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Battery Powered Crossover for In-Ear Monitors

Helsinki Metropolia University of Applied Sciences

Bachelor of Engineering

Electronics

Thesis

27.1.2015

Helsinki Metropolia University of Applied Sciences

Author(s) Title Number of Pages Date	Arttu Valtteri Nurmi Battery Powered Crossover for In-Ear Monitors 47 pages + 8 appendices 27 January 2016
Degree	Bachelor of Engineering
Degree Programme	Electronics
Specialisation option	
Instructor(s)	Kai Lindgren, Senior Lecturer

The goal of this thesis was to design and implement a portable active three way audio crossover for use with high end earphones. This crossover system is housed in its own separate enclosure and is connected between an audio source and earphones.

A 24dB/octave Linkwitz-Riley response was chosen for the electronic filters making up the crossover. A functional schematic and small size printed circuit board layout were then designed for the crossover. Great care was taken in component selection to find the optimum balance between performance and footprint.

Three versions of the design were built, each an improvement on the previous one. The initial goals set for overall audio performance were met and surpassed by a large margin. The result is an active crossover far exceeding any passive systems in use currently for this application, with the only drawback being the obvious one; the need for an external unit between source and earphones.

With all its core functionality better than necessary, any improvements or further study should look into either making the unit smaller, thus more practical, or adding additional features such as wireless connectivity to source. These features could help make the system a more attractive consumer product.

Keywords	crossover, in-ear monitor, earphone, audio, filter, op amp

Tekijä(t) Otsikko Sivumäärä Päivämäärä	Arttu Valtteri Nurmi Akkukäyttöinen jakosuodin korvamonitoreille 47 sivua + 8 liitettä 27 Tammikuuta 2016
Tutkinto	Insinööri (AMK)
Koulutusohjelma	Elektroniikka
Erikoistuminen	
Ohjaaja(t)	Kai Lindgren, Lehtori

Tämä lopputyön päämäärä oli suunnitella ja rakentaa kannettava kolmitie aktiivi jakosuodin korvamonitoreita varten. Tämä jakosuodin toimii erillisessä kotelossaan ja kytketään äänilähteen ja korvamonitoreiden väliin.

Jakosuotimen elektronisia filttereitä varten valittiin 24dB/oktaavi Linkwitz-Riley vaste. Kytkentäkaavio ja pienikokoinen piirilevy suunnitelma tehtiin jakosuotimelle. Jokaisen komponentin valinta tehtiin huolella, jotta saavutettaisiin paras tasapaino suorituskyvyn ja koon välillä.

Kolme versiota suunnitelmasta valmistettiin, joista jokainen täytti määritellyt alkuspesifikaatiot edeltäjäänsä paremmin. Lopputuloksena on aktiivijakosuodin, joka ylittää suorituskyvyllään kaikki yksinkertaiset passiivijärjestelmät, jota korvamonitoreissa käytetään.

Tämän jakosuotimen toiminta ylitti kaikki esitetyt parametrit laajalla marginaalilla, joten aiheen jatkokehitys tulisi keskittää joko laitteen pienentämiseen tai uusien ominaisuuksien kehittämiseen. Tällaisia ominaisuuksia voisi olla esimerkiksi langaton yhteys äänilähteeseen. Näin tuotteesta voisi saada entistä houkuttelevamman kokonaisuuden käyttäjille.

Avainsanat	jakosuodin, korvamonitori, kuuloke, operaatiovahvistin

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1 Introduction

The subject of this thesis is the design of a practical portable battery powered three way active crossover unit, to be used in conjunction with purpose built, application specific multi-driver in-ear monitors. It is to be an externally housed crossover system between an input source player such as a mobile phone or an equivalent portable music player, and custom built earphones.

This is an adaptation of an idea that is commonly used in full size desktop active hi-fi speakers, but miniaturized, and re-designed for use in a portable pocket sized hi-fi system, roughly the size of a small headphone amplifier and in fact can be considered a six channel headphone amplifier, with each channel supplying a separate audio band. This type of ear phone specific system has been attempted in just two previous commercial products, one of which was quickly discontinued and the other receiving less than positive reviews, due to numerous problems. The design in this paper attempts to correct the mistakes that made those earlier products fail.

Due to its specialized nature, high cost and a very niche appeal, this product is by no means intended as a mass market item, but as a study into extending the capabilities of portable in-ear monitors as far possible, with the fewest sacrifices in terms of audio quality, portability and ease of use. This thesis focuses on the design of the active crossover and the accompanying ear phone are not discussed in detail.

2 Project Overview

2.1 In-Ear Monitors

In-ear monitors or IEMs for short (also sometimes called canal phone), are small type of earphone that is inserted deep into the ear canal for better fit and isolation than earbud types. Often used by musicians and recording industry professionals, though rapidly growing in consumer usage as they start to replace the very common earbud. In addition to the insertion depth, one of the main differences between IEMs and traditional consumer earbuds are the transducer technology used inside the housing. The common earbud transducer is a dynamic driver, which is essentially a miniature version of the conical shape moving coil speaker found in virtually all full size speakers. IEMs often employ a balanced armature speaker, which differs quite a bit in construction from dynamic drivers, although dynamic drivers are used in some IEMs as well. IEMs can contain two or more of these balanced armature drivers for a wideband audio output. In fact some of the very top end custom made models can have up to a dozen balanced armatures inside each earphone housing. Figure 1 shows an example of the internal components of a multi driver custom IEM.

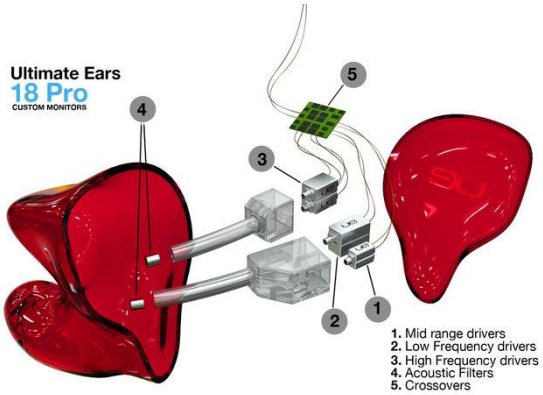


Figure1: Exploded view of a multi driver custom in-ear monitor. Copied from Ultimate Ears [1]

2.2 Balanced Armatures

Balanced armatures can have very small physical dimensions compared to moving coil drivers. They are very sensitive, thus require very little input power to drive to adequate listening volumes. This small size and high efficiency make them ideal for portable audio applications.

Compared to a traditional dynamic speaker the internal operation of balanced armature is somewhat different. It contains a moving armature between two magnets. This armature is coupled to a diaphragm or membrane via a drive pin. As current passes through the internal coil it creates a changing magnetic field that moves the armature, which in turn transfers the movement to the membrane that pushes air out of the armature. The internal components can be seen in figure 2 below:

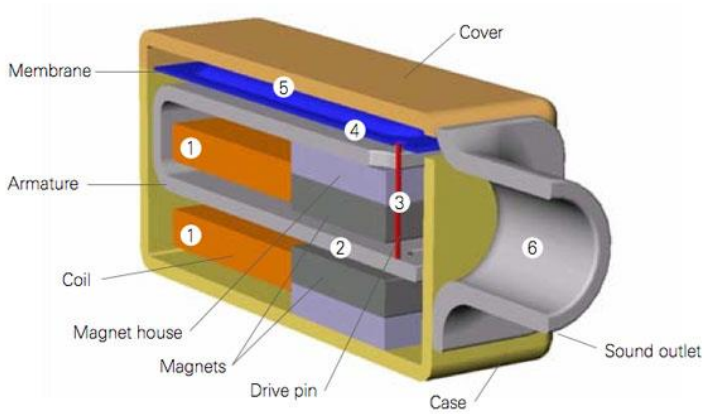


Figure 2: Internal structure of a typical balanced armature driver. Copied from Knowles Acoustics [2]

This structure enables the manufacture of very small drivers. However, they do often exhibit a fairly narrow frequency response band, excelling mainly in the mid-range of audio frequencies (roughly 200Hz-3000Hz). It is essentially a physical impossibility for a single driver to cover the entire audible spectrum of 20 Hz-20 kHz evenly, thus it is often necessary to combine two or more drivers into an earphone to achieve a suitable wide-band output for accurate music reproduction. Because simply combining several drivers in parallel or series would result in an extremely congested mid-range, a need for some type of crossover between these drivers is required, to filter out overlapping frequencies.

2.3 Crossovers

Crossovers combine several electrical filters, creating a wideband frequency response with minimal overlap from combining multiple audio drivers into one system. An ideal crossover would allow each driver to reproduce a certain frequency range and completely block them from outputting at all other frequencies. In practice this is never achievable.

The traditional method for crossovers in IEMs is to use passive components to create simple first or second order RC filters. This is done due to the size limitations of the earphone, which restricts the use of more complex crossovers. In fact, often not even true crossovers are used. Even with ear phones containing more than two drivers, band pass filters are rare and designers rely heavily on acoustic filtering to tune the output for a somewhat acceptable results.

There are a number of drawbacks with this method. First order filters have a very slow roll off into the stopband, which often causes overlap between two drivers; they both reproduce some of the same frequencies near the crossover region causing a hump in the response curve. Also the impedance of balanced armatures is not linear with frequency (see figure 5 in section 2.4.4), creating uneven loading of the filter stages. With passive filters this means their calculated cut-off points will vary depending on frequency, causing an uneven filter response. Since passive components are then load dependent, it is necessary to use high value reactive components that have a high time delay. This causes a deterioration of the circuit impulse response. First order filters also introduce a phase shift of 90 degrees which is detrimental to sound quality.

These problems can be partly overcome with increasing the order of the filters to get a steeper cut off curve, and a phase shift of 180, which can be corrected by reversing the wiring of one of the drivers. However, adding passive components in front of the drivers will reduce its sensitivity potentially causing problems with portable audio sources that have a relatively low output voltage. Figure 3 illustrates a simplified block diagram of the placement of commonly used crossovers in in-ear monitors.

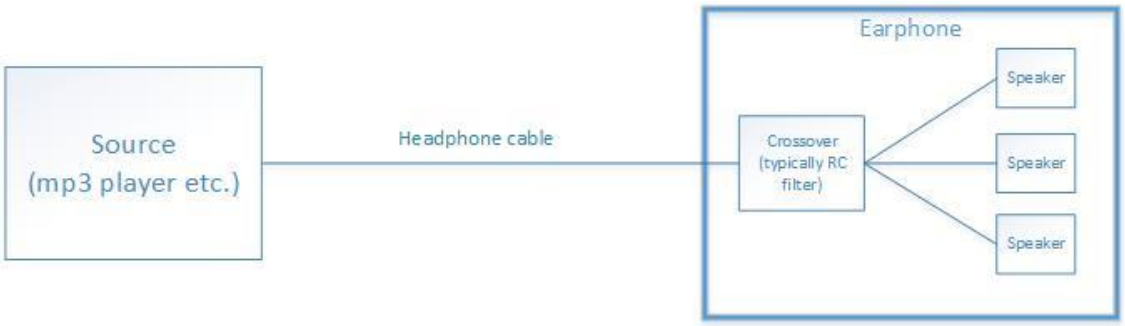


Figure3: Block diagram traditional filter placement inside earphone

Active filters can solve most of these problems. The use of operational amplifiers (or op amps for short) as buffers will isolate the filters from the varying impedance in the drivers, as op amps have practically constant input and output impedances in the frequencies of interest, when designed correctly. Also since active filters can be designed for a specified gain, there is no sensitivity loss from the filter components, even with higher order filters.

The main design problem lies in applying active filters to small portable earphones. The complexity of the circuit and the need for a power source means that it has to be implemented externally. The crossover circuitry is moved out of the earphone and into a separately housed unit that the earphones then plug into, hence the topic of this thesis. When the crossover components are entirely removed from the earphone, this leaves only the drivers inside. Each driver is connected to the external crossover via its own conductor, meaning a headphone cable with seven wires is needed for two earphones (three signals per earphone and signal return). This is also the reason why the earphones must specifically be manufactured for this system and the crossover cannot be universally. The crossover system also contains a six channel unity gain amplifier to drive the output loads. This proposed system block diagram is illustrated in figure 4 below:

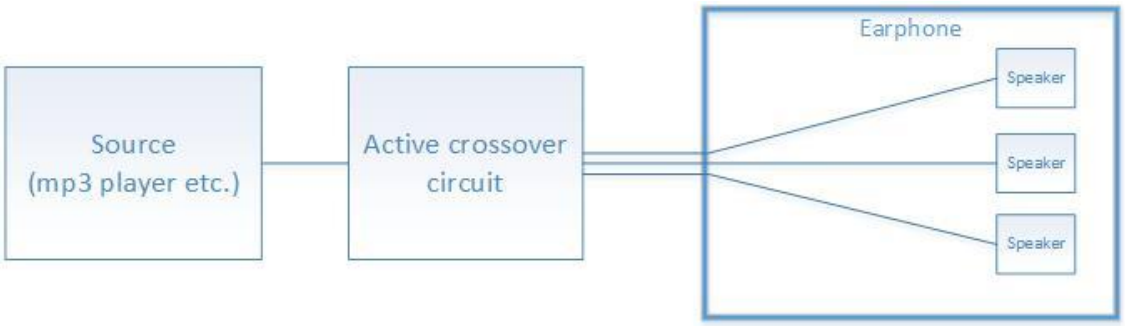


Figure 4: Block diagram of proposed external filter circuit

With the vast majority of IEMs being produced with internal RC filters, there have been only two prior attempts to create the type of external active crossover system proposed in this thesis. However, numerous user opinions indicate serious flaws in the design of these products, and they have since been discontinued. Problems included audible background noise, poor volume control adjustable range, loud turn on “pops” through the earphones (caused by circuit turn on transients), poorly adjusted frequency outputs (especially bass levels deemed unreasonably high) and a fairly low battery life of about 8 hours. These are unacceptable traits in high end audio products, especially considering that they cost well over 1000 €. Overcoming these issues and creating a product with none of these problems is the primary goal of this thesis.

Going through the specifications of these two active crossover systems, it was evident that both were attempted with digital filtering techniques.

The approach in this paper is quite different, using analog active filters to achieve overall better performance. The reason behind this choice is that most portable sources, though digital in data storage, have internal digital to analog converters (DAC) and output audio in analog format. Thus, it seems counter intuitive to use digital filters where the audio signal would first be converted to analog in the source, then back to digital in the crossover, and again back to analog at the output of said crossover. The differences between analog and digital filtering will not be discussed in detail in this paper.

2.4 Design Specifications

The first part of designing a product is to specify all design features and capabilities that need be accomplished.

2.4.1 Output Power

The typical sensitivity to 1mW for balanced armatures is between 96dB SPL and 123dB SPL (decibels to sound pressure level relative to 20µPA) [3]. A sound pressure level of above 123 dB is already considered very loud and not suitable for continuous listening. Thus, it can be concluded that rarely is there a need for a lot over 1mW of power per output channel in this application.

2.4.2 Battery Life

Battery life is an important specification for the general feasibility of the design. If the product performs well while only achieving a few hours of playback it would be virtually unusable as a commercial product. Since there are no set guidelines on what the battery life should be, the run time of an existing commercial product was used as the number to beat. This is would mean a run time greater than 8 hours [4].

2.4.3 Noise

Due to the high sensitivity mentioned above, low output noise is also crucial since balanced armature very easily pick up even the lowest of noise levels. The design goal here is a completely “black background”, meaning that no noise is audible while no audio signal is playing. To demonstrate with an example, one of the balanced armatures used in this design has an SPL of 106 dB at 100mV. A quiet room has a background noise SPL of roughly 20 dB and it would serve as decibel value to stay below in order to make the noise of the circuit indistinguishable from background noise. Putting these values together:

$$20\log\frac{100mV}{V_{noise}} \leq 106dB - 20dB \tag{1}$$

And rearranging to solve V noise:

$$Vnoise \leq \frac{100mV}{\left(10^{(86dB/20)}\right)} \tag{2}$$

$$\rightarrow Vnoise \leq\ 5\mu V$$

This means a noise voltage of less than 5µV is required for the generated noise to be below 20dB SPL. The reality is more complex than that, as the human ear does not perceive all frequencies equally, but it does give a ballpark figure aim for.

In order to effectively reduce noise, the different types of noise sources must be identified. These are:

1/f Noise

1/f noise is inversely proportional to frequency, meaning it rises as frequency falls. Because of this spectral distribution, noise in most circuits tends to be higher at lower frequencies. This type of noise appears in active and passive components, depending on their construction. The important value here to look for is the corner frequency, the lower being always better, as this would ideally keep the 1/f noise below audible frequencies.

Johnson Noise or Thermal Noise

The primary source of Johnson noise in circuits are resistors, as it depends on temperature and resistance of a component. All resistors generate Johnson noise and it can be calculated as a noise voltage source:

$$V_{NR} = \sqrt{(4kBTR)} \tag{3}$$

Where k is Boltzmann's Constant (1.38×10–23J/K), T is the absolute temperature, B is the bandwidth, and R is the resistance.

Example:

T = Absolute Temperature = T(°C) + 273.15

B = Bandwidth (Hz)

k = Boltzmann’s Constant (1.38 x 10–23J/K)

A 1000Ωresistor generates 4nV / √Hz @ 25°C [5]

Op Amp Current & Voltage Noise

All op amps exhibit varying degrees of noise. These can be determined from data sheets and are generally divided into **noise voltage** and **noise current** values. The noise voltage is expressed in the same units as resistor noise (Johnson noise) and can be summed accordingly. Noise current, however, is heavily dependent on the resistor values surrounding it. As ohms law states, voltage is current times resistance. Thus, the benefits of low noise op amps can be completely undone by combining them with high value input resistors.

Determining The Dominant Noise Source

In complex systems it is beneficial to identify the dominant cause of internal noise. RMS noise voltages are summed by root sum squaring:

$$V_{noise} = \sqrt{(V_{n1}{}^2 + V_{n2}{}^2 + V_{n3}{}^2)} \qquad (4)$$

Because of this relationship, the noise voltage source that is three to five times higher can be considered dominant and the rest of the noise source can be ignored to simplify calculations. Below in figure 4 is an example of this relationship demonstrated with a simple op amp circuit.

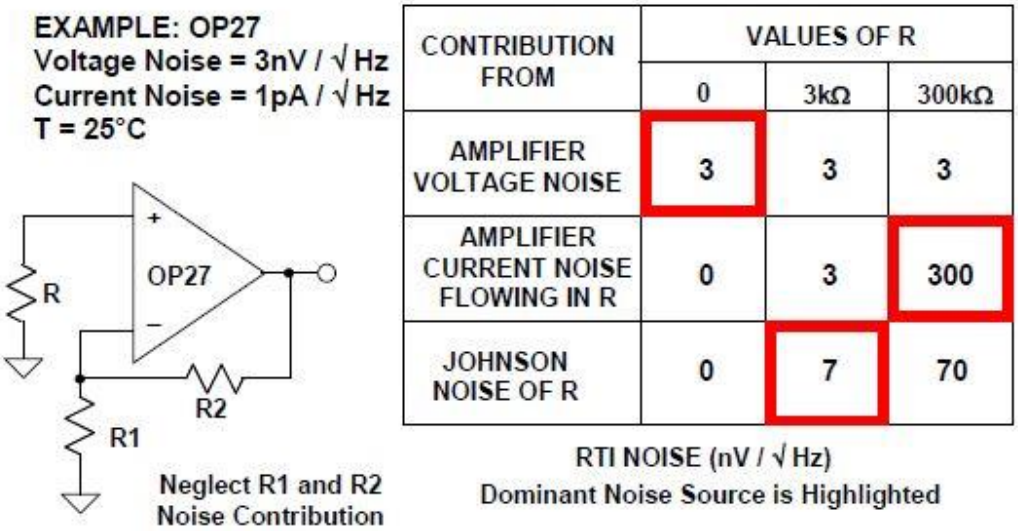


Figure 4: Example of dominant noise source in an op amp circuit. Copied from Analog Devices (2008) [5]

What this shows is that it is ultimately the choice of resistors that dictates the overall noise, depending on the resistance values used, not only because of the Johnson noise, but also due to the op amp current noise relationship. [5]

EMI

Electromagnetic interference can also introduce noise into a circuit, with the exception that here it is considered as coming from an external source. It is equally, if not more important to limit susceptibility to external noise sources than it is to design for low internal noise. The methods used for reducing this susceptibility are [6, 26]:

1. Shielding – Refers to using a grounded enclosure to minimize outside noise coupling.
2. Filtering – Filter components used in signal or power path either globally or locally, to block already coupled noise on traces or interconnects from entering sensitive circuits.

3. Grounding – Solid ground planes and short ground traces used to reduce common impedance coupling of noise and isolation of sensitive circuits from noisier ones.

Total Harmonic Distortion or THD

This is typical specification for audio applications as it is often used as one of the main parameters to evaluate the quality of audio equipment. Distortion is nonlinearity of the signal, with harmonic distortion being the frequency multipliers of an input signal. The measured value is the ratio of the harmonic amplitude to the input signal amplitude, quoted as decibels or a percentage. There is no clear consensus what level of distortion is audible. Some sources state 1%, while others as state that as low as 0.1% can be distinguished. And also the order of the harmonic distortion matters, as well as whether or not one is listening to pure sine waves or actual music [7, 9]. The target for this project will be a distortion figure as low as possible, definitely below 1%, preferably below 0.1%.

2.4.4 Output Impedance

A low output impedance is important in maintain a high damping factor. This is because the output impedance together with load impedance forms a voltage divider reducing the available output voltage swing. To make matters worse the impedance of balanced armatures is never constant, but varies with frequency. Thus, the voltage divider also varies with frequency, causing a non-flat frequency response.

$$20\log\left(\frac{20\Omega}{20\Omega+5\Omega}\right) - 20\log\left(\frac{115\Omega}{115\Omega+5\Omega}\right) = 1.49dB \tag{5}$$

The impedance varies between 20Ω and 115Ω for frequencies of interest. If we had an output impedance of >5Ω, it would create a varying voltage divider of the range of 1.49dBV. Clearly not a linear output. For example, in figure 5 below it the impedance variations of a balanced armature can be seen.

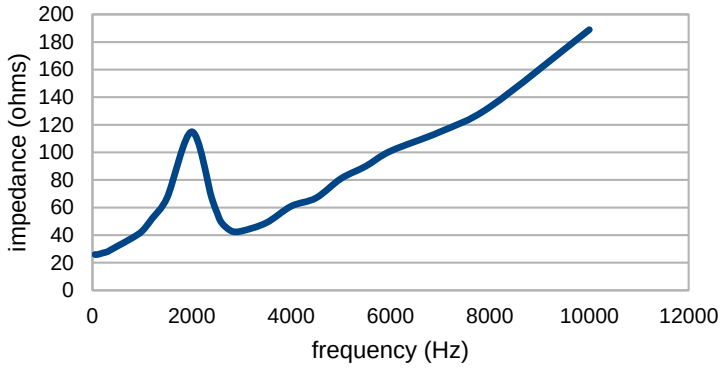


Figure 5: Example impedance curve of typical IEM

To further demonstrate the effect of varying series resistance connected to a balance armature, figure 6 shows a graph from a manufacturer of balanced armatures, which very clearly show the difference even a small amount of resistance can make.

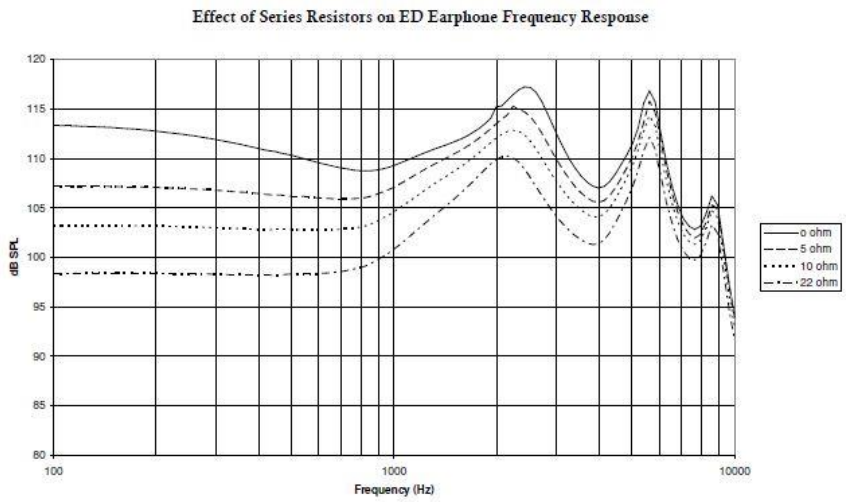


Figure 6: Effect of series resistance in one type of balanced armature. Copied from Wickstrom [8]

A difference of around 5 dB in SPL on the low frequency response with only 5 Ω of resistance added in series with the balanced armature. All this means that the circuit should have an output impedance as close to zero as possible.

2.4.5 Low DC Offset

To prevent damage to headphones, typically almost no DC voltage should be present at the output. This becomes even more crucial with highly sensitive balanced armatures. Any excess DC voltage becomes wasted output power and is summed with any signal

fed to the balanced armature increasing the possibility to overload them with excess power. DC voltage can be blocked at the outputs with capacitors. However, due to the amount of outputs (three per channel, six in total), and the low impedance of the connected loads (some balanced armatures can be lower than 10 Ω at some frequencies), these capacitors would have to be unreasonably large considering the limited circuit board space available.

Because of this the rest of the circuit must be designed with a low dc offset in mind, most obviously by selecting low offset op amps and taking into account the input bias currents that together with op amp input resistors generate offset voltages.

3 Filter Design

In this section the calculation and design of basic filters is covered and the available options for implementing the crossover circuit is explored. First some basic terminology used with regard to filters.

3.1 Terms

Transfer function and frequency response

The transfer function H(s), is the Laplace transform of a system's impulse response. It is the systems output divided by the input in the S domain and it is a description of how a system responds to an input signal. It can be used as the basis for designing any type of filter.

The frequency response is a representation of the transfer function in the frequency domain H(jω), giving the magnitude and phase responses with respect to frequency. A typical method for displaying frequency response is the bode plot.

Common transfer functions for filters are:

High-pass filter – Attenuates frequencies below its cut off frequency (fc), allowing high frequency signals to pass

Low-pass filter – Attenuates frequencies above its cut off frequency, allowing low frequency signals to pass

Band-pass filter – Is a combination of high- and low pass filters, having two cut off frequencies, attenuating both above and below the frequency band it allows to pass.

Notch (or band stop) filter– Is the inverse of a bandpass and attenuates the frequency band between its cut off points.

These ideal transfer functions are illustrated in figure 7. These are the types of filters we would like to create, but in reality are not possible to achieve.

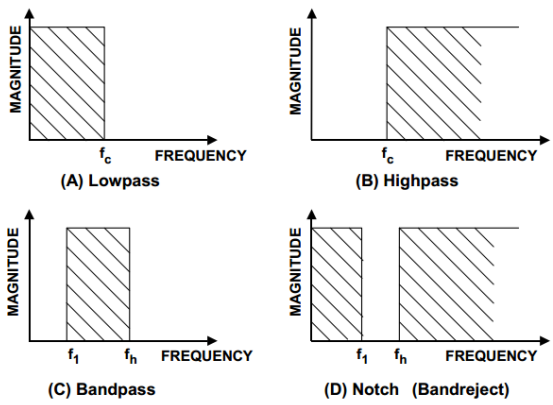


Figure 7: Ideal responses of different filter types. Copied from Analog Devices [9]

Common filter response parameters are described as follows:

Pass Band

The passband of a filter is the frequency range where the output is above the cut-off point (fc or -3 dB in most cases) and can be described as the part of the filter that is passed through with minimal attenuation. In an ideal filter the attenuation at the passband is zero.

Stop Band

The stopband of a filter is the frequency range defined by the stopband corner frequencies where the attenuation is at least the specified level. In an ideal filter, this attenuation is infinite, meaning no amount of the input signal is passed through.

Transition band

The transition band is defined by the area between the pass- and stop band corner frequencies. It is essentially as slope, which is defined by the steepness of the attenuation. In an ideal filter, the transition from passband to stopband is instantaneous and the transition band would not exist.

Ripple

Ripple is variations in signal amplitude and can exist in both pass bands and stop bands. It can be defined as Amax – A min in a given frequency band. In an ideal filter no ripple exists and the amplitude is constant.

A visual representation of these parameters can be seen in figure 8.

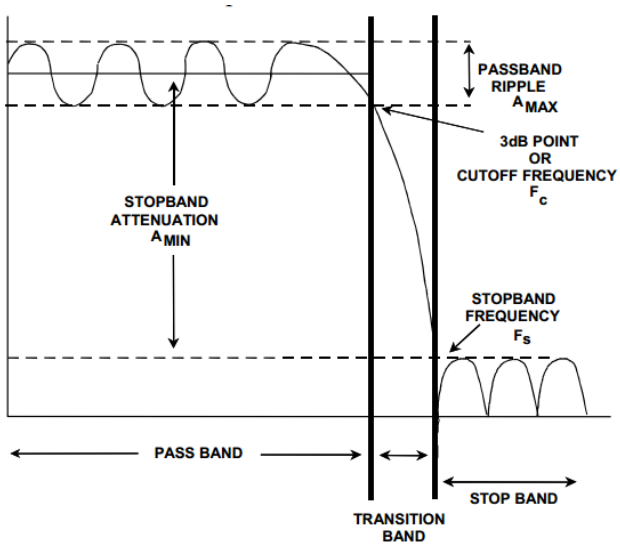


Figure 8: Key filter parameters. Copied from Analog Devices. [9]

Quality factor or Damping ratio

Describes the damping of oscillations in resonant circuits the quality factor can be defined with the function:

$$Q = \omega \frac{energy\ stored}{energy\ dissipated} \tag{6}$$

In practice, filters with high Q have steeper initial roll off near the cut-off point, but have higher overshoots or ringing, causing ripple in the pass- and stopbands. Quality factor is often divided into the following categories:

Over damped (Q < 0.707) – This system does not oscillate, but suffers from slow response.

Critically damped (Q = 0.707) – This system does not oscillate and results in the fastest approach to a steady state as possible without oscillations.

Under damped (Q > 0.707) – This system does oscillate, creating overshoots in transient response, but also achieves a faster response than the overdamped or critically damped systems. [10]

Poles & Zeroes

Derived from the transfer function of a system, the roots of the numerator are called zeros and the roots of the denominator are called poles. A system can have one or more poles and zeroes. Every pole provides a response of -20dB/decade and every zero +20dB/decade, respectively.

Order

The order of a filter describes the order of the polynomial of the transfer function or the number of poles. Or in more general terms, it determines the eventual maximum steepness of the attenuation after initial “knee” of the curve [11, 264]. Several stages of a filter can be cascaded for a steeper attenuation. However, as each successive stage loads the previous stages output, the Q values for each stage will change and must be re calculated when adding stages. This allows for quicker roll off of the attenuated signal while maintaining the desired filter characteristics. Each successive increase in filter order increases the sharpness of the attenuation. For example a first order filter attenuated at a rate of 20dB/decade or (6dB/octave as is commonly used in audio applications), then a second order filter is attenuated 40dB/decade, a third order filter 60dB/decade and so forth.

3.2 Designing a Filter

From the information above it is possible to design many kinds of theoretical filters by specifying a type of filter required (high pass, low pass etc.) and taking a suitable transfer function and transforming it into the response type needed. After that component values can be calculated if a frequency cut off point and Q values are known. It happens however that many types of filter responses have been studied and calculated by various mathematicians over the years, meaning that in practice an engineer can choose one of those types instead of designing a filter from scratch. Thus, it is common to use one of the well-known filter transfer functions in when designing a system, with the selection done based on the desired filter characteristics, depending on system priorities.

3.2.1 Butterworth

Perhaps the most common types of filter, the Butterworth is known as maximally flat, because it is designed for the flattest passband response. The downsides being a less steep transition into the stop band and phase delay characteristics. By increasing the number of poles, the stopband falloff can be steepened. As can be seen in figure 9, the passband of a Butterworth filter has no ripple, which makes it ideal in audio filters.

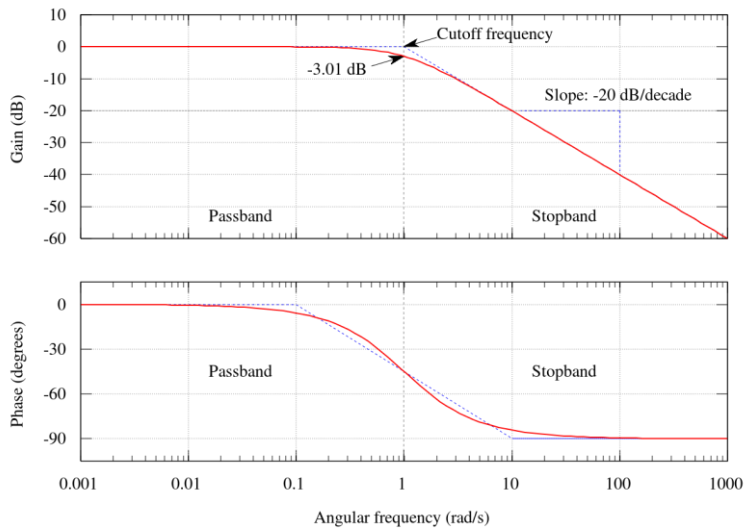
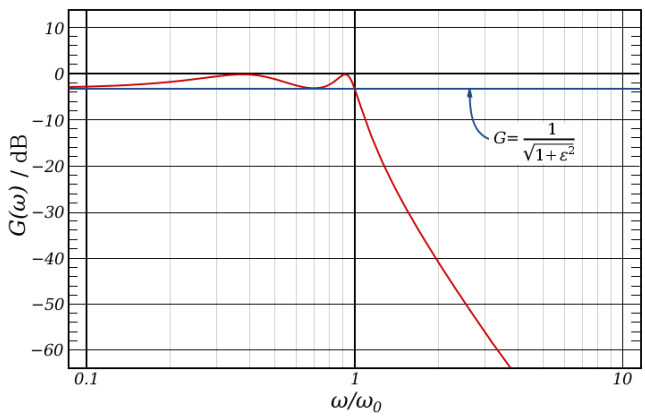


Figure 9: Bode plots of single pole Butterworth filter. Copied from Wikipedia [12]

3.2.2 Chebyshev

Chebyshev filters are characterized by a steeper roll-off than Butterworth filters, with the down side of having more passband (Chebyshev type I) or stopband (Chebyshev type II) ripple. They are generally used when a very steep attenuation is required and ripple in the signal is less important.

A Chebyshev filter of order n will have n-1 peaks or dips in its passband response. Thus, steeper the filter, the more ripple it will have. When designing a Chebyshev filter the nominal ripple must be specified as it affects the passband gain also. The cut off frequency of a Chebyshev filter is not assumed to be the -3 dB point as with the Butterworth filter, but instead, the cut-off point is normally the frequency at which the specified ripple is exceeded. [13]



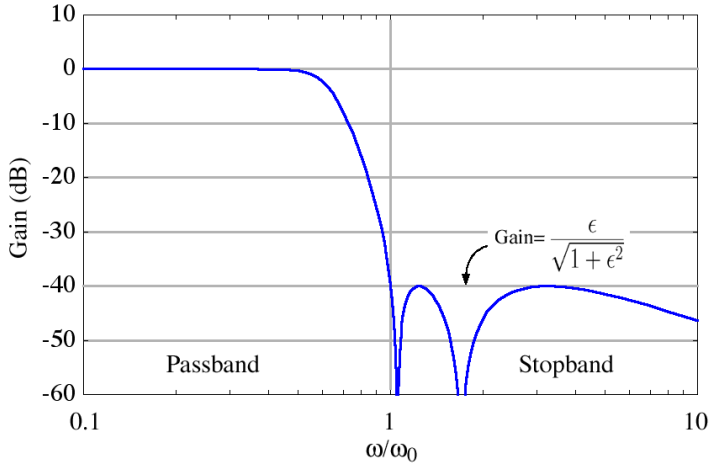


Figure10: Amplitude plots of Chebyshev type I (above) and type II (below) filters. Copied from Wikipedia [14]

However, in this application, the flatness of the output frequency bands are the primary concern, making Chebyshev filters a poor choice in this application. This ripple can be easily seen from figure 10 above. A case could be made for using type II filters as the passband is flat, but almost no information was found on using Chebyshev type II filters for audio, possibly due to poor phase performance, so using it was ruled out as well.

3.2.3 Bessel

The third common type of filter is the Bessel filter. Out of all the types here, it has the slowest roll off curve, but it is specifically designed for maximally flat phase delay, meaning a linear phase response and, thus the best transient response characteristics. This is an important feature in crossover systems as when several signals are summed at the output of the system, any large phase differences between them can certainly become audible. Because of this, the Bessel filter is often used in crossovers. [15]

3.2.4 Linkwitz-Riley

The Linkwitz-Riley infinite impulse response filter was developed to overcome the problems caused by these prior filter types in crossover systems. It is based on the Butterworth filter and is also called Butterworth squared [16].

It consists of two cascaded Butterworth filters, each stage with a cut-off point of -3dB. This creates a total of -6dB at the cut-off fc, that when summed, the low pass and high

pass filter have a gain of 0dB at the crossover point, resulting in a flat passband throughout. Compared to a Butterworth, the Linkwitz-Riley filter also has a linear phase response on the passband. Since one second order Linkwitz-Riley section has a phase difference of 180° between high and low pass signals, the cascaded fourth order filter appears fully in phase with a 360° phase delay. This results in an overall zero phase difference between all output signals. [17] The difference between Butterworth and Linkwitz-Riley amplitudes at the crossover point is illustrated in figure 11.

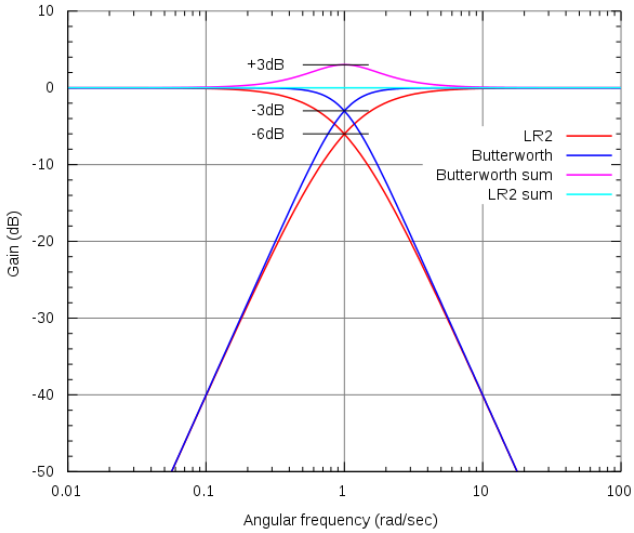


Figure 11: Comparison of Linkwitz-Riley filters and Butterworth filters in crossovers. Copied from Wikipedia [16]

Because of these properties a fourth order Linkwitz-Riley type filter was used for this circuit. The calculations required for designing a Linkwitz-Riley filter are as follows [18]:

Low pass:

$$Q = \frac{1}{2}\sqrt{(C_1/C_2)} \tag{7}$$

$$f_0 = \frac{1}{2\pi * R * \sqrt{(C_1 * C_2)}} \tag{8}$$

$$R = \frac{1}{2\pi * 2Q * f_0 * C_2} \tag{9}$$

$$C_1 = 4 * Q^2 * C_2 \tag{10}$$

And similarly for high pass:

$$Q = \frac{1}{2}\sqrt{(R_2/R_1)} \tag{11}$$

$$f_0 = \frac{1}{2\pi * C\sqrt{(R_1 * R_2)}} \tag{12}$$

$$R = \frac{1}{2\pi * 2Q * f_0 * C} \tag{13}$$

$$C = \frac{2Q}{2\pi * f_0 * R_2} \tag{14}$$

These equations together with the follow Q values in table 1 allow the design of any order of Linkwitz-Riley filter.

Table1: Q values for Different-Order Linkwitz Riley Filters. Copied from Linkwitz [18]

	LR2	**LR4**	**LR6**	**LR8**	**LR10**
Q_0 of stage 1	0.5	0.71	0.5	0.54	0.5
Q_0 of stage 2	-	0.71	1.0	1.34	0.62
Q_0 of stage 3	-	-	1.0	0.54	1.62
Q_0 of stage 4	-	-	-	1.34	0.62
Q_0 of stage 5	-	-	-	-	1.62
dB/octave slope	12	24	36	48	60

3.3 Active Filter Topologies

There are several different ways of implementing the chosen filter response with active circuits. This section explores a few of the most common active filter topologies. As with response types of the previous section, each has its own benefits and drawbacks, with no clear all around best type.

3.3.1 Sallen & Key

Possibly the most common active filter layout is the Sallen & Key. They are simple to design and implement, with the fewest possible components. A good choice for simple circuits.

The drawbacks include, difficult tunability as two pairs of passive components must be closely matched, and high bandwidth requirements for the op amps used, because as the frequency rises, so does the effective output impedance. This changes the way the feedback component (R and 2C in figure 12 below) behaves. With low pass filters the attenuation bottoms out at some frequency, before starting to rise again due to the added impedance in the feedback loop [19, 141]. A Sallen & Key implementation of a two way crossover can be seen in figure 12.

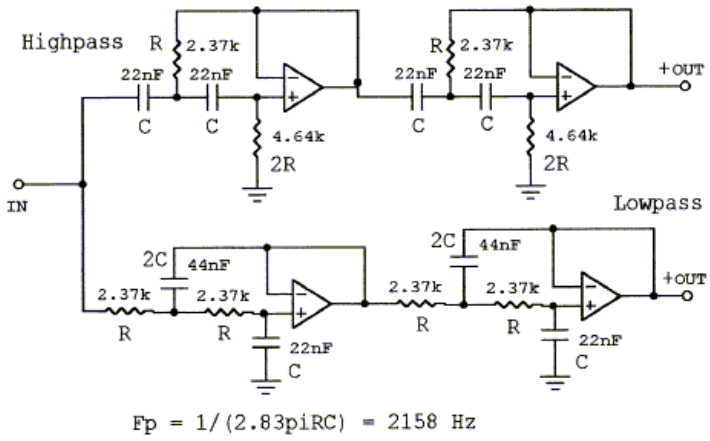


Figure 12: Sallen & Key layout of a two-way crossover (high- & low pass). Copied from Linkwitz [18]

3.3.2 State Variable

Another type of filter implementation considered here is the state variable. It has the benefit of providing easy tunability, as only one resistor controls the Q and only one passive component needs to be changed for changing the frequency cut off on each second order stage. This also means smaller passive component count. Figure 13 shows a Linkwitz-Riley implementation of a state variable topology.

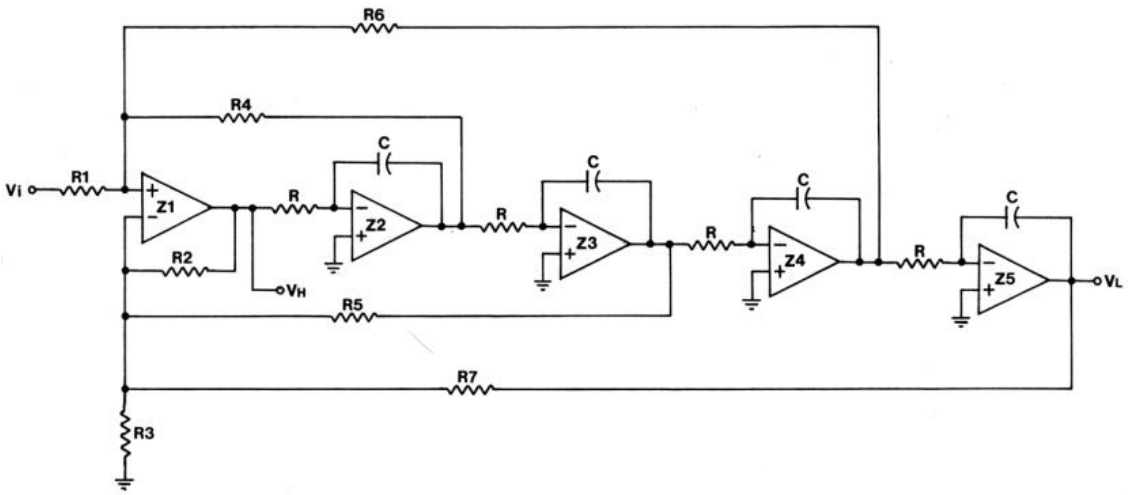


Figure 13: 4th order Linkwitz-Riley 2 way crossover (high- & low pass). Copied from Bohn [20]

The drawbacks include increased complexity of the feedback circuits and an additional active component required per channel. These drawbacks make the PCB design significantly more troublesome with the small board size used. That is the reason why the state variable design was abandoned for this project, after some attempts at a PCB layout, in favor of Sallen & Key for this design

4 Schematic Design

With the core elements (that is the filters) now specified, it is possible to design the entire schematic around them. Starting off with a general block diagram of the entire circuit, each stage is then shown and explained in more detail. The schematic and successive PCB layout (detailed in chapter 5) were done with Cadsoft Eagle. The full schematics can be seen in appendix 1.

4.1 Block Diagram

The following block diagram illustrated in figure 14 is the planned layout of the core functionalities of the signal path of the circuit, starting from the input through the filters and terminating to the outputs to be connected to the in-ear monitors. Only one channel is shown here and it must be duplicated for stereo operation.

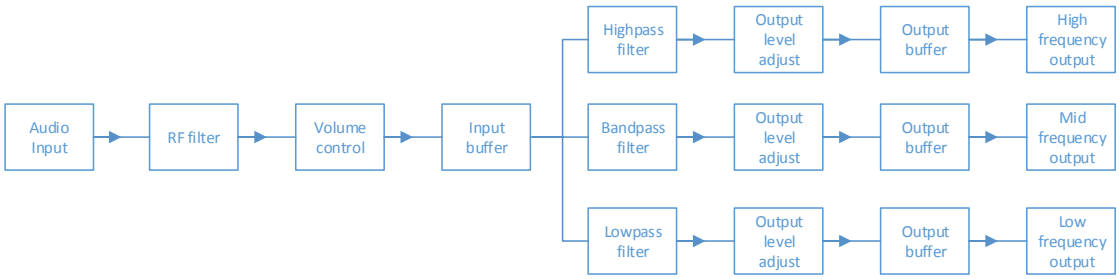


Figure 14: Block diagram of the crossover circuit (one channel shown)

4.2 Input

The input section contains a common 3.5mm headphone jack followed immediately by RF filtering components: a series ferrite and a shunt capacitor. This forms a radio frequency filter to block RF interference coupling into the circuit through the input interconnect cable. It is followed by DC blocking capacitors, a volume control potentiometer and input buffer. Because this circuit is designed to be connected to a wide array of audio sources, some of which can have harmful DC offset voltage at their output, the DC blocking capacitors are there to ensure no DC offset voltage enters the circuit from the source. The capacitor values must be chosen so that the high pass filter it forms together with the potentiometer resistance parallel with the input buffer resistor (10kΩ) gives a corner frequency well below 20Hz to pass all of the audible frequency range. In the worst case of the volume pot turned fully up we have roughly 5kΩ across the input to ground. In this case we calculate it as:

$$C > \frac{1}{2\pi * 20Hz * 5k\Omega}[F] \tag{15}$$

This tells us that a minimum of 1.6µF is required.

The input buffer is simply a unity gain operational amplifier to provide a high impedance input to be seen by the source equipment to ensure the circuit can be drive from virtually all audio sources. Figure 15 illustrates this input buffer.

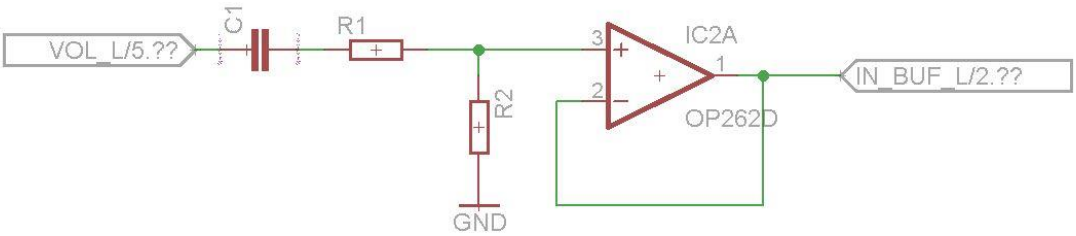


Figure 15: Input buffer

The output of the input buffer is then divided into the various filters that make up the main part of the circuit.

4.3 High Pass Filter

The high pass filter provides frequencies above a set corner frequency to be fed into to the tweeter element (transducer with good high frequency response) of the in-ear monitors. It is a Sallen & Key active filter topology with two second order filters combined to create a fourth order filter with 24dB/octave attenuation. The high pass section can be seen in figure 16.

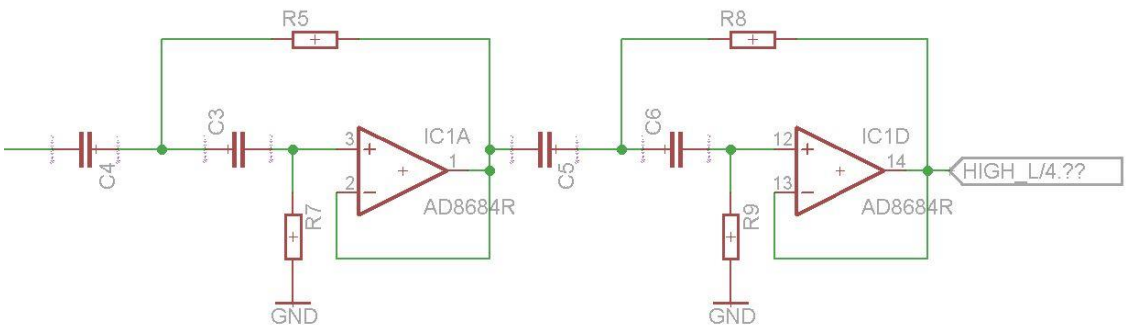


Figure 16: Fourth order high pass filter

From the Linkwitz-Riley equations in the previous section, the equations of R and C can be derived.

$$Q = \frac{1}{2}\sqrt{(R_7/R_5)} \tag{16}$$

Since Q = 0.707 (Table 1):

$$\frac{R_5}{R_7} = 0.5 \tag{17}$$

All capacitor values are equal, and the values of R and C are:

$$R_7, R_9 = \frac{1}{2\pi * 2Q * f_0 * C} \tag{18}$$

$$C = \frac{1}{2\pi * 2Q * f_0 * R_5} \tag{19}$$

Where R5 and R8 are twice the value of R7 and R9.

Now we are free to choose either a resistance or capacitance value and calculate the other for whatever frequency is needed. In practice it is probably easier to choose a capacitance first since they are available in fewer components values, often following the E24 table (1, 2.2, 3.3, 4.7 10 etc.)

4.4 Low Pass Filter

The low pass filter provides frequencies above a set corner frequency to be fed into to the woofer element (driver with good low frequency response) of the in-ear monitors. It is of the same Sallen-Key design as the high pass filter but of a low pass type. The input of the low pass filter is taken from the output of the band pass filters low pass section, instead of directly from the input buffer. This is done to keep the correct phase alignment between the different output signals [21]. Figure 17 shows the low pass filter.

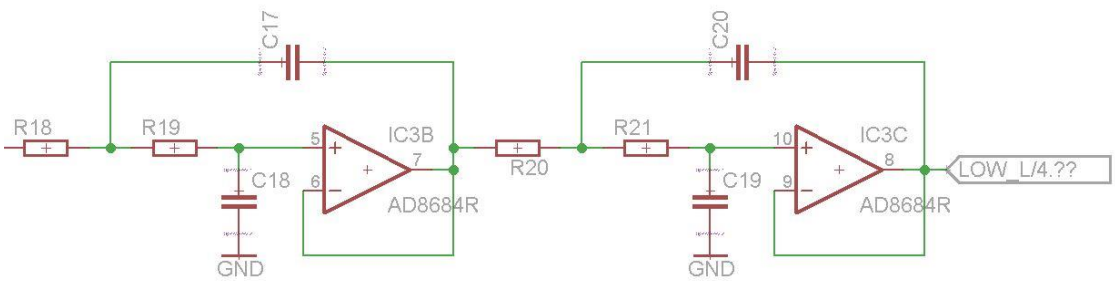


Figure 17: Fourth order low pass filter

The R and C values are calculated in a similar way than the high pass, with the exception that this time the ration of the capacitors determine the Q value:

$$Q = \frac{1}{2}\sqrt{(C_{17}/C_{18})} \tag{20}$$

And again Q = 0.707 (Table 1):

$$\frac{C_{18}}{C_{17}} = 0.5 \tag{21}$$

And the equations for the components are (all resistors are equal valued):

$$R = \frac{1}{2\pi * 2Q * f_0 * C_{18}} \tag{22}$$

$$C_{18}, C_{19} = \frac{1}{2\pi * 2Q * f_0 * R} \tag{23}$$

Where C17 and C20 are twice C18 and C19. Again it should be easier to start by selecting capacitors first and then calculating the needed resistance.

4.5 Band Pass Filter

The band pass filter is simply a combined set of high- and low pass filters described previously to provide the frequency range between the other two outputs (high and low outputs). The band pass filter is set so that its upper corner frequency(low passed out) equals the corner frequency of the high pass filter described above (in section 4.3) and its lower corner frequency (high passed output) equals the low pass filter's (4.4) corner frequency. This gives a total of three different outputs with two crossover frequencies in between them and should give a summed flat frequency response in total. Only one channel was described here. A second identical set of filters is required for stereo operation.

In appendix 2, all possible crossover frequencies are calculated with available chip components and limitations of those values are discussed more in the component selection chapter. To adapt this design to different frequency cut-off points, one can simply choose from that table.

4.6 Output

Each of the six outputs consists of one op amp set to unity gain to buffer the individual filter outputs so that they “see” a constant high impedance load and thus are not affected by the impedance variations of the balanced armature loads and with the exception of the input voltage divider value, all six outputs are identical. This voltage divider at the input of each op amp (resistors R22 and R23 in the picture above), is there to provide the ability to individually adjust the level of each output. This is important as the different balanced armatures have varying sensitivities and some need to be attenuated more than others. The output buffer can be seen in figure 18.

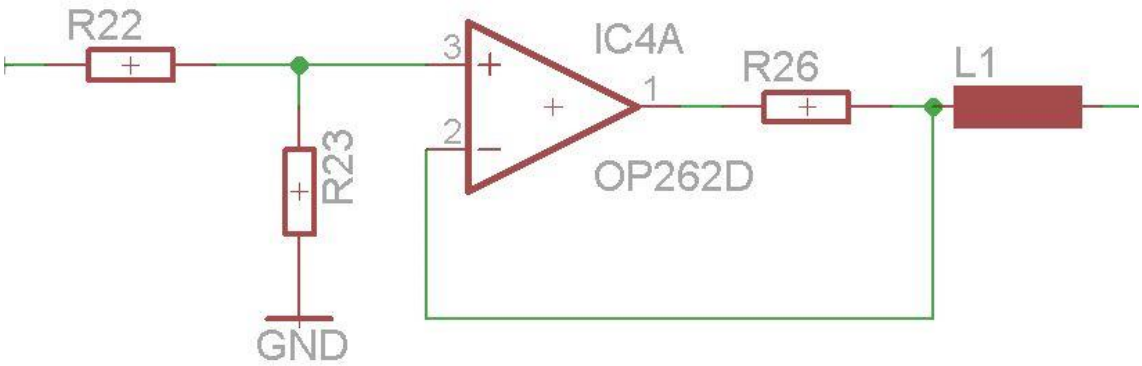


Figure 18: Output buffer

There is also a small value resistor on the output of the op amp to provide short circuit protection when the earphones are connected or disconnected. When disconnecting the headphone connectors, the audio jacks can be briefly short circuited. The resistor should be placed inside the feedback loop to keep it from raising the effective output impedance [22]. This resistor also serves to help stabilize the op amp in case of capacitive loads, by increasing its phase margin by a small amount [23].

A ferrite bead at the output (L1 in the image above) was meant for further helping the op amps deal with high load capacitances [22], however proto board testing showed no discernible benefit in using ferrites in this part of the design, so the ferrites were discarded from the circuit.

4.7 Power supply

Power is provided by a 3.7V lithium-ion battery. Since batteries have rather clean voltage output with very little voltage ripple or noise, it was deemed unnecessary to provide additional voltage regulation to its output. A 220µF reservoir capacitor with a small bypass capacitor is placed parallel to the battery to provide for instantaneous power draw in order to maintain adequate supply at all times.

Operational amplifiers generally need a positive and negative voltage supply in audio applications to provide an output voltage swing that is centered on half the total supply voltage. To provide for this dual supply from a single battery, a virtual ground was created for the circuit. This means using a reference voltage of exactly half the supply voltage for the signal path.

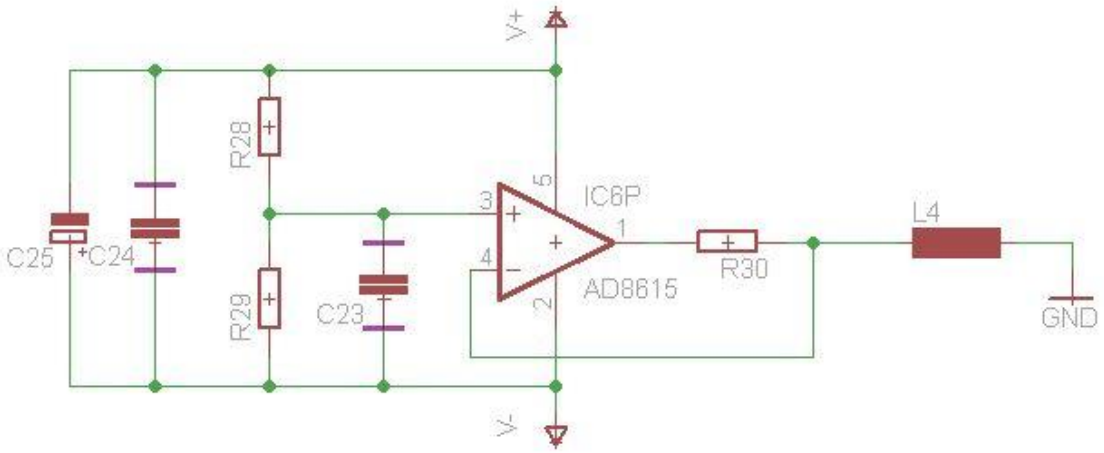


Figure 19: Virtual ground divider

Virtual ground circuits can be implemented simply by using a resistive divider in the supply to create a reference of VCC/2. In this circuit the resistive divider is also buffered with an operational amplifier to ensure reference point stability and a low impedance ground and is illustrated in figure 19. A capacitor is placed parallel to one of the dividing resistors (C23 in figure 19), providing a low impedance path for AC to ground, so any voltage ripple will not shift the reference point and no resistor noise will enter [24].

When selecting this op amp it is important to check that it is able to sink the current of the entire load to be connected, thus the entire circuit and connected loads, since all load return currents flow through this circuit.

Battery charging is provided with a Microchip MCP73831 lithium ion charge controller. It is specified for a charging current 450mA with a programming resistor of 2.2kΩ. It is utilized using Microchip's own reference design from the component datasheet. R32 sets the charging current with the following formula:

$$I = \frac{1000V}{R_{32}} = \frac{1000V}{2.2k\,\Omega} = 454mA \tag{24}$$

4.7uF electrolytic capacitors are placed at the input and output of the charge controller to ensure stability, as the reference suggests. A LED in series with a 470Ω resistor provides a charging indicator. The LED glows while charging and is turned off once the battery charging is complete. The charging circuit is shown in figure 20.

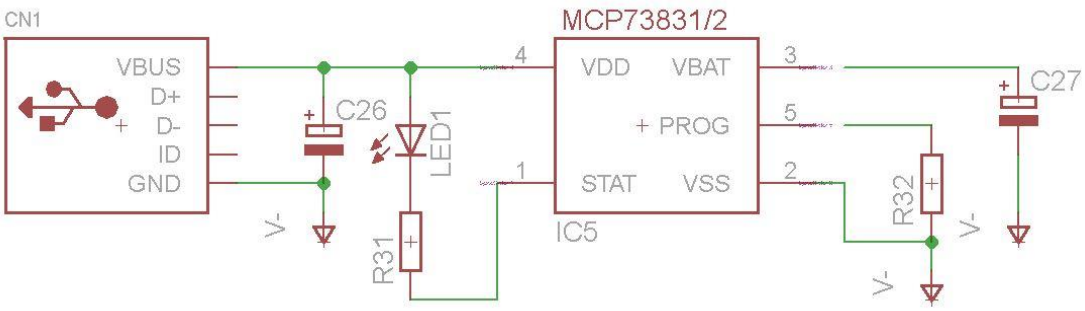


Figure 20: Battery charging circuit

4.8 Simulation

Before commencing with the actual PCB design, this schematic was modelled in National Instruments' Multisim software and simulated to check crossover functionality. This was done to minimize any mistakes prior to spending numerous hours on the PCB layout. It effectively limits any problems encountered to real world effects such as component tolerances and parasitic effects of layout traces, external noise sources and generally, any unplanned issues. As can be seen from the simulated bode plots in appendix 3, the circuit does work at least in theory as specified.

5 PCB Design

Chapter 5 describes the process and considerations made during the layout process of the circuit board. As with the schematic, it was also made using Cadsoft Eagle software. Numerous revisions of the circuit board layout were eventually made, not all of which will be covered in this paper, and only the differences of the ones manufactured will be discussed in the next chapter. This chapter will go through the common design aspects of all the revisions. All PCB board designs can be found in appendix 4.

5.1 Circuit Board Size

Since this is meant to be a portable design, small form factor must be attained. Small is an arbitrary quantity, so the following criteria was applied. Length and width of the device should not greatly exceed the size of an average smart phone or mp3 player. Additionally the width should only be just enough to fit the necessary panel components (output jacks and volume pot being the largest components on the front, and thus dictate the minimum width). The height should be only enough to accommodate the highest internal component, which in this case is the volume potentiometer (11.35 mm).

Browsing through commonly available electronics enclosures, a Hammond 1455xx series enclosure was the most suitable choice due to its dimensions and availability. It has internal slots sliding in and holding in place a circuit board with the dimensions of 80 mm x 50 mm. This sets the design goal for the board size.

5.2 General Layout Practices

Whenever possible, good layout practices where attempted to uphold when designing the PCB. Compromises had to be made along the way; this section details these layout practices and whether or not it was feasible to implement them.

5.2.1 Ground Planes

A solid ground plane is the preferred design method mentioned in many works [9], [25]. However due to the high component count and the manufacturing of prototypes on two layer boards, this is not at all possible to implement here. It is attempted to connect grounds at several points with vias in order to achieve some semblance of ground plane with as many low impedance return current paths as possible. This is to reduce common impedance coupling in the ground path (as mentioned in the design specifications) [6].

5.2.2 Short Traces

Minimizing impedance and thus coupling from other traces and noise sources is important so trace lengths and thus inductive loop sizes should be kept as small as possible. This was the main factor considered when placing components on the board: certain parts like input and output connectors and volume control had to be placed in certain logical spots for usability, so the rest of the components were placed in a manner to have the signal go from A to B in as short a route as possible. For example input buffers are placed close to the volume control and output buffers are close to the output connectors.

5.2.3 Power Rail Decoupling

All operational amplifiers use decoupling capacitors placed as close as possible to their power pins. Since this is a single supply design only a single decoupling capacitor is used for each device from the power rail to ground. No definite research results were found on whether or not the virtual ground should also be decoupled to the positive supply and ground, so due to space limitations this option was not explored. Also, in order to save board space some op amps are placed directly opposite each other (one on the top layer, other on the bottom) in order to share a decoupling capacitor. This might end up degrading high frequency ripple rejection of the op amps and maybe increasing crosstalk, but compromises have to be made in order to fit the circuitry on the board.

5.2.4 RF Rejection

In addition to attempting a good grounding scheme as mentioned above, also shielding and RF filtering was employed to reject outside noise from coupling into the circuit. The signal input of the design employs a single pole RC filter with a corner frequency in the MHz range (values vary with design version).

Since practical usage experiments showed some susceptibility to cell phone TDMA noise from time to time the location of this RC filter was moved significantly closer to the input connector. Previously any noise conducted on the input signal lines had the opportunity to travel the entire length of the circuit board before reaching the RF filter. This means it is possible that noise was radiated from those long signal traces inside to other parts of the circuit before filtering could take effect.

The supply ground is also connected to the system enclosure, which is made of aluminium. This provides some shielding against outside sources as well. In early versions of the product an enclosure with plastic end panels was used. This degrades shielding effectiveness, so for further versions (version 3 onwards) aluminium end panels will be machined for a more complete shield.

6 Component Selection

With a functional circuit board done, all that is left for a finished product is to populate the board with components. With a seemingly endless variety of different suitable active and passive components, based on different manufacturing methods, technologies and raw material, choosing the optimal parts can be just as impactful to the end result as the schematic design and circuit board layouts. In order to not deteriorate the performance of this design a careful look at all components should made before placing them on the PCB.

6.1 Operational Amplifiers

Selection criteria for operational amplifiers in this application as follows:

Gain-Bandwidth Product should be 40dB or 100 times above your maximum gain at the used bandwidth, when designing active filters with a Sallen & Key configuration [26]. Doing so ensures enough open loop gain for the feedback loop, through the entire bandwidth [27]. The required GBW can be calculated with the following formula:

$$GBW = 100 * 20kHz * Q * Gain \tag{25}$$

Which in our case gives a GBW of 1.42 MHz

Slew Rate refers to an op amps maximum rate of change, meaning its ability to react to quick changes in the signal voltage. It limits maximum usable frequencies and output voltage. In this design the maximum frequency is 20 kHz and the maximum output peak voltage is roughly 4.2 V. So the slew required slew rate can be calculated as:

$$\mathrm{SR} = 2\pi \times 4.2\mathrm{V} \times 20\mathrm{kHz} = 0.58\frac{\mathrm{V}}{\mu\mathrm{s}} \tag{26}$$

Noise, here again lower is always better. Since the passive components contribute to the most the noise in the circuit the limit for this value can be set at roughly the same as for the resistors. A value of 15 nV/Hz is the Johnson noise of a 10kΩ resistor (chosen as the maximum desirable value in the resistor selection section), and it is also a good limit for op amps. In addition the 1/f noise corner should be as low as possible, because the 1/f noise starts to rise very rapidly below its corner frequency. A value well below 100 Hz would be preferable. [27]

Total Harmonic Distortion + Noise. Most op modern op amps have very low THD values, often well below 0.1%. That is why this figure is not critical, since most of the THD in this circuit is expected to originate form passive components, especially chip capacitors.

Minimum supply voltage less than 3V. Most operational amplifiers require at least 5V supply voltage to work. Since this application is utilizing only a single lithium ion cell, that can drop as low as 3V, the op amp must be rated to operate below 3V.

Rail to rail output. With the operating voltage limited to only 3.7V, rail to rail operation is important to ensure maximum possible output voltage swing from the operational amplifiers. In this case the maximum attainable RMS voltage output with the nominal battery voltage of 3.7V is

$$\frac{3.7V}{2\sqrt{2}} = 1.3V \tag{27}$$

This value will be higher when the battery is fully charged and will drop considerably when almost depleted. Luckily, most low voltage operational amplifiers are rail to rail by design, making maximal use of the limited supply voltage.

Unity gain stable. In this design all operational amplifiers are configured for unity gain (a gain of 0dB), but some operational amplifiers are unstable at unity gain and can oscillate. Here chosen op amps must be specified unity gain stable.

DC Voltage offset. Since the output of this units are dc coupled, care must be taken in choosing op amps with minimal dc offset. Excess dc voltage at the output could easily damage the very sensitive balanced armature speakers. Values below 0.5mV are deemed acceptable here.

Low Iq, or quiescent current. With a total of 25 op amps in this crossover system, it is important to keep the total current consumption low to conserve battery life. Iq values below 2mA are deemed acceptable here because it will keep the overall idle current below 50mA, ensuring a long battery life from any decent sized lithium battery.

The output buffers and virtual ground divider have a few additional criteria to meet as they connected directly to the load. The following additional criteria apply to the op amps connected directly to the load.

Low output impedance. Because the output of the circuit will be connected to a low impedance load (range between 5Ω-100Ω), the output impedance of the output buffer must be below 1Ω at frequencies up to 20kHz to ensure proper damping for the load (as explained in the design specifications).

Output current. The operational amplifiers must be able to source (output buffers) and sink (virtual ground divider) the current drawn by the load in all circumstances. For the output buffers this is not a problem since each one only has to supply current for a single output, but the virtual ground divider must sink the combined return currents for all six loads. To estimate the current requirement we need to do some calculations. The currents are higher with lower impedance loads, so we can calculate with the worst case scenario, or in other words the lowest impedance. Having measured impedance curves of various balanced armatures, it has been found that the lowest impedance tend around 15 ohms at low frequencies with averages around 25 to 60 ohms. These were calculated with a maximum output voltage of 500mVAC RMS. Realistically this voltage will not ever be reached as 200mV is often more than enough to drive all balanced armatures at high volume levels, but it is prudent to have plenty of overhead. Putting all of this together we get:

$$6 * \frac{500mV}{15\Omega} = 200mA \tag{28}$$

Thus, we would need an op amp with at least 200mA of output current capability to meet this criteria. Now this is grossly overestimated value and an output current capability of half that would most likely be sufficient, because the average load impedances are more than twice the ones used in the calculation above and the output voltages significantly lower.

Phase Margin refers to an op amps phase relative to -180° at its unity gain frequency. All op amps have phase delay, meaning the output signal lags behind the input by some amount. The phase is highly important with op amp stability when using feedback (as nearly all op amps circuits do). If the total phase delay exceeds 360° the feedback becomes positive, causing the op amp to oscillate. Since the feedback is connected to the inverting input, which adds -180 compared to the input, phase margin is measured relative to -180°. Even though any number above zero is stable other factors such as capacitive loads can add to the phase shift causing instability. In practice phase margins above 45° is considered adequate and 60° is ideal [27]

Ability to drive capacitive loads. As explained above, some operational amplifiers can become unstable if connected to capacitive loads. In this application long interconnect wires can create such a capacitive loads and cause oscillations in some operational amplifiers. The amount of acceptable load capacitance is sometimes mentioned in datasheets and is useful to keep in mind when selecting components. The higher the acceptable capacitance, the better.

Based on the criteria set above, a number of possible op amps for this design were gathered and compiled in table 2 below. There are nearly endless more possibilities that cannot all be explored, and here the selection was limited to Texas instruments and Analog Devices op amps.

Table 2: Operational amplifier choices based on criteria described above, for filters and buffers.

Op amp	**Noise**	**DC offset (typical)**	**Output impedance**	**I q**	**Price***
AD8694	8 nV/√Hz	400 uV	<1Ω @ 100kHz	0.95 mA	2,45 €
AD824	16 nV/√Hz	200uV	1.5 Ω @ 20kHz	500 uA	9,78 €
OP484	3.9 nV/√Hz	300uV MAX	<10Ω @ 100kHz	2 mA	9,40 €
OP462	9.5 nV/√Hz	50 uV	~25Ω @ 1MHz	500 uA	8,24 €
AD8604	33 nV/√Hz	350 uV	<5 Ω @ 20kHz	680 uA	2,42 €
AD8618	10 nV/√Hz	80 uV	3Ω @ 1MHz	1.7 mA	5,63 €
AD8608	8 nV/√Hz	20uV	1Ω @ 1MHz	1.2mA	3,79 €

*Price in Euros at www.mouser.com as of November 2015

As we can see from the chart there is a compromise between performance, price and power consumption. It was at the designer's discretion to prioritize between these parameters.

In different versions of the design, both AD8694 and AD8608 were used. Both op amps have very similar specifications, the main difference being a much lower DC offset in the AD8608, at the expense of slightly higher current draw and cost.

6.2 Capacitors

Considering the size limitations of the printed circuit board, all passive components are to be kept as surface mount components as they have a smaller footprint than through hole types. Surface mount capacitors are available in several different dielectric materials, but due limitations on cost, size and available capacitance values, ceramics were considered as the only viable choice. The capacitance values needed to produce the desired crossover frequencies are in a range between a few nano Farads to a few micro Farads. Smaller capacitance values would be affected by parasitic capacitances in the PCB, that could affect crossover performance and higher capacitances in ceramic capacitors aren't readily available.

6.2.1 High K Ceramic Type

The most common dielectric type for ceramic capacitors is High K type commonly known as X5R or X7R. They are cheap and easily available in a wide range of sizes and capacitance values. The drawbacks of High K ceramics, are their distortion characteristics, and piezo electric effects. A general consensus exists, that these types of capacitors should not be used in the audio signal path [28]. These can still be used as power supply decoupling capacitors where the harmonic distortion is not an issue.

6.2.2 C0G/NP0 Ceramic Type

Less common and more expensive, the C0G/NP0 type capacitors exhibit far less distortion than the High K capacitors, making them a more suitable choice for high quality audio applications where a small form factor is required. These capacitors however are many times more expensive and not readily available in higher capacitance values potentially needed for filters at lower frequencies. This capacitance limitation necessitated a need to employ parallel capacitors to create a filter with a low frequency cut-off point (low frequency output of the crossover) without using very high resistances. See appendix 1 for details on all attainable crossover frequencies with different capacitor types.

6.3 Resistors

As with the capacitors, only surface mount resistor are suitable in this application. Resistance values used in this application range from around 1kΩ to 50kΩ to achieve desirable filter corner frequencies. Using higher value resistors introduce more Johnson noise, thus it is preferable to keep the values as low as possible. However, low resistance values require the use of higher capacitances to reach lower filter cut-off frequencies, which might only be possible to implement with the lesser performing High K capacitors. This creates the need for a compromise between the noise from resistors and distortion from the capacitors. Also the loading of the previous filter stage should be kept in mind, as low resistances can create a high current draw from the previous stage. An ideal balance can only be found out through experimentation. In most parts of the circuit values of less than 15k Ω are used to minimize noise.

Surface mount resistors are available in several sizes. In addition to taking up space the size of the component affects its performance too. Noise and linearity characteristics change with the resistor power rating, which is also directly proportional to its size. The larger the component, the less noise and distortion it causes, this can be seen in the charts in appendix 5. For this reason 1206 size resistor are used in the audio signal path of this design

SMD resistors can be divided into two types: thick film and thin film. Generally easily available in a wide range of values with low cost the thick film resistors are by far the most common type of surface mount resistor. They can be used in designs where a high degree of accuracy or noise are not critical.

More costly than thick films, thin film resistors exhibit more stable temperature coefficients, less deviation through aging and lower current noise values and better linearity. In this application thermal coefficients are not crucial as a very low power device, no significant heating up of the components is expected. Whereas the thermal noise is not affected by the type of resistor used, the current noise however is. The current noise has a 1/f spectral distribution, meaning it is more pronounced in lower frequencies. From figure 21 below we can see that above 1 kHz the thermal noise becomes the dominant noise source and the effect of the current noise can effectively be ignored. [29]

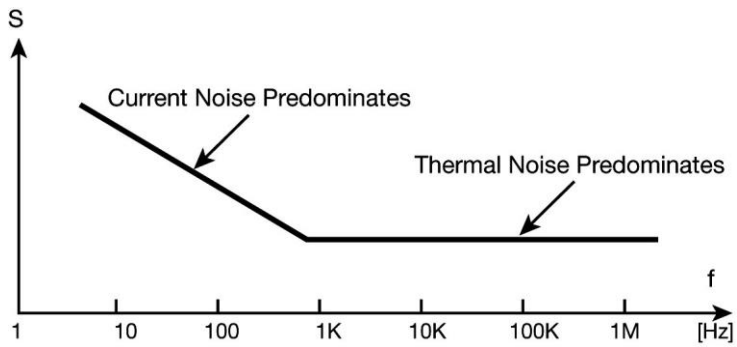


Figure 21: Spectral distribution noise in resistors. Copied from Vishay [29]

Since, in an audio circuit frequencies below 1 kHz are important the use of thin film resistor is beneficial of thick film types. For maximizing cost efficiency a case could be made for using thick film components in the high passed signal chain as the noisier low frequencies would be filtered out, at the expense of a more complex bill of materials.

Linearity is also a factor to be considered when choosing resistors. Thin film types exhibit a more linear frequency response, which means less distortion or THD. A more detailed comparison between thin film and thick film resistors can be seen in the resistor comparison charts in appendix 5.

6.4 Connectors

While the input connector is a standard 3.5mm audio jack, the output connectors were somewhat more problematic to implement. This is because there are six output channels to drive and no standard audio connector accommodates that. What was needed is a solution that is small enough to fit on the enclosure and PCB, and also be practical on the headphone cable. As mentioned in the introduction, this system requires the earphones to be purpose-built for it. This allowed the designer to freely choose whatever type of connector was deemed most practical. After exploring various options such as

DIN connectors (no proper locking mechanism, thus prone to coming loose) and mini XLR (7 pin version are rare and expensive, and also male connector very bulky), two 4-pin 3.5mm audio jack connectors were the final choice. With two smaller connectors instead of one large one, each will be connected to one IEM (left and right channels).

The cable used in the prototype IEMs consisted of two thin 3.5mm plug audio cables connected together with a braided sleeve. The result was barely thicker than a normal headphone cable, thus a practical choice.

All the components in each version of the system can be seen on the bill of materials in appendix 6.

7 Measurement and analysis

7.1 Test Methods and Equipment

The initial specifications from section 2 of this thesis are to be measured and analysed in order to conclude whether or not this project is a success. Here the tools and methods are explained for each spec to be measured, with the full specifications of the test equipment in appendix 7.

Brymen BM867s Multimeter
This very accurate multimeter was used for measuring the output DC offset, output impedance, channel balance, total current consumption, virtual ground reference voltage and power supply voltages.

Velleman PCSU 200 USB Oscilloscope
The Velleman USB scope is a handy tool for general testing and trouble shooting. Its specs are not good enough for accurate testing of noise and THD figures, but it does an adequate job for measuring the output bode plots (crossover cut-off frequency points and phase difference between outputs) and maximum output levels before clipping.

ESI ESP1010 Sound Card & ARTA Software
Noise and distortion figures were measured using ARTA spectrum analyser software, with ESI ESP1010 computer soundcard providing the necessary AD and DA conversion.

Some of the measurements are quite basic and require no detailed explanation on how they concluded. A few of the more complex measurement procedure will be explained here in more detail.

7.1.1 Bode Plots

Bode plots were made with the Velleman USB oscilloscope, as it has the functionality to automatically graph the frequency response of a system, making the measurement very quick and easy. Both amplitude and phase responses were graphed. Each frequency band output is separately measured in its own graph, for both left and right channels. In addition a combined graph of the full bandwidth output for both left and right channels was made. Another thing to note is that the outputs are not equal in amplitude to one another. This is because the different balanced armature used in testing have very different sensitivities and the outputs were adjusted to suit the designers' subjective preference. Below in table 3 are all crossover frequencies and output levels for the different versions.

Table 3: Crossover frequencies and output levels

	Low End	Mid-Range	High End
1st version	112Hz (adjustable level)	112Hz - 2800Hz (-12.45dB)	2800Hz (-0dB)
2nd Version	85Hz (-12.7dB)	85Hz-1023Hz (-10.1dB)	1023Hz (-2.5dB)
3rd version	68Hz (-12.7dB)	68Hz-2000Hz (-9.5dB)	2000Hz (0dB)

7.1.2 Output Impedance

To measure the output impedance, the input of the device under test connected to the function generator of the Velleman oscilloscope outputting a sinewave. The frequency of the sine wave was chosen to be in the pass band of each output under test (50 Hz for the low pass, 1 kHz for the band pass and 10 kHz for the high pass). The output was adjusted to about 100mV. Then a 10Ω (measured exactly 10.00Ω with the multimeter) resistor was connected to the output and the new output voltage recorded. As this load resistor forms a voltage divider with the output impedance, using ohms law the following equation can be derived:

$$Z_{OUT} = R_{LOAD} * \left(\frac{V_{UNLOADED}}{V_{LOADED}} - 1\right) \qquad (29)$$

And with that we calculate all output impedances from all design version into the table below (full table with intermediary voltages in appendix 8).

7.1.3 THD and THD+N

The distortion was measured by hooking up the device to the Inputs and outputs to the ESP1010 sound card and generating a 1 kHz signal to the input and measuring the output with the ARTA software. Since the crossovers output are divided into three different frequency bands they were all tied together with 470Ω resistors to create a passive summing network. This full range output was then measured. Normally audio equipment is measured with around -3dBV (0.707V RMS) output or so, but since this device will not be with used signals of such high amplitude, THD was measured with an output of -20 dBV (100mV) as this is provides a reasonable listening level. Using higher signal levels would provide better specifications on paper, but it was considered more honest to measure levels that could conceivably be used with an actual balanced armature load.

7.1.4 Noise Floor

The **noise floor** or **background noise level** was measured with the same setup as the distortion, but with the exception that no input signal was used. The resulting spectrum analyser view will yield the noise floor of the system. This could be used to compute a **signal-to-noise** ratio, which is the ratio between the average noise level and average signal level. Moreover the **dynamic range** value can also be concluded which is the ratio between the noise floor and maximum level before clipping. [30] There are some problems with this though. As each individual output is attenuated to a different signal level, each would have their own signal to noise ratios. Furthermore those output levels are rather significantly attenuated compared to the input and other outputs (a difference of -12dB or more), the dynamic range of the system would be considerably higher than the signal-to-noise ratio. As the dynamic range depends on the battery voltage level of the system which varies, and since there would be several signal to noise ratios between outputs, it was decided to provide only the noise floor figure as that is common to the entire circuit and all outputs.

7.2 Design Revisions

Even though the very first prototype worked, throughout the design process some trial and error has provided additional insights, based on which improvement ideas were developed. Here the variations and improvements are explained and test results for each revision are shown. At least a dozen different design revisions were laid out on cad software, some with branching objectives differing from the initial specifications and out of those, three different circuit revisions were actually produced.

7.2.1 Initial Prototype

The initial prototype or first manufactured version was made basically as a proof of concept. The main goal was to see that the crossover would work with the calculated frequencies and relatively little consideration was spent on component selection and detailed circuit layout.

The passive components are of unknown brand and technology, sourced cheaply through eBay. The components values were not considered beyond the calculation of the frequency cut-off points. The operational amplifiers are partly the same as in later version, and thus their specifications were thoroughly checked to meet the criteria laid out in chapter 6.

7.2.2 Second Version

This second version corrects some of the mistakes made in the initial prototype. The main problem was some audible background noise when no music was playing. To correct this all resistor values were lowered, with an absolute maximum value set at 50kΩ and ideally attempting to keep most of the values below 10kΩ.

Another point was to source all components from a reputable supplier (www.mouser.com) and use well-known brands. All resistors are thin film types to keep noise and distortion to minimum. Chip capacitors are High K ceramics (X7R dielectric), as this enables a very wide choice of crossover frequencies. The crossover frequencies were changed for experimentation purposes, but it is entirely possible to use the same frequencies as in version 1 if necessary.

The PCB design was modified to improve grounding and incorporates some limited use of ground planes for the virtual ground and power supply ground. This was almost entirely missing in the first version.

The op amps were partly changed. In version 2 all op amps are AD8608s as opposed to just having them as the input and output buffers. These op amps are slightly better performing and are expected be the reason behind the reduce output DC offset and impedance.

Other smaller changes include the elimination of trimmer resistors at the low frequency output level adjust. They were placed in the initial version to easily adjust low frequency output, but were deemed unpractical and took up too much valuable board space. Also a quite loud “thump” could be heard during power up and power down in the first version. These occur during powering transient period and while they were too small to be accurately measured with the equipment on hand, they can easily be heard through the IEMs connected to the output.

In version 2 it was attempted to correct it by placing the main power supply reservoir capacitor before the power switch, so they would not draw high currents during power on [22]. And in version 2 this thump, though still slightly audible, it was considerably reduced from version 1.

7.2.3 Final Version Including Added DAC Section

The third version is a major overhaul of the design, since it includes an on-board USB digital to analog converter. This DAC was not part of the initial plan laid out, but rather the idea for its inclusion came about after the battery charging connector was changed to a more practical mini USB connector from the initial DC barrel connector. Since the DAC was not part of the initial specifications, but rather is an extra feature, it will not be discussed in detail in this paper. Only the analog sections of all designs are compared and analysed.

The analog portion of the circuit underwent a number of changes too in version 3. This time around, after studying the difference between different types of ceramic capacitors (as explained in chapter 6), a change to C0G capacitors were made for the entire analog

signal path. This also necessitated some changes in the crossover frequencies as the choices become more limited with these components.

The PCB layout is also very different, not only due to the inclusion of the DAC section, but the ground connections were again improved from version 2. Because of the inclusion of a far greater amount of components, and the resulting tight layout of components, the third board was not attempted to be milled at Metropolia, but instead it was manufactured at a professional board house.

7.3 Test Results

Here is a summary of the test results and brief discussion on what they mean in practice. The full test results for all versions can be found in appendix 8.

7.3.1 Frequency Response

The core crossover functionality worked very well from the very first version. Bode plots of all outputs show flat passbands and a consistent 60dB/decade slope down to -60dBV when the oscilloscope noise floor prevented further accuracy. There is no measurable phase shift in version 1, precisely as planned. In version 2, however there is a noticeable phase difference of about 21 degrees between the mid-range and high outputs. Because there is no phase shift between the low end and the mid-range, it was concluded that this error was due to component value tolerances and not a problem in the circuit itself. Checking and changing the component values would solve the phase error. Version 3 fixes the phase discrepancy, as it only has 6 degrees of phase difference at the higher crossover point and less than 1 degree at the lower crossover point. These small phase differences in version 3 will not be audible.

Detailed bode plots (with amplitude and phase) of all individual outputs of all versions can be found in appendix 8. One thing to note is that the phase plots are highly inaccurate on the stop band of the signal and should be disregarded.

7.3.2 Output Impedance

Versions 1 and 2 have a very small output impedances, but in a few of the outputs of version 1 the impedance goes above 1Ω which can create a small change in the output frequency response when connected a low impedance load. Output impedances show

an improvement in version 2, where the result are excellent. All the output impedances can be seen from table 4 below.

Table 4: Output Impedances

Output Impedance				
		1st version	2nd Version	3rd version
Left Channel	High @ 10kHz	1.48 Ω	0.09Ω	0.046 Ω
	Mid @ 1kHz	1.51Ω	0.09Ω	0.064 Ω
	Low @ 50 Hz	0.57Ω	0.29Ω	0.082 Ω
Right Channel	High @ 10kHz	0.19Ω	0.16Ω	0.048 Ω
	Mid @ 1kHz	0.13Ω	0.14Ω	0.092 Ω
	Low @ 50 Hz	0.23Ω	0.15Ω	0.054 Ω

7.3.3 DC Offset Voltage

The DC offsets of all outputs were very low from the first version on. Again version two showed improvement in this category where the maximum offset voltage is only 0.055mV, ten times lower than in version 1, where the result was already a very decent value of 0.552mV. Version 3 is practically equal to the second version. Below in table 5 are DC offset values from outputs.

Table 5: DC Offset Voltages

DC offset (No Input Connected)				
		1st version	2nd Version	3rd version
Left Channel	High	-0.027mV	0.021mV	-0.041mV
	Mid	0.183mV	-0.011mV	0.020mV
	Low	-0.141mV	-0.001mV	0.007mV
Right Channel	High	0.552mV	0.09mV	0.020mV
	Mid	0.417mV	0.022mV	-0.005mV
	Low	0.38mV	0.055mV	-0.034mV

7.3.4 Distortion and Noise

Distortion and noise figures were perhaps the most problematic, since they require a great deal of accuracy from the test setup, and defining the test parameters is not as a clear cut as the other measurements. The test method was explained previously and was applied successfully to version 2 which measured a decent THD+N figure of 0.072%. Version 1, however started to distort before the specified output level of -20dBV, the reason for this is unknown as the distortion should only rise rapidly once the system start

to clip, meaning the input voltage exceeds the power supply voltage level. The output was taken down to -25dBV, where version 1 measured a THD+N level of 0.33%. Version 3 is the best of the bunch, measuring at 0.023% THD+N. The decrease in distortion can likely be attributed to the use of C0G capacitors in the audio path.

The noise floor was then attempted to test for both version 1 and 2, but the noise floor of the test equipment at -93.8dBV proved to be higher than both versions, so it could not be accurately measured beyond that limit. When connecting IEMs to the crossovers however, a slight difference in the background noise floor can be heard. In version 1 this is very faint and can be only heard in quiet surroundings. Version 2 improves upon this further, and its background noise is almost completely inaudible. In the noise floor version 3 is again an improvement on the previous versions, and the background noise through the IEMs can only be heard in complete silence, and only if the listener is concentrating on hearing it.

The full distortion and noise graphs can be seen in the test results in appendix 8.

7.3.5 Current Consumption and Battery Consumption

The main differences between the first two versions here is the change of four of the quad op amps to ones with slightly higher current consumption and the lowering of most of the resistance values in version 2. Thus, it was expected for the current consumption to go up a few milliamps to compensate for increased performance. And that is exactly what happened. Version 3 uses the same op amps as the second one, which results in the current consumption being roughly the same. Below in table 6 is the comparison between idle current consumption of the versions and a calculated battery idle runtime with a 1600mA battery.

Table 6: Battery Consumption

Battery Consumption		
	Iq (Idle Current)	Idle Runtime (1.6Ah battery)
1st version	22.24mA	71.94 h
2nd Version	24.03mA	66.58 h
3rd version	24.70mA	64.77 h

8 Conclusion

This portable crossover system was a challenging design, not due the complexity of the circuit itself, but rather the very demanding loads it has to be able to handle. The high sensitivity of balanced armatures means that system going to be a low gain one, with the low voltages signals being highly susceptible to noise interference. In many instances, it has turned out that the balanced armatures are more sensitive to picking up noise than the test equipment used. Taking this into account, the very reasonable noise and distortion figures from the system are at the very least, satisfactory.

The main crossover functionality had no major issues in any version and worked almost precisely as simulated. Were it not for the noise and distortion issues in version 1, this project would have been a success from the very first prototype. Version 2 addressed these issues and so met all the initial specifications. And finally version 3, which was manufactured on a proper PCB and utilized better passive components, produced remarkably good test results. To summarize, this is a very high performance system, with high DC accuracy, linearity and dynamic performance. The battery life of over 60 hours is major improvement over 8 hours mentioned in the beginning of this paper. Technically this project is a resounding success and worthy to be called a high end audio crossover.

As for actual commercial feasibility of the product, one has to consider the intended user. Not many casual music listeners would be willing to add another piece of equipment between their music players and earphones to carry around. This device was meant for audiophiles or professional musicians, many of whom already often use headphone amplifiers of equivalent size and weight in their audio systems.
Further development of this system could go in two ways. First, keeping only the current functionality and focus on making the unit smaller. As the inclusion of the DAC section shows, the analog portion of the system could fit in a smaller PCB, thus a smaller enclosure. The connectors and volume control would be the biggest components to fit in. The second option would be to include additional functionality. The DAC converter in version 3 opens up the possibility for adding a wireless Bluetooth module. Another idea the author has considered is active noise cancelling. The possibilities for further development are wide.

In any case, market research is not in the scope of this technical paper, but to conclude, the author of this thesis uses this crossover system now on a daily basis and has no intention to go back to lesser performing IEMs. The finished system can be seen in its housing and with prototype custom built IEMs in figure 22 below.

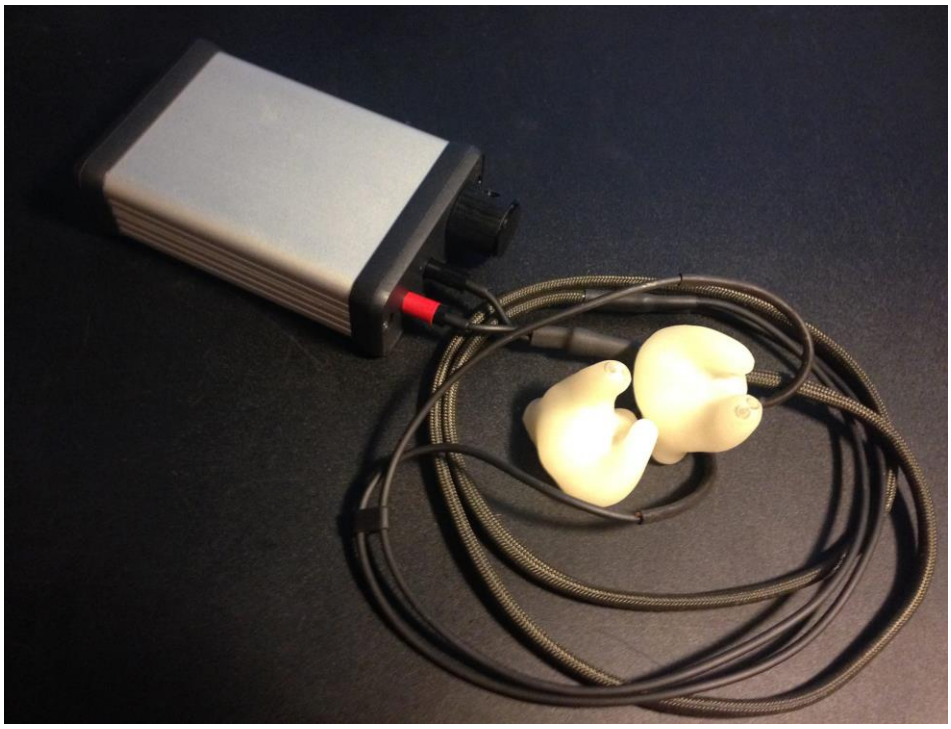


Figure 22: Finished Crossover with IEMs

References

1 Ultimate Ears. In-ear monitor diagram [online]. Hifi plus website; 22 December 2011.
URL: http://www.hifiplus.com/articles/ultimate-ears-ue-18-pro-in-ear-monitors-playback-52/. Accessed 5 January 2016.

2 Hertsens T. Balanced armature diagram [online]. Innerfidelity website; 15 December 2014.
URL: http://www.innerfidelity.com/content/how-balanced-armature-receiversdrivers-work. Accessed 15 January 2016.

3 Knowles. Product information page [online]. Knowles website.
URL: http://www.knowles.com/eng/Applications/Specialty-components/Hearing-aid-components/Balanced-Armature-Products. Accessed 10 December 2015.

4 Unique Melody. PP-6 tech specs [online]. Unique Melody website.
URL: http://uniquemelody.co/lineup/platformpure6/. Accessed 10 January 2016.

5 Analog Devices. MT-047 Op Amp Noise [online]. Analog Devices Inc.; October 2008.
URL: http://www.analog.com/media/en/training-seminars/tutorials/MT-047.pdf. Accessed 20 December 2015.

6 Ott HW. Noise Reduction Techniques in Electronic Systems. 2nd ed. John Wiley & Sons; 1988.

7 Self D. Audio Power Amplifier Design Handbook. 3rd ed. Newnes; 2002.

8 Wickstrom T. Earphone Design Guide Application Note [online]. Knowles Acoustics.
URL: www.knowlesacoustics.com. Accessed 31.5.2014.

9 Analog Devices. Basic Linear Design [online]. Analog Devices website.
URL: http://www.analog.com/library/analogDialogue/archives/43-09/EDCh%208%20filter.pdf. Accessed 15 December 2015.

10 Wikipedia contributors. Q Factor [online]. Wikipedia website; 23 November 2015.
URL: https://en.wikipedia.org/wiki/Q_factor. Accessed 28 December 2015.

11 Horowitz P, Hill W. The Art of Electronics. 2nd ed. Cambridge University Press; 1989.

12 Wikipedia contributors. Butterworth Filter [online]. Wikipedia website; 28 November 2015.
URL:https://en.wikipedia.org/wiki/Butterworth_filter. Accessed 28 December 2015.

13 Lacanette K. Application Note 779 A Basic Introduction to Filters - Active, Passive, and Switched-Capacitor [online]. National Semiconductor; April 1991.
URL: http://www.ti.com/lit/an/snoa224a/snoa224a.pdf. Accessed 30 December 2015.

14 Wikipedia contributors. Chebyshev Filter [online]. Wikipedia website; 11 November 2015.
URL:https://en.wikipedia.org/wiki/Chebyshev_filter. Accessed 28 December 2015.

15 Bohn D. A Bessel Filter Crossover, and Its Relation to Others [online]. Rane Corporation; April 2006.
URL: http://www.rane.com/note147.html. Accessed 4 November 2015.

16 Wikipedia contributors. Linkwitz–Riley filter. Wikipedia website; 7 April 2014.
URL: https://en.wikipedia.org/wiki/Linkwitz%E2%80%93Riley_filter. Accessed 28 December 2015.

17 Bohn D. Linkwitz Riley Crossovers: A Primer [online]. Rane Corporation; October 2005.
URL: http://www.rane.com/note160.html. Accessed 4 November 2015.

18 Linkwitz S. Active filter descriptions [online]. Linkwitzlab website; 28 June 2014.
URL: http://www.linkwitzlab.com/filters.htm. Accessed: 5 November 2015.

19 Self D. Small Signal Audio Design. Focal Press; 2010.

20 Bohn D. A Fourth-Order State Variable Filter for Linkwitz-Riley Active Crossover Designs Specifications [online]. Rane Corporation; October 1983.
URL: http://www.rane.com/pdf/linriley.pdf. Accessed 9 November 2015.

21 Elliot R. Linkwitz-Riley Electronic Crossover [online]. Elliot Sound Products website; 22 August 2012.
URL: http://sound.westhost.com/project09.htm. Accessed 20 November 2015.

22 Kan T. Mini3 circuit description [online]. AMB website; 2011.
URL: http://www.amb.org/audio/mini3/. Accessed 15 November 2015.

23 Semig P, Claycomb T. Capacitive Load Drive Solution using an Isolation Resistor [online]. Texas Instruments; November 2014.
URL: http://www.ti.com/lit/ug/tidu032c/tidu032c.pdf. Accessed 5 January 2015.

24 Kitchin C. Biasing and Decoupling Op Amps in Single Supply Applications [online]. Analog Devices; 2002.
URL: http://www.analog.com/media/en/technical-documentation/application-notes/AN-581.pdf. Accessed 8 September 2015.

25 Glenewinkel M. AN1259 System Design and Layout Techniques for Noise Reduction in MCU-Based Systems [online]. Freescale Semiconductor; 1995.
URL: http://cache.freescale.com/files/microcontrollers/doc/app_note/AN1259.pdf. Accessed 20 December 2015.

26 Kugelstadt T. Active Filter Design Techniques. In: Carter C, Mancini R, editors. Op Amps for Everyone. Burlington, MA; Newnes; 2009. p.365-438.

27 Bohn D. High-Quality Low-Voltage Audio [online]. Rane Corporation; February 2001.
URL: http://www.rane.com/hilovolt.html. Accessed 9 November 2015.

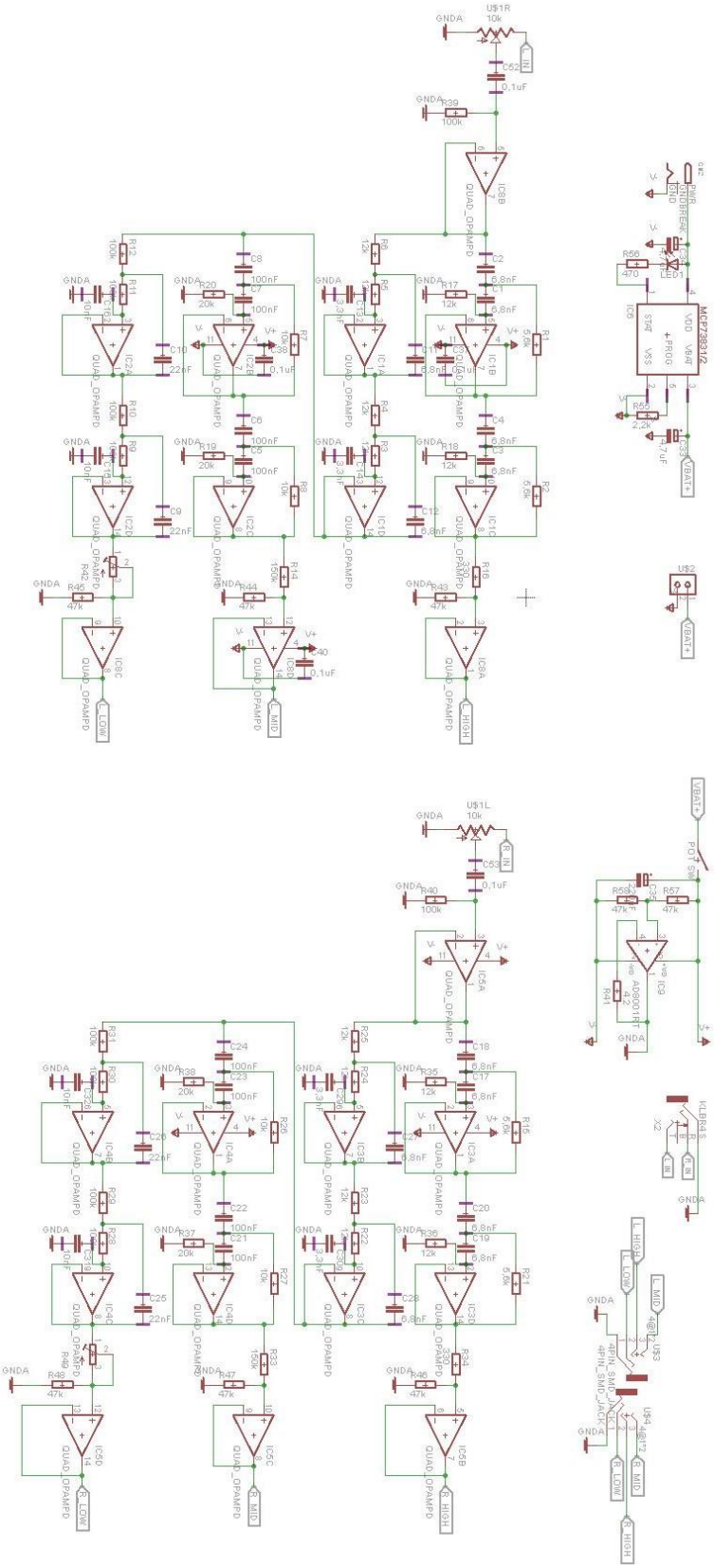
28 Bateman C. Capacitor Sounds part 3 [online]. Electronics World; March 2003. URL: http://linearaudio.nl/sites/linearaudio.net/files/Bateman%20EW%2010%202002%20mar2003%2010%20nF%20and%20less.pdf. Accessed 16 January 2016.

29 Hernik Y. Selecting Resistors for Pre-Amp, Amplifier, and Other High-End Audio Applications [online]. Vishay Precision Group; January 1 2011. URL: http://www.vishaypg.com/docs/49997/49997.pdf. Accessed 20 December 2015.

30 Brown, P. The Differences Between Dynamic Range & Signal-to-Noise Ratio [online]. Prosound website; 18 December 2014. URL: http://www.prosoundweb.com/article/the_differences_between_dynamic_range_signal-to-noise_ratio/. Accessed 16 January 2016.

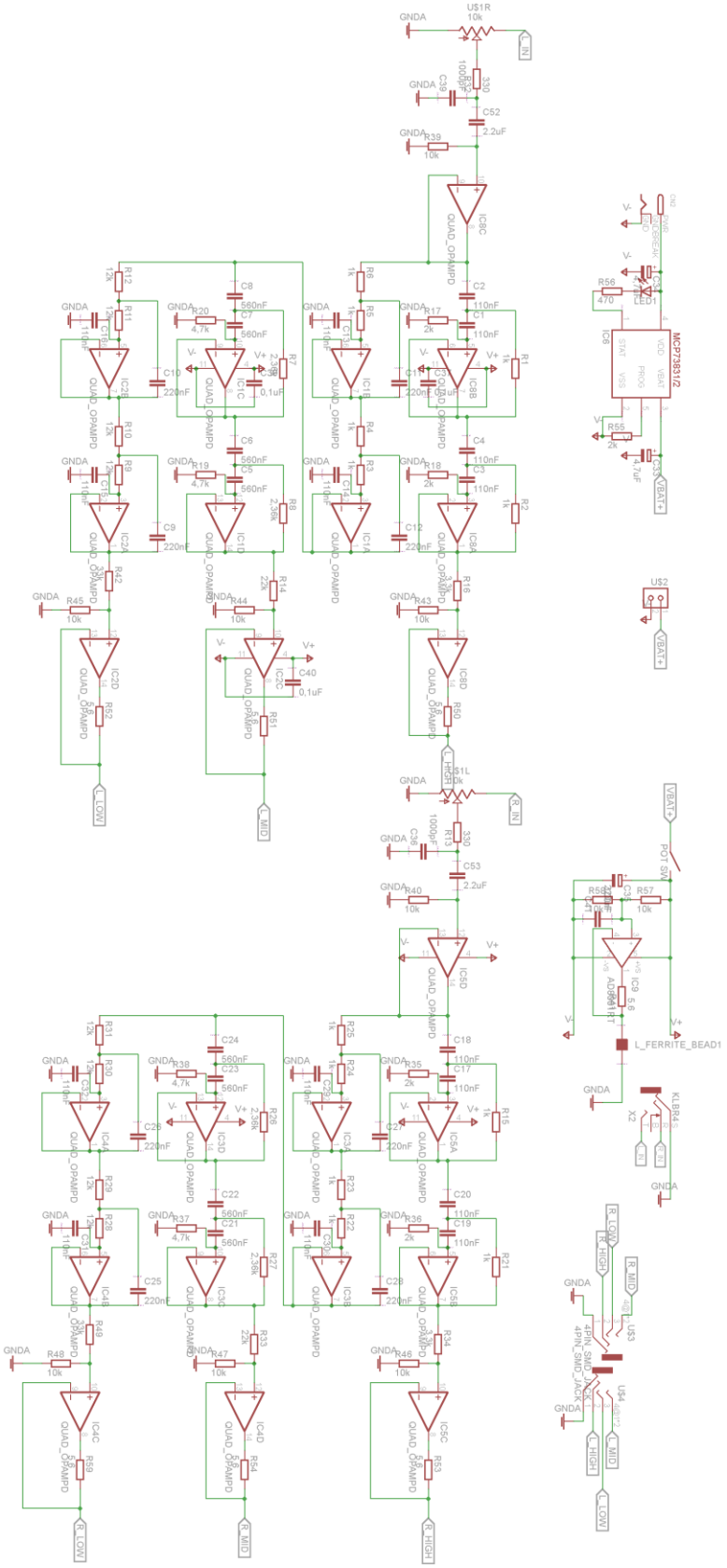
Schematic Diagrams

Version 1



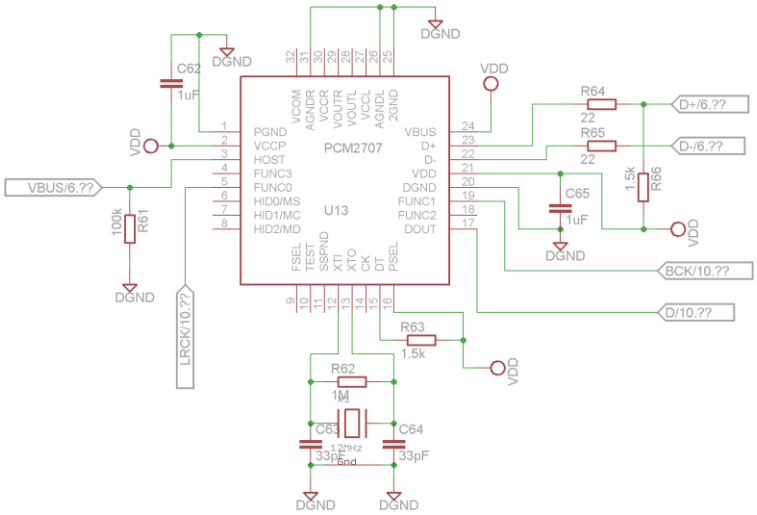
Schematic Diagrams

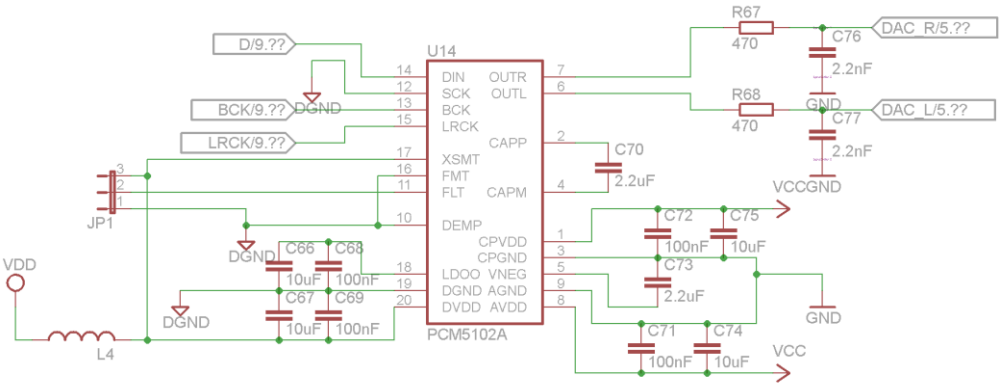
Version 2

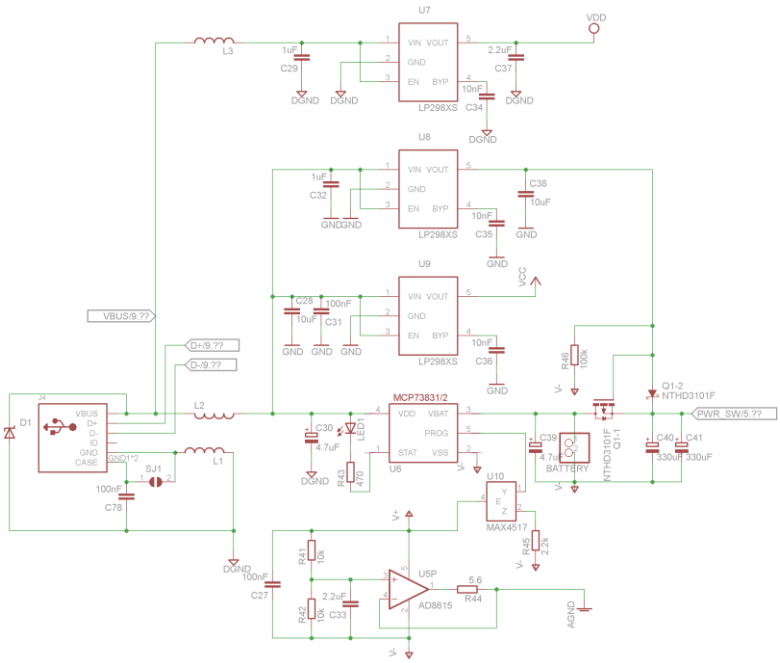


Schematic Diagrams

Version 3 (DAC and power supply sections only)







Possible Crossover Frequencies

Using X7R capacitors

	590Ω & 1.18k Ω	1k Ω & 2k Ω	1.1k Ω & 2.2k Ω	7.5k Ω & 15k Ω	10k Ω & 20k Ω	16.5k Ω & 33k Ω
1nF & 2nF	190871	112614	102376	15015	11261	6825
1.5nF & 3nF	127247	75076	68251	10010	7508	4550
1.8nF & 3.6nF	106039	62563	56876	8342	6256	3792
5nF & 10nF	38174	22523	20475	3003	2252	1365
15nF & 30nF	12725	7508	6825	1001	751	455
50nF & 100nF	3817	2252	2048	300	225	137
110nF & 220nF	1735	1024	931	137	102	62

Using C0G(NP0) capacitors

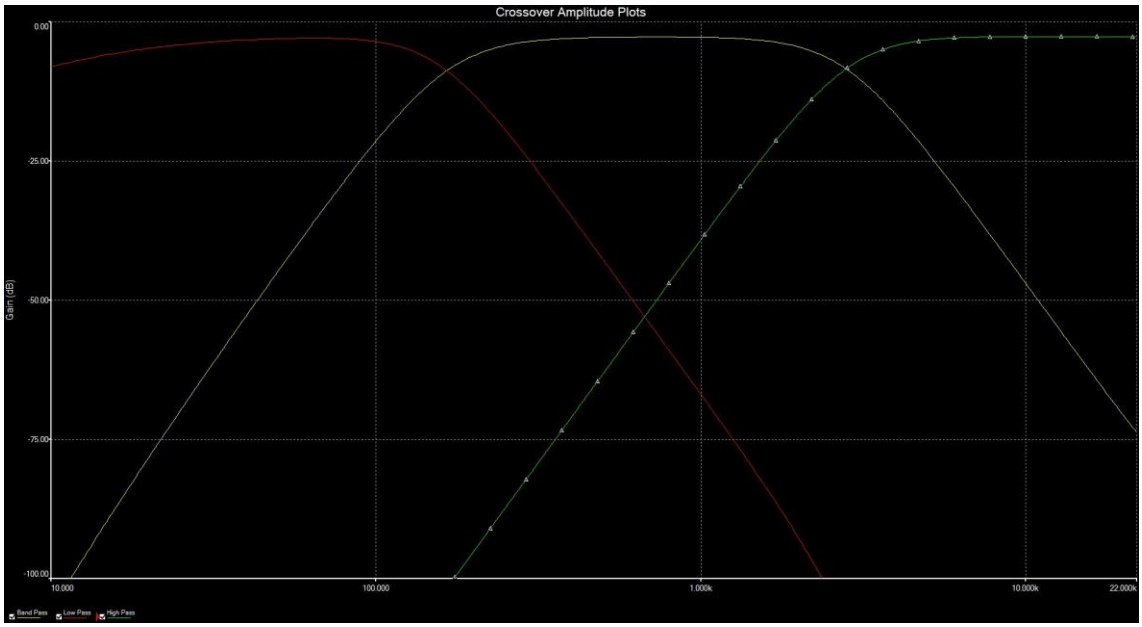
	590Ω & 1.18k Ω	1k Ω & 2k Ω	1.1k Ω & 2.2k Ω	7.5k Ω & 15k Ω	10k Ω & 20k Ω	16.5k Ω & 33k Ω
1nF & 2nF	190871	112614	102376	15015	11261	6825
1.1nF & 2.2nF	173519	102376	93069	13650	10238	6205
1.2nF & 2.4nF	159059	93845	85313	12513	9384	5688
1.5nF & 3nF	127247	75076	68251	10010	7508	4550
1.8nF & 3.6nF	106039	62563	56876	8342	6256	3792
7.5nF & 15nF	25449	15015	13650	2002	1502	910
10nF & 20nF	19087	11261	10238	1502	1126	683
11nF & 22nF	17352	10238	9307	1365	1024	620

Using parallel capacitors

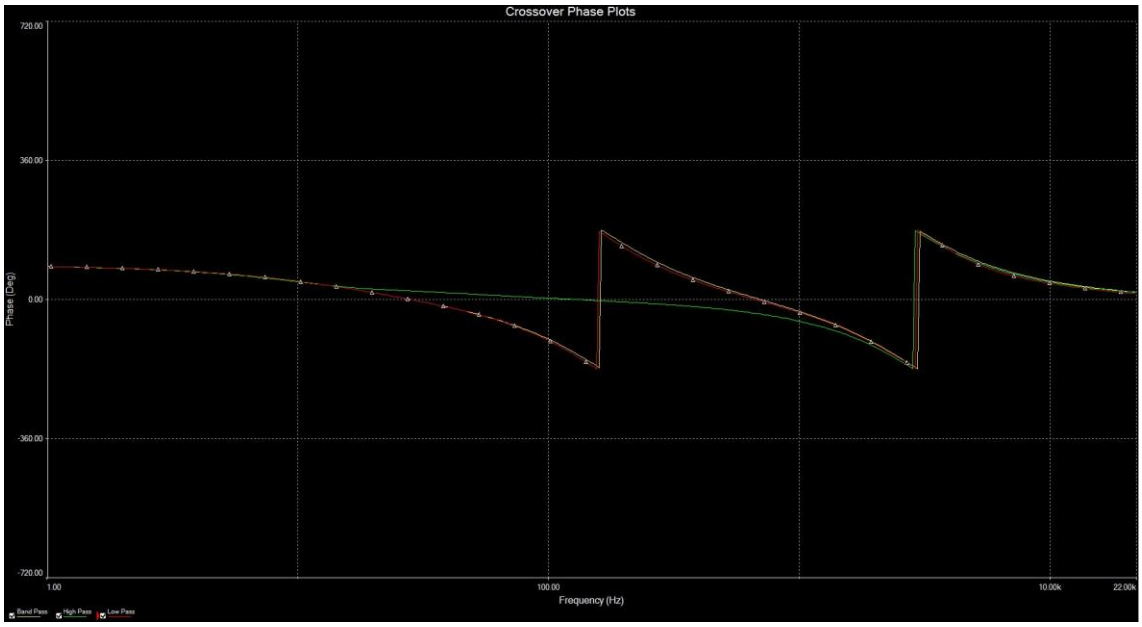
	590Ω & 1.18k Ω	1k Ω & 2k Ω	1.1k Ω & 2.2k Ω	7.5k Ω & 15k Ω	10k Ω & 20k Ω	16.5k Ω & 33k Ω
27nF	7069	4171	3792	556	417	253
33nF	5784	3413	3102	455	341	207
47nF	4061	2396	2178	319	240	145
68nF	2807	1656	1506	221	166	100
82nF	2328	1373	1248	183	137	83
100nF	1909	1126	1024	150	113	68

*frequencies in red are not suitable and orange ones only in marginal cases, where a very high crossover point is desired

Simulated Bode Plots



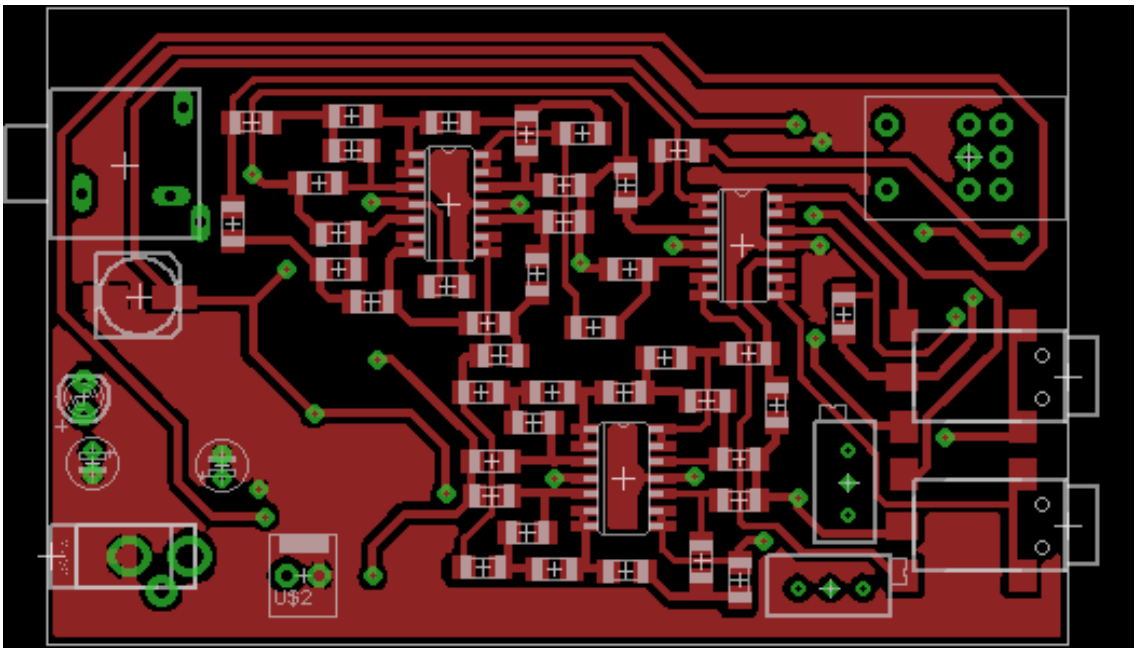
Amplitude Plots Simulated in Multisim



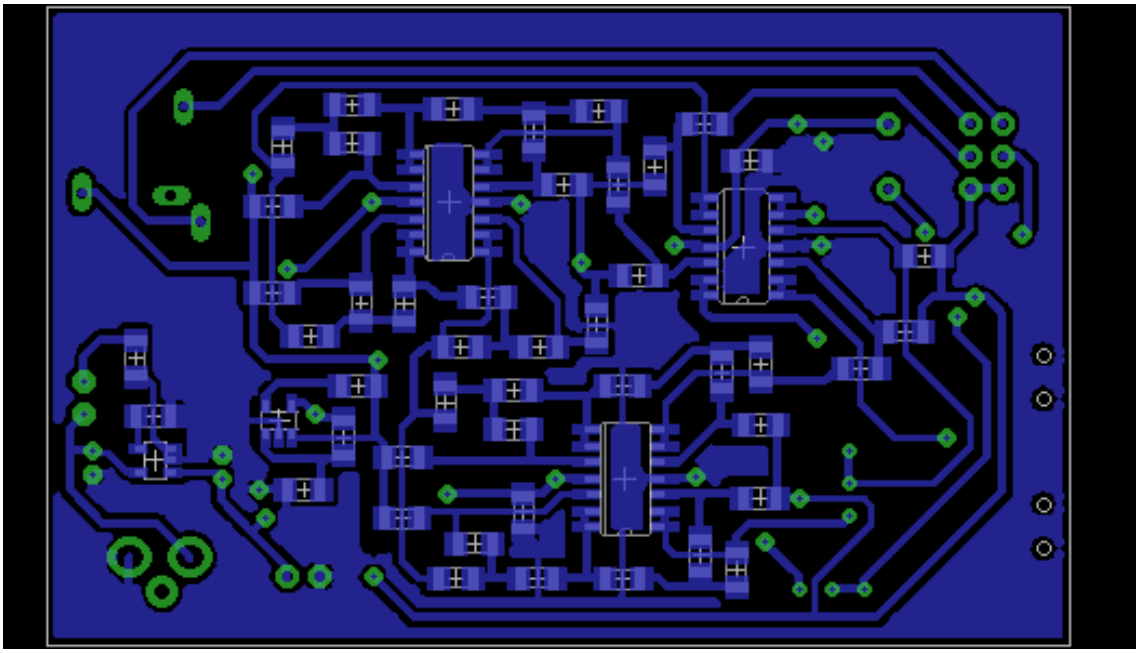
Phase Plots Simulated in Multisim

PCB Layers

Version 1



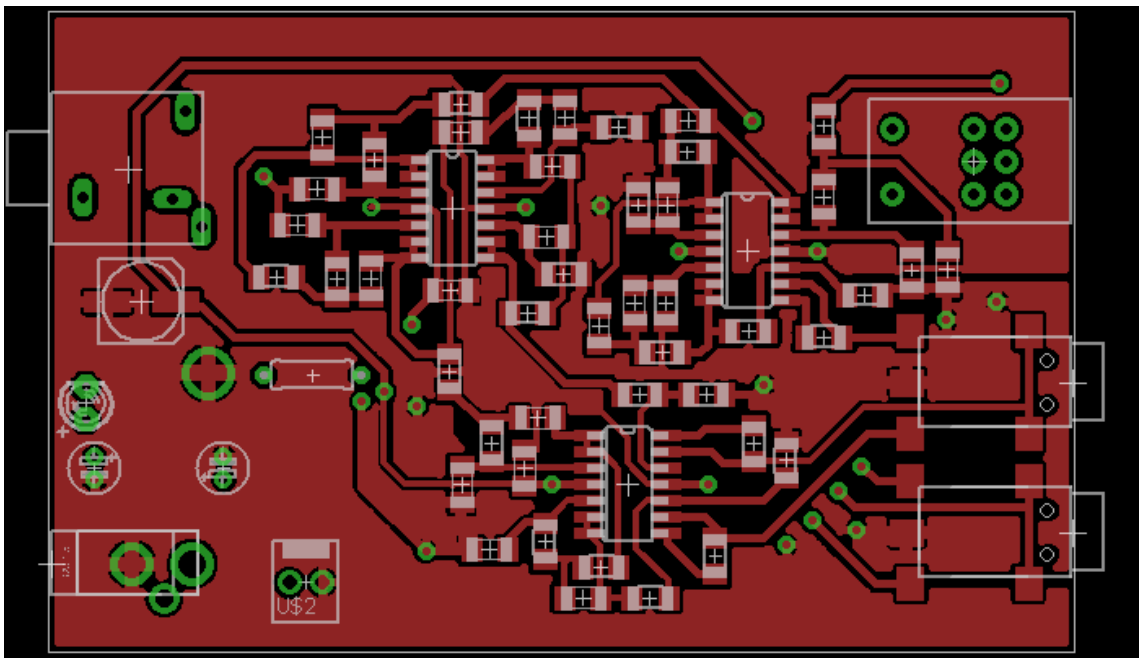
Top Layer



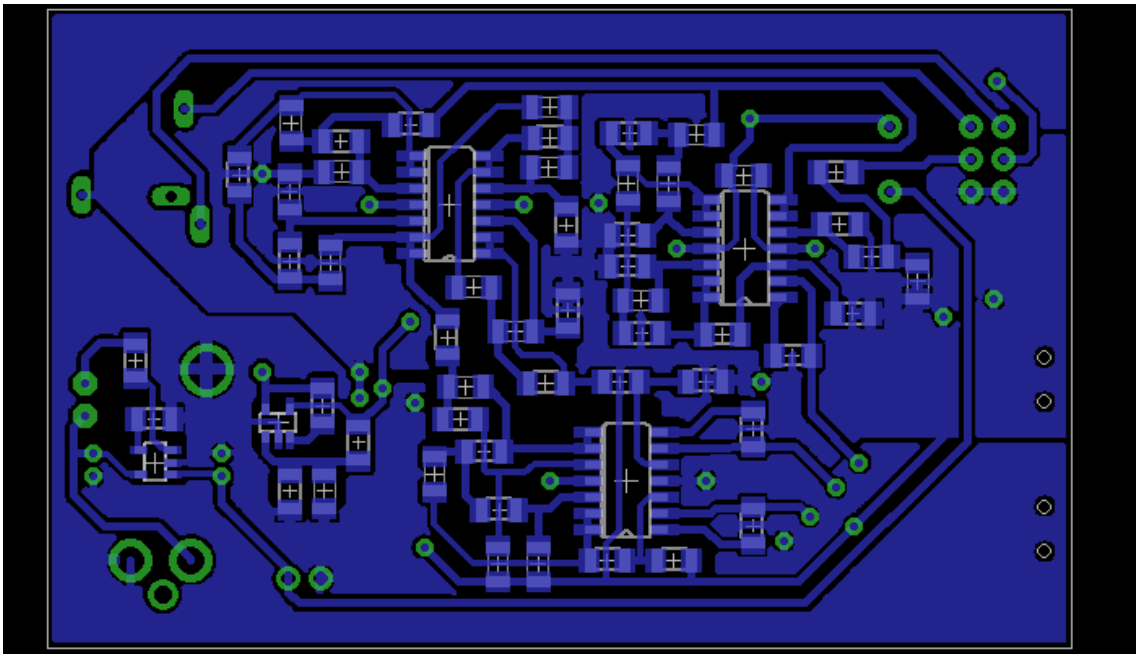
Bottom Layer

PCB Layers

Version 2



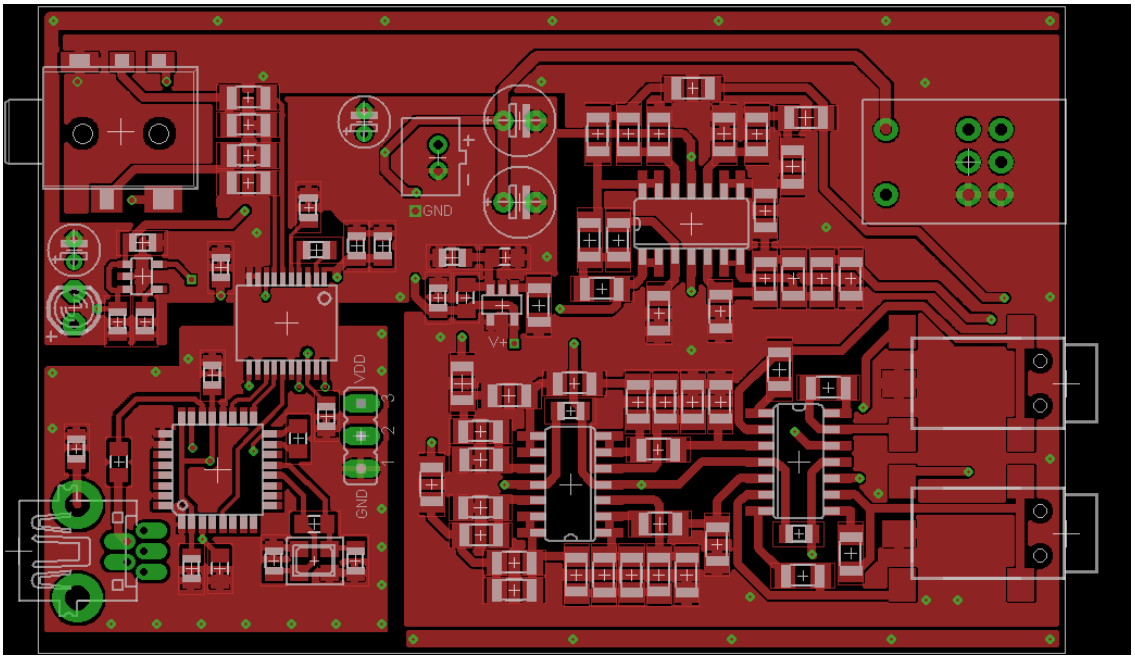
Top Layer



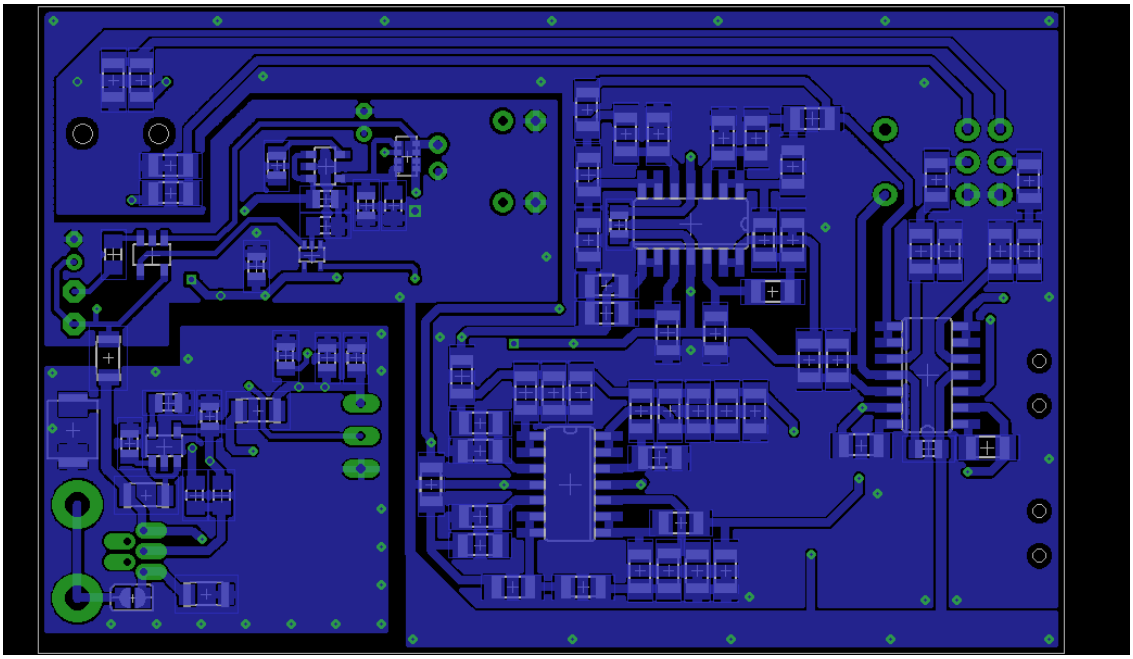
Bottom Layer

PCB Layers

Version 3



Top Layer



Bottom Layer

Resistor Comparison Charts

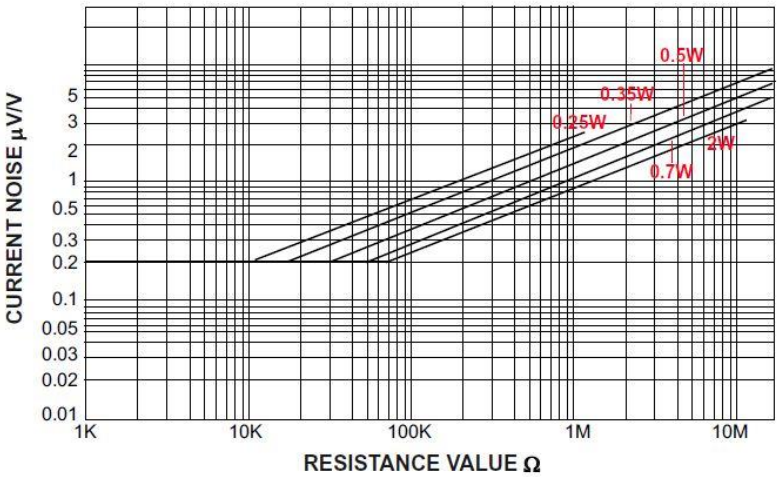


Chart: Resistance vs current noise. Copied from Self (2010) [19]

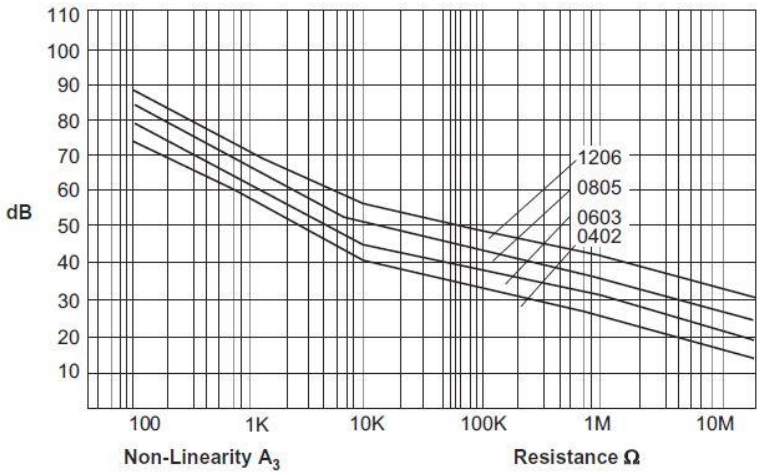


Chart: Thick film resistor nonlinearity. Copied from Self (2010) [19]

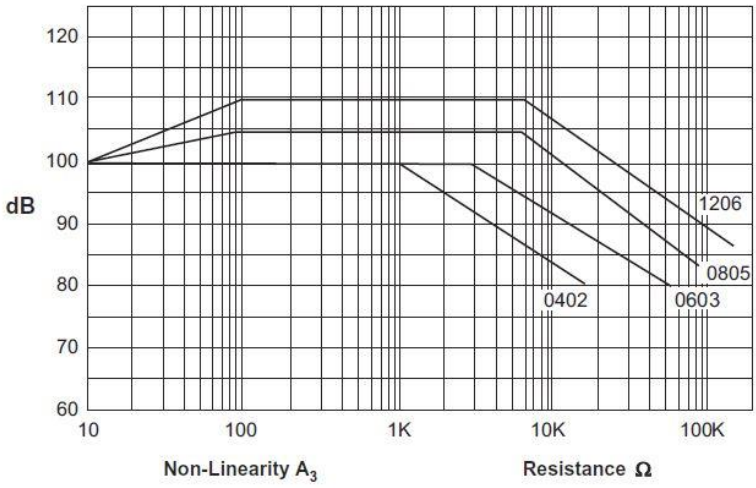


Chart: Thin film resistor nonlinearity. Copied from Self (2010) [19]

Bill of Materials

Version 1

Qty	Value	Package	Parts	Description
12	6.8nF	C1206	C1, C2, C3, C4, C11, C12, C17, C18, C19, C20, C27, C28	CAPACITOR
4	3.3nF	C1206	C13, C14, C29, C30	CAPACITOR
4	10nF	C1206	C15, C16, C31, C32	CAPACITOR
2	4.7uF	E1,8-4	C33, C34	POLARIZED CA-PACITOR
1	220uF	153CLV-0605	C35	POLARIZED CA-PACITOR
5	0.1uF	C1206	C37, C38, C40, C52, C53	CAPACITOR
8	100nF	C1206	C5, C6, C7, C8, C21, C22, C23, C24	CAPACITOR
4	22nF	C1206	C9, C10, C25, C26	CAPACITOR
1		DCJACK_1.3MM	CN2	2.0mm DC Barrel Jack
4	AD8694	SO14	IC1, IC2, IC3, IC4	OP AMP
2	AD8608	SO14	IC5, IC8	OP AMP
1	MCP73831/OT	SOT23-5L	IC6	MCP73831/2
1	AD8615	SOT23-5	IC9	OP AMP
1		LED3MM	LED1	LED
4	5.6k	R1206	R1, R2, R15, R21	RESISTOR
2	150k	R1206	R14, R33	RESISTOR
2	330	R1206	R16, R34	RESISTOR
4	20k	R1206	R19, R20, R37, R38	RESISTOR
12	12k	R1206	R3, R4, R5, R6, R17, R18, R22, R23, R24, R25, R35, R36	RESISTOR
1	4.2	R1206	R41	RESISTOR
2	100k	RTRIMT93XA	R42, R49	Trimm resistor
8	47k	R1206	R43, R44, R45, R46, R47, R48, R57, R58	RESISTOR
1	2.2k	R1206	R55	RESISTOR
1	470	R1206	R56	RESISTOR
4	10k	R1206	R7, R8, R26, R27	RESISTOR
10	100k	R1206	R9, R10, R11, R12, R28, R29, R30, R31, R39, R40	RESISTOR
1	10k	ALPS_RK097	U$1	ALPS RK0971221Z
1	KK2VRND	KK-V-2-RND	U$2	Molex KK
2	4PIN_SMD_JACK	1503_02	U$3, U$4	3.5mm 4 Pin jack connector
1	KLBR4	KLBR4	X2	Jack connector, 3.5 mm

Bill of Materials

Version 2

Qty	Value	Package	Parts	Description
16	110nF	C1206	C1, C2, C3, C4, C13, C14, C15, C16, C17, C18, C19, C20, C29, C30, C31, C32	CAPACITOR
2	4,7uF	E1,8-4	C33, C34	POLARIZED CAPACITOR
1	220uF	153CLV-0605	C35	POLARIZED CAPACITOR
2	1000pF	C1206	C36, C39	CAPACITOR
3	0,1uF	C1206	C37, C38, C40	CAPACITOR
1	100nF	C1206	C41	CAPACITOR
8	560nF	C1206	C5, C6, C7, C8, C21, C22, C23, C24	CAPACITOR
2	2.2uF	C1206	C52, C53	CAPACITOR
8	220nF	C1206	C9, C10, C11, C12, C25, C26, C27, C28	CAPACITOR
1		DCJACK_1.3MM	CN2	2.0mm DC Barrel Jack
6	AD8608	SO14	IC1, IC2, IC3, IC4, IC5, IC8	OP AMP
1	MCP73831/OT	SOT23-5L	IC6	MCP73831/2
1	AD8615	SOT23-5	IC9	OP AMP
1		0207/7	L_FERRITE_BEAD1	A Ferrite Bead
1		LED3MM	LED1	LED
12	1k	R1206	R1, R2, R3, R4, R5, R6, R15, R21, R22, R23, R24, R25	RESISTOR
2	330	R1206	R13, R32	RESISTOR
2	22k	R1206	R14, R33	RESISTOR
2	3.3k	R1206	R16, R34	RESISTOR
5	2k	R1206	R17, R18, R35, R36, R55	RESISTOR
4	4,7k	R1206	R19, R20, R37, R38	RESISTOR
10	10k	R1206	R39, R40, R43, R44, R45, R46, R47, R48, R57, R58	RESISTOR
7	4.6	R1206	R41, R50, R51, R52, R53, R54, R59	RESISTOR
2	33k	R1206	R42, R49	RESISTOR
1	470	R1206	R56	RESISTOR
4	2,36k	R1206	R7, R8, R26, R27	RESISTOR
8	12k	R1206	R9, R10, R11, R12, R28, R29, R30, R31	RESISTOR
1	10k	ALPS_RK097	U$1	ALPS RK0971221Z
1	KK2VRND	KK-V-2-RND	U$2	Molex KK
2	4PIN_SMD_JACK	1503_02	U$3, U$4	
1	KLBR4	KLBR4	X2	Jack con-nector, 3.5 mm

Bill of Materials

Version 3

Qty	Value	Package	Parts	Description
1		JST-PH-2	BATTERY	JST 2-Pin Connector
11	100nF	C0805K	C1, C12, C23, C24, C27, C31, C68, C69, C71, C72, C78	CAPACITOR
20	100nF	C1206	C13, C14, C15, C16, C17, C18, C19, C20, C21, C22, C52, C53, C54, C55, C56, C57, C58, C59, C60, C61	CAPACITOR
2	10uF	C1206	C2, C3	CAPACITOR
2	22pF	C1206	C25, C26	CAPACITOR
6	10uF	C0805K	C28, C38, C66, C67, C74, C75	CAPACITOR
4	1uF	C0805K	C29, C32, C62, C65	CAPACITOR
2	4.7uF	E1,8-4	C30, C39	POLARIZED CA-PACITOR
4	2.2uF	C0805K	C33, C37, C70, C73	CAPACITOR
3	10nF	C0805K	C34, C35, C36	CAPACITOR
12	7.5nF	C1206	C4, C5, C6, C8, C9, C10, C42, C43, C45, C47, C48, C50	CAPACITOR
2	330uF	E2,5-6E	C40, C41	POLARIZED CA-PACITOR
4	15k	R1206	C44, C49, R13, R17	RESISTOR
2	33pF	C0805K	C63, C64	CAPACITOR
4	15nF	C1206	C7, C11, C46, C51	CAPACITOR
2	2.2nF	C1206	C76, C77	CAPACITOR
1		SMBJ	D1	Suppressor diode
1	SJ-3515-SMT	SJ-3515-SMT	J1	3.5 Stereo Audio Jack (SMT)
2	4PIN_SMD_JACK	1503_02	J2, J3	
1		USB-MINIB	J4	USB Connectors
1		JP2	JP1	JUMPER
4		L3216C	L1, L2, L3, L4	FERRITE
1		LED3MM	LED1	LED
1	NTHD3101F	CHIPFET-1206A	Q1	Power MOSFET and Schottky Diode
10	330	R1206	R1, R2, R3, R4, R27, R28, R29, R30, R39, R40	RESISTOR
12	7.5k	R1206	R11, R12, R14, R15, R16, R18, R47, R48, R49, R50, R51, R52	RESISTOR
12	16.5k	R1206	R19, R20, R22, R23, R24, R26, R53, R54, R56, R57, R58, R60	RESISTOR
4	33k	R1206	R21, R25, R55, R59	RESISTOR

2	10k	R0805	R41, R42	RESISTOR
1	470	R0805	R43	RESISTOR
1	2.2k	R0805	R45	RESISTOR
2	100k	R0805	R46, R61	RESISTOR
8	10k	R1206	R5, R6, R7, R8, R31, R32, R33, R34	RESISTOR
1	1M	R0805	R62	RESISTOR
2	1.5k	R0805	R63, R66	RESISTOR
2	22	R0805	R64, R65	RESISTOR
2	470	R1206	R67, R68	RESISTOR
7	5.6	R1206	R9, R10, R35, R36, R37, R38, R44	RESISTOR
1		SJ	SJ1	SMD solder JUMPER
6	AD8608	SO14	U1, U2, U3, U4, U11, U12	OP AMP
1	MAX4517	SC70-5	U10	Single Analog Switch
1	PCM2707	TQFP32-08	U13	
1	PCM5102A	SSOP20BU	U14	SoundPlus(TM) Stereo Audio DAC
1	AD8615	SOT23-5	U5	Single Operational Amplifier
1	MCP73831/OT	SOT23-5L	U6	MCP73831/2
3	LP298XS	SOT23-5L	U7, U8, U9	
1	10k	ALPS_RK097	VR1	ALPS RK0971221Z
1	12MHz	NX3225	X1	4-pin 3.2 x 2.5mm crystal

Test Equipment Specifications

Brymen 867s Multimeter Specifications

Manufacturer	BRYMEN
Meter type	digital multimeter
Sampling	5x/s
DC voltage measuring range	0,01m...500m/5/50/500/1000V
DC voltage measuring accuracy	±(0.03% + 2 digits)
AC voltage measuring range	0,01m...500m/5/50/500/1000V
AC voltage measuring accuracy	±(0.8% + 60 digits)
DC current measuring range	0,01µ...500µ/5m/50m/500m/5/10A
DC current measuring accuracy	±(0,1% + 20 digits)
AC current measuring range	0,01µ...500µ/5m/50m/500m/5/10A
AC current measuring accuracy	±(1% + 40 digits)
True effective value measurement	True RMS AC, True RMS AC+DC
Resistance measuring range	0,01...500/5k/50k/500k/5M/50MΩ
Resistance measuring accuracy	±(0.1% + 6 digits)
Capacitance measuring range	10p...50n/500n/5µ/50µ/500µ/5m/25mF
Capacitance measuring accuracy	±(2,5% + 3 digits)
Frequency measuring range	10...200kHz
Frequency measuring accuracy	±(0,02% + 4 digits)

Test Equipment Specifications

Velleman PCSU200 USB Oscilloscope Specifications

General:

Markers for: amplitude/voltage and frequency/time
Input coupling: DC, AC and GND
Supply from USB port (500mA)
Dimensions: 100 x 100 x 35mm / 3.9 x 3.9 x 1.4"

Function Generator:

Amplitude range: 200mVpp to 8Vpp @ 1kHz// no load
Vertical resolution: 8 bits
Square wave rise/fall time: 0.3µs
Sample rate: 25MHz
Typical sine wave distortion (THD): < 1%
Output impedance: 50ohm
Frequency range: from 0.5Hz to 500 kHz (sine 1MHz)

Oscilloscope:

Bandwidth: two channels DC to 12 MHz ±3dB
Input impedance: 1 Mohm / 30pF
Maximum input voltage: 30V (AC + DC)
Time base: 0.1µs to 500ms per division
Input range: 10mV to 3V/division
Input sensitivity: 0.3mV display resolution
Record length: 4K samples / channel
Sampling frequency: 250Hz to 25MHz

Transient Recorder:

Timescale: 20ms/Div to 2000s/Div
Max record time: 9.4hour/screen
Max number of samples: 100/s
Min number of samples: 1 sample/20s

Bode Plotter:

Voltage range: 10mV, 30mV, 0.1V, 0.3V, 1V, 3V
Frequency range: 1kHz , 10kHz, 100kHz, 500kHz
Frequency start: 1Hz,10Hz, 100Hz, 1kHz, 10kHz, 100kHz

Test Equipment Specifications

ESI ESP1010 Sound Card Specifications

<Analog Audio>

1. Analog Inputs
   1) Connector Type: 1/4" female TRS-type, balanced or unbalanced (ch 1~8)
   2) Peak level: 0dBFS @ +6.5dBV (-10dBV nominal)
   3) Impedance: 10k ohms minimum
   4) Att. & Gain Control:-60dB ~ +15dB (0.5dB step size) *ch1, 2ch only

2. Analog Outputs
   1) Connector Type: 1/4" female TS-type, unbalanced (ch 1~8)
   2) Peak level: +6.2dBV @0dBFS (-10dBV nominal)
   3) Impedance: 100 ohms
   4) Attenuation Control: -60dB ~ 0dB (0.5dB step size)

<Digital Audio>

1. Internal 20ch /36-bit Digital Mixer (Input 10ch/Output 10ch)

2. Sample rate supports: (22.05,24)*,32,44.1,48,88.2,96kHz : *analog only

3. A/D Converter
   1) Signal to Noise Ratio: 107dB (A-weighted) @ fs=48kHz
   2) Dynamic Range: 107dB (-60dBFS with A-weighted) @ fs=48kHz
   3) S/(N+D)(-1dB):100dB @ fs=48kHz
   4) Interchannel Isolation: -110dB
   5) Resolution: 24-Bit

4. D/A Converter
   1) Signal to Noise Ratio: 112dB (A-weighted) @ fs=44.1kHz
   2) Dynamic Range (S/N): 112dB (60dBFS with A-weighted) @ fs=44.1kHz
   3) THD+N: -94dB @ fs=44.1kHz
   4) Interchannel Isolation: -100dB
   5) Resolution: 24-Bit / 96kHz

5. Digital Input
   1) Connector Type: RCA(provided via I/O cable)
   2) Format: IEC-60958 Consumer (S/PDIF coaxial)
   3) Sampling Rate: 44.1,48,88.2,96kHz
   4) Resolution: 24-Bit

6. Digital Output
   1) Connector Type: RCA(provided via I/O cable), Optical (on board)
   2) Format: IEC-60958 Consumer (S/PDIF coaxial)
   3) Sampling Rate: 44.1,48,88.2,96kHz
   4) Resolution: 24-Bit

Test Results

Full Output Impedance Calculation Tables

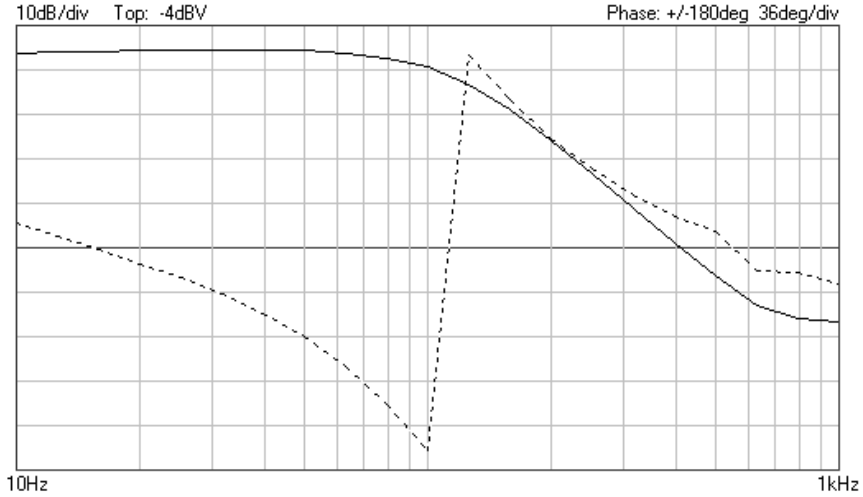
	Output Impedance					
	Left Channel			**Right Channel**		
	High @ 10kHz	**Mid @ 1kHz**	**Low @ 50 Hz**	**High @ 10kHz**	**Mid @ 1kHz**	**Low @ 50 Hz**
1st version						
Vu (un-loaded voltage)	100.23mV	100.04mV	100.48mV	100.7mV	100.55mV	100.07mV
Vl (with Rload = 10.00 Ω)	87.34 mV	86.89 mV	95.1mV	98.8mV	99.23mV	98.42mV
Imped-ance (Rl * (Vu/Vl-1)	1.476 Ω	1.513 Ω	0.566 Ω	0.188 Ω	0.133 Ω	0.232 Ω

2nd Version						
Vu (un-loaded voltage)	100.55mV	100.44mV	100.03mV	100.4mV	100.23mV	100.8mV
Vl (with Rload = 10.00 Ω)	99.68mV	99.51mV	97.18mV	98.86mV	98.86mV	99.3mV
Imped-ance (Rl * (Vu/Vl-1)	0.087 Ω	0.093 Ω	0.293 Ω	0.156 Ω	0.139 Ω	0.151 Ω

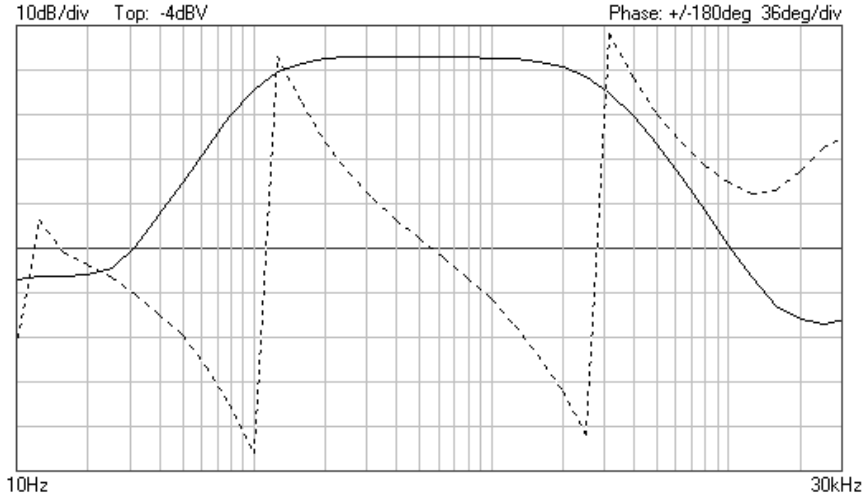
3rd version						
Vu (un-loaded voltage)	100.59mV	100.52mV	100.16mV	100.43mV	100.36mV	100.53mV
Vl (with R load = 10.00 Ω)	100.13mV	99.88mV	99.35mV	99.95mV	99.45mV	99.99mV
Imped-ance (Rl * (Vu/Vl-1)	0.046 Ω	0.064 Ω	0.082 Ω	0.048 Ω	0.092 Ω	0.054 Ω

Test Results

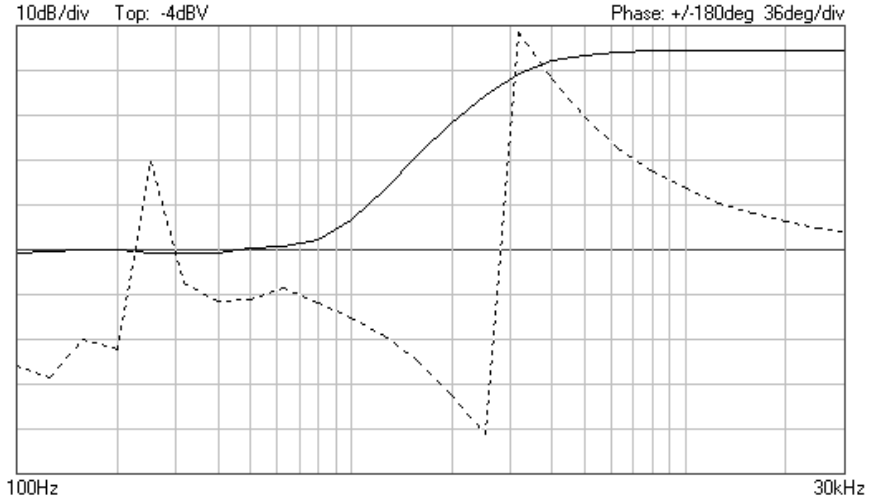
Version 1 – Crossover Output Bode Plots



Left channel low pass output



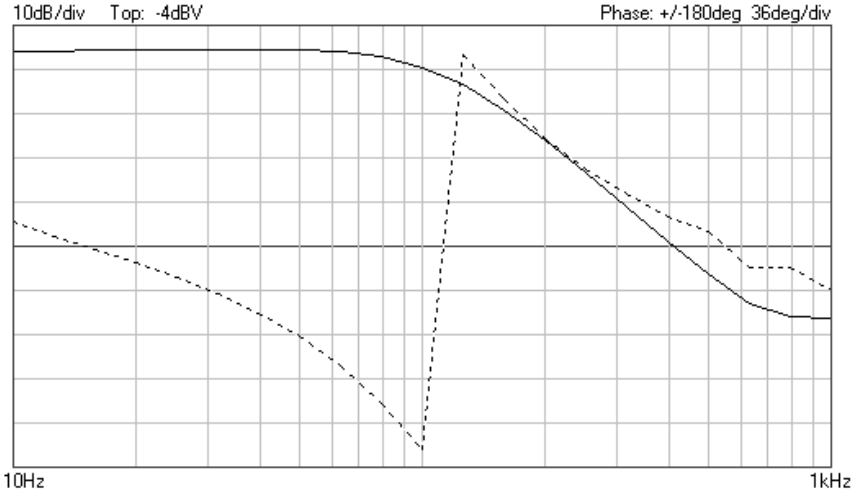
Left channel band pass output



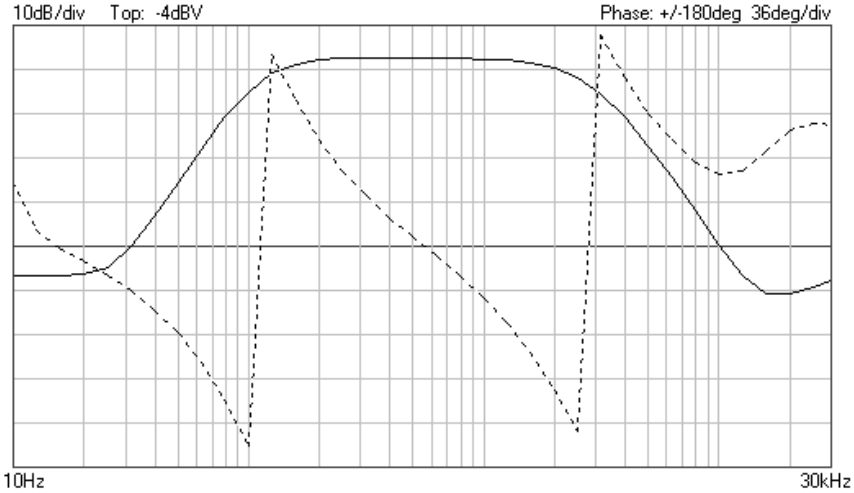
Left channel high pass output

Test Results

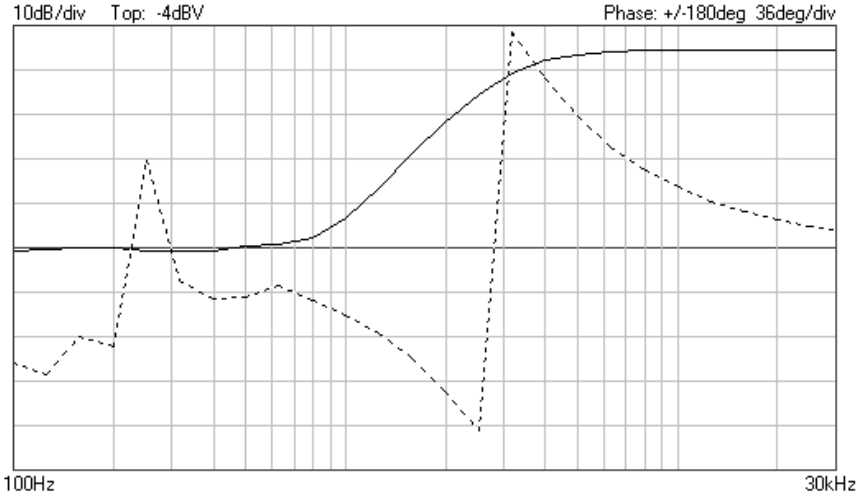
Version 1 – Crossover Output Bode Plots



Right channel low pass output



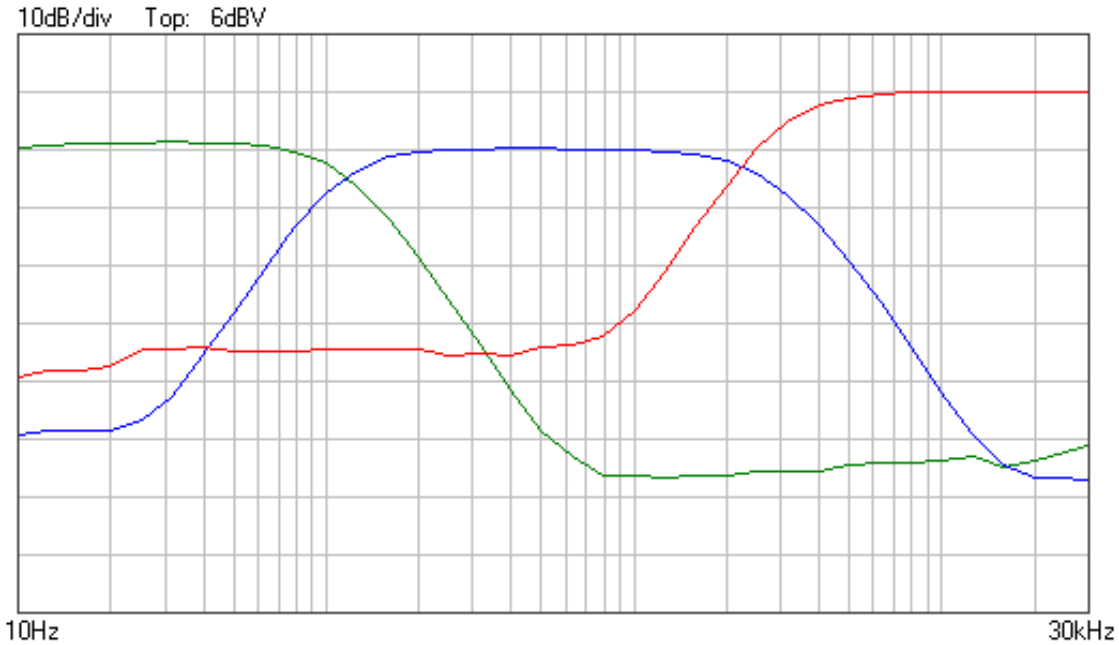
Right channel band pass output



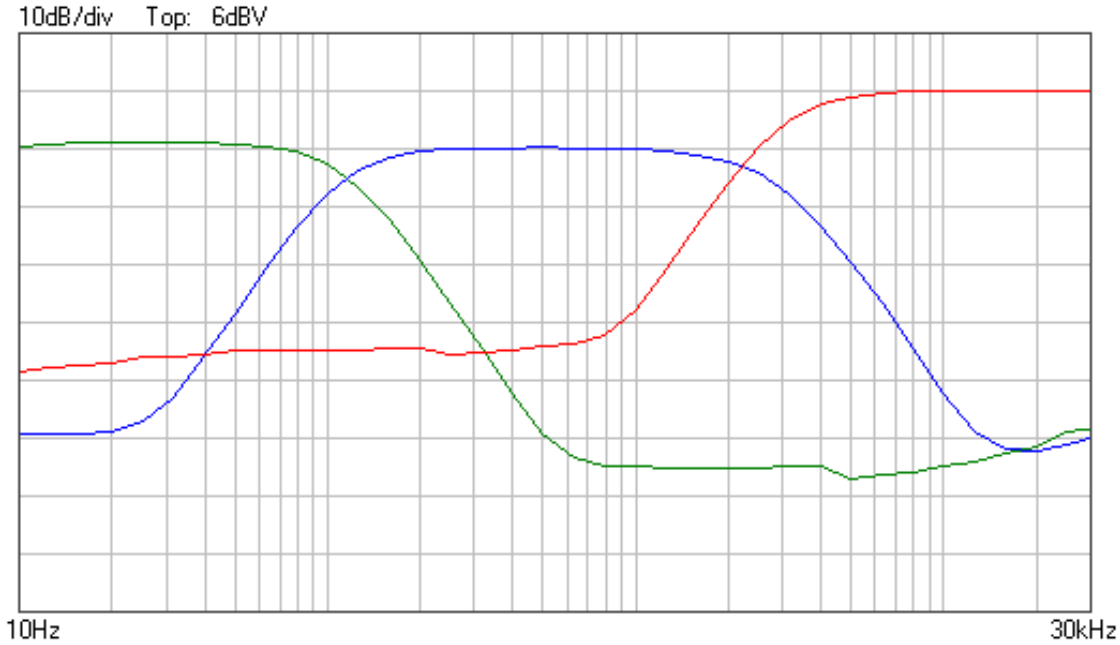
Right channel high pass output

Test Results

Version 1 – Crossover Output Bode Plots



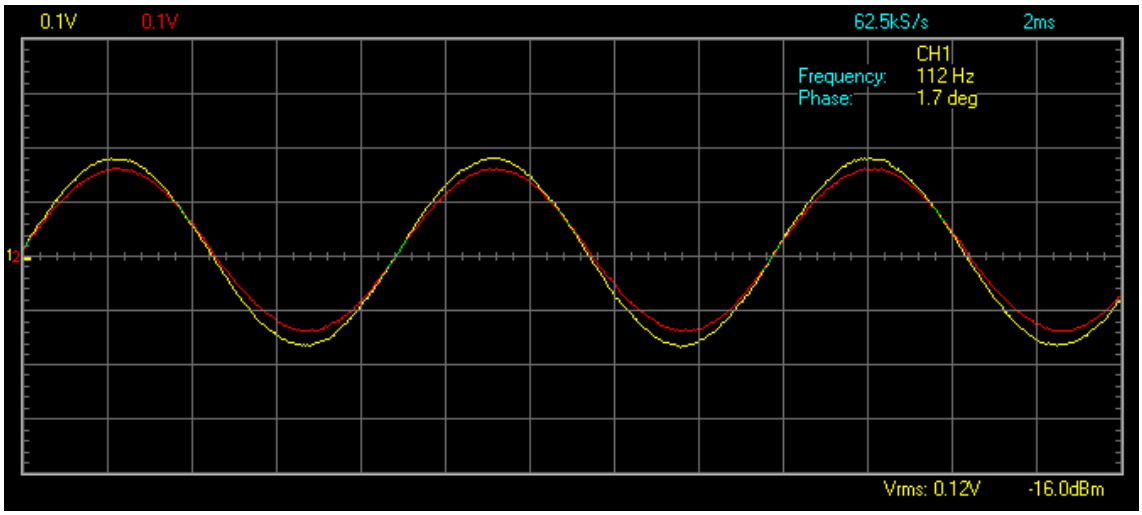
Left channel all outputs



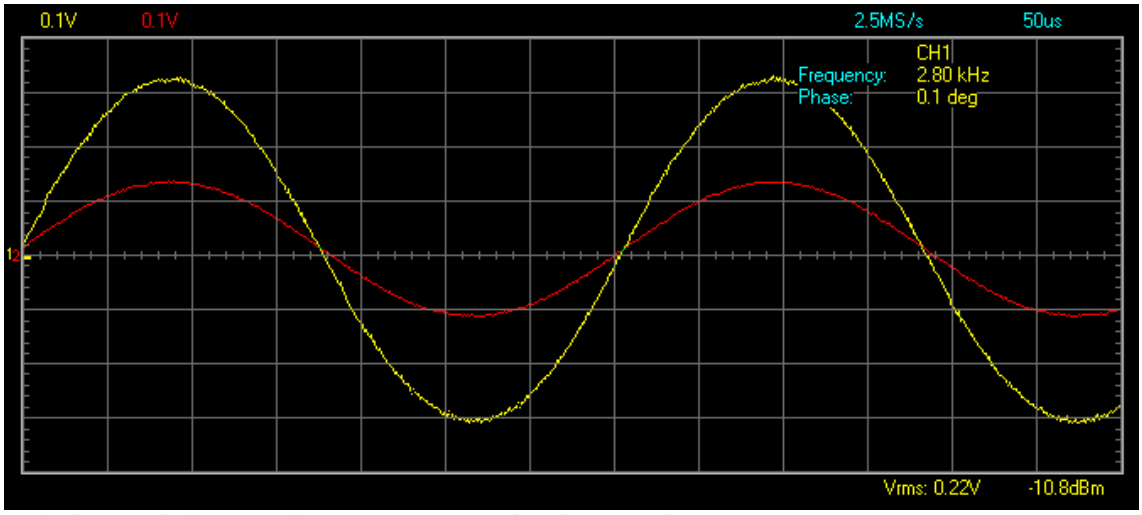
Right channel all outputs

Test Results

Version 1 – Phase Differences Between Outputs



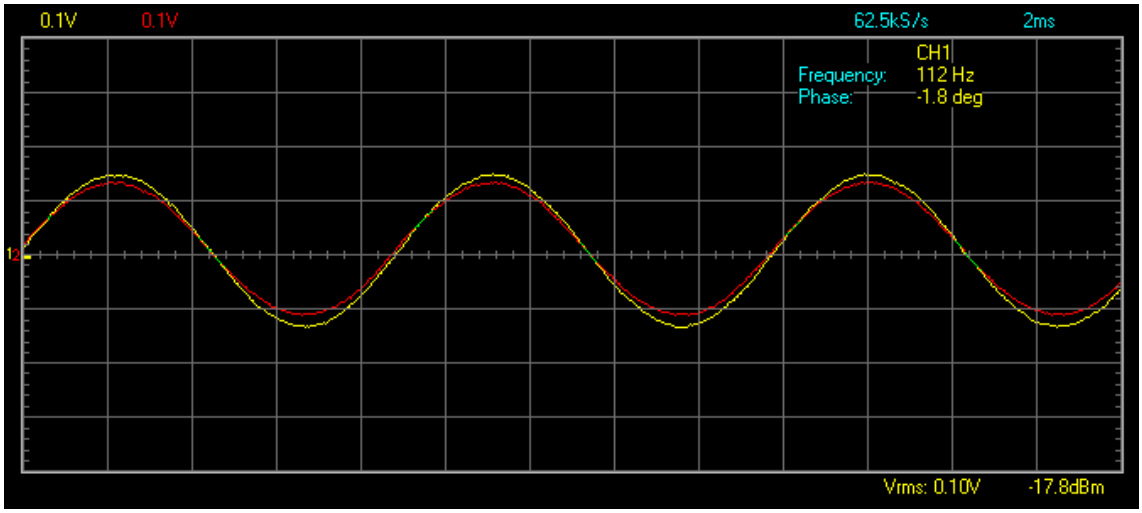
Left channel phase at low crossover point (112Hz)



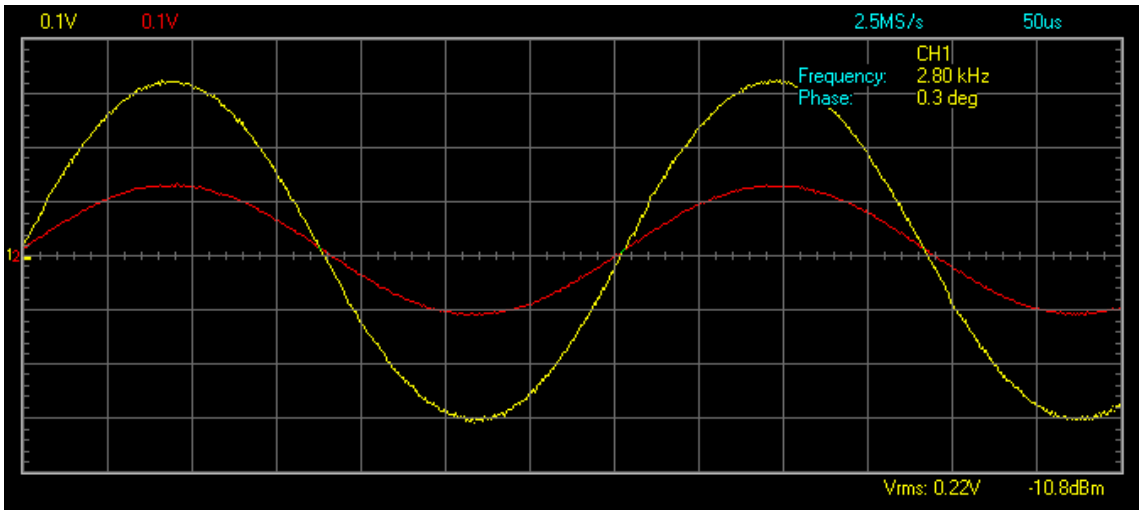
Left channel phase at low crossover point (2800Hz)

Test Results

Version 1 – Phase Differences Between Outputs



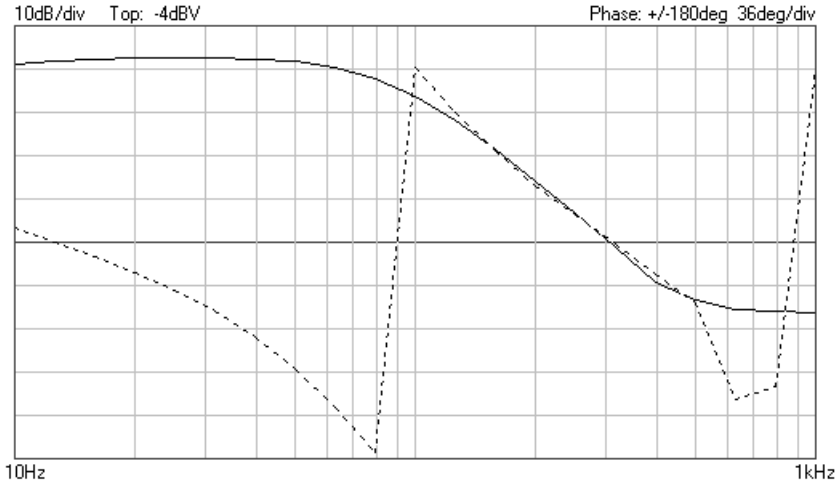
Right channel phase at low crossover point (112Hz)



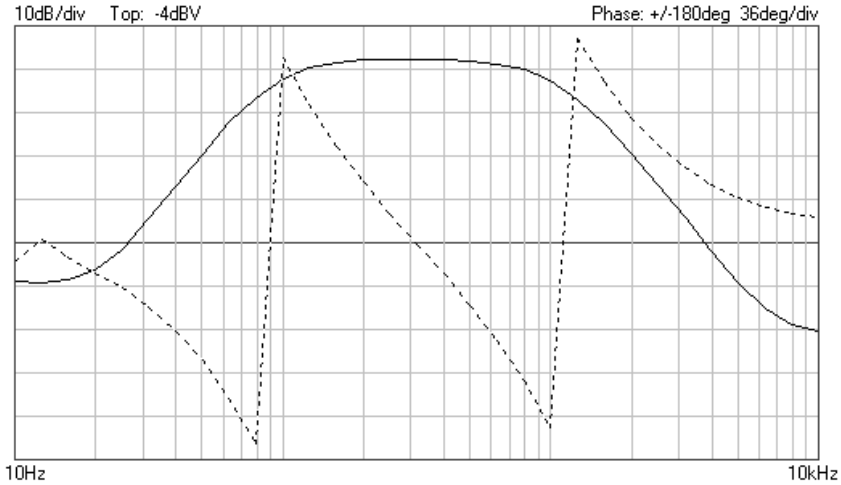
Right channel phase at low crossover point (2800Hz)

Test Results

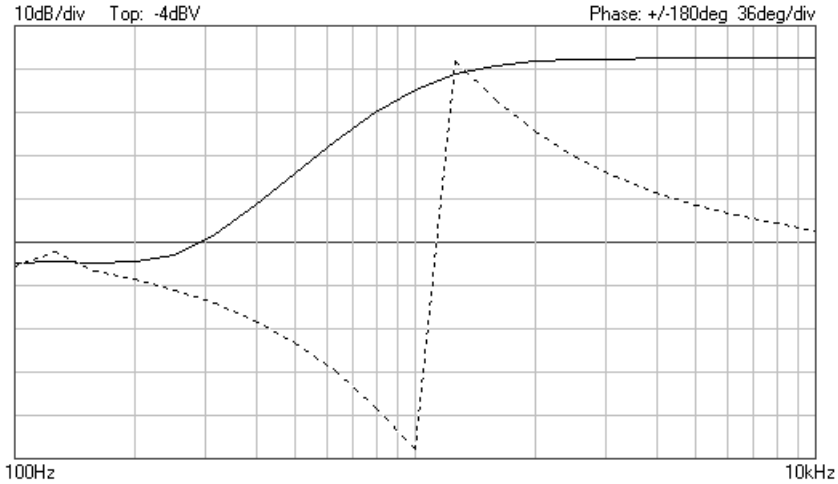
Version 2 – Crossover Output Bode Plots



Left channel low pass output



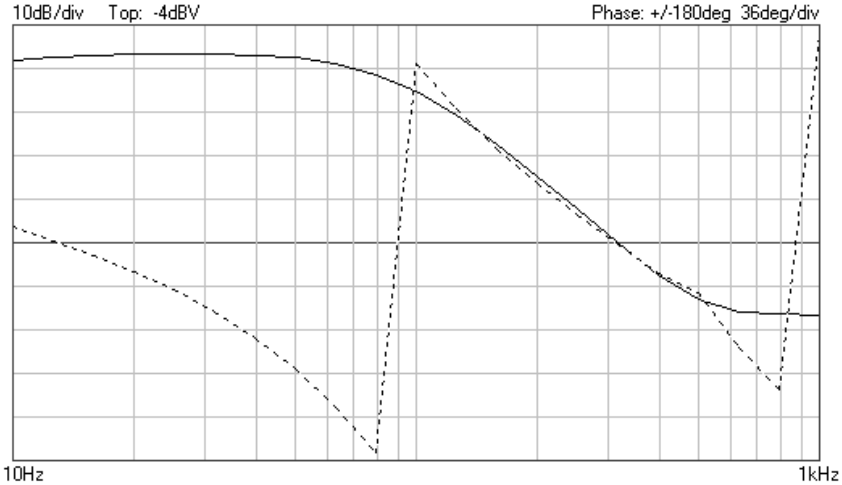
Left channel band pass output



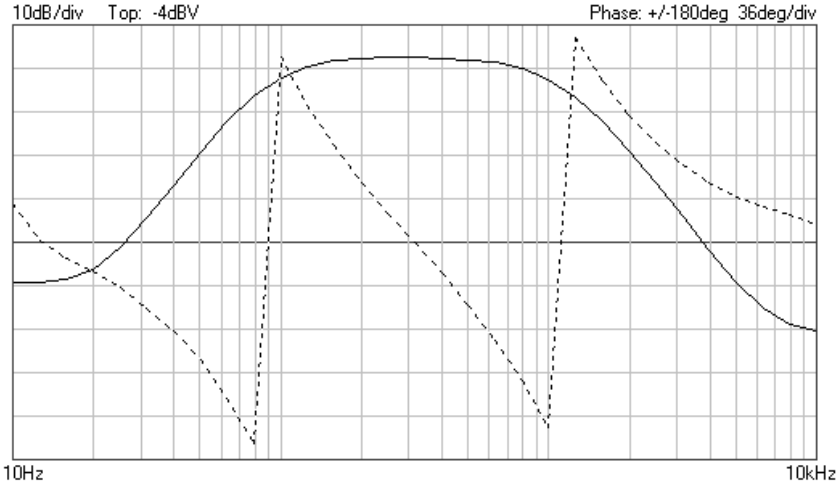
Left channel high pass output

Test Results

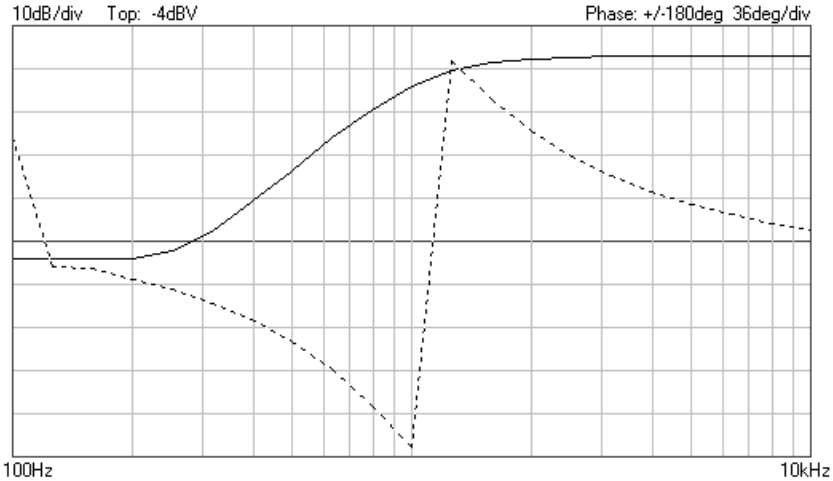
Version 2 – Crossover Output Bode Plots



Right channel low pass output



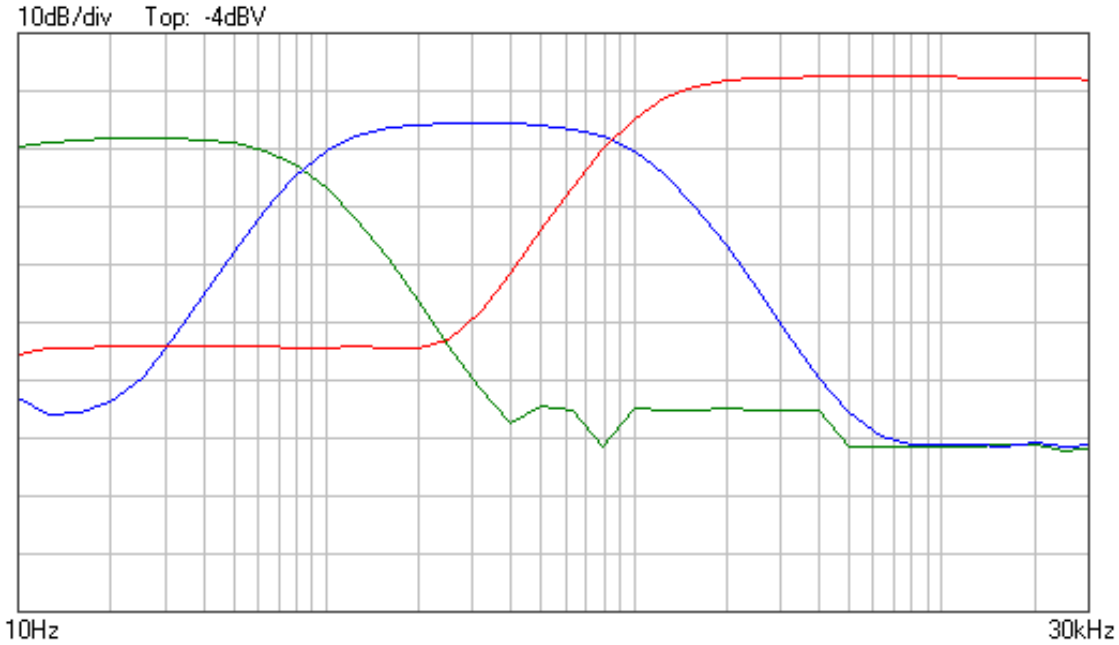
Right channel band pass output



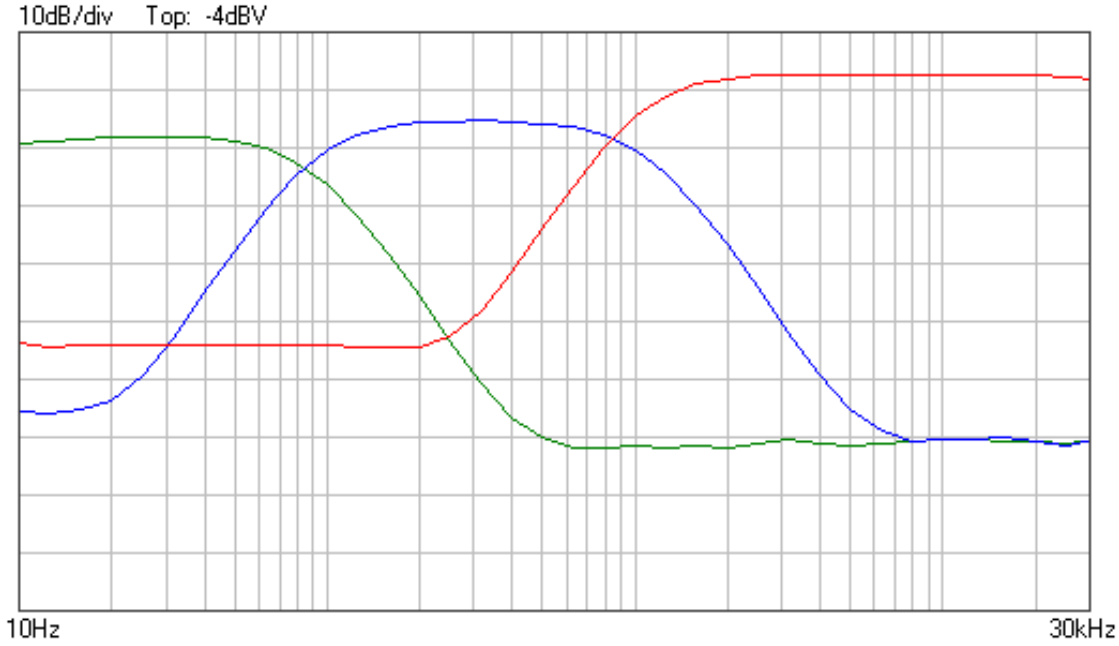
Right channel high pass output

Test Results

Version 2 – Crossover Output Bode Plots



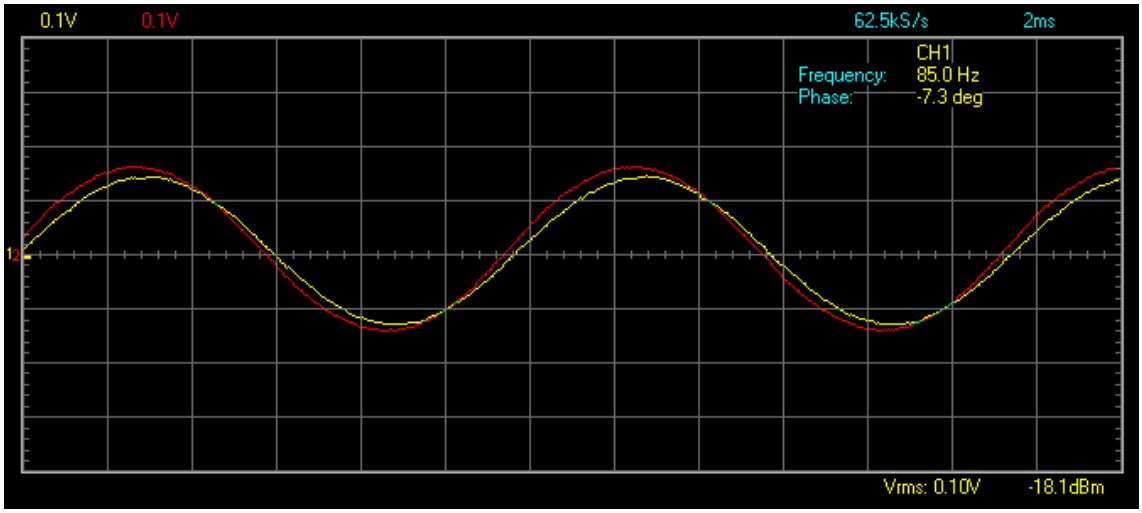
Left channel all outputs



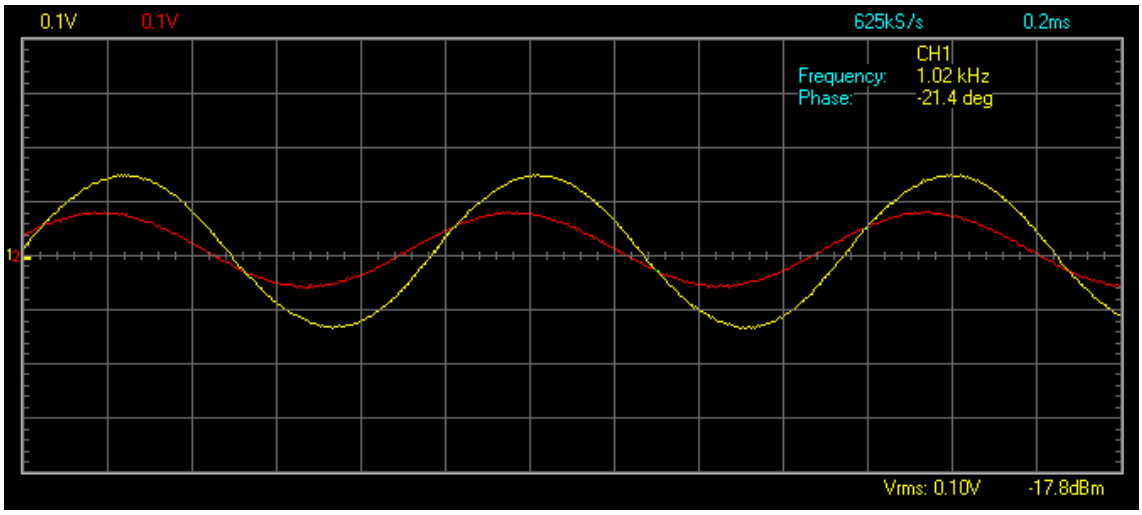
Right channel all outputs

Test Results

Version 2 – Phase Differences Between Outputs



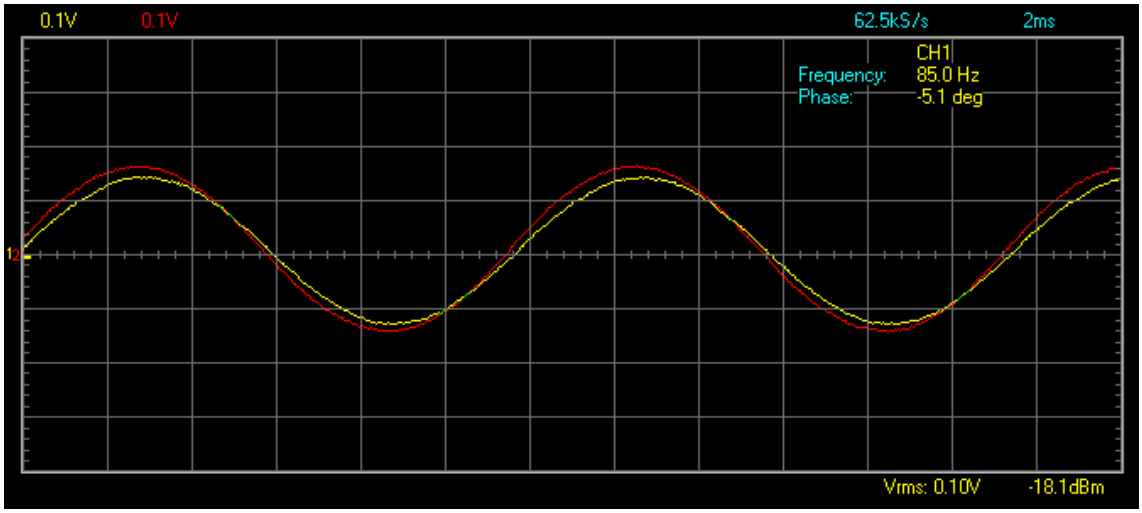
Left channel phase at low crossover point (85Hz)



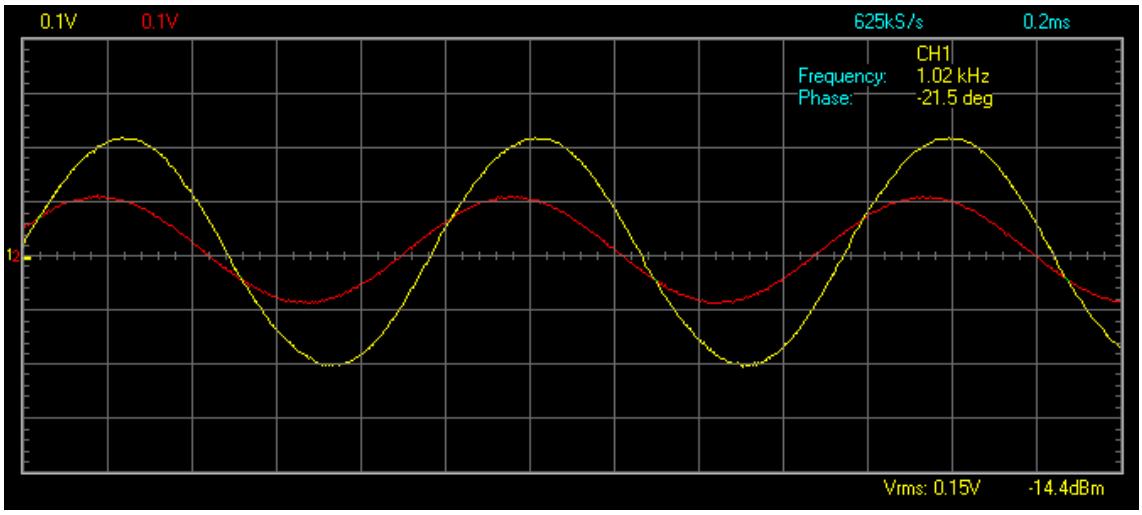
Left channel phase at low crossover point (1023Hz)

Test Results

Version 2 – Phase Differences Between Outputs



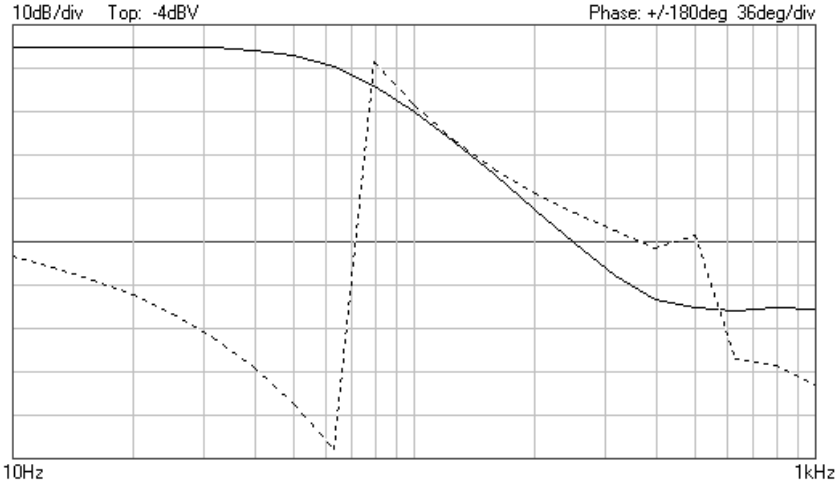
Right channel phase at low crossover point (85Hz)



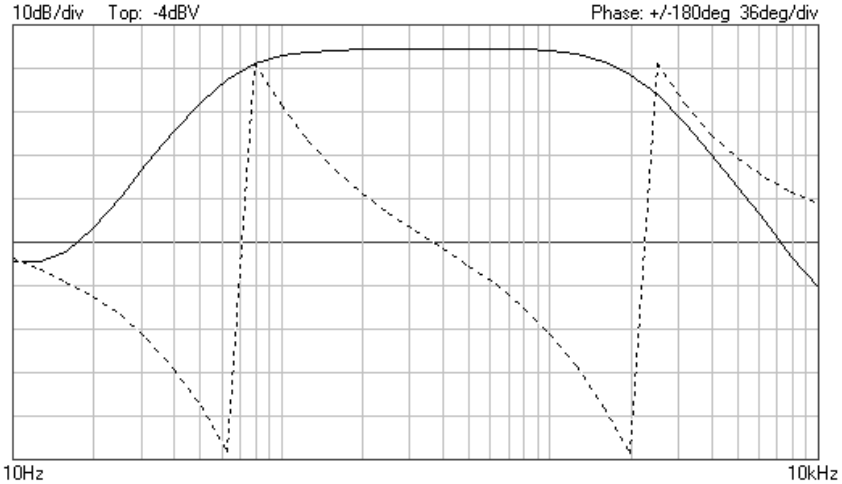
Right channel phase at low crossover point (1023Hz)

Test Results

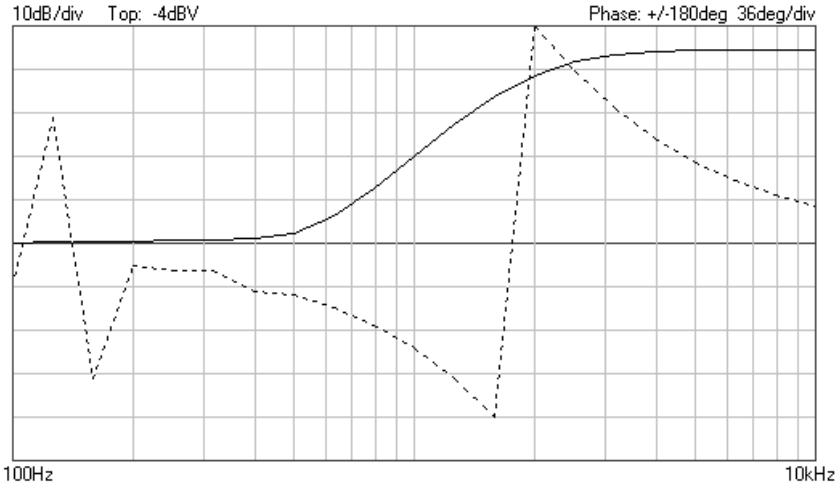
Version 3 – Crossover Output Bode Plots



Left channel low pass output



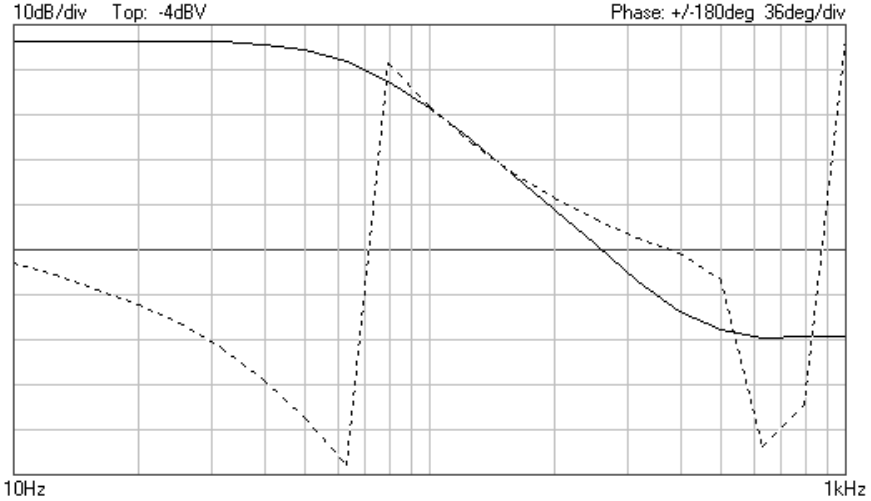
Left channel band pass output



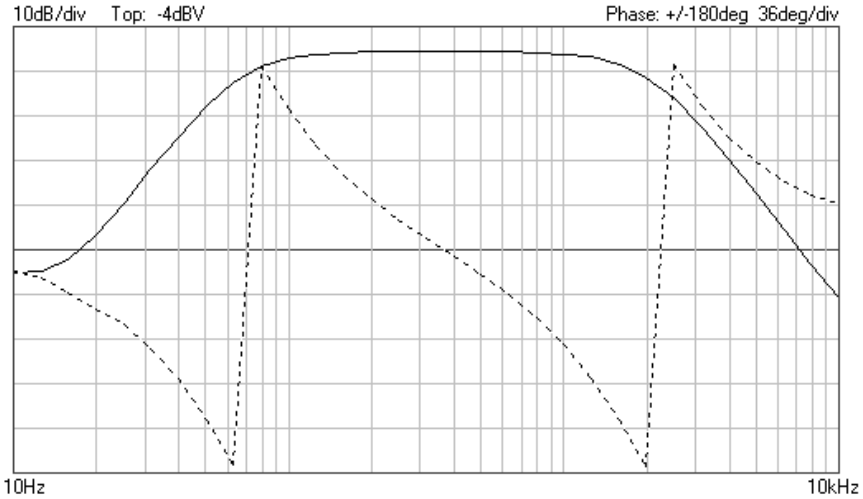
Left channel high pass output

Test Results

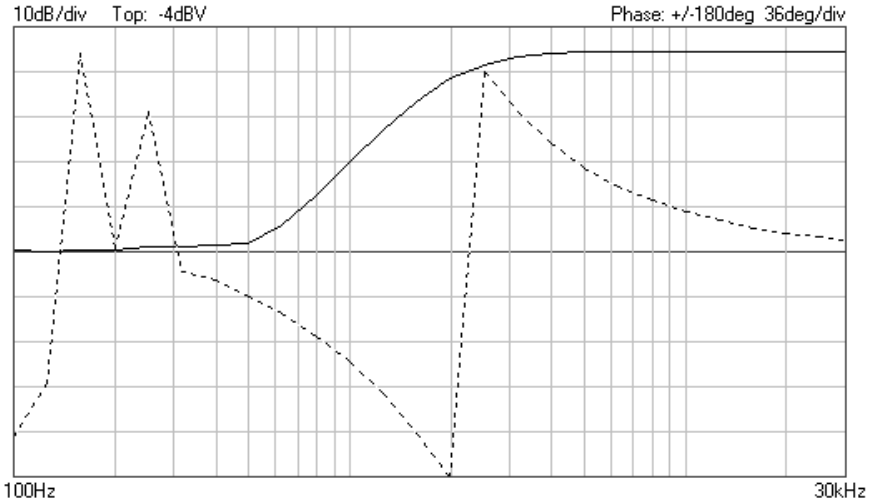
Version 3 – Crossover Output Bode Plots



Right channel low pass output



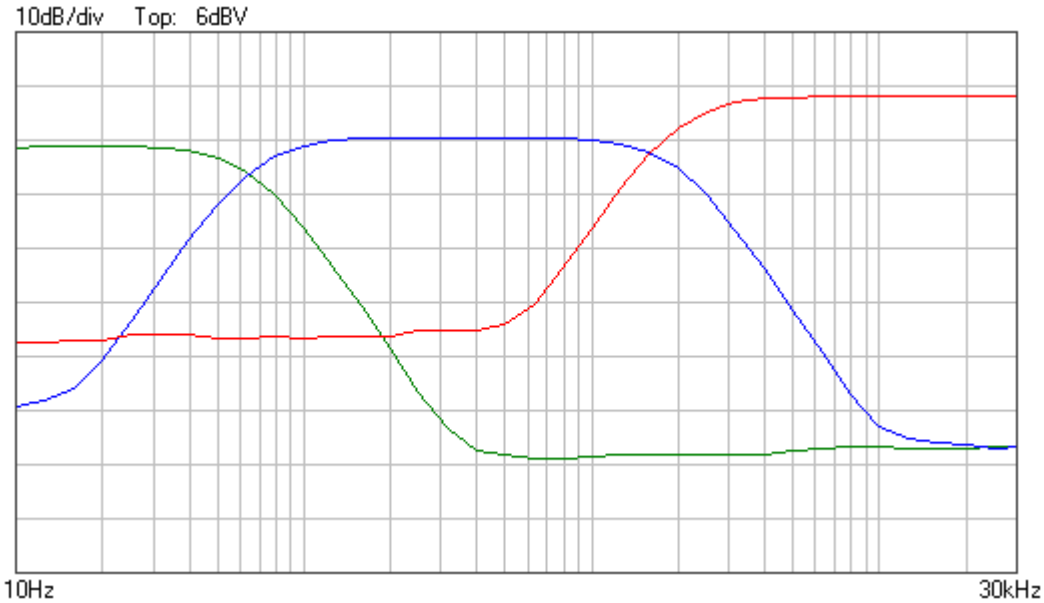
Right channel band pass output



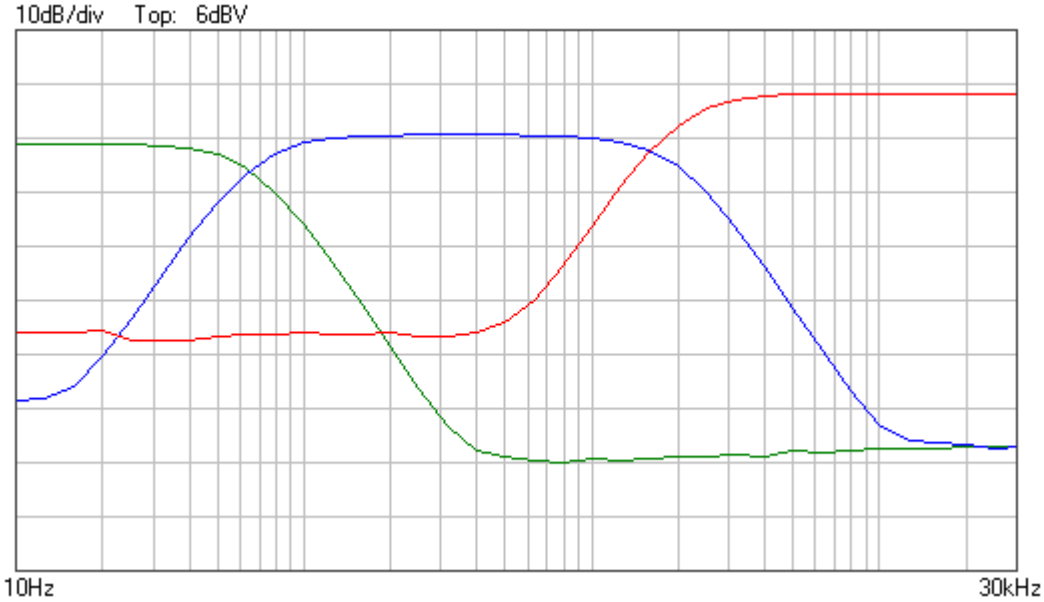
Right channel high pass output

Test Results

Version 3 – Crossover Output Bode Plots



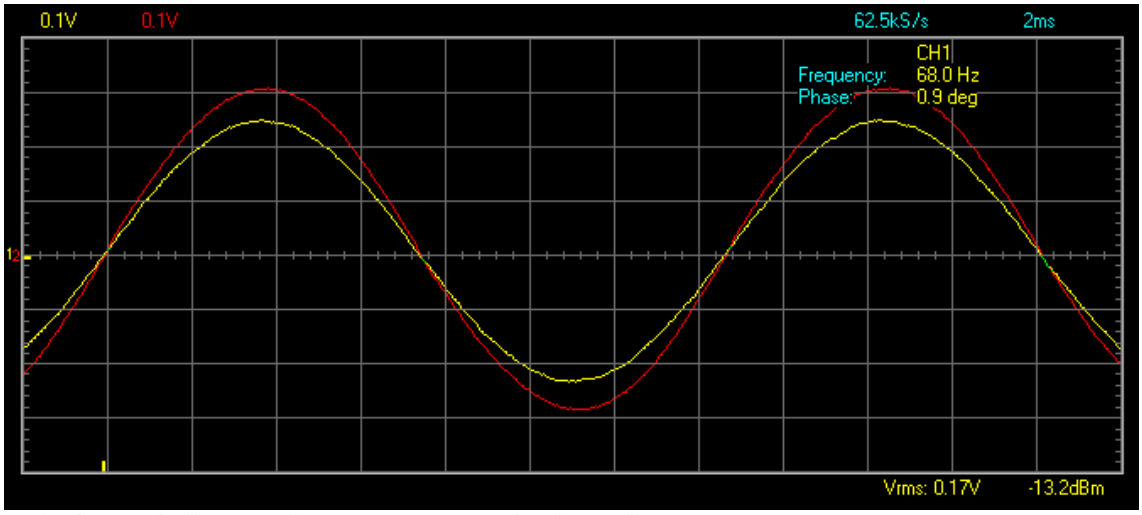
Left channel all outputs



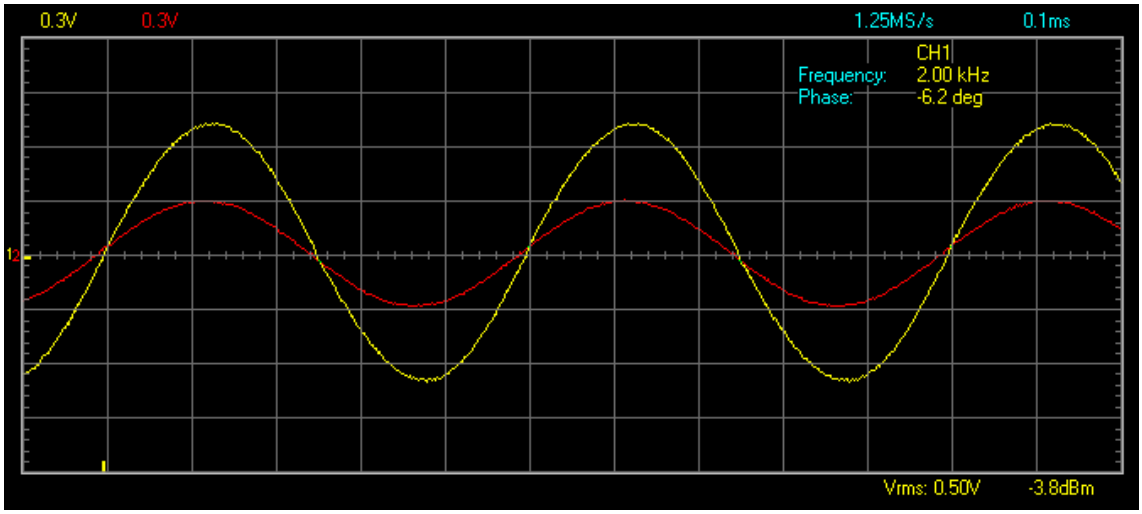
Right channel all outputs

Test Results

Version 3 – Phase Differences Between Outputs



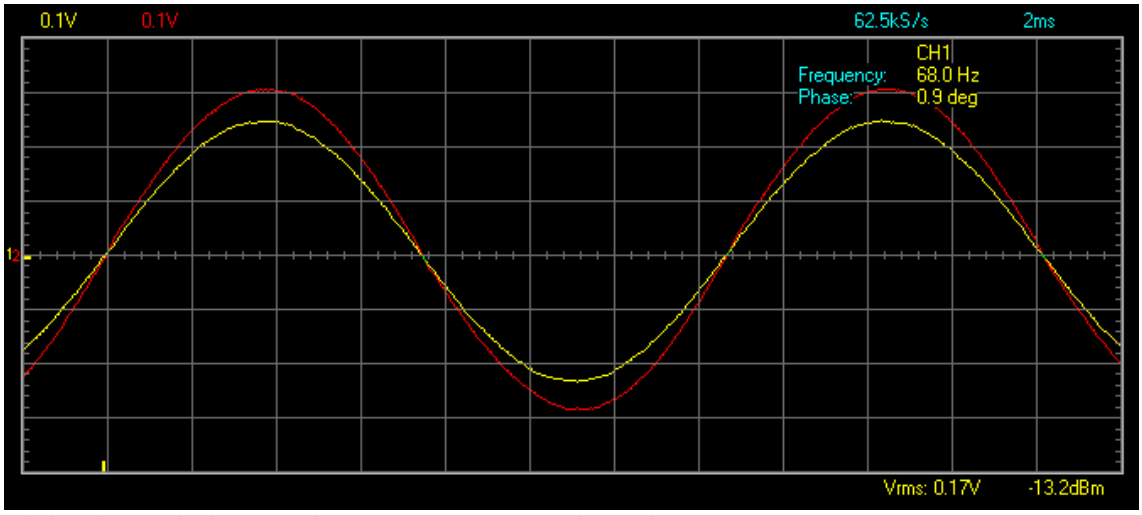
Left channel phase at low crossover point (68Hz)



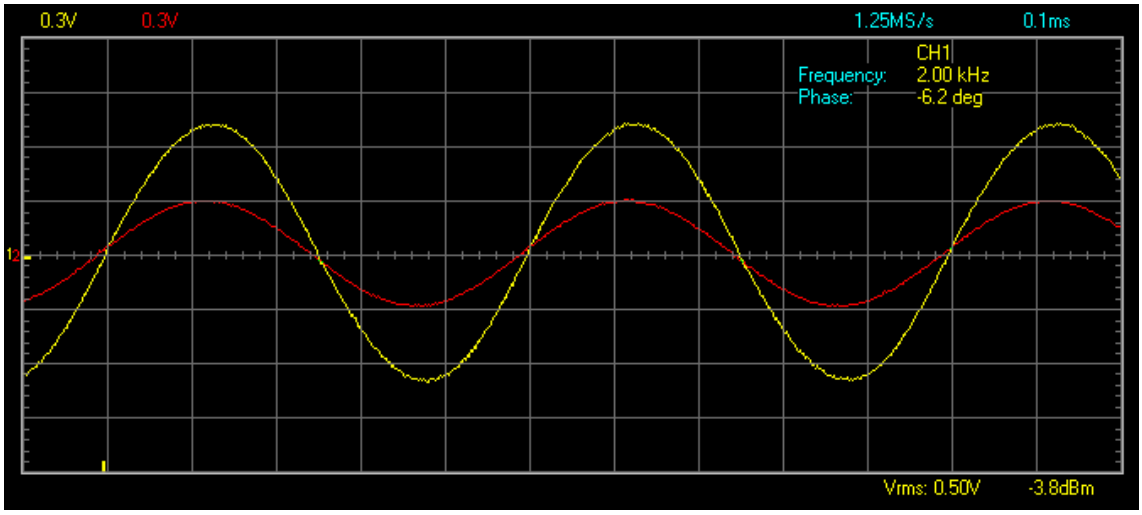
Left channel phase at high crossover point (2002Hz)

Test Results

Version 3 – Phase Differences Between Outputs



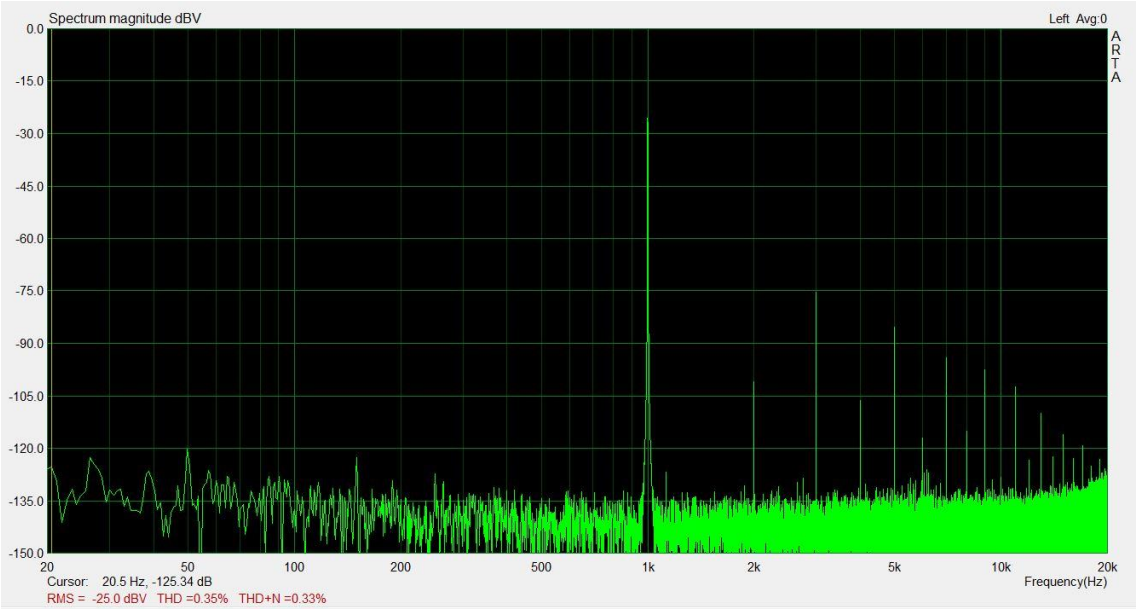
Right channel phase at low crossover point (68Hz)



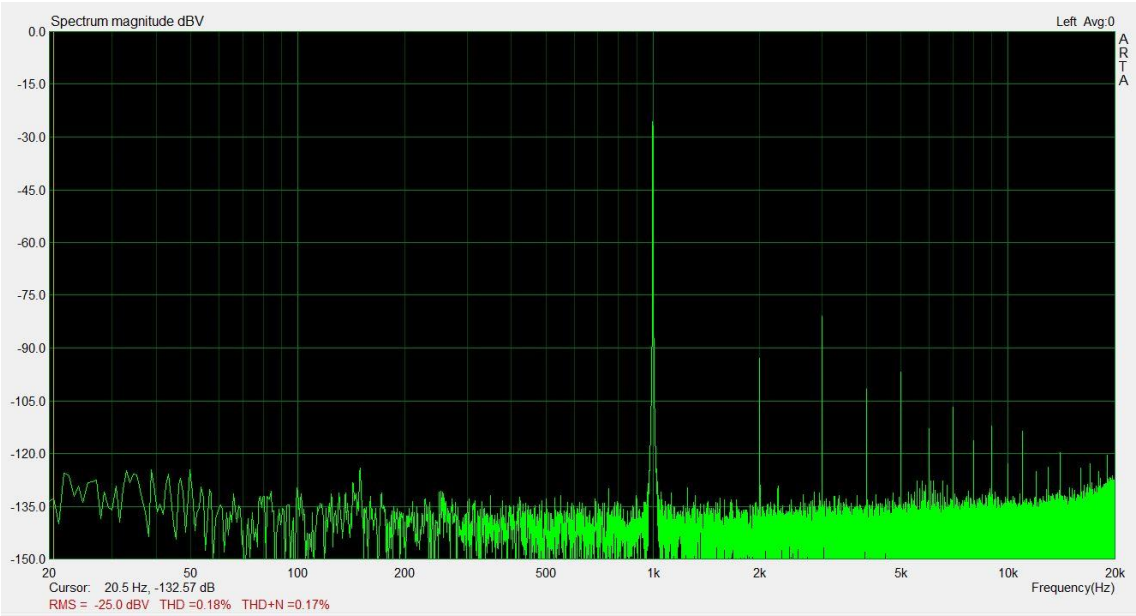
Right channel phase at high crossover point (2002Hz)

Test Results

Version 1 – Distortion and Noise



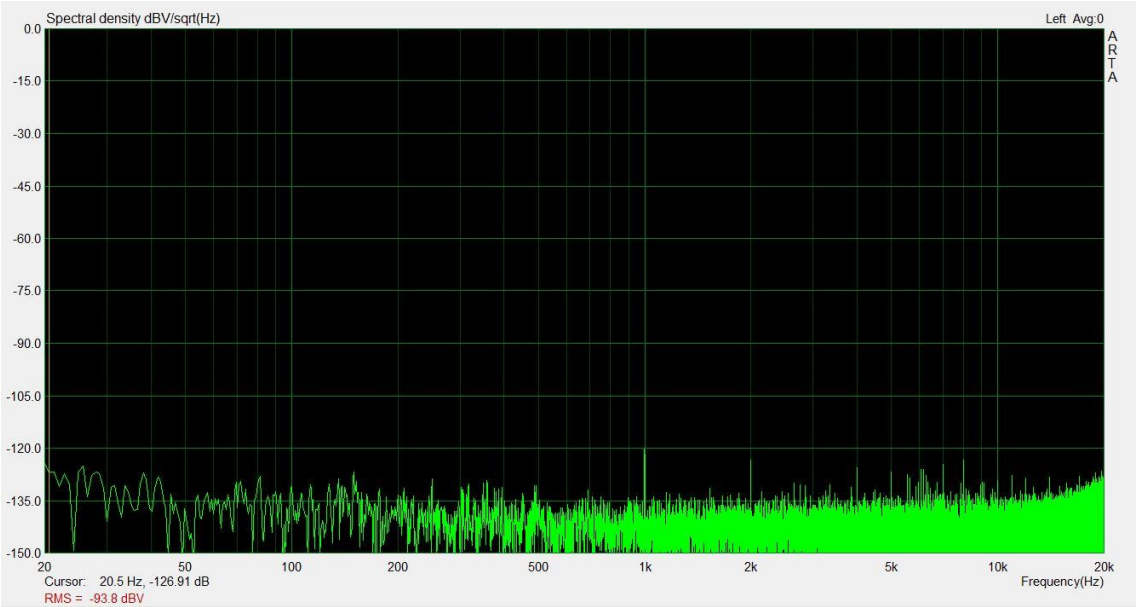
THD and THD+N left channel, 1kHz @ -25dBV



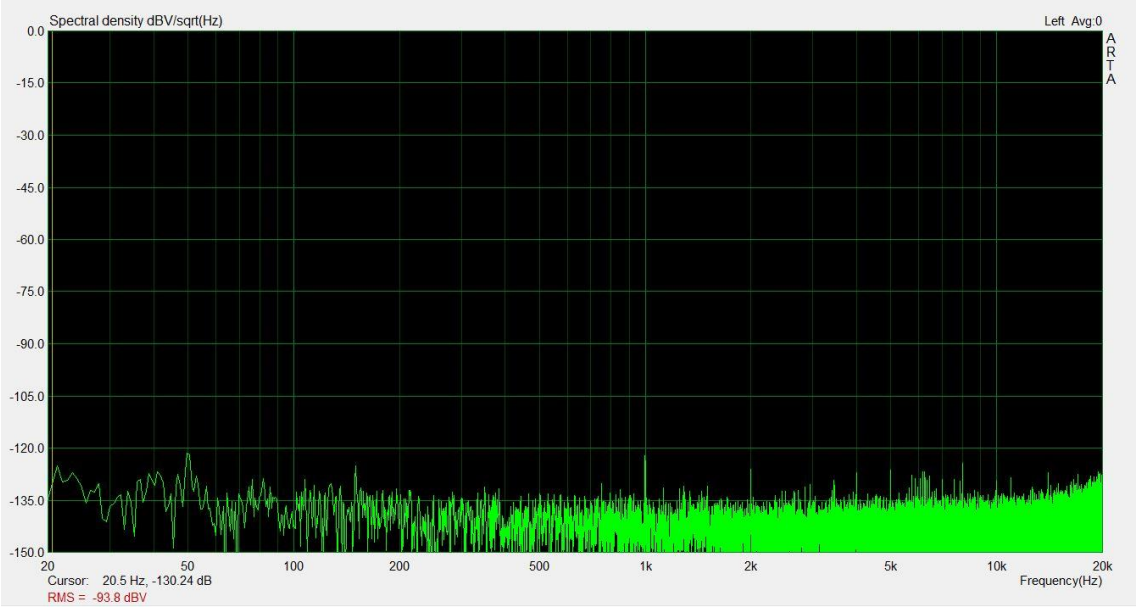
THD and THD+N right channel, 1kHz @ -25dBV

Test Results

Version 1 – Noise



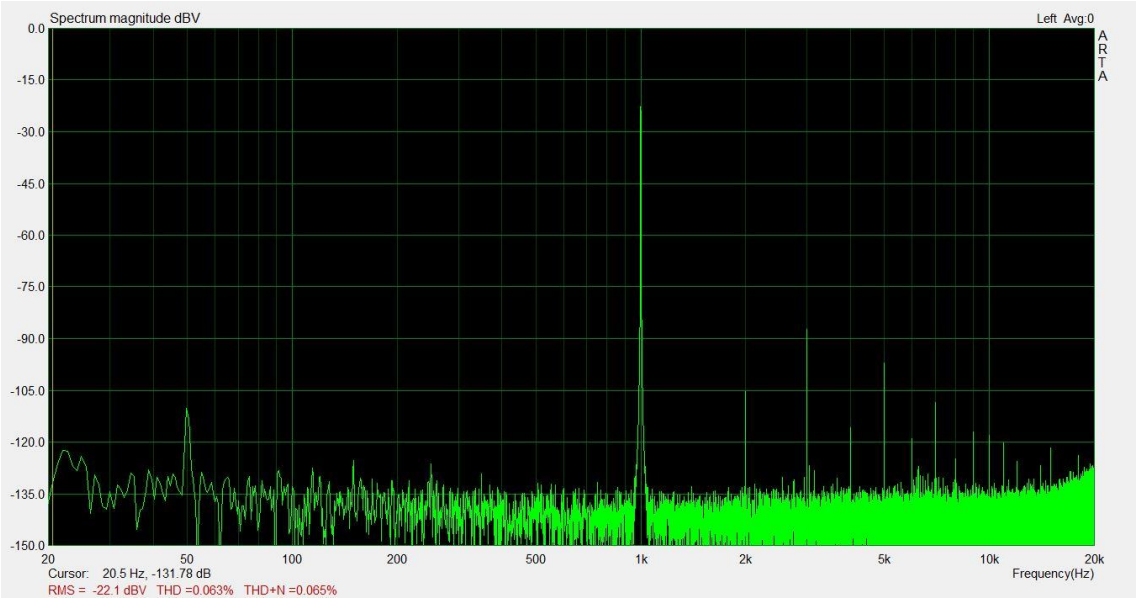
Noise floor version 1, left channel (limited by test equipment)



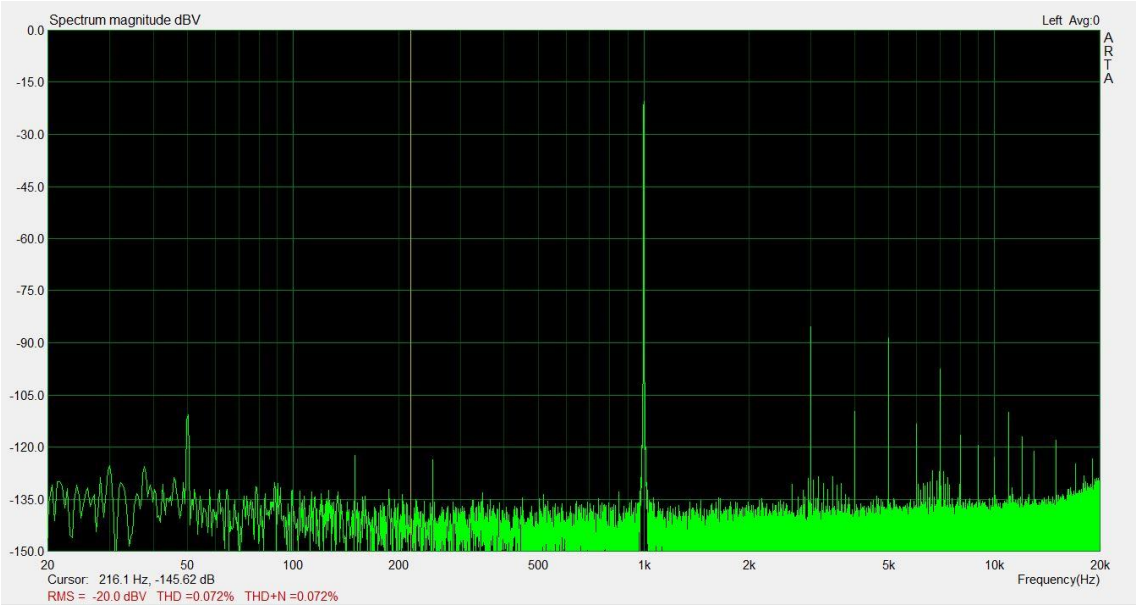
Noise floor version 1, right channel (limited by test equipment)

Test Results

Version 2 – Distortion and Noise



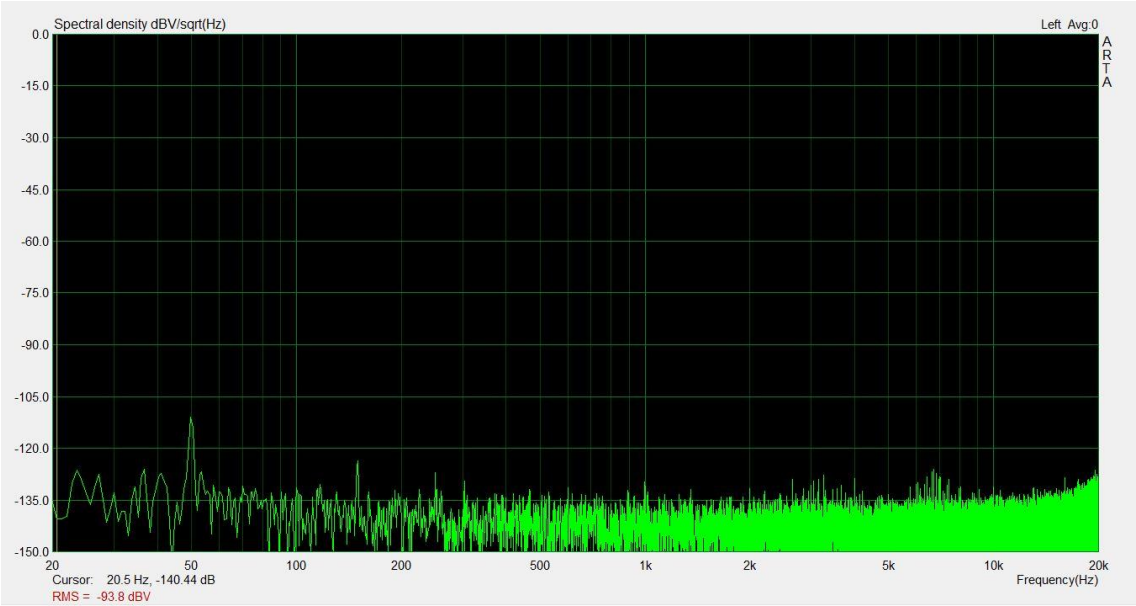
THD and THD+N left channel, 1kHz @ -20dBV



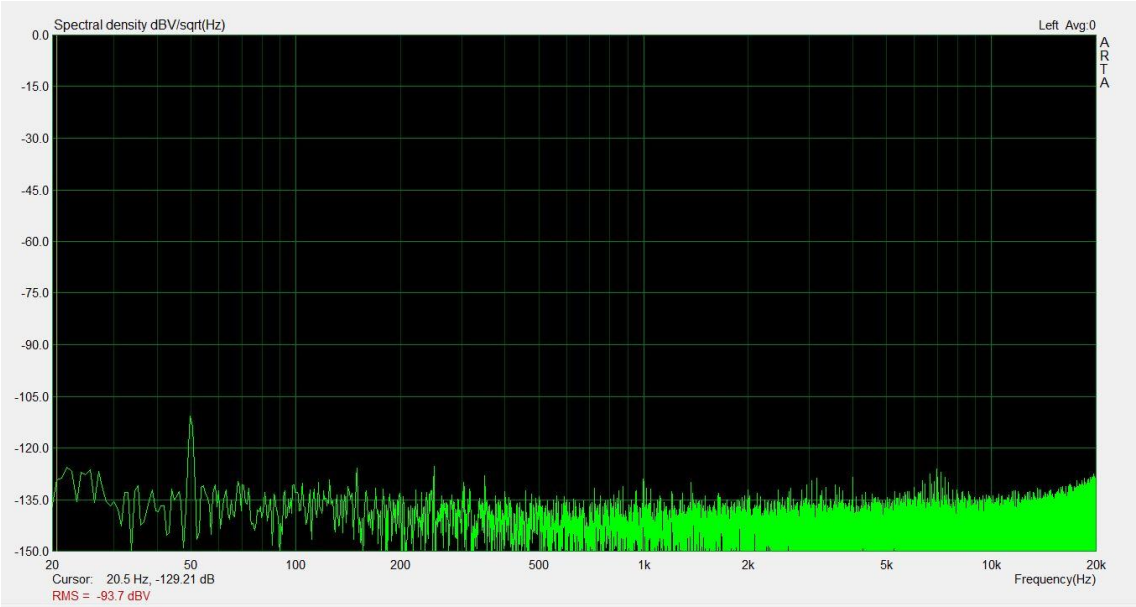
THD and THD+N right channel, 1kHz @ -20dBV

Test Results

Version 2 – Noise



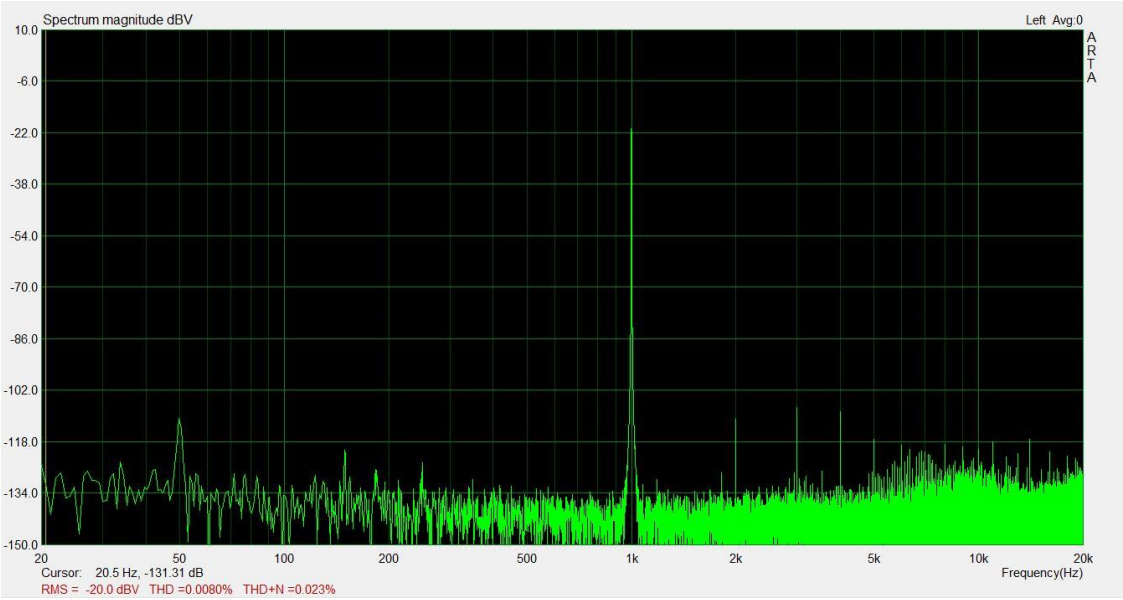
Noise floor version 2, left channel (limited by test equipment)



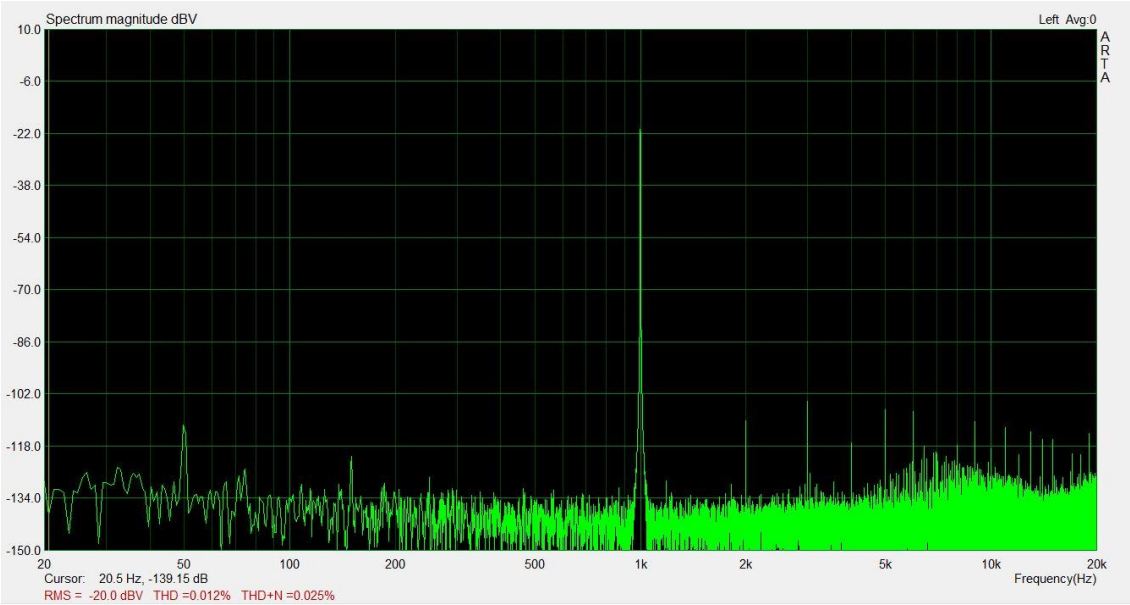
Noise floor version 2, right channel (limited by test equipment)

Test Results

Version 3 – Distortion and Noise



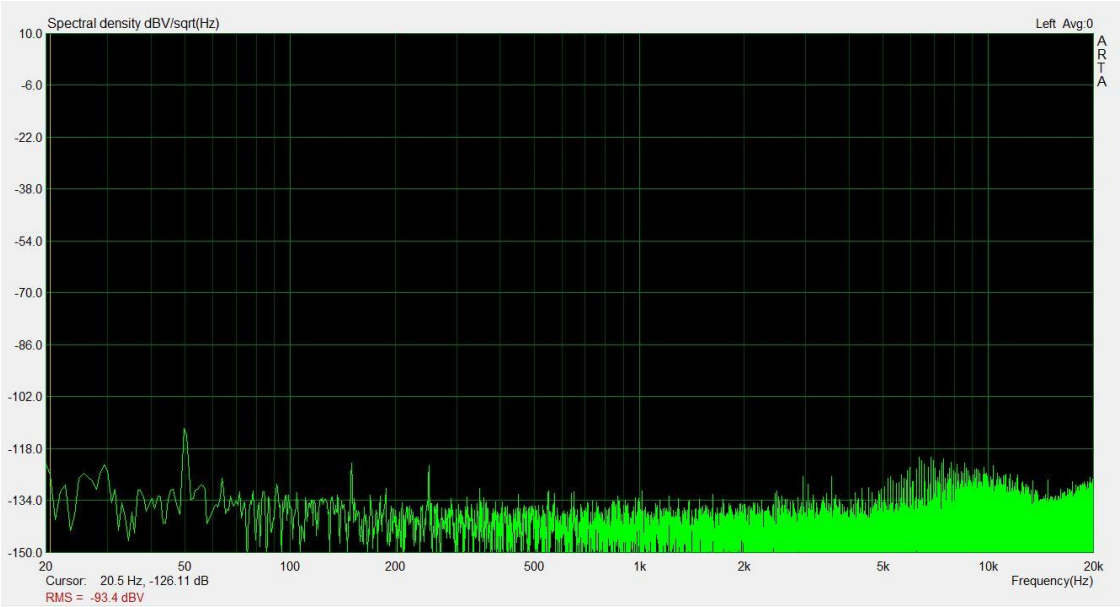
THD and THD+N left channel, 1kHz @ -20dBV



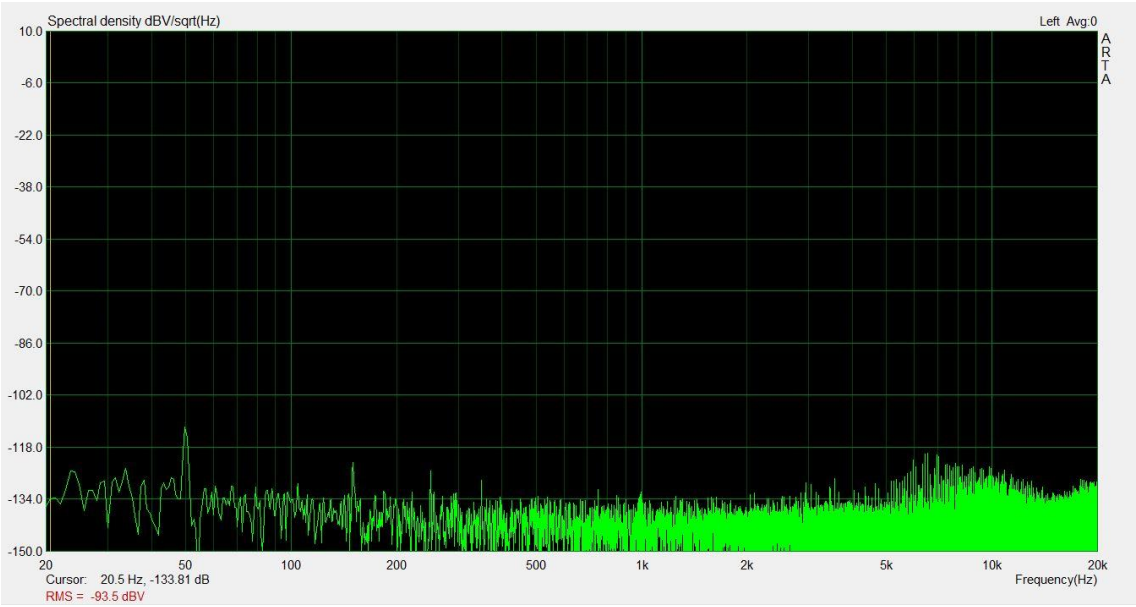
THD and THD+N right channel, 1kHz @ -20dBV

Test Results

Version 3 – Noise



Noise floor version 3, left channel (limited by test equipment)



Noise floor version 3, right channel (limited by test equipment)