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R&D Stories

The Return of the Original Western Electric 300B Vacuum Tube

By Richard Honeycutt

Audio Electronics

Using Chaffee's Method of Load Lines

By Scott Dorsey

You Can DIY!

Experimenting With Tube Sound

By Bill Reeve

The Cranberry

A Versatile Stereo Vacuum Tube Amplifier

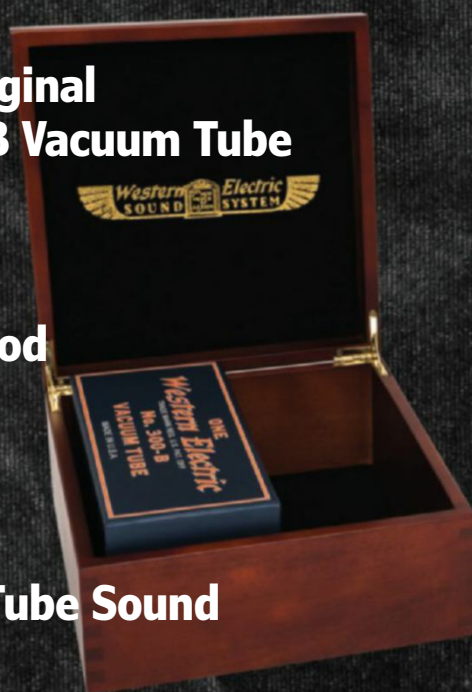
By Bruce Heran

Audio Electronics

Noise in MC Step-Up Transformers

By Burkhard Vogel

GLASS AUDIO
Special



Practical Test & Measurement

Measuring Vibration with Piezoelectric Sensors (Part 2)

By George Ntanavaras

Audio Electronics

High-End Mic-Preamplifier Design (Part 4)

By Mike Meehan

Hollow-State Electronics

Vibrato and Tremolo

By Richard Honeycutt

Audio Scope

The Real Output Current

By Michael Steffes

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A New Window for Vacuum Tubes

The original *Glass Audio* periodical was launched in 1989 and was published twice a year by Edward T. Dell Jr. "in the interest of high-quality audio reproduction using vacuum tube technology. Issue '0' was published in 1988 to promote and introduce the new concept, generating useful suggestions from readers—and the positive response encouraged Ed Dell to move ahead. This was a seemingly strange endeavor for the small Peterborough, NH-based publisher, known then for *The Audio Amateur* and *Speaker Builder* periodicals. The compact disc (CD) had replaced vinyl as the mainstream format for music distribution, but tube amplification was still inspiring a select group all over the world.

As Richard Painter wrote in the first issue of *Glass Audio*, "I believe that although the tube may be equaled or even bettered by transistors in high-current, low-impedance applications, they are unequalled in small signal voltage amplifying circuits (preamps, pre-preamps, and so forth)."

Most tube preamplifiers and amplifiers introduced then, in the late 1980s, had another distinctive feature—they were relatively affordable. Also, many new companies focusing on high-fidelity tube amplifier designs were appearing in the US, United Kingdom, and Japan. *Glass Audio* #1 featured a directory of Tube Electronics Manufacturers, listing 57 brands, including many relatively new names, at the time. The reason for this resurgence was directly linked to the availability of low-cost Russian, East European, and Chinese tubes, largely focusing only on audio applications—one of the few remaining markets. In contrast, most tube manufacturers from the United States and Western Europe had already ceased their activity or were about to, while the prices for the most recognized new-old stock (NOS) tubes continued to increase dramatically.

In his editorial for the first issue of *Glass Audio*, Ed Dell highlights this "tube renaissance" in both high-end consumer audio and also for professional audio equipment.

In 1989, the year Sony sold 50 million Walkmans and celebrated the 10th anniversary of the portable cassette player, it was also possible to find precious tubes and tube equipment being sold cheap at garage sales everywhere—magnets for DIY audio and tube enthusiasts. The enthusiasm around tube audio grows in direct contrast with the transition to other "cheaper" or "more convenient" forms of music distribution, as manufacturers gradually find ideal market application niches for tubes.

Fifteen years later, in May 2005, when *audioXpress* celebrated the anniversary with a special issue of *Glass Audio*, the magazine was still filled with many ads from tube suppliers and specialized amplifier manufacturers from the US, Europe, and Japan. An interesting article from Eric Barbour—a regular *Glass Audio* contributor since 1991, and also co-founder and former associate editor of *Vacuum Tube Valley*, the only other magazine that focused on tubes—discussed a visit to the MU, Inc. tube factory in Oceanside, CA. Or what was left of it, as Barbour documented with his photos of this operation mainly focused on radar, radio transmitters, and industrial applications. "I already knew that MU was probably the last factory in North America capable of making glass vacuum tubes without depending on numerous outside parts suppliers," he stated.

Today, as tube audio equipment continues to enjoy cult status, keeping a handful of tube companies busy, there are not many other applications left. The unique properties of vacuum tubes are mainly valuable for high-end audio, studio recording and processing, or music instrument amplification and effects, and that is where opportunities remain for a few manufacturers.

As we reach another May edition and we celebrate *audioXpress'* roots in *Glass Audio*, we are obviously very glad to once again display a tube entirely manufactured in the United States on our cover. As *audioXpress* reported back in 2018, the 150-year-old Western Electric reintroduced its flagship product—the famous 300B electron tube. Against major obstacles, the Rossville, GA-based company is back in production, as Charles Whitener, the company's CEO and main promoter explains in this edition. And *audioXpress* has already received a pair of new WE 300Bs to review.

J. Martins
Editor-in-Chief

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OUR NETWORK



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Acoustic Sciences Corp.	39	Menlo Scientific, Ltd.	59
All Electronics Corp.	65	New Sensor Corp.	34-35
Amplitrex Audio Products, LLC	51	Parts Express International, Inc.	68
Apex Tube Matching	31	Purifi Audio	19
BTB Elektronik-Vertriebs GmbH	33	QuantAsylum USA, LLC	17
Celestion	15	Redco Audio	47
COMSOL, Inc.	9	SB Audience	25
Hammond Manufacturing, Ltd.	3	Solen Electronique, Inc.	63
Hypex Electronics B.V.	53	Triad Magnetics	67
JJ Electronic	11	Warkwyn	39
KAB Electro-Acoustics	19	Western Electric	2
Lundahl Transformers AB	13	ZUMI Systems	45

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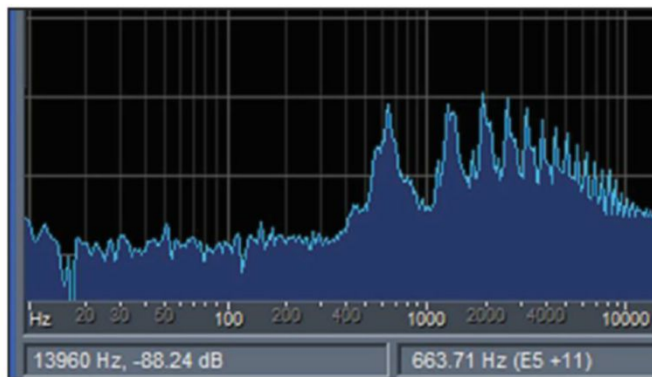
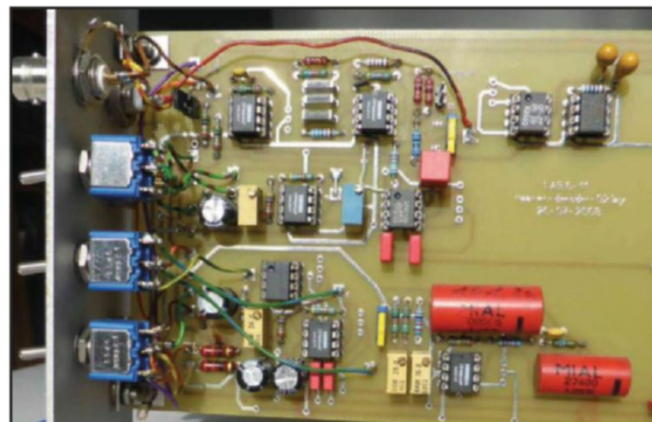


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● Glass Audio

8 The Original Western Electric 300B Vacuum Tube

Now Shipping Worldwide

By Richard Honeycutt

audioXpress provides an inside look at Western Electric, which has reestablished operations due to the resolve of entrepreneur Charles Whitener. The company now based in Rossville, GA, recently created a new and re-imagined electron tube works facility with the goal to resume manufacturing the 300B single-ended triode tubes, and reintroduce its flagship product to market.

14 Using Chaffee's Method of Load Lines

By Scott Dorsey

In this article, Scott Dorsey recalls a rapid and ingenious way to use load lines to easily predict tube circuit operations.

20 Experimenting with Tube Sound

By Bill Reeve

Learn how automatic gain control applied to solid-state and vacuum tube audio paths enable interesting design, construction, and listening experiences in this easy-to-build stereo preamp.

26 The Cranberry

A Versatile Stereo Vacuum Tube Amplifier

By Bruce Heran

In an effort to expand his DIY amplifier collection, Bruce Heran designed this stereo vacuum tube amplifier using a concept that originated in the 1940s and was initially intended for a budget amplifier.

● Columns

AUDIO SCOPE

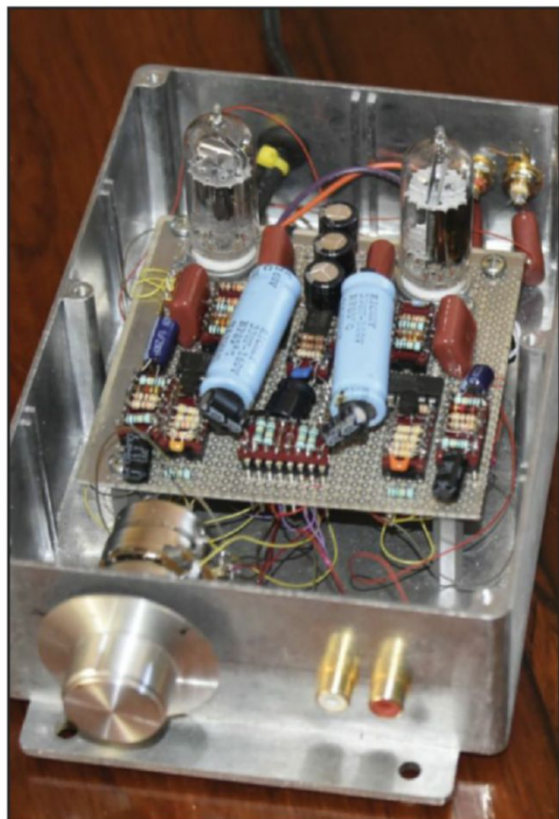
36 Determining the Real Output Current for Op-Amps and Fully Differential Amplifiers

By Michael Steffes

HOLLOW-STATE ELECTRONICS

60 Vibrato and Tremolo

By Richard Honeycutt



● Features

40 Measuring Vibration with Piezoelectric Sensors (Part 2)

Design and Construction

By George Ntanavaras

In this article series, George Ntanavaras describes his research on measuring vibration using Piezoelectric Sensors. In the second part, he discusses the design of such a system, configurations of the charge amplifiers circuits that he built and tested, and the connections to the piezoelectric sensors. He also presents some additional sensors that could be used for vibration measurements.

48 Noise Is Everywhere—Even in MC Input Transformers (Part 1)

A Special Survey for the Skeptics

By Burkhard Vogel

In this article, Burkhard Vogel discusses the signal-to-noise (SN) measurements of moving coil (MC) phono amps. He thinks it is good practice to measure a transformer with an input load and he wants to prove that an MC step-up transformer (SUT) produces noise because of its insignificant impedance.

54 The Science and Physics of High-End Mic-Preamp Design (Part 4)

Different Approaches

By Mike Meechan

In previous articles, Mike Meechan considered the fundamental physics that play a rudimentary part in microphone preamplifier design, and explored its inhabitancy in the real world to retain the fidelity of the mic signal along its path from transducer to output. At this stage, he continues to look at what (theoretically) contributes to perfect active mic preamps.

● Departments

4 From the Editor's Desk

5 Client Index

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The Original Western Electric 300B Vacuum Tube

Now Shipping Worldwide

Western Electric has reestablished operations due to the resolve of entrepreneur Charles Whitener. The company now based in Rossville, GA, recently created a new and re-imagined electron tube works facility with the goal to resume manufacturing the 300B single-ended triode tubes, and reintroduce its flagship product to market.

By
Richard Honeycutt

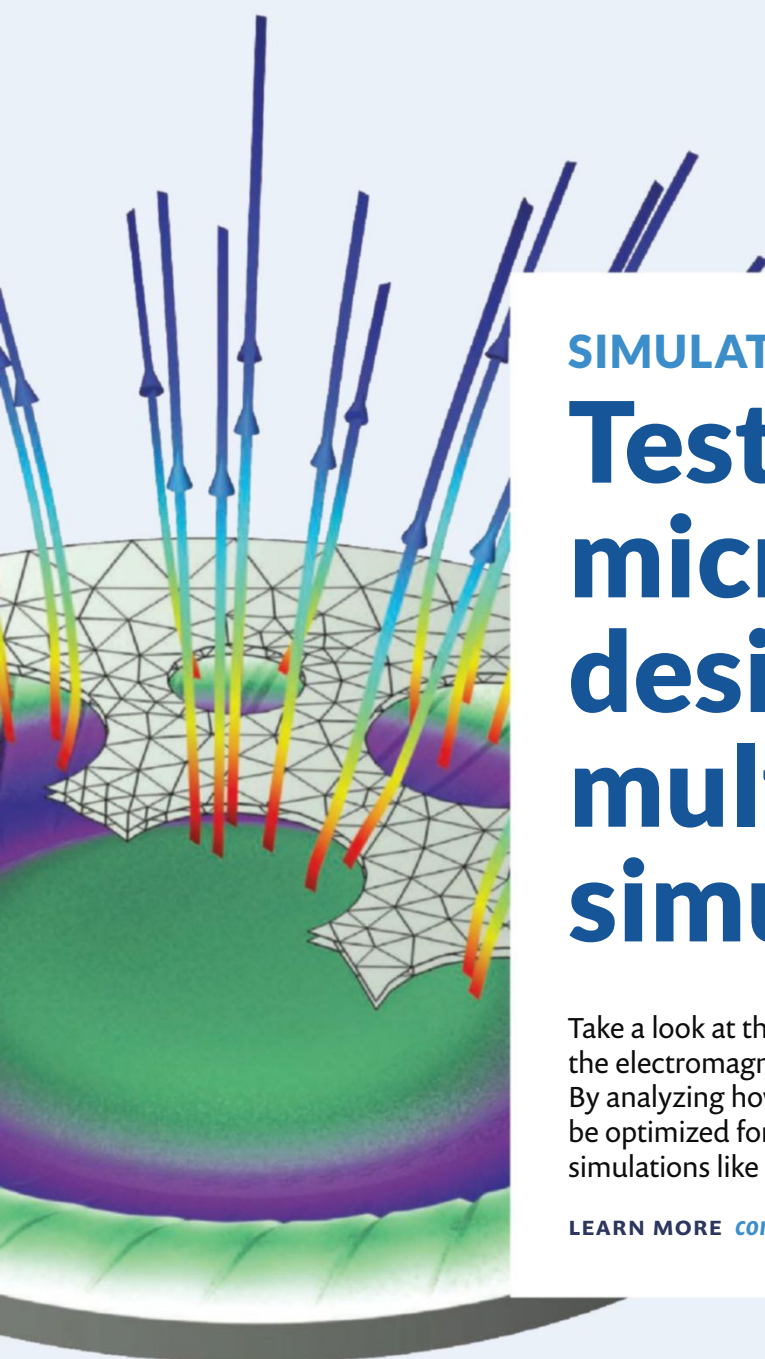
The original 300B vacuum tube, a longtime favorite in high-fidelity, is now available for direct purchase from Western Electric and authorized dealers worldwide. After two years of factory construction, assembly line setup, and staff training, early batches of electron tubes shipped from October 2020 to January 2021. Western Electric has since shifted to regular production in order to meet record demand from overseas distributors and to stock up for a new online store. The price for one matched pair is \$1,499. Single tubes and matched quad sets are also available.

"We're honored to continue serving music listeners and amplifier designers who've trusted the quality of our electron tubes for years," says CEO Charles Whitener. First introduced by Western Electric in 1938, the 300B is now praised among audiophiles for its ultra-long life and sonic purity. "Making sure our customers get the 'real thing' is a great challenge. We rely on decades of experience and a strict adherence to quality standards set in stone long before our time," Whitener adds.

With the return of the famous triode comes the re-establishment of Western Electric manufacturing—now at the company's new facility located in Rossville, GA. Officially established in 2018, the Rossville Works facility contains an entire wing dedicated to continuous 300B production. On the production floor, state-of-the-art, modern machinery harmonizes with original machinery—some first commissioned in the 1940s. Easily the most advanced tube operation in company history, the Rossville Works represents a commitment to the future of the vacuum tube industry and high fidelity audio.

Anticipating the official announcement of the return of the 300B and Western Electric's commitment to quality manufacturing, *audioXpress* had long planned a visit to the new facility, which unfortunately was not yet possible in time for this edition. However, Whitener was kind enough to grant a brief interview and he supplied some exclusive photography, which we used to put together this article.

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For the new Rossville facility, Charles Whitener confirmed they installed new hydrogen reduction ovens, automated cathode cleaning lines, new laser welding systems, a state-of-the-art water deionization plant, and an updated testing system with improved accuracy. However, the 300B is still manufactured from its original tooling and assembly specification standards. “Although we’ve invested in a modernized production line, many of its aspects will of course remain true to tradition. For example, the core material, the so-called “secret sauce” embedded in each cathode is derived from an original Hawthorne Works 1963 melt. Western Electric’s attention to each delicate detail and historic promise of quality place the WE 300B in a category of its own.”

The Interview

Richard Honeycutt: Can you give us a summary of the history of Western Electric’s vacuum tube manufacturing from the New York, Hawthorne, Allentown, and Kansas City Works to the present?

Charles Whitener: Founded in 1869, Western Electric is an American manufacturer with an enduring relationship with quality. Today, we make stunning listening experiences through innovative audio

equipment and legacy hi-fi like the 300B vacuum tube. The history of our company has strong ties to the history of the vacuum tube itself. In 1912, Western Electric engineers witnessed a demonstration of Lee de Forest’s “Audion,” a primitive yet ingenious device that showed them—and subsequently the world—the amplifying power of the high vacuum tube. AT&T purchased de Forest’s patent and Western Electric’s Engineering Department went to work, setting up a small tube shop on Hudson Street in New York. The operation was small, but grew to employ more than 1000 people by 1929. As demand exploded, tube manufacturing was expanded to the Hawthorne Works and later the Allentown Works. Western Electric supplied millions of tubes of various types to fuel the electronics industry in the US, and as Westrex internationally.

As the era of the vacuum tube faded, all remaining operations were moved to the Kansas City Works. This move was completed in 1970. Eighteen years later, Western Electric employees manufactured what they thought would be their last 300Bs. I approached AT&T in 1992, finally re-establishing manufacturing at the same Kansas City location in 1996. The Western Electric name, all of the tube specs, tooling, and machinery—not to mention decades of employee know-how—was going to disappear. In a way, we rescued it all from being put on a shelf forever. We’ve been manufacturing legacy products like the 300B and new high-fidelity ever since then. Our latest manufacturing facility, the Rossville Works is our finest yet. Our staff operates state-of-the-art machinery alongside much of the old machinery to produce the same incredible product people have known for years.

Richard Honeycutt: How is the 300B different from other power triodes, other than its iconic status?

Charles Whitener: The real magic of the 300B is in the filament. An equipotential cathode allows transconductance to be sustained symmetrically across the anode, when excited with direct current. The nickel filament core is from a 1963 melt from the Hawthorne Works. It’s coated with a proprietary, double carbonate mixture, which contributes significantly to the tube’s legendary long life and overall performance. In fact, the tube’s metallurgy is based on designs originally intended for underwater sea cable transmission, where minimal maintenance is obviously essential.

I would also add that because the 300B has a directly heated cathode, as opposed to indirectly heated, it offers superior impulse response. Thermal latency is reduced. So, when we experience the pleasurable sound of a 300B, we’re experiencing the low thermal inertia of a directly heated cathode.



The new Western Electric 300B power triode is manufactured in the Rossville Works Plant in Rossville, GA.



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Vacuum Tubes



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Each pair of Western Electric 300Bs is shipped in a classic wooden box.

Richard Honeycutt: In general, how do your 300Bs compare with other ones now on the market?

Charles Whitener: Those who love the 300B know there's nothing like the real thing in terms of performance, longevity, sound, and overall quality. As we're now set up for continuous 300B manufacturing, we hope people are less and less inclined to resort to replica tubes. Some out there are better than others. But the 300B is, as the saying goes, often imitated, never duplicated. When you see the name on the base, you know you can trust the quality. We guarantee it with a five-year warranty.



A new 300B is shown in front of the aging rack.

Richard Honeycutt: If a customer buys a pair or quad of your 300Bs, how much advantage will they notice if hand-matching is chosen?

Charles Whitener: All pairs and quad sets that we sell are computer-matched based on a range of performance characteristics. We perform this test as part of our final inspection. Offering tubes that are as close to identical as possible gives the end user confidence. They have put lots of careful thought into getting the best sound possible with every other component in their setup, so we'd hate for our product to throw off the balance! Particularly with push-pull configurations.

Richard Honeycutt: How big is the market for premium vacuum tubes presently?

Charles Whitener: Since we re-established manufacturing in 1996, the market has remained niche. But we also have reason to believe it's growing. When we made the official announcement that we were bringing back the 300B in 2017, we did this with the idea to reach new 300B lovers. We then launched an online pre-order campaign and were met with more interested people in one week than we had been in the past 10 years combined. We're now in regular production and look forward to seeing how far our reach as a manufacturer can go. Niche? Yes. But the demand for good sound will never go away.

Richard Honeycutt: What does your crystal ball show for the future of hollow-state audio equipment?

Charles Whitener: As long as there are new amplifier designs, particularly single-ended, people will need 300Bs. It seems every week we learn about something new and exotic from customers and friends in the industry. As a tube manufacturer, we're always encouraged by these new designs. It produces steady demand for 300Bs and shows the world that it's not always about nostalgia and legacy products.

I'd also like to add that these new products, our 91E Integrated Amplifier included, produce an outflow of conversation and memorable listening room moments. This is the world of hi-fi. It's up to us to influence and educate younger generations about good sound and encourage them to enter the conversation. Anyone who has listened to their favorite music with a good tube amplifier knows it's an experience worth sharing. 

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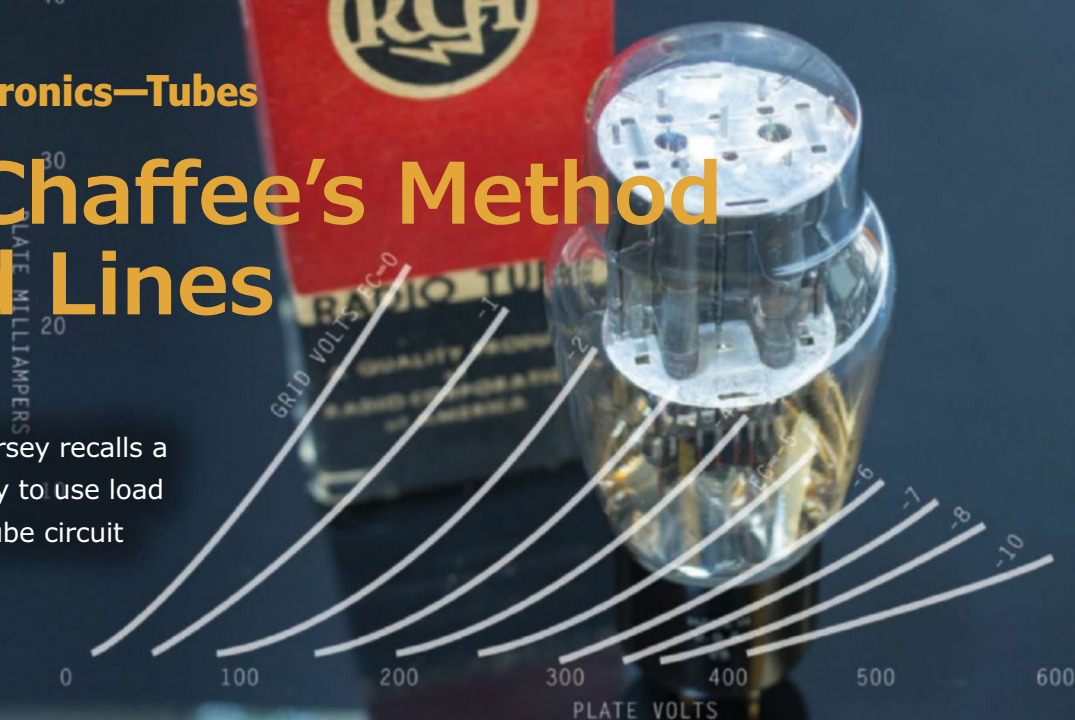
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Using Chaffee's Method of Load Lines

In this article, Scott Dorsey recalls a rapid and ingenious way to use load lines to easily predict tube circuit operations.

By
Scott Dorsey



If you are like me, you think of a triode tube as a simple transfer function. One independent variable goes in: the cathode-to-grid voltage. One dependent variable comes out: the plate current. So, since we think about devices that way, you'd expect the datasheet for a tube to have a simple graph laid out with one plotted against the other. Maybe you might want to have a family of different curves at different plate voltages. That's what you'd want to see. (Sometimes you do see this as an addition to the characteristic curve plot, and it's called the transfer curve or the grid curve and it is sometimes more clear for things like estimating linearity.)

But you don't. Instead, you see a rather confusing looking family of traces, with the plate voltage plotted against the plate current with a whole array of different grid voltages. This seems totally counterintuitive. How do you tell at what voltage the tube turns on? How can you see how wide the linear region is? All of those look difficult with the

characteristic curves, sometimes called the "plate curves" or "plate family of curves."

But, it turns out that the reason why this is the way the curves are laid out is because this allows a tremendous number of useful calculations to be made entirely with graphical methods, using a pencil and a straightedge. Today with modern computers, the ability to make these calculations numerically is much easier, but the graphical method will also help you get a sense of how the circuit is operating.

Chaffee's Method of Load Lines is rapid and ingenious and allows you to try out a tremendous number of different biasing and drive arrangements. It may have been invented by E. Leon Chaffee at Harvard University in the 1920s, or maybe not. My freshman circuits class instructor called it Chaffee's Method so that is how I will forever think of it.

Let's get a characteristic curve out of the RCA *HB-3* handbook. Although the tube handbooks sold to service shops have curves on some devices, the *HB-3* has a complete directory of curves for everything RCA sells, and more than that.

For years, the loose-leaf set was fairly rare and sold for a premium, but now you can go to Pete Millett's website and print out the curves for any tube you want, full size on a big sheet of paper to make it easy to see what is going on. See, we will be using computer technology here. **Figure 1** shows the curve for the 12AT7. I picked it because the 12AT7 is a common tube with typical but interesting curves. More about that in a bit.

Let's say we want to build a standard voltage amplifier gain stage, as shown in **Figure 2**. This is our baseline circuit for this entire article. It's actually half a differential amplifier and everything in this article can be generalized to a differential amplifier,

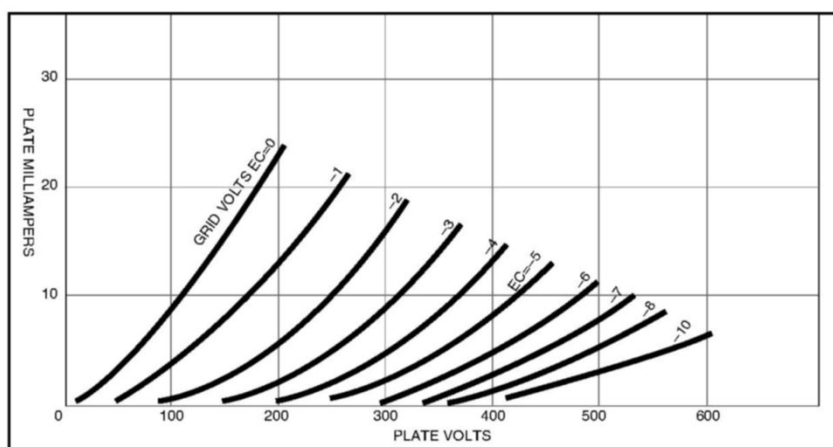


Figure 1: This graph shows the curve for the 12AT7.

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but because life is short and art is long that will be a subject for another day.

For now, we have this simple gain stage and we want to know what the values of the resistors should be. We're going to use a B+ supply voltage of 250V because it's a nice number that makes the math easy. We are also going to have an input signal that is $2V_{pp}$, that is the positive and negative peaks are 1V from ground. That also is picked to make the math easy, but I assure you this method works just as well for higher and lower signal levels.

Figure 2: This baseline circuit, which is typically used to build a standard voltage amplifier gain stage, is used as an example.

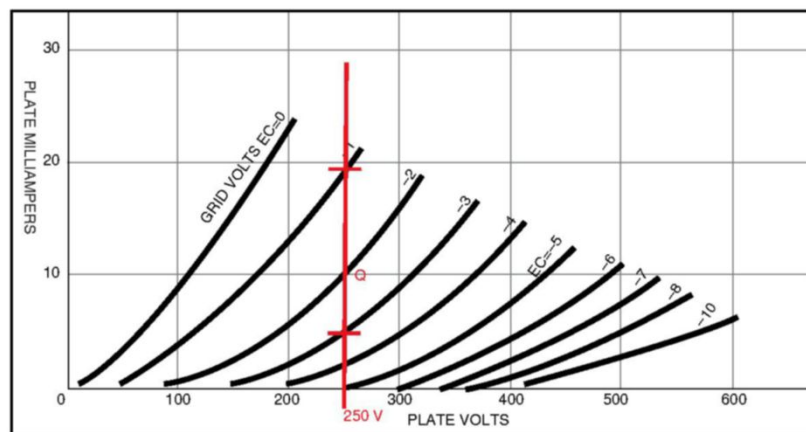
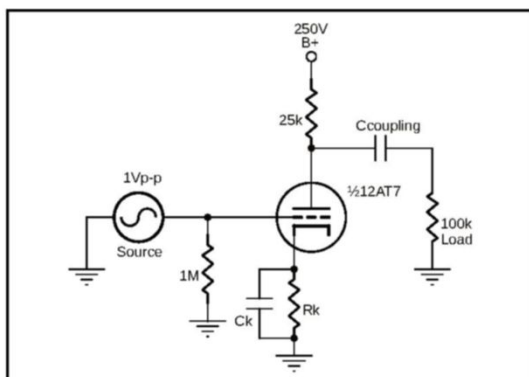


Figure 3: In this example, a vertical line is drawn at the B+ voltage we are using, and we can examine the curves that intersect it.

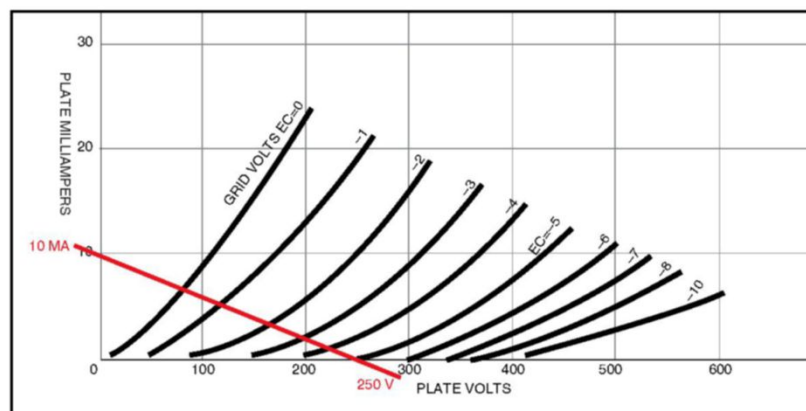


Figure 4: Here, a diagonal line is drawn between the 10mA on the vertical axis and the 250V on the horizontal axis.

Step One: Finding the Operating Point

So, we have a Class A amplifier with a $2V_{pp}$ signal, and we want it to be as linear as possible for a range a volt above and a volt below the quiescent operating point. So how do we find that point?

As demonstrated in **Figure 3**, we draw a vertical line at the B+ voltage we are using, and examine the curves that intersect it. This tells us the amount of current that the plate will carry for each different B+ voltage, and the distance between the curves on that line should be uniform in the region where we are operating.

Now, the best that we can do is to set the value to about 2V, and as we look at the curve intersections we can see that the linearity isn't really very good. If we set the quiescent voltage at -2V, the quiescent plate current will be at 10mA. At a negative peak, we'll be at -3V and the plate current will be at 19mA. At a positive peak, we'll be at 1V and the plate current will be at 5mA. The two horizontal bars indicate the operating region. It isn't quite linear; the gain is a little more on the positive direction than the negative direction. This is because the 12AT7 isn't really a very linear tube.

If we had picked a 12AX7 or a 6SN7, we would have had much better symmetry around the operating point. On the other hand, if we had reduced the signal level, it would have helped as well. But we've already chosen the 12AT7 and the 2V signal level (or maybe it was chosen for us), so we're just going to accept that nonlinearity and see where it takes us.

I then checked the tube handbook to see what it recommends for a bias point for an A1 amplifier for a 12AT7. It recommends 2V, so it seems we're doing as well as we can do.

If this was a push-pull amplifier that we wanted to bias into Class B, we could get away with reducing the bias by quite a bit because we would only care about linearity in one direction from the quiescent point, not both. If that were the case, we could likely get away with a -3V or even -4V bias and reduce quiescent current way down. But we're doing Class A today.

Step Two: Finding the Cathode Resistor Value

What is the maximum current that could flow if the tube was a total short? Well, 250V through a 25kΩ resistor would be 10mA. That's our maximum point. So in the second step (**Figure 4**), we draw a diagonal line between the 10mA on the vertical axis and the 250V on the horizontal axis. This is our DC load line and that line intersects the 2V curve at 4mA.

That is to say when the tube is idle, it should have a plate current of 4mA. We also know that the value of the cathode resistor is such that it needs to drop 2V

with a plate current of 4mA, which means it needs to have a value of $2/0.004 = 500\Omega$.

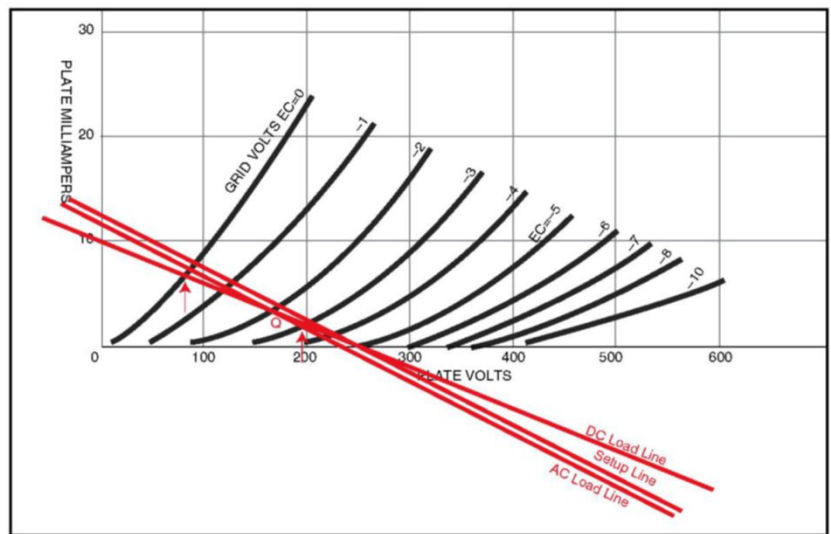
The current through the tube is not linear with voltage, so we can't just use simple arithmetic to figure this out using the numbers off the datasheet. But, we can use the graph combined with simple arithmetic to figure it out without having to calculate all the interactions.

If you look up the datasheet recommendation for the cathode resistor for a Class A amplifier, the Sylvania people recommend 200Ω , not 500Ω . They are giving a good medium value that will work for a variety of plate resistor values. With the graphical method, you can instead figure out the best value for your specific circuit values.

Step Three: Drawing the Dynamic Load Line

So far we have looked at the circuit's DC behavior and determined what the DC values should be when the device is idle and at positive and negative peaks. Now we need to look at the same circuit's AC behavior in a slightly different way (**Figure 5**).

C1 is a DC blocking capacitor and isolates the DC values of the circuit from the load. But it passes AC,

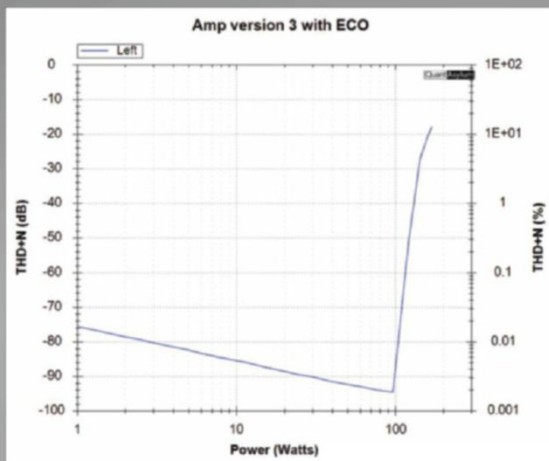


and so when we think about the circuit's AC behavior, we can think of the load as being in parallel with the plate resistor. How can we do this when the load is a resistor to ground and the plate resistor is a resistor to B+? Because from the standpoint of AC, the power supply is just a short. In this step, we don't care about the DC levels, we suddenly are only worrying about AC.

So now we have a $20k\Omega$ impedance on the plate (because $1/100k\Omega + 1/25k\Omega = 1/20k\Omega$ we just treat it as two resistors in parallel). This means the AC flow at 250V on the plate through a $20k\Omega$ resistor

Figure 5: Drawing the DC load line, setup line, and dynamic load line.

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would be 12.5mA, a little higher than the DC value of 10mA. So we draw the second line, the dynamic load line between 250V and 12.5mA. We will call this the setup line.

Mark the quiescent point on the DC load line, that is, the point where the 2V curve touches the load line. Now, we draw a third line, parallel to the setup line, but passing through the quiescent point. This is the AC load line. This line can be used to determine the plate voltage given the grid voltage, which is a very useful thing to know.

If the load is infinite, the AC and the DC load lines will be the same, and you will sometimes see directions to use the DC load line for various calculations where the load impedance is high. It takes so little time to draw two more lines that the added accuracy is worth the trouble. Note that as the load impedance drops, the two lines get farther and farther apart. Another way of thinking about the AC load line is that it is the same as the DC line, corrected for load impedance.

To begin, since we are working with the grid voltage between -1V and -3V, we can look at the points where the AC load lines intersect the -1V and -3V grid curves, which are at 90V and 200V, respectively. So a 2V swing on the grid produces a 110V swing on the plate, which is a gain of 55 for that stage.

You could have worked that out approximately by knowing the μ of the tube and the plate resistance, or you could have worked it out precisely by solving a set of differential equations, but with the graphical method you can work it out with a pencil and simple arithmetic. You can do this in reverse too, to figure out what value of resistor you will need in order to get the correct gain for the circuit.

There is a numerical way of drawing the AC load line, which many textbooks will describe, but frankly the graphical method is easier. However, if you are reading a book and the authors suddenly just take some resistances and plug in some numbers and get the load line, that's what is going on. You can do that too, if you like arithmetic.

Resources

E. Leon Chaffee, *Theory of Thermionic Vacuum Tubes*, McGraw-Hill Book Co., Inc. 1927.

F. Langford-Smith, *Radiotron Designer's Handbook*, Radio Corporation of America, 1953

Pete Millett's DIY Audio pages, <http://pmillett.com>

RCA Electron Tube Handbook. HB-3, Radio Corporation of America, First Edition January 1, 1940.

J. F. Rider, *Inside the Vacuum Tube*, John F. Rider Publishers, 1945.

Digression: Calculating Some Constants That Aren't Constant

You'll notice that the tube handbook lists a lot of useful tube parameters that people who use numerical methods need, such as the plate resistance, R_p , the free gain μ , or the transconductance G_m .

The plate resistance value on the datasheet is actually the slope of the characteristic curve at the quiescent point. This can be worked out by drawing a tangent line on the curve and measuring. Since the quiescent point you select might not be the same one the people who wrote the tube handbook selected, the R_p you get with the graphical method will be much more accurate.

We already described how to calculate the μ at a given operating point for the tube. Given these two measurements, you should be able to figure out for yourself how to calculate G_m .

Step Four: The Cathode Bypass Capacitor

You can see here that we are using a cathode bypass capacitor to give the highest possible gain into the load. The actual loss through a 200 Ω resistor into a 100k Ω load is pretty minimal, but we can eliminate it completely with a bypass capacitor. How big a value do we need? A good rule of thumb is that the bypass capacitor should have an impedance of 10% of the cathode resistor. In this case, that would be 20 Ω , so using the reactance chart in your handy handbook, we get around 300 μ F at 25Hz. That's a very large value and will require an electrolytic, large enough to suggest that perhaps we don't want to bother with cathode bypassing on this stage today.

Step Five: Calculating Output Power

In the process of calculating the gain, we worked out that we had a 110V swing at the plate. We can use this along with the load impedance to determine the output power. Since $P=IV$ and $V=IR$, $P=V^2/R$ and $110^2/100,000$, the peak power is 0.12W. Which is about where we'd expect to be for a small signal device. This calculation is more useful for power stages with lower impedance loads, obviously, but just as easy to do. Divide by $\sqrt{2}$ to get RMS power.

Step Six: Calculating Harmonic Distortion

A good estimate for harmonic distortion can be found using Equation 1:

$$\frac{(I_{p \max} + I_{p \min}) - I_{p \text{quiescent}}}{(I_{p \max} - I_{p \min})} \times 100$$

We know the maximum and minimum plate currents over that 2V input swing are 5.5mA and 1mA respectively, by looking at the AC load line. We

know quiescent current is 6mA. So we get:

$$\frac{6.5 - 60.5}{4.5 \times 29} = 0.06 = 6\%$$

See, I told you earlier that the linearity didn't look very good, but now you can calculate what it's going to be. You can also try moving the bias point and recalculating. But it doesn't get much better with such a large signal. Try using a 6SN7.

Additional Thoughts

Can this be used for pentodes, whose curves are shaped very differently? You bet! You'll have to go through an extra step to set the screen grid voltage, but it all works.

Can this method be used for junction field effect transistors (JFETs) and metal-oxide-semiconductor field-effect transistors (MOSFETs)? Yes! You can't use the equations in the *Radiotron Handbook* because the parameters are measured differently today, but the curves are the same and you can still work everything out from the curves.

Can it be used for bipolar transistors? Sort of. The method for transistors is slightly different and requires twisting your head around a bit but there

About the Author

Scott Dorsey has a degree in electrical engineering, during the pursuit of which he worked in the broadcast and recording industries. After several years working at a major studio, he took a job with a defense contractor. This left him time to do live concert recording for acoustical music and to design and build audio devices for personal use and on contract to several audio manufacturers and importers. He is a regular contributor to several audio magazines. He has been publishing equipment reviews and DIY projects since the mid-1980s. He is probably best known in the general audio community for his retrofit electronics designs in inexpensive Oktava, AKG, and Feilo microphones.

is a similar graphical method for transistors too.

Can you extend the curves on the datasheet out if they don't go far enough? Maybe, but the curves are extended throughout the normal operating envelope of the tube. So if you're wanting to work in a section where the datasheet doesn't have curves, you are becoming a test pilot and you can no longer expect all tubes of the same type to operate the same way out there.

Conclusion

The curves provide the parameters of the system inside the tube. The load line provides the parameters of the system outside the tube. The two of them together describe how the tube functions in-circuit. Using this method, you can easily predict tube circuit operations. 



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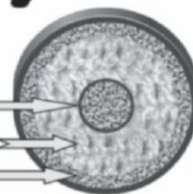
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Experimenting with Tube Sound

Automatic gain control applied to solid-state and vacuum tube audio paths in this easy-to-build stereo preamp allow interesting design, construction, and listening experiences.

By
Bill Reeve



Photo 1: This project is an easy-to-build stereo preamplifier with matched vacuum tube and solid-state signal paths.

The difference in sound reproduction between vacuum tube and solid-state amplifiers is much discussed. However, it is difficult to provide the opportunity to compare the two sounds. A solid-state amplifier can accurately reproduce recorded sound. However, the nonlinearity of a vacuum tube's plate characteristic curves creates low-level harmonics that can give music a sound described as "warm," "full," or "rich." The effect has been described as being similar to music played on a grand piano in which the instrument's strings vibrate at harmonic frequencies of each struck note, creating a subtle chord-like effect.

This project is an easy-to-build stereo preamplifier with matched vacuum tube and solid-state signal paths. A single knob allows the listener to transition between the two sounds (**Photo 1**).

To provide an accurate comparative listening

experience, the paths must be relatively well matched in phase, frequency response, and passband gain. Stable phase and frequency matching is achievable during amplifier design and testing. However, gain tends to drift.

Matched channel volume (gain) is very important in a sound comparison because people reliably prefer louder (even slightly louder) music. Having previously built tube and solid-state comparison systems, I found that the biggest challenge is maintaining accurate volume matching between the tube and the solid-state paths. In this design, a simple analog automatic gain control (AGC) circuit keeps the two channel volumes matched.

Design Overview

This preamplifier is intended to be both easy to build and easy to use. To avoid the dangerously high B+ voltages needed by most vacuum tubes, I used 12AE6A triodes. These low-voltage vacuum tubes were developed in the 1950s for use in 12V automobile radios. To provide circuit power, I repurposed a regulated (and ground-isolated) 12V, 1A AC-to-DC power supply that I believe once powered a desktop printer.

The only external interfaces are input and output RCA jacks and a blend knob. Volume can be adjusted at the input signal source. I used wire-wrap sockets to make building the circuitry relatively easy, and each path is implemented with a single quad op-amp.

To begin, I'll describe the circuit sections in the order they were designed and tested. I'll only describe one channel, as the stereo signal paths are identical.

Figure 1 shows a conceptual block diagram of one stereo path. The AGC circuit, which is a proportional

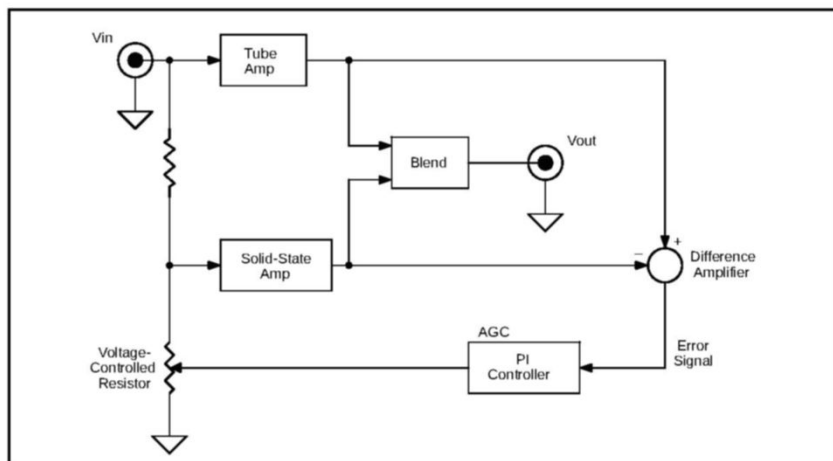


Figure 1: A conceptual block diagram of a signal path

and integral (PI) controller, attenuates the solid-state amplifier's input signal so that the path's output volume matches the output volume of the vacuum tube path. The voltage-controlled resistor is a JFET transistor. The difference amplifier includes rectification and low-pass filtering to create an error signal proportional to the average difference in volume between the two paths. This error signal drives the AGC.

Mechanicals

Since the amplifier is intended to be moved from place to place, I built it into a cast-metal enclosure. I recessed the tubes into the lid of the enclosure, and supported them with rubber grommets, to enable the amplifier to more easily survive a life of traveling.

Photo 2 shows the enclosure before the circuit was installed showing the mechanicals only. **Photo 3** shows how the inside of the preamp looked when it was completed.

Circuit "Ground" Reference Voltage

The 12V input is divided into +6V and -6V by adding a "ground" reference with a single op-amp buffering the reference voltage. The 2Ω resistor adds stability to the op-amp, since it is driving a capacitive load (**Figure 2**).

Blend Circuit

The solid-state and vacuum tube signal outputs are each applied to either end of a $500k\Omega$ potentiometer. A non-inverting amplifier buffers the wiper. This enables the output to transition between the vacuum tube and solid-state signal paths. The right and left stereo paths each use a $500k\Omega$ potentiometer, and these two potentiometers are ganged together, so one knob controls both (**Figure 3**).

Vacuum Tube Audio Path

Once the power supply voltages were in place, I built the vacuum tube channel. After the vacuum tube path was characterized, I could design and build the solid-state path to match it. A relatively low gain (between $2\times$ and $5\times$) would make the amplifier generally useful. Since common cathode amplifiers have high gain, an input voltage divider was necessary.

For maximum signal swing, the nominal plate voltage should be at the power supply midpoint (0V). This dictated a $24k\Omega$ plate resistor. A 750Ω cathode resistor provided a plate-cathode current of about 0.5mA current (see the operating point diagram from the tube datasheet shown in **Figure 4**). The vacuum tube path is shown in **Figure 5**.

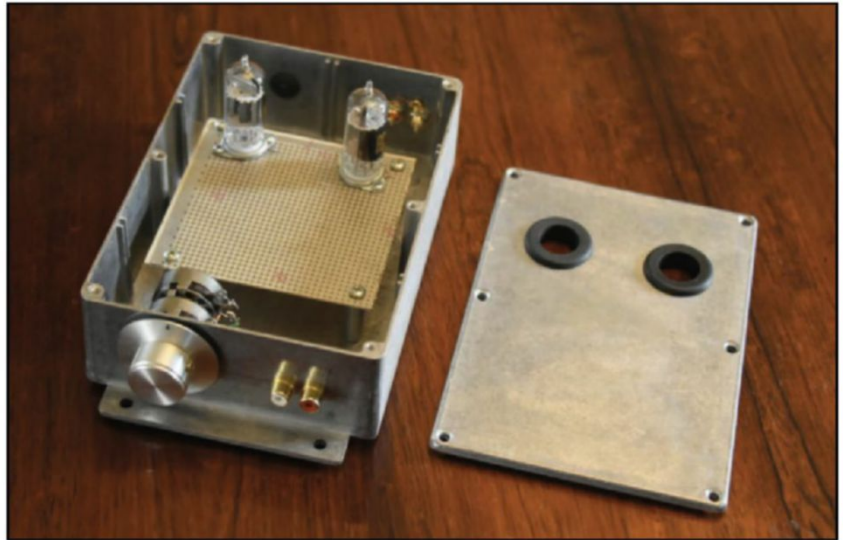


Photo 2: Mechanical enclosure prior to installing the circuitry



Photo 3: An interior view of the completed project

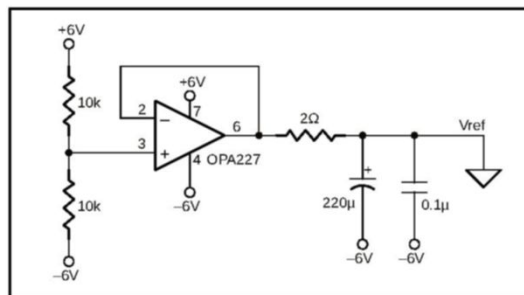


Figure 2: Circuit that provides the system "ground" reference

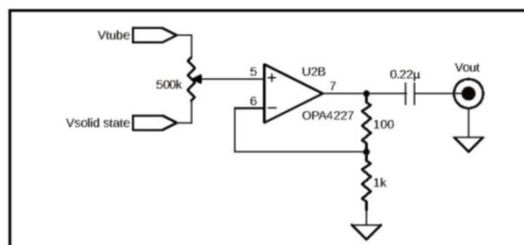


Figure 3: Signal blending circuit

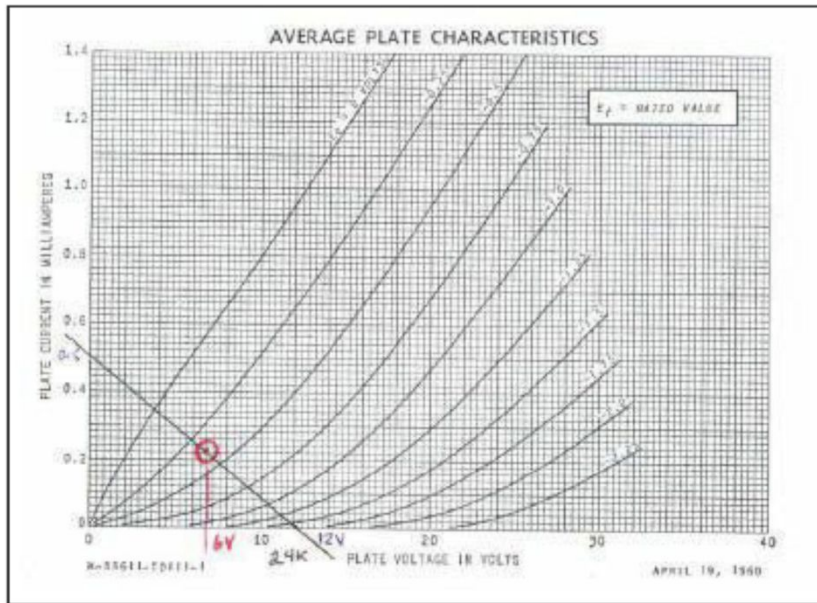


Figure 4: Operating point selection

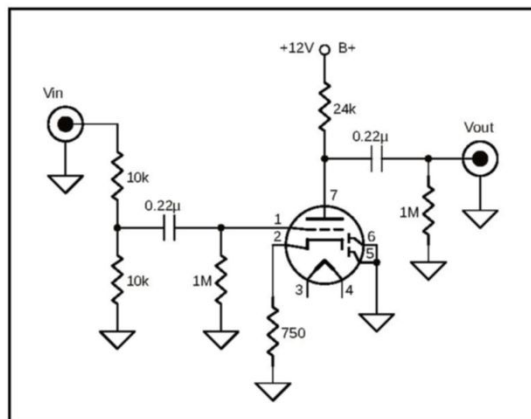


Figure 5: Vacuum tube path schematic

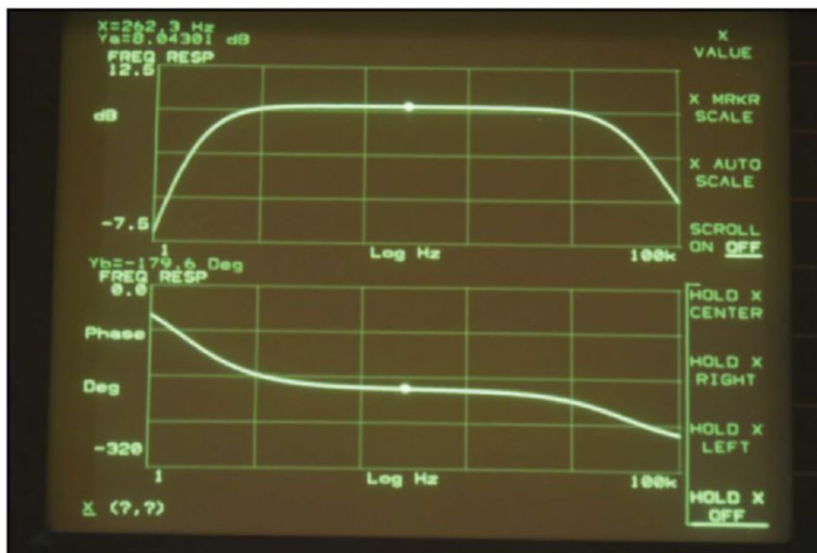


Figure 6: Tube path transfer function (gain above and phase below)

An output-over-input sweep of this circuit yielded a low-frequency roll-off at about 3Hz and a high-frequency pole at about 30kHz (created by stray capacitance). The passband gain is 2.5 (8dB). **Figure 6** shows the transfer function of this vacuum tube circuit.

Solid-State Audio Path

Figure 7 shows the schematic of the solid-state path. The input to the solid-state path is a voltage divider made up of an upper resistor and a lower 2N5457 JFET. The junction field-effect transistor (JFET) functions as a variable resistor. The two 470kΩ resistors linearize its response. As the output voltage from the AGC (which is the JFET gate-source voltage) goes more negative, the JFET drain-source resistance increases. **Figure 8** shows a graph of that relationship which I measured for one of the JFETs used in the circuit.

A nominal operating point of about -1V, or about 500Ω, determined the 2kΩ voltage divider's value. Such a low input impedance drove the input coupling capacitance up to 330μF to approximately match the vacuum tube stage's low-frequency performance.

Since the vacuum tube path is inverting, the solid-state amplifier also needs to be inverting. We want the voltage supplied to the gate of the JFET by the AGC amplifier to nominally operate at about -1V, so the solid-state path gain needed to be less than the vacuum tube path gain without AGC feedback (when the JFET's gate-source voltage was 0). The 47pF feedback capacitor approximately matches the vacuum tube path's high-frequency roll-off.

Difference Amplifier

The difference amplifier provides the AGC circuit with an error voltage representing the average (rather than instantaneous) volume difference between the two output signals. The schematic for the difference amplifier is shown in **Figure 9**. The small-signal Schottky diodes rectify, and the difference amplifier then low-pass filters and subtracts these two audio signals to create this error signal for the AGC. The difference amplifier's low-pass time constant should be longer than the AGC controller's.

The controller's ability to match the solid-state output volume to the tube signal volume can only be as good as the error signal it receives. Error signal bias will result in a residual volume difference between the two outputs (a control engineer would call this "hangoff").

Bias can be caused by component variations. For example, differences in input resistor values or the Schottky diode forward voltages. A significant source of bias in this circuit is the fact that the difference

amplifier is fed by an op-amp (from the solid-state path) and a common-cathode amplifier (from the vacuum tube path).

Bias can be compensated by selecting appropriate values for the difference amplifier's input resistors. The tube path's 820k Ω value was determined by brief experimentation. I simply reduced that resistor from its nominal 1M Ω value until the controller matched the two output signal volumes to my satisfaction.

Automatic Gain Controller (AGC)

As mentioned previously, the AGC is a proportional and integral controller. It receives the error voltage signal from the difference amplifier, and applies a control voltage to the JFET's gate, which modulates the JFET's drain-source resistance. The JFET's drain-source resistance works as part of a voltage divider to change the signal amplitude at the input to the solid-state path. Thus, the solid-state path's average output voltage amplitude will match the tube path's average output voltage amplitude.

Bill of Materials

Quantity	Item
Capacitors	
2	47pF
2	0.1 μ F
6	0.22 μ F
2	1 μ F
8	220 μ F
4	22 μ F
2	330 μ F
ICs	
1	OPA227
2	OPA4227
Potentiometer	
1	500k Ω , 2 ganged
Resistors (0.25W)	
1	2 Ω
2	100 Ω
2	750 Ω
2	1k Ω
4	2k Ω
10	10k Ω
2	24k Ω
2	120k Ω
4	470k Ω
2	820k Ω
10	1M Ω
Schottky Diodes	
4	SD101A
Transistors (JFETs)	
2	2N5457
Vacuum Tubes	
2	12AE6A

The schematic for the PI controller is shown in **Figure 10**. The controller has a free integral (so it can track an average gain error, a zero at about a

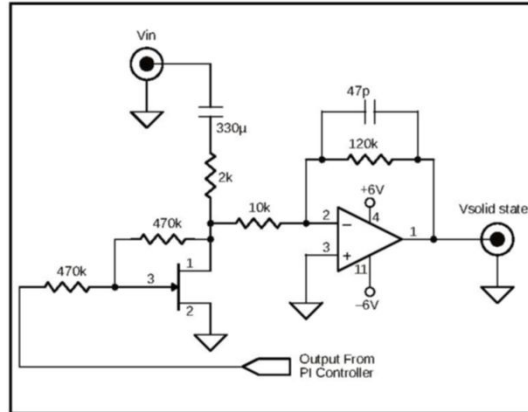


Figure 7: Solid-state path schematic

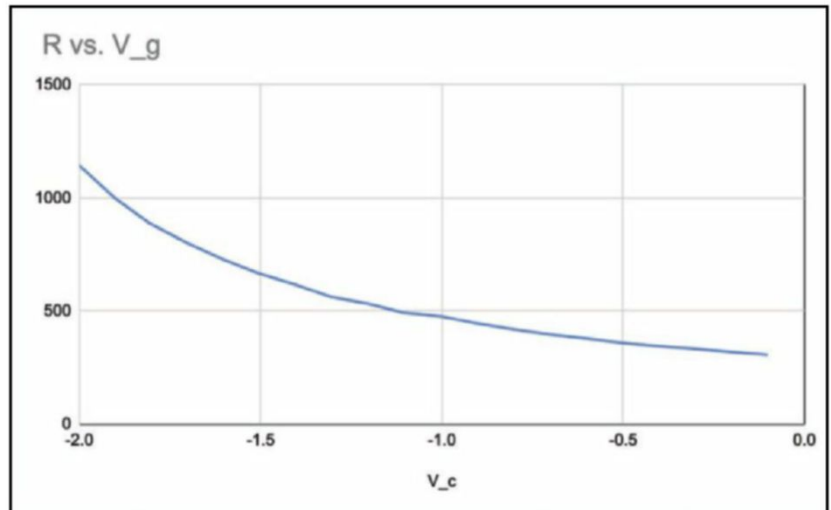


Figure 8: The JFET drain-source resistance in ohms (Ω) as a function of gate-source voltage

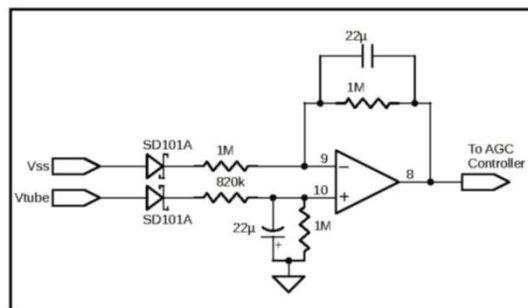


Figure 9: Schematic for the difference amplifier

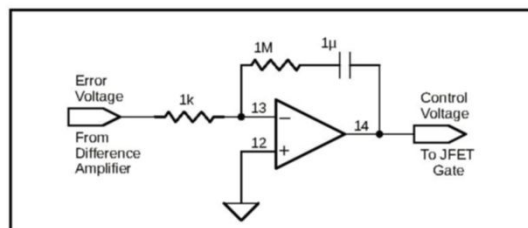


Figure 10: Schematic for the PI controller

sixth of a Hertz and a high-frequency gain of about 1,000. Combined with the difference amplifier, the automatic gain control system works well, and there is no noticeable volume difference as the listener changes between the two signal paths. The circuit

analysis is somewhat involved, but this component selection provides both good response and stability.

Some Final Thoughts

As mentioned previously, power, input, output RCA jacks, and the blend knob are the only external interfaces. Volume is managed at the signal source, and circuit power needs to be unplugged to be disconnected. These are, of course, up to the designer's discretion. **Figure 11** shows a complete schematic of one of the stereo paths.

A difference between the solid-state and vacuum tube outputs can be seen in the frequency domain. The power spectral density plots below show the harmonic content of the outputs when the input to both paths is a 10kHz sine wave.

In **Figure 12**, the upper trace is the vacuum tube path, and the lower trace is the solid-state path. The larger harmonic content in the vacuum tube path is evident.

About the Author

Bill Reeve has been engaged with electronics since the days of slide rules and vacuum-tube shortwave radios. Bill has been a Technical Program Manager at Google for the last seven years, after retiring from a 28-year career as a Program Director at Lockheed Martin. In the 1970s and 1980s, he worked as an engineering geologist in Colorado and California, and for two years in the 1990s as an Engineering Director at the DeWalt Industrial Tool company. Bill has Masters degrees in Geology, Mechanical Engineering, and Electrical Engineering. His interests include rock climbing, wood working, motorcycling (on- and off-road), designing audio electronics, sailing, and repairing vintage audio amplifiers and test equipment (oscilloscopes, spectrum analyzers, power supplies, and meters). He has also designed spacecraft electronics and mechanisms.

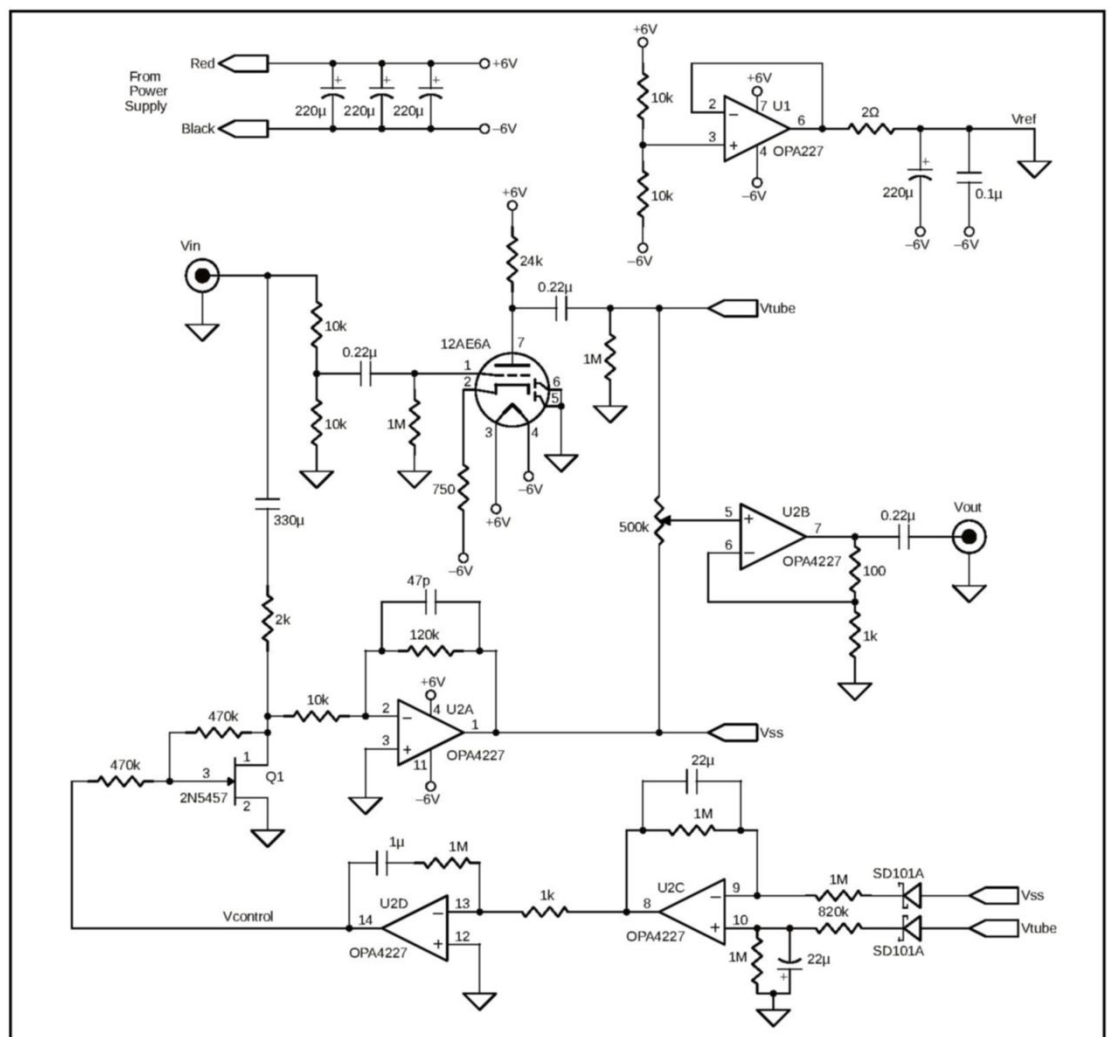
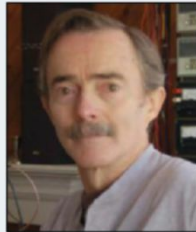


Figure 11: This schematic shows the power distribution and one of the two stereo paths.

You will find that real-time use of the blend knob is a remarkably ineffective way to discern differences between two audio outputs. As you might already know, only very pronounced differences in audio quality can be detected by relatively quickly switching back and forth between audio sources.

To evaluate subtle differences (i.e., the difference between two good-quality amplifiers), a person needs to listen to one source long enough to have the sound of that specific music from that specific source become familiar. This forms a personal baseline for that music. Then, when listening to the same music through a different source, one can notice changes to their "baseline." This is a way to become aware of subtle differences in clarity, bass depth, and low-level harmonics.

This experiment in tube sound can be an enjoyable project to design, build, and use. It makes a conversation-starting desktop headphone amplifier, or a preamp for a power amplifier system.

In your own design, you can experiment with (among other things) the vacuum tube bias point to a less (or more) linear region, changing the parameters

of the AGC and the differential amplifier filters, as well as modifying the audio amplifier passbands.

I hope you enjoy the design, building, and listening experience. 

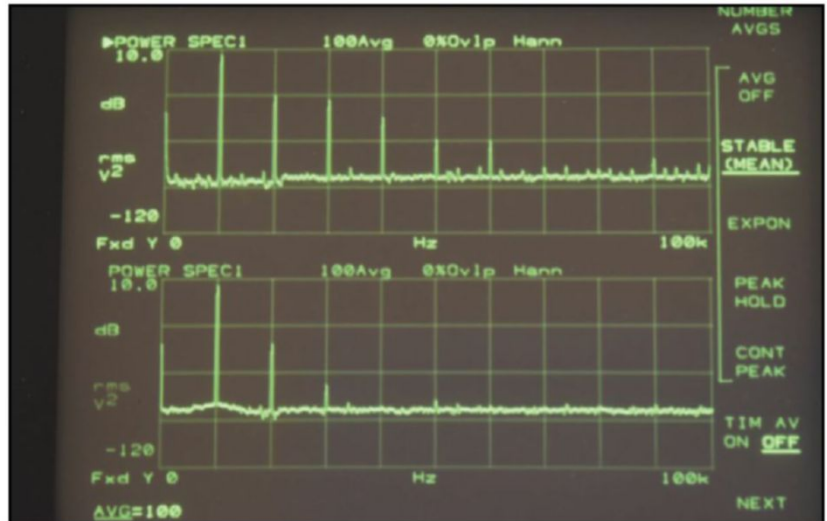


Figure 12: Comparison of a vacuum tube (upper) and a solid-state (lower) frequency response to a 10kHz sine wave input

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The Cranberry

A Versatile Stereo Vacuum Tube Amplifier

By
Bruce Heran



Photo 1: For this project, Bruce Heran built a versatile stereo vacuum tube amplifier that has a single chassis and a shared power supply that he calls the "Cranberry" due to its unique color.

In an effort to expand his DIY amplifier collection, Bruce Heran designed this stereo vacuum tube amplifier using a concept that originated in the 1940s and was initially intended for a budget amplifier.

I admit it. I am a diehard DIY enthusiast. Worse, I love high-quality audio. Because of that I have more amplifiers than I can possibly use. That doesn't stop me from designing and building more. It is something I like to do. I like them to be distinctive as well. I picked the color of this one and then named it The Cranberry (**Photo 1**). I am not sure I would pick the same color again. It just doesn't match anything else I own. Fortunately, the color of the chassis does not impair the sound in any way.

Thoughts Prior to the Build

This amplifier is a bit of a variation for me. I generally prefer mono block amplifiers as they are easier to place near the speakers they power. That said, many folks find a single central amplifier better suited to their needs. I have received many requests to design a stereo version using some of my middle-sized mono blocks. Just sticking two together would work, but it would certainly not be all that efficient so I passed on that option.

A single chassis with a shared power supply was clearly the best choice. The need for a way to control gain and input selection was clear as well. The active circuitry is quite similar to the mono blocks I designed and have widely posted on the Internet. There are several variations that have evolved from the original ones first built in 2008. Many similar-sized amplifier designers seem to be engaged in a "horsepower" race to see who can make the most power output.

I personally think that it is more important to place sound quality over maximum power. This results in some design consequences. I find that push-pull Class "A" operation seems to provide the quality of sound I like and indeed hope you will as well. Additionally, I find that ultra-linear operation is advantageous. I tried both pentode and triode modes and neither provided the quality of sound I sought. Naturally, there are some aspects of this determination that are less desirable. For one, efficiency is lower. Typical efficiency of the ultra-

linear mode is about 33%. Pentode mode is a bit more efficient and triode mode less efficient. A related consequence is heat generation. This amplifier generates a significant amount of it. Still, I believe the ability to reproduce musical source material in a high-quality manner is well worth it.

The Circuit Design

The actual circuit is different from many other amplifier designs. The concept originated in the 1940s and was intended for a budget amplifier. Component parts back then were scarce, costly, and not as varied as we have now. So in the search for a little more power, eliminating a few parts made sense. The design was for a push-pull amplifier without a phase inverter stage. That design slaved one of the output tubes to the other. It sounds complicated and seems like it would not work well. I suspect that back then that was probably true. We now have many choices for components and the limitations of the early design are actually pretty easy to overcome.

The output stage in the original design used a resistor in the cathode circuit of the output tubes. One tube was fed the signal and the second had its grid at ground potential. The second tube received its signal between the cathode and ground. Thus, it was out of phase with the tube receiving the input signal directly. This provided the phase inversion needed for a push-pull output. This forced both tubes to conduct all the time. The linearity of the stage was relatively poor as there was no provision to match the current between the two tubes. An additional consequence is that voltage gain of the stage was low. If you look

at fairly common phase inverter stages now used in many amplifiers you can see some similarities.

So how is this design different? First, it uses a constant current source (CCS) in the cathodes in place of the resistor. I use a type LM317HV integrated circuit (IC) regulator for that. A number of individuals have been critical of that component, mostly about inherent noise levels and nonlinear performance. Regardless, they do work well and in this design and do not exhibit those problems. The only limitation I have discovered is that the voltage across them cannot drop below about 4V. Distortion results if you do.

A second potential limitation would seem to be the maximum voltage rating of the IC. The high-voltage version is rated at 57V. It would appear that this would be easy to exceed at full power output of the amplifier as the peak cathode voltage could exceed that. Either the ICs are considerably under rated or AC signal peaks don't affect them the same as a DC voltage would.

Another modification from the original design is the use of a balance control between the cathodes. This allows for adjustment of the current flow between the tubes at idle. This is accomplished by effectively altering the DC bias of each tube in a sort of see-saw manner. Since there is a CCS, if you increase the current in one tube, the other is forced to decrease. This does point to an issue though. The tubes need to both be reasonably well matched and have reasonable thermal stability. Most power tubes are, but I have run into some that refuse to stay balanced. It presented itself as a thermal issue.

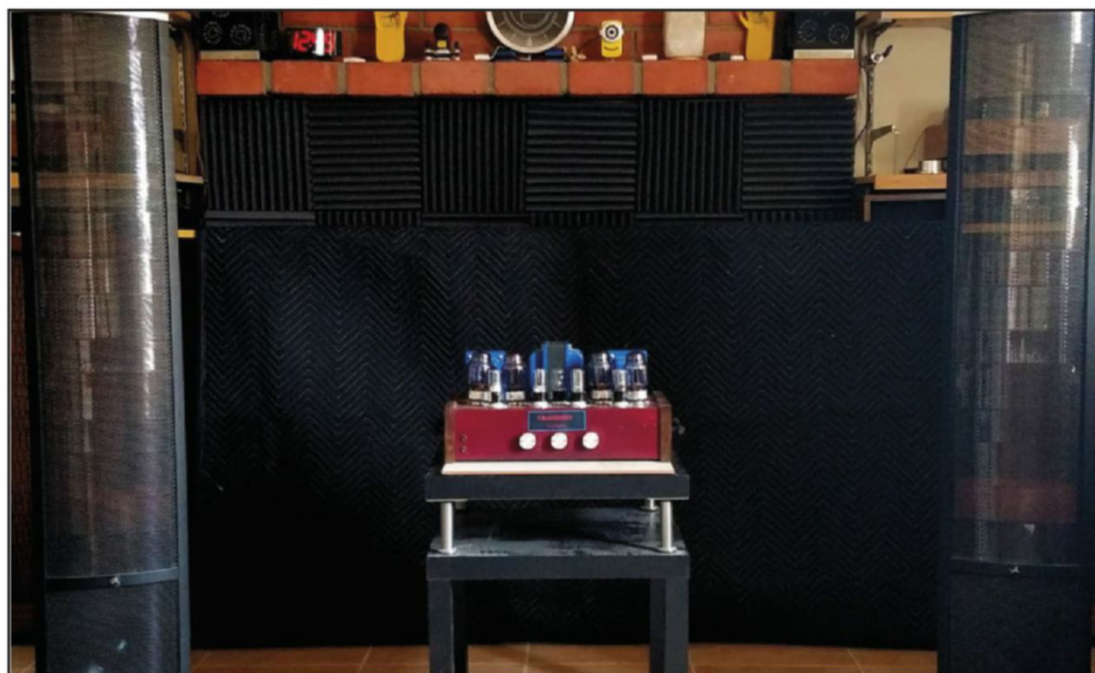


Photo 2: The amplifiers were tested to drive a pair of MartinLogan Electrostatics.



Photo 3: The interior layout shows the partial modular design with most wiring point to point.

One tube would start to get hotter and thus alter its bias requirements, and the whole stage would become unbalanced.

Careful selection of the resistors from the tube grids to the source and ground eliminates this with the majority of tube types and brands. Once set, the balance between the tubes will be fine for several months. A drift of up to about 10% has no effect on the sound. A reduction in output will occur above that level, but the sound is still fine.

A number of individuals will question the source of bias for the tubes. The CCS establishes it based on the programmed current flow. There is no adjustment for bias per se in this design. A third change from the original design is the use of impedance matching components in the grid circuits. The original design had the slave grid directly tied to the ground. I have

placed a capacitor and resistor between the grid and the ground that match the ones in the other tube's grid. It is not a perfect match as the output impedance of the driver circuit is not zero. However, it is close enough.

The Driver Stage

The driver stage is a fairly common type. It is known as a SRPP. I have seen at least three different translations for SRPP. The most common seems to be series regulated push pull. The variation I use is a fully symmetrical one. I find that when carefully designed it will provide a very large linear voltage swing, have low distortion, and reasonable power supply noise rejection. The biggest trade-off is that the gain is about half of the maximum possible for the type of tube used.

A complication of all similar stages is that it is possible to exceed the tube ratings for voltages between any of the heaters and any of the cathodes. Using dual triodes like I do, this consideration is important. The solution has been well documented. It involves isolating the heaters from the ground and providing a DC voltage reference to the entire heater circuit. I use a reference of approximately one-third of the voltage applied to the anode of the upper tube for this. There are some limitations and it restricts the actual stage anode voltage to about 300V to 350V.

I have additionally found that the manufacturer's ratings for heater-to-cathode voltages only apply to new tubes. As they age, the value drops significantly so a juggling act is needed between maximum anode voltage and the heater reference voltage to keep all the electrodes in the safe zone. A heater-to-cathode failure or arc over will result in a massive

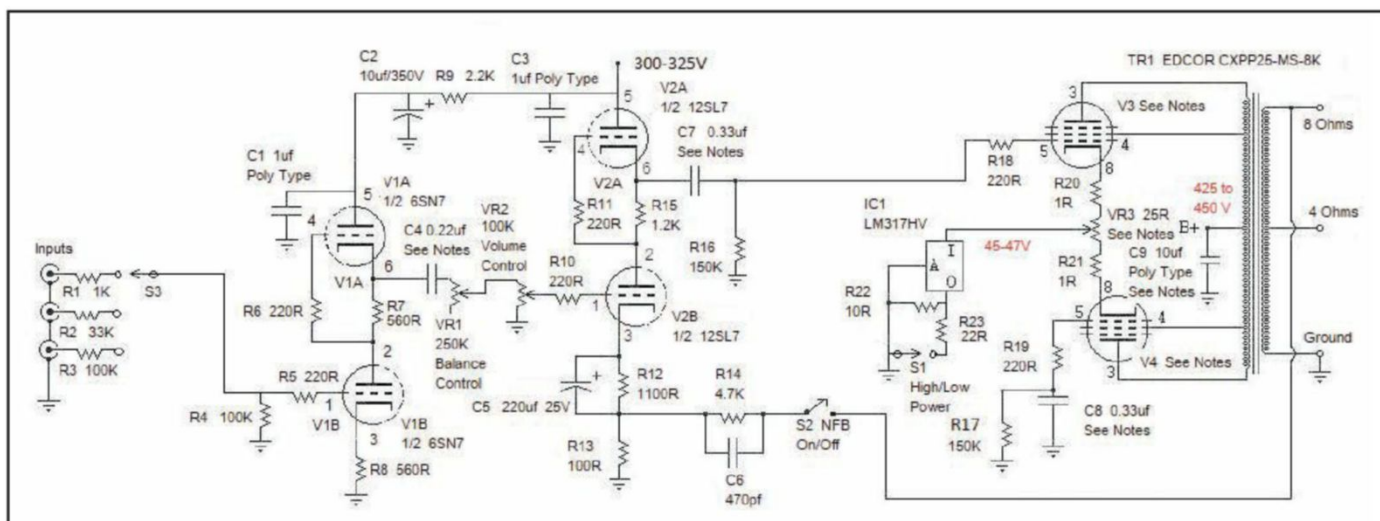


Figure 1: This schematic shows one channel of the active circuitry.

spike to the output stage and in turn to the speakers. Besides being quite loud, it might possibly cause speaker damage. A side benefit of the isolation and reference voltage is that the heater circuit tends to scavenge electrons that don't go toward the anodes. This makes for a slight but measurable reduction in tube noise.

Additional Gain Stage

Another difference between earlier designs is the incorporation of an additional gain stage. There were three main reasons for doing this. First, some signal sources were not strong enough to drive the power amplifier stages to full output. Second, I wanted to add a channel balance control and have it isolated from the input signal. Last, I wanted to add the ability to select input sources. These features would normally be included in a line stage preamplifier for use with a pair of mono blocks. Effectively, the extra stage eliminates the need for a line stage preamplifier.

I have received a number of questions about this design. A question that is often asked is: What tube types can be used? Having the ability to set the idle current to any value desired makes the amplifier quite versatile. Pretty much any type of pentode power tube that can operate with an anode voltage between 400V and 475V will be fine. This includes the KT66, the KT77, the KT88, the KT120, the EL34, the 6550, and numerous similar tubes.

One tube that would seem to be overkill is the KT150. It certainly can be used at higher dissipation than any of the others. However, I have personally found it works extremely well at a lower power setting. Using about 65mA each and 450V on the anodes, a pair will deliver approximately 20W RMS

and in my opinion sound better than any other combination. My personal system is configured that way.

A common second question is: How often is it necessary to replace the tubes? It is something I can't accurately answer. The reason is simple; I have never had to do it yet. The tubes shown in Photo 1 of the amplifier (blue glass JJ KT88s) are the first ones I bought in 2008 and have thousands of hours use. They test and perform fine. They were my "daily drivers" until last year when I switched to using the KT150s in my main system.

Another common question is: How sensitive is the amplifier to speakers? In short, it really doesn't matter what they are. I use some of the arguably most difficult to drive main speakers, a pair of MartinLogan Electrostatics (**Photo 2**). They look electrically like huge capacitors and their impedance drops from 4Ω midband to 1.2Ω at 20kHz. They are a tough load for many amplifiers, but no problem for these.

While I am not convinced that specifications will tell you how something sounds, I reluctantly will provide the following data. The typical power output is in the 15W to 18W range for tubes running the low power setting (except the KT150s) and 25W to 33W range for the high power setting. Measured response at 1W and full power levels is within 0.5dB from below 20Hz to past 20kHz. Distortion at half power (depends on tube choice to some extent) is under 2% maximum. Levels below 1% are usual in the 1W RMS region. The amps shown in Photo 1 have KT150s measured at 0.34% at 1W. Worst case for any combination at full power is under 4%. The signal-to-noise ratio (SNR) is typically about -85dBW to -90dBW.

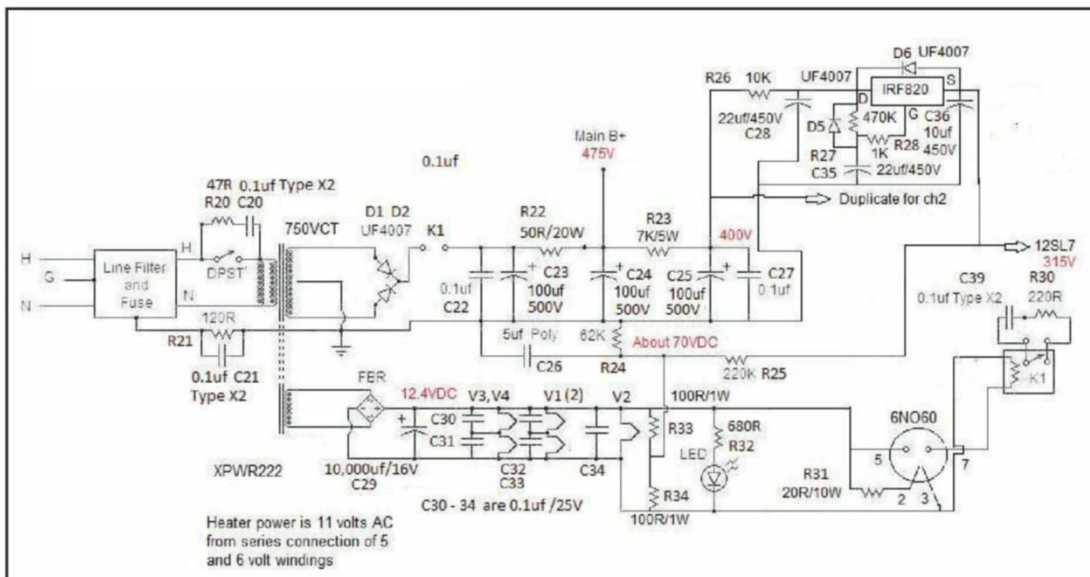


Figure 2: The power supply is shared between two channels.

Components for The Cranberry

Active Circuitry Section (all to be parts duplicated for second channel)

Resistors all are in ohms and 0.5W or greater unless specified

R1	1kΩ	C1, C3	1μf poly 350V
R2	33kΩ	C2	10μf /350V
R3, R4	100kΩ	C4	0.22μf *
R5, R6, R10, R11, R18, R19	220Ω	C5	220μf /16V
R7, R7	560Ω	C6	470pf
R9	2.2kΩ/1W	C7, C8	0.33μf*
R12	1.1kΩ	C9	10μf /500V poly*
R13	100Ω	VR1	250kΩ balance control
R14	4.7kΩ		Bourns Dual MN series
R15	1.2kΩ	VR2	100kΩ stereo volume
R16, R17	150kΩ		Control (I used Alps)
R20, R21	1Ω	VR3	25Ω 1W linear
R22	10Ω /1W		
R23	22Ω/1W		
TR1	EDCOR CXPP25-MS-8K		
V1	6SN7		
V2	12SL7		
V3, V4	Various *		

* See text

Power Supply Component Listing

R20	47Ω	C20, C21, C39	0.1μf type X2
R21	120Ω/1W	C22, C27	0.1μf 500V poly
R22	50Ω/20W	C23, C24, C25	100μf 500V
R23	7kΩ/5W	C26	5μf/350V poly
R24	62kΩ	C28, C35	22μf/500V
R25	220kΩ	C29	10,000μf/25V
R26	10kΩ	C30, C31, C32, C33, C34	0.1μf/25V
R27	470kΩ		10μf/350V
R28	1kΩ	C36	EDCOR XPWR222
R30	220Ω	Transformer	Generic IEC input
R31	20Ω/10W	Line Filter	UF4007
R32	680Ω	D1, D2, D5, D6	IRF820
R33, R34	100Ω/1W	IC	50V/5A
GNO60	Amperite Thermal Delay Relay	FBR	Your Choice
K1	Generic 12V coil relay	LED	

Alternative Timer Components

IC	LM555 (or NE555)	C1	220μf/25V
R1	220kΩ	C2	0.2μf/25V
R2	100kΩ	D1, D2	1N4001
K1	As in other circuit		

All the components and their values are detailed in here.

About the Author

Bruce Heran lives in Sierra Vista, AZ, and is an avid DIYer. He has been involved in electronics for 59 years. He built his first project—a one-transistor amplifier—when he was 13. He has a BS in Biology and studied electrical engineering for two years. Electronics has always been his hobby. In May 2011, he retired from his work as a project manager of a logistics contractor with the Federal government. Now he is co-owner of Oddwatt Audio, an electronics company that specializes in vacuum-tube audio components. He is vice president of design and support, and all of the company's products are kits and assembled audio equipment based on his personal designs. He is an avid collector of vinyl records and vintage turntables.



The Build

Building this amplifier is relatively easy (**Photo 3**). Only one channel of the active circuitry is shown in **Figure 1**. The power supply is shared between the two channels (**Figure 2**). Any DIY amplifier using tubes involves some concerns. Foremost is safety. The voltages used internally are potentially lethal and builders need to be familiar with working with such hazards. The soldering technique is also important. Poor solder connections seem to be one of the most common causes of misbehavior of DIY builds. Some specific details for The Cranberry's construction are provided.

Coupling capacitors C4, C7, and C8 have a great influence on the sound. They should be quality components. Boutique ones are not necessary, but low quality ones will result in a reduction in sound quality. Capacitors C1, C2, C3, and C9 should be "poly" types for best audio performance. The power supply is actually one-half of the sound chain and these capacitors insure a low impedance signal return path. Only one C9 is needed and it is shared by both channels.

All resistors can be 0.5W unless specified otherwise on the schematic (Figure 1). Either metal or carbon film is satisfactory. Carbon composition resistors should not be used. Resistors R22 and R23 "program" the current for the output tubes. They set it at 65mA or 91mA per tube. The total through the LM317HV is thus twice that value. You can use other values by calculating different resistances. The approximate value would be 1.25 divided by the resistance in ohms. This will give an approximate value for the current for two tubes.

When selecting current values, it is important to keep the tube dissipation within specifications. I like to keep it below 85% of ratings. The LM317HV needs a heatsink. The dissipation can get to about 8W when running tubes at the higher power level. I like to mount the heatsink on top of the chassis as pair of LM317HV will quickly heat up the inside of the chassis if it is not really well ventilated. VR3 is a 25Ω linear variable resistor and is used to balance the idle current between the output tubes. It needs to be at least 2W RMS rated and larger is better. A wire wound one is OK. The worst case dissipation is at the extreme positions and could reach 1W when using the high power settings. VR1 is a 250kΩ balance control. In the center position, there is no resistance between either source and its respective output. The input stage is run open loop. The resistors R1, R2, and R3 act as voltage dividers with the input signal to allow for sources that might over drive the tube. They are optional and can be eliminated or if needed altered to higher values.

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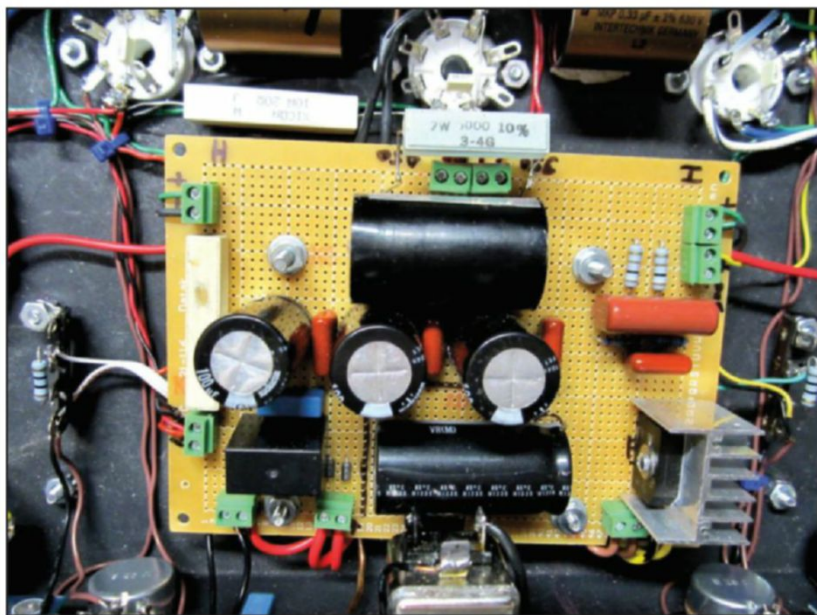


Photo 4: The power supply provides the necessary high voltage and filament power.

The specified output transformer is important. However, others may be used. The one used is an overachiever. It easily handles the nearly 200mA DC current and has an unusually wide response. Most builds will show a -3dB point at about 10Hz and an upper -3dB point well past the audio band. The phase of the output transformer is important. It is used to invert the signal. This makes the amplifier non-inverting. It also is necessary for the negative feedback (NFB) circuit operation. If it is backward, the feedback will be positive and the sound will be shrill.

An easy way to tell is if it gets louder with the NFB on, then the transformer is wired backward. You have a choice of swapping the leads between the output tube anodes and screens or changing

which tube is driven and which one is slaved. Either way works. The NFB is switchable. The amplifier will run fine without it, as it is not used for correction of sonic issues. I included it to guard against possible instability at well above audio frequencies. When built with the suggested components there is a resonance peak in the 60kHz to 80kHz range. It is large enough that I felt it might be possible to cause oscillation with some unusual speaker loads (like my ESLs). I have never seen it occur, but I would rather have the provision available just in case.

The power supply provides the necessary high voltage and filament power (**Photo 4**). Other designs are fine as long as they can provide over 250mA at 450VDC and 12VDC for heater power at sufficient current for the types of tubes used. Filtering of the high voltage is important. This is particularly the case for the input and driver stages. A capacitive multiplier is used for them (**Photo 5**).

The power supply is generally standard in design. I used solid-state rectifiers for efficiency although a vacuum tube rectifier variation could be used. I also used an "old school" thermal delay relay to prevent the high voltage from reaching the circuitry immediately. It works in combination with a normal relay and does not by itself handle the high voltage. With solid-state rectifiers the peak DC voltage can reach about 1.4 times the RMS rating of the power transformer virtually immediately. This would exceed the filter capacitor ratings and possible cause arc over in the driver and preamplifier tubes. The thermal delay relay delays the high voltage for about 60 seconds. These relays look like octal tubes and blend in well with the design. They are getting quite scarce and costly though. An alternative is to use the IC-based timer design shown separately in **Figure 3**.

AC Mains Input Filter

I always use an AC mains input filter. I find that typical AC mains now have a rather large amount of trash on them and the less that gets into the amplifier that way the better. I also directly attach the AC mains "earth" ground directly to the chassis. The actual signal grounds internal to the amplifier are not directly connected to the chassis.

The chassis ground and signal grounds are joined only via a 120Ω resistor with a parallel type X2 capacitor. This eliminates ground loops while allowing the chassis to act as both a safety shield and EMI shield. A rectifier bridge is often used in guitar amplifiers in place of the resistor and capacitor. It performs the safety function well, but I feel that it is less satisfactory in the rejection of electromagnetic interference (EMI). This is because of the forward voltage drop penalty of the rectifier.

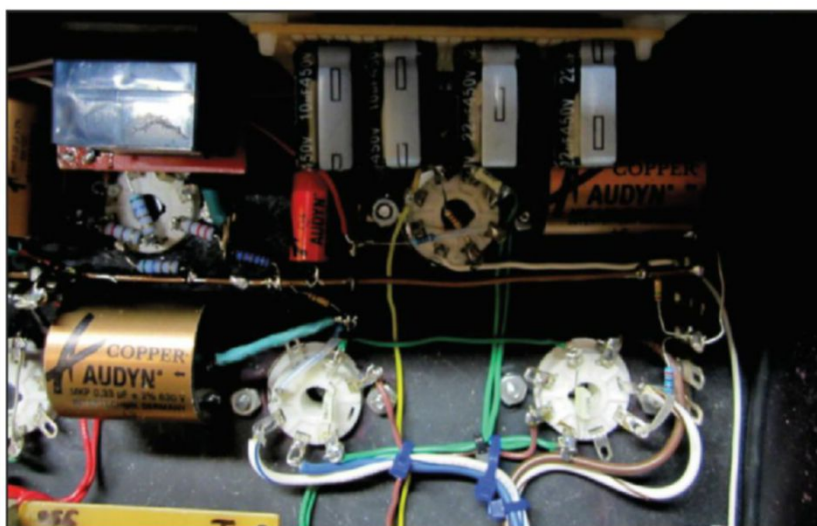


Photo 5: Filtering of the high voltage is important. This is particularly the case for the input and driver stages. A capacitive multiplier is used for them.

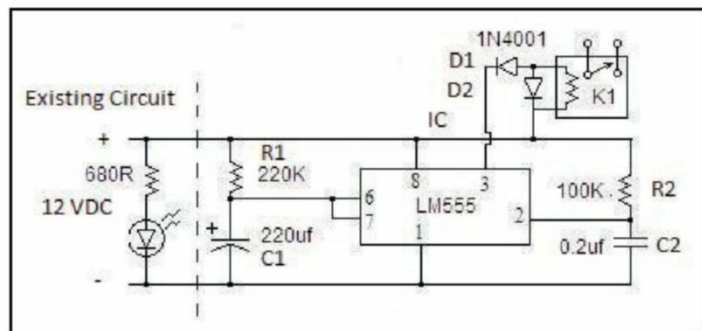


Figure 3: Thermal delay relays delay high voltage for about 60 seconds. However, due to their scarcity and increasing cost, an alternative is to use the IC-based timer design.

The Chassis

The selection of the chassis is also important. I recommend one at least one-and-a-half times as big as you think you need. This makes for easier building, a less cluttered arrangement, and much easier troubleshooting if needed. Side benefits are a cooler operation and easier component placement. I had to put handles on mine as it was quite heavy. I also consider a bottom plate mandatory. The screen-like material I used is what is known as a screen door kick panel (**Photo 6**). It comes in sheets about 50cm x 100cm. You can get it from hardware stores that carry screen door parts. There are several designs and colors. It cuts easily with scissors and can be bent into shapes to enclose sub-assemblies inside gear.

The Setup

Set up is easy. Plug in the tubes and turn it on. If there is smoke, arcing, or something else dramatically wrong, turn it off. It will often be easy to find the offending components. If this sounds odd, well it sort of is. I get numerous inquiries on similar amplifiers where the voltages are not correct on "testing." About 95% of the time it is because it was "tested" without the tubes in place.

It is not possible to accurately determine if the amplifier is functional without the tubes inserted. A common belief is that tubes are somehow delicate and will be harmed if

there is a wiring error. This is really rare. So, plug them in and turn it on.

For a quick bench check of the voltages, you should attach some sort of load as well. Voltages ought to be within about $\pm 10\%$ of what is shown on the schematic. Some variations are expected depending on the choice of output tubes and your actual AC mains voltage.

Tubes running at low power will pull down the high voltage less than ones that are running at higher power levels. A really good indication of correct operation is the balance of the output tubes. If they balance then in all likelihood the amplifier will be fine. Any other issues will be minor. If the tubes balance, then feed in a signal and listen for output. If it is not satisfactory then you need to start troubleshooting. First thing is to verify that it is wired correctly and all solder joints are sound. After that you need to get serious and hunt for the problem. I am not going to go into all the possibilities for this. Such procedures can be found online.

The Result

How does the amplifier sound? This is naturally a personal opinion. I believe you will be much surprised at the quality. I find them quiet, detailed, and very robust. They seem to be much more powerful than expected and have superior bass response. It really sounds great on my electrostatics. 

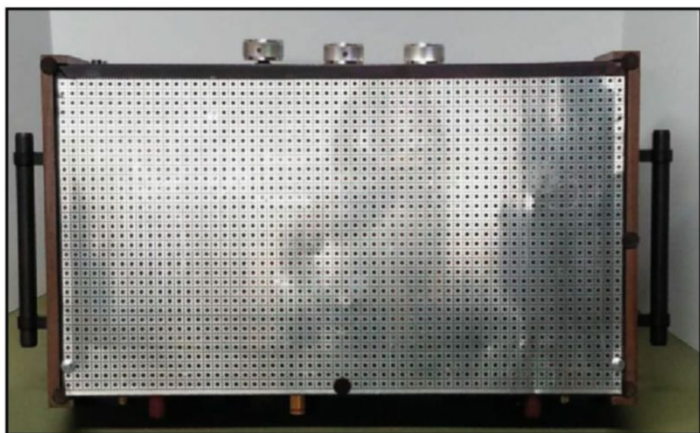


Photo 6: I also consider a bottom plate mandatory for this design and the screen-like material I used is what is known as a screen door kick panel.



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Determining the Real Output Current for Op-Amps and Fully Differential Amplifiers

By
Michael Steffes

In this article, our audio expert uses various approaches to specify output current and the evolution in op-amps and fully differential amplifier specifications. A critical parameter that should be well characterized when designing audio amplifiers.

Every recent op-amp and fully differential amplifier (FDA) release reports some sort of output current specification—but what do these really mean? Pretty much whatever that particular development group wants it to mean. The marketer wants the biggest number possible, the test engineer wants a 100% test yield number, and the designer is caught between them, while the end customers wonder what they can depend on.

Starting from the IC designers' tools, they work with the full Process Design Kit (PDK) offering process variation models with over temperature shifts for all of the separate elements in the IC design. Every final product parameter has a typical process, nominal temperature, simulated response value to produce the typical specification.

One evolution has been to take a typical 25°C specifications as T_j (junction temperature) vs. the legacy ambient temperature. A second nuance is the limit specifications for Automatic Test Equipment (ATE) needs to use only the ambient (and its range) as there is usually no die self-heating in these very fast final test screens (see Reference 1, Footnote 5). The final IC design review is largely stepping through PDK simulations showing a typical process case with

$T_j=25^\circ\text{C}$ numbers (for typical specifications), over process case typical ambient temperature range (for first cut ATE limits), and full over process case and T_j range for worst case limits—all of this is usually generated at the final IC design review—a small portion of this gets into the datasheet.

Different Output Current Definitions and Operating Assumptions

Audio amplifiers have progressed from legacy devices with no output current specification (NE5532) to only a short circuit current specification (OPA2132) to a stated maximum output voltage and current swing into a typical (usually 600Ω) load as shown in **Table 1**.

A short circuit current specification is implicitly at 0V output and obviously very nonlinear. Working through the typical 600Ω load numbers in Table 1 (at ±15V supplies), that ±13.5V swing implies a 1.5V required headroom to deliver $\pm 13.5/600\Omega = \pm 22.5\text{mA}$. Going on to the maximum output current line, that ±25mA into 600Ω implies a ±15V output swing, which works with the stated ±17V supplies. While consistent, these specifications really don't tell you what the maximum supported voltage swing

V_{OUTMAX}	Maximum Output Voltage Swing	$R_L = 600\Omega$	±13.5	±12.5	V (min)
		$R_L = 2\text{k}\Omega$	±14.0		
		$R_L = 10\text{k}\Omega$	±14.1		
I_{OUT}	Output Current	$R_L = 600\Omega, V_S = \pm 17\text{V}$	±25	±21	mA (min)
$I_{\text{OUT-CC}}$	Instantaneous Short Circuit Current		+53 -42		mA

Table 1: Output voltage and current specifications for the LME49723

would be into perhaps a 100Ω load. Is it ±2.5V implied by the 25mA maximum output current specification—or more?

An improved output V/I curve emerged over time as illustrated in **Figure 1** from the OPA1602 datasheet.

This is called a “claw” curve as it resembles the inner profile of a lobster claw. There is a lot of good information in this ±18V supply, typical process case, presentation.

- At 20mA you can see a small headroom required (0.6V for this “rail to rail output” device)
- As the output current increases (source or sink), there is a linear increase in the required headroom, indicating a resistive component to the supplies in the output stage.
- At some point, the output stage collapses and knees over in the available output voltage at higher currents.
- That “knee” has strong temperature dependence, reducing about 30% from -40°C to +125°C
- There is far less output sinking current available than sourcing current.
- There is more voltage headroom required on the sourcing current side.

This available output voltage vs. current characteristic has actually been incorporated into the most recent simulation models. This curve is generated averaging a limited number of devices from a single process lot used for characterization. In theory, you should still guardband the breakpoints down by some amount for process variation—perhaps by 20%. Looking at the 25°C curve, and assuming you need a symmetric output voltage swing using the sinking current restriction, to stay below the knee would suggest a safe ±47mA maximum output current where you could swing ±17V looking at the higher headroom sourcing

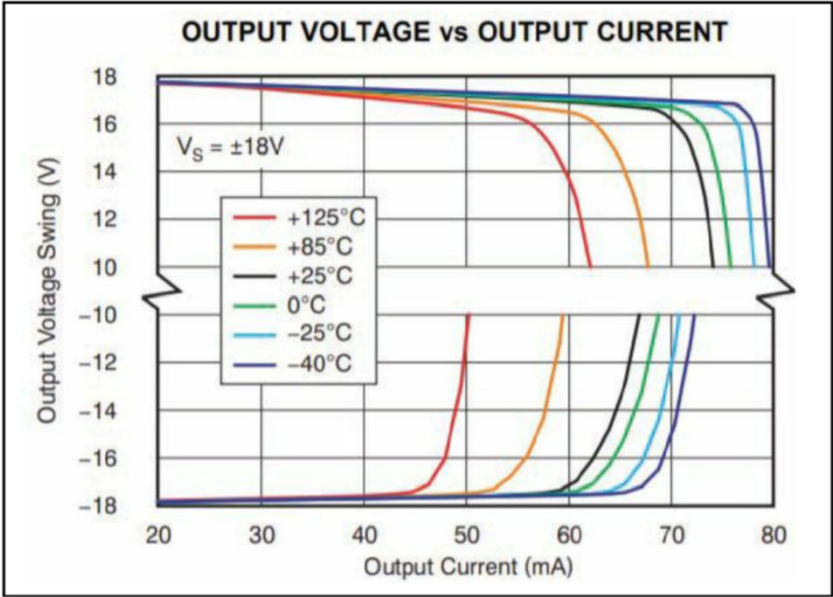


Figure 1: Maximum output voltage vs. output current (Reference 2, Figure 29)

current side. Putting those together, suggests a minimum load of 360Ω for this full unclipped output voltage swing on ±18V supplies. But this is close to hard clipping—well into an almost useless operating region for low harmonic distortion.

Estimating a “Linear” Output Current Capability

For high open-loop gain op-amps and FDAs, the most fundamental determinant for output harmonic distortion is the open-loop linearity of the output stage (for the maximum voltage and output current desired) corrected by the loop gain at the fundamental frequency. For this to apply, it is imperative to keep the output voltage swing below the clipping limits shown by Figure 1—usually by at least 0.2 to 0.5V to each swing limit. It is also imperative to ask for a peak output current that

OUTPUT						
V _{OL}	Output voltage, low	T _A = 25°C, G = 2	0.05	0.1	V	A
		T _A = -40°C to +125°C, G = 5	0.05	0.1		B
V _{OH}	Output voltage, high	T _A = 25°C, G = 2	4.9	4.95	V	A
		T _A = -40°C to +125°C, G = 5	4.8	4.9		B
	Maximum current into a resistive load	T _A = 25°C, ±1.6 V into 27 Ω, V _{IO} < 2 mV	±58	±70	mA	A
	Linear current into a resistive load	T _A = 25°C, ±1.7 V into 37.4 Ω, A _{OL} > 80 dB	±45	±50	mA	A
	Linear current into a resistive load overtemperature	T _A = -40°C to +125°C, ±1.31 V into 37.4 Ω, A _{OL} > 80 dB	±35	±45	mA	C

Table 2: Output swing ranges for the OPA837 on +5V total supply

does not significantly degrade the open-loop gain for the op-amp or FDA.

A recent trend has been to establish a safe output voltage and current condition into a resistive load and re-test the device A_{OL} for a minimum allowed value to validate a minimum linear output current in final test. This is still somewhat subjective, but better than nothing. A recent effort at this approach shows up in the OPA837 (Reference 1) as shown in **Table 2**.

Some key insights and caveats to the information shown in Table 2 include:

- The light load output voltage swing test is with a $2k\Omega$ load and a $2k\Omega$ feedback R_F at gain of +2. The total load is then $2k\Omega || 4k\Omega = 1.33k\Omega$ where the room temperature headrooms are 0.1V to each supply for this rail-to-rail output op-amp. The overtemp specification is at a gain of +2 (not +5, typo in the original).
- These output current tests are performed with $\pm 2.5V$ supplies
- If the far right column shows an "A", this is 100% tested at an approximate 25°C junction temperature.
- The maximum continuous current line is intended for applications less concerned about the lowest harmonic distortion.

This A_{OL} test is easily implemented at any combination of output voltage and current using the classic ATE op-amp servo loop (Reference 3). Typically, the A_{OL} test is performed with a small voltage swing centered on the specified average. For instance, a $\pm 100mV$ swing would be performed

where an 80dB minimum A_{OL} limit implies no more than a resulting $\pm 10\mu V$ input offset (or error) voltage swing—a trivial measurement with modern test loops. These specifications would suggest a maximum usable output current of $\pm 45mA$ if the application is over a narrow temperature range reduced to $\pm 35mA$ over the full temperature range. This is actually very good power gain for this $600\mu A$ (trimmed) supply current device. This is more useful output current information than the implied nonlinearity points in the "claw" curve of Figure 1.

A similar set of specifications appears in the more recent precision FDAs as shown in **Table 3** from the THS4551 datasheet (Reference 4). The voltage swings are with a $1k\Omega$ differential load showing about 0.23V worst case headroom requirement to each power supply.

The linear output current test is producing a $\pm 2.1V$ differential swing into a 50Ω differential load to generate a $\pm 42mA$ output current. For an FDA, this output voltage would be a 2.5V common mode voltage with $\pm 1.05V$ around that on each output pin.

Conclusions

If you set out to create an industrywide parametric matching and sorting tool for op-amps and FDAs, how would you handle assigning some kind of "linear" output current number to each device? First, most of the parameters that would go into such a database are pretty definitive. Usually, we would use the typical supply voltage table at room temperature for everything—then things such as maximum input offset voltage are usually trivial to locate and mean the same thing across all parts.

PARAMETER	TEST CONDITIONS	MIN	TYP	MAX	UNIT	TEST LEVEL ⁽¹⁾
OUTPUT						
Output voltage, low	$T_A = 25^\circ C$	$(V_{S-}) + 0.2$	$(V_{S-}) + 0.23$		V	A
	$T_A = -40^\circ C$ to $+125^\circ C$	$(V_{S-}) + 0.2$	$(V_{S-}) + 0.22$			B
Output voltage, high	$T_A = 25^\circ C$	$(V_{S+}) - 0.23$	$(V_{S+}) - 0.2$		V	A
	$T_A = -40^\circ C$ to $+125^\circ C$	$(V_{S+}) - 0.22$	$(V_{S+}) - 0.2$			B
Continuous output current	$T_A = 25^\circ C$, $\pm 2.5 V$, $R_L = 40 \Omega$, V_{OCM} offset $< \pm 20 mV$	± 60	± 65		mA	A
	$T_A = -40^\circ C$ to $+125^\circ C$, $\pm 2.1 V$, $R_L = 40 \Omega$, V_{OCM} offset $< \pm 20 mV$	± 50				B
Linear output current	$T_A = 25^\circ C$, $\pm 2.1 V$, $R_L = 50 \Omega$, $A_{OL} > 80 dB$	± 40	± 45		mA	A
	$T_A = -40^\circ C$ to $+125^\circ C$, $\pm 1.6 V$, $R_L = 50 \Omega$, $A_{OL} > 80 dB$	± 30				B

Table 3: Output swing specifications on single +5V supply for the 150Mhz, precision, THS4551 FDA

There is probably no specification more uncertain across suppliers and decades than “linear” output current. To arrive at some semblance of order, the following was used in a database I had built some years back:

- If there is nothing more than a short circuit current, use half of that number for linear output current.
- If there is something like a claw curve (but nothing more), use the minimum source or sink current that causes an increase of 0.5V from the no load headrooms.
- If there is a minimum A_{OL} test, use that 25°C minimum number.

The best specifications (Reference 1) will have both a claw curve and minimum A_{OL} test line. The claw curve can be used to guardband a required voltage headroom to each supply, while the minimum A_{OL} test conditions can give you the maximum source or sink output current before hard nonlinearity might result. 

About the Author

Michael Steffes has worked since 1982 in high-speed signal path IC development spanning design roles for Comlinear Corp. (the original current feedback originator) with product definition and launch roles through National Semiconductor, BurrBrown, Texas Instruments, and Intersil. He continues to focus on High-Performance ADC interface solutions, DSL/PLC line driver design, and signal path optimization tools. Providing new product data sheet and applications collateral for more than 150 industry leading amplifier products has led to hundreds of contributed articles and application notes since 1988.



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[2] “Soundplus High Performance, Bipolar Input Audio Operational Amplifiers,” Texas Instruments OPA1602 Datasheet, revised November 2011, www.ti.com/lit/ds/symlink/OPA1602.pdf

[3] D. Baum and D. Hiser, “The Basics of Testing Op Amps, Part 1: Circuits test key op-amp parameters,” EDN, November 16, 2011, Texas Instruments, www.edn.com/the-basics-of-testing-op-amps-part-1-circuits-test-key-op-amp-parameters

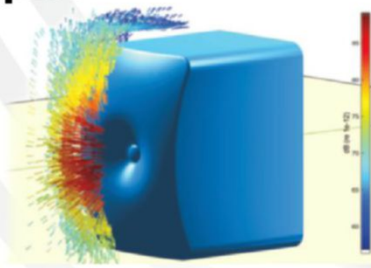
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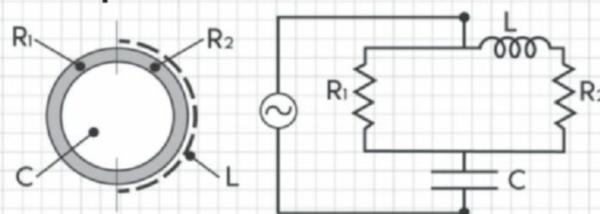


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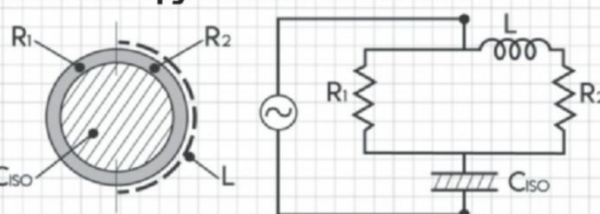
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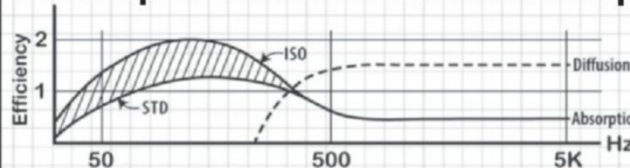
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Measuring Vibration with Piezoelectric Sensors (Part 2)

Design and Construction

By
George Ntanavaras



In this article series, George Ntanavaras describes his research on measuring vibration using Piezoelectric Sensors. In the second part, he discusses the design of such a system, configurations of the charge amplifiers circuits that he built and tested, and the connections to the piezoelectric sensors. He also presents some additional sensors that could be used for vibration measurements.

Configurations of the Charge Amplifiers

Trying to eliminate the mains interference, I decided to introduce the benefits of the differential amplifier. This type of amplifier is suitable for use in measurement and test equipment since it amplifies the voltage difference between its two inputs by some constant factor and rejects the part of the signal that are common to both inputs.

Figure 1 shows the electronic diagram for four circuits that I tested. All of them, are based on the LMC6482 CMOS operational amplifier (op-amp). It has an ultra-low input current of only 0.02pA and a very high input impedance of 10TΩ. **Photo 1** shows the construction of these circuits on prototype PCBs.

Circuit 3 is a charge amplifier based on only one op-amp U1 with differential inputs. Resistors R5 and R6 provide ESD protection. The capacitors C1, C13 and C2, C12 adjust the gain of the charge amplifier and resistors R1 and R3 provide the bias. The second op-amp U2 provides some additional gain.

Circuit 4 uses two single-ended charge amplifiers around U1 and U2, one for each side of the signal

while a third op-amp U3 transforms the balanced signal to a single-ended signal.

Circuit 5 is the classical differential voltage amplifier using three op-amps. And Circuit 6 uses two similar voltage buffers U1 and U2 for the positive and negative side of the input signal and a low-cost instrumentation amplifier U4 to transform the balanced signal to a single ended signal.

The circuits were supplied by an external 24VDC switched mode power supply followed by a capacitance multiplier circuit to produce a very low noise supply voltage. I thought that the circuits would show different behavior, but all of them measured well and their output noise was almost identical while the mains interference was reduced. I decided to use the Circuit 3, mostly because of its simplicity.

Sensor Arrangements

I also tested different sensor arrangements trying to increase the output level and possibly find a layout that will reduce the mains interference.

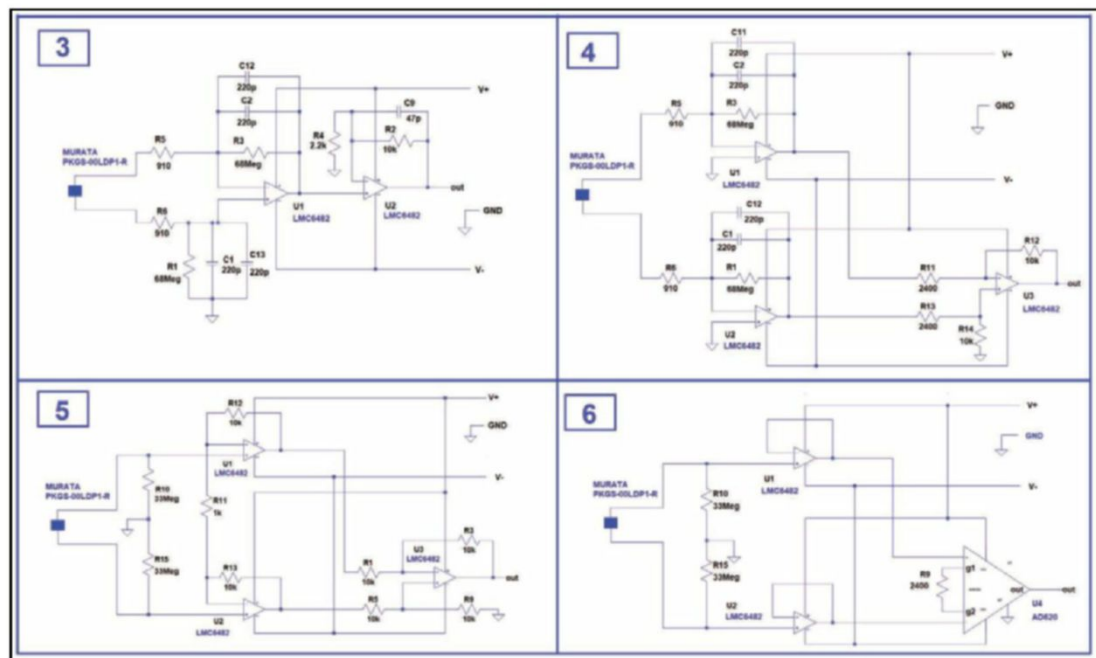


Figure 1: I tested four circuits and here are the schematics for each of them.

Photo 2 shows three of them. The first consists of two parallel sensors connected to the amplifier with a twisted pair to improve the rejection of the external electromagnetic interference.

The second arrangement is similar to the first one but with four parallel sensors to further improve the sensitivity. The third consists of two sensors connected in series with their common point connected through the shielding of the twisted pair cable to the ground of the amplifier circuit. The other two points are connected to the amplifier's positive and negative inputs.

As expected, the use of the shielded cable considerably improved the immunity to the external electromagnetic interference. I had even better results when I completely shielded the sensor with a metal can. I will give more detailed information about this in the next paragraphs.

The Sensor Connecting Cable

In every application, where the sensor is not next to its amplifier circuit, the cable is a major item effecting the system's noise performance. Especially for the piezoelectric acceleration sensor, the cable is much more critical and it is very important to select the proper cable to carry the high impedance signal.

"Tribo-electric" noise originates from capacity and charge changes due to dynamic bending, compression, and tension of the layers making up the cable. Common cables, when flexed or bent, generate an electrostatic charge output that is due to the layers making up the cable. This charge may have greater amplitude than the signal being generated by the accelerometer and the charge amplifier cannot separate whether the charge-generated signal is from the accelerometer or the

cable, so any disturbance of this kind should be eliminated by choosing and using the proper cable.

It is also equally important to make the cable as vibration-free as possible since the cable movements generate noise that interferes with the original signal. I use masking tape to firmly attach the cable because it allows easy removal without leaving residue or damaging to the loudspeaker cone. A good example of my technique is shown in **Photo 3**.

I have to admit that I spent more time looking for and testing the proper cable than I spent on the sensor itself. A lot of problems that I had during the first experiments with the sensors were due to the connecting cable and it took me some time

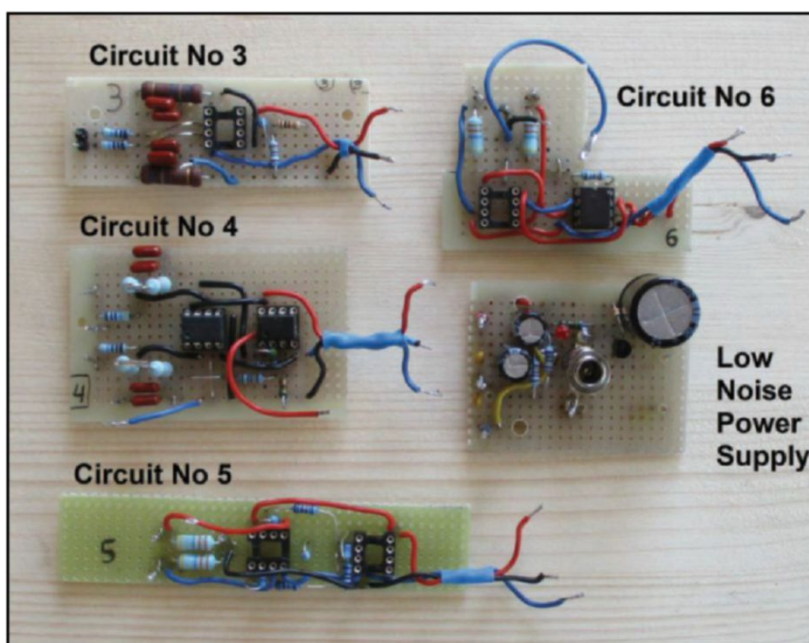


Photo 1: This is the construction of these circuits on prototype PCBs.

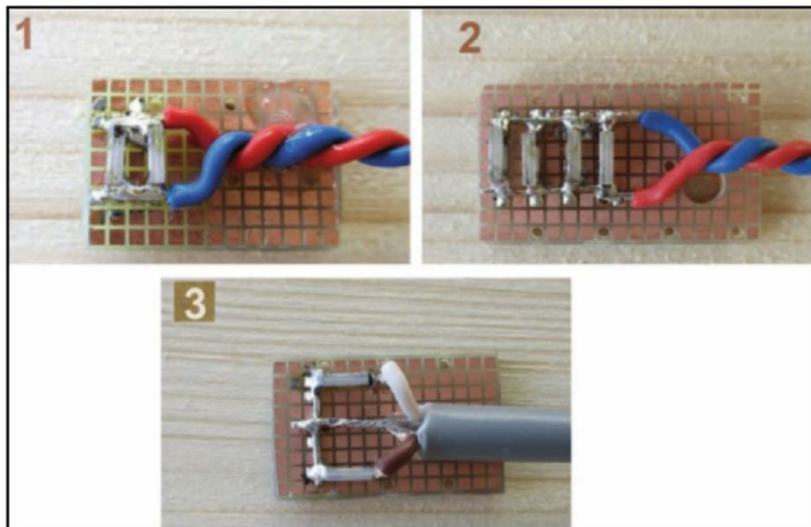


Photo 2: Here are three of different sensor arrangements that I tested while trying to increase the output level and possibly to find a layout that will reduce the mains interference.



Photo 3: Masking tape is used to firmly attach the cable.

to understand and figure out how to solve these problems. For example, in the measurement shown in **Figure 2**, the difference in response between the blue and the red curve is caused because a small part of the connecting cable was not supported well and it was free to move around.

When I noticed the cable used by the ACH-01 was produced by Mogami, I found the company's catalog and start searching for the proper cable. The problem was that the ultra-flexible and soft miniature cables that probably were suitable for the sensor could only be ordered in very large quantity, usually a 1000' reel.

After some research, I found a vendor from

whom I could order a few types of Mogami cables in small quantities. I purchased and tested several of them and I ended up with the Mogami W2901 cable. This miniature, soft, and flexible cable that features mechanical strength is for use mainly in lavalier microphones. It has a outer diameter of only 2.16mm and a balanced configuration using two 0.054 mm² (#30 AWG) conductors and a double shield. I used a short length of about 60cm of this cable to connect the sensor to the amplifier.

The PZ-1 Sensor

As I mentioned, my initial tests were performed with the Murata PKGS-00LDP1-R shock sensors. I was just starting to design a small PCB for the sensor when Murata announced that it was going to discontinue the PKGS-00LDP1-R.

Fortunately, I did find that there was a possible replacement, the Murata PKGS-00GXP1-R (Mouser, #81-PKGS-00GXP1-R-R). The Murata PKGS-00GXP1-R is also a surface-mounted component with smaller dimensions (3.8 mm × 2mm × 1mm).

Its sensitivity is lower at 0.35 pC/g (±15%) and the capacitance is only 390pF (±30%). The resonance frequency is higher at 31kHz (±16%) but it is very strong in amplitude, almost +40dB and with a very high Q. This resonance will be compensated by the amplifier's electronic circuit. **Photo 4** shows a comparison between the two sensors.

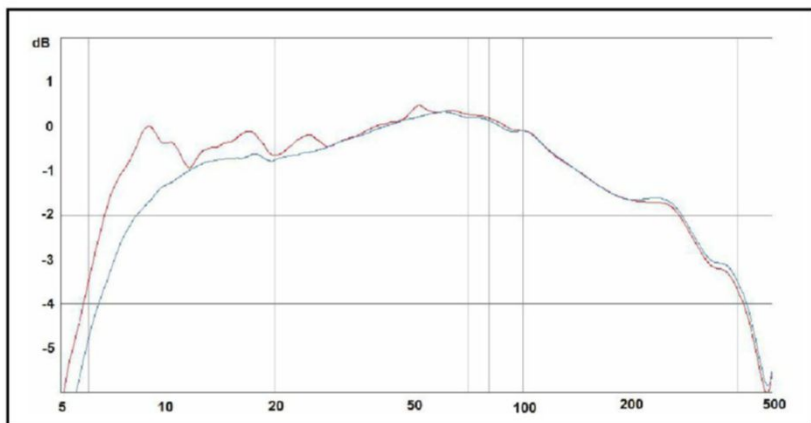


Figure 2: The difference in response between the blue and the red curve is caused because a small part of the connecting cable was not well supported and it was free to move around.

I designed a small double-sided PCB for easy mounting of the Murata sensors and the cable. The PCB has a ground plane on its bottom side and the Murata sensors are placed on its top side. Small pads are included for the connection of the connecting cable as shown in the right side of **Photo 5**. The two sensors are connected in parallel. The ground plane of the sensor PCB is connected through the shielding of the cable to the ground of the amplifier. The sensor has a polarity marking indication on its body and each sensor should be placed on the PCB with the same polarity, otherwise cancellation of their outputs will occur because they are connected in parallel.

Photo 5 shows how the assembly of the sensor was done. On the left side (a), the cable was prepared for soldering on the PCB of the sensor. In the middle (b) are the pins and the connector that was used for the connection of the cable to the amplifier. The right side (c) shows the small PCB of the sensor with the Murata sensors soldered on its position. The left side of **Photo 6** shows the completed sensor, which I named PZ-1.

The PZ-1-sh Sensor

The PZ-1 sensor with the Mogami cable gave very

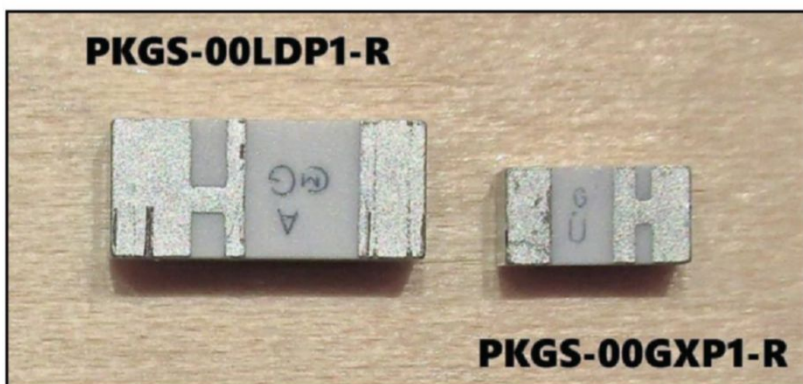


Photo 4: I compared the Murata PKGS-00LDP1-R shock sensors, which has discontinued and its potential replacement, the Murata PKGS-00GXP1-R.

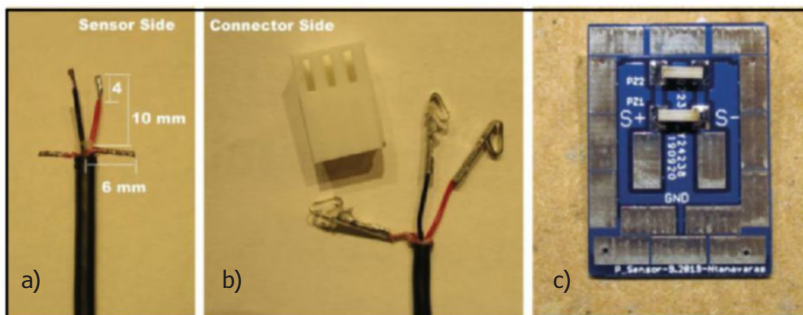


Photo 5: To assemble the sensor, I prepared the cable for soldering on the PCB of the sensor (a); in the middle are the pins and the connector that was used for the connection of the cable to the amplifier (b); and the right side shows the small PCB of the sensor with the Murata sensors soldered on its position (c).

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Photo 6: The PCB of the PZ-1 sensor was designed so that the metal can could be soldered directly to its ground plane. Here, the sensor is shown with the metal can before it was soldered to the PCB.

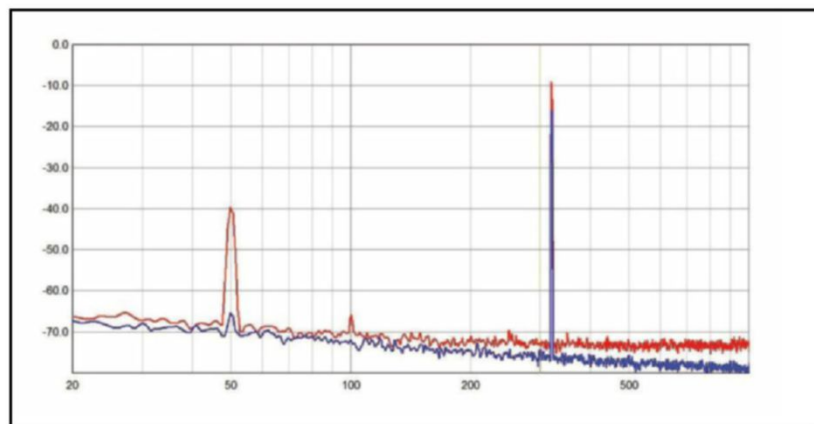
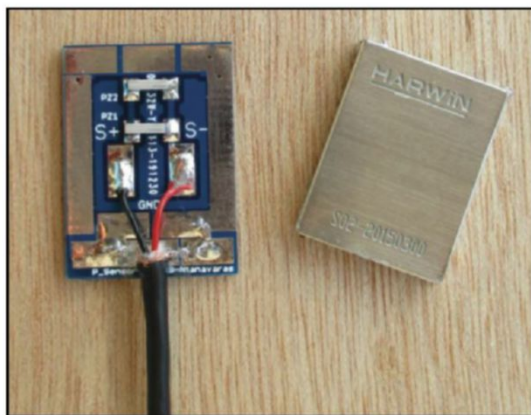


Figure 3: This graph shows a comparison of the frequency spectrum at the output of the piezoelectric amplifier between the PZ-1 and the PZ-1-sh sensors.

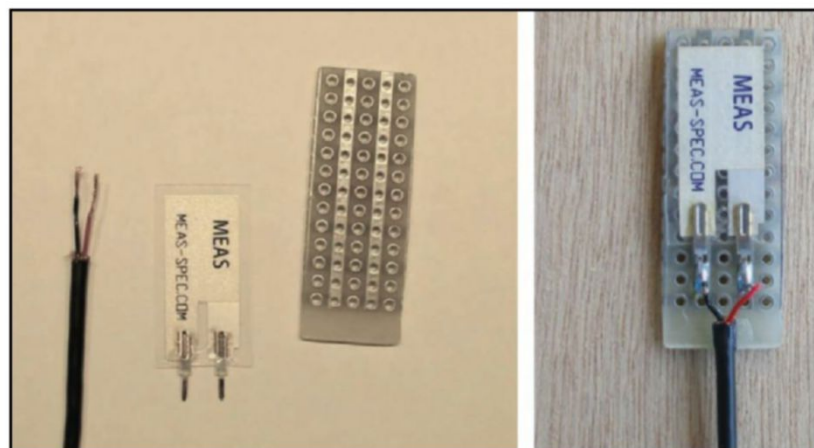


Photo 7: A small PCB was cut and permanently glued to the sensor using Loctite 401 instant adhesive (PZ-2-Parallax sensor).

good results concerning the suppression of the mains interference. I found out that the results were further improved when I used a small electromagnetic compatibility (EMC) shield can to completely shield the sensor. This is a EMC shield nickel can with dimensions of 20mm × 15mm × 3mm and thickness at 0.2mm (Harwin S02-20150300).

I designed the PCB of the PZ-1 sensor so that the metal can could be soldered directly to its ground plane and I named this version of the sensor PZ-1-sh. Photo 6 shows the sensor with the metal can before it was soldered to the PCB.

The metal can further eliminate the mains interference and also reduced the total output noise of the amplifier. **Figure 3** shows a comparison of the frequency spectrum at the output of the piezoelectric amplifier between the PZ-1 and the PZ-1-sh sensors. Both of them were attached to the Peerless cone, which reproduced a 320Hz sine signal. With the shielded sensor, the 50Hz mains interference is 25dB lower in level while the 100Hz is totally eliminated. The total noise floor is reduced by 2dB to 3dB.

The PZ-2-Parallax Sensor

After I had very good results with the PZ-1 and PZ-1-sh sensors, I decided to look for some other ready-made sensors that could be used. The Parallax Vibration tab sensor #605-00004 (Mouser 619-605-00004) consists of a flexible piezoelectric film laminated to a polymer substrate. When the sensor is bent or displaced from its neutral axis, the strain within the piezoelectric film generates voltage. According to its datasheet, the device has a high sensitivity of 50 mV/g and can generate very large voltages up to 70V if bent to 90°.

I cut a small PCB with dimensions of 38mm × 14mm as shown on the left side of **Photo 7**, where I permanently glued the sensor using Loctite 401 instant adhesive. In this way, the sensor cannot bend and produce high voltages.

I used a Mogami W2901 cable to connect the PCB to the amplifier. The two pins of the sensor were connected to the positive and the negative input of the amplifier. The shield of the cable was connected only to the ground of the amplifier. I also glued the cable to the sensor PCB using Loctite 401. I named this sensor assembly PZ-2-Parallax.

The amplifier that I will present in the third part of the article series was specially designed for the PZ-1 and PZ-2-sh sensors. I conducted all the tests for this sensor using this amplifier without any problems, but I used low vibration levels. Probably at higher levels, some protection will be needed and the following modifications will be sufficient, although I have not applied them.

About the Author

George Ntanavaras graduated from the National Technical University, Athens, Greece, in 1986 with a degree in Electronic Engineering. He currently works as a Technical Consultant for Greek electronics companies. He is interested in the design of preamplifiers, active crossovers, power amplifiers, and most loudspeakers. He also enjoys listening to classical music.



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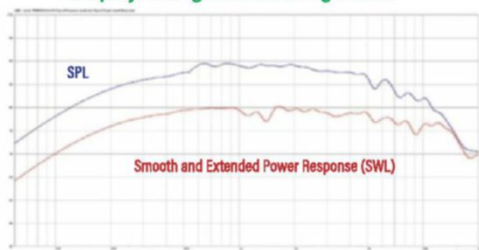
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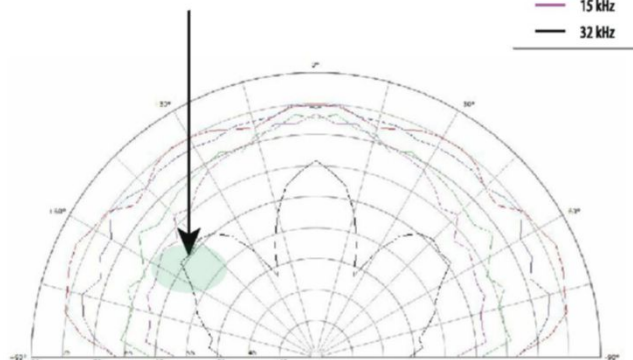


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The datasheet of the Parallax sensor recommends that a series 220 Ω resistor with a parallel clamping Zener diode of 5.1V should be connected at the output of sensor to protect the next circuit. I think the amplifier will be also effectively protected if its input resistors R1 and R20 are increased to 1k Ω and two parallel connected ultra low leakage diodes with reverse polarities are placed between the (+) and (-) pins of the IC1A. Also the gain of IC1B should be reduced to 0dB by omitting resistor R30.

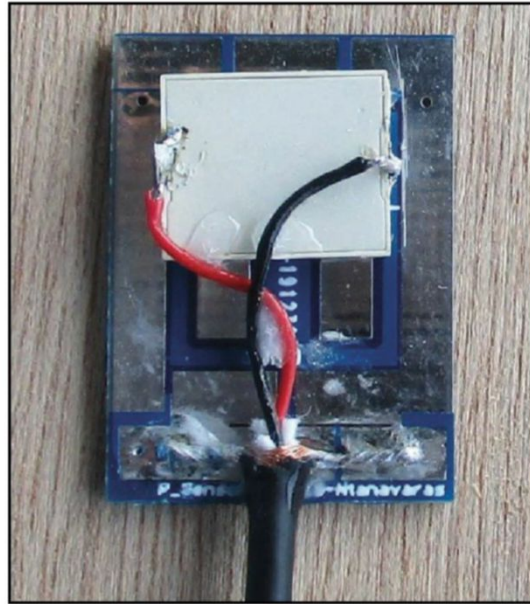


Photo 8: A Mogami W2901 cable was used to connect the PCB to the amplifier (PZ-3-PUI sensor).

The PZ-3-PUI Sensor

A piezoelectric speaker is a loudspeaker that uses the piezoelectric effect for generating sound. The motion that produces the sound is created by applying a voltage to a piezoelectric material. Typically they operate in the range of 1kHz to 5kHz, and they are used to generate sounds in electronic devices or as tweeters in computer speakers and portable radios.

PUI Audio SMT-1141-T-3-R (Mouser, #665-SMT-1141-T-3-R) is a surface-mounted piezoelectric loudspeaker with very small dimensions (11mm $\times$ 9mm $\times$ 1.7mm). It has a resonance frequency at 4.1kHz and a very high capacitance of 12000pF ($\pm$ 30%). It weighs only 0.3 grams.

I permanently glued it with Loctite 401 instant adhesive on the PCB that I designed for the PZ-1 sensor and I used a Mogami W2901 cable to connect the PCB to the amplifier as shown in **Photo 8**. The cable was also glued to the PCB.

The measurements proved that this type of sensor was a lot more sensitive than the others, having an output almost 35dB higher than the ACH accelerometer. Taking into consideration that the capacitance of this sensor is more than 30 times bigger, this is something that should have been expected. I have to test it with very low vibration levels to avoid overloading the amplifier.

Unfortunately, just before I finished writing the article, I found out that this transducer was obsolete. The SMT-1141-T-5-R (Mouser #665-SMT-1141-T-5-R) has exactly the same characteristics and although I have not tested it, I think it will probably give similar results. The modifications for the protection of the amplifier as described in the paragraph about the PZ-2-Parallax sensor are also valid here. Additionally, the gain of IC1A should be reduced.

Other Transducers

All four sensors that were described in this article are shown in **Photo 9** for comparison. Of course, there are a lot of other piezoelectric sensors on the market that I have not tested it and they would probably also give good measurements results.

Some other sensors that I ordered for testing that gave good measurements results are shown in **Photo 10** (from left to right): Measurement Specialties SDT-028K (Mouser, #824-1-1000288-0), Measurement Specialties FDT1-028K (Mouser, #824-1-1002785-1), and Measurement Specialties LDT1-028K sensors (Mouser, #824-1-1002910-0).

The SDT1-028K is an interesting sensor. It is fully shielded for use in applications in high electromagnetic interference (EMI) environments. It consists of a rectangular element of piezo film

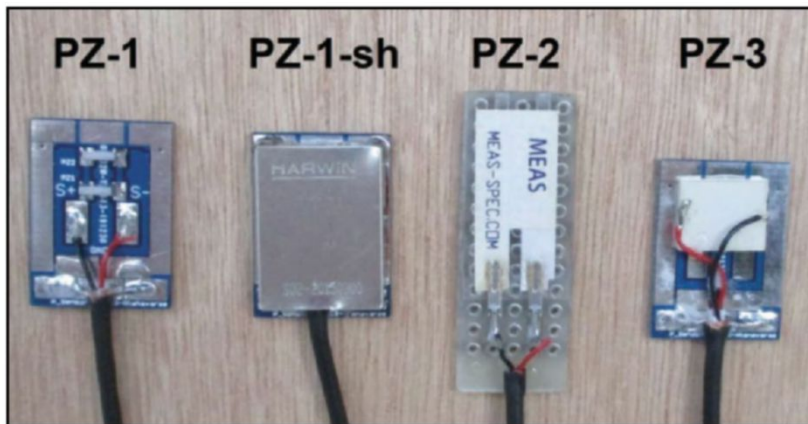


Photo 9: Here are all four sensors that were described in this article.

Sources

Mogami Cables | www.mogamicable.com/category/bulk

Murata Manufacturing Co., Ltd. | www.murata.com/en-sg/products/sensor/shock

Parallax, Inc. | www.parallax.com/product/605-00004

PUI Audio, Inc. | www.puiaudio.com/products/smt-1141-t-5-r

together with a molded plastic housing and 18" of coaxial cable. The film element, screen printed with silver ink, is folded over on itself, providing self-shielding of the sensor area. The dimensions of the sensor are: Length=28.6mm, Width=11.2mm, and Height=0.13mm.

The FDT1 Sensor is a rectangle element of Piezo film with with 28µm thickness and silver ink screen printed electrodes. Rather than making the lead attachment near the sensor, the piezo polymer tail extends from the active sensor area as flex circuit material. This gives a very flat, flexible lead with a connector at the end.

The LDT1 sensor has a piezo film element laminated to a sheet of polyester (Mylar). The dual wire lead attached to the sensor transfers the signal.

All these sensors are suitable for vibration measurements but depending on the surface, special attention will be needed to properly mount them.

Next Month

In the final part of this article series, I will describe the design and the construction of the piezoelectric amplifier and the measurements that will enable us to evaluate the whole system. 

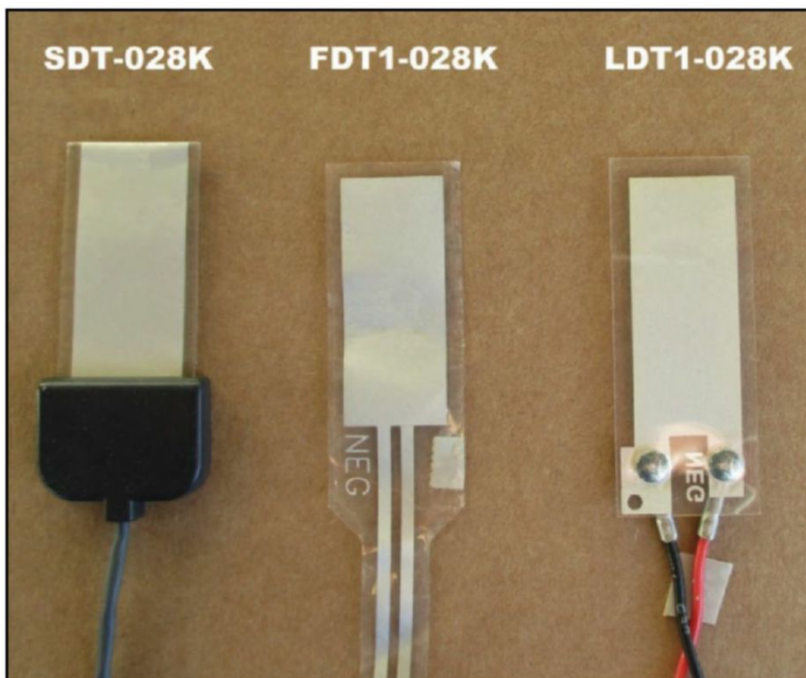


Photo 10: These are some other sensors that were used for testing and provided good measurements (from left to right): Measurement Specialties SDT-028K, Measurement Specialties FDT1-028K, and Measurement Specialties LDT1-028K sensors.

GETTING THE MOST OUT OF
VACUUM TUBES
By Robert B. Tomer

Vacuum Tubes—The Important 4/5ths of Tube Circuits
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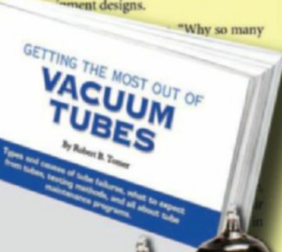
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Noise Is Everywhere—Even in MC Input Transformers (Part 1)

A Special Survey for the Skeptics

In this article, Burkhard Vogel discusses the signal-to-noise (SN) measurements of moving coil (MC) phono amps. He thinks it is good practice to measure a transformer with an input load and he wants to prove that an MC step-up transformer (SUT) produces noise because of its insignificant impedance.

By

Burkhard Vogel

Last summer I had an interesting discussion with the test center manager of one of Germany's top hi-fi magazines. The issue: signal-to-noise (SN) measurements of moving coil (MC) phono amps. The problem: He insisted on measurements with shorted input. That contrasts with my position to measure with an input load of 10Ω or 20Ω , which is the only way to include the input current noise of the phono amp into the SN measurement.

During these discussions he came up with the additional (and for me, unusual opinion) that an MC step-up transformer (SUT) does not produce noise because of its insignificant impedance. This man is not alone. During the past 40 years I have often read the same wrong opinion in test magazines.

Now, after reading Gary Galo's remarkable December 2020 *audioXpress* article [1] about the usage of Jensen Transformer's MC SUT JT-44K-DX, I thought it would be good practice of engineering to argue against that test center manager's view by taking a deeper look into the matter. I intend to prove that this kind of transformer always produces more or less significant amounts of—at least—thermal (Johnson) noise, depending on their specs or measured data, their input load R_0 , and the

input noise situation of the following moving magnet (MM) phono amp. By the way, microphone input transformers suffer from the same noise problem. Here, we will concentrate on thermal noise only.

There are many design and build-up efforts that have led to the RIAA Phono Amp Engine I and II in my books (Part V [2], Part II [3]), and in *Linear Audio* [4], including the application of several types of SUTs, produced by the US company Jensen Transformers (JT) and the Swedish company Lundahl (LL):

- JT-44K-DX with $n = 10$ see datasheet specs
- JT-346-4-AXT with $n = 4$ see datasheet specs
- JT-346-12-AXT with $n = 12$ see datasheet specs
- LL9226 with $n = 11$ measured via "C" configuration

Additionally, I have checked the SUTs from Sowter (UK), Haufe (Germany), and Pikatron (Germany). However, they did not fall into my scope of quality. It also became obvious that JT's datasheets get first place when it comes to precision

and extensive content. Unfortunately, as with all other manufacturers, it does not offer Spice models. However, its products offer outstanding quality.

Well, all my already-mentioned JT transformers are “hidden” in living amplification chains. The only “free” ones were a second pair of LL9226s. Therefore, I decided to work through the entire matter on two different tracks:

- Track 1 is the theoretical one that checks the JT-44k-DX with specs of its datasheet [5], together with a single-ended (un-balanced) following MM phono amp input, by comparing the simulated results with the calculated ones.
- Track 2 is the practical one that checks the LL9226s with measured data, (data that does not always exactly equal the datasheet specs [6]), together with a following balanced MM phono amp input, by comparing the simulated and calculated results with the measured ones.

A further test rounds out the entire exercise: What I call “following phono amp” is nothing but the simulation of an MM phono amp input by three different op-amp versions in single-ended (one op-amp) and balanced (two op-amps) mode with these op-amps:

- LT1128C with very low input voltage noise and rather high input current noise
- OP37G with acceptable (MM-usable) average input noise values (close to the values of the NE5534A)
- OPA627B with rather high input voltage noise and extremely low input current noise (close to the input noise generated by, for example, a b-configured E88CC tube at roughly 90V/3 mA [11] or [12])

By following the two tracks and because of cross-check reasons, the application of three very different supporting methods is very useful:

- Simulation with Linear Technology’s LTSpice [7]
- Calculation with a math software (Mathcad or MCD)
- Measurements with my CLIO-based measurement setup ([2] Chapter 24, [3] Part IV)

To proceed further, we must find solutions for the following two situations: A (opposing opinion) and B (my opinion).

A. A noiseless transformer is the one that has DC coil resistances according to its datasheet specs; however, it does not create thermal noise. I call it a cold SUT.

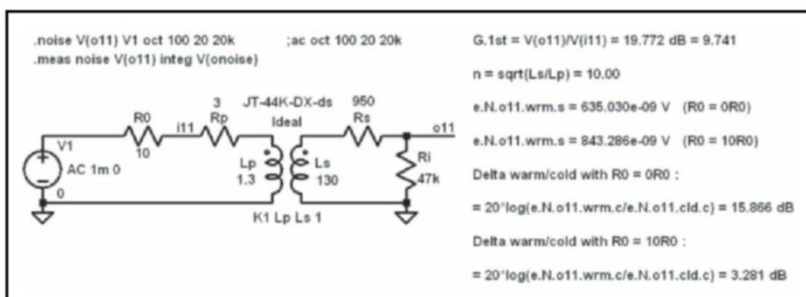


Figure 1: Simulation arrangement of the warm JT transformer

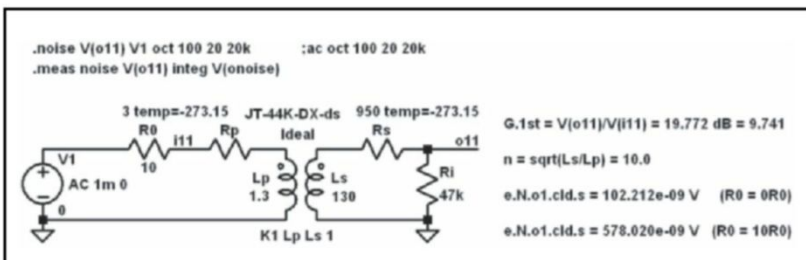


Figure 2: Simulation arrangement of the cold JT transformer

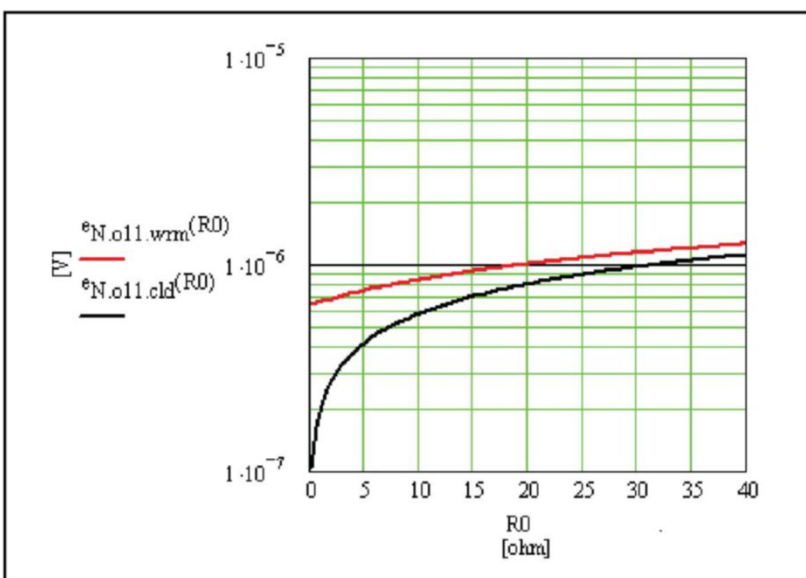


Figure 3: Output noise voltages of Figure 1 and Figure 2: red = warm = top trace, black = cold = bottom trace

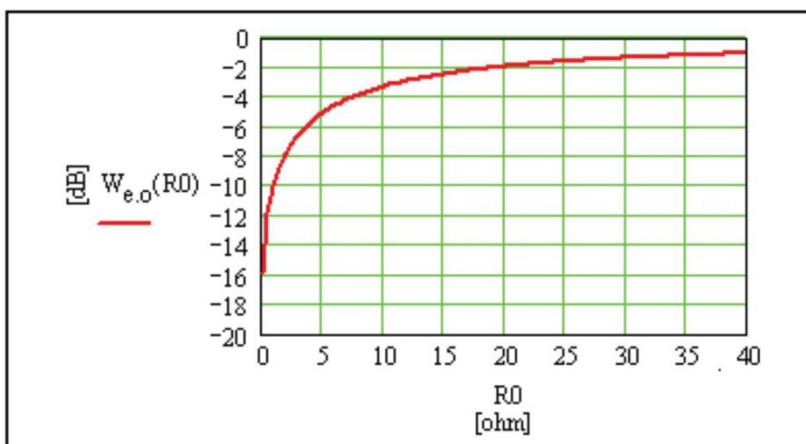


Figure 4: The deltas in Figure 3 expressed as worsening of the cold situation = $We.o$ in decibels (dB)

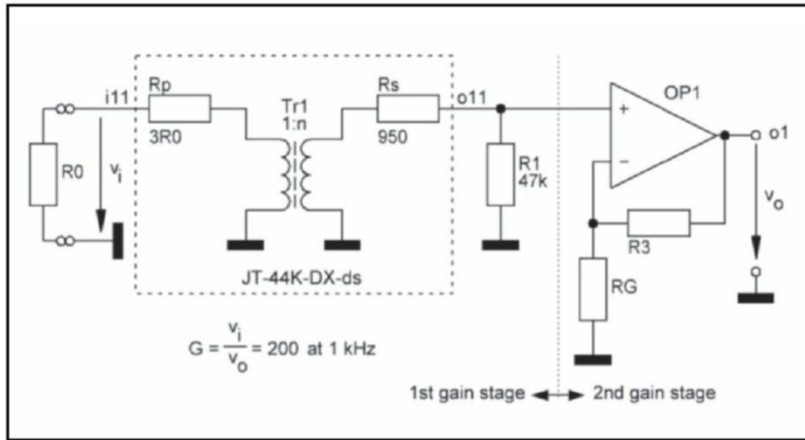


Figure 5: Single-ended test configuration with a JT transformer and op-amp based single-ended amp input

B. A noisy transformer is one that has DC coil resistances according to its datasheet specs; however, contrary to A, it does produce thermal (Johnson) noise [8], [9]. I call it a warm SUT. Note: “Warm” means: the transformer’s internal primary and secondary resistances R_p and R_s are kept at room temperature = plus 300.15°C (like all other LTSpice components); hence, the transformer’s resistances add thermal noise to the arrangement. “Cold” means: the transformer’s internal resistances R_p & R_s are theoretically kept at minus

273.15°C (easy to do in Spice and in math); hence, the transformer’s resistances do not add thermal noise.

I call the delta between the transformer’s cold and warm states the worsening W of the cold state because the warm results always look worse than the cold ones. In other words: The transformer adds noise to the input of the following MM phono amp. How and how much will be discussed in the following sections.

All simulations, calculations, and measurements were performed in the audio band B20k = 20Hz ... 20kHz. Two rather detailed MCD worksheets can be reviewed on *audioXpress*’ website under “Supplementary Material.” Finally, the main results will then be expressed in dB, dB(A), and dBV, and shown in three tables and six graphs.

Theoretical Approach with JT-44K-DX

JT-44K-DX Alone

First of all, with a simple simulation schematic I want to demonstrate that a SUT alone creates noise (which is also the case with step-down transformers, by the way). **Figure 1** shows it, based on the resistive and turns ratio figures of JT’s datasheet. I could not measure the real transformer parameters because of the above mentioned “hidden” situation. That is why I use the transformer simulation approach with an ideal transformer à la Giles Brocard’s *The LTSpice IV Simulator* [7, page 508 f].

Table 1: Simulated and calculated results with JT-44k-DX-ds and three different op-amps in single-ended (= un-balanced) mode as shown in Figure 5, $R_0 = 0\Omega$

1/A	B	C	D	E	F	G	H
2	Amp : type of op-amp	Calculated & simulated signal-to-noise ratios in B20k at the o/p of the un-balanced Amp - warm & cold JT-44K-DX-ds and excl. transformer			calc vs. sim = D - E [dB]	units	Remarks about action possibilities
3		configuration	sim	calc			
4	LT1128C sim with LT model	Amp & warm trafo	-97.09	-97.11	0.02	dBV	meas+sim+calc
5		Amp & cold trafo	-105.23	-105.35	0.13		sim or calc
6		Worsening = line 4-5	8.14	8.25	-0.11	dB	calc
7		Amp no trafo & i/p short to gnd	-108.87	-108.88	0.01	dBV	meas+sim+calc
8		CNI = line 7 -5	3.64	3.53	0.12	dB	calc
9	OP37G sim with buvo model a	Amp & warm trafo	-95.76	-95.76	0.00	dBV	meas+sim+calc
10		Amp & cold trafo	-100.01	-100.01	0.00		sim or calc
11		Worsening = line 9-10	4.25	4.25	0.00	dB	calc
12		Amp no trafo & i/p short to gnd	-100.31	-100.31	0.00	dBV	meas+sim+calc
13		CNI = line 12 -10	0.30	0.30	0.00	dB	calc
14	OPA627B sim with buvo model b	Amp & warm trafo	-94.41	-94.41	-0.00	dBV	meas+sim+calc
15		Amp & cold trafo	-97.06	-97.06	0.00		sim or calc
16		worsening = line 14-15	2.65	2.65	-0.00	dB	calc
17		Amp no trafo & i/p short to gnd	-97.15	-97.15	0.00	dBV	meas+sim+calc
18		CNI = line 17 -15	0.10	0.10	0.00	dB	calc
19	a = change of AD’s model with 3.0nV/rHz to the G version’s 3.2nV/rHz i/p voltage noise density						
20	b = change of TI’s model to the one with noise curves of the datasheet						

Figure 2 shows the same simulation arrangement; however, the coil resistances R_p and R_s are set to -273.15°C . Hence, the transformer becomes fully noiseless.

On the right side of Figure 1, we find the simulation results for the gain G_{1st} (= in line with the datasheet values 19.65dB ... 19.82dB), the RMS noise voltages at the output o11 for two different input loads: 0Ω and 10Ω , and the delta between the cold (Figure 2) and warm state. With $R_0 = 0\Omega$ we get a worsening of $W = 15.9\text{dB}$ rounded and with $R_0 = 10\Omega$ we get $W = 3.3\text{dB}$ rounded.

I have chosen $R_0 = 10\Omega$ because with this result readers may get an impression about the noise impact of the input load situation of the MC cartridge DL103R || 47R5 plus SUT in Galo's article [1].

Because there is no difference between the simulated and calculated results, we use Mathcad to generate the two graphs shown in **Figure 3** and **Figure 4**, which show the evolution of the output noise voltages (Figure 3) and the worsening figure $W_{e.o}$ (Figure 4), both versus the input load R_0 . All necessary equations are given on MCD Worksheet 1, points 0. ... 4.

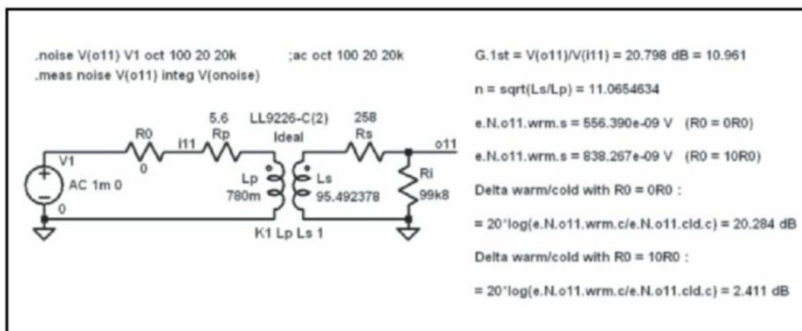


Figure 6: Simulation arrangement of the warm LL transformer

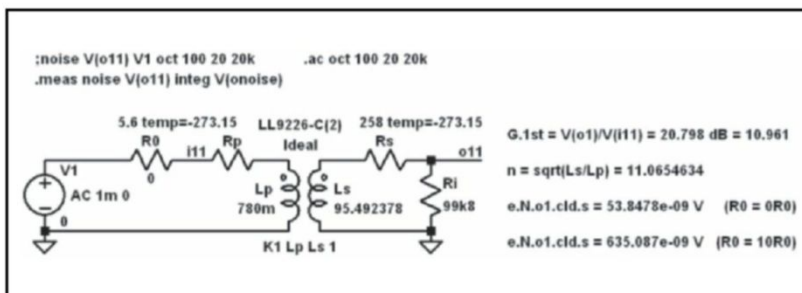


Figure 7: Simulation arrangement of the cold LL transformer

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JT-44K-DX Connected to the Input of a Single-Ended MM Phono Amp

Thus, without any measurement to prove the given findings, this theoretical exercise triggers the question: What would happen with worsening W if we connect the transformer to a real MM amplifier input? I will demonstrate this with a well-known

active and un-balanced amplifier configuration called single-ended.

Figure 5 shows the theoretical test configuration with JT-44K-DX-ds (ds = datasheet based). The corresponding math and simulation results on MCD-WS 1 were created with an OP37G op-amp. **Table 1** presents the overall non-equalized signal-to-noise (SN) results for the three op-amp types.

I hope Table 1 is self-explanatory. We can see that there is practically no difference between the calculated and simulated SN results. In lines 6 and 11 and 16 we find the worsening figure W, indicating that a rather high input voltage noise corresponds with a lower W and vice versa. I guess this is a logical result because the high input voltage noise of the op-amp has the tendency to mask the lower voltage noise at the transformer's output.

In the second part of this article, I will explain the steps to get the numbers in the tables and what they have to do with the current noise impact (CNI) of a transformer.

Practical Approach with LL9662

LL9662 Alone

The balanced configuration of Lundahl's top MC input transformer presented here is based on the recommended "C" input and output layout of its six different coil windings. The simulation schematics are given in **Figure 6** and **Figure 7**. They show the measured coil resistances and coil ratio n that could

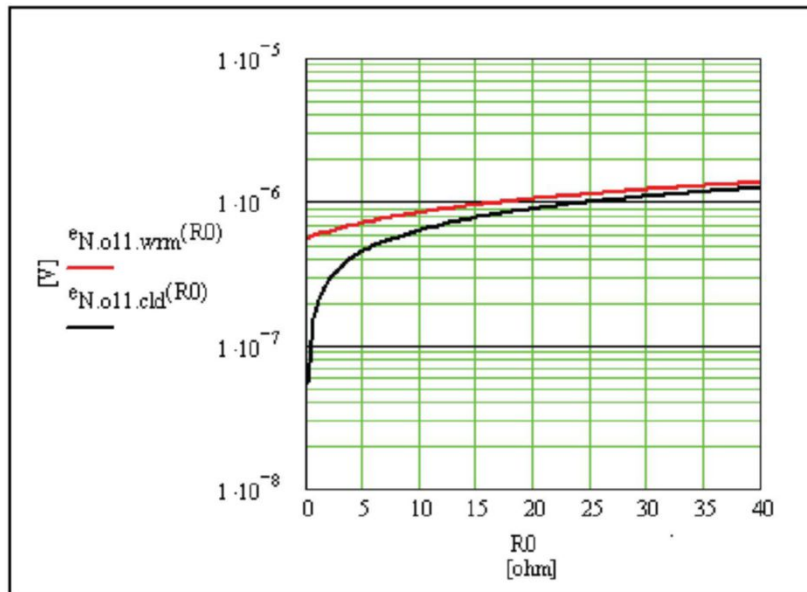


Figure 8: Output noise voltages of Figure 6 and Figure 7: red = warm = top trace, black = cold = bottom trace

Project Files

The Mathcad Worksheets found in the Supplementary Material section of the audioXpress website present the complete math around the two different step-up transformers (stand-alone under points 1. ... 4.) and their connection to two different and active amplification methods, un-balanced and balanced (under points 5. ... 12.):

MCD-Worksheet 1: JT's transformer JT-44K-DX configured alone with datasheet values of its DC resistances and turns ratio plus its connection to an OP37G driven un-balanced MM phono-amp input

MCD-Worksheet 2: LL's transformer LL9226 "C" configured alone with measured values of its DC resistances and gain plus its connection to an OP37G driven balanced MM phono-amp input

To download the Mathcad Worksheets, visit:

<http://audioxpress.com/page/audioXpress-Supplementary-Material.html>

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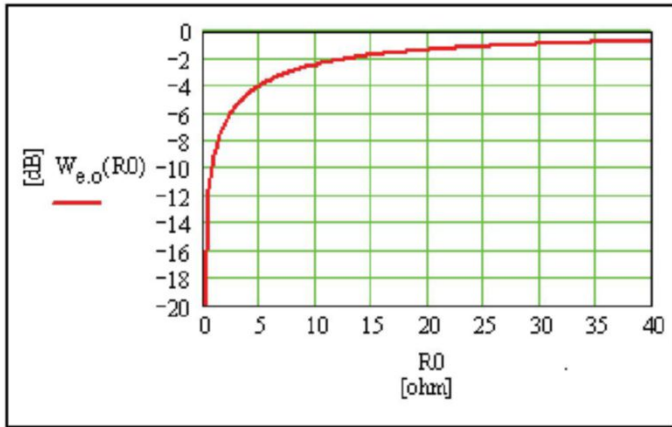


Figure 9: The deltas in Figure 8 expressed as worsening of the cold situation = $W_{e.o}$ in decibels (dB)

be calculated with the measured gain $G_{1st.m}$ of the arrangement. MCD-WS 2 (points 0. ... 4.) shows the way how we can calculate them and what could also be achieved by simulation. I found no difference between the two results.

On the right side of Figure 6 there are the simulation results for the gain G_{1st} , the RMS noise voltages at the output o11 for two different input loads: 0Ω and 10Ω , and the delta between the cold (Figure 7) and warm state. With $R_0 = 0\Omega$, we get a worsening of $W = 20.3\text{dB}$ rounded and with $R_0 = 10\Omega$ we get $W = 2.4\text{dB}$ rounded.

Because there is no difference between the simulated and calculated results we use Mathcad to generate the two graphs in **Figure 8** and **Figure 9**, which show the evolution of the output noise voltages (Figure 8) and the worsening figure $W_{e.o}$ (Figure 9), both versus the input load R_0 . All the necessary equations are given on MCD Worksheet 2, points 0. ... 4.

Next Month

Next month we will proceed with the practical approach on Lundahl's LL9226, followed by a step by step approach to noise measurements, simulations and calculations, with conclusions. 

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, LAs in EWW, LA, and D&D. Finally, edited by the Springer Nature publishing house he has six books (currently number 7 is "under construction") that round up the gained knowledge.





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The Science and Physics of High-End Mic-Preamp Design (Part 4)

Different Approaches

By
Mike Meechan



In previous articles, Mike Meechan considered the fundamental physics that play a rudimentary part in microphone preamplifier design, and explored its inhabitancy in the real world to retain the fidelity of the mic signal along its path from transducer to output. At this stage, he continues to look at what (theoretically) contributes to perfect active mic preamps.

For many years, there seemed to have been a widely accepted policy of low-level analog signal transfer. Cost, complexity, and the lack of bandwidth of digital transmission equipment, coupled with the lack of suitable converters during this period, obviously precluded the use of the digital domain. However, the reasons for the more widespread pursuit of other plausible analog alternatives in professional audio seemed unclear for many years. Common sense has since prevailed—in no-compromise recordings or sound-reinforcement situations, modern-day philosophy in this respect is now represented by dual solutions:

- Transmit the signal at a level sufficiently high that interference—of whatever its nature—has any effect minimized
- Place signals in a domain so that malign effects can be mitigated as much as possible

Note that while this is the case, much ingenuity and development work has been expended in the design and manufacture of cables intended specifically for connecting microphones. The Starquad arrangement is particularly highly regarded, where a balanced (quad) cable core with capacitance unbalanced characteristics is used as a twisted pair (**Figure 1**). This capacitive imbalance— theoretically— nulls interference of whatever sort (especially noise—since the conductors are twisted together

with a very short lay length). Additionally, insulation thickness of each conductor is closely controlled so that capacitance is matched as closely as possible. While this reduces crosstalk, long lengths can roll off high-frequency because the capacitance of the cable (in combination with the resistance of the cable) creates a high-frequency filter. However, this cable type's main advantage is in its magnetic immunity properties—important when signal cables are routed next to power cables radiating 50Hz or 60Hz.

One Option

One solution is to place the preamp onstage—at the immediate point of signal capture. Thus, the cable length which a low-level signal travels is optimally minimized. These remote amplification devices would be placed as near as practical to microphone take-offs—traditionally, either in a rack at the stage edge, perhaps residing inside stage-box- like constructions somewhere nearby, or sometimes even as dedicated units embedded in the stage flooring itself. Serial digital communications links, daisy-chained across the units, terminate at a master control unit residing in close proximity to the sound mixer. From this master control, functions such as individual channel gain (usually divided between coarse and fine), 48V phantom powering, phase reverse and even high-pass filter (HPF) settings can be realized, stored and recalled at will, often under some form of automated control. In this way, the audio signal applied to the

mixing desk channel input is high-level, high-fidelity and gain-controlled, being more akin to that received from a digital sound source (**Figure 2**).

Choosing a Transformer or Active Device First Input Stage

Rather interestingly, the remote-recording company Effanel Music (no longer in business) once had 52 channels of mic preamps located on stage. However, they were not remotely operated (at least not in the conventional sense). Instead, there was a so-called “Hardy operator”: the operator sat beside the racks of M-1 preamps and was always in contact with others by headphone so adjustments could be made quickly (and sort-of remotely—in an obtuse way).

Note that some engineers prefer to run individual XLR cables directly from the mic output to the mixer input, obviating the need for stage boxes. In reality, while this might be practical (and desirable) in a small venue, it is not practical when the front-of-house (FOH) desk is perhaps 100 meters or more from the stage.

In the Effanel Music situation previously noted, the racks of M-1 preamps were grounded on stage, while the remote recording truck was in the alley, or down the block, and often grounded to the nearest fire hydrant. So in this instance, there were very different “ground” potentials between the stage and that of the fire hydrant some long distance away. However, the floating output of the JT-11-BM output transformers completely isolated the audio signal from “ground,” so there couldn’t be a ground loop to cause hum or buzzing. In this respect, along with complete galvanic isolation, the transformer was without peer. Furthermore, using the Jensen JT-11-BM output transformer after the output of the 990 op-amp, (particularly in the Effanel situation), provided ground isolation and also helped to protect the 990 from full short-circuits. (All of the Effanel M-1 preamps were equipped with the JT-11-BM output transformers.)

Good pro audio industry practice persisted in the acclaimed Hardy mic preamps over the years since the first M-1 design debuted. In this vein, the key orthodoxy requires the preamplification of the mic signal as near to the mic itself (on the stage in a live environment), then transmitting it to the FOH mixing desk inputs at as high a level as possible.

The John Hardy Version of the Iconic 990 Hybrid Op-Amp

The first comes in its guise using the Hardy 990C+ as a line driver in this instance, a task to which it is eminently suited. The 990 has current

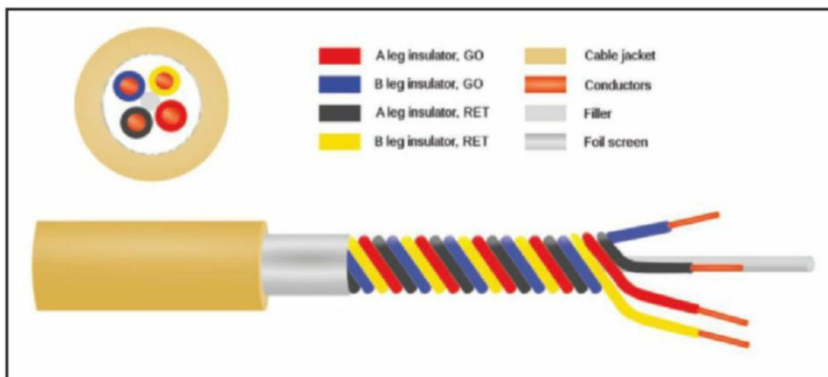


Figure 1: The physical properties and electrical characteristics of the Starquad cable

limiting circuitry in the output stage. The two output transistors, the MJE171 and the MJE181, are at the edge of the circuit board between the row of four pins (+V, 0V, -V and Output) and the wall of the anodized aluminum potting shell, with the metal surfaces of those transistors directly against the wall of the potting shell for maximum heatsinking; the epoxy potting compound fills in the tiny gap between the output transistors and the potting shell for thermal coupling. And this is where it gets rather clever.

The Coefficient of Thermal Expansion (CTE) of the epoxy potting compound is matched as closely as possible to the CTE of the components. There are variances between the CTE of various components, the PCB, the potting shell, etc., but they are all confined to a fairly narrow range. The closer the CTE of the epoxy is to the various parts, the less physical stress there will be between everything as the modules heat up and cool down, and the

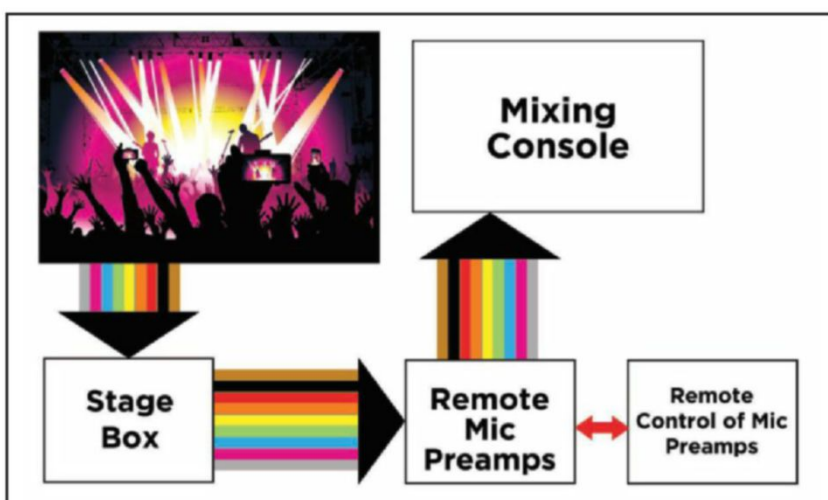


Figure 2: This is an example of a conventional wholly analog approach (remote amplification reduces noise, crosstalk, and CM susceptibility are reduced compared to straight mic-mixer cabling XLR-XLR connection). Problems with this approach are the expense (remote preamps are expensive), the high-level analog signals are still prone to interferences especially over long distances, and extra cable is necessary for the remote preamplifier control interface, plus additional cable is required to feed preamp outputs to mixer.

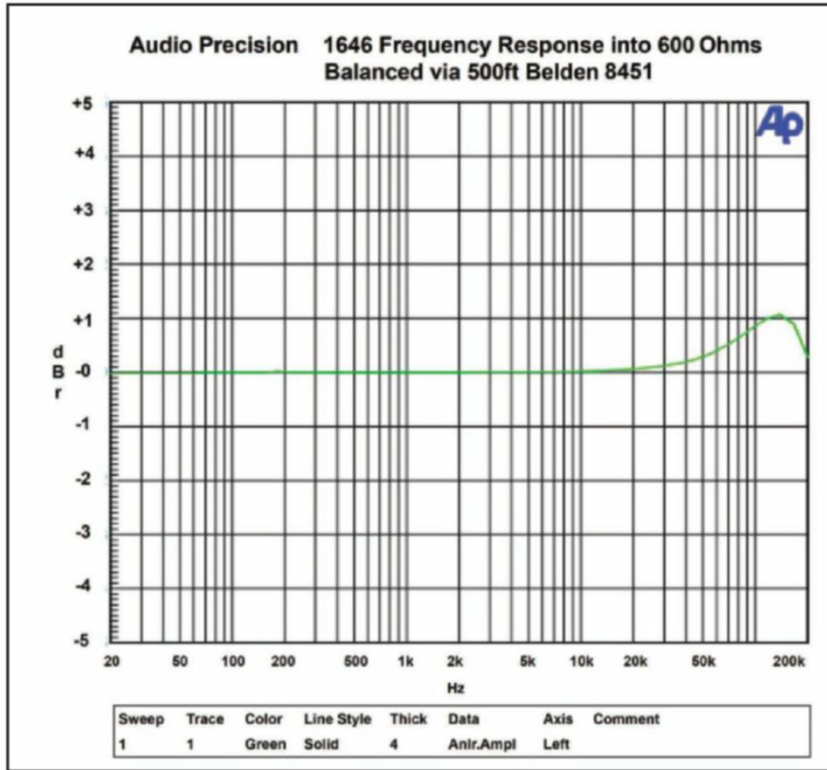


Figure 3: The 1646 frequency response when driving 600Ω at the end of 500' of Belden 8451 cable (Image: THAT Corp.)

thermal conductivity of the epoxy is relatively high to help distribute and dissipate heat. Note that there are certain components that should thermally track each other, so they are located next to each other on the circuit board and the epoxy provides thermal coupling. The high thermal conductivity of the aluminum potting shell helps to keep everything at approximately the same temperature while dissipating heat; the circuit board material has a high glass transition temperature to avoid issues of odd and excessive expansion beyond that temperature. Thus, it is pretty bomb-proof, even when an output cable has a dead short (with the proviso as noted earlier re added protection via Jensen JT-11-BM output transformers).

The Ubiquitous Chip

Another solution comes from the pro-audio IC experts of THAT Corp., in the guise of the THAT 1646 OutSmarts differential line driver. This employs monolithic differential line drivers, which markedly improve performance over conventional cross-coupled designs. The IC design is based on a high-performance, fully differential op-amp and laser-trimmed thin-film resistors, and so exhibits low noise and distortion, high slew rate rate, and provides wide output swing (13Vrms). Furthermore, the parts are stable when driving difficult loads, and

Parameter	Symbol	Condition	Minimum	Typical	Maximum
Maximum Capacitive Load	–	Stable operation	–	–	Unlimited
Quiescent Supply	I_S	–	–	4.9mA	5.75mA
Output Short Circuit	I_{SC}	Both outputs to 0V	–	–	70mA
Supply Rails	V_{SUPPLY}	–	±4.5V	±18V	±20V
Maximum Output Level	–	0.1%THD+N $R_L=600\Omega$	–	27.5dBu	28dBu
Output Noise	$NOISE_{OP}$	Balanced, 22Hz–20kHz	–	–101dBu	–
THD+N (differential)	–	$V_O=10 V_{RMS}$ $R_L=600\Omega$ $f=20\text{Hz}–5\text{kHz}$ $f=20\text{kHz}$	–	0.0007% 0.002%	0.009%
THD+N (single-ended)	–	$V_O=10 V_{RMS}$ $R_L=600\Omega$ $f=20\text{Hz}–5\text{kHz}$ $f=20\text{kHz}$	–	0.001% 0.006%	0.009%
Gain	G	$R_L=600\Omega$ Balanced; Single-ended	5.8dB 5.76dB	6dB 5.96dB	6.2dB 6.16dB
Gain Error	G_{ERROR}	$R_L=100\text{k}\Omega$ per output Balanced	–	0.02dB	0.2dB
Output Z	Z_{OUT}	–	40Ω	50Ω	60Ω
Output CMRR	$CMRR_{OP}$	F=1kHz, BBC method	46dB	65dB	–
Slew Rate	SR	$C_L=50\text{pF/output}$	–	15V/μS	–
Small Signal Bandwidth	–	$C_L=50\text{pF/output}$	–	10MHz	–

Table 1: Key electrical characteristics of THAT 1646 OutSmarts differential audio driver IC (abridged)

have short-circuit-protected outputs. The 1646 has a sister IC, the 1606, which has a useful differential input in place of the single-ended one of its family sibling. Both models incorporate THAT's patented OutSmarts technology, a dual-feedback-loop design that prevents the excessive ground currents typical of cross-coupled output stages (CCOS) when clipping into single-ended loads. The IC is more-than-capable of driving 500' long cables and is stable under any capacitive load (**Figure 3**).

Basically, the factor, which ultimately limits how much cable length can be driven, is degradation and the high-frequency roll-off that results from the $50\Omega Z_{\text{OUTPUT}}$ and the cable capacitance. We consider this IC solution in greater depth in the last part of the series.

Another Solution

This step forward meant that it was only a matter of time before the concept of early pre-amplification of the microphone signal was brought up-to-date, and as it did, migrated toward early analog conversion to digital (**Figure 4**). Accordingly, Kai Conrath introduced the very first digital microphone in his university thesis, back in February 1996—and the development results were taken to the patent office. DE 19606261 C2 was granted soon afterward: two prototypes of digital microphones were then presented at the 1996 AES Convention at Copenhagen. In his system, line-level outputs were converted (internally to the unit or via a cascaded outboard A/D converter) into an AES-EBU digital audio serial bitstream. High-quality amplification was effected at source, and the problems that a long cable presented both to a critical low-voltage signal—and to the electronics at the destination intended to discriminate this wanted signal from any interfering ones—were circumvented to a large extent. Polar pattern and much more could be controlled remotely.

The Neumann soon-to-be-defunct digital equivalent of the U83, the JLM-103-D, uses more-modern technology than Conrath's to achieve f_{SAMPLING} of 44.1kHz to 192kHz, 24-bit output using a 28-bit AD, and an AES-42 digital data stream 25 years later. Advances in microelectromechanical systems (MEMS) have made this approach much easier to implement. DACs such as the Burr-Brown PCM1798 Bit Perfect device, which has 32-bit resolution/768kHz f_{SAMPLING} , or Analog Digital Inc.'s (ADI) AD5791 BRUZ device (which uses multiladder technology), provide the necessary good compression-less conversion from one domain to the other. Both will support DXD1024, MADI, and AES-EBU outputs (**Figure 5**).

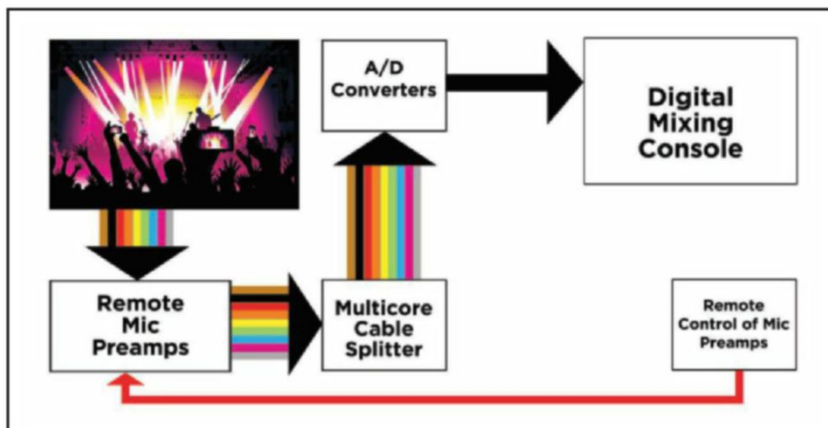


Figure 4: The remote mic signal amplification/digital conversion obviates many signal transfer difficulties. In the diagram, microphone signals are amplified by remote preamp on or very near the stage. Line levels are then combined via stage box into a multicore cable. These signals are much less susceptible to interference, crosstalk, etc, and the quality of high-quality preamps is maintained. With AES/EBU, bitstreams could be multiplexed into a single carrier and conveyed over fiber. Main problem with this approach is the higher cost.

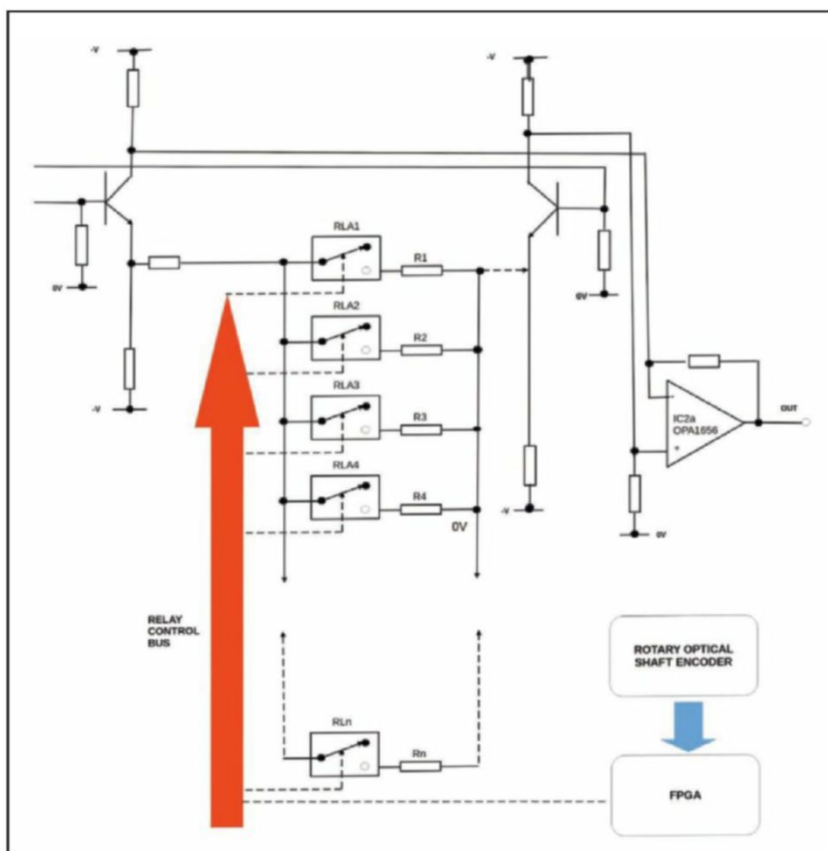


Figure 5: Relays with switched resistors enable remote control of gain.

Table 2: RG resistor gain values for the THAT 1510 and THAT 1512 ICs. Note that RG values are identical for both the INA217 and SSM2019, and these ICs are pin-for-pin-compatible (as is the INA163)

R_G	THAT 1510	THAT 1512
∞	0dB	-6dB
10k Ω	6dB	0dB
5k Ω	9.5dB	3.5dB
1k Ω	20.8dB	14.8dB
500R	26.4dB	20.4dB
100R	40.1dB	34.1dB
50R	46.1dB	40dB
10R	60dB	54dB
5R	66dB	60dB

For those of a DIY inclination, visit the website: www.beis.de/Elektronik/Electronics.html for a range of AD/DA modules to build into your own designs. As an aside, I've used the company's multi-configurable 51-segment VU/PPM LED level meters on the conceptual design.

A configuration very similar to the one shown in Figure 5 is used in the highly acclaimed consumer audio preamp/headphone amp from the acclaimed consumer ultra-fi US company, Benchmark Media. The 64 bank relay/switched precision metal-film resistive resistor network, which this controls (and

provides 256 steps in 0.5dB increments) derives its control from a FPGA rather than an MPU. This means that the gold-plated relays open-and-close in a carefully-timed make-before-break fashion, to avoid audible clicks. (Rotary gain control switches must also be make-before-break types, for the same reason). I have included R_G resistor gain values for the THAT 1510 and THAT 1512 ICs in **Table 2**. Note that R_G values are identical for both the INA217 and SSM2019, and these ICs are pin-for-pin-compatible.

Current Developments

Current development (in some respects) is focused on the wider-spread integration of preamplifier into the actual microphone body, such that the "pseudo" microphone output is a line-level balanced analog signal or—ideally—with the addition of a suitable A-D converter, an AES-EBU digital audio bitstream – conveyed by an appropriate transmission medium (wired or wireless). Obviously, several difficult-to-realize factors of the design could be circumvented, either if the signal pre-amplification were undertaken within the microphone structure itself, or in a "bolt-on" box that interfaced directly with the microphone body. Long transmission lines, isolation, crosstalk, and RFI, all become much less troublesome, and so Z_{LOAD} load impedance required by the capsule could be matched exactly, thus affording ideal frequency and THD+N performance.

The counter-argument here is that many engineers prefer the particular signature "sound" of a beloved preamp/EQ, be it tube or whatever, and so will stick dichotomously with the legacy analog way of doing things. Each to their own.

I suppose it's what makes music distinctive and... well... music.

Next Month

Next month, we look more closely at various—sonically excellent—IC solutions that there are out in the marketplace and currently available.

About the Author

Mike Meechan studied electrical/electronic engineering in Glasgow and has both BS and MS degrees. Until 1996, he worked as senior engineer at the BBC's World Service, Bush House, London, then at BBC Monitoring. A period as a freelance designer/author in the professional audio field followed, and he founded Audio Elixir, (an audio design house specializing in low-noise pro-audio analog solutions). From 1998-2002, he was a documentation manager for a large organization providing end-to-end digital broadcast solutions, to BSkyB, London, UPC EuroHITS, Amsterdam and Prague, and DirecTV, Los Angeles. For 10 years he was chief architect and led the Motorsport Knowledge Exchange project for Oxford Brookes University (School of Technology). Since 2011, he's been providing audio engineering consultancy to the industry. Over a period of 30 years, Meechan has authored 80 articles/papers on pro-audio/broadcast-related matters, and designed, built, and published numerous acclaimed pro-audio products. His research interests include low-noise audio design, particularly pre-amps, EQ, and dynamics processing. He previously held memberships to the Audio Engineering Society, the Institute of Engineers & Technologists, and PLASA.



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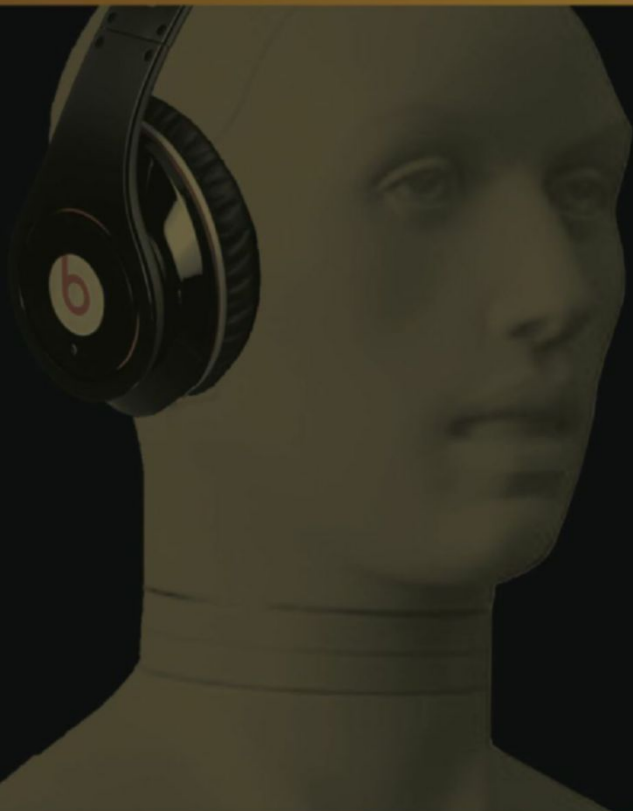
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Vibrato and Tremolo

By
Richard Honeycutt
(United States)

In this article, Richard Honeycutt discusses the differences between vibrato and tremolo and how those effects are produced in guitar amps and Hammond organs.

During my pre-teen years, my family listened to many concerts both in person and on TV. I noticed that many of the singers' voices seemed to quiver, an effect that I later learned was called vibrato. I found that some singers' vibrato added musicality, but others were annoying. My older sister and I characterized the annoying vibrato as "fake vibrato," and the pleasant form as "natural vibrato." I have since learned that there is indeed a natural vibrato that is a result of proper vocal technique, although vibrato can be produced unintentionally. Vibrato can also be produced by musical instruments. The technique for doing so varies with the instrument.

A musical effect that is often confused with vibrato is tremolo. Originally, the term tremolo was used of multiply plucking the string of an instrument (e.g., a mandolin) in order to give the effect of sustain from an instrument that naturally has little sustain. More recently, tremolo has come to mean rapid variation in the loudness of a note; whereas vibrato denotes variation in pitch. With electrical or electronic instruments, pure vibrato or tremolo is possible, but for acoustic instruments including the voice, vibrato often includes variation of both pitch and loudness.

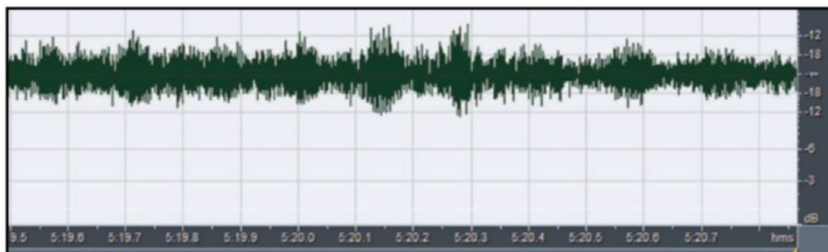


Photo 1: The amplitude of a tenor's voice signal varies with time.

Photo 1 shows the waveform of a tenor singing a sustained note from "Comfort Ye" in Messiah with vibrato. The accompanying tremolo is clearly discerned. Although it is not easy to tell from the image, the instantaneous pitch is varying also. Since pitch variations in musical vibrato comprise only a small fraction of the central pitch, they are difficult to perceive from an image of the waveform.

However, if we perform a Fast Fourier Transform (FFT) of the waveform, we can see whether each note is represented by a narrow line for the fundamental and each overtone, or by a wider line showing a range of pitch variations during the analyzed portion of the note.

Photo 2 shows the FFT of the waveform shown in Photo 1. The tenor is singing a G# at about 207Hz, but the fundamental's amplitude is below that of the second and third harmonics, which can be seen as fairly wide frequency spreads surrounding 414Hz and 621Hz, respectively.

The fact that the human voice does not produce a pure sine wave, and that the tenor in the recording from which the waveform in Photo 1 is taken was accompanied by an orchestra playing softly makes it difficult to clearly see the effects of the tremolo and vibrato.

Photo 3 shows the waveform of a solo cornet playing a note with almost no vibrato. Notice how narrow the spikes are. **Photo 4** shows the waveform produced by a cornet trilling from D5 (587Hz) to E5 (659Hz), which is similar to a wide vibrato. Note how much wider the frequency spikes are, indicating acoustic energy being radiated in a wider band of frequencies than shown in Photo 3.

Parameters of Vibrato and Tremolo

As players of electric guitar know, two vibrato and/or tremolo controls are generally available on an amplifier: depth and rate. The depth control affects how much the frequency varies when vibrato is used, or how much the amplitude varies when tremolo is used. The rate control affects the speed of the variation. Vibrato and tremolo produced by the voice and the acoustic musical instruments generally has a rate between about 5Hz and 15Hz. For reference, the tenor whose waveform appears in Photo 1 has a vibrato rate of about 13Hz, and the cornet whose waveform is shown in Photo 3 has a rate of 8.5Hz in another passage in the same recording (not shown in Photo 3). I once knew a vocal instructor whose students were taught to create a very slow, very wide vibrato whose rate was about 1.6Hz and whose pitch variation was over a semitone above and below the center pitch. The effect was very unpleasant. 1960's folksingers Sylvia Tyson and Buffy Sainte-Marie both had unusually fast vibrato rates that were a natural characteristic of their voices.

Pipe organs often have tremulant controls that add both vibrato and tremolo to the sound. Some of these allow for variation of depth and rate of the effect.

Electronic Circuits to Produce Tremolo

Changing the amplitude and/or frequency of a wave is called modulation. The equation for the instantaneous voltage of a sine wave is:

$$v(t) = v_0 \sin(2\pi ft + \phi)$$

where $v(t)$ = the instantaneous voltage as a function of time

v_0 = the initial voltage (DC in the case of a mixed DC+AC signal)—this is the peak amplitude

f = frequency in Hertz (Hz)

t = time, and ϕ is the initial phase angle

If we multiply the right-hand side of the equation by a number or a wave function, we change or modulate the amplitude. **Figure 1** shows what happens if we modulate a high-frequency wave (e_c) by a lower-frequency wave (e_i), producing an amplitude modulated (AM) wave. If e_c is taken to represent the audio from a musical instrument note and e_i represents the tremolo control waveform, the AM wave shown at the bottom row of waveforms of Figure 1 is the resulting note with tremolo.

Tremolo circuits used in guitar amps usually consist of a variable-frequency oscillator whose frequency of operation ("tremolo rate") is well under

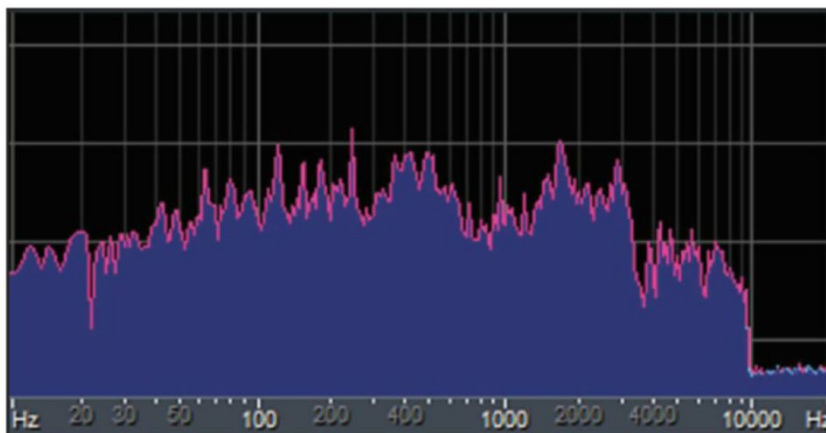


Photo 2: The Fast Fourier Transform (FFT) of a note sung with vibrato shows a spread of frequencies for the fundamental and each harmonic.

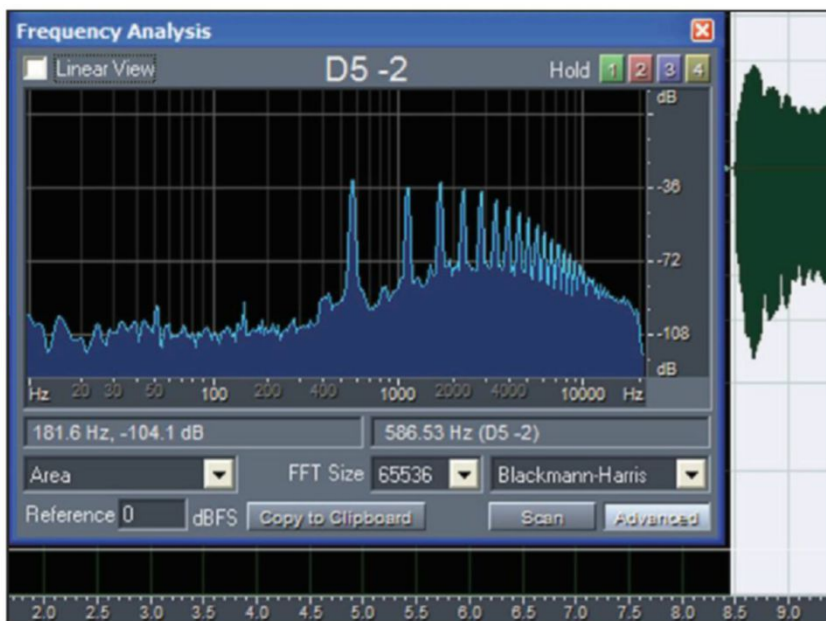


Photo 3: The spikes at the fundamental and overtone frequencies of a cornet played with little vibrato are quite narrow.



Photo 4: The spikes at the fundamental and overtone frequencies of a cornet playing a trill are wider.

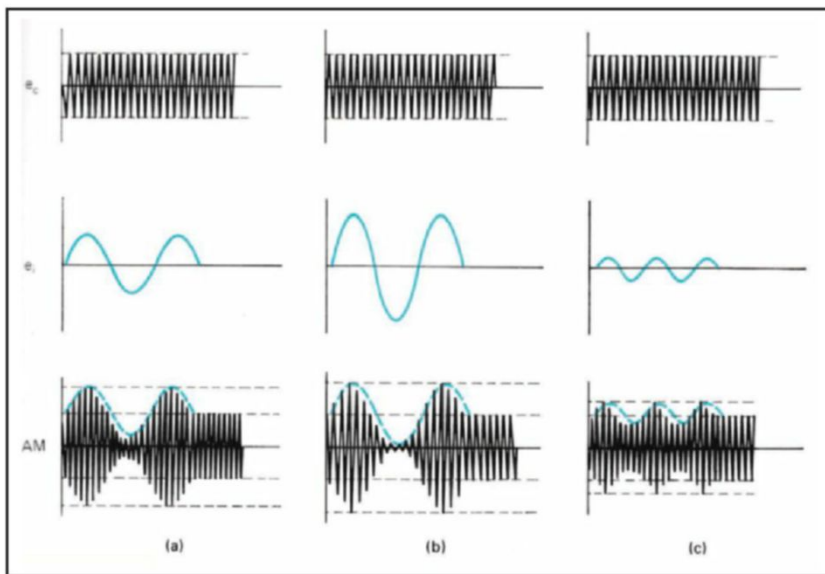


Figure 1: These figures show three levels of amplitude modulation (tremolo depth).
(Source: G. Miller, Modern Electronic Communication)

10Hz, either driving an optoisolator or varying the control-grid bias of the output tubes. Optoisolators used in classic amps usually comprised a neon lamp illuminating a cadmium sulfide or cadmium selenide light-dependent resistor (LDR). The neon lamp was placed in the plate circuit of a triode whose grid was driven by the oscillator's output signal.

The resistance of the LDR varies more-or-less inversely with the light level, and neon lamps' brightness is proportional to the current in the lamp. However, since the internal resistance of a neon lamp is extremely low after the voltage across it has exceeded its "striking voltage" (~90V), control of current depends upon the circuit in series with the lamp. Maximum current for a small neon lamp used in optoisolators is typically about 1.5mA to 2mA.

Figure 2 shows the partial schematic of a Fender Twin Reverb amplifier, with the tremolo circuit outlined in blue. The left-hand triode is the oscillator that produces the tremolo control signal. The circuit includes three RC networks: the 0.02μF capacitor, the SPEED control, and the 100kΩ resistor; and

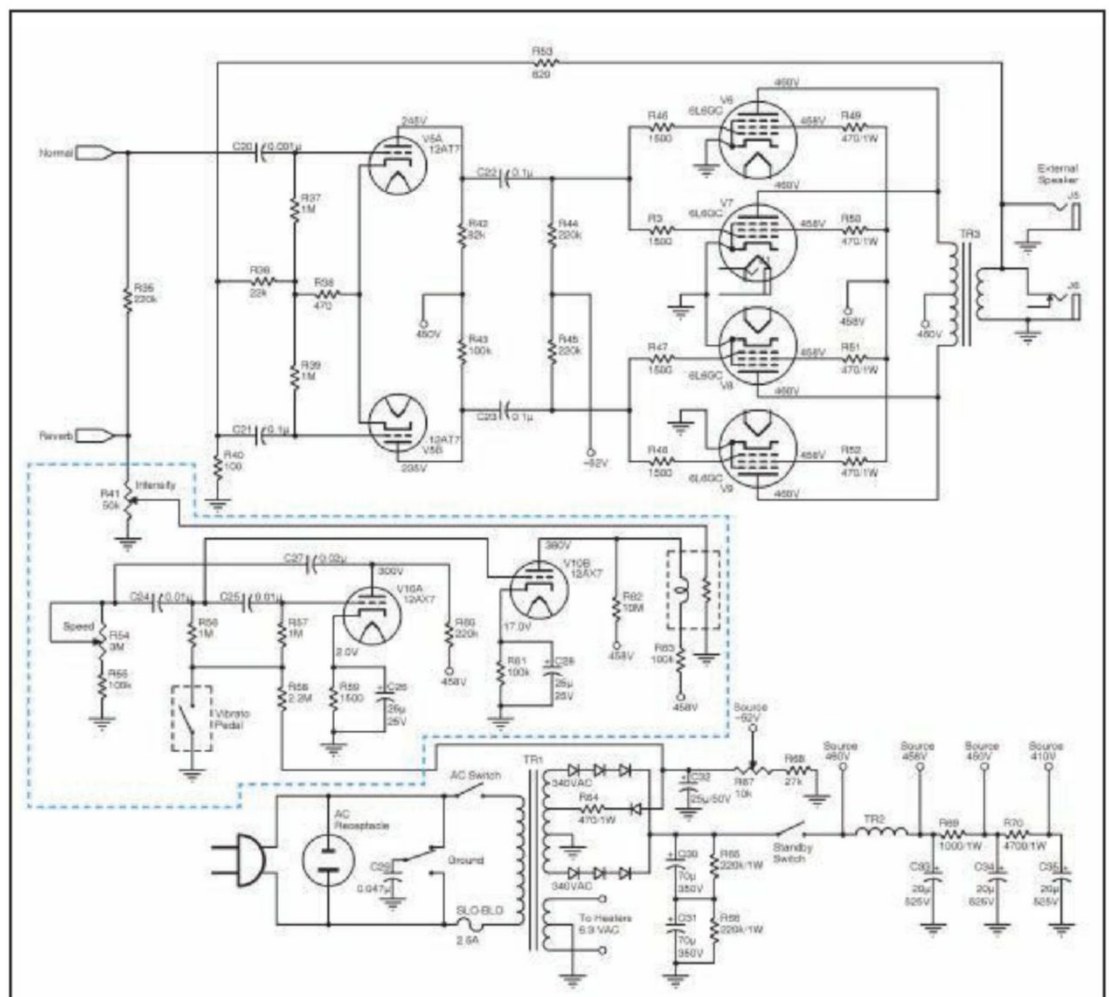


Figure 2: The tremolo oscillator of the Fender Twin Reverb amp is outlined in blue.

each of the 0.01 μ F capacitor-1M Ω resistor combinations. At some frequency, the total phase shift provided by these three networks will add up to 180°, and there will be positive feedback, cancelling the 180° phase shift caused by the triode stage and causing oscillation.

The SPEED control affects the frequency at which this occurs, within limitations provided by the other components in the phase-shift network. (A common complaint about the Fender tremolo circuit is that guitarists are not able to slow the tremolo as much as they would like. Replacing the 0.01 μ F capacitors with 0.02 μ F units is a solution.) Most Fender amps using the optoisolator-style tremolo have pretty much the exact same oscillator as the Twin, and the rest of the tremolo circuit is nearly the same. A good reference for helping you design a phase-shift oscillator can be found on the Aiken Amplification, Inc. website (see Resources).

In the Twin Reverb circuit, the right-hand triode drives the optoisolator, which is connected so as to vary the amplitude of the “vibrato-channel” signal fed to the 12AT7 phase inverter. Once the triode’s plate current is low enough that 90V appears across the neon lamp, the lamp will begin to glow and then the voltage across the lamp will drop to about 60V.

At this point, the tremolo oscillator wave applied to the grid modulates the plate resistance of the tube and that resistance, plus the 100k Ω resistor in series with the neon lamp will control the brightness of the neon lamp. That undulating brightness, in turn, controls the resistance of the LDR of the optoisolator in an inverse fashion: the brighter the neon lamp at any instant, the lower the resistance of the LDR, and the more the signal is attenuated. Thus, the guitar signal being amplified is amplitude-modulated by the tremolo oscillator.

As an aside, as one tries to understand the operation of a neon-LDR optoisolator, it helps to remember that a triode is a voltage-controlled current source (VCCS). The grid-to-cathode voltage controls the plate current.

If a neon lamp is connected between the B+ supply and the plate, with or without a resistor in series, the current in the neon lamp will be controlled by the tube, even though the voltage across the neon lamp remains constant at about 60V once conduction has begun. Thus, it is not helpful to think in terms of the voltage gain of a triode used as a neon lamp driver, but rather in terms of the control of a plate current.

The sound character of any particular tremolo circuit depends upon the tremolo waveform, the way in which the optoisolator tube is driven, and where in the amplifier circuit the amplitude modulation is applied. **Figure 3** and **Figure 4** show the waveforms produced by the tremolo oscillators of a Fender Twin Reverb amp and a Silvertone model 1484 amp, respectively. The two amps have very different tremolo sounds, and the difference in waveforms is part of the reason.

Another part of the reason is the way the tremolo intensity is controlled. In the Silvertone, the amplitude of the oscillator waveform is varied before being applied to the tube stage driving the neon lamp, whereas in the Fender, the control is on the output side of the optoisolator. In the Fender, that lets the tube drive stage run full on, while in the Silvertone, the drive voltage to the neon

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Figure 3: The tremolo control wave of a Fender Twin Reverb amp has significant distortion.

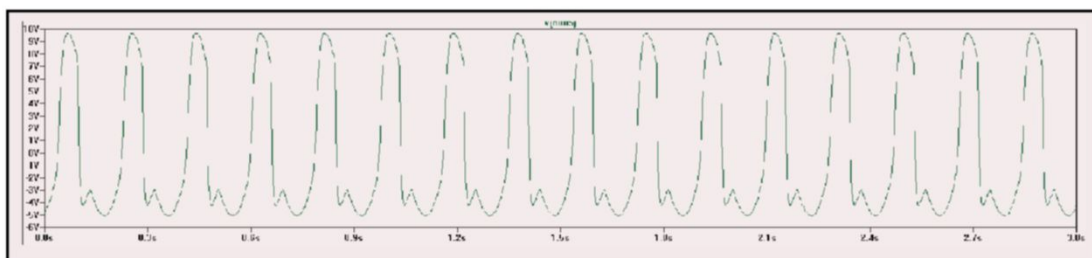
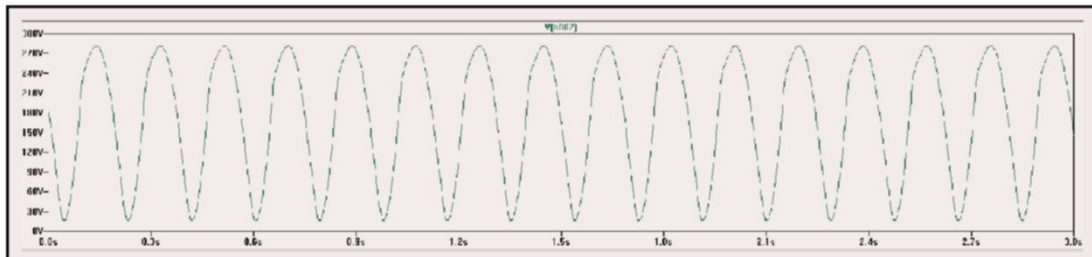


Figure 4: The Silvertone Twin-Twelve's tremolo control wave is closer to being a sine wave.



lamp is controlled by the tremolo depth control. The signal amplitude and impedance at the point where the tremolo acts can also affect the sound of the tremolo.

Another way to provide tremolo is to feed the wave from the tremolo oscillator to the bias network of the output tubes, varying the gain of the output stage. The Princeton reverb and several other Fender amps including the Vibroverb use this “bias wiggling” approach.

Yet another approach to tremolo involves applying the tremolo wave to the screen grid of a voltage amplifier, or even a power tube, as in one of the small tube amps using a single-ended 6AQ5 or a similar pentode as the power output tube.

Some amps apply the output of the tremolo oscillator to a phase inverter, producing mirror-image outputs. The guitar signal is then split into low- and high-frequency parts, which are handled by different triodes. The mirror-image tremolo wave then varies the amplitude of the lows and highs inversely to each other. This has sometimes been

mistaken for true vibrato, but is really more similar to a stereo chorus effect.

Electronic Circuits to Produce Vibrato

As we previously mentioned, vibrato involves changing the instantaneous frequency of a wave. This can be done either by direct frequency modulation (FM) or by phase modulation (PM). Since frequency is the time derivative of phase, FM and PM are equivalent but not identical. In FM, the deviation, or instantaneous variation in frequency compared to the unmodulated wave, depends only upon the modulating signal amplitude; whereas in PM, it depends upon both the modulating signal's amplitude and its frequency. So if the modulating signal is low-pass-filtered (integrated) before being used to phase-modulate the “carrier” wave—in this case, the notes produced by the electrical or electronic musical instrument—true FM results. AM and FM are compared in **Figure 5**.

The Magnatone 260 is one of very few amps—perhaps the only amp—that provides actual vibrato. It does this by phase-modulating the signal using a unique patented approach. **Figure 6** shows a schematic of the Magnatone 260. The two 6CG7 sections appearing near the right center of the schematic function as the two-stage phase modulator. Each section has a varistor connected to the plate, and another to the cathode. Since the cathode and plate signals are 180° out of phase, and have approximately equal voltages in this circuit, whichever varistor has the lowest resistance passes more signal current, making the phase of the signal at that varistor dominant.

The resistance of a varistor depends upon the voltage across it, and the voltage is controlled

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.



by the oscillator waveform applied to the 6CG7 triode section. This vibrato using phase modulation should not be confused with a phase-shifter effect, which first became popular during the 1970s. A “phase shifter” was actually a device that phase-modulated the signal and then mixed the result with the original signal. The characteristic sound resulted from swept comb filtering resulting from constructive and destructive interference between the original and the phase-shifted wave. This effect is similar to flanging, except that flanging involves delaying—not phase-shifting—the signal before mixing the result with the original signal.

The Magnatone vibrato circuit is expensive to build because it requires more parts, and while some people really love its sound, others prefer tremolo, perhaps because they are accustomed to it.

Hammond Organ Vibrato

Hammond organs produce two vibrato effects: “vibrato” and “chorus.” The vibrato is produced by phase-modulating the output from the organ’s tone generator. This is accomplished by using a “vibrato

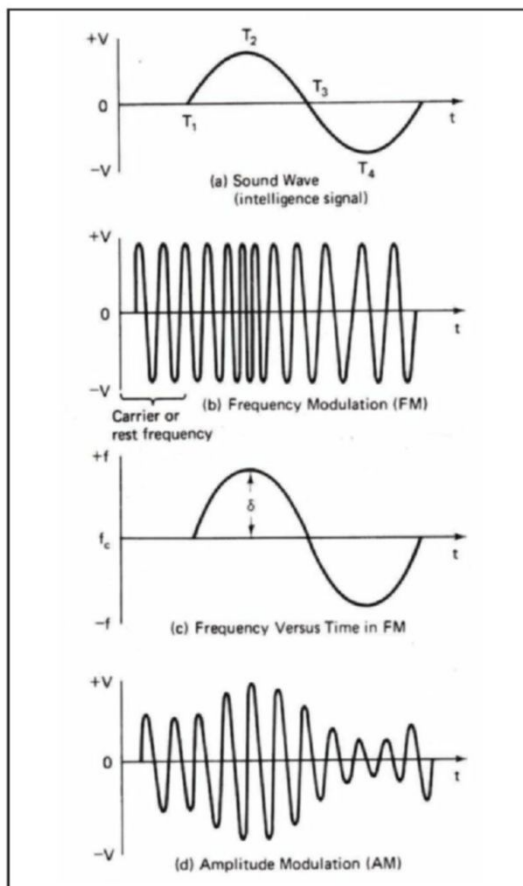


Figure 5: This figure compares tremolo (AM) and true vibrato (FM). (Miller, op. cit., p. 156.)

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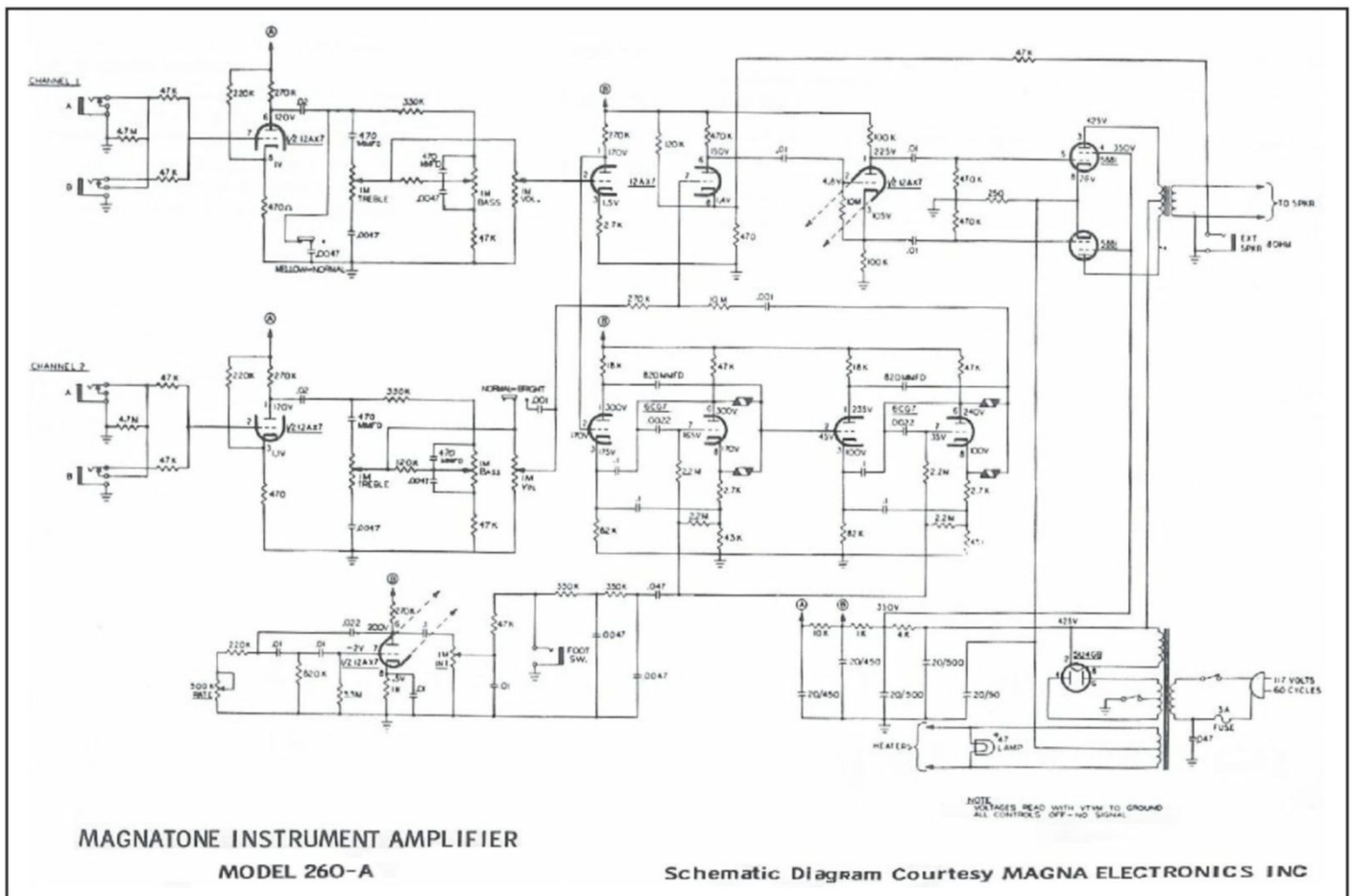


Figure 6: The Magnatone 260 provided true vibrato.

scanner that noiselessly switches the signal among plates of a rotating variable capacitor used in an all-pass filter circuit. (All-pass filters produce phase delay but do not affect frequency response.) The vibrato scanner runs at about 7 Hz (~420 RPM). Some models have switching arrangements that allow the intensity ("depth") of the vibrato to be controlled. The chorus effect is created by blending some of the "dry" (un-vibratoed) signal with the vibrato signal.

For a short while, Hammond produced the chorus effect by using a second tone generator with the frequencies detuned by about 4% to 8% (depending upon which frequencies). The output of the chorus generator was mixed with the main generator output when the chorus effect was switched in.

Leslie Speakers for Tremolo and Vibrato

Donald Leslie began working in the late 1930s to create a speaker for a Hammond organ that more successfully emulated a pipe or theatre organ. He discovered that baffles rotating along the axis of the speaker cone gave the effect he was seeking. Hammond was not interested in marketing or selling the speakers, so Leslie sold them himself as an add-on, targeting other organs as well as Hammond. Leslie made his first speaker in 1941.

The Leslie model 122 speaker has a rotating drum with one side opening, placed below the 15" woofer, plus a pair of rotating horns driven by a compression driver. It produces a combination of amplitude and phase modulation, using the Doppler Effect. The drive motor has three speeds: off (no effect), "chorale" or low speed, and "tremolo" or high speed. In addition to their use with Hammond organs, Leslie speakers have been employed to add tremolo and vibrato to guitars and other instruments, and even voices (e.g., "Just Dropped in to See What Condition My Condition Was In" by The First Edition, 1967).

Resources

"Designing Phase Shift Oscillators for Tremolo Circuits," Aiken Amplification, Inc., 2018, www.aikenamps.com/index.php/designing-phase-shift-oscillators-for-tremolo-circuits

G. M. Miller, Modern Electronic Communication, Prentice-Hall, Englewood Cliffs, 1993, p. 51.



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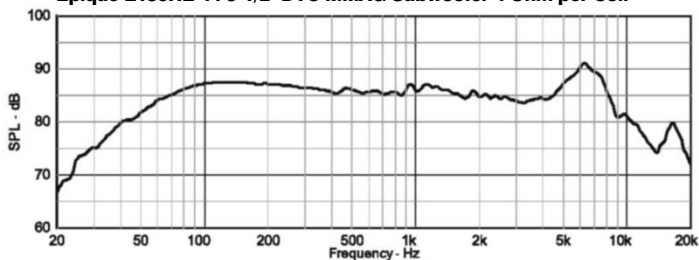
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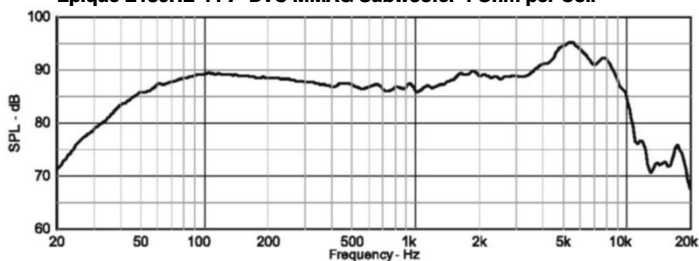
Patented MMAG motor system

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- Damped woven carbon fiber cone delivers remarkably smooth response
- Inductance control for extremely low distortion

Epique E150HE-44 5-1/2" DVC MMAG Subwoofer 4 Ohm per Coil



Epique E180HE-44 7" DVC MMAG Subwoofer 4 Ohm per Coil



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