

CLASS D AMPLIFIER MEASUREMENT FILTER

Design and test notes



23-Feb-18

Rev 3.0

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VERSION HISTORY

This document was started as a lab notebook to keep track of results during development. It therefore was more of chronological log. Once the project was done it was decided to go back and add additional information so that someone else could not only copy the design but see the decisions that led to this particular design. There are some design dead-ends and side experiments from the original notes in this document.

REV

- 1] 14-Feb-18. Initial
- 2] 15-Feb Rev 2 add Tr measurement
- 3] 16-Feb Rev 3 added measurement with 1K load, tweaked model.
- 22 Feb Added final results

ABBREVIATIONS

- AC. Alternating current.
- ADC. Analog to Digital Converter.
- ADI. Analog Devices (semiconductor manufacturer). <http://www.analog.com/en/index.html>
- AES. Audio Engineering Society. <http://www.aes.org/>
- AP. Audio Precision. A well known manufacturer for audio testing products. <https://www.ap.com/>
- ASD. Amplitude spectral density. Units of amplitude per root-Hz.
- BNC. Bayonet Neill–Concelman connector. A standard connector system in electronics.
- BTL. Bridge Tied Load. The most common output driver configuration in power amplifiers as it provides 4 times the power from the same supply voltage as a single ended or half bridge design.
- BW. Bandwidth. Generally this would be taken to mean the -3dB point, though when calculating RMS signal level from noise spectral density it's $f_2 - f_1$. usually $f_2 \gg f_1$ and BW is taken to mean f_2 .
- DAC. Digital to Analog Converter.
- dB. Decibel. Throughout this paper voltage, not power, levels are used so $\text{dB} = 20 \log (X/Y)$.
- DC. Direct current.
- DUT. Device Under Test.
- EMI. Electro Magnetic Interference.
- EQ. Equalization.
- f. Frequency.
- FFT Fast Fourier Transform. A mathematical function for converting discrete time domain samples to the frequency domain.
- Hz. Hertz.
- I/O. Input/Output.
- kHz. kilohertz (1000 Hertz)
- SPICE. Simulation Program with Integrated Circuit Emphasis. A standard software for circuit simulation. The specific version used here called TINA-TI.

- THD. Total Harmonic Distortion. Expressed as a percent or a dB value. As a dB value it would be negative, but many times the minus sign is dropped as it's understood that the value can't be bigger than 0 dB which represents 100% distortion.
- THD+N. Total Harmonic Distortion plus Noise.
- XLR. Originally called Cannon XL connectors. A 3 pin style is used for balanced audio.

OVERVIEW

Class D audio amplifiers produce switching noise in their outputs in the 200 to 400 KHz range (and harmonics of those). While harmless to a speaker, audio test equipment can be confused by the presence of these signals. A passive filter is used to reduce this noise to a level that won't affect measurements with typical test gear.

While AP¹ and others makes an accessory specifically for this purpose it is somewhat expensive for having sit around when just an occasional measurement is needed.

The general characteristics of the filter are taken to be:

- Low insertion loss
- Flat (± 0.1 dB) to 20 kHz
- F_c in the 50 to 80 kHz range
- Attenuate by > 50 dB at 250 kHz
- High enough impedance to not load the amplifier (we'll say $> 10x$ the highest load impedance a power amp might encounter, 32 ohms meaning the filter input impedance > 320 ohms).

The AP filter has very low insertion loss (fraction of a dB). This makes the filter design more complicated; the approach here is to have an insertion loss of 6 dB when driven by the typical < 1 ohm impedance of an amplifier. However this also means the AP filter only works in to a high impedance load (200 k ohm in the case of the AP515 input). The proposed filter's output impedance of 500 ohms allows it to work in to sound cards and similar devices with input impedance in the 10K range without worry of changing the characteristics.

Flatness isn't that critical as results can always be scaled with a calibration curve. Generally it's preferred to not have to add that step to the measurement process. However the final design did end up with some unexpected roll-off at 20 Hz.

The filter cutoff F_c is driven by the trend in systems to have a 96 kHz sampling rate so the frequency response is extends up to the 40 kHz or so area. The reasons for that are outside of the scope of this report.

The attenuation amount is based on the guidance from AP about using their test equipment with class D power amplifiers.

While it's unlikely a power amp will see a 32 ohm load, it's not an unreasonable worst case assumption.

¹ See <https://www.ap.com/analyzers-accessories/accessories/aux-family-switching-amplifier-measurement-filters/>

FILTER DESIGN TOOL

The analog filter design tool Elsie from Tonne Software was used for initial design (the basic free version):

<http://www.tonnesoftware.com/elsie.html>

Analysis of the design and of modifications made during testing to reflect actual component values was done in TINA-TI:

<http://www.ti.com/tool/TINA-TI>

BALANCED FILTER

High power class D amplifiers are almost always run in the BTL configuration, meaning there is no output reference to ground. Therefore the filters are designed for operation in a balanced configuration. Most of the initial testing is done on a single side, so the Fc point appears shifted from the final specification.

OTHER

A lot of commentary will make reference to this TI app note about Class D amp filters:

<http://www.ti.com/lit/an/sloa107/sloa107.pdf>

However it is felt that this simple RC filter doesn't provide enough attenuation to make accurate measurements on very low THD+N amplifiers; that paper was written in 2004 and technology has changed since then. Increasing the filter order isn't practical with RC networks in this application.

Other devices for measurement filtering can be found but at least some of them use ferrous core inductors, which presents a source of distortion unless the test conditions are tightly controlled.

THE COILS

The theme of this experiment was to use low cost prewound coils used for speaker crossover applications as a way to avoid the time/complexity to design custom coils. Parts Express offers a wide range of values and seemed like the cheap way to base the design off of. Particularly [they have some 20 AWG based ones](#) –thinner wire could be used but it's not available from them.



Figure 1 Example of coil used

It did turn out on another project that getting custom coils made isn't a huge extra expense, though for only 4 (the number needed to build one unit) it's not known what the price is.

An issue that was discovered is that the flat coils with a lot of layers will show (spoiler alert) issues due to the proximity effect. Thinner wire and a wider profile would eliminate that problem (a comparison was made with a custom coil with 3 layers) in the bandwidths present here.

The coils also proved to be very adept at picking up 60 Hz. Even with all surrounding equipment off (other than laptop and AP) and being several feet away a simple loopback test through a coil picked up 60 Hz and harmonics. This happened even when placed inside the metal enclosure, so it is assumed the radiated field being detected is magnetic in nature. Changing coil orientations in free air could change the measured peaks by a factor of 3. Two coils in series were tried to see if having one reversed might cancel some of the interference but no arrangement was found that helped.

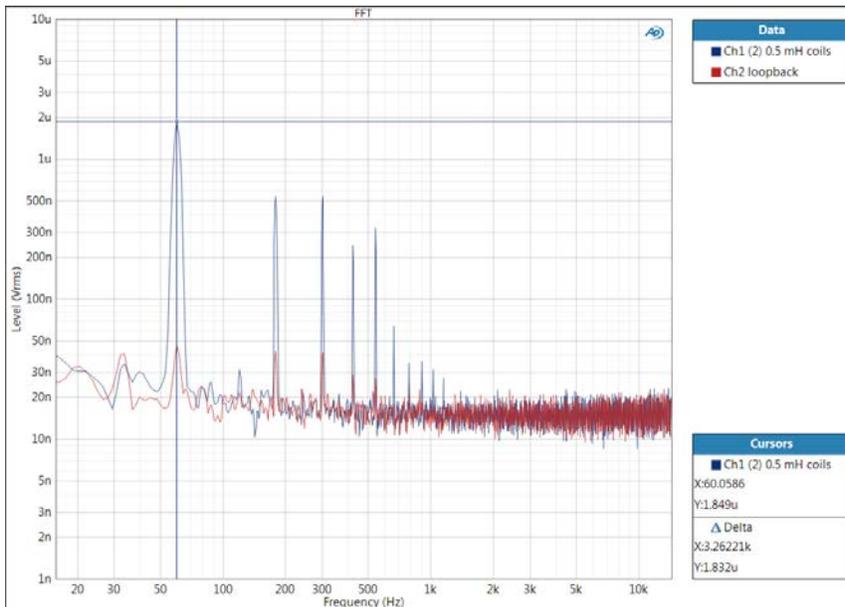


Figure 2 Two 0.5 mH coils in series in the "-" lead (AES17 filter, 2.2 uV RMS total noise vs. 1.1 uV on the loopback connection)

What did help was placing one coil in each leg of the balanced connection to/from the AP, the 60 Hz peak dropping from 1.8 uV to 400 nV. Presumably in this configuration the 60 Hz is appearing as a common mode voltage which the AP will not respond to. However this best result is only obtained when the two coils were right next to each other. Separating them by a few inches the 60 Hz level was 1 uV; total noise was 1.9 uV RMS.

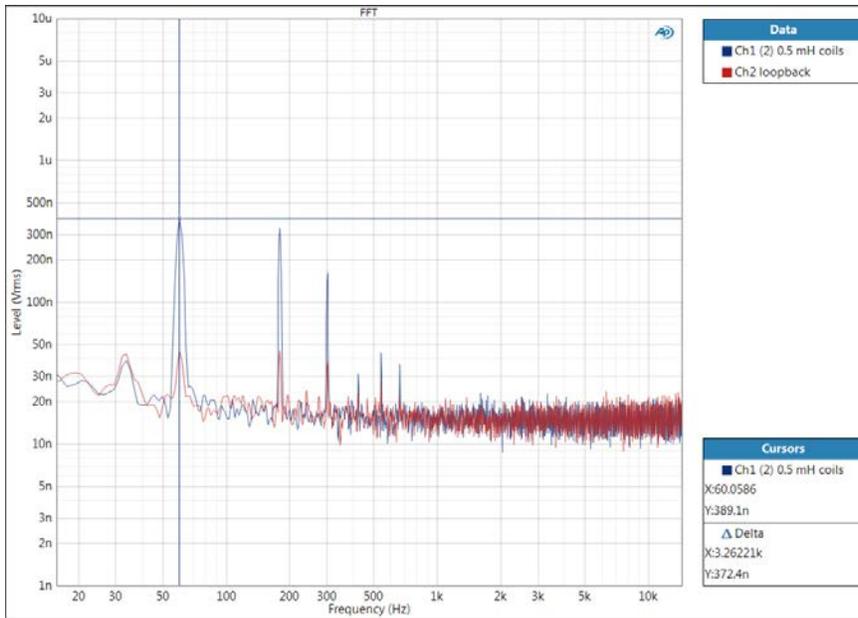


Figure 3 One 0.5 mH in each leg and coils in close proximity. Total noise about 1.2 uV

Returning to the final filter box, Figure 4 shows the 32 K pt FFT (AP-Equi filter, 8 averages) obtained. Despite the large 60 Hz spikes the total noise is not a large number. However the experiment with the coils showed the importance of maintaining symmetry in the design to avoid common to single ended noise conversion. Upon inspection of the unit it was realized one of the four coils was installed upside down relative to the other 3.

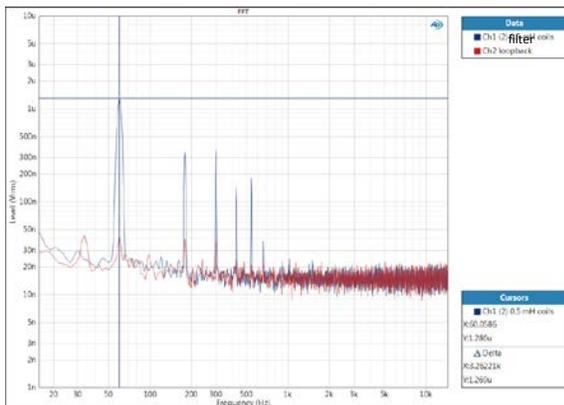


Figure 4 Almost final filter box. 1.9 uV total noise (vs. 1.1 uV on the loopback) AES-17 filter*

Correcting that coil by flipping it over to match the other three the results of Figure 5 were obtained. Total noise measured (eyeballed) as about 20 nV higher than the loopback input. The 60 Hz peak is 150 nV, the 180 Hz one a little over 200 nV.

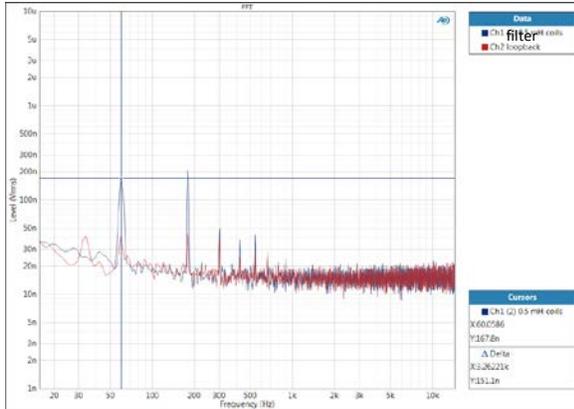


Figure 5 Final filter with coil fixed (see text)

* Note there was a labeling error in the above two plots, Ch 1 is the filter box output.

EXISTING FILTER

The AP filter has been reverse engineered and the schematic made available, though there is some debate if the values reported are correct. <http://cranialcrumbs.blogspot.com/2015/05/audio-precisions-class-d-filter.html>

A TINA-TI simulation was created (of one side) and produces results that are close to available published data, so lacking better information this is taken as the equivalent filter design.

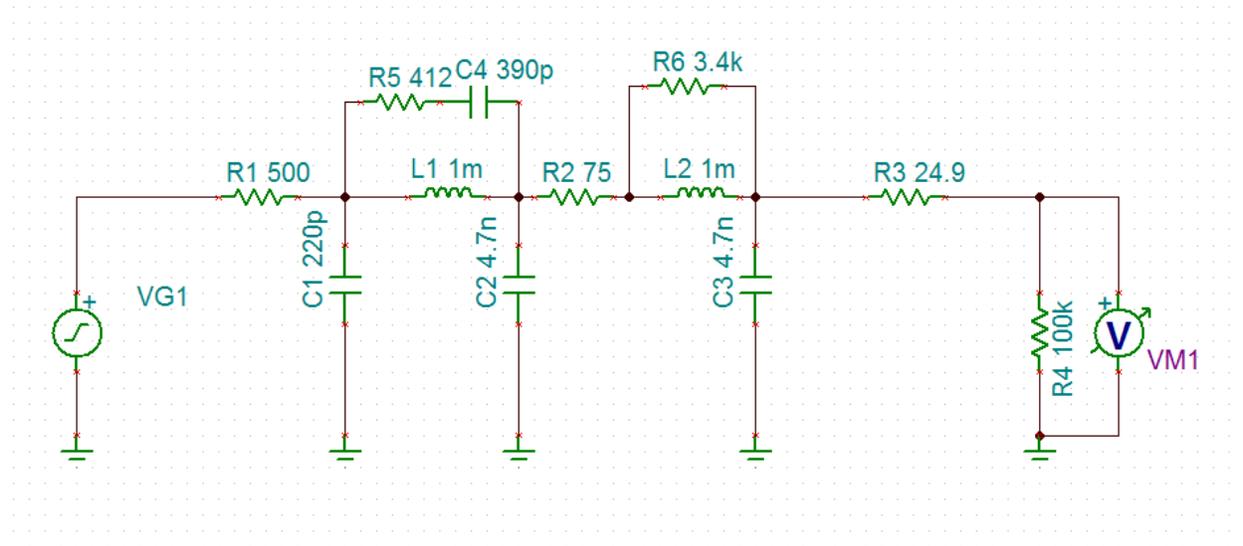


Figure 6 Filter design from website

Typical filter design assumes the input impedance (R1 in the above) is the same as the load impedance as these types of filters are typical of RF applications, and maximum power transfer is desired. The AP design assumes high load impedance, and by some simple experiments the parallel components look to be there to keep the overall response curve monotonically decreasing with the high impedance load.

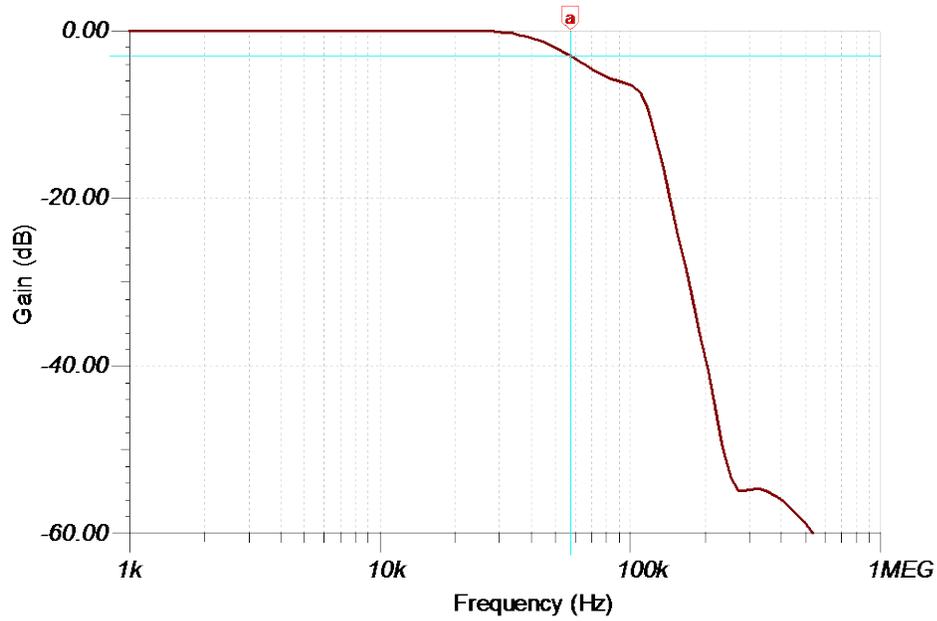


Figure 7 Reference filter response (amplitude vs. frequency)

The 3 dB point (of the one side) is about 58 kHz. Note that the filter has almost no insertion loss. It's down about 55 dB at 250 kHz.

Phase response is well behaved in the audio band but must be considered if looking at the phase response of the amplifier:

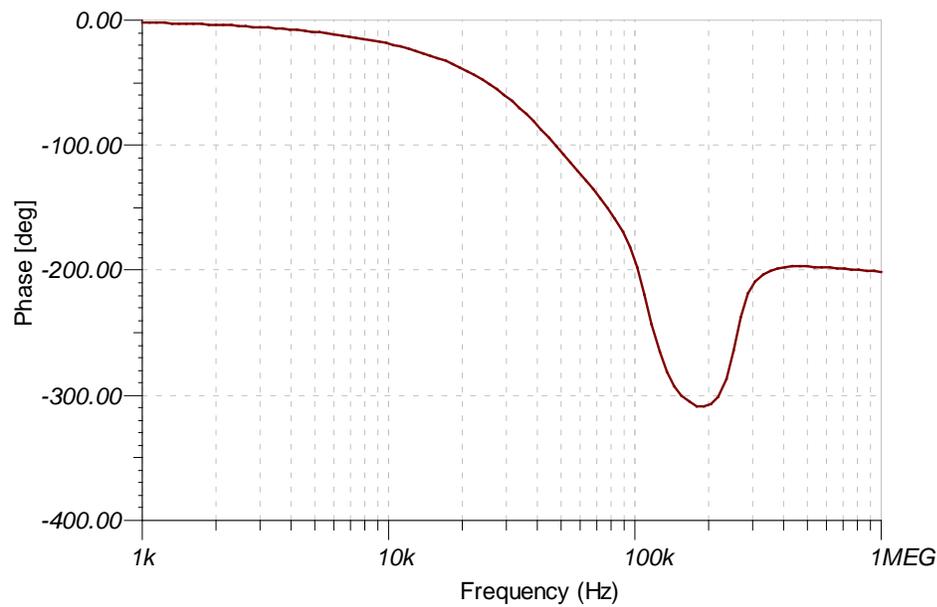


Figure 8 Reference filter response (phase vs. frequency)

PROPOSED FILTER

One simplification made was to have the source and load impedance be the same. While this results in a 6 dB signal loss this is not a major concern with the power amplifiers that will be tested. As the measurement tools all have auto-ranging inputs the measurement of amplifiers with low noise floors (< -110 dB) is still easily achieved.

The Elsie tool was used in an iterative process based on standard available components. The design was recaptured in to TINA-TI for further analysis.

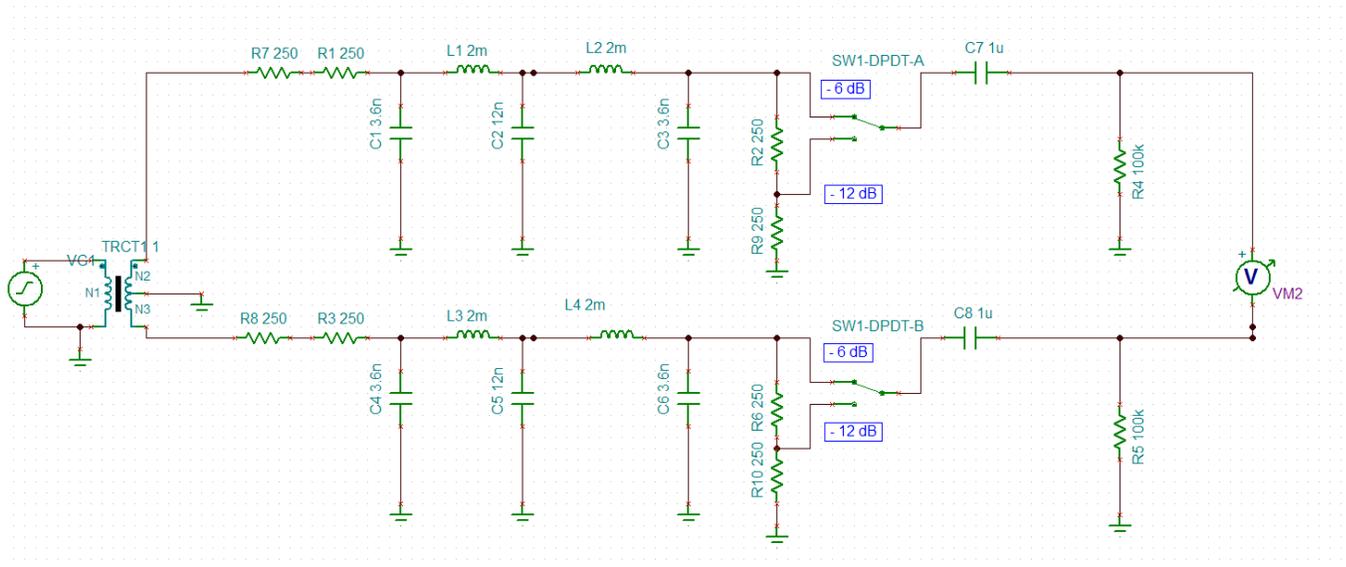


Figure 9 proposed filter simulation model

The initial version of the model in Figure 9 used 466 ohm resistors. However component availability and the desire for a 6 dB cut switch at the output change the resistors to 500 ohms, formed from two 250 ohm resistors in series. 3W non-inductive wirewound parts were used for the low distortion and power handling capability. This change made no appreciable difference to the filter function.

In some of the initial measurements the original design value of 466 ohms was used and there are references to that value in the text.

The input transformer is an ideal one and is needed² to correctly generate ground referenced balanced sources in TINA-TI so that AC Analysis can be performed. A 1:2 ratio could be used to avoid the scaling issues.

The filter has a -3 dB Fc of 59 kHz and is 68 dB down at 250 kHz.

² There may be some better way...

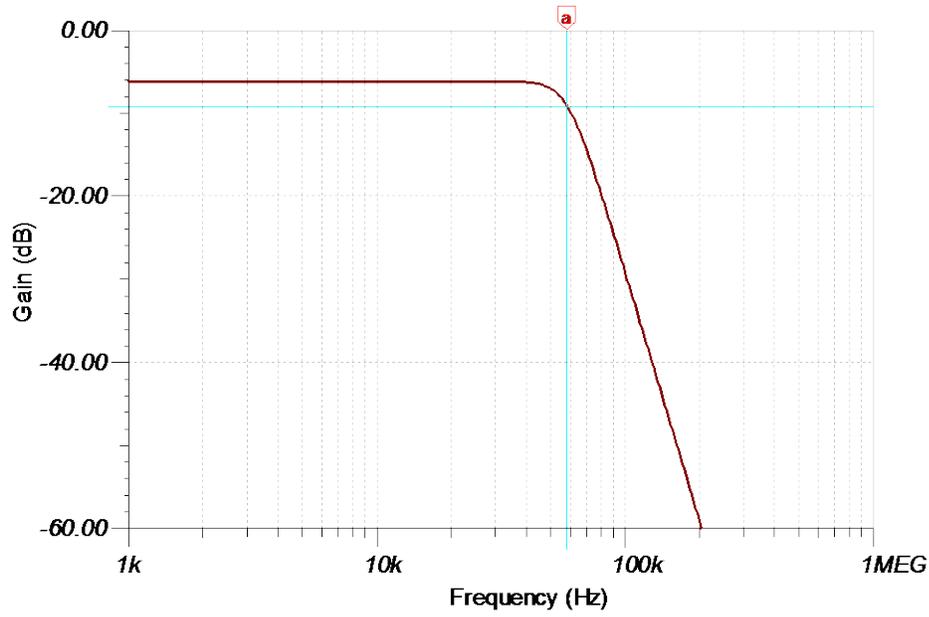


Figure 10 Proposed filter gain

The phase response does change faster in the audio band compared to the AP design, but either way if measuring the phase response of an amplifier a correction would need to be applied to remove the filter's contribution.

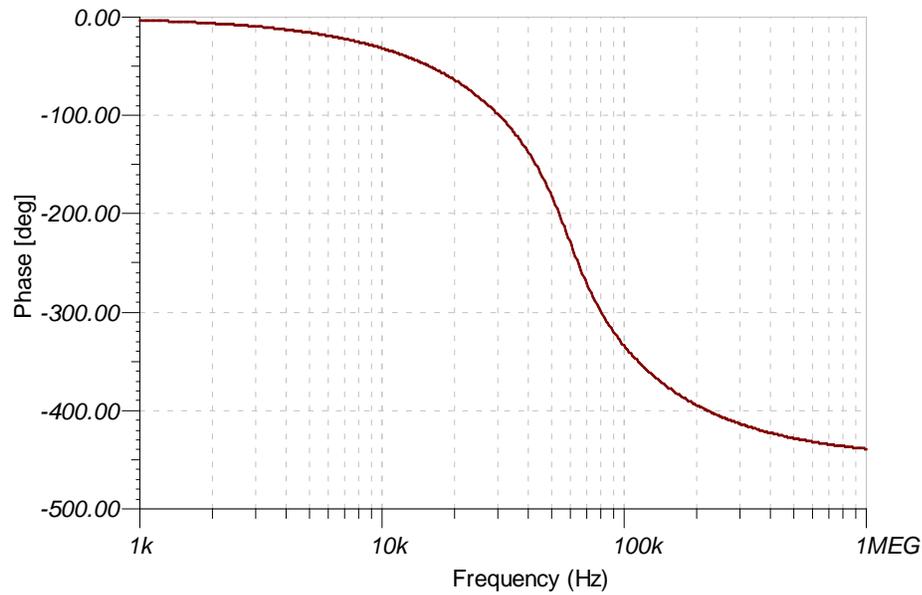


Figure 11 Proposed filter phase response

INITIAL EXPERIMENTATION

The 2 mH coils were received prior to the other parts so a filter was prototyped with parts on hand to see how well the measured performance matched the design. Differences were noted so the design was deconstructed and measurements made to determine where/why simulation and reality disagreed – or if it was pilot error. Ultimately the reasons for the differences are still not clear but are assumed to be some overlooked parasitic effect.

The AP is limited to signal generation up to 80 kHz; being able to analyze out further might have uncovered more information but would have required going to manual measurements with the test gear on hand.

Most plots are presented for f of 1 kHz to 1 MHz, and -60 dB to 0 dB amplitude. This includes the AP plots even though they stop at 80 kHz as the consistent plotting scale allows for easier visual comparison.

These test circuits were built on a breadboard and test lead pigtailed to BNC used for connection to the AP; there was some noise pickup – mostly 60 Hz and harmonics. RMS noise level measured about 7 μ V, not enough to worry about for most of the measurements (though will set a floor for THD+N).

An LCR meter was not available at the time that this was written to verify values; this would obviously help remove some uncertainty in explaining the measured values and the deviation from the simulated ones.

To make comparison easier all attenuation is reported relative to the attenuation at 1 kHz, not relative to the input.

COIL TEST

To move F_c down in to the range the AP can accommodate two 2mH coils in series were used. A generic TINA-TI simulation was set up to allow adding additional parts by changing values from 'almost' zero to the values in use. The AP generator has an impedance of 50 ohms which results in the 0 Hz value being -6.5 dB instead of -6 dB when the R_s value at the input is used (R_1 in the TINA-TI schematic).

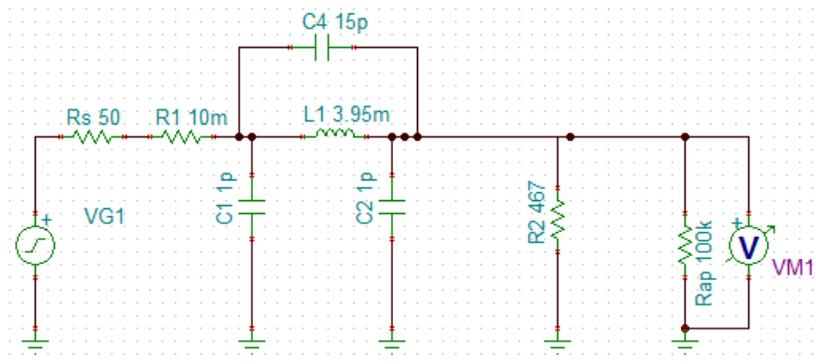


Figure 12 Generic simulation circuit for filter component testing (shown with the values used for two 2 mH coils in series)

C4 was added to represent the unknown parasitic capacitance of the coils. From some crude measurements (resonance around 650 kHz) the C_p of the 2 mH coil is around 30 pF. With two in series C_p is half that, or 15 pF. The effect on response is out in the 100s of kHz.

The above simulation has a – 3dB point of 20.8 kHz and -12 dB down at 80 kHz

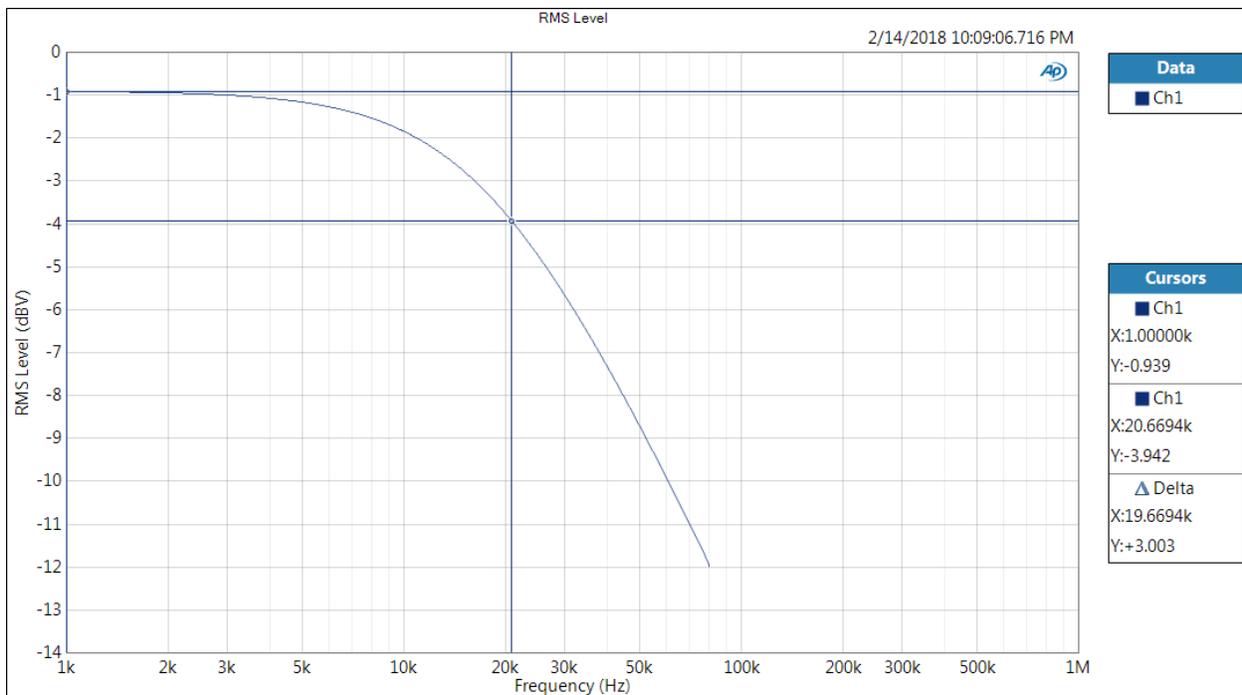
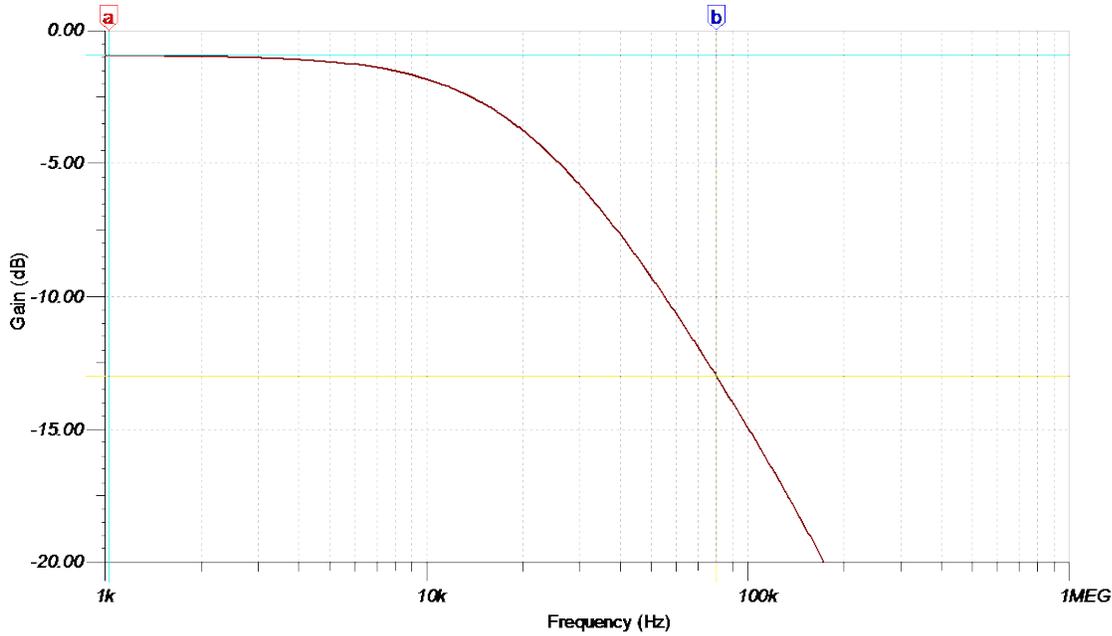


Figure 13 4 mH coil (two 2mH in series) simulated (top) and actual (bottom) responses (not on the common scale used elsewhere)

The measured -3 dB point was 20.8 kHz, the simulation values shown earlier being modified to match this measurement to give an inductor value of 3.95 mH for two in series.

At 80 kHz the response was -11.0 dB where was the simulated value was -12 dB. An error in the resistor values might explain that except the 1 kHz values for simulation and measurement agree (< .01 dB difference). The parasitic capacitance only has a small impact at 80 kHz and can't explain the 1 dB difference. It was noticed there

was about +/-0.1 dB variation in the measurement depending on where the inductors were connected/placed on the breadboard, but there was no obvious pattern (which included placing them up to several coil diameters away from each other in case there was some mutual coupling). This variation is probably within the bounds of error from the connector fingers of the breadboard.

THD was close to the AP limits, measuring just 0.0002% at 1 kHz and rising to 0.0005% at 20 kHz, though there were some spikes and other roughness in the plot which is assumed to be a function of the previously reported noise and low level EMI.

A quick look at possible error sources: The estimated (i.e. for a straight wire with no magnetic induced crowding) AC resistance of the coil is .01 ohm³ at 20 kHz based on an estimated 1250" of 20 AWG wire. The actual resistor values were measure at 466 and 467 ohms⁴, but as the output is a ratio of those two values at the lower frequencies this won't explain a difference between simulation and measurement in the final circuit with the divider. Simulations were updated to use 466/7 ohms to eliminate this minor difference between the standard 470 ohms value and the measured values. There is a change in effective inductance as the frequency increases, but at 80 kHz the change is calculated to be small (< 1%).

The AP's loopback with just a 466 ohm resistor load is flat within +/- 0.02 dB (and actually within +/- 0.005 dB except right around 79 kHz).

To get the same attenuation at 80 kHz an L value of 3.4 mH is needed in the simulation; this puts the 3 dB at 24 kHz. At the moment a reasonable explanation for the mismatch between the simulation and the actual circuit isn't known.

A single inductor is 1.975 mH, very close to the specified 2 mH value (with a DCR of 1.06 ohms). Using just a single inductor the calculated -3dB point is 41.4 kHz and at 80 kHz the attenuation is -6.92 dB.

The measured 3 dB point is 43.3 kHz, which is an unexpected deviation. An inductor value of 1.9 mH is needed in the simulation to produce the correct -3 dB point.

At 20 kHz the difference between simulation and measurements is 0.07 dB. The AP measured value (with either of the inductors used previously, so the values are equal) is -5.87 dB at 80 kHz, or a response that's 1 dB higher than expected.

In summary, at this point in testing there was an overlooked characteristic of an actual air coil inductor that the simulation doesn't include. Or there's some error in the way measurements are being performed, though that turned out to not be the case. For an explanation see the section: Revised coil only test (added in Rev 3) on page 18.

COIL TEST MEASURING TR

The 10% to 90% rise time is related to the time constant as:

$$t_r = \tau \cdot \ln 9 = \frac{L}{R} \cdot \ln 9 \cong \frac{L}{R} \cdot 2.197$$

³ The DC resistance of the coil is given as 1.06 ohms, a SPICE value of 2 ohms was used to provide an approximation at 20 kHz.

⁴ As noted previously the final design uses 500 ohm resistors.

See https://en.wikipedia.org/wiki/Rise_time#One-stage_low-pass_LR_network for derivation.

Using a Tek FG503 (50 ohm source) as square wave generator for rise time measures it shows 44 usec with a 50 ohm load resistor (so 100 ohms total), so plugging in that value yields 2.0 mH. Probably +/-5% in the measurement uncertainty.

ADDING BACK THE CAPS FROM THE CIRCUIT

The capacitor values are +/-5%.

Just one inductor L_s now used, for simulation we chose 1.9 mH as the value based on the F_c measured with a single inductor in the prior steps..

The input filter is now added back. The additional resistance shifts the operating point considerably, the -3 dB point (simulated) is 80.7 kHz and the response at 20 kHz is calculated as -0.2 dB. The AP measures -2.4 dB at 80 kHz.

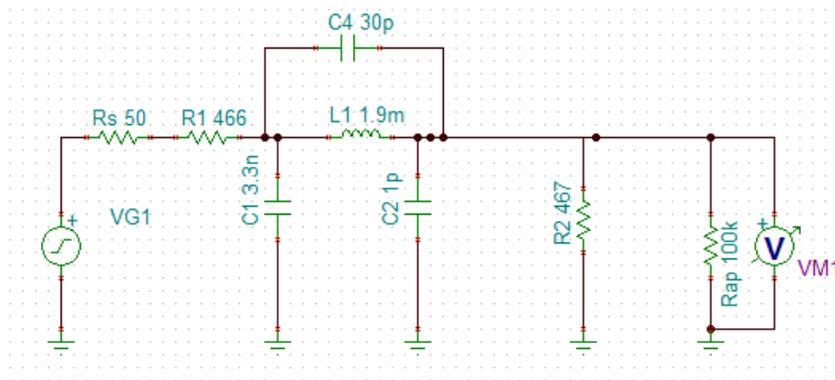


Figure 14 intermediate simulation circuit

Next the 12 nF cap is added (C2). The simulation values are a -3 dB point of -55.8 kHz, -8.6 dB at 80 kHz, and -0.26 dB at 20 kHz. Measured values were a -3dB point of 50 kHz, -8.1 dB at 80 kHz, and -0.49 dB at 20 kHz.

FINAL FILTER

As done in the previous section attenuation is reported relative to the value at 1 kHz, not relative to the input.

The absolute value of the attenuation at 1 kHz is -6.5 dB.

The rest of the components are added back to the simulation and actual circuit. The simulation calculates a (relative) -3 dB point of 59.6 kHz, -12 dB at 80 kHz, and -.02 dB at 20 kHz.

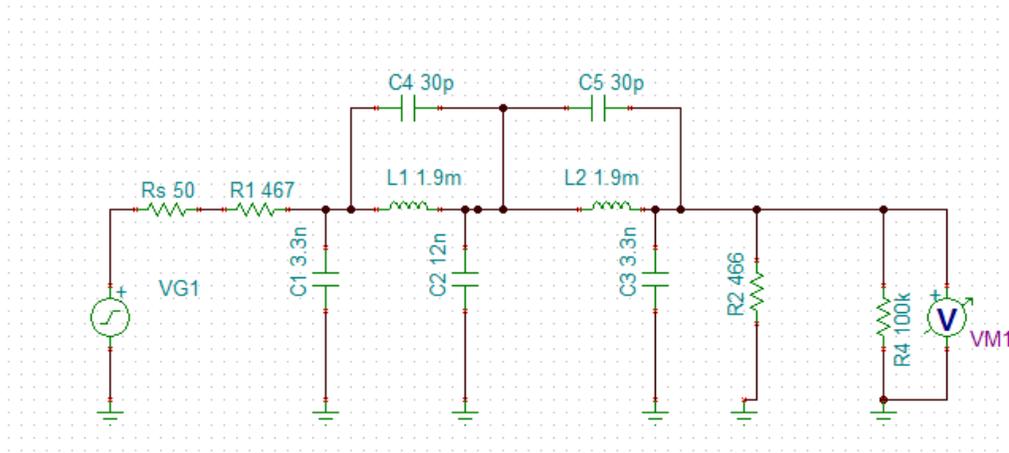


Figure 15 Simulation based on assumed actual L values

The measured (relative) values are -3 dB point of 52.1 kHz, -11 dB at 80 kHz, and -0.39 dB at 20 kHz.

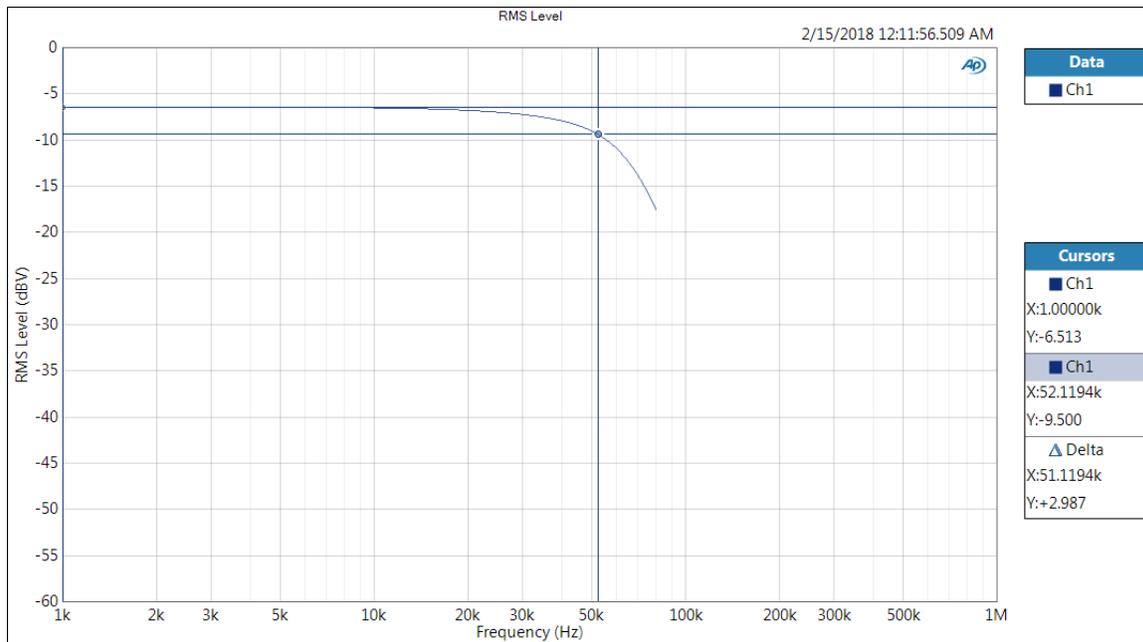
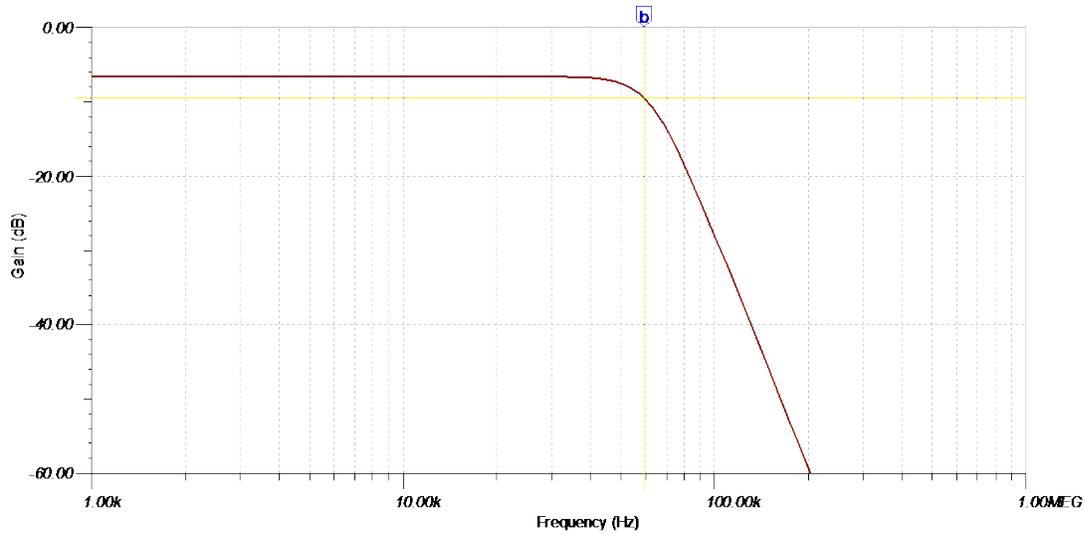


Figure 16 Simulated and actual filter plots with markers on -3dB point.

While the simulation and actual circuit don't fully agree they do appear to be close enough to trust that the simulation's result of > 60 dB attenuation at 250 kHz matches the real world. The resonance frequency out around 650 kHz simulates at -150 dB and the level never rises to more than -120 dB.

The error at 20 kHz and -3 dB point mismatch are troublesome from the perspective of understanding the actual filter's performance. It is easy to hand-wave and blame the inductors, but an actual explanation is needed. [Later this question is answered].

From a practical standpoint the filter will solve the problem.

REVISED COIL ONLY TEST (ADDED IN REV 3)

The breadboard was set up with two inductors in series. The TINA-TI model was enhanced with a couple more known (or estimated) values, though generally they would not be expected to affect the frequency range of interest (i.e. < 100 kHz for determining why the response at 20 kHz is off from the simulation).

The load resistor was changed to a 1K 1% non-inductive wirewound part.

The measured value of the –3dB point (47.6 kHz) was used to set the L value in the simulation to produce the same frequency. The simulation includes the inductor DCR of 1.06 ohms.

Note that in all of the measurements and simulations the AP's source resistance of 50 ohms reduces the response at DC to less than 0 dB.

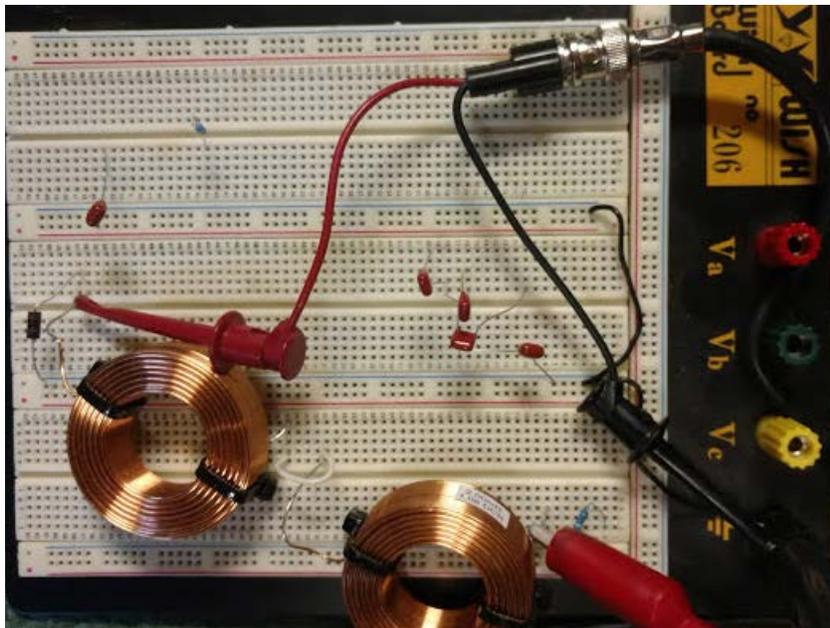


Figure 17 Breadboard with two inductors and 1K load

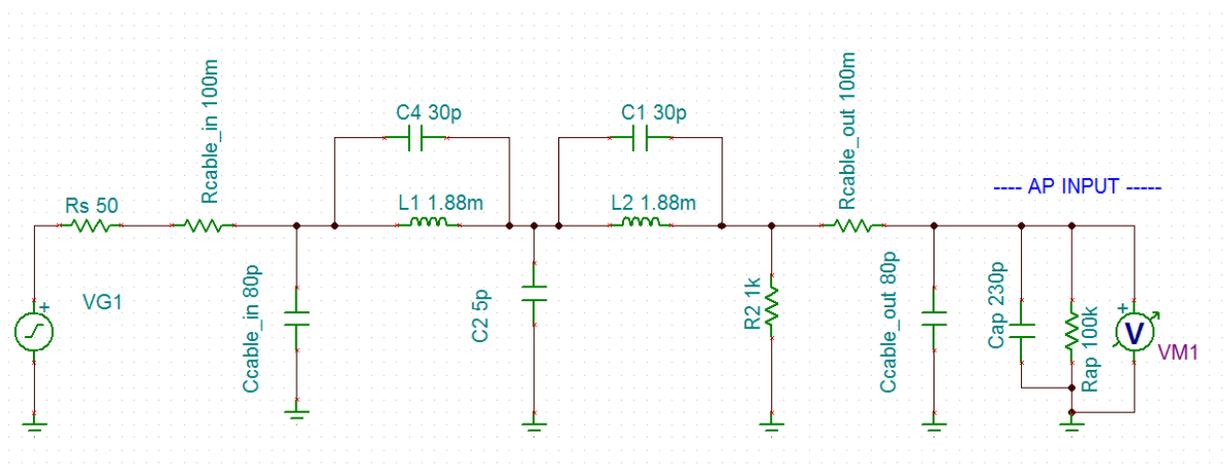


Figure 18 updated simulation model for LR filter with secondary effects added (cable resistance, capacitance, breadboard capacitance)

The addition of the small parasitic values to the model does not affect audio band performance, but does affect performance at the harmonics of the switching frequency, i.e. from 500 kHz and higher. Though beyond the AP's measurement range it was desirable to know how the final circuit would perform, though it presumably would have smaller parasitic values; as well as the actual filter capacitor values are orders of magnitude larger than the parasitics. Ultimately the only important one seems to be the coil's parasitic capacitance.

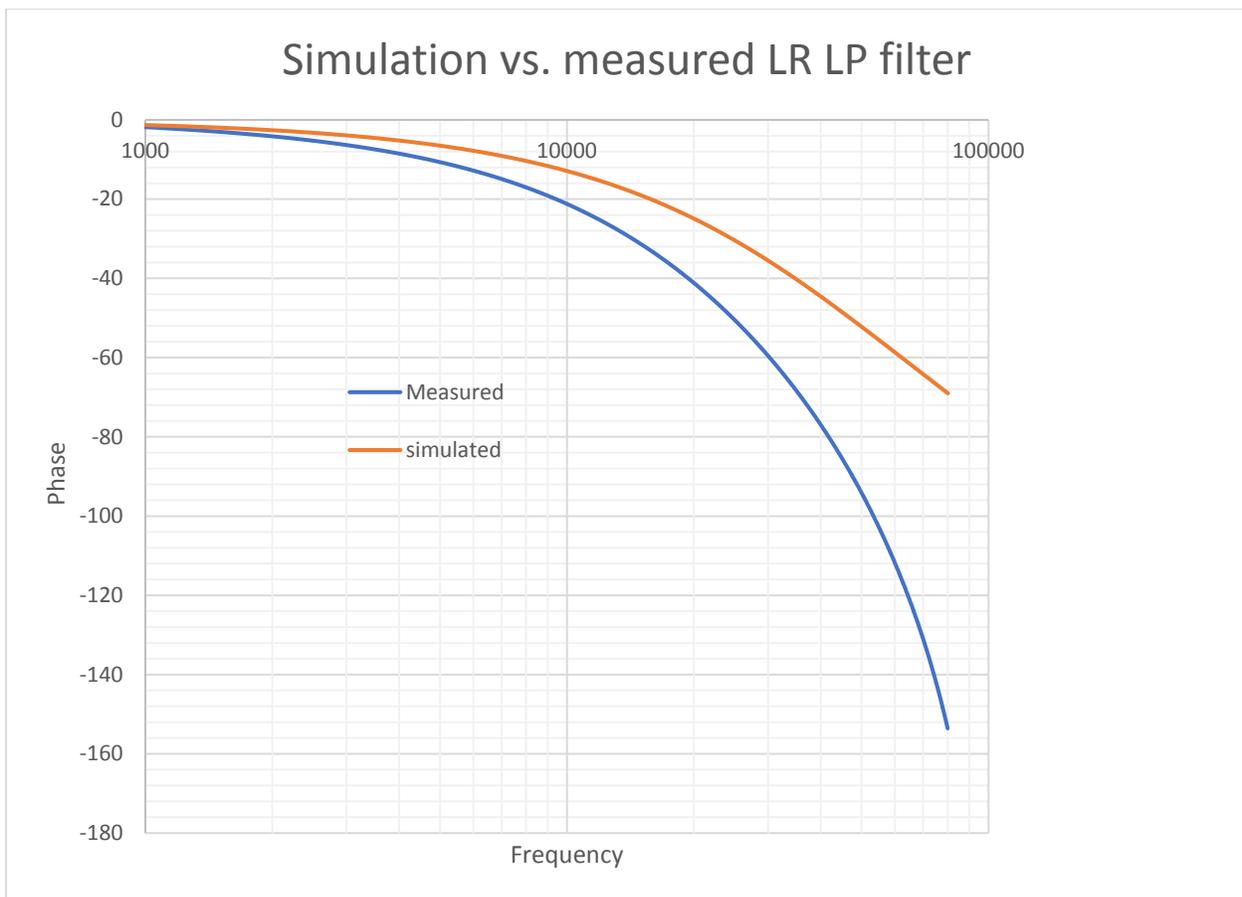


Figure 19 Measurement versus initial simulation of a simple first order LR LP filter

However this model still failed to produce a curve that accurately matched the measured values. Returning to research the parasitic effects of large multi-layer air core inductors, the issue of the proximity effect had not been considered as it was viewed as something of a concern after skin loss, and that wasn't an issue here.

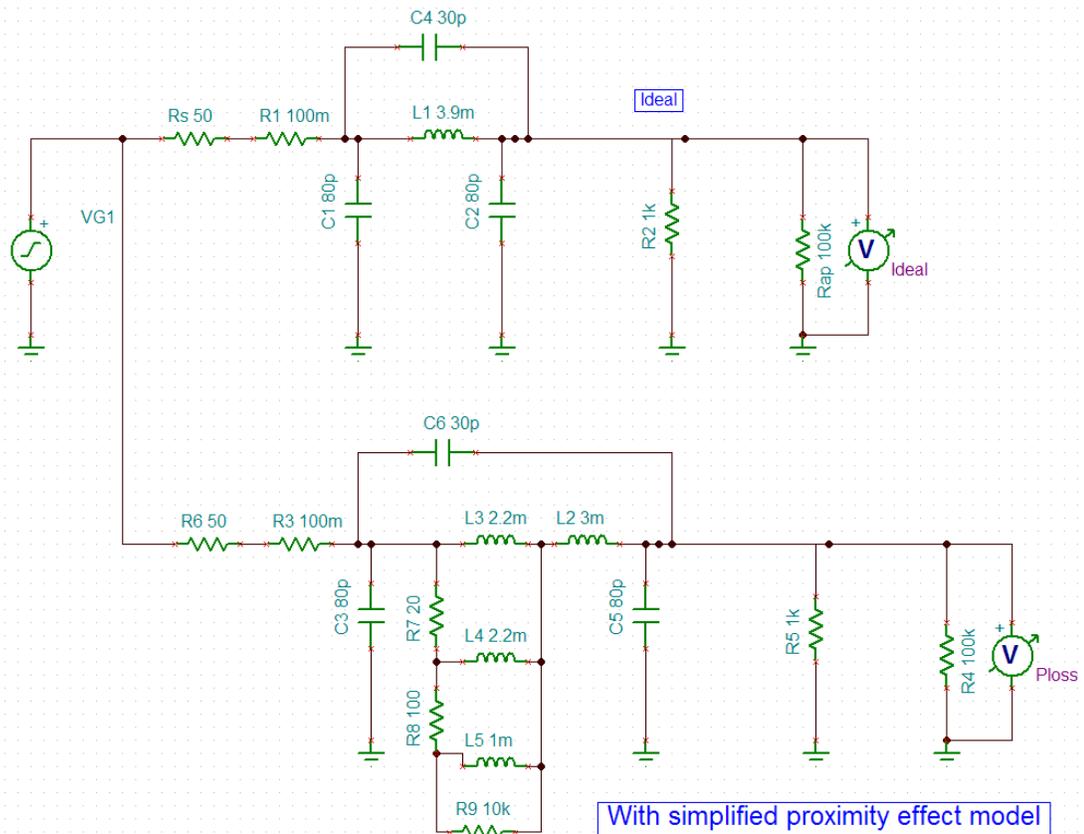
Further research showed that assumption was wrong, and for an inductor of this size the proximity actually comes in to play at the top end of the audio band. Of course in the application of a speaker crossover this wouldn't be an issue so there's no fault here associated with picking a cheap inductor.⁵

A search of inductor proximity effect mostly turns up research related to making planar inductors in ICs or PCBs. It seems most relevant work on this for large coils dates back to the 60s and therefore not as prone to success via a google search. However this website <http://ridleyengineering.com/design-center-ridley-engineering.html> offers a number of useful papers on inductors. This one in particular: <http://www.ridleyengineering.com/design-center-ridley-engineering/39-magnetics/163-097-linear-circuit-model-for-predicting-magnetics-winding-proximity-losses.html> (registration may be needed) provides a simple way to model this in SPICE.

The models on the Ridley Engineering website also include effects from the core losses/saturation; those are not a consideration with the air core inductors.

As the interest here was in a model for just a limited range of frequencies (i.e. audio band) the ideas from that paper were simplified to produce the following TINA-TI model to evaluate the difference accounting for proximity effect makes.

⁵ There are other styles of air core inductor that might not have the same problem – for example copper foil – but they cost 10x the price and the goal was a low cost filter.



<http://ridleyengineering.com/design-center-ridley-engineering/39-magnetics/163-097-linear-circuit-model-for-predicting-magnetics-winding-proximity-losses.html>

Figure 20 Model with proximity effect partially modeled (includes test setup parasitics as well)

The simulation results more closely matched the actual measurements being observed, with the proximity effect serving to increase the attenuation at 20 kHz where as the more ideal model does not, as shown in Figure 20. Note that this simulation result doesn't align with the exact values measured, but shows the relative shape of the two curves correctly.

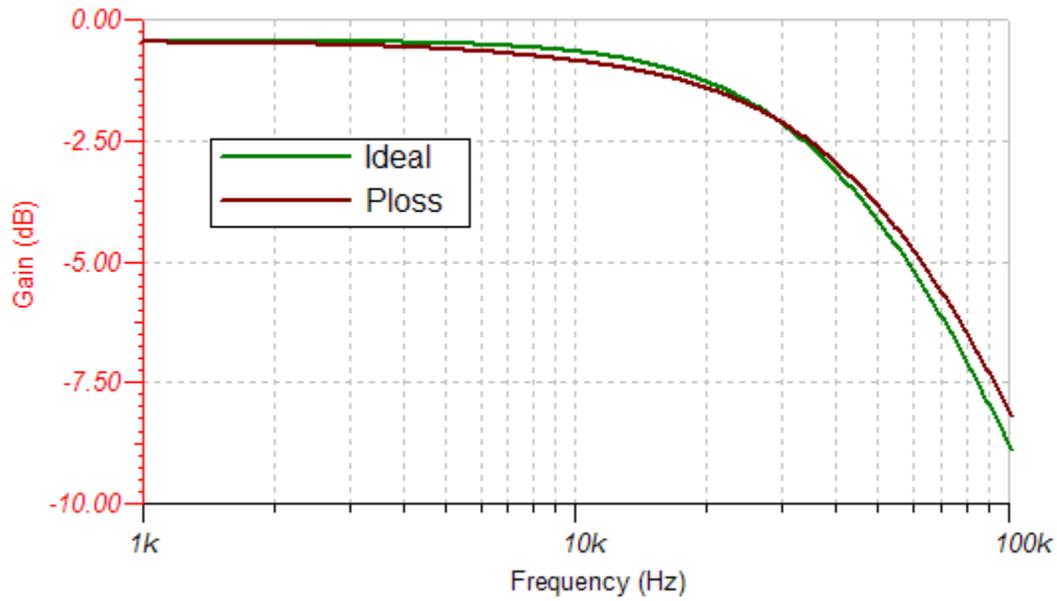


Figure 21 Simulation of idealized inductor and model with proximity effect loss included

Comparing the new simulation results with the actual measurements, we can see that it's a much closer fit. Further modification of the proximity effect model values would no doubt result in even closer match, but that was not deemed necessary as even with the depressed response at 20 kHz the filter is still useable.

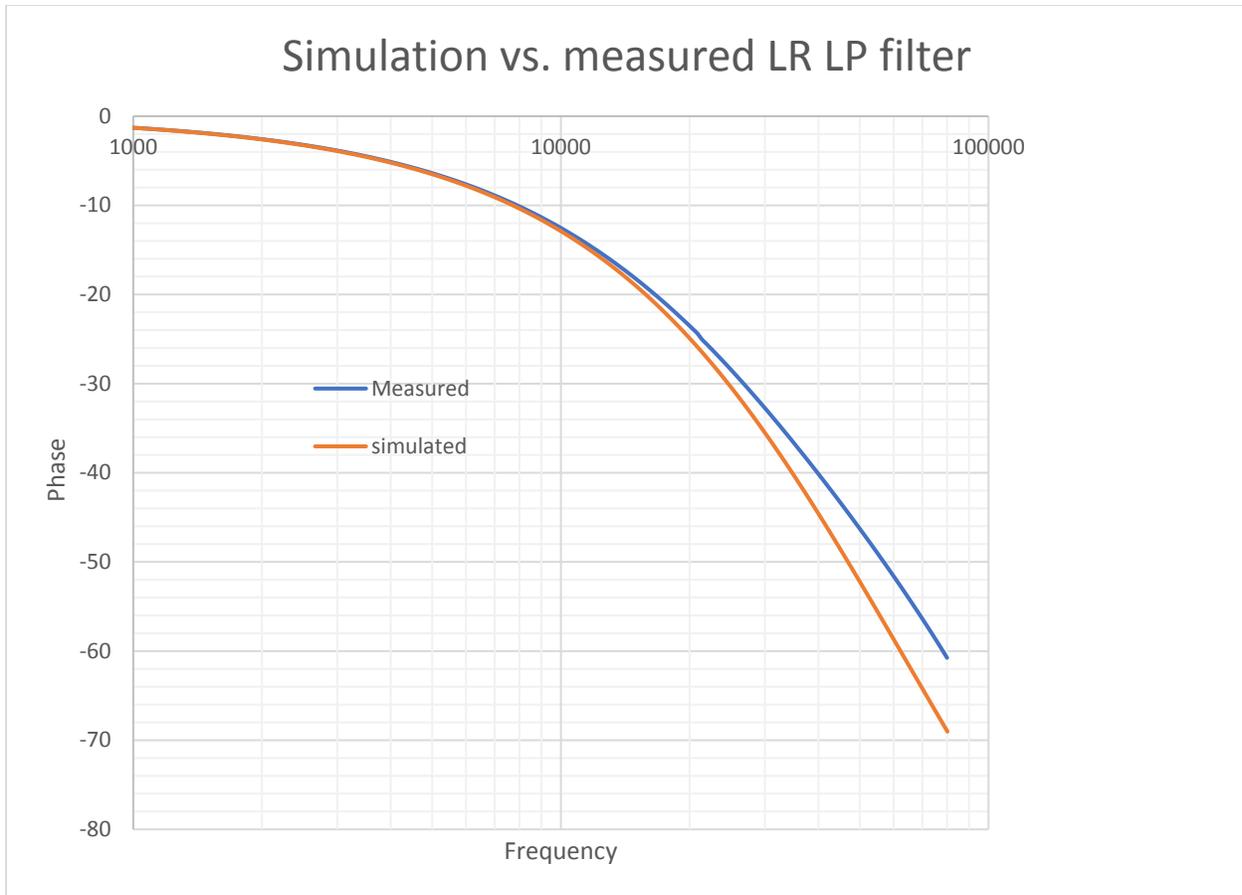


Figure 22 Model with approximated proximity effect included compared to the measured response

Comparing Figure 19 with Figure 22 it can be seen that the model with the proximity effect is a much better match in the audio band. Further model refinement was not needed as it seemed unlikely that some other mystery effect need be called upon.

CONSTRUCTION DETAILS

As only one filter was being constructed and the range of frequencies is low, the circuit was built on to a piece of perf board with point to point wiring.

The entire unit is constructed in a steel case with a hinged lid. While a cookie tin would also work, the strong construction of the steel, a lid that can't be lost, and better shielding of the steel were all a consideration. The downsides were a higher price (about \$15) and no snacks to eat during construction.

In addition to the basic filter function, a connection for a monitor speaker is provided, along with a switch to select the level: Soft, Loud, or off. With high power amplifiers the attenuation will not be enough and an additional series resistor should be added. Turning the monitor speaker on will slightly affect measurements; testing showed that

the changes from the load are at most a percent or two of the measurement value and not going to create problems compared to the other sources of measurement uncertainty.

A switch is added to the output to select the attenuation level in case the measurement instrument can't handle the 70 to 100V output of some high power amplifiers. For very high power amplifiers additional attenuation may be needed for some measurement equipment.

Fuses were added to prevent an accidental short from doing serious damage. ½ amp fuses are fine.

This design includes a 1 uF series output capacitor to block DC from reaching the measurement equipment. With a high impedance load (100 K ohm) that f_c is about 2 Hz; with 10 K it's about 16 Hz and that would be too high for audio use. A 22 uF cap would be more appropriate for use with mid impedance inputs.⁶

The capacitors used are rated at 600V or better (which actually means an AC rating of 200V for the parts used).

With high power amplifiers a few watts will be dissipated across the 1K (DC) impedance of the input. A 1 KW amp will typically produce 80V output which results in 9 Watts in the filter input. As they are rated for 12 W total running at full power for extended periods would not be good. If planning to use this with large amps 5W or 10W resistors should be used. For non-inductive wirewounds these become expensive.

Make sure all painted surfaces are sanded to bare metal where ground connections are needed.

Though measurement showed the cover was grounded through the hinge, a wire was added anyway to ensure long term conductivity. The top of the case/bottom of the cover was not sanded at the contact line as the cover does not sit tightly. If a RF project was done then sanding and RF gasketing would be needed in this type of box.

⁶ The unit was built partially with on-hand materials and I didn't think about the larger bipolar cap until later

PARTS LIST

The parts are generally not that critical and reasonable substitutions can be made.

quan	manuf PN	Manuf	Description	Distri PN	Distri
<u>Main board:</u>					
1	169P59XXXP	Vector	Breadboards 6.00 X 17 X .062 042" Hole Diameter	574-169P59XXXP	Mouser
8	WNE250FET	Ohmite	Wirewound Resistors - Through Hole 5W 250 ohm 1%	588-WNE250FET	Mouser
2	F462AK123F1K0A	Kemet	Film Capacitors 1000V 12000pF 1%	80-F462AK123F1K0A	Mouser
4	ECW-HC3F362J	Panasonic	Film Capacitors 3.0kV 0.0036uF 5% MPP L/S=22.5mm	667-ECW-HC3F362J	Mouser
4	257-054	Dayton	2 mH 20 AWG coil DCR 1 ohm	257-054	Part Express
2		Xicon	75 Ohm 10W Resistor Wire Wound 5% Tolerance	016-75	Part Express
1			3AG / AGC Screw Terminal Fuse Holder 5-Pack	070-605	Part Express
2			1/2A 3AG fuse		
as needed			Solder posts		
as needed			bus wire		
<u>Installed in chassis:</u>					
2	6883	Pomona	Test Jacks Dual Binding Posts	565-6883	Mouser
1	WTA-60R-4PK	Jameco	Green binding post (out of a 4 pack) 1 uF (film) or 22 uF (bipolar electrolytic) cap. See text.	2205755	Jameco
1	AC3MMDZ	Amphenol	XLR "D" Flange Panel Mount Male Nickel	092-0154	Part Express
1	1MD1T6B11M1QE	Jameco	SWITCH,TOGGLE,DPDT,ON-ON, SOLDER,5A@120 & 2A@250 VAC	317472	Jameco
1			SPDT Mini Toggle Switch Center Off	060-514	Part Express
as needed			heat shrink for isolating wires/components (like the bannana post studs)		
as needed			Standoffs (4-40 1/2" typical - 8 used)		
as needed			Screws for standoffs (4-40 1/4" typical - 16 used)		

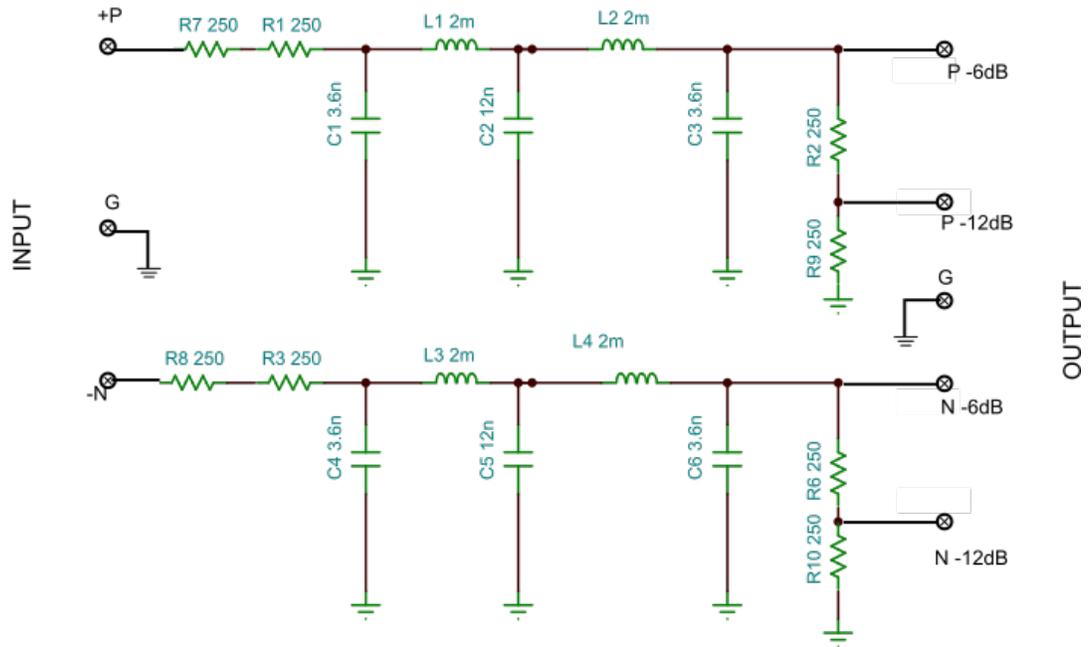
as needed
as needed
1
as needed

226-95

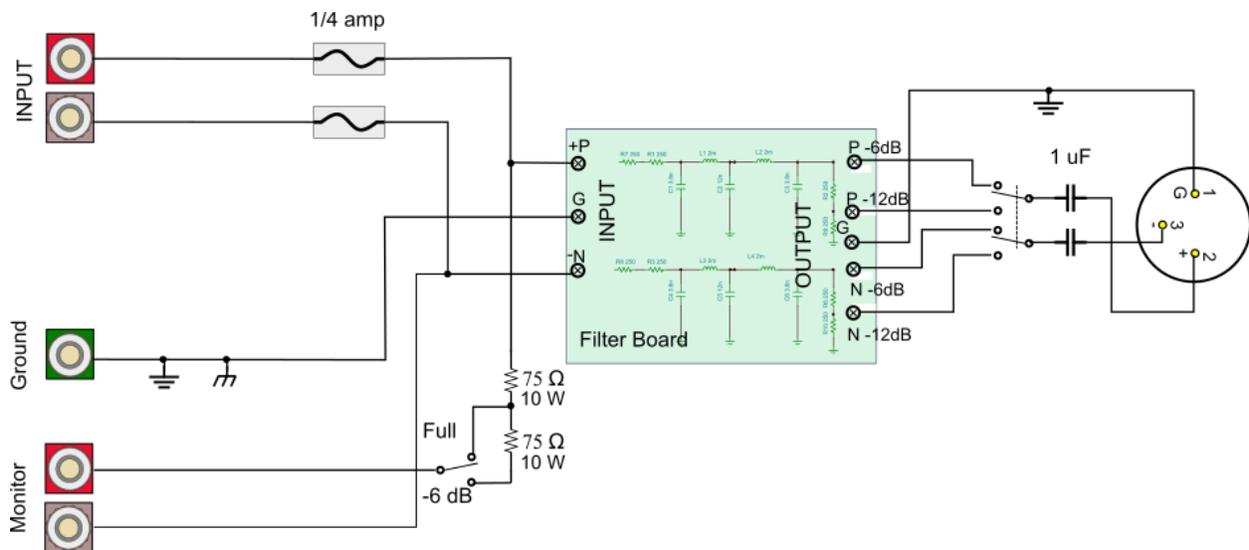
Durham

Interconnect wire
crimp connectors
Chassis 13x9x2 (inch) 340x235x51 (mm) [or similar]
Labels

SCHEMATIC



All resistors above are 3 W. If measuring high power (> 400W in to 8 ohm) amps 5W or 10W parts should be used.



See earlier note about using 22 uF bipolar caps at the output for mid impedance loads. If measuring high power amps the 75 ohm resistors in series with the monitor speaker should be doubled or an additional cut range added.

FINAL MEASUREMENTS

See the discussion associated with Figure 5 regarding 60 Hz EMI issues and how it was resolved.

Final noise measurements were 1.1 μ V RMS (AES-17) which is the noise floor of the AP515 being used.

A full measurement BW (90 kHz on the AP515) actually measures less noise on the filter than the AP loopback as the filter is cutting some of the AP's noise. The filter measured 2.7 μ V RMS and the AP loopback 4.4 μ V RMS.

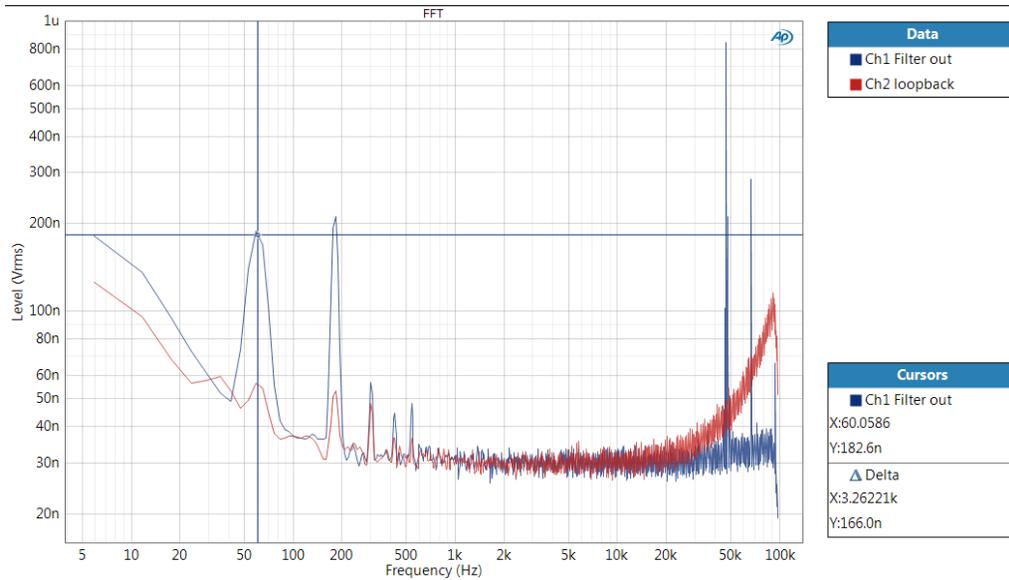


Figure 23 Wide bandwidth measurement (32 K pt FFT AP-Equi filter, 64 averages)

A large spike at 48 kHz is assumed to come from the AP. It's interesting to note the AP515's output contains increasing noise over 20 kHz.

THD+N was at the measurement limits.

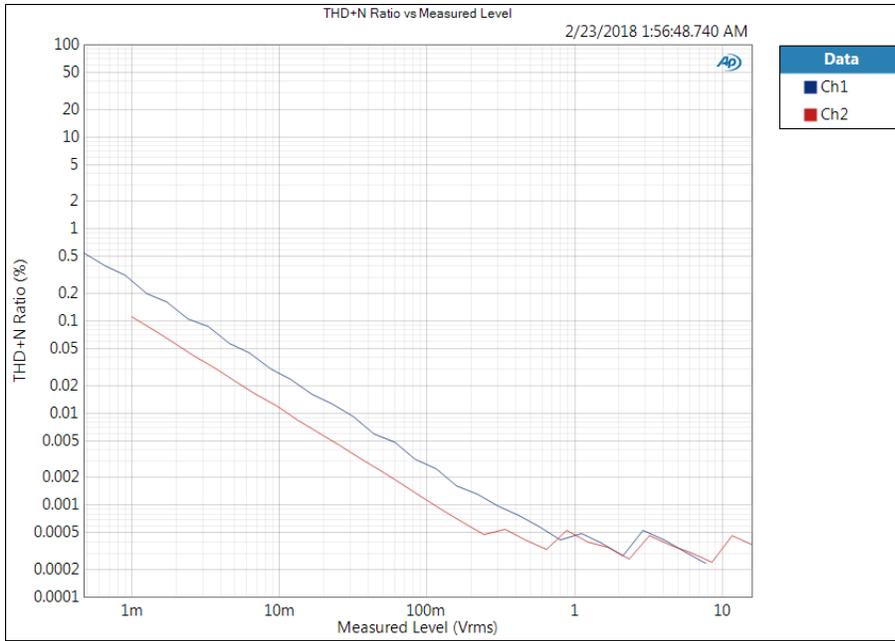


Figure 24 THD+N of filter (Ch 1) and loopback (ch 2). Offset at lower levels is due to the 6 dB attenuation of the filter

Response is 0.5 dB down at 20 kHz due to the previously described proximity effect losses in the inductors. The -3dB point is around 50 kHz. The additional (with respect to the filter's -6dB insertion loss) 0.5 dB attenuation at low frequencies is from the 50 ohm impedance of the AP generator.

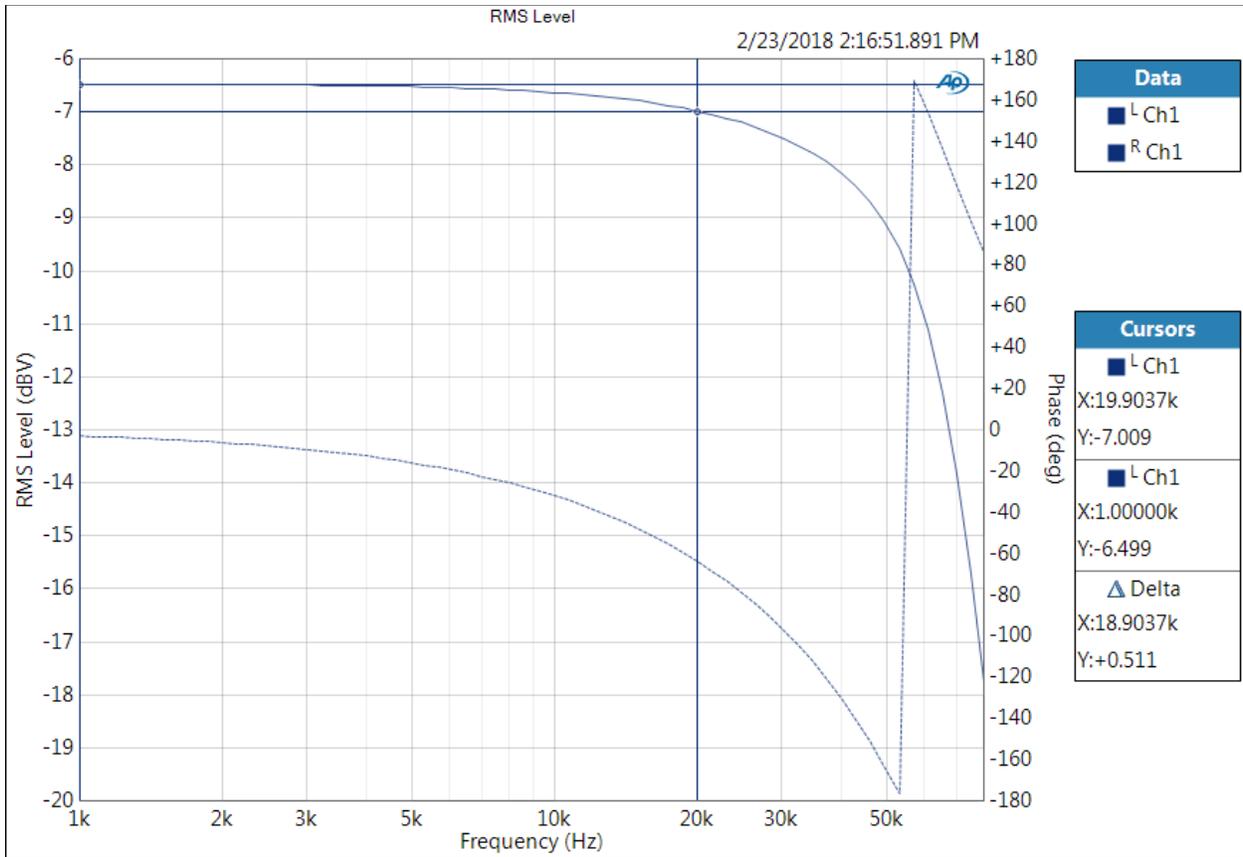


Figure 25 Frequency and phase response

The AP is limited to 80 kHz so manual measurements were made with a signal generator and scope, though beyond 200 kHz the attenuation was too great to get a reliable measurement with the available equipment. There was no evidence of any increase in output levels over 200 kHz up through 3 MHz.

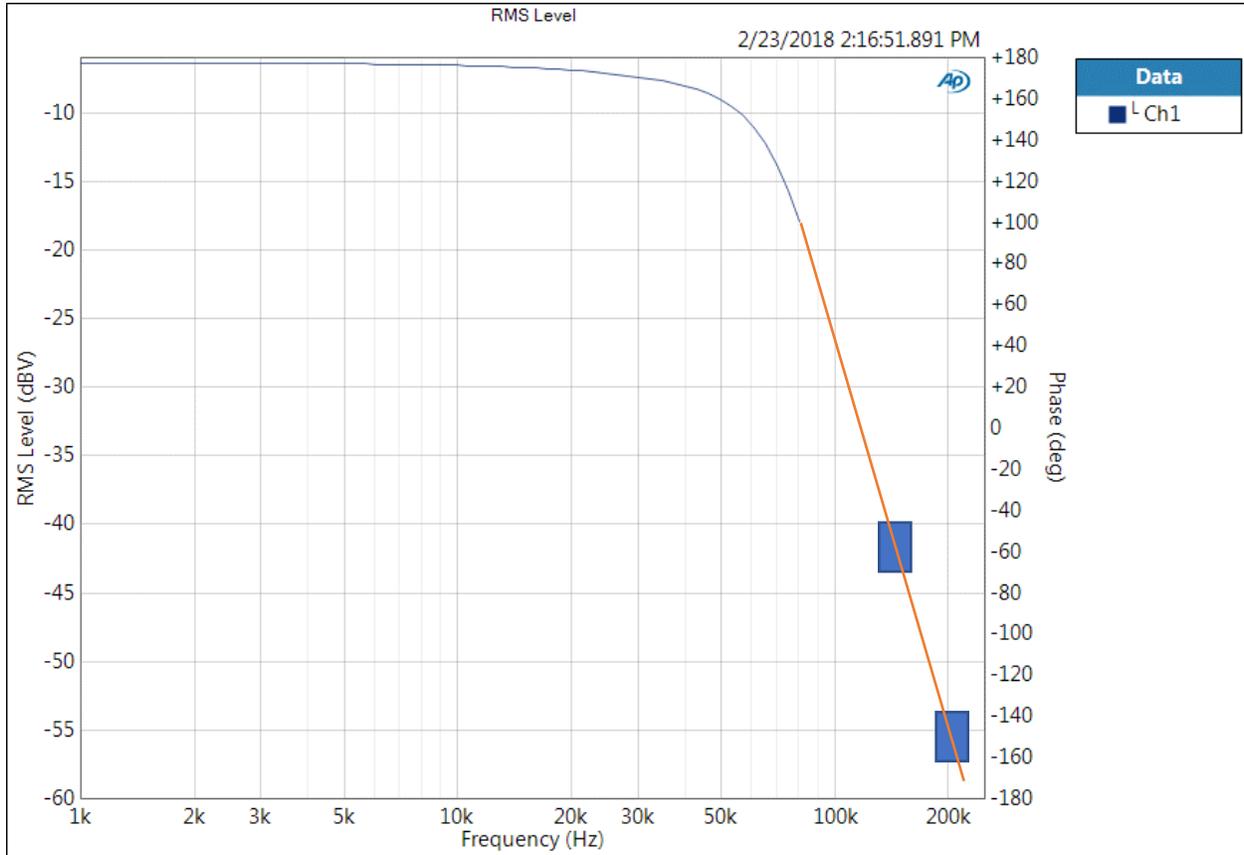


Figure 26 Extended manual measurement of frequency response to 200 kHz (with crude blue error boxes and brown line)

PICTURES



Figure 27 Front view showing convenient handle (a major design feature!) and cover that can't be lost compared to cookie tin based construction



Figure 28 The 70V is just made up. The parts used could go higher except the resistors will melt.



Figure 29 Low tech assembly of the insides

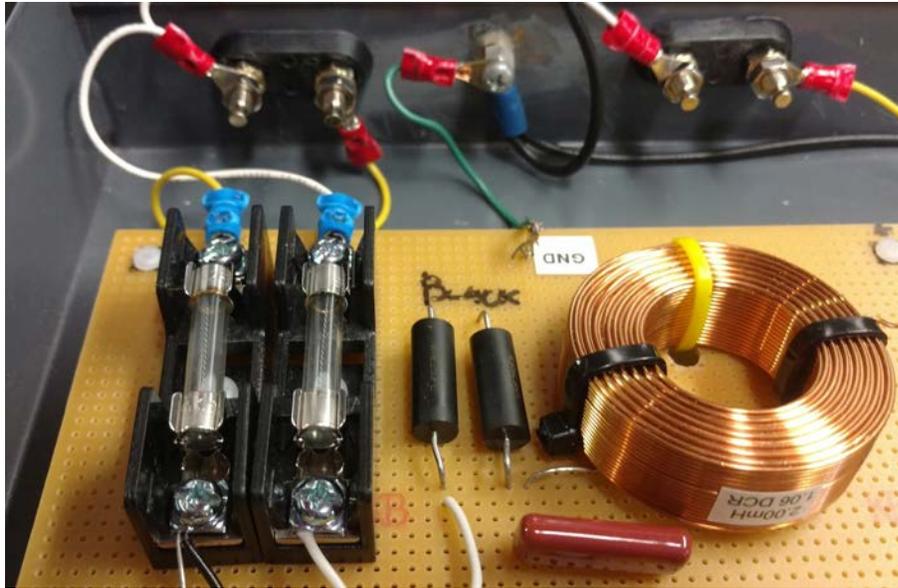


Figure 30 Close up of input section and fuses

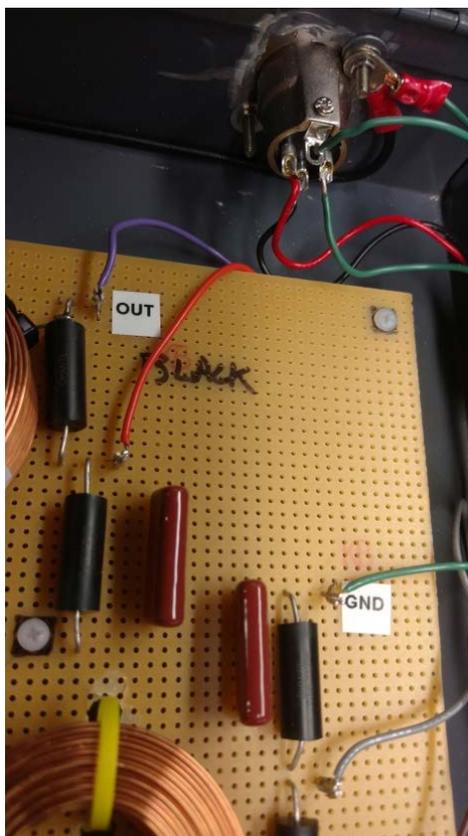


Figure 31 Output section. Note ground wires added to shell connection.



Figure 32 Output range switch and DC blocking caps. See note in text about using 22 uF bipolar instead.