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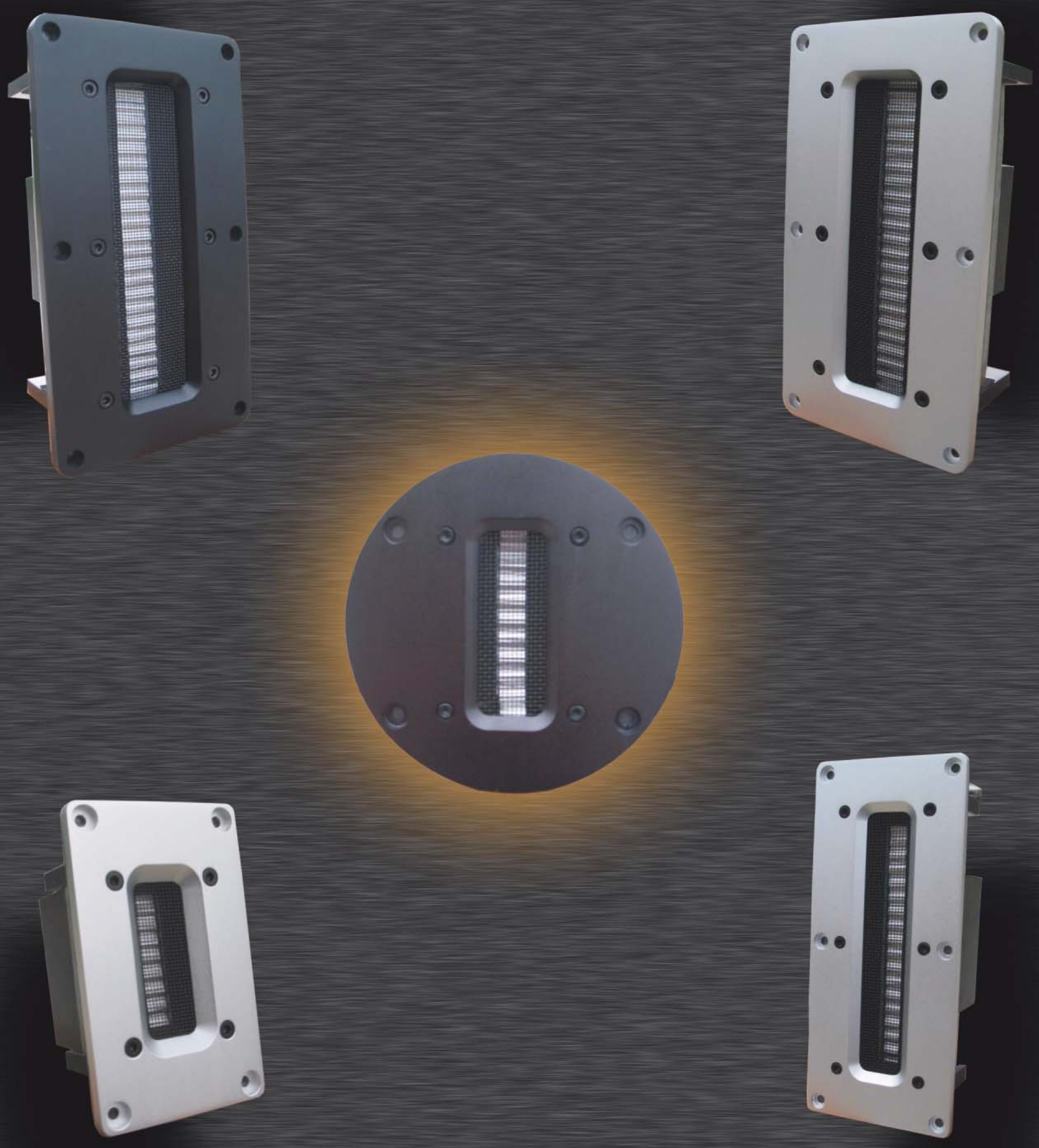

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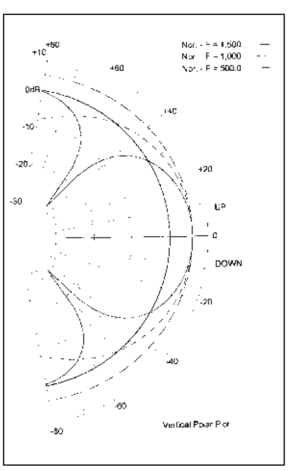

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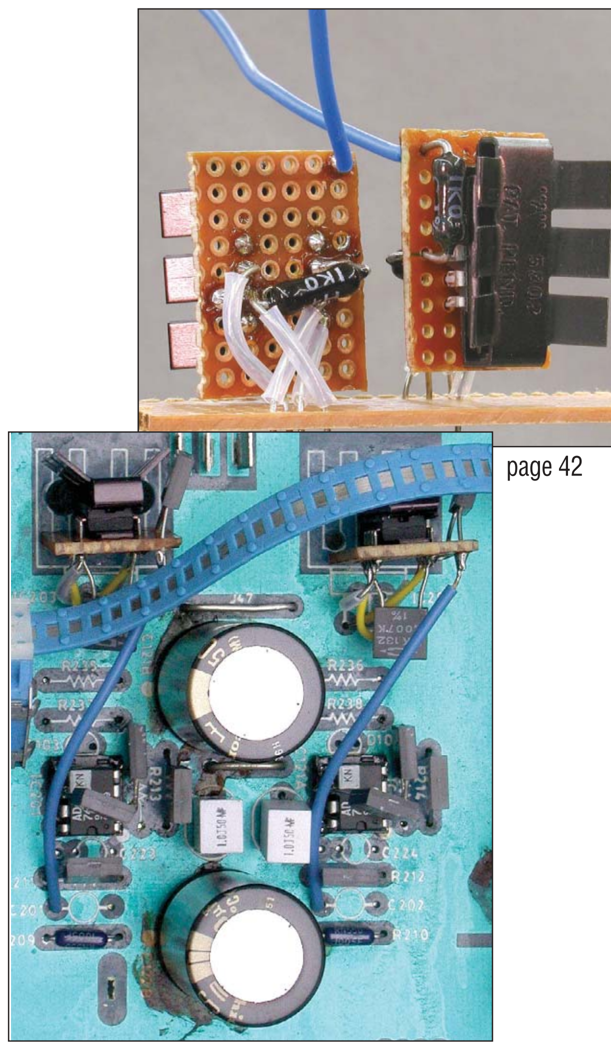

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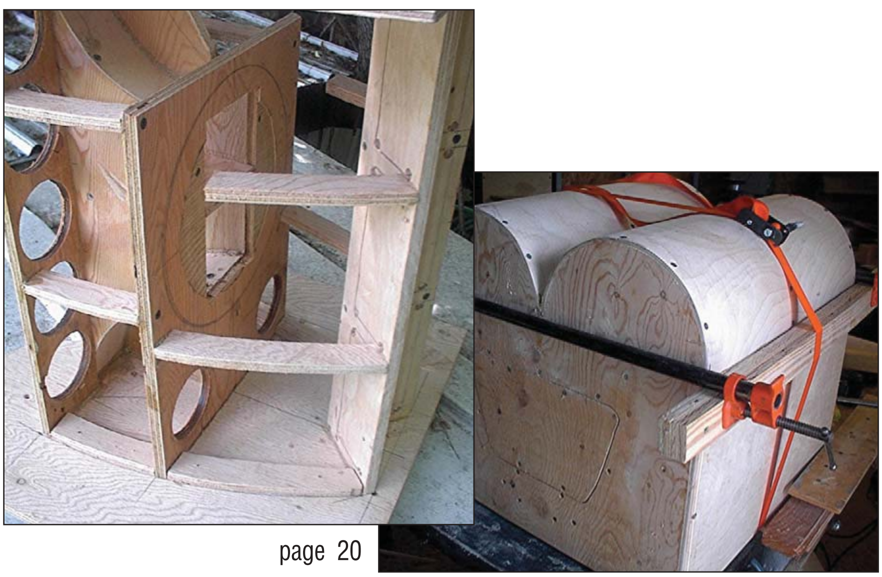

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JOHN STUART MILL

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Guest Editorial

Surround Sound Flawed

By David J. Weinberg

High-end audiophiles prefer a specific layout of the surround system speakers: the face of each identical full-range speaker (regardless of number—typically five, but can be more) is located tangent to a circle whose center is the listener's sweet spot. This puts each speaker the same distance from the listener and, assuming room effects are sufficiently damped, results in the same perceived sound quality from each speaker.

Assuming the engineer has produced a surround recording that is in sync channel-to-channel, this speaker setup results in a time-coherent sound field at the listener's head, reproducing the recording engineer's intent. However, this is only effective for a single listener, and the playback system *owns* the room.

While some recordings put instruments in the surround channels for artistic reasons, most surround recordings, particularly of classical music, use the surround channels for hall ambience. It is that concept I address.

Most of us interested in surround sound prefer an installation that allows us to use the room for multiple functions, to move about during playback, and to share the experience with others not *all* sitting in our laps. This means that for those surround recordings intended to bring the listener into the concert hall, the spatial effects (including the implied location of instruments on the "stage" and the recording hall's ambience characteristics) need to vary appropriately with the listeners' positions relative to the speakers.

It is well known that amplitude-panning and precedence effects between speakers are far from perfect in producing a phantom image. However, with fairly dry listening room acoustics—and full-range speakers with a fairly flat frequency response that is uniform versus horizontal angle—imaging can be adequate over a reasonably large listening area, particularly when you consider that in a concert hall, perfect imaging fades quite close to the stage.

Ambience is another matter. Consider a choral, chamber group, or organ performance in a fairly reverberant location, such as a church (to make the effect more obvious). Once the listener gets beyond the critical distance into the far field, the spaciousness of the reverberant (surround) sound is quite consistent over a very large area (naturally, next to a wall is an exception).

Thus, there is a fundamental flaw in surround sound as implemented for the home. As the listener moves toward a surround speaker, its level relative to the front speakers increases (as it should, but usually by too much), and its location becomes identifiable. This is clearly because there are so few sources of surround channels' sound, and their location allows the listener to get too close to them.

For some installations, an imperfect solution I have found is to place the surround speakers on the floor in the rear corners, facing up (since most people use smaller speakers for surround channels, this enhances their bass output, and, to me, reverberant bass is critical to a sense of space). This enlarges the area over which the surround-generated ambience is more uniform and the speakers less localizable, since most of their sound must bounce off the ceiling and walls before it reaches the listening area. Putting the surround speakers near the ceiling might also partially solve the problem, but I believe the typically 8′ ceiling height isn't sufficient, and I prefer bass reinforcement to include the floor.

A better solution is found in *properly implemented* surround sound in movie theaters (not found very often), which uses a much larger number of sources (even though each surround channel shares its signal among a group of speakers) and locates them such that the listener can't get close enough to an individual speaker to identify it as a source. This is generally difficult to accomplish in the home environment, but can be approached by using several identical speakers spaced along the back and side walls for the surround channels. By proper series/parallel connection, each of these arrays will present an acceptable load to the amplifier.

For now, there is no reasonably-priced truly effective solution. Research by experts such as Tomlinson Holman (www.tmhcorp.com; whose 10.2-channel system addresses this and other issues) will likely improve the situation in years to come. ❖

David J. Weinberg (Tobias Audio, Silver Spring, Md., (301) 593-3230; WeinbergDa@cs.com) is an engineering consultant and technical journalist in audio, video, and film technology. He provides audio and home theater engineering consultation and professional on-location digital recording services to companies, radio stations, and individuals. He brings to his work an MSEE, a first class Radiotelephone license and over 35 years of continued study and active involvement in the audio, video, and computer industries. He is a member of the Audio Engineering Society and of the Society of Motion Picture and Television Engineers. David has authored a number of articles on various phases of audio for video and film, and is also the Boston Audio Society's (www.BostonAudioSociety.org) membership officer and editor of its journal—"The B A S Speaker."

Practical MOSFET Testing for Audio

This noted developer investigates the use of MOSFETs in our designs and offers some trade secrets on building better amps. **By Nelson Pass**

The quality of individual parts is a particular concern to audio "do-it-yourselfers" (henceforth known as DIYers). Many of them lie awake at night agonizing over choices of capacitors, resistors, wires, and so on, in the belief that the characteristics of these passive parts greatly influence the quality of sound passing through the circuit. Maybe so.

But what about the active components? If there is anything that the objectivist and subjectivist camps should be able to agree on it is that there is considerable difference between different active gain devices and that they produce measurable, if not audible, results.

As a minimalist, I personally like to work with MOSFETs, and I do so because they give me the most performance with the fewest parts. It is my experience that particularly with simple circuits and minimal feedback, the specific character of the individual MOSFET makes a real difference.

With that in mind, I set out to measure some of these parts to see whether I could select the best for use in the signal path.

WHAT IS A MOSFET?

First off, a MOSFET is an electronic amplifying device, more specifically, a transistor. It is a little part that has three connecting pins coming out of it, known descriptively as the source, the drain, and the gate. Using a hydraulic metaphor, if you think of the MOSFET as a faucet, then the source and drain are what the water flows through, and the gate is the lever that turns the flow on or off, fast or slow. Except, of course, that the water is electrons, and the lever is operated by electronic charge.

Continuing the water analogy, assuming that there is some water pressure (voltage) across the faucet (transistor), then water (electronic current) will flow through it when the valve is turned on (the gate is charged with voltage). In the case of a MOSFET, if there is drain pin voltage relative to the source pin voltage, electric current will flow from the source to the drain, provided enough voltage is placed between the gate and the source.

There are two basic kinds of MOSFETs, the N channel and the P channel types, differing by the voltage polarities they work with. In the case of an N channel MOSFET, you can make a fine operating example (that you can do yourself) by attaching the source pin to circuit ground, a positive voltage source to the drain through a current meter, and a variable positive voltage source to the gate—in this case a potentiometer from V+ to ground with the wiper at the gate (*Fig. 1*).

When the potentiometer is at full counterclockwise, the gate voltage is at 0V (grounded) and no current flows through the transistor. As you turn the potentiometer clockwise, the voltage on the gate becomes positive with respect to the source pin, and current starts to flow from the V+ of the power supply to ground through the transistor. The higher the gate voltage, the more current there is. The less gate voltage, the less current there is. The point of conduction varies from device to device, but current will start flowing anywhere from about 2 to 5V on the gate.

Practical lesson 1: if you turn the potentiometer sufficiently clockwise, you may hear a popping sound and smoke will come out of the MOSFET (if you're lucky). For this reason, I recommend that you keep an eye on the current meter when you try this.

The P channel device works the same way, but with the drain at V- instead. The existence of both N and P polarity devices in transistors is very handy, and gives them a big flexibility advantage over tubes, which have only the "N" characteristic.

THE THREE WAYS

As with the other three pin gain de-

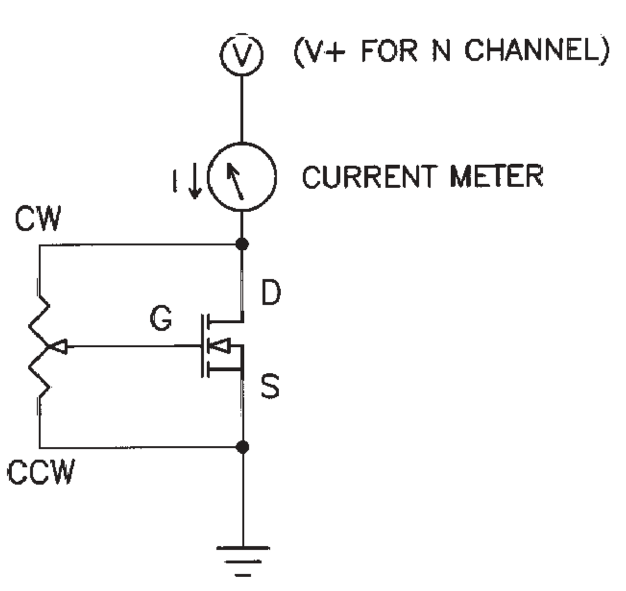

FIGURE 1: Simplified test setup.

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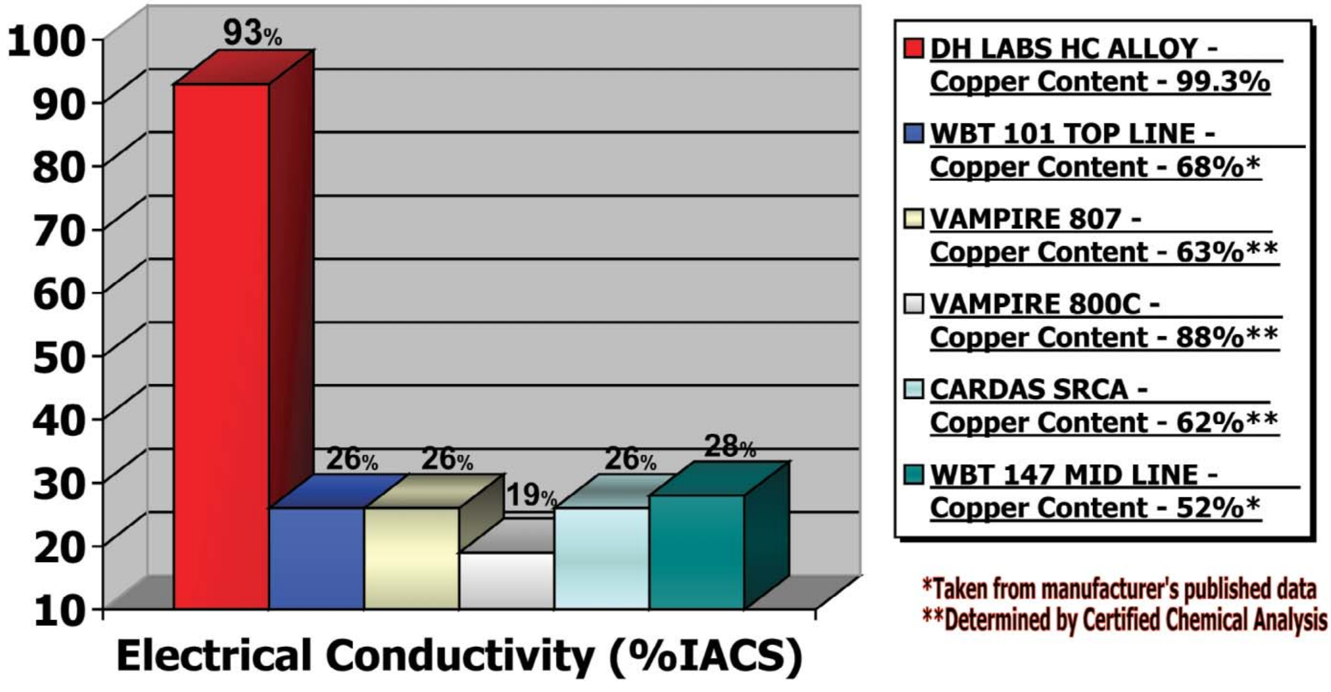

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vices, there are only three ways to hook up and use a transistor. With a MOSFET they are known as common source, common drain, and common gate.

Figure 2 shows three examples. In the common source example the signal goes into the gate, the source is grounded, and the signal comes out the drain with its polarity reversed and both its current and voltage amplified. The input impedance is high, and the output impedance is essentially the value of the drain resistor.

Notice in *Fig. 2* that I have gone to the trouble to indicate the phase of the output signal with a little "sine wave" icon next to the output. Of the three ways, only common source reverses the phase.

Common drain, also known as a source follower, is where the input signal goes in the gate and the source pin follows it, providing current gain, but not voltage gain. The output voltage is almost the same as the input voltage. The input impedance is high, and the output impedance is fairly low, being the inverse of the transconductance figure of the MOSFET.

Common gate, usually seen as a "cascode" connection, is where the signal goes into the source pin and comes out the drain. This connection has no current gain—the output current equals the input current. It can produce output voltage gain in phase with the input. The input impedance is the inverse of the transconductance of the MOSFET, and the output impedance is the value of the drain resistor.

Of course, the transistor itself has no idea how you are using it. It blindly reacts to the variations in voltage and current on its pins without any true knowledge of its place in the scheme of things. We all experience something like this now and again.

If you reverse bias on the drain and source, the ordinary MOSFET will behave like a silicon diode. That is to say, if you take an N channel device and put positive voltage on the source and negative on the drain, current will flow, and you'll see something like 0.7V drain to source. P channel devices, of course, do the same thing with opposite voltage polarity.

WHAT CHARACTERISTICS ARE THERE?

MOSFETs come with numerous numbers describing their characteristics:

- Operating Wattage (chip temperature)—How hot can you run it?
- Maximum Voltage (pin-to-pin)—What voltages between pins will break the chip?
- Maximum Current (drain-to-source)—What high current will melt the connections?
- Current vs. gate-source Voltage (Vgs)—What is the specific gain character?
- Gate Capacitance—The input gate resistance is nearly infinite, but the gate capacitance is substantial and often nonlinear.

There are tons more numbers, enough to fill four to ten pages with lists and charts and graphs for each type of device. International Rectifier seems to be the dominant supplier of power MOSFETs, so check out the information at their web site (www.irf.com).

So what characteristics are important? Look at the data sheets of one of my favorites, the IRFP240, an N channel power transistor in a plastic case. First off, it's rated at 200V drain-source (that's good), a maximum current of 20A (also good), and a maximum wattage dissipation of 150W (case temperature at 25° C). The maximum voltage between the gate and source is 20V. These are important specs for selecting such a device, but you can pretty much figure that you aren't ordinarily going to be testing them, as you are not usually looking to break the part.

As you go down the long list of numbers, some of interest for linear audio stand out: Vgs, the "threshold gate voltage," where conduction starts (anywhere from 2 to 5V); the transconductance (somewhere between 0 and 12, depending); and the gate capacitance (anywhere from 1200 to 2400pF, also depending). These figures vary with voltages, current, temperature, and also by device. (I would reproduce these figures and curves here, but there isn't that much space, and I'd just as soon not have to deal with IR for permission). The transconductance is measured in siemens, and is simply how many more amps flow from drain to source for each additional volt on the gate-source pins.

The gate capacitance is important, for while the input impedance looks close to infinity at DC, there is what looks to be a capacitor of quite a few picofarads attached from gate-source and gate-drain; both of which start produc-

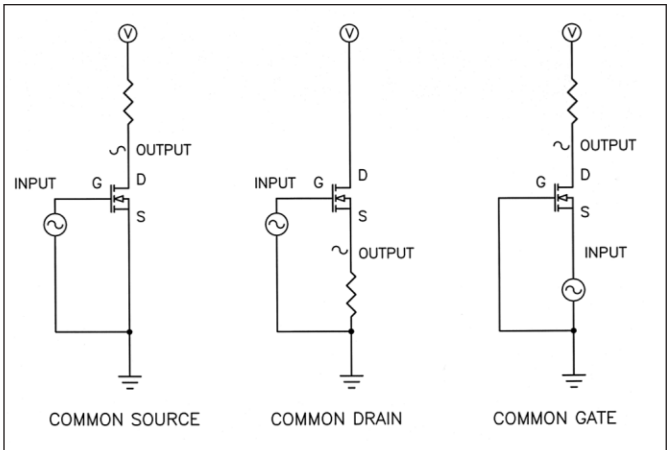


FIGURE 2: The three ways to use a MOSFET.

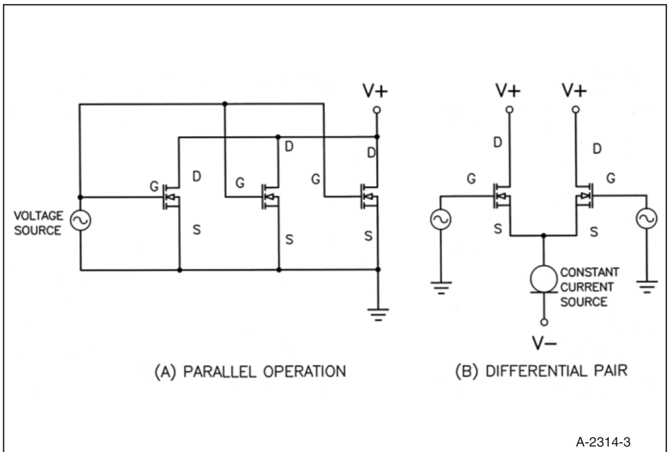


FIGURE 3: Circuits that like matched parts.

ing a load at high frequencies for whatever is trying to drive the gate.

Distortion? Not mentioned.

It is clear, however, that if you want to know what the performance will be for a given device in a circuit, you'd best be measuring it yourself, not only because the manufacturer is not likely to be duplicating your exact circuit, but also because there is often wide variation between devices.

WHY MATCH?

Often the reason for measuring MOSFET characteristics is to match devices to each other. Most often this consists of measuring the Vgs of the devices, and occasionally extends to measuring the transconductance.

In the case of a power output stage, you often want to parallel power de-

FIGURE 4: Test everything circuit.

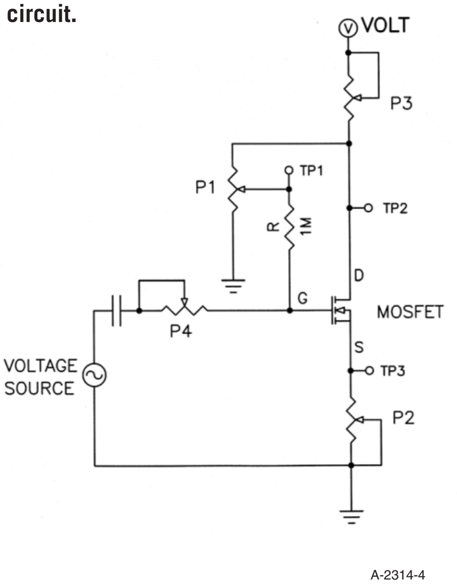
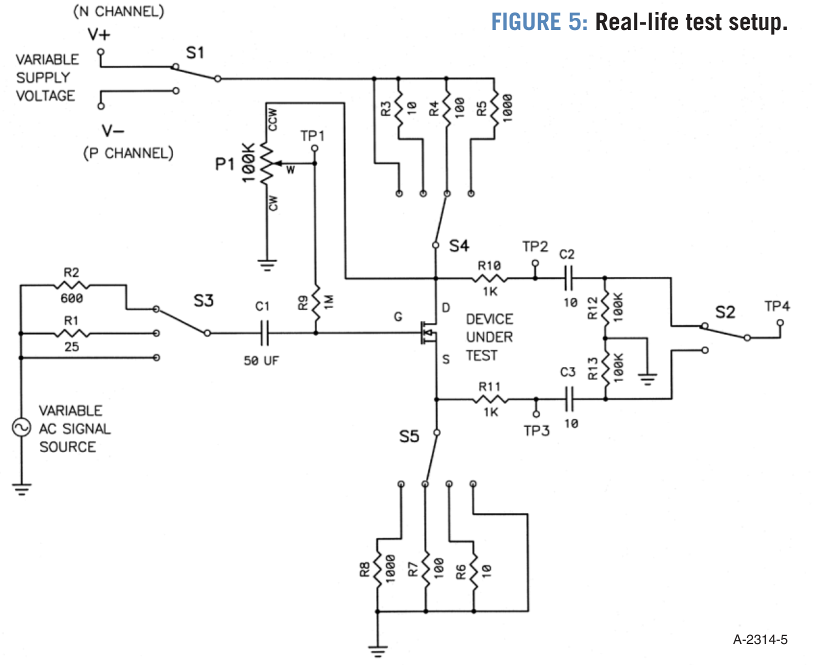


FIGURE 5: Real-life test setup.



vices to share the current and heat dissipation, as in *Fig. 3a*. Here it is desirable that the devices be identical so that the current is shared equally. If the Vgs of each device is different, some device will find itself doing the disproportionate share of the work, while some other device might not be turned on at all. Even if the Vgs is matched at a given current for all devices, they might vary at different currents, so you either want to sample and match Vgs at different current and temperatures, or perhaps simply measure and match the transconductance at the operating point and cross your fingers.

A popular way around this is to match Vgs between devices having the same manufacturing lot code, meaning they were made at the same time under the same circumstances. This technique is very effective, and if you sequentially measure the Vgs of a batch of smaller chips of the same lot code, you can actually visualize the position of the particular chip on the silicon wafer by the Vgs pattern that develops. These devices are usually extremely well matched in all characteristics.

The other part of a typical amplifier circuit where matching is important is the circuit of *Fig. 3b*, the "input differential pair," which compares the input signal to the amplifier's output signal and amplifies the difference. This is known as feedback. Matching these parts allows for low DC offset at the amplifier output, good characteristic tracking over a range of temperatures, and (you hope) lower distortion. I say "you hope" because ideally you want the distortion characteristics of the matched pair to cancel exactly. You are well advised to not assume that matching will actually do this.

Matching is a key to performance in monolithic circuits such as op amps, where it can be conveniently assumed that transistors made next to each other will be very similar. To some extent this is used to lower distortion, but it appears that mainly the intent is to retain constant DC offset and other characteristics over a range of temperatures.

CIRCUIT FOR MEASUREMENT

First you must decide what you're interested in—being regular guys who want to build up the best possible circuit without producing an expensive laboratory.

You should trust some of the numbers the manufacturer gives, such as the maximum voltage, current, wattage, and temperature ratings. These are not difficult to test for, but the average DIYer isn't going to want to burn up a bunch of parts to get this information. We trust the manufacturer, but usually we divide all his numbers by two, unless they're bad numbers, in which case we multiply by two. This approach aids your success and satisfaction, and improves relations with transistor vendors, all of whom are probably doing the best they can.

So what will you test? The Vgs for turn-on is my first thought. It's necessary, easy to do, and therefore popular. It would be nice to see what the gain figure really is under a particular set of circumstances. While you're at it, it would be nice to see the input capacitance inferred from the high-frequency rolloff of the gain, also under varying circumstances. Also, you might as well measure the distortion vs. both frequency and amplitude.

You will want to test in both common source and common drain modes, which make up the bulk of usage where you care very much about quality and matching. You can ignore measurements for the usual common gate mode of operation, inferring their values from the others, and noting that as Cascode devices, they contribute little of their own character to the circuit performance.

Figure 4 shows a simplified circuit for measuring some of the practical characteristics of MOSFETs with an eye toward their use in audio circuits, and it will allow measurement of the characteristics just identified. The ranges of values of the adjustable resistors in this circuit are from 0Ω to infinity, and of course this covers just about any contingency. The DC and AC voltages to be measured would be those at the three pins of the MOSFET and the supply voltage, and from these everything can be calculated. The capacitor must be a low leakage film type.

MEASURING VGS

Suppose you simply want to measure the DC value of Vgs at a given current. Referencing *Fig. 4*, you must set P1 full clockwise. Adjust P3 and the voltage at (V) for the current you want and measure the DC voltage between the drain and ground, noting and trimming the voltage across P3 to ensure that the correct amount of current is flowing.

P4 doesn't matter here, as the DC is blocked by the capacitor, but it is always nice to have a few ohms in series

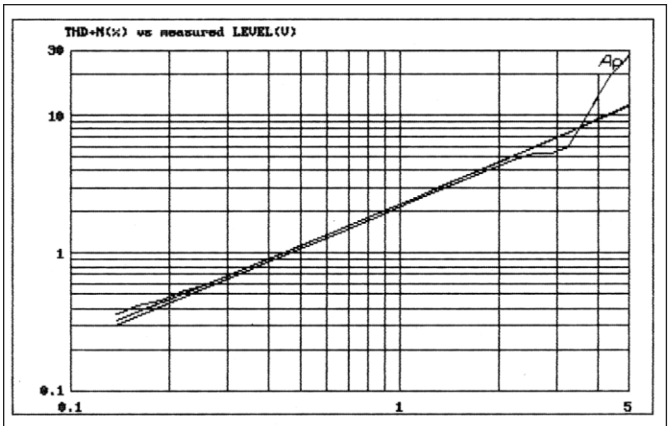


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FIGURE 6: **IRFP240 distortion/amplitude at 5V at 1A.**

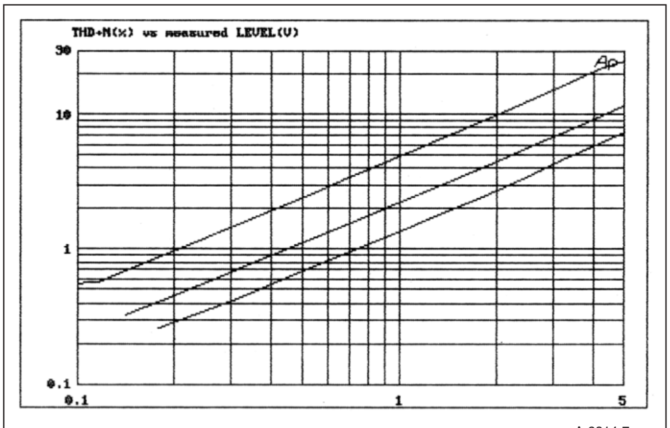

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FIGURE 7: **IRFP240 distortion/amplitude at 10V at 0.5, 1.0, 1.5A.**

with the gate of any MOSFET to prevent high-frequency (parasitic) self-oscillation, which will alter the voltage reading. Of course, if you are simply trying to match Vgs of various devices, you only need to trim P3 and (V) once, and then simply group devices that match within the tolerance you want.

Most of the time you are looking for less than 0.1V variation in Vgs between devices, and from experience, you'll see that it is fairly easy to get 0.01V variation if you have a reasonable population of transistors to work with. In production quantities, you get this all the time. If you have only a few devices, you might have to settle for the 0.1V matching figure.

The Vgs is temperature dependent, which means that the parts tested should all start out at the same temperature, usually room temperature. (I mention this because at Pass Labs we don't heat our inventory area much in the winter, and we must let the devices sit in the test area to warm up to room temperature.)

As you run the test, you must be aware that you are heating the devices both by running some current and voltage through them, and also by touching them, and the Vgs will alter with temperature. Ideally, you test the devices under conditions identical to the intended use, and this means mounting the devices in thermal contact with a heatsink of the appropriate temperature and allowing for the temperature to settle.

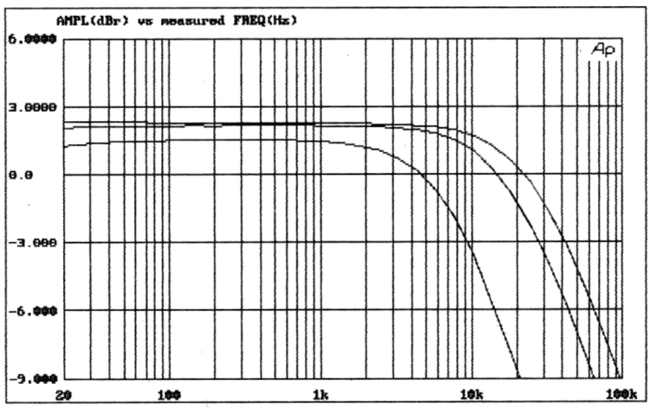

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FIGURE 8: IRFP240 transconductance/frequency at 1A at 5, 10, 15V.

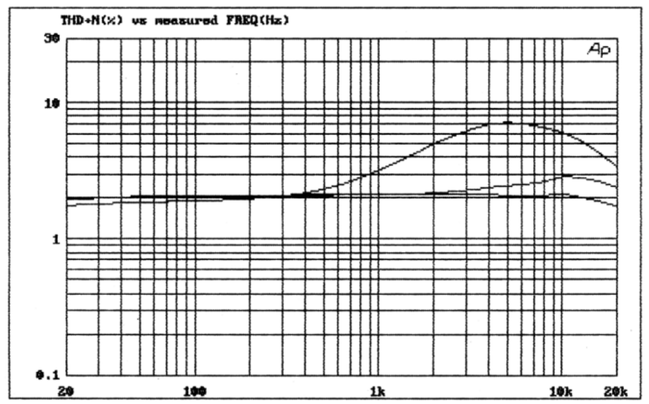

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FIGURE 9: IRFP240 distortion/frequency at 1A at 5, 10, 15V.

MEASURING TRANSCONDUCTANCE

Using the circuit of *Fig. 4*, you can measure transconductance under arbitrary conditions. In common drain mode, if you short P3 and measure across an arbitrary DC and AC value for P2 with a given input signal and supply voltage, you can compare the AC variation of gate-source voltage versus the output voltage. The Vgs variation divided by the current through P2 gives the transconductance, so you would be dividing the AC current occurring across P2 (which is the value of the voltage across P2 divided by the AC voltage), and you would be dividing this by the AC voltage which appears across the gate and test point TP3. TP1 is used only for DC gate measurements, being connected through 1MΩ.

In a common source setup, we often measure the transconductance using the voltage appearing across a resistive load:

Transconductance (siemens) = Output volts/Input volts/drain load

P1 is adjusted to set up the proper DC bias for the circuit. In common source mode, you short P2 and measure across P3, comparing this voltage to the Vgs, once again setting P1 and (V) to the appropriate bias values. The AC current through P3 divided by the AC voltage of Vgs is again the transconductance figure, and once again you can vary P5 to observe the effects of the apparent input capacitance.

Computing the apparent input capacitance is easy enough. Find the frequency where the transconductance drops 3dB (=0.7 times the low frequency figure) and use the formula

$C = 1/(R * F * 2 * \pi)$

C is in farads

F is in Hz

R is the sum of ohms of R5 and the AC source

$\pi = 3.14.$

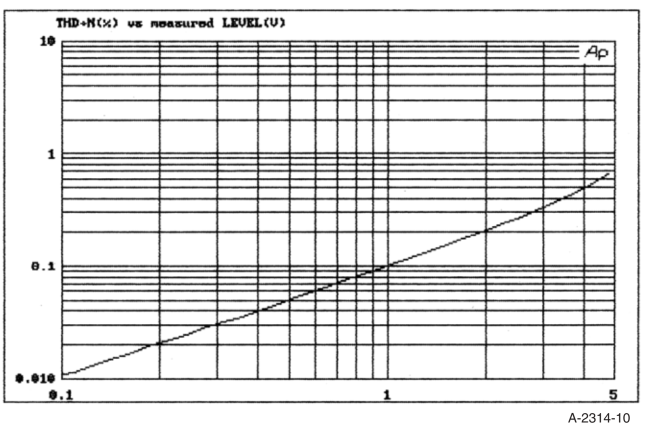
For example, at low frequencies the transconductance might be 10 at 100Hz, and if we find it is 7 at 100kHz with a 1kW AC source impedance, then

$C = 1/(1000 * 100{,}000 * 2 * 3.14) = 1.6 * 10^{-9} = 1.6\ nF = 1{,}600\ pF$

MEASURING DISTORTION

Note that all the parameters you measure will depend on the conditions the transistor experiences—all the voltages, currents, temperature, and frequency. You will find that transconductance and capacitance are among these, and these are crucial to audio performance.

That these figures vary in value gives rise to distortion in the transistor. You can view all distortion in this manner—variations in gain or loading, which in the case of a MOSFET is the transconductance, and input capacitance. The input capacitance is a particular problem in power MOSFETs where the capacitance can get fairly high and varies with voltage and current.



A-2314-10

FIGURE 10: IRFP240 distortion/amplitude (com drain) at 10V at 1A.

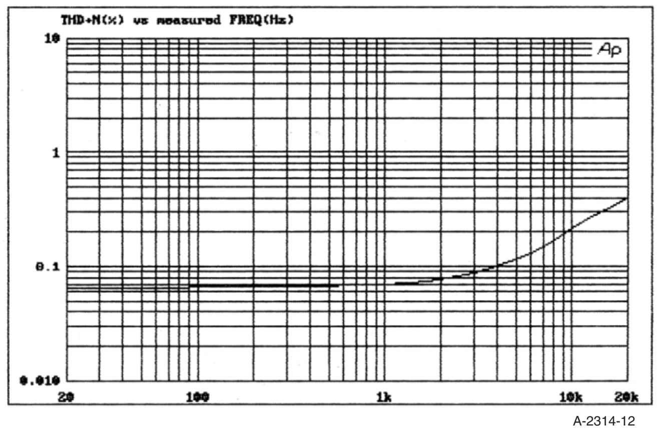


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FIGURE 12: IRFP240 distortion/frequency (com drain) at 10V at 1A.

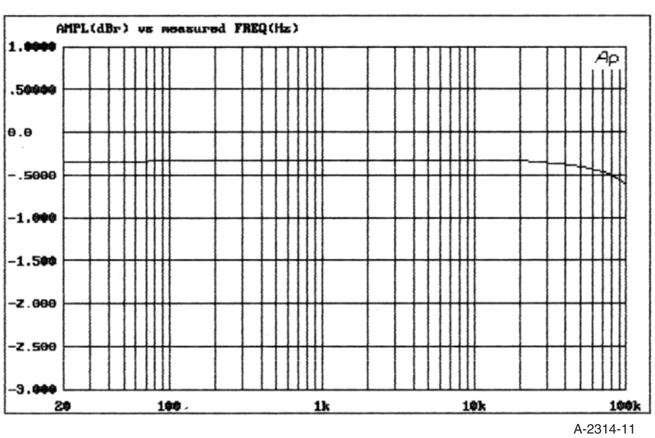

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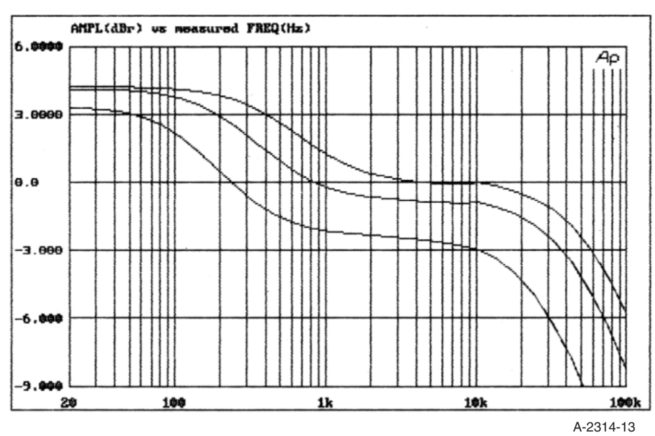
FIGURE 11: IRFP240 amplitude/frequency (com drain) at 10V at 1A.



A-2314-13

FIGURE 13: IRFP9240 amplitude/frequency at 5, 10, 15V at 1A.

Measuring distortion is like measuring the transconductance, except that you also look at the output harmonic distortion versus level and frequency. As you would expect, the distortion, which is primarily second harmonic, increases with amplitude and frequency. The amplitude-dependent distortion comes from the variation in transconductance versus voltage and current, and the frequency-dependent distortion is the result of nonlinearity in the input capacitance.

INTRODUCING FIGURE 5

Figure 5 shows a practical version of the circuit of *Fig. 4* using switches and power resistors where real current is likely to flow instead of potentiometers. It assumes the use of a variable regulated supply designed to provide the voltages and currents you might find interesting. If you have a dual voltage supply, it makes the use of S1 convenient in reversing the supply polarity. Otherwise, you must use a two-pole switch to reverse both supply terminals.

If you don't have an Audio Precision™ test rig, you can use an ordinary audio oscillator, wideband voltmeter, and distortion analyzer, but of course it won't be nearly as much fun. The drain resistance is varied by S4 from 0Ω to 1kΩ, similarly for the source resistance. Other niceties include some output capacitance and resistance for connection to the input of an Audio Precision test rig. If using the Audio Precision, you can dispense with S3 and R1 and R2 since the AP has 25 and

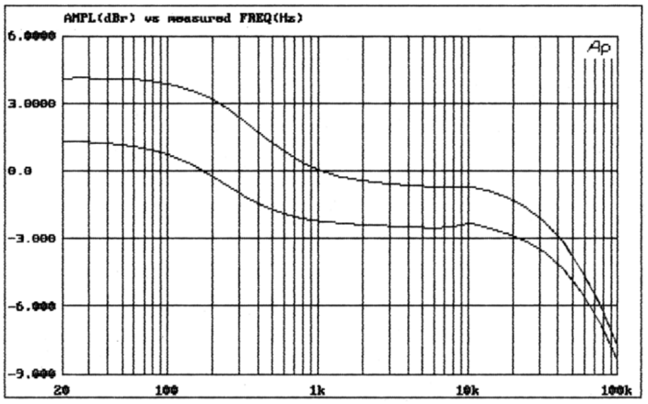


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FIGURE 14: IRFP9240 amplitude/frequency at 0.5, 1.0A at 10V.

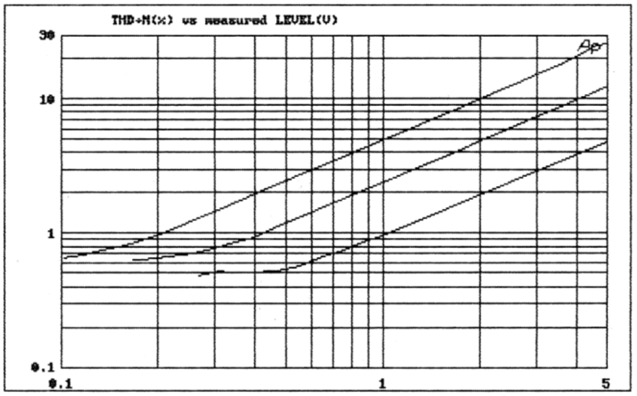

FIGURE 15: IRF610 distortion/amplitude at 0.05, 0.1, 0.15A at 10V.

A-2314-15

600Ω selectable source resistance.

For common source testing, S5 is set to 0Ω and S2 is set to the drain of the MOSFET. For common drain testing, S4 is set to 0Ω, and S2 is set to the source of the MOSFET. P1 is used to provide DC bias to the gate. Test points TP2 and TP3 are isolated through 1k resistors to avoid parasitic oscillation of the MOSFET when voltmeter probes are used to measure DC junction voltages.

IRFP240 MEASUREMENT

Big power MOSFETs were tested with 10Ω on the drain, medium (TO-220 type) devices were tested with 100Ω, and small plastic types (TO-92, for example) were tested with 1kΩ, as befitting their probable use.

First up on the block is the venerable IRFP240, a 200V 20A power transistor of which I personally use lots. I set the drain resistance at 10Ω and measured the Vgs for three random samples with Vgs of 4.64, 4.52, and 4.66V. Two of these had the same lot code.

Figure 6 shows distortion versus output amplitude of one of the three samples of IRFP240, taken at 1kHz. There is a considerable similarity between the parts, the ones with the same lot code being virtually indistinguishable from each other.

On this curve the distortion is quite high at the higher amplitudes because the drain is both biased at 5V, which is quite low, near Vgs, which is about 4.6V. As a result, the circuit naturally clips at less than about 3V rms (±4.2V peak). This is a low voltage, atypical of audio use, and so *Fig. 7* shows P1 set for 10V across the drain-source, but a family with increasing current from an increased power supply voltage. It is clear that the distortion plummets when offered greater bias current.

At voltages above a few volts, the distortion figures for these MOSFETs tend to converge, meaning that for low frequencies, the drain-source voltage does not heavily influence the transconductance. What does have a heavy influence is the amount of DC bias current through the device. *Figure 7* clearly shows a rough inverse proportionality between distortion and bias current. This particular device has a low frequency transconductance of about

Transconductance (siemens) = Output volts/Input volts/drain load
Transconductance = 1.39V /0.050V/ 10Ω = 2.78

Meanwhile, *Fig. 8* clearly shows that this transconductance will experience a high-frequency rolloff determined by the resistance of the driving input coupled with the apparent input capacitance. In *Fig. 8* we see curves associated with the previous test, but with a constant 1A bias, 600Ω signal source, and with drain-source DC bias held at 5V, 10V, and 15V. The capacitance of the input declines with increased D-S voltage, dramatically at lower voltages, and with declining returns as the voltage goes above about 15V or so.

The capacitance comes from two main sources, the gate-source capacitance, which is high but sees lower voltage, and the gate-drain, which is lower in value but often sees higher voltages. For *Fig. 8*, a 6kHz rolloff with 600Ω im-

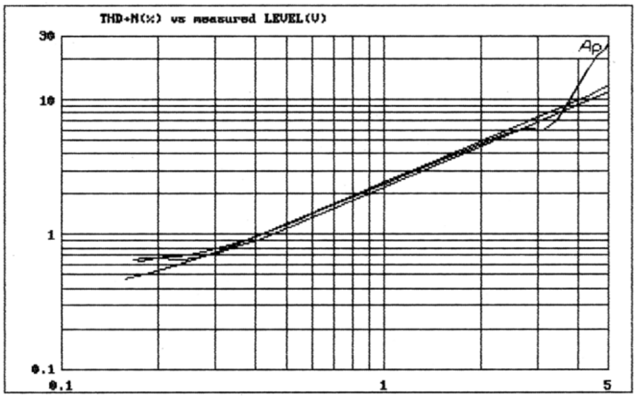

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FIGURE 16: IRF610 distortion/amplitude at 5, 10, 15V at 0.1A.

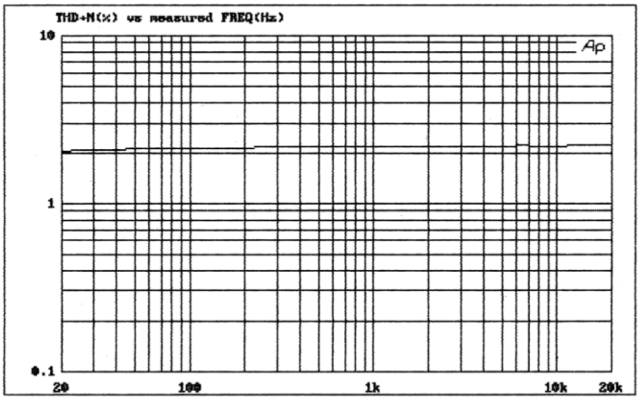

FIGURE 18: IRF610 distortion/frequency at 10V, at 0.1A. A-2314-18

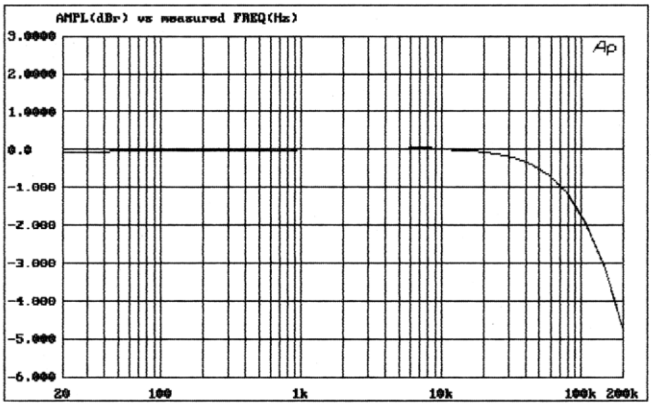

FIGURE 17: IRF610 amplitude/frequency at 10V, at 0.1A. A-2314-17

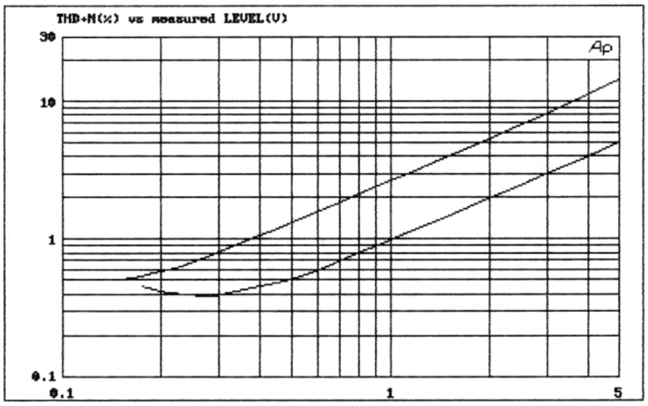

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FIGURE 19: ZVN3310 distortion/amplitude at 5, 20V at 10mA.

plies around 44nF. This is the same as 0.044μF, or 4400pF. If you increase the voltage across the MOSFET to 15V, this figure is divided by about 5, and now looks more like 8800pF.

The calculation: $C = 1/(R * F * 6.26)$
where C is in farads
R is in ohms (the total impedance of the input source)
F is frequency in Hz

When you compare these results to the International Rectifier data book, you see a quoted figure of about 1500pF at 10V for gate-source, and a gate-drain of about 300pF. The voltage gain at the drain of this circuit multiplies with apparent gate-drain capacitance. As this figure is about 28, you expect 300pF × 28 to be the contribution from that connection, which would be 8.4nF. Adding 8.4nF to 1.5nF totals about 10nF, which is close enough.

The drain-source voltage figures into this heavily, too. *Figure 9* clearly shows the distortion versus frequency improvement to be had when you have a higher voltage supply. Here you see three curves, all 1A bias and voltage output into 10Ω. The good-looking two curves are at 10 and 15V across the device, and the not-so-good-looking curve is taken at 5V. If you compare distortion with higher currents through the device, you see minor improvements at high frequencies.

What does this say? If you want the transconductance to be linear, you run a high bias current. If you want more bandwidth out of the MOSFET, give it more voltage (and a lower impedance source). If you want both, run the device hot.

You might get the impression from these curves that an IRFP240 is a high distortion device, but in fact we are not necessarily showing it in its best light. *Figure 10* shows the distortion vs. amplitude for the part in common drain (follower mode), *Fig. 11* shows the frequency response of that output, and *Fig. 12* shows the distortion curve vs. frequency; the latter two curves taken at 1V input and a 10Ω load. These are good figures. Want to see bad? I could show you some bipolar or tube curves worse than these.

THE IRFP9240

As a concept this part is supposed to provide an inverse polarity complement to the IRFP240, and it more or less does this. The voltage and current ratings are similar, the transconductance is somewhat less, but otherwise it looks fine. Until you run a frequency response curve on it. *Figure 13* shows the surprising result. The transconductance dips down as you approach the midrange. Then it shelves off flat for a bit, and proceeds on its ordinary decline at high frequencies.

Where is this from? Who knows? The effect does decline with higher voltages as you see in *Fig. 13*, which has 5, 10, and 15V across the drain. Higher current does not reduce it, as you see in *Fig. 14*, taken at 0.5 and 1A currents. How bad is it? I've seen worse. Oddly, the Harris version of the same part does not show this.

Does this mean you should avoid the P channel IR parts for linear use? No, I don't think so—I've been building commercial amplifiers with the emphasis on N channel parts for many years, but I still find the need for some P channel components, and I have not had many difficulties with this. You need to keep the characteristic in mind and work around it, and you also need to remember that these curves represent some worst-case test scenarios. Having had the opportunity to compare IR versus Harris in real amplifiers, you have seen advantages and disadvantages both ways, and so don't become too excited about it. If you want to make some lemonade out of this particular lemon, use it to design an amplifier with a little more control to the bottom end of an amplifier and less feedback over the mid and high frequencies.

IRF610

If the IRFP240 is the popular "poppa bear" of the MOSFETs, then the IRF610 is the "momma bear." Housed in a TO-220 package, it does medium power duty at 200V, 3A, and 40W. It was tested with a drain resistance of 100Ω.

Figure 15 shows a family of curves of distortion versus amplitude for a 10V drain voltage and with current at 0.05A, 0.1A, and 0.15A. Of course, the lowest distortion comes from the highest bias current.

Figure 16 shows a family of curves of distortion versus amplitude with the drain biased at 5V, and also 10 and 15V. Not a lot of difference, as with the larger IRFP240.

Figure 17 shows that the bandwidth with a 600Ω source gives a 150kHz rolloff, with a 1V output and a voltage gain of about 30 (transconductance = 0.3), and by our previous example calculation implies about a 1.7nF input capacitance. Comparing this to the rated 140pF gate-source capacitance, and 30× the rated 9pF drain to source capacitance, you surmise that perhaps this sample offers more input capacitance than the "typical" specification.

Figure 18 shows the distortion versus frequency for this part at 10V, 100Ω, and 1V out. The distortion is the same across the audio band. The IRF9610 exhibited the same sorts of variation vs. frequency as you saw in the IRFP9240. This was the IR part, and I did not have a Harris example on hand to test.

ZVN3310

Now you come to a nice small signal-type MOSFET, the Zetex ZVN3310. You will be loading the drain with 1kΩ. *Figure 19* shows the distortion vs. output for two 10mA current bias and 20mA bias. Gee, the distortion goes down with more bias!

The transconductance is about 0.03 or so, and the bandwidth makes it out to about 300kHz or so, which surprises me as a bit low, probably an artifact of our setup. The distortion is flat across the band (yawn). Nice part, and the complement ZVP3310 looks about the same.

CONCLUSION

So you've seen how to do some tests on MOSFETs, and that their performance is limited and varies from part to part. This variation is much more than what you see when you measure passive parts such as resistors and capacitors, and deserves much more attention.

Commercial linear amplifier manufacturers can't take the time to carefully measure each component—they depend on high gain and feedback to smooth out the bumps in the performance of individual parts. Only the high-end designer or hobbyist might put the time into individually testing active devices.

If you are building only one amplifier, you can set up the circuit to substitute parts in and out and evaluate the performance of the permutations. Whether by measurements or listening, if you know what you want when you get it, this approach will work for you.

If you are building one amplifier and don't want to take the infinite-number-of-monkeys-with-typewriters approach, you measure a population of parts and pick what appear to be the best ones. This practical procedure works well, but misses the occasional symbiotic relationship that can occur between two particular parts.

For example, the notion that matched parts is the best approach doesn't always hold up. We have had success swapping parts in and out of production amplifiers at random until we achieved the best performance, then soldering those in permanently. Doing this can often result in much less noise and distortion, if you judge with a distortion analyzer. Alternatively, it can give you better subjective sound, if listening is your preference (and you have lots of time!).

As a manufacturer, I don't design with building only one amplifier in mind, so I like to proceed as follows: Obtain a population of parts for the amplifier and measure each, cataloging the results. Then swap the parts in and out of the prototype, two channels at a time, and measure and listen. And continue to listen until someone complains about time-to-market.

Once you have figured out what combination (or sets of combinations) of characteristics you want, it's not that tough to select them for production. It takes time and energy, but saves time at the test bench, and your customers get a more consistent product—one that sounds more like the prototype on which you lavished all that attention.

What about the parts that don't make the cut as signal bearing devices? They make good active constant voltage and current regulators, or you can sell them to your competitors. Is this approach applicable only to MOSFETs? Of course not. It works with any part, but in particular has great value with active gain devices of any sort—those parts that suffer the most variation and are most important to the sound.

One last time: the active devices with which we build amplifiers have more influence over the sound than the quality of the passive parts, and also have much wider variations between pieces. The emphasis in the marketplace on the differences between resistors, capacitors, and wire is misguided if you ignore the transistors or tubes. If you want a better amplifier, measuring and selecting the individual gain devices is a good place to start. ❖

The DR250a Horn

High performance, power, and portability—it's all here in this author's latest in his line of horn cabinets.

By Bill Fitzmaurice

Last spring—out of the blue—I received an e-mail from someone who was starting a new loudspeaker company and wanted me to be his designer. He'd seen my Siamese Snail project (*SB* 2/99, p. 26), but he had not seen my DR series of horns. When I sent him copies of *audioXpress* containing the DR projects, he was just about ready to go with those designs as-is. I wasn't. As good as the DR cabinets were, I knew that I could still make a few improvements.

The DR250a (*Photo 1*) is the first of the prototypes for a new line of loudspeakers tentatively to be marketed under the Delta Sounds brand. Based upon the DR10a (*aX*, June 2002, p. 16), the DR250a is suitable for electric bass, keyboards, or PA. It's an exceedingly powerful, yet small and lightweight box, transportable in the back seat of a sedan. The "250" refers to its 250mm (10″) woofer, while "a" refers to its intended use as a backline or club-sized PA cab. Two notable features differentiate it from the DR10a.

NEW FEATURES

First, there are no parallel surfaces within the cabinet whatsoever. Parallel surfaces cause phase anomalies that hurt response, so the flat panels that make up the throat horn sides and the cabinet top and bottom are angled in a trapezoidal arrangement. The result is an improvement in the woofer response curve of the DR250a compared to the DR10a (*Fig. 1*), with an additional 1kHz of usable bandwidth. Another benefit of this configuration is that the box is angled upward when placed directly on a stage, for better intelligibility in the backline mode, while multiple cabinets can be splayed when in a line array mains arrangement.

The second alteration to the older design involves the tweeter array. Traditional tweeter placement, with elements aimed on-axis, gives falling high-frequency content as you move off-axis. For instance, the beamwidth (–6dB from on-axis) of a typical high-end pro-sound speaker, the EAW™ KF850, ranges from 86° at 500Hz to 60° at 2kHz to 51° at 16kHz, for an average nominal horizontal dispersion angle of 55°.

PHOTO 1: The completed DR250a horn.

Theoretically, you could widen the angle of dispersion by using cabinets "clustered" side by side. In fact, the KF850 (along with speakers from virtually every manufacturer) has a trapezoidal shape intended for just that purpose. However, this places mid- and high-frequency elements on a horizontal plane, a situation guaranteed to cause phase-sourced comb filter effects, along with excessive vertical dispersion that leads to early reflections off the floor and ceiling. While horizontally arraying mid- and high-frequency drivers intuitively would seem the logical route to wide horizontal dispersion, in fact, the opposite is true.

MEASUREMENTS

The DR250a improves horizontal dispersion with two vertically arrayed banks of tweeters that are aimed inward, cross-firing at a 90° angle. A radial reflector placed at the junction of the two banks minimizes phase anomalies to smooth response both on- and off-axis while adding a measure of horn loading. As the frequency rises, and the distance between the radiating planes of the two vertical tweeter banks approaches and finally exceeds a half-wavelength, phase anomalies cause a reduction in output. To counter this effect a pair of semicircular baffle extensions flank the tweeter array, giving additional horn loading in the upper register.

Phase cancellations finally win out above 10kHz, reducing SPL to below that of an axially mounted configuration, but the additional power required to correct the situation via EQ is minimal, so there is no problem in achieving flat response to 16kHz. The average drop in response at 30° off-axis is only 2dB right out to 16kHz. At 45° off-axis the average drop in SPL is 4dB, and again that figure holds true across the entire audio spectrum. In fact, at some frequencies the SPL is a dB or two higher off-axis than on-axis (*Fig. 2*).

With a –6dB horizontal dispersion angle of no less than 100°, this pro-

sound cabinet has horizontal dispersion characteristics that very few audiophile grade speakers can match. At the same time, HF vertical dispersion is very tight, on average down 6dB at only 10° off-axis, virtually eliminating early reflections on the vertical plane.

The summed result is a box somewhat less than 7ft^3, about 40 lb, with an average 106dB SPL/2.83V/1m output from 100Hz to 10kHz that sounds the same no matter where you sit in the audience. Response peaks from 320 to 500Hz and 3 to 6kHz may be tamed with L-C filters, but in pro-sound applications that chore would be accomplished via EQ, as would the required correction above 8kHz.

Note the two traces in *Fig. 1*, for 50Hz and 80Hz box tuning. Using different duct lengths enables you to tune the box for specific performance. As a stand-alone box, a 50Hz Fb gives maximum bass bandwidth, but if you intend to use a subwoofer for the deep bass anyway, an 80Hz Fb gives higher sensitivity in the mid-bass. For that matter you may forgo the ducts entirely, enhancing response above 100Hz by a few dB at the expense of response below 100Hz. Build it whichever way works best for you.

PARTS SELECTION

The woofer I used is one of the newer neodymium magnet drivers, the PAudio™ SN10-MB. Pertinent specs are Fs 62Hz, V_{AS} 36 ltr, Q_{TS} .33, Z 8Ω, and 200W. Another driver with specs reasonably close should work OK. My SN10-MB came through with an initial Fs of 80Hz, which obediently came down to 66Hz after 24 hours with a 32Hz/8V break-in signal fed to it, and it will probably get down to 62Hz with use. This shows that it's a good idea to test driver Fs and break it in if required.

There are two advantages to neodymium magnet drivers. The obvious one is weight. The SN10-MB, for instance, weighs only 5 lb, which is about half that of the Carvin™ PS-10 in my DR10a, and a quarter that of a premium driver, such as an EV™DLX.

The other advantage is greatly reduced frame size, which is most welcome in the compact DR design. In fact, the driver chamber of the DR250a is even tighter than that of the DR10a, so a driver with a ferrite magnet probably would not fit. The one disadvantage to the PAudio™ is availability, because the company is new to the US and distribution is limited. Searching the Web should turn up a source, though, as more and more outlets pick up the line.

I would have used my usual tweeters of choice, horn-loaded piezos, but my new-found client wanted to try something more exotic: ribbons. Though they have superb sonic characteristics, ribbons have a couple of shortcomings that have kept their use in pro-sound rather limited. For one, they're not as efficient as horn-loaded tweeters, and for another they can be very pricey.

I managed to get around those obstacles by using six Dayton™ Pt2 Planar tweeters, available from Parts Express, which achieved the necessary sensitivity and power handling. At less than $30 this ribbon isn't any more expensive than most dynamic tweeters, though using six of them does add up, and you'll need to shell out for a crossover, too. If you want performance cheaply, one alternative is the CTS™ model KSN1167A, also available from Parts Express. Note that the Pt2 tweeters are very shallow and don't interfere with the woofer frame; with deeper horn-loaded tweeters, make sure you lay them out to avoid the woofer.

Speaking of crossovers, the SN10-MB works fine up to about 5kHz or so in this box, so for electric bass use only, you may opt to forgo the tweeters; if so, alter the tweeter baffle to a flat panel. However, the horizontal dispersion of the woofer horn declines above 2kHz, so if you are using tweeters, crossing over at 2kHz makes sense. That could be a bit dicey with the Pt2, because its recommended crossover is 2.5kHz with a third-order network, so I used a fifth-order high-pass (*Fig. 3*). This is coupled with a third-order low-pass on the woofer, which keeps phase relationships between the woofer and tweeters correct. If you use piezos, which require no crossover, the recommended KSN1167As are a good match, because they don't kick in much below 4kHz and won't suffer from excessive crosstalk with the woofer output.

I've found that horn-loaded boxes magnify response irregularities more than direct radiators, and that making the cabinet interior sonically inert is an absolute must. In this box I used two types of damping material—fiberglass and polyester. The fiberglass is a thin (⅝″), high-density variety that is used for backing 2′ × 4′ ceiling tiles, which I used to line those areas of the cabinet other than the driver chamber. For the

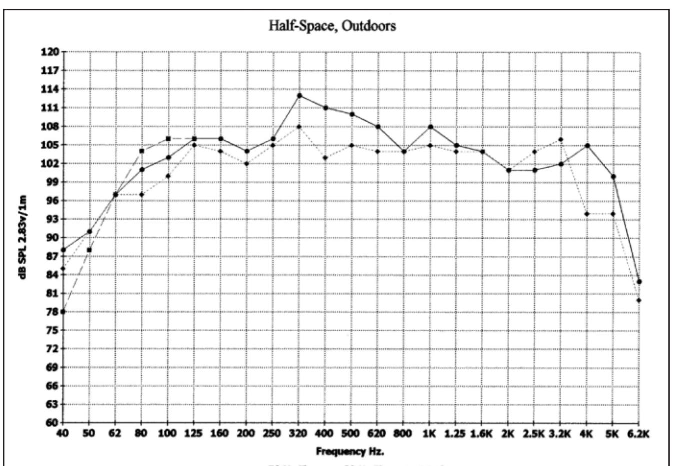

FIGURE 1: DR250a woofer SPL response.

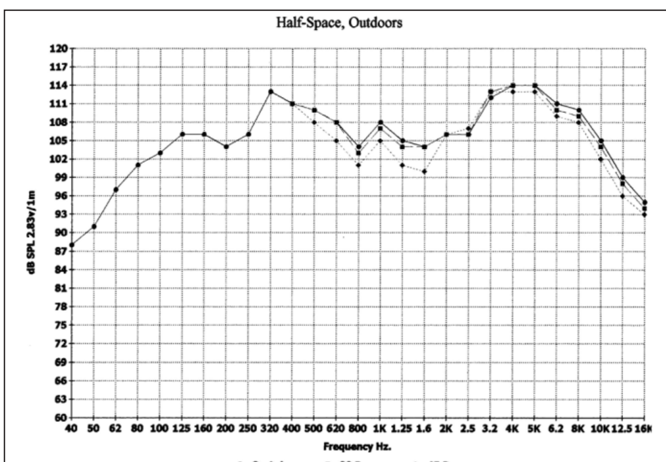

FIGURE 2: DR250a axial response.

driver chamber, where working with the components might lead to a serious case of fiberglass itch, I used polyester batts. Normally used for upholstery, these high-density inch-thick pads work very well for cabinet lining, and in this case have a decided advantage over loose polyfill stuffing.

The neodymium magnets used in both the PAudio™ woofer and the Dayton™ ribbon tweeters are heat sensitive, and will demagnetize with excessive heat exposure. The PAudio™ woofer—and all other neodymium woofers I've seen—uses a heatsink on the magnet, and the cast aluminum frame serves to dissipate voice coil heat as well. Likewise, the Dayton™ tweeters also have a large metal back plate that draws heat away from the magnet. Stuffing the driver chamber might allow polyfill to rest against the woofer and tweeter frames and prevent them from shedding heat. Lining the chamber with

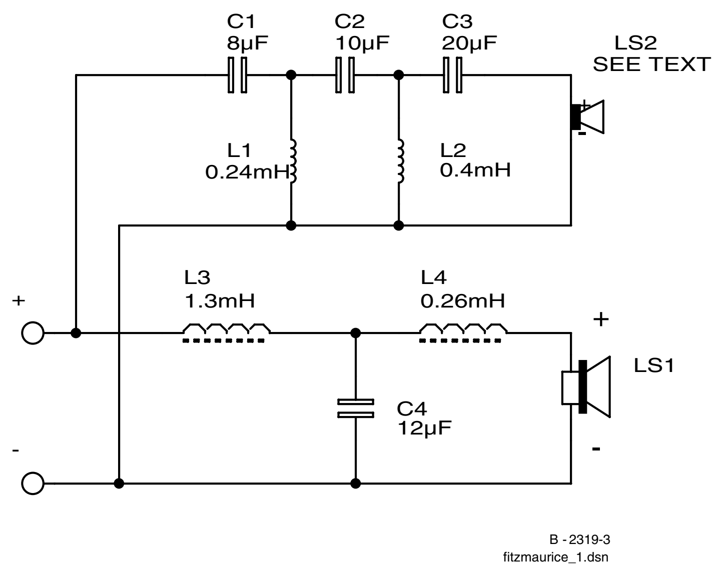

FIGURE 3: Crossover. Fifth-order high-pass/third-order low-pass filters. 5.6Ω tweeter/8Ω woofer.

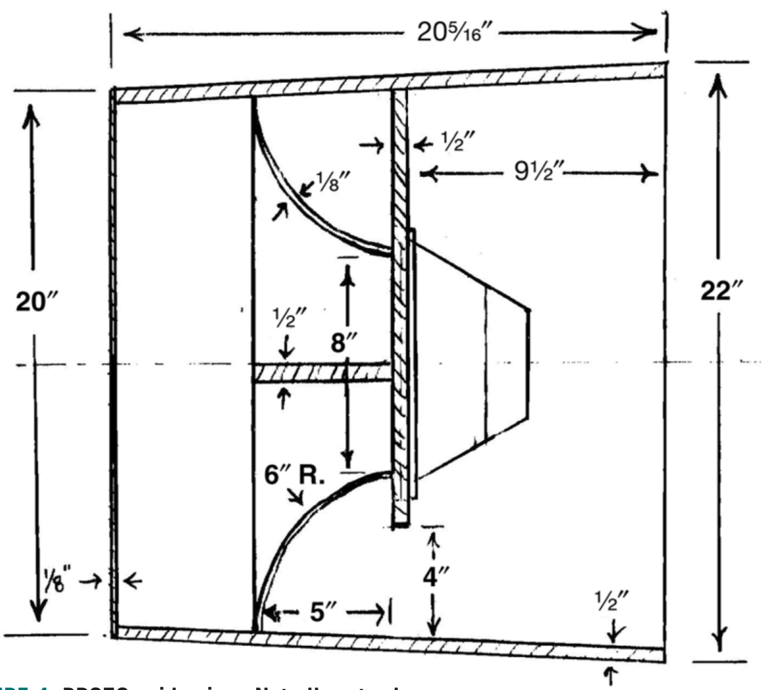

FIGURE 4: DR250a side view. Not all parts shown.

high-density polyester batts damps internal reflections while allowing the driver frames to dissipate heat.

For the most part, I used ½″ plywood to build the prototype, with ¼″ used for the mouth horn panels and ⅛″ for the throat horn panels and back. You may use ⅝″ or even ¾″ plywood for the flat parts for ease of joinery, but the self-bracing design does not require it for vibration control. Baltic birch, nice for its consistency and stiffness, is an option for the flat panels, but is denser than spruce plywood and will add a few pounds. Baltic birch is probably your only option for the ⅛″ parts.

For the ¼″ parts, avoid 5-ply Baltic, which is too stiff to bend, as well as Lauan plywood, which usually consists of two thin veneer plies over a thick core and also doesn't bend well. The best bet for ¼″ plywood is A/C spruce with three equal thickness plies. All part sizes and layout measurements are nominal, with the exact sizes determined by the finished thickness of the stock being used.

All joints are secured with drywall screws, piloted and countersunk, and adhesive. I prefer caulking-gun-applied construction adhesive of the polyurethane variety, which expands as it cures to fill voids. You'll also need a heat-melt glue gun for low-stress joints that need to set quickly.

CONSTRUCTION

Start by cutting out the throat horn sides (*Fig. 4*). Save the trimmings. Cut the leading and trailing edges with the saw blade set 5° off vertical to take into account the taper of the horn.

Cut the trapezoidal throat divider and attach both horn throat sides to it, with the divider offset to one side of the middle, so that the two halves of the horn won't be quite identical (*Photo 2*). Use the trimmings from the throat horn sides as stiffeners for the throat horn sheaths, attached right down their middles (*Photo 3*). Secure the throat horn assembly in a temporary jig (*Photo 4*) to hold the parts square while you attach the sheaths.

Cut the throat horn supports, again with a 5° angle on one edge, and drill holes through them to minimize both cabinet weight and flat internal surfaces. Make the supports about an inch longer than required. Attach the supports to the throat horn assembly, using clamps to hold the parts in alignment during the process (*Photo 5*).

After the adhesive has set, run the assembly through the table saw atop a panel-cutting jig, this time with the blade set 3° from vertical (to accommodate the shape of the box), trimming the assembly to finished length. Use a sander to trim the stiffeners to the same angle.

Cut out the top and bottom. While you can cut the arcs with a saber saw, a bet-

PHOTO 2: The throat horn sides and divider.

PHOTO 3: The throat horn sheath stiffener.

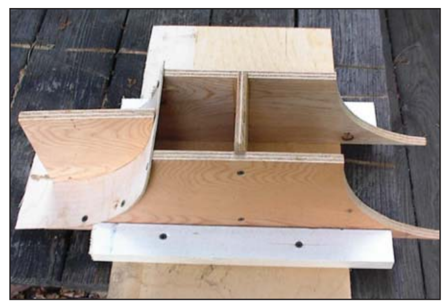

PHOTO 4: The throat horn on an assembly jig.

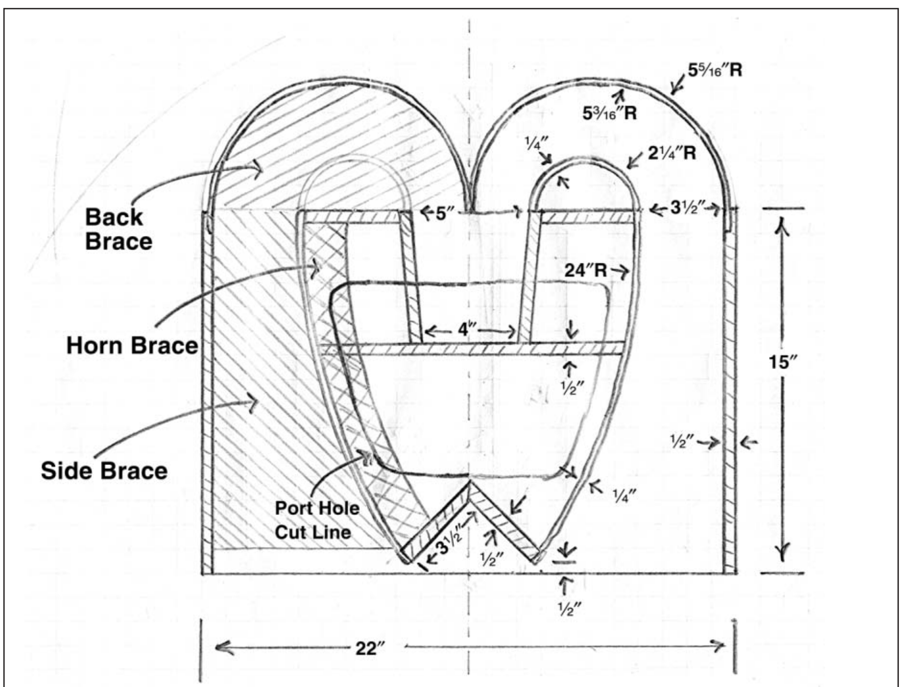

FIGURE 5: DR250a top/bottom view. Not all parts shown.

PHOTO 5: Attaching throat horn supports.

PHOTO 6: Making circles with a router.

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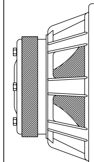

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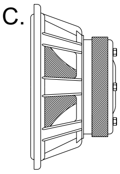


ter method to achieve a perfect cut is to use a router mounted on a board (*Photo 6*), with a screw through the center of the circle into the mounting board serving as a pivot to rotate the part being cut. Also note that the radii of the arcs depend on whether you plan to make the back halves from a single or a double layer of $\frac{1}{8}''$ plywood. Lay out the locations of all mating parts and cut the access porthole through the bottom, using a saber saw starting with a plunge-cut.

PHOTO 7: Tweeter baffle first length cut.

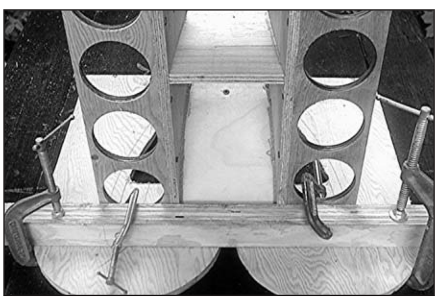


PHOTO 8: Using clamps and guideboard to align parts.

PHOTO 9: Baffle in place.

Cut out the tweeter baffle halves, making them at least an inch too long. If you're using rectangular tweeters, mark their locations on the baffle halves and cut out their mounting holes now, with the exception of the last cut at the baffle leading edges. You can drill holes for round tweeters later with a hole-saw. Assemble the tweeter baffle, trimming it to finished size and angle by running the assembly across the table saw, blade set 3° off vertical (*Photo 7*). If the cut doesn't reach all the way through, flip the assembly over to finish the job with the aid of the panel cutting jig.

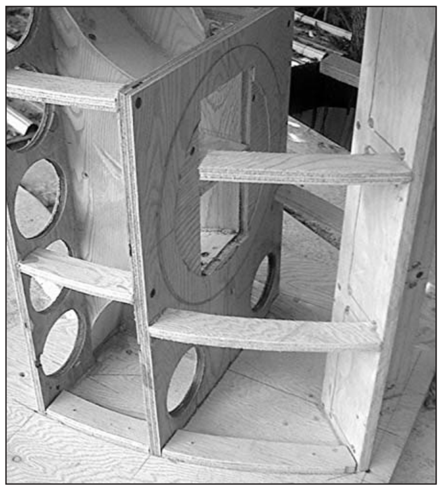


PHOTO 10: Horn braces. Note the "incomplete" brace.

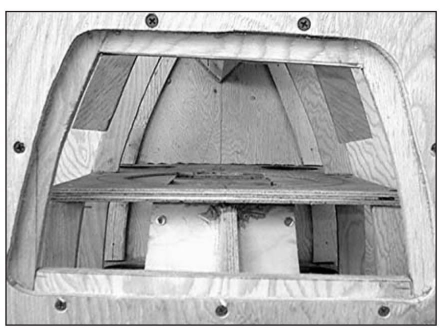


PHOTO 11: The porthole flange.

PHOTO 12: Lining with damping fiberglass.

Using clamps and guideboards to hold parts in alignment (*Photo 8*), attach the throat horn and tweeter baffles in turn, first to the top and then the bottom. The baffle is about an inch longer than necessary so the driver can fit on it at the bottom; cut at the required 3° angle where it mates to the top. Place it in the assembly, tracing through the throat horn the location of the hole. Cut the driver hole and a couple of venting holes as well and attach it to the assembly (*Photo 9*), using plenty of adhesive where it mates the thin sheathing and screws cannot be used.

Cut and install the horn braces, bisecting them as required where they meet the baffle (*Photo 10*), making sure they won't interfere with the tweeters. Where they would interfere with the installation of the woofer, it's not necessary for the braces to completely span from the tweeter baffle to the woofer baffle. The braces form part of the porthole mounting flange; use plywood scraps to complete the flange (*Photo 11*), making sure that the woofer will slide into place easily.

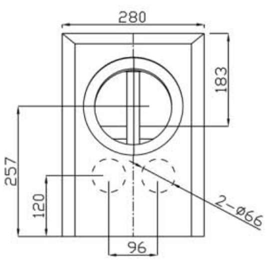
Line those parts of the assembly that will later be inaccessible with damping

PHOTO 13: Beginning the sheathing of the horn.

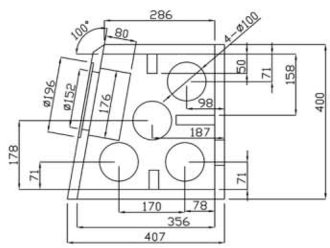
PHOTO 14: Make sure that driver fits!

Swans Tempus Kit

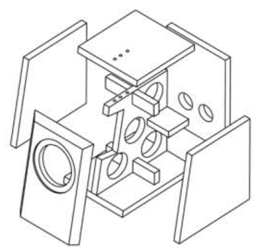
Swans proudly presents their first European-style, independently developed loudspeaker kit. Due to exceptionally high standards for a kit speaker, TEMPUS is a top-level, world-class performer in all regards.
German sound magazine KLANG+TON featured the Tempus in the June 2002 issue, commending its excellent acoustic response.



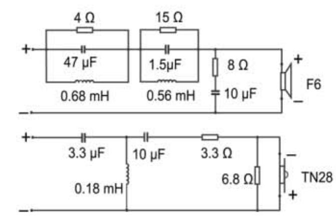
Front View



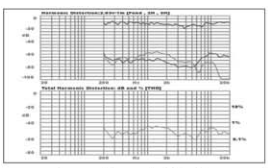
Section Side View



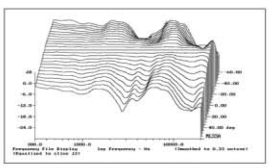
Explosion View



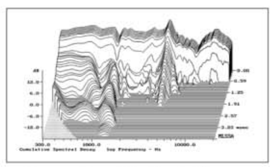
Filter Circuit



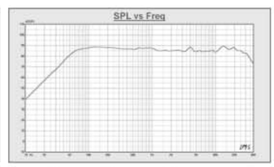
Tempus Harmonic Distortion



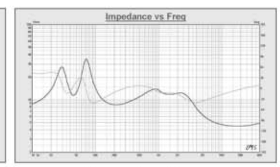
Tempus Horizontal response +60deg ~ - 60deg



Tempus Cumulative Spectral Decay



Tempus Frequency Response



Tempus Impedance

The Tempus project was initially conceived as a private work of the acoustic arts, to be executed independent of all professional affiliations. Despite a lack of traditional marketing and initial editorial commentary, the design has since received wide acclaim in the independent press and has gone on to tremendous commercial success. Tempus' reputation is therefore solely the result of unexpectedly high fidelity from a speaker of its modest origin and cost.

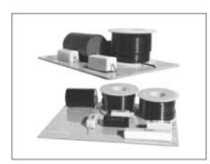
For Tempus' development, Swans agreed to supply premium components for what evolved into a rigorous electro-acoustics design program, executed to the highest European standards. The resulting design's performance was subjected to stringent laboratory confirmation and completely documented in the audio press, including revealing measurements not typical for any loudspeaker much less one of this type. Tempus has since gone on to claim top performance awards from acclaimed KLANG+TON magazine of Germany, during which time Swans cancelled all advertisements and further commentary.

Tempus is now one of Europe's most favored loudspeakers, leaving listeners admiring its naturally musical sound and excellent value. The design is already a classic, destined to sweep the audio world in a fashion similar to its current success.

The new Swans Tempus set includes:

- 2 x 1.1"/28mm Swans TN28 soft-dome tweeters with open-air housing, Neodymium motor, ferrofluid cooling, decompression chamber, integrated grille and German-made silk dome;

- 2 x 6.5"/165mm Swans F-6 shielded high efficiency midbass drivers, with Kevlar/paper curvilinear cone, phasing plug, and alloy basket;
- 2 x completed crossover networks with premium-grade components;



- Two sets heavy input terminals;
- Two pairs internal cable sets;
- Complete manufacturer's instructions.

All Tempus kits are thoroughly pre-tested by Swans and come in matched acoustical sets.

Price: $599 with Tempus parts kit only;
$999 with Tempus kit and cabinets;
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Sensitivity (2.83V/m):	86dB
Nominal Impedance:	8 ohms
Power Handling:	10-120W
Dimensions(HxWxD):	406x286x408 mm

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material (*Photo 12*). Starting at the tweeter baffle, attach the mouth horn panels (*Photo 13*). Use either long rigid clamps or webbing clamps and wooden cauls to pull the sheaths into place as you drive screws into the supports. This is your last chance to make sure that the driver fits into place (*Photo 14*).

After the adhesive has set, complete cutting the tweeter mounting holes (*Photo 15*), removing screws that are in the way and filling their holes with adhesive, and sand-flush all joints. Cut the back supports and glue them in place, using clamps and wooden cauls to keep them in alignment until the adhesive sets (*Photo 16*). Using an abrasive blade and a jig, halve a two-foot section of 4″ Schedule 40 PVC pipe for the reflectors. Cut the reflectors to size, stuff them with damping material, and install with hot-melt glue, being careful to make all joints airtight (*Photo 17*).

PHOTO 15: Tweeter holes completed.

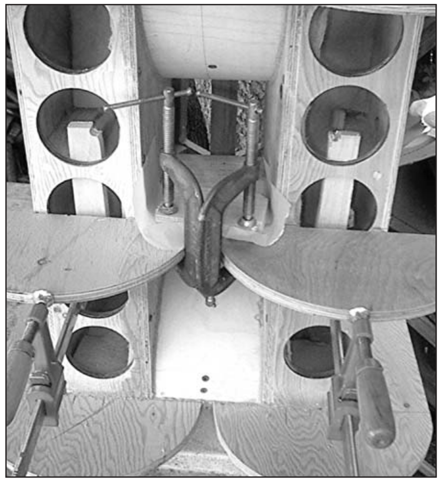

PHOTO 16: Installing the back braces.

Using a hole saw mounted on a long bit, drill the duct mounting holes; this will require more than one step to drill to the saw's maximum reach—remove material from the hole and continue drilling until complete. Cut the side braces a bit oversized and use dead-reckoning (*Photo 18*) to mark their finished sizes for final trimming before attaching them to the assembly. Note that the side braces extend only to within about an inch or so of the horn mouth.

TAKING SHAPE

You may use a single layer of ⅛″ ply-

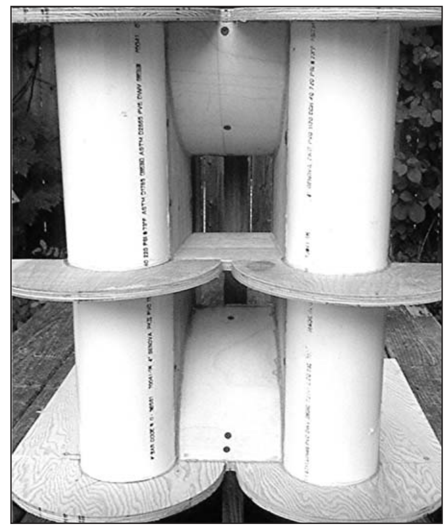

PHOTO 17: Reflectors in place.

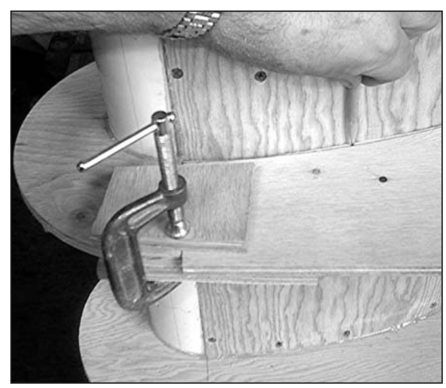

PHOTO 18: Marking a side brace for final trimming.

PHOTO 19: "Book-leaving" the back.

wood for the back halves, but because this prototype would be traveling cross-country, I laminated two layers for extra strength; the installation job is almost identical. Make sure you align the plywood on its easier-bending axis before marking and cutting. Make them long enough to extend at least an inch past the curved section of the supports. Cut the requisite pieces of plywood, gluing and screwing them together at one edge (*Photo 19*).

Insert the assembled "book-leaves," holding them in place with clamps and cauls (*Photo 20*). If you're laminating the back, liberally apply adhesive between the two layers. Use long clamps to begin pulling the halves into place as you drive screws into the braces, top and bottom (*Photo 21*), finishing off the job using webbing clamps and cauls to pull the assembly together. After the adhesive has set, use a circular saw with the blade set at minimal depth to finish-trim the end of the two layers.

Install the ducts using hot-melt glue, the majority of which you apply from the inside of the box. When you initially cut the ducts to length, cut one end at a 60° angle to approximate the shape of the horn. The duct lengths depend upon your tuning preference. For 50Hz

PHOTO 20: Beginning attachment of the back.

PHOTO 21: Butter-flying the book-leaves.

tuning make the two top ducts 6″ long (usable length, not including the angled section) and the two bottom ducts 1½″ long. For 80Hz tuning make all the ducts 1½″ long, and for 100Hz don't use ducts at all. After the glue has set, sand the ends flush to the cabinet.

Now is a good time to begin applying the finish—be it carpet (*Photo 22*), paint, or spray-on truck bed liner, because the innards of the horn mouth are much easier to access now rather than later.

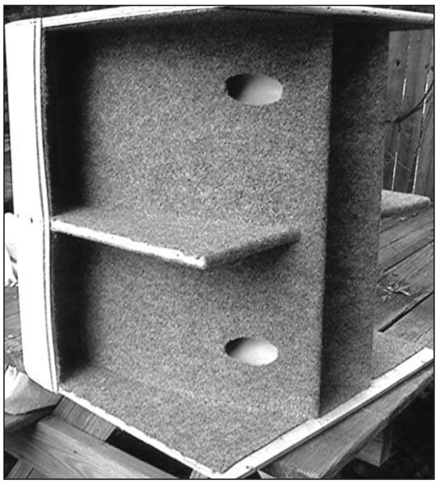

PHOTO 22: Finishing the horn interior.

Next cut the sides, sawing or routing a recess where they will overlap the back. Install them, using webbing clamps and cauls to apply pressure on the glue line where they mate the back halves (*Photo 23*). When the adhesive has set, trim and sand the sides and back flush to the top and bottom and complete the application of the cabinet finish, protective hardware, handles, and jack as desired. Wire the jack, making sure to caulk airtight where the wire passes through to the driver chamber.

If you're using dynamic tweeters, install the crossover on the inside of the porthole cover (*Photo 24*). As for the crossover components, size them as close as possible to *Fig. 3*, using only high-quality capacitors (no electrolytics) throughout. The bypass inductors of the high-pass filter may be inexpensive small wire gauge air-cores, but the inductors for the low-pass filter must be large gauge/low DCR ferrite or iron-core inductors to ensure minimal insertion loss.

Install the lightweight neodymium woofer, using screws to hold it in place; you can easily reach through the holes in the tweeter baffle to drive them. In-

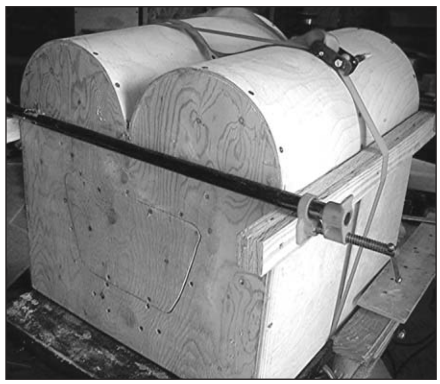

PHOTO 23: Melding the sides to the back.

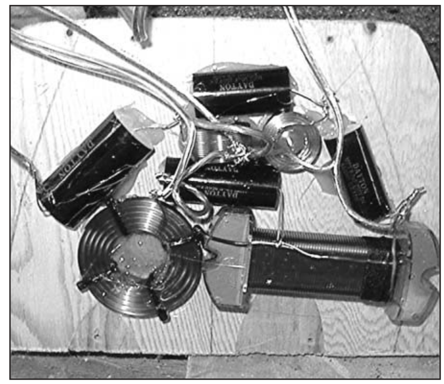

PHOTO 24: Crossover components in place.

stall the tweeters (*Photo 25*). If you opted for the ribbons, use a quartered piece of 1½″ PVC, screwed to the cabinet, for the diffuser. You can paint or carpet this as desired to match the cabinet finish (*Photo 26*). The baffle extensions are made of halved 1″ PVC which you can cut in one pass on the table saw using an abrasive blade, rip fence, and finger board. I found that these extensions should not be carpeted, so paint them to match your finish and screw them to the assembly.

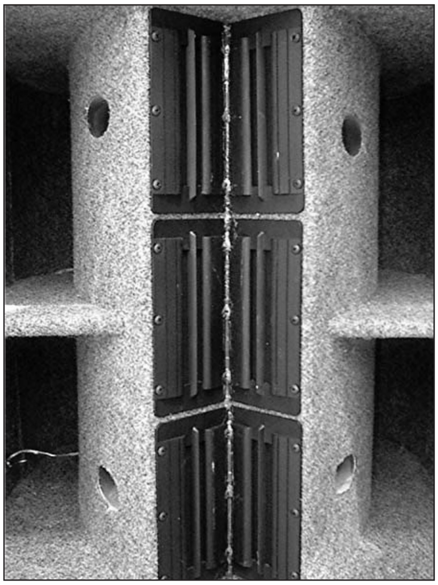

PHOTO 25: Ribbon tweeters installed.

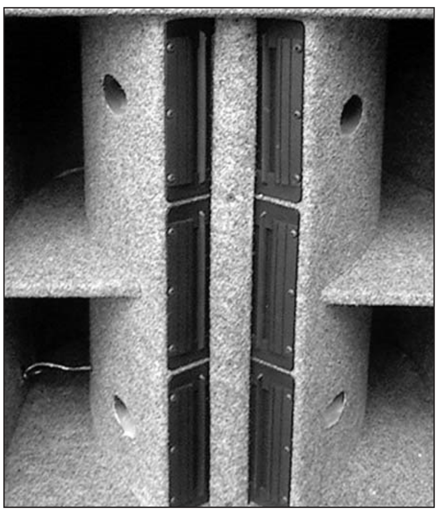

PHOTO 26: Tweeter deflector in place.

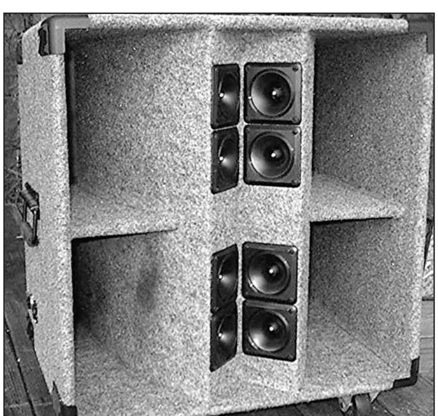

PHOTO 27: The DR250a, with CTS KSN 1167A tweeters.

The Pt2 tweeters have two slots in their faceplates through which you can see the diaphragms; mount the baffle extensions so that about one-half of the outer slot is covered for the best loading. Wire each bank of three tweeters together in parallel, wire the two banks together in series, and then wire the tweeter bank and woofer to the crossover, in phase. Line the remaining cabinet interior surfaces with damping material, making sure not to block the ducts or cover either the woofer or tweeter frames. Weather-strip the porthole flanges and screw the cover in place. Plug it in and rock.

A RINGING ENDORSEMENT

Upon completion, I sent the prototype across country to the headquarters of Delta Sounds, where it was immediately pressed into service as a demo for bass players in the Pacific Northwest who might be potential customers. The first player to try it found it to be far superior to his David Eden™ and SWR™ speakers in every respect, so much so that he volunteered his immediate endorsement if we would build him a signature model. I have no doubt that his offer to do whatever he could to have a DR250a of his own will not be the last.

If you want your own DR250a, you don't need to wait until the retail version (estimated price $800) appears. Build your own with the described drivers for less than $300; with piezo horns it costs about a hundred less.

Next up is my first venture into a true pro-sound subwoofer. Named after the acoustic instrument that it closely resembles, the Tuba 24 gives performance that other speaker designers say can't be done. Don't let your subscription lapse, because you'll definitely want to see this one.

Since building the original, I have constructed another DR250a that contains no ducts, and includes CTS KSN 1167A tweeters instead of the Dayton ribbons (*Photo 27*). ❖

Canto Sirena, Pt. 1

The author presents a balanced input differential phono preamplifier that uses octal tubes. In this first part he discusses techniques for obtaining accurate, stable equalization. **By James Lin**

A full generation after the introduction of CDs, people are still playing phonograph records and companies are still putting out new releases on vinyl. Presumably then, there is still some degree of interest in phono preamplifiers. Another old technology that seems to be holding up well is tubed electronics. Sometimes, it seems, newer is not necessarily better. Because many of the best LPs were recorded with tubed electronics, it seems natural to use a tubed preamplifier to replay them.

I had a pretty good commercial 6DJ8-based tube preamp, with one annoying trait, however: with a moving magnet cartridge, its output was about 20dB lower than my CD player or FM tuner. Fortunately, I managed to avoid switching from playing records to FM or CD and having to dive for the volume control as a blast of sound from the speakers reminded me to turn it down. Usually.

In the meanwhile, I had seen a J.C. Morrison design called the Siren Song[1]. Its gain was about 15–17dB higher than my phono section, but still designed for use with moving magnets. That seemed promising. Then, too, who can resist a design whose author says is "a phono preamplifier for hedonists" and who proclaims that "every effort has been taken to ensure that the performance of this device will be sentimental, agitated, maudlin, tender," and "tries, above all, to lure the joy out of your records[1]."

I thought the basic circuit topology was really cool—always a strong consideration—consisting of two stages of balanced differential amplifiers, followed by a single-ended output stage, with passive RIAA equalization, and the design was not commercially available. Why build a me-too preamp? But, I thought the Greek hedonist (the siren is, after all, a Greek myth) might benefit from a little Roman engineering rigor. I also decided on a rather unusual physical layout, which I think worked well (*Photo 1*).

DIFFERENTIAL AMPLIFIERS

A differential amplifier, or diff amp, uses two devices to amplify the signal. As you can see from the schematic (*Fig. 1*), two tube sections are connected to each other by their cathodes, the input is applied across the two grids, and the output is taken off between the two plates. The advantage of this topology is that it amplifies the difference between the two phases of the input signal—hence the name differential amplifier—while discriminating against anything that the two have in common, such as power-supply noise and hum, or radio frequency interferences and other induced noise.

The better the match between the two devices, the better the common-mode rejection. Its relative insensitivity to power-supply changes was the reason that Morrison chose to use a differential input stage. It is an especially good topology for a balanced signal, such as is naturally generated by a phonograph cartridge. The disadvantages are the cost of an extra tube and noise, from two tubes instead of one.

Perhaps because of this, tube diff amps have not been used for phono preamp circuits—at least until recently. Instead, published tube phono stage designs have almost uniformly been single-ended, with the negative lead of the phono cartridge tied to ground. Even in "high end" audio, the use of this topology has been notably rare. It was only when transistors became common, and cheap, that diff amp topology became very common. For instance, it is almost universal in integrated circuit op amps.

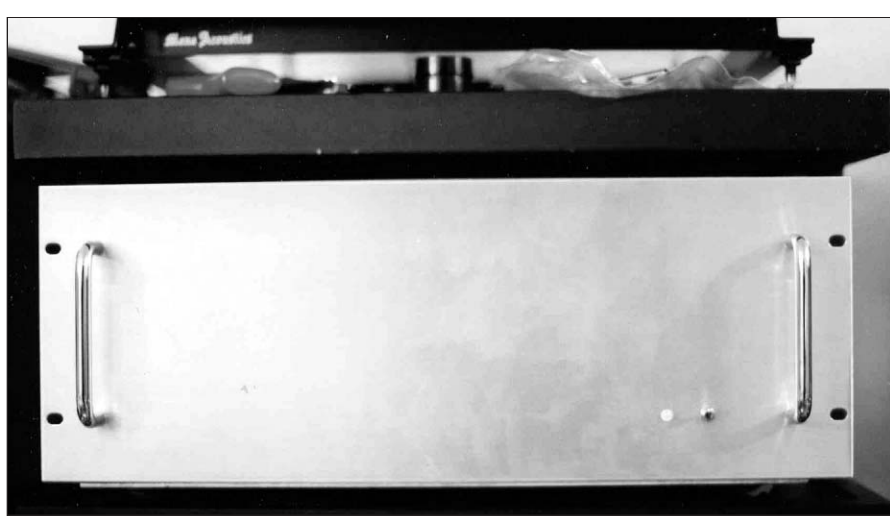

PHOTO 1: Front view of Canto Sirena phono preamp.

ABOUT THE AUTHOR
James Lin became interested in building audio equipment from reading *The Audio Amateur*. He has written articles for *The Audio Amateur*, *Speaker Builder*, and *Glass Audio*. This is his first article for *audioXpress*.

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According to W. Marshall Leach, the first application of a differential amplifier as a phono preamplifier was by Dan Meyer[2] in a transistor op-amp design published in *Wireless World* and *The Audio Amateur.* A few years later, Tomlinson Holman also used a differential amplifier topology for the first stage of the phono preamplifier he designed for the Advent 300 receiver[3]. Leach himself used a differential amplifier topology in a design in *Audio* in 1997[4], but he cascaded two differential amplifiers in series. The second differential amplifier helped reject the noise contributed by the first stage. However, all of these designs used the diff amp in a single-ended manner, with the negative lead of the phono cartridge tied to ground, and active feedback applied to the negative diff amp input.

Morrison published the Siren Song design in the now defunct *Sound Practices* magazine in 1992[1]. By the way, the original article is well worth reading because it addresses matters from a unique point of view. Like Leach's design, it used two cascaded differential amplifier stages to help reject noise; however, it had a balanced input stage. It also used "weird" octal tubes, and was somewhat unusual in breaking up the passive RIAA equalization into two simple RC networks between the stages.

RIAA CONSIDERATIONS

Modern phono cartridges are movement detectors. The strength of the signal is proportional to the velocity at which the stylus moves, which in turn is proportional to the size of the wiggles times their frequency. Unfortunately, this means that the size of the wiggles would be too large to be practical at the low end, whereas at the high end, they would be so small that they would be lost in the noise.

To avoid this, the signal engraved on the record is decreased at low frequencies to limit the size of the wiggles, and boosted at high frequencies to improve the signal to noise ratio. The preamplifier has a complementary boost at the low end, and cut at the high end, which also nicely decreases audible high-frequency noise. This is known as equalization, and the overall result should ideally be a flat frequency response.

Back in the "good old days" of 78 rpm records, and into the early days of the LP, each record company used a different equalization curve. Sometimes the same company even used different curves at different times! If you look at early mono and stereo preamplifiers, such as the Marantz 7, you will find that they had controls to choose several different curves to match whatever record was being played.

In the 1950s, the boost and cut was standardized by the Record Industry Association of America (R.I.A.A.), with time constants of 3180, 318, and 75µs, or in frequency terms, 50.05, 500.5, and 2122Hz. Despite this, many preamplifiers well into the 1970s (and beyond) were designed with RIAA equalizations that were inaccurate, including some very highly regarded and expensive models[5].

In 1979 Dr. Stanley Lipshitz published a series of tables to calculate accurate RIAA equalization networks[5] in the *Journal* of the Audio Engineering Society. Since then, there has been no excuse for anyone to design a phono preamplifier with inaccurate RIAA equalization. It is not a matter of cost, because an accurate design uses the same number of parts as an inaccurate design. Today, anyone with a PC spreadsheet program can type in the equations and quickly generate resistor-capacitor combinations that will provide accurate results.

STABILIZING THE RIAA TIME CONSTRAINTS

Although active equalization has been the most popular option in the past, passive equalization seems to be gaining favor. Paul Stamler has discussed the reasons for this—as well as some of the subtleties involved in achieving accurate equalization—in *audioXpress*[6]. RIAA equalization with tubes requires some adjustments, because a tube amplifier stage has limited gain, less than infinite input impedance, and significant output resistance. Further, both the input impedance and output resistance will vary depending on the gain of the circuit. Because of this the RIAA network values will differ significantly from those calculated for an ideal amplifier.

You might think that the input impedance of a tube is infinite because

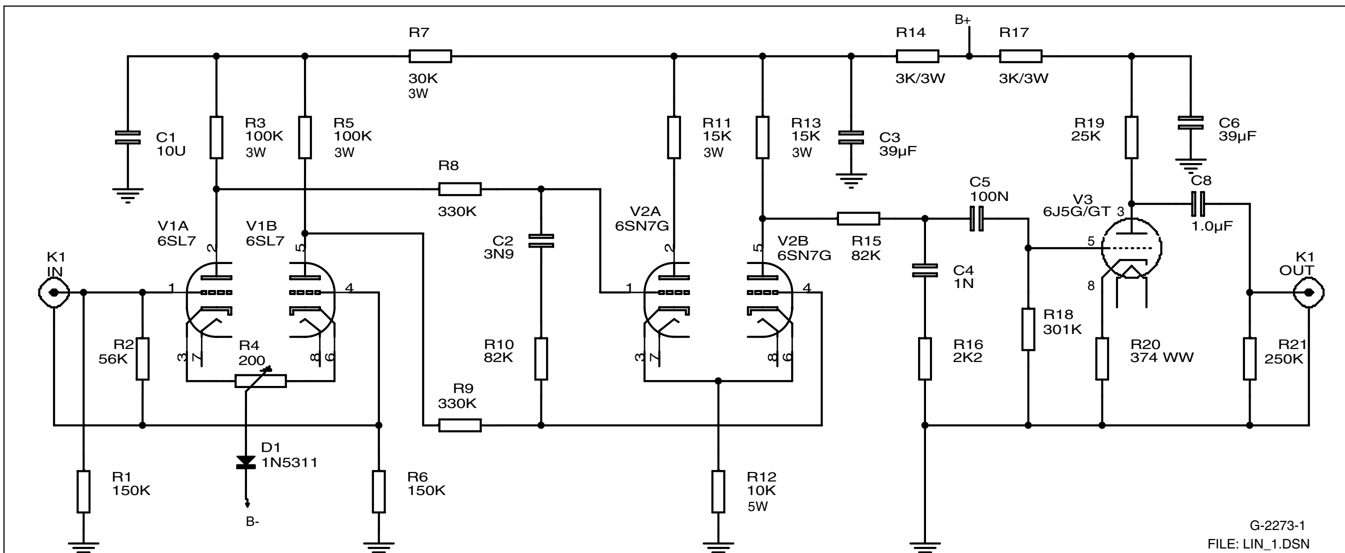

FIGURE 1: Canto Sirena schematic.

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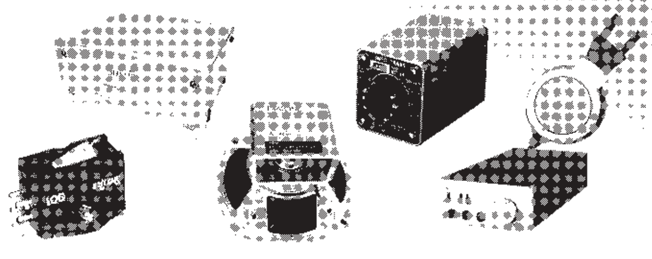

■Cartridge

Model	Price (US$)	Postage (Air Economy)	
Denon DL-102 (MONO)	150	Area I $18	China, Korea, Hong Kong, Taiwan
Denon DL-103 (STEREO)	200	Area II $22	Singapore, Malaysia, Indonesia
Denon DL-103R (STEREO)	250	Area III $27	North America, Oceania, Europe
Denon DL-103 PRO (STEREO)	350	Area IV $34	Africa, South America
Shelter Model 501 II (CROWN JEWEL REFERENCE)	750		
Shelter Model 901 (CROWN JEWEL SE)	1,400	These Area I ~ IV are for all products except book.	

■Japanease Audio Book

Postage $15

Title		Price (US $)
Attractive Tube Amps Vol. 1&2	(Isamu Asano)	30 each
The Joy of Vintage Tube Amps 1&2	(Tadaatsu Atarashi)	30 each NEW
Direct & Indirect Tube Amps	(Kiyokazu Matsunami)	40 NEW
SE Amps by Transmitting Tubes	(Kouichi Shishido)	50
The Remembrance of Sound Post	(Susumu Sakuma)	30
Classic Valve	(Hisashi Ohtsuka)	40
MJ Selected 300B Amps	(MJ)	30
Top-Sounding Vintage Power Tubes	(Stereo Sound)	30
Output Trans of The World	(Stereo Sound)	30
20TH CENTURY OF AUDIO	(Stereo Sound)	30
Vintage Speaker Units	(Stereo Sound)	30 NEW
Tube Amp Craft Guide	(MJ)	30

■MC STEP UP TRANS

Model	Specifications			Price (US$)	Postage**
	Pri.Imp (Ω)	Sec.Imp (kΩ)	Response		
Shelter Model 411	3~15	47	20Hz~50kHz	980	Area I $25
Jensen JE-34K-DX	3	47	20Hz~20kHz	550	Area II $30
Peerless 4722	38	50	20Hz~20kHz	300	Area III $40, Area IV $50

**Air Economy

■STAX

Model	Price (US$)
OMEGA II System (SR-007+SRM-007t)	Ask
SRS-5050 System W MK II	Ask
SRS-4040 Signature System II	Ask
SRS-3030 Classic System II	Ask
SRS-2020 Basic System II	Ask
SR-001 MK2 (S-001 MK II+SRM-001)	Ask

■Speaker

Model	Specifications					Price* (US $)	Postage** (US $)			
	D (cm)	Ω	Response	db	w		I	II	III	IV
Fostex FE208Σ	20	8	45Hz~20kHz	96.5	100	296	62	74	120	156
Fostex FE168Σ	16	8	60Hz~20kHz	94	80	236	42	50	73	98

*Price is for a pair **Air Economy

■TANGO TRANS (ISO) [40models are available now]

Model	Specifications				Price (US $)	Postage** (US $)				
	W	Pri.Imp (kΩ)	Freq Response	Application		I	II	III	IV	
XE-20S (SE OPT)	20	2.5, 3.5, 5	20Hz~90kHz	300B,50,2A3	396	47	56	84	113	Price is for a Pair
U-808 (SE OPT)	25	2, 2.5, 3.5, 5	20Hz~65kHz	6L6,50,2A3	242	42	50	73	98	
XE-60-5 (PP OPT)	60	5	4Hz~80kHz	300B,KT-88,EL34	620	62	74	115	156	
FX-40-5 (PP OPT)	40	5	4Hz~80kHz	2A3,EL34,6L6	320	47	56	84	113	
FC-30-3.5S (SE OPT) [XE-60-3.5S]	30	3.5	20Hz~100kHz	300B,50,PX-25	620	62	74	115	156	
FC-30-10S (SE OPT) [XE-60-10SNF]	30	10	30Hz~50kHz	211,845	620	62	74	115	156	
X-10SF [X-10S]	40	10W/SG Tap	20Hz~55kHz	211,845	1160	90	110	180	251	
NC-14 (Interstage)	—	[1+1:1+1]5	25Hz~40kHz	[30mA] 6V6(T)	264	30	40	50	70	
NC-16 (Interstage)	—	[1+1:2+2]7	25Hz~20kHz	[15mA] 6SN7	264	30	40	50	70	
NC-20F [NC-20] (Interstage)	—	[1:1]5	18Hz~80kHz	[30mA] 6V6(T)	640	42	50	73	98	
NP-126 (Pre Out)	—	20,10	20Hz~30kHz	[10mA] 6SN7	264	30	40	50	70	

** Air Economy

■TAMURA TRANS (All models are available)

F-7002 (Permalloy)	10	3.5	15Hz~50kHz	300B,50	836	60	70	110	145	Price is for a Pair
F-7003 (Permalloy)	10	5	15Hz~50kHz	300B,50	836	60	70	110	145	
F-2013	40	10	20Hz~50kHz	211,242	786	70	84	133	181	
F-5002 (Amorphous)	8	3	10Hz~100kHz	300B,2A3	1276	65	80	120	160	

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the grid is not connected to anything else, so there is nowhere for a current to flow. This is true at DC; however, since the elements of a tube are actually small pieces of metal, they form little capacitors. When the tube is used as an amplifier, you can show that the effective capacitance that the circuit sees between the grid and the plate is the actual capacitance multiplied by (A+1), where A is the voltage amplification of the tube in the circuit[7]. This phenomenon is known as the Miller effect, and it affects the input impedance of any active device, tube, transistor, or FET (see sidebar for a more detailed discussion).

The output resistance of a tube amplifier stage is also non-zero, being formed by the plate resistance of the tube, and the plate resistor in parallel[7]. If the cathode resistor is not bypassed, it also needs to be added to the effective plate resistance, by a multiplying factor of (A+1)[7].

In addition, there is some variation in amplification from tube to tube, and also as a tube ages, leading to further potential frequency response alterations. Typically, tube parameters will generally vary by 10–20% between samples. A good design should minimize the effects of these variations by allowing the closer tolerance and more stable passive elements to predominate.

One way to stabilize the equalization is to choose the values of the passive elements that will swamp the values inherent in the active device itself. For example, if the tube stage has an output resistance of 10kΩ, you can place a series 100kΩ resistor at the output, giving a total output resistance of 110kΩ. Now, if the tube ages, or you put in a new tube with an output resistance of 11kΩ—a difference in the tube parameters of 10%—the total resistance goes up to 111kΩ, a change of less than 1%.

Similarly, you can place a large capacitor value across the input to swamp any variations in Miller capacitance. Of course, there is a limit; thus, the values you choose will generally be a compromise to try to minimize the overall effects of variations. This technique should also minimize any variations due to stray capacitances, and so on, which result from the physical layout of the design and are not predictable on paper.

How much does this matter? Well, in an equalization network, an alteration of 3–4% can change the frequency response by 0.3dB over a broad range of frequencies, which is just perceptible, so that sets the maximum desirable variation.

To summarize, you can model a tube as an ideal amplifier with a limited gain—a Miller effect capacitor on the input side and an output resistor on the output side. You must then add the Miller effect capacitance and output resistance to the real resistances and capacitances in the circuit when calculating an equalization network. These values change depending on the circuit, so begin by calculating the gain and output resistance of each stage[7]. Fortunately, there is software available from John Broskie at Glassware that can do most of this[8].

TUBE CHOICE

Another unusual feature of the Siren Song was the use of octal tubes instead of the much more common miniature tubes such as the 12AX7 and 6DJ8s. As Eric Barbour has pointed out[9], the almost universal use of nine-pin miniature tubes in preamplifier circuits is more due to economics and fashion than sonic merit. The 12AX7s so ubiquitous in the 1950s "Golden Age of Hi-Fi," and the recently popular 6DJ8 were less expensive to make originally than octal tubes—although, ironically, demand has since made some 12AX7s and 6DJ8s more expensive to buy these days.

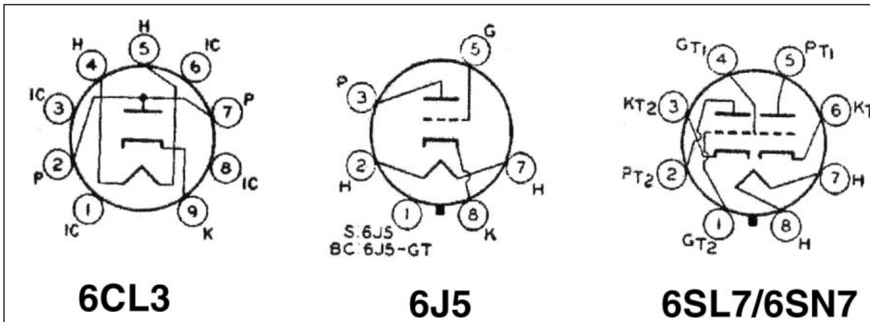

FIGURE 2: Base diagrams for tubes used.

MILLER EFFECT

To see why the Miller effect exists, consider, at a basic level, what a capacitor is. If you have two pieces of metal suspended in a vacuum, you can place a charge upon them. This charge will then produce a voltage between one element and the other. As you alter the charge, you alter the voltage. The relationship between the charge Q and the voltage V defines the capacitance C, or in mathematical terms, $Q = C * V$.

Physically, a tube consists of metal elements (cathode, grid, and plate) suspended in a vacuum. Thus, any two elements, such as the grid and the plate, form a small capacitor, amounting to a few picofarads. In itself, this capacitance would seem to have little effect at audio frequencies. If you use a capacitance meter to measure the capacitance between, say, the plate and the grid in a triode vacuum tube, you get the grid-to-plate capacitance (Cgp).

Suppose you put the tube in a common-cathode circuit, but leave the power off. If you increase the voltage on the grid by one volt, then you must inject a small amount of charge q between the grid and plate so that $1V = Cgp*q$, where Cgp is the grid-to-plate capacitance.

What happens when you turn the power on? Suppose you put the tube in a grounded cathode circuit that produces a voltage gain of A (actually, –A, since the plate voltage changes in the opposite direction to the grid voltage—when the grid voltage goes up, the plate voltage goes down). Then, if you increase the grid voltage by 1V, the plate voltage changes by –A volts, so the total change between grid and plate is (A+1) volts, rather than the 1V change you had when the power was off.

To accomplish this voltage change, you need to inject a charge of $(A+1) * q$ onto the Cgp capacitor. Thus, the tube capacitance looks as though it is (A+1) times larger when it is turned on compared to when it was turned off. This is the Miller effect. The same thing happens with a transistor or FET, but the physical process is easier to understand with a tube since you can actually see the metal bits that make up the capacitor.

In addition, once there is a commercially successful design using a particular tube, there seems to be a herd instinct that prompts other designers to rush to use the same tube. Thus, the high-end market is full of single-ended phono preamp designs using nine-pin miniature tubes whose main virtues are cost and availability. Both are important commercial considerations; however, a do-it-yourself builder who is interested in making one unit is not as restricted.

The octal base 6SL7 and 6SN7 are two of the most linear and best-sounding common triode tubes around[9]. A 6SN7 typically has several-fold less distortion than a 12AX7 or 6DJ8. Inherent linearity is particularly important in a no-feedback phono preamp design because the active devices must provide extra amplification to overcome the losses in the equalization networks, and there is no feedback to lower distortion. More linear tubes should provide better sound. The layout of the bases for the tubes used are shown in *Fig. 2*.

CIRCUIT DESIGN

Since I thought the operating conditions of the tubes were satisfactory in the original design, I left them unchanged. The first stage is a 6SL7 in a diff amp topology with a gain of about 44×, or 33dB. The current source diode 1N5311 increases isolation from the negative power supply, but it is a potential noise source, so a quiet one is essential.

Although the current source improves common-mode rejection, it actually worsens the high voltage power-supply rejection at the output, since the constant current means constant voltage across the plate resistors[8]. Thus any voltage variations or noise in the power supply will be transmitted as a common-mode signal in the output. The use of a second diff amp stage will save you here, as any transmitted power-supply noise from the first stage will be viewed as common signal and mostly rejected by the second stage. The operating conditions are very conservative—plate dissipation is less than 20% of maximum.

High amplification is important here because you need to raise the signal above the noise, and the first RIAA network will attenuate the output of this stage up to ten-fold going into the second stage. The choice of the input stage tube also has a major effect on the ultimate noise level of the preamp; with moving magnet cartridges the 6SL7 is slightly quieter than the widely used 12AX7[10]. However, the differential stage will worsen the noise by 3dB compared to a single-ended design. For this reason, this design is not very suitable for a low-output moving coil cartridge.

The first stage is direct-coupled via an RC network with the second stage 6SN7, also in a diff amp topology. The gain would be 13×, or 22dB, with the signal taken off both plates; however, since you are using only half the signal, the gain is 6.5×, or 16dB. Incidentally, notice which half of the diff amp feeds the single-ended output stage—this is chosen so that the preamp is non-inverting from input to output. There is a –0.3dB rolloff at 20Hz due to the interaction between C5 and R18, which will give a little bit of protection against record warp signals.

Nearly all phono systems will have a little bit of rise in this area due to the subsonic resonance peak resulting from the stylus compliance interaction with the effective mass of the tonearm and cartridge. Thus, a little rolloff will actually tend to flatten the overall response. Again, operating conditions are very conservative—less than 40% of maximum plate dissipation.

For the output stage I substituted a 6J5GT for the 7F8 or 6N7 tubes used in the original design. The 6J5GT is equivalent to half of a 6SN7, and Barbour opines it is one of the best sounding common triodes[9]. It has 2–3dB less gain than a 6N7 tube, but in return offers a lower output impedance.

The output impedance is still significant at about 8kΩ so the output cable should have a low capacitance. For example, a combined cable plus line stage input capacitance of 200pF would result in about a 0.25dB rolloff at 20kHz, and this is about the upper limit of acceptable—100pF results in about 0.1dB rolloff, which is better. The line stage input resistance should be about 10× the output resistance, or around 100kΩ to minimize loss of gain.

Also, the use of the 6J5GT does not require paralleling two sections as the 7F8 and 6N7 tubes do, for those of you who are paranoid about such things. Although the RCA tube manual gives the grid-plate capacitance specifications as 4pF, I measured several makes of 6J5GT and found grid-plate capacitances averaging 7pF, which I used in calculating the RIAA networks. Except for the change in tubes, the cathode and plate resistor values are identical to the original design, with conservative plate dissipation less than 60% of maximum.

If you choose to try the original 6N7, you can wire the socket for the 6N7 connecting the grids and plates in parallel. This will allow you to use either tube since the 6J5 has the same pin layout as the 6N7, with the exception that pins 4 and 6 are absent in the 6J5, whereas they are connected to a grid and plate, respectively, in the 6N7. The gain with the 6N7 is increased to 16× or 24dB, but the output resistance is also increased to nearly 12kΩ.

RIAA NETWORK

Other than the change in the output tube, the big change in the circuit occurs in the RIAA equalization, which is split into two networks. This has the advantage that it is much easier to find standard resistor capacitor values that will give accurate results. Some have suggested that another advantage of this scheme is that you can more easily trim for exact equalization[1,11]. However, in his RIAA equalization article, Lipshitz argues persuasively that it is easier to do an accurate design in the first place[5].

Morrison's design places the 75µs RC network after the first diff amp stage, and the 3180/318µs network between the second diff amp and the output stage. Since the output resistance of the 6SL7 stage is relatively high, the first RIAA network will need to use a high resistance value to minimize the effect of variations in the tube output resistance. However, you also need to use as large a capacitor as possible to swamp the Miller effect variations at the input of the second stage 6SN7. High resistance into a high capacitance results in a low-frequency turnover point. Therefore, it makes more sense to use the first RC network to set the 3180/318µs turnover points.

Because of the high resistance used in this section, there is an additional rolloff due to Miller effect capacitance interaction with the 82kΩ resistor across the 6SN7 inputs at around 72kHz. This would result in about a 0.3dB rolloff at 20kHz; however, you can correct this in the second network. Variations in this rolloff point can not be compensated; however, due to the high frequency the effect of variations is limited in the audible range. If you are thinking that the final values of resistor and capacitor don't give the correct turnover points, remember that the 6SL7 output resistance and 6SN7 Miller capacitance are also in circuit.

The 75µs network is then located between the 6SN7 and the 6J5 tubes. Again, the network components are chosen to minimize the effect of tube variations on RIAA accuracy. This network has about a 2.2dB insertion loss, and rolls off at 6dB per octave above 2122Hz. The 2.2kΩ resistor in series with the capacitor in this network stops the rolloff at about 72kHz, compensating for the Miller effect rolloff of the first RIAA network. Interestingly, designer Allen Wright claims that placing the RIAA networks in this order also sounds the best[11].

I replaced the 250kΩ pot at the output of the original design with a fixed resistor, so the full output gain better matched my other sources. If you wish to use this preamp to drive a power amp directly, it would be better to have the volume control at the power amp input, because such a high-value pot at the preamp output will roll off the high frequencies due to interaction with the connecting cable capacitance.

The final circuit design is shown in *Fig. 1*. The calculated RIAA equalization constants assuming the nominal values for tubes and passive components are within 0.6% of ideal, resulting in a deviation from flat of ±0.03dB. Even with unselected 1% components, the equalization should be within ±0.2dB or less. These changes in the RIAA networks result in more accurate, and more stable equalization. While some may question whether this degree of accuracy is necessary, since it doesn't cost anything, why not?

As tubes age, their output impedance increases and Miller effect capacitance decreases. With this split RIAA design topology, these two effects tend to offset each other. Thus, a 20% decrease in emissions of all three tubes from nominal values should alter the RIAA equalization by less than 0.05dB over most of the audio band.

Part 2 details component and tube selection and construction of the phono preamp. ❖

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Adcom's GFP 565 Preamp Mods, Part 3

Now it's time to add a new line stage to the mods described in Parts 1 and 2. **By Gary Galo**

In Part 2 of this series, we completed the installation of a new power supply based on Walt Jung's *Improved Regulator*. In this installment, we'll replace the line stage with a new, high-performance design. But, before beginning the line stage, there are a few other options to consider.

During Victor Campos' tenure at Adcom, they called him The Buffer Man, and with good reason. Every preamp Adcom produced under his direction had actively-buffered tape outputs to isolate the signal path from nonlinear loads. The GFP-565 preamp used a pair of dual op amps configured as unity-gain voltage followers to buffer the two pairs of tape outputs.

At that time, many of the IC op amps Adcom used were manufactured by Linear Technology, and bore the familiar LT logo. But, Adcom had its own part numbers printed on these devices to conceal their identity. Now that the preamp has been discontinued, I won't be giving away any trade secrets by identifying them. The tape output buffers—labeled Adcom 7A—were LT1057 FET-input dual op amps.

During my evaluation of the 565 preamp for *Stereophile* back in 1989, one of the LT1057 op amps in my review sample failed. The output of the device went to the negative supply rail (–18V DC), and the non-inverting input was at a potential of several volts negative. This produced a potentially dangerous thump when you moved the recording selector switch to the same position as the listening selector. Subsequently, another LT1057 failed in my other 565 preamp.

Linear Technology did failure analyses on both chips, and found that the positive supply connections had failed internally. This did not give me a great deal of confidence in the LT1057, so I replaced all of them with the Analog Devices AD712JN, one of the best dual op amps available at that time (www.analog.com).

The AD712 is still a solid performer for this application, but if you have really high-quality recording equipment, you may wish to use something even better. My first choice is the TI/Burr-Brown OPA2604AP (www.ti.com). Other high-end replacements include the TI/Burr-Brown OPA2134-PA.

Adcom used 7.5k pull-down resistors on the LT1057s, which forced the output stages into Class-A operation. Retain these resistors if you use the AD712. Op amp output stages have improved considerably since the AD712 was designed, so remove these resistors if you use any of the other devices. In particular, the OPA2604 is an all-FET device, including the output stage, so you should definitely remove the pull-down resistors for this device. The modified tape output buffer circuit is shown in *Fig. 5*.

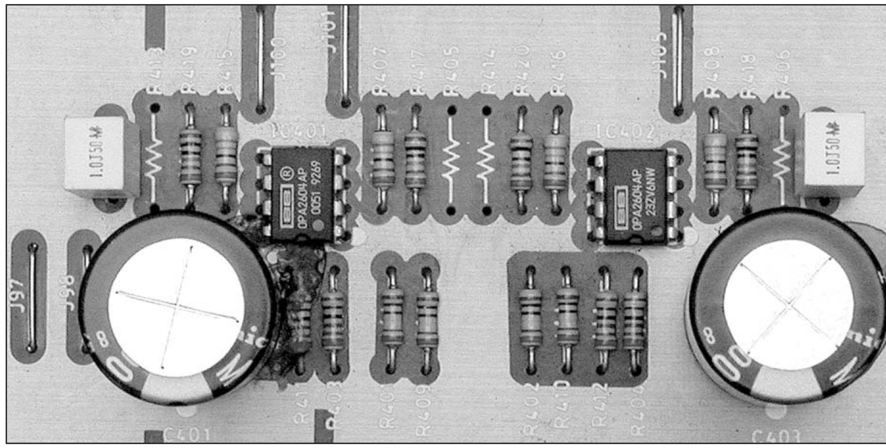


PHOTO 22: The modified tape output buffers, using OPA2604 op amps. The pull-down resistors are removed for the OPA2604.

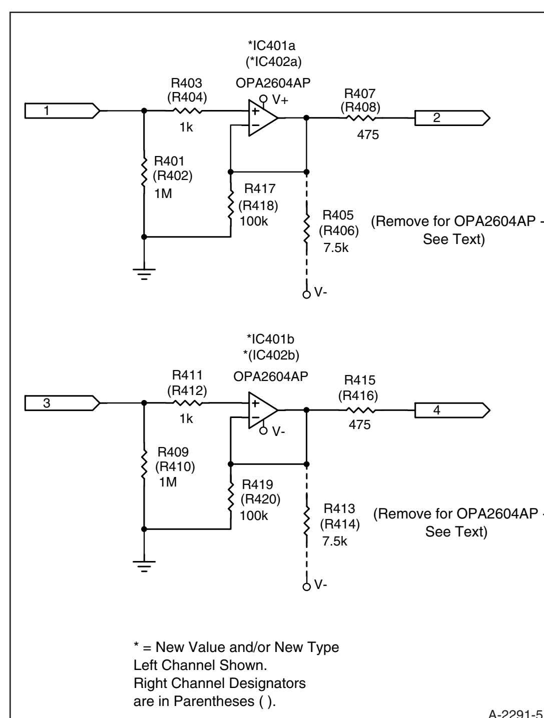


FIGURE 5: The modified tape-output buffers. OPA2604 dual op amps replace the original LT1057s, and pull-down resistors R405, R406, R413, and R414 are removed. The pull-down resistors should be retained if AD712 op amps are used.

- Remove IC401 and IC402. Replace these with OPA2406s, OPA2134s, or AD712s (*Photo 22*).
- Remove the 7.5k pull-down resistors—R405, R406, R413, and R414—if you use any of the recommended devices other than the AD712.

RESISTOR CHOICES

Before proceeding, you will need to decide what type of resistors you intend to use for the remainder of this project. Adcom used Roederstein/Resista MK2 resistors in the GFP-565, but they have steel end caps, and are now sonically outclassed by a number of other resistors. I recommend using non-ferrous resistors for all of the circuitry described later, and in Part 4 (which will cover the phono preamp).

The most cost-effective choice is the Vishay-Dale CMF Type RN60 sold by Mouser and Welborne Labs, which I discussed in Part 2. Ron Welborne says that he spot-checks the Vishay-Dale resistors to make sure that they are non-magnetic, and all that I have received from Mouser have also been non-magnetic. The Vishay-Dale RN60 resistors will drop right into the existing resistor footprints on the Adcom PC board.

Michael Percy Audio and Welborne Labs still have some old, all-copper Holco H4 resistors in stock. E-mail them to see whether they have old-stock Holcos in the values you need before purchasing. Holco H4 resistors will fit the existing Adcom footprints if stood on end, as shown in several of the photos later in this article. If you can get them, they are excellent.

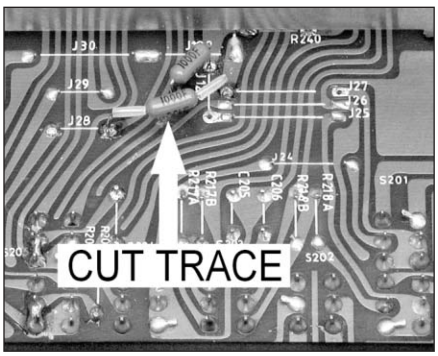


PHOTO 23: Output resistors for the optional Aux Outputs. These resistors connect the line stage outputs to PC traces previously used for the signal processor outlines.

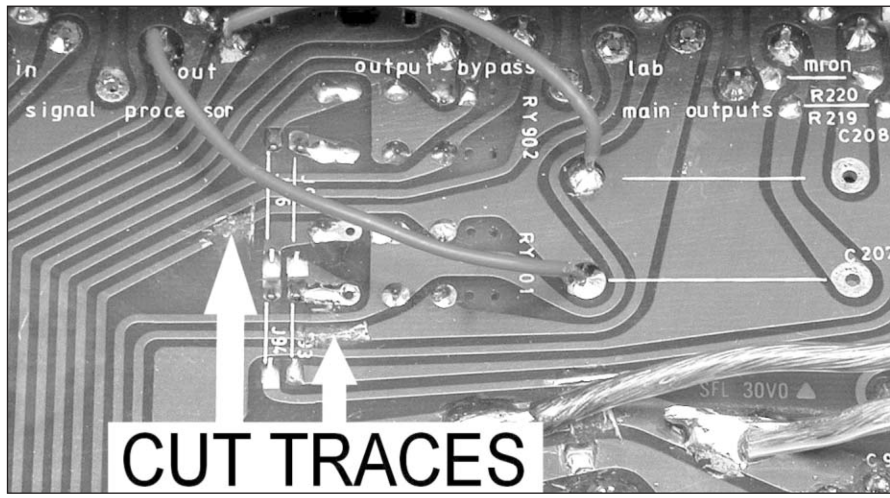


PHOTO 24: The modified output muting relay. Relay RY901, previously used to mute the Norm and Lab outputs, is now connected to the Aux Outputs with two insulated jumpers.

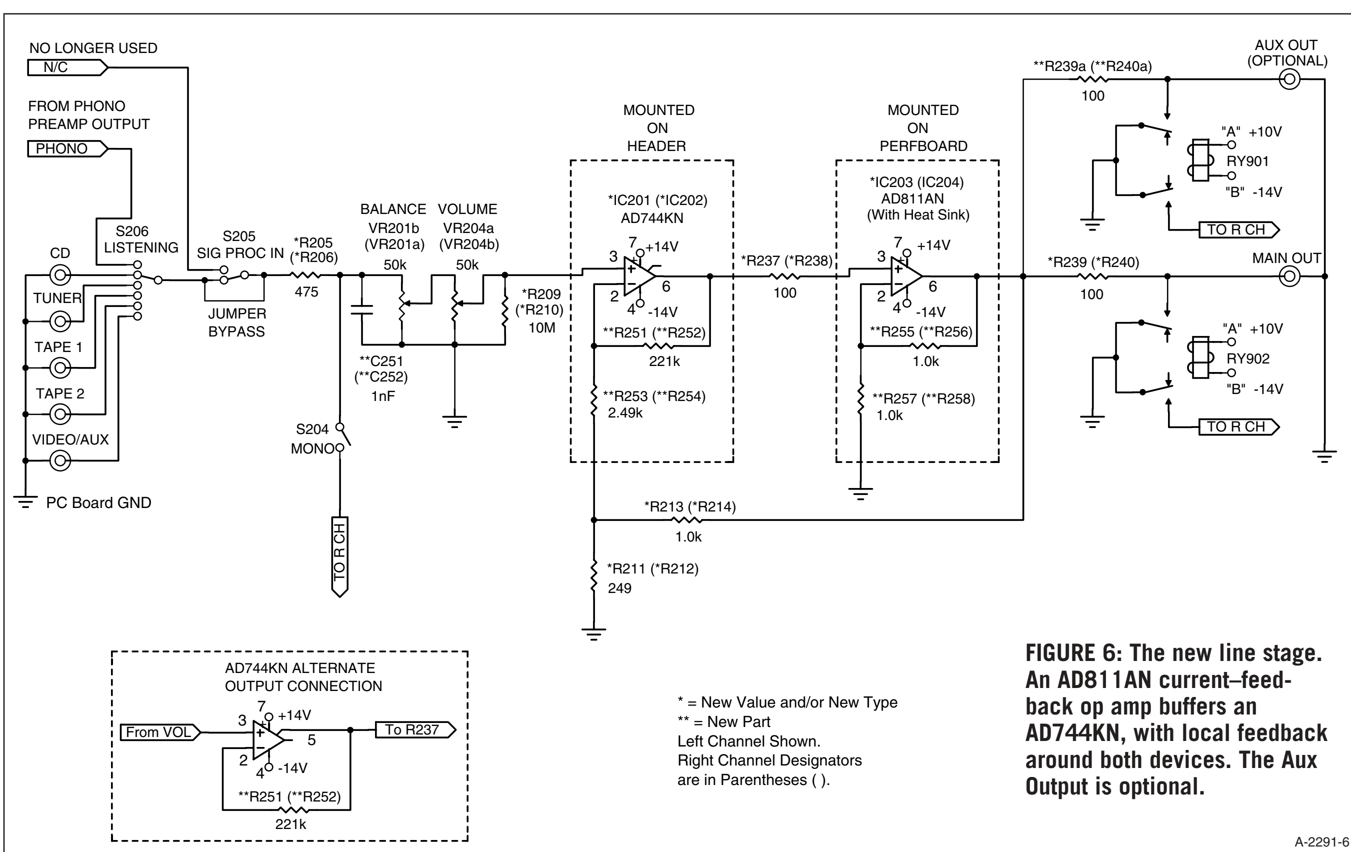


FIGURE 6: The new line stage. An AD811AN current–feedback op amp buffers an AD744KN, with local feedback around both devices. The Aux Output is optional.

I built my "Number 1" preamp with Caddock MK132 precision film resistors in all of the audio circuits. The price tag will be a deterrent for some builders, but on a high-resolution system they are well worth the expense. Michael Percy, Welborne, and Parts Connexion stock these. If cost is no object, the Vishay S-102-series bulk foils are considered by some to be the finest audio resistors available. But at over $11 a pop, this will be an expensive preamp project!

I won't refer to specific resistor types from now on (with one exception)—the final choice is yours, based on your listening requirements and budget. Michael Percy Audio stocks the Vishay resistors. To fit the Adcom resistor footprints, the leads on the Caddock and Vishay resistors should be bent out at a slight angle, and then bent straight down.

AUX OUTPUTS

In Part 1, I suggested adding a pair of Aux Output jacks for biamping. If you don't need the extra pair of outputs, you can skip this section. In the following steps, I assume that you have installed the jacks and connected them to the PC board as outlined in Part 1. Now add a second pair of output resistors and route them to the PC traces previously used for the signal processor outlines (*Fig. 6* and *Photo 23*).

- Remove jumper J28.
- Make a ¼″ cut in the PC trace between S205 and jumper J59 in the same front-to-back location as the J25 jumper. Be sure the cut area is completely clean and free of copper trace material.
- Scrape ¼″ of lacquer from the end of the cut trace leading to J59, and tin the bare copper.
- Solder a 100Ω resistor—R240a—between this tinned PC trace and the rear-most hole previously occupied by J124.
- Connect a second 100Ω resistor—R239a—between the hole left vacant by the removal of J28, and the nearest hole left vacant by the removal of J27. These are not the J27 and J28 holes that go to the switch bank—they're the other ones. Use sleeving on the resistor leads.

This completes the connection of the Aux Outputs. But, they have no muting relay. This is an easy problem to fix—simply connect the muting relay previously used for the Lab and Norm outputs (*Photo 24*).

- Make ¼″ cuts in the two PC traces that go from the left and right Lab output back to J62 and J63. Cut them between RY901 and J62/J63, as close as possible to relay RY901. Make sure to remove all copper material.
- Remove C207 and C208.
- Add jumpers made from insulated hookup wire between the processor out left and right PC pads (now the Aux Outputs) and the C207 and C208 hole closest to RY901.

The Aux Outputs will now be muted on turn-on and turn-off just like the main outputs.

VIDEO/AUX INPUTS

My system hasn't been biamped in several years, so I no longer need the Aux Outputs. I found another use for the pair of high-quality RCA connectors mounted in the old processor loop holes. I am now using a DVD player that plays regular CDs, SACDs, and DVD-Audio discs. The player must be used as a stand-alone when playing SACDs, but I use my outboard digital processor for CDs and all DVDs, including DVD-Audio discs. So, I need an additional pair of high-quality line input jacks.

I decided to replace the Adcom

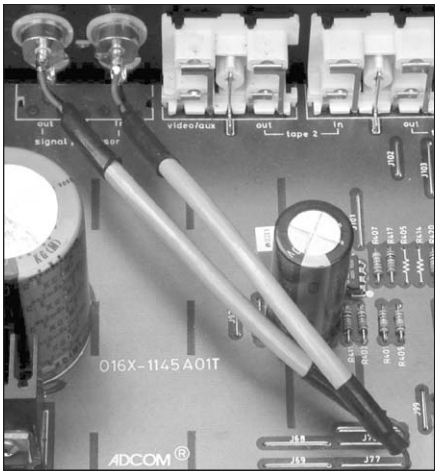


PHOTO 25: The Video/Aux inputs connected to new input jacks. This is an alternate use for the new RCA jacks if the Aux Outputs are not needed.

PC-mount Video/Aux jacks with the new jacks, connecting the new jacks to the Video/Aux traces on the Adcom PC board (*Photo 25*).

- Remove jumpers J76 and J77. This interrupts the signal paths between the Video/Aux jacks and the selector switches S206 and S207.
- Connect the new jacks to the selector switches using short lengths of D.H. Labs BL-1 interconnect. The shields and black center conductors are soldered to the ground washers on the new RCA jacks (the ground washers should already be connected to the PC board). On the other end, the shields and black center conductors should float—cut these leads short and use heat-shrink tubing to insulate them. Connect the red center conductors to the J76 (right) and J77 (left) holes closest to S207. You'll probably need to enlarge the holes slightly.

Note that the old PC-mount Video/Aux jacks are not removed from the PC board—they have simply been disconnected. The new jacks will allow you to connect your DVD player's output to the high-quality Video/Aux inputs, and your digital processor's output to the CD inputs, or vice versa.

THE LINE STAGE

Adcom's original line stage consisted of a Linear Technology LT1056 op amp (Adcom 6A) buffered by an LT1010 (Adcom 1A) placed inside the feedback loop. The LT1056 is an FET-input device with low enough DC offset to allow direct coupling of the output. This line stage was an excellent performer at the time it was designed, but there are a number of op amps and buffers that offer better performance today.

A complete schematic of the new line stage, including input switching, balance and volume controls, and output muting is shown in *Fig. 6*. This line stage is based on the Walt Jung design I published in the Philips DAC960 mod series in *TAA* back in 1992[1]. A high-performance op amp—IC201—provides the gain. IC201 is buffered by a high-speed, current-feedback op-amp—IC203—placed inside the feedback loop, which provides 100mA of output current capability. In the original Adcom line stage, the voltage gain was set at 11. The new circuit has a voltage gain of 5, which is plenty due to the much-improved dynamics of the new line stage powered by the new supply regulators.

The design shown in *Fig. 6* has one important refinement: the addition of local feedback around the gain op amp, IC201. This circuit was first described by Walt Jung in his column "Walt's Tools and Tips," for a time a regular feature in *Electronic Design*[2]. More recently, Walt described this topology in the 2002 edition of Analog Devices' *Op Amp Applications*[3].

The local feedback around IC201 serves two purposes. First, the effective open-loop bandwidth of IC201 is increased. Second, resistor R253 is approximately equal to the impedance of the volume control, set to a normal listening level, at the non-inverting input of the op amp. This helps match the source impedances at the op amp's inverting and non-inverting inputs. Local feedback is also used around the AD811.

The active devices are the same as those used in the DAC960 project—an AD744KN op amp buffered by an AD811AN high-speed current-feedback amplifier. The AD711KN has a unique feature: the output can be taken from compensation pin 5, bypassing the Class-AB output stage. This is how the chip was used in the DAC960 project. Walt informed me that pin 5 is capable of driving the high-impedance local feedback network in the new topology. *Figure 6* shows this alternate output connection, in addition to the conventional connection to Pin 6.

In the Analog Devices *Op Amp Applications* article, Walt suggested the more recently designed AD825AR, which, on paper, seems to offer even better performance than the AD744. I don't yet have enough experience with the AD825AR in this design to recommend it, but I may report on it in a future issue.

OLD CIRCUIT REMOVAL

Begin by removing some of the old line stage components.

- Remove IC201 and IC203 in the left channel.
- Remove IC202 and IC204 in the right channel.
- Remove D201 (L) and D202 (R).
- Remove R235 (L) and R236 (R).
- Remove C201 (L) and C202 (R).
- Remove C223 (L) and C224 (R).

Now change the input and output series resistors (*Photo 26*).

- Replace R205 (L) and R206 (R) with 475Ω resistors.
- Replace R239 (L) and R240 (R) with 100Ω resistors.

Next, you must replace a few more resistors.

- Replace R237 (L) and R238 (R) with 100Ω resistors.
- Replace R213 (L) and R214 (R) with 1.0k resistors. Do not put the new resistors in the old locations. Instead, put the new resistors in the holes adjacent to the original R213 and R214 footprints. This will allow adequate clearance when you install the op amp assemblies. (These adjacent holes are electrically identical to the

ones marked "R213" and "R214.")

- Replace R211 (L) and R212 (R) with 249Ω resistors.

R209 and R210 are 10M bias return resistors for the op amps. These resistors keep the DC offset at safe levels in case the volume control wiper opens or becomes intermittent. The currents through these resistors are so small that the steel end caps in the original Roederstein MK2 resistors are probably quite benign. However, if you want to be a real purist, you can replace them with Vishay-Dale VMF type RN55 resistors (I did). Don't waste money on the expensive Caddock or Vishay resistors in these locations—they're unlikely to yield any further improvement.

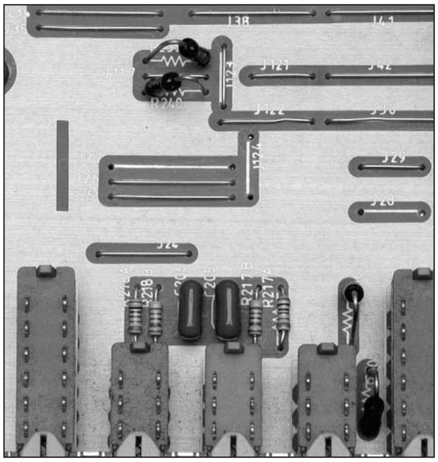


PHOTO 26: New 100Ω main output resistors are shown at the top of the photo. Replacement 475Ω input resistors are in the lower right, between two of the switches.

- Replace R209 (L) and R210 (R) with 10M Vishay-Dale CMF type RN55.

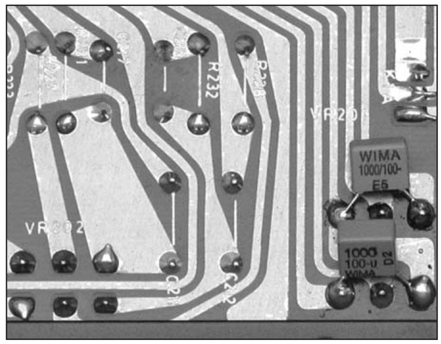


PHOTO 27: 1000pF capacitors are installed across the left and right sections of the balance control for bandwidth limiting. Wima FKP–2 polypropylene caps are shown here. Reliable MultiCap PPMFX types are an even higher–quality alternative.

Bandwidth limiting for the new circuit is set to 330kHz by the R/C network R205/C251 (R206/C252; in the assembly steps that follow, all components designators in parentheses refer to the right channel). C251 and C252 are soldered across the left and right sections of the balance control, on the bottom of the PC board (*Photo 27*). The best capacitor for this application is the Reliable MultiCap PPMFX polypropylene type; a lower-cost polypropylene alternative is the Wima FKP-2.

- Solder C251 (C252)—1nF—across the left and right sections of the balance control, on the bottom (trace) side of the PC board.

The new op amps and buffers won't simply drop into the existing PC footprints. Gain op amp modules IC201 and IC202 must be built up on an 8-pin DIP header with the local feedback resistors R251 and R253 (R252 and R254). I prefer the Tyco Plug Adapter Assembly sold by Allied, which has gold-plated pins. The Aries header sold by Digi-Key has tin-plated pins, and will also work. Both have forked contacts on the top, which facilitates soldering the op amp and resistor to the header. Assembling the headers is relatively easy (*Photo 28*).

- Cut off the thin part of the lead from pin 2 of the AD744 op amp. Leave the wide portion of the lead intact.
- Solder the op amp to the header, except for pin 2.
- Solder R251 (R252)—221k—between pin 6 and pin 2 of the op amp.
- Solder R253 (R254)—2.49k—between pin 2 of the component carrier and the R251 (R252) lead that is already soldered to pin 2 of the op amp.
- Solder the two assemblies in the IC203 and IC204 footprints. Pay careful attention to orientation, which is the same as the original op amps.

If you decide to try the AD744 with the output taken from pin 5, you can easily move R251 (R252) from pin 6 to pin 5 of the op amp. In this case, I suggest putting 100Ω resistors R237 and R238 on the bottom of the PC board. This will make the output connection to pin 5 easier. Note that the 221k feedback resistors must be connected to the same pin as the output, as shown in *Fig. 6*. After the line stage is completed, you may wish to conduct some listening experiments to see which connection you prefer.

You can build the AD811 modules on two small pieces of perfboard, ⅝″ × 15⁄16″, or 6 solder pads wide by 9 solder pads long (*Photo 29*). Here's how to build the modules:

- Solder unused pins 1, 5, and 8 of the AD811 op amps to pads on the perfboard. This will hold the op amp in place while you connect leads to the other pins. Orient the op amps along the length of the perfboard, leaving two rows of holes between pins 4

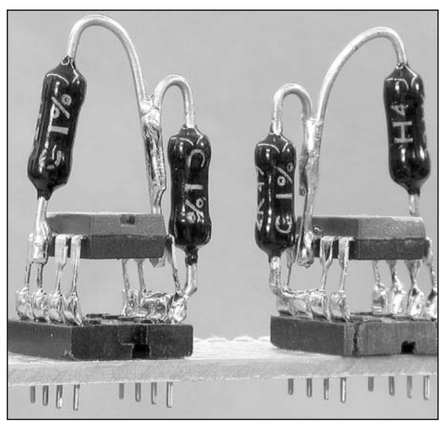


PHOTO 28: Two views of the AD744 op amp modules using Holco H4 resistors. The op amps are soldered to a header, along with the local feedback resistors. R253 (R254) is soldered between pin 2 of the op amp and pin 2 of the header.

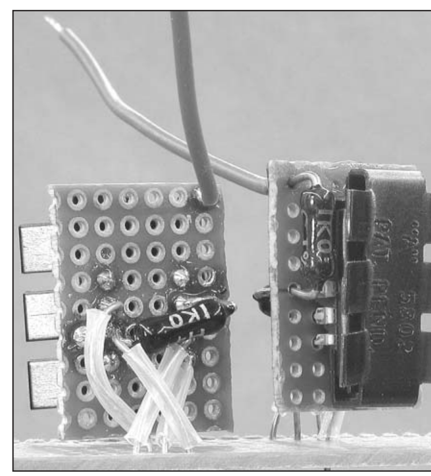
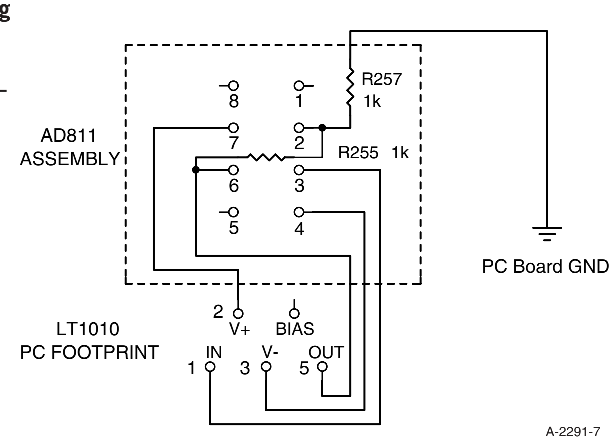


PHOTO 29: Front and back views of the AD811 modules, showing the local feedback resistors. A clip–on DIP heatsink is used on the AD811.

FIGURE 7: Diagram showing connections from the AD811 modules—assembled on small perfboards—to the LT1010 footprint on the Adcom PC board. The bias connection for the LT1010 is not used.



and 5, and the edge of the perfboard.

- Solder R255 (R256)—1k—between pins 2 and 6 on the back side of the perfboard.
- Mount R257 (R258)—1k—on the op amp side of the perfboard. Solder one lead to pin 2 of the op amp, or the corresponding lead of the R255 (R256). Solder the other lead to a spare solder pad, and solder a 4″ length of insulated hookup wire to this lead.
- Solder four 2″ leads of 22AWG solid hookup wire to pins 3, 6, 4, and 7 of the op amp.
- Slide clip-on DIP heatsinks onto each AD811. 8-pin DIP heatsinks are fine, but they are hard to attain from most distributors. I used the Aavid heatsinks made for 14-pin DIP applications. Center the heatsink on the AD811 and use a dab or two of five-minute epoxy to keep the heatsink in place. Allow the epoxy to cure before proceeding.

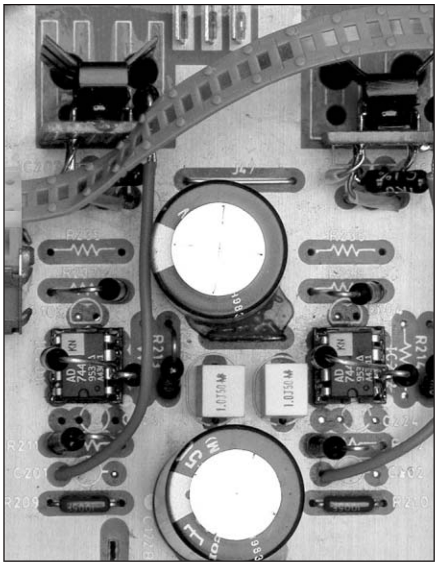


PHOTO 30: The completed line stage using Holco H4 resistors. The insulated wires from the AD811 local feedback resistors are soldered to the ground pads previously occupied by C201 and C202.

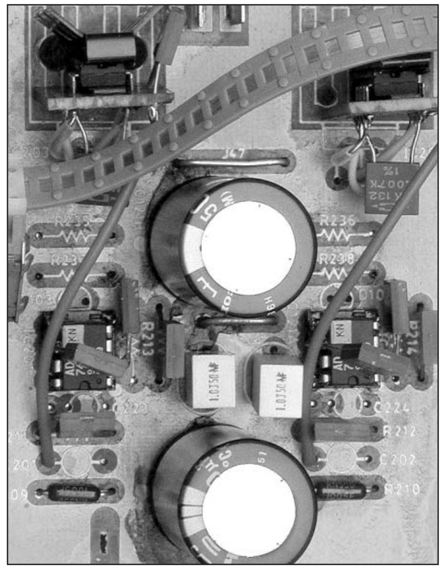


PHOTO 31: The completed line stage built with Caddock MK132 resistors. R237 and R238 have been soldered to the bottom side of the PC board, allowing easy connection to pins 5 or 6 of the AD744.

Now, install the modules on the Adcom PC board. *Figure 7* shows how the modules connect to the LT1010 footprint.

- Bend the four bare leads of the modules as shown in *Photo 29*. Put short lengths of sleeving over the leads and install the leads in the LT1010 PC footprints, making sure that the connections match the wiring diagram in *Fig. 7* (*Photo 30*). Solder the four leads to the bottom of the Adcom PC board. Note that the bias connection in the LT1010 footprint is not used.

- Connect the insulated leads from R257 and R258 on the AD811 modules to the ground holes on the Adcom PC board previously occupied by C201 and C202. These are the C201 and C202 holes closest to the preamp's front panel. Trim the leads so that they are just long enough to reach comfortably. Solder the leads in place (*Photo 30*).
- Alternately, you can route these leads through screw holes in the PC board that were used to hold the LT1010 heatsinks in place and sol-

TABLE 1
MODIFIED GFP-565 PREAMPLIFIER

MEASUREMENTS ON SAMPLE BUILT FOR C. VICTOR CAMPOS		
S.N. AC14755		
LINE STAGE THD – LEFT	**WIDEBAND**	**w/80kHz LP FILTER**
(2V in > 2V out)		
20Hz	0.0037%	0.0017%
1kHz	0.0037%	0.0018%
10kHz	0.0042%	0.0028%
20kHz	0.0055%	0.0046%
LINE STAGE THD – RIGHT	**WIDEBAND**	**w/80kHz LP FILTER**
(2V in > 2V out)		
20Hz	0.0037%	0.0017%
1kHz	0.0037%	0.0018%
10kHz	0.0043%	0.0031%
20kHz	0.0055%	0.0048%
LINE STAGE IMD	**SMPTE (4:1)**	**1:1**
60Hz + 7kHz		
(2V in > 2V out)		
LEFT	0.0014%	0.0013%
RIGHT	0.0015%	0.0012%
LINE STAGE FREQUENCY RESPONSE		
–0.2dB @ 10Hz, –2.1dB @ 100kHz (Left and Right Channels Identical)		

All measurements made with Sound Technology 1700B
by Gary Galo, 11/2/2002

PARTS LIST

TAPE OUTPUT BUFFERS:	
(2)	TI/Burr–Brown OPA2604AP op amps, Digi–Key OPA2604AP–ND,—or—
	TI/Burr–Brown OPA2134PA op amps, Digi–Key OPA2134PA–ND,—or—
	Analog Devices AD712JN op amps, Newark 05F7277, or Analog.com
LINE STAGE:	
(2)	10M, ¼W resistors, Vishay–Dale CMF Type RN60, Mouser 71–RN55D–10M (R209, R210)
	All other resistors are builder's choice:
	Vishay–Dale CMF Type RN60 (Welborne Labs, Mouser 71–RN60D–Value)
	Holco H4 (older non–ferrous type, if available, Welborne Labs, Michael Percy Audio, Parts Connexion)
	Caddock MK132 (Welborne Labs, Michael Percy Audio, Parts Connexion)
	Vishay S–102 (Michael Percy Audio)
(2)	475Ω (R205, R206)
(2)	221k (R251, R252)
(2)	2.49k (R253, R254)
(6)	1.0k (R213, R214, R255, R256, R257, R258)
(2)	249Ω (R211, R212)
(4)	100Ω (R237, R238, R239, R240) (six are needed if you add the Aux Outputs, R239a, R240a)
(2)	1nF (1000pF, 0.001μF) Polypropylene Capacitors,
	Reliable MultiCap PPMFX (Welborne Labs, Parts Connexion)—or—
	Wima FKP–2 (Welborne Labs) (C251, C252)
(2)	Analog Devices AD744KN op amps, Newark 05F7342, Analog.com
(2)	Analog Devices AD811AN op amps, Newark 05F7667, Analog.com
(2)	Aavid/Thermalloy DIP Heatsink, Digi–Key HS179–ND
(2)	Tyco 8–pin Plug Adapter, Allied 905–3114—or—
	Aries 8–pin DIP Header, Digi–Key A101–ND
MISC.	
Insulated hookup–wire	
22AWG bare hookup wire	
Sleeving	
Perfboard with solder ringed holes (Radio Shack 276–147)	
5–minute epoxy (local hardware store)	
D.H. Labs BL–1 Interconnect for optional Aux Inputs	

der them to the main ground traces on the bottom of the PC board. You'll need to scrape enough lacquer off the PC traces to make the ground connection.

Photos 30 and *31* show the completed line stage with both Holco and Caddock resistors. At this point, check all wiring and component placements very carefully. Be particularly careful in checking the wiring of the AD811 modules and the connections from these modules to the Adcom PC board. Carefully clean the Adcom PC board around the line stage components with the CaiKleen TRP DG7S-6 cleaner recommended in Part 2.

TESTING

Power up the preamp and check the DC-offset at the Main output jacks, after the muting relays have timed out. Offset will normally be 1.25mV or less. Analog Devices specifies the typical input offset voltage for the AD744KN as 0.25mV, and the line stage has a voltage gain of 5. Maximum offset is specified as 0.5mV, so the worst-case situation in this line stage should be 2.5mV at the output.

Check the voltage gain with a 1kHz sine wave. Set your generator output to 0.5V RMS and check the output of both channels with the volume control fully clockwise. It should be 2.5V RMS. If you have a distortion analyzer, set the generator output to 2V RMS, and advance the preamp volume control for an output level of 2V RMS. This is the "unity-gain" setup that Victor Campos used to measure preamp THD when he worked at Adcom.

This setup minimizes the effects of noise on the THD measurement. *Table 1* gives the measurements of the GFP-565 preamp that I modified for Victor Campos. They were made on my Sound Technology 1700B analyzer, and are typical of what you should measure with this preamp. These measurements are quite close to the Adcom original, and I don't believe that they are much help in quantifying the sonic performance of the modified preamp. But, they will tell you whether you have made any serious errors in performing these mods.

I suggest "cooking" the preamp on the bench—with the cover and bottom plate installed—for 24 hours and re-checking the measurements. If all is well, hook the preamp up to your system, and enjoy! We plan to have at least one third-party evaluation of the modified preamp in a future issue of *aX*, and I welcome your comments and reactions. Part 4 will feature the new, servo-controlled phono preamp. ❖

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1. Galo, Gary. "Pooge 5: Rite of Passage for the DAC960"–Part II. *Audio Amateur*, 3/1992, pp. 34–39, 58.

2. Jung, Walt. "Op-Amp Audio: Minimizing Input Errors," *Electronic Design*, Dec. 14, 1998, pp. 80–82.

3. Jung, Walt. "Composite Current Boosted Line Drivers," *Op-Amp Applications*, Analog Devices, 2002, pp. 6.62–6.64.

SOURCES

Allied Electronics

7410 Pebble Drive
Fort Worth, TX 76118
817-595-3500
817-595-6444 (FAX)
http://www.alliedelec.com/

(Remaining sources are listed in Part 1)

Why Speakers Have Slanted Fronts, Part 2

Part 2 continues our examination of vertical polar plots for common crossover functions with a pair of spaced ideal drivers. **By G. R. Koonce**

Figure 5 is the vertical polar plot for the third-order Butterworth (B3) CO at the CO frequency. The results are very similar to that of the B1 CO. The main difference is that for normal tweeter polarity the B1 main lobe is below zero degrees, while for the B3 it is above zero degrees. Again the system response on-axis is at 0dB. The horizontal polar plot is still a circle at 0dB.

Figure 6 is the vertical polar plot for the fourth-order Linkwitz-Riley (LR4) CO at the CO frequency. As expected, the LR4 shows a null on-axis with the top driver inverted polarity. With both drivers normal polarity, the system response is a wide, symmetrical lobe centered on-axis and with an on-axis response of 0dB. The horizontal polar plot with normal tweeter polarity is a circle at 0dB.

It is clear that CTC spacing causes a major change in the system response. This is true with ideal drivers even without any CO. It is well known that if you use dual drivers it is important not to try to use them too high in frequency.

Figure 7 shows the vertical polar responses at a variety of frequencies for two ideal 6″ woofers mounted vertically at the minimum possible 6″ CTC spacing. Note that by 1,500Hz response nulls are appearing by about 49° off-axis. For a sound velocity of 1128ft/s the nulls would be at 48.76°.

When you start stacking drivers vertically, you produce an equivalent driver with a large effective vertical "diameter." The vertical directivity of this equivalent driver causes its vertical radiation beam to start to narrow at a relatively low frequency. The nulls produced can modify the main lobe response produced by any CO used with these dual drivers. The +6dB response on-axis occurs because two drivers driven in parallel with identical signals show a 6dB response rise. The horizontal polar plot is a circle at +6dB (with the ideal drivers).

ELECTRICAL VERSUS ACOUSTIC CROSSOVER SLOPES

To this point we have been looking at ideal drivers. Thus the acoustic frequency response of the driver with its CO network shows the same shape as the CO network alone. This is no longer true for real drivers, which have their own response slopes that make the acoustic slope different from that of the electrical CO alone.

Many builders choose to build with the LR4 CO, which has ±24dB/octave slopes, and the two CO outputs are in-phase at the CO frequency. As you have seen, this produces a wide system response lobe that is centered on-axis. However, if you simply build an electrical LR4 CO (active or passive) and use it to drive real drivers, you will likely not end up with the desired LR4 acoustic CO.

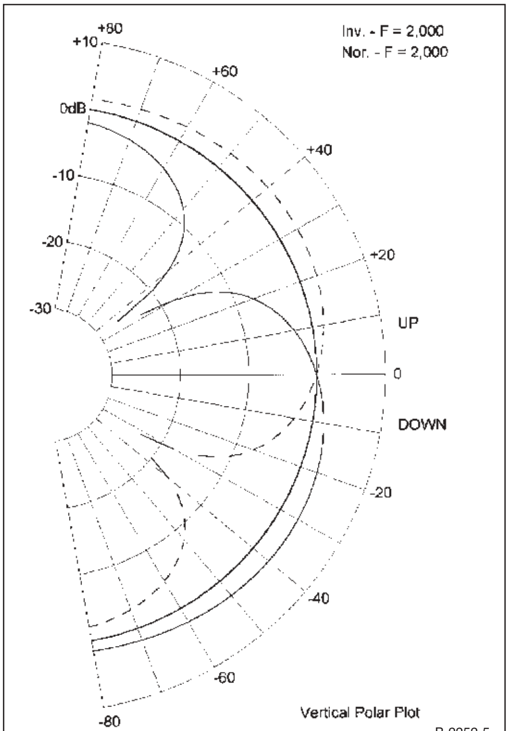

FIGURE 5: Ideal third-order Butterworth CO—solid curve is inverted tweeter polarity; dashed curve is normal tweeter polarity.

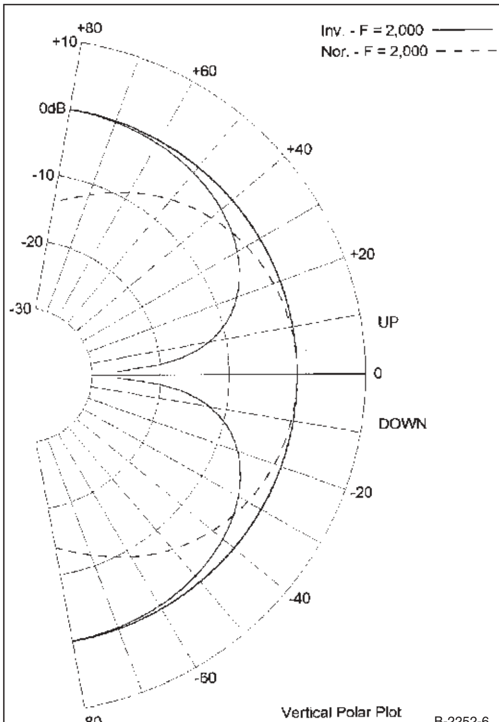

FIGURE 6: Ideal fourth-order Linkwitz-Riley CO—solid curve is inverted tweeter polarity; dashed curve is normal tweeter polarity.

Working with CO design via modeling, I have found that many times you can come very close to an acoustic LR4 CO by using an electrical CO of lower order and taking advantage of the additional slopes introduced by the drivers themselves. This is because at the limit of their response most drivers tend to rise or fall at the first- or second-order rates of ±6 or ±12dB/octave, respectively (see reference 3 of Part 1 for coverage of this topic). It is for this reason that the "rule of thumb" to connect the tweeter with inverted polarity with a second-order electrical CO is not always valid. In modeling many systems with a second-order electrical CO, I find about 50% are best with the tweeter inverted polarity and an equal number best with normal tweeter polarity.

POLAR PLOTS WITH SOME REAL DRIVER FEATURES

Now consider the performance of systems with drivers having some "real" features. The drivers are still considered as having a flat frequency response and 8Ω resistive input impedance, but they now have some of the characteristics of real drivers. We are still assuming a 4″ woofer and 1″ dome tweeter, but mounted at a more realistic 5″ CTC spacing. The directivity of each driver is included.

Many years of testing and modeling have shown that the best piston diameter to use for modeling a driver's directivity is not always the same as the actual cone/dome diameter. I have seen 1″ domes that modeled best (agreed best with test results) with a piston diameter in the range 0.8″ to 1.3″. We will use 1″ for our example system.

Similarly, the best piston diameter to use to model a woofer may be greater or less than its actual cone diameter, depending on cone shape and dust cap or phase plug shape. For example, I have seen 8″ woofers that modeled best using a piston diameter over the 6″ to 9″ range. We will use 4″ for our theoretical 4″ driver and also include the fact that the driver acoustic center is located behind the mounting plane using a horizontal offset of back 2″ for the woofer and back 0.25″ for the tweet-

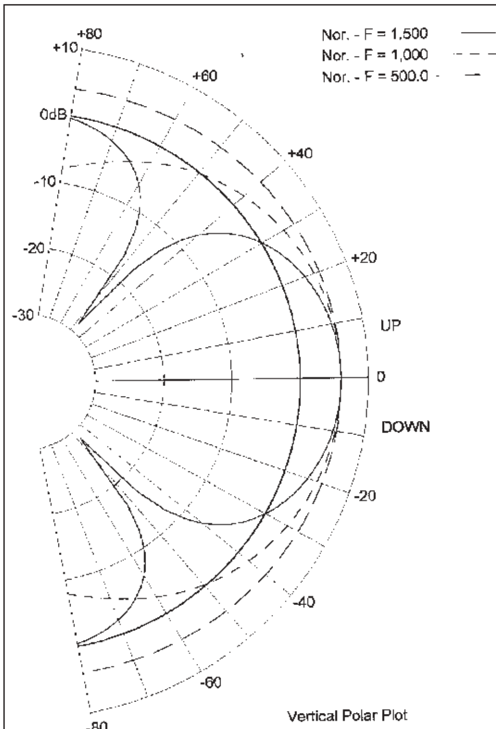

FIGURE 7: Two 6″ diameter ideal drivers mounted 6″ CTC with no crossover. Curves identified on plot. B-2252-7

er. Even though the drivers are not contributing any frequency response change, the system frequency response is considerably changed.

Figure 8 shows the results for the LR2 CO with the tweeter having inverted polarity at several frequencies about the 2kHz CO frequency. The best (flattest with frequency) response is no longer at zero degrees, but has been shifted to near –20°. The enclosure for such a system should place the seated listener at an angle of about –20° to the tweeter axis; i.e., you should listen somewhat below the tweeter height. The response stays fairly flat from only about –10° to –30°, a spread of 20°. This is typical of the usable vertical lobe you might get with a two-way system, and this is with perfectly flat driver responses.

Figure 9 shows plots for the LR4 CO used with the same drivers. Results are virtually the same as with the LR2 CO—the vertical lobe's center is depressed to near 20°. Note the LR4 is used with both drivers wired normal polarity.

Figures 10 and 11 show results for an LR2 CO system assuming an 8″ woofer and 1″ dome tweeter with a 4″ faceplate. This forces a minimum CTC spacing of 6″. Remember the driver responses and input resistances are still assumed flat across the audio band.

Figure 10 shows a spread of plot frequencies, while *Fig. 11* shows a group of frequencies around the CO frequency. Performance is slightly worse than the smaller woofer systems in *Figs. 8* and *9*. This same system was examined with the LR4 CO with nearly identical performance, so the plots are omitted.

The 8″ woofer two-way system is one of the most difficult design problems and is only practical if you can find a tweeter that you can use at a rather low frequency to keep the CO frequency low. The tweeter in this system is moved laterally off the woofer centerline by 4″ to the left. Since our reference is always

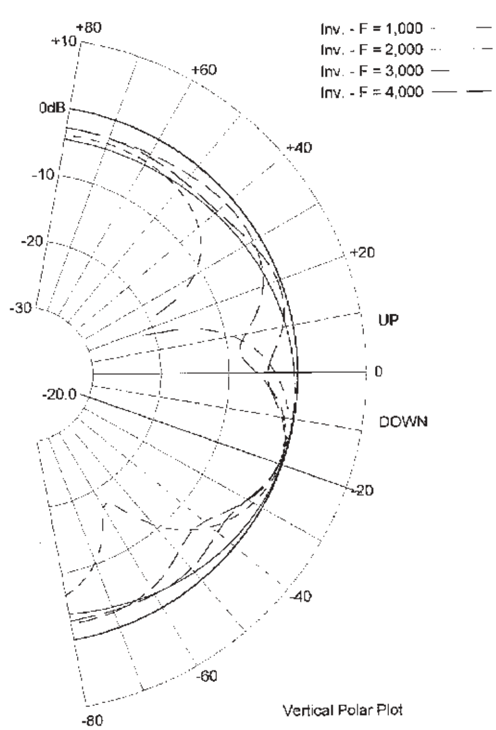

FIGURE 8: Polar plot for LR2 crossover with some real driver effects. B-2252-8

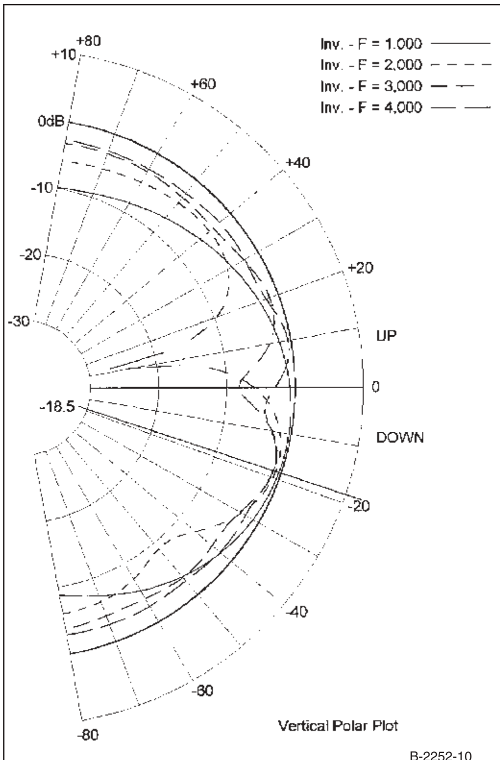

FIGURE 10: Polar plot for 8″ woofer system with LR2 CO—wide frequency spread in plots.

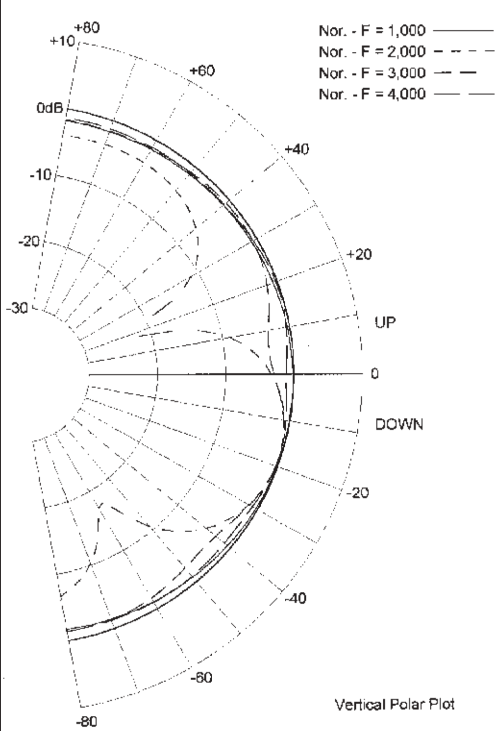

FIGURE 9: Polar plot for LR4 crossover with some real driver effects. B-2252-9

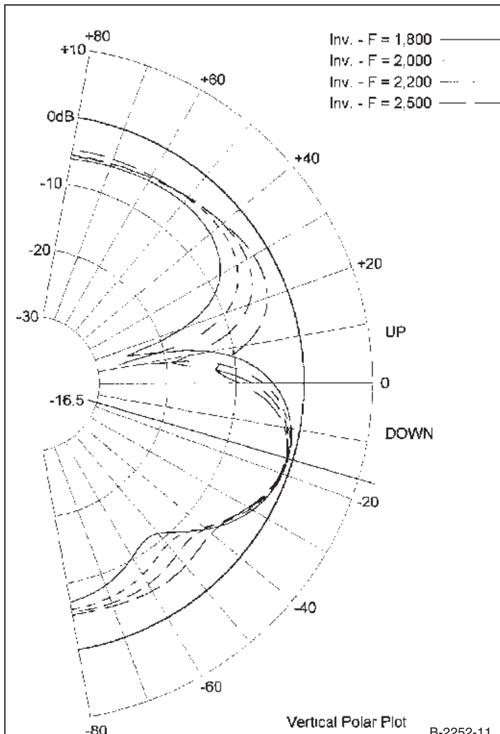

FIGURE 11: Polar plot for 8″ woofer system with LR2 CO—plot frequencies near CO frequency.

the tweeter's center (*Fig. 2* of Part 1), this is equivalent to moving the woofer 4″ to the right. This has no effect on the vertical polar plot, but as *Fig. 12* shows it sure bothers the horizontal polar plot, which would indicate a system with possible imaging problems. *Figure 13* shows that a 1″ tweeter lateral offset causes little problem, but large offsets should be avoided.

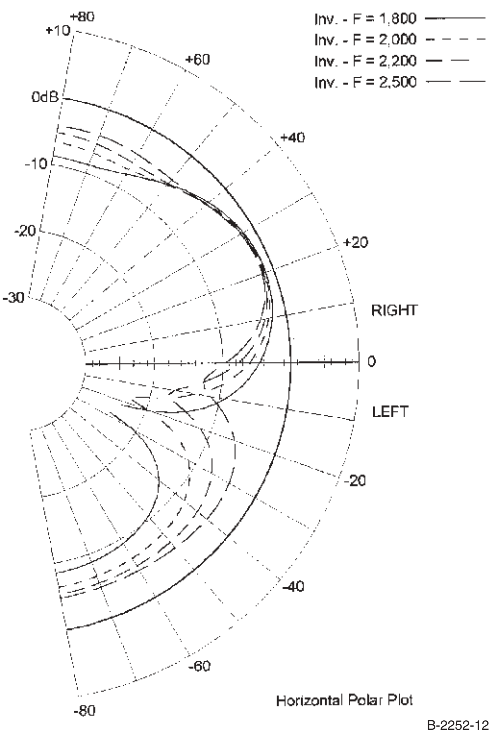

FIGURE 12: Horizontal polar plot for 8″ woofer system with LR2 CO—plot frequencies near CO frequency.

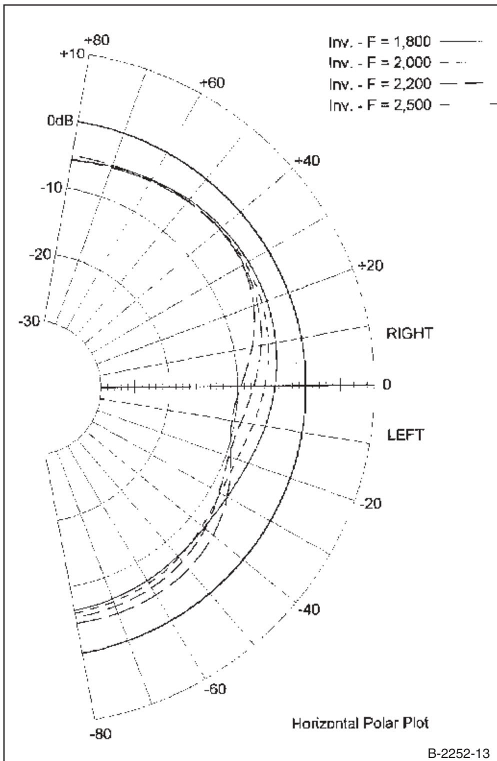

FIGURE 13: Horizontal polar plot for system in *Fig. 12* with tweeter lateral offset reduced to 1″.

SUMMATION DISTANCE

So far the polar plots have been generated with summation at infinity. This assumes the summation point is so far away that the angle to each driver from the listening point is the same. While summation at infinity makes the math easy, it does not represent how you normally listen to or test a speaker system.

In the real world you normally listen at a distance from the front panel that is much less than infinity, perhaps about 10′ (120″). For testing purposes you may be limited to a distance even less than this, possibly limited by your room size. These polar plots can thus be slightly misleading if you are going to test your CO design at a relatively small distance.

When you generate polar plots at a fixed distance it will make some change in the system response. The output from each driver is summed at a point on a plane a fixed distance out from the front panel. A line through the tweeter's center at the specified angle off-axis indicates the location of this point (*Fig. 14*).

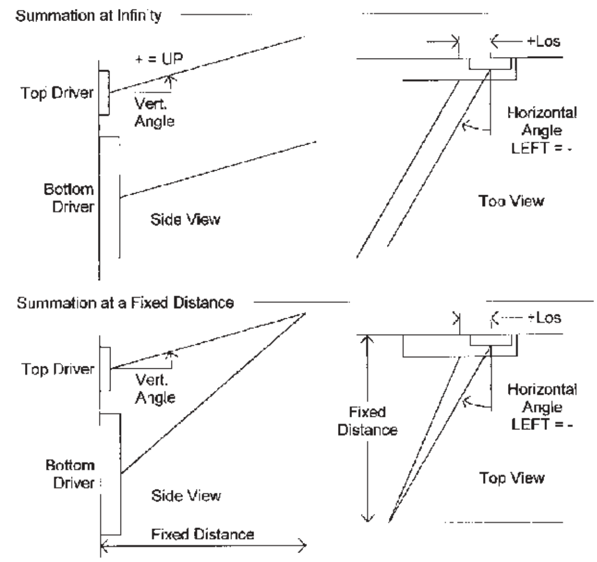

FIGURE 14: Comparison of summation at infinity and at a fixed distance

B-2252-14

Any driver lateral offset (Los) or CTC spacing means the woofer is looking at this point with a different angle than that specified for the tweeter. This produces a difference that will be very important to quasi-anechoic testing, but probably has less effect on the perceived listening response. *Figure 15* shows results for the conditions of *Fig. 10* when summation occurs at a fixed 60″ from the front panel. The major change shown is that the center of the main lobe has moved down a few more degrees.

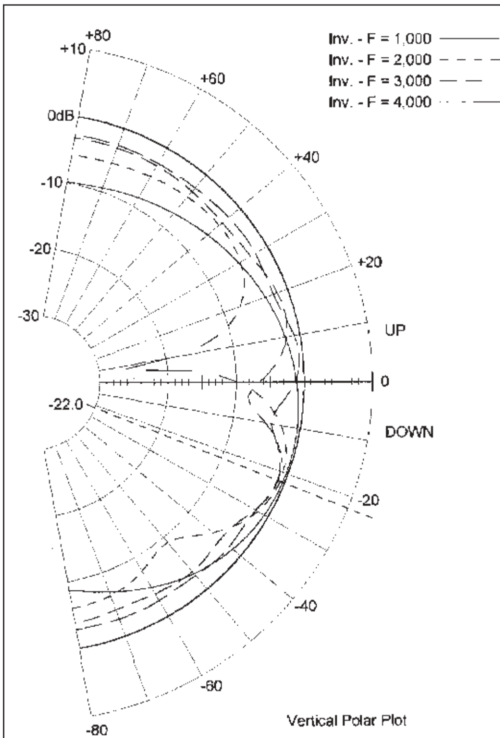

FIGURE 15: Polar plot for system of *Fig. 10* with summation at 60″.

B-2252-15

This article has now introduced most of the effects of real drivers on the system performance. The missing items are the effects of the driver's actual frequency response (magnitude and phase) and of the driver's actual input impedance on the CO network. These will be introduced later when we study some actual systems.

THE SYSTEM DIRECTIVITY PLOT

You are dealing with three variables when you try to understand the system over space: the system response, the frequency, and the angle indicating the point of interest. You have seen the polar plot contains a series of curves showing the system response versus angle with each curve being for a fixed frequency. The system directivity plot, which slices things the other way, is a series of curves showing system response versus frequency with each curve at a different angle.

Figure 16 shows a typical horizontal directivity plot as my software generates them. The center curve is the system response on the axis of interest (which may be angled off the tweeter's axis). Each plot moving forward represents the system response 10° further to the left (facing the system) of the reference axis. Similarly, each curve behind the reference axis curve represents the system response ten additional degrees to the right.

I'm sure readers have seen system directivity plots in test reports on speaker systems. They may look different from *Fig. 16*. Many times the plots show curves for each 5° angle in the horizontal plane from on-axis and may cover the full ±180° range.

Figure 16 plots actual responses; i.e., each curve is the system response at that angle. Many prefer to plot the relative response. Here the on-axis response is shown as a straight line and what is plotted at the other angles is the difference from the on-axis response. I prefer the actual response type plot, which I use in my software.

The vertical system directivity curve is similar. The center curve is system response on the reference axis, and curves in front are moving vertically in one direction and curves behind are for the other direction. Again these plots may be actual or relative to the on-axis plot.

Since, as you have seen, systems tend to go bad in the vertical plane much more quickly, most vertical directivity plots cover less total angle range and may use less angle change between curves. My vertical directivity plots use 5° steps and cover ±20° from the reference axis. My software allows the reference axis (0°) to be angled off perpendicular at the tweeter's center in the horizontal plane if desired.

Next month, in Part 3, we'll get down to specifics with a look at some real-world system examples and their respective polar and directivity plots. ❖

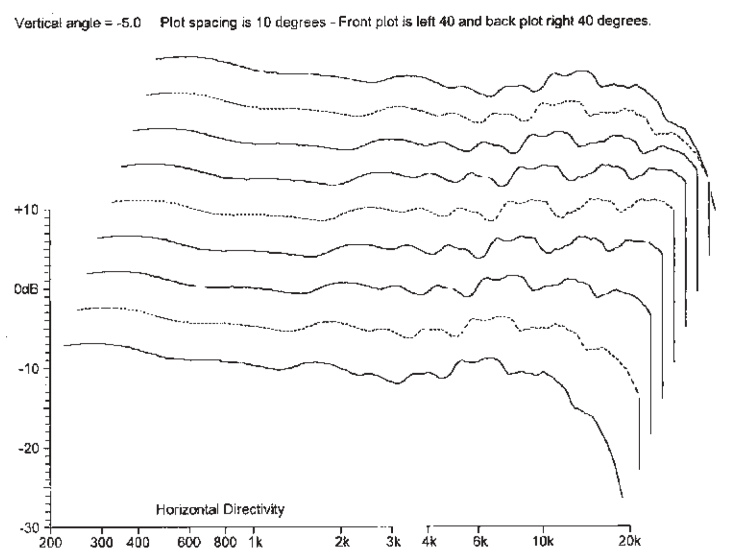

FIGURE 16: A typical system horizontal directivity plot.

B-2251-16

Product Review

Thorens TD 295 MK III

Reviewed by Charles Hansen

Thorens Export Co. Ltd., IM Junkholz 44 • CH-4303, Kaiseraugst-Basel, Switzerland, phone: ++41 (0)61 813 03 36 • FAX: ++41 (0)61 813 03 39, www.thorens.ch (info@thorens.ch), $899, warranty 2 years. Supplied by Trian Electronics, Inc., 5816 Highway K, Waunakee, WI 53597, 608-850-3600. Dimensions: 435mm W × 358mm D × 145mm H (dust cover closed), weight: 8.5kg.

The TD 295 Mk III is a two-speed belt-driven turntable with a low voltage motor. The TP 42 tonearm is supplied "plug and play ready" with an Ortofon OMB 10 cartridge, and will accept any cartridge with standard ½″ mounting holes. The upper surface of the turntable base is available in three colors: black, mahogany, or anis wood, all with an elegant high-gloss multi-layer "piano" finish (*Photo 1*). The dust cover sits on two pivoting posts at the rear of the base.

PHOTO 1: TD 295 MK III front view.

CONSTRUCTION

The motor, platter, and the tonearm assembly are all solidly mounted to the upper section of the base assembly. Four sandwich layers of MDF make up the lower section of the base. The base assembly sits on four elastomer cones with felt-bottomed plastic feet.

The 15cm inner platter is die-molded from a rigid but lightweight fiber material and has an integral spindle. The outer rim of the inner platter is belt driven from the motor (*Photo 2*). The turntable spindle sits on a robust Thorens Safeguard TM ball bearing. The 30cm 2.3kg main platter is zinc alloy and is dynamically balanced. It sits on top of the inner platter and provides the mass inertia to ensure that vibrations are absorbed, and motor speed fluctuations and torque pulses are kept in check. A felt mat is provided for the platter playing surface.

A 15V AC 0.19A wall adapter that plugs into a 5.5mm jack on the rear of the turntable powers the 10-pole two-speed synchronous motor. The adapter feeds a built-in electronic oscillator that provides for either 33⅓ or 45 rpm operation, selected by a rotary switch on the left side of the base. A 45 rpm spindle adapter is provided. In the center stop position, the electronic controller applies dynamic braking to the motor windings to rapidly bring the turntable to a halt and hold it in position.

A push-button on-off switch sits to the right of the speed selector switch. In the off mode, the motor voltage is disengaged, so you can manually rotate the turntable. An optical sensor silently shuts off the motor when the tonearm reaches the lead-out grooves of a record. This optoelectronic LED sensor does not mechanically load the tonearm as would a micro-switch auto-stop sensor.

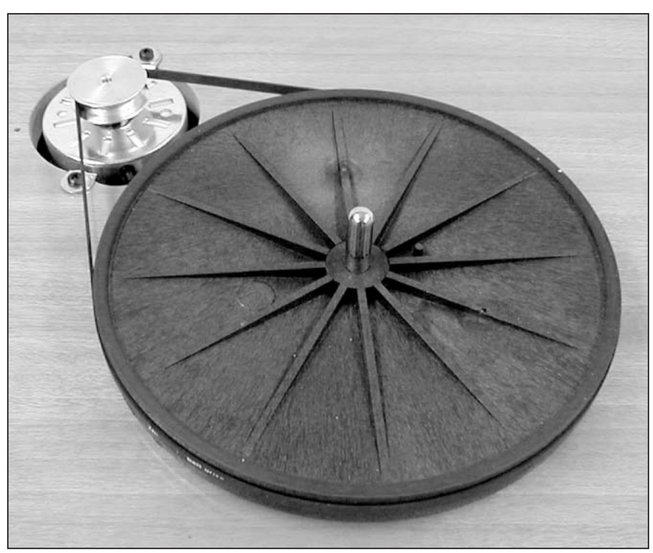

PHOTO 2: TD 295 MK III motor and inner platter belt drive.

TONEARM AND CARTRIDGE

The TP 42 tonearm is a straight hollow metal tube with an effective length of 229mm. The cartridge is mounted with a 25° offset and 18.8mm overhang. The tonearm is supported by two concentric gimbal needle bearings. The counterweight has an elastomer isolation insert, and threads onto the counterweight shaft for tracking force (TF) adjustments. The tonearm lift/cueing lever has a well-damped action that sets the stylus gently on the record.

Anti-skating is implemented by means of a small cylindrical weight.

You place its thin monofilament line in one of three notches in the anti-skating post at the rear of the tonearm. This is best done with tweezers. The weight then loops through a wire support arm that directs the monofilament line away from the tonearm and allows the weight to hang vertically above the base. Notch 1 is used for the OMB 10.

The tonearm is hard-wired to a 1m pair of shielded interconnects with gold-plated RCA phono connectors. An integral turntable ground wire is included to allow you to ground the TD 295 MK III to your phono preamp chassis. The tonearm wiring uses litz wire from the cartridge clips to the phono interconnects. Cable capacitance is listed as 160pF.

The standard Ortofon OMB 10 cartridge is pre-mounted to the tonearm, with VTA pre-adjusted at the factory. The OMB 10 has an elliptical diamond stylus with 8/18µm radii. Output voltage is 4mV at 5cm/s, 1kHz. The cartridge specs are given in *Table 1*.

If you choose to install a different cartridge, Thorens supplies a cardboard alignment protractor that you set onto the platter spindle. The pattern has reference points for 13° and 25° offsets that allow you to adjust the cartridge mounting screws so that the stylus-to-pivot and cartridge/headshell alignment is in accordance with specified dimensions.

Tracking force is very easy to set up, even without a stylus force gauge. You remove the stylus guard, and adjust the counterweight so the cartridge just floats at the record playing level. Then you hold the counterweight in place and turn the force dial ring until the top of the scale reads zero.

TABLE 1
ORTOFON OMB 10 CARTRIDGE SPECIFICATIONS

Output voltage	4mV at 5cm/s, 1kHz
Channel balance	1.5dB at 1kHz
Channel separation	22dB at 1kHz, 15dB at 15kHz
Frequency response	20–20kHz, +3/–1dB
	20–24kHz, ±3dB
FIM distortion	<1%, 1.5gm TF, DIN 45.542
Tracking ability	70µm, 315Hz, 1.5gm TF
Compliance, lateral	25µm/mN
Stylus	Elliptical, 8/18µm radii
Equivalent stylus tip mass	0.5gm
Tracking force range	1.25–1.75gm
Vertical tracking angle	20°
DC resistance	1kΩ
Inductance	580mH
Recommended load	47kΩ, 200–400pF
Cartridge weight	5gm

The TF ring is graduated from 0 to 30, for 0-30mN (milli-newtons) force. Finally, you turn the counterweight to the recommended TF reading (1.5gm for the Ortofon). When I checked the TF with my stylus force gauge, it read 1.5gm, so the counterweight scale appears to be sufficiently accurate.

Unfortunately, the 8-page owner's manual mixes grams and mN force without providing any conversion factor. The TF is given in gm, while the anti-skating reference is mN. Since 1gm = 9.8066mN, if you use 1gm = 10mN you will be close enough. While the manual is generally quite thorough in describing assembly and setup, it makes no mention of installing the main platter. It sits on the inner platter and requires no further alignment (there is no positive engagement between platters).

LISTENING TESTS

I used a number of jazz, pop, and classical music LPs for my listening audition. I had the Hagerman Bugle and DacT CT 100 phono preamps on hand for testing, which I used with a passive volume control. I also used the phono stage in my own control preamp (*Audio Electronics* 6/97, p. 8). The power amplifier is a Parasound HCA-1000A power amp. Loudspeakers are NHT SuperOnes and an SW2P subwoofer. After some time with each preamp, I decided on the Hagerman Bugle in conjunction with the passive volume control.

The Thorens table and Ortofon cartridge had no trouble negotiating the Shure Audio Obstacle Course test LP, or the high velocity tracks on the Stereo Review SR12 Stereo Test Record. There was no evidence of rumble on the unmodulated test record "silent" tracks.

Massed strings and choral voices were clear and well-defined. I would classify the Ortofon cartridge tonal balance as neutral, with an excellent midrange. Bass response was solid and well extended, so much so that it accentuated any low-level rumble I found recorded on certain LPs. If your phono preamp permits, I suggest engaging the 1380µs turnover filter when you encounter such recordings.

The turntable itself was completely free of rumble, flutter, and wow. It was quite insensitive to mechanical shock applied to the turntable shelf of my equipment rack. The elastomer feet and the dense mass of the base effectively absorbed vibrations and the air waves from the subwoofer before they reached the tonearm.

The TD 295 MK III brought out all the fine details in instruments such as brushed cymbals, triangles, acoustic guitar, strings, and the woody tone of the acoustic bass. Separation between instruments was very good and the soundstage was wide and stable.

After my extended playing interval, the AC adapter transformer was only slightly warm to the touch.

MEASUREMENTS

I used two test records: Stereo Review's SR12 Stereo Test Record (1969) and Hi-Fi News & Record Review's Test Record (HFN-001, 1996). I used the DacT CT 100 phono preamp for measurements due to its versatile cartridge load switching and selectable EQ time constants. I set the CT 100 for 47kΩ/200pF load and flat (3-point) RIAA response.

The frequency response for the TD 295 MK III is shown in *Fig. 1*, measured at the CT 100 phono preamp output jacks. I established 0dBr as 400mV preamp output using the 1kHz reference track on the SR12 test record. The overall response was determined by measuring the output voltage for the 19 warble tracks covering 20kHz down to 20Hz.

HF response rose at each end of the audio spectrum, but remained within the +3/−1dB specification for the OMB 10. The DacT CT 100 has an optional 3.18µs (50kHz) rolloff time constant switch that would level off the HF response. This time constant is standard in the Hagerman Bugle.

Channel separation for the six test tracks is shown in *Table 2*. Results were essentially the same for the R-L and L-R tests. Ortofon specifies −22dB for the OMB 10 at 1kHz.

My LED strobe showed both platter speeds to be absolutely spot-on. The results were the same whether a record was being played or not. While the motor does take several revolutions to overcome the platter inertia, once at speed there is no movement of the bars on the strobe disk to indicate any flutter or wow. The strobe emits very narrow light pulses that are synchronized to the 60Hz power line.

I measured the frequency of three reference frequency tracks on the test record. The 440Hz "A" measured 440Hz, the 1kHz track measured 989Hz, and the 3kHz track measured 2968Hz. The latter two tracks are only 1.1% below the specified frequencies if the test record tracks were done perfectly.

The inner platter and motor pulley diameters have about a 6.86:1 ratio. The motor would need to turn about 229rpm for 33⅓ rpm at the platter, and 308rpm for 45rpm at the platter. For a 10-pole synchronous motor, this requires an oscillator frequency at the motor stator windings of 19Hz and 25.7Hz, respectively.

I used a magnetic flux pickup placed near the motor end plate to try to determine the oscillator frequency. Using an oscilloscope, I found the flux waveform frequency to be very close to those values. The flux had a high percentage of second harmonic, indicating the stator may use two slots per

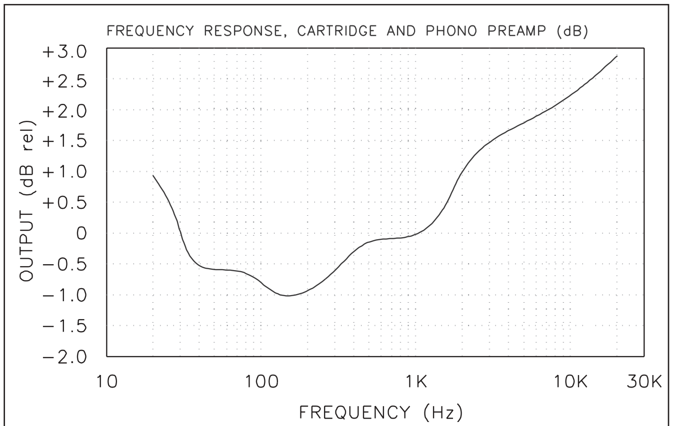

FIGURE 1: Frequency response. A-2284-1

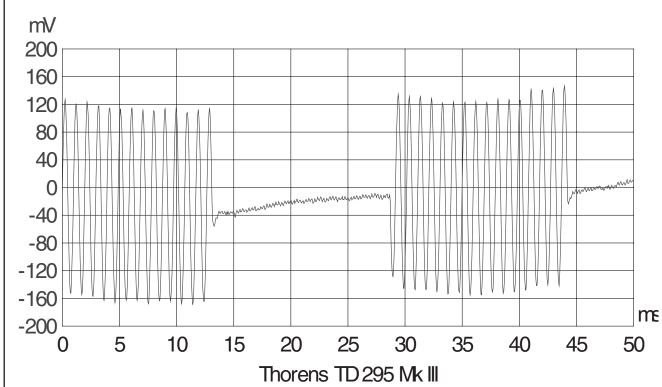

FIGURE 3: 1kHz tone burst response. A-2284-3

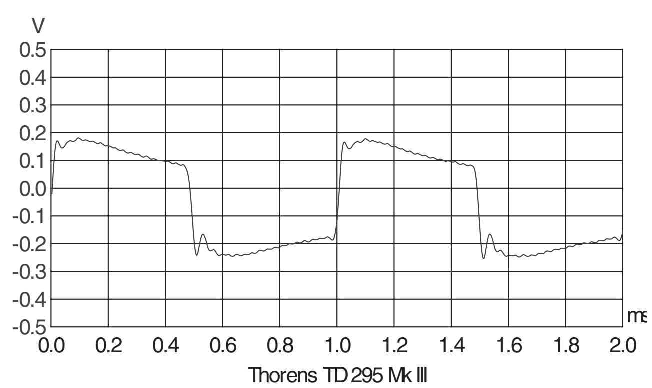

FIGURE 2: 1kHz square wave response. A-2284-2

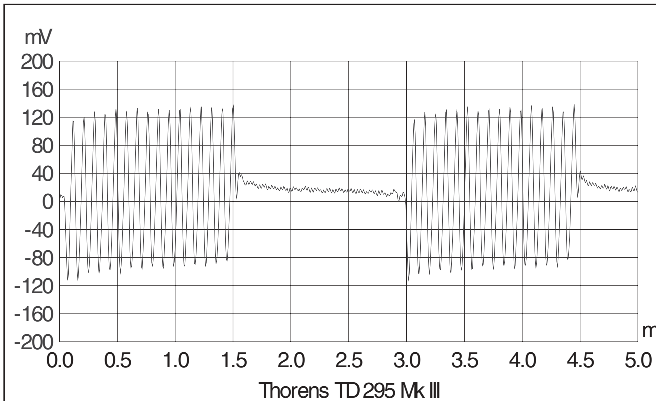

FIGURE 4: 10kHz tone burst response. A-2284-4

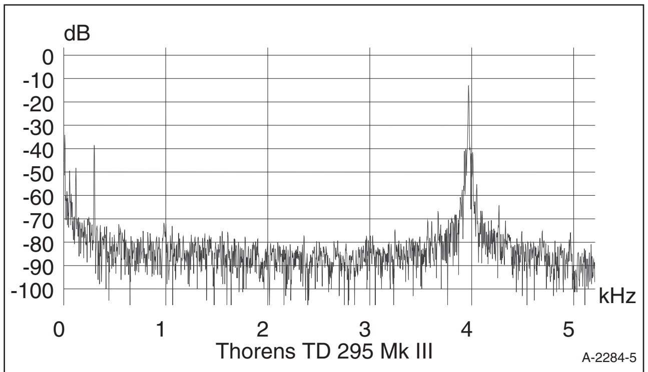

FIGURE 5: Spectrum of 300Hz + 4kHz intermodulation signal.

rotor pole. The motor reluctance torque also produces a second harmonic component in a synchronous motor stator at near full load.

When the speed select rotary switch is in the center stop position, the flux waveform appears to be 120Hz, perhaps indicating that a full-wave rectified "DC" voltage is applied to the stator windings to freeze the rotor position. The low motor frequencies allow the TD 295 motor pulley to be larger than a 60Hz line powered motor, providing more belt contact surface.

Figure 2 shows the cartridge response to a 1kHz square wave test track. The illustration in the test record booklet shows a slight tilt and cupping in the "ideal" as-recorded waveform, so it does not represent a perfect square wave shape. I have found that this rolled off wave-shape with a slight oscillation at the leading edges is the usual performance result in my testing.

TABLE 2
CHANNEL SEPARATION

WARBLE FREQUENCY	SEPARATION DB
6.4kHz–12.8kHz	–18.7
3.2kHz–6.4kHz	–23.5
1.3kHz–3.2kHz	–25.5
800Hz–1.2kHz	–27.9
400Hz–800Hz	–30.3

Figure 3 shows the response to a 1kHz tone burst. The OMB 10 achieved very accurate results here, with no evidence of response dips, resonance, or spurious response. *Figure 4* shows the exemplary response to the 10kHz tone burst track.

Figure 5 shows the Ortofon OMB 10 output spectrum reproducing a combined 300Hz + 4kHz intermodulation distortion (IMD) signal. The 4kHz signal is recorded at 7.5cm/s (–50dB rel), and the 300Hz signal is recorded at 9cm/s (–25dB rel). The 4.3kHz IMD product is –66dBr, and the 3.7kHz product is –67dBr, with no other spurious products evident. There are additional response peaks at 60Hz (–49dBr) and 120Hz (–48dBr).

I think these power line artifacts are due to residual hum pickup in my test setup, since I also noted them during the tests on the DacT CT 100 phono preamp. The CT 102 power supply is a high-frequency switcher, and does not bring the 60Hz line onto its power-supply board.

The manufacturer's specifications and measured results are shown in *Table 3*.

CONCLUSION

The TD 295 MK III is a substantial and stable record playing platform, with excellent speed performance and very good noise isolation. The Ortofon OMB 10 cartridge is a good match for the tonearm/table combination and offers fine performance. ❖

TABLE 3
MEASURED PERFORMANCE

PARAMETER	MANUFACTURER'S RATING	MEASURED RESULTS
Frequency Response:	20Hz–20kHz +3/–1dB	20Hz–20kHz +2.9/–1dB
Channel Separation:	22dB at 1kHz 15dB at 15kHz	18.4dB, 1.3kHz
Channel balance	1.5dB at 1kHz	1.1dB at 1kHz
Recommended Loading:	150–400pF, 47kΩ	
Compliance:	16µm/mN	
MM Output:	4mV at 5cm/s, 1kHz	4.0mV at 5cm/s, 1kHz
Wow and Flutter:	<0.045%, DIN 45507	
Rumble, unweighted:	>70dB, DIN 45-539-B	
Weighted:	>70dB, DIN 45-539-B	

XpressMail

CORRECTIONS

In my article "Considering Negative Feedback" (*aX*, Dec. 2003), I'd like to apologize and clear up some confusion, and also point out some errors:

1. On p. 24, the L and C scaling was omitted from Fig. 1. Following from the 100:1 impedance and the 1:10 frequency scaling, simulator capacitance = 0.1 times actual, simulator L = 1000 times actual. Now for the confusion: (a) op-amp feedback resistors and the 990W on IC1B don't follow the 100:1 R scaling (they don't matter; they just set the gain); (b) the simulated tube amp feedback resistors from S1 to the 616Ω R aren't scaled; they're approximately the tube amp values; the 616Ω R is roughly the cathode input R of the 12AX7; (c) C_T and L_T (transformer parasitics) are not only scaled as stated, but are also normalized to the secondary side.
2. Not shown on Figs. 2 and 3 are a 12AT7 gain stage and direct-coupled 6SN7 cathode follower. The amp's gains stated on p. 24 are from input (each balanced half) to transformer primary (each half), with 5Ω load on transformer secondary.

   Photo 3b caption should read "compared with 2b," not 2a.
3. On p. 29, center column, third line from the bottom, "out-of-phase feedback" was poorly worded; what I meant was "phase-shifted away from the desired 180° (negative) feedback." Likewise, near the top of the right column, "reasonably in phase" means "reasonably close to 180°."

Articles Worth Serious Consideration
The following articles should be required reading for anyone designing a tube amp, especially because many long-existing proven concepts are so often ignored or forgotten. I've used such concepts from all three articles, and they work very well.

1 "Odds and Evens" by Graham Dicker (*aX* July 2001), about using complementary transfer curvatures for linearizing cascaded stages. SET designers take note: you don't need to SETtle for 5-10% THD at full power!
2. "The Future of Vacuum Tubes in Audio" part 2 (*GA* 4/00) by Lynn Olson, regarding clean limiting.
3. "Designing Your Own Amp" part 5 (*GA* 6/00) by Norman Crowhurst. Fig. 1 shows transformer primary feedback. His whole series is of legendary excellence.

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

COMPONENT VALUES FOUND

I greatly enjoy your magazine! A recent article, "Hybrid Solid-State Power Amplifiers" by Charles Hansen, appears in the July 2003 issue. Four figures and four tables accompany the article. Tables 1, 2, and 4 all contain identical information. This is an obvious error, as the tables omit several component values in the accompanying figures. As I am interested in building one of the circuits, could you please provide the correct information? Thanks in advance!

Greg Bergmann
Columbus, OH

Charles Hansen responds:

Thanks for the letter from Greg Bergmann pointing out the errors in the tables for my article. All four tables have errors (my fault). Corrections on next page.

HOME THEATER UPDATE

Thanks to the American marketplace for video products, yesterday's high-end DVD players can now be obtained inexpensively and spare the user from applying any component upgrades. As an example, one recent evening my brother brought over a Pioneer Elite DV37, a $1000 player he found as a factory-refurbished item, for which he spent $160 plus tax and shipping, and which carried a new product guarantee.

My modified Sony uses an 8-bit video D/A converter running at 27MHz, while the DV37 uses a 10-bit converter at 54MHz. A comparison of familiar movies run on both players shows that intensity gradation in shaded objects is more accurate, and edges are slightly sharper, from the Pioneer. Colors are essentially identical.

These differences, although not large, are easily noticeable to anyone and genuinely come closer to the cinema sensation. And all without plugging in the soldering iron or applying more vibration control than four Sorbothane feet. The DV37's higher computational speed and numerical accuracy produce visibly more accurate wide-screen conversions for viewing on my Zenith's 4:3 screen.

Since superb picture quality is a given in players like this, it frees the videophile to vent his electronic curiosity on securing the best video from his TV monitor[1,2,3].

By having the chance to view this new player's graphic renderings, I can

TABLE 1
SIMULATION PART VALUES FOR FIG. 1

C1 – 1μF
C2, C6, C7 – 100pF
C3 – 10pF
C4 – 39pF
C5 – 47μF
R1 – 1k
R2, R8 – 47k
R3, R6 – 1k5
R4 – 22
R5 – 100
R7 – 2k2
R9, R10 – 47
R11, R12 – 0R33

TABLE 2
SIMULATION PART VALUES FOR FIG. 2

C1 – 1μF
C2, C6, C7 – 100pF
C3 – 10pF
C4 – 47μF
C5 – 39pF
R1 – 1k
R2, R8 –47k
R3, R6 – 510
R4, R5 – 15k
R7 – 2k2
R8 – 47k
R9, R10 – 330k
R11, R12 – 47
R13, R14 – 0R33

TABLE 3
SIMULATION PART VALUES FOR FIG. 3

C1 – 1μF
C2 – 100pF
C3 – 1nF
C4 – 47μF
C5 – 39pF
R1 – 1k
R2 – 47k
R3, R4 – Hundreds of ohms, depending on your selected output bias point
R5 – 1k5
R6 – 8k2
R7 – 2k2
R8 – 47k
R9, R10 – 100
R11, R12 – 10k
R13, R14 – 0R33
R15 – 220

TABLE 4
SIMULATION PART VALUES FOR FIG. 4

C1 – 1μF
C2, C6, C7 – 100pF
C3 – 10pF
C4 – 47μF
C5 – 39pF
R1 – 1k
R2 – 47k
R3, R6 – 1k5
R4 – 22
R5 – 100
R7 – 2k2
R8 – 47k
R9 not used
R10, R11 – 47
R12, R13 – 0R33

now say that all the effort spent on upgrading the monitor has been richly rewarded. I have seen new movies in theaters, and later purchased the DVD. The theater certainly has a bigger screen, but the colors on my CRT are better, and the brightness and contrast ratio are clearly superior.

DVD soundtracks heard through my stereo come from only 10′ away and produce abundant movie thrills, thanks to the electronic upgrades chronicled in the pages of this magazine. It's a great time for home audio/video.

REFERENCES

1. "A DVD Rescue," *audioXpress*, July 2002, pp. 24–27
2. "Detrashing Television," *audioXpress*, June 2003, p. 4
3. "Capacitor Violence for the Unintimidated," *audioXpress*, July 2003, pp. 67-68

Darcy Staggs
Orange, Calif.

SAFETY NOTICE

There are about six Tube Analyzers in the field that were supplied with metal base Svetlana 6550 tubes. The metal base is connected to pin 1, which in turn is connected to pin 3—the plate. The metal base is therefore hot with high voltage on it.

Anyone with a Tube Analyzer with a metal base 6550 should wrap electrical tape around the base or contact me so I can replace the tube with a plastic base unit.

Bruce Rozenblit
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bruce@transcendentsound.com

TUBE CHOICE

I thank Mr. Scott Reynolds very much for "A Simple, High Quality AM Tube Receiver" (Aug. 03 *aX*, p. 6). However, V1, the 6BA6 does not seem to have nearly as much gain as 6DC6 or 6BZ6 valves, for example. The Gm of the 6BA6 is 4,400, while that of the 6BZ6 is 6,100. Gain in the first RF stage is particularly important for listeners in fringe areas such as ours, where most medium-wave stations are 100km, or more, away and low-powered (5kW or less).

According to the Engineering Report published by the Collins Radio Company on the classic R-390A Communications Receiver, the 6DC6 and 6BZ6 valves were the two valves that they found had the best characteristics of low noise and low cross-modulation. The 6DC6 valve was chosen because it afforded better A.G.C. control—perhaps not an issue in this design.

The only disadvantage would be the fact that the 6DC6 costs considerably more than the 6BA6.

Terry Robinson
Victoria, Australia

Scott K. Reynolds responds:

Good idea! I originally chose the 6BA6 for V1 in my receiver because it is the classic tube for such circuits and I had a number on hand. But your suggestion of using the 6DC6 or 6BZ6 should give a bit more RF gain. I've also experimented with the 6EH7, a semi-remote-cutoff pentode with a Gm of 12500µS. The higher Gm gives 8–9dB more gain, which can indeed help on weak stations.

The values for R1 (180Ω) and R2 (33kΩ) given in my article for the 6BA6 seem to work OK for the 6EH7 as well, but unfortunately the 6EH7 requires a 9-pin socket instead of the 7-pin used for the 6BA6, so it is difficult to substitute one for the other. I wired a 9-pin socket to a 7-pin plug so that I could swap them. But if you do this, be sure to keep the wires as short as possible and keep the plate lead away from the grid lead.

The 6DC6 and 6BZ6 are more convenient substitutions for V1 since both use a 7-pin socket, and although the cathode and suppressor pins on both tubes are swapped relative to the 6BA6, you can simply plug them into the 6BA6 socket on my receiver, since I connected the cathode and suppressor together. But based on their published Gm, the 6DC6 and 6BZ6 would appear to offer only 2–3dB more gain (I have not tried them). Of course, that is still worthwhile.

If you substitute the 6EH7 for the 6BA6, you will almost certainly need to add the RF gain control mentioned in the article. The 6EH7, with its higher gain, is not as tolerant of strong signals.

I'm not familiar with the report by the Collins Radio Company that you mention. The 6EH7 was developed for use in the intermediate-frequency (IF) amplifier stages of color television receivers and may not have been available when Collins Radio did the study.

Thanks for your letter. I hope that readers will feel free to experiment with my radio in this and other ways (although beginners might wish to build it as described in the article and get it working first).

AMP TESTER

I am very pleased to see your reactive load tester (Aug. '03 *aX*, p. 48). Since I've not yet done such testing, your results of nearly doubled THD w/capacitive load confirms for me the need for this testing. I have three comments/ questions:

1. I'm very impressed with your achievement of nearly constant phase angle over a wide bandwidth. I'm curious as to how you designed the elegantly simple networks.
2. You mentioned a 2.83V level (2W @ 4Ω). Was this a tube or solid-state amp? (The latter is likely to have more distortion at low power.) I'm now building a 300W stereo tube amplifier, so I'm curious about what kind of amplifiers you've reactively tested. Also, if you've taken advantage of your tester's 100W capability.
3. I'd be very interested to see how the distortion spectrum changes with reactive load. If reactance causes extension of distortion order, for example, then of course there's even more concern.

Congratulations on your excellent design. I must build one to test my amplifier with, since it saves me from needing the "closet full of parts" you referred to, and that I had thought I needed!

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

Dick Crawford responds:

Thanks for your interest in my article "Build a Reactive Load." In answer to your queries:

1. *I designed the reactive networks in two steps: first, the schematic using an analysis of the requirement of constant phase angle over a wide range of frequencies; second, the values of the components in the schematic by simulation using a program called "B squared Spice." I am a big fan of simulation, because it allows me to quickly try different values for the components in a circuit, and thus choose the*

values that give desired characteristics with real-world values.

2. I used a solid-state amplifier for my measurements at 2.83V. I've briefly tested the inductive load at 100W at 20Hz. The reactive load box became hot, particularly the 4Ω load resistor, but seemed to behave properly. At 20Hz I measured little difference between the inductive load and the resistive load.
3. As I mentioned in the article, THD measurements with the inductive load at 1kHz were just slightly worse than with a resistive load, and the THD spectrums did not seem very different. However, with a capacitive load the THD measured about twice as much as with a resistive load. Furthermore, the THD spectrum did change with a capacitive load, with the higher order distortion products increasing.

HUM HUNT

I read with interest the article on upgrading the Dynaco FM-5 power supplies in the August 2003 issue (p. 28). A few months ago I decided to try to solve a 60Hz hum problem that my FM-5 exhibited. One channel was worse than the other. I originally attributed the hum to power-supply ripple and added considerable capacitance to the power supplies to reduce it. I did not record the ripple improvement but it was nearly unmeasurable when I was done. Still, the hum persisted.

I reasoned that the hum was being induced by radiation from the transformer and began my search for the point of entry. My first thought was that the entry point was the fairly long wiring path to and from the volume control. So I removed the volume control from the circuit and installed direct jumpers from pads 18 and 19 to pads 17 and 16, respectively, on PC26. There was no improvement.

I next loosened the mounting screws for the power-supply transformer so that I could alter its orientation to the circuit boards. I found that various orientations of the transformer reduced or increased the hum, and it was not necessarily the same change on both channels. By moving the transformer closely along PC26, I was able to find the points of injection. They were the two 19kHz filters, FL53 and FL54, which are potted modules, but I surmise that they have an unshielded inductor which picks up the stray magnetic field from the transformer.

I was unable to find any orientation that completely cured the hum in both channels, so I took an extreme action to remove the hum field. I moved the transformer outside of the chassis. To avoid making a new hole in the back of the chassis, I removed the unused, unswitched AC outlet. I added about 3′ of wire to the transformer leads, so I could place the transformer well away from the chassis. This cured the problem completely.

The FM-5 is now free of any audible hum even at high volumes. This is an essentially free modification that made a very audible improvement in performance.

Greg Benecke
Torrance, Calif.

Daniel Dufresne responds:

I have not experienced any noticeable hum with my Dynaco FM5 tuner in my systems. You did not specify whether the volume control affected the hum level. I made the following hum measurements and obtained *these results: with the volume control at minimum, output is 128µV, RMS, 22.4Hz to 22.4kHz, and 24µV, RMS "A" weighted.*

With the volume control at maximum and a 400Hz signal modulating a 1mV 100MHz carrier with 75kHz deviation, the output is 2.5V. The noise without modulation is 2mV, giving a 62dB signal to noise ratio. The published test reports give about 65dB signal to noise.

The results of this test depend somewhat on the signal generator used; mine is a Hewlett-Packard 8640A. However, you did succeed in finding a solution to your hum problem. Maybe others can benefit from your experience.

A POOGE FOR POOGE

I received a nice e-mail from Walt Jung concerning the filter design for "A Digital Pooge for HP's 339A Test Set" (March '03, p. 16). He came up with the following suggestions, and I made some THD+N tests, described in detail later.

The extremely small 10pF filter cap C4 used in the U1C stage of the filter should really be larger, maybe about 10× (100pF). This would give you much better predictability overall. You can, of course, easily scale any of these Sallen-Key circuits by multiplying the caps and dividing the resistors by the same factor. For instance, in Figure 2, change the U1C stage values to: R4 and R5 to 8k66, C3 to 8n2, C4 to 100pF. You can do the same scaling for the U1D and U1B stages as well, if you like.

Raising the caps and lowering the resistors will also suppress whatever amount of nonlinear input capacitance distortion is inherent in the FET input op amp. Walt thought that the THD+N of this filter could be improved some by doing this.

Also, note that the forward path of the LT1058 FET op amp is asymmetric in gain characteristics, and has unequal ±slew rates. This will show up as rising second harmonic distortion with a fixed level and increasing frequency. For general filter design, a cleaner and simpler topology is used in the LF351/LF411, TL071, or AD711. (Note: the TL074 is a quad of the TL071, and the AD713 is a quad of the AD711. Both are direct plug-in replacements for the LT1058.)

I noted to Walt that since the HP 339A filter runs at a low 10mV, slewing issues would seem to be a rather remote potential (the LM348 used in the standard filters only slews at 0.5V/µs versus minimum slew rates of 8V/µs for the LT1058 and 5V/µs for the TL074). So noise is the paramount issue, both for the op amps and the passive component values. To put this in perspective, note that the bipolar OP27 voltage noise of ~3nV/√Hz is about on a par with a single 1k resistor (4nV/√Hz). So, you will need to work carefully to reduce the R values.

The bipolar devices don't do as well, apparently because their higher input current noise interacts with the higher Z of the filter. Lower impedances will help that.

But all of these FET input op amps wouldn't help noise at very low levels, since their own noise voltage is many times that of the OP27. The likely best choice would be the AD743/AD745, which has both very low current noise *and* voltage noise: 6.9fA/√Hz and 3.2nV/√Hz, respectively (both at 1kHz).

The unity-gain-stable AD743 has a slew rate of 2.8V/µs, while the sister part AD745, stable at a gain-of-five, has a slew rate of 12.5V/µs. The AD743 is available as an eight-pin DIP, or 16-pin SOIC, while the AD745 is available only as a 16-pin SOIC (www.analog.com).

Walt then suggested that it might be possible to eliminate an op amp or two, and thus lower noise even further. For example, the subsequent stage in the 339A after C5 is the non-inverting buffer for the RMS-DC converter, with a Z of 1M, so the U1B stage can be eliminated by going to much lower values for R6 and C5, for example, 1k00 and 22nF.

Likewise, I installed R1 as a safety precaution in case of a wiring mistake on my added board, so R1 and the U1A buffer might also be eliminated, helping noise. The filters are driven from either an LM310 (the voltmeter function) or an LM318A op amp (the distortion test function), so the U1B buffering is not a functional requirement. Walt's suggestions will get you down to a dual or two single op amps without altering the filter response. The reader will need to verify this performance.

I ran some THD+N tests with various op amps on a quasi-third-order 27kHz LP filter purpose-built for a different application (*Fig. 1*). The power source is ±9V batteries. At low levels noise dominates the THD+N readings.

Beyond the minimum point, the rising curve indicates that distortion products dominate the THD+N. The results at 1kHz are shown in *Fig. 2*. The 27kHz results are in *Fig. 3*. The noise, GBW, and slew rate specs for the chosen op amps are in *Table 1*. Note these are the worst-case

TABLE 1
SELECTED OP AMP PARAMETERS

DEVICE	VOLTAGE NOISE	CURRENT NOISE	GAIN-BW	SLEW RATE
OP27	3.8 nV/√Hz	0.60 pA/√Hz	5.0 MHz	1.7 V/µs
TL071	18 nV/√Hz	0.01 pA/√Hz	3.0 MHz	8.0 V/µs
LF351	25 nV/√Hz	0.01 pA/√Hz	4.0 MHz	13 V/µs
LF411	25 nV/√Hz	0.01 pA/√Hz	3.0 MHz	10 V/µs
AD711	18 nV/√Hz	0.01 pA/√Hz	3.4 MHz	18 V/µs
LM741	30 nV/√Hz	3.0 pA/√Hz	0.4 MHz	0.3 V/µs
AD743	5.0 nV/√Hz	0.0069 pA/√Hz	4.5 MHz	2.8 V/µs
LT1056	22 nV/√Hz	0.004 pA/√Hz	5.5 MHz	9.0 V/µs
LT1792	6.0 nV/√Hz	0.01 pA/√Hz	4.0 MHz	2.3 V/µs

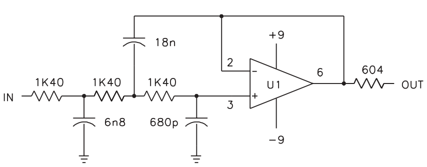


FIGURE 1: Revised filter design.

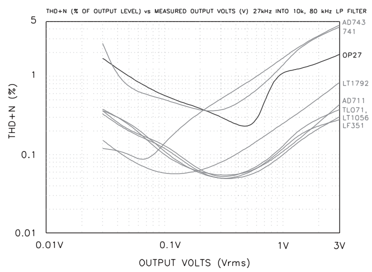


FIGURE 2: 1kHz THD+N results.

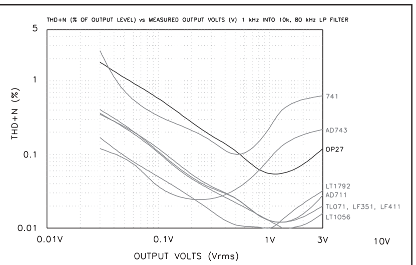


FIGURE 3: 27kHz test results.

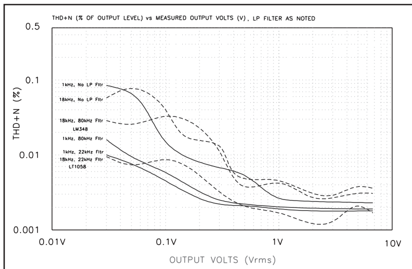


FIGURE 4: 22kHz results.

numbers, or typical values if the worst case isn't listed. The 741 is listed because the LM348 is a quad version of the 741.

Finally, tests on the LT1058 fifth-order 22kHz LP filter in the article, performed by feeding the output of the HP339A oscillator into its distortion test input, are shown in *Fig. 4*. The only changes from the circuit in the article are in U1C: R4 and R5 to 8k66, C3 to 8n2 MPP, and C4 to 100pF NP0. The tests were run at 1kHz (solid lines) and at 18kHz (dashed lines), which is the frequency at which the 22kHz response is still flat and filter distortion is highest. I ran the tests with the modified 22kHz LP filter (LT1058 op amp), with the standard 80kHz LP filter (LM348 op amp), and then with all filters disengaged.

Chuck Hansen
Ocean, N.J. ❖

Book Review

Rider: Inside the Vacuum Tube

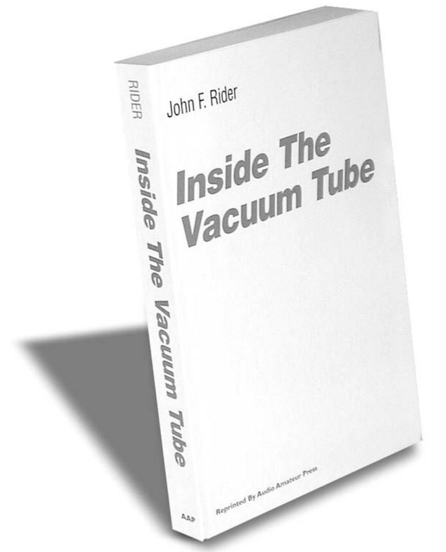
Reviewed by Scott Frankland

John F. Rider was a natural teacher who rose to become one of the biggest names in electronics publishing during the golden age of the vacuum tube. Rider's stature in the publishing world ranks alongside Hugo Gernsback and Howard W. Sams. All three managed, in their own way, to be at once technical yet popular.

Rider, in particular, has a trademark down-to-earth teaching style that at first glance seems a kind of "electronics for dummies." What he does, rather, is to make electronics accessible—and even fun. The average engineering text, in contrast, is dry as dust. There are exceptions, but amusement is infrequent as a rule[1].

Typically, Rider aims to portray the action of electrons within tubes and circuits. In so doing, he is not averse to using whimsy to flag our attention (*Fig. 1*). Rider wants his readers to clearly *visualize* the flow and dynamics of electrons. Only then does he try to chart these effects within tubes. This parallels what more formal authors such as Jacob Millman teach[2]: that the first step is to grasp the physical process involved, and only then to try to formalize it quantitatively.

Rider's approach tries to instill in the reader an intuitive feel for invisible processes by means of creative graphics—an approach that has evolved into modern computer animation[3]. In 1945, Rider even went so far as to provide 3D illustrations in the original edi-



tion of his book (with 3D glasses included). Unfortunately, these illustrations could not be reproduced in this new edition by Audio Amateur Press[4] (available from Old Colony Sound Lab at 1-888-924-9465 or from www.audioXpress.com). Nonetheless, the bulk of the illustrations remain, and there is much that remains to conjure in this thorough-going handbook of vacuum-tube fundamentals.

THE BOOK

Like few other books of its kind, *Inside the Vacuum Tube* excels at the graphical approach to vacuum tubes[5]. The major aspects of tube action are portrayed in just about every way possible. Following a thorough, if somewhat droll, introduction to basic tube physics, Rider explains how to plot mutual conductance (g_m) graphically. This is the tube factor used by "tube checker" machines to indicate the strength of a tube.

Rider shows how to use the grid family of curves as well as the plate family of curves to obtain g_m. He does the same for the other two tube factors—amplification factor (μ) and plate resistance (r_p). He then shows the interrelationships—and interdependence—of the three factors. This thoroughgoing approach ties together the basic elements of static tube action and shows why the dynamic transfer curve is the leading indicator of tube performance.

After this groundwork, Rider deals with voltage amplification in all its fundamentals. This makes up the central core of the book. Rider's expansive treatment fills a welcome niche, as voltage amplifiers are used throughout audio electronics: in line stages, phono stages, amplifier front-ends, DAC back-ends—virtually everywhere amplification is needed.

Rider shows how to determine voltage amplification for both triodes and pentodes step-by-step. Math is used sparingly, and is kept simple. Recognizing that equations are necessary for calculations of circuit values, Rider uses charts to explain the math, and vice versa. In nearly every case where an equation appears, it's accompanied by a corresponding graph, along with a circuit diagram. At each step, Rider methodically describes in plain English what occurs in the charts.

Rider's emphasis is on constructing a dynamic transfer curve for any given load line, which shows at a glance the linearity of the stage. This allows you to find the optimum operating point. Rider's extensive treatment of the transfer curve allows you to see graphically how different values of load resistance affect linearity.

Figures 9–11 on page 214 of the book show the transfer curves for five different plate loads, and may be considered

(to page 72)

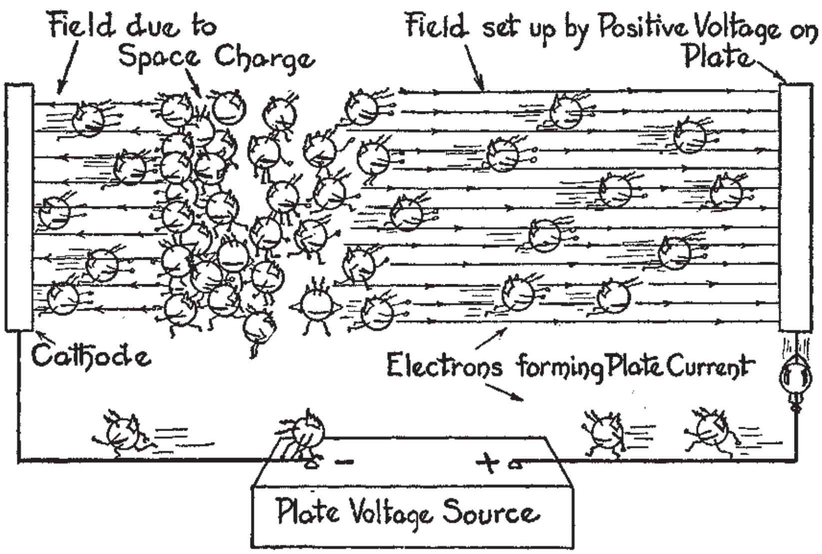
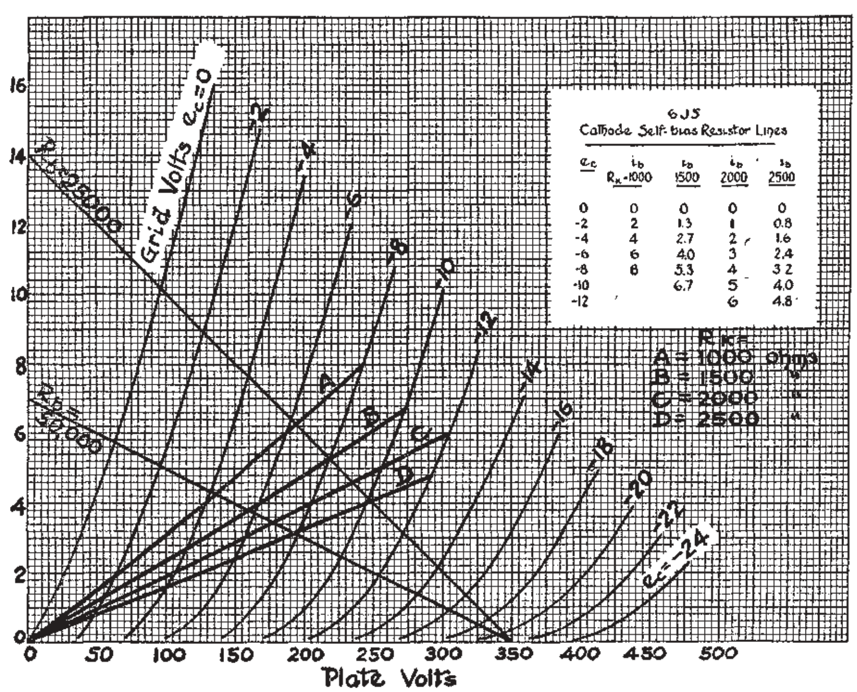


FIGURE 1: Electrons at work in Rider's book. G-2219-1



e_c	i_b R_k=1000	i_b 1500	i_b 2000	i_b 2500
0	0	0	0	0
-2	2	1.3	1	0.8
-4	4	2.7	2	1.6
-6	6	4.0	3	2.4
-8	8	5.3	4	3.2
-10		6.7	5	4.0
-12			6	4.8

FIGURE 2: Sample grid from *Inside the Vacuum Tube.* G-2219-2

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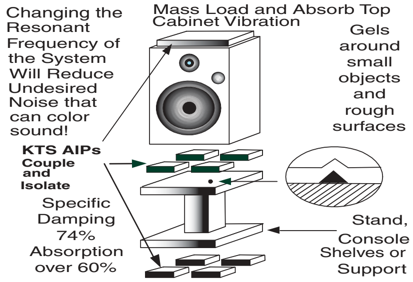

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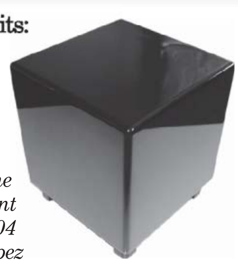

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Inside the Vacuum Tube
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the hub of Rider's narrative. For the more advanced reader, this is a good place to start. Other considerations relevant to the selection of the optimum plate load and operating point are summarized on pages 268–269.

At each step, Rider is less interested in expounding theory than he is in delineating the actual dynamics involved. His approach cuts through the often excessive formality of college-level electronics texts to get at the essential nuts-and-bolts of what you want to accomplish in a working circuit. Rider's presentation, however, stops short of multistage circuit design. He concentrates instead on the proper construction of a single gain stage.

Besides the emphasis on tube physics and voltage amplification, one of the most valuable aspects of this book is the chapter on cathode circuits. Virtually every aspect of the cathode circuit is discussed: including the effect of the cathode resistor on gain, output impedance, and linearity. In particular, the effect of the cathode resistor on bias is thoroughly explained (*Fig. 2*). All of which leads naturally to the cathode follower. Rider's thoroughness and clarity in explaining this relatively abstruse device is exemplary, approaching that of John D. Ryder's classic presentation[6,1].

In addition, there's a chapter on power amplifiers, although it's short and cuts to the chase rather quickly. Nonetheless, Rider gives an uncommonly thorough explanation of dissipation in power tubes. This is another area where he rivals John D. Ryder in the clarity of his exposition. The predominant strength of *Inside the Vacuum Tube*, however, remains the voltage amplifier.

SUMMING UP

When I first began my study of the voltage amplifier as a circuit element some 25 years ago, I found most texts that treated this subject either too formal or too simple. One exception is Daley's *Principles of Electronics*[7]. But I still needed to read several other books to shore up the gaps in Daley's exposition.

By then I had found an original copy of Rider's book; but before I could finish it, somebody borrowed it and that was the last I saw of it! I never could find another copy—until now. Rider's book is so clear and thorough, and its self-teaching quotient so high, that it might be all you need to master this central aspect of tube-circuit design.

If your aim is to better understand the basic principles of vacuum tubes—their characteristics, load lines, operating points, and above all, *voltage amplification*—then Rider's book is among the most accessible you can find. Outward appearances notwithstanding, *Inside the Vacuum Tube* provides the most thorough yet simple exposition of voltage amplification that I've seen. Its entertaining yet methodical teaching style, while not the shortest route to finding quick answers, is nonetheless the surest route to a fuller understanding. (*The Radiotron Designer's Handbook*[8] is still probably the quickest way to find solutions to specific problems in tube-circuit design. This book is long out-of-print, but is available on CD from Old Colony. It is, however, far more advanced mathematically than *Inside the Vacuum Tube*.) ❖

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4. John F. Rider, *Inside the Vacuum Tube*, © 1945, reprinted by Audio Amateur Press, Peterborough, NH, 2002, 407 pp., $29.95.
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7. J.L. Daley, Ed., *Principles of Electronics and Electronic Systems*, United States Naval Institute, 1956.
8. F. Langford-Smith, ed., *Radiotron Designer's Handbook*, Amalgamated Wireless Valve Co., Sydney, Australia, 4th ed., 1953, reprinted on CD by Audio Amateur Press, Peterborough, NH, 1996, $29.95.