BASS GUITAR PREAMP DESIGN

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I'd like to dedicate this work to my long term partner Louise. Without her love, patience and encouragement this work would not have been possible.

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CHAPTER 1

1.1 HISTORY

The first occurrence of an external bass guitar preamplifier came in the form of passive direct inject (DI) boxes. In the 1960's The Beatles were unable to be heard over their loud audiences, the DI box was invented to give public address support to their backline. Although bit strictly a preamplifier, this was the first instance of interfacing the bass guitar with line level equipment.

In 1975, Sterling Ball tested a Leo Fender bass design, the Stingray. This bass is regarded as the first bass utilising 'Active' electronics. Containing a 9v battery and a 2 band Equaliser, the bass outputted at line level removing the need for a DI box. Passive DI boxes were still used to convert the unbalanced jack signal into a balanced XLR signal.

As technology has progressed over the past 30 years, the opportunity for new preamps designs has constantly emerged. Filters, distortion and compression are common characteristics found within today's units.

1.2 AIMS & OBJECTIVES

The aim of this project is to gain a thorough understanding of the electrical design elements of an audio application, specifically bass guitar preamps. The artefact will be a functional bass preamp designed to operate within the parameters of a technical specification discussed in this report. The artefact will address the problem of interfacing the bass guitar with audio line level instruments. Thorough testing of the unit will be undertaken and documented in a second report.

1.3 SUMMARY OF REPORT STRUCTURE

In chapter 2, considerations regarding the theoretical and physical design are discussed. With reference to existing solutions, the individual sections of a bass preamp are looked at closely. Chapter 3 outlines an overall plan for work to be carried out on the artefact. Initial theoretical schematic design, specification and bill of materials based on the design considerations discussed in chapter 2 will also be included. Chapter 4 looks at the build process and the issues raised during production. Chapter 5 documents the testing and test results of the preamp. This report concludes with Chapter 6. Chapter 6 discusses future development plans and looks to summarize the entire project.

Chapter 2

Fundamental electronic theory referenced throughout this section can be found in Appendix 1.

2.1 Existing Solutions

Commercial products are readily available to perform the task of bass guitar pre-amplification. Designs address different problems but fundamentally are the same solution. A full specification of the leading bass preamps is listed in appendices 2. These solutions will be used to create a specification for the artefact.

2.2 Input stage

2.2.1 Input Signal

In order to design the input stage of any electronic circuit, the output of the previous stage needs to be studied.

Datasheets and specifications for pickups often don't give peak voltage readings (Vpmax). Using a 10M probe of an oscilloscope peak voltages have been recorded from 3 bass guitars. This method is included in appendix 3. Figure 1 shows the peak output voltage of 3 bass guitars. It is noted that the following readings are worst case scenario measurements. They were recorded with significant attack on the bass strings.

Bass	Pickups	Vpmax
Fender Precision Bass	Split Coil Humbucker	0.5189
Fender Jazz Bass	2 x Single Coil	0.3894
Warwick P/J	Split Coil Humbucker, Single Coil	0.3567

Figure 1: Peak Voltage of 3 bass guitars.

The amount of gain needed to bring instrument level to line level can be calculated from the following equation.

$$dB_{gain} = 20 \times log\frac{V_{out}}{V_{\text{in}}}$$

Where:

$V_{in} = Instrument\ Level = 0.6V_{pk}$

$V_{\text{out}} = Line\ level = 1.737V_{pk}$

$\therefore$

$$dB_{gain} = 9.232$$

I have rounded the largest reading up to 0.6V as an absolute worst case scenario reading to cover the many different bass guitar designs on the market. As active basses will produce significantly higher output swings than passive basses, attenuation is needed at the input stage.

2.2.2 Attenuation

"An attenuator or pad is an arrangement of non-inductive resistors in an electrical circuit used to reduce the level of an audio-or radio-frequency signal without introducing appreciable distortion." (Ballou, 2005)

It is commonplace to add a switchable 20dB pad to account for active basses. Figure 2 shows an unbalanced L-type attenuator pad:

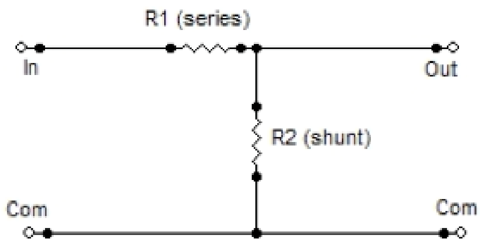


Figure 2: L-Type Attenuator, Chinn, R (n.d.) [online image]

The L-type attenuator pad is simply a voltage divide. In appendix 4, the component, Impedance and attenuation calculations are listed.

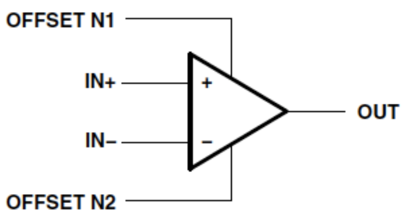
2.2.3 Impedance Bridging

"If the load impedance is 10 times more than the source impedance, it is called a "bridging" impedance. Bridging results in maximum transfer of voltage from source to load." – (Barlett, B. & Barlett, J. 2002)

The input impedance of the preamp must be set high enough in order to transfer maximum signal from the bass guitar. Typically instrument output impedances are around 600Ω with typical input stages set at 1MΩ. This is known as impedance bridging and is a common technique to ensure the signal is suitably transferred from one device to another. The alternative approach is called impedance matching where both input and output impedance are set to be equal.

2.3 Operational Amplifiers

Figure 3 and 4 show the modern schematic layout physical layout of an Operational Amplifier (Op amp) respectively.



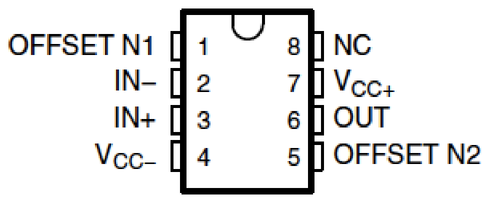


Figure 3 (Left): Op Amp schematic, Figure 4 (Right): Physical Op Amp layout schematic.

National Instruments (n.d.) [online image]

Pins IN+ and IN- refer to the non-inverting and inverting input respectively. Vcc+ and Vcc- are the inputs for the power supply. Offset N1 and N2 control the offset null of the op amp so the input transistors are perfectly balanced.

Operational amplifiers can vary significantly operation and quality. The following sub-sections look at the individual characteristics and their suitability for an audio application.

Slew Rate

Although not directly describing op amps, Ramsey and McCormick define slew rate as:

"The ability of an amplifier to respond to high level transients" (Ramsey & McCormick, 2009)

Slew rate is represented in V/us (Volts per microsecond). Insufficient slew rate may cause distortion at higher frequencies due to their inability to react quickly enough to a sudden change in voltage. Using the calculations stated in Appendix 5, the theoretical minimum slew rate for an audio Op Amp = 0.43V/us. Realistically a much higher slew rate would be implemented, as the sharpest of transients cannot be accounted for. Consideration must be paid to the slew rate and bandwidth in audio circuitry.

Unity Gain Bandwidth

George B. Rutkowski defines the perfect bandwidth of an opamp:

$$BW = \infty$$

"An infinite bandwidth is one that starts and dc and extends to infinite cycles/second (Hz)" (Rutkowski, 1993)

Common Mode Rejection Ratio (CMRR)

The CMRR defines the rejection of common signals at 2 inputs of operational amplifiers. For example, a good op amp would reject DC offset by a ratio upwards of 100dB. This would be of importance in a microphone preamp design where the input is balanced and a differential amplifier is required to unbalance the signal. As a bass guitars output is unbalanced, the CMRR characteristics of an operational amplifier are of less concern.

Number of Channels

Individual Op Amps can process multiple channels saving space and cost. An example is the National Instruments TL074 that has 4 channels. If a circuit requires many Op Amps, then it may be desirable to use integrated circuits that can offer more channels. This will save on cost and PCB/Strip board space.

Max Supply Voltage

An Op amp has a maximum voltage that can be applied to its ±Vcc pins. Too much applied voltage can damage the Op Amp. If an operational amplifier can't accept a large enough supply voltage, then there is a chance that an audio signal will clip the Op Amp. For example, an Op Amp with a supply of +/- 5v will clip an audio signal that has an amplitude of +/1 6v.

2.3.1 Specifications

Appendix 6.1 shows the study of the most common op amps found in commercial and DIY audio applications. Appendix 6.2 includes the specification of the ideal op amp.

In reference to this Appendix 6.1, the appropriateness of different Op amps for audio applications can be reviewed. The TL series of op amps proves value for money characterised by its lower slew rate and bandwidth. The National Semiconductor LM617X shows much more superior specifications to the OPA627 at a fraction of the price.

The sonic difference of these Op Amps is subjective. The quicker slew rate of the LM617X and OPA627 will reduce high frequency distortion. That said, the TL07X data sheet states that harmonic distortion is at 0.003%, human hearing is not sensitive enough to recognise this. This is also a common THD+N reading for audio products working at 20Hz-20kHz +4dBu. Therefor, appropriate for an audio circuit.

2.3.2 Op Amp Rules

Op Amp functionality is based upon the following rules that must be recognised when designing with Op amps.

1. Op amps will always attempt to make the difference between inverting and non-inverting inputs zero.
2. The input of an op amp draws no current.

(Horowitz & Hill, 1989)

2.3.3 Amplification

There are two main methods of amplification with op amps. Inverting and non-inverting.

Non-Inverting Amplifier (Figure 5):

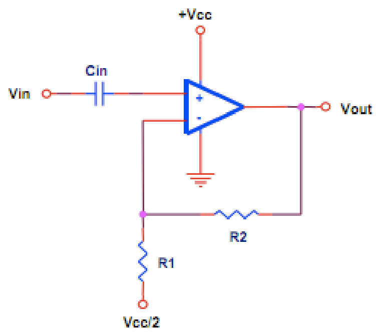


Figure 5: Non-Inverting Amplifier

National Instruments (n.d.) [online image]

The signal is fed into the non-inverting input. A feedback loop is sent to the inverting input. The voltage divider controlled by R1 and R2 controls the amount of gain. Gain is calculated by:

$$gain = 1 + \frac{R_2}{R_1}$$

The Op amp tries to make the inverting input the same as the non-inverting. As the signal first passes through a voltage divider, the output is raised to compensate for the voltage loss.

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INVERTING AMPLIFIER (FIGURE 6)

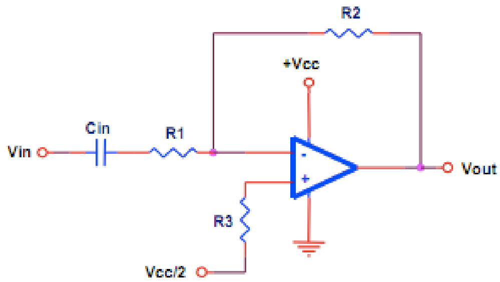


Figure 6: Inverting Amplifier

National Instruments (n.d.) [online image]

The Signal is coupled with a feedback loop into the inverting input of the op amp. The voltage divider set by R1 and R2 control the gain. The gain is calculated by:

$$gain = -(\frac{R2}{R1})$$

Inverting amplifiers are said to have negative gain. This does not mean attenuation; it refers to the 180 degrees change in polarity. The output of the op amp will always try and make the inverting input equal to Vcc/2. Therefore, if the input is positive, the op amp will output negative and vice versa.

2.3.4 BUFFER AMPLIFIER

Buffer stages (Shown in Figure 7) are an integral part of preamplifier design. Horowitz and Hill state the buffer amplifier:

"is simply a non-inverting amplifier with R1 infinite and R2 zero (gain = 1)."

(Horowitz & Hill, 1989)

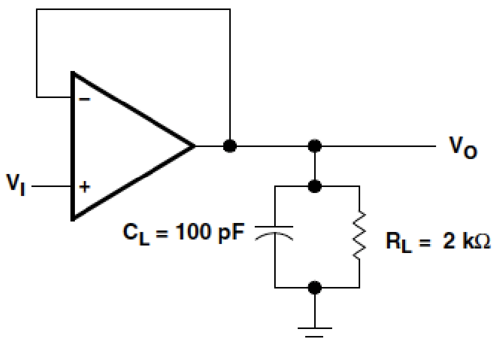


Figure 7: Non-Inverting Buffer Amplifier

National Instruments (n.d.) [online image]

It isn't uncommon to see several buffer stages in a circuit; examples of this can be seen on the Triangular Ear Bass 21 Schematic (Appendix 14). They are very useful at the input stage of circuitry as they have a high input impedance and a low output impedance. This draws maximum voltage from the preceding circuit and supplies maximum current to the proceeding circuit.

2.4 Filtering

"A filter is a device or network for separating waves on the basis of their frequency"

(McMannus, 2005)

Appendix 7 includes filter terminology referenced throughout this section.

2.4.1 Reactance

In order to change the frequency response of a signal, linear components sensitive to a change in frequency are needed. Linear components don't change frequency but they do dictate frequencies contained within a signal, in higher terms, tonality. Horowitz and Hill state:

"The output of a linear circuit, driven with a sine wave at some frequency f, is itself a sine wave at the same frequency with, at most, changed amplitude and phase"

(Horowitz & Hill, 1989)

Linear components such as capacitors and inductors have frequency dependent resistance known as reactance. Reactance is treat similarly to resistance but as it is frequency dependent, calculations become more complex. Calculations are found in Appendix 8.

In order to manage the complexity of reactance the term impedance is used to represent the overall resistance of a network. Ohms and Kirchoff's laws still apply but all terms need to be referenced as impedance and not resistance.

2.4.2 Passive Filters

High Pass

Figure 9 shows a typical High pass Circuit, a voltage divider formed by a capacitor and a resistor.

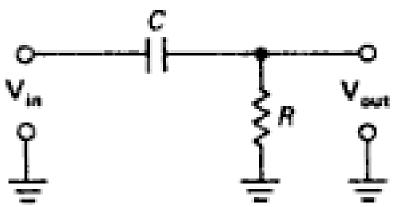


Figure 9: Passive, first order High Pass Filter

[Image] In: Horowitz, W. & Hill, P. (1989)

Using the following equation, the voltage output at a specific frequency can be determined. See appendix 9 for complete equation list.

$$Vout = Vin\frac{2\pi fRC}{[1+(2\pi fRC)^2]^{\frac{1}{2}}}$$

At low frequencies the series capacitor has a high reactance creating and open circuit. As the frequency increases towards its cut-off frequency, the capacitor's reactance decreases and shorts the circuit. Figure 10 shows a magnitude response plot of a high pass filter.

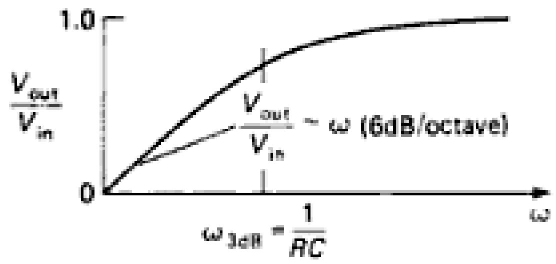


Figure 10: High Pass Magnitude Response.

[Image] In: Horowitz, W. & Hill, P. (1989)

Low Pass

By reversing the order of R and C a low pass filter is formed (Figure 11). The reactance within the capacitor is at a high resistance at lower frequencies causing the curve to roll off at the other side of the cut-off frequency (Figure 12).

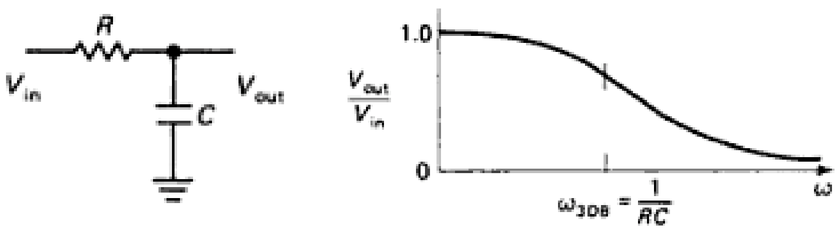


Figure 11 (Left): First order Low Pass Filter. Figure 12 (Right): Low Pass Magnitude Response.

[Image] In: Horowitz, W. & Hill, P. (1989)

The cut off frequency is standardised as the point where 3dB of attenuation occurs as described by Wayne Corr:

"When this (Frequency Cutoff) occurs the output signal is attenuated to 70.7% of the input signal value or -3dB (20 log (Vout/Vin)) of the input."

Storr, W. (n.d). [online].

Passive filters are stable due to their simplicity but to create steeper roll offs, more orders must be used. The reactive components will interact with each other and circuits become undesirably large. Their inability to produce gain also causes problems for creative equalisation. In order to address these problems, we can apply passive components to an active network.

2.4.3 Active Filters

Active filters have long been the standard in audio instrument design. Lancaster states:

"An Active filter is some combination of integrated circuit operational amplifiers, resistors and capacitors that does things that normally could be done only with expensive inductor-capacitor passive filter combinations"

(Lancaster, 1996)

Using capacitors in tandem with negative feedback filters is used to mimic the operation of inductors. These "gyrator" networks bring the best of active and passive filters together.

2.4.4 Multiple-Feedback Bandpass Filter Topology

Figure 14 shows a Multiple-Feedback Bandpass (MFBP) filter configuration credited to L.Huesman. Figure 14 uses an inverting amplifier with two feedback loops, one with high pass circuitry and the other with low pass. MFBP filters are popular in filter design for their flexibility, tolerance to sensitivity errors in components and can be tuned to almost any shape with a quality factor under 20.

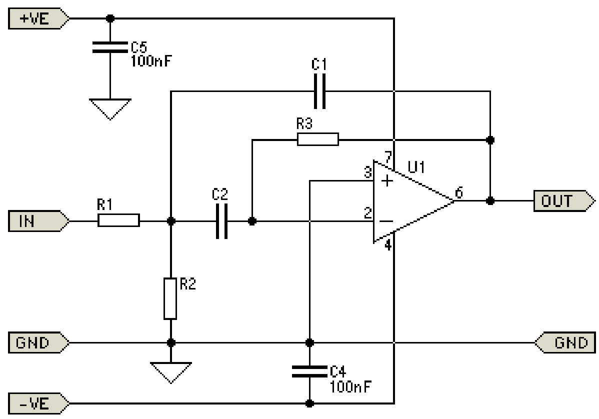


Figure 14: Multi-feedback implementation

Elliot, R. (2000). [online image]

The gain is controlled through the R1, R2 voltage divider. C1 = C2 and control the pass band.

Williams and Taylor state the transfer function of this design of MFBP filter:

$$T_{(s)} = \frac{sR_3C}{s^2R_1R_3C^2 + s2R_1C + (1 + \frac{R_1}{R_2})}$$

$Where:$

$T_{(s)}$ = Transfer Function

s = Complex Variable representing Amplitude ad Phase

Williams, A & Taylor, F. (1995)

Appendices 10 show equations and design parameters essential to component values and filter stability for the MFBP topology.

2.4.5 Boost/Cut Topology

Adjustable gain is a common and expected parameter for filters. This can be achieved by adopting a circuit patented by Kenneth James of Dolby Laboratories The band pass filter becomes isolated from the boost and cut summing amplifiers. Figure 15 is a full schematic layout of this topology.

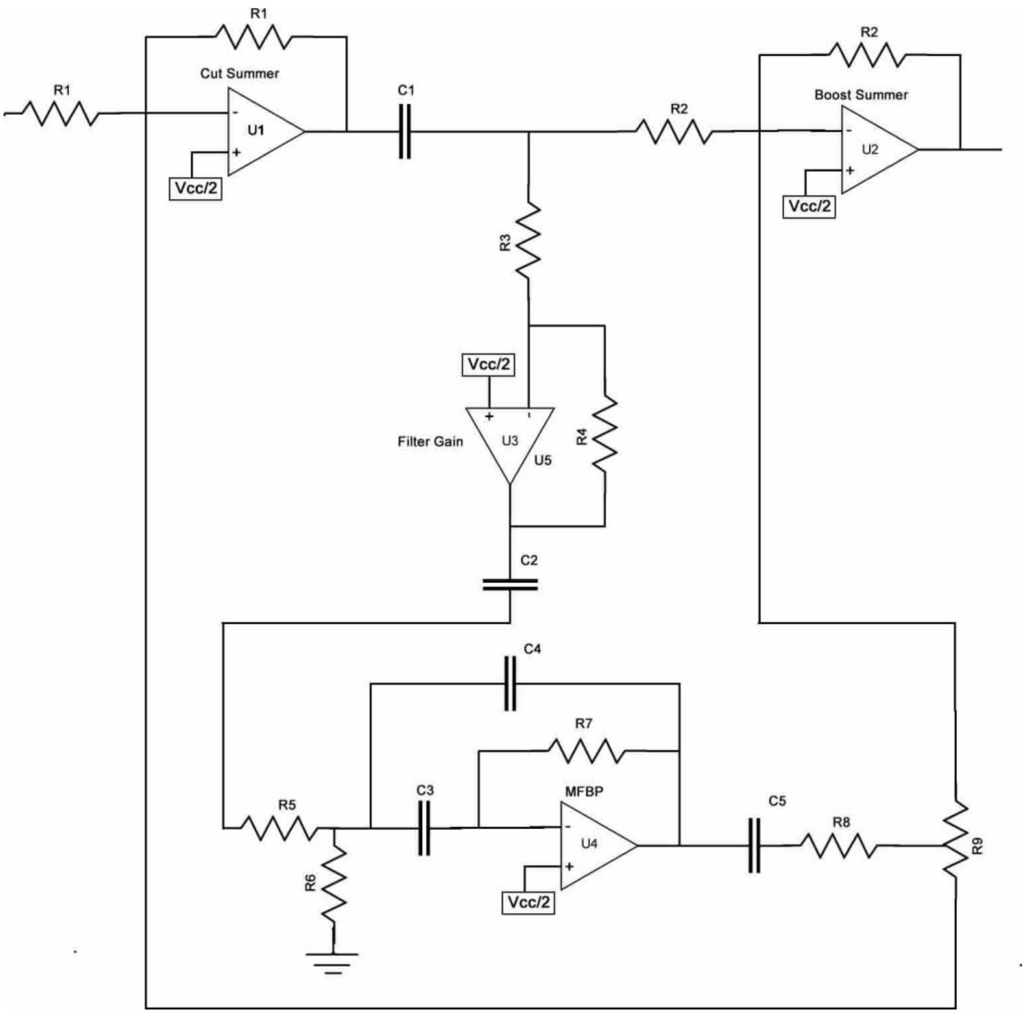


Figure 15: Boost/Cut topology

Brundy, K (1975).

This system is based on the phase relationships between the filters and the summing amplifiers.

Dennis A. Bohn correctly states that at the Boost Summer (U2):

$$Boost\ Output = 1 + kBP$$

And at the Cut Summer (U1):

$$Boost\ Output = 1 - kBP$$

$Where:$

$1 = Normalized\ Full\ Frequency\ Input$

k = Gain

BP = Band Pass Signal

Rane Coperations, Bohn, D (1986)

All of the op amps are inverting so when the signals are summed, their phase interacts. The signal summed to U1 causes phase cancellation at the pass band. The signal sent to U2 causes phase reinforcement at the pass band. U3 controls the gain of the pass band. As seen in Figure 15, U3 is a unity gain amplifier so when summed the output will sum to double or half (±6dB) the input. U4 is the MFBP topology as discussed in 2.4.4. R9 dictates how much signal is sent to U1 and U2. Many filter sections can be used in parallel under this topology as long as mixing resistors (R8) are used before the summers.

Using the filter specifications discussed in Chapter 3, Figure 16 shows the predicted magnitude response of a 4-band equaliser.

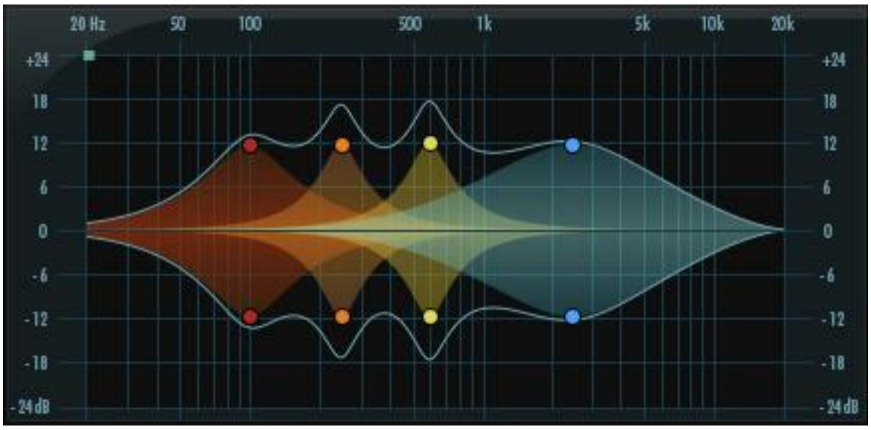


Figure 16: Expected Response of initially designed filters.

2.4.6 Filter implementation

Commercial Preamps offer ideas regarding how flexible filters can be implemented without sacrificing simplicity of use.

The John East J-Retro unit has a unique treble control. When boosted, the centre frequency is 3kHz. When cut, the centre frequency falls to 1.5kHz. Using the boost/cut topology this could be achieved by using a dual gang potentiometer. One gang boosting 3kHz and grounding the cut, the other cutting 1.5kHz and grounding the boost.

The initial schematic discussed in Chapter 3 and referenced in appendix 12, shows a switch selecting between 2 frequencies controlled by the mid control.

An in-depth look at equalisation tendencies in bass guitarist could open up a wealth of opportunities for improving filter functionality in bass preamps.

2.5 Power

2.5.1 Power Supply

Power supplies step down the mains voltage to give acceptable headroom for the given application. The 9v DC regulated power supply is a very popular supply for supply guitar pedals. Using a Parallel filter bank and series DC jack links it is possible to power 30 pedals from a single mains supply.

2.5.2 DC Regulation

Referring to the linear DC regulator, the basis of regulating power supplies, Marty Brown states:

> "It relies on the variable conductivity of an active electronic component to drop voltage from an input voltage to a regulated output voltage."
>
> (Brown, 2001)

This can prove inefficient, as power is lost across a DC regulator in the form of heat. Despite this, regulated supplies show good performance at producing a clean voltage source for sensitive circuits to run off. As audio circuits can be susceptible to external noise, the DC regulated power supply is seen as the most efficient way to supply power from mains.

2.5.3 Split Power Rail

A split power rail is used to create a DC bias when using a single power rail. As seen in Figure 18, a voltage divider is used to half the supply voltage to properly bias the Op amp.

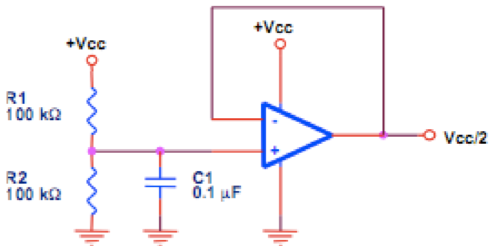


Figure 18: Split supply coupling.

National Instruments (n.d.) [online image]

Op amps need to swing between a positive and negative rail with reference to ground. In the above example, a virtual ground is created at Vcc/2. The signal swings between Vcc+ and 0v with reference to Vcc/2.

Op Amps are unable to swing all the way between their +Vcc and –Vcc rails. Therefore it is important that the split supply is as accurate as possible to get the most out of the op amp. If a bias is set at 75% of the power supply it is possible that little or no positive swing will occur. Using smaller voltage divider resistors can help achieve a more accurate division however, the amount of current drawn increases as resistance gets smaller. If this current is needed elsewhere in the circuit, more accurate voltage divider may have to be sacrificed by higher resistor values.

Integrated circuits such as the Texas Instruments TLA2426 are readily available to split single supply lines. They simply contain refined voltage dividers and buffer stages. The use of integrated circuits at this stage of the bass preamp circuit is more than likely unnecessary.

2.5.4 Decoupling

A final point to recognise is the use of capacitors to decouple the DC offset from the AC signal. This is important from protecting the bass or line level application from DC voltage. A capacitor in series is used to remove the DC from a signal. Precision values aren't required, a 0.1uF capacitor will suit audio applications fine. Smaller capacitors may be needed for higher frequency instruments.

2.6 Output

2.6.1 Operational Amplifier Based

By using buffers it is possible to output at low impedances, a very important characteristic if long cables are used. Inverting and non-inverting buffers can be utilised for creating a balanced output (pins 2 and 3) of an XLR connection.

Op amp output stages are often seen as an alternative to the transformer output stage. Transformer output stages offer good isolation between devices as well as internally setting output impedance. Unfortunately they are expensive to be considered in the initial specification.

2.6.2 Ground Lift

If a ground loop is contained within a system a switch must be installed to break the loop. A switch disconnecting the balanced ground connection is necessary and can be seen on the initial schematic in appendix 12.

2.6.3 Output Impedance

Typical line inputs and amplifier inputs have a minimum impedance of 1M. Therefore, an output impedance of 100K or lower is required. Practical outputs have an even lower impedance of around 600 Ohms. The output Impedance is set by attenuation networks tuned to attenuate as little as possible while still containing the required impedance.

2.7 Design Considerations

As discussed in previous sections of this chapter, various techniques are used to achieve gain, filtering, impedance balancing and power supplies. Consideration needs to be paid to the way these sections interact.

A 20dB L-Type pad and a Decoupling capacitor will form a High Pass Filter. Component values need to be chosen that will not interfere with the audio spectrum.

If two parallel resistive paths to ground (or virtual ground) are placed before the input buffer, the overall input impedance will be limited by the lower resistance value.

The circuit load cannot draw more current than available from the power supply. This shouldn't be an issue as most regulated power supplies at least 1.3A of current available. If too much current is drawn from a battery, the life span of said battery will be poor.

Chapter 3

3.1 Specification

A specification sheet for the preamp is included in Appendix 11. This gives rough design parameters that were used to create an initial schematic. Many of the parameters are based upon existing models that have proved successful. The specifications will be practically improved upon where possible.

3.2 Schematic

Appendix 12 is a schematic designed to the initial specification using the theory discussed in Chapter 2. The individual sections from the schematic will be mocked up on breadboard, gradually adding different sections to the circuit will help diagnose any hardware problems.

Appendix 13 includes all the necessary component calculations.

3.3 Bill of Materials

Figure 20 is the Bill of materials required to build the initial schematic. Over time as the design changes the list of materials will also change.

Component	Quantity	Notes
Resistors		
9K1	2	
15K	1	
2K	1	
30K	1	
470	1	
13K	1	
27K	1	
220	2	
11K	3	
22K	2	
330	1	
1M	2	
10K	11	
39K	1	
100	3	

1K	4	
Potentiometers		
10K	2	Log
100K Lin	4	Linear. Centre indent preferable
Capacitors (Ceramic)		
0.1uF	13	
0.01uF	2	
0.22uf	2	
0.18uF	2	
10uF	9	
100uF	2	
Op Amps		
TL07X	13	Various combinations of TL071,72 and 74.
Other:		
DPDT Switch	2	
SPST Switch	1	
Breadboard		
Jack Sockets	2	
XLR Socket	1	

Figure 20: Initial Schematic Bill of Materials.

Chapter 4

4.1 Modular Design

The circuit was initially to be built on one piece of strip board; this would make troubleshooting the circuit difficult. In order to maximise the efficiency of testing and troubleshooting, the circuit was designed in a modular fashion. The circuit was split into nine separate boards; power, input, input pad, input gain, filter summers, filter amps, passive filter networks, output gain and output. Appendix 15 shows photos of the individual boards. Removing modules from the design became crucial when testing the preamp; it made examining the interaction between boards more practical and offered the opportunity of experimentation. The modules were built on 2.54mm spaced strip board, proving useful for prototyping as well as containing permanent components. The filter modules and gain stages, due to their increased complexity were first built and tested on breadboard. Tracks were broken using a drill bit, this helped maximise the space available on the board.

The modules were connected through molex connectors and terminals. This solution offers a stable electrical connection with the ability to safely reconfigure boards. With no connector present, the terminals provided good testing pins. A combined total of 95 molex terminals are used within the circuit. Common practice is to use a crimp tool to secure a wire in a crimp terminal. Without access to a crimp tool, wires were soldered into the crimp terminals. Appendix 16 shows the molex wire colour coding system used throughout the circuit. At every input and output there is a 4 way molex terminal. Appendix 17 shows the pin layout for this. This was standard across all the boards ensuring they are all compatible with each other, meaning that the power board can be connected to any other module for testing. There is a test pin installed on the power board to input a signal onto the 4th output pin.

All components were tested to be within their ±5% tolerance before they were placed on the boards. This ensured that all components were of the same value as the designs, a precaution taken to maximise productivity during construction. An exception to this is the potentiometers used; these were rated at ±20%. During the discussion of test results in section 2, component tolerances are assumed to be a contributing factor to minor discrepancies in results.

4.2 Enclosure

The circuit was enclosed in a tin box. All holes were made in the tin with a step drill bit. The only issue came when drilling out the XLR terminal. A step drill bit big enough to cut the 25mm hole required was unavailable. The only alternative was to use a flat wood bit. This made the hole untidy and sharp. Efforts were made to remove all sharp edges with a file.

The conductive strip board tracks were covered with PVC LX tape in order to stop the circuit boards shorting. The internal surface area of the enclosure was covered in duct tape to make sure the copper didn't short to the chassis. The wires connecting the molex terminals were twisted and became rigid, this was enough to hold the boards in place. As the outer tin is conductive, this offers protection from the coupling of interference and noise. The tin makes conductive contact with the DC jack, giving external noise a route to ground, bypassing the signal path.

4.3 Issues During Construction

At the input of the preamp, only half of the bias voltage was received at the non-inverting input of the operational amplifier. After debugging the system no solution was found and in the best interest of completing the artefact, work continued beyond this problem. Figure 21 shows a similar circuit to the input section of the final design. Using a diode model of a transistor, the schematic shows how the bias current flows through the circuit.

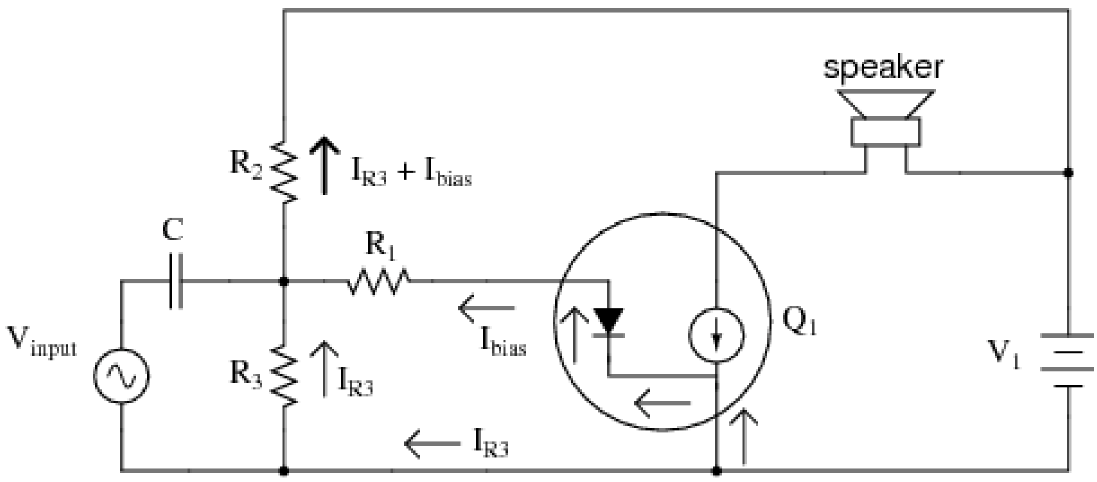


Figure 21: Bias current draw. Kuphaldt, T (n.d.) [online image]

As Tony R. Kuphaldt states with relation to Figure 21:

> "A voltage divider's output depends not only on the size of its constituent resistors, but also on how much current is being divided away from it through a load. The base-emitter PN junction of the transistor is a load that decreases the DC voltage dropped across R_3"
>
> (Kuphaldt, T, n.d)

The DC voltage dropped across R3 would upset the balance of the voltage divider giving a lower than expected bias voltage. This could be the cause of the low input bias at the input of the buffer. This issue would only become a problem if a lower value supply rail were used. At the input, the bias voltage is around half of its intended value. A 5v supply rail would only produce 1.25v bias. A strong bass signal would clip the supply rail. When working with a 9v supply the bias voltage is 2.25v, which is more than enough to pass a clean signal through the buffer stage. If it wasn't for the worst case scenario planning, this may have been an unavoidable problem.

The pad, power and filter modules utilise a double pole double throw (DPDT) switch to change between circuits. When operated there is a voltage spike at the output, which creates a popping sound. Jack Orman offers and explanation for this problem.

> "Capacitors will leak a tiny voltage which will build up on the input (or output) as a small voltage potential. When the pedal is switched on, the voltage is discharged into the signal path."
>
> (Orman, J, 2008)

By adding pull down resistors at various positions around the DPDT switches, this rogue DC voltage could be dropped from the signal. At the filter stage, bias voltage is present through the DPDT switch, meaning a pull down resistor would lower the bias and make the filter amplifier un-operational. By moving the DPDT switch to after the mixing resistors, the bias could be isolated and the switch would not suffer from this issue. The initial response to this problem was to use capacitors to block DC from the switch. This unfortunately led to some undesirable filtering of the audio signal. The late consideration of this problem left no opportunity to implement a permanent solution.

When building on strip board, shorting connections occurred. Continuity testing with a multimeter helped pinpoint the problem but finding the exact point of contact proved difficult. On several occasions, complete boards were rebuilt to resolve a shorting issue. This was a time consuming exercise, lacking in efficiency. A second issue with strip board came with the wearing down of the copper tracks. By the later stages of the artefact, strip board that had been used in the initial stages started to oxidise. Tinning wash or varnish was not used to protect the copper tracks as all of the boards were under development right until the end of the build. The solder joints were not defluxed for this same reason.

Decoupling capacitors throughout the circuit formed high pass filters in conjunction with resistive loads. During the testing of the circuit, it was immediately noticeable that the cut off frequency of these high pass filters was too high (around 400Hz). The 0.68uF decoupling capacitor after the second gain stage was the main issue, it was brought up to 6.8uF. This helped bring the practical cut-off back down to the sub frequencies. Work continued with other decoupling capacitors until the values were raised enough to lower the cut off to an acceptable level.

Overall, a lot of lessons were learnt during the production phase of this project. Looking forward to building the next prototype, more time will be allocated to production. The issues I have experienced will be taken into consideration and improved upon.

4.4 Final Schematic

The final schematic shows significant differences from the initial schematic proposed in the literature review. Throughout the construction of the artefact, every opportunity to improve the design was taken. Appendix 18 shows the initial schematic and Appendix 19 shows the final schematic.

Instead of two non-inverting gain stages, two inverting gain stages were used. This was implemented because the non-inverting gain stages by nature, reference ground at the inverting input of operational amplifier. This posed difficulties as the circuit is biased at half the supply voltage.

The input pad was redesigned because the -20dB bass dropped bias voltage as well as AC signal voltage. The bias drop became problematic at the first gain stage, the bias was too close to ground for the amplified to avoid. The signal was decoupled before the switch and then reapplied after. This made sure that the resistive network was only applied to the AC signal.

The resistive networks at the line output stage were reconfigured to decrease the loading effect of a long XLR cable. Capacitance present in a long cable can alter phase shift of higher frequencies around the feedback loop. This can cause oscillations due to positive feedback at the inverting input. This consideration was made after the initial design of the instrument stage, and so the larger resistive network was not adapted in the same way.

CHAPTER 5

5.1 TEST NOTES

Throughout the testing an input sine wave at 1.2Vpp was used, this is the worst case input signal as stated in the literature review. This level proved to represent an extreme worst-case peak and should be taken into consideration in analysis of results.

The vast majority of testing was done using Metric Halo's Spectrafoo software and Mobile IO 2882 hardware. In an ideal world, tests would be carried out on specialist hardware such as a prism dScope. The Mobile IO 2882 was running at 96kHz, its maximum sampling frequency. In order to avoid aliasing, there will be a low pass filter before the analogue to digital converter (ADC). According to the Nyquist theory (Appendix 31) the filter cut-off will be at half the sampling frequency (48kHz). If the circuit is producing high frequency oscillations or is inducting high frequency noise, they would not be present in the results.

Spectrafoo uses Fast Fourier Transform (FFT) algorithms to convert the digital data from the time domain and into the frequency domain. As stated in the Spectrafoo manual:

"The spectrum analysis within Spectrafoo utilizes a multichannel, high resolution, real–time FFT engine. At its highest resolution setting, Spectrafoo utilizes 64k point FFTs, allowing you to see features with widths as small as 2/3 Hz."

(Metric Halo, n.d)

Realistically, the resolution of the final graphs look worse than suggested in the manual. At lower frequencies it would seem that Spectrafoo has only been able to estimate the frequency response, later smoothing it with an averaging algorithm.

As suggested in the mobile IO 2882 Users Guide, a pseudo cable was made to utilise the differential line input stage of the unit. Figure 22 shows how this cable was wired.

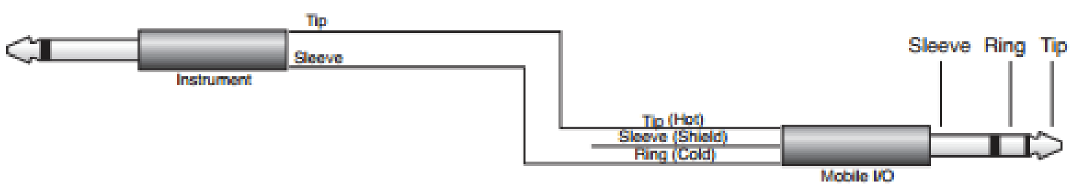


Figure 22: Pseudo jack wiring

[Image] In: Metric - Halo. (2012)

Hugh Robjohns comments:

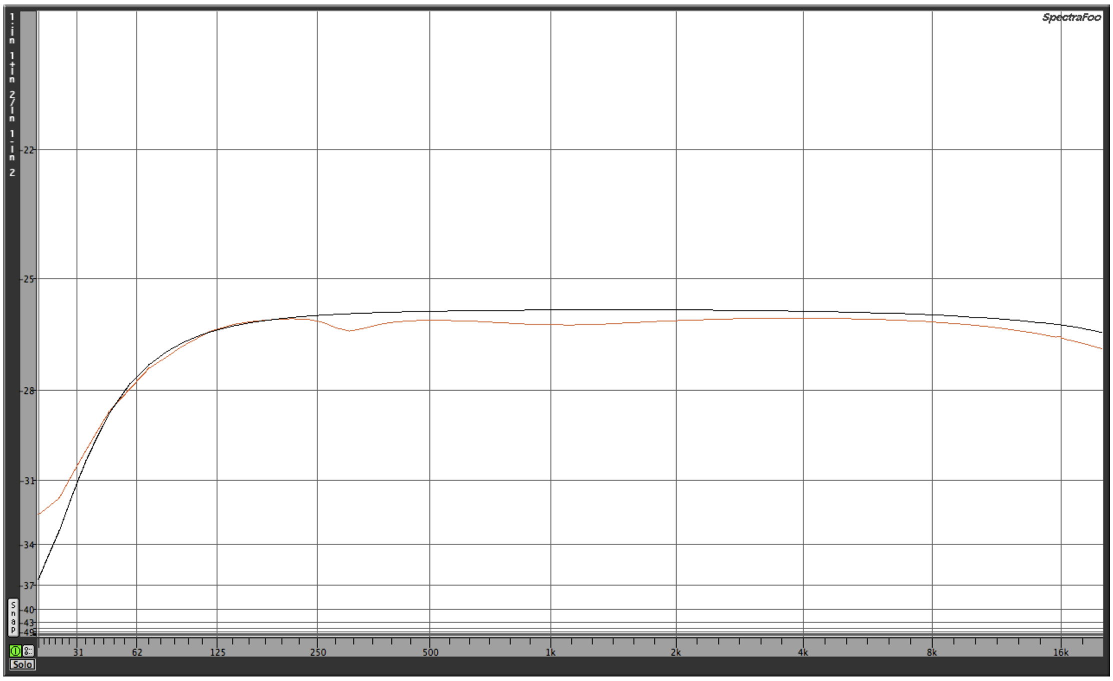
"The Impedance to ground of the two signals is very different. The 'Hot' wire will have an impedance of a few tens of Ohms, while the 'cold' wire has an impedance of zero Ohms. So any interference that gets into the cable won't be a true common-mode signal."

Robjohns, H (2008)

As there is electrical isolation between source and load and full shielding of the signal, the pseudo jack cable ensured the signal coming into the 2882 was as noise free as possible. The 2882 User's guide also states its input as having 110dB signal to noise ratio. Specifications like this made the 2882 as the most ideal hardware testing solution. The Metric Halo IO 2882 specifications can be found in Appendix 32.

5.2 Frequency Response

Appendix 20 shows the test procedure for the frequency response tests. Figure 23 shows the frequency response of the entire circuit.



Black: No EQ Connected

Red: EQ measured flat

Figure 23: Frequency Response

Without the filters connected we see a steeper roll off around 50Hz. This could be the unintentional loading effect the filters have on the circuit. If we use figure 24 as an example, we see a high pass filter formed by a resistor and a capacitor.

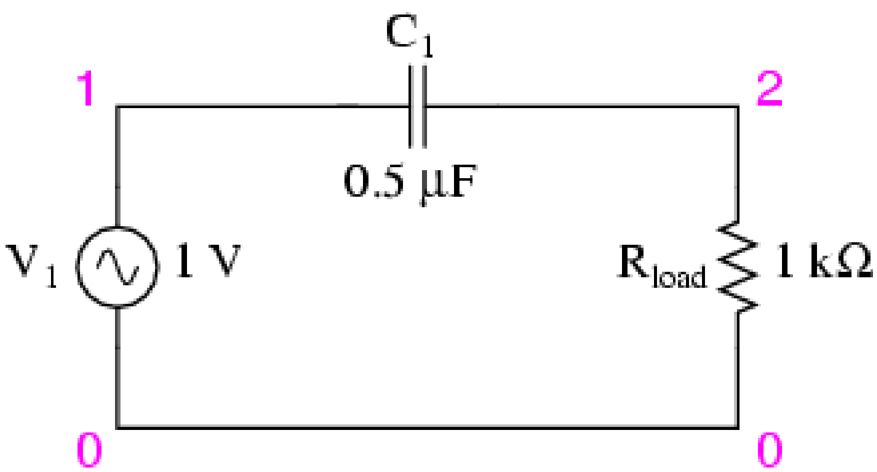


Figure 24: Passive High Pass Filter. Kuphaldt, T (2007). [online image]

If Rload is a representation for the total impedance of the filters we can see how the filters would affect the frequency cut off. By increasing R (adding filters), the frequency cut off is decreased, Figure 25 proves this:

$$f_{cutoff} = \frac{1}{2\pi RC}$$

Figure 25: Fc calculation

All high pass filters formed by the resistor/capacitor networks were designed to have a maximum cut off of 40Hz. practically the cut off seems higher, lying at around 100Hz. As mentioned in Chapter 1, the initial cut-off has been improved significantly since the first build.

The frequency response graph shows that with the EQ connected, there are some slight contours in the shape of the response. This is most notable in the region of 250Hz were there is a cut in frequency. Despite the pot being tested to be in its middle position, the pot would still appear to be in a counter clock wise "cut" position. This could also be the cause of a slight dip in the 1kHz region. Using pots with centre indents would improve the test procedure for a flat EQ as they are much more accurate in centre position. It is also worth considering the ±20% tolerance rating on the potentiometer; this could bias the centre point towards the cut position. Apart from this, we see slight attenuation when the filters are connected. The preamp shows good frequency performance up to 20kHz. At no point in the direct signal path (the circuit excluding filters) is a low pass filter formed. This makes the high-end frequency response of the circuit extremely high. The operation amplifiers (op amps) in the circuit are the only high frequency limiting components in the design.

Figure 26 shows the large-signal differential voltage amplification (Avd)/phase shift vs. Frequency plot from the Texas Instruments TL07X data sheet.

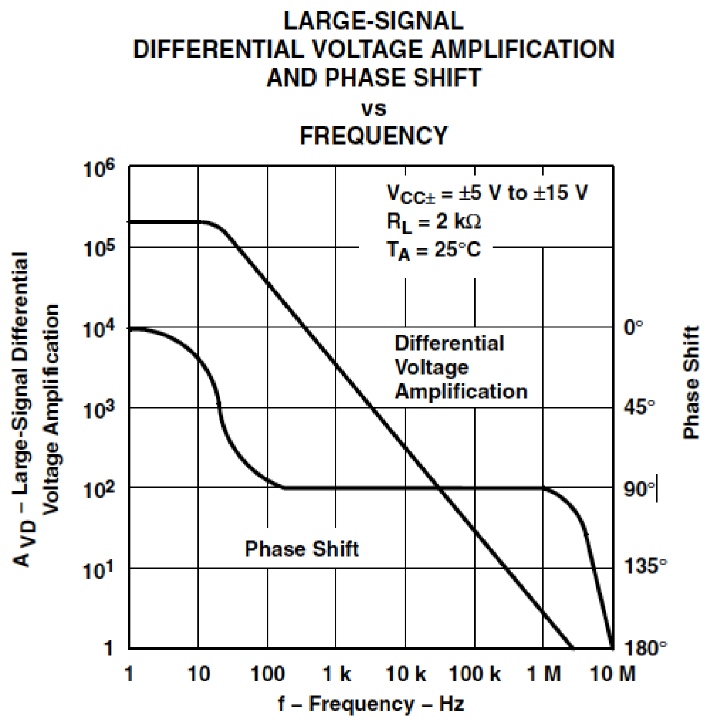


Figure 26: Avd Vs. Phase Shift of TL07X

Texas Instruments, (1978).

Bruce Carter describes the Avd as:

> "Large signal differential voltage amplification, AVD, is similar to the open loop gain of the amplifier except open loop is usually measured without any load."

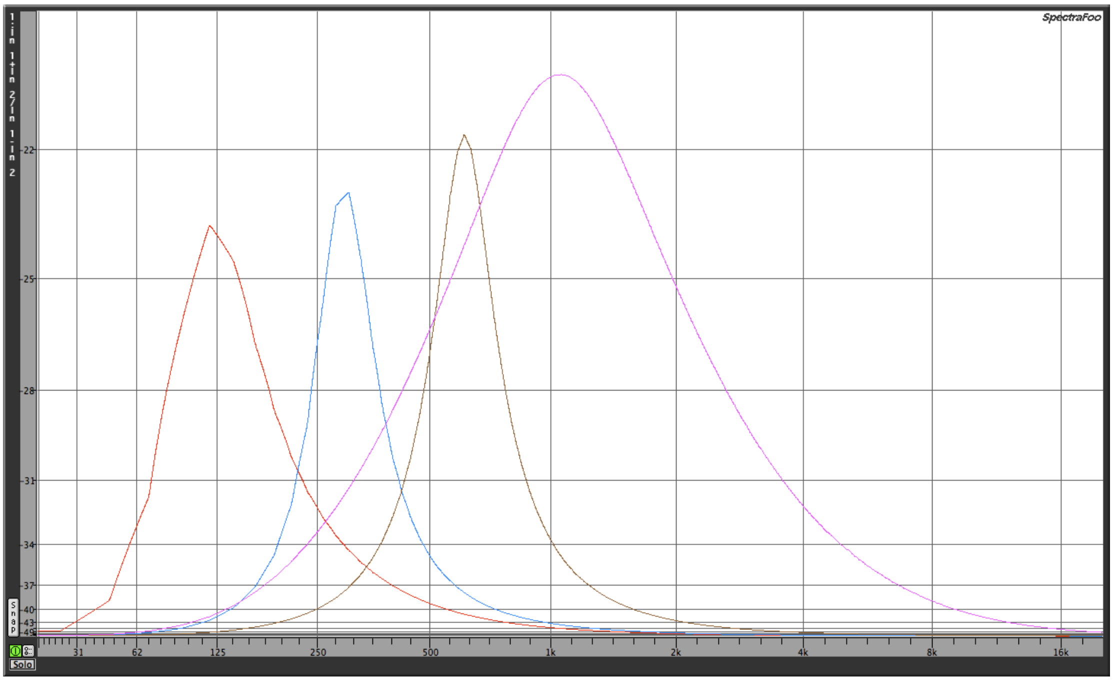
Carter, B (n.d)

At maximum drive, the highest gain value of an amplifier is 2 (6dB). This correlates to a 3MHz cut off frequency. This leaves the circuit susceptible to interference from radio and telecommunication signals. During the testing period of the artefact, this never posed a problem and the design was not updated to include a filter to suppress this issue.

5.3 Filter Performance

5.3.1 Bandpass Frequency Response

Each individual band pass filter circuit was tested. Figure 27 shows the 4 results of this test.



Red: 100Hz

Blue: 250Hz

Brown: 600Hz

Magenta: 1000Hz

Figure 27: Bandpass filter responses

The filters were all designed to be at unity gain at their centre frequency (Fc). The graphs show that there is variation (above and below unity) between the four filters. In the literature review, the multi-feedback band pass filter component values were calculated. It is seen here that significant rounding of ideal values were needed to meet standard resistor and capacitor values. Furthermore, some of these values were not available at the time of construction so further compromising was needed. Combined with component tolerances, this offers reasonable explanation to this. As the passive band pass networks were built on a separate board, it was possible to swap out different filters for testing. After several tests with various different filters, the original design was used as it provided the most stability and performance.

5.3.2 Summed Filter Response

Figure 28 - 13 show the frequency response graphs of the filters summed to the original signal.

Black: No EQ Connected

Blue: 12dB Boost

Red: 12dB Cut

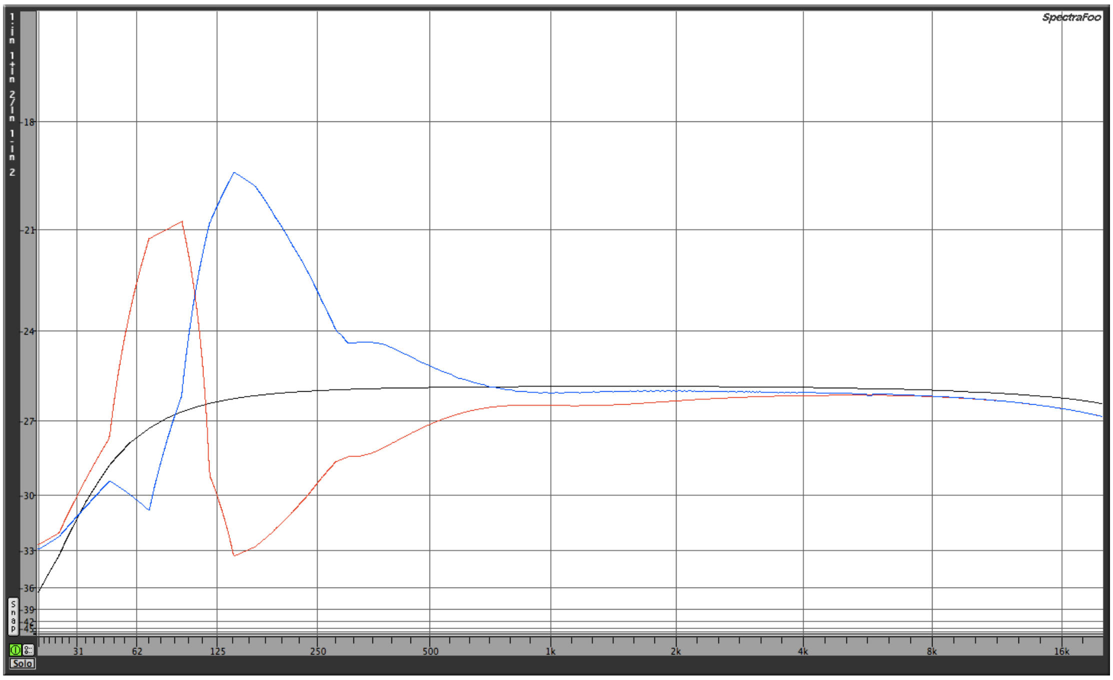


Figure 28: 100Hz

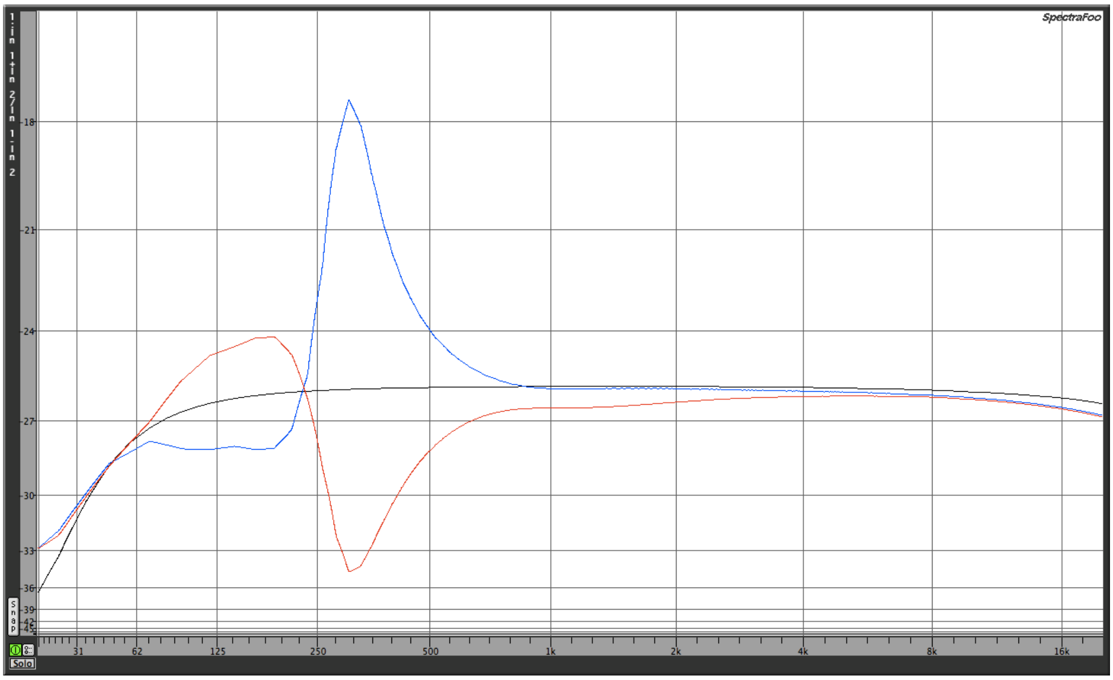


Figure 29: 250Hz

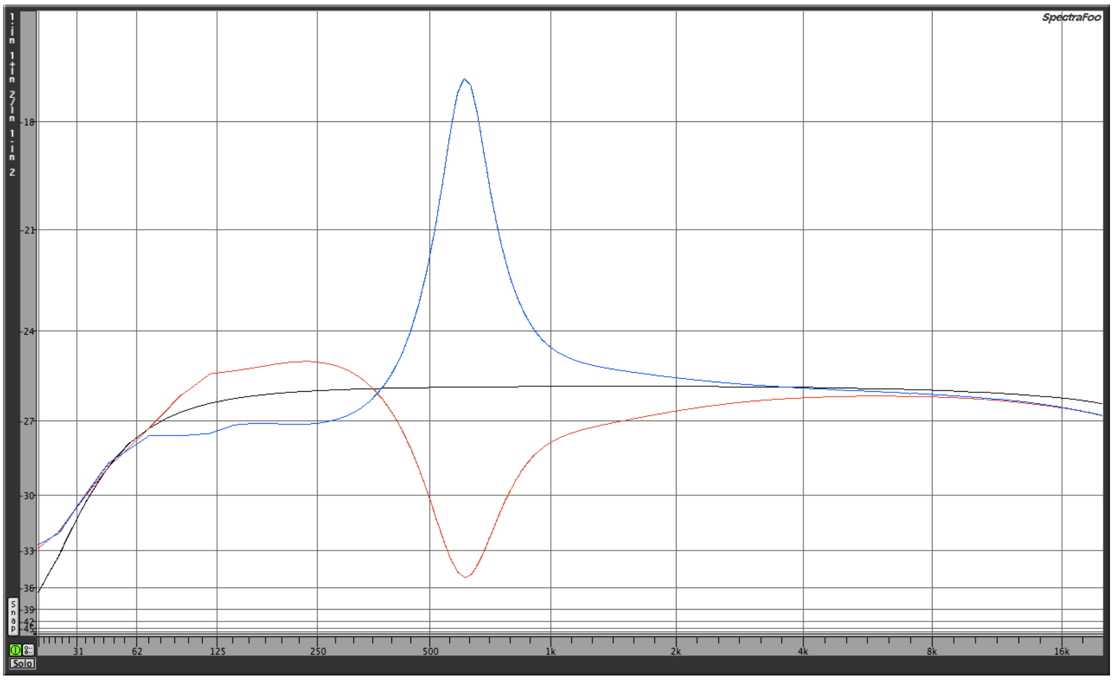


Figure 30: 600Hz

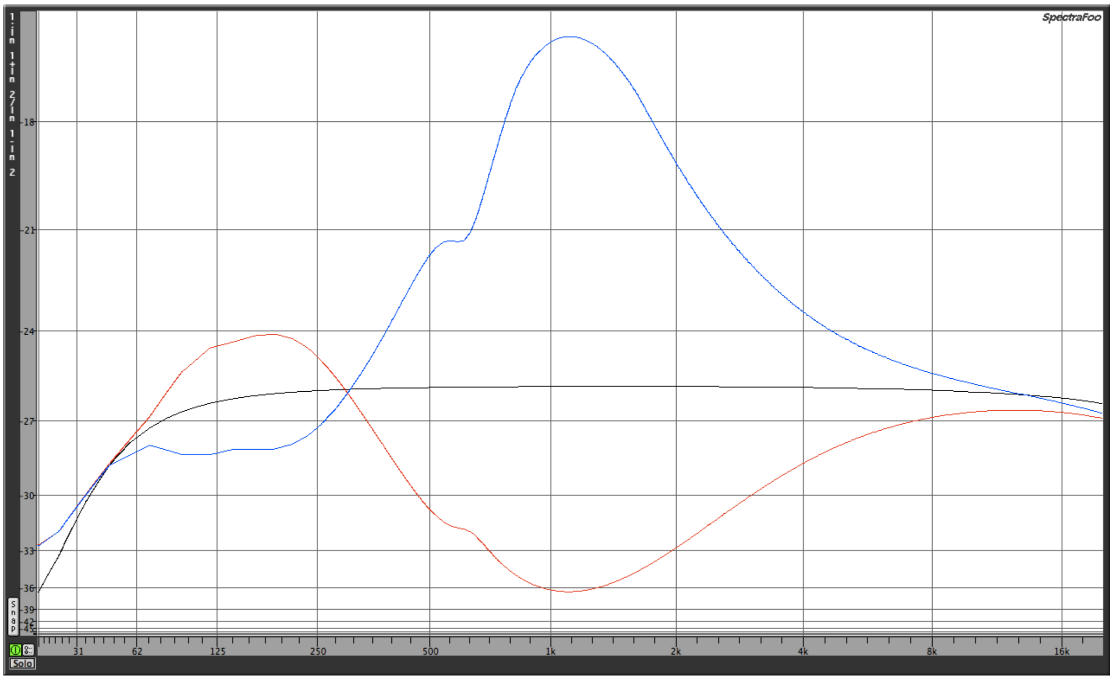


Figure 31: 1kHz

A common issue with all the filters is an issue at the unity gain point below the Fc. When the filter is boosted the unity gain point cuts and vice versa. It's a property that is most evident on the 100Hz filter were the situation becomes serious. Figure 32 is the phase response of the bandpass filters. Appendix 21 shows how this graph was produced.

Red: 100Hz

Blue: 250Hz

Brown: 600Hz

Magenta: 1kHz

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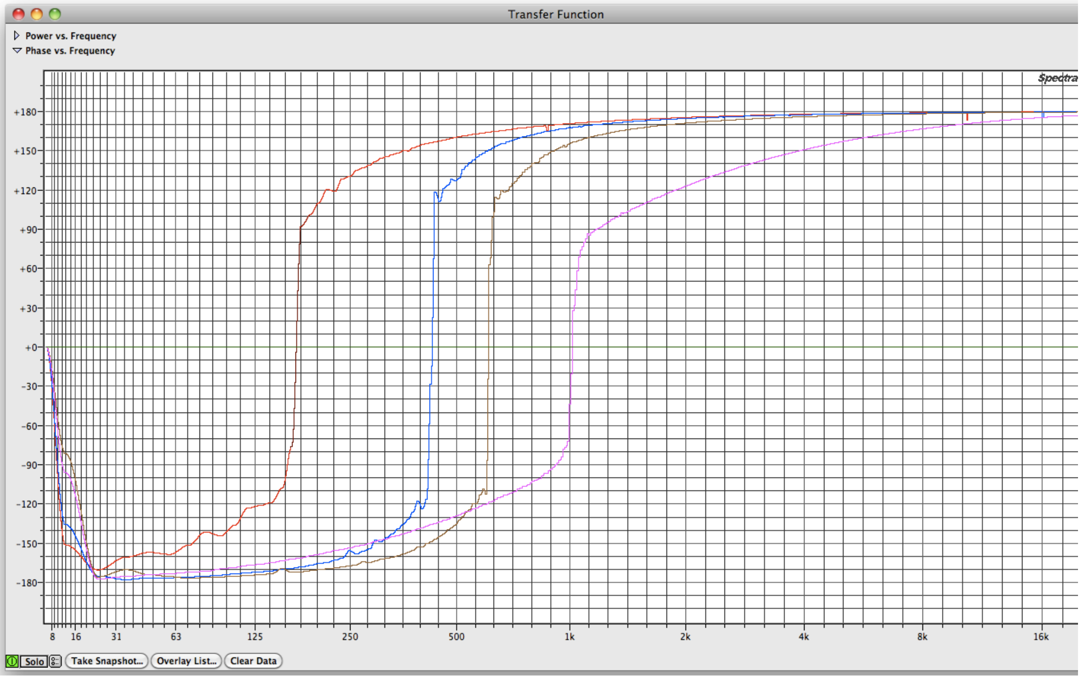

Figure 32: Bandpass Phase Response

Looking at this graph alongside the phase shift vs. frequency graph of the TL07X (Figure 33) shows the potential source of the problem. At these low frequencies, both signals are shifting in phase over the 0-100Hz frequency band. As these two phase shifts are occurring in the closed loop of the op amp, this could lead to the instability within this frequency band.

A potential improvement could be the use of a Sallen and Key band pass topology instead of a multi-feedback band pass. Figure 34 shows the schematic layout of a Sallen & Key band pass filter.

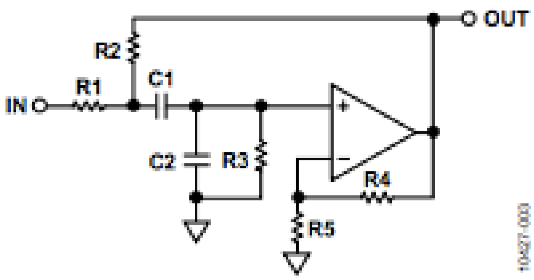

Figure 34: Sallen & Key Bandpass Filter

Zumbahlen, H (2008)

The Sallen and Key topology does not use filter circuitry in conjunction with negative feedback. Keeping the filter and amplification circuitry separate entities means that the filter performance is not limited by the amplifier performance.

The 250 and 600Hz were identified in research as problem frequencies when refining a bass guitar sound. They have a higher Q that was designed for cutting “boxy” and “Nasal” frequencies around the 250-600Hz regions. The bass tone is made smoother and stronger when these frequencies are cut. When boosted, the 250Hz filter can offer a degree of “growl”. The 1kHz filter is the strongest and most versatile filter. This is clearly evident in the frequency response graphs and the sonic output of the circuit. When boosted the sound is opened up with a considerable amount of string and fret noise amplified, a good characteristic of a rock bass sound. When cut, a vintage tone becomes available. With significant attenuation in the high frequencies, the sound becomes well rounded, deeper with a tone that would be found in soul, funk and blues music.

5.3.3 Interaction

The frequency response of the summed signal was tested with all four filters connected in parallel and with unused filters disconnected. This was done to observe the interaction between filters. Figures 35-42 show these graphs.

Black: No EQ Connected

Red: Connected in Parallel

Blue: Other filters Disconnected

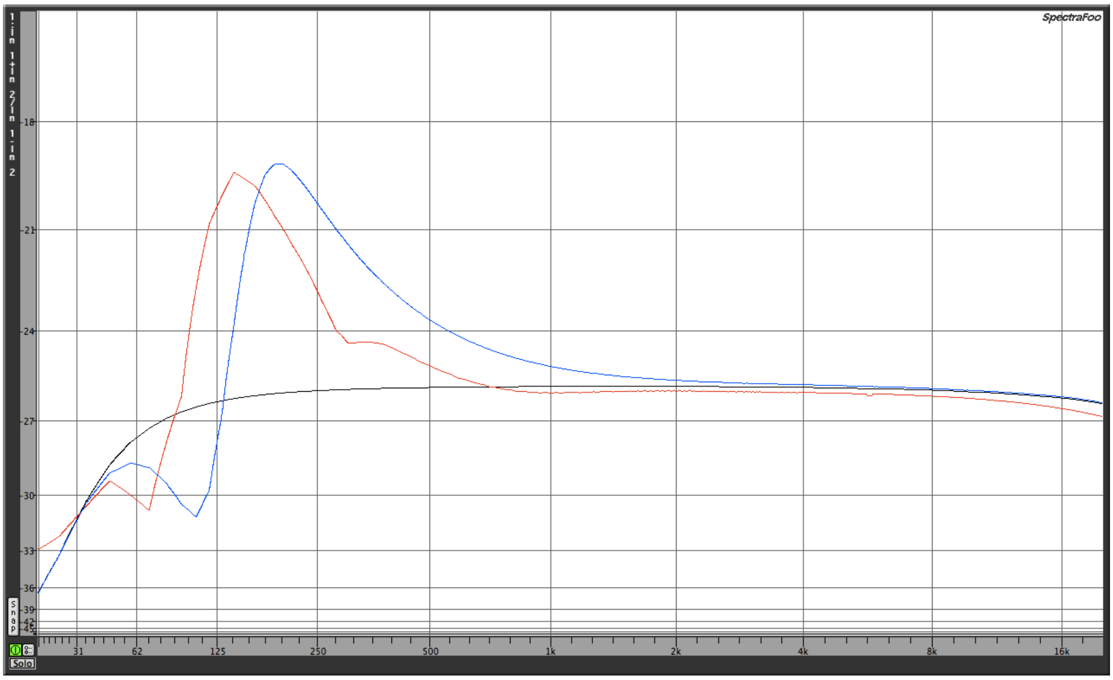

Figure 35: 100Hz Boost

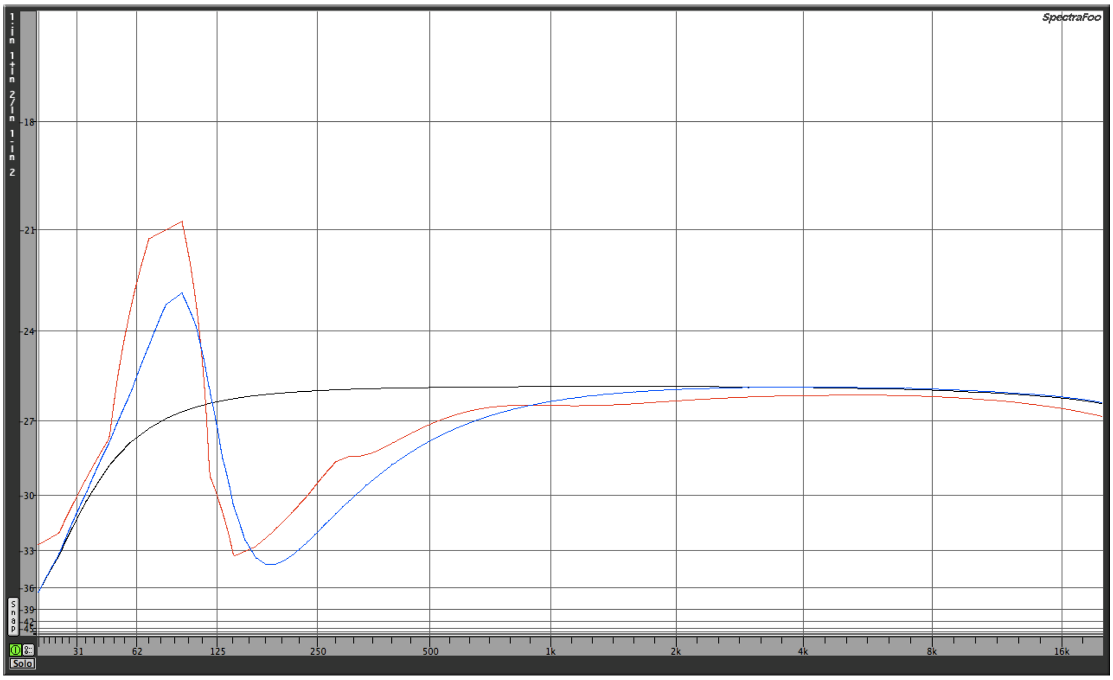

Figure 36: 100Hz Cut:

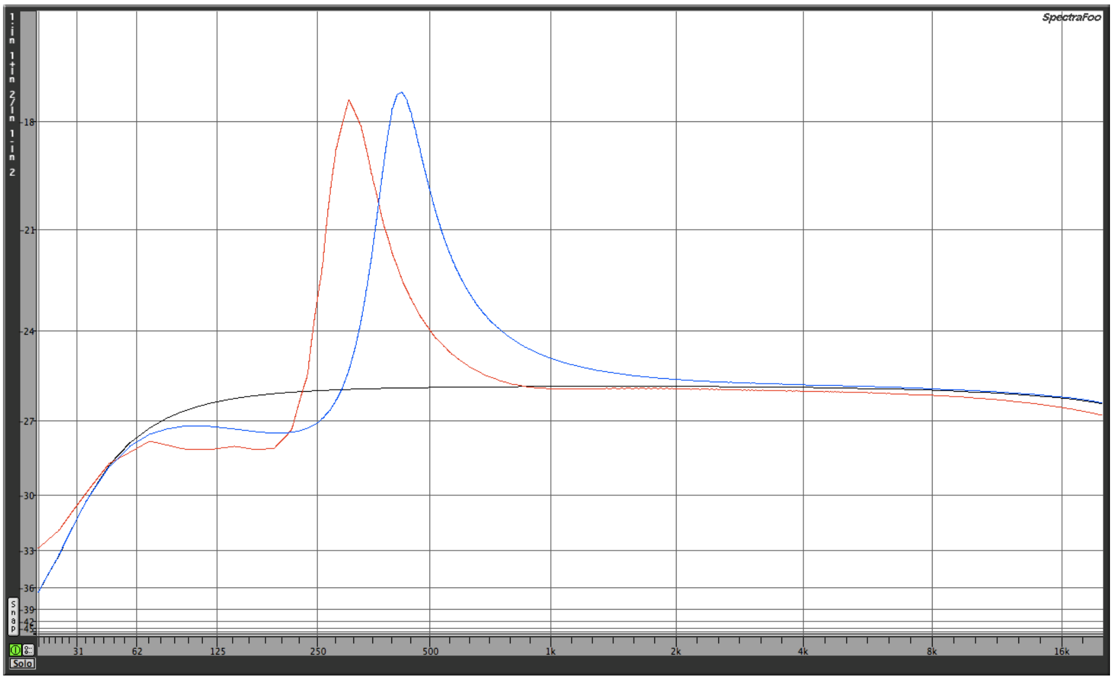

Figure 37: 250Hz Boost

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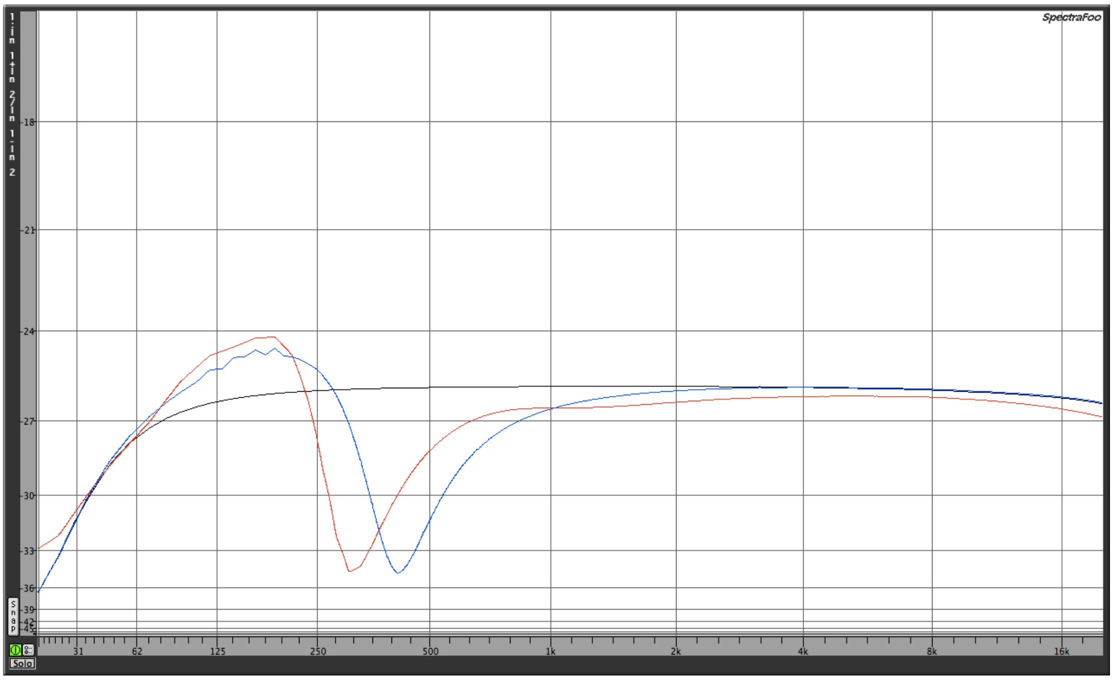

Figure 38: 250Hz Cut

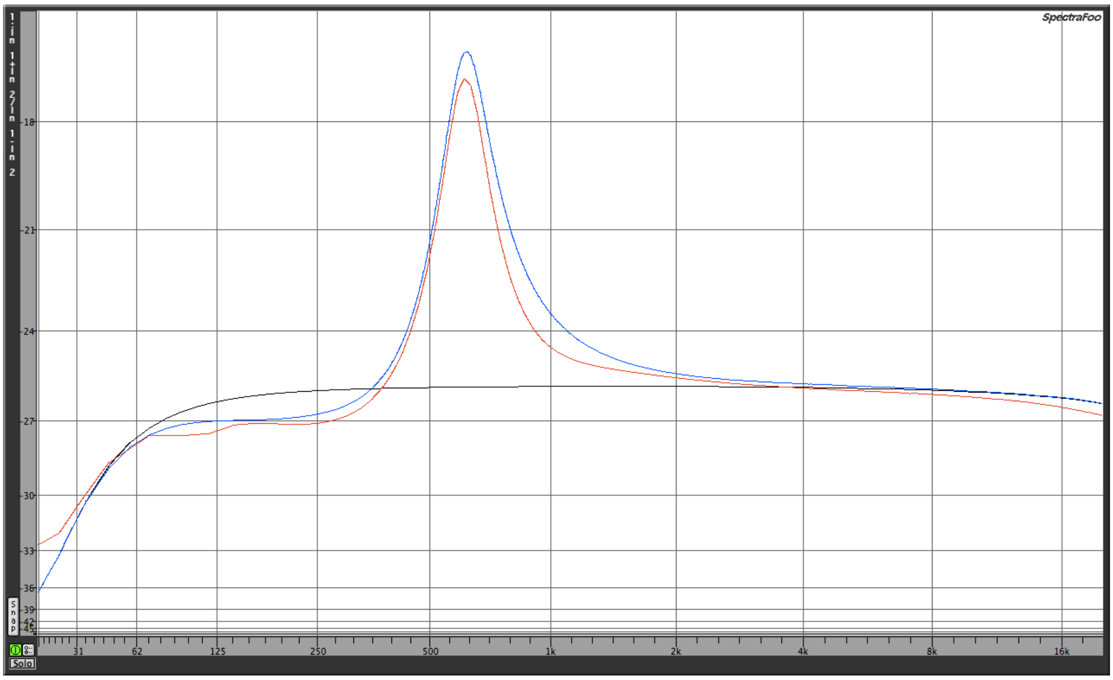

Figure 39: 600Hz Boost

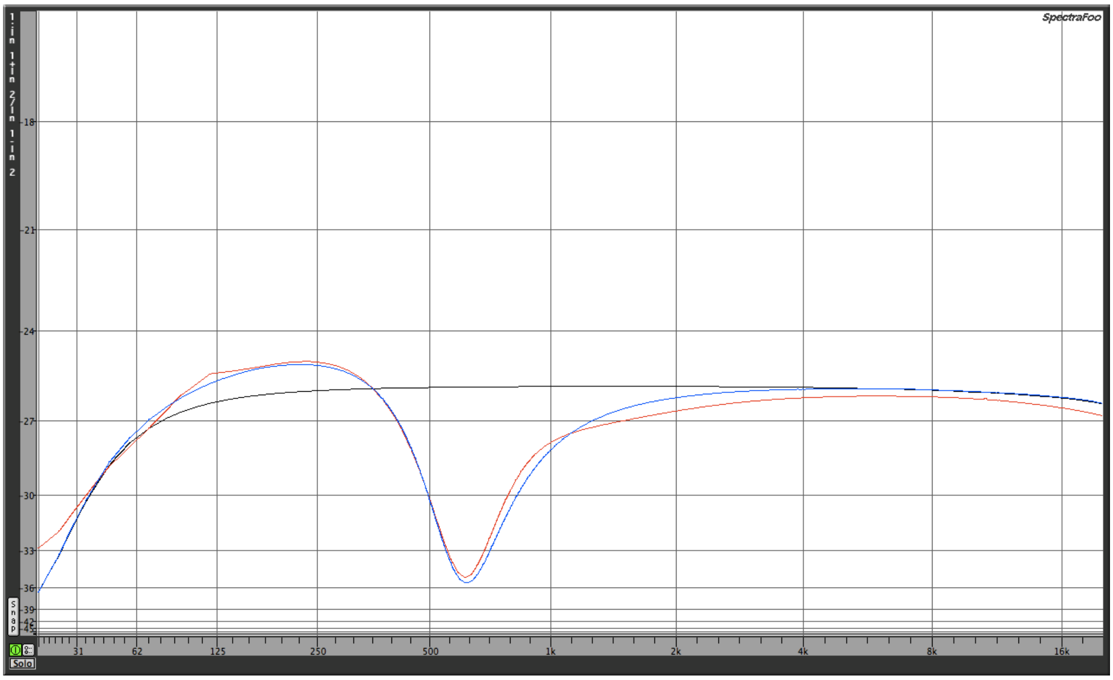

Figure 40: 600Hz Cut

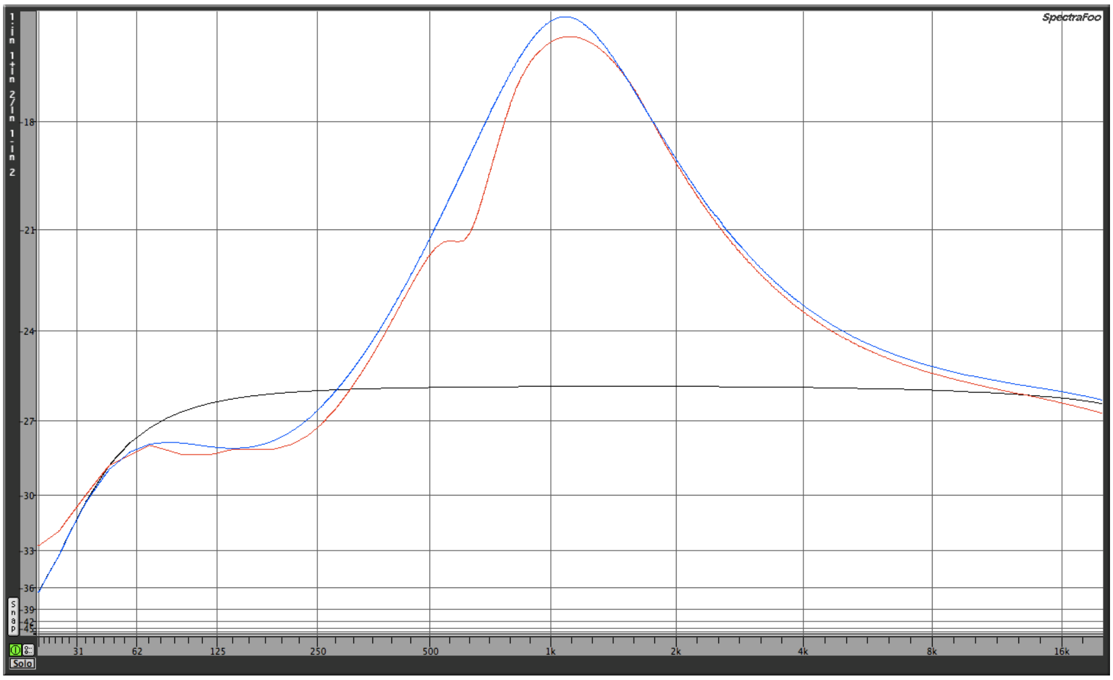

Figure 41: 1kHz Boost

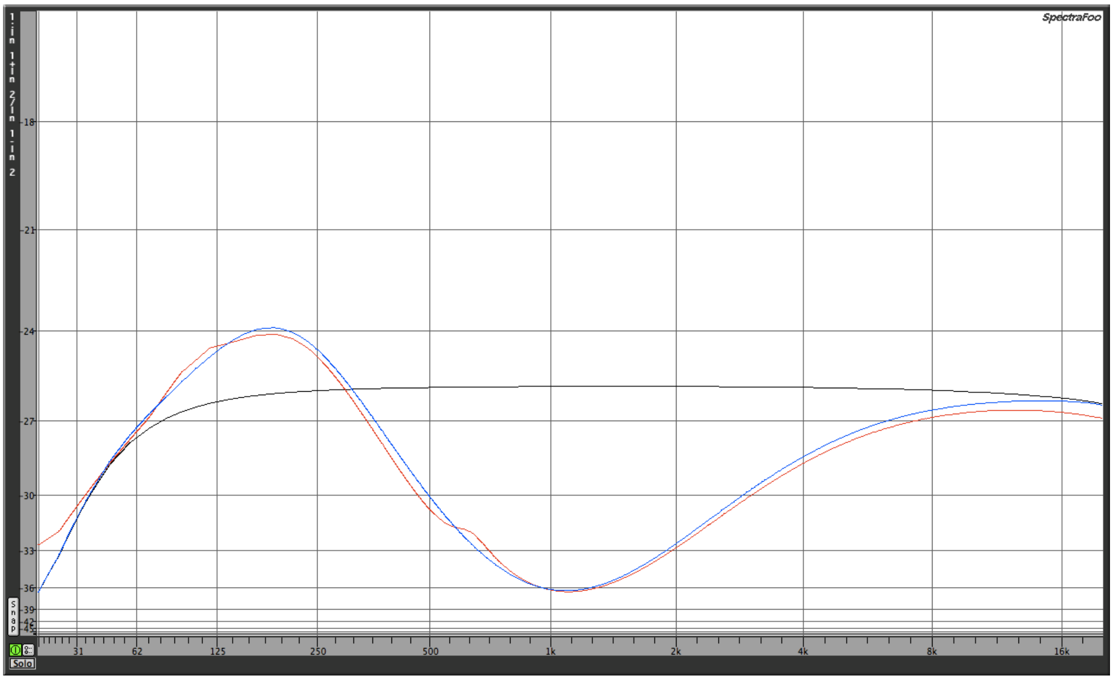

Figure 42: 1kHz Cut

When connected in parallel, the 100Hz and 250Hz filter show a lower Fc. This suggests that the filters are loading each other, lowering Fc. The unconnected filters tend to perform better than the connected ones; the known issues are still present but look less extreme and smoother. This may offer another partial explanation for poor filter performance when summed back to the original signal. John Bohn offers a potential solution to increase summing performance.

"A technique used almost universally to improve the combined responses of graphic equalizers involves two series summing circuits. This way two adjacent bands are not added together by the same summer"

Bohn, J (1986)

This creates physical separation of filter modules and reduces the chance of interaction between them. Figure 43 is a graphical representation of this.

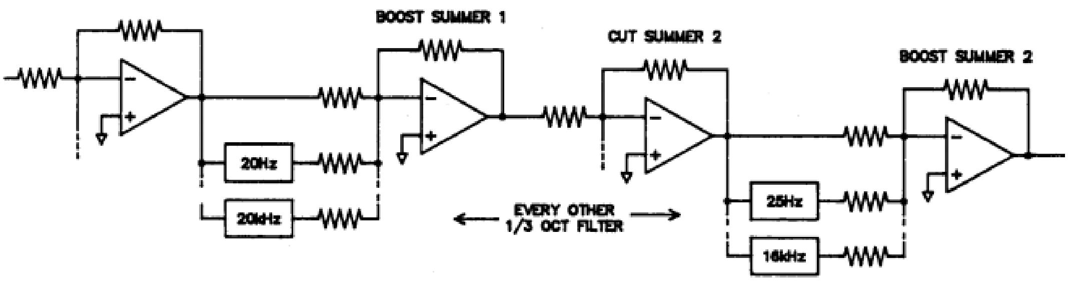

Figure 43: Dual Summing Filters

Bohn, J (1986)

5.4 Total Harmonic Distortion + Noise

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Total Harmonic Distortion + Noise (THD+N) measures harmonically related content and other noise at the output, that isn't present at the input. As Dennis Bohn from Rane corporation states:

"Distortion analysers make this measurement by removing the fundamental (using a deep and narrow notch filter) and measuring what's left using a bandwidth filter. The remainder contains harmonics as well as random noise and other artefacts."

Bohn, J (2000)

Without the other artefacts, a distortion reading may not give a true representation of artefacts inducted into the signal through the circuit. Appendix 22 shows the test set up for measuring THD+N. The 0.005% THD+N reading shows good performance in relation to standard THD specifications for professional audio equipment. The preamp was tested at unity; this significantly improves the THD+N reading. If the preamp were pushed towards saturation, a lot more harmonically related content would be present at the output.

5.5 Signal to Noise Ratio

The signal to noise ratio gives the ratio between the noise floor of the circuit and a reference level. The reference level used during this test is 1.2Vpp (Volts peak to peak). A 20-20kHz band pass filter was added before an oscilloscope. This was to remove any signal that would be measured unnecessarily. Appendix 23 shows the signal to noise ratio testing procedure and calculations.

The noise worked out at 59.233dB. This is a fair way from professional audio standards. The reading was taken before the circuit was fitted into its enclosure, this could offer increased signal to noise performance by reducing interference. A significant use of molex terminals is another contributing factor to the noise in the circuit. As will be discussed in section 3.1, a printed circuit board would eliminate the large volume of molex connections. Although this measurement could be improved, Berry Papin offers a more realistic and practical incite into how this reading could be perceived when he comments:

"Many people don't realize how much a 30 or 40 dB ratio is. A piece of electronics equipment with a S/N ratio of 80dB may be good enough for all but the best systems."

Papin, B (n.d)

This opinion puts a better perspective on the reading. Although an extra 20dB would bring the signal to noise reading to professional standards, a reading of 60dB of is acceptable.

5.6 Impedance

The testing procedure for input and output impedance can be found in Appendix 24 and 25. The measured 1.35M input impedance is higher than the designed value. This has no effect on the operation of the input stage. As long as the input impedance is ten times as large as the source impedance, the circuits will be bridged (maximum voltage transfer). The output impedance of a bass guitar varies depending on pickups, potentiometers and internal preamps. It's a safe assumption that the output impedance of a bass guitar isn't higher than 10kΩ.

With two outputs comes the need for two separate output impedances. Instrument and line inputs commonly have input impedances of around 1M. Because the instrument output requires more attenuation, a larger output impedance had to be factored into the design. As long as the output impedance does not exceed around 100K it should have no effect on output bridging. 13.605k and 2127 Ω output impedances are ideal and found amongst the industry standards.

5.7 Current Draw

Appendix 27 shows the technique for measuring the current draw. With the preamp in it's quiescent state, the circuit draws 32.1mA from the power supply. With the preamp fully amplifying a 1.2Vpp signal, the circuit draws 32.9mA from the power supply. Appendix 28 is an excerpt of the TL07X data sheet; it states the maximum current draw from the amplifier is 2.5mA. 13 amplifiers are used; this gives a theoretical quiescent maximum state current draw of 32.5mA.

Based on a 550 mAh 9v PP3 battery the following lifespans have been calculated. With the preamp in its quiescent state, the battery will last 17.1339 hours and running at full capacity at 16.71713 hours. These are acceptable values but by no means are they exceptional. For example, the Sansamp Bass Driver consumes only 6mA. This is incredibly low and would result in 91.6 hours battery life.

All 13 op amps in the circuit are biased at half the supply rail, this keeps the transistors inside the op amps turned on all the time (Class A). This is inefficient as power is dissipated as heat energy. Using the more efficient class B or AB push pull system may improve the expected battery life.

5.8 Output Level

The following subsections discuss specifics regarding the maximum output capabilities (including 12dB boost at 1kHz) of the circuit. Everything below the maximum value is controllable with via the gain stage potentiometers. Appendices 29 and 30 show all output readings and calculations.

5.8.1 Line Level

As the preamp can operate up to an 18v supply, the output gain can be pushed as high as 14.2Vpp. This maximum dBu reading of +16.2275 shows a safe signal level for professional audio equipment to accept. Most professional audio systems have a maximum capacity of +24dBu at the input. This puts the maximum output of the bass 12dBu above the +4dBu line level standard and 8dBu below a level that would be considered unworkable. At a normal operating level the preamp outputs much closer to line level.

During testing, the preamp was pushed to use the entirety of its headroom powered by 9v supply. The maximum output of this test gives +8.3895dBu, a closer reading to the intended +4dBu line level output. Maximum gain before clipping of 13.6248dB is acceptable and only little over the average of many commercial products.

5.8.2 Instrument Level

The instrument level is the normal line level signal through an increased pad. The pad lowers the output to a lower level expected by bass guitar amplifiers. The line level output can be used if the input stage of the amplifier has a pad. With an 18v supply the output can be as high as +11.5004dBu and at 9v as high as +3.70dBu. Both levels are high for an expected instrument signal (-5.2346dBu).

CHAPTER 6

6.1 FURTHER WORK

In order to improve noise performance, a review of the use of molex connectors is necessary. Once all improvements to the circuit were made, having a PCB etched would make the entire circuit more stable. The circuit wouldn't necessarily need to be placed on one board; a motherboard with all processing could be connected to a separate board for controls. This is often seen in audio design and an idea that could easily be adopted in the preamplifier design.

A simple phase lead capacitor in parallel with the gain stages could be installed. Paying more attention to high frequencies outside of the range of human hearing would bring the circuit more in line with standard practice within professional audio design. These capacitors are unlikely to have any impact on the tone of the preamp, as they will produce cut off points above 20kHz. It would however give the op amp more protection against noise, interference and op amp instability. Figure 44 shows C and R2 forming a low pass filter around a negative feedback network.

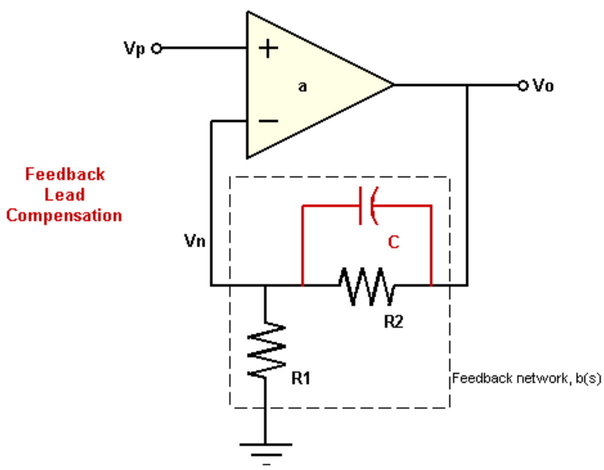

Figure 44: Phase Lead Capacitor (C)

Math Works (n.d). [online image]

Apart from the obvious improvements needed on the 100Hz filter, a redesign could improve the filters dramatically. For example, moving the 100Hz filter down towards the sub regions of the audio spectrum would help give the preamp more weight and much more low end versatility. The 250 and 600Hz filters perform very well when cut but when boosted, they can sound unnatural. To improve this, a dual gang pot could be used. That way, different parameters could be used to sum between the boost and cut summers. A lower Q could be achieved when boosted, without affecting a high performance of its cutting function.

The preamp has a transparent effect on the overall tone of the bass signal. With experimentation, a more in depth analysis can be made of a characteristic bass tone. These studies could become the basis of fixed filters within the signal path, adding a characterizing tone before the filters are tweaked. Alternatively they could be wired into the summing amplifiers and faded in at a given level. This would give the user a master control for a hard-wired tone made up of several filters. Furthermore, due to the filter topology used, both sides of a linear pot could be used to control a different set of filter parameters.

For example, a set of filters that would give a vintage tone could be wired to the "cut" half of a potentiometer. A modern set of filters could be wired to the "boost" half of a pot.

The preamp was originally designed to operate under guitar pedal parameters, including designs for running off a 9v battery. The process of splitting the supply and biasing the circuit puts a lot of limitation on redesigns. Whether adopting virtual ground integrated circuits or transformers from mains voltage, forming dual supply lines would remove any biasing and decoupling issues in the current design. This would change the preamp from class A to class B but wouldn't necessarily have any detrimental effect on the sonic performance of the preamp. A class B push-pull system may compromise the good THD+N performance due to cross over distortion but given the frequency content of the bass signal, it is unlikely to be audible in the output. The dual supply would allow more efficient running of the op amps, using all of the available transistors inside the amplifier. This more efficient use of power could improve battery life, a specification that is poor in the current design. There is also an option of combining the above solutions to create a class AB system; this is often the preferred method in audio amplifier design. Adding small bias current to the input of a transistor will move the bias point above 0 so crossover distortion is greatly attenuated. The transistors still share the work so energy isn't dissipated as heat.

The power consumption of the circuit needs to be further analyzed. The circuit performs fairly poorly in the respect. Given the timescale of the project it was always likely that the circuit wouldn't be as efficient as desirable.

It of my personal desire to include reamp functionality into the input of the preamp. This would lead to supreme versatility within the studio environment. The inclusion of insert sends and returns is also a likely inclusion in further designs. This would be beneficial for using third party compression units (or other processing).

6.2 Conclusions

Overall the preamp is versatile and stable across all available parameters, offering acceptable sonic output. The circuit is capable of accepting any instrument signal and outputting to any line or instrument level device. The results of electrical testing are representative to the sonic output of the preamp.

The project has been a success on an academic and personal level. Further practical work and redesigns have continued during the writing of this report and will continue into the future.

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Appendices

1: Electronic Circuit Equations

1.1: Ohms Law

Ohm's law states that voltage is proportional to the current through a load.

$$\boldsymbol{V = IR}$$

Figure X: Ohms Law

Where:

$V = Voltage$

$I = Current$

$R = Resistance$

1.2: Kirchoff's Current Law

Kirchoff's current law states that the amount of current flowing into a node (physical connection) has to flow out of that same node. Using Figure X we can state:

$$I_1 + I_2 = I_3 + I_4 + I_5$$

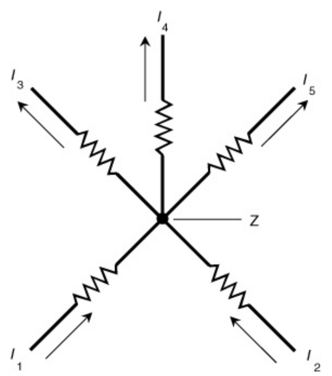

Figure X: Kirchoff's Current Law

Education.com (n.d.) [online image]

1.3: Kirchoff's Voltage Law

Kirchoff's Voltage Law stages that the sum of voltages within a closed loop circuit, must equal the voltage source.

$$V_{source} = V_1 + V_2 \ldots\ldots + V_n$$

1.4: Thérevin's equivalent circuit

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Any network of resistors and voltage sources can be made equivalent to 1 voltage source and 1 resistor. This theorem is used heavy to simplify complicated resistive networks. It is applied heavily in modular designs for distinguishing the ability of different sections to work together in a circuit.

1.5: Series Resistance:

$$R_{total} = R_1 + R_2 \ldots.. + R_n$$

1.6: Parallel Resistance:

$$\frac{1}{R_{total}} = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} \ldots.. + \frac{1}{R_n}}$$

2: Commercial Preamp Specifications

2.1: Tech 21, Sansamp Bass Driver

RRP: £165	Technical Specifications Presence: Brings out upper harmonic content. Drive Controls Overdrive and Gain Treb/bass +/- 12dB Blends between unaffected and effected signal Impedances: Input: 1M Output: Low Z	http://www.tech21nyc.com/products/sansamp/bassdriverdi.html
Features: Level Blend Treble Bass Drive Presence Line/Instrument Outputs, Both XLR and Jack Parallel out Footswitch – Full Bypass Accepts 48V Ground Lift		

2.2: East, STMP - 01

RRP: £216.40	**Technical Specifications:** 6dB gain for output jack 12dB for XLR (line) Bass – Boost only. 50Hz, 18dB.	http://www.east-uk.com/
Features: Bass Mid		

Treble Footswitch for Mid Frequency Ground Lift Footswitch – Full Bypass Bright Switch + 8dB @ 7Hz	Treble +12dB @ 3K, -12dB@1K Mid with Variable Frequency +/- 12dB, 150Hz-3KHz	

2.3: EBS, MICROBASS 2

RRP: £286	Technical Specifications: Frequency Response+0 / -3 dB20 - 20k Hz Tone Controls Bass+/- 15 dB @ 100 Hz Treble+/- 15 dB @ 10 kHz shelving Bright+ 12 dB @ 10 kHz,15dB gain Midrange+/- 12 dB, 50 –2.000 Hz ,Q=0.80 Impedances: Input Impedance A: 10M Input Impedance B: 1M Output Impedance: 10K	http://www.bass.se/2009/microbassII.htm
Features: Mute Footswitch Boost Drive Bass Treble Mid w/variable frequency Volume FX Mix Tube Sim GND Lift Parallel Output Headphone Output Enhancer Filter Bright Switch FX Send/Return Speaker Sim		

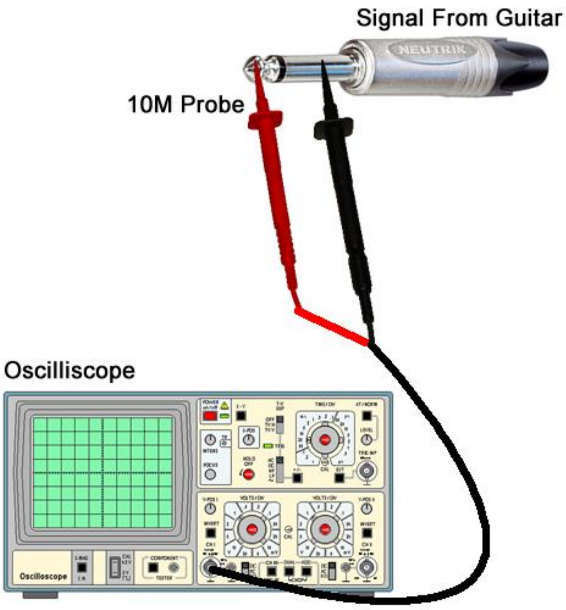
2.4: RADIAL: BASSBONE

RRP: £245	Technical Specifications:	

Features Boost Bass Mid Treble Level 1 Level 2 GND Lift Channel 1 Contour Shaping Boost footswitch Input selection footswitch 2xinst IN Balanced DI Out Unbalanced inst Out TRS Insert	Contour Shape 1: + 45dB @ 110Hz -5.1dB @ 650Hz +2.7dB @ 4.3KHz Contour Shape 2: +3.7dB @ 75Hz -6.3 dB @ 350Hz +11dB @ 5.6kHz Bass:+/- 12dB @ 75Hz Mid: +/- 10dB @ 470Hz High: +/- 12dB @ 5.6kHz Impedances: Input: 470K Output: 10K DI Out: 600Ohms	http://www.tonebone.com/tb-bassbone-features.htm

3: Bass Output Measuring

Using the 10M probes on the oscilloscope, the instrument was sensitive enough to get good readings from the bass guitars. The following diagram shows the test setup.



The peak output voltage was held on the oscilloscope and noted. To determine the absolute work peak possible the bass was played extremely aggressively.

4: Voltage Divider Equations

A full list of equations used to find the output voltage of a resistive or reactive voltage divider.

$$Vout = Vin\frac{R_2}{R_1 + R_2}$$

$$R_1 = \left(\frac{10^{\frac{A}{20}} - 1}{10^{\frac{A}{20}}}\right)Z$$

$$R_2 = \left(\frac{1}{10^{\frac{A}{20}} - 1}\right)Z$$

$Where$:

$A = Attenuation$

$Z = Series\ Impedance$

$$I = \frac{Vin}{Z_{total}}$$

$$Z_{total} = Z_1 + Z_2$$

$$V_{out} = \frac{Z_2}{Z_1 + Z_2}$$

$$X_C = -j/\omega C$$

$$j = \sqrt{-1}$$

$$Vout = Vin \frac{R}{\left[R^2 + \left(\frac{1}{\omega^2 C^2}\right)\right]^{\frac{1}{2}}}$$

5: Slew Rate Equations

$$Slew\ Rate \geq 2\pi V_{pp} f_c$$

Baker. B. (n.d).

$Where$:

$V_{pp} = Peak\ to\ Peak\ Voltage$

$f_c = Cuttoff\ Frequency$

6: Op Amp Specifications

6.1: Real

Manufacturer	Model	Number of Channels	Maximum Supply Voltage	
			+Vcc	-Vee
Texas Instruments	TL071	1	+18V	-18V
	TL072	2	+18V	-18V
	TL074	3	+18V	-18V
	OPA627	1	+18V	-18V
National Semiconductor	LM6171	1	+36V	-36V
	LM6172	2	+36V	-36V
	LM833	2	+36V	-36V
	LM4562	2	+17V	-17V

Maximum Input Voltage	Differential Input Voltage	Slew Rate	Band Width	CMRR	Price
±15V	±30V	13V/us	3MHz	100dB	£0.67
±15V	±30V	13V/us	3MHz	100dB	£0.77
±15V	±30V	13V/us	3MHz	100dB	£0.99
±20V	±22V	55V/us	16MHz	100dB	£19.35
Unknown	±10V	3600V/us	100MHz	100dB	£3.15
Unkown	±10V	3600V/us	100MHz	100dB	£3.79
±15V	±30V	7V/us	15MHz	100dB	£0.43
Unknown	Unknown	20V/us	55MHz	120dB	£2.90

6.2: Ideal

- Infinite Bandwidth
- Infinite Gain
- Infinite Input Impedance
- 0 Output Impedance
- 0 Noise
- Infinite CMRR
- Infinite Slew Rate

7: Filter Definitions

Term	Definition
Pass band	The band of frequencies passed by a filter network.
Stop Band	The band of frequencies rejected by a filter network.
Cut off frequency	The frequency at which the signal has been attenuated by 3dB.

Band Width	The difference between 2 cut off frequencies in a band pass or stop filter.
Quality Factor	A numerical representation of bandwidth in relation to Centre frequency.
Centre Frequency	The centre frequency of a pass band.
Order	The amount of orders in a filter is determined by the amount of reactive components.
Transfer Function	The Difference between the output and input of a signal.

8: Impedance Equations

8.1: Impedance:

$$Z = R + X_c + X_L$$

$$X_L = j\omega L$$

$$X_c = 1/j\omega C$$

8.2: Ohms Law

$$V = IZ$$

8.3 Series Impedance

$$Z_{total} = Z_1 + Z_2 \ldots\ldots + Z_n$$

8.4 Parallel Impedance

$$\frac{1}{Z_{total}} = \frac{1}{\frac{1}{Z_1} + \frac{1}{Z_2} \ldots\ldots + \frac{1}{Z_n}}$$

9: Passive Filter Equations

9.1 Low Pass Transfer Function

$$Vout = Vin\frac{2\pi fRC}{[1 + \omega^2 R^2 C^2]^{\frac{1}{2}}}$$

9.2 High Pass Transfer Function

$$Vout = Vin\frac{2\pi fRC}{[1 + (2\pi fRC)^2]^{\frac{1}{2}}}$$

9.3 Centre Frequency

$$f_c = 1/2\pi RC$$

9.4 Band Pass Q

$$Q = \frac{f_c}{f_{bw}}$$

10: MFBP Equations

$$T_{(s)} = \frac{sR_3C}{s^2R_1R_3C^2 + s2R_1C + (1 + \frac{R_1}{R_2})}$$

$$R_1 = \frac{R_3}{2A_r}$$

$$R_2 = \frac{R_3/2}{2Q^2 - A_r}$$

$$R_3 = \frac{Q}{\pi f_r C}$$

$$A_r = K_{f_r}$$

Where:

$A_r = Gain\ at\ Resonant\ Frequency$

$s = Complex\ variable\ representing\ amplitude\ and\ phase$

$Q = Quality\ Factor$

$A_r < 2Q^2$

Williams, A & Taylor, F. (1995)

11: Initial Specifications

Op Amps	TL071, TL072, TL074. Acceptable Slew Rate Low Noise 100dB CMRR Cheap Low Power consumption
Controls	Controls: Volume Bass Mid

	Mid Frequency Switch (DPDT) Treble Pad Switch (DPDT)
Equaliser	Bass +/- 12dB @ 100Hz. Mid +/- 12dB Freq 1: 250Hz Freq 2: 600Hz Treble +/- 12dB @ 2.5 kHz Active Multi-feedback networks
Input	Impedance: 1M Switchable -20dB Pad. (L-Pad Design) Buffer Stage Unbalanced Jack
Gain Stage	IC Based, Non Inverting 0-12dB Variable Gain.
Line Output Stage	Impedance: 600 Ohms Ground Lift XLR
Instrument Output	Impedance 10K Unbalanced Jack 20dB pad
Power	9v Battery DC Jack DC Jack cuts battery Power Bias at 4.5V

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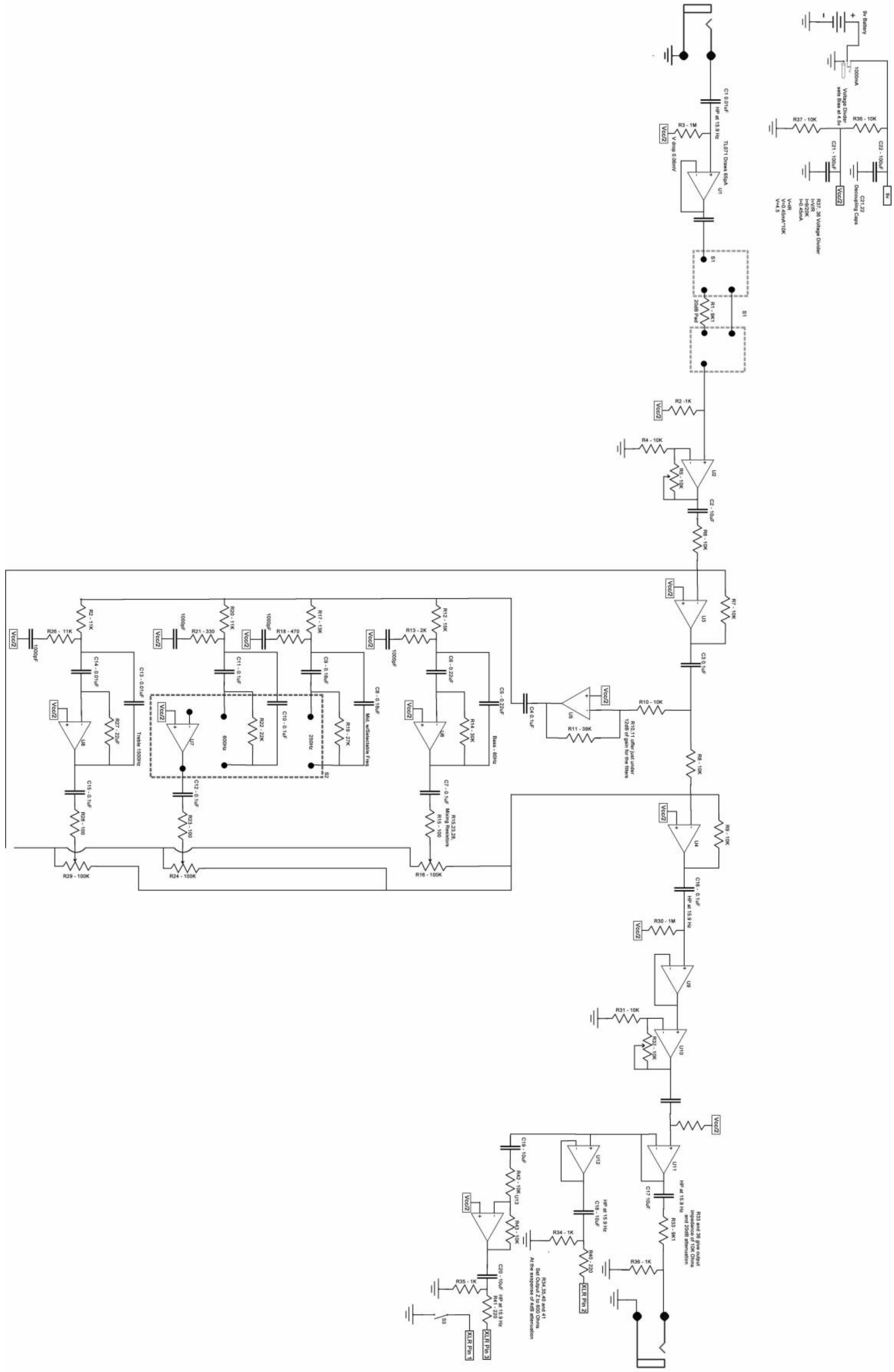
12: Initial Schematic



13: Schematic and Component Calculations

13.1: L Type Attenuation Pads

Equations	$R_1 = \left(\frac{10^{\frac{A}{20}} - 1}{10^{\frac{A}{20}}}\right) Z$	$R_2 = \left(\frac{1}{10^{\frac{A}{20}} - 1}\right) Z$
Instrument Input & Input Pad $Impedance = Z = 10000$ $Attenuation = A = 20$		
Ideal Value	$R_1 = 9000$	$R_2 = 111.11$
Actual Value	$R_1 = 9K1$	$R_2 = 1K$
Balanced Output $Z = 600$ $A = 4$		
Ideal Value	$R_1 = 221.43$	$R_2 = 1025.83$
Actual Value	$R_1 = 220$	$R_2 = 1K$

13.2: Gain Stages

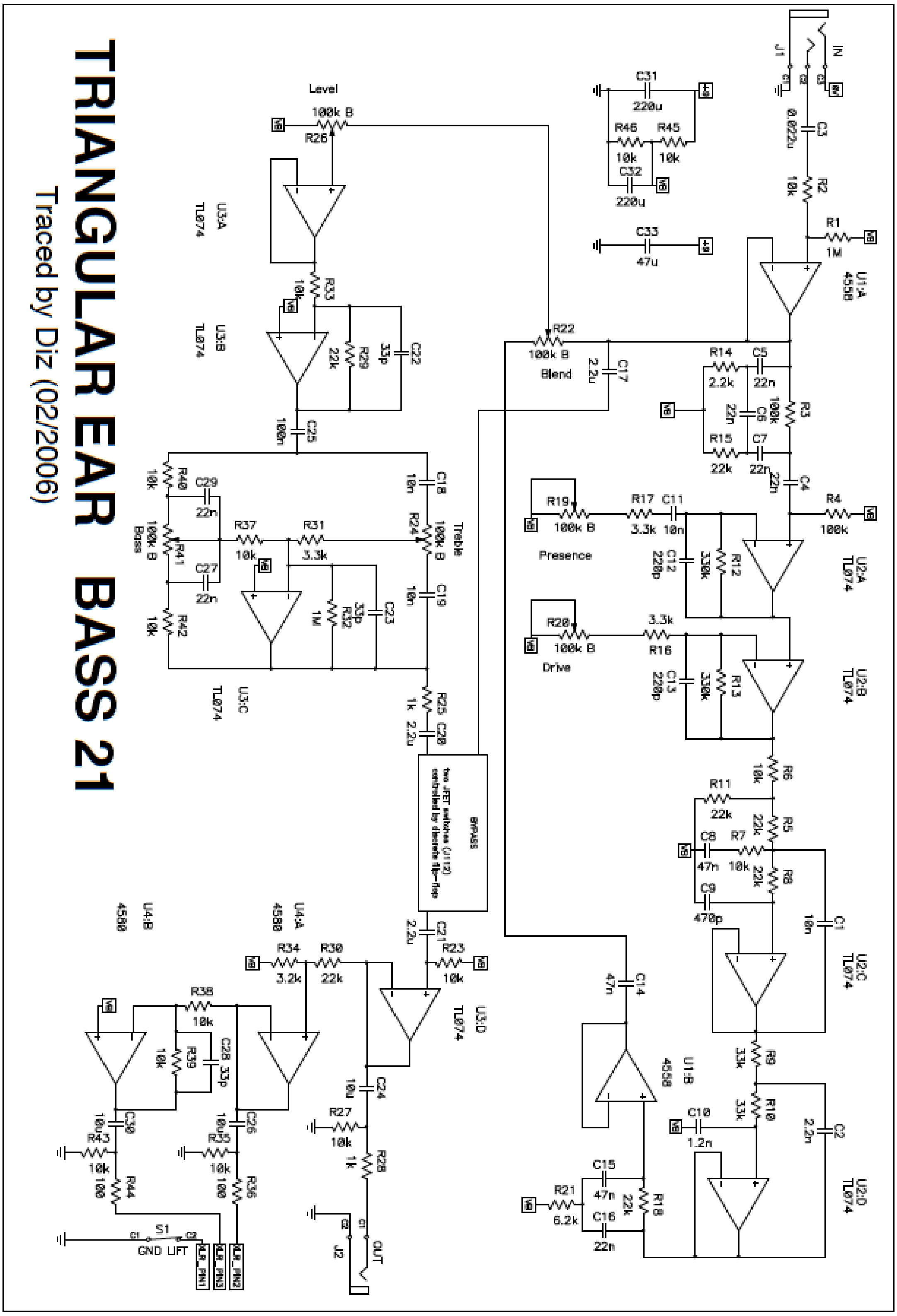
Non-Inverting	$\boldsymbol{Gain = 1 + (\frac{R_2}{R_1})}$
Gain Stages	
Gain = 2 (approx. 6dB)	$R_2 = R_1$
Inverting	$\boldsymbol{Gain = -\frac{R_2}{R_1}}$
Filter Gain	
Gain = 4 (approx. 12dB)	$R_2 = 4 \times R_1$

13.3: Multi-feedback Components

Equations	$\boldsymbol{R_3 = \frac{Q}{\pi f C}}$	$\boldsymbol{R_1 = \frac{R_3}{2A_r}}$	$\boldsymbol{R_2 = \frac{R_3/2}{2Q^2 - A_r}}$

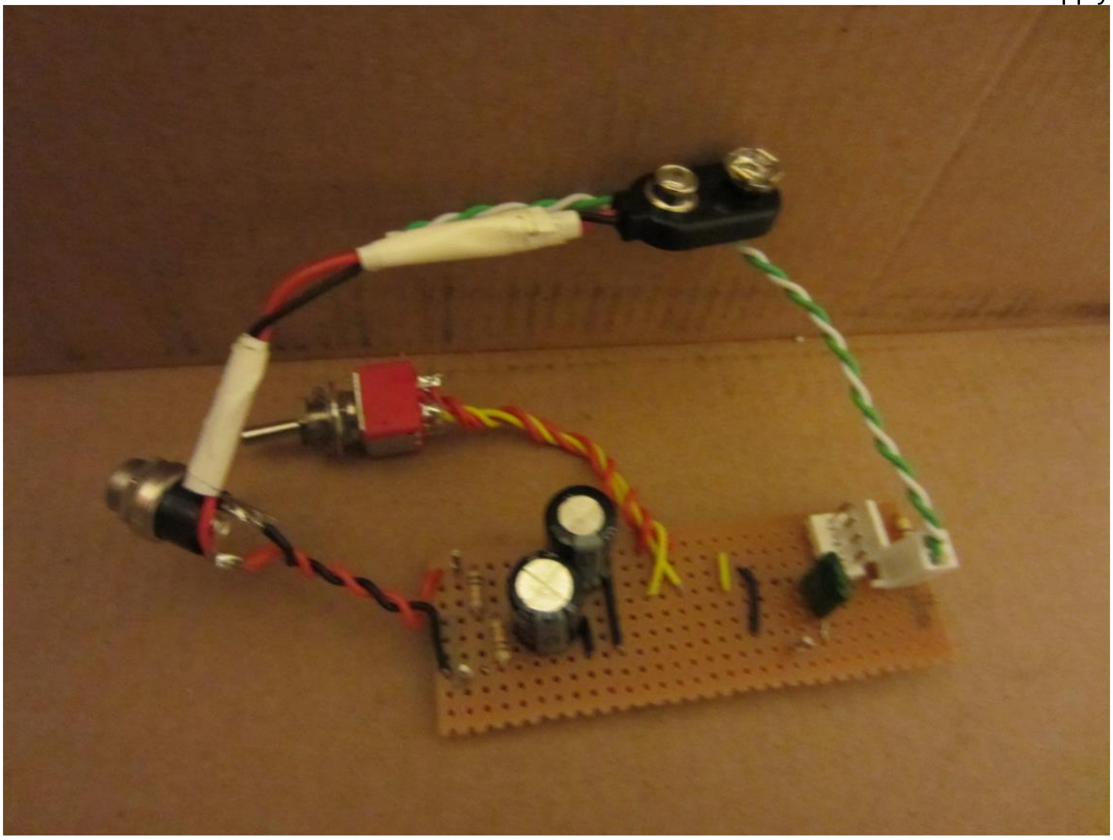
Bass $Q = 2$ $f = 100$ $C_1 = C_2 = 0.22uF$ $A_r = 1$			
Ideal Values	$\boldsymbol{R_3 = 28937.262}$	$\boldsymbol{R_1 = 14468.6311}$	$\boldsymbol{R_2 = 2066.947}$
Actual Values	$\boldsymbol{R_3 = 30K}$	$\boldsymbol{R_1 = 15K}$	$\boldsymbol{R_2} =$ 2K
Mid a $Q = 4$ $f = 250$ $C_1 = C_2 = 0.18uF$ $A_r = 1$			
Ideal Values	$\boldsymbol{R_3 = 28294.2121}$	$\boldsymbol{R_1 = 14147.1060}$	$\boldsymbol{R_2 = 456.3583}$
Actual Values	$\boldsymbol{R_3} =$ 27K	$\boldsymbol{R_1} =$ 13K	$\boldsymbol{R_2} =$ 470
Mid b $Q = 4$ $f = 600$ $C_1 = C_2 = 0.1uF$ $A_r = 1$			
Ideal Values	$\boldsymbol{R_3 = 21220.6590}$	$\boldsymbol{R_1} =$ 10610.32954	$\boldsymbol{R_2} =$ 342.2686984
Actual Values	$\boldsymbol{R_3} =$ 22K	$\boldsymbol{R_1} =$ 11K	$\boldsymbol{R_2} =$ 330
Treble $Q = 1$ $f = 1500$ $C_1 = C_2 = 0.01uF$ $A_r = 1$			
Ideal Values	$\boldsymbol{R_3 = 21220.6590}$	$\boldsymbol{R_1} =$ 10610.32954	$\boldsymbol{R_2} =$ 10610.32954
Actual Values	$\boldsymbol{R_3} =$ 22K	$\boldsymbol{R_1} =$ 11K	$\boldsymbol{R_2} =$ 11K

14: Triangular Ear 21 Schematic

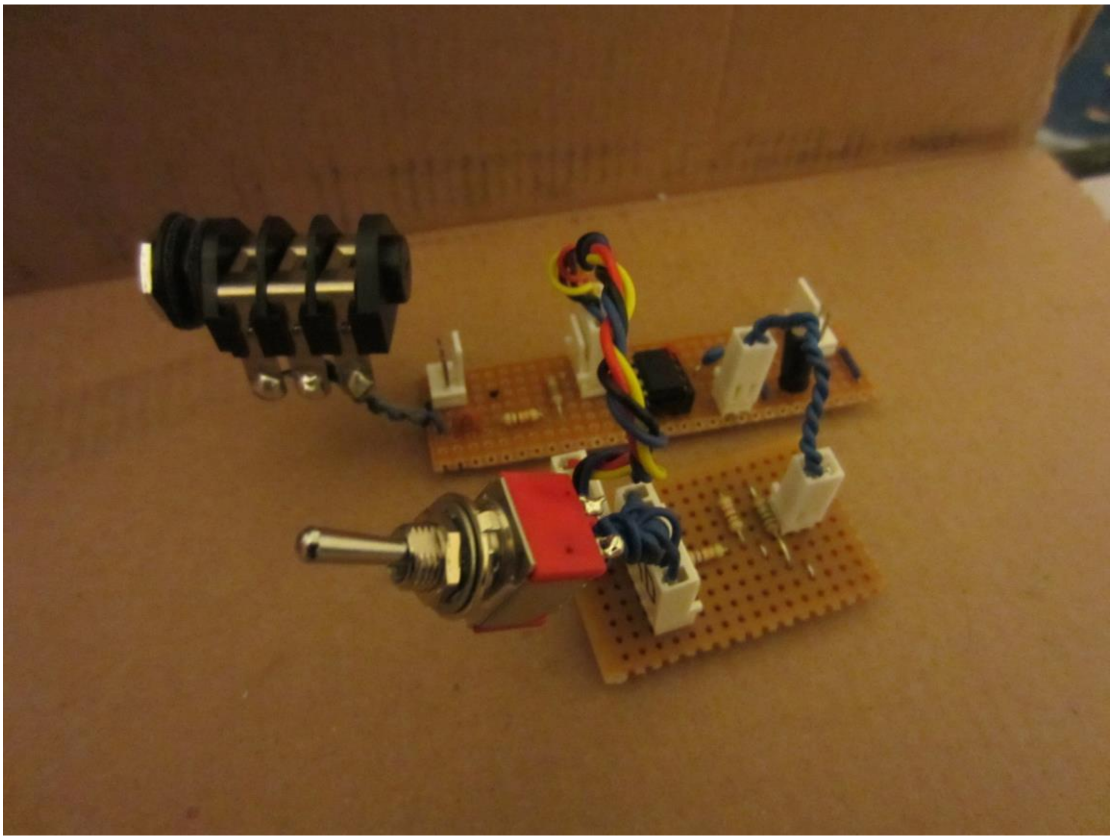


15: Module Photos

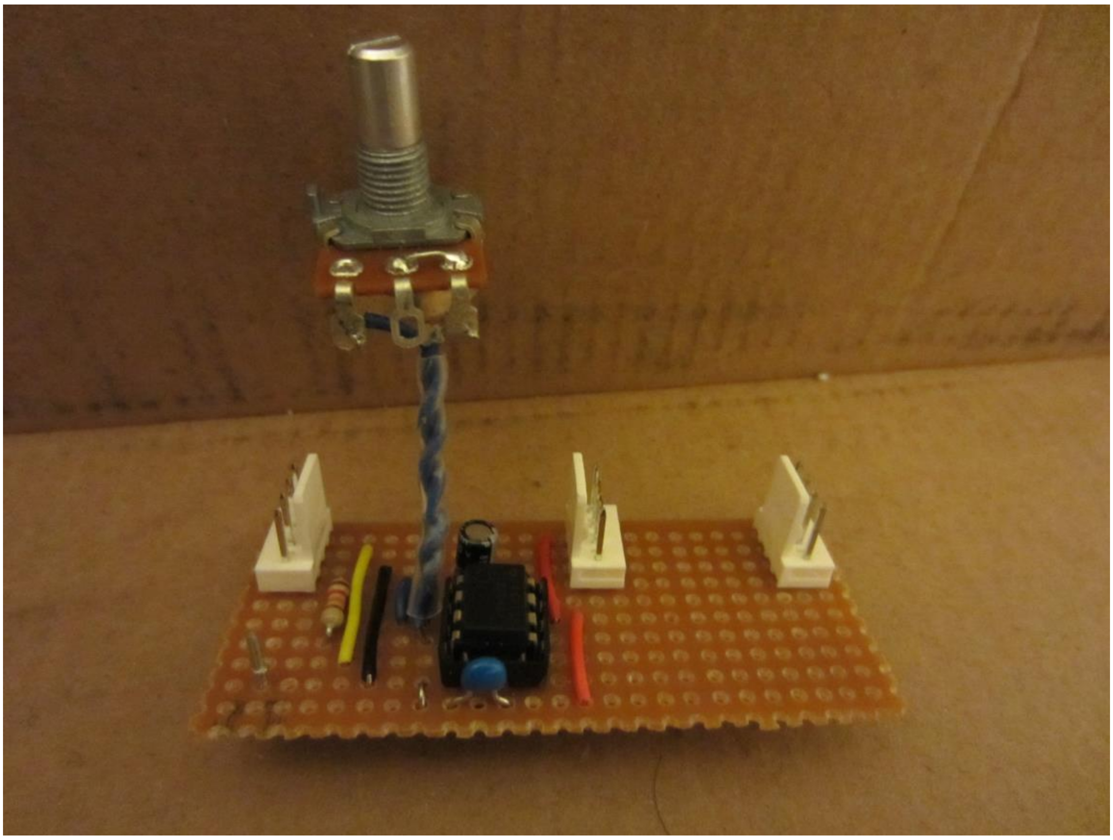
Power Supply



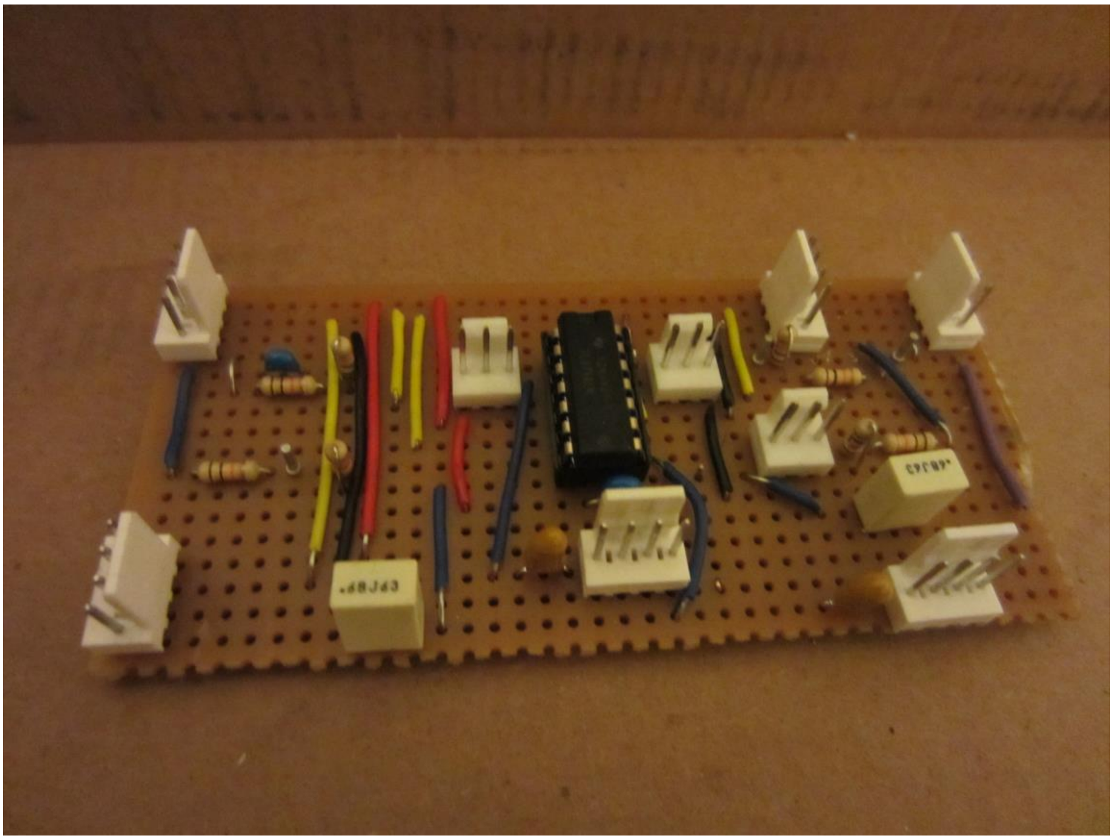
Input stage and Pad



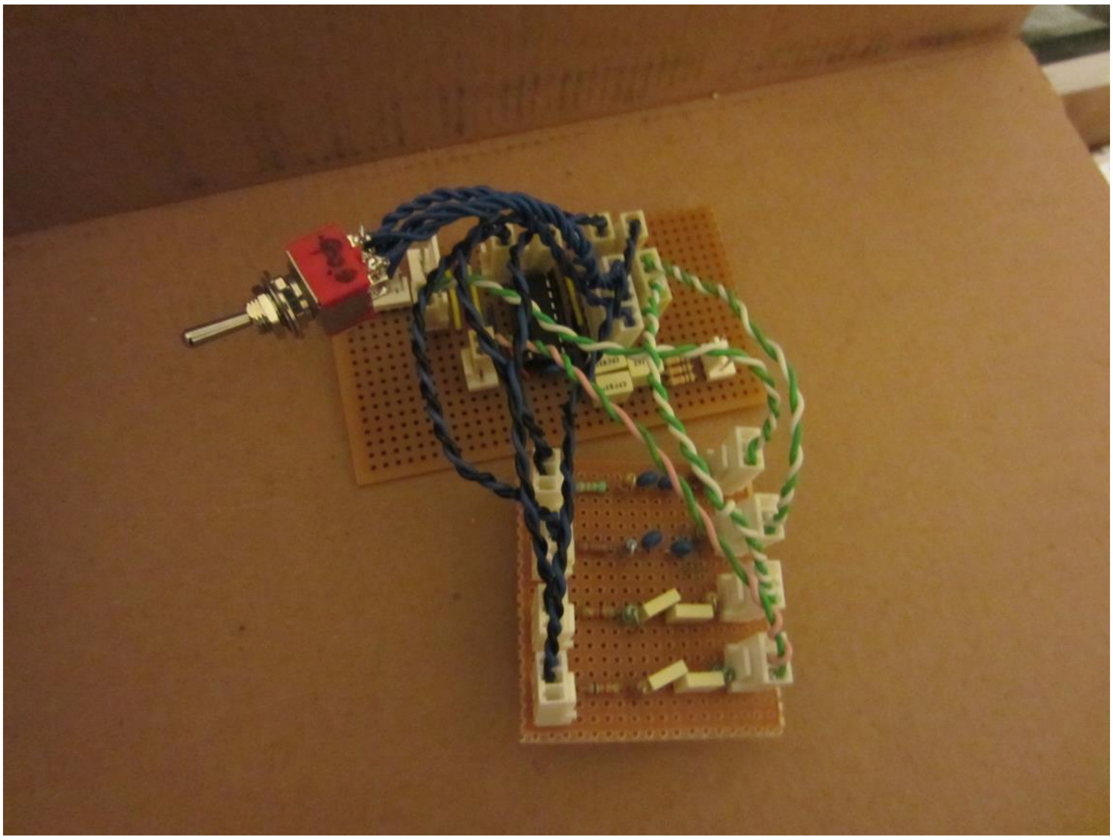
First gain stage



Summing Amplifiers



Band pass filters and amplifier



Second Gain stage

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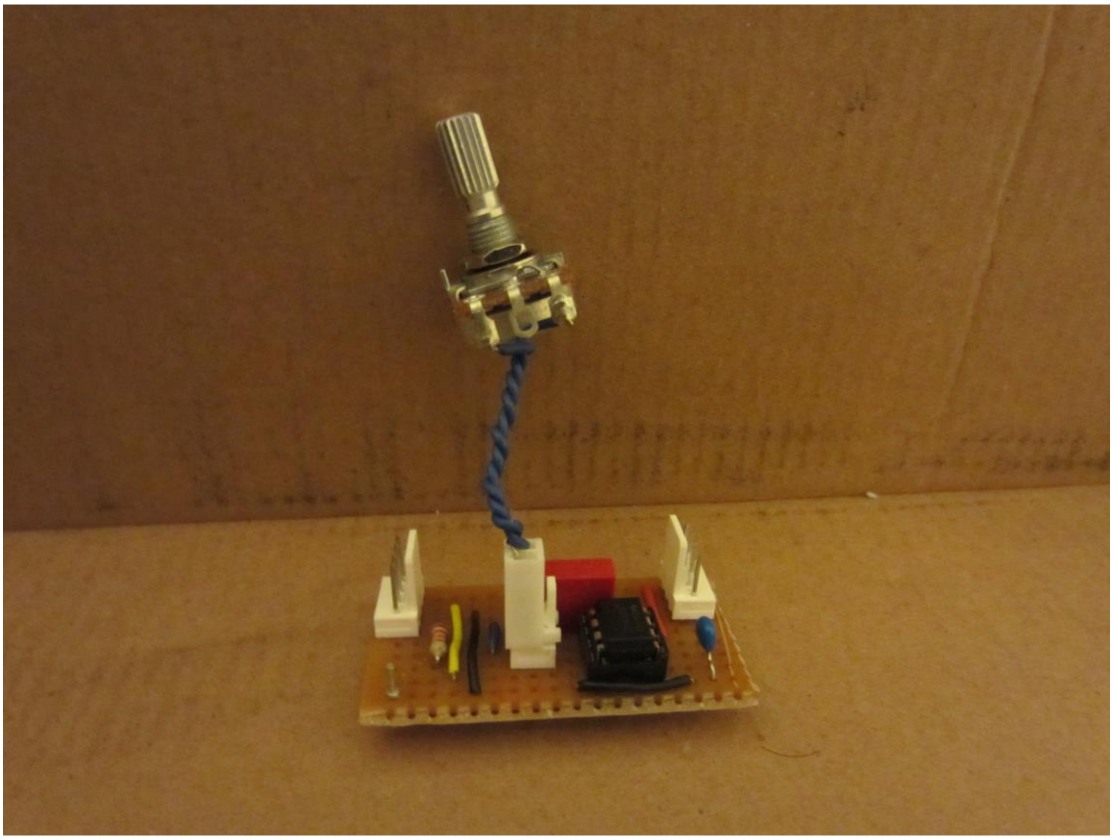
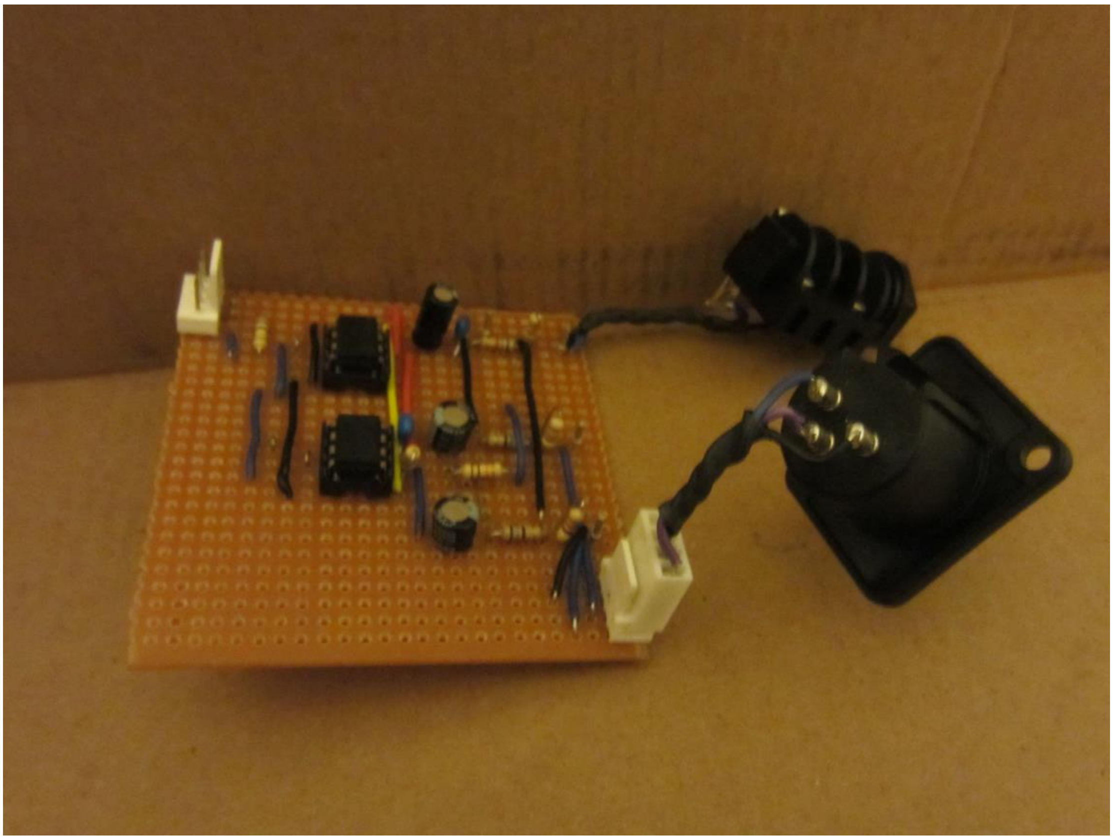


Output



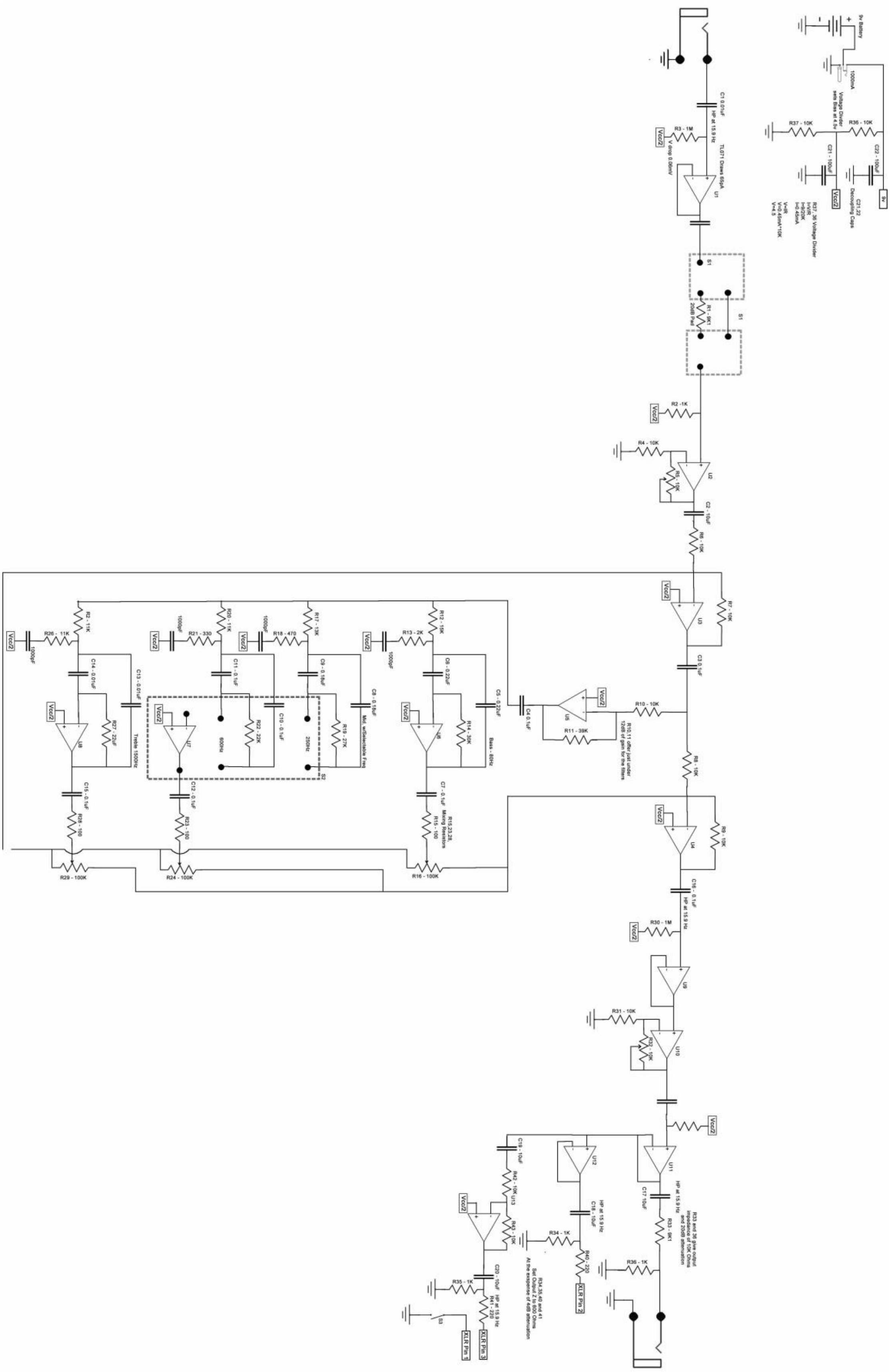
16: Wire Colour Coding

Colour	Carried Signal
Red	+Vcc
Yellow	+Vcc/2
Black	Ground
Blue	AC Signal
Green	Filter Amp Input
Pink	Filter Amp Feedback
Orange	100Hz
Purple	250/600Hz
White	1kHz

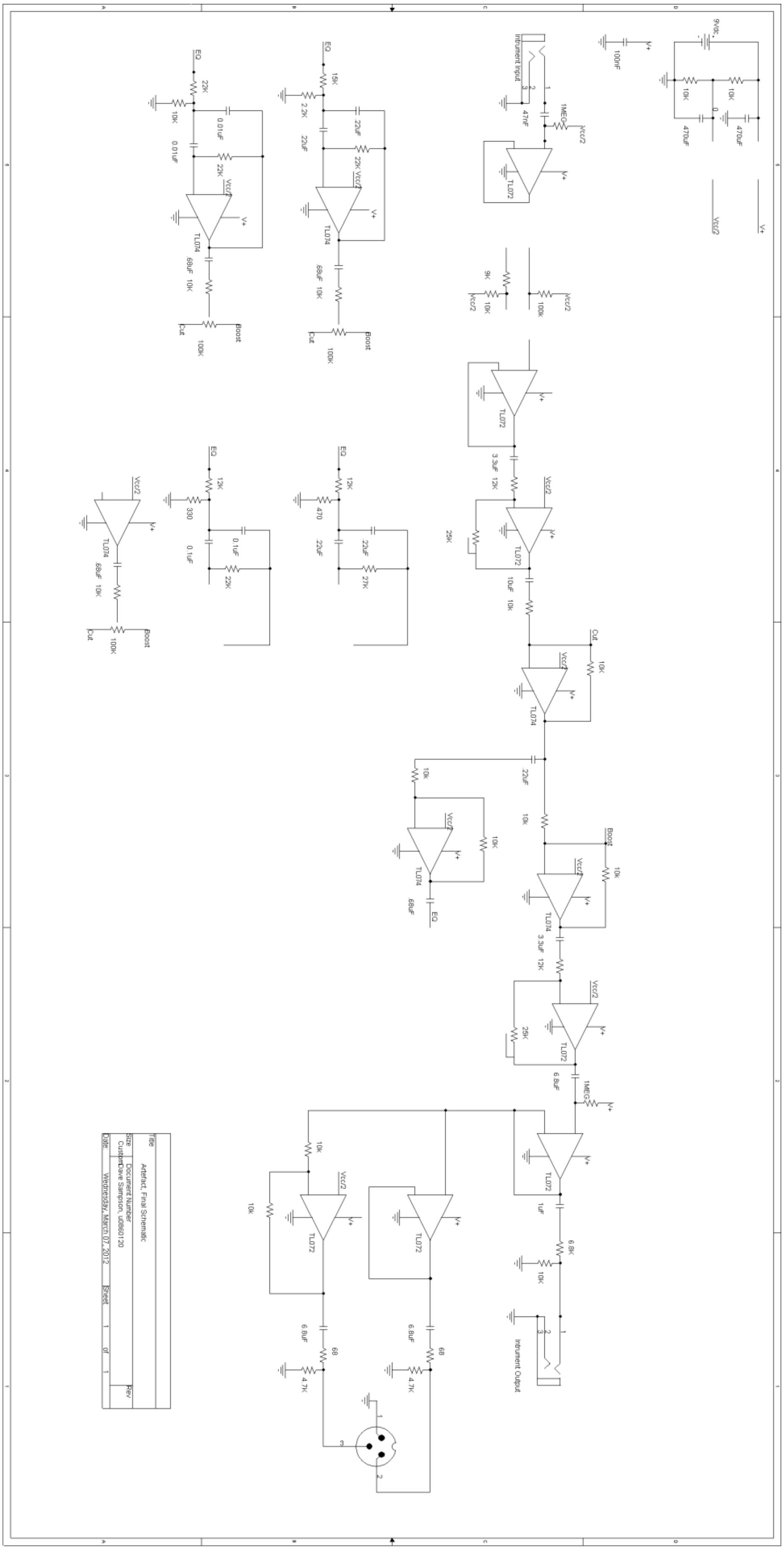
17: Molex Pin Layout

1. V+
2. V+/2
3. 0v
4. Signal

18: Initial Schematic



19: Final Schematic



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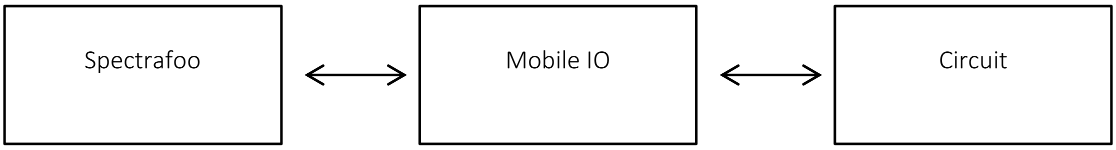
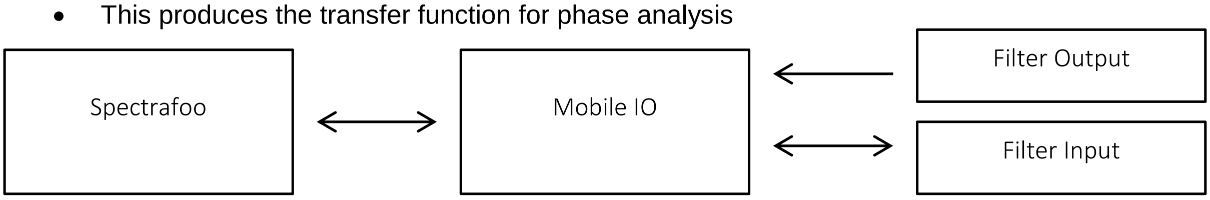
20: Frequency Response Measurement Technique

- Spectrafoo produces sine sweep
- Sine sweep parameters:
    - Start Frequency: 20Hz
    - End Frequency: 20kHz
    - 1.2Vpp Output
    - Duration: 2 Minutes
    - Swept Twice



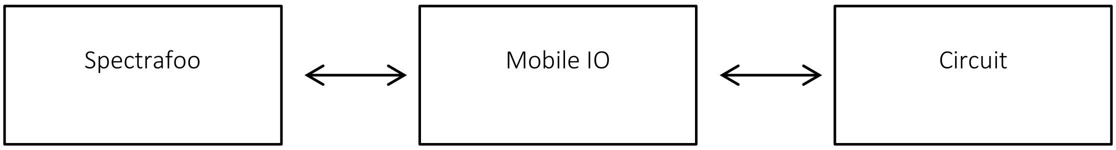
21: Phase Response Measurement Technique

- Spectrafoo sweeps the circuitry and compares the output to the input
- This produces the transfer function for phase analysis



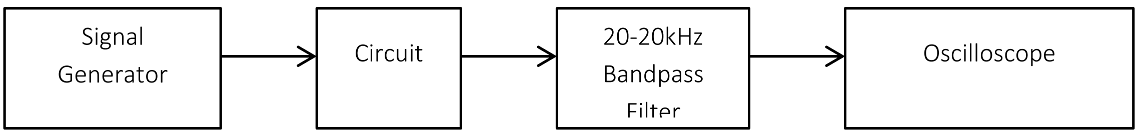
22: THD+N Measurement Technique

- Spectrafoo produces a pure sine wave at the circuit input
- Spectrafoo uses a tight notch filter to remove this signal from the output signal
- Spectrafoo then measures the leftover signal for harmonic distortion and noise
- Circuit receives 1.2Vpp and outputs 1.2Vpp



23: Signal to Noise Ratio Measurement Technique

- Signal generator produces 1.2Vpp Sine wave at the input
- 1.2Vpp measured at the output
- Bandpass filter limits the measurement to an audible frequencies
- Oscilloscope measures noise while circuit is in quiescent state.
- Oscilloscope set to trigger highest voltage, obtaining a worst case measurement



Reference Level: 1.2Vpp

Measured Noise Level: 0.00131Vpp

Calculation:

$$S/N_{db} = 20log\frac{V_r}{V_n}$$

$$S/N_{db} = 20log\frac{1.2}{0.00131}$$

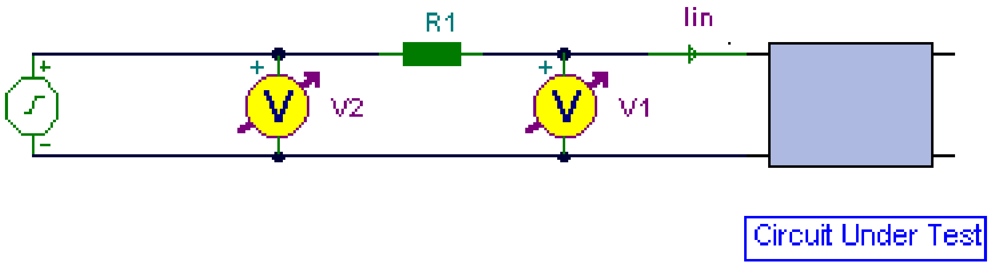
$$S/N_{db} = 59.233$$

Where:

Vr = Reference Voltage

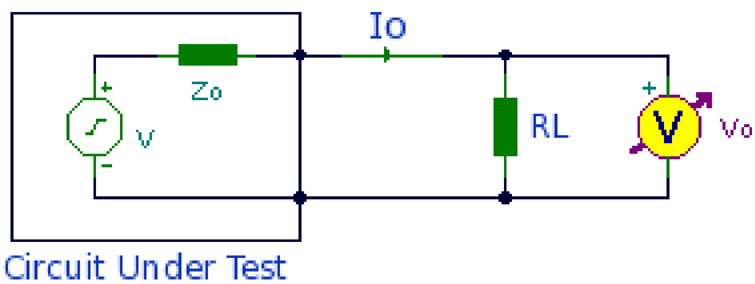
Vn = Noise Voltage

24: Input Impedance Measurement Technique



http://www.zen22142.zen.co.uk/Theory/images/zin.gif

25: Output Impedance Measurement Technique



http://www.zen22142.zen.co.uk/Theory/images/zout.gif

26: Impedance Calculations

26.1: Input

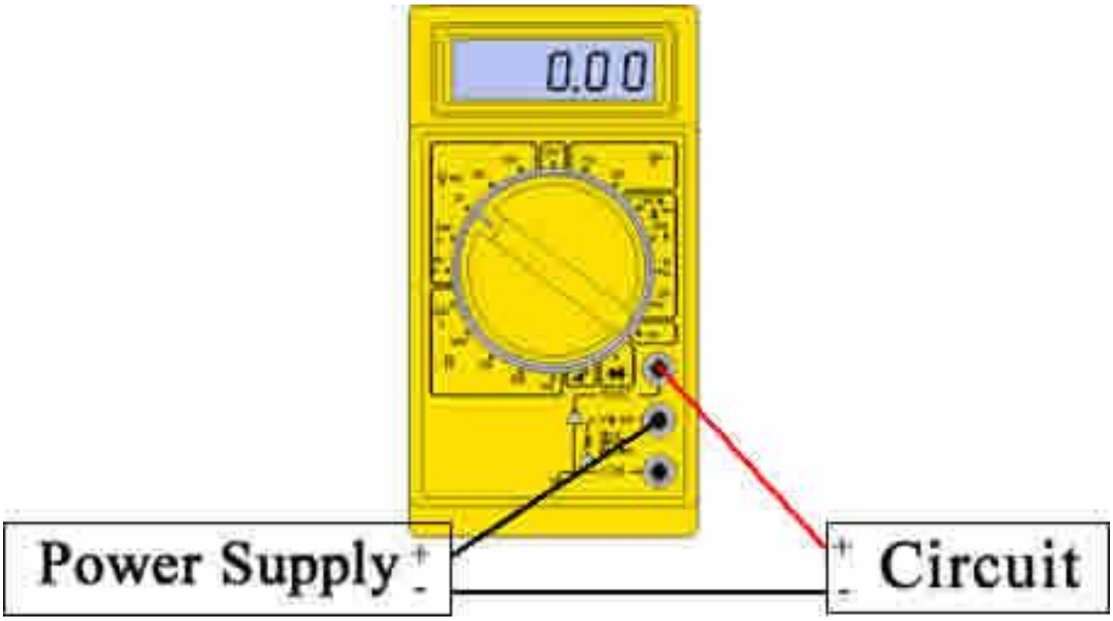
$I_{in} = \frac{V_2 - V_1}{R_1}$ $Z_{in} = \frac{V_1}{I_{in}}$	$I_{in} = \frac{0.529 - 0.526}{10k}$ $Z_{in} = \frac{0.526}{0.0000389} = 1.35M$

26.2: Output

$Z_{out} = \frac{R_{load}(V - V_{load})}{V_{load}}$	Instrument Output $Z_{out} = \frac{10k(1 - 0.42363)}{0.42363} = 13.605k$ Line Output $Z_{out} = \frac{10k(1 - 0.824562)}{0.824562} = 2127.65$

27: Current Draw Measurement Technique

- Multimeter set to 20mA



28: TL07X Electrical Characteristics

electrical characteristics, $V_{CC\pm} = \pm15$ V (unless otherwise noted)

PARAMETER		TEST CONDITIONS†	T_A‡	TL071M TL072M MIN	TL071M TL072M TYP	TL071M TL072M MAX	TL074M MIN	TL074M TYP	TL074M MAX	UNIT
V_{IO}	Input offset voltage	$V_O = 0$, $R_S = 50\ \Omega$	25°C		3	6		3	9	mV
			Full range			9			15	
α_{VIO}	Temperature coefficient of input offset voltage	$V_O = 0$, $R_S = 50\ \Omega$	Full range		18			18		µV/°C
I_{IO}	Input offset current	$V_O = 0$	25°C		5	100		5	100	pA
			Full range			20			20	nA
I_{IB}	Input bias current‡	$V_O = 0$	25°C		65	200		65	200	pA
						50			50	nA
V_{ICR}	Common-mode input voltage range		25°C	±11	−12 to 15		±11	−12 to 15		V
V_{OM}	Maximum peak output voltage swing	$R_L = 10\ k\Omega$	25°C	±12	±13.5		±12	±13.5		V
		$R_L \geq 10\ k\Omega$	Full range	±12			±12			
		$R_L \geq 2\ k\Omega$		±10			±10			
A_{VD}	Large-signal differential voltage amplification	$V_O = \pm10$ V, $R_L \geq 2\ k\Omega$	25°C	35	200		35	200		V/mV
				15			15			
B_1	Unity-gain bandwidth	$T_A = 25°C$			3			3		MHz
r_i	Input resistance	$T_A = 25°C$			10^{12}			10^{12}		Ω
CMRR	Common-mode rejection ratio	$V_{IC} = V_{ICR}min$, $V_O = 0$, $R_S = 50\ \Omega$	25°C	80	86		80	86		dB
k_{SVR}	Supply-voltage rejection ratio ($\Delta V_{CC\pm}/\Delta V_{IO}$)	$V_{CC} = \pm9$ V to ±15 V, $V_O = 0$, $R_S = 50\ \Omega$	25°C	80	86		80	86		dB
I_{CC}	Supply current (each amplifier)	$V_O = 0$, No load	25°C		1.4	2.5		1.4	2.5	mA
V_{O1}/V_{O2}	Crosstalk attenuation	$A_{VD} = 100$	25°C		120			120		dB

29: Maximum Gain Readings

	Without EQ (9v)			With EQ (9v)			With EQ (18v)		
	Vpp	dB gain	dBu	Vpp	dB gain	dBu	Vpp	dB gain	dBu
Instrument	3.24	8.6272	3.392	3.36	8.9432	3.70	8.24	16.734	11.500
Line	5.44	13.128	7.893	5.76	13.624	8.39	14.2	21.462	16.227

30: Maximum Gain Calculations

The input signal was a 1.2Vpp 1kHz sine wave. This reading includes the 12dB of gain offered by the 1kHz boost.

30.1: 18v Maximum Line level Output

dB

$$dB_{gain} = 20log_{10}\frac{V_2}{V_1}$$

$$dB_{gain} = 20log_{10}\frac{14.2}{1.2}$$

$$dB_{gain} = 21.4621$$

dBu

$$14.2Vpp = 7.1Vp$$

$$Vrms = 7.1Vp \times 0.707 = 5.0197$$

$$dBu = 20log_{10}\frac{5.0197}{0.775}$$

$$dBu = 16.2275$$

30.2: 9V Maximum Line Level Output

dB

$$dB_{gain} = 20log_{10}\frac{5.76}{1.2}$$

$$dB_{gain} = 13.6248$$

dBu

$$5.76Vpp = 2.88Vp$$

$$Vrms = 7.1Vp \times 0.707 = 2.036$$

$$dBu = 20log_{10}\frac{2.036}{0.775}$$

$$dBu = 8.3895$$

30.3: 18V Maximum Instrument

dB

$$dB_{gain} = 20log_{10}\frac{8.24}{1.2}$$

$$dB_{gain} = 16.7349$$

dBu

$$8.24Vpp = 4.12Vp$$

$$Vrms = 4.41Vp \times 0.707 = 2.20584$$

$$dBu = 20log_{10}\frac{2.20584}{0.775}$$

$$dBu = 11.5002$$

30.4: 9V Maximum Instrument

dB

$$dB_{gain} = 20log_{10}\frac{3.36}{1.2}$$

$$dB_{gain} = 8.9431$$

dBu2

$$6.72Vpp = 3.36Vp$$

$$Vrms = 3.36Vp \times 0.707 = 1.68$$

$$dBu = 20log_{10}\frac{1.68}{0.775}$$

$$dBu = 3.7085$$

30.5: 1.2VPP Reference Signal

dBu

$$1.2Vpp = 0.6Vp$$

$$Vrms = 0.6Vp \times 0.707 = 0.4242$$

$$dBu = 20log_{10}\frac{1.68}{0.775}$$

$$dBu = -5.2346$$

31: Nyquist Theory

$$f_{smin} = f_{max} \times 2$$

Where:

Fsmin = Minimum sampling frequency

Fmax = Maximum frequency needed

32: Metric Halo Mobile IO 2882 Specifications

- 4 channels balanced XLR inputs. Each input has:
  - 24-bit 96kHz A/D converters (110dB SNR)
  - remote controllable pre-amps with 40 dB of gain
  - remote switchable input impedance characteristics
  - remote switchable 20dB pad
  - remote switchable 48v Phantom power, with 10mA current limit
- 4 channels balanced TRS inputs. Each input has:
  - 24-bit 96kHz A/D converters (110dB SNR)
  - remote controllable pre-amps with 40 dB of gain
  - remote switchable input impedance characteristics
  - remote switchable 20dB pad
  - remote switchable 48v Phantom power, with 10mA current limit
- 8 channels balanced TRS outputs. Each output has:
  - 24-bit 96kHz D/A converters (120dB SNR)
  - remote controllable output gain (from -12dBV up)

Mic/Line Inputs	
Line +4 Gain Range	-2 dB – +40.5 dB
Line -10 Gain Range	-13.8 dB – +28.7 dB
Inst Gain Range	0 dB – +42.5 dB
Mic Gain Range	0 dB – +42.5 dB
Mic Pad Gain Range	-20 dB – +22.5 dB
Line Input Impedance	10k Ω
Instrument Input Impedance	200k Ω
Mic Input Impedance	200k Ω (12k Ω with phantom)
Mic Pad Input Impedance	10k Ω (6k Ω with phantom)

Maximums	
Max Gain	42.5 dB
Preamp Headroom	20 dB above Digital Clip
Phantom Power	+48v Regulated, high current, individually switch-able, P48 test compliant, short circuit/ hot-swap pro-tected
Output	+26 dBu