



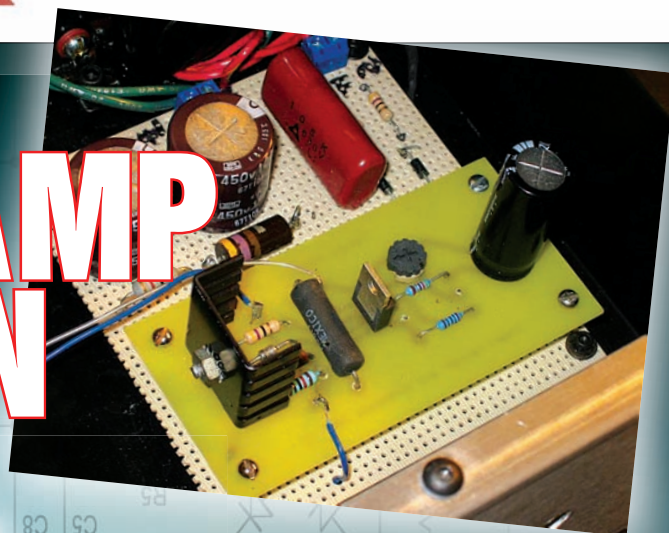
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TUBE PREAMP DESIGN

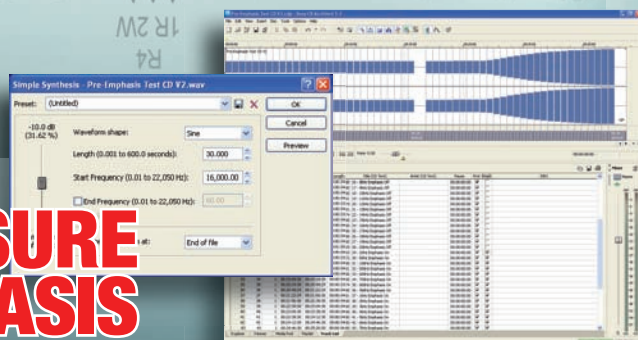


SHOWCASING A 20W PUSH-PULL AMP

- Circuit Simulation Tips From a Pro
- Noise Measurements Of Linear Systems' JFET
- Testing a Power Supply Design



CD TO MEASURE DE-EMPHASIS



The ImPasse Preamplifier

This author unveils his design of a tube preamp to drive the Pass F4 and other low-to-unity-gain power amplifiers.

It all started with the Burning Amp Festival (BAF) '07 in San Francisco. My friend was building a balanced version of Nelson Pass's First Watt F4 and asked for a recommendation on how to drive it. The Pass F4 is a unity gain power buffer, meant to take a high-level output from a preamp and drive speakers with it—an interesting concept. The schematic suggests that the F4 can swing about 20V peak into an 8Ω load single-ended, but 40V peak in balanced mono.

A diagram of the balanced mono connection for the F4 is shown in **Fig. 1**. A preamp with balanced outputs is used for all of the voltage gain and to drive the cables to the power amps, along with the power amp input impedance. The F4s buffer the balanced preamp outputs and provide a balanced signal to the speakers. Because a unity gain power stage can be designed to run open loop with quite low distortion, the performance of the electronic path rests almost entirely on the balanced-output preamp, which ought to be low distortion and low source impedance to reduce the criticality of interconnects.

So rather than recommend a preamp for the task, I took the totally impractical step (I don't own an F4!) of designing and building one for his demonstration at BAF '07. And, of course, it needed to be called The ImPasse.

THE ROAD NOT TAKEN

I will first examine the design requirements. In the balanced mono arrangement, your preamp must be able to swing 20V peak on each of its output

polarities, and do that driven from a typical line-level source (2V RMS or 2.8V peak). To accommodate CDs and records that are cut at a lower-than-normal level, you should build in some extra gain, arbitrarily and conveniently arriving at a target gain of 26dB ($A = 20$). In the balanced mono connection, the power amplifier input impedances are identical and symmetrical. Thanks to the FET input, their values are also somewhat arbitrary—the intrinsically low gate current of a FET means that voltage offsets due to gate current can be

very low even with relatively large gate resistors. This relieves the preamp from the need to deliver much signal current; it just needs to swing many volts very cleanly.

For me at least, the requirement to swing a lot of volts at low current with a simple circuit suggests... tubes. And what's even more compelling is that tube stages that give balanced outputs at high voltage levels are already well-known and characterized—the input and driver stages of a push-pull power amplifier fit the ticket perfectly (I'll return to this

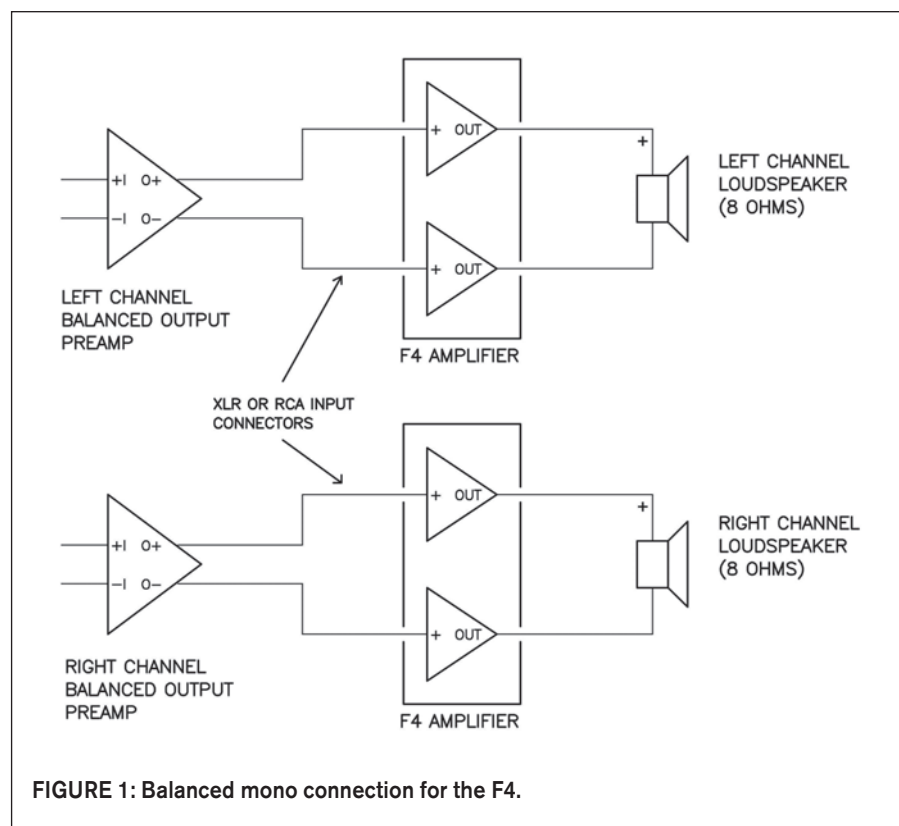


FIGURE 1: Balanced mono connection for the F4.

point at the end).

Given the symmetrical loads and the requirements for balance, you have several options. The first that comes to mind is the classic differential amp (**Fig. 2**). With a constant-current sink in the tail and load that is equal on both polarities, the balance is excellent. Likewise, the output impedance, Z_{out} , can be made relatively low; with symmetrical loads, it is roughly equal to the plate resistance in parallel with the load resistance. Power supply rejection is excellent.

To get to the target gain of 20, you need a tube with a gain somewhat larger than 40; from a single-ended source, gain to each plate of a differential amplifier is half the gain that the same circuit would have in common cathode. Restricting yourself to easily available tubes, you could look at 12AT7/ECC81, 6SL7, or 12AX7/ECC83. Being more exotic, you could check some of the high gm European pentodes, triode connected, in order to achieve the requisite gain with a low plate resistance.

Consider the common tubes. Of the three candidates, the 12AT7/ECC81 has the lowest plate resistance; at 10-15k for normal operating currents, the preamp's output impedance will be 20-30k, which is somewhat higher than is comfortable. Any cable capacitance will be liable to cause HF losses. The other tubes produce a much worse problem. It

seems that you'll need to follow the differential amplifiers with a pair of buffers, probably cathode followers, so a well-designed preamp based on that topology will have four tubes at a minimum.

You can do somewhat better with the exotics. A D3a, triode connected, will have a plate resistance of slightly greater than 2k at 20mA operating current. This will result in an output impedance of about 4k5, considerably better but still quite high. But with care and attention paid to the cables and amplifier load, this could be a viable option, albeit at some expense—D3a is not cheap, and the current to run two channels of preamp will be 80mA or more.

'TIS A GIFT TO BE SIMPLE

Turning to a different topological option well-trod by power amplifier designers, consider a simple common-cathode voltage amplifier feeding a split-load inverter (**Fig. 3**). You can couple the two stages in various ways: direct coupling, RC coupling, or a combination of the two are all simple and effective.

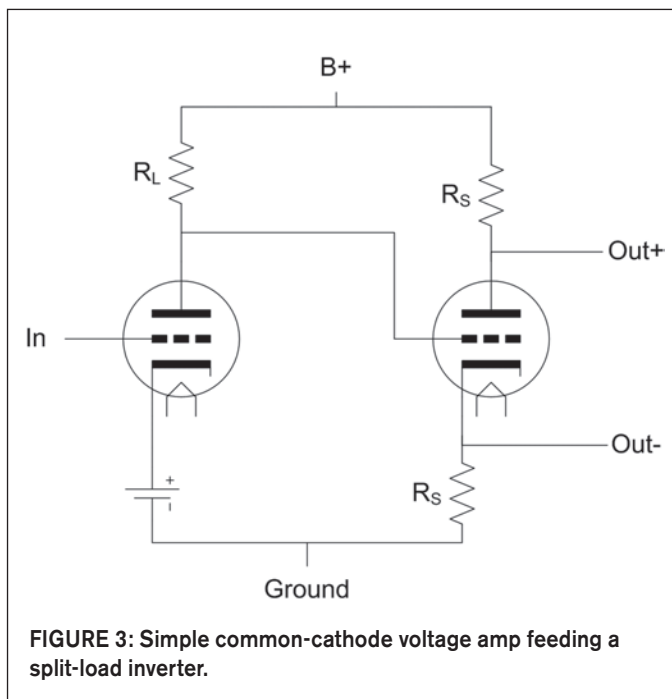
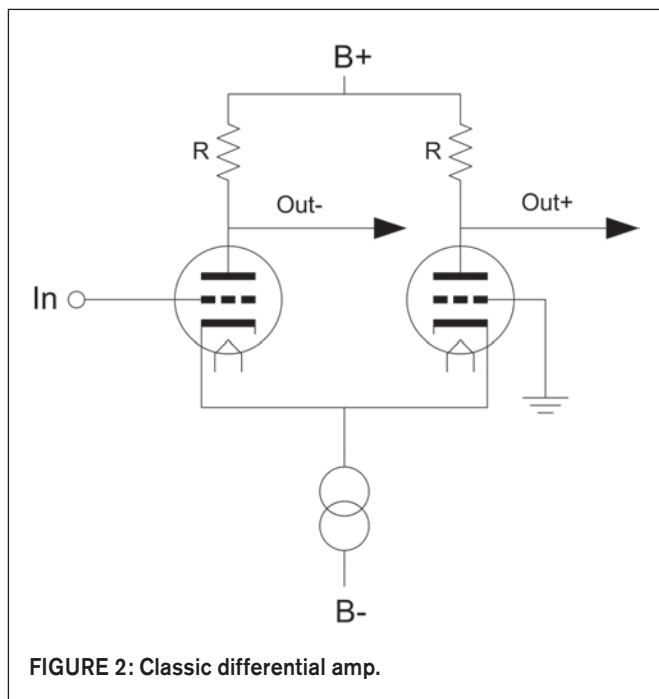
Contrary to myth, the split-load inverter has symmetrical source impedance from both the plate and cathode sides as long as the load is also symmetrical. That source impedance is lower than that of a comparable diff amp, being approximately 1/gm (similar to a simple cathode follower). For a common tube

such as ECC88, for example, Z_{out} will be approximately 100-200 Ω , a much more attractive figure than even the exotic differential amplifier.

Better yet, the overall gain of the combination of grounded cathode voltage amplifier and split-load inverter will have double the gain of the same voltage amplifier tube in differential mode. This extends the range of tubes that you can use for the first (voltage amplifier) stage to include medium-mu triodes such as 6SN7.

Medium-mu triodes also have the advantage of potentially lower input capacitance due to reduced Miller effect. If you use a high mu triode such as a D3a as an input tube, the input capacitance for a single-ended source will be nearly 300pF. This might upset some driving sources, but these days that would be a minority. Still, it is something to consider, especially when choosing the value of the volume control—assuming a low source impedance driving the volume control, the worst-case source impedance is at the halfway (-6dB) setting and is about half the volume control total resistance.

This impedance, combined with the tube's input capacitance, forms a pole, which will often impinge on the audio band. A 10k potentiometer combined with a D3a input will give a -3dB point of about 100k, which is a satisfactory



ry bandwidth, but higher values could begin to be problematic. Gain is also significantly higher than target, so some careful reduction measures would need to be designed in.

Fortunately, a very common medium-mu triode, the 6SN7, has demonstrably excellent linearity at these voltage swings—the requisite gain, sufficiently low input capacitance (80pF)—and is widely and easily available in an impressive variety of flavors. Distortion performance is comparable to the exotics, and input capacitance is significantly lower. Among common medium-mu tubes, it is the most linear in voltage amplifier service, and the μ of 20 is spot-on for this application.

FINAL ANSWER

So, given the design choices here, I've gone conventional and will use a 6SN7 as a single-ended common-cathode voltage amplifier, and an ECC88 as a split-load phase splitter/buffer. It's cheap, easy, and as good/better performance than the exotics in differential mode. And a whole two-channel preamp can be done with two tubes. The fact that I had a

single CV1988 (a rare British military version of the 6SN7) begging to be used did not influence my choice of tube here, no sir!

I'll now move to the step-by-step practicalities of the design. The input stage provides all of the voltage amplification and is hence the most critical regarding distortion and available voltage swing. It needs to swing slightly more than the desired 20V peak output because the phase splitter will have slightly under unity gain. It will be lightly loaded—a split-load inverter has the same low input capacitance as a cathode follower, so can have a high input impedance.

A 6SN7 in grounded-cathode mode with an 8mA constant-current source plate load will show less than 0.04% THD at 20V peak; selected samples will show even less. This distortion is overwhelmingly second harmonic. At 2.83V RMS out, the THD is better than 0.01%, again predominantly second-order. Not bad for a single device, open loop!

Biasing in common-cathode voltage amplifiers typically involves a series re-

sistor in the cathode circuit, with the cathode bypassed to ground. A high-quality bypass cap is bulky and expensive, so an interesting and very viable alternative is the use of forward-biased diodes for developing a constant voltage at the cathode. LEDs are particularly suitable because of their low source impedance, low noise, and reasonably high forward voltage drop. A cheap surplus red LED will typically show 1.7V drop with an AC impedance of 5Ω or less at 10mA current. And it has the side benefit of visually indicating that the tube is indeed drawing current.

Looking at the characteristic curves for the 6SN7, you see that for 8mA of plate current and 160V on the plate, the grid must be about -3.5V with respect to the cathode. Or the cathode can be 3.5V positive with respect to the grid. So two red 1.7V LEDs in series between cathode and ground will automatically provide about the right bias while maintaining a low (10R) AC impedance. With constant-current loading of the tube, the current through the LEDs will also be constant (because plate current equals cathode current), so any nonlinearity in

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the LEDs' impedance with current is suppressed.

Another bonus of CCS loading of the plate and LED biasing of the cathode is the greatly increased power supply rejection. Any noise on the B+ rail is divided down at the output of the 6SN7 by about the ratio of plate resistance to CCS source impedance. The low impedance of the LED and the high impedance of the CCS mean that this ratio will be very small. For example, with a 7500Ω plate resistance and a 1000M CCS output impedance, any power supply noise or instability is reduced by over 100dB. That's a very nice way of removing power supply criticality. The outline of the voltage gain stage is shown in Fig. 4.

One notable feature is the use of an input transformer. In an earlier article, I justified its use in some detail; basically, it provides outstanding common-mode hum immunity and galvanic isolation, while at the same time facilitating either balanced or single-ended drive. A 1:1 unit is appropriate here. The gold standard is the Jensen JT11P-1, which I used in the prototype, but other units from Cinemag, Sowter, and Edcor should work equally well. All of these have very low distortion and excellent bandwidth.

CONSTANT CRAVING

A very good choice for a constant-current source (CCS) is a cascode, and among cascodes, one of the best performers is a FET cascode. Finding JFETs that will withstand the voltages and currents involved is not trivial. High voltage MOSFETs are plentiful and, in cascode, can make an excellent CCS. Unfortunately, the biasing arrangements necessary for enhancement mode devices can complicate things.

There is relief at hand—depletion-mode MOSFETs are becoming easier to get, and the DN2540 is cheap, available, and works perfectly in this service. A simple yet high-performance CCS using these wonderful devices is shown in Fig. 5. This two-terminal, self-biasing cascode CCS with a source impedance in the 1000M range can withstand

400V without breaking a sweat.

To understand the operation of this cascode CCS, first consider Q1. Its current is set by the value of R5. For a fixed current, the voltage drop across R5 is fixed. Because to first-order the gate current is zero, the drop across R6 (the gate stopper, put there to prevent oscillation) is zero, so Q1's gate-to-source voltage is fixed at I_d times R5.

Considering those flat I_d versus V_d FET characteristic curves, that means that the current through the FET is largely independent of the drain-source voltage. If the current tended to rise, the voltage across R5 would rise, which would tend to drive the FET toward lower current. The opposite argument can be made for the FET tending toward lower current. So, because of that feedback effect, I_d is constant.

The second FET has a more subtle role. As you modulate signal across the CCS, the drain-source voltage varies. Though the curves are quite flat (i.e., the output resistance is very high), they are not perfectly so. Additionally, the drain-source capacitance is significant, especially at low voltages, limiting the source impedance at high frequencies. Worse yet, at lower drain-source voltages, the capacitance is very nonlinear.

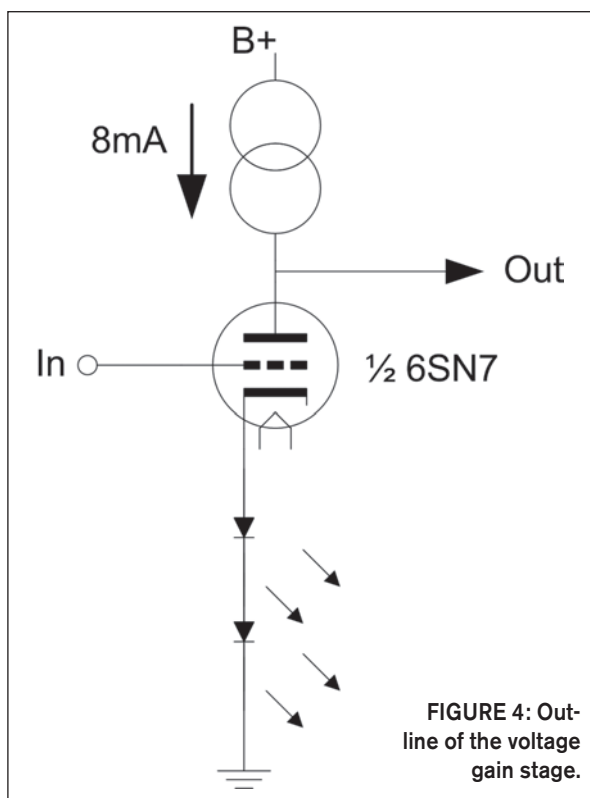
Ideally, you would hold V_{ds} constant. And that is the function of Q2, which is connected as a source follower with the gate driven by the source of Q1. R7 is a gate-stopper, as before, with no significant voltage drop across it. Q2's source follows the signal, so that the voltage between Q1's drain and source is held constant. Thus, the drain-source capacitance is not modulated, and the constant V_{ds} across Q1 means that current is held more constant—equivalently, you can say that the output impedance is higher.

The complete schematic is shown in Fig. 6. You can now start assigning component values. The gate-stoppers' value is largely non-critical—a few hundred ohms will do. Because I had a bag of 300R carbon resistors, that was my chosen value. If you use 330R or 270R or even 470R, the operation of the CCS will not be affected; it is important to have the body of the resistor as close to the gate as possible. The value of R5 is more critical—that's what sets the CCS's current. And the CCS sets the operating current of the voltage amplifier tube.

FETs vary. A lot. It is not unusual to find two devices with the same part number having I_{dss} or g_m values that differ by a factor of three. I bought several tubes of the DN2540 and kept one.

Those parts were all relatively similar, with two or three outliers from the batch of 100 tested. But I cannot guarantee that other lots of these parts will be so closely controlled. So, although I give the value of R5 that worked for this particular lot, you will need to adjust its value for the devices you have to get the desired operating current.

To test the operating current and check whether the initial value of R5 must be raised or lowered, you can use the circuit in Fig. 7. This jig is powered by two standard 9V batteries in series; of course, you can use any other convenient power supply. The current is calculated from the measured voltage across R_{test} divided by R_{test} 's value. A convenient value is 1k, which will give an 8V drop at the nominal operating current. Now, the absolute value of the current isn't extremely critical, but



you will need to match both channels. Although 8mA is the design current here, going a little bit higher or lower (say, between 7.5 and 9mA) won't ruin the performance.

There's one more little trick up my sleeve—R4. The CCS needs to carry all of the 6SN7's plate current and drop a lot of volts from the B+ rail (you'll see later why the B+ rail needs to be fairly high, about 350V). So because the CCS has such a high impedance, adding some series resistance doesn't much affect its operation, but R4 will dissipate some of the power necessary in the plate circuit. This will eliminate the heatsinking requirement for Q2, which would otherwise need to remove more than a watt of power. I chose its value to drop about half of the necessary voltage. It serves a more subtle purpose, too—it isolates any of the stray capacitances of the power FETs from the plate circuit. For thermal reasons, it should be generously rated—a 2 or 3W unit is ideal.

Volume controls are a common weak point in analog pre-amps. You should use a high-quality one; a stepped attenuator would be my first choice. For my prototype, I selected parts on hand wherever possible, so I used an Alps Black Beauty. Personal choice should rule here. A too-large value will interact with the Miller capacitance of the 6SN7 to cause treble losses, and too small can cause loading issues for the input transformer. You need to pick the input circuitry as a team!

You must first and foremost cater to the whims of the prima donna, the input transformer. It is more than a bit fussy about how its secondary is loaded; otherwise, it will make you sorry you ever bought it. And here's where you'll need to do a bit of experimenting if you use a different input transformer and want the absolute best performance the pre-

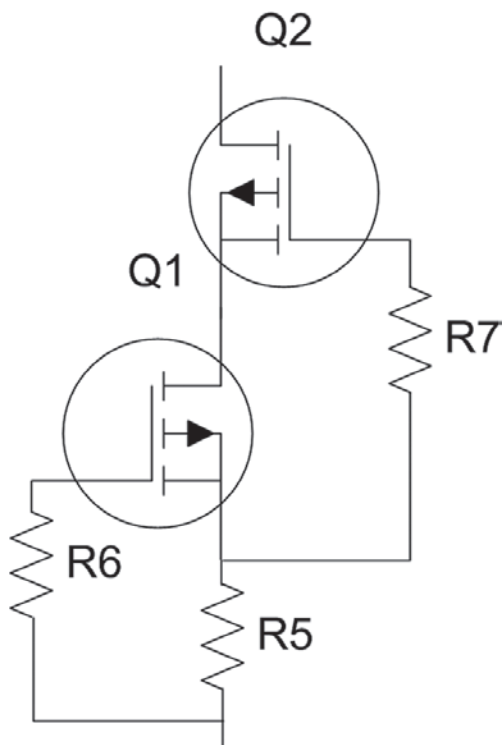


FIGURE 5: A simple yet high performance CCS.

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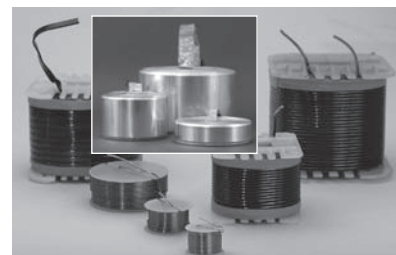
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amp is capable of. If you use the Jensen, I've saved you some time wrestling with scopes and square waves; a 100k control with a 15k resistor shunting it (no capacitor needed) works nearly perfectly. There is some minor variation depending on the position of the volume control (this is because of the ~80pF Miller input capacitance of the 6SN7), but it is still pretty close to optimal at all positions. For other transformers or other values of volume control, either follow the manufacturer's recommendation or adjust the values of R1 and C1 using a square wave source.

OUT HERE IN THE FIELDS

With the voltage amplifier stage

wrapped up, you can move on to the output stage, which is tasked with taking a single-ended signal and converting it to balanced (the so-called phase splitter), then driving the power amp and associated interconnects.

The ECC88/6DJ8 in the output stage offers a very low source impedance (100R, typical) and it is an easy tube to find in all sorts of flavors. Because of the 100% degeneration in this circuit, there's little advantage to using any of the collectors' items in this spot. And other tubes will also work well here—an ECC99, for example, would be a fine choice. I have substituted 7308, 6KN8, and 6922 with no measurable or audible difference. The key virtues are reason-

able current capacity and high transconductance.

Now, how do you set up the output stage? In a split-load inverter, the current through cathode and plate resistors is equal, so the voltage drops across them are equal. I want to run the current high enough to keep the tube's transconductance high, but not so much as to threaten reliability. As with the first stage, 8mA is a good compromise. Now we need to choose the load resistors. Too low and the distortion goes up, too high and it's difficult-to-impossible to get 8mA of current at any reasonable supply voltage.

Because I've used ECC88s in line amps with a 30k load and got reasonably

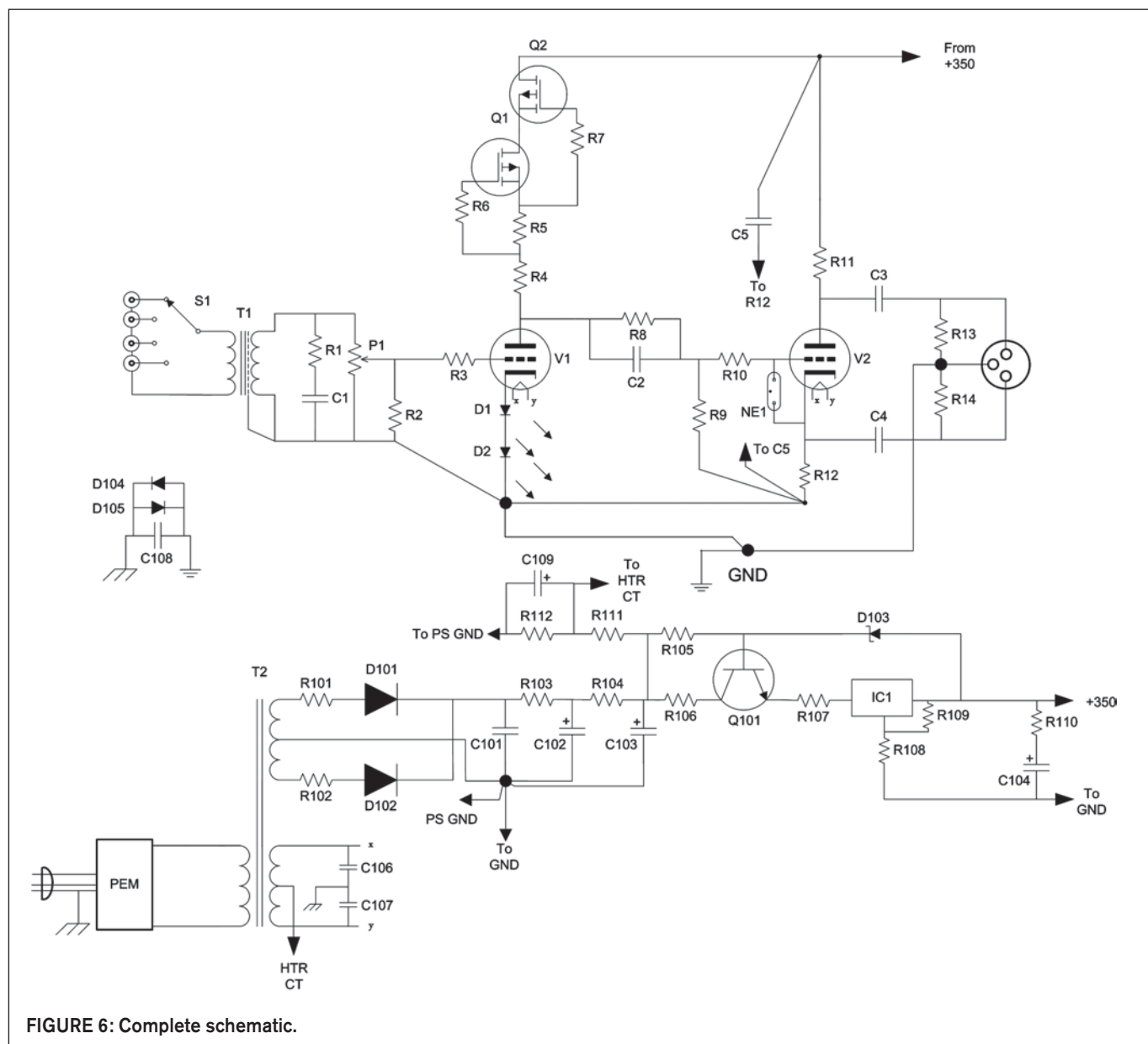


FIGURE 6: Complete schematic.

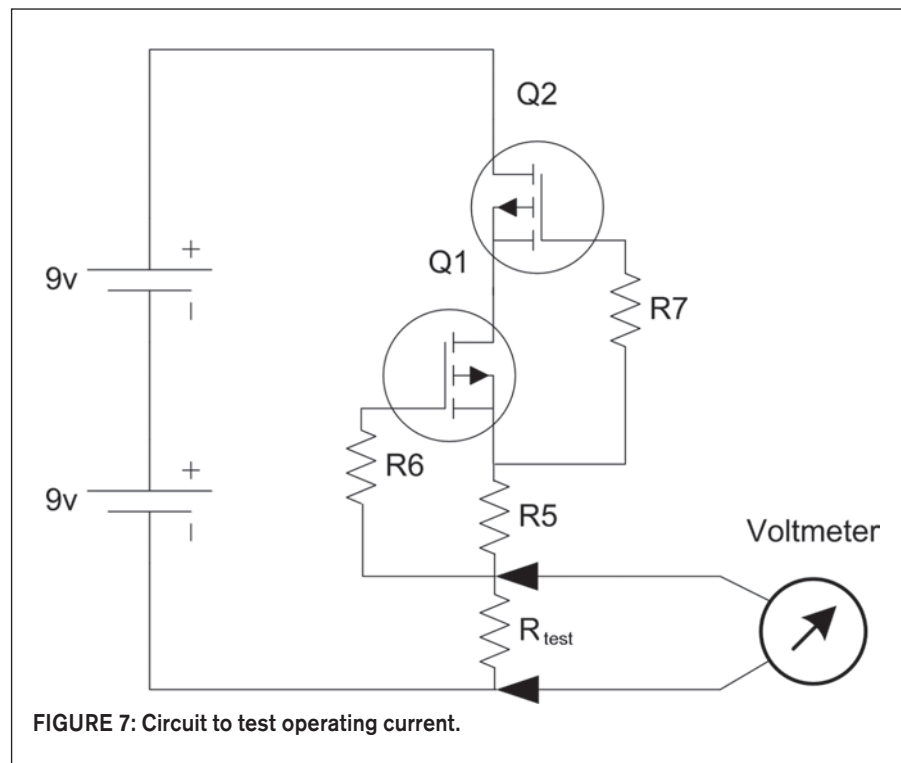
good distortion performance, and 8mA gives a 240V drop (plenty to supply our swing requirements), I will declare that the Official Value. Splitting the load means that the cathode and plate resistors, R11 and R12, are each 15k, with a drop of 120V or so across them at the nominal 8mA. You can now see why the B+ needs to be at 350V or more. R11 and R12 each drop about 120V, and you want 100V or so across the tube to ensure proper operation at the high voltage swings required.

R11 and R12 should be very tightly matched so that the balance and distortion are optimized. If you buy a dozen or two 1% resistors, it should be easy to find a couple of pairs that match to within 0.1%; the absolute value isn't critical, the relative values are, so the calibration of your ohmmeter isn't going to be the determining factor in matching.

With the cathode resistor at 120V, the grid will need to be slightly negative of that value, about 118V or so. This presents a small problem—the 6SN7 will likely have 170-180V or more on

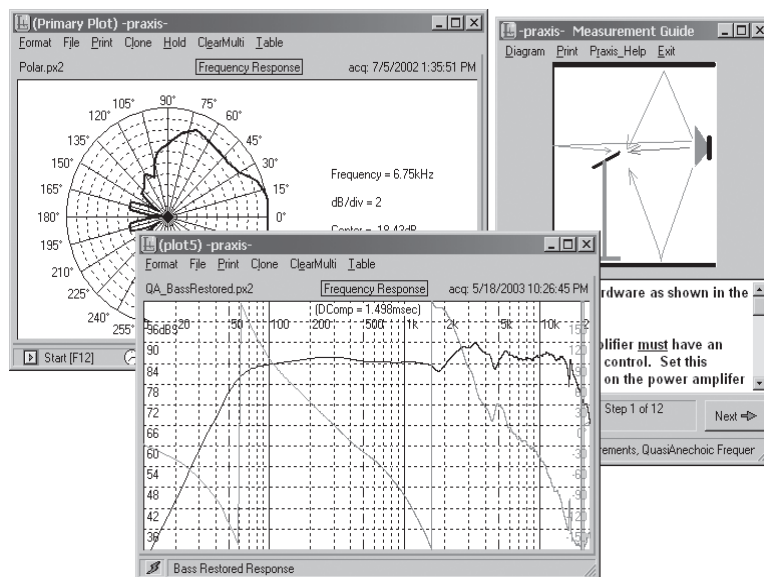
its plate, so direct coupling becomes difficult. In order to get the output stage grid to the right voltage, you need to

lose 40V or so of DC. There are several ways to do this—for the sake of simplicity I used a voltage divider (R8 and R9)



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with the upper leg bypassed by C2. The voltage divider resistors slightly load the voltage amplifier, but at 1M minimum, the loading is pretty insignificant, more than a hundred times higher than the 6SN7's plate resistance.

The choice of output coupling caps, C3 and C4, will depend on the expected load; their value sets the bass rolloff corner, with a 3dB down point calculated from $f_3 = 1/(6.28RC)$, where R is the expected load resistance. For a 100k load, a very moderate 0u47 capacitor will give an f_3 of about 3.5Hz. Although the F4's stock input resistor is 47k, its value can safely be increased to 100k without worries about offset voltage or RF pickup. The important thing is to have the input resistors as well matched as possible because they are part of the AC load of the split load inverter. As before, 0.1% matching is easy to achieve by buying more than you need and spending time with an ohmmeter.

R13 and R14 hold the output voltage at ground in absence of a load. Their matching is less critical than the lower value load resistors and power amp input resistors. 1% is good enough.

NE-1 serves to limit the grid-to-cath-

ode voltage of the ECC88 at power-up. This is often accomplished by the use of a diode, forward-biased at power-up, but reverse-biased after warm-up. And if a neon bulb is difficult to source, you can use a diode, with its cathode (banded end) connected to the cathode of V2. The advantage of the neon bulb, besides a visual indicator of power-up, is that its capacitance is lower and more linear than a reverse-biased diode. It lights up with a nice orange glow when you first turn on the power switch, then extinguishes as the preamp warms up and the bias LEDs start to light.

POWERHOUSE

I have endeavored to make this circuit uncritical of the power supply, but it is still worthwhile to try to get to the point of diminishing returns. The huge power supply rejection afforded by the CCS loading of the input stage and the large voltage swings make this a very straightforward exercise. You need a stable 350V for the B+ rail; the constant-current of the voltage amplifier and 100% local degeneration of the output stage go a long way to ease the demands for ultra-low impedance. Nonetheless, it is easy to

make a supply that is orders of magnitude better than what you strictly need for not much more complication and cost than a supply that just does the minimum. So I will don my 16th piece of flair and use an active regulated supply with reasonably low noise and source impedance.

The Maida regulator, based on National Semiconductor's LM317, has made guest appearances in such designs as Joe Curcio's Stereo 70 conversion, Morgan Jones's Crystal Palace, and my own Red Light District. Because of the signal levels involved and the inherent power supply rejection of the signal circuitry, I've used a simplified version closer to Curcio's than Jones's or the original Maida circuit. If you want to go crazy and add all the extra bypass caps and the protection diodes needed to deal with the consequences, you can boost the regulator's measured performance, but the sonic effect will be minimal-to-nonexistent. The Maida regulator is easy, low-cost, quite stable, and reliable.

Because the design is discussed thoroughly in Maida's paper, I won't rehash it in detail. I used a TIP50A as the

Parts List

Resistors (Signal circuit, two of each needed for stereo)

R1.....	18k, metal film (see text)
R2, R9, R13, R14.....	1M, 0.125W
R3, R10.....	1k, carbon composition, 0.25W
R4.....	12k5, metal film or wirewound, 2W
R5.....	300R, metal film, 0.125W
R6, R7.....	120R, 0.125W
R8.....	680k, 0.5W
R11, R12.....	15k, metal film or wirewound, 1W, matched to 0.1%

Controls

P1.....	100k stereo potentiometer, audio taper
S1.....	2 pole 4 or 5 position rotary switch

Capacitors (Signal circuit, two of each needed for stereo)

C1.....	Not used, jumpered (see text)
C2, C5.....	0.1µF, 630V polypropylene
C3, C4.....	0.47µF 400V polypropylene

Semiconductors and lamps (Signal circuit, two of each needed for stereo)

D1, D2.....	Red LED, 1.7V
Q1, Q2.....	DN2540N5 Depletion mode MOSFET
NE-1.....	NE2 neon bulb

Transformer (Signal circuit, two required)

T1.....	Jensen JT11P-1 or equivalent 1:1 input transformer
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Resistors (Power Supply)

R101, R102.....	47R, 0.5W
R103.....	470R, 3W, wirewound or metal oxide
R104.....	1k, 3W, wirewound or metal oxide
R105.....	20k, 0.5W
R106.....	100R, 0.5W
R107.....	47R, 0.5W
R108.....	56k, 5W, wirewound
R109.....	200R, 0.5W
R110.....	4R7, 0.25W
R111.....	270k, 0.5W
R112.....	47k, 0.5W

Capacitors (Power Supply)

C101.....	1µF 630V polypropylene
C102, C103.....	100µF 450V electrolytic
C104.....	47µF 400V
C105.....	Not used
C106, C107, C108.....	0.01µF 600V ceramic disc
C109.....	10µF 100V electrolytic

Semiconductors (Power Supply)

D101, D102, D104, D105.....	UF4007 fast recovery rectifiers
D103.....	Zener diode, 12V 1W
Q101.....	TIP50A or equivalent
IC1.....	LM317 adjustable regulator

Transformer and Miscellaneous (Power supply)

T2.....	Allied 6K3VG, 650VCT at 40mA, 6.3VCT at 2A
PEM.....	Power Entry Module, SAE PM1B-6 (included with DIY1712 case)

pass device, but there's no reason why any other moderate power transistor wouldn't work there. A power MOSFET could also usefully serve in that position.

Under operation, assuming a 400V raw supply (a good compromise between headroom and pass transistor dissipation), the regulator needs only to drop 50V, but remembering that the regulator will be turned on and off with the preamp, do not try using a low voltage device for the pass transistor! The regulator draws about 6mA, the signal circuits for a stereo version another 32mA. So if you estimate the total current to be 40mA, the pass transistor and LM317 dissipate $(400 - 350 = 50)$ volts times 0.040A, or 2W. Most of that is dropped by the pass transistor, so it should have some heatsinking.

The output voltage is set by 1.25 times the ratio of R108 to R109. R108 dissipates a bit over 2W, so size it generously. It may be useful to make it from a series or parallel combination of two 2W resistors.

The output capacitor, C104, has a re-

sistor in series with it, basically to flatten out the inductive output impedance of the LM317 regulator. This series resistor means that money spent on an ultra-low ESR cap for C104 is better directed toward booze and gambling. The signal circuitry is locally bypassed by C5.

How do you get the 400V for the raw supply? Looking in the iron pantry, I found a nice power transformer from Allied, the 6K3VG, rated at 650VCT, 40mA. It's overkill for this job, but it's reasonably cheap (\$25), has a shield layer, and is easy to find. There's no reason why a surplus unit won't work just as well. All you ask is that you end up with 400V at reasonably low ripple to feed the regulator. This particular transformer has a conventional E-I core; other constructions will also work—some even better, but avoid toroids. With the universal adoption of compact fluorescent lights and switching supplies, power lines are noisier than ever, and toroids diligently couple that noise into the power supply.

High-speed rectifiers are cheap these days, so why not use them? R101 and R102 keep the ripple currents under

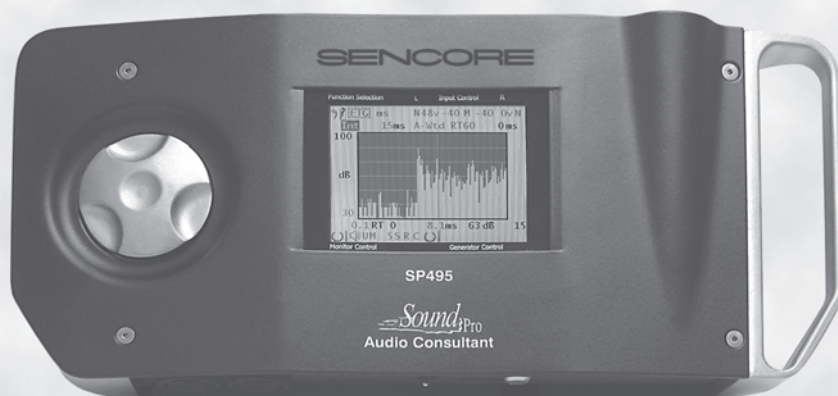
control and give the diodes some padding during power-up. I chose the value of C101 to give 400V at the raw supply output; its relatively small value again keeps ripple currents low. This is followed by two RC filters; the ratings of the resistors are reflective of RMS ripple currents. Wirewound resistors are preferred, but the metal oxide type will also do well.

On the primary side, I used a power entry module for the fuse, switching, and IEC line cord plug. You can use individual bits for this, too. Whichever approach you take, make sure that the earth ground lead from the power line is firmly connected to the chassis. Safety is the number one concern! Of the millions of ways of making a raw supply, this is certainly one of them.

As for heaters, the ideal solution is a separate heater transformer so that rectifier noise from the high voltage supply isn't coupled to the heaters. I took a shortcut and used the 6.3V winding on the power transformer. There is no need for DC here—the signal levels are high, and with some care in layout and lead

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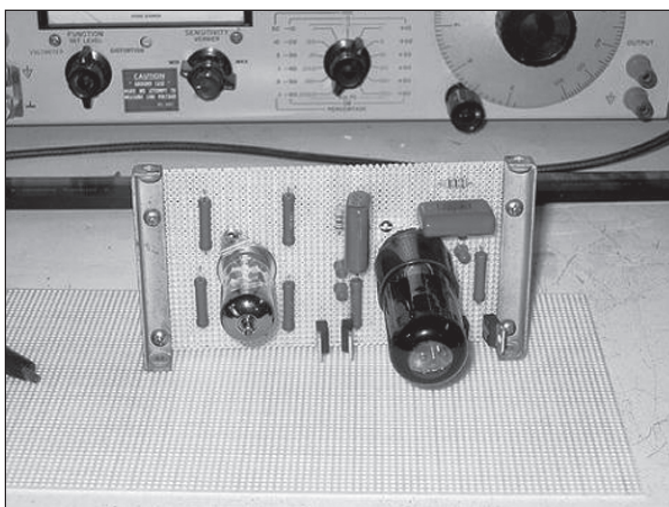


PHOTO 1: Signal circuitry built on a perfboard.

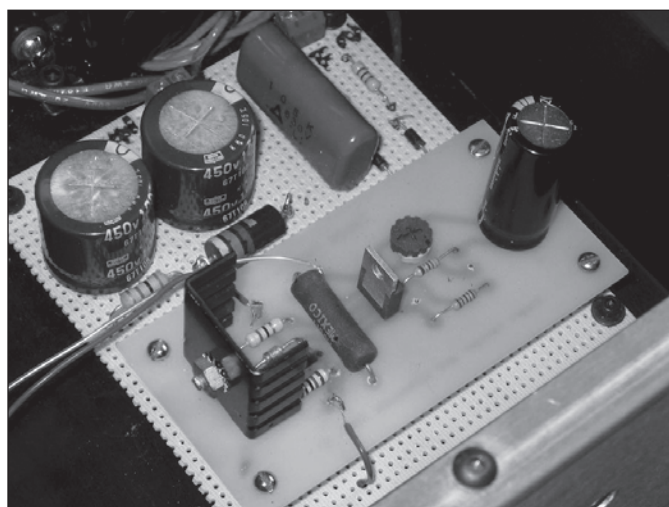


PHOTO 2: Raw supply and regulator.

dress, you can avoid hum. Capacitors C106 and C107 bypass any common-mode noise to chassis ground—they are optimally placed right at the 6SN7 heater pins, though I sited them closer to the transformer. All heater leads should be tightly twisted and dressed close to the chassis, away from any low-level signals.

As is customary, the heater supply center-tap is DC biased above ground, but AC bypassed to ground by C109. This minimizes hum pickup in the 6SN7 and keeps the ECC88 heater-to-cathode voltage within the specified maximum.

WE CAN BUILD IT, YES WE CAN

For the prototype, I used a mix of modular boards with different construction techniques. The circuit is simple enough to be done point-to-point or on a perfboard; if there is interest, it would be relatively easy to put the whole thing on a PCB. **Photo 1** shows the signal circuitry built on a perfboard. Because of their size, I put the output caps, C3 and C4, on a separate board, positioned between the main signal circuitry board and the output connectors.

Photo 2 shows the raw supply and regulator, built on a mixed platform of perf and PCB. As is evident,

there's nothing exotic in the componentry, nor (given the pains we took in the circuit design) is there any need.

Grounding is straightforward—I used a set of individual stars all brought to a common point, much as indicated on the schematic. The common point is connected to the chassis (safety) ground at one point via a groundbreaker, which helps prevent the formation of ground loops with other equipment. The groundbreaker is bypassed at RF by C108, whose value is small enough to have negligible conductance at hum frequencies.

I found a great source of casework (my own weakness) at DIY Enclosures LLC (www.diyenclosures.com). I used their DIY1712 cabinet, which had way

more space than I needed. Unfortunately, the inside height is just too small to accommodate the CV1988 and the room needed to clear the pins from the bottom of the case. This is the reason for the vertical mounting of the PCB. I chose the position so that the tubes were under the vent slots in the case's cover. I drilled a few holes in the bottom plate to aid circulation, and, as a result, the preamp runs only moderately warm.

Output connectors are XLR, just to emphasize the balanced nature. They also look cool and work flawlessly.

ON FURTHER EXAMINATION

One nifty feature of the split-load inverter is the opportunity to pick off a single-ended signal from the cathode. So if you want to drive an F4 (or other amp) single-ended, you can do so. For the output stage to run as a cathode follower, the plate should be at AC ground. Conveniently, there's already a capacitor there, C3. So by connecting the output side of C3 to ground, you have bypassed the plate to ground and made a single-ended output available at C4.

This is most easily done by making up a set of interconnects with XLRs at one end and RCA plug at the other. Inside the XLR, wire a jumper between the

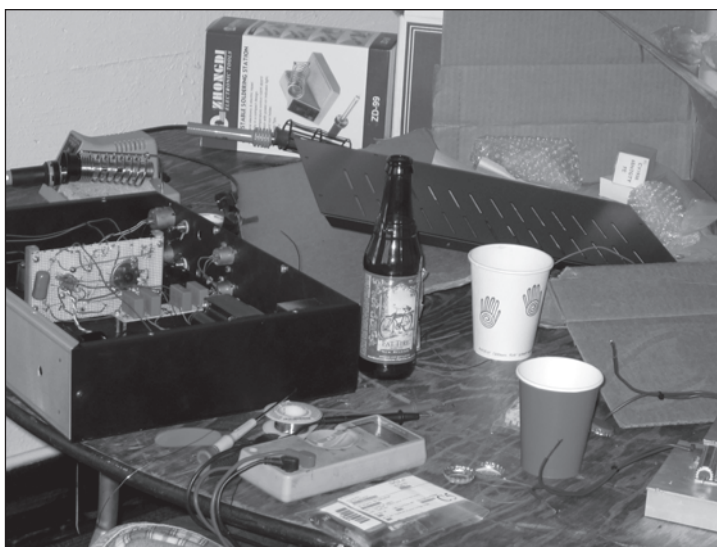


PHOTO 3: Final wiring in a back room at the Burning Amp Festival (and before clean-up).

plate output pin and ground. That way, you can go back and forth between balanced and single-ended just by plugging the appropriate interconnects in and out. Note that the preamp run single-ended inverts polarity.

The test bench performance was all I could hope for. Mid-band THD was 0.04% at 20V peak (40V peak-peak) out into a 10k load, single-ended. This broke down as -68dB second harmonic, -92dB third, and all other harmonics below -100dB. Distortion at 25Hz was slightly worse, with third harmonic measuring -89dB. Hum and noise were below 0.5mV, nearly -100dB from full output. At 2.83V, frequency response was flat ± 0.5 dB from 5Hz to 48kHz (my test limits).

If the 26dB of gain is not an issue, this can serve as a good line amplifier in other installations. But really, it needs to stretch out and swing some volts. It has not escaped my notice that this preamp looks very much like the input stage and phase splitter for a push-pull tube power amp. And there's no reason not to build a tube unity gain power amp with a bal-

anced input.

Though I have not bench-tested with other 6SN7s, a quick listening test showed that the preamp's sonics were not degraded by the substitution of a 5692 or a button-base 6SN7GT.

Ever in a rush to get things done, the final wiring took place in the back room of the Burning Amp Festival (**Photo 3**). This occasioned derisive hoots from passersby and curious stares from those who wondered why we'd be soldering instead of doing sensible things like drinking, socializing, and listening to music. Nonetheless, when the preamp was plugged into the balanced F4s, driving a pair of Basszillas, the rush and stress all were worth it—dead silent background and no artificial sweetening or compression. It did everything sonically that you could want from a preamp, driving the Pass amps without breaking a sweat.

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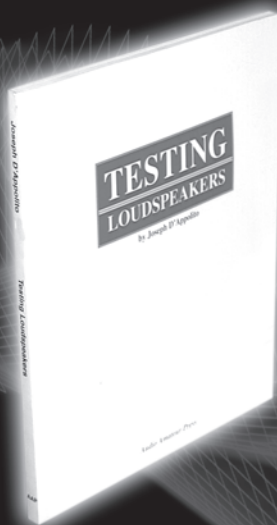
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ACKNOWLEDGMENTS

As usual, the gang at diyAudio.com were a great inspiration and help. Mark Cronander really got the project started, and Chris Bridge did far more than a fair share of construction work. I thank Nelson Pass for the use of **Fig. 1** and for his support and encouragement, Morgan Jones for his usual perceptive critique (and the CV1988!), and Cynthia Wenslow of Rising Moon Studio and Steve Eddy of Q-Audio for assistance with drawings and organization. *aX*

The book on loudspeaker testing from MTM inventor, Joseph D'Appolito!



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Simulating Inductors and Networks

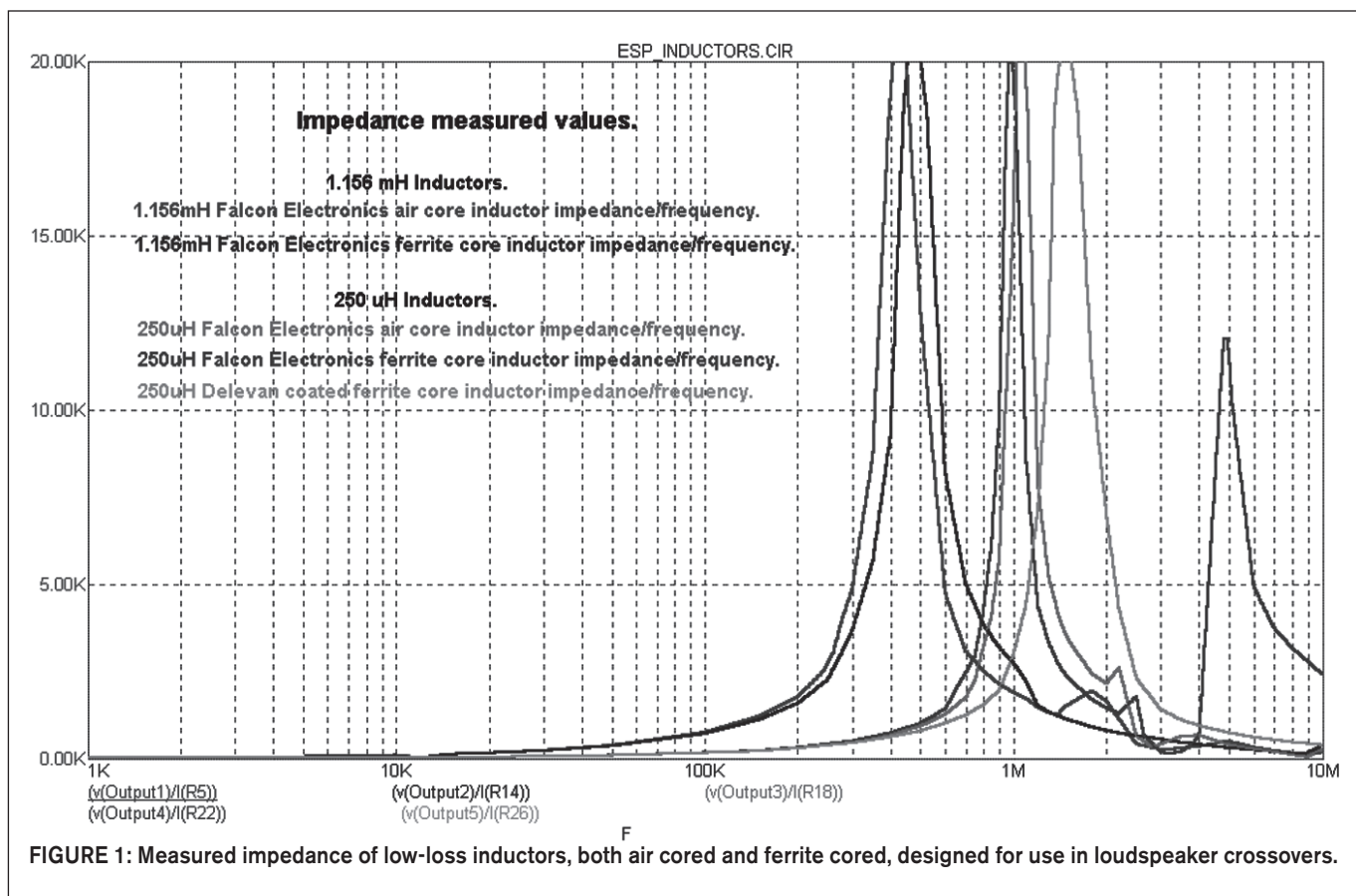
Using the Micro-cap7 software, the author introduces a hands-on approach to SPICE circuit simulation to devise new, improved, user models, able to accurately mimic inductor behavior by frequency.

SPICE, the “Simulation Program with Integrated Circuit Emphasis,” has become by far the most widely used circuit simulation program. It is provided with competent transistor macromodels, so performs well when simulating integrated circuits. Basic capacitor, resistor, and inductor models are also included, but only as idealized, theoretically perfect components, suited for use with the small values used when

modeling integrated circuits, but far removed from almost all discrete, real-world components. This article shows how vastly improved, realistic inductor models can simply be produced using only the basic primitive elements found in every SPICE simulator.

Most SPICE analyses are made using large signal transient simulation to produce a time domain waveform of the circuit’s behavior, just like probing the circuit

using an oscilloscope. When modeling in the time domain, you can modify SPICE models to account for amplitude nonlinearity. You can also perform small signal AC, frequency domain simulations, and when using SPICE or a SPICE equivalent simulator, you can modify capacitor, resistor, and inductor models to also account for parameter changes with frequency. However, within SPICE, these frequency and amplitude dependent pa-



rameters are mutually exclusive; you cannot use both parameter sets within one simulation.

Of course, SPICE is not the only type of simulator; other simulators model principally in the frequency domain and many can combine both time and frequency domains together by using convolution as in the "Harmonic Balance" simulators. Such simulators are usually provided with a library of truly competent inductor models able to replicate real-world components, even to very high frequencies.

INDUCTOR MODELS

Unfortunately, few component makers provide realistic "SPICE" models, unless their components are specifically intended for use at high frequencies, for example, Coilcraft, which provides accurate simulation models for use in the "Microwave Office" software. As a result, simulations of larger value inductance using the basic "SPICE" model, even at modest frequency, result in large errors and misinterpretation of circuit behavior. Let me examine actual measurements of a few "real" inductors (Fig. 1).

All practical inductors include self-capacitance between adjacent wire turns and from the start and finish end-to-end turns. When using toroid cores, these end-to-end turn capacitances dominate unless you take care to leave sufficient unwound space, typically about 90° between start and finish winds. Next I examine typical "SPICE" plots using the default

inductor models provided as standard.

The SPICE default inductor model (Fig. 2) agrees with my measured values at audible frequencies and continues to agree up to 100kHz. However, by 200kHz I find the 1.156mH inductor simulation suggests an impedance of 1454Ω, whereas the measured value was 1779Ω. Clearly these errors then increase rapidly with frequency. Many writers have used SPICE default models to simulate amplifier Nyquist response up to 1MHz, driving into a simulated loudspeaker cable and simulated crossover/speaker loads, using only basic SPICE models for cable and speaker, with invalid, unreliable conclusions.

Every inductor includes parasitic capacitance between adjacent coil turns and between start and finish windings. These capacitances result in the parallel resonances shown in Fig. 1. Inductors intended for use in loudspeaker crossover networks will exhibit a major resonance, typically between 100kHz and 2MHz; larger values resonate at even lower frequency. At higher frequency, as shown in Fig. 1, other lesser resonances should also be anticipated. For example, the 3.5mH inductors used in my horn-loaded speaker crossover resonated at 210kHz, but exhibited many smaller high frequency resonances.

AN IMPROVED INDUCTOR MODEL

Using SPICE basic components and with

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The **FF165K** is run full range with a frequency response out to 15kHz. The **T90A** super tweeter has been added to cover the upper frequencies. The **T90A** is a top-mount horn tweeter with an **Alnico magnet**. The tweeter is rolled off on the low end with a **Fostex Tin & Copper foil capacitor**. The system frequency response is 55Hz to 35kHz at 95dB.



The **BK-16** cabinet is made from **Baltic Birch** plywood and is sold flat, unassembled, unsanded and unfinished. Cabinet dimensions are 9.75" W x 14.75" D x 29" T.

The kit includes:

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- Pair of FF165K - full range
- Pair of T90A - horn tweeter
- Pair of DB-Cup - Input cup
- Pair of Crossovers
- Nordost 2-Flat wire for tweeter
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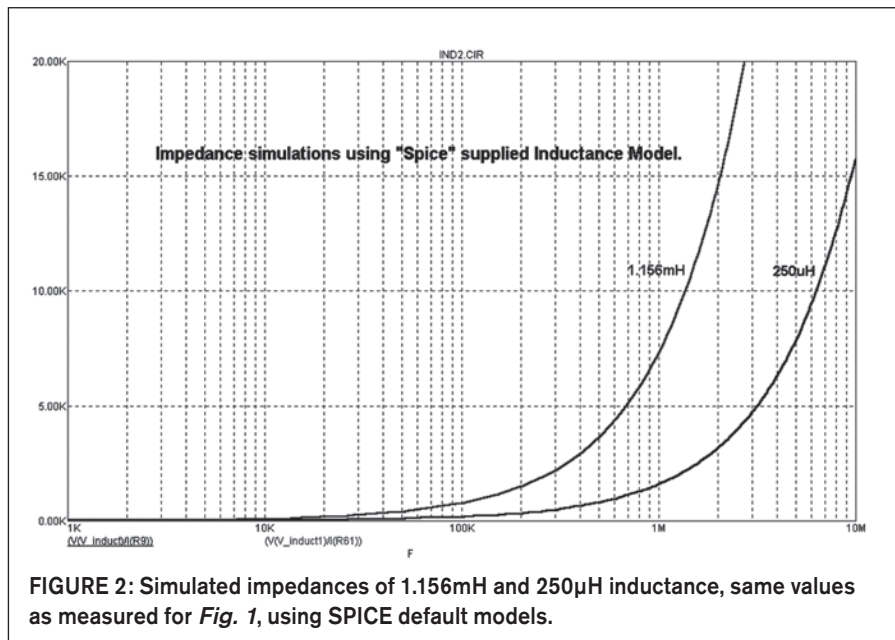


FIGURE 2: Simulated impedances of 1.156mH and 250µH inductance, same values as measured for Fig. 1, using SPICE default models.

a knowledge of an inductor's resonant frequency, you can easily devise a more appropriate computer model. You can calculate the effective self-capacitance using the equation:

$$C = \frac{1}{4\pi^2 F^2 L}$$

derived from the more common equation

$$F = \frac{1}{2\pi\sqrt{LC}}$$

Using the 400kHz resonance for the 1.156mH inductor (**Fig. 1**), you find its equivalent capacitance, the combined effect of its turn-to-turn and end-to-end capacitances to resonate at 427kHz must be 120pF. Using just these values results in too sharp a resonance, because the real inductor exhibits a measurable DC resistance of 2.184Ω; almost all is effectively in series with the inductance. Also because the resonating capacitances are complex, they will exhibit resistive losses; some loss will be in series with this capacitance, which affects the resonance width, together with a large shunt loss resistance value, which limits the peak impedance value.

These series and shunt resistances are best devised by running a few simula-

tions, adjusting values as needed to approximate measured values. I find that a peak impedance shunt resistance of 40kΩ makes a good starting value. The measured DCR in series with the inductance, less about 0.1Ω for those lead wires which lie outside the main winding, provides an excellent fit to measured slope values, which can be finally tuned by adjusting a small resistance, in series with the shunt capacitance. These “cut and try” adjustments take little time, but how well do they work (**Figs. 3 and 4**)?

With a few extra components, you can add an additional “node” to also model secondary resonances.

Loudspeaker drivers can be treated as an inductive component, but for accuracy to high frequency they require several (many) additional nodes. With the impedance of a driver measured to about 10MHz, in addition to the usual audible frequency resonances which are usually well modeled, the drivers I measured ranged from an 18” 250W Goodmans Power series with a peak impedance of 495Ω at 400kHz to a Kef B110 with peak impedance of 570Ω at 1.1MHz. In my ESP_replica network, the bass driver

used measured 843Ω at 500kHz, and the T27 tweeter measured 825Ω at 2.4MHz. Within this frequency band, other drivers all measured as intermediate impedance peak values. With inductors typically peaking around 40kΩ, clearly the combined impedance of crossover and speaker driver at high frequency must approximate 5-600Ω when resonant, reducing with further increase in frequency.

I had two quite different crossover/speaker cabinets available. The first was my horn-loaded two-way system (**Figs. 5 and 6**), which measured a peak impedance of 525Ω at 900kHz. The other, the ESP replica assembly as used for my cable evaluations paper, measured a peak impedance of 575Ω at 1.5MHz.

At frequencies higher than this inductive resonant peak, the inductive leading phase no longer applies. The measured phase angles now lag, and the inductor impedance reduces with frequency, just like a capacitor. This often results in the circuit drawing unexpectedly high and capacitive currents.

MODEL DEVELOPMENT

The simplistic, idealized models for ca-

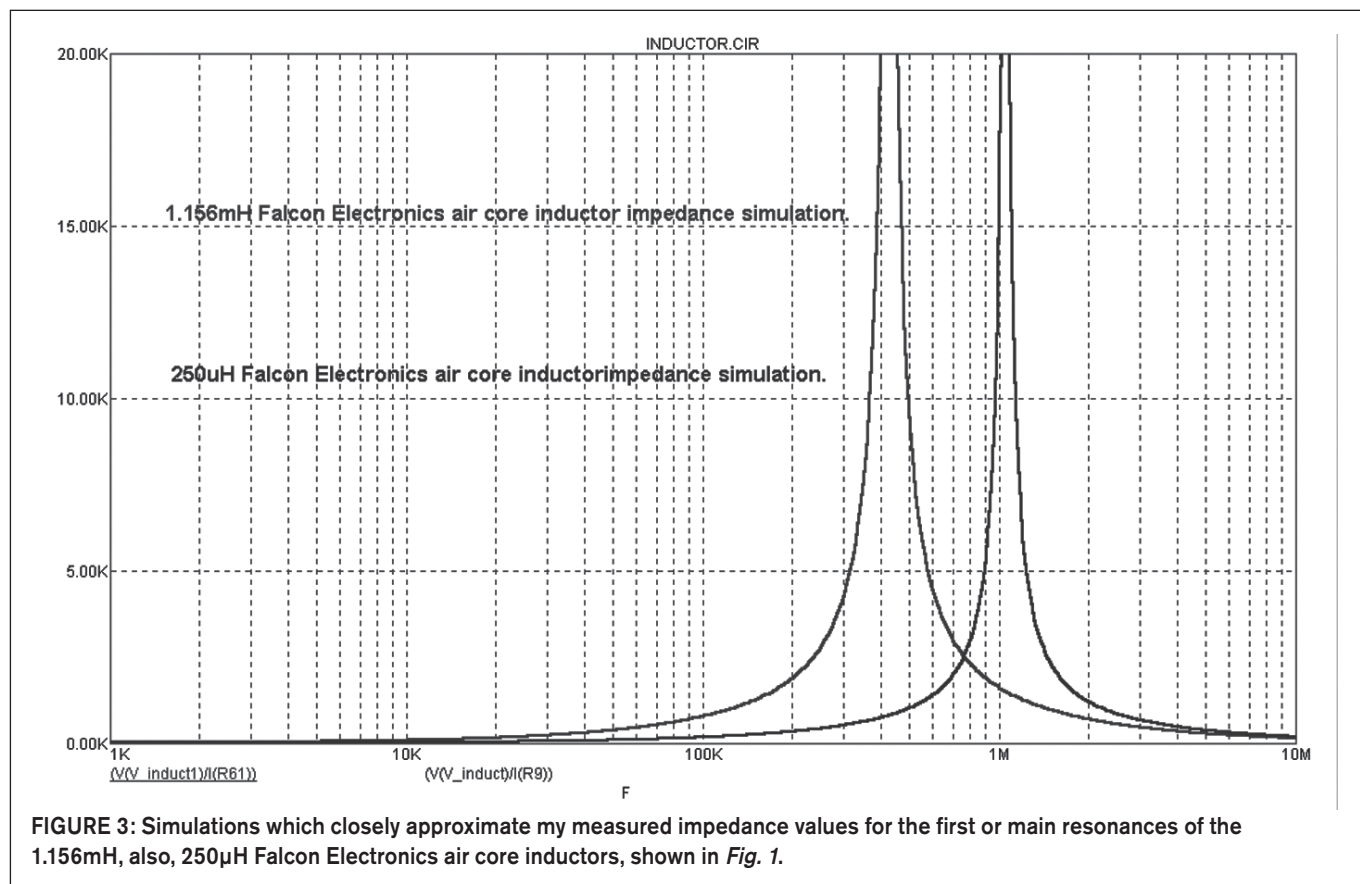


FIGURE 3: Simulations which closely approximate my measured impedance values for the first or main resonances of the 1.156mH, also, 250μH Falcon Electronics air core inductors, shown in *Fig. 1*.

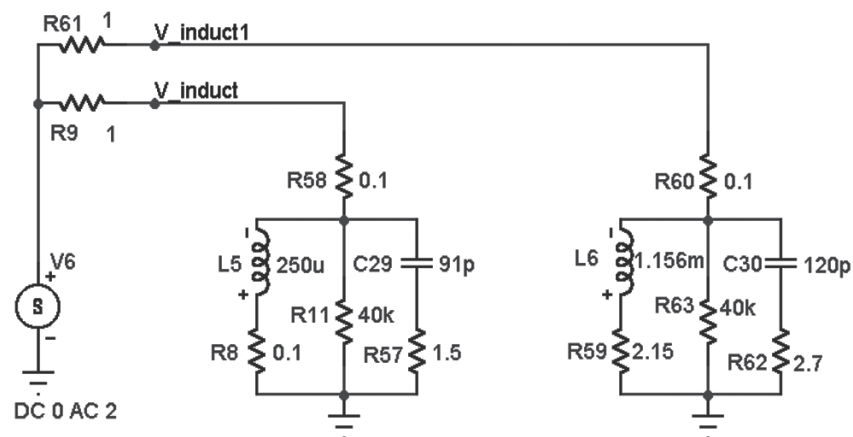


FIGURE 4: The simple SPICE models as used for the *Fig. 3* realistic simulations, of the first or main resonances for the 250 μ H and 1.156mH Falcon Electronics inductors.

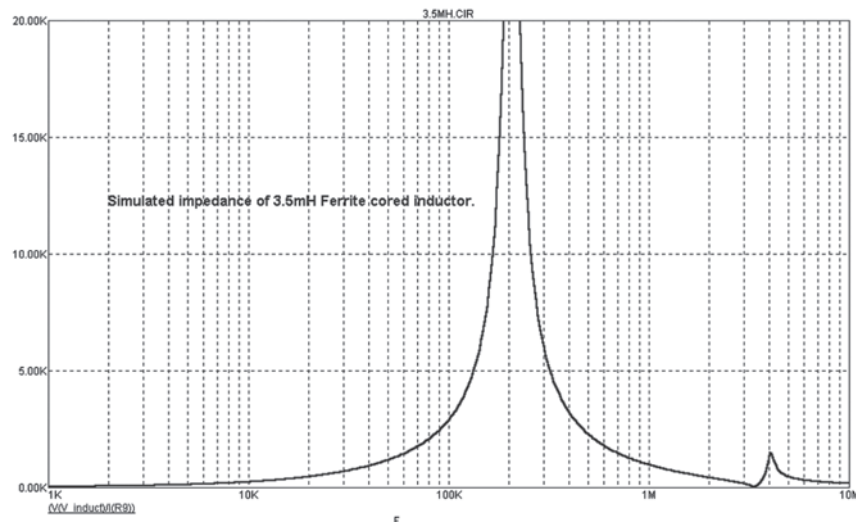


FIGURE 5: Simulated impedance of 3.5mH, ferrite cored inductor as used in my horn-loaded speaker crossover network. Actual measured peak impedance was 40k Ω at 210kHz.

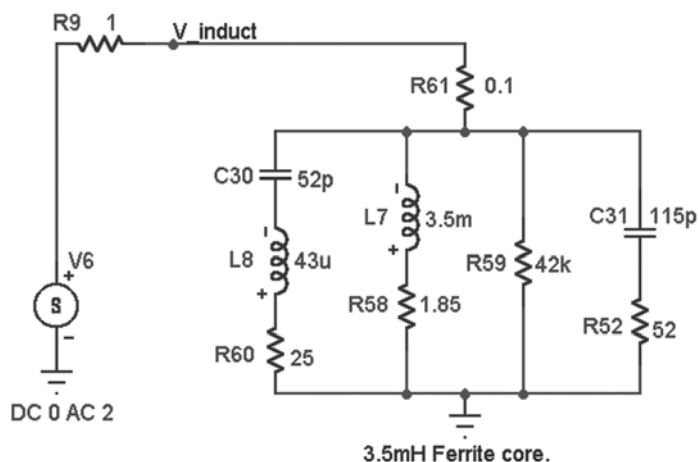
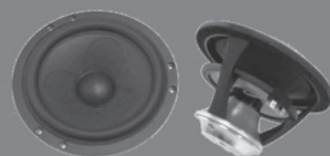


FIGURE 6: Model used for *Fig. 5* simulation.



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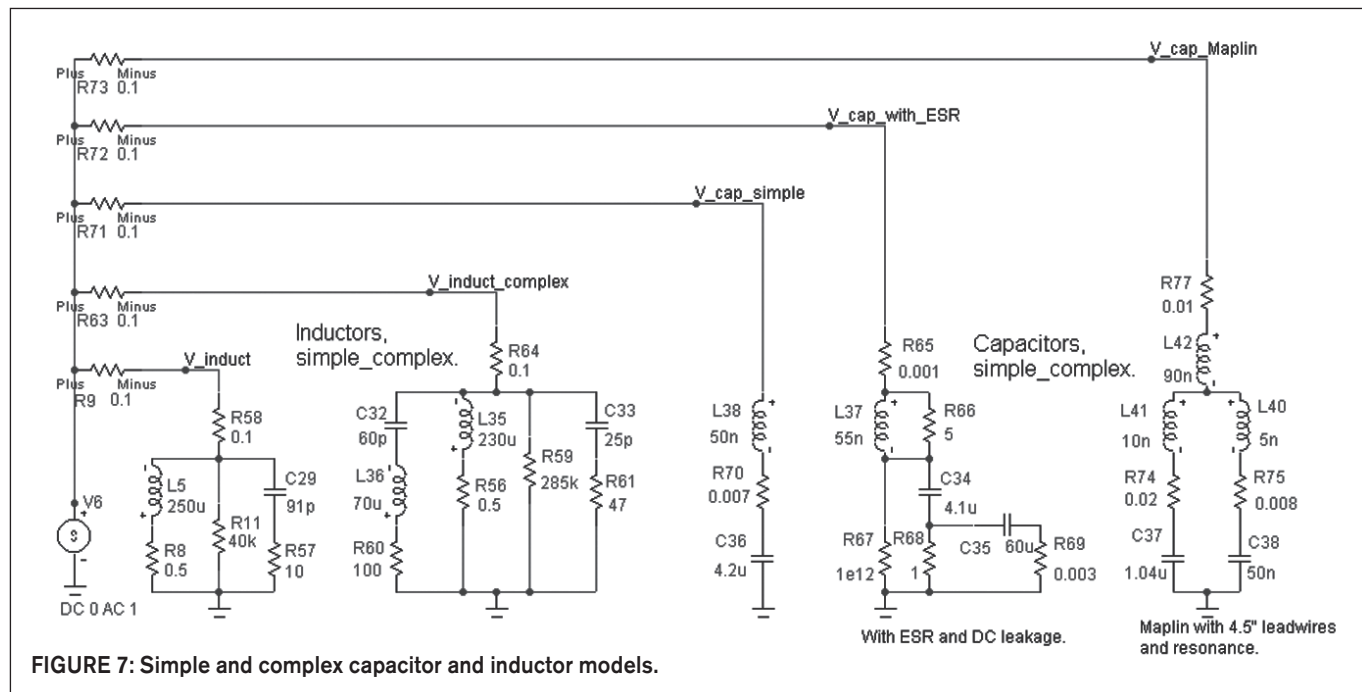
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capacitors, resistors, inductors, and transmission lines provided in most SPICE-based simulators can be useful for DC, transient, and low-frequency AC modeling; they do not, however, represent any

practical components. In the real world, capacitors, resistors, and especially inductors, measure quite different from their SPICE predictions.

All components include parasitic ele-

ments, so are better described at least to moderate frequencies, by combining all three elements: capacitance, resistance, and inductance for each device. For realistic high-frequency simulation, every com-



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ponent is then better described using a distributed, transmission line like, model. To complicate matters further, in real components the values of these parasitic elements invariably change with measurement frequency. SPICE allows model values to be modified using the “.model” statement for transient or frequency “F” factors for AC simulations, but these two options are mutually exclusive.

CAPACITORS

Begin by examining an improved capacitor model (Fig. 7). The ideal capacitor exhibits a 90° phase angle between its applied voltage and through current, but every practical capacitor must use metallic electrodes and metallic external connections, both of which introduce resistance and inductance. Any solid dielectric insulator exhibits its own dielectric losses. Combined together, these result in a phase angle at low frequency less than 90° and increasingly so with an increase in frequency. This degraded phase angle can be represented by using the appropriate, very high value shunt resistor, which, placed across a perfect capacitor, results in the same phase angle. This high resistance value is usually measured as conductance in siemens, the reciprocal of ohms.

This degraded phase angle is more usually represented by the equivalent very low value resistor in series with the perfect capacitor. This ESR value or equivalent series resistance can also be derived from measured values for capacitance and $\tan\delta$ by multiplying $\tan\delta$ by X_c , the capacitive reactance at that specific frequency; $\tan\delta$, of course, is frequency dependent. The capacitor electrodes and connection resistances exhibit inductance, which also acts in series with the capacitor.

A reasonable estimate for self-inductance is to allow 7.5nH for each 1cm length of straight leadout wire, plus a slightly lesser amount for each 1cm of the capacitor body length. Should the maker's impedance plot be available, then you can more accurately calculate self-inductance knowing capacitance value and the capacitor's self-resonant frequency. At self-resonance, the capacitor's reactance and self-inductance become of equal value but opposite phase angle, so they cancel out.

$$\text{Hence, } L = \left(\frac{1}{4 \times \pi^2 \times F^2 \times C} \right)$$

The most nearly perfect capacitor would exhibit near constant degraded phase angle or $\tan\delta$ with frequency. Because for each doubling in frequency, capacitive reactance halves, the ESR for the near perfect capacitor must also halve. In a practical capacitor, ESR is frequency dependent, almost but not quite halving for each octave increase in frequency. When using the shunt resistor or G, its equivalent conductance, both values are strongly frequency dependent.

In this case, to maintain this degraded measured phase angle, the shunt resistance must more than halve and conductance (G) must more than double for each octave increase in frequency. At some frequency the metallic electrode and connection resistances dominate the dielectric contribution. ESR then reaches a minimum value, after which it slowly increases due to “skin” effects. It is certainly never a constant as many wish to believe.

The simplest fixed-value three-component capacitor model shown in Fig. 7 can suffice over a narrow frequency range both for AC and transient simulations. Making these parameters frequency de-

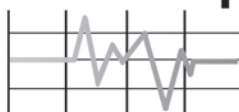
pendent extends the model useful range but negates use for transient simulations. By adding a second, larger capacitor and resistor, you can realistically model over a more useful range, both for AC and transient simulations, while retaining fixed element values. The required values are easily calculated either from makers' published graphs and data or from measured values (Fig. 8). Similar schematic models are provided as standard in the better, non-SPICE-based simulators.

INDUCTORS

In a similar fashion—provided the maker's impedance curves are available—you can quickly calculate the values needed for an inductor model. Unfortunately, for inductors used in crossover networks, such data usually does not exist.

You could measure the resonant frequency and impedance at resonance using a signal generator and suitable test meter, or, lacking this equipment, you could even make a vague estimate for resonant frequency based on the inductor value and the inductor plots shown. Despite not producing an accurate model, even that

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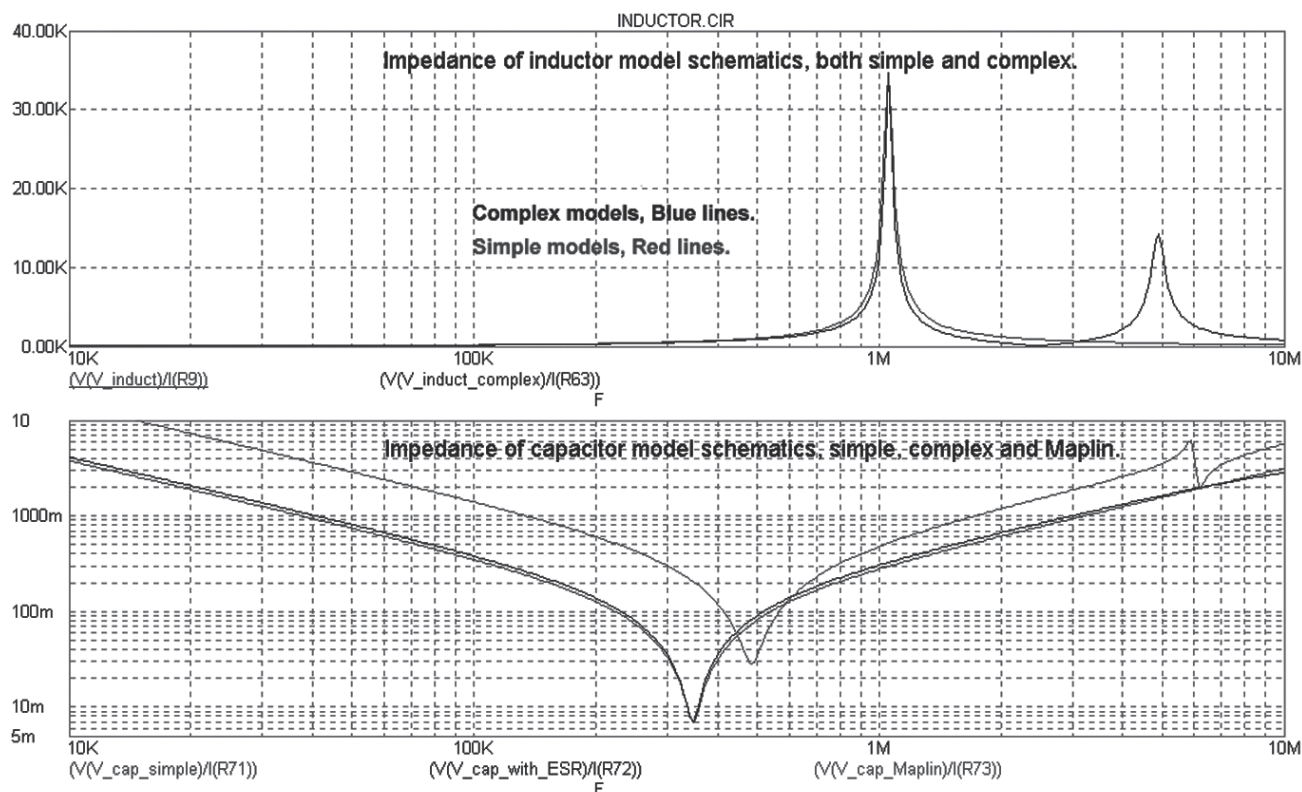


FIGURE 8: Using the schematic models in Fig. 1, you can quickly produce far better, more realistic simulations, closely matching actual measured values, for both capacitors and inductors.

\$ ESP replica crossover measured Direct, impedance and phase.
\$ No amplifier zobel or cable used.
\$ Impedance magnitude and phase measured.
\$ Frequency checked using freq counters.

```
.define ListVal (0,0,0)
+(1000,4.89,11.3)
+(2000,10.61,19)
+(2300,13.1,12)
+(2470,13.49,1)
+(2700,13.2,-7)
+(3000,10.77,-12.6)
+(4000,5.48,-10.6)
+(5000,3.92,-0.8)
+(6000,4.01,9.3)
+(7000,4.51,15.6)
+(8000,5.20,20.4)
+(9000,5.86,24.4)
+(10000,6.58,28.8)
+(11000,7.18,31)
+(12000,7.83,33.5)
+(13000,8.41,35.2)
+(14000,8.98,36.6)
+(15000,9.56,38)
+(16000,10.11,39.6)
+(17000,10.66,40.8)
+(18000,11.20,42)
+(19000,11.75,43)
+(20000,12.31,44)
```

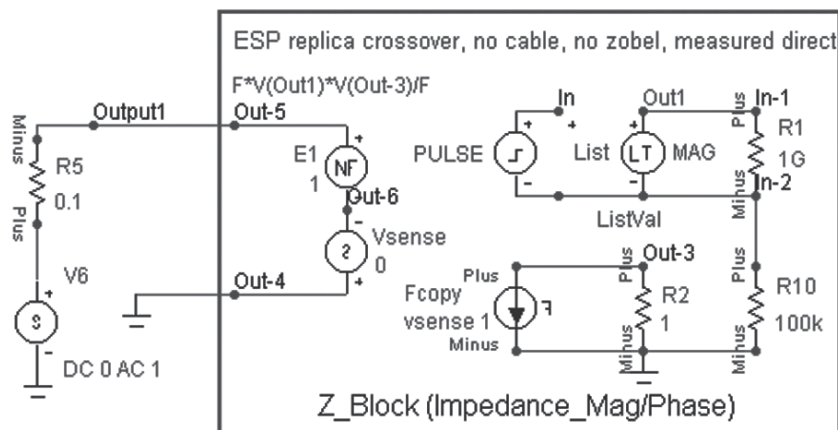


FIGURE 9: SPICE “One Port” or “Z_block” model subcircuit which you can use in simulations.

rough “guesstimate” model would be far more representative of a real inductor at high frequency than relying on the simple, basic SPICE default “perfect” inductor.

LOUDSPEAKERS

The usual published equivalent circuit for a loudspeaker driver has been devised to provide simulations only at audible frequencies. Such models then assume the speaker driver continues to act as a perfect inductor, so they do not include the inevitable self-capacitance. As a result, these models assume that speaker impedance continually increases with frequency. Unfortunately, that ever-increasing impedance is not found when measuring actual speaker drivers even at high frequency. For this article I measured impedance and phase angle of a representative range of speaker drivers, from 1kHz to 10MHz.

The T27 tweeter, used to assemble my ESP_replica circuit, was peak resonant with 825Ω at 2.4MHz, followed by a low of 105Ω at 7MHz to 390Ω at 10MHz. A Kef B110B speaker resonated with 570Ω at 865kHz, then a low of 75Ω at 3.5MHz, a small 108Ω peak at 5.85MHz to another low of 72.5Ω at 9.5MHz. I also tested 4”, 8”, and 10” low-cost full-range drivers. All exhibited a remarkably similar series of impedance peaks followed by a number of lower impedance ripples. The bass driver used to assemble my ESP_replica network peaked with 843Ω at 500kHz, followed by multiple much smaller resonances at 1.4MHz, 2.5MHz, 3.63MHz, before finally settling to a reasonably steady 120Ω .

A Goodmans 250W 18” “power” bass unit also exhibited multiple resonances, with 495Ω at 400kHz, 138Ω at 800kHz, 400Ω at 1.4MHz, 272Ω at 2.05MHz, 360Ω at 2.5MHz, 213Ω at 3.6MHz, and finally 160Ω at 6MHz, before settling down around 100Ω at 10MHz. While each of these plots differed, all measured speaker impedances increased up to a first major resonant peak, resulting from the inductance and self-capacitance of the voice coil and at frequencies usually between 500kHz and 1MHz, followed by a series of notably lesser impedance troughs and peaks, like those found measuring most inductors.

By 10MHz all resonances had flattened to a relatively consistent, moderate impedance, typically around 100Ω . None of these drivers measured as a particularly high impedance following this first resonant peak.

The Z_block model (Fig. 9) allows a CSV listing of measured frequency, impedance, and phase angle parameters to be displayed on screen or used together with other components in SPICE simulation.

You may wonder why I chose to use the Z_block to represent my test inductors. Why not simply model their schematics using SPICE? At audible frequencies with modest component values, that can work quite well, however, at higher frequencies every component used—whether inductor, resistor, capacitor, and especially speaker drivers—must use complex, multicomponent models, to accurately match resonant frequencies. Every inductor or speaker voice coil includes significant self-capacitance and resonant frequency peaks and troughs. Simplistic SPICE simulation of an inductor shows impedance continually

increasing with frequency, quite unlike the measured values’ resonant peaks and troughs, so this can lead to false conclusions.

However, you must always remember that large value capacitors are series resonant at audio frequency. The Elna 4700 μ 63V aluminum electrolytic used in my power amplifier was series resonant at 7.5kHz.

Accurately measuring an inductor, capacitor, or a complete speaker system—simply inputting measured values of impedance and phase angle by frequency into the Z_block as shown—is quicker, simpler, and most important, error free, producing the most accurate simulations possible for any component or even a complex speaker connected via its speaker cable. **ax**

Cyril Bateman’s extensive work on capacitors as originally published in *Electronics World*, is now available on CD from Old Colony Sound Lab, PO Box 876, Peterborough, NH 03458, 888-924-9465, e-mail: custserv@audioXpress.com.

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Noise Measurements of the LSK389B Dual JFET

Good noise performance with reasonable consistency was observed on a shipment of 20 units.

The LSK389 dual JFET by Linear Systems (800-359-4023 or 510-490-9160, Fax 510-353-0261, www.linearsystems.com) has been advertised for some time in *audioXpress*. (See ad in this issue on p. 41.) In this ad, the voltage noise density (input reference) at a drain current (I_d) of 3mA is graphed at $1.3\text{nV}/\sqrt{\text{Hz}}$ at 1kHz, and $1.0\text{nV}/\sqrt{\text{Hz}}$ at 20kHz.

I've heard some concern about the consistency of this part. On Sept. 22, 2006, I ordered ten units of the LSK389B-TO71. The "B" category is an I_{dss} (saturation drain current, at zero gate-source voltage) range of 6-12mA. (There's also the "A" with 2.6-6.5mA, and the "C" with 10-20mA.)

Of the ten units, three had a voltage noise density (from now on called simply "noise") ranging between 2 and $3\text{nV}/\sqrt{\text{Hz}}$, while the other seven were between 1.0 and $1.8\text{nV}/\sqrt{\text{Hz}}$. This was a measurement averaged over 1kHz-10kHz. Meanwhile, the datasheet specifies 0.9 typical, 1.9 maximum $\text{nV}/\sqrt{\text{Hz}}$ at 1kHz. From the downsloping noise versus frequency graph shown in Linear Systems' ad, the 1-10kHz averaging I used will show a lower noise voltage than the level at 1kHz. Therefore, the three noisiest of my ten units were significantly out-of-spec.

MY TWO APPLICATIONS, THUS FAR

I used this JFET in my "Low-Noise Measurement Preamp" (*aX*, April '07, p. 26) and in "The LP797 Ultra-Low Distortion Phono Preamp" (*aX*, Sept. '07, p. 6). I was concerned—particularly in the phono preamp—about consistently obtaining the very-low noise performance these JFETs are capable of. Note that the ad is titled "1nV Low Noise Dual JFET."

FAST-FORWARD 14 MONTHS

On November 29, 2007, I ordered 20 units of the LSK389B-TO71. I received them three days later—lot no. JF300-214-4, date code 0551. As **Table 1** shows, the 1-20kHz averaged noise density (with I_d per FET ranging from 4.4-6.8mA; more on this later) was: range 0.941-1.84 $\text{nV}/\sqrt{\text{Hz}}$ (a 3.35dB

range), median 1.1625; average 1.1646. The closeness of the median and average values is indicative of a well-behaved symmetrical distribution (more on this also later). **Table 2** shows the four units left from the 2006 group.

In both my measurement and phono preamps, I use both FET sections in parallel; this divides the noise voltage by $\sqrt{2}$,

TABLE 1 Linear Systems LSK389B-TO71 Dual JFET

Bias and Equivalent Input Noise

Ship date 11/23/07 Date Code 0551 Lot # JF300-214-4 Test Date 12/4/07

Sample No.	V_D V	I^* mA	Noise 20Hz-20kHz nV RMS	Noise "A" nV RMS	Noise 1kHz-20kHz nV RMS	Noise, each FET 1kHz-20kHz Avg. $\text{nV}/\sqrt{\text{Hz}}$
1	5.79	10.55	164.2	134.6	140.4	1.321
2	6.51	8.85	177.8	145.8	146.0	1.384
3	4.31	12.00	138.8	112.9	120.8	1.099
4	4.64	11.51	157.3	129.1	132.5	1.232
5	6.00	9.59	147.3	120.1	127.1	1.171
6	6.18	9.32	126.3	105.1	114.1	1.020
7	3.27	13.48	120.9	99.3	108.4	0.953
8	5.46	10.36	153.2	128.1	132.9	1.237
9	5.32	10.59	144.3	120.1	126.8	1.168
10	5.20	10.73	154.3	128.8	133.2	1.240
11	4.95	11.08	144.1	119.0	125.9	1.157
12	5.40	10.47	144.6	120.4	126.9	1.169
13	4.76	11.36	156.1	131.1	134.8	1.258
14	5.23	10.69	147.9	124.4	130.3	1.207
15	5.14	10.81	141.9	118.5	125.5	1.153
16	3.22	13.55	118.8	97.4	107.4	0.941
17	5.97	9.59	156.2	131.4	135.5	1.266
18	4.77	11.33	136.2	112.9	120.7	1.097
19	3.95	12.51	120.9	99.3	108.8	0.957
20	6.30	9.16	156.0	131.1	135.1	1.262

$V_{\text{dd}} \approx +10\text{V}$ $R_d = 509.8\Omega$ $R_S = 10.00\Omega$ gates grounded both FET sections, paralleled * combined current
10.877mA average 1.1646 $\text{nV}/\sqrt{\text{Hz}}$ average

Table 2 Linear Systems LSK389B-TD071 Dual JFET

Bias and Equivalent Input Noise

Ship date 9/22/06 date code 0447 Lot # JF 300-4-3 Test Date 12/4/07

Sample no.	V_d V	i^* mA	Noise 20Hz-20kHz nV RMS	Noise "A" nV RMS	Noise 1kHz-20kHz nV RMS	Noise, each FET 1kHz-20kHz Avg. $\text{nV}/\sqrt{\text{Hz}}$
1	6.00	8.89	372.5	314.3	278.1	2.795
4	4.26	11.32	296.1	244.7	219.4	2.177
5	4.16	12.95	122.9	95.5	108.8	0.957
7	4.75	11.56	117.6	97.8	108.1	0.951 Tested 12/25/06

$V_{\text{dd}} \approx +10\text{V}$ $R_D = 509.8\Omega$ $R_s = 10.00\Omega$ gates grounded both FET sections, paralleled * combined current

a 3dB reduction. But I also use a 10 Ω common source resistor, for both DC bias stability and a return point for negative feedback (from the AD797 op amp, which is cascaded with the JFET). The thermal noise of this 10 Ω resistor (at +25 $^{\circ}$ C) is 0.4057nV/ $\sqrt{\text{Hz}}$, which is RMS-added (square root of sum of squares) to the input-referred parallel FET noise.

AS-USED NOISE EXAMPLES

I measured the FET's noise voltages in a circuit (Fig. 1) similar to that in the pre-amps, then calculated backwards, RMS subtracting the 10 Ω resistor noise; and then multiplying by $\sqrt{2}$ to obtain the per-FET noise. (Two in parallel will halve the equivalent noise resistance.)

The column in Table 1 labeled "Noise, 1kHz-20kHz" shows the directly measured integrated noise voltage, input referred, over this bandwidth. I followed the test circuit of Fig. 1 (overall gain stabilized at close to 100 by the feedback) with the low-noise measurement preamp, its gain set to 1000. The latter's selectable noise bandwidths were used to measure 20Hz-20kHz, "A" weighted, and 1kHz-20kHz noise levels. Figure 2 shows the overall noise test setup.

Taking the 1kHz-20kHz values and dividing by $\sqrt{19,000\text{Hz}}$, which is 137.84 $\sqrt{\text{Hz}}$, gives the 1kHz-20kHz averaged noise density values, for the configuration of both FETs paralleled and the 10 Ω source resistor. In the 20 unit sample, this ranges from 0.779 to 1.059nV/ $\sqrt{\text{Hz}}$. The equivalent thermal noise resistance (+25 $^{\circ}$ C) ranges from 36.9 to 68.1 Ω . Note that these resistances would be 10 Ω lower if the FET sources were directly grounded.

JFET CURRENT IN TEST CIRCUIT

Note that there's no overall negative feedback (NFB) at DC, because of the coupling cap C2. But the 10 Ω source resistor R4 provides "source degeneration" (analogous to emitter degeneration in a bipolar transistor). This simply means negative feedback, and here the NFB extends to DC, because an increase in FET current generates a greater voltage drop across R4; this biases the sources more positive. This is equivalent to biasing the gates more negative regarding the sources (more negative Vgs). And this decreases the current, which has the

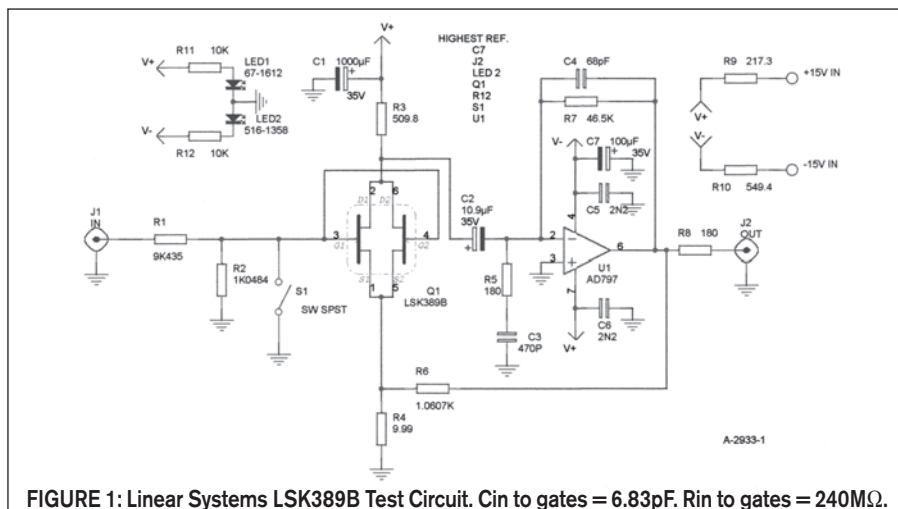


FIGURE 1: Linear Systems LSK389B Test Circuit. Cin to gates = 6.83pF. Rin to gates = 240M Ω .

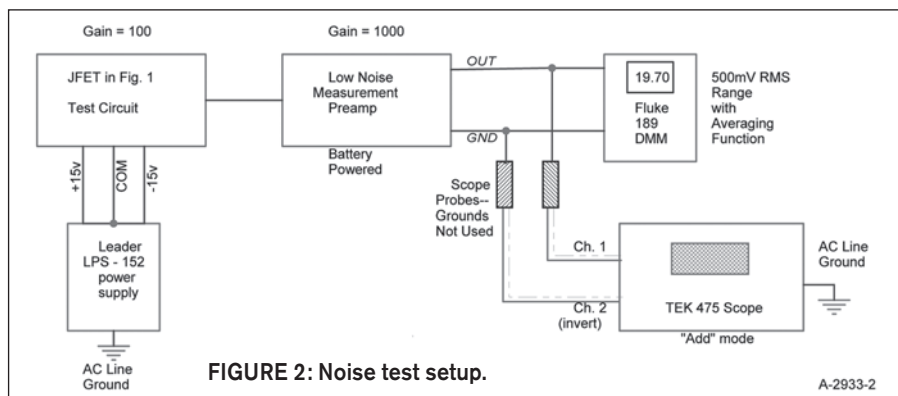


FIGURE 2: Noise test setup.

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effect of opposing the initial assumed attempt to increase the current.

This is useful in stabilizing the bias current (I_D) against unit-to-unit I_{DSS} variations. The LSK389B is specified with an I_{DSS} range of 6-12mA, a 2:1 range. I didn't measure I_{DSS} on these 20 units, but I did on the previous shipment of ten units; I_{DSS} was in-spec on all ten units.

In this test circuit, the FET current ranged from 8.85mA (unit #2) to 13.55mA (unit #16). This is for both FET sections paralleled, so if the two FETs in the package were matched, the range of per-FET currents would be 4.425-6.775mA. I had previously found that increasing the current above 3mA produced very little decrease in noise. Also, with a minimum of +10V at the top of the drain resistor (R3 in the test circuit), the current can be as high as 8mA per FET (16mA package total), while providing a drain voltage (V_D) of at least 2V.

The range of FET currents measured here is comfortably within the range that's allowable for proper, and low

noise, operation in the LP797 phono preamp. In the low-noise measurement preamp, the FET sources (after the 10 Ω resistor) are AC-coupled to ground, the current being highly stabilized with a large source bias resistor from a negative voltage supply. I didn't do this in the phono preamp because I didn't want an electrolytic cap (large value, 1000 μ F required) in the audio path. But for the noise measurement preamp, that's fine.

FURTHER NOTES ON TEST CIRCUIT

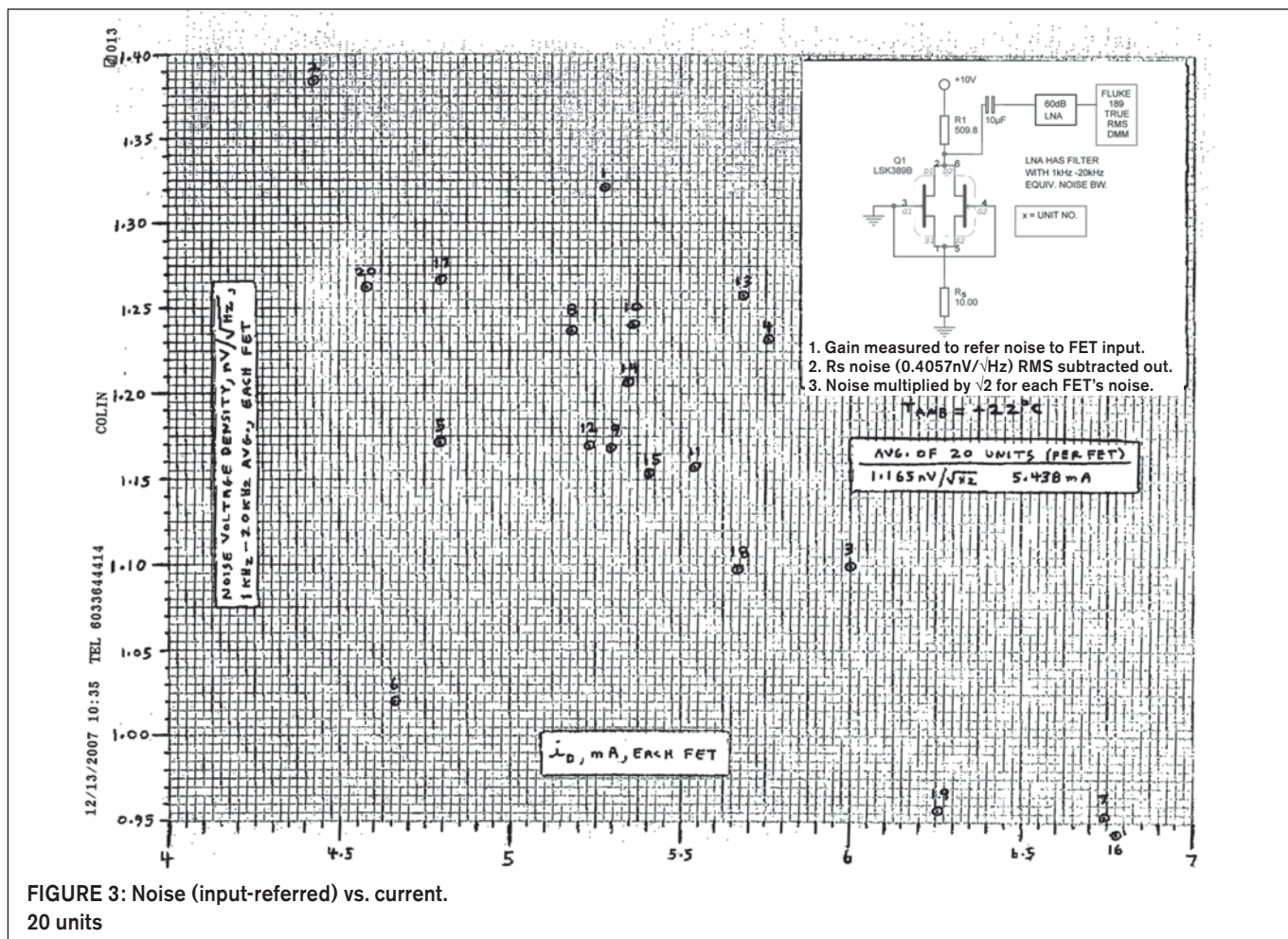
The FET's transconductance is specified at 20ms typical at 3mA (each FET), for the two in parallel that would be 40ms. With this value, the circuit's overall feedback loop gain would be 12.4 (21.9dB). With this loop gain, a 2:1 variation in FET sample transconductance (say, from 14 to 28ms per FET) produces a 5.4% overall closed-loop gain (0.46dB) variation. I observed much less variation. The circuit has a -3dB bandwidth of 1.4Hz-800kHz. Therefore, the noise measurement BW is, for all practi-

cal purposes, determined by the settings of the low-noise measurement preamp following the test circuit.

To achieve freedom from AC line hum corruption, I used a very well-filtered ± 15 V supply. The low-noise measurement preamp and the Fluke 189 true RMS meter are battery powered, avoiding ground loops.

I measured the level of 60/120Hz line ripple by temporarily shorting the Fig. 1 test circuit output, and then connected the measurement preamp output to an oscilloscope. I used the scope's (Tektronix 475) inputs differentially: channel 1 connected to the preamp output, while channel 2 connected to the preamp's ground. Then I used the scope's subtraction ability by inverting channel 2 and setting to "Add" mode.

The scope's ground was not connected to the test circuit, except through the AC line plug's safety ground. The ± 15 V supply was also line grounded. This prevented large common-mode voltages from appearing at the scope



(which could occur if the $\pm 15\text{V}$ supply and/or scope ground were floating).

The observed AC line ripple was at least 20dB below the measured amplified FET noise, even with noise BW extending to 20Hz. Because noise is uncorrelated with AC line ripple (and any other signal, for that matter), an AC ripple level 20dB below the measured noise produces an error of only 0.04dB.

NOISE VERSUS CURRENT CORRELATION GRAPH

Figure 3 contains 20 circled points, numbered with the LSK389B test samples. The horizontal scale is the per-FET operating current (I_o), and the vertical scale is the 1kHz-20kHz noise voltage density of the FET (for each one in the package). If a single FET were thus plotted over a range of currents, the graph would be a downsloping curve, because (within limits) increasing current increases transconductance because the FET's channel resistance becomes lower. This reduces thermal noise voltage.

But in **Fig. 3** the 20 points are fairly randomly scattered. There is a small amount of lower-noise versus higher-current correlation, but obviously other semiconductor physics effects influence the noise level. Notice, though, samples #2 and #16: unit #2 has the highest noise and lowest current, while it's vice versa for #16.

Parts List	
C1	1000 μF
C2	10.9 μF
C3	470P
C4	68pF
C5, C6	2N2
C7	100 μF
J1	IN
J2	OUT
LED1	67-1612
LED2	516-1358
Q1	LSK389B
R1	9K435
R2	1K0484
R3	509.8
R4	9.99
R5, R8	180
R6	1.067K
R7	46.5K
R9	217.3
R10	549.4
R11, R12	10K
S1	Switch SPST
U1	AD797

Noise with Q1 sample(7):

NBW	out μV RMS	in nV RMS	In, average nV/ $\sqrt{\text{Hz}}$	per FET average nV/ $\sqrt{\text{Hz}}$
20kHz-100kHz	22.31	223.1	0.789	
1kHz-100kHz	24.79	247.9	0.788	
1kHz-20kHz	10.81	108.1	0.784	0.951
20Hz-20kHz	11.76	117.6	0.832	1.029
"A"	9.78	97.8		
20Hz-1kHz	4.63	46.3	1.479	2.012

with Q1 sample (7)
1kHz gain = 10.0032
(100.027 regarding Q1 gates)
 $v+ = +10.64\text{V}$, 20.1mA
 $V- = -10.29\text{V}$, 8.6mA
Q1 $V_d = +4.748\text{V}$, $i_o = 11.56\text{mA}^{**}$

Frequency response: -3dB BW = 1.4Hz - 800kHz
1kHz maximum out = 5.14V RMS
 V_o DC = -2.0mV
 ** Both FETs combined
*lot JF300-4-3
date code 0447

SUMMARY FROM TABLE 1 DATA

Here, I'll use the average values of the 20 samples: Integrated noise voltages of the two paralleled FETs with the 10 Ω source resistor are 145.355nV (20Hz-20kHz BW) and 126.655nV (1kHz-20kHz BW). RMS subtracting these gives 71.320nV; this is the integrated (total) noise in a 20Hz-1kHz band. Dividing this by $\sqrt{980\text{Hz}}$ gives a noise density of 2.278nV/ $\sqrt{\text{Hz}}$, averaged over the 20Hz-1kHz band.

Then, RMS subtracting the 0.4057nV/ $\sqrt{\text{Hz}}$ thermal noise of the 10 Ω resistor, and multiplying the result by $\sqrt{2}$ for each FET of the paralleled pair, gives 3.170nV/ $\sqrt{\text{Hz}}$ average noise voltage density in a 20Hz-1kHz band, for a single FET, at an average current of 5.44mA. This is 8.7dB higher than the 1.165nV/ $\sqrt{\text{Hz}}$ figure averaged over the 1kHz-20kHz band. This reflects the normal low-frequency noise increase of semiconductors, FET and bipolar.

However, as the "A" weighting curve shows, the ear is relatively insensitive to low-level, low-frequency sounds.

"A" WEIGHTED NOISE

This data in **Table 1** ranges from 97.4-145.8nV RMS, with an average (of the 20 samples) of 120.47nV RMS. This is for the as-tested configuration of two paralleled FETs and the 10 Ω source resistor. Calculation of the "A" noise for a single FET (without source resistor) is difficult without knowing the resistor's "A" weighted noise, and will not be estimated here.

But because the as-tested configuration has proven to have reliable low-noise performance, the above "A" weighted input-referred noise voltage measurements are relevant. These levels range from

-140.2dBV to -136.7dBV. The average is -138.4dBV.

CONCLUSION

Compared to my 2006 order (in which three of the ten units had out-of-spec noise), all 20 units of my 2007 order were comfortably within the noise spec.

Linear Systems offers low-noise screening for a fee (I don't yet know the cost), but if these 20 units are representative, that shouldn't be necessary.

If you build the LP797 phono preamp, I recommend using the specified DIP sockets for the JFETs. Then, by shorting the preamp inputs and measuring the output noise, you can select JFETs for lowest noise. For small-quantity orders from Linear Systems, it probably costs less to buy a few extra JFETs (so you can select the lowest noise units) than to pay Linear Systems for noise screening.

At this point, I'd say that the LSK389B is an excellent product, very suitable for low-noise audio applications. *aX*

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Tri-Way Low Voltage Supply, Pt. 2

To show the versatility of this supply, the author puts the Tri-Way board through its paces.

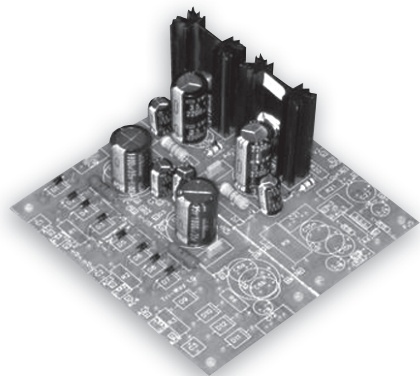


PHOTO 1: The Tri-Way board, with most of the parts installed for use in a compressor.

I'm going to go through three very different designs using the Tri-Way. I'll do the first one the long way, showing all the math; I'll whiz past the other two, stopping only to point out how they differ from the first. To keep the equations straight, I've numbered them.

Before I start, I must remind readers that a good power supply design should not only work under normal conditions, but also should be built to keep working under worst-case conditions. These might include low or high line voltages, components at the limits of their specified tolerances, and high operating temperatures. (Ever work a festival in 100° heat with the mixer in the sun? I have.) I design for multiple worst-case scenarios, and my supplies very seldom fail or misbehave.

EXAMPLE 1: THE COMPRESSOR

This was the *ur*-circuit, the one for which the board was conceived. It's pretty standard stuff: a dual-channel compressor which requires a maximum of 0.25A from $\pm 15V$ supplies. (Seem like a lot? The compressor itself pulls only 65mA; the rest is to light the metering LEDs.) I'll refer to part numbers for the + supply; the - supply will use the same values.

The first step is to specify a few givens. I'm using two separate transformer windings, so I'll use D1-D8 to produce two bridge rectifiers. R4 will be 1 Ω ; C4 will tentatively be 1000 μF , and C13 will be

3300 μF . (If one cap is bigger, I like to put it in the second stage of filtering to minimize turn-on current draw.)

An LM317 drops a nominal reference voltage of 1.25V between its output and adjust terminals. If R16 = 200 Ω , then it will draw a total nominal reference current of:

$$1. I_{ref} = (V_{ref}/R16) + I_{adj} = (1.25/200) + 0.00005 = 0.0063A \text{ (6.3mA)}$$

(What's I_{adj} ? That's current coming out of the adjust terminal, as specified in the datasheet.)

For an output voltage of 15V, the adjust terminal should be at:

$$2. V_{adj} = V_{out} - V_{ref} = 15 - 1.25 = 13.75V$$

That voltage comes from dropping the reference current through R13, which should be:

$$3. R13 = V_{adj}/I_{ref} = 13.75/.0063 = 2183\Omega$$

The nearest E96 value is 2.21k. If the current is 6.3mA, the power dissipated in R13 will be:

$$4. P = (I_{ref})^2 * R13 = (.0063)^2 * 2210 = .088W$$

A quarter-watt metal film resistor should be able to handle that without problems. So far, so good.

NOT QUITE NOMINAL

In the real world, though, not everything runs at nominal ratings. Resistors have tolerances, and so do regulators' reference supplies. I've specified 1% resistors; if R16 is 1% low, then its real resistance will be 198 Ω . If the regulator, meanwhile, is running at the high ends of its V_{ref} and adjust-terminal current tolerances, then V_{ref} will be 1.3V, and the reference current will be:

$$5. I_{ref-hi} = (V_{ref-hi}/R16-lo) + I_{adj} = (1.3/198) + 0.0001 = 0.00667A \text{ (6.67mA)}$$

If, meanwhile, R13 is 1% high, its real resistance will be 2232 Ω . Pass the above current through it and you get:

$$6. V_{reg-hi} = V_{ref-hi} + (I_{ref-hi} * R16-hi) = 1.3 + (.00667 * 2232) = 16.2V$$

And R13 will dissipate, in the worst case:

$$7. P = (I_{ref-hi})^2 * R13_{hi} = .0992W$$

Still pretty reasonable for a quarter-watt resistor.

So under the worst combined conditions, the regulator will be putting out a bit over 16V. This won't hurt the audio circuits any, but it'll affect the design process.

What happens when everything goes in the opposite direction— V_{ref} is at its low limit of 1.2V, R16 is at its high limit of 202 Ω , and R16 is at its low limit of 2188 Ω ? (There's no minimum spec for adjust-terminal current, so I'll use the "typical" figure.) Then (sparing the math):

$$8. I_{ref-lo} = 5.99mA$$

$$9. V_{reg-lo} = 14.3V$$

I'll use those numbers later.

OK, I've presented the worst case for regulator current draw; it'll be 6.67mA. Since the specified load current is 250mA, total current drawn from the supply will be 256.67mA. Those are more significant figures than necessary; let's call it 257mA, or 0.257A.

Next you need to figure how much voltage the regulator needs across it to keep regulating properly, known as the "dropout voltage." That comes from a chart in the spec sheet, and, interpolating a little, at this current draw it wants at least 1.8V at a chilly temperature of 10° C. (I've worked gigs like that, too.)

If the dropout voltage is 1.8V, then you'll need at least:

$$10. V_{unreg} = V_{reg-hi} + V_{dropout} = 16.2 + 1.8 = 18.0V \text{ to maintain regulation under worst-case conditions. I normally add a fudge factor of 0.25V to allow for supply ripple, so I really want:}$$

11. $V_{unreg} = 18.25V$

That's the bottom-line number: the raw supply must feed the regulator at least 18.25V.

COLD AND RAW

For the raw supply design, I'll start with the transformer, choosing one with 18V AC. (That's the voltage of each winding; remember that I'm using separate windings for the two halves of the supply.) Nominally 18V AC, that is.

Unfortunately, this is another spot where things don't stay nominal. In the US, fluctuations of $\pm 10\%$ in the volts coming out of the wall are pretty typical, and in most of the country summertime voltages are near the lower end of that tolerance.

This means that I must design the power supply for a transformer that's really putting out, not 18V AC, but 16.2V AC. This runs through the full-wave diode bridge and into a filter capacitor. I make the (deliberately pessimistic) assumption that I'll drop 2.5V in the diodes, and, of course, I'll drop some voltage, V_{drop1} , in $R4$:

12. $V_{drop1} = R4 * I_{total} = 0.257V$

The peak voltage produced by a sine wave is $V_{ac} * \sqrt{2}$, so taking the losses (and low line voltage) into account,

13. $V_{pk-lo} = (V_{ac-lo} * \sqrt{2}) - 2.5 - V_{drop1} = (16.2 * 1.414) - 2.5 - 0.257 = 20.15V$

We've specified that $C4 = 1000\mu F$, and $C7$ is $100\mu F$, so the total input-section capacitance is $1100\mu F$, give or take a bit. How much ripple will there be? This will depend on several factors, but a good rule-of-thumb calculation, when C is expressed in μF and I in amperes, is:

14. $V_{ripple1} = 2400 * I_{total} / C = 0.561V$ AC

A respectably low number. How to compute the DC voltage? Another rule of thumb calculation, again with capacitance in μF and current in amps:

15. $V_{DC} = V_{pk} * (1 - ((4170 * I_{total}) / (V_{pk} * C)))$

For this design, at -10% wall voltage, that works out to:

16. $V_{DC-lo} = 20.15 * (1 - ((4170 * 0.257) / (20.15 * 1100))) = 19.18V$ DC

I said earlier that the minimum voltage the regulator wanted to see, V_{unreg} , was 18.25V DC. So:

17. $V_{drop2} = V_{DC-lo} - V_{unreg} = 19.18 - 18.25 = 0.93V$

18. $R7 = V_{drop2} / I_{total} = 0.93 / 0.257 =$

3.602Ω

The first choice would seem to be a 3.6Ω resistor, but what about its tolerance? If it's 5% high, that's 3.78Ω ; the voltage drop will be 0.971V. That would make $V_{unreg} = 18.21V$. Hmm... the difference is only 0.05V under worst-case conditions. I think it's safe to specify a 3.6Ω 5% resistor; the power through that resistor will be:

19. $P = (I_{total})^2 * R7 = (0.257)^2 * 3.6 = 0.238W$

A 1W resistor will work, and so will a 2W. Being cautious, I'll go for 2W.

What about the ripple? $R7$ and the capacitors which follow it ($3400\mu F$, more or less) form a low-pass filter; its cutoff frequency F_c is:

20. $F_c = 1 / (2 * \pi * R7 * C13 || C16) = 13.0Hz$

The ripple will be mainly 120Hz (on a 60Hz electrical grid, anyway), and will be reduced by approximately:

21. $F_c / F_{ripple} = 13.0 / 120 = 0.108x$

Since the ripple at the first capacitors was 0.561V AC, the ripple on the second capacitors will be:

22. $V_{ripple2} = 0.561 * 0.108 = 0.061V$ AC

That's plenty low enough to ignore.

DISSIPATION IS RUINATION

The last thing to calculate is the power the regulator will dissipate, worst-case. This happens when the regulated voltage is at its lowest possible value, V_{reg-lo} , while the wall voltage is at its hottest. We already know $V_{reg-lo} = 14.3V$ and $V_{drop1} = 0.257V$ DC for this circuit; if the wall voltage is 10% high, the transformer's secondary will be putting out $V_{AC-hi} = 19.8V$ AC. So:

23. $V_{pk-hi} = (V_{AC-hi} * \sqrt{2}) - 2.5 - V_{drop1} = (19.8 * 1.414) - 2.5 - 0.257 = 25.25V_{pk}$

24. $V_{DC-hi} = V_{pk-hi} * (1 - ((4170 * I_{total}) / (V_{pk-hi} * C))) = 25.25 * (1 - ((4170 * 0.257) / (25.25 * 1100))) = 24.27V$ DC

25. $V_{unreg-hi} = V_{DC-hi} - (V_{drop2}) = 23.34V$ DC

So the voltage across the regulator will be:

26. $V_{regdrop} = V_{unreg-hi} - V_{reg-lo} = 23.34 - 14.3 = 9.04V$ DC

And the power dissipated by the regulator, worst-case, will be:

27. $P = V_{regdrop} * I_{total} = 9.04 * 0.257 = 2.32W$

With a Wakefield 637-10ABP heatsink, the thermal rise should be about $37^\circ C$;

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CUSTOM WORK

assuming an ambient temperature of 40° C, the chip temperature should be 77° C. That should be fine; the LM317T is rated to a chip temperature of 125° C.

The compressor will be attached to the regulator all the time; it draws a minimum current of 65mA. This is enough to keep the regulator regulating happily, so no additional load resistor is required; I'll leave out R19.

EXAMPLE 2: THE FILAMENTS

What about using Supply 2 to produce 6.3V DC at 1.2A, which will power four 12AX7s (wired for 6.3V) or two 6SN7s? To avoid tedium, I won't show most of my work this time.

Looking at the regulator, if $R_{o-a} = 200\Omega$ again, then $R_{adj} = 800\Omega$ (nearest E24 is 806 Ω). The regulator's dropout voltage (from the datasheet) is 2.25V, so under worst-case conditions the unregulated voltage at the regulator's input needs to be (mumble, mumble) 9.23V DC.

Using the same filter capacitors (1100 μ F at the input, 3400 μ F in stage 2 of the filter) and a 12V AC wall-wart transformer, $V_{pk-lo} = 11.57$ V DC, ripple voltage will be 2.63V AC, and the volt-

age at the first filter capacitor will be 6.99V DC.

TILT! That's not going to work; the voltage at the first cap is less than the voltage regulator needs to see at its input.

What to do? A higher-voltage transformer would work, of course, but I'm already using a 12V AC transformer to get 6.3V DC; higher seems extravagant. Instead, the best approach is to increase the filter capacitance.

What if I make both big caps 4700 μ F (plus 100 μ F from the little electrolytics in parallel)? Now the voltage at the first filter cap is 10.52V DC, and the ripple is about 0.6V AC. That's better; doing the arithmetic for the dropping resistor R8, it should theoretically be 1.07 Ω ; 1.0 Ω is the next lowest value.

Putting that much current through a 1.0 Ω resistor will dissipate 1.21W, so R8 needs to be a hefty one; that's why I left room on the board for 5W dropping resistors. (Come to think of it, I should probably juice up R5, too; there's *just* enough room on the board for a 3W Panasonic metal oxide resistor.)

With these values the ripple at the regulator's input will now be 0.17V AC,

a level I can live with.

What about heat? At high line voltage the worst-case regulator dissipation will be (more mumbles) a bit over 8W. Uh-oh; with the little heatsink I used in Example 1, the temperature rise will be 128° C, and if the ambient temperature is 40° C the chip will be sizzling at 168° C.

This clearly calls for a bigger heatsink. The Wakefield series I specified for this board goes up to 2.5" high, which provides significantly lower thermal resistance, but with this kind of power I think it's time to punt by putting the regulators off-board. The LM317K and LM337K come in TO3 packages (available from Mouser), and something like a Wakefield 680-125A heatsink mounted on short spacers would dissipate 8W handily. I could mount this heatsink on the outside of the cabinet, provided I used a plastic insulating cover on the regulator itself, since there is voltage on the case.

EXAMPLE 3: THE PHANTOM WALKS!

For the third example, I'm going to use Supply 3 for +48V DC phantom power

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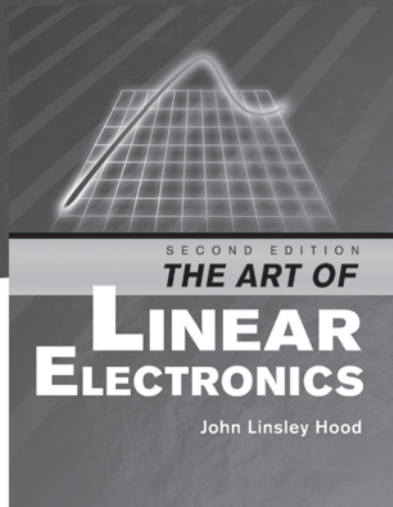


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to run a couple of microphones. Phantom-powered mikes draw varying currents, but the worst case would be dead shorts, which would draw 14mA apiece. (Not that you would care about the regulation if the mike is a dead short.) I'll also want to draw another 15mA to pre-load the regulator (I'll get to that in a bit), for a total of 43mA.

With a voltage this high, I'll use a TL783 regulator instead of an LM317, and that changes a few things. The TL783 has higher dropout voltages and adjust-terminal currents, and its nominal reference voltage is 1.27V instead of 1.25V. (Oddly, the tolerance limits are still the same, 1.2V and 1.3V.) Texas Instruments recommends that the adjust terminal not be bypassed with a capacitor, so I can leave out C24.

Setting the reference resistor R18 to 1k, the nominal current is:

$$28. I_{ref} = (V_{ref}/R18) + I_{adjust} = (1.27/1000) + 0.000083 = 0.00135A \text{ (1.35mA)}$$

To get 48V DC from the output I'll need:

$$29. V_{adj} = V_{reg} - V_{ref} = 48 - 1.27V = 46.73V \text{ DC}$$

And the adjusting resistor will be:

$$30. R15 = 46.73/0.00135 = 35,071\Omega$$

The nearest E96 value is 34.8k. Maximum current through the adjust string will be 1.42mA, and maximum power dissipation in R15 will be 0.071W, so a quarter-watt resistor will be adequate.

The dropout voltage for a TL783 drawing 43mA is roughly 5V; making the usual allowance for regulator and resistor tolerances (and my 0.25V fudge factor), the voltage at the regulator's input should be at least 57.0V DC.

I'll begin with a 48V AC transformer, $C6 = C15 = 470\mu F$ and $C9 = C18 = 33\mu F$. Doing the usual arithmetic, the DC voltage on the first stage of the filter will be +58.2V DC under low-line conditions. Dropping this to 57.0V DC would require a 28.5Ω resistor; the next-lowest value is 28.0Ω . That'll dissipate 0.055W; a quarter-watt resistor will do. (The nearest E24 value is 27Ω .)

The rest of the design is pretty straightforward; maximum voltage under high line conditions would be 72.1Vpk, and it's prudent to add another 10% fudge factor for a lightly loaded transformer. That means the maximum voltage on the first capacitor could be about +79V DC, so the caps should be rated at

100V working voltage. With the capacitors I specified (the largest 100V units that would fit in the space), ripple voltage at the regulator input will be about 20mV, which is negligible.

What about idle current? Unlike the previous examples, this is a circuit where the current drawn by the load will vary a lot. Some microphones, such as Neumann KM 84s, pull as little as 0.5mA from the phantom supply, while others draw as much as 6mA. Ideally, the regulator should keep regulating under all circumstances, even when there's no microphone connected.

Texas Instruments specifies a minimum current draw of 15mA for the TL783 to perform properly. Under worst-case conditions ($V_{ref} = 1.2V$ DC, $R18 = 1010\Omega$), the adjustment string will draw 1.27mA; to draw 15mA total the load resistor R21 will need to pull an additional 13.7mA. The worst-case low regulated voltage will be 45.0V DC, so a 3278Ω resistor will do the job. The next-lowest E24 value is 3k, which actually draws 15mA under nominal conditions.

How much heat will R21 dissipate? If the regulator is at its highest possible voltage (50.1V DC) and the resistor is at the low limits of its tolerance (2850Ω), then:

$$31. P_{diss} = ((50.1)^2)/2850 = 0.88W$$

A 2W resistor is the minimum for reliable performance, and since there's room on the board for a 3W Panasonic metal-oxide resistor, that's what I'd use.

Finally, the regulator will dissipate 1.14W under worst-case circumstances, which the smallest specified heatsink will handle easily.

BRIEF DIVERSIONS

It might occur to you that the above calculations are the sort of thing that spreadsheets do well. That's correct; I've designed a spreadsheet which does them, but it's not quite ready for prime time yet. Watch this space.

Speaking of spreadsheets, I've produced an Excel file with parts lists for the three examples above; it's posted on the magazine's website, www.audioXpress.com. The lists include the power transformers and all the parts on the boards except mechanical bits (insulating washers, standoffs, pins, and so on).

WINDING UP

The Tri-Way board can be useful for a variety of projects; I've left open as many options as possible, in the hope that readers will find new ways to make it sit up and do tricks.

As I said when I wrote up the Gamp supply, I no longer fabricate my own circuit boards; the chemicals are more toxic and messy than I want to deal with, and I can at least hope that commercial manufacturers have better facilities to handle and dispose of the stuff safely. You may not agree, so I've made a board layout that's posted on the *audioXpress* website; it's as close to the fabricated layout as my graphics programs will allow.

Or, of course, you could buy your board(s) from me; like the Gamp boards, the price will be US \$24 per unit (there'll be a small discount for quantities >2), plus shipping/handling. You can e-mail me at pstamler@pobox.com, and no, I don't expect to get rich from this project either.

Many thanks to Cory King of the Webster University Audio Construction Club. I hope you have as much fun using this board as I had designing it! *aX*



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When the compact disc was under development, Sony and Philips built an optional treble pre-emphasis curve into the Red Book specifications for the format. Initially the CD was intended to be a 14-bit medium, which pushed the limits of storage and signal processing at that time. By the time the CD was actually introduced to the public in 1982, the resolution had been increased to 16 bits. Recordings made with a resolution of 14 bits had very poor linearity at low signal levels, particularly in the high frequencies, and even the 16-bit converters in the 1980s had shortcomings in this regard.

The Red Book *pre-emphasis* specification applied a high-frequency boost ahead of the analog-to-digital converters, which ensured that low-level, high frequencies would be recorded in a more linear fashion. This high-frequency boost was applied with an analog equalization circuit, because it needed to be applied *prior* to A-to-D conversion in order to overcome the limitations of the converters. A complementary *de-emphasis* equalization curve was applied in playback, usually with an analog filter *after* digital-to-analog conversion. Because low level, high frequencies remained boosted during the D-to-A conversion process, linearity problems in those converters were also reduced.

By the early 1990s many manufacturers of digital conversion chips were implementing de-emphasis in the digital domain, usually in the playback digital filters. At that point, low-level linearity of D-to-A converter chips had im-

proved to the point where it really was not necessary to keep the signal pre-emphasized during the conversion process.

The Red Book pre-emphasis/de-emphasis standard has often been referred to as a noise-reduction system, but this is a simplistic and incomplete explanation. True, the high-frequency boost in record and complementary cut in playback does reduce quantization noise, but this was probably not the greatest sonic benefit. The greatest benefit was the improved high-frequency linearity at low signal levels. By the early 1990s, the entire process had become a moot point due to improved linearity of both A/D and D/A converters.

Very few CD manufacturers actually implemented the Red Book pre-emphasis standard. Nearly all of the CDs I own with pre-emphasis are discs manufactured in Japan by Denon in the 1980s, either for their own label (the entire Eliahu Inbal/Frankfurt Radio Sym-

phony Mahler cycle, for example), or discs they made for other labels (Music and Arts Programs of America had many of their early CDs manufactured in Japan by Denon). In recent years, many manufacturers of CD players and outboard D/A converters have stopped implementing playback de-emphasis—the Monarchy M24 I reviewed in the Oct. 2007 *aX* is a case in point. This is a problem for those of us who have been collecting CDs since the 1980s. I believe that all CD playback hardware should be backwards compatible.

RED BOOK STANDARD

The Red Book pre-emphasis curve is shown in **Fig. 1**. Time constants are specified as 50 μ S and 15 μ S, corresponding to frequencies of 3183Hz and 10610Hz. Relative to the low end of the spectrum, the +3dB point for the boost is 3183Hz, with the boost shelving at 10610Hz.

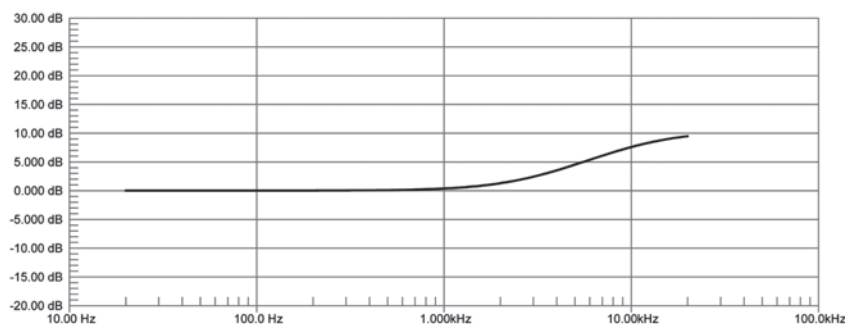


FIGURE 1: The Red Book pre-emphasis curve specifies time constants of 50 μ S and 15 μ S. The +3dB point for the high-frequency boost is 3183Hz, shelving at 10610Hz. Maximum boost is +9.49dB at 20kHz.

Only a handful of test CDs have been made with tracks for checking de-emphasis in playback. I have used *Hi-Fi News and Record Review Test CD II* (HFN15), which has pre-emphasis tones at 1kHz, 4kHz, and 16kHz. Chuck Hansen has used the *CBS Test Disc* (CBS-1), which has tones at 125Hz, 1kHz, 4kHz, 10kHz, and 16kHz. These discs are adequate for determining whether de-emphasis has been implemented, but they don't have enough tones to give meaningful data about the accuracy of the de-emphasis. Some discs, such as the *Pierre Verany Digital Test* (DV 788031-32) and the *Denon Audio Technical CD* (38C39-7147), have sweeps with pre-emphasis from 20Hz to 20kHz, but these are only useful if you have measurement equipment that can be synchronized with a sweep generator.

ROLLING YOUR OWN

I have always been frustrated with my inability to measure the accuracy of playback de-emphasis, so I decid-

ed to take matters into my own hands and make my own test CD. To do so, I needed a precise model of the Red Book pre-emphasis curve, which I produced using CircuitMaker 2000. **Figure 2** shows the simulation model. $T = RC$, so the $50\mu\text{s}$ time constant is produced by $(R1 + R2) * C1$, and the $15\mu\text{s}$ time constant is produced by $R1 * C1$. Scaling for V_{cVs1} is set to the value of $R2$, and I_{cVs1} is set to unity. $R3$ is a load resistor for I_{cVs1} , arbitrarily set to 100k (this value is unimportant).

I used this simulation model to generate the curve shown in **Fig. 1**. CircuitMaker 2000, like most schematic capture programs with simulation, will allow you to put cursors on the generated graph and measure one level relative to another. **Table 1** shows the measurements I made on the simulation in **Fig. 1**. The "FREQ." column lists the frequencies I decided were appropriate for a truly useful de-emphasis test CD.

I set the first cursor at exactly 20Hz, then moved the second cursor to the remaining frequencies and noted the dif-

ference relative to 20Hz. Those results are plotted in the second column. The maximum Red Book high-frequency boost in record is about 9.5dB. To avoid clipping, the 20Hz reference should be recorded at a level of -10dB, which is why that column is labeled "LEVEL REF TO -10dB."

The third column gives those levels relative to 0dB, in which case 20Hz is now at -10dB. Finally, the fourth column lists the track numbers for each test tone. The CD I made actually has two sets of tracks: 1 through 28 are recorded without pre-emphasis—in other words, flat—at -10dB. Tracks 29 through 56 duplicate the previous tones, but with pre-emphasis applied according to the levels indicated in **Table 1**.

I produced the test CD with Sony Creative Software's Sound Forge version 9.0, the digital audio editor I use on an almost daily basis (www.sonycreativesoftware.com). Sound Forge has a Simple Synthesis function that allows you to generate sine waves at any frequency, for any duration you specify (**Fig. 3**). I used Simple Synthesis to gen-

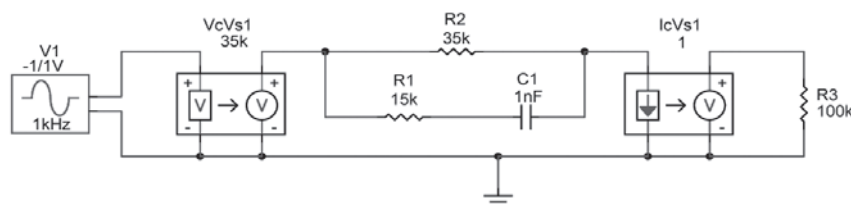


FIGURE 2: Simulation model for generating the Red Book pre-emphasis curve. The $50\mu\text{s}$ time constant is the product of $(R1 + R2) * C1$; $15\mu\text{s}$ is equal to $R1 * C1$.

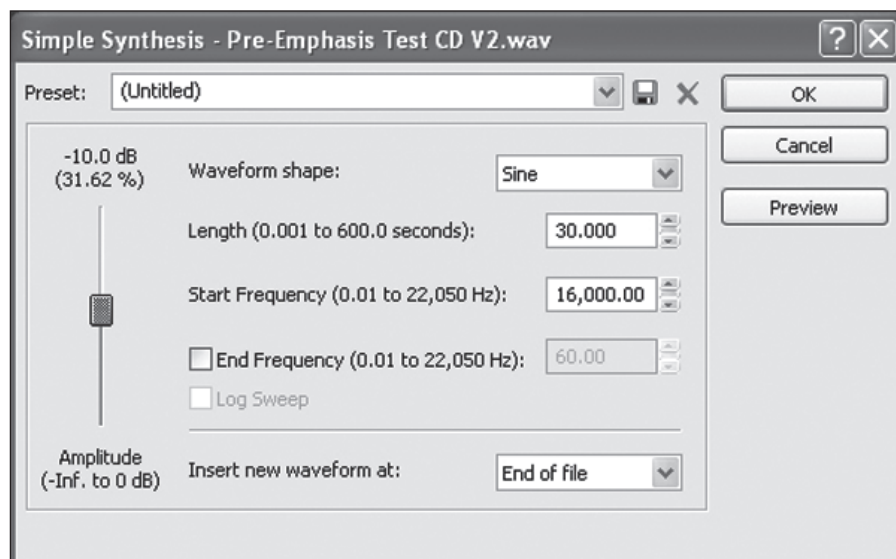


FIGURE 3: The Simple Synthesis module in Sony's Sound Forge will generate a sine wave at any frequency, for any duration, at any level.

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erate each tone in **Table 1** for a length of 30 seconds, at a level of exactly -10.0dB, for a total of 28 tracks. I put 4 seconds of silence at the end of each tone, plus 1 minute of silence at the end of the last track. Then, I marked the entire file and performed a simple copy and paste, duplicating all 28 tracks again. These duplicated tracks—29 through 56—are the ones that will have pre-emphasis added.

Sound Forge has two different “Volume” functions that can boost or cut volume by any level you choose. The one under the “Process” menu allows adjustments in 0.01dB increments, which is a bit too coarse for the lowest frequencies. There’s another “Volume” plug-in, under the “FX Favorites” menu, that allows adjustments in increments of 0.001dB, which is the one I used (**Fig. 4**). You’ll need to highlight each tone from track 29 through track 56 and boost that tone by the amount indicated in the “LEVEL REF TO -10dB” col-

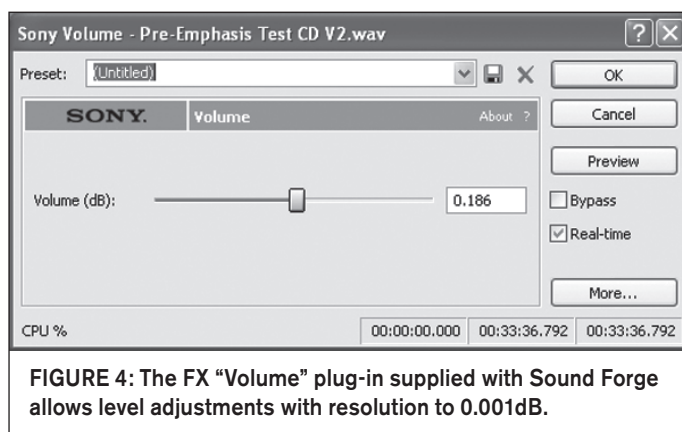


FIGURE 4: The FX “Volume” plug-in supplied with Sound Forge allows level adjustments with resolution to 0.001dB.

umn. When you finish, if you zoom all the way out, your Sound Forge screen will look like **Fig. 5**.

You can then use Sound Forge’s “Statistics” function (under the “Tools” menu) to check the level of each tone relative to 0dB (**Fig. 6**). Highlight nearly all of the tone in each track, one at a time, but don’t highlight the silence on either side. Then click on “Statistics” and look at either “Maximum sample value” or “Minimum sample value.” They should be the same and should also match the level given in column 3 of **Table 1** for that particular tone.

Occasionally, I found that the tones were not at the level they should be. In this case, I highlighted the entire tone and normalized it to a level of -10.0dB, then repeated the above procedures. It always worked on the second attempt.

Sound Forge 9.0 comes with a Mastering Equalizer plug-in made by iZotope, which includes pre-emphasis and de-emphasis curves.

However, the algorithms don’t seem to have the precision I get from manually adjusting each tone according to the simulation. As an example, **Table 1**, column 3 says that 11kHz should be at -2.05dB; if I pre-emphasize all 28 tracks using iZotope’s plug-in, 11kHz is at -1.65dB.

CD AUTHORIZING

With Sound Forge, you first produce tracks by putting markers in the file where you want them. You also need a marker at the end of the file. When you’re all done, right-click on the Re-



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Table 1 Pre-Emphasis Simulation Levels

FREQ.	LEVEL REF TO -10dB	LEVEL REF TO 0dB	Tracks
20Hz	0	-10.000	1, 29
50Hz	0.001	-9.999	2, 30
100Hz	0.004	-9.996	3, 31
500Hz	0.096	-9.904	4, 32
600Hz	0.138	-9.862	5, 33
700Hz	0.186	-9.814	6, 34
800Hz	0.241	-9.759	7, 35
900Hz	0.303	-9.697	8, 36
1kHz	0.37	-9.630	9, 37
2kHz	1.29	-8.710	10, 38
3kHz	2.43	-7.570	11, 39
4kHz	3.54	-6.460	12, 40
5kHz	4.53	-5.470	13, 41
6kHz	5.38	-4.620	14, 42
7kHz	6.1	-3.900	15, 43
8kHz	6.69	-3.310	16, 44
9kHz	7.19	-2.810	17, 45
10kHz	7.61	-2.390	18, 46
11kHz	7.95	-2.050	19, 47
12kHz	8.25	-1.750	20, 48
13kHz	8.49	-1.510	21, 49
14kHz	8.71	-1.290	22, 50
15kHz	8.89	-1.110	23, 51
16kHz	9.05	-0.950	24, 52
17kHz	9.18	-0.820	25, 53
18kHz	9.3	-0.700	26, 54
19kHz	9.4	-0.600	27, 55
20kHz	9.49	-0.510	28, 56

Pre-emphasis levels based on the circuit simulation of **Figs. 1** and **2**. The test CD has 28 tracks recorded flat at -10dB, and 28 more recorded with the 50µS/15µS Red Book pre-emphasis curve.

gions List and change the markers to regions. Sound Forge comes bundled with CD Architect 5.2, a professional “disc at once” CD authoring program that provides full Red Book PQ encoding and editing. CD Architect makes tracks from the regions you’ve already produced. Open the .WAV file with CD Architect’s “Open Media” function, and save the file as a CD Architect (.CDP) project. Click on “Track List” in the lower window (Fig. 7).

You should see all 56 tracks you’ve produced in Sound Forge. On the right, you’ll see a column called “Emph.”—there’s a box for each track, all unchecked. Check each box from track 29 through track 56. This will tell CD Architect to write the pre-emphasis flag for each of those tracks. This flag tells your CD player or outboard DAC to turn on the de-emphasis circuit.

It’s important to understand that there are two processes involved in producing a CD, or individual CD tracks, with pre-emphasis. The first step is to apply the correct high-frequency boost to your .WAV file. I did this with Sound Forge, one track at a time, adjusting the level of each tone according to my circuit simulation.

But, altering the frequency response

according to the Red Book 50µs/15µs time constants won’t tell your CD player to apply the correct de-emphasis. The pre-emphasis flag must be recorded on the disc by your CD authoring program. CD Architect allows you to add a pre-emphasis flag to each track, individually.

FINISHED TEST CD

Once you’ve finished adding the emphasis flag to tracks 29 through 56, save your changes in the .CDP file and burn the CD. You’ll now have a de-emphasis test CD that is far more useful than any published test discs that I’ve seen. To check your CD player or outboard DAC, first play tracks 1 through 28, monitoring the player or DAC output on an AC voltmeter with a dB scale. These tracks should show a flat frequency response from 20Hz to 20kHz. Now do the same with tracks 29 through 56.

If your player or DAC supports the Red Book de-emphasis specification, you should also get the same flat frequency response you got with tracks 1 through 28. If you get a response that rises with frequency, with 20kHz at about +9.5dB relative to 20Hz, your player or DAC doesn’t support de-emphasis. For the most accurate measurements, I use a digital AC voltmeter with resolution to

1mV. After making the voltage measurements, I convert them to dB in a spreadsheet using the formula $\text{dB} = 20 \log E1/E2$.

De-emphasis errors are similar to RIAA equalization errors in phono preamps. Because errors are sometimes spread across an octave or more, even errors of a few tenths of a dB can be audible, if they occur between 1 and 5kHz, where the ear is especially sensitive. The de-emphasis graph Chuck Hansen prepared for the Benchmark DAC1 USB review (Fig. 2 in his review, published in

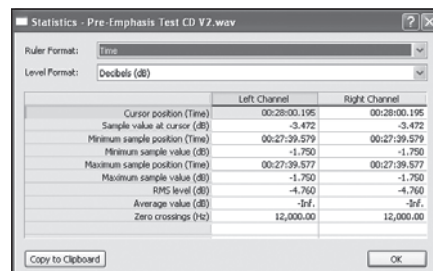


FIGURE 6: The Sound Forge Statistics module allows verification of the level of each tone relative to 0dB. Use the minimum or maximum sample values, which should be the same.

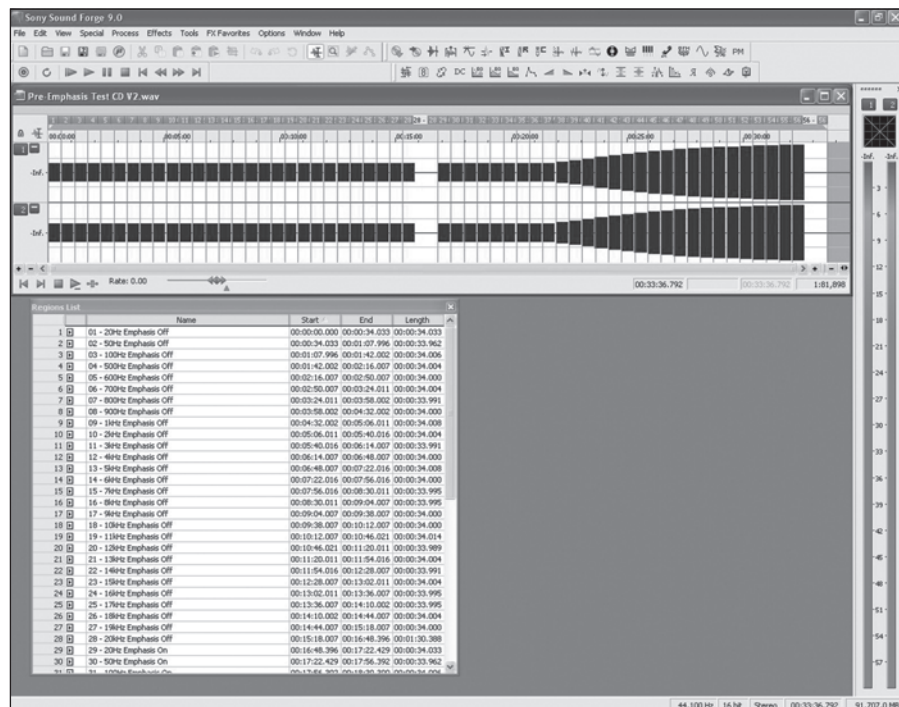


FIGURE 5: The Sound Forge screen with the completed test CD. Tracks 1–28 are recorded flat; tracks 29–56 have been adjusted in level to correspond to the Red Book pre-emphasis standard.

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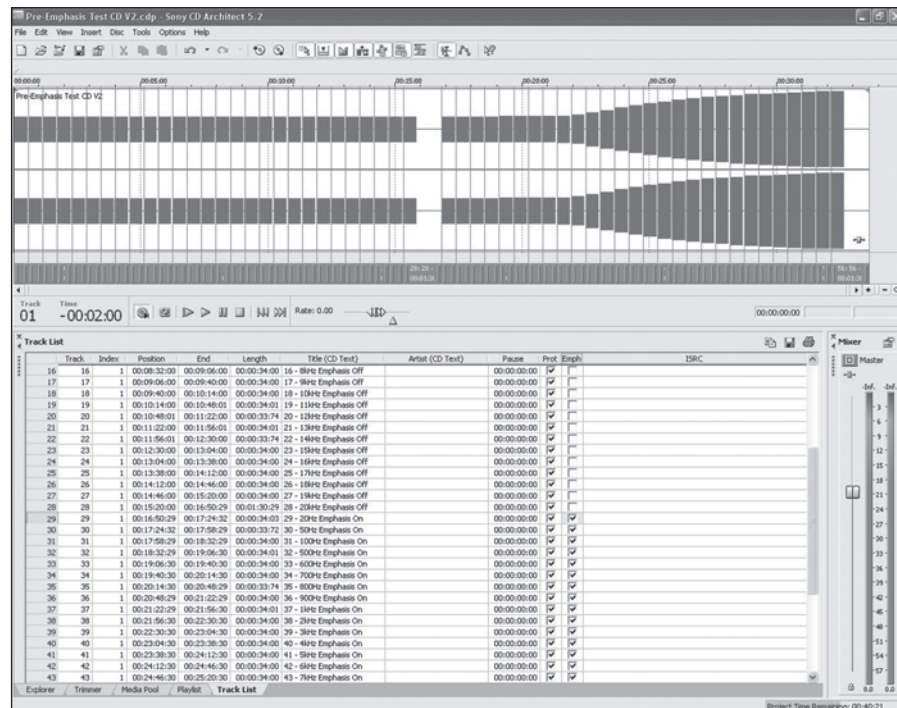


FIGURE 7: Sony's CD Architect 5.2 allows full Red Book PQ encoding and editing, including the addition of a pre-emphasis flag to any track.

Jan. '09 issue, p. 32), was produced from data I measured with this test CD, and shows excellent accuracy, $\pm 0.09\text{dB}$, 20Hz to 20kHz.

The worst errors are at the top of the spectrum, where they are unlikely to be audible. I also measured the DAC1 USB using the *Hi-Fi News and Record Review* test disc mentioned previously. The *HFNR* disc showed 16kHz to be at $+0.399\text{dB}$ (left) and $+0.451\text{dB}$ (right), relative to 1kHz. The measurements with my test CD showed 16kHz to be at $+0.002\text{dB}$ (left) and 0.058dB (right), relative to 1kHz.

The nearly half a dB error shown by the *HFNR* test CD is suspicious. In the datasheet for the AD1853—the DAC chip used in the Benchmark DAC1 USB—Analog Devices specifies the de-emphasis error as $\pm 0.1\text{dB}$. The measurements made with my test CD show the Benchmark DAC to be well within Analog Devices' tolerance. After sending Chuck Hansen a copy of my test CD, he said he would use it for his de-emphasis measurements. I thank him for his feedback and words of encouragement during the preparation of the test CD and this article. I hope other readers will find this disc as useful as we have.

The De-Emphasis Test CD can be purchased from the author for \$20 each including Media Mail shipping in the US. Send a check or money order payable to Gary Galo, 211 May Road, Potsdam, NY 13676.

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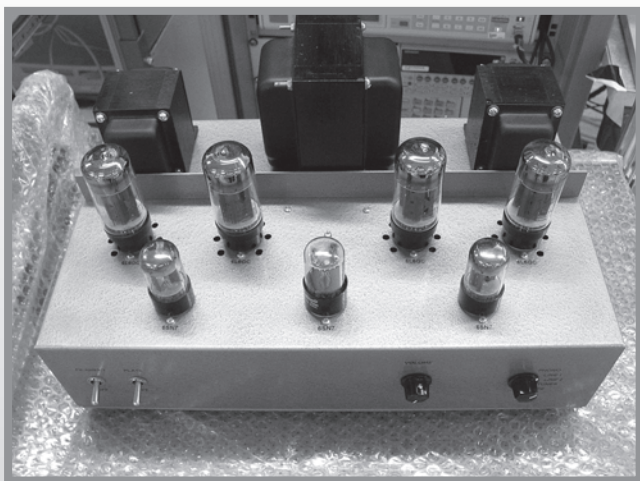
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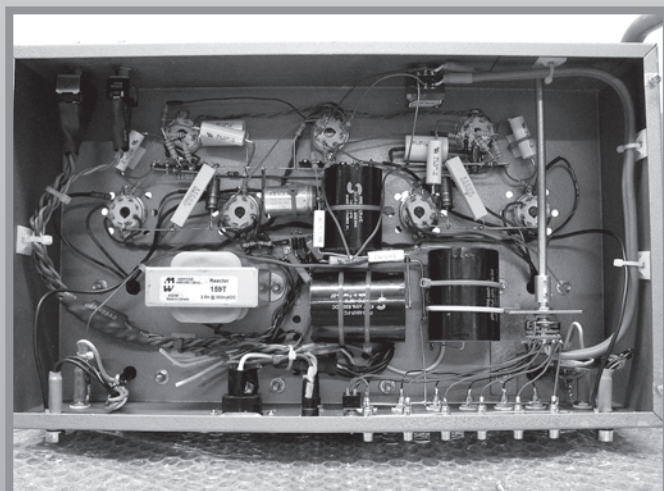
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July 2008
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20W Push-Pull Amp



Here are some photos of an amp I have built from your magazine article “A 20W \$260 Amplifier” (*Glass Audio 5/00* by Joseph Still). I made a few modifications (choke filter in the power supply, better components, and so on). Sound is very good, distortion is low, and there is no hum even with my sensitive horn loudspeakers. ■



BOOK REVIEW

Sound FX

Reviewed by Ed Simon

More money is spent on toothpaste each year than on professional audio equipment. That makes us a fairly small community. So it is a bit unusual for me to get a new book and read two dozen or so names in the acknowledgment section and personally know only one of them. OK, so some of the names are family and the book is on the edge of the work I do, but, still, only one common source is rare.

Alexander U. Case has written a book titled *Sound FX, Unlocking the creative potential of recording studio effects*—a clever title, considering that there are so many “complete” recording books that picking one area where most people just past novice would want to improve their skills will help sell books. The book contains much more than the title implies.

COVERING THE TOPIC

The author has an excellent grasp of what audio starts from, where it can go, and what he wants to do with it. In many ways, this book shares some important characteristics with the film cartoons of the late 40s. It can be enjoyed as a basic work, but those with more insight will be amazed at how much more there is.

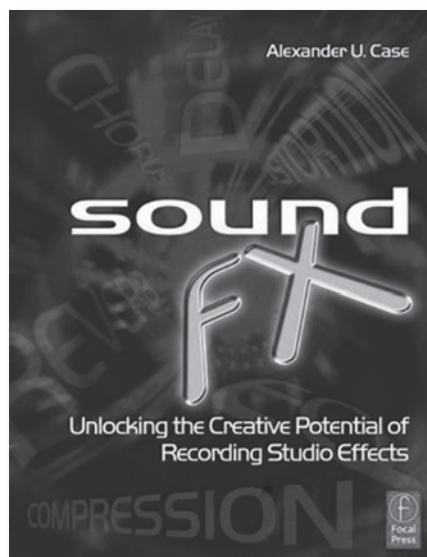
Case approaches his presentation a bit differently than I would expect in an introductory book. He first presents the cold, hard equations, even if it is the type of math that I would not expect to see until after at least one or two calculus courses. Then he presents the explanation in language and images that allow the beginner to grasp the concept.

The book also assumes that the reader has knowledge of how things really work and so skips what many basic books would address. For example, he doesn't mention microphones until chapter 13. He assumes that once you know how air is modulated in chapter one, you can figure out how this gets into the signal flow of chapter two.

This undoubtedly is because Case is really teaching this subject. He understands

what the student needs to add to his/her knowledge to do useful work.

The first section of the book includes three chapters about the basics. The first—how air is transmitted by sound waves—is represented graphically and mathematically, and explains why these principles concern us musically. The author then details how this is handled in the mixer and multi-track recording process. Concluding the introduction section is a chapter packed with information on how we perceive these sounds and can use this to advantage. This is the best single explanation I have ever seen of this material!



The next section of the book is on what he terms amplitude effects. The presentation in the fourth chapter on distortion is the start of this title material. The chapter is too short from an audiophile's point of view, but does offer insight into the recording artist's deliberate use of distortion for musical effect and some of the problems to try to avoid. A nice feature here and throughout the book is the listing of specific musical selections to illustrate the written examples.

The next three chapters cover equalization, compression and limiting, then expansion and gating. Although presented from a recordist's point of view, these

are a quite complete and understandable explanation of the technologies, the actual devices, the limitations of these processes, and how to use them for both recording and, to a lesser extent, for live work. Specific examples given of where and how to use the processes are clearly based upon the author's actual experience. This gives a much more hands-on feel to the book.

ENTERING THE STUDIO

Chapter eight on volume controls is probably the single most important chapter in the book. You would have expected the explanation of what a volume control is and how to use it to be much earlier in the presentation. It begins with an explanation of the types of controls and how they work. This material contains a brief explanation of how to mix music in a contemporary pop style. This chapter makes very clear to the journeyman the difference between older styles of recording, live recording, and current studio practice.

The next section deals with time-based effects. The three chapters cover delay, pitch shift, and reverb. A musical theory approach is taken to many of the explanations of how to use these processes. There are some charts relating tempo to time in chapter nine that will undoubtedly be a constant reference for some of the folks who work a mixing console for fun or profit.

Chapter 11 on reverb is aimed squarely at recording technology, devices, and techniques. But it will give most readers an insight into what the still emerging digital technology will do to live performances and the performing space.

Section four, which covers the basics of mixing, is where this work really becomes a more complete studio book. It explains very well why there are both pre-fader and post-fader sends on mixing boards. The basics of where to assign which signal and how to get started are reasonably

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CROSSOVER APPROACHES

I want to thank G.R. Koonce for his recent article "Passive Crossover Linear Phase Speakers" (June and July '08). This topic is of interest to me, and it is noteworthy that your conclusions and suggestions are similar to mine.

In my article titled "Tweeter Setback" (Sept. '08), I reduced the "delay dispersion" by putting the front of the tweeter behind the front of the woofer. Your article and mine have a common goal: to reduce the effects of phase distortion. Your article is more pristine in that it approaches zero phase distortion. My article acknowledges phase distortion, but asks what can be done to reduce it to inaudibility.

I was very interested in your comments on the audibility of different crossovers. You display unusual (for hi-fi) honesty. Your praise of the third-order crossover agrees with my assessment. In my opinion, fast crossovers help in reducing lobing.

There is one crossover design you did not mention that can result in a linear phase loudspeaker. In the supplementary crossover (CO), one CO (say, the low-pass) is subtracted from the input, with

the "supplement" being the other CO (the high-pass). However, I've worked at making a good supplementary CO for years now, with little success, the major problem being that the supplement is a low-order CO, often with ears. This results in a lot of destructive interference, and real-world drivers don't do destructive interference well.

I agree with your recommendation to use a digital electronic CO, because it can give arbitrary delay and a wide CO choice. I also agree that good drivers are very important. I have a pair of Lowther knockoffs that sound delightful in spite of their measured limitations.

I appreciate your good work, and look forward to your next article.

Dick Crawford

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G.R. Koonce responds:

I would like to thank Mr. Crawford for his kind comments on my article and look forward to reading his article. Tweeter setback generally reduces time dispersion, but can have practical problems. In my systems the tweeter location increased the distance between the tweeter and the grille cloth. After

I wrote the article, I learned that increasing this distance raises the trouble the grille cloth can cause in terms of adding echoes to the tweeter response. I spent considerable time rectifying this problem, finding thin felt rings on the tweeter faceplate virtually mandatory.

I think Mr. Crawford and I agree on the time dispersion issue. You want to keep it limited, but there is no need to drive it to zero. Thus, minimum-phase crossover designs are probably of more merit than true linear-phase designs. Mr. Crawford brings up the supplementary crossover. I believe this approach did not fit the intent of my article because I know of no passive implementation using a single amplifier per channel. It does not sound as though Mr. Crawford has had much luck with this approach.

Since my article, George Augspurger has brought to my attention a minimum-phase crossover approach he had developed. With his permission, I will document it here for readers to consider. It is amazingly simple, being a low Q second-order design with wide overlap. You simply design a second-order low-pass (LP) with a Q of 0.5 at twice the intended crossover frequency and pair it with a second-order high-pass (HP) with the same Q of 0.5 at half the intended cross-

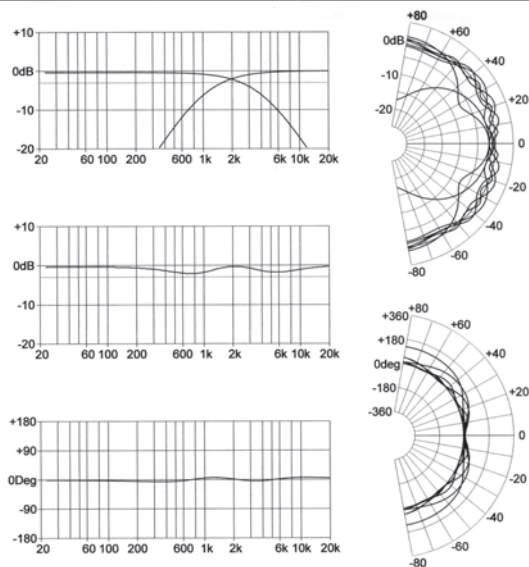


FIGURE 1: Performance of two-way minimum-phase crossover with ideal WTW array.

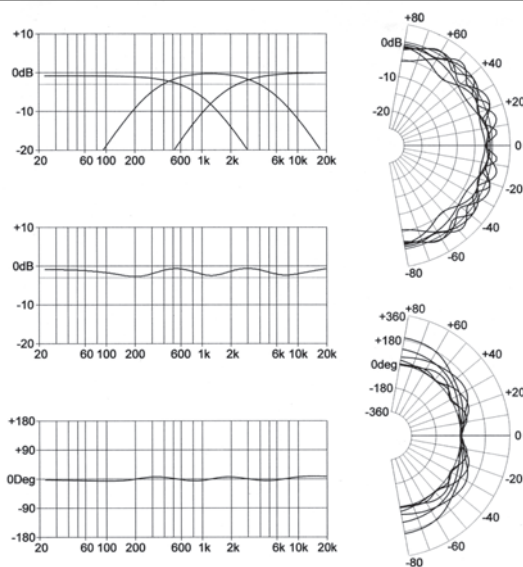


FIGURE 2: Performance of three-way minimum-phase crossover with ideal WMTMW array.

over frequency.

Figure 1 shows the shape of the LP and HP responses and the system performance with a symmetrical WTW array assuming ideal drivers for a design CO frequency of 2kHz. The magnitude anomalies are about 2dB and the phase holds to about $\pm 7^\circ$. The approach can be extended to a three-way by designing the bandpass to the same Q and overlap as for the other sections.

Figure 2 shows the performance for a three-way five-driver WMTMW symmetrical ideal array. The performance is not as good if the two crossover frequencies become close together. This is a simple approach with low parts count that ultimately gives 12dB/octave slopes, but still wants drivers with good response overlap. I recommend using symmetrical arrays to limit lobing problems. **Table 1** shows the design parameters for the crossovers in the figures to clarify the approach.

I hope all those working on time dispersion effects and limiting designs will publish so we may all make progress in this area. Right now, I admit I'm still confused about what constitutes the tolerable limit and how it varies with frequency and other variables.

The linear phase speaker article by Mr. Koonce was of great interest to me because I am building an almost identical system. I noted that the 18dB per octave crossover is -6dB at the crossover point, while the 6dB crossover is -3dB. Is this correct? Also, there is a 6dB per octave "Solen split" which is down 6dB at the crossover point. It is supposed to reduce the "problems" of a first-order filter. Any help or comments would be greatly appreciated!

Vincent Mogavero
VMogavero@aol.com

G.R. Koonce responds:

Mr. Mogavero has asked about the inconsistency in the crossover point between the first-order (6dB/octave) and third-order (18dB/octave) crossovers (COs) shown in my article. I suspect many readers may not be familiar with why the various CO orders have different crossover points, so let me give a brief review.

I will discuss the first-order Butterworth (B1), the second-order all-pass (AP2), the third-order Butterworth (B3), and the fourth-order Linkwitz-Riley (LR4). All of these COs

TABLE 1: GEORGE AUGSPURGER'S MINIMUM-PHASE SECOND-ORDER CROSSOVER APPROACH.

System	Crossover Frequencies	LP Q	Design f_{co}	BP Q	Design f_{coL} f_{coH}	HP Q	Design f_{co}
Two-Way	2k	0.5	4k			0.5	1k
Three-Way	500, 3k	0.5	1k	0.5	250 6k	0.5	1.5k

are capable of summing to unity across the frequency band, the definition of an all-pass CO. In all-pass configurations, they do not all have the same system phase characteristic.

Each CO is covered in the accompanying multi-graph figures. The plots are the electrical outputs of the CO when loaded with fixed resistors. You may alternately think of the plots as representing the on-axis acoustic output of an "ideal" coaxial driver.

The top graph is the low-pass (LP) magnitude output (solid line) and the high-pass (HP) magnitude output (dashed curve). The second graph shows the phase shift for the LP and HP outputs. The third graph shows the system output for both drivers wired the same polarity (solid line) and for the tweeter (HP) polarity inverted (dashed curve). The fourth graph is the system phase shift for the same conditions. All plots are for a 1kHz CO frequency.

The B1 CO (6dB/octave) is shown in

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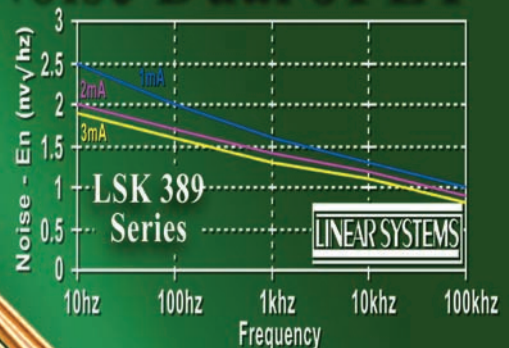
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Fig. 1. As Mr. Mogavero notes, this type CO has both responses down 3dB at the CO frequency. Note that the phase difference between the LP and HP is 90° throughout.

As **Fig. 2** shows, when two equal signals are vector-added with 90° spacing, the result is 1.414 times each signal, or up 3dB. This is why this CO has the LP and HP down 3dB at the CO point. Note that with either tweeter polarity the system sums to unity. However, only with normal tweeter polarity does the system show zero degrees phase shift and is thus phase linear.

The AP2 CO (12dB/octave) is shown in **Fig. 3**. The LP and HP phases are 180° apart, so when summed with normal tweeter polarity they cancel, producing the infamous second-order notch. With inverted tweeter polarity the system sums to unity because now the LP and HP are in phase. Two equal in-phase signals sum to twice each signal, or +6dB. So all-pass second-order COs cross over at 6dB down. Note that the system phase shift shows a phase reversal occurs at the frequency of the response notch with normal tweeter polarity.

The B3 CO (18dB/octave) is shown in

Fig. 4. The LP and HP are 270° apart which, because phase repeats at 360°, is the same as 90° apart. Thus again the two sum to 1.414 times each signal and must thus cross over at -3dB. Both tweeter polarities sum to unity. Note, however, the inverted tweeter system has much less phase shift across the frequency band than the system using normal tweeter polarity.

Finally, the LR4 CO (24dB/octave) is shown in **Fig. 5**. The LP and HP are 360° apart, which is the same as being in phase. Thus the normal tweeter polarity sums to unity while the inverted tweeter polarity has the notch. The system phase shift once again shows a phase inversion occurs at the

frequency of this notch.

So with “ideal” all-pass COs the pattern is clear: Odd-order COs LP and HP outputs should be down 3dB at the CO frequency, while for even-order COs they should be down 6dB. As Mr. Mogavero has spotted, the third-order CO I tried (Fig. 29 in my article) was down more than 6dB at the CO frequencies. Review of the second-order CO (Fig. 27 in my article) shows the lower CO point is down about 8dB. This is because when the CO drives real drivers things do not behave as predicted in the ideal.

The COs used in my article were developed by modeling the acoustic response of the system without regard for using ideal AP2 or B3 CO shapes. The drivers’ impedance variations and their response anomalies make the acoustic LP and HP responses much different from the expected electrical response of the CO components. It is the acoustic CO response that you care about, and using “ideal” electrical CO networks will likely not produce what is needed in the way of acoustic CO responses. How the acoustic CO responses sum is further modified by the drivers’ physical positions and their relative acoustic origin locations. Thus

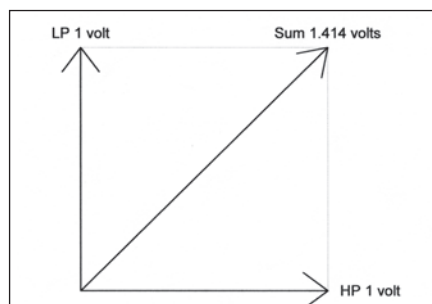


FIGURE 2: Vector summation of two signals at 90° apart.

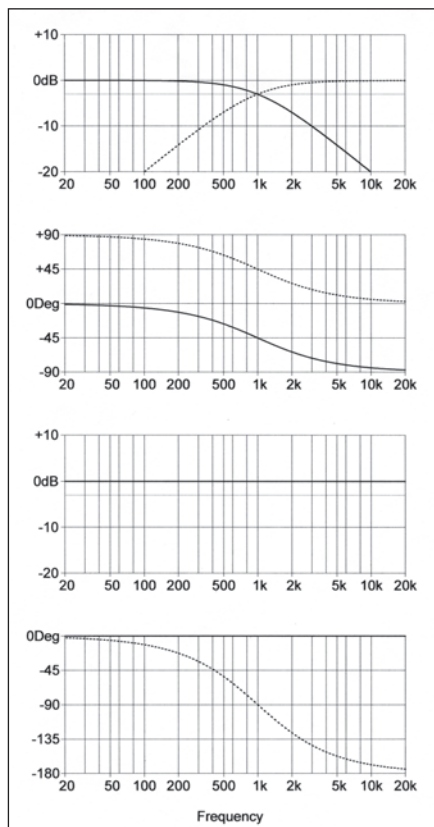


FIGURE 1: Performance of first-order Butterworth crossover.

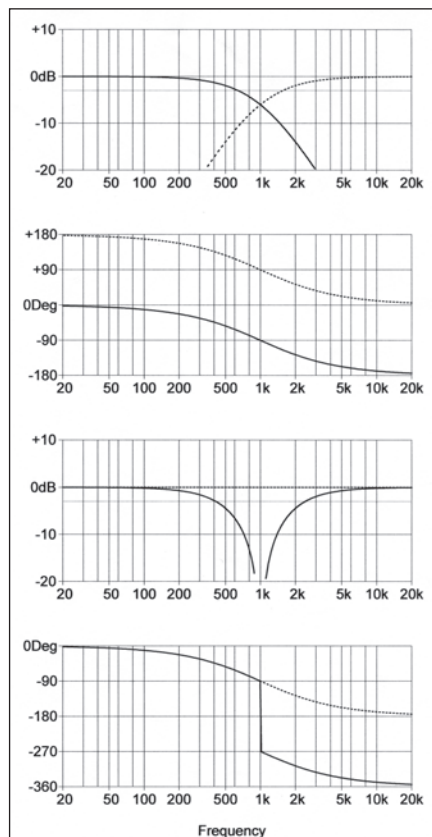


FIGURE 3: Performance of second-order all-pass crossover.

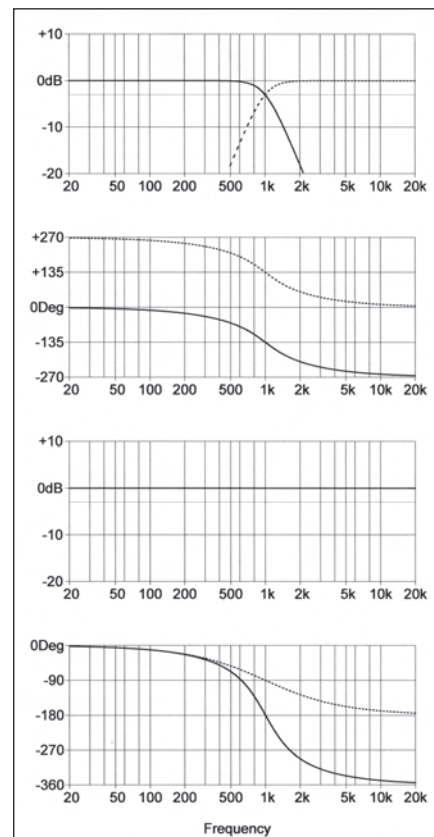


FIGURE 4: Performance of third-order Butterworth crossover.

to produce the desired system response the crossover points are not what ideal CO theory predicts.

I know nothing about the "Solen split" CO and can therefore not comment on it. I hope this answers Mr. Mogavero's questions and clarifies the different CO points. I wish Mr. Mogavero good luck with his speaker project.

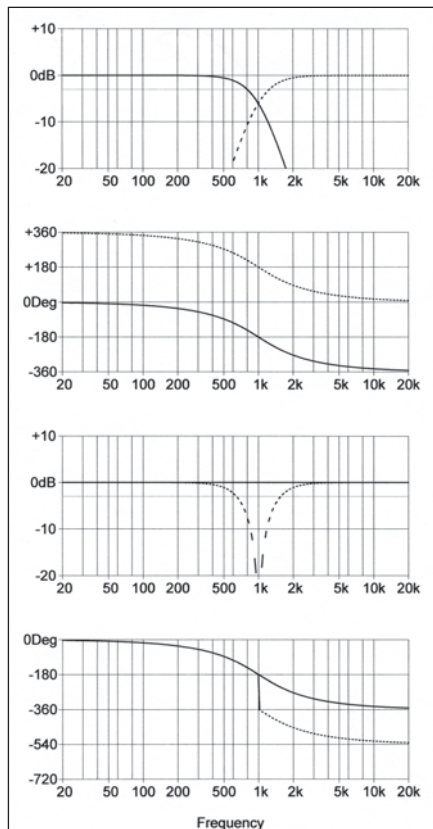


FIGURE 5: Performance of fourth-order Linkwitz-Riley crossover.

GUITAR AMP

I have some questions for Mr. Joseph N. Still about his tube integrated stereo amp (Sept. '08). I am a musician, and would like to build this amplifier to use with my tube preamp. Here are my questions:

1. Can the bias system, which is cathode bias, be replaced by the system you used in your 60W ultralinear amplifier from *audioXpress* 6/04? This system has an adjustable/balance bias system.
2. Can I bridge both output transformers with a switch to get a mono output (parallel them) when not using them in stereo?

I have all the parts to build your 60W unit, but I see you have a new 50W one that uses fewer parts. I know this is a hi-fi amplifier, but would love to use it in my guitar rig; it would give me a clean warm tube output power amplifier.

Dusk Ball

mosfetman20@yahoo.com

Joseph Norwood Still responds:

Thank you for your inquiry regarding my article. I hope you'll enjoy it and have little trouble building it!

1. Yes, you can use the bias system from the amplifier in the 6/04 article.
2. You can bridge the 8Ω outputs by using a switch, if the guitar speaker is 4Ω. If you use an 8Ω speaker, parallel the output

transformer with 16Ω outputs.

I found a matched pair of 6550s in a fixed bias configuration gave a balanced output, so you might choose to try this configuration before going with the more complex bias system.

The amplifier in the 9/08 article contains an error. The feedback of 12dB is incorrect; the correct feedback is 9dB. I'm very sorry for the mistake!

I assume you intend to build the amplifier on a single chassis, which is available from Mouser Electronics (800-346-6873). The size of the aluminum chassis is 17 × 10 × 3".

TUBE AMP

This is in reference to Alexander Arion's "Greek Triad" article in the October '08 issue. First, I can't see the advantage of the cascode stage. Maybe I just don't understand it well enough. Second, the coupling capacitor to the grid of the output stage is so big that the response would be down 3dB at about 0.6Hz. I pity the output transformer!

Third, the input is directly coupled without any coupling capacitor, so transients from switching inputs are applied directly to the grid. Looks like a poor design to me. Lack of negative feedback probably is because there is not enough gain in a single 6SL7 stage to permit it. Of course, maybe the author thinks negative feedback is bad.

Ron Anderson

Hendersonville, N.C.

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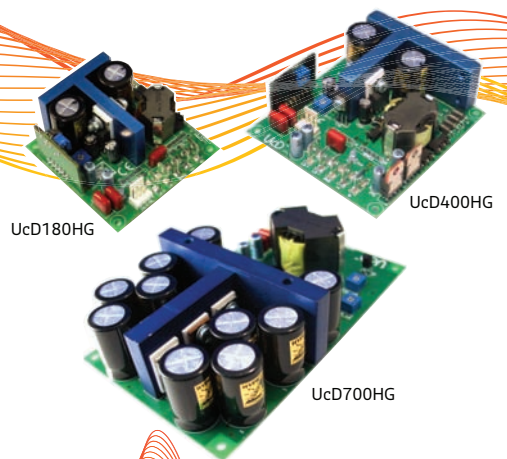
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Alex Arion responds:

Thank you for your inquiries about the Greek Triad. The first stage is not a “cas-code” but a SRPP, which is good enough for the job. I selected the 6SL7(6H9C) instead of the classic 6SN7 because of its M factor, bigger than 6SN7(6H8C). The coupling capacitor (1MF) was selected especially to transfer a lot of low frequencies—bypassed by a 47NF. The output transformer is doing well, without problems.

I am not against negative feedback, but if the electronic work is very carefully done, with no noise and good acoustic performance, why do it? Anyway, the amp has been working well from about one year in one of my friends’ home.

VOLUME CONTROLLER

I’m not quite sure what to make of “An Automated Level Control” in your October issue (p. 18). The author states that “It does not ‘compress’ the peaks, but simply ‘turns down the volume’ as you would do with your volume control.” I fail to see the distinction between this circuit and any active compression cir-

cuit with a predetermined threshold. This circuit is a compressor, plain and simple.

The author weds an LED and a CdS (not CDS) cell with epoxy and shrink sleeve. Silonex (www.silonex.com) makes a series of “Audiohm” optocoupler devices that perform this same function and have much improved audio characteristics over CdS. You can see a similar compressor circuit in their Technical References.

With the part values as shown, a steep attenuation of high frequencies throughout the circuit starts at the input where even at full volume control setting R18 and C9 reduce the HF -3dB point to just 3.8kHz. It will be reduced even further as the volume control is lowered. By the time the audio signal makes it to the output, the overall HF -3dB point is only 2.7kHz.

Op amps U1 and U2 have limited gain-bandwidth and should have small capacitors across their feedback resistors R14 and R2 to avoid oscillation. Any variations in R4 will change the LF -3dB point as the RC product of R3 +

R4 and C2 change.

The whole project seems more geared toward ham radio enthusiasts, where the extremely low (by hi-fi standards) high-frequency turn-down could make Morse code tones and voice more intelligible in the presence of a high noise background. I’m not familiar with ham radio conventions; perhaps that is why the schematic is shown with input on the right and output on the left. Overall, the project is fine for ham operators, I guess. However, given the limited performance of the AD820 op amps and that LM386, I just don’t think that one of the applications would be “music systems.”

Cal Jonstone

cjonstone1231@yahoo.com

J.R. Laughlin responds:

If you study the circuit a bit better to understand how it works, you will see that it does not *compress* or *limit* the peaks but simply divides down the entire audio signal such as a volume control will do. D1 and C6 store the approximate peak value, and the LDR divides down the volume so that the determined peak value is not exceeded. R9 was not used. R10 can be used to adjust the time constant of the circuit. Some information about compression and limiting can be read at sound.westhost.com/compression.htm.

The Audiohm couplers use the CdS cell like mine does. I set the high-end rolloff for my particular application; suit yourself for yours. The basic -3dB bandwidth of the amps is around 1.9MHz, which for any gain value used here would provide sufficient bandwidth for the audio range.

Read the “ADJUSTMENTS” section for how to adjust the low- and high-frequency rolloff. Note that use of the LM386 is entirely optional; you can connect any amp of your desire in place of it, externally. I can find no need for caps across R14 and/or R2.

HELP WANTED

I need help finding a source for ½” bi-tuminized felt. If you remember, Leak and others used it in their speakers to reduce wall vibrations. I have used layers of shingles, which work OK, but would prefer the right materials.

Vincent Mogavero

VMogavero@aol.com

aX

CONTRIBUTORS

Stuart Yaniger (“The Impasse Preamplifier,” p. 5) resides in Benicia, Calif.

Cyril Bateman (“Simulating Inductors and Networks,” p. 16) has done extensive work on capacitors, and resides in Norfolk, England.

Dennis Colin (“Noise Measurements of the LSK389B Dual JFET,” p. 24) graduated with a BSEE from the University of Lowell (MA) and is currently an Analog Circuit Design Consultant for microwave radios. Previously a band keyboardist and a recording engineer, he has been published in the *Journal of the Audio Engineering Society*. Colin has demonstrated the audibility of phase distortion at Boston Audio Society, and has designed the “Omni-Focus” speaker (bipolar coincidental with phase-linear first-order crossover), ARP 2600 analog music synthesizer, 1kW biamp and PWM supply at A/D/S, and Class D amps.

Paul J. Stampler (“Tri-Way Low Voltage Supply, Pt. 2,” p. 28) is a recording engineer/producer, musician, and technical writer; he also hosts a radio program, “No Time to Tarry Here,” featuring traditional folk music and related stuff. He has delighted in 78s since he was a boy, when they were still being made.

Gary Galo (“A De-Emphasis Test CD,” p. 32) is Audio Engineer at The Crane School of Music, SUNY Potsdam, where he also teaches courses in music literature. A contributor to AAC since 1982, he has authored over 230 articles and reviews on audio technology, music, and recordings. He has been the Sound Recording Reviews Editor of the *ARSC Journal* (Association for Recorded Sound Collections) since 1995, was co-chair of the ARSC Technical Committee from 1996 to 2004, and has given numerous presentations at ARSC conferences (www.arsc-audio.org). Mr. Galo is also a frequent book reviewer for *Notes: Quarterly Journal of the Music Library Association*, has written for the *Newsletter of the Wilhelm Furtwängler Society of America*, and is the author of the “Loudspeaker” entry in *The Encyclopedia of Recorded Sound in the United States*, 1st edition.

Arto Terho (“Showcase: 20W Push-Pull Amp,” p. 38), resides in Finland.

Ed Simon (Book Review: *Sound FX*, p. 39) received his B.S.E.E. at Carnegie-Mellon University. He has installed over 500 sound systems at venues including Jacob’s Field, Cleveland, Ohio; MCI Center, Washington D.C.; Museum of Modern Art Restaurants, New York; The Coliseum, Nashville, Tenn.; The Forum, Los Angeles; Fisher Cats Stadium, Manchester, N.H.

John Sunier (“Super Fidelity,” p. 46) is a CD reviewer for *Australian HiFi and Home Theatre Technology*. His website is www.audaud.com.

Continued from p. 39

but briefly, presented.

Case then moves on to two special chapters, the snare drum and the piano. The discussion of control and imaging here should be of interest, or perhaps controversy, to most readers. The specifics should guide many of the readers in the studio.

The final chapter is on mixing down the multi track recording in basic terms and with forms of automation. Today this is the conclusion of the studio recording process. In the days of vinyl there would have been a chapter on actually getting the sound to fit into the groove. It is assumed the digital files from the studio can be transferred unmodified to the distribution media with the current technology.

This book, although clearly aimed at current pop music recording techniques, offers many excellent explanations of basic to advanced audio theory and technique. Much is applicable to live sound, and the basics should help many audiophiles better understand why they hear the things they do.

Case does not cover many of the other recording issues in this book. There is very little on microphone technique. The actual selection of the gear and its interconnection is reasonably avoided, because that would limit the life of the book. So in one sense the title is correct; it is not intended to be a complete recording

guide. It does contain more than enough information that this book, with the use of the references, could be the core of an excellent modern audio education.

The one shortfall that bothered me is the lack of attribution for some of the data presented. As an example, figure 3.1 is a very valuable chart of audio thresh-

olds related to levels and frequency, but it bears no reference as to where this information came from.

This book hits its target reader dead on, but goes way beyond that. At \$39.95, this book is a very low-cost way to get a seat in the first-class compartment on the audio clue train. *aX*

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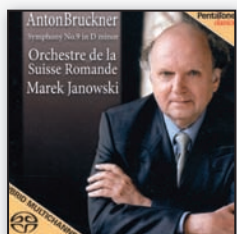


Claviers Mozartiens|Pierre Goy, vis-à-vis/clavichord/piano carré |Lyrinx SACD LYR 2251

During Mozart's short lifetime there was a transition from the hegemony of the harpsichord and clavichord to the new pianoforte. Christofori developed the gravecembalo col piano e forte, while in Germany Hebestriest created his dulcimer-inspired pianos known as "pantaloons." The vis-à-vis was a unique instrument with a harpsichord action at one end and a piano with bare wooden hammers at the other. The Sonata in B major is heard on the vis-à-vis. An unfretted, double-strung clavichord, based on a 1772 instrument, is heard in three short Mozart selections. Two very different square pianos also feature on this CD, because Mozart especially liked this type of instrument. The enhanced resolution of SACD makes it easier to distinguish the subtle differences between the four keyboard instruments.

French Romantic Organ Music arranged for Brass Quintet & Organ|Guilmant, Vierne; Lefebure-Wely, Boellmann|Elmar Lehnen, Seifert Organ, Kevelaer/International Brass Quintet|Audite MC SACD 92.556

The organ of the papal Basilica of St. Mary in Kevelaer, Germany, is a monumental instrument with 135 stops. The opening five-movement Sonata by Guilmant (No. 5/Op 80) is over a half-hour length, and is one of the composer's eight such organ sonatas. It blends ancient and modern elements and ends with a grandiose final double fugue. My pick of the disc is Boellmann's Gothic Suite, which also harks back to earlier music but avoids liturgical chant or even counterpoint. After an Introduction and Chorale, a minuet and a prayer to Notre Dame, the work ends with an absolutely spectacular audiophile-ecstatic Toccata finale which closes out the musical experience in stunning fashion.



Bruckner|Symphony No. 9 in E minor|Suisse Romande Orchestra|PentaTone Classics SACD PTC 5186 030

This incomplete (it lacks a finale) masterpiece was the shy Austrian composer's final symphonic utterance. The most modern of all his works, the second movement Scherzo with its incessant hammering rhythm could fit right into the futurist/mechanized music style which developed in the 1920s. The last movement he actually

completed—the Adagio—is the longest of the three and abounds in many intense orchestral climaxes. It is so monumental sounding that it brings the symphony to such a very logical conclusion that you don't miss the missing finale at all. This recording is superior to the Vienna Philharmonic SACD that was one of my Best of the Year picks last year. The Suisse Romande orchestra sounds richer, gutsier, and less strident, the low bass support is stronger, and the vital over-structure of Bruckner's massive blocks of sound flows more naturally and smoothly.



Frommermann|Music of the Comedian Harmonists|Channel Classics SACD CCS SA 26807

The unique male vocal sextet known as the Comedian Harmonists was founded in Berlin in 1927 by Harry Frommermann. That sextet combined crack vocal technique and arrangements—which led the way to many later innovations in jazz and classical vocal ensembles—with a wonderful

sense of humor and a charming presentational style. The Dutch vocal group Frommermann named itself after the founder of the Comedian Harmonists.

Established in 2004, Frommermann has the same voice blend as well: three tenors, two baritones and a basso, plus their pianist. Several of the 19 songs are in German but complete lyrics in English are in the note booklet. Their version of the music from Rossini's Barber of Seville must be heard to be believed. The sextet is arrayed in front of the listener and sings a variety of classical lieder, folk songs, pop tunes, and originals.

Art Pepper|The Way It Was!|Contemporary Records/Mobile Fidelity SACD UDSACD 2034

A lavish reissue on SACD from Mobile Fidelity is always a special thing, and when the music attains such heights as it does on these early Art Pepper sides, jazz lovers can't lose, particularly since Pepper himself dictated the liner notes. Some odd spatial placements occur, as happened in the early days of stereo when engineers were feeling their way with it. On the first tracks Pepper is playing cozy alto sax over on the left channel together with Warne Marsh on tenor while the drums and bass are way over on the right-hand channel. . . with absolutely nothing in between. Then on some of the solo Pepper tracks he is heard coming from the phantom center channel. Many tracks run more than six minutes, leaving plenty of time for some great solo work. The sound is clean, the tunes are all classics, and the individual approaches of the four different pianists make for added interest.



Bamboo|John Kaizan Neptune, shakuhachi/Arakawa Band|First Impression Music K2HD LIM K2HD030

This is one of the many albums from transplanted American shakuhachi master John Kaizan Neptune which have been extremely popular in the Far East. Neptune was enchanted with the sound of the simple Japanese end-blown bamboo flute and studied it formally in both Hawaii and Japan, becoming a master teacher, performer and composer as well as developing improvements on the instrument's design and building new instruments himself. This album was released 26 years ago and won a Japanese Best Recording of the Year award in 1980. The five tracks are a mixture of jazz, funk, and blues combined with Japanese folk music influences. The second and fourth tracks are the longest at over 12 minutes each and feature some exquisite sounds from the various shakuhachi Neptune has built. Fun listening.