Abstract

Our analog laser harp is a unique combination of a synthesizer, laser harp and Theremin. Like a laser harp, it consists of several lasers which, when broken, create a different sound. Like a synthesizer, the pitch of these different voices. And like a Theremin, the pitch is controlled without physically touching the instrument. Jordan was responsible for designing the distance sensing and laser detection scheme, Matthew designed the different synth voices, and Chad designed a power-efficient Class G amplifier to connect the harp to an external speaker.

Introduction

A laser harp, as it currently exists, is an electronic instrument that consists of several lasers (usually 5-10), projected from the base of the instrument up to the ceiling. When the user interrupts a particular beam with his hand, a unique sound is produced. Our analog laser harp will also be an electronic instrument that is controlled by breaking laser beams, but has the added bonus of having continuous pitch control. Like a Theremin, the pitch of our harp changes based on the vertical position of the hand and, unlike a Theremin, four different synthesizer voices can be created by using
the 4 separate lasers/sensors. The resulting project is a unique electronic instrument interface combined with fun, distinct synth voices and a power efficient Class G amplifier capable of driving a large, high end speaker.

1. Laser Detection and Distance Sensing
   by Jordan Addison

1.1 Introduction
The front end of the laser harp is responsible for allowing the input (the user’s hand position) to accurately and consistently control the synthesizer. The two inputs into the synthesizer module are: which beam (string) has been broken by the hand, and the hand’s vertical position along the string. These two inputs will be handled separately, and will be converted into voltages to be sent to the synthesizer module. In order to identify which laser has been broken, I chose to implement a ‘beam-break’ circuit using photodiodes and transimpedance amplifiers. The output from the laser module is used to activate the correct distance sensor, and is also sent to the synthesizer module to turn on the corresponding effect. I designed an ultrasonic distance sensor in order to detect the vertical position of the hand. The output of the distance sensor is used as the input to the voltage controlled oscillators (VCO) that control the pitch of each synth. A simple sketch of the physical layout of the laser harp is shown in fig. 1. The physical structure for the laser harp was built with the help of the EDS lab staff using 4" X 1/4" acrylic beams.

1.2 Laser Detection
The beam break detection design was fairly simple and straightforward. As pictured in figure 1, the lasers were positioned at the base of the laser harp, pointing upwards at their respective photodiodes. With the use of a transimpedance amplifier, an output voltage indicative of whether the laser beam is blocked can be easily obtained. The lasers used in the set up were 650nm, 3mW lasers. I chose red lasers because they were the cheapest and most readily available. The laser modules I purchased were already fitted with a constant current driver circuit, and needed only to be connected to a 3-5V supply. 3mm photodiodes (Digikey part #1080-1140-ND with a spectral range of 400nm-1100nm were used in the transimpedance amplifiers).
Figure 1: A diagram of the structure of the laser harp. The blue pill boxes at the base of the structure represent the transmitter/receiver pair for each string.

Figure 2: A schematic of the beam detection circuit. Four of these are included in the final design, one for each synth voice. The output $v_{\text{laser}}$ is used to turn on the string’s corresponding distance sensor and synth voice.
Though the peak detection wavelength of the photodiodes is listed at 940nm, I found that they worked quite well with the 650nm lasers. I deduced from the photodiode spec sheet (using the expected light current at the peak wavelength along with the spectral sensitivity curve) that they could nominally output $2.4 \mu A$ of current with incident light from the lasers. With this in mind, I chose to use a 5.1MΩ resistor in the transimpedance amplifier, expecting to see a 12.4V output with a laser beam incident on the photodiode. After some preliminary characterization I found this resistance value to be quite sufficient, observing 0V on the output in ambient light and 7-15V on the output with the laser turned on at a distance of 2.5ft. Although my original design idea called for the use of an ambient light comparator, this seemed unnecessary since the photodiodes did not respond at all to the ambient light in the lab and was therefore omitted from the final design. Instead, I compared the transimpedance output to a threshold voltage of 5V using an LM311 comparator IC and used the rail-to-rail output as the laser indicator voltage. The 5V input was taken directly from the power supply, and was also used to power the laser modules, but could also easily be obtained from the +15 supply using a voltage divider.

1.3 Distance Sensor

1.3.0 Overview

The general principle behind the operation of the distance sensor is not far-fetched. The transmitter sends out a short pulse, while the receiver (positioned side-by-side with the transmitter) waits for the signal to echo off of an object. When the first echo is received, a pulse is generated to trigger a sample and hold (S&H) circuit. The sample and hold samples a ramp voltage, so that its output voltage linearly corresponds to the time at which it was triggered - i.e. how far away the object was that created the echo. The farther away the object, the larger the output voltage.

1.3.1 Transmitter

The only thing necessary for correct transmitter operation is a short pulse at 24 kHz. I was able to achieve this with the use of an MC14073B 3-input AND gate. The three inputs into the gate are:

1. A constant 24 kHz square wave with 50% duty cycle to ensure the transmission is at the correct frequency
2. A periodic 'one-shot' pulse to regulate the width and timing of each transmission

3. $v_{laser}$, from the beam-break circuit, to ensure that the transmitter is on only when the corresponding laser is interrupted

The steady 24 kHz square wave was generated using a 555 timer configured as an astable oscillator. At first, I considered using a monostable 555 timers to create the one shot pulse. But, because of the intricate timing needed between the transmitter and receiver, this implementation would have called for very finely tuned timers with exactly matched frequencies that seemed very difficult to perfect. Instead, I utilized the same sawtooth waveform sampled by the S&H. Two comparators that compare the sawtooth waveform to two different reference voltages can be used to create a regular pulse with consistent width. The first comparator outputs 15V when the sawtooth wave ($v_{ref}$, fig.3) reaches 7.5V, and the second goes high when $v_{ref}$ reaches 7.56V. The two outputs are then subtracted to generate a finite width pulse with the same period as $v_{ref}$, 39ms. I chose the first threshold to be 7.5V out of convenience, I was already using a 7.5V rail for a few other parts of the circuit. With that as a reference, I calculated $\frac{dv}{dt}$ of $v_{ref}$ to be about $0.128V ms$ and chose a pulse width of 450$\mu$s (enough time for
Figure 4: A schematic of the transmitter circuit. Section (a) is the 24 kHz oscillator, and section (b) is the comparator/one-shot pulse generator. These two signals, along with \( v_{\text{laser}} \), are inputs into a 3-input AND gate whose output is a 24 kHz pulse with a 450\( \mu \)s duration that occurs every 39ms. In practice, 4 AND gates and transmitters must be used (one for each sensor) with the first two inputs being the same timing signals pictured and the third being the corresponding \( v_{\text{laser}[1-4]} \).

about 12 cycles at 24 kHz) which gave me the second threshold of 7.56V.

1.3.3 Receiver

When the echo reaches the receiver, a small signal oscillatory signal (a few tens of millivolts in amplitude) appears across it. The first task of the receiver circuit is to amplify the signal and demodulate it so that it can be more readily processed. To achieve this, I used an inverting amplifier with a gain of 20db, and a simple peak detector. In order to detect when an echo is received, I compare the resulting signal to a threshold voltage of 7.29V. The received echo can vary in amplitude depending on the object that is causing it (some surfaces reflect sound more readily), the angle at which its facing (something very hard to control with a hand), and its distance (closer objects produce a much larger echo). I chose the threshold by inspection to be high enough to prevent any random noise from registering as a received signal, but low enough to detect the smallest peak when an object is present.

As can be seen in figure 6, a large burst is seen at the receiver (yellow
waveform) just as the transmitter (blue waveform) is activated. To solve this problem, I generated a pulse to control when the receiver is allowed to trigger the sample and hold. I generated this pulse in the same fashion as the transmission pulse, using two comparators and $v_{ref}$. I chose the width of the pulse by inspection. The width of the transmission burst is around 2ms, so I decided to begin the receiver pulse 2ms after the transmission ends - which gave me a threshold of 7.8V. I decided on the length of the pulse according to the farthest distance I would need to measure. The second peak seen on the receiver waveform is from the sound bouncing off of the ceiling. I chose to end the pulse just before the reflection from the ceiling is observed, or 7.2ms after the initial burst ends, resulting in a second threshold of 8.8V.

Similarly to the transmitter circuit, digital logic is used to enforce the timing of the receiver. Only when the corresponding beam-break voltage, the receiver timing pulse, and corresponding $v_{rcv}$ signal are high should the sample and hold be triggered. One problem that arose was that the width of the sample and hold trigger pulse depended greatly on the character of the echo signal at the receiver, and these signals are very hard to control. If two objects are at the same distance from the sensor but one reflects sound better than the other (maybe a hand with closed finger vs. splayed fingers), the resulting echoes will have different amplitudes. The echo with

**Figure 5:** A schematic of the receiver circuit. Section (a) contains a 20dB inverting amplifier, and section (b) contains a peak detector. In practice, four of these circuits are needed, one for each sensor and each the respective $v_{rcv}[1−4]$. 
Figure 6: The blue waveform shows the transmission pulse, and the yellow waveform is the signal at the receiver after being amplified and low pass filtered. The first peak on the yellow waveform is the burst from the initial transmission; the second peak is the reflection of the signal off of the ceiling.

Larger amplitude will create a longer sampling pulse because of the RC time constant of the peak rectifier (fig. 5), and different voltages will appear at the output of the sample and hold despite the objects being at the same distance. To combat this problem, I added a monostable 555 timer just before the sample and hold trigger. This way, the 555 timer will output a pulse of uniform width regardless of the strength of the echoed signal. I chose the width of the monostable pulse to be 11ms, long enough to ensure the sample and hold cannot be triggered more than once per cycle.

To save on space, I used a couple of additional logic gates in my implementation. The \( v_{rcv} \) signal (fig. 5) from each sensor is input into an AND gate along with the corresponding \( v_{laser} \) signal. These four voltages (of which at most one can be high at any time) are sent through an OR gate, to create collective receiver signal. That signal is then sent to the AND gate along with the delayed receiver timing pulse, and the output of this AND gate is tied to the trigger of the monostable timer. This design uses two MC14081B quadruple 2 input ANDs, one MC14071B quadruple 2 input OR, and one LM555 timer.

**Areas for Improvement**

With more time, there are a few key things about this design that I would change. Most notably, I found that the laser beam detection was not practical
Figure 7: A schematic of the timing circuit for the receivers. Section (a) shows the CMOS gates managing the signals from each receiver. Section (b) shows the voltage comparators generating the delayed timing pulse. Section (c) shows the monostable pulse generator that is triggered only when an echo has been detected on the activated sensor while the timing pulse is high. Section (d) shows the sample and hold circuit that samples $v_{ref}$. This output is low pass filtered, and then sent on to the VCO.

Figure 8: A high level block diagram of the distance sensing module.
when combined with the ultrasonic sensors. When I attempted to put the acrylic beam containing the photodiodes atop the structure, I found that closing the box created an unacceptable amount of noise as the ultrasonic signal bounced from wall to wall. I would have liked to find another way to attach the photodiodes, but this is tricky since they must be very well aligned with the lasers in order to work reliably. Another improvement could be on the LPF at the output of the sample and hold circuit. I added this RC circuit to get rid of the ‘sawtooth-y’ look of the output voltage that would have caused unpleasant pitch oscillations. The trade-off was that the rate of pitch change became pretty limited. Another solution that I fiddled with was adding (another) sample and hold, that would only sample the ‘sawtooth-y’ output at times when it is known to be constant (i.e when it is holding rather than sampling, so any time after the receiver timing pulse has ended), but could not get this scheme to work. With more investigation, I believe a better solution can be achieved.

2. Analog Synthesizer  
by Matthew Okabue

2.0 Overview
The analog synthesizer portion of the circuit contained a few important modules. The synthesizer receives two important inputs, the first one is the effect control which is decided by a laser break beam system. The other is the tone control for the VCO which is determined by an object’s distance from the sound emitter. The first of these being the input to the system which used an array of MOSFET switches to determine which effect circuit to activate. This worked because the first part of the project, the laser harp structure will give me a high voltage when one of the lasers is broken, which is applied to the gate of an n-type MOSFET. When this voltage is on the MOSFET will conduct current and allow the input voltage to pass through to the voltage controlled oscillator (VCO). In the VCO, the DC control voltage will cause the VCO to emit a signal at a frequency that is related to the input voltage. Either the square or triangle wave output is used as the input for the effect circuits. Then this modified waveform is sent to the Class G Amplifier.
2.1 Tremolo

There were four effects circuits in total, the first one was the tremolo circuit. The tremolo circuit is unique in that it oscillates in volumes so it has two frequencies of oscillation. It has its normal oscillation frequency that determines the pitch and then slower frequency that determines the rate of volume oscillation. Two potentiometers in the circuit allow the "depth" of the oscillation to be controlled, which decides how close to zero it oscillates in voltage. In addition, a potentiometer controls the speed of oscillation that can swing the frequency from a few Hz all the way to a few hundred Hz. This circuit was the first one I built and it was a little difficult getting the oscillation circuit set up, but other than that, it was straightforward. Until I connected the VCO (voltage controlled oscillator). Once I connected the voltage controlled oscillator, the triangle wave output from the VCO caused the tremolo to demonstrate abnormal behavior. It started alternating the duty cycle instead of alternating the voltage. There was also the issue of the output voltage. The way I designed it caused the output voltage to be around 16 volts peak to peak, which was too large to be an acceptable input to our Class G Amplifier, so I had to attenuate the output.
2.2 Armstrong Green Ringer

This circuit was intended to emulate the ring modulation effect, despite the fact that it was not an actual ring modulator. I found this out after I had built the circuit so I decided to make the best of it. It made an interesting sound when the input waveform was a square wave, so I decided to use that output of the VCO. When a square wave was the input, the ringer put out a consistent pattern of beats. Building this circuit was fairly straightforward, but it was hard to test, because I was not sure what the output waveform should look like. I ended up doing a lot of tests and examining the output waveform on the oscilloscope until it looked like a unique sound. When I actually played it, despite the way the waveform looked the sound was just a shrill tone. This circuit was difficult because it seemed to work differently every time I tried it.

2.3 Big Muff Pi

The Big Muff Pi distortion circuit is a very popular effect circuit and has been used as a guitar effect since the late sixties. This circuit consists of four transistor stages. The first stage is a boost stage, the next two stages are
clipping stages, and the last stage is a volume recovery stage. This circuit was a bit complicated in the beginning due to the large amount of feedback connections that had to be made. There was also the issue that the circuit was much more effective on low frequency signals than higher frequency ones. The biggest problem by far was that the output voltage was very small, only a couple hundred millivolts. This causes the signal to be very noisy due to the large amount of 60 Hz noise coming from the wall outlets.

2.4 Phaser

The final circuit was the phaser. This circuit was fairly straightforward to build since the design was modular. I added one additional stage so that both sides of the 353 would be full and stabilizing the side that is not in use.
would not be a concern. After constructing the circuit I started testing and I ran into this odd issue where the amplitude of my signal would drop off completely in the middle of my range. The range for all of my effects circuits was supposed to be 0-5 volts. This circuit got up to one volt and then fell silent until the voltage reached 4 volts. This silence in the middle was not useful, so I tried to scale the range so that the 0 to 1 volt section could be spread over the range, but amplifying the signal introduced more noise into the system. I was not sure what to expect the circuit to sound like from the waveform, but it looked interesting. When it played it sounded very similar to a fart noise, so this circuit became known as the fart effect.

2.5 MOSFET Switches

To control the switching between the different effect circuits, a simple MOSFET switching set up as used. If the laser string was broken, it would send a voltage to the gate of one of the MOSFETs and that MOSFET would begin conducting and send the VCO control voltage through to the input of the VCO that controlled that specific effect. This used an on/off decision system so it was purely digital, but a useful way to decide which circuit to use. As shown in figure 14, the on/off input is at the gate of the MOSFET. I ran into some issues because I accidentally connected the MOSFETs in the wrong orientation and since there was a diode bypass, the circuits were always conducting and I could not turn my effect circuits off.
2.6 VCO

The voltage-controlled oscillator was designed with two outputs, one square wave output and the other was a triangle wave output. The frequency of the output waveform increases as the DC input voltage increases in magnitude. It used the LM358 which was a dual package Op-Amp. I ran into some issues with the VCOs initially because the output frequencies were very slow and limited to the range from 0 to 400Hz. This problem was addressed by using a capacitor with a small capacitance value. I was able to get the maximum up to 4.5 kHz, but after that point using a smaller capacitor had little to no effect.

2.7 Final Integration and Improvements

Once everything was completed and tested it was time to integrate our system. So we brought together the laser, supersonic apparatus, the synthesizer, and the amplifier. This caused many headaches in my circuit. Once we connected everything the output voltages from my effect circuits were highly attenuated, this was not ideal, because that introduced a lot of noise into the system.

If I were to do this project again, I would start with testing the sounds on the speakers because even if the waveform looks interesting on the scope, it may be unexciting to the ear. I would also have wanted to make sure I know the specs of my teammates’ modules. I also would try to avoid making four VCOs, I think that 1 or 2 would be more than you will ever need. I would
have also tried to do more with the mixing of two or more of the effects, because I think if that is well handled it could become something really great. Thought would have to be put into what sounds go well together, because random mash ups can be good, but having a planned mash up where the two sounds complement each other can definitely be a positive thing.

3. Class G Amplifier Module  
by Chad Uyehara

3.0 Overview

The amplifier module for Laser Harp consisted of a Class G audio amplifier capable of delivering about 100 Watts of continuous power. A Class G amplifier is similar to a class B amplifier, but incorporates multiple power rails in order to increase the efficiency. A class G amplifier exploits the property that music has a high crest value. Crest value is the ratio of peak amplitude to the RMS value of the waveform. This property shows that music does not continuously play at a peak value, but rather oscillates to produce a sound. Therefore, a Class G amplifier can switch to a low

Figure 15: A schematic of the VCO.
Figure 16: Class G Amplifier Block Diagram. Input stage, voltage amplifying stage, output stage, and the negative feedback loop.

power rail for the non-peak signals, and then switch to a higher rail when the waveform exceeds the lower rail. This can greatly increase efficiency because the amplifier does not need to be powered constantly by a higher voltage rail.

My implementation of the Class G amplifier was designed in a typical 3-stage configuration. This consists of an input stage, voltage amplifying stage, and the output stage. The amplifier will be powered by a ±15V rail and a ±30V rail from an existing power supply and will be designed to power an 8-ohm, 100 Watt mid-woofer in order to play the sound coming from our laser harp.

3.1 Input Stage

The input stage was designed with a differential pair driven by a current source and balanced with a collector current mirror. The differential pair allows the feedback signal to be subtracted from the input and generate an error signal that drives the output. A differential pair input was chosen over the single transistor input stage because of its low DC offset and the improved linearity. The collector current mirror increases the transconductance of the input stage and balances the current through the differential pair at the input. The transfer characteristic of the differential pair goes as:

\[ I_{out} = I_c \tanh \left( \frac{V_{in}}{2V_t} \right) \]

[1, pg. 79]

To reduce the nonlinearity, the tail current is increased, thus increasing IC, and then a resistor is added to compensate for the drop in internal resistance. The external resistor must be added in order to bring the transconductance back to its original value (since \( g_m \) was increased by the increase in tail
current). Increasing the tail current also helps to increase the slew rate which is directly proportional to $I_e$. A tradeoff for the increase in linearity and slew rate is the increase in DC offset. An additional DC offset is produced because the extra resistors generate a Johnson noise given by the equation:

$$V_{off} = \left( \frac{a_{100}}{100} \right) \left( \frac{k \times R_e}{2} \right)$$

[1, pg. 87]

With the design of the input stage with a tail current of 6mA and emitter resistors of 100Ω, the offset is a small 3mV. Therefore, showing the benefits of increasing tail current drastically outweighs the disadvantages as 3mV is small compared to other sources of offset voltage. For the current mirror, a basic configuration was chosen to decrease complexity for the tradeoff of slight differences in the base voltages. A Wilson mirror could have been designed for this application, but if a transistor with a high beta was chosen, then the difference would be negligible (<1%). The current mirror chosen for this application mirrors current with a beta dependency by the equation:

$$I_{out} = \frac{I_m}{1+\frac{1}{\beta}}$$

[1, pg. 84]

The Wilson mirror has the dependency of beta squared which improves the mirroring. The MPSA06 transistor was chosen for its beta value of 100 which would mirror the current accurately. The current mirror transistors also had emitter degeneration resistors. The tail current source was designed with a typical pnp current source. The collector current equation for the current source used is as follows:

$$I_C = \frac{V_t}{R_e}$$

[1, pg. 86]

Therefore, the resistor for a 6mA current source is approximately 100Ω and for a 10mA current source is approximately 60Ω. The 10mA current source will be used in the voltage amplifying stage.

A challenge with the input stage was optimizing the current source and emitter degeneration resistors for the differential pair. The optimization was performed through simulations in LTspice where multiple values for the current source were chosen and emitter degeneration resistors calculated to match the current, then tested to see if a desirable output was produced. This testing involved many variables including slew rate and DC offset, as they both are dependent on the tail current and emitter resistors.
3.2 Voltage Amplifying Stage

The voltage amplifying stage consists of a Darlington configuration inside the Miller loop driven by the dominant pole capacitor, or $C_{dom}$. $C_{dom}$'s main function is to set the dominant pole frequency such that the total loop gain falls below unity to prevent high frequency oscillations. $C_{dom}$ is determined by the equations below for high frequency gain and the dominant pole frequency. $C_{dom}$ also determines the slew rate and linearizes the voltage amplifying stage. The Darlington configuration is used to enhance the beta which in turn increases the amount of local negative feedback by increasing the low frequency gain.

$$LF_{gain} = g_m \beta R_c$$
$$HF_{gain} = \frac{g_m}{\pi C_{dom}}$$
$$DominantPole = \frac{1}{\pi C_{dom} \beta R_c}$$

[1, pg. 62-3]

3.3 Output Stage

The output stage is where the class G operation is implemented. A typical output stage consists of a pair of emitter follower transistors that are biased slightly on in order to reduce crossover distortion. This idea was then expanded in order to make a second pair of emitter follower transistors that could be triggered on at certain times (peak power points). Going over the
Figure 18: Voltage amplifying stage of the amplifier. Darlington configuration for beta enhancement and full voltage swing (bottom) driven by a current source (top). The bias generator is in between the current source and voltage amplifier.

In the typical configuration, there are two emitter followers, one being the driver and the second being the output. The inner transistors are the drivers which share an emitter resistor and the outer transistors have emitter resistors of low values. For the higher voltage rail transistors, the driver has an emitter resistor that establishes the current in the base of the output transistor. The output is then tied to the collector of the lower voltage output transistor.

The bias generator for a typical class B output stage is a $V_{BE}$ multiplier configuration. For class G operation, another set of bias generators needed to be present in order for the switching to occur at the right time. To achieve a bias for the high voltage output stage, Zener diodes were used to get an appropriate voltage drop. The bias was calculated to be approximately 1.4V for the two low voltage rail transistors, and 3.5V for the high voltage rail transistors. The $V_{BE}$ multiplier is a well-known circuit and its resistor values can be found with a simple equation. For the lower voltage rails switching capabilities, a Schottky diode was used because they have fast switching times and low voltage drops, which in turn reduces distortion at the output. All the driver transistors are powered by the upper rail in order to maintain its linearity. If the drivers were powered by their own respective rail, when switching occurred, the driver $V_{CE}$ would jump, causing a jump.
3.4 Feedback Loop

The negative feedback consists of the output signal driving one leg of the differential input. A method called bootstrapping the input was used which consists of a resistor network from the input leg and a bootstrapping capacitor to the other leg. Bootstrapping the input implies that the input will have a low resistance DC path, but a high AC impedance. On the schematic, C3 acts as the bootstrapping capacitor while R3 and R4 balance the input by creating an equal resistance to that of the feedback resistor R16. An isolation resistor, R9, was added in order to isolate the feedback point from stray capacitance. Finally, R15 and C8 help to maintain low frequency response as a consequence of the low impedance feedback loop.
3.5 Protection Circuitry

The diode configuration (D1-4, fig. 21) protects C8 from the amplifier saturating in either direction. If the voltage in either direction is offset by more than 1.2V, the diodes will turn on and ground the input. There is also a single slope VI limiting protection circuitry (fig. 22) that protects the output from being overloaded. Q12 and Q13 in the design take care of this protection circuit. The $V_{CE}$ and current through the output stage is now sensed by the VI limiters and can pull the base current from the driver down to protect the output transistors and speaker load.

3.6 Simulation

Using the parameters explained through each of the previous sections, a simulation with LTspice was performed to verify correct operation. (See figures 23-4.)
Figure 22: Single Slope VI Limiting.

Figure 23: From the simulation, the class G operation is shown by the upper transistors turning on at 15V (the lower rail value). The output shows a swing of at least 20V into an 8 ohm load, giving a power rating of 50 Watts at this operation. For the maximum rating, the voltage swing is the value of the upper rail which is 30V, giving a continuous power rating of 112 Watts. For a 1V input at 1kHz, the gain is about 20.
3.7 Construction

The class G amplifier was built on a breadboard for ease of modifications. Since this is a power device, the components have a chance of overheating and frying, so a breadboard was essential to use in the testing phase. The difficulties with running the amplifier off the breadboard were the heatsinking capabilities and added distortion due to stray capacitances of the breadboard. The heatsinking was an issue because the output transistors needed to be mounted onto a large heatsink, so the solution to this problem was using jumper wires to connect from the pins of the transistors to its relevant spot in the breadboard. This solution may have caused distortion because of the long wires running to and from the transistor in the output stage. Using a breadboard in general will cause noise in a very sensitive application such as audio, so after testing was finished, a PCB was designed and ordered. The PCB did not come in on time for the presentations of the final product, but it would have solved the difficulties mentioned earlier. The output transistors would be able to easily mount to a heatsink straight from the PCB and signal traces were taken into consideration of the PCB layout to minimize distortion.

Figure 24: Simulating the bandwidth with a 20kHz input. There was no attenuation so the amplifier’s slew rate is high enough for audio application.
3.8 Testing

Testing consisted of checking each stage of the amplifier and the different modules that contributed to correct operation of the stage. The testing that was performed was identical to the checkoff list that was made for the final presentation. For overall performance, the amplifier should produce at least 50 Watts of continuous power into an $8\,\Omega$ load. The class G operation should be functional, shown through probing the output transistor and seeing the switching states. With switching, there should be little to no timing errors or distortion from going to the high voltage rail and back down to the low voltage rail. The bandwidth of the amplifier should range from 20Hz-20kHz and beyond such that there is no attenuation in the audible range. The current sources for the input stage and voltage amplifying stage should be producing a constant 6mA and 10mA, respectively. Finally, the overload protection (single slope VI Limiting) should be functional. This was tested by lowering the load impedance ($4\,\Omega, 2\,\Omega$) and then measuring the current/voltage through the output transistors to make sure they are not overloaded.

Further Improvements

Some designs that were considered would increase linearity and decrease distortion of the current design, but would add multiple levels of complexity. If more time were permitted, a double input stage with a "push-pull" voltage amplifying stage (VAS) could have been implemented. The purpose of a double input stage (two differential stages) would be to drive the top and bottom of the VAS. A push pull system is considered to be the most efficient for current delivery which is why it would be an improvement. The complexity it adds to the design is the addition of a differential pair for the input and the design of a push pull VAS system with the emitter follower configuration.

As for protection circuitry, a dual slope VI limiter could have been implemented other than a single slope. The addition of a Zener diode and a couple resistors make the difference between single and double slope, but the protection circuit was not the main focus of the project. Therefore, the less complex, single slope VI limiter was designed such that there would at least be some protection circuitry for the output.
References