Shhhhhhhh! – How to Quiet Cabinet Panels

JANUARY 2006
US $7.00 • Canada $10.00

audioXPRESS

The Audio Technology Authority

- **Finishing Touches for an Improved Phono Preamp**
- **An Enlightened JFET Design**

REVIEWED: Sound Pax Speaker System
a cardboard speaker that will astonish

Universal OPT for Tube Guitar Amps

www.audioXpress.com

Back

Universal Output Transformer for Guitar Amplifiers

The author proposes a new universal output transformer for use with various tube amp topologies.

By Menno van der Veen

Almost a century after their introduction, vacuum tube amplifiers are still appreciated for their special qualities. Even now, many companies design new valve amps which are highly rated by audiophiles. Musicians—especially guitarists—also like the special musical character of vacuum tube amplifiers.

Earlier research[1-7] about toroidal output transformers has shown that the toroidal transformer is very well suited to function as the impedance converter between high impedance vacuum tubes and the low impedance loudspeaker. The major reasons are the inherent low nonlinear distortion and the very small leakage inductance, which allows for a precise tuning of the high-frequency rolloff and resonances inside the output transformer. However, the toroidal transformer is expensive to manufacture and is not often used in musical instrument amplifiers, which are under high market pressure. So mostly less expensive EI transformers are used.

This new research focuses on these EI transformers, leading to a new concept of a universal output transformer, especially intended for musical instrument amplification.

FIGURE 1: Where to connect an output transformer?

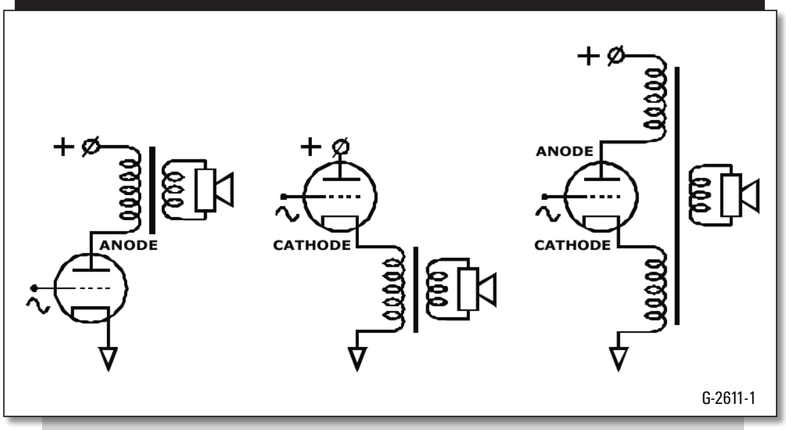
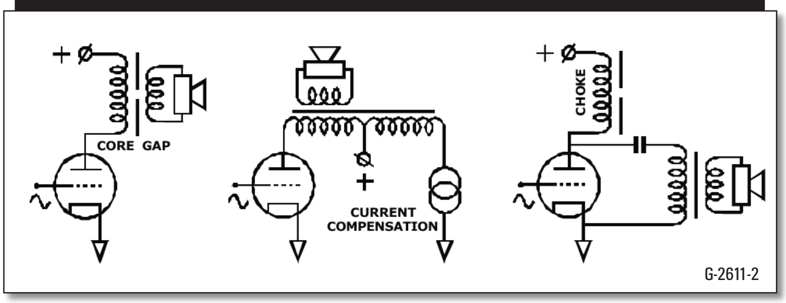


FIGURE 2: OPT connections for SE applications.



BASIC QUESTIONS

There are many tube amplifiers in the world, with many different topologies, such as single-ended (SE), push-pull (PP), paralleled push-pull (PPP), and cathode follower (CF). All tube amps are of different design, concept, and quality. Do they have the same basic concept? If yes, what is the nature of this concept?

Every tube amplifier has its own specific output transformer. Do these output transformers have the same characteristics? If so, what is the general nature of those characteristics?

One final question: Is it possible to design a universal output transformer that will fit into all the valve amplifiers?

If so, then there exists a very good reason to design such an output transformer. But let's first start with the first question about the basic concept of tube amplifiers.

SYSTEMATICS IN DESIGN

In a tube, the current runs from anode to cathode. The voltage on the control grid determines the amount of current and in most situations the control grid does not draw any current. When you wish to extract power from a tube, you should connect the output transformer (OPT) to where the current runs, also to the anode or to the cathode or to a combination of both. Please note that this first systematic approach delivers three possible choices. **Figure 1** shows the situation.

You can use one power tube and connect the output transformer to this tube. The quiescent current of the tube runs through the primary of the transformer. Consequently, the core of the OPT must be gapped, or a kind of balancing current compensation should be applied, or capacitor coupling. This again delivers for three possible single-ended choices (**Fig. 2**).

The next classification considers push-pull amplifiers, in which two power tubes are driven with inverted signal phase at their control grids, while their DC quiescent currents cancel out in the center tapped primary. These circuits with all their variations can be divided into basic push-pull and paralleled push-pull (or configurations such as the Circlotron with two separate power supplies).

For details, see **Fig. 3**. For push-pull, you now have two new choices.

When using a pentode as power tube, you can construct three kinds of local feedback with the screen grid. When the screen grid is connected to the anode, you have a triode. When the screen grid is connected to a tap on the primary winding, you have the ultralinear configuration. When you connect the screen grid to a voltage supply (which might be equal to the anode voltage supply), you have a pure pentode configuration. Again, you have

FIGURE 3: Basic push-pull topologies.

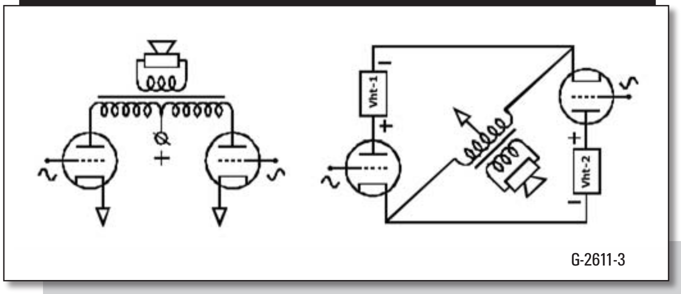


FIGURE 4: Local feedback with the screen grid.

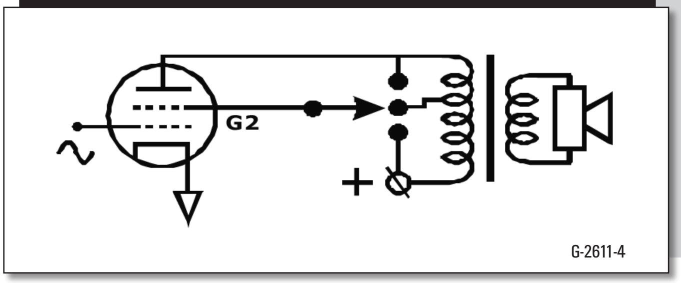
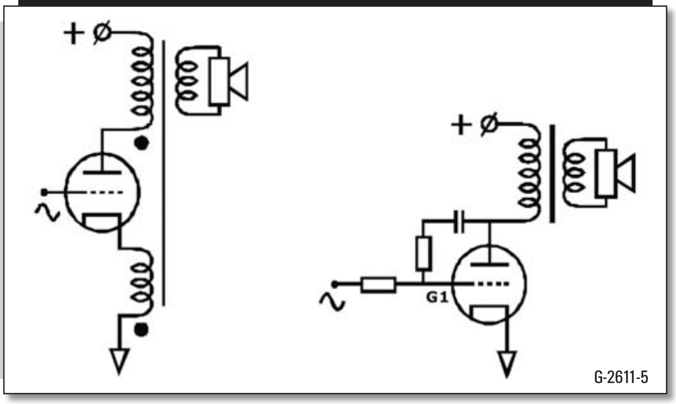


FIGURE 5: Local feedback with the control grid.



three new choices available.

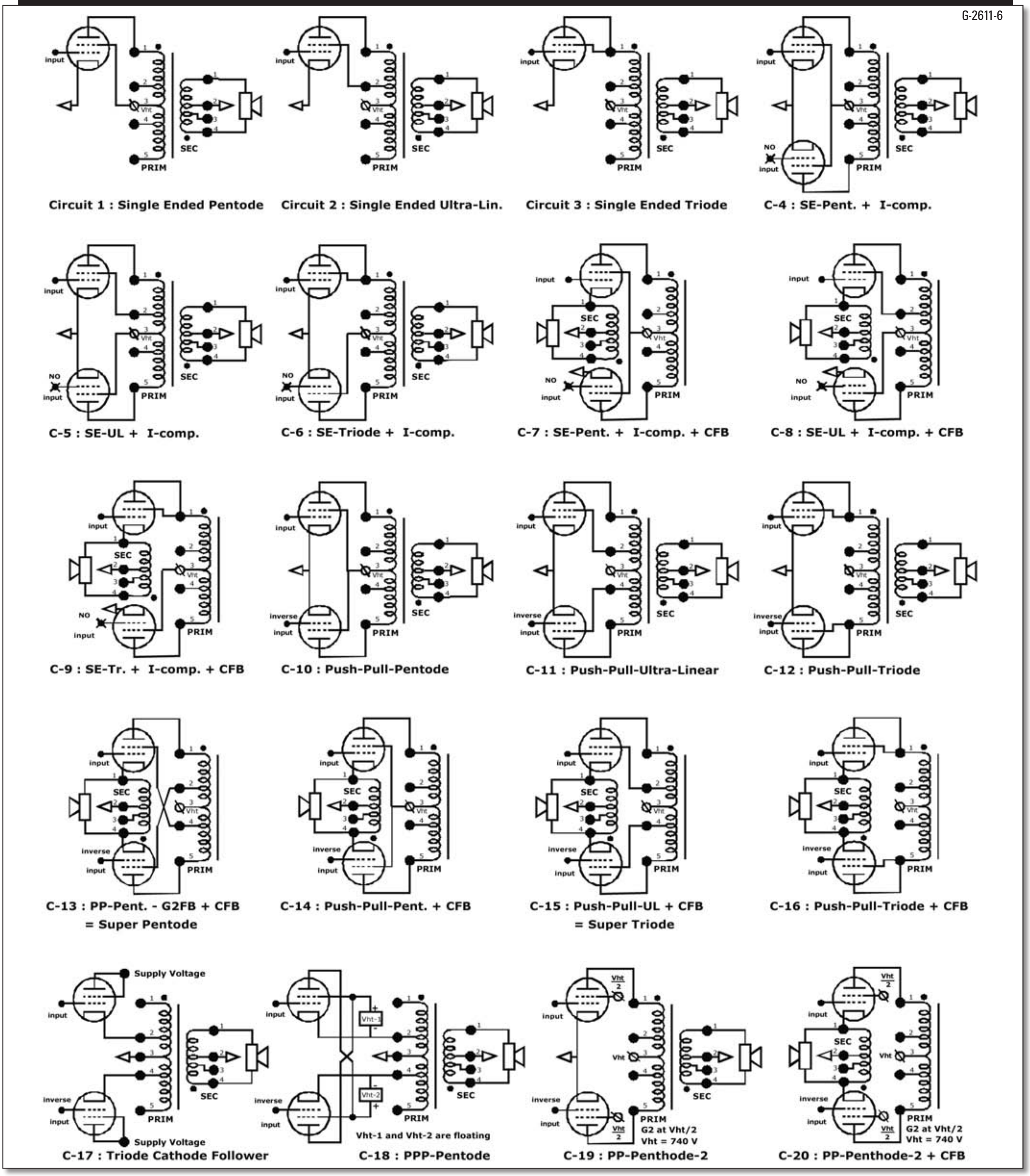
Or you can apply other sorts of local feedback: at the cathode through a separate winding at the output transformer, or at the control grid by means of local feedback with resistors. This delivers two new choices.

When you multiply the number of choices available, you get 54 for SE and 36 for PP. In summary, this approach delivers 90 different amplifier concepts, which all are based on one systematic approach of the possibilities offered by tubes. **Figure 6** shows 20 of these 90 amplifiers. I built these 20 amps to check the validity of the system, and I will tell more about the results of these tests.

SPECIFICATIONS OF THE NEW OPT

I have shown that a general approach delivers a universal system for all tube

FIGURE 6: Twenty amplifiers resulting from the general system.



amplifiers. The next question is: Is it possible to derive the specifications of a universal OPT from this system? I approach this question from the practical side by investigating common features of today's musical instrument amplifiers.

One of the most common tubes in the guitar world is the EL34, with its American equivalent 6CA7 (which, in fact, differs from the EL34 because it is a beam power tetrode, while the EL34 is a real pentode). Valves such as the 6L6 and 6550 are also often seen in the guitar world. The high-end scene adds the KT66 and KT88 to these choices, but for now, let's just consider guitar applications.

In most amps the high voltage supply is somewhere between 300 and 500V, and the power to be delivered by two tubes in push-pull should be around 40W. Many guitar amplifiers are over-specified in regard to power. Very often the peak power is mentioned, which is a factor 1.5 times the continuous output power, depending on the regulation of the power supply transformer. So, 60W in guitar world can be equal to 40W continuous power. But who is able to hear on stage the difference between 40 and 60W (= 1.8dB)? All these conditions lead to a primary impedance (for push-pull) around 4kΩ with a power capability of 40W.

When you consider high voltage supplies around 700 to 900V, then a primary impedance of 8kΩ is optimal. For instance, with 2 × EL34 at 740V, you can expect a continuous output power around 70-80W.

The maximum output power of an OPT is directly related to its lowest full-power frequency. The lowest guitar tone is at 90Hz, and you can expect to hear some resonating of the guitar body. Then the OPT should be able to deliver *80W power at 8kΩ primary at 40Hz* to fulfill all these demands. In a 40W amp the same core can handle this power at a frequency 40/√2 = 28Hz. Then the –3dB power bandwidth starts at 20Hz, which opens the door for high-end applications.

If you choose cathode-follower configurations in push-pull topology, then *the primary impedance should range from 500 to 1000*Ω.

Suppose you wish to apply this push-pull transformer to single-ended applications. Then you should gap the core to compensate for the DC-magnetization of the core. Gapping the core means that the primary inductance Lp is much smaller than with a non-gapped core. Consequently, the exciting current is much larger in the gapped core (Iex = V / (2.pi.f.Lp). This explains why SE amplifiers often have difficulties in reproducing low frequencies.

I decided to use a gapless core, and to employ the possibilities of DC current compensation by means of a current source. Such a source is not so difficult to construct with modern semiconductor design, and it removes the huge disadvantage of large, useless exciting currents (which, in fact, exhaust the single-ended tube at low frequencies). Consequently, I can apply the standard push-pull output transformer for SE applications. I only need to offer a reasonably wide range of SE primary impedances between 1000 and 4000Ω in practical situations.

When you add to this list of demands a –3dB frequency range of at least up to 20kHz, you should carefully model leakage inductance and internal primary capacitance, which puts a demand on the way the windings are placed on the core.

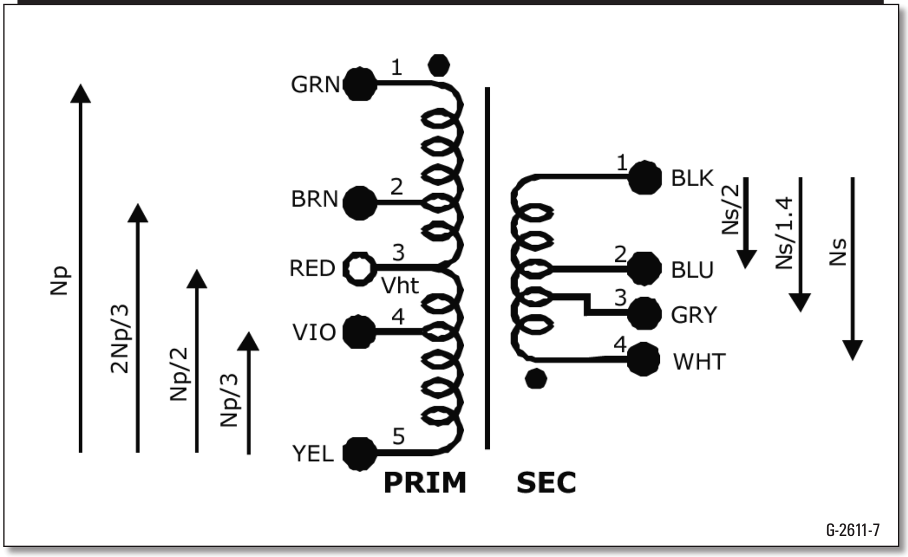
The last important demand is that the insertion losses in the magnet wire resistances should be less than 0.2dB, because in guitar amps the OPT is often used in overdrive, thus preventing the OPT to end its life in flames.

Is it possible to meet all these demands in one transformer? Yes, else I would not have written this.

THE ACTUAL UNIVERSAL OPT

Figure 7 shows the actual universal output transformer. The total number of primary turns equals Np.

FIGURE 7: The universal output transformer.



The primary is tapped in the center (power supply connection), and there are two ultralinear taps (at 1/3 and 2/3 times Np), which can be used for 33% ultralinear feedback or for cathode-follower applications. The total number of secondary turns equals Ns, and the secondary is tapped at 1/2 and 1/√2 times Ns.

The final specs of the universal transformer are: Np/Ns = 22.4; Lp, max = 900H; Rip = 210Ω (total primary winding); Ris = 0.13Ω (1-2 = Ns/2); Ris = 0.23Ω (1-3 = Ns/1.4) and Ris = 0.47Ω (1-4 = Ns); primary leakage inductance Lsp = 20mH; primary effective capacitance Cip = 800pF; power capability of the core is 80W at 40Hz at the onset of core saturation.

When you consider push-pull applications of this transformer, refer to **Table 1** for the range of impedances that can be handled. **Table 2** shows the single-ended application between taps 1 and 3 (with the current compensation on tap 5), while **Table 3** shows the impedances for cathode-follower application (cathodes connected to the ultralinear taps

2 and 4). These tables are based on the fact that impedances from secondary to primary transform by $(Np/Ns)^2$ and vice versa.

Knowing that most loudspeakers have 4 or 8Ω impedances, you can select from these tables the secondary tapping to adapt the optimal value to the application. Because the impedance of a loudspeaker is frequency dependent, you should select a tap with an impedance closest to the nominal impedance of the loudspeaker and not worry whether or not the match is 100% perfect. A match between 70 to 130% is fine in real applications. Personally, I always select the secondary tap impedance to be somewhat smaller than the nominal speaker impedance for lowest harmonic distortion.

TABLE 1 PUSH-PULL APPLICATION

Zaa 1->5	Zs 1->2	Zs 1->3	Zs 1->4
2000	1	2	4
3000	1.5	3	6
4000	2	4	8
5000	2.5	5	10
6000	3	6	12
7000	3.5	7	14
8000	4	8	16
[Ohm]	[Ohm]	[Ohm]	[Ohm]

TABLE 2 SINGLE-ENDED APPLICATION

Za 1->3	Zs 1->2	Zs 1->3	Zs 1->4
1000	2	4	8
1500	3	6	12
2000	4	8	16
2500	5	10	20
3000	6	12	24
3500	7	14	28
4000	8	16	32
[Ohm]	[Ohm]	[Ohm]	[Ohm]

TABLE 3 CATHODE-FOLLOWER APPLICATION

Zcc 2->4	Zs 1->2	Zs 1->3	Zs 1->4
500	2.3	4.5	9
600	2.7	5.4	10.8
700	3.2	6.3	12.6
800	3.6	7.2	14.4
900	4.1	8.1	16.2
1000	4.5	9	18
1100	5	10	20
[Ohm]	[Ohm]	[Ohm]	[Ohm]

EXPERIMENTAL SETUP

As I previously mentioned, I tested the 20 amplifier topologies as shown in **Fig. 6**. In circuits 1-17 I used EL34 power tubes at 370V high voltage supply with a quiescent current of 50mA per tube. In circuit 18 I used two separate supplies of 370V. In circuits 19 and 20 I added the two supplies to 370 + 370 = 740V while the screen grids were at 370V.

In circuits 19 and 20 the quiescent current was set to 25mA per power tube. The power supply is not shown, but this supply contains two fully separated 270V windings, rectified and buffered by 330µF, delivering 370V DC under quiescent conditions and 340V DC under full power operation. The supply also delivered Vn = –55V for the trimming of the quiescent current of the power tubes and for the "current source" R6 of the ECC81 driver and phase splitter tube. See **Fig. 8** for the "audio part" of the universal amp.

An example of the interconnections between the power tubes and the OPT is given in **Fig. 9**. It concerns circuit 20, which explains how amplifier and OPT can be systematically connected to each other.

To change the amplifier topology, you need to rearrange only some interconnections, and with **Figs. 6** and **8** combined it will be clear how to do so.

I tested all these amps on the following issues: maximum power (Pmax) at 1kHz at the onset of clipping (to check the primary load line), the lowest –3dB frequency (f-3L) at the loaded output at 8V peak to peak level (to check the primary inductance Lp), the same for the highest –3dB frequency (f-3H) (to check Lsp and Cip), the output impedance (Zout) at the speaker terminals using a 4Ω dummy load, and the gain from input to loaded output. With Zout and gain, I was able to determine the amount of local feedback in each circuit.

In circuits 1-9 I connected a 4Ω load between secondary taps 1-2 in order to produce a primary impedance Za = 2kΩ.

In circuits 10-16 I connected a 4Ω load between secondary taps 1-3 to produce a primary impedance Zaa = 4kΩ.

In circuits 17-18 I connected a 4Ω load between secondary taps 1-2 and loaded the primary between the UL-taps 2-4 in order to produce a primary impedance of Zcc = 900Ω.

In circuits 19-20 I connected a 4Ω load between the secondary taps 1-2 in order to produce a primary impedance Zaa = 8kΩ.

Subjectively, I evaluated the amps in stereo, with a high-quality preamp plus CD-player and loudspeakers and interlinks. With a restricted number of tracks of some very good CDs, I was able to get impressions about total harmonic distortion, speaker damping (Zout), dynamics (Pmax), bass response (f-3L), high-frequency range (f-3H), and preference of each amp for high-end applications. After these stereo tests I evaluated the amplifiers with intensive guitar sounds. Some of the subjective findings are mentioned in remarks in **Table 4** and in the evaluation following.

RESULTS AND DISCUSSION

Table 4 shows the results of my many experiments. Changing from one amp to another was very easy with this concept. It was just a matter of soldering new wires to the different taps, and a new amplifier was born.

Circuits 1-3: there is no DC-compensation. The core is saturated, resulting in a small Lp value and consequently f-3L is high, also depending on the effective plate resistance of the driving power tube. The sound character is soft and not loud, with much 2nd harmonic distortion—too much to produce an open soundstage.

Circuits 4-6: now the Lp value is large again. There is no core saturation, and the bass is strong. The amount of power is not large yet because the lower current compensating EL34 eats power. Its plate impedance is not infinity. However, these amps sound loud and silky.

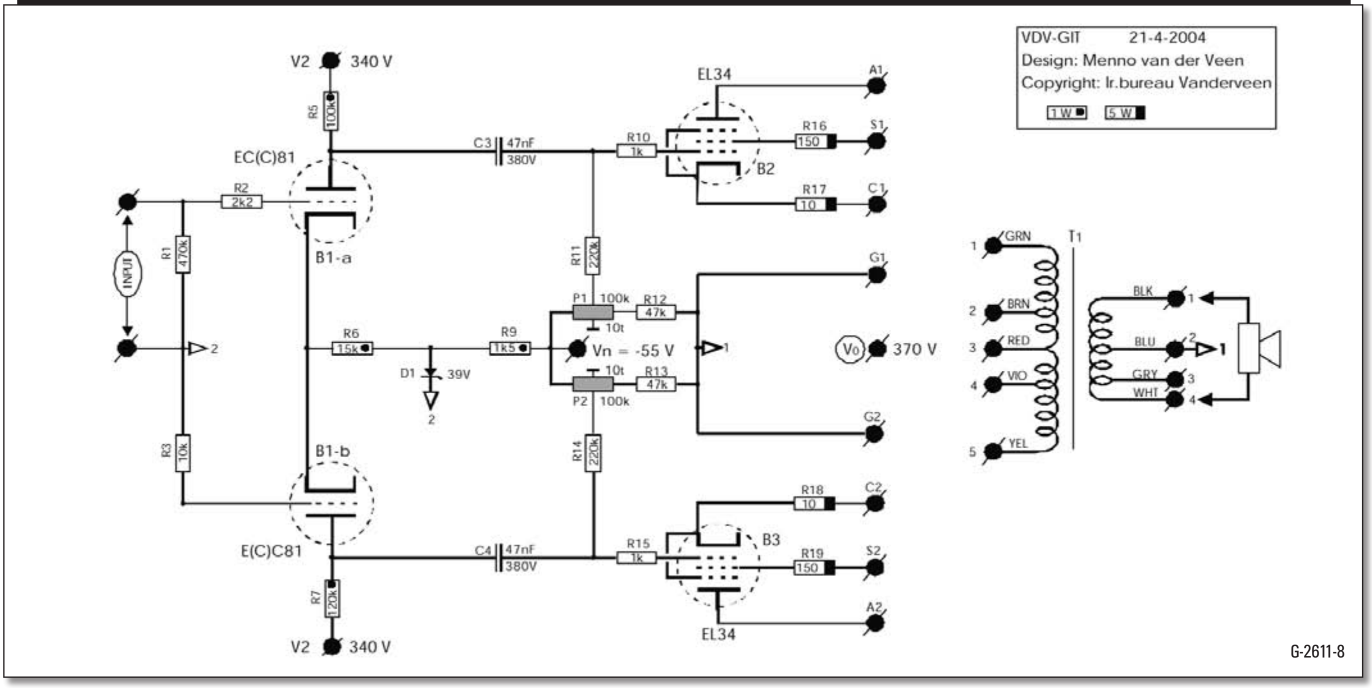
Circuits 7-9: with cathode feedback the output impedance is extra small, with good bass control, but the sound character becomes harsh because the clipping behavior is less soft.

Circuits 4-9: an EL34 in pentode mode is used as a current supply. This circuit is not ideal. Two things are needed for improvement. First, remove any ripple voltage at Vo; otherwise, there is too much hum at the speaker because the current source and amplifying tube have different plate resistances. Consequently, the ripple voltage does not cancel out at the balanced primary winding.

Second, add a cathode resistor to raise the effective plate resistance of the current source (or any other clever trick). This will prevent the current source from eating power. Also a 50mA quiescent current is not the optimal choice, but was kept constant due to good comparison between the different amps. I leave it to the reader to improve this further.

Circuits 1-9: In all these amps the primary impedance was set at 2kΩ. There was no research for the optimal load line or quiescent current setting. Consequently, there is too much total harmonic distortion. This makes the sound character friendly and silky and soft, but not open or detailed. However, each SE builder

FIGURE 8: The universal amplifier setup.

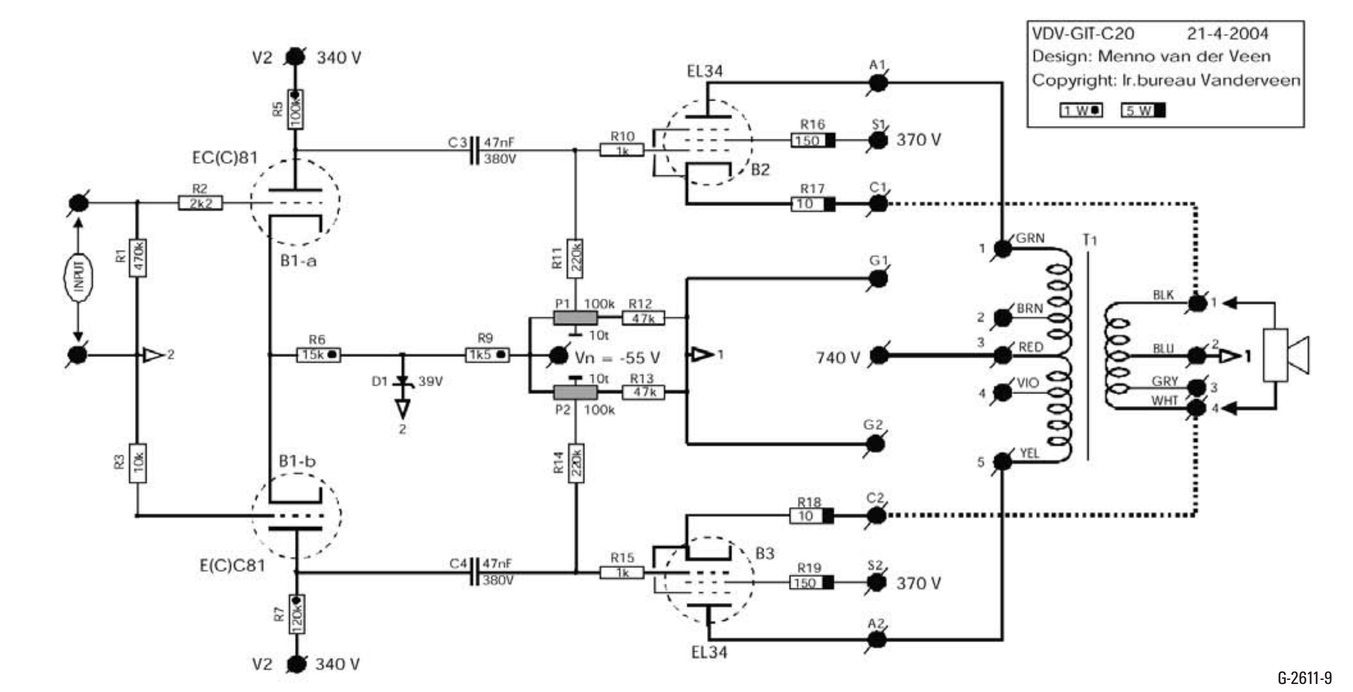


knows how to optimize these SE circuits, which was not the goal for this research.

Circuits 10-12: the sky opens, a curtain is removed, there is power and dynamics and little distortion. Triode circuit 12 is especially well-detailed.

Circuits 13-16: the cathode feedback improves the loudspeaker damping. However, in circuit 13 the subjective result has a little (not yet

FIGURE 9: Example of connections for circuit 20.



understood) harsh character. Circuit 15 sounds superior and identical to circuit 12, but has twice the power, which is an important feature.

Circuits 17-18: In these two circuits the driver stage did not have enough headroom to get to the clipping level, resulting in the very small gain. However, these two circuits measure excellently and very clean. Because of the driver stage limitation, there was no subjective evaluation.

Circuits 19-20: these circuits show how to get as much power as possible from 2 × EL34. Circuit 19 shows a very clear transition from class A to class B. This improves largely in circuit 20 through the cathode feedback. Circuit 20 is especially dynamic.

For high-end applications the two circuits 12 and 15 are the real winners. For guitar I prefer circuit 10 and 19 for bluesy and jazzy styles, while 20 is excellent—clean and loud and open and even friendly in overdrive, making it the loudest amp for heavy on-stage applications.

TABLE 4 TEST RESULTS

amp	P@1kHz	f-3L	f-3H	Zout	Gain	Remarks
1	3.1	57	31	45	20	SE-pentode, small Lp, little bass, silky
2	4.5	35	23	20.6	13.8	SE-Ultra-Linear, small Lp, little bass, silky
3	4.5	23	18.6	3.8	8.9	SE-triode, small Lp, some bass, silky
4	5.3	8	15.7	45.2	6.7	SE-pent with current compensation, loud
5	5.3	8	15.7	18.9	5.9	SE-UL with current compensation, good bass, silky
6	5.3	7	19.8	3.3	3.9	SE-triode with current comp, deep bass and silky
7	9	8	26	4.9	4.9	SE, as 4, with CFB, hard clipping, not open
8	9	8	24	3.4	4.0	SE, as 5, with CFB, hard clipping, not open
9	6.1	8	23	2.0	3.0	SE, as 6, with CFB, hard clipping, not open
10	29	8	47	53	13.3	PP-pentode, clean, loud, dynamic
11	28	8	26.4	9.3	9.5	PP-UL, clean, loud, dynamic
12	14	8	24	3.4	6.2	PP-triode, very clean, superior
13	26	8	56	12.8	11.1	PP-Super Pent, powerful, somewhat harsh character
14	30	8	50.8	6	8	PP-Pent., CFB, powerful, not harsh
15	29	8	42	3.8	6.8	PP-Super Triode, powerful, extreme clean and open
16	13.4	8	28	2.4	4.9	PP-Triode with CFB, less power but cleanest of all
17	8	8	59	1.93	1.5	Triode Cathode follower, driver stage is limiting
18	11	7	68	1.33	0.2	PPP-pentode Cathode follower, driver stage is limiting
19	78 (91-peak)	8	20	45	12	PP-Pentode VO = 737V, loud, not very open
20	72 (91-peak)	8	38	4.2	7.1	PP-Pentode VO = 737V, with CFB, dynamic loud, open
[-]	[Watt]	[Hz]	[kHz]	[Ohm]	[-]	

SUMMARY

This systematic approach to vacuum tube amplifiers shows that a system can be defined which includes all possible variations in topologies. The system contains 90 variations (at least), and some of them have never been constructed until now. You can use this new universal output transformer for all the amplifiers of the system, especially for musical instrument (guitar) amplification with valve amplifiers.

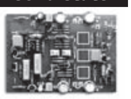
Tests with this output transformer in 20 amplifiers of the system show the validity of this approach, making it extremely easy to change from one amplifier topology to another, simply by rearranging some wires. This approach is a very valuable tool, which enables me to quickly and clearly gain insight in the character of different amplifier topologies, without having to deal with tremendous costs of a new output transformer for every new amplifier. aX

REFERENCES

1. Menno van der Veen, "Theory and Practice of Wide Bandwidth Toroidal Output Transformers," 97th AES convention 1994, preprint 3887.
2. Menno van der Veen, "Modeling Power Tubes and Their Interaction with Output Transformers," 104th AES convention 1998, preprint 4643.
3. Pierre Touzelet, Menno van der Veen, "Small signal analysis for generalized push-pull tube amplifier topology," 112th AES convention 2002, convention paper 5587.
4. Menno van der Veen, Pierre Touzelet, "New Vacuum Tube and Output Transformer Models Applied to the Quad II valve amplifier," 114th AES convention 2003, convention paper 5748.
5. Menno van der Veen, "Modern High-End Valve Amplifiers Based on Toroidal Output Transformers," *Elektor Electronics*, ISBN 0-905705-63-7.
6. Menno van der Veen, "Measuring Output Transformer Performance," *Glass Audio*, 5/97, p. 20.
7. Menno van der Veen, "Lab Report Specialist Range Toroidal Output Transformers," www.plitron.com and www.amplimo.nl for the Dutch language version; see also (5) chapter 8: "Special Output Coupling Techniques."
8. Menno van der Veen: For new developments in this research, visit: www.mennovanderveen.nl

This article was originally published as a paper at the 118th AES convention, May 28-31, 2005, Barcelona, number 6347.

Menno van der Veen *has a degree in physics with a specialization in electronics. He is a professor of physics at a teacher training college as well as a reviewer of high-end audio for various magazines. He is a designer of high-end toroidal output transformers (Plitron, Amplimo), and is a member of the review board for the JAES. He is also director of Vanderveen bv, an audio electronic research and consultancy firm.*



The Power JFET Amplifier: Zen Variations #8

By Nelson Pass

How to achieve enlightenment and a JFET power amp using the Zen approach.

Thanks to a nice person on the Pass Labs Forum (**www.diyaudio.com**), I became aware that high current power JFET transistors are again available. You can check them out at **www.lovoltech.com**, which offers a small variety of high current N-channel vertical JFETs in TO-251 and TO-252 packages. They don't hold much voltage (20–24V) or dissipate much power (69W, derated at 1.8W/degree), but they do deliver up to 100A peak. Clearly designed as switches, they nevertheless offer a linear operating region and are attractive for audio amplifier design.

FIGURE 1: Simplified Yamaha B1 JFET amplifier.

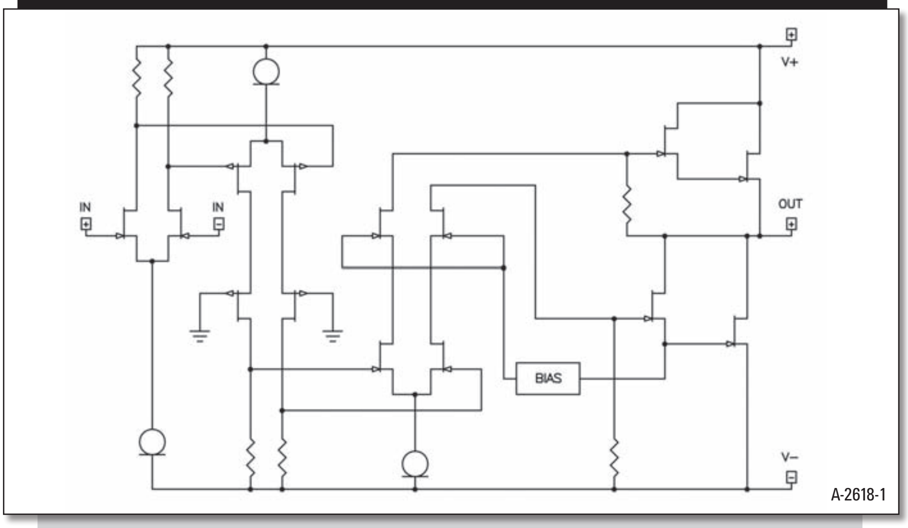
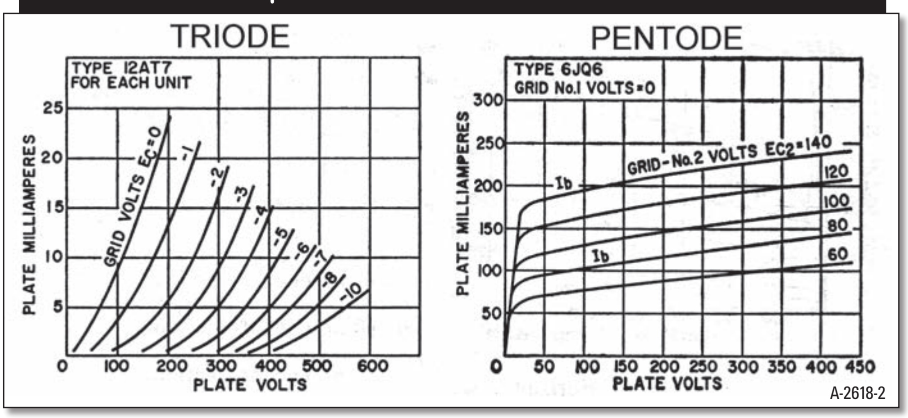


FIGURE 2: Triode and pentode character.



Small-signal JFETs are routinely used in the input stages of the finest solid-state amplifiers and preamplifiers, where they give very high input impedance, high linearity, and very low noise. JFETs are often extolled in promotional literature for their "tube-like" qualities. Their characteristic curves, like MOSFETs, largely resemble pentodes.

In the 1970s, Sony and Yamaha offered a few JFET power amplifiers, but discontinued them for unknown reasons. **Figure 1** shows a simplified schematic of the Yamaha B1, which remains a highly regarded product.

Why power JFETs, you ask? Don't we have enough transistors already? We do have lots of choices in devices, but a vertical JFET brings some particular characteristics to the table that you don't get elsewhere. When you look at characteristic curves for devices, you see what I call the "triode character" and the "pentode character," as shown in **Fig. 2**.

Here you see that the triode character has a partially exponential curve of current versus voltage across the device, while the pentode has a more logarithmic shape. You could think of the triode as "concave" and the pentode as "convex," and by comparison, a resistor would have a straight line.

Since no gain device does such a good job of approximating a resistor, triode designers linearize the tube circuit by operating the voltage/current load line to get some cancellation between competing voltage and current distortion. This can be done with either single-ended or push-pull circuits.

The guys with pentode type devices, which includes solid-state as well as non-triode tubes, take a different approach, which is to try to work them in the more voltage-invariant area and use other methods to correct for remaining nonlinearities, usually through negative feedback.

I'm not here to argue the merits of one approach over another, but as I look at the characteristic lines for these new power JFETs, I notice that there is a narrow region at the bottom where there is less convex curvature. In fact, it looks concave, *kind of like a triode*.

With Lovoltech part LU1014, note that with a gate voltage of –1V, the curve is concave below about 5A and 4V. In this range it has that triode character, and this is the area of interest to us here.

There are some other potential advantages to these JFETs. For a given bias current, they are more linear than comparable MOSFETs, and while they have comparable capacitances, the values are more linear over the operating voltages. Depletion mode is convenient, as it offers self-biasing with no particular circuitry to set the bias point, and the devices have a small negative temperature coefficient, so they hold a stable bias point easily. Last but not least, JFETs are not sensitive to static discharge.

THE SINGLE APPROACH

If you have a device that behaves a bit like a triode, then it is natural to try it out in a popular triode amplifier topology. For many aficionados, that topology would be single-ended Class A operation. 300Bs and 211s operated single-ended and coupled to an output transformer have been held by many as a low power musical standard. Single-ended Class A tube amplifiers are not very powerful, and their measurements are nothing to write home about,

FIGURE 3: LU1014.

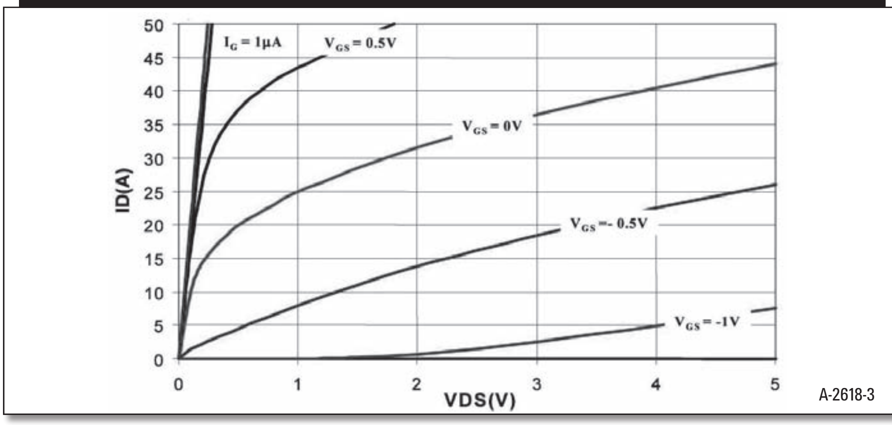
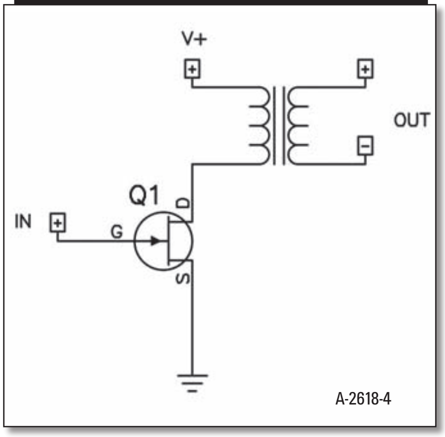


FIGURE 4: Plug it in and see what happens.



but there is no denying that they have strong musical appeal to a sizable segment of the audiophile population.

Well, it seems easy enough—just pop one of these self-biased little buggers in series with a transformer, as shown in **Fig. 4**. The gain and input impedance of the transistor are such that you won't even need more than this single stage—much easier than the three gain stages of your better tube amps, and no fussing around with filament supplies, either.

THE PROBLEMS

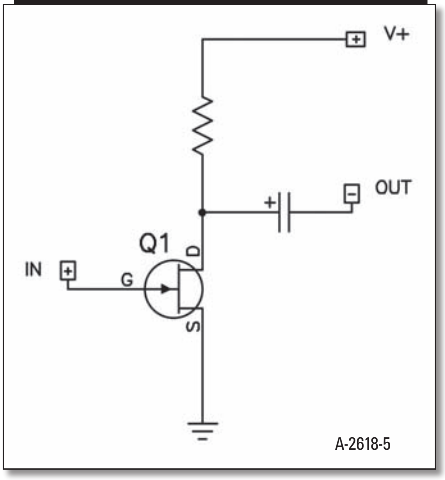
There are some reasons the circuit of **Fig. 4** might not be appropriate.

First, as shown, the JFET will draw a tremendous amount of current, and quickly take up smoking. You can fix that with some resistance on the source to ground. Looking at the curves, maybe an ohm or so.

While you're fixing things, perhaps you can do without the transformer at all. In tube circuits the transformer converts the high voltage/low current of the tube's terrain to the relatively low voltage/high current of the speaker. Again looking at the JFET's curves, you see that perhaps it isn't necessary, and you can replace the transformer with a resistor and output capacitor, as in **Fig. 5**. If you think that a capacitor is not an improvement over a transformer, then you're welcome to your opinion, and in fact the transformer will raise the efficiency of the amplifier, but we are going to leave that for another time.

As you slowly crank up the supply voltage, you see that this certainly works, but the distortion is kind of

FIGURE 5: This one worked for a minute.



high, and when you raise the voltage enough to get a good listen, you notice that the high end is a little deficient. After a few exciting seconds, the amplifier makes a popping noise, and then:

Nothing.

OK, you're making progress anyway. Our problems (hereafter referred to as opportunities) are fairly simple. First, the transistor in question has limited dissipation. Even though it's rated at 69W, realistically I'd be afraid to operate the package at more than about 10W or so. A TO-251 is a small device, and you must work to get the heat away from it.

Second, the transistor is rated at only 24V. That's just about enough to make a 2W amplifier with this circuit, and under these circumstances the transistor will be dissipating something approaching 16W. You might get away with this, but you still have pretty high distortion, and the top end is still down.

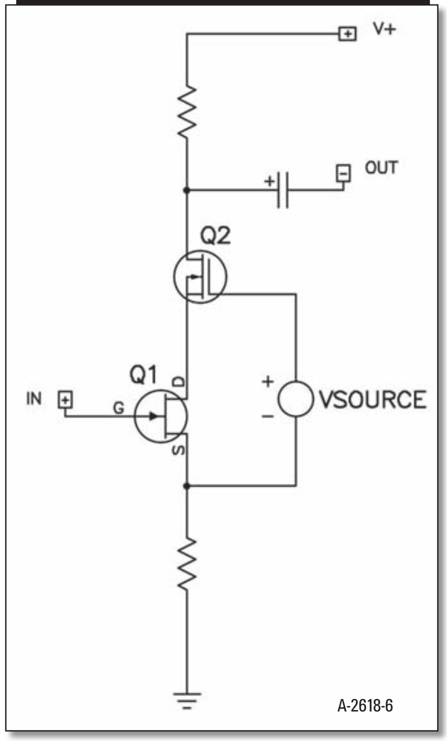
You can fix the top end—just get a preamp with a low output impedance. The 400pF capacitance between the gate and the drain of the JFET is being amplified by the gain of the amplifier, and looks like more than 3200pF. What? You say your preamp has a 10k output impedance? We can fix that.

Even with a low impedance preamp, the distortion is still high. Some opportunities just seem insurmountable.

THE SOLUTION

There is a single solution to these problems—cascode operation. It is easy to cascode the drain of the JFET with a partner device, as illustrated in **Fig. 6**. The partner device, operated common gate, contributes so little of its

FIGURE 6: Simplified final circuit.



own character to the output that it doesn't rate as a gain stage per se, but it shields the actual gain device from the output voltage. The gain variation due to fluctuations of drain-source voltage (the music) has nearly vanished, as has the need to charge and discharge the capacitance from drain to gate.

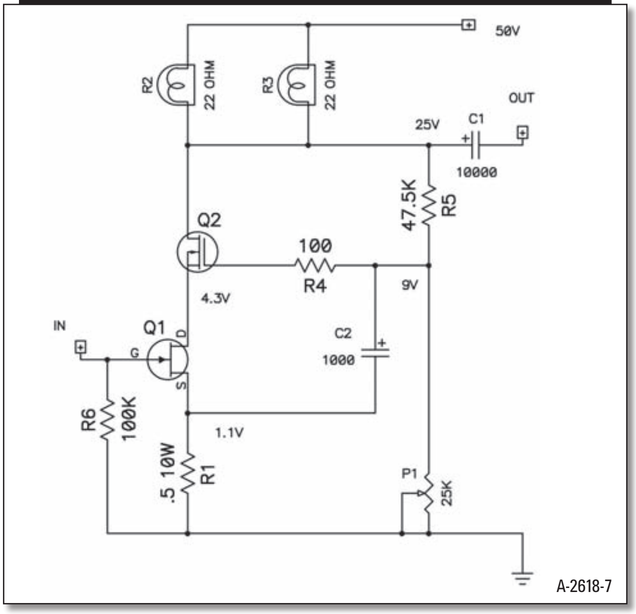
Even better, the drain to source voltage is not only nearly constant, but it is low, as low as you might want in order to keep the transistor reasonably cool. The power MOSFET doesn't have anything to say about the output signal, but it must take all the heat. Now you can crank this circuit up.

Figure 7 shows the Full Monty.

The current source resistor of the simplified circuit could be an ordinary power resistor, cooking along at 40W, or you can get a pair of 300W (120V) light bulbs, which are cheap and don't need heatsinks. I like the clear ones because they look swell.

Just about all you want to know about this circuit is in the schematic, as I have

FIGURE 7: Amplifier without feedback.



posted the relevant voltages. There is wiggle room with regard to supply voltage and current, but you want to watch Q1—its dissipation is the voltage across it and the current draw of the circuit, which as shown is about 3.2V and 2.2A, equaling 7W. The voltage and current are functions of the part itself, but you can adjust the voltage with P1. As you increase the value of P1, the voltage across the JFET increases, and the current will increase with it.

Construction of the amplifier is as straightforward as any other amplifier. The power supply doesn't require regulation, but it does require filtering to avoid supply noise from bleeding into the output. It is also very helpful if the supply comes up slowly on turn-on. You can obtain both these qualities with a capacitance multiplier as outlined in previous Zen articles. Q2 is mounted on a heatsink like any other power device, but we pay special attention to the mounting of Q1.

I mounted mine directly on an anodized sink, using thermal grease and about a pound of pressure from a clip. I relied on the anodizing to insulate the case of Q1 from the heatsink. If you do the same, you will want to be certain that it is insulated, otherwise I recommend mica and grease.

Of course, you will want to put in audiophile-approved bypass capacitors in parallel with C1, but I leave you to choose the part.

Before you fire up the amplifier, check everything twice, (once in the evening, once in the morning) and set P1 to the minimum value. It should turn on drawing no current. Monitor the voltage, the drain and source of the JFET, and the drain of Q2 to ground. Slowly, very slowly, increase the value of P1.

After a while, the drain of Q2 should start to drop as the drain and source of Q1 start to rise. With a little luck, you'll arrive at the voltages shown. The two light bulbs in parallel should measure about 11Ω, so that the correct output voltage will give about 2.2A of current, and you'll see about 1.1V at the source of the JFET, where the current goes to ground through a 0.5Ω power resistor.

The crucial test is to get the 25V at the output while maintaining a relatively low voltage across the JFET. You want to keep the dissipation on the

PHOTO 1: Second harmonic distortion.

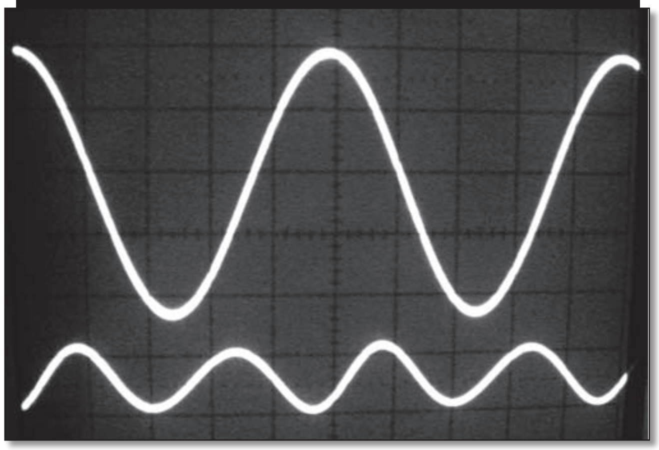


FIGURE 8: THD vs. output. No feedback.

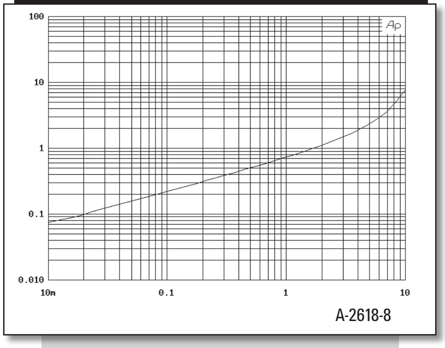
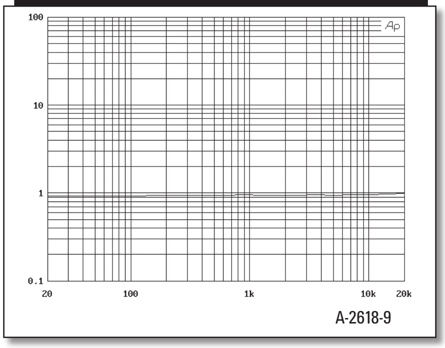


FIGURE 9: THD vs. frequency. No feedback.



JFET at less than 10W, and depending on the individual JFET, you should be able to manage that.

The distortion of the circuit versus output is shown in **Fig. 8**. It rises with the classic second harmonic character, as shown in **Photo 1**. Compared with a typical MOSFET in the same circuit, it shows about one-third the distortion.

Figure 9 shows the distortion at 1W across the audio band. It is flat and remains flat for source impedances up to several kΩ. The bandwidth with a 600Ω source impedance is down a fraction of a dB at 100kHz.

By virtue of the resistance of the light bulbs, the damping factor for **Fig. 7** is slightly less than 1. With an 8Ω load, the gain is about 18dB.

Naturally, we look for any opportunity to improve the performance of this circuit, and if you don't mind throwing away a little gain, you can consider using some negative feedback on the circuit.

FIGURE 10: Amplifier with 6dB feedback.

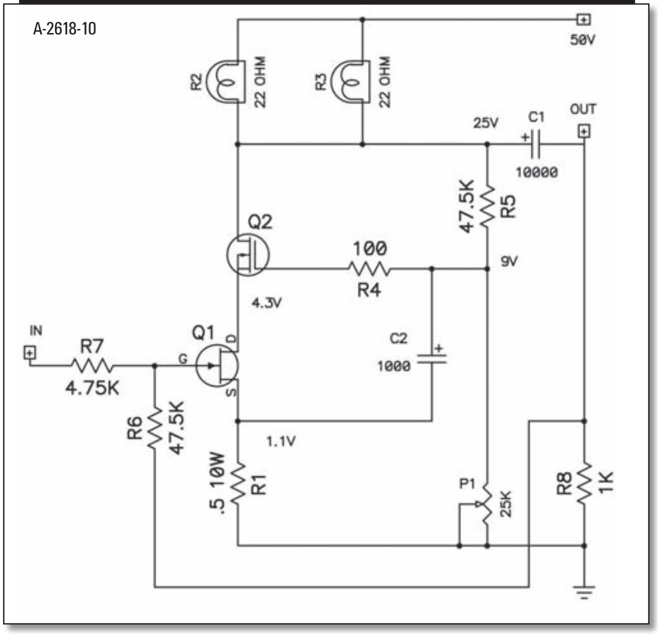


Figure 10 shows the circuit with feedback implemented. You have found a new connection for R6 and added R7 to form a feedback loop giving about 6dB of feedback. The input impedance is now about 10kΩ, and the damping factor is increased to 2. Due to the higher source impedance, the amplifier is down a bit more at 100kHz, and the distortion climbs a little bit at 20kHz, but nothing to become excited about.

In the circuit of **Fig. 7**, a resistor to ground at the input ensured that the gate of the JFET was referenced to ground, and in this circuit R8 now provides that function, giving the gate a ground reference when there is no DC source impedance.

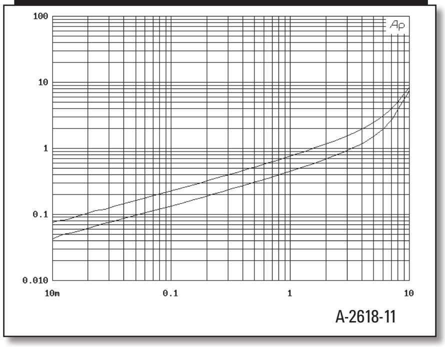
Figure 11 shows the distortion with and without feedback, and you see that the distortion is reduced by nearly half over most of the operating range. As you reach clipping, nothing can help you, not even feedback.

CONCLUSION

Well, there you have it, an actual power JFET amplifier. There are many things that you will be able to do to make it better, and other designers will come along and add lots of parts and make it increasingly complex. In the meantime, you can get your feet wet with this simple circuit.

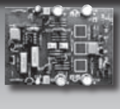
I'd like to spend some time talking about how great it sounds, but I would rather be listening to it, so I end here for the time being. *aX*

FIGURE 11: THD vs. output. With and without feedback.



ACKNOWLEDGEMENTS

Fig. 2 was adapted from the *RCA Receiving Tube Manual*, and **Fig. 3** was obtained from a Lovoltech data sheet.



A Phono PrePreamp for the (SA) CD Age, Part 3

By Norm Thagard

In the final part of this series, the author resolves some critical power-supply requirements.

This design does not offer exceptional power-supply rejection. Therefore, it requires a good power supply to realize its potential. For this reason, in addition to the usual full-wave rectifier/capacitor-input filter, there is an additional RC filter section and shunt voltage regulator for each supply. In order to physically separate the power transformer from the preamplifier circuitry, the power supply is partitioned into a raw supply contained within its own enclosure and voltage regulators located within the preamplifier enclosure. This technique is frequently used with sensitive preamplifiers.

POWER-SUPPLY SETUP

All components through the RC filter stage are located within the raw supply enclosure (**Fig. 5**). The remaining circuitry of the four voltage regulators was constructed on a single board physically located within the preamplifier enclosure. The regulator circuitry is shown in **Fig. 6**.

The two enclosures are connected with approximately 3′ of cabling suitable for conducting DC current in excess of 150mA, which is the approximate preamplifier current draw per channel from the positive power supply. The per-channel draw from the negative supply is about 50mA. Due to additional current drawn by the shunt voltage regulators, 18-gauge stranded wire is recommended for the interconnections.

The transformer is common to both channels, but from that point separate components are required for right and left channels. Only one channel is completely depicted.

The shunt regulators are based upon a design found on the Internet[25]. In reviewing back issues recently, I discovered that the design is similar to one that appeared in *TAA* a number of years ago[26]. While simple, their performance is excellent, far exceeding that possible with the standard IC regulators widely available. The published design is slightly modified by replacing the BJT shunt element with a MOSFET and sourcing collector current for Q_{101} and Q_{102} directly from the regulated bus rather than the well-filtered but nonetheless unregulated source from the raw power supply.

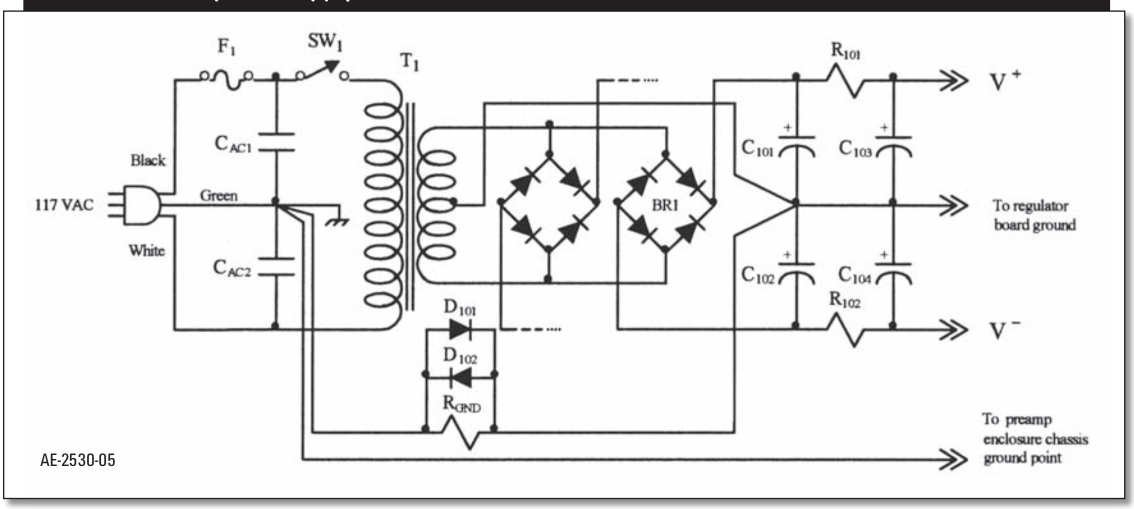
Initially, I used medium-power Darlington transistors in lieu of M_{101} and M_{102} in order to obtain sufficient voltage drop across R_{109} and R_{110} to keep those resistors at 1.8kΩ while still providing adequate collector current for Q_{101} and Q_{102}. Per a suggestion by Brad Wood, who also kindly proofed this article and offered several useful suggestions for its improvement, I replaced the Darlington transistors with MOSFETs. The gate-source voltages of the MOSFETs are even higher than the Darlington base-emitter voltage, which results in higher stage gain and consequently a greater level of negative feedback. He also recommended placement of C_{107} and C_{108} for stability reasons.

REGULATION OPERATION

I suspect that just about any medium-power MOSFETs will work. If dissipation can be limited to less than 1W, then you can substitute even low-power MOSFETs. As in the first stage of the preamplifier, I chose Q_{101} and Q_{102} for their noise performance. Substituting 2N4401 and/or 2N4403 types should give similar results.

At DC, these are essentially zener diode-based shunt regulators. For the

FIGURE 5: Raw power supply.



positive regulator:

$$V_{reg}^{+} = V_{BE-Q_{101}} + V_{Z_{101}} + V_{R_{105}}.$$

Regulation is degraded by the presence of R_{105}, but its inclusion greatly reduces the effect of zener diode noise. The value chosen is a reasonable compromise between the two antagonistic requirements. If some trimming of regulated positive or negative rail voltage is desired, it can be raised by increasing R_{105} or R_{106}, respectively, and lowered by a decrease in the values of those resistors. The values used in the prototypes yielded rail voltages of about 17.5V.

Zener currents are set, ignoring the small error due to base current, by R_{107} and R_{108} according to:

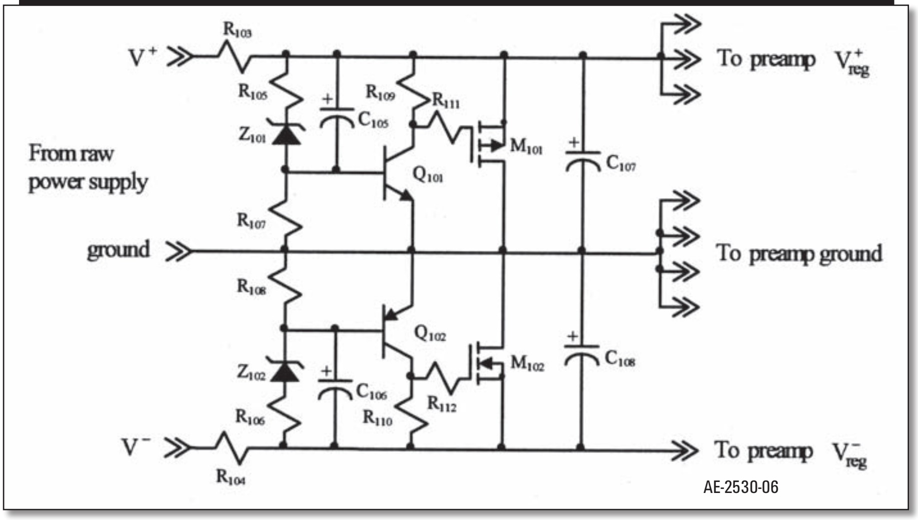
$$I_Z = \frac{V_{BE-Q_{101}}}{R_{107}} \cong \frac{V_{BE-Q_{102}}}{R_{108}} \cong \frac{0.7}{100} = 7 \text{ mA}.$$

This value was chosen because it is close to the nominal current on which 1N5246B zener diode specifications are based. For example, dynamic resistance is guaranteed to be no more than 17Ω at I_Z = 7.8mA[27].

$Q_{101}M_{101}$ and $Q_{102}M_{102}$ form feedback amplifiers with unity AC gain, thanks to coupling to the Q_{101} and Q_{102} bases through C_{105} and C_{106}, respectively. Open-loop gain is almost that of first-stage gain since the MOSFETs are source followers. First-stage gain on the positive regulator is shown in **Equation 1**.

The effective dynamic resistance

FIGURE 6: Voltage regulators.



Equation 1.

$$-g_{m-Q_{101}}R_{109} = -\left(I_{C-Q_{101}}/V_T\right)R_{109} = -\left[\left(V_{GS-M_{101}}/R_{109}\right)/V_T\right]R_{109} \cong -3.5\text{V}/25\text{mV} = -140 \text{ (43 dB)}$$

in shunt across the rail is greatly reduced by this feedback and the favorable $R_{dynamic}/(R_{dynamic} + R_{103})$ voltage divider ratio. This feedback is effective over the whole audio frequency band, reducing the effect of noise and to a lesser extent, ripple.

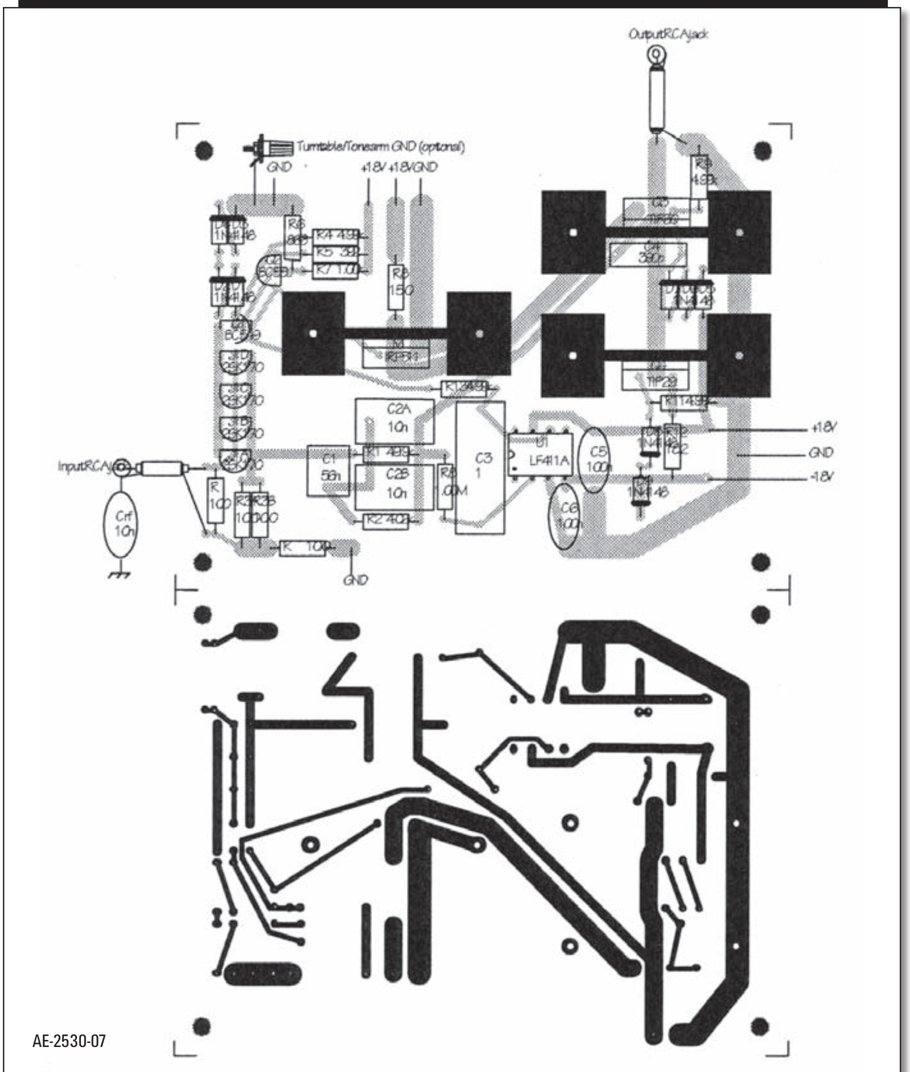
These are shunt regulators, so don't neglect series resistors R_{103} and R_{104} as I inadvertently did, forgetting the big capacitors of the RC filter back in the remotely-located raw power-supply enclosure. These resistors improve performance as their values increase. However, if they are too large, the regulator will "drop out" as voltage falls below that necessary to keep the zeners operative.

Although a power transformer with 40V CT secondary is specified, the prototype used one with 36V CT rating. While this placed the shunt regulators rather close to drop out, reducing R_{101} and R_{103} to 22Ω and R_{10} and R_{104} to 62Ω did allow proper regulator operation. Although power wastage is ecologically unsound, higher secondary voltages allow increased resistor values that improve performance.

On the other hand, lowering the values of these resistors too much can result in excessive power dissipation in the MOSFETs. I mounted the MOSFETs on small heatsinks, just in case. Fortunately, the source-follower topology places the drains at ground, so no insulation between transistor and heatsink is required even if, as was the case for the prototype, the heatsinks are grounded.

It's not a bad idea for a Variac® to be available to monitor regulator voltage while running the line voltage up to 110V RMS. The regulators should

FIGURE 7: Preamplifier board parts placement and pattern.



"kick in" below this voltage to guarantee regulation even in periods of line voltage drops. If the regulators kick in at voltages substantially less than, say, 100V RMS, then perhaps some thought should be given to increasing the values of R_{101} – R_{104}.

STAR GROUNDING

The pattern of interconnections from the regulator board to the preamp boards is important. Improper connections, which often can work under less demanding conditions, may wreak havoc with high-gain audio circuits such as this one. The original layout used a common ground bus. There was an intermittent motorboating that was extraordinarily dependent upon physical location of the preamp enclosure. Breaking this bus and supplying four separate ground interconnections seemed to eliminate this problem.

Perhaps I should have paid more attention to the warning in one reference that read ". . . or, in audio circuits, (daisy-chaining of ground returns can cause) low-frequency "motorboating" oscillation . . . The extra investment in wiring is just about always worth it."[28] At a minimum, ground returns should be separate for the first and second stages. For what it's worth, the use of R_{iso} seemed to make the preamp less sensitive to grounding variations.

The same reference also notes that "The rule also applies to the power rail feed as well as to its return, and, in fact, to any connection where current is being shared between several circuits[29]." For this reason, three separate interconnections were used for the positive rail. The negative rail required only a single connection to the regulator board because that power is only supplied to the op amp and the output (second) preamp stage.

I have read that a ground "mecca" typical of the star grounding technique is just a crutch whose necessity is readily obviated if proper attention is paid to the problem[30]. I am sure that application of the technique can be overzealous, but use of separate power and return leads is sometimes essential. Anything that reduces IR voltage drops on the buses will help. Keeping wiring inductance low also helps. Broad buses and large diameter wires reduce *IR* drops and, to some extent, inductance. The best technique for reducing inductance is to shorten traces and wires.

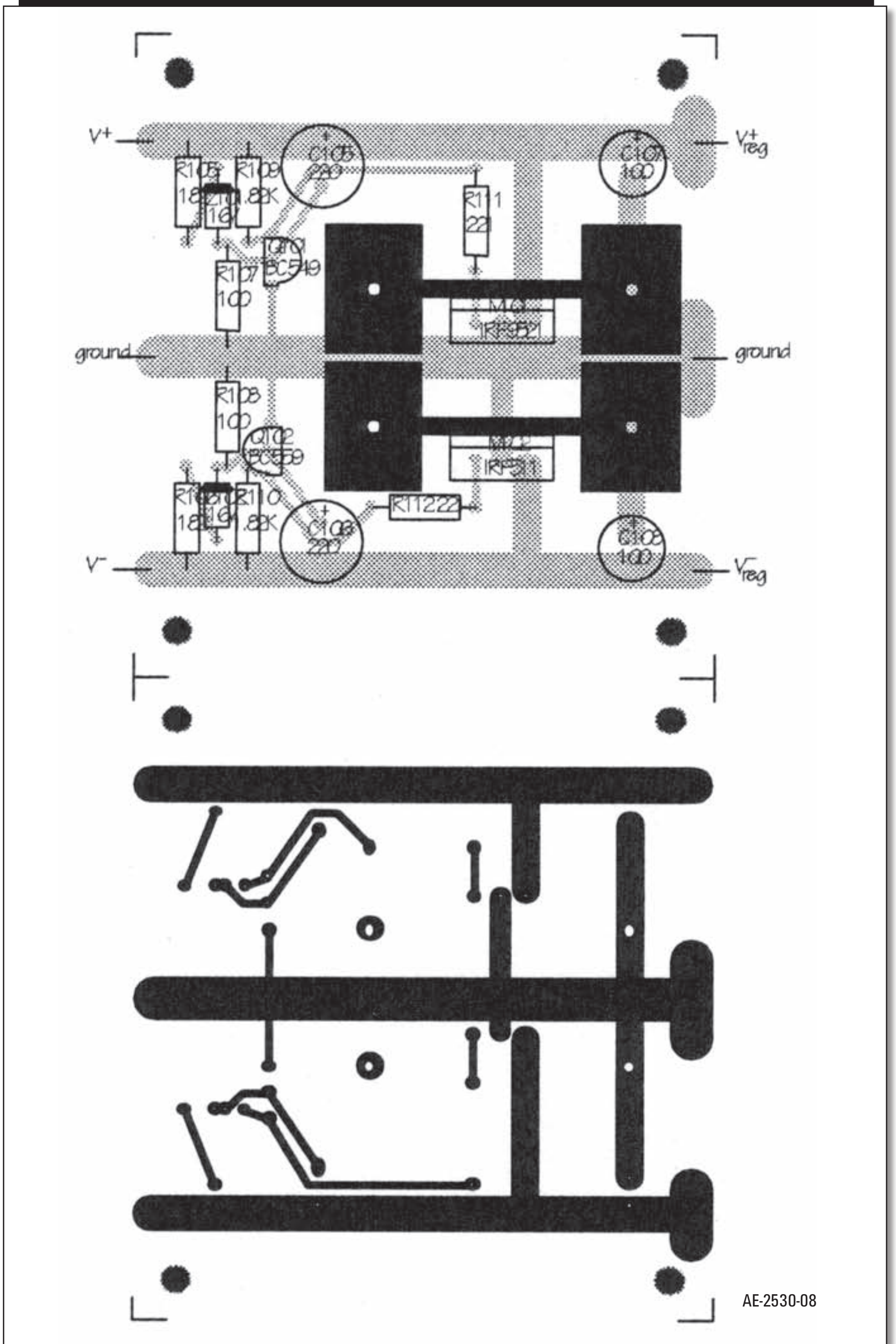
In the breadboard step of the design process, the input stage was initially supplied through an additional zener-based shunt regulator. This is a good way to avoid interstage coupling via the power supply if you attend to the zener noise. A simple additional RC filter between the output stage supply and the first stage power bus may be similarly efficacious.

Despite this discussion, the actual culprit proved to be capacitive coupling to the very sensitive circuitry with its low-frequency gain of 10,000. Grounding the preamp enclosure to the power-supply enclosure and hence the power-line ground did not fully protect against this coupling. However, connecting one of the right-channel PC-board grounds to preamp enclosure ground at the binding post as shown in **Fig. 7** completely eliminated the low-frequency oscillations. The binding post was grounded to the preamplifier chassis to allow for an optional grounding connection from the tonearm or turntable. With this ground connection in place, the power-supply enclosure should *not* then be connected to the preamp enclosure, as this will result in a ground loop and audible hum.

A simpler power-supply and ground interconnection scheme between regulator board and preamp boards might, in fact, work. However, after the modification, I was unwilling to return to the original common bus configuration.

Please refer to **Figs. 7** and **8** for the details of the interconnections between preamp and regulator boards. The prototype circuitry was placed on prototyping boards rather than the PC board patterns shown. I found no errors in proofing the board designs, and the PC patterns generally follow the layout of the prototyping boards. Even so, you should confirm that there are no errors in the patterns as published,

FIGURE 8: Regulator board parts placement and pattern.



because an important verification step is missing.

Although I originally employed a grounding scheme identical to that of my previous phono preamp, there was an audible hum unless a three-wire-to-two-wire AC plug adapter was put in place. Isolating the circuit ground from chassis and power line grounds with 10Ω resistor R_{GND} eliminated the hum and the need for the adapter by reducing ground loop current. I have used this technique to eliminate ground loops at the inputs to my amplifier designs, and some commercial preamps use this scheme as well. The resistor is subject to failure due to potential differences between circuit and power-line ground. Back-to-back diodes D_{101} and D_{102} protect R_{GND} and provide mutual self-protection against voltages that would exceed their peak inverse voltage (p.i.v.) rating by clamping resistor voltage to one diode drop.

Be especially careful that the RCA input and output phono jacks are isolated from the chassis. I found that the RCA input jack of the right channel was contacting the chassis. Although initial continuity checks indicated no problems, after a number of insertions and removals of interconnects into the jack, the motion caused a metal washer on the jack to break through a mica insulator and contact the enclosure at the mounting site. The result was a large hum caused by the ground-loop that was produced.

Do not despair, however, because I overcame these problems. If you follow the bus suggestions and use components and values specified in the parts list, you should have no problems with the power supplies.

MEASUREMENTS

The preamplifier is objectively a reasonable performer. I've previously discussed noise measurement results. THD is just 0.004% at 1kHz, rising to 0.014% at 20kHz. Both figures are taken at 10V RMS output into a 10kΩ resistive load. THD at 20Hz was less than 0.04%. Higher distortion at 20Hz was not unexpected due to the lower feedback at low frequencies.

In all cases, distortion did not increase appreciably until output voltage increased above 12V RMS. Above this voltage, noticeable "graceful" clipping became evident. Asymmetrical overload signals will upset the DC servo, but it is highly unlikely that you will encounter such inputs unless the MC version is used with a very high output MM cartridge.

RIAA tracking errors were no more than 0.1dB. I did not use a reverse RIAA network; rather, I measured output at 24 discrete frequencies from 20Hz to 20kHz[31]. The technique was to begin at 20Hz with the input amplitude that produced 10V RMS output and then measure output at various frequencies up to 1kHz with the same input amplitude. Because my distortion analyzer does not work with inputs less than 100mV, I then input a 1kHz signal whose amplitude yielded 10V RMS output and repeated the process at various frequencies up to 20kHz.

This is time-consuming and does not measure errors at all frequencies. It has the advantage of avoiding measurement error due to tracking errors in the reverse RIAA network, itself. It would be unlikely, given the close spacing of the measurement frequencies, to have tracking errors significantly greater than would those seen at the discrete measurement points.

CONCLUSION

There is not much left to explain. It was necessary to add C_4 to suppress a high-frequency oscillation. The value chosen was empirically determined as optimal from the standpoint of providing the necessary compensation while having immeasurable effect at the upper end of the 20–20kHz audio band.

Capacitors C_5 and C_6 are disc ceramic capacitors designed to suppress potential oscillation in the LF411 op amps. These were placed per National Semiconductor's recommendation[32]. In parallel with the ceramics, some would add small tantalum capacitors in the range 2.2–10μF. I did not use these in this design because I have had too many bad experiences with tantalum capacitors.

This preamplifier is essentially composed of two compound stages, to wit, two cascodes, which was also the case for my earlier design in *audioXpress*. However, the topology differs in the way these two stages are connected. In the earlier article, they were cascaded independent gain blocks. Here they are cascaded, but then both are included in a frequency-dependent feedback loop that implements RIAA equalization.

You could achieve further improvement in mid- and high-frequency noise performance by reducing the value of R_3. This would probably cause slew-rate limiting at high-frequency, high-output levels. Increasing second-stage bias currents could eliminate this problem. Improvement in low-frequency noise performance would require different transistors. Switching to BJTs would offer the most likely improvement, but I have noted the interfacing problem with BJT inputs.

For those with a MOSFET fetish, a version replacing Q_3 with a p-channel MOSFET worked fine in PSpice simulation. The substitution did require a current source in lieu of R_8, but neither noise performance nor distortion suffered from the modification. However, I was satisfied with performance of the design shown here, and there is a philosophical argument that better is the enemy of good.

SOUND IMPRESSIONS

I should mention that I embarked on this project as a way to drive my power amplifiers directly from the output of this phono preamplifier, bypassing the system preamplifier altogether. These amplifiers require only 1.5V RMS input for 480W output. Thus, I could easily reduce R_3 for further noise reduction because onset of slew-rate limiting would not occur at a level these amplifiers would ever see. The point is, along the way I modified the design to be more generally useful, in the process sacrificing some advantages specific to my use.

Some strongly opinionated people notwithstanding, there are arguments for and against any scheme. I prefer the topology of the earlier design because gain can be increased or decreased by a change in the value of a single resistor in the second stage. With the earlier design, I increased overall gain and altered the gain distribution between the two stages allowing the phono preamplifier to drive the power amplifiers directly. The sound was, to my ear, noticeably cleaner. In any event, there is no way that passing the phono pre-preamp signal through the two additional amplifier stages of the system preamp can improve the sound.

For those who built or contemplate building the original design, I feel even better about its performance after several years. Although not designed for them, it performs respectably with both low output moving-magnet and moving-coil cartridges. It was not optimized for noise, yet noise was low enough to be inaudible at the listening position.

The sound was gorgeous, better than I had previously ever heard from vinyl. There was no volume control because the unattenuated gain was just sufficient to drive the speakers to a comfortable level provided the recording level was medium to high. Those records in my collection with low recording levels did not yield a high enough (to me) volume.

The major advantages to the new design are better noise performance, higher closed-loop gain, and, in my humble opinion, better sound. With a system in which the amplifiers are driven differentially via a phase splitter, the higher gain necessitated modifying the phase splitter to allow insertion of a 10kΩ dual potentiometer at its input for use as a volume control. For those who may have constructed the phase splitter[33], I will mention that the modification was to replace the BJT with a MOSFET as described in the referenced article.

As of this writing, the preamp has been in my system for over a year. Given how many times I have mounted and dismounted the preamp boards from the chassis and the number of times I've placed voltmeter and oscilloscope probes on the boards with the

POWER SUPPLY

Resistors *are 1%, metal film, 0.25W resistors unless otherwise specified. Values are in ohms.*
R_{GND} – 10, 2W metal film
R_{101}, R_{103} – 25, 3W wire-wound or metal film
R_{102}, R_{104} – 75, 1W wire-wound or metal film
R_{105}, R_{106} – 182
R_{107}, R_{108} – 100
R_{109}, R_{110} – 1.82K
R_{111}, R_{112} – 221

Capacitors *are 35V aluminum electrolytic capacitors unless otherwise specified. Values are in farads.*
C_{101}, C_{102} – 4.7m (1mF = 1×10^{-3}F)
C_{103} – 10m
C_{104} – 4.7m
C_{105}, C_{106} – 220μ, 25V
C_{107}, C_{108} – 100μ, 25V
C_{AC1}, C_{AC2} – 10n, 1kV disc ceramic

Diodes
BR1 – 1A, 200 PIV bridge rectifier
D_{101}, D_{102} – 100 p.i.v., 3A (1N5401 or equivalent)
Z_{101}, Z_{102} – 16V, 500mW (1N5246B or equivalent)

Transistors
Q_{101} – BC 549C
Q_{102} – BC 559C
M_{101} – IRF 9521
M_{102} – IRF 511

Transformer
T_1* – Power transformer designed for operation from 120V AC, 60Hz mains. 40V CT, 1A secondary.

Fuse
F_1* – 200mA normal blow

Switch
SW1* – Mains switch, 120V AC, 5A

*Common to both channels. Only one required.

inevitable occasional slip, I believe the design to be robust and forgiving of moderate abuse. This is a not-inconsequential consideration in a DIY project.

Gain cannot be easily changed with the present design. If gain is to be changed, five component values must be recalculated and the components swapped out. Even so, I have impressed myself with this current design. I wasn't at all sure at the outset that a reasonably performing preamp for low-output phono cartridges could be produced from two gain stages. Yet noise performance, RIAA tracking, and distortion are all perfectly acceptable.

Typically, moving-coil systems implemented with just two electronic stages require a step-up transformer. The noise benefits are certainly worthwhile, but when I look at transformer specifications for distortion and frequency response, I am moved to avoid them. If these factors were not enough to preclude my use of a transformer, the price tag of a good transformer is.

I am coming to the view that the amount of ambience heard in sound reproduction is somehow a measure of overall accuracy. If that is the case, this MC phono preamp is exquisitely accurate. Reproduction of big band and rock music is attention-getting. The most noticeable improvement, though, is in the reproduction of classical, i.e., orchestral music. Orchestral recordings have always seemed to lack something. No more is this true. There is both improved detail in the precise location of instruments and increased dynamics in, for example, the visceral impact of percussion.

As a member of the board of and season subscriber to the very good Tallahassee Symphony Orchestra, I have periodic opportunities to hear live performances. In some cases, the sound in my listening room is better than that in the concert hall, probably due to favorable positioning of the recording microphones and perhaps a better concert hall in the recorded version of a given work. This could not be the case, though, if the system were not up to the task.

No system can provide the impact of sitting on the stage during a performance of Gustav Holst's *The Planets*," which I narrated for the Chattanooga Symphony Orchestra a short time back. Nonetheless, if you place this article's preamp in a good system, I have no doubt that you will like what you hear. aX

REFERENCES

24. Werner Ogiers, "Simple Voltage Regulators," **www.tnt-audio.com**.

25. Andreas Schubert, "Parallel Regulators for Audio," *TAA*, 4/90, p. 40.

26. *Discrete Semiconductor Products Databook*, National Semiconductor Corporation, 1989, p. 2–16.

27. Tim Williams, *op. cit.*, p. 9.

28. *Ibid.*, p. 9.

29. Horowitz and Hill, *op. cit.*, p. 457.

30. Kerry Lacanette, *op. cit.*, Table 1, p. 822.

31. Wanda Garrett, *Linear Applications Handbook*, National Semiconductor Corporation, 1994, pp. 906, 907.

32. Norman Thagard, "Single-Ended to Differential Mode Made Easy," December, 2004, *audioXpress*, pp. 8–19.

Lowering A Loudspeaker's Mechanical Noise Floor, Pt 1

By Mark Sanfilipo

Discover how panel behavior affects speaker box responses.

When I design a loudspeaker system for home or studio use, with the goal of maximizing sonic accuracy, I focus on a handful of key areas in judging how successful the design is. Among the top three subjective criteria are soundstaging and transparency. One of the things that I as the designer can do to maximize transparency is to lower the noise floor of the system as much as possible, given a particular design.

In a past issue of *Speaker Builder* I presented a simple, easy way to lower the electromagnetic noise floor of loudspeaker systems that make use of passive crossover networks. In this article I look at cabinet panel vibration and its contribution to the mechanical noise floor of the system and present a method to mathematically model the Insertion Loss (IL) of a cabinet's panels and apply the results to actual driver response curves.

MODEL FEATURES

The test subject I model is a totally enclosed mid-bass system, operating with a bandpass response. The low-pass -3dB point is established by a crossover network at 430Hz, with the particular driver/cabinet alignment chosen providing for a high-pass -3dB point at 40Hz.

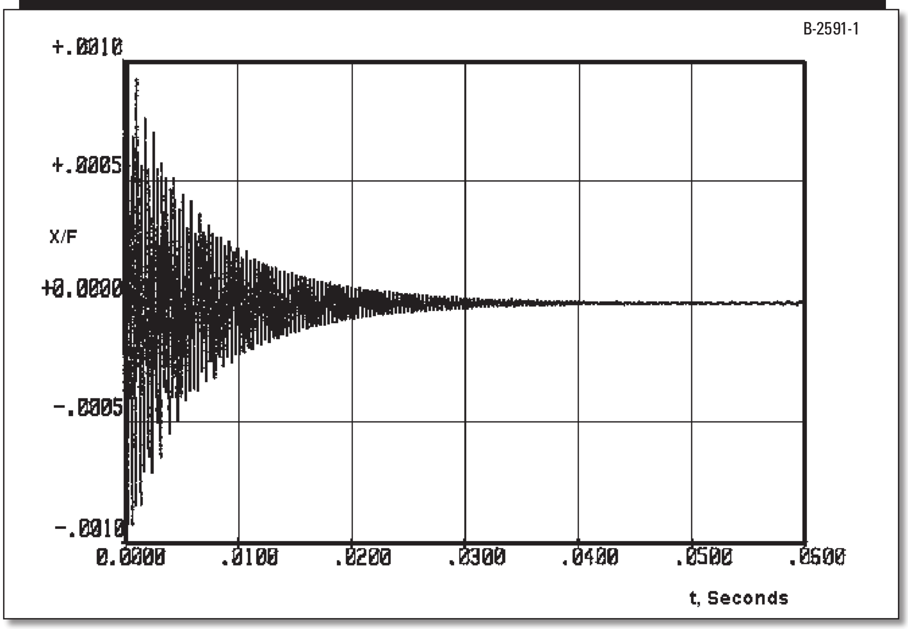
Whenever appropriate, I place each driver in its own cabinet. One of the benefits is that it usually becomes easier to identify and put in place appropriate design solutions to lower the system's mechanical noise floor. With the mid-bass cabinet's bandpass -3dB points set at 40Hz and 430Hz, my design goal becomes clearer and therefore easier to define: What panel output that can't be pushed outside the bandpass region of the system's response will be minimized to the degree that the design (and budget) will allow?

The cabinet's panels will, virtually, be built of a polymer composite. I chose the composite not only for its well-understood mechanical properties, but also because of its strength, excellent dimensional stability, and superior vibration damping properties (**Fig. 1**). Indeed, the material I used for this model has the ability to damp vibration four times faster than granite! Knuckle-rap a cabinet made of this composite and it feels—and sounds—as though you're knocking on a block of solid rock.

I could just as easily have used a wood product to construct the virtual cabinet simulations, but wood products—such as plywood—are orthotropic (its measured mechanical properties depend on the direction being considered) by nature, which would have added a needless degree of complexity to the following analyses. Compare the birch mechanical properties (**Table 1**) with that of the table outlining the mechanical properties of the composite material given in **Table 2**.

As well, if you'd like to run FEA simulations using wood products, you may find it a challenge to track down all the information (such as a full spectrum of mechanical properties) needed to run accurate, dependable simulations.

FIGURE 1: Composite sample impulse response.



Having said that, you may still find it's worth the effort, because running FEA simulations prior to actual construction can provide insight to your proposed project's mechanical and acoustical behavior, thus enabling you to make better-informed design decisions while the project is still in its formative stages.

TABLE 1: Sample Wood (Birch) Properties

Physical properties	Metric	English
Density	.55g/cc	0.0199 lb/in^3
Mechanical properties	Metric	English
Hardness, side indentation	3.3N	0.742 lb-f
Hardness, end indentation	3.6N	0.809 lb-f
Ultimate tensile strength (⊥ to grain)	6.3 MPa	914 psi
Modulus of rupture in bending	0.115 GPa	16.7 ksi
Flexural modulus	13.8 – 13.9 GPa	2000 – 2020 ksi
Flexural yield strength at elastic limit	31.7 MPa	4600 psi
Compressive yield strength, | | to Grain	56 – 56.3	Mpa 8120 – 8170 psi
Compressive yield strength, ⊥ to Grain	6.7 MPa	972 psi
Poisson's ratio, µTL	0.024	0.024
Poisson's ratio, µRL	0.043	0.043
Poisson's ratio, µTR	0.426	0.426
Poisson's ratio, µLR	0.426	0.426
Poisson's ratio, µLT	0.451	0.451
Poisson's ratio, µRT	0.697	0.697
Shear strength	13 MPa	1890 psi
Toughness, radial	56.5 J	41.7 ft-lb
Toughness, tangential	70.1 J	51.7 t-lb

Panel acoustic output results from varying internal surface pressure arising from the motion of the driver's diaphragm as well as the mechanical vibration induced by the motion of the driver's frame, along with that of other panels. (The latter case is easy to envision when you consider the high degree of impedance matching that will exist between panels made of the same material, unless measures are taken to provide for an impedance mismatch between the panels; doing so will reflect vibrational energy back to the source panel.)

In either case, you can minimize panel vibration by a variety of means including isolation, absorption, a combination of both, or other such engineered solutions. In this first part of a two-article series, I'll focus directly on

TABLE 2: Mechanical Properties of Isotropic Polymer Composite

Properties	Metric Units	English Units
Density	2.32g/cm^2	0.084 lb./in^3
Compressive strength	110 – 117MPa	16,000 – 17,000psi
Tensile strength	15.2MPa	2,200psi
Flexural strength	17.2MPa	2,500psi
Young's modulus	35,982MPa	5.23E6psi
Poisson's ratio	0.23	0.23
Thermal expansion	16.9E-6 mm/mm/°C	9.4E-6 in/in/°F
Thermal conductivity	0.244W cm/°C	14.1 BTU/hr ft^2 °F/in
Specific heat	0.23 – 0.26 cal/g °C	0.23 – 0.26 BTU/lb. °F

the panels themselves and how you can minimize this sort of secondary acoustic radiation.

THE ANALYSIS

Loudspeaker cabinet panels are typically rectangular shaped and the cabinet itself can be considered—to a good first approximation—an enclosure comprising a collection of plates (defined as a construct whose lateral dimensions are much greater than its thickness), clamped at each edge. Therefore, for all analytical purposes of the following simulations I've run, I'll consider the individual panels to be clamped, rectangular plates. As previously mentioned, the vibrations that naturally set up within each panel arise from both airborne pressure fluctuations produced on the interior panel surface by the action of the driver's diaphragm as well as structural-borne mechanical vibration set up as a reaction to vibration of the driver's frame.

To understand the basics of panel resonance as well as panel acoustic output, it's useful to resort to a lumped-parameter, acoustic-equivalent circuit modeling driver/cabinet interaction (**Fig. 2**). The usual caveats for lumped-parameter simulation are applicable: driver operating in piston mode, all cabinet dimensions are smaller than the wavelengths of interest, and so on,

where:

Pas = acoustic pressure generator

Zas = acoustic equivalent of loudspeaker impedance

$Mw_{1,2,3}$ . . . = equivalent acoustic mass of individual cabinet panels and air load

$Cw_{1,2,3}$ . . . = equivalent acoustic compliance of individual cabinet panels

$Rw_{1,2,3}$ . . . = equivalent acoustic resistance of individual cabinet panels owing to damping

Cab = acoustic compliance of air in enclosure

Cab' = added compliance owing to cabinet filling and/or lining with fibrous material

Rar = acoustic radiation resistance of air load on the front of the cone and box wall(s)

Ralb = leakage around box compliance

Rab = acoustic resistance of box losses owing to absorption

Below the fundamental resonance (the frequency at which Cw resonates with Mw), the effect of Cw provides for the primary control of the circuit's impedance. At resonance, impedance is controlled by Rw, the panel's equivalent acoustic resistance owing to damping. At resonance the Mw-Lw-Cw circuit effectively shunts the Rar-Cab-Cab'-Ralb-Rab branch of the circuit, causing a dip in response. Just above resonance, where Mw resonates with the Cab branch of the circuit, the volume velocity through Rar is increased and you get a peak in the response plot.

In both cases (interestingly enough) a large enough Rw value would minimize the shunting effect, in turn minimizing the magnitude of the response peak or dip. The best practice, therefore, is to use a material with as high an Rw value as the project budget allows.

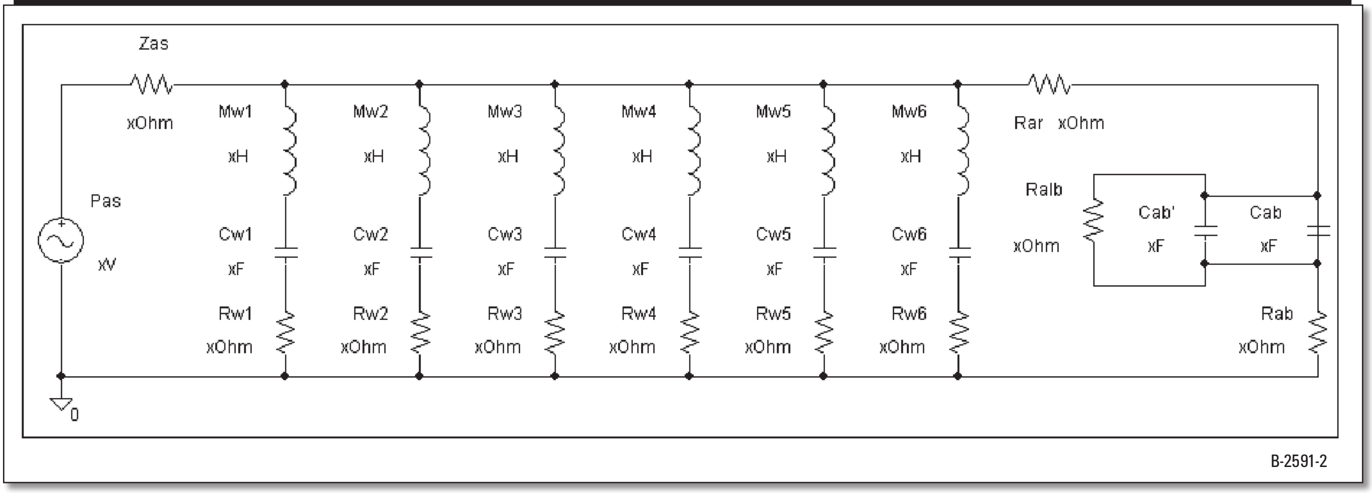
Any panel—whether simply supported or clamped at any (or every) edge—will exhibit a clearly defined, vibrational mode series known as eigenvalue, which are theoretically infinite in number and, in general, are not harmonically related. To demonstrate this vibrational series, I ran a FEA (finite element analysis) simulation, modeling a rectangular panel, clamped at all four edges, with the following physical dimensions:

Dimension	Named Variable	Value
Length	a	20″ (.508m)
Width	b	10″ (.254m)
Thickness	h	.75″ (.01905m)

In general, when working up a simulation series, I tend to model only the first 30 or so modes, because the resonance frequencies that occur above that range, as I already mentioned, generally contribute little to the system's total acoustic output. Phase-cancellation, owing to different segments of the panel moving in opposite directions and diminishing panel volume-velocity, all add up to a lowered panel contribution to total system acoustic output, when the panel is operating in the upper reaches of the mode series.

Figures 3-6 show the results, in graphical form, illustrating the magnitude of peak panel displacement at various mode numbers. The panel displacement,

FIGURE 2: Driver/cabinet interaction circuit.



as modeled for each simulation series, is exaggerated in appearance for the sake of visual clarity.

It's clear from this series of models that rectangular panels do exhibit resonance modes, each with a distinct pattern of panel deformation. Indeed, note that different sections of the panel can, at the same moment, be moving in opposite directions.

In general, the higher modes will move different areas of the panel in opposite directions, effectively diminishing the net acoustic radiation of the panel owing to the diminished net panel area capable of producing the acoustic radiation in the first place. In other words, phase cancellation, aris-

FIGURE 3: Panel displacement. Mode 1 = 67.778Hz.

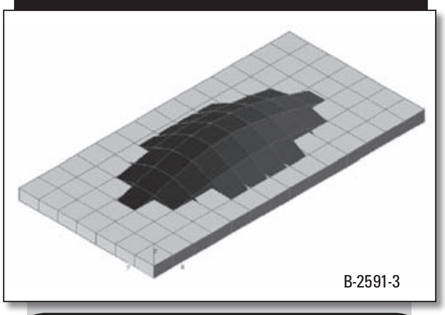


FIGURE 4: Mode 4 = 85Hz.

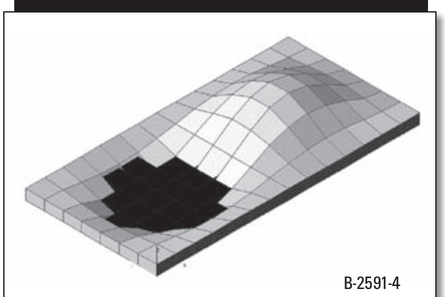


FIGURE 5: Mode 9 = 176Hz.

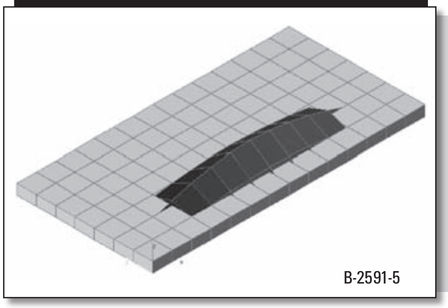
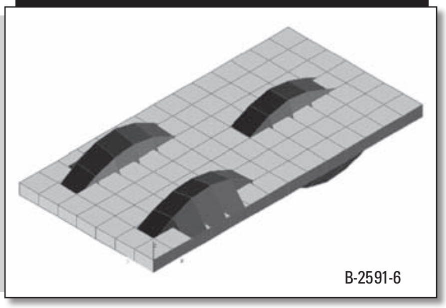


FIGURE 6: Mode 27 = 344Hz.



ing from different panel areas moving in opposite directions, limits the net amount of total acoustic radiation emitted. Add to this the fact that higher order modes are easier to damp, and you see these self-same modes are generally not in a position to affect system response to the degree the lower order modes can.

RESPONSE FACTORS

To deal with these problematic lower resonance modes, the best practice of the heavy-panel school of resonance control (as opposed to the light-panel school, which advocates using materials such as Aerolam) requires using panels of appropriate thickness, made of a stiff material exhibiting a high degree of internal damping. Further reductions result from shifting the fundamental resonance frequency of the panel high enough so that it no longer presents itself within the passband portion of the system's response spectrum, in this case 40Hz to 430Hz. In past projects, I've found cross-bracing cabinet panels—usually using a material stiffer than that from which the panels are made—a useful and effective means for shifting these natural panel resonance frequencies upward.

I've run some FEA simulations to look at the effects of panel cross-bracing. I re-ran the above simulation, this time modeling a series of bracing techniques. For the first simulation, I braced the panel vertically (across the mid-section of the panel parallel to the front-edge Y-axis, with the results in **Figs. 7-10**).

Bracing the panel vertically produced a series of modes clustered around 96Hz. This upward shift of the fundamental resonance is a step in the right direction, but not nearly high enough for my purposes.

I then modeled the panel braced horizontally, with the brace located at .5y, parallel to the x-axis and running the full length of the panel. The results are shown in **Figs. 11-15**.

In this series of simulations the upward shift of the natural panel resonance frequencies is quite apparent, producing superior results to that of the vertically positioned brace. For a cabinet to be used for the reproduction of the mid-bass portion of the total system response, this is again a step in the right direction, but still not high enough for my purposes.

In actual practice I use multiple horizontal braces for a mid-bass system of this size. Modeling two braces located approximately one-third and two-thirds up the y-axis (front) edge of the panel and running to the back edge of the panel produces the results

FIGURE 7: Vertical bracing. Mode 1 = 96Hz.

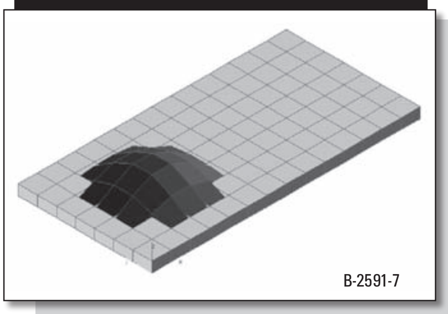


FIGURE 8: Mode 2 = 96Hz.

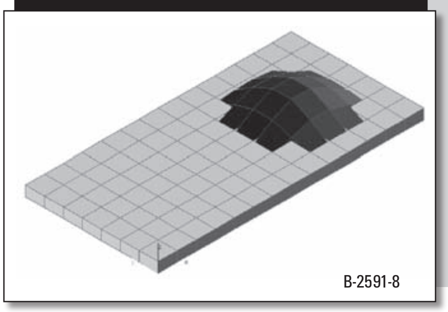


FIGURE 9: Mode 3 = 96Hz.

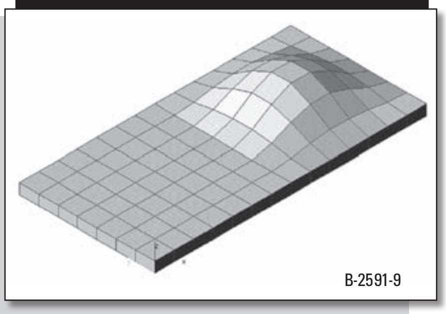


FIGURE 10: Mode 4 = 96Hz.

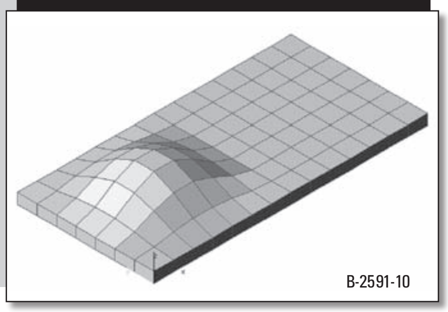


FIGURE 11: Horizontal bracing. Mode 1 = 195Hz.

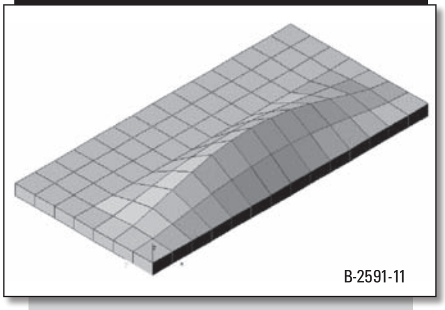


FIGURE 12: Mode 3 = 203Hz.

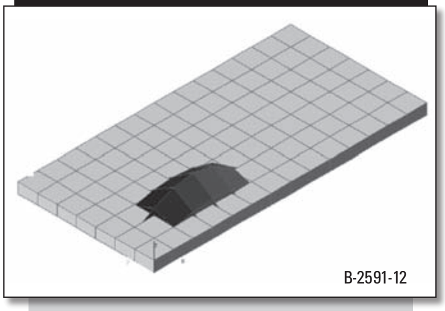


FIGURE 13: Mode 4 = 203Hz.

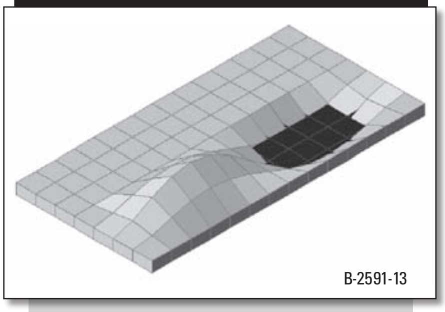


FIGURE 14: Mode 10 = 291Hz.

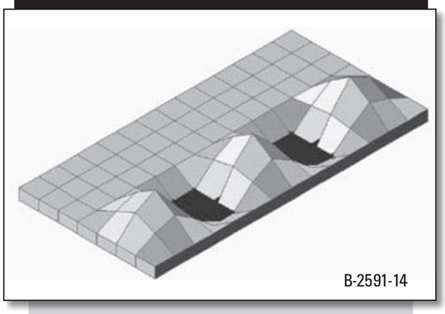


FIGURE 15: Mode 15 = 349Hz.

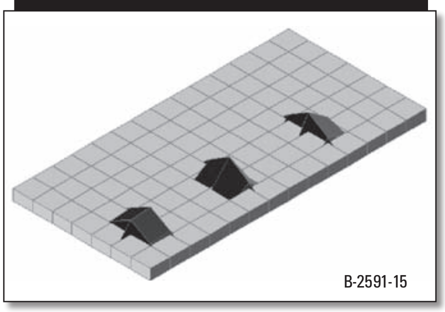


FIGURE 16: Two braces. Mode 1 = 344Hz.

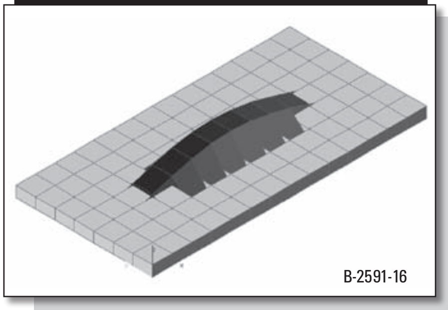


FIGURE 18: Mode 10 = 387Hz.

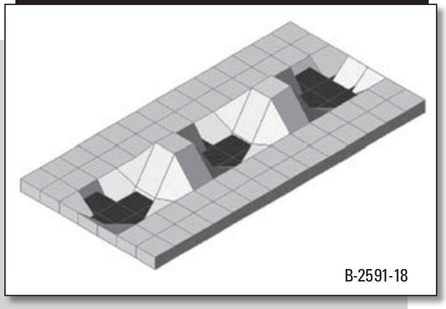


FIGURE 17: Mode 4 = 345Hz.

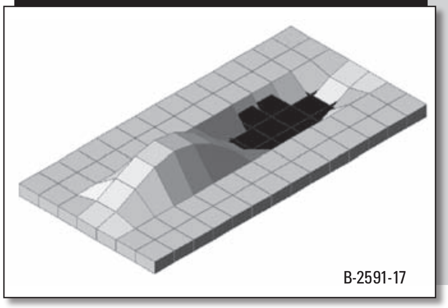
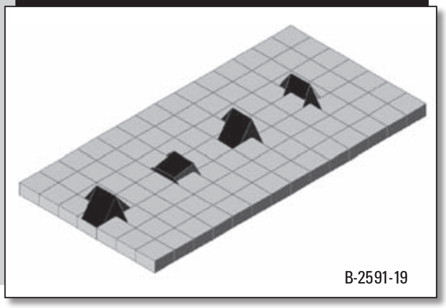


FIGURE 19: Mode 15 = 542Hz.

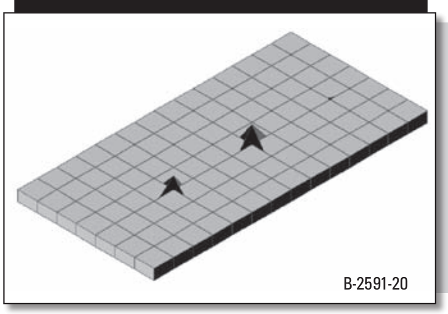


in **Figs. 16-19**.

The frequency at which the fundamental natural panel resonance occurs is now in the neighborhood of the 430Hz low-pass point. However, I'd like to see the mode series pushed higher. Modeling three braces, placed at .25y, .5y, and .75y and running the full length of the panel from front to back produces the results in **Fig. 20**.

The first mode of vibration is now above the low-pass cutoff frequency, but only by about a half octave. I could add another brace, which would certainly push the mode series higher. Because I already know, by virtue of the preceding simulations, that bracing works well, I chose at this point in the series instead to re-run the simulation, keeping the three braces in place, but now doubling the thickness of the panel to 38mm (1.5″). Doing so produced a first mode resonance frequency of 701Hz (**Fig. 21**).

FIGURE 20: Three braces. Mode 1 = 658Hz.



Not bad. Doubling the pane thickness again produced the result in **Fig. 22**. Here you see the fundamental mode frequency increase a bit to 720Hz. There are, of course, practical limits to how much bracing you can employ or how thick panel walls can be made in a design.

Note that both methods used—brac-

FIGURE 21: Three braces, panel thickness doubled. Mode 1 = 701Hz.

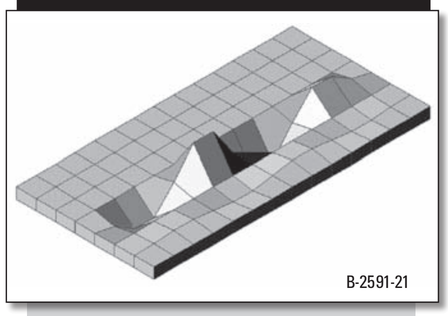
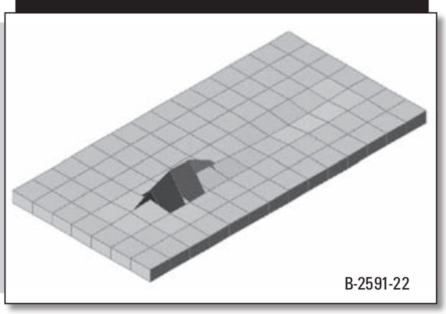


FIGURE 22: Three braces, panel thickness quadrupled. Mode 1 = 720Hz.



ing and increasing panel thickness—also increase Rw, thus increasing damping. There also exists another approach to panel resonance control worth mentioning: constructing the cabinet in such a way as to stagger the resonance mode frequencies, so that no two panels exhibit the same acoustic behavior.

Thus far you have seen that by using the appropriate panel material, with high enough damping, properly braced and of the correct thickness, a designer can effectively shift the mode series high enough and/or otherwise minimize the effect panel resonances can have on final system output and do so to the point they no longer pose response problems for the system. Sometimes, though, shifting panel resonances upward can be counter-productive. For example, if doing so shifts them into a portion of the response spectrum where they will have a deleterious effect on the system's output, doing more harm than good, then shifting them downward, or staggering the response aberrations is called for.

In a case where the techniques illustrated thus far don't sufficiently resolve panel resonance issues, there are other means you can try. For example, resorting to strategically located constrained layer damping CLD panels (typically a metal sandwich made up of a base layer and constraining layer, bonded by a visco-elastic material filling layer), which dissipates vibration energy via sheer deformation, can be very effective. Extensional damping (adding a single layer of visco-elastic material to the inside of the panels), which dissipates vibration energy by converting it into heat by way of friction arising due to material flexure, works well but depends upon the amount of flexure it is exposed to in order to do what it does—and the amount of flexure it's exposed to at any given moment may be very, very small. This works well against impact noise. Of course, simply choosing a different material better suited to your design requirements is another appropriate alternative.

THE MATHS

The classic equation for determining the resonance frequencies of a uniform, isotropic, rectangular panel, with boundary conditions set as all edges, simply supported (ss), is:

$$f_{SS} = \frac{1}{2*\pi} * \sqrt{(\frac{D}{\rho})} * \left[(\frac{m\pi}{a})^2 + (\frac{n\pi}{b})^2 \right]$$

and for the same panel with boundary conditions set as all edges clamped (c), the fundamental resonance frequency can be found by:

$$f_C = \frac{12}{2*\pi} * \sqrt{(\frac{D}{\rho})} * \sqrt{\frac{7}{2} * \left(\frac{1}{a^4} + \frac{4}{7} * \frac{1}{a^2 * b^2} + \frac{1}{b^4} \right)}$$

where:

f = Resonance frequency, Hz

$D = \frac{E * h^3}{12 * (1 - \eta^2)}$, flexural rigidity or bending stiffness of plate

m, n = mode numbers, integers

ρ = panel mass per unit area

E = Young's modulus, modulus of elasticity

h = panel thickness

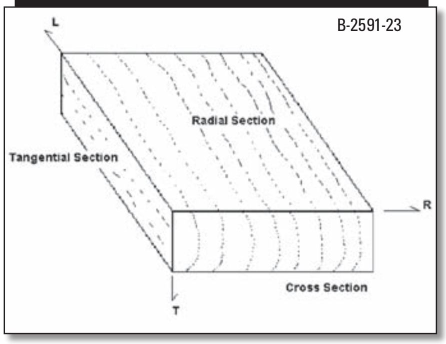
a = panel length

b = panel width

η = Poisson's ratio (ratio of orthogonal straining)

A few caveats about these equations are in order. First, note the phrase "isotropic," which means a material that exhibits mechanical properties (such as Young's modulus) that are independent of which direction the measurements are taken from which the mechanical properties are calculated. A good example of an isotropic material is MDF or, as mentioned previously, a polymer composite material (**Fig. 23**). A material such as plywood would not be considered isotropic, but rather orthotropic, which means (as you've likely already guessed) its measured mechan-

FIGURE 23: Direction affects measured box material properties.



ical properties depend on the direction (laterally, transversely, or radially) being considered.

A second caveat is: When you start cutting openings in the panel, the formulas are no longer applicable, and measured versus calculated resonance values will differ substantially. This doesn't mean the formulas aren't useful in speaker building, they are. But I would use them only as a starting point for working up a rough estimate of panel resonance frequency for those panels that don't have openings cut into them and only if I were using an isotropic material such as MDF or a polymer composite.

As I mentioned earlier, the contribution to total system acoustic output made by the panels does not arise solely from the mechanical vibrations set up in each panel by the motion of the driver's frame and other adjacent panels. You need to consider another factor: the outward transmission by the outer side of a panel due to fluctuating pressure incident to the internal side of a cabinet panel. For the following analysis, again assume that the panels behave as plates, clamped at each edge and under uniform load by the driver diaphragm-induced acoustic pressure fluctuations.

In any case, the internal pressure fluctuations found within a loudspeaker cabinet under normal operating conditions aren't very large to begin with and further decrease as frequency increases. The internal box pressure produced by an operational driver bolted into one panel of the cabinet can be found, to a good first approximation, by:

$$P_B = \frac{\sqrt{\frac{P_A * 2 * \pi * c}{\rho_0} * \frac{\rho_0 * c^2}{V_B}}}{(2 * \pi * f)^2}$$

where:
P_B = Rms box pressure, pascals
P_A = radiated acoustic power, watts
c = velocity of sound in air, ~345 m/s
ρo = density of air, 1.18 kg/m^3
V_B = net internal box volume, liters
f = frequency, Hz

Figure 24 shows the results of a series of simulations, holding P_A constant at .1W and varying V_B, using the above equation.

One common mathematical method used for expressing the degree to which a panel can reduce the outward expression of these internal pressure fluctuations is by transmission loss, which, in dB, is expressed mathematically by:

$$TL = 20 * \mathrm{Log}_{10}\left(\frac{1}{\tau}\right)$$

where τ, the transmission loss coefficient, is expressed by:

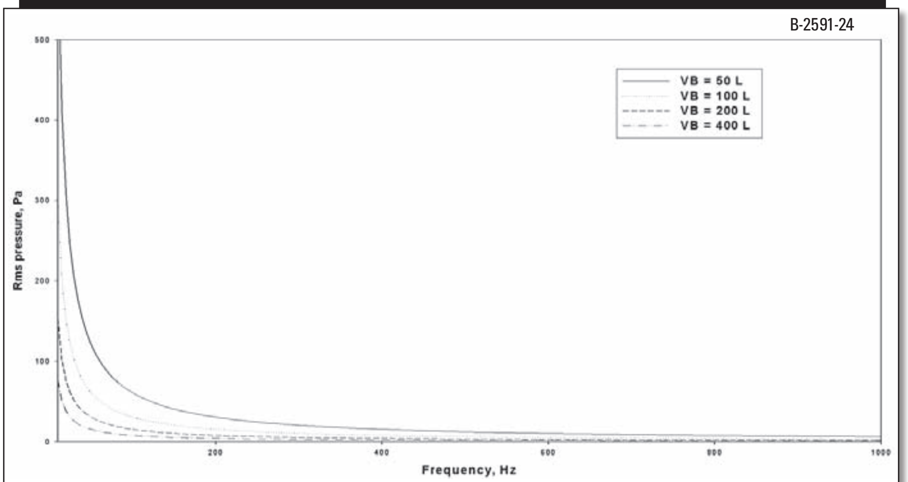
$$\tau = \frac{|P_i|^2}{|P_t|^2}$$

P_i = incident pressure
P_t = transmitted pressure

Factoring in panel mass and frequency, the TL equation then becomes:

$$TL = 10 * \mathrm{Log}_{10}\left(\left(\frac{m * \omega}{2 * z}\right)^2 + 1\right)$$

FIGURE 24: Steady-state RMS pressure vs. frequency.



where:
m = panel surface density, (ρ * h)
ρ = material density
h = panel thickness
Ω = radian frequency
z = panel material impedance

Taking into account m, and solving for B, the bulk modulus or effective stiffness of the panel is found by:

$$B = \frac{E_{Panel} * h^3}{12 * \left(1 - \eta_{Panel}^2\right)}$$

And from the results of the preceding equation you can now calculate the panel's critical frequency:

$$F_c = \frac{c^2}{2 * \pi} * \sqrt{\frac{m}{B}}$$

Finally, you can calculate the transmission loss slope below $F_c/2$:

$$TL = 20 * \mathrm{Log}_{10}\left[\left(\frac{2 * \pi * \left(\frac{F_c}{2}\right) * F_c * m}{2 * \rho * c}\right)^2\right] - 5$$

And the slope above F_c:

$$TL = 10 * \mathrm{Log}_{10}\left(\left(\frac{2 * \pi * F_c * m}{2 * \rho * c}\right)^2\right) + 10 * \mathrm{Log}_{10}\left(\frac{2 * \eta}{\pi}\right)$$

Once you have calculated the TL values, assessing the effective reduction of transiting acoustical energy by the panel becomes a simple matter of subtracting TL from the dB SPL level of the source—in this case, those levels in the immediate vicinity of the cabinet-internal side of the panel.

These expressions are, of course, panel-focused, and the majority of transmission loss theory found in the literature—upon which these equations are based—is for wall-sized (or larger) "panels" dividing room-sized "enclosures" sporting ideal (or completely lacking) boundary conditions. The results are not necessarily applicable to panels of the material type dimensions and boundary conditions commonly found in loudspeaker enclosures.

INSERTION LOSS

Searching for a better enclosure-focused approach to characterize the outward re-radiation of acoustic energy by an enclosure's panels leads to the concept of insertion loss. Mathematically, the general expression of insertion loss (IL) is defined as:

$$IL = \mathrm{Log}_{10}\left(\left(\frac{SPL_{External}}{SPL_{Internal}}\right)^2\right)$$

where:
$SPL_{External}$ = acoustic sound pressure level, external to cabinet panel
$SPL_{Internal}$ = acoustic sound pressure level, internal to cabinet panel

The equation is instructive but nevertheless descriptive in nature.

Fortunately, there exists a predictive mathematical expression for determining the insertion loss of a loudspeaker cabinet panel that takes into account various factors such as enclosure volume impedance, panel mechanical impedance, and so forth. For a close-fitting cabinet (one in which the distance from internal sound source to any panel is 1m or less), with clamped boundary conditions, IL can be mathematically stated by:

$$IL = 10 * \mathrm{Log}_{10}\left[\left(\mathrm{Cos}(k * d) + \left(\frac{\pi^2}{4 * K * \omega * \rho_0 * c}\right) * \mathrm{Sin}(k * d)\right)^2\right]$$

where:
k = acoustic wave number, $\frac{2 * \pi}{\lambda}$
λ = wavelength
d = distance from source to panel
ρ_0 = density of air
c = speed of sound in air

and

$$K = \frac{1.35}{\left[3.86 * D_i\left(\frac{129.6}{a^4} + \frac{78.4}{a^2 * b^2} + \frac{129.6}{b^4}\right) - \omega^2 * \rho * h\right]}$$

where:
a = panel length
b = panel width
h = panel thickness
ρ = density of panel material
Ω = frequency of interest
and

$$D_i = \left(\frac{E * h^3}{12 * (1 - \upsilon^2)}\right) * (1 + i\eta)$$

which is the complex bulk modulus of the panel material
η = internal damping coefficient of the panel material
E = Young's modulus or modulus of

elasticity
ν = Poisson's ratio (ratio of orthogonal straining)

For a polymer composite panel with the material properties as listed in **Table 2** and with the dimensions: a = 20″ (.508m), b = 10″ (.254m), and h = .75″ (.01905m), you can calculate D_i, K, and finally IL.

Figure 25 shows the IL graph. Here

FIGURE 25: Insertion loss vs. frequency.

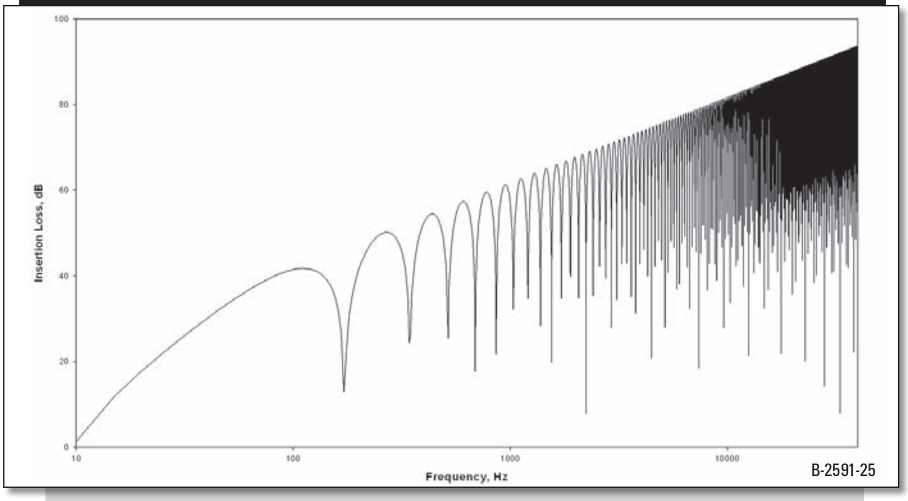


FIGURE 26: Varying panel length.

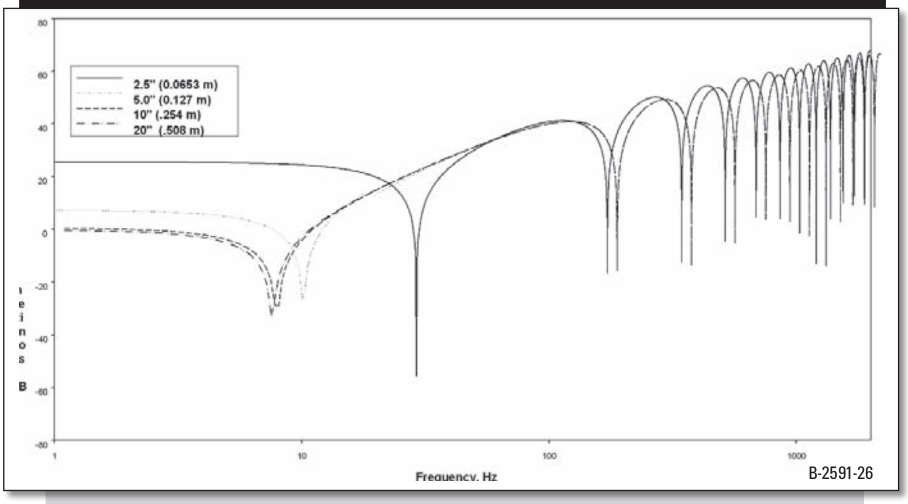
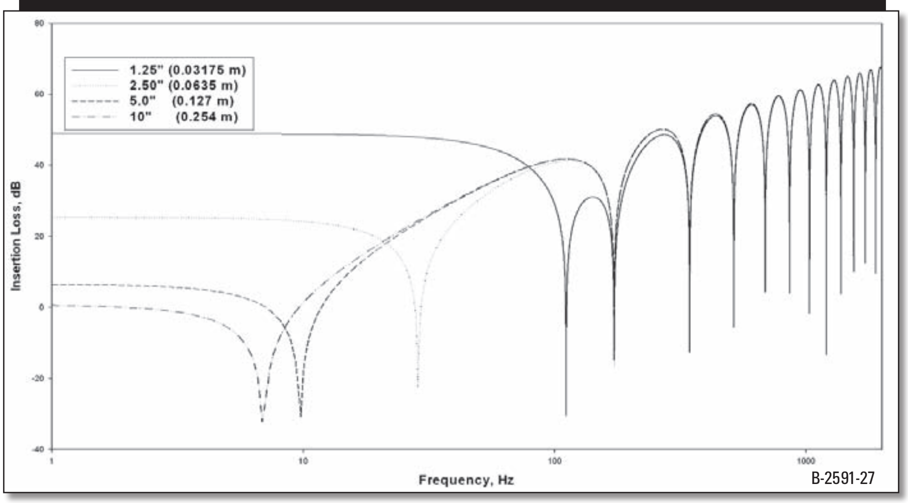


FIGURE 27: Varying panel width.



you see a series of nulls alternating between peaks in IL values. It's clear the responses of a cabinet's panels depend on both the frequency of the various excited modes and the frequency generated by the driver enclosed within the cabinet. As well, the IL response at higher frequencies is controlled by various material properties along with the distance between the panel and driver. When panel vibration is in phase with the incident pressure wave, the panel becomes acoustically "invisible" to the incident pressure wave. Standing waves within the enclosure have the characteristic values of:

$$\omega_x = \frac{c * \pi}{lx} * n_x, \omega_y = \frac{c * \pi}{l_y} * n_y, \omega_z = \frac{c * \pi}{l_z} * n_z$$

and frequencies can be approximated by:

$$f = \frac{c}{2} * \sqrt{\left(\frac{n_x}{l_x}\right)^2 + \left(\frac{n_y}{l_y}\right)^2 + \left(\frac{n_z}{l_z}\right)^2}$$

where:

c = speed of sound in air,

nx, ny, nz = 0, 1, 2, 3 . . .

lx, ly, lz = internal length, width depth

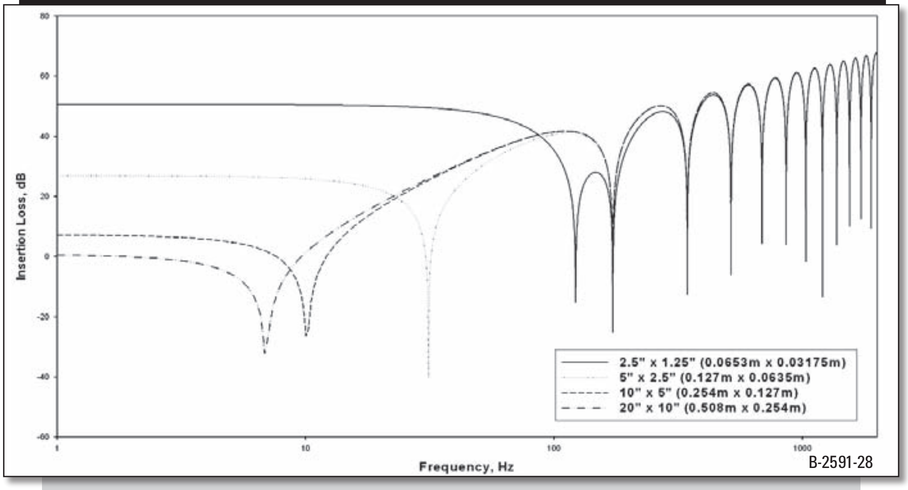
These standing waves can produce acoustic wave pressures greater than that generated by the driver itself. This can in turn, under certain conditions—such as when panel vibration is in phase with the incident pressure wave—result in panel vibration of a magnitude sufficient to be considered a sort of negative insertion loss.

Curious to see how varying the values of various parameters would affect the IL dB levels, I ran a detailed series of simulations. First I varied panel length (**Fig. 26**). Then I varied panel width (**Fig. 27**). Varying length and width, individually, did indeed alter the IL response spectrum. I then varied length and width simultaneously (**Fig. 28**). In all three cases, as the panel became smaller, you see an increase in the frequency of the first null as well as an increase in IL.

I then varied panel density (**Fig. 29**). Across the IL response spectrum you see that increasing density increases the amount of insertion loss. All else being equal, you can thus expect that the denser a given panel's material, the quieter it will be.

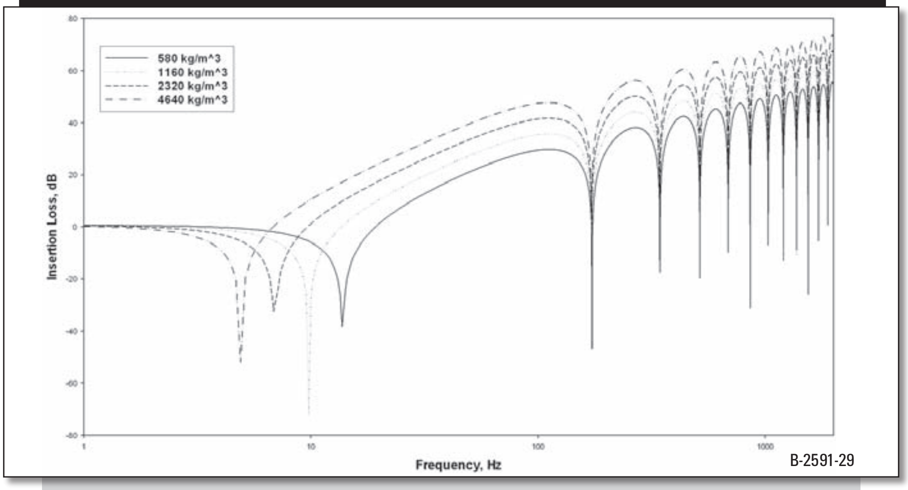
Then, I varied panel thickness (**Fig.

FIGURE 28: Varying panel length and width.



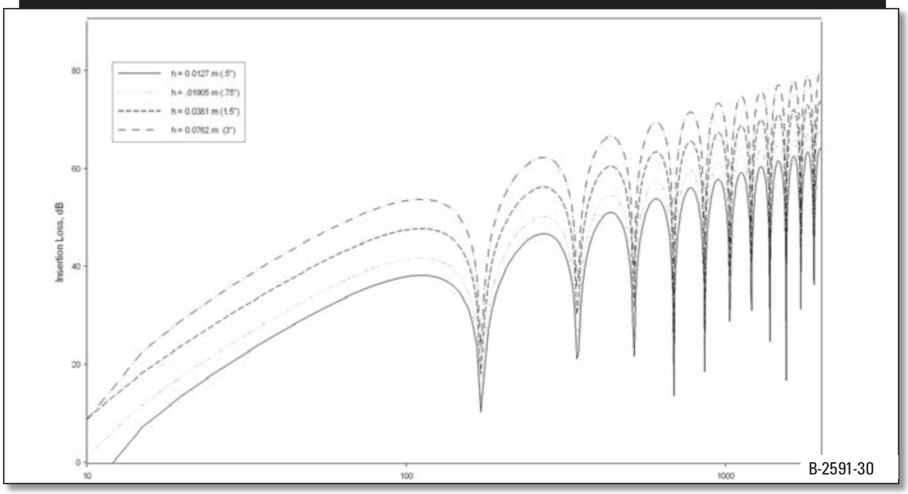
B-2591-28

FIGURE 29: Varying panel density.



B-2591-29

FIGURE 30: Varying panel thickness.

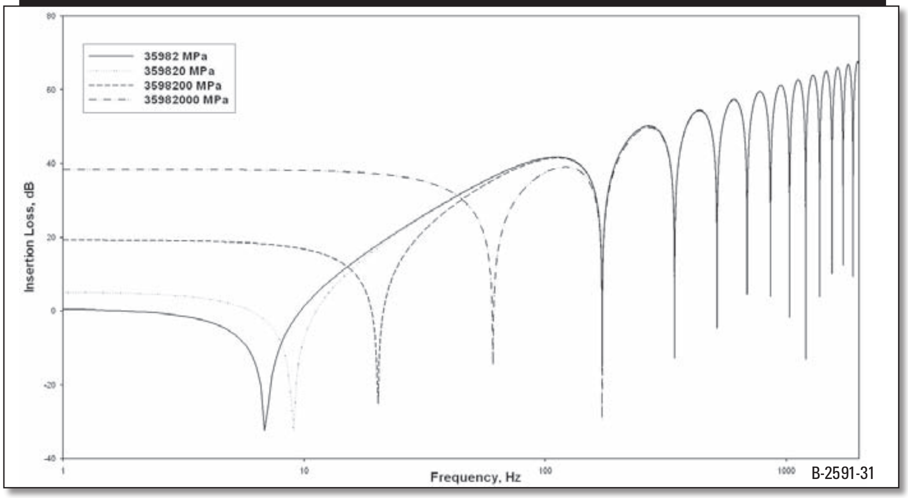


B-2591-30

30). Increasing panel thickness increases effective stiffness, which results in increased insertion loss, particularly noticeable at the lower end of the IL response spectrum, where insertion loss is controlled directly by the panel's effective stiffness.

Last, I varied the panel material's Young's modulus, a mechanical parameter that effectively characterizes a panel's stiffness (**Fig. 31**). As expected, you see that IL response variations happen primarily in the lowest decades of the IL response spectrum; IL increases as E increases. Note too that with increased stiffness you see the first,

FIGURE 31: Varying panel Young's modulus.



lowest frequency IL minimum shift upward in frequency.

To complete the simulation run, I applied various simulated IL response spectra to actual db-SPL driver response curves, which resulted in predicted panel acoustic response, when driven by the acoustic pressure waves generated by the driver enclosed *within* the cabinet.

Specifically, I measured a raw driver, then simulated a totally enclosed cabinet, made of the polymer composite with the driver placed within the cabinet (not bolted into the face of one of the cabinet's panels). The simulations produced predicted results of dB-SPL measurements, taken at 1m, outside the cabinet (**Fig. 32**).

BASELINE

The top curve is the actual dB-SPL response of the raw driver, measured at 1m. The bottom curve is that of the predicted response as measured outside the simulated cabinet that totally encloses the driver, at a distance of 1m.

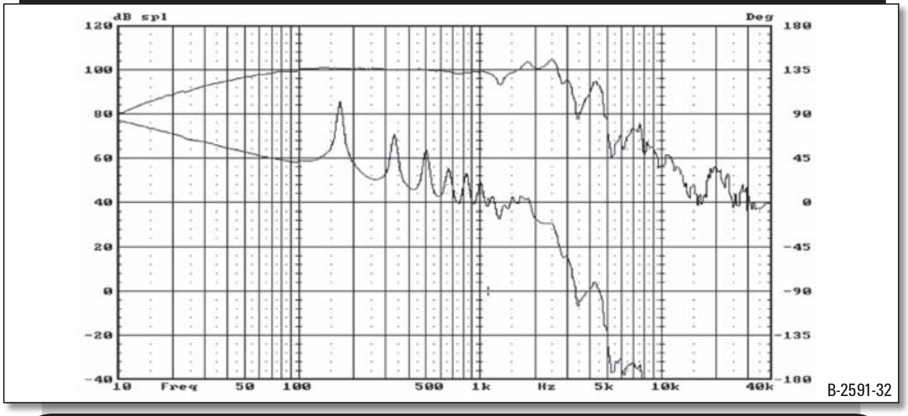
Keep in mind these results are for an enclosure that contains no sort of sound-absorbing filling or liner; thus the thermodynamic condition within the enclosure is adiabatic. Using an absorptive filling and/or liner would alter the predicted results, likely lowering the actual dB SPL measured outside the cabinet walls. As well, the simulation does not account for panel acoustic output resulting as a consequence of mechanical vibrations originating at the driver's frame.

At any rate, once having established a performance baseline, I then ran a further set of simulations, each with the purpose of exploring (and illustrating) the effects of varying mechanical properties of the cabinet panel's material.

Continued on p.52

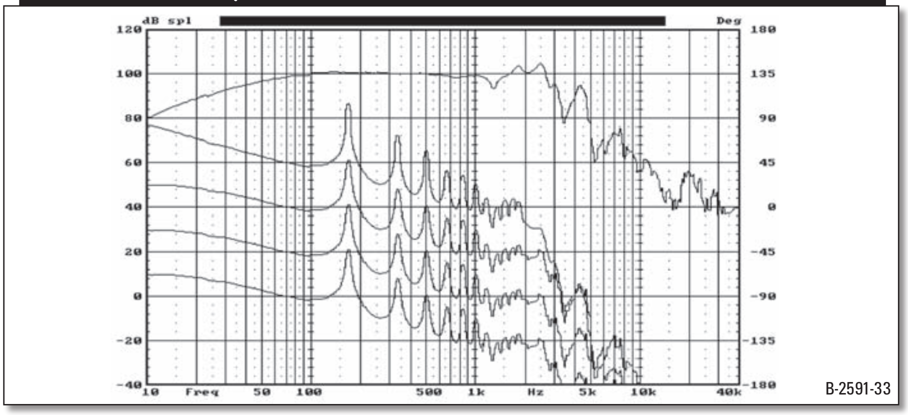
Continued from p.49

FIGURE 32: Raw driver simulation.



B-2591-32

FIGURE 33: Density variations.



B-2591-33

My first simulation series (**Fig. 33**) held all properties constant, except for density. Varying density produced substantial differences in the amount of insertion loss and in turn substantial effects on panel acoustic output. The resulting curves show that the more dense the material, the higher the insertion loss, resulting in decreased panel dB SPL levels.

Next, I re-ran the simulation, this time holding all parameters constant, save stiffness (**Fig. 34**). Increasing stiffness did affect IL, but it took quite a substantial increase in panel stiffness before significant changes in IL and thus panel acoustic output became evident. You can see from **Fig. 33** that IL did increase as stiffness increased, and at the lower end of the response spectrum, just as predicted by theory.

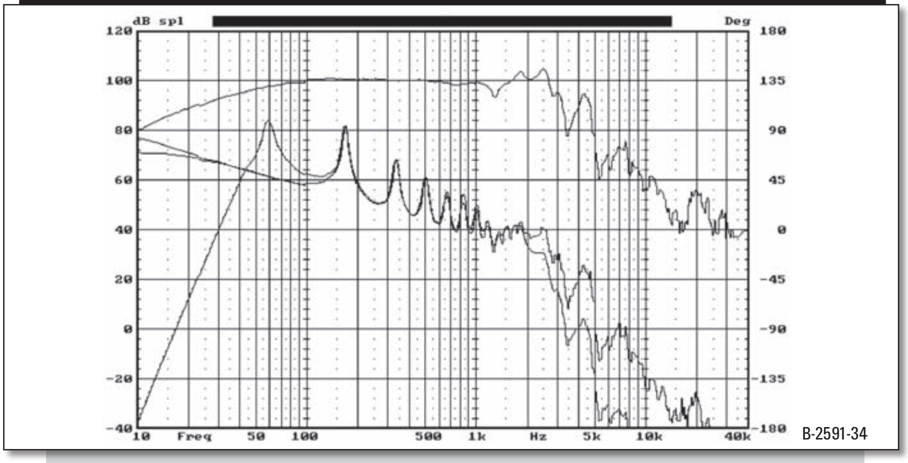
Last, I ran a series of simulations holding all parameters constant, save thickness. As you can see in **Fig. 35**, increasing panel thickness increases IL.

CONCLUSION

In this article you have seen that panel resonances—either structural or airborne in origin—can have effects, usually deleterious in nature, on system response. You have also seen that the designer has at his or her disposal a number of means to damp, disperse, or shift these resonances out of the frequency range of interest. I think it also fair to say that the use of such modeling techniques as FEA can be an invaluable, efficient tool for exploring various design solutions when seeking to lower the mechanical noise floor of the system.

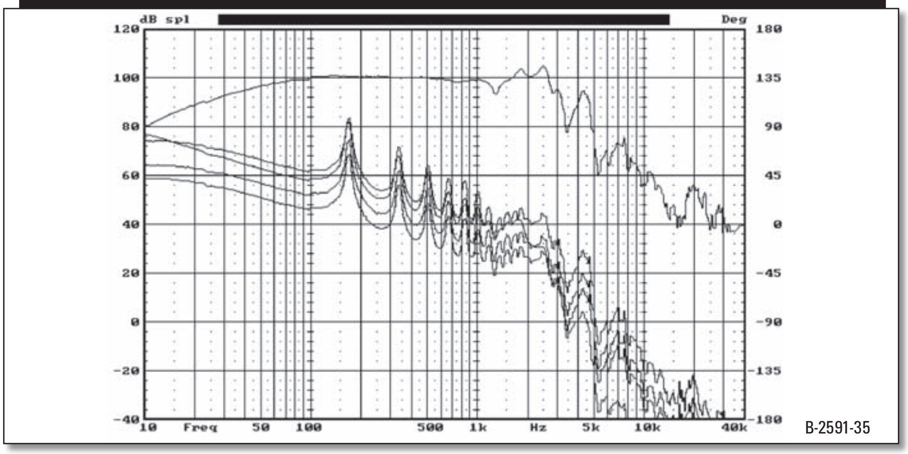
In Part 2, I'll look at additional panel-quieting measures such as constrained layer damping. *aX*

FIGURE 34: Stiffness variations.



B-2591-34

FIGURE 35: Thickness variations.



B-2591-35

VENDORS

Pro audio kits: mic pre, mixers, buffers, crossovers, power amps.
Also custom designs.
www.borbelyaudio.com
Selected Borbely Audio kits in Japan:
http://homepage3.nifty.com/sk-audio

WANTED

stereo gear old/new/quality, speakers, amps, preamps, turntables, reel to reels, vacuum tubes, tube testers, guitars, parts, etc. **850-314-0321**
Email: **sonnysound@aol.com**

AD INDEX

Yard Sale

WANTED

A crystal 8412 receiver and a Pacific Microsonics PMD-100 chip for a home-brew DAC, and Allen Bradley 120Ω 1W carbon composite resistors. Rick (406) 721-9463.

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

Back

PRODUCT REVIEW

Sound Pax Speaker System

By Dennis Colin

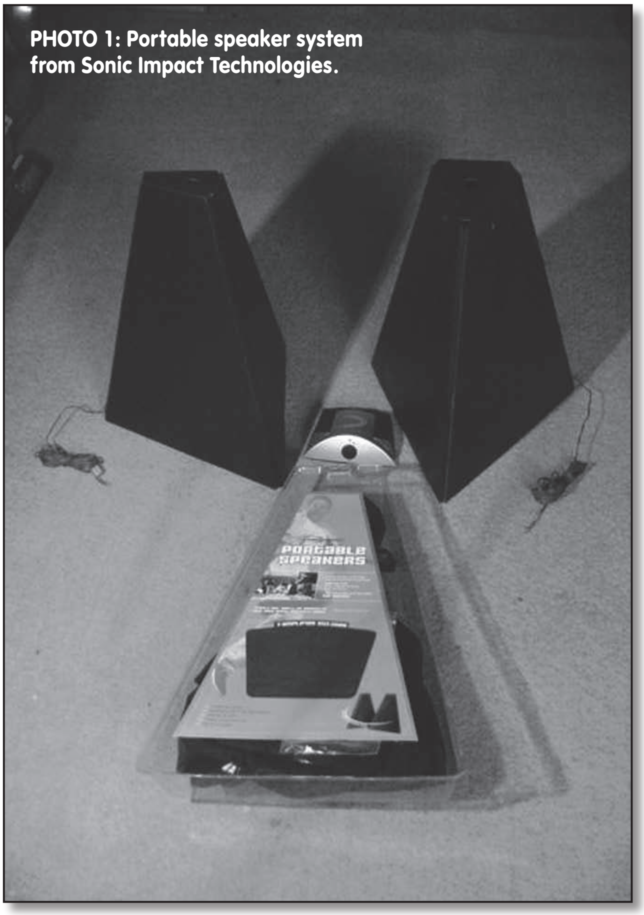


PHOTO 1: Portable speaker system from Sonic Impact Technologies.

What kind of sound would you expect from a pair of 1 lb. 19½″ tall cardboard speakers driven by a 14 oz. battery-powered amp, the combination costing $69? And what would you think of this amp (6.5 oz. without the batteries) claiming 11W per channel 4Ω, 6W per channel 8Ω, at 0.1% THD, true continuous average power? Finally, how about a piece of cardboard driven by a single acoustic "exciter" motor, whose specifications include "sensitivity: high," and "power handling: excellent. Loud enough to dance to."?

I remember what kind of sound *I* expected. Then I connected the system to a CD player and pressed "play." No, I didn't observe 30Hz bass knocking flowerpots off tables! But surprisingly, I also didn't observe any unpleasantness to the sound—no serious resonances, no cardboard rattling, and so on.

What I *do* hear from this $69 wonder is genuinely astonishing—reasonably balanced response I'd estimate to cover 80Hz to at least 15kHz, natural voice reproduction, smooth highs, good image focus coherence, and "loud enough to dance to," perhaps 90dB SPL. When I then listened to my "audiophile" system (Swans M1 with Scan-Speak subs, "Mad Katy" 125W per channel tube amp), I not surprisingly thought of the Sound Pax system as "not audiophile grade." But when just listening to the Sound Pax, I could forget the "audiophile" stuff and enjoy the music, be it a first-rate classical SACD such as *Stereophile*'s K622 (Mozart clarinet concerto with Antony Michaelson, Mu-

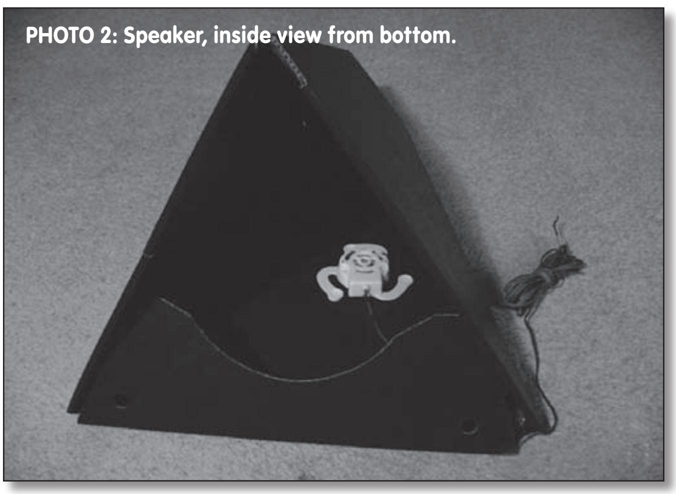


PHOTO 2: Speaker, inside view from bottom.

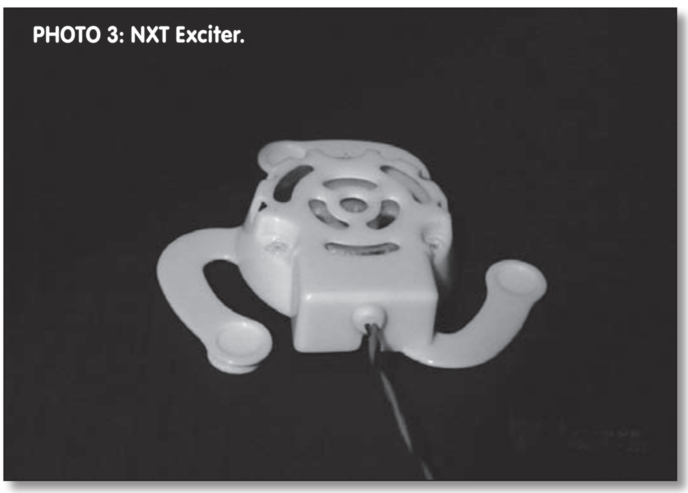


PHOTO 3: NXT Exciter.

sical Fidelity MF-SACD017) or driving electric rock such as "That's It" (David Raitt, Jimmy Thackery, Blu Rock' It BRCD132).

Comparing the two systems (grossly unfair: $3500 in *materials cost* vs. $69 *retail*!), the Sound Pax had less resolution, a mildly resonant lower midrange, and, of course, no deep bass and perhaps 90dB maximum SPL. But the sound was smooth, natural, undistorted, focused and spacious, and free of objectional coloration. "Computer speakers" I've heard at twice the price cannot even begin to compare.

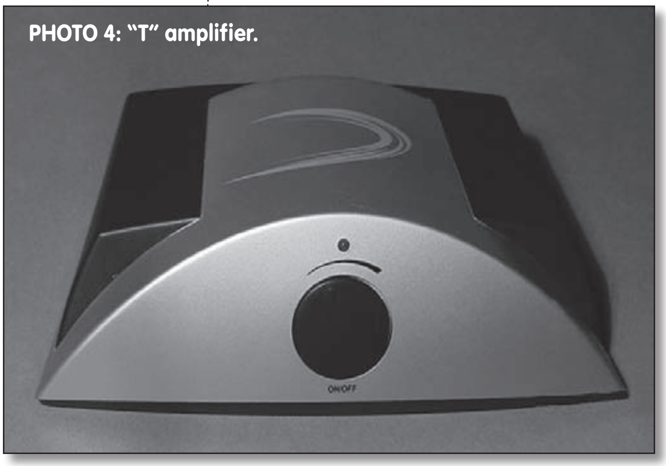


PHOTO 4: "T" amplifier.

PHOTO 5: "T" amp, rear view.

COMMENTS ON CARDBOARD

After some thought, I believe there's nothing inherently wrong with the idea of a cardboard speaker:

1. Cardboard is made of paper, which forms the conical diaphragm of many a fine conventional driver. Paper has a high stiffness-to-mass ratio, and can be well-damped.
2. I've made many cardboard prototype speaker enclosures for midrange and tweeter drivers. Often they were less resonant than the dimensionally-copied wooden enclosures, above the (200Hz or so) cardboard resonance limit.
3. The high-tech-sounding term "constrained-layer damping," applied to multiple wooden (or other material) layer cabinet construction, has been touted as almost revolutionary. But this term also applies to good old-fashioned cardboard.

In the Sound Pax speaker, the cardboard sheet is both the driven diaphragm and the enclosure—no discontinuity there. Also, the triangular pyramid shape is naturally rigid, and the tapered width smoothes the distribution of resonances both in the vibrating material and from acoustic wave diffraction. And the off-center placement of the "exciter" motor further ensures a smooth distribution of whatever resonances escape the cardboard's intrinsic damping. Most important, all of this theorizing is supported by the pleasant, natural sound reproduction.

For $69, including an apparently good-quality 20W stereo amp, the performance-to-price ratio borders on the ridiculous (ridiculously good, that is)!

For $207 you can have five-channel surround plus one extra channel, whose amp can power a small subwoofer. This price, plus some old enclosed woofer you probably have collecting dust, buys you a complete 5.1 channel setup good for a small room or the beach!

An absurdly good and novel product—why didn't I think of it first?

FIGURE 1: Sound Pax speaker, frequency response. Speaker on floor of 2000ft³ room, mike 12″ on-axis. 1/6 octave smoothing. Response is ±3.5dB, 85Hz-25kHz.

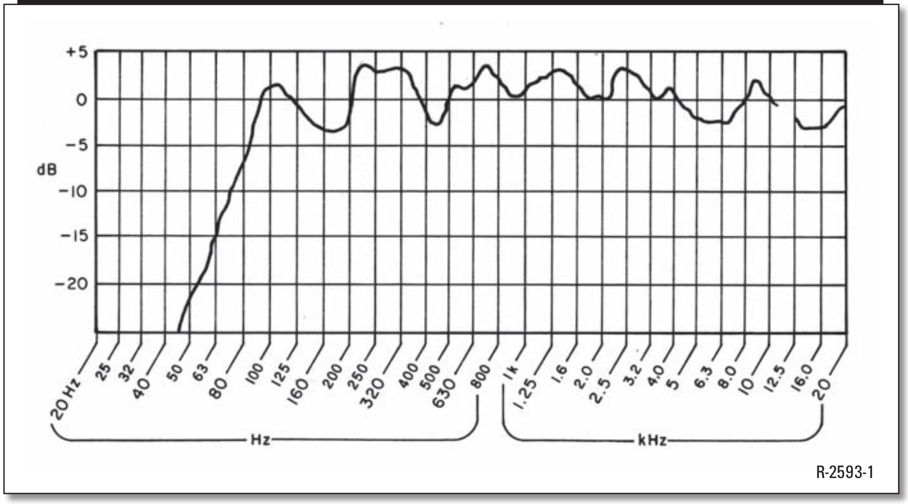
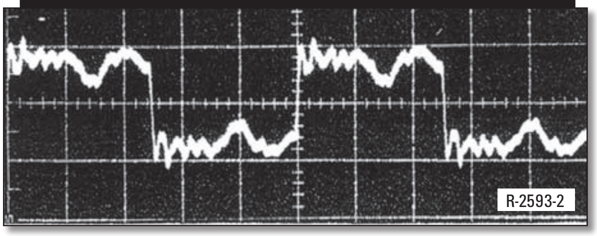


FIGURE 2: Sound Pax speaker, 400Hz square-wave response. Speaker on floor of 2000ft³ room, mike 5″ on-axis.



MEASUREMENTS
Sound Pax Speaker:

While certainly not as comprehensive and accurate as Joe D'Appolito's measurements usually included in *aX* reviews, my two close-miked in-room measurements used the Panasonic WM-60AY capsule (used in Mitey Mike II, typ. ±1dB 20Hz-20kHz). **Figure 1** shows the frequency response at 12″ on-axis with speaker on the floor (as it's likely to be located—the pyramid taper aims the direct sound upward). The response, ±3.5dB 85Hz-25kHz, would be good for a conventional speaker in a room—but for an "excited" sheet of cardboard, it's nothing short of astonishing. I'm more "excited" than the cardboard!

Figure 2 shows the 400Hz square-wave response at 5″ on-axis. The response is not only time-coherent, but is a cleaner square wave than that from most high-quality multi-way dynamic speaker systems. NXT (inventor of the flat-panel speaker and exciter products) certainly knows what it's doing!

"T-AMPLIFIER":

As **Table 1** shows, this little jewel works very well; I heard no "digital switching glare" (why would I at 1MHz switching frequency?) or other amplifier anomalies. 20W from a 6.5 oz. amp is 49W per pound.

The output has about 56mV RMS of

the 1MHz switching frequency, feeding a 12′ antenna (the supplied speaker wires). Although the differentially-driven twisted-pair configuration should suppress much of the 1MHz radiation, interference to a nearby AM radio should be likely. (If so, ferrite suppression cores around the amp end of the speaker wires should help.) A portable AM radio I tried received interference around 1MHz, but only if within about 10′ of the speakers, amp, or wires. *aX*

TABLE 1: MEASUREMENTS

Sound Pax Speaker

Response, 12″ on-axis, in-room, 1/6 octave smoothing: ±3.5dB, 85Hz–25kHz
Sensitivity: approximately 84dB SPL, 2.83V, 1m
DC resistance: Stereo pair, 9.1Ω, 8.4Ω
Dimensions: 19.5″ × 12.75″ × 12.25″
Weight: 15.5 oz.
Wires: 12′ twisted pair.

Sonic Impact "T-Amplifier"

With 10Ω load; eight fresh Duracell batteries (1.612V each, no load):
Power output: 8.1W per channel just before clipping
Sensitivity: 304mV RMS input with volume control maximum, for maximum output
Frequency response: -3dB at 28Hz, 63kHz
Switching frequency output: ≈1MHz, 56mV RMS
Output configuration: Differential (balanced) "H" Bridge, both terminals at ≈ 50% of DC supply (Do Not Ground!)
Dimensions: 6¼″ × 4″ × 2″
Weight: 6.5 oz., 14 oz. with batteries

TABLE 2

Sonic characteristics ratings	
Presence	8
Freedom from distortion*	8
Frequency response smoothness	7
L-M-H balance	8
Treble quality	8
Midrange quality	7
Bass quality	6
Bass extension	4
Immediacy and transient response	8
Image focus	9
Stereo soundstage realism	9
Ambience	8

** within limits of ≈ 10W overall, ≈ 1W bass*

A note on perspective regarding this scale:
1 = worst rock band PA system you've heard
3 = boom box
5 = mass-produced stereo speakers
7 = just acceptable for "high fidelity"
9.5 = Joe D'Appolito's "THOR" (*aX* May 2002)
10 = virtually perfect: limited only by recording/playback technology limits

FOR ORDERING INFORMATION CONTACT
SONIC IMPACT TECHNOLOGIES
www.si-technologies.com

XPRESS MAIL

A TINY VINTAGE GARAGE AMP

Could John Stewart post the schematic for his 3W AU6/6AQ5 garage amp ("An Affordable SE Triode Amp," *GA* 4/00)?

Jim Petersen
Springfield, Oreg.

John Stewart responds:

I built this little amplifier back in 1968. Since then it has been used in quite a wide variety of locations. For the past few years it has been in my workshop. Program sources are an Eico HFT-90 FM receiver and a matching Eico HFT-94 AM receiver.

I managed to build the entire amp, tone controls, and preamp into a Hammond 8 × 12 × 3 chassis. Most of the major components are on a 3 × 12 sub-chassis inside. That includes the power supply, power amp, preamp, and tone circuits. I assembled the sub-chassis on the bench. The controls, input, and output connections are mounted on the front and rear panels of the chassis. These were all wired together at the final assembly stage.

For more on this simple 3W amplifier, go to **www.audioXpress.com**. Once there, click on the current issue and follow the link title "articles and addenda."

Visit **www.audioXpress.com**

SE 211 AMP

I enjoyed the clever use of the interstage transformers in the "Push-Pull for Single-Ended 211 Amp" article by Lee Roitberg in the November *aX* (p. 30). I do have two comments.

On page 30 under the heading "The Light Goes On," Mr. Roitberg says "Amorphous core, nickel core, cobalt core—these are the words many an audio builder lust (sic) after."

AlliedSignal called the amorphous metal product line MetGlass® when they introduced it, and amorphous cores are available in many different alloys with square-loop or round-loop annealing, and grain-oriented and non-oriented strip configurations. Not all of these alloys are suitable for audio use. Nickel alloys have very low core losses and a wide operating frequency range, and are most desirable for low-level audio signal transformers. The relatively low peak saturation flux density of nickel alloys generally precludes their use in power and output transformers.

Cobalt alloys are not at all suitable for audio use, except as permanent magnet materials in loudspeakers (Alnico, samarium cobalt, and so on). Cobalt electromagnetic alloys, typically the permendur® alloys, are characterized by a 30% higher peak flux density than conventional 3% silicon iron. However, the core losses are also proportionally higher when utilizing this Bmax advantage. It is brittle and difficult to punch into laminations, and special annealing is needed to achieve the optimum properties. Cobalt is also a very expensive metal, at over $100/lb as a core material. It is rarely used outside the aerospace industry, where the need for low weight and the ability to withstand high operating temperatures mandates its use in transformers and generators.

On page 33 under the heading "Ring. Ring. Ring!," Mr. Roitberg says the output "peaked at +4dB at 14.8kHz, then dropped like a rock." The vertical axis of the upper graph in Figure 3 on page 34 is in output V RMS rather than the customary dB used in a frequency response curve. While the output peaks at about 4V RMS relative to 1kHz (I read 11.8V RMS), this is only a 3.38dB increase. If the peak were +4dB higher it would have reached 12.7V RMS. I only point this out on the chance that Mr. Roitberg had calculated dB but accidentally wrote V RMS on the vertical axis.

Chuck Hansen
Ocean, N.J.

Lee Roitberg responds:

Thank you, Mr. Hansen for taking the time to read my article and, more important, pro-

aX Online Exclusive

DISTORTED REFERENCES

By David J. Weinberg (Tobias Audio)

Two books published by Berklee Press ("a publishing activity of Berklee College of Music," **www.BerkleePress.com)**, one written by David Franz (engineer and Berklee College graduate student) and the other by Daniel M. Thompson (Berklee College of Music Assistant Chair of Music Production and Engineering), purport to provide enough information to enable effective setup and use of a home or professional recording studio. I give each a failing grade and hope they are not used in the classroom.

Recording and Producing in the Home Studio: A Complete Guide (©2004, ISBN 0-87639-048-3, 236 pages, $25) is Franz's attempt. Franz introduces the book as "for artists, songwriters, bands and aspiring engineers and producers who want to better understand how to engineer and produce their own recordings." This book is of limited use to a musician who is starting to produce and create his own recordings, and who wants an overview of what topics he needs to learn about, plus how they interact. It reads as though written by a musician whose ego led him to write, but in spite of his engineering degree seems without the depth of knowledge, or the ability to teach that knowledge, that is necessary to fully inform the budding musician/producer/engineer.

The most useful and informative part of the book are sections 2-4 (chapters 4-14), which address practical operational techniques useful in producing, recording, and mixing. Franz explains the role of the producer versus that of the engineer, acknowledging that in many home studios, the musician does it all. He suggests microphone selection and techniques, channel assignments, various traditional mixing board functions, how to take advantage of MIDI, editing techniques, and mastering. These seem to be based on his preferences with conventional mixing boards. He doesn't make clear that there are other ways to approach each task that might be better suited to the way other musicians work or think. The novice musician could become confused if he uses this book as a tutorial and his equipment doesn't offer the same capabilities. ■

To Continue

***Visit* www.audioXpress.com**

LARGE SIGNAL ANALYSIS OF DIFFERENTIAL TRIODE TUBE AMP

By Muhammad Taher Abuelma'atti

The author presents a mathematical model (basically a sine-series function) for the input/output voltage characteristic of a differential vacuum-tube triode amplifier. The results show that, similar to transistor-based differential amplifiers, the vacuum triode differential amplifier generates only odd-order harmonic and intermodulation components. The results also show that the amplitudes of these components are strongly dependent on the amount of cathode feedback and the amplitudes of the input tones. Moreover, the third-order intermodulation component is dominant and is higher than the third-harmonic component by about 12dB over a wide range of the input voltage. ■

To Continue

***Visit* www.audioXpress.com**

viding some feedback.

Regarding your first comment, you bring up a very interesting point on different core materials and the reality that some of them are not usable for audio applications, but I fail to see how this has any direct relevance to my article considering that I do not discuss the merits of any of the core materials or their intended use. While the science of core material can be fascinating, this article is not the proper forum for that discussion. Not to diverge from the article further, but please take a look at **www.magnequest.com**. I believe you will find a few transformers made of both nickel and cobalt that are readily available for audio use.

Regarding your second comment, although it is customary to see frequency response curves displayed in dB, the visual impact of a 4dB spike would have been lost in the graph and some people may see the anomaly as minor. By presenting the graph in V RMS, an emphasis was placed on the spike. With the limited data points presented on the graph, the highest peak is only 3.3dB. In reality, the graph continues to rise between the two highest data points provided to just over 4dB.

The idea that you took the time and effort to calculate the peak from a 2.25″ by 2″ graph just amazes me. If I had been aware of your intentions, I would have included the exact 14.8kHz data point. I hope you use this tenacity when you build this amplifier.

HAPPY DIYER

I am executive editor of *IEEE Spectrum* magazine. I have had wonderful success building an amplifier highlighted in the June 2004 issue of *audioXpress* (Satoru Kobayashi's "KT88 Hybrid PP Stereo Power Amp").

I finished the amplifier about a week ago, and have been absolutely amazed and delighted with its performance. I have been testing it with a cheap, portable Sony CD player and a pair of 14-year-old, two-way B&W speakers, and even with this modest setup the amplifier's beautiful response and tone shines through. It is certainly a match for my no-expenses-spared 300B single-ended monoblock amps that I have been using for the past few years, and which cost me probably twice as much to build as did Mr. Satoru's KT88 push-pull.

I first contacted Mr. Satoru almost immediately after reading the article, and have been corresponding with him regularly ever since. I have found him to be incredibly patient, kind, helpful, and even witty. The smallest question or inquiry on my part resulted in an immediate reply—sometimes multiple replies—and often quite late at night. I often got the impression he was staying up late simply to make sure he had answered all my queries to my satisfaction. This kind of interest and assistance is pretty much unprecedented, in my experience.

During the project, there were very few parts I had to order from Japan. All were easily obtained, but perhaps more important, all could have been substituted for with US or European parts. The only exception was the beautifully designed chassis.

I had it made by Mr. Shibuya, at Mr. Satoru's suggestion.

Mr. Shibuya tailored the chassis for the RCA jacks, the binding posts, and the power-entry module I selected. Mr. Satoru forwarded all these details to Mr. Shibuya for me and ensured that Mr. Shibuya understood what I wanted, and even sent me a CAD-blueprint of the final chassis for my inspection prior to the work being done. In the end, I found that the KT88 chassis cost about one-sixth of what I had paid for the two less-complicated stainless-steel chassis that I had had made for my monoblocks. I wouldn't hesitate to order another chassis from Mr. Shibuya.

By the way, I recently started building the "High End Preamp for Today's Audiophile" (*audioXpress*, February and March '05), which I intend to use with Mr. Satoru's KT88 amp. The contrast between the two designers, Mr. Satoru and Mr. Ben Hinrichs, has been stark.

I went to Mr. Hinrichs' website and from there to a link where I bought a kit for this preamp. The kit came with no instructions and no schematic whatsoever. It was also lacking a few necessary connectors, which I had to hand-fashion myself (I could not make the German man who sells the kit understand what I was lacking). I wrote to Mr. Hinrichs twice with a few straightforward questions, and he declined to respond in both cases. Nevertheless, I'm confident I'll be successful in the end.

I'll close by saying you have another happy customer, and that I hope I can look forward to more articles by Mr. Satoru in the future. He has certainly made a believer out of me.

My compliments to you and your staff. Keep up the good work.

Glenn Zorpette
Sea Cliff, N.Y.

aX