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Myths in Tube Circuit Designs
By Peter Dieleman

Hollow-State Electronics

Down Memory Lane Electric Guitar Amps
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Fresh From the Bench

The Hypex NCORE NCx500 Amplifier Module
Doubling Down on a Flagship

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Vintage Op-Amp Noise Simulation — Part 2
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Power Transformer Parameters, Selection, and Testing Part 7 — Transformer Cores
By Chuck Hansen

You Can DIY!

Monoblock Power Amplifier with Precision Bias Control Part 2 — Construction and Performance
By Bruce E. Gillingham

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Make It Worth It!

In this Glass Audio Special issue of *audioXpress* we are proud to offer two high-quality tube amplifier projects from two renowned tube experts from both sides of the Atlantic. The first project by Menno van der Veen is another amplifier designed under the auspices of the author's TubeSociety initiative. The other comes from Thomas Perazella and in this case details a remake of a famous David Berning ZOTL amplifier. Both are of remarkable quality and value to all tube/valve enthusiasts at a time when it becomes increasingly more risky (and expensive) to build your own projects, unless we already know that the results are truly exceptional.

We are also privileged to be able to share the knowledge of renowned tube audio expert and book author Peter Dieleman as he demystifies some common myths in tube circuit designs. To have original articles of this level and quality on tube design is getting to be increasingly difficult even though the enthusiasm around tubes has not diminished in any way.

A year ago in this May issue I was sounding the alarms for the dramatic supply chain constraints affecting the availability of new tubes, as the invasion of Ukraine and the sanctions imposed on Russia had a significant impact on the global supply of raw materials and packaged tubes to the audio industry. There are very few tube manufacturers left, and those remain constrained by the supply of parts and materials. A year later, those few suppliers still active have been able to re-establish barely sustainable stock levels but, as predicted, with significant price increases. So, it's not surprising that we see fewer new builds being submitted for publication.

It's increasingly not worth it to build another average amplifier when it has become painfully obvious that once we burn through existing new-old stock (NOS), it will never be possible to replace the tubes. Some might say that there are still plenty of tubes to last decades of keeping up with a fun and rewarding hobby—and that might be true. But maybe the results of such efforts are better delivered by following designs with recognized merits.

In *audioXpress* we review and evaluate projects that involve tubes on a regular basis, but increasingly few are of interest for publication since they are fundamentally a review of designs that have previously been addressed and published. A word of appreciation is due to our technical editor Jan Didden for the diligence and persistence to go through that review process and have a much greater perspective of the rich history of this publication to find valuable concepts and ideas.

Alternatively, as found in this issue, you might as well pursue a rewarding build of the extraordinary high-power (300W into 4Ω), bipolar, Class-AB, monoblock audio power amplifier that Bruce E. Gillingham concludes this month with details of the complete system, featuring an original precision Bias Controller, and its performance measurements.

And if objective, measurable audio quality is truly the goal, consider jumping directly to Class-D and the Hypex NCx500 amplifier module that Stuart Yaniger reviews. As his measurement and analysis in this issue reinforce, this is a new reference performer that just needs a power supply, wiring, and your choice of chassis. The result is so close to perfection that one would likely degrade performance to add some originality or color—maybe by combining a tube preamp stage?

If it is of any consolation for tube audio enthusiasts, neither transistors nor the latest Class-D modules are easily sourced and manufactured these days. Many of the most popular devices used in audio amplifiers have long been discontinued and are increasingly difficult to find in consistent quantities. Even from the larger semiconductor brands, it's important to know what is actually still being manufactured and what is simply available from existing stocks.

Given the remaining component supply challenges and manufacturing delays that resulted from the global pandemic, we are now seeing the products announced precisely a year ago finally becoming available. Most of the Class-D amp module manufacturers still offer six months or more lead times for volume orders.

So, whatever the chosen path, make your design truly worth it.

J. Martins
Editor-in-Chief

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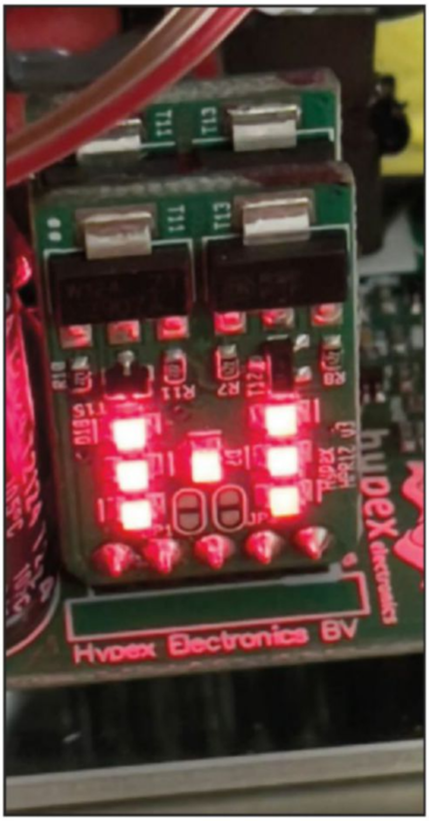

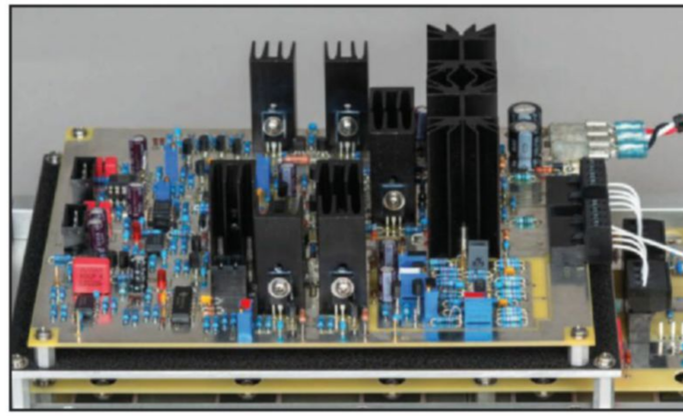

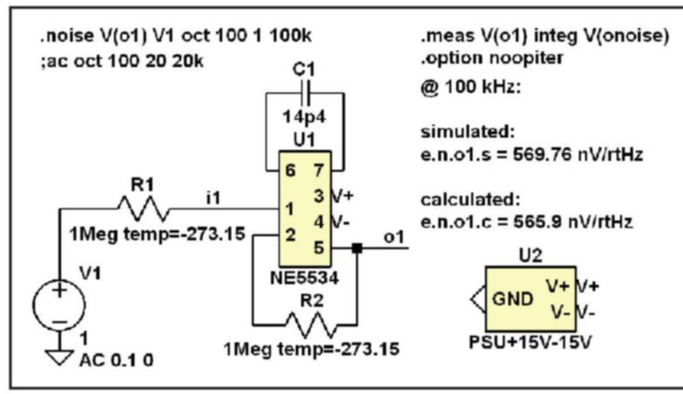

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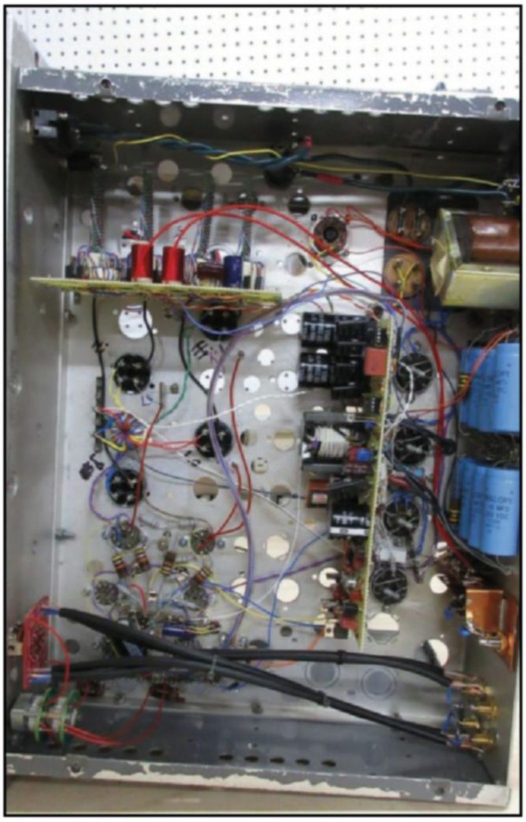

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The Hypex NCORE NCx500 Amplifier Module

Doubling Down on a Flasghip

By
Stuart Yaniger

NCORE (NC) amplifier modules are the flagship product from Dutch manufacturer Hypex Electronics and its success among the high-end audio brands is well documented. In 2022, Hypex announced the introduction of its improved NCOREx technology, starting with the NCx500 OEM module, which *audioXpress* received for review.

Audio power amplification hit an asymptote some years back and we're now on the flat part of the quality curve. We long ago passed the point of distortion reduction to inaudible levels. The same thing occurred with frequency response deviations due to reactive loudspeaker loads interacting with amplifier output impedance. The next hurdle was size, weight, and heat, and the advent of mature Class-D designs has taken us much closer to perfection from both performance and efficiency points of view. Although Class-D amps have been around for quite a while, the technology reached the point where the amps had performance levels competitive with traditional Class-A/AB amps only about 10 to 15 years ago.

A renaissance occurred, and as is often true, it seemed to happen all in one place—northern Europe, which has been a traditional home for audio innovation. One name that keeps popping up is Bruno Putzeys who has been an integral part of Hypex, Purifi, and Kii, and originated the first practical self-oscillating Class-D amplifiers, which kick-started the rapid evolution of the genre. The various companies selling these products have been competing to see how flat, quiet, and linear they can make their amps. After having several of the top brands pass through here whose performance exceeded that of my test gear, I have been of two minds regarding further development: Does it really matter once we have comfortably exceeded any questions of audibility? On the other hand, isn't it desirable to have absolutely the best performance that can be engineered into a product if it can be done without compromising practicality (e.g., cost, size, reliability, and efficiency)?

The Hypex NCx500 is the latest missile launched in the Power Amplifier Performance War.

"What's the Motivation?"

Traditionally, if you wanted to go into the high-performance audio electronics business, you could start from scratch and design a circuit, design the circuit boards, the case, the power supply, and the wiring, then either build a final product or have it assembled to your specifications. In years past, designers learned to shortcut the process by building so-called chip amps (notably the various Gainclone amps), which were based on hybrid integrated circuit (IC) power amps; add heatsinking, a power supply, I/O connections, and relatively simple surrounding circuitry, drop them in a nice case, and assuming basic designer competence in layout and grounding, you could have a very good performer. It won't be state of the art, but it will be a competent performer. And it's not design, really, but integration, and differences in performance between integrators can still be significant.

In the wonderful modern era, one can still buy power ICs suitable for constructing chip amps, but for really serious performance goals and high power, the integrator can now opt for Class-D modules, of which there is now a very broad choice. They are somewhat larger than their IC predecessors but still quite small. And for a fair comparison, we have to look beyond the size of the amplification itself (chip vs. Class-D module). There's a physical limitation because making a device smaller also makes it more difficult to pull heat out of it, and thermal dissipation requirements means that the actual size is chip plus support circuitry plus heatsink versus Class-D module plus its heatsink. Class D, of course, has much higher efficiencies, so the heatsinking requirements are less onerous.

Of course, size reduction has its limits. With marginal gains in efficiency at higher powers, the challenge becomes heat removal.

Even with a 92% efficiency (and that's an optimistic number), a 200W amplifier will still have to dissipate 16W, which is easy when there's enough radiating area, but harder if the amp were, say, the size of a postage stamp. But if it's the size of a playing card, this is certainly small enough for any practical purpose and sufficient area to be able to draw out the heat without resorting to fancy tricks. More about this in a bit.

Integrators have the option of incorporating them into power amplifiers, where, in a manner analogous to the linear amplifier-based chip amps, a complete product can be made with the addition of I/O, power, and heatsinking, or thanks to the low emissions and simple thermal management requirements, can be included in more all-in-one packages such as AV receivers or streaming-based amplifiers. Examples of the former include power amplifiers from March Audio, Buckeye Audio, and BoxEm. An example of the latter is the NAD M10 streaming amplifier that I use in my main stereo system. High-performance modules for integrators are sold by, among others, Hypex, Orchard, Purifi, ICEpower, and Pascal.

And with this level of efficiency, compactness, and low noise, the modules can easily be integrated into active loudspeakers.

The NCx Products

Among the high-end options, the Hypex modules are compact, high performance, nearly turnkey solutions to power amplification. As a result, they have been quite popular, and for good reason: Their performance is excellent, they have a good reputation for reliability, and they tend to have low electromagnetic interference (EMI) radiation.

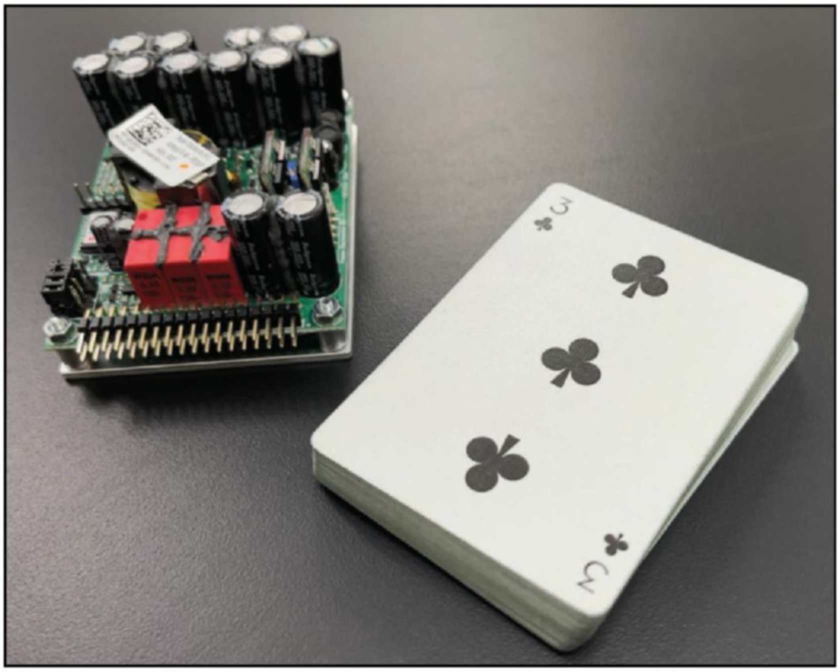


Photo 1: The NCx500 power amplifier module is built on a compact circuit board bolted to a solid metal heat spreader.

Hypex has several series of available modules for different performance levels and applications. The NCORE NC and now the NCx series are its flagships, with ridiculously low distortion, low audible noise, and high power delivery. From a design engineering point of view, they feature a self-oscillating loop to generate the carrier, extremely high levels of negative feedback while maintaining stability, and proprietary gate-drive circuitry to minimize dead time and distortion.

The new "x" series squeezes in another 6dB of feedback by increasing open-loop gain (and presumably adjusting compensation to retain stability), which in theory will lower the distortion and source impedance by a factor of two compared to the previous NC series.

Frank Veldman of Hypex explained, "Both NCORE and NCOREx revolve around a way in which the distortion of the power stage is isolated, amplified, phase reversed, and added to the original input signal fed into the UcD like power stage. Differences between NCORE and NCOREx are in the stage where the isolated distortion is amplified. This section has been improved where we have now been able to add 6dB to 20dB of gain (frequency dependent) without sacrificing stability. This improvement has been put to use not just for better overall distortion figures, but also to counter any increase in output stage open-loop distortion originating from the improved efficiency we were after."

Overall feedback is taken at the speaker output to minimize any distortion contribution from the output filter's series inductor. Both input and output are run as balanced differential to reduce the effects of any ground loops or common-mode noise pickup.

Hypex also states that it upgraded the quality of some of its passive components. It's unclear how much that could enhance the performance compared with its implemented changes in feedback, but that could make a small difference and perhaps enhance reliability as well.

The NCx500 module is indeed about the size of a playing card (**Photo 1**), and the circuit board and components are bonded to a solid piece of metal plate that can be used as a heatsink at lower powers, but use of the modules at higher powers and with lower impedance loads will require external heatsinking to limit the temperature rise to safe levels. The metal plate is really meant to provide an efficient thermal interface to an external heatsink.

For my testing and ultimate amplifier fabrication, I bolted the modules to a small but solid finned heatsink (**Photo 2**), and even with extended high power use and testing, the module was only mildly warm to the touch (perhaps 45°C). Hypex published an application note detailing thermal calculations to determine adequate heatsinking—it's interesting that the module-to-heatsink thermal resistance can usually be neglected because of the size and efficiency of the module's metal plate for thermal coupling. In many applications, it will be sufficient to bolt the modules to the amplifier enclosure and use that as a heatsink.

I/O to the module is accomplished through a 36-pin wire-to-board crimp connector usually used for ribbon cables. Alternately, Hypex sells an evaluation board that plugs directly into the module connector and allows rapid connection of the power supply and the signal I/O connections. For my evaluation, I used a ribbon cable so that various provided non-audio functions could be accessed and to give me flexibility once I did a more permanent installation and build.

Besides power and signal connections, that I/O connector has provisions for remote feedback in the manner of a Kelvin connection, with separate feedback leads for hot and cold. The intention is to compensate for the resistance of the connector potentially increasing the source resistance (i.e., lowering damping factor) by connecting the feedback leads to the module outputs on the far side of the 36-pin connector.

Although one might be intrigued by the idea of taking the feedback from a point further downstream (e.g., from the speaker terminals, bypassing the speaker cables), Hypex cautions that the feedback leads be limited to a few centimeters. "Using FBH/FBC to sense remotely (e.g., at the end of a speaker cable) does not work well."

The predecessor NC500 modules did have an interesting and little-used feature: a current monitoring output (0.1V/A) derived from a 100mΩ resistor in series with the hot output lead. This could be used to detect high current conditions, which could threaten the safe operating area of the output devices, or potentially to provide a series-derived voltage for feedback that would increase the output impedance if desired, or even make the amp act as a current source rather than a voltage source. Unfortunately, this feature was not used much and was removed from the NCx500. Frank Veldman points out that the resistor and monitoring circuitry could still be added by the integrator, and if an OEM purchaser wanted that feature back, it could easily be implemented.

For ease of integration, the modules now have a built-in preamp stage (the word "buffer" is used a lot, but is not exactly accurate), which accepts a balanced audio input and raises the module's rated voltage gain from 12.4dB (4.2V/V) that was standard in the NC500 to 27dB (22.4V/V), the latter mode eliminating the need for an external preamplifier. The preamp stages are powered via discrete regulator modules, which are also available as standalone regulators, the HPR-12 and the HNR-12 (+12V and -12V, respectively, **Photo 3**). The regulator modules use an array of red LEDs, presumably as references—I have used LEDs in the past and have found that good ones have exceptionally low noise

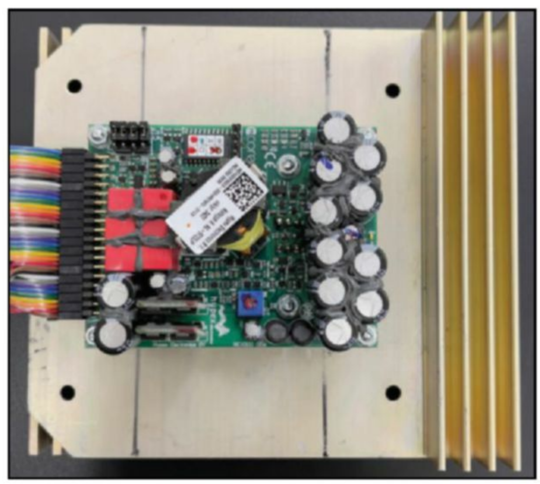


Photo 2: To effectively dissipate heat from extended high power operation, the module should be bolted to a heatsink. The one shown here was probably overkill

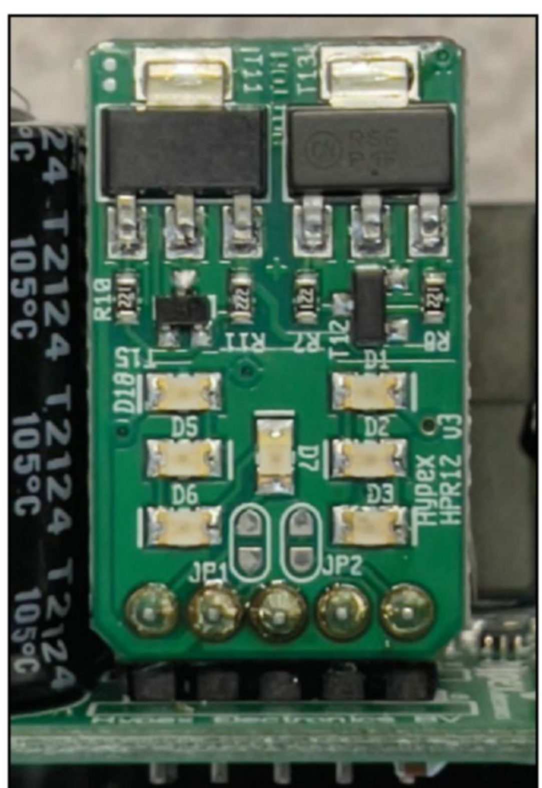


Photo 3: The HPR-12 and the HNR-12 discrete voltage regulator modules provide low noise, highly regulated voltage to the preamp stage.

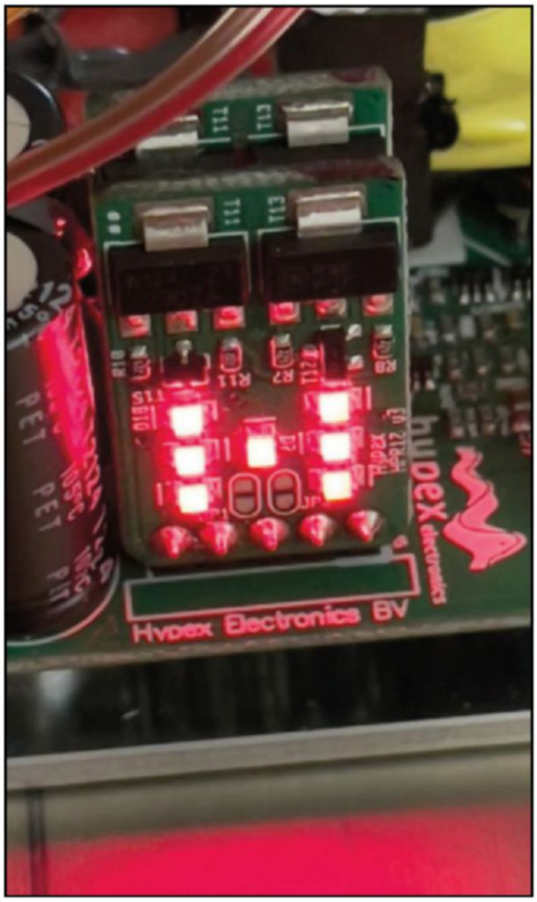


Photo 4: The LED references on the regulators give a jolly glow and look suspiciously like the company's monogram.

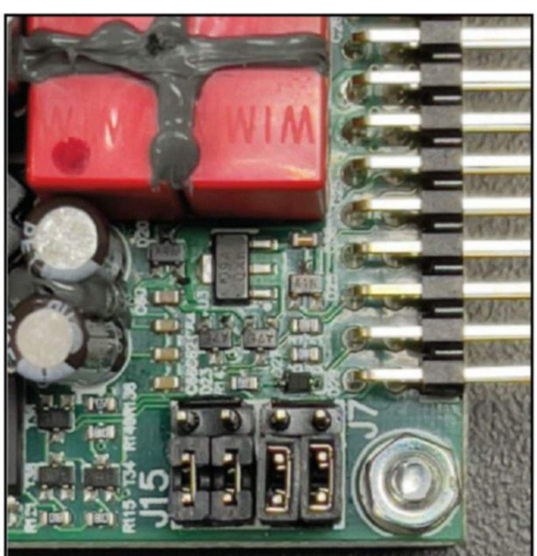
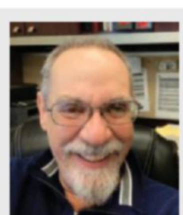


Photo 5: Gain and input impedance can be set by moving four jumpers, a new feature of the NCx series.

About the Author



Stuart Yaniger has been designing and building audio equipment for nearly half a century, and currently runs a technology consulting agency in western New York. His professional research interests have spanned theoretical physics, electronics, chemistry, spectroscopy, aerospace, biology, and sensory science. One day, he will figure out what he would like to be when he grows up.

and excellent voltage stability. One might notice that the brightly illuminated array is shaped like a capital H (**Photo 4**), but Hypex denies that this was deliberate branding.

Of course, an integrator might opt for using a lower gain option by resetting four jumpers (**Photo 5**), the gain can be reduced back to 12.4dB. To achieve full output from the modules with an 8Ω load, the drive to the module's input needs to be 13V to 14V. So for most common system gain structures in home audio (typically 2V to 2.5V to the power amp to achieve full output), an additional gain stage must be provided. As we will see, it will be tough to achieve higher performance than the included preamp stage, but for marketing and aesthetic reasons, someone might want to use a front end made from vacuum tubes or special field-effect transistors (FETs), though the distortion and noise performance is likely to be degraded. To be fair, if the modules were capable of unity gain, I would be tempted to build a tube or hybrid differential amplifier, or use something like the ImPasse design I published in *audioXpress* about 15 years ago. It might not have the same stunning distortion and noise performance as the Hypex preamps, but it would be good enough to avoid any audible degradation and have a glowing glass charm.

The modules can be operated in two modes, hardware and software, selectable by either grounding or floating one of the I/O connector pins. Software mode allows addressing the module via an I^2C bus using a standard 100kHz data rate. Hardware mode allows the module to act as a standard analog amplifier. There are also pins on the I/O connector to indicate amplifier failures from overload, thermal, or DC offset. The amp may be remotely placed in active or standby as well. Reading register locations, the integrator can also offer temperature and bus voltage monitoring so that additional protection circuitry can be implemented.

Speaking of voltages, the NCx500 requires symmetric bipolar high voltage for the output stage (±35V minimum, with reduced output power capability and ±85V nominal for full power), ±15V at 35mA for the low level circuitry (which is regulated down to ±12V on the NCx500 board as previously mentioned), and +15V referenced to the negative high voltage supply at 75mA to drive the gates of the output FETs. If the integrator doesn't want to design a new supply, these voltages can all be supplied by Hypex's SMPS1200A700 switching supply (**Photo 6**). This supply is good for 1200W short term, and can be connected to the NCx500 module's control lines to allow automatic shutdown in case of excessive DC offset. Additionally, there are lines for amplifier

and power supply standby and enable so that the integrator can design for avoidance of start-up and shut-down pops or other unwanted noises. The supply is mounted in a metal frame, which is significantly larger than the amp module, but it doesn't require significant external heatsinking.

The low-level circuit supply can also be selected via jumpers to put out a regulated ±12V if the NCx500's built-in preamp is not used so that an integrator's external preamp can be powered.

Because the switched-mode power supply (SMPS) is set up for a two-wire mains power connection, the integrator must design the mechanicals to meet Class II double-insulation requirements. As supplied, the power supply's circuit board is spaced 6mm from the metal frame, so that the metal frame can still be bolted to the case to provide extra heatsinking while maintaining the minimum required insulation distance between PCB and case to meet Class II standards.

The NCx500 In Use

Because I literally mounted the amps on a breadboard, doing the setup was slightly tricky. But the balanced topology and high common-mode rejection (CMR) saved the day, and despite the crudity of the setup, the amps were dead quiet with my speakers. They were fed by the preamp output of an NAD M10 streaming amplifier through a single-ended-to-balanced connector. My loudspeakers are a biamped four-way dynamic system based on the NHT 3.3, with electronic crossover at 125Hz, so are moderately sensitive, but play very loudly when fed with enough power. Their impedance runs between 4Ω to 6Ω through most of the audible range.

The built-in nCORE amps in the M10 are usually sufficient, but every once in a while, especially with heavy metal music, I can drive them out of their comfort zone. For example, when I play Sleater-Kinney's "Dig Me Out" at a reasonable clip for that genre, the distortion of Corin Tucker's voice when she turns on the energy is prominent despite the built-in distortion of the source material. Switching over to the Hypex NCx500s, the fuzzy edge on her voice vanishes, leaving behind, in a weird way, clean distortion (i.e., what was actually on the recording). Ditto some big band material—Buddy Rich's take on "Birdland" from the *Class of '78* album played perfectly cleanly while rattling the glass windows and objects sitting on my bookcases. And yet after all that, the amps were still barely warm.

On more moderately demanding music (acoustic jazz, Americana, bluegrass, and classical), the amps sounded exactly like what the measurements indicated, that is, absolutely clean and uncolored. The low source impedance means that frequency response coloration due to interaction with the loudspeakers basically vanishes. And I should note that, at least with my loudspeakers, the amps had no audible noise—even with my ear pressed against the midrange or tweeter, I could not tell if the amp was on or off.

Although most of my testing used the NCx500 modules to power the upper part of my speakers, I did swap them for the 1ET400A Purifi amps I've been using to power the bass. I was unable to hear any significant difference, not surprising given the low distortion, high power delivery, and low source impedance of both amps. As a side note, the common legacy of the Purifi and Hypex modules is reflected in the I/O connector pinout, which is almost identical between the two products. The visual appearance of the boards is also remarkably similar.

Photo 6: All module power requirements can be supplied by Hypex's SMPS1200A700 switched-mode power supply.

On the Bench

As usual, my test setup comprises an Audio Precision APx525 audio analyzer, a Vicnic 1kHz low distortion oscillator, and various home-built high power dummy loads. Unfortunately, the metal enclosure for the amplifier had not arrived by measurement time, so these were taken with the module and the power supply on the previously mentioned breadboard with no shielding. It's safe to assume that the already-low noise in these measurements would be reduced even further with a shielded build.

First, the basics. Gain with the preamp enabled measured 27.8dB; with the preamp disabled, the gain dropped to 12.5dB. Frequency response was flat until a small roll-off (-0.2dB) at 20kHz. Source impedance at midband was below my reliable measurement capability (<20mΩ). This could be measured with a Kelvin connection, but suffice it to say that it's likely that in a real-world setup, the speaker cables' resistance will vastly dominate.

With the preamp in place, the differential (+ to -) input impedance measured 94.3kΩ. With the preamp bypassed, input impedance was 2.8kV.

Next, I shorted the input to measure the output noise spectrum, which is shown in **Figure 1**. Typical of self-oscillating amplifiers, the principal noise component is at the switching frequency, roughly 450kHz at a few ticks above 400mV. The other notable discrete frequency component is a 2mV spike just under 100kHz, presumably originating from the switching frequency for the SMPS 1200A700 supply. In the audio range, total noise summed to under 10μV.

Unless otherwise noted, the following measurements were made with the preamp enabled, a 96kHz sample rate, and 256k data points.

Balanced inputs are advantageous because of CMR. CMR is very dependent on input stage impedances and topology. **Figure 2** shows the CMR versus frequency of the NCx500 module, with the upper curve representing fully balanced input and the lower curve obtained by driving pins 2 and 3 of the input with identical signals (i.e., common mode). The CMR at low frequencies is extremely good at 100dB (high hum and buzz rejection), and even at high frequencies, it remains better than 50dB (i.e., over 99% of the common mode noise is rejected). The jiggle at 60Hz is due to some stray hum pickup in the measurement setup.

Turning to power, I first ran THD+N versus output power sweeps at 100Hz, 1kHz, and 10kHz into an 8R dummy load, the results of which are shown in **Figure 3**. Note the remarkably low distortion floor! The distortion starts rising at about 250W. The 10kHz distortion is a higher than at midband or bass but still only reaches 0.015% at 300W.

Repeating this measurement for a 4R load results in the data shown in **Figure 4**. In the bass, distortion reaches 1% distortion at 650W; in midband, the amp puts out 700W at the same point; and the 10kHz power at 1% distortion is about 620W. I can see why I was able to get higher clean SPL from these amps in my

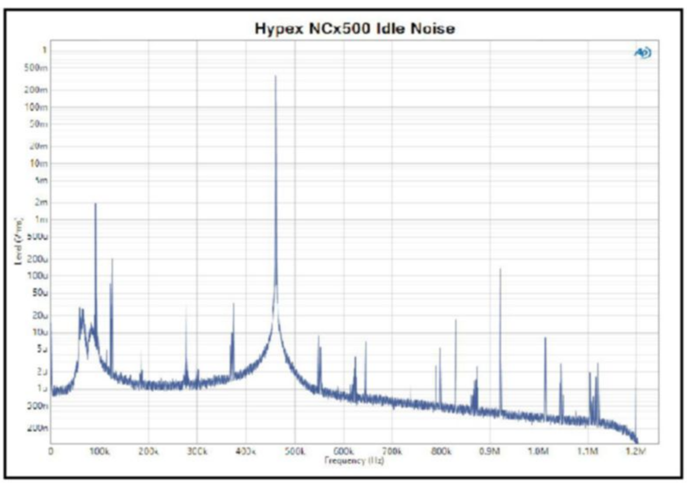


Figure 1: The zero-signal noise of the NCx500 is quite low other than the typical spike at the switching frequency.

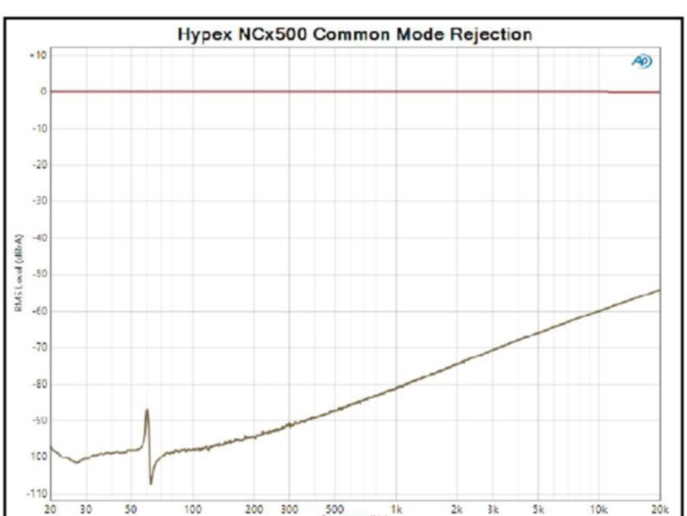


Figure 2: The NCx500 has very high common mode rejection (CMR), especially at low frequencies.

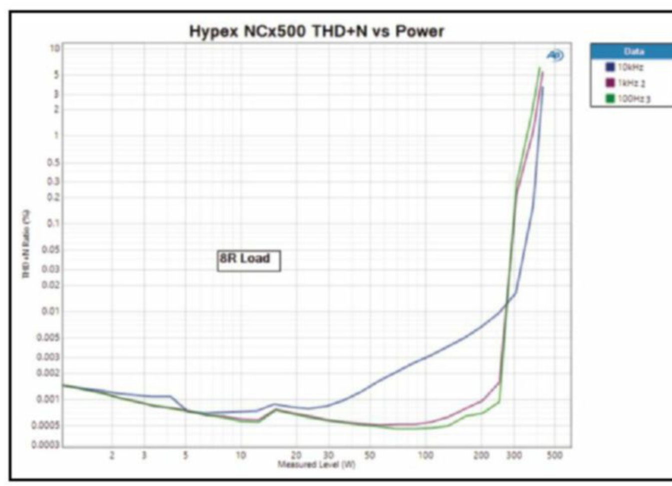


Figure 3: The plot of distortion versus power shows over 340W at 1% THD for an 8Ω load.

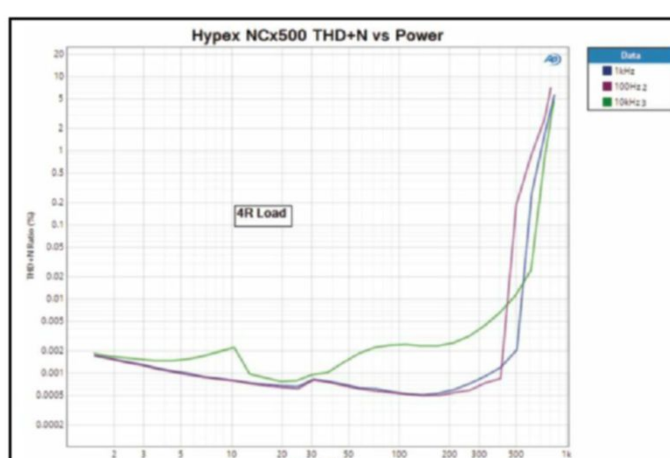


Figure 4: With a 4Ω load, maximum power increases to over 700W midband.

Fresh From the Bench

system! An 18kHz/19kHz IMD test at 100W didn't faze this amp at all (**Figure 5**); the intermodulation product is -120dB down, which is to say a single part per million. Sidebands are slightly higher, reaching -110dB. This is remarkable performance to say the least.

Equally impressive is the dependence of distortion on frequency. **Figure 6** and **Figure 7** show the distortion as a function of frequency at 1W and 5W into 8R and 4R, respectively. There is some rise at higher frequencies, but even the worst case is an order of magnitude better than what the best amplifiers of 10 years ago could manage. Most of the midband distortion in the 8R results is the residual of the Audio Precision analyzer.

The harmonic distortion spectra also indicate remarkably high performance. The spectrum of a 1kHz sine wave at 1W into an 8R load is shown in **Figure 8**. Third harmonic is dominant at -113dB below the fundamental, which is edging pretty close to the test equipment residual. Increasing the power to 100W gives the spectrum of **Figure 9**, where the third harmonic has increased to -109dB, still an amazingly low level. Higher frequencies do nearly as well. **Figure 10** shows 10kHz distortion at 1W into an 8R load; the third harmonic is still near test residual, but compared to the

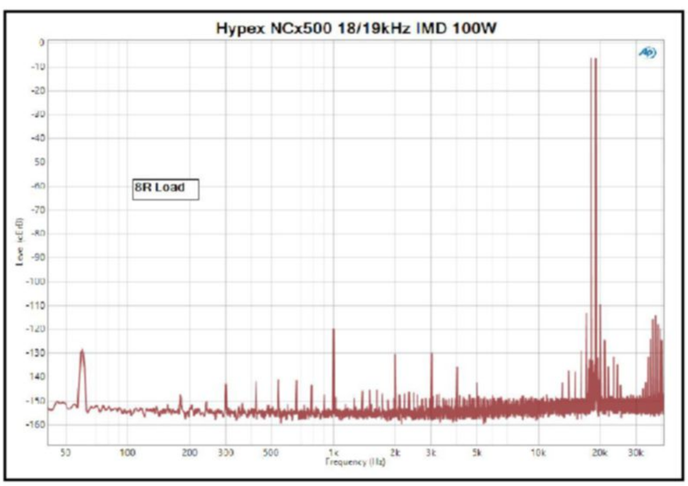


Figure 5: The high-frequency intermodulation at 100W out is superbly low, about 1ppm.

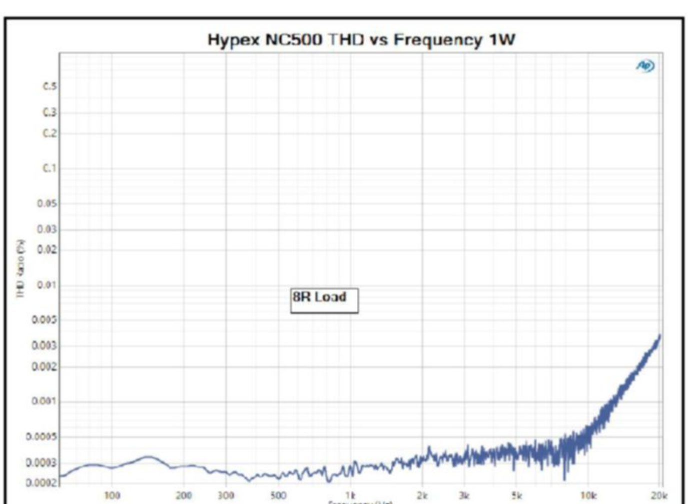


Figure 6: The distortion versus frequency plot at an 8Ω load shows ultra-low distortion through most of the audio band and only a small rise at the highest frequencies.

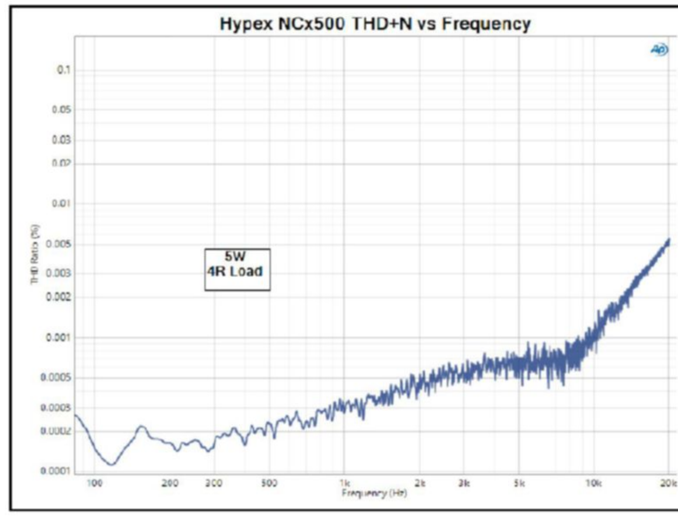


Figure 7: With a 4Ω load, the distortion versus frequency remains extremely low.

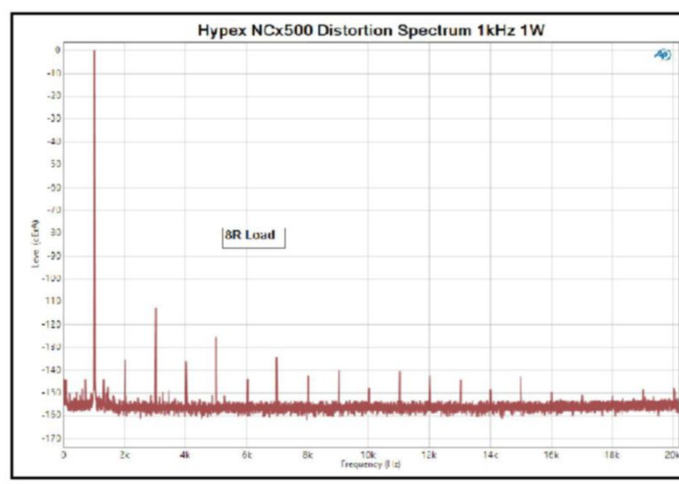


Figure 8: The distortion spectrum at 1W into 8Ω is dominated by the third harmonic at a level challenging the audio measurement gear.

Resources

Hypex Electronics BV "Heatsinking Requirements of Class D Amplifiers," www.hypex.nl/img/upload/doc/an_wp/AN_Thermal_design.zip

Hypex Electronics BV "High Efficiency High Power SMPS," www.hypex.nl/documenten/download/1848

B. Putzeys, "NCORE Technology White Paper," www.hypex.nl/img/upload/doc/an_wp/WP_Ncore_Technology.pdf

B. Putzeys "The F-Word, or why there is no such thing as too much feedback," *Linear Audio Volume 1*, https://linearaudio.net/sites/linearaudio.net/files/volume1bp.pdf

S. Yaniger "A Tale Of Two Class D Amplifiers," *audioXpress*, July 2020, https://audioxpress.com/article/fresh-from-the-bench-a-tale-of-two-class-d-amplifiers-orchard-audio-bosc-and-purifi-audio-eigentakt-eval1

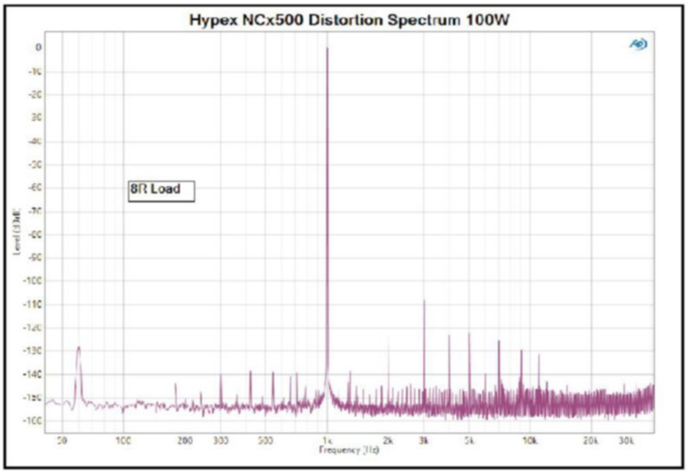


Figure 9: Increasing the power to 100W, the signal spectrum is still remarkably clean.

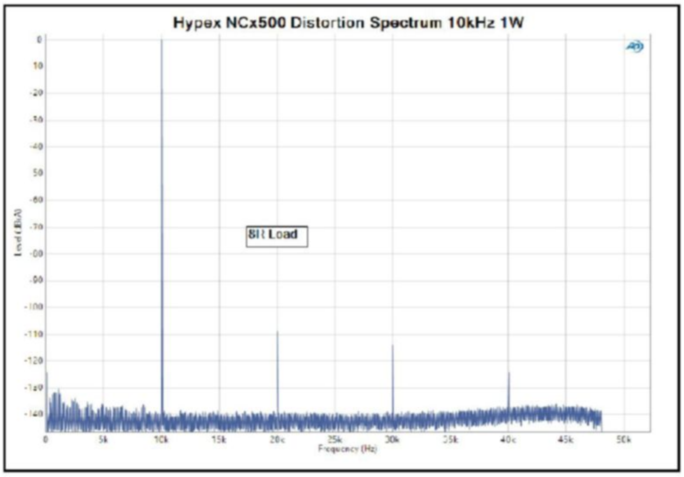


Figure 10: The distortion spectrum at 10kHz is still very clean, with monotonically decreasing harmonics.

1kHz result, second harmonic is now dominant at -109dB. All in all, these are absolutely superb measurements.

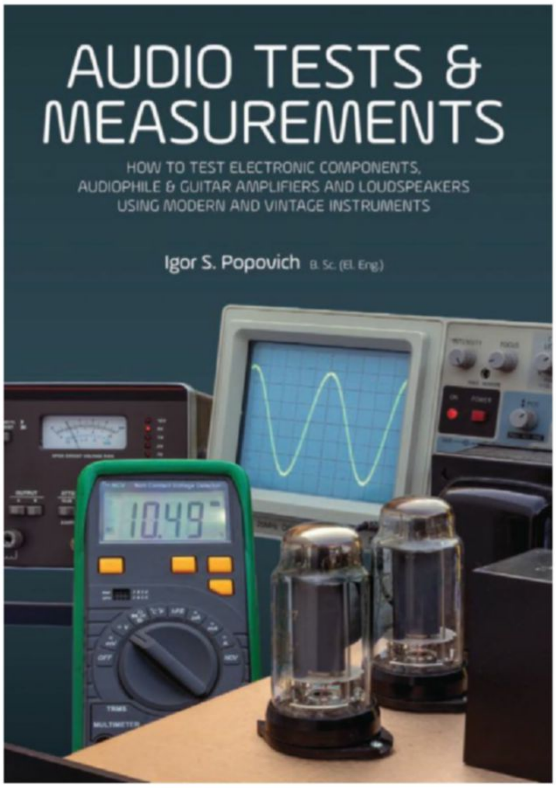
Conclusions

Although the big selling point here is the lower distortion and source impedance, in reality these two characteristics had already long ago passed the point of diminishing returns. So it's great that the evolution of the NC series engineering is continuing, but it honestly has little or no impact on the sonic performance of these amplifiers, which remains impeccable. The real differentiator here from earlier generations, to my mind, is the built-in preamplification, which is at a performance level that integrators will have trouble equaling: These boards are the entire package, no extra circuitry needed to make an ultra-high-performance product, just box, power supply, and wiring.

The extra power and performance I get from these modules has me sifting through my bank account to see how I can buy a couple for myself. The only unrequited longings I have are for a unity gain version and current monitoring outputs. Other than that, this is as close to a perfect product as I can imagine.

Learning Audio Testing and Measurements

Review by
Jan Didden



Our technical editor Jan Didden explores Igor Popovitch's book, *Audio Tests & Measurements*, and provides his thoughts on its value to our readers.

Books about designing and building audio equipment are a dime a dozen (figuratively speaking), treating solid state, tubes, analog, digital, what have you. But a book that focuses on measuring all those designs is rare. This is one of those books, written by a relatively unknown (in our neck of the woods) author Igor Popovitch. Popovitch, it turns out, is quite a prolific author, as he also wrote many books about designing audio equipment—a quick look at Amazon lists at least 10 books authored by him.

Audio Tests & Measurements
By Igor S. Popovitch
Revised edition 2022
Career Professionals, Canning Vale, Australia
ISBN-13: 9780980622393
$60 from Amazon or Barnes & Noble

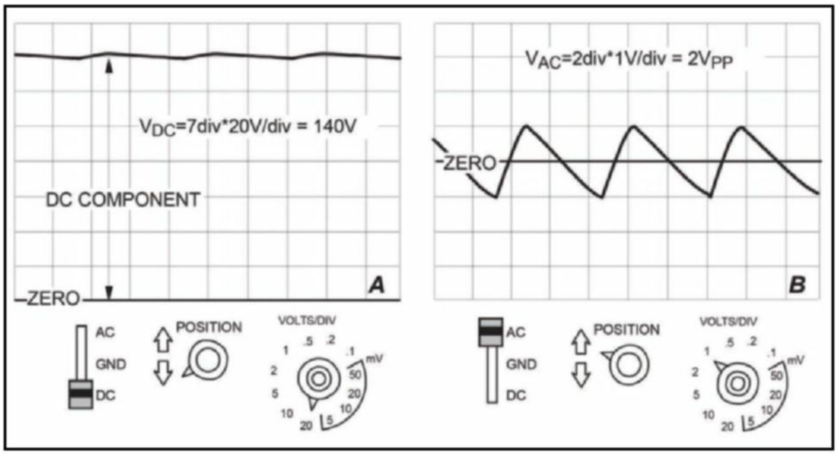


Figure 1: A sample problem: how to set the scope controls to find the DC and ripple voltage on a DC power supply output.

This particular book, *Audio Tests & Measurements*, runs about 180 pages, but that is deceiving as the book's format (A4 size) is about twice the size of most soft cover books. So as to the contents, it is better considered a 360-page book, or even more as the margins are quite small. So let's call it a 400-page book.

This book covers a very wide field of measurement-type methods. Sensibly, it kicks off with a chapter on test instruments, errors, limitations, and safety issues. Subsequent chapters discuss test equipment such as signal sources, signal tracing, power supplies, filters, multimeters, and oscilloscopes.

Then Popovitch moves on to actual tests and measurements: testing passive components, (pre) amplifiers, distortion measurements, transformers, loudspeakers, curve traces (solid state as well as tubes), and tube testing.

When I started reading this book, I had some ambivalent feelings about its value. The author makes it very clear at the beginning of the book that he focuses on explaining measurement methods and not necessarily providing user guides for test equipment.

The examples and use cases often show what he calls vintage test equipment, such as the venerable HP89903A analog distortion analyzers and even Heathkits. He briefly mentions digital oscilloscopes, but not PC-based oscilloscopes or instruments. On the other hand, the chapter titled, "Oscilloscopes, how they work and how to use them" is a gem and almost worth the price of the entire book. And, the section he calls "A crash course on oscilloscope controls" should be mandatory reading for anyone entering this field (**Figure 1**).

The chapter on distortion measurements treats all the usual measurements such as total harmonic distortion (THD) and THD+N (noise), using spectrum analyzers, selective voltmeters or fast Fourier transform (FFT) processing to study individual harmonics, and the concept and measurement of intermodulation distortion (IMD) as shown in **Figure 2**.

And that leads us to the question: At which audience is this book aimed? Popovitch himself gives the answer. The book is a great resource for the audiophile interested in the technical side of hi-fi equipment and its testing. If you are a guitar player interested in modifying, repairing, or even building your own equipment, you can make a head start with this book. But even electrical engineering students should find lots of value in this book as it explains test and measurement concepts in a straightforward no-nonsense way, with emphasis on practical measurements instead of overwhelming theory. This also means that the test equipment discussed is generic and leaves the prospective readers free to use or select the equipment they might have accessible.

So, as I worked through this book, my initial ambivalence gave way to enthusiasm. If you recognize yourself in the intended audience as discussed, this book will help you get a fast start in this field. As your experience and understanding grows, you probably want to go into more detail, but then you've obtained a solid knowledge base and you can build upon that. Check it out.

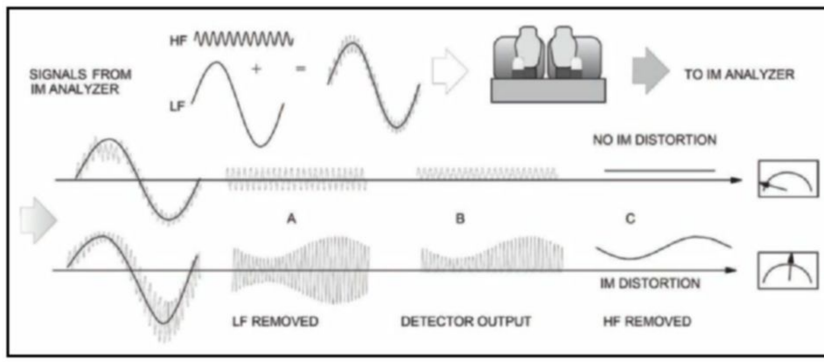


Figure 2: Conceptual view of measuring intermodulation distortion

The Vanderveen Trans PP80 Valve Amplifier

This new push-pull (PP) valve amplifier, featuring the Trans technique, delivers 80W power, refined resolution, valve-sound envelopment, and low distortion plus good speaker damping. And... the amp is simple to construct.

By
Menno van der Veen

Photo 1: A completed version of the VDV Trans Push-Pull 80 Valve Amplifier.

This valve amplifier is the next step in my Trans research, with earlier single-ended Trans designs published in previous *audioXpress* editions [1,2,3,4]. In this article, we present the new push-pull version (**Photo 1**). I say "we" because my TubeSociety students, Erwin Reins and Hans Gubbens, strongly supported this effort by making prototypes.

To start, I will provide a short summary of the Trans essentials [4]. Trans uses local feedback only around the power tube while the output transformer is outside this feedback loop. The power tube is driven by a voltage controlled current source plus one resistor R** between the anode and the control-grid. Then all the voltage amplification of the power tube is used to correct its internal distortions. **Figure 1** shows a visualization of the Trans concept. The local Trans-feedback also makes the power tube's plate resistances very small, thus preventing magnetic distortions inside the output transformer [5].

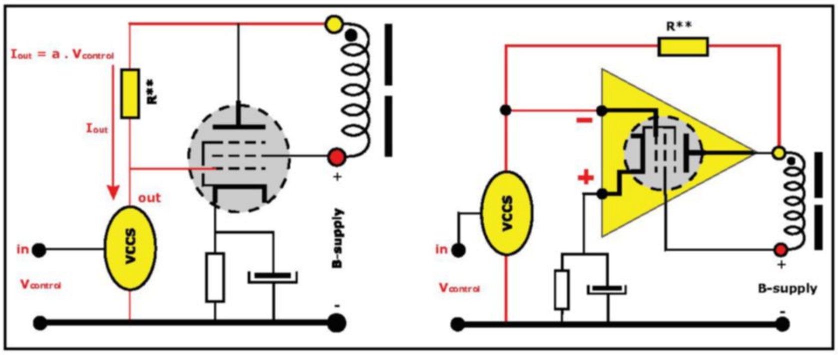


Figure 1: The Trans principle is detailed with a tube and an equal tube op-amp circuit.

In summary about Trans: Its local feedback is a soft local feedback technique that lowers power tube and OPT distortions. Years ago, when I started with Trans, I said: "I have gold in my hands." This observation has not changed during the research that has followed.

Power Supply Demands

The great advantage of my earlier published single-ended amplifiers is their constant current demand. The power supply can be simple, however, it needs good hum-rejection. The B+ voltage will not change under loud or soft music reproduction.

However, the situation in Class-AB push-pull amplifiers is totally different. With soft music (Class-A), we only have to deal with the constant quiescent currents. However, with louder music in Class-B, the tube's peak currents can rise up to eight times. So, the power supply encounters a large and changing power demand and consequently its B+ voltage will change/sag.

The Trans circuit uses DC-coupling between the anode and the control grid. If B+ changes in Class-B,

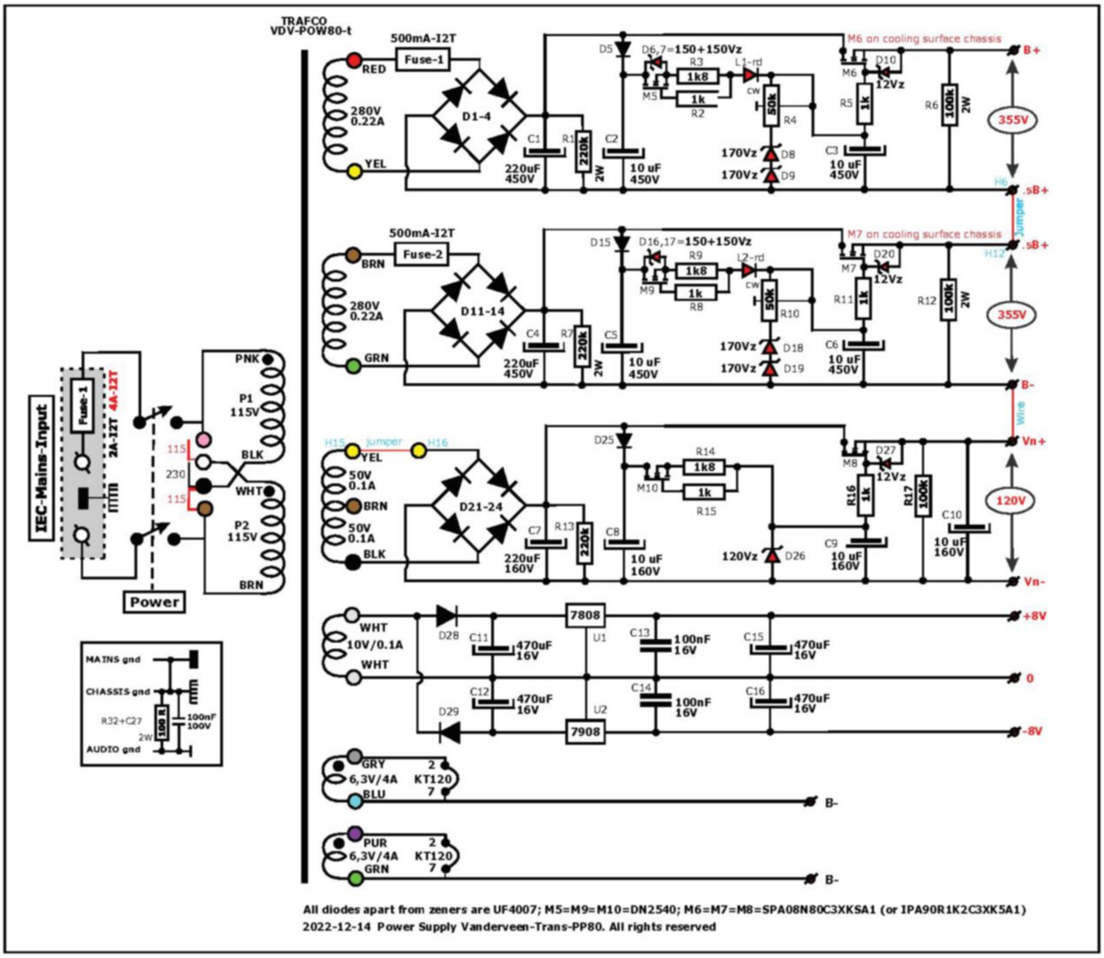


Figure 2: This schematic shows the VDV Trans PP80's power supply.

the quiescent operating point will change as well, pushing the tube away from its operating point. This means: we need a new B+ supply, which I discuss next.

The Power Supply Circuit

The power supply circuit is shown in **Figure 2**. Here, again, I applied a MOSFET for soft start and suppression of hum components. New are the Zener diodes, which create voltage stabilization.

Using high voltage (280V AC) winding, a small current source drives a constant current of 1mA traveling through a red LED (to show that the supply functions), then through a trimmer potentiometer (trimpot) of 50kΩ, and finally through a 340V Zener diode (2 × 170 Vz in series). Because the 1mA current is constant, we get a constant voltage at the tappers of the trim pots R4 and R10. This stabilized voltage is fed to the gates of the M6, and 7 power field-effect transistors (FETs).

When the trimpot is at maximum (clockwise), the supply output is about 390V DC unloaded. When we trim this output voltage to the requested constant 355V DC, we have enough headroom to deal with transformer regulation and hum, and all the variations in mains-voltage and Class-A-B current demands. Because we have twice this circuit in series, a total B+ = 710V DC is available to be fed to the center tap of the output transformer. Also ½B+ = 355V DC is available to power the screen grids of the power tubes. **Figure 3** shows an impression of the Power Supply PCB.

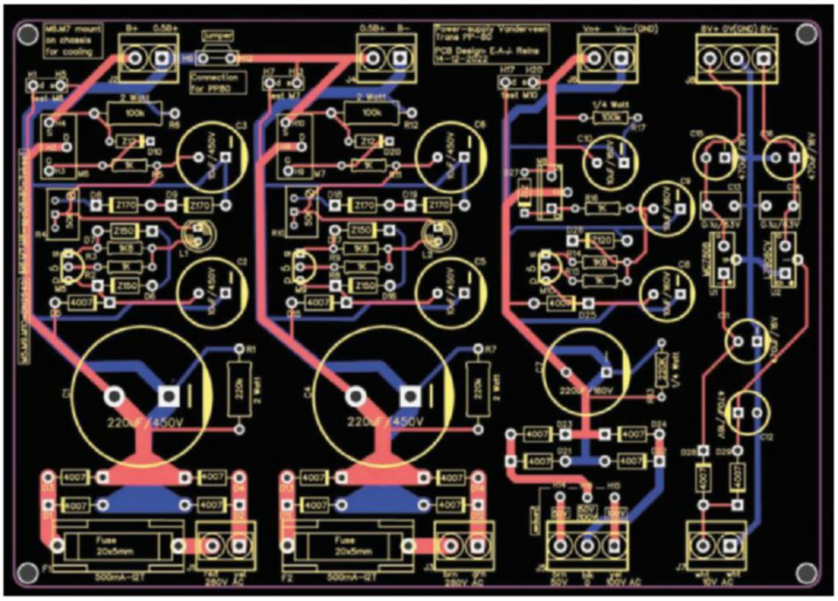


Figure 3: The power supply's PCB, shown here, is 150mm × 105mm.

Balanced Input Plus Phase Splitter

The power stage of the amplifier needs to be driven by a high-quality phase splitter with amplification factors of +1 and -1. **Figure 4** shows the circuit.

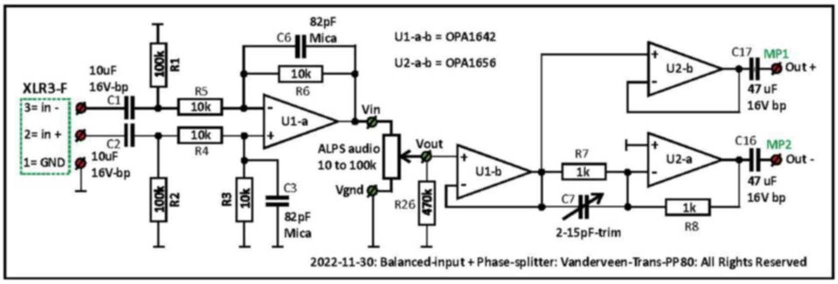


Figure 4: This is the schematic for the balanced input plus phase splitter.

The circuit around U1-a is a textbook balanced input circuit. Not much to say about it except that its high-frequency range is limited by C3,6 to prevent a square wave overshoot.

When you apply unbalanced music sources (RCA), you don't need the U1-a circuit. In that case, connect your input signal directly to Vin (to the top of the Alps volume control, which is not connected to the output of U1-a). Or use a toggle switch (see more about this later) to choose between unbalanced and balanced inputs.

Compared to the simple U1-a circuit, the Alps volume potentiometer plus U1-b circuit is special. The OPA1642 is a dedicated op-amp with constant input capacitance (6pF), which is not reacting on the amplitude of the input signal. Its f-3H high-frequency roll-off will not react on the amplitude of the signal nor will it create changing phase delays. This prevents phase and harmonic distortions, which trained ears can notice. This small circuit largely improves the quality of the volume control. I have never heard such clean Alps behavior.

The actual phase splitter is realized with U2, where U2-a amplifies -1× and U2-b +1x. Any difference in time delay inside U2-a and U2-b can be trimmed to zero with the trim capacitor C7, to create exactly 180 degrees phase difference up to 1MHz. I will discuss later in the article how to trim C7 to its optimal setting.

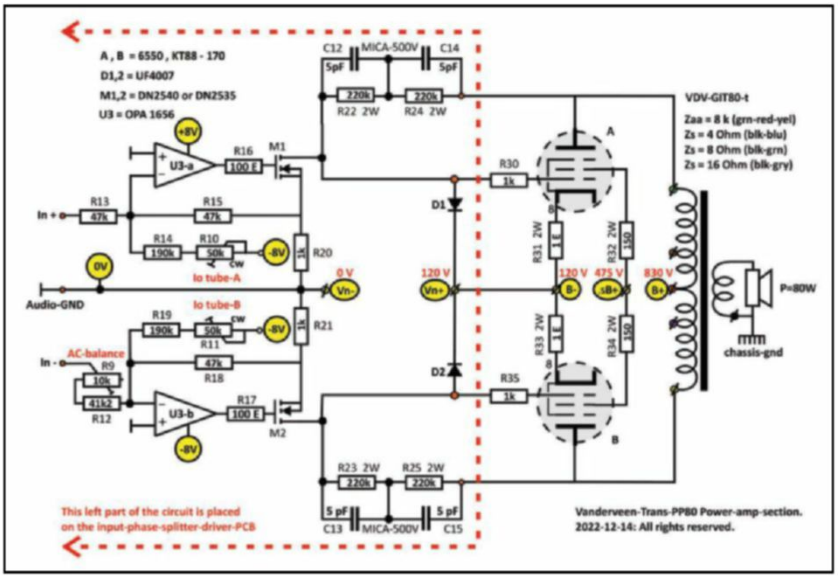


Figure 5: This schematic shows the Trans PP80's power amplifier section.

The Trans PP80 Power Amplifier Circuit

Seeing the circuit shown in **Figure 5**, you might wonder: Is that all there is? Yes, this simple circuit delivers 80W and gives you all the details you might need. Let's discuss it.

The actual voltage-controlled current sources are the FETs M1,2, each being totally controlled by the low distortion op-amps U3-a,b (= OPA1656). This results in extremely linear current sources, which are temperature compensated as well.

Because of the DC-coupling to the control grids of the power tubes, the anode quiescent currents can also be controlled by U3-a,b and set with R10 and R11 to Io = 60mA per tube, meaning 60mV DC over the 1Ω cathode resistors R31 and R33.

The resistance numbers of the upper PP section are used in the next formulas. Each op-amp amplifies A = R15/R13 = 1x. Therefore, the transconductance of each complete current source is given by:

$$g = (Vin/R20) / Vin = 1/R20$$
$$= 1\ mA/V$$

Photo 2: These Trans monoblocks were designed by Hans Gubbens.

GLASS AUDIO

The amplification from each input to each power tube anode is:

$$A_{1/2} = (R22 + R24) \cdot g = 440 \text{ x} = 52 \text{ dB}$$

The voltage amplification of the total push-pull (PP) amp from both inputs to the Zs = 4Ω secondary tap of the output transformer needs its opt-turns-ratio correction plus a factor of 2 because the primary Zaa = 8kΩ is push-pull driven:

$$A_{T,4} = 2 \cdot A_{1/2} \cdot \sqrt{(Zs/Zaa)} = 19.7 \text{ x} = 25.9 \text{ dB}$$

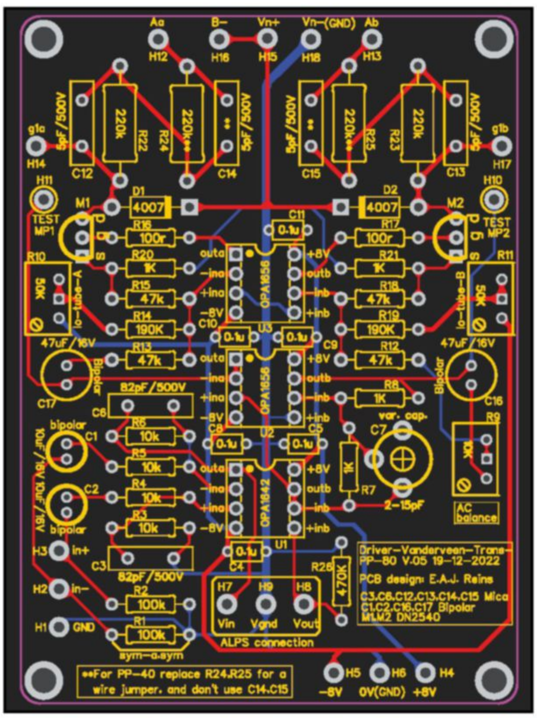


Figure 6: This image shows the PCB for the input phase splitter and driver section (90mm × 65mm).

Photo 3: Here, the SOIC8-DIP8 adaptors have been added to enable easy op-amp exchange or repair.

Next, we shall subtract the losses of the winding resistances Rip = 100Ω and Ris4 = 0.16Ω of the Git80-t output transformer resulting in:

$$A_{eff,4} = A_{T,4} \cdot Zs / (Rip \cdot Zs/Zaa + Ris + Zs)$$
$$= 18.7 \text{ x} = 25.4 \text{ dB}$$

Please notice that the voltage amplification of the total Trans PP80 amplifier is only determined by the resistors mentioned in this section and the turns ratio and resistances of the Git80-t output transformer windings. You can imagine that high-quality resistors and the Git80-t opt (no resonances and low leakage) are mandatory in this design. [For the diehards: in $A_{eff,4}$ the small Trans plate resistances of the output tubes are not taken into account, but the validity will be shown in the section entitled "Measurements."]

I have already heard comments about this Trans design such as: "Menno, you are using ugly semiconductor stuff to control a valve amplifier: shame on you!" My response: With highest quality op-amps with effective gain of 1x, I create clean current sources with ultra-small distortions. Notice that the first-order-low-pass open-loop-gain of the op-amps is fully available for all these corrections. Next, each power tube corrects itself, only using its own tube-transfer function. So, the actual sound character is from the power tubes only. Even the sound character of the OPT is excluded because of the small Trans plate resistances of the power tubes [5]. For these reasons, I call this Trans PP design a pure valve design.

[For the diehards: I am fully aware of the different opinions about "the sound of circuits." I conducted a lot of research in this area. For instance, I compared this new design with my earlier Trans SE versions and earlier Trans PP amps [6]. There no op-amps were applied, just the fewest discrete components, even all tube circuits, and so on. My observation is that this newest Trans PP80 design sounds superior in details and envelopment]. Enough for now. Let's start constructing this amplifier.

Construction of the Amplifier

This article shows several photos detailing how Erwin Reins made the amplifier, the case he used, and how he placed the components. Let it inspire you. Monoblocks are also a good choice. **Photo 2** shows the versions built by Hans Gubbens. As I often say to my students: "A new amp starts with dreaming about how it looks."

Visit my website, www.mennovanderveen.nl, [7], to download all the extra information about the PCBs plus the detailed PCB schematics in which all connection points are clearly indicated, as well as where to find the transformers, the start-up procedure, and so on. Assuming that you have studied that material, I will now focus on the major construction issues.

The Input Phase Splitter Driver PCB

Figure 6 shows the actual driver PCB in detail. This image shows how to place all its components. Please notice the orientation of the Io trimpots (turn completely counter-clockwise) and the AC balance trimpot (leave in mid-position) and of the high-frequency trimmer C7. **Photo 3** shows the SOIC8-DIP8 adapters that we applied to enable easy op-amp exchange or repair. Do not yet connect the Alps volume potentiometer.

Figure 7 shows how to trim C7 to its optimal position. You need a 1MHz-sine-1Vrms function generator signal and a two channel oscilloscope (for instance the Pico-2204). Only for this test: use two 9V DC batteries to deliver power to the PCB. Wire the signal cables as indicated.

Trim C7 until a straight line is obtained in the Lissajoux A-B scope presentation. Having done that, leave C7 alone. The phase-splitter will stay long term stable up to 1MHz. Next remove all cables and batteries.

Figure 8 shows how you can wire your inputs to balanced (XLR) or unbalanced (RCA) with the help of a double-pole miniature toggle switch (on-none-on). The wiring of the Alps is also shown.

Wiring the Power Amp Audio Section and the Supply

Next you can place the driver PCB in your amplifier case, preferably between the power tubes (or very close to them) for the shortest wiring to the power tube sockets to prevent any oscillation. **Figure 9** offers more details.

You might place the tubes, the transformers, and the PCBs at other positions. However, please lay all wires to their connecting points as indicated. This drawing (Figure 9) carefully takes all the currents and ground-loops into account, preventing the large power tube currents from entering the sensitive driver section. I have not explained anything about the wires from the Pow80-t transformer to the power supply PCB because I think it is self-explanatory. However, the comments made about FETs M6,7 are very important. These FETs become hot (4W of heat at least) and do need the cooling of the aluminum or steel chassis.

Also not shown are the 6.3 filament wires (indicated with f). Twist them and connect to pins 2 and 7 of each power tube socket. The primary winding of the Git80-t has two unused UL-taps. These Brown and Orange wires need proper isolation. Don't cut them as you might wish to use them later.

When you study the images of the Erwin Amp (**Photo 4**), you see that he applies a mains switch-on delay circuit. In my amps and drawings, I applied special I2t- or T-fuses, which can handle the large inrush currents of toroidal mains transformers.

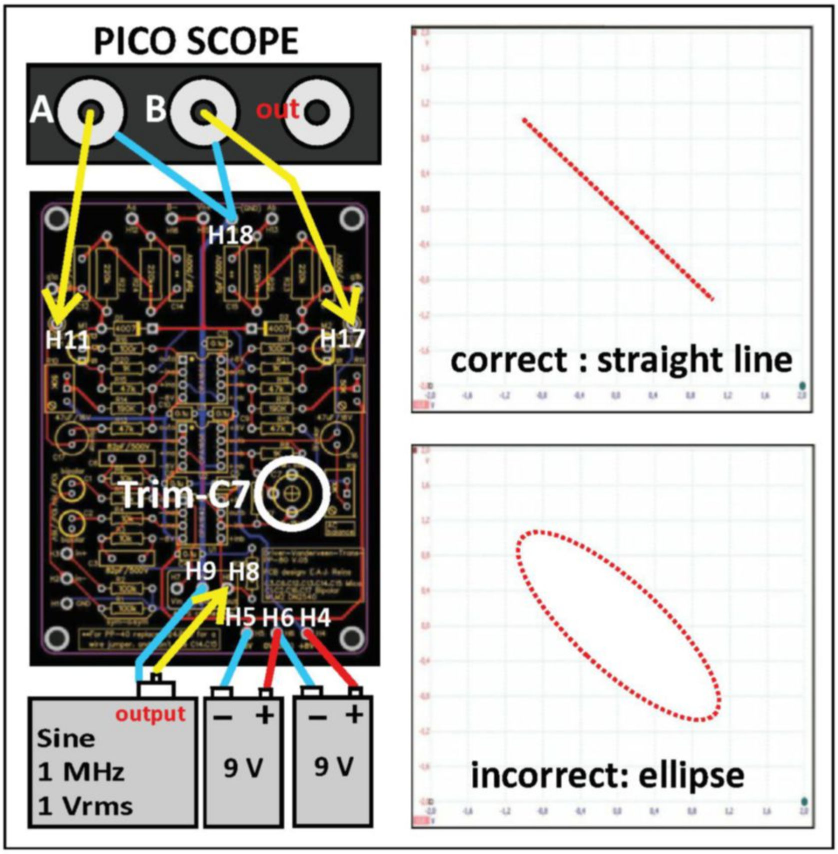


Figure 7: This diagram details how to trim C7 with a 1MHz sine wave and an oscilloscope.

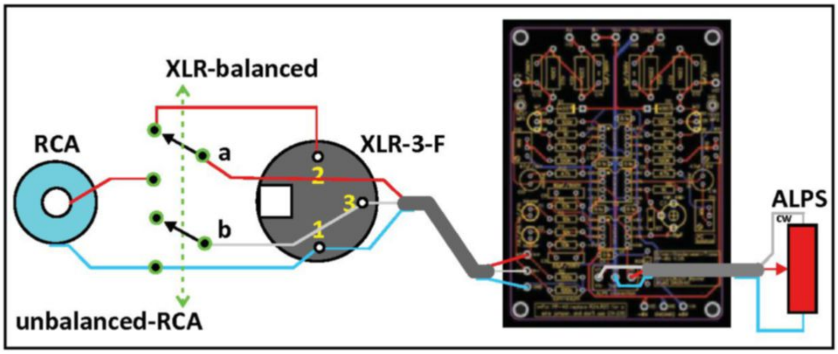


Figure 8: You can wire your inputs to balanced (XLR) or unbalanced (RCA) with the help of a double-pole miniature toggle switch.

You Can DIY!

About the OPT and Some Safety Tips

Don't place the toroidal OPT very close to (or on top of) the mains-toroidal. The recommended minimum distance is at least 2cm to prevent noticeable hum in your loudspeakers. Please don't forget to connect the secondary black wire (the null wire) to the chassis. I have never experienced voltage sparks between the primary and secondary windings inside my toroidals OPTs; that's not the issue. It is the capacitive coupling between the primary and the secondary that can cause AC voltage coupling into the secondary, especially at higher frequencies. It is not harmful, but it can influence the results of your measurements. Proper grounding prevents such issues.

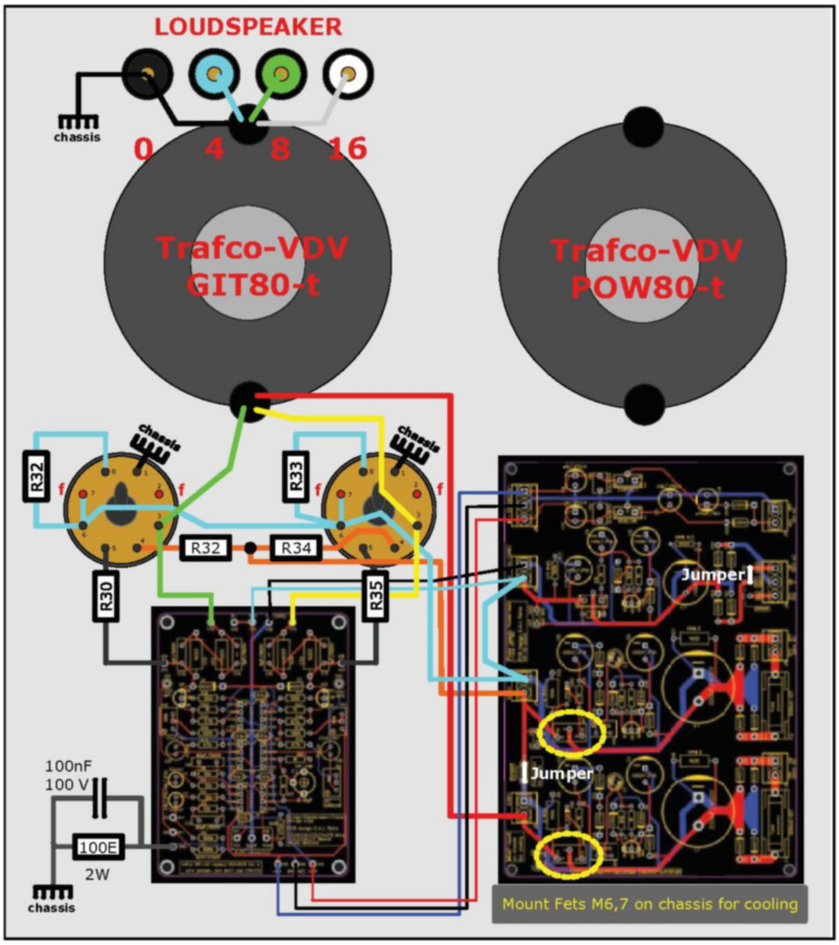


Figure 9: Place the driver's PCB in your amplifier case, preferably between the power tubes.

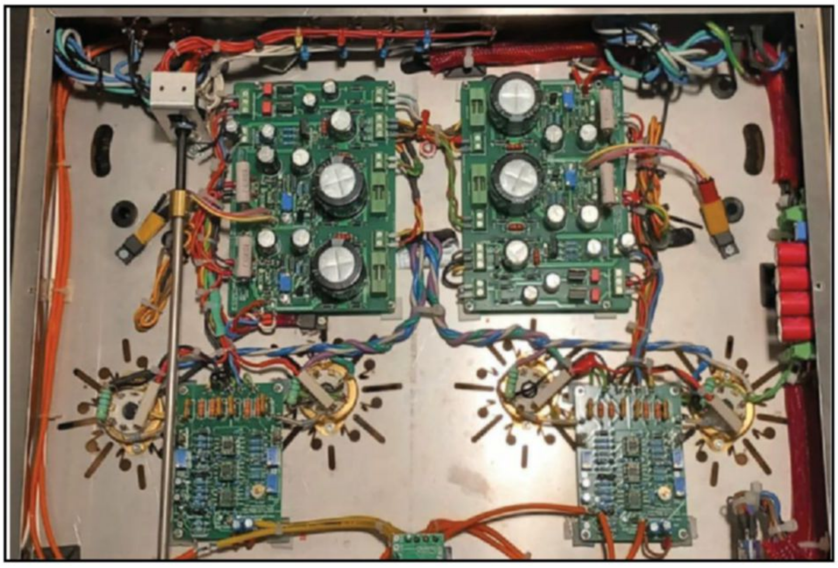


Photo 4: Here is a bottom view of the Erwin Reins amplifier, in this image you see that he applies a mains switch-on delay circuit.

If you apply KT88 to KT170 power tubes, then the foot section of the tube contains a silver-colored shielding ring, which is connected to pin 1 of the tube socket. Connect pin 1 to the chassis to prevent any possible voltage difference between this ring and the chassis, which might be harmful if you touch both the chassis and the ring at the same time.

Start-Up Document

I tried to make the start-up procedure a short document, but describing all the precautions and the proper actions and sequence takes too many words and too much space. Therefore, I have decided to make this important start-up document available on my website [7].

Measurement Results

Unless otherwise indicated: All measurements are performed in a 4Ω load at +6 dBV = 2 Vrms = 1W in 4Ω load. We mostly listen and judge at that level and below.

Figure 10 shows the gain and frequency range of the amplifier. Measured is 25.1dB gain while 25.4dB is calculated earlier. This small difference will be discussed in the output impedance measurement. The -3dB frequencies start below 5Hz and are above 60kHz. Figure 10 shows no resonances in the amplifier, and **Figure 11** confirms that result. Only a small negligible 50Hz residual is visible.

Next, **Figure 12** shows the harmonic components of a 1kHz sine test signal. You see the declining character while all are small. For me, it remains remarkable that the simple Trans technique is so powerful in reducing distortions! Also nice to notice that harmonics above three have a much smaller magnitude. As I stated earlier, this proves that my op-amps have enough open-loop gain to firmly reduce higher harmonics and to maintain the tube character of the sound. But I have a bit more to say about harmonics, which is not visible in Figure 12. Please look at **Figure 13**. The distortion components are measured for all frequencies from 5Hz to 20kHz, providing extra insight.

From 5Hz to 20Hz, all harmonics are larger than at 1kHz, caused by the magnetic properties of the core of the OPT (core distortions are proportional to 1/f). Due to the small Trans plate resistances of the power tubes, these distortions are not excessive, only 0.04% THD at 20Hz. This is the next bit of proof of the Trans miracle.

Above 1kHz, the components H2 and H3 and H4 rise with frequency, which is caused by a tiny imbalance inside the power tubes and OPT. Should they be matched to an equalness better than 0.05% at 10kHz? An impossible heck of a job. Because of their small magnitude, I paid no further attention to this. (Global negative feedback corrects such imbalances, but at a cost: You lose the free open tube sound character, but that is another story).

The last measurement, **Figure 14**, shows the output impedance of the amplifier at the 4Ω output taps. The result is 0.36Ω at 1kHz, meaning a damping factor DF4 = 11. This is large enough for most tube amp applications, and we get this result without any global negative feedback. [For the diehards: Backward calculation now

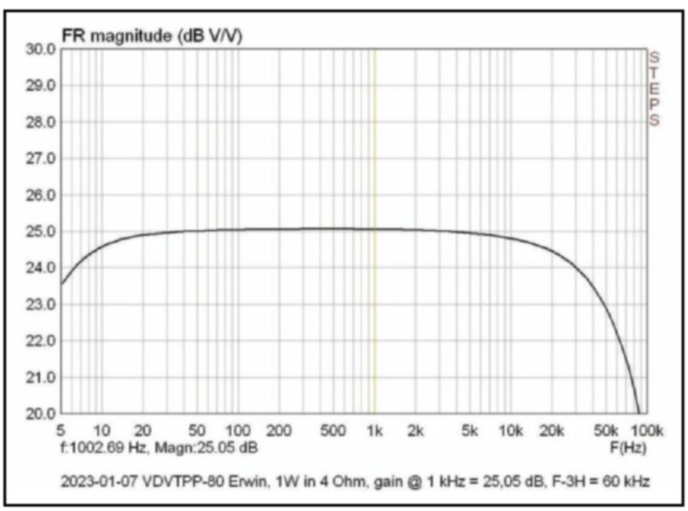


Figure 10: This graph depicts the gain and frequency range of the Trans PP80 amplifier.

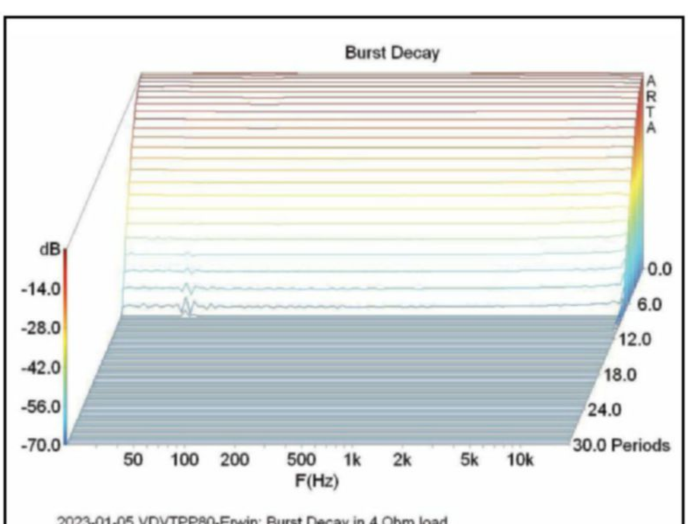


Figure 11: This measurement clearly shows the absence of any nasty resonances.

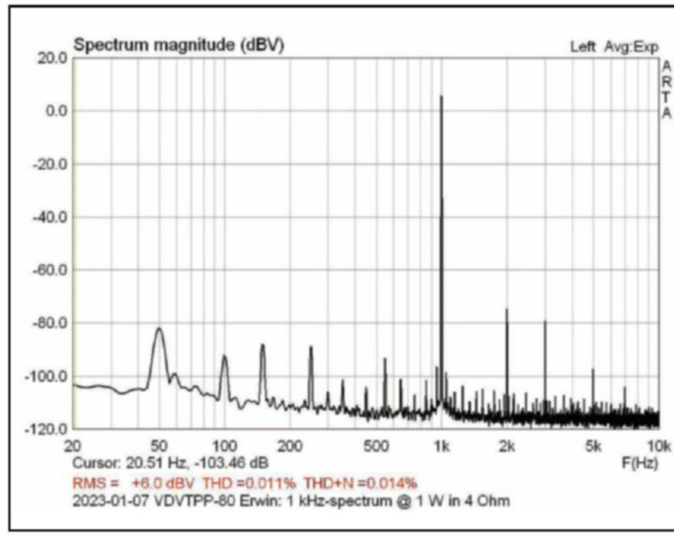


Figure 12: This measurement reveals the 1kHz distortion spectrum; notice the declining magnitude of higher harmonics.

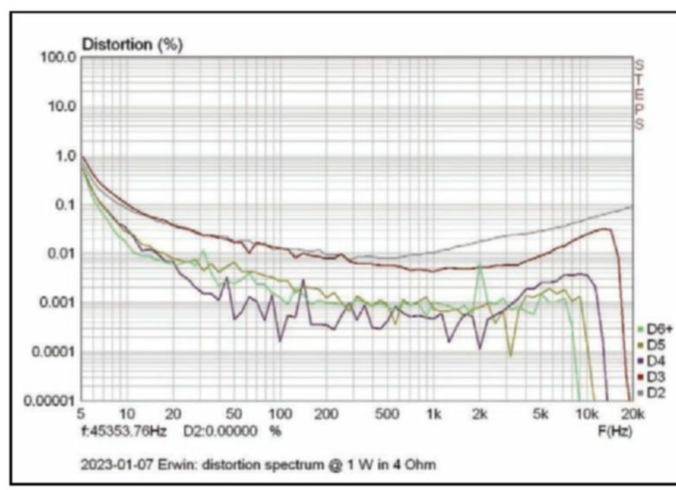


Figure 13: The harmonic spectrum from 5Hz to 20kHz is shown in this graph.

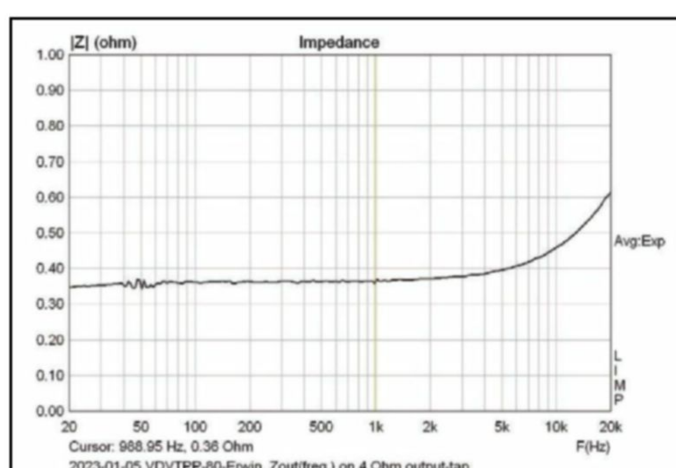


Figure 14: The output impedance Zout is shown at the 4Ω taps.

About the Author

Menno van der Veen studied Physics at the Groningen-University in the Netherlands and graduated on measuring doting-concentration in semiconductor wafers. After that he taught physics for a number of years. Meanwhile, he wrote many review articles for high-end audio magazines. He also was a sound engineer-coordinator in a theater and at outdoor festivals. In 1985, he founded his engineering firm with a focus on tube amplifiers and toroidal output transformers. He published his research in Audio Engineering Society (AES) papers and books and was chairman of the Dutch AES section. In 2005, he started his bi-weekly TubeSociety Saturday school. His Trans focus began around 2011 and the results were published in ***audioXpress***. In his private time, he plays and researches electronic guitars, their valve amplifiers, and effect pedals. Here lays the basis of his love for music and electronics and art.

delivers an effective plate resistance per power tube of rp = 150Ω. This is far better than any triode plate resistance].

Was I right to neglect these tube plate resistances in the gain calculation? Recalculating the last formula with 2 × rp = 300Ω added, results in $A_{eff,4}$ = 25.2dB (measured 25.1dB). The 0.2dB difference between 25.4dB and 25.2dB equals only a 2.3% linear mismatch. When you take component tolerances into account you get the same mismatch. That's why I dared to neglect 2 × rp.

What causes the rise of Zout above 5 kHz? It is the leakage inductance Lsp of the OPT. Trans does nothing to correct that. So, Lsp must be small, which is the case in my toroidal Git80-t. Try this with EI-type transformers with their larger leakage and you will be disappointed when applying Trans. With EI you absolutely need global negative feedback (NFB), including the OPT to create a Zout behavior as in this Trans design.

The Subjective Side

I hope you were able to read the subjective comments about my earlier Trans SE designs. They all have in common the idea that Trans sounds "Tube." All have more details and space than my previous global feedback or no-feedback designs. The new Trans PP80 has exactly the same properties plus better refined resolution and space and details as well as a quicker cleaner bass response (which is logical because Lp is much larger in push-pull). That is the essence—a major leap forward. Understandable? I think so, thanks to the contribution of the clean current sources, while maintaining the specific tube sound from the power tubes. Is more power a contributor to these observations? Yes, but normally I do not play my music that loud. I think improved control is the major factor.

Conclusions and Summary

The research question was: Will Trans work well in a push-pull configuration? The answer is: Even better than expected. The new amp has 80W of output power, is very stable, provides good speaker damping with little distortion, and sounds "Tube." This exercise taught me a new way of designing. Think in component transfer functions and research the influence of each transfer on the final sound character. Chose which transfer you wish to dominate and make all others super-clean. The Trans PP80 proves that such an approach works well. This raises my next research question. Nowadays, we have amps with 1ppm distortion. I clearly hear their good qualities. However, my ears prefer minimal amounts of distortion. Then I better hear space and environment. Is this like the idea that a little salt or grease improves the taste? Does minimal distinct distortion improve hearing?

Author Acknowledgements: The creation of this new Trans amplifier was only possible thanks to the intensive support of my TubeSociety students Erwin Reins and Hans Gubbens. Without their work, the realization of the amplifier would not have happened. I also thank the European Triode Festival (ETF) organization and Audio Vereniging Midden Nederland (AVMN) and all my TubeSociety students. There I found the platforms to expose my new concepts to trained ears. Their comments were of the utmost importance, replacing my pride by solid knowledge and preventing me to be too subjective.

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A (Z)OTL with 300Bs

Part 1 — The Makeover

Rebuilding an amplifier for reasons of better reliability, better aesthetics, or bringing it up to the latest standard can be very rewarding. And if it is a classical, technological grandstand, even more so. Thomas Perazella embarked on such an adventure with a famous David Berning ZOTL amplifier. In Part 1 of this article, Tom presents the amplifier and discusses its unique circuitry. In Part 2, he will proceed to dismantle and rebuild the amp for the new owner and place it in a new chassis.

Here is a preview of the 300B's updated look, shown with the tube cage in place.

By
Thomas Perazella

It has been intriguing to watch the renewed interest in the 300B triode for audio circuits. A descendant of the 300A originally produced by Western Electric, it has in recent years taken on an almost mystical reputation with certain audiophiles. There are many interesting stories of the 300B on the Internet.

One of my friends, David Berning, has been building very high-quality vacuum tube audio amplifiers for many years. He has used many different tube types, but the fundamentally outstanding feature of his designs is his patented ZOTL technology. One of the factors that is imperative in getting good performance from a vacuum tube amplifier is the process of converting the high impedance of the tube to the low impedance of the speaker in a manner that preserves the frequency response, phase, and low distortion of the active device. The most common method of getting this impedance match is the use of a transformer. We will take a closer look at transformers and the factors such as leakage inductance, interwinding capacitance, hysteresis, saturation, and others that make it very difficult and costly to build a wide bandwidth transformer free of gremlins.

Transformer Basics

Over the years, *audioXpress* has had several excellent articles describing transformers and their applications in audio. The most recent article series is listed in Resources and it references a number of additional articles. They contain a wealth of knowledge on transformer design, materials, construction, and compromises.

Two of the most important aspects of transformers in audio use are for electrical isolation and impedance conversion. Let's look at isolation first.

Isolation

In the early days of electrical transmission, the power and return lines were not directly referenced to ground although the generators were. Around 1913 that changed when the safety hazards of ungrounded systems were recognized. For example, if an electrical appliance such as a washing machine were fed ungrounded power and lightning struck a power line, it would travel through the washing machine to the water supply that was connected to ground, causing a fire. By referencing the mains power to ground external to the house, that type of hazard was minimized. Grounding the metal chassis parts of equipment also gave the current a path to ground instead of through a human body in case of a fault. That was the good part of grounding. However, if the power was not isolated from any exposed parts of the equipment, a severe shock hazard would exist for a person touching both the equipment and ground.

Considering that a lot of audio equipment uses unbalanced connections where one side of the signal is directly connected to the chassis, local isolation from the mains power is essential. A transformer is most often used to provide that isolation. In a transformer, the power is transferred from the mains primary side to the secondary signal side through the change of magnetic fields within the transformer. Therefore, even though the primary side is referenced to earth ground, the secondary side is isolated, preventing a shock hazard to earth ground.

Impedance Matching

In an electrical circuit, maximum power can be delivered when the impedance of the source matches the impedance of the load. If the impedance of the load is much higher, then it cannot accept all the power that is available from the source. If the impedance of the load is much lower, the source will not be able to provide all the power that the load can accept. The primary reason for using an output transformer in a vacuum tube audio amplifier is to match the high impedance of the tubes, which are the source to the low impedance of the speakers that are the load. It does this by having different numbers of turns in the primary and secondary windings. In a transformer, the impedance between windings varies with the square of the turns ratio. For example, if the primary side of an output transformer had 200 turns and the secondary had 10, the turns ratio would be 200/10 or 20 and the impedance reflected to the secondary from the primary would be 20 × 20 or 400. If the impedance of the tubes on the primary side was 3,200Ω the reflected impedance on the secondary would be 8Ω. Therefore, the tubes could provide substantially more power through the transformer than if the 8Ω load were directly connected to the 3,200Ω tubes. This impedance matching does come with problems due to the physical factors in the transformer construction. Articles mentioned in the references go into the limitations in detail. Those limitations grow as the transformer is designed to operate over a wide frequency range.

Avoiding Output Transformer Problems

Over the years, various designs have been used such as Output TransformerLess (OTL) amps using many paralleled tubes and no output transformer to lower the effective impedance delivered to the amplifier. At best, they were "band aid" solutions to the impedance matching problems suffering from cost and efficiency limitations. The documentation on the Berning website (see Resources) describes his use of a single-frequency transformer to interface between the tube and the output section of the impedance converter, eliminating the output transformer in the traditional sense while acting as a bidirectional low distortion transformer. This is the essence of the ZOTL concept.

Being an experimenter as well as a designer, Berning decided to make a push-pull 300B. Part of the reason was also to test the validity of using a vacuum tube switch-mode power supply. The slower turn on and off times of the tubes could result in less high-frequency switching noise, which is always an advantage. Since it was a proof-of-concept subject to the inevitable changes necessary in the development process, style was not the prime consideration. He found an old chassis that had been from a Hewlett-Packard tube digital counter and used that.

The resulting amp had very good performance. Berning used it for a while but someone in the audio industry heard about it and decided to purchase it. After a time, one of the members of the local Philadelphia Area Audio Group (PAAG) bought it and was likewise very pleased with the performance. However, with time, the rough looks began to cause him concern. At one of the PAAG meetings I spoke with him and mentioned that if he liked the amp so much, I could move all the electronics to a new chassis with improved aesthetic. He agreed and so I started the makeover process.

With a project such as this, there are four fundamentals necessary to provide a chance at a successful implementation. First, you must have a thorough knowledge of how the circuit operates. I had worked with Berning on several projects and in addition to studying the patent about the ZOTL circuit, had many discussions with him on the concept and operation. Second, you need to have a schematic of the circuit for the details on signal flow. Berning was kind enough to provide that. It was originally hand drawn but it has been redrawn and broken into two sections for clarity, courtesy of *audioXpress*. Third, you must photograph the existing piece from every conceivable angle. Fourth, and probably most important, you need lots of patience.

Circuit Operation

In addition to the ZOTL circuit described here, this amplifier has several other important characteristics necessary for optimum operation, including proper recovery from clipping. Except for a blocking capacitor at the input to prevent entrance of externally present DC, the amplifier is DC coupled throughout. In a traditional AC-coupled amplifier, clipping can induce asymmetric charge on the coupling capacitors leading to poor performance when the amplifier clips. Also, a unique auto bias circuit is used to insure proper bias of the output tubes.

Details of the ZOTL operation are on the Berning website. In short, a high-frequency oscillator drives a bidirectional low-voltage switching power supply, the impedance converter, that can both source and sink current to and from the speaker. That power supply is modulated by the vacuum tubes through a small transformer. This is not the common pulse width modulation (PWM) mechanism, but rather a double-polarity pulse amplitude modulation. Information on that operation is included in the Resources. Since that transformer always sees the same frequency, the problems associated with a wide bandwidth transformer are eliminated. The power that goes to the speaker never has to navigate its way through a typical output transformer. The small control transformer only modifies the voltage levels of the individual pulses in the impedance conversion process resulting in the voltages that the low voltage power supply delivers to the speaker. So, you have the best of both worlds. The characteristics of the output tube are preserved by passing through this quite substantial low impedance power supply. **Figure 1** shows the redrawn drawn schematics from the one that David Berning supplied.

Mains power is first supplied to a multi-winding transformer that has the dual primaries wired in parallel for 120V operation. There are four secondary voltage windings. The high voltage winding is center tapped and feeds a 5U4 rectifier tube in a full-wave center-tapped configuration. Filament power for

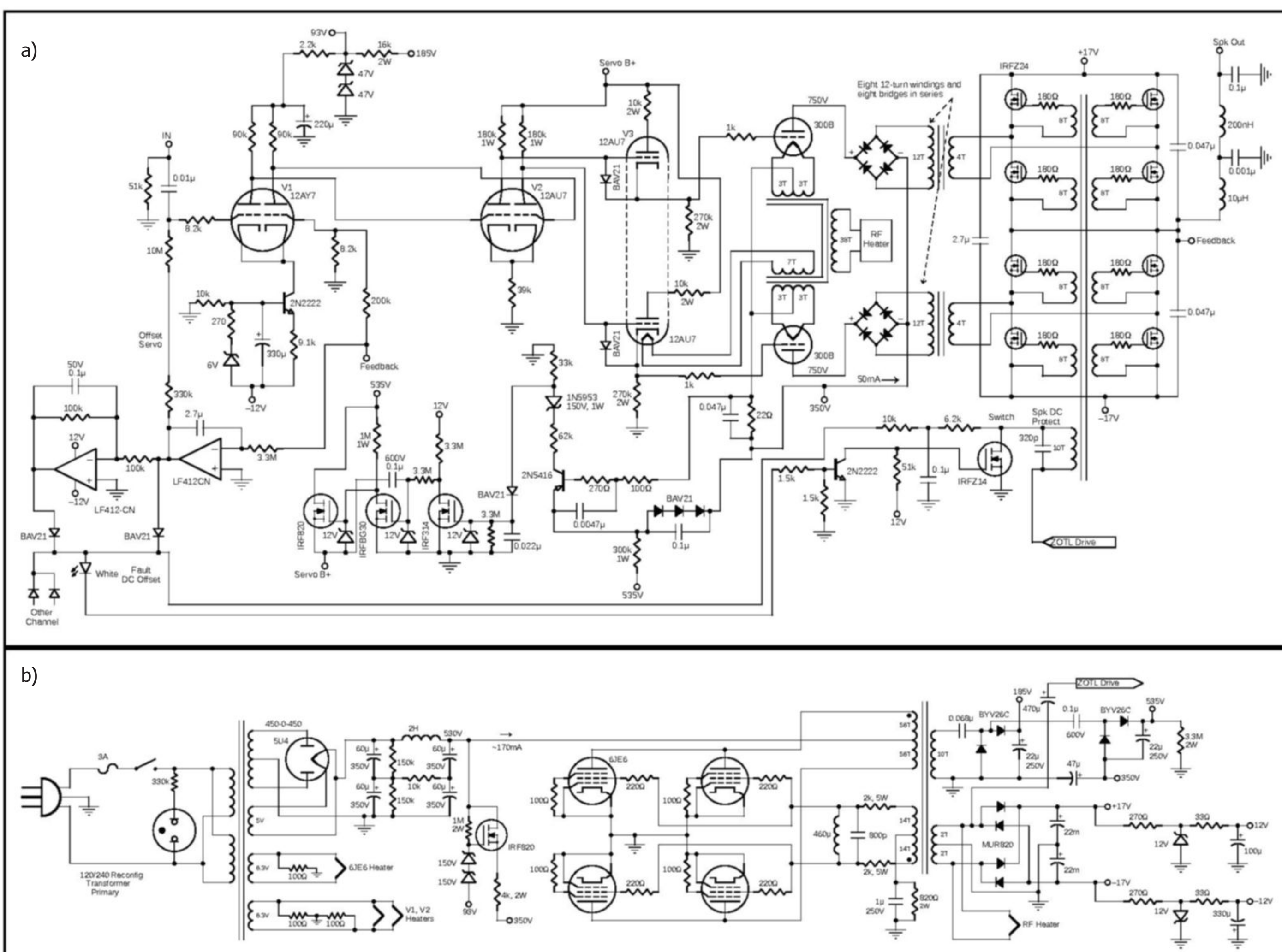

Figure 1: This is the schematic for the input power supply section (a); and the schematic for the audio and ZOTL section (b)

the 5U4 comes from a 5V secondary winding. The other two 6.3V windings provide filament power to the 6JE6 beam power tubes that are used to create a high-frequency switching power supply and the first two signal tubes.

The high voltage supply is conventional with a first section of two capacitors in series for handling higher voltages as filters with balancing resistors, a large 2H inductor and then another two series filter capacitors. Another resistor connects the center of the two output capacitors with the balancing resistors of the first stage.

Directly off the positive voltage of the second stage capacitors is a voltage regulating circuit that supplies 350V to bias the output tube directly heated cathodes above ground.

Unlike most power supplies that would take the tube voltages from the rectified and filtered secondary power transformer windings, power that is supplied to the impedance converter, the filaments of the 300B output tubes and the 12AU7 driver tube, comes from the switching power supply that draws that power transformer derived DC and feeds it into the center tap of the primary of the switcher transformer. The operating frequency of this switcher is determined by the LC network across the transformer winding that feeds grid 2 of the 6JE6 tubes. It was just over 240kHz. The B+ voltage for the first stage signal gain tube comes from the power transformer voltage, but the second signal gain tube and driver B+ comes from a servo circuit that is used to provide symmetrical auto bias.

The ZOTL

The Zero-Hysterisis Output Transformerless (ZOTL) circuit is both a simple yet complicated design. It is essentially a bridge-configured switching supply with an externally derived fixed switching frequency. For each half of the push-pull circuit, there are four MOSFET transistors in a bridge configuration. Detailed descriptions of the operation are available on the Berning website, but here is a simple description. Looking at the section that corresponds to the top output tube of the push-pull configuration, each MOSFET is driven by a winding connected to a secondary winding of the main switching power supply transformer. The polarity of the windings on the gates of the MOSFETs is such that one side of the top of half of the bridge conducts, while the MOSFET on the bottom half of the other side of the bridge conducts. That results in current flowing one way through the primary of small transformer connected to the bridge rectifier array that provides B+ voltage to the output tube. The current that goes through that primary is the same current that goes through the speaker. The same process occurs on both the opposite side of the bridge and the lower bridge.

Since the current pulses that come from the bridge sections and are then filtered before reaching the speaker pass through the transformers that link to the output tubes, the impedance of the tubes is reflected to the switching power supply. If an output tube is fully off, the tube is in a high impedance state and the secondary of the coupling transformer reflects that high impedance to the primary, which is in series with the bridge operation. Therefore, the instantaneous pulse amplitude going to the speakers is very low. As the tube begins to increase conduction to follow the audio signal, more current can flow through the tube, reducing its impedance, which is reflected to the switching power supply increasing the voltage available to the speakers. This following of the impedance changes of the tube as it goes through changes with the varying audio signals is the mechanism used to provide impedance matching without the limitations of a wide bandwidth output transformer.

About the Author

Thomas Perazella is a retired Director of IT. He received a Bachelor of Science degree from the University of California, Berkeley campus. Audio has been his passion for more than 50 years and he is a member of the Audio Engineering Society, the Boston Audio Society, the Philadelphia Area Audio Group, the DC HiFi Group, and the DC Audio DIY Group. He has written for *Speaker Builder* and *audioXpress* magazines, authored several articles in professional audio journals, and taught commercial lighting at the Winona School of Photography. Recently, he received a patent on a cost-effective high-efficiency LED lighting system for commercial and residential buildings. He is also a Past President and Treasurer of the Rockville Chapter of the Izaak Walton League of America, one of the oldest national conservation organizations in the US.

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DC Servo

With a DC-coupled amplifier, close matching of the two differentially configured sections of the tubes and other components can yield reasonably low DC offset voltages at the output. However, any shifts in the parameters of the tubes or other components with age can result in high offsets that could reach 1V. In this design, there is an op-amp integrator that links to the same feedback line that controls the AC feedback of the second section grid of the input differential pair. That integrator ignores AC signals but as the frequency at the output goes down to DC, if a DC voltage is detected, it introduces an offset voltage to the first section grid of the differential input pair, bringing the DC to very low levels at the output. The second section of the op-amp functions as a speaker protection circuit. If the DC offset remains high due to a failure in any part of the circuit, it will turn off the MOSFET switch in the primary of the ZOTL drive winding, preventing the bridges from functioning and stopping any current from going to the speakers.

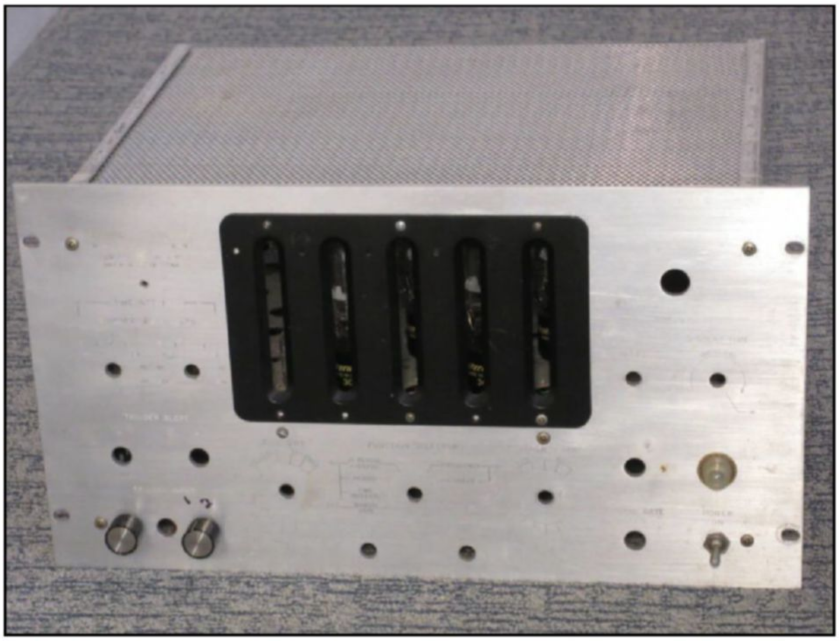


Photo 1: Here is the front of the original chassis.

Photo 2: This is a rear view of the original chassis, with the cover removed.

Auto Bias

The biasing circuit for the output tubes is also sophisticated. For balancing between the two tubes, since the whole internal circuit is DC coupled, any bias difference will be detected by the DC servo circuit and corrected. For the actual level of the bias, a 22Ω sensing resistor is in series with the return line of the two connected cathodes. However, the bias is not just measured under no signal conditions but while any level of signal is present. The voltage pulses across the resistor as the tubes operate are fed through the reference resistor and they are compared to a reference voltage.

If the minimum voltage across that reference resistor at any point goes too low, the 2N5416 PNP transistor turns on and feeds a signal to the three MOSFETS connected to it. Those transistors integrate the pulses and control the B+ to the two 12AU7 tubes. If the PNP is turned on because any pulse is too low, the B+ increases and the bias increases. If the PNP turns off because all the pulses are too high, the B+ will decrease and the bias will be reduced. Since there is signal integration, this adjustment is done in a way so that stability is maintained and hunting is eliminated.

As you can now see, there is a lot of sophisticated design in this amplifier with the goal of extracting the maximum performance possible.

Original Chassis Condition

With the original chassis, the power switch, pilot light, input selector switch, and volume control were mounted on the front panel. There were additional non-functional holes and a bezel with five vertical slots for the previous display.

As you can see from **Photo 1**, it was definitely not a chic look. From a rear view, you can see the layout of the transformer and tubes on top of the chassis along with the IEC power receptacle, output terminals, and input jacks on the rear. Not visible behind the transformer was an input power fuse holder. Again, there were lots of unused holes. A perforated cover enclosed the components on the top of the chassis (**Photo 2**).

Upon closer inspection of the chassis, I noticed that probably sometime during shipping, the amp had been dropped and the chassis under the power transformer had bent. As a result, the transformer was leaning forward and actually pressed on the 5U4 rectifier tube.

Underneath were two hand-wired circuit boards assembled on perforated board, a choke, power supply filter capacitors, and miscellaneous components directly wired to the tube sockets and terminal strips, as shown in **Photo 3**.

Next Month

So, there you have it. A unique amplifier design waiting to find a fitting home on a new chassis for its owner. That will be the subject of Part 2 of this article.

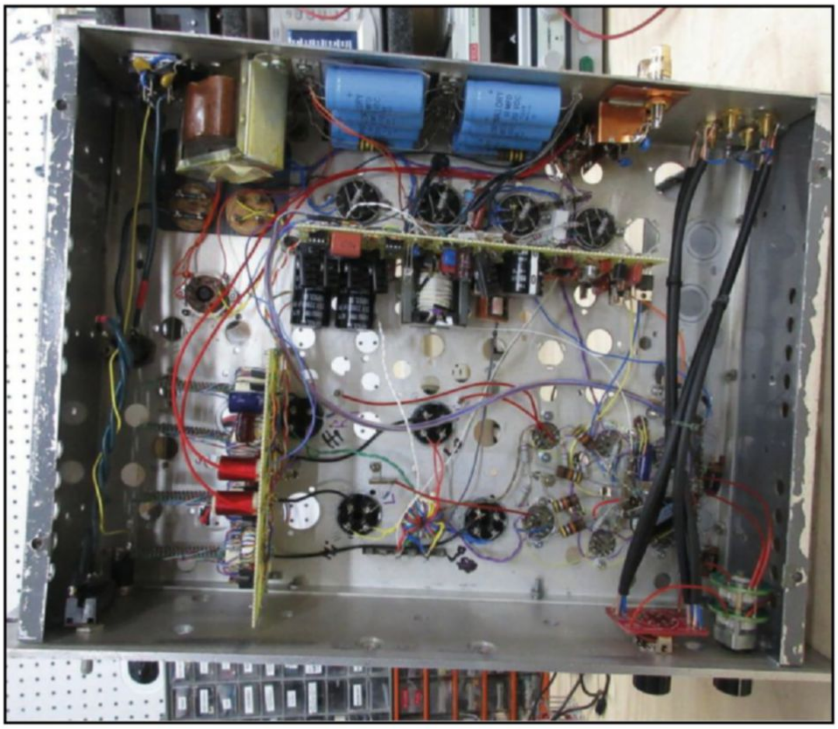
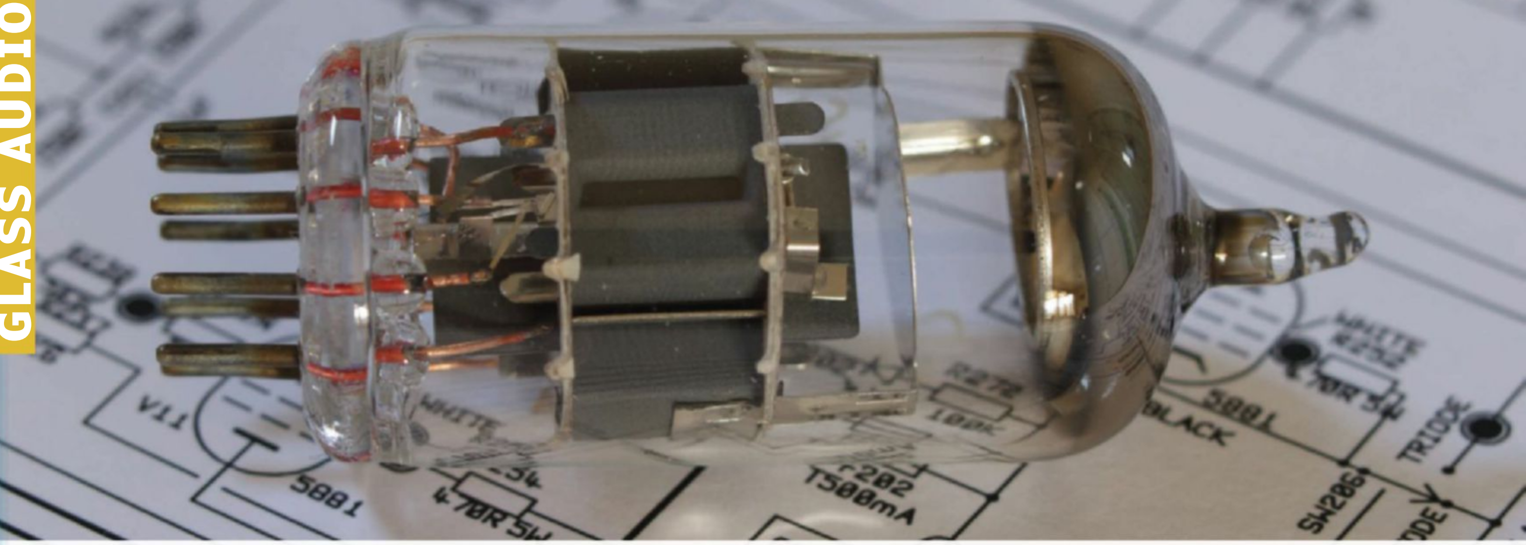


Photo 3: I was careful to document the wiring inside the original chassis, shown here.

GLASS AUDIO

Myths in Tube Circuit Designs



In the world of tube amplifiers, a number of myths and misunderstandings repeatedly show up. To mention just a couple—load (mis) match in output stages is by definition a bad thing; and the efficiency of a triode output stage is 25%, where a pentode shows 50% efficiency. In this article, Peter Dieleman will demystify and debunk them.

By
Peter Dieleman

Author's Note: A few times references are made to equation derivations from books that are no longer available on the market. They were written in the heroic days and 20 years ago. We make these derivations available in the Supplementary Material section of the audioXpress *website (see the Project Files link). As a rule, these mathematical derivations are "unfriendly," if not offensive.*

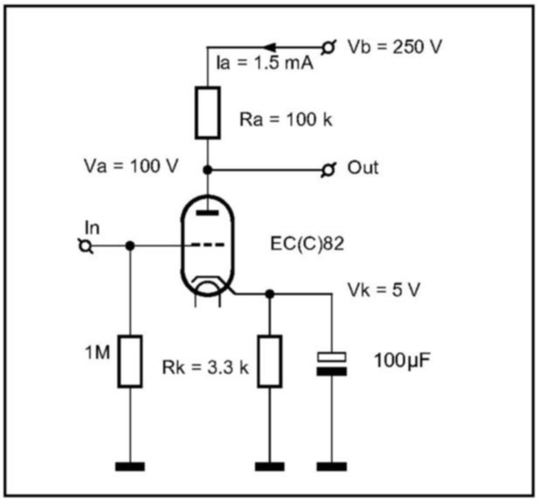


Figure 1: ECC82 with probable mismatch

Introduction

In the past century a lot of knowledge on the operation and design of tube circuits has been developed. Please realize that in the years 1945 to 1965 there were tubes only ("them phony transistor things" were not yet appreciated by everyone). After the decline in the use of tubes, replaced by the much more powerful and easier to handle transistors (you cannot drop a tube and go unpunished; you can drop a transistor), this knowledge has partly been forgotten.

But you don't need all this ancient knowledge to create a reasonably good tube circuit design. There are several good books and articles out there. However, it can be observed that from time to time statements are made by designers that are wrong. I have already mentioned two of them.

The first one refers to the so-called load match. For instance, output stages use an output transformer that has to fit the behavior of the output tubes. It usually does not do this completely. This is often marked as bad, but this opinion is not well founded. I will explain.

There are two excellent reasons to choose impedance mismatch. The first one is you happen to have a transformer and you are willing to use it. A load mismatch will result in a small loss of output power. Fact of the matter is that most likely you will not hear this small difference. The second reason is that you can choose mismatch in order to reduce distortion. This proves to be an extremely efficient weapon.

What Is Load (Mis)match?

Figure 1 shows an ECC82 (12AU7) circuit that you probably have seen many times.

The supply voltage is V_b = 250V. The anode resistance R_a is 100kΩ. The chosen current is 1.5mA. Therefore the anode voltage is V_a = 100V.

According to the characteristics shown in **Figure 2**, an anode voltage of 100V and an anode current of 1.5mA indicate that we need a cathode voltage of V_k = 5V, and hence the cathode resistor has a value R_k = 3.3kΩ. The cathode resistor is decoupled with an electrolytic capacitor (i.e., an elco) of 100μF.

Now look at the voltages shown in Figure 1. During the positive part of a sine wave input signal, we can pull the anode voltage down to 40V, assuming that we want at least 40V across the triode (neglecting V_k). This is an AC negative peak voltage of v_a = (100 – 40) = 60V.

During the negative part of a sine wave input signal the anode voltage can rise up to as much as 250V (the pinch-off situation of the triode). This is a positive AC peak voltage of v = (250 – 100) = 150V. There is a large asymmetry between the two signal parts. You may call this non-symmetrical, non-optimal or a mismatch.

Changing the current to 1mA would result in an anode voltage of V_a = 150V. This requires a cathode voltage of 9V, and therefore a cathode resistor of 9kΩ. We could then make a positive v_a swing of 100V and a negative one of also 100V, where V_k = 9V. Now this is symmetric and there no longer is a mismatch. The AC swing at the anode now is $200V_{pp}$.

The question is whether this is a better design. When we want to output a signal of $200V_{pp}$, it definitely is. When we want to amplify a microphone signal in the millivolt world, both circuits would do.

Thus, matching and mismatching do not exist in themselves, but they depend on the usage of the application. There is no such thing as an optimal match in itself. The concept of matching implies at least two actors.

Operating Points of Output Triodes

A First Design

Instead of discussing things, it is better to just design an output stage here and conclude a few things from the results. The whole purpose of a good output stage design is to pull as much power out of the tubes as possible. In this instance, we will use an EL34 (6CA7) as an example. However, our design freedom is limited by a few maximum values of the tube.

This EL34 tube is a pentode. It can be dressed as a triode by connecting G_2 to the anode. Also G_3 can be connected to the anode. With a 6L6, you can't because its G_3 is internally connected to the cathode.

In forum discussions, we see opinions whether to connect G_3 to the cathode, as usually it is the case with pentodes, or connect it to the anode.

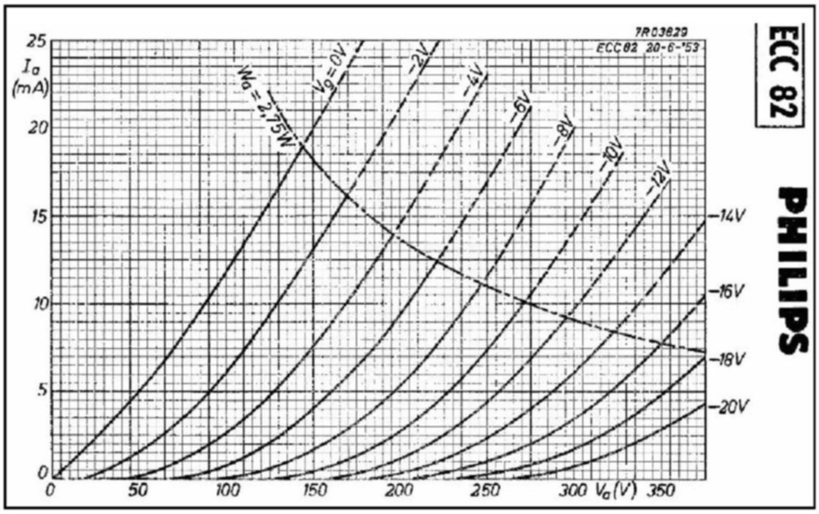


Figure 2: Anode characteristics of an ECC82 or 12AU7

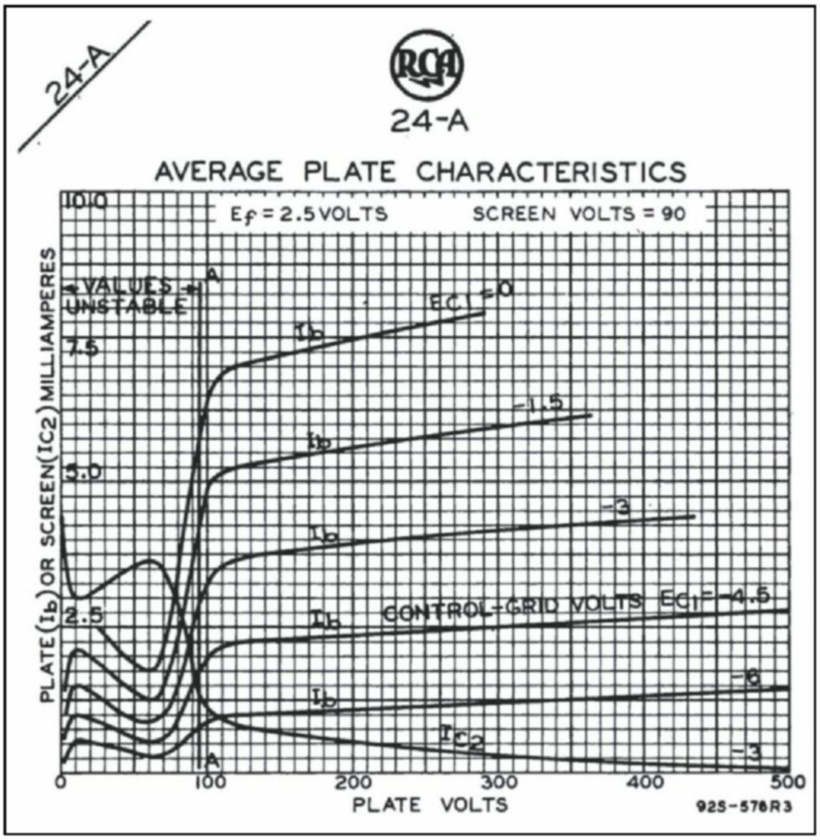


Figure 3: 24-A tetrode characteristics. Please note the funny anode current drops (known as kinks) when V_a < 100 V.

A tickling question is what made the EL34 designers decide to make G_3 available on pin 1? The answer is, that a separate G_3 connection is needed when not wanting to have the tetrode behavior (**Figure 3**), but the pentode behavior instead.

The physical distance between G_2 and G_3 is small. Therefore a maximum value of V_{g2} is specified.

A too high voltage on V_{g2} connected to the anode, where G_3 has cathode potential, may result in sparks. You would not want that. In our case, we don't have tetrode behavior, therefore, we do not need the healing influence of G_3. Let us lump all metal beyond G_1. In a pentode configuration, V_{g2} will not get that large anyway—usually is it fixed with an elco.

These characteristics, shown in **Figure 4**, are constructed by myself from earlier less readable graphs. I apologize for not giving the needed scientific credit. I do not know where I got this early material 25 years ago.

Step 1: Choose a dissipation and a quiescent V_{ak} voltage. I have chosen V_{ak} = 410V and $I_a = I_0$ = 75mA. This is a common choice. The quiescent point E is where, in Figure 4, the red, the blue, and the green colored lines intersect. The power consumption is P_a = 410 × 75 = 30.75W.

The specs (see the EL34 datasheet [1]), say P_a <= 24W and P_{G2} <= 8W. This adds up to 32W. But be careful, it is not guaranteed that with an anode current as given, that when the anode dissipation is 24W, the G_2 dissipation is less than 8W. I have chosen to limit the sum of these dissipations to 29W. The 30.75W would be somewhat larger than my chosen maximum for the EL34 triode, so this configuration may not be on the safe side. For the reasoning here, this is not relevant.

Step 2: Now the tube's anode current can drop from I_0 to zero. In a Class-A configuration, it can increase to 2 × I_0. This maximum value of course is reached when V_g = 0. This maximum current, according to the specs [1], is 150mA. The red load line, shown in Figure 4, is drawn from the point (410, 75) toward a current of 2 × 75 = 150mA. The point D of 150mA is chosen where the red line crosses the curve of V_g = 0.0V in the point (142.5V, 150mA). Extending it to the right side, it happens to hit the I_a = 0 line at point G, where V_a = 678V. We now know two values: the anode voltage swing is $v_{a\_}$ = (410-142.5) = 267.5V. The anode current swing is 75mA. But we can tell a lot more.

The power consumption is P_a = 410 × 75 = 30.75W. The power transferred to the output transformer is:

$$W_a = \frac{1}{2} \times (267.5 \times 75) = 10.03W$$

The factor ½ results from working with peak values instead of V_{eff} and I_{eff}. A push-pull output configuration would produce twice this value giving 20.1W.

The efficiency then is:

$$efficiency = \frac{10.03}{30.75} \times 100\% = 32.6\% \tag{1}$$

Busted this 25% myth!

The internal resistance of the EL34 (look at the blue line in Figure 4) in this configuration is:

$$r_i = \frac{(490 - 320) = 170V}{140mA} = 1.21k\Omega$$

The required transformer impedance is:

$$r_a = \frac{410 - 142.5}{75} = \frac{267.5}{75} = 3.6k\Omega$$

In a push-pull output transformer we will have:

$$r_{aa} = 2 \times r_a = 7.2k\Omega$$

This is somewhat larger than we usually find. An EL34 should have r_{aa} = 6kΩ, so they say.

Point E shows V_{gk} = -36V. Therefore, we can allow a sine wave input signal of $72V_{pp}$.

This brings us to point F, where V_{ak} = 597V.

Let's take a sneak preview at the distortion. The distortion by the second harmonic, according to the well known equation, is:

$$d_2 = \frac{1}{2} \times \frac{v_{a-} - v_{a+}}{v_{a-} + v_{a+}} =$$

$$\frac{1}{2} \times \frac{(410 - 142.5) - (597 - 410) = 267.5 - 187 = 80}{597 - 142.5 = 454.5} = 8.8\%$$

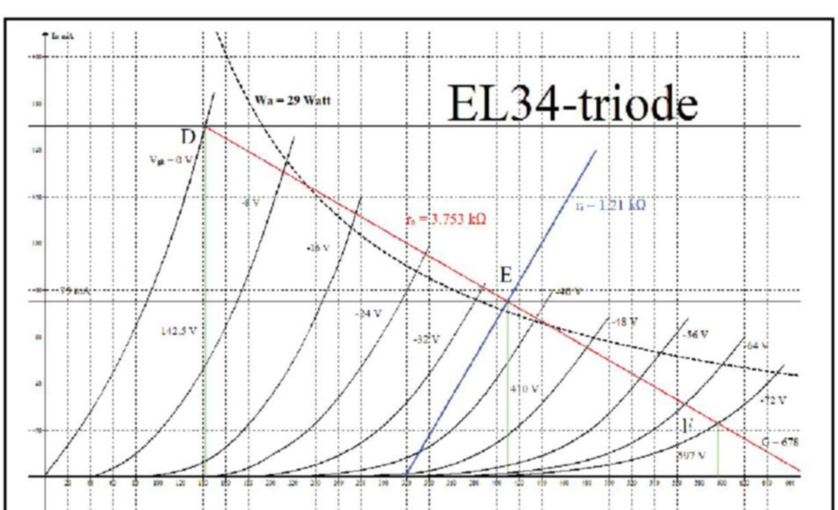


Figure 4: The EL34 with operation point V_a = 410V and $I_a = I_0$ = 75mA

In the nominator, we find the difference of the two anode signal half swing values. In the denominator we need the full swing value. The distortion is 8.8% when v_{aa} = 454.5V_{pp}.

You may find this distortion a bit too large. It will be shown in the sequel that this distortion will be nullified.

A Second Design

The design shown in **Figure 5** has a lower power consumption. Its V_a = 344V. Its I_0 = 70mA.

We can repeat all calculations: The red load line is drawn from the point (344, 70) toward the point D with a current of 2 × 70 = 140mA. Extending it to the right side, it happens to hit the I_a = 0 line at the point G, where V_a = 551V. We now know the anode voltage swing is v_a = (344 - 136.5) = 207.5V. The current difference is 70mA. The power consumption is P_a = 344 × 70 = 24.08W. The power transferred to the output transformer is:

$$W_a = \frac{1}{2} \times (207.5 \times 70) = 7.3W$$

The efficiency is then:

$$efficiency = \frac{7.3}{24.08} \times 100\% = 30.3\% \qquad (2)$$

Again, busted this 25% myth!

The internal resistance of the EL34 in this configuration is :

$$r_i = \frac{(420-270) = 150V}{140mA} = 1.07k\Omega$$

The required transformer impedance is:

$$r_a = \frac{344-136.5}{70} = 3k\Omega$$

With a push-pull output transformer we will need:

$$r_{aa} = 2 \times r_a = 6k\Omega$$

This is what we usually find. As generally stated, an EL34 should have r_{aa} = 6kΩ. As an example of what is available, the Amplimo/Plitron transformer VDV6040 has an r_{aa} = 6kΩ, while being able to handle an output power of as much as 40W.

The second harmonic distortion is:

$$d_2 = \frac{1}{2} \times \frac{(344-136.5)-(487-344) = 207.5-143}{487-136.5 = 350} = 9.2\% \qquad (3)$$

Again, this 9.2% occurs when v_{aa} = 350V_{pp}.

Using an Existing Transformer

In the previous design, I chose the quiescent anode voltage and derived the corresponding r_a. If your existing transformer is designed to provide a specific r_a , just change the order of things.

Choose the maximum value of I_a = 2 × I_0. Observe the corresponding V_{min} in point D (Figure 4).

Now take V = R × I or, v_a = r_a × I_0. Add this AC peak value to V_{min} and you will have the required value of V_a and V_b.

A Mathematical Model of a Class-A Triode

There is a lot of theory on the configurations of triodes. The needed theories are not derived here. The derivations can be found in [2], [3], [4] and in the Supplementary Material section on the *audioXpress* website or in one of the better books.

The equations cited here give:

A triode in a balanced push-pull output stage in Class A has an optimal so-called power match when r_a = 2 × r_i and thus, in a balanced stage r_{aa} = 4 × r_i. The maximum AC peak anode voltage is half the V_{ak} value.

The efficiency in general is:

$$W_a = \frac{1}{2} P_a \times \left(1 - 2 \times \frac{P_a \times r_i}{V_{ak}^{\ 2}}\right) \qquad (4)$$

This last equation is quite unknown. P_a is the anode dissipation for the triode. W_a is the power transferred to the transformer and thus to the loudspeaker. These values combined give us the efficiency.

So, the triode's efficiency also has a maximum of 50%, just like the pentode. This is in contrast to what is often published, and I have already busted this myth. We approach this maximum efficiency by combining a high value of V_{ak} and a triode with a low r_i.

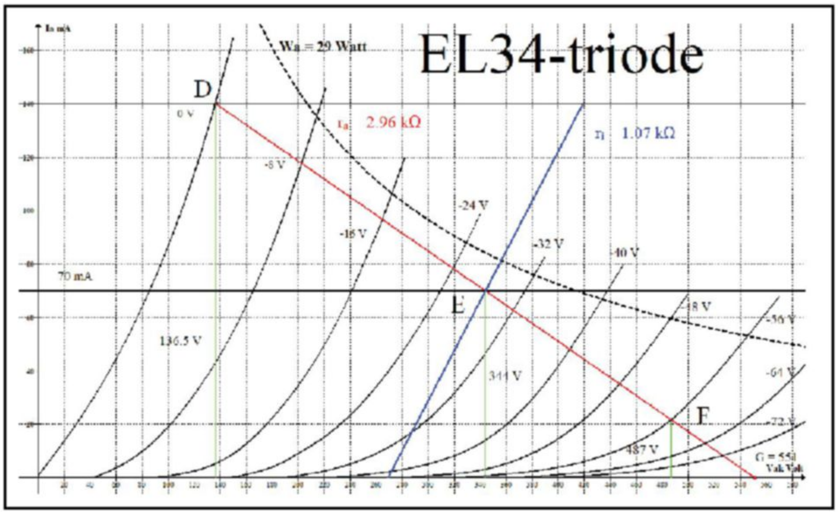


Figure 5: EL34 is shown at 344V and 70mA.

Putting two triodes in parallel to get half the value of r_i does not solve a thing because the value of P_a between the two parentheses doubles. The product of P_a and r_i remains the same! We really need a triode with a low r_i by design.

We can now substitute the values found in Figure 5 into Equation (4) and see what we get:

$$W_a = \frac{1}{2} P_a \times \left(1 - 2 \times \frac{P_a \times r_i}{V_{ak}^2}\right) \quad (4)$$

$$W_a = \frac{1}{2} \times 24.08 \times \left(1 - 2 \times \frac{24.08 \times 1.07}{344^2}\right) = 6.8W$$

The efficiency is 28.2%.

Working with the characteristics, which is always more accurate than any model, we saw that the efficiency was 30.3%.

For r_a we found r_a = 3.0kΩ and for r_i we found r_i = 1.07kΩ. The theory said $r_a = 2 \times r_i$ .The v_a swing should have been v_a = ½ × V_a or, 172V. It is 207.5V. The conclusion is that the theoretical model is not that sound. It is derived from an idealistic representation of the curves from the characteristics. Stick to the approach of working with the characteristics; they are always correct.

The Influence of Mismatch on the Distortion

Cancelling Out Even Harmonics

First of all, balanced output transformers have a beautiful, unexpected property. Most designers know that the magnetic fields caused by the two DC anode currents cancel out each other.

What is less known is that they also cancel out the even (second and fourth …) harmonics. They add up the odd harmonics. A triode produces mainly even harmonics, which sound as octaves of the fundamental. After the cancellation, there is only minimal distortion left. This is the strong card for triodes when we want a low distortion. The price is a lower efficiency.

The Influence of R_a/r_i on Distortion

There are two less known equations about the influence of the ratio R_a/r_i on distortion. Once more, these theories are not derived here. The derivations can be found in [2], [3], [4] or on the *audioXpress* website or in one of the better books.

These equations are:

$$d'_2 = \frac{d_2}{\left(1 + \frac{R_a}{r_i}\right)^2} \quad (5)$$

and

$$d'_3 = \frac{d_3}{\left(1 + \frac{R_a}{r_i}\right)^3} \quad (6)$$

In these equations d_2 and d_3 are respectively the second and the third harmonic distortion levels when R_a = 0. (In this situation, we can measure the distortions in the anode current). These distortions are large because we deal with real triode curves. They are the results of the triode's 3/2 power law. To convince yourself, substitute R_a = 0.

The left-hand side numbers with an accent indicate the distortions that result, after applying an anode resistor R_a. These distortions can be significantly lower.

A triode mainly has second harmonic distortion. A triode also has a low r_i . The triode benefits from a large R_a; this gives a high value in the denominators of Equation (5) and Equation (6).

A pentode with its third harmonics distortion has a high r_i, which is written as r_p in this article. (Its name, traditionally, is plate resistance). The value of r_p is usually much larger than R_a, therefore pentode distortion does not benefit from a larger R_a.

A Computational Example

To see the healing property of an impedance mismatch, at first, we take R_a = 50kΩ and after that R_a = 200kΩ. For simplicity we state that in these two cases we have the same r_i = 10kΩ.

With R_a = 50kΩ we see:

$$d'_2 = \frac{d_2}{\left(1 + \frac{50}{10}\right)^2} = \frac{d_2}{6^2} = \frac{d_2}{36}$$

With R_a = 200kΩ we see:

$$d'_2 = \frac{d_2}{\left(1 + \frac{200}{10}\right)^2} = \frac{d_2}{21^2} = \frac{d_2}{441}$$

Increasing R_a results in a decrease in distortion of 441/36 = 12.25 times.

Project Files

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

References

[1] EL34, https://tube-data.com/sheets/010/e/EL34.pdf

[2] P. Dieleman, *Theorie en praktijk van buizenversterkers*, Segment B.V. 2004.

[3] P. Dieleman, *Théorie et pratique des amplificateurs audio à tubes*, Segment B.V. Elektor, 2005.

[4] P. Dieleman, *Theorie und Praxis des Röhrenverstärkers*, Segment B.V. 2004.

I_o	V_{min}	V_a	V_{max}	v_a	r_a	W_a	P_a	eff	d_2	r_i	Figure
75	142.5	410	597	267.5	3.6	10.05	30.75	32.6	8.8	1.21	4
70	136.5	344	487	207.5	3.000	7.3	24.08	30.2	9.2	1.07	5
50	111	450	682	339	6.78	8.5	22.5	37.8	9.4	1.4	

Table 1: Comparing three designs.

Taking a closer look, we may conclude that increasing R_a wins from the increase of r_i. Unfortunately, R_a cannot be increased infinitely. When we do this, we get a too high DC voltage drop across this resistor. The anode current drops and we end up in a position in the characteristic where r_i has increased. To maintain the same value of r_i, we need a higher supply voltage.

The moral is that when choosing a small power mismatch, a too high r_a will result in some loss of power, but at the same time in a reduced distortion. Mismatch is not bad per se.

Finally, **Table 1** shows the comparisons of the designs. (I added a third design with a lower current, i.e., a high r_a. There is no corresponding figure for this).

Conclusions

The maximum power was obtained by extending the EL34 beyond its limits. The minimum power and the maximum differ by 73%; one can distinguish this by ear. The efficiency is always larger than 25%; so that myth is also busted. The distortion is always too large, but the even-order harmonics will be cancelled out to a large extend in a push-pull stage.

About the Author

Peter Dieleman is now 71 years of age, retired but still active every day, consulting for the Dutch Government. During the past 10 years he has become a specialist on IT structures for safety, management, and exploitation of tunnels.

At age 17, he went to study electric engineering ('Electrotechniek') at Delft Technical University in the Netherlands. He then earned his living as a television repair man (tubes in those days). At age 30, he gradually switched to digital electronics because television sets got better and there was less need for repairmen. He was a Lecturer at Groningen and Twente universities in the Netherlands for about 25 years. And, he has always kept his love for tubes alive. In 2003, he wrote his first book on the theory and practice of tube amplifiers. In 2009, he finished his second book on tube designs like SRPPs, mu-followers, and OTLs. The first books was translated into German and French; the second one into German.

Vintage Op-Amp Noise Simulation

Part 2 — Calculations and Noise Sources

By
Burkhard Vogel

In Part 1 of this article, Burkhard Vogel presented the deviations from realistic noise figures in datasheets, based on measurements and calculations for real op-amps. In this second and final part, the findings are expanded through the use of Mathcad calculations, culminating in an accurate Spice model for the noise sources, and a conclusion about noise source correlations.

Let's begin Part 2 of this article with a discussion about voltage noise. Taken from Chapter 15 of my book *Slopes and Levels – Spice Models to Simulate Vintage Op-Amp Noise*, the Mathcad worksheet MCD-WS 1 in **Figure 9** shows the calculated graphical result for NE5534AN's voltage noise density. **Figure 10** presents the simulation outcome, generated with the **Figure 11** simulation arrangement. At the right side of Figure 9, we also find copies of the calculated numbers at selected frequency points, rounded to one digit after the decimal point.

There is no big difference between the numbers in Column C of Table 1 and the ones of Figure 9. Further succ-apps attempts won't drastically improve the RMS noise content in B1k, B20k, and B100k. MCD-WS 1 tackles this issue too. Thus, the simulated voltage noise curve shown in Figure 10 equals the ones of Figure 6 (detailed in Part 1, *audioXpress,* April 2023) and Figure 9.

Current Noise

Taken from Chapter 15 of my book, Mathcad worksheet MCD-WS 2 in **Figure 12** shows the calculated graphical result for NE5534AN's current noise. **Figure 13** presents the simulation outcome, generated with the **Figure 14** simulation arrangement. At the right side of Figure 12, we also find copies of the calculated numbers at selected frequency points, rounded to one digit after the decimal point.

There is no big difference between the numbers in Column C of Table 2 and the ones of Figure 12. Further succ-apps attempts won't drastically improve the RMS noise content in B1k, B20k, and B100k. MCD-WS 2 tackles this issue too. Thus, the simulated current noise curve shown in Figure 13 equals the one shown in Figure 7 (detailed in Part 1, *audioXpress,* April 2023) and Figure 12.

Further down in the section about the correlation of the Current Noise Sources and in MCD-WS 2's Section 5, I also went through the calculation concerning the correlation of the current noise sources in the Texas Instrument (TI) model. The result: The two noise sources seem to be 100% un-correlated.

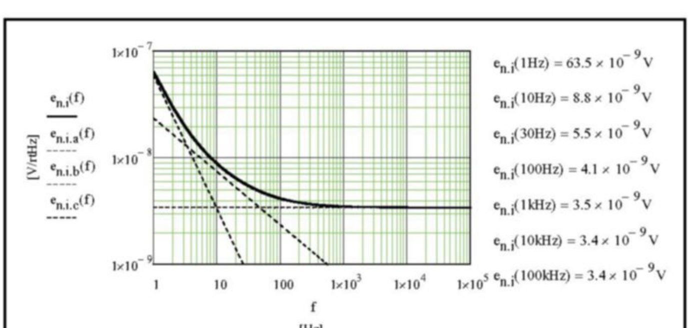


Figure 9: This is the calculated spectrum of NE5534AN's voltage noise (solid), including tangents (a, b, c - dotted) and levels.

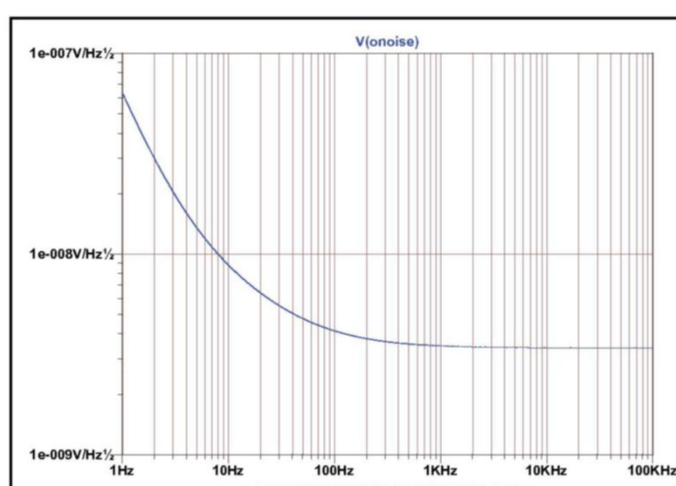


Figure 10: Here is the simulation result of (1) and Figure 9 and Figure 11; V(onoise) in V/rtHz (left ordinate) at point "oe" of Figure 11.

Further Material of Noise Traces

It's not easy to get good measurement results of OPA noise. That's why I asked Paul Horowitz and Winfield Hill for permission to also show Figure 8.110 and Figure 8.111 from their book, *The Art of Electronics*, Third Edition. The figures provide a range of measured voltage noise density and current noise density traces.

The caption to Horowitz and Hill's Figure 8.110 (shown here as **Figure 15**) states:

"Measured voltage-noise-density spectra of a selection of op-amps. The boldface parts are auto-zero amplifiers which circumvent the 1/f "flicker-noise" demon by repetitive offset correction; the italic OPA627 is a JFET-input op-amp. See also Figures 8.60 and 8.61."

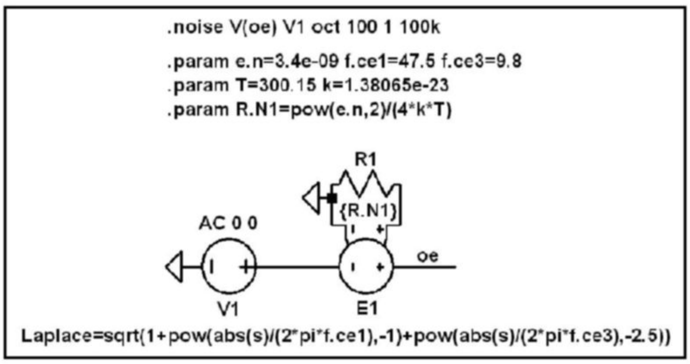


Figure 11: This is the simulation content of NE5534AN's voltage noise generator.

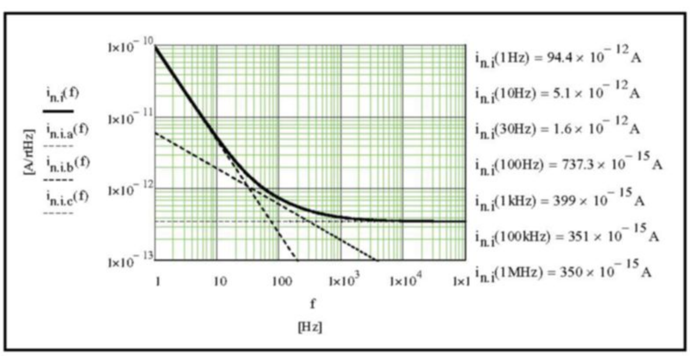


Figure 12: Here is the calculated spectrum of NE5534AN's current noise (solid), including tangents (a, b, c - dotted)) and levels.

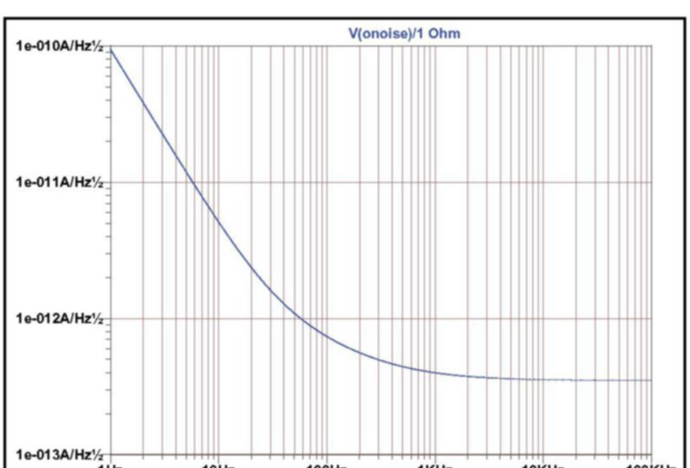


Figure 13: This is the simulation result of (2) and Figure 12 and Figure 14; V(onoise)/1Ω in A/rtHz (left ordinate) at point "oi" of Figure 14.

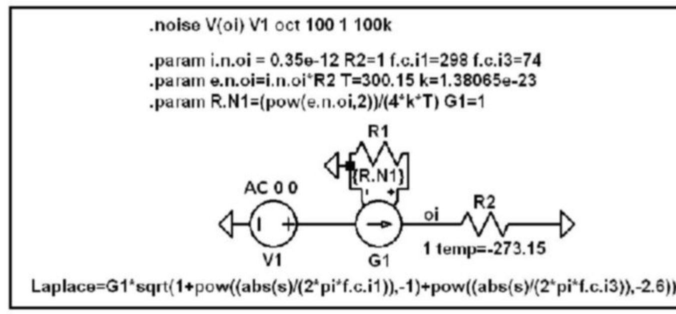


Figure 14: This is the simulation content of NE5534AN's current noise generator.

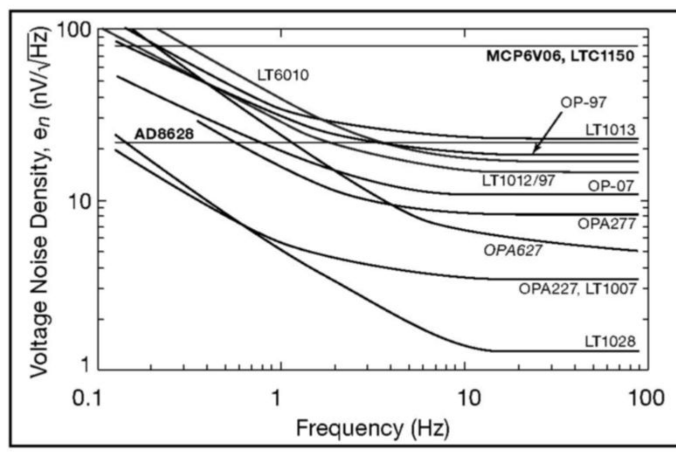


Figure 15: This graph shows the measured voltage noise density spectra of a selection of new and older OPAs from Paul Horowitz and Winfield Hill's book, *The Art of Electronics*, Third Edition.

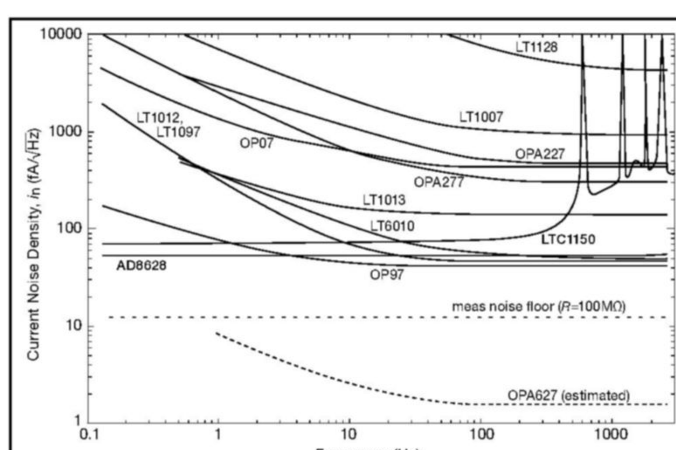


Figure 16: This graph shows the measured current noise density spectra of a selection of new and older OPAs from Paul Horowitz and Winfield Hill's book, *The Art of Electronics*, Third Edition.

The caption to Horowitz and Hill's Figure 8.111 (shown here as **Figure 16**) states:

"Measured current noise density spectra for most of the op-amps of Figure 8.110. The boldface parts are auto-zero amplifiers, which circumvent the 1/f "flicker-noise" demon by repetitive offset correction, but exhibit switching induced clock noise at higher frequencies (see the expanded plot in Figure 5.52). See also Figures 8.60 and 8.61."

There is one problem that I could not fix. In Figure 16 the top trace is supposed to be for an LT1128. This OPA is an LT1028 with less GBW. Concerning noise, the datasheet shows no differences. Additionally, Figure 8's trace A presents the white noise level according to the datasheet, roughly 1pA/rtHz. Hence, because of its 100% correlated current noise sources, this trace could represent the one generated by fully unequal input loads.

The Final Simulation Model for the New NE5534AN

By following the findings in the before-mentioned section about Current Noise and the next one "What About the Correlation of the Current Noise Sources of the Original TI Model," the non-existance of a correlation of the current noise sources becomes a fact. Thus, this new NE5534AN model is based on three independent noise sources without any correlation. N at the end of all my noise-adapted OPA's indications stands for my manipulation to get datasheet-reflecting input noise for simulation purposes.

If we study the internal circuitry of the 5534 [8] shown in **Figure 17**, we do not see any indication of what we can find for example in the circuitries of the LT1028/LT1128 or OP27/37: bias-current-cancellation, a measure that leads to correlated input current noise sources. "This cancellation circuitry injects two correlated current noise components into the two inputs." [10], [11]

NE5543AN's internal circuitry is shown in **Figure 18**. Based on Figure 11 and Figure 14, **Figure 19** and **Figure 20** show the content of the noise sources plus their netlists. Finally, **Figure 21** gives the symbol of the newly created model.

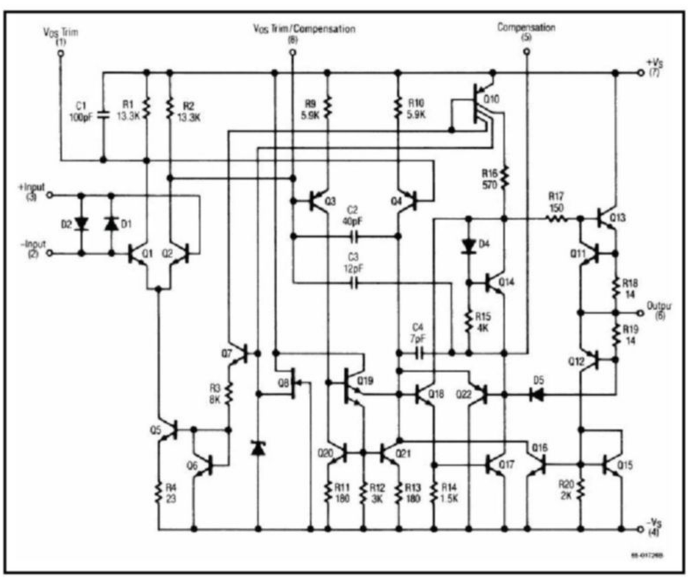


Figure 17: This is the internal circuitry of the type 5534 OPA.

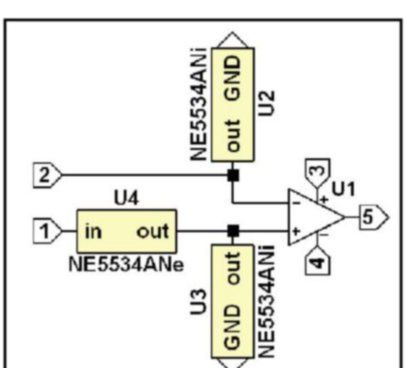


Figure 18: The internal circuitry of the NE5534AN is shown here.

Netlist NE5534AN:

```
.subckt NE5534AN 1 2 3 4 5
XU1 N001 2 3 4 5 level.2 Avol=0.1Meg GBW=15Meg
  Slew=13Meg ilimit=25m
+ rail=0 Vos=0 phimargin=45
+ en=0 enk=0 in=0 ink=0 Rin=500Meg
XU2 0 2 NE5534ANi
XU3 0 N001 NE5534ANi
XU4 1 N001 NE5534ANe
.lib C:\Program Files (x86)\LTC\LTSpiceIV\OPAs\
  NE5534AN\Noise current 5534A.net
.lib C:\Program Files (x86)\LTC\LTSpiceIV\OPAs\
  NE5534AN\Noise voltage 5534A.net
.lib UniversalOpamps2.sub
.ends NE5534AN
```

Netlist voltage noise generator:

```
.subckt NE5534ANe in out
E1 in out N001 0 Laplace=sqrt(1+pow(abs(s)/(2*pi*f.
  ce1),-1)+pow(abs(s)/(2*pi*f.ce3),-2.5))
R1 N001 0 {R.N1}
.param e.n=3.4e-09 f.ce1=47.5 f.ce3=9.8
.param T=300.15 k=1.38065e-23
.param R.N1=pow(e.n,2)/(4*k*T)
.ends NE5534ANe
```

Netlist current noise generator:

```
.subckt NE5534ANi GND out
R1 N001 0 {R.N1}
G1 0 out N001 0 Laplace=G1*sqrt(1+pow((abs(s)/
  (2*pi*f.ci1)),-1)+pow((abs(s)/(2*pi*f.ci3)),-2.6))
.param i.n.oi = 0.35e-12 R2=1 f.ci1=298 f.ci3=74
.param e.n.oi=i.n.oi*R2 T=300.15 k=1.38065e-23
.param R.N1=(pow(e.n.oi,2))/(4*k*T) G1=1
.ends NE5534ANi
```

The symbols for the noise sources U2, U3, and U4 (2 × NE5534ANi + NE5534ANe) are already presented in Figure 18.

What About the Correlation of the Current Noise Sources of the Original TI Model?

Concerning the headline's question, the answer is as follows:

=> Practically, there is no correlation!

The results of the application of the following equation, including the correlation factor (CF) [12]:

$$e_{n.o1}(f,CF) = \sqrt{e_n(f)^2 + i_{n1}(f)^2 R1^2 + i_{n2}(f)^2 R2^2 + 2CFi_{n1}(f)i_{n2}(f)R1R2} \quad (3)$$

via its simplification by setting $i_n(f) = i_{n1}(f) = i_{n2}(f)$:

$$e_{n.o1}(f,CF) = \sqrt{e_n(f)^2 + i_n(f)^2 \left(R1^2 + 2CFR1R2 + R2^2\right)} \quad (4)$$

and with the data gained in the second section of Part 1 (*audioXpress*, April 2023), the results of the simulation of TI's original NE5534 à la **Figure 22** show practically no difference. MCD-WS 2 and its Section 5 gives the details. Additionally,

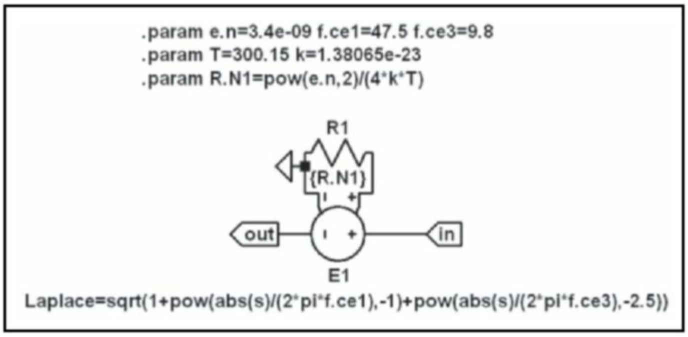


Figure 19: Here is the NE5534AN's voltage noise generator

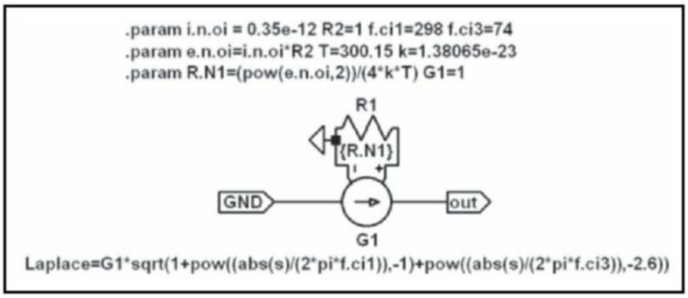


Figure 20 The NE5534AN's current noise generator is shown here.

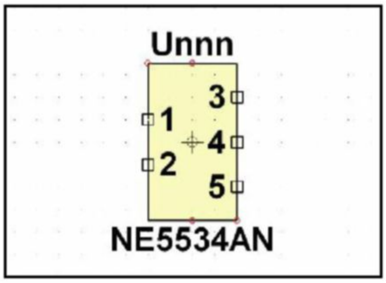


Figure 21: This image is the symbol for the NE5534AN.

Figure 23 shows the evolution of Equation (4) with CF = -1 … 0 … +1 (red). The crossing with the black curve, which is the constant value of TI model's voltage noise density in B100k, leads also to the fact, that there is practically no correlation between the current noise sources. In the course of the calculation, the tiny difference between the calculated and the simulated results comes from not taking into account the input resistance of the OPA.

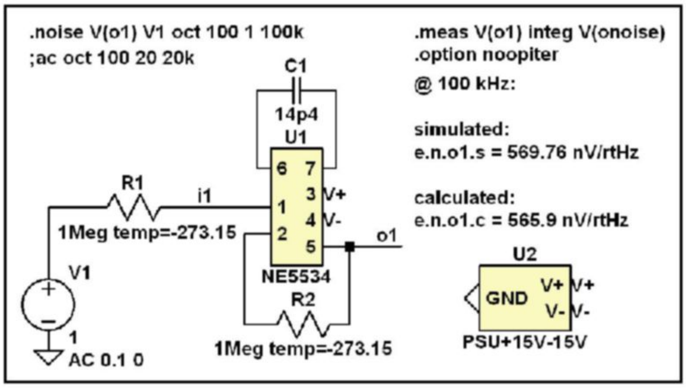


Figure 22: This simulation schematic is used to check the OPA's correlation behavior.

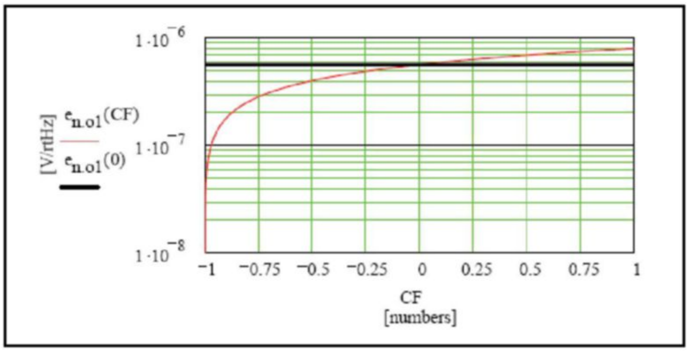


Figure 23: Calculated output of TI's model (black) vs. the CF-dependent trace of equation (4) is shown here.

About the Author

Burkhard Vogel obtained his academic degree Dipl.-Ing. (Telecommunications) in 1973 from Darmstadt University of Technology, Germany. He then decided to follow a general management career path in the high-tech and IT industry in Germany, Switzerland, and Austria. In the recent decades fruitful discussions on tube audio and electronic noise topics with customers, employees, colleagues and friends led to a broad range of works on these issues. They were published in several magazines including *Electronics and Wireless World* (EWW), *Linear Audio* (LA), and *Design & Design* (D&D), together with many additional comments in EWW, LTEs in EWW, LA, and D&D. Finally, Burkhard has seven books edited by the Springer Nature publishing house (currently number 8 is "under construction") that round up his gained knowledge.

What About the OPA Input Resistance Rn?

At the end of the second line in the netlist of the NE5534AN, we find the differential input resistance Rn=500Meg. Although the datasheet indicates an input resistance of 100kΩ, I kept the shown value.

There are two reasons for that. First, with a practically infinite input resistance the math without input resistance will be less complex, and second, the difference between the two values does not lead to significant differences in the input referred voltage noise densities. Even with the lower value, with the Avol shown, and with, for example, a gain of 100 the resultant differential input resistance is always > 10MΩ in a GBW of 100kHz.

Thus, I recommend using the datasheet's Rn value only in cases of BJT-based inputs with extremely low input resistances ≤ 50kΩ and Avol values ≤ 50,000.

However, I also recommend using the datasheet's given Rn values in all cases of JFET-based inputs with their extremely high input resistances of Rn ≥ 1T = 10^{12}Ω.

The Gain-of-Three Question

TI's simulation model does not show any measures to limit the range of gains to ≥ 3 à la the datasheet. It also works fine with all gains < 3. Therefore, I've treated the proposed NE5534AN solution the same way.

Project Files

Two Mathcad Worksheets can be found in the Supplementary Material section of the *audioXpress* website: MCD-WS 1 (voltage noise) and MCD-WS 2 (current noise)
To download the Mathcad worksheets, visit:
http://audioxpress.com/page/audioXpress-Supplementary-Material.html

References

[10] "1990 Linear Databook," Linear Technology Corp., 1989.

[11] The remark shown in Figure 8 (a) underneath the graphs: nearly all shown OPAs have bias-cancellation circuits, exceptions are G (5534) and Q (AD8021).

[12] C. D. Motchenbacher and J. A. Connelly, *Low-Noise Electronic System Design*, Wiley, 1993, p.21

Correction for Part 1

In the April 2023 issue of *audioXpress*, one of the figures in "Vintage Op-Amp Noise Simulation—Part 1 Deviation from Realistic Noise Figures in Datasheets," by Burkhard Vogel, was incorrectly labeled. The Y-axis was shown as V/RtHz$^{1/2}$, that should have been A/RtHz$^{1/2}$. See the corrected Figure 2 below. Also, in Figure 5, the top Y-axis legend of 100dB was missing, although obvious from the scale values. The online and PDF editions of the April 2023 issue have been updated with the corrected figures.

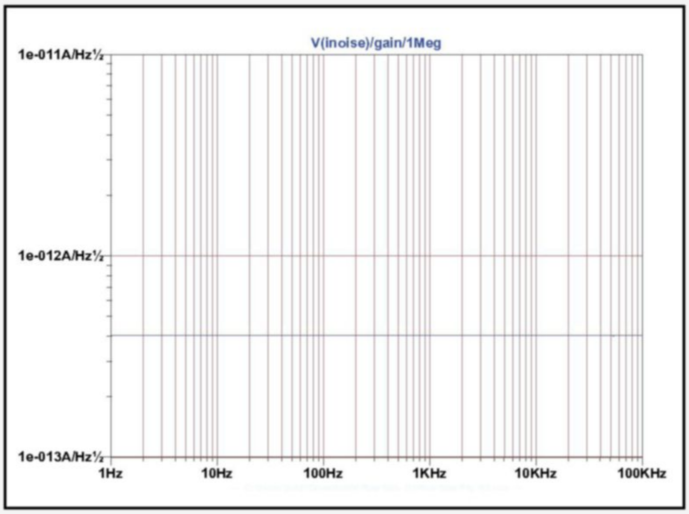


Figure 2: Current noise trace of TI's NE5534A simulation model

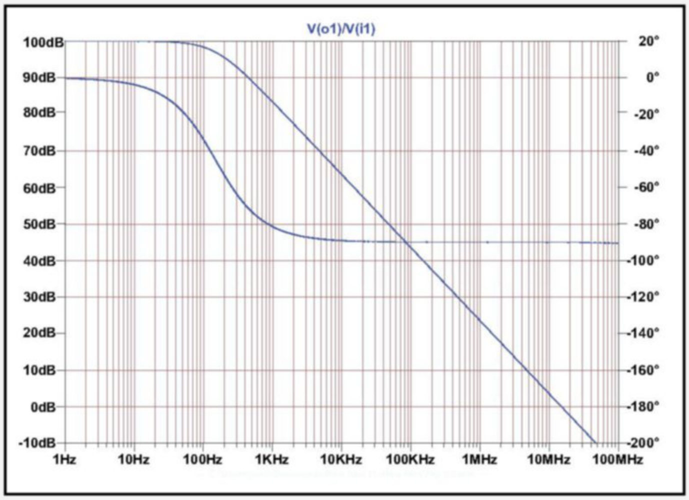
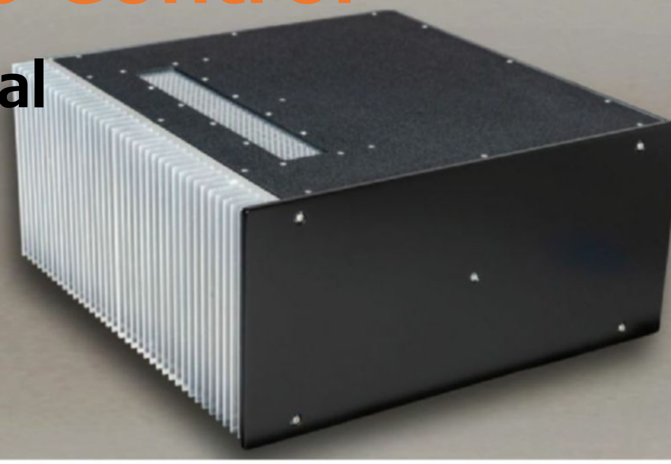


Figure 5: Open-loop gain with Test OPA-M

Monoblock Power Amplifier with Precision Bias Control



Part 2—Circuitry, Mechanical Design, and Measurements

In Part 2, Bruce Gillingham discusses the remainder of the MB1's circuitry, its mechanical design, and its performance measurements.

By
Bruce E. Gillingham

In Part 1, I provided an overview of my MB1 amplifier and discussed its audio path in detail. However, there is more to a successful power amplifier than just the audio path design. Proper power supplies are required and there should be a monitoring/protection function to prevent, or at least minimize, damage to the amplifier and speaker in the event of abuse or component failure. The complete power amplifier is a system. Good system design involves making design trades just as is done with the basic amplifier design, but at a higher level. The mechanical design also provides a contribution to success. Let's start here with the power.

AC Switch

AC power enters the chassis through an IEC inlet filter mounted on the rear panel. The hot side is protected by a 5A panel-mounted slow-blow fuse.

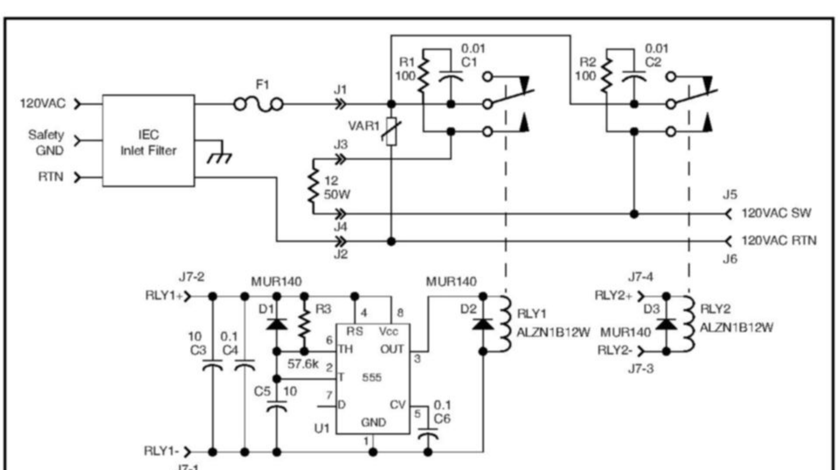


Figure 6: AC power enters the chassis through an IEC inlet filter mounted on the rear panel. This schematic shows the AC Switch circuit card assembly (CCA).

On/Standby switching is provided by the AC Switch circuit card assembly (CCA) (**Figure 6**). It contains two high-current (16A) power relays controlled by the Low Voltage Regulator/Protection CCA.

When 12V is applied to the RLY1 inputs, the output of timer U1 goes High, pulling in RLY1. This connects 120V AC to the main power transformer though an off-card 12Ω, 50W power resistor. The resistor reduces the initial turn-on surge to 10A rms maximum. After a delay of ~320ms, 12V is applied to the RLY2 inputs. RLY2 pulls in and bypasses the series resistor. There is a secondary surge but it is less than the initial one. After another ~330 ms, U1 times out and its output goes Low. RLY1 drops out, disconnecting the resistor as it is no longer needed. This reduces power dissipation in the 12V regulator and protects the power resistor in the event of a failure of RLY2 or its driver circuitry. R1/C1 and R2/C2 are snubber networks to reduce contact arcing and electromagnetic interference (EMI) emissions. VAR1 is a surge protector.

High Voltage Power Supplies

The ±65V DC supplies are unregulated, consisting of a 35A bridge rectifier with 40,000µF of filter capacitance per voltage. Careful consideration was given to charging and ground return currents in the design of the Capacitor Bank PCB. Andrew C. Russell's Ovation e-Amp filter board was used as a starting point [16].

It is designed with a "T" grounding scheme and separate charging and load buses. The buses overlay each other on opposite sides of the PWB. The ground traces are replicated on both sides of the PCB. Not all ground terminals were used in the MB1 as the final design has a secondary star ground on the Output CCA.

The ±75V DC supplies are on the High Voltage Regulator CCA. They have a full-wave voltage doubler fed from the main transformer and two-stage regulation. **Figure 7** shows the positive regulator. The negative regulator is identical except for component polarities.

The pre-regulator is a simple open-loop emitter follower (Q1) with a Zener diode string (Z1-Z3) in the base biasing. It reduces the voltage to 90V, shares power dissipation with the final regulator, improves performance, and provides a safety function. Should the final regulator fail, the pre-regulator will hold the output voltage to a level that the IPS and VAS can survive.

Final regulation is performed by a discrete linear regulator. Q2 and Q4 form a Darlington series pass transistor. Q3 is configured as a current source to bias the Darlington. The current source improves input ripple rejection and maintains a high impedance at the base of Q4, which maximizes the open-loop gain of the regulator. Q6 is the error amplifier. It compares Z4's voltage with a fraction of the output voltage. R11 allows adjustment of the output voltage and C8 provides loop compensation for the regulator. Q5 adds short circuit protection.

Floating Power Supplies

Figure 8 shows the floating power supplies used by the Bias Controller. The ±6.2V outputs are produced by simple shunt Zener regulation (Z2, Z3). The Zeners are fed from current sources referenced to the ±75V regulated supplies used by the IPS and VAS. Q1 and Q2 form the positive current source; Q3 and Q4 the negative. Potentiometer R1 balances the two current sources to minimized the current into, or out of, the ground reference. Q1 and Q4 are mounted close together to improve thermal tracking. Ideally, they would be in the same package or cemented together. Zener Z1 is a fail-safe, in the event one of the other Zeners fail.

Low-Voltage Power Supplies

Figure 9 shows the three low voltage DC power supplies on the Low Voltage Regulator/Protection CCA. All are sourced from the off-card idle power transformer. Its two 15V AC windings are connected in series, creating a 30V AC center-tapped input to the CCA. This is full-wave rectified by BR1 and filtered by C8 and C13, producing approximately ±24V.

Linear regulators U1, U2, and U4 drop this to +12V, +15V, and –15V, respectively. The three regulators are mounted to on-card heatsinks. The +12V regulator is always on, producing "12V." A switched version, "12V_SW," is produced elsewhere on the card. Resistors R1 and R3 reduce the power

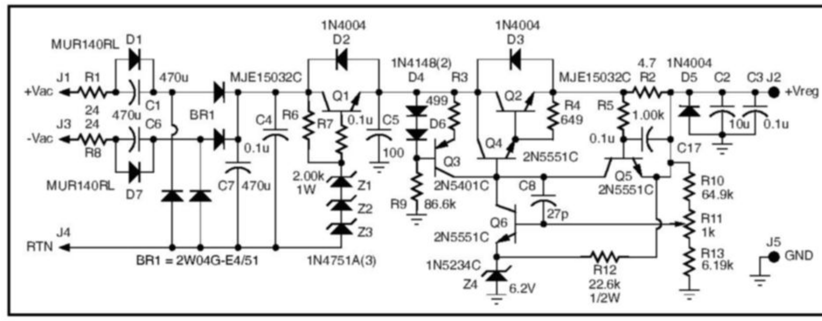


Figure 7: Here is the positive high voltage regulator. The negative regulator is identical, except for component polarities.

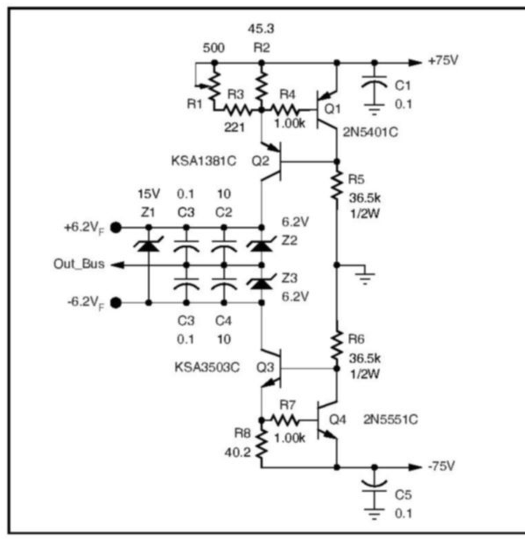


Figure 8: This schematic shows the floating power supplies on the Driver CCA used by the Bias Controller.

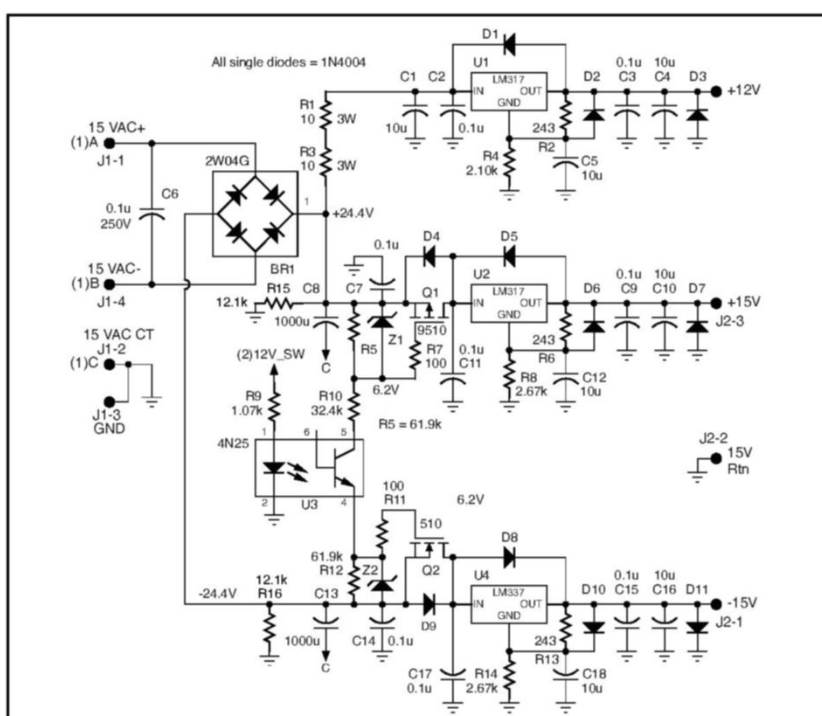


Figure 9: Shown here is the schematic for the three low voltage DC power supplies on the Low Voltage Regulator/Protection CCA.

dissipation in U1 under high current load conditions. The unregulated input voltages to the ±15V regulators are switched on and off by FETs Q1 and Q2. These are controlled by the presence or absence of 12V_SW via optocoupler U3.

Control, Monitoring, and Protection

Clipping/distortion and DC fault detection circuitry is on the Driver CCA. The processing of detected faults from the Driver CCA and the other functions discussed in this section are on the Low Voltage Regulator/Protection CCA.

AC Power Control

Figure 10 shows the AC power control circuitry. Putting the Power switch in the up (On) position, or applying 5V DC to 12V DC across the Trigger inputs with the Power switch in the center (Trigger) position, applies current to the base of Q4 from the always on +12V. Q4 turns on and, in the absence of a thermal fault, pulls the gate of FET Q1 down, turning it on. This enables 12V_SW.

The 12V_SW energizes relay RLY1 [on the AC_SW CCA], which pulls in and connects 120V AC to the main power transformer through a series resistor. This limits the turn-on surge (soft start). The application of 12V_SW also enables timer (2)U1a. When U1 times out in approximately 315ms, its output goes High, turning on Q3, energizing RLY2 [on the AC_SW CCA]. This provides a direct path from the 120 VAC to the main power transformer.

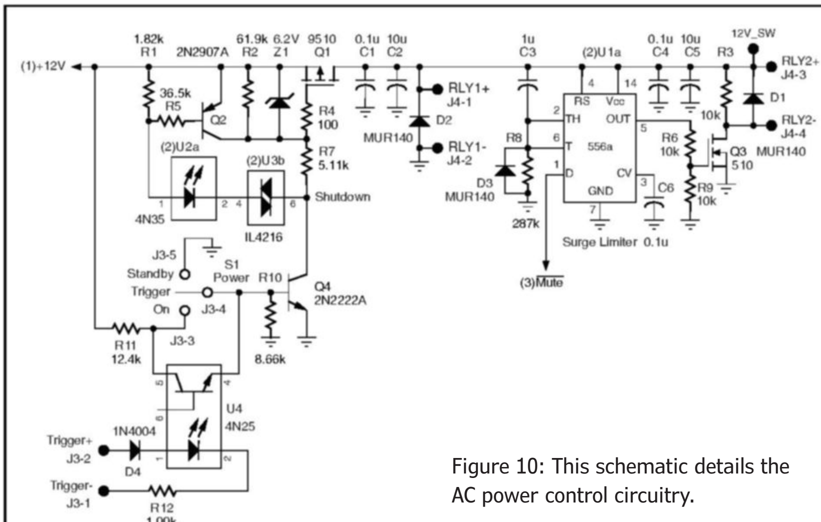


Figure 10: This schematic details the AC power control circuitry.

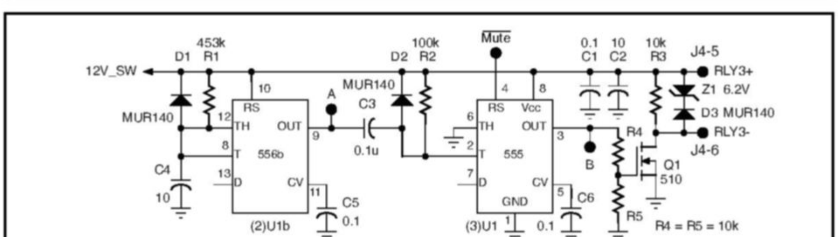


Figure 11: The speaker turn-on delay timer and loudspeaker protection flip-flop are described in this schematic.

The discharge pin of (2)U1a is Low during the timing interval. This holds the Mute(–) node Low, prohibiting speaker connection. This provides a redundant Power On Reset (POR) function and protects the speaker and amplifier if (2)U1a fails to time out. The discharge pin goes High at the end of the timing interval, releasing the Mute(–) node.

When an over-temperature condition is detected, the TRIAC in optocoupler (2)U3b is turned on and latches if Q4 stays on. The current through the TRIAC turns on Q2, depriving Q1 of its gate drive. Q1 turns off, and 12_SW rapidly decays, turning off the speaker relay, ±15V DC, and AC to the main transformer, which, in turn, shuts off ±75V DC and ±65V DC. The current through the TRIAC also energizes the LED in optocoupler (2)U2a. (2)U2's associated transistor is used for front panel display control.

If a Standby request is made during a thermal fault, Q4 turns off. This removes the gate drive to Q1 keeping 12V_SW Low, unlatches the TRIAC, and turns U2a's LED off. If the thermal fault is still active, an attempt to transition to Power-up mode will fail because the TRIAC will be on when Q4 turns on. When the thermal fault clears, the amplifier will be able to transition to Power-up mode.

Speaker Control

Figure 11 shows the speaker turn-on delay timer, (2)U1b, and loudspeaker protection flip-flop, (3)U1. As 12V_SW goes High, flip-flop (3)U1 is held in reset by the POR circuit, which keeps the Mute(–) node Low for approximately 52ms. Any uncleared fault will continue to hold this node Low after the POR pulse goes away. When reset, the flip-flop's output is Low, Q1 is off, and there is no voltage across the speaker relay connections (RLY3+, RLY3–).

When 12V_SW goes High, the trigger pin of timer (2)U1b is initially held Low by C4. This drives the output (Node A) High. It stays High until C4 charges to the threshold voltage through R1 (~5s). When the threshold is surpassed, (2)U1b's output is driven Low. This transition is differentiated by C3/R2, applying a negative going pulse to the trigger input of flip-flop (3)U1. If Mute(–) is High when the pulse occurs, the flip-flop is set. (3)U1's output goes High, Q1 turns on, and power is applied to the off-card speaker relay. If Mute(–) is Low when the pulse occurs, the flip-flop is not set. Since the (2)U1b timer will not re-trigger without cycling the

About the Author

Bruce E. Gillingham retired in 2018 after a 45-year career with a major US defense contractor. With an education in electrical engineering, he began his career as an analog and RF designer and progressed to systems engineering, providing technical leadership for numerous development programs. Bruce has been designing, building, and listening to his own preamplifiers and power amplifiers since high school. Early efforts were with tubes; most have been solid state. His other hobbies include listening to live and recorded music (mostly classical), photography, and wood/metal working. Bruce is married and lives on Central Florida's East Coast.

power on and off, the flip-flop will stay in reset and the speaker disconnected.

Power On Reset (POR)

The purpose of the POR (**Figure 12**) is to ensure that the speaker control flip-flop (3)U1 has a hard Low on its Reset pin at 12V_SW power-up. This function is implemented by comparator U1a. Point A is biased at 6V, which holds U1a's output and the Mute(–) line Low until 12V_SW goes High and C2 charges to > 6V. This takes approximately 52ms. After the brief delay, U1a's output goes High. Note, however, that several open collector devices are wire-ORed at the (3)Mute(–) node, so this point may still be held Low.

AC Dropout Detect

AC power line dropouts are detected by the circuitry in **Figure 13**. Diodes D3 and D4 monitor the AC input to the CCA. When present, the full wave rectified AC turns on Q1 which discharges C3 and drives the output of U1b High. If the AC drops out, Q1 turns off and C3 charges rapidly to the 6V threshold through R4. This causes U1b to pull Mute(–) Low. This latches the speaker output Off before any of the supply voltages sag excessively. The loss of a single cycle of 60Hz (16.7ms) triggers the turn-off.

DC Fault Detect

The DC fault detection circuitry on the Driver CCA is shown in **Figure 14**. R4 and C2 form a low-pass filter to attenuate audio frequency signals from the output bus waveform. The residual DC is monitored by the U1a/U1b window comparator. When the DC exceeds ±0.7V, a fault is declared (output Low). This illuminates LED1 and removes the pull-up from J3-1 that is connected by cable to the Low Voltage Regulator/Protection CCA. Diodes D2 and D3 protect the capacitor and comparator inputs from over voltage.

Comparators U1c and U1d monitor the ±65V DC rails applied to the Driver CCA. If the magnitude of either of these has not yet risen above, or later falls below, 40V, a fault is declared. This catches low line, blown rail fuses, and excessive load current. This fault is wire-ORed with the DC offset fault so it also illuminates the LED and removes the pull-up from J3-1.

Clipping/Distortion Detect

Figure 15 shows the clipping/distortion detection circuit on the Driver CCA. Instrumentation amplifier U2, an INA111, measures the voltage between the bases of the amplifier's input differential pair. Since the amplifier has a finite open

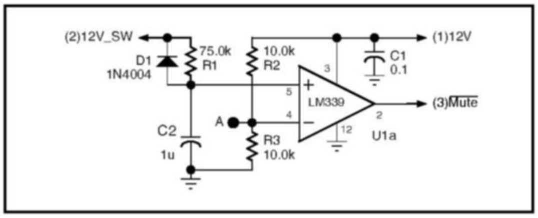


Figure 12: The purpose of the Power On Reset (POR) is to ensure that the speaker control flip-flop has a hard Low on its reset pin.

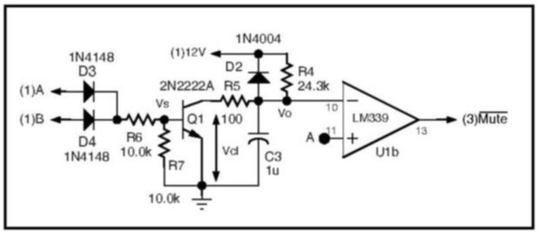


Figure 13: AC power line dropouts are detected by this circuitry.

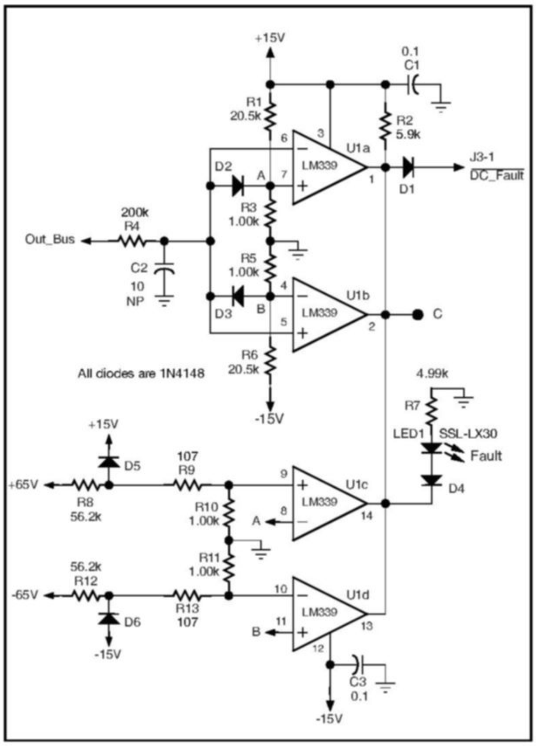


Figure 14: This is the DC fault detection circuitry on the Driver CCA.

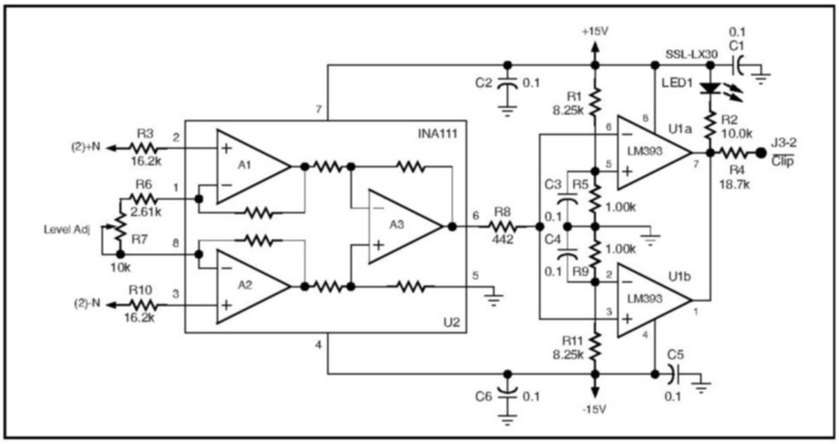


Figure 15: This schematic details the clipping/distortion detection circuit on the Driver CCA.

loop gain (~550x), there will be an "error" voltage across its inputs whenever it has an output voltage. This error voltage is proportional to output level and does not represent an undesirable condition. If, however, the amplifier goes into clipping or some other strong distortion condition, the error voltage will rise dramatically and this can be detected by the U1a/U1b window comparator.

R7 is adjusted so the highest anticipated normal output level does not trip the comparators. When clipping is detected LED1 illuminates and connector J3 Pin 2 goes Low. This pin is connected by cable to the Low Voltage Regulation/Protection CCA for processing. During brief non-repetitive faults, illumination of board-mounted LED1 may not be noticeable to the eye.

Driver CCA Fault Processing

The circuitry shown in **Figure 16** on the Low Voltage Regulator/Protection CCA monitors the two fault lines from the Driver CCA and processes the information for front panel display and amplifier/speaker protection. The DC Fault(–) line goes Low if the Driver CCA detects ±65V under voltage or excessive DC offset on the amplifier's output. This, in turn, drives U2b's output Low, muting the speaker. The Clip(–) line goes Low if the Driver CCA detects excessive error voltage across the bases of the input stage differential transistor pair. Timer U1 is wired as a one shot, extending fault pulses to a minimum of ~0.5 seconds. Its output "Clip" turns the front panel indicator to red when High. With Clip High, C5 is released to charge via the current source built around Q1. This current source is on whenever Clip(–) is Low. The switched current source and C5 form an integrator. If C5 charges to 6V by one or multiple fault pulses before it is reset by Clip going Low, U2a's output will go Low, turning the speaker off. This is an "excessive clipping" event.

Thermal Shutdown

The Thermal Shutdown circuitry is shown in **Figure 17**. Heatsink temperature is monitored by sensor U2, which is mounted on the main heatsink. Comparator U1d compares U2's output with a voltage derived from stable voltage source Z1. A heatsink temperature > 70ºC drives the output of U1d Low. This sinks current through the LED of optocoupler U3a, turning on the associated TRIAC (2)U3b. This initiates Thermal Shutdown mode. U1d being Low also sinks current through the LED U2a. The associated transistor is used for front panel display control. R15 applies hysteresis to comparator U1d. This keeps U1d's output Low until the sensor U2 drops below 60ºC. Capacitor C4 prevents erroneous initiation of Thermal Shutdown mode when the amplifier is first plugged into the 120V AC power source.

Status Indication

The front panel status display is a single tricolor LED, but it can convey a considerable amount of information. Control is performed by discrete diode-

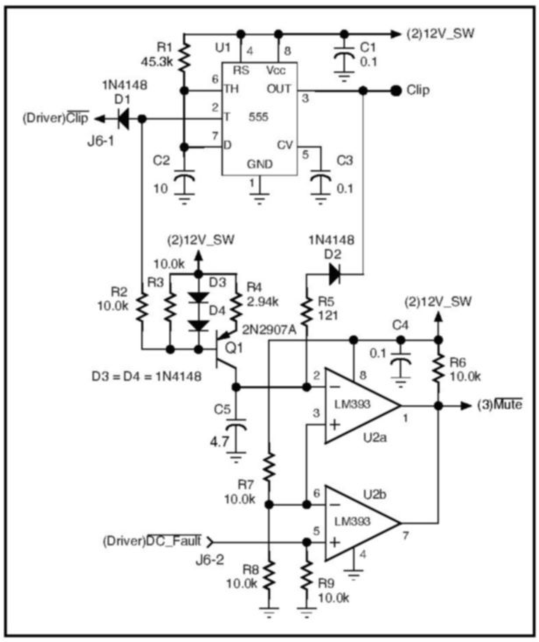


Figure 16: This circuitry on the Low Voltage Regulator/Protection CCA monitors the two fault lines from the Driver CCA.

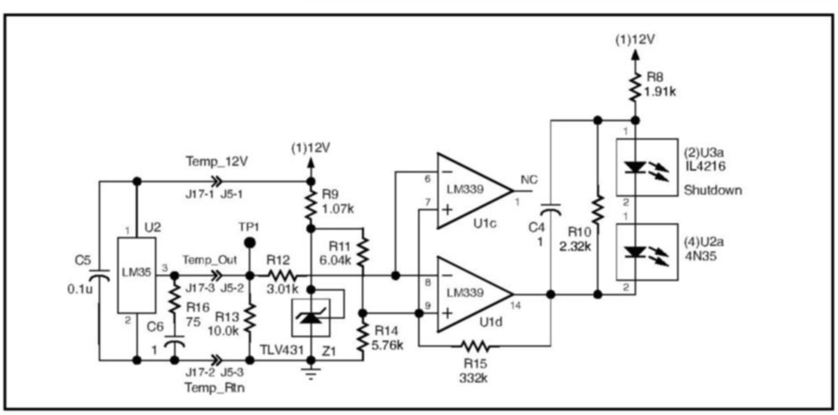


Figure 17: This is the Thermal Shutdown circuitry.

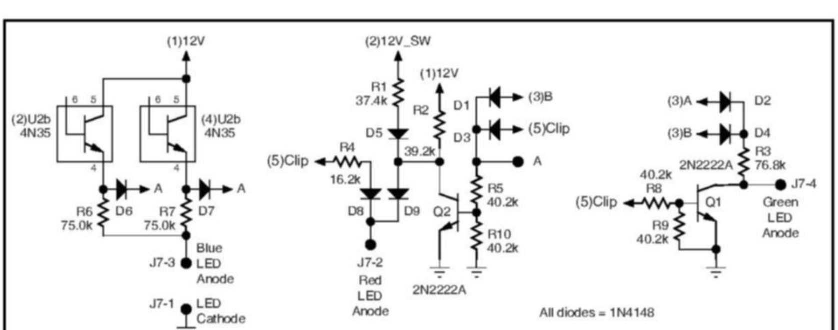


Figure 18: Control is performed by the discrete diode-transistor logic.

transistor logic shown in **Figure 18**. The red LED is illuminated at low-intensity in Standby mode by R2 and D9 from the always on 12V supply.

When 12V_SW goes High, additional current is provided by R1 and D5, illuminating the red LED at high- intensity during Power-up Mode. Both of the current sources can be shunted to ground by Q2, turning the red LED off. Q2 is turned on by (3)B in Normal mode, or by (5)Clip during clipping. The red LED is illuminated at high intensity by (5)Clip by R4 and D8.

The green LED is turned on at high-intensity by (3)A during Power-up mode. Along with the high- intensity red LED, a "yellow" indication is given. During Normal mode, the green LED is illuminated at high-intensity by (3)B. Q1 turns the green LED off during clipping events by (5)Clip.

The blue LED can be illuminated at low-intensity by either (2)U2b or (4)U2b. If they are both on together, the blue LED will illuminate at high-intensity. Diodes D6 and D7 turn off the red LED when there is a thermal fault. The green LED goes off when +12V_SW is turned off in Thermal Shutdown.

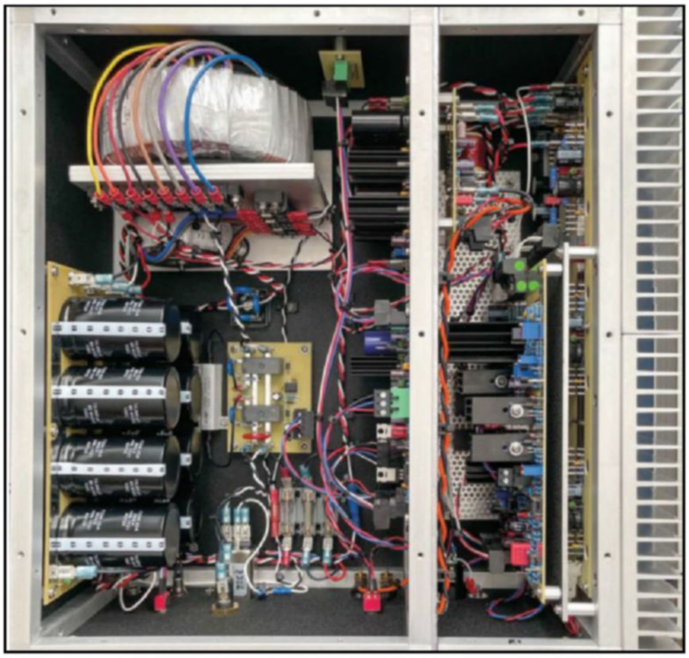


Photo 5: The MB1 is shown from above with the top cover removed.

Mechanical Design

MB1's chassis is constructed from 1/8" thick rectangular aluminum plates. All four edges of each plate are fastened by stainless steel machine screws to 1/2" × 1/2" solid aluminum bars. This forms a very sturdy chassis. **Photo 5** shows the MB1 with the top cover removed showing the internal construction.

The plates and bars were cut using a carbide- tipped aluminum cutting blade on a table saw. Tapped (threaded) holes are used in the bars for seldom, if ever, removed screws. All other holes have threaded stainless steel inserts. Most fasteners are flat head screws in counter sunk holes. All plates are completely painted on both sides with the exception of areas that contact the unpainted bars. This is to maintain a good electrical bond between the various chassis pieces. The plates were sanded with an orbital sander and then painted with a black wrinkle automotive paint and baked in an oven to get a uniform wrinkle. The wrinkle paint hides machining marks and improves cooling via radiation. The bars were just sanded.

A "decorative" 1/4" thick secondary front panel (not visible in Photo 5—see Photo 2 from Part 1 in the April issue of *audioXpress*) is bolted to the front of the chassis with four #6 cap head screws. The cap head screws sit flush in counter-bored holes. The decorative front panel is finished with black satin enamel paint. The MB1's front panel is 15" × 7" high. It has a centered hole for the indicator LED to shine through. The finished amplifier is 16.65" deep, including rear handles.

The chassis has an internal 1/8" thick partition to separate the audio circuitry from 60Hz AC, transformers, and power supplies. The audio compartment is the smaller one and has one side (left from the front) fabricated from a commercial extruded aluminum heatsink, rather than a flat plate.

The audio compartment is ventilated by rectangular cutouts on the top and bottom covers. The openings are covered on the interior sides with 1/16" thick perforated aluminum sheet. The output power transistors, Output CCA, and, Driver CCA are all mounted on the heatsink, creating a functional amplifier module. Power supplies, speaker interface, and protection circuitry are mounted elsewhere. The module's CCAs are on threaded standoffs. The Driver CCA is separated from the Output CCA by a 1/8" aluminum plate that serves as a shield. The amplifier module is shown in **Photo 6**.

Another internal design feature is the vertical mounting of the two toroidal power transformers. An upside down T-bracket is used. It mounts to the chassis bottom panel with three 1/4" machine bolts. Vibration isolation is provided by neoprene shoulder and flat washers. Vertical mounting was chosen to optimize the use of the internal volume and keep the transformers as far as possible from the input stage of the amplifier.

The four 1/2" high feet were fabricated from a 2" diameter solid round aluminum rod. Rubberized cork pads were bonded to the exposed sides of the feet. The feet attach to the bottom plate with two #6 screws. A centered, #10-24, threaded hole in each foot allows it to be attached to an amplifier stand.

Performance Details

Square Wave Response

The MB1's large-signal (80 V_{pp}) 10kHz square wave response into a 4Ω load is shown in **Figure 19**. The 10-90% rise and fall times

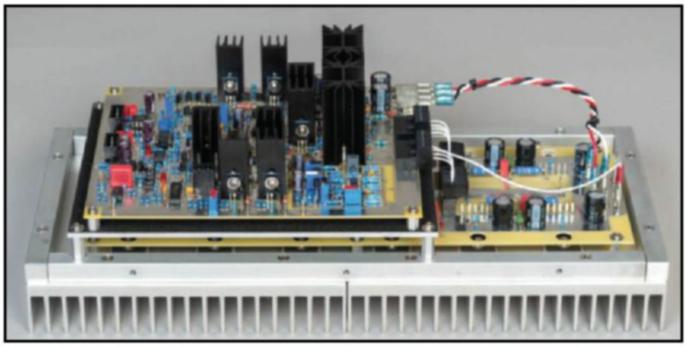


Photo 6: Here is the MB1's amplifier module.

are approximately 3µs with no overshoot, ringing, or "funnies," despite the ±10A current swings.

Clipping Behavior

Figure 20 shows the MB1 driven with a 10kHz sine wave into clipping with an 8Ω load. Clipping is clean with no sign of "sticking."

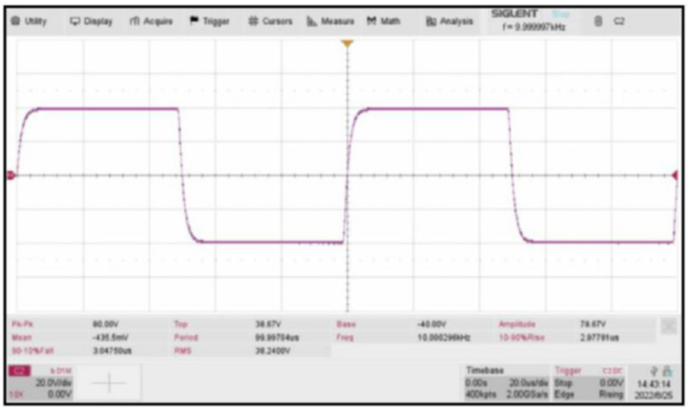


Figure 19: This graph shows the MB1's large-signal 10kHz square sine wave into 4Ω.

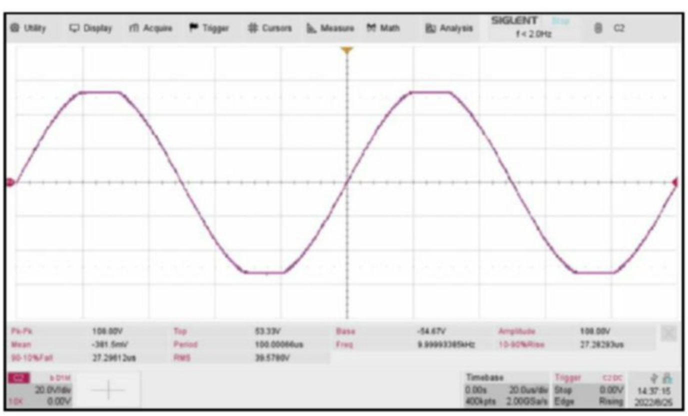


Figure 20: The MB1 is being driven with a 10kHz sine wave into clipping with an 8Ω load.

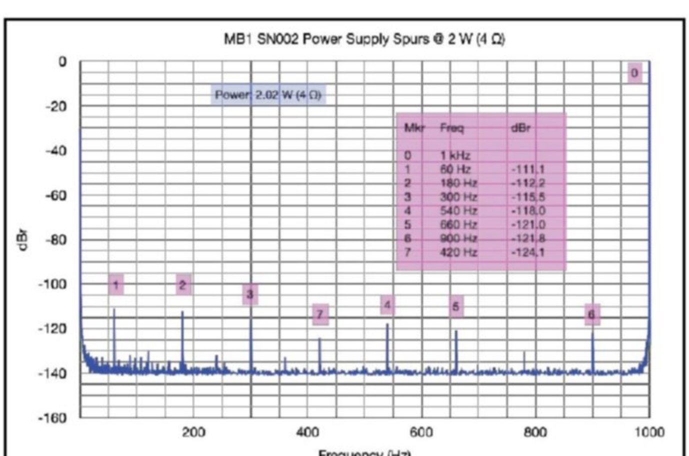


Figure 21: The graph here details the spectrum of a 1kHz sine wave, DC to 1kHz at 2W into 4Ω.

Power Supply Spuria and Noise

Figure 21 shows the spectrum of a 1kHz sine wave, DC to 1kHz, at 2W into 4Ω. The spurs are all multiples of the power line frequency (60Hz), and all are at least 110dB below the fundamental. Increasing the output power increases the absolute level of the spurs, but the performance relative to the fundamental actually improves.

MB1's power supply spurious performance is excellent. This is especially true considering that the amplifier's grounding scheme was not optimized for bench testing, but rather for use in a system that has multiple devices connected to power safety ground. Following the Audio Engineering Society (AES) recommendation to ground the input XLR connector pin 1 directly to the chassis greatly improved the in-system performance but degraded the bench measurements somewhat (~5dB at 60Hz).

With both inverting and non-inverting inputs terminated in 300Ω to ground, the unweighted output noise measured in the 20Hz to 20kHz band was -94dB relative to 1W into 8Ω. This an excellent number. Relative to the rated power into 8Ω (150W), the number becomes -116dB. The MB1 is a very quiet amplifier.

Total Harmonic Distortion (THD)

Most often, THD+N is presented. Here, the measurements are THD and do not include non- harmonically related noise. This allows one to assess noise (previous section) and harmonic distortion performance separately. **Figure 22** shows THD-1k at 150W into 8Ω. Performance is quite good (< 0.0013%) with the early harmonics decreasing monotonically. Current demands are low and there is a fair amount of margin before clipping. **Figure 23** shows THD-1k at 300W into 4Ω. As this is very near clipping, ±120Hz sidebands appear on the harmonics. They are at a low level and do not impact the THD numbers. The sidebands are gone if the drive is reduced by just 0.5dB (267W output). Note that I did not maintain the line voltage at 120Vrms. **Figure 24** shows MB1's THD at 1kHz into 4Ω vs. level. The horizontal scale is in dBV (dB referenced to 1V). This is the voltage across the load. Conversion to watts is:

$$\text{Power (W)} = (10^{(\text{dBV}/20)})^2/R = (10^{(\text{dBV}/10)})/R$$

The distortion curve is fairly flat until near clipping and remains below 0.003% up to 250W. These levels, though not spectacular, are very good. Thanks to the Bias Controller, there is no indication of crossover distortion.

Figure 25 shows THD vs. frequency and load resistance. A fixed output of 20Vrms was used for all loads. This ensured the amplifier was well above the Class-A region but safely below clipping. The measurement bandwidth was 85kHz. This allowed at least three harmonics to be measured. Distortion changes little with frequency, but doubles with a halving of load resistance. The former characteristic is due to the near constant level of feedback, the latter is likely due to a current- related non-linearity in the output stage. All the distortion measurements in this test were below 0.006%.

Intermodulation Distortion (IMD)

Figure 26 shows the CCIF (IUT-R) IMD two-tone (19kHz and 20kHz) performance of the MB1. The peak of the sum of the two tones is the same as the peak of a single tone producing 300W into 4Ω. All spurious products are more than 91dB below the individual tones and 97dB below their sum. Both even and odd products drop rapidly with order. The MB1 performs very well on this test, even at rated output.

As an additional check to ensure that the amplifier has no slewing problems, I repeated the CCIF IMD test but with 39kHz and 40kHz tones, doubling the test waveform's slew rate. The

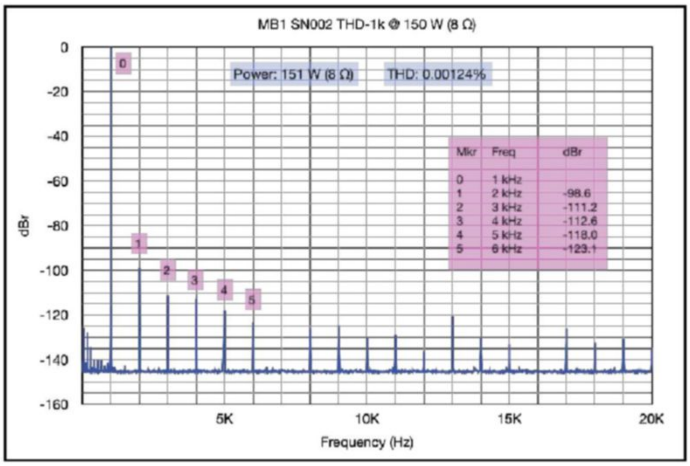


Figure 22: This measurement shows the THD-1kHz at 150W into 8Ω.

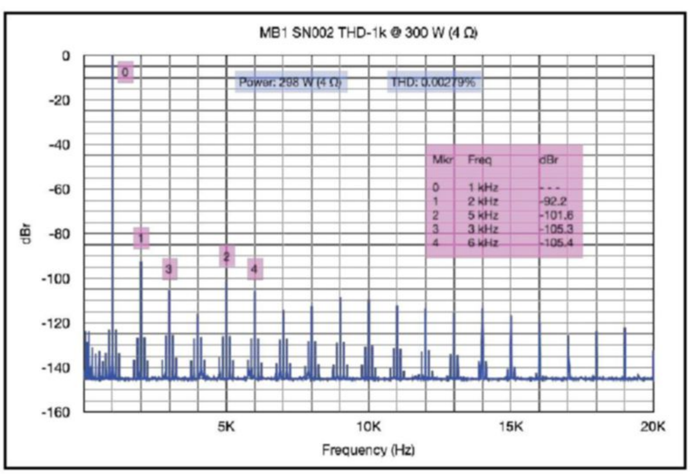


Figure 23: This measurement shows the THD-1kHz at 300W into 4Ω.

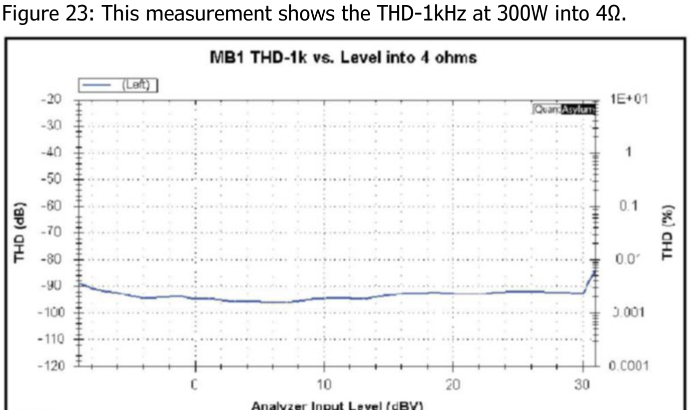


Figure 24: This measurement shows MB1's THD at 1kHz into 4Ω vs. level.

results, shown in **Figure 27**, are almost as good as the ones with 19kHz and 20kHz tones (Figure 26). The largest spur (1kHz) is 97dB below the peak power. This is excellent performance and gives no indication of slewing problems.

Summary/Final Thoughts

The MB1 was developed over a long period of time. Circuit design and analysis and the mechanical layout started as slow background activities in 2014. A serious level of effort began in 2019 and progress was actually aided by the pandemic shutdown in the following year. I deemed the effort complete in January, 2022 and installed two MB1s in my dedicated home audio system, replacing a much older stereo amplifier.

The MB1 is relatively complex and expensive but it was a very satisfying project. I am a retired Electrical/Systems Engineer, who enjoys studying the prior works of others, "borrowing" pieces I find appealing and combining them in new ways and adding new and unique design where possible.

Originally an analog/RF designer, I delight in being able to perform as many functions as possible without resorting to digital design or software. This may explain some of the design choices I made, especially in the control, monitoring, and protection areas. I also enjoy doing all of the work that I can.

Not everything was done in total isolation, however. I had the printed circuit boards manufactured commercially by ExpressPCB (www.expresspcb.com) and my good friend Steve Semenko (a master machinist) designed and fabricated a beautiful lead bending jig for the power transistors. One of my oldest friends, Bob Smith (also a retired engineer), served as a sounding board for design ideas over the years. Also, as can be seen from the References list in Part 1 (*audioXpress*, April 2023), Bob Cordell was both an inspiration and a guide.

If anyone has the interest to replicate all or part of this design, I have some suggestions. The mechanical aspects of the project were a large part of the total effort. Purchasing a chassis, most likely with an integrated heatsink, would greatly reduce the effort involved and could result in a more attractive product. The monitoring/protection logic could be implemented in an Arduino or Raspberry Pi device, providing more flexibility. The double-regulated front end high-voltage power supplies are, perhaps, overkill and the voltage doubler approach is wasteful thermally. Additional windings on the main transformer or adding a small transformer are options. The floating power supplies for the Bias Controller are wasteful thermally and could possibly be replaced by modern, low-noise, isolated switching regulators. I chose not to take the risk of switching noise.

The MB1 met all of my design goals and was an excellent platform for developing the Bias Controller. The two units that were constructed have been installed into my audio system since January, 2022, and have been performing admirably. They are essentially silent mechanically and electrically, have great power reserves, and run cool.

As an electrical engineer, I tend to believe that all well-designed audio power amplifiers sound much the same. If pressed, however, I would describe the MB1's sound as clean, very clean. Perhaps this is due to its well-controlled bias.

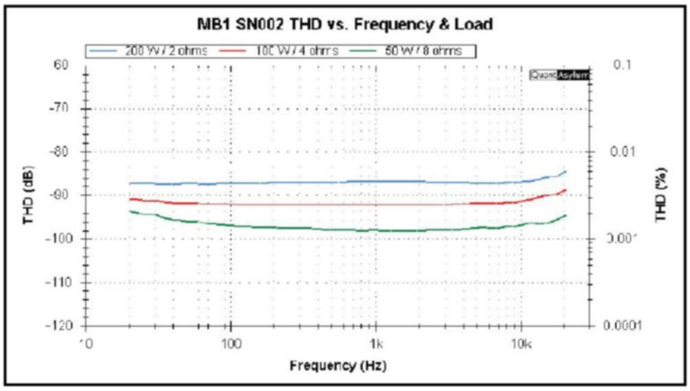


Figure 25: Here is the THD vs. frequency and load resistance.

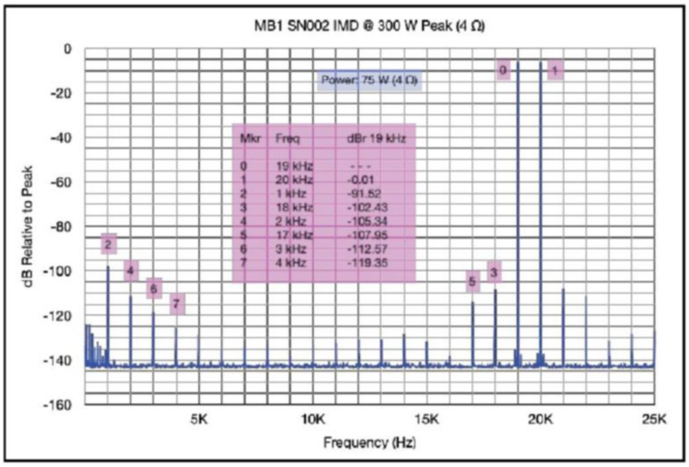


Figure 26: This graph details the MB1's CCIF IMD two-tone performance.

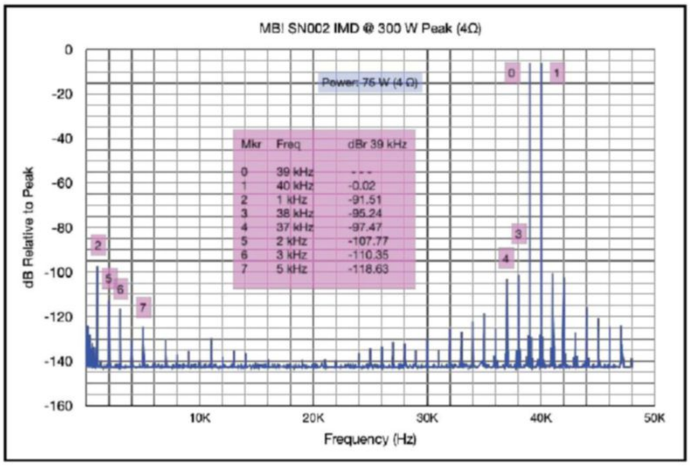
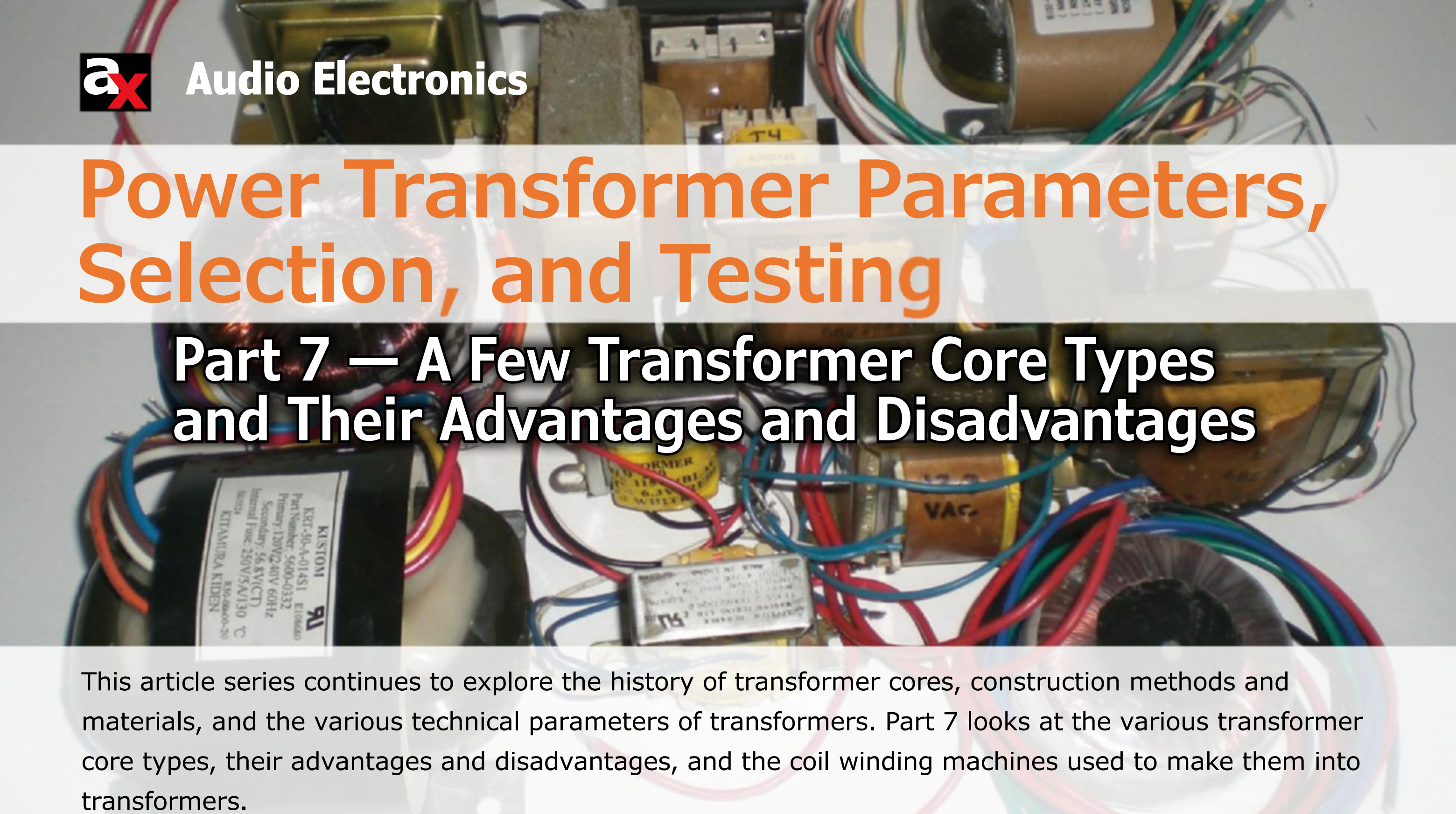


Figure 27: These results, using 39kHz and 40kHz tones, are almost as good as almost as good as the ones shown in Figure 26.

References

[16] A. Russell, "Ovation E-Amp: A 180 Watt Class AB VFA Featuring Ultra Low Distortion," HiFiSonix, https://hifisonix.com/wordpress/wp-content/uploads/2011/03/The_e- Amp_V2.03.pdf.



Power Transformer Parameters, Selection, and Testing

Part 7 — A Few Transformer Core Types and Their Advantages and Disadvantages

This article series continues to explore the history of transformer cores, construction methods and materials, and the various technical parameters of transformers. Part 7 looks at the various transformer core types, their advantages and disadvantages, and the coil winding machines used to make them into transformers.

By
Chuck Hansen

Let's begin this article with stacked laminated cores. Stacked laminations are cut from flat sheets of magnetic material. They can be simple rectangles like the "I" in E-I cores, or complex shapes such as the circular rotor and stator laminations used in motors and generators. They are either punched with steel dies, or cut by means of lasers or electrical discharge machining (EDM).

Shell-type E-I cores are the most widely used type for general purpose low voltage transformers (**Figure 18**). All the windings are wound on a bobbin or former and the E and I laminations are alternately interleaved into the center of the wound coil assembly. The center legs of the E laminations need to be twice the width of the I and the two outside E legs, to create a uniform width magnetic circuit, where the magnetic flux Φ generated by the coil in the center leg is divided in half through the outer legs (Φ/2). Since there are small air gaps in the interface between E and I laminations, the alternate stacking of the laminations also reduces the leakage flux and core losses.

Another arrangement, called the core-type, uses L-L or U-I laminations (**Figure 19**), where the primary and secondary windings are wound on different legs of the core. There is no center leg as with the E-I core. The assembled L-L core is a square, so every other lamination pair can be rotated by 90° to distribute the air gaps between solid core material. This is not always as efficient as the E-I core since the leakage flux is higher and there is reduced magnetic coupling between the windings.

This can be somewhat compensated for by winding half of each of the total number of primary and secondary winding turns on each leg. This also allows the primary and/or secondary windings to be connected in either parallel or series, for either 120V AC or 240V AC operation, for instance.

For high-voltage utility transformers, the L-L transformer is preferred because the primary and secondary windings can be installed on separate legs of the core for the maximum electrical isolation. The utilities aren't as concerned with the weight of a large power transformer, so their suppliers use more core area to maximize permeability. Core loss is proportional to core volume and approximately the square of the peak flux density, so there is higher efficiency with a larger core volume operating at a lower flux density.

For lower power low voltage transformers, the U laminations of the U-I core shown in Figure 19 has clipped corners, which allows some saving in magnetic core material as long as the corners have at least the same area as the straight section. If the U and I sections are to be interleaved, the corners cannot be clipped.

The E-I, L-L, and U-I lamination assemblies have small distributed air gaps that limit the maximum

Figure 18: E-I core laminations (Image Source: siliconsteelstrip.com)

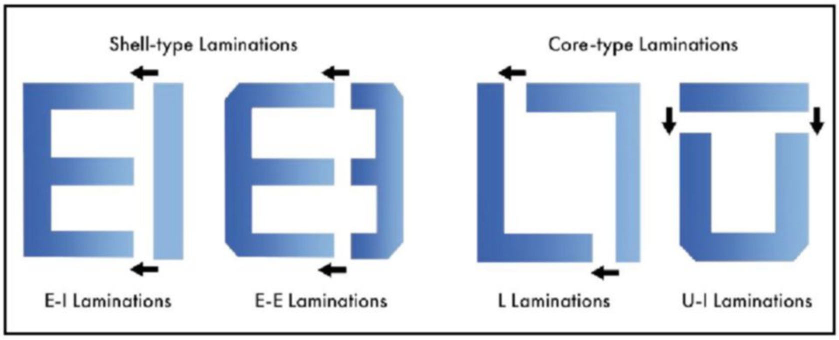


Figure 19: E-I and E-E shell-type, plus L-L and U-I core-type laminations.

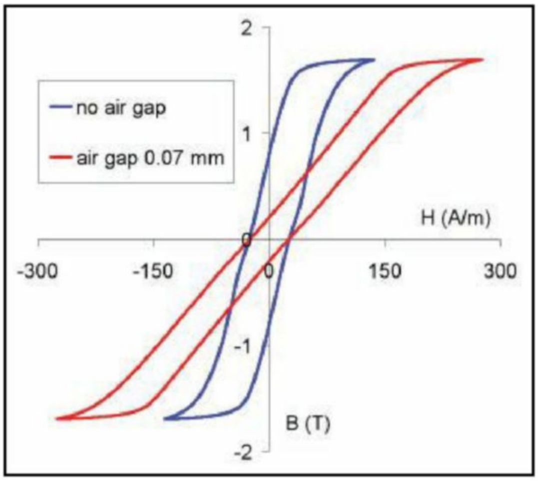


Figure 20: B-H curves with and without air gap (Image Source: Stan Zurek, Encyclopedia Magnetica)

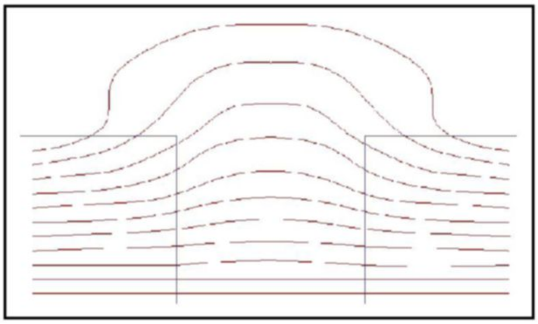


Figure 21: Fringing flux lines around air gap (Image Source: Static Electromagnetic Devices)

flux density and reduce the transformer efficiency. The total air gap is in series with the magnetic path and has a larger effective area. The majority of the magnetizing force, H, is required to overcome the reluctance of the air gap. This expands the area inside the B-H curve and takes more H to generator the same flux density, B, than if there were no air gap (**Figure 20**).

The larger the length of the air gap, the less likely the flux will cross the gap in a straight line and follow the magnetic core as if there were no gap (the path of least reluctance). Once the flux lines are in the gap air-space, they are no longer forced to follow any specific path, and are free to spread out. **Figure 21** shows an approximate view of the flux around the edges of the top half of an air gap. The lines become curved near the gap edges and spread away from the horizontal magnetic path. This is known as fringing.

The E-I and U-I laminations position the air gaps outside the area the windings occupy. If the laminations were made more complex such that the windings would entirely cover the gaps, this could improve efficiency by making use of the gap fringing flux. Other very interesting E-I core configurations were shown in an 2018 *audioXpress* article written by Gerhard Haas (see Resources).

Large utility transformers distribute the air gap positions over many different locations in the core laminations. This way the gap fringing flux is absorbed by the laminations above and/or below the gap. These large transformers are hand-built since the dimensions of the legs of the various laminations vary quite a bit. There are also curved segmented core laminations that can be woven through a finished coil assembly to resemble a toroidal core.

Advantages and Disadvantages of the various Transformer Cores

E-I Lamination Transformers

The advantages of E-I Cores include:

- A well-established manufacturing base exists for laminations, winding bobbins, sheet metal frames, insulating materials, etc.
- E-I steel stampings produce very little scrap, making very efficient use of the magnetic sheet steel, which reduces cost.
- E-I cores are flat stampings with no winding, bending or forming required.
- The E-I shell-type core needs only one bobbin, and it needs less iron.
- The transformers are tolerant to some amount of DC offset due to the air gaps.

The disadvantages of E-I Cores include:

- They are not as efficient as tape wound C and E cores, O cores, R cores and toroidal cores due to the distributed air gaps in the stacked E-I, L-L and U-I laminations. This results in a higher stray leakage flux.
- E-I core typically uses only 60%-80% of the grain orientation.
- A core made from stacked laminations sheared from fully annealed material may require a stress-relief anneal before stacking. This is especially true for thinner laminations.
- E-I laminations increases their cross-sectional area through the square corners, which results in a non-uniform core area. Holes are sometimes punched in the corners and the center top and bottom of the laminations. These are used for mounting screws, rivets or dimples that hold the transformer frame or mounting bracket against the laminations. The hardware through the core is a solid conductor that will increase the local core losses, as compared with laminations without holes, due to their eddy currents.
- E-I cores produce more audible noise than tape-wound cores.

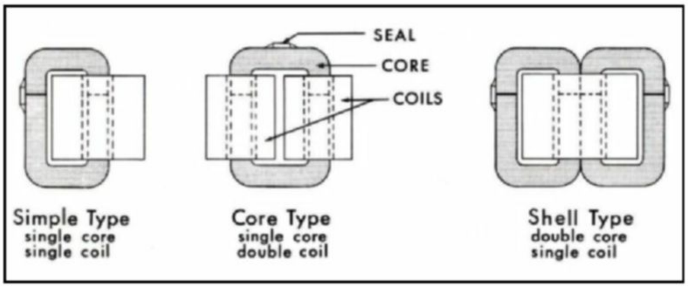


Figure 22: C core transformers: Simple type (l), core type (c), shell type (r) (Image Source: Arnold Silectron Core Bulletin SC-107A).

Tape-Wound Cores

Tape-wound cores are made in a continuous strip, wound under tension like a clock spring. They are then annealed to relax the molecular structure. Unlike the standard E-I core, with about 40% of the grain in the wrong (right-angle) direction, tape-wound cores have all the grains in the optimum magnetic direction. Except for the very small air gap in C and E cores, there is no air gap in the

Resources

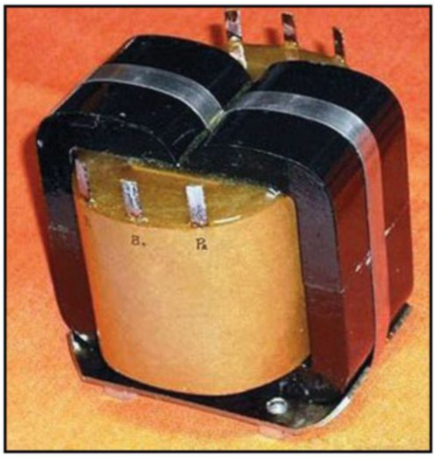
Arnold Silectron Core Bulletin SC-107A, 1966, pg. 51.

G. Haas, "Transformers for Tube Amplifiers," *audioXpress*, May 2018.

H. Hunt, Jr. and R. Stein, *Static Electromagnetic Devices*, Allyn & Bacon, Inc., 1963; pp. 43-45.

Stan Zurek, *Encyclopedia Magnetica*, Encyclopedia-Magnetica.com

Photo 17: Double C core transformer and a mounting bracket held together with steel bands (Image Source: James Transformer)



other types of tape-wound cores, resulting in a core stacking factor of 97.5% of its height for 14-mil core tape.

C and E Core Transformer Construction

C core transformers consist of two matched core halves and a coil, along with a metal band. The band is required to keep the two transformer core halves aligned and held tightly together to minimize the air gap. The gaps can be located under one or more coils, so the gap fringing flux can add to the total usable core flux.

E core transformers are used primarily for compact balanced three-phase transformers with one phase winding on each leg. If there is any load imbalance, an air gap will be required for each leg of the E core. Three individual C core transformers are usually used instead, which is more efficient since no added air gap is required, and the third C core has less magnetic length than the outside section of an E core. The core losses are larger for a three-phase wye-delta connected E core transformer because of the third harmonic flux.

A single C core transformer is configured as a Simple-type with the windings all on one leg, or as a Core-type with the primary and secondary divided between two legs. A single band holds the two core halves together and can also be used to retain a mounting bracket.

If two identical C cores are used the transformer can be configured as a more efficient Shell-type with the windings on the inside legs of the two cores (**Figure 22**). The bands can also be used to retain a mounting bracket (**Photo 17**).

Notice here that the C core halves shown in Figure 22 are not symmetrical. This was an alternate style available from core manufacturers to allow easier visual alignment of the two core halves inside the core box or winding bobbin, for minimum leakage flux.

Commercial C and E core transformers use bands made of low carbon full-hard cold-rolled steel, nickel-plated steel, spring-tempered phosphor bronze or hard brass. Small retaining seal clips are required to maintain the band pressure, and they are available in tin-, nickel- and silver-plated steel for good solderability.

Aerospace transformers use 316 stainless steel bands and special seal clips. After positioning the mounting bracket, mechanical or pneumatic band tensioning tools are used to pull the band into proper tension, and then crimp the seal clips to the bands to maintain the band tension.

The advantages of C and E cores include:

- Like E-I cores, a wide variety of winding forms and bobbins are available for C and E cores.
- C and E cores have a lower stray magnetic field flux than E-I cores.
- C cores weigh less than E-I cores of the same power rating due to their curved rather than rectangular profile, and have a high power-to-weight ratio.
- C and E cores fully utilizes the grain orientation of the steel for greater magnetic efficiency.

The disadvantages of C and E Cores are:

- Tape wound cores do not tolerate DC content in the AC waveform very well. You can add thin air gap material between the legs of the C or E cores if a small amount of DC content is unavoidable. This reduction in permeability increases the magnetizing current for a given flux density and reduces Bmax.
- There are a limited number of available core sizes compared to E-I laminations. C core cross-sections are not necessarily square, depending on the ratio of the core tape width to the wound outer diameter. The inner diameter (ID), outer diameter (OD), and height only come in 0.125″ increments.
- Alignment of the two core halves is critical to achieve the maximum efficiency. Any misalignment results in a lower effective core area and causes wasteful leakage flux.
- C and E cores are typically priced higher than traditional E-I cores due to less flexibility to adjust the core size of a C core. Each core requires a custom winding mandrel that accommodates a specific range of tape width and core profile.

Next Month

Next month we will look at additional tape wound transformer types, their advantages and disadvantages, and the specialized machines needed to make them into transformers.

About the Author

Chuck Hansen is an Electrical Engineer and holds five patents in his field of electric power engineering. He has worked in both the electric utility and aerospace electric power industries. He was assigned as the Instrument and Controls Startup Group Lead Engineer for two nuclear power generating units. He was the Electrical Systems and Controls Supervisor for a number of military programs as well as commercial and business jet electric power programs. He has written two books for Audio Amateur publications, and more than 260 magazine articles. He began building vacuum-tube audio equipment in college, and enjoys restoring and modifying audio equipment and test equipment. In his most recent roles as Senior Design Engineer and Lab Calibration Manager, he designed specialized test equipment particular to the aerospace electric power sector.

A Trip Down Memory Lane

Electric Guitar Amps

By
Richard Honeycutt

Photo 1: This 50W Valco amplifier with two Jensen 12" speakers was built for Sears, Roebuck, and Co.

The history of tube application began with radio receivers, the telephone system, and cinema sound systems. Later, tubes began to be used in home stereo systems, and in amplifiers for electric guitars. Today, the primary markets for tubes are high-end stereo and musical instrument amplifiers. In this article Richard Honeycutt remembers his first contact with "starter" guitar amps as a teenager, progressing throughout his days as a semi-pro musician.

I bought my first electric guitar in 1962, at the age of 14. It had no brand name, not even a Sears-Roebuck or Montgomery Ward house brand. It was made of poplar, including the neck and fingerboard, and painted an ugly shade of yellow-green. I refinished it with a cherry-red sunburst, shading to a walnut brown on the back and edges and re-strung it. That guitar served for me to learn the differences between playing an electric guitar and playing the Silvertone arch-top acoustic that I had used up to then.

I bought the guitar and an amp for less than half the going price of a Silvertone "starter" electric guitar. The amp's maker was as anonymous as the guitar's maker. The amp itself certainly dated back to the early 1950s, or perhaps even the 1940s, because the tubes were all loctals. If memory serves, it used a 7F7 high-mu twin triode, a 7A5 beam power tube, and a 5Y3GT rectifier. The circuit was similar to that of an early Fender Princeton (**Figure 1**), except for the tube complement. Maximum output power from a single-ended 7A5 was about 3.3W, which in this case fed an 8" Magnavox speaker. It had one channel, one volume control, and a tone control. No tremolo or reverb. And one other prominent characteristic: hum—lots of hum.

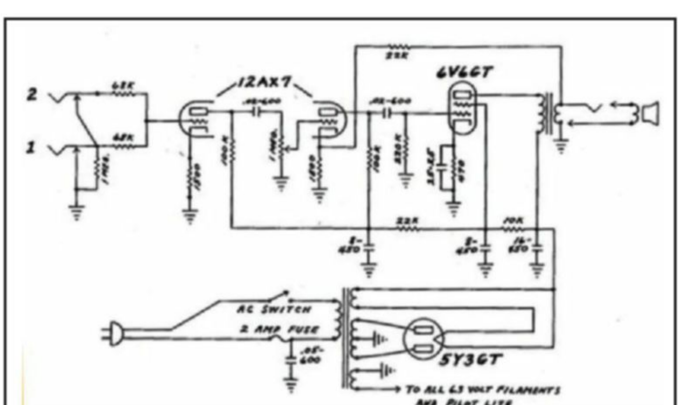


Figure 1: My anonymous first guitar amp's circuit was much like this old Fender Princeton circuit, except that it used a 7F7 twin triode and a 7A5 beam power tube instead of the 12AX7 and 6V6GT.

At that point, I had studied Supreme Publications' *Radio Servicing Course Book* and Van Valkenburg, Nooger, and Neville's five-volume Basic Electronics course commissioned by the US Navy, so I thought that by rewiring the low-level signal path, twisting the filament leads, and adding a hum-balance potentiometer to the filament supply, I could eliminate the hum. Not so: I had yet to learn about common grounding impedance, often mis-called "ground loops," caused in the case of this amp by randomly soldering ground points to the chassis rather than using a single ground point located near the input jack. So I just tried to ignore the annoying hum for two years, until I could afford a better amp.

The Silvertone Model 1484 Twin Twelve

In 1964, I bought that better amp: a Silvertone Model 1484 Twin Twelve (**Photo 1**). The cost equated to a little more than 100 hours' pay from my part-time radio announcer job. As you can see from the schematic shown in **Figure 2**, it had two channels, each with its own bass, treble, and volume control; a spring reverb unit similar to the Hammond-designed one; a tremolo; and a standby switch. The latter was useful when the guitarist had an intermission during a performance, but didn't want to have to spend the time required for the tube amp to warm up when the intermission was over. [The 1996 Movie *That Thing You Do* included shots showing the Twin Twelve being used by the young Rock and Roll Band, the Oneders (renamed "the Wonders")].

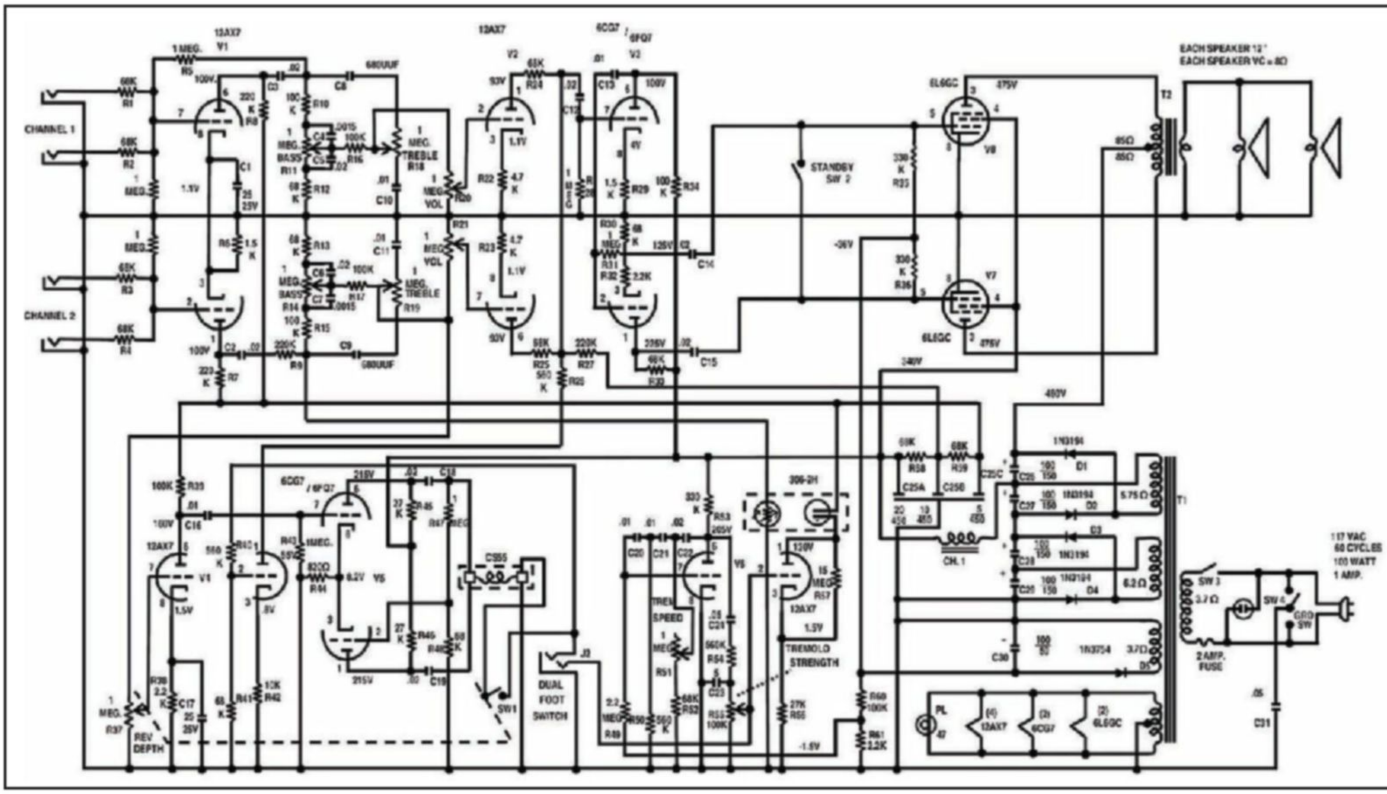


Figure 2: The Silvertone Model 1484 "Twin Twelve" was a popular guitar amp in the 1960s.

Advertised as a 50W amp, the Twin Twelve actually put out about 44W at the clip point, according to my old DuMont oscilloscope. (**Photo 2**). Valco held down the weight and parts cost by incorporating a voltage doubler B+ supply, allowing the use of a smaller and lighter power transformer than the more common full-wave supply would have required. The tone controls were a modified Baxandall design, giving a near-flat response (**Figure 3**) when carefully adjusted (knobs set with bass at ~2.5 and treble at ~ 3), plus a fairly wide range of audible tone adjustment. The tremolo circuit used one triode of a 12AX7 as a phase-shift oscillator driving the second triode, which varied the gain of channel two's first preamp stage in an almost sinusoidal fashion (**Figure 4**), creating a much more musical tremolo effect than most other popular amps of the time (**Figure 5**). The reverb was driven by the two triodes of a 6CG7/6FQ7 tube, operated in parallel. This dual-triode tube had a maximum plate dissipation of 4W, compared to 1.2W for the 12AX7. The reverb spring output was amplified by a 12AX7 and then mixed with the plate signal of channel two's second preamp stage. The hum of the Twin Twelve was much lower than that of my anonymous first amp, and what hum remained largely resulted from the wiring and single-coil pickup of my old guitar. During high school, I built my own solid-body electric guitar using properly shielded wiring, which reduced the hum, but the DeArmond pickups were still single-coil units,

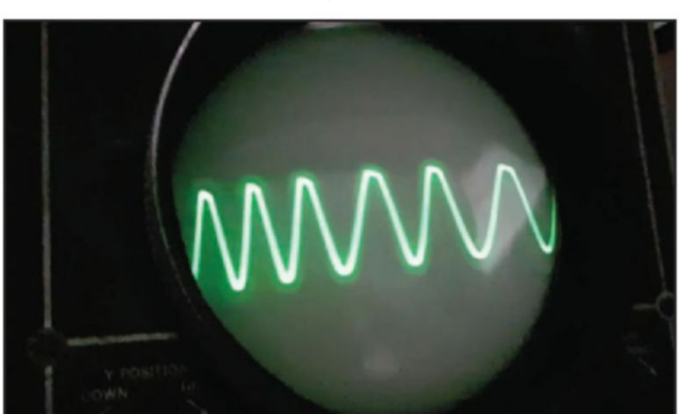


Photo 2: At the onset of clipping, The Twin Twelve put out about 44W.

About the Author

Dr. Richard Honeycutt bought a *Popular Electronics* magazine from the rack at a bus station in 1960. It contained an article on building a transistorized audio amplifier, which captured his interest. He followed up by subscribing to several electronics magazines. To avoid the cost of parts, he built up a "junk box" by disassembling trade-in TVs from his uncle's furniture store. His dad bought him the Van Valkenburg, Neville, and Nooger Basic Electronics book set, so he learned about tube circuits as well as transistors. He started repairing electronic devices about 1965, earning his First-Class Commercial FCC license in 1969. He has worked part-time in broadcast radio engineering and audio electronics repair since 1968, in addition to full-time work in acoustical and audio system design, plus a 20-year stint teaching electronics at the college level.

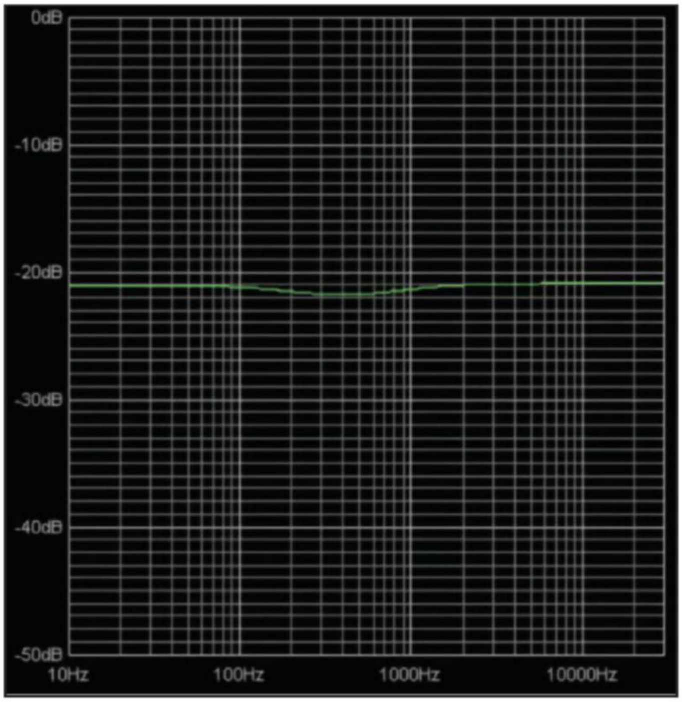


Figure 3: The Twin Twelve's tone controls could be set for an almost flat frequency response.

Photo 3: The Silvertone Six Ten was the final "step-up" amp model in the mid-60s Silvertone line.

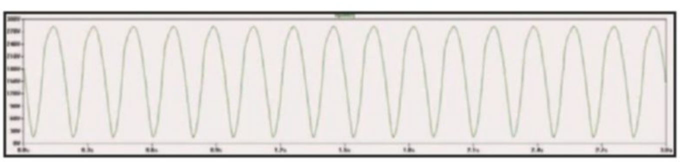


Figure 4: The Twin Twelve's tremolo wave was almost sinusoidal.

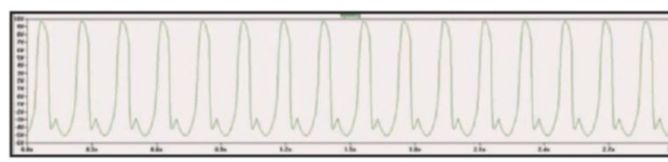


Figure 5: the tremolo of other popular amps of the time had significant distortion.

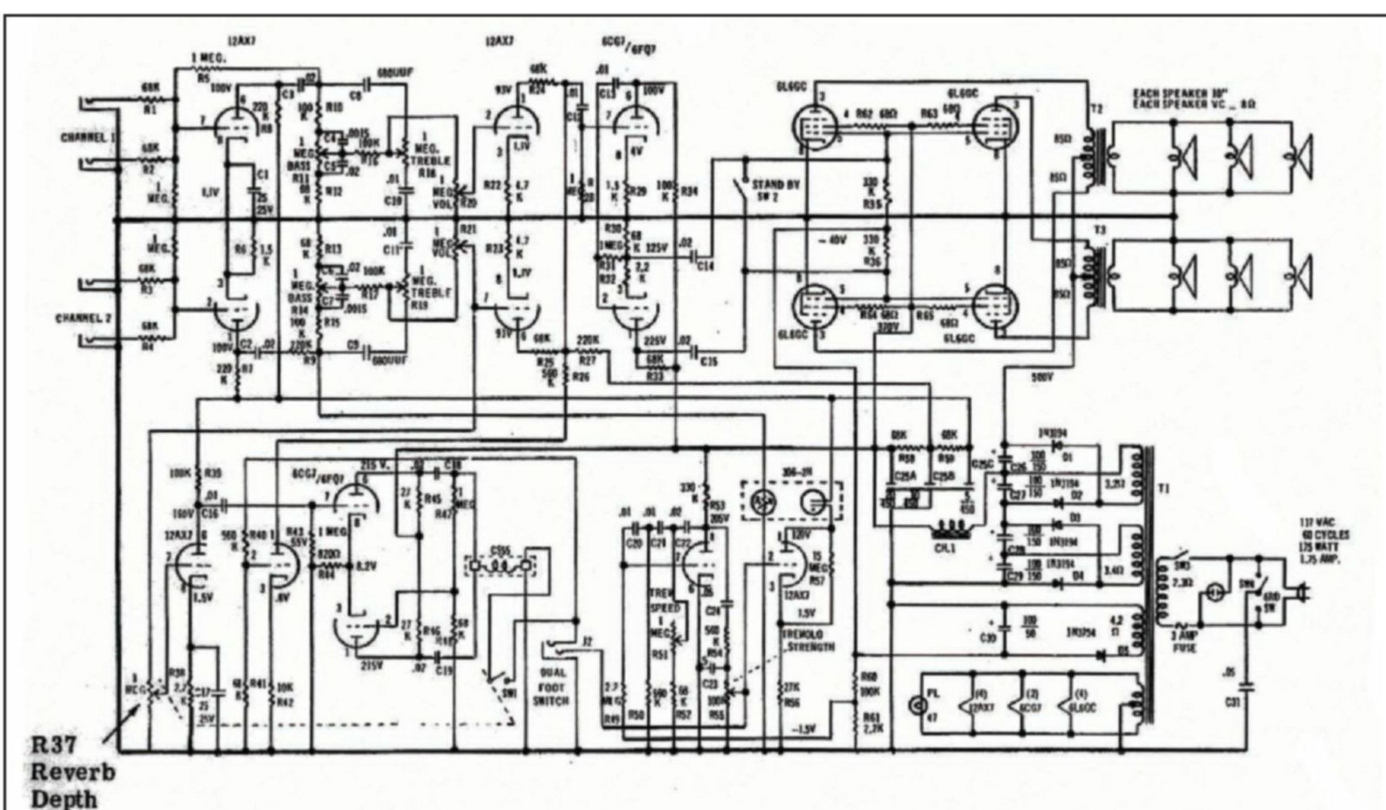


Figure 6: The Silvertone Six Ten circuit was similar to that of the Twin Twelve.

Photo 4: My 1970s guitar/bass amp was a Rickenbacker B115. (Image Source: www.rickresource.com)

so some hum remained. A year or two later, I was able to buy a Guild Starfire V guitar with humbucking pickups, and the long battle with hum was finally over.

Like many high-school rock bands of the early 1960s, my band started out using my Twin Twelve for two guitars, a Wurlitzer electric piano, and PA all at once, our bass player's father bought him a Silvertone Six Ten amp that we used to carry part of the load.

Photo 3 shows a Silvertone Six Ten amp, and as you can see from the schematic in **Figure 6**, the Six Ten was essentially a 100W (advertised power) version of the Twin Twelve, using six 10″ Jensen Special Design guitar speakers instead of the two 12″ ones of the Twin Twelve. Unlike the typical parallel push-pull output tube configuration of most 100W hollow-state guitar amps, the Six Ten split the speakers into two groups of three, feeding each group from a push-pull pair of 6L6GCs,each with its own output transformer, driving a nominal load of 2.7Ω from each output transformer. The tone control, tremolo, and reverb circuitry of the Six Ten and the Twin Twelve were identical.

The Rickenbacker B-115

In 1972, I joined a lounge band in which I alternated between electric bass and lead guitar. I ran my Rickenbacker 4001 bass in the stereo mode, using a GBX 215 for the fingerboard pickup and a guitar amp for the bridge pickup. At first, I used the trusty Silvertone Twin Twelve as my guitar amp, but in 1974, I switched to a newly released Rickenbacker B-115 (**Photo 4**) for the guitar and the bass's bridge pickup. This 100W amp used 7025 preamp tubes, a 12AT7 phase splitter, and two paralleled push-pull pairs of 6L6GCs for output tubes (**Figure 7**). It had no tremolo or reverb. The bass/middle/treble tone controls used a topology similar to Fender's three-control circuits, but with different capacitor values, making for a more pleasing tone, to my ears.

After I quit playing in lounges, I sold my guitar amps and when I played bass, I usually ran the Rickenbacker 4001 in mono, using only the fingerboard pickup for a mellow sound, feeding either the GBX 215 or the smaller GBX Bass Bug. (A confession with apologies to my strict hollow-state amp devotees: I have always preferred a well-designed solid-state bass amp to a hollow-state one, although the Ampeg SVT and the Sunn 2000S hollow-state amps sound quite good).

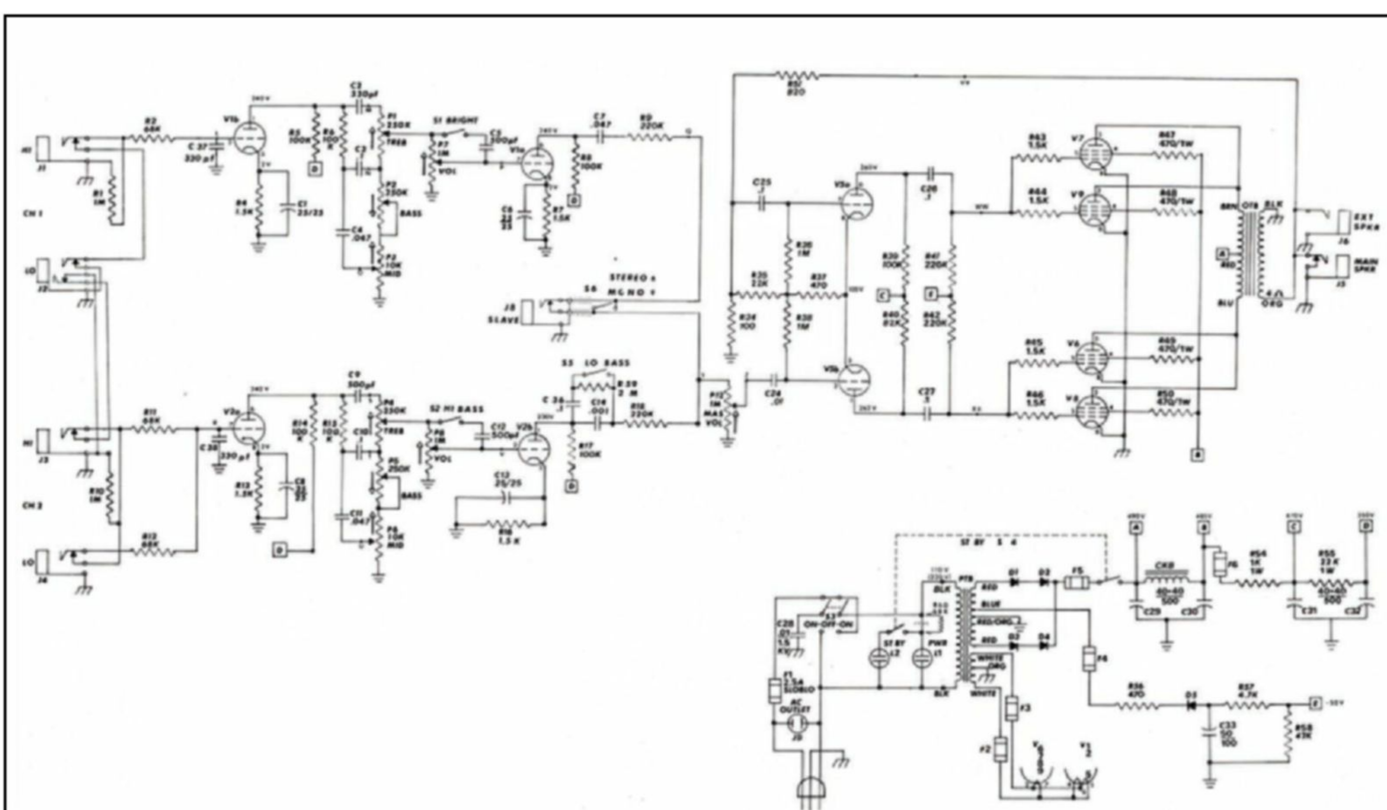


Figure 7: The Rickenbacker B115's circuitry was similar to that of several Fender amps of the day, but with differences in some component values.