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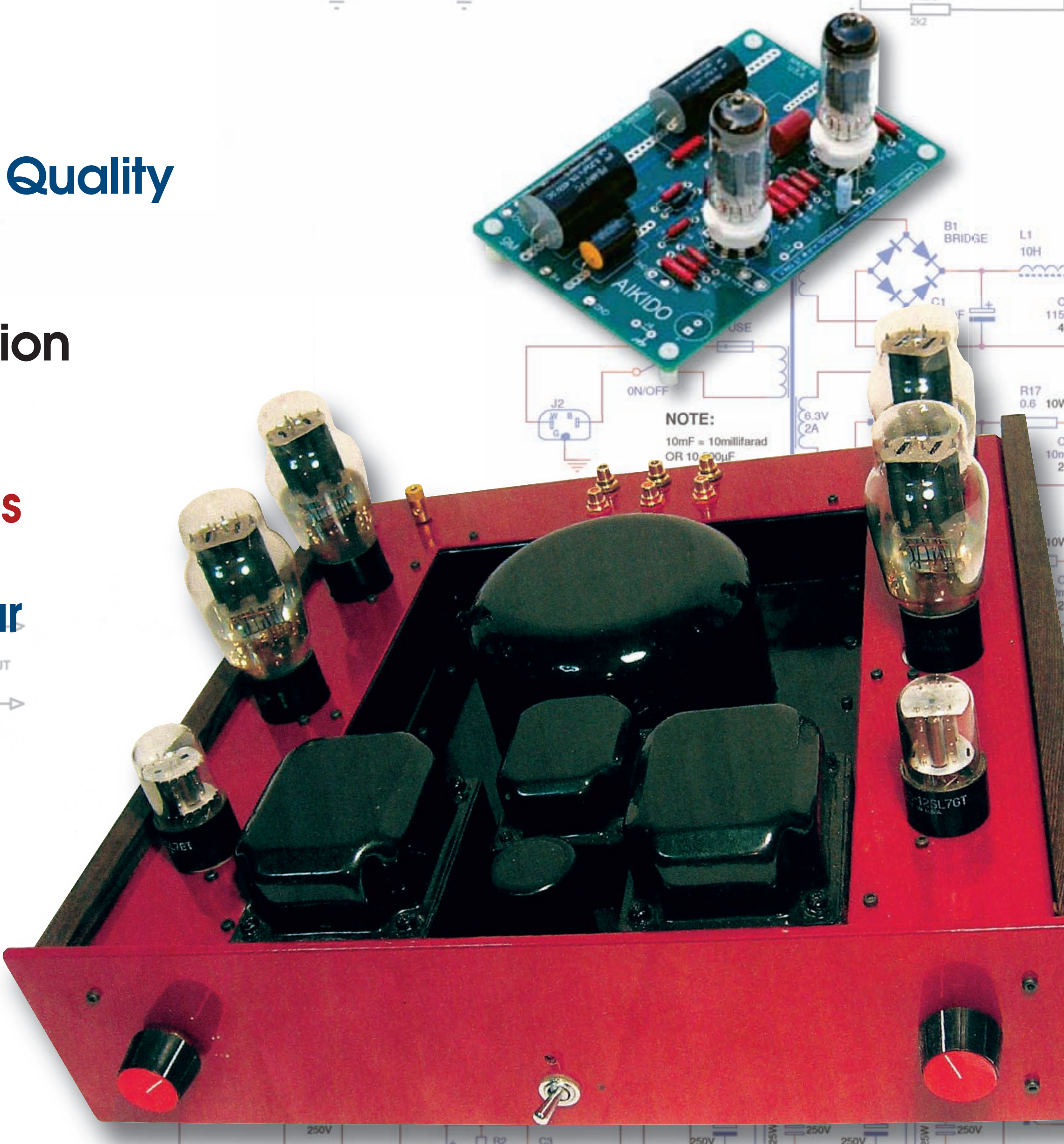
## How to Design A Power Amp



## A Truer Test Of Component Quality

## Examining Time Dispersion in Speakers

## Ear Intricacies and How You Hear

11

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## Editorial

# Paper vs. Screens?

By Edward T. Dell, Jr.

What Benjamin Franklin began back in 1775 is now rapidly being destroyed. Ben, a youngest son from Boston, ran away to Philadelphia, became the discoverer of electricity, founder of the first fire department, negotiator for funds to financially enable the American Revolution, and our first assistant Postmaster General.

As one of the staunchest of supporters of American independence, he believed it essential that if the powers of government were derived from the consent of the governed, then the governed ought to be well informed. His postal service set up a special rate for periodicals to make their distribution possible, and more affordable.

No one these days seems to have any memory or regard for this vital public support for citizen information. The United States Postal Service is a quasi-governmental entity which operates on a fiction that it is not part of the government. For some time, up until a couple of years ago, this was belied by the tags adorning USPS-owned vehicles which read "US Government." These have since disappeared. Now, the vehicles have no tags at all. Naturally not, since the USPS is an "independent" agency whose deficits are funded by the US Congress, and which pays no taxes of any kind, not on gasoline, or vehicles, or anything else.

The USPS has raised rates for delivering periodicals by approximately 18%. At least that is the rumor. These rates have remained somewhat mythic for several months since even after they went into effect, no publisher knew exactly what his or her costs were to be since the postal people did not know either. No rate forms, no mailing forms for periodicals were available either and still are not, three months after going into effect. Publishers pay at two levels for mailing magazines and newspapers, one for editorial content, and the other for advertising pages. The latter is scaled in cost based on distance.The larger the publication's circulation, the cheaper the rate, since they can pre-distribute palletized batches by truck to distribution centers.

The Internet has changed everything. First-class mail is used primarily for moving checks, and that is shrinking. Advertising is moving from newspapers and magazines to the web. It appears likely that USPS is repairing its first-class shortfall by charging periodical readers higher rates. Inevitably, publishers must pass along these rate rises to the subscriber. We are holding steady in our rates, except for overseas readers, for the moment. Rates for 2008 will be higher, however. How much will depend on what we eventually hear about what the rates are.

Internet and periodical rivalries have raised the inevitable question of whether periodicals will survive. Our 400-year-old reading habit of holding paper in our two hands will take time to disappear—if ever. The vote is ongoing. Most periodicals have expanded their offerings to the web.

You can watch a politician in a movie attached to his quote in an Internet news story. This certainly enriches what periodicals are able to offer readers and is another advertising revenue stream. But it remains true that periodicals cannot survive without advertising. This periodical cannot. Technology has helped us do more with less. We remain efficient and have continued to improve our productivity.

As always, you are in charge. What you value you will pay for. You have made *audioXpress* a unique, worldwide vehicle for information about the technicalities of audio. It is also a community where we are able to explore a dynamic technology together. It is a showcase for our best collective thinking.

Your choices about subscribing and patronizing our advertisers will determine our fate.—*Ed.*

# A Mains Analyzer—How Clean Is Your Juice? Part 1

In the process of building a fast and safe way to test mains quality, the author discovers some dirty secrets of the mains supply you take for granted.

This project started out as a simple need to know how much current was being drawn from a particular electrical outlet by a client's equipment rack, together with an innocent curiosity as to whether there was any detectable noise on the electrical supply in general. What I learned along the way has radically changed my opinion of the quality of daily power use. While the project may be a bit over the top for one-time use, the various electronic blocks should be useful to many in a multitude of applications. The measurements I present should be required reading for anyone designing mains powered equipment.

## STOCKING UP

I wanted to measure the line current being drawn from a particular outlet in a safe and convenient way. Not having a mains plug to banana adaptor or other suicidal device in my toolkit, measuring mains voltage and current was difficult, to say the least. I could have purchased a current clamp, but even these need a special mains cord or pigtail, so the live or neutral can be separated out for the clamp to loop around. You also still need a galvanic connection to measure voltage. A search on the Internet failed to find what I had in mind—a simple in-line device—so I decided to build my own special box.

In true engineering fashion I started by taking a step back to see the big picture; a simple ammeter was just a wasted opportunity. I figured that while I was at it, I should measure line voltage. And why not ground leakage current to highlight any dangerous fault conditions? Detection of outlet wiring faults would be useful, as would a Ground Fault Circuit Breaker (GFCB) tester.

As an aid to debug any problems uncovered, an isolated monitor output for connection to a scope or spectrum analyzer would be essential. To know when to use the monitor, you need to know whether there's noise or hash on the line, so let's add a glitch detector to increase the cool factor.

At the end of this brainstorming session, my behemoth ammeter included the following:

Line voltage display
Line current display
Ground leakage current display
Indication of outlet wiring faults
Indication of glitches or noise on line
GFCB test capability
Isolated line monitor output
20A load capability

## POTENTIAL PROBLEM

As usual, I started by sketching a block diagram on a handy napkin and quickly made a few realizations. The first was that there is no obvious common "zero volt" reference. I wanted to pick a common reference point for the two current measurements (line and leakage), but concluded that the earth leakage circuit needed to be completely isolated from the line current and voltage circuits. If I referenced the line current circuit to neutral and the leakage current circuit to ground, and then connected to a faulty outlet that had neutral swapped with live, then I'd have 120V AC between the two input amplifiers, which is enough to cause any IC to go bang in a big way.

The next realization was that it would be very bad if something went wrong with this test box and interrupted the neutral and ground wires, causing all the attached equipment to float off the live wire. The use of wimpy wire, low power rating components, or fuses was unacceptable; this would need to be an exercise in overkill. I ended up with the block diagram in **Fig. 1**.

## POWER CIRCUIT DESIGN

Looking on the Internet, I found some heavy-duty shunts rated at 50mV for 25A (i.e., 2mΩ). At rated current these shunts would dissipate 1.25W, and with their whopping 5″ × 1″ dimensions, I was confident that any residential circuit breaker would pop before these would. 14-gauge copper has a fusing current of around 160A so this is an appropriate cabling choice. I used the plug end of a damaged extension cord (from the last time the high school lad next door trimmed my hedge!) and quality outlet sockets to complete the power section.

## METERS

I wanted digital panel meters to display the voltage and current measurements, and I found some 3½ digit LCD modules with ½″ high numerals for under $8 each at my usual surplus stores. I bought three. The meters had a full-scale input of 200mV, so I was going to need some amplifiers. I produced **Table 1** as a design aid.

The sensing signal input is the signal from the power circuit corresponding to the maximum expected voltage or current to be measured. In the case of the current, it is the signal from the 25A shunts at 2mΩ. This is a pretty small signal requiring rather large input gains. The voltage sense signal is from a resistor divider, and I chose the resistor values to yield the ideal amplitude signal, hence the input gain requirement is 1. I wanted the biggest signal through the precision rectifier to minimize errors.

Because I was working with ±5V supply rails, I needed to allow for peak versus RMS levels and also a hundred millivolts or so because the op amps can't quite hit the rails. I picked 3.1V as the optimum full-scale amplitude. The voltage sense circuit needs to cater to a mains supply 20% above nominal, so I chose the resistor divider to provide the 3.1V RMS for an input of 144V RMS.

This all means that to bring this 3.1V full-scale signal down to the 200mV the displays expect, I need an output attenuator, which can most easily be obtained with another resistor divider around a calibration pot. With the system level planning complete, I could get down to the fun part.

## CURRENT METERS

The current sense signal from the 25A shunts is 2mV per amp, so significant gain is required: 78 or about 38dB for the line current and 781 or about 58dB for the ground leakage current. 38dB is achievable in a single op amp stage, but 58dB is best split into two stages to ensure sufficient bandwidth and minimize offset voltage problems. For the line current sense amplifier, I used a standard differential stage with a Kelvin connection to the 25A shunt; the amplifier is shown in **Fig. 2**.

I DC-coupled the amplifier to ensure that I had good common-mode rejection at 60Hz and because the low source impedance of the shunt mitigated any input offset or bias problems. To ensure stability and to reduce noise, I rolled off the high frequency with C16 and C17 at 340Hz, so there was less than 1dB rolloff at 150Hz. Low-frequency rolloff is provided by C18 with a –3dB point at 5Hz.

For the ground leakage current sense amplifier, I needed 10× the resolution, so I chose a two-stage instrumentation amp (**Fig. 3**).

The signal I wanted to measure was the same order of magnitude as the input offset voltage for the LM324 op amps I used, so to minimize offset problems, I set the gain of the first stage slightly lower at only 26dB. This allowed me to use fairly low-value resistors, which also helped minimize noise. The classic instrumentation amplifier circuit used has excellent common-mode rejection, and, coupled with a Kelvin connection to the 25A shunt and twisted pair interconnects, the induced noise level should be acceptably low. A simple inverting op amp stage around U9C with a gain of 32dB provides the appropriate signal for the RMS measurement stage. Again, I rolled off at 400Hz to minimize noise.

My biggest mistake here was setting the amplifiers' input impedance way too high. It turned out that there was a significant leakage current flowing through the input amplifier from the mains supply to ground. This current developed a differential signal at the input due to the difference between the currents in the positive and negative amplifier input resistors (and to some extent the difference in the 1% resistor values). When your source impedance is 2mΩ, a high impedance is 1Ω. Why was I using a 10MΩ input impedance amplifier?

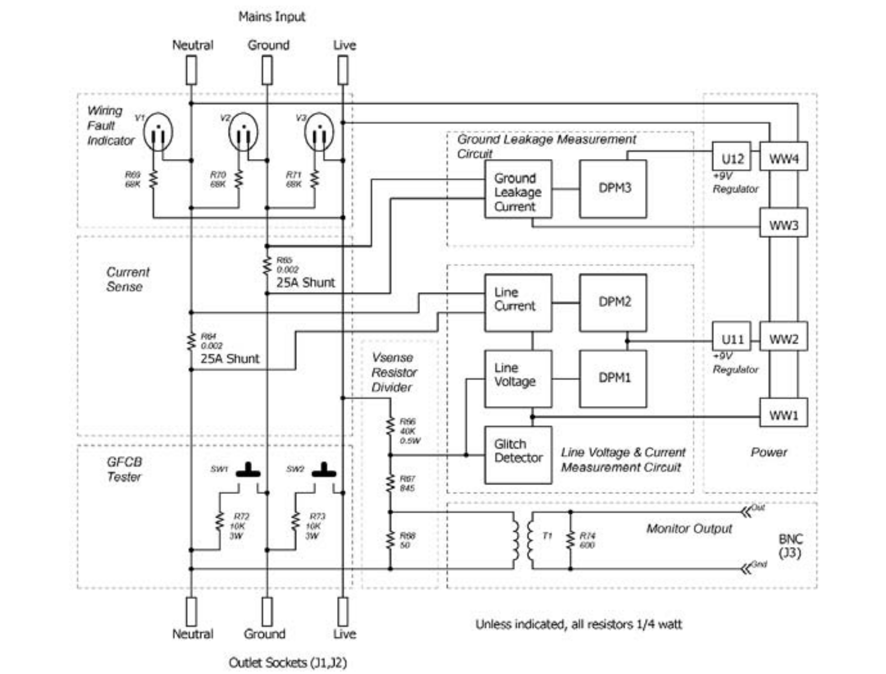


**FIGURE 1: Top level diagram of the mains tester.**

**TABLE 1 Input and output gains for RMS**

| | max. expected signal | sensing signal input | input gain | output gain |
|---|---|---|---|---|
| line voltage | 150V | 3.125V | 1 | 0.064 |
| line current | 20A | 40mV | 78 | 0.064 |
| ground leakage | 2A | 4mV | 781 | 0.064 |

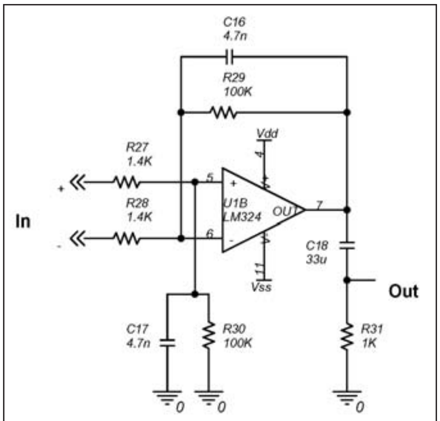


**FIGURE 2: Line current input amp.**

I tied both inputs to amplifier ground with 50Ω resistors and the improvement was impressive. The noise voltage developed due to the leakage current flowing in the input resistors was reduced so much that I was now able to zero out the display where before the offset had been too great. All that noise had been running up the RMS detector past where the offset null could adjust.

## POWER SUPPLY

It wasn't until I got the panel meters home and played with them for a while that I discovered a design quirk with them; they need a 9V supply that floats with respect to the voltage you are trying to measure. Not a problem for a battery-powered hand-held meter, but definitely a wrinkle in the design when you need additional input circuitry for gain and RMS averaging. It looked as though I needed two supplies for each meter.

When I discovered the ground-neutral isolation issue, the power supply requirement increased to two wall-warts (AC adapters). Now with floating display supplies I needed a grand total of four wall-warts. This was becoming ridiculous. After much hand-wringing and pacing, I convinced myself that it was better than using batteries, the wall-warts were free (I bought a gross some years ago), and the extra weight would lend an air of quality to the final product (I should be in marketing!). I chose to use a favorite trick of cracking open the wall wart, cutting the plug prongs off with my large dykes, and hot-gluing the transformer to the base of the (insulated

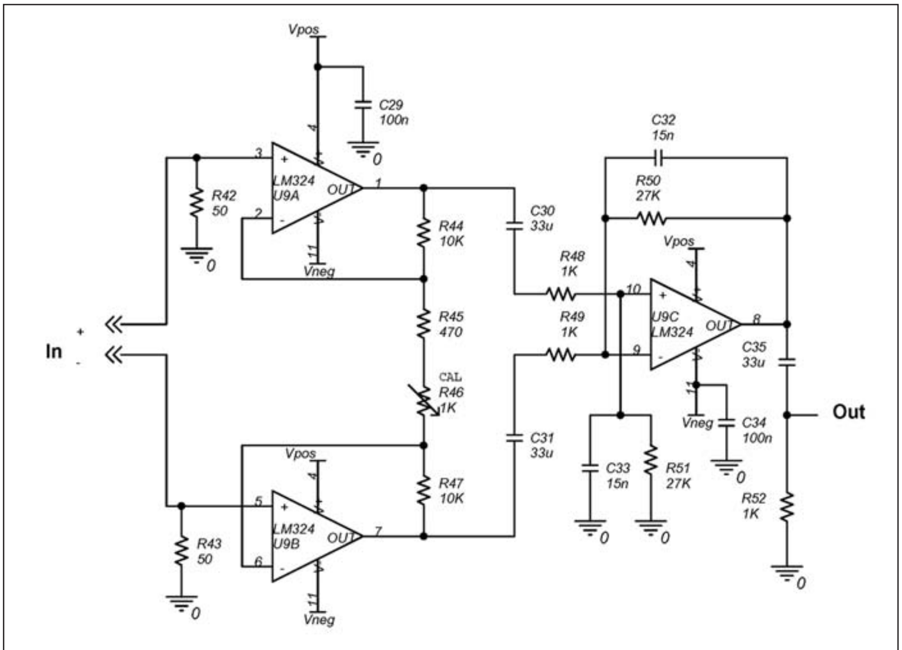


FIGURE 3: Ground leakage current input amp.

plastic!) enclosure for a fast, cheap power supply.

For the ground leakage meter circuit I used the wall wart output unregulated. There's not much to this circuit, and all stages are seeing the same signal. I even split the supply with a couple of 1k resistors for a Vcc/2 virtual ground with no problems.

For the line current and voltage meter I needed a regulated supply because the thresholds for the glitch detector are derived from the supply lines and I wanted consistent, repeatable trigger levels. Also, the glitch detector and meter circuits are seeing different signals and so to prevent any possible interaction, I buffered the Vcc/2 split with an op amp to provide a controlled, low impedance ground reference.

The datasheet for the panel meters indicated that 11V was the maximum supply and I know these wall warts can put out 15V no load. Just to be safe, I threw in a couple of 7809 regulators I had lying around to regulate the panel meter supplies to 9V.

Grabbing a bunch of wall warts, I assigned four as follows.

- A 7.5V, 200mA rated wall wart delivering 10V no load to drive the ground leakage current meter circuit. Even at 200mA this will provide >±4V to keep the LM324 in control.
- For the panel meters, a pair of 9V rated 500mA wall warts to provide a minimum of 11V even under line sag/brown out conditions. I grouped the line voltage and current panel meters on one wall wart. The 11V minimum is because the 9V regulators need at least a 2V drop to stay in regulation. The quality of the panel meter reading stability depends on the quality of this supply.
- For the line voltage and current meter and glitch detector circuits, I used a 12V rated 500mA wall wart with a no load voltage of 17V. Because I wanted ±5V regulators for this circuit, I needed a wall wart that could guarantee at least 14V for the circuit load. Pulling 100mA (a pessimistic estimate), this wall wart delivered over 16V with a supply of 123V. A quick bit of Ohm's law led me to believe the source impedance of this transformer was around 5Ω, so I would get 14.5V for 96V in (-20% line sag). Your performance may vary.

## MEASUREMENT CIRCUIT

The next piece to design was the measurement circuit—there are three of them. I wasn't after extreme accuracy—I'll use my calibrated multimeter if 5V or 100mA is going to make a difference—but I wanted a useful, linear range. The problem with just using a diode bridge is the forward voltage drop. 1.4V may be OK in a power supply design, but when my supply rail is only +4.5V, this represents a huge chunk of the available range. An Internet search produced several favorable references to the circuit in **Fig. 4**.

The way I see it, for the positive half cycle presented to the input, D4 conducts, D3 blocks, and the pair of op amps acts together as a unity gain buffer producing a positive half cycle at the output equal to the input. On negative half cycles, D3 conducts, D4 blocks; U3A therefore acts as a buffer and U5 is now an inverting unity-gain amp, again producing a positive half cycle equal in magnitude to the input. Both diodes act within the feedback loop of the op amps so their forward voltage drop is canceled. In reality, the exponential V-I curve of the diodes causes significant errors close to zero, so you want to put the largest signal you can through this circuit for best linearity and precision.

As has been done for centuries, I approximated the RMS averaging circuit with a simple R-C low pass filter (R36, C22) by assuming that I would only ever be measuring pure sine waves (which, as I discovered much later, is not actually the case for mains power). I chose the time constant of this circuit to be half the update rate of the LCD panel meter (which is 500ms), so I would get a quick but stable reading. I picked R to be 10kΩ to keep C reasonable because the 500kΩ input impedance of the LCD panel meter would allow such a high impedance at this point. Besides, I had an X0.064 attenuator to add in between.

## WIRING FAULT INDICATOR

Nothing clever or innovative here. I just wired three neon bulbs with their limiting resistors across the three mains wires as follows: amber live-neutral, amber live-ground, red neutral-ground. You want to see both amber neons come on; you'll want to fix anything else immediately.

Many neon indicators come with a limiting resistor already attached. The ones I used from Radio Shack had a 6.8kΩ resistor, which makes for a nice bright neon. However, during testing in the case where the ground line is floating, the live-ground and neutral-ground neons both came on when I really wanted only the live-neutral neon to glow. I think the neons fire at about 45V and then need only microamps to sustain; the two neons in series still had over a milliamp flowing through them.

Fast-forwarding to the answer, I cut open a commercial wiring tester and discovered that they used 68kΩ resistors. I removed the Radio Shack-supplied resistors and replaced them with 68kΩ. The neons still come on, but now they're very dim and it's easy to tell the difference.

## GFCB TESTER

GFCBs are excellent safety devices and are required by code wherever water is nearby. However, they can be a source of nuisance trip problems, especially when used with equipment that has Electro-Magnetic Interference (EMI) filters on the mains input connector. These filters have capacitors from live and neutral to ground to filter high-frequency noise from the mains line. The problem is that a 100nF capacitor will also pass over 4mA of 60Hz leakage current into the ground line. Two such devices, and you're really close to the trip current of your GFCB.

Other household appliances such as negative ion generators also have high ground leakage currents. Just to complicate things, an installed GFCB can also protect regular, non-GFCB outlets further downstream, so you may be connected to a protected line without even knowing it. The last thing you want is the GFCB popping right in the climactic final action scene of your Saturday night movie. Fortunately, you can use this GFCB tester function to safely see whether your home theater outlet is protected.

The tester simply places a resistance across either live-neutral or live-ground to cause a leakage current to flow. I have a figure of 8mA stuck in my head for the trip current of a GFCB, but I couldn't quote a source. I decided to inject around 12mA because that required a very standard 10k resistor. The worst-case power dissipation is 2W (for +20% line voltage). I paralleled up four ¼W, 47.5k resistors to make a 1W, 12k resistor and called it good enough. I know, if I keep pressing the GFCB test button long enough, I can make smoke come out.

## GLITCH DETECTOR

The idea behind this circuit was to indicate whether there was any noise riding on top of the 60Hz mains waveform. This noise could be high-frequency hash from switching supplies, load noise from electrical machinery, or just the transients generated whenever any piece of electrical equipment in the neighborhood is turned on or off. The presence of such noise on the mains supply would place an additional burden on the power-supply filtering of any sensitive audio gear and potentially cause some unexpected behavior or audio artifacts. Choosing a detection threshold for this noise proved to be more difficult than I thought, but I started out as follows.

If any power-supply-borne noise was to be inaudible on a typical line-level signal, then it would need to be well below –90dBV. Having measured some of my circuit designs in the past for their ability to reject noise on the power rails (known as power-supply rejection ratio, or PSRR), I knew that 90dB rejection was difficult, but possible, to achieve with careful design, so it seemed as though 1V of noise was a good threshold for concern. Technically, this should be 1V on the secondary, but because the

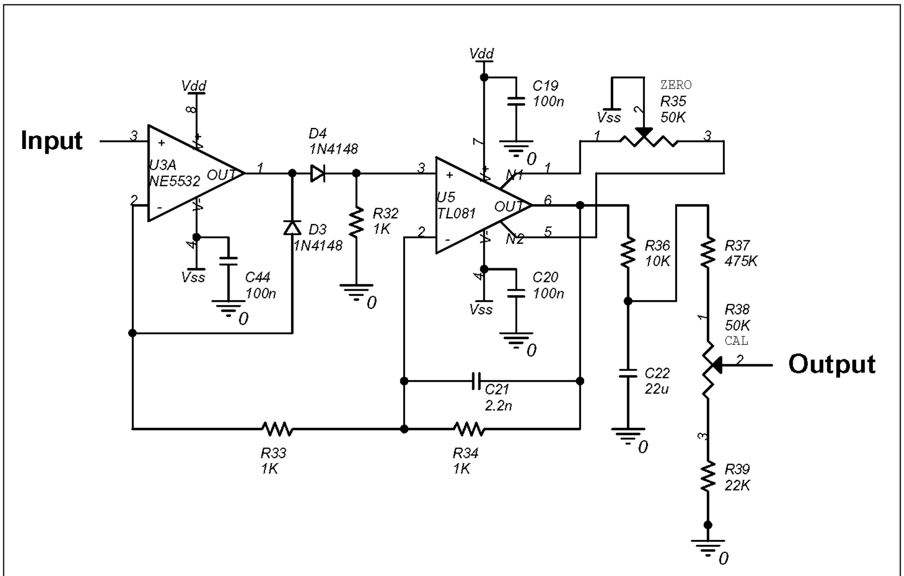


FIGURE 4: Precision full-wave rectifier.

secondary could be anything from 5V to 500V, I decided to just use this threshold straight on the mains voltage directly. Detecting a 1V signal in the presence of a second 120V signal takes some careful design; I took out my napkins.

I started with a simple R-C high-pass filter feeding a window comparator to detect positive- or negative-going spikes. Using the same voltage sense signal as the line voltage meter gave me 2.5V RMS 60Hz signal as input. The equivalent threshold of 1V on 120V was therefore 21mV on this 2.5V input signal, and the idea of running comparators with ±21mV thresholds made me nervous.

Additionally, the circuit would need to be operated at high impedance, so as not to load the line voltage meter. This made applying a mV or two of hysteresis on the comparators tricky. Adding the requirement to attenuate the 3.5V peak of the 60Hz signal down to below the 21mV threshold meant that the high-pass break frequency needed to be 8 octaves above 60Hz (8 octaves at 6dB/octave = 48dB attenuation), or 15.4kHz. This wasn't going to detect much audio band noise!

I needed a better filter. One of my favorite tricks for steep slope filters is to design a peaked (high Q) second order filter and then add an RC on the output at a slightly higher frequency (for a high pass, lower for low pass) to flatten out the passband. I chose the multiple-feedback design[1] shown in **Fig. 5** because it allows you to adjust the Q and the break frequency independently with resistor value changes.

This third order filter achieved the –42dB minimum attenuation at 60Hz with a break frequency of 400Hz, which I deemed to be acceptable. One small detail about this filter: You need a controlled source impedance. Raising the signal source impedance to even 100Ω degraded the Q of the filter so there was no way I could use the signal off the roughly 1kΩ resistor divider. I was already committed to using multiple packages of op amps, so I added a buffer to guarantee a stable low source impedance.

With that pesky 60Hz signal out of the way, I could now gain up the residuals to allow a reasonable detection threshold level which I arbitrarily set to ±1V by a resistor divider chain R8, R9, R10. The detector is shown in **Fig. 6**.

Adding 1% hysteresis on the comparators ensured clean switching when a glitch was detected. The LM339 comparators that I used have an open collector output that can sink a respectable current, typically 16mA. This meant that I could add a capacitor, C10, to ground on the output and stretch any detection output pulse so that it produced a visible flash on the indicator LED.

When one of the comparators detects a glitch, it turns on its output transistor, which discharges the capacitor. The capacitor then must charge up through the pull-up resistor, R15, once the output transistor turns back off. Because I had a pack of four comparators, I used one more, U2C, to clean up the capacitor charge waveform and drive an LED.

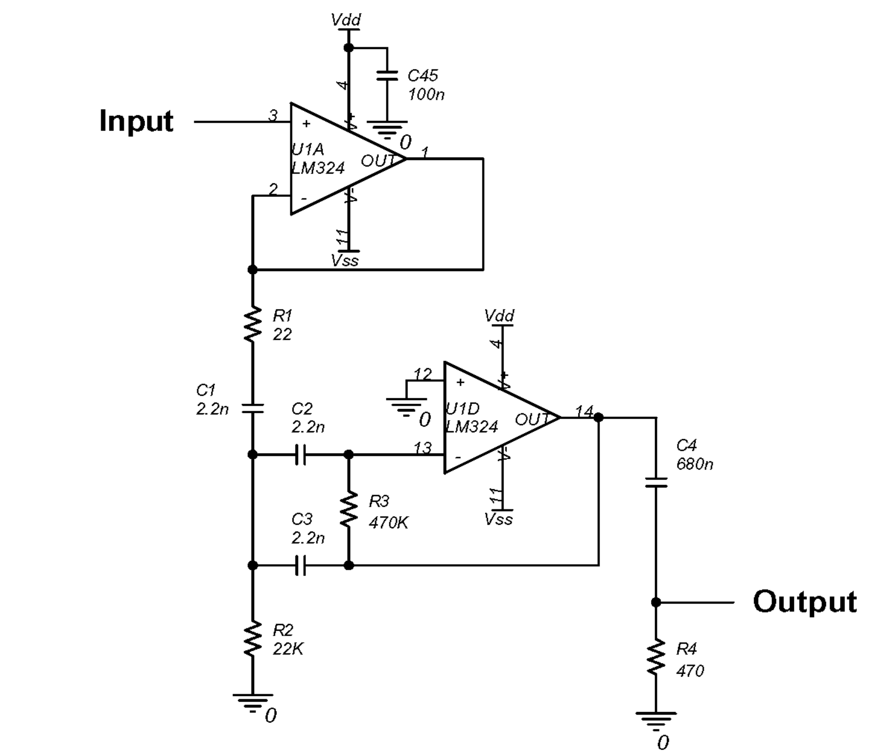


**FIGURE 5: Third-order high-pass filter.**

## MONITOR OUTPUT

Finally, I wanted to be able to quickly hook up a scope or analyzer to look at the mains waveform without risk of electrocution. I chose a 600Ω audio transformer that I knew had primary-to-secondary isolation exceeding the worst-case 205V peak you could expect from the wall outlet. I actually used a 1500V rating transformer for my own personal safety. It seemed like an easy job: 1V RMS signal from a 600Ω source resistor + 1:1 transformer + 600Ω load = perfect mains reproduction.

What I forgot to consider was that the primary inductance of the transformer I was using was only about 37Ω at 60Hz, and this loaded the 600Ω source resistor so much, it actually dropped the reading on the line voltage meter. A few napkin scribblings later, it dawned on me that there's no way I'm going to drop a 1V signal onto a 4Ω resistor to get flat bandwidth down to 60Hz. That would be a power dissipation of 30W! Not wanting to buy another (higher) quality audio transformer, I finally decided to compromise as follows:

As a mains monitoring device, it's OK to relax bandwidth to 120Hz or even 180Hz, especially if you have a known, constant attenuation at 60Hz that can be compensated for.

Reducing signal level actually increases measurement bandwidth because saturation at low frequency is the limiting design factor for the transformer. However, I must maximize signal level and optimize operating impedances to combat both inductive and capacitive coupling of mains-related signals.

A quick redesign of the resistor divider to drive the transformer with a 50Ω source, using the preexisting 845Ω resistor and 40kΩ (modified from 70kΩ) composite dropper resistor, and I was back in business. I call the dropper resistor "composite" because it's made from multiple paralleled resistors. This increases the power rating of the composite resistor. The worst-case line voltage sense signal error due to transformer impedance loading was limited to under 1%.

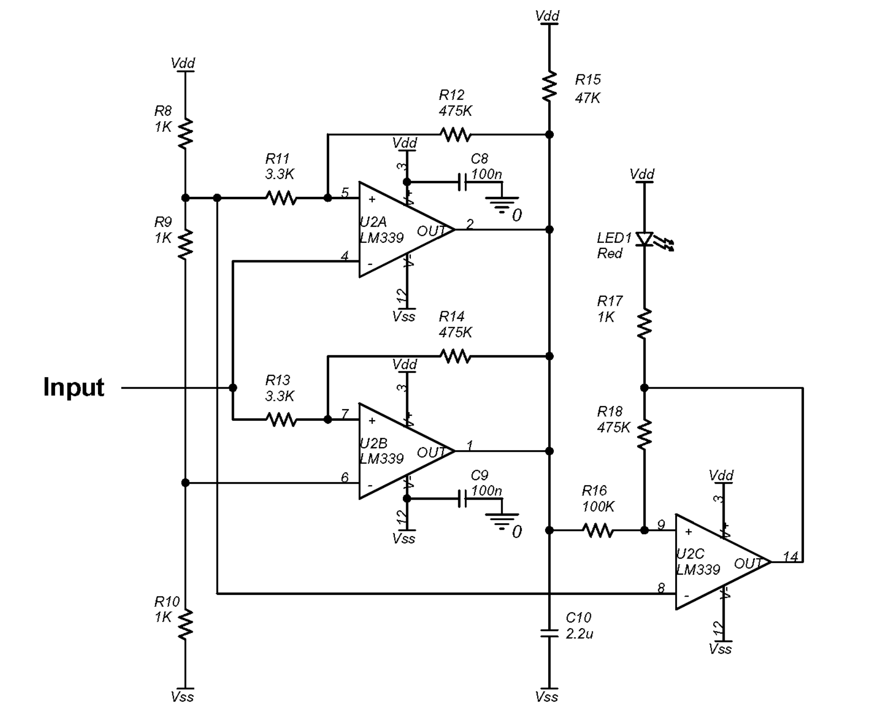

FIGURE 6: Window comparator.

Warning: Mains voltages are lethal. Please use caution and intelligence when working with exposed mains circuits. Consider this project only if you are comfortable working with high voltages. As a precaution, make sure you are not working alone and keep one hand in your pocket when probing live circuits. I accept no responsibility for any damage or destruction caused by careless interaction with mains electricity. *aX*

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# Frequency Delay Dispersion

The phase nonlinearity of most second- and higher-order speaker crossovers is illustrated with the step response of cascaded all-pass networks. Audibility issues are discussed.

In my article "Waveform Phase Distortion" (*SB* 1/97)[1], I described how most speaker crossovers (COs) generate delay dispersion, where lower frequencies are delayed more than higher frequencies. Bill Waslo[2] described the theory in excellent detail.

The audibility threshold depends on many factors; reports vary from about 100µS to 2mS. A sufficient amount, though, is certainly audible, as visualized by the "reductio ad absurdum" example of the midrange being delayed by one minute relative to the treble!

## CROSSOVERS

A true first-order (neither driver polarity inverted) CO produces a square-wave (step) response limited in perfection only by the drivers. For second- and higher-order COs, there is one type that can also (ideally) have a perfect step response, as described by John Kreskovsky in "A Transient-Perfect 2$^{nd}$ Order Passive Crossover" (*aX* May '01[3]). This type uses extensive frequency overlap and/or amplitude peaking. But most standard high-order COs (e.g., Butterworth and Linkwitz-Riley) produce an all-pass-type phase nonlinearity of order one less than the CO order. For example, a third-order CO has a second-order all-pass response.

In this article, I show the step responses of various numbers of cascaded first-order (non-resonant) all-pass stages. These have a flat amplitude response (DC-50kHz) but a nonlinear phase versus frequency curve, which produces a varying delay versus frequency relation.

## THE ALL-PASS CIRCUIT

**Figure 1** shows the op amp circuit, ten of which I cascaded on a breadboard with a jumper wire selecting 1 through 10 stages (first- through tenth-order all-pass response). **Figure 2** shows the phase shift and group delay of a single stage on linear frequency scales. Group delay is the negative of the phase versus frequency slope, and generally (but not always) corresponds to the propagation delay of a group (band) of frequencies in a transmitted signal.

**Figure 3** shows the same data with log frequency scales. Note that the curves are now symmetrical about $f_o$, where the phase is –90° and the group delay (T) is down to half the LF value ($T_o$) of $1/\pi f_o$, which T approaches for $f \ll f_o$.

Thus, low frequencies are delayed relative to high frequencies (which approach zero delay). Correcting this in real time is physically impossible over an unlimited bandwidth (required for perfect step response), without an infinite number of correction stages or a negative-delay ("crystal ball") circuit! However, it *is* possible over a limited bandwidth by progressively delaying the higher frequencies to match the LF delay.

This can be done digitally, but Dick Crawford ("The Phase Redeemer," *SB*

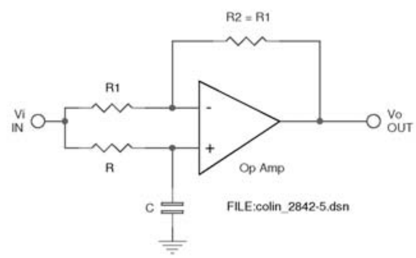


**FIGURE 1: All-pass stage.** $f_o = \dfrac{1}{2\pi RC}$ = -90° frequency. $A_v = \dfrac{V_0}{V_1} = \dfrac{1-j\omega RC}{1+j\omega RC}$.

$A_v = \dfrac{1-j\dfrac{f}{f_0}}{1+j\dfrac{f}{f_0}}$ $|A_v| = 1$ (unity gain at all frequencies). Phase: $\varphi = -2\tan^{-1}\left(\dfrac{f}{f_0}\right)$. Group delay: $T = \dfrac{\dfrac{1}{\pi f_0}}{1+\left(\dfrac{f}{f_0}\right)^2}$,

LF group delay: $T_0 = \dfrac{1}{\pi f_0}$.

Values used: R1 = R2 = 22.1k, R = 68.1k, C = 10.06nF.

Then $f_0$ = 232.3Hz, $T_o$ = 1.37mS. For n stages, $T_o = \dfrac{n}{\pi f_0}$

(e.g., 13.7mS for ten stages). For $f \ll f_0$, phase approaches 0°, delay approaches $T_o$. For $f \gg f_0$, phase approaches –180°(n), delay approaches 0.

5/97)[4] has done this with an optimized combination of all-pass phase-shift networks. He used five cascaded second-order stages to reduce the delay dispersion of a second-order CO from 57µS to 18µS, a reduction of 39µS. The compensation circuit itself delays the highest audio frequencies 50µS more than the lowest frequencies. Photos 1 and 2 of Mr. Crawford's article (p. 36) show the CO's first-order all-pass step response greatly improved by the compensator. The HF wiggles are of no consequence, because they range from about 27-40kHz.

Mr. Crawford used resonant second-order all-pass networks. Besides having double the phase excursion (0° to 360°) of a single stage, using resonance (Q greater than 0.5) sharpens the phase versus frequency curve. This allows higher frequencies to be delayed more than lower frequencies over a limited bandwidth, which is exactly what most speaker COs need.

In this article, I've used only first-order all-pass networks, which are non-resonant both singly and in cascade. The following waveform descriptions serve to illustrate the step responses of idealized (flat summed amplitude versus frequency) higher-order speaker COs.

## ALL-PASS TIME RESPONSE IMAGES

**Figure 4** shows the input 83Hz square wave and the outputs of 1, 2, and 3 stages of cascaded all-pass networks (that shown in **Fig. 1**). For up to three stages, I used the 83Hz square wave and a 2mS/div. time scale, so the responses' initial spike (of opposite polarity to the input step for odd number response order) would be (at least barely) visible.

In **Fig. 5** I slowed the driving frequency to 33.3Hz and lengthened the time scale to 5mS/div. This is so that even with ten stages (lower right waveform), the response would have time to settle before the next step transition arrives. The tenth-order all-pass (ten cascaded **Fig. 1** circuits) clearly shows the instantaneous HF transient response, followed by progressively delayed lower frequencies. It's important to remember that the amplitude/frequency responses are all very flat (hence the name "all pass"). Using TL074 quad op amps, this response even through ten stages is ±0.09dB from DC –50kHz.

If you turn **Fig. 5** upside down, the

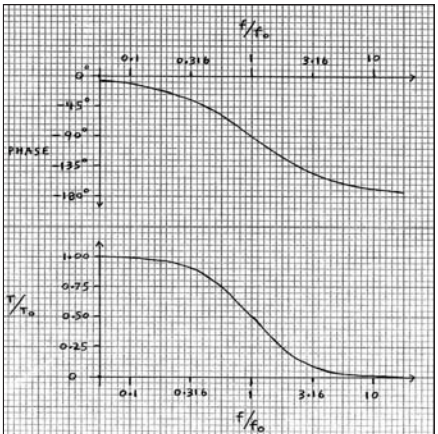

FIGURE 3: All-pass phase and group delay, log frequency scale.

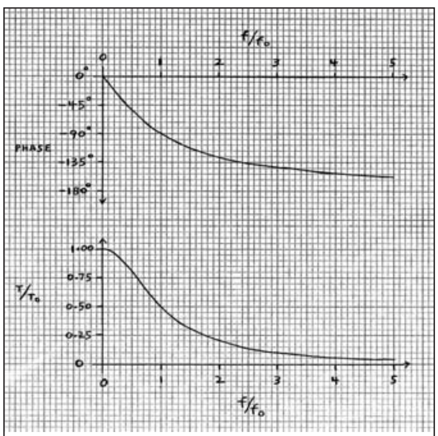

FIGURE 2: All-pass phase and group delay, linear frequency scale.

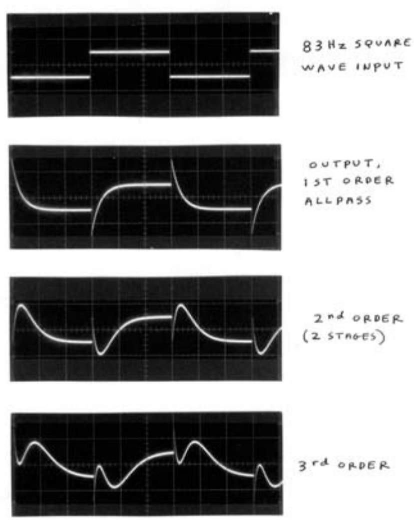

FIGURE 4: 83Hz square-wave responses, all-pass networks centered on 232Hz. 2mS/div., 1V/div.

time-reversed waveforms are what you would need to perfectly compensate this delay dispersion. Even the first-order response shows the physical impossibility of real-time, zero delay, unlimited bandwidth correction: The hypothetical "compensator" would need to "know" when the input step was going to arrive, and anticipate it by starting a positive (growing) exponential curve (mathematically starting from the infinite past!), and then magically stopping "on a dime" at the precise input step moment. Twilight Zone music, please!

Once, in the 1970s, I did compensate an all pass, but not in real time: I passed a square wave through a first-order all-pass, recorded it on tape, then played the tape backwards, feeding the time-reversed signal through the same all-pass. Within the limits of tape reproduction, lo and behold, out came a reasonable square wave!

The initial response transient, barely visible in the first- and second-order responses in **Fig. 5** (viewed right side up), become invisible in the higher-order waveforms. But they are there, of opposite polarity to the input step for odd-numbered orders, and of the same polarity for even-numbered ones. This is because each stage has a –180° shift (same angle as +180°) for frequency components well above the $f_o$ of 232Hz. One rule to remember: After reaching steady-state response, the number of voltage-direction excursions in response to an input step is one more than the number of stages of all-pass.

For example, in the first-order response (one stage), the output initially jumps in the opposite direction to the input, then reverses direction, exponentially decaying to the input voltage level. The higher-order step responses resemble a non-uniformly stretched spring. Listening to electronic pulses (e.g., a 1Hz square wave) through ten stages of all-pass shows a dramatic effect: The sharp input click becomes a downward swooping "teeooup" sound. The instant attack sound of the input is almost completely smeared.

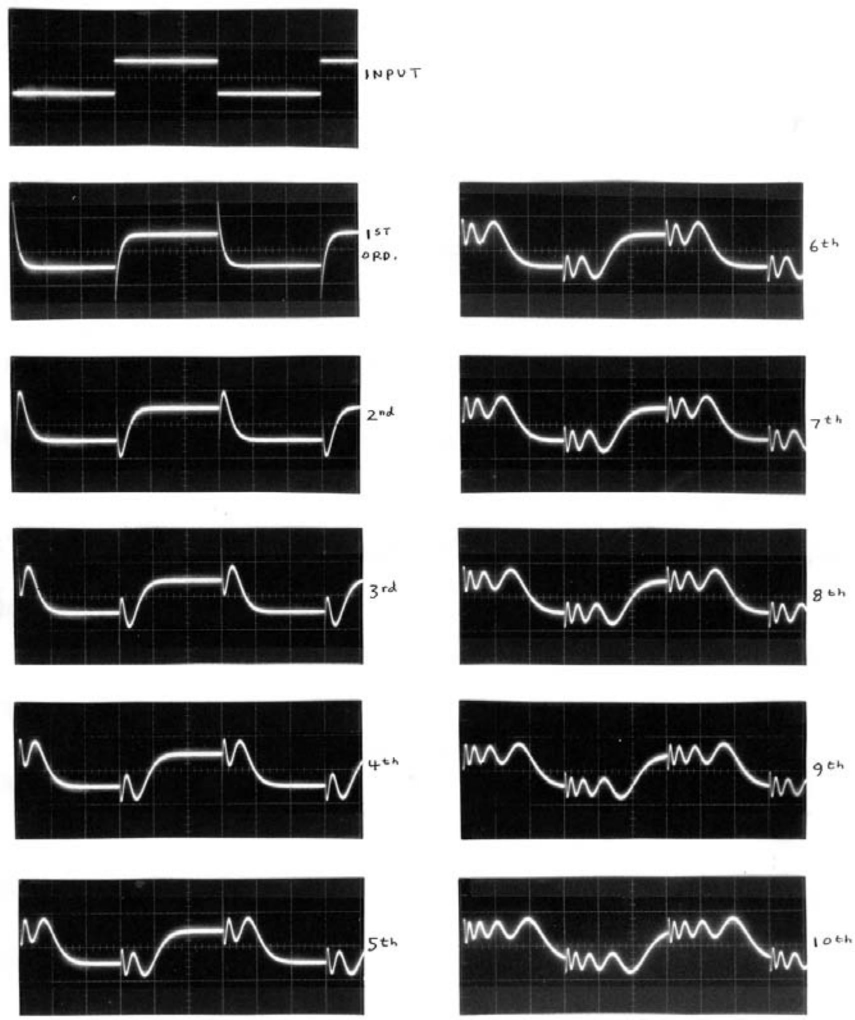


**FIGURE 5: 33Hz square-wave responses, all-pass networks centered on 233Hz. 5mS/div., 1V/div.**

## GROUP DELAY

As Bill Waslo pointed out in his article, group delay doesn't necessarily correspond to actual time delay; nor does a constant group delay ensure distortionless waveform transmission. Constant group delay means a linear phase versus frequency relation. But for waveform reproduction fidelity, this straight phase/frequency line must pass through 0°, or an integral multiple of 180°, at zero frequency (DC).

**Figure 6** is an example of a constant 90° phase shift (Hilbert transform) acting on an ideal (zero rise time) square wave. The phase/frequency relation is linear, so group delay is zero. But because of the constant 90° offset, the square wave is obviously not preserved!

This ideal 90° constant-phase processing is not physically possible in real time: First, note that the curved response must "crystal ball" anticipate the square wave steps. Second, at the input transitions, the output reaches plus and minus infin-

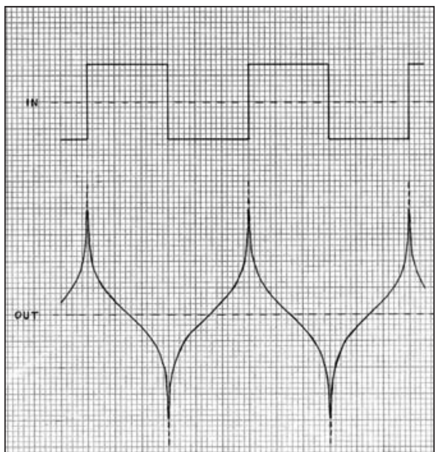


**FIGURE 6: Square wave fed through network having flat amplitude/frequency response but constant 90° phase shift. Network group delay is zero, but waveform is obviously altered. Harmonic amplitudes and RMS voltage are unchanged.**

ity! Yet the output's harmonic amplitudes are identical to those of the square wave. The RMS value is also unchanged.

Using the backwards tape recording method previously described with a phase-shift network, I was able to generate a reasonable approximation of this waveform. It did sound almost the same as the square wave. **Figure 7** shows the effect on a triangle wave.

The point is that constant group delay doesn't guarantee waveform fidelity. However, for the simple all-pass networks used in this article (and for well-aligned speaker COs), the calculated group delay corresponds reasonably well with actual signal delay. Because a crossover of order "n" has the phase characteristic of (n – 1) cascaded all-pass stages, the maximum frequency dispersion of an nth-order CO, with CO frequency of $f_o$, is $T_{max} = (n - 1)/\pi f_o$. For example, a fourth-order CO at 200Hz has a $T_{max}$ = 4.77mS, and a third-order

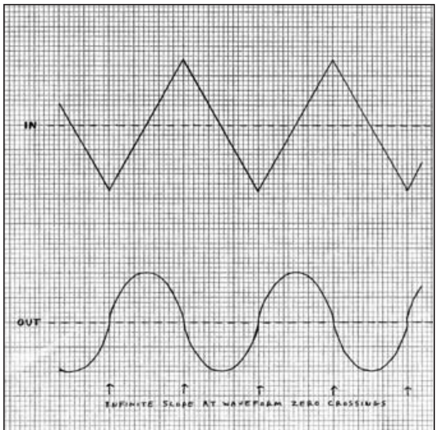


**FIGURE 7: Triangle wave fed through network having flat amplitude/frequency response but constant 90° phase shift. Harmonic amplitudes and RMS voltage are unchanged.**

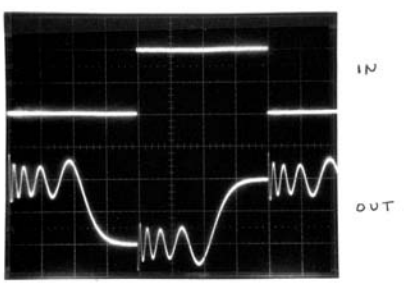


**FIGURE 8: 25Hz square-wave response, ten-stage all-pass centered on 232Hz. 5mS/div., 1V/div. LF group delay = 13.7mS.**

CO at 2.5kHz has a $T_{max}$ = 255µS.

**Figure 8** shows the ten-stage all-pass response to a 25Hz square wave. The initial response transients, of the same polarity as the input steps, are too quick to be visible with the 5mS/div. time scale. Note that the output waveform crosses zero about 12.6mS after the input step. This is close to the network's LF group delay of 13.7mS.

In **Fig. 9** a 25Hz triangle wave was used to drive the ten-stage all-pass. The time scale is 10mS/div. The triangle wave's (odd) harmonics roll off at −12dB/octave, twice the rate of the square wave harmonics. That's why the HF transient response wiggles are much lower in amplitude than with the square wave.

The triangle's harmonics up to the ninth are lower than the 232Hz all-pass center frequency. But even the fifth harmonic is 28dB below the fundamental amplitude. At its frequency of 125Hz, the all-pass group delay has only dropped to 77.5% of the initial (very low frequency) value.

Because of this, the triangle's shape is reasonably reproduced, delayed by close to the network's 13.7mS LF group delay value. This demonstrates how a large number of cascaded all-pass stages can produce a reasonably faithful waveform delay over a useful bandwidth proportional to the number of stages.

## STEEP CROSSOVERS

Some newer speaker designs use very high-order COs, such as sixth and higher. One manufacturer advertised so-called "infinite slope" COs. If taken literally, there would be infinite LF delay; that is, the bass would never come out!

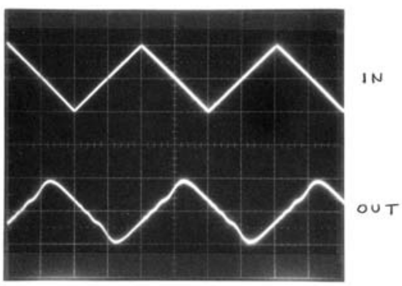


**FIGURE 9: 25Hz triangle wave response, ten-stage all-pass centered on 232Hz. 10mS/div., 1V/div. LF group delay = 13.7mS.**

I think that speaker used transmission zeroes (notch filters) to generate very steep slopes over a very narrow band. In any case, the speaker's transient response should be interesting!

## COMPENSATION OF DELAY DISPERSION

Note: Although I've used non-resonant all-pass responses to represent speaker CO summed response dispersion, the all-pass characteristic of many COs has resonant Q values (greater than 0.5). This is typical of third- and higher-order Butterworth COs, for example. A resonant all-pass Q doesn't (in a well-aligned CO) produce amplitude/frequency errors; it steepens the phase versus frequency slope around the CO frequency. This, for high enough Q values, can increase the group delay around the CO frequency to a greater value than the LF delay.

But regardless of Q value, all second- and higher-order COs (except the special class of transient-perfect COs referred to earlier) produce a delay that drops toward zero at frequencies much above the CO frequency. Therefore, a delay compensator must delay higher frequencies more than lower frequencies, over the full audio band, or at least over the frequency range where delay dispersion is considered to be possibly audible. (For me, this range is about 30Hz to 5kHz.)

This can be accomplished by the use of cascaded resonant all-pass networks,

## AN EXPERIMENT WITH HIGHER FREQUENCY DELAY DISPERSION

Being most familiar with woofer CO effects, I wanted to hear the effects at higher frequencies. So I replaced all ten 10nF caps in the breadboard with 2n2 values. This raised the all-pass phase shift $f_o$ to 1060kHz, while decreasing $T_o$ to 300µS per stage, or 3mS for all ten stages.

Listening to individual step pulses (1Hz square wave) on Sennheiser HD650 phones, I could (barely) hear the effect of even one all-pass stage: a very slight dulling of the sharp clicks. With all ten stages, I could easily hear the blurring into that "teeooup" sound. With square waves around 20-100Hz, the effects were similar, and the "raspy" character was progressively reduced with more stages.

Then I tried an asymmetrical sawtooth wave (containing all harmonics, even and odd, in a –6dB/octave decreasing series). From about 200-500Hz, the sound somewhat resembles a violin tone (but nonresonant and much more "buzzy"). I was surprised by how much the all-pass time dispersion muted the sound, but only at SPLs above about 70dB. The latter confirms that it's the ear's nonlinearity that generates its own added harmonics, and these (as well as whether they reinforce or cancel the source's harmonics) are sensitive to phase shifts (between the source harmonics).

In fact, as Lipshitz and Vanderkooy (in an AES article whose identity I don't remember) stated, simply inverting an asymmetrical waveform, at moderate or higher SPL and on phones or very coherent speakers, can make a distinct difference on its tone. You can easily hear this if you have two sine wave oscillators, an oscilloscope, and a pair of headphones: sum the two oscillators (e.g., with a pair of 10k resistors joined to the scope and headphone amp). Set one to about 200Hz, and the other as close as possible to twice the first one's frequency. Set the octave-higher frequency level to about one-third the amplitude of the lower frequency. Watching the scope, carefully tune one frequency so the waveform becomes almost stationary. Try to tune so the waveform variation repeats about once every two seconds.

Then listen at a fairly loud volume (but not loud enough to distort the phones or amp). You will hear a truly dramatic cycling tonal change, but it's only the phase relation between this synthesized fundamental and second harmonic that's changing! You can very easily rule out possible distortion effects by moving the phones several inches away from your ears—the tonal changes will vanish! ■

with an optimized distribution of $f_o$ and Q values, as Dick Crawford has done. Designing such a compensator requires consulting a filter design book that includes delay equalizers, or else a long time spent doing trial and error experimentation with a good simulation program. (If you do this with actual circuit breadboarding, be prepared to spend a *really* long time!)

## A RESONANT SECOND-ORDER ALL-PASS CIRCUIT

**Figure 10** shows a circuit I built, based on Dick Crawford's article[4]. I cascaded four of these to get an eighth-order resonant all-pass. Resonance (Q greater than 0.5) steepens the phase versus frequency slope, but does not produce an amplitude peak. (It's still an all-pass, meaning flat amplitude versus frequency response.)

Because it's second-order, the circuit of **Fig. 10** has a phase shift that varies from 0° to –360° as frequency varies from zero (DC) to infinity. Of course, -360°, like +360°, is the same (non) angle as 0°. But –360° here means that as frequency is increased from zero, a sine wave is progressively delayed (phase shifted) until, at frequencies much higher than $f_o$ (where the phase is –180°), the wave is delayed by an amount approaching a full cycle.

**Figure 11** shows the phase and group delay (solid curves) of the cascade of four second-order stages. At the $f_o$ of 577Hz, the phase shift is –720°; that is, a 577Hz sine wave is phase-shifted by two full cycles. The group delay starts at 1.556mS (the $T_o$ value) and increases to a peak around 530Hz of about 6.4mS. Then the group delay drops steeply, for example, to 0.332mS at 1.5kHz.

The dashed curve is the group delay of this same eight-stage resonant all-pass additionally cascaded with two stages of the first-order (non-resonant) all-pass circuit previously described. Note how the delay variation from DC to 500Hz is reduced.

The eighth-order resonant all-pass is by no means an optimized compensator for the decreasing delay versus frequency of the two first-order cascaded all-pass stages (the group delay of one of which is shown in **Figs. 2** and **3**). Between about 220 and 520Hz, the resonant

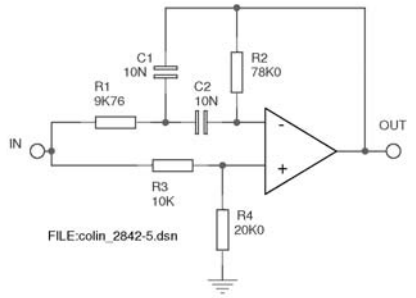


**FIGURE 10: Second-order resonant all-pass. For flat amplitude response:**

$$\frac{R2}{R1} = 4\frac{R4}{R3}, \text{ C1 = C2. Gain} = \frac{R4}{R3+R4}.$$

$f_0 =$

$$\frac{1}{2\pi C\sqrt{(R1R2)}} = 577\text{Hz}, Q = \frac{1}{2}\sqrt{\frac{R2}{R1}} = 1.41.$$

$$\text{Phase} = -2\tan^{-1}\left[\frac{a}{Q\left(1-a^2\right)}\right] \text{ where } a = \frac{f}{f_0}. \text{ Group delay:}$$

$$T = \frac{\frac{1}{Q\pi f_0}\left(1+a^2\right)}{a^4 + a^2\left(\frac{1}{Q^2}-2\right)+1}.$$

**LF Group delay:** $T_0 = \frac{1}{Q\pi f_0} = 0.391\text{mS}$, **gain = 2/3 = -3.52dB. Group delay at** $f_0$: $T_{fo} =$

$$\frac{2Q}{\pi f_0} = 1.556\text{ms}, \frac{Tf_0}{T_0} = 2Q^2 = 3.976.$$

**For four cascaded circuits:** $T_0 = 1.565\text{mS}$, $Tf_0 = 6.22\text{mS}$, **Gain = 0.1975 = -14.09dB.**

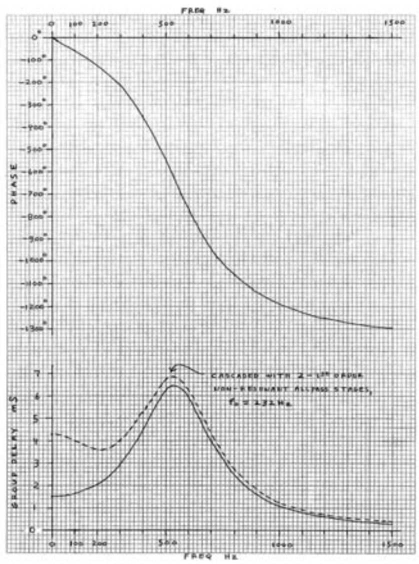

**FIGURE 11: Phase and group delay, four cascaded 2nd order resonant all-pass stages.**

eighth-order all-pass is overcompensating the non-resonant all-pass' downsloping delay.

However, I didn't attempt to design a delay compensator; rather, I simply wanted to be able to view step responses of—and listen to the effects of—a high-order resonant all-pass. But once having done this, I was curious about cascading it with the existing 1-10 stage selectable first-order all-pass breadboard. Despite the overcompensated group delay (dashed curve in **Fig. 11**), using only two first-order stages produced the most symmetrical square-wave response.

## DELAY COMPENSATION EXPERIMENT

**Figure 12** shows the setup. For the test signal, I filtered the input 40Hz square-wave with a second-order low-pass filter, with a Q of 0.61 giving a maximally flat step response (Bessel or linear phase approximation). The filter's $f_o$ of 371Hz is sufficiently below the cascaded all-pass' peak delay frequency of 530Hz, so as to minimize step response wiggles from higher frequencies where the group delay drops.

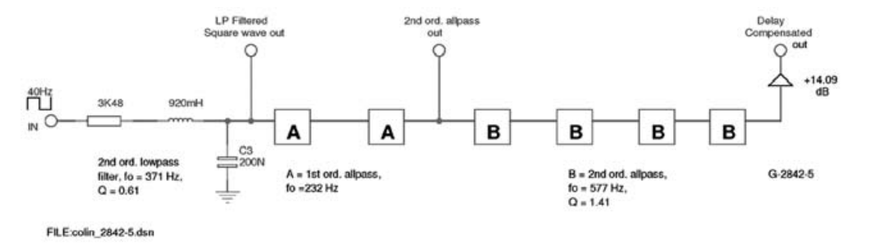


**FIGURE 12: Delay compensation setup.**

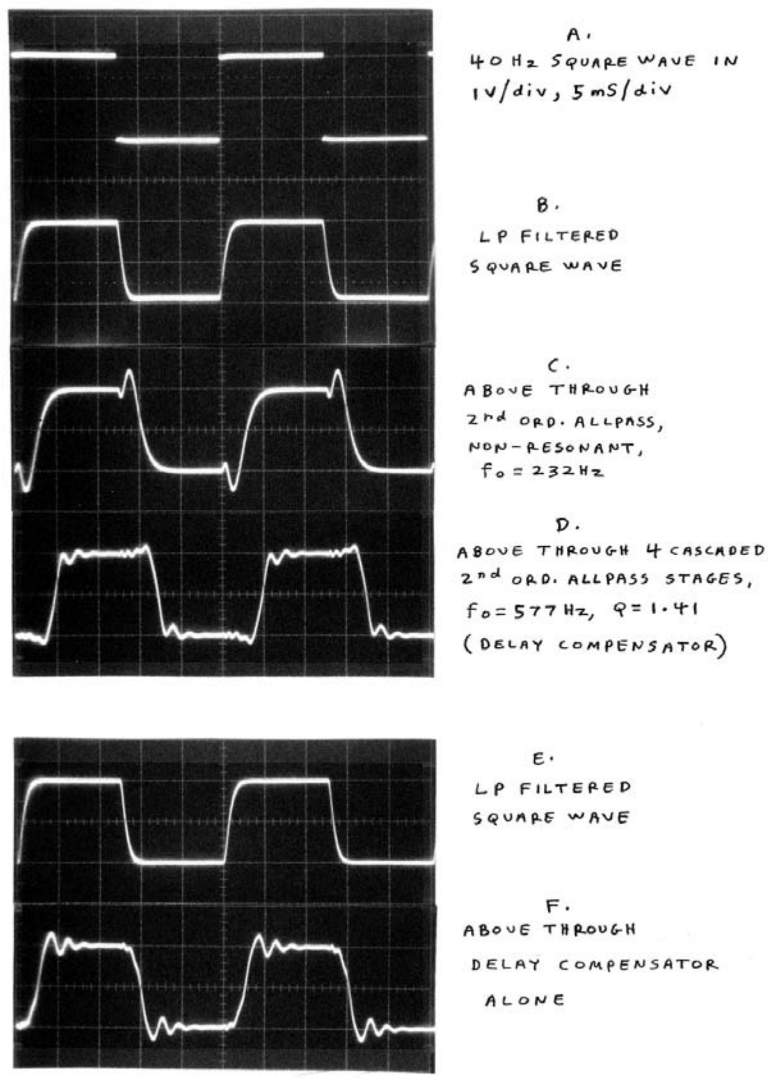


**FIGURE 13: Delay dispersion compensation.**

**Figure 13** shows the results. The filtered square wave in waveform B is the test signal. Waveform C shows the response of the second-order all-pass circuit, similar to a third-order speaker CO. This is then fed through the eighth-order resonant all-pass, whose output is waveform D. Note how, compared to C, the shape is more "square" overall, with steeper sides and reduced overshoot. The small HF wiggles in D are from the crude, far from optimized delay compensation I mentioned. But you can see the basic principle at work here. Note the delay of about 4mS.

Interestingly, selecting only two stages of first-order all-pass produced the best square-wave response, even though using three or four stages produces less overall delay variation from DC to 500Hz. This could be because, as previously mentioned, group delay doesn't always correspond to actual signal delay.

For example, a first-order high-frequency boosting shelf response (such as that produced by speaker diffraction spreading loss) actually has a *negative* group delay around the center of the upward slope region. If this corresponded to actual signal delay (negative), your speaker would produce music *before* the input signal arrived! Just think: Enough of this could allow you to hear tomorrow's winning lottery number today! OK, back to the real world.

Note that waveform D is more symmetrical than B. The extra HF delay compensation of the eighth-order resonant all-pass appears to be compensating the LF (re HF) delay dispersion of the low-pass filter used in the test signal.

Waveform F in **Fig. 13** is the response to the filtered square wave (E) of the eighth-order resonant all-pass by itself (without the second-order non-resonant all-pass). Note that the lower frequency filtered step excursion responds before the most prominent higher frequency wiggles. This is from the rising delay curve in **Fig. 11**, the lower solid curve from DC to 530Hz. **Figure 14** shows this same effect with a 50Hz duty cycle offset sine wave from an old HP8116A function generator.

## RESPONSE TO UNFILTERED SQUARE WAVE

In **Fig. 15**, waveform A is a 40Hz square wave, and B is the response from the eighth-order resonant all-pass. C is the response through the combination of the eighth-order resonant and second-order non-resonant networks. B and C display the group delay variations shown in **Fig. 11**, bottom, the solid and dashed curves, respectively. In waveforms B and C, the first arrival is the faint short spike, containing the highest step pulse frequencies that are delayed very little.

In C, you can see this spike "springing" into ripples of decreasing frequency. This represents the increasing delay as the frequency in **Fig. 11** decreases from above 1.5kHz to 530Hz. Then (back to **Fig. 15C**), the signal makes its large step excursion toward the opposite square-wave polarity. Notice that the delay from the initial spike to the step excursion midpoint (at zero volts) is about 4mS. In **Fig. 11**, dashed curve, the LF group delay (from DC to about 300Hz) is close to this 4mS value.

From about 5 to 7mS after the initial spike, you see another set of prominent wiggles (after each step transition). This corresponds to the 380 – 660Hz region in **Fig. 11** (dashed curve), where the group delay ranges from 5mS to the 6.8mS peak value.

## SOUNDCHECK

Listening to low-frequency step pulses

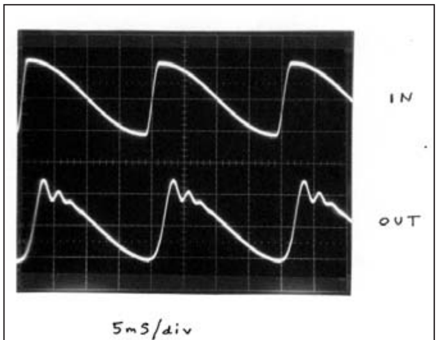


**FIGURE 14: 50Hz 10% duty pulse FM sine wave through four second-order resonant all-pass stages, $f_o$ = 577Hz, Q = 1.41. LF group delay = 1.565mS, 577Hz group delay = 6.22mS.**

(1Hz square wave) through the eighth-order resonant all-pass, I could easily hear the effect. Unlike the cascaded first-order non-resonant all-pass' downward sweeping "teeooup" sound, the resonant all-pass turned the sharp 1Hz square-wave's clicks into a sort of "chuwik" sound. Of course, lasting only 6.4mS, the sound was much quicker than trying to say "chuwik," but this invented word describes the effect of the group delay dispersion, where I could hear the sound start with the highest frequencies, which quickly swept downward, and then partway back up. It sounded like some kind of quick bird chirp.

If you think this analysis is "for the birds," remember that they're probably the original inventors of music!

## TRIANGLE WAVE RESPONSE

Because the triangle wave has much lower harmonic amplitudes than the square wave, the 50Hz triangle response in **Fig. 15E** has much lower HF ripple amplitude. The cascade of both all-pass networks then displays a reasonably clean waveform delay of about 4mS, which agrees with the LF group delay value.

## AUDIBILITY ISSUES

1. Delay dispersion itself, heard on a single driver (e.g., headphones) is probably not audible below 1-2mS. One exception, though, is with loud steady-state tones having a coherent but low-order harmonic spectrum, such as flute and organ tones. Especially with asymmetric waveforms having even harmonics, you can hear phase shifts and even simple polarity inversion. This is because of the ear's asymmetric nonlinearity—about 1% at 90dB SPL below 1kHz, and much greater for louder and lower frequencies. Phase shifts including inversion can change the ear's own added harmonics and IM tones. Furthermore, the auditory nerve cells fire predominantly on negative waveform half-cycles, at least for the lower frequencies.
2. Spatial driver separation enhances the audibility of delay dispersion, because in a fast, wideband transient the higher frequencies come out of, say, the tweeter first; then the delayed lower frequencies come out of a different position, say, the woofer. A fast, swooping motion can be sensed. And even if the two drivers are close compared to the listener's distance, inter-driver phase shifts enhance the sense of separation. This is similar to left-right stereo phase differences near 180° producing a sound stage perception extending beyond the speaker positions (used in some boom boxes and TVs with built-in closely spaced speakers for "enhanced stereo").
3. Delay dispersion can increase or decrease a waveform's peak amplitudes (while not affecting the RMS levels). Note how even the first-order all-pass in **Fig. 4** (second photo from top) *triples* the square waves's peaks from ±0.5V to ±1.5V. Some musical waveforms can have their peaks reduced, depending on the time and spectral structure.

As mentioned, the ear's nonlinearity, especially for loud sounds below 1kHz, is sensitive to waveform peaks. Therefore, delay dispersion can affect the impact and realism of drum strikes, string attacks, and other powerful transients, even if the delay dispersion itself is inaudible.

These effects can be quickly revealed with the pulses from the Sound Strobe, reviewed by Ed Simon in *aX* April '07. Because lower frequency COs have higher time delay dispersion (it's inversely proportional to frequency), a fourth-order woofer/mid CO, having (at best) the step response shown in **Fig. 4** (lower photo), could easily produce audible degradation. With a CO frequency of 200Hz, the

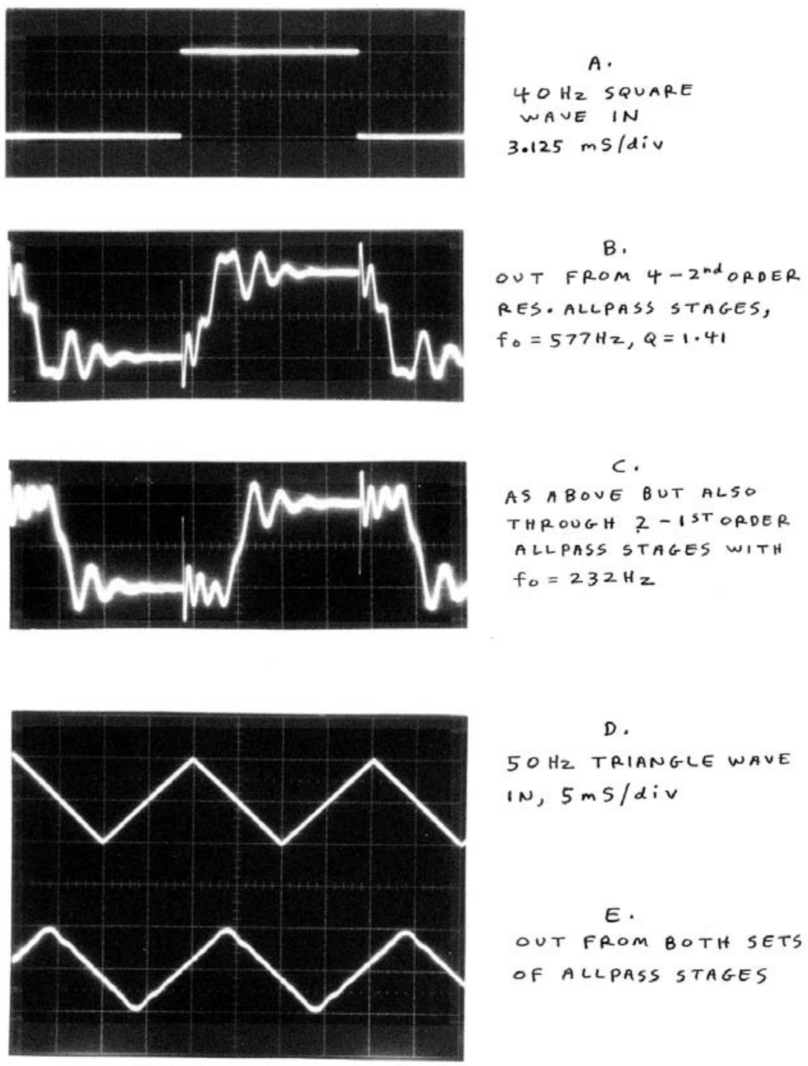


**FIGURE 15: Various all-pass responses.**

woofer output is delayed 4.8mS relative to the midrange, not counting extra woofer time lag. This is equivalent to a 5.4´ path length difference.

Now consider the effect on a musical transient of this and a woofer/midrange spacing of 2´: A live bass string plucking transient emanates from a few inches of string and soundboard. (The rest of the string's length and the soundboard area produce delayed resonances, but the sharp initial transient radiates from a small area.) But from the speaker, the transient is frequency smeared over a 2´ high by 5´ deep area! I have no problem hearing this effect even with a second-order woofer/mid CO, in comparison with a waveform phase-preserving first-order CO.

For example, my two best speaker systems are the Swans M1 with Scan-Speak sub (*aX* Sept. '05)[5] using a second-order CO, and the Venue Speaker (*aX* Nov. '06)[6] with first-order woofer CO. Both have a very flat in-room response down to about 30Hz. But the attack transients of electric or acoustic bass, low piano notes, drums, and low-pitched guitar notes sound naturally precise (clean "bite" or "snap") on the first-order Venue. On the second-order Swans/Scan-Speak, these attack transients still sound clean, but not quite as naturally "biting." When I first tried a third-order woofer CO on the Venue proto, bass and drum transients were "recessed behind the scene," sounding "not quite there." On p. 20, Fig. 15 of the Venue article, note the clean 80Hz square wave.

## MIDRANGE/TWEETER CROSSOVERS

In the Venue article, Fig. 18 (p. 21) shows the step response on a 200µS/div. time scale. The response resembles **Fig. 4**, second-order all-pass. A second-order all-pass is characteristic of a third-order CO. While the Venue mid/tweeter CO is *electrically* second-order, the *net acoustic* CO is close to third-order.

**Figure 18** in the Venue article shows about a 200µS delay from the initial tweeter spike to the point where the midrange response reaches the exponential waveform rise time level (63% of steady-state). This agrees with Joe D'Appolito's excess group delay maximum of 200µS (Venue article, Fig. 14, p. 19).

My threshold for delay dispersion is about 1mS, so I can't hear the 200µS mid/tweeter time spread of the Venue. What I found to be important in this CO frequency range (2.5kHz in the Venue) is to have the two drivers' outputs in phase with each other, around the CO frequency. This is a different thing from absolute waveform phase accuracy. Minimizing inter-driver phase differences at typical tweeter CO frequencies is important for spatial image coherence and sharp focus.

To my ears, a well-designed D'Appolito (MTM) configuration can sound virtually transient-perfect, even with (again, well-designed) high-order COs. I think this is because the symmetrical MTM arrangement makes the midrange and treble acoustic centers spatially coincident. This eliminates the delay-dispersion audibility enhancement with separated drivers that I discussed. The best example I've heard is Joe D'Appolito's Thor speaker[7].

## WHAT I HEARD WITH THE TEN-STAGE ALL-PASS

With the sawtooth wave around 200Hz, switching in the all-pass *dramatically* (and I mean that literally) muffled the tone, all but removing the buzzy edge sound. But when I moved the phones 6˝ away from my ears (greatly lowering the SPL but not changing anything in the phones or amp), lo and behold the buzziness returned! Switching the all-pass in and out made no difference.

## EFFECT ON MUSIC

This confirmed what I had thought: Although I couldn't directly hear the 3mS time dispersion strongly (except with xylophones and some drums), I easily heard the ear's nonlinear reaction to phase shifts. The entire musical sound became "covered," "veiled," muted, dulled, and de-focused. With this higher $f_o$ of 1060Hz, I needed three all-pass stages to notice the effect. That's a delay dispersion of 1mS, about the same as my threshold with the lower fo of 232Hz (where one stage has a $T_o$ of 1.37mS).

All of this establishes (for me) that the phase nonlinearity (time dispersion) of most midrange/tweeter COs is inaudible, while I can hear the bass transient blurring of woofer COs from 100-300Hz if second- or higher-order.

## CONCLUSION

I hope this article has provided some insight into the phenomenon of time dispersion, particularly regarding its audibility (or lack thereof) in speaker crossover phase nonlinearity. *aX*

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# A Single-Ended Class A Hybrid Amp, Part 2

Part 1 looked at a MOSFET single-ended unity-gain source-follower amplifier output stage. This follow-up adds an appropriate gain stage.

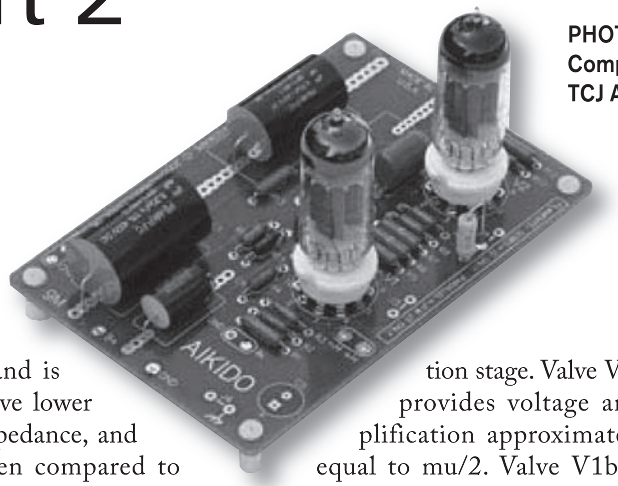


**PHOTO 1: Completed TCJ Aikido.**

Designing a power amplifier requires a wholistic approach. The output stage needs to be delicately coupled with the input stage for best results.

To review, you should choose each component for optimal linearity and low noise. The output stage features a MOSFET, which is great for handling high current, but questionable for voltage gain. The design used the MOSFET as a source follower to maximize inherent local feedback and minimize the effects of gate capacitance. The fewer components, the easier it is to concentrate on getting each one right. The output stage has only one active device to worry about. I played a couple tricks (capacitance multiplier and Pass Aleph current source) to earn some bonus points.

The input stage will follow a similar philosophy. The vacuum tube can easily handle the required voltage swings (±20V) to drive the source follower and is well-regarded in audio circles. The triode has outstanding built-in inherent local feedback and proven linearity. It doesn't work so well with high currents, which is why valve amplifiers have output transformers, but this won't be a problem as an input stage.

## INPUT STAGE

The triode features a high degree of inherent local feedback. As the grid voltage increases, the plate current also increases, causing the plate voltage to drop from where it would have been (thereby also inverting the signal). This is negative feedback and is why triodes usually have lower gain, lower output impedance, and superior linearity when compared to pentodes, which defeat the inherent local feedback by inserting extra grids to decouple the plate current and voltage. Pentodes have much higher gain (less is lost to local feedback) and higher output impedance.

Because linearity and low noise are valued, the John Broskie Aikido input stage is an excellent choice (**Fig. 1**). I have named all of the component designators to match the Tube Cad Journal (TCJ) printed circuit board.

The Aikido consists of two legs that function (mostly) independently. The first leg (**Fig. 2**) is the voltage amplification stage. Valve V1a provides voltage amplification approximately equal to mu/2. Valve V1b is an active plate load for V1a. It loads V1a with a copy of itself, so the active load reflects the nonlinearities of the tube being loaded. It is convenient if V1a and V1b are two identical triodes packaged in a common envelope (dual triode).

If you use identical triodes, V1a and V1b will be balanced, and the first leg's output will rest at B+/2. This also means half of any power supply noise will be seen on the output.

Cathode resistors R2 and R4 should also be identical to maintain vertical symmetry. Set these depending on tube

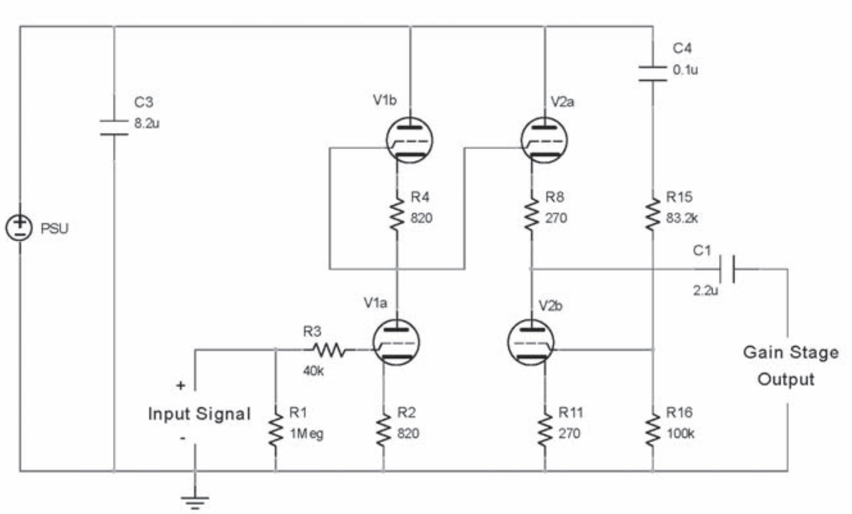


**FIGURE 1: Tube Cad Journal Aikido gain stage.**

type. I chose the Electro-Harmonix 6CG7EH, which is essentially the venerable 6SN7, but in a 9-pin package. With a 300V power supply and 330Ω cathode resistors, the tubes run at about 9mA current. I have used everything from 270Ω to 870Ω resistors with good sonic results.

This tube has mu = 20, which results in a gain of about 10. The MOSFET output stage is capable of peak-to-peak ±18V (assuming 50V PSU). My signal sources are Musical Fidelity and Theta DACs, which easily output 2V p-p. Therefore, I need a gain of only about 10 to push the output stage to full power.

If you need more gain, try the 6DJ8 (750Ω cathode resistor for 6mA and gain of 14), although audiophiles will argue about sound quality of the type. For 12V heaters, try the 12AU7 (gain of 9), 12FQ7 (gain of 10), or 12AT7 (gain of 20). Others may choose to try the octal variant, using the 6SN7 (recommended).

The first leg's output is DC coupled (no coupling capacitor) into a second leg (**Fig. 3**), which is a unity-gain (actually, a little less than unity) cathode follower (V2a) loaded by an active cathode resistor (V2b). Again, the active load V2b loads the cathode follower V2a with a copy of itself, reflecting nonlinearities. The cathode follower takes great advantage of the triode's inherent degenerative local feedback (therefore, unity gain). With a twin triode, V2a and V2b will be balanced, and the output will rest at B+/2.

Here's the Aikido magic: Any power supply noise is AC coupled through C4 to the grid of V2b. Resistor divider R15 and R16 adjust the noise injection ratio. V2b inverts the PSU noise and modulates V2a bias current, effectively canceling out any PSU noise that would have otherwise been on the output. It actually works! It is dead quiet. You can calculate the proper relationship with R15 = R16 × (mu - 2)/(mu + 2).

Cathode resistors R8 and R11 should be identical to maintain vertical symmetry between V2a and V2b. I also chose the 6CG7EH for the second leg. The important parameter is output impedance, which is roughly 1/gm + R8. With 300Ω, the 6CG7EH has about 500Ω output impedance, which is low enough to drive the beefiest MOSFET in the output stage, even big TO-264 monsters.

Along with the active devices (V1, V2, and Q1), the Aikido coupling capacitor C1 is sonically important. I believe your money is best spent on active devices (tubes), before capacitors. Choose C1 with whatever budget you have left over after buying your tubes, or save C1 as an

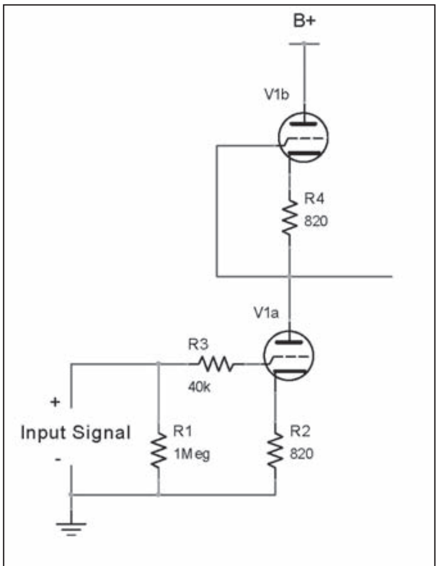

FIGURE 2: First leg.

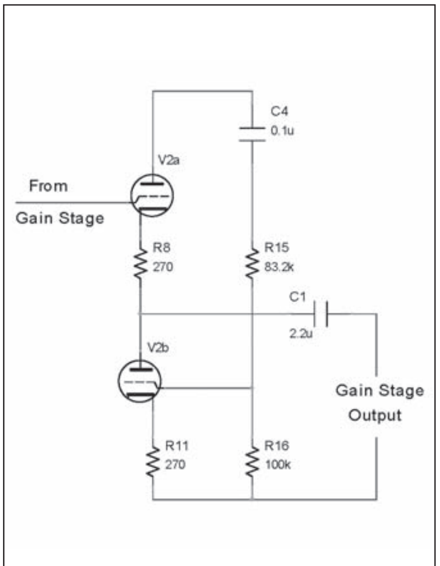

FIGURE 3: Second leg.

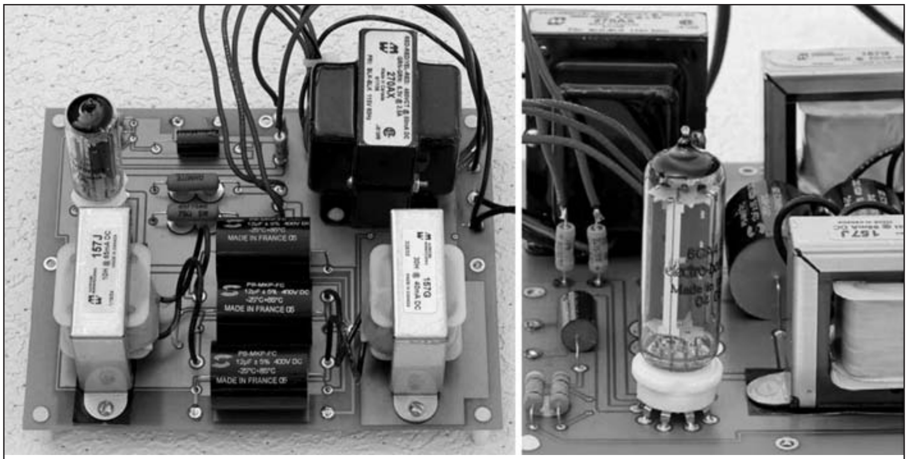

**PHOTO 2: Assembled 300V B+ PSU.**

upgrade. I used a Solen 2.2µF polypropylene.

Capacitor C4 is also in the audio signal path (it influences the grid of V2b). I chose the Dayton film and foil polypropylene 0.1µF. Make sure all voltage ratings exceed your power supply's maximum (unloaded) output.

With octals (8-pin), I'd use the 6SN7. You can try a huge variety of tubes here, making sure the pinouts and heater voltages are correct. With 9-pin minis and 6V heaters, some options are 6AQ8, 6BQ7, 6DJ8, 6CG7, 6FQ7, 6GM8, 6H30. . . the list goes on.

Extra resistors R5, R6, R7, and R12 help keep each leg's output centered on B+/2. These have a number of interesting advantages, including preventing damage if powering up with one socket empty or if a tube goes bad. The resistors maintain the proper B+/2 voltages, even with the heaters cold! These resistors are 1MΩ, so they have negligible effects on the circuit's sonic performance.

You can get John Broskie's Tube Cad Journal Aikido PCB at www.tubecad.com. You can purchase Aikido boards for both octals (8-pin) or mini (9-pin) valves.

## 300V B+ PSU

Be forewarned: High voltages are dangerous. The audio world needs to preserve the few brave pioneers who read about and want to build amplifiers like this. Acknowledge the dangers of working with high voltages, and do a favor to all of us: we need you. If you are unpracticed with high voltages, seek assistance.

I prefer the nature of a valve rectified, choke filtered B+ power supply. I chose the 6CA4/EZ81 rectifier because it is not grossly oversized for the application and uses a standard 6.3V heater. It is available new from Electro-Harmonix or JJ Electronics (among others). The power supply can be as sophisticated or as simple as desired. I experimented with some regulators, but found the Aikido topology is incredibly resilient to power supply issues. Heroic efforts simply aren't needed.

The schematic is shown in **Fig. 5**. The Hammond 270AX power transformer provides 50mA of 480V AC with center tap. Use a dedicated fuse for the AC feed to the power transformer. R18 and R19 are plate resistors for the rectifier. Voltage after rectification is about 325V DC (unloaded), filtered by two stages of pi filters.

Capacitors should be C9 ≥ C10 ≥ C11, or the pi filters will oscillate during the turn-on process as the capacitors charge. The first choke is 10H, rated at 65mA to handle peak currents. The second choke, being downstream and seeing smoother current, is rated at 30H and 40mA. I chose Solen 400V 12µF capacitors throughout (C9 = C10 = C11).

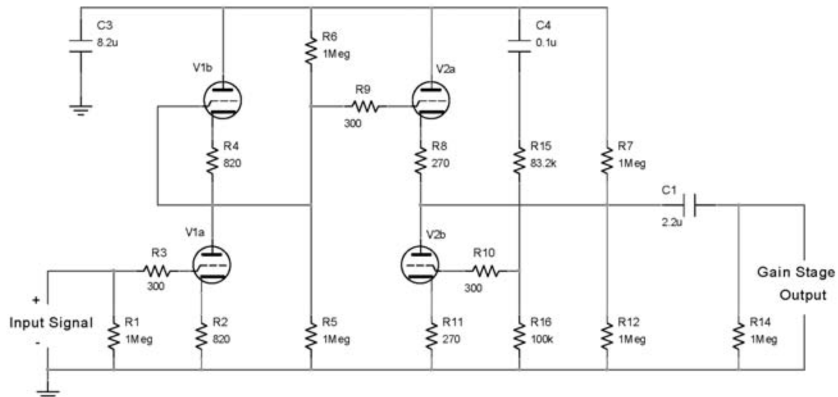


**FIGURE 4: Full TCJ Aikido.**

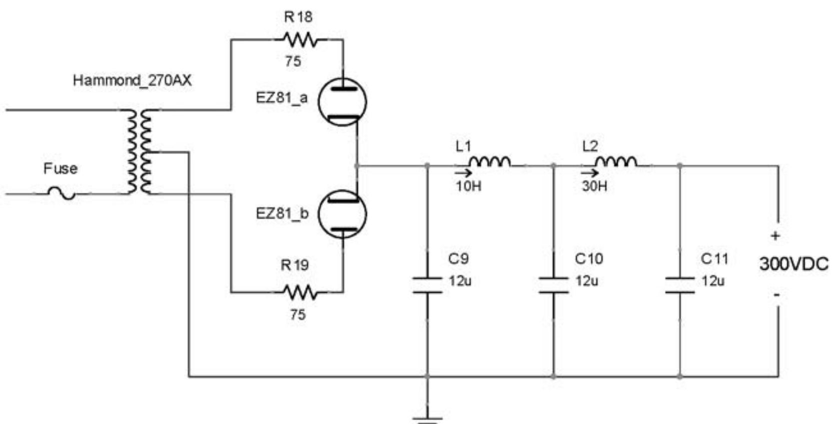


**FIGURE 5: 300V B+ PSU.**

Not wanting a rat's nest of high voltage spaghetti, I put the power supply on a printed circuit board (**Photo 2**). I also placed some extra solder pads for the option of replacing C9 and C10 with Mundorf 33μF Tubecap high-value electrolytics, but I've been quite happy with the 12μF polypropylene Solens (although, this doesn't seem to be enough capacitance, especially when I'm accustomed to 10,000μF or more in solid-state designs).

The 6.3V secondary on the 270AX runs slightly high, so two 0.75Ω power resistors buffer the heater. This also has the benefit of slowing down the heater warmup (inrush current and thermal shock), prolonging the life of the rectifier.

This power supply has plenty of juice (40mA, limited by my selection of L2) for ambitious experimentation, and can be used for a variety of preamp designs. You can easily obtain different voltages by changing the 270AX with an alternate Hammond model.

## HEATER PSU

I used a separate transformer for the Aikido's heaters. The Hammond 270AX in the B+ power supply didn't have enough current for two 6CG7s heaters. At first, I used a 6V switching power supply brick. Unfortunately, you could hear the high-frequency switching noise. Ultimately, I achieved the best sound with a DC power supply (**Fig. 6**). The constant-current feature is probably overkill, but I wanted to try it.

The heater supply provides 600mA at 12.6V for the two Aikido 6CG7 triodes (filaments connected in series). The LM317 is connected as a current regulator, and tries to maintain a 1.25V drop across resistor R20 (1.25V/2.1Ω = 0.6A). To compensate for the voltage drop across the LM317 and R20, the raw voltage (before the LM317) needs to be 18V.

The Hammond 161G28 provides 14V AC with the secondaries in parallel, which is rectified and filtered with capacitor C12 to about 19V. Resistor R20 dissipates 0.75W (1.25V × 0.6A), and the LM317 dissipates about 3.5W. Heatsink appropriately. As a matter of practice, I include a fuse for the transformer. I also use an NTC on all high-fidelity power supply circuits that directly feed filter capacitors. The TR5 fuse is the little red button behind the transformer in **Photo 3**.

## ASSEMBLING THE AMPLIFIER

Now you have a unity-gain, single-ended MOSFET output stage and an ultra low-noise, high-linearity valve gain stage. The output stage was designed to use the MOSFET's inherent local degenerative feedback to eliminate the dreaded Miller capacitance, making it much easier to drive. The gain stage was engineered around triodes to get linearity, low noise, and the correct amount of gain. The triode's output impedance can easily drive the MOSFET in this configuration.

The MOSFET output stage has an output impedance of 0.25Ω, being dominated by the 0.22Ω sense resistor for the Pass Aleph circuit. Removing the resistor and disabling the Aleph drops the output impedance to 0.1Ω. For my Triangle Acoustics Solis loudspeakers with a nominal impedance of 6Ω, that translates into a damping factor of 32 with the Aleph and 80 without. I happily trade damping factor for the ben-

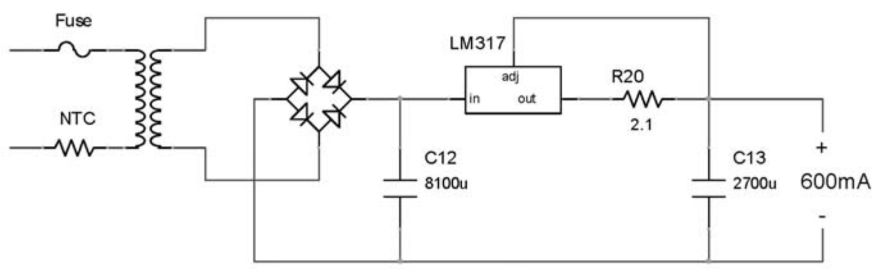


**FIGURE 6: Heater PSU.**

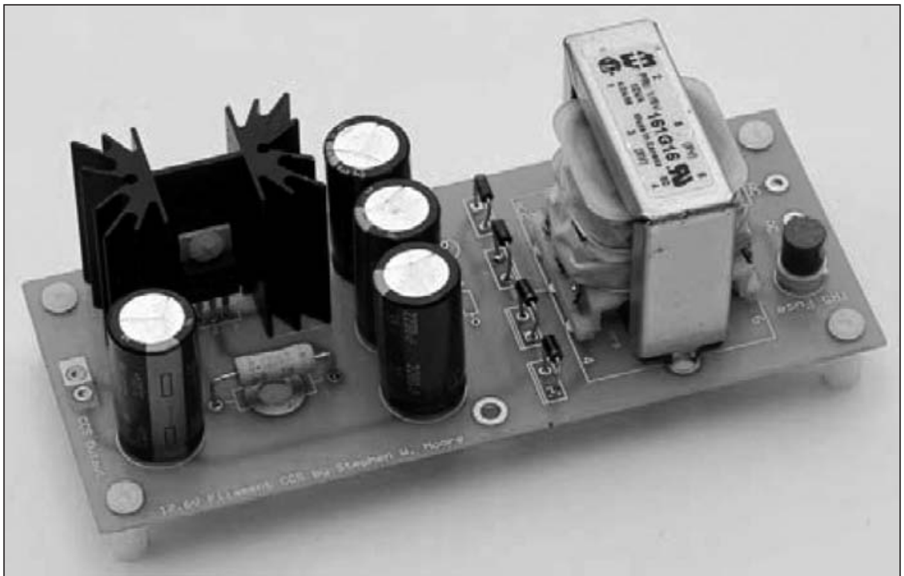

**PHOTO 3: Heater PSU.**

efits of Pass's Aleph (Pass discusses the Aleph's benefits in Zen Variations Part 4, Dec. '02, p. 16). I have also found my Solis start to sound dry and lose soundstaging with too high damping factors. A side benefit of the Aleph is it increases output power. I highly recommend auditioning some of the Pass Labs amplifiers, because you'll learn a lot from listening to the fruits of his labor.

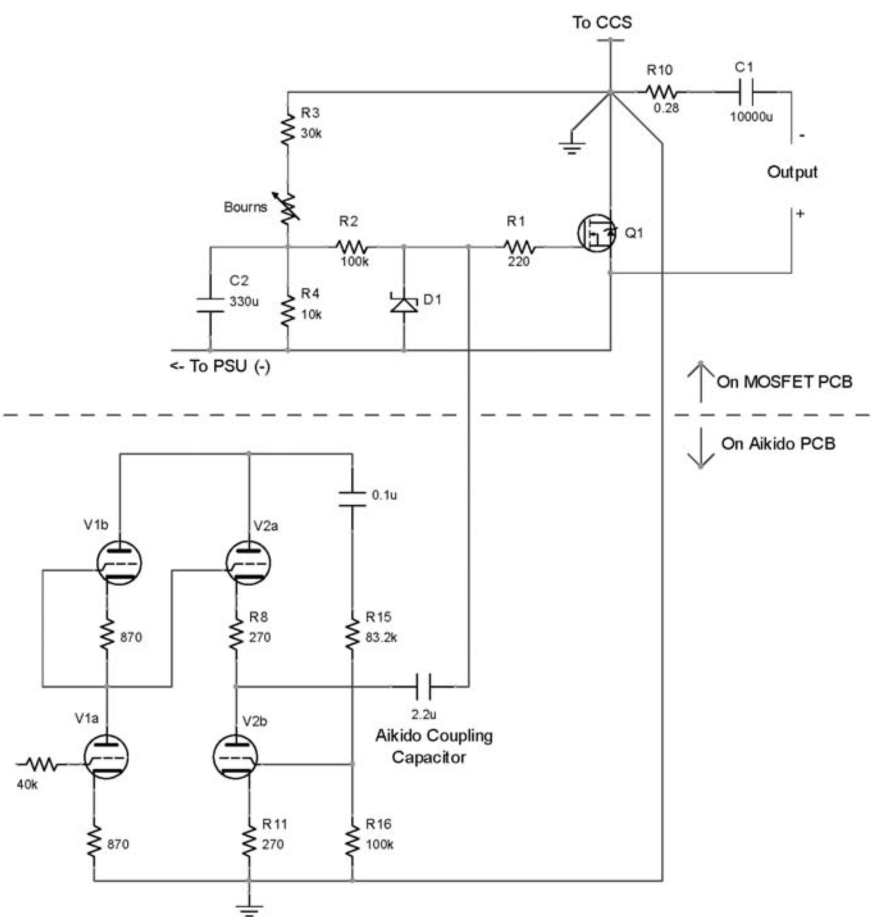


**FIGURE 7: Connecting the Aikido to the source-follower MOSFET.**

The Aikido is coupled to the MOSFET with capacitor C1 on the Aikido PCB. With the IRFP044, you can achieve good bandwidth with a 0.47µF or larger capacitor. I used a Solen 2.2µF polypropylene, and I am happy with the result. Audiophiles can use paper-in-oil, beeswax, or pizza grease (John Broskie recommends the wet slug approach).

As a hat-tip to the Nelson Pass Zen amplifiers and the John Broskie Aikido valve stage, this amplifier has been dubbed the ZenKido. Full output power into 8Ω is 21V p-p for about 27W. For lower impedances, the output stage runs out of current. For higher impedances, the amplifier runs out of voltage. For example, maximum output into 6Ω is 16V p-p for 21W, and into 10Ω is 22V for 24W.

Some could claim that the coupling capacitors may dominate the amplifier's sonic signature. I find it is actually a combination of V1, V2, and Q1. As Pass pointed out in his article "Practical MOSFET Testing for Audio" (Jan. '04, p. 8), the dominant sonic signature is almost always the active device. Spend more time and money considering the valves than considering C1. For Q1, the IRFP044N is a very good choice, and is also cheap. After choosing the valves, the choice of C1 becomes the next priority.

There are many parts to pay attention to: the output stage power supply, the source follower PCB, the B+ power supply, the Aikido PCB, and the heater supply. Even a "simple" amplifier is a very complex affair!

## LISTENING

The ZenKido was designed around a low-noise, low-distortion voltage amplification stage combined with an easy-to-drive, single-ended, unity-gain, Zen-like output stage.

I found that matching between the amplifier and loudspeaker is quite important, contrary to what some noted contributors to this magazine (such as Doug Self) have theorized. My Solis loudspeakers tend to "dry up" and lose soundstaging with highly load-invariant amplifiers. The sound becomes less engaging and more fatiguing. I've also noted the same effect with several other loudspeaker brands. Finding the right damping factor for your particular loudspeaker appears to be important.

This flies in the face of the notion that loudspeakers are the dominant contributors to the sonic character in the audio chain. I can't subscribe to that theory anymore. A "straight wire with gain" may not be the best solution for you. (You could also argue this is the loudspeaker's fault, but there is no perfect loudspeaker, just as there is no perfect amplifier.)

The Aikido gain stage imparts sophisticated and refined dynamics. The ultra low noise performance is very nice, especially with my relatively sensitive Triangle Acoustics Solis loudspeakers. The source-follower MOSFET output stage in Part 1 imparts good tonality and a lack of offensive aggressiveness. The combination of the two delivers an overall satisfying combination. The only thing this amplifier lacks is high output power, so 90dB/watt speakers (or better) will be needed to compensate. Look for flat impedance curves, which help maximize this amp's potential, especially if the speaker lacks big capacitive phase angles that correlate with low impedance.

This amp loves a tube-friendly loudspeaker. I've tried this amplifier with a variety of speakers, including many DIY creations, and some from Fostex, Paradigm, Definitive Technology, Altec Lansing, and Infinity.

My listening tests validate the hypothesis: tube-friendly loudspeakers are the best fit. I plowed through my music collection, listening for accuracy and soundstaging. It's very tempting to play this amplifier loud, reveling in otherwise hidden nuances and subtleties. Solo vocalists are a joy, due to the soundstaging, profound tonality, and lack of aggressiveness. I have found some large symphonic pieces to get lost in a muddle, presumably because the amplifier runs out of steam at loud listening levels or compression due to loudspeaker capacitive (negative) phase angles, which this amplifier has trouble dealing with.

Simply put, this amplifier likes small groupings of musicians. If you listen to a lot of contemporary vocals (Mark Knopfler, Nora Jones, Alison Krauss), jazz, blues (Jimmy Rodgers, Diana Ross), fusion (Bela Fleck), country (Johnny Cash, Randy Travis, Merle Haggard), bluegrass (Ricky Skaggs), gospel, big band, folk, new age (Enya, Enigma), or small orchestral/chamber music, then this amp is worth a try. If your bag is the classic large-format symphony, amplifier/speaker matching will be challenging.

That being said, when my system (including this amplifier) is combined appropriately with DIY Linkwitz dipole woofers, I've never heard the massive St. Eustache organ for Ceasar Frank's Piece Heroique (Jean Guillou on Dorian label) replicated quite so. I've heard the organ live numerous times, so I have a good feel for the damp, earthy tone the recording should reveal. This was a fantastic match.

I'm enjoying the 6CG7EH (9-pin) in the Aikido. I expect the 6SN7 (octal) should also sound very good. This is a tube-roller's paradise, and along with the Aikido coupling capacitor C1 and MOSFET Q1, there's plenty of room for experimentation. *aX*

# **Sounds** and Hearing, Pt. 2

We often take our hearing for granted. Find out how the ear works and what we can do to keep it functioning properly.

The idea of a mechanism that took hundreds of millions of years to develop is staggering. It was accomplished so slowly that you cannot observe any changes in your lifetime. In its present-day form, the ear, with its resulting construction, is so complex as to be overwhelming. To make matters more complicated, medical terminology can make a somewhat simple understanding even more intimidating.

## THE EAR: A BIOLOGICAL TRANSDUCER

The hearing mechanism is very elaborate, and the brain's processing of information is more like a science-fiction story with an unfinished ending. Although volumes have been written about the hearing process, perhaps a few simpler explanations may help to understand some of the intricacies of the ear and how you hear.

The brain has a sound memory center that begins accumulating sounds almost at birth, perhaps even before. In normal adults, the brain has been able to establish a library of about 400,000 sound patterns to relate to the outside world. You can play tunes any time from your mental library, even complete symphonies with all the instruments playing together. In a similar way, nerve impulses from other sense organs, such as visual images or impulses from other parts of the body, can also be stored for comparison with new sensory input.

The ear is an incredibly sensitive device. The threshold of audibility corresponds to a pressure variation of less than one billionth of one atmosphere. This is commonly referenced as 0dB sound pressure level at 1000Hz, but varies with frequency.

At the other extreme, +130dB is considered the threshold of pain, corresponding to a voltage ratio of more than 3,000,000 to 1 (some sources quote +120dB as the pain threshold). Even a loud noise only causes microscopic movements of the eardrum. For a high-frequency sound, the motion may be only one-tenth of the diameter of a hydrogen molecule[8]. Hydrogen is in a diatomic form.

Although the ear magnifies a wide range of sound intensities, the transmitting equipment is too stiff to respond to the very weakest tones, and thus they are not heard. If the range were not limited, you would be assailed by such sounds as your own body's muscle contractions and bone movements.

A microphone can be called a transducer because it transforms energy from an acoustical system through a diaphragm to a mechanical system that is then converted to an electrical system. The ear functions as a biological transducer.

**Figure 9** shows a cross-section of the ear[9]. There are three main sections: the outer, middle, and inner ear. The outer and middle ear function as an impedance transformer, converting energy from the low impedance of the air to the high impedance of the fluid in the cochlea.

Acoustic resonances in the auditory canal of the outer ear can double the sound vibration force. The mechani-

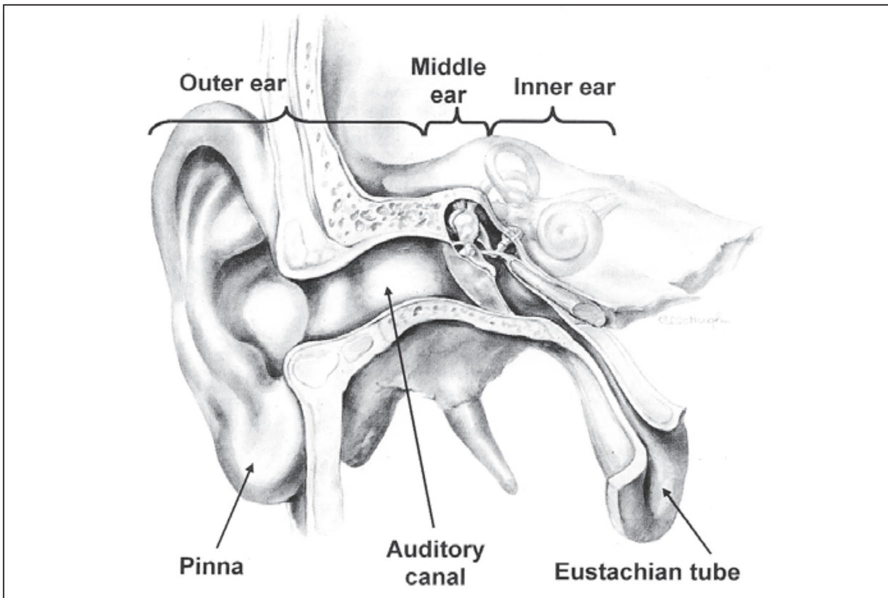


**FIGURE 9: Cross-section of the ear.**

cal advantage of the bone-lever system of the middle ear can triple it. Pressure then transmitted to the cochlea at the inner ear can increase it 30 times. The total result can be an amplification of up to 180 times before a sound wave sets the fluid of the inner ear in motion.

The principal parts of the ear can be seen in the three-dimensional model of **Photo 3**[10]. The sensitive parts of the inner ear are well protected inside the skull bone, shown as the porous structure. Sound travels from the auditory canal at the bottom left and ends at the eardrum near the bottom center. This converts the acoustic energy to mechanical energy that is transferred from the eardrum to the mechanical system of the middle ear consisting of three bones—the hammer, anvil, and stirrup (known collectively as the ossicles). These are shown just above the eardrum and are the smallest bones in the human body.

You can see the stirrup as a U-shaped bone at the center of the picture. The three loops at the top center of the picture are the semicircular canals that provide our sense of balance. They are part of the same bone structure as the cochlea, which resembles a snail shell and can be seen at the right side.

The stirrup transmits vibrations to the fluid-filled cochlea, which converts the energy into electrical pulses that are sent to the brain. The auditory nerves are shown just above the cochlea. The Eustachean tube is at the bottom right and is used to equalize air pressure on the eardrum. To put size in perspective, the cochlea is about the size of a pea.

The inside of the cochlea is hollow and is the most complex part of the hearing mechanism. A cross section is shown in **Fig. 10**. It is divided into three fluid-filled parts—the tympanic (lower) canal, the vestibular (upper) canal, and the cochlear duct (middle canal). They are separated by thin membranes. The upper and lower canals are filled with a fluid called perilymph, and the middle canal is filled with endolymph. These fluids have different chemical compositions and electrical charge.

The stirrup transmits vibrations to the oval window of the upper canal. Vibrations travel through this canal to the tip of the cochlea at its center, where the cross section is the smallest. The waves then enter the lower canal until they reach the membrane-covered round window at the large end of the lower canal. This window dampens the vibration.

The small rectangle enclosing the organ of Corti in **Fig. 10** is shown magnified in **Fig. 11**. The organ of Corti is a gelatinous mass located in the middle canal. It performs the most complex and interesting of the transformations and contains about 7,500 interrelated parts.

There are four longitudinal rows of auditory sensor hair cells. Three of the rows are outer hair cells and one row contains inner hair cells. They are embedded in supporting cells of the basilar membrane that forms the floor of the middle canal. Bundles of stiff stereocilia (much smaller hairs) project from the hair cells. Incidentally, stereocilia and stereo sound are not directly related.

The tectorial membrane shown in **Fig. 11** is suspended from the bone surrounding the cochlea and projects over the hair cells and stereocilia like a shelf.

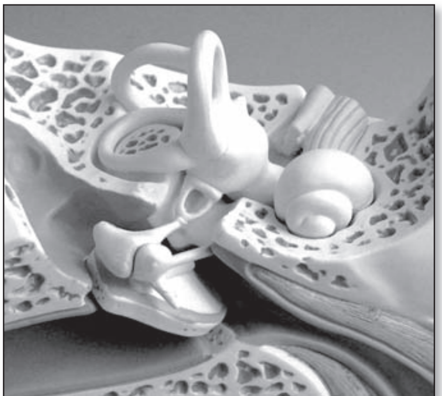
**PHOTO 3: Inner ear model.**

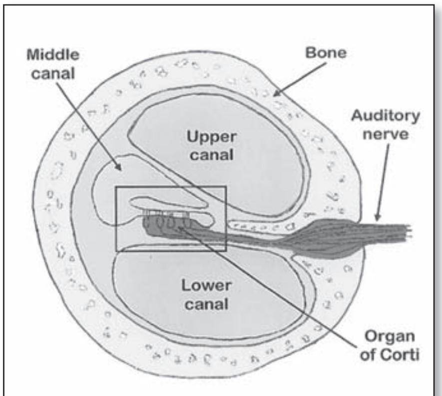

**FIGURE 10: Cross-section of the cochlea.**

It is coupled to the organ of Corti by the stereocilia bundles attached to the hair cells. There are over a million stereocilia in each ear. The fluid pressure waves produce motion of the basilar membrane that bends, twists, pulls, and pushes the hairs. This motion causes the stereocilia to move against the tectorial membrane.

The basilar membrane, also shown in **Fig. 11**, is light and taut near the wider stirrup end and thick and loose at the smaller end. Waves induce a ripple in the membrane. High tones produce their greatest crests where the membrane is tight; lower tones crest where the membrane is slack. The membrane can also pick up vibrations from the skull. When hearing is impaired, a bone conduction transmitter can sometimes be used to couple audio vibrations to the skull and restore partial hearing.

The chain of events for hearing then becomes even more involved with the conversion to electrical signals. The most intricate process occurs when motion of the stereocilia causes a chemical process to take place that alters the permeability of the membranes of the hair cells and allows positive ions to enter the corresponding cells. As a result, the respective cells develop a receptor potential and release more neurotransmitter molecules at its synapse with a sensory neuron. The resulting increase causes electrical discharges that are generated by the nerve cells. The cochlea-to-brain transmission system contains 30,000 nerve fibers issuing from the organ of Corti. The sound frequency interpretation depends on which fibers are activated.

Although there has been some correlation with nerve-firing rate and audio frequency, there is a recovery time for each nerve that limits the maximum firing rate to 1000 firings per second. This is a much lower rate than our upper limit of hearing at 20,000Hz. Nerves can take turns and increase the effective rate to 3000 firings per second, but information about most sound frequencies depends on which fibers are activated and not on the rate of firing.

Auditory nerve axons convey signals to the thalamus and cerebral cortex of the brain, where the auditory signals are interpreted as sound. Some signals are even fed back. The enormous amount of information transmitted to the brain can be selected and interpreted in an unbelievable number of ways. One that is commonly known is our ability to filter out a single conversation in a room full of people who are all talking.

## SAFETY DEVICES OF THE EAR

The middle ear has inherent safety devices that help to protect the inner ear from loud noises and large changes in ear pressure. Loud noise in excess of 80 to 85dB triggers two sets of muscles. One tightens the eardrum and restricts its ability to vibrate, while the other pulls the stirrup away from its link to the inner ear. Sounds that are most damaging are those above 500Hz and protection mainly affects the range below 2kHz. Of course, there is a release time for the muscles as well. However, the protective system is relatively slow to act, being from 10 to 30ms for very loud sounds and up to 150ms for levels near the trigger threshold[11].

Typically, protection response is more effective for younger people. Nevertheless, hearing damage can still occur because there is a tendency to play music even louder, overriding the efforts of our ears to protect themselves. Hearing damage can occur at any age for very short duration high-level sounds such as a rifle shot or other acoustic spikes. The muscles do not respond fast enough.

You may have noticed that as you play music louder that the sound changes. Not only is there an apparent frequency shift, but the quality of the sound also seems to change. This is very apparent for some people, but others may not have noticed the effect at all. I attribute it to the action of the muscles of the inner ear and a listening level above 80 to 85dB.

The second safety device is the Eustachean tube, which connects the air-filled middle ear with the mouth cavity and serves as a pressure equalizer. This is why I was advised, when firing an anti-tank weapon, to open my mouth so that pressure would then be equalized at the eardrum and possible hearing damage could be avoided. This is similar to the operation of a noise-canceling microphone in which equal pressure at both sides of the diaphragm reduces diaphragm displacement. You may more commonly experience this when ascending or descending in an airplane.

## HEARING DAMAGE

A normal hearing loss due to aging consists of a gradual decrease in high-frequency sensitivity over time. This means that for a man at age 35, hearing sensitivity could be reduced as much as 11dB at 8kHz. In comparison, for a woman at that age the reduction might be only as much as 5dB. This varies, of course, from person to person, but you can infer that sensitivity would be down even further at 20kHz[12].

On the other hand, hearing damage is an acquired symptom. Despite safety devices built into the human hearing mechanism, survival did not depend on tolerance to loud sounds. On the contrary, it depended on man's ability to hear and locate faint sounds. Humans are not equipped to tolerate exposure to so many loud sounds encountered in our present environment. Whether it is machinery, engines, or places of entertainment, the danger is there.

When it comes to car audio, home audio, and even headphones, you can be inadvertently attracted to the sensation of hearing and even feeling loud sounds and the sensation that the louder it is, the better. Unfortunately, prolonged exposure to excessively loud sounds or even short-term peaks can damage or destroy part or all of your delicate hair cells.

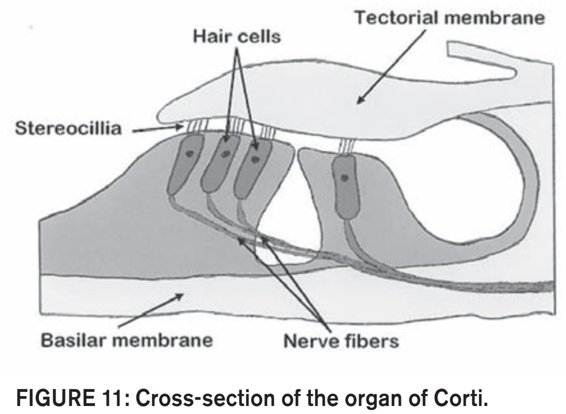

FIGURE 11: Cross-section of the organ of Corti.

They can become twisted, bent, and/or fused and are no longer able to respond properly to the incoming sonic vibrations. The ear's performance is degraded even when only a relatively small number of hair cells is damaged. The damage can be permanent in part or all of the frequency range for the rest of your life.

Hearing aids, which simply amplify frequency areas, might help to restore hearing to some degree depending on the severity of hearing loss. In cases of total hearing loss, cochlear implant surgery can be performed. Often, it is the stereocilia that are damaged and not the nerves. An implant can be effective in some cases and is becoming more common.

A typical operation involves inserting a tiny 22-channel electrode under the basilar membrane so that it curls around inside the cochlea. The electrodes stimulate the auditory nerves, bypassing the damaged hairs. An internal receiver is added near the cochlea with an antenna loop. An external microphone and amplifier are used outside the body.

The amplifier converts the audio signals to pulses, which are inductively coupled by placing the transmitting coil at the side of the head next to the receiver antenna implant. This operation can restore varying degrees of hearing, but not total hearing ability. Even if it restores only partial hearing, it is a tremendous help in reducing or lessening the feeling of isolation that total or near total deafness can cause.

A symptom of exposure to extra loud sound is ringing in the ears. This may go away after a period of rest, but repeated occurrences may someday cause permanent ringing.

For example, at one place of entertainment that used a loud public address system, I came prepared with a sound level meter and earplugs. It didn't take long for the sound to become very unpleasant, so I put my earplugs in and measured 85-90dB on the A scale and slow response. On the C scale with fast response, I measured 90-100dB. The PA system was used much of the time for about an hour.

I noticed later that one of the foam earplugs didn't fit tightly in my ear. Although I thought it cut down the intensity, when I removed both of them at the end of the performance, I could only hear well out of the ear with the completely sealed plug. Fortunately, this was only a temporary loss and my hearing recovered after 15 minutes or so. No warning was given to us that we would be exposed to uncomfortable or damaging sound levels. To quote from ASHA, "Sounds louder than 80 decibels are considered potentially dangerous. Both the amount of noise and the length of time of exposure determine the amount of damage."

Hearing damage is not just confined to adults. This can be even more prevalent with younger people, particularly those who use iPods with earbuds that are inserted directly into the ear canal. In a paper presented before the Audio Engineering Society[13], one researcher cited several audiological studies of students. In one case, it was found that 3.8% of the sixth graders failed a high-frequency hearing test, while 11% of 9th graders and 10.6% of high school seniors failed. A survey of incoming college freshmen yielded a 33% failure rate. The next year, 60.7% of the new incoming class failed. These figures clearly imply an accumulation of damage[14].

## THE EMOTIONAL APPEAL OF LOUDER SOUND

Several difficult protective devices could be used but they can take away from the physical and emotional sensation of loud sound. To quote one disco enthusiasnt, "I like to feel the sound." The reasons people prefer louder sound have received little attention[15]. Loudness commands our attention. It's a survival trait. Louder sounds are usually associated with danger or excitement. They make the pulse go faster. They take your attention away from other lesser stimuli. In a sense, it is like a drug, providing an escape to a different space controlled by sound like the rhythm or the inventiveness of the music. The louder the sound, the more immersion there is in this other space. This immediate gratification would seem to be without penalty because the gradual increase in damage is not at first apparent. Warnings are then easy to ignore, and, like drugs, loud sound can even be habit forming.

## CONCLUSION

In my opinion, sound in several dimensions has a long way to go to capture real acoustic events. However, so many improvements have been made in the last 50 years that music reproduction has become thoroughly enjoyable. For me, I only wish it had occurred at an earlier time.

There are two sides to pleasing sound. There is the re-production of reality and then there is the production of many alternate realities, all of which offer a different path to sonic ecstasies and all of which can provide many varied experiences. After all, the pleasure of music in various forms is what mankind has sought and expressed for many thousands of years. However, combined visual effects distract from a sonic experience and limit your imagination.

In addition, after this brief introduction to some of the delicate and very complicated functions of the ear, I hope that you will take care of your hearing and your children's hearing so that you may all continue to enjoy music and sound throughout your lifetimes. aX

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By Dennis Colin

# Intrinsic Fidelity Testing

"What's wrong with conventional listening evaluations?"

Glad you asked. Although the performance of a preamp, line stage, or power amp in your system is of primary interest (at least for now), there are so many variables that such evaluation is difficult, uncertain, and likely to change. These variables include:

1. The rest of your audio system. You might like a low damping factor tube amp because its speaker Z-dependent bass peak compensates a speaker/room bass attenuation. Or you might like your preamp's HF rolloff because it "tames" the HF edginess and graininess of your CDs.
2. Recordings. Most popular music, and even some classical recordings, have been "produced" with bass rolloff, compression, and various midrange/treble "equalization." This is because of the infamous "loudness wars"; sales are proportional to dB SPL.
3. LF hearing loss at the usual lower-than-live playback SPL (Fletcher-Munson effect). Together with variable #2, this is a good reason to use (good) tone controls, instead of colored components (expensive tone controls).
4. The uncertainties in evaluating a component. Preconceived notions about what a different or modified component "should" sound like, listening from a different position (moving your head 1″ can make over 1dB difference at 10kHz), and relying on memory are notorious in their ability to influence perception.
5. Not having heard live music (especially unamplified acoustic instruments and singers) for too long. How many times have you tweaked your system and been satisfied with it for some time, and then went to a concert and thought, "so this is what it should sound like"?

If, on the other hand, the component in question has the highest possible intrinsic fidelity to its input signal, it might not currently provide the best overall system sound. But when you upgrade other components and/or get better recordings, or upgrade from CDs to SACDs, DVD-As, or vinyl, that component's superior transparency will become evident.

## INTRINSIC FIDELITY TEST

**Figure 1** shows the general setup. The input/output levels should be matched to 0.05dB, because wideband differences of 0.2dB are audible. If the component has significant frequency response variations, the best you can do is level match at 1kHz, or some frequency at which the component's gain is around the 300Hz-3kHz average.

A power amp should be loaded with a reactive network approximating real-world speaker loads. An actual speaker would be best, but it would need to be acoustically isolated (e.g., wrapped in blankets in a large box in another room). Preamps can be load-sensitive, so if the attenuator and listening amp don't provide a low enough load impedance, add a parallel load resistor. The selection of attenuator pot value can help.

The component should be driven to several dB below clipping if a power amp, or to a representative normal-use level if a preamp. The component might need input attenuation or gain. Phono preamps can be tested if preceded by a very accurate inverse RIAA network, with the "In" position of the A/B switch coming from the inverse RIAA network's input.

To ensure near (but not at) clipping level drive for a power amp, an oscilloscope at the output is best. But as a substitute, you can set the power amp's gain (with its own input attenuator if it has one, or an external one) if you know the source player's peak output. With CD players, the Red Book full-scale output standard (0dBFS) is 2.00V RMS sine (±2.83V peak). The CD and SACD players I've measured (Sony) produce up to ±3.2V peak with "hot" recordings.

## LISTENING

In this test, any audible difference is undesirable. If the output sounds better than the input, you have the "expensive tone control" effect. (Or, as some ads would have you believe, some magic time-machine circuit that removes recording studio distortion!)

With a very transparent, neutral, and quiet component, you might not hear any difference. If this is so, congratulations! Following are some suggestions for maximizing the sensitivity of the test, while also guarding against the influence of thought processes (when listening for

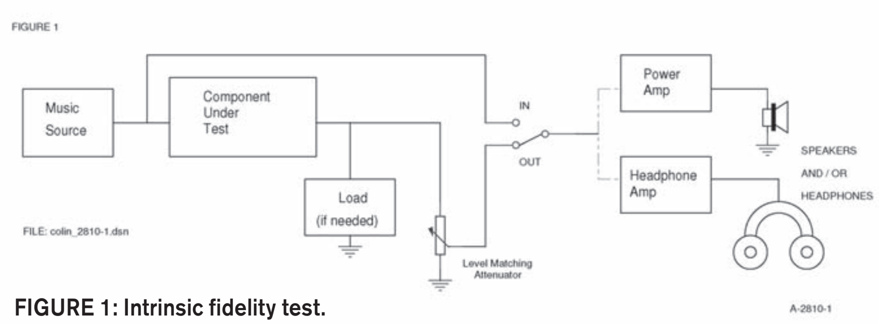



**FIGURE 1: Intrinsic fidelity test.**

very subtle effects, thoughts, expectations, and so on can make you believe you hear something that disappears with enough patience, relaxation, and persistent repetition):

1. Use high-quality headphones, which can remove room acoustics, eliminate frequency response changes due to head movements, and have much higher resolution than most speakers. It's good to also listen with your speakers, because a very slight effect heard on high-quality phones could be inaudible in normal speaker listening. And I suppose it's possible that the room-enhanced speaker stereo ambience sound field could be more revealing of subtle spatial effects that the tested component could be producing.
2. Choose the best available recordings. Vinyl, SACD, and the rare DVD-A that uses its full resolution format are the most revealing. Conventional 16/44 CDs just don't have as much resolution and grain-free transparency. However, because in this test it's only differences, not preferences, that we're listening for, even pink noise and electronic impulses can reveal some effects. But the best musical sources will, of course, be the most revealing.
3. Be prepared to spend at least an hour with the A/B switch. I recommend several switching methods:
   a) listen to an entire musical piece in each switch position.
   b) switch after the first few seconds of a piece (preferably one that "starts with a bang"), then quickly reset the track to the beginning. Repeat at least 20 times.
   c) Find a sustained violin ensemble harmony, piano chord, sung note, horn tone, reverb tail, and so on, and rapidly switch back and forth. Listen without thinking about which position the switch is in. It's best to allow all thoughts to evaporate and just enjoy the music. If you notice a difference, that's when it's time to focus. But remember that focusing on one thing excludes others, so after analyzing one perception, "remember to forget" again, and proceed.
4. The following might seem silly, but it can be revealing: temporarily connect both switch positions to the same point (input or output). Then perfect your acting ability by pretending you're still comparing two different signals. You might be surprised at the influence of the normal music variations, or of head motions if using speakers. If nothing else, this silly exercise can provide a "noise floor calibration" for establishing a limit on reliable and consistent audibility thresholds.

## THE EFFECT OF SMALL FREQUENCY RESPONSE VARIATIONS

As I mentioned, 0.2dB variations can be audible. However, they're not necessarily perceived as such. Through a phenomenon that can be called "psychoacoustic mapping," frequency response variations can affect imaging, focus, and spaciousness. Generally, bass/lower midrange emphasis increases perception of space, and treble emphasis increases perception of tonal details and image focus.

A little experimentation with an equalizer or tone controls will help you recognize these effects. Then if, say, a preamp appears to improve detail, it could be a simple and mild HF boost. Because this also increases any edginess in the source material, a simple treble tone control could make this preamp's use satisfactory. This sounds obvious, but it serves as an example of the potentially overwhelming number of variables affecting overall sound quality.

## POSSIBLE EFFECTS OF THE SWITCH AND ATTENUATOR

1. I mention this in the interest of true science. Remember, "science" doesn't mean just measurements. The term literally means "the open-minded search for truth." I personally don't believe the switch and attenuator (if of good quality) have any perceivable effect, but my mind is open to the possibility.
2. The music signal has already traveled through many resistors, capacitors, attenuators, cables, amplifying devices, and possibly transformers (and switches, connectors, and so on). And

the recording medium most likely has far more degradation than all of the above components used in recording.

3. If you still think the switch and attenuator could have effects, you can perform an intrinsic fidelity test on them alone (e.g., use a second switch for the "component under test" in *Fig. 1*). The same applies to cables and connectors. Note: I use Caig "Deoxit Gold G5" (formerly called "Pro Gold") spray on all connections. Using this on a clean tin-plated connector provides a better contact than a dirty gold-plated one!

## RESULTS OF SOME TESTS

First, a note on my hearing. For the scientists, my threshold is within 3dB of the standard "good hearing for young people" audiogram. I have a good friend whose threshold is 20dB above (less sensitive than) the standard audiogram, yet he can easily hear the superior resolution of vinyl and SACDs (in that order) over 16/44 CDs. For the audiophile, my hearing has been praised by Ed Dell and Joe D'Appolito regarding my speaker reviews in *aX*.

I've performed the intrinsic fidelity test on two components so far:

1. "Mad Katy" 250W stereo tube amp (*aX* June/July '06): There were infrequent times when I thought I might have heard a difference, but if so it was on the edge of my perceptibility. Furthermore, very extensive listening made me doubt there was any audible difference.
2. "The LP797 Ultra-Low Distortion Phono Preamp, (Sept. '07 *aX*)" preceded by a stereo pair of very accurate (±0.03dB) inverse RIAA networks ("Tweaking the Passive Inverse RIAA Network," *aX* Aug. '07), overall (combined) gain trimmed to ±0.01dB at 1kHz. I heard absolutely zero difference. No coloration, distortion, noise, hum, nor loss of spatial or tonal resolution, nor change in dynamics.
3. While not an intrinsic fidelity test, I performed a conventional A/B comparison of the outputs of "Mad Katy" and a 70W per channel solid-state amp. Using both recorded music and live miked voice and a bell, I easily heard the difference. The tube amp had that natural voice "roundness of tone," and very natural reproduction of the bell's transient "clink" and resonating "clang." (I also compared the reproduced sound on excellent ribbon-tweeter speakers, to the live bell and voice). The solid-state amp, by comparison, made vowel sounds slightly "constricted," and made the bell sound slightly "electronic," instead of like real metal vibrating.

## COMMENTS

The following are my personal opinions, based on much observation—but nevertheless they are beliefs, and as such are open to debate. I welcome your (favorable or unfavorable) responses. All I ask is that they are respectful; please don't accuse me of being biased or having tin ears!

1. In the phono preamp test, with no audible difference, the signal traveled through eight caps (two in the inverse RIAA network), many resistors, one JFET, three op amps, and some ordinary (but good quality and clean) RCA cables.

   All resistors were 1% metal film (but not "exotic" brands). All caps were metalized polypropylene. All connections were cleaned and sprayed with G5. The unity-gain trimpots (in the inverse RIAA network) were precision 14-turn cermet units.

   Because of the no-sonic-difference result (very carefully auditioned for an hour of switching), I conclude that claims of the audibility of (ordinary but good quality) resistors, film caps, wire, and the best op amps are unfounded.
2. I believe that such claims are the result of one or more of these factors:
   a) The claimants' assertions are based on memory (listening, soldering in a different resistor, and so on, then after many minutes listening again). Memory is notoriously unreliable (ask any courtroom judge), plus you're probably not re-listening in exactly the same position in the room; this strongly influences the frequency response you hear. Also, the SPL is probably not

matched to the required 0.2dB or better.

b) The influence of beliefs and expectations. Please don't be insulted; these influences are well known by audiological researchers. There was a story in *Stereophile* of an experienced audiophile who bought a simple wall clock that the manufacturer claimed (if you can believe this) would "purify the AC line from timing imperfections in the whole house!" (and I'll sell you the Golden Gate Bridge for $10). It was the infamous "Tice Clock."

Anyway, the audiophile installed this clock behind his refrigerator out of sight. After about a month, he told a friend that the clock improved the sound of his very expensive music system—better resolution, less noise, and so on. Then his friend revealed the truth to him: on the day the audiophile installed the clock, the friend secretly unplugged it! The poor audiophile wouldn't speak to his friend for several months!

c) Here's a factor that's less controversial: Many circuit designs can be overly sensitive to a component's value and its parasitics (e.g., the Effective Series Resistance of a cap). This can be caused by potentially unstable feedback loops, power supply interactions, excessive tube or transistor bias-point changes (with signal amplitude, line voltage, or temperature), RF interference, and so on. Such factors can also make a component overly sensitive to cable capacitance and inductance. You can merely replace a 5″ piece of wire, then notice a difference.

But the heat of soldering could permanently change a resistor's value or the leakage of a cap, not to mention the surface leakage of a PC board that had been high due to humidity. If the circuit is overly sensitive to such things, the audible difference would be wrongly attributed to the atomic structure of the replaced wire! (Another factor could be different ground capacitance if the new wire position is different).

3. If you really believe that, say, a 5″ piece of wire can make an audible difference, please perform a careful intrinsic fidelity test on it. If you then hear a consistently reliable difference, I'll believe you and not use that kind of wire!
4. I realize that it's possible to have (as an example) ten elements (resistors, wire, and so on) that individually have a degradation that's below audibility, but when combined these degradations (especially if of the same nature and polarity) sum to an audible level. This is, of course, a good reason to use the highest affordable quality parts, at least up to a point.
5. The opposite side of that argument is the extremely revealing nature of a very careful and extensive intrinsic fidelity test. If the differences heard are very small, the component's effect will likely not be audible in normal use.
6. There's some opinion that after listening normally for some time—say several months—you can notice subtle effects that aren't revealed in, say, an hour of even rigorous evaluation. I find this to be untrue; I've listened with some particular components for over a year, and any "accumulated observations" were revealed in that first hour of intrinsic fidelity testing, and with greater sensitivity.

Furthermore, hearing has evolved to be very adaptable in "tuning out" benign anomalies such as small amounts of 2nd harmonic distortion and mild frequency response deviations.

## CONCLUSION

Intrinsic fidelity testing serves two purposes: It allows the most sensitive evaluation of an electronic component, free of other (system and source) variables, and it can be very informative about the effects of various circuits, topologies, and parts, on sound quality. *aX*

### NOTE

This article was written before, but is now published after, my article "An A/B/I Switch" (June '07). The philosophy of that switch is described here.

**review** | By Charles Hansen

# Tube Imp Update

Following our first review of the Tube Imp Mini Tester[1], British Audio Products sent an updated version of the Tube Imp for follow-up evaluation. I am happy to report that the modified unit produced results closer to those I obtained with the Audiomatica Sofia. As of this writing, there was no word on the availability of updates for existing Tube Imp owners.

## MEASUREMENTS

I used the exact same vintage Mullard ECC83 as last time, checking the gain and transconductance using the tube for the two Class A amplifier conditions specified in the RCA Receiving Tube Manual RC-30. As before, I set the specified plate voltage and then adjusted the grid voltage to produce the specified cathode current. For the second Class A amplifier condition I needed to use Va = 200V DC rather than the 250V DC specified in the RCA manual due to the Va limitation of the Tube Imp. I compared them to the same data points from the plate curve data file I ran on the Audiomatica Sofia.

The results of these tests on section 1 of my Mullard ECC83 are shown in **Tables 1** and **2**. I included the earlier Tube Imp test data for reference. The Sofia displays μ, gm, and Rp directly along with plate current Ia. The Tube Imp measures gain (μ), mA/V (gm), and cathode current (Ik), so I calculated Rp from the formula Rp = μ/gm.

**British Audio Products/
Moth Group**
10 Dane Lane, Wilstead, Bedford, Bedfordshire, UK MK45 3HT
www.britishaudio.co.uk or www.tube-imp.co.uk
e-mail: info@mothgroup.com
++44(0)123 474 1152
Fax: ++44(0)123 474 2028
Price: £299 U.K.
Net weight with carrying case: 4.4 lbs (2kg)
Available online at http://store.securehosting.com/stores/sh204131/shophome.php?itemprcd=tubeimp

**TABLE 1 Measurements, Mullard ECC83 section 1**

RC-30 Class A: Va = 100V, Vg = -1V Ia = 0.5mA

| Parameter | Data Book | Sofia | Tube Imp 5/07 | Revised Tube Imp |
|---|---|---|---|---|
| μ | 100 | 95.2 | 72.4 | 89.4 |
| gm (mA/V) | 1.3 | 1.17 | 1.3 | 1.14 |
| Rp kΩ | 80 | 81.2 | 55.7 (calc) | 78.4 (calc) |

Note 1: Rp calculated from Rp = μ/gm for Tube Imp

**TABLE 2 Measurements, Mullard ECC83 section 1**

RC-30 Class A: Va = 200V, Vg = -2V Ia = 1.2mA

| Parameter | Data Book | Sofia | Tube Imp 5/07 | Revised Tube Imp |
|---|---|---|---|---|
| μ | 100 | 98.1 | 72.1 | 88.2 |
| gm (mA/V) | 1.6 | 1.64 | 1.2 | 1.53 |
| Rp kΩ | 62.5 | 59.8 | 58.1 (calc) | 57.6 (calc) |

Note 1: Rp calculated from Rp = μ/gm for Tube Imp
Note 2: Va held to 200V due to Va limit of Tube Imp

## CONCLUSION

The revised Tube Imp understated the gain by only 6% for the Va = 100V test in comparison to the Sofia, and 10% for the Va = 200V test. The gm was only 3% low in the first test and 7% low in the second test.

The revised Tube Imp now produces better absolute values of μ and gm than the first review sample. I can also now read gm with a resolution of two decimal places rather than one.

## REFERENCE


1. "Review: Tube Imp Mini Tester," Hansen, C., *aX*, pp. 33-35, May 2007.


*We believe it is appropriate to congratulate this British company for their traditional quality of response to critical reviews, in sharp contrast to some US manufacturers who behave petulantly in aggrieved fashion when responding to such review criticism.*—**E.T.D.**

**Audio Aid** By Rickard Berglund

# Build a New Tone-Control Circuit

This tone control is a modification of the original Baxandall "Negative feedback tone control" published in *Wireless World*, October 1952. The potentiometer P1 controls the whole range 1kHz-20kHz. The control ranges of the potentiometers P2-P7 are as follows:

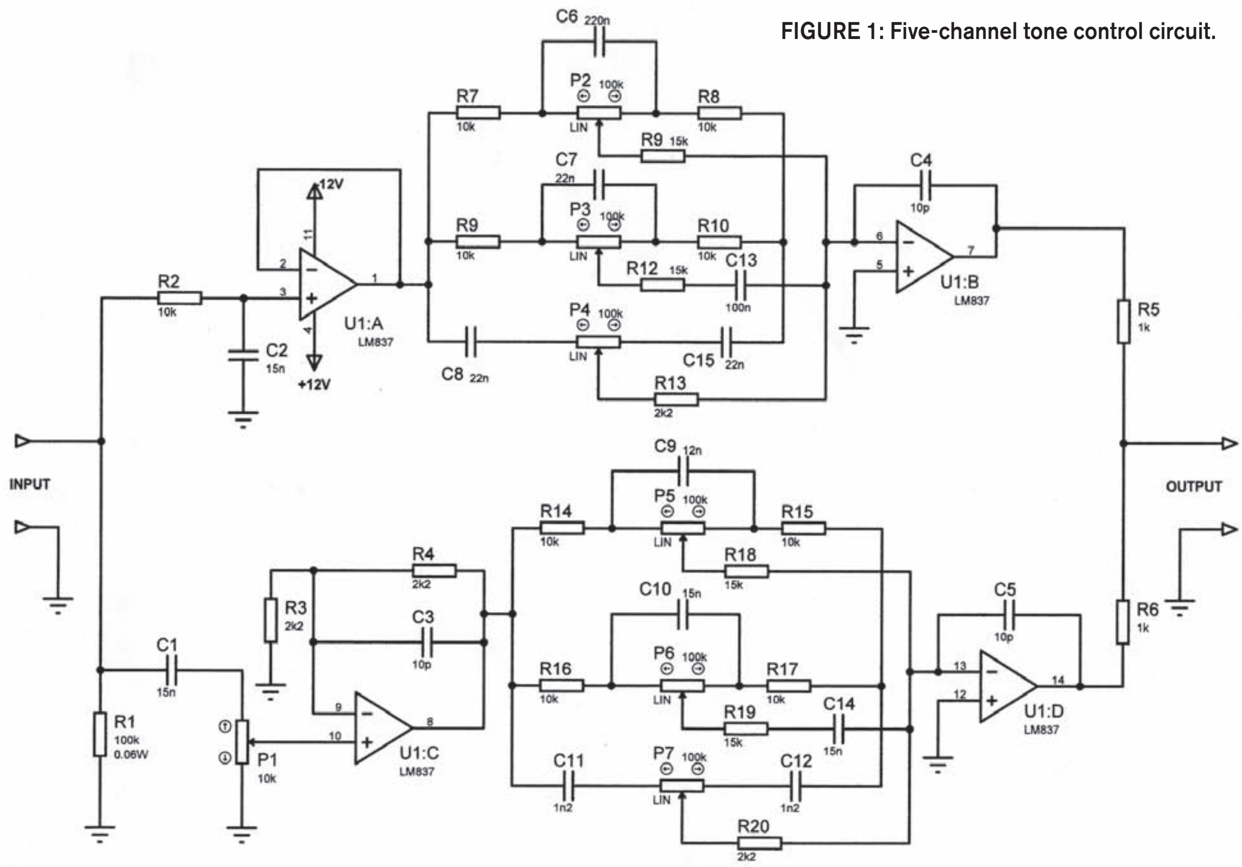
| | | |
|---|---|---|
| P2 | low bass | 100Hz |
| P3 | high bass | 250Hz |
| P4 | low midrange | 500Hz |
| P5 | high midrange | 2000Hz |
| P6 | low treble | 4000Hz |
| P7 | high treble | 10,000Hz |

The output impedance is 500Ω, and the input impedance is 9kΩ. For IC1-IC4 you can use one LM837 or two NE5532s. You can use the same power supply as used for the phase meter in *audioXpress*, November 2006 ("A Phase Meter Calibrator," p. 22). *aX*

**Parts List**

| | |
|---|---|
| C1, C2 | 15nF |
| C3-5 | 10pF |
| C6 | 220n |
| C7 | 22n |
| C8, 15 | 22n |
| C9 | 12nF |
| C10 | 1.2nF |
| C11-12 | 1nF |
| C13 | 100nF |
| C14 | 5.6nF |
| P1 | 10kΩ lin. |
| P2-7 | 100kΩ lin. |
| R1 | 100kΩ |
| R2 | 10kΩ |
| R3, 4 | 2.2kΩ |
| R5, 6 | 1kΩ |
| R7-10, R14-17 | 10kΩ |
| R11, 18 | 15kΩ |
| R12, 19 | 6.8kΩ |
| R13, 20 | 2.7kΩ |

FIGURE 1: Five-channel tone control circuit.



# Yard Sale

## For Sale

Loudspeakers: Focal 5N313, 5″ neoglass, $50/pair;
Morel MDT27 horn tweeter, $30/pair;
Scan-Speak 18W8545K00 kevlar, $50;
Audax HM210X aerogel $40/pair;
Audax HP210Z12 aerogel $30/pair;
A-R 3/4T tweeters, $20/pair;
Radio Shack FE103, 4″ paper, $40/4;
MCM 55-2320, 5″ fiberglass laminate, $60/8.
Tektronix 2205, 2-channel, 20MHz scope, $100.
e-mail: renatus@britten.com.

L. F. Garbage filter partially assembled with instructions, all parts and article (Jung *AA* 3/82), $8;
Used Morrey super buffer used as DC servo loop one cap broken, no article, $6.
Both $12. Free Shipping! (US only).
Vincent Mogavero. Vmogavero@aol.com (718) 672-7478.

## Wanted

Original Empire DB 208 belt in usable condition (I can adjust for tension.) (NOS OK) (Shiny surface OK). for $8 or ?.
Vincent Mogavero. Vmogavero@aol.com (718) 672-7478.

"Yard Sale" is published in each issue of *aX*. For guidelines on how subscribers can publish their free ad, see our website.

# VENDORS

Audiophile components: JFETs, MOSFETs, Tantalum, Caddock, Dale resistors, Black Gate, ELNA Cerafine, SILMIC II, TONEREX, Nichicon Fine Gold Muse, KZ, Jensen four-pole capacitors, Stepped attenuators, Teflon cables, Connectors
Custom designs for OEM customers
Custom Assembly
**www.borbelyaudio.com**
Selected BORBELY AUDIO kits in Japan:
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## What's New on the *aX* website?

### Supplemental Information:

Corrected Fig. 2 and Fig. 3 from Dennis Colin's "The LP797 Ultra-Low Distortion Phono Preamp" (*aX* 9/07).

Specifications from Tom Perazella's "The Kiss Bass Project" (*aX* 9/07).

Gerber File from "Analog Bass Control for Magnepans" by George Danavaras (*aX* 8/07).

Circuit Cards for "Cascoding National's LM12 for 140W" by Bill Christie (*aX* 8/07).

### Articles from Past Issues:

"A Hybrid Valve MOSFET SE Amp, Part 1," by Stephen W. Moore (*aX* 10/07).

"Hideaway TL Sub Revisited," by Bjorn Johannesen (*aX* 10/07).

"The Six Tee Nine Tube Amp," by Randy Miller (*aX* 10/07).

"The LP 797 Ultra-Low Distortion Phono Preamp," by Dennis Colin (*aX* 9/07).

"More on the Sound Strobe," by Ed Simon (*aX* 9/07).

"Calibration of the Hickok 539C Tester," by Daniel Schoo (*aX* 8/07).

"Tweaking the Passive Inverse RIAA Network," by Dennis Colin (*aX* 8/07).

"Optimal Stages for Minimal Distortion," by Dennis Colin (*aX* 8/07).

David Weinberg's review of HIFICRITIC (*aX* 7/07).

For information on these and others, visit www.audioxpress.com. And don't forget to check out the links to our other magazines, *Voice Coil* and *Multi Media Manufacturer*!

## Ad Index

## CORRECTION

Within the September 2007 *audioXpress Letters* discussion on pp. 58-60, there are a couple of problems, specifically in the letter from Marc Whitney and my reply.

On pages 59 and 60, the graphics between Figs. 1 and 2 got swapped in this letter, while the captions are correct as printed. Thus, if the graphics data from these two figures is swapped, everything appears in proper context. Also, the fonts within the schematic figure got lost somehow, so what should be "Ω" symbols show up as "W" (proving Murphy's law once more).

*Walt Jung*
*WaltJung@waltjung.org*

## VOODOO ENGINEERING

I read with interest Mr. Kwinn's suggestion for a construction article about a power regenerator ("Xpress Mail," July '07, p. 43). I felt obligated to come forward and clear up one or two minor misconceptions Mr. Kwinn (and many other people) seem to be under.

Virtually all audio equipment operates internally on DC power. The power supply takes in the AC line current and converts it into clean DC power of an appropriate voltage or voltages to power the circuitry. During this process the power supply removes all trace of the AC (and any noise present along with it). If the internal circuitry is critical about receiving a specific operating voltage, a regulated supply is used so that variations in the line voltage don't affect the supply voltages. This also removes any variations that might result if the input line voltage actually varies (possibly due to other devices on the line).

The purpose of a power conditioner (or filter) is to remove some or all of the noise on the AC line. This serves to "help" a power supply that isn't quite well enough designed to do a good job by itself. Power supplies frequently benefit from a little help of this type. It also blocks any "unreasonable" noise—such as the surge from a lightning strike or motor transient.

A power regenerator goes one step further and "makes all new line voltage" for the equipment to run on. This allows it to ensure that the voltage remains constant and that there is no noise present. This level of control should *not* be needed if the attached equipment is reasonably well designed. The sole way in which a power conditioner or regenerator can positively affect the sound of the devices connected to it is to *remove* any odd noises or "muddiness" that might occur as a result of noise leaking through their power supply and "annoying" the circuitry—or by providing stable voltage to equipment that has poor regulation and must run off particularly bad lines. It should also protect them from damage by major transients, but that will not be audible.

Any power regenerator such as the "famous" PS model, which actually produces power *other* than the 60Hz AC sine wave (50Hz in Europe) on which all equipment is intended to run, is an example of "voodoo engineering." There is no technical reason for varying the line frequency to produce any beneficial effects whatsoever on properly functioning equipment. If anything, there is the distinct possibility of causing a degradation in performance or actual damage. [If a transformer is defective, and buzzes at line frequency, it is *possible* that changing the line frequency *might* reduce this while not otherwise damaging anything. It might also reduce the interactions between hum in different components—not by eliminating the problems, but by shifting one so they no longer "line up." Both of these situations, however, presuppose one or more defective pieces of equipment to begin with.]

If there is, somewhere out there, a piece of equipment that benefits from a power regenerator, it would almost certainly benefit *more* from a careful redesign of its own power supply—which would also probably be much less costly.

The only exception to this is with *very* powerful devices such as large power amplifiers. Because of the way most power supplies draw current, they draw most current during the peaks of the sine wave. Under some circumstances this can allow the power waveform to become distorted in such a way that less peak voltage is available to the device—which can limit the maximum power output. This only occurs with devices that draw high current (power amplifiers of several hundred watts per channel), under some power line conditions, and requires a regenerator capable of supplying 20A or so to correct—which is far outside the context that was suggested.

The best way to improve (possibly) the sound of small signal equipment is to provide them with clean, well-filtered power, and to upgrade their power supplies (if they are deficient) with better regulation and high-frequency filtering. Of course, the actual circuitry might be a candidate for improvement as well.

*Keith Levkoff*
*kLevkoff@panix.com*

## SUGGESTION

How about a survey comparing coupling capacitors? I'd love to set up a group of Hi-Fi legends in the same room using a controlled setup and swap in different coupling caps from Sprague Orange Drops to "Unobtainium" Jensen Silver/Gold Foils. The survey could rate various aspects on a scale of 1-10 and add comments.

*Craig Myers*
*cragtone@sbcglobal.net*

## CORRECTION

A while back (in May) an error was found in the silkscreen pattern of the circuit board supplied in the kits that Old Colony Sound Lab is marketing to match my sub article ("Simple Servo Sub 100/120 Modification," Dec. '06, p. 8). I found this when a customer who couldn't get his sub working correctly sent me his sub (at my request).

The silkscreen markings for diode D1 (a 5.1V zener diode) and capacitor C1 (a tantalum capacitor) are backwards. The silkscreen shows the cathode side (with the bar) of the diode and the [+] leg of the capacitor going to ground. The bar of the diode and the [+] lead of the capacitor both need to be connected differently from the way they are marked.

The anode (without the bar) side of the diode and the [-] lead of the capacitor should go to ground. In other words, just turn each of these around.

Diode D2 and C2 (same types as D1 and C1) of the board are labeled correctly. Also, the schematics both in the article and as supplied with the kit are correct—the error is only in the silkscreen pattern of the kit printed circuit board. Fortunately, building to the incorrect orientation will not cause any damage to equipment or board components—just remove C1 and D1 and put them back in the opposite way and all should be well. With these parts backwards, the sub will function somewhat, though it will sound really lousy!

*Bill Waslo*
*www.libinst.com*

## CHIP UPGRADES

In response to Mr. LeBeck's letter in the May '07 issue ("DVD/SACD Players," p. 58), I upgraded my Philips CD 650 and built my own headphone amp. I had some good experience with OP275 and the new National Semiconductor LM4562 (both dual) in my old Philips CD 650 player. In my WNA MKII headphone amp I use AD843 (single OP) in signal path. Another good single OP is the National Semiconductor LME49710. My equipment consists of Sennheiser HD 600 and (preferred) AKG K 701 headphones, a modified Philips CD 650 CD player, and Marantz SA 7001 KI.

I plan to get some AD828 and AD827 to compare them.

*F. Stockhammer*
*www.stockhammer.eu/hifi/*

**R.K. LeBeck responds:**

This interesting letter puts me in mind of my need to (finally) investigate building a real headphone amp, as well as upgrading/replacing/supplementing my headphone collection (Sennheiser HD580, Beyer DT-990 Pro). I, too, had wanted (some few years ago) to investigate the OP275, but never got around to it. The LM4562 was first mentioned to me by a Silicon Valley friend sometime late last year—its datasheet is extraordinarily impressive and I do hope to try it out. To make things clear, my experiments with the AD827 [unity gain stable] and AD828 [minimum gain stability is 2] were provoked by my frustration with the sound quality of "affordable" DVD/SACD/CD players.

Years ago, I modified a Philips DAC960 for a friend [see the 1992 POOGE articles by Jung and Galo]—the ultimate op amp upgrades were AD811 [I-V conversion] and BUF-03 for the output buffer/filter. I later applied the same devices and some of the methods to a Philips CD-80 CD player (wholly-derived [??] from a Marantz-Japan design, it seemed built like a tank). I worked on these units several times between 1991 and 1998.

Both of these units use the same Philips chipset (16-bit DAC and digital filter) as the DAC960 and the CD-80 as well as other Philips/Magnavox (typically in the US) 4×-oversampling CD players (examples: Magnavox model numbers CDB460, 560, 582 and probably others unknown to me). I was particularly impressed by the performance of the AD811 during that period, and also by the BUF-03.

In 1998, I also tried the BUF-04 in a Magnavox CD player and it then seemed like a very good alternative to the BUF-03. However, I don't believe either item is in production anymore.

In 2001, I procured an Assemblage DAC2.7 with the ordinary op amps installed because the old "Parts Connection" had a sale on many similar items back then. I soon modified the DAC2.7 [an outboard DAC] with the op amps shown in its user manual's schematic: AD811 and OPA627A. In that unit, those devices definitely improved the audible performance—later in 2003, I tweaked it a bit more, using ideas I remembered from the POOGE articles by Jung/Galo.

Thus, I have never forgotten the Jung/Galo usage of the AD811 in 1992 (it's a high-speed video op amp and is a "transimpedance" amplifier requiring some attention in its use). Here, I might also mention that a friend (in 2004) was able to get good results with an AD812 (a sort-of dual version of the AD811 with slightly diminished specs, but also a "transimpedance" amplifier). He used it as a unity-gain buffer in an old Nakamichi CD player's buffer/output-filter stage—but had to follow the "transimpedance" amplifier's (easy to do in the unity-gain configuration) requirements.

With all this history behind me [use of the AD811], I eventually thought of applying the "normal" (i.e., they are *not* "transimpedance" devices) video-oriented op amps: AD827 and AD828 (they are dual versions of the AD847 and AD848).

Finally, I first applied the AD827 (with great difficulty, because of its size as a normal dual-op amp) to the Sony DVP NS500V SACD/DVD/CD player. The results were very gratifying and SACD discs finally sounded like I'd always hoped they would.

A bit later I used a SOIC package of the AD828 in a second and identical Sony unit and it worked just as well [the Sony unit's technical documentation showed that the op amp stage had a gain of ~2, thus making the AD828 a good choice], and was much easier to install (if you forget the problems due to size and handling of the much smaller SOIC chips which were what the Sony unit's audio PC board was designed for—and is typical of almost all modern equipment).

Finally, I'd recently heard of a low-cost universal disc player, the Oppo DV970HD DVD-V/DVD-A/SACD/CD/and so on disc player, which had begun getting attention in the magazines I normally read every month. Wanting to extend the experiment beyond the Sony players, I opened up the OPPO unit and saw that it used the same op amps as the older Sony unit [the ubiquitous 4558 dual op amp]. Fortunately, use of a test disc (CBS CD-1) and an oscilloscope made it easy enough to find the op amp I was most concerned with (the two front stereo channels) and thus replacement of the SOIC 4558 with an SOIC AD828 (sourced from Digi-Key) was reasonably easy enough to do.

Subsequent tweaking: replacement of the electrolytic coupling capacitors in the front stereo channels—in the Sony *and* the OPPO units—with the economical Black Gate; 16V rated substitutes (source-able from Michael Percy Audio and other suppliers) improved the sound quality even further (all of these new capacitors were bypassed with small—because easy to use in these modern units using so many relatively tiny electronic parts—Panasonic 50V rated polypropylene units from Digi-Key).

## TUBE BOARDS

The article titled "Three New Prototyping Boards" (June '07, p. 38) was great to help DIY builders know what's available. My **Photos 1-3** show the JCO tube

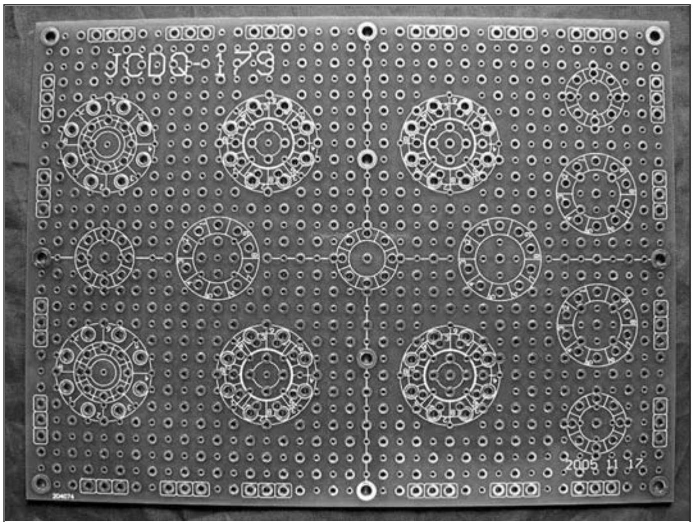

PHOTO 1: JCO tube prototype board.

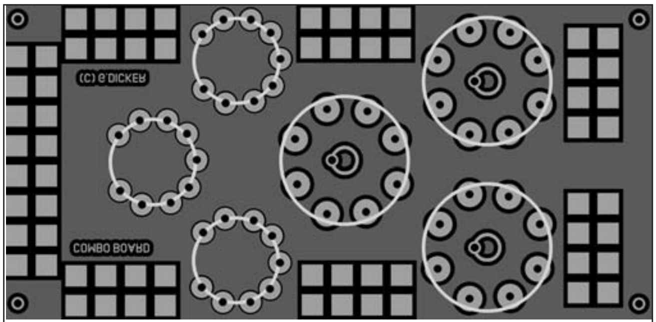

PHOTO 2: Author's protoype board.

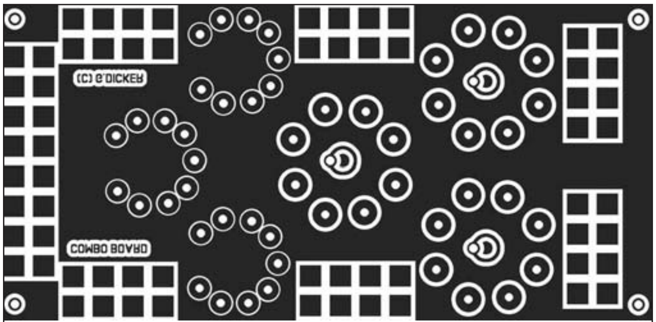

PHOTO 3: PCB pattern.

proto board available on eBay for around $25US posted, and my own proto board available for $15US including worldwide postage. The PCB pattern is also included for those who may wish to make their own.

*Graham Dicker*
*Blackwood, Australia*
*gtd@bettanet.net.au*

## IC AMP UPDATE

In reference to Jack Walton's "High End 120W MOSFET IC Driven Amp," *aX* 7/07:

1. Unless the amplifier is powered from a 240V line, it is against NEC to switch both input lines. The grounded neutral on a 120V line may not be switched (or fused). UL will not list a device with a switched neutral.
2. How does the power supply provide ±56V from a 44VCT winding on T1? 88VCT would work.
3. The Fig. 2 schematic shows device U1 as a loudspeaker. The parts list identifies the LM4702 as U1. Where does the lead connected to R13, R3, L1 actually go?
4. Starting component ID with "1" on each schematic leads to confusion. I suggest that you sequentially assign each schematic a number. Component ID then starts with C11, C12; C21, C22, and so on. This also ties components shown on the silkscreen to the proper schematic diagram.

*Charles Stillhard*
*Bisbee, Ariz.*

**Jack Walton responds:**

1. As a ham radio operator for many years, I posed this concern to the technical experts at the American Radio Relay League in Newington, Conn. Reader Stillhard is correct. In 120V AC lines ("Hot Neutral Ground") the ground should not be fused or switched, which can pose a shock hazard. In 220V AC lines ("Hot Hot Neutral") both lines may be switched.
2. Each side of the center-tapped transformer I used delivers approximately 44V AC. After rectifying and filtering, and given the losses in the surge protectors and diodes, this averages out to somewhat less than the theoretical value you would expect.
3. I am sorry for the confusion with respect to the connection of the driver board and the output board. The U1 and U5 designations are the same device. While a loudspeaker schematic symbol is shown, in actuality this is just to illustrate where the speaker should be hooked up. The node that joins the "load sharing" resistors on the output board (R3-6) is the same one that conjoins L1, R1, R3, R13.
4. I lifted a page from the work of Erno Borbely ("New Power Amp Modules, Parts I and II," *Audio Amateur*, 3 and 4/93) and decided that a more flexible layout would use separate driver and output boards, allowing for bipolar transistors or MOSFETs in the final stage. A board for the output stage is not really necessary; just keep in mind, however, that the gate stopper resistors should be as close to the MOSFET gate as possible.
5. I apologize for the component ID numbering, but my CAD program automatically numbers components.

It seems there is an error in Fig. 2 of Jack Walton's article. I believe that C13 (the 470nF negative power supply bypass cap) should be connected to the side of R16 nearest the LM4702.

I don't build many projects, but I still find construction and design articles interesting and informative. I was wondering when people would start using the LM4702.

*Keith Levkoff*
*kLevkoff@panix.com*

**Jack Walton responds:**

As reader Levkoff correctly points out, capacitor C13 is connected directly to Vee pin of the LM4702. The connections are correct on the printed circuit board, incorrect on the schematic.

The Gerber Files for the driver board have been posted to the *aX* website (www.audioXpress.com) along with images of the top and bottom copper layers and silkscreen. These files and images are for Version 1.4 boards. The easiest way to get started is to use the images (which were produced at 100%) and print onto 3M transparency film.

Subsequent to publishing the article, I had some observations for improving the performance beyond the statistics published at the end of the article:

1. Before installing the RC network consisting of R2/C14 (330Ω, 10pF), check to see whether the amplifier actually oscillates. If there is no problem with oscillation, you can leave out these components and the THD% improves by a hair.
2. On the output stage schematic I showed 330pF capacitors (C1, C3) connected from gate to source of the output MOSFETs. If you eliminate these capacitors, the slew rate of the amplifier will decrease, but stability will be enhanced. Your mileage may vary depending upon the load characteristics and the precise configuration of the output stage.

In the amplifier I used for my day-to-day listening, C1/C3 are in place, but the R2/C14 RC network is disconnected.

After the article was published, Troy Huebner of National Semiconductor wrote an application note which is required reading: "*LM4702 Driving a MOSFET Output Stage*," AN-1645. In particular, Huebner demonstrates an output biasing scheme for the Renesas Lateral MOSFETs which simply consists of a 1K potentiometer across the source and sink pins of the LM4702 and a 47K resistor from sink to ground. He also uses 10pF capacitors (which I show as 30pF silver micas, C4 and C11 on the output stage) which "hot up" the slew rate of the AN-1645 amplifier. I have used this simpler bias scheme which works well.

Huebner further describes the method for explicitly determining the value of gate-stopper resistors. Use of other output devices, such as the Vishay IRFP240/9240, Magnatec BUZ901/906, and Toshiba 2SK1530/2SJ201 MOSFETs, are described in this application note. National promises an application note that will discuss the optimization of the LM4702 compensation capacitors. The LM4702 is soon to have several brothers and sisters in a large family of high-end driver devices.

Version 1.6 driver boards, the 2SK1058 and 2SJ162 lateral MOSFETs, and some of the ancillary bits and pieces are available on my webstore, www.tech-diy.com/LM4702_Kits.htm. I did not want to flog the product on *audioXpress* (and the PCBs were not yet available at press time). I promise to keep interested readers up to date on developments via my website. The amplifiers based upon the LM4702 are very listenable, and I am gratified by the enthusiastic response to the article.

## METER MEASUREMENTS

I can confirm the 10dB peak of the Radio Shack digital sound level meter, in an interesting story. While working on my current speaker project (a sort of reincarnation of the JBL Paragon using a LE14A woofer and a LE85 compression driver on a home made tractrix horn), I received tremendous help from the JBL Forum (see thread: http://audioheritage.org/vbulletin/showthread.php?t=12420). I was having a problem with a large peak in the response of the LE85 at 6kHz, which turned out to be my RS meter, not the LE85. That reminded me of the previous speaker I built which *also* had a peak at 6k (measured by the RS meter).

I then produced a simple LCR trap with a rheostat for the R to adjust the peak and successfully "tamed" the 10dB peak to a reasonably flat response. The problem was that in listening to the speakers, especially on female vocals, both my wife and I preferred the speaker without the trap (with the 6k peak). At the time I found this curious, but built the final version to what our ears liked. Now with my new measuring setup (see above thread), I find the "ear tuned" speaker is very nearly flat. So much for measurement versus "golden" (or not so) ears!

*Mark Parker*
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## MAN AND MUSIC

After reading the review in *audioXpress* (7/07, p. 6), I ran out and bought *This Is Your Brain on Music*. When reading this book, I wished there were a companion CD with some of the sounds and songs that were referred to in the text. I just now discovered that the author has a website with these sounds: www.yourbrainonmusic.com.

*Mark Rumreich*
*mark.rumreich@thomson.net*

## MUSICAL INFLUENCES

The review of *HiFiCritic* (July '07, p. 45) was disappointing, as was my rough impression from their website. I don't doubt the integrity of Mr. Colloms and their other reviewers, especially in light of their no ad policy. However, I question their ability to distinguish physically caused sonic differences from mindset influences.

Case in point: the "colouration" from the Quad speaker's illuminated logo. Peter Walker, if he can divert himself from enjoying the Great Concert Hall in the sky, will laugh his head off!

Circa 1957, *Life* magazine published an article on "synaesthesia," in which some people could supposedly "hear colors," "smell sounds," and so on—an "intersensory crosstalk" phenomenon. The lighted Quad logo certainly has *visual* coloration (unless it's pure white, of course). But a sonic influence?

Now, over decades of observing my mind and the tricks it can play (which the great writer Jiddu Krishnamurti said can be great fun), I've become well aware of the influence on musical perception of many factors, both mental (beliefs, and so on) and physical (my health, comfort, room lighting, and so on). Regarding the latter, it's no secret that dimming or darkening the ambient light enhances musical pleasure; concert halls and seductive lovers do this! The ears have less competition from the eyes.

I suggest that *HiFiCritic*'s reviewers

try evaluating the Quad light's "influence" when blindfolded (a new raison d'etre for "blind testing"). Then, if they hear no effect of the light, they could show some semblance of conducting scientific research by investigating the effect of visual distractions on musical perception!

*Dennis Colin*
*Gilmanton, I.W., N.H.*

## MORE ON LM12 AMP

Thank you for printing Bill Christie's article describing the LM12 amp. I usually build tube amps, but the LM12 looks like an interesting first solid-state project. I have a couple of questions:

1. I haven't found a source for the special Vishay/Sprague 22,000µF caps. Can you point me to where I might order some?
2. Mouser has a MJ15015G and a NTE388 cross-referenced to a MJ15015. Would these parts be OK for Q1?
3. I wasn't going to bother building the clipping indicator circuit. I'm assuming it's not really needed. Am I wrong?
4. I found an AVEL toroid (made in Conn., where I live) that has a 40V secondary rated at 1000VA. I know with tubes this would be close enough. Except for a possible size difference, would this work in your LM12 circuit? Or would I need to change some circuit values? If so, I'll just get the exact part you recommend.
5. After I've built this amp, do you have recommendations for the first power-up, such as using a Variac, or key voltage checks? I was going to use dummy 8Ω loads, but do you have any other thoughts?

Thanks again for a very interesting article. And thanks in advance for your time.

*Tom Ottman*
*Berlin, Conn.*

**Bill Christie responds:**

Thank you for your interest in the article. I hope these responses will answer your questions satisfactorily.

1. I apologize for a typo in the article regarding the part number for the Vishay/Sprague filter capacitors. The part number should read 36DY223F0075CB2A, which is available from Mouser Electronics.
2. The cross-referenced transistors should work fine.
3. The clipping indicator is entirely optional. Sometimes I like bells and whistles.
4. The transformer I specified has two 38V windings forming a 76V center-tapped secondary. If the AVEL toroid is 80V center-tapped or has two 40V windings, it should work fine.
5. It's always good to use a Variac for initial testing and possibly a lower power supply fuse.

I just read Bill Christie's article "Cascoding National's LM12 for 140W." I am in Greece and am unable to read the whole article about this amp; I can only see the pdf files existing in your site.

From what I've seen, this design is much different than the original posted by National Semiconductor. Can you tell me why? Also, why does your design sound better? The recommended rail voltage in the National book is 100+100V. How much is yours?

A note about proper thermal design in the National book recommends keeping the tip 31-32 drivers outside heatsink. Maybe they don't require cooling, but is there a chance that they need thermal compensation?

Finally, the original article mentions a swing of 90+90V and output of ±10A. Do you think something like that is possible?

*Sakis Petropoulos*
*east_electronics@yahoo.gr*

**Bill Christie responds:**

The circuit that I have adapted for my design is taken from National Semiconductor's application note AN446 and is shown in Figure 11 of that publication. The circuit uses a higher voltage power supply, so I needed to adjust some of the component values in the cascode circuit. The only other change was to lower the gain for an inverting input buffer with higher than unity gain. I added the input buffer primarily to raise the input impedance. National's circuit has an input impedance of 1kΩ, which would be too low for many preamplifiers.

I redrew the schematic to make it easier for me to read, but is circuitwise identical to that in Figure 11 except for the changes in the component values just mentioned. The power supply that I used is ±56V. This gave me all the power I wanted. The National circuit shows a ±100V supply, which theoretically could output 900W of power. I assume it will work as advertised and don't see any reason why it wouldn't sound as good as mine provided the preamplifier can handle the low input impedance.

The TIP31 and TIP32 transistors are used with MJ15015 and MJ15016 in darlington configurations to increase the current gain. They dissipate very little power and do not require mounting on the heatsinks.

I very much enjoyed Bill Christie's article on the LM12 ("Cascoding National's LM12 for 140W," 8/07, p. 38) and would like to build several of these amps. I have not been able to find the artwork you mentioned as being available on **audioXpress.com**. Since you obviously have built a number of these and have the artwork, you also probably have a supplier to manufacture them. Why not offer them to other *audioXpress* readers?

I see several projects a year that require circuit boards, but it's difficult to find someone to make them. The setup costs for the first set can be prohibitive for small boards.

*Bruce Brown*
*TuninFork@aol.com*

**Bill Christie responds:**

I etched the circuit boards myself. I'm forwarding a 600 dpi file of the artwork to you and to *audioXpress* for inclusion on their website (www.audioXpress.com). I printed it mirror image on transparency film with an inkjet printer with the ink volume set to heavy and quality set to best. PC board material with a photosensitive coating is available from Parts Express. A 6″ × 9″ board will handle two channels.

## BIAS SCHEMES

After reading Tim Smith's article on building a pentode amp ("Pentode Petite," *audioXpress* 6/05), I put it on my list of projects to build. When I came across a set of power and output transformers from a Harman Kardon A 300, I decided it was time to build the Pentode Petite.

The A 300 used 7355 output tubes and a voltage doubler's HV power supply, but they looked like they would work perfectly. The power supply I built used the voltage doublers connected to Smith's design for the rest of the HV section. It put out 335V. The filament supply paralleled all the tubes to use the 6.3V secondary of the power transformer. I built the bias supply as presented in the article with a separate transformer. It produced 24V unloaded, so everything seemed to be fine.

The sound of the amp was very nice, with no hum or oscillation, but I had trouble keeping the output tubes balanced, and the power and output transformers became warm after 3-4 hours of use, even at moderate levels. The bias voltage at the grid of the EL34s was –18 to -19V and the cathode current was 105-125mA (not even close to the 60-80mA specified in the article). I even tried several sets of EL34s and still couldn't solve the issue. The tubes ran very hot, but no glowing on the plates. I even tried using different secondary taps to change the plate load on the output tubes, all to no avail.

My friend Larry builds many tube amps and never uses fixed bias. He prefers the automatic bias setups for simplification and less maintenance, so I decided to change the pentode to an auto bias system.

The mods to do this are as follows:

1. Remove the complete bias system from the amp (T3, D2, D3, C12, C13, C14, R8, R9, LED 1, C15-18, PA1, and PA2).
2. Remove R12 and R13 from the cathodes of each of the output tubes.
3. The ends of R18 and R19, which used to go to the bias pots, now need to be grounded.
4. From the cathode (pin 8) of each EL34, connect a 520Ω 5W resistor to ground and parallel it with a 200-250mF 100V electrolytic cap (+ to pin 8).

You now have an amp with auto bias. So how did it turn out? The amp sounds every bit as good as it did before, but even after 8 hours of listening the power transformer just becomes warm to the touch. The output tubes run much cooler and the measured voltage at each cathode is within 1-2V of each other. I suspect the power output is lower than with fixed bias, but for most situations, this really isn't a problem.

I love building and modifying vintage audio equipment and encourage everyone who reads *audioXpress* to consider submitting articles. For me this is the best thing about the magazine. I very much liked Tim's article and think it is a superb amp with lots of neat and innovative features. My modifications are in no way due to a problem with his design, but an improvement that worked for me.

*Bruce Brown*
*Valley Springs, S.D.*

**Tim Smith responds:**

I'm sorry that Mr. Brown had difficulty with the bias supply with his version of the Pentode Petite amplifier. However, it seems that he has implemented a very useful cathode bias scheme that I would recommend as an alternative biasing scheme for anyone who wishes to build this amplifier. I have always used fixed bias because it essentially enables more power output and better balancing of the output tubes. However, in retrospect, I think that a cathode bias scheme may be a better choice for a novice project such as this one.

He noted that the voltage applied to the E34L grids was in the −18V range. This is insufficient and reveals that there may be a problem in the bias supply as he built it. The bias supply is based on a small Radio Shack 12.6VCT transformer which feeds a voltage doubler supply. Output from this supply, measured at the negative end of C14, should be about −34V with 120V supply.

R8, R10, P1, P1, and R11 form a series resistive chain which places the wipers of P1 and P2 at about -22 to −23V when both pots are centered (perhaps Mr. Brown inadvertently used higher values for R8 and/or R10 than indicated on the schematic). This voltage, with about 325V on the E34L screens, provides about 70mA idle current for each tube. If less idle current is desired, you can replace R11 with a 3.9 or 4.3k .5W resistor. However, this modification will slightly lessen the amplifier's gain and increase its distortion.

I'm glad to find out that a number of people have decided to build this amp or a version of it. Mine is now with a friend who uses it a lot and is happy with it. For those who are interested, I am currently working on an enhanced design that uses a custom power transformer which will let the amplifier use the full 30W capability of the output transformers. This power transformer, wound by Heyboer, probably costs less than the combined cost of the three power transformers in the original Pentode Petite design.

## OP-AMP STAGES

Mr. Colin's article ("Optimum Stages for Minimal Distortion," Aug. '07, p. 50) appeared just as I was pondering whether or not an op amp gain stage ahead of an LM3886 (or other IC amplifier) might be a good idea. The application note for the LM3886 and similar devices shows sample circuits typically with a gain of 20, which might be a little less than desired for some applications. I had feared that adding another stage perhaps with a gain of 5 would degrade the sound more than simply changing the NFB resistor values to achieve a desired gain of 25.

Although Mr. Colin's focus was phono

amps (and probably equally applicable to line-level preamps), I believe he addressed my concern.

There may be an issue with which [the op amp or the LM3886] reaches clipping first, but that could be worked out. I would be interested to hear the author's thoughts on the matter.

I would also have liked Mr. Colin to include noise into the equation, because I can conceive of that being as much or even more of an audible concern than distortion. In any case, the article is much appreciated.

*Marc Whitney*
*Phoenix, Ariz.*

**Dennis Colin responds:**

Thanks for your interest. The LM3886 is a 68W audio power amp, and as such, its distortion is likely higher than a good line-level op amp such as the OPA 134PA (or dual unit OPA 2144pA). Consider that power amps generally have an optimum amount of NFB (too little might not attenuate low-order distortion enough, while too much can produce excessive high-order distortion, particularly with reactive load phase shifts).

Therefore, it's probably best to use the manufacturer's recommended amount of NFB—e.g., the closed-loop gain of 20 (26dB) that you stated for the LM3886—and then drive it with a good op amp whose gain you can set as desired. But if you need a gain of 25, which is only a 1.9dB increase over the gain of 20, the LM3886 shouldn't be compromised noticeably, if at all.

68W output power is 23.3V RMS into 8Ω; peak voltage is 33.0V. With an LM3886 gain of 20, then, its full-power input voltage requirement is 1.65V peak, 1.17V RMS. Now, while CD players tend to provide 2.8V peaks with "hot" recordings (2.83V peak or 2.0V RMS is the 0dB full-scale Red Book standard), many sources might not provide the 1.65V peaks needed for full power output.

So it appears that an op amp preceding the LM3886 would be desirable. The OPA2134PA (dual for stereo) has excellent performance: 0.00008% THD at unity gain, or 0.0008% with a gain of 10 (20dB); 8nV/√Hz voltage noise, negligible current noise (3 femtoamps/√Hz), 8MHz GBW, 20V/µS slew rate, and it costs only $2.63 from Digi-Key. Using this op amp offers additional benefits:

1. Non-inverting overall polarity can be restored if the power stage is (or performs better when configured as) inverting.
2. The power stage is driven from a very low impedance.
3. Ultrasonic and/or infrasonic rolloffs can be easily implemented.
4. Overall distortion can be lower because the power stage's NFB can be set to the optimum level. Note: I found the OPA2134PA's input and output sonically indistinguishable in an "intrinsic fidelity" test (see "Intrinsic Fidelity Testing," June '07).
5. Noise will likely be lower. In a two-stage amp, if both stages have the same net equivalent input noise voltage, and the input stage gain is 3.16 (10dB), the overall input-referred noise is only 0.41dB higher than a single-stage amp. *However*, here, the OPA2134PA (or OPA134PA) noise is probably *much* lower than that of the power stage. As an example, suppose the power stage's input-referred noise is 10dB higher than that of the op amp, and the op amp gain is set to 3 (9.5dB). Then the overall two-stage input-referred noise is 6.8dB *lower* than that of the power stage used alone.

This assumes that the op amp's feedback resistor is low enough to not contribute much added noise. With a 1k value (4.06nV/√Hz noise at +25° C), the OPA2134PA noise will be degraded by 1.0dB. With this, a non-inverting configuration, and a 499Ω resistor from inverting input to ground (for a gain of 3), the wideband unweighted noise will be about 103dB below full output signal.

These low resistor values require a non-inverting op amp configuration to allow a high input impedance (a 100k resistor from the non-inverting signal input to ground is good). Then, if the power stage is inverting, you can simply reverse the speaker leads to maintain correct overall polarity (absolute polarity can sometimes be audible because of the ear's asymmetric nonlinearity at high SPL).

I recommend an OPA2134PA input stage with a gain of about three, followed by a passive volume control, which feeds the LM3886. The low-z op amp output can easily drive a tapped volume pot with a loudness-compensation circuit (such as I used in "A High-Quality Solid-State Headphone Amp," Nov. '05). The result should be a very nice and inexpensive 68W per channel amp. *aX*

## CONTRIBUTORS

**Edward T. Dell, Jr.** ("Paper vs. Screens," p. 6) is owner and publisher of Audio Amateur, Inc., publishers of *audioXpress*, *Voice Coil*, and *Multi Media Manufacturer* magazines.

**Iain McNeill** ("A Mains Analyzer–How Clean Is Your Juice? Part 1," p. 8), an avid reader since the days of *Speaker Builder*, resides in Aptos, Calif.

**Dennis Colin** ("Frequency Delay Dispersion," p. 16, and "Intrinsic Fidelity Testing," p. 36) has demonstrated the audibility of phase distortion at Boston Audio Society, and has designed the "Omni-Focus" speaker (bipolar coincidental with phase-linear first-order crossover), ARP 2600 analog music synthesizer, 1kW biamp and PWM supply at A/D/S, and Class D amps.

**Stephen Moore** ("A Hybrid Valve MOSFET SE Amp, Part 2," p. 26) has extensive experience in the advanced energy field from both the science and business perspectives. He has written multiple patents relating to control theory and the energy field and is widely published on topics of lithium battery systems. He is a registered expert in the International Electrotechnique Commission (IEC) regulatory and standardization body SC/TC21A.

**Roger Russell** ("Sounds and Hearing, Pt. 2," p. 32) has a degree in electrical engineering from Rensselaer Polytechnic Institute. He has written several articles about loudspeaker design, testing, and other audio devices. Several patents have been awarded for some of his designs. Although retired now, he was a member of the Audio Engineering Society and the International Society for General Semantics for many years. His interests range from stereo sound to photography, new-age music, mystery clocks, and science fiction. He has a web page at *www.roger-russell.com*.

**Charles Hansen** ("Tube Imp Update," p. 40) holds five patents in his field of engineering and enjoys modifying guitar amplifiers and effects to reduce noise and distortion. With hundreds of articles to his name, he is a frequent collaborator to the magazines of Audio Amateur Publications.

**Rickard Berglund** ("AUDIO AID: Build A New Tone-Contol Ciruit," p. 42) has done research in analytical chemistry and industrial electronics design in Sweden. He is now engaged in construction and manufacturing of tube amplifiers.

The audio hobby for **Alex Arion** ("Classic Circuitry: Greek Style," p. 52) began in the magic '60s in Communist Romania in his grandmother's attic, where, together with two friends, he installed a small "illegal" laboratory. Their first attempt was a beginner's guitar amplifier and a professional guitar pickup. The tubes used included ECC40s and EF40s. The rectifier was a 5µ4, Russian equivalent. They mounted all in a hat box that they found in the attic. After finishing his studies and many years of research, Alex moved to Greece, where he continues to build different tube-based and solid-state amps.

# Classic Circuitry: Greek Style

My latest tube amp (**Photo 1**) features 6A3 tubes (2A3, with 6.3V heating), preceded by 12SL7 tubes. I call this unit Black Cherry, due to the color scheme of the chassis. The unit's dimensions (**Fig. 1**) are in metric, and the circuit (**Fig. 2**) includes nothing new under the sun, mainly due to the work of American and European pioneers who have gone before (primarily between 1938 and 1970). I chose simplicity in the design, because I'm sure that the best amp is simply a wire connected between a source and the speaker. Unfortunately (or not), we need to put something in between.

**PHOTO 1: The Black Cherry tube amp.**

## CIRCUITRY

The power stage (**Fig. 3**) includes two 6A3s in parallel, working in class A, with automatic bias tube amps. All the parts are good quality, carefully selected. The output transformer (OTL, **Fig. 4**), the most important part for all tube amplifiers, is manufactured here in Athens by Tim, who has a lab specializing in customized transformer construction. The first stage is a classical SRPP based on a 12SL7 double triode with a good enough amplification factor (I also tried the Russian 6H9C, with the same results). The amp could work with zero feedback or with different amounts of feedback, controlled by the switch.

There is nothing special about the power supply, except maybe the fast diodes for the high voltage rectifier and the special capacitors. No stabilization and no slow start—this to keep the price down! Of course, the power transformer (toroidal type) and the choke were

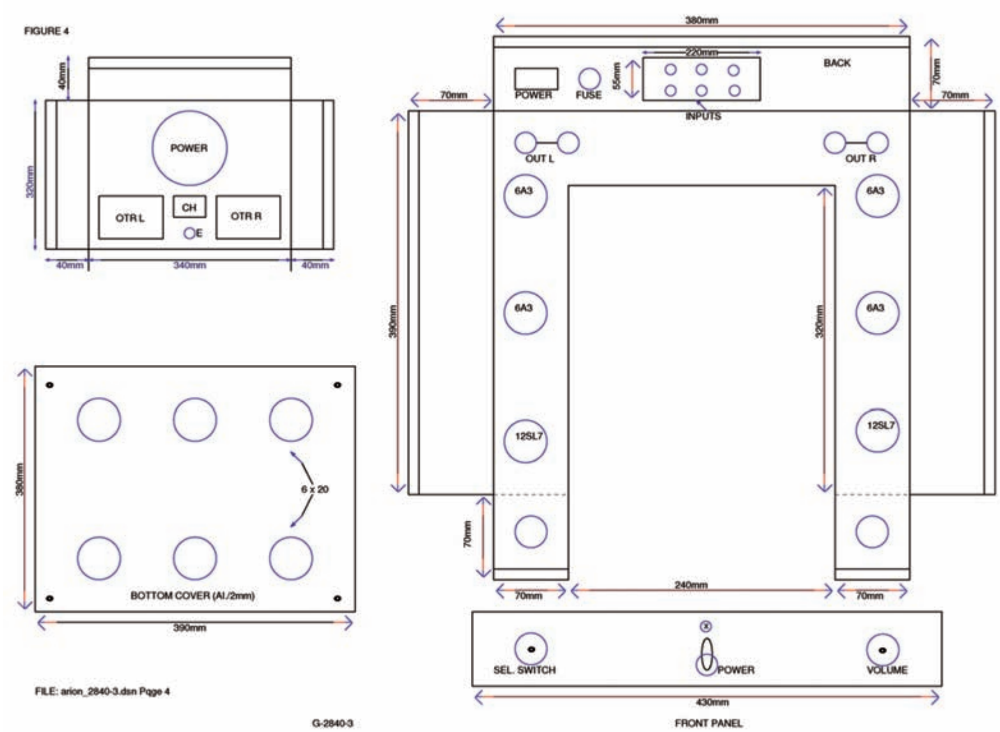


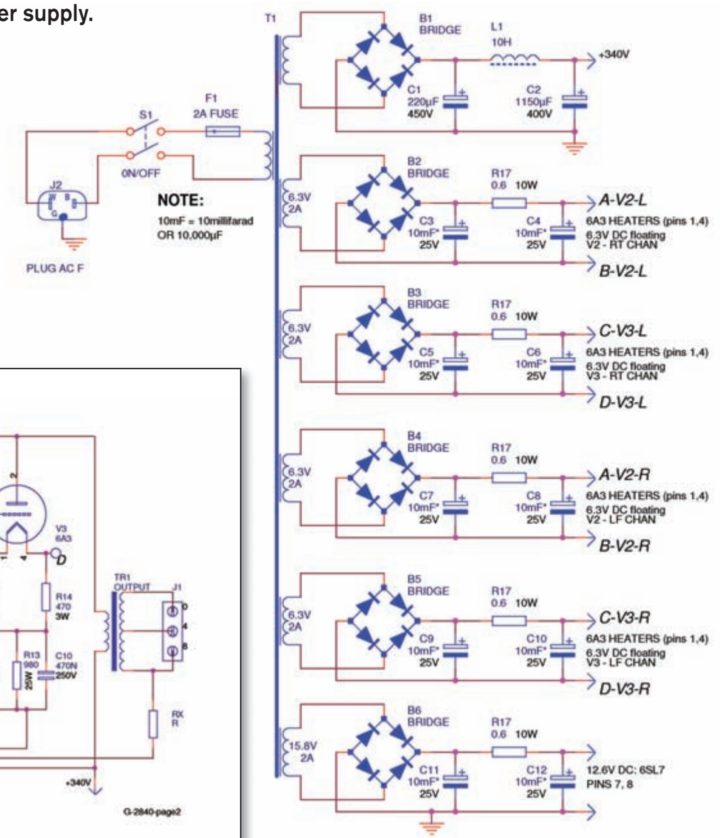

**FIGURE 1: Chassis dimensions.**

made by Tim, as we requested.

## CHASSIS AND TUBES

The chassis, at least, is an original design. In fact, there are two chassis mounted one on top of the other, made from aluminum 2mm soldered with Argon. The bottom one contains all the transformers and the big capacitors, while the top one includes the tubes and the in/out connectors. The 6A3 and the 12SL7 are Sylvania (NOS), and came from a for-

FIGURE 3: Power supply.



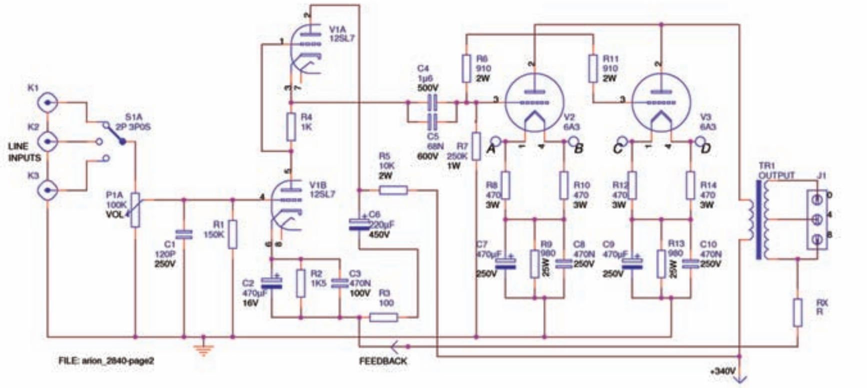



FIGURE 2: Amp circuit (one channel).

mer American military base. On a similar chassis, I used the impressive 6336 power double triode, with good results. (Side note: A few years ago I returned to my former country of Romania, for family reasons. A friend took me to visit a military base, where I observed some soldiers amusing themselves by shooting at tubes and watching the little explosions. I stopped the massacre and purchased the survivors, at a good price. The brand names included Philips, Siemens, Tesla, and others, in original boxes!)

I would like to thank my collaboration team: Hristos Diamantopoulos (metallic and special electronic construction), Aris Tsiklis (architect design), and George Provataris (metal/wood construction).

*aX*

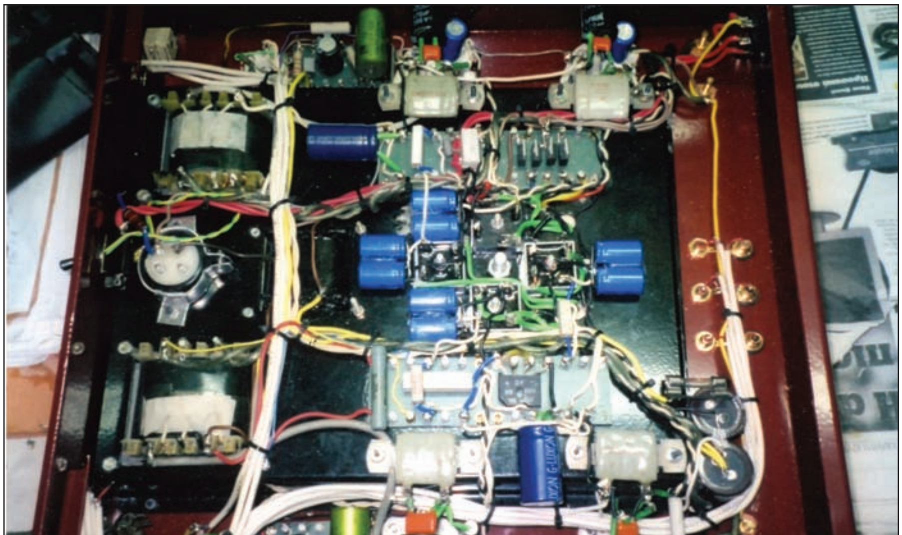


**PHOTO 2: Amp innards.**

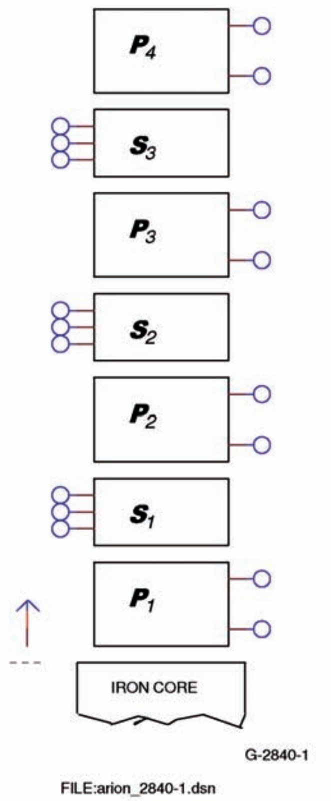



**FIGURE 4: Output transformer data.**