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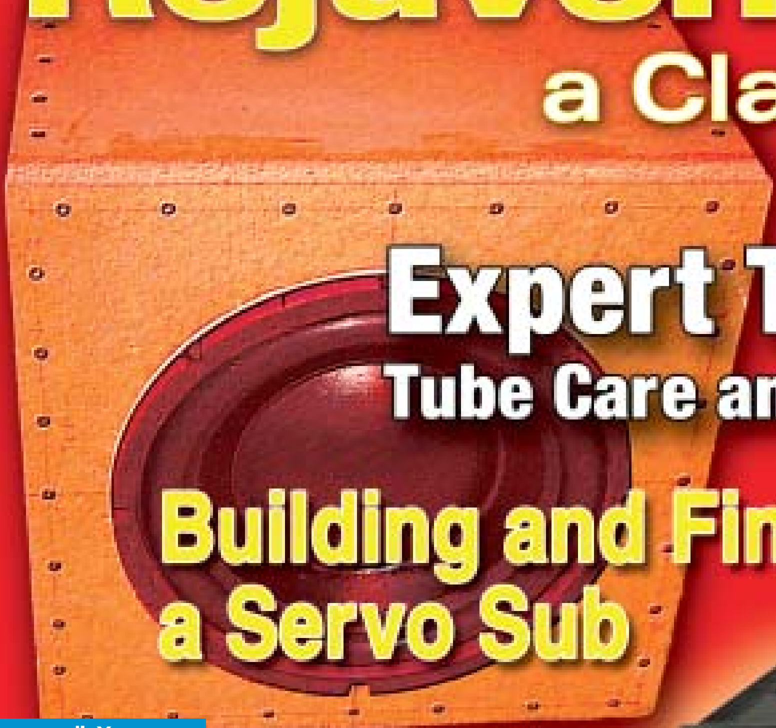

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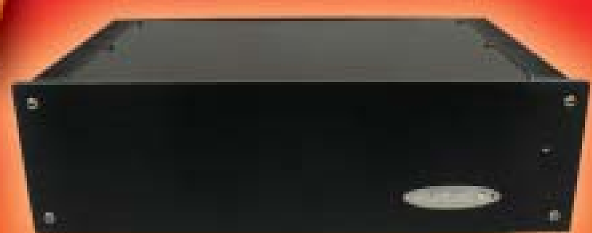

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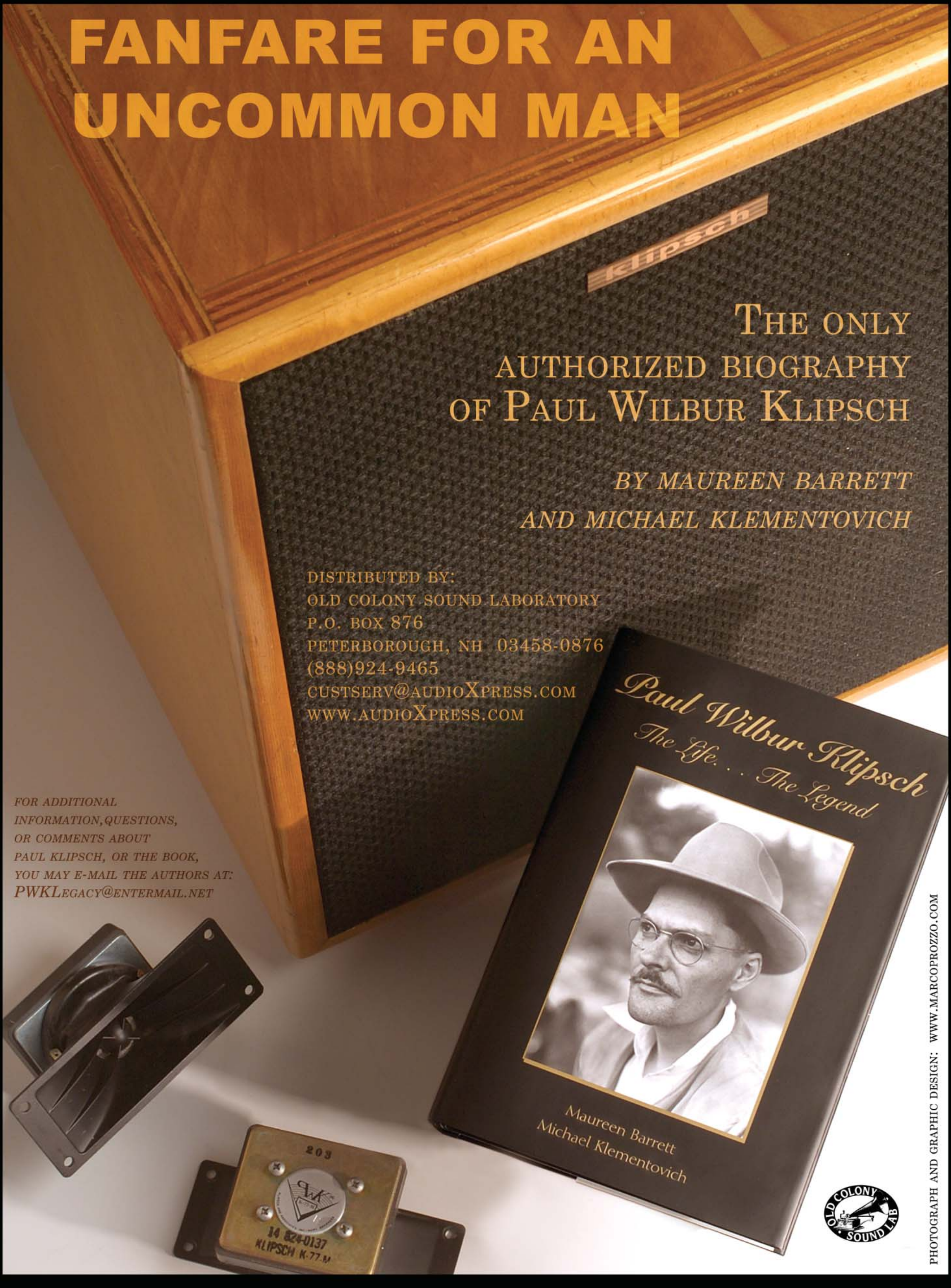

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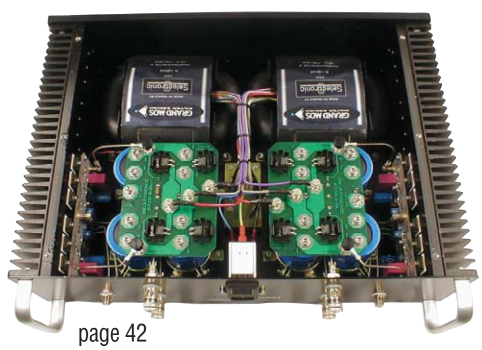

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12″ Dual Voice Coil Servo Subwoofer System

Now you can build your own subwoofer system, featuring a dual voice-coil driver design. **By Daniel L. Ferguson**

My article in the November '03 issue of *audioXpress* ("A Servo Dual Voice Coil Subwoofer," p. 18) presented research I had done on a servo subwoofer which used a dual voice coil driver with one coil driven and the other used as a velocity sensor. The article explored the relationships between velocity, acceleration, and sound pressure level, and showed that the sensor voice coil output voltage could be representative of cone velocity if the crosstalk from the driven voice coil could be eliminated. While I achieved some degree of success with the concept, I was not ready to go forward with a construction project.

The main drawback with the design at that point was distortion at higher volume levels during large transients such as kick drums. Since then, I have continued to experiment to eliminate that problem and have progressed enough to proceed with building the first working model. This article gives the results of that effort and the details to allow you to construct a similar unit.

In the original version, the feedback signal contained two components—the derivatives of the sensor voice coil voltage and driven voice coil current—which are the defining variables in the basic equation for the sensor voice coil acceleration transfer function. As it turns out, it is very difficult to match the phase relationships between these two to mimic the physical system. At this point, I must report that I have had only modest success after many hours of experimentation. I found that the system was very stable when either sensor voice coil voltage or driven voice coil current was used as feedback. When combined, system stability was significantly reduced.

PHOTO 1: Rear view of Dayton 12″ DVC driver.

Therefore, I have set aside the driven coil current component—at least for the present—and have developed a servo system that uses only the sensor

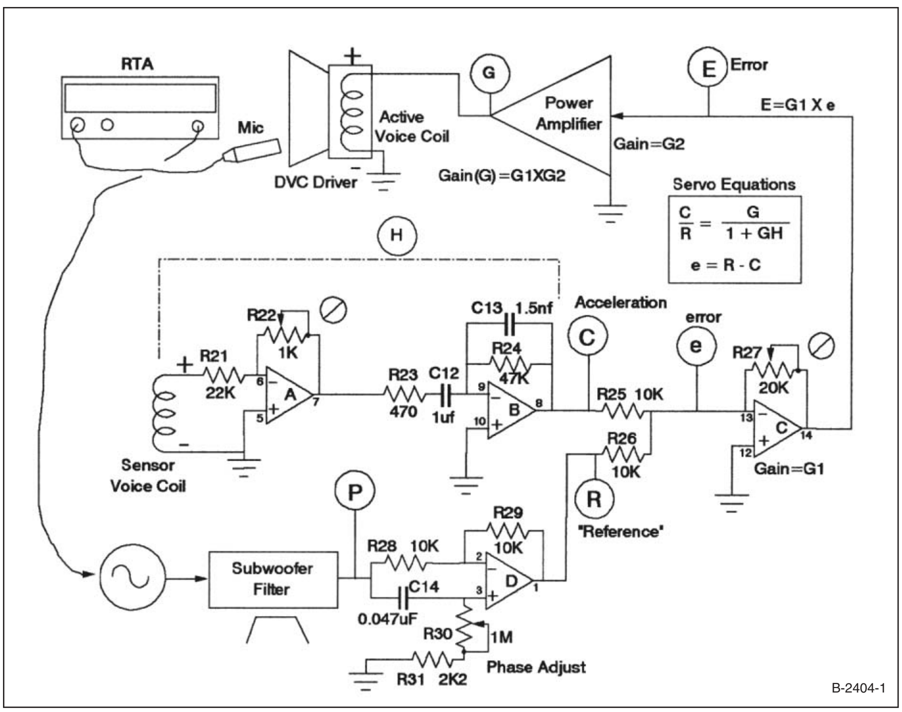

FIGURE 1: Servo circuit diagram.

ABOUT THE AUTHOR

Dan Ferguson has a BS and MS in mechanical engineering from Clemson University with a specialty in automatic controls. He is a long-time speaker builder and the author of three books on auto sound (the latest, *Car Stereo Speaker Projects Illustrated* is available from Old Colony Sound Lab) and several articles for *Speaker Builder/audioXpress*. He has been employed by Kimberly-Clark Corporation in various management positions for the past 22 years and is currently an operations consultant on a large MIS project. He and his wife have three grown children and five grandchildren, and reside in Appleton, Wis.

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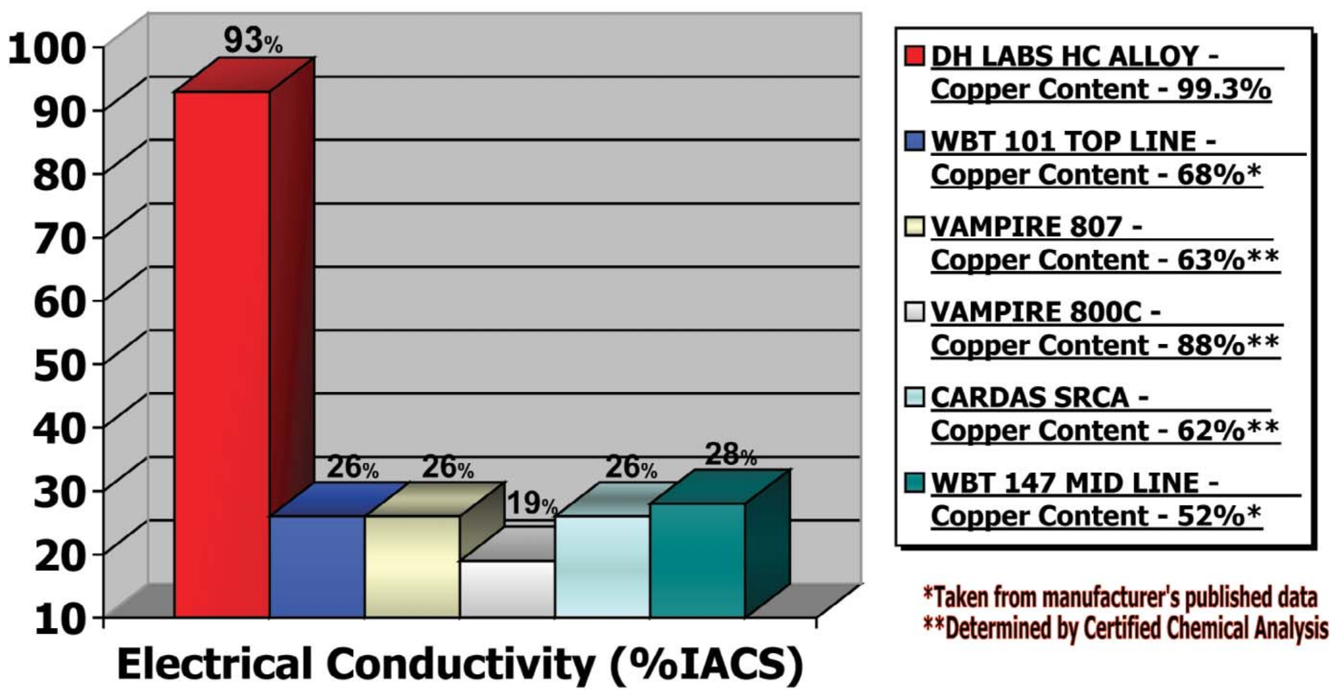

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voice coil voltage as feedback. While this is a theoretical compromise, it seems to work reasonably well. The reason for this is that the current-induced crosstalk from the driven voice coil appears to be minimal within the bottom decade. At very low frequencies, it is present, but only at low levels. While I intend to continue to experiment with the theoretical model, the more practical version presented in this article is still a significant improvement over open-loop subwoofers and appears to be very stable and reliable.

CONCEPT

Figure 1 shows the diagram of the simplified servo system. The sensor voice coil output voltage is scaled and inverted in op amp A and fed to an improved differentiator circuit driven by op amp B. The differentiator output at point C (the "controlled variable") is representative of the driver cone acceleration within the 20 to 120Hz or so bandwidth. Collectively, the system consisting of the sensor voice coil and the differentiator make up the "sensor," referred to in control terminology simply as "H."

Since an unused op amp section was available on the quad chip used, I added an optional phase adjustment circuit driven by op amp D. Depending on the position of R30, the subwoofer's phase can be offset from the input signal by approximately 0 to –180°. While this is somewhat of a "luxury" item, it has the potential for significantly improving system integration with the main speakers. The phase shifter output at point R represents the "reference" which the servo attempts to reproduce.

Op amp C has a variable gain from 0 to 2 depending on the position of R27 and is adjusted to allow setting the power amplifier volume control at a convenient level. The summing junction of this inverting op amp is the sum of the sensor output (which has been inverted by op amp A) and the input signal. The sum of these two is the system error correction signal. The job of the servo is to attempt to drive this to zero.

Overall system gain is the product of op amp C and the power amplifier gains and is referred to simply as "G." Later on, you'll see how these servo variables are applied in classical automatic control theory.

DRIVER SELECTION AND TESTING

The starting point was selecting the proper dual voice coil (DVC) driver. The ideal candidate should have low Qts, low free-air resonance frequency (Fs), low Vas, high excursion (Xmax), and

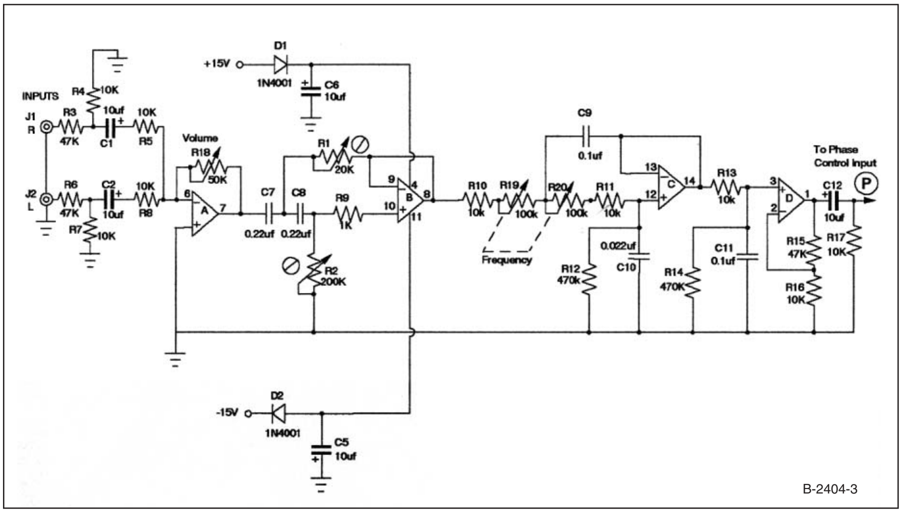

FIGURE 3: Servo subwoofer filter schematic.

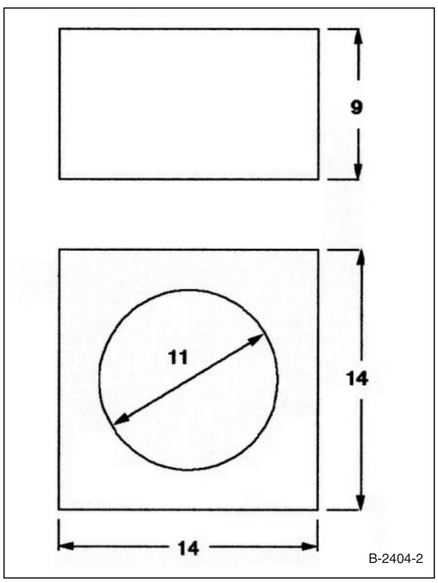

FIGURE 2: Subwoofer enclosure inside dimensions.

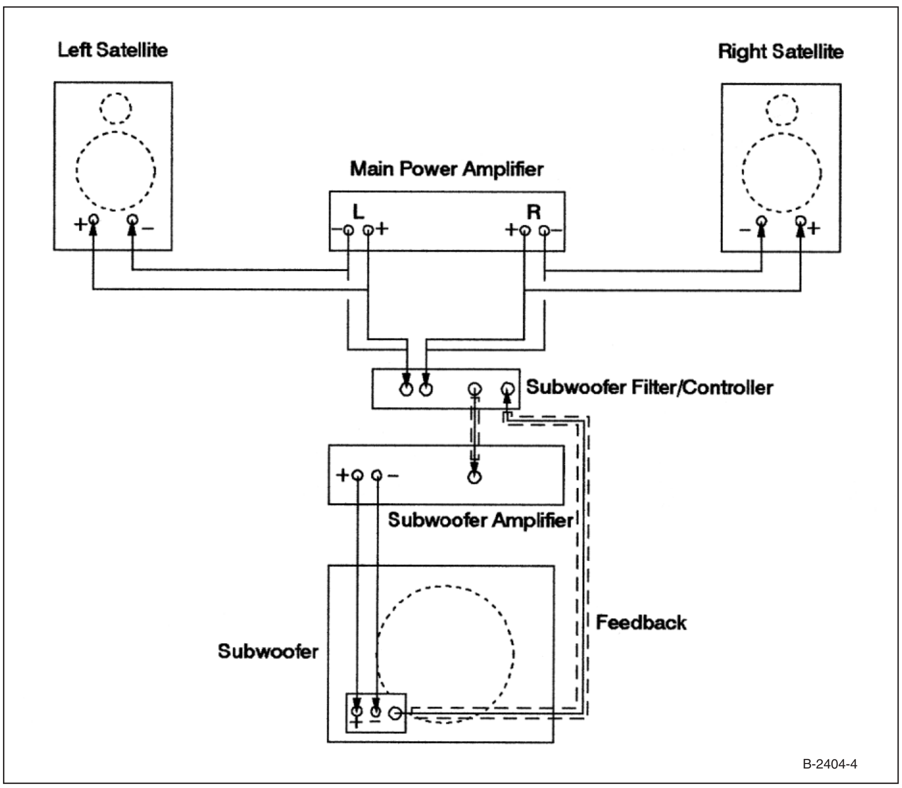

FIGURE 4: System wiring diagram.

good power handling. Last of all, it should be reasonably priced.

Taking all these factors into consideration, there are not many suitable drivers to choose from. The one I settled on is manufactured by Dayton and distributed as catalog number 295-185 by Parts Express, who graciously lent a sample unit for this article. The published specifications for this driver are as follows:

Power handling: 350W RMS/per coil, 600W total
Voice coil diameter: 2″
Voice coil inductance: 1.81mH
Nominal impedance: 8Ω per coil/4Ω total
DC resistance: 2.69Ω (both coils connected in parallel)
Magnet weight: 112 oz.
Fs: 21.7Hz
SPL: 87.4dB 1W/1m, 90.4dB @ 2.83V/1m (both coils connected in parallel)
Vas: 4.25ft^3
Qms: 12.53
Qes: 0.38
Qts: 0.37
Xmax: 15.1mm
Net weight: 18 lb
Catalog price: $119.80

Collectively, these appear to be an excellent set of specifications. The ±15.1mm (0.59″!) excursion rating held prospects for a linear magnetic system operation at high sound pressure levels, which is crucial for a servo mechanism. All the other parameters are similarly favorable.

Photo 1 shows the bare driver. It's a brute with a stiff, heavy, treated paper cone reinforced by an equally stiff 6½″ diameter domed "dust cover." The relatively thick foam surround is approximately an inch wide and appears consistent with the driver's high-excursion rating. Connection terminals are heavy-duty spring types able to accommodate large diameter wire and are plated to resist corrosion. The magnet assembly is covered with a rubber boot and is free of sharp edges. Altogether, the unit is impressive in appearance and heft.

After a couple of hours of break-in with a 3V 25Hz signal, I measured the Thiele/Small parameters and attained the following results:

Fs: 20.6Hz
Re: 2.6Ω
Qms: 9.96
Qes: 0.38
Qts (voice coils in parallel): 0.365
Vas (added mass method): 3.50ft^3
Vas (closed box method): 3.47ft^3

The critical parameter, Qts, was dead on. Having a somewhat lower Fs and Vas is usually an advantage. Lower Fs means lower cutoff frequency, while lower Vas translates to reduced box size for any given alignment. All parameter measurements were consistent and repeatable. Clearly, this was the most "linear" woofer I have experimented with to date.

With one voice coil driven, the measurements were essentially the same, except that Qts is now double what it was with both coils driven. And, of course, the DC resistance (Re) of a single voice coil is double that of two coils in parallel.

Fs: 20.7Hz
Re: 5.2Ω
Qms: 10.89
Qes: 0.82
Qts: 0.76
Vas (added mass method): 3.50ft^3
Vas (closed box method): 3.47ft^3

ENCLOSURE CONSTRUCTION

In order to minimize any nonlinearities the enclosure might introduce into the servo loop, I decided to make the structure more rigid by constructing it out of 1⅛″ thick counter top material. While this material is not readily available, I was able to find a local kitchen counter shop that not only could supply the material but was also willing to precision-cut the pieces to size for a nominal fee. As a result, I paid a total of $27 for the initial six pieces for the trial 1½ft^3 box, which is shown under construction in *Photos 2* and *3*.

The completed box is rigid, to say the least. Since the resistance of a material to flexure varies with the cube of the thickness, this box should be 3.4 times

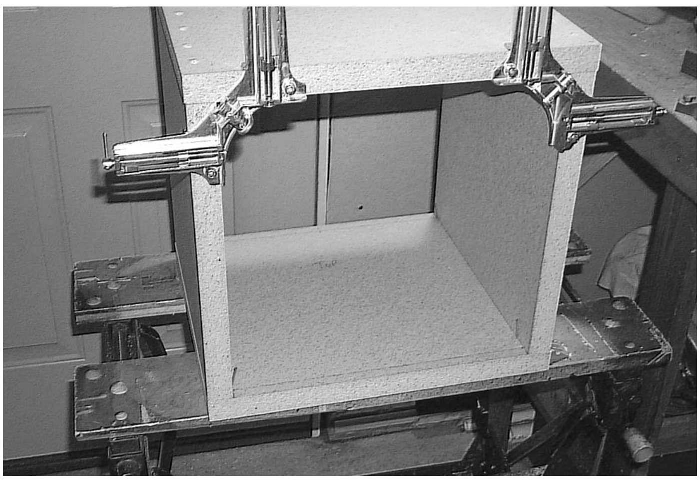
PHOTO 2: Subwoofer enclosure under construction.

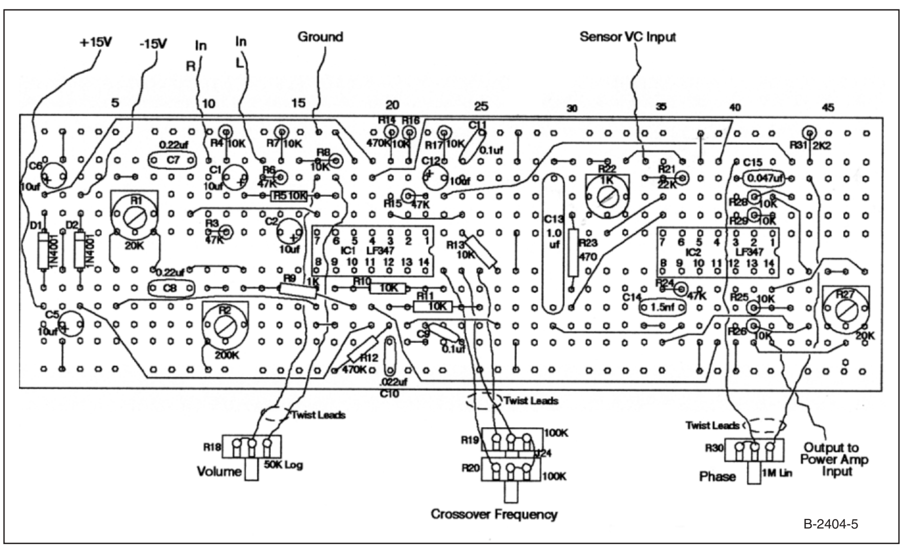

FIGURE 5: Subwoofer controller PC board layout.

stiffer than an equivalent box constructed from ¾″ thick material.

As has become my standard practice, I sized the pieces for a 1⁄16″ overlap at each joint and glued and screwed the pieces together. I needed to buy a new flush cutting router bit to handle the thicker material. The one I chose is made by CMT (#806.630.11), which is able to trim off stock up to 2″ thick like a lightsaber. I also used my router with a straight plunging bit to accurately cut the driver mounting hole using a simple jig made from a piece of scrap bolted to the router's bottom plate and a nail for a center pivot. The accuracy of the hole diameter is dependent only on how accurately the center pivot nail is placed.

After testing and experimentation by adding blocks to reduce volume, I determined that one cubic foot would be adequate to achieve the target closed box Q of 1.5. I cut the box down and installed a new back. The result is shown in *Photo 4* with dimensions given in *Fig. 2*. As long as box volume is maintained,

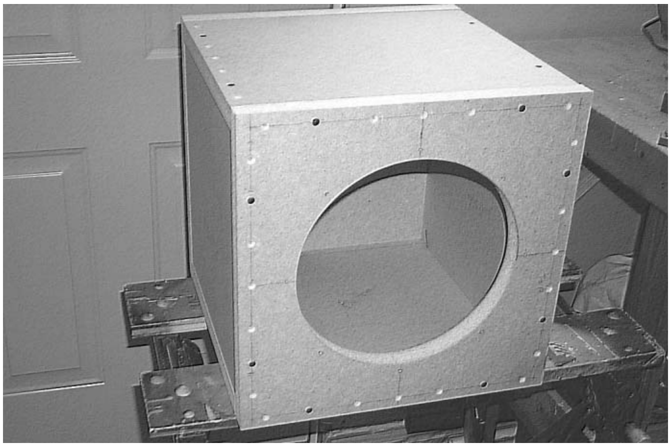

PHOTO 3: Subwoofer enclosure with front and rear panels.

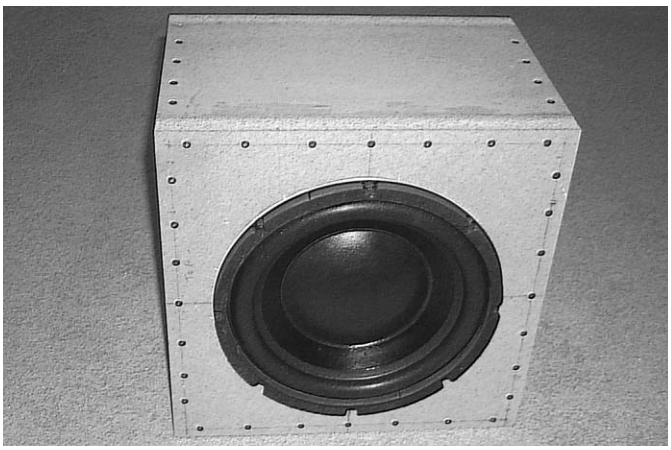

PHOTO 4: Assembled subwoofer front view.

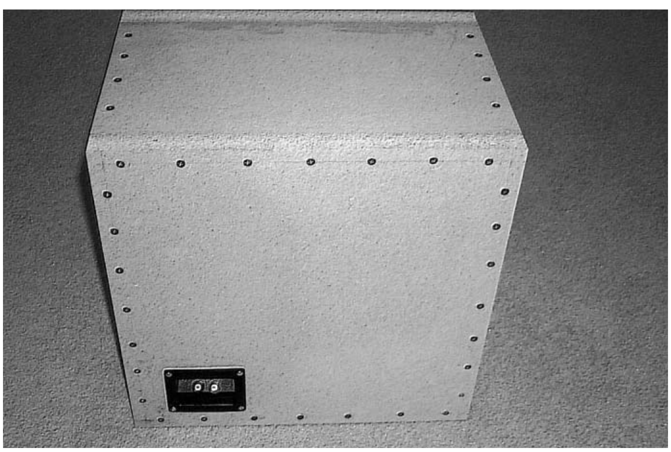

PHOTO 5: Assembled subwoofer rear view.

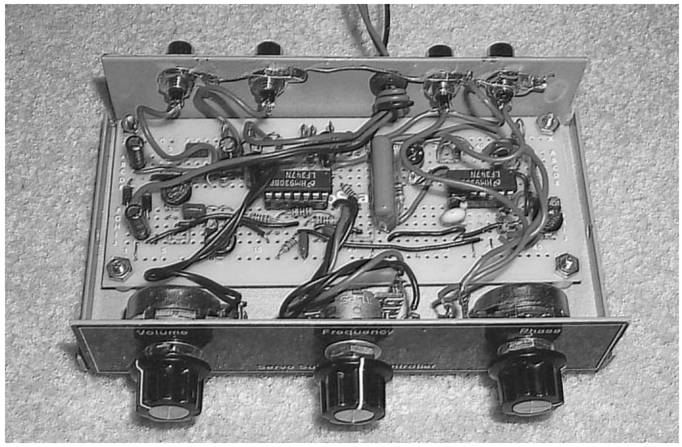

PHOTO 6: Assembled subwoofer filter/controller.

PHOTO 7: Subwoofer filter/controller front view.

PHOTO 8: Subwoofer filter/controller rear view.

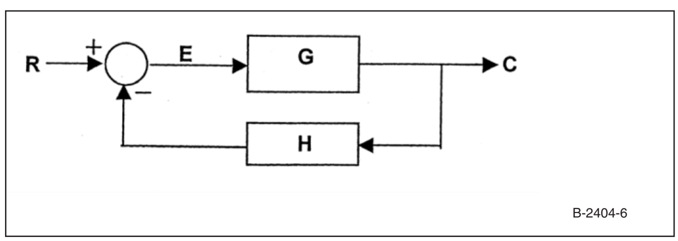

FIGURE 6: Basic block diagram for a closed-loop system.

none of the dimensions are critical except for the driver mounting hole diameter. With the dimensions shown, there is plenty of room to install a grille frame if you desire one.

Photo 5 shows the terminal cup installed in the rear of the enclosure near a corner, which is the strongest point and shouldn't weaken the back as much as it would near the center. I also installed an RCA jack in the terminal cup to provide a convenient sensor voice coil connection.

CONTROLLER

To ensure repeatability of results, I decided to include all of the subwoofer control functions on one circuit board. All that's required for a complete servo is a regulated ±15V power supply and an external amplifier. The subwoofer filter section schematic is shown in *Fig. 3*. You can possibly omit this portion of the circuitry if you have a suitable filter available with similar response characteristics.

For maximum flexibility, I designed this version of the filter with high-level inputs, which are connected to the right and left satellite speaker terminals. *Figure 4* shows the complete external wiring diagram.

Light-gauge speaker wire with an RCA plug on one end is the preferred interconnect for the controller inputs. Standard, shielded audio patch cords connect the sensor voice coil to the feedback input terminals and send the controller's output to the power amplifier (Note: It may not even be necessary to use shielded cable for the feedback connection). Nothing fancy is required here. The tiny operating bandwidth and

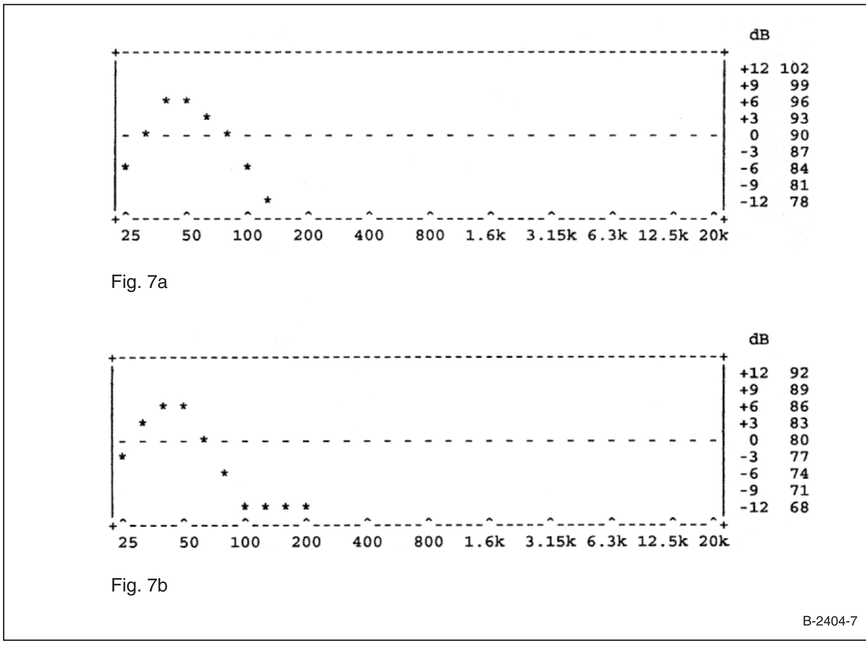

FIGURE 7: Subwoofer loop response. 7a: close-miked response. 7b: accelerometer response.

ultra-low frequency range just doesn't justify it.

The controller circuit board is built on a standard prototyping board available from Radio Shack or a number of other suppliers. The layout is shown in *Fig. 5*. If you use the same cabinet I did, you will need to trim off approximately ¼″ from each end of the board and drill new mounting holes before getting started. The completed control unit is shown in *Photos 6*, *7*, and *8*. I made front and rear "faceplates" using Lotus Freelance and printing them on full-page label stock. After that, I covered them with transparent shipping tape before cutting them out and placing them on the cabinet.

SOME BASIC SERVO THEORY

The basic block diagram for a closed-loop system is shown in *Fig. 6*.

The classic equation (expressed in Laplace transforms), which describes the closed-loop transfer function of a simple servo, is

$$\frac{C}{R} = \frac{G}{1+GH}$$

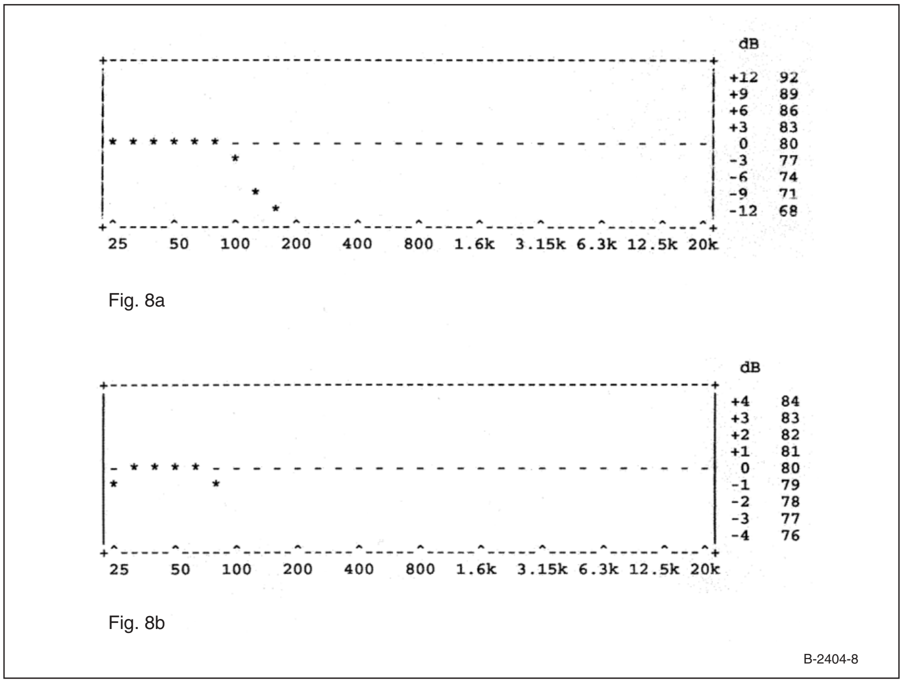

FIGURE 8: Subwoofer filter response. 8a: 3dB per division scale. 8b: 1dB per division scale.

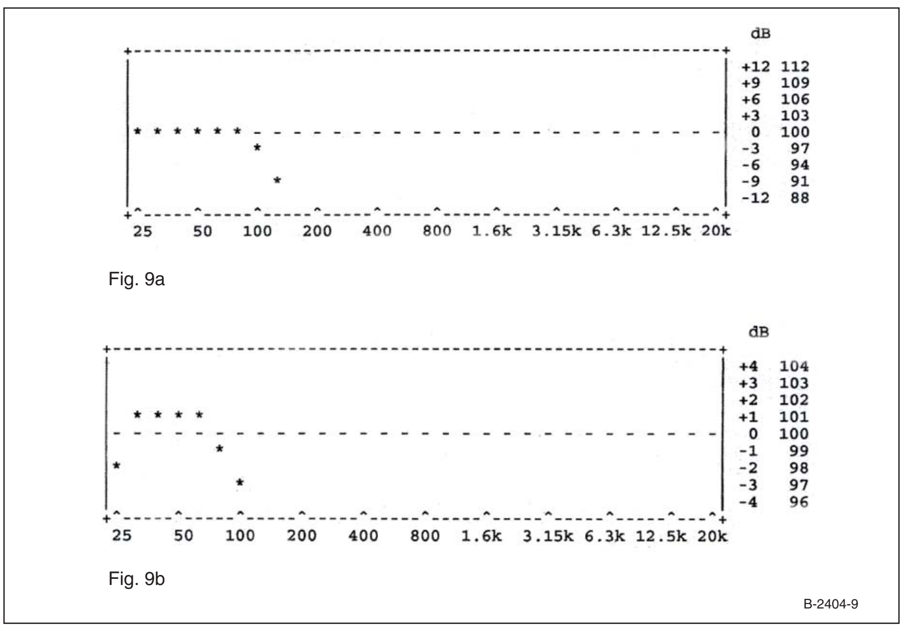

FIGURE 9: Subwoofer closed loop response. 9a: 3dB per division scale. 9b: 1dB per division scale.

This states that the controlled variable, C, varies with the reference variable, R, according to the ratio of the transfer function of the controller/amplifier, G, divided by 1 + the product of G and the transfer function of the feedback sensor, H.

Ideally, C/R would always equal 1 so the controlled variable would equal the reference. Since it's an imperfect world, the challenge for the servo designer is to come as close as possible to this ideal.

It can also be shown that the error-correction signal transfer function, E, can be described by

$$\frac{E}{R} = \frac{1}{1+GH}$$

where $E = R - HC$

If you want the error to be small, GH must be much greater than zero. However, as you increase GH, you will eventually reach a point where the system becomes unstable.

Therefore, if H is too high, you will need to reduce the amplifier gain to maintain stability. This could reduce the forward gain so much that the speaker output is insufficient. Clearly, a balance must be reached so that the power amplifier gain setting is near midpoint for good signal-to-noise and adequate sensitivity.

After much experimentation, I settled on the following servo settings for the best combination of accuracy and stability. With the system operating in steady-state at 40Hz, I took the following measurements:

R27 set at its maximum of 17K
G1 = R27/R25 = 17K/10K = 1.7
G2 = Vout/Vin = 0.97V/0.2V = 4.85
Total G =(G1 × G2) = (1.7 × 4.85) = 8.25
With R22 set at = 333Ω
H = 0.142V = 0.71R
Therefore GH = (8.25 × 0.71) = 5.85
Since GH is much greater than 1, the servo should have a positive effect on error reduction.

OPEN-LOOP SYSTEM RESPONSE

Prior to final testing, I lightly stuffed the enclosure with 14 oz of poly fiber fill, which reduced the closed box Q from 1.5 to 1.4 and closed box resonant frequency from 40.2Hz to 37.3Hz. Directionally, this should improve system performance since another basic servo design principle is to develop the best open-loop response possible before closing the loop.

Using the hookup shown in *Fig. 1*, I applied pink noise to the system and measured the response with my Audio Control SA3055 real-time analyzer. *Figure 7a* is the open-loop response measured with the RTA microphone. This is clearly the response of a grossly undersized box.

Figure 7b is output of the sensor voice coil after taking its derivative (point C in *Fig. 1*) under the same system conditions as previously stated. There are some differences between *Figs. 7a* and *7b* that are cause for concern. While the peak is at the right frequency, the slopes on either side are not the same as those measured with the mike. Although these differences appear to be genuine, the servo still performs well, as the following data shows.

CLOSED-LOOP RESPONSE

If the servo is working, the subwoofer will reproduce the response curve of the subwoofer filter.

With the subwoofer filter trimpot settings at R1 = 17.2K (the maximum setting for the particular trimpot I used) and R2 = 70K, I applied pink noise to the filter inputs and obtained the response shown in *Fig. 8* from the filter output (point P in *Fig. 1*). *Figure 8a* is with the resolution set to 3dB per divi-

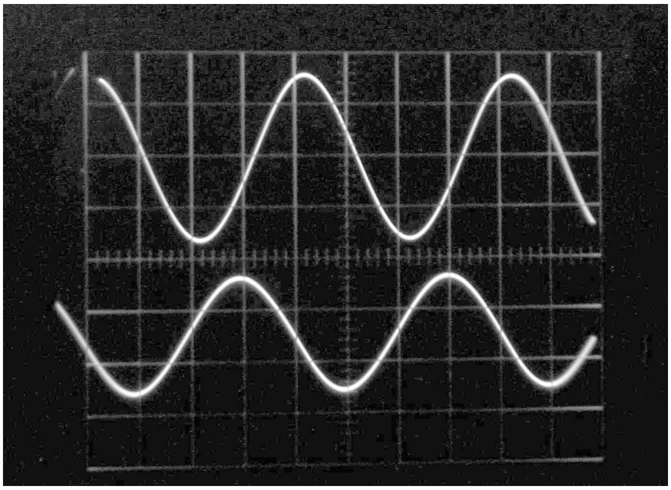

PHOTO 9: Subwoofer closed-loop waveform at 25Hz.

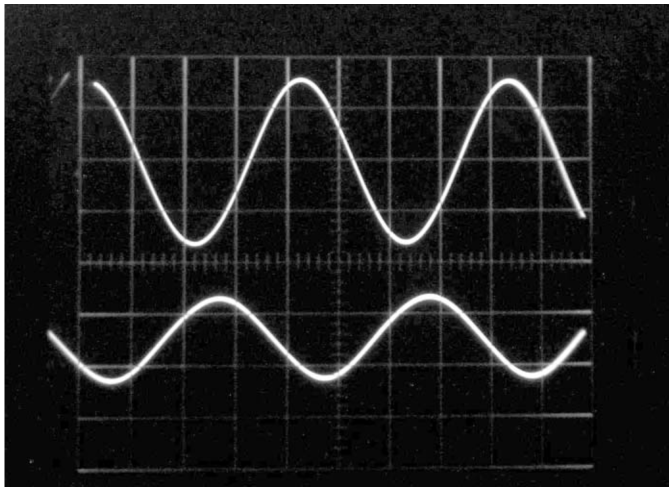

PHOTO 10: Subwoofer open-loop waveform at 25Hz.

sion. *Figure 8b* is the same response curve with the resolution set to 1dB per division. Clearly, the curve is ruler flat within the passband and is a proper reference for the servo.

Next, I connected the complete servo without changing any settings and achieved the acoustic, close-miked response curve for the subwoofer shown in *Fig. 9*. *Figure 9a* is with the resolution set to 3dB per division. It is identical to the reference curve.

Figure 9b is the same curve with the resolution set to 1dB per division. It is very similar to the reference curve and is identical within the passband. The slopes on either end of the passband are, in fact, steeper than the filter and are consistent with the sensor coil's response noted previously. However, the difference is on the order of only 1dB, which should make it inaudible.

For this portion of the test, I judge the servo to be a success.

CLOSED-LOOP WAVEFORMS

Photo 9 shows the subwoofer sine waveform at 25Hz with the amplifier output set at about 2V AC. The upper trace is the input reference signal taken at Point R in *Fig. 1*, while the lower trace is the close-miked subwoofer output. The waveform appears to be sinusoidal and distortion-free.

For comparison, *Photo 10* is the same operating point with the feedback signal unplugged. While the amplitude is significantly lower, the waveform still appears to be sinusoidal. This is clearly credited to the quality of this driver. Higher frequencies within the subwoofer filter passband are considerably easier to reproduce.

Photo 11 is a sample at 30Hz. At this point, the amplitude is the same as the upper trace reference. *Photo 12* is the same operating point with the feedback unplugged. The amplitude has clearly fallen off.

Another test of the servo is its ability to track a complex waveform. *Photo 13* shows the closed-loop subwoofer response to a 25Hz square wave. The upper track is the reference signal from the subwoofer filter; the lower track is the close-miked subwoofer response. After a square wave passes through the subwoofer filter, it's not too square any more.

Now, it's important to understand that this is *not* distortion. It is simply what happens when the higher-frequency components are filtered out of the square wave.

In any case, the irregular shape of this curve makes for a good gymnastic workout for the servo to try to track, and it does a fairly creditable job of doing so. *Photo 14* shows the identical condition with the feedback unplugged. There is some deterioration in both shape and amplitude.

HOW LOUD WILL IT PLAY?

I placed my RTA mike at a distance of 1m from the subwoofer at a height of 1m off the floor. At a power amplifier voltage

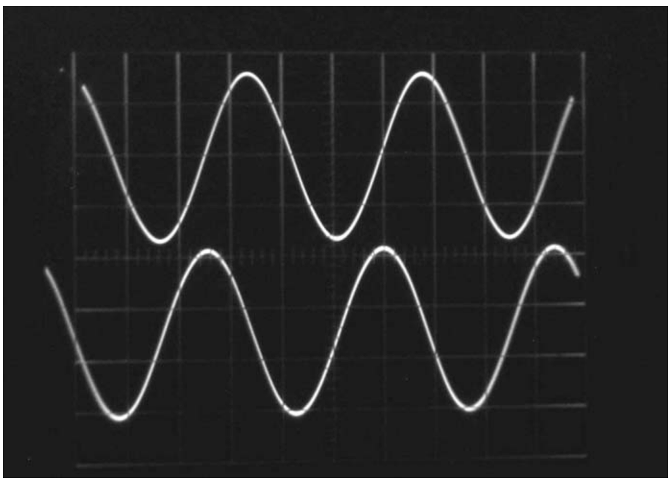

PHOTO 11: Subwoofer closed-loop waveform at 30Hz.

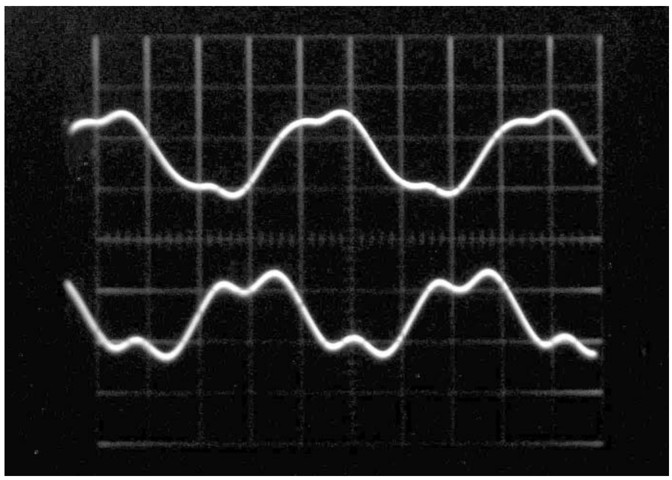

PHOTO 13: Subwoofer closed-loop waveform—25Hz square wave.

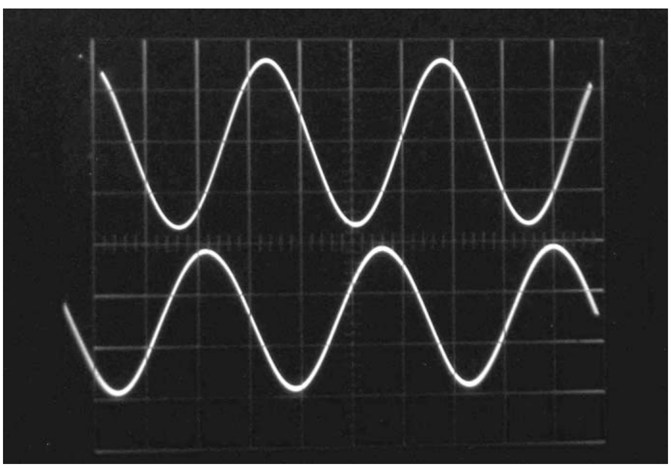

PHOTO 12: Subwoofer open-loop waveform at 30Hz.

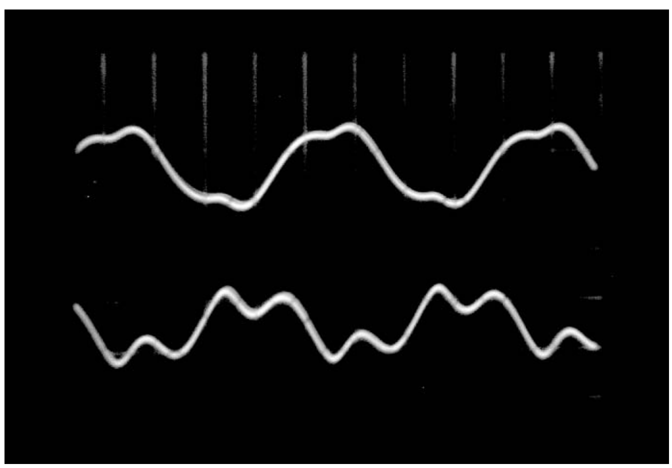

PHOTO 14: Subwoofer open-loop waveform—25Hz square wave.

of 22V RMS, the sub produced an output of 90dB at 30Hz before the waveform began to show signs of becoming triangular. At 35Hz, the sub would handle 25V RMS before visible distortion and the output reached 100dB.

HOW DOES IT SOUND?

There is only one word necessary to describe the sound of this sub—smooth. When low frequencies are reproduced without distortion, they become almost subliminal. In addition, the servo forces the sub to react faster to transients than an open-loop sub, so kick-drum impacts are immediate. It just might be a good match for a pair of electrostats or magnaplanars. Since the cabinet is so compact, you could build one for each channel.

Before closing, a word about setting up the phase control for a seamless blend with the main speakers is in order. The easiest way I have found to optimize the phase control setting is to run a single tone through your entire system at the crossover frequency you are aiming for—preferably somewhere around 80Hz. Position yourself between the sub and the main speakers and adjust the sub volume until it appears to be equal to the main speakers. Then sweep the phase control up and down until the tone is the loudest.

At this point, the sub is in phase with the mains. You can make this test more definitive with a sound-level meter.

CONCLUSION

My design goals for this project were to apply what I learned from the original experimentation and build an accurate, compact subwoofer with adequate output within the passband of 20 to 100Hz. In my opinion, it meets these goals. More importantly, this subwoofer makes beautiful music. So, while it's a little bit of a technical compromise, it's no slouch. I recommend it for a fun project. ❖

PARTS:

RESISTORS	VALUE	QTY	PART NO.	SUPPLIER
R4, R5, R7, R8, R10, R11, R13, R16, R17, R25, R26, R28, R29	10k	13	10.0KXBK-ND	Digi-Key
R9	1k	1	1.00KXBK-ND	Digi-Key
R12, R14	475k	2	475KXBK-ND	Digi-Key
R3, R6, R15	47.5k	3	47.5KXBK-ND	Digi-Key
R21	22.1k	1	22.1KXBK-ND	Digi-Key
R23	475	1	475KXBK-ND	Digi-Key
R31	2.21k	1	2.21KXBK-ND	Digi-Key
R1	20k Trimpot	2	3306P-1-203	Digi-Key
R2	200k Trimpot	1	3306P-1-204	Digi-Key
R27	1k Trimpot	1	3306P-1-102	Digi-Key
R18	10k Pot (Log)	1	271-1721	Radio Shack
R19, R20	100k Dual Pot	1	271-1732	Radio Shack
R29	1M Pot (Linear)	1	271-211	Radio Shack
DIODES				
D1, D2	1N4001	1	1N4001GI	Digi-Key
CAPACITORS				
C1, C2, C5, C6, C12	10µF/35V Electrolytic	5	493-1077-ND	Digi-Key
C7,C8	0.22µF/50V Metal Film	2	P4667	Digi-Key
C9, C11	0.1µF/50V Metal Film	2	P4525	Digi-Key
C10	0.022µF/50V Metal Film	1	P4517	Digi-Key
C12	1.0µF/50V Metal Film	1	P4675	Digi-Key
C13	1.5nF/100V Metal Film	1	495-1092-ND	Digi-Key
C14	0.047µF/100V Metal Film	1	495-1101-ND	Digi-Key
INTEGRATED CIRCUITS				
IC-1, 2	LF-347	2	LF347N	Digi-Key
HARDWARE				
Circuit Board		1	276-170	Radio Shack
Knobs	¾″ Dia.	2	274-415	Radio Shack
Cabinet	5½ × 3 × 1¼	1	537-139-P	Mouser
RCA Jacks		4	161-1052	Mouser
Grommet	5/16″ Dia.	1	534-731	Mouser
Nylon Spacers	¼″ long	4	561-K4.25	Mouser
Screws	4-40 × ½″	4	H146	Digi-Key
Nuts	4-40	4	H216	Digi-Key
IC Socket	14 Pin	2	A24808-ND	Digi-Key
WIRE				
Red, Green, Black	22 gauge Stranded	1 pkg.	278-1224	Radio Shack
Bus Wire	24 gauge	1 spool	278-1341	Radio Shack

SUPPLIERS

Parts:

Parts Express
725 Pleasant Valley Dr.
Springboro, OH 45066-1158
1-800-338-0531
www.partsexpress.com

Mouser Electronics
958 N. Main
Mansfield, TX 76063-4827
1-800-346-6873
www.mouser.com

Digi-Key
701 Brooks Ave. South
Thief River Falls, MN 56701-0677
1-800-344-4539
www.digikey.com

Radio Shack
Local Store
1-800-THE-SHACK
www.radioshack.com

Lumber and building supplies:
Local countertop contractor
Local home improvement stores

PARTS LIST MATERIALS:

1 sheet countertop material (30″ × 8′ × 1¼″)
1 box 2″ long course thread sheet rock screws
Rope caulk
3′ twin lead speaker wire
Carpenter's glue

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Saga of a Tube OTL Amp

This author presents an update of a classic output transformerless (OTL) tube amplifier of the 1950s. **By Glen Orr**

PHOTO 1: Completed amplifier, strictly utility.

I am not an electronics professional. I am a hobbyist. Although I have designed and built amplifiers and other electronic gear, most have been solid-state, and all have been for my own pleasure only.

I have been profoundly interested in electronics, especially audio electronics, since I was a freshman in college, when one day in the basement of my dorm, among the ping-pong and pool tables, an upperclassman set up his home-built high-fidelity system. It was in the days before the advent of stereo, and was composed of one single ported, unfinished plywood baffle, standing about 6′ high, housing what I think was an Altec Lansing 15″ woofer with a midrange and a tweeter. He had a tube amplifier built on a raw aluminum chassis driving the speaker with pretty much all it could take. On the turntable was Duke Ellington's "Skin Deep." I was hooked.

A little more than four years later, during Christmas vacation in my first year as a dental student at Northwestern University, I built a tube output transformerless (OTL) amplifier from an article and circuit diagram (*Fig. 1*) published in *Audio Engineering* in June 1954. I copied *Fig. 1* from the original article. The amplifier was designed to drive a 16Ω loudspeaker directly without the use of an output transformer, or, for that matter, without a power transformer. Later the same circuit was published in the "Classic Circuitry" column of *Audio Amateur* and then still later in Volume Three of *Audio Anthology*. The latter also includes the original article.

THE ORIGINAL CIRCUIT

The amplifier can be divided into two parts. The first is the voltage amplifier, which is comprised of two sections of the first 12AT7, and provides all the voltage gain for the amplifier. The second part is the power amplifier or current amplifier. It has no voltage gain and makes up the rest of the amplifier, not counting the power supply. The output tubes are in a single-ended push-pull arrangement and operate close to class B with 100% global negative feedback from the output to the cathode of the first section of the second 12AT7. The phase inverter is the second section of the same 12AT7, and the 6SN7 provides the drive for the output tubes.

The power supply is a line-operated ±140V, with voltage doublers for the negative output tube bias and +250V supply. One leg of the bias is adjustable to balance the direct-coupled output to 0V. The original circuit used a

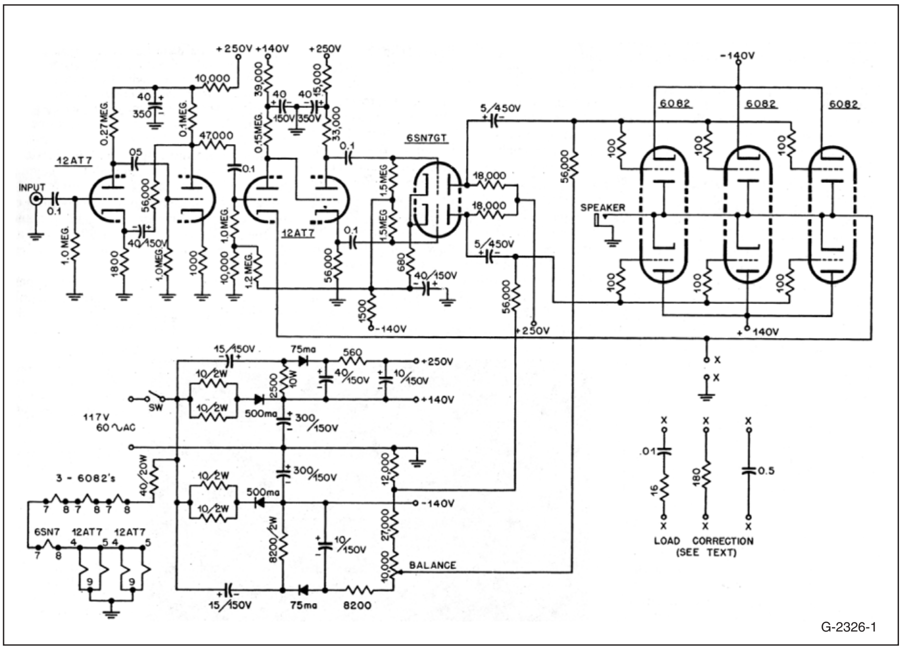

FIGURE 1: The original amplifier (published in 1954).

ABOUT THE AUTHOR

Glen Orr has a BA from Luther College, Decorah, Iowa, in chemistry and physics, a DDS from Northwestern University in Chicago (Evanston), Illinois, and is a practicing dentist. His hobbies, besides audio electronics, include aviation and duplicate bridge. He is a commercial pilot and flight instructor. He built a PJ-260 open cockpit biplane in his garage from a set of plans (as opposed to a kit), and flew it in 1996 after 17 years of construction. He is an avid Experimental Aircraft Association member and became a Life Master in the American Contract Bridge League in 1977.

series/parallel arrangement for the filaments from the AC line. The amplifier is RC-coupled, except between the first section of the power amplifier and the phase inverter, which is direct-coupled.

FIRST MODIFICATIONS

When I built the amplifier in the early '60s, I had little equipment and virtually no experience. Earlier I had built a Knight Kit audio amplifier, and the year before, in undergraduate school, I took one four-hour course in electronics. I at least had the foresight to order a Heathkit VTVM kit, which I built first. The construction site was a large desk in my bedroom. Armed only with this background and hardware (and youth), I still had the audacity to modify the circuit (*Figs. 2* and *3*):

1. I changed the output tubes from 6082s to 6AS7Gs. The 6AS7G has a filament voltage of 6.3V as opposed to the 26.5V of the 6082—otherwise they are electrically identical. I also paralleled the original three output tubes with three more, hoping to increase the output power.
2. I modified the power supply (*Fig. 3*). The original amplifier was line operated, which I didn't like. I had an old console radio that I had taken apart (now I wish I still had it intact) that had the parts for a +450V power supply using an old type 80 tube. I used the power transformer from the unit and replaced the type 80 with an electrically identical 5Y3. This changed the plate supply to both sections of the 6SN7 to 450V.
3. I changed the bias arrangement around both sections of the 6SN7 in the power amplifier, but left the voltage divider for the fixed bias on the first section of the second 12AT7. I removed the negative supply and ran the cathodes of the 6SN7 to ground through the existing 680Ω resistor.
4. I used a transformer and full-wave rectifier for the +155V and –155V supplies and silicon diodes in place of the selenium rectifiers. This is depicted on the schematic as +140 and –140V, but I always had 10 to 15V more than that. I attribute this disparity to a voltage drop across the selenium rectifiers and the 5Ω protective resistors, which I omitted. I may have also picked up a little extra voltage from the power transformer I used.
5. I kept the associated voltage doubler circuit for the bias supply, but again substituted a silicon diode for the selenium rectifier. I added a second potentiometer to make both legs of the bias totally adjustable.
6. I used a separate filament transformer for all the tubes in the amplifier, except, of course, for the 5Y3. The 6AS7G requires 2.5A per tube, so I added a 20A 6.3V transformer.

Well, so much for the original "transformerless" design!

At the time I was happy—the amp sounded great and had plenty of volume with the efficient 16Ω loudspeakers, which were popular in those days. In any event, a few years later, after I built an Eico 460 oscilloscope, I was somewhat disappointed when my amplifier—even with the three extra output tubes—clipped at slightly below 19W RMS. The amplifier had been described as a "25W amplifier" with only three output tubes. On further analysis, perhaps the authors of the original article exaggerated the performance of the amplifier.

The presence of the original half-wave 500mA selenium rectifier circuit was marginal at best. After all, in order to put 25W RMS into 16Ω, you need 1.25A. However, there were no such limitations with the upgraded power supply I had built.

The amplifier withstood very hard use for a few years, until I graduated from dental school and later enlisted in the U.S. Army. When I was discharged in late 1969, transistors were becoming the rage for high-end audio equipment, so as I resumed my electronics hobby, I stored the OTL amplifier in a dusty corner of my basement.

NEW MODIFICATIONS

I have been a subscriber of *Audio Amateur* and its descendants since the early 1970s. Unfortunately, I never subscribed to *Glass Audio*. However, when *Audio Electronics* and then *audioXpress* were born, I started reading about tubes again, rekindling my interest in my OTL amplifier. So one day I dusted off the old amp and started experimenting with it. The following is what I came up with (*Figs. 4* and *5*):

1. I made my first modification because the power transformer for the ±155V supply gave up the ghost (I caused an inadvertent short that may have had something to do with it). When I replaced the power transformer, I decided to make the bias supply for the output tubes independent of the ±155V supply, so I dispensed with the voltage doubler, and added yet another transformer, another rectifier bridge, and filter capacitor.

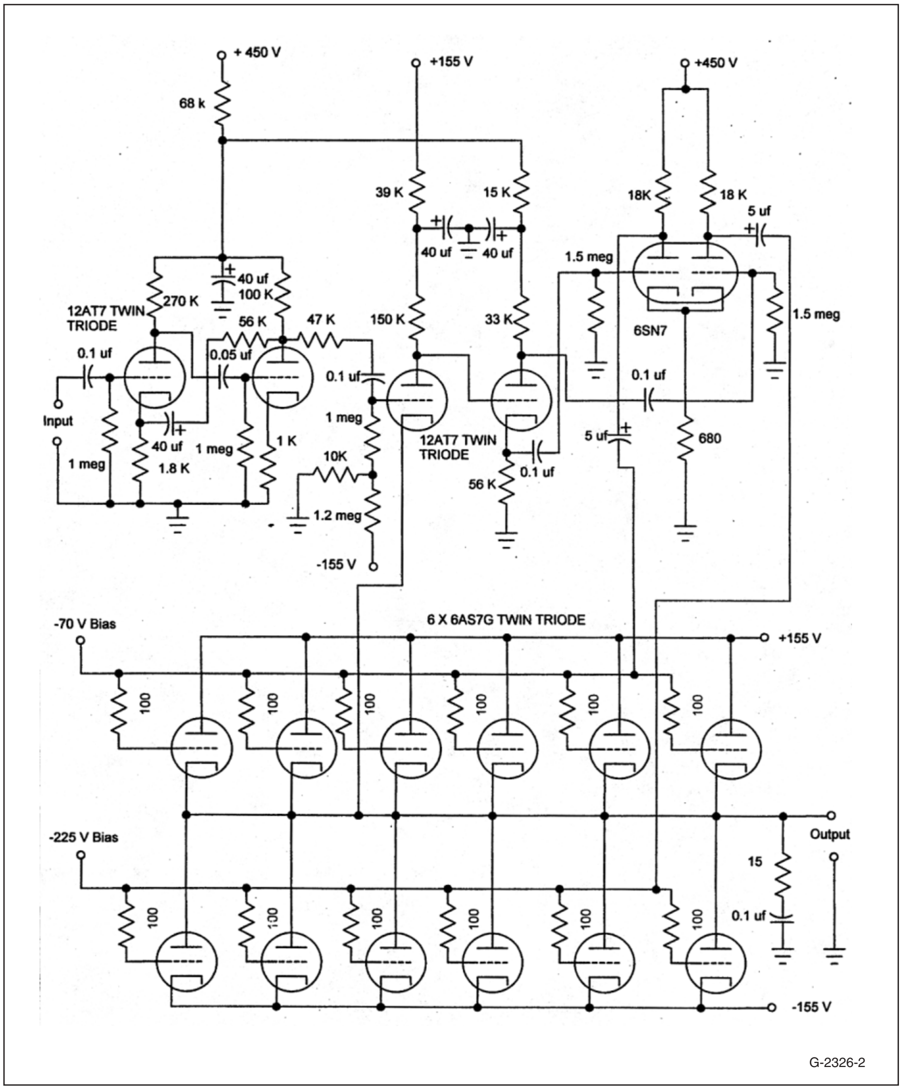

FIGURE 2: First modification circa 1961.

2. I changed both 12AT7s to 12AX7s, now V1 and V2 in *Fig 4*. The 12AX7 has an amplification factor of 100, as opposed to 60 for the 12AT7, which translates into more potential voltage gain.

TABLE 1
AMPLIFIER PARTS LIST

RESISTORS (¼W UNLESS NOTED OTHERWISE)	
R1	4700Ω
R2	470k, ½W
R3, R8, R12	1 meg
R4	12k
R5	91k, 1W
R6	100k, ½W
R7	220k
R9	1500Ω
R10, R11	470k
R13	2200Ω
R14	10k
R15, R20	330k, 1W
R16, R19	1.5 meg
R17, R18	18k, 3W
R21	39Ω
R22-R33	100Ω
R34	15Ω, 1W
CAPACITORS	
C1, C5, C7, C8, C9, C12	0.1μF, 400V
C2	2.2μF, 400V polyester
C3	0.05μF, 400V
C4	40μF, 450V electrolytic
C6	22pF, 400V
C10	5μF, 700V electrolytic
C11	5μF, 700V electrolytic
TUBES	
V1, V2	12AX7
V3	6SN7
V4-V9	6AS7
SEMICONDUCTORS	
Q1	2N3906
Q2	MPSA92
Q3	2N5551
Q4, Q5, Q6	2N3904
ZD1	62V, ½W zener diode
POWER SUPPLY PARTS LIST	
RESISTORS	
R35, R36	390Ω, 1W
R37, R38	150k, 3W
R39	24k, 1W
R40, R42	15k pot, 1W
R41	27k, 1W
R43	15k, 1W
R44, R45	56k, ¼W
CAPACITORS	
C13, C14	80μF, 500V electrolytic
C15, C16	400μF, 200V electrolytic
C17	16μF, 400V electrolytic
SEMICONDUCTORS	
D1-D4	3A rectifier diodes, 600 PIV
D5-D8	1A rectifier diodes, 600 PIV
TUBES	
V10	5Y3
TRANSFORMERS	
T1	6.3V, 20A
T2	750 CT 50mA, 5V AC at 2A
T3	240V ct 3A
T4	220V, 50mA
Misc. fuses and fuse holders, on-off switch, line cord	

3. I believed that the old power amplifier circuit did not provide adequate drive to the output tubes and that was the main reason the power output was less than I expected. So I changed the method of phase inversion using a cathode-coupled amplifier instead of the original configuration in the first stages of the power amplifier section. This also provided a convenient way to apply feedback from the global negative feedback loop.

Instead of going from the output to the cathode of the first tube in the loop as in the original amplifier, the feedback now goes from the output to the grid of V2b, analogous to modern transistor power amplifier or op-amp topology. I used a constant-current source in place of a cathode resistor in V2 and replaced the 680W cathode resistor with a constant-current source in the V3. The current sources act like a very large resistor, which maximizes the output and increases the linearity of the cathode-coupled amplifiers.

I selected current sources of about 0.3mA for V2 (0.15mA for each section) and rather large plate resistors at 470k. At first I used smaller resistors in the plate circuit, but I developed problems with grid current in the high mu triodes. I used 16mA for V3 (8mA for each section), with the same 18k plate resistors used in the original amplifier.

4. Perhaps getting carried away with current sources, I also replaced the load resistor of V1a with a current source. The current source, again, acts like a very large plate resistor and maximizes the gain of the circuit, which in turn allows an increase in the feedback and thus reduces distortion. I chose a plate current of about 0.15mA for the plate current of V1a. I left V1b in a conventional configuration.

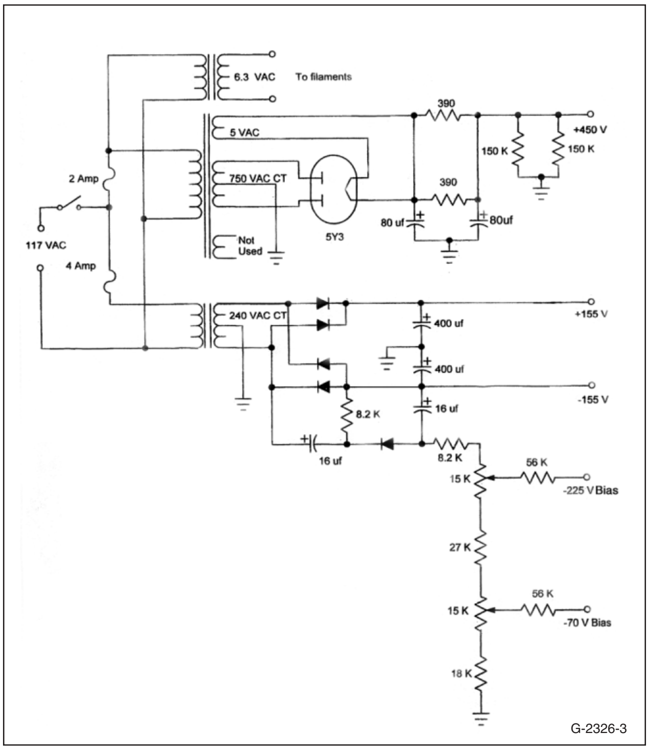

FIGURE 3: Power-supply modification circa 1961.

Some purists may say that this is no longer a tube amplifier. I suppose you could argue that. However, all the gain of the amplifier (voltage and current) is derived from tubes. The transistors that are used for the current sources are actually functioning as resistors. Anyway, after all is said and done, most of the current in the amplifier must run through the PN junctions of the rectifier diodes, and if you replace the 5Y3 with a pair of diodes (which I will probably do if the 5Y3 in the amplifier dies), all of the current will need to pass through silicon diodes.

Surprisingly, I have had very few problems with high-frequency oscillation and no problem with motorboating. The only issue I had with high-frequency stabilization was solved by the placement of a 22pF capacitor, C6, in parallel with the grid resistor on V2a.

One further comment: There is a mismatch in the output stage caused by the fact that half of the 6AS7 sections have the load in the plate and half in the cathode. The original circuit in *Fig. 1* tried to compensate for this by using unequal resistors in the phase inverter, which was the second section of the second 12AT7. When I tried to allow for this after my modifications, any method I used decreased the performance of the circuit (asymmetrical clipping and reduced power output), so I finally gave up and let the 100% negative feedback handle the compensation.

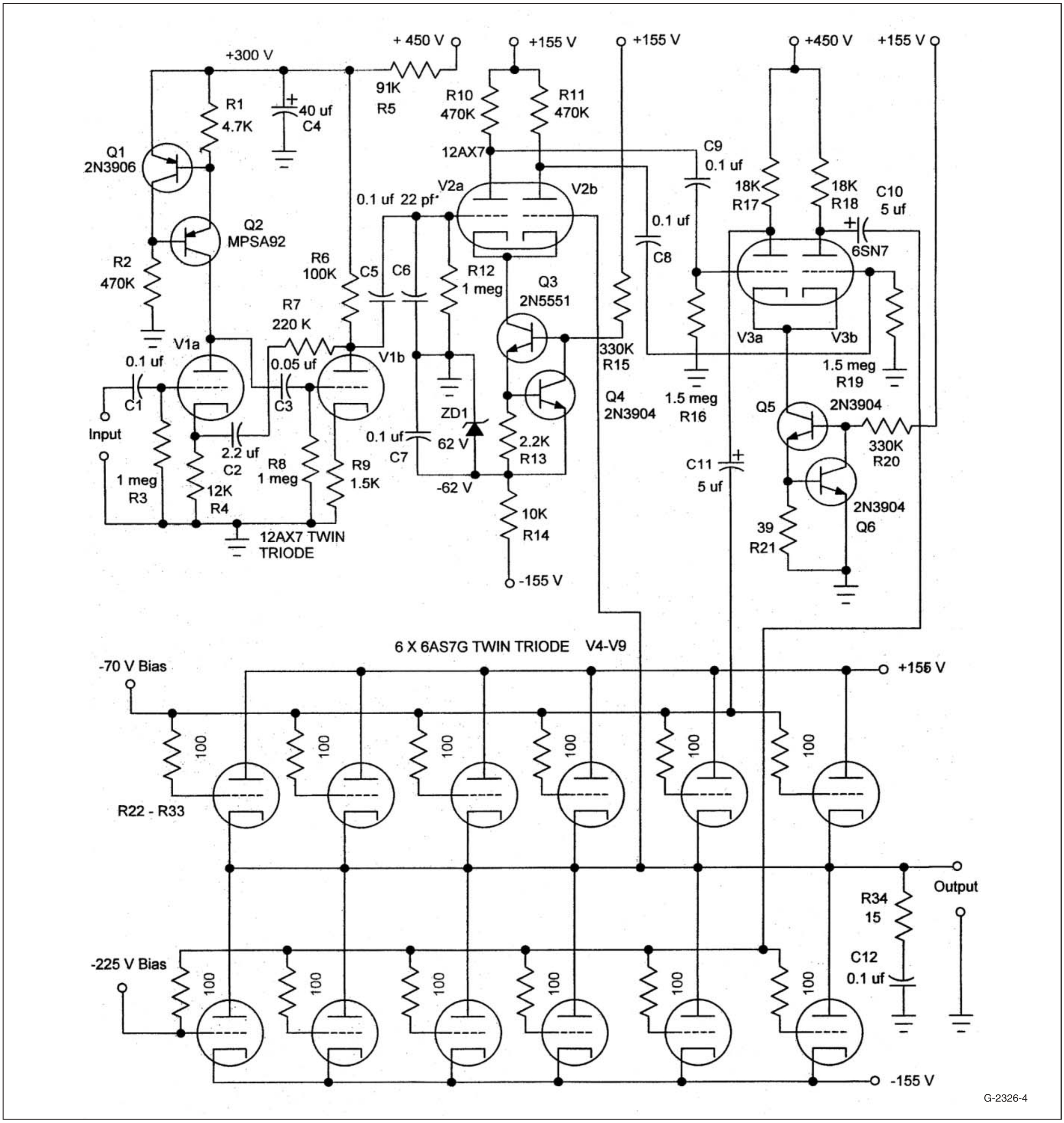

FIGURE 4: Completed amplifier with cathode-coupled amplifiers and current sources.

RESULTS

I was very pleased to find that the amplifier clipped at a little more than a whopping 50W RMS output into 16Ω, almost 40W into 8Ω, and 25W into 4Ω. The output impedance is around 0.5Ω. The amplifier is very quiet, producing less than 0.002V RMS of noise at the output with the input shorted to ground. The voltage gain is a little over 17 (almost 25dB) and the amplifier needs 1.6V RMS for full output.

At 4W output, the frequency response is dead flat from 10Hz to 60kHz, and is 1dB down at 100kHz. At 45W it is flat from 10Hz to 25kHz, and is 2dB down at 55kHz; it drops off rapidly to 7.5dB down at 100kHz.

Harmonic distortion is depicted in *Fig. 6* and shows an improvement over the original amplifier. Unfortunately, I did not perform distortion tests on my old amplifier before I modified it. (It

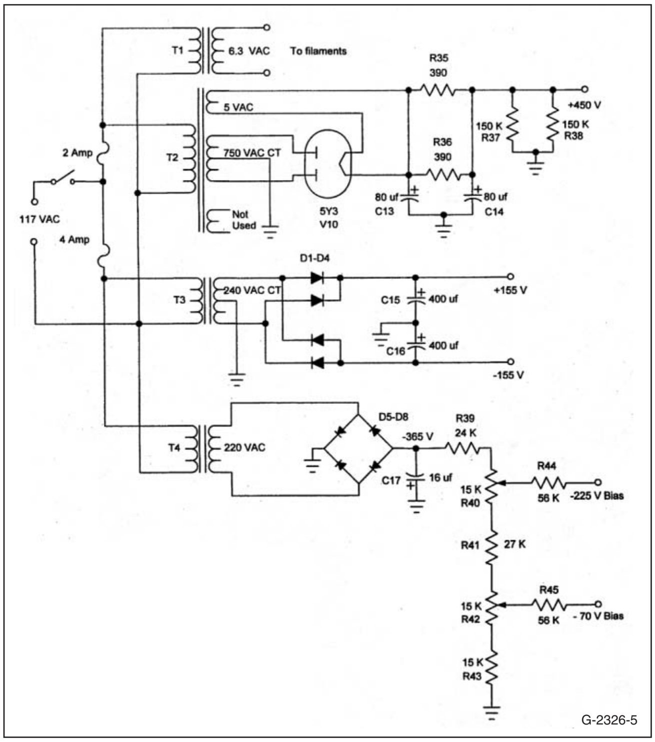

FIGURE 5: Latest power-supply modification.

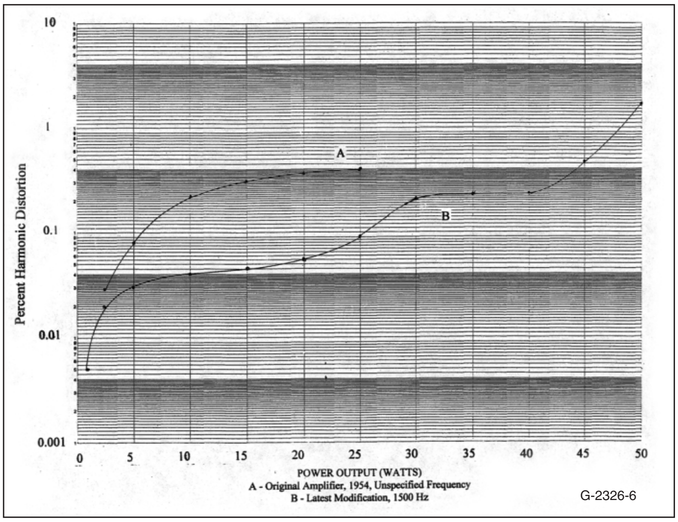

FIGURE 6: Distortion comparison.

never occurred to me that I might write this article.) The graph of the old amplifier was reproduced from the graph in the original article.

The frequency used for the test was not specified in the article, although you could assume it would be in the neighborhood of 1 to 2kHz. The distortion tests I have recorded here for the new amplifier were done at 1500Hz.

CONSTRUCTION

I constructed the original amplifier on two different chassis as was the custom in the early '60s. I built the power supply on one chassis and the amplifier on the other, with a cable connecting the two. All the wiring is point-to-point, except for the added current sources which I built on two small pieces of perfboard attached to the amplifier chassis in convenient locations.

Having two chassis caused a problem for me from the beginning. If I were starting again from scratch, I would build it all on one. I used an eight-conductor cable that plugged into an octal socket on the power supply. Six of the conductors were used for each of the power supply components: ground, +450V supply, +155V supply, –155V supply, –70V bias, and the –225V bias. This left the 6.3V filament supply.

Unfortunately, I had not adequately allowed for the fact that each of the 6AS7Gs requires 2.5A of filament current. The wire size in the cables was too small and became pretty warm when I first tested the amplifier, so I paralleled more wires and taped them to the cable. There was a second octal socket I originally installed to power a Dynakit preamp I built along with the amplifier. Some early Dynakits did not have their own power supply and tapped into the supply of the power amplifier. Later, when I quit using that preamp, I used that socket to further parallel the filament supply.

Photo 1 shows the completed and modified amplifier. As you can see, it is strictly utility.

If you wish to build the amplifier, as I mentioned, I suggest you build it on one large chassis. *Table 1* is a parts list for the amplifier and power supply. If you have trouble finding the 5µF 700V electrolytic capacitors listed there, you can use two 400V 10µF capacitors in series for each one instead.

ADJUSTING THE OUTPUT BIAS

The best way to adjust the bias when the amplifier is first tested is as follows:

1. Attach a power resistor of 10 to 30Ω

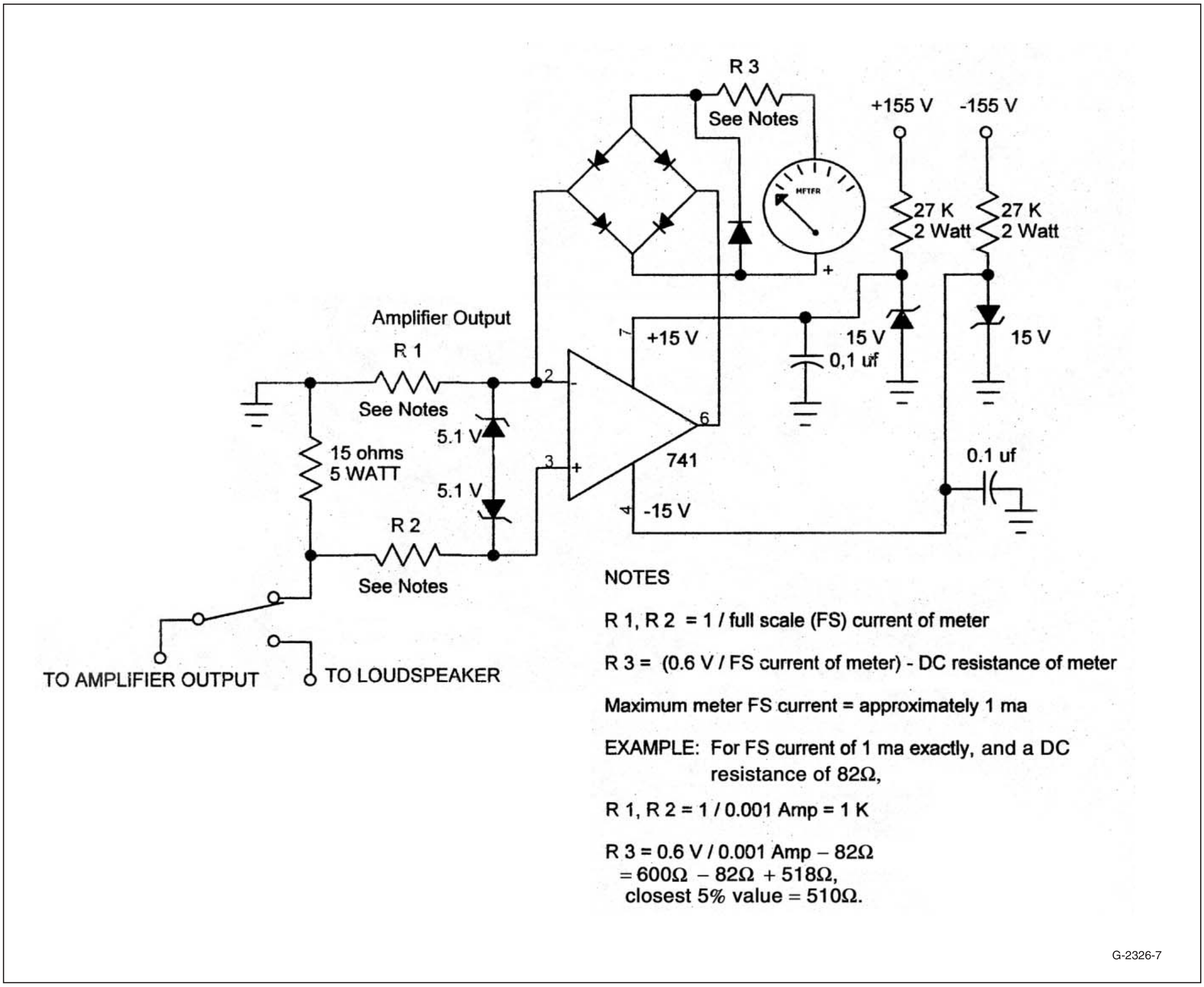

FIGURE 7: Meter driver circuitry (see text).

to the output.

2. Remove the 6AS7Gs from the amplifier, then turn the amplifier on.
3. Adjust the bias on the 6AS7s with the output load on the cathode to –70V.
4. Adjust the bias on the 6AS7s with the output load on the plate to –225V.
5. Turn the amplifier off and replace the 6AS7Gs in the sockets.
6. Attach a voltmeter across the output on a scale of about 15V. I like to use an analog voltmeter for this task. Turn the amplifier on and let it warm up for a few minutes.
7. Looking at the meter, move one of the bias potentiometers so that the needle moves toward zero. When you get the needle close to zero, I suggest you wait about 5 minutes. Then change to a 1 or 1.5V scale on the meter and adjust it as close to zero as possible. You should be able to get it within 100mV or so.
8. Leave the amplifier on for a few hours and then readjust it again. Plan on readjusting it from time to time because as the tubes age it will tend to drift a bit. After a month or so of use, it drifts very little once it warms up.

I found that it is handy to build a meter permanently onto the chassis for periodically balancing output tube bias. I have included a schematic (*Fig. 7*) for a driver circuit for an analog meter with a full-scale (FS) current of about 1mA or so. Included in *Fig. 7* are formulas for calculating resistor values for the circuit depending on the actual FS current of the meter.

Except as indicated, the resistors and zener diodes are all ½W and the capacitors are ceramic disks. The other diodes can be low-power silicon rectifiers such as 1N4001 series, and the op amp is a 741 or equivalent. You then adjust the bias for the lowest possible reading on the meter.

If anyone has any ideas about how to further improve the amplifier, I'm all ears! This story is probably not over. ❖

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Filament and High Voltage Power

Further hints on the care and feeding of tubes from an expert.

By A.J. van Doorn

Reading Mr. Kornacker's article entitled "Current Regulated Heater Supply" in *aX* of April 2003 prompted me to look a bit closer at the behavior of heaters and filaments.

FILAMENT DATA

Calculating the hot heater resistance from the tube specs and the inrush current from the measured cold resistance produced the results shown in *Table 1*.

The inrush current varies from 2.7 to 7.2 times the nominal value. The calculated inrush current values will be a bit lower than listed because the power-supply inner resistance and the cold resistance are adding up. The 813 and 211 have directly heated filaments with a low mass and a high inrush current, so they heat up quickly. Still, these Sylvania and RCA tubes have a long lifetime.

The 12AX7 and ECC83 each have two well-matched heaters. The Softek tubes (EL34, 6H8C, 6922, and 12AX7) have a higher resistance when cold than their continental mates and consequently less inrush current and an increased lifetime. But when applying current, the difference in behavior is rather small and all tubes follow the same curvature. For both the four subminiature and four octal tubes, I listed the voltage with respect to the heater current and produced the list in *Table 2* and the curves in *Fig. 1*.

At 300mA the subminiatures 6111WA and 6021WA have a voltage from 6100 to 6250mV, meaning a heating power between 1830 and 1875mW, which is only a little less than the nominal 1890mW (6300mV × 0.3A). At 600mA the octal tubes 6SN7GT and 6H8C have a voltage from 5600 to 6250mV, and a heating power from 3360 to 3750mW, which is also a little below the nominal 3780mW (6300mV × 0.6A).

The specs given by Svetlana allow a voltage of 6300 ±600mV (±9%). This is the same tolerance as in the mains supply voltage in other countries.

As you can see in *Fig. 1*, the change in heater resistance counteracts the change in voltage. This means that the

TABLE 1

TUBE	HEATER		RESISTANCE OHM		INRUSH AMPS	
MODEL	V	A	HOT	COLD	A	X I NOM
813	10	5.0	2.0	0.3	33	6.6
211	10	3.5	2.9	0.4	25	7.2
KT88	6.3	1.6	3.9	0.8	7.7	4.8
EL34	6.3	1.5	4.2	1.24	5.1	3.4
EL84	6.3	0.76	8.3	1.8	3.5	4.6
EZ80	6.3	0.9	7.0	1.5	4.2	4.7
6SN7	6.3	0.6	10	1.8	3.5	5.8
6H8C	6.3	0.6	10	2.4	2.6	4.4
6021WA	6.3	0.3	21	3.0	2.1	6.8
6111WA	6.3	0.3	21	3.1	2.0	6.7
6922	6.3	0.3	21	5.7	1.1	3.6
E88CC	6.3	0.3	21	3.2	2.0	6.7
12AX7	6.3	0.3	21	8.0	0.8	2.7
ECC83	6.3	0.3	21	3.4	1.9	6.2

TABLE 2

TUBE	6111WA		6021WA		6SN7GT		6H8C	
NR	1A	1B	2A	2B	3A	3B	4A	4B
Heater mA	mV	mV	mV	mV	mV	mV	mV	mV
100	600	650	595	590	225	240	270	280
150	1500	1600	1480	1440				
200	2850	2950	2800	2800	730	800	760	800
250	4380	4450	4320	4320				
275	5200	5300	5160	5180				
300	6150	6250	6100	6100	1630	1750	1580	1650
325	7150	7230	7080	7080				
400					2850	3000	2700	2800
500					4350	4500	4100	4250
570					5500	5650	5100	5350
600					6000	6250	5600	5850
630					6550	6750	6100	6400

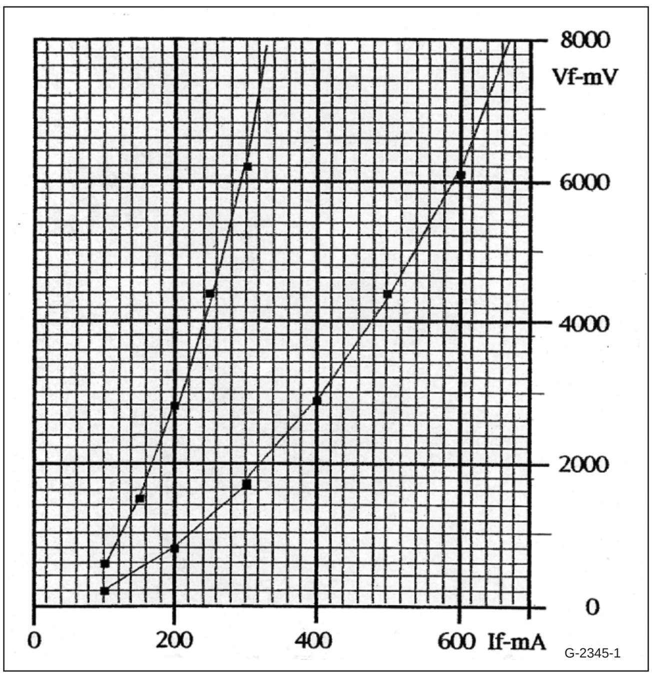

FIGURE 1: Tube voltage/heater current.

cathode also works very well below the nominal heater voltage. A change in heater voltage of 10% will give a 5% change in heater current and a 15% change in energy instead of 20% with a common resistor.

The easier an element can provide an electron emission, the lower its so-called work function. To lower the work function of pure tungsten, filaments are thoriated and work at a temperature of 1800K. By carbonizing the emitter, the rate of evaporation of the thorium layer is reduced by a factor of 6.

The oxide-coated cathode is very efficient (20× better than tungsten) and provides a high emission current at a temperature of about 1000K. It contains a metal sleeve of konal that is coated with the oxides of barium and strontium, which combine efficient operation with a long life.

According to Langmuir-Child's law, the space charge current is independent of the temperature and the work function of the cathode. Thus, no matter how many electrons a cathode can supply, the tube's geometry and the ap-

TABLE 3

TUBE	VOLTAGE		6300MV		6900MV		5700MV		
EL34	A	mW	%	A	mW	%	A	mW	%
nominal	1.50	9450	100						
1	1.50	9450	100	1.58	10900	115	1.4	7980	84
2	1.50	9450	100	1.58	10900	115	1.4	7980	84
3	4.56	9830	102	1.64	11315	120	1.47	8380	89
4	1.49	9440	99.9	1.57	10830	114	1.4	7980	84
KT88									
nominal	1.6	10080	100						
1	1.64	10330	102	1.73	11940	118	1.54	8780	87
2	1.64	10330	102	1.73	11940	118	1.53	8720	87
3	1.67	10340	103	1.73	11940	118	1.56	8890	88
4	1.85	11655	116	1.92	13250	131	1.75	9975	99
6550									
nominal	1.5	9450	100						
1	1.50	9450	100	1.59	10970	116	1.41	8040	85
2	1.53	9640	102	1.63	11245	119	1.44	8210	87
3	1.54	9700	103	1.64	11315	120	1.44	8210	87
4	1.53	9640	102	1.60	11040	117	1.42	8100	86

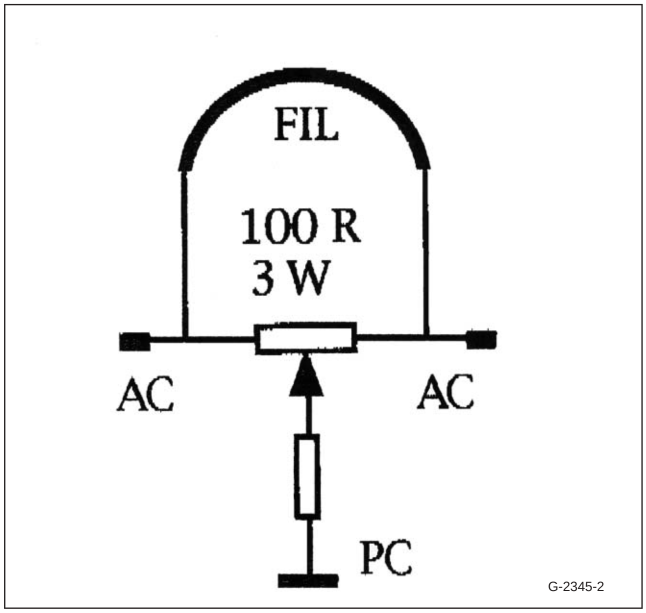

FIGURE 2: Potentiometer 100R-3W set for lowest hum.

plied potentials will determine the amount of current that is collected by the anode.

NON-REGULATED POWER SUPPLY

As mentioned previously, the energy in the heater follows the changes in voltage. I made a test with three sets of four power tubes. The results are shown in *Table 3*.

Only four samples may not be representative for the lot, but it gives me the impression that—apart from some exceptions—the heaters of power tubes are well matched to their specs. In most countries, the mains voltage is stable enough and a regulator is not needed. With an AC supply, the heat capacity of the cathode prevents its temperature from following the instantaneous variation in heater current. In preamps, you may hear the induced noise to the cathode, but you may get rid of it by giving the heater a positive voltage with respect to the cathode (no emission from filament to cathode).

For directly heated power tubes as the 211 or 813, a stabilized supply is seldom used. These tubes only allow for a 5% tolerance in the filament voltage. Mostly using them in SE class A mode filament noise is a major problem. With its low gain, the 211 might work with AC for the filament (*Fig. 2*). The potentiometer across the filament must be adjusted for minimum hum. Using high gain pentodes such as the 813 or with sensitive speakers on a DC supply— even non-regulated—is a must.

The 10V DC over the length of the filament will produce an uneven bias and thus an uneven emission of the filament. While relatively small with a high bias voltage, it may become important at low bias and high currents during modulation of the grid. In high power applications, sometimes the filament polarity is changed at regular intervals to balance out the filament wear.

REGULATED POWER SUPPLIES

As already mentioned, the regulated DC heater supply prevents noise. You can regulate the voltage, the current, or the energy to the heaters. With proper regulation, the filament energy is steady and independent of variations in the mains supply. Without risking a too-low emission, you can lower the heating energy. The filament will like it and last longer.

Before regulation, the AC voltage is changed into DC by a bridge rectifier, followed by a filter capacitor (*Fig. 3*). To reduce spikes and strain on the bridge and the filter capacitor, you can put a resistor in the AC line. This will also lower the high inrush current. For proper operation, most regulators need a minimum voltage drop of 3V; hence the regulator input voltage must be at least 3V higher than the output voltage of the regulator, even at a low mains supply.

The power loss in the regulator is its voltage drop times the passing current, so use a small heatsink. With two or more heaters, series connection is more efficient than in parallel mode.

REGULATION

Voltage Control

Voltage regulation supplies a steady voltage to the heaters (*Fig. 4*). In accordance with *Table 2* and at 6250mV tube 1b consumes 1875mW and tubes 2a and 2b 1900mW, a deviation of 1.3%. From the older octal, tube 3b gets 3650mW and tube 4a 3950mW, still only 5.3%

If you want to connect two similar 6V tubes, put the filament in series to have a low current through the regulator and less heat loss as well. With the newest switching regulators from National, it is quite easy to have a powerful DC stabilizer, as needed for the direct-heated 211, and so on. Look for it at www.national.com.

Current Control

Current control supplies a steady current to the heaters (*Fig. 5*). Here tube 1b gets 1875mW and nr 2, 1830mW, a

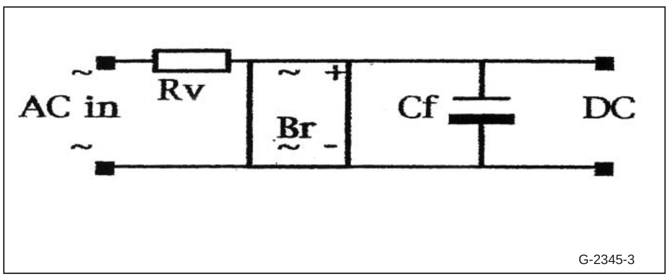

FIGURE 3: Small ripple on Cf will not pass the regulator.

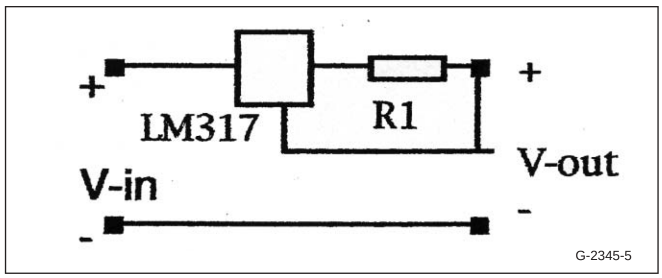

FIGURE 5: Iout = 1250: R1mA. 8R2 = 153mA, 4R1 = 305mA.

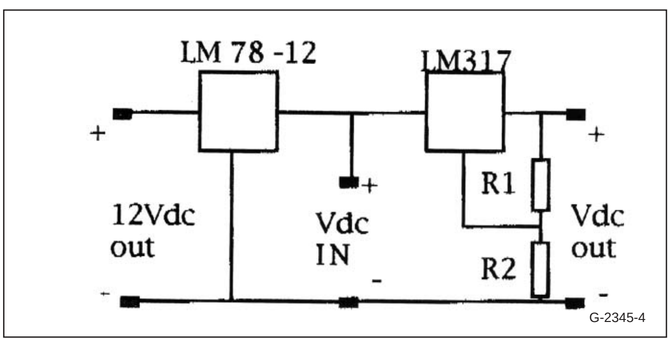

FIGURE 4: Left fixed 6/12V reg. Right adjustable reg. Vout = 1.25× (R2 + R1): R1V DC.

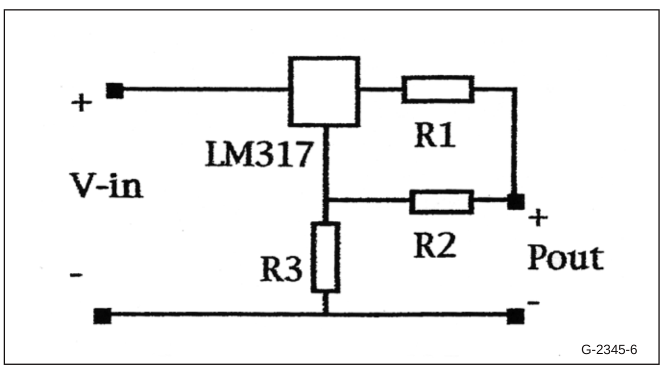

FIGURE 6: I–out × (R1 + R2): R2 = 625mV. I–out × R1 = 625mV.

change of 2.4%. Also, tubes 3b and 4a have 3750 resp. 3360mW, a change of 10.4%. The advantage of current control is its soft start capability. The temperature of and the voltage on the heater rise slowly until stabilization has been reached. In Mr. Kornacker's example two 150mA heaters are connected in parallel to a 300mA source; here if one heater gives up the other gets the full 300mA. Thus, with current control heaters are in series.

Compound Control

Compound (power) control can be done as in *Fig. 6*. With this simple circuit, you can make a very precise heat control. I wanted an output of 1A at 10V, hence the nominal filament resistance of 10Ω. The energy output is 10W. For a drop of 625mV at 1A, R1 must be 625mΩ. With R2 at 625mV the output voltage is 10,000: 625 = 16 × higher (than on R2) so R3 must be 15 × R2, if R2 = 1k then R3 = 15k. A lower load resistance will increase the drop in R1 and consequently lower the voltage on R2. With a load changing 10% both current and volts will change 5%.

Example:

Rload becomes 10Ω less 10% = 9Ω.
Amps increase 5% to 1.05A and drop in R1 = 657mV.
R2 automatically drops to 1250 − 657 = 593mV, corresponding to
an output of 16 × 593 = 9488mV, equal to 1.054A in 9Ω.
The power is 9.5 × 1.05 = 9.97W, a variation of only 0.3%.

SAFETY

In the good ol' days, radios and amps didn't have any precautions against high inrush currents or cathode stripping and still worked very well for a long time (but not the cheap stuff). Why a nice soft-start for small preamp tubes and not for the power tubes? Some people use the pretty expensive 300B, and I use 211, 813, and KT88, but forget the soft-start. In the DC supply for the 211-813, you can use small series resistors to cut the high inrush current in half.

By using a separate heater transformer, you can reduce the AC inrush current with a series NTC resistor in its primary—a bit like in the old and ugly AC-DC radios (*Fig. 7*). In balanced mode, a power amp has two tubes that have the same anode current, often sharing a bias circuit. Here you can use a 12.6V AC supply with the filaments of both tubes in series, so if one filament fails, the other tube stops conducting as well.

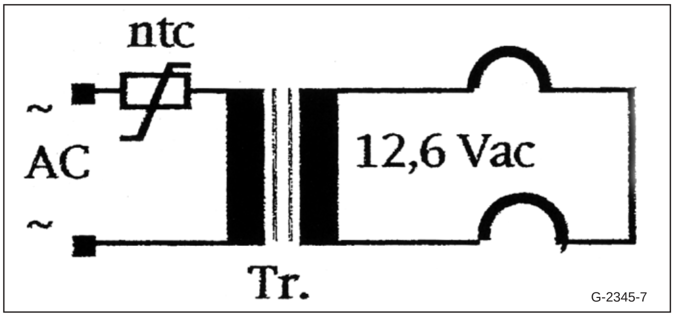

FIGURE 7: Using a separate heater transformer, as in an AC-DC radio.

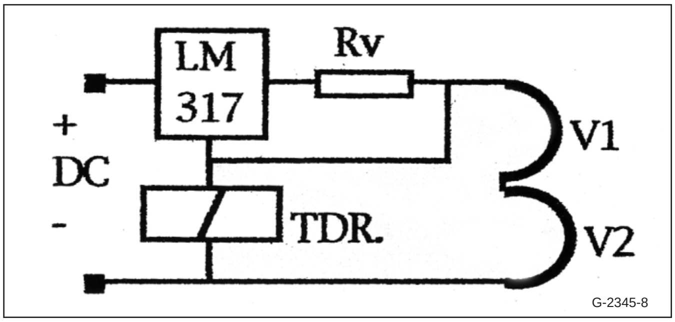

FIGURE 8: After slowly reaching 12V DC, the relay pulls and switches on the HT.

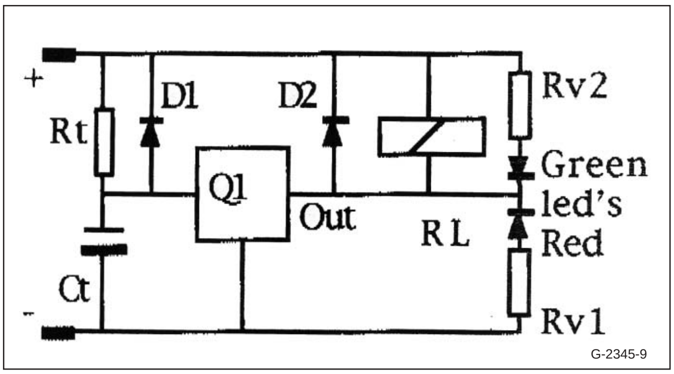

FIGURE 9: You can make a very precise TDR with an LM431A.

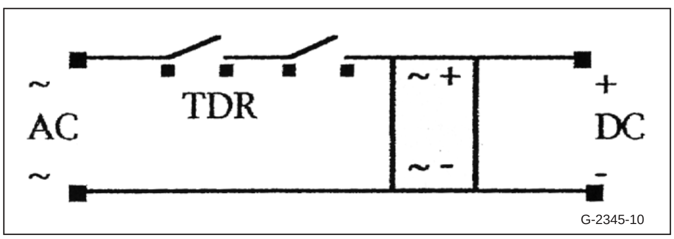

FIGURE 10: Using a phi filter.

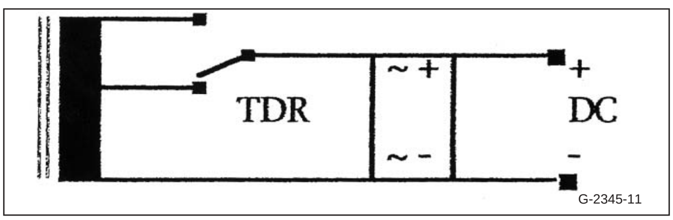

FIGURE 11: Use two relay contacts in series if you need to switch high voltages.

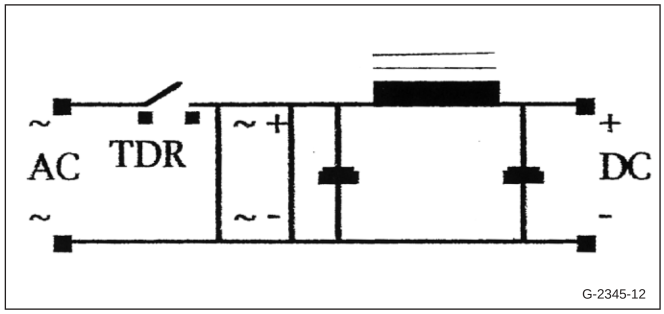

FIGURE 12: Using a split AC supply voltage.

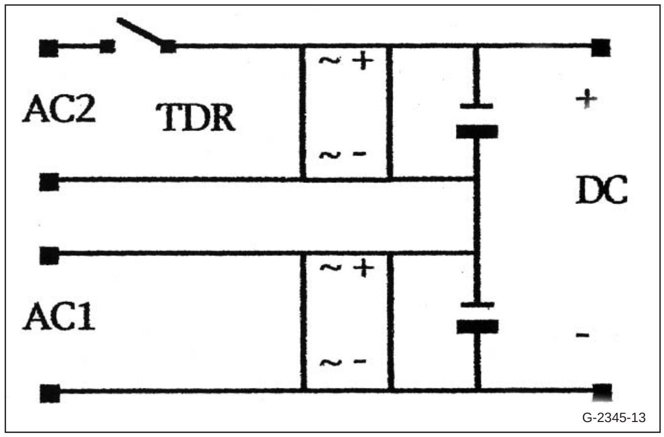

FIGURE 13: When TDR switches on, the current of the power tubes reaches the full value.

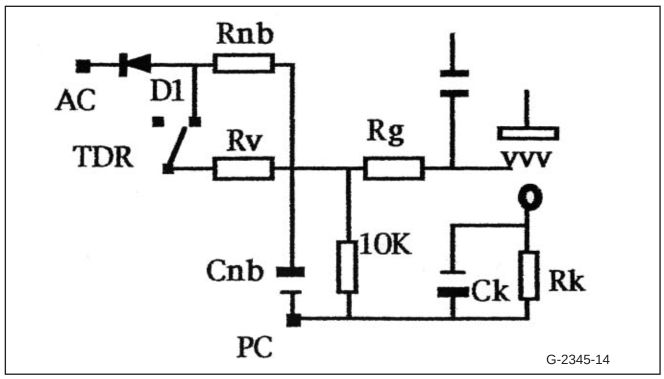

FIGURE 14: Applying negative bias.

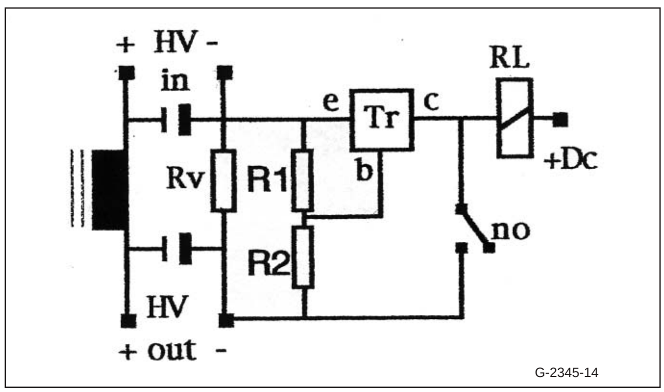

FIGURE 15: Set voltage divider R1-R2 to trip Tr at the desired maximum current.

TIME DELAY

To prevent the high voltage from coming in too quickly, you can use a commercial time delay relay (TDR). With a current control for the heater, you can make a TDR by connecting a relay in parallel to the heater. After slowly reaching 12V DC, the relay will pull and switch on the HT (*Fig. 8*). With an LM431A, you can make a very precise TDR (*Fig. 9*).

At switch-on, Ct charges through Rt and LED R is on through Rvl and RL. Ct charges up to the reference voltage of 2.5V and QI conducts, RL pulls, LED G lights up through Rv2, and LED R is off. At switch-off, Ct discharges through D1, and D2 protects Q1 + RL against spikes.

HIGH VOLTAGE

Analog to *Fig. 3* DC is made from AC. The HV-DC passes through a filter to eliminate the voltage ripple. SE, Class A amps need a far better filtering than the balanced ones where the ripple currents are cancelled out. Filter elements can be capacitors, resistors, chokes, or MOSFETs. Mostly, a phi filter is used (*Fig. 10*) but all have a high inrush current in common.

In general the standard relay contacts cannot switch high AC voltages and you may need to put two contacts in series to prevent arcing of the contacts (*Fig. 11*). However, both the rectifiers and filter caps suffer from this rude switching-on, and it is far better to use a split AC supply voltage as in *Fig. 12*. With the initial low HV and a fixed bias, the current of the power tubes starts at almost zero and will reach the full value when TDR switches on (*Fig. 13*).

Often the DC-HV is so high that two filter capacitors are series-connected. To equalize the voltage on the caps, bleeder resistances are used—not a very good practice. It would be better to use a split DC power supply as you see in *Fig. 13*.

If all filter capacitors can withstand the maximum HV (1.5×V AC), you can use the TDR to control the negative bias instead of the incoming AC voltage. A high enough negative bias will prevent the power tubes from conducting; it can be applied as you see in *Fig. 14*. TDR removes the high negative bias voltage. Rv sets the bias value. Rk cares for the auto bias and is—in general—a high wattage resistor.

I use a combination of a fixed and auto-bias in such a way that both have the same voltage. For example, with balanced EL34s in triode mode, Rk is 100Ω, Ck has 15V DC, and Rnb is set to have 15V DC on Cnb as well. Please note that in this case TDR only switches the AC supply (*Figs. 12* and *13*). There is no need to change the negative bias voltage of, say, 15V DC; as with this bias, no current flows at start-up.

OVERLOAD PROTECTION

For protection, fuses are used in the power supply, generally in the primary or secondary of the PT. This is quite useless for protecting the power tubes. A fuse in the HV line of the OPT is better, as capacitor charging currents don't pass through the fuse. It is best to measure the total supply current. If the voltage of resistor Rv rises to 600mV, Tr will conduct, and RL will close, resetting TDR, and removing the power. RL stays on by its holding contact, if the currents drop afterwards. ❖

Room Correction, Part 2

A closer look at room correction and loudspeaker response in this four-part DSP preamp project. **By Rune Aleksandersen**

With the advent of the DSP processor, it is now possible to correct defects in the loudspeaker response. To some degree it is also possible to correct deficiencies that the room adds to the loudspeaker response. The goal of a correction system should be to improve the perceived sound of the stereo system.

I started to use some correction curves and pre-filter wave-files, and soon realized that making a DSP filter that works well is a challenge. I needed to consider psychoacoustic properties as well as technical limitations to what a DSP processor can do.

One important property of stereo is the Haas-effect[2]. The balance of loudness of two sources is perceived as directional info within the first 5–20ms. Because of this, amplitude matching of the direct sound from a pair of sound sources is important in order to get a good stereo perspective. This is what DSP is good at. You can match frequency response of a pair of loudspeakers within fractions of a dB with the amplitude corrected to a flat response.

At first, you would think that correcting for a flat frequency response and perfect impulse response at the listener's position would be the solution. But there are several problems with this approach: Psychoacoustic research has determined that peaks are more objectionable than notches in a frequency response[3,4,5]. If I make a filter that fills in all the tiny notches in a frequency response, this filter will sound horrible. The notches will be removed at the measured position only, with peaks occurring when you move your head just a little bit. Smoothing the measured response helps a lot.

Quality of the stereo image is very dependent on the ratio of direct sound to diffuse sound[6,7]. The ratio depends greatly on dispersion patterns of the loudspeaker since more directive de-

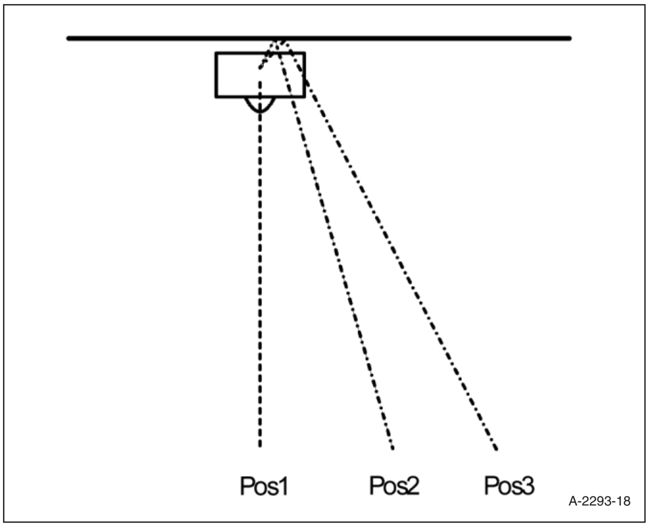

FIGURE 18: Back wall reflections path for three positions with speakers close to wall.

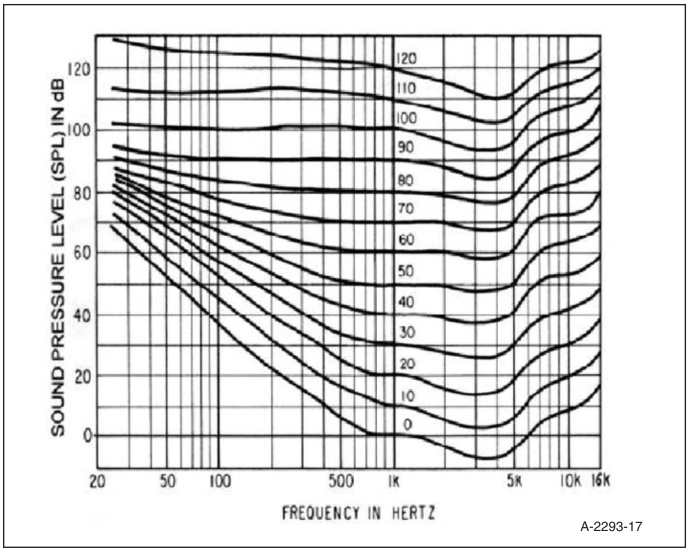

FIGURE 17: Fletcher-Munson curves.

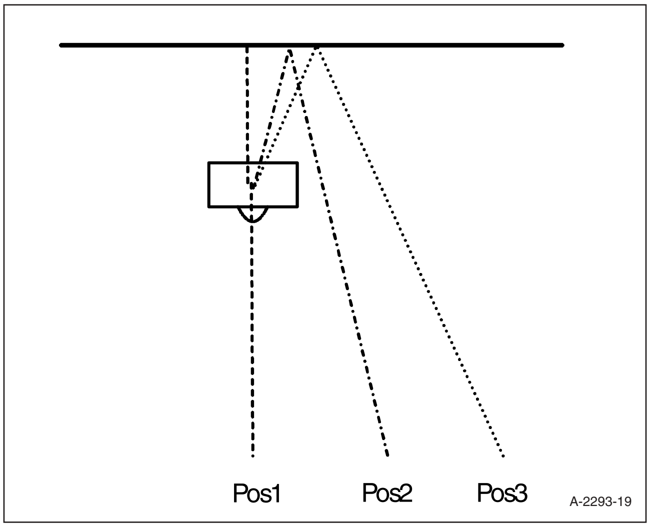

FIGURE 19: Back wall reflections path for three positions with speakers away from wall.

signs have a better ratio. You can improve the ratio by using digital signal processing for optimizing the direct sound and removing early reflections.

The diffuse sound is important to perceived spectral balance and the reverb added to the sound coming from the loudspeakers. This diffuse sound field varies with the directional properties of the loudspeakers as well as the shape and size of the room and listener's position. There will be varying degrees of absorption of high and low frequencies depending on construction materials used in the room. Since there is always absorption of the high frequencies, attempting to equalize the in-room response flat will fail, leading to a too-bright sound.

The human ear is not linear, as you can see in the Fletcher-Munson curve (*Fig. 17*). The perceived balance of low and high frequencies depends on the playback level. When playing back music at low levels, the bass and treble weaken. (This is the reason for inventing the loudness button.) While at higher levels, the bass and treble are perceived stronger compared to the midrange. Target curves used in the high- and low-frequency ranges might help in balancing the playback to the room and the listening levels to be used.

Low frequencies, meaning frequencies below the Schroeder frequency (150–200Hz), are mainly influenced by room modes. Because wavelengths are similar to room dimensions, in some spots there will be amplification; other spots will be sucked out. Because of this, loudspeaker placement has much

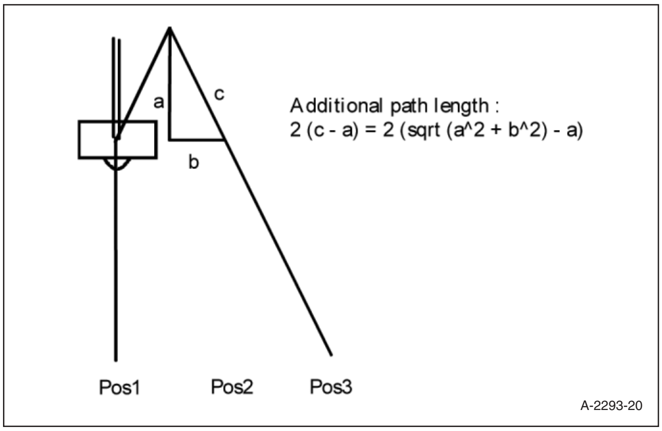

FIGURE 20: Geometry for added path length.

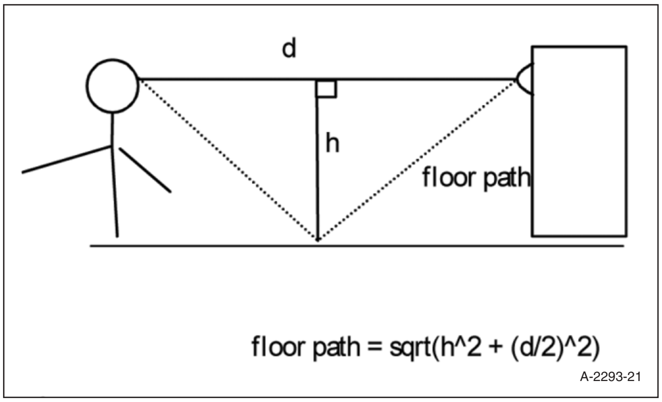

FIGURE 21: Floor reflection geometry, assuming listener in height with loudspeaker.

influence on response for the very low frequencies. In this band, there will typically also be room gain of 3–6dB towards the low end. Low frequency peaking is annoying and should preferably be removed if possible.

In trying to figure out a practical approach to loudspeaker and room correction, I will discuss different methods in the following groups:

1. Direct sound (including early reflections)
2. Diffuse sound and spectral balance
3. Low frequencies

DIRECT SOUND

The direct sound is of primary importance to image localization, stereo perspective,[7] and perceived coloration.[8] The following factors influence the direct sound:

- driver response
- loudspeaker box shape and construction
- edge diffraction
- nearby objects
- nearby walls, floor, and ceiling

About 100 years ago, [135 now.—Ed.] Lord Rayleigh[9] developed his Duplex Theory. Image localization is a combination of two processes in the horizontal plane: a time-domain process that is called Inter-aural Time Difference (ITD) and a frequency domain Inter-aural Intensity Difference (IID). Basically, the mind measures both the time differences between our ears as well as the intensity difference for determining direction. For frequencies below about 1.5kHz the ITD is dominant, while above this frequency, IID is most important.

Perceived coloration of the direct sound is found to be up to 25ms after the initial sound, and the directional location information is mostly in the 0ms to 3–5ms range.[8] New research has also shown that the first millisecond after the direct sound arrives is the most important for determining which direction the sound came from[10].

A single reflection produces a comb-filtering effect in the frequency domain. The audibility depends greatly on the delay of the reflection as well as the loudness. Audible comb-filtering should be corrected if possible. Listener-position dependent comb-filtering might not be possible to correct successfully. Correction of reflection spikes is interesting and challenging. If not done properly, this can lead to sharp resonances in the frequency domain that are unpleasant to the ear.

I think measuring methods for obtaining a correction curve is crucial to getting good results. Since our ears operate both in the time and frequency domains, I will try to analyze the measured data in both domains. I've taken a look at two mainly different ways of placing the loudspeakers—near the wall and away from the wall.

The trigonometry in *Fig. 18* shows that placement near the wall adds little to the path length for the three illustrated listener positions. The differences in path lengths (*Fig. 19*) with a speaker placed well away from the wall are much greater and will cause listener positional dependency. Since DSP corrections are performed in the time domain, the reflections need to be placed at the same spots in the time domain for all listener positions to be most effective.

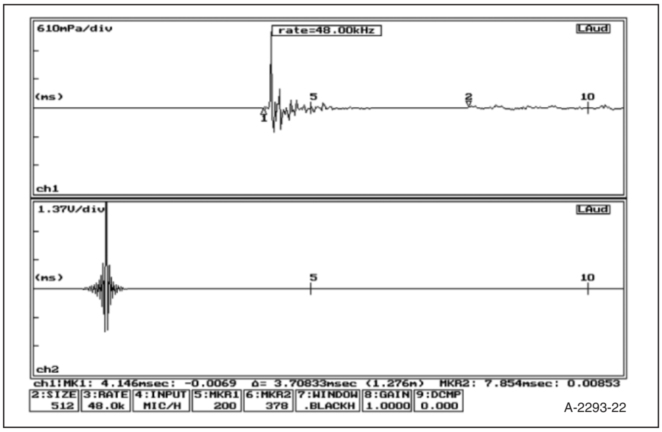

FIGURE 22: 1m semi-anechoic response.

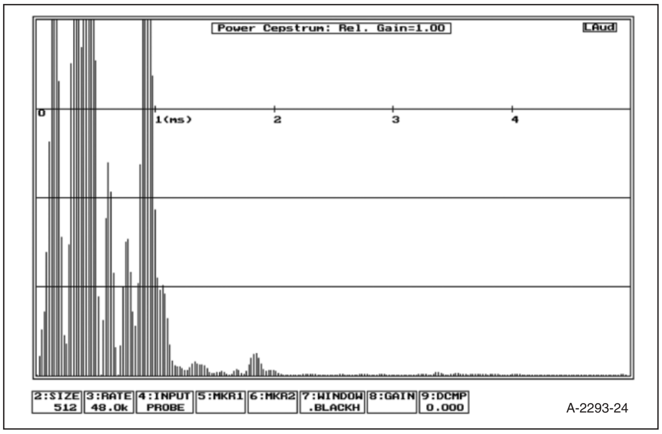

FIGURE 24: Cepstrum of the 1m semi-anechoic response.

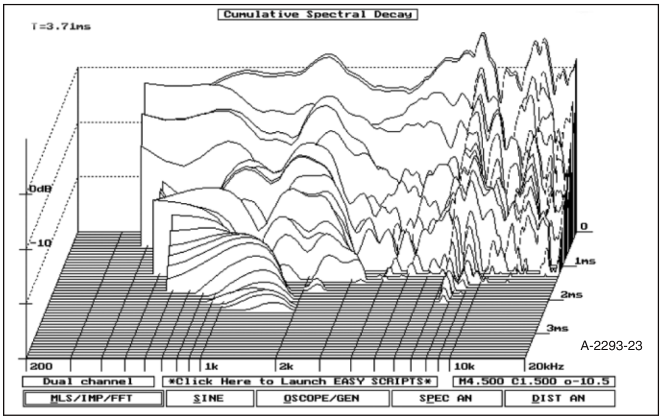

FIGURE 23: Waterfall of 1m semi-anechoic response.

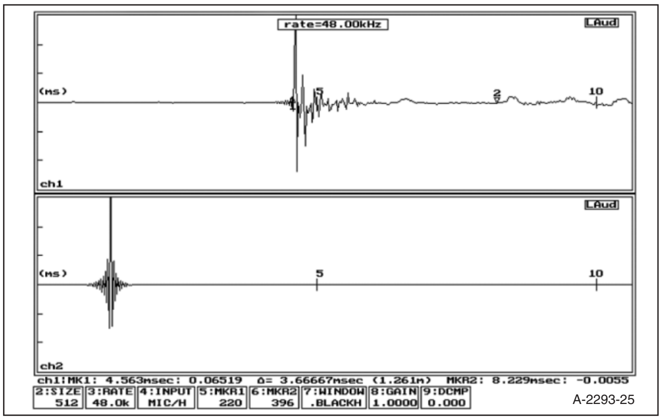

FIGURE 25: 1m response placing speaker near the back wall.

Geometry demonstrating what adds to the path for a distance **a** from the back wall and the listener position is shown in *Fig. 20.* As the distance to the back wall increases, the more positionally sensitive (the value of **b** has more significance) the path length becomes.

Assuming the same height for the listener and the speaker, the geometry is shown in *Fig. 21.* The floor path is highly dependent on listener distance, which might make floor and roof reflections difficult to correct.

The reflections from the back wall cause significant ghost images and coloring comb-filter effects. Moving the speakers well away from the walls reduces the energy of the direct sound compared to the reflected sound. This is the traditional way of placing speakers, which I think has a major shortcoming: The center image becomes weak and stereo perspective does not seem real, lacking weight for images that are in between the speakers.

By placing the loudspeakers close to the back wall and angling them strongly inwards, there is more coloration due to the back wall reflections. Also, ghost images appear. On the positive side, there is a nice feel of wholeness and 3D effect to the image as well as the illusion of the loudspeakers disappearing.

Subjectively, this makes a stronger stereo perspective; images in the center especially are more in focus. I think that coloration and ghost images can be greatly reduced by using DSP techniques and deploying more directional

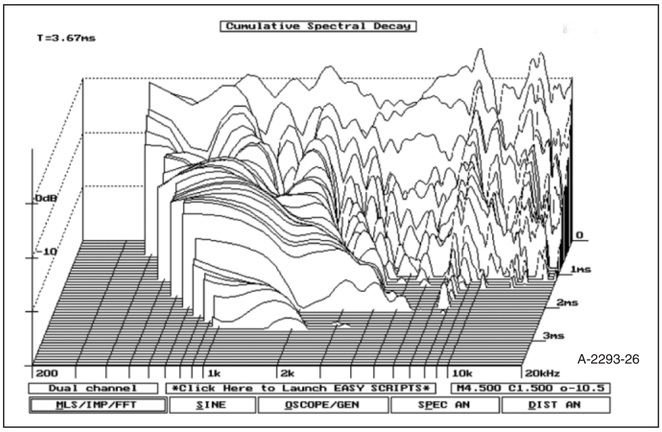

FIGURE 26: Waterfall of 1m response placing speaker near the back wall.

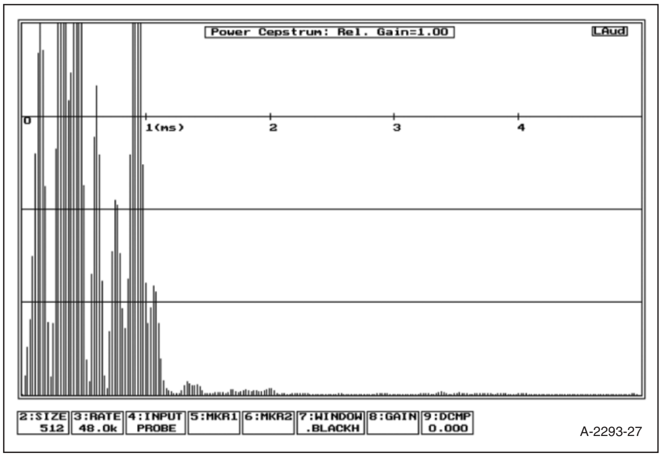

FIGURE 27: Cepstrum of 1m response, speaker next to wall.

speaker patterns. There should be several advantages to doing correction with loudspeakers placed close to the back wall:

- Position-dependent comb-filtering effects from the back wall can, you hope, be cancelled out.
- Room gain allows for less driver excursion at low frequencies, improving and giving less distortion.
- More precise and natural stereo image.

Figures 22 and *23* show the semi-anechoic impulse response and waterfall plot for the loudspeaker. Checking the geometry closer, I made some measurements with the speakers close to a 2×3m mirror wall. Compared to the near wall measurements of *Figs. 25* and *26*, I observed that there was a back wall reflection occurring after 2ms. I saw this reflection in the waterfall plot as a ridge from the lower limit of the measurement up to around 2–3kHz. This ridge of approximately –10dB will appear as coloration to the listener.

Coloration might be checked using the cepstrum transform as shown for the anechoic response. This transform basically tells something about the wiggles of the frequency response or echoes in the time domain that translates into coloration to our ears. It can also be seen as a measure of reflections or diffraction effects that are present. All cepstrum plots are done with the frequency mid-band response set to 0dB level.

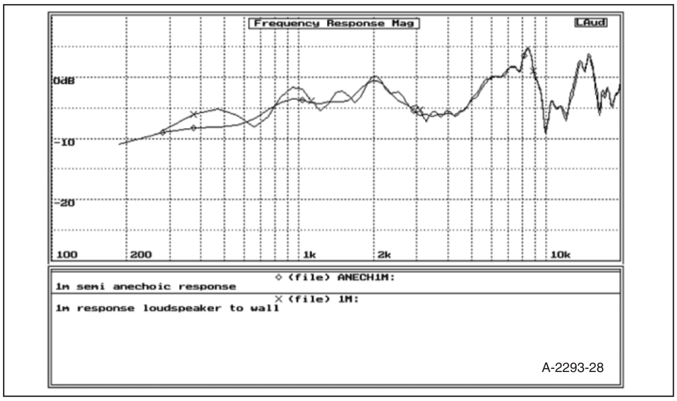

FIGURE 28: Comb-filtering effect on amplitude of single back wall reflection.

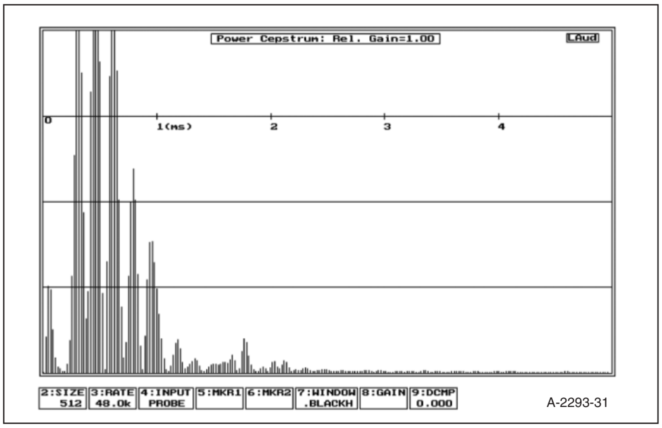

FIGURE 31: Cepstrum of 1m off-axis response.

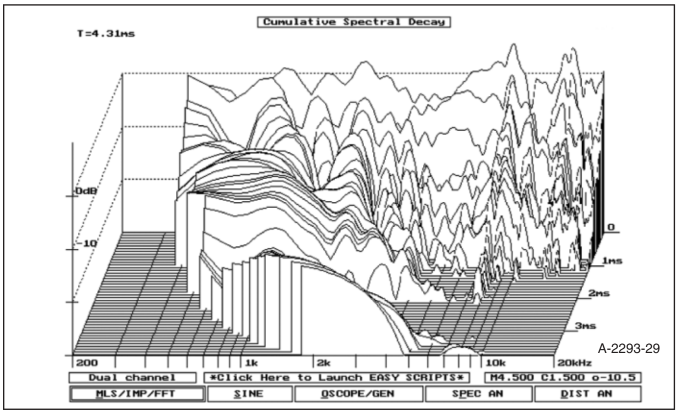

FIGURE 29: Waterfall of 1m response placing speaker near the back wall, including the floor reflection.

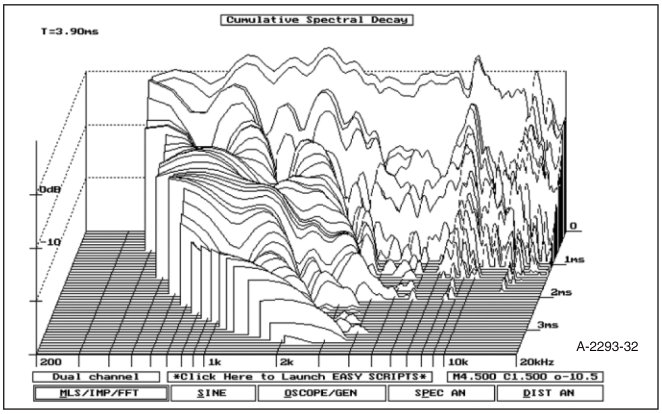

FIGURE 32: Waterfall of 1m off-axis response placing speaker near the back wall.

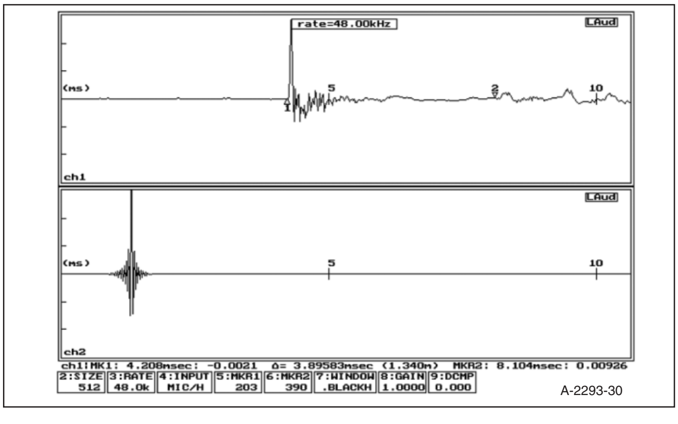

FIGURE 30: 1m off-axis response placing speaker near the back wall.

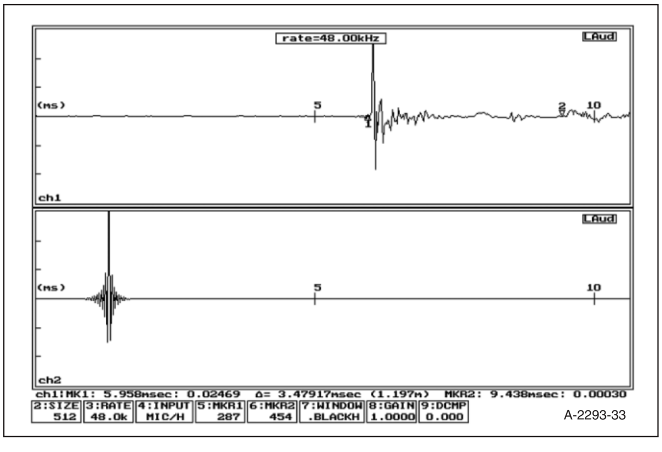

FIGURE 33: 1.6m response placing speaker near the back wall.

Figure 28 shows the comb-filtering effect comparing the 1m anechoic response with a response including the back wall reflection. You can see a comb-like frequency response in a broad frequency span from 300Hz to around 4kHz. This comb-filtering is similar for all the measured positions. The cepstrum plot of *Fig. 27*, including the back wall reflection, is very similar to the semi-anechoic response cepstrum plot.

The floor reflection is included in *Fig. 29*. The energy of the signal has almost died –30dB, and now starts to rise again.

From *Table 2*, you can see that the back wall reflection is rather constant in time. The small variation is caused mainly by the off-axis measurement angle. Floor and roof reflection placements are dependent on listener distance. In *Figs. 34* and *36*, the floor and roof reflections overlap!

The longer the distance that is measured (*Fig. 33*), the shorter the distance from the initial wave front to the floor and roof reflection will be. If I measure too far away from the speaker, the impulse response will include the floor and roof reflections in the time window of interest. I must avoid this, since those reflections obviously are listener-position dependent. For a time window up to around 3.5ms there are no floor reflections present, and this includes the back wall reflection. Therefore I will use a measurement with this time window when making correction filters.

Cepstrum of the longer distances (*Figs. 35* and *38*) seem to have somewhat less reflective energy present for the longer measurement distance. Longer distances are of such a geome-

TABLE 2
POSITION OF REFLECTIONS ACCORDING TO LISTENER POSITION

Position	Back wall reflection	Floor reflection	Roof reflection
1m on-axis	2.0ms	3.9ms	5.0ms
1.6m on-axis	2.0ms	3.6ms	3.9ms
1m off-axis	1.9ms	4.0ms	5.3ms
1.6m off-axis	1.9ms	3.5ms	3.9ms

try that reflected energy becomes stronger compared to the direct sound. I think you can see this in the cepstrum of the off-axis responses. There is a small energy peak around 2ms, being the back wall reflection.

DIFFUSE SOUND AND SPECTRAL BALANCE

The diffuse sound (or reverberant sound) is the sound occurring 20ms and longer after the direct sound. It is composed of reflected sound that comes from all kinds of directions not having any valuable phase information. Perceived timbre or spectral balance is the complex aural summation of direct and reverberant sound.

The axial response of a loudspeaker is not sufficient for determining this timbre. The whole system of loudspeaker, room, and listener all influence it. Reflected sounds from floor, roof, and side walls depend on the building construction materials. Loudspeaker directivity is also important. A directive speaker will have less reflected energy from the surroundings than an omnidirectional design.

The reverberation on the recording and in the room adds up to the listener experience. A dead room, such as an anechoic chamber, is a terrible listening room, as is a concrete apartment with no furniture in it. Reverberation response times of around 0.4 seconds

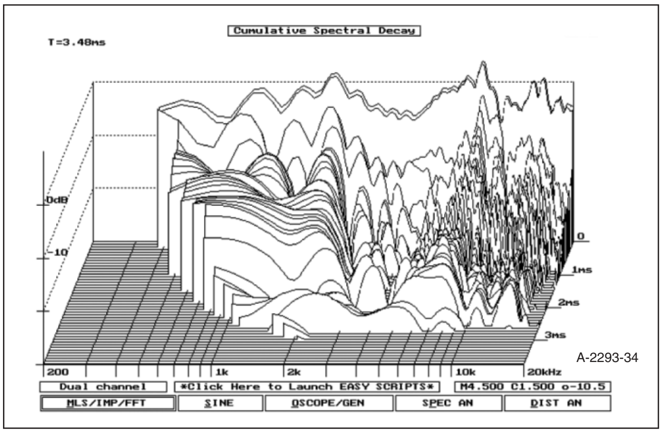

FIGURE 34: Waterfall of 1.6m response placing speaker near the back wall.

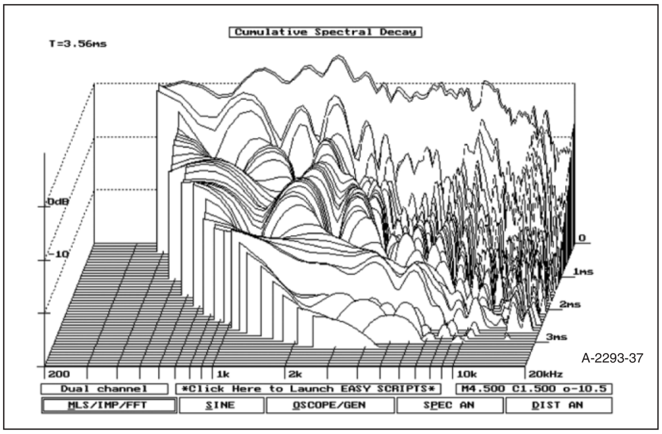

FIGURE 37: Waterfall of 1.6m off-axis response placing speaker near the back wall.

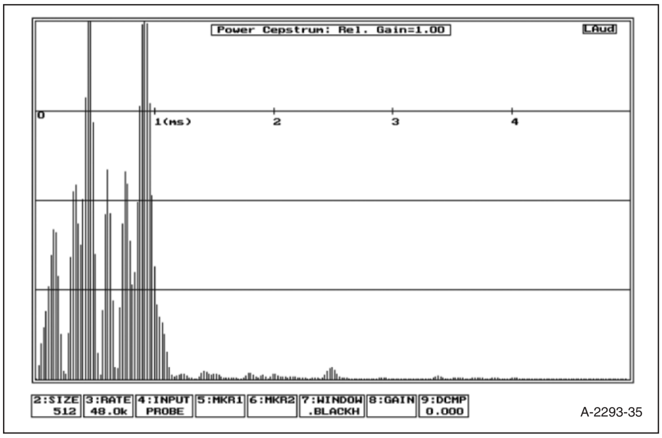

FIGURE 35: Cepstrum of 1.6m response.

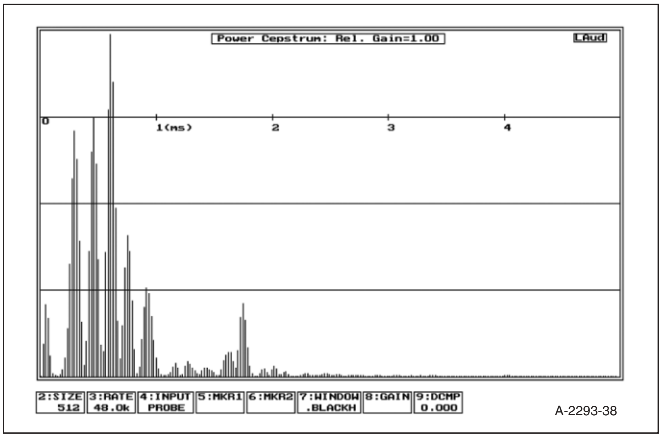

FIGURE 38: Cepstrum of 1.6m off-axis response.

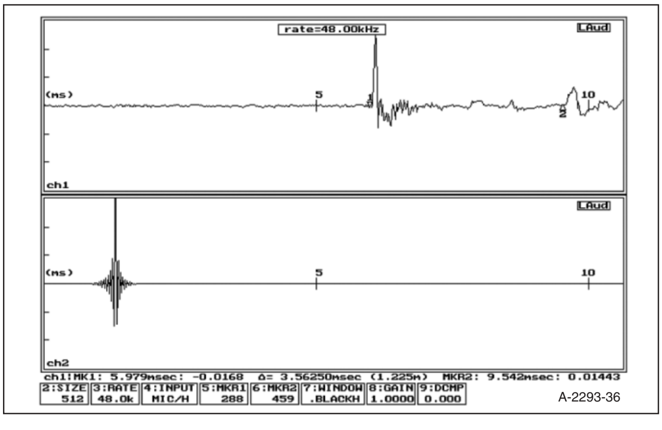

FIGURE 36: 1.6m off-axis response placing speaker near the back wall.

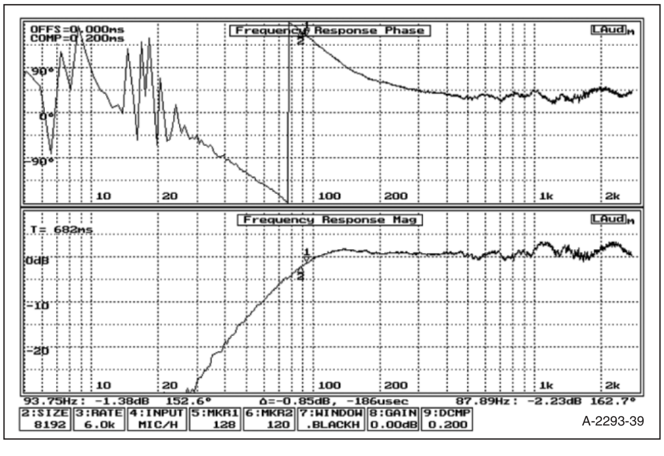

FIGURE 39: Near-field response of JX92 driver.

seem very good, and are commonly found in recording studios.[8] The reverberant field in a recording might be the sound captured at the recording venue, or artificial reverb added by the recording engineer.

How to handle the reverb is an interesting question. There seem to be several approaches. The two outer extremes could be:

- The recording has all reverb recorded and needs no sound from the room.
- The reverbant field should reproduce the recording venue.

These two cases might be handled or solved like this:

- Remove the sound of the room—either by acoustic damping, reflection handling, or by DSP techniques.
- Make a reverbant sound in the room, possibly by adding surround speakers or artificial reverb.

Removing the reverbant field of the room is difficult, since the response will differ greatly from listening positions in the room. DSP FIR filters are easy to implement but do not handle reverbant fields well. IIR filters might be better suited to this task, especially in the low-frequency region[11].

The room absorbs much of the high frequencies radiated by the loudspeaker. You can easily see this by measuring in-room response and in RTA measurements. Response in the lower part of the spectrum has similar challenges. Too much power added in the low-frequency region might cause the sound to be strained, approaching the X_{max} of the driver.

Using a target or tilting correction curve is a good way to finely adjust the loudspeaker spectral balance in the room to subjective taste[12]. This method of adjusting spectral balance is implemented in many of the commercial loudspeaker and room correction systems.

How much to add or subtract using target curves depends on several factors, such as the audibility curve. Music played loudly has more bass and highs

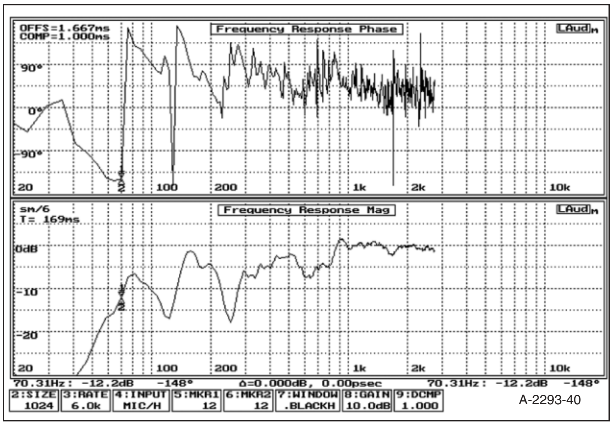

FIGURE 40: Room response of JX92 studio monitor.

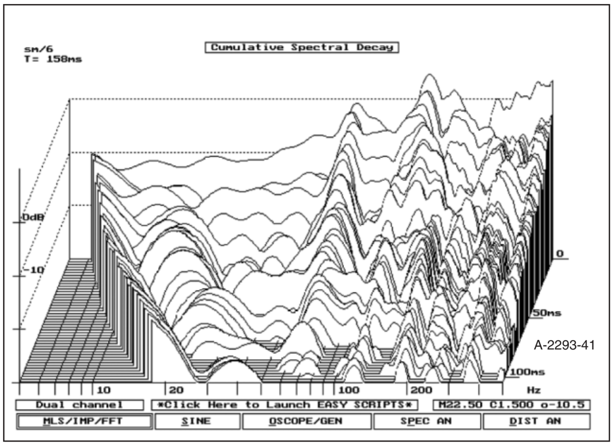

FIGURE 41: Waterfall response of room response.

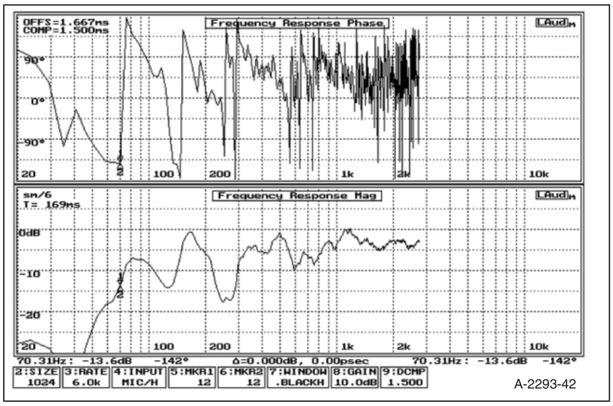

FIGURE 42: Frequency response of another room position.

in it because of this. (The loudness button was invented to compensate when listening at low levels by boosting highs and lows.)

Measuring the spectral balance is typically done using an RTA analyzer. Usually, the measurements are done averaging over ⅓ or ⅙ octave. From such measurements, you can see that loudspeaker response falls off above 4–6kHz. The rate of this falloff influences the perceived brightness of the loudspeaker.

It might be tempting to correct the response so the RTA response becomes all flat up to 20kHz. However, this does not sound good. The speaker will sound too bright. Because of this, I think it is a good idea to establish a target curve that you can set up yourself. In this way, you can adjust the brightness.

LOW FREQUENCIES

Wavelengths longer than the dimensions of the room will induce room modes. This results in amplification at certain frequencies and cancellation of others. You can see an example of this in *Fig. 40* showing the room response for one listener position of the JX92S studio monitor. Other listening positions will give other responses. Another response a meter away is shown in *Fig. 42*.

These frequencies below the Schroeder frequency, around 100–200Hz of most rooms, depend on the listener's position. Traditionally, the method of correction for the frequencies in this range is the use of absorbers, which can be Helmholtz resonators matched to the frequency to absorb and placed at mode maximas.

Peaks are more audible and offending than the dips in the frequency response. Since response differs from listener positions, the measured data should be processed in some way. It could be averaged, which can help to identify common peaks and dips at the desired listening positions. Reducing peaks that occur in several places in the room might be a good idea. Dips are more challenging to cope with. Boosting dips in the response can cause annoying peaks elsewhere in the room.

There will always be a phase rolloff in the low-frequency region. This is because the phase of moving coil drivers turns near the driver rolloff region. Also, the box causes additional phase turning. Phase rolloff adds group delay and affects the impulse response. The phase response of the JX92S driver is shown in *Fig. 39*. (I have seen some research that has pointed out that correcting the phase turning at low frequencies is worthwhile. However, I have not been able to re-find this reference.)

Correcting phase response can only be done using an active filter compensation circuit[13] or DSP techniques. FIR filters easily correct the abnormalities in the phase response of a loudspeaker while keeping the amplitude the same.

The room will always induce some room gain. For an ordinary room, this can typically be 3dB per octave below 200Hz. I would think this room gain is not a minimum phase function. In this case the phase response should not depend on this gain in amplitude.

Resolution of FIR filters is a big challenge when correcting for low frequencies. The spacing of the filter is linear with frequency, while our ear is more to a logarithmic scale. This makes it impossible to use a 48kHz sampling rate FIR filter for low-frequency correction because the computing power required is too formidable, requiring filter lengths of 10–20 thousand taps. There are several possible solutions:

- Use a warped FIR filter (does not correct phase)
- Use FIR filters with multirate sampling techniques
- Use frequency domain filtering

Frequency domain filtering is available on the Internet with the Aurora[14] software plug-in for CoolEdit. Other packages implementing a correction engine on the PC are the Linux-based NWFIR[15] and the Windows program Sinc Audio RCS[16]. Common to all this software, the response is FFTed to the frequency domain first. Then the amplitude and phase is corrected. This is an easy operation in this domain. Afterwards, the IFFT is used to transfer the corrected signal back to the time domain.

Multirate techniques use downsampling for the low-frequency filters to improve resolution. High resolution is required in the low-frequency area, which typically translates to 1–2Hz. This can be solved with an FIR filter of some hundred coefficients at a rather low sampling rate of some hundred hertz.

Combining a low-frequency corrector with a steep low-pass filter seems attractive. If I were to implement a steep loudspeaker filter at 500Hz, the minimum sampling rate for this filter according to Nyquist should be above 1kHz. (The reason that it must be higher is the requirement for a transition into the stopband of the filter.)

In the next part of this article I will explore different correction strategies for direct sound correction, diffuse sound correction, and low-frequency room correction. ❖

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PRODUCT REVIEW

Selectronic GRAND MOS Silver Design Amplifier

By Charles Hansen and John and Sandra Schubel

Selectronic
BP 513
84/86 rue de Cambrai
59022 Lille Cedex, France
Phone (33) 328.550.328
Fax (33) 328.550.329
www.selectronic.fr
Overall dimensions: 455 × 350 × 130mm (17.9″ × 13.8″ × 5.12″)
Weight: 25kg (55 lb)

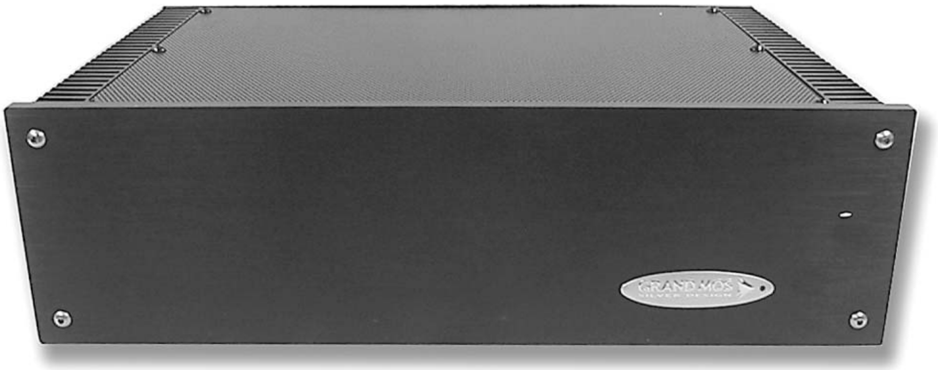

PHOTO 1: GRAND MOS stereo amplifier front view.

The Selectronic GRAND MOS is a stereo power amplifier rated at 100W per channel, and is rated for a minimum load impedance of 2Ω. A bridge-tied monophonic version is also available, rated at 400W (8Ω). Both factory assembled and kit versions of the amplifiers are available. The kit includes all components, hardware, and a pre-drilled and finished chassis.

Although the review sample of the GRAND MOS was pre-assembled, the kit assembly guide was included in the documentation. While only 11 pages long, the manual is complete and includes color photos and drawings of the entire assembly process. All you need are a few tools, and a good digital multimeter to set the bias. Temporary 1Ω fuse/shunts are included for this purpose. I recommend that you have prior kit building experience since there is none of the novice hand-holding that was the hallmark of the Heathkit assembly manuals.

Photo 1 shows the front panel of the amplifier, which has only a cool blue LED indicator inside the nameplate. The front panel is 9mm thick black anodized aluminum.

The back and bottom of the amplifier are constructed of 2mm black-painted steel. Heavy finned aluminum heatsinks are used along each side, and the 1.5mm steel top is perforated to enhance cooling. The heatsink fins have no sharp edges to cut your hands as you move the amplifier about. The unit is very stiff and rugged, even with the top removed. Two satin-finished aluminum handles are located at each side of the rear panel for ease in lifting this heavy amplifier.

The rear panel (*Photo 2*) has an IEC power receptacle with integral fuse and RF line filter. The unit is furnished with a heavy power cord. The power transformer primaries are factory-wired for 120V mains. The third pin of the AC receptacle is connected to the chassis.

Audio signals are input to two high-quality silver-plated Teflon™-insulated RCA input jacks. Two pairs of high-quality silver-plated Neutrik Speakon binding posts provide the connections for the speakers. These binding posts, in accordance with EU requirements,

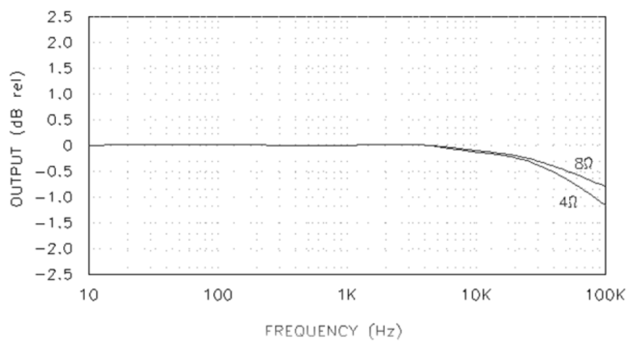

FIGURE 1: Frequency response. A-2378-1

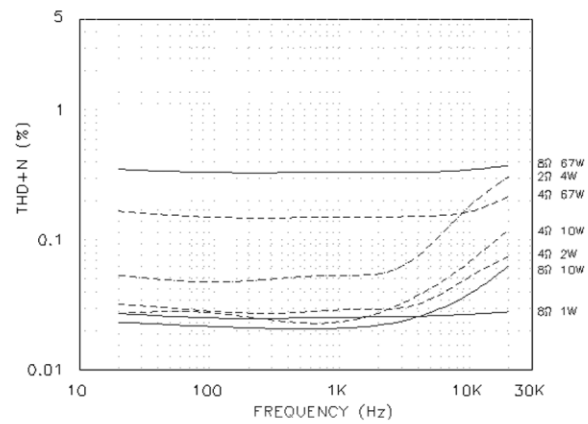

FIGURE 2: THD+N vs. frequency. A-2378-2

are not on US 0.75″ spacings, so you cannot use dual banana plugs.

Photo 3 shows the amplifier with the cover removed. The amplifier is a dual-mono design. Two large 500VA Huiran R-type power transformers[1] occupy the front of the chassis. Very little interconnecting wiring is used. The two epoxy power supply PC boards use silver-plated screw connections for the eight 10,000µF reservoir capacitors and the power transformer secondary windings. Four discrete BY239 diodes per board rectify the transformer output into ±52V DC (no load) power rails.

The two double-sided Teflon® amplifier PC boards are located along the sides of the unit. You can see the six MOSFETs for each amplifier channel attached to the heatsink with hex socket screws and keratherm insulators.

Connections between the rear panel jacks, the power supply boards, and the amplifier circuit cards are made with large-gauge Teflon®-covered solid wire and silver-plated screw connections. The factory wiring is very neat and should be easy for the constructor to duplicate in the kit version.

TOPOLOGY

A schematic was furnished as part of the GRAND MOS assembly guide. The amplifier is a true dual-mono design, sharing only the AC line input between channels. The amplifiers are electrically floated from the chassis. The input jack shell and speaker negative are connected together at each amplifier PC board, but the GRAND MOS does not share a common ground between channels as do many other stereo amplifiers.

The input stage of each amplifier comprises symmetrical single-ended JFETs cascoded with bipolar transistors. Selectronic uses matched JFETs for best performance. The collectors of the BJTs are cascaded to symmetrical

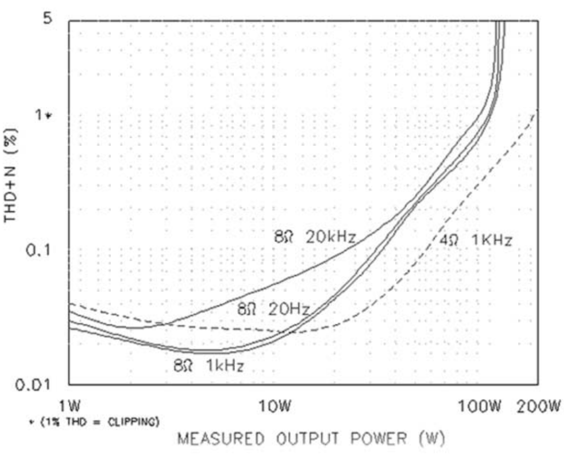

FIGURE 3: THD+N vs. output power. A-2378-3

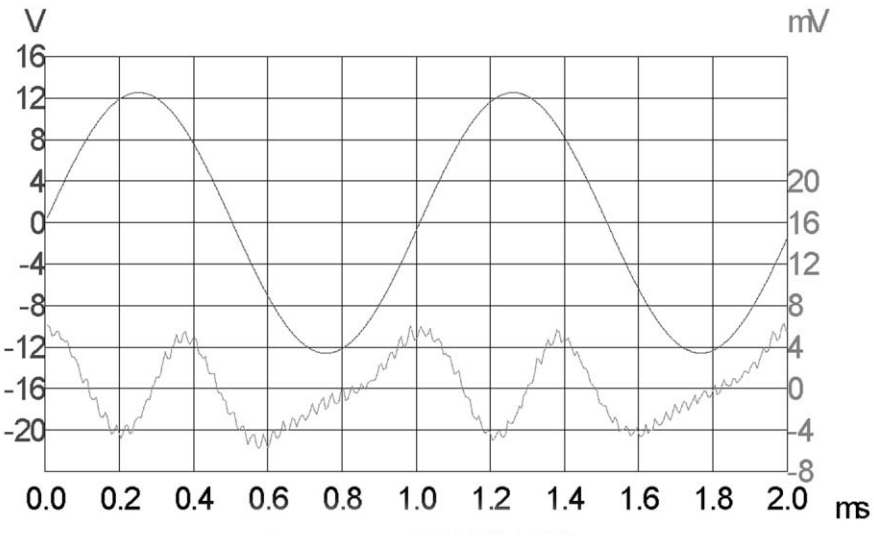

FIGURE 4: Residual distortion. A-2378-4

PHOTO 2: GRAND MOS stereo amplifier rear view.

common-source MOSFETs that provide 31dB voltage gain and drive three pairs of parallel complementary-symmetrical source follower output MOSFETs (types J162 and K1058), with low quiescent bias current. 10pF dominant pole compensation capacitors are used at the driver/amplifier MOSFETs.

Individual pots in the source of each JFET allow for adjustment of the bias. DC feedback is taken from the speaker output to the common side of the bias adjust pots at the input JFETs. There are absolutely no series capacitors in the signal path, so the GRAND MOS can amplify any DC component in the audio input signal. There is no zobel network or series R-L stability network at the output of this minimalist design.

LISTENING CRITIQUE: SELECTRONIC TRIPHON II QUAD AMPLIFIER

By John and Sandra Schubel

The Selectronic Triphon II amplifier we tested was a quad 16W class A amplifier, packaged in a black rack mountable case. The Triphon II has a companion amplifier, the GRAND MOS, which provides two 100W channels designed to power the woofers of a speaker system when controlled by a Triphon crossover network (see review in the August issue). When used with a Triphon crossover network, the Triphon II and its companion GRAND MOS amplifier can be used to drive independently the two woofers, two midrange speakers, and two tweeters of a two-channel system. Since our speakers have their own crossover networks and do not provide direct access to the tweeters and midrange speakers, we used two channels of the Triphon II amplifier to drive our two channels of speakers.

We used an HHP CDR-800 (reviewed in *Audio Electronics* 2/00) to play selections from *Hi-Fi News and Record Review*'s CD Test Disk III and directly connected it to two of four channels of the Triphon II amplifier using a passive gain control. Each amplifier channel drove a NHT SL-2 subwoofer connected to a NHT Model 1.3 via the subwoofer's internal crossover network. We used a NAD Stereo Power Amplifier 214 as a reference amplifier and a Realistic sound level meter to set listening levels, to assure that the sound levels from each amplifier were approximately the same. Our listening preference put the level at approximately 72dB, C weighted and measured from a distance of 12′. We noted that the signal level needed to drive the Triphon II was significantly greater than that required by the NAD 214.

We allowed the amplifiers to warm up for several hours before performing listening tests. We could not help, however, playing one of our favorite CDs in the meantime. We picked a CD with our favorite clarinetist, Kenny Davern, titled "Bob Wilbur - Kenny Davern - Summit Reunion," Chiaroscuro CR (D) 311. The CD has some excellent brush and cymbal work by drummer Bobby Rosengarden.

We were immediately impressed with the sound. Sandra described it as smooth. Bobby Rosengarden's cymbals were very crisp and Milt Hinton's bass was deep and solid. We've noted that on some lesser amplifiers, you must listen closely to discern which instrument is the clarinet, and which is the soprano saxophone. Not so with the Triphon II, as each instrument stood out with great clarity. It was like being at a front table in a jazz club, but without the smoke.

Clearly, we were impressed. Our only concern was that when touching the RCA input plug in the top left rear of the amplifier, we noted a buzz. We noticed this phenomenon as well when touching the red speaker terminal on that same channel. This did not occur when touching the corresponding connectors on the lower left rear of the amplifier.

We also noticed that this amplifier dissipates a lot of heat. I took a laser temperature probe and measured 133° F on the finned heatsinks that form the sides of the amplifier's case, a 61° rise over the temperature of the room measured with the same device. Even the front face of the amplifier was hot, measuring 114° and uncomfortable to touch.

With amplifiers thoroughly warmed up, we proceeded to do the formal listening test, using the aforementioned *Hi-Fi News and Stereo Record Review*'s CD Test Disk III.

JERUSALEM/PARRY

John believed that the sound from the Triphon II was much brighter and less raspy than experienced from the reference NAD. The sound is definitely very clean. The stereo separation is excellent and helps bring the soundstage forward. Sandra noted that the brass is brilliant and pleasing in sound. The highlights of the brass and cymbals sizzle. The drum sounds were much more distinct one from the other. And the sound stage is much closer perceptually.

The power supply transformers use a faraday shield between the primary and secondary windings to reduce the interwinding coupling capacitance. Four low ESR computer-type 10,000μF filter capacitors provide the energy reservoir for each channel.

Assuming the schematics are complete, I didn't see any DC servo-control or speaker protection circuitry in the GRAND MOS schematic. Since the amplifier is DC-coupled from input to output, this could place your speakers in jeopardy if a small DC offset in the preamp output were amplified by the 31dB gain of the amplifier. In case

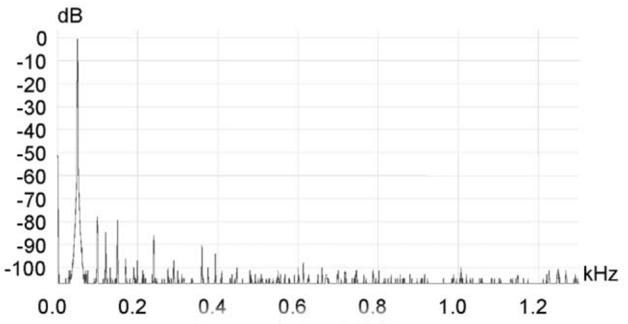

FIGURE 5: Spectrum of 50Hz sine wave. A-2378-5

HENRY V EXTRACT/DOYLE

The sounds from the Triphon II and from the reference NAD are remarkably similar on the first listening of this piece. On the Triphon II, Sandra noticed the clear, tonal sound of a block struck by the percussionist during the opening measures of the piece. This sound of the block is dull and not tonal when amplified by the NAD.

When the actor begins to recite, the sound is brighter on the Triphon II, with an emphasis on breath sounds. Sandra preferred the sound of the voice as amplified by the NAD, thinking that there was too much "s" sound on the Triphon II. She believed that she was closer to the impassioned actor with the NAD, particularly when he shouts his final command "Play!" John believed that the listener was closer to the actor on the Triphon II, because the breath sounds and "s" sounds were more pronounced.

The sound of the trumpet concerto on the Triphon was brilliant. The instrumentation is largely brass and violins, and you felt that you were in the third row. The sound was more muted with the NAD, as if you were in the middle of the hall.

We played the selection several times, and John noticed that the harpsichord was barely discernible on the Triphon II. By comparison, the harpsichord was detailed and in balance with the other instruments when played on the NAD. Sandra would rather hear this selection on the Triphon II, where pianissimo passages seemed more intense and clearer than on the NAD. Sandra did not miss hearing the harpsichord (her least favorite instrument), but not hearing it bugged John, who was undecided as to which was the better listen. The sound of the double bass was solid, and roughly the same on both amplifiers.

PETER AND THE WOLF/PROKOFIEV

Sandra and John agreed that the sound from the Triphon II is brighter, but not abrasively so. The triangles and other high percussion are crisper on this amplifier. Otherwise, the sound from the two amplifiers is roughly equivalent.

The mixture of the instruments does not become muddy on either amplifier. Double bass and drums are solid on both amplifiers. So we did not have a strong preference for either amplifier on this selection. Both were enjoyable to listen to.

At this point, we reverted back to the Bob Wilbur - Kenny Davern - Summit Reunion, and played selections "Black and Blue" and "Should I." Sandra still liked the Triphon II better, because of the intense amount of information revealed. She liked the bass, the high piano notes, and the varying qualities of the clarinet as it went from its straining high notes to its thick vel-

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of a shorted MOSFET, only the 15A slow-blow amplifier line fuse stands between the considerable output power of this amplifier and your speakers.

MEASUREMENTS

Selectronic recommends that the GRAND MOS be powered 15 minutes before listening. I initially had problems with ultrasonic oscillation in the left channel. I needed to isolate the AC line grounds at my signal sources to achieve stable operation.

I operated the GRAND MOS amplifier at 10W into 8Ω for 1 hour. The heatsink temperature increased to 48°C. The THD reading at the end of this run-in period was the same as when I began: 0.017% in the left channel and 0.021% in the right. Accordingly, the distortion measurements of the right channel are presented here. There is no noise at all when starting up or shutting down the amplifier. Output hum and noise measured 0.39mV (–84dBr, input shorted) and was inaudible with my ear against the speaker. I also measured –54mV of DC offset.

The GRAND MOS does not invert polarity. Input impedance measured 15.0k. The gain at 2.83V RMS output into 4Ω and 8Ω loads was a high 31.16dB and 31.22dB, respectively. This could help compensate for the low 15k input impedance if a preamp with a relatively high output impedance were used. The output impedance at 1kHz was 0.05Ω, increasing slightly to 0.07Ω at 20kHz.

The frequency response (*Fig. 1*) was within –1dB from DC to 116kHz, at an output of 2.83V RMS at 1kHz into 8Ω. It wasn't down to –3dB until 247kHz. There was just a bit more HF rolloff with a 4Ω load. The response with a complex load of 8Ω paralleled with a 2μF cap (a test of compatibility with electrostatic speakers) was essentially the same as the 8Ω load alone. The IHF speaker load, which has an impedance peak at 50Hz, produced no change in

cont. from 45

vet low notes. The Triphon II is a very precise amplifier.

John preferred the NAD for its balance. John noted, however, that the guitar was much clearer on the Triphon II, and the snare drum and cymbals had more sizzle as well. We had not reset the listening levels from those used with the *Hi-Fi News and Record Review*'s CD Test Disk III. This resulted in an average listening level of 80dB with peaks over 86dB—fairly loud for John's taste. This may have skewed his preference toward the NAD, which was generally not as bright in the upper frequencies.

We next played "Palasteena" on the Summit Reunion disk. The bass is clear and solid on the Triphon II, if not as resonant and pronounced as on the NAD. Piano passages (with bass and cymbal accompaniment) were very clean on the Triphon II, and made the soundstage seem very close. The piano was more distant on the NAD.

The bass was not as strong on the Triphon II but seemed to have more purity of sound and more detail. Sandra preferred the sound of the bass on the NAD, and John preferred the Triphon II, even though it did not give the perception of deepness that came from the NAD. The final tally on this piece: Sandra preferred the NAD; John still prefers the Triphon II at lower listening levels, but agreed that it sounds strident at the 80 to 90dB listening level.

We returned to the *Hi-Fi News and Record Review*'s CD Test Disk III. The fireworks on La Réjouissance/Handel produced dramatically different results on the two amplifiers. The fireworks are more clearly discernible and realistic on the Triphon II; they are colored and muted on the NAD. The difference with which the fireworks were rendered made a lasting impression.

WELCOME, WELCOME/PURCELL

Sandra found the lower of the two voices more pronounced on the Triphon II. John perceived that the Triphon II had a crisper sound, particularly the voices. The harpsichord was more distant on the Triphon II than on the NAD. Correspondingly, the voices are very immediate on the Triphon II, and more distant on the NAD. Again, the listener feels very close to the soundstage with the Triphon II, and farther back in the audience with the NAD.

MAHLER 8, GLORIA PATRI DOMINO

In the opening passages, the straining and urgency of the instruments is very present in the Triphon II. It is much less so in the NAD. The female voices are much clearer on the Triphon II, and the choir is immediately in front of the listener.

Playing this selection on the NAD, the listener is, in comparison, toward the back of the hall. The soprano voices are less detailed. The organ and orchestra are also more distant, and there is more resonance in the bass. The crescendos are more exciting on the Triphon II than on the NAD. Sandra and John both chose the Triphon II as the preferred amplifier in this bakeoff.

SONIC CHARACTERISTICS RATINGS

Presence	JS	8
	SS	8
Stereophonic Effect	JS	6
	SS	7
Soundstaging	JS	9
	SS	8
Ambience	JS	9
	SS	7

the frequency response. The GRAND MOS amplifier will be insensitive to variations in speaker impedance with frequency.

At low frequencies the noise floor limited the excellent crosstalk performance. The dual-mono design assures that minimum coupling occurs between channels (see *Table 1*).

TABLE 1
CROSSTALK

Frequency	R to L	L to R
100Hz	-86dB	-86dB
1kHz	-84dB	-84dB
10kHz	-77dB	-77dB
20kHz	-74dB	-74dB

THD+N vs. frequency is shown in *Fig. 2* for the loads indicated at the right side of the graph. During distortion testing, I engaged the test set 80kHz low-pass filter to limit the out-of-band noise.

Figure 3 shows THD+N vs. output power for the loads and frequencies shown in the graph. There was absolutely no strain right up to the point of maximum power. The amplifier, with both channels driven, reached its 1% clipping point at 117W with the 8Ω load (for 0.68dB of headroom) and 192W with the 4Ω load. The negative half-cycles clipped just slightly before the positive half-cycles. The heatsinks

CORKHILL FIVE SHORT PIECES/PRINCE CONSORT PERCUSSION

The nature of the sound is different on the two amps. The first short piece is very quiet, and Sandra believed that the unusual percussion effects of this piece were ghostly and mysterious on the Triphon II, and by comparison sounded slightly "electronic" on the NAD. The crescendos on the Triphon II were dramatic, but more muted on the NAD.

In the aftermath of the final, crashing crescendo of the second short piece, Sandra heard pleasant, faint tendrils of sound, where she found the NAD sound to lack that detail. As a whole, both amps did a beautiful job on this selection. Both had the ability in the crescendos to move the drums forward toward the listener.

CONCLUSION

So which amplifier would we own if we could have only one? Sandra and John both agreed that we would choose the Triphon II. The Triphon II, even without the aid of the GRAND MOS amplifier to drive the woofers, had sufficient power to drive our speaker system beyond comfortable listening levels. We had an unexpected reality check on our opinion when we attended a concert at Carnegie Hall the day after conducting these listening tests. The Triphon II, with its brightness, is more consistent in sound with a live performance in that hall.

What can we say about the Triphon II's companion amplifier, the GRAND MOS? We are listening to Kenny Davern on the GRAND MOS as we write, and it is a beautiful full-range amplifier in its own right. The cymbals are crisp and bright, at least as bright as on the Triphon II, and the clarinet and tenor sax are well defined.

We compared the amplifiers on several selections, and threw in some theater organ for good measure, and found the GRAND MOS to have similar characteristics to the Triphon II, with perhaps not quite the same clarity of sound but slightly brighter with similar soundstaging and ambience. It is far better than I would expect of an amplifier intended to drive woofers alone. We also noted that it did not produce very much heat. It is an excellent match and companion to the Triphon II.

John's advice to a prospective Triphon system owner is to first make sure that you have adequate air conditioning for summer use in the listening room as the Triphon II is a real heat generator. Also, although we placed it on an open surface we still experienced front panel temperatures too hot to touch comfortably (114° F) even though the room ambient was low. (We had turned off the home heating systems so that there would be no stray sounds in the room.) If you keep your audio system components in a cabinet, we strongly suggest that you find a place for this amplifier outside of the cabinet reached a maximum temperature of 57°C at 67W into 4Ω.

The distortion residual waveform for 10W into 8Ω at 1kHz is shown in *Fig. 4*. The upper waveform is the amplifier output signal, and the lower waveform is the monitor output (after the THD test set notch filter), not to scale. This distortion residual signal shows mainly the 2nd overlaid with noise. THD+N at this test point is a low 0.021%.

The spectrum of a 50Hz sine wave at 10W into 8Ω is shown in *Fig. 5*, from zero to 1.3kHz. The THD+N here measures 0.022%. The 2nd, 3rd, 4th, and 5th harmonics measure –77dB, –81dB, –101dB, and –101dB, respectively. Low-level power supply rectification artifacts are also present at 120Hz and 180Hz. The spectrum of a 1kHz sine wave (not shown) had a nearly identical distribution of harmonics.

Figure 6 shows the amplifier output spectrum reproducing a combined 19kHz + 20kHz CCIF intermodulation distortion (IMD) signal at 34Vpp into 8Ω. The 1kHz IMD product is –86dB (0.005%), and the 18kHz product is –81dB. Repeating the test with a multitone IMD signal (9kHz + 10.05kHz +20kHz, shown in *Fig. 7*) resulted in a 950Hz product of –88dB and a 1050Hz product of –83dB. This gives a better indication of the amplifier's nonlinear response, since it is a closer approximation to music than a sine wave. The GRAND MOS produced excellent IMD results.

2.5Vp-p square waves of 40Hz and 1kHz into 8Ω were almost perfect. The leading edge of the 10kHz square wave was slightly rounded, but also contained a very low level of 330kHz ringing. When I connected 2μF in parallel with the 8Ω load, there was significant ringing at 70kHz (*Fig. 8*). The spikey appearance of the sinusoidal ringing is due to the limited high-frequency response of my DSO. A comparison of the measured results and the manufacturer's ratings is shown in *Table 2*. ❖

TABLE 2
MANUFACTURER'S RATINGS AND MEASURED RESULTS

Parameter	Manufacturer's Rating	Measured Results
Power Output:	100W RMS (sic) 8Ω	117W 8Ω, 192W 4Ω, 1% THD
Frequency Response, 10W, 8Ω:	DC – 200kHz power bandwidth	DC – 247kHz ±3dB
Total Harmonic Distortion:	<0.05%, 10W	0.021%, 10W 8Ω
IMD –CCIF (19+20kHz):	N/S	0.005% CCIF
MIM (9+10.05+20kHz):		0.004% MIM
Input impedance:	15kΩ (775mV)	15.0k
Signal to Noise Ratio:	N/S	–84dB
Gain:	775mV for 100W RMS (sic)	31.2dB
Output Impedance:	N/S	0.05 Ω 1kHz

REFERENCE

1. The oval laminated core of an R-type transformer is continuous like a toroid, but has a circular cross-section made up of varying widths of grain-oriented silicon steel laminations. This gives the minimum iron weight for a given core area. The winding coil forms are also circular in cross-section. This allows a shorter wire length per turn and less overall coil resistance.

PHOTO 3: GRAND MOS stereo amplifier interior view.

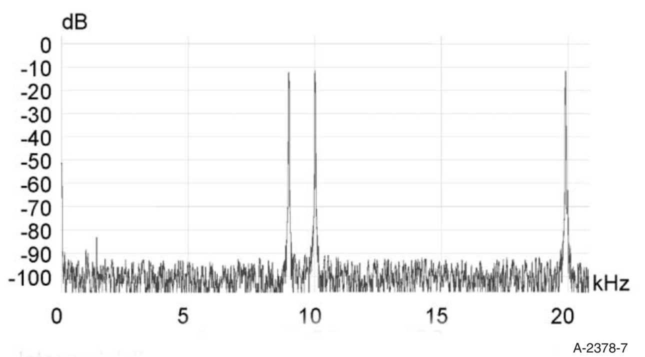

FIGURE 7: Spectrum of 9kHz + 10.05kHz + 20kHz intermodulation signal.

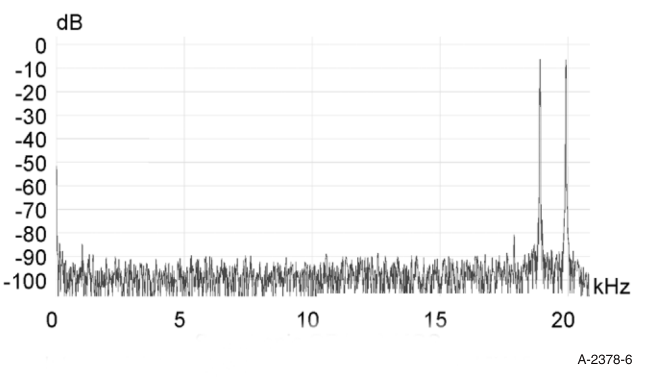

FIGURE 6: Spectrum of 19kHz + 20kHz intermodulation signal.

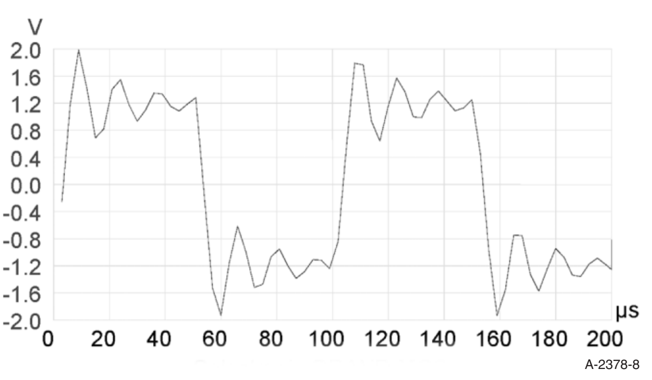

FIGURE 8: 10kHz square wave into 8Ω in parallel with 2μF.

New Chips on the Block

The Silonex Audiohm Photoresistor/Coupler

By Charles Hansen

Back in 1997 I received a sample of a new optocoupler from Silonex called the Audiohm. Resistive optocouplers use an LED whose light output is optically coupled to a photoresistor. In the past I have used the Clairex CLM6000 in some guitar effects projects. This device used cadmium sulfide (CdS) as the photoresistive material (see sidebar entitled Photosensitive Materials). The Audiohm uses a proprietary mixture of CdS and cadmium selenide (CdSe), and is reported to have lower noise and distortion than either a purely CdS or CdSe device alone.

Silonex has listed a number of advantages for use of the Audiohm as a control element in analog audio circuits. These include:

- A fully isolated controlled variable resistor
- Relatively low drive current
- Good distortion performance
- Zero charge injection (no clicks)
- Inherent time constant that offers smooth on/off switching

Until recently, there was no application information for the devices. There is now a study that looks at the properties of these devices from the perspective of an audio engineer. The website also offers a number of interesting application notes and circuits:

- Compressor
- Limiter
- Soft Switching
- Level Controls and Cross-Faders

The site says additional applications are coming soon: VCA replacement, expander, and noise gate. I also found a Leslie Effects Rotor Adapter on the Internet (search on LERA).

ANALOG OPTICAL SOUND

Modern optocouplers are not the first use of optical audio processing. The most common method once used an optical process whereby a transparent line was recorded along one side of movie film. The strip varied in width according to the frequency of the sound, and was known as a variable-area soundtrack.

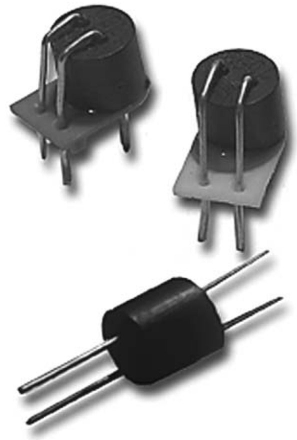

PHOTO 1: Audiohm available packages.

As the film passed by an audio pickup, bright light from an exciter lamp was focused through the transparent line by a lens. The light that passed through the film shone on a photocell. The light was changed to electrical current by the photocell. Wider parts of the strip allowed more light, which caused the photocell to produce more current.

A variation of this method was known as variable-density soundtrack. It used a strip that varied in transparency rather than width. The more transparent the strip, the more light that shone through.

The biggest problem with these methods was that the natural graininess of the film and any scratches from handling produced a lot of background noise. In the 1950s magnetic recording became popular and eventually supplanted optical soundtracks. Photoelectric sound films were decades ahead of the optical analog LaserDisk first demonstrated by Philips and MCA in 1972.

Phototubes had a large-area light-sensitive cesium (or other alkaline metal) cathode surface and a collecting anode contained in a glass vacuum tube bulb. When light impinged on the cathode it emitted photoelectrons in proportion to the intensity of the incident light, which were then focused on the plate. The early image orthicon television cameras used a variation of this device called a photomultiplier tube.

AUDIOHM ANALOG MODE

In analog mode, the Audiohm optocoupler performs the function of a current-controlled variable resistor. An LED emits light that is directly proportional to the LED control current. The CdS/CdSe photoresistor changes its resistance in proportion to the light level. When used in analog mode the photoresistor reportedly has considerably less distortion than an FET and is more cost effective than a voltage-controlled amplifier (VCA) or digital potentiometer.

The Audiohm product line consists of 27 devices that are available in either platform or axial packages, both for through-hole installation (*Photo 1*). Sorted devices are available for applications where tracking is important, such as multi-channel volume control.

Silonex employs highly developed pro-

PHOTOSENSITIVE MATERIALS

The liberation of electrons from materials when exposed to light is known as the photoelectric effect. Heinrich Hertz recorded the first experimental evidence of the photoelectric effect in 1887 when he was investigating the production and reception of electromagnetic (EM) waves.

The first experiments with the photoelectric effect were performed on contact rectification materials such as copper oxide and selenium (the first semiconductor rectifiers were made from these compounds). Later it was noticed that if light fell on a semiconductor junction, its conductivity increased. This was first noted in the early 1950s when experimentation with diode and transistor applications became widespread. Germanium diodes were about 200 times more sensitive to light than selenium cells. Analog designers found that if light fell onto the glass cases of their circuit diodes, the analog performance was thrown off.

The main types of photosensitive semiconductor devices are:

1. Photoconductive cells (light-dependent resistors) consist of a semiconductor substance (cadmium sulfide, cadmium selenide, cadmium telluride, lead sulfide, and so on) whose electrical resistance varies with the intensity of illumination falling on the cell. Connections to the cell are by means of two lead electrodes.

Photoconductivity occurs in semiconductor materials due to the higher electrical conductivity that results from increases in the number of free carriers that are generated when photons are absorbed. Many of these semiconductors are compounds of the II-VI element groups in the periodic table.

In darkness, the photoconductor resistance is very high. If a voltage is applied to the cell material, only a small dark current will flow. This is the characteristic thermal equilibrium current. When light shines on the photoconductor and is absorbed by the crystal, its electrons are excited, decreasing the cell resistance.

For cadmium sulfide (CdS) material, the light must have a wavelength shorter than 515 nm. At shorter wavelengths, the light is absorbed near the surface of the crystal, inducing electron-hole recombination. If a CdS cell is left in total darkness for more than 10 hours, the value of its resistance is maximized. This is the true, or equilibrium, dark resistance. In practical applications where the CdS cell is used at various light levels, the previous light levels affect the dark resistance. This is called the light history effect.

CdS cells require a certain amount of time to respond to incident light. This limits the ability of the cell to detect rapidly changing light levels. The response rise time is expressed as the time required for the resistance to reach 63% of its saturation value after illumination. The decay time is the time required for the resistance to decay to 37% of its saturation value after the light is removed.

An amorphous selenium alloy is used on the drums of laser printers and copiers. CdS cells are used in flame detectors, in the exposure meters for automatic cameras, for counting moving objects, and in automatic door-opening systems.

2. Photovoltaic cells convert light directly into electrical energy without the need for an external source of current. Silicon solar cells convert sunlight directly into electric energy. They are usually connected in panels as a source of electric power in satellites, and may someday provide electricity for private homes.
3. Optocouplers (opto-diodes, transistors, and thyristors) are used to provide galvanic isolation between circuits to prevent shock hazards and to transfer digital data.

cessing of the CdS and CdSe photoconductive layer. They claim this guarantees the lowest possible ohmic contact and reduces parasitic capacitance, resulting in devices with ultra low distortion.

MEASUREMENTS

I connected one of the Audiohm NSL-32R3S devices and an old Clairex CLM6000 CdS cell, with their LEDs in series with a DC current source. The resistance change with LED control current is shown in *Fig. 1*. The change is very logarithmic in nature, much more so than a log (audio) taper potentiometer. The CdS cell has a different slope than the Audiohm.

Next I took four samples of the sorted NSL32 devices and ran the same test (*Fig. 2*). I also included the sample device from 1997. The four sorted devices track within about 5.5dB at the highest resistance, and improve to less than

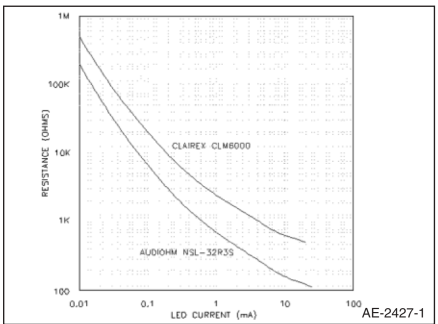

FIGURE 1: Resistance versus LED current.

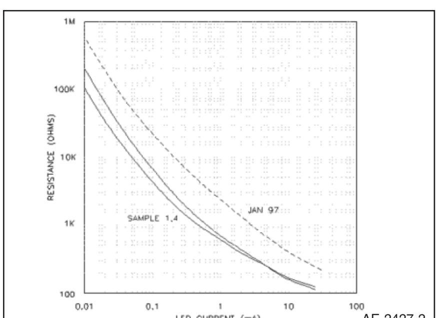

FIGURE 2: Resistance tracking versus LED current.

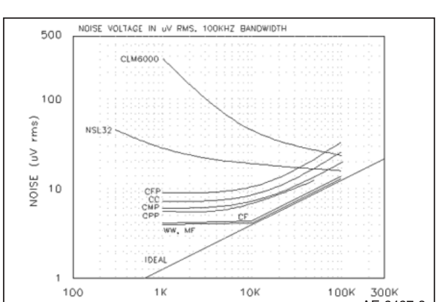

FIGURE 3: Noise versus resistance.

1dB below 1k resistance. Note that the curves cross each other near 250Ω. The earlier device has a similar slope, but does not track the four sorted samples.

Since the optocoupler is a current injection semiconductor device, it has higher noise than that of a resistor. I connected a number of fixed and variable resistors across a 6V DC battery and measured the AC noise over a bandwidth of 100kHz. I measured the potentiometer noise from their wipers to one of the end terminals.

The graph in *Fig. 3* shows the Johnson noise of the resistors (plus any noise from the battery) and the noise generated by the Audiohm and Clairex optoisolators. Note the reversed slope of the optoisolator curves. The discontinuity below 4μV is the noise floor of my test setup.

The curves are compared with theoretical "ideal" resistance that has no excess noise. From bottom to top at 10k resistance, the tested devices are:

- Ideal fixed resistor (theoretical)
- Wirewound and metal film fixed resistors (WW, MF)
- Carbon film fixed resistor (CF)
- Conductive plastic pot (CPP)
- Cermet pot (CMP)
- Carbon composition fixed resistor (CC)
- Carbon film pot (CFP)
- Audiohm NSL32
- Clairex CLM6000

Figure 4 shows the shunt attenuator application circuit from the Silonex web site. The audio input signal is applied to a series 10k limiting resistor. The Audiohm is connected at the audio output side to ground. LED current from some control interface (current source, digital-analog converter, and so on) varies the resistance of the photoresistive element.

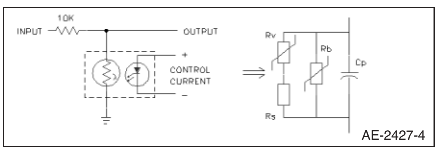

FIGURE 4: Shunt attenuator application and photoresistor equivalent circuit.

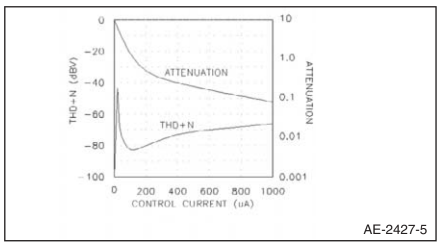

FIGURE 5: Attenuation and THD+N versus control current.

The equivalent circuit of the photoresistor is shown on the right side of *Fig. 4*. Rv is considered to be an "ideal" resistance that varies with the LED light intensity. Cp is the parasitic capacitance of a few pF. Rb is the bulk dark resistance of the cell layer, and Rs is the minimum resistance of the cell layer at saturation light intensity. Rb and Rv show a marked negative voltage dependency, such that resistance decreases with increasing voltage across the photocell.

Figure 5 shows the graph from the Silonex data for the shunt attenuator circuit in *Fig. 4*. When LED control current is first applied, the cell resistance is so high that the nonlinear response has no effect on distortion. As attenuation starts to increase, there is a peak in the distortion due to resistance nonlinearity and noise.

As the cell resistance drops further, less of the input voltage appears across the cell. Noise is inversely proportional to cell resistance, and cell nonlinearity is proportional to the cell voltage, so the distortion drops rapidly. From there the distortion rises gradually until it reaches the second (but much lower) peak, where cell resistance is at or below the series resistance.

There are other graphs on the web site for series attenuation (the worst distortion case) and different configurations of series/shunt attenuation. Silonex also compares the various methods of audio switching in terms of voltage, isolation, drive requirements, and on/off resistance.

The optocoupler cell layer has an inherent response time delay. The rise time, T_R, is 5ms to 63% of final conductance at I_F = 5mA. The decay time to 100kΩ after removal of I_F is 10ms. The NSL32SR2 has a much slower decay time of 500ms. This may be advantageous, depending on the end application. ❖

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Book Review

Sound Recording Advice

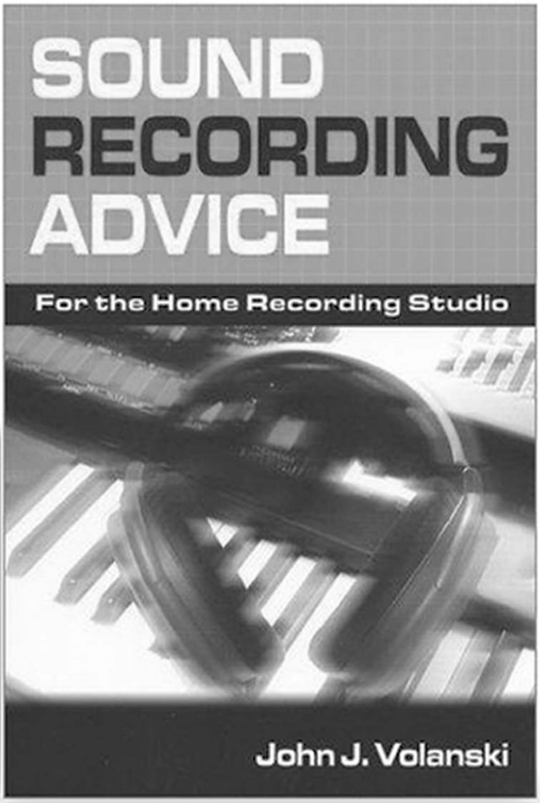
Reviewed by Richard A. Honeycutt

Sound Recording Advice *by John J. Volanski. Pacific Beach Publishing, PO Box 90471-A001, San Diego, CA 92169. http://www.jvolanski.com. $19.95.*

John Volanski, like most of us interested in audio recording, entered through the "hobby" door as a youngster. When his father brought home a reel-to-reel recorder with sound-on-sound capabilities in the 1960s, Volanski was hooked. (For those readers unacquainted with reel-to-reel terminology, sound-on-sound was a method of building up multiple separately-recorded tracks onto a single 2-track recording. Unfortunately, it also built up noise, and at its best, required prodigious ability of the recordist to achieve good quality, since the balance among tracks could not be altered once the track had been recorded. I do not lament its replacement by true multitrack!)

Volanski's career has been in engineering design and management in avionics simulation and virtual reality. His articles have been published by a number of internationally-circulated magazines. He has owned a home studio for over 20 years. Thus, in many ways, he is highly qualified to write an introductory book on audio recording.

The book commences with a definition of general terms, such as "talent," "medium," "flanging," and so on, that have particular meanings in the world of audio recording. Next, the various recording formats are described along with their strengths and weaknesses. Then studio AC power systems, mike preamps, MIDI, and studio interconnections are discussed. The "equipment" section of the book concludes with sage advice and recommendations for the choice and purchase of new or used equipment. This section contains a great deal of information presented in a very practical, down-to-earth style, much in the fashion of an old pro mentoring a newbie.



Section two addresses studio layout and furniture, and begins with six pages on studio acoustics. Equipment location, noise avoidance and removal, and test equipment are surveyed. The third section gives examples of home-brew equipment modification projects, including a list of parts suppliers. Section four discusses actual recording techniques, including the physics of sound, mike placement, bouncing tracks, mixing, and mastering. The book concludes with a long collection of tips and techniques.

The author's presentation is straightforward, fair, and to the point. His selection and organization of topics are well thought-out. And his long experience shows through. I would recommend this book to anyone who is fairly new to recording, or even to those who have been at it for a while, but need to fill in gaps in their knowledge.

That said, I must mention some shortcomings, all of a nature that could have been eliminated by good technical editing. [Warning: Rant coming up.] Unfortunately, in an effort to cut costs, few books other than textbooks receive any technical editing these days. In fact, it is obvious that many books do not even receive grammatical editing. This fact naturally increases errors in publications, since not even we authors are perfect.

[Warning: Shameless self-promotion coming up.] The notable exception to this rule is trade and specialty magazines such as *audioXpress* and *Voice Coil*. In these publications, at least one technically qualified person reads each article as part of the editing process, and thus far fewer technical slips occur than in the general press.

Here are some examples:

- Reel-to-reel tapes are said to be available in 2″ thickness. These would be a real (reel?) problem! What is meant is 2″ width.
- In the discussion of DAT, oxide shedding is incorrectly referred to as "oxidizing." The particles on a recording tape are already oxidized.
- After correctly warning of the unsubstantiated hype that is often associated with vacuum tube electronics[1], the author mentions, with a seemingly straight face, cables "designed to pass audio and video in one direction only." But both audio and video electrical signals are alternating current, which must therefore go both ways. I really don't want to hear the distortion that would result from a one-way cable (also known as a diode).
- Proximity effect is incorrectly attributed to dynamic mikes. In fact, proximity effect is present in any directional mike, whether dynamic, ribbon, capacitor, piezo, or whatever.
- Ribbon mikes are incorrectly described as "another name of figure 8." Ribbon mikes use a ribbon as the diaphragm, rather than a dome attached to a moving coil. By nature they have a figure 8 response, but cardioid ribbon mikes are available, as are figure 8 (more properly called "bidirectional") mikes using any of the other generating mechanisms.
- Connectors are not adequately discussed (e.g., pin 1 ground on XLRs), yet the reader is advised to disconnect the ground from one connector to eliminate a "ground loop." Also, the need to maintain the ground connection for phantom-powered mikes is not mentioned.
- The discussion of room modes is oversimplified, since it is one-dimensional. Actual room modes involve 1-D, 2-D, and 3-D sequences of frequencies. Further, the idea of eliminating the effect of room modes using a parametric equalizer is naïve, since the effect upon frequency response of any mode is spatially dependent.
- To feed from a balanced source to an unbalanced load, the reader is advised to connect pins 2 and 3 of an XLR connector to hot, and pin 1 to ground. This will short out the balanced signal. The correct method of connection actually depends upon the output configuration of the balanced source: either pin 2 to hot and pin 3 to ground, or pin 2 to hot and pin 1 to ground.
- The inverse square law is mentioned, but the reader is not told that it applies only to near-anechoic environments (such as outside).

Besides these, which I believe are most of the actual errors, there are four points on which I wish information had been included.

The first concerns the use of noise gates. If the threshold is mis-set, they introduce bothersome artifacts at the onset and ending of a sound. And even if the threshold is set correctly, if the input signal is very noisy, the ear/brain interprets the noise as distortion after gating, since the noise is not always present, and therefore is not ignored by the ear/brain.

Second, although the electrical noise generated by CRT monitors is discussed, the best solution—using an LCD monitor—is not suggested.

Third, the reader should be warned about the tendency of many blank CDs to produce errors (showing up as audio pops and clicks) near the end of a long recording. I need to be very particular with blank CDs I use, because the concert recordings I often make push close to the 73-minute limit of a standard CD. Auditioning a completed CD is important if you are to avoid unpleasant surprises later.

And fourth, the section on equipment modifications assumes a familiarity with schematics and basic construction techniques. A reference to a book such as Charles Hansen's would be helpful (*The Joy of Audio Electronics,* available from Old Colony Sound Lab, 888-924-9465, www.audioXpress. com).

If you have read this far, I hope you have noticed that my complaints are about small things, not major issues. I included them primarily so that the novice reader of Volanski's book will not be misled by these small errors. On the whole, the book is very well-done, and fills a need that has long existed for a good introduction to audio recording. And if everyone who needs such a book buys a copy, then the publisher will contract for a second edition, which will be even better! Let's look forward to it! ❖

REFERENCE

1. Unfortunately, the entire audio field is rife with unsubstantiated hype. Since we certainly cannot perfectly correlate measurements with hearing, each of us must exercise "sound" judgment, combined with careful listening, in choosing equipment. In general, if it sounds better to us, then it is better—for us, at this particular stage of our ear training. But remember that in the 1950s a whole roomful of table radios tuned to the same AM station sounded wonderful to most non-critical listeners. In the 1960s a speaker system design was published using the "whole bunch of cheap speakers sounds really good" principle. For good reason, you will not find this speaker system used as a studio monitor today! The bottom line is that we must use our brains, our ears, and the advice of others—in that order. And remember that both brains and ears can be fooled by our expectations: placebo effect is real, even for those of us who are not aware of it.

Xpress Mail

PROJECT HELP

Dick Crawford's article, "A 1PPM Distortion Analyzer" (*aX* March '04, p. 8), was a challenging construction project, noted to be so by the author. Reader Frank Glabach contacted the author for help and has recorded his adventure for other constructors. The extensive list is posted on our website for reference. Those without access to the web may send a #10, stamped envelope to "Distortion Analyzer" at the magazine's address.—ETD

AMP VARIATIONS

The article by Joseph Norwood Still in the June 2004 issue of *audioXpress* ("35W Triode and 60W Ultralinear Control Amp," p. 26) is the latest in a long series of construction articles by Mr. Still that have appeared in *Glass Audio* and *audioXpress* over the last six years. The first article by Mr. Still appeared in 2/98 *Glass Audio* and was a revised version of an article originally published in 1959.

Since the initial article, additional articles by Mr. Still have covered numerous variants of this same basic amplifier incorporating alternative tube lineups and offering different control functions. Why do you continue to publish essentially the same article by Mr. Still over and over and over again? Please use the limited editorial space for material that is new and stimulating for audio hobbyists.

Dean Hiebert
dhieber@mchsi.com

Articles by Mr. Still are a good example of variations on a theme. His articles have been among the most popular with more evidence of his projects being built by readers than almost any other.—E.T.D.

CROSSOVER DIP

James Moriyasu's test of the Usher speaker (*audioXpress* Feb. 2004) is the best I have ever read. Even in discussing the knee of the transfer function, how much more thorough can you get? He also uniquely showed the amplitude dip at the crossover point when the polarity of one of the drivers is reversed—in my opinion very important.

In my experience (I've been building speakers for more than 30 years) getting a dip as deep as possible at the crossover point when the polarity of one of the drivers is reversed is absolutely necessary, even if this causes the frequency plot to be a little less flat. When you design a speaker system, do you try to achieve this, and do you have any tips on how you do this?

Ernie Brunings
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James Moriyasu responds:

I agree that getting a dip with reversed driver polarity is important. In general, I try for a dip of at least 12–15dB. Dips of 5dB or lower are less satisfactory, and I'll press on with a different design until realizing a dip of 12–15dB. I've never performed any subjective testing as to whether a crossover with a shallow dip sounds as good as one with a deep dip. However, Vance Dickason—if my memory of our conversation is accurate—recommends a minimum dip of 12–15dB, so that has been my standard.

As for tips, generally, a more complex crossover, which means more parts, seems to model the dip better. I think this is because it allows closer tracking of the target crossover slope.

For example, you can achieve a typical two-way driver crossover with a 4th-order Linkwitz-Riley characteristic with 2nd-order electrical circuits. This is possible because the driver rolloffs contribute to the overall crossover slopes. You can sometimes achieve a little better crossover with third-order on the highpass and/or lowpass. So, the higher-order electrical circuits tend to produce deeper dips. However, I usually go with the crossover with the fewest parts, as long as the dip is deep enough.

REVIEWS

I must confess that on reading the review of the DacT CT 100 preamp (June 2003, p. 58), my feelings were a mixture of bemusement and déjà vu. Here we have three separate assessments of the same product, none of which agree with one another in major respects. What does that tell me, using simple logic? That all three are wrong, and this is a basic premise from which it is useful to start examining this disastrous assessment.

It's a small wonder the manufacturer was clearly upset! It took me back to the days—all of three decades ago—when in addition to my design work I also undertook reviews of audio products. So let me make clear where these opinions are coming from—a good half a century of experience.

Conscious that it will cause great controversy, I must dismiss out of hand the contributions from John and Sandra Schubel. Theirs is a collection of entirely subjective opinions, using the most unreliable of test instruments—the ear and brain. To translate those opinions means resorting to subjective language that can only mean something to the expressor and usually very little to anyone else. Subjective assessments of this case are perfectly valid when listening and evaluating musical instruments and, to a lesser extent (although even this is open to debate), all audio transducers such as loudspeakers.

Even so, it is still a matter of opinion and very remote from an objective evaluation. Long ago—or at least I thought so—a plethora of carefully thought-out test procedures existed to check the performance of any part of the active chain, e.g., preamps and power amplifiers, that would reveal any artifacts liable to corrupt the signal.

The late Peter Walker famously quipped that, "all an amplifier is required to do, is make the signal bigger." Clearly, the methodology used by both the makers of the product and the reviewer demand examination, and, frankly, my suspicions are aroused when some degree of importance is implied to a difference of 0.004% in THD between channels by a reviewer. Why deface the ICs when it is certain they are off-the-shelf devices?

However, I would like to be constructive, particularly since at least three decades ago, disparity between any objective measurements and the aural result was causing some concern, not only to me but to fellow reviewers. In the case of phono input stages, the penny finally dropped.

It was the belated realization that the signal-source, i.e., the phono cartridge, was not the obliging constant-voltage source that we were all using to check the stage, but from a complex mixture of reactances as well. Moreover, it was also in the inverse feedback path of the input stage. It then seemed logical to include in the test procedure a typical MM cartridge, and inject the test signal in via a low-value resistor, e.g., 0.5Ω in the ground lead of the cartridge. Using pure tones, sometimes actual program signal and even square waves revealed artifacts hitherto undetected.

The phono stage then performed exactly as it would with a typical MM cartridge, divorced from the artifacts contributed by the mechanical parts of the cartridge, i.e., stylus and tracking problems. My fellow workers also tried it, confirming my discovery, and it became standard practice. Interestingly, there was little difference with MC cartridges whether feeding the pre-equalized signal in direct or via the injection technique. This may explain why they enjoy greater popularity and consistency in performance.

Reg Williamson
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Publisher Ed Dell responds:

I must take the strongest objection to my friend's (to me absurd) suggestion that subjective opinions are irrelevant in evaluating audio equipment. I do not believe people go into an audio store with a sheaf of measurements in their hands and then decide against listening to whatever they decide to buy. It may have pleased the late Peter Walker to say that neither he nor his staff needed to listen to equipment they manufactured before shipping it, but most ordinary mortals require some assurance that only their ears can supply.

John and Sandra Schubel are experienced listeners, and they report what they heard. No one, including our friend Reggie Williamson, can tell them they did not hear what they heard. Both they and our measurement engineer are well qualified for what they do. John's years of experience at Bell Labs do not exactly disqualify him as a reliable listener. Chuck Hansen's many years of engineering experience at AlliedSignal and other companies has made him the most careful and scientific expert I know of anywhere.

If listening evaluation of audio equipment, relating to how it sounds, is irrelevant, what is the point of quality reproduction? When ab-

solute scientific measurement principles overtake common sense, we get a serious distortion of the whole question of quality sound.

Over the 35 years of publishing periodicals about quality sound reproduction, we have tried to give balanced consideration to both ways of evaluating what quality sound is like. The same old subjective/objective, science/intuition battle continues here, as it does in almost every human discipline. It is not now resolved, and, given the human condition, it never will be.

Our purpose in reviewing equipment is to give readers some idea of what the equipment is like and how it performs. We could, of course, serve our own commercial purposes by suppressing less-than-positive reviews, as has been done by many other audio periodicals. This avoids alienating advertisers. But if we provide reliable tests and experienced listeners, then we owe it to readers, and ultimately to advertisers as well, to let the process proceed and publish what we find. To do otherwise undermines the integrity of the publication, and eventually readers become disillusioned and believe nothing of what they read.

As I read other audio periodicals both from the US and from overseas, I simply regard them as essays about how much fun it is to use and listen to the equipment, in many cases with no measurements of any kind. The reviewers seem to be performing each month, with absolutely no comparison to what they said the previous month, nor what they will say to applaud what they will be using in the month to come. This is entertainment, as has happened to so much these days in the news media and politics. But even this disservice to readers does not invalidate the idea of a subjective opinion reported by an experienced listener. Bad use, as I learned in formal logic, is no argument.

I hope readers who are interested in the product we reported on will try it for themselves. I have been impressed with the very high quality and ingenious design of DacT products since they appeared years ago. I believe they are serious about audio and attempt to offer outstanding value and functionality.

At least your response to this review effort belies one of the claims of now extinct audio magazines: "No one wants to read bad reviews."—ETD

The manufacturer's response by Allan Isaksen to your DACT CT 100 RIAA preamp review (aX 5/04) raised some interesting questions that were not completely addressed in Charles Hansen's response. I was surprised by the differences Mr. Isaksen found between Mr. Hansen's test measurements and DACT's on the same piece of equipment.

1. Mr. Isaksen believes John and Sandra Schubel's passive preamp volume control might be deficient. What type of passive preamp did they use, and was this same passive preamp used with all the available phono preamps and turntables in the course of their listening test?
2. Has Mr. Hansen measured the deviation errors in his inverse RIAA testing network? Is there a reason for not using the comparison to the RIAA curve suggested by Mr. Isaksen?
3. I find it hard to believe a preamp with low-power buffer transistors in the output stage can actually have an output impedance of 0.1Ω, thus rivaling a number of power amplifiers tested previously in your pages, as DACT claims. But it would be interesting to have Mr. Hansen describe the method by which he determined it to be 30Ω.
4. In the discussion of distortion in Fig. 5, Mr. Hansen presented THD+N, while Mr. Isaksen refers to THD (presumably without the +N factor). Mr. Hansen does say the THD alone is 0.0022%, while THD+N is 0.029%. It seems noise (+N) might play a very important part in an RIAA preamp due to the low level of the phono cartridge signals. I would think there is a bottom limit on noise performance in moving-magnet RIAA stages due to the high 47k input load resistor.

Cal Jonstone
cjonstone@yahoo.com

Charles Hansen responds:

I think there was probably more discussion on the DacT CT-100/CT-102 test reviews than all the other tests I've run for aX *combined.*

1. *The "passive preamp" (this seems to be the accepted terminology for a volume control or stepped attenuator with input switching) in this case was an Alps "Black Beauty" stereo control with a value of 50kΩ. This same control was used for all three phono preamps and both turntables used during the audition.*

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Classifieds

VENDORS

JENA LABS
Our websites are our information centers for interconnects and cables, AC power solutions, DIY and OEMparts, Immersion Cryo services, and dealer list. **jenalabs.com jenatek.com**

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WANTED

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Xpress Mail

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2. A very important part of the testing process is periodic calibration of all test equipment. This prevents errors due to the invariably slow drift of the measurement equipment over time, and is only fair to the manufacturer who has entrusted their product to the test regime.

I perform my calibrations on a yearly basis with a suite of Fluke calibration equipment whose own calibrations are traceable to NIST. Jim Hagerman made the inverse RIAA network I currently use, and I carefully made a table of its deviations from the published RIAA equalization. It is flat over the audio frequency range to ±0.1dB. The deviation errors are –0.1dB from 20 to 50Hz, +0.1dB centered at 500Hz, –0.1dB from 5kHz to 10kHz, and +0.06dB at 20kHz. I always correct my frequency response curves for these minor deviations.

The inverse RIAA network is a time-saving tool that allows rapid measurement of the phono preamp frequency response by allowing for a fixed input signal at the input of the network, in this case 0.995V RMS input for 10mV at 1kHz into the phono preamp MM input with a 47kΩ cartridge load. All other input levels are then automatically adjusted for frequency in accordance with the RIAA equalization standard. If the phono preamp equalization has the same accuracy as the inverse network, the results should cancel and produce a straight line response at the phono preamp output. The passive inverse RIAA network is sensitive to both signal generator source impedance and phono stage input resistance.

3. My method for measuring output impedance for line-level signals is to apply an accurate resistive decade box (another form of stepped attenuator) to the output connector, and adjust it until the measured voltage is precisely one-half that of the open-circuit voltage. The output impedance is then equal to the decade resistance. In the case of the DacT CT100, because of its current-limiting power supply, I made sure the output impedance was stable beyond the half-voltage point (it was). This is performed at 20Hz, 1kHz, and 20kHz.

Things are not as straightforward for power amplifier measurements. This involves a calculation based on the measured difference between the output terminal voltages at 4Ω and 8Ω load. For solid-state amplifiers I add an open circuit voltage measurement as well to check for output impedance linearity with load.

4. The +N portion of THD+N is indeed an important factor where low-level input signals are involved. This is why many control preamps that use the same active devices in their phono and line stages may have a level of THD+N that is an order of magnitude higher than the lower gain line stages. You can also see this in the measurements of THD+N vs. input or output level. The distortion curve initially slopes downward as the signal level increases and overcomes the noise component, then shifts abruptly upward (or more gradually in the case of tube equipment) as the clipping level is reached.

While a 47k resistor at 25C has a noise voltage of about 3.9µV over the audio bandwidth (27.8nV/root–Hz), it is shunted by the RL impedance of the phono cartridge (I use 1k3 and 500mH). At high frequencies the rising impedance of the cartridge (62k at 20kHz for the above values) is compensated for by the input loading shunt capacitor, if any. ❖

The MCUTracer, Part 2

Now you can automatically determine the voltage necessary to test tube characteristics with this DAC/voltage amp controlled by the Basic Stamp, using a StampPlot graphical user interface. **By Jack Walton and Martin Hebel**

To get the most out of the MCUTracer, you need an adjustable power supply capable of B+ and C– voltages of +400 and 50. The digitally controlled power supply described here uses off-the-shelf parts, including high-voltage MOSFETs as the "pass" element for the B+ supply. For the example shown here, we used a Heathkit IP-17 high-voltage regulated power supply as the carcass—employing the chassis, meters, and transformers, but discarding all of the control circuitry. The design of the power supply is such that you can use it for the basis of upgrading the IP-17 without the microprocessor if you wish.

Co-author Martin Hebel wrote additional code (The codes and instructions for how to use them for parts 1 and 2 are both located at www.audioXpress.com. Once there, click on the current issue, and follow the link title "articles and addenda.") which employs StampPlot, dedicated Supervisory Control And Data Acquisition Software (SCADA) for the BASIC Stamp. The new code for the Basic Stamp II allows you to control the microprocessor from your PC, specifying the range of plate and grid voltages and maximum current for the device under test. You can download an evaluation version of StampPlot from www.stampplot.com.

THE POWER SUPPLY

The power supply consists of three modules: a digital-to-analog controller connected to the Basic Stamp in the

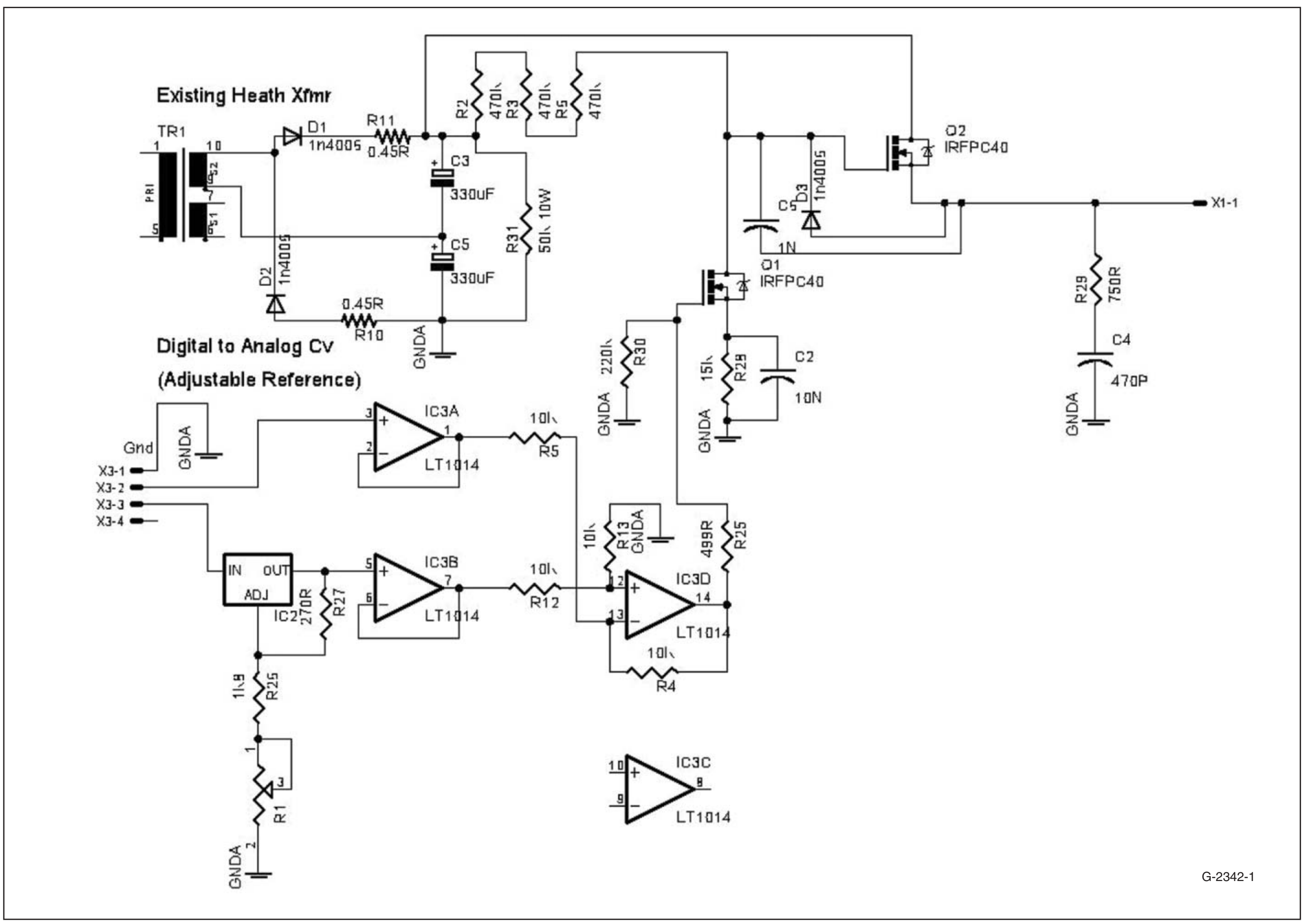

FIGURE 1: B+ supply circuit.

MCUTracer via a 4-conductor ribbon cable, and separate B+ and C– modules.

The high-voltage B+ supply is a variant of that shown in *The Art of Electronics*[8], using instead a pair of beefy IRFPG40 HexFETS. I kept the topology of the Heathkit supply and used a voltage doubler consisting of a pair of 1N4007 rectifiers, 330μF/450V elec-

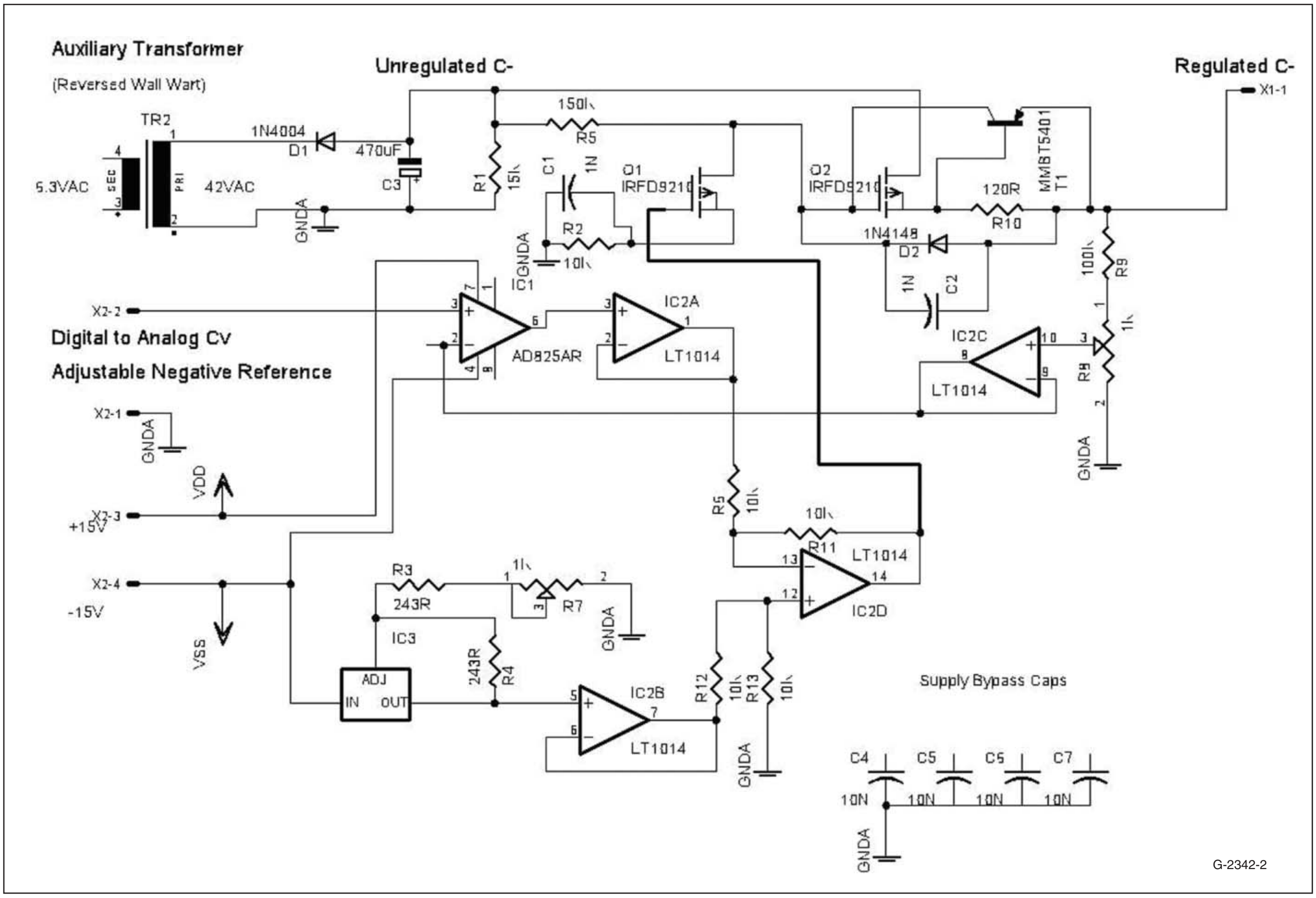

FIGURE 2: C– supply circuit.

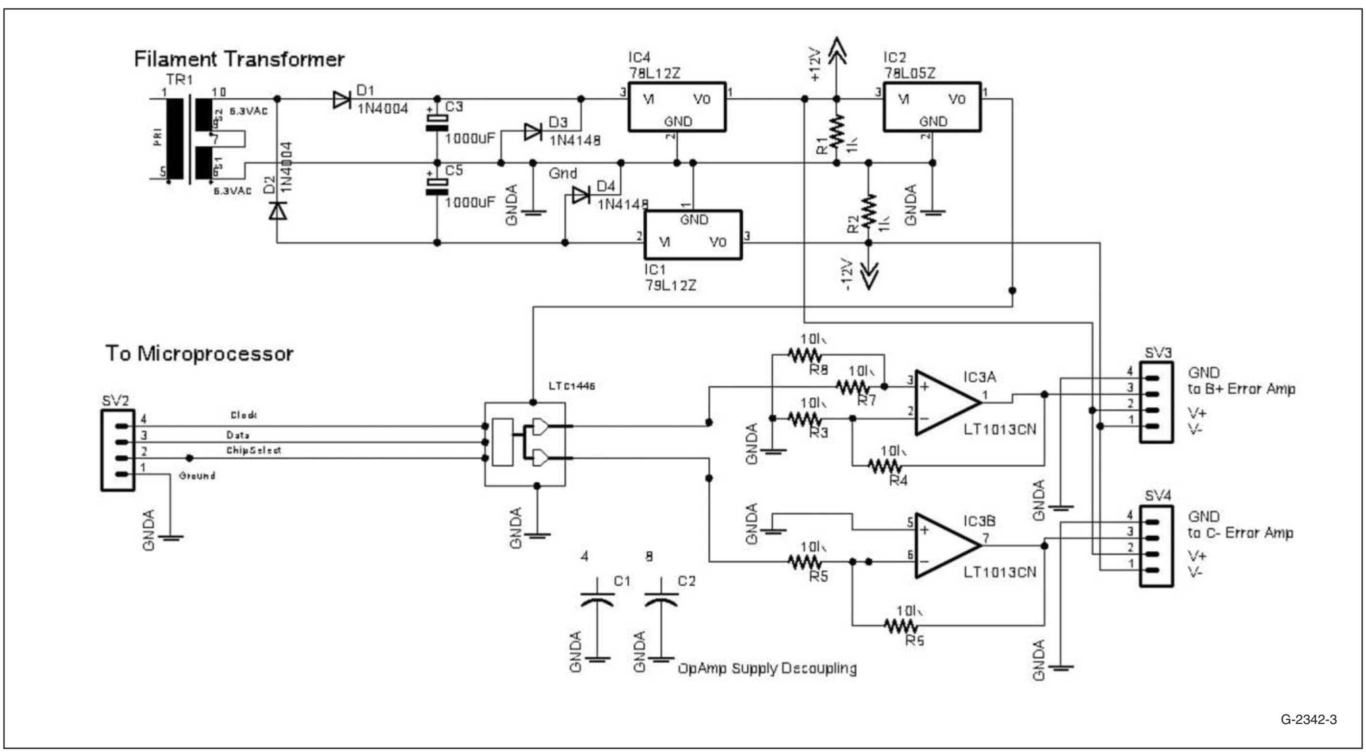

FIGURE 3: Operational amp/DAC supply circuit.

trolytic capacitors, and 0.47R current limiting resistors. Instead of a fixed reference as shown in the text, I used a 12-bit 0 to 4.1V signal from the LTC1446 DAC as an adjustable reference to the inverting input of the op amp.

This voltage to the gate of the first HexFET is biased with an adjustable reference using an LM317LZ linear voltage regulator; while this isn't the quietest bias source, it was easy to implement. The trimmer resistor can be used to set Q1's gate threshold and will serve to set the "Zero Intercept" for the supply. The B+ supply in *Fig. 1* is not regulated. Because the signal to the first HexFET is biased with the LM317Z and its associated circuitry, the output of the power supply increases monotonically with the input from the DAC.

The C– supply is also a variant of *The Art of Electronics* design using the International Rectifier IRFD9110 DIPFET. An Analog Devices AD825AR is used as the error amplifier, and a Linear Tech LT1014 is used to buffer the error and bias signals used to drive the gate of the first HexFET. I had expected to use one of the windings of the Heath power supply transformer to power the C– supply but this proved to be a little impractical without the use of a resistive divider to bring the voltage down. Instead I borrowed the transformer from a wall wart, cracked the case, and mounted it securely on perfboard and spacers on the rear of the IP-17 cabinet. I attached the secondary of this auxiliary transformer to one of the 6.3V windings on the IP-17 filament transformer. The primary of the transformer now became the secondary and delivers about 42V AC to the C– supply rectifier.

The controller board houses a voltage doubler circuit, 78L15 and 79L15 regulators for the operational amplifiers, and a 78L05 regulator for the DAC. An LT1013 error amplifier is shown buffering the output from the LT1446 digital-to-analog (DAC) converter, and inverting the voltage for the C– supply. The DAC has its own voltage reference, which simplifies construction. The con-

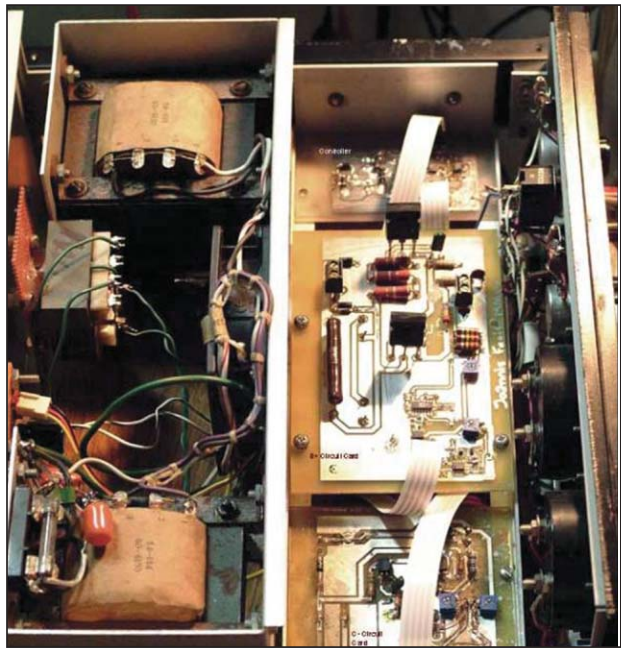

PHOTO 1: Inside the MCUTracer cabinet.

troller board, B+ and C– supplies shown here use a mix of surface-mount and through-hole devices, but you could just as easily use through-hole parts. Note that the pinouts of the surface mount and through-hole devices differ.

The schematics for B+, C–, and operational amplifier/DAC supplies are shown in *Figs. 1*, *2*, and *3*.

IMPLEMENTATION

As mentioned at the outset, I chose to cannibalize a Heathkit IP-17 high-voltage adjustable power supply as the carcass for this venture. These supplies routinely show up on eBay for $30 or so. (If you can't find an IP-17, there are high-voltage supplies by Eico and Lambda, as well as earlier Heath supplies which you could press into service.)

The IP-17 has switchable meters for voltage and current, 5-way binding post connectors for B+, C–, and 0.0–6.3–12.6 filament voltages. Use of the IP-17 will save a lot of time and metal bending, searching for transformers, and so on. The IP-17 has a 12.0 × 3.7″ saddle on which the original and new modular circuit boards rest. The new boards fit right into this saddle.

It's a simple matter to snip the connections, remove the saddle, drill out the rivets, and install the new circuit boards. Connections to the circuitry of the IP-17 are pretty much as they had been, except that you remove the connections to the control potentiometers along with the surplus wiring which went to the original IP-17 regulator board. Since Heath did not include a connection from the chassis ground at

PARTS LIST

Name	Alias	Shape
C1	100N	KERKO4X3R2_5
C2	330U/25V	ELKO6_3R5
C3/25V	330U/25V	ELKO6_3R5
C4	100N	KERKO4X3R2_5
C5	100N	KERKO4X3R2_5
C6	100N	KERKO4X3R2_5
C7	100N	KERKO4X3R2_5
C8	100N	KERKO4X3R2_5
C9	100N	KERKO4X3R2_5
C10	100N	KERKO4X3R2_5
C11	100N	KERKO4X3R2_5
C12	100N	KERKO4X3R2_5
C13	4.7U	TAN KO4_5R5
C14	4.7U	TAN KO4_5R5
C15	100N	KERKO4X3R2_5
C16	100N	KERKO4X3R2_5
C18	100N	KERKO4X3R2_5
C19	10N	KERKO4X3R2_5
D1	1N4001	DIO_DO35
D2	1N4001	DIO_DO35
DB1		DB9FL
POT1	10K	SPITRI64Y
POT2	10K	SPITRI64Y
POT3	10K	SPITRI64Y
POT4	1K	SPITRI64Y
POT5	1	SPITRI64Y
POT6	10K	SPITRI64Y
R1	330K/0.5W	R0411R5
R2	330K/0.5W	R0411R5
R3	330K/0.5W	R0411R5
R4	330K/0.5W	R0411R5
R5	330K/0.5W	R0411R5
R6	330K/0.5W	R0411R5
R7	100K	R0204R7_5
R8	10K	R0204R7_5
R9	10K	R0204R7_5
R10	10K	R0204R7_5
R11	10K	R0204R7_5
R12	10K	R0204R7_5
R13	10K	R0204R7_5
R14	10K	R0204R7_5
R15	10K	R0204R7_5
R16	10K	R0204R7_5
R17	10K	R0204R7_5
R18	10K	R0204R7_5
R19	10K	R0204R7_5
R20	1K	R0204R7_5
R21	1K	R0204R7_5
R22	51R	R0204R7_5
R24	51R	R0204R7_5
R25	51R	R0204R7_5
R26	51R	R0204R7_5
R27	51R	R0204R7_5
R28	51R	R0204R7_5

(to pg. 68)

the rear of the cabinet to the binding post on the front panel, I added it in my version. In addition, the space once occupied by the 6L6GC regulator tubes has been "nibbled" out and replaced with a fan to cool the high-voltage transistors.

I mounted the LTC1446 DAC, its power supply, and operational amplifiers on a PCB (this version uses surface-mount components), and the B+ and C– supplies on separate PCBs. Connections from the DAC to the non-inverting inputs of the error amplifiers and the error amplifier power supplies are via small ribbon cables fitted with 0.100″ Molex connectors. Connection between the microcontroller of the MCUTracer and the power supply is also with a 4-conductor ribbon cable. The controller, B+, and C– supplies mount right onto the IP-17 saddle (*Photo 1*). Note that in order to allow the filter capacitors (mounted on the bottom of the PCB) to clear the bottom of the IP-17 chassis, I needed to mount the B+ supply on ½″ insulated spacers.

Where the 6L6GC regulator tubes were in the original, I mounted a 12V fan to cool the pass MOSFETs. On the rear of the cabinet is a small power supply which borrows 6.3V AC, and rectifies and filters the voltage for the fan. I also replaced the neon bulbs (they weren't functional) on the front of the cabinet with a red and green LED to indicate whether plate or grid voltages were being measured by the front-panel meters. These are powered from the fan supply. Note that heatsinks are used on the B+ supply but these were omitted for clarity.

GRAPHICAL USER INTERFACE (GUI)

The most difficult part of the project was to provide an interface which would be easy to implement and effective in use. In part 1, we used Parallax's StampDAQ™ interface between the Basic Stamp II and Microsoft Excel. For part 2, Martin developed a macro which implements the GUI using StampPlot®, a more sophisticated program. StampPlot and the macro are freeware available on his website, www.stampplot.com. To use StampPlot, copy the Basic Stamp II code in the appendix to your microprocessor, download StampPlot, install, and run once. Then download the HVSupply macro, save to your desktop, double-click to open, and you're good to go!

Several "text boxes" in which you can input data are shown in the screen shot of the StampPlot HVSupply GUI (*Fig. 4*). The plate and grid text boxes supply data to the Basic Stamp for the two 12-bit words used by the LTC1446 DAC to control the power supply. The PVStep and GVStep text boxes tell the Basic Stamp how to increment the plotting of data. Finally, I Max, highlighted

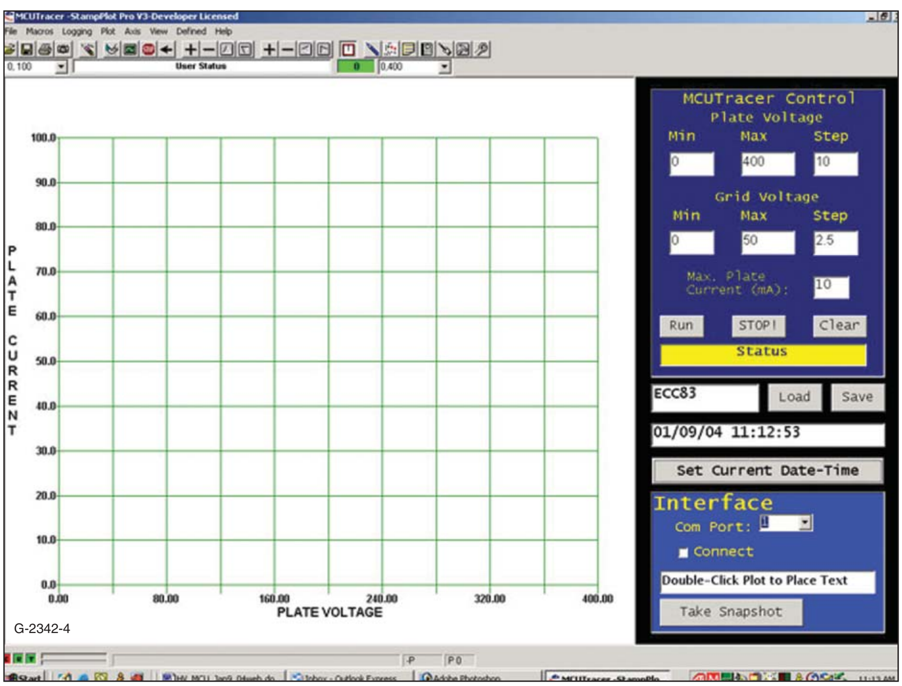

FIGURE 4: Sample StampPlot text box.

in red, provides you with a threshold which—if reached—causes the program to stop operation and brings the power supply to "StandBy" and all voltages to zero.

Testing the power supply: do not connect the B+ and C– power supplies to their regulator circuitry yet. First, test the DAC and error amplifier control circuitry to see that they are working correctly. Enter 400V into the maximum plate voltage and 50V into the maximum grid voltage text boxes, 100 into the plate voltage STEP box, and 10 for the grid voltage step box. Attach oscilloscope probes to each of the DAC outputs. Power up the power supply and press the CONNECT check-box and mouse-click the run button.

Examine the resulting waveforms, which should appear as the ramps in *Fig. 5*. This chart illustrates that, with the grid voltage initially set at zero, the plate voltage will be stepped from 0 to 400V (five steps including zero). The grid voltage is then reduced 10V and the plate voltage stepped again.

TESTING THE HIGH-VOLTAGE CIRCUITS

In working with high-voltage circuits, it's always a good idea to keep one hand in your pocket, or at least out of harm's way. This will prevent any inadvertent (and catastrophic!) application of current across your heart! You should attach the heatsinks to the transistors using fiber shoulder washers and an insulation pad.

Make the high-voltage transformer connections to their respective regulatory circuits. Set the trimmer potentiometers to their midpoint positions. Put the midpoint supply voltages (200V for the B+ supply and 25V for the C– supply) into the respective boxes on StampPlot. By typing a "1" into each of the grid and plate STEP boxes, the power supply goes into a "static calibration mode," that is, the DAC will not ramp, but provide only one voltage to the amplifier on the power supply board.

Attach a digital voltmeter to the B+ and C– outputs on the power supply. Switch on the power supply and examine the peak B+ and C– voltages. Click on the CONNECT and RUN checkboxes and the voltage on the meters should approximate half of full scale. You will undoubtedly have to adjust the trimmer potentiometers so that the range of voltages covered linearly tracks the output of the DAC on the controller board. The potentiometer, which attaches to the power supply output, the "Error Amplifier," potentiometer, adjusts the slope of the power supply output. The pot, which is attached to the LM317 or LM337LZ bias supply, adjusts the "Zero-Intercept." A scope is very handy in making these adjustments, but with a little patience, you can do them by hand.

USING MCU POWER SUPPLY AND MCUTRACER

With the MCUTracer and MCU power supply connected, input the correct plate and grid voltages for the proper range of B+ and C– voltages. Enter the increment of voltage steps in the appropriate text box and the maximum plate current in the text box where indicated.

To connect the HV power supply and

(from p. 66)

R29	2K2	R0204R7_5
R30	2K2	R0204R7_5
R31	2K2	R0204R7_5
R32	2K2	R0204R7_5
R33	2K2	R0204R7_5
U1	LT1014	DIP14A
U2	LT1013	DIP8A
U3	LTC1093	DIP16A
U4	BS2	DIP24
VREF	LM4040-4.1	LM78LXX_
VR_POS	7805	LM78XXV
V_NEG	79L05	LM79LXX
QNT	**ALIAS**	**SHAPE**
13	100N	KERKO4X3R2_5
1	330U/25V	ELKO6_3R5
1	330U	ELKO6_3R5
2	4.7U	TANKO4_5R5
1	10N	KERKO4X3R2_5
2	1N4001	DIO_DO35
1		DB9FL
4		HOLE
2		HDR1X8
7		HDR1X2
4	10K	SPITRI64Y
2	1K	SPITRI64Y
6	330K/0.5W	R0411R5
2	100K	R0204R7_5
12	10K	R0204R7_5
2	1K	R0204R7_5
6	51R	R0204R7_5
5	2K2	R0204R7_5
1	LT1014	DIP14A
1	LT1013	DIP8A
1	LTC1093	DIP16A
1	BS2	DIP24
1	LM4040-4.1	LM78LXX_
1	7805	LM78XXV
1	79L05	LM79LXX

MCUTracer to the PC, click the "Connect" check-box. Next click the "Run" button once to begin graphing plate current (Y-axis) versus plate voltage (X-axis). To halt operation, click the STOP! button.

If the maximum plate current value is exceeded during operation, the MCUTracer will halt operation and bring the plate and grid voltages to zero. If this situation occurs, you will need to make the appropriate adjustment in one or more of the parameters you have chosen and try again. As an example, perhaps the grid voltage was too close to zero, in which case you must choose a different minimum value.

You can use StampPlot to take a JPEG screen shot of the data. You can also save the data and download it into a spreadsheet program for later analysis or reference. You'll find more information on the operation of StampPlot on the website www.stampplot.com.

THE BASIC STAMP PROGRAM

Using the "Read" function in StampPlot, the PC reads data from the text boxes for plate and grid voltages, increments, and plate current. These values are transmitted to the Stamp using the "SERIN" command.

The number of iterations and lines drawn are calculated from user-supplied input "PV Steps" for plate voltage steps and "GV Steps" for grid voltage steps. Every time a new grid voltage is calculated, another line is plotted in a different color from the previous.

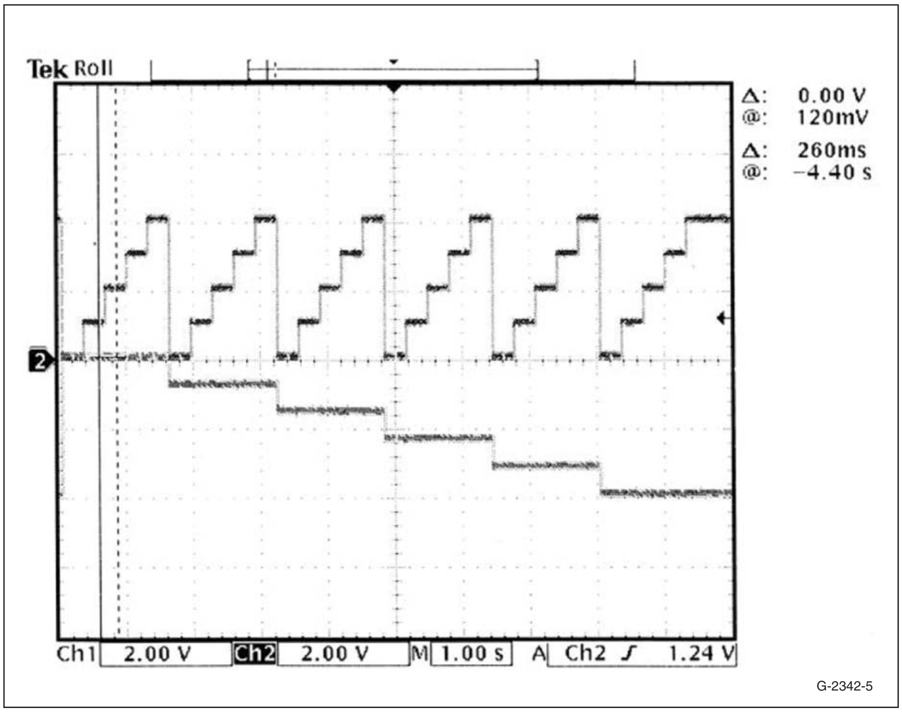

FIGURE 5: Waveforms in power-supply test.

The code will first take the plate voltage and current for device 1 and then plot a small circle on the chart at this location. When the plate voltage is incremented, another circle is plotted and a line is drawn between the two points. The Basic Stamp then resets the grid voltage to its initial setting and performs the same routine for another device.

For the second device, however, only a circle is plotted. If no second device is attached, it will just plot a bunch of circles along the X-axis. You can remove the code lines for a second device if you wish, snipping where necessary! When the program has made one complete iteration for a particular grid voltage, the grid voltage is plotted just to the left of the ending point.

Changes and improvements in the code will be posted on the StampPlot website, http://www.selmaware.com/stampplot/pubs_products/MCU_Tracer/home.htm. ❖

REFERENCE

8. Horowitz, Paul, and Hill, Winfield, "High Voltage Regulated Supply," *The Art of Electronics*, Cambridge University Press, Cambridge. Second Ed., 1989, p. 369.