{s^2 + 2\gamma\omega_{n1}s + \omega_{n1}\omega_{n2}}$$
$$\frac{V_b}{V_i}(s) = \frac{\omega_{n1}s}{s^2 + 2\gamma\omega_{n1}s + \omega_{n1}\omega_{n2}}$$
$$\frac{V_h}{V_i}(s) = \frac{s^2}{s^2 + 2\gamma\omega_{n1}s + \omega_{n1}\omega_{n2}}$$

Where $s = j\omega$

These equations show that by altering the gains ω_{n1}, ω_{n2}, and γ, one has complete control over the position of the poles in these transfer functions. Further it can be seen that setting $\omega_{n1} = \omega_{n2}$ allows one to move the gains together but still retain full control over the poles. When $\gamma = 1$ the transfer functions all have a real double pole at $s = -\omega_{n1}$, when $|\gamma| \geq 1$ the transfer functions will have two distinct but real poles, however when $|\gamma| < 1$ the poles of the transfer functions are a complex conjugate pair, resulting in resonance. Thus, in order to dynamically control the cutoff frequencies of the filters and how much resonance they have, one must control the gains ω_{n1} and γ. The design detailed in this report implements these variable gains using Operational Transconductance Amplifiers.

U12 is an LM13700 operational transconductance amplifiers with a current output on pin 5 that, in this configuration, is equal to the inverting input on pin 4 amplified by a transconductance gain, set by the bias current on pin 1, i.e. $i_o = -V_- \times g_m$, where $g_m \sim i_{bias}$.

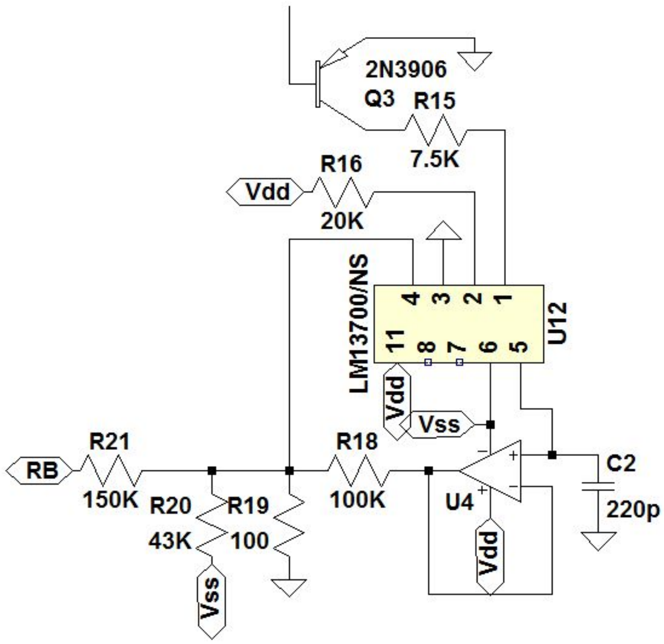


Figure 20: The Auto-Wah's variable gain integrator block

U4 is configured as a voltage follower with capacitor C2 and the output of U12 tied to its non-inverting input. Since the current through a capacitor is proportional to the derivative of the voltage across it and there is no input bias current into the op-amp under ideal conditions, the output of U4 is approximately proportional to the integral of the current into C2, which is the integral of the voltage on pin 4 of U12 scaled by a gain proportional to the bias current into pin 1 of U12. In this way U12, U4, and C2 (and with the same derivation, U11, U8, and C5) form a variable gain integrator block needed by the state variable filter outlined above.

The rest of the schematic is attached in the appendix. C6 and C7 are decoupling capacitors used to separate the output stages from the heart of the state filter. U10 is a non inverting amplifier with gain of roughly 5, chosen to boost the heavily attenuated signal back closer to line level. U5 is configured as an inverting summer that adds the lowpass and input signals but weights the low pass output more to account for attenuation by filter circuitry. The output to Bandpass is a lag or step down filter with DC gain of nearly 2 and a high frequency gain of unity, however the step down occurs in the MHz range and effectively functions as an amplifier on audible range signals. R18 and R33 are feedback resistors that change the math derived above slightly to give more control over the filter zero locations. R19 and R32 form two summing voltage dividers and R20 and R31 are pull down resistors. U9 is an inverting amp with gain 3 and outputs to RA, the input to the circuit that controls γ .

In figure 21, the circuitry to the left of U13 is bias control that determines the feedback gain γ , while U13 applies the gain and completes the loop. Expecting -5 to 5 volts for each control signal, U3 sums and inverts the resonance control voltages which, for demonstration purposes are the output of the envelope follower and the LFO signal. POT5 sets the initial resonance gain or γ value. U7 is a unity gain inverting amplifier with a virtual ground of about -12V in order to properly bias the LM13700, since the bias current pin is two diode drops above the negative rail. The non-inverting pin 3 of U13 sees 1/100 of the bandpass voltage. R56 turns the output current into a voltage, and the on chip transistors on pin 7 buffer the output voltage before it is added to

V_h by R21 in figure 20. An interesting performance note about this circuit: when the control signal is made very large, the imaginary part of the complex conjugate poles also becomes large, giving the filter a very high Q and essentially turning it into an oscillator that boosts a specific frequency and heavily attenuates all others.

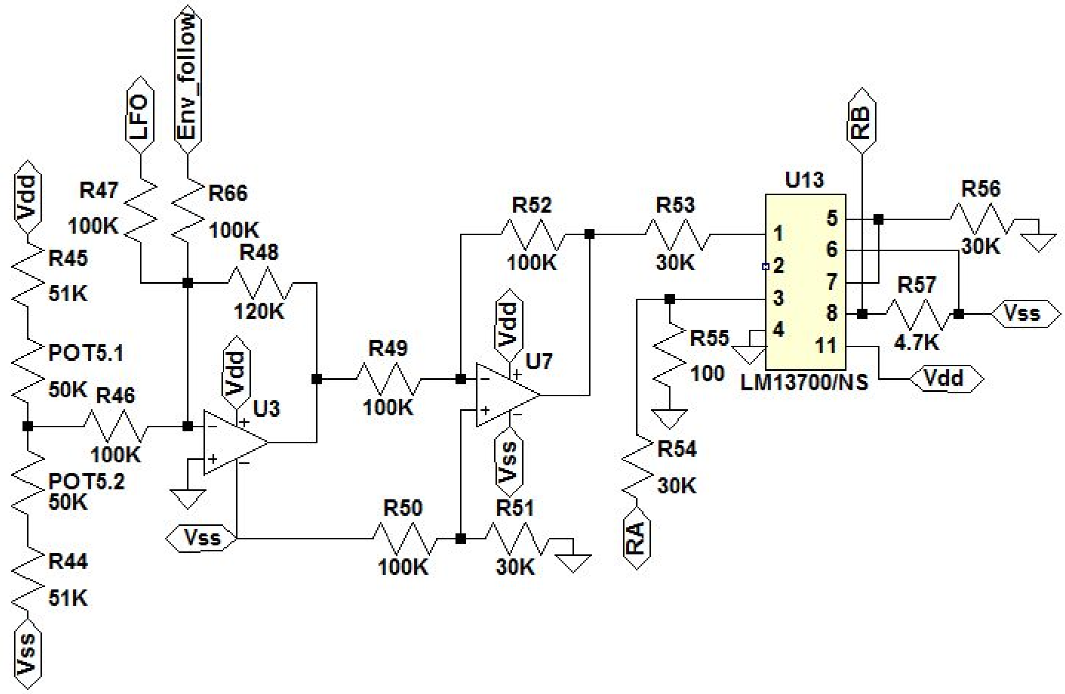


Figure 21: Resonance feedback control circuit

The audio signal is first high-pass filtered and buffered before being rectified by an op-amp precision rectifier, whose feedback loops compensate for the diode drops across D2 and D3, allowing for extremely low level signals to be properly rectified. The combination of a rectifier followed by a low pass filter like this is often referred to as an envelope follower because it produces a voltage output that describes the envelope of a waveform. In this case the envelope follower will track the amplitude of the bass guitar signal, outputting a peak during the attack of the note and decaying slowly with the sound of the note. A pot can be added in series with R72 to control the gain factor and thus sensitivity of the envelope follower to input signals. The output of the envelope follower is then optionally inverted and summed with other optional control signals such as an LFO. The new control signal is then passed through a resistor divider that determines the sensitivity of filter cutoff frequency to control voltage. U2 is in feedback and will thus force $V_- = V_+$ thereby keeping Q1 in a emitter follower configuration so the voltage on the emitter will be proportional to the control voltage. Q2 and Q3/Q4 form a cascode current source that biases U12/U11 with a current waveform that is proportional to the control voltage. POT4 controls the high frequency roll off of the biasing circuitry.

This multi-staged circuit creates an Envelope Controlled Filter, which allows an input signal to simultaneously bias the gain of two integrator blocks whose configuration allows that same signal to be dynamically filtered based on its own volume. Furthermore, the signal can control the resonance of said filter, accentuating certain frequencies during the attack, while sweeping filter frequencies and becoming a richer, fuller sound as the note decays.

Testing and Lessons Learned (Mark and Israel)

Since the project is inherently modular, each module is tested with the function generator before integration with the rest of the system. Although problems arose during individual testing, the full integration with the bass guitar and the commercial amplifier experienced few set backs. Integration worked so well because all modules contain input and output buffers, so with the correct signal level, one module will not interfere with the other.

Even though integration provided few setbacks, much time was spent on tuning the modules to achieve more recognizable and distinctive sounds. During this phase many adjustments to the original designs were made. One of the main challenges was tuning the flanger; the initial unreliability of the module was mainly related to the MN3207 BBD clone and the replacement of the PLL with discrete components and comparators. The original chip is old and not well documented. The clone requires a higher DC offset for the input to function properly. Furthermore, the quality requirement for the clock signal is difficult to achieve with lab parts. However, with proper filtering and comparator usage described in the individual module's section, the limits of the available devices are really pushed and a recognizable flanger effect is achieved.

Future students who want to utilize analog delay in their circuit are recommended to purchase the MN3007 clones instead of the MN3207. The flanger schematic originally uses MN3007 and no input DC offset is specified. The MN3007 also appears in most of the existing designs involving analog delay. These reasons make the MN3007 a simply superior version of the BBD.

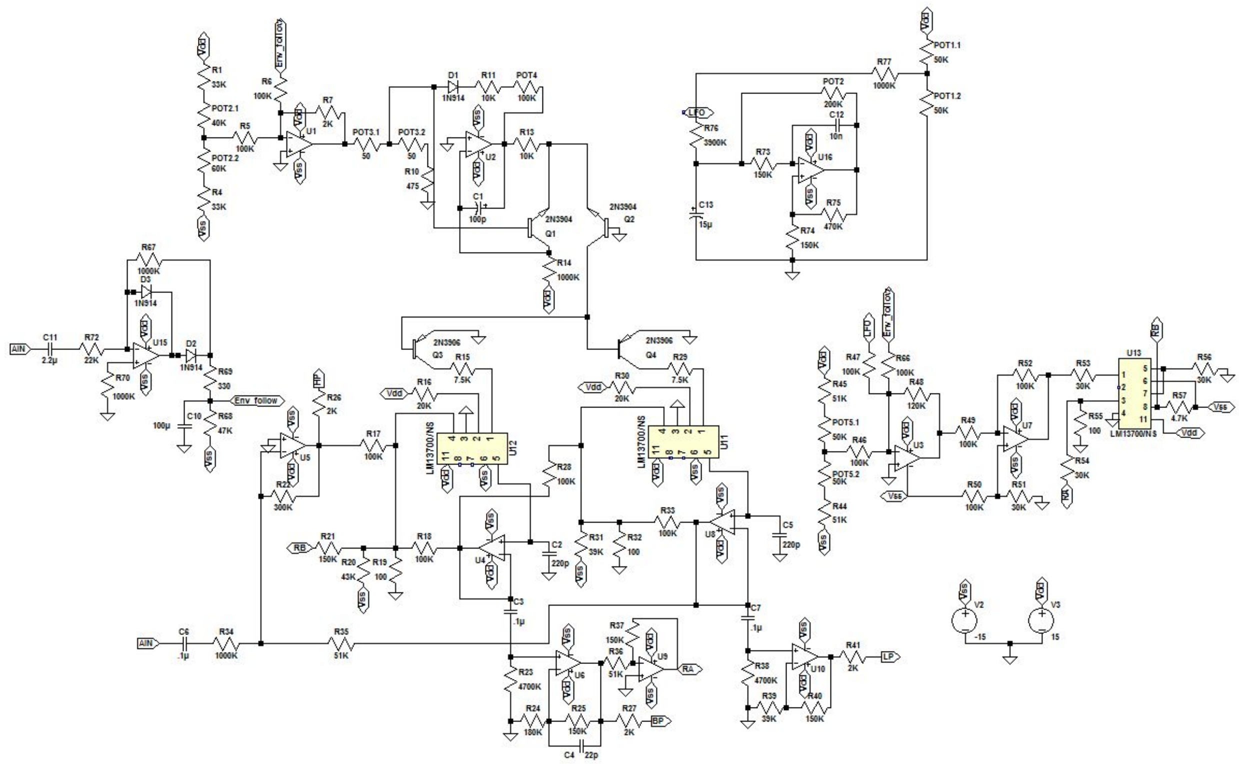
The Auto-Wah was also a particularly difficult module to get sounding just right. The complexity of the schematic made it a tough circuit to debug on a breadboard. Once it was working, the sheer number of tuneable parameters led to hours sunk just turning knobs and seeing what sound would come out. In addition the parasitics of the breadboard and some self-resonance led to an audible hum when filter cutoffs were set too high.

Conclusion

The “Bass Guitar Effects Bank” project was a great experience for us to solidify 6.101’s class material. We worked extensively with filter design, oscillator design, and learned to push the limits of available parts in order to achieve success. During device integration, we learned that physical circuits can really behave much different than spice simulations, one such example is the slower than expected switching speed of the 2N7000 MOSFET. In the end, we are able to meet all expected goals and stretch goals and, most important of all, we both become better engineers.

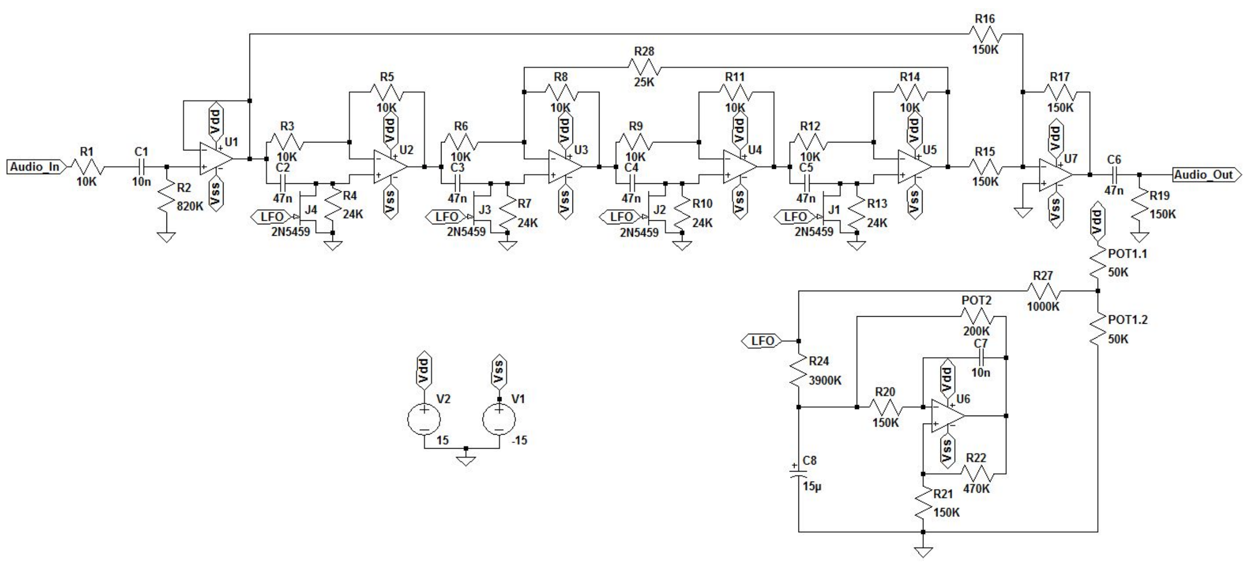
Appendix

Auto-Wah w/Precision Rectifier, LFO, State Variable Filter, and Resonance Control



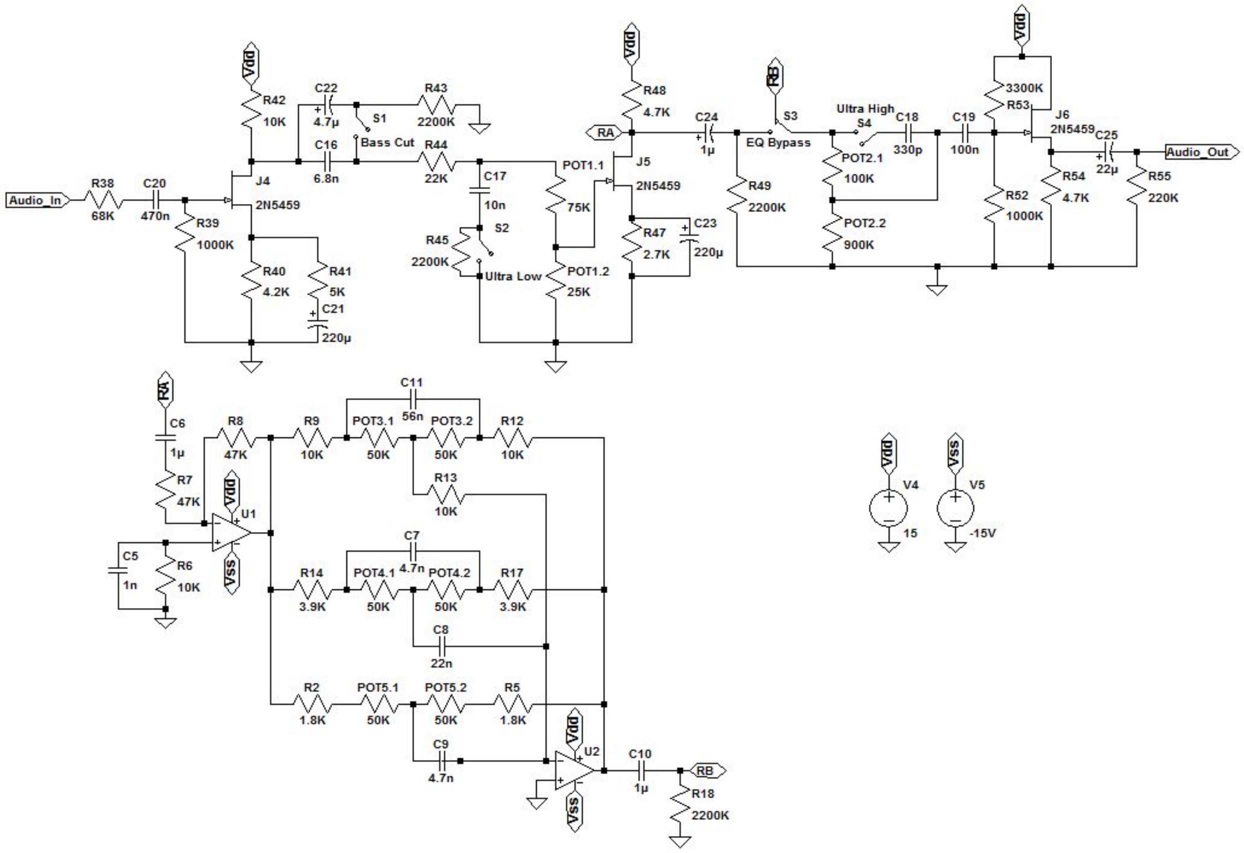
Variable State filter + resonance control schematic credit: Ray Wilson on musicfromouterspace.com

4 stage Phaser w/LFO and feedback



schematic credit: mxr phase90 on electrosmash.com

2 Stage JFET Pre-Amplifier w/3 Channel Equalizer and tone control switches



schematic credit: Albert Kreuzer on albertkreuzer.com

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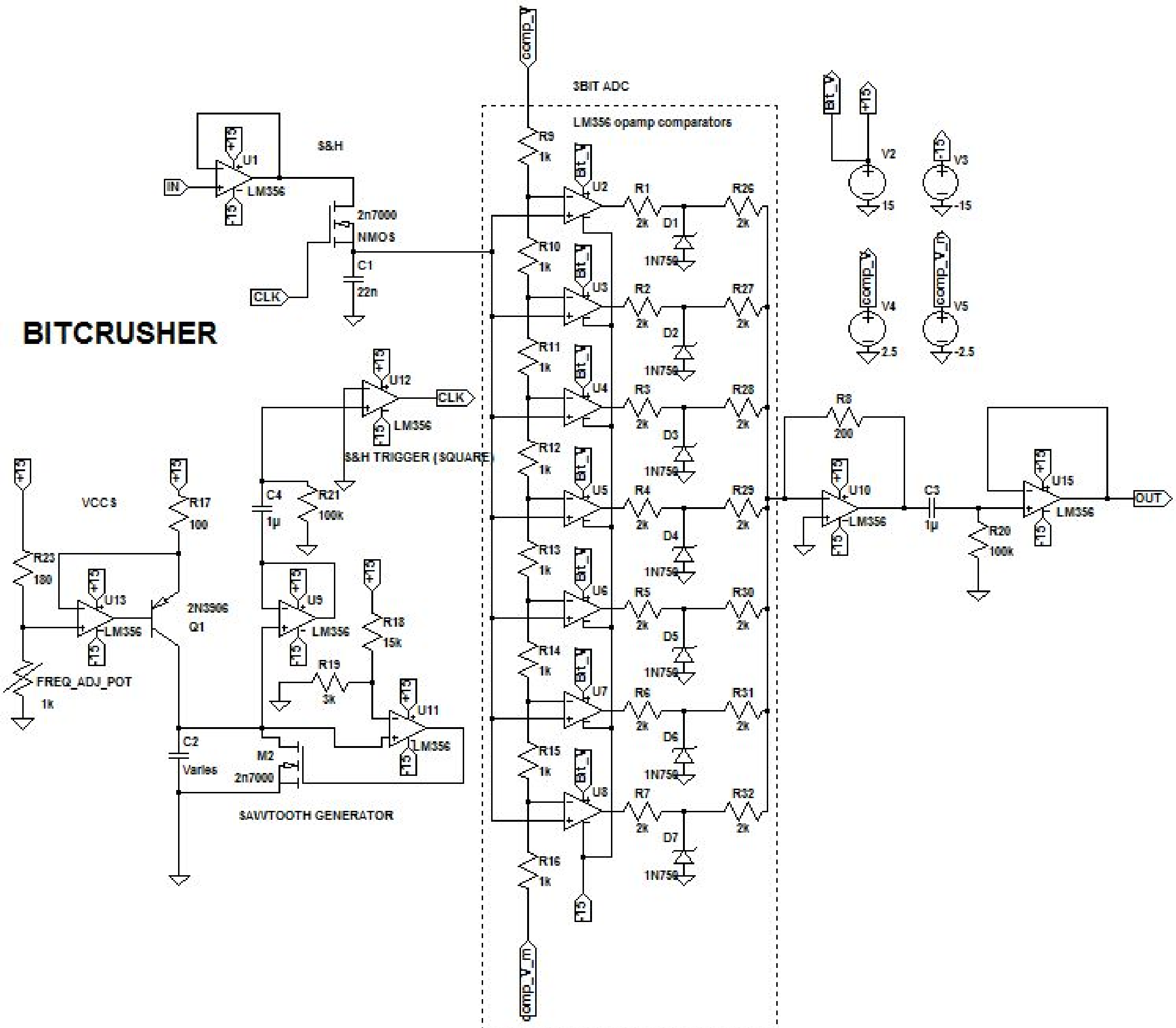
3BIT ADC
LM356 opamp comparators
comp_V
comp_V_m
S&H
IN
U1
LM356
2n7000
NMOS
C1
22n
CLK
U12
S&H TRIGGER (SQUARE)
VCCS
R17
100
R23
180
U13
2N3906
Q1
FREQ_ADJ_POT
1k
C2
Varies
M2
SAWTOOTH GENERATOR
C4
1µ
R21
100k
U9
R18
15k
R19
3k
U11
R9
R10
R11
R12
R13
R14
R15
R16
U2
U3
U4
U5
U6
U7
U8
R1
R2
R3
R4
R5
R6
R7
2k
R26
R27
R28
R29
R30
R31
R32
D1
D2
D3
D4
D5
D6
D7
1N750
Bt_V
+15
-15
V2
15
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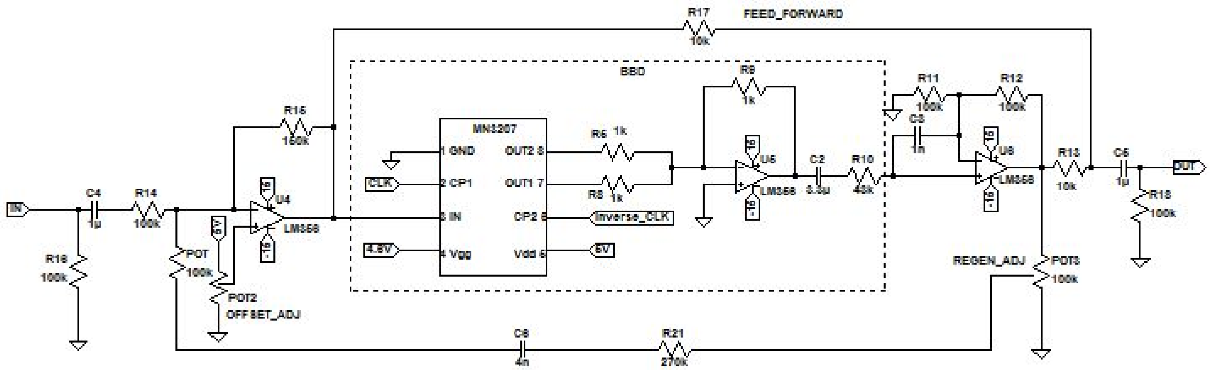
V5
-2.5
R8
200
U10
C3
R20
U15
OUT



FLANGER w/ REGEN and REVERB