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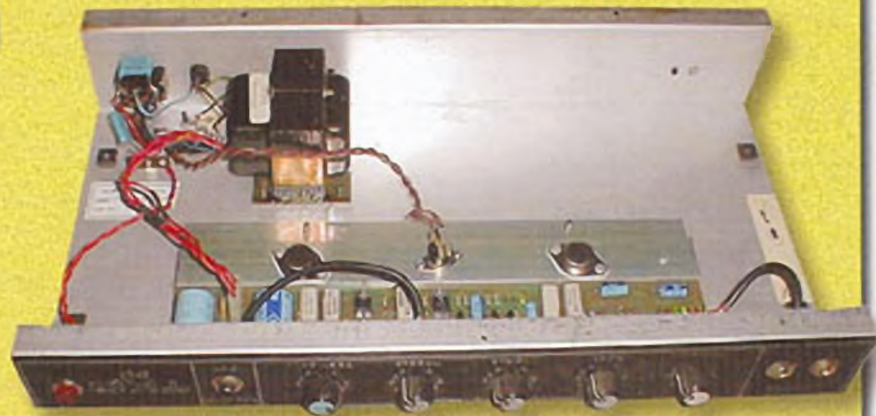

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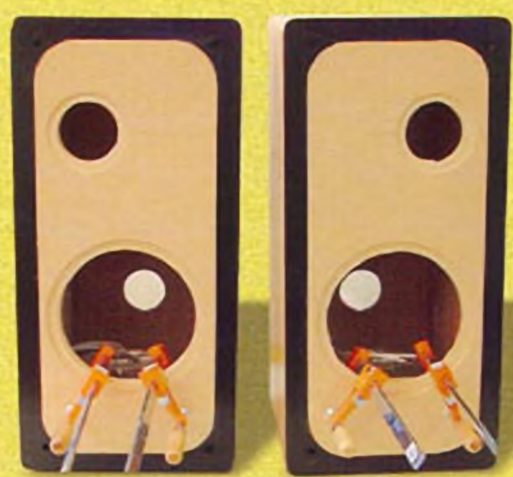

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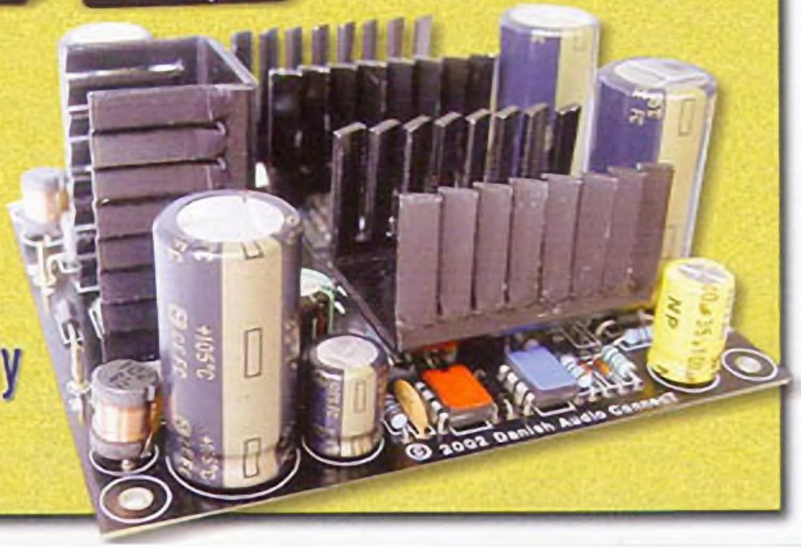

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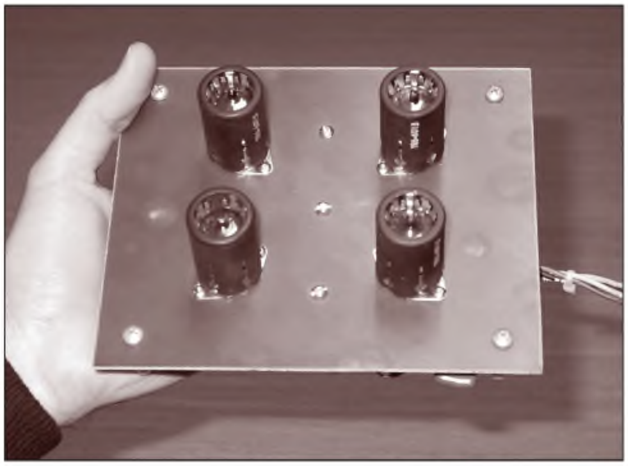
CONTENTS

VOLUME 34 NUMBER 10 OCTOBER 2003

FEATURES

page 40



page 36

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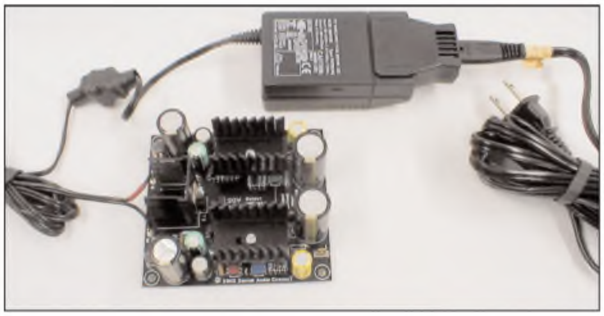

page 56

DEPARTMENTS

IN EVERY ISSUE

page 16

CODE Articles and letters identified with the following symbols:
solid state tubes speakers

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JOHN STUART MILL

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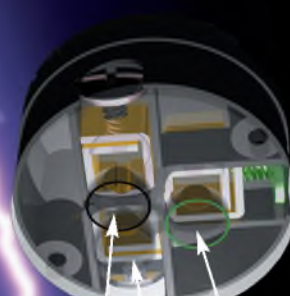

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Zen Variations, Part 5:
The Complementary Zen

Sticking with single gain-stage amps—this time with a push-pull design—this series proves once again that simpler is better.

By Nelson Pass

The Zen Amp concept explores how well you can make a simple audio amplifier perform. It makes for an interesting illustration of design technique, since more complex circuits are inevitably composed of collections of simple circuits, and it is well recognized that the optimization of these simple sub circuits is usually the key to getting the most out of the larger circuit.

Apart from that, there is aesthetic pleasure in rendering a device in a simple way. Also you find that—other things being roughly equal—simple circuits sound better musically. The Zen series has so far demonstrated that single gain-stage amplifiers have enough stand-alone performance to justify their use in audiophile applications. There is no reason why we can't go on from here to build two and three gain-stage circuits, but we have not yet exhausted the possibilities for single gain-stage amplifiers, and so we will explore some more.

COMPLEMENTARY DESIGN

Previously we have limited ourselves to single-ended Class A amplifiers, but it's time to try a complementary (push-pull) design. Since this is supposed to be a tutorial series, we need to ask the primary question, "What is meant by complementary operation and why do I want it?"

If there is only one gain device, there cannot be complementary operation. If there are two gain devices, you can operate them in opposition—one of them conducting more current while the other one conducts less current and vice versa. While this still constitutes one gain stage, the two devices give more gain and power, and in addition certain types of distortion (principally second harmonic) tend to cancel.

The reason for distortion in an amplifier is pretty much due to the variation of gain of the active devices over their operating curve. As a transistor or tube experiences greater and lesser voltage and current over the audio waveform its gain changes, producing distortion. If an oppositely operated device has the same sort of character but is operated in reverse, the distortions of the two devices tend to cancel each other out, and you get better performance. This effect is useful, although it is largely limited to second harmonic (non-symmetric), and does not help out at all on odd-order (symmetric) distortions.

Technically, the Son of Zen project (*TAA* 2/97) is a complementary stage by virtue of having two single-ended stages, one on each side of the loudspeaker. This has been occasionally referred to as a "balanced single-ended" amplifier, and isn't what you normally think of when you refer to a push-pull amplifier. However, it does benefit from the same distortion cancellation mentioned previously, and its character is partly the result of the complementary operation of the transistors.

COMMON DRAIN OPERATION

Routinely, designers produce single-stage complementary voltage follower amplifiers by using devices of opposite polarity types with both a plus and a minus supply voltage (*Fig. 1*), where two MOSFETs operate in "Common Drain" mode, and the signal goes in the Gate and comes out the Source. The circuit has current gain, but not voltage gain, and is similar to many of the output stages of ordinary multi-stage amplifiers in which a previous stage has done all the voltage amplifying you need.

It operates by having the Source pins of the MOSFETs connected to the output, where they tend to follow the voltage presented to the Gates. A bias voltage is required between the input of the amplifier and the Gates of the MOSFETs in order to put the Gates of the MOSFETs in the proper DC position. A MOSFET requires about 3–4V DC between the Gate and Source before it will start conducting, and by adding this to the input, you avoid losing the first 3 or 4V of the input signal.

For driving loudspeakers, you usually want the amplifier to have both volt-

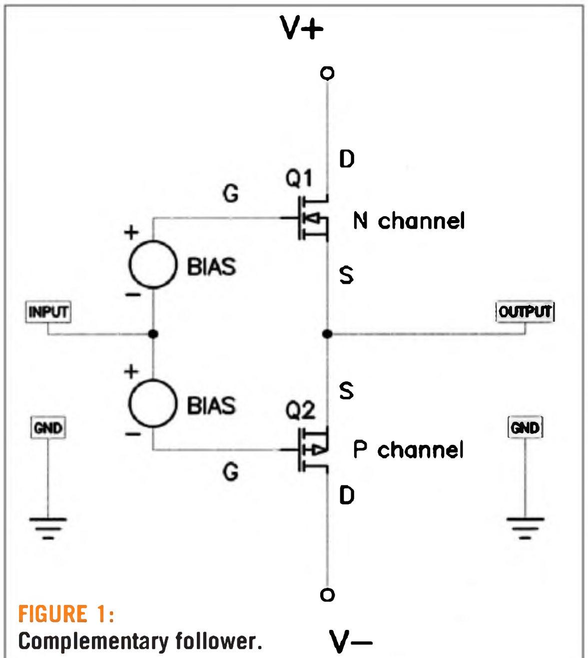

FIGURE 1: Complementary follower.

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Jensen JE-34K-DX	3	47	20Hz~20kHz	550	Area II $30
Peerless 4722	38	50	20Hz~20kHz	300	Area III $40 Area IV $50

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

■Speaker

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

*Price is for a pair **Air Economy

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

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■TAMURA TRANS (All models are available)

Model	W	Pri.Imp(kΩ)	Freq Response	Application	Price (US $)	I	II	III	IV	
F-7002 (Permalloy)	10	3.5	15Hz~50kHz	300B,50	836	60	70	110	145	Price is for a Pair
F-7003 (Permalloy)	10	5	15Hz~50kHz	300B,50	836	60	70	110	145	
F-2013	40	10	20Hz~50kHz	211,242	786	70	84	133	181	
F-5002 (Amorphous)	8	3	10Hz~100kHz	300B,2A3	1276	65	80	120	160	

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age and current gain. It is a rare preamplifier that will output the full voltage needed, and this leaves the follower circuit of *Fig. 1* useful only in a limited number of cases.

COMMON SOURCE OPERATION

To get both voltage and current gain, the amplifier will be operated in "Common Source" mode, in which the input goes into the Gate and comes out the Drain. *Figure 2* shows a simplified complementary version of such an amplifier. Q1 and Q2 are both driven by the same input signal, but because one is an n-channel device and the other a p-channel type device, their reactions to the input are opposite. If the input voltage is positive in polarity, Q1 conducts less and Q2 conducts more, and the output voltage goes negative, making this an inverting type amplifier.

R1 and R2 form the input and feedback network of the amplifier, and the gain of the circuit will be slightly less than the ratio of R2/R1. On the positive side, R3 and R5 form a network that sets up the proper DC voltage bias for the Gate of Q2, meaning 3 to 4V across R3. Similarly on the negative half, R4 and R6 set up bias for the Gate of Q1.

C1 and C2 conduct the input signal to the Gates of these MOSFETs while blocking this DC Gate bias.

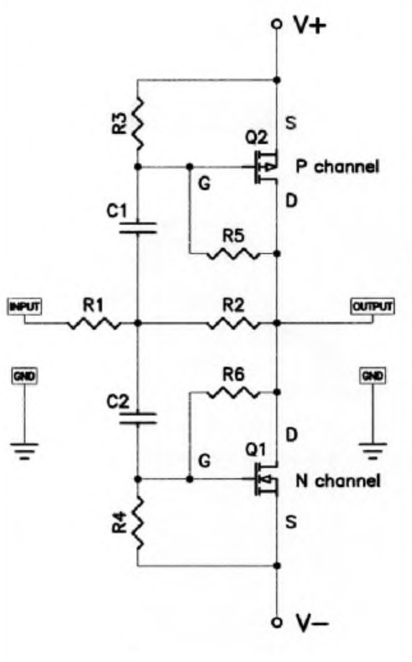


FIGURE 2: Simplified circuit.

MORE PARTS!

Figure 2 will work just fine as it is, but it's usually more convenient to add some little touches for extra stability. *Figure 3* shows a more detailed schematic designed to make life easier for those of you fearless enough to build this amplifier (and I certainly hope you do). First off, never use MOSFETs without some resistance in series with the Gate pins, else they might just fly away with parasitic high-frequency oscillation all by themselves, and so we have R7 and R8 to damp out that possibility. The value is not at all critical—somewhere between 100 and 500Ω will do the job.

Also, you will notice that we have added Source pin power resistors to Q1 and Q2. These help stabilize the DC bias a bit, and also give us convenient spots across which we can measure the current flow through the circuit.

In this case, I have chosen two parallel .47Ω 3W resistors to form a 6W .235Ω resistor (because the Digi-Key catalog offers only the inexpensive metal film Panasonic power resistors down to .47Ω). Like the Gate resistors, a wide range of values will work fine. I suggest between .1 and .5Ω, keeping in mind that they should be able to handle as much as 4A without overheating. All the rest of the resistors will be happy at .25W ratings, and none of them are critical in value except R3 and R4.

In a real amplifier, R3 and R4 assume critical values, as they sensitively control the bias current through the amplifier. We want this bias current to be in a well-defined range, as it bears strongly upon the performance and also the heat dissipation of the circuit. Too little bias current gives high distortion, and too much will cook the transistors. As a consequence, R3 and R4 are adjustable resistors (potentiometers) which will be used to set up the bias current exactly when the finished amplifier is first tested.

Note in *Fig. 3* that C1 and C2 have now assumed some values and are polarized electrolytic-type capacitors. I have no objection to film-type capacitors, nor is there any reason not to bypass the 47μF electrolytic caps with your favorite lower value film types. C1 and C2 should be rated at least the value of the supply voltage, and I recommend 35V minimum for this project.

Q1 and Q2 are not particularly critical as to type. I use IRFP240 and IRFP9240, which work fine, are commonly available, and are not too expensive. Many other types will work fine. Matching Q1 and Q2 is not important.

The heatsinking for Q1 and Q2 will be an important issue, as we intend to run them between 30 and 75W each, depending on the supply and bias. This is somewhat higher than we have seen in the other Zen projects, and even more effort will need to go into keeping these devices at reliable temperatures. In general, we can accept 50 to 60°C on the heatsink near the transistor mounting point.

The circuit of *Fig. 3* is quite sensitive to power-supply noise and fluctuation. V+ and V– will need to be stable and quiet; otherwise, the bias will wander and noise will appear at the output of the amplifier.

PERFORMANCE ANXIETY

We built up two separate channels of the circuit of *Fig. 3* and ran a series of bench tests, measuring characteristics at different supply voltages and bias points. *Figure 4* shows the total harmonic distortion and noise (THD+N) for the amplifier, measured at 1kHz and driving 8Ω. The regulated supply voltages V+ and V– have been set at 20V each, and three separate curves have been run, showing THD+N versus output power in watts for bias currents of 1.5A, 2.0A, and 2.5A. These bias figures produce heat dissipation for each device at about 30W, 40W, and 50W, respectively.

Figure 5 shows the THD+N for the amplifier with the regulated supply voltages V+ and V– at 25V each and bias currents of 1.5A, 2.0A, and 2.5A, dissipating about 37W, 50W, and 62W, respectively, for each transistor. In *Fig. 6* you see the same test, but with the supply voltages raised to ±30V, and the dissipation at about 45W, 60W, and 75W for each transistor.

From these curves we find performance that is roughly similar to the Zen Variations 2 project (July '02), with a bit more distortion but also a bit more power. There is not a lot of variation in performance between 1.5A, 2.0A, and 2.5A bias, although the higher bias tends to sound better and is also better with low impedance loads. Bias at only 1.0A gave much higher distortion under all conditions, and so 1.5A is the minimum recommended bias current.

The harmonic character of distortion is primarily second harmonic at lower wattages, in spite of this being a complementary circuit, and reflects the imprecise matching of the complementary devices. Because the devices are slightly different in gain variations, asymmetry appears in the distortion waveform, producing even-order harmonics. At power levels above a couple watts, third harmonic and other odd harmonic components start to dominate, and of course at some point the circuit simply clips pretty much like any other amp.

Figure 7 shows the THD plus noise from 20Hz to 20kHz measured at 1W and without filtering. The distortion climbs at the highest frequencies and is essentially doubled at 20kHz. The frequency response was measured as –3dB at about 10Hz and 100kHz.

The output impedance of the amplifier is relatively high, approximately 2Ω, giving a damping factor of about four. This is gain dependent, and improves inversely with gain, so that a higher value of R1 proportionately increases the damping factor.

QUIET, PLEASE

The ZV5 is much more sensitive to supply noise than other amplifiers, and this is because the input AC signal is referenced not to ground, but to the AC value of the supply rails. To the extent that the supply ripple noise is symmetric, there is some cancellation, but what remains is still amplified by the gain of the amplifier. Typically you will find that the value of the noise on the supply is comparable to the noise that appears at the output.

This means that the Complementary Zen requires a very quiet supply. Given a quiet supply, however, the noise figures are astonishingly good on our samples—on the order of 18μV over the audio band. As a rule, commercial manufacturers are happy with ten times that amount.

Providing the appropriate supply for this circuit is an exercise in big hardware. Usually when you say this you mean big transformers, capacitors, and heatsinks, but in this case you mostly mean big capacitors.

Alternatively, you could design an active supply using switching, feedback, and other techniques to get the appro-

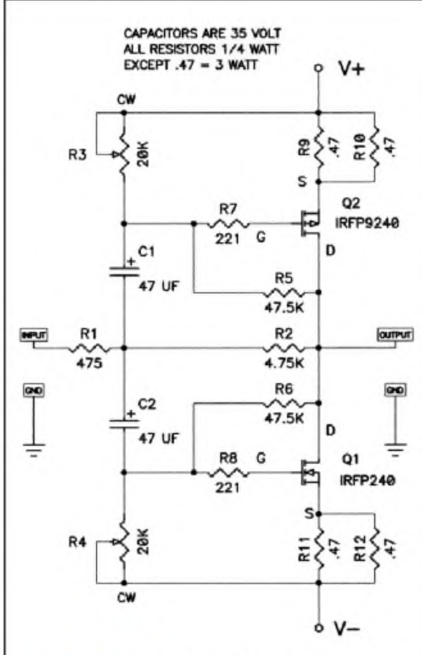


FIGURE 3: Complementary Zen.

priate voltage and current at the noise levels you want. You want a source impedance around .05Ω or less and noise around 10μV or less. I leave this approach to you as an exercise; this article does it the big, dumb way.

Figure 8 shows the schematic of such a power supply. In many ways this is typical of the power supplies for the Zen amps, but it uses all the tricks to get low noise and low source impedance. Big as it is, this supply is meant for one channel only, as any signal impressed on the supply by one channel will show up on the other channel.

PRIMARY SCHOOL

Because the design of this circuit is flexible enough to take advantage of a range of supply levels, we cannot specify a particular transformer, and so we will go through a quick exercise to determine what we need.

The power transformer will need to have the voltage and current rating appropriate to the draw of the circuit. In general, you should use a transformer whose VA rating is more than twice that of the circuit draw.

The rule-of-thumb formula for the selection of the transformer is as follows:

VA rating (watts) = 2.5 * 2 * (regulated voltage + 8) * bias current

For the example of ±20V at 1.5A (the smallest example) you get:

VA = 2.5 * 2 * (20 + 8) * 1.5 = 210W

For the example of ±30V at 2.5A (the largest example):

VA = 2.5 * 2 * (30 + 8) * 2.5 = 475W

Remember, this is per each channel!

SECONDARY SCHOOL

You want the unregulated supply coming off bridge rectifiers B1 and B2 to be about 8V higher that the desired regulated supply, and that is why you see the number eight added to the regulated voltage in the previous calculation. Of this 8V, about a volt will be lost to ripple, about a volt will be lost to DC across the inductors L1 or L2, about 4V will be lost from the Gate to Source of regulator transistors Q3 and Q4, and you will need the 2V left over to run a couple of milliamps current through R13 and R14 to drive the Zener diodes Z1 and Z2.

A simple way to calculate the secondary voltages required is to start with the regulated rail voltage, add the 8V, add another 3V for bridge and thermistor losses, and divide by 1.4, which will give you the AC voltage rating of each secondary winding. For the 20V example, you have (20 + 8 + 3)/1.4, which gives 22V AC.

This figure should be under load; in other words, this transformer should deliver this secondary voltage while operated at 200W. If a 10% regulation figure is called out on the transformer specs, then consider adding 10% to that 22V AC. When in doubt, buy a heavier transformer with a bigger VA rating.

You may not get exactly the voltage that you want, but this is not likely to be a real problem. You may need to adjust your expectations for the voltage rating of the Zener diodes Z1 and Z2 to be about 2V less than the voltage appearing at the Drains of Q3 and Q4, and everything will probably be fine. If you want the difference voltage to be other than the 2V, adjust R13 and R14 so that the current flowing through them to the Zener diodes is about 2mA.

Fuse F1 is standard slow blow type rated at 4A or so (for 120V operation). If this proves to be too small, then try 5A. If you are running 240V AC line power, then you will need to halve that figure, as well as run the transformer primary coils in series rather than parallel.

S1 should be a nice, big, high current rated switch. While it's true that thermistor TH1 will help reduce large turn-on current surges, the rating of the switch should be at 25A and 250V or better. Thermistor TH1 is a Keystone CL60 type, rated at 5A steady-state current. Of course, you can leave both the switch and/or the thermistor out if you wish; they are not truly required for operation.

C3 should be a line-rated capacitor, designed to sit across an AC power line in safety. You can leave it out also, if you wish.

B3 is a power diode bridge, and can be rated the same as B1 and B2, which are 35A at 200V or so. B1 and B2 clearly have the function of rectifying the secondary AC voltage into the unregulated rails. Some of you will wish to explore the variety of higher-end types of diodes that are available for this purpose, and of course you are encouraged to do so. The prototype was built with big, dumb, cheap rectifier bridges with no pretension whatsoever.

In the case of B3, the function is to provide a connection to earth ground (the one provided by your AC outlet) with a small voltage barrier to prevent

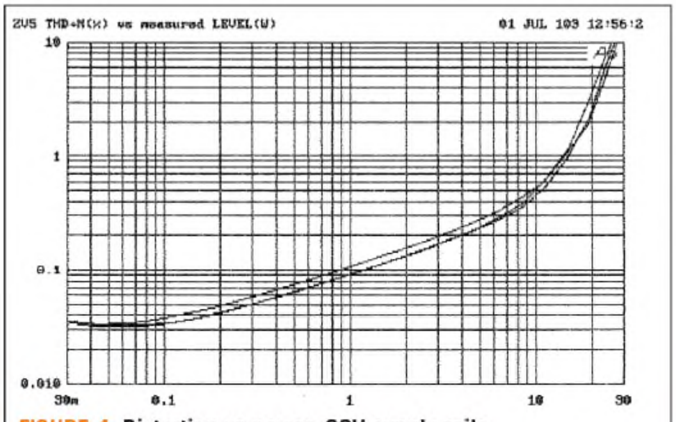


FIGURE 4: Distortion vs power; 20V supply rails.

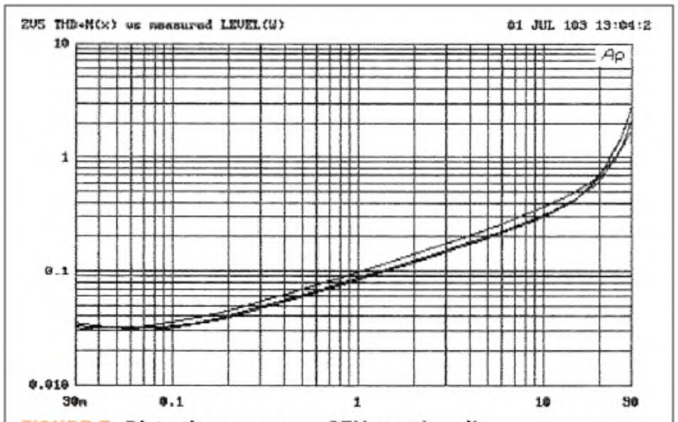


FIGURE 5: Distortion vs power; 25V supply rails.

the ever-present ground loop that haunts many systems. If your internal ground starts exceeding about .6V above earth ground, B3 will conduct it off to earth, keeping you safe from shock, I hope, in case of transformer failure or some other mishap which makes the ground more lively than is wanted. There are other approaches, and perhaps you can bypass B3 altogether without picking up noise, but for safety reasons you should never disable the earth ground connection of your system.

You will note in *Fig. 8* that provision is made on B3 for connection to the ground of another channel sharing the same chassis. If there is no other channel sharing the same chassis, you can leave this pin unconnected, or better yet, attach it to the same circuit ground as the other pin.

Capacitors C4 through C8 are rated at about 30,000µF (or more!) at a voltage exceeding the unregulated rail voltage (remember: regulated voltage +8V). You cannot skimp too much on these capacitance values and still get a sufficiently quiet supply for this amplifier. You can parallel smaller values if you like, and of course you can add the always popular film capacitors for additional bypass.

REGULATION RULES

The MOSFET regulators in this circuit are simply followers, driven by the voltage appearing at their Gates. On a long-term DC basis this is determined by the value of the Zener diodes, but fairly large capacitors C10 and C11 are provided not only to help filter noise, but also to provide a very slow startup voltage.

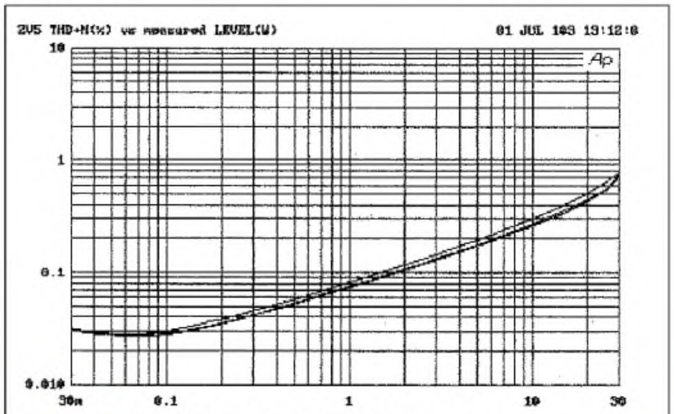


FIGURE 6: Distortion vs power; 30V supply rails.

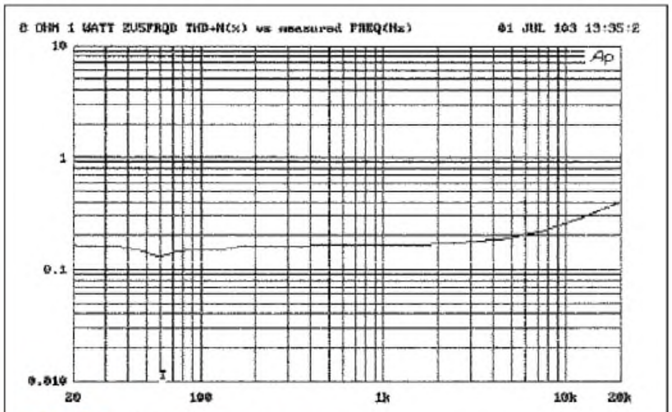


FIGURE 7: Distortion vs frequency.

ZV5 can draw quite a bit of extra current if awakened rapidly due to C1 and C2, and so the 2200µF values for C10 and C11 give it a gentle wake-up call.

You might imagine that large capacitance would not be needed coming off the output of the MOSFET regulators, but in fact they are. Q3 and Q4 are operated without feedback, and their intrinsic "resistance" (the inverse of their transconductance) is about .25Ω, enough to produce some undesirable modulation at AC frequencies. This is squashed by the 30,000µF of C8 and C9 so that it doesn't become an issue until down about 10Hz or so, and by then your woofer has much bigger problems than the amplifier.

Note that Zener diodes Z1 and Z2 are chosen to be 4V higher than the regulated voltage desired. The appropriate voltages can be formed by placing Zener diodes in series, but you should consider the dissipation each diode experiences. Typically, each Zener diode will have about 2mA current running through it, and this multiplied by the Zener voltage forms the dissipation figure. For example, a 34V Zener with 2mA current dissipates 78mW (.078W). A nice .5W Zener diode would do well here, and would be unlikely to break.

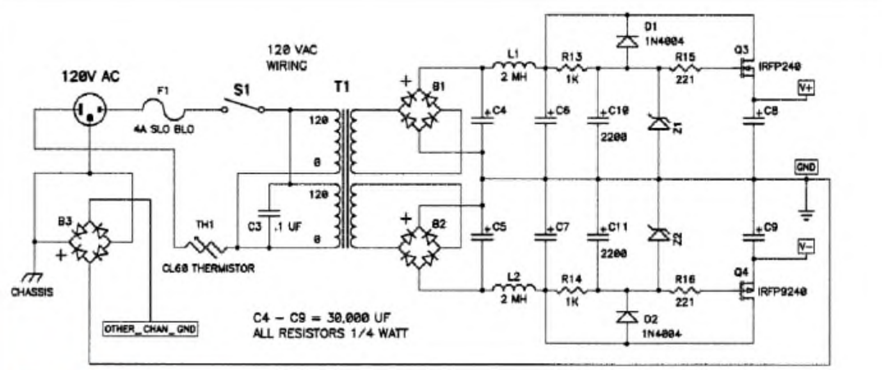


FIGURE 8: Power supply (single channel).

Diodes D1 and D2 are provided to drain off C10 and C11 quickly when the power is turned off, else the 20V Gate to Source ratings of Q3 and Q4 might be exceeded.

The particular characteristics of L1 and L2 are not very important. They simply need to hold about 2mH of inductance at a DC current up to 3A or so without saturating or burning up. Your standard big coil with fat wire will do the job, and remember not to locate it near the amplifier's input. In fact, with the exception of C8 and C9, the power supply in general is best kept at a distance from the audio portion of the circuit. No need to get carried away—a foot or so usually does the trick.

BIAS CHOICES

In previous Zen projects we have tended to limit the dissipation on the gain devices to around 25W or so, but in ZV5 we are clearly playing a bolder game, with dissipation figures as high as 75W per device.

The distortion curves of *Figs. 4, 5,* and *6* clearly show that there is not a big difference in distortion versus bias figures between 1.5A and 2.5A, as the curves tend to cluster together for these figures.

It's true that in general the sound does become better with the higher bias figures, but the heat to be removed climbs proportionately, and keeping these devices as cool as possible becomes critical when you start exceeding 25W or so dissipation.

Let's just say it outright: There's no such thing as too much heatsinking. Whatever you think will do the job, go ahead and double it. Your friends and neighbors (if not your wife) will stand in awe of your massive heatsinks, and probably your transistors will survive a few years.

And if you find that you underestimated your heatsinks, don't despair; any kind of fan blowing air across the sinks will work wonders.

CONSTRUCTION TIPS

This is a relatively simple circuit you can easily build with point-to-point wiring. We will be offering a circuit board and maybe some parts for sale at www.passdiy.com, and there is very little else that you can't obtain from Digi-Key and similar suppliers.

As always, MOSFETs are sensitive to static electricity, so you should handle them with some care until you solder into the circuit. You need to mount the gain devices to the heatsink with some care, because the wattage is higher than usual for these projects, and you want to remove as much heat as you can from the chip. Remember that the metal of the transistor cases are Drain connected, and unless you have made special arrangements, you must electrically insulate them using silicone material or mica and thermal grease. If you can devote an electrically isolated heatsink to just Q1 and Q2, you can dispense with the insulator, as the circuit connects the Drains of Q1 and Q2 anyway, and just use thermal grease (computer geeks love Arctic Silver).

Virtually none of the parts in this circuit are critical in value, with the exception of R3 and R4, which are adjustable. Take care in the selection of potentiometers for R3 and R4, since an open wiper on either of them will result in high current and probable failure. If you get to the point where you can precisely replace R3 and R4 with fixed resistors, feel free to do so. Outside of R3 and R4, feel free to alter the values of the parts or use what you happen to have on hand.

Similarly, a large variety of complementary MOSFET parts will work perfectly well. Don't be afraid to try them out. Watch out for the wattage ratings, and by this I mean never run a part higher than half of its power rating. You may find that in rare cases the Vgs figure is different enough to require a higher potentiometer range than the 20kΩ of R3 and R4, but this is unlikely.

START ME UP!

At this point, assuming that you have carefully constructed this circuit, checked everything twice, and then again the next day, it's time to fire it up.

First, set the values of R3 and R4 to the counterclockwise (minimum resistance) setting. Don't assume anything; check it with an ohmmeter, because if anything will kill this amplifier it will be R3 and R4 set at maximum resistance.

Ideally it is nice to start an amplifier up with a full range of test equipment, monitoring all the voltages and currents, analyzing the distortion, temperature and so forth. I like to keep about eight multimeters, three oscilloscopes, an Audio Precision System 1, four soldering irons and a fire extinguisher on my bench. However, I'm going to suggest that you can do this one with a single Radio Shack voltmeter with a DC scale that goes from 1mV to 100V.

Also it will be helpful to have some kind of load; some power resistor or perhaps an old disposable dinky speaker (like the crummy one that came with your computer). We aren't planning on sending any real power through it; we just find that adjustment goes faster with any kind of load.

You can fully adjust the amplifier by alternately measuring the voltage across two points: Output to Ground, and across the parallel R9/R10 resistors (going from the Source of Q2 to the positive regulated supply). The voltage across the R9/R10 will tell you how much current is going through the transistor Q2, and the voltage at the output to ground will tell you when Q1 is operating at the same current.

Start out with the DC voltmeter on the 2V scale, and attach the probes across R9/R10. Turn the AC power on, and stand back. After there is no explosion or fire, check to see that the voltage read by the voltmeter is 0 or maybe only a millivolt or so. If this is the case, check to see what the voltage at the output to ground is. It also should be quite low—0 to a few millivolts.

If these voltages are 0 and the fuse hasn't blown, then you are halfway home. Check the unregulated and regulated supply voltages and confirm that you really did turn the amplifier on. Confirm that there is about 2V across R13 and R14 and the voltages across Z1 and Z2 and that the regulated supply voltages are about 4V less than the measured Zener voltages. It takes a little while for C10 and C11 to charge, so don't become excited if these voltages take 10 or 20 seconds to settle in.

Go back to R9/R10 and begin slowly turning R3 clockwise until a few millivolts DC appears across R9/R10. Then, looking at the DC voltage from output to ground, slowly turn R4 clockwise until the voltage goes from some millivolts positive to about the same number of millivolts negative. Repeat this procedure, increasing the value of R3 slightly while measuring the DC across R9/R10 and then increasing the value of R4 while measuring the DC output voltage.

Ultimately, you will want to arrive at some stable DC voltages across R9/R10 while also having an output voltage near 0. Assuming that you have chosen R9 and R10 as .47Ω each, the following voltages will correspond to the desired range of bias current:

353mV = 1.5A   470mV = 2.0A
588mV = 2.5A

As the amplifier warms up, you will need to trim these values, alternately adjusting for the voltage across R9/R10 and then for 0V DC at the output. This procedure could take a half hour, maybe more, to get just right.

How much output offset voltage is too much? More than 100mV, in my opinion. As a rule, you should be able to stabilize this circuit stable within 50mV after warmup.

CONCLUSION

This is a very good-sounding amp. It sounds most like ZV2, but with a bit more power and a little more punch on the bottom. Whether it is actually better or not is a matter of taste. At higher bias currents, it definitely delivers more into lower impedance loads. As of this writing I'm driving a set of Jordan 92s mounted into rear-loaded horns (the upcoming J-Horn project). I believe I could live with this for a couple of years.

ACKNOWLEDGMENTS

As always, many thanks to Karen Douglass and Desmond Harrington for their valuable help, and Sponge Bob Squarepants for inspiration.

"Ohhhhhhh, Who lives in a pineapple under the sea?..." ❖

Modifying Peavey's TKO 80 Bass Amp

Part 1

Two audio veterans pool their knowledge and experience to restore a classic bass guitar amp. **By Charles Hansen and G. R. Koonce**

Finished TKO 80 bass amp.

BY CHARLES HANSEN

I recently bought a "pre-owned" Peavey TKO 80 bass guitar amp, vintage 1980. It is certainly well-used, with a tear in the grille cloth, but the electronics are in working order. The driver is dusty and has mildew residue. The Tolex covering is shot, and most of the nickel-plated parts have rust on them (*Photo 1*). It's what *Guitar Player* magazine might call a "pawnshop prize." However, I have owned two other Peavey guitar amps, which have a well-earned reputation for bulletproof reliability. Therefore, I thought it was worth restoring.

The Parts Express catalog is filled with musical instrument, hi-fi, and car audio loudspeakers. Confusion reigns when it comes to choosing, so I contacted G. R. Koonce, who has much more experience in driver selection.

Peavey has a more recent 80W version of the TKO 80, which uses their 12″ "Blue Marvel" driver in a vented cabinet. My first attempt to order the TKO 80 user manual and service schematic resulted in the documentation for the newer amp. A quick call to Peavey's excellent customer support line verified that they still had the manual and schematic for the older unit, which they shipped right away. (Neither TKO 80 is in production any more. The latest incarnation is the TKO-115 with a 15″ driver.)

PHOTO 1: TKO 80 bass guitar amplifier as purchased with grille removed.

THE STOCK TKO 80 DESIGN

The 20″ × 20″ × 11″ enclosure is made of ¾″ particleboard and plywood. It seems to be a very stiff and solid enclosure. The amplifier chassis (*Photo 2*) sits in the top 3″ of the enclosure. The internal closed-box volume for the driver is $1.24ft^3$. There is no stuffing or acoustic absorber in the box, and it is not perfectly sealed. The speaker leads pass through a 5/8″ hole in the enclosure to the amplifier chassis.

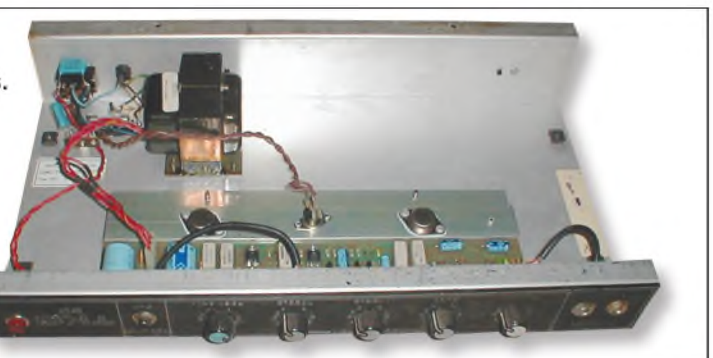
With the bass plugged in, I found that for the notes below C_3 (55Hz) the 12″ musical instrument driver was doubling the fundamental. At the lowest E_3 note (41Hz), the 82Hz second harmonic was very prominent. While this had the advantage of maintaining the volume level of the higher notes, it was hollow and unpleasant sounding.

I chose to replace the existing 12″ driver with one with more extended bass response, even at the cost of some volume. My primary use for this will be as a practice amp and for moderate-volume jazz jam sessions with friends. I didn't want to change the physical dimensions for several reasons. The amp already weighs 46 lbs. The custom grille board would no longer fit if anything were added beyond the existing front surface area. The power cord is stored on two L-shaped brackets on the rear of the chassis, so changing the depth would limit the access. The speaker leads plug into the circuit board from the bottom of the amp chassis, and the chassis would not be easily accessible if the enclosure were deeper. Any modification must be limited to adding a vent or an adapter ring (but preferably not both).

MEASUREMENTS

The four-string bass guitar covers the piano keyboard range of E_3 to E above middle C, or 41.2Hz to 329.6Hz (*Fig. 1*). The TKO 80 amplifier is factory rated for 50W into 8Ω at 5% THD, 40W at 1% THD. I ran some THD versus output power

PHOTO 2: Amplifier chassis.



measurements on the amplifier (*Fig. 2*), and it puts out only 30W at 3% THD, 1kHz into 4Ω. The power supply rails drop from ±37V DC NL to ±28V at maximum load, 4Ω. At 41Hz, it makes 43W into 8Ω and 28W into 4Ω, 3% THD. The power transformer looks a bit small, putting this unit into the "practice amp" category.

With all the circuitry on one PC board, there was no available connection to the input of the power amp section. There is a preamp out stereo phone jack, however. By rewiring this jack, we were able to use the short center sleeve to access the power amp input stage. This is what I used to obtain the 1kHz 8Ω power amp's only curve.

At low frequencies and low volume, the THD is limited by 60Hz hum rather than clipping (*Fig. 3*). The dip in THD at 60Hz is a result of the distortion test set notch filter removing the hum as well as the 60Hz fundamental. This doesn't seem to be power supply noise—that would produce 120Hz hum due to the full-wave rectification. There may be pickup or ground-induced noise somewhere.

The harmonic distortion products from the spectrum of a 50Hz sine wave at 10W into 8Ω (*Fig. 4*) show the presence of the 60Hz power line at −61dB. The 20-year-old filter caps (2200μF per rail) are probably tired. You can now get 4700μF in the same size package. The amp is an old quasi-complementary design, and you can see severe crossover spikes in the 1kHz distortion residual signal (*Fig. 5*). The distortion at this test point is 0.1%.

The existing bass/mid/treble controls are passive, and interact like mad. There is no position where they are flat even within 2.5dB. The solid line in *Fig. 6* shows the frequency response with the tone controls as flat as possible. You can also see a peak in the power amp's response (dashed line) at 60kHz, and the limited LF response without tone control action (−5dB at the E_3 41Hz fundamental). Testing with a 10kHz square wave (not shown) showed underdamped ringing, which verifies the HF peaking. The pre gain control also reduces the LF response at higher gain by increasing the −3dB point of the input op amp.

Someone had done some work on the circuitry since it was built in August 1980. The preamplifier section has an NE5532 dual op-amp, where the service schematic shows a TL072. The two "62792" NPN output transistors (selected 2N3773s according to the Internet references) had been changed to MJ15024s, with a 1992 date code. All the 100nF and 100pF caps in the audio path are ceramic Z5U types in axial packages (*Photo 3*).

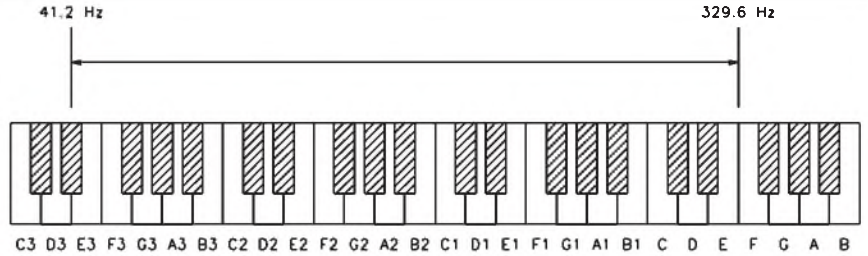


FIGURE 1: Bass guitar frequency range.

B-2247-1

The final test of the stock performance was to measure the acoustic frequency response with a sound level meter. I also measured the sound level with the TKO 80 amplifier driving my NHT SW2 subwoofer. This sub uses an external amplifier with electronic crossover, so the sub unit itself has no CO (*Fig. 7*). The big surprise here is the high-frequency range of the NHT sub. Note that the tests are at a fixed 1V RMS output at 1m rather than the usual 2.83V/m. At 1kHz and 2.83V RMS the sound level with the TKO 12″ driver was unbearable, even with hearing protection.

In light of the amp's poor 4Ω performance, we decided that we needed to select an 8Ω driver. I also plan to do some design improvements to the electronics, which we will cover later. With the basic requirements set, we started digging through the Parts Express catalog to find a replacement 12″ bass driver. We found that most of the "bass guitar" speakers have a "response" of 50Hz to maybe 3.5–4.5kHz. The non-musical-instrument woofers all have lower responses: 26–35Hz.

BY G. R. KOONCE
SELECTING A NEW DRIVER

Selecting an available driver to meet given system requirements is always a difficult task. When you add the requirement that the driver must work in an existing box of fixed volume the job

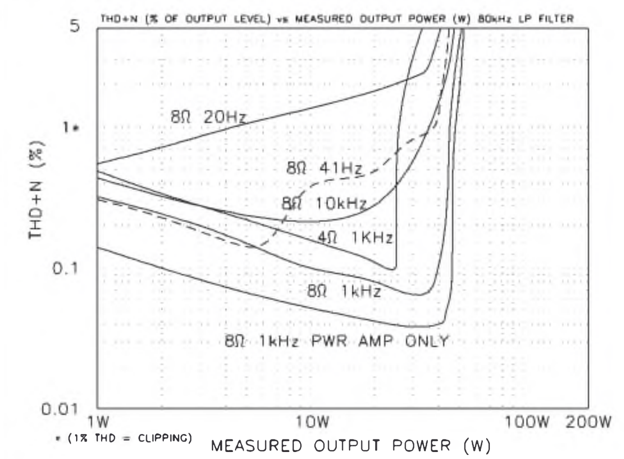


FIGURE 2: THD+N versus output power, stock TKO 80.

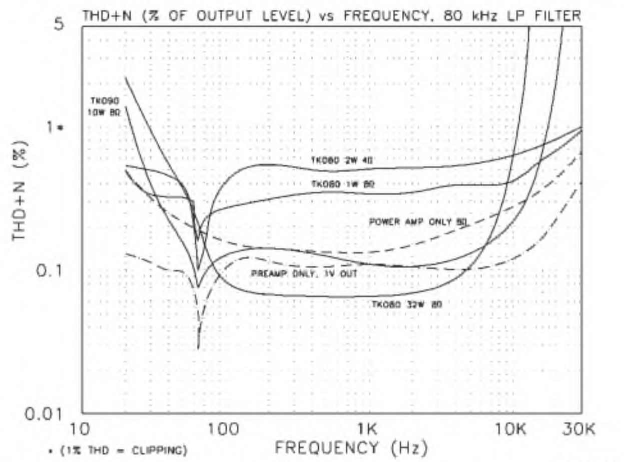


FIGURE 3: THD+N versus frequency, stock TKO 80.

becomes much tougher. Consider the items. In this case the gross box volume was fixed, because the external dimensions of the Peavey TKO 80 were not to be modified. The design box volume to be used was established at 1.24ft^3. One advantage is that the driver is connected directly to the amplifier (no crossover), so you need not consider any noticeable source resistance.

PHOTO 3: Amplifier PC board.

The new driver must, of course, fit in the existing box. This probably restricts you to a driver the same diameter or a bit smaller than the original. Generally, you can make an adapter ring to allow mounting a smaller driver. The other important items for consideration are the system specifications: frequency response, maximum SPL level, and efficiency.

This project required a system that would go down to low E_3 for the bass guitar, or 41Hz. This required selecting a driver that would produce a response down to near 41Hz in the 1.24ft^3 box volume. Some equalization was possible to "flatten" the response down to 41Hz if needed. The spectrum of the initial E_3 note attack from the unamplified electric bass pickup, played normally, is shown in *Fig. 8*.

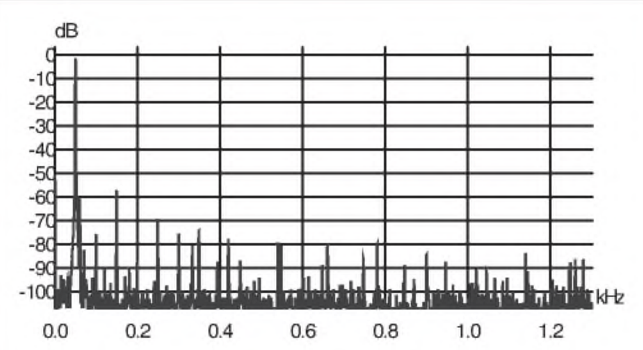


FIGURE 4: Spectrum of 50Hz sine wave, stock TKO 80.

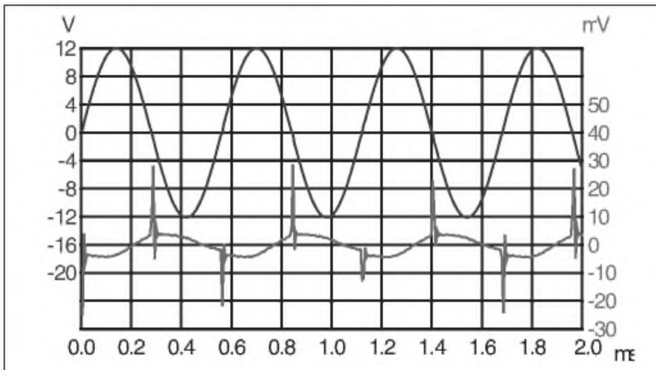


FIGURE 5: 1kHz residual distortion signal, stock TKO 80. B-2247-5

Here you can see that the second and third harmonics are higher than the 41Hz fundamental. This is undoubtedly why fundamental doubling in the stock 12″ driver was so effective in maintaining volume down to the low E_3. You can also see low-level 60Hz power line harmonics. The bass pickup is of the "humbucking" type, but some interference is unavoidable with an electromagnetic pickup, and the amp itself has some 60Hz susceptibility (*Fig. 3*).

As the string vibration decays, the fundamental increases while the harmonics dampen out (oscilloscope graph in *Fig. 9*). Another interesting phenomenon is that the initial pitch is a bit sharp, converging to 41Hz as the string vibration decreases in amplitude.

The high-frequency requirement was a bit vague. The highest required note is middle E, or 330Hz. To carry the third harmonic at full amplitude would mean about 1kHz on the high end; the fifth harmonic would push you to 1,650Hz. A study of specifications for current bass guitar amplifiers indicated that they were about 3dB down by 2kHz. A spectrum of the highest unamplified fretted note initial attack is shown in *Fig. 10*. From this, 2kHz looks like a reasonable upper-frequency limit for driver selection.

Initially, we did not know the required maximum SPL capability and required efficiency. The amplifier power available and the intended application indicated that extreme efficiency was not required. Bill Fitzmaurice designs music speakers that require very high efficiency, and he thus uses horn-loaded systems. Fortunately, this system did not require such high efficiency, which greatly simplifies the problem. Initial testing of the amplifier (about 40W to 8Ω, 28W to 4Ω) and the required playing level indicated a minimum sensitivity of about 87dB/W/m would be sufficient.

QUICK RESPONSE ESTIMATES

Looking through lists of T/S parameters of affordable drivers can be a bit confusing. Just how will any driver perform in a given box? You can achieve some quick estimates without resorting to design software.

Dr. Small defined an efficiency-bandwidth product (EBP) for drivers used in a direct radiator application; i.e., no horn or other acoustic load on the driver. The EBP is equal to the driver resonance (f_S) divided by the driver electrical Q (Q_{ES}): EBP = f_S/Q_{ES}. If the EBP is 50 or below, the driver will probably do better in a closed-box (CB). If the EBP is 100 or higher, the driver will do better in a vented box (VB). Between 50 and 100 you have your choice.

When dealing with the driver used in a VB you can get a quick estimated optimum design just by the total driver Q (Q_{TS}). If Q_{TS} is about 0.4 (for a total box Q of 7), then the optimum box will be a true Butterworth fourth-order (B4) re-

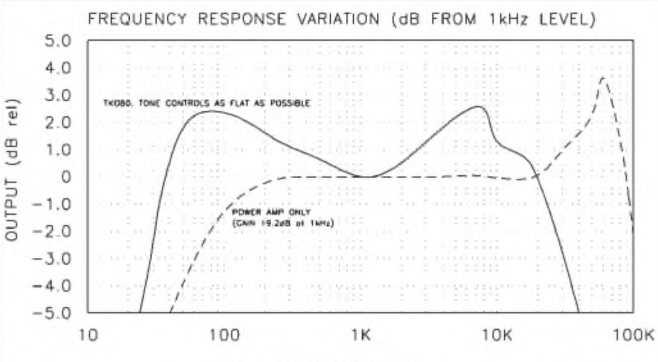


FIGURE 6: Amplifier frequency response, stock TKO 80.

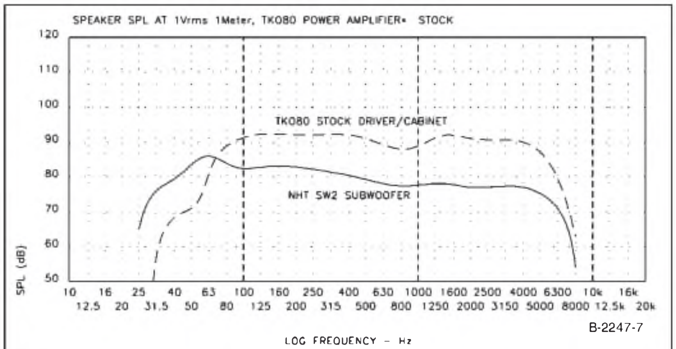


FIGURE 7: Loudspeaker frequency response, stock TKO 80 and NHT SW2.

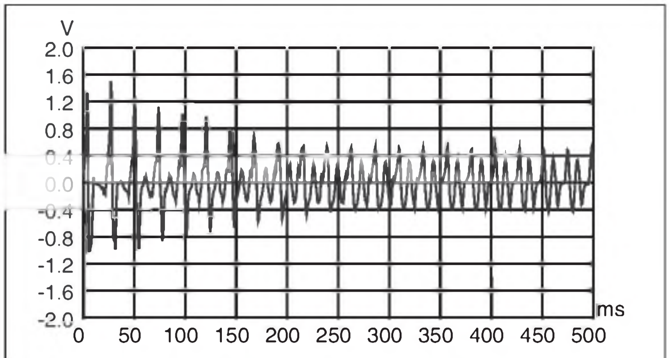


FIGURE 9: Waveform of bass guitar E_3, 41Hz.

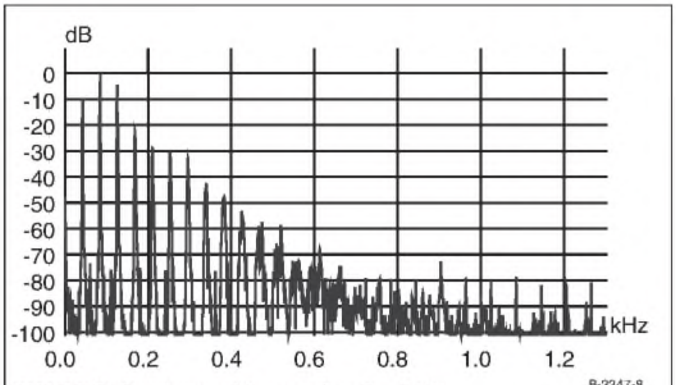


FIGURE 8: Spectrum of bass guitar E_3, 41Hz.

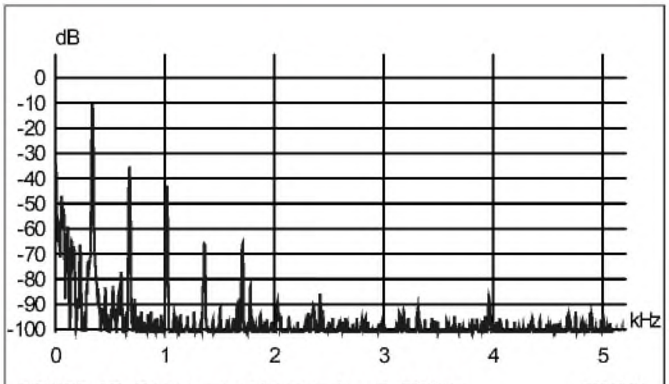


FIGURE 10: Spectrum of bass guitar E, 330Hz.

sponse. This means the box volume will match the driver equivalent compliance volume (V_{AS}), and the box tuned frequency (f_B) will match f_S and the system –3dB frequency (f_3): $f_B = f_3 = f_S$.

As Q_{TS} goes above 0.4, then you tend toward a Chebyshev (C4) response with a bit of ripple. The optimum box size will be bigger than V_{AS}, while f_B and f_3 will be somewhat below f_S. If the Q_{TS} drops below 0.4 then you tend toward the quasi-Butterworth third-order (QB3) responses. The box volume will be below V_{AS}, but f_3 and f_B will be above f_S, possibly way above.

I know some music people use the boom-box 4 (BB4) alignment, with which I have no experience. It has a simple (without ripple) peak in its response, but I don't know how it is for box size or f_3.

You can also estimate how a driver will do in a closed-box (CB) based on driver Q_{TS}. Generally, you would design a woofer CB with a total system Q (Q_{TC}) somewhere in the range of 0.7 to 1.0. The f_3 will be at its lowest if you use the highly damped $Q_{TC} = 0.707$ (B2) alignment, but f_3 is a slow-moving function of Q_{TC}. The closer the driver Q_{TS} is to the Q_{TC} you want, the bigger the box will be relative to V_{AS}. The CB system resonant frequency (f_C) will be equal to $f_S \times Q_{TC}/Q_{TS}$ and, over the Q_{TC} range we are discussing, f_3 will be about 80 to 100% of f_C.

It is thus clear that low Q_{TS} drivers will produce a small CB, but with a high f_3. High Q_{TS} drivers designed into a CB at the low end of the Q_{TC} range will require a large box. Clearly, computer software is the way to get exact values, but these tips can give you a quick hint.

TABLE 1
DRIVER IDENTIFICATION

NO.	DRIVER SIZE, MAKE, AND MODEL	Z_{IN} (Ω)	F-RANGE (Hz)
1	12″ Eminence Delta-12LF #290-416	8	45–3,000
2	12″ Eminence Kappa-Pro-12 #290-424	8	35–2,500
3	10″ Dayton Series II HD #295-110	8	35–3,000
4	12″ Pyramid Super Pro SPC #290-115	4	35–3,500
5	10″ Audax PR240ZO #296-105	8	27–2,000
6	10″ Dayton Pro Sound #295-060	8	35–3,000
7	12″ Dayton Treated Paper Cone #295-320	8	28–2,500
8	12″ Martin Sound Poly Cone #3130	8	18–5,000
9	11″ Hi-Vi D10G #297-430	8	23–2,000
10	10″ Peerless 850146 #297-636	8	23–1,800
11	10″ Peerless 831727 #297-616	8	22–2,800
12	10″ Vifa M26WR-09-08 #297-315	8	26–1,000
13	11″ Focal 11K7512 #297-535	8	38–2,000
14	10″ Phoenix Gold QX108 #293-550	8	32–1,200
15	12″ Peavey TKO 80	8	

Notes: Size is nominal inches. # is Parts Express Catalog number. Zin is the rated driver impedance in ohms. F-range is the catalog-specified frequency range.

DRIVER TYPE OPTIONS

The system requirements for this project mean three driver types are candidates for solving the problem.

The first is music speakers, which generally have very high efficiency and thus big magnets. This typically means they have a low Q_{TS}, which makes getting a low f_3 difficult. Most affordable music speakers are not known to have a high linear displacement (X_{MAX}) capability. Finding a music speaker to deliver the required output level at 41Hz might be a problem.

Normal woofers are also candidates, as the efficiency requirements are not so high as to exclude them. They should certainly be able to meet the upper-frequency-limit requirement. The allowable 1.24ft³ box volume is a bit small for a 10″ or 12″ woofer to get down to 41Hz.

Subwoofer drivers are also candidates at the required efficiency. Their main advantage is a large linear displacement capability (X_{MAX} rating), meaning a high output SPL capability at the 41Hz lower frequency limit. The question is whether you can get such a driver to work with a 1.24ft³ box and whether it would go high enough in frequency.

Many large drivers, of all classes, have a dip in their response somewhere in the 400Hz to 1kHz range. This is normally referred to as an "edge hole" and is said to be caused by the surround becoming out of phase with the cone center. Such dips are normally not identified unless a true driver frequency response plot is available. These dips can be rather deep, and if they fall on a fundamental or harmonic frequency that you need, they can be a problem.

Large drivers also have a tendency not to go away "nicely." Generally at the top end of their range, the response peaks due to the development of cone break-up modes. Again, if one of your harmonics falls on such a peak, the sound quality could suffer.

REFERENCE POWER AND EFFICIENCY

There is a lot of confusion in these two areas that you must avoid. When comparing the rated SPL of drivers, you must be careful. The efficiency of the driver (acoustic power output versus electrical power input) is defined in sensitivity as a certain SPL at 1W input at 1m distance (dB/W/m). This is the value you need to consider when thinking about the input power required for a given SPL output.

Many drivers are given a voltage sensitivity rating of a certain SPL at 2.83V input at 1m distance (dB/2.83V/m). This rating is very useful for matching drivers in a multi-way system, but is sometimes used by manufacturers to make their driver appear to have a higher efficiency. An 8Ω driver will have the same SPL in dB/W/m and dB/2.83V/m, since 2.83V will deliver 1W to an 8Ω driver. However, a 4Ω driver rated at 90dB/2.83V/m will really have a power sensitivity of 87dB/W/m, since 2.83V will deliver 2W to a 4Ω driver. Watch for this confusion factor when comparing driver efficiencies.

In the "good old days" when most speaker conceptual work was developed, the measurement of true power into a device with a complex impedance, such as a driver, was difficult. Thus most developmental work was done with "reference power," which is the power an input voltage delivers to a fixed resistor; normally the voice coil DC resistance (Re) or the nominal impedance value is used. This may upset some people, but in practice it makes things easier to study and to understand. Thus if a driver has a given rated input electrical power (Pe), its maximum input in the high-frequency range will be a straight line at Pe.

You know this is not actually true because the input impedance varies, but it is handy for discussing the maximum power capability versus frequency for a driver. Generally, when using reference power the input power is labeled Per and the output power labeled Par. This

TABLE 2
DRIVER THIELE/SMALL PARAMETERS

NO.	DRIVER	F_S	V_{AS}	Q_{ES}	Q_{MS}	Q_{TS}	R_E	EBP	SPL
1	Eminence 12LF	45.0	2.40	0.45	7.28	0.42	6.06	100	97
2	Eminence Pro-12	37.0	4.27	0.25	6.93	0.24	5.46	148	98
3	Dayton Series II	34.0	2.78	0.33	9.50	0.32	6.2	103	92
4	Pyramid Super Pro	30.0	3.89	0.36	7.98	0.34	3.1*	83	93
5	Audax PR240ZO	27.0	3.23	0.49	9.35	0.46	6.2	55	89
6	Dayton Pro Sound								
7	Dayton Treated Paper	28.0	5.11	0.42	2.94	0.37	6.6	67	92
8	Martin Sound Poly	18.0	7.49	0.82	2.39	0.61		22	87
9	Hi-Vi D10G	22.0	5.88	0.50	2.23	0.41	6.7	44	88
10	Peerless 850146	22.6	5.10	0.40	2.56	0.35	5.5	57	88
11	Peerless 831727	22.3	4.81	0.38	2.62	0.34	5.4	59	88
12	Vifa M26WR-09-08	26.0	4.59	0.36	2.82	0.32	5.9	72	89
13	Focal 11K7512	38.0	1.83	0.43	1.97	0.35	5.75	88	90
14	Phoenix Gold QX108	32.9	2.07	0.60	6.27	0.54	6.78	55	87
15	Peavey TKO 80	52.0	1.07	0.78	3.30	0.63	6.71	67	89

Notes: f_S in Hz. V_{AS} shown in cubic feet. Re in ohms. EBP = Efficiency bandwidth product. SPL in dB/W/m, converted from dB/2.83V/m if needed. *Indicated value was estimated as not shown in catalog. Values for driver #15 are measured/calculated values for original TKO 80 driver.

TABLE 3
BOX DESIGNS AND LARGE SIGNAL PARAMETERS

NO.	DRIVER	BOX	F_B/Q_{TC}	F_3	ANOMALY	41Hz	X_{MAX}	P_E	CONE DIA
1	Eminence 12LF						4.80	300	9.6
2	Eminence Pro-12						4.80	400	9.6
3	Dayton Series II						5.50	250	8.0
4	Pyramid Super Pro	VB				–6dB	3.70	300	9.6
5	Audax PR240ZO	CB				–4dB	4.00	100	8.7
6	Dayton Pro Sound								
7	Dayton Treated Paper						4.50	80	9.6
8	Martin Sound Poly	VB						150	9.6
9	Hi-Vi D10G	CB	0.98	42			8.25	150	8.8
10	Peerless 850146	CB	0.78			–5dB	9.00	200	8.05
11	Peerless 831727	CB	0.75			–5dB	9.00	220	8.15
12	Vifa M26WR-09-08	VB		32		–6dB	6.50	130	8.5
13	Focal 11K7512	VB		42		–5dB	9.00	175	8.8
14	Phoenix Gold QX108	CB	0.89	45		0.7dB	11.60	150	8.1
15	Peavey TKO 80 (measured)	CB			61	–21dB		80	9.6

Notes: f_B/Q_{TC}: for a VB f_B is shown and for a CB Q_{TC} is shown. f_3 in Hz. Anomaly is peak or ripple in dB. 41Hz indicates how far down the response is at 41Hz. X_{MAX} is in mm. Pe is electrical power limit in watts. Cone diameter in inches is estimated as 80% of outside diameter.

article uses reference power when discussing the maximum power capability of the drivers.

CLOSED-BOX VERSUS VENTED-BOX CONSIDERATIONS

One of the prime considerations in the use of a CB or VB is that a CB will have a larger net box volume. The requirement is that the outside dimensions of the TKO 80 enclosure not be changed. Thus any port must be internal to the box and will reduce the net box volume. To do 41Hz at a reasonably high level will require a large port area. Several articles have appeared in *aX* recently demonstrating that the port area must be larger than had commonly been thought. This would greatly reduce the available net box volume if a VB design is used.

There is some confusion over CB versus VB efficiency that must be clarified. Everyone is aware that in theory the VB has a 3dB efficiency advantage over the CB, as it uses output from both sides of the driver cone. Many people mistakenly believe this means a given driver will show a 3dB sensitivity advantage in a VB over use in a CB. This is not true.

The reference efficiency is designed into the driver, and at high frequency the driver will show this efficiency in any direct radiator application. The box type and alignment can change the shape of the driver response at low frequency, but at the top end of the driver response the SPL level will be the same. To increase the SPL you must go to some loading technique such as the use of a horn.

So where does the 3dB advantage of a VB appear? With a given driver it will show up as a smaller required box size or a lower f_3 compared to the same driver in a CB. Don't be confused here.

If you can develop a VB design in the

TABLE 4
LARGE SIGNAL PERFORMANCE PREDICTIONS

		MAX AT HF			MAX AT 41HZ			OTHER LIMIT		
NO.	DRIVER	SPL	P_{AR}	P_{ER}	SPL	P_{AR}	P_{ER}	SPL	P_{ER}	Hz
1	Eminence 12LF									
2	Eminence Pro-12									
3	Dayton Series II									
4	Pyramid Super Pro	116	2.5	300	108	0.4	200	106	58	48
5	Audax PR240ZO	108	0.4	100	95	0.02	13	102		60
6	Dayton Pro Sound									
7	Dayton Treated Paper									
8	Martin Sound Poly									
9	Hi-Vi D10G	109			103		80			
10	Peerless 850146	111		200	102		75			
11	Peerless 831727	110		220	105		80			
12	Vifa M26WR-09-08	112		220	105				98	75
13	Focal 11K7512	113		175	108		175	111	175	50
14	Phoenix Gold QX108									
15	Peavey TKO 80 (measured)									

Notes: Max at HF shows the maximum values in the upper frequency range. Max at 41Hz shows the maximum values at 41Hz. Other Limit shows maximum values for any limiting frequency between 41Hz and the upper frequency range. SPL is dB/m level. Par is the acoustic power output. Per is the reference electrical power input.

available net box volume, it may produce a higher output SPL capability at low frequency than is available with a CB. At the higher frequencies the SPL limit is normally set by the electrical power rating of the driver (Pe). As frequency is reduced you will reach a point where the limiting factor is the linear driver displacement capability (X_{MAX}). With the VB, the driver is highly damped near the box tuned frequency (f_B) and thus has a low cone displacement.

Thus by proper selection of the VB alignment, you can develop a design in which the port is doing the radiating down to near the 41Hz lower limit. Such a design would have a much higher output SPL capability at low frequency, provided the port is large enough to sustain this SPL level.

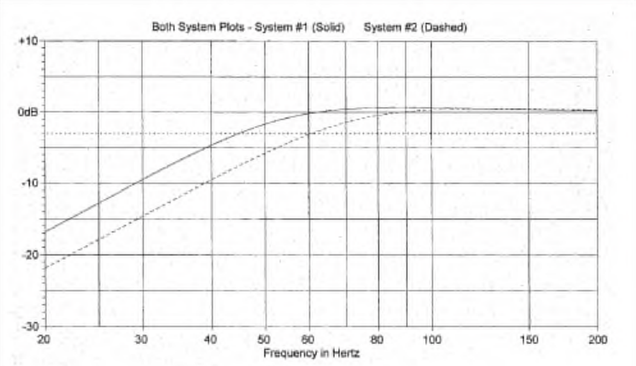


FIGURE 11: Box design software plot of QX108 versus stock 12″ driver. B-2247-11

TABLE 5
QX108 RECOMMENDED ENCLOSURE DESIGNS

TYPE	CLOSED-BOX	VENTED BOX			
	V_B	V_B	F_B	PORT DIA.	PORT LENGTH
"Musical"	1.50	2.25	35Hz	3″	3.75″
"SPL"	0.60	1.75	45Hz	3″	2.25″

EQUALIZATION

If you can't develop a design that is sufficiently flat down to 41Hz, it is possible to correct the response somewhat via equalization. You need to keep in mind that you may be modifying the response in a region where X_{MAX} is setting the output power limit. This is almost a certainty with a CB design. If this is true, then any equalization may reduce the maximum average drive level you can use, thus reducing the output capability above the equalized frequency range.

SOFTWARE

Small-signal analysis was done on privately developed software based on the equations published by Dr. Benson[1], whose work includes consideration of the Qs associated with the port, box leakage, and box absorption. All VB analysis was done using the default Qs that are equivalent to a total box Q of 7. Benson's equations have the advantage that they plot the responses of the cone and port independently for a VB. This allows manipulating the design to make the port the major radiator in the low-frequency range.

VB designs were generally analyzed via non-optimum design techniques at the specified net box volume. Initial work ignored volume loss to the port. CB designs were via very high Qs emulating a lossless box as is normally done in CB design. The available net box volume was entered and resulting system Q_{TC} noted. The desire was to maintain the final Q_{TC} below about 0.9.

Large-signal analysis was also done on privately developed software based this time on the work of Dr. Small[2]. This software worked with reference power and predicted the power sensitivity (dB/W/m) for comparison with the driver specification. This software can plot the maximum input power capability versus frequency and identify whether electrical power (Per) or displacement (X_{MAX}) sets the limit. The software also plotted the maximum acoustic power output capability (Par) versus frequency and showed the corresponding SPL level at 1m. The reference used is 1W acoustic producing 112dB/m.

Note that one critical parameter in establishing the maximum power capability is the cone area, entered into the software as cone diameter. This parameter is normally not published in catalogs, so was taken as 80% of the published driver outside diameter. If you are close to the true cone diameter, then the computed sensitivity will agree with the driver specification, so this was checked. In most cases agreement was good, but the large signal analysis should be taken as an "indication" and not a "guarantee."

PHOTO 4: Enclosure routed for adapter ring.

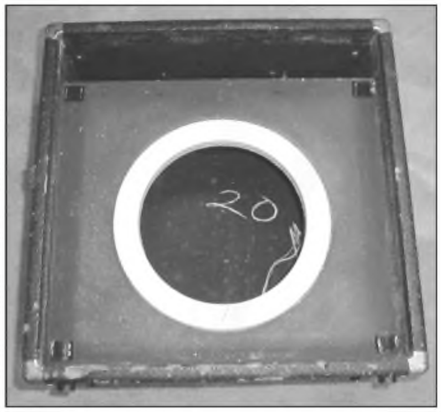


PHOTO 5: Adapter ring fit check.

PHOTO 6: Adapter ring installed and painted, Tolex removed.

Again, remember that this work is based on reference power, not actual power delivered to the driver. If you are using the driver at a power level where this difference becomes significant, you are flirting with disaster.

RESULTS OF DRIVER STUDY

We evaluated a variety of 10″ and 12″ driver data from various catalogs to find an acceptable one. (We found no acceptable 8″ drivers.) *Table 1* identifies the units that seemed to offer a possibility, along with data on the original 12″ driver in the TKO 80 (No. 15). *Table 2* shows the T/S parameters listed for each driver and the computed efficiency bandwidth product.

When the driver sensitivity was rated in dB/2.83V/m it was converted to dB/W/m based on the nominal driver impedance. The T/S parameters for the original 12″ driver were calculated from free air and CB measurements with a 1mA constant current source.

The box designs used with each driver are shown in *Table 3*. If it is a CB design, then the total system Q (Q_{TC}) is shown. For a VB design the box tuned frequency (f_B) is shown. Remember in all cases the box volume is fixed at 1.24ft^3. The design f3 along with the amount the response is down at 41Hz, as read from the plot, is shown.

Table 3 also includes the driver parameters required for large-signal analysis. Keep in mind that the cone diameter is not shown in the catalogs and was taken as 80% of the driver diameter.

Table 4 shows the results for large-signal analysis without any equalization of the frequency response. The maximum Per that may be used and the expected Par and SPL for this input are shown for three cases. The first case is at high frequency where the limit is set by electrical power constraints. The case for the minimum needed frequency of 41Hz is also shown.

In many cases a VB will have a region of limitation below that at 41Hz and, in such cases, that information is also shown. For a CB design some intermediate frequency above 41Hz may be shown for additional information.

None of these drivers quite makes it "flat" (above –3dB) down to 41Hz, but several are within a couple of dB. At this point we started to factor in other criteria. For example, I have used drivers from this vendor before and thus had some experience that they meet their published specifications. Also, you must consider price. These all become weighed along with the large-signal analysis results to select a driver.

The best candidates were deemed to be numbers 10, 11, and 14. All of these are nominal 10″ drivers, which give better performance in a 1.24ft^3 box than the 12″ drivers we located. Note also that all of these are CB designs, removing the requirement to make any allowance for volume lost to a port and thus a redesign effort as a VB.

Weighing all the variables, we decided to go with driver #14, the Phoenix Gold QX108 10″. The deciding factor was that this is a subwoofer driver with an X_{MAX} of 11.6mm (almost ½″) and thus designed for large excursion work. It would be capable of the required output level even if equalization was used and certainly can sustain all the power the amplifier is able to produce over the application frequency range.

Figure 11 shows a comparison of the CB performance of the QX108 versus the stock 12″ TKO 80 driver.

TABLE 6
PHOENIX GOLD QX108 MEASURED PARAMETERS

PARAMETER	MEASURED	SPECIFICATION
Re	6.42Ω	6.78Ω
f_S	34Hz (Z = 60.7Ω)	32.9Hz
Q_{MS}	4.18	6.27
Q_{ES}	0.49	0.60
Q_{TS}	0.44	0.54
V_{AS}	2.08	2.07
f_C	55Hz (Z = 61.1Ω)	
Q_{MC}	6.93	
Q_{EC}	0.81	
Q_{TC}	0.73	

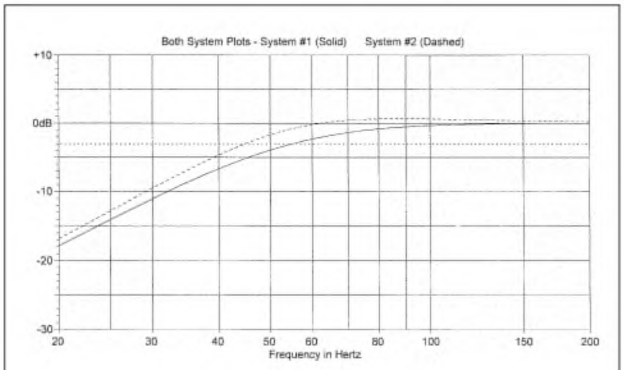


FIGURE 12: Box design software plot of QX108 actual versus measured data. B-2247-12

MANUFACTURER'S RECOMMENDED ENCLOSURES

The Phoenix Gold QX108 driver came with an instruction sheet that recommended some enclosure designs (*Table 5*). The instruction sheet said that "Musical" was a box of superior sound quality, and "SPL" was a higher output with less bass. Don't be confused by this; as you know, the SPL is designed into the driver, and all these enclosures will show the same SPL level at the upper frequencies.

We evaluated these designs via privately developed software based on the work of Dr. Small. This software will design ports for a VB based on the classic equations for a bare (unlined) box, but will also design a more realistic port length for a box with lined walls based on many years of studying this topic by David Weems. Based on these design evaluations with the published driver specifications, we decided to forgo the recommended designs and use the initial CB design with the Phoenix Gold QX108 driver.

MODIFICATIONS TO MOUNT THE NEW DRIVER

Experienced speaker builders are, no doubt, familiar with this section of the article. You need an adapter ring to allow mounting the 10″ QX108 in the opening occupied by the stock 12″ driver. First, we routed the enclosure cutout with a ⅜″ rabbet to accept the ring (*Photo 4*), cut the adapter ring from high-quality ¾″ plywood, and routed it with a rabbet to fit the one in the enclosure. A fit-check of the adapter proved it to be tight and accurate (*Photo 5*).

Next, we drilled the adapter for the existing 12″ driver bolt pattern, and for the QX108 bolt pattern, with the holes 22.5° apart. We painted the adapter flat black to match the front of the TKO 80 cabinet and installed eight 10–32 screws held by T-nuts. The speaker leads go through a hole in the cabinet on the amplifier shelf. A little silicone rubber sealed the hole to make the enclosure airtight. I added acoustic foam to both sides, the top, and the rear of the interior of the cabinet.

Now that the woodwork was finished, we removed the tattered vinyl covering (*Photo 6*) and glued new black Tolex™ to the bare wood cabinet. We removed the torn grille cloth from its frame, replaced it, and ordered new metal corners and rubber feet to finish off the cabinet restoration.

With the QX108 in hand, we ran it in for 12 hours at its specified resonance frequency, then calculated its actual T/S parameters based on free air and CB measurements with the 1mA constant-current source. We used the HP 339A oscillator to drive the current source box, for a pure sine wave rather than the 3% THD the function generator produces at low frequency. The f_S measured 35Hz with a resonance impedance of 61.5Ω.

After free-air testing, we installed the QX108 with the cone facing into the TKO 80 enclosure. With this method, you can make voltage measurements right at the driver terminals, eliminating the added resistance of the enclosure's speaker leads. In order to ensure accurate CB results, you must add the air volume in front of the cone (0.0128ft^3) to the 1.24ft^3 enclosure volume.

Before measurements, Chuck performed the leak test that I suggested, with a 1.5V battery applied to the speaker leads. In this test, the driver will take an initial jump, then travel the last 10% of its displacement in about a second if there is no adverse air leakage from the enclosure. This ad-hoc air leak test turned out to be satisfactory.

Next, we measured the CB parameters to get V_{AS}. This requires some patience. It takes about two minutes for the mV readings to settle to a final value, always increasing upward toward the stable number.

With the driver removed, we re-measured the free air data. We found f_S dropped to 34Hz and 60.7Ω. After cranking the data through the spreadsheet, the calculations revealed the parameters shown in *Table 6*.

Welcome to the world of driver variations. I ran this measured data versus the spec data in the design software. The f_3 is 54.9Hz versus 44.8Hz using the spec sheet data. The curve looks flatter using the measured parameters, but the SPL is 2dB lower than the published data predicts. Are these discrepancies between measured data and the QX108 specification within the acceptable error ballpark?

Note the large difference in the Q_{MS} measurement from the QX108 spec. Q_{MS} is the Q due to mechanical resistances, the losses in the surround and rear suspension. These losses vary with time, temperature (a lot), and humidity. Thus that parameter is very variable. Fortunately, it has only about a 10% effect on Q_{TS}, so the variations are not a problem. The Q_{MS} value does vary in the testing, but the accepted approach is to make it a "don't care" condition.

LISTENING TESTS—QX108 SUB

The next test was the important one: how does it sound?

Once we connected and installed the amplifier, there was plenty of volume. However, there was an audible peak at A_3 (55Hz) and a slight bit of doubling from E_3 41Hz (82Hz doubling) to F$\#_3$ 46Hz (93Hz doubling). This enhances

(to p. 67)

The SOAP Factor

Check your power amp's performance with this handy method to estimate clipping power by measuring the quiescent DC supply.

By Benjamin L. Poehland

It's 7:10 PM on a typical weeknight, and I'm comfortably esconced on the sofa watching "Nightly Business Report." Piped through a ten-band equalizer and an external power amp, the voice of anchorman Paul Kangas is clean and crisp, in sharp contrast to the content of the day's business news, which is anything but clear. My mind wanders as I contemplate the wisdom of bailing out of a mutual fund while there are still a few pennies left in it to save.

BOOM-BOOM CARS

From some distant quarter I sense a faint throbbing. Aw geeze, is the stock market giving me another headache? No . . . this is real.

The throbbing grows nearer, louder, and is now well defined: boom-boom-bonk, boom-boom-bonk! I sit up and glance instinctively at the heavy drapes concealing a large picture window in my living room. BOOM-BOOM-BONK! The huge glass pane is resonating, adding its own buzz to the annoying throb invading my home through walls and drapes.

Then, mercifully, the phenomenon gradually subsides and fades into the night, leaving me once again in peaceful contemplation of my investment decisions. It was just another annoying boom-boom car.

ABOUT THE AUTHOR

Ben is a retired research scientist with eleven publications in the field of natural product drug discovery and molecular structure elucidation. No stranger to these pages, Ben has contributed a number of articles on audio restoration to *TAA* and *Speaker Builder*. His first audio project was the assembly of a Dynaco ST-400 kit in 1977, and he's been addicted to "roll your own" audio ever since. His personal motto is "Give me Ohm's Law and a soldering iron, I can do anything!"

IMPOSSIBLE POWER

The boom-boom car phenomenon began in the mid-1980s, grew exponentially in the 1990s, and today is a standard feature of our roadways. We've all experienced these vehicles, typically driven by a male in his late teens or 20s, whose mating call is the thunder of monster subwoofers and the screech of rubber tires on asphalt. My house is a good 50′ from the street, yet the pulsating power of a boom-boom car is enough to rattle my doors and windows, and can easily be heard over a block away even when the offending vehicle has its windows closed.

The audio power available in automotive stereo systems today is incredible. Systems rated 50W RMS per channel, even 100W or more, are readily available. My first encounter with a boom-boom car left me in a state of shock and bewilderment. Being an educated man, I knew such power levels were quite impossible, and I could prove it.

As a young audio enthusiast in the 1970s, I took an evening course in audio electronics at a local community college, taught by a colorful engineer with a day job at a posh audio salon. Car stereo was a subject of keen interest to the class, but we were all rather disappointed to learn the maximum power you could ever hope to generate from car speakers was a whopping 6W/channel. Enough to annoy your passengers and perhaps the car in front of you, but nowhere near enough to flatten the neighborhood with a sonic boom.

THE IMPOSSIBLE PROOF

The arithmetic of Ohm's law is undeniable. You start with the basic relationship for power in DC circuits:

$$P = V^2/R \quad (1)$$

where P is power in watts, V is DC voltage, and R is DC resistance in ohms. Since we're dealing with AC waveforms and want power expressed as RMS (root-mean-square, the AC equivalent of DC power), some adjustments are necessary. First, substitute speaker impedance Z in place of DC resistance. Then substitute RMS volts for DC volts. That relationship is

$$V_{RMS} = 0.707V_{DC}/2, \text{ or } 0.3535V_{DC} \quad (2)$$

ignoring reactive effects and assuming the peak-to-peak excursion of the AC

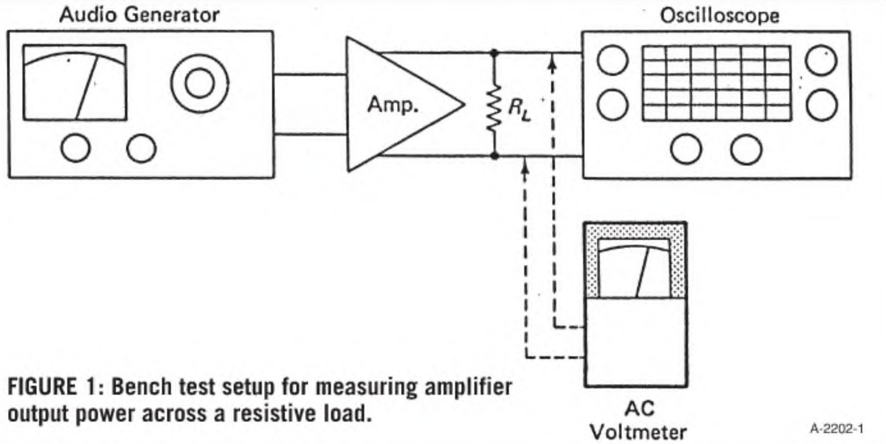


FIGURE 1: Bench test setup for measuring amplifier output power across a resistive load.

waveform oscillates from 0V to +V. (Actually, it doesn't, but I'll address that later.) Thus a simple relationship for RMS clipping power relative to DC power in AC circuits becomes

$$P_{DC} = (0.3535V_{DC})^2/Z \quad (3)$$

where P_{DC} is the RMS clipping power based on quiescent DC supply measurement V_{DC}.

Now plug in real-life values and verify the result. The nominal 12V DC electrical system in most automobiles actually measures around 14V, and typical automotive speakers have a nominal impedance value Z = 4Ω. Plugging these values into equation 3 and doing the arithmetic,

$$(0.3535 \times 14)^2/4 = 6.1\text{W}$$

gives the magic 6W value I learned in my class. I was so impressed by this exercise that for many years I routinely used it as a quick-and-dirty means of estimating the clipping output of power amplifiers simply by measuring the DC supply voltage.

Having thus demonstrated that no car stereo system can produce more than 6W, how is it that 100W car stereos are possible? Digital trickery, of course.

Coincident with the rise of the computer industry in the early 1980s came digital switching power supplies. Originally designed to provide high-current/low-voltage supplies to computer circuits without the cost and weight of a large power transformer, it was only a matter of time before the same technology would be applied to applications requiring high voltage derived from a high-current/low-voltage source (such as a high-powered audio system operated from a car battery).

Simply put, switching power supplies employ power transistors to drive a high-frequency/high-current oscillator from a low-voltage DC source (car battery) through a small but highly efficient step-up transformer to produce higher voltage at lower current, which is then rectified to provide the step-up DC supply for those monster car amplifiers. At the time I took my audio class, this technology was still in development and did not appear in the commercial audio market for another eight years.

PRESERVING THE CLASSICS

Following my work with Tigersaurus[1], my audio interests drifted increasingly toward the repair and restoration of classic solid-state audio gear. I define "classic" as equipment manufactured from 1968 through 1981, which I regard as the Golden Era of solid-state audio.

Numerous improvements and circuit innovations—many now routinely implemented in today's designs—were introduced during that time. Most of the equipment from that era employs discrete through-hole components (which are replaceable), as compared to the cheaper IC blocks and surface-mount parts in much of today's gear. And finally, the overall quality of construction and craftsmanship in equipment from that era puts to shame the flimsy plastic digital offerings that populate today's mass-market store shelves.

It grieves me beyond words to see so much of this older equipment callously tossed into landfills. An entire generation of audio innovation is disappearing, adding insult to injury by contributing to the pollution of our environment. I can't save them all, but I do what I can.

Over the past few years I've begun participating in local municipal recycling projects, and I'm proud to say that I've personally rescued dozens of classic pieces from the heartless clutches of landfill operators. My own collection of vintage gear has expanded nicely, and I make the effort self-supporting by offering surplus items at the occasional yard sale or flea market.

WAKE-UP CALL

My initial encounter with boom-boom cars jolted me only briefly. Once I learned that their apparently impossible performance didn't violate any laws of nature, I confidently resumed my usual practice of employing equation 3 to estimate power amp output from DC supply measurements.

At a recent sale of surplus restored amplifiers, a pugnacious buyer informed me that a particular Vector Research receiver I was offering delivered only 30W, while I was claiming 50W based on DC supply measurement using $Z = 8\Omega$ in equation 3. The gentleman initiating this conflict being an obnoxious know-it-all, I felt motivated to prove him wrong and settle the issue by making a standard bench test using a verified 8Ω resistive load and an AC signal source. The result was a wake-up call.

We were both wrong. My bench test indicated the amp delivered 40W at the onset of clipping, so my claim was off by 20%. I was vaguely aware that my DC-estimation technique was a little sloppy, but I didn't think it was *that* sloppy.

After making my DC-based power estimate, I'd typically try to throw in a fudge factor to account for the fact that peak-to-peak excursions never actually reach the maximum limit defined by the DC supply rails. Stated mathematically,

$$V_{p\text{-}p} < |V_{DC}| \qquad (4)$$

so the 6W value for my car stereo exercise is truly only a theoretical maximum. In practice, it is less.

Equation 4 represents the real-life situation for virtually all audio power designs. Designers always allow a "safe operating area" (SOA) that takes into account the voltage, current, and thermal specifications for the output transistors, as well as constraints imposed by the Federal Trade Commission in making power claims for a given level of distortion. Not to mention, any amplifier whose output amplitude spanned the entire range of the DC supply would likely be so unstable as to oscillate itself to death.

GOING FOR SOAP

What the altercation over my Vector Research amp revealed was that I never knew exactly what fudge factor to use to account for the SOA, or SOAP factor as I eventually came to call it (Safe Operating Area Power). Design engineers determine the SOAP limits, and they are human beings whose personal philosophies span the range of humanity in general. Your straight-laced conservative types might limit SOAP to no more than 50% of the available DC, while wild-west buccaneer types might push the envelope to the very edges of the DC supply. While there are design equations that govern SOA, you can't always be sure how rigorously the engineer applied them. There's no way short of an actual bench test to find out.

The uncertainty gnawed at me. I finally decided the only way to resolve the issue was to perform DC supply voltage and AC power measurements on as many amplifiers as possible in an effort to obtain a statistical value to plug into equation 3 to make it more realistic, i.e.,

$$P_{SOAP} = C_{SOAP}(P_{DC}) \qquad (5)$$

where P_{SOAP} is the estimated RMS clipping power of the amp based on simple quiescent DC supply measurement, P_{DC} is the value from equation 3, and C_{SOAP} is a statistical constant that takes the safe operating area, or SOAP factor, into account. I needed measurements on enough different amps to give reasonable statistical validity to the SOAP factor. My growing collection of vintage gear proved the perfect resource for conducting a study of this nature.

At this point you might wonder, why bother with this? Why not perform an actual bench test on every unit and be done with it? Mainly because bench testing is a pain in the neck, and most households are not equipped with a full-fledged test bench. Taking a DC supply measurement is something I can easily do on the kitchen table using only a portable DMM and a screwdriver to loosen the cover.

Whereas a bench test requires me to lug the amp up several flights of steps to my bench, hook up a lot of wires, fire up three pieces of test gear, follow a tedious test procedure, and pray that I don't destroy the test unit through some mishap. (Once you've blown up a 250W amp on the bench, the experience stays with you[2].) If you can establish a reasonable relationship between quiescent DC supply voltage and RMS clipping power output, you'd have the basis for a very handy "field test without the mess."

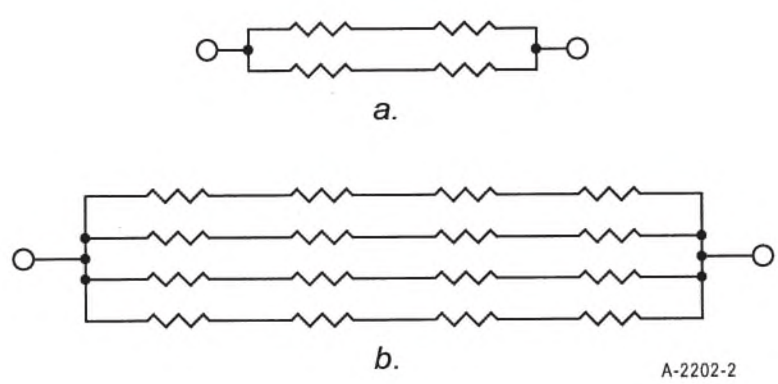


FIGURE 2: Practical 8Ω load resistor networks for audio power testing. a. 80W load; b. 320W load. All resistors are 8Ω/20W 5% non-inductive (see text). A-2202-2

A MEASURE OF AMPS

Figure 1 shows the basic test bench setup I used for testing all the amplifiers listed in *Table 1*. Just about any oscillator will suffice, as long as its distortion is below 2% and it has a variable output. My oscillator is a Leader LAG-120B. You can use any oscilloscope, though a dual-channel unit with bandwidth >1MHz is preferable. In my setup I use the Ramsey BS-601[3] with an extra parallel connection directly from the oscillator output to channel B of the scope, giving me a direct onscreen comparison of input and output waveforms during the test.

You can use any decent AC voltmeter in the test setup. While a true-RMS meter is preferred, older meters whose circuitry is average-responding but calibrated for RMS readout is acceptable since you are measuring essentially pure sine waves. The Sabtronics 2010A DMM I used in this study falls into the latter category. To allay any paranoia, I periodically check the RMS calibration of the Sabtronics meter with a portable Beckman Tech-360, which is a true-RMS instrument.

Load resistor R_L is a critical component that you cannot omit from the test setup. Its nominal value should be 8Ω, and you should carefully verify its exact value with a quality ohmmeter or DMM, taking test lead resistance into account. My test load is a custom unit of my own design, measuring 8.0Ω, with a trim switch to permit fine-tuning the nominal value to account for the 10% resistors I used in its construction. My load is rated 240W, the resistive elements mounted in a fan-cooled enclosure with heavy banana jack connectors and 14-gauge wiring throughout.

My load box is rather a kludgy affair, designed and built some years ago when the necessary resistors weren't readily available. Today you could build your own high-power audio load quite easily using the schematics in *Fig. 2* as a guide. The necessary 8Ω/20W resistors are readily available at Radio Shack (catalog #271-120, $1.49) and have the added virtue of being non-inductive.

All the amplifiers used in this study were thoroughly cleaned, repaired, subjected to listening tests, and verified fully functional prior to testing. All pots, switches, and jacks were treated with Caig Laboratories (website: www.caig.com) DeoxIT-D100L (reformulated Cramolin CR-10[4]). Only one test was made of each channel, separately, at a single frequency of 1kHz. The AUX input was used on integrated amps and receivers, with volume and balance at midpoint and all tone controls set flat.

I began my test procedure by first making all the connections and then allowing at least a ten-minute warm-up for both the measuring instruments and the unit under test. I then slowly increased the output level from the oscillator while observing the waveform on the oscilloscope.

As soon as the waveform began to clip, I backed off the oscillator and made fine adjustments until I judged the output signal to be just on the verge of clipping. At that point I recorded the reading on the AC voltmeter. I repeated this process for the other stereo channel. In virtually every case the outputs of the two channels were within a few percent of each other, so I averaged the two readings to report a single value for each amp.

TABLE 1
VOLTAGE AND POWER DATA FOR 30 AMPLIFIERS

BRAND	MODEL	VINTAGE	DESIGN	VDC	V_{RMS}[4]	P_{RMS}[6]	P_{DC}[7]	P_{RMS}/P_{DC}	P_{SOAP}[8]	% ERROR[9]
H/K	SC-2520	1968[3]	QCS	+ 55	14.4	26	47	0.55	28	+ 7
Lafayette	LA-125TA	1968	QCS	− 63	16.3	33	62	0.53	37	+ 11
Dynaco	ST-80	1969	QCS	+ 65	18.5	43	66	0.65	39	− 10
Allied	498	1970	QCS	+ 82	23.3	68	105	0.65	62	− 10
Dynaco	ST-120	1971	QCS	+ 75	23.6	70	88	0.80	---	---
Sansui	55001	1972	QCS	+ 90	23.4	68	127	0.54	75	+ 9
Marantz	2230	1972	QCS	+ 67	18.4[5]	42	70	0.60	41	− 2
JVC	VN-700	1972	FCS	± 34	18.4	42	72	0.58	43	+ 2
Sony	STR-7055A	1974	FCS	± 41	22.1[5]	61	105	0.58	63	+ 3
Technics	SA-5550	1974	FCS	± 47	26.5	88	138	0.64	83	− 6
SWTPC	210/A	1976	FCS	± 75	46.0	265	351	0.75	---	--
Onkyo	TX-2500	1977	FCS	± 32	17.9	40	64	0.62	38	+ 5
Fisher	RS-1058	1977	FCS	± 56	29.8[5]	111	196	0.57	118	+ 6
Pioneer	SX-450	1977	FCS	± 26	14.2[5]	25	42	0.60	25	0
Marantz	2230B	1977	FCS	± 34	18.6[5]	43	72	0.60	25	0
Marantz	2226B	1977	FCS	± 31	17.3	37	60	0.62	36	+ 3
Nikko	NR-715	1979[3]	FCS	± 35	19.6	48	77	0.62	46	− 4
Vector Rsch.	VR-2000	1980[3]	FCS	± 33	17.8	40	68	0.59	41	+ 2
JVC	R-S33	1980	FCS	± 38	20.3	52	90	0.58	54	+ 4
H/K	hk385i	1986	FCS	± 39	21.2	56	95	0.59	57	+ 2
Pioneer	VSX-9500S[1]	1991	FCS	± 70	37.5	176	306	0.58	184	+ 4
Pioneer	VSX-9500S[2]	1991	FCS	± 40	22.0	61	100	0.61	60	− 2
Technics	SA-400	1978	DIC	± 42	21.9	60	110	0.55	61	+ 2
Technics	SA-202	1980[3]	DIC	± 33	18.0	41	68	0.60	37	−11
Onkyo	TX-1500MkII	1978	CIC	± 27	14.5	26	46	0.57	25	− 4
Sony	HST-48	1978	CIC	± 27	13.5	23	46	0.50	25	+ 8
Sansui	R-611	1985	CIC	± 40	20.5	53	100	0.53	55	+ 4
Philips	FR-920[1]	1992	CIC	± 48	25.8	83	144	0.58	79	− 5
Technics	SA-GX100	1991	CIC	± 43	22.1	61	116	0.53	64	+ 5
Technics	SA-GX490[1]	1995	CIC	± 67	34.3	147	280	0.52	154	+ 5

Notes:
1. Main channels only.
2. Rear channels only.
3. Estimated date.
4. Measured at onset of clipping, 1kHz, RL = 8Ω, avg. of both channels.
5. Inverted output.
6. $(V_{RMS})2/RL$
7. Using Eq. 3, Z = 8Ω.
8. Using Eq. 5, with C_{SOAP} values from Table 2.
9. $[(P_{SOAP} - P_{DC})/P_{SOAP}]$ x 100

POINTS TO PONDER

As I dove into the daunting task of power measurements on 30 amplifiers, I was confronted with some issues I hadn't previously considered. Foremost among these was dealing with four different types of BJT (bipolar junction transistor) audio power designs: quasi-complementary symmetry (QCS), full-complementary symmetry (FCS), discrete integrated circuit (DIC), and combined integrated circuit (CIC). The basic differences between these designs are displayed in *Fig. 3*. Not wishing to place myself in the position of comparing apples to oranges, I segregated the data according to the type of design.

Another issue was how to account for amplifiers with split power supplies in comparison to units with single-rail supplies. In making the theoretical calculations, I believed it was legitimate to simply consider an amp with split supplies of, say, ±30V to be the same as an amp with a single-rail supply of +60V. A peak-to-peak waveform excursion between −30V and +30V in a split-supply system is no different from a similar waveform oscillating from 0V to +60V in a single-rail system.

One amp surprised me with a negative single-rail supply voltage. Again I simply treated that as though the supply were positive. Also, several amplifiers exhibited phase inversion. I made note of that, but it doesn't affect the data.

Obviously I didn't perform all these measurements in a day. This project was strung out over a period of several weeks. As I refined my ideas and techniques, I found it prudent to re-test a few amps I had tested earlier.

I was annoyed to discover that DC supply measurements—and the corresponding power measurements—varied somewhat from day to day and even varied according to the time of day. All except one of these amps have unregulated power supplies, so consistency from one day to the next is no better than what my local electric utility provides from the wall socket. Fortunately, I was seeking ratio relationships in this study, which renders my results relatively immune to daily fluctuations in AC line voltage.

Obtaining consistent measurements of the DC supply voltages drove me nuts. Even after a ten-minute warm-up, they all fluctuated somewhat. The range of fluctuation was typically 0.2V to 1.0V, the variance being greater on the higher-powered units.

To preserve my sanity, I recorded the DC readings as nominal values to the nearest volt. This annoyance occurred in all the test units, and was most likely caused by ripple currents, supply noise, and minor variations in AC power. DC line noise probably does contribute a trivial amount of error to my results, but since the relationship I'm exploring is rough to begin with, I have no problem ignoring it. In similar fashion I've rounded off the calculated power figures to the nearest watt.

As I began working with these amplifiers I noticed that none of them clipped symmetrically. Virtually all of them commenced clipping either at the top or bottom of the waveform, but not both halves simultaneously. For the purposes of this study I elected to define "clipping" as the very first sign of peak flattening regardless of in which half of the waveform the flattening first appeared.

SOAP FACTORS REVEALED

The numbers in *Table 2* sum up everything I sought in this exercise. It turns out the SOAP factor for QCS and FCS amplifiers is about the same number, 0.6, while the SOAP factor for the IC amplifiers is somewhat less, 0.55.

There were anomalies. In *Table 1*, you see I omitted reporting a P_{SOAP} value for the Dynaco ST-120. As I examined the data for the QCS amps, the P_{RMS}/P_{DC} ratio for the ST-120 stood out like a sore thumb, 0.8, while all the rest of the QCS figures are closer to 0.6. If I had included the value for the ST-120 into the SOAP factor for the QCS amps, it would have skewed the QCS SOAP factor to a higher number and given large errors for the P_{SOAP} values for all the other QCS amps.

I instinctively thought the ST-120 should be omitted from the study, but not without justification. Is there something unusual about the ST-120 compared to all the other units tested? Indeed, there is. The ST-120 is the only amp in the test group that has a regulated power supply.

The designer of the ST-120 obviously was comfortable pushing the SOAP limit on this amp, because he was assured the power supply wouldn't exhibit any DC droop even at maximum output of both channels. Whereas, with the more common unregulated supplies, a more conservative SOAP factor is required because of anticipated DC supply sag at full power output.

While I was compelled to exclude the ST-120 from the results reported in this article, it doesn't mean there is no SOAP factor for that amp. It just means there is probably a SOAP factor, but it will need to be determined independently for amps with regulated supplies as a group. Since only one such amp appeared among my test units, it represents an insufficient statistical sample.

You'll note a similar situation pertains to the SWTPC 210/A amp when you look at the FCS data in *Table 1*. Again, the ratio is unusually high at 0.75, and I excluded the data for this amp in calculating the SOAP factor for the FCS amplifiers. I did so on the basis that the 210/A is the only monoblock amp in the crowd. Whereas all the other amplifiers must supply current demands to another stereo channel in addition to the one that's being measured, the 210/A has only one channel which permits the full capacity of the DC supply to be harnessed without any other drains.

In addition, the DC supply of this particular amp is stiffer than that of the original design, since this was a modified unit in which I enhanced the supply by increasing DC filtration to 40,000µF by adding extra caps. That gives the 210/A an unfair advantage over the other amplifiers, which are all stock units.

Regarding the data for the DIC and CIC amps in *Table 1*, I was initially tempted to separate them into two groups. However, the P_{RMS}/P_{DC} ratios for these amps are similar, and from a design standpoint the two groups are functionally equivalent. I therefore lumped all the IC amps together and derived a single SOAP factor for ICs as a group. With a value of 0.55, the SOAP factor for the IC amps is notably lower than that for the QCS and FCS units, which are discrete-transistor designs. This makes sense, on the basis of thermal considerations alone.

With the large power transistors scrunched up into the same package along with the more delicate devices used in the lower gain stages, the probability of thermal damage to the smaller devices is increased. The solution is to apply a more conservative safe operating area to the IC device as a whole, resulting in a lower SOAP factor compared to discrete designs where the high-current output devices are physically separated from the lower gain stages.

OTHER SOAP CLASSES?

As I considered the data for the 210/A, I wished I had more "super-amps" available to include in my study. The 210/A is a classic super-amp, and I wonder if such amplifiers might also constitute a separate group with its own SOAP factor. It would have been much more interesting if I could have included test data for an Ampzilla, Dynaco ST-416, Heath AA-1800, Bose 1801, McIntosh 2300, Adcom GFA-1, Carver M-400, or Phase Linear 400 in comparison to the 210/A.

There is yet another class of amplifiers deserving of a SOAP study: MOSFET amps. I do have a Van Alstine MOSFET 120 but excluded it from my study since I regard MOSFETs as a separate class; with only one unit I lacked a sufficient statistical sample. MOSFET amps incorporate the best features of tubes and transistors, but they're still distressingly rare. The Yamaha B-2, Acoustat Trans-Nova 200, and Hafler DH-200 come to mind, but it would be beneficial for audiophiles everywhere if there were as many MOSFET models as BJTs available for SOAP testing.

CLAIMS AND CAVEATS

At this point it's time to place this work in perspective and summarize what can, and cannot, be accomplished with quiescent DC supply measurements. On the basis of the work described in this article, I'm comfortable making the following claim: Using equation 5 and the SOAP constants from *Table 2*, it is possible to predict within an error of about 10% the RMS clipping power of a stereo amplifier using only a quiescent measurement of the DC power supply. It's a handy method for audiophiles,

TABLE 2
SOAP FACTORS FOR AMPLIFIERS

(STATISTICAL AVERAGE OF P_{RMS}/P_{DC})

DESIGN TYPE	C_{SOAP}
QCS	0.59
FCS	0.60
DIC, CIC	0.55

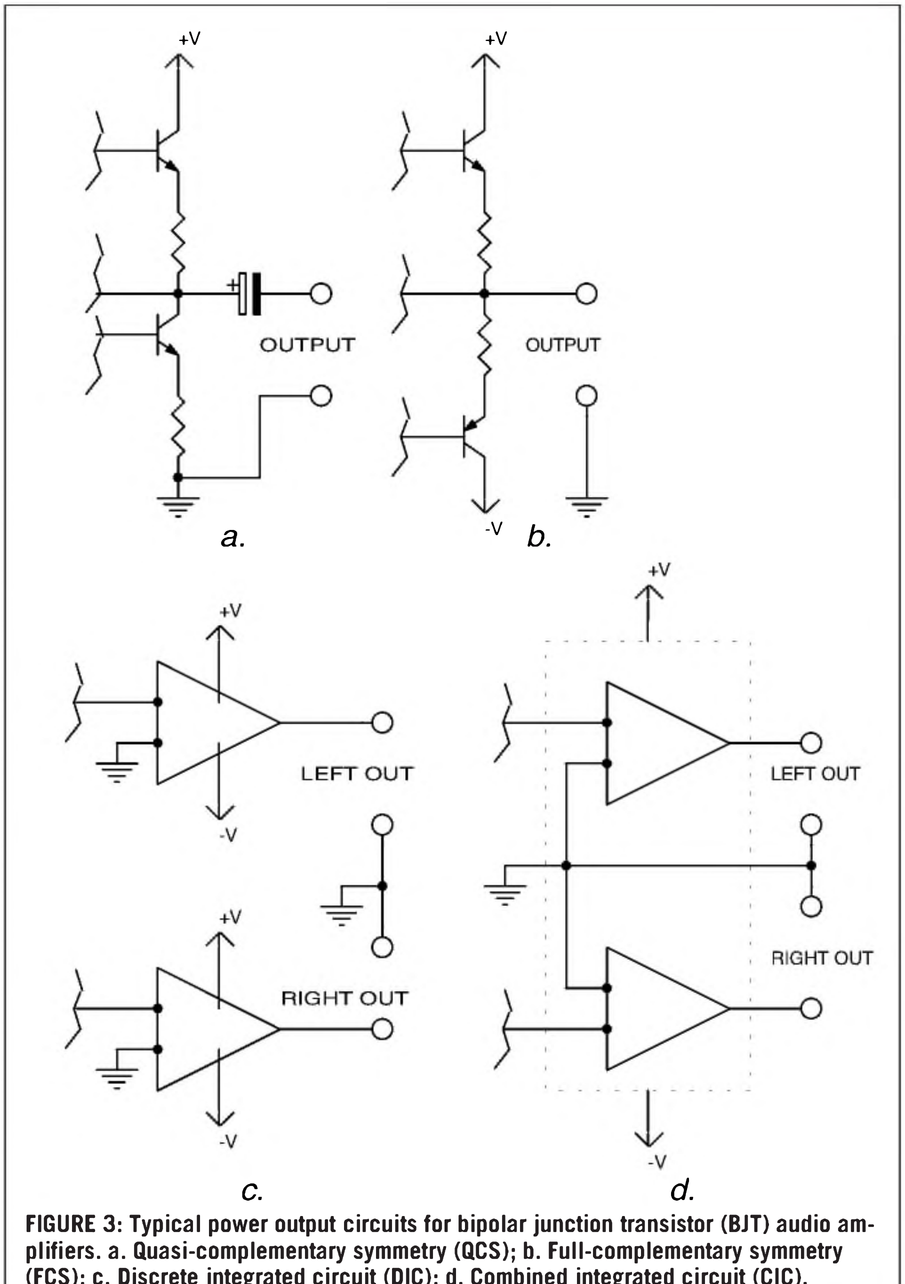


FIGURE 3: Typical power output circuits for bipolar junction transistor (BJT) audio amplifiers. a. Quasi-complementary symmetry (QCS); b. Full-complementary symmetry (FCS); c. Discrete integrated circuit (DIC); d. Combined integrated circuit (CIC).

technicians, and engineers to estimate the approximate output of a power amp that turns up with no schematics, manuals, or spec sheets, in circumstances where the only available test gear is a DC voltmeter.

Now for the qualifiers. For starters, the amplifier must be in good working order. It's senseless (and possibly dangerous) to make any kind of measurement on equipment with corroded switches, burnt fuses, leaking filter capacitors, or blown outputs. A power claim based on DC supply measurement cannot be used as a substitute for an actual bench test, especially if the unit is being offered for sale.

Equation 5 isn't presently valid for super-amps (>200W/ch), MOSFET amps, monaural power amps, tube amps, or amps with regulated power supplies. A DC-based power estimate doesn't tell you anything about the frequency response of the amp, nor does it give you an idea of what level of distortion to expect at the onset of clipping.

There's a final caveat I'll address separately. The number you generate using equation 5 most closely correlates to the manufacturer's stated performance with one channel driven. You can't use it to predict power output with both channels driven. However, I believe that performance specs with one channel driven may actually be closer to real-world behavior than the more conservative specification with both channels driven.

I reason that power specs with both channels driven are really only valid for a common monaural input signal to both stereo channels, resulting in maximum stress on the DC supply since both channels are making identical instantaneous current demands. But monaural use in the typical home installation is relatively rare. More likely, the inputs to the two channels will be fed with non-identical stereo signals whose peak power demands will converge only occasionally. Thus, each channel reproducing a stereo signal will likely behave as though it is the only channel being driven.

GOING FORWARD WITH SOAP

In this article I've given SOAP constants for three types of amplifiers, but I've also identified at least five other classes of amps for which no valid SOAP constants are available. Obviously there's a lot of work still to be done.

If readers would like to measure V_{DC} and V_{RMS} at 8Ω for their own amplifiers (taking care to employ the procedures described in this article), I would be pleased to collect the raw data and periodically revise the existing SOAP constants as well as add new classes of amplifiers to the list. You can e-mail data to me at: benpoe1@juno.com. If enough good data is submitted, I'll revise the SOAP constants and update the list in this publication with an occasional Letter to the Editor. The greater the statistical base for these figures, the better off we all shall be. ❖

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Borbely's RIAA Preamp With Tubes

This approach to RIAA equalization results in a high-quality phono preamp featuring the high-transconductance 5842 tubes. **By Joe Tritschler**

In 1985, the illustrious Erno Borbely published an article in *The Audio Amateur*[1] describing the design and construction of a high-quality audio preamplifier. His design included a unique approach to the RIAA-equalized phonograph disc stage.

Traditionally, the phono preamp has been implemented using an equalization network in the feedback loop of a high-gain, cascade feedback pair (CFP) amplifier. Citing poor equalization accuracy, dynamic range problems, and overall bad sound, Richard Marsh[2] and others developed passive solutions to RIAA equalization—usually a network sandwiched between two linear amplifier stages. In an attempt to utilize the best features of each approach, Borbely devised a half-passive/half-active scheme for implementing the RIAA equalization curve in a very low-noise, high-headroom phono preamplifier.

Being of the persuasion that insists on tubes for highest quality sound, I believed that a vacuum tube implementation of the Borbely RIAA architecture might make a superb-sounding phono stage. Thus, when I needed a new preamp for critical evaluation of other components in early 2001, I began work on the design in this article.

ABOUT THE AUTHOR

Joe Tritschler, 24, received his M.S.E. from Wright State University (Ohio) in June 2003 with a major in Electrical Engineering. He plans to begin work on his Ph.D. in the fall. When Joe's not busy conducting his latest hi-fi experiments, you'll find him working on his 1960 Cadillac Fleetwood, blasting his potato cannon near his parents' home in Enon, Ohio, and singing and playing guitar in his rockabilly trio, Crazy Joe and the Mad River Outlaws (www.madriveroutlaws.com).

BACKGROUND

For those unfamiliar with the playback requirements of phonograph records—specifically the long-playing microgroove record (LP)—perhaps a brief overview is in order. The process of cutting discs is not a trivial one. The modern feedback-coil cutting head operates as a "constant-velocity" device, in which the recorded signal produces a groove modulation with constant velocity with respect to frequency. This is in contrast to the "constant-amplitude" system, where the groove modulations exhibit constant amplitude with respect to frequency. In the constant-velocity system, groove amplitude is in inverse proportion to frequency when driven with a constant voltage.

The net result of this is a 6dB-per-octave rise in groove amplitude towards low frequencies. To ensure that record cutover (where groove amplitude is so large that the cutting stylus blasts through the groove wall) does not occur, the RIAA specifies that all discs cut in the US be equalized with a first-order attenuation of low frequencies below 500Hz, flattening at 50Hz. To improve signal-to-noise ratio, a high-frequency pre-emphasis is added with an initial time constant of 75µs (approximately 2.12kHz). The RIAA playback equalization curve is, therefore, the inverse of this characteristic, with a low-frequency boost starting at 500Hz and leveling off at 50Hz, and a high-frequency cut starting at 75µs.

Allen Wright has pointed out that there is obviously a limit as to how high in frequency the recording boost can prevail,[3] and this is generally accepted to be in the neighborhood of 50kHz. Therefore, in most modern phonograph preamplifiers, the high-frequency de-emphasis levels off above this point. It would seem that a first-order differentiator spanning the range of 20Hz to 20kHz during cutting would be much easier to implement and decode, and might sound better. Perhaps some more studied readers could enlighten

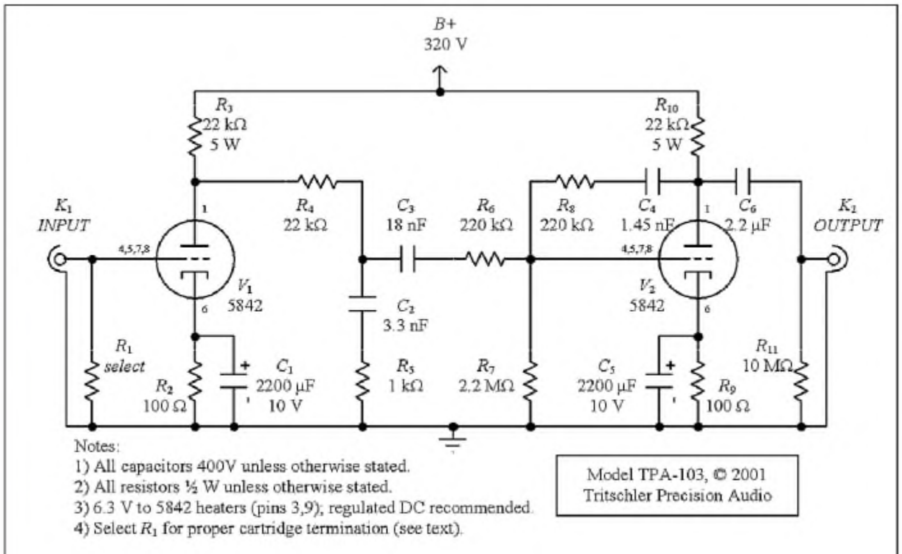

FIGURE 1: Preamp circuit.

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me about why the RIAA decided on two first-order transfer functions separated by only a couple of octaves.

Implementing the RIAA equalization characteristic in a low-noise disc playback amplifier has always been a formidable task, and the two principal approaches have been hotly debated. Borbely's solution is to use a wideband voltage amplifier direct-coupled to a passive low-pass filter for the 75µs rolloff, and then an amplifier with active shelving EQ for the 500Hz boost. Erno Borbely deserves kudos for his extremely original and effective solution. Please refer to his original article[1] for a detailed account of his design considerations.

THE DESIGN

My phono preamp (*Fig. 1*) begins with a common-cathode triode amplifier used as a linear voltage gain stage. I chose the 5842 for its outrageously high transconductance (g_m) of 25mA/V at the published nominal operating current (25mA). High transconductance is absolutely necessary for low noise, and is important when the signal source is good for only 3-4mV$_{RMS}$ at 1kHz (typical output for a moving-magnet phonograph cartridge).

Quiescent plate voltage is set to approximately 100V with a plate current of 10mA. This lower current results in slightly lower g_m (20mA/V), but it is still much better than the typical low-g_m, high-µ twin-triodes such as the (miserable) 12AX7. Amplification factor (µ) at this operating point is 42, and dynamic plate resistance (r_p) is 2kΩ. Plate load resistor (R_L) is 22kΩ, selected to be more than ten times the plate resistance for excellent linearity and high gain. B+ is 320V, regulated.

I used a 100Ω cathode resistor to bias the tube to the required grid-to-cathode potential of −1V, and inserted a 2200µF

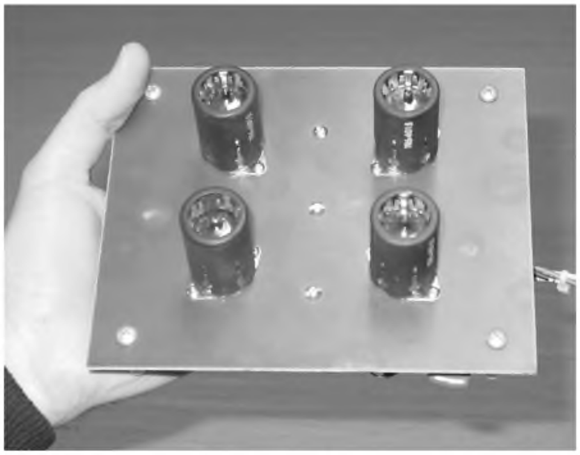

PHOTO 1: Prototype preamp.

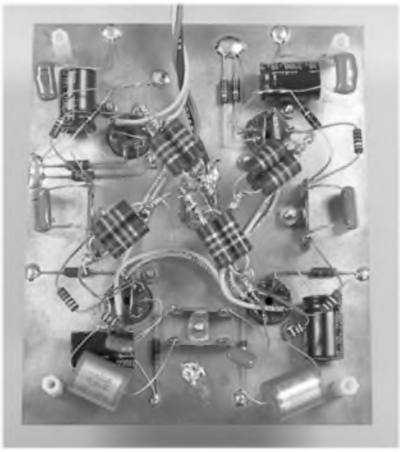

PHOTO 2: Underside of board.

bypass capacitor to maintain low effective plate resistance and encourage maximum gain. Input grid-leak resistor can be 47kΩ or whatever the cartridge manufacturer specifies for proper loading; the same applies to any capacitance added to the input. Just remember that amplifier input capacitance due to the Miller effect is already around 75pF, and be sure to include cable/tonearm capacitance, which is effectively in parallel with the input, when computing the necessary remainder. For the definitive analysis on proper cartridge termination, refer to Raymond A. Futrell's excellent article.[4]

This stage is direct-coupled to a shelved low-pass filter consisting of a series 22kΩ resistor and shunt RC network. Admittedly, a larger series resistor would have presented a lighter load for the first stage at high frequencies, but only at the expense of midband gain, and an effective high-frequency load of five times the plate resistance is still quite good. The RC network consists of a 3.3nF capacitor and 1kΩ resistor.

Assuming second-stage input loading of 220kΩ and bearing in mind that the effective series resistance includes the output resistance of the stage ($r_p \parallel R_L$), the turnover point is within 1% of 75µs with a shelf within about 3% of 50kHz. Midband gain is approximately 34.7, or 31dB, again assuming the aforementioned loading conditions, and including midband loss due to the 75µs filter.

Fans of the legendary Charles Boegli will immediately recognize the second amplifier stage, which is configured as an anode follower[5], or inverting voltage amplifier with shunt-derived/shunt-applied feedback. I chose shunt/shunt feedback for its profoundly better subjective sound quality over conventional series-applied feedback. Input series resistor is 220kΩ in accordance with previous assumptions, and this is in series with a decoupling capacitor to block the standing plate voltage from the first stage. The feedback resistor is also 220kΩ for approximately unity midband gain, and the boost below 500Hz is courtesy of a series 1.45nF capacitor, which can be selected from 1.5nF capacitors.

For the 50Hz shelving frequency, I encountered problems with insufficient open-loop gain; neither a shelving resistor across the feedback RC network nor tuning the input network to 50Hz gave adequate results because of this. After much agony and frustration (and a stern refusal to add more gain stages), I discovered empirically that a bit of "English" applied to the decoupling capacitor gives remarkably accurate conformation to the RIAA curve. The final value of 18nF gives a predicted shelf below 36Hz, but the actual turnover is very nearly 50Hz. Grid-leak resistor is 2.2MΩ, selected to be ten times the input and feedback resistances to minimize the possibility of interaction. Output capacitor is 2.2µF, and the entire phono stage will drive loads down to about 50kΩ without affecting gain or RIAA accuracy.

CONSTRUCTION

I built the prototype on a piece of copper-clad epoxy PC board material (*Photos 1* and *2*). The copper makes a very good ground plane, which is important for a low-hum preamplifier. The circuit is point-to-point wired using terminal strips, which I believe has far superior sound to PC wiring. Resistors are Allen-Bradley carbon composition with 5% tolerance, which I prefer over film types for their sound despite less-than-stellar noise performance (although if de-rated to approximately one-third their nominal power rating, noise and stability are dramatically improved).

Each plate load resistor is composed of three 68kΩ, 2W resistors in parallel to make a composite resistor within 3% of the specified value. The remaining resistors are all ½W types. Decoupling and equalization capacitors are metallized polypropylene Sprague "Orange Drops," which I use here because they are cheap, available in surplus, and pretty good in sound quality. The 2200µF cathode-bypass capacitors are Nichicon electrolytics. IERC heatsinks are used, which also serve to shield the tubes.

TESTING

I tested the preamplifier for frequency response and equalization accuracy using a Fluke 407DR laboratory power supply, Hewlett-Packard 33120A 15MHz signal generator, and Tektronix AA501 distortion analyzer used as a high-precision AC voltmeter (*Table 1*). Standard values are taken from a modified version of a computer program found in Morgan Jones' excellent volume, *Valve Amplifiers*.[6]

As you can see, the equalization is accurate within 0.4dB from 10Hz to 100kHz. Most of the variation occurs between 50 and 500Hz, probably because I made no concentrated effort to select parts. The 1.45nF capacitor previously mentioned is most likely the culprit, being directly responsible for the 500Hz turnover frequency. Gain at

TABLE 1
PREAMP RESPONSE

FREQUENCY (Hz)	STANDARD GAIN (dB) (REF. 1kHz = 0dB)	MEASURED GAIN (dB) (REF. 1kHz = 0dB)	ERROR (dB)
D.C. (0Hz)	+19.9	−∞	−∞
10	+19.7	+19.5	−0.2
20	+19.3	+19.7	+0.4
50.05	+16.9	+17.0	+0.1
70	+15.3	+15.1	−0.2
100	+13.1	+12.8	−0.3
200	+8.2	+7.9	−0.3
500.5	+2.6	+2.5	−0.1
700	+1.2	+1.2	0.0
1k	0	0	0.0
2k	−2.6	−2.6	0.0
2.122k	−2.9	−2.9	0.0
5k	−8.2	−8.3	−0.1
7k	−10.7	−10.8	−0.1
10k	−13.6	−13.7	−0.1
20k	−19.0	−18.9	−0.1
50k	−24.5	−24.4	+0.1
70k	−25.7	−25.7	0.0
100k	−26.6	−26.7	−0.1
200k	−27.3	−28.8	+1.5

1kHz is 30.8dB, resulting in only 140mV output from a 4mV cartridge, so a line stage will be required to drive most power amplifiers. This situation is similar to that encountered with tube-type FM tuners.

I then tested the preamplifier for sonic quality using a Lambda Model C-481M-505 regulated power supply, Thorens TD-125 turntable, Premier MMT tonearm, and Shure M97xE moving-magnet cartridge through home-brew tube power amps and speakers and a variety of (mostly jazz) LPs. While this is a simple setup by audiophile standards, the sonics can hardly be described as modest. The sound of this preamplifier is clean, natural, and balanced, without a trace of harshness or exaggerated "tube warmth." Detail is excellent without being obnoxiously analytical.

The Sheffield Labs direct-to-disc recording of Harry James, the King James Version, yields so much realism that the Chair of the EE department at Wright State University commented, "I've never heard recorded music sound like this before." Hum and noise are truly negligible. In fact, this preamp has been a fixture in my main hi-fi system for over two years, shattering the normal lifespan of a component by at least 18 months.

If and when time permits, I plan to construct a second stereo preamplifier using non-inductive wirewound precision resistors and oil-type Sprague Vitamin-Q capacitors. I strongly encourage any reader looking for a high-quality phonograph preamplifier to try this one.

TABLE 2
PARTS LIST

COMPONENT	DESCRIPTION
C1	CL
C2, C6	2200µF, 10V
C3	3N3
C4	18N
C5	1N45
C7	2µ2
K1, K2	RCA
R1	RL
R2, R9	82
R3, R10	22k, 5W
R4	22k
R5	1k
R6, R7	220k
R8	2M2
R11	10M
V1, V2	5842

For those not willing to use NOS 5842's, a kit version of this preamp using a paralleled 6922 in place of each is available from Tritschler Precision Audio (www.tritschlerprecisionaudio.com). If there is significant interest in the project, a PC board will be made available through Old Colony Sound Lab.

ACKNOWLEDGMENTS

Special thanks to the EE department at WSU for their encouragement and use of their test equipment. ❖

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An Unorthodox Two-Way, Part 2

The author provides cabinet design and construction, driver measurements, and crossover design and construction in the detailed second part of this two-way speaker project. **By Jon Mark Hancock**

I chose the Woodstyle WS123REV cabinet because it had sufficient volume to accommodate bracing and driver and port volume (*Photo 4*). I used the REV version instead of the standard version because it reverses the width and depth ratios, having a more narrow front (12″) and greater depth (14.5″), which has several benefits. First, it's possible to fit a 3″ ID port with the required length necessary for a moderately low cabinet tuning (range of 32Hz). Next, the narrower baffle moves the baffle step frequency up to a slightly higher range, which eases the crossover baffle step compensation.

Also, my experience is that the greater cabinet depth allows the greater absorption of the mid-woofer back wave in the midrange without overdamping the bass. Lastly, the proportions are esthetically more pleasing to me, particularly since it results in a smaller front panel—which is more apparent than the cabinet depth from most listening positions. This makes the speaker appear smaller—rarely a shortcoming, in my experience.

If you prefer to start from scratch in building this, use the overall dimensions in my detailed drawings as a guide (*Figs. 12* and *13*). I recommend using a locking dado construction for the front and rear panel. To match the precision of the Woodstyle enclosures, you'll need to do repeatable setups to 1⁄32 of an inch. For those interested in working with the Woodstyle enclosure, I'll detail step by step how to modify it for this speaker system.

FRONT SUB PANEL

Starting with the Woodstyle enclosure, my first step was to mask off the area where the supplementary front sub panel is attached. I used a scribe to mark off the area where it will be attached (*Photo 5*), then masked the outer area using Scotch safety release masking tape (the heavy flat white paper masking tape, not the beige crinkly stuff more often used for quick painting tasks).

Then, using a small hand-held orbital sander, I removed the black paint finish in this area. This is probably the single most annoying task in this project, because the clear coat applied to the black paint on the front panel tends to clog even coarse 100-grit paper. Using a wire brush on the sandpaper will help keep it clean.

I made the supplementary front sub panel from ¼″ hardboard; not the standard Masonite, but a hardboard more nearly resembling HDF, and finished smooth on both sides. I obtained mine from a local home improvement center (the one with the orange paint theme). This material is similar to ¼″ MDF in color, but seems stronger and stiffer. When gluing and clamping this panel to the speaker front, be careful to line it up per the drawing dimensions (so it will clear the grille panel), and if you don't have bar clamps, use extra wood pieces behind the clamps at the rear panel.

You can use standard wood glue such as Titebond, but for all bracing and supplementary panels I used two-part epoxy (30 minutes drying time). You'll probably find the best price on epoxy from places selling supplies for fiberglass repair or lay-up. Local hardware stores are expensive for smaller sizes, such as a dual tube plunger.

PHOTO 4: Finished speaker on stand.

Some national discount stores such as Wal-Mart are about 35% less.

Epoxy, though expensive, seems to produce some of the best-sounding joints (minimal sound)[4], and is relatively non-critical of glue joint thickness—so that's what I use. Feel free to experiment, but realize that if you do, you may wind up with a very different result than I did.

I evaluated the tweeter and mid-woofer positions using Baffle Diffrac-

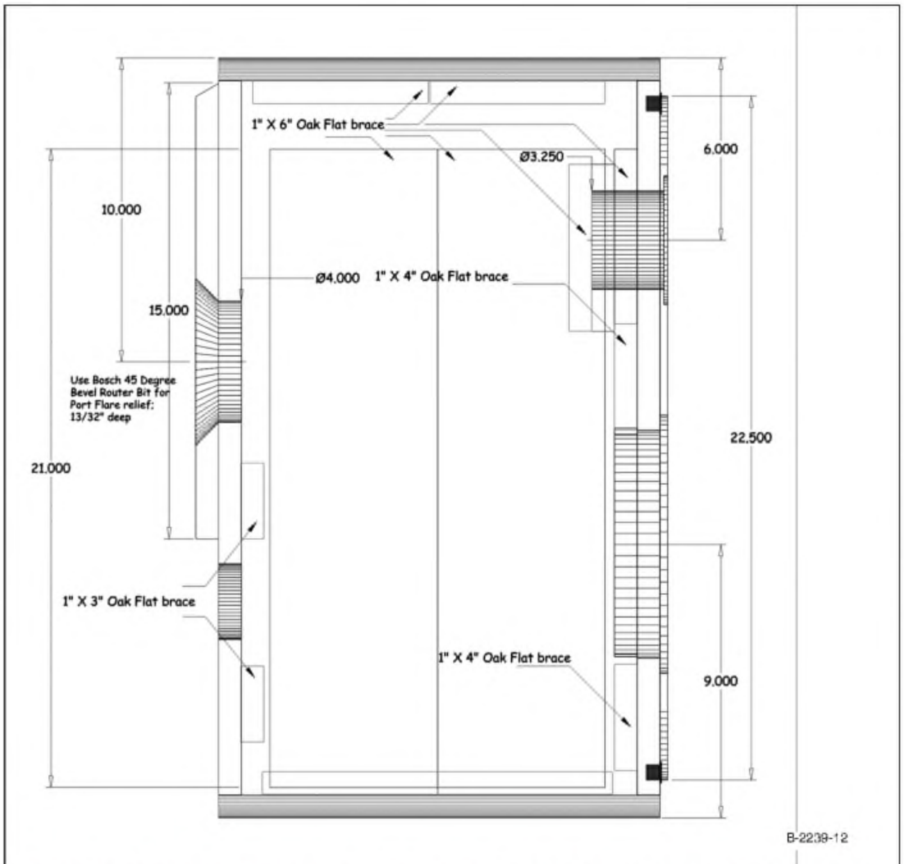


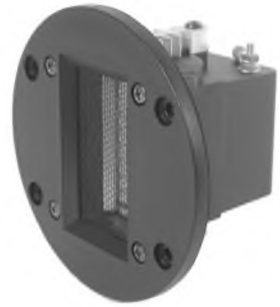
FIGURE 12: Side construction view of modified Woodstyle enclosure with dimensions.

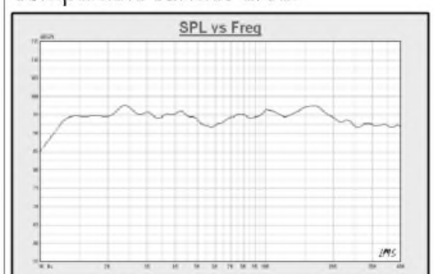
tion Simulator from Paul Verdone, available as freeware from the FRD consortium website[7]. The mirror image offset for the tweeter location distributes the edge diffraction effects over a wider frequency range and reduces the impact at any one frequency.

Next, I marked the position for the woofer and tweeter centers (see *Fig. 13* for front panel dimensioning) and, using the world-famous Jasper jig, routed out the mounting flanges, setting up for an outside diameter on the tweeter hole of 4.25″, and 8.5″ for the woofer (*Photo 6*). I used a ¾″ router bit and adjusted the location of the hole size pin to compensate for the extra bit size compared with the nominal ¼″ for the Jasper jig calibration. The mounting recesses were 1/16″ deep. After cutting out the woofer and tweeter holes, check the driver fit (*Photo 7*).

BRACE INSTALLATION

First, I installed the top to bottom side braces. Refer to *Fig. 12* for dimensions and positioning; they are mounted flush to the inside of the front panel and the bottom of the enclosure. Their





length is limited by how long a piece you can insert through the woofer hole and still clear installing it to the side.

I found this easiest to do by laying the enclosures on their side, taking care to protect the finish, and first installing both side wall braces for one side. I provided clamping force while the epoxy set using some Plitron power transformers and inductors which I had on hand, but I'm sure something more pedestrian, such as bricks, would work just as well. Install braces for the other side wall in the same manner after the first ones have set up.

PHOTO 5: Masked and sanded, ready for attaching front panel build-up.

PHOTO 6: Driver recesses routed in front panel build-up.

I prefer not to have tweeter mounting plates subject to back wave pressure from the low-frequency drivers, so I used 1 × 6 (which is actually 0.75″ × 5.5″) oak to produce a tweeter "sub enclosure." Cut identical pieces 8.5″ long, but cut one with the same 3.2″ diameter hole as for the front baffle tweeter cutout. The easy way to do this is to hold the piece in position by hand and, using a marker pen, trace the hole outline in the front panel. First, I glued and clamped the brace with the tweeter clearance hole, then—after it sets—secured a second solid brace.

After mounting the tweeter backing panels, the next step in reinforcing the enclosure is to glue the 1″ × 3″ braces above, below, and to the sides of the woofer mounting hole. Standard clamps work well for gluing these braces in place (*Photos 8* and *9*). After the front panel braces had set up overnight, I used a router with a roller bit to follow the panel cutout and cut out the braces (*Fhotos 10* and *11*).

Following this, you can glue the bottom and top wall braces in place after fabricating them from 1 × 6″ material. Note that the top wall braces can run the full width of the internal wall, so they can be cut to about 10″, while the bottom wall braces must allow for the side wall braces, and so should be the same width as the tweeter braces, about 8½″.

Last I mounted the internal braces for the rear panel (above and below the pre-cut cup hole—see *Fig. 12*) and cut an additional MDF board for an external rear panel brace. With this brace external, you can accommodate a longer port and avoid reducing the box volume. For one set of the MkIIIs, I made this panel full length of the top to bot-

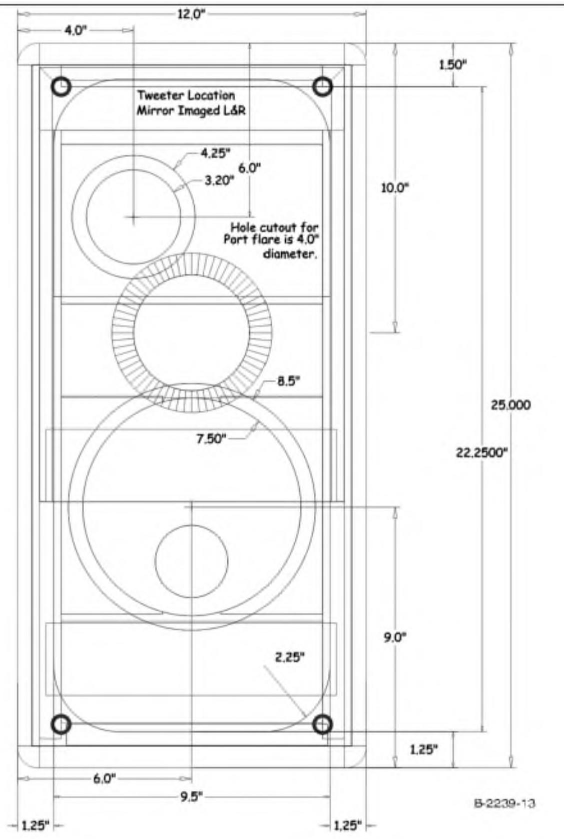


FIGURE 13: Front construction dimensions, showing tweeter and woofer locations (mirror right and left), with front panel build-up using ¼″ HDF.

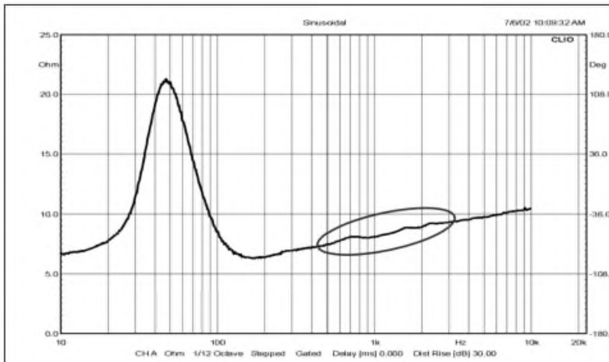


FIGURE 14: Impedance plot for 21W8554, highlighting the "wrinkles" in free air curve. B-2239-14

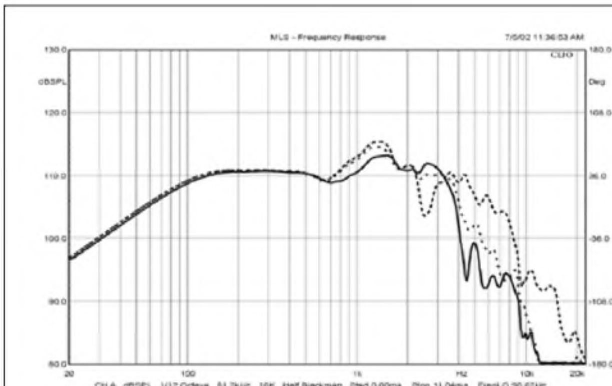


FIGURE 15: Near-field measurements of 21W8554 (½″ to cone) at center, mid-way from center to cone edge, and cone edge near surround. B-2239-15

tom, but for the other and the MkIV, it was shorter (*Photo 12*). I beveled the side and top cuts by 30° from the vertical, and, like the front baffle panel, I sanded the cabinet to remove the original paint before gluing the new panel.

PORT INSTALLATION

After gluing the internal bracing in place, the next step is to mark the position for the mounting hole for the bass reflex port, and to cut this port. I used a 3″ diameter Precision Port assembly, with flared inner and outer pieces. In principle, you can cut the baffle hole for the port to the inner radius of the mounting flange and then install the assembled port through the baffle hole. However, this results in a relatively large cutout hole, which weakens the enclosure wall. I decided to minimize the size of this hole, which requires a bit more care in the assembly and installation of the port.

The preferred method is to use a hole saw and drill press. I have an old Craftsman setup I keep around for jobs like this. Otherwise, mark the 4″ diameter circle with a compass and use a fine scroll cutting blade in a saber saw, taking care not to overheat the blade by cutting too fast. To use a mounting hole

PHOTO 7: Checking driver fit after cutting holes.

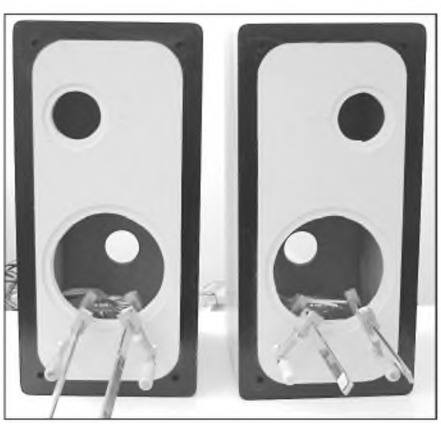


PHOTO 8: Clamping woofer opening interior braces.

PHOTO 9: Completing clamping the front panel braces in the woofer area.

PHOTO 10: Ready to mill out woofer opening in front panel braces.

this small for the port, you'll also need to provide some relief on the port hole to clear the flare. The simplest way to do this is with a 45° bevel router bit with a roller bearing follower. I've found Bosch makes a nice bit for this kind of job.

I set the bit to cut 0.4″ deep, then checked it against the flare and took a little more off until the port flare would sit flush. With the roller follower, alignment of the cut is a non-issue; it's all controlled by the depth and your original cutout.

Photo 13 shows the cabinet back panel after routing the port bevel. Note that you should cut the center piece so that the total port length, including flares, is about 1″ longer than the calculated length from your design program or manual calculations. For this design, you must cut the center pieces 3.5″ long. Using PVC welding cement, glue the rear flare and the cut tube together. I didn't install the port yet, because there's a few more steps to be completed first.

Now it's time to reach for the safe release masking tape again. As shown in *Photo 14*, mask off the painted areas of the front panel and the sides of the cabinet in preparation for painting the unfinished portion of the front panel. Use your favorite satin or gloss black enamel, with sufficient drying time between several light coats. Even though this is "safe release" masking tape, I still remove it fairly soon after the last coat is reasonably dry. Let the paint harden at least overnight, then mask and paint the rear walls in a similar fashion.

Before attaching the main port flare, I check the position and fit, mark drill holes with a center punch, and drill pilot holes for the "grabber" screws, which have extra large flat heads. I quickly apply a healthy bead of hot glue on the bevel area and a small bead on the flat area at the edge of the bevel, then install the flare and seat it firmly to spread the glue. After installing the mounting screws, I use the welding cement to attach the body of the port to the port flare.

Keep in mind the clearance between the port and the rear of the front panel. I carefully controlled the components and total height of the tweeter Zobel network board, which I mounted on the rear of the built-up front panel behind the tweeter, so it's important to have enough clearance to slide it into place. If you're not sure about the room you have, or the total board height, you may wish to leave the final port assembly until after you have mounted and wired the Zobel board.

The final step is to install a brace side to side to further reduce the side panel flexure. Quite simply, cut a 1″ × 4″ oak 9″ in length for a snug fit between the braced cabinet walls.

I mounted mine between the second set of side wall braces (toward the rear), just below the hole for the reflex port. I used a rubber mallet to tap it into place, first coating the ends with some epoxy and locating one end firmly in the position I wanted it in the cabinet, then tapping the other end with the mallet to scoot it into position. *Photo 15* shows how it looks through the woofer hole after installation. This brace isn't absolutely necessary, but I found it does

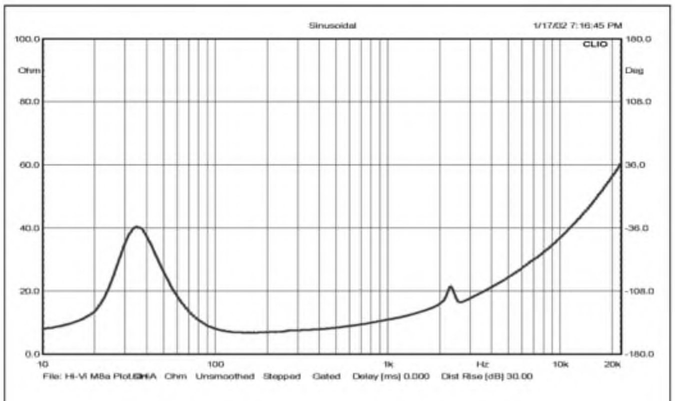


FIGURE 16: Free air impedance plot of HiVi Research M8a.

B-2239-16

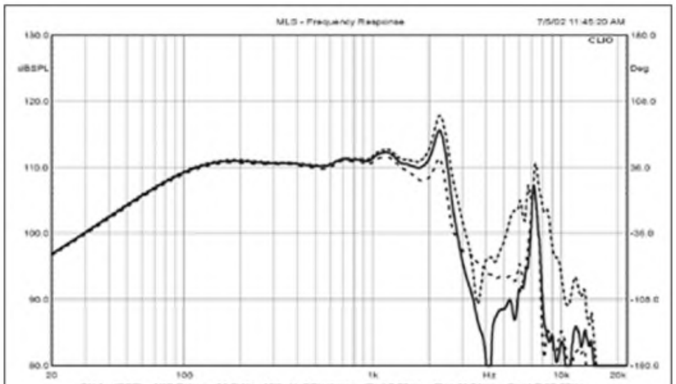


FIGURE 17: Near-field measurements of M8a at center, mid to edge, and edge near surround.

B-2239-17

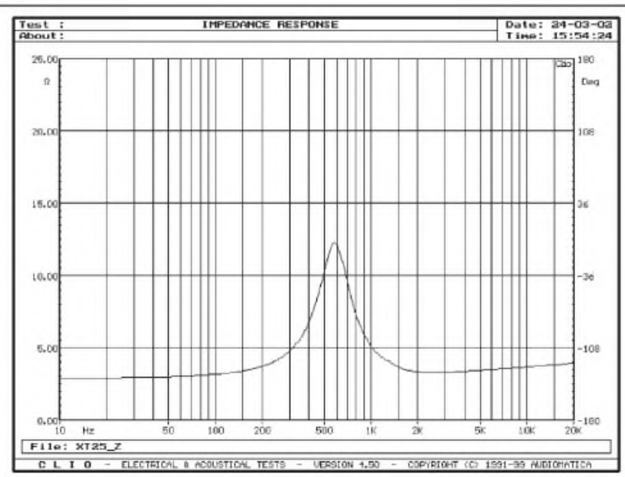


FIGURE 18: Impedance plot for Vifa XT-25.

B-2239-18

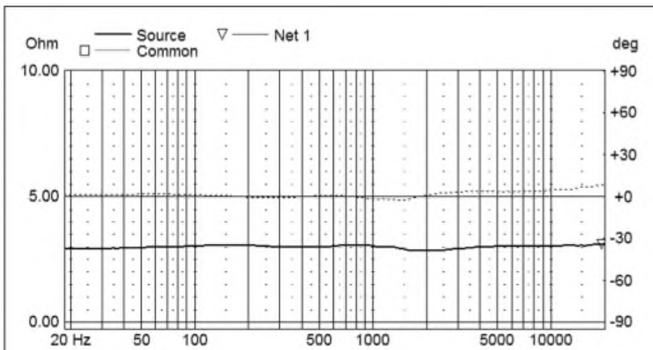


FIGURE 19: Predicted impedance of Vifa XT-25 with calculated Zobel network.

B-2239-19

further stiffen the sides and reduce the panel resonance.

LF DRIVER MEASUREMENTS

I evaluated drivers and worked with the first crossover concepts using a simple particleboard sealed box, as well as an older Woodstyle enclosure. First, consider an 8″ driver I used previously for a three-way system, the Scan-Speak W21/8554 (*Photo 16*). *Figure 14* shows a measured impedance curve, while *Fig. 15* is a graph of the near-field measured frequency response at three positions: the center, midway towards the surround, and at the edge near the surround.

The first impedance glitch at 700–800Hz seems to correspond to a slight dip in the response, followed by the "classic" Scan-Speak rising amplitude characteristic. The near-field response tracks quite closely across the cone until this mode is encountered, then diverges somewhat up until about 1600Hz. Then, above 2kHz there are large perturbations in the response, as additional modes kick in.

Based on my experiences testing drivers and commercial speakers, I prefer operating drivers within their pistonic mode and seeing cone modes attenuated by at least 20dB or more—preferably 40dB. For this reason, the crossover frequency should stay below the first cone breakup in the mid-woofer passband. This is often not the first obvious peak in the response curve, which is why a careful examination of the impedance curve and the near-field response can be very helpful.

I eventually settled on the Hi-Vi Research M8a (*Photo 17*), an 8.5″ frame mid-woofer with a concave aluminum magnesium cone. The impedance curve looks clean until the first major mode at about 2.3kHz (*Fig. 16*), and the near-field response tracks well (*Fig. 17*). The response rise at the first breakup mode at 2.5kHz is relatively low in Q compared to some other 8″ metal cone drivers I tested, and I hoped it would require less extreme crossover measures to tame.

The response curves measured at

PHOTO 11: Front panel braces routed and tweeter sub-baffle installed.

PHOTO 12: Back.

three near-field positions track well, showing close tracking up to about 3kHz. With this type of driver, and the challenge of mating it as a two-way, the elliptic crossover design proved to be a very important factor in the success of the project, and, as we'll see, enabled the response to be tailored closely to the overall requirements.

CROSSOVER DESIGN

The basic process for measuring driver response and importing data into PC-based software for crossover design has been described at some length by many others, so I won't recapitulate very much of that here. There are a few points I've found which make the process go more easily and consistently, so perhaps a brief discussion of those issues will be helpful for others.

For ease of data import to LspCAD, I use the DOS version of the Audiomatica CLIO software when making measurements for import, as its output data format is supported directly by LspCAD as well as other crossover design software. I also use CLIOWIN, whose flexibility I find very useful for driver investigations—but its data format for MLS output isn't compatible with many programs expecting the output from the DOS version.

If I had measured the drivers in the final enclosure design at "far field" (2m or more), then the required compensation for baffle step would be straightforward, as it's a function of the measured response versus the desired acoustical response. But since I was measuring somewhat closer to reduce room reflections for MLS measurements with 200Hz bandwidth, this required that I estimate the compensation for baffle step using a combination of the BDS simulator by Paul Verdone, and comparing against my near-field driver results.

The next step is to produce a file for the tweeter that is just for impedance compensation—a single driver project. For tweeters such as the Vifa XT25 which do not use any LF damping with ferrofluid, the impedance rise at resonance is substantial (*Fig. 18*) and will interact with the crossover filter unless its corner frequency is several times higher.

I prefer to start the tweeter network design with a full impedance compensation Zobel, optimized for as close to a ruler flat impedance curve for the tweeter as I can get. This also has the benefit of assuring very consistent frequency response when an L-pad is used with a wide range of attenuation

PHOTO 13: Back panel after routing for port mounting flare.

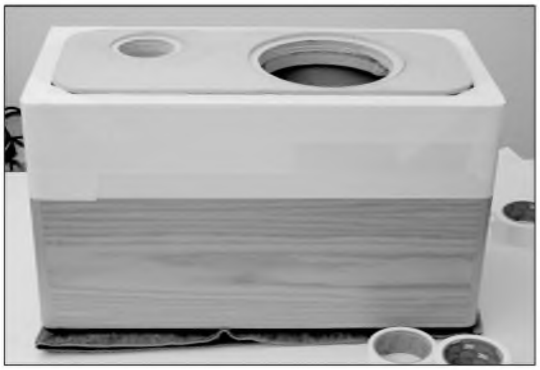


PHOTO 14: Front masked and ready to paint.

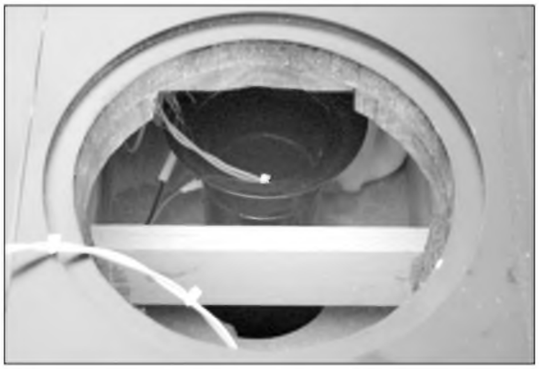


PHOTO 15: View showing side to side brace and interior view of port.

variation. While this is unlikely to happen in a single design, if you like to "recycle" your network designs, it's a must.

Since from the outset I knew there was interest in a boundary loaded smaller version of this speaker, as well as an MTM, this flexibility was mandatory. While it's not uncommon for "production" speakers to use a single resistor for tweeter attenuation and optimize all the component values so that the de-

PHOTO 16: Scan-Speak 21W8554 sample.

PHOTO 17: HiVi Research M8a mid woofer.

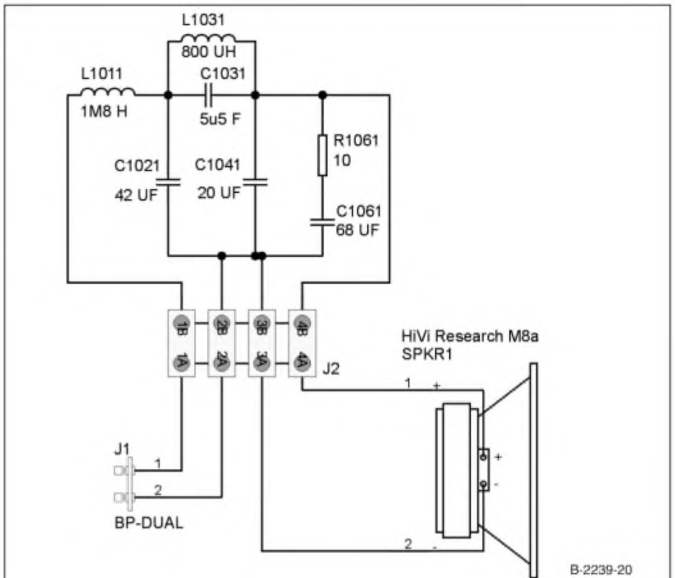


FIGURE 20: Low-pass crossover, final design for LP board.

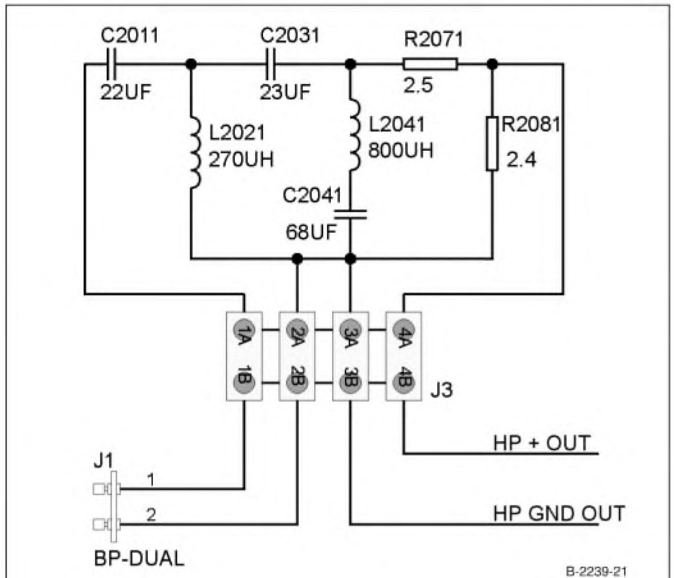


FIGURE 21: High-pass board crossover schematic.

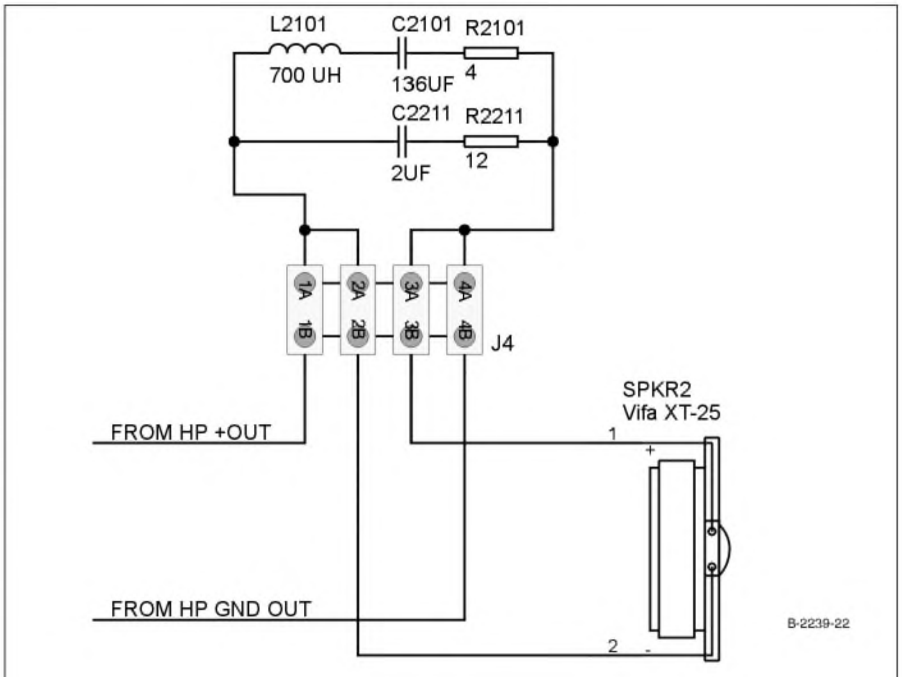


FIGURE 22: Tweeter Zobel board network, connected between HP network and tweeter (note tweeter polarity).

sired acoustic transfer function is achieved with a single resistor, any need to adjust the tweeter level up or down may result in the shape of the filter transfer function being altered by the change in impedance seen by the network. My experience has been that minimizing these kinds of interactions in the early design phase pays benefits in finishing quickly a consistently performing design—you can optimize for cost reduction later, and for most DIY constructors, the price of a few resistors is a non-issue anyway.

Figure 19 shows the predicted impedance curve after selecting the Zobel values. No, this isn't a typo or mistake—the resistance and impedance of the XT25 are rather low, and I would probably even call it a 3Ω tweeter myself, not a 4Ω as it is specified. It is probably not the tweeter to use for a SET friendly amp design!

Even when I know well the approximate level I'll be using and required padding, I start with no pad and set up the network topology and optimize for the acoustic transfer function. Then, I freeze component values, add the L-pad, reset the level, and let the optimizer dial in the attenuation. This results in L-pad values which are "classic," and avoids impedance interaction in the network.

I've found that if I optimize the circuit at once, it may use the impedance interaction of a specific set of attenuation resistors to control part of the slope, and result in a network that works properly at only one attenuation level. Since the optimizer is "dumb" about these issues, you must be the smart one.

Other considerations for the woofer network and optimization, beyond the nominal acoustic transfer function target for LP network, include baffle step compensation and compensation for driver acoustical distance behind the tweeter, plus attenuation of the cone resonant peak. In general, the network topology is that of a conventional fourth-order LP network, with the addition of a zero in the transfer function by means of C1031. L1031 and C1031 are selected to optimize the slope of the LP cutoff to the acoustic target (eighth-order L-R), and also to tune the notch in the response to the driver peak at roughly 2.5–2.8kHz.

You achieve baffle step compensation by the interaction between the primary inductor, L1011, and the RC network formed by C1061 and R1061 in parallel with the driver impedance. Within a limited range, you can adjust the baffle step compensation by changing the values of this RC network. The

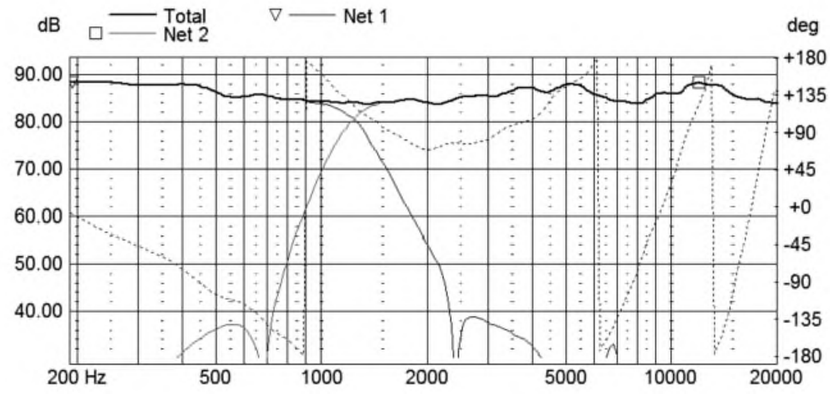


FIGURE 23: Predicted 1m response using LspCAD.

change in the effective load impedance presented to the filter network as a result of the use of the baffle step compensation requires adjustment of other component values, which is how the optimizer "pays its keep."

As a final step, I try adjusting inductor values to the nearest conventional available value, and re-run the optimizer once again to finalize the capacitance values. Last, I double-check the optimizer to make sure I haven't produced a monster network with unusual impedance dips that only an Aragon, Krell, or Bryston amplifier will drive comfortably (*Figs. 20–22*).

RESPONSES

Figure 23 shows the predicted composite response; note that the tweeter measurements were made in a test box which didn't use any diffraction suppression techniques, and I believe the roughness in the 6kHz and 12kHz areas is due to this. I don't think this is something to address in the crossover design, but rather by the baffle layout and diffraction suppression techniques, which I'll describe next month in the final assembly. Note the expanded vertical scale compared with typical crossover plots, covering a range of 60dB. A more conventional plot range of 30dB would not show the notch and "bounce-back" of the filter network.

This predicted response includes the effect of adding a 60mm acoustic center offset in LspCAD for the mid-woofer, to model explicitly the difference in acoustic origin relative to the front panel, and modifying the tweeter slope and woofer slope slightly to achieve optimum summing on the design axis, which is midway between the woofer and tweeter. ❖

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The ABCs of Filters, Part 2

In Part 1 we became acquainted with filter terms, basic equations, and types. Now let's put this theory to work in part 2.

By Richard Honeycutt

Figure 9 shows the important elements of a common speaker driver. Basically, it is a complicated damped mass-spring system. The obvious mass is the combined mass of the cone, dome, and voice coil, plus about 1/3 of the mass of the surround, spider, and lead wires. The spider, with some help from the surround, provides the stiffness (newtons of force per meter of resulting motion).

Damping results from internal friction in the surround and spider plus electromagnetic damping provided by the interaction between the magnetic system and the voice coil. (These constitute a generator, with the load being the amplifier's output resistance, the voice coil resistance, and the resistance of the wires from amplifier to speaker. Thus any motion of the cone results in countervoltage being generated in the voice coil, producing current in the coil/wire/amplifier circuit, and resulting in heat being dissipated in the wires and in the amplifier's output devices.)

Less obvious mechanical elements that affect the speaker are (1) the mass of the air close to the cone, whose motion follows that of the cone, and (2) the equivalent resistance (called radiation resistance), represented by the acoustical energy radiated from the cone. Numerous authors, notably Harry Olson[1] and Leo L. Beranek[2], have presented equivalent circuits for loudspeaker drivers. Beranek's is shown in *Fig. 10*. These equivalent circuits are valid only for low frequencies, because at high frequencies, the cone does not move as a single mass, so you must use a much more complex model. Beranek defines low frequencies as those for which the wavelength is greater than three times the cone diameter.

The values shown for the circuit elements are as follows, from left to right:

- Voltage source: the force resulting from the input voltage (e_g), and magnetic field strength (B), interacting with the length (l) of wire in the voice coil, in the presence of the amplifier and voice-coil resistances, R_g and R_E, respectively.
- The damping provided by the same electromagnetic system.
- The mass of the coil, cone, and so on (as described previously).
- The mechanical resistance of the spider and surround.
- The compliance of $\left(\frac{1}{\text{stiffness}}\right)$ the spider and surround.
- Twice the mass of air that moves with each side of the cone mounted in a baffle.

This circuit ignores the inductance of the voice coil, which is important only at high frequencies, and the radiation resistance, which is usually much smaller than M_{M1}.

Notice the force generated by the speaker's motor system. It is just the product of the electrical current

$$\frac{e_g}{R_g + R_E},$$

the magnetic flux density (B), and the length of wire (l) in the voice coil. Thus

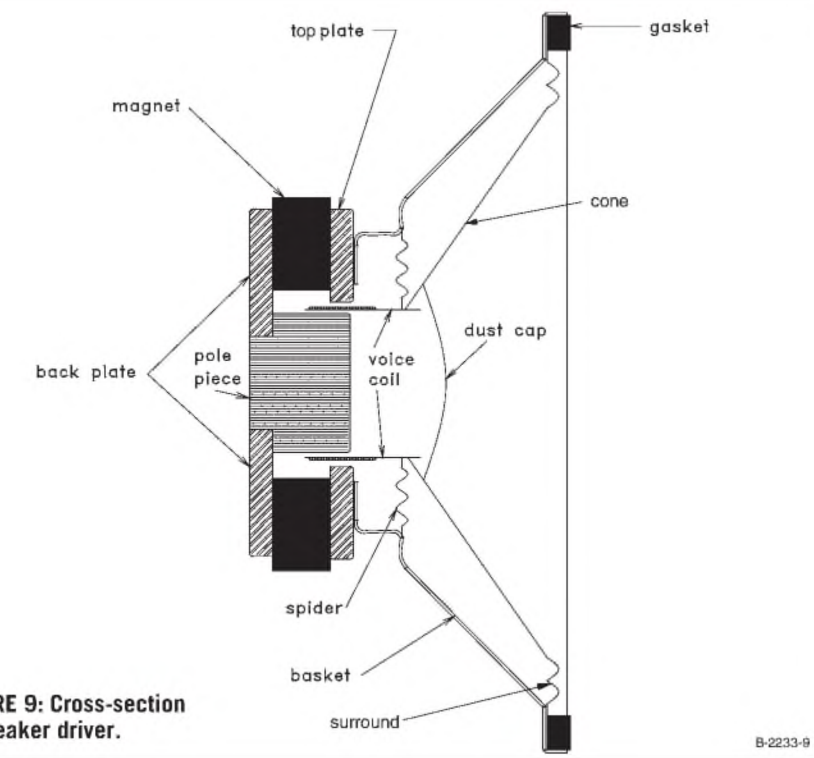

FIGURE 9: Cross-section of speaker driver.

the term Bl acts analogously to the turns ratio of a transformer from the electrical to the mechanical domain. A resistance that appears on one side of a transformer is reflected on the other side by the product of that resistance and the turns ratio squared; thus, the B^2l^2 multiplier. The electrical resistances appear in the denominator because, from the mechanical side, they act as though they were in parallel with the force source.

Understanding these analogous circuits is easier if you know the meaning of the subscripts. They are: generator (g), electrical (E), mechanical (M), diaphragm (D), suspension (S), and the "radiating side of the cone" (l).

Notice that if you combine similar elements in this circuit, it is a series combination of a voltage source, inductance, capacitance, and resistance. You can predict the low-frequency behavior of the driver by plugging this circuit into a circuit simulator such as Pspice®, provided that you use the correct values for each circuit element. (Showing how to calculate these values is beyond the scope of this article.)

Did you notice that the circuit of *Fig. 10* is just a second-order LRC filter? Since it is a series circuit, the Q is defined as:

$$Q = \frac{X_L}{R} = \frac{2\pi fL}{R} = \frac{2\pi fM}{R} \qquad (9)$$

where M and R are the total mass and total resistance, respectively.

SPEAKER IN A SEALED CABINET

When you enclose the speaker in a sealed cabinet, three things happen. First, the sound radiated from the rear of the cone is contained so that it cannot interfere constructively and destructively (adding to and subtracting from) the sound radiated from the front of the cone. Interference from the back wave is not included in the equivalent circuit model for a speaker driver.

The second effect of the sealed cabinet is that the trapped air acts as a spring, which adds to the stiffness of the speaker. The third effect is that any acoustically absorbent material in the cabinet, plus any small air leaks, add resistance to the circuit. *Figure 11* shows the equivalent circuit of a speaker driver in a sealed box.

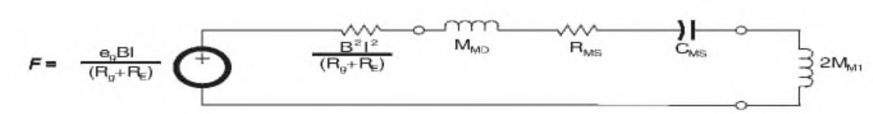

FIGURE 10: Simplified electromechanical analogous circuit for loudspeaker driver[3] in a baffle. (Radiation elements have been omitted.) B-2233-10

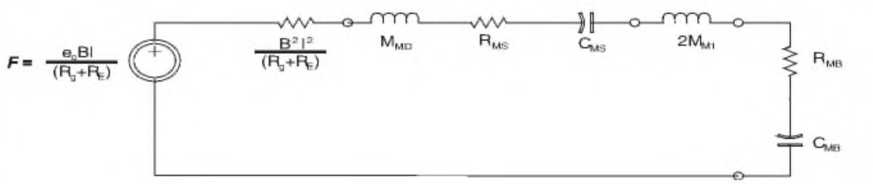

FIGURE 11: Simplified electromechanical analogous circuit for loudspeaker driver in sealed cabinet[4]. (Radiation elements have been omitted.) B-2233-11

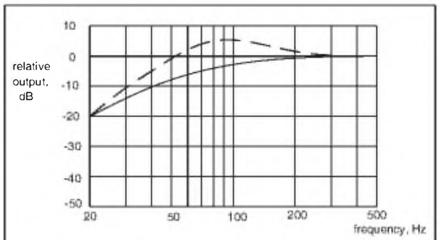

FIGURE 12: Sample frequency responses for driver in sealed cabinet (solid line: Q = 0.5, dashed line: Q = 1.2). B-2233-12

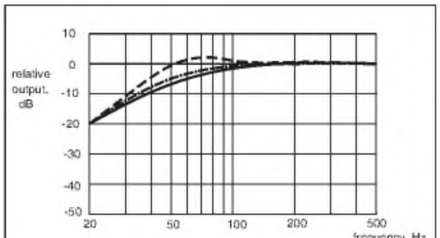

FIGURE 13: Frequency responses for sealed cabinets of different sizes. B-2233-13

Notice that the only change between *Fig. 10* and *Fig. 11* is the addition of the mechanical resistance and compliance of the cabinet, R_{MB} and C_{MB}. The circuit is still a second-order filter fed by a voltage source. The only difference is that the combined resistance is slightly larger than for the driver alone, and the combined compliance is smaller (since compliance is modeled as capacitance, and

$$C_T = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \ldots + \frac{1}{C_n}} \text{).}$$

Since the speaker driver in a sealed cabinet is essentially a filter, you can control the filter damping (mainly through choice of a driver with the appropriate Q_{TS}) so as to obtain a Bessel, or Butterworth, or Chebychev, response. Or, for that matter, if your name is Ogwally, and you wish to document a particular response that has not previously been studied, you could build the first speaker system with an Ogwally alignment!

Notice that I sneaked in the term "alignment" here. Previously, I discussed choosing certain filter families. However, Thiele and Small refer to this process as choosing an alignment. While this terminology was enlightening for electrical engineers who were familiar with filter design, for others, it seemed mainly to add an element of mystery to speaker design.

Prior to the publication of Thiele's and Small's work, we just built a box big enough to obtain the low-frequency cutoff we wanted, lined it with fuzz, and listened and bragged. After the new terminology was introduced, we had to worry about choosing an alignment, for goodness sake! Well, now I hope the mystery has been cleared up.

Now that I have an equivalent circuit for a sealed-box speaker, how do I predict the frequency response? I could go through the math for the equivalent circuit, then convert the resulting volume velocity (analogous to current, remember?) to SPL; but once again, Beranek has rescued us.

The rather complex equation for the SPL output from a speaker splits into two parts: the "source" part and the "filter" part. The "source" part includes such factors as the electromagnetic motor strength, but is independent of frequency in the range in which my analogous model is valid. So all I need to deal with is the "filter" part of the equation. While this part is still algebraically messy, Beranek simplified it by normalizing the various $2\pi f$ terms into ω/ω_0 terms, where $\omega = 2\pi f$ is the so-called angular frequency, and ω_0 is the angular frequency at resonance (2π times the critical frequency). Since you can completely specify the response of a second-order filter if you know the critical frequency and the Q_{TS}, Beranek reduced the equation for frequency re-

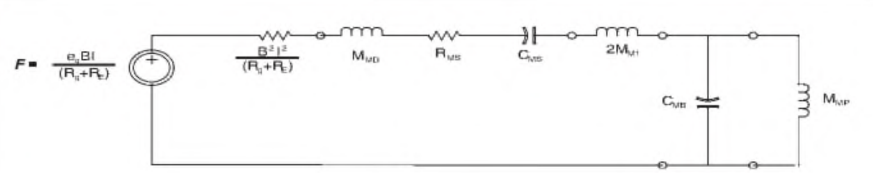

FIGURE 14: Simplified electromechanical analogous circuit for loudspeaker driver in vented cabinet. (Radiation elements have been omitted.)

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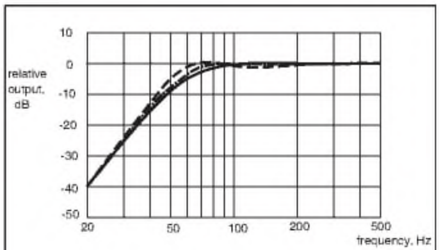

FIGURE 15: The effects of varying α. (From top to bottom, the curves represent $V_B = 1.4V_{AS}$, $V_B = V_{AS}$, $V_B = 0.7V_{AS}$; for all curves, h = 1, $Q_{TS} = 0.4$, and $Q_L = 7$)

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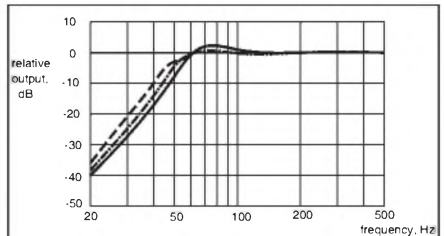

FIGURE 16: Effects of vent tuning. (From top to bottom, the curves represent $f_{BOX} = 1.1f_{SPKR}$, $f_{BOX} = f_{SPKR}$, and $f_{BOX} = 0.9f_{SPKR}$; for all curves, $\alpha = 0.7$, $Q_{TS} = 0.4$, and $Q_L = 7$.)

B-2233-16

sponse shape to:

$$SPL = 20\log\left\{\frac{\omega/\omega_0}{\sqrt{\frac{1}{Q_T^2}+\left(\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega}\right)^2}}\right\} \quad (10)$$

For example, if you design a speaker system that resonates at 60Hz, and you try one design with a system Q_T of 0.5 and another with Q_T of 1.2, you can plug equation (10) into a graphing program such as MathCAD® or a spreadsheet or a graphing calculator, and come up with the responses shown in *Fig. 12*.

Two items are left to discuss. First, you probably noticed that I plotted the curves based upon the system Q_T rather than the Q_{TS} of the driver. Since the equivalent circuit for the driver in the box is different from that for the driver alone, the two Qs will not be the same. The system Q_T is related to the driver Q_{TS} by the equation:

$$Q_T = \sqrt{1+\alpha}\,Q_{TS} \quad (11)$$

where α is the compliance ratio C_{MS}/C_{MB} and Q_{TS} is the driver Q. (While speaker compliance is not usually published, the related value V_{AS} usually is, and $C_{MS}/C_{MB} = V_{AS}/V_{AB}$, where V_{AB} is the volume of the cabinet.)

Second, ω_0 is the system resonance frequency, not the speaker resonance frequency. You can find it using

$$\omega_0 \cong \sqrt{1+\alpha}\,(2\pi f_S).$$

Figure 13 shows the effect of box size, with sizes chosen to represent Bessel, Butterworth, and 2dB Chebychev responses. In each case, the driver Q_{TS} is 0.5. (However, the three speakers do not have the same value of f_0.)

Conceptually, the reason that a higher compliance ratio increases the Q_T is that a high α implies a small box, which, in turn, means that more of the system stiffness is provided by the very-high-Q air spring rather than the lower-Q spider and surround.

VENTED CABINETS

Adding a hole or open tube to the cabinet opens additional options. With a sealed cabinet, you can change the cabinet volume or internal damping (which mainly acts as a change in cabinet volume). But with a vented cabinet, you can also change the vent tuning and the vent Q, although lowering the Q of the vent basically undoes the purpose of the vent and changes the system into a leaky sealed system. The greater system complexity shows up in the simplified equivalent circuit for the vented system.

Beranek advises calculation of response at three critical frequencies, and then assembles the response from curves on graphs corresponding to the behavior at those frequencies. Novak[5] simplified Beranek's analysis by ignoring the resistance of the vent (which should be very small) and also ignoring the effects of mutual coupling between the cone and the vent. This latter simplification is valid if the cone and vent are of significantly different size and/or are not closely spaced. I will use Novak's approach.

Figure 14 shows the equivalent circuit for a vented system. In addition to vent resistance, cabinet resistance is also ignored; it, too, is usually small.

I now have a fourth-order system, because I cannot add M_{MP} or C_{MS} to any of the other elements, since they are not in series with any of them. Besides the additional method of affecting the

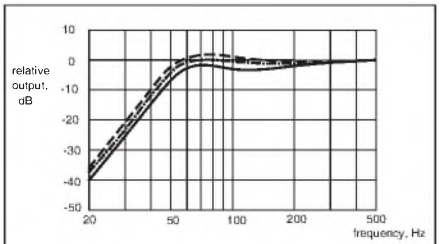

FIGURE 17: Effects of loudspeaker Q. (From top to bottom, the curves represent $Q_{TS} = 0.52$, $Q_{TS} = 0.4$, $Q_{TS} = 0.28$; for all curves, $\alpha = 0.7$, h = 1, and $Q_L = 7$.)

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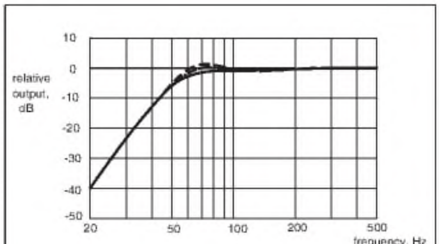

FIGURE 18: Effects of box Q. (From top to bottom, the curves represent $Q_B = 9.8$, $Q_B = 7$, $Q_B = 4.9$; for all curves, $\alpha = 0.7$, h = 1, and $Q_{TS} = 0.4$.)

B-2233-18

frequency response mentioned earlier, having a fourth-order system also means that the ultimate stopband slope is 24dB/octave.

Until the publication of Thiele's paper[6], the design flexibility vent tuning provided generally went unappreciated. Even excellent treatments such as those of Beranek and Novak dealt with vent tuning using a statement such as "The vent is usually tuned at the resonance frequency of the speaker, although it need not be." Novak developed a simplified equation to predict the frequency response of a vented system, but it assumes that the vent is tuned to the speaker's resonance.

Now let's get our progress into perspective. Olson wrote down much of the theory of speaker enclosures in the 1930s. Beranek expanded on Olson in the mid-'50s. Novak expanded on Beranek in the late '50s. Then in 1961, Thiele presented two landmark papers to the I. R. E. Radio and Engineering Convention in Sydney, Australia. Thiele recast Novak's equation into a form involving time constants and coefficients, so that it looks very different, but the physics are the same as in Novak's equation. (My equation (12) is still another way to state the same physics.)

At this point, Thiele hadn't done anything very new, but then he pointed out that any fourth-order filter—be it a speaker or a simple LCR filter—could be described in terms of its alignment or filter family. Then anyone who understood filter theory could design vented cabinets to perform as desired. In fact, you could add options by adding external filtering components, thus producing fifth- and sixth-order speaker systems. His Table 1 summarizes this revolutionary approach. In order to provide an idea of what was contained in Thiele's Table 1, I reproduced a small section of it as *Table 1*.

Here's an explanation of the information in the table. The first column is simply reference numbers to make discussion of the various alignments easier. The second column identifies the alignment as—in this case—Butterworth fourth-order or Chebychev fourth-order. The third column is the ratio of the system cutoff frequency to the speaker resonance frequency.

Column 4 is the ratio of the system cutoff frequency to the frequency to which the vent/box are tuned. The column 3 value divided by the column 4 value provides the tuning ratio *h*. Column 5, C_{AS}/C_{AB}, is also the ratio of V_{AS} to box volume. Column 6 is self-explanatory.

By looking at Thiele's Table 1, you get a quick idea of what useful enclosures are possible for a given driver, and what sort of driver you would need to choose to obtain certain performance. For example, if you want a fourth-order response with a driver whose Q_{TS} is 0.4, the box size should be somewhere between 1.055 and 1.414 times the speaker's V_{AS}, and the tuning ratio will be between 1.000 and 0.867/0.935.

In the days before personal computers, not to mention speaker-design software, this kind of knowledge was an incredible time-saver! Even today, charts that are more user-friendly versions of Thiele's Table 1 are extremely helpful; such charts are included in Dickason's *The Loudspeaker Design Cookbook*, for example.

In Marshall Leach's very useful book *Introduction to Electroacoustics and Audio Amplifier Design*, there is an equation for the SPL produced by a vented speaker system[7]. By taking the magnitude of this complex transfer function, you derive the following equation:

$$SPL = 20\log\left[\frac{(\omega/\omega_0)^4}{\sqrt{\left[(\omega/\omega_0)^4 - a_2(\omega/\omega_0)^2 + 1^2\right] + \left[\left(-a_3(\omega/\omega_0)^3 + a_1(\omega/\omega_0)\right)^2\right]}}\right] \quad (12)$$

where

$$a_1 = \frac{1}{Q_L\sqrt{h}} + \frac{\sqrt{h}}{Q_{TS}},$$

$$a_2 = \frac{\alpha+1}{h} + h + \frac{1}{Q_L Q_{TS}},$$

$$a_3 = \frac{1}{Q_{TS}\sqrt{h}} + \frac{\sqrt{h}}{Q_L} \quad (13)$$

and α is the compliance ratio, h is the vent tuning f_{BOX}/f_S, Q_{TS} is the total Q of the speaker, and Q_L is the enclosure Q. The angular frequency ω_0 is, as with the closed box, the system resonance frequency. For a vented box,

$$\omega_0 = \sqrt{(2\pi f_B)(2\pi f_S)} = 2\pi f_S\sqrt{h}\ .$$

Using equation (12) allows you to examine the effects of varying each of the four parameters affecting frequency response in vented-system design. *Figures 15–18* show these effects.

Here you see that α affects both the cutoff frequency and the alignment or filter family. Varying box volume by about 40% changes the low-frequency cutoff by about 15%, and changes the alignment from something similar to a Bessel to something similar to a 0.5dB Chebychev. In this figure, as in the following ones, fairly common values are used for the other parameters.

You can see that changing the vent tuning by about 10% primarily affects the alignment; the low-frequency cutoff is almost unchanged. In the common situation in which you design a cabinet for a particular speaker, *Figs. 15* and *16* show that α will determine the cutoff frequency, and you can then play with the

TABLE 1

FOURTH-ORDER SPEAKER ALIGNMENTS FROM THIELE'S TABLE 1

NO.	TYPE	RIPPLE (DB)	F_3/F_S	F_3/F_B	C_{AS}/C_{AB}	Q_{TS}
5	B_4	—	1.000	1.000	1.414	0.383
6	C_4	—	0.867	0.935	1.055	0.415
7	C_4	0.2	0.729	0.879	0.729	0.466
8	C_4	0.9	0.641	0.847	0.559	0.518
9	C_4	1.8	0.600	0.838	0.485	0.557

vent tuning to adjust the response shape near cutoff.

Here you see that speaker Q, like vent tuning, affects filter family but not cutoff frequency. Of course, two other parameters of the speaker affect cutoff frequency as well: compliance (V_{AS}) and resonance frequency (f_S). Compliance shows up as the denominator of α, and f_S helps to determine ω_0, as shown in equation (13).

Varying the box Q by 40% makes little difference. Most designers use a figure of seven as a typical box Q. Only if the box is very well-braced and caulked, with minimal stuffing, or if it is leaky and overstuffed, is the Q_B likely to make much difference.

Figure 19 shows the effects of f_S upon response, with α, h, Q_{TS}, and Q_B held constant. The effect of V_{AS} is already revealed in *Fig. 15*, since it affects the response equation by changing α. The figure assumes that box size is constant, and α is constant at 0.7, although choosing real speakers with different f_S and the same Q_{TS} may well result in different values of V_{AS}, which would change either α or box size.

As you would expect, changing f_S changes the cutoff frequency of the system, but only by a factor of approximately

$$\sqrt{f_{S_{NEW}} / f_{S_0}}\,.$$

SUMMING UP

Having examined all the possible ways of controlling the response of a vented speaker system, you can appreciate Thiele's codification of all those options, which makes possible the compilation of charts such as those in Dickason's *The Loudspeaker Design Cookbook*. Now that you are familiar with the origins and meaning of the filter terminology that Thiele and his successors use, you can better understand how to use the design charts.

But having studied the effects of various design parameters upon the frequency response of a speaker system, you will be able to make judgments of what parameter to change, and in what direction, to achieve a desired goal without resorting to charts. By using the equations I've provided with a graph-maker of some sort, or by using any of the various speaker design programs, you will be able to evaluate any design changes you make as you go. The end result will be better speakers with less wasted plywood! ❖

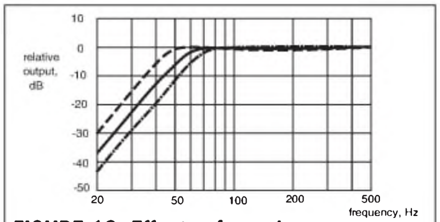

FIGURE 19: Effects of speaker resonance frequency. (From left to right, the curves represent f_S = 40Hz, f_S = 60Hz, f_S = 80Hz; for all curves, α = 0.7, h = 1, Q_{TS} = 0.4, and Q_B = 7.) B-2233-19

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Product Review

DacT CT102 Power Supply

Reviewed by Charles Hansen

Danish Audio ConnecT A/S, Skannerupvej 14, DK-6980 Tim, Denmark, US eFax (+1) 248 282 0645, www.dact.com. Price: EUR 282.00, Dimensions: 10cm W × 9cm D × 4cm H.

The DacT CT102 is an audio power supply specifically designed to be used with DacT's CT100 phono stage module and CT101 line stage module. It can also be used with other high-end audio electronics circuits. The features described in the user guide are low dynamic output impedance, extremely wide bandwidth, and fast regulation. The regulated output voltage can be jumper selected for either ±15V DC or ±20V DC. Maximum output current is 200mA per rail, with output short-circuit protection on each rail.

INSIDE THE POWER SUPPLY

Photo 1 shows the PC board top view with the heatsinks for four bipolar power transistors. There are six 8-pin DIP linear ICs on the PC board whose identities have been masked by red, blue, or green paint. In the event of problems the CT102 would need to go back to DacT for service.

The PC board is a high-quality single-sided design with gold-plated PC tracks. You can see the attention to the power and grounding track layout in *Photo 2*. The many parallel traces produce a single-point connection for all the filter capacitors on the PC board. The board has four isolated mounting holes and gold-plated input/output pins.

PHOTO 1: Top view of CT102 board.

TOPOLOGY

A schematic was not furnished with the power supply, but the block diagram in *Fig. 1* will convey the principle of operation. A 100–250V AC mains adapter with 48V DC output—supplied with the CT102 module (*Photo 3*)—is a switching power supply that operates at about 28kHz. The 48V DC enters the CT102 card through low-pass LC EMI filters and is impressed across Q1 and Q2, a pair of common-emitter bipolar power transistors, which are biased such that they produce a low-impedance output ground reference point at their collectors, forming ±24V DC supply rails for the output regulator transistors Q3 and Q4.

The output power transistors each have a control loop amplifier that monitors the output voltage and compares it to a 2.5V precision reference. The control loop also monitors the output current and provides for fold-back current limit in case of an overload or short circuit. A 4700µF filter capacitor on each regulated rail provides energy storage for transient current needs. Diodes are used for both input and output reverse polarity protection.

DC REGULATION TESTS

My first task was to determine the DC regulation of the CT102 power supply with its AC power adapter. The initial no-load voltages when I first powered up the CT102 were ±14.92V DC. With a balanced ±200mA maximum rated load, the voltages at the CT102 output terminals were +14.78 and −15.00V DC. I operated the CT102 at full load, 200mA into 75Ω for one hour.

At the end of this period, all three heatsinks were cool to the touch. The no-load voltages after the heat run were still ±14.92, while the full-load voltages had dropped to +13.48 and −13.16V DC. Decreasing the load to ±22mA produced DC voltages of ±14.84.

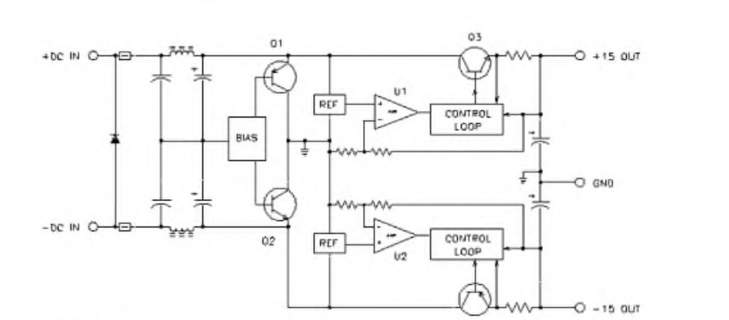

FIGURE 1: CT102 block diagram.

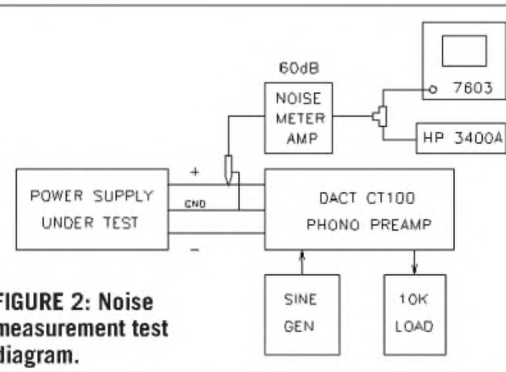

FIGURE 2: Noise measurement test diagram.

Next I applied unbalanced loads, with one supply fully loaded and the other at no load. With only the positive supply loaded, the output voltages were +13.84 and −14.97V DC. Loading just the negative supply yielded +14.45 and −13.25V DC.

The heatsink for the two power transistors that form the phantom neutral became a bit warmer as the transistors worked to maintain a balanced input voltage ground reference. The mains adapter output voltage into the CT102 was 46.5V DC for both unbalanced load conditions. The DC output impedance for the power supply worked out to be 0.25Ω for the positive supply and 0.39Ω for the negative supply.

The AC adapter used with the CT102 is a switching power supply. The broadband ripple from the mains adapter on the input DC bus was 15mV RMS, composed primarily of the 28kHz switching frequency component. There was virtually no evidence of any 60Hz power line artifacts using an analog scope. The adapter no-load DC voltage was 47.8V DC, dropping slightly to 46.6V DC when the ±15 outputs were full loaded to 200mA.

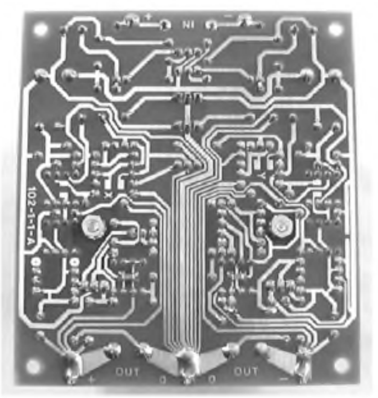

PHOTO 2: Bottom view of CT102 board.

Its line regulation was very good. In response to my isolated variable AC supply, it jumped into action at 28V AC line voltage, producing 30V DC with some noticeable switching noise. It dropped out at 20V AC with a decreasing line voltage. The CT102 was able to produce ±15V DC at 200mA with a line voltage to the mains adapter of 34V AC. The minimum ΔVin-Vout was 2.33V DC on the positive output and 1.71V DC on the negative output, in order to maintain that 200mA output.

TESTING FOR NOISE

Measuring the noise of a power supply involves isolating the very small AC noise signal (less than 1mV RMS) from the 15V DC supply rail. The measurement needs to be performed over a wide bandwidth, 1MHz or more. One means of displaying the AC noise content is with a high-quality analog oscilloscope in AC input mode. But determining the RMS value is another matter.

My most sensitive wideband meter is an HP 3400A true-RMS AC voltmeter, but its lowest range is 1mV full-scale (FS), and there are no scale markings below 0.1mV. When I attempted to measure the RMS noise on the CT102, it was a tentative 0.05mV.

Clearly, I needed something to amplify the AC noise signal so I could read it on one of the higher scales on the 3400A. I designed a 60dB gain (Av=1000) "Noise Meter Amplifier" (NMA) for this purpose, and powered it

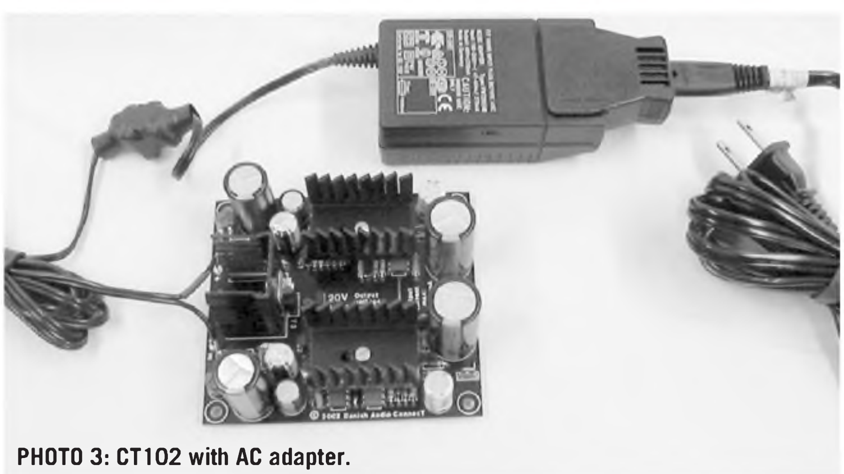

PHOTO 3: CT102 with AC adapter.

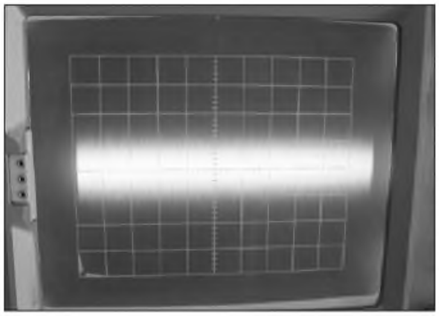

PHOTO 4: Noise meter amp scope, input open (10µV/div RTI, 2ms/div, 4.7µV RMS).

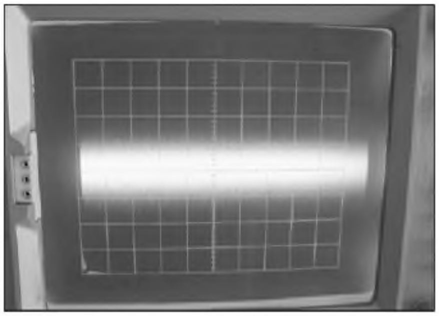

PHOTO 5: Battery noise scope (10µV/div RTI, 2ms/div, 4.9µV RMS).

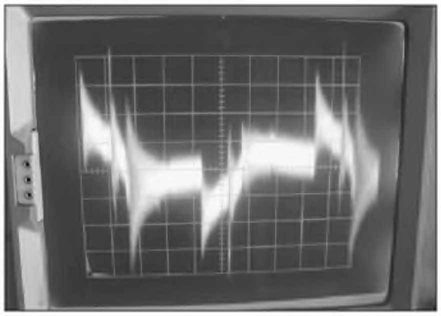

PHOTO 6: CT102 noise scope (20µV/div RTI, 2ms/div, 18µV RMS).

with two 9V batteries to keep the noise introduced by the amplifier itself to a minimum. AC coupling is used at the input, and the gain of 1000 allows me to directly read the noise in µV on the HP 3400A mV scales.

The noise amplifier circuitry will generate noise of its own through the input noise voltage and current of the op amps (e_n and i_n) and the thermal Johnson noise from all its resistors, so care is needed in the selection of components. I used two low-noise AD745N op amps that Walt Jung graciously provided from his sample stock (the 8-pin DIP version is no longer available), each with a gain of 30dB (Av=31.62). I built the circuit on a ground plane PC board, and at 60dB gain it proved to have 3.6mV RMS residual output noise with the input shorted, and 4.7mV RMS with the input open (using the HP 3400A). The noise floor referred to the input (RTI) is thus 3.6µV RMS and 4.7µV RMS, respectively. This is the output noise divided by the noise gain of the amplifying circuit.

In addition to the CT102, I wanted to test some other power sources to obtain comparative data. I selected the following supplies:

- Battery power (two 12V 1.3Ah sealed Panasonic lead-acid cells, with 2200µF 35V caps in parallel as recommended by DacT).
- The CT102 supply set for ±15V DC and using its AC power adapter.
- The ±15V DC regulated preamp power supply by Gary Galo in *TAA* 4/90, page 47, using LT1085 and LT1033 TO-220 linear regulators.
- A ±15V DC regulated power supply using LM340-15 and LM320-15 TO-220 linear regulators (equivalent to 78T15 and 79T15).
- An HP 6236B supply with dual-tracking transistor regulated outputs (vintage late 1970s), set for ±15V DC.

Initially, I loaded each supply rail with Caddock MP915 TO-style non-inductive power film resistors. I used a 200Ω and 500Ω in series to get ±21mA load current. The Mouser part numbers are 684-MP915-200 and 684-MP915-500, respectively. Next, I connected each supply to the DacT CT100 phono preamp, which draws ±22mA quiescent current.

Finally, I connected the CT100 to my sine-wave generator to measure the amount of amplified signal that was passed to the power-supply rails at various frequencies. This is a function of the power-supply rejection ratio (PSRR) of the phono preamp as well as the dynamic impedance of the power-supply regulator circuitry.

The test setup block diagram is shown in *Fig. 2*. A Tek scope probe connects the power-supply rail to the input of my NMA. The output of the NMA is T-connected to a Tek 7A16A vertical amplifier plugged into a Tek 7603 100MHz scope mainframe and the HP 3400A true RMS AC voltmeter.

I have a Tek 7A22 differential amplifier whose lowest range is 10µV/div, but it is limited to ±1V input at this range. While the AC noise was lower than this, I was worried that the initial connection to the 15V rail would couple an overvoltage transient into the 7A22. It was safer to amplify the noise, then connect it to the standard vertical amplifier.

As I mentioned earlier, the wideband residual noise referred to the input (RTI) for my NMA with the input open is 4.7µV RMS, or –113dB relative to a preamp output of 2V RMS (dBr). A scope photo of the open circuit NMA noise is shown in *Photo 4*. I connected the 12V DC battery to the NMA input and measured a broadband RTI noise level of 4.9µV RMS (–112dBr), with a scope photo in *Photo 5*.

Both photos are with the scope vertical amplifier set at 10mV/div, which is equivalent to 10µV/div RTI. I chose a time base setting of 2ms/div to make any 60Hz or 120Hz power line ripple easy to spot. There was essentially no difference whether I used the Caddock resistors or the CT100 phono preamp as the load.

The next step up in noise level is with the CT102 power supply and its AC adapter. The wideband noise on the positive rail measured 18µV RMS RTI, while the negative rail showed 19µV RMS RTI noise (–100dBr).

The scope photo of the positive rail noise is shown in *Photo 6*, with the vertical scale increased to 20mV/div (20µV/div RTI). While it looks quite messy, this noise is actually a low 160µV p-p. Again, there was no difference whether I used the Caddock resistors or the CT100 phono preamp as the load.

OTHER POWER SUPPLIES

The regulated DC supply designed by Gary Galo had an RTI noise of 42µV RMS on the positive rail and 50µV RMS (–92dBr) on the negative rail. *Photo 7* shows the scope trace, which has more of a white noise characteristic than the CT102. The vertical scale here is now 100mV/div (100µV/div RTI). The 4700µF input and output filter caps are Panasonic TS.

The next supply was the LM340/LM320 design. This has a similar topology to Gary Galo's design, except it uses regulator ICs with a fixed ±15V DC output rather than adjustable regulator ICs. Both rails showed 105µV RTI noise (–86dBr). The scope photo, again at 100mV/div, is shown in *Photo 8*. A bit of 120Hz full-wave rectification ripple is now present. This may be due in part to the less expensive Xicon 4700µF filter caps used in this particular supply.

My final supply test used the HP 6236B. Despite the large 2N3055 TO-3 power transistors and Mallory filter caps, the RTI noise was the highest of the bunch at 340µV RMS RTI (–75dBr). This is actually below its specified 0.35mV RMS or 1.5mV p-p ratings. *Photo 9* shows the resultant scope trace, which was made at 500mV/div (500µV/div RTI).

My final test series was to apply a sine-wave signal to the CT100 phono preamp and drive it to 2V RMS output into 10k. I monitored the power-supply rails for any sign of the sine-wave test signal. I restricted my tests to the CT102 and Gary Galo supplies. Neither supply showed any disturbance beyond its residual level until about 10kHz.

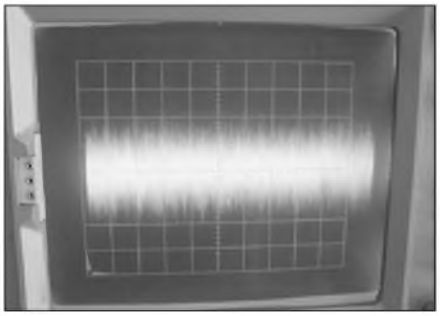
PHOTO 7: Galo supply noise scope (100µV/div RTI, 2ms/div, 42µV RMS).

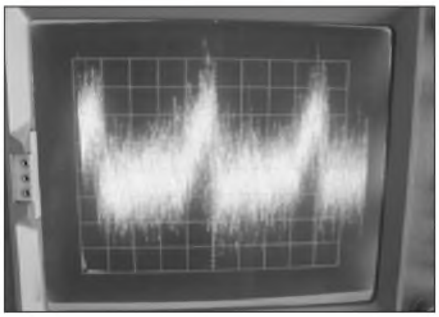
PHOTO 8: LM340/LM320 supply noise scope (100µV/div RTI, 2ms/div, 105µV RMS).

From there up to 50kHz, the amount of signal on the power-supply rails increased steadily, reflecting the falloff in PSRR (and CMRR) of the preamp and the increasing impedance of the power-supply itself.

I took two photos with a 20kHz 2V RMS output signal with the scope time base set for 50µs/div. In *Photo 10*, the scope photo of the Gary Galo regulator, the 20kHz signal is now high enough to modulate the wideband noise, which has increased from 42µV RMS to a still very respectable 68µV RMS RTI (–89dBr). The vertical scale is 100mV/div or 100µV/div RTI.

Photo 11 shows the +15V rail noise at 20kHz from the CT102 at 20mV/div (20µV/div RTI). While the wideband noise is still 19µV RMS RTI (–92dBr), the ghost of the 20kHz signal is just noticeable above and below the broadband noise.

The CT102 noise is specified with an IHF "A" weighting filter, so I also measured the rail noise with "A" weighting. It was 3.5mV RMS (3.5µV RTI, or –115dB relative to 2V RMS). The noise relative to 0.775V over a 22kHz bandwidth is –118dBu.

CONCLUSION

The discrete transistor regulators in the CT102 produce a very low noise level. While not on a par with a pure battery supply, the CT102 has wideband noise levels –8dB better than the excellent Gary Galo IC regulator supply. This margin increases with frequency. *Table 1* shows the CT102 ratings compared with the measured results.

Manufacturer's response:

1. CT102 has a built-in current limit at 200mA. We are using a constant-current limiting principle rather than the fold-back current limiting principle that is mentioned in the review.

If CT102 is asked to deliver (determined by the load impedance) a current right at or above the current limit, it will no longer function as a voltage generator (which it is intended to be) but will function as a constant-current source of around 200mA. Therefore, in testing or using CT102, it's important to ensure that it operates within its voltage generator load range. It's our opinion that when the reviewer uses 200mA output current for several of his tests, he might have gone just to the limit or outside the intended operating area of CT102.

We are uncertain whether this is the case or whether there was something wrong with the switch mode wall adapter that we supplied the reviewer. We have indications that some of the initial switch mode adapters were not able to supply the full output current when operated at 110-115V mains.

2. Our output noise figures are different from the reviewer's noise measurements, so we need to communicate with him about what causes the differences.

Specifically, when testing various CT102 specifications using a load of 75Ω, the risk is that CT102 goes into its current source mode. In other words, the excellent voltage source properties of CT102 are lost. We are questioning the 75Ω load resistor itself. First of all, it has a tolerance and could turn out being lower than 75Ω, leading to a current draw higher than 200mA. Even more critical, when it comes to the long-term test, is where the power dissipated by the 75Ω resistor will lead to it becoming hot. This normally results in a drop in resistance, meaning a larger current than 200mA.

So while testing the maximum output current that CT102 will deliver is a natural part of a review, we suggest testing other specifications to be carried out at a current below maximum current for the reasons given. Our published data is factory tested at 190mA and specified at 100mA output current. Normally, we would expect measuring output impedance, and so forth, carried out at somewhat below short-circuited output current; for instance, 150mA.

The result here is that the reviewer's measured DC output impedance goes way high compared to what it would be if CT102 was operated in its voltage generator range. We have verified and still reach the figure saying that the CT102 output impedance typically is 0.001Ω from DC to 5kHz, slowly rising above 5kHz to 0.006Ω at 100kHz.

TABLE 1
CT102 KEY SPECIFICATIONS, MEASURED PERFORMANCE

PARAMETER	MANUFACTURER'S RATING	MEASURED RESULTS
Output voltage (jumper selectable)	±15V DC or ±20V DC	Verified
Maximum steady-state output current	200mA	Verified
Peak output current	10A (short-circuit protected)	Not tested
Output impedance	0.001Ω at 1kHz 0.002Ω at 20kHz 0.006Ω at 100kHz	0.25Ω at DC
Output noise (wideband)		19µV RMS (–100dB relative to 2V RMS)
Output noise (IHF A)	–126dB	–115dB
Full-load regulation	No spec	±11%/–14%

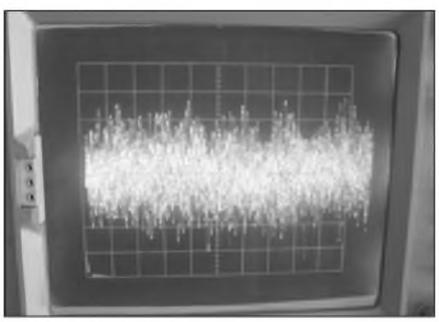
PHOTO 9: HP6236B supply noise scope (500µV/div RTI, 2ms/div, 340µV RMS).

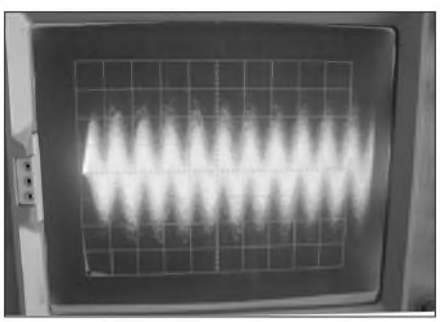
PHOTO 10: Galo supply noise at 20kHz scope (100µV/div RTI, 50µs/div, 68µV RMS).

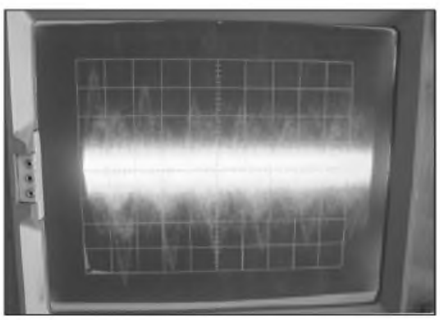
PHOTO 11: CT120 supply noise at 20kHz scope (20µV/div RTI, 50µs/div, 19µV RMS).

Adding to this is the uncertainty about the current output capability of the switch mode adapter that we shipped for the review, which means that the CT102 output current and specs might have been degraded when operated at 110VAC (we are suspecting this problem might occur only when operating the adapter at 110V mains).

The fact that the reviewer finds a rather higher DC output impedance also at 22mA out does indeed indicate that there is something wrong with the switch mode adapter we supplied.

Other corrections to the review are as follows:

Under the section titled Topology, foldback current limiting should be constant-current limiting.

Under Other Power Supplies, we anticipated 50µV noise to be –100dBr rather than being –92dBr. Maybe a misunderstanding on our side?

When measuring CT102 output noise we are deducting the analyzer's own noise. Our typical noise figures are:

- *Measured noise (IHF-A weighted): 0.9µV.*
- *Analyzer own-noise (IHF-A, input shorted): 0.8µV.*
- *Calculated CT102 output noise:* $\sqrt{[\sqrt{(0.9)} - \sqrt{(0.8)}]} = 0.41\mu V = -127.7dB$ *relative to 1V.*
- *Measurements carried out using the Panasonic VP-7722P Audio Analyzer.*
- *This leaves a big difference between the reviewer's and our noise figures, and we do not find any obvious explanation on the differences.*

The review initially states that the features we claim for CT102 are low dynamic output impedance, extremely wide bandwidth, and fast regulation. This is correct. It would have been interesting if the review had discussed whether these parameters are considered relevant for an audio power supply.

Also, the review does not conclude in detail whether the claimed features have been obtained. A comparison between CT102 and the other kinds of power supplies already discussed in the review would have been very interesting. The review does mention in the conclusion that ". . . the CT102 has wideband noise levels –8dB better than the excellent Gary Galo IC regulator. This margin increases with frequency." We had hoped that the review would document more precisely how the margin increases with frequency, as we are especially claiming high-frequency features.

The conclusion does not deal with our claim that CT102 has a very low dynamic output impedance over a large frequency range. Although we find this feature extremely important for audio electronics circuits, the output impedance is only measured at DC (and we find the measured DC value to be different from what they ought to be, possibly because of a faulty switch mode adapter). For instance, high-frequency output impedance measurements might have put the specifications of CT102 in perspective even when compared to battery power supplies.

Finally we trust CT102 is a world-class audio power supply and that its abilities have not been fully and correctly expressed in the review. If the worse-than-we-expected measurements are caused by a faulty mains adapter supplied by DACT, it is, of course, our responsibility. However, we hope to be given a chance to show that the tested CT102 including mains adapter is not a typical specimen. ❖

Allan Isaksen
Danish Audio ConnecT

Xpress Mail

PHILIPS MANUALS

Here is a description of my quest to purchase a Technical Service Manual for the Philips SACD-1000 SACD/DVD/ CD player.

Early in November 2002, after consulting the Philips website, I obtained an address for the Philips Corporation, and wrote a letter requesting the price and cost of shipping and handling of the Technical Service manual.

After waiting a week or two or longer, I contacted Philips through their website. I sent a query asking the same question as had been in my initial letter ("what would be the cost of a Technical Service Manual including shipping and handling?"), and received a reply on Nov. 30, 2002. This e-mail reply came from Philips USA Customer support:

"Unfortunately, the service manual is not available to customers."

I replied in return, explaining that I was/am an experienced electronics and audio systems repair person (which I am, having worked in electronics for about 20 years, and having been an audio enthusiast and constructor for even longer). I tried to explain that I have been making use of legitimate equipment documentation for quite a while now . . . often for professional reasons.

I received an e-mail reply on Dec. 2, 2002, apologizing for the inconvenience and giving me a phone number for their ASC Technical Assistance department where ". . . customer service professionals will be happy to assist you."

A day or so later, I called the number provided and was placed in a waiting queue—which should be very familiar to most people who have made phone calls to almost any of today's organizations. The friendly voice advised me that a service person would become available as soon as possible. After I waited about 5 or 10 minutes, a phone began to ring (it seemed), so I let it ring, hoping to contact service personnel. But the phone continued to ring, until it stopped, to be replaced by a busy signal. I then attempted to call the same number later that day, and for about two or three days after, always getting a busy signal.

On Dec. 5, 2002, I sent an e-mail, describing my inability to make contact via the supplied phone number. On Dec. 9, 2002, I received another reply from Philips USA Customer support, apologizing for the inconvenience and supplying me with another (different) phone number.

Calling this phone number, I did get through to customer service personnel (off-shore?) and was advised that I would need to call a different phone number to obtain information on obtaining service documentation, and that the service person I was speaking to ". . . did not have that number available at this time."

(Around this time, I did receive a reply to my original "snail-mail" letter [dated Nov. 26, 2002, postmarked Dec. 2, 2002], in reply to my original sent earlier in November, from the Philips Customer Care Center in Miami, Fla., which provided a phone number for their Technical Support Line. Unfortunately, this phone number did not appear to be in-service.)

Thus, I replied to the Dec. 9, 2002, e-mail describing the situation and asking for a phone number to enable me to contact someone at Philips who could supply the information I needed for ordering a Technical Service Manual.

The final reply, received Dec. 11, 2002, from Philips USA Customer Support, informed me that "Unfortunately, service manuals are not available to customers. These manuals are only available to authorized servicers."

I have occasionally obtained service manuals (by ordering and paying for them) over the years—in several cases, for a variety of Philips/Magnavox CD players and related devices (about six different items)—from around 1987 up to around 1992 and had never encountered any kind of difficulty. The Philips Corporation's service division supplied the info on cost and postage, and I sent payment and received the goods in question—no problems of any kind. During that same period I experienced very good service from SONY.

In more recent times, I have ordered and paid for service manuals for items of interest (Panasonic, Pioneer, and Sony DVD players) over a period from early 2000 up until only a month or two ago, and have received the required information promptly (via e-mail in these more recent years), sometimes ordering the Tech Manuals by telephone from the various companies' service departments (most recently, from SONY, which was very efficient, effective, and professional).

Disappointingly, Philips has not behaved the way I assumed they would, in the light of past years' experience, as I have described here.

Thus, to round this up, does anyone know how this situation can be reversed, and how a professional relationship can be re-established between an important equipment manufacturer and its customers, as was once the case in the past?

Robert K. LeBeck, Jr.
Sequim, Wash.

FROY USER

I've been using the Froy III speaker for about three years now and love it. Your comments on the sound of the speaker sound correct; it is balanced almost like the THOR (I've heard a reflex-loaded version of THOR with Joe's crossover), except it is set up to be a little more forward than the THOR because the designer prefers that type of sound believing live music is rarely proper. I know that because I have been good friends with Murray Zeligman for decades (yes, that does make me biased) who has designed some very interesting electronics and speakers over the years.

If you're interested in learning about the Froy you can contact Murray at Zeli@erols.com. I believe you'd find his thoughts on the Froy speaker and audio very interesting.

Allen Edelstein
hahax@rcn.com

DIY FORMAT

I have been a reader and subscriber of *Speaker Builder/audioXpress* for over three years. The current series of projects done by the editor Ed Dell seem much improved for someone like myself. The full-size diagrams and multiple forms of assembly instruction—picture, schematic, and layout diagram—should be used in all the articles submitted.

So thanks for this format and I hope it continues and is applied to all projects.

Lon Ponsehock
lon@athenet.net

CABINET THICKNESS

I see the May '02 article about the THOR displays a different sidewall thickness than the SEAS website. At the end of the article there is a clarification about this, which lists a website from SEAS that doesn't come up. They do have the THOR plans, with the 1″ thick cabinet. I see it is 9″ wide at the front. The *aX* May '02 article also shows the front width at 9″, but with ¾″ walls. There is an internal difference of ½″! Which is correct?

Neal Egbert
djcruel1@interl.net

The website THOR is an early version where SEAS assumed that D'Appolito's design included 1″ walls. That is incorrect. D'Appolito's design calls for a 1″ front panel and ¾″ walls.—E.T.D.

TUBE SELECTION

Regarding Neal Haight's comments about the K-12M amplifier made by S-5 Electronics (Xpress Mail, 5/03, p. 64), the 11MS8 tube was used for the vertical oscillator and output stages in many Japanese import portable black and white TVs, vintage 1970. Many of these sets used an 11BM8, with the pentode used as the sole audio stage, while the triode served as the video synchronizing pulse separator. A huskier brother to the –BM8 is the 6GW8/ECL86, but its pin configuration is considerably different from the –BM8, and adapting the K-12M printed circuit board to it may not be feasible.

Frankly, if I were making a small, low-cost power amplifier, I would use a 12AX7 for the voltage amplifier/phase splitter and a pair of 6AQ5s for the push-pull output. Why were the vertical sweep 11MS8s chosen for the K-12M?

Michael Kiley
Crestwood, Ill.

UNDERESTIMATED VARIABLE

The Fletcher-Munson curve has a profound effect on component voicing that is very underestimated. Unless you listen to music at 90dB, you lose up to 10dB of bass and get a prominent high-mid peak at 3–5kHz, which makes music sound bottomless and glaring.

Thus, a loudspeaker that is voiced at 90dB will sound fine when listened to at that level but not at the normal 70–75dB level (a reggae album will lose all deep bass and guitar solos or Bob Dylan's harmonica will be piercingly loud), while a speaker voiced at 75dB will sound full at that level and have prodigious bass output at 90, which is why most people turn it up anyway!

The effect of the curve should be taken into account at all stages: voicing of each component, monitoring, mixing, remastering, and so on, or else balance problems will arise. Would a 75dB standard level be a solution?

Robert Laberge
Québec, Canada

AM ANTENNAS

Thanks to Chris Campbell for his comments on my simple review of two commercially produced AM antennas ("Xpress Mail," 4/03, p. 63). My unstated motivation was to attract the interest of knowledgeable individuals who would write about AM reception! Publication of the review must have been regarded as such by Mr. Dell . . . a trial balloon to open up a fascinating field, and to attract in-depth articles and construction projects!

I would appreciate Mr. Campbell contributing some of his research on the Internet. You could possibly write an article about building an AM antenna to the builder's choice of frequency.

I am interested in 1030kHz, where WOSO transmits from San Juan.

Alternately, you could direct readers to sources of pertinent information. Both contributions would be heartily appreciated!

Carlos E. Bauzá
Gyama, PR

PHONO PREAMPS

In reading your product review of budget phono preamps (April 2003, p. 40), I was dismayed to see three of the seven products are crude transistor designs typical of the mid-to-late 1960s. Anyone still wishing to use this topology should see the article "Low-Level Tape or Phono Preamp" by Richard Moore (*The Audio Amateur*, 4/74, p. 4). Here, Mr. Moore has improved the design by using a 2MΩ collector load resistor on the first stage, a Darlington pair for the second stage, and a higher than usual impedance feedback RIAA EQ network. These revisions increase the open loop gain of the preamp, reducing distortion and improving RIAA equalization accuracy.

In the IC op-amp method, I use the design of the RIAA phono section of Walt Jung and Dave White's "A PAT 5 Modification" (*The Audio Amateur*, 1/78, p. 7). Even cheapening the circuit by using TL071 ICs and nearest standard value 5% carbon film resistors and polyester film or silver mica capacitors, the sound quality is much better than that of most magnetic phono preamps included in mass-produced receivers.

Michael Kiley
Crestwood, Ill.

CIRCLOTRON

The article on the circlotron amplifier of April 2003, p. 32, sounds interesting, and I intend to build the amplifier since I already have the tubes. However, I have noted there is a printing error between the bridge rectifier BR1 and fuse F3 (secondary 115 AC for HT). The amplifier schematic (Fig. 2) shows the two cathodes of BR1 (positive HT supply for anode of V3A) are linked to fuse F3. The line between the fuse F3 to the two cathodes of the BR1 bridge rectifier should not be there.

The local mains supply voltage is 230V–50Hz. In regard to the output transformer, the author recommended the use of a 120/17V step-down transformer. There is a possibility that such transformers can be obtained from the scrap yard, but I am just wondering whether a 230/34V could be used in its place? If so, how will this affect the output in regard to impedance and sonic aspect? Is it necessary that one side of the secondary for the loudspeaker be grounded? It will be very interesting for me as well as those who intend to build this amplifier.

A.T. Hoh
Singapore

Monny Nisel responds:

Thank you, Mr. Hoh, for the interest shown in my article, "6AS7/6080 Circlotron Amp." As you rightly mention in the first paragraph of your letter, an error has slipped into the schematic, an error I missed. The link between the F3 fuse and the positive output of the rectifier bridge should be removed, as it shorts, practically, one leg of the bridge. The circuit should look identical to the lower one using the BR2 rectifier bridge.

Regarding the transformer questions, I shall mention first that there is another small error in the schematics, as TR1 and TR3 notations are reversed to fit the parts list.

The transformer you mention, 230/34V, used with an 8Ω speaker, will reflect into the secondary an impedance of 366Ω; the transformer I recommended reflects a 450Ω impedance (or close to that), which seems to be the optimum from the point of view of power output, so you may expect a drop in power, though not significant.

For the same power, a 230V transformer has almost double the turns of a 120V one, which translates in a higher primary inductance with its pros and cons.

A higher primary inductance will favor the lower side of the frequency range but at the same time will drop the output at the higher end; a higher primary inductance will also reduce transformer distortions. Again, all those may not be significant as the total number of turns and inductance is low in a power-supply transformer compared to an output transformer.

What I suggest, Mr. Hoh, is the following: if you already have the 230/34V transformer, give it a try. The measurements and especially the listening tests will indicate whether you need to buy the one I recommended.

Regarding the last question, there is no need to ground one side of the secondary of the output transformer, which I tried once myself. However, the results were not so great, according to my taste, and I do not recommend it.

IONOVAC

Regarding "Build a Plasma Tweeter" (4/03, p. 4), if I remember correctly, the Ionovac used helium rather than air as the working gas. This gave the whole metal horn chamber a blue glow when it was working. I think the theory was that helium was the lightest non-combustible gas and would be easier to pressure-modulate with the audio signal.

Charles Hansen
Ocean, N.J.

Colin Jcye responds:

Thank you for bringing up this item for discussion. While I can only claim through word of mouth that I've heard of such a helium-powered ion tweeter being used, I'm also a bit curious as to the theory behind its intended operation.

The one good reason I can think of is that helium is a very inert gas, so, in this absence of oxygen, there is no ionization of oxygen, and thus no ozone production. However, every high voltage hobbyist I've shown the ion tweeter to has been very amazed by the surprisingly low amount of ozone production.

From my days of playing with sparks, I remember the sharp smell of ozone that lingered in the air after the occurrence of such a sudden air breakdown. I believe the reason for this low ozone production may be the fact the local flame region is not a short burst, in which case the air molecules are suddenly ionized and immediately pushed away from the spark region by the intense air pressure gradient. It seems to me that the steady, more spherical, region of ionization affords the ionized oxygen molecules the chance to recombine as they drift more toward the cooler outside boundary of the flame.

The fact that the ozone production is so low does not mean that it is non-existent, though. I can definitely smell it after about 30 minutes of operation in my non-ventilated dorm room here at MIT. In a more ventilated room, I could run the unit for about an hour before noticing the ozone smell, and even then, it was not objectionable.

The expense and inconvenience of using *helium does not quite justify its use in the ion tweeter the way I just explained it, so there must be other reasons. The surprising thing to me is the ionization potential of these different gases. For nitrogen (N_2), the ionization potential is around 14.5 electron-volts (eV), for oxygen (O_2) 13.6eV, and for helium (He) 24.5eV. This seems backwards at first, because it implies that you need around twice the energy to ionize He (and hence twice the power) as you do for N_2 or O_2.*

However, He is a monatomic molecule, which generally means that the recombination rate of the ions and electrons is much lower than that of O_2 and N_2, which are both diatomic molecules. Furthermore, N_2 and O_2 can have multiple ionization states (even at this low temperature plasma), which may lead to nonlinear effects. Helium appears to have an advantage in both of these arenas. Since helium is of significantly lower density than air and it has a lower recombination rate, perhaps the plasma flame is larger for helium than air for the same amount of oscillator power. I believe the recombination rate is key, because you could ask why they didn't just use dirt-cheap N_2 gas instead of He if ozone was the problem.

About two years ago, this topic came up through e-mail. I have lost contact with the people involved, but I still have some information from the following e-mail:

"I found this [helium-driven plasma tweeter] in the October 1978 edition of Wireless World. *It comes from a report on the High Fidelity Show in Atlanta and the Consumer Electronics Show in Chicago . . . the Plasmatronic Hill Type 1 plasma driver, which claimed to provide exceptional clarity of reproduction (zero coloration and laser-like phase coherence) from 700Hz to above the limits of audibility, and operated via a bi-amped system with a 12" subwoofer and 5" mid-range unit. The plasma driver requires topping-up with helium in cylinders from your local welding supplies company, every 300 hours of listening time. The cost is claimed not to exceed 30 to 40 cents per hour."*

From another e-mail:

"I almost purchased a Plasmatronics Hill Type 1 unit in the US a few years ago . . . The design uses five electrodes fed from five tubes. The five electrodes sit in a ceramic cavity which is about 2cm or so in diameter. Each electrode has inert gas carried to it through the electrode."

Both e-mails indicated the Plasmatronics was the producer of this helium-mystery. I actually spotted an original Ionovac on eBay.com about a year ago, which made no mention of helium gas, nor did the Ionovac schematic. Good sources of information about the theory of operation using helium might be found in patents at http://uspto.gov/.

Indeed, this is a very mysterious device. This is the extent of what I know, but I would like to try experimenting with this concept of helium gas (and maybe even its brother, the much cheaper argon). Perhaps it will lead to another article in a future issue of aX.

Interesting project in the April 2003 issue ("Build a Plasma Tweeter"). I've actually heard fairly broadband sound coming from a failed FM antenna—20kW of RF made a plasma about 18" long. But this is not something that you should run in your house.

I want to point out that the plasma tweeter as described may be violating Federal Communications Commission rules regarding radiated radio frequency energy. There are strict limits for what is properly called an "Unintentional Radiator" in Part 15 of the rules.

The frequency mentioned in the article, 40MHz, is used in the land-mobile industry as "Low Band." Many different users including Public Safety use that band. If any RF energy is radiated from the tweeter—and the description of interference to CD players leads me to believe it is radiating RF energy—the plasma tweeter could be a source of interference that ultimately could lead to a large involuntary donation to the Federal Government. Would you want to learn that your plasma tweeter caused someone's death because the radio interference prevented an ambulance from getting there in time?

While there are clear audio advantages to this device, anybody contemplating building one should take great pains to prevent any RF radiation from the device and fully comply with FCC rules. This requires the use of a field intensity meter to measure the radio frequency non-ionizing radiation to demonstrate compliance with Part 15, not something that is commonly and/or inexpensively available.

Bill Ruck
billruck@earthlink.net

Colin Jcye responds:

Thank you for your challenging and valid concerns. I am aware that the FCC does not look kindly on anyone not complying with their regulations, even if it is in the name of science or hobby, and I encourage experimenters out there to be especially conscious of the problem of unintentional interference if you are attempting this project. Nonetheless, I hope this doesn't scare anyone away from developing this type of acoustic transducer. The issue of interference can be easily handled.

The FCC regulation you mention can be found in section 15.209 at http://www.fcc.gov/Bureaus/Engineering_Technology/Documents/cfr/1999/47cfr15.pdf. It states that the radiation emission limit for the electric field in the band from 30.0MHz to 88.0MHz is 100µV per meter at a distance of three meters.

While I tried twice to get the devices tested in an anechoic chamber (both chances fell through), I must admit that I do not know whether they fully comply with FCC regulations at this time. I know that theory is never a good substitute for measured data, but I would like to point out that for my tea ball square mesh size of ~0.5mm at 35MHz (~8.5m wavelength), the wire mesh essentially looks like a thin conductive shell. For this material, the thickness of the conductive tea ball is no less than four penetration depths, blocking around 99.97% of the power radiated from the flame at that high-Q dominant frequency. Practically, there is no power emitted through the tea ball compared to your microwave oven, which operates at 2.5GHz (0.12m wavelength), with only a thin metal mesh between you and 1000W of RF power.

Scaling appropriately, it would be very easy to shield all 35MHz activity. I did try measuring radiated power from the plasma tweeters using a cheap hand-held meter, which is essentially accurate to an order of magnitude. I had a difficult time reading anything but noise except for the exposed part under the coil, which gave me ~10µW/cm² at a 1m distance with my flame height set at a little over 1cm. Actually, this meter indicated fluorescent lamp emissions at 1m that were only about an order of magnitude less than my plasma tweeter. The shielding could be much enhanced simply by adding some more grounded copper wire arches around the exposed portion of my coil, at which point I would expect it to look a lot like a fluorescent lamp in terms of radiation—probably a lot better.

I would like to experiment with an argon plasma, since the recombination rate for monoatomic gases is much lower than that of diatomic ones. This may allow for a more efficient flame (less power), as well as to prevent ozone formation.

BUILDING IN PROGRESS

As a builder, I keep following the DIY articles and found one project that really struck a chord, Pete Millet's headphone/line amp (Nov. '02, p. 20). Most of the projects involve odd or expensive parts, and I really don't need another exotic power amp or line amp. But this one was different for a special reason. I could put together "kits" for friends who have never built with tubes. The availability of the PC board—and a well-done board it is—was key. The parts suppliers are great to work with, and, of course, the amp works great.

I'm using mine as a line amp in an all solid-state system to add "tube sound" to let listeners hear the difference. Pete's emphasis on the experimental nature of his design is what makes this ideal. One of these kits went to a teen that my wife tutors who has decided on a vocational track in school. He did a good job building it, having just been exposed to basic soldering techniques, and with a little assistance put it together. This was both an eye opener (golly, I built something) and a real confidence builder (it works!). It helped reinforce his career choice, since schools really don't think too highly of working with your hands.

It's my hope that you receive or can encourage your authors to write in this direction. If circuit boards for smaller projects could be more readily available, this might encourage more building as well.

Thanks for a unique magazine.

Jim Albanowski
jimalbanowski@comcast.net

Pete Millett responds:

I've been surprised by how well-received the hybrid headphone amplifier project has been. I've received dozens of e-mails from builders who, like Mr. Albanowski, have been really happy with not only the result of their effort, but the whole experience of building something. It's encouraged many people who've never built an audio project, let alone something with tubes, to give it a try. Without exception, those I've heard from who have tried it have been very pleased with their results.

I think many of us—and I've been guilty of this as well—tend to get caught up in audiophile exotica, and forget about the joy of a simple home-built project. The high cost of high-end components and the complexity of many projects must discourage many would-be audio hobbyists from even trying to build something.

I've also heard from many potential DIYers who don't think they're up to the challenge of doing point-to-point wiring in a project. I've always believed that a well-done PCB can perform as well as hand wiring, and a PCB is a whole lot easier to assemble. The results are also more predictable, I think.

Based on the responses I've received, I certainly will try to come up with more easy-to-build projects such as the hybrid headphone amp. Thanks to Jim Albanowski and all the other builders out there for the great positive feedback! ❖

(from p. 27)
loudness, but is slightly objectionable, like it was with the stock 12″ driver. I was most concerned about the driver not being solid at 41Hz, since any sub driver should handle that with no problem.

With the different Qs the response will probably be down at 41Hz, but the driver should stay clean if you are not really pushing it too hard. There was also a bit of cone or surround breakup in the E_3 to G_3 note range (41Hz–49Hz) at high volumes. Given the low amplifier power available, it cannot be overpowering the driver.

The high notes are not as bright as Chuck would like, even with the treble turned full up. Even though the bass is designed to anchor the bottom end of the rhythm section, it should still have its full range voice, including harmonics.

THE REAL PHOENIX GOLD QX108 DRIVER REVEALED

As you can see, the CB tests showed the QX108 driver has T/S parameters that did not match the published specifications. Using the measured parameter produced a CB design in a 1.24ft^3 box with f_3 at 55Hz and a Q_{TC} of 0.72. Testing in the box verified this design showing that the Q_{TC} measured 0.73. A plot of the design system showed that the response would be down about 7dB by 41Hz (*Fig. 12*).

To eliminate the bass guitar as a variable, Chuck connected the TKO 80 with QX108 into his stereo system, replacing the NHT SW2 subwoofer. This confirmed there was very little useful output below the 49Hz G_3.

With these initial test results in hand, we decided to contact Parts Express, wondering whether they had any experience with the QX108 driver measured performance versus its specifications. The QX108 driver performed poorly in the TKO 80 system, just as the actual measured driver parameters suggest.

At this point we considered going to the second candidate on the driver list, the Peerless 850146 (PE #297-636). Assuming Chuck measured everything correctly and the Phoenix Gold specified data is not reliable, how much more accurate can we expect the Peerless data to be? Chuck mailed the results to Parts Express and asked whether they had tested the QX108 themselves.

Darren Kzuma of Parts Express informed us that Phoenix Gold is a new manufacturer more in the car audio camp, with all their drivers made in China, and the specs and data may not be reliable. The T/S parameters might even vary from driver to driver.

In order to achieve the goal of selecting a driver for the existing fixed volume CB, where the validity of the specifications are important, he recommended we go with the next candidate on our list. He thought the Peerless data could be trusted, so Chuck ordered the 850146 driver. Darren said he would arrange for the return of the QX108.

THE PEERLESS FIX

The Peerless #850146 10″ woofer, Parts Express #297-636, is driver No. 10 in *Tables 1–4* and was originally considered one of the better candidates. We ran it in for 12 hours at its specified resonance frequency, then measured the free-air data. The QX108 used eight #10 mounting screws, the Peerless uses six #8, so the adapter ring has six new holes. We plugged the other eight holes with the gasket material PE sent with the driver.

After installing the driver facing into the TKO 80 cabinet, the ad-hoc air leak test again turned out to be satisfactory. Next came the CB tests to obtain the T/S parameters. The test results are shown in *Table 7*.

A repeat of the free-air tests after we removed the driver showed they had not changed from the first set of measurements. Note again the large difference in the Q_{MS} measurement from the Peerless spec.

LISTENING TESTS—PEERLESS

Chuck installed the Peerless driver in the TKO 80 cabinet, and did some subjective listening. All the notes on the bass are now nicely voiced, without any noticeable peaks or dips in the response. He again connected the TKO 80 in place of the NHT SW2 sub in his stereo system.

While NHT has nothing to fear from the modified TKO amp, the Peerless still acquits itself quite nicely. It reaches down to the 32.7Hz C_3 and 30.9Hz B_4 bass notes on reference CDs with 5-string bass, but the 27.5Hz A_4 in the *Titanic* sound track CD came up silent. It doesn't sound as tight as the NHT, especially at higher volumes, but at normal listening levels it is acceptable. Of course, one of the problems in using the bass amp in an existing stereo system is the limited slope of the TKO passive tone controls compared to the versatile electronic crossover in the NHT subwoofer amp.

The QX108 couldn't deliver the 32.7Hz C_3 in the earlier test. There was a pronounced muddy, fuzzy sound. The improvement in low-frequency extension with the Peerless driver, although not quite as deep as the NHT SW2 subwoofer, is readily apparent. The highest notes are more realistic with a nice midrange presence. The amp is now much clearer and more satisfying to listen to throughout the playing range.

Testing showed the Peerless driver met its specified parameters and that it produced superior sonic performance in the TKO 80 system. Part 2 covers the amplifier and preamp, as well as more on the importance of speaker specifications. ❖

TABLE 7
PEERLESS 850146 MEASURED PARAMETERS

PARAMETER	MEASURED	SPECIFICATION
Re	5.20Ω	5.50Ω
f_S	24Hz (Z = 73Ω)	22.6Hz
Q_{MS}	5.29	2.56
Q_{ES}	0.41	0.40
Q_{TS}	0.38	0.35
V_{AS}	5.27	5.10
f_C	51Hz (Z = 38Ω)	
Q_{MC}	6.27	
Q_{EC}	0.99	
Q_{TS}	0.86	

REFERENCES

1. Bensen, J. E., "Theory and Design of Loudspeaker Enclosures," published by Synergetic Audio Concepts, 1993.

2. Small, R. H., "Vented-Box Loudspeaker Systems, Part II: Large-Signal Analysis," JAES, July/Aug. 1973, Vol. 21, No. 6, pp. 438–444.

SOURCES

Antique Electronic Supply
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www.tubesandmore.com

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Parts Express
800-743-1677
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Tools, Tips, & Techniques

Free Filament and Bias Voltage

It has been more than three decades since I built my last tube-type amp from scratch. It was an integrated stereo unit with 5881 output tubes (ultralinear configuration) and a handful of 12AX7s and 12AT7s or 12AU7s; it weighed a ton and would heat a small room during the winter. The good old days!

One trick that I found useful—but haven't seen mentioned in recent *aX* articles—is to eliminate the negative bias supply and get free DC supplies for the preamp tubes, as well as eliminate the filament turn-on stresses. You know when the filament supply is first turned on and the filaments flare up very brightly.

A BETTER IDEA

On one amp I recall not having a satisfactory tap on the high voltage transformer, nor having enough room for a bias transformer and supply components. So I went with a big, high-wattage cathode resistor. This was a good fix except that it heated the under chassis area to an unacceptable level, even with lots of added ventilation holes. The solution: use the filaments of the preamp tubes as the resistor.

The advantages of this scheme are multiple:

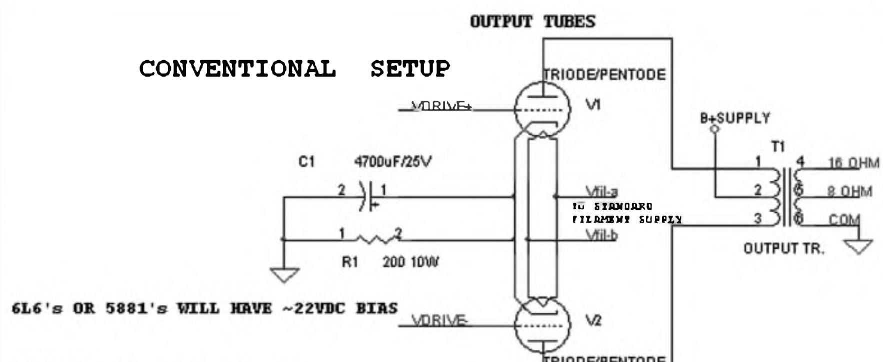

FIGURE 1: Conventional setup.

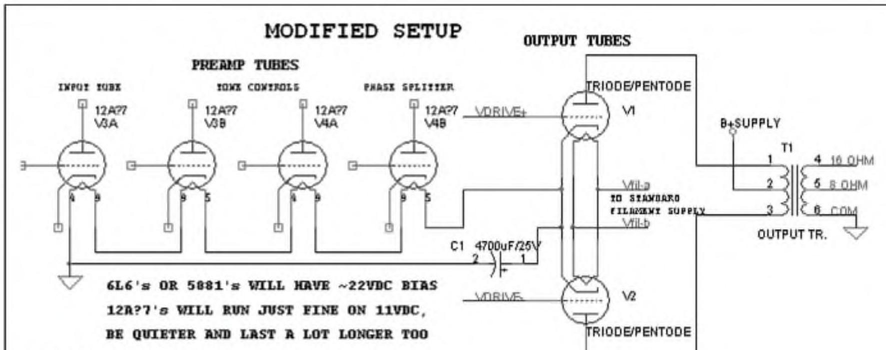

FIGURE 2: Modified setup.

1. The heat is dissipated above the chassis.
2. The preamp tubes have DC on the filaments.
3. The preamp tubes' filaments heat up slowly (as the output tubes warm up and get plate current). There's no super bright turn-on surge as you see in many types of tubes attached directly to a filament transformer.
4. Lower noise from the preamp tubes.
5. Fewer parts.
6. No need for a negative bias supply or a tap on the HV windings for the supply.
7. Smaller filament transformer current requirements.

It is still desirable to have a large-value filter cap across this filamental "cathode resistor." When configuring the filaments, I found it best to have the input stage tubes' filaments closest to ground. You can series or parallel filaments based on your output tubes' current requirements, but if in doubt the total of the filament chain making up the cathode resistor can have less than rated voltage applied and work well.

That is, if you put two 12AX7s' filaments in series, they could run with only 18 to 20V DC across them. If you need more current, parallel the filaments. An integrated amp can easily have five or six 12A*7 tubes, so your application should have a lot of flexibility. And you can always add a small cathode resistor to cover any small bias voltage shortages.

One minor caveat: cathode-to-heater voltage ratings. Check whether any tubes used for the "cathode resistor" have been in a circuit that has a high cathode voltage (as you see on some types of phase-splitting circuits). Be sure not to exceed the tube's cathode-to-heater voltage ratings.

I hope this proves of some use to readers. I am sure that some will shoot holes in this idea, but to each his own. ❖

Patrick M. Brunner
Brooksville, Fla.

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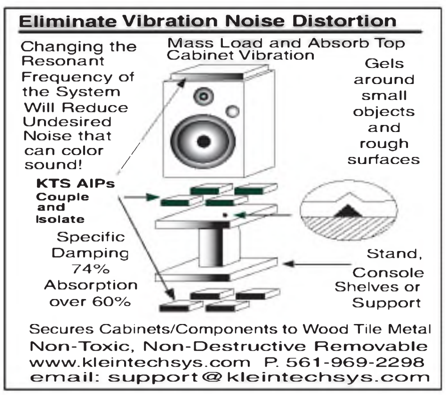

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David Lucas manuals and materials for electrostatic speakers. If you or someone you know has purchased David Lucas manuals and kits, please contact: Angel Rivera 941-798-9227 (Florida), rivera_angel@hotmail.com, or alrivera@tampabay.rr.com.

Complete workshop manual for Marantz 10B Stereo FM tuner. Original or photostat. Will pay in US $ or £s sterling. Clive Kavan, 9B, Ashbourne Parade, Hanger Lane, Ealing, London, W5 3QS., England. +44 20 8998 8127, FAX +44 20 8997 0608, Mobile: +44 7860 519748.

Schematic for Luxman C-1010 solid state preamp. Tom Tutay, Transition Audio Design, 24 Elm Avenue SE, Ft. Walton Beach, FL 32548, 850-244-3041, FAX (850) 244-2498.

AD INDEX

New Chips on the Block

By Charles Hansen

Pacific Microsonics PMD-200 HDCD

Pacific Microsonics has introduced their new PMD-200 high definition compatible disc (HDCD) chip for CD and DVD audio and video players. The device provides HDCD decoding, filtering, and processing, accepting up to 24-bit input data at sample rates from 44.1kHz to 192kHz. The chip is based on the Motorola DSP56364 digital signal processing engine, and can handle 100 million instructions per second (mips).

In addition to featuring HDCD decoding and precision filtering for the over 4,000 HDCD-encoded CDs available today, the PMD-200 adds HDCD precision conjugate filtering for DVD-video and DVD-audio. These multi-format HDCD capabilities make the PMD-200 ideal for consumer electronics products such as DVD players and A/V receivers that must be capable of playing audio from DVD-video discs, HDCD CDs, conventional CDs, and HDCD DVD-audio discs.

Pacific Microsonics, Inc.
Contact: Andy Johnson
510-475-8000, ext. 102 (US and Europe)
ajohnson@hdcd.com
www.hdcd.com

Analog Devices AD1833/AD1836

The Analog Devices AD 1836 is a high-performance, multi-channel, single-chip coder/decoder (Codec). It provides six digital-to-analog converters (DACs) and four AD1855 audio analog-to-digital converters (ADCs) using ADI patented sigma-delta techniques. It supports multiple digital stereo channels with 24-bit conversion resolution and a 96kHz sample rate. Thus, it is fully compliant with the newest Digital Versatile Disk (DVD) audio specifications. The AD1836 also includes six independent volume controls adjustable via an SPI-compatible serial control port.

The AD1836 is designed for home theater equipment, automotive audio systems, and digital audio-effects processors. The AD1833 has a similar architecture and is targeted for DVD players.

The AD1836 and AD1833 feature a 108dB dynamic range for each of the six DACs, and a 104dB dynamic range for the four ADCs. It means that designers can build smaller, lower power-consuming systems without having to compromise on the accuracy of the sound. The multi-bit Sigma-Delta architecture reduces idle tones, and the patented data-directed scrambling minimizes sensitivity to jitter.

To reduce design complexity and reduce overall cost, the devices have been optimized to interface directly with the ADI SHARC® family of digital signal processors (DSPs).

The AD1836 52-pin MQFP (PQFP) packaged Codec is priced at $9.50 US in 1000-piece increments. It requires a single +5V supply, but it also has a separate supply pin for its digital interface, allowing it to be used with +3.3V devices.

The AD1833, a similar 48-pin LQFP packaged device, will cost $6.80 US in 1000-piece quantities. It, too, requires a single +5V supply, with a separate supply pin for its digital interface, allowing it to be used with +3.3V devices. ❖

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