

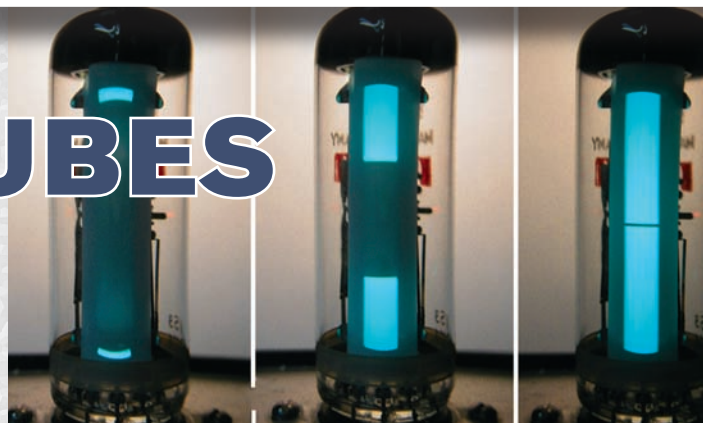
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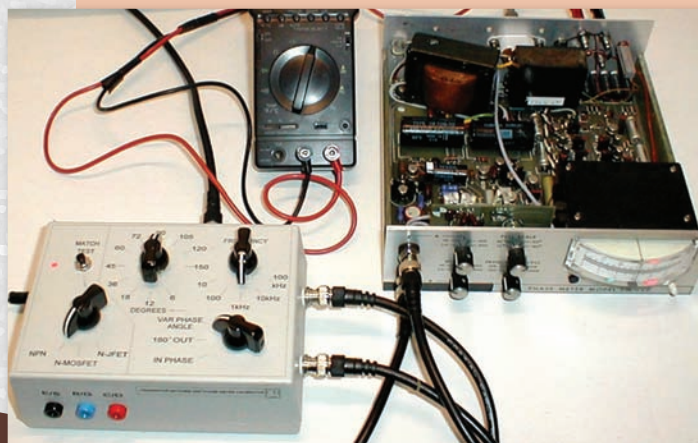


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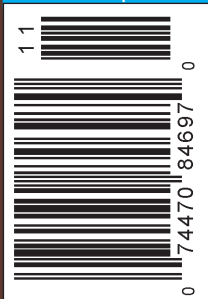
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Chuck Hansen was almost shivering with obvious delight as he wrote the following to the editor in a recent e-mail.

My lovely niece, a newly-minted nurse of 1 year and member of the iPod generation, stopped by to see us while I was checking out the Parasound Zamp. I sat her down to listen to my system. Once I explained what to listen for, she could easily pick out the difference in soundstaging, imaging, and frequency response between the Zamp, the Parasound HCA-100A, and my modified Scott 222C. She was quite fascinated with the Scott, since she had never seen or heard tube amplifiers before.

She went out to her car and got her Pearl Jam and Springsteen CDs and happily spent time listening to "her" music. I also played some vinyl for her—she had never really seen that up close before either, having been born about when the CD format came out. Her audio "systems" are her car and her iPod.

It's a real treat when watching as someone of her generation gets her first exposure to high-end audio.

We audiophiles tend to keep the great secret of splendid sound to ourselves. The same tendency seems to pervade the audio industry as well. It is, I believe, an ill-founded form of false modesty—or laziness, or whatever. In audio's earliest days when we talked about high fidelity, there were a few who believed that "hearing is believing" and took the examples to the public.

G.A. Briggs, who started his career as a cloth salesman, delighted in live demonstrations of good sound. He manufactured some fine drivers which were elegant examples of British craftsmanship. He wrote extensively and entertainingly about hearing, audio, music, and sound systems. He brought salesmanship to his passion for good sound.

Peter Walker provided public demonstrations of good sound to the early audio shows of Britain. Significantly,

Walker played saxophone in a local band while building Acoustical Manufacturing and a series of components from amplifiers to electrostatic speakers which have rarely been equaled. Like Briggs, his craft grew out of his enthusiasm for sound, without the benefit of an engineering degree.

The people at Acoustic Research were the first to offer the public an accessible (if you happened to be in New York City) audition booth for your own choice of recorded music. It was surprising how well known that relatively small center became. I suppose the folks at AR had some idea of the traffic to the room over time. It took advantage of the acuity of the human ear to hear the difference in what they were hearing, compared to what was available at home.

These days the whole industry is lying awake nights with the MP3 specter roiling their imaginations. All except the people at Apple, of course. But I believe the time is long past for the audio industry to stir itself to make the obvious benefits of excellent sound available to the generality of the citizenry. Would ten well-managed audition rooms in selected cities of North America be beyond the means of the industry? The magic works.

My nephew visited me the other day. He manages a wire manufacturing facility nearby for a big multinational. He grinned rather sheepishly as he stepped into my listening room. "I have to admit," he said, "that almost all my listening these days is MP3." He was intrigued by my Thor speakers and the Pass amplifiers driving them. Then he sat listening intently for a few minutes. Finally he said, "Those are the best speakers I have ever heard in my life. I thought mine were pretty good, but these are much better."

I hope we will keep in mind that Apple's iPod sets a benchmark for people's ears that can be shellacked by five minutes of auditioning the real McCoy. Convenience is great, but not entirely nourishing, after all. —E.T.D.



Precise dB Monitoring with Eye Tubes

This circuit for use with eye tubes eliminates variations

for precise and accurate signal measurements.

By Joe Sousa

Tuning eye, or indicator, tubes came into use before World War II to help tune radio receivers by indicating signal strength. This was helpful to reduce distortion and adjacent channel interference that would result if the radio were tuned too far from the center of the channel. The eye tubes were developed as a cheaper alternative to the needle movement meters. It was not until the 1960s that needle meters were made economically enough in Japan to displace indicator tubes.

During the decades from the 1930s to the 1960s, when tuning indicator tubes were in production, they found wide use in instrumentation as a low-cost alternative to the needle meter. Capacitor measurement bridges and tube testers were among the more common instruments making use of these inexpensive indicator tubes.

By the 1950s, indicator tube design had matured enough that good precision could be obtained for use as recording level meters in open-reel tape recorders. The most accurate indicator tubes were the European EM84/6FG6, EM87, and similar tubes, which indicate signal strength with two green fluorescent rectangular sections on the cylindrical wall

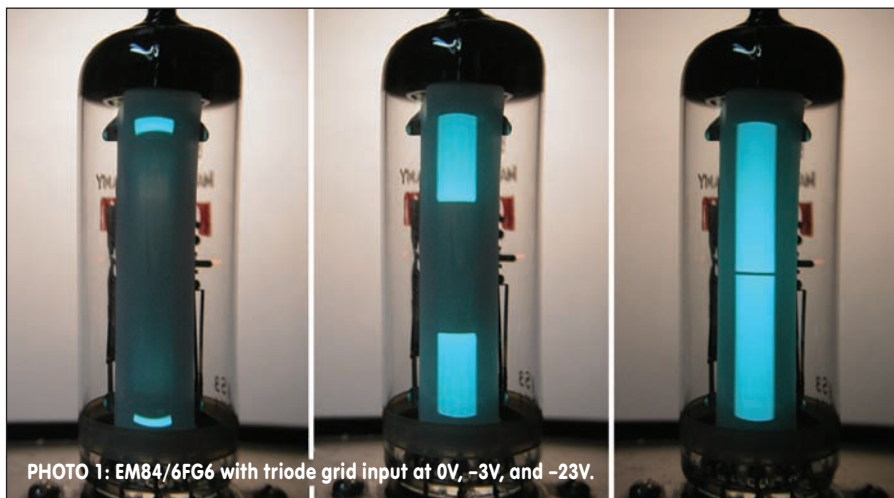


PHOTO 1: EM84/6FG6 with triode grid input at 0V, -3V, and -23V.

of the tube. The green rectangles grow longitudinally, as the input signal grows more negative (Photo 1). At full signal strength, the two rectangles meet in the center, forming a solid rectangle 1.5" long.

Figure 1 shows the internal arrangement of the EM 84 / 6 F G 6 tubes, which were built with two integrated sections sharing a common cathode. One section is a small triode that amplifies the 0 to -24V input swing into a 0 to +250V swing that, in turn, drives the output deflection node. The plate of the triode has a 470k load to +250V and drives the deflection node on the other section

to deflect the output beam on the fluorescent target painted on the side of the glass envelope. The 470k plate resistor

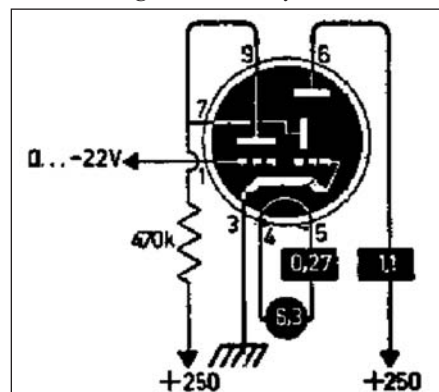


FIGURE 1: Standard wiring for EM84/6FG6 eye tube.

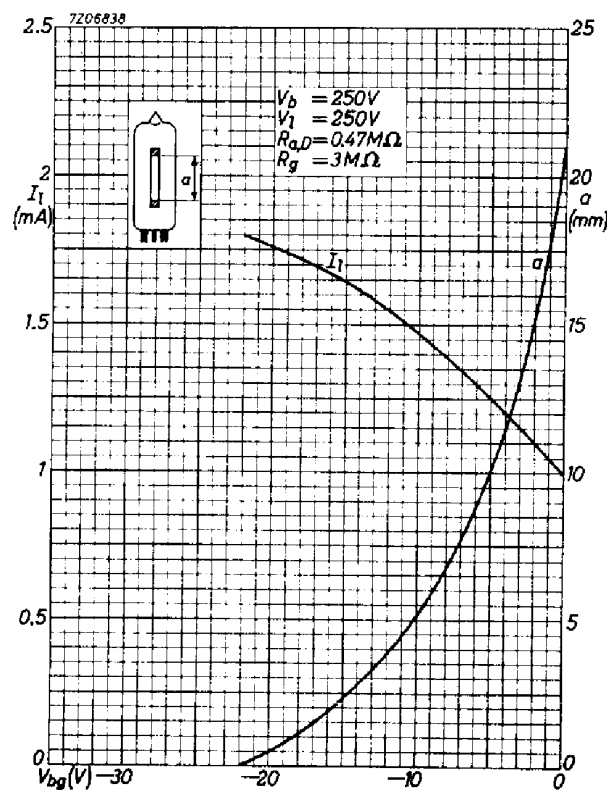


FIGURE 2: Transfer characteristics from input triode grid voltage to target deflection for a Philips EM84/6FG6 eye tube.

sets the amplitude of this swing.

The supply voltage also affects the amplitude of the target deflection. Reducing B+ from 250V down to 200V to save wear on the phosphor also increases deflection. Some data sheets show curves for reduced supply operation.

The data sheets from various manufacturers show a carefully designed transfer curve from the input voltage at the grid of the internal triode to the size of the gap between the two glowing rectangles. This curve (**Fig. 2**) follows a logarithmic characteristic very closely, covering 30dB of input voltage range. This accurate decibel response and the large bright display made these tubes suitable for use as fast, precise indicators of recording levels in tape recorders of the 1950s and 1960s. The logarithmic characteristic was also beneficial for use in radio tuning service, as it permitted extended indication of tuning at lower signal levels.

PRODUCING DECIBEL STEPS

When I discovered this logarithmic transfer curve, I was so impressed that I

decided to test it. I built a circuit to feed a repetitive series of negative voltage steps staggered in dB increments to the input grid of the EM84. **Figure 2** shows these steps in 6dB increments, as applied to the input of the EM84.

For this article, I chose three selectable dB step settings of 3dB, 6dB, and 10dB. This sequence of fixed dB steps

refreshes at a 250Hz rate and appears on the fluorescent target as a superposition of evenly spaced fluorescent bars of decreasing lengths.

You can make several adjustments to have this graduated pattern fill the fluorescent target exactly. First, you can adjust the 250V supply voltage or the value of the 470k pull-up resistor so that

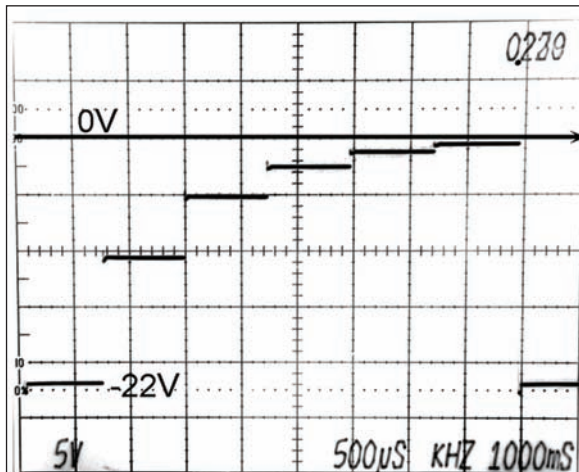
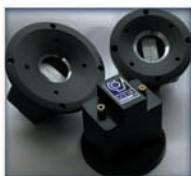
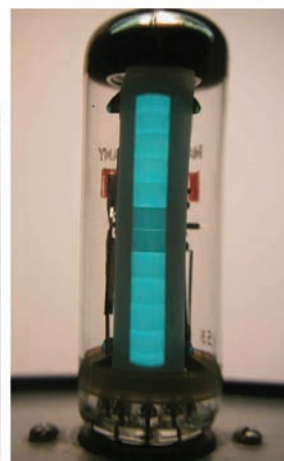


FIGURE 3: Exponential negative voltage step sequence in 6dB steps to drive input triode grid of EM84. Each step has 50% of the voltage amplitude of the previous step. Photo: EM84 6dB steps 30dB range.



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the two fluorescent bars meet exactly at the center when the maximum negative level of -22V or more is achieved and the input triode is cut off. You can adjust the 470k pull-up from 200k to 1Meg to make the rectangles overlap slightly in a bright line at the center, or you can leave a slight gap, to suit your preference. Compare the center overlap in **Fig. 3b** with the gap in **Fig. 6b**.

You can also adjust the start of the deflection to open up the deflection pattern, so that the target is nearly completely blank when there is no input signal. A small adjustable DC offset voltage is added with pot U11 in **Fig. 4** to the input drive to the grid, and is varied between -1V and $+1\text{V}$ to nearly extinguish the minimum deflection pattern when there is no input signal.

The last adjustment to be made to accommodate variations from tube to tube, supply voltage variations, and tube aging is the maximum deflection at the center of the target. Here, you simply need to adjust the gain of the circuit driving the input grid with U6 pot in **Fig. 4** so that the maximum level of the dB step sequence makes the step pattern close precisely at the center.

Once the tube is calibrated for the starting and ending points of the pattern, you can configure the step size. You can calibrate the size of the dB step for any step size inside the 30dB range of the tube fluorescent target. My three switchable step sizes of 3dB , 6dB , and 10dB are selected by a three-position selector switch on the left side of the schematic diagram in **Fig. 4**, and are calibrated individually with an oscilloscope.

A fixed step dB sequence is an exponential sequence, in which each step is a fixed percentage of the previous step. The circuit in **Fig. 4** generates an exponential step sequence to drive the input grid of the tuning eye tube. I used a simple method to obtain this sequence by sampling the decaying voltage of a precharged RC network.

The 555 timer U1 tells the LF398 sample-and-hold IC to sample the continuous exponential decay of C3 at fixed intervals, every few

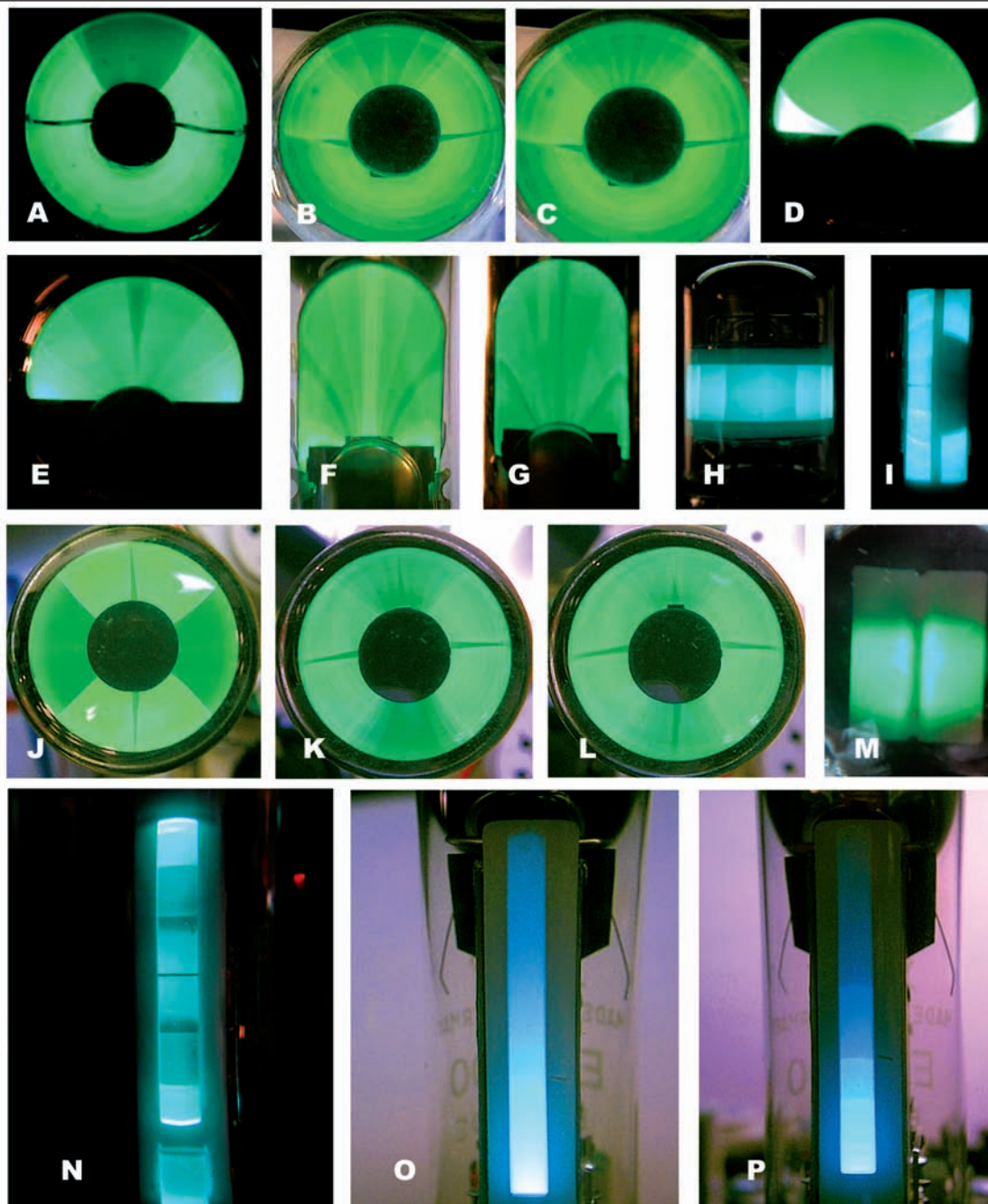


PHOTO 2: Displaying dB steps on different phosphor targets. a: 6U5/6E5 input grounded. b: 6U5/6E5 with 6dB steps. c: 6U5/6E5 with 3dB steps. d: EM71 input grounded. e: EM71 with 10dB steps. f: EM80 with 6dB steps. g: EM81 with 6dB steps. h: EAM86 with 10dB steps. i: IM801 with 10dB steps on one input and a floating input on the other. j: UM4 input grounded. k: UM4 15dB in 6dB steps. l: UM4 30dB in 6dB steps. m: 6AL7 with grounded inputs is too blurry for step display. n: EMM803 10dB steps, with phosphor burn. o: EM800 with fixed -10V input. p: EM800 linear display with 6dB steps.

dB with each new sample, and are adjusted as a group, simply by adjusting the



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sample rate, or frequency of 555 timer U1. The dB step selector switch simply connects different adjustable timing resistors to this 555 timer U1.

The remaining task in step generation

is to recharge the capacitor periodically. The 555 timer U2 sends a $30\mu\text{s}$ pulse to recharge C3 after it has discharged to 3% or less of the initial charge (3% corresponds to a 30% decay). At the end

of the step sequence, the capacitor has discharged to approximately 150mV.

This charge decay is needed to cover the 30dB range of the EM84 tube. Ini-

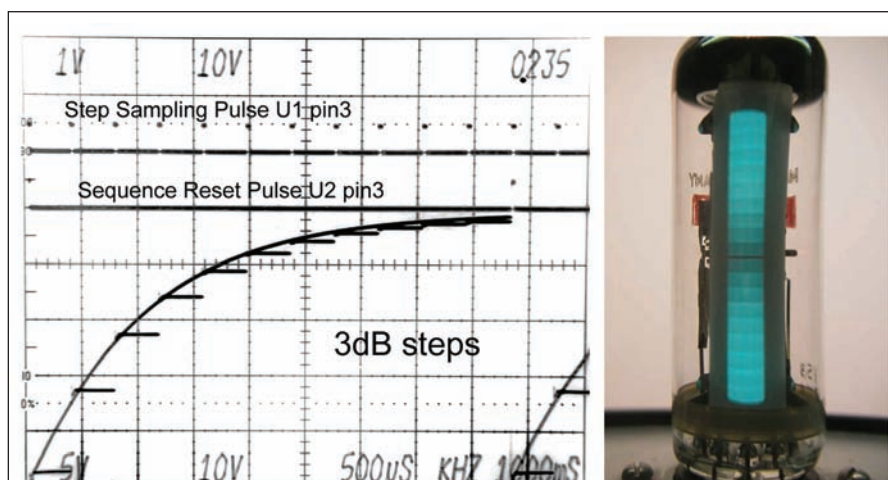


FIGURE 5: Two pulse trains sample and reset linear RC exponential decay to produce precise negative voltage steps in $\sim 3\text{dB}$ increments for EM84 triode grid input. Photo: EM84 with 3dB steps shows 30dB range.

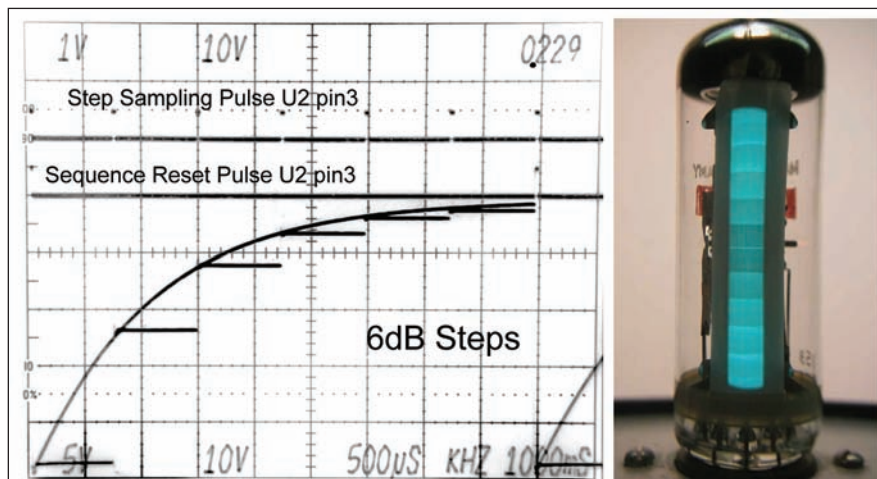


FIGURE 6: Slowing the step sampling pulse rate with a single adjustment yields 6dB steps. Photo: EM84 with 6dB steps shows 30dB range.

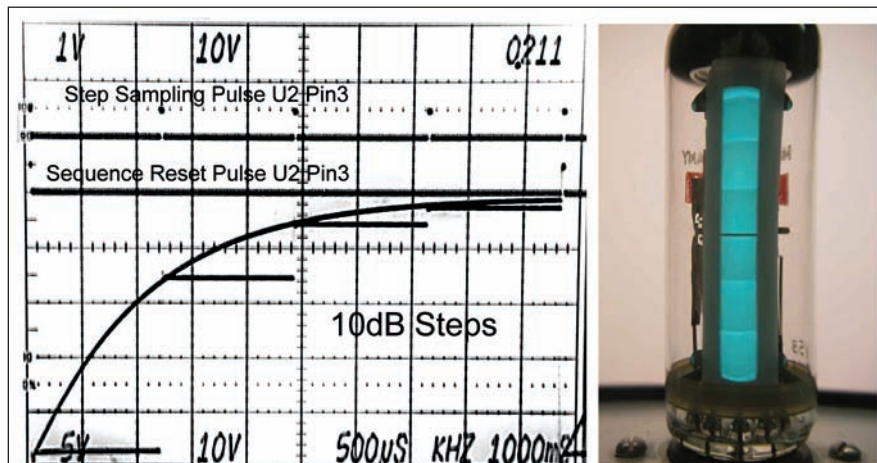


FIGURE 7: Three 10dB steps fill the 30dB deflection range of the EM84. Each step is only 31.6% of the previous step. Photo: EM84 with 10dB steps shows 30dB range.



PHOTO 3: Top view of dB step generator and VU meter. a) output offset control to adjust start of deflection. b) output gain control to adjust maximum deflection. c) VU display mode; stepped bars over blank background or smooth bars over stepped background. d) power switch and LED power-on indicator.



PHOTO 4: Step generator controls. a) input attenuator. b) calibration switch. c) decay: 6dB/sec or 6dB/100ms. d) dB step size. e) phono jack audio input. f) submini jack dB step output.

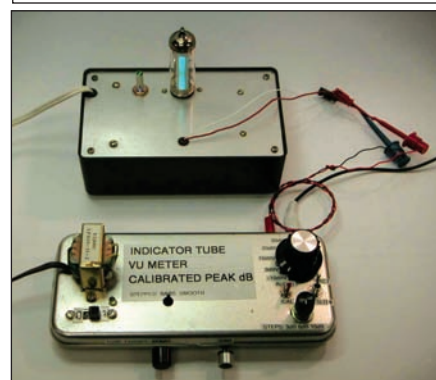


PHOTO 5: dB step generator and VU meter driving EM84 fixture.

tially, C3 is charged to 4.4V through D1. The two 555 timers are coupled with R5 to ensure that the step pattern and the discharge cycle remain synchronized.

The output of the LF398 sample-and-hold IC swings from -4.8V toward 0V. The LT1057 op amp U5 with Q1 and Q2 form a high voltage amplifier to extend the -4.8V step swing up to an adjustable maximum of -40V to drive the triode grid of the EM84.

Because of working with dB, the number of dB between steps will not change, while adjusting the overall step amplitude to fit the maximum input voltage range of the indicator tube. The negative input voltage needed to maximize deflection of the fluorescent pattern is specified at -22V for the EM84, and -10V for the pin-compatible EM87. In practice, these voltages can vary up to several volts, so adjustment of the input amplitude may be needed to get maximum deflection.

The dB selector switch selects between three preadjusted timing resistors of the first 555 timer U1 to sample the decaying exponential voltage on C3 more frequently for small dB steps, or less frequently for large dB steps. Compare **Figs. 5-7**.

The following describes the calibration method for the dB step size:

For the 3dB step, adjust RU10 trim pot to get steps that are 70.7% of the voltage of the previous step. For the 6dB steps, adjust RU7 with a slower step rate to get each new step at 50% of the previous step, and for 10dB steps, slow down the step rate with RU8 to

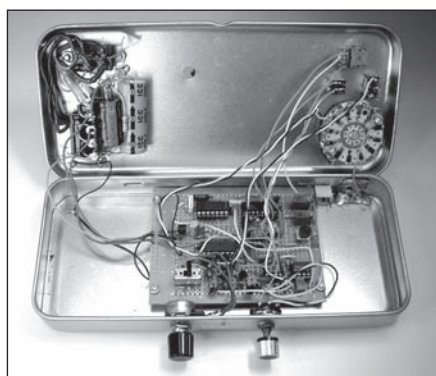


PHOTO 6: Interior of dB step generator and VU meter, with power supply on the left side of the lid and input attenuator on the right. Output offset knob on the lower left and the output gain knob on the lower right.

TABLE 1: DYNAMIC RANGE IN DB AND VOLTAGE RANGE FOR VARIOUS EYE TUBE TYPES.

Eye Tube	Target	Measured range in dB (specified range in V)
EM84/6FG6	2 rectangles	30dB, 0...-22V
EM87	2 rectangles	28dB, 0...-10V
UM4	2 sector round	30dB over 2 sections
EM81/6DA5	bathtub	15dB, 0...-15V
EM80/6BR5	bathtub	20dB, 0...-15V
EMM801	dual, 2 rectangles	30dB, 0...-20V
6U5/6G5	1 sector round	15dB, 0...-8V
1629	1 sector round	15dB, 0...-8V
EAM86	2 horizontal rectangles	22dB, 0...-8V
EM71	round fan sector	24dB, 0...-20V
EMM803	2 rectangles, 1 small square	30dB, 0...-15V linear -1V...-4V
EM800	1 rectangle	linear 0...-10V

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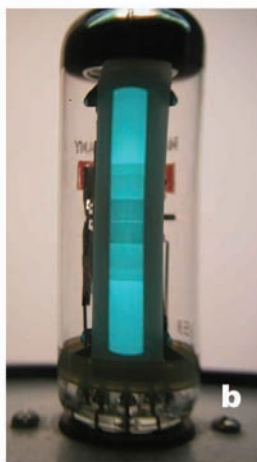
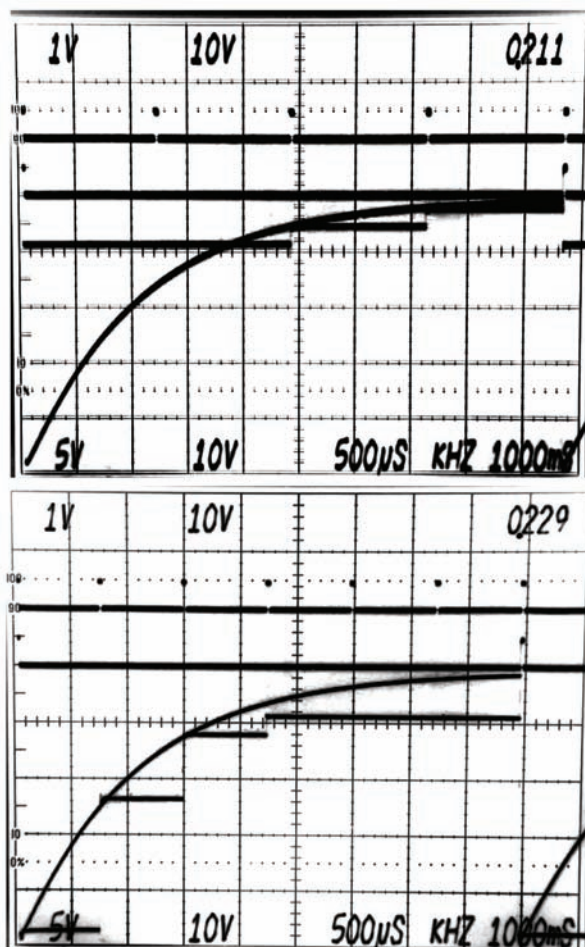


FIGURE 8: VU meter with 6dB graduated scale. Photos: Two display modes on EM84 with the same input: a) graduated bars over a dark background; b) smooth bars over a graduated background.

(Continued from p. 13)

obtain steps that are only 31.6% of the previous step. This is best done with an oscilloscope, but you can make these adjustments using the EM84 by applying known DC voltages with these dB ratios, and then adjusting the timing resistors for 555 timer U1 until the step size shows the same deflection as the known DC voltages.

DISPLAYING DB STEPS

The dB step generator is useful to check various indicator tube types, and to compare various tubes of the same type from different manufacturers or different levels of wear (**Photo 2**). Other tube types, such as the classic round eye 6U5/1629, the bathtub target EM80/EM81/EM85, or the off-center round fan deflection EM71, have a logarithmic deflection characteristic, but the smaller deflection area makes them harder to read. The dark portions of the concave targets tend to glow slightly.

As a result, concave metal targets offer a lower contrast than the phosphor targets that are painted on the glass. In this group of round eye tubes, the EM71 is the most suitable with its wide deflection area. The EAM86 has a rectangular side view pattern like the EM84, and the deflection is horizontal rather than vertical.

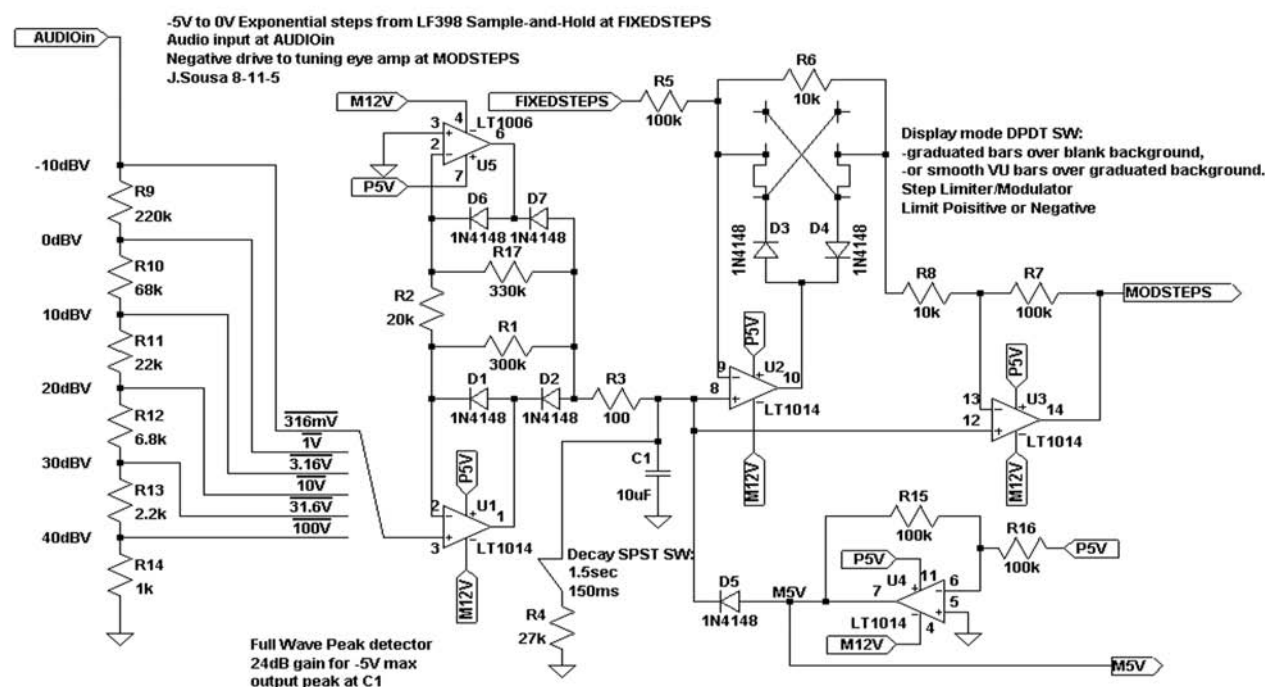


FIGURE 9: Schematic diagram for dB step modulator. AUDIOin is peak detected and limits the dB step sequence input at FIXED STEPS, to indicate VU level at MODSTEPS.

The EM801 is a variant of the EM84 that shows two identical channels on the same phosphor target. It has two rectangular patterns adjacent to each other, but they are narrower, slightly distorted, and are harder to read.

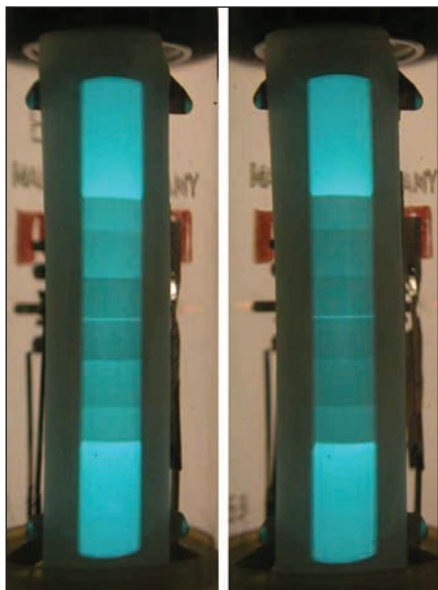


Photo 7: EM84 VU display. Increase U2 period to darken the background graduated steps.

The EM800 is the only eye tube I have found with a linear deflection scale, which makes it unsuitable for use with the dB step generator, as you can see in **Photo 2-p**. It is pin-compatible with the EM84, and the deflection pattern is only one single bar that covers a target area similar to the EM84. The EM800 requires a 200k plate load on the triode.

The EMM803 has two independent and dissimilar target deflection areas. The longer deflection area is logarithmic and very similar to the target of the EM84. The shorter deflection area is linear and extends no more than 1/4" just below the main target. You could use this small target area as a multiplex FM stereo indicator.

The 6AL7 tube requires about -10V to +20V applied to its deflection inputs, and it does not include a triode pre-amplifier. It also has a blurry beam and short deflection. The deflection response is fairly linear, which makes it unsuitable for use with the log step generator.

ADDING THE VU METER FUNCTION

After working with the dB step genera-

tor for a while, I decided to combine the precise steps of the step generator with a live VU signal to produce a graduated VU display on the target of the EM84. To do this, I added the graduated scale provided by the dB step generator to the VU signal in one of two selectable display modes as shown in **Fig. 8**. One display mode adds graduated dB steps to the otherwise smooth green rectangles.

The other display mode adds a background pattern of fixed steps, with the original smooth bars deflecting over this pattern to indicate signal level. The equal dB steps make it easy to accurately measure the strength of the signal in dB or volts.

The first mode, with stepped bars over a dark background, is generated when the peak-detected VU signal on C1 (**Fig. 10**) combines with precision dB steps in U2 and U3 to limit the *maximum* dB step excursion on the phosphor target. Alternatively, flipping the display mode switch and the VU signal on C1 limits the *minimum* dB step excursion, resulting in a display of smooth VU bars over a graduat-

ed dB step background. You can increase the contrast between the smooth bars and the graduated background steps by dimming the background steps slightly (**Photo 7**). Just increase the value of C2 to 0.01μF to lengthen the period of the second 555 timer U2 so that the background dB steps spend less time on the target.

The VU circuit in **Fig. 9** generates the VU signal with a peak detector and a calibrated attenuator to obtain full-scale ranges of -10dBV = 316mV peak,

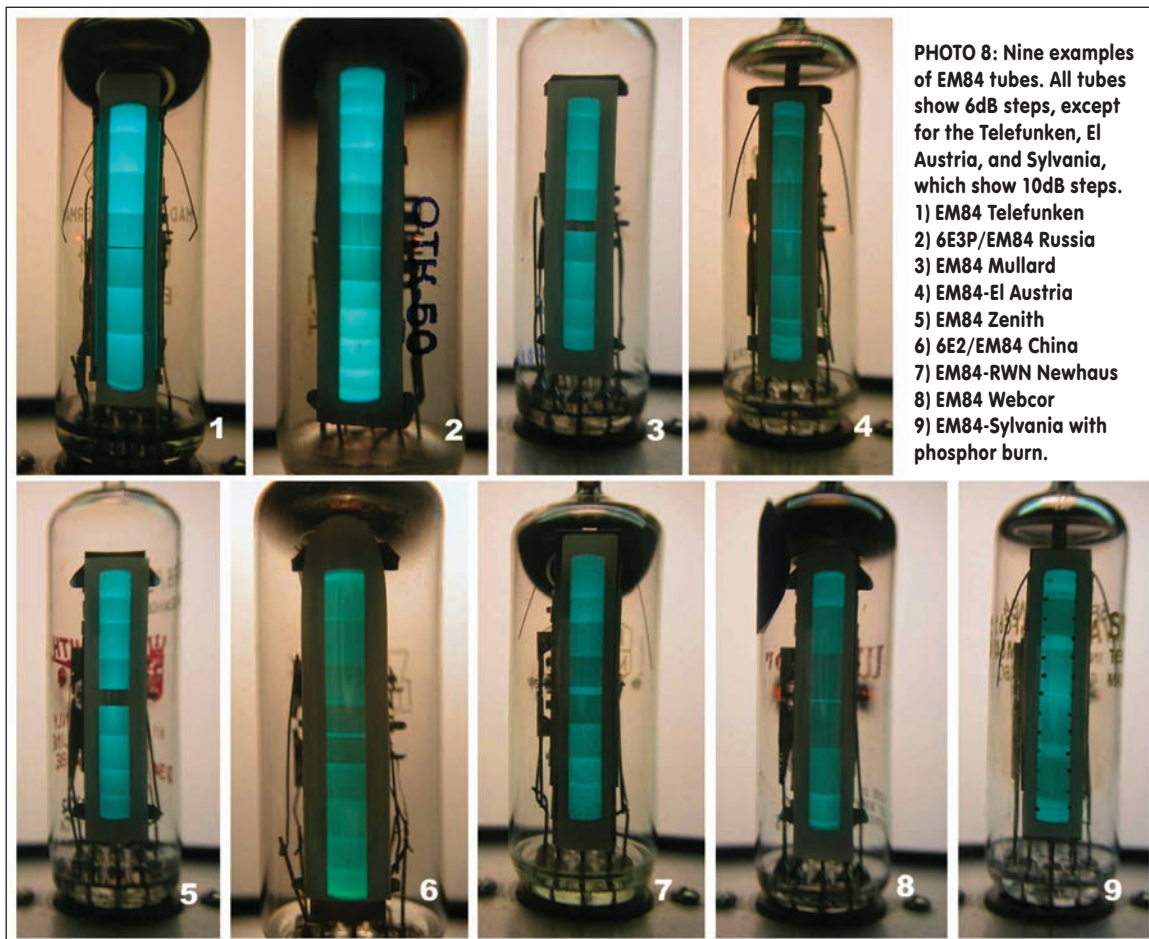


PHOTO 8: Nine examples of EM84 tubes. All tubes show 6dB steps, except for the Telefunken, El Austria, and Sylvania, which show 10dB steps.

- 1) EM84 Telefunken
- 2) 6E3P/EM84 Russia
- 3) EM84 Mullard
- 4) EM84-El Austria
- 5) EM84 Zenith
- 6) 6E2/EM84 China
- 7) EM84-RWN Newhaus
- 8) EM84 Webcor
- 9) EM84-Sylvania with phosphor burn.

0dBV = 1V peak, +10dBV = 3.16V peak, +20dBV = 10V peak, +30dBV = 31.6V peak, and +40dBV = 100V peak. If you want to use this circuit to monitor maximum power at speaker terminals, you can monitor the power of amplifiers ranging from 12mW peak into 8Ω with the -10dBV range, as may be the case with a headphone signal, to 1250W peak into 8Ω with the +40dBV scale. You can also set the decay rate of the VU detector with R4 to 6dB/sec for careful monitoring of maximum swings, or 6dB/150ms to gauge the dynamic behavior of the sound material.

USING THE METER

I found the 6dB step setting most useful to monitor dynamic content in music. The higher resolution of the 3dB step is best suited to check frequency response in an amplifier or speaker system. The 10dB step setting has been handy to check the total dynamic range of EM84 tubes.

One possible setting that can be useful in recording or precise monitoring of maximum signal levels is to adjust step size to obtain just one or two steps for comparison with the incoming VU signal. In the case of a recording level meter, you accomplish this by adjusting the indicator gain to obtain maximum deflection for a +3dB maximum record level. Then, you could adjust the next step to 70.7% of this level for 0dB.

Finally, lower the value of R4 to shorten the period of the second 555 timer U2 to shorten the step sequence to just these two steps. This way, you can see very precisely when the 0dB level is exceeded, without any parallax error caused by external bezel markings. In this application, the preferred mode is the smooth VU bars over a graduated background.

Another advantage of combining dB step levels with the signal voltage is that the VU voltage will always be accurately measured with respect to the dB step graduation, even with a change of tubes, or when supply variations affect the deflection factor of the tube. When you change tubes, or as the tube ages, you simply readjust the high voltage amplifier gain to make the green bars meet precisely in the center and readjust the offset control to minimize deflec-

tion with zero input and good dB step conformity at the ends of the deflection target. These adjustments will not affect the relationship between the VU signal and the graduated dB steps, so that measurement accuracy is preserved.

The nature of dB measurements is such that two stepped dB VU meters could be driven by the same audio signal with their two input attenuators set 30dB apart, such that the combined dB measurement range of the two tubes would be doubled from 30dB to 60dB. For example, if one VU meter is set to 0dBV, it will indicate sound peaks from -30dBV up to 0dBV, then another unit could measure the same audio signal with its input attenuator set to +30dBV, so that it can monitor signals from 0dBV to +30dBV.

The VU meter circuit in **Fig. 9** includes D5 to limit the output drive when the input is overdriven, but the -10dBV range should not see input signals greater than 5V to avoid damage to the input stage. This expanded dynamic range of 60dB using two eye tubes is particularly useful to monitor classical

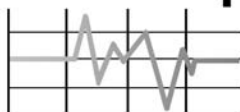
music, in which the quiet passages are often more than 30dB quieter than the loudest musical peaks.

NOT ALL EM84 TUBES ARE CREATED EQUAL

There is a great variation in brightness and phosphor smoothness. All tubes are driven with 6dB steps, except Telefunken, EI-Austria, and Sylvania, which are driven with 10dB steps. You can eliminate the gaps at the center of some of these tubes with a slight increase in input voltage amplitude. You could even out the start of the pattern at the ends of the phosphor target on the EIAustria tube with the offset adjust knob. All tubes were able to display the full 30dB range with very good accuracy.

The quality and smoothness of the phosphor target makes some tubes better than others. The brightest ones were RCA, Telefunken, and Russian tubes, which had a somewhat rough target, but were still quite usable. Some house brand tubes such as the RWN-NEUHAUS and WEBCOR, were barely usable, with a dim and rough phosphor target. *ax*

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The Venue Loudspeaker

The author maintains that, with this loudspeaker, you hear instruments as the solid 3D objects that they are, in precise focus, surrounded by a spherical “night sky” shell of natural venue ambience.

By Dennis Colin

Some of you might remember the so-called “3D” comic books of the 1950s, with red and blue stereo glasses. I say “so-called 3D” because the images weren’t rounded and solid, but rather like flat sheets of cardboard. The only thing 3D about these images was the depth separation between the “cardboard cutouts” and the equally flat background.

So it is with many conventional (forward-only radiating) speakers. They may have the flattest frequency response, lowest distortion, and most precise imaging this side of Mars; instrument sounds could be completely realistic. But those images still sound physically flat, like those “3D” comic book “cardboard sheets.” Like sitting inside giant headphones.



PHOTO 1: The Venue full-range bipolar speaker.

DESIGN PHILOSOPHY

The Venue (**Photos 1 and 2**) could be described as a coherent, full-range bipolar speaker. It is *not* (and not intended to be) an omnidirectional radiator; such units spread too much playback room acoustics.

But in a room with reasonably good and well-damped acoustics, of at least 2000ft³ volume, the Venue’s controlled upward-angled full-spectrum rear output stimulates an optimum amount of well-dispersed room ambience, which accomplishes two things:

1. Recorded venue ambience (concert hall long-delay reverberation or intimate nightclub ambience) is heard coming from all directions, because

(a) it is

(b) the time-blended lack of transients doesn’t allow Haas (precedence) effect localizing of this recorded ambience at the speaker positions or their



PHOTO 2: The Venue features upward-angle rear output.

horizontal stereo image plane, whereas the moderate amount of this room stimulation doesn’t interfere with the direct (forward) image precision. Rather, the latter is reinforced, making the direct soundstage images sound 3D solid, as the following shows:

2. Images sound like real sound sources because of “ray tracing”: Since the rear speaker output is spatially and (reasonably) time coherent with the front output, the arrival angles of ambience reflections are consistent with the locations of the direct image soundstage positions. This is analogous to reflections and shadowing of sunlight reinforcing the sun’s position in the sky.

ADDITIONAL DESIGN FEATURES

1. High quality drivers: 10” Scan-Speak woofer in sealed enclosure, 5” Focal midranges (front and back) in sealed compartment, and 1” Morel soft-dome tweeters (front and back).

2. Close spacing of all five drivers with tapered enclosure top, approximating a coherent spherical radiator, but with controlled (yet wide) dispersion.

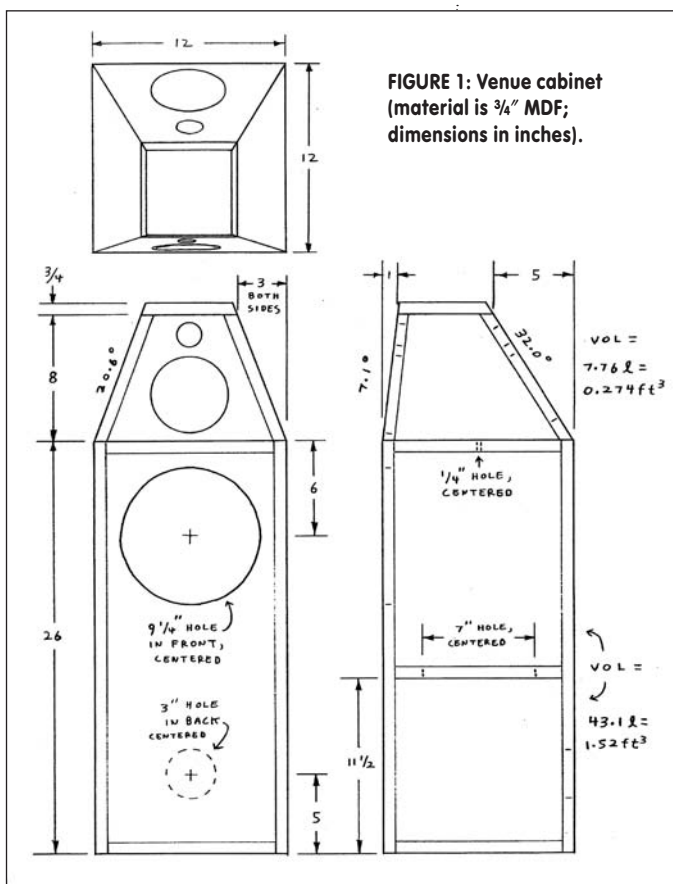
3. Angled and tapered cabinet top for smoothest dispersion and minimum diffraction.

4. First-order phase-preserving woofer/mid crossover (co) and quasi-third-order mid/tweeter co. Speaker transient response is time-coherent to within 200μs (**Fig. 18**).

5. Frequency response both anechoically and in a good room is smooth and with good bass extension (**Figs. 8 and 9**).

6. Power handling is high (about 200W typical music) and distortion is low even with deep bass.

7. Overall coherence: The woofer/mid co frequency is about 220Hz, meaning that the rear mid/tweeter combination rolls off (smooth first order) below



that frequency. However, below about 220Hz the woofer's output becomes nearly omnidirectional, reaching behind the enclosure to fill in where the rear midrange starts to roll off.

Thus, both the direct and rear outputs are full spectrum.

THE SOUND (IN A GOOD ROOM)

You hear images that not only have excellent tonal accuracy (my personally most important speaker criterion), but also sound very much like real musical instruments and voices—solid sound sources precisely located in their original ambient venue—hence the name.

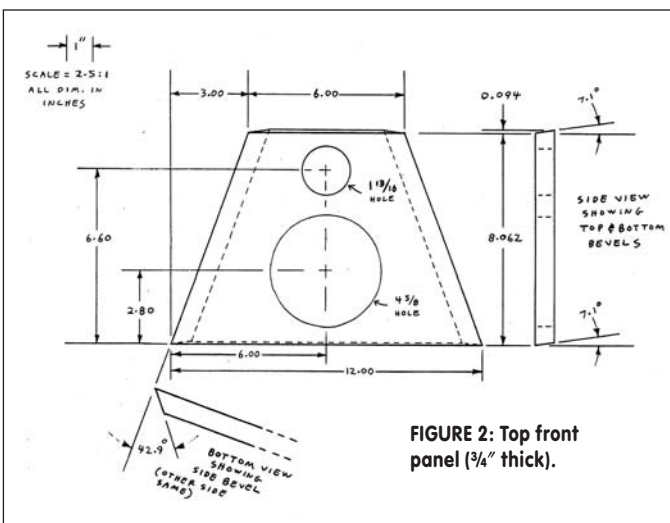
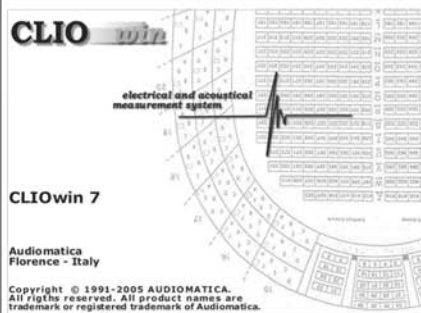


FIGURE 2: Top front panel (3/4" thick).

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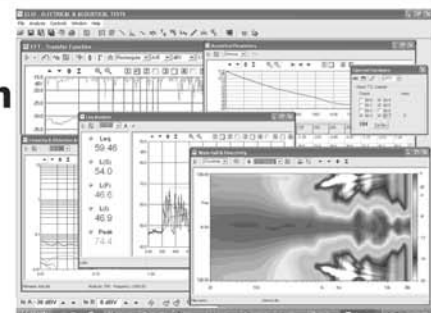


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ENCLOSURE DESIGN

I first built a full-size cardboard model, cutting the tapered pyramid top pieces by trial and error. The top angles are different—front, side, and rear; so I had hoped the cabinetmaker could figure out the bevel angles by experiment.

Wrong—he asked me to precisely calculate them! [expletive deleted], I thought!

So I spent a full day resurrecting my high school trigonometry memories, finally arriving at all the angles and dimensions. Some notes:

1. I recommend sealing all internal

wood joints with silicone rubber—particularly the top of the woofer enclosure, to prevent woofer back pressure from leaking into the midrange/tweeter compartment. You should test this by driving the woofer with about 30W at 20Hz (13.4V RMS across its $\approx 6\Omega$); feel the midrange cones to make sure they're barely (at most) moving.

2. Use 1 lb. of

Acousta-Stuf damping in the woofer compartment; pack as much as possible into the upper chamber.

3. When performing step 1's leakage test, also check for any signs of leakage to the outside—listen for any “puffing” sounds and feel around the entire enclosure for any air leakage.

4. For those who prefer flush-mounting the drivers, the countersink depths are 3mm (tweeter), 5mm (midrange), and 6mm (woofer).

Figures 1-5 show the detailed cabinet construction. **Figure 6** is a mounting template for the midrange and tweeter. Note: The scale values shown in **Figs. 1-6** are not correct; my original drawings have been reduced for publication.

CROSSOVER

Woofer = Scan-Speak 25W/8565-01,

43 ltr sealed

Midranges = 2—Focal 5W 4211,

7.7 ltr sealed

Tweeters = 2—Morel MDT40

Woofer: 1st order co (L1), zobel (C1, R1)

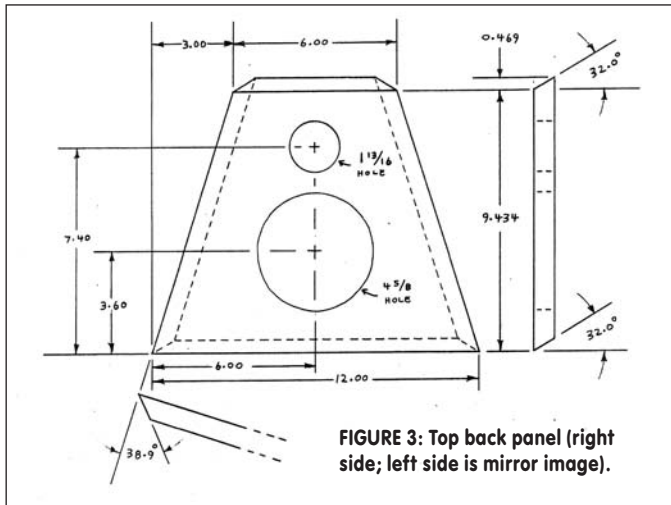


FIGURE 3: Top back panel (right side; left side is mirror image).

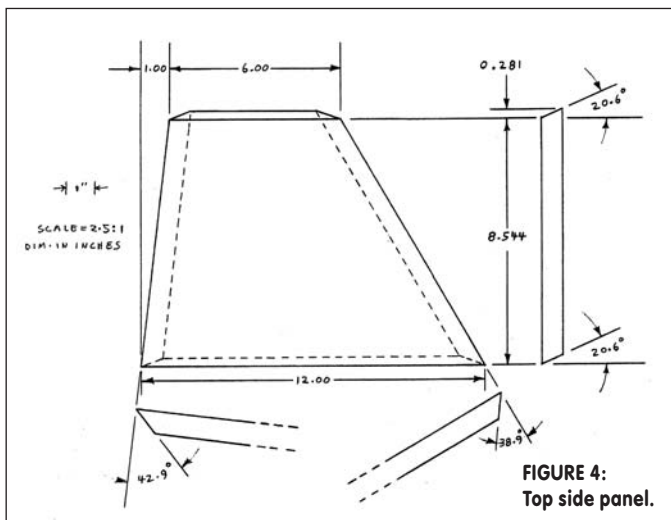


FIGURE 4: Top side panel.

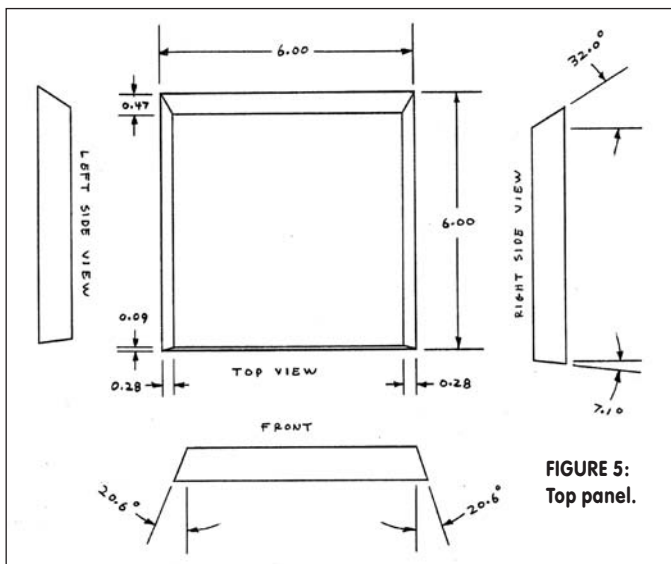


FIGURE 5: Top panel.

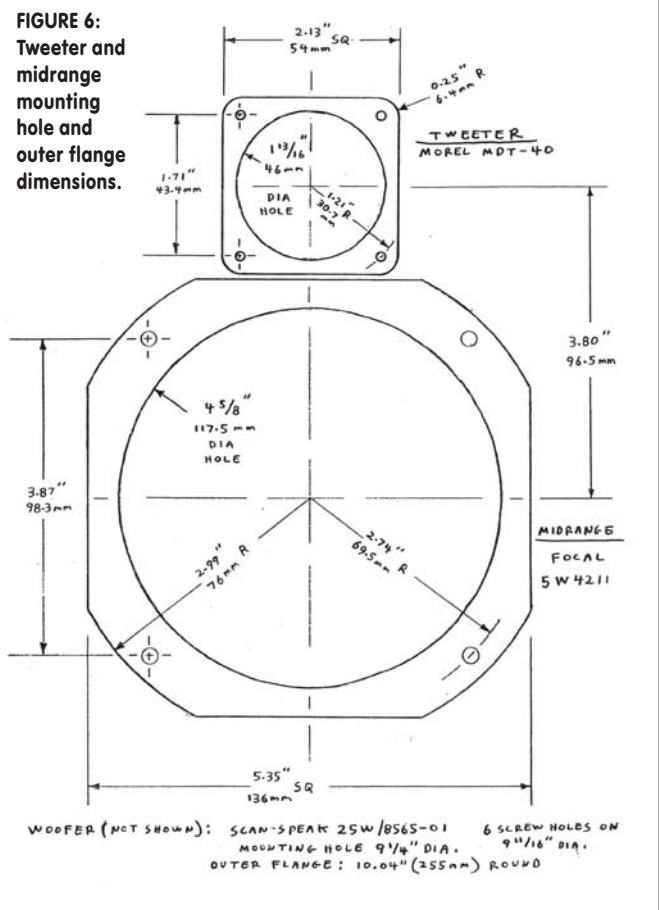


FIGURE 6: Tweeter and midrange mounting hole and outer flange dimensions.

- Midrange: 1. 1st order LF co (C2)
 2. Damping of 85Hz Z peak (R2, L2; also R4)
 3. 1.1kHz dip (L3, R3, C3);
 -5dB, half octave at -2.5dB
 4. 2nd order (elect) HF co (L4, C4)
 with R4 damping
 Acoustical slope $\approx$ 16dB/octave
- Tweeter: 1. 2nd order (elect) co (C5, L5);
 acoustic $\approx$ -16dB/octave
 2. 5dB attenuator (R5) with
 20kHz boost (L6, C6)
 3. Zobel (R6, C7)
 4. 6kHz dip, -2dB (R7, L7, C8)

- Notes: 1. Woofer/mid co at 220Hz,
 -2.5dB re acoustic sum
 2. Mid/tweeter co at 2.5kHz,
 -5dB re acoustic sum

MEASUREMENTS

Quasi-anechoic on-axis response at 4', mike 37" high, performed by Joe D'Appolito. Anechoic $f_{-3} = 37\text{Hz}$ (Fig. 8).

RTA in 2000ft³ live room (Fig. 9). For room photo and dimensions, see "Stereo Subs for the Swans M1," *aX* Sept. '05, Photo 2 and Fig. 1 (pp. 6, 8). Note that f_{-3} is 27Hz; 20Hz is only 7dB down.

Composite of three in-room measurements (Fig. 10).

Acoustic co responses (Fig. 11).

Electrical co responses (Fig. 12).

Impedance magnitude and phase (Fig. 13), courtesy of Joe D'Appolito.

Note that the impedance reaches lows of 2.77Ω at 400Hz at 19kHz, and is around 3Ω from 4.5-20kHz. This is unfriendly to many tube amps. With my Mad Katy 250W stereo tube amp (*aX* June '06), Ed Dell, Joe D'Appolito, and I noticed a slight "softening" of the treble when using the amp's 4Ω output

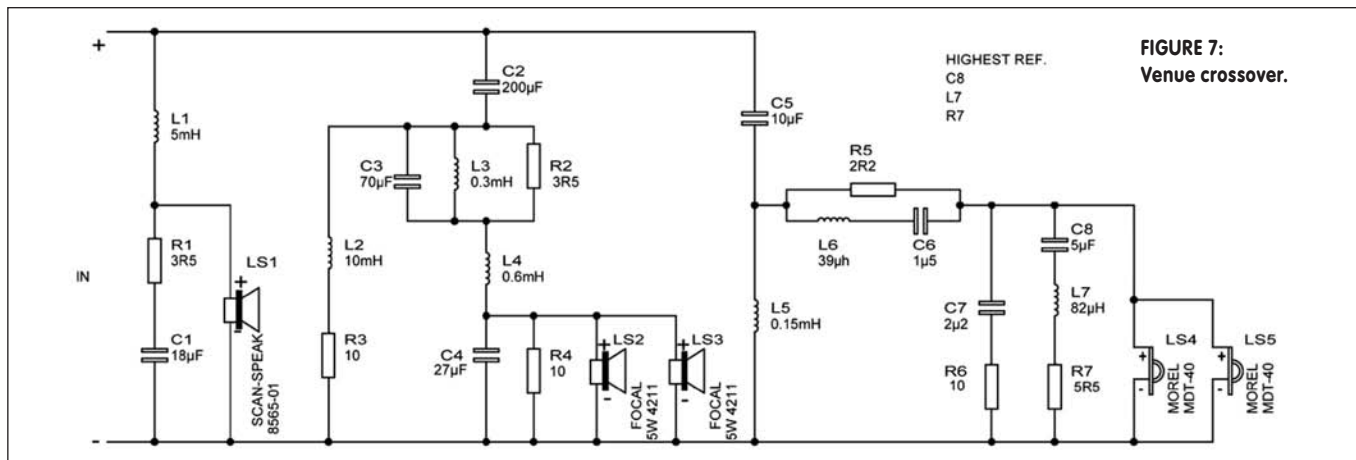


FIGURE 7:
Venue crossover.

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taps, when compared with a solid-state amp. When I later switched to the tube amp's 2Ω taps, I heard no such softening; everything sounded crystal clear,

including live-miked voices and a bell.

Excess group delay (Fig. 14), courtesy of Joe. Note that the peak is only 200μs, equivalent to a 2.7" acoustic path differential.

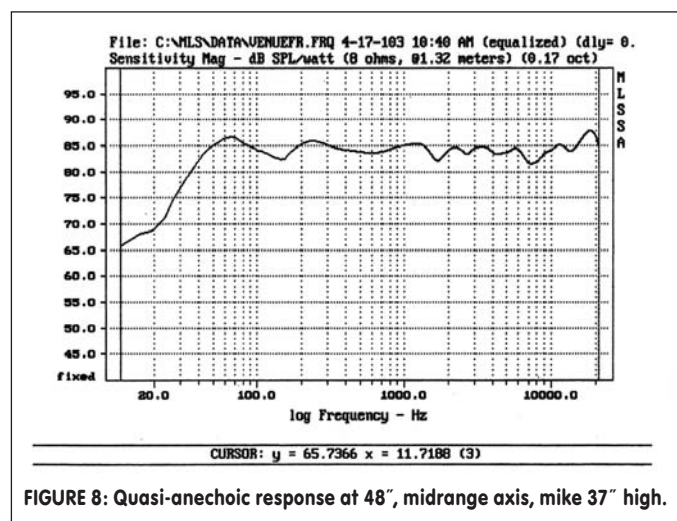


FIGURE 8: Quasi-anechoic response at 48", midrange axis, mike 37" high.

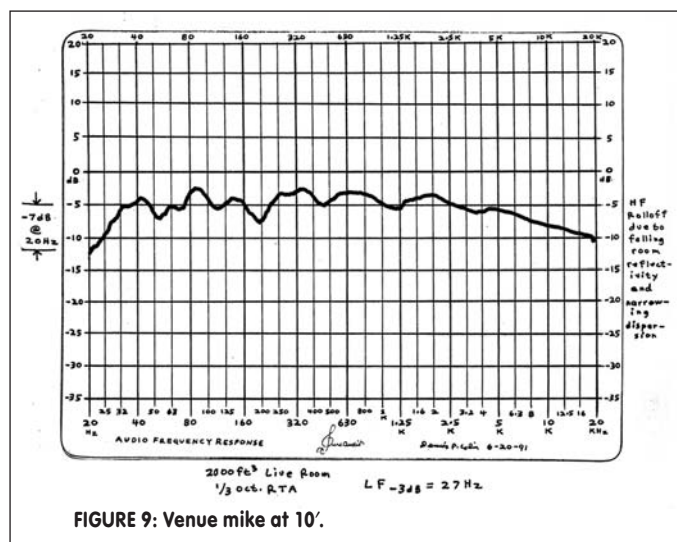


FIGURE 9: Venue mike at 10'.

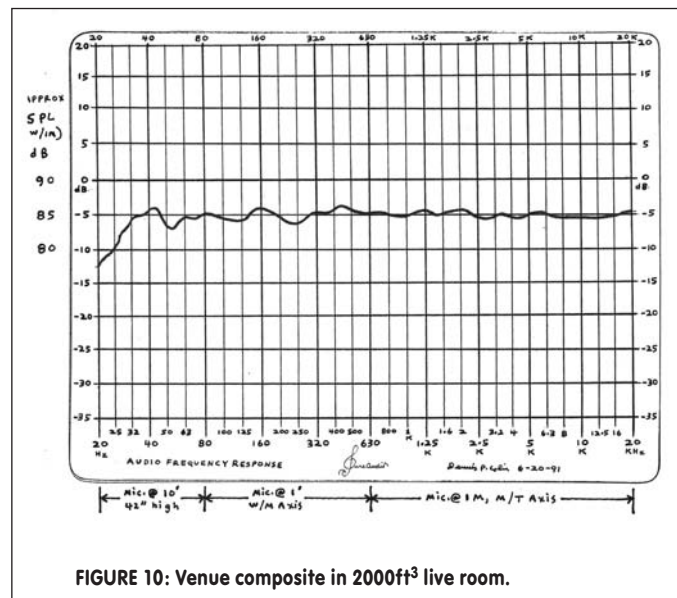


FIGURE 10: Venue composite in 2000ft³ live room.

OSCILLOSCOPE WAVEFORM PHOTOS

80Hz square-wave response at 1' on mid/tweeter axis; 2ms/div (Fig. 15). This shows the first-order woofer/midrange co's transient-preserving ability. You hear, for example, the sudden impact of a drum,

and a bass string's "pulsiness" and harmonic structure with precise timing.

3.1kHz linear triangle wave (Fig. 16). Admittedly, this frequency was "hand-picked" because the mid/tweeter co is not time-coherent to less than 200μs. But the linear response demonstrates the drivers' capacities for clean, linear response. 1' mid/tweeter axis, 0.1ms/div.

10Hz step response in room (Fig. 17). 1' on midrange axis, 2ms/div.

Expanded (0.2ms/div) step response (Fig. 18); 24" on mid/tweeter axis, 200μs/div.

The first impulse is the tweeter response. Its 10-90% rise time (not discernible on this figure) is 15μs, corresponding to a BW-3 of approximately 0.34g/15μs ≈ 23kHz. The time from

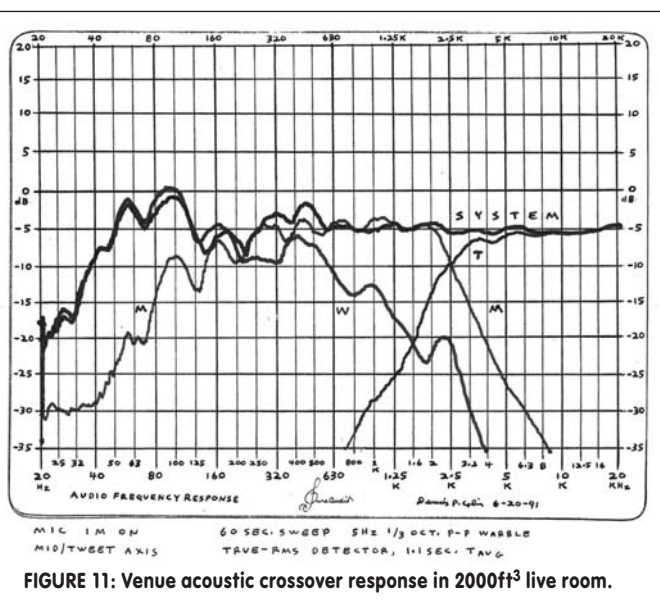


FIGURE 11: Venue acoustic crossover response in 2000ft³ live room.

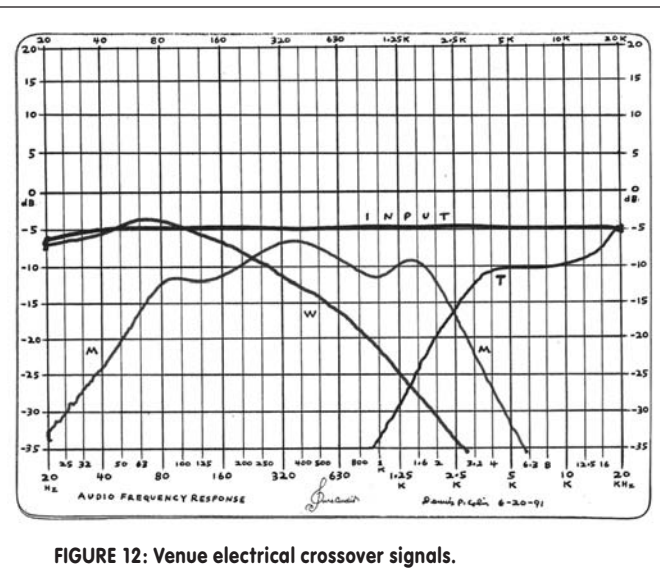


FIGURE 12: Venue electrical crossover signals.

this impulse peak to when the mike voltage again reaches that peak is about 200 μ s, agreeing with Joe's excess group delay measurement in **Fig. 14**.

After this first 200 μ s (typical third order co time response), the waveform reasonably (considering an in-room response) follows the exponential decay seen in **Fig. 17**, expected from the low frequency rolloff (second order at about 40Hz).

ROOM PLACEMENT

Note that the woofer center is 20" off the floor (for close woofer/midrange coherence). This causes a "floor dip" (from reflection cancelling) at about 169Hz. But you can avoid this by placing the speakers so that the woofer cone is twice that distance, or 40" from the room's front wall; position the enclosure back 28" from that wall. Then, at 169Hz the wall reflection will be a full wavelength

round trip, resulting in a reinforcement. In practice, this fairly well "cancels the cancellation" from the floor.

The speakers should be at least 28" from the side walls, and toed in to face a centered listener. However, with the Venue speakers' coherent wide-angle dispersion, the entire room (if good) is the "sweet spot."

In a five-channel surround (no ".1" sub needed), the sound is extremely lifelike,

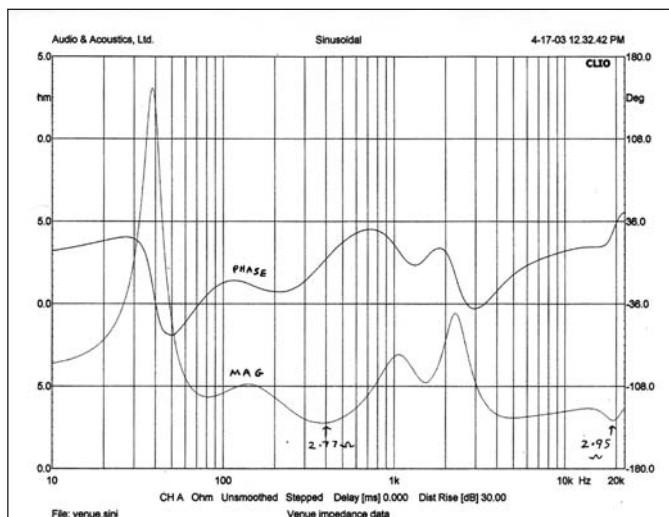


FIGURE 13: Impedance and magnitude phase.

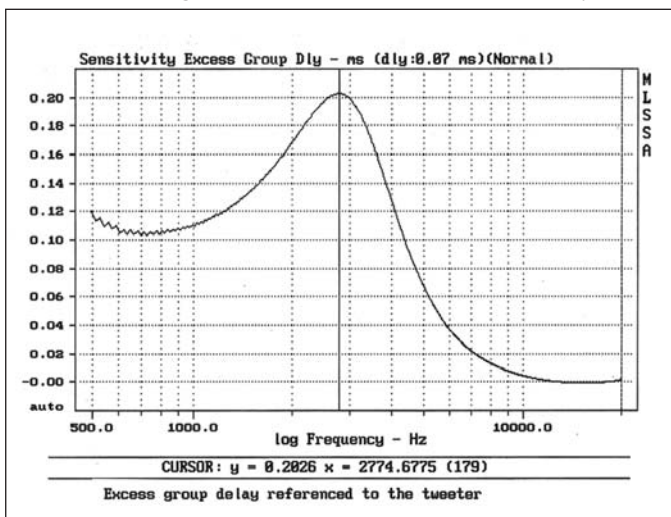


FIGURE 14: Excess group delay.

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SENSITIVITY AND POWER HANDLING

Figure 8 shows an average sensitivity of about 84dB at 2.83V, 48". Corrected for the standard 1m distance, this results in about 86dB.

Usually, the 2.83V (1W with 8 Ω) sensitivity is about what you would hear 2-3m away, because of room reinforcement and the contribution of both stereo speakers. But the full-range rear driver output adds about another 3dB in a room (a forward-only radiating speaker is omnidirectional in the bass only, resulting in a bass-heavy room reinforcement), but the Venue tends to maintain full spectrum reinforcement matching the direct output.

However, the Venue's average impedance is about 4 Ω , so 2.83V RMS produces 2W. The net result is this: In a good room, a power level of 1W per channel produces an SPL of about 86dB. A reasonable power-handling figure is about 200W peak sine-averaged (as power amps are specified). As my article "Musical Waveform Peak and Average Power Frequency Distributions" (*aX* Nov. '05) and letter from Robert Cordell ("Peak Audio Levels," *aX* Mar.

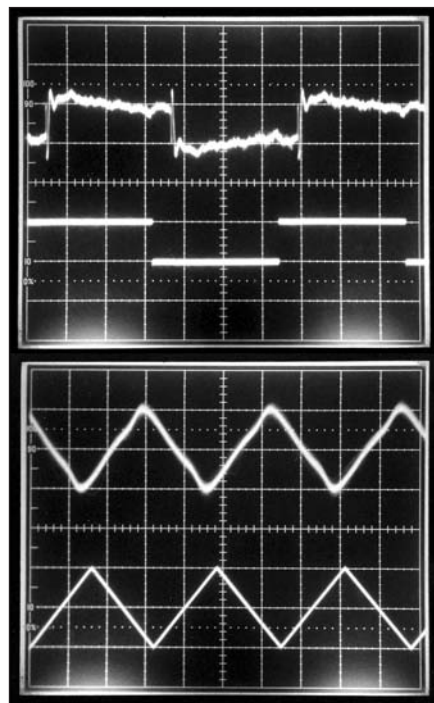


FIGURE 15 (top): 80Hz square-wave response.
FIGURE 16 (bottom): 1kHz linearm triangle wave.

TABLE 1.

Ref	Description	Size	Mfr	Mfr.PN	Dist	Dist. PN
	woofer	10"	Scan-Speak	25W/8565-01	Madisound	
	midrange (2)	5"	Focal	5W 4211	Zalytron	614211 5W4254
	tweeter (2)	1"	Morel	MDT-40?	Madisound	
	Acousta-Stuf (1.5 lb)					
C1	cap, 18 μ F, poly		Solen		Madisound	
C2	cap, 200 μ F, poly		Solen		Madisound	
C3	cap, 70 μ F, poly		Solen		Madisound	
C4	cap, 27 μ F, poly		Solen		Madisound	
C5	cap, 10 μ F, poly		Solen		Madisound	
C6	cap, 1.5 μ F, poly		Solen		Madisound	
C7	cap, 2.2 μ F, poly		Solen		Madisound	
C8	cap, 5 μ F, poly		Solen		Madisound	
L1	Ind, 5mH				Madisound	Sledgehammer
L2	Ind, 10mH				Madisound	Sledgehammer
L3	Ind, 0.3mH				Madisound	
L4	Ind, 0.6mH				Madisound	
L5	Ind, 0.15mH				Madisound	
L6	Ind, 39 μ H		Delevan	4590-393K	Digi-Key	DN4507
L7	Ind, 82 μ H		Delevan	4590-823K	Digi-Key	DN4511
R1, 3	Res, 3.5 Ω , 25W					
R2, 4, 6	Res, 10 Ω 25W					
R5	Res, 2.2 Ω , 15W					
R7	Res, 5.5 Ω , 15W					
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'06) show, the ratio of peak-to-average power on recorded music (if not clipped in playback) ranges from at least ten to over 20dB (a power ratio of 10:1 to

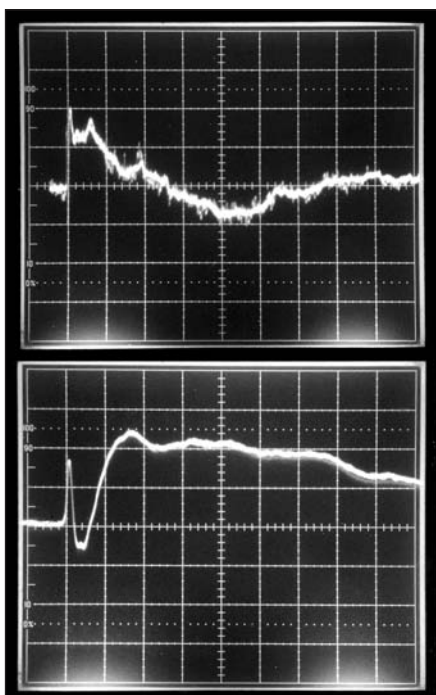


FIGURE 17 (top): 10Hz step response.
FIGURE 18 (bottom): Expanded step response.

over 100:1). Thus, the Venue speaker can comfortably handle the output of a 200W per channel amp if sensibly used.

The resulting peak SPL would then be about 109dB, with a maximum long-term average SPL of about 99dB. This is close to the maximum SPL generated by a large symphony orchestra.

One caution: while a pair of Venue speakers sounds very good and reasonably powerful with rock music, be careful with highly-compressed or synthesizer recordings; I wouldn't recommend using an amp of greater than 100W/channel, and that's assuming the amp isn't clipping.

But for most music listening, a pair (or four or five in multichannel) of these speakers will provide reasonably high SPL, powerful bass, very clear and natural tonality, precise imaging, and unusually deep and natural rendition of instrument and voice solidness and original venue ambience. *ax*

REFERENCE

1. "Loudspeaker Directionality and the Perception of Reality," by Bert Berlant, *Audio Eng. Society Journal*, May 1985.

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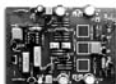
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By Charles Hansen

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- Finding the phase angle between voltage and current to determine the phase shift and impedance of a loudspeaker over its frequency range.
- Finding the phase shift between the input and output voltage of a tube amplifier to establish the HF (high frequency) and LF (low frequency) cut-off points needed to avoid instability in feedback amplifiers.

If you need to measure phase angle, you can do it with an oscilloscope, but it is a tedious process. You must display the two sine waves on the scope, and place the time base in the uncalibrated

mode with the reference sine wave positioned so its start and finish zero-crossings exactly span the full ten horizontal divisions on the screen. Each division represents 36° . For small phase angle differences, you can position a half-cycle over the ten horizontal divisions, so each major division will represent 18° .

Next, while triggering on the reference sine wave for display stability, you measure the time difference between the reference and variable sine wave same-direction zero-crossings and convert that number of divisions to degrees of phase shift. You must also determine—by relative position—whether the variable sine wave is leading or lagging the reference. You must use the chop mode—not alternate—so the two sine waves are synchronized in time.

Some very expensive digital storage oscilloscopes (DSO) have the processing

power to do this work for you, but for most hobby-oriented analog scopes, you need to perform the above steps for each frequency of interest. The

scope needs to have a wideband frequency response of at least 100MHz for a 50kHz sine wave measurement range. At low frequencies, the relatively long display blanking time will make the measurement difficult. At high frequencies the dimmer trace intensity will make the measurement equally difficult. A PC-based storage oscilloscope eliminates these problems and, if it has cursors you can position on the time axis, can eliminate any rough interpretation of an analog display and provide a more exact determination of the phase shift.

Over a year ago I won a Phase Meter PM-720 analog phase angle meter on eBay. The date codes on the parts indicated it was made about 1965. Because I couldn't find any documentation on the unit, I spent a fair amount of time reverse-engineering the schematic from the circuit board layout. Some of the circuitry was sealed in a potted module that plugs into the main circuit board.

I intended to restore the PM-720 to full working condition, but needed a method of accurately varying the phase angle of one sine wave with respect to another fixed reference sine wave. This article describes the fairly simple circuit that I made for this purpose. Parts

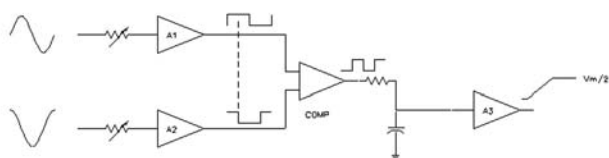


FIGURE 1: Phase Angle Meter block diagram.

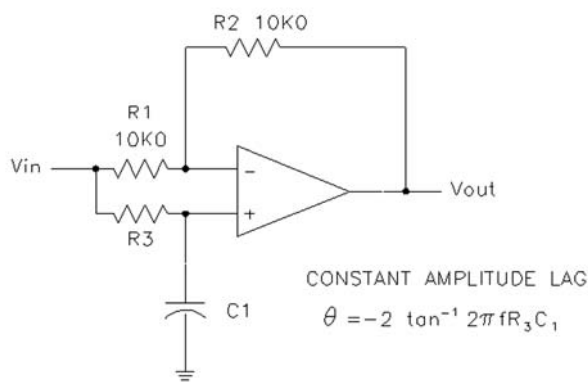


FIGURE 2: Generalized all-pass circuit.

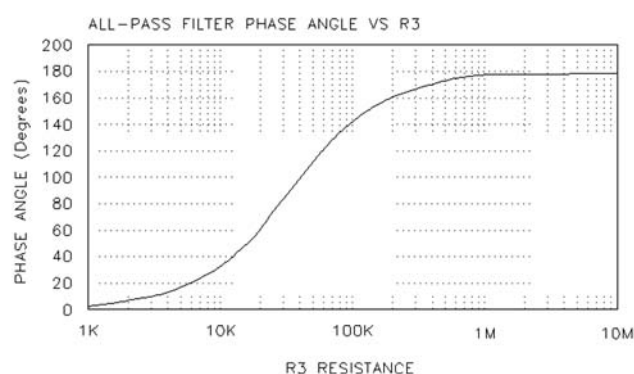


FIGURE 3: All-pass lag angle vs. R3.

cost is about \$41. You can also use it to calibrate the Analog Phase Angle Meter by Vern Mastel in the 4/1980 issue of *TAA*¹. Another analog phase angle meter candidate is the HP 8413A phase-gain meter, often available on eBay.

PHASE METER BASICS

The block diagram for the PM-720 in **Fig. 1** is fairly typical of the operation of an analog phase angle meter. The reference and variable sine waves are input to A1 and A2, which amplify (if necessary) the sine waves and convert them to phase-accurate square waves. The variable cosine wave in the diagram lags the sine reference by 90°. A comparator extracts the phase angle difference and applies the result to

an integrator that produces a DC voltage that is linearly proportional to the phase angle between the two sine waves.

A3 buffers and scales the integrated signal and applies a zero-level (in-phase) calibration offset if required. Logic circuit flip-flops (not shown) receive the same square waves that are input to the comparator and determine whether the variable sine wave input leads or lags the reference sine wave.

Modern digital phase angle or gain-

phase meters can do this and so much more with their DSP capabilities. You can also determine power factor, lead-lag compensation, phase-gain margin, watts, vars, voltage ratios, Bode plots, and so on.

The PM-720 designers had a very limited availability of ICs in order to perform the various steps needed to produce the phase angle measurement. As such, most of the circuitry is discrete, with over 170 parts in the unit, many of

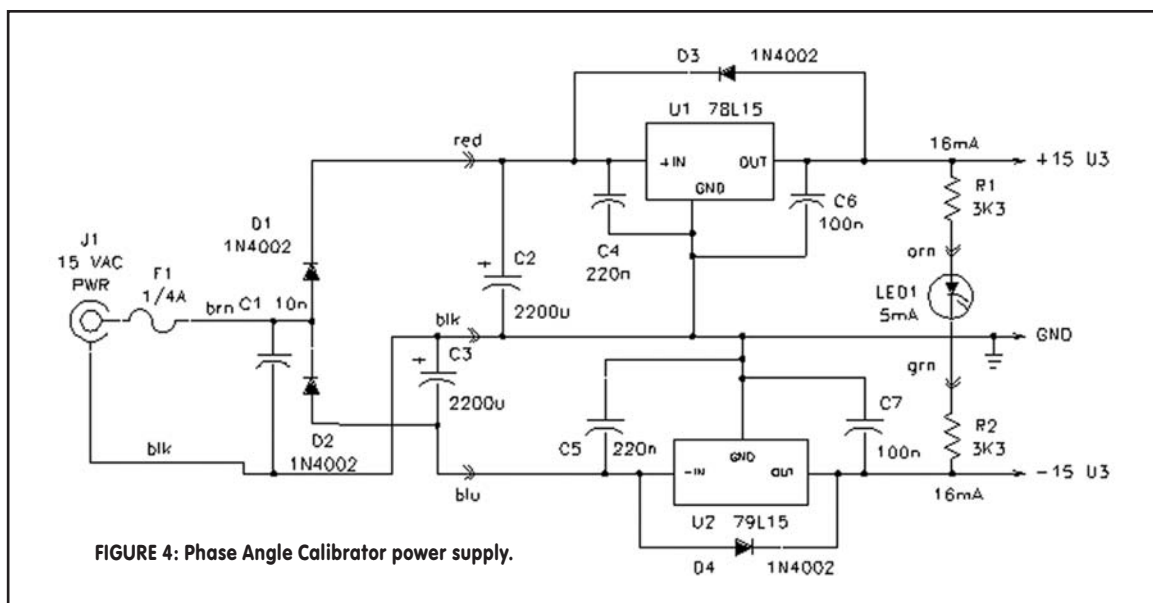


FIGURE 4: Phase Angle Calibrator power supply.



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The lead/lag indicating circuit is composed of five Fairchild μ L91429 dual 2-input NOR gates of RTL (resistor-transistor logic) in 10-lead TO-5 can packag-

The reference and variable square waves are then applied to level shifting circuits that convert the reference square wave to 0 to +1.7V, and the variable input square wave to -10V to +0V. These two square waves are applied to the potted module along with the zero adjust offset derived from the +12V supply. The integrator 0 to +3.6V DC voltage comes out the back of this module, and represents a 0 to 180° meter reading. There are four trimpots that calibrate the full-scale readings for each of the

The first, designed by linear IC genius



Bob Widlar, was the $\mu A702$ produced by Fairchild Semiconductor (FSC) beginning in 1963. It had the same +12 and -6V supply voltages as the $\mu A710$ comparators used on the PM-720 main PC board, and the date codes on the 710s were 532 (32nd week of 1965, using the generally accepted industry date coding). The Fairchild high gain-bandwidth $\mu A709$ (also designed by Widlar) had just come out in 1965, so it probably wouldn't have been available during the PM-720 design cycle. Either op amp would have been quite expensive in 1965. If I am correct that this module contains one of the early op amps, it may have been potted inside to hide the integrator design details from perceived competitors. Or I could be totally wrong.

Bob Pease of National Semiconductor said these original $\mu A702$ and $\mu A710$ FSC designs used crude NPN monolithic technology, which was all that was available back then. They used +12V and -6V supplies so as to not exceed the overall 18V breakdown rating. The $\mu A702$ was a poor performer, and was out of production by the mid-1980s (the last listing I have in my 1987 Fairchild Linear catalog shows support for only the Military $\mu A702QB$ package).

The 702 and 710 circuits used mostly load resistors pulling up to +12V, and pulldown resistors to -6V. Most of the signal processing was done between the zero volt ground pin and +6V. The output swing was in the range 0 to +6V. For its day, it was considered fairly bright.

When the $\mu A709$ came along, with its 36V breakdown ratings it could happily run on $\pm 15V$ supplies.

Texas Instruments also made versions of these devices. Bob didn't mention whether National ever made the 702, but Widlar went to work for NSC in 1966 and designed the much improved LM101 and LM107 op amps.

HOW THE CALIBRATOR WORKS

The Phase Angle Calibrator makes use of an op amp filter circuit called the all-pass circuit, which takes a sine-wave input and produces a constant amplitude phase-shifted sine wave output. The lag output version is shown in **Fig. 2**. You can produce a leading phase-shifted output by interchanging R3 and C1.

The theory behind the all-pass filter is available in many reference books and texts, but I found one by Walt Jung³ that I believe is the easiest for a novice to understand. The phase shift angle is varied by R3 and C1 in accordance with the formula

$$\theta = -2 \tan^{-1} 2\pi f R_3 C_1$$

where θ is the phase angle, and f is the frequency. After selecting a suitable value for C_1 , you can solve for R_3 by rearranging the formula:

$$R_3 = \tan(-\theta/2) / 2\pi f C_1$$

This is hardly a linear relationship, as you can see by the graph in **Fig. 3**. Large

changes in resistor value produce very little change in phase angle as you approach 0° or 180°. It is much easier to apply the input signal to both inputs of the phase angle meter for 0°, and use an op-amp inverter to generate the 180° signal. You also need to multiply the value of C1 by 10 for each decade decrease in frequency to retain the constant

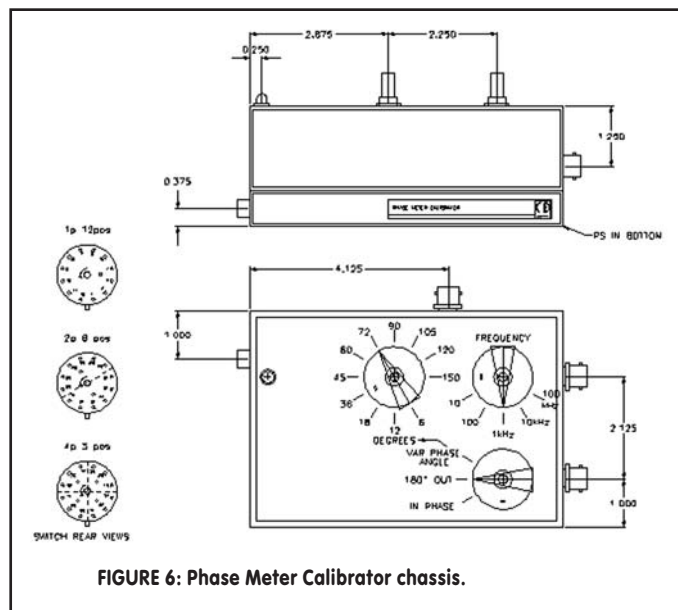


FIGURE 6: Phase Meter Calibrator chassis.

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amplitude feature of the all-pass filter.

The schematic diagram for the Phase Meter Calibrator power supply is shown in **Fig. 4**. The J1 power source is a 15V AC plug-in AC adapter, so there is no

on-off power switch.

D1 and D2 each half-wave-rectify the 15V AC into roughly $\pm 21V$ DC. C2 and C3 are the main reservoir caps, and U1 and U2 are linear regulator ICs that provide $\pm 15V$ DC at the supply rails for the all-pass phase shift circuits. C4 and C5 ensure stable operation for the regulators. LED1 is the power-on indicator. Fuse F1 protects the power supply AC adapter. C6 and C7 are the local bypass caps for the all-pass circuit op amp.

The schematic diagram for the all-pass phase shift circuit is shown in **Fig. 5**. A sine-wave signal generator is applied to J7. The calibrator can handle up to 7V RMS of input signal. This becomes the reference sine wave that is connected to J8. Op-amp U3A is the all-pass filter.

S3 selects a phase shift angle from 6° to 150° , and S4 selects a frequency range from 10Hz to 100kHz. The phase angles I chose for S3 are optimized for the four ranges of operation of the PM-720, but you can change the resistors to meet your individual requirement, in accordance with the formula shown earlier to solve for R_3 .

S5 applies the input sine wave to the variable output jack J9 for the zero phase shift condition. In the second 180° phase shift position, the input sine wave is processed through unity-gain inverter U3B. The variable phase shift position of S5 selects the output of the all-pass circuit U3A. The two output jacks have series 100Ω resistors to protect against accidental shorts. The input source signal generator should have a low output impedance of no less than 600Ω .

CONSTRUCTION

The drawing for my chassis is shown in **Fig. 6**, along with the rear views of the rotary switches I used. I built the calibrator circuitry on two wire-wrapped perfboards for this project. In keeping with good assembly practice, solder in the least sensitive parts first (resistors, then capacitors). Then install the regulator ICs and other semiconductors last. Double-check the orientation of the polarized components. I used a low-profile IC socket for the op amp, and checked for the proper power supply voltages at the socket before plugging in U3.

Photo 1 shows the top view of my Phase Meter Calibrator chassis. The switches and banana jacks on the left side are associated with a transistor matching circuit that is not part of the subject calibrator (for details see www.passdiy.com/articles.htm, transistor matching). The lettering was made on transparent adhesive-backed drafting appliqué film, as is my usual practice. The sine wave input jack J7 is at the top, while J8 and J9 are on the right side.

Photo 2 shows the internal wiring. The raw DC supply, power jack, and fuse are in the bottom part of the chassis on the left in the photo. The switches, DC regulators, and op amp are located in the top half of the chassis, on the upper-right side of the photo. The unrelated transistor matching jacks and switches are on the lower right in the photo.

I used tapped nylon spacers to mount the perfboards to the chassis. I used $\frac{1}{4}$ " spacers on the raw DC supply board and cut the wire-wrap pins short because of the depth of the rotary switches. The op-amp board uses $\frac{3}{4}$ " spacers and full-

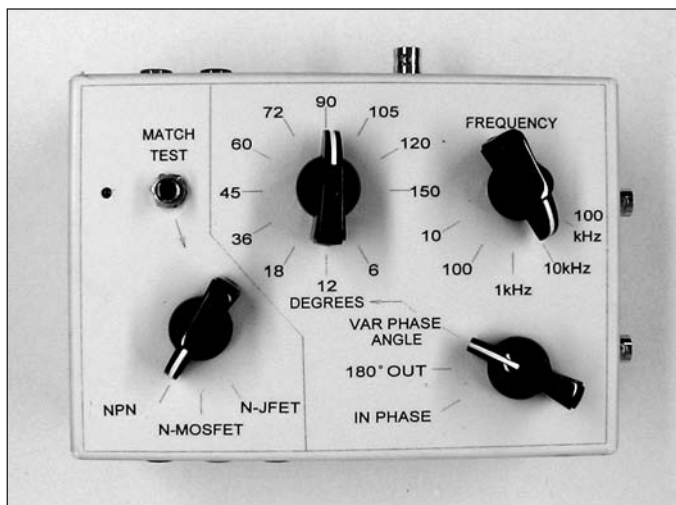


PHOTO 1: Phase Meter Calibrator top view.

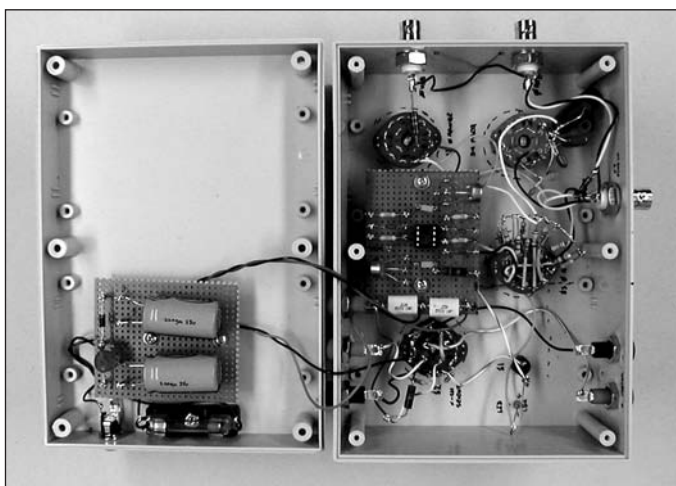


PHOTO 2: Phase Meter Calibrator interior view.

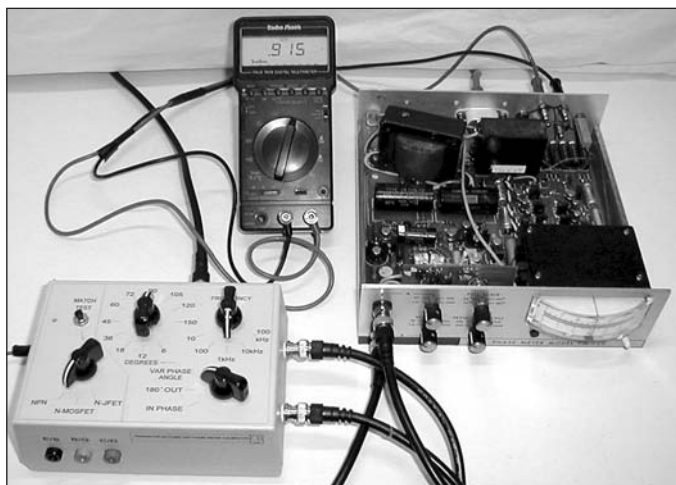


PHOTO 3: Phase Meter Calibrator with PM-720 phase meter.

height pins. The resistors and caps that control the phase shift are mounted on their respective rotary switches.

USING THE PHASE METER CALIBRATOR

Photo 3 shows the calibrator in use with the PM-720 Phase Meter. The input signal is a 1kHz sine wave, and the calibrator is set for 90°. The PM-720 shows the needle centered on the 180° FS range. My DMM is connected to the analog output jacks and shows a reading of 91.5° (0.915V DC). The red power lamp and blue LAG lamp can also be seen below the meter.

The PM-720 still has its original aluminum filter caps at this point. As soon as I verified that the unit was fully func-

tional, I replaced all the aluminum caps with Mouser Xicon TG high temperature axial lead caps.

R9 is always in-circuit and gives approximately 180° phase shift, although at this extreme the gain is dropping from unity and the distortion is higher. Because of the asymptotic approach of the required resistor values to the 0° and 180° phase angles, you may need to select different values of R10, R11, R19, and R20 while using an oscilloscope to verify the phase shift. C14 (100kHz range) is only 47pF, and suffers from any parasitic capacitance in the wiring. Again, I needed to select C14 from a number of different NP0 caps to get a reasonable 3% accuracy at the highest range. **aX**

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Mouser Electronics

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www.mouser.com

Radio Shack

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www.tandy.com

TABLE 1: PARTS LIST.

Symbol	Value	Description	Vendor	Part No.
C1	10nF 50V	Ceramic X7R	Mouser	21RX510
C2-C3	2200µF 35V 20%	Aluminum	Mouser	140-XRL35V2200
C4-C5	220nF 100V 10%	Polyester	Mouser	140-PF2A224K
C6-C7	100nF 50V	Ceramic X7R	Mouser	21RX310
C8-C9	27pF 50V 5%	Ceramic NP0	Mouser	140-50N2-270J
C10	470nF 250V 10%	Polypropylene	Mouser	1429-2474
C11	47nF 250V 10%	Polypropylene	Mouser	1429-2473
C12	4n7 100V 10%	Polypropylene	Mouser	1429-1472
C13	470pF 50V 5%	Polystyrene axial	Mouser	23PS147
C14	47pF 50V 5%	Ceramic NP0	Mouser	140-50N5-470J
D1-D4	1N4002	1A 50V	Mouser	625-1N4001
F1	¼A 3AG	Fuse	RS	270-1002
J1		Power jack, 5mm/ 2.5mm mini	Mouser	502-TR-2A
J7-J9		BNC, insulated	Mouser	523-31-10-RFX
LED1	LED	Red T-1 ¼	Any	
R1-R2	3k3 ½W 5%	Carbon film	Mouser	293-3.3k
R7, R8, R21, R22	10k0 1%	Metal film	Mouser	271-10.0k
R9	10M 5%	carbon film	Mouser	29SJ250-10M
R10	127k 1%	metal film	Mouser	271-127K
R11	59k0 1%	metal film	Mouser	271-59.0K
R12	44k2 1%	metal film	Mouser	271-44.2K
R13	33k2 1%	metal film	Mouser	271-33.2K
R14	24k3 1%	metal film	Mouser	271-24.3K
R15	19k6 1%	metal film	Mouser	271-19.6K
R16	14k0 1%	metal film	Mouser	271-14.0K
R17	11k0 1%	metal film	Mouser	271-11.0K
R18	5k36 1%	metal film	Mouser	271-5.36K
R19	3k57 1%	metal film	Mouser	271-3.57K
R20	1k78 1%	metal film	Mouser	271-1.78K
R23	4k99 1%	metal film	Mouser	271-4.99K
R24-R25	100 5%	carbon film	Mouser	29SJ250-100
S3	1P 12 POS	Rotary non-shorting	Mouser	10YX112
S4	2P 6 POS	Rotary non-shorting	Mouser	10YX026
S5	4P 3 POS	Rotary non-shorting	Mouser	10YX043
U1	78L15	Reg. +15V, TO92	Mouser	513-NJM78L15A
U2	79L15	Reg. -15V, TO92	Mouser	513-NJM79L15A
U3	NE5532AP	op amp, single audio	DK	296-7040-5-ND
		Enclosure, bud 6.89 × 4.89 × 2.5	Mouser	563-PT-11557-G
		15V AC 400mA power adapter	Mouser	41AC415
		Fuse clip, single	RS	270-739
		Socket, 8-pin gold WW DIP	Mouser	575-213308
		Wrap-ID, DIP8	DK	K470-ND
		hardware, perfboards, standoffs, wire, etc.		

REFERENCES

1. "An Analog Phase Meter," Vern Mastel, 4/80 *TAA*, pp. 12-16.
2. "Simple Phase Meter Operates to 10MHz," Ron Mancini, *Intersil Application Note AN9637.1* (10/27/2004).
3. *Audio IC Op-Amp Applications, 2nd Edition*, Walt Jung, Howard Sams, Para. 5.3.7 All-Pass Circuits, pp. 164-165.

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By Joseph Norwood Still

This Class A1 single-ended, triode/ultralinear power amplifier (**Photo 1**) is an excellent performer designed to leave the high distortion associated with single-ended amplifiers relegated to the history books. The 1kHz, 0.6% distortion at 20W triode mode and 25W, 0.64% distortion ultralinear (UL) mode is not an aberration but the actual distortion of this amplifier, which is achieved through careful design to obtain the lowest distortion of each stage throughout the amplifier. The design requires selecting the proper operating point for the tubes and increasing the operating plate current of the paralleled 6550s power tubes from a typical 60mA to a more aggressive 95mA.

By increasing the plate current, the distortion of the 20W triode amplifier is reduced from 9% to 4% (with 6550s cathode bypassed). The next step is to leave the cathode resistors of the power tubes unbypassed. This produces degeneration within the triode output stage and reduces the distortion of the power amplifier to 2.5% at 20W. The final step

in the distortion reduction process is obtained by applying 7dB of loop feedback to reduce the distortion of the power amplifier to 0.6%.

These three events describe the primary reason for the dramatic reduction of the distortion from 9% to 0.6% for the 20W amp. The amplifier sensitivity is 0.46V RMS for 20W triode mode and 25W UL mode.

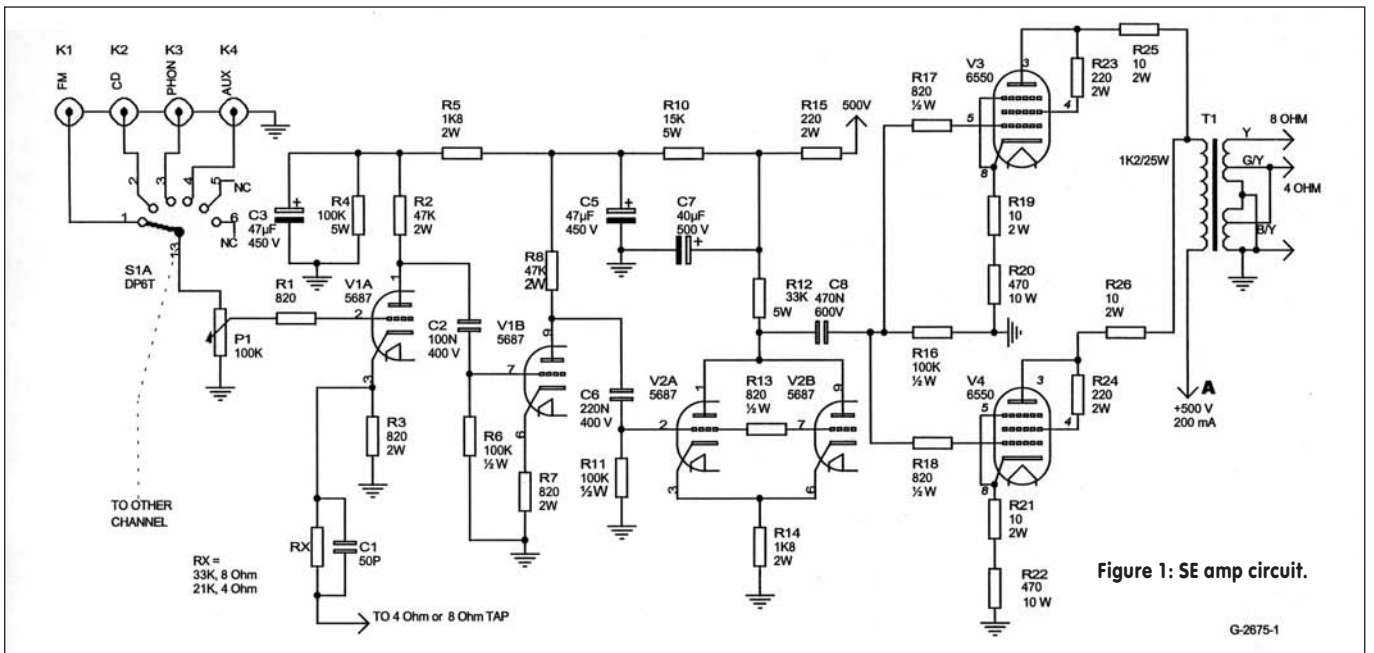
When operated in the UL mode at 25W/1kHz, the amplifier has a distortion of 0.64% or 0.8%/1kHz distortion

at 30W. The 25W level is the rating limit of the Hammond output transformer. For operation in the UL mode, disconnect the 220Ω resistors from the plates of the 6550s and connect them to the screen tap of the output transformer. The sensitivity of the amp is 0.46V RMS for an output of 25W. This is the only circuit change required to modify the amplifier from triode to UL operation.

Of Interest: If you look at the sine wave of a single-ended amplifier (with



Photo 1. Finished unit.



cathode bypassed or fixed bias) on an oscilloscope, you can see the unsymmetrical nature of the sine wave. This is greatly improved by providing degeneration in the cathode of the output stage. The distortion obtained by employing cathode degeneration and increased plate current of the 6550s greatly reduces the distortion in this SE amplifier. The final distortion mimics or improves the distortion levels typically found in push-pull amplifiers.

The 6550 power output tubes are operated at the design center limit to obtain the lowest distortion, and no color of the plate is evident. Also, at this level of operation the sonics of the sound, which is commonly referred to as a "sweet spot," is greatly improved. With an un-bypassed cathode resistor in the cathode, the plate resistance of the triode output tubes is increased, which reduces the damping on the voice coil of the speakers. This is not a major problem, because triodes naturally have an inherently low plate resistance. Thus, the inherent low impedance of the triode is still present, although diminished.

The damping of the triode and ultralinear amplifiers is satisfactory. When the 8Ω secondary winding is adjusted for a 2.8V RMS output and connected to an 8Ω load, the voltage increases to only 4.0V RMS (triode) and 6.2V RMS (ultralinear) when the load is removed. This damping is sufficient, and the bass frequencies of the woofer are effectively controlled.

At 10W (normal listening level), the triode amplifier has less than 0.5% distortion from 30Hz to 15kHz, and the ultralinear amplifier has less than 0.56% distortion from 30Hz to 15kHz.

FEATURES

The amplifier is a complete unit, and two amplifiers are required for stereo operation. A line amplifier/preamplifier is not required to drive the amplifier, which has a right- and left-channel volume control. A double section, six-position rotary switch is required to select FM, CD, SACD, Phono, and so on.

Important: I recommend that you construct the amplifier on a single aluminum $17 \times 10 \times 3$ " chassis and not as shown in **Photo 1**. Also, do not use a steel chassis box, because grounds re-

quire stripping the painted steel chassis to bare metal. Furthermore, drilling and punching holes is much more difficult with a steel chassis. The front of the power amplifier chassis has a heater and high voltage on-off switch and two volume controls. The rear of the amplifier chassis has a power cord and fuse, four speaker terminals, and eight phono jacks.

I further recommend that you use an 80-cfm fan, located directly behind the plate transformer, and with the air circulating around the transformer directed at the 6550 power output tubes. I also suggest that you "notch-out" the chassis at the lower end of the fan to provide under-chassis cooling from the fan. These recommendations will ensure proper cooling of the tubes, transformers, and electrical components.

CASCADE PREAMPLIFIER

The first stage of the cascade amplifier (**Fig. 1**) has a flat frequency response of 20Hz to 30kHz, distortion of .08% at 1V RMS, and voltage gain of 10. The DC supply is +305V, the plate voltage is +85V, and cathode voltage is +3.7V.

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The second stage of the cascade amplifier has a flat frequency response of 20Hz to 30kHz, distortion of 0.54% at 10V RMS, and voltage gain of 10. The DC supply is +320V, the plate voltage is +88V, and cathode voltage is +4V. Do not attempt to bypass the cathode resistor with a capacitor, as instability of the amplifier will occur. Note: disconnect the feedback loop when making these tests.

DRIVER STAGE AMPLIFIER

The driver stage amplifier (Fig. 1) is a paralleled 5687 with a flat frequency response of 20Hz to 20kHz. It uses a plate resistor of 33k Ω and a cathode resistor of 1800 Ω . The DC supply is +500V DC and the plate voltage via 33k Ω plate load resistor is +250V DC. The cathode voltage is +14V DC, and the distortion is 1.2% at 1kHz with 70V RMS output (6550 tubes plugged-in, F.B. loop not connected).

TABLE 1: TRIODE POWER AMPLIFIER DISTORTION %.

	5W	10W	15W	20W
32Hz	0.48	0.46	0.62	1.0
100Hz	0.46	0.44	0.54	0.64
1kHz	0.46	0.44	0.5	0.6
10kHz	0.44	0.42	0.5	0.62
15kHz	0.4	0.38	0.46	0.31

TABLE 2: ULTRALINEAR POWER AMPLIFIER DISTORTION %.

	5W	10W	15W	20W	25W	30W
32Hz	0.46	0.46	0.64	0.8	1.5	2.3
100Hz	0.45	0.4	0.5	0.52	0.7	0.94
1kHz	0.44	0.42	0.48	0.5	0.64	0.8
15kHz	0.42	0.38	0.4	0.46	0.6	0.9

TABLE 3: 6550 TUBE CHARACTERISTICS (SINGLE TUBE, UNLESS TWO TUBES SPECIFIED).

	Triode Mode	Ultralinear Mode
Ebb	500V DC	500V DC
Eb	485V DC	485V DC
Eg2 (220 Ω screen resistor)	483V DC	487V DC
Ek (485 Ω cathode resistor)	48-50V DC Ek	48-50V DC
Ib + Ig2 (zero signal)	93mA	93mA
Ib + Ig2 (maximum signal)	96mA	98mA
Grid Signal (maximum)	70V RMS	70V RMS
Plate dissipation (zero signal)	40W	40W
Plate dissipation (maximum signal)	32W	35W
Plate load (two tubes)	1250 Ω	1250 Ω
Power input (two tubes)	100W	100W
Power output (two tubes)	20W	25W

TABLE 4: 5687 VOLTAGE REQUIREMENTS.

	5687 Cascade Amplifier		5687 Paralleled Driver
	First Stage	Second Stage	Both Sections
Ebb	305V	320V	500V
Eb	85V	88V	250V
Ek	3.7V	4V	14V

Do not attempt to bypass the 5687 cathode resistor, as instability of the amplifier will occur. Furthermore, distortion reduction is obtained by the unbypassed cathode resistor of the 6550s and the higher plate current of the 6550s. This is verified by referring to the distortion data in **Tables 1** and **2**. Note: disconnect the feedback loop when making these tests.

POWER OUTPUT STAGE

The power amplifier stage (Fig. 1) consists of two 6550s connected in a Class A1 single-ended configuration. The 6550 degenerative output stage (unbypassed cathode) requires 70V RMS signal drive for the grids of 6550s to obtain a 20W output in the triode mode and 25W output in the UL mode. The reason for classifying the 6550s as operating in a

Class A1 mode is: The amplifier's plate current is virtually constant and the grid current at maximum signal is only 1mA. This amount of grid current is a typical drive requirement to overcome the capacitance encountered in paralleled high transconductance 6550 power tubes.

Of the 70V RMS drive signal, 35V RMS is required to develop the degeneration (6dB of negative feedback) in the cathode circuit of the 6550s. The other 35V RMS of the 70V RMS signal is required to drive the 6550s to obtain a 20W output. The grid bias of the 6550 is -50V DC and the signal is 35V RMS. This signal is required to drive the 6550s to a 20W triode output and a 25W UL output.

The 6550 output tubes employ isolat-

PARTS LIST OF AMPLIFIER (DOUBLE FOR STEREO EXCEPT AS NOTED WITH #).

Part	Quantity	Source
10 Ω 2W 500V metal oxide 282-10	4	Mouser
220 Ω 2W 500V metal oxide 282-220	2	Mouser
470 Ω 10W wire-wound 280CR10-470	2	Mouser
820 Ω ½W carbon 30BJ500-820	3	Mouser
820 Ω 2W metal oxide 282-820	2	Mouser
1800 Ω 2W metal oxide 282-1800	2	Mouser
15k Ω 5W metal oxide 286-15K	1	Mouser
22k Ω 1W for 4 Ω speaker 281-22K	1	Mouser
33k Ω 1W for 8 Ω speaker 281-33K	1	Mouser
47k Ω 2W metal oxide 282-47K	2	Mouser
100k Ω 5W 500V metal oxide 30BJ500-100K	1	Mouser
100k Ω ½W carbon 30BJ500-100K	3	Mouser
33k Ω 5W 500V metal oxide 286-33K	1	Mouser
100k audio taper RV38-100KA	1	AE
#S1-Rotary switch, 2-pole, 6 position 275-1836	1	RS
50pF mica capacitor 500V C-SM 50	1	AE
0.1 μ F 400V polypropylene C-PD1-400 (orange cap)	1	AE
0.22 μ F 400V polypropylene C-PD22-400 (orange cap)	1	AE
0.47 μ F 400V polypropylene C-PD47-400 (orange cap)	1	AE
40 μ F 500V electrolytic CSA40-500	1	AE
47 μ F 450V, electrolytic, C-ET47-450	2	AE
5687, double triode (V1, V2)	2	AE
6550C-WC power pentode Winged C (matched-V3, V4)	2	AE
T1- Output Transformer, P-T1640SE Single-End, 25W, 200mA maximum, primary impedance 1250 Ω , secondary 4, 8, 16 Ω , 20Hz to 20kHz ($\pm$ 1dB maximum at 25W), weight 12 lbs.		AE
9-pin ceramic tube socket PST9-162	2	AE
Octal tube socket PST8-209MIP	2	AE

PARTS LIST OF POWER SUPPLY

Part	Quantity	Source
Fan 80 cfm, 4.7 in ² $\times$ 1" thick, "quiet"	1 (Specifications Typical) - -	
Diode Rectifier 1A, 1000V, P-Q1N4700	4	AE
Switch, Rocker, 6A at 250V SPST, Mtg. Hole-.71"	2	112-R13-130A Mouser
Resistor, Wire Wound, 20 Ω 25W	1	280-CR25-20 Mouser
75k Ω 500V 2W 282-75K	2	Mouser
0.2 μ F 600V, CSD2-600	1	AE
470 μ F 400V, CR-474	4	ALEL
Heater transformer, 120V to 12.6V, 10A, Hammond	1	P-T166S6 AE
Plate transformer, 120V to 200V, 870mA, Hammond	1	P-T167H200 AE
Chassis box, aluminum, 17 $\times$ 10 $\times$ 3", P-H1444-32	1	AE
Fuse holder, 3AG	1	--
Fuse, slow blow, 3AG, 5A 250V	1	--
Line cord (3 prong plug, 14 AWG)	1	--

tion resistors on all the tube elements to ensure the plate, grid, and cathode elements are free of unwanted signals. A plate "stopper" 10Ω metal oxide resistor, which measures static plate current, is used. The screen grid is connected to the plate via a 220Ω 2W metal oxide resistor for triode operation. A 470Ω 10W resistor is connected in series with a 10Ω 2W metal oxide resistor to provide a +48V DC bias voltage. With maximum signal drive the self-bias voltage is increased to 50V DC.

The two 6550s can be balanced for equal plate currents by interchanging the four 6550s until balance is achieved. If this exchange does not provide balance, replace the 10Ω resistor of the tube with high plate current with a 22 or 33Ω resistor.

I did not have this problem using matched output tubes. The current of each 6550 output tube is 95mA. The Ebb is + 500V DC, the Eb is +485V DC, and cathode-to-plate voltage is +435V DC. The 6550s are operated with 40W plate dissipation (no signal), and no color of the plates is evident. At 20W power output, the plate dissipation is 32W. At an output of 20W triode and

25W ultralinear, the 6550 control grid has a 70V RMS driving signal and a typical AC driving current of 1mA.

Note: 1.) The distortion of the power amplifier does not fall off rapidly as the power output is decreased but tends to remain relatively constant at approximately 0.5%. It is also evident that the distortion at 5W is slightly higher than at 10W (Tables 1 and 2). This is also due to the degenerative effects of the 6550 cathode circuit. Because the feedback effect is more pronounced as the frequency is increased, the distortion at 10kHz and 15kHz is decreased. This, of course, affects the shape of the 10kHz squarewave, which has slanted/rounded sides.

2.) The power amplifier shows no sign of ringing with a 10kHz square wave applied to the input when you place a 1000N capacitor across the 4 and 8Ω secondary windings of the output transformer.

3.) A single-ended amplifier connected to an oscilloscope reveals a non-symmetrical sine wave; as is typical for

single-ended amplifiers, the bottom end of the sine wave is broader than the top end. This condition does not occur in push-pull amplifiers. The unsymmetrical state of the sine wave is eliminated in self-biased, single-ended amplifiers when using an unbypassed cathode resistor.

4.) Before applying power to the amplifier, make certain a 100kΩ grid resistor is connected between the 6550 grid and ground. This precaution will prevent catastrophic failure of the 6550s.

5.) Hammond transformers fit well with my philosophy of designing affordable hi-fi equipment for the audiophile.

POWER SUPPLY

The power supply contains the power transformer, heater transformer, capacitors/resistors, switches (heater/plate), and a low noise blower fan (cools the power transformer). The front of the chassis has heater and high voltage switches for applying heater power prior to applying high voltage. The power supply provides power for two of the single-ended power

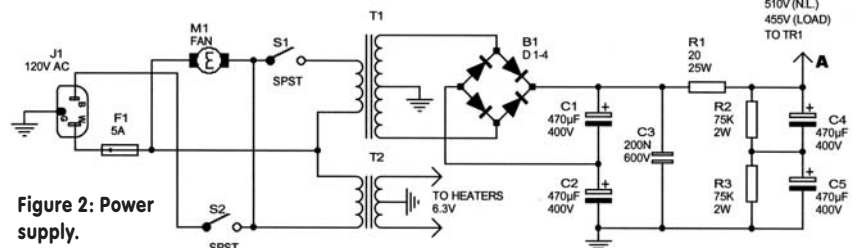


Figure 2: Power supply.

G-2675-2

WARNING: Lethal voltages are present. Exercise extreme caution when constructing and testing the amplifier and never leave the amplifier upside down when children are present.

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amplifiers described in this article.

The power supply (**Fig. 2**) has a 6.3V AC-C.T., 10A heater transformer, and a 200V AC 870mA plate transformer. The plate supply features a full-wave bridge, voltage doubler circuit, which uses two 470 μ F capacitors. The current rating of the transformer is reduced to 435mA. Filtering of the output voltage from the bridge rectifier is provided by series-connected 470 μ F electrolytic capacitors and a 20 Ω isolation resistor.

Two series-connected 75k Ω bleeder/voltage divider resistors are included in the output of the supply. The 200N capacitor (C3) eliminates AC hash. The AC ripple of the +500V DC output of the supply is 1.5V RMS. The plate and heater transformer furnishes power to two 20/25W power amplifiers. The 6.3V AC from the heater transformer is directed to the tubes of the right- and left-channel amplifiers via separate heater wires.

FREQUENCY RESPONSE TUBE CHARACTERISTICS

At 20W the triode amplifier is flat from 30Hz to 20kHz. At 25W the UL ampli-

fier is flat from 30Hz to 20kHz (**Tables 1 and 2**).

The stereo amplifier requires four 6550s and four 5687s. I recommend that you order six 5687s to ensure replacement of a failed tube or selection of low distortion tubes, if you have a distortion meter (**Tables 3 and 4**).

RECAP OF DESIGN

This low distortion design is the exclusive domain of a power output stage operating in the triode or ultralinear mode. Tubes that are especially suited for the degenerative cathode circuit are triodes having high cathode bias and low plate impedance.

The design is compromised with a pentode output stage and its inherent high plate impedance. The ultralinear mode also compromises the damping. If this amplifier used the conventional output stage with a bypassed cathode biased for 60mA, the distortion at 20W/1kHz would be 9%.

Currently, most single-ended amplifiers operate with between 5 and 10% distortion, which level is very noticeable and sonically produces a very unpleas-

ant sound and eliminates any positive features attributed to a single-ended amplifier. This design with the degenerative output stage and more aggressive operating plate current changes the whole design concept for single-ended amplifiers.

With its 0.6% distortion at 20W, this amp produces outstanding distortion numbers. As the frequencies are increased from 1kHz to 10kHz, the distortion is reduced at all power levels (**Tables 1 and 2**). This phenomenon is typical for a power output stage with self-bias having an unbypassed cathode resistor. The UL amplifier provided 25W with 0.64% distortion and 30W with 0.8% distortion.

I recommend the triode amplifier over the UL version, as the former has better damping. Try both modes and make your own selection based on listening observations. To confuse the issue, you might like to know I operate my amplifier in the UL mode.

The amplifier has excellent "sonics" as typified with SE amplifiers. The unpleasant sound of the high distortion SE amplifier is vanquished! The approximate cost to build the stereo amplifier and power supply is \$450.

Important: This amplifier design is presented to show that you can obtain low distortion in SE amplifiers with a very simple and logical design approach. It is intended for SE devotees and is not recommended as a DIY construction article for the typical audiophile. For the typical audiophile, I recommend and prefer my fixed bias push-pull AB1 35W triode/60W UL amplifier design in *audioXpress*, June 2004. **ax**

Joseph Norwood Still, retired from the Electronics Industry, is designing affordable high-quality amplifiers for the dedicated audiophile. He resides in Towson, Md.

SOURCES

Mouser Electronics 1-800-346-6873, mouser.com

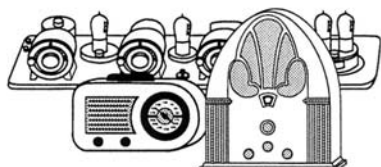
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Build a Flat Panel Speaker

This article investigates distributed-mode loudspeakers (DML), a new development in loudspeaker technology, with many desirable attributes and a multitude of practical applications.

By Daisuke Koya

Dr. Ken Heron¹ at the United Kingdom's DERA (Defence Evaluation and Research Agency) worked on composite panels for helicopters. In his research, he found that the panels radiated significant output in the audio bandwidth. Although this was not desired in aircraft, it was suggested that this could be used in loudspeakers.

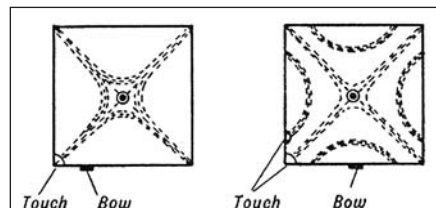


FIGURE 1: Two examples of modal resonance patterns in a free plate (after Chladni 1787)².

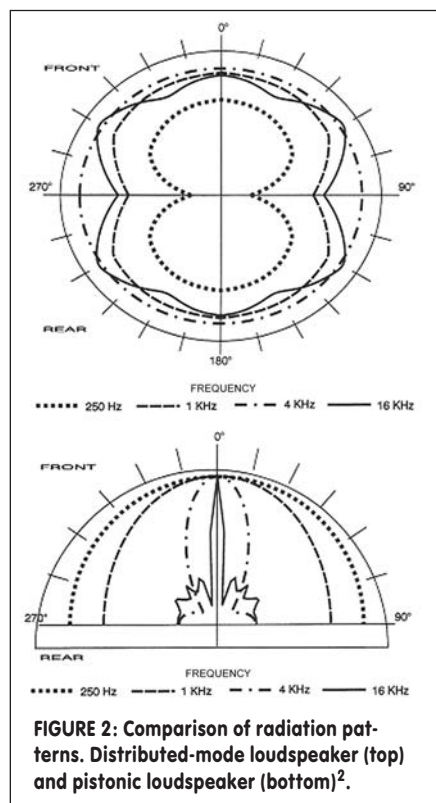


FIGURE 2: Comparison of radiation patterns. Distributed-mode loudspeaker (top) and piston speaker (bottom)².

COMMERCIAL EXAMPLES

There are quite a few commercial examples of DMLs (distributed-mode loudspeakers) today. I own Sonic Impact's SI-5, Gen2 SI-5, and the amazingly good-sounding cardboard SoundpaX DML speakers. Brookstone, the specialty consumer products store, has a product called the "Wafer-Thin CD System with MP3," which also sounds very nice.

Teac has mini stereo systems that feature DML speakers. NXT, the primary licensee of DML technology, has developed SoundVu, which allows a video display device to double as a loudspeaker providing up to three channels of sound. NEC currently sells desktops and notebook per-

sonal computers with LCD displays that incorporate SoundVu technology.

THEORY

Chladni² in the late 18th century demonstrated that resonant modes were produced in a glass or metal plate (Fig. 1). The plate was fixed at its center, excited by a violin bow, and damped by nodal points by touch, which displayed modes through the medium of sand dusted on the surface.

A DML³ is a bending wave panel with resonant modes randomly distributed in frequency throughout the vibrating panel. Although for piston technology, resonance is damped or generally

avoided to prevent unpredictable frequency response, distributed-mode loudspeaker design encourages as many resonances as possible.

A DML has these characteristics:

- Wide bandwidth⁴—An optimized DML design can have eight octaves of bandwidth.

- Wide directivity—Compared to piston speakers, the DML displays wide directivity (Fig. 2), which is generally independent of frequency.

- Boundary reaction¹—For the

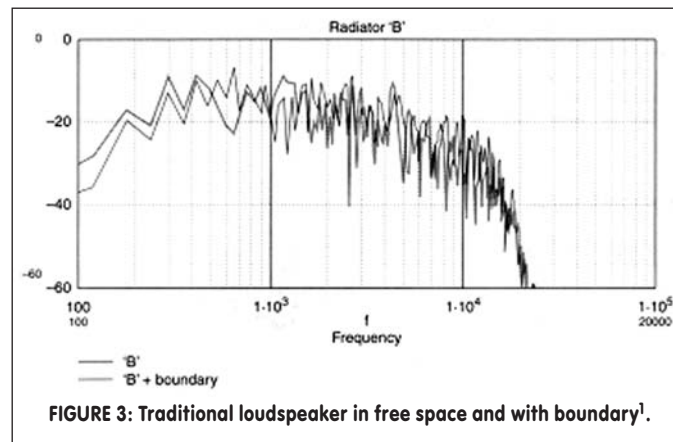


FIGURE 3: Traditional loudspeaker in free space and with boundary¹.

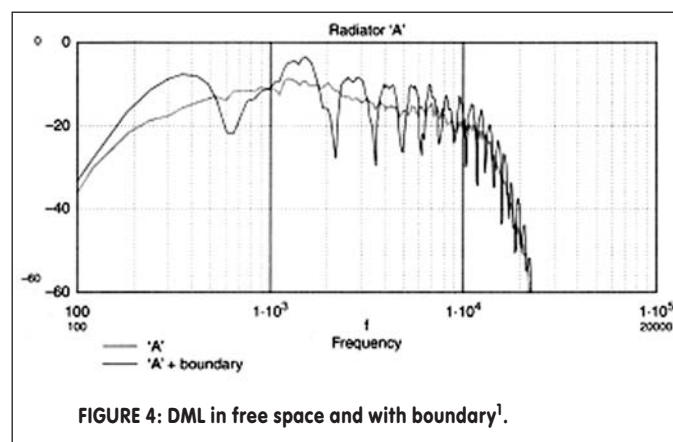


FIGURE 4: DML in free space and with boundary¹.

DML, compared to a traditional loudspeaker (**Fig. 3**), there is insignificant correlation between the frequency response measured in free field compared to the frequency response measured with a nearby boundary (**Fig. 4**). This results in improved clarity and intelligibility.

- Maintains loudness better with distance—The SPL reduction of a DML does not follow the inverse-square law. For moderate distances (e.g., 3' ~ 16' or 1m ~ 5m), the sound pressure level falls roughly as a linear function of distance². Loudness drop in sones from the front to the back of a room is less for a DML as compared to a conventional direct-radiator loudspeaker. Improved distribution of loudness is also observed for the DML¹.

- Lower distortion¹—Because the DML is a velocity device, the sound pressure is proportional to mean panel velocity, whereas in a conventional moving-coil loudspeaker, the sound pressure is proportional to acceleration. Thus, at low frequencies, the DML has lower distortion due to lower displacement. For optimized DML designs, the distortion products are typically 10dB lower than a pistonic loudspeaker.

- Acoustic feedback margin¹—In the presence of microphones, the feedback margin in sound reinforcement systems using DML technology is higher than that of conventional cone loudspeakers.

- Electroacoustical efficiency²—The real-world electroacoustical efficiency compares well with moving-coil systems (0.5% ~ 2%).

- Shallow depth—DML construction is generally shallower than conventional direct-radiator moving-coil loudspeakers.

DESIGN

Here are a few points regarding how to optimize a DML design:

- The selection of a light and stiff panel, implying the use of custom composites, will ensure the best results.

- The objective in the design of distributed-mode loudspeakers is to maximize the amplitude and density of resonances.

- For optimal results, the location of the drive point should couple to all the panel modes while ensuring de-correlation between panel velocity and drive point¹.

Si-5, which manufactures the SoundPad exciters, recommended foam-core posterboard as the panel material. Although a professional DML engineer recommended against using foam-core posterboard due to its less-than-ideal

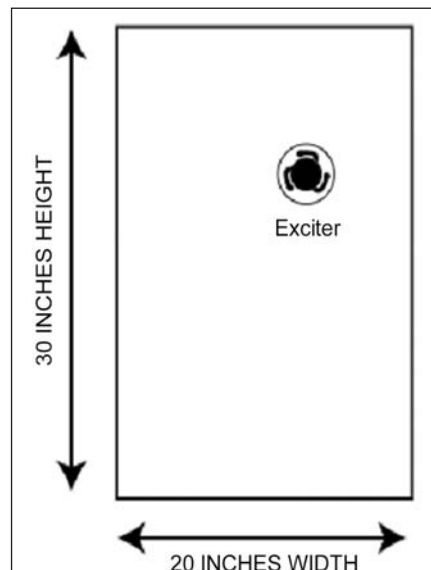


FIGURE 5: Foam-core posterboard and exciter diagram.



PHOTO 1: Foam-core posterboard with SoundPad exciter attached.



PHOTO 2: SoundPad exciter.

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high-frequency response characteristics, you cannot purchase better panel materials such as Acoustic M1 and Rohacell foam at local office-supply stores, which precluded their use for this project.

I used the following items to construct the DIY DML:

1. Foam-core posterboard—20" × 30" (50.8cm × 76.2cm), 3/16" (5mm) thickness.
2. Sonic Impact SoundPads.

The 20" × 30" foam-core posterboard

(**Photo 1**) is available for around \$5 per sheet from most major office-supply stores. Art-supply stores may have these posterboards for even cheaper. For the exciter, the Sonic Impact SoundPads (**Photo 2**) are available for \$20 from Sonic Impact Technologies (www.si-technologies.com). I actually found my SoundPads at a local Target discount store.

Because I do not have access to the FEM (Finite Element Modeling) software by NXT called PanSys, which can perform exciter drive-point optimization

given the panel properties, I determined the drive point by a heuristic method of DML theory, subjective impressions, and acoustic measurements. The SoundPad instruction manual provided a place to start: "Press the SoundPad, before exposing the sticky surface, against

one of the intended speakers and move it around until you are satisfied with the sound balance of treble and bass."

The objective here was to try to get the best distribution of modes horizontally and vertically. The drive point of the exciter for this project was based on moving the exciter (SoundPad) around the foam-core posterboard while listening to familiar pieces of music (**Fig. 5**). I next performed acoustic measurements based on subjectively preferred exciter locations for confirmation. These preferred exciter drive-point locations and subjective comments are as follows:

- Exciter 1/2 width, 1/3 height—Good high-frequency response, but low frequencies suppressed.
- Exciter 1/2 width, 1/2 height—Good low-frequency response, but high frequencies suppressed.
- Exciter 1/3 width, 1/3 height—Best subjective balance of low and high frequencies.
- Exciter 3/7 width, 4/9 height—This well-defined drive point recommended by NXT³ seemed to sound just as good as the 1/3 width, 1/3 height exciter position.

In the end, I settled on the final exciter drive-point location of a 3/7 width, 4/9 height location, as the subjective performance was just as good as the 1/3 width, 1/3 height exciter position with less peaking in the measured high-frequency response.

CONSTRUCTION

Construction was easier than any speaker I've ever built in my life, past and future (probably). After exposing the adhesive surface of the SoundPad, it was attached to the pre-determined location on the foam-core posterboard. Ta da!

MEASUREMENTS

I performed all measurements with the 3/7 width, 4/9 height exciter drive-point location. The DIY DML was suspended (not clamped) from a speaker stand.

Figure 6 displays the 1/6-octave smoothed full-range frequency response of the DIY DML at 1m and 2.83V drive level. Because the frequency response is far from flat (27Hz~11.5kHz ±7.5dB), it is difficult to ascertain the nominal voltage sensitivity for the DIY DML. However, it can be said that at least for this

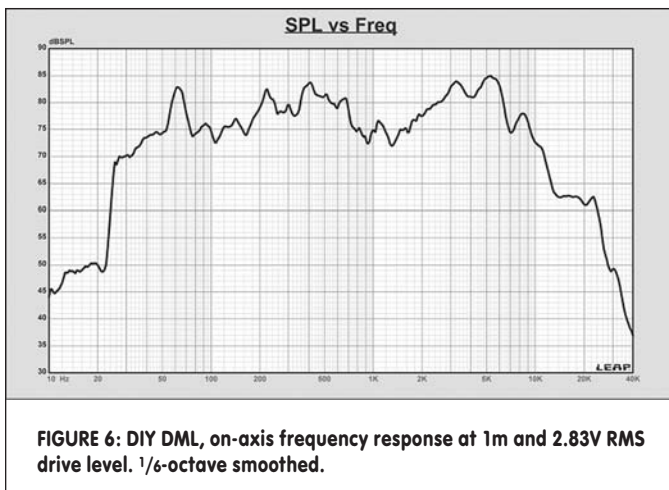


FIGURE 6: DIY DML, on-axis frequency response at 1m and 2.83V RMS drive level. 1/6-octave smoothed.

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example of a distributed-mode loudspeaker that its voltage sensitivity falls shy of most two-way bookshelf speakers. Although the low-frequency response is excellent, the high-frequency response drops precipitously above 10kHz.

Figure 7 displays the on- and off-axis (15°, 30°, and 45°) frequency response. Between 500Hz and 10kHz, you can observe that the frequency response does not change dramatically. Above 10kHz, the frequency response becomes more directional as the responses for the 30° and 45° curves are gradually more attenuated.

Figures 8 and 9 display the 360° polar response curves for the horizontal and vertical orientations of the DIY DML. The curves span from 700Hz to 20kHz. Generally, you can observe omnidirectional behavior between 1kHz and 7kHz for both the horizontal and vertical polar responses. You can see for both horizontal and vertical polar responses that omnidirectional behavior does not hold true at frequency extremes. **Figures 10A and 10B** are plots of the horizontal and vertical directivity index (DI) and is another method to view the polar data in **Figs. 8 and 9**.

Figure 11 displays the impedance response of the DIY DML, and it looks nothing like a conventional moving-coil loudspeaker! The impedance plot is rife with multiple resonances, which is expected behavior in a distributed-mode loudspeaker. It is surmised that the peak lowest in frequency (26Hz) is the fundamental resonance of the DIY DML. The modal density increases at higher frequencies, which you can see by the multiplicity of peaks in the impedance response. Above 2kHz, the impedance response increases dramatically due to the inductance in the voice coil of the SoundPad exciter.

SUBJECTIVE IMPRESSIONS

Equipment used:

- Sony DVP-650V SACD changer

- Onkyo TX-SR501 receiver
- Cheap interconnects and speaker cables

Most of the listening was done with the Onkyo receiver's 12dB/octave Dolby Digital high-pass crossover set to 80Hz, because sending a full-range signal to the DIY DML tended to overload the exciter and panel, which resulted in perceived harmonic distortion. Despite its 25Hz low-frequency extension, the DIY DML was not capable of playing low frequencies at realistic levels. Alleviat-

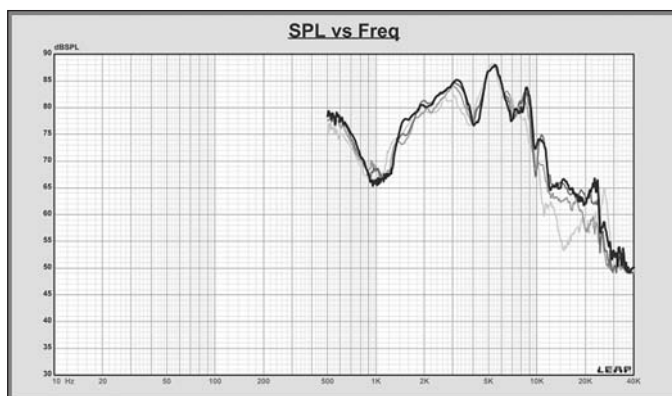


FIGURE 7: 0°, 15°, 30°, 45° on- and off-axis frequency response at 1m and 2.83V RMS drive level.



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ing the DMLs of low-frequency duties and having a subwoofer take over may be more effective in full-range sound reproduction.

It is important to note, though, that some home theater receivers, for some reason, disengage the high-pass crossover to the main speakers in stereo (2-channel) mode. Fortunately, the Onkyo does not, but if this is the case, you can try the Dolby Pro Logic "Music" modes, which may restore high-pass crossover functionality.

The following is a list of stereo music tracks I used during the listening session, followed by comments:

Flim & The BBs, *Tricycle*

- Very nice visceral low-frequency transient response and extension
- Lacking some high-frequency extension

Paula Cole, *Tiger*

- Excellent low-frequency extension but levels must be kept low to prevent

distortion

- Again, lacking some high-frequency extension

M-Pact, *Prince Medley* (recording not commercially available)

- Somewhat defined soundstage, doing justice to the original recording which was done, I believe, with five Shure SM 58 microphones panned across the five a cappella singers

Madonna, *Holiday*

- QSound reproduction was accurate (i.e., synthetic widening of soundstage)

Rebecca Pidgeon, *Spanish Harlem*

- The acoustic double bass tends to overload the DML panel and exciter at realistic levels, even with the 80Hz high-pass crossover
- The phantom image was solid even when I moved my head $\pm 15^\circ$ horizontal from between the speakers, demonstrating relative insen-

sitivity to listener placement

Ludwig van Beethoven, *Symphony No. 9 in D Minor ('Choral'), Op. 125, Movement 4: Ode To Ludwig* as played by Don Dorsey

- The transient attacks in this piece were very accurately reproduced

CONCLUSION

This article examines an approachable method to distributed-mode loudspeaker (DML) construction and evaluation. For the cost of two CDs, you can construct a pretty decent example from readily available parts. What is most impressive about the DIY DML is its stable imaging and visceral transient response.

Although my example was not faithful to the best examples of DML speakers, this is unsurprising, because I could not utilize the best materials and computer optimization. The future holds promise for this new technology, and I hope to see more applications for home, portable

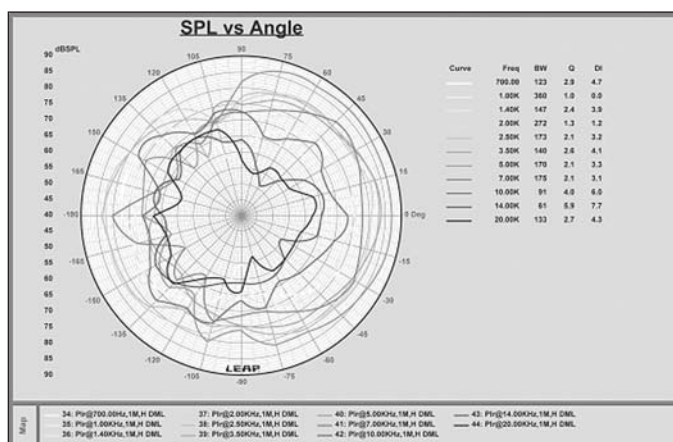


FIGURE 8: Horizontal polar response.

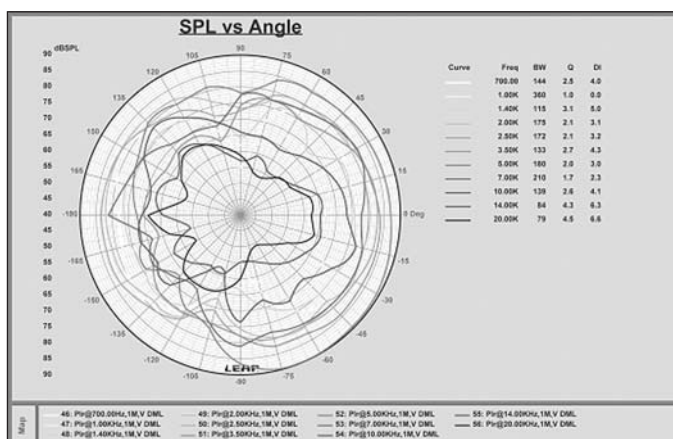


FIGURE 9: Vertical polar response.

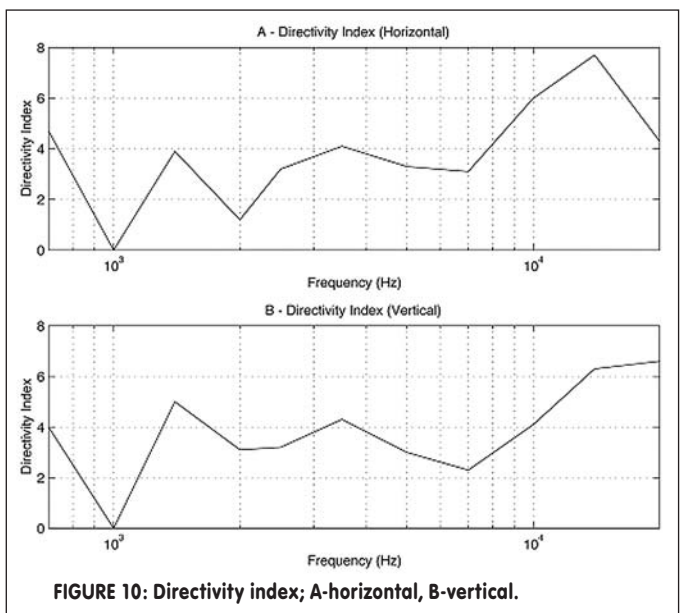


FIGURE 10: Directivity index; A-horizontal, B-vertical.

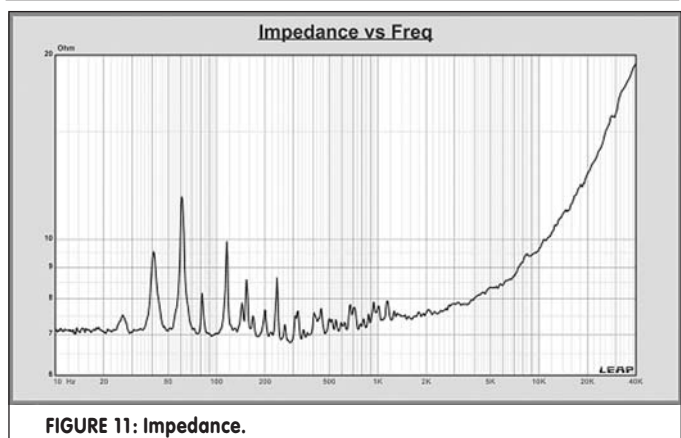


FIGURE 11: Impedance.

sound, personal computing, automobile, and sound reinforcement. **ax**

Daisuke Koya is currently on leave from his Ph.D. in audio engineering due to Chronic Fatigue Syndrome, which has severely affected his health since 2000. When his health allows, he works at HSU Research, a speaker company located in California. In his spare time, Daisuke enjoys attending museum exhibits and drinking espresso.

I would like to thank HSU Research for allowing use of their measurement facilities, and I also thank Vian Li for providing insight into DML theory and design.

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Improving the Cross-Coupled Inverter

A simplified version of the Van Scoyoc Circuit of 1948.

By Craig Kendrick Sellen

In these days of extreme amplifier bandwidth, a phase inverter with good balance and low reactive phase shift over a wide frequency range becomes a necessary part of any amplifier design. The circuit that most nearly meets these requirements is the cross-coupled inverter developed in 1948 (**Fig. 1**).

FEATURES

Phase shift is held to a minimum by the small number of phase shift points within the inverter. The low output impedance of the first stage minimizes high-frequency shift, and direct coupling eliminates it at the low end. The only point of low-frequency shift is in the output capacitors and the following grid resistor; the bulk of the high-end shift is also due to the 30k Ω output impedance of this stage.

Theoretically perfect balance is obtained by applying the same signal to different elements of the inverting triodes as shown in **Fig. 2**. Other circuits either depend on inherent imbalance and feedback for operation (cathode-coupled and floating paraphase) or require precision components for approximate balance (split load and voltage dividing). Another feature of this circuit is constant balance at all frequencies. This occurs because both halves of the signal pass through the same number of tubes and networks and both signal outputs are taken from similar elements of the output tubes (as opposed to the split load).

There are, however, some incorrect notions about the balancing in the cross-coupled inverter. First, with a single-ended input, the balancing control between the cathodes of the first stage does

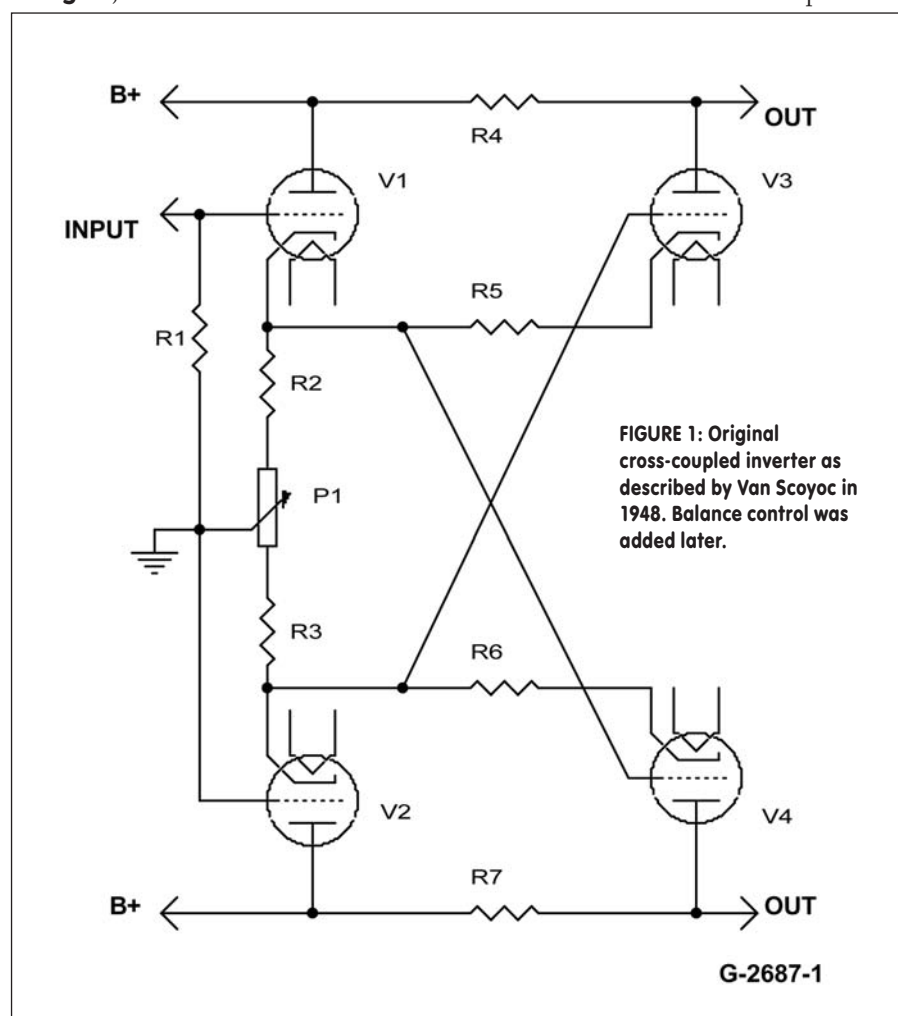
not balance the dynamic characteristics of the amplifier. It merely balances the tube bias and equalizes the sensitivity of the two inputs. Because neither of these is very critical in the improved model of the inverter, you can omit this control, as it was in the original Van Scoyoc circuit. Provision for dynamic balance is included elsewhere in the new circuit.

Also, the term “inherent balance” is somewhat misleading. Referring again to **Fig. 2**, note that both halves of the

signal are equal only at the input of the second stage. This leaves variations in tubes and load resistors of the second stage, thus upsetting the balance. Another problem with the design is the high (30k Ω) output impedance, which limits the high-frequency response and causes clipping as grid bias reaches zero in the power output stage.

CIRCUIT SOLUTION

You can solve both of these problems



simply by adding another cathode follower lower to the circuit. Because of the low DC voltage at the output of the cathode follower, it can be direct-coupled to the power stage without many of the problems of bias and static balance that usually accompany direct coupling. Direct coupling here eliminates a low-fre-

quency phase shift point, thus maintaining the stability of the amplifier. The low output impedance reduces high-frequency shift and allows some power to be delivered to the power stage.

It can also be shown by the design equations for cathode followers that the gain of the stage can be varied by vari-

ations in the cathode resistance. This provides a very convenient method for balancing the dynamic characteristics of the amplifier, as shown in **Fig. 3**. Operation of this control will cause a slight static imbalance in the power stage. It should be negligible, but the purist may compensate for it by a balance control of the power stage, if one is provided.

The main fault of this inverter is its low gain due to the two cathode followers. Most modern tuners and preamps, however, deliver enough voltage to drive it to full output. In general, this inverter should satisfy every requirement in amplifiers where less than 15V drive is needed at the power stage grids. **ax**

Craig Kendrick Sellen, of Carbon-dale, PA, has written dozens of articles on the field of electronics. These articles have appeared in both American and foreign periodicals, and on the Internet. He was trained in solid-state, but has a side interest in vacuum tube circuits of all types. Craig has his own website at www.msnusers.com/103pennywhistler, and his e-mail address is kendricksellen@hotmail.com.

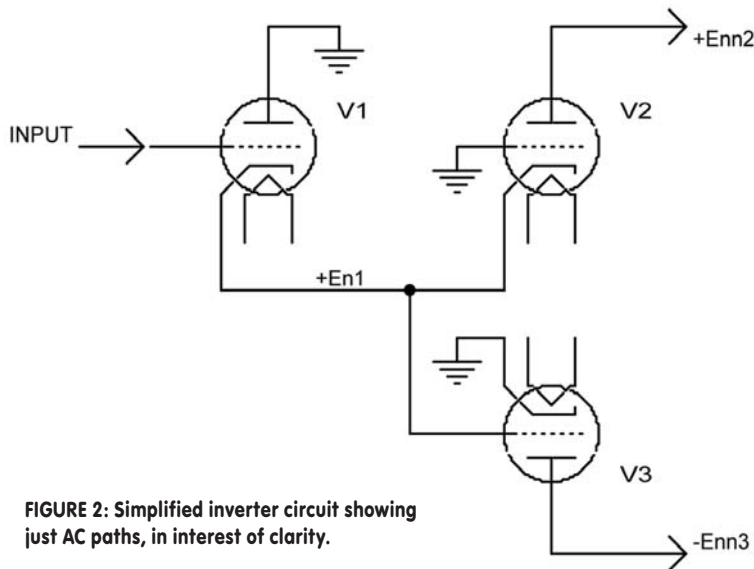


FIGURE 2: Simplified inverter circuit showing just AC paths, in interest of clarity.

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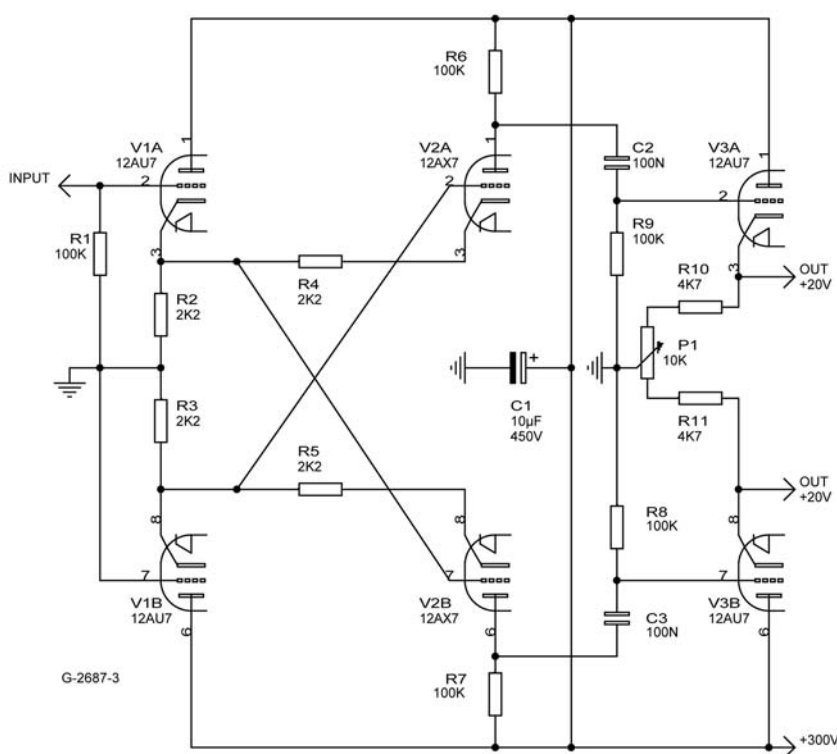
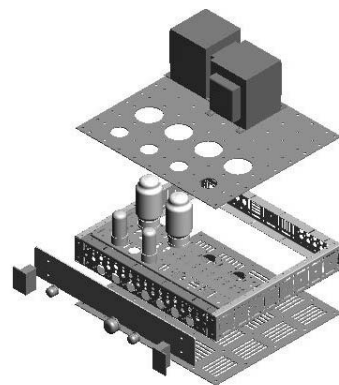


FIGURE 3: Cross-coupled inverter with cathode-follower output. Output leads should be connected directly to power tube grids. Cathode voltage of power tubes must be increased by 20V. Balance control illustrated will compensate for as much as a 6% imbalance.

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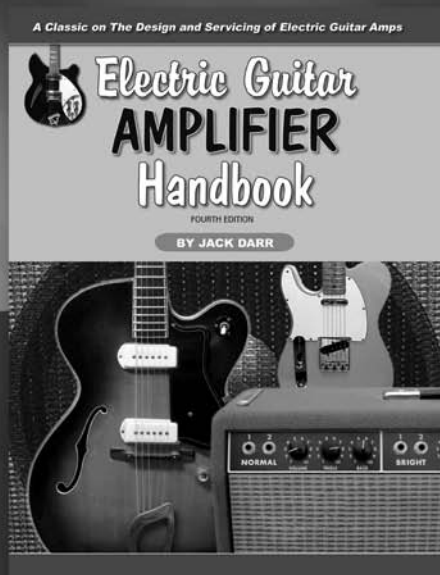
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GREAT SCOTT!

Charles Hansen's extensive refurbishment and modification of his Scott 222C (May–Sept. '06) is a wonder and a great credit to him. The finished job looks just gorgeous and I bet the sound is vastly improved.

However, the task looks most daunting. Can he tell us about how many hours he put into this work? I have a 222C, but I'm not sure I could last the distance!

*Bart Shepherd
Epping, NSW, Australia*

Charles Hansen responds:

Thanks for the compliment, Mr. Shepherd. I have no idea of the man-hours I put in. I started in November 2004 and finished in October 2005, but it was only a part-time effort. Much of that time was spent waiting for parts, manuals, investigating design changes, computer simulations, and so on. If you want to do the minimum amount of work and just restore the 222C to original working condition, I suggest the following:

- Check all the solder joints and touch up those that need it.
- Replace all the ceramic-tube paper American Ceracaps with modern polypropylene film caps.
- Replace the selenium bias supply rectifier with a silicon rectifier (mandatory!).
- Replace the aluminum can capacitors with modern versions made by CE manufacturing.
- Replace any obviously overheated resistors and wax-coated capacitors.

PUSH-PULL AMP

I greatly enjoyed the tube construction article in the August issue by Claudio Rosada ("Tuono: PP PL504 Amplifier," p. 18). Although this is a carefully designed, meticulously tested project, I did find one error.

Mr. Rosada's ideal output transformer would load each tube with 1200 Ω , making the total plate-to-plate load 4800 Ω , not 2400 Ω , as stated. The reason for this is that transformers follow a square-law relationship for impedance. A push-

pull transformer has twice the number of turns as a single-ended one, so the plate-to-plate impedance is 2 squared, or four times the impedance of half of the winding.

In other words, each tube sees one-quarter of the plate-to-plate load impedance. Mr. Rosada's actual transformer (2750 Ω) will present a load to each tube of 687 Ω . This is lower than the target value of 1200 Ω , but is not too serious, as the ideal load should be about twice the plate resistance which, for a PL504, is about 350 Ω . The actual measurements seem to vindicate Mr. Rosada's design.

Also, the PL504 is somewhat hard to find in this country. The electrically similar PL500/27GB5 is inexpensive and readily available, but I don't have detailed specifications. Does Mr. Rosada have any experience using this tube?

*Bob McIntyre
musitronix@yahoo.com*

Claudio Rosada responds:

In order to answer Mr. McIntyre, I must briefly review the theory of push-pull amplifiers.

First of all, you must consider the total dynamic load applied to each tube in the push configuration; in fact, from the perspective of one primary winding, there are two "secondary" windings. The first one is the real secondary winding, and the second one is the other half of the primary.

So the dynamic load seen by one tube will be the parallel of the secondary load transferred to the primary and the dynamic load connected to the second primary winding. The final result will depend on the load connected to the other primary winding.

You can consider two specific types of this load (other possibilities are less interesting):

- 1) The load is another tube with the same bias current, driven by the same input signal in opposite phase.
- 2) The load is a DC current source and its value equals the bias current of the tube.

The first case corresponds to a push-pull stage polarized in class A. It means the bias current is sufficiently high that each tube

never reaches the cutoff condition when the maximum signal is applied to the grid. In this case the signal path connects the two tubes "in series." Together they see a total load corresponding to the plate-to-plate load, so each tube will see just one-half of this load.

The second case corresponds to another two possible applications of a push-pull transformer:

a) A single-ended stage is driving one primary winding, while a current source (e.g., a pentode) is connected to the other primary winding in order to perfectly balance the DC current. With this configuration you can use a push-pull transformer in a single-ended application.

b) Two tubes are connected in a push-pull stage polarized in class AB1. For low signals the stage works as previously stated, but when the input signal exceeds a certain limit, one of the two grids is so negative that its tube goes in cutoff and becomes dynamically an infinite load. The tube bias point will define the signal level at which this condition will occur, and from now on the dynamic load drops to one-quarter of the plate-to-plate load.

Tuono is polarized with high current in class AB1, so it works in pure class A for a big part of the total signal dynamics. For simplicity in my article I was referring only to this condition.

As long as the tube stays in the A region, the effective load is one-half of the plate-to-plate load, while when one tube reaches the cutoff, the load effectively drops to one-quarter of the plate-to-plate load.

In **Figs. 1** and **2** the output signal (upper track) on 8 Ω load and the cathode current (lower track) of one tube are shown with a 1kHz input signal at different levels. **Figure 1** corresponds to 1W output power, and a perfect class A is visible. In **Fig. 2**, corresponding to 6W output power, the cathode current approaches the cutoff condition and is highly distorted. In this way the class B operation is used to provide an additional power boost during high peaks, while in normal condition both tubes are working.

The push-pull amplifiers theory is completely detailed in *Radiotron Designer's*

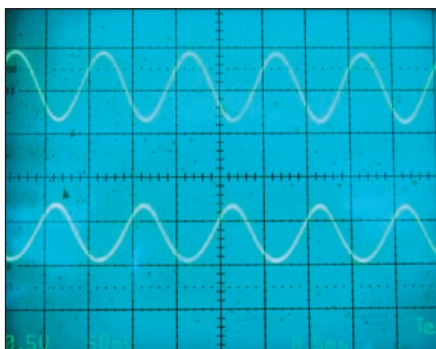


FIGURE 1: Tube measurement at 1W.

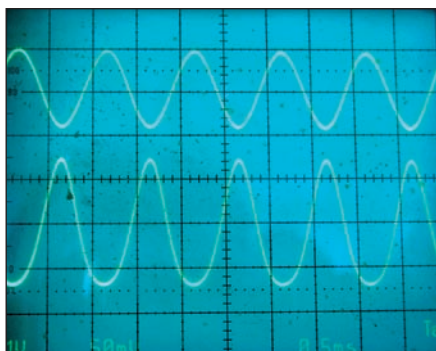


FIGURE 2: Tube measurement at 6W.

Handbook Fourth Edition, cap. 13.5, "Push-Pull Triodes Class A, AB1." PL500 data sheets are available in Frank's database at: <http://frank.pocnet.net>. PL504 looks to be an evolution of this tube. PL500 has a W_a power limit between 12W and 17W (the same values for PL504 are 16W and 22.5W), so you can expect a consequent reduction in output power when used in a similar circuit. I never used this tube.

TEST GEAR

The article titled Mad Katy by Dennis Colin (June '06) contains oscilloscope images showing tone burst responses. I am interested in finding out what equipment the author used to generate the tone bursts, and would appreciate finding out the manufacturer and model of the equipment.

Tom Russell
Richardson, Tex.

Dennis Colin responds:

Thank you for your interest. I used an HP8116A pulse/function generator (vintage about 1980, when HP was HP!). I've seen used ones for sale, unfortunately for over \$1000. It has a gate and a trigger input, which allow clean tone bursts to be set up, that start at the sine wave zero crossing, then end after a number of cycles deter-

mined by the pulse width feeding the gate input.

The 8116A is expensive (new about \$40,000!) because it covers one cycle every 1000 seconds, to 50MHz. You should be able to find a sine wave generator covering the audio range, with a gate input that's zero-crossing synchronized, for much less cost. Also, you could use an analog switch IC to pulse any sine wave source from any pulse or square wave source. To make the tone burst start and stop at the zero crossings, the pulse would need to be triggered or synchronized from the sine wave's zero points. An open-loop op amp would turn the sine wave into a square wave (or use a sine/square wave generator); the square edges could then be used to trigger the pulse generator.

I can't afford expensive equipment—I was lucky to find a sale of the 8116A, the Tek 475 scope, and more for about 2% of the original value.

TRANSFORMER MECHANICAL NOISE

While I am an avid reader of *audioXpress*, I don't do all that many construction projects. When I do undertake a



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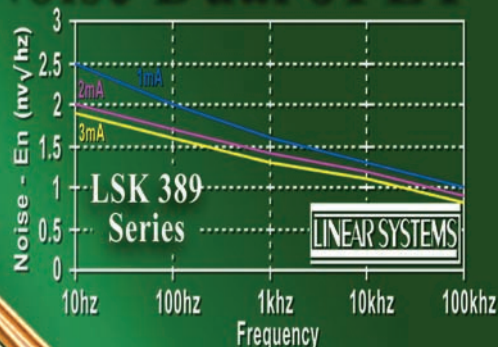


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project, I've found that the most difficult part to source is usually the power transformer. While it's easy enough to choose a unit that will do the job, there is no way to judge how much mechanical noise it will generate.

Many manufacturers use epoxy filling and other methods to "limit mechanical noise," but the only way to actually determine which units are noisy is to buy them and see (listen). To me, nothing ruins a nice quiet amplifier worse than a transformer that hums so loudly you have to shut it away in a closet. I would love to hear from other readers about their good (and bad) experiences with noisy transformers—especially from the vendors who advertise in *audioXpress*.

Keith Levkoff
kLevkoff@panix.com

CORIAN SPEAKER?

I would like to congratulate Peter Smith on a well-executed design and clear, thorough write-up ("NeoNoir: A Ported Two-Way," 9/06 *aX*). Not having enough time to work on my own designs, I enjoy living vicariously through the adventures of others; I found myself agreeing with just about every one of Peter's design decisions.

The Corian baffle was a tease though. Ever since remodeling my kitchen with the stuff, I have wanted to build a Corian speaker cabinet. Being the rock equivalent of MDF, it seemed that it should be denser and more passive than wood products.

Could Peter give us some details on any audible improvements with the Corian baffle? What tools were used—are carbide router bits OK for Corian or do you need special cutters? How thick was the Corian lamination? Any special finishing for the rounded edges?

iajn@impaudio.biz

Peter J. Smith responds:

Thank you for the kind words on the article. Unfortunately, I did not do the millwork on the Corian baffle. A speaker builder friend kindly offered to cut the baffles for me out of sink cutouts he had lying around. I do know there are a number of special blades available for Corian and people have mentioned slow cutting speeds. I also know there is

finish work starting with 220 grit wet paper and progressing to rubbing compound.

I did some basic drilling into my baffles and found the Corian brittle with a tendency to chip around the entry and exit points of the drill. Placing masking tape over the holes and drilling through the tape helped. I also struggled with scratches whenever hard tools got near the Corian. I would head to the Internet for specific guides on working with Corian.

As for sound and stiffness, the NeoNoir has the Corian glued with Liquid Nails (flooring) to a 3/4" piece of MDF. This makes for a stiff composite baffle, which always helps sound quality. I would not build an entire speaker cabinet out of Corian, because I would not trust the glue to hold, plus Corian is only 1/2" thick. The harder and thinner material may actually be more excitable than regular MDF. So, the bottom line is that Corian was a nice experiment, but I am moving back to wood.

PREAMP ERRORS

The preamp construction article by Cornelius Morton in the September issue of *audioXpress* has four errors that I have found. There may be more, but I am not experienced enough to have seen them. The Figure 3: Power Supply Schematic, and the Figure 5: Front and Rear Panel Layout, are switched from their accompanying diagrams. On page 6, under Line Stage: "The output of V2A is coupled through C7,..." should be C8. On page 7, center column: "A full-wave bridge rectifier, D1, rectifies the 6.3V AC, R1 limits..." should be R5.

I am also interested in power-supply changes the author might suggest were I to build this as a high-level only preamp, eliminating the phono stage tube requirements.

Steve Perlmutter
stavis@chilitech.net

Cornelius Morton responds:

Thanks for your interest and inputs. Yes, the captions for Figs. 3 and 5 are reversed. I noticed the same; links between the images and caption are correct.

C7 should be C8 as you indicate. The bridge diode limiting resistor should be R5, not R1, which was properly identified earlier. Concerning power-supply changes required to run the line stages only, they are relatively simple.

B+. Delete C11 and R21 for both channels. The voltage will increase a couple of volts but this will not be a problem.

Filaments. You may need to adjust R5 but the testing procedure covers this. I would expect that R5 will be 0.47Ω.

The line stages on this preamp are easy to listen to. Distortion is low and increases linearly with level as noted in the article. The distortion curves taken for different frequencies are virtually identical. There are no bumps or discontinuities in these curves that would tend to grab the ear. I have recently run a frequency analyzer on this. The output at 0.7V RMS indicated 0.15% second harmonic with a third harmonic visible at 30dB below the second. Nothing else was visible.

TUBE PREAMP TIPS

As an avid reader and author dating back to *Glass Audio* days, I read Cornelius Morton's article (*aX* 9/06) with interest. It is a very well-written and thorough article, even detailing the construction details.

Because I have built and tested many of these valve phono preamps, I have a couple tips that will even further improve the preamplifier design:

1. Operate the first stage with a grounded cathode. Check out the curves (<http://tdsl.duncanamps.com/pdf/vm322.pdf>), note the improved linearity when operating at zero volts! Operation with the grounded cathode has two advantages; first it is simpler, eliminating the (cough, cough) electrolytic capacitor in the signal path, and second, the gm is maximized while the plate resistance and the resulting gain is maximized. The result is better sound and lower noise, and you save money to boot!

2. Add a bleeder resistor to the output. 330-470K is fine.

3. Add local bypass of film capacitors at the plate resistors.

4. For the second stage it is better to direct-couple to the first stage. It eliminates parts in the signal path to improve the sound. From the curves you see that the first stage is going to run at about 35V. If you wish to run about 10mA in the second stage, the Rk is 3.3k (getting rid of C4 and R7). Then you set your first zero for the bass rolloff with C5 at 30Hz = 10-15μF. Use a tantalum or low ESR electrolytic

for this; 50-100V rating. Again bypass with a 1 μ F film cap.

David Wolze
dwolze@comcast.net

Cornelius Morton responds:

Thank you for your interest and comments. I have a few comments regarding your suggestions which I will try to keep in the order they occur.

Grounded cathode operation, zero bias. This would raise the possibility of a bit of grid current on heavy passages, which would be mitigated by the relatively low DC resistance of a moving magnet cartridge. The operating point of 35V DC E_p and 10mA I_p reflects a gm of 11500 S with an equivalent noise resistance of 222 Ω , $R_n = 2.5/\text{gm}$, and a best case noise floor of -85.3dB referenced to a 5mV input signal. For a gm of about 9900 S the equivalent noise resistance is about 254 Ω , resulting in a best-case noise floor of -84.5dB referenced to a 5mV input signal. The measured noise was better than -80dB with the phono input open and measured at the preamplifier output; shorting the input might gain a dB. Actually, I thought that getting that close was not too bad. The best noise floor would be -88.5dB, achieved by paralleling the two sections of the 6922 at an E_p of 35V DC and a total I_p of 20mA, though for safety I would still like cathode bias.

Eliminating the electrolytic. Current electrolytic design maintains capacitance well into the megahertz region and shows low hysteresis losses when the polarizing voltage is large compared to the signal voltage as is the case here. With the low signal levels involved, I believe that the differences in distortion would be very small and offset by the self-bias correction effects for tube aging.

The bleeder is there in the form of the 250k phono level control.

You could add local bypass caps, but again the current crop of electrolytics exhibit low levels of inductance and maintain capacitance over a wide range of frequencies. Looking at the first phono stage, the plate load R2 and C11 form a voltage divider from the plate to ground for the AC signal. R2 has a value of 8200 Ω , while the impedance of C11 is only several ohms at best and usually much lower. For a plate signal of 1V AC the resulting voltage across C11, at an X_c of 2 Ω , would be about 0.25mV. If the impedance of C11 increased by 100%

that voltage would only increase to 0.5mV. In either case it is good filtering.

DC coupling between the two phono stages. I considered and then discarded this for several reasons. To maintain plate voltage for the second stage, the source voltage would be increased by the plate voltage of the first stage—35V DC for your suggestion and 60V DC for the design. This would complicate the power supply and make it nearly impossible to provide sufficient positive bias to the filaments to prevent filament to cathode conduction without exceeding filament to cathode voltage specs or really complicating the filament circuits. Your desire to replace C4 and R7 with a tantalum C5 is not good. Tantalum caps are about the most nonlinear caps made by man.

All resistors in this design are 1% metal film and all non-electrolytic caps are 5% film except the caps in the RIAA compensation circuits which are 3%. The values and accuracies of the RIAA circuit components are critical in maintaining a $\pm 0.25\text{dB}$ accuracy in the RIAA compensation. A good reason to keep C4 and R7.

One way of slightly decreasing distortion and decreasing cost would be replacing all of the line-level controls with 58k resistors and the phono level with a 470k resistor. Then move the volume control to the input of the first line stage with it being driven with the selector switch output. This would require volume adjustment between sources which was not my original intent.

Thanks again for your interest.

BASS RESPONSE EQUALIZER

I enjoyed, and concurred with most of, George Danavaras' article on low-frequency electronic equalization of rooms (Aug. '06, p. 36). A couple of comments:

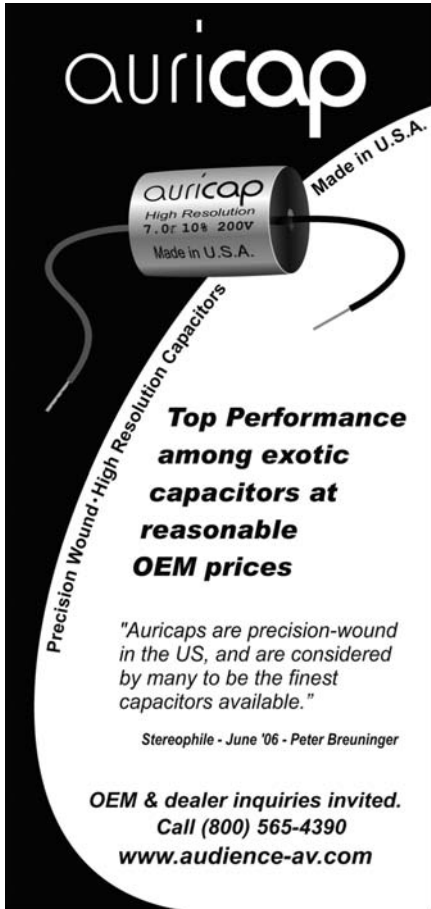
The equalizer seems to assume a monaural signal source. This source might be a simple passive mix of right and left line-level channels, or maybe is already a dedicated subwoofer output from a pre-amp or receiver/amp. In any case, there should be an awareness of the significant loading of some sources by the rather low impedances at the input of the equalizer (R1, R2, C1, C2) which for many line-level sources would attenuate the high frequencies in the source signal—desirable for the sub channel but not for the signals potentially used for the full-range ones. If the input is a dedicated LF channel, then you're likely alright—but then

the high-frequency signals are probably not present to begin with.

He states that the low-frequency cut-off due to C3 is "about 1Hz." I calculate about 18Hz (-3dB) due to the loading by R3 and the attenuator network in series to common, a mite high for a subwoofer. If you moved C3 inboard a bit and placed it between the attenuator taps and the 160k, R40, the -3dB point would instead be at about 3Hz.

When I read "self-filter" and "self equalizer," I thought for a moment I'd missed some eponymous trick circuit of Doug Self's—but inspection of the schematic revealed that it should read "shelf equalizer."

The phase change circuit can indeed be useful, but you should understand that, for that simple circuit, the phase shift is frequency-dependent—so the description given of phase versus potentiometer setting is correct at only a particular frequency. As well, for sufficient generality, I'd recommend the inclusion somewhere in the chain of a polarity inversion capability—although, if you have ready access to them, flipping the wires at the sub driver itself could do the job.



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Or, for that matter, if you are not concerned with absolute polarity, you could flip relative polarity by reversing the connections at the full-range speakers.

The one-op-amp synthetic inductors are tried-and-true topologies, although they do tend to be a bit noisy as the $R4/R3$ ratio becomes large—typically this will not be a problem for low-frequency processing because the noise bandwidth is so small. Among other options, circuits based on GICs (General Imittance Converters) use two op amps per inductor but if done right have better noise performance.

Also, when concatenating sections of such *dip* circuits, with a lot more work you can often eliminate the buffers separating the sections, reducing the number of op amps in the signal chain.

All in all, the proof is in the listening, and supported by the helpful measurements. Thanks for all of the hard work, George!

Brad Wood
Canoga Park, Calif.

George Danavaras responds:

I would like to thank Brad Wood for the

Look!!

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interest in my article and the kind words.

Concerning his comments: It is true that I use the equalizer for a dedicated subwoofer, which is an open baffle design as described in the *audioXpress* article ("Double Dipole Subwoofer," November 2004). This is why the -3dB frequency for the low-pass filter which is realized with R1, R2, C1, and C2 is about 10kHz (supposing that the output impedance of the previous stage is about 100Ω). The low-pass filter at the input is also used to reject any possible EMI signals due to the long cable used to connect the equalizer to the system. When the equalizer is going to be used for the full audio spectrum, then the C1 and C2 should be decreased to about 470pF or 1nF.

Your calculation for the low-frequency cutoff due to the high-pass filter forming with the capacitor C3 and the attenuator network is correct. Initially I used a value of 4.7μF for C3 which gives a cutoff frequency close to 1Hz. I used the value of 330nF when I installed the open baffle subwoofer in order to eliminate the "hard drive" at the very low frequencies.

The correct expression is "shelf-equalizer." I am sorry for the confusion.

Although I included the phase change circuit in the design of the equalizer, later I found out that the use of such a circuit at least in my listening room is limited. So usually I bypass this circuit. Maybe it is more useful in other cases.

Finally, I don't measure any noise at the output of the equalizer that is due to the inductor realized with the op amp.

HEAD AMP DESIGN

I read with much interest the article by Paul Rossiter ("A Head Amp for Very Low Impedance Moving Coil Cartridges" *audioXpress* 9/06, p. 22) about the design of a high-performance head amp for moving coil cartridges. I have a few questions regarding the design and implementation.

1. I have not been able to find a source for the 2SC3329 and 2SA1316 transistors. Can you suggest a source for them?

2. Mr. Rossiter performed some performance analysis of the amplifier using SPICE. I have been unable to find SPICE models for these devices (also for the MPSA12 and MPSA62 Darlington transistors). Is there a way to obtain these models? I have access to SPICE, and if I have the models I can

analyze the effects of some changes I have considered.

3. One design change that I would like to model (and on which I would solicit Mr. Rossiter's opinion) is to replace the packaged Darlington transistors with a pair of complementary Darlington transistors (also known as a Sziklai pair) constructed from pairs of the 2SA1513 and 2SC3329 transistors. This complementary Darlington pair is constructed from an NPN and a PNP transistor as opposed to two devices of the same type.

An advantage of this configuration is a lower V_{be} drop (only one base-emitter junction in the compound base-emitter connection versus two for the MPSA12 and MPSA62 transistors). This will result in better V_{be} matching over temperature changes than the standard Darlington connection. This approach will require sorting and matching transistor pairs, but may result in improved performance. A concern here would be potential instability due to increased parasitic impedances caused by the compound connection of the two devices.

4. I would also like to explore the addition of a second parallel transistor to each of the input devices (thus reducing the effective value of r_{bb}). While theoretically capable of a 3dB improvement in input noise, I believe that because the parallel transistors are unlikely to be perfectly matched, the effect will not be as great as 3dB. Mr. Rossiter discusses this as a possibility with the potentially adverse effect of higher distortion; therefore, I would be interested in analyzing it before changing the design. I also believe that the values of R5 and R6 would need to be adjusted to change the emitter currents in the parallel pairs of transistors. Again, Mr. Rossiter's opinion would be of interest.

5. One last (small) point is that the schematic of Fig. 7 in the article seems to have the transistors in the MPSA62 Darlington pair (Q4) incorrectly connected. The two base-emitter junctions should be in series, not parallel, and the collectors should be in parallel, not series, I believe. This is a small matter because the devices are packaged together in one package. It is mostly a cosmetic concern.

Wayne Torrey
waphylz@cox.net

Paul Rossiter responds:

I thank Wayne for his interest and comments on the article. I will try to address each of the points he has raised.

1. I obtained the 2SC3329 and 2SA1316 transistors from Erno Borbely (who will be very familiar to the readers of *audioXpress* for his many excellent FET-based designs). I have checked with him and he still has a number of these parts available (sales@borbelyaudio.com).

2. The SPICE package I used (Micro-Cap 8) had the models for the 2SC3329 and 2SA1316 already included in the model library, so I just used those. For the MPSA12 and MPSA 62 I used the model for the MPSA77 also from the Micro-Cap 8 library, with only the gain and polarity parameters changed to match the MPSA12/62 data-sheet values.

Unfortunately, due to changed circumstances since I wrote the article, I no longer have access to the library and so cannot provide the actual model parameters. However, as I noted in the article, the (audio frequency) distortion results from the simulation were not particularly dependent upon the actual model used for the first transistors and so anything close, suitably adjusted for the rbb' and current gain of the 2CS3329/2SA1316 (given in *Table 1* in the article), should suffice. Similar comments apply to the Darlingtons.

3. Replacing the Darlington transistors with complementary feedback pairs (Sziklai pairs) is an interesting idea because the 100% local voltage feedback is then applied around the pair rather than each transistor separately. In amplifier output stages such an arrangement offers better large signal linearity than the Darlington equivalent and also better thermal stability because the base-emitter voltage drop of the output device is then within the feedback loop and only the base-emitter voltage drop of the driver affects the quiescent current (and this can easily be compensated). In the present application thermal stability is not an issue and the small signal distortion seems to be totally insignificant (and is certainly way below the noise floor). But please don't let these comments stop you from further experimentation (and the pursuit of perfection), and I would be keen to hear the outcome. I imagine that the high-frequency domain might need some attention to ensure that there is no instability.

4. I did discuss using paralleled transistors

for Q1 and Q2 with the attendant lowering in noise contribution. However, in this design the noise ultimately becomes determined by the feedback resistor (R2 in *Fig. 6*), and so the gain in overall noise performance to be obtained by paralleling the input transistors is rather limited (remember, there are already effectively two transistors in parallel in the design). As discussed in the article, you can obtain a 1.5–1.8dB reduction in noise by reducing the 10Ω feedback resistors R2/R3 to 4.7 or 3.3Ω. However, the attendant reduction in other impedances required to maintain the gain leads to an increase in distortion, and so (as is usually the case) there is a tradeoff to be made.

5. Wayne is quite correct about the error in the symbol for the MPSA62 (Q4) in *Fig. 7*. That will teach me to not simply use a library symbol that looks about right! However, as he says, it is of no dramatic consequence because the elements are (I hope!) properly connected within the package. *aX*



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Audio Aid: Build a One Transistor MC Step-up

By Rickard Berglund

With modern low-noise transistors, you can build an MC step-up with only one transistor. The MC step-up is powered by a 12V lead-acid battery. The value of R4 depends on the hfe of the transistor used.

Transistor	R4	hfe	Relative Noise for transistor
BC 639 Philips	180k	150	126
BC 487 Motorola	360k	320	85
BC 489A Motorola	180k	165	72
BC 637 Siemens	150k	130	80
BC 639 Motorola	150k	125	68
MPSA 06 Fairchild	220k	200	118
MPSA 42 Motorola	180k	160	74
2N4401 Motorola	220k	200	136
MPSA 42 Philips	180k	150	250

The brand of the transistor is very important. I found that Philips MPSA 42 and Philips BC 639 are 5-10dB more noisy than Motorola. Distortion of the amplifier is low due to local feedback. The gain of the amplifier is 19-20dB. If you have a problem with radio frequency

interference, place a 22nH inductance in series with C1 and a 22pF capacitor in parallel with R4. *ax*

COMPONENTS

C1 2.2μF polypropylene
C2, C3 1μF polypropylene
R1, R2 120Ω wirewound
R3 1200Ω

FIGURE 2: Pinouts for different transistors. Bottom view.

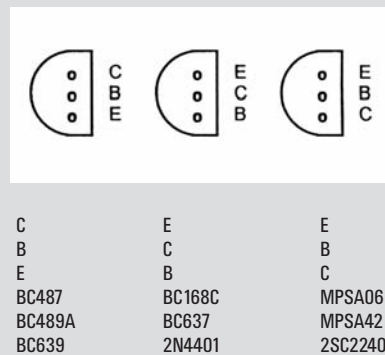
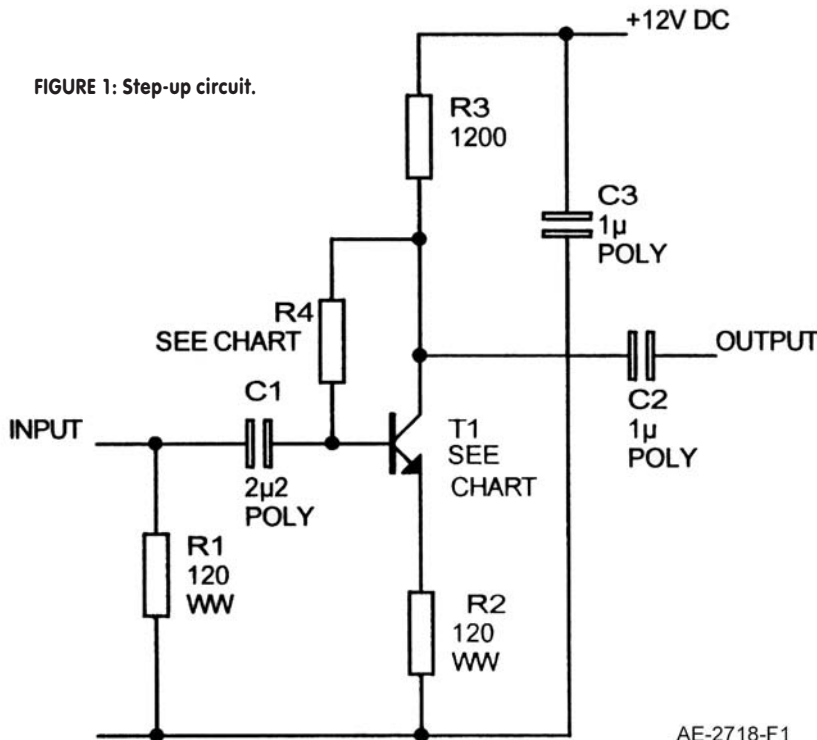


FIGURE 1: Step-up circuit.



CONTRIBUTORS

Edward T. Dell, Jr. ("Editorial: Undernourished Ears," p. 6) is the publisher of *audioXpress*. He has written many articles and editorials throughout the publications of the many magazines produced by Audio Amateur Publications.

Joe Sousa ("Precise dB Monitoring with Eye Tubes," p. 8) has written three design Idea winners with *EDN*, a feature article for *Radio Electronics*, and a design idea article for *Electronics World* magazine. This is his first article in *audioXpress*.

Dennis Colin ("The Venue Loudspeaker," p. 14) has been a band keyboardist and a recording engineer, and is a frequent contributor to *audioXpress*.

Charles Hansen ("A Phase Meter Calibrator," p. 22) holds five patents in his field of engineering and enjoys modifying guitar amplifiers and effects to reduce noise and distortion. With hundreds of articles to his name, he is a frequent contributor to the magazines of Audio Amateur Publications.

Joseph Norwood Still ("Two SE Power Amps: Triode or Ultralinear," p. 27), retired from the electronics industry, designs affordable high-quality audio amplifiers for the dedicated audiophile. He has written many articles for Audio Amateur throughout the years.

Daisuke Koya ("Build a Flat Panel Speaker," p. 34) is currently on leave from his Ph.D. in audio engineering due to Chronic Fatigue Syndrome, which has severely affected his health since 2000. When his health allows, he works at HSU Research, a speaker company located in California. This is his first article for *audioXpress*.

Craig Kendrick Sellen ("Improving The Cross-Coupled Inverter," p. 40) was trained in solid-state, but has a side interest in vacuum tube circuits of all types. He has written dozens of articles on the field of electronics, in both American and foreign periodicals, and on the Internet.

Rickard Berglund ("Audio Aid: Build a One Transistor MC Step-up," p. 54) has worked in research in analytical chemistry, and in industrial electronics design. He is now engaged in construction and manufacturing of tube amplifiers.