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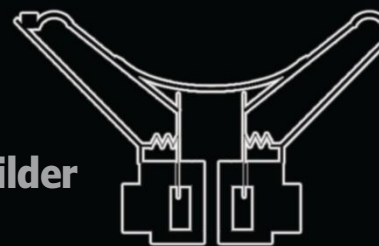
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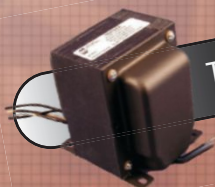
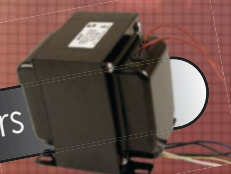


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How We Began

The origins of this Glass Audio Special that *audioXpress* features in May every year, goes back to our magazine's roots when Edward T. Dell, Jr. founded a quarterly periodical called *Glass Audio*. That magazine reflected an increasing interest for tube-based audio electronics. There are many reasons why tube audio made such a strong comeback since that period and has remained strong. Some of those reasons are explained in the pages of the original *Glass Audio* magazine. In 1994, Dell wrote an editorial, titled "How We Began" to remind everyone of why he did it.

"A veteran electronics show watcher reported just a few days ago that a walk through the aisles of the Winter CES (1994) while making a count of the audio tube electronics versus solid state ended with a tally of 60% tubes, 40% transistors. My friend was full of admiration at the fact that I had launched this periodical in the late months of 1988.

Before anyone makes any assumptions about whether I am a prophet, or the son of one, let me talk about how this magazine came into being. During 1987 I began to notice that interest in tube electronics seemed to be attracting more and more attention by many of our nearly 300 authors. I also noticed that despite the popularity of solid-state gear, a thin stream of dissatisfaction with it continued to appear in reviews in many magazines.

When more and more tube articles from authors started to land on my desk, it just seemed that a magazine dedicated to tube technology was a good idea. Thus, the first issue No. "0" of Volume 1 was mailed late in 1988.

Another major effort is the marriage of digital techniques to tube topologies. We are all aware that solid-state ancillary devices can enhance good tube sound, and that today's components for building tube gear are giving us results, which have heretofore been impossible. Add to that the former secret research in the old USSR, which produced brand-new tubes with unprecedented capabilities, and we have some exciting adventures ahead of us."

In the same January 1994 issue of *Glass Audio*, Norman Tracy submitted a project for a "Tube DAC," where he commented, "After a huge buildup prior to their official launch, the first CD players were a tremendous disappointment to me. I can recall gritting my teeth and wincing through demos, trying in vain to enjoy music through the new medium. (...) I now know the missing ingredient: emotional involvement. Somehow, though, I managed to stumble onto the means to extract this most elusive virtue from digital audio."

And he goes on from there to describe the discovery of external, improved, quality digital-to-analog converters, how to approach digital filters, and the benefits of oversampling. Interestingly, his Burr-Brown based DAC project with a tube analog stage took most of the 54 pages in that same edition of *Glass Audio*! This project was subsequently improved with an updated tube output circuit by Joe Curcio—a regular contributor for *Glass Audio*, who maintains a tube-focused website to date (www.curcioaudio.com).

Fast forward 24 years (yes, it's been that long) and we are still discussing the merits of digital versus analog audio, while the mainstream music industry is now based on the streaming of digital music files, which are mostly mastered in high-resolution 24-bit and 96 kHz sampling, frequently going through some tube-based stage for adding those "euphonic" effects resulting from even-order harmonic distortion. Guitars and many musical instruments mostly continue to be recorded using tube-based preamps, while Neumann-Berlin just re-issued its famous U67 Tube Microphone.

Recently, I was quite surprised (and pleased) to discover the Copper Reference Premium Stereo Audio Interface (www.tracktion.com/products/copper-reference), a modern USB Type-C recording interface from Tracktion, a software company based in Seattle, WA. The Copper Reference features a 12AX7 tube stage on the analog inputs, designed by Mike Soldano (www.soldano.com), followed by a dual AKM AK4396 ADC (summing quad to stereo for increased dynamic range), and ESS 9038 Pro Sabre digital-to-analog conversion.

The fact that a software house addressing the studio and musicians market decided to make a detour into hardware to create a no-compromise solution they thought was missing for them, is very telling. I have no doubt that the reasons why tubes returned in strength in the late 1980s, as discussed by Ed Dell, remain valid. Most importantly, the *Glass Audio* and *audioXpress* content legacy remains an important tube audio reference for future generations.

J. Martins

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COLUMNISTS

Vance Dickason has been working as a professional in the loudspeaker industry since 1974. He is the author of *Loudspeaker Design Cookbook*—which is now in its seventh edition and published in English, French, German, Dutch, Italian, Spanish, and Portuguese—and *The Loudspeaker Recipes*. Vance is the editor of *Voice Coil: The Periodical for the Loudspeaker Industry*, a monthly publication. Although he has been involved with publishing throughout his career, he still works as an engineering consultant for a number of loudspeaker manufacturers.

Dr. Richard Honeycutt fell in love with acoustics when his father brought home a copy of Leo Beranek's landmark text on the subject while Richard was in the ninth grade. Richard is a member of the North Carolina chapter of the Acoustical Society of America. Richard has his own business involving musical instruments and sound systems. He has been an active acoustics consultant since he received his PhD in electroacoustics from the Union Institute in 2004. Richard's work includes architectural acoustics, sound system design, and community noise analysis.

Mike Klasco is the president of Menlo Scientific, a consulting firm for the loudspeaker industry, located in Richmond, CA. He is the organizer of the Loudspeaker University seminars for speaker engineers. Mike specializes in materials and fabrication techniques to enhance speaker performance.

Steve Tatarunis has been active in the loudspeaker industry since the late 1970s. His areas of interest include product development and test engineering. He is currently a support engineer at Listen, in Boston, MA, where he provides front-line technical support to the SoundCheck test system's global user base.

Ron Tipton has degrees in electrical engineering from New Mexico State University and is retired from an engineering position at White Sands Missile Range. In 1957, he started Testronic Development Laboratory, which became TDL Technology, to develop audio electronics. All product sales and services were terminated on December 31, 2015, but the TDL website is still online with a variety of audio information and downloads.

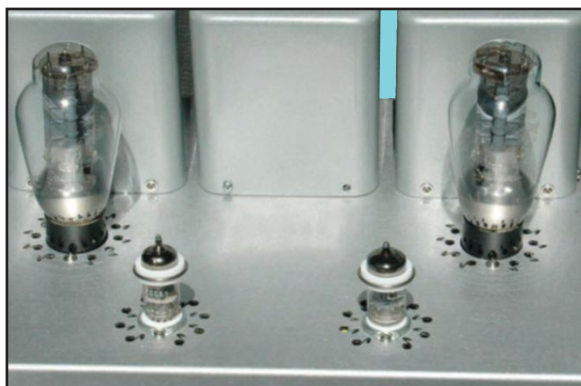
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● Glass Audio Special

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By Gerhard Haas

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42 DIY Tube Amplifier Heat Management Options

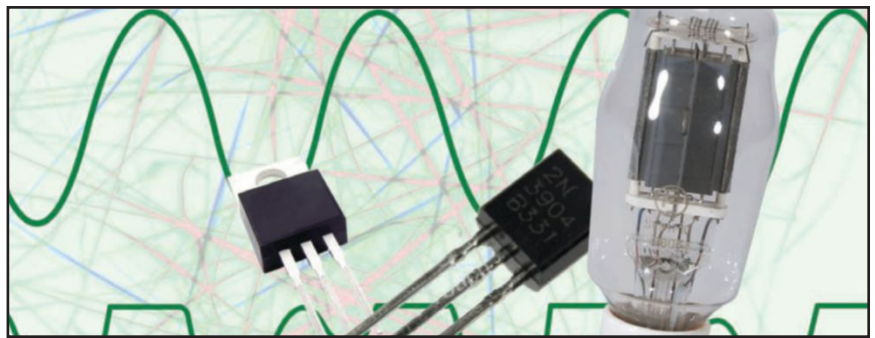
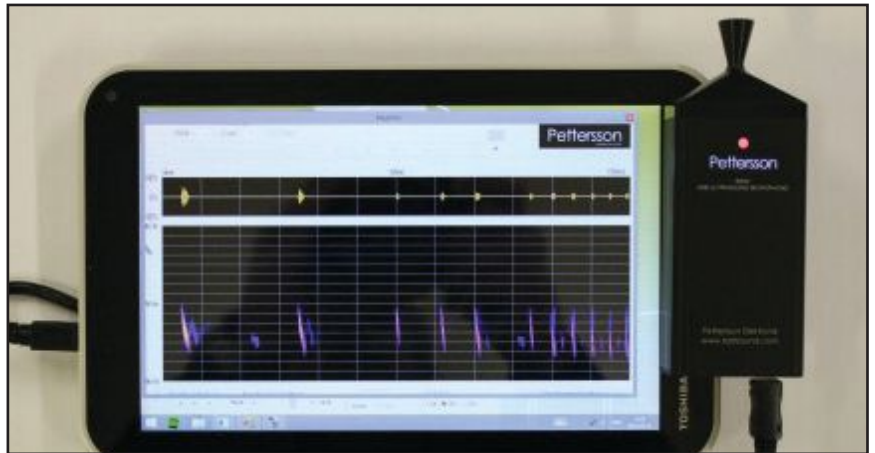
By Patrick Brunner

Tube amplifiers generate the most heat per watt of audio output. Our author shares some methods to greatly increase the performance and the reliability of your DIY tube amplifier.

46 TubeSociety 2A3-300B SE Amplifier Project (Part 2) Saving the Best for Last

By Menno van der Veen

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Learn more about this book authored by *audioXpress* columnist Richard Honeycutt. In *Acoustics in Performance*, Honeycutt details "all you need to know about good acoustics and sound systems in performance and worship spaces."

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By Oliver A. Masciarotte

Oliver Masciarotte has an opportunity to sit down with Jason Stoddard, to discuss his company Schiit Audio, a high-value brand available at reasonable prices.

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Exploring Master Quality Authenticated (Part 1)

Measuring the Difference

By now, I expect every music enthusiast has heard about Master Quality Authenticated (MQA) and how the high-frequency content is folded, origami fashion, into the low-frequency region. An MQA decoder (DAC) unfolds the highs back to their proper spectral location. The theory sounds great but because we can't directly hear this upper register, how do we know what's actually there?

By
Ron Tipton

(United States)

I played my source file through an MQA DAC (Meridian Explorer 2) and a non-MQA DAC (TEAC UD-301) to another computer running the TrueRTA spectrum analyzer set to a 20 Hz to 50 kHz display, and exported the spectral data to two text files for plotting. The results are graphed in **Figure 1** with the black line detailing the MQA decoding. There is

definitely more MQA energy in the 22 kHz to 50 kHz region—10 dB to 20 dB above the system noise level.

My source, a FLAC MQA file, was the "From a Distance" track from the album *Distance* written and performed by Martin Tingvall, chosen because I enjoyed listening to it. The album is available as a download from www.highresaudio.com. (The tracks are MQA in a FLAC container file: 24-bits and 48 kHz)

The Explorer 2 is a bit "fussy." It seems to work fairly well with drivers such as Direct Sound, WASAPI, and "Speakers" (the default for many sound cards). But an ASIO driver is required for bit-perfect playback and the Mode LED stays white if this need isn't met. Meridians' tech support sent me an email detailing the setup for the free foobar 2000 media player. It's also worth noting that it's easy to make a file non-bit-perfect. It is best to use the original downloads without changing the bit resolution, sample rate, or even the file name.

The Next Step

By using the magic of a product developed for listening to bat sounds, we have a tool to view and listen to what is produced by a high-frequency speaker (tweeter). This product is the model M500 USB ultrasonic microphone and BatSound Touch software from Pettersson Elektronik AB in Sweden. (This software is available only for Windows 7, 8, and 10.) The M500 is shown in **Photo 1** along with the tripod adapter that I added to make it more useful for my measurements.

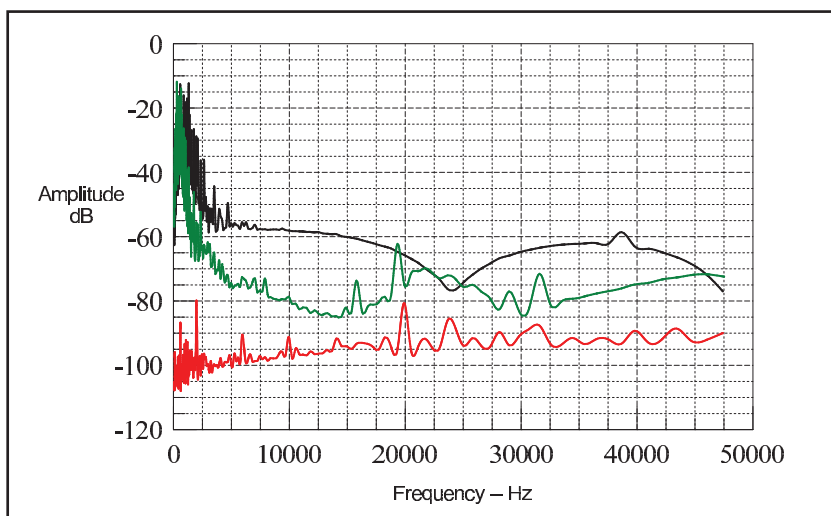


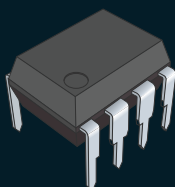
Figure 1: The MQA encoded track was played through the Explorer 2 and its spectrum was captured by TrueRTA running on another computer. The display was captured as a text file for plotting. The track was played again through the UD-301 and captured. A third capture of the system noise was made with nothing playing. The black line belongs to the Explorer 2, the green line to the UD-301 and the red line is the system noise. I used a linear frequency scale to expand the 20 kHz to 48 kHz region.

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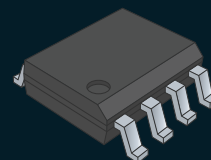
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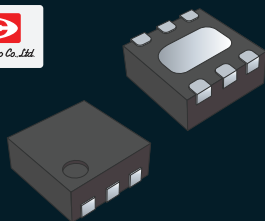
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The Measurement Setup

The block diagram in **Figure 2** shows my measurement setup. I again used the Meridian Explorer 2 MQA DAC. As of now, there are several DAC options, most outside of my price range. (Pioneer, Onkyo, Bluesound, and others have MQA-enabled players but not standalone DACs.) The off-the-shelf items: the model 414A stereo to mono converter, the model 412 variable gain amplifier, and the power amplifier were checked for a flat amplitude response (± 0.5 dB) up to 100 kHz.

The High-Pass Filter

Revisiting Figure 1, the ultrasonic amplitudes are generally much lower than the low- and mid- audible frequencies, so a high-pass filter is needed as a crossover network to keep the audible frequencies from “wiping out” the tweeters. I spent a lot of time and effort on this filter—building five different versions. The first was a 14-pole active

Butterworth using $\pm 1\%$ metal film resistors and $\pm 5\%$ film capacitors—the response was terrible. I did a 10-pass Monte Carlo simulation in TopSpice and I knew the response would not be too good, but it was worse than predicted.

I’ll skip the next two failures and continue with the fourth version. It was a 10-pole active Butterworth using $\pm 1\%$ metal film resistors and $\pm 1\%$ NPO ceramic capacitors. This one worked fairly well but not quite good enough in my preliminary tests.

For the fifth try, I added another filter in its own metal enclosure with its own power supply. With its cutoff frequency at 19 kHz, its amplitude response was down only 0.3 dB at 22 kHz, so the series combination was still down about 3 dB at the 22 kHz design frequency. Although not as steep as an actual 20-pole filter, it’s still “pretty darn” good and much easier to actually build. **Figure 3** shows the simulated and measured responses. There are five different colored lines: the measurement line is black with the others from four Monte Carlo simulation runs. The circuit diagram and SPICE simulation file are included in the Supplementary Material found on the audioxpress website.

Figure 4 depicts the filtered ultrasonic spectrums from the two DACs and the residual filtered noise with no input signal. Both were measured at the output of the model 412 amplifier with the TrueRTA analyzer. The noise is at least 5 dB below the signal minimums.

Power Amplifier

I wanted a flat response to 100 kHz, so I built an amplifier using a pair of circuit boards designed around a Texas Instrument (TI) LME49830TB integrated driver. Depending on the components used in its compensation circuit, the power bandwidth can be set as high as 400 kHz. I found an already-built board on eBay set to about 110 kHz. I bought a pair of them and used them in my design.

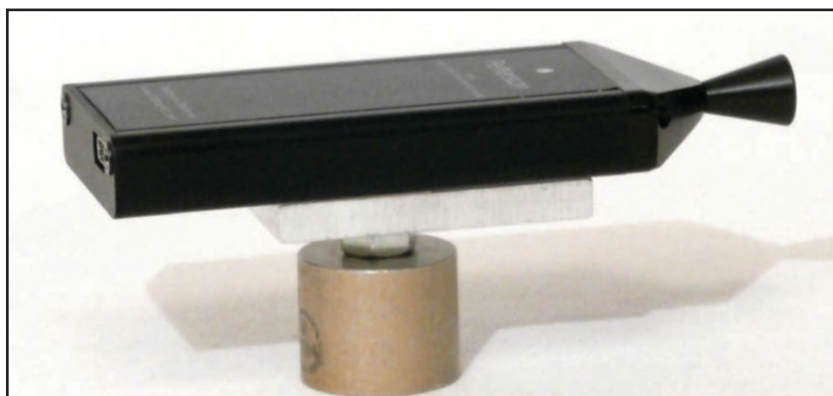
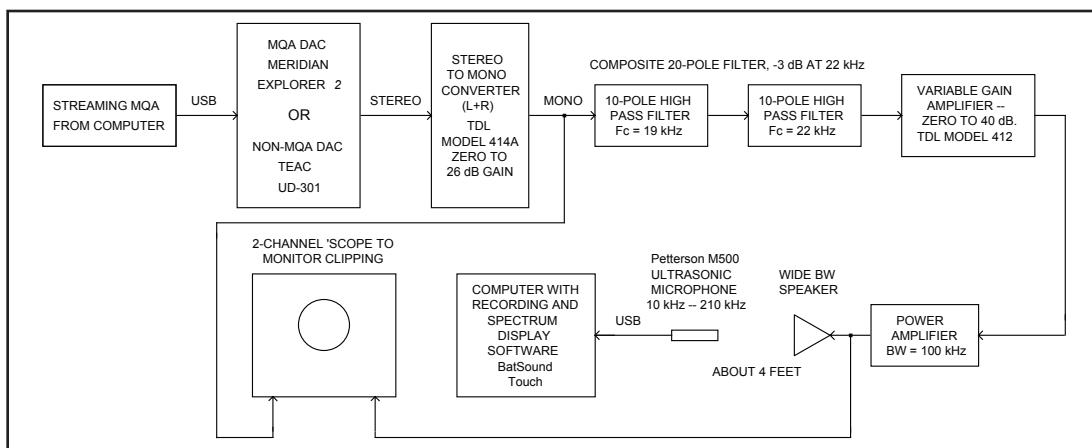


Photo 1: This is the BatSound M500 ultrasonic microphone with the tripod mount adapter I added. The aluminum plate is attached to the microphone’s bottom plate with double-sided tape. It is drilled and tapped with a $\frac{1}{4}$ -20 thread, the cylinder screws into this hole. The USB digital output is a 16-bit stream at a 500 kHz sample rate. A pdf copy of the user manual is included in the Supplementary Material

Figure 2: My measurement setup block diagram. The MQA encoded track was played with foobar2000 running on Windows 10. Both the Explorer 2 and UD-301 DACs have a USB input. A Windows 7 computer was running the BatSound Touch program.



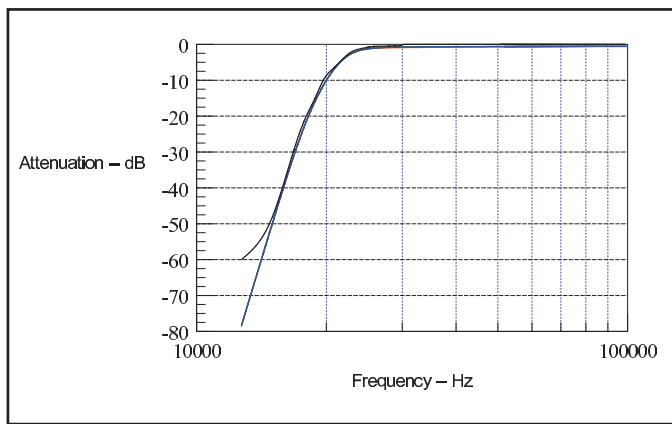


Figure 3: The amplitude vs. frequency response of the composite 20-pole high-pass filter. The multi-colored line is composed of four Monte Carlo simulations using TopSpice. These runs are closely spaced because of the $\pm 1\%$ resistors and capacitors. The black line is the filter's measured response.

I plan to write an article about the amplifier's design and construction so I won't go into more detail here except to mention that getting the heat out of the power MOSFETs proved to be a challenge, which I solved by constructing a copper "L" bracket because I wanted to use a 3.5" high enclosure rather than a taller one in which I could have vertically mounted the board.

The Tweeters

I tried three different tweeters connected one at a time to the power amplifier (see **Photo 2**). On the left side is a BESTON model RT001A ribbon tweeter, with a 3 kHz to 40 kHz response and a 6 Ω input impedance. In the center, a TB model 25-1983 1" titanium dome; 500 Hz to 40 kHz and 8 Ω . On the right, a BCT model L010 piezo electric; 2 kHz to 60 kHz with a measured capacitance of 150 nF. Using the advice of a couple of websites, I built the network shown in **Figure 5** to help tame

the harshness that seems to be inherent in purely piezo tweeters.

Listening To Noise

Using the setup shown in Figure 2, I found I obtained the best results with the L010 piezo electric tweeter. Perhaps its sensitivity is higher than the other two or the higher frequency response made the difference. The BatSound Touch software displays a spectrogram and also records the microphone's output as a BatSound wav file, 16-bits at a 500 kHz sample rate. **Photo 3** is a screen capture of a portion of the MQA "From a Distance" track as decoded by the Meridian Explorer 2 and high-pass filtered.

The Touch program has the ability to play the captured .wav file in the audible range by down sampling. The audible sound was captured by another computer and recorded as a 24-bit, 96 kHz .wav file. A 30-second segment is included in the

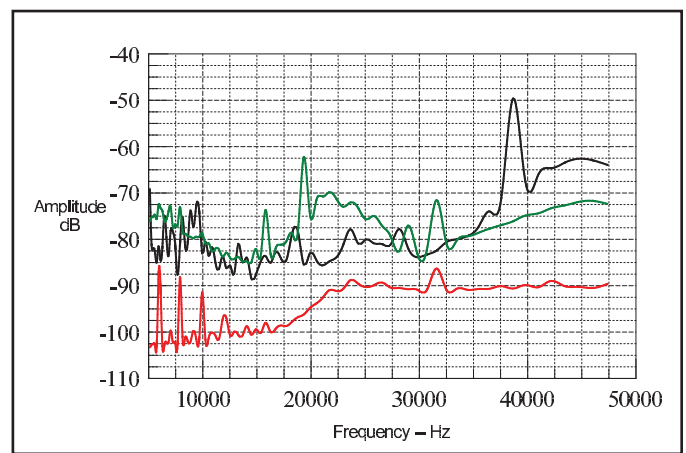


Figure 4: The black line shows the MQA encoded track decoded by the Explorer 2 DAC after passing through the high-pass filter and amplified by the model 412 Audio Attenuator-Amplifier. The filtered signal has much improved amplitude resolution over what can be seen in Figure 1. The red line shows the noise out of the high-pass filter with the same model 412 gain as above. The noise level is at least 5 dB lower than the filtered signal's minimums.

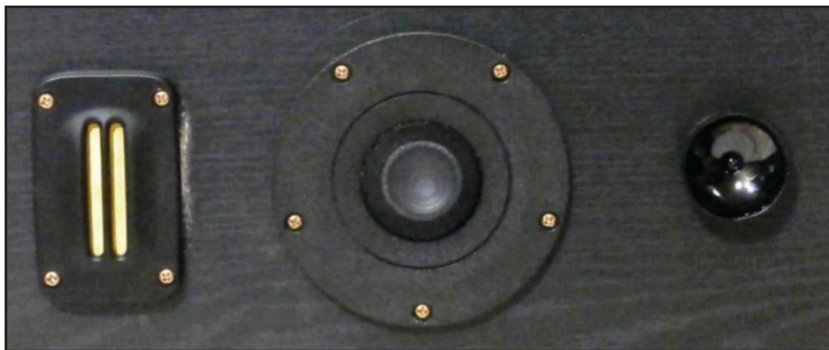


Photo 2: The three high-frequency tweeters are mounted on a 13" $\times$ 14" sheet of 0.5" MDF which is supported by a 0.75" thick MDF base plate. I got the best measurement results using the piezo electric tweeter on the right. It is very light weight, about 6 grams, and is simply stuck to the MDF with a piece of double-sided tape. Its compensation network (Figure 5) is housed in a small aluminum box fastened to the MDF base plate.

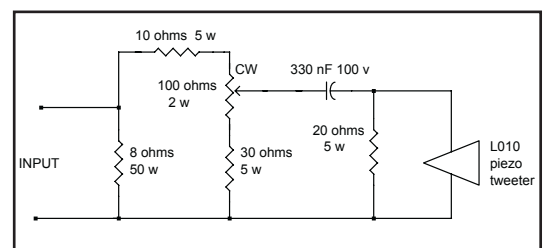


Figure 5: The circuit diagram of my compensation network that tames the harshness of a purely capacitive piezo electric tweeter. The single-turn adjustable resistor (pot) aids in impedance matching and does function as a volume control. The circuit is housed in an aluminum box fastened to the MDF base plate behind the speakers.

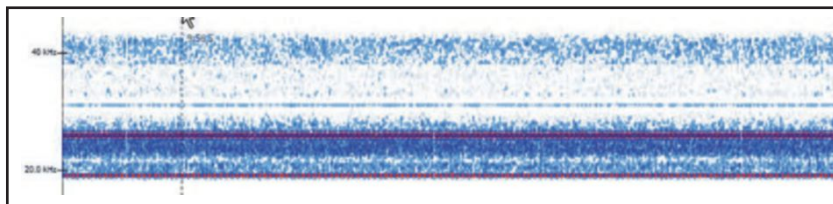


Photo 3: This is a screen capture of a portion of the MQA track spectrogram as decoded by the Explorer 2 DAC and filtered as displayed by the BatSound Touch software. The darker blue indicates higher amplitude and the red spots are even higher. The highest amplitudes are at and just below 20 kHz where this high-pass filter response has only about 7 dB loss.

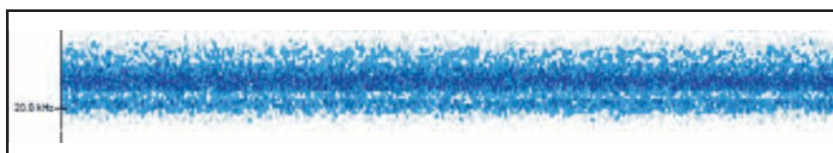


Photo 4: The screen capture using the non-MQA decoding TEAC UD-301 DAC displays just random noise.



Photo 5: This TDL model TSMD1 speaker is half of a stereo pair. The closed box measures 23.75" tall x 9" wide x 12" deep. Construction is 0.75" thick MDF for sides, top, bottom and back with a 0.75" thick 13-ply birch plywood front panel. The drivers include a pair of Dayton #295-337 7" aluminum woofers and a single Dayton PT2 planar tweeter crossed over at 3 kHz with a passive second-order Linkwitz-Riley network. Response is flat to within 3 dB from 100 Hz to 20 kHz.

Supplementary Material. Remember that the actual frequencies are about two times higher. It sounds very "noise-like" and according to Meridian, it is shaped noise. Just how this enhances the sound is probably another example of psychoacoustics in action. That is, our hearing adds the ultrasonic cues to the audible music. To get the full MQA benefit, you need a sound system able to reproduce frequencies to 40 kHz.

A playback through the TEAC DAC and high-pass filtered produced the screen capture shown in **Photo 4**. As expected, it seems to be just plain noise.

Final Thoughts

During the course of making these measurements I listened to "From a Distance" many times through both DACs and both sounded excellent. The playback system I used is pretty good with its Marco Ferretti hybrid power amplifier (*audioXpress*, February 2001), the speakers shown in **Photo 5** and a Daniel Ferguson 12" dual voice-coil feedback subwoofer (*audioXpress*, September 2004). However, the Dayton planar tweeters are fairly flat to only 20 kHz so I wasn't getting the full MQA sound.

Next month, I will conduct a similar review of an MQA track streamed from TIDAL. 

Project Files

To download additional material and files, visit <http://audioxpress.com/page/audioXpress-Supplementary-Material.html>

Sources

RT001A tweeter

BESTON Technology | www.ribbonspeaker.com.tw/RT001A-980514.htm

foobar2000

foobar2000 | www.foobar2000.org

BCT L010 tweeter

Kemo Electronic | www.kemo-electronic.de/en/Car/Speaker/L010-Piezo-Loudspeaker.php

Explorer 2

Meridian | www.meridian-audio.com/en/products/dacs/usb-dacs/explorer

TopSpice

Penzar Development | www.penzar.com/topspice.htm

BatSound Touch Software

Pettersson Elektronik AB | www.batsound.com

TB 25-1983 tweeter

Tang Band Speaker | tb-speaker.com/en/products/25-1983

TrueRTA

True Audio | www.trueaudio.com



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Back EMF Phase Relationships in Moving-Coil Loudspeakers (Part 1)

Identifying, Quantifying, and Modeling Back EMF

As a loudspeaker's voice coil travels through the magnetic field in the gap, a voltage is produced, referred to as an electromotive force (EMF). Because it impedes input, the adjective "back" is used when referring to this opposing voltage in electric motors, including moving-coil loudspeakers. In a loudspeaker, the phase relationships between the input signal and the Back EMF cause the impedance of the unit to be inductive, resistive, or capacitive, depending on frequency. This article will explain the sinusoidal relationships that make this so.



By
Andy Lewis

(Acme Sound, LLC)

To begin, let me make it clear that this article is not about loudspeaker design. These observations might not be particularly useful to any loudspeaker designer, including myself. Its purpose is to point out something I noticed about how moving-coil drivers work, when visualizing the position of a speaker's moving-assembly lagging in time behind its input signal.

I had been considering how it was possible for the impedance of a driver to be purely resistive at resonance, such as is that of a DC rotary motor. In a moment of clarity, as I visualized the phase relationship at that frequency graphically, I realized that the resistive impedance itself revealed where the cone had to be at any instant in order for the impedance to be resistive.

After that realization at resonance, which is the simplest permutation, I had only to "slow things down," by doing "thought experiments," to see the similar relationships below resonance, and "speed things up" to reveal the driver's behavior above resonance.

After I qualitatively realized what was happening, it was then a matter of reconciling my observations and impedance measurements with what I know

about reactive impedances and Newtonian physics. By the time I finished, it all seemed to make sense, and I present it for your perusal.

I hope this article will successfully communicate the only description of this activity I've seen in my 50-plus years of paying attention.

When originally completed, this article was five times as long. On the advice of people I trust, I present this condensed version. I hope you can get a fairly clear picture of what I'm attempting to convey. You might notice subtle references to content not otherwise in evidence.

If it seems as though anything has been under-explained or unsupported (including, but not limited to, the restorative force of the cone's suspension or voice-coil self-inductance), it has. If necessary, I can post further content online, as necessary, or could respond to questions, time permitting.

Abstract

In this article, we explore the generation of Back EMF in the region of a loudspeaker's resonance. It examines EMF produced in a rotary DC generator, and expands to model the Back EMF in a DC motor.

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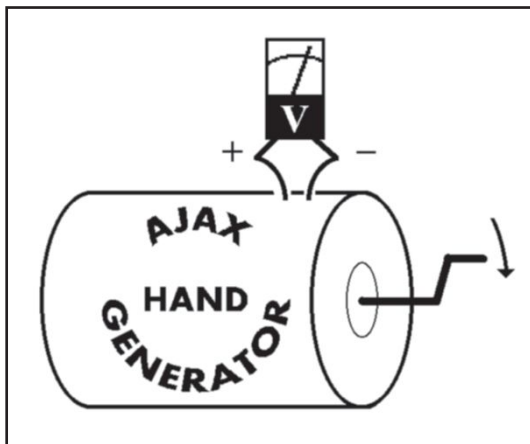
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Figure 1: This is an example of a hand-crank powered DC generator and a voltmeter.



It then discusses the loudspeaker as a linear AC generator, before considering it as an AC-driven moving-coil loudspeaker. It then shows how to model the Back EMF of a driver with a parallel LCR filter, and specifies the values of the components for a particular woofer. In the article, I also demonstrate a method for calculating the position of the moving assembly, or "cone," relative to input signal, using complex impedance measurements.

The second part of this article defines the phase relationship between input current and cone position as "slip." It then explains how the degree of slip can be used to explain how and why the Back EMF of a driver manifests itself as an inductive reactance below resonance, a resistance at resonance, and a capacitive reactance above resonance.

The article also illustrates how and why the Back EMF at no time "drives" a current, but at most, shifts

Figure 2: A "Generator" driven by an external voltage is a motor.

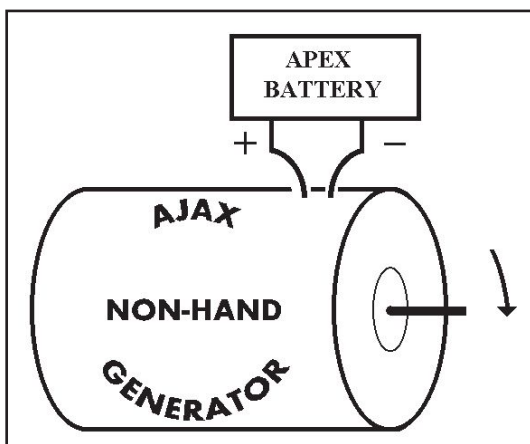
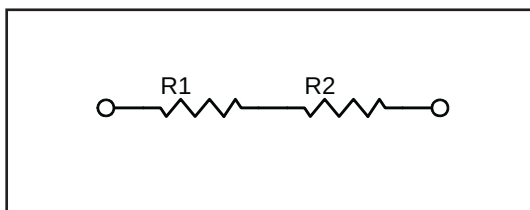


Figure 3: The DC motor is represented as two resistors.



the phase of input current relative to input voltage, just as any reactive impedance.

Precis

In a simple DC generator, wire coils turn on an axis, through a magnetic field, and produce a voltage ($V = Blv$). The armature travels in a constant direction, and the voltage thus produced is DC. This DC voltage drives a current through the resistance of the coils, in series with whatever external load is present.

When the same device is driven by an applied DC voltage, as opposed to an external physical force, the generation of EMF is unchanged ($V = Blv$), but the EMF produced opposes the input current, just as if it were a resistor. For this reason it is referred to as a "back" EMF, and can be modeled as a resistor.

Therefore, the DC motor can be modeled by using two resistors in series, one being the resistance of the coils, and the other being the resistance of the Back EMF.

Just as an external physical force can turn a rotary DC generator to produce a DC voltage, a loudspeaker driver can be driven by an external driving force to produce an AC voltage. When we do so, we can observe that the constantly changing position of the cone has a fixed 90° derivative relationship to its velocity (rate-of-change-of-position), which is phase with resulting EMF. Velocity precedes position.

Just as the Back EMF present in the DC motor presents a resistance to input voltage, and thus can be modeled by a resistor, the Back EMF in a moving-coil driver (loudspeaker) presents an impedance to input voltage, and can be modeled by an RCL parallel filter in series with the loudspeaker's DC resistance. Just as the Back EMF of the resistor in the DC motor has a specific value, these reactive components have specific, identifiable values.

Due to inertia, the position of the loudspeaker's cone lags behind input current. We will define this delay in time as "slip," which can be expressed as a phase angle. The precise phase angle between the two can be calculated from a driver's complex impedance information.

Degree of slip at a given frequency is what determines the magnitude and phase of the Back EMF, which opposes input voltage. This causes the phase angle of the impedance of the Back EMF to be inductive below resonance, resistive at resonance, and capacitive above resonance.

EMF in a Simple DC Generator

A loudspeaker driver is a motor, albeit an AC linear motor (to and fro). We will start by comparing

it to a less complex motor: The DC rotary motor (round and round). Even more fundamentally, we will begin by looking at the DC motor as a generator, when driven externally by a hand-crank.

In a rotary DC generator, the wire coils turn on an axis, through a magnetic field, and produce a voltage:

$$V = Blv \quad (1)$$

where:

V = voltage generated

B = magnetic field strength

l = length of conductor crossing magnetic field

v = velocity at which conductor moves through the magnetic field.

The armature travels at constant angular velocity, and the voltage is proportional to velocity. This DC voltage drives a current though the resistance of the coils, in series with any external load (see **Figure 1**).

DC Motor

When the same device is driven by a DC voltage, as opposed to an external physical force, the generation of EMF is the same ($V = Blv$).

In the presence of a driving voltage, the EMF produced resists the input current, just as if it were a resistor (see **Figure 2**). For this reason, it is referred to as a "back" EMF, and can be modeled with a resistor. Consequently, the DC motor can be modeled by using two resistors in series, one being the resistance of the coils, and the other being the resistance of the Back EMF (see **Figure 3**).

The total resistance of the DC motor is equal to the internal resistance of the motor plus the effective resistance of the Back EMF generated as the motor turns. Back EMF manifests itself as, and can be expressed and measured as, a resistance. When this resistance is added to the motor's inherent resistance the sum equals the total resistance of the working motor.

$$R_1 + R_2 = R_t \quad (2)$$

where:

R_1 = motor internal resistance

R_2 = resistance of back EMF

R_t = total resistance to DC voltage source

First Phase Relationship: AC Linear Generator Creates Alternating EMF

Just as an external physical force (a hand-crank) can turn a rotary generator to produce a DC voltage, a loudspeaker driver can be driven by an external "wheel-and-stick" mechanism to produce an AC voltage. When we produce an AC voltage in this way, we can observe that the constantly changing position of the cone has a fixed 90° derivative relationship to its velocity (rate-of-change-of-position): velocity precedes position.

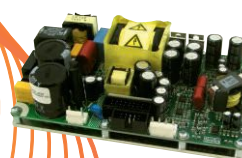
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NC250MP



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NC502MP

Add-on Module NC100HF
The NC100HF is a dedicated tweeter-amplifier which fits the NCxxxMP series.



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Qualitative Description

We drove our DC generator with a hand-crank. For the purpose of comparison, we will drive an AC generator using an alternating external source.

Figure 4 shows the AC generator being driven by a shaft connected to a wheel. One end of the shaft is fastened to a fixed point on the wheel, and the other end is attached to the moving assembly of the AC linear generator.

Figure 4: Here, the AC generator is driven by a wheel and a shaft (not drawn to scale).

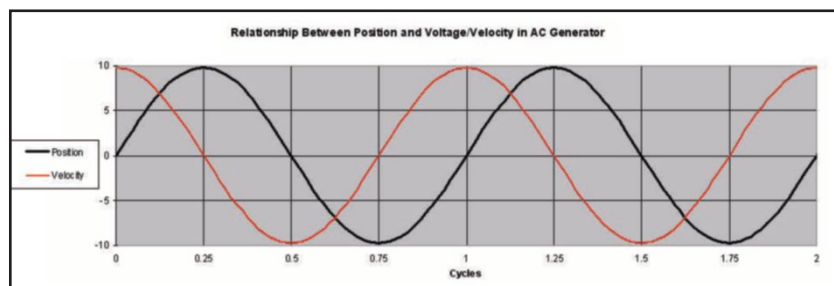
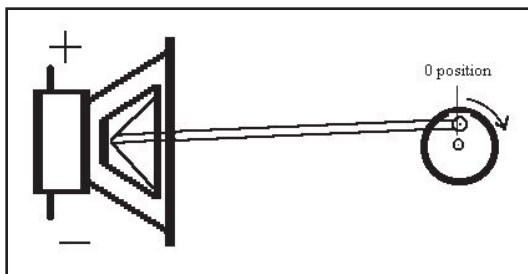


Figure 5: The graph shows the phase relationship between the cone position, the velocity, and the voltage.

Figure 6: In the moving-coil loudspeaker, the AC linear generator is driven by external voltage.

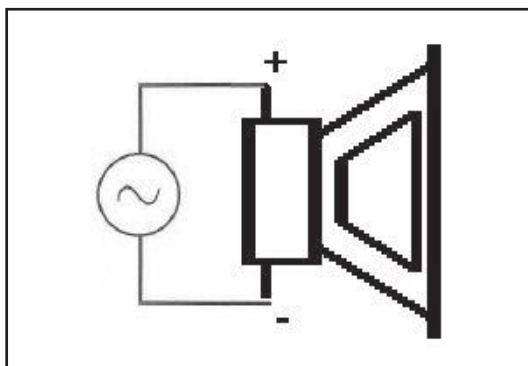
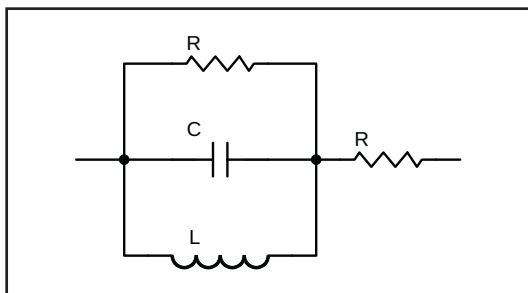


Figure 7: The LCR filter represents Back EMF and the resistor represents the voice coil DC resistance.



Phase Relationship in Linear AC Generator

Back EMF can be understood using differential and integral sinusoidal relationships. This will illustrate the first of them.

Note that the wheel driving the shaft (shown in Figure 4) has a designated zero position. This zero position corresponds to the moving assembly's point of zero excursion. Importantly, when the wheel is turning, this is also the moving assembly's point of maximum or minimum (negative) velocity, as the moving assembly travels back and forth. In other words, there is a 90° phase-shift between position and velocity. Velocity is in phase with the instantaneous voltage produced by the generator.

The position of the cone at any instant is a given, based on the position of the wheel driving the shaft. The first derivative of position is velocity, which is defined to be rate-of-change of position, and is represented by a cosine curve, as shown in **Figure 5**, and voltage generated at any instant is proportional to the cone's velocity. You can see that the point of zero position coincides with the point of maximum (absolute) velocity. We will see that the moving assembly of the voltage-driven loudspeaker driver has the same relationship of position to velocity.

AC Linear Motor: The Moving-Coil Loudspeaker

Just as our DC generator can become a motor when an external DC voltage is applied, our AC generator becomes an AC motor when exposed to an AC voltage source, such as an audio amplifier. We can use such a device to reproduce sound. We commonly refer to an AC motor used in this fashion as a moving-coil loudspeaker (see **Figure 6**).

And, as the Back EMF produced by a rotary DC motor can be modeled by using a resistor, the Back EMF produced by the AC loudspeaker driver can be modeled by using an LCR filter. The single resistor representing the voice coil's resistance is analogous to the resistance of the coils in the DC motor, and the Back EMF is represented by an inductor, a capacitor, and a resistor in parallel with one another, as shown in **Figure 7**.

In the region of its resonance, the impedance of the loudspeaker is the vector sum of the impedance of this LCR filter and the DC resistance of the voice coil. At any frequency, impedance can be described by its magnitude and phase angle. We will use a specific woofer when describing the behavior of Back EMF, the impedance characteristics of which are illustrated in **Figure 8**.

We use a specific real-world woofer to illustrate the concepts discussed in this paper. It is an 8" woofer from a Fisher system built in about 1965. It is a good quality driver. The characteristics of this woofer that are relevant to the article are shown in **Table 1**.

Calculate Magnitude and Phase of Back EMF from Impedance Measurements

We will calculate the impedance and phase angle of the Back EMF at three different frequencies: below resonance (f_1), at the resonant frequency (f_s), above resonance (f_2). The frequencies below and above resonance are the frequencies f_1 and f_2 , which are commonly used when measuring a driver to calculate its motor parameters, Q_{TS} , Q_{ES} , and Q_{MS} .^[1] The choice of these frequencies, while entirely arbitrary, is interesting because of the symmetrical nature of the impedances of at f_1 and f_2 , as will be seen.

Because these impedances are complex, having a magnitude and a phase angle, we will use complex numbers to calculate them, and vector diagrams to illustrate them.

Determination of Magnitude and Phase Angle of Back EMF Below Resonance: f_1

The frequency f_1 for this driver is 23 Hz. We will illustrate how to isolate Back EMF at this frequency (see **Figure 9**). As was the case with the DC motor, we will subtract the DC resistance of the wire windings from the total impedance, in order to solve for the impedance of the Back EMF:

Frequency: 23 Hz

The magnitude and phase angle of the driver's impedance at this frequency were measured using test equipment:

Impedance Z: 13.64 Ω
 Impedance Phase Angle (RAD): 1.07 or 61.3°

This complex impedance has both an inductive component (X_l), and a resistive c component (R_t).

$$X_l = 13.64 \sin(1.07) = 11.97 \, \Omega$$

$$R_t = 13.64 \cos(1.07) = 6.55 \, \Omega$$

Therefore, the complex impedance, can be expressed and modeled using these values:

Complex Impedance (ohms): 6.55 + 11.97j

Next, as we did with the DC motor, we subtract

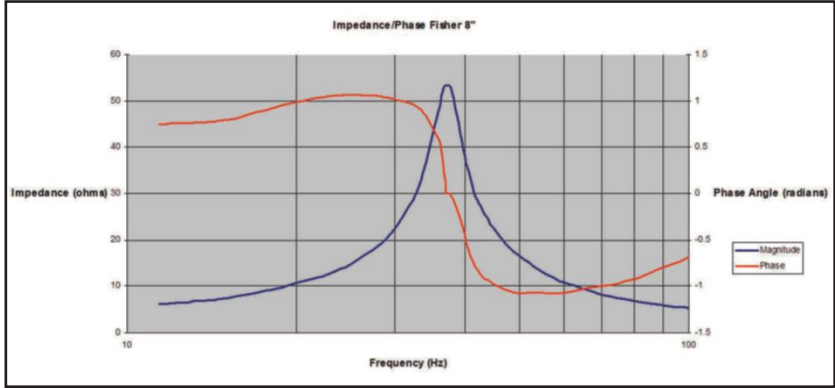


Figure 8: The impedance curve of the Fisher woofer shows the magnitude and the phase.

Fisher 8" Woofer Characteristics	
F_s	36.93 Hz
DCR	3.50 Ω
Z_{MAX}	53.50 Ω
f_1 ^[1]	23 Hz
Impedance at f_1	13.64 Ω , at 61.3° phase angle
f_2 ^[1]	59.30 Hz
Impedance at f_2	13.64 Ω , at -61.3° phase angle

Table 1: These are characteristics of the Fisher 8" woofer that are relevant for this article.

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the DC resistance of the voice coil to solve for Back EMF, this time using complex numbers:

Voice Coil DC Resistance R_{DC} : 3.5 Ω

The DC resistance is, by definition, non-reactive, and therefore, has a phase angle of zero.

Complex Expression of R_{DC} (ohms): 3.5 + 0j

So, magnitude and phase of Back EMF can be found by subtraction:

$$\text{Back EMF} = Z - R_{DC} = (6.55 + 11.97j) - (3.5 + 0j) = (3.05 + 11.97j)$$

Complex Expression of Back EMF Z_{EMF} (ohms): 3.05 + 11.97j

The Back EMF at this frequency has a resistance (R_{EMF}) of 3.05 Ω , and an inductive reactance (X_L) of 11.97 Ω . In order to determine the net value of the Back EMF, we use the Pythagorean Theorem:

$$Z_{EMF} = \sqrt{(R_{EMF}^2) + (X_L^2)}$$

Or, in this case,

$$\sqrt{(3.05^2) + (11.97^2)} = 12.35$$

Net Impedance of Back EMF: 12.35 Ω

The phase angle has a trigonometric relationship:

$$\text{Phase Angle} = \text{atan}(X_L/R_{EMF})$$

Or, in this case,

$$\text{Phase Angle} = \text{atan}(11.97/3.05)$$

Back EMF Phase Angle (RAD): 1.32 or 75.6°

Figure 9: The impedances of the resistor and the inductor model are shown at 23 Hz.

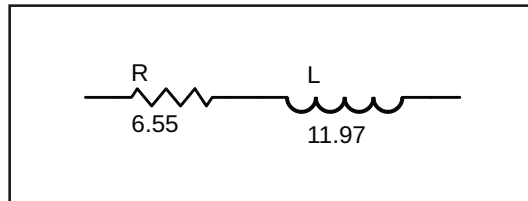


Figure 10: The vector diagram summarizes impedance at f_1 , below resonance.

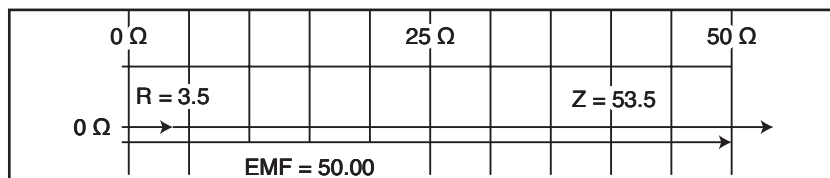
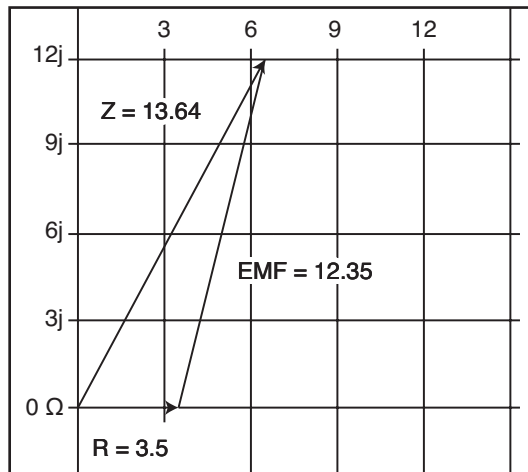


Figure 11: The vector diagram summarizes resistance impedance at Resonance (f_s): $R_{DC} + EMF = Z$

Figure 10 shows the vector diagram of the unit's impedance at f_1 , with R and EMF adding up to Z.

Determination of Magnitude and Phase Angle of Back EMF at Frequency of Resonance: f_s

The frequency of resonance (f_s) of our Fisher woofer is 36.93 Hz. At resonance, just as at f_1 , we can subtract the DC resistance of the voice-coil windings from the total impedance in order to solve for Back EMF.

As before, the impedance of the driver at this frequency was measured on the bench:

Impedance (Z): 53.50 Ω

Impedance Phase Angle: 0

Unlike the reactive impedance at f_1 , this impedance is purely resistive, and therefore, has a phase angle of zero. As with the rotary motor, we can describe the voice coil resistance as R_1 , the Back EMF as R_s , and the total impedance as (R_t):

$$R_1 + R_2 = R_t \quad (2)$$

Because the voice-coil resistance and the impedance of the Back EMF have the same phase angle (zero), the magnitude of the Back EMF can be found with a simple arithmetic subtraction, just as with the rotary motor:

$$R_{EMF} = R_t - R_1 \quad (3)$$

or,

$$R_{EMF} = 53.5 - 3.5 = 50 \Omega$$

The impedance of the Back EMF at 36.93 Hz is 50 Ω , purely resistive. **Figure 11** shows the vector diagram of the unit's impedance at resonance (f_s). Again, R and EMF add up to Z. At resonance, the impedance is purely resistive. This tells us that Back EMF is resistive. More on this later.

Determination of Magnitude and Phase Angle of Back EMF above Resonance: f_2

These calculations are a mirror-image of those performed at f_1 , below resonance. The form of the calculations is the same. For that reason, and for the sake of brevity, we will only provide the results and the vector diagram.

The frequency f_2 for this driver is 59.30 Hz. We will subtract the DC resistance of the voice coil windings from the total impedance, at this frequency, as we have done at the lower frequencies.

Back EMF Phase Angle (RAD): -1.32 or -75.6°

Figure 12 shows the vector diagram of the unit's impedance at f_2 , with R and EMF adding up to Z. Above resonance, the impedance is decreasing

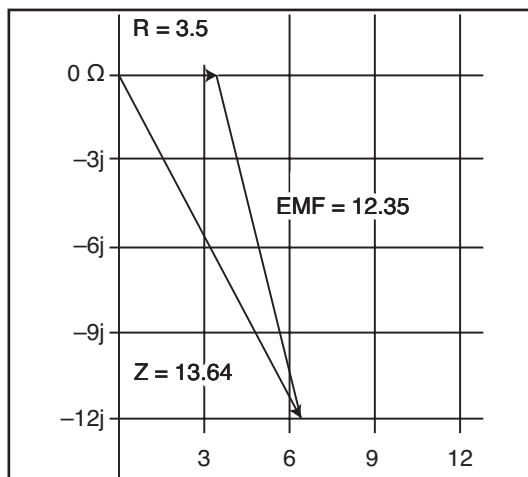


Figure 12: The vector diagram summarizes impedance at f_2 , above resonance.

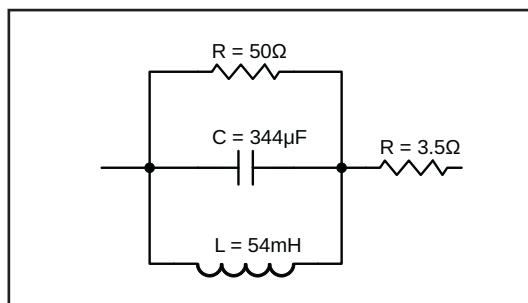


Figure 13: Here are equivalent real-world values of impedance for the Fisher woofer.

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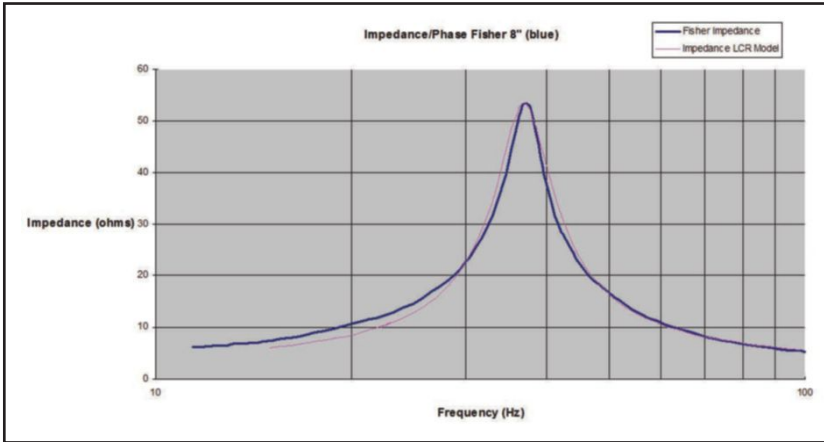


Figure 14: The graph shows the impedance curves of the Fisher woofer and the equivalent LCR filter.

with respect to frequency. This tells us that the impedance and the Back EMF are capacitive.

LCR Model of Fisher Woofer

In the section on the DC motor, we were able to model the resistance and Back EMF using resistors

in series. We were able to quantify the Back EMF by subtracting the coil resistance from the total resistance. Similarly, the Back EMF of our Fisher woofer can be modeled using specific components, as follows:

Effective Resistance of Back EMF: 50 Ω

Effective Capacitance of Back EMF: 344 μF

Effective Inductance of Back EMF: 54 mH

Figure 13 shows the LCR filter model, with the effective component values. This filter's impedance curve is comparable to that of the Fisher woofer. **Figure 14** shows the impedances both of the modeled filter and the actual woofer.

Intersection of Mass/Compliance and Back EMF as Modeled

As it happens,

$$LC = MC_{MS} \quad (4)$$

where,

L = inductor in LCR model

C = capacitor in LCR model

M = mass of loudspeaker moving assembly

C_{MS} = loudspeaker's compliance

Back EMF Voltage and Stall Ratio

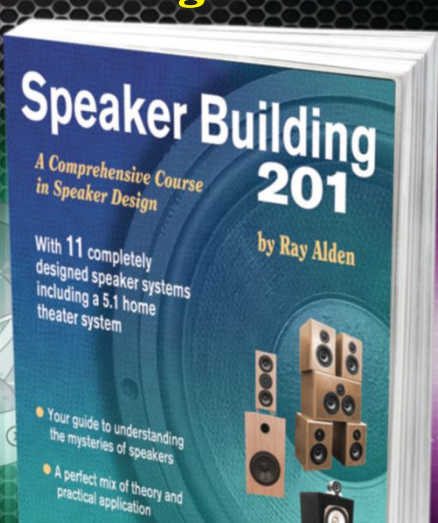
Because Back EMF manifests itself as impedance, we have been expressing its magnitude in ohms. Ultimately, however, Back EMF is a voltage, and it can be interesting to consider it as such. The ratio of the Back EMF to input voltage is equal to the ratio of Back EMF impedance to total impedance, just as with any impeding series components. This can give us an idea of the significance of Back EMF at a given frequency. First we will calculate this ratio at F_1 , 23 Hz, for an input voltage of 10 V.

- Net Loudspeaker Impedance: 13.64 Ω (see "Linear Motor")
- Current at 10 V: 7.33 A ($I = E/Z$)
- Back EMF Impedance: 12.35 Ω (see "Linear Motor")
- Back EMF Voltage: 9.05 V
- Ratio Back EMF to Input Voltage: 0.905

So, at this frequency, for a 10 V input, we produce a Back EMF of 9.05 V.

At resonance, 39.93 Hz, the current is diminished by a high resistive Back EMF, but the magnitude of

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the Back EMF is barely changed:

Impedance: 53.5 Ω

Current: 0.187 A ($I = E/Z$)

Back EMF Impedance: 50 Ω

Back EMF: 9.35 V

Ratio Back EMF to Input Voltage: 0.935

Interestingly, we see a change in Back EMF of roughly 3%. (9.05 V to 9.35 V). For a given input signal, the magnitude of the Back EMF is nearly unchanged with frequency, with this woofer. (These figures will be duplicated at f_2 , above resonance, where the numbers are the same, but the phase angles are mirror-image.)

Stall Ratio

Bob Carver and others have found it useful to express Back EMF in volts, and have defined a "stall ratio," which is seen as related to efficiency of energy conversion.^[2]

$$\text{Stall Ratio} = \frac{V_{EMF}}{V_I - V_{EMF}} \quad (5)$$

where,

V_{EMF} = voltage of Back EMF, and

V_I = applied input voltage

"Stall," would be a condition occurring when the driver operates with a stall ratio of less than 1, and the Back EMF is less than half the applied voltage. The stall ratio of this woofer at f_1 and f_2 is 9.53, and at f_s is 8.7. Both figures are well above Carver's benchmark figure of 1.

Next Month

The second part of this article series will define the delay in time between input current and moving-assembly position as "Slip," and examine how, because Slip increases with frequency, the resulting Back EMF manifests itself as it does in its three different modes of generation: below resonance, at resonance, and above resonance. 

About the Author

Andy Lewis lives in Englewood, CO, plays drums, and produces "Acme Sound Low-B" low-coloration loudspeakers for bass players. He is also involved in Special Olympics and other disability activities with his son, Collin.

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[1] R. Small, "Director Radiator Loudspeaker System Analysis," *Journal of the Audio Engineering Society (JAES)*, June 1972.

[2] I. Berger, "The Stall Ratio and Back EMF," embedded in "Sunfire True Subwoofer Mark II, D.B. Keele, *Audio*, November 1997.

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Acoustics in Performance

In this book, Richard Honeycutt details “all you need to know about good acoustics and sound systems in performance and worship spaces.” *audioXpress* and *Voice Coil* readers will be familiar with Richard A. Honeycutt’s writings, including *audioXpress*’ Hollow-State Electronics column, dedicated to exploring all tube-related electronics, and his Sound Control column, dedicated to everything in acoustics. Richard has also contributed to *Voice Coil* since 1994, so the author knows his subject matter.

Review By
J. Martins
(Editor-in-Chief)



Before we begin, I want to disclose the fact that I had access to an earlier version of Richard Honeycutt’s book, and that I actively engaged with him on the preparation of the final work, which I am happy to see finally published. Because of my involvement, I also feel the need to share that this book is essentially a small part of Richard’ Honeycutt’s work, because the book’s aim was essentially to provide an introduction to acoustics to anyone who needs to deal with sound and performances. *Acoustics in Performance* is a technical book written for everyone.

There are many “audio engineering” books available that quickly dive into the mathematics, and the schematics, and because of that are quickly placed on the shelf or put aside for future reference. There are also many works discussing acoustics, which in most cases look like Albert Einstein’s notebook and mostly discuss sound from all the

theoretical angles, except the ones that matter to all of us, which is: “given the understanding of sound and acoustical phenomena, what should we do (or not do) to deal with it in practical terms.”

Of course, acoustical consultants like to jump on this point and say, “just call me.” Or you can buy and read this book. Even for simple things such as knowing where to place the speakers in the room and obtain the best possible conditions to enjoy music listening to dealing with an orchestra playing in a auditorium, or knowing when and how to apply sound reinforcement because you have a few hundred people that need to understand what’s being said. That’s what *Acoustics in Performance* is about. Understanding the basics and what you need to know to deal with “acoustics and sound systems in performance and worship spaces!” as the subtitle of the book says.

As Honeycutt writes, “Everyone knows that the ability to hear music in balance and to understand speech is essential in any space used for performance or worship. A basic, non-engineering understanding of acoustics will help building owners, theater managers, ministers and teachers of music, performers, and other professionals to achieve their goals of excellent acoustics in venues with which they work. Performers having a basic understanding of acoustics will be able to make the most of the acoustics of the venue in which they perform. This

Acoustics in Performance
Richard A. Honeycutt
Publication Date:
February 2018
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Publisher: Elektor International Media BV
124 pages
Format: Soft cover

book helps those responsible for providing good acoustics in performance and worship spaces to understand the variables and choices entailed in proper acoustic design for performance and worship."

Knowing the author's level of technical knowledge, don't expect this book to be all "light reading." Honeycutt's knowledge is rock-solid and that's why he could structure the book based on the essential understanding of the basic science and "need to know." So, in the first two chapters "The Science of Acoustics" and "Physical Acoustics of Rooms," most students will find excellent material for study, presented in a way that will always provide an excellent quick reference guide—from physics to psychoacoustics to road-ready rock 'n roll basics. Again, understanding sound and acoustics first and then doing something about it.

And "doing something about it," is what Honeycutt discusses in the next three chapters of the book, "Attaining the Goal," "Controlled Acoustics," and "Working with Your Room." He also provides excellent overviews of the basics, and advice about the things everyone should know and always remember. When I read this work the first time, I

was reminded of the time when I started playing in a band and had to perform live for the first time. Even after too many years, chills go down my spine when I remember the awful mistakes and the stupid things we could have avoided. This book was written so you can start with confidence and avoid the "wish I had known that before" experience.

Yes, I think Honeycutt wrote this book thinking about a few young and inexperienced people he met doing sound, the minister that complained about his worship space, or the people that hire—or should have hired—an acoustical consultant because they "have a problem." But I think that all *audioXpress* readers will enjoy and learn something with his book—and they will know to whom they should also gift this book.

Just remember, Richard A. Honeycutt earned a B.S. in Physics, and holds a Ph.D. in Electroacoustics, serves as chair of the North Carolina regional Chapter of the Acoustical Society of America (ASA), and owns his own acoustical consulting business (meaning, he is normally paid to explain to clients the things that he explains in this book... don't tell anyone!) 

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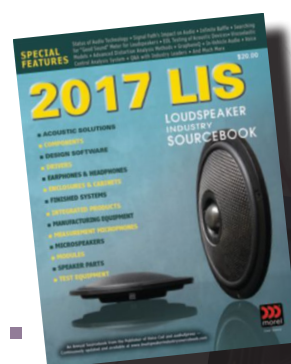
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Schiit Audio's High-Value Ideas

Unless you've been away from hi-fi for the past few years, you've probably heard of Schiit Audio, a high-value brand available at reasonable prices. Over the years, Schiit Audio owner Jason Stoddard and I have talked at trade shows and I recently caught up with Jason to talk about his company.

By
Oliver A. Masciarotte
(United States)

Jason Stoddard (pictured) started Schiit Audio in 2010 when he and audio industry veteran Mike Moffat decided it was time to shake things up a bit.

Oliver Masciarotte: Can you give us a brief intro to you and the company Schiit Audio (www.schiit.com)?

Jason Stoddard: Hey all, maybe a few words to set the scene might be helpful. About eight years ago, I started Schiit Audio with Mike Moffat [founder of Theta Digital and a pioneer in digital audio], initially in one third of our garage. Our original products were a pair of headphone amps—one tube, one solid state. Since then, we've been expanding the lineup, down, and sideways—with less expensive products, more expensive products, DACs, and recently, a pair of preamps and a speaker power amp. Today, our line spans products from \$49 to \$2,399, and our garage has grown to 15,000 square feet of industrial space in Valencia, CA. I do most of the analog design, and Mike does the digital.

My history with audio starts in college, where I made loudspeakers and eventually had a small speaker company, Odeon, which didn't go very far. From there, I was at Sumo in the Polaris 2/ Andromeda 3/Diana/Theorem/Artemis era, if anyone remembers that. Then, almost two decades as an evil marketer with my own ad agency (Centric/Agency of Change) gets us to the Schiit era.

Masciarotte: What is Schiit's engineering philosophy?

Stoddard: Oh boy. How do I put this? "Stubborn pragmatists?" "Flexible idealists?" Because, you know, we focus on low-cost products...but stick mainly to fully discrete designs on the analog side, and mainly to multibit DACs with our own digital filters on an Analog Devices SHARC DSP. In general, I'd say it goes something like this: "Deliver the highest value you can on the board, without putting too much cost in the box." We all know about high-end products that are absolutely beautiful, flawlessly machined out of a single block of aluminum...and yet have extreme compromises in terms of their implementation, using consumer spec volume control chips, mid-line op-amps, or low-performance Class D amp modules.

Yeah, I know; lots of opinions here, which is why I mention low performance. Sure, there are good volume control chips, good op-amps, and good Class D implementations out there. There are also plenty of not-so-hot designs.

Masciarotte: Tell us about epic development hurdles. What has been some particular pain points? Why and how did they get solved?

Stoddard: A lot of our epic fails...er, I mean hurdles... come about when we crash up against components that actually have significant, and unavoidable, cost. Large transformers are expensive. Heavy heatsinks are expensive. The AD5791 DACs we use in Yggdrasil are expensive. And, unlike fancy chassis or boutique capacitors, we can't simply decide not to use these parts. So the costs are unavoidable.

When we know we're looking at an unavoidable cost, a lot of times we'll start by trying to sidestep it. With Vidar, our 100 W [into 8 Ω] stereo power amp, we knew we'd be staring at least two of these costs—a big transformer and lots of heatsinking. At first, it seemed like it made sense to start with low-cost heatsinks in a forced convection heat tunnel. But heat tunnel costs start to add up fast, too. The extrusion cost may be low, but what about a truly quiet fan? What about the additional chassis bits to keep the air flowing where it should? What about the cost of gaskets that will last much longer than a five-year warranty period?

Yeah. We built a heat-tunnel design, tweaked it until it worked...then costed it out and threw it away. We then tried to do a passive heat chimney design with high efficiency heatsinks, but again, reality intruded...we would have needed a "panic fan" to meet the FTC 1/8 power testing requirement. That one went in the garbage as well. We ended up with a dead-simple design with lots of exposed heatsinking, because once you factored in the additional costs of fans and sensors and chassis parts, the large exposed heatsink was the way to go. It also had the advantage of not irritating customers with fan noise. It should have been an obvious solution.

Sometimes looking at things sideways works, though. In the case of our upcoming turntable, we were stuck on cost, no matter how we looked at production. Machined plinth and platter? Oh holy moly, no wonder so many turntables are \$2,500 and up. MDF plinth? Well, with a skeletal design, you're looking at tons of waste, and also the need for custom 2.5" thick MDF. Again, super pricey. Delrin platter? Oh boy, those cost like a Magni 3 [a \$99 headphone amplifier, explained later in the article]. So what we did was go sideways—to aluminum die casting. Die casting means a complicated plinth is easy and inexpensive, and the platter can be formed with most of the mass on the edge, for better stability. Die casting means we'll be able to do a turntable with amazing specs and capabilities, for not much more than an entry-level rig. But it also means big investments in tooling, and a big production run.

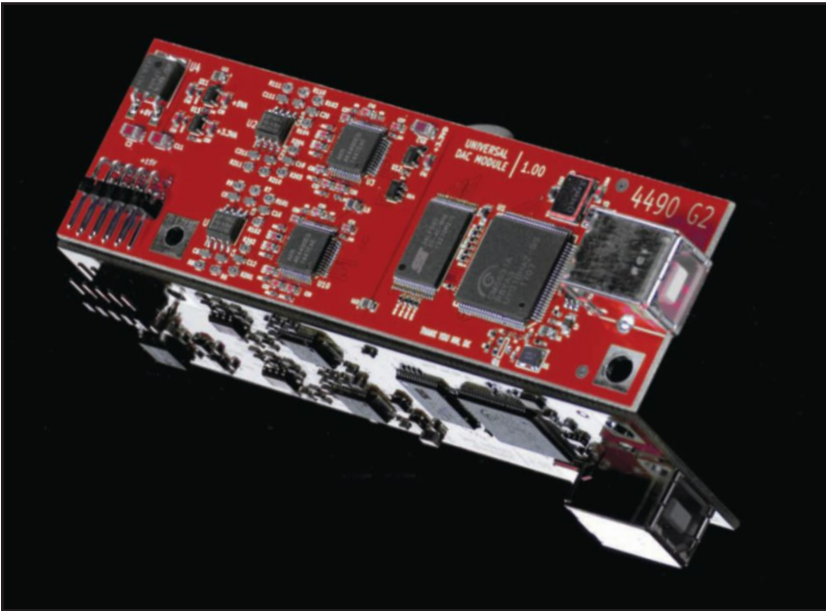
Masciarotte: Unique design choices—what differentiates you guys from the also rans?



Saga is a passive or active preamp with a relay-switched stepped attenuator volume control for perfect channel matching, unmeasurable distortion that also has an unique active stage that uses a hybrid tube buffer, and includes a remote control.



To improve on Lyr 2, Schiit Audio had to completely reinvent the amp, introduce two new technologies, and make it modular. Not only is the new Lyr 3 a big sonic upgrade, it can be configured with the DAC or phono preamp of the user's choice.



Lyr 3 also takes an internal card, which can be a AK4490 DAC card (pictured), phono preamp card, or Multibit DAC card—a modular design to reduce obsolescence.

Stoddard: This may sound funny, but I think it's in focusing on "no compromise" features, together with some unique solutions. Bear with me, I know "no compromise" is a loaded phrase. By "no compromise," I mean, "discrete design, tubes (done right, not starved-plate-just-for-cosmetics), linear power supplies, relay-stepped attenuation, things you expect to see on much higher ticket items." Consider Saga, a remote-controlled, relay-switched tube preamp at \$349, or Magni 3, a \$99 fully discrete current-feedback topology headphone amp and preamp, as examples.

We also focus on unique solutions, where they make sense. Our Lyr 3 modular headphone amp/DAC/preamp introduces our first constant-transconductance BJT (Bipolar Junction Transistor) output stage, using both PNP and NPN devices on both sides of the rail (but not with gain). It's not an entirely new idea—thank Bob Cordell to opening our eyes to the need, and John Broskie for proposing many different ways to achieve the same objective. In this case, a constant transconductance stage makes sense, because it allows us to run at less than full-out

Class A bias, and still have great performance. Lyr 3 also refines our compound tube-BJT hybrid topology that uses a current mirror and small anode resistor to enable the tube to use virtually all the B+ voltage on the plate and return feedback to the cathode and, with an operating point servo, to eliminate interstage input and output coupling capacitors. Lyr 3 also takes an internal card, which can be a AK4490 DAC card, phono preamp card, or Multibit DAC card—a modular design to reduce obsolescence. I could go on—Pivot Point, the whole Multibit DAC thing using medical-grade ICs that nobody else is using, like the AD5791, the 5781, and the 5547, the time and frequency-optimized digital filter, etc., but I think you get the picture.

Masciarotte: Cost-effective as opposed to state-of-the-art. What compromises must your company make?

Stoddard: I'm tempted to say, "It depends. If you think \$160 capacitors and \$50 fuses make a difference, then we have to make lots of compromises." The real answer (at least in my mind): not many, other than designing for large runs and not getting embroiled in software we don't control.

Since we do larger runs, up to 2,500 to 5,000 units at a time, maybe not big in the larger world, but pretty big in audio, we need to pay extensive attention to production engineering. I mentioned the upcoming Sol turntable and die casting, but it goes far beyond that. Stuff has to go together fast if we're making it in the US. That's why you won't see lots of wires or complicated hand assembly in our products, and it's why we're now tooling up to replace all of our leaded LEDs with custom light pipes, and all of our screw-on knobs with push-on knobs. Sounds basic? It'll save thousands of hours over the course of a year.

And software we don't control? Yeah. That's why you won't be seeing us do DAPs or streamers. We don't control all the software or protocols. We don't control when we have to update our apps. That's a level of complexity we want to stay away from. We already have 32 different firmware versions we use across our product line, and it's challenging enough keeping them all straight. If we decide to become a de facto software company, that's a big risk — one that might be a very costly distraction.

About the Author

Oliver A. Masciarotte has spent more 30 years immersed in the tech space, working on facilitation, optimization, marketing, and product development for clients worldwide. As an author and speaker, he enjoys informing folks about technological best practices. More information is available at seneschal.net and othermunday.com.

Masciarotte: Since both hollow and solid state can be used to create anything from ultra-wideband, ultralow distortion gear to narrowband, hyper-colored "vibey" products, what do you consider to be the merits of each?



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The as-yet unreleased Sol turntable will be made from aluminum die casting, which means Schiit Audio will be able to produce a turntable with amazing specs and capabilities, for not much more than an entry-level rig.

Stoddard: We're more on the accurate/low-distortion side of the fence for both tubes and transistors, though we're never going to win any awards for "lowest distortion of all time." Consider that our all-tube headphone amp, Valhalla 2, isn't considered particularly "tubey" or "lush", and you'll see where we're coming from. We do aim for a more forgiving character at lower price points, though. Vali 2, a \$149 tube hybrid, is arguably our most "tubey" sounding product. Magni 3 was tweaked, in terms of operational point, until we considered it to be very slightly "warm."

Honestly, I'm most interested in hybrids, though some people consider them to be the worst of both worlds. With a hybrid, you can deliver the best qualities of a solid-state output stage, low output impedance and high output current, together with the best qualities of tube gain, like huge voltage swing capability and excellent linearity (in tubes like the 6SN7 and, to a lesser extent, the 6DJ8). The problem is the interface, of course, and that's where we're heading away from more conventional resistor-loaded tube designs and moving towards a current mirror approach. Of course, then you have both tube and transistor gain in the picture, so we can argue about the merits of that as well.

Masciarotte: What about op-amps? Do you use them and, if so, do you opt for discreet or do you go with what's appropriate and/or cost effective in a particular stage or price point?

Stoddard: We're not adamantly opposed to op-amps, but I do like the control and flexibility of discrete designs. Especially if you're talking current-feedback or balanced differential designs. Yes, I know there are CFA op-amps and differential op amps, but I like to be able to tailor the open-loop characteristics of the design.

That said, I don't think using op amps is a bad design choice. I'm very happy with our Fulla 2, which uses a modestly-priced OPA1662 for gain and paralleled LMH6643s for output buffering. Paralleled OPA1688s are also very nice for outputs, and quieter than the 6643. We also use some very high-spec op amps in our products that use multiplying Multibit DACs, like the AD5547. There's no way a discrete design will fit in that application.

Masciarotte: Your thoughts on balanced versus unbalanced I/O... Are balanced lines beneficial given the short cable runs normally encountered in a home environment? Are there some power amps that benefit from being driven by a balanced source?

Stoddard: I'd be happiest if balanced simply went away, because it's a source of much confusion amongst the customers. In the absence of that, I'd prefer that anything using balanced connectors really is balanced, rather than simply using the connectors, which is what we do. If it has XLRs, it's balanced, differential, and maintained all the way throughout.

Beyond that...(sigh). I think there's far too much focus on "balanced" versus "unbalanced." It seems that many people think audio nirvana is as close as rooting out all the unbalanced gear in their system. But it's not that simple. I've achieved great sound with both balanced and unbalanced systems. So I'm agnostic.

As far as some amps doing better from balanced sources, sure. Well, no. Maybe. If it's coming from a real differential source (hardware balanced DAC with one channel per phase, for example) and it's a real differential amplifier, then yes. But such amps are few and far between. FWIW, Yggdrasil, and Gungrir are hardware-balanced DACs, and Jotunheim, Mjolnir, and Ragnarok are balanced, differential amps.

Masciarotte: How about balanced versus unbalanced internals?

Stoddard: WeeeeIIIIII...if you're good with the additional complexity of balanced, and can take the higher thermal noise floor (power supply gunk might cancel, but not thermal noise), then personally I like the potential of balanced. But I can't provide hard evidence that it's "better" than single-ended.

Again, though...if it's not balanced internally, it shouldn't use balanced connectors. At least in my opinion. Yes, I know, I'm stubborn. And I have a crazy Loki Maxi equalizer design that would take 120 relays as balanced versus 60 relays as single-ended. So there are points at which you have to bend the rules. Maybe.

Masciarotte: Can you share your thoughts on switched mode versus linear power supplies?

Stoddard: It's simple: I'd never use a switcher if I didn't have to. Again, like my opinion on op-amps, this isn't due to any great animosity, and I don't think using a switcher is universally "bad." And yes, I know, switchers are getting better, and can beat linear supplies in some metrics...but switchers mean FCC certification. And I'd rather avoid having more products that need more certifications. And, when you get right down to it, [you] don't want 250 kHz to 3 MHz noise running around everywhere. Yes, you can filter it. Yes, you have PSRR (Power Supply Rejection Ratio). Yes, I know. And yes, I may change my mind in the future.

As of right now, we have exactly one product using a switching supply: the Fulla 2, where getting a ± 5 V rail off of a USB power input, with good efficiency and current output, means a switching rail generator, not a linear supply.

Masciarotte: What can you share that will help our readership better understand the role of the manufacturer, and what aspect of audio requires more attention from the consumer than it's getting?

Stoddard: If I could say one thing, it's this: chill out. Sometimes it seems there's lots of argument about what's "best" and "right," but it comes down to this: if it doesn't work for you, it doesn't work for you. If you think that everything should run switchers and op-amps are the best way to audio nirvana, do it! If you think that absolutely everything should be balanced from start to finish, have at it! Share what you learned, absolutely, but always be willing to learn. And always, always, take time to sit back, listen, and enjoy the music. 🎸

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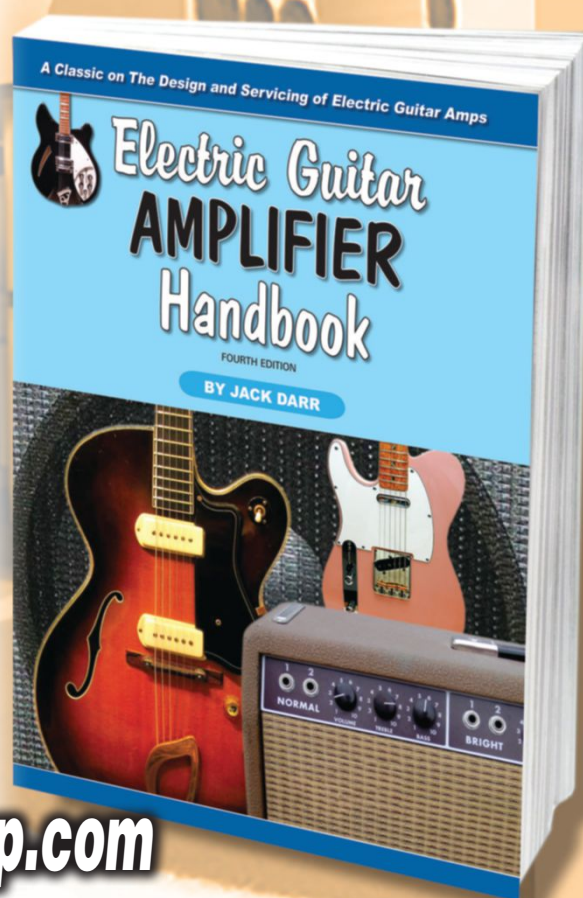
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Transformers for Tube Amplifiers (Part 1)

Transformer Basics

The most important component of a tube amplifier is the output transformer. An amplifier system's quality depends on that component.

This article series explains the fundamental properties of transformers, examines the important criteria for rating and building a transformer, and analyzes how the transformer circuitry affects the overall result. This article also discusses the necessary features of transformers, which are important for selection and practical use.



By

Gerhard Haas

(Germany)

Basically, a transformer has the task of translating currents and voltages in a way to meet desired requirements. A mains transformer is nothing more than a transformer that is specialized in 50 Hz or 60 Hz and a mains voltage of 230 V, 115 V, or the like. For example, tube filaments require the low voltage of 6.3 V at a high current, in the case of anode voltages, it is the contrary. The tube output transformer must convert high voltages at low current into high currents at low voltages to operate loudspeakers with an impedance of 4 Ω or 8 Ω . In contrast to a mains transformer which only needs

to master just one frequency, the low-frequency transformer must cover a spectrum ranging from 20 Hz up to 20 kHz. Unlike the mains transformer, the engineering effort is clearly greater.

The Operating Principle of Low-Frequency Transformers

To understand the operating principle of a low-frequency transformer, its fundamental properties need to be considered. That can be done using some equivalent circuit diagrams, formulas, and the related explanations. In principle, voltages and currents are translated in the desired ratio, as it is done with the main transformer. **Figure 1** shows the schematic diagram of the transformer. The formulas specified in Figure 1 show how voltages, currents, and resistances interrelate in ideal conditions. The reason why the resistances change in square according to the transformation ratio can be clearly explained using an example. If

Author's Note: All physical units specified in this article refer to the international SI system, the basic specifics of which are meters, kilograms, seconds, amperes, and the units derived from them or parts or multiples of them respectively. This article does not represent any kind of building instructions for transformers, that would go way beyond the given scope.

a voltage source has 100 V and a load resistance of 1 k Ω is connected, a current of 0.1 A flows. The transformed power output is 10 W. If the voltage is decreased to 10 V with a ratio of 10:1, a current of 1 A will flow with the same power output. This comprises a resistance of 10 Ω . The ratio between the currents and voltages is 1:10, the resistances show a ratio of 1:100, therefore n^2 . This calculation can be easily understood using Ohm's law.

This explanation applies to the ideal transformer. In reality, one needs to consider some disturbing aspects, which are impossible to avoid.

Figure 2 shows the equivalent circuit diagram of the real transformer.

Imagine an ideal transformer Tr , which is to be wired. U_S represents the signal source, which has an internal resistance R_1 . R_2 is the secondary load. These are the same ratios as specified in Figure 1. Now, the inevitable copper resistance R_{CuTot} is added summarizing the primary winding resistance and the copper resistance of the secondary winding, translated on the primary side. Ohmic losses occur here. At the copper resistance, electrical power is simply transformed into heat.

What is even more important for high-quality low-frequency transformation are the components such as the primary inductance L_1 , the stray inductance σL_1 , where σ (sigma) specifies the stray coefficient, and the capacitance C_{Tot} . C_{Tot} contains the layer capacitances of the windings and the wiring capacitances. All three components affect the frequency response to a greater or a lesser extent. The equivalent circuit diagrams in **Figure 3a** show which component has what influence at low frequencies. **Figure 3b** shows that information for high frequencies. For convenience only, n and R_2 are left out here. Also, R_{CuTot} has not been considered as the copper resistances of the primary and secondary winding are relatively small in contrast to R_1 if the transformer is rated properly.

For low frequencies, σL_1 doesn't have any influence because the stray inductance is very low in relation to the primary inductance. The primary inductance, the winding capacitance, and the wiring capacitance together form a parallel resonance circuit, the resonance frequency f of which is specified by the circular frequency $\omega T = 2 \times \pi \times f$. When looking at the resonance frequency, the impedance of the parallel resonance circuit is theoretically almost infinite. In practice, this resonance point appears, if at all, as small waviness at best, as it will be heavily dampened by the losses occurring in the transformer as well as by the source resistance R_1 .

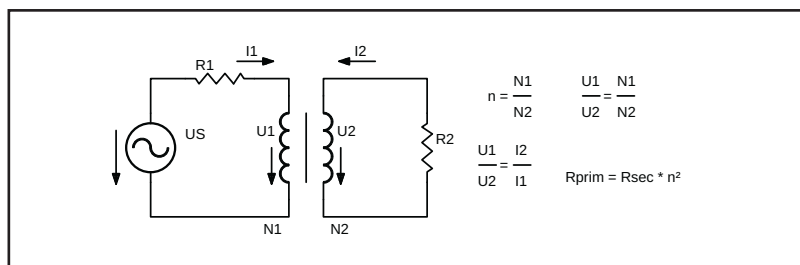


Figure 1: To understand the operating principle of a low-frequency transformer, its fundamental properties need to be considered. This is a schematic of a transformer.

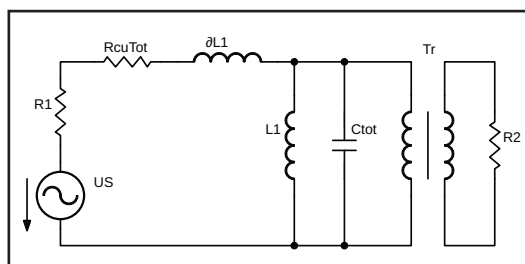


Figure 2: Here is the equivalent circuit diagram of the real transformer.

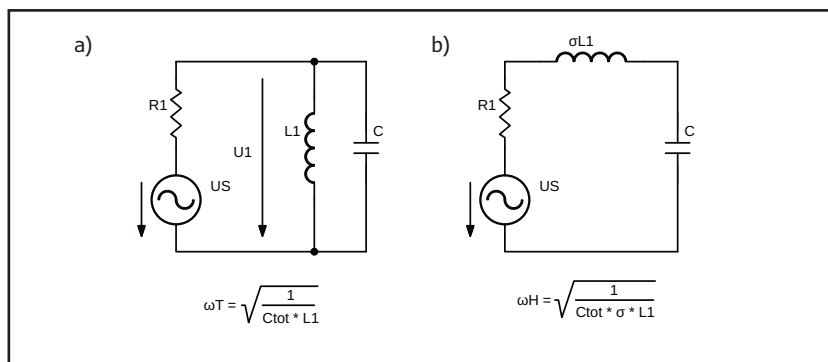


Figure 3: The equivalent circuit diagrams show when which component has what influence at low frequencies (a) and that same information for high frequencies (b).

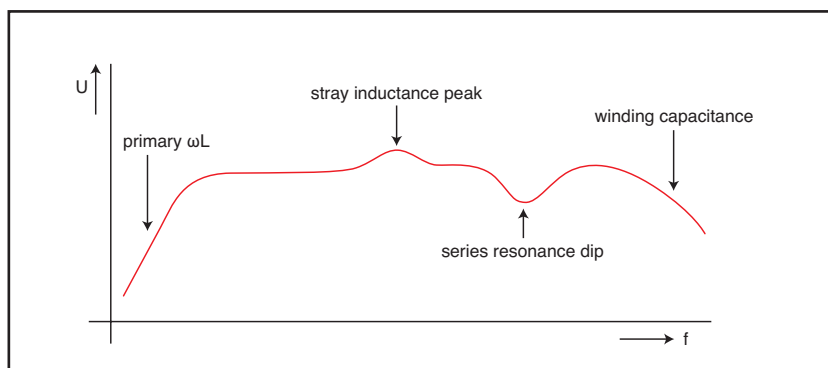


Figure 4: It is possible that the frequency response will increase above the resonance frequency of the absorption circuit to eventually drop because of the influence of the winding and wiring capacitances. This graph shows the general frequency.

Figure 5: For the lowest frequencies, the stray inductance, and the winding capacitance can be disregarded because they don't have any influence since they are small.

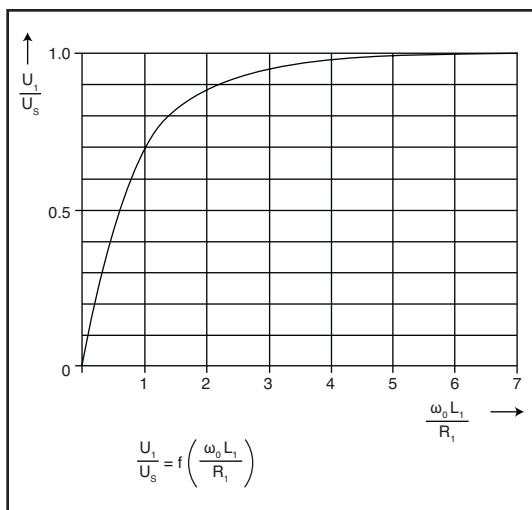


Figure 6: The transformer capacitances C_{Tot} from Figure 3 are justifiably split up into C_L and C_W .

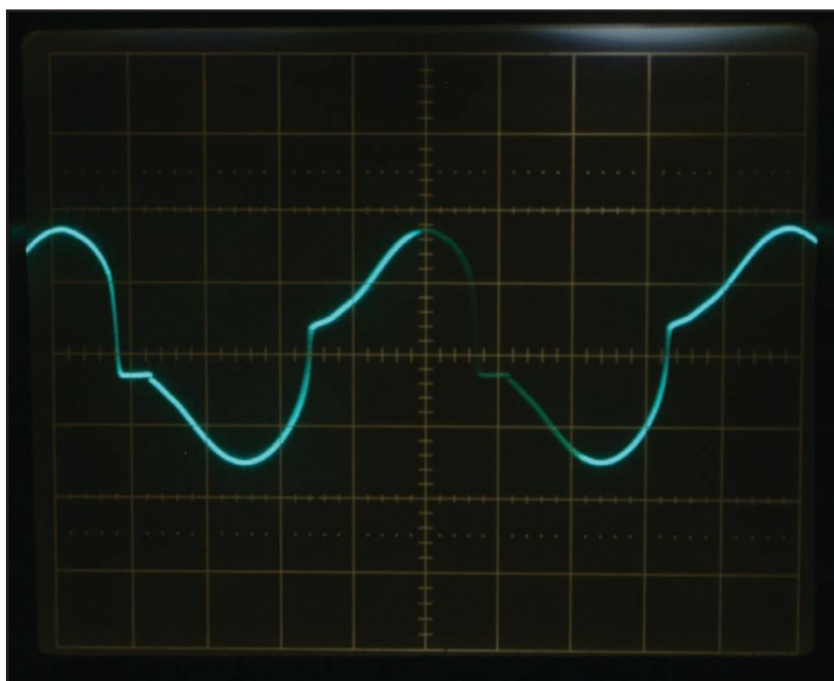
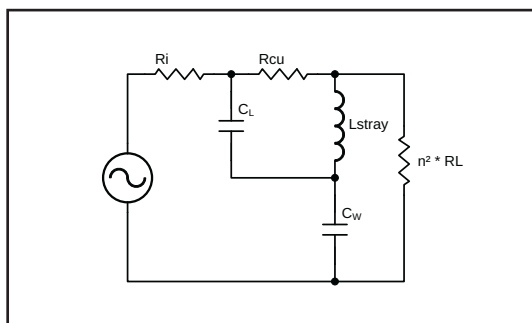


Figure 7: This image clearly shows what the signal looks like with a saturated iron core.

For high frequencies, the impedance of the primary inductance is infinite and, therefore, ineffective. The leakage inductance σL_1 and the wiring capacitance C_{Tot} from Figure 2 form a series resonance circuit with the resonance frequency ωH , as shown in Figure 3b. At this resonance point, the transformer's alternating current resistance approaches zero. In this way, theoretically, the current becomes infinite. However, this resonance point is also dampened most of all by the internal resistance R_1 of the signal source, by the transformer losses (copper resistance and windings), which are not specified here, and by the terminating resistance R_2 . If R_1 is very low, the resonance point will be much larger than expected with the transformation ratio. It is possible that the frequency response will increase above the resonance frequency of the absorption circuit to eventually drop because of the influence of the winding and wiring capacitances.

Figure 4 shows the general frequency.

At the lower end of the frequency range, there is also a clear drop. The reason for that is explained in **Figure 5**. In case of the lowest frequencies, the stray inductance and the winding capacitance can be disregarded because they don't have any influence since they are small. The primary inductance L_1 and the source resistance R_1 form a frequency dependent voltage divider. In Figure 5, the same designations are used as in Figure 3a. The source voltage U_s is divided by the voltage divider consisting of the ohmic resistance R_1 and the frequency dependent impedance L_1 . The factor $\omega_0 \times L_1$ will decrease as the frequency drops. The lower the frequency, the higher the voltage drop at R_1 and the lower the voltage at L_1 . The ratio between U_1 and U_s is a frequency dependent function of the ratio $\omega_0 \times L_1$ to R_1 . In the case of the normalized frequency in Figure 5, the voltage drop at 1 (x-axis) is 3 dB (≈ 0.7).

The Operating Principle of High-Frequency Transformers

To explain how a transformer behaves with high frequencies, **Figure 6** shows an equivalent circuit diagram that slightly differs from the standard representation. With low frequencies, the only effective components would be the voltage divider ratio of the internal resistance R_i (referred to as R_1 in the previous figures) of the signal source, of the copper resistance R_{cu} and of the load resistance R_L transformed using the transformation ratio n . In the case of low and medium frequencies, the impedances of the coil capacitance C_L and the winding capacitance C_W are so high that they don't have any influence here. Even the low impedance of the stray inductance L_{STRAY} is of little importance,



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that's also because C_W has a very high impedance at low frequencies. The transformer capacitances C_{Tot} from Figure 3 are justifiably split up into C_L and C_W as shown in Figure 6.

When buying a high-frequency choke for suppression, the typical values for the inductance, the maximum current carrying capacity, the ohmic resistance, but also the resonance frequency

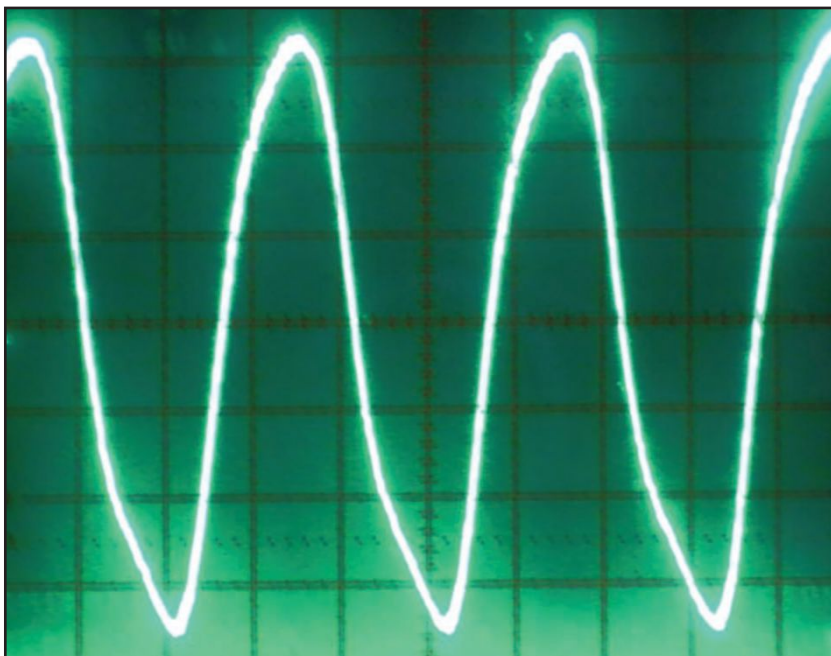


Figure 8: This graph shows what the signal could look like with high frequencies.

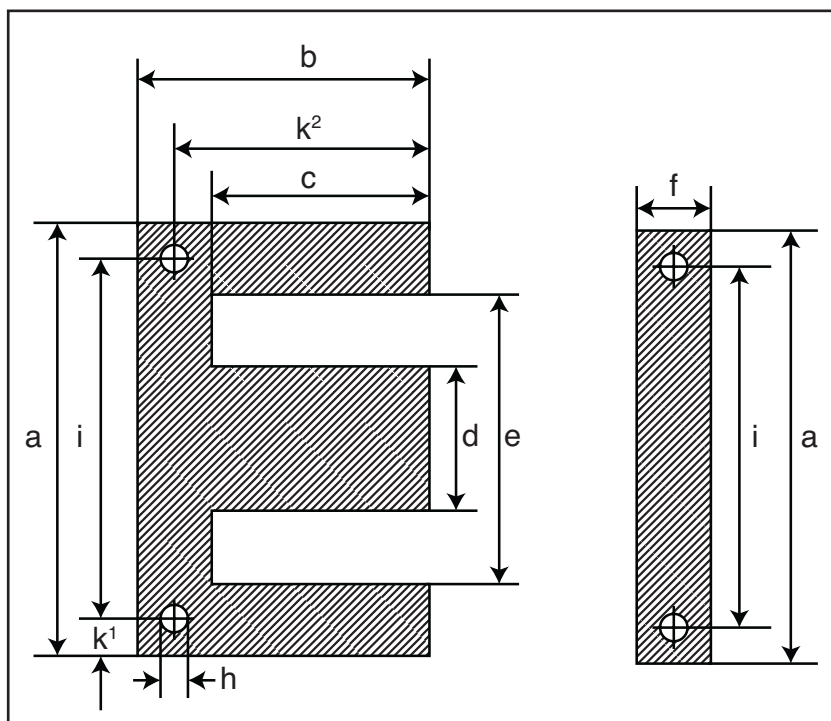


Figure 9: This drawing shows a waste-free normal EI-type cut.

are specified. The capacitance of the coil and the inductance form a resonant circuit that is dampened by the copper resistance. Figure 6 shows the capacitance C_L of the coil itself. The value C_W contains the layer capacitances and all other capacitances and C_W is higher than C_L . As the frequency increases, the series resonance circuit, which is formed by L_{STRAY} and C_W and attenuated by R_i , R_{Cu} , and the load resistance transformed by n will set the tone. The frequency response shows a clear drop.

For higher frequencies, the level increases as in this case L_{STRAY} and C_L form an attenuated parallel resonance circuit. With even higher frequencies, the impedance of L_{STRAY} will increase resulting in its loss of influence. However, C_L and C_W represent an increasing load reducing the output level. It needs to be considered that in Figure 6, the components C_L , C_W , R_{Cu} , and L_{STRAY} are cumulated in a simple manner. Actually, it is all about the subcomponents, which are spread all over the entire winding and actually require a far more complex equivalent circuit diagram. For the observations made here and for a better understanding, the simple illustration is sufficient.

Now, we want to clarify all this information using an example. When measuring an output transformer (e.g., at 1 dB under full modulation and with a properly connected load), its typical frequency range can be from 20 Hz (-0.5 dB) to 60 kHz (-1 dB). Often, a well-dimensioned transformer holds almost the full level up to the lowest limit frequency. Its iron core goes into saturated state, which can be noticed by an increasing signal distortion. **Figure 7** clearly shows what the signal looks like with a saturated iron core. In the case of even lower frequencies, the level can drop as I already explained. If significantly decreasing the output level, the frequency can range from approximately 15 Hz (-1 dB) to 80 kHz (-1 dB). This is not a wondrous multiplication but solid physics.

For high frequency, the factor $C \times U$ has its impact. Since the winding and wiring capacitances can be assumed to be constant, the voltage present at the capacitor has a significant influence. The formula $Q = C \times U$ defines the charge of the capacitor. The higher the voltage is at fixed capacitance, the more charge has to be moved. The higher the frequency, the faster the capacitance needs to be charged. As the frequency increases, the charging time constant $\tau = R \times C$ comes into play. This can be seen using the oscilloscope. With a sine signal, more or less exponential charge and discharge curves of a R_C element become visible. The phase shifts and inductances also have their impact, they cause further signal deformations.

Figure 8 shows what the signal could look like with high frequencies. A tube layout does not give much possibility of improvement. The internal resistance of the signal source can't be arbitrarily reduced, but is determined by the tube line-up and the circuit design. The transformer's characteristics are also fixed. Minor influences within narrow bounds can be exerted by setting the operating point and the idle current. The feedback factor also has an influence. As the feedback factor increases, the signal distortion is reduced, which is at the expense of other amplifier characteristics. The higher the feedback, the more tube amplifiers sound like transistor amplifiers.

These considerations explain why the transmittable frequency range becomes larger at a lower signal level. At the lower band end, the iron core of the transformer is not driven into saturation, at the upper end a lower charge is required for shifting the entire capacitances. When you listen to music at lower or medium volume levels, the entire bandwidth is available. For high modulation, the frequency range is limited.

Creating a Square Wave Signal

At this point, another important factor must be considered. Amplifiers are often checked for how they deal with square wave signals. Before going into detail, one needs to realize how a square wave signal is created. Every periodic oscillation can be described with a Fourier sequence. For a symmetrical square wave signal,

$$\frac{4 \times A}{\pi} \left(\cos x - \frac{1}{3} \cos 3x + \frac{1}{5} \cos 5x - \frac{1}{7} \cos 7x + \dots \right)$$

shall apply.

In this formula A refers to the actual amplitude value and x refers to the frequency. As you can see, only odd harmonics with their amplitudes dropping according to their order are added and subtracted from the fundamental $\cos x$.

Harmonics up to the 15th order are necessary for a good square wave. So, if a signal with 1 kHz is applied to an amplifier that features a sufficient bandwidth, the square wave signal will be transmitted cleanly. If a digital system features a bandwidth of up to about 22 kHz, as it is the case with CDs, a square wave signal with 1.5 kHz can barely be transmitted cleanly.

When checking an amplifier together with its output transformer at 20 kHz using a square wave signal, its bandwidth should be 300 kHz. If power amplifiers feature such bandwidths, they are vulnerable to disturbances caused by radio stations

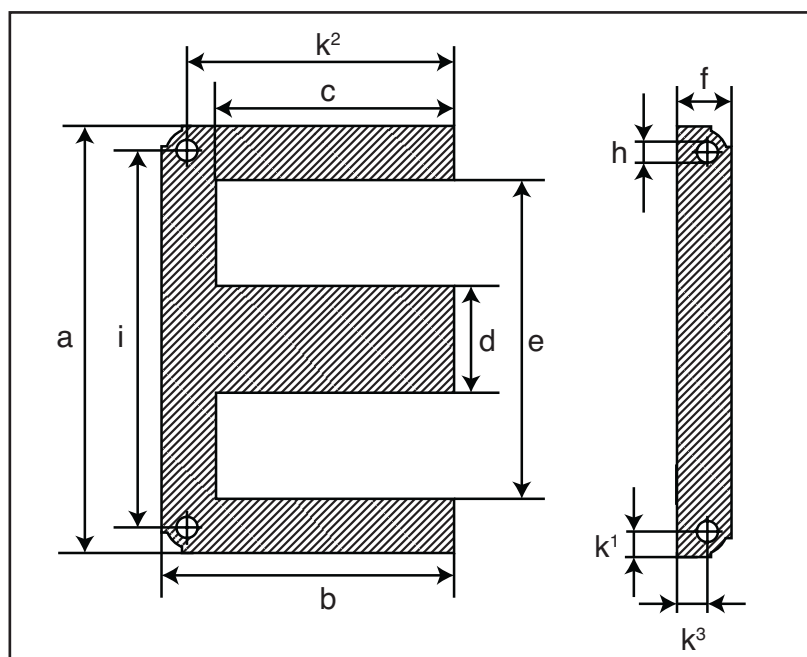


Figure 10: This image shows the low-waste EI-type cut.

and other sources that generate high frequencies. As already explained in Figure 3 and Figure 6, the inductances and capacitances of the transformer play an important role—they affect the square wave behavior. If a square wave signal of 5 kHz is still transmitted cleanly, this means that the amplifier has a bandwidth of 75 kHz.

During the test using square wave signals, it can always be seen that the edges of the square wave signal during the rise and fall are not as steep as with the original signal. This can be easily verified using a two-channel oscilloscope and superimposing the input and output signal. Likewise, greater or lesser overshoots can be determined. Those may not only be caused by the transformer itself, the interaction between the amplifier part and the transformer can also be the reason for those overshoots. If the transformer's basic characteristics are good, these overshoots can be more or less compensated by means applied to the amplifier.

One thing has still to be considered, music signals consist of a complex frequency mixture and never of square wave signals. Furthermore, a speaker membrane is not able to move from its idle position to its extreme position and then to the other extreme position and back to the idle position within short time intervals. This is a spring-mass-system subject to mass inertia. It is essential that the amplifier, in combination with its output transformer, have a significantly faster rise time than the fastest music signal.

The Iron Core

Now, it is time to examine the effect of the iron core and its qualities. Let's first start with the conventional tube output transformers and their setup. In historical circuits and their remakes, one always finds indications for the necessary core dimensions of transformers. For a power amplifier with an output of 30 W to 50 W, for example, transformer cores of the sizes M 102, EI 130, or even EI 150 are recommended with the indication that only those can guarantee the desired amplifier qualities. Therefore, the quality of a tube output transformer is often judged according to its size and weight. According to many opinions, a big and heavy transformer is good, a small and light one is rather bad.

The reasons can be found in basic literature on the topic of low-frequency transformers from the 1930s and 1960s. There is hardly any more recent substantial literature. To build good transformers,

the so-called 0.35-mm thick dynamo sheet IV made from hot-rolled silicon-iron alloy was preferably recommended. The common laminations suitable for transformers are also described.

The conventional EI-type waste-free cuts were used as the preferred type of lamination. Waste-free means that during a punching process, two Es and two Is fall out from the sheet. All other types of lamination result in more waste, which is reflected in the price. The waste-free cores have sharp corners. The low-waste sheets have rounded corners, different mounting hole dimensions and differing winding space. **Figure 9** shows the drawing of a waste-free normal EI-type cut. **Figure 10** shows the low-waste EI-type cut. The dimensions of a and b are the same with both laminations, all other dimensions differ. **Figure 11** shows the M-type cut. **Figure 12** shows the MD-type cut. Here, all dimensions are the same, except for b, f and i. For MD-type cuts, the same coil bodies are used as with the M-type cuts. One thing you will notice is that dimension f is by 50% greater with the MD-type cut as with the M-type cut.

The M-type cut has a slotted center tab for sticking metal sheets into the coil body. And with that, we have found the disadvantage of this core section. Exactly where the main magnetic flux is located, there is an air gap that causes a stray field and losses. For the MD-type cut, the "D" stands for "diagonal." Here, the slotting necessary for stacking is not located in the main magnetic flux, but where it is divided into the two cross legs. Normally, the two outside legs are half as wide as the center tab, since only 50% of the magnetic flux of the center tab need to be absorbed on each side. To be able to fully use the advantages of the MD-type cut, the cross tabs must be by 50% wider than with the M-type cut. Furthermore, the MD-type sheets are grain oriented in main magnetic direction.

To top all that off, only M111-quality sheets measuring 0.35 mm in thickness are used. In this way, the MD 102 B core can achieve up to 320 VA as the mains transformer. This is about 18% more than the M 102 B, also using M111-type sheets measuring 0.35 in thickness (maximum 271 VA), or 43% more than using 0.5 mm thick standard M330 sheets (maximum 224 VA). (The number following the "M" defines the iron losses of the relative core material. For the M330 it is 3.3 W per kilogram, for the M111 only 1.11 W.) In the MD-type cut the losses and the leakage fields are reduced and the performance is increased by four major characteristics:

- No air gap in the magnetic main axis
- Cut into the area with lower magnetic load

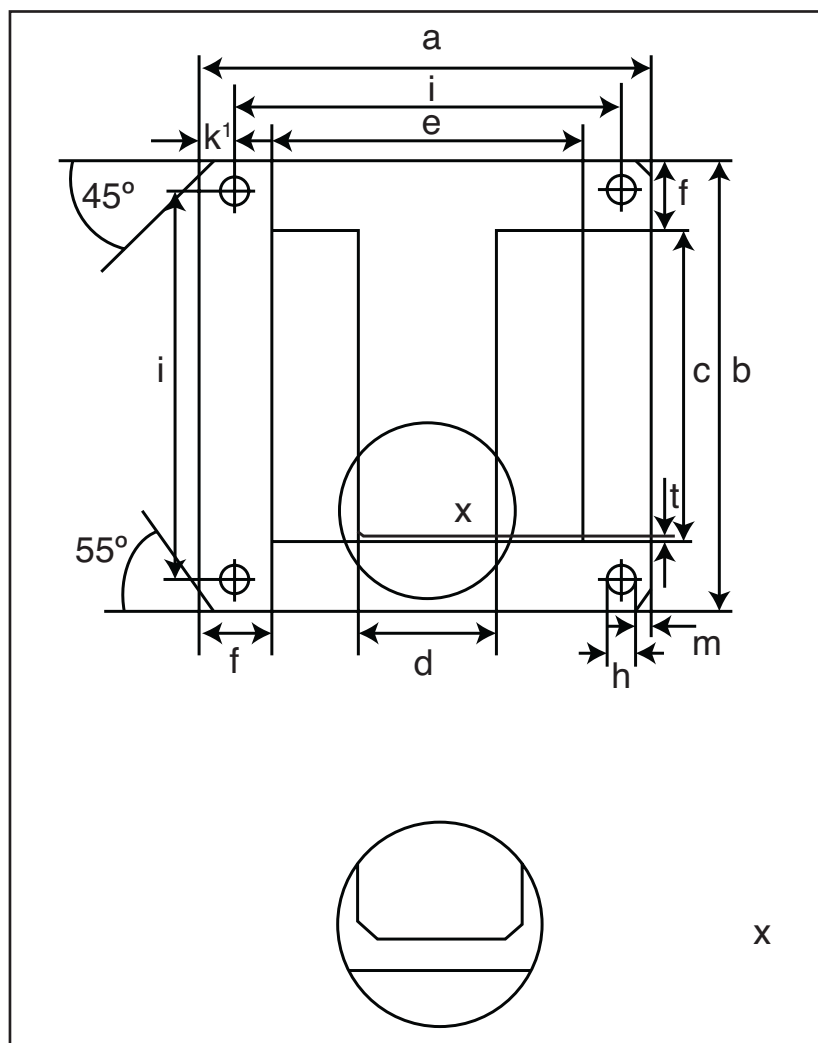


Figure 11: For comparison, here is an M-type cut.

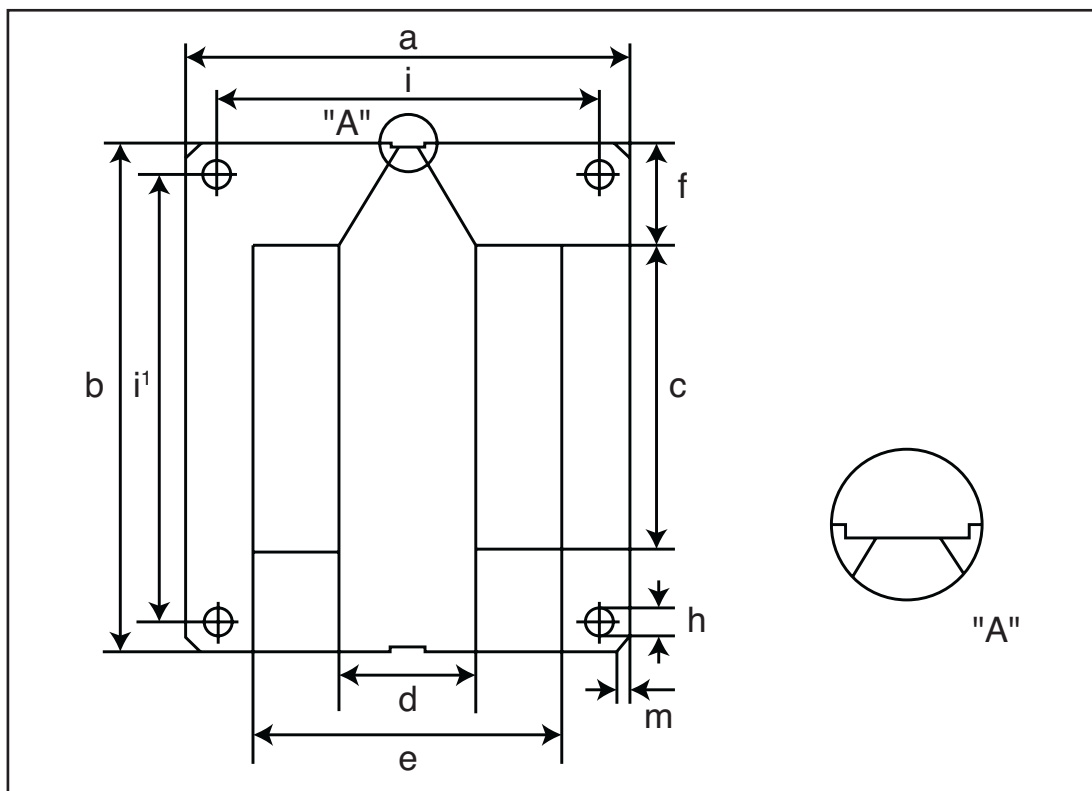


Figure 12: This is a MD-type cut. In the case of the MD-type cut, the "D" stands for "diagonal."

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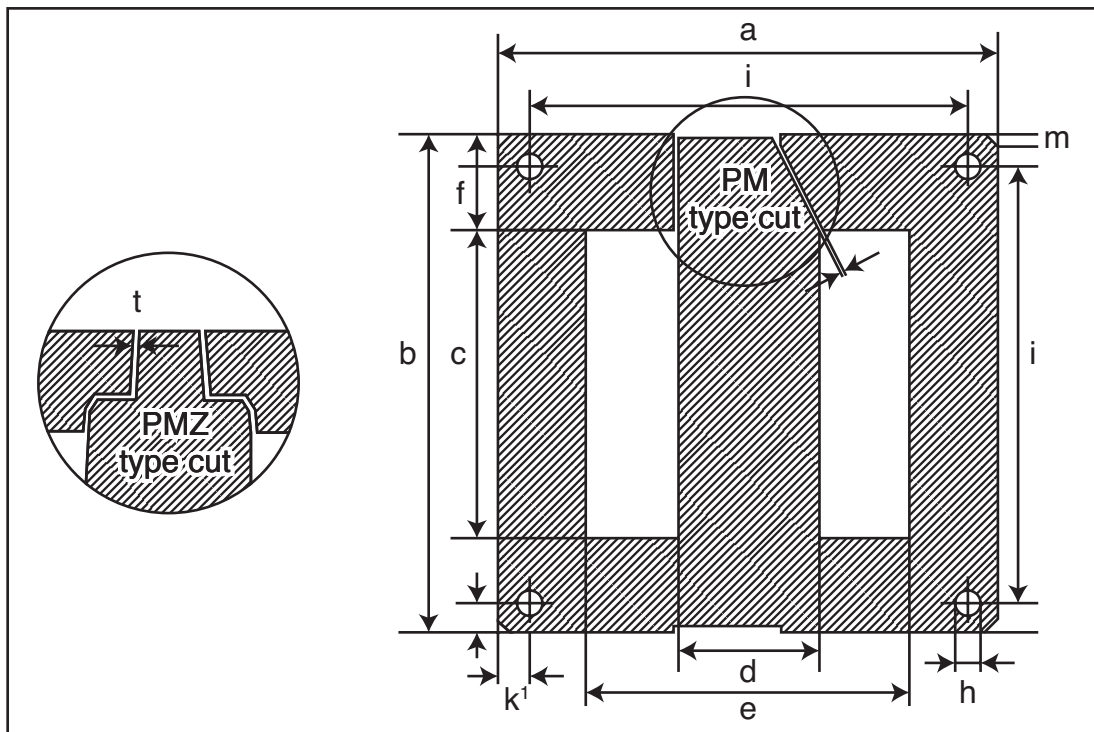
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Figure 13: This image shows the PM-type or the PMZ-type cut. The two cuts differ in their slotting, which allows for stacking the sheets.



- Low-loss, grain-oriented metal sheets with low magnetic resistance
- Due to the higher magnetic resilience of the core, a lower number of windings per volt is required, and the copper losses are reduced by the lower DC resistance

With grain-oriented sheets, the iron particles are aligned according to the main magnetic axis which reduces the losses and increases the performance. **Figure 13** also shows the PM-type or the PMZ-type cut. The two cuts differ in their slotting, which allows for stacking the sheets. The PMZ-type cut is very similar to the MD-type cut and completes the output range up to about 650 VA. The MD-type cut covers a performance range from about 10 VA to 320 VA which is absolutely suitable for low-frequency applications either as mains transformer or also as a low-frequency transformer.

Compared to the MD-type and PM- or PMZ-type cuts, the EI-type cuts have an enormous air gap because there are three joints. An iron cross-section of only 50% is expected at the joints. For special applications, there is also an M-type cut with defined air gap (see the detail X in Figure 11). EI-type cuts are well suitable for applications that require an air gap, as DC-biased transformers in single-ended power amplifiers and mains filter chokes. EI-type cuts cover an output range of about 0.4 VA up to 4,000 VA.

Several years ago, most of the transformer and core sheet manufacturers have switched to EI-type cores. The M, MD, PM, and PMZ laminations are hardly used anymore. These core formats can be stacked (bonded, welded, etc.) by machine only partially. In the case of EI-type cuts, there are machines which insert the sheets automatically into the coil body and therefore save a lot of working hours. Despite the savings on labor costs, the big disadvantage is that significant technical advantages are sacrificed in favor of a lower cost of manufacturing.

In the second part of this article series, we will discuss the useful core size and core type. Also, we will share our thoughts about the electric performances needed to get a good audio transformer.

Editor's Note: This article includes a summary of information previously published in two articles in Elektor Roehren Sonderheft and contents extracted from the book High-End Mit Roehren.

About the Author

Gerhard Haas has written books as well as articles for Elektor Germany since 1995. *The Roehren Sonderheft* magazine was his creation, along with some partners. The first issue came out in 2005. Now, there are 10 issues containing a collection of interesting articles. Gerhard has also developed many electronic circuits and equipment. In the last two decades, he primarily worked on tube amplifiers and the circuitry around them, including the audio transformers, filter chokes, power transformers. He sold his company, Experience Electronics, several years ago to focus on designing tube amps and audio transformers, write articles for *Sonderheft* and finish his book *High-End Mit Roehren*. The first copy of his book was printed in 1995, and he is now working on an updated follow up.

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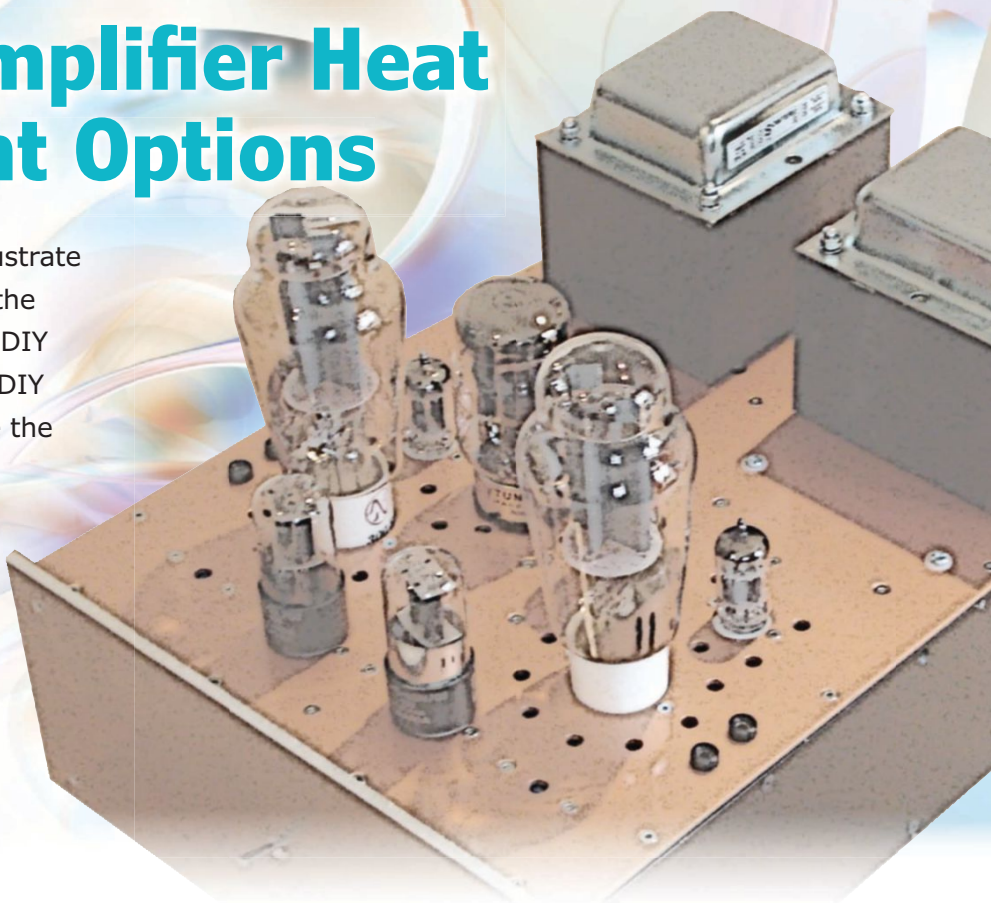
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DIY Tube Amplifier Heat Management Options

The objective of this article is to illustrate some methods to greatly increase the performance and reliability of your DIY tube amplifier. I'm only discussing DIY tube amplifiers since they generate the most heat per watt of audio output and are hopefully of the most interest to you. Most designs that I have seen in *audioXpress* have the tube and the transformers on top of the chassis and everything else under the chassis. Most of the amplifiers look nice and work okay, but I think there is a better way!



By
Patrick Brunner
(United States)

Excessive heat is the known killer of tubes, transformers, resistors, and especially electrolytic capacitors. For any given tube amplifier design, the quiescent heat (actually wasted energy) is nearly fixed, so the actual solution comes in the form of heat management—or how to get excess energy from where you don't want it to some place where it will do less or no harm. And when you build a nice compact tube amplifier things get even hotter! Heat/energy gets transferred or moved by radiation, conduction, and (in this case forced air) convection.

In a confined space, like the under chassis of a typical tube amplifier, there often is little or no ventilation so the heat builds over time, cooking the under chassis parts, especially the electrolytic capacitors. Drilling a few holes around the output tubes is a bit futile (where is the air source in a typical closed bottom chassis?)—unless there are a lot of ventilation holes in the elevated chassis bottom plate.

Ways to Decrease the Temperature

One option is to buy more expensive higher-temperature-rated capacitors (or even more expensive polymer film capacitors). Or in the case of electrolytic capacitors, epoxy seal the crimped

ends to reduce the rate of drying out from heat. Also be aware the capacitor's equivalent series resistance (ESR) determines how much internal heat will be generated inside the capacitor—lower ESR = less heat. But these options do not address the problem, they just postpone it.

Resistors are more tolerant of heat and buying a larger, higher wattage part will reduce the surface temperature but it will have no effect on the amount of heat energy dissipated by the part. $P = I^2 R$ doesn't change regardless of the part's rated wattage.

Transformers are a big cost item and cost reducing short cuts on the power transformer usually make for a hot transformer. Consider removing the end caps and adding sheet aluminum heat radiators, painted flat black except where they make contact with the core, and mounting them with thermal paste. This may help in a retrofit, and might be useful in a new amplifier build if you are pushing the transformer's rating close to or beyond its ratings.

The Solution

The real solution is to get the ambient air moving in a controlled manner. Consider a fan mounted on a uniform tube, with the far end capped, and with



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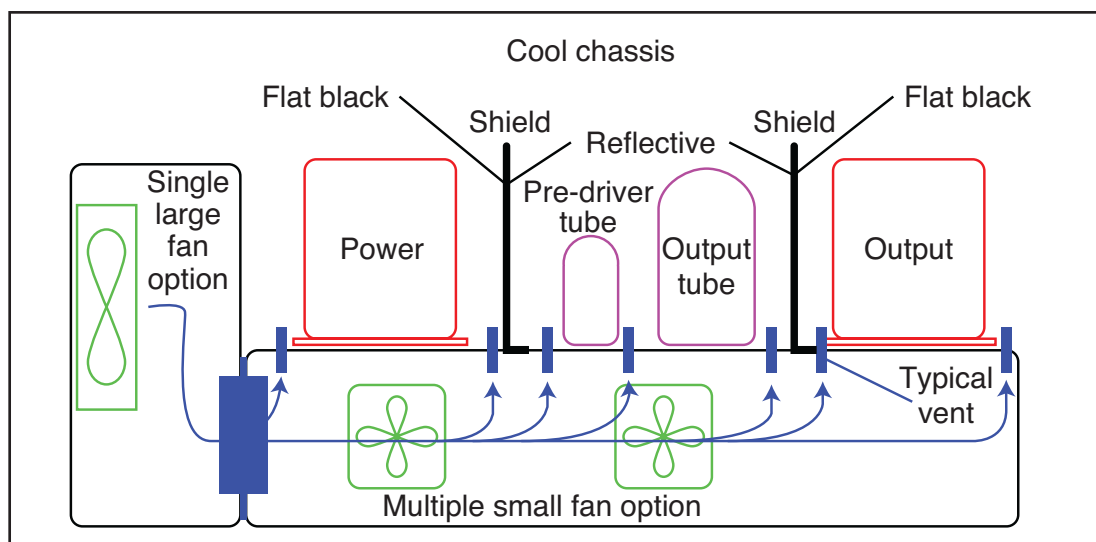
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Figure 1: This concept can be applied to a tube amplifier.



uniform series of holes drilled at equal intervals. The holes nearest the fan will have the highest pressure and velocity while at the opposite end the pressure and velocity will be reduced due to air venting and friction.

Use the typical improved layout (see **Figure 1**) for a mono-block (or even a stereo version) amplifier, thinking left to right and the near edge as the "front." there would be power transformer (hot), tubes (hottest), and output transformer (comparatively cool). This assumes a standard rectangular chassis with matching base plate (without vent holes) and some appropriate mounting feet.

Adding a standard box fan in a vertically mounted 2" or 3" deep chassis with base plate (same width as the main chassis) with (optionally) a fan filter or at least a protective grill is an excellent source for controlled forced convection. The chassis should be taller than the main chassis to protect tubes and transformers from damage, and the fan should be mounted closer to the top, the opposite side from the main chassis. Cut a hole (as large as practical) in the end of the main chassis and a matching one in the fan's chassis and mechanically join them (try speed nuts). You now have a pressurized under-chassis compartment.

At the power transformer end (left) drill a series of holes (1/4" diameter should be fine) around the periphery of the power transformer. Spacing will depend on the transformer's weight and your chassis' material and thickness. Remember adding holes is far easier than removing them. For very large transformers (larger than 10 lbs.), I would suggest a 0.093" thick aluminum chassis. If your mechanical design uses a cover (ideally made of steel) to totally enclose the transformers, be sure to add vent holes in the top or at least very near the top of the cover.

If you have gone with a tube rectifier locate it near

a long axis edge of the main chassis at the power transformer end. If you are using chassis mounted can capacitors put them on the opposite long axis edge side of the power transformer (keep the tube's heat away from the capacitors). Add more ventilation holes for the rectifier tube, or mount the tube socket recessed (1/4" should be okay) below the chassis and use an enlarged clearance hole (1/4" larger diameter than the tube base/socket) so the rectifier tube gets 360° forced ventilation.

If you went with solid-state rectification and under-chassis capacitors, only the transformer will need the ventilation holes. If you used an aluminum block-style bridge rectifier, the ventilated aluminum chassis will offer a superior heatsink (use thermal paste). NOTE: stainless steel chassis conduct heat only ~30% as much as an aluminum chassis. Consider stainless as a thermal insulator. Also at this end on the chassis side wall, I would suggest an IEC-filtered power entry with a filter module, an On-Off switch, and maybe a pilot light.

Moving to the center section where the tubes are located, use the same approaches as described for the rectifier tube. Input tubes on the left, output tubes on the right nearest the output transformer. The output tubes should have a greater overall vent area than the power transformer, therefore, more cooling for the hottest parts—the output tubes.

Additionally consider a steel, chassis wide, transformer high, "L" partition painted silver on the tube side and flat black on the transformer side, located between the power transformer and the tubes. The silver paint will reflect the radiating heat from the tubes and the steel will reduce the magnetic field from the power transformer and hopefully reduce hum issues. The partition will also strengthen the chassis in the area of the power transformer. Don't use aluminum,

if at all possible. It is non-ferrous so it won't affect any magnetic field.

The center section of the chassis side wall is ideal for the input jacks and any volume or other controls you want to add. Last, at the output transformer (right end), consider another identical steel partition between the output transformer and the tubes to reduce radiation heating of the output transformer. Even though the output transformers run cooler than the power transformer, a few ventilation holes will maximize cooling of your costly output transformer.

This end section of the chassis side wall is ideal for the output jacks and/or binding posts. The idea here has been to minimize wire lengths to the chassis side wall mounted components.

Miscellaneous Fixes

Fans only need to move a little air to greatly improve cooling. They do not need to run at full, noisy speed. But more air moving means better cooling.

Fans can be connected in series for slower quieter operation, or simply connected to a lower than specified voltage source. Be sure to check for stalling at reduced voltages. Two 120 VAC fans in series or a 240 VAC fan running on 120 VAC are good options.

DC fans can be implemented using a filament supply and 1N4007 series diodes (there is little cost between a 50 PIV and 1000 PIV 1N4000 series diode—stock up on 1N4007s) and minimal filtering—a fan does not need super clean DC to work. A 6.3 VAC source will yield about 8 VDC, enough for a 12 VDC fan, or with a 12.6 VAC source will yield about 16 VDC, enough for a 24 VDC fan. If you used solid-state rectification, a spare 5 VAC source with a voltage doubler might be another good option.

In severe heat load cases, two (or

more) fans could be used. Another fan plus a chassis could be added to the output transformer end, or compact fans side by side using a long chassis connected to the “back” vertical edge of the main chassis. A 3” tall main chassis could accommodate some smaller (~70 mm) fans or compact small computer/surplus fans internally.

Buzzing transformers can often be quieted by hammering a wooden shim between the core and the windings, sometimes needing shims in all four possible locations. Use a chisel to cut the shims flush with the core and add some epoxy to finish this fix. Be very careful and don't hit the windings.

High power output cathode resistors—could they be mounted through or above the chassis, or heat sunk to the chassis? Could the preamp tubes' filaments serve as all or part of the cathode resistors? Free DC preamp filaments as well (Harman Kardon A500)! Fixed bias would probably be better, cooler and usually provides more output power. Ever consider an auto bias circuit? Replace the cathode resistors with a constant current source (CCS) power transistor mounted to the aluminum chassis (with insulator and thermal grease if needed). Examine past *audioXpress* issues for typical articles and details.

If you don't abhor solid-state parts, consider a chassis-mounted and heat-sunk solid-state device(s) for B+ and/or screen voltage regulator pass elements, which usually reduces distortion. Don't forget to use shielded wire for the input line as well as feedback line, if used.

All of this advice is given without any guarantees, but with good intentions and best wishes for your success. Hope some of these suggestions prove useful and make your “cool” DIY amp a lot cooler. Best regards, Pat 

About the Author

Patrick Brunner is an independent design consultant. He started Applied Technical Services in 1982, designing primarily in the medical electronics field. Prior to 1982, he worked at various military, industrial, and medical electronics companies. Patrick resides in Florida.



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TubeSociety 2A3-300B SE Amplifier Project (Part 2)

Saving the Best for Last

In the first part of this article, I described, four 300B single-ended (SE) amplifiers. Now, I will discuss the fifth amplifier, detailing the power supply section and the construction. At the end of the article, I will discuss my conclusions about this experience.

By
Menno van der Veen
(The Netherlands)



The first three amplifiers discussed in Part 1 all exhibited a general behavior. They had rather large total harmonic distortion (THD), where the second and third and higher harmonics decline in amplitude. The fourth amplifier (the 300B-Trans) strongly deviated from this behavior by showing remarkably less THD, higher speaker damping, and the second harmonic was much smaller than the third. This behavior allows the micro details to be clearly heard. In Amplifier 5 (see **Photo 1**), I attempted a different route to reach for the same excellent results I achieved with Amplifier 4.

Use a Tube Driver

The blown Amplifier 4 taught me that DC-coupling can be very nasty. So, back to input or coupling capacitors, but they needed to be of excellent quality. So, for Amplifier 5, I decided to use a high-quality tube driver. I did many experiments, with mu-stage and the like, but these circuits did

not fully satisfy me, mainly because of the lack of available research time.

Then, I thought, “suppose I use tube circuitry with real current sources instead of cathode and anode resistors, would that be the right way to go?” **Figure 1** shows the resulting schematics. I also changed the operation point of the 300B.

My earlier experiments had shown that the 2A3 setting might be great for the 2A3, but not for the 300B. Now I use $V_{ak} = 340\text{ V}$, $I_o = 90\text{ mA}$, $V_{gk} = -70\text{ V}$ and $Z_a = 3.5\text{ k}\Omega$ primary impedance by loading the 4 Ω secondary tap with an almost 6 Ω speaker. Remember $Z_a/Z_s = (N_p/N_s)^2$, where the turn numbers N_p and N_s are fixed. Changing Z_s automatically changes the primary impedance Z_a in ratio.

The BSP135 current source (1.2 mA) in the first anode creates an almost horizontal load line, thus lowering the distortion of the first amplifier stage (see **Photo 2**). The second tube section acts as a cathode follower. Its distortions are minimized by

Photo 1: Amplifier 5 was the final amplifier Menno van der Veen made for the TubeSociety 2A3-300B SE Amplifier Project.

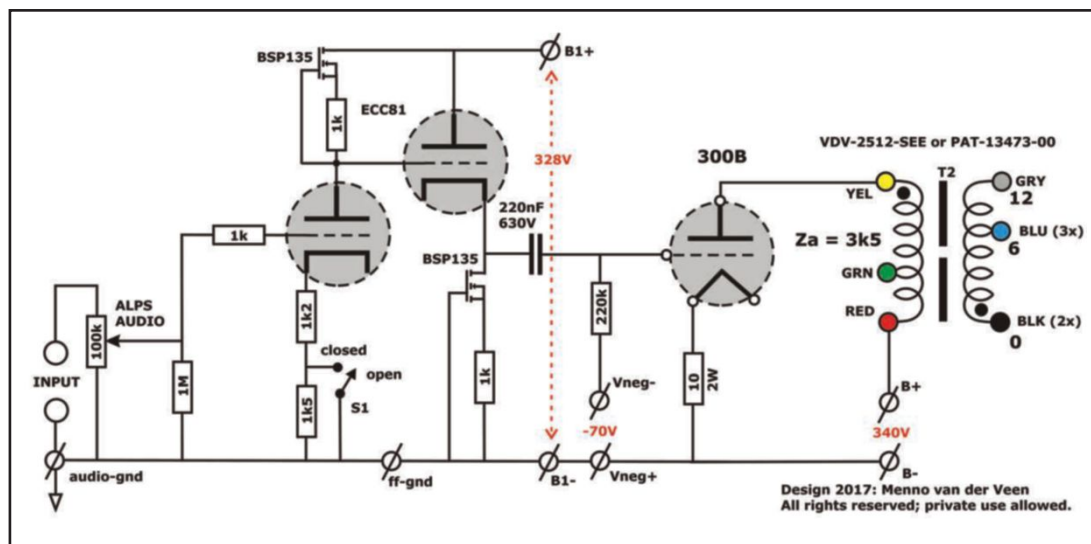


Figure 1: I used this schematic to build Amplifier 5.

the second BSP135 current source. This current source is also applicable in Amplifier 4.

The operating point (anode voltage) of the first triode can be changed by closing S1. With open S1 the cathode follower output is at 190 V above ground; with closed S1 the output is at 120 V. The circuit is able to drive the 220 k Ω grid resistor of the 300B. I know that 50 k Ω is prescribed for setting the operation point with negative grid voltage. However 50 k Ω would be a too heavy load for this driver section. With my Svetlana 300B, the operation point stays stable with $R_g = 220$ k Ω . To explain the functionality of S1, I will discuss this amp with S1 = open and with S1 = closed.

S1 Open

When you compare **Figure 2** with the measurement shown in Figure 7 from Part 1 of this article series, you see that the THD has gone



Photo 2: Here is the simple construction of the constant current source.

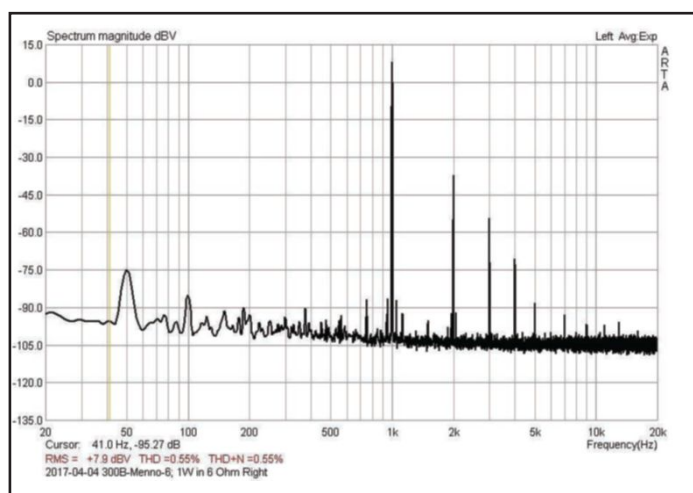


Figure 2: Amplifier 5 with S1 = open is measured at 1 W in 6 Ω load.

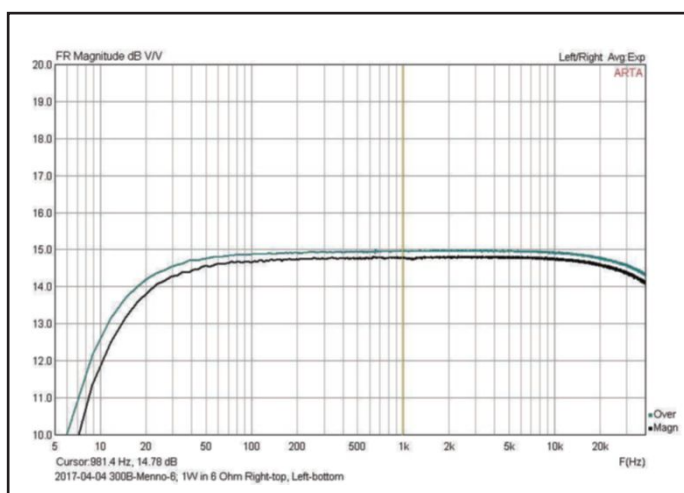


Figure 3: Amplifier 5 with S1=open is shown with left and right channels.

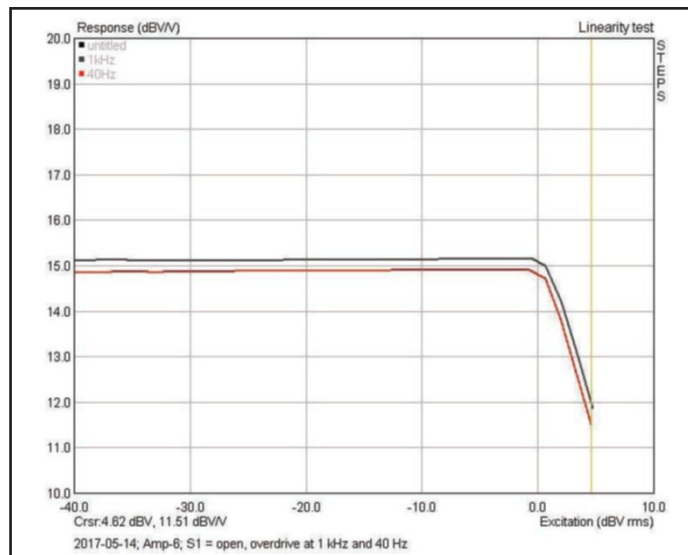
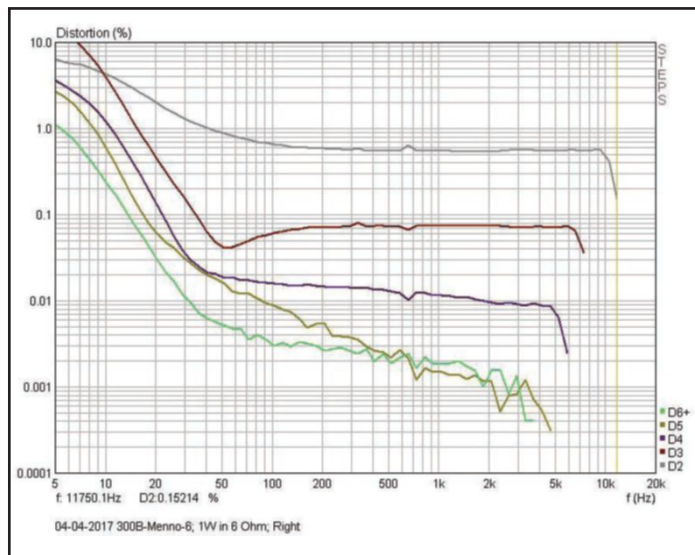


Figure 4: Amplifier 5 is measured with distortions at 1 W in 6 Ω .

Figure 5: Amplifier 5 is shown with linearity at 1 kHz and 40 Hz (red).

down by a factor of two. This is mainly caused by the improved operating point of the 300B.

The frequency range shown in **Figure 3** is as wide as before and gives no reason for concern. However, I also want to mention to please measure the f-range at full power, to check that no slew limiting occurs (that happens when dV/dt is too small) in the audio range up to at least 20 kHz. I checked this and there is no reason for concern with Amplifier 5.

However, the reason why I mention this is because I noticed with some other amps a strange low-to mid-frequency blur when playing a CD containing a nightingale recording. With a nightingale recording, there is often louder high-frequency information present. If dV/dt is too

small, the amplifier can't handle those signals, resulting in the blur that I mentioned. It took me a while to discover its cause. That's why I share this experience now.

Harmonics as function of frequency are shown in **Figure 4**. It is rather standard behavior and we have seen this before.

Figure 5 shows the linearity ($= 20 \cdot \log[V_{out}/V_{in}]$) expressed in dBV/V as function of V_{in} for 1 kHz and 40 Hz. Overdrive occurs around $V_{in} = 0$ dBV ($= 1$ Vrms). The overdrive knee is rather soft (it also sounds "soft"), and no core saturation is detected at 40 Hz.

This is a healthy 300B design—7 W output power (at 10% THD) and an output impedance of 1.5 Ω at the 6 Ω tap. There is also a standard accepted

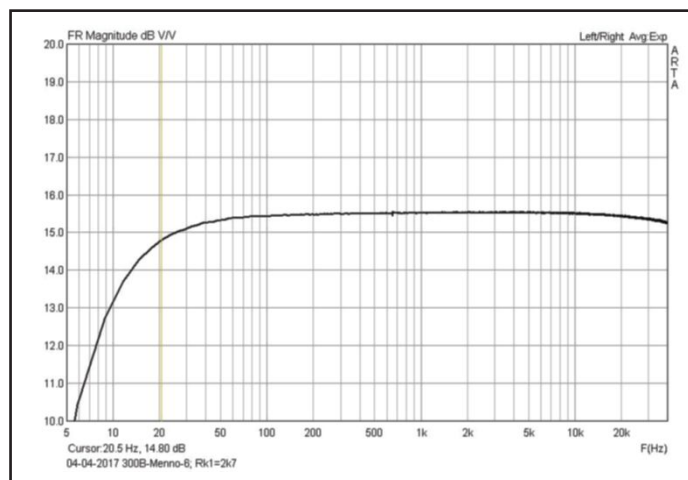
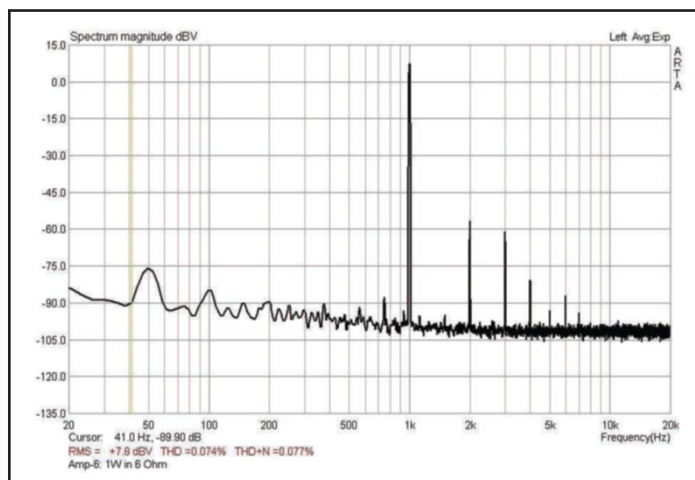


Figure 6: Amplifier 5 with S1 = closed is measured at 1 W in 6 Ω load.

Figure 7: This is the frequency range measured for Amplifier 5.

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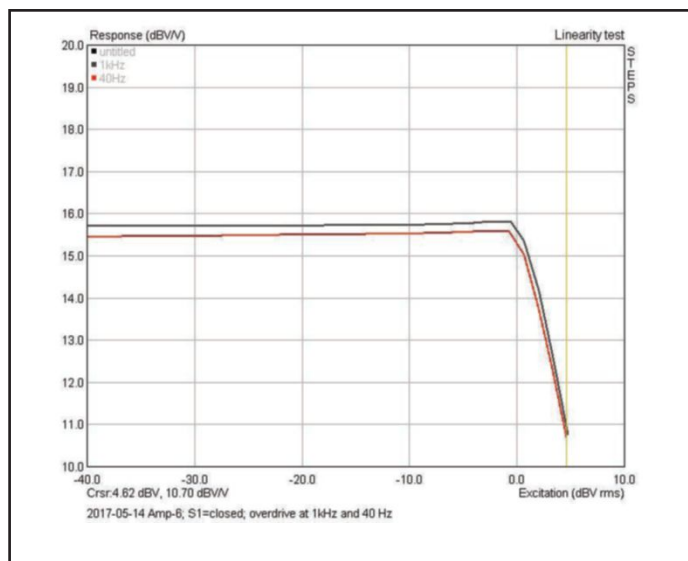
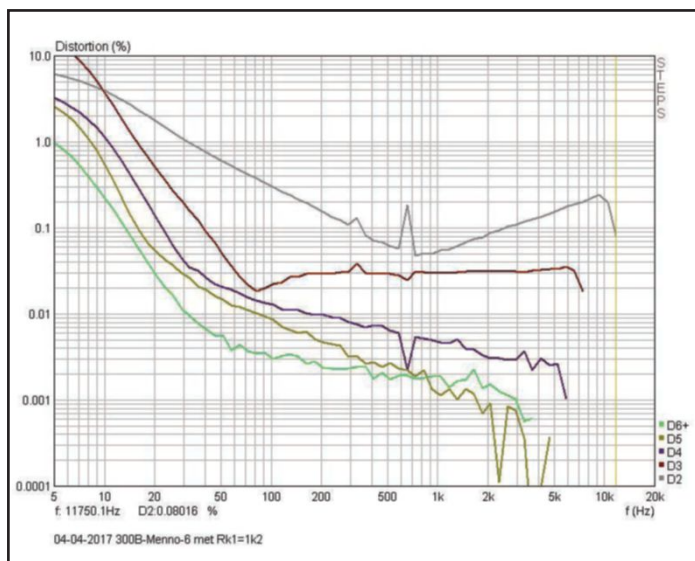


Figure 8: Amplifier 5 is measured with distortions at 1 W in 6 Ω .

Figure 9: Amplifier 5 is shown with linearity at 1 kHz and 40 Hz.

decline in harmonics, which stay under the 1 kHz masking curve plus a wide enough frequency range.

How does this amplifier sound? Warm, an embracing open sound stage, easy to listen to, but again a curtain caused by the second harmonics. I have heard better, especially with Amplifier 4. So, let us now close S1.

S1 Closed

Figure 6 already shows that the harmonics are largely suppressed. The frequency range and amplification in **Figure 7** are almost the same as with S2 = open (compare to Figure 3). This is as expected because the first input triode has a current source at the anode. Then the amplification is hardly influenced by how large R-cathode is and is almost equal to the mu-factor of the first triode.

Figure 8 shows in full detail what is happening. The harmonics 2, 3, and 4 are smaller than with S2 = open (compare to Figure 4). This behavior is almost equal to what the Trans was doing, although I do not apply any negative local feedback. What happens is that the driver section pre-distorts in an inverse manner the signal to the grid of the 300B. The driver and the 300B distortions together compensate for each other, resulting in a very small effective total distortion. Again we see below 20 Hz the dominant core distortion. From 20 Hz to 1 kHz, we see in the second harmonic the decreasing oval load line effect of Lp. Above 1 kHz to 20 kHz, we again notice the oval load line effect of Cip. This pre-distortion technique is a very effective way to suppress the distortion of the complete amplifier.

Figure 9 shows the linearity, which we compare to Figure 5. Now the overdrive-knee is more abrupt. The tonal character of overdrive sounds "sharper" than it did with all the other amplifiers discussed in Part 1 of this article. In my opinion, this is the only disadvantage. However, the solution is simple: Don't play it that loud.

In my subjective opinion, this fifth amplifier with S1 = closed has many fine details, there is

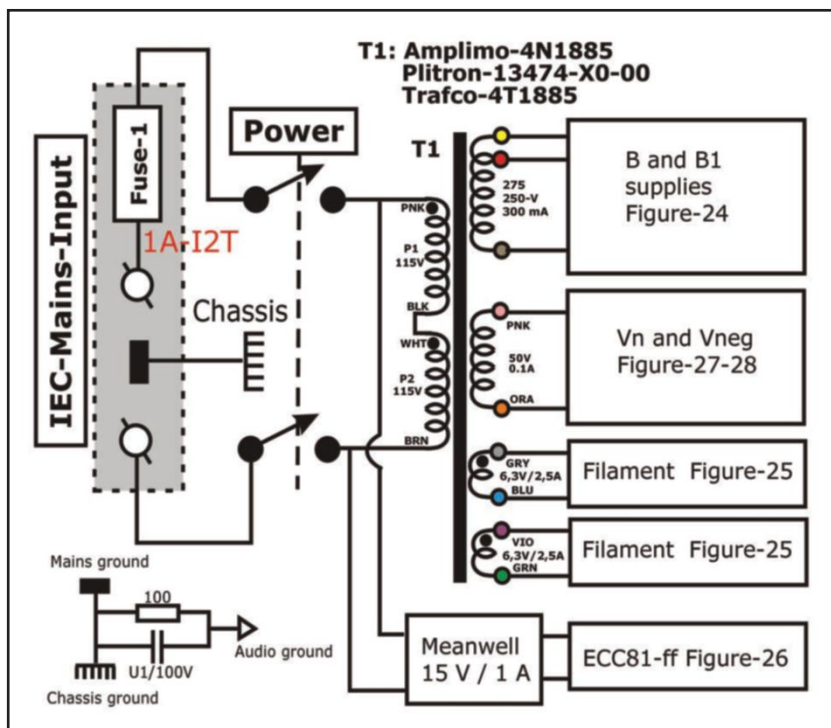


Figure 10: Here are the primary mains connections and grounding. The 230 VAC connection is shown. For 115 VAC connect BLK to BRN and connect WHT to PNK.



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Photo 3: Here is the R//C-coupling between chassis and audio-ground.



Photo 4: This is the toroidal power transformer for Amplifier 5. On top of it, the Meanwell AC/DC converter plus slow-start for the ECC81 filament.

no curtain, a wide and deep open soundstage, it's embracing, has excellent tonal balance, and is almost comparable to all the goodies of Amplifier 4. Even after late hours listening into the night and over the next few days, Amplifier 5's character stays convincing. Now, with classical tube driver section and modern current sources, I am again close to my goal. This soundstage is honest, usable for critical evaluation of a recording and enjoyable listening with total focus on the music. This design, as with Amplifier 4, pleases me.

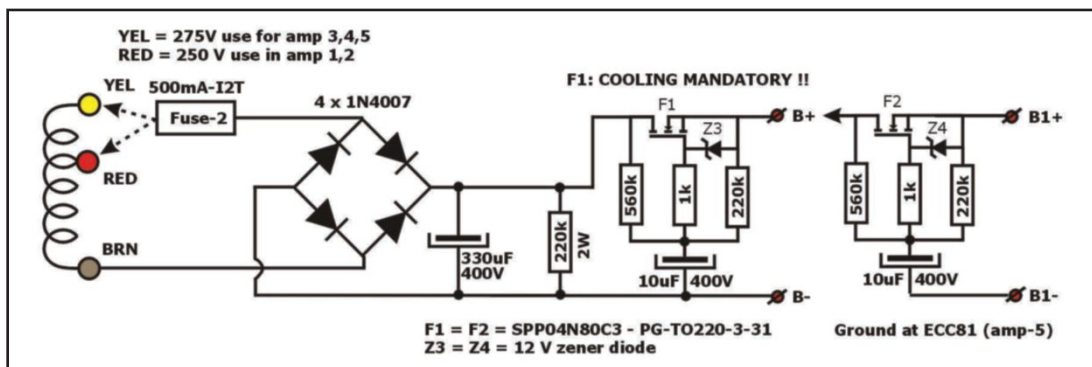
This last amplifier completed my support to my students. Next, I will discuss the power supplies and how I constructed my Amplifier 5, followed by some conclusions.

Power Supplies and Construction

A general overview of the complete power supply section is shown in **Figure 10**. The power transformer^[1] is a toroidal. How the mains are connected is shown as well. Mark the regional voltage in your country. Here, in Europe, it is 230 VAC (two primaries in series, as drawn). America has 115 VAC, where the two primaries shall be in parallel.

The bottom left side of Figure 10 shows how to ground this amplifier. The mains-ground is directly connected to the chassis for reasons of safety. The audio-ground is connected (at the input of the amplifier) to the chassis via a parallel R-C circuit of 100 Ω and 100 nF (see **Photo 3**). This special

Figure 11: This shows the high voltage supplies B and B1 for Amplifier 5.



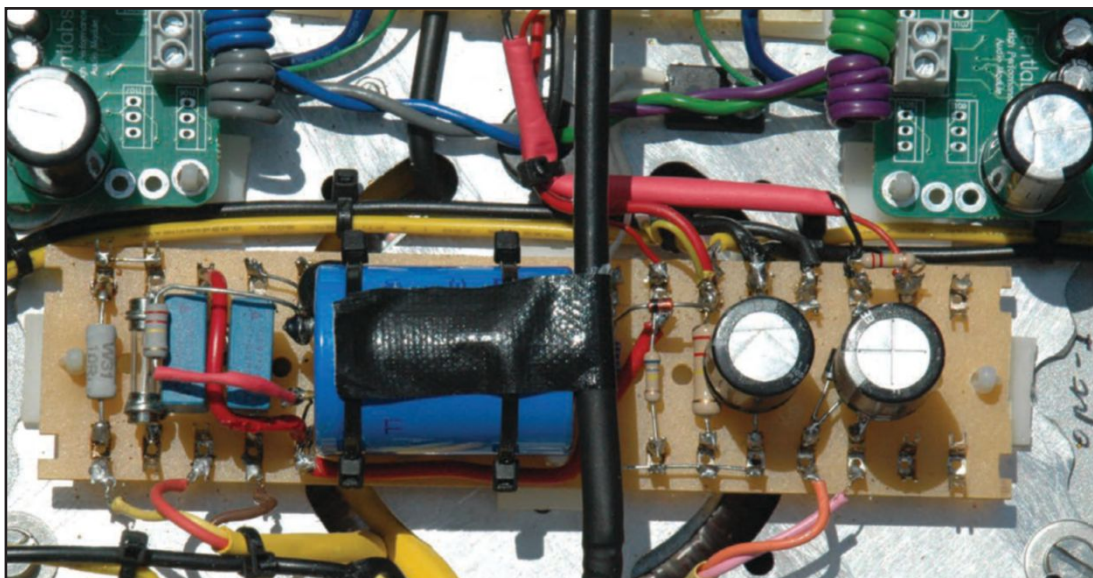


Photo 5: Practical realization of the B+ and Vneg supplies.

About the Author

Menno van der Veen studied engineering physics at the university. He taught physics at an upper secondary school and teacher trainings college. At the age of 40, he founded his engineering firm with a focus on tube amplifiers and toroidal output transformers. He wrote many articles and books about these subjects. About 12 years ago, he also started his TubeSociety academy. Tube-amp research, consultancy, design, teaching and playing the guitar are the main activities of his present work.

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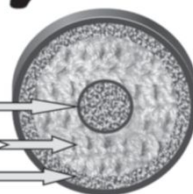
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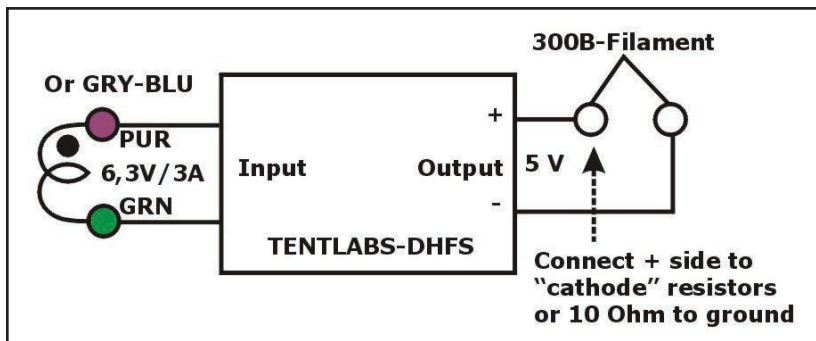


Figure 12: This is the 300B 5V filament supply (two of them are needed for stereo).

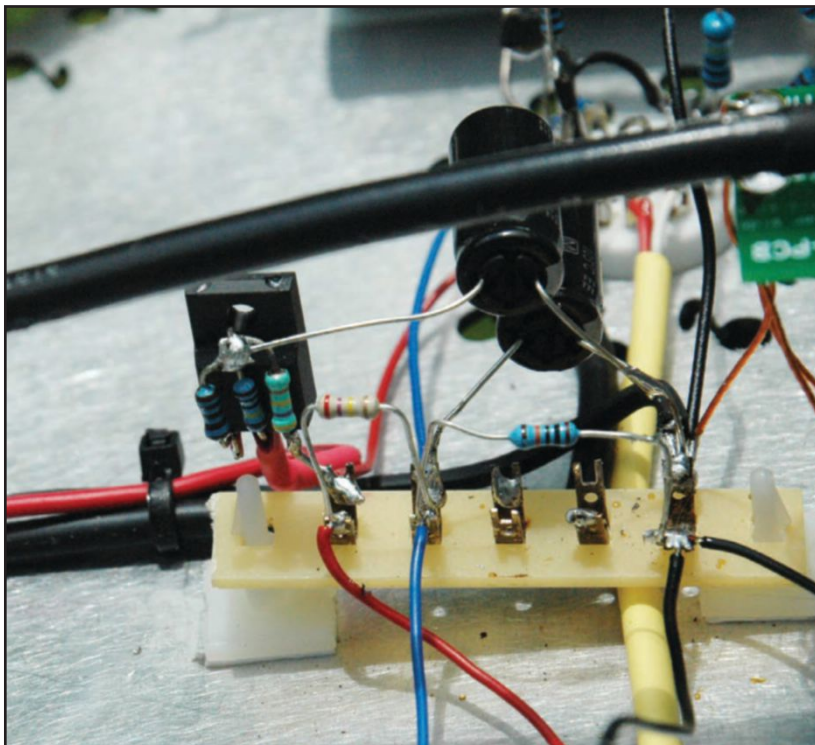


Photo 6: The B1+ supply is placed close to the ECC81 tubes.

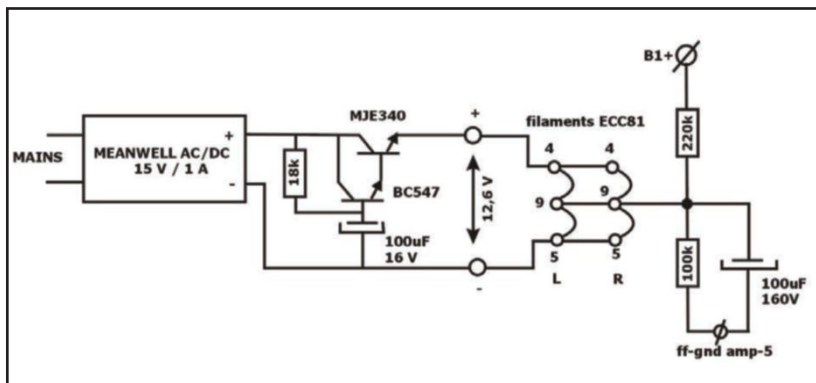


Figure 13: This schematic shows the ECC81 ff supply in Amplifier 5.

connection prevents humming ground loops. **Photo 4** shows the toroidal power transformer.

Figure 11 shows the schematics of the B+ and B1+ supplies. **Photo 5** shows the interior of Amplifier 5.

The high voltage winding (250 V or 275 V) is rectified, buffered by 330 μ F/400 V and fed into SPP04N80C3, the special TO220-isolated power field-effect transistor (FET). The capacitor at the gate of 10 μ F/400 V is slowly loaded by the drain-resistor of 560 k Ω . This creates a very slow start-up of the high voltage B+.

The first FET after rectification should be cooled on the chassis. It dissipates a maximum heat of 2×90 mA times 12 V between drain and source. The heat equals 2.2 W, which demands this extra cooling. Please note that the 1 kHz gate-resistor should be placed very close to the gate of F1 (and also at F2), to prevent nasty oscillations. The B1+ supply repeats the circuit of the first FET, but dissipates almost no heat: Some milliamps times 12 V is 0.1 W; no cooling mandatory. The output impedance of both FET circuits is 2.5 Ω at frequencies above 20 Hz. Therefore, no further electrolytic caps between B+ and B- are needed. The ripple voltage at the B+ output is smaller than 1 mVRMS under full current load.

Figure 12 shows the schematic of the 300B filament supply. **Photo 6** shows the actual filament supply placed on Amplifier 5.

On the power transformer are two 6.3 V secondary windings. Per 300B, a winding is used and fed into the Tentlabs Direct Heated Filament Supply (DHFS)^[2]. The output is trimmed for 5 V and connected to the 300B filament. The positive (+) output shall be connected to the cathode resistors or to the 10 Ω grounding-resistor, as indicated in the amplifier schematics. The Tentlabs DHFS is silent and acts as a current source, preventing audio currents from entering the supply circuit. I am impressed by its silence and resulting clear sound. Remember: Each 300B has its own 6.3 V secondary winding and its own DHFS. There are no interconnections between the two independent 300B filament circuits.

The filament of the two ECC81 tubes in Amplifier 5 *cannot* be connected to the existing 6.3 V windings, because these winding are only meant for the 300B and should stay completely free of any other connection or use. Therefore, I applied a Meanwell AC/DC converter, which converts 90 to 230 VAC mains into a DC voltage of 15 V (see **Figure 13**).

Its output got a slow start circuit (no flash-on

in filaments) around MJE340 plus BC547. Some cooling is necessary for the MJE340 because the total heat is 0.6 A times (15 to 12.6) V = 1.44 W. The filament output is lifted to B1+/3 V to prevent too large voltages between filaments and cathodes of the ECC81. This circuit is also extremely silent and no influence on the amplifiers output is detected.

In Amplifier 3 and Amplifier 4 (described in Part 1 of this article series), an extra clean voltage V_n is applied to lift the B-supply 60 V above ground (see **Figure 14**). The 50 V winding on the power transformer is used.

In Amplifier 5, a simpler version of V_n is applied, called V_{neg} (see **Figure 15**). Photo 5 (the right side) shows how it actually looks on the amplifier. Negligible current is asked from this supply. That is why this circuit is very simple.

Never try to improve these circuits by means of Zener diode clamping or DC stabilization. If the mains voltage rises, so will $B+$ and so will the V_n or V_{neg} , thus compensating drift of the operation point through changes in V-mains.

Also note that in the V_n and V_{neg} supplies no surges have been used. These two supplies

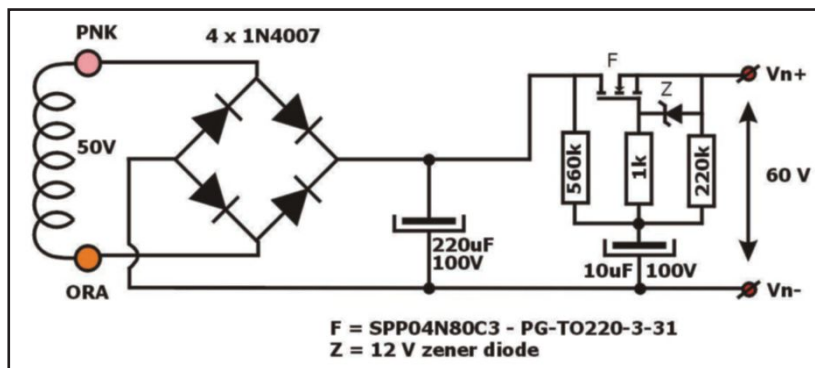


Figure 14: $V_n = 60$ V supply for Amplifier 3 and Amplifier 4 (detailed in the first part of the article). Place 1 k Ω very close to FET F.

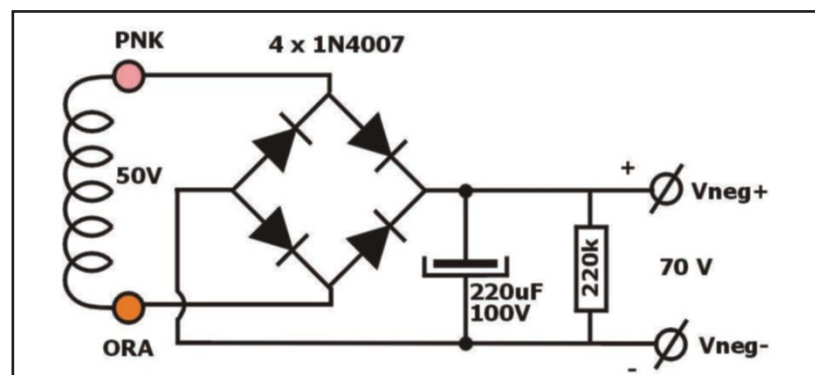


Figure 15: Amplifier 5 is shown with $V_{neg} = -70$ V supply.

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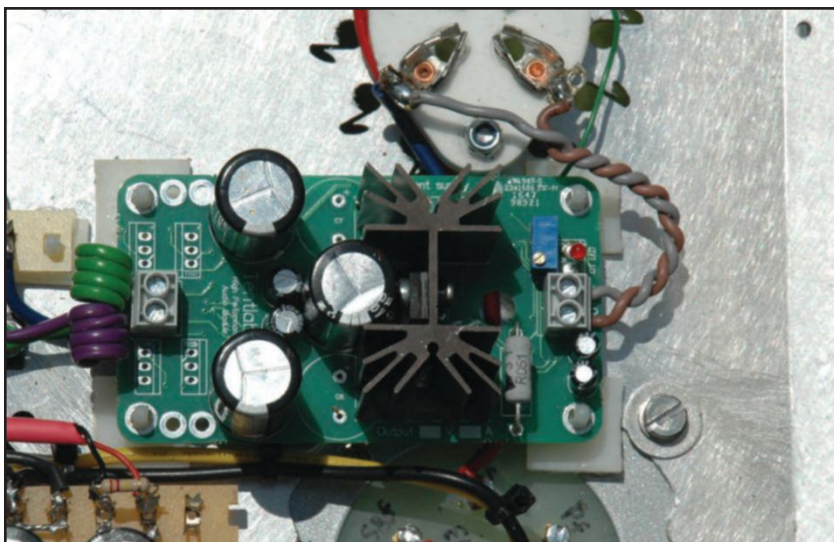


Photo 7: I used the Tentlabs filament supply for 300B.

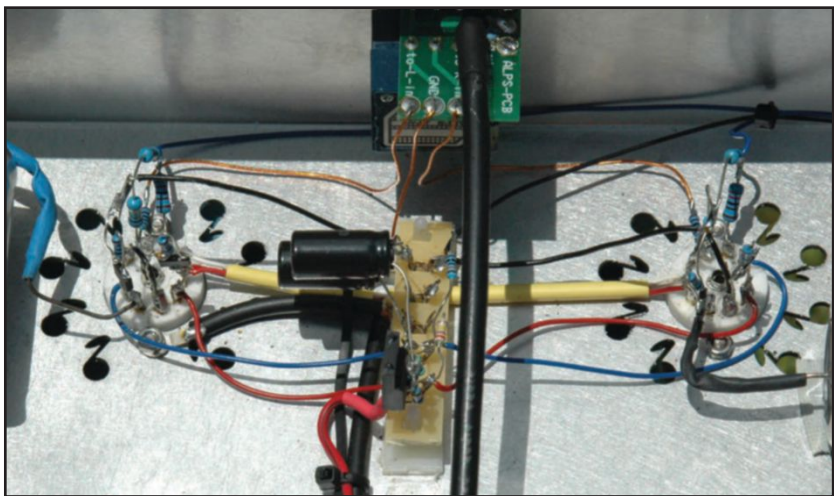


Photo 8: These are the ECC81 sections. The CCS circuits are very small, hardly visible, and mounted directly on the tube sockets.

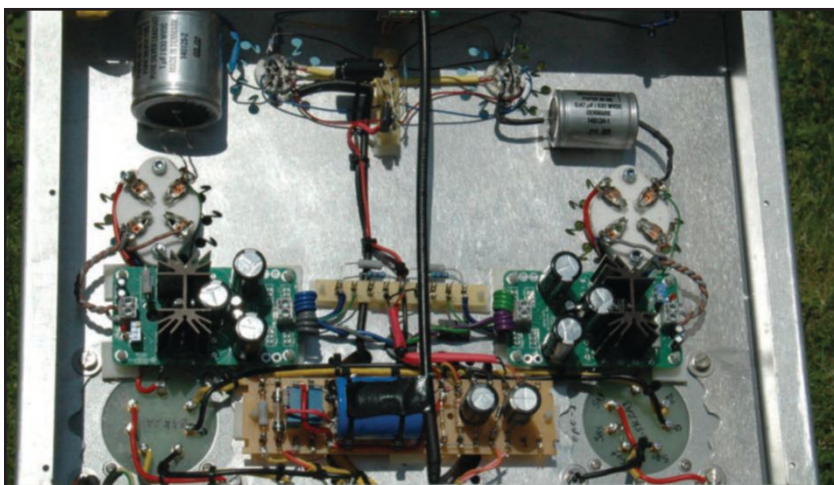


Photo 9: This is the overview of the inside of Amplifier 5. I used two extra-large Jensen silver in oil capacitors because I had them in stock. The prescribed $\mu 22/600$ V versions are smaller.

always should function, otherwise the 2A3 or the 300B valves will certainly see their life shortened because their anode currents will surpass the maximum limits due to the possible absence of V_n or V_{neg} .

Photos 7–9 give a conclusive general overview of Amplifier 5's construction.

Conclusions

Based on my long experience with push-pull valve amplifiers, plus the findings of this SE research, I would like to draw some conclusions.

If your goal is to hear as many details as possible, then the amplifier's THD should be as low as possible. You can apply massive overall negative feedback (Bruno Putzeys^[3]), or local feedback around the output tube (as in Amplifier 4), or massive local feedback (Frank Blöbbaum^[4]) or pre-distortion compensation (as in Amplifier 5 with S1 = closed). Knowing what the magnetic hysteresis in OPT-core does with the sound, I prefer local feedback in front of the OPT. Please don't include the OPT in the feedback path.

If you wish the music to embrace you, then THD may be present, especially the second harmonic, but details get lost.

The coupling capacitor's brand and construction are of the utmost importance to hear the embracing and details in a non-feedback design. Exactly the same can be said about the quality of the resistors and OPT-core material. For this research project, I did not proof it by my own measurements. In the literature, there is an abundance of the highest quality proof^[5,6] available to show low-level defects caused by components. The ear can detect them.

Designing a fine ("fine" is a subjective property) amplifier is an art. The general understanding is that (in SE amplifiers) local feedback or pre-distortion weakens the second harmonic and hardly the third. This research has proven that the third and higher can be attenuated as well, although not as much as the second. Peter Baxandall's research (discussed in Part 1 of this article series) shows the harmonics above 2 are the first to rise in level with the amount of feedback. However, that was measured on a JFET. In tube circuitry, apparently, things are a little different. It is not the first time that I notice the third and higher harmonics to decline directly with the amount of negative feedback.

There is a general opinion that the harmonics in a fine amplifier should decline regularly with their number. They all should stay under the masking curve of a constant tone. My research

Resources

- [1] Power transformer -T1
For Western Europe: <https://www.ringkerntrafo.nl/shop/high-end-audio/voedingstrafo-s/719/4n1885-voedingstrafo-voor-buizenversterkers-103va.html>
For Nafta: Plitron 13474-X0-00, www.plitron.com
For Eastern Europe: Trafco 4T1885, www.trafco.rs
- [2] TentLabs, <http://tentlabs.com/Components/Tubeamp/Tubefilament/index.html>
- [3] Linear Audio, www.linearaudio.nl
- [4] F. Blöbbaum, "Multiple Transconductance Amplifiers Part 1," *Linear Audio*, Volume 6; ISBN: 978-949092-9077, www.linearaudio.net
- [5] B. Hofer (Audio Precision), "Master Class about Quality of Components," <https://youtu.be/HsPNGVwCrhw>
- [6] J. Prymac, "Capacitor EDA models with Compensations for Frequency, Temperature and DC Bias," 2010 Capacitor and Resistor Technology Symposium (CARTS) Conference, March 2010.
- [7] AudioMagic, www.audiomagic.nl

does not support that rule. Lowering the second harmonic (as in Amplifier 4 and Amplifier 5) below the level of the third largely improves the hearing of details without effecting the embracing character of the sound. I surely agree that higher harmonics (above the third) should decline regularly. Personally, I consider the second harmonic to be a warm "curtain" creator.

More and more I give my tube amplifiers a damping factor below 10. Higher values by means of overall negative feedback give the amplifier a "flat" character. Then the sound stays close to the speakers and does not freely vibrate everywhere in the listening room.

Testing your amp at a 1 W level and below is a sensible thing to do. But please don't forget to measure the power bandwidth up to the highest frequencies to detect presence of slew rate limiting induced distortion. Full power bandwidth up to 20 kHz stays mandatory. 

Author's Note: Teaching students really is a joy. The greatest reward is to see the gladness in their

eyes. Many thanks to Peter van Willenswaard^[7] for proofreading this manuscript and sharing his valuable experience. My wife Annemieke made a fantastic lunch for every TubeSociety meeting—a real feast. And I want to praise my students, who constructed such high-quality amplifiers!



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Comparing Output Waves of Tubes, FETs, and BJTs

In most music playback systems not using horn speakers, the amplifier is driven into distortion at least some of the time, even though the distortion may only last for a few cycles of a transient, and thus, not be consciously perceived as distortion.

By
Richard Honeycutt
(United States)

In the March 2018 Hollow-State article, we examined the characteristic curves for vacuum triodes and pentodes and field-effect transistors

(FETs) and bipolar junction transistors (BJTs), with a view toward understanding why they have different distortion signatures. While it is true that input signal level, gain, output loading, and biasing all affect the distortion produced by an amplifier stage, we can still note trends in the ratio of odd-order to even-order distortion, and in the amount of intermodulation distortion (IMD) produced by a stage in comparison with the harmonic distortion components.

When we input a sine wave to an amplifier stage and the output is distorted, if we notice asymmetrical distortion (positive half-cycle distorted more, less, or in a different way than the negative half-cycle), we can conclude that the harmonics are predominantly of even order. Even-order harmonic distortion tends to have a soft or blanketed sound; whereas, odd-order distortion sounds razzier.

Triodes

Many music listeners contend that some audiophiles' preference for triodes rather than pentodes, FETs, or BJTs stems from the

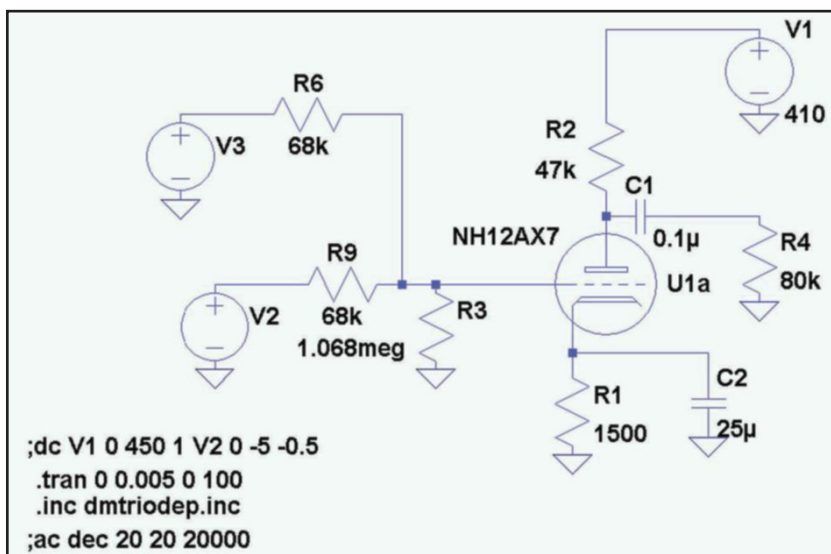


Figure 1: This LTSpice model allows us to look at harmonic distortion and IMD for a single-triode stage.

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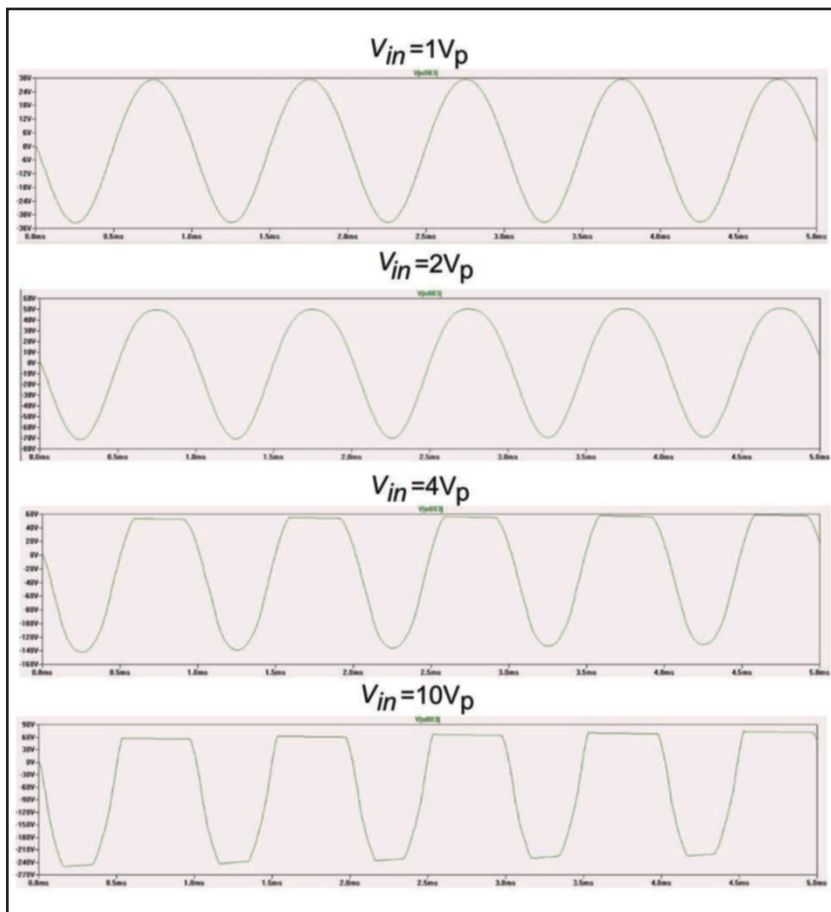


Figure 2: The development of harmonic distortion can be seen as the input voltage is increased.

predominantly even-order distortion produced by triodes when the amplifier is driven into distortion. One of the advantages of SPICE-based circuit simulators such as LTSPICE is that they enable us to model single stages or complete amplifiers and test them for harmonic distortion with sine-wave inputs, and for IMD with two-tone inputs. **Figure 1** shows an LTSPICE model of a preamplifier stage using one triode of a 12AX7. The stage is set up for a two-tone input, and the levels and frequencies of the tones can be set as desired, or we can choose to use only one input signal.

Figure 2 shows the output wave from the triode stage, using various input voltages. With a 1 V input, the output is a near-perfect sine wave. When we increase the input voltage to 2 V, we can barely make out a slight compression of the positive peaks, indicating the onset of even-order harmonic distortion.

The distortion increases as we raise the input voltage to 4 V, at which point the positive peaks are clipped, making the asymmetrical distortion more obvious. With a 10 V input signal, both peaks are clipped, although the positive peak is clipped more.

The top curve in **Figure 3** shows the output wave for a two-tone input: one 10 V_p sine wave at 1 kHz and another at 1.6 kHz. In the lower curve, which is the Fast Fourier Transform (FFT) of the upper curve, the first peak at the left represents the 600-Hz difference between the two input tones' frequencies, and has an amplitude of about 29 dBV. The second and third peaks represent

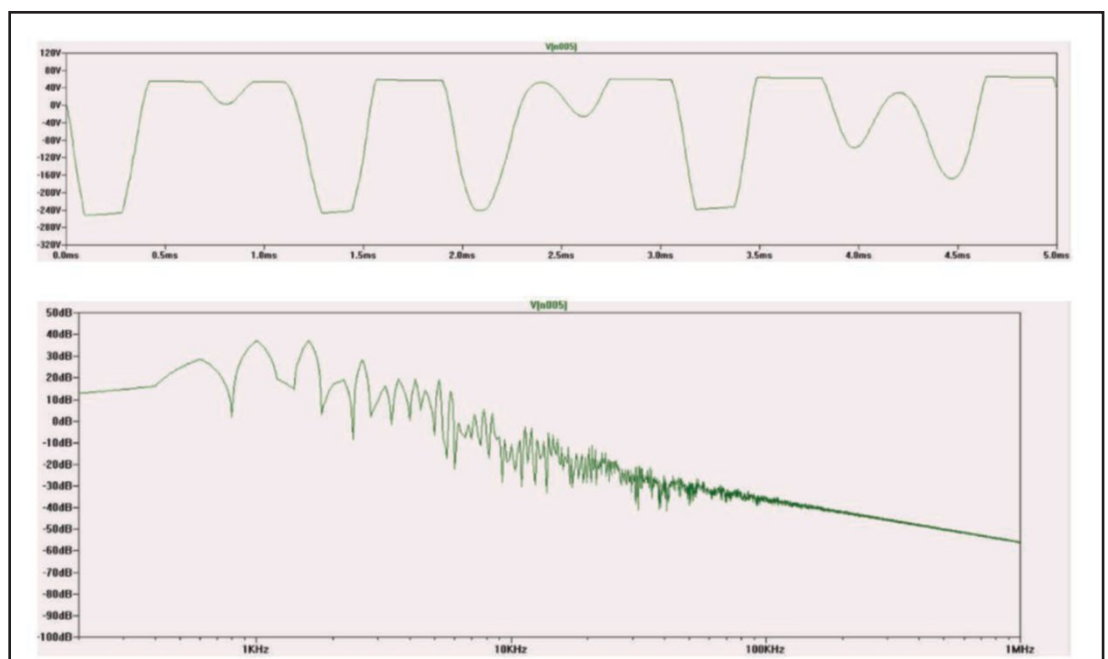


Figure 3: FFT Analysis shows the harmonic and IMD components of the output wave with a 2-tone input.

Sound thinking in coaxial design

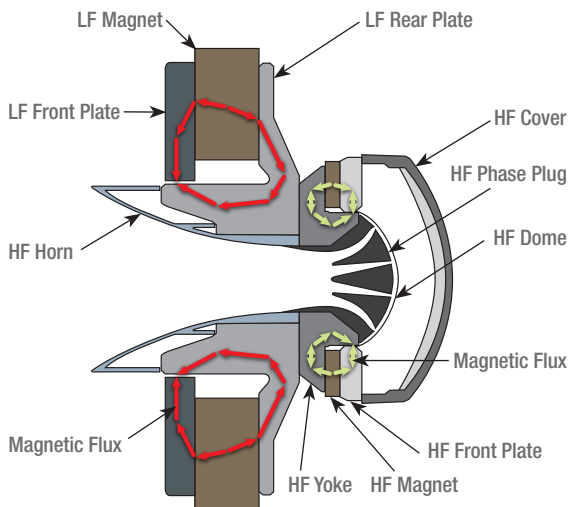
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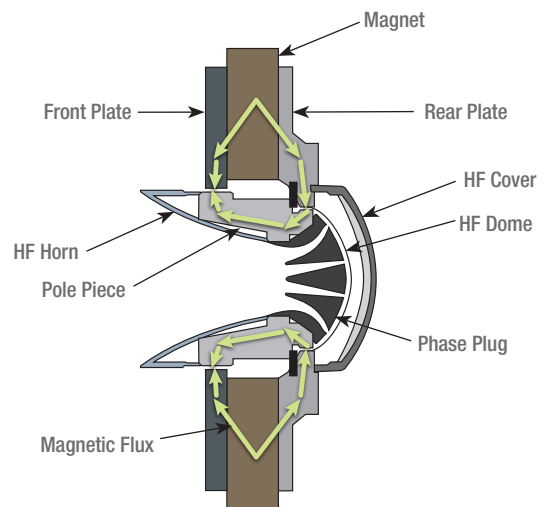
The magnetic flux is optimally distributed between the two elements (calculated using FEA), to achieve the best balance between HF and LF response (see diagrams below). Additionally, the use of a single magnet assembly also means lighter weight and a more compact profile, compared to more conventional dual motor designs.

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CELESTION

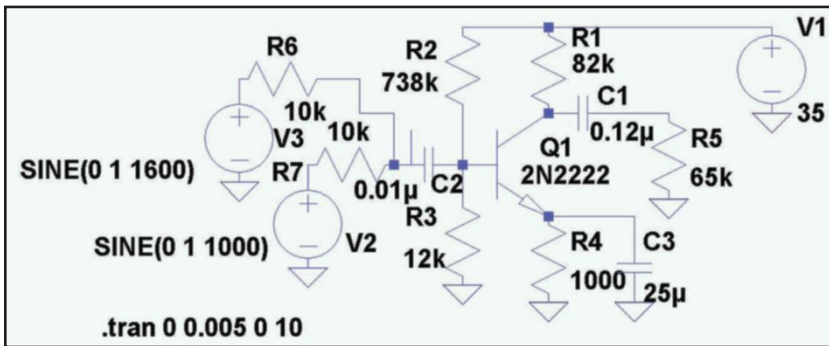


Figure 4: This is an LTspice model of a BJT preamp stage.

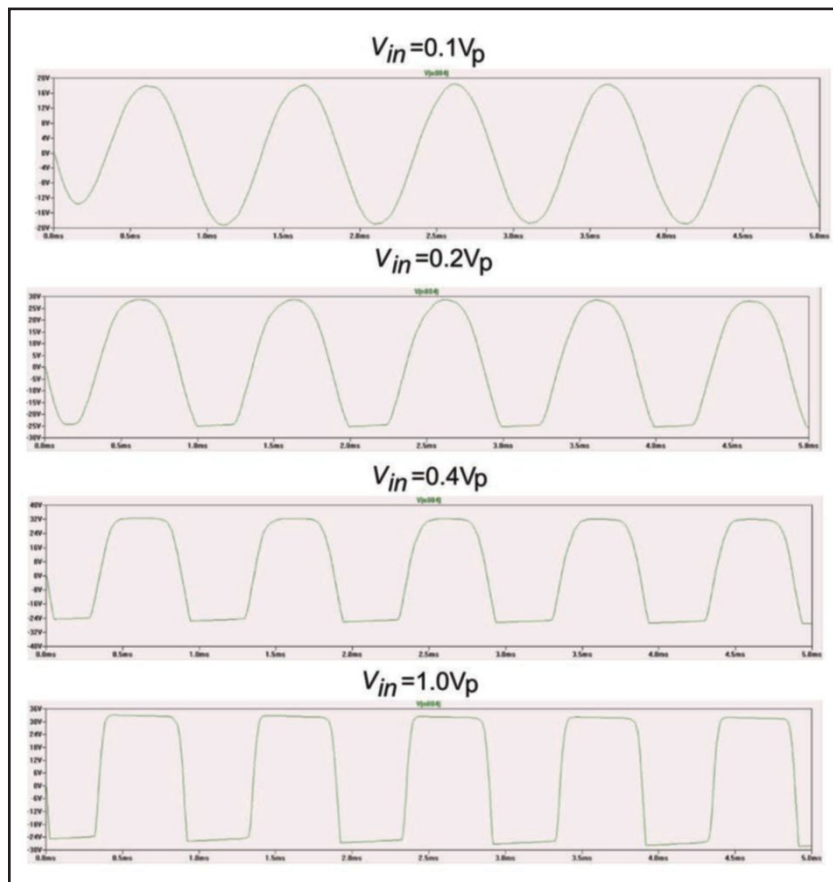


Figure 5: The output of the BJT circuit clips more abruptly than does the triode circuit.

the input frequencies, with amplitudes of about 35 dBV. The fourth peak is the second harmonic of the 1 kHz input wave, and has an amplitude of about 18 dBV. The fifth peak represents the sum of the input frequencies, at about 28 dBV. The sixth peak is the third harmonic of the 1 kHz input wave, with an amplitude of about 16 dBV. The seventh peak is the second harmonic of the 1.6-kHz input wave, at about 17 dBV.

Higher-order harmonics and IMD products can be traced, with decreasing amplitudes, as you follow the curve to higher frequencies. Note that in this case, the IMD components are only a few decibels below the levels of the input waves. (If we approximate the IMD components' amplitudes as 28 dBV each, the two would power-sum to $28 + 3 \text{ dBV} = 31 \text{ dBV}$, which is 5 dB below the levels of the 1 kHz and 1.6 kHz components.)

The harmonics are at a substantially lower level: 17 dBV to 18 dBV for the second harmonic, versus 12 dBV to 16 dBV for the third harmonics. Thus as we would expect for a triode stage, the second harmonic is several decibels greater than the third. If we follow the sequence of peaks, we will find that both the second and third harmonics' amplitudes decrease as frequency goes up. In fact, the amplitude of the distortion components decreases at a rate of about 25 to 30 dB/decade.

BJTs

Figure 4 shows a BJT preamp stage. Like the triode stage of **Figure 1**, the BJT stage has no negative feedback: The triode's cathode resistor and the BJT's source resistor are bypassed. In this simple circuit, a BJT begins distorting at a much lower input voltage. Thus, the waveforms shown in **Figure 5** represent outputs when the peak input voltages are 0.1 V, 0.2 V, 0.4 V, and 1 V, unlike the input voltages for the triode circuit, which are 10 times higher.

With the lowest input voltage tested (1 V for the triode, and 0.1 V for the BJT), both the triode and the BJT produce clean outputs. But when the input is raised to twice that level, the BJT clips the negative half-cycle.

Although the triode squashes the positive half-cycle a bit at twice the lowest input voltage, it does not clip until four times the lowest input voltage is reached. The triode clips the positive half-cycle, whereas the BJT clips the negative one. (Whether the positive or negative half-cycle is distorted first depends upon the AC load resistance connected to the plate, collector, or drain of the active device.)

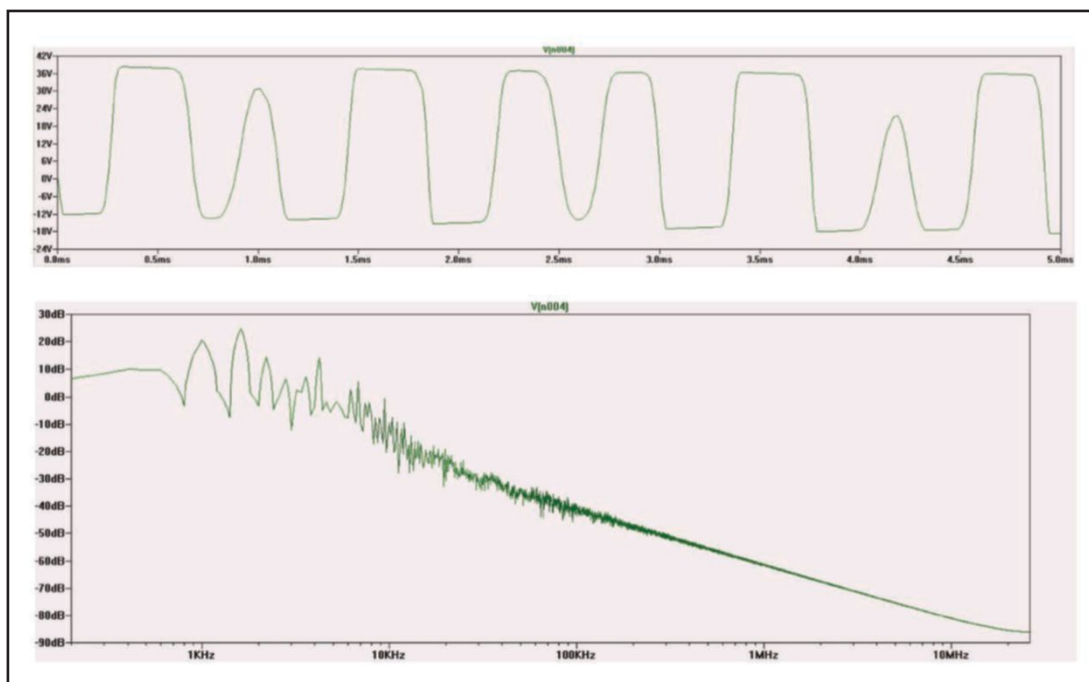


Figure 6: The output wave and the FFT results for the BJT stage are different from those of the triode stage.

Many arguments have been engaged over whether this difference is audible. Both the triode and the BJT are seen to clip asymmetrically indicating a predominance of even harmonics.

When we apply a two-tone input signal—1 V at 1 kHz and 1 V at 1.6 kHz—we get the waveform shown at the top in **Figure 6**. Comparing the shape of the FFT of this waveform (lower part of Figure 6) shows higher levels of third and fourth harmonics from the BJT circuit. (It is important to compare the shapes of the FFT curves, not the decibel values of the different frequency components, since the input voltages—and hence the output voltages—are very different.) The level of the distortion components decreases by about 23 dB/decade as frequency increases, indicating that the distortion of the BJT would be more razzy and less smooth than that of the triode.

JFETs

Figure 7 shows the output waves for inputs of 0.1 V, 0.2 V, 0.4 V, and 1 V. With a 0.1 V input, the JFET stage is just beginning to produce some asymmetrical distortion. The schematic of a JFET preamp stage is shown in **Figure 8**, and it also has no negative feedback. To provide a range of distortion similar to that of the other two circuits, the same input voltages were used as for the BJT stage. As with the BJT, the JFET distorts more on the negative half-cycle. With a 1 V input, the positive half-cycle is beginning to approach hard clipping, as was the tube triode with a 4 V input.

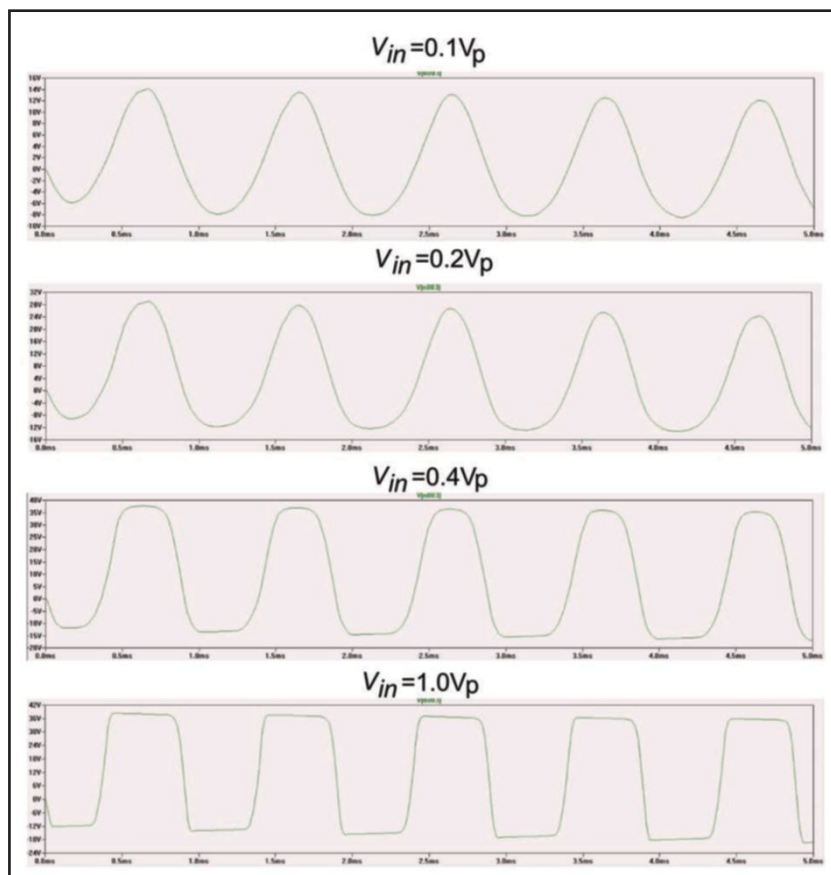


Figure 7: The output waves for a JFET circuit clearly differ from those for a triode tube or BJT circuit.

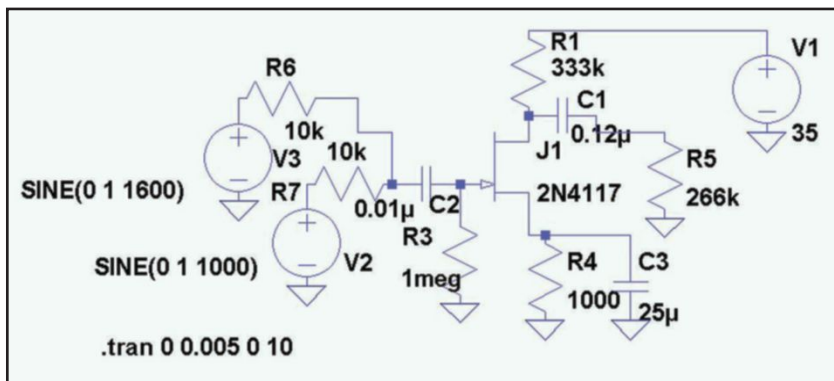


Figure 8: This LTSpice model of a single-stage JFET preamplifier circuit produces somewhat different output waves than the tube circuit did.

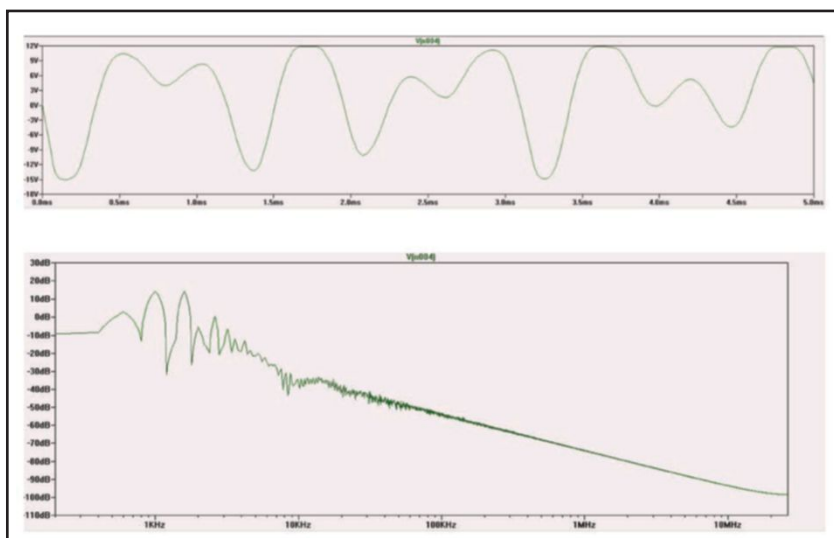


Figure 9: The output wave and FFT analysis are quite different for the JFET, compared to the vacuum triode or BJT.

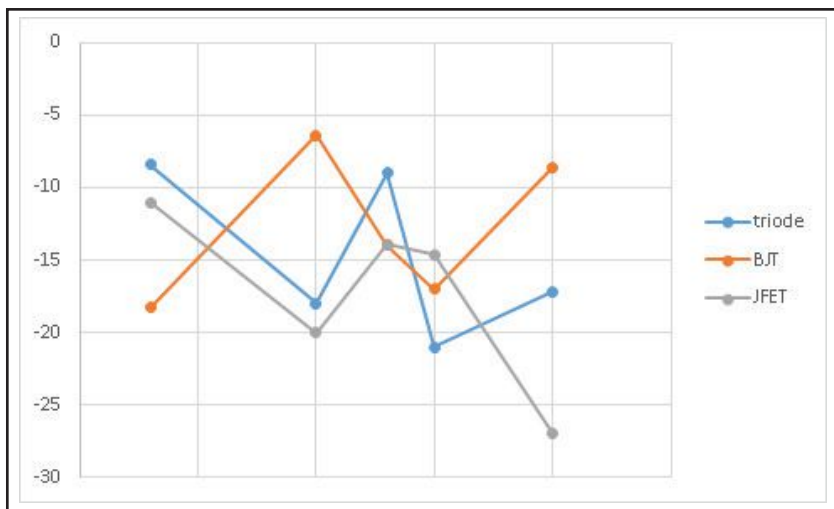


Figure 10: This graph compares the IMD and harmonic distortion components of the triode, BJT, and JFET stages.

Applying a two-tone input to the JFET stage, we get the output wave and the FFT trace shown in **Figure 9**.

The Results

To make it easier to compare the behavior of the three circuits, I have graphed the levels of the distortion components for each (see **Figure 10**). The plot points, from left to right, represent the IMD difference frequency, the second harmonic amplitude of the 1 kHz wave, the IMD sum frequency, the third harmonic of the 1-kHz wave, and the fourth harmonic of the 1-kHz wave. The levels of the frequency components are normalized to the input voltages, so the “-decibel” values indicate how many decibels below the level of the fundamental frequencies each distortion component is.

As indicated earlier, the triode stage has more IMD than either of the transistor stages. The BJT stage has higher second-harmonic levels and lower third-harmonic levels than the triode stage. The three lowest-frequency JFET distortion components pretty much track those of the triode tube. The JFET has proportionately more third harmonic. But next comes what I suspect may be a surprise: The fourth-harmonic level from the JFET stage is even lower than that of the triode tube.

It is certainly true that the biasing of the active devices and the signal levels affect the distortion, as would the presence of negative feedback. But the abrupt change from peak compression to clipping, and the greater degree of clipping for the highest input levels in the BJT compared to the triode stage would produce sharper corners of the clipped waveform. This wave shape is associated with more high-frequency distortion components. (Compare the three stages’ output waveforms from the two-tone test.)

In a complete amplifier, there are not only single-ended stages such as these, but there also may be balanced (perhaps Class AB or B) driver and output stages as well. The distortion signatures of these stages is different from that of a single-ended stage. And the gain structure of the amplifier determines which stage gets overdriven first, another factor affecting the sound of an amplifier when driven into distortion.

Editor’s Note: All audioXpress articles from 2001 to present can be found on the aX Cache, a USB drive available from www.cc-webshop.com.

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Here are a few places where you might find a copy of *audioXpress* and possibly meet one of our authors and staff members.

May 4–6, 2018

Lonestar Audiofest 2018

The Embassy Suites Dallas Park Central, Dallas, TX

www.lonestaraudiofest.com



While US Hi-Fi shows have been gradually more commercial and professionally organized, a few other events for audio enthusiasts refuse to follow trends and remain laid back and faithful to their roots of camaraderie and self-supported. Intended to be informal and non-profit, and targeting DIY experts as well as established manufacturers, Lonestar Audiofest has been consistently small and focused. There are a few seminars confirmed already and many of the Texas-based brands and builders are promising a strong showing. Great reasons to attend.

May 7–11, 2018

175th Meeting of the Acoustical Society of America

Hyatt Regency Minneapolis, Minneapolis, MN
acousticalsociety.org



The 175th Meeting of the Acoustical Society of America (ASA) will be held Monday through Friday, May 7–11, 2018. A block of rooms has been reserved at the Hyatt Regency Minneapolis at discounted rates. Information about the upcoming meeting is now available on the new ASA webpage at AcousticalSociety.org. The Exhibition will include computer-based instrumentation, scientific books, sound level meters, sound intensity systems, signal processing systems, devices for noise control and acoustical materials, active noise control systems, and other exhibits on acoustics. An instrument and equipment exhibition will be located near the registration area and meeting rooms.

Technical Committees will hold open meetings on Tuesday, Wednesday, and Thursday evenings. A “Hot Topics” session sponsored by the Tutorials, Short Courses, and Hot Topics Committee will cover the fields of Engineering Acoustics, Signal Processing in Acoustics, and Structural Acoustics and Vibration. A tour of the recently completed U.S. Bank Stadium, home to the Minnesota Vikings, will be held during the week.

May 10–13, 2018

HIGH END 2018

MOC, Munich, Germany
www.highendsociety.de



Already set to be the largest gathering of high-quality audio experiences in the world, the Munich High End show, will be once again the meeting place for all those interested in the latest in consumer audio. As one of the best-organized consumer audio shows in the world, visitors will find all kinds of products related to high-quality consumer electronics as well as a unique variety of services at Munich’s MOC Event Center.

This year, the internationally acclaimed exhibition will mark its 37th consecutive staging, making it Europe’s oldest trade fair of its kind. More than 500 exhibitors from nearly 50 countries will offer a complete market overview, with systems for those just getting started, centrally-controlled multi-room systems, headphones, turntables and, of course, the ultimate peak of technical perfection.

The 2018 edition of High End Munich will also be the first in partnership with CanJam Munich 2018. On the public days of High End 2018 (not on the trade visitor day), the CanJam headphone show will be staged concurrently in a location next to the MOC. This exhibition caters to consumers having a special interest in headphones and mobile audio devices. That is why the admission tickets and trade visitor passes for the High End 2018 are also valid for the CanJam Munich 2018. The venue for this special trade show is the so-called “Kohlebunker” (coal bunker) diagonally across from the MOC.

In addition to the wide array of innovative products, the event will once again feature an abundantly diverse special events program consisting of live music, workshops, lectures, and music presentations.

May 23–26, 2018

144th Audio Engineering Society (AES) Convention Milan 2018

NH Hotel Milano Congress Centre, Milan, Italy

www.aes.org/events/144



The Audio Engineering Society (AES) has confirmed the 144th International AES Convention and Exhibition program, this year taking place in Milan, Italy. The convention will cover the full spectrum of professional audio topics including 3D Audio, Sound Reinforcement, Sound for Cinema, Sound for VR, Multichannel Audio, Recording and Production, and Room Acoustics.

Embracing this year’s theme—“The Power of Sound”—the convention will be co-chaired by Alberto Pinto (Professor, CESMA, NUCT University, Stanford University) and Nadja Wallaszkovits (Chief Audio Engineer, Phonogrammarchiv, Austrian Academy of Sciences and AES President-Elect), who have assembled a team of leading audio professionals to organize the events and presentations for Europe’s largest professional audio education, networking, and gear exhibition of the year. The four-day technical program at AES Milan will run May 23–26, with the exhibition taking place May 24–26. The technical program sessions include more than 160 research paper presentations, covering the latest in audio research, technologies and techniques aimed at educating audio professionals and students, alike.

The companion exhibition will showcase leading manufacturers with the latest gear on-hand for demonstration and evaluation, along with educational presentations and panels taking place on the show floor. The expo will also feature dedicated demo rooms and other unique listening experiences, complementing the full schedule of special events and social opportunities going on throughout the show.

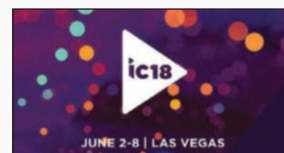
Malcolm Hawksford (Emeritus Professor within the School of Computing Science and Electronic Engineering at Essex University, Colchester, and a recent AES Gold Medal Winner) will present the 38th Richard C. Heyser Memorial Lecture, speaking on the topic “Understanding High Quality Audio – A Personal Journey.”

June 2–8, 2018

InfoComm 2018

Las Vegas Convention Center (North and Central halls), Las Vegas, NV
Show Dates: June 6–8

www.infocommshow.org



InfoComm organizers, AVIXA (the Audiovisual and Integrated Experience Association, www.avixa.org) have announced new initiatives, content program, and exhibition for InfoComm 2018, which returns to the Las Vegas Convention Center June 2–8, 2018. The show will kick off on June 5, with the opening reception at OMNIA Nightclub Las Vegas.

The AV Tech Tours will take visitors to view tech marvels, such as the nightclubs at MGM and Aria. There are more than 80 new education sessions, training, and workshops, from the Emerging Trends Day to the new Integrated Life @ InfoComm 2018 (www.parksassociates.com/integratedlife2018), a special track of workshops and sessions, and a dedicated area on the InfoComm 2018 show floor, for spotlighting smart building, smart home, mobile, security, and related audiovisual solutions that create integrated experiences wherever people use technology.

AVIXA also announced that developers of C2 Montréal, a groundbreaking business conference merging commerce and creativity, will bring their innovative brand of immersion and collaboration to the InfoComm show floor.



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