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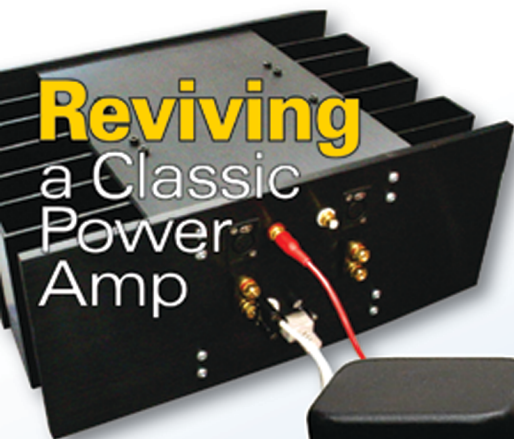
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The PLH Amplifier

By Nelson Pass

This solid-state magician demonstrates his latest design.

In 1969 John Linsley Hood wrote in *Wireless World*:

During the past few years a number of excellent designs have been published for domestic audio amplifiers. However, some of these designs are now rendered obsolescent by changes in the availability of components, and others are intended to provide levels of power output which are in excess of the requirements of a normal living room. Also, most designs have tended to be rather complex.

In the circumstances it seemed worthwhile to consider just how simple a design could be made which would give adequate output power together with a standard of performance which was beyond reproach, and this study has resulted in the present design.

He then described a Class A power amplifier using three gain stages of bipolar transistors in a topology which continues to be admired for its elegant simplicity and sound quality.

The centerpiece of this design is the middle stage, an NPN transistor used as a phase splitter, simultaneously driving the positive half of the output stage and the negative half with symmetric signals of opposite phase.

Figure 1 shows a simplified version of the JLH topology. Signal input appears at the base of Q1, and is amplified and inverted to drive the base of Q2. Q2 acts as a gain device and also a signal splitter, driving both Q3 and Q4 simultaneously, but out of phase with each other. Q3 and Q4 form the output transistors, Q3 operating as a common emitter gain device, contributing current and voltage gain,

and Q4 operating as a common collector device contributing only current gain. The resistors provide bias for the system, and R1 and R2 feed the output of the amplifier in a loop back to the emitter of Q1.

Q2 is the heart of the design, and in my opinion, it is the elegant economy with which it performs the complementary gain to drive the output devices that gives the circuit its classic beauty.

The JLH was designed at a time when “the tube era was in decline” and the new generation of designers was pulling out all the stops to produce big science amplifiers—pure voltage sources with high power and infinitesimal distortion—complex circuits with lots of feedback.

Thirty-six years and a little progress later, we can perhaps appreciate the simple charm of the JLH topology as an exercise in minimalism, but if you haven’t listened to one, you might be very surprised by the quality of sound, which is extraordinarily good within its power limitations. If you have efficient speakers and you like to listen to two-channel sound at reasonable levels, the JLH is still in the top rank.

The amplifier has reasonable specifications; nothing special that isn’t wildly exceeded by a \$3 chip, but it produces real music. Its flaws are not irritating and it does a wonderful job wringing more music out of modern recordings and even MP3s. I can’t think of another transistor design from that era that works as well.

Figure 2 shows the circuit rendered more completely, but for more extensive documentation on the versions of the JLH amplifier, I recommend The Class A Amplifier site www.tcaas.btinternet.co.uk.

In **Fig. 2** additional details of setting up the DC bias for each device are shown, where capacitors are used to separate DC values from AC values. C1 separates feedback from bias current. C2 separates input signal from input DC bias volt-

FIGURE 1: Simplified JLH amp.

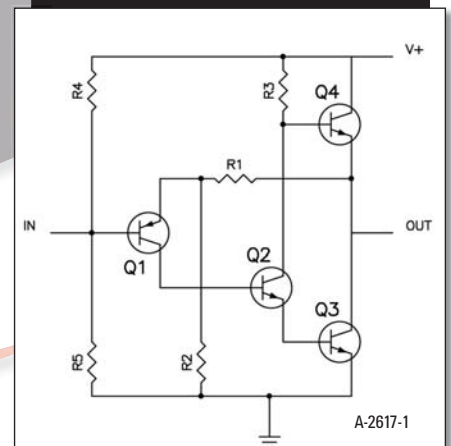
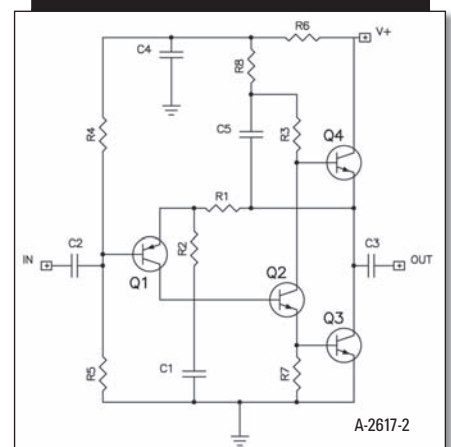


FIGURE 2: JLH amplifier 1969.



age, and C3 blocks the output DC of the amplifier from the load. C4 removes supply noise from the voltage powering the front end of the amplifier, and C5 forms a “bootstrap” circuit, making resistors R5 and R6 behave more like a constant-current source at audio frequencies.

The original JLH amplifier has approximately 55dB of open loop gain divided into 22dB of amplifier gain and about 33dB of feedback. As detailed in the original article, it delivered 10W at approximately 0.1% harmonic distortion or less.

The amplifier’s popular longevity speaks volumes about the quality of its sound, and this is understandable given its simplicity coupled with excellent measured performance. It has a particularly tube-like quality compared to the more complex solid-state designs of the era and since.

FIGURE 3: JLH amplifier 1996 (simplified).

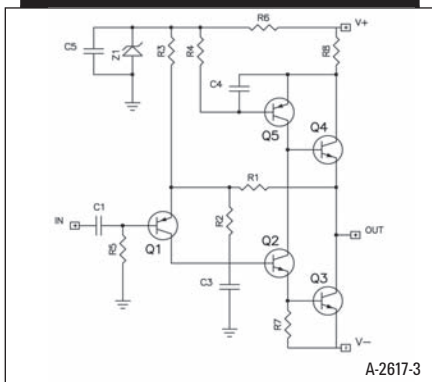
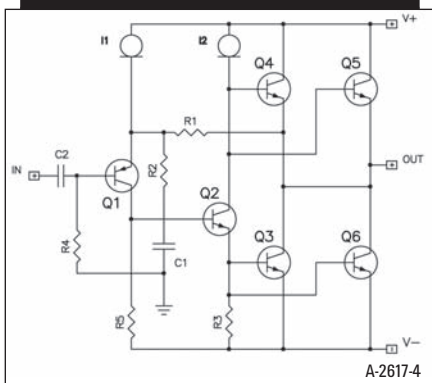


FIGURE 4: JLH variant 2000.



The distortion is largely 2nd harmonic, and is closely proportional to the output voltage. This means that .01% distortion at 0.1W becomes 1% at 10W, and you can draw a pretty straight line between the two points on a logarithmic graph. Such a curve is characteristic of a single-ended output topology, and there have been arguments regarding whether or not the output stage is single-ended Class A, push-pull Class A, or a mixture of both. We will be having some fun with that later.

One flaw in the original JLH design was that its bias current, that idling current which flows through the parts of the circuit, had some dependency on the power supply voltage, resulting in altered performance for different AC line voltages. Power supply regulation solves this problem neatly, but there were other ways this was addressed in later versions of the circuit.

NEWER JLH CIRCUITS

John Linsley Hood published an update to the amplifier in 1996 that addressed bias stability issues, parts substitutions,

and provided a version that had a direct-coupled output, eliminating the output capacitor. At the same time, it was in many ways the same amplifier, the measured performance being very similar.

The JLH circuit continues to be interesting to the audiophile community and has been the subject of several updates. **Figures 3 and 4** show simplified schematics of later generations of JLH amplifiers.

Figure 3 shows a simplified schematic of the 1996 version published by John Linsley Hood that fixes the bias stability issue with the addition of Z1 and the Q5 portion of the circuit. This version also direct-coupled the output of the amplifier, using dual supply rails.

In 2000 someone else produced the circuit seen in **Fig. 4**, in which constant-current sources are used to bias the first two gain stages, giving good power supply rejection to the circuit. This version also doubled up on the number of output devices. You will notice that **Figs. 1-4** have their feedback loop addressing the emitter of the feedback transistor. Nowadays they have gotten fancy and call it "current feedback."

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Just for your entertainment, I cobbled together the circuit of **Fig. 5** as an example with a differential input. An obvious variation, but I haven't seen it used. You can drive this input stage with a balanced signal by lifting C1 from ground and driving it as a negative input. At 100 Ω each, R6 and R7 will give this input

By June Linsley Hood

John Linsley Hood was born in 1925, and educated at Reading School (UK). He became a student assistant at GEC Research Laboratories, London, before joining the RAF, where he worked on radar, first at Telecommunications Research Establishment at Malvern, then at an experimental TRE unit at Malabar Hill, Bombay.

After a course at Reading University, he got a job in the electronics laboratory at Windscale in Cumbria, in the Department of Atomic Energy. Next he went to British Cellophane, in Bridgewater, Somerset, as head of the electronics research laboratory. He stayed there nearly 25 years working on varied instrumentation projects.

His interest in radio and audio began at the age of eight when his grandfather gave him an old crystal set to play with, and was boosted further when he made friends with a railway signalman who gave him useful information. While at Windscale he began his quest to design amplifiers which would have greater power and less background noise and distortion, which resulted in his Class A amplifier which was described and published in *Wireless World* in 1969.

He wrote many articles for *Wireless World* (later *Electronics World*), also for *Hi-Fi News*, *Electronics Today*, *Short Wave Magazine*, and *The Gramophone*.

He was asked to contribute chapters to Ian Sinclair's *Audio + Hi-Fi Handbook*, which encouraged him to write a book on his own, and in the late 1970s he joyfully took voluntary redundancy from British Cellophane in order to concentrate on writing and consultancy work.

He was also a keen photographer, an ARPS with a reputation for high quality monochrome landscape work. He was interested in photographic chemistry and in testing and comparing developers, lenses, and printing papers.

Mr. Linsley Hood died March 11, 2004.-Ed.

about the same open loop gain as the original input transistor with the 220 Ω degeneration of the original feedback impedance.

I recently measured the performance of a working copy of the circuit of **Fig. 4**. It had 17.5V supply rails and was biased at about 2A per channel. The open loop gain is also about 55dB into 8 Ω , and its measured performance is comparable to the original circuit.

Figure 6 shows the distortion versus output power. The bandwidth of the amplifier is -3dB at 100kHz, the damping factor is about 35, and the distortion versus frequency is fairly flat, rising slightly at 20kHz.

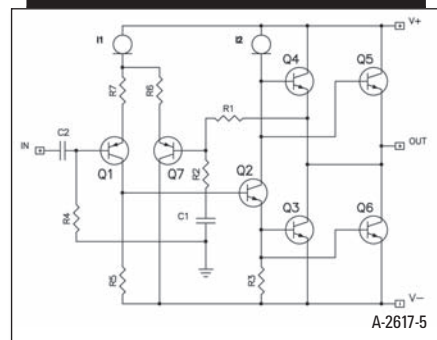
THE PLH AMPLIFIER

One of the issues that arises from adding gain stages to amplifiers is that while they increase the open loop gain and allow more feedback correction, they themselves are the source of additional distortion. While the extra feedback can lower the distortion numbers, usually the additional circuitry is reflected in a more complex distortion character having higher order harmonics and inter-modulation components. These are generally agreed to be less musical sounding.

Michael Cunningham wrote "Novelists must usually decide what degree of slavish accuracy would make their stories more alive, and what degree would make them less." The amplifier designer has a similar problem to solve. It's not hard to make an amplifier that measures well—it's comparatively hard to please audiophiles.

My own approach is to make the signal path as simple as possible, work to lower the distortion of that basic circuit before feedback is applied, and then apply minimal (or no) feedback, largely in agreement with the comments in Linsley Hood's original article. The result is not always the best objective measurements, but the

FIGURE 5: Differential input.



sound is often interesting.

The three-stage topology of the JLH amplifier routinely uses simple Class A operation and about 33dB negative feedback to achieve this performance, and this lured me to consider what kind of amplifier I could achieve with an even simpler circuit and less feedback. The output stage and the intermediate phase splitter cannot be dispensed with and still resemble a JLH, but you can certainly remove the input transistor.

By thumbnail calculation, the input transistor of the JLH contributes about 27dB of voltage gain. Remove it and the open loop gain of the amplifier drops into the region of about 28dB. If you scale back the gain of the amplifier from 22dB to 18dB, you end up with about 10dB of feedback—a very minimal amount.

Unfortunately, only 10dB of feedback means that the original amplifier will likely deliver something more like 1.5% distortion at 10W. Since such a figure is better than that seen on many single-ended Class A tube amplifiers (SET), this might be an acceptable amplifier. Actual-

FIGURE 6: THD vs. output at 1kHz at 8 Ω .

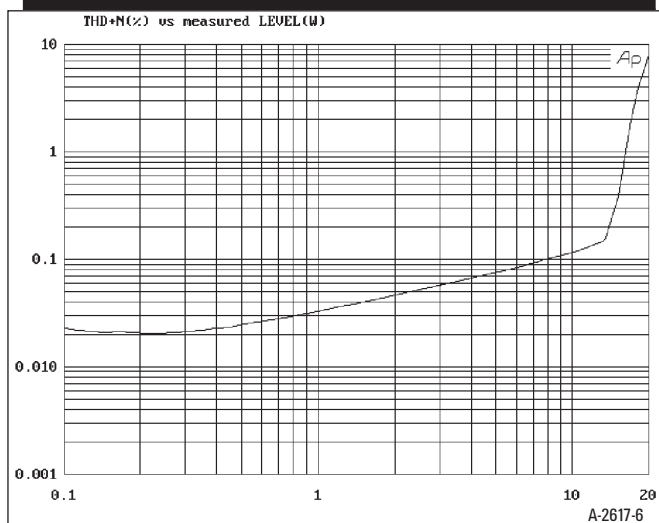
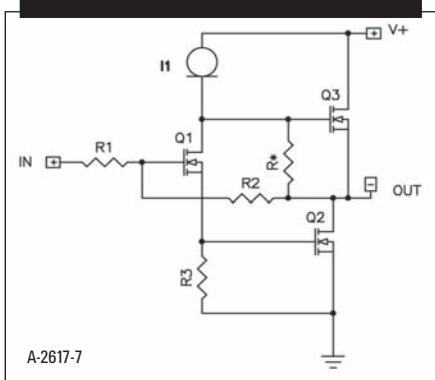


FIGURE 7: Simplified PLH amplifier.



A-2617-7

ly, since the input transistor is no longer contributing to the distortion figure, you would expect the performance to be better than that, and maybe that's the point.

With that in mind, I stripped down the JLH to the simpler MOSFET topology shown in **Fig. 7**, and (ahem) added my initial to the name.

I chose MOSFETs for their high input impedance and because they really deliver their most linear operation in Class A mode. Since the JLH performs two phase inversions in the signal path, removing the input leaves the amplifier phase inverting, so you can label the output node as "minus" and the output ground as "plus."

The feedback path of R1 and R2 now addresses a "virtual ground" at the gate of Q1. Q1 is biased by a current source I1, and drives the gates of Q2 and Q3 simultaneously with voltage of opposite phase.

As a concept, you can get about 35dB open loop gain from this circuit with the intended MOSFETs. You would only see

it at lower frequencies, however; because the gate capacitance of the MOSFETs will come into play at the higher audio frequencies.

To give a more uniform open loop gain to this circuit across the audio band, and a more interesting comparison to the original circuit, choose values similar to the original JLH without the input transistor, which means that for the circuit of **Fig. 7**, you would be adding 0.47Ω source resistors to Q2 and Q3, and this will give you about 26dB open loop gain.

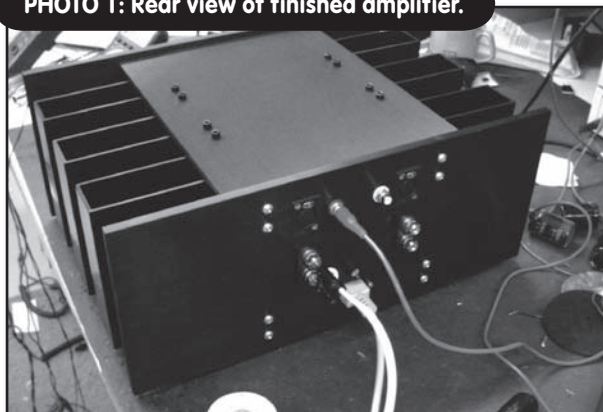
Like the original JLH, this circuit operates between a positive voltage and ground, so you will need input and output coupling capacitors. There is no reason that it can't be made direct-coupled using dual supplies, but I will save that for another time.

Figure 8 shows the actual circuit. As with the simplified circuit, Q1 is the input transistor and the feedback is accomplished through R1 and R2. The source of Q1 follows the gate input signal and drives the transistor Q2 in common source mode (voltage and current gain). The drain of Q1 delivers an inverted and amplified version of the input signal to drive transistor Q3 in common drain mode (current gain).

The idealized current source I1 and R* of the simplified circuit is replaced by the network of P2, R4, and C2. C2 produces a "bootstrap" connection to the output that makes P2 look like a constant-current source in parallel with R*, approximately the resistance between the wiper and the counter-clockwise connection of P2. P1, R6, and C5 form a filter to take out power supply noise, and adjusting P1 will set the bias current of the amplifier.

P2 will be used to set the relative gain contribution of the upper transistor Q3 versus the lower transistor Q2, but does not affect the bias current

PHOTO 1: Rear view of finished amplifier.



or output DC value.

Most of the resistors are ¼W types, but I recommend a 3W value for R4, R8, and R9. The capacitors are all rated at 50V. None of the values needs tight tolerance. P2 is best chosen as higher wattage type. You can get by with 2W, but 5W is preferred.

I have shown nominal voltages on the schematic as guides. The circuit will work with supply voltages from 35 to 45V without modification using P1 and P3 potentiometers to adjust the bias current and output DC. This particular amplifier is biased at 2A, and the output voltage is set at 20V, or one-half of the 40V supply value.

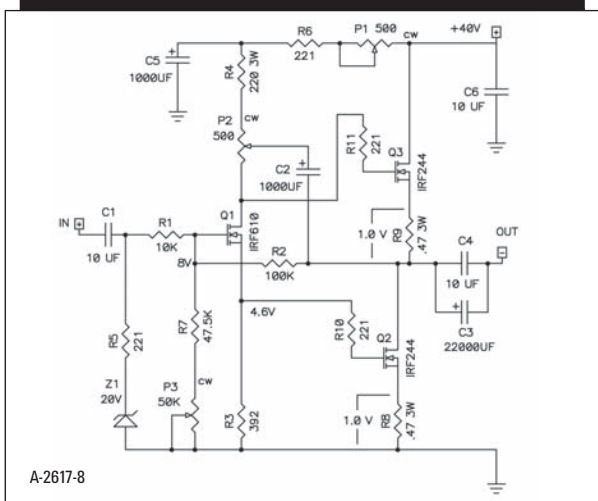
The MOSFETs used are fairly arbitrary, and in general you can substitute similar types. Since virtually all the power MOSFETs are rated at least 40V, you are left a wide latitude of choices of vertical types. Lateral MOSFETs would also work with adjusted resistor values.

Remember that all MOSFETs are static sensitive. It is not necessary to match the Q2 and Q3.

You will need a power supply that happily sources 35 to 45V at 2A per channel. A regulated supply is most ideal, as the bias will vary in rough proportionality to supply voltage, although we built four units without a regulated supply and they work fine. The heatsinks should be able to dissipate about 70-90W per channel with a rise of about 25° C. You can find examples sprinkled about at www.passdiy.com and elsewhere.

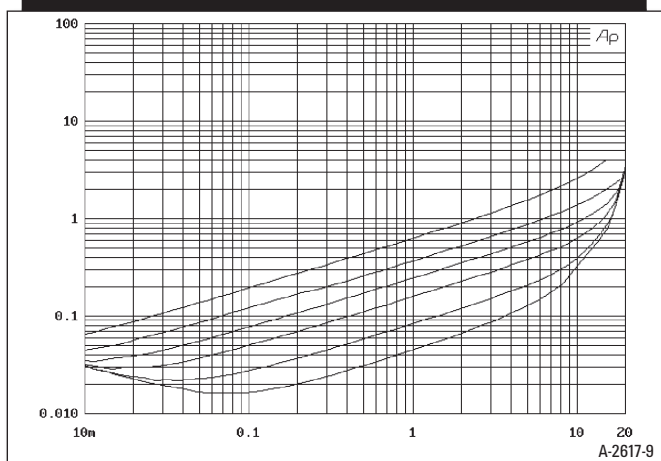
This summer, Chris and Matt Williams ("The Kleinhorn, Pt. 1, 1/05, p. 24) helped me to build the amplifiers which worked more or less the same. Lucky boys, they got away with two of them. These were made from machined plates of aluminum, bolted together, with heatsinks

FIGURE 8: PLH power amplifier.



A-2617-8

FIGURE 9: Distortion vs. output, one pair output device.



made from square aluminum tubing and all of it anodized black. The wiring was all point-to-point on an output stage PC board (**Photo 1**).

We used a transformer with a 35V secondary rectified into a 30,000 μ F capacitor. We filtered that supply voltage through a 0.5 Ω at 25W resistor to another 30,000 μ F capacitor, which reduced the ripple noise by about 20dB. This powered both channels.

Remember to use good grounding—a star ground at the (-) terminal of the second capacitor is a good idea, and keep the input and grounding leads away from the power supply components, or you will pick up noise. The chassis should always be earth-grounded to the AC power plug, and the circuit ground was connected to the chassis through a 5A power thermistor.

ADJUSTMENT

At a minimum, adjustment of the ampli-

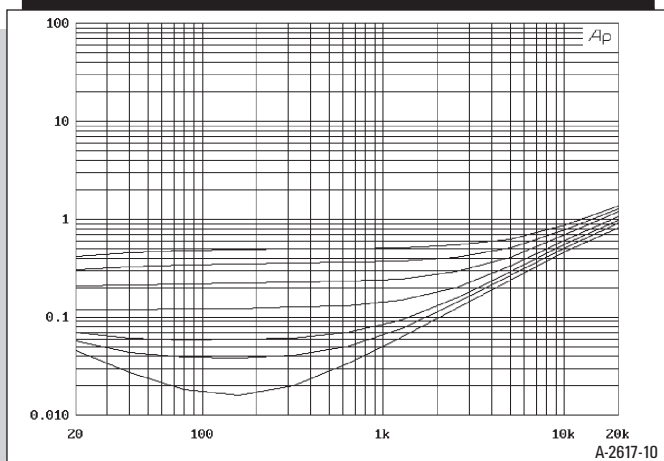
er requires a DC voltmeter. I recommend the use of a Variac® to slowly turn on the AC power for your initial check-out. Before you apply power to the amplifier, set the value of P1 to its maximum resistance, which should set the bias at a minimum current. Set P2 and P3 at their midpoints.

I recommend firing up (unfortunate phrase, that) one channel at a time, with the other channel disconnected from the supply rail. It's a good idea to put a 3A fast blow fuse in series with the positive rail of each channel, and to keep some spare fuses handy.

When power is applied, the output of the amplifier, the voltage across the source resistors R8 and R9, should be less than 1V. The output DC voltage, as seen on the drain of Q2, should be somewhere around 20V.

If you can bring the supply rail up to 40V without having the bias go above 1V, then adjust P3 to put the output voltage

FIGURE 10: Distortion vs. frequency, one pair output device.



at about one-half the supply voltage (20V with a 40V rail). Now slowly decrease the value of P1 until the source resistor voltage approaches 1V.

As the channel warms up, adjust P1 and P3 in small steps so that the output is at 20V and the voltage on one of the source resistors is 1V. Watch the output voltage and current draw for a half hour or so, readjusting as needed as the circuit warms up. The bias will tend to drift up, but the DC output value will tend to be more constant. At the end of this procedure you should have a stable channel.

If you share two channels off a common supply, you will be taking another look at these settings later because the supply voltage will drop a volt or more when both channels are connected.

PERFORMANCE

The final adjusted circuit has an open loop gain of about 26dB, a closed loop gain of 16dB, and uses about 10dB of neg-

FIGURE 11: PLH with parallel outputs times 2.

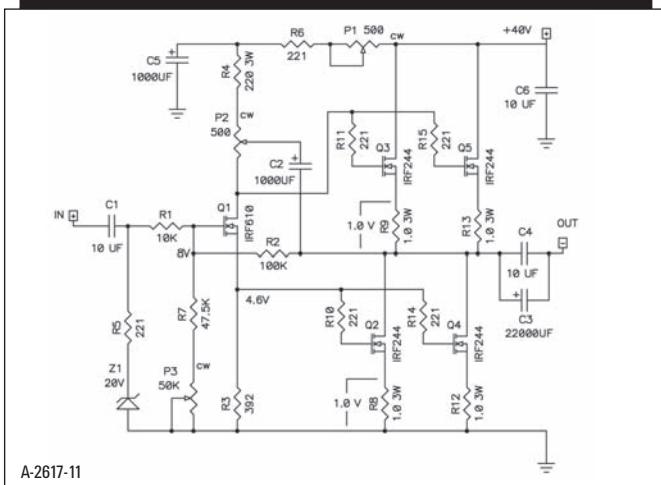
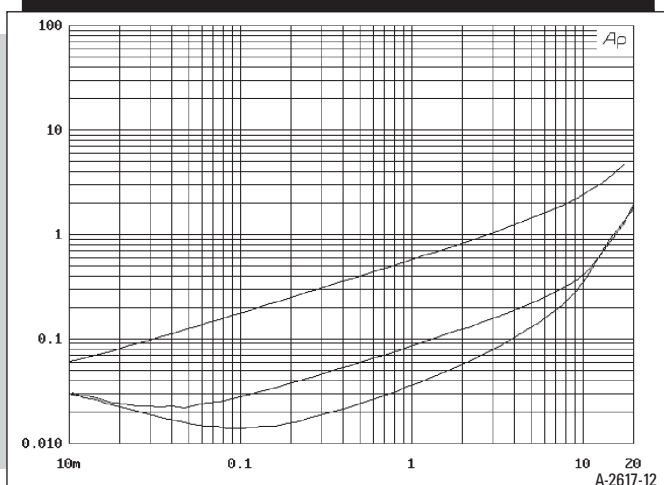


FIGURE 12: Distortion vs. output, two parallel outputs.



active feedback. The bandwidth is -3dB at about 1Hz and 100kHz. The (unweighted) noise is about 80µV. The input impedance is about 14kΩ, and the output impedance is about 3Ω (a damping factor of 2.5).

The upper and lower output devices share the same value of DC current, but as mentioned before, this circuit can be adjusted to vary the AC current contributions from each. When the values are set equal, then the classic push-pull balance is achieved. By adjusting P2, you can shift this balance so that one device takes a larger share of the output.

As you turn P2 clockwise (note the "cw" on the schematic), a greater percentage of current comes from the upper device Q3, and as you turn it counterclockwise, a greater percentage comes from the lower device Q2. Fully counterclockwise results in the upper transistor bank functioning as a constant-current source, with all the gain in the lower bank, operating as a purely single-ended Class A circuit. An 80% clockwise rotation of P2 gives about a 1:1 ratio between the devices (as the pot rotates through about 10 hours of a clock, with full clockwise as 11 PM, you would set the pot to 9 PM).

Linsley Hood spent some time experimenting with his circuit using mismatched output devices and noted that if you don't have equal gain, it is better to place the devices with greater gain on the bottom of the circuit. I confirmed that this was the case by adjusting the balance between the positive and negative halves of the output stage, measuring and listening to various settings.

I spent a great deal of time playing with this balance, trying to reconcile the best curves to the best sound. At low and mid-band frequencies, setting P2 to a value that gave equal contribution to the output current resulted in the lowest distortion versus output power. By comparison, setting the value so that the lower half delivered two-thirds of the output current gave about twice the distortion, but it remained more constant over the audio band.

This more "single-ended" output stage showed a more pure second harmonic type distortion than the more "push-pull" output stage, which contained more third and higher harmonics. This was not a completely apples-to-apples comparison, though, as increasing R5 also had the

effect of slightly increasing the open loop gain, and thus the amount of negative feedback.

I encourage you to try varying P2. Remember that you can confirm the ratio of upper versus lower output current by measuring the AC voltage across the source resistors R8 and R9, while running a low-frequency sine-wave into a resistive load at about 5W or so.

Figure 9 shows the distortion versus power into 8Ω at 1kHz, with the lowest curve showing the 1/1 ratio (push-pull operation) and the highest curve showing

a 0/1 ratio (P2 fully counterclockwise), where Q3 operates as a constant-current source and the circuit is operated as single-ended Class A.

Figure 10 shows the distortion versus frequency at 1W, and you see that the differences between the two settings tend to go away at the highest frequencies, where the variations in capacitance of the devices do not tend to cancel. Also, there are some harmonic content differences over the range of settings. The lowest distortion curves have greater content of higher order harmonics, which tend to go



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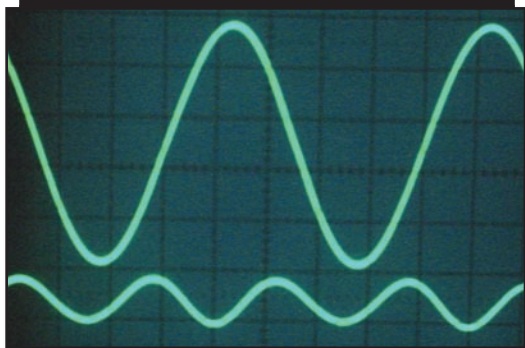
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PHOTO 2: Distortion waveform at 1W at 1kHz.



away as you move toward single-ended operation.

Photo 2 shows an example of the harmonic content with the ratio of top/bottom gain of 2/3, which was one of my favorite settings.

I suggest that you mount P2 on the front panel of the amplifier, and listen to it at the various values. Over most of the range, the open loop gain of the amplifier does not alter significantly until you get close to the counterclockwise position (0/1), where it drops off about 5dB, resulting in about 5dB of negative feedback. At the 1/1 position, the open loop gain is about 26dB, giving about 10dB of feedback. Note that the differences you will hear are not strictly feedback related, but are also a function of push-pull cancellation of distortion.

When the amount of power and heat starts getting big, we often parallel out-

put devices in a circuit in order to share the dissipation. In this case the dissipation is about 40W per device, and it is tempting to use another pair of devices in parallel with the originals to take that value down to a more reliable 20W per device. **Figure 11** shows the circuit that does this. Note that I have increased the values of the source resistors to 1.0Ω.

Figure 12 shows the performance of this circuit with the 1/1 setting being the lowest curve, the 0/1 setting the highest, and a 2/3 ratio shown in the middle. There is somewhat less distortion than the curves of **Fig. 9**.

Figure 13 has a little bit different story to tell. At 1W over the audio band, the 1/1 ratio might not be considered the best choice, as the 2/3 value has a more uniform performance.

The question often comes up, “How many output devices in parallel is optimal?” The best answer often lies in trial and error. I built a version with four parallel devices and 2.0Ω source resistors. After measurement and listening, I decided that four pairs were too many.

It’s interesting to look at a casual comparison of the results with the original JLH, PLH, and two of the Zen amplifiers in the context of simplicity, gain, and feedback applied. All of the amplifiers

amount of feedback on some amplifiers helps them to measure better, so you can assume that an otherwise equal amplifier would have some proportionality between measurement and feedback. I was curious about this—if it were true, then how would these amplifiers measure if they all had the same amount of feedback?

I normalized the output impedance and distortion figures to what we would expect if all the amplifiers had 20dB of feedback and assumed that the difference would be due to feedback (**Table 1**). Here you can see some interesting things. First, for unknown reasons, the normalized output impedances of the JLH and PLH are about twice that of the Zen amps.

Second, you see that the original Zen amp stands out as having the highest distortion. This is partially explainable by the fact that the other three designs all have some mechanism to cancel some distortion without negative feedback. In the JLH and PLH, the output stages are driven out of phase by a phase splitter, and in the ZV4 (Zen Variations #4), the P channel input buffer and Aleph current source offer some cancellation.

While the distortion numbers for the JLH are better than the PLH, it operates at about 23dB more negative feedback. Normalizing these numbers to a 20dB feedback figure, the PLH appears at least four times more linear. Of the potential explanations, the first is the possibility that the more heavily biased MOSFETs are more linear than the bipolar devices.

Another possibility is that the input device of the JLH, used to produce much of the feedback, is contributing significantly to the distortion. It could be that the data is in error (like *that’s* never happened), or perhaps the calculation assumption that the distortion would be inversely proportional to the negative feedback is flawed. Maybe it’s all these things.

AND NOW FOR SOMETHING COMPLETELY DIFFERENT

Somewhere in the last 26 articles, I grew tired of telling the readers how wonderful it sounds. So I decided not to do that anymore. If you want to know, go build it.

Comments and questions are best directed to the Pass Labs forum on www.diyaudio.com *aX*

FIGURE 13: Distortion vs. frequency, two parallel outputs.

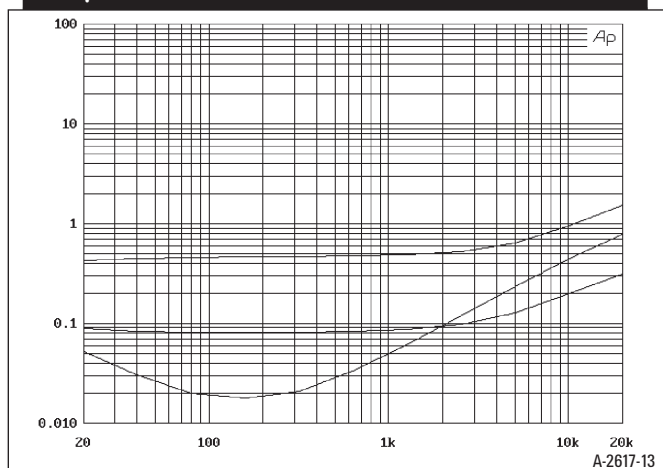


TABLE 1

	Gain (dB)	Open Loop (dB)	Feedback (dB)	Zout ohms	Normalized Ohms	THD % @ 1 Watt	Normalized @ 1 Watt	THD % @ 10 Watt	Normalized @ 10 Watt
JLH	22	55	33	0.23	1.03	0.03	0.13	0.12	0.54
ZEN	9	23	15	1.00	0.53	0.60	0.32	2.50	1.33
ZV4	13	33	20	0.60	0.60	0.06	0.06	0.20	0.20
PLH	17	27	10	3.00	0.95	0.04	0.01	0.35	0.11



The Short Line Array

This line array will fit your space requirements as well as your budget.

By Bill Fitzmaurice

I've had several requests for a small high-sensitivity speaker good for either hi-fi mains, control room monitors or, most especially, home theater (HT) center channel. One of those requests came from me. Ever since I outfitted my HT with TLAH mains and a T18 sub, I haven't been able to hear the center channel well enough, as the 87dB sensitivity of my MTM center just wasn't enough to keep up. Obtaining more sensitivity with a line array was the obvious answer, but like most of you I also don't have room for another 3' long plus cabinet in my entertainment center.

I decided that the largest box I could tolerate was 26" long, about 7" high, and 6" deep. So was born the short line array (SLA). Short it may be in size, but in sensitivity and re-

sponse it's right up there with the big boys, as you can see in the SPL chart (Fig. 1).

FEATURES

One advantage of a line array is that you can use many inexpensive low-powered drivers to produce high sensitivity. Another is that you can have a very good-sounding cabinet for very little money. When searching out drivers to use in this project, check out the web for specials and closeouts.

The woofers I used in the original were 4" Pioneer "NSB" closeouts from Parts Express, part # 269-568, which I purchased for 49 cents apiece. Specs are f_s 105Hz, Q_{TS} .76, V_{AS} .1ft³. The tweeters were Onkyo closeouts, part # 269-780, 24 cents apiece. These tweeters have no mounting flanges,

so you must glue them in place. There are many variations of this Onkyo tweeter available, some with coaxial mounting frames that you can discard.

Also note that these tweeters have three wiring lugs; the middle lug is not connected to the voice coil, so ignore it. If you use different drivers, it is critical that the woofers have squared-off "pincushion" style frames and that the tweeters are no more than 1.5" diameter.

A center-channel home theater speaker is usually horizontally mounted (Photo 1). This can be problematic with all alignments, but particularly with line arrays, which are by design intended to be vertically mounted. However, you can still obtain very good results with the SLA. By having the woofer frames literally touching each other, and with the minimal frame width of the pincushion design, the outputs of the woofers integrate into a single cohesive wave front up to at least 4kHz.

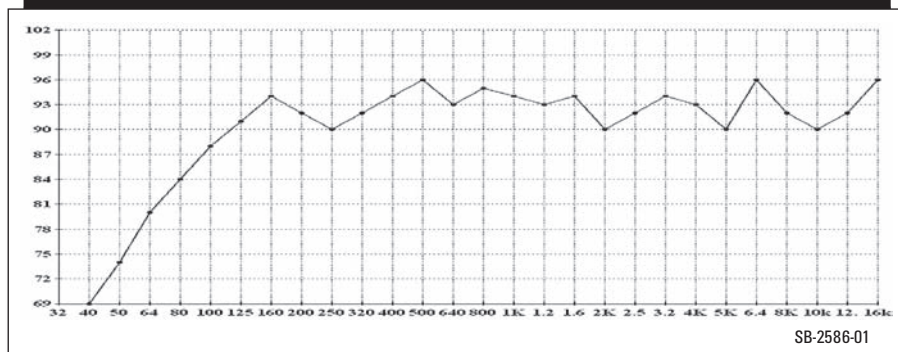
Because the main output of HT center channel is voices, with a primary bandwidth from 150Hz to 5kHz, there is no discernible comb-filtering. Even in music applications, combing is not a serious problem when this cab is lying down, but if at all possible it should nonetheless be upright.

The tweeters are mounted on the rear of the baffle, which is normally not preferred. However, the SLA uses this scheme to its advantage. The tweeters fire through 1" diameter holes that are chamfered with a 1/2" radius quarter-round router bit, transforming each hole into a small horn. This increases the effective radiat-

PHOTO 1: The SLA.



FIGURE 1: Frequency response.



ing plane of each driver, minimizing the frame space between them to aid in their wave-front integration. Even without this feature, the close-spacing of their acoustic centers guarantees wave-front integration to at least 7.5kHz, but with it they work even better.

CONSTRUCTION

Because the woofers are so close that their mounting holes overlap, and because of the routing of the tweeter micro-horns, you should make the SLA baffle out of solid wood. Its

size is a standard 1×8 , which actually measures $0.75" \times 7.25"$. Any wood will do, though you'll probably prefer to use something that matches the rest of your system. You can use solid wood for the rest of the cabinet, but plywood or MDF is OK as well. I used red oak for my baffle, along with $0.75"$ red oak veneer plywood for the rim of the box and scrap plywood for the back.

With the close spacing of the drivers, it would be very difficult to wire them and seal them to the baffle with cabinet sides in place, so construct

the cabinet with a permanently installed back and a removable baffle. This requires a bit of specialized hardware, though not so special that you can find it at any well-stocked hardware store: threaded inserts (Photo 2). Drill a hole in the edge of the box and drive in the insert with a screwdriver or Allen wrench—depending on the design—and you have an easy way to bolt the baffle to the rest of the box.

Start construction by cutting out the baffle. Its finished size is $26"$ long, but cut it slightly longer to allow for trimming after assembly. Draw two

FIGURE 2: Baffle view.

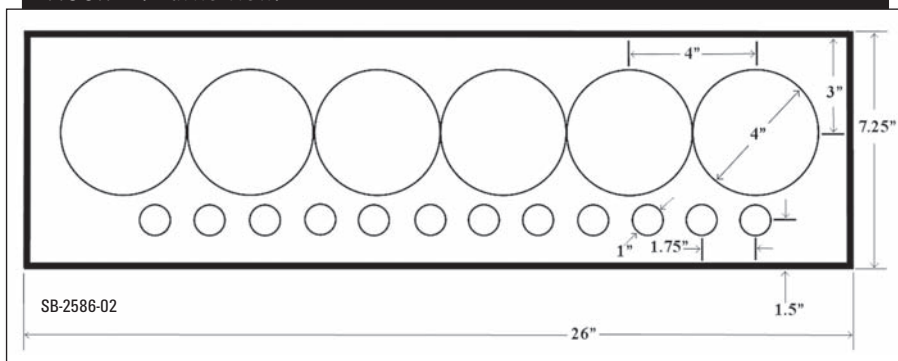
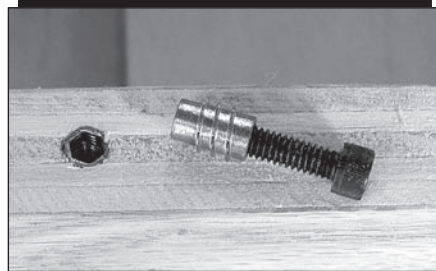
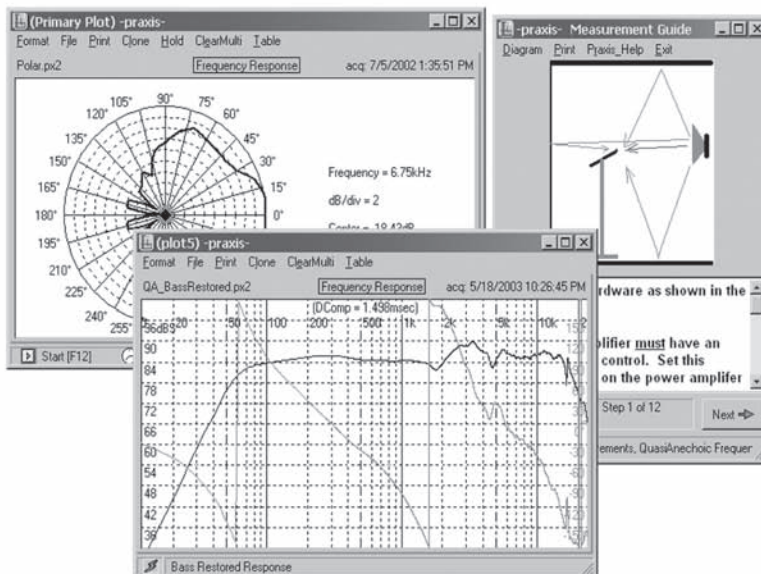


PHOTO 2: A threaded insert, Allen head bolt, and insert mounted in the box side.



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horizontal lines on it—one 3" from the top edge and one 1½" from the bottom edge. These form the axes for the holes for the woofers and tweeters,

respectively. The woofer holes are 4" diameter, 4" on center, the tweeter holes 1" diameter, 1¾" on center, as shown in Fig. 2.

PHOTO 3: Baffle rear view. Note the reduction of the baffle thickness where tweeters will be mounted, and chamfering of woofer holes.

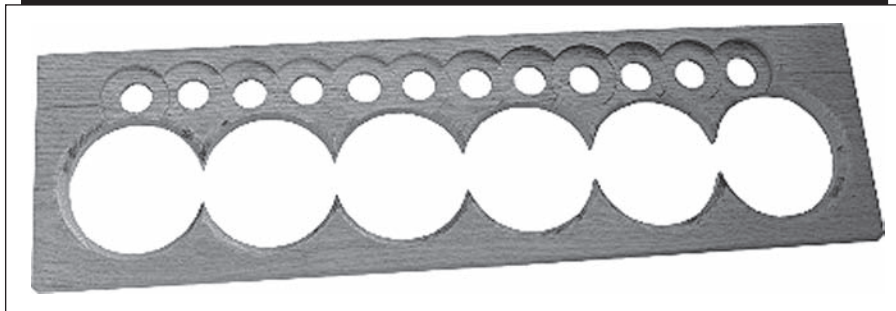


PHOTO 4: Assembling with biscuits



Drill the first tweeter hole with its center 3" from one end of the baffle, as is the first woofer hole. If you're building a pair of these as stereo mains, you should make them mirror-images of each other. Before drilling the tweeter holes, you need to reduce the thickness of the baffle where they are to be mounted by approximately 3/16". You can do this with a router or a drill-press-mounted Forstner or spade bit.

I used a 1½" diameter spade bit, leaving ¾" of material at the edge of the baffle undisturbed. Then drill out the 1" diameter tweeter holes and rout them from the front side with a ½" radius quarter-round router bit. Using the same router bit, from the inside of the baffle, put a chamfer on the woofer holes as well, without which it would be almost impossible to get at the woofer frames later on when they need to be sealed to the baffle (Photo 3).

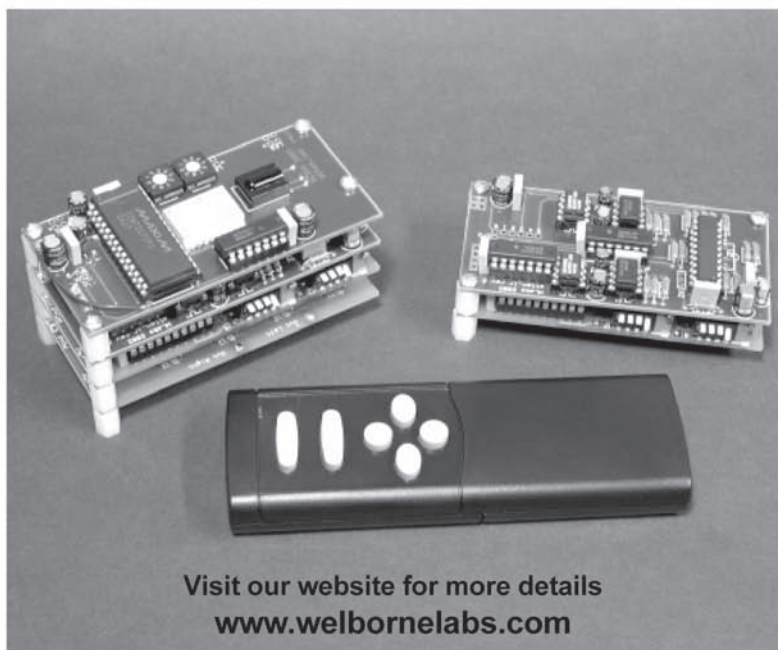
Now cut out the back, sides, and ends. Their dimensions depend on the thickness of the material they're made

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from. The back is the same nominal size as the baffle less the thickness of the top, bottom, and sides. Therefore, with $\frac{3}{4}$ " material the back will be 26" less 1.5" long, for a net of 24.5", and 7.25 less 1.5" wide, for a net of 5.75" wide.

Cut the back perhaps a sixteenth of an inch undersize in both dimensions, so that the top can be trimmed to the exact size of the box after assembly. You can make the back from whatever scraps you may have, even jointing it with a backer piece over the joint if required. Cut the ends to the exact width of the back by 5" deep, so in the

case of $\frac{3}{4}$ " material that would be 5" by 5.75". They attach to either end of the back by whatever means you prefer, be it glued and screwed, nailed and screwed, or even just glued. On my prototype I used a biscuit jointer, which gives very strong perfectly aligned joints with no visible fasteners.

Next come the sides, cut to be as long as the finished length of the back/ends assembly, and 5" wide. Attach these to the assembly via the same method as with the ends (Photo 4). After applying the second side,

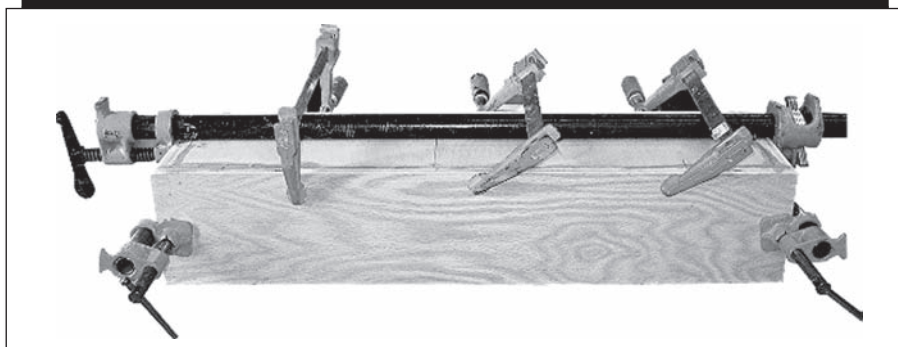
clamp the unit thoroughly while the glue sets (Photo 5).

ASSEMBLY

When the glue has set, use a couple clamps to hold the baffle in place. On its face, mark where the attaching bolts will pass through it, evenly spacing six to eight of them around the box perimeter. Using a drill bit the same diameter as the bolt, drill through the baffle and about 1" beyond into the box frame, preferably using a drill press to ensure the holes are perpendicular. Remove the baffle and re-drill the holes in the box frame, this time with a bit large enough to accept the inserts. Screw the inserts in place, driving them slightly below the wood surface.

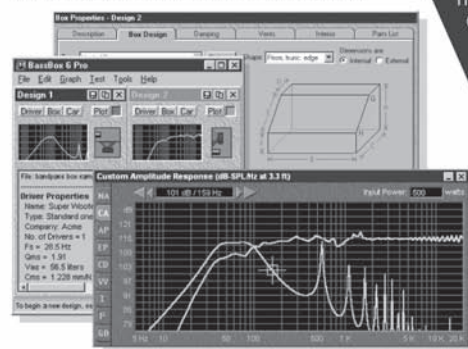
Coat their exterior with a few drops of epoxy before installation to better secure them, but don't use so much that it oozes and fills the threads; if that does occur chase them clear after the epoxy sets with a thread chaser/tap. Using a drill bit one size larger than the bolt head, countersink

PHOTO 5: Clamps, clamps, and more clamps.



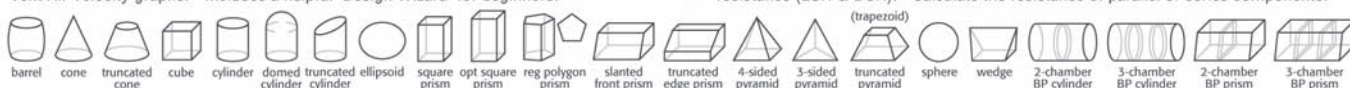
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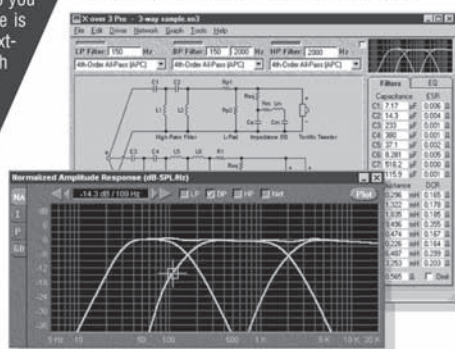
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PHOTO 6: Cutaway of a counter-sunk baffle bolt hole.



the holes in the baffle about halfway through, so that even after chamfering the baffle edge the bolt heads won't extend beyond the surface (**Photo 6**). Bolt the baffle to the box.

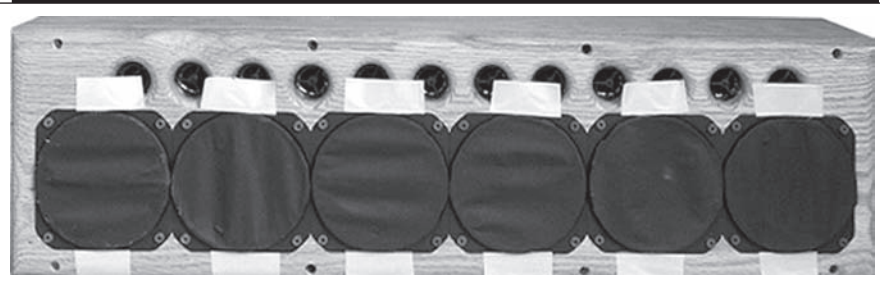
Use a router with a flush trimming bit to trim the baffle flush with the box; if you don't have one, use a sander instead. Likewise, use a quarter-round bit to chamfer the baffle edges to a $\frac{1}{2}$ " radius; otherwise, use a sander. In either case proceed to installing a port. Depending on how you plan to mount the cabinet, you can put the port in either end of the box or on the back.

The port is simply a 2" diameter hole; there is no duct required. If your port goes through $\frac{1}{2}$ " cabinet walls, a simple hole will do, but if it goes through $\frac{3}{4}$ " material, chamfer it on the outside to a $\frac{1}{2}$ " radius. Also decide on a location for a jack or binding posts, again depending on how you plan to mount it, and drill the necessary hole or holes in the box. Sand the box completely and apply your finish of choice. I did mine with Tung Oil, which looks very good on hardwoods and is as easy and foolproof a finish as you'll find.

Remove the baffle and install the drivers. The woofers install from the front, attached with $\frac{1}{2}$ " screws. The tweeters install from the back, glued in place with a hot-melt glue gun.

Install the tweeters first. Hold a tweeter in place, checking its alignment while looking through its hole in the baffle, and then tack it into position on the baffle rear with a couple of drops of glue, holding it steady until the glue cools enough to secure it. Then fully seal the tweeter frame to the baffle, leaving no gaps. Do each tweeter in kind, being sure to have the lugs aligned in the same direc-

PHOTO 7: Masking the woofer/baffle joint for sealing.



tion, and so that they don't hit either the woofers nor extend into the joint between the baffle and the box.

When you have done the tweeters, mount the woofers. You may wish to spray-paint the woofer frames first. If so, shield the cones with masking tape, leaving the tape in place when you're done to protect the cones until the cabinet is completed.

FINISHING TOUCHES

It's likely that there will be some gaps between the woofer frames and the baffle, so use masking tape to fully cover those gaps (**Photo 7**). From the inside of the cabinet, caulk the joints of the woofers and baffle and those of the woofers to each other as well. A small tube of almond tub and tile caulk works fine. After the caulk has set, remove the masking tape from the front of the baffle.

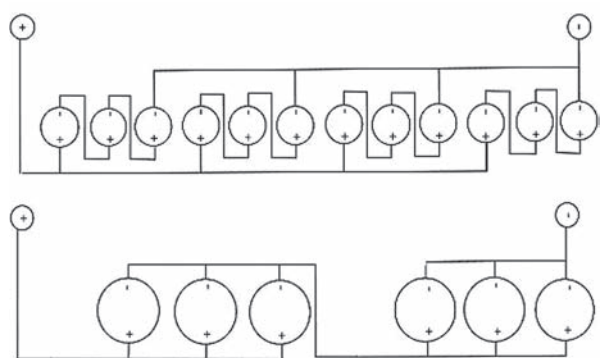
Wire the drivers, using polarized loud-speaker wire. Connectors are OK, but I prefer soldered joints, and that's much easier to do if you stay with 16 or even 18 gauge wire. There's not enough current flowing in this box to justify anything heavi-

er, especially with the tweeters.

The woofers are marked with a + on the positive lug, but the Onkyo tweeters are not. Instead you'll see a spot of blue paint on the magnet near the positive lug. The impedance of the woofers is 8Ω , so they are wired as two series-wired banks of three woofers in parallel, for a total 5.6Ω load.

The Onkyo tweeters are 6Ω each. They run a bit less sensitive than the woofers, so they're wired as four parallel-wired banks of three series tweeters, giving an impedance of 4.5Ω and matching the sensitivity of

FIGURE 3: Wiring schematic.



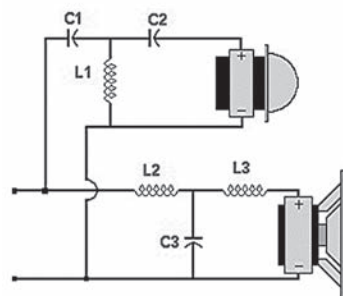
SB-2586-03

FIGURE 4: Crossover schematic.

3rd Order Butterworth

6000 Hertz

4.5 Ohm Tweeter / 5.3 Ohm Woofer



Parts List

Capacitors

C1 = 3.93 μ F
C2 = 11.79 μ F
C3 = 6.67 μ F

Inductors

L1 = 0.09 mH
L2 = 0.21 mH
L3 = 0.07 mH

SB-2586-04

the tweeter line to that of the woofer line (**Fig. 3**).

The crossover is a third-order at 6kHz (**Fig. 4**). When ordering your components, you can safely go within 10% of the listed values. C1 and C2 must be poly caps, but since C3 is merely a shunt you may use a less expensive electrolytic there if you wish.

None of the inductors are of a high enough value that their DCR is of much concern, so 18 to 20 gauge air core coils will do just fine. Using hot-melt glue, mount the crossover components in the box, keeping the coils well apart and oriented on different axes to minimize crosstalk (**Photo 8**). Wire the crossover input to your input jack/binding posts, and use a pair of fairly long wires to connect the outputs to the drivers.

Reflections inside the cabinet will cause ragged response, so loosely fill it with polyester pillow stuffing to damp them. Place the baffle over the box and check to be sure that no wires or driver lugs are caught in the joint between the baffle and the box.

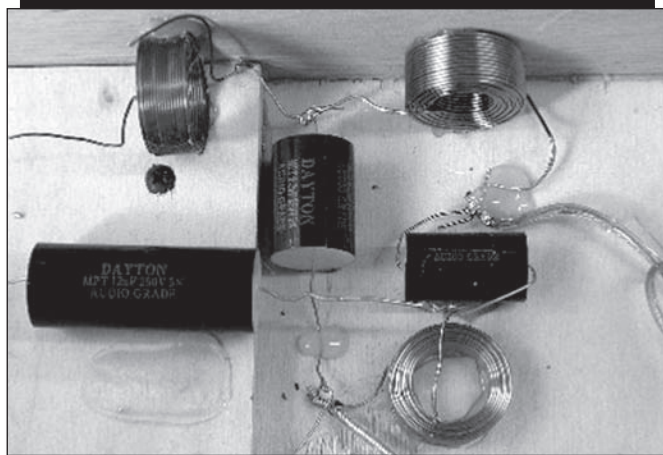
If the fit is not quite perfect, use a strip of thin foam weather strip to ensure an airtight connection. Bolt the baffle in place. It's done.

For center channel use, the f3 at 125Hz is quite adequate. For music or for HT mains, SLAs must be used with a sub, with a crossover between 100 and 125Hz for the most part. However, if your sub happens to be a Tuba 18 it has much higher sensitivity in the crossover area than the SLA, so you will find that setting the low-pass frequency on the sub amp at 80Hz or even less will give the best results.

The woofers I used have shielded magnets, obviously intended for center-channel duty, but the tweeters do not. As small as they are, the tweeter

magnets still are strong enough to disturb a CRT if placed too close, especially if you don't add bucking magnets. To help prevent this, place the tweeter line to the outboard side of the cabinet with reference to the TV screen. In my case no amount of fiddling prevented the speaker from interfering with the TV until I put it above the screen and laid it on its back, aimed upward. Interestingly enough, it sounds just fine. *ax*

PHOTO 8: The crossover.





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Junk Box Tube Amp Jubilee

This tube amp project celebrates the joy of producing a high-quality product for very little investment.

By Bruce W. Brown, RPh.,
and Larry DeRuyter

I recently finished a pair of complicated monoblocks ("70 Watts of McIntosh Power," 1/90 *Glass Audio*), a project that took almost a year to complete. Not only did it take a long time to find all the parts and do the layout and wiring, but the final testing and setup also dragged on. I had lost nearly all enthusiasm for the project before I finished.

Luckily, the final product turned out to be superb: powerful, quiet, and extremely musical. These are the best-sounding amplifiers I own! When my buddy Larry and I were auditioning them, we discussed the problem of running out of steam before completion of a major project.

A PROJECT TAKES SHAPE

He proceeded to pull out a couple of small monoblocks and hooked them up for me. (Larry and I spend lots of time discussing different types of tube and transistorized equipment. We frequently evaluate each other's projects and suggest improvements.) Larry tends to be very biased about his power amps. He prefers the unity-coupled McIntosh transformers and frequently restores McIntosh tube amps. He has not been impressed with the ultralinear Dynaco or Heathkit amps I have restored or built.

His small monoblocks sounded wonderful and impressive. I assumed they were McIntosh circuits, but was quite surprised when he told me they were

built using some Fisher 400 output transformers I had given him. He based these amplifiers on a circuit for a Pilot AA-410 amp that appeared in Feb. '02 *aX* in the Classics Circuitry section (p. 72). See **Fig. 1**.

His amps used 1½ 12AU7s, a 5Y3GT, and two 6L6s. I was surprised that such a simple circuit could sound so good and decided I needed to build a pair. Larry said he had experimented with a variety of preamp and phase inverter tubes and found that most common ones worked very well with few modifications.

My first inclination was to order some transformers, but the more I thought about it, the more it made sense to build these from what I had on hand and see how "non-critical" they really were to the final sound. I was fortunate enough to acquire an original Pilot AA-410 and after listening to it for a while, I decided to build a stereo amplifier (**Photo 1**). Because this is a "junk box" project, I will give several design options, and I hope you will find most of the parts already in your shop. I think you will be impressed with your efforts.

You're probably asking yourself, "Why are these guys building so many of these things?" The answer is simple: You can build them with a wide variety of different parts, the wiring is very simple, and they sound wonderful!

If you wish to build the amp as shown in the original schematic, you can order all the

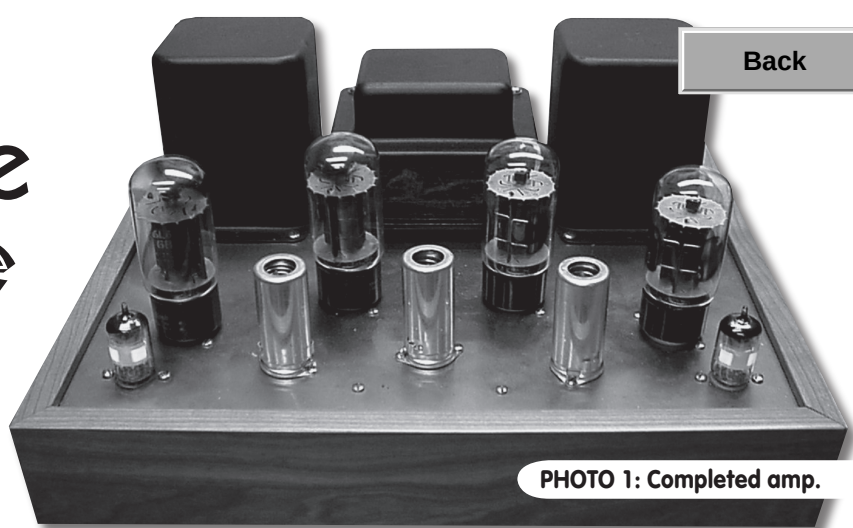
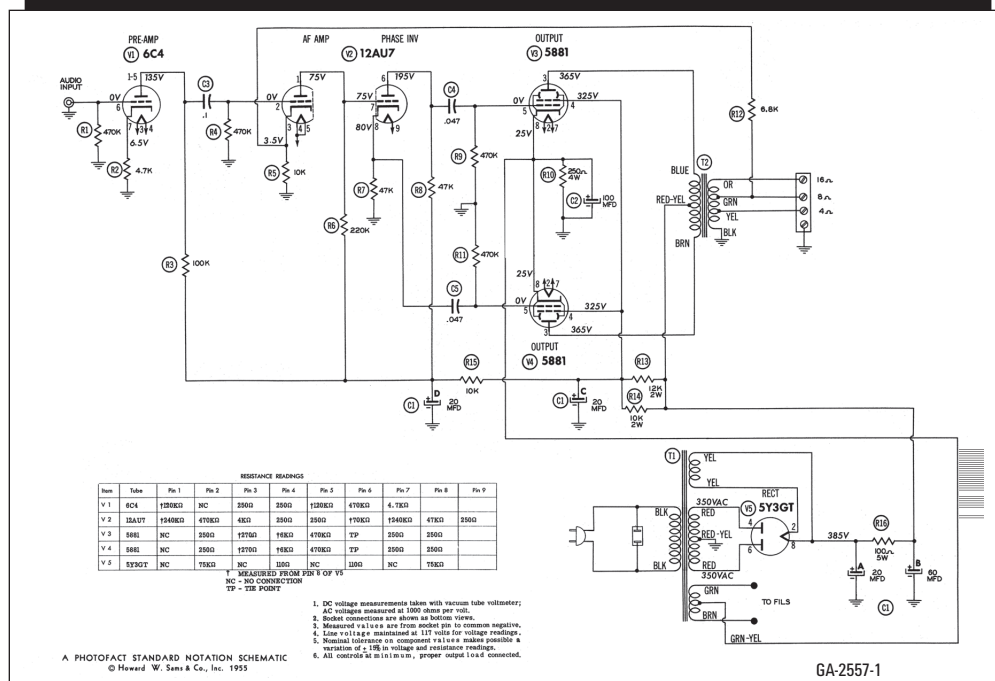


PHOTO 1: Completed amp.

FIGURE 1: Original Pilot AA-410 amp circuit.



parts from a variety of suppliers, or dig into your extra-parts box and see what you can come up with. You can use virtually any output transformer designed to be used with push-pull 6L6s, 5881s, 7591s, 6BQ5s, or even an EL34 transformer (although the heavier iron is more than you need). You can use ultralinear-type transformers and just tape off the extra taps.

I had a very nice pair of potted transformers from some '50s-vintage Heathkit A-9 mono integrated amps I had picked up for a reasonable price. You can probably locate some Heath, Fisher, Sherwood, or other manufacturer transformers from a junked unit or one with a bad power transformer. As I mentioned earlier, Larry used a pair that I salvaged from a Fisher 400 receiver with an open power transformer. The chassis was missing all the tubes and the face was gone when I picked it up, so it was not a candidate for restoration.

HEATH 9C OUTPUT AND FISHER POWER TRANSFORMERS

If you cannot locate a set of output transformers (**Photo 2**), you can order new ones, such as Hammond P-T1620s or P-T1650Fs, from Antique Electronic Supply (though this sort of defeats the purpose of the project!). The kind of power transformer(s) you select depends on whether you are going to build a set of monoblocks or a stereo amplifier (the route I chose for this article).

If you are building monoblocks, you will need a power transformer that will supply about 700V CT at about 125-150mA, double the current for a stereo

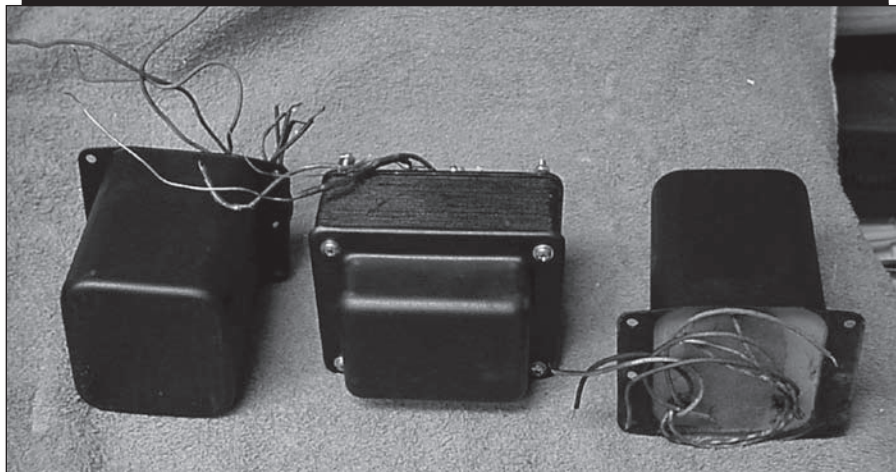
amp. It will also need to supply the filament voltages for the tubes you select as well as the rectifier tube, if you use one. Again, if nothing you have will work, you can use a new Hammond power transformer, such as P-T273CZ or P-T273BX. When you have selected your preamp, phase inverter, rectifier, and output tubes, find the transformer current needed from **Tables 1 and 2**.

You can use miniature, octal, or even loctal tubes. These medium-mu triodes are commonly used and readily available. The original design used a 6C4 as the AF amp, but you can use one-half of a 12AU7 (or if you are building a stereo amp use one-half for each channel). The second tube is a phase inverter; the original used a 12AU7, which is what I used, but feel free to select whatever medium-mu dual triode you prefer.

I chose three 12AU7s plus the four 6L6s for a total filament draw of 4.5A. I prefer to be conservative, so I tried for a 5A 6.3V tap. There are several options for the power-supply section (discussed later). If you use a tube rectifier, you will also need a 5V supply at the appropriate current. **Table 2** lists some appropriate tube rectifiers; you probably have one in your spare tube box.

I hope you have suitable transformers. If not, you need one that supplies around 700V center tapped, or 350V or 175V with no tap. The options allow use of any of these voltages. Make sure they supply enough current for your mono or stereo application. If you do not have the exact high voltage, you can tailor the power supply to deliver the voltages you need.

PHOTO 2: Sample transformers.



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TABLE 1*

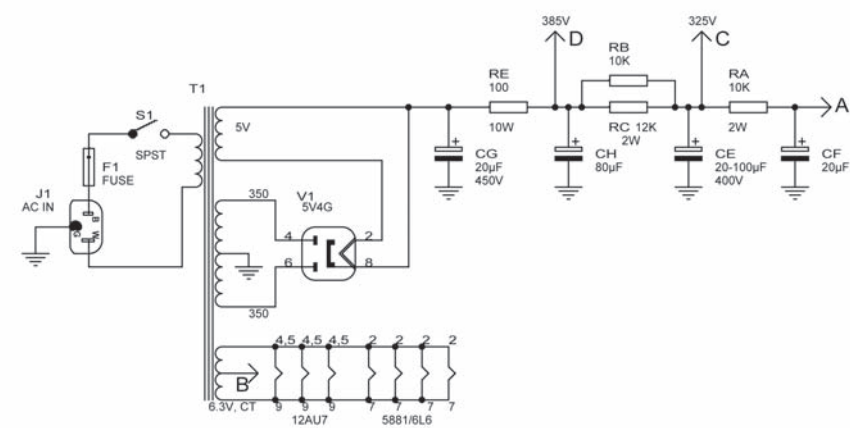
TUBES	Filament	Max Plate	Max Plate	Base
	Current	Current	Voltage	
6C4	0.15A	11mA	300V DC	miniature triode
6J5	0.3A	20mA	300V DC	octal triode
12AU7	0.3A	11mA	330V DC	miniature dual triode
6SN7	0.5A	9mA	300V DC	octal dual triode
7N7	0.8A	9mA	300V DC	loctal dual triode
6L6/G/B	0.9A	88mA	360V DC	beam power tube
6L6-GC	0.9A	116mA	450V DC	beam power tube
5881	0.9A	80mA	400V DC	beam power tube

TABLE 2*

Tube	Filament	Max Plate	Max Plate	Voltage Drop
	Current	Voltage	Current	
5AR4	1.9A	450V	250mA	
5Y3	2.0A	350V	125mA	-50 at 125mA
5Y4	2.0A	350V	150mA	-60 at 125mA
5U4	3.0A	450V	225mA	-44 at 225mA
5V4	2.0A	375V	175mA	-47 at 175mA
5R4	2.0A	750V	650mA	-67 at 250mA
5Z3	3.0A	350V	225mA	-58 at 225mA
5V3	3.0A	415V	1200mA	-46 at 350mA

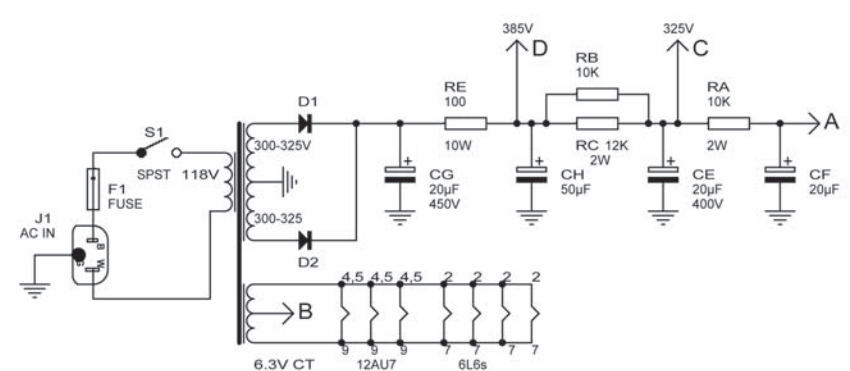
*RCA receiving tube manual RC-20/essential characteristics-GE

FIGURE 2: Power supply option 1.



G-2557-2

FIGURE 3: Power supply option 2.



G-2557-3

The original circuit (Fig. 1) specified a capacitance of 20mF, 80mF, followed by 20µF and 20µF. I used 20µF, then 100µF, and 50-100µF for the last two sections. If you have an assortment of transformers, it is a good idea to breadboard the power supply, and decide which combination works best. You only need to complete the power supply to the output transformer center tap, because if the voltage here is right, everything else will fall into place.

POWER SUPPLY OPTION 1

Figure 2 uses the stock type circuit and a transformer of 650-700VCT on the HV secondary. If you use a secondary close to 650 VCT, you need a 5AR4 rectifier, which is very efficient and will pass the most voltage. Likewise, if your transformer is above 725, use a 5R4, which will drop the voltage. I use 5R4s in Dynaco Mark 3s to lower the operating voltage of the power supply. These amps originally used 5AR4s with 525V electrolytics in the power supply and were manufactured when line voltage was typically 110V.

With today's line voltages of 120-125V, it is common to see Mark 3s with 550V on the electrolytic; this is not a healthy situation. By using a 5R4 instead of a 5AR4, I can drop the power-supply voltage under 500V. This preserves electrolytics as well as tubes and transformers.

Remember, the rectifier you choose must be able to handle the current your amplifier will draw. Do not try to use 5Ys or 5Vs for a stereo amp (Table 2). Unloaded, the voltage at D should be around 475-490V DC. To test your actual loaded voltage, you can temporarily connect a 5kΩ 10W resistor between point D and ground and measure the voltage at D to ground. If this reading is above 400V, you will need to increase the value of resistor RE.

You could also insert a choke, 7-15H, at the appropriate current rating instead of RE. If your voltage is really high, you can insert a choke before C6. This will drop voltage considerably, but the choke may have a mechanical hum. Ideally, you would like 385V DC loaded. If your voltage is lower than 360V, you can decrease the value of RE.

If your transformer has the center tap on the 6.3V winding, great; if not, you will need two 100Ω 2W resistors for the

biasing. If things are good at this point, you can skip to actual construction; if not, go on to Option 2.

POWER SUPPLY OPTION 2

If you have a transformer with 600 to 650 VCT, you can use semiconductor rectifiers instead of a tube (Fig. 3). Semiconductors are more efficient, so you will get higher DC voltages.

Is there a difference in sound between tube and silicon rectifiers? Yes, there sometimes is, but whether it justifies buying a new power transformer so that you can use a tube is completely up to you. Follow the same procedure as in option 1 for fine-tuning your HV supply.

POWER SUPPLY OPTION 3

Sounds strange, but you may have a toroidal transformer with two 175V secondaries or a transformer with a 315-350V untapped secondary. In this situation, you would use a full-wave bridge rectifier (Fig. 4). Use 600V 1A diodes and continue to fine-tune, as in options 1 and 2.

POWER SUPPLY OPTION 4

This option uses a voltage doubler and a transformer with a secondary of 150 to 175V (Fig. 5). I used this option because I had a really nice Fisher transformer with a 110 and 187V tapped secondary at 1A current (a little overkill for this amp, but it worked fine). It also had a 4.7A non-tapped 6.3V secondary as well as a 6.0V 1A secondary. I was able to use the 1A for the indicator tubes and the 4.7A winding for the 12AU7s and the four 6L6s.

You can use a variety of values for CA, CB, CC, and CD; I used 200 μ F -300V electrolytics. Remember that putting these in series is the equivalent of 100 μ F at 600V. My initial unloaded voltage, at point D, was over 500V DC, so I needed some drop.

I used 10H chokes instead of RE, and built duplicate high-voltage supplies from the voltage doublers circuit so I could use two small chokes, because I did not have one big enough to handle the total current. At first this seems excessive, but for the price of a couple resistors and a few electrolytics, I saved myself the cost of a single big piece of iron. This is the fun part of this project: making use of the parts you have on hand (Fig. 8).

You may need to increase or decrease

FIGURE 4: Power supply option 3.

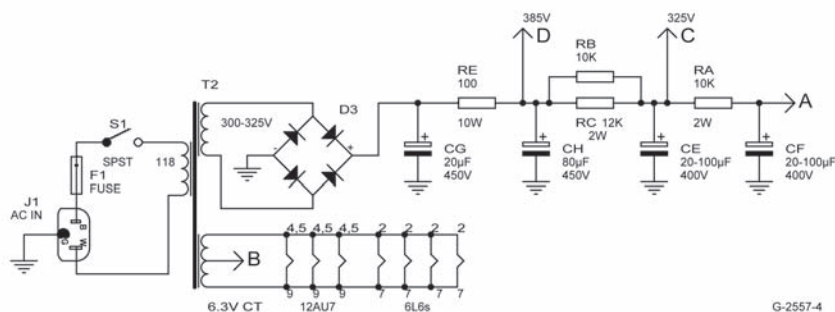
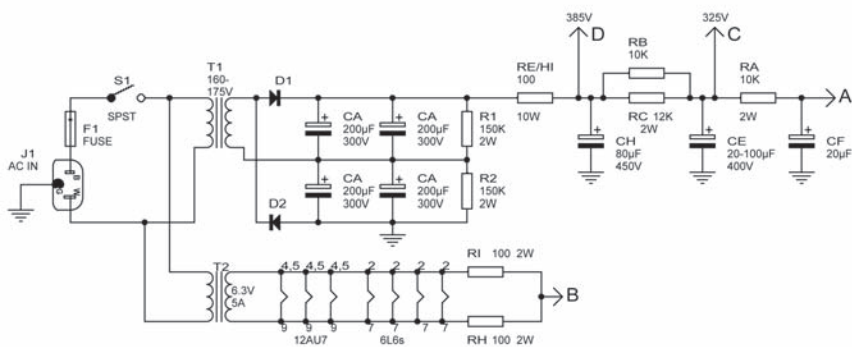


FIGURE 5: Power supply option 4.



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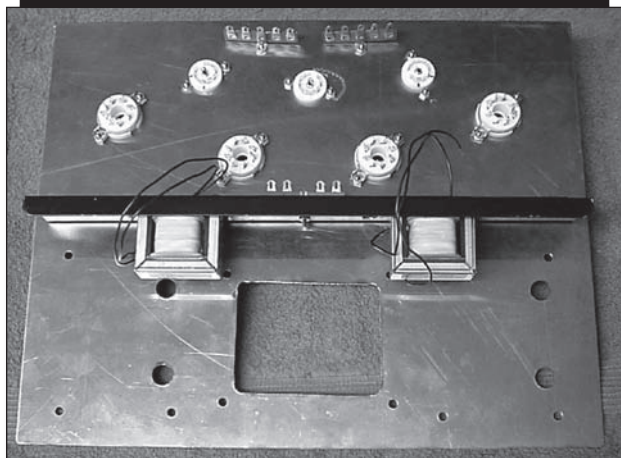
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PHOTO 3: Stainless-steel chassis machined (no level indicator sockets).



the value of RE/H1 to get close to the 385V loaded power-supply voltage. The original circuit uses self-biasing, and for this it uses the center tap of the 6L6 filament supply. Because my transformer did not have a center tap, I used two 100Ω 2W resistors, as shown in **Fig. 5**.

STEREO AMP

Now that you have selected your power transformer—and, I hope, your out-

put transformers—you need to prepare your chassis. Use a piece of cardboard to lay out your transformers and tube sockets. Try not to crowd everything in a small chassis, which can make wiring and service difficult. You can use a new commercial chassis, a used empty chassis, or you can roll your own.

I built my amp on a flat plate and then made a box to fit (**Photo 3**). Use mild steel, aluminum, copper, brass, or just about any metal heavy enough to support your iron. Larry provided me with a very nice piece of stainless steel. (Luckily I was able to convince him to do the machine work for me—stainless is nice to look at, but tough to drill and machine.) You need sharp drill bits, a good drill press, and a carbide die or Dremel™-type grinder to cut your holes. I recommend you measure

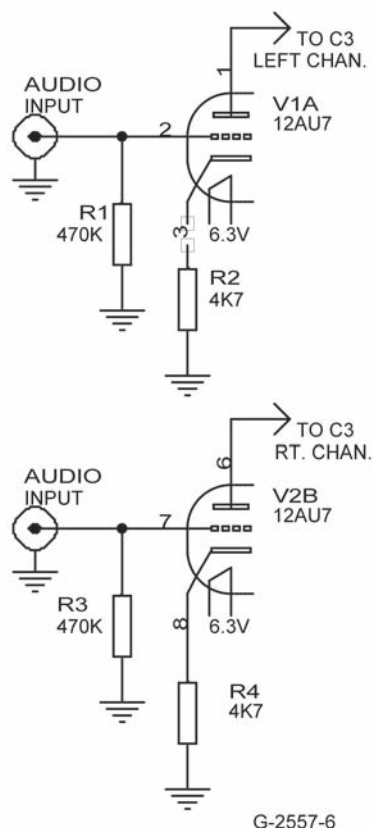
twice before you cut and drill!

I normally mount my chokes on top of the chassis, but Larry likes to place shielding in his amps, so he mounted my chokes on the shielding. Another option for this amp is to wire for audio level indicators. I have always liked the looks of tube level meters but have never built an amp with them, so I decided to add them to this one (**Fig. 7**).

You also need to consider input and output connectors and your options for level controls and input switches. I use level controls on my amplifiers because they allow me to use the amp directly with a CD player or tuner without the use of a preamp/control center. They also allow adjustment of the balance between the speakers (with monoblocks). You may choose to add an input selector with jacks for additional inputs.

You can build a simple box with butt-joined corners to hold the flat chassis, but I used ½" solid cherry, cut to the width that would allow me to recess the top and the bottom and still allow for clearance of the shield and construction. I

FIGURE 6: Stereo amp.

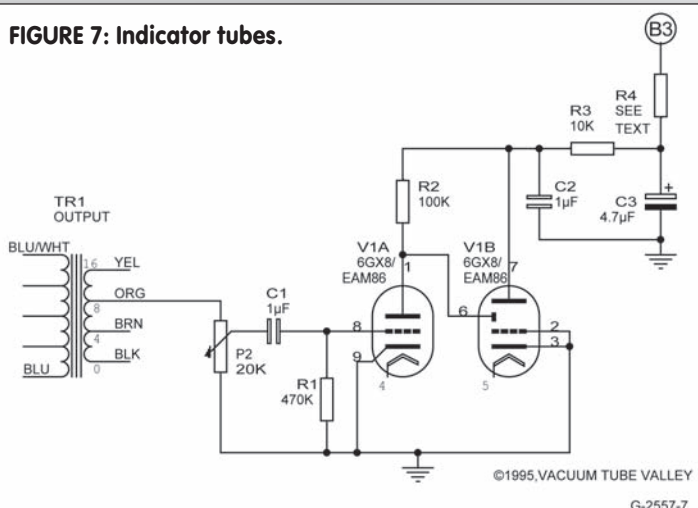


TUBE LEVEL METERS

Indicator tubes can add character to any tube amp. There are a variety of tubes still available and many different circuits for their use. I encourage you to research and locate the style you want to use before you punch holes in the chassis.

I include the circuit plan designed by Daniel Steele for a Stereo 70 (**Fig. 7**). His article was featured in *Vacuum Tube Valley* (Issue 3, Volume 1, Winter 1995). It uses 6GX8/EAM86 indicators. I was able to find several of them listed in online auctions and purchased them for less than \$10 each. In the schematic the input is connected to the 8Ω tap on the output transformer. The 20k pot is used to get the level of indication you desire for the volume set. Adjust the value of R4 to give you approximately 200V on terminal 7 tube socket. ■

FIGURE 7: Indicator tubes.



started with pieces longer than I was going to need and ran them through the tablesaw to cut a 1/4" deep slot for the chassis. Once you get the fit you like, glue the front and sides together. I used a picture frame clamp and an air brad gun to tack the front and sides together. I used screws to attach the back, which allow for disassembly to slide the chassis out. I also glued and nailed some blocks in the inside corners to strengthen the wood joints.

When the glue was dry, I removed the back and used a saber saw and router to cut an opening for another small stainless plate for the input jacks and level control. I used some surface-mount output strips, but you can use binding posts if you prefer. Once all the pieces fit, I sanded the cherry and applied three coats of polyurethane, sanding lightly between coats. Set the case aside and begin work on wiring the amp.

I prefer to wire the power supply first, and then work from the output transformers forward. Because I used a transformer with voltage doublers, my layout may be significantly different from yours. I had a large assortment of electrolytics, which I purchased from Electronic Goldmine. I recommend that you twist the filament wires together to minimize AC hum.

As I said earlier, the transformer I used did not have a center tap on the filament winding, and because it is necessary to use this to bias the output tubes, I used two 100Ω 2W resistors hooked as shown in **Fig. 4**. Once the power supply is wired, you can perform a power-up test to check voltages. You can also insert the tubes at this point to make sure the filament wiring is correct.

CAPS SELECTION

Before proceeding to the amp wiring, I would like to discuss the selection of coupling capacitors and resistors in the audio section. I use 1 and 2W carbon film or metal film resistors in audio circuits. They are stable and quiet, and I prefer having a little safety margin.

Today's metal film resistors are considerably smaller than carbon composition resistors of the same wattage, even though space is not an issue in this amp. (Some people swear by NOS carbon composition resistors; I generally swear at

them. They can be an elusive source of noise and frustration in a finished amp.) Use what works for you.

I use solid 18- or 20-gauge wire with high temperature insulation, trying to color-code my wiring using green for AC, black for grounding, red for high voltage, and yellow and white or blue for audio.

Coupling capacitors are another issue. I did not believe that capacitors sounded different until I started trying various types. You can hear a difference between a 40-year-old "Black Beauty" and a new film capacitor. I have tried all kinds of newer caps, and prefer Illinois™ film caps or Solen Fast™ caps. I have also had good luck with white "Wonder Caps" in triode amps. The Illinois types are very inexpensive, Solens are reasonable, and Wonders, a little more expensive. Larry prefers "Orange Drops." I would not recommend spending much money on premium caps.

I have found that this amplifier circuit is not very critical, and because there are only three coupling caps in the circuit, you can experience a savings here. I believe that capacitors tend to change a little as they "break in." After about 20 hours of use, film caps tend to mellow a little; I think they do sound better after a little use. Believe what you want and build it the way you want, it's your baby!

POWER-SUPPLY WIRING

It is time to wire the amplifier portion (**Photo 4**). As I mentioned earlier, I prefer to start at the output and work forward. If you cannot reach the sockets with the

leads, you can use pin 1 (which is not tied to tube elements) of the 6L6 sockets as a tie point for the 0.047 coupling caps that go to the phase inverter pins 6 and 8. I used a terminal strip to remotely mount resistors R3, R6, and R8 with wires running power to the input and phase inverter tubes.

Preventing ground loops is very important, and the best way to avoid them is to use a master ground bus wire. This bus wire runs from the input of the amplifier and is grounded to the chassis only at this point. Attach each resistor and capacitor ground to this wire. Once you have completed the amplifier section(s), check your wiring; or better yet, have someone else check your wiring. (Larry was kind enough to check mine.)

Photo 5 shows my completed stereo amplifier; as noted, your actual amplifier will vary from this, depending on the power supply you use.

There is a wrong way, an adequate way, and an optimal way to fire up a new amp for a first-time checkout:

1. **WRONG WAY**—Stuff in the tubes, hook up expensive speakers and an audio source, and turn on the power switch!
2. **ADEQUATE WAY**—Stuff in the tubes, hook up inexpensive speakers and an audio source, and hook up the power through an ordinary 100W household lamp and turn on power.
3. **RIGHT WAY**—Stuff in the tubes, hook up inexpensive speakers or dummy load, and the audio source, and plug the amplifier into a Variac. I like to use speakers instead of a dummy

PHOTO 4: Power-supply wiring.

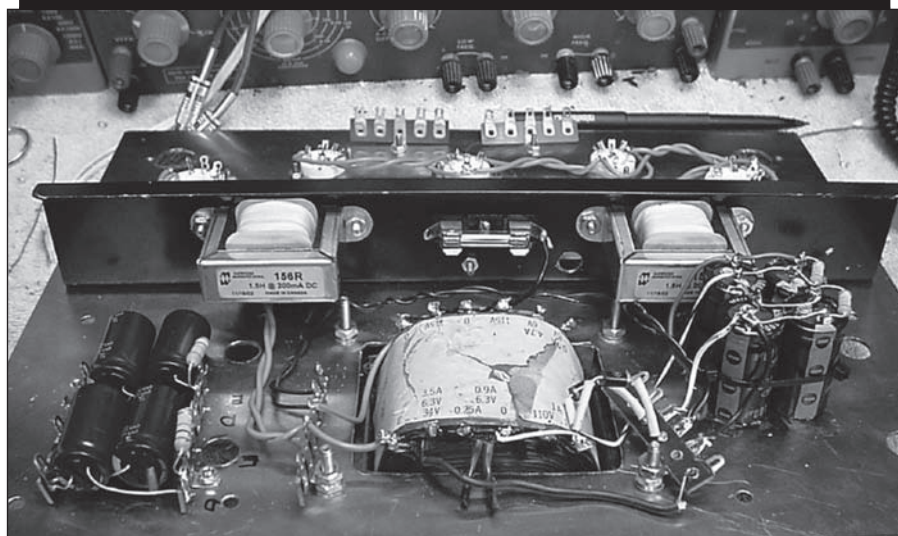
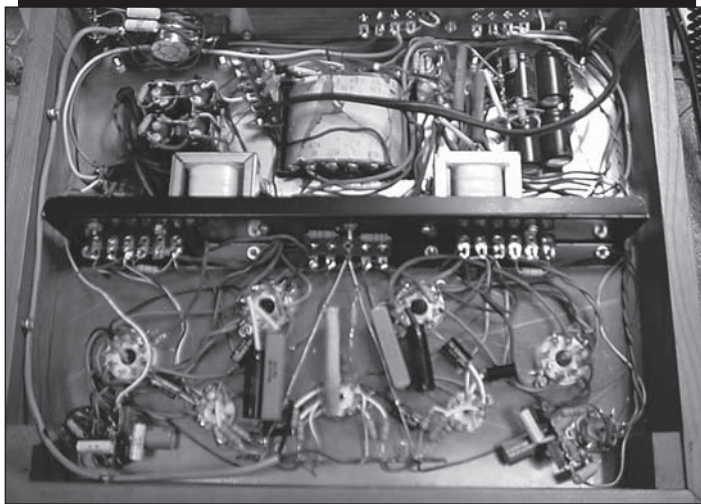


PHOTO 5: Completed stereo amplifier.



load, because sometimes the primary and secondary output windings of the output transformers will be out of phase, which will result in an obnoxious howl as you first fire up the amplifier. Attach your voltmeter to the power supply and bring up the voltage slowly. Watch and listen for signs of problems, such as burning, arching, popping, crackling, hum, feedback, or, worst of all, smoke.

I usually bring line voltage up to about

90-100V and check everything. I then reach 120 and check the power-supply voltages again. If the voltage at point D is between 375-395V, everything else should fall in line. If the voltages are close and none of the other "bummers" have occurred, slowly bring the volume level up. You should be rewarded with music! If the howl I mentioned earlier occurs, reverse the primary winding connections of the output transformers to the plates of the output tubes.

Some amplifier layouts, especially home-built units, can result in high-frequency oscillation that is difficult to isolate without the use of an oscilloscope. It usually occurs in the input high gain stage. Dynaco, Heath, and even Marantz

used a very simple solution that I sometimes find useful. If you are concerned about this issue, connect a 15k resistor in series with a 50-100mF ceramic disk or mica high voltage capacitor from the plate of the first input tube to ground. This does not affect the sound of the amplifier, but will prevent this high-frequency oscillation, which can be very damaging to the tweeters of your speakers.

At this point you can fine-tune the voltages. If the voltage at point D is low, decrease the value of the 10W resistor pair RE. If the voltage is high, you can increase the value. Once the voltage is right at D, the rest should fall in line. If they are not exactly where you want them, adjust the values of RA, RB+C, R8, R6, and R3. Most amps will tolerate a 10-20% voltage range.

After running the amp for 20 minutes, check voltages again and use the "hand touch test" on the transformers. If the power transformer is more than warm, you may have excessive current draw. Look carefully at the output tubes for hot spots.

If your filament current is too close to the maximum, you can shorten the life of the power transformer. Also, if you use a rectifier tube, make sure the transformer current rating is appropriate for the tube you are using. Use of a rectifier tube that draws 3A on a 2A winding will destroy the transformer after a short period of time.

Further tweaking can be helpful; Larry recommends tailoring the feedback resistor R12 and the bias resistor R10. By changing the value of R12 from 6.8k to 10k-12k, you can substantially increase the bass your amp produces. Larry also recommends increasing the value of R10 from 250Ω to 390Ω, which will lower the bias on the output tubes and extend their life without a significant change in sound or power output. These are great little amplifiers, and you should have fun experimenting.

Be aware of the high voltage current draw at heavy amplifier loads. My torture test involves using my JBL 200s at fairly high levels. I have noted some transistor amps perform perfectly in my shop, but do not do well with the JBLs.

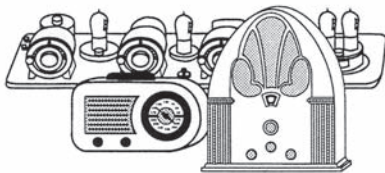
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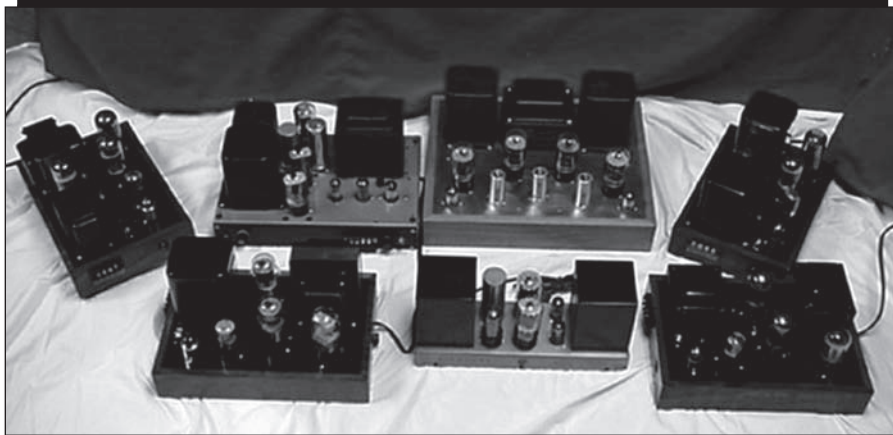
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PHOTO 6: Author's finished amplifiers.



mended minimum drive of 100W and will generally uncover weaknesses in amplifiers very quickly. I have found that the Pilot clones vary considerably in the "hand test" as described previously, and that changing the value of R10 cools down the power transformer accordingly.

Larry and I have each built a number of these amplifiers (**Photo 6**). The photo shows L to R rear: Larry's monoblock with Fisher Transformers, Bruce's monoblock built on Heath W 5 chassis with Heathkit A-9 output and Baldwin organ power transformer, Bruce's stereo Amp

from article, Bruce's monoblock with Knightkit integrated amp power and output transformers; front, L to R: Larry's monoblock with Heathkit 9A output and unknown power transformer, Bruce's original pilot AA 410, Larry's monoblock with Heathkit AA 100 output and power transformer. The long skinny one in the center is an original Pilot AA 410. As you can see, they are quite different in appearance and makeup, but they all perform very well. I hope that yours do, too. With a set of vintage speakers, such as Dynaco A25s, KLHs, or Advents and a good input source, you will enjoy the sound. *ax*

PARTS LIST (ONE INDICATOR-DOUBLE FOR STEREO)

1-20k linear pot or trimmer
1-470k ½W resistor
1-100k ½W resistor
1-10k ½W resistor
1-50-150k 1W resistor (see text)
1-1µF 450V film capacitor
1-0.1µF 450V film capacitor
1-4.7µF 450V electrolytic capacitor

PARTS LIST SINGLE MONOBLOCK PILOT 410

R1, R4, R9, R11- 470K ½W resistor (AES R-E470K)
R2- 4.7k ½W resistor (AES R-E4.7K)
R3- 100k ½ or 1W resistor (AES R-E100K)
R5- 10k ½ or 1W resistor (AES R-E10K)
R6- 220k ½ or 1W resistor (AES R-E220K)
R7, R8- 47k ½ or 1W resistor (AES R-E47K)
R10- 250-390Ω 5W resistor (SEE TEXT) (R-Q250-R-Q390)
R12- 6.8k-12k 2W resistor (SEE TEXT) (R-F6.8K to R-F12K)
R13- 12k 2W resistor (this is a duplicate from the power supply section) (R-F12K)
R14- 10k 2W resistor (this is a duplicate from the power supply section) (R-F10K)
R15- 10k 2W resistor (this is a duplicate from the power supply section) (R-F10K)
R16/H1- 100Ω 10W resistor (this is duplicated from power supply schematic, see text) (RR100)
C2-100-500µF 300V electrolytics (see text and select based on power supply)
C3- 0.1µF 630V film capacitor (AES Illinois C-TD1-630 or Solen C-FSD1-630)
C4, C5- 0.047µF 630V film capacitor (AES C-TD047-630 OR Solen C-FSD47-630)
V1- Preamp tube- medium mu triode (see text)
V2- AF amp/phase inverter tube medium mu dual triode (see text)
V3, V4- Output tubes-5881 or 6L6
V5- Rectifier 5Y3GT-see text (this is a duplicate from the power supply section-see options 1, 2, 3, or 4)
T1- Power transformer (see text; this is a duplicate from the power supply section)(AES P-T274AX)
T2- Output transformer (see text) (AES P-T1650G)

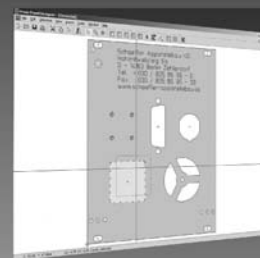
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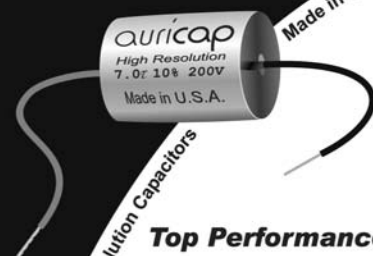


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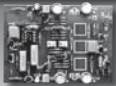


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A Phono PrePreamp for the (SA) CD Age, Part 2

By Norm Thagard

With design considerations and components selection covered in the first part of this series, we now proceed to the design specifics for the moving-coil preamplifier.

Figure 3 shows the moving-coil preamplifier schematic. Beginning at the input side of the circuit, note that the RCA input jack should be insulated from the chassis and the ground terminal of the jack bypassed to the chassis via a 10nF capacitor that is labeled C_{rf} . This avoids a potential ground loop while bypassing the ground terminal to chassis to suppress RF that might otherwise enter the circuit via the cable shield. C_{rf} can be a ceramic capacitor.

CIRCUIT COMPONENTS

The resistor labeled R_{in} sets the input resistance of the preamplifier. Its value should be equal to the load resistance recommended by the cartridge manufacturer for input into a "head amp." For the prototype, the value used was 100 Ω based on the cartridge manufacturer's recommendation. Some design-

ers provide an array of input resistors with DIP switch selection of the recommended value.

My approach was to solder two gold-plated pins removed from a good-quality DIP IC socket into the PC board. These pins readily accept the leads of a 1/4W precision resistor whose value is that recommended by the cartridge manufacturer. Should a new cartridge require a different resistor value, it is a simple matter to pull the old resistor from the pins and insert the new resistor.

I used no input capacitor, because I believe that moving-coil cartridges are relatively insensitive to capacitive loading. Neither Audio-Technica for their OC9/MLII nor Shelter for its 501, for example, specify a particular capacitance loading for these two cartridges.

R_{iso} was added based on a suggestion in *The Art of Electronics*¹⁹. It lets

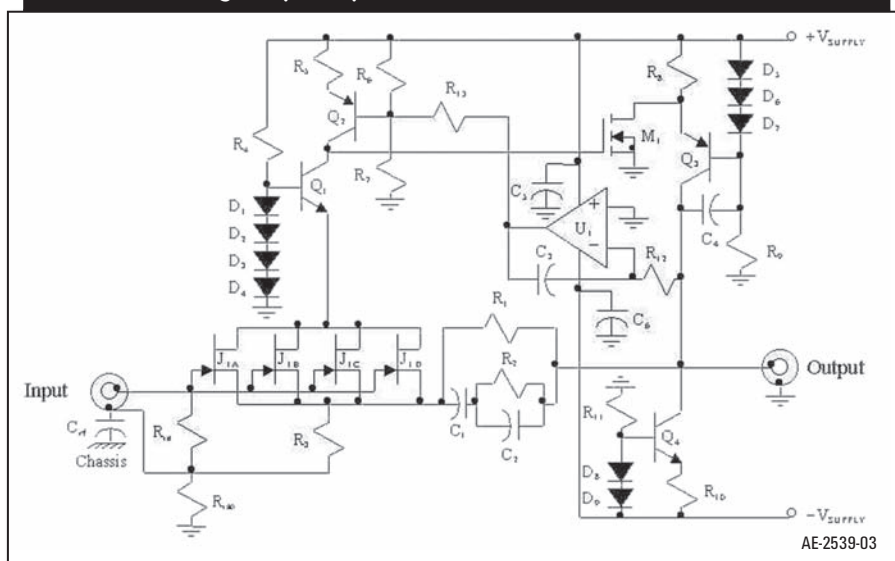
the phono cartridge ground lead set the potential at that node, reducing (you hope) the effect of noise on the preamp ground bus. Any noise coming in on that lead appears unamplified at the output while cartridge output is amplified by the rather large closed-loop gain of the phono preamp, yielding a "pseudo-differential input configuration." With one exception, I saw no objective change in preamp performance with or without R_{iso} , so if its presence offends a builder's sensibilities, by all means delete it.

A typical 2SK170 JFET will have about 45pF input capacitance, so four of them would present almost 200pF to the cartridge in the absence of Miller effect. $J_{1A} - J_{1D}$ constitute the common source (CS) portion of a cascode, which minimizes but does not, as some texts claim, eliminate Miller effect. Miller capacitance here would be $C_{Miller} = 4C_{gd}(1 - a)$, where $C_{gd} \cong 20$ pF is a typical value for gate-to-drain capacitance in a 2SK170 JFET. In this preamplifier, CS gain is $a = -g_{m-JFETs}/g_{m-Q3} < -1$ because the transconductance g_{m-Q1} of bipolar junction transistor (BJT) Q_1 is approximately 1S*, while the combined transconductance of $J_{1A} - J_{1D}$ is perhaps a little more than 150mS.

Although the 2SK170 has relatively high transconductance for a JFET, it is nonetheless significantly lower than a similarly operated BJT. The upshot is that C_{Miller} is slightly greater than $4C_{gd}$ or about 100pF. To this must be added $4C_{gs} \cong 100$ pF. Note that although the intrinsic gate-to-source capacitance is slightly greater than the intrinsic gate-to-drain capacitance, the latter potentially contributes more to the input capacitance of the preamplifier due to Miller effect.

Based on the reasoning in the previous two paragraphs, if a cartridge

FIGURE 3: Moving-coil preamplifier schematic.



requires significantly less than 200pF capacitive loading, then it would seem to be incompatible with this preamplifier. However, the input impedance is more complicated than this due to global negative feedback, which I have neglected due to its analytical complexity. You could sacrifice noise performance for reduced input capacitance by eliminating one or more of $J_{1A} - J_{1D}$, but noise performance is very important in the first stage of a gain-of-10,000 preamplifier.

If the cartridge requires more than 200pF, simply add an extrinsic capacitor C_{in} in parallel with R_{in} so that $C_{in} + 200\text{pF} = \text{recommended capacitive load}$. I hope that any capacitive loading requirement is a loose one because the previous calculations, while theoretically sound, are based on typical values for JFET parameters. Actual values can sometimes vary greatly from the typical.

PSPICE

In this preamp, quiescent drain current I_{DQ} for the input JFETs is set by the DC voltage drop across R_3 . Assuming matched JFETs, then $V_{GSQ} = -4I_{DQ}R_3$. The JFETs are therefore self-biased and will assume a value of I_{DQ} that is a function of their intrinsic parameters I_{DSS} and V_P according to the square-law equation $I_{DQ} = I_{DSS}(1 - 4I_{DQ}R_3/V_P)^2$.

I think it's difficult in the new millennium to be a serious designer without a working knowledge of PSpice. It might be illustrative, then, to look at how the simulation program deals with JFET bias. JFET models in PSpice directly use pinch-off voltage V_P , although it is called the threshold voltage and symbolized as V_{TO} . While I_{DSS} is not directly specified, it can be determined from $I_{DSS} = V_{TO}^2 \cdot \text{BETA}$, where BETA is a PSpice parameter with units of A/V^2 called the transconductance coefficient.

The PSpice model for the 2SK170 has $V_{TO} = V_P = -0.5024\text{ V}$ and $\text{BETA} = 0.05986\text{ A/V}^2$, so $I_{DSS} = V_{TO}^2 \cdot \text{BETA} = (0.5024)(0.05986) = 15.11\text{ mA}$. To complicate matters, the intrinsic JFET model with these values of V_P and I_{DSS} is augmented by a series source resistance $R_S = 4.151\Omega$ that will increase $|V_{GS}|$ by $I_D \times R_S$. This slightly increases the

complexity of the square-law equation, but even so JFET bias drain current in PSpice for the input stage can be predicted from **Equation 1**.

Numerically: $I_{DQ}^2 - 0.0702I_{DQ} + 0.000433 = 0$. From the quadratic formula in **Equation 2**, you can immediately rule out the former result because it is much greater than I_{DSS} . PSpice therefore should, if our calculations are correct, yield a drain current of 6.83mA. However, the simulation value was 6.86mA, a discrepancy that might be thought to be due to rounding error, but is not.

In real FETs, I_{DQ} will be increased above the square law value by a factor $1 + \lambda V_{DSQ}$, where λ is the channel-length modulation. PSpice accounts for this with the parameter LAMBDA, which for the 2SK170 is 0.001923 V^{-1} . The effect of LAMBDA is analogous to the Early effect in BJTs. In fact, Early voltage, designated by V_A , is analogous to $1/\lambda$. In the saturation region, drain current will increase linearly with increasing drain-source voltage, exactly as though there were an intrinsic resistor r_d from drain to source. This resistance, sometimes designated r_o in textbooks, is determined by triangle relationships on the $i_D - v_{DS}$ graph to be $1/(\lambda I_D)$, where I_D is the drain current calculated from the square-law equation neglecting the effect of channel modulation. In the present example, $r_d = 1/[(0.001923)(0.00683)] \approx 76\text{ k}\Omega$.

In general, intrinsic collector-emitter resistance in BJTs is lower than r_d . The result is that in their respective normal operating regions, FET $i_D - v_{DS}$ curves are "flatter" than BJT $i_C - v_{CE}$ curves, i.e., an FET more ideally, which is to say more closely behaves like a pure transconductance device (voltage-controlled current source).

Using $V_{DSQ} = 1.47\text{ V}$ from the simulation to adjust the above result gives $I_{DQ} = 6.85\text{ mA}$. The above result indicates that while addition of r_d to the JFET model improves its fidelity, drain current errors are minor if this resistance is omitted. There is also a drain ohmic (contact) resistance denoted RD which would change V_{DSQ} by a few tens of mV and affect the result even less than the omission of r_o .

In point of fact, the typical values



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in the manufacturer-provided PSpice model may be vastly different than those for the actual physical device the builder has in hand. In the case of 2SK170s, the manufacturer presorts them into three I_{DSS} groups. This is convenient for the builder since the parameter spread is reduced. Unfortunately, a given distributor may not have the group that would be wanted for a given design. I matched JFETs on a curve-tracer for $I_{DSS} = 12\text{mA}$ and $I_{DQ} \cong 6\text{mA}$ was observed for $J_{1A} - J_{1D}$.

On a closing note, it not only is possible to display PSpice device parameters, it is possible to alter them. Thus, knowing that my 2SK170 parameters differed from the PSpice model values, it was easy to edit the “instance” (simulation) values to correspond to the actual values in order to force the simulation to behave more like the real circuit.

This is as far as we’ll explore this topic here. I could not resist the temptation to expand the topic, but this author cannot—nor would the editor tolerate coverage of all of analog electronics in one article.

CIRCUIT OPERATION

I mentioned previously that the input stage topology is BiFET cascode with JFETs in the CS portion. If, as transistors are added in parallel, the total drain current is maintained constant, then g_m increases and hence improves voltage noise performance by the factor $\sqrt{N}$, where N is the number of paralleled devices.

No source resistors are used for $J_{1A} - J_{1D}$. While this avoids the degradation in noise performance that would accompany the resistors, it also removes the negative feedback that would tend to equalize drain currents among the four devices. Therefore, matching of JFETs is recommended and is not difficult to do, even without a curve tracer²⁰. After matching, I verified that in-circuit I_D between devices differed by no more than 10%, even after extended operation.

The only way to achieve the necessary open-loop gain is to use active loads. Q_2 and R_5 – R_7 form an active load for the first stage cascode. The R_6 – R_7 voltage divider sets Q_2 base voltage.

Q_2 emitter voltage will be about 0.7V higher and $R_5 = V_{R5}/I_{R5} \cong V_{R5}/(I_{DQ} - J_{1A} + I_{DQ} - J_{1B} + I_{DQ} - J_{1C} + I_{DQ} - J_{1D}) \cong (V_+ - V_{EQ-Q2})/(4I_{DQ} - J_{1A}) = \{[R_6/(R_6 + R_7)]V_+ - V_F\}/(4I_{DQ} - J_{1A})$, where $V_F = -V_{BEQ-Q2} \cong 0.7\text{V}$. The subscript “Q” in, e.g., V_{EQ} , implies that it is the “quiescent” or DC bias value of the parameter that is specified.

In the present case, $I_{R5} \cong 4I_{DQ} - J_{1A} \cong 24\text{mA}$, i.e., I_{R5} is determined by the combined drain current of the self-biased JFETs. Increasing R_5 will reduce shot noise from the active load via increased degeneration²¹ (local negative feedback) but will also simultaneously move Q_2 closer to saturation, a condition that will occur when emitter voltage falls to within 0.3V or so of the fixed collector voltage. Global negative feedback will set MOSFET M_1 ’s gate voltage as required, which in turn will fix Q_2 collector voltage to a level as high as 4V. If the value of R_5 is constrained to keep $V_{EQ-Q2} \geq (V_{CQ-Q2})_{\max} + 1\text{V} \cong 5\text{V}$, then $V_{CEQ-Q2} \geq 1\text{V}$, guaranteeing active region operation. This means that R_5 should not drop more than $V_+ - 5\text{V} = 13\text{V} \Rightarrow R_5 \leq [18 - (4.3 + 0.7)/0.024] = 542\Omega$.

If $V_{EQ-Q2} = 5\text{V}$, then $V_{BQ-Q2} \cong 5\text{V} - 0.7\text{V} = 4.3\text{V}$. The R_6 – R_7 voltage divider sets V_{BQ-Q2} according to Equation 3.

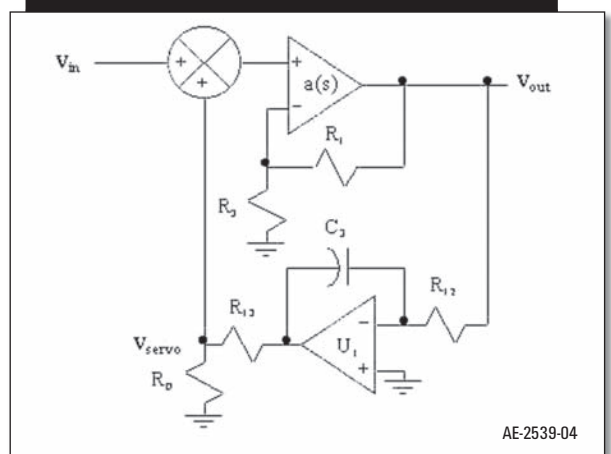
While this consideration dictates the R_6/R_7 ratio, it does not constrain their absolute values. Reducing the value of $R_6||R_7$ will reduce the active load noise contribution and, as you will see, lower the DC servo cutoff frequency, which is desirable. However, the “stiffer,” i.e., more constant, voltage division may make it impossible for the current-limited servo to adjust Q_2 base voltage sufficiently to maintain

the output quiescent voltage at zero. Another negative aspect of the stiffer voltage divider is increased power consumption. Reducing $R_6||R_7$ below approximately 350Ω gave negligible improvement in noise level and it was unnecessary to do so in order to maintain low-frequency RIAA curve tracking.

By adjusting base voltage of current-source transistor Q_2 , the inverting integrator of subcircuit $U_1C_3R_{12}R_{13}$ controls Q_2 conductivity. Since V_{EQ-Q2} is fixed, this means that the servo will adjust Q_2 collector voltage and hence M_1 drain current to that required to null quiescent output voltage.

This may sound a bit complicated, but a couple of points should help. First, drain and collector currents in active region transistors are to a first order approximation set by—and only by—gate-source and base-emitter voltages, respectively, according to a square-law equation for FETs and the exponential Ebers-Moll equation for BJTs. Second, avoid use of either of these equations in design work if at all possible. This is usually possible because the transistor equations must simultaneously be compatible with the circuit equations à la the old tube load lines. Simplest design is therefore accomplished by having the circuit equations dictate transistor behavior.

FIGURE 4: Simplified DC servo configuration.



AE-2539-04

Equation 1.

$$I_{DQ} = I_{DSS} \left(1 - \frac{(4R_3 + RS)I_{DQ}}{V_P} \right)^2 \Rightarrow I_{DQ}^2 + \left(\frac{2V_P}{4R_3 + RS} - \frac{V_P^2}{(4R_3 + RS)^2 I_{DSS}} \right) I_{DQ} + \frac{V_P^2}{(4R_3 + RS)^2} = 0.$$

Equation 2.

$$I_{DQ} = \frac{-(-0.0702) \pm \sqrt{(-0.0702)^2 - 4(1)(0.000433)}}{2} = 0.0351 \pm 0.5\sqrt{0.0032} = 0.0351 \pm 0.0283 = 63 \text{ mA or } 6.83 \text{ mA}.$$

I typically design using Kirchhoff's voltage and current laws in circuits that force desired transistor behavior. Negative feedback helps, too. In fact, use of active loads as is done here is only possible with either DC feedback or servo control. Otherwise, there is nothing to force the transistors to assume the proper bias conditions.

Finally, analyze the prospective behavior of your circuit under the assumption that the transistors are in their active regions unless a given transistor is intentionally to be in cutoff or saturation (ohmic or triode region in the case of FETs). Your analysis will lead to certain voltage and current conditions that should be compatible with active region operation. If these voltages and/or currents are not so compatible over the possible range of transistor parameters, locate the problem and try again.

DC SERVO DESIGN

I have designed several DC amplifiers. It would be nice if this phono preamp could respond to DC, but even a differential design would have problems maintaining the desired bias stability in this case. Perhaps some elaborate thermal compensation scheme would provide the necessary stability. Alternatively, an AC-coupled design could ease the problem, but this would place capacitors in the signal path, which seems to be a no-no in high-end designs. Also, the use of active loads requires DC feedback to maintain the bias point.

A DC servo offers an attractive method for bias point stability while avoiding coupling capacitors. As I will show, however, such a servo will pose design problems in its own right. The major problems are achieving a sufficiently low cutoff frequency so as not to affect low-frequency RIAA curve tracking while maintaining sufficient control range. Fortunately, these are surmountable as the current design proves.

But for an article in TAA²², I might never have begun using DC servos. The article offered without derivation the design equations. The phono preamp is noninverting as is the signal path from Q_2 base to the output. From this it follows that for application of

low-frequency negative feedback the appropriate servo configuration is that of a noninverting amplifier with an inverting integrator. I therefore offer the derivation for component values for the configuration in Fig. 4.

The schematic is overly simplified in that gain v_{out}/v_{in} is greater than gain v_{out}/v_{servo} . This occurs because inputs at the base of Q_2 and the gates of $J_{1A} - J_{1D}$ are amplified by what is shown in Equation 4, respectively, where R_L is the effective resistance at the node common to the collectors of Q_1 and Q_2 . It should be recognized that the formulae above are those for CE or CS stage gain in the presence of emitter or source degeneration, respectively. Even though the JFETs constitute the CS portion of

a BiFET cascode, the gain expression is identical to that for a CS stage with drain load resistance R_L .

Since gain from this node to the output is the same for either input, the ratio of open-loop gains between the two signal paths is:

$$\frac{a_{J_{1A}-J_{1D}}}{a_{Q_2}} = \frac{1/g_{m,Q_2} + R_5}{1/g_{m,J_{1A}-J_{1D}} + R_3}$$

Although transconductance of Q_2 has already been noted to be several times that of the combined transconductance of $J_{1A} - J_{1D}$, $R_5 \gg R_3$ nonetheless results in $v_{out}/v_{in} > v_{out}/v_{servo}$ as asserted. In fact, the ratio of gains is roughly 30 so local feedback provided by emitter degeneration resistor R_5 eases the difficulty of lowering the servo cutoff

Equation 3.

$$V_{BQ-Q_2} = \frac{R_7}{R_6 + R_7} V_+ = 4.3V \Rightarrow \frac{R_7}{R_6 + R_7} = \frac{4.3V}{18V} = 0.239 \Rightarrow R_7 = 0.314 R_6.$$

Equation 4.

$$a_{Q_2} = \frac{R_L}{1/g_{m,Q_2} + R_5} \text{ and } a_{J_{1A}-J_{1D}} = -\frac{R_L}{1/g_{m,J_{1A}-J_{1D}} + R_3},$$

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frequency. This need not be accounted for separately since the same effect could be obtained by reducing v_{servo} by a change in the voltage division provided by $R_D R_{13}$.

The **Fig. 3** schematic component designations have been retained for ease of understanding the servo configuration that is actually operative in this phono preamplifier. The summing junction provides negative feedback due to the sign reversal provided by the inverting integrator. R_D is the parallel combination of $(\beta_{Q2} + 1)(R_5 + r_{e-Q2})$, R_6 , and R_7 , which is approximately $R_6 || R_7$. The bottom terminal of the summing junction represents the Q_2 base. It is not necessary to know what $a(s)$ is, except that its magnitude at any frequency of interest should be and is very high.

Only R_1 and R_3 are shown in the preamp's feedback loop because the servo, if properly designed, will be unable to respond to audio frequency stimulus. Thus, we are interested only in the low-frequency gain $A_0 = 1 + R_1/R_3 \approx 10,000$.

Noninverting terminal voltage is the output from the summing junction, which is:

$$v_+ = v_{in} + [R_D/(R_D + R_{13})]v_{int}$$

where $v_{int} = -[1/(sR_{12}C_3)]v_{out}$ is the output voltage of the integrator. In turn: $v_{out} = (1 + R_1/R_3)v_+ \Rightarrow v_+ = v_{out}/(1 + R_1/R_3)$.

Therefore: see **Equation 5**.

Grouping terms: see **Equation 6**.

Substituting $j\omega$ for s in the above expression, the gain v_{out}/v_{in} is found to be that shown in **Equation 7**.

Gain magnitude is $|A_V(jf)| = (1 + R_1/R_3)/[1 + (f_0/f)^2]^{1/2}$. From this, it follows that $|A_V(j0)| = 0$, $|A_V(jf_0)| = (1 + R_1/R_3)/(2^{1/2})$, and for $f \gg f_0$, $|A_V(jf)| \approx 1$

$$+ R_1/R_3 = A_0.$$

These results imply the existence of a zero at DC and a pole at f_0 . Of course, $|A_V(jf)|$ actually tracks the RIAA curve, so these results merely manifest the simplifying assumption made at the outset. The integrator will have no noticeable effect on circuit operation for frequencies much greater than f_0 . To preclude deviations from the RIAA curve due to the servo, it is necessary to ensure that $f_0 \ll 20\text{Hz}$.

GOOD NEWS, BAD NEWS

The gain equation describes a high-pass filter with cutoff frequency f_0 , so the servo does not allow the preamplifier to respond to DC. Although the gain expression may, at first glance, look a little complex, closer examination reveals that the integrator time constant $R_{12}C_3$ will be divided by DC closed-loop gain $1 + R_1/R_3$ and by the voltage divider ratio $R_D/(R_D + R_{13})$ at the integrator output.

The integrator time constant should be as large as practical. To avoid use of electrolytic capacitors, it seems reasonable to restrict the value of C_3 to no more than $10\mu\text{F}$. Op amp U_1 is a JFET-input op amp that should work well if $R_{12} = 1\text{M}\Omega$. The fundamental cutoff frequency due to the integrator would be $1/(2\pi R_{12}C_3) = 0.016\text{Hz}$. This is great, but the bad news is that it is multiplied by the low-frequency closed-loop gain of the preamplifier, which is $A_0 = 10,000$.

The good news is that the voltage divider formed by $R_D R_{13}$ can potentially salvage the situation. However, there are confounding considerations for this divider. Resistance R_D must be as small as possible, but smaller values increase power consumption and lower the offset range over which U_1 can maintain null output. There is an upper limit on resistance values as well if Q_2 base

bias current requirements are to be met. As mentioned, local feedback from R_5 has the same effect as lowering the voltage divider ratio, which eases the constraint on the divider. Ain't electronics wonderful!

Due to this bonus effect of R_5 , the time constant $1/(2\pi R_{12}C_3)$ can be ten times greater without compromising the low-frequency cutoff. The obvious thing to do, then, is to choose a smaller value for C_3 because a good-quality $1\mu\text{F}$ capacitor is much smaller, cheaper, and more widely available than one of $10\mu\text{F}$ value. The fact that the preamp tracks the RIAA curve with 0.1dB accuracy down to 20Hz with output offset voltage of approximately 1mV proves that the servo does its job.

The values used in the $R_6 R_7$ voltage divider worked fine for JFETs with $I_{DSS} = 12\text{mA}$. Should I_{DSS} be significantly greater than this, then the divider ratio should be modified to lower Q_2 base voltage. Of course, if I_{DSS} is significantly less than 12mA, then base voltage should be increased. Alternatively, R_5 could be reduced in the former or increased in the latter case. A decrease in the value of R_5 could, however, affect RIAA tracking at low frequency.

As a matter of interest, you could use a zener diode in lieu of R_6 or R_7 , taking advantage of the diode's low dynamic impedance, which can be about 2Ω for a 1N4735 operated at 41mA ²³. The disadvantage is that the ability of the servo to null output would be restricted to an impracticably small offset range. Placing a small resistor in series with the diode would loosen this restriction and reduce the noise contribution of the zener (zeners can be very noisy) but would also partially nullify the benefit of using the zener in the first place. Even with the resistive divider, it may be necessary to alter either the value of R_5 or the R_6/R_7 ratio should the JFET drain currents differ significantly from those obtained in the prototypes, as mentioned in the previous paragraph. In fact, the values offered here seemed to be variable over quite a range while still giving normal operation.

As a final note, you could add a capacitor shunting R_6 to reduce noise at middle and high audio frequencies. Perhaps it should be added to increase

Equation 5.

$$\frac{v_{out}}{1 + R_1/R_3} = \frac{R_D}{R_D + R_{13}} \frac{-1}{sR_{12}C_3} v_{out} + v_{in} \Rightarrow v_{out} = -\frac{R_D}{R_D + R_{13}} \frac{1 + R_1/R_3}{sR_{12}C_3} v_{out} + (1 + R_1/R_3)v_{in}$$

Equation 6.

$$v_{out} \left(1 + \frac{R_D}{R_D + R_{13}} \frac{1 + R_1/R_3}{sR_{12}C_3} \right) = v_{in} (1 + R_1/R_3)$$

Equation 7.

$$A_V = \frac{v_{out}}{v_{in}} = \frac{1 + R_1/R_3}{1 + \frac{R_D}{R_D + R_{13}} \frac{1 + R_1/R_3}{j\omega R_{12}C_3}} = \frac{1 + R_1/R_3}{1 - jf_0/f}, \text{ where } f_0 = \frac{1}{2\pi R_{12}C_3} \left[(1 + R_1/R_3) \frac{R_D}{R_D + R_{13}} \right]$$

power supply rejection, because variations in the positive power supply are otherwise amplified. Unfortunately, the capacitor would provide little benefit at low frequency where JFET flicker noise dominates and will produce instability in servo operation unless a zero is introduced into the integrator²⁴. I chose the alternative of a fairly elaborately filtered and regulated low-noise positive power supply.

CONSTANT-CURRENT SOURCES

There are two constant-current sources in this preamplifier, both serving as active loads. As we just saw, the current source formed by Q_2 serves the additional role of completing the DC servo loop. Its emitter resistor R_5 enhances performance by increasing the Norton-equivalent resistance of the source from r_{o-Q_2} , which is the intrinsic collector-emitter resistance of Q_2 , to $(1 + g_{m-Q_2}R_5) r_{o-Q_2}$. Current-source current will be $\{[R_6/(R_6 + R_7)]V_+ + V_{BE-Q_2}\}/R_5 \cong \{[1,000/(1,000 + 866)]18 + (-0.7)\}/392 = 23\text{mA}$, anticipating that $J_{1A} - J_{1D}$ will require a total of 20–25mA. As discussed previously, current-source current will be modulated by the servo to set Q_3 emitter current at the level needed to maintain output offset voltage near zero.

The second stage is a folded-cascode whose current source/active load consists of $Q_4D_8D_9R_{10}R_{11}$. It will command Q_4 collector current of $(V_{D8} + V_{D9} - V_{BE-Q_4})/R_{10} \cong 0.7/18.2 = 38.5\text{mA}$. While this current command does not directly set Q_3 collector current, it will indirectly do so. Always keep in mind that active region transistor collector or drain current is set by base-emitter or gate-source voltage, respectively.

DC feedback via the servo is the key as it sets M_1 gate voltage and hence MOSFET quiescent drain current to that value necessary to null output quiescent voltage. It accomplishes this by adjusting the emitter voltage of Q_3 through the resulting R_8 voltage drop. This voltage drop is produced by current through R_8 , which will be $(V_{D5} + V_{D6} + V_{D7} + V_{BE-Q_4})/R_8 \cong 1.4/15.0 = 93\text{mA}$, of which about half will flow into the emitter of Q_3 and half into the drain of M_1 . These bias currents were set, by the way, on the basis of the minimum

required to avoid slew-rate limiting at full output from 20Hz–20kHz.

This is an interesting proposition. You can determine Q_3 collector current using the $Q_4D_8D_9R_{10}R_{11}$ current source and the quiescent M_1 drain current will, ignoring Q_3 base current, be $1.4V/R_8 - I_{CQ-Q_3}$, i.e., M_1 quiescent drain current is set by the value of R_8 . The servo-controlled first-stage active load takes care of the rest, producing both the desired second-stage bias currents and near-zero quiescent output voltage. What a bargain!

There is some loss of signal current in R_8 , a loss that could be eliminated with a current source. Little difference in performance was seen on the breadboard. Anyway, as long as the impedance “looking” into the emitter of Q_3 is significantly less than the value of R_8 , the signal current loss is inconsequential. Obviously, a resistor is cheaper and occupies less board space than a constant-current source.

Without the use of active loads for Q_1 and Q_3 , the preamplifier could not meet the low-frequency open-loop gain requirement, $a(jf) > 100\text{dB}$. Of course, the advantage of an active load is lost unless impedance at the output node is high, i.e., provided the input impedance of the following stage is high. At the output of the first stage, impedance remains high due to the high impedance “looking” into the M_1 gate. The second stage must drive the input impedance of a control preamplifier or perhaps a potentiometer. Such an impedance is typically tens of kilohms at a minimum.

To be sure this design met reasonable interface requirements, testing was done driving a 10kΩ resistor. Voltage sampling at the output reduces output resistance by the factor $1 + \beta a$. The net effect is that little fall in output voltage is seen over the audio band when the output is shunted by 10kΩ, indicating that output resistance is significantly less than this.

For those who think they hear degradation in sound reproduction due to active loads, I would say that the alternative is added gain stages, which no one likes. In addition to setting current-source current, R_{10} raises the Norton-equivalent resistance “looking” into

the collector of Q_4 . This latter effect potentially boosts gain because no-load second-stage gain is approximately $-g_{m-M_1}R_{\text{Norton}}$.

Next month, in the final part of this series, we'll examine power-supply considerations. *aX*

* S = Siemens, new name for mu (μ).

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PARTS LIST

Parts are specified for one channel only, except as noted.

PREAMPLIFIER

Resistors are 1%, metal film, 0.25W resistors unless otherwise specified. Values are in ohms.

R_{in} — See text
 R_{iso} — 10.0
 R_1 — 49.9K
 R_2 — 4.02K
 R_3 — 5.00 (two paralleled 10Ω resistors)
 R_4, R_9, R_{11} — 4.99K
 R_5 — 392
 R_6 — 600
 R_7 — 1.00K
 R_8 — 15.0, 1W
 R_{10} — 18.2
 R_{12} — 1.00M
 R_{13} — 20K

Capacitors are 2%, 50V polypropylene film capacitors unless otherwise specified. Values are in farads.

C_{if} — 10n, 100V ceramic disc
 C_1 — 56n
 C_2 — 20n (formed by paralleling two 10nF capacitors)
 C_3 — 1μ
 C_4 — 390p
 C_5, C_6 — 100n, 25V ceramic disc

Diodes

D_1 – D_9 are 1N4148 silicon signal diodes.

Transistors

J_{1A} – J_{1D} — 2SK170 JFETs
 M_1 — IRF511 MOSFET
 Q_1 — BC549C BJT
 Q_2 — BC559C BJT
 Q_3 — TIP30C BJT
 Q_4 — TIP29C BJT

ICs

U_1 — LF411A

NAD L53 DVD/CD Stereo Receiver

By Charles Hansen

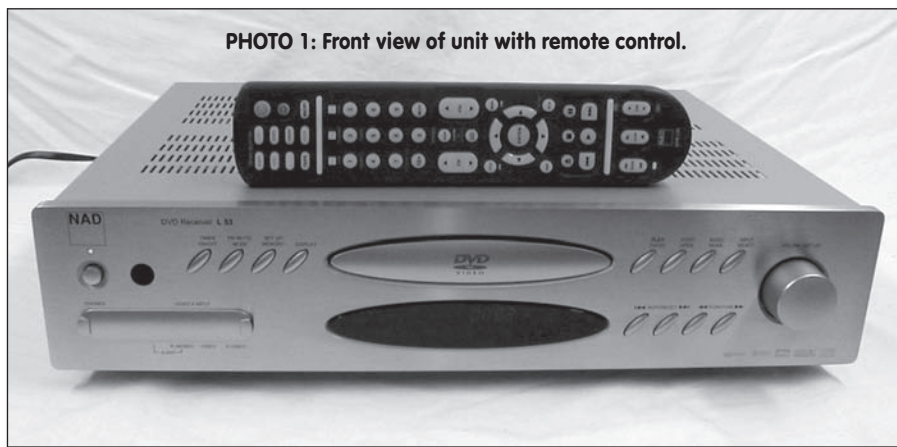
The L53 is a 50W per channel stereo AM/FM receiver that also includes a CD/DVD player. The only other gear required are stereo speakers and a TV monitor. You can dispense with the monitor if you only want to listen to audio. The L53 includes a learning remote with illuminated keys that you can program to replace your TV, and/or cable and satellite remotes, and you can even program it to activate your entire system with only two keystrokes. A clock/timer function is included in the front panel display.

The L53 uses the latest generation DVD-video decoder with progressive scan output to a digital TV display through its component video connections. Composite and S-video connectors are provided as well, although an RF (modulator) output is not available. If your TV is an older model without A/V input jacks, you can always use a VCR as the RF modulator. The CD/DVD drive will play DVD-V, VCD, SVCD, CD, CD-R, CD-RW, MP-3, JPEG, and WMA data. The L53 will not play DVD-Audio or SACD high-resolution audio formats.

The L53 features Sound Retrieval System (SRS) technology to provide a surround sound simulation from only two speakers in rooms that cannot accommodate additional surround speakers. A subwoofer output that tracks the stereo level control allows full compatibility with any of the many powered subwoofers currently available.

The stereo power amplifier section is a discrete transistor design, and all the power supplies use conventional iron transformers and linear regulators rather than the more common computer-style switching power supplies. As such, the L53 is a hefty 18.5 lb.

PHOTO 1: Front view of unit with remote control.



INSIDE THE AMPLIFIER

Photo 1 shows the front panel of the L53 along with its HTR-L53 remote control. The power switch is located on the left side just below the blue LED standby indicator. Pressing the drive open/close button will also turn the unit on. After a startup delay, the standby indicator will go out and the main display will come on. There is no off position, so when the power switch is cycled off the unit remains in standby mode. The remote control infrared sensor is to the right of the power switch.

The access door just below the power switch covers a 1/4" headphone jack and the Video 4 composite and S-video input jacks. When headphones are plugged in, the speaker outputs are muted.

To the left of the drive tray are four slanted buttons that control the clock/timer, the FM mute/mode button, the setup/memory button (used to store up to 60 station presets or access the DVD

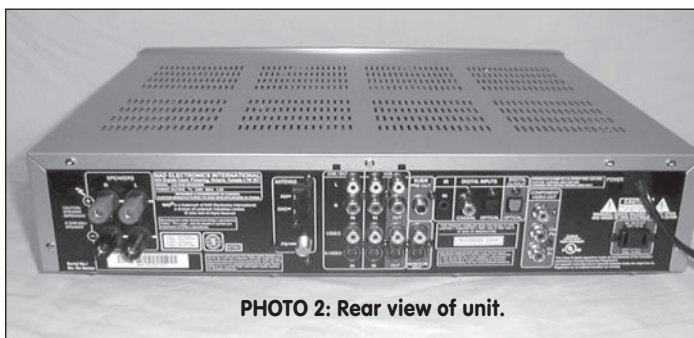
options), and the display format button. The blue vacuum fluorescent main display is located just below the drive tray.

To the right of the drive tray are the play/pause button, the drive stop/open button, the audio mode select, and the input select. The four buttons below them control the forward/back/skip/scan/tune functions. The large knob on the right is the volume control, which is not a dual analog pot, but a servo control that also actuates the bass, treble, and balance controls in conjunction with the setup menus.

The HTR-L53 is an eight-device illuminated learning remote control with macro function. The manual is a bit thin with regard to all its features, but it seems straightforward enough, and the lighted color-coded buttons are a welcome respite from the sea of black-buttons-on-black-background-with-tiny-printing all too common in remotes these days.

The L53 is constructed of heavy gauge gray painted steel, with cooling slots on the top and bottom. The front plate is 10mm-thick aluminum rather than the plastic used on many of the consumer A/V products. Four thick hemispherical rubber mounting feet provide ample finger room for

PHOTO 2: Rear view of unit.



lifting the L53.

Facing the rear panel (**Photo 2**), the left side has two pairs of speaker binding posts. In accordance with CE requirements, these are not on $\frac{3}{4}$ " USA spacings, so you cannot use dual banana plugs. Each threaded post has a chassis-mounted plastic guide bushing that accepts heavy gauge wire. The manual says speakers should be rated at 8Ω minimum.

Next in line are the clip terminals for the supplied AM loop antenna and a 75Ω F-connector for the FM input. A 300Ω dipole is supplied along with a matching 300Ω : 75Ω balun.

Composite and S-video input connectors are supplied for a cable/satellite box and a VCR. There are also composite and S-video output jacks that you can use to send A/V signals to a VCR or DVD recorder. Next to the A/V jacks are the subwoofer output jack and the composite and S-video monitor out-

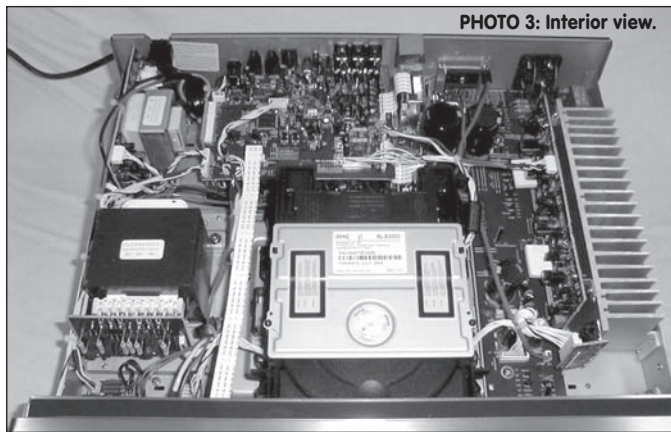
put connectors. All the RCA jacks have gold-plated shells, but the center connectors appear to be nickel-plated.

To the right of all these connectors are a Xantech-style infrared repeater input, S/PDIF coax and optical PCM digital inputs, and an optical PCM output jack. Just below these connectors are SCART/RGB connectors for the European PAL television signals. Finally, the three vertical connectors on the right side of the connector stack supply component video Y/Cb/Cr signals to video monitors that accept them.

The polarized two-prong power cord is permanently attached, with a polarized two-prong accessory outlet rated for 100W/1A just below.

Photo 3 shows the amplifier with the cover removed. A schematic was not furnished with the unit, but service manuals are usually available from NAD at extra cost. Looking just inside the left rear panel, you'll note the power cord is threaded through a toroidal ferrite EMI filter core. The left rear of the L53 has an AC power input board with a MOV transient suppressor, film caps, and the line fuse.

The small transformer supplies power to the front panel display/control board just under the front lip of the chassis (out of view in the photo). The larger transformer at the left front supplies the high current requirements for the power amplifier, and the lower voltages needed for the logic, video, digital audio, and analog line-level audio circuits. The



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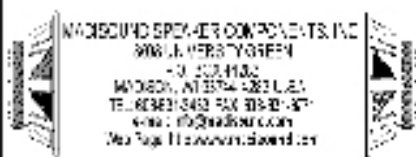
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location of these two power transformers makes the L53 much heavier on the left side.

Front and center is a Raymedia CD/DVD drive unit with an A/V control/interface PC board incorporated in the bottom of its housing. The DVD format is handled by an ESS Vibrato 2 MPEG-4 chip set with a 54MHz 10-bit video DAC. A Cirrus Logic CS4391A 24-bit/96kHz stereo DAC decodes 16-bit PCM to analog audio.

Stacked just behind the drive are two horizontal PC boards. The bottom board is the video interface and switching board, while the audio and logic control is on top. The various functions are controlled by means of an STMicroelectronics ST92F124V1T6 8/16-bit microcontroller chip with 128K of flash memory. A minimal number of discrete 74ACT chips provide TTL-to-CMOS glue logic.

The analog audio is processed through an NJR NJW1153P six-channel electronic volume control IC with input select, tone control, mute functions, and three output record channels. Based on the data sheet for this chip, it appears to be entirely analog in the signal path. A Philips SAAA6579T chip demodulates the Radio Data System (RDS) information from the tuner section. An SRS Labs SRS31D0 chip functions as the Sound Retrieval System (SRS) surround ambient decoder.

There are two dual op amps on the audio board. One is an NJM2068M, roughly identical to ye olde RC4558 from the mid-1970s. The other, despite the fact that the silk-screen ID in its location also reads 2068M, is a high-quality Burr-Brown OPA2134 located adjacent to the left/right channel audio inputs to the power amplifier board. The 2068 op amp appears to provide summing and low-pass equalization for the subwoofer output.

Dual supply voltages are provided for the NJW1153 by L7808 and L7908 linear regulator ICs. The rear panel audio input and output jacks are all coupled with aluminum electrolytics. The video board uses discrete RF transistors for video output buffering.

Additional regulated DC supplies (+5, ± 12 V) for the audio board come from the front area of the power amp

supply board on the right side of the unit. The three linear regulator ICs are located on the forward edge of the heat-sink. The power amplifier DC supply is at the rear of the supply board.

Main reservoir capacitors are two 6800 μ F 50V Koshin electrolytics. Supply rail fuses and the speaker Zobel networks and output inductors are located nearby. Just in front of the speaker binding posts is a 5A startup delay/protection relay. The discrete power amplifier circuitry is on the vertical PC board attached by metal clips to the large aluminum heatsink.

The AM/FM tuner and PLL are located in the shielded enclosure directly in front of the antenna jacks. All-in-all, the L53 has a compact, well-thought-out component arrangement. NAD also makes the L73 Surround Sound Receiver with five channels of audio amplification and adds DVD-Audio playback and HDTV-compatible video output.

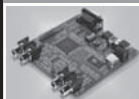
MEASUREMENTS

Before making any measurements, I set up the L53 for some selected listening to make sure all the audio modes were operating properly. It was at this point I found that the "included" remote control batteries had not been included, and that the L53 was very easy to operate from its front panel controls. When you insert a disk, the unit must determine the format before playback begins, because all shiny silver 120mm discs initially look the same. It took about 15 seconds from the "LOADING" display until the format was recognized and accepted for play. Initially, the drive (Korean-sourced, but made in China like the rest of the L53) was fairly noisy up close to the unit, but the noise gradually subsided until it was no louder than my Rotel CD player.

Because the L53 can also play MP3 files, I listened to the 128K-sampled compressed MP3 CD-R that I had previously made for the Pioneer DV-563A test (aX November 2004, p. 36). My perception was that the MP3 tracks were a bit louder than the CD tracks of the same songs. Imaging was better than with the DV-563A, but the soundstage was still flat, with most of the verb or ambience that was present in the CD version lost in the MP3 encoding. This

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external source options need to flow through the NJW1153P six-channel control IC, so its performance will set the baseline for the audio performance of the L53. There is very little in the way of important audio parameters in the NJW1153 data sheet, other than gain (always near unity) and THD+N curves versus input voltage. Important specifications such as slew rate, unity gain bandwidth, input impedance, bias current/voltage, and noise performance go missing.

I put the L53 in CD playback with a CD test disc for 1 hour before making any line level measurements. The player was just warm to the touch above the heatsink after this period at 1W into 8Ω.

FIGURE 1: Frequency response (line level sources).

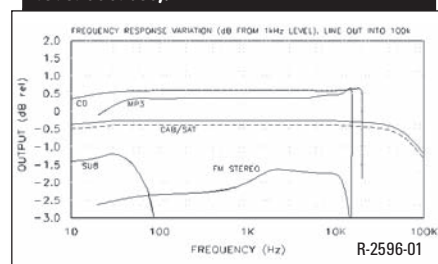


FIGURE 2: THD+N vs. frequency (line level sources).

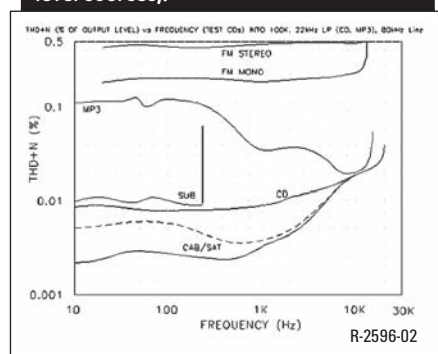
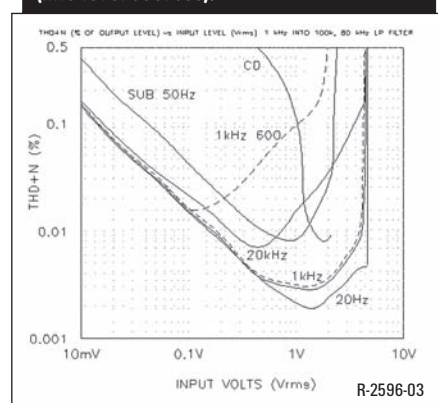


FIGURE 3: THD+N vs. input voltage (line level sources).



Input impedance for the CAB/SAT and VCR audio line inputs was approximately 90k. The S/PDIF digital coax input impedances measured the specified 75Ω. Output hum and noise with the analog inputs terminated in 600Ω measured -89dB (ref 2V RMS), decreasing to -107dB A-weighted. Front left and right channel separation measured a fine -80dB at 10kHz, unweighted.

The output impedance for the two-channel stereo VRC outputs measured 460Ω from 1kHz to 20kHz, and a higher 700Ω at 20Hz where the output coupling capacitor adds to the impedance. L-R channel balance was essentially perfect. The stereo outputs do not respond to variations in the volume, balance, or tone control settings, and are passed through from the sources at nearly unity gain (-0.26dB), although this will change with load.

The subwoofer output does follow the volume control setting, and had an output impedance of 110Ω, and showed unity gain with a volume control indication of "-1" (100k load). The S/PDIF digital coax and the three component video output impedances are all specified to be 75Ω. All input and output jack shells and the speaker negative binding posts are connected to chassis ground.

LINE LEVEL SOURCE MEASUREMENTS

Frequency response for line level sources (CD, MP3, FM tuner, and the VCR line input) into 100k is shown in **Fig. 1**. The VCR line level output is also shown with a 10k load (dashed line). The 1kHz 0dBfs CD line output was 2.12V RMS at 1kHz, or 0.60dB higher than the CD Red Book standard of 2V RMS.

There was just 0.2dB of de-emphasis response error at 1kHz when playing pre-emphasized CD test tracks. The MP3 frequency response with identical CD test tracks ripped to 128k MP3 format was 0.38dB higher through the midband before dropping off just above 15kHz. The subwoofer output response is shown with 2V RMS input and the volume at "-2" to improve clarity on the graph.

I measured the FM tuner frequency response with an RF signal level of 65dBf and 100% (±15kHz) modulation.

In this measurement there are three responses that can contribute to the overall error: the 75 μ S pre-emphasis and steep 16kHz LP filter on the audio that is fed to the FM signal generator, and the 75 μ S de-emphasis in the tuner under test. FM stereo crosstalk performance was 42dB at 1kHz.

Note that the FM stereo line level output at 1kHz with 100% modulation is 750mV, or -8dB relative to 2V RMS. Accordingly, the FM frequency response curve has been shifted up by +6dB to keep it within the graph.

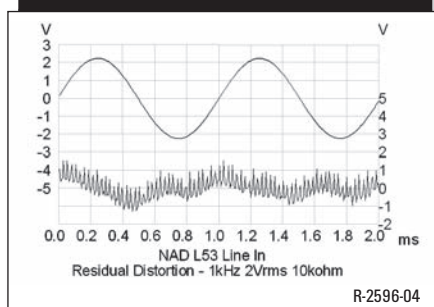
THD+N versus frequency is shown in **Fig. 2** for each audio line level mode. I engaged the 22kHz LP filter in my distortion test set for the CD and MP3 tests, and the 80kHz LP filter for the line input and FM tuner. The analog CAB/SAT input voltage was held to 2V RMS because this is the 0dB level on the CD test disc and the VCR line output does not respond to the volume control. Distortion for the FM tuner section is shown for mono and stereo. The analog line input distortion with a 10k load is again shown

with a dashed line.

The THD+N for the MP3 test tracks shows a bit of a roller-coaster ride. There is also a dip at 60Hz that may indicate that some 60Hz AC line pickup is being notched out by the distortion test set at that frequency.

THD+N vs. input voltage is shown in **Fig. 3**, at 1kHz for each line level audio mode into 100k (I excluded the FM curve because reducing the modulation level greatly increased the distortion, which was near 0.5% to begin with). I measured the subwoofer THD at 50Hz. This time I added a very low 600 Ω line level load at 1kHz (dashed curve).

FIGURE 4: Line input 1kHz residual distortion with 80kHz LP filter.

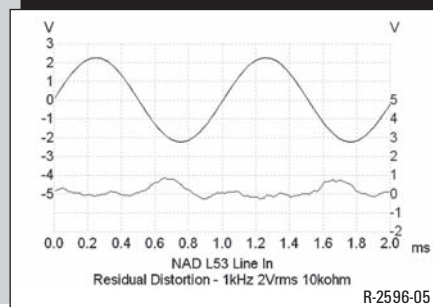


ANALOG CAB/SAT LINE INPUT PERFORMANCE

The residual distortion signal with an analog input of 1kHz at 2V RMS into 10k is shown in **Fig. 4**. The upper waveform is the VCR line output signal, and the lower waveform is the distortion test set monitor output after the test set notch filter, not to scale. THD+N at this test point is a low 0.0029%. Note the series of 44.1kHz spikes riding on the low-level noise, perhaps due to coupling from the digital audio section (although the CD player is off at this time).

Figure 5 shows the same data, but with the steep 22kHz low-pass filter

FIGURE 5: Line input 1kHz residual distortion with 22kHz LP filter.



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rather than the gradual 80kHz low-pass filter engaged. This confirms that the spikes do not produce any noticeable artifacts within the audio band. The negative half-cycles appear to have a predominant second harmonic, while the positive half-cycles are a mix of higher harmonics and noise.

The spectrum for a 50Hz sine wave at 2V RMS into 100k is shown from DC to 1.3kHz in **Fig. 6**. The THD+N measures only 0.0024%. The levels of the second through fifth harmonics were -108dB, -96dB, -98dB, and -103dB, respectively, for a calculated THD of 0.0021%. Decreasing the load to 10k produced a uniform increase in all the harmonics for a THD+N reading of 0.0061%.

Repeating the spectrum at 1kHz (not shown) saw a similar distribution of harmonics with a THD+N of 0.0031% into 100k and 0.0039% into 10k. The slight rise in THD near 60Hz in **Fig. 2** may be caused by low-level coupling of

the power line, but it was not visible in **Fig. 6**.

In view of the presence of coherent HF spikes in the residual distortion of **Fig. 4**, I extended the 1kHz spectrum to 166kHz to see whether they were present (**Fig. 7**). Here you can see the 44.1kHz spike at -90dB and an additional 88.2kHz spike at -95dB. Again, there are no visible artifacts in the audio band, but the presence of these two frequencies with the CD player off is puzzling, if benign. The logic clock remains running for the microcontroller, of course.

Figure 8 shows the L53's output spectrum reproducing a combined 19kHz + 20kHz CCIF intermodulation distortion (IMD) signal just below the $\pm 6.8V$ clipping level of the NJW1153 chip into 10k Ω . The 1kHz IMD product is -96dB

FIGURE 6: Line input spectrum of 50Hz sine wave.

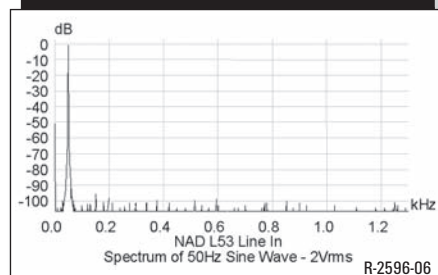


FIGURE 7: Line input extended spectrum of 1kHz sine wave.

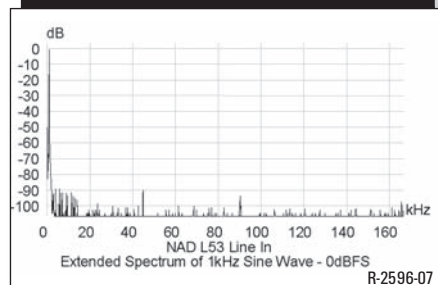


FIGURE 8: Spectrum of 19kHz + 20kHz CCIF intermodulation signal.

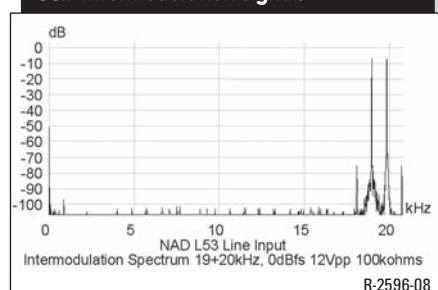


FIGURE 9: Spectrum of 9kHz + 10.05kHz + 20kHz multi-tone intermodulation signal.

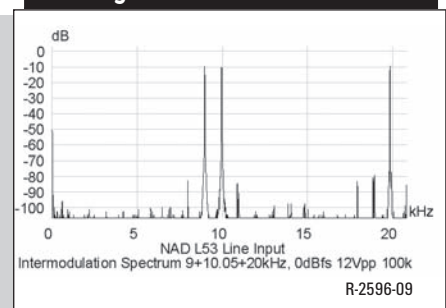


FIGURE 10: Square-wave response at 40Hz into 10k.

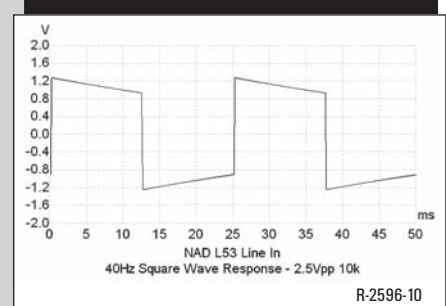
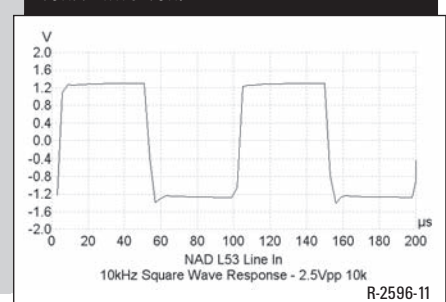


FIGURE 11: Square-wave response at 10kHz into 10k.



(0.0016%), with the 18kHz and 21kHz products at -85dB, or 0.0056%. Repeating the test with a multi-tone 9kHz + 10.05kHz + 20kHz IMD signal (Fig. 9) resulted in a worst-case 18kHz product of -79dB or 0.011%. There is also a picket fence of low-level 1kHz products throughout the spectrum.

I viewed the response of the L53 to three square-wave test frequencies using a 10k load. The response at 40Hz (Fig. 10) showed a moderate and acceptable amount of tilt, not all that different from the same 40Hz into 100k. The 1kHz square wave (not shown) had a bit of rolloff on the positive-going edges but a single half-cycle peak at the negative-going edges. This is more apparent with the 10kHz square wave shown in Fig. 11. Because there is no high-frequency response peaking for the analog input signals into resistive loads in Fig. 1, I can't really explain this asymmetry. I ran it long enough to ensure it wasn't due to aliasing with my DSO.

SUBWOOFER OUTPUT

The subwoofer line output frequency

response (Fig. 1) is fixed and cannot be altered by the tone controls. Matching subwoofer level with the main speakers can be performed using the level control on a powered sub, after which the L53 subwoofer line output level tracks the volume control in the L53. The NJW1153 chip incorporates a subwoofer level control that, like the other five channels, tracks the six-channel Amp Gain setting, but the chip does not have any subwoofer low-pass crossover EQ capability. I surmised earlier that the subwoofer stereo summing and LP filter functions are performed by the 2068 op amp located adjacent to the subwoofer line output jack.

CD AND MP3 PLAYER MODES

I performed measurements on the L53 CD player section using my usual test discs. I did not attempt to measure the Dolby Digital or DTS audio performance because I do not have any DVD-V test discs. The analog VCR line outputs showed normal polarity, a positive-going test pulse producing a positive-going output. The digital black test



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tracks measured -100dB, indicating that the L53 was probably muting the outputs during this test.

The channel separation test tracks consist of a 0dB sine wave on one channel, and digital black on the opposite channel whose crosstalk is being measured. The L53 also appears to mute the analog output of the silent channel during this test. I measured the response at the VCR audio line output jacks, as I did with the other sources.

In CD mode, the player performed perfectly in the track defect dropout tests out to track 32 on the Pierre Verany test disc, which contains a 1.25mm gap (Red Book requirement is 0.2mm). Track 33, with a 1.5mm gap, produced clicking sounds and some audio muting.

The residual distortion signal in CD mode for 1kHz at 0dBfs (Fig. 12) shows the same harmonic asymmetry as the line level residual in Fig. 5. The MP3 residual distortion signal (not shown) was similar, with a bit higher noise level.

The spectrum of a CD 50Hz sine wave at 0dBfs is shown in Fig. 13, from DC to 1.3kHz. The calculated THD based solely on harmonics was 0.0078%, while the THD+N measured 0.009% with the distortion test set. It stayed at this level through 1kHz before rising at higher frequencies as shown in Fig. 2. The second harmonic measured -91dB, the third was -96dB, the fourth -115dB, and the fifth at -82dB. Since the CD spectrum is characterized by a much higher noise level than that of the line input alone in Fig. 6, I trust that the spectrum analyzer has properly isolated the actual harmonics hidden within the noise. There are no 60Hz power line harmonics or other spurs evident in the spectrum.

The MP3 spectrum in Fig. 14 shows a higher concentration of noise along with harmonically unrelated spikes as high as -73dB. The THD+N measures 0.11%, although it drops to 0.034% at 1kHz (Fig. 2).

Figure 15 shows the CD spectrum of response to equal level CCIF 11kHz and 12kHz intermodulation test signals, each at -6dBfs, from DC to 20.8kHz. The 1kHz IMD difference product measures -72dB with a matching product at 2kHz. There are no products evident

at 10kHz or 13kHz, but they may be buried in the noise.

When the same IMD signals are ripped to MP3 format, the results are not quite as good (Fig. 16). The 1kHz and 2kHz products are higher at -71dB and are joined by visible 10kHz and 13kHz products. The rising skirts around the 11kHz and 12kHz stimulus signals indicate noise that may be due to the MP3 audio compression. There

FIGURE 12: Distortion residual of CD 1kHz sine wave.

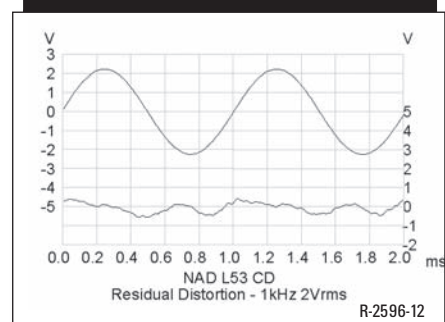


FIGURE 13: Spectrum of CD 50Hz sine wave.

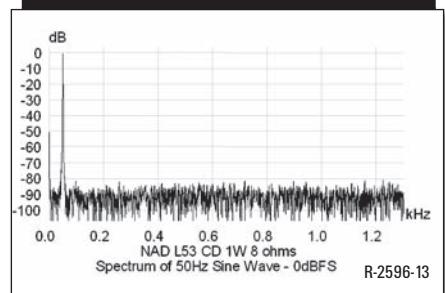


FIGURE 14: Spectrum of MP3 50Hz sine wave.

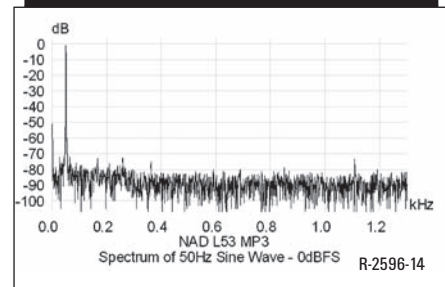
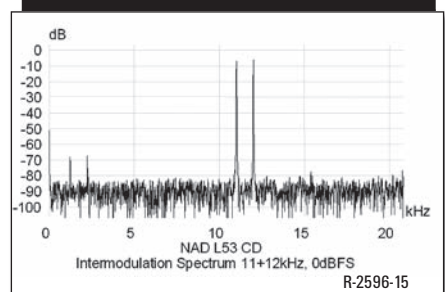


FIGURE 15: Spectrum of CD 11kHz + 12kHz intermodulation.



are also additional spikes that have no relation to the IMD test signals.

A repeat of the test in CD mode with the more difficult 19kHz + 20kHz IMD test track (**Fig. 17**) shows the 1kHz, 18kHz, and 21kHz intermodulation difference products to be about -80dB (0.01%). But for the noise level, this is decent performance.

Figure 18 shows the CD reproduction of an undithered 1kHz sine wave at -90.31dBfs. At this level the signal consists of ± 1 bit of data, producing two different voltage levels that are symmetrical about the horizontal axis (time). These discrete voltage steps are obscured in the -80dB noise floor, so the effective bit rate is probably about 14 bits.

Noise also took its toll on the $\frac{1}{3}$ octave spectrum of the same undithered 1kHz sine wave at -90.31dBfs shown

in **Fig. 19**. There is a slight rise in the spectrum around 120Hz, which may be a power supply rectification artifact, but a discrete 1kHz peak is nowhere to be found. A $\frac{1}{3}$ octave spectrum of the -90.31dBfs test track ripped to MP3 showed a series of peaks at multiples of the 60Hz line and higher initial noise floor (**Fig. 20**).

The CD playback of a 0dBfs square wave at 997Hz (**Fig. 21**) exhibits a slightly unsymmetrical Gibbs phenomenon damped sinusoid ringing associated with the steep digital filters used in the L53. Reproduction of the same square wave ripped to MP3 is shown in **Fig. 22**. Note that the Gibbs pre and post echo ripple has a total of 16 "pulses" per full cycle, compared with the 22 pulse CD playback.

You can approximate the PCM Nyquist frequency response limit by mul-



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FIGURE 16: Spectrum of MP3 11kHz + 12kHz intermodulation.

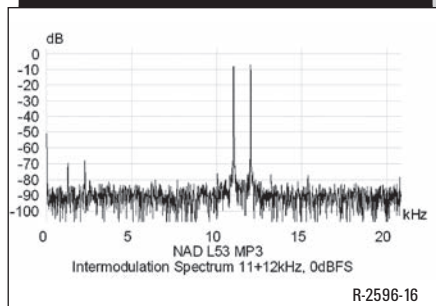


FIGURE 19: 1/3 octave spectrum of CD 1kHz sine wave at -90.31dBfs.

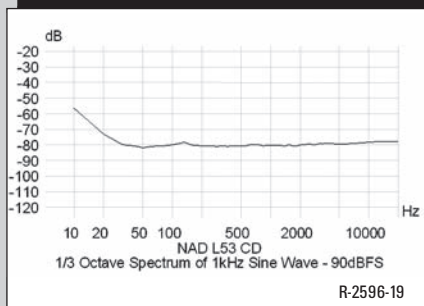


FIGURE 17: Spectrum of CD 19kHz + 20kHz intermodulation.

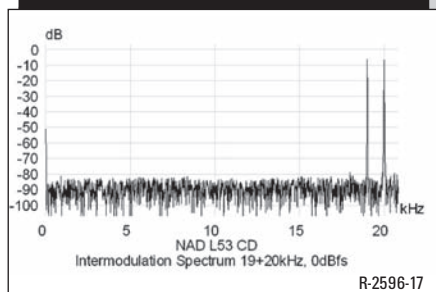


FIGURE 20: 1/3 octave spectrum of MP3 1kHz sine wave at -90.31dBfs.

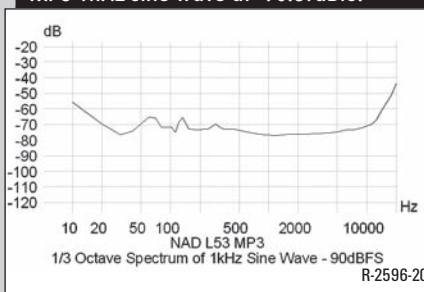


FIGURE 18: Undithered CD 1kHz sine wave at -90.31dBfs.

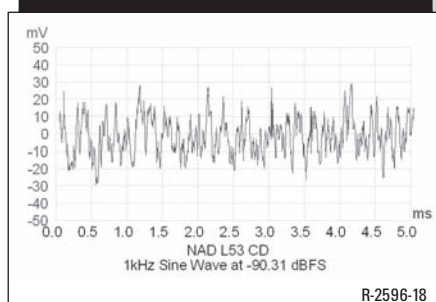
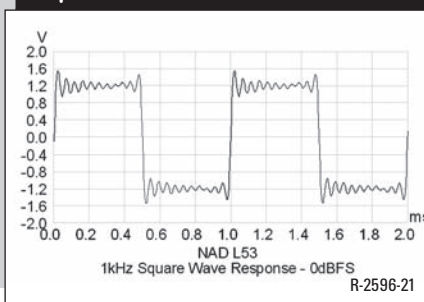


FIGURE 21: 997Hz CD square-wave response.



tipling the number of pulses by the fundamental square-wave frequency. This demonstrates the more limited high frequency response available as a result of the MP3 compression algorithm. The MP3 waveform is even less symmetrical, and appears to be slightly modulated by 60Hz.

MEASUREMENTS — FM TUNER SECTION

I did not run any tests on the AM section of the tuner, other than making sure it was functional. The FM tuner has two modes: FM and MUTE. The display shows “Tuned” when the tuner

locks to a valid FM signal, although there is no meter or bar graph to show signal strength. The L53 FM tuner inverts the audio polarity.

The L53 stops its manual station scan every 100kHz, even though only the odd 0.1MHz channels are assigned in North America. Same reason that the SCART video output connector is provided, I guess, except these additional FM frequency stops are a bit more time consuming and less convenient to the user during manual tuning. (Manufacturer’s Note: the SCART connector is supplied in the European version only).

The L53 tuner quieting character-

istics are shown in **Fig. 23**. I needed to switch to FM mode (not MUTE) to measure mono sensitivity. The minimum subjective usable sensitivity using manual tuning was 10.6dBf, with some visible noise breakup in the audio signal on the scope. The mono signal became free from visible noise artifacts at 16dBf.

The station auto search threshold, where it would stop at the signal generator’s frequency during a station scan, was 19dBf. The red stereo indication came on at 22dBf. The 50dB stereo quieting signal strength was approximately 30dBf. There was no overload at the maximum RF input of 100dBf. The audio output disappeared when I moved the FM test signal 200kHz to either side of the tuned center frequency, so adjacent channel rejection should be adequate.

When I engaged the FM MUTE mode, the auto search threshold increased to 34dBf. Using the supplied FM dipole antenna and 300Ω to 75Ω balun, I could pick up only about half the FM stations in FM MUTE as in FM mode with its lower auto search threshold. Performance was much better in my audio system with its dedicated 75Ω FM antenna.

The FM performance is certainly adequate, especially in a strong signal area, but not quite as good as my venerable NAD 4155 dedicated tuner.

POWER AMPLIFIER SECTION MEASUREMENTS

I operated the L53 amplifier with a line input signal of 0.5V RMS 1kHz and adjusted the volume for 10W into 8Ω. For most of the intended one-hour period, the THD for both channels remained near 0.01%. The temperature above the case cooling slots became quite hot after this period, measuring 58° C. After 45 minutes, the THD suddenly rose to 0.15%. The distortion residual showed a large increase in crossover spikes, so I assume the amplifier bias was decreasing to reduce temperature.

I lowered the power level to 1W and after another 15 minutes things returned to normal. Admittedly, a continuous 10W sine wave is a difficult operating condition. It is close to the ⅓ power or 16.7W that produces maximum dis-

FIGURE 22: 997Hz MP3 square-wave response.

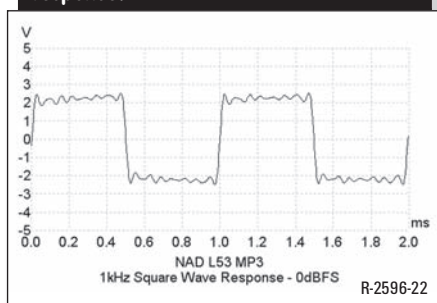
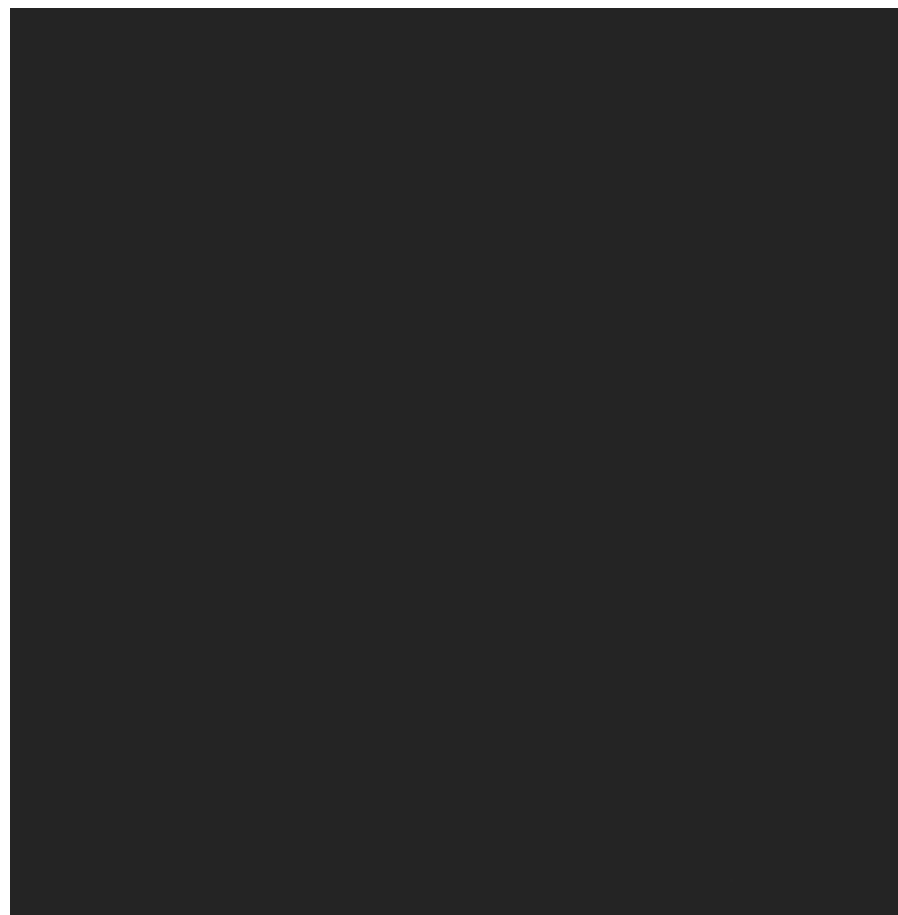
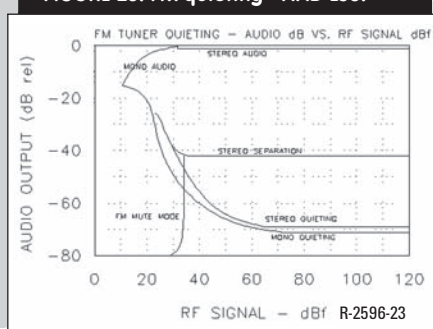


FIGURE 23: FM quieting - NAD L53.



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sipation in the output transistors.

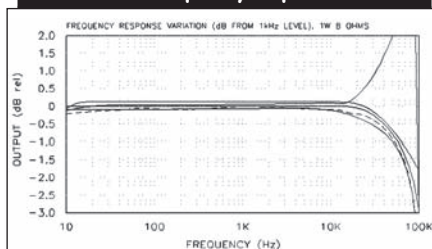
The measurements for the right channel are presented here. There is no noise at all when starting up or shutting down the amplifier. Output hum and noise measured -89dB, input shorted (-109dB with A-weighting), and some hiss was audible at full volume with my ear against the speaker. I also measured -40mV (L) and -13mV (R) of DC offset.

The L53 does not invert overall (line in to speaker out) polarity. The gain at 2.83V RMS output into 4Ω and 8Ω loads was a high 35dB and 36dB, respectively. I noticed that the right channel output was about 0.5dB higher throughout the entire volume control range, while the line level outputs tracked perfectly. The output impedance at 1kHz was 0.12Ω, increasing only slightly to 0.13Ω at 20kHz.

The frequency response at 2.83V RMS for loads of 4Ω, speaker (dashed line), 8Ω, no-load, and 8Ω paralleled with 2μF (bottom to top at 30kHz) is shown in **Fig. 24**. Response was within -1dB from 10Hz to 57kHz, at a reference output of 2.83V RMS at 1kHz into 8Ω. It wasn't down to -3dB until 110kHz. There was just a bit more HF rolloff with a 4Ω load.

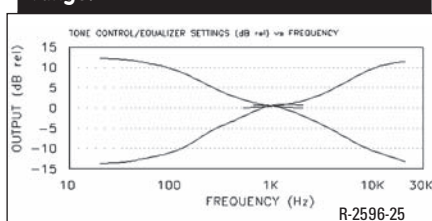
The response with a complex load of 8Ω paralleled with a 2μF cap (a test of compatibility with electrostatic speakers) produced a pronounced high frequency peak exceeding +5dB at 70kHz, at which point the L53 tripped into its

FIGURE 24: Frequency response.



R-2596-24

FIGURE 25: Tone control frequency range.



R-2596-25

protection mode. The flat response in the audio band tells me the L53 amplifier will be insensitive to variations in speaker impedance with frequency, although it may protest electrostatics.

At low frequencies the noise floor limited the excellent crosstalk performance, but there is adequate isolation between the stereo channels at higher frequencies (**Table 1**). The tone control frequency response range is shown in **Fig. 25**. The tone controls only operate on the power amplifier output and are inactive at the line or subwoofer outputs. The maximum spread at the low and high frequency extremes exceed the specified ± 10 dB by a few dB.

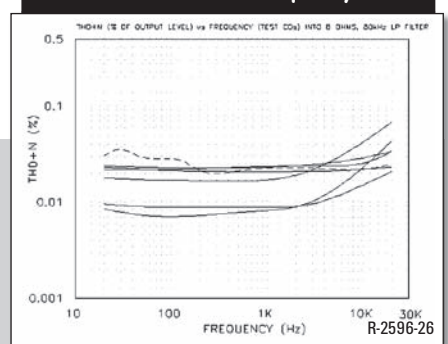
I made all the amplifier distortion measurements with the tone controls flat and the SRS turned off. In addition to the fixed test loads, I added an NHT SuperOne loudspeaker¹ for this measurement.

THD+N versus frequency is shown in **Fig. 26** for outputs of 8Ω 10W, 8Ω 1W, 8Ω paralleled with 2μF, 4Ω 2W,

TABLE 1: Crosstalk.

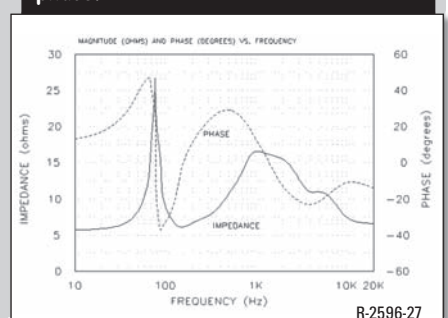
Frequency	R to L	L to R
100Hz	-77dB	-77dB
1kHz	-77dB	-76dB
10kHz	-73dB	-72dB
20kHz	-71dB	-70dB

FIGURE 26: THD+N vs. frequency.



R-2596-26

FIGURE 27: Speaker impedance and phase.

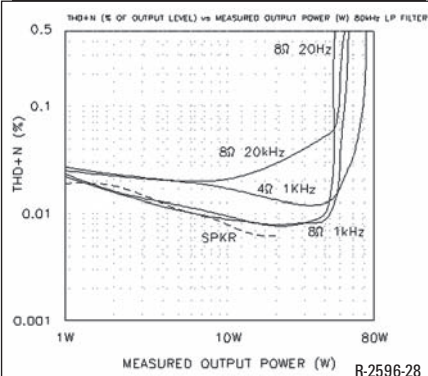


R-2596-27

8 Ω 33W, and 4 Ω 10W (bottom to top at 20kHz). The speaker load is shown with a dashed line, and its distortion is inversely proportional to the speaker impedance. The impedance and phase of the NHT SuperOne is shown in **Fig. 27**. During distortion testing, I engaged the test set 80kHz low-pass filter to limit the out of band noise.

Figure 28 shows THD+N versus output power for the loads and frequencies shown in the graph. The L53, with both channels driven, reached its 1% clip-point at 55.8W with the 8 Ω load

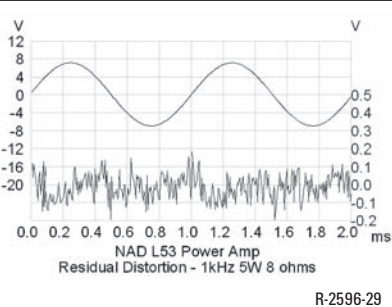
FIGURE 28: THD+N vs. output power.



(for 0.47dBW of headroom) and 73.4W with the 4 Ω load (1.67dBW). The positive half-cycles clipped just slightly before the negative half-cycles, and both showed a scoop-out at the peak of the sine wave as if the top 5% of the peak were inverted back into the waveform. This was a bit unusual, as most solid-state amplifiers clip flat across the top. The chassis above the heatsink reached a maximum temperature of 61° C after 15 minutes at 16.7W into 8 Ω , so I reduced power to avoid another protective trip.

The distortion residual waveform for 5W into 8 Ω at 1kHz is shown in **Fig. 29**.

FIGURE 29: Residual distortion 1kHz 5W.



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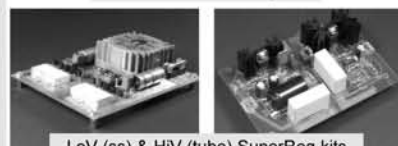
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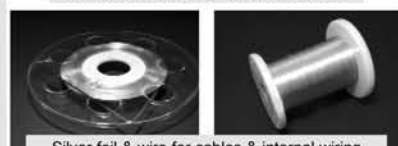
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The upper waveform is the amplifier output signal, and the lower waveform is the monitor output (after the THD test set notch filter), not to scale. This distortion residual signal shows mainly noise with perhaps a hint of higher order harmonics. THD+N at this test point is a low 0.01%.

Increasing the output power to 33W (Fig. 30) showed a bit of periodic scoop-out in the noise, suggesting second harmonics on the negative half-cycles. Distortion here is even lower at 0.0082%.

The spectrum of a 50Hz line input sine wave at 5W into 8Ω is shown in Fig. 31, from DC to 1.3kHz. The second, third, fourth, and fifth harmonics measure -87dB, -90dB, -105dB, and -93dB, respectively. The calculated THD based on the distribution of harmonics is 0.0053%, while the THD+N measured was 0.01%. Repeating the 5W spectrum with the 50Hz 0dBfs CD test track (Fig. 32) produced a spectrum with a higher noise floor. THD+N is also higher at 0.023% with a more prominent second and fifth harmonic.

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Figure 33 shows the amplifier output spectrum reproducing a combined 19kHz + 20kHz CCIF IMD signal at 20Vpp into 8Ω. The 1kHz IMD product is -82dB (0.008%), and the 18kHz product is -79dB (0.011%) with the 21kHz product just off the graph. Repeating the test with a multi-tone IMD signal (9kHz + 10.05kHz + 20kHz, shown in Fig. 34) resulted in 950Hz and 1050Hz

products of -82dB, a group of spikes around the two lower tones and above 15kHz of -95dB, and an 18.95kHz product of -92dB. This test gives a better indication of the amplifier's nonlinear response, because it is a closer approximation to music than a sine wave. The L53 power amplifier produced good IMD results, even better than the line input signals of Figs. 8 and 9, which were taken at a higher voltage and not

FIGURE 30: Residual distortion 1kHz 33W.

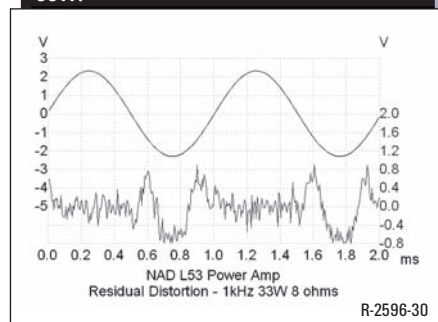


FIGURE 31: Spectrum of 50Hz line input sine wave.

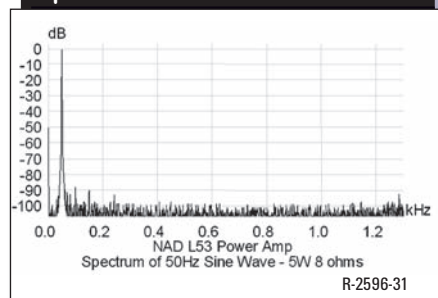


FIGURE 32: Spectrum of 50Hz CD sine wave.

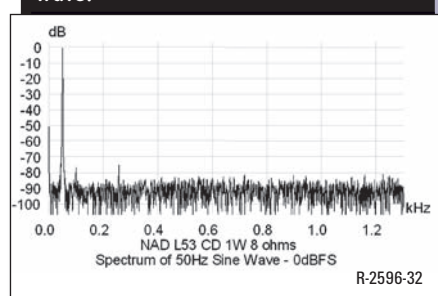


FIGURE 33: Spectrum of 19kHz + 20kHz intermodulation signal.

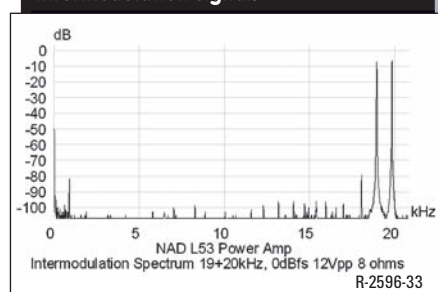


FIGURE 34: Spectrum of 9kHz + 10.05kHz + 20kHz intermodulation signal.

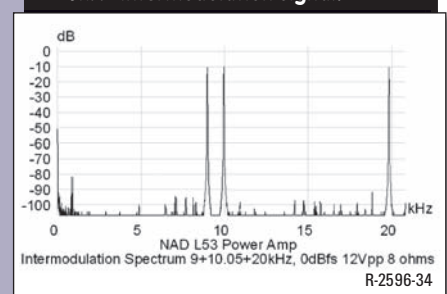


FIGURE 35: Square-wave response at 40Hz into 8Ω.

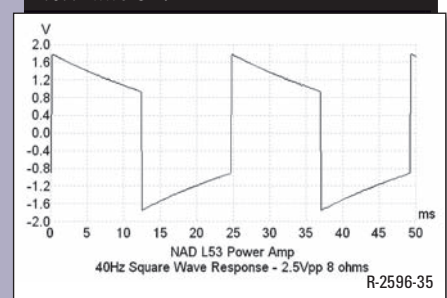


FIGURE 36: Square-wave response at 10kHz into 8Ω.

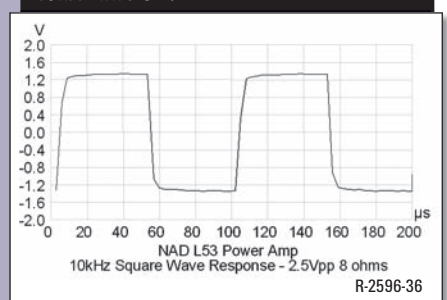
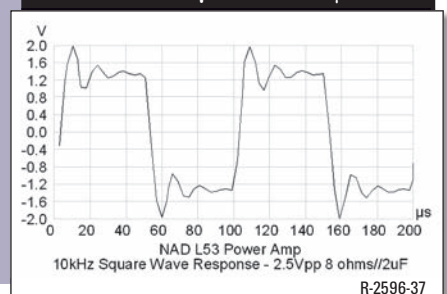


FIGURE 37: Square-wave response at 10kHz into 8Ω in parallel with 2μF.



at the lower level that reflects the additional voltage amplification by the power amplifier.

The 2.5Vp-p 40Hz square wave in **Fig. 35** shows more tilt than that of the line input alone in **Fig. 10**, but the power amplifier does add its own tilt to the initial tilt of the preamp 40Hz. The square-wave response to 1kHz into 8Ω (not shown) was just about perfect. The leading edge of the 10kHz square wave in **Fig. 36** was slightly rounded, but when I connected 2μF in parallel with the 8Ω load (**Fig. 37**), there was significant ringing at 70kHz, conforming to the response peak in **Fig. 24**.

Overall, the L53 is a nicely performing piece of equipment that incorporates a lot of features at this price point. Full disclosure here: I own a number of NAD products: a 214 stereo amplifier, a 304 integrated amplifier, and

a 4155 AM/FM tuner. The L53 met or exceeded its published specifications, except for the performance of the CD section, although it could be argued my distortion measurements are within the statistical variation for independent tests. NAD also specifies a different noise test method (EIAJ DVD tests) than I used, and I could not duplicate their test because I don't have a DVD video test disc in my collection. A comparison of the manufacturer's ratings and my measured results is shown in **Table 2**. *ax*

[Watch for listening critique results of this unit in a subsequent issue. —Eds.]

REFERENCES

1. "The case for more testing," Dennis Colin letter, *Stereophile*, June 2004, p. 11.

TABLE 2: Measured Performance.

Parameter	Manufacturer's Rating	Measured Results
Power Output, Stereo Mode	2×50W, 0.08% THD - 8Ω	2×50W, 0.03% THD - 8Ω
Frequency Response, 8W, 4Ω	20Hz - 20kHz ±0.5dB	10Hz - 34kHz ±0.5dB, 1W 8Ω
Total Harmonic Distortion	0.08% - 50W - 1kHz	0.08% - 55W - 8Ω
IMD - CCIF (19+20kHz)	0.08% at Rated Power	0.011% CCIF 20Vpp
MIM (9+10.05+20kHz)		0.0079% MIM 20Vpp
Damping Factor	>60	73 (1kHz), 60(20kHz)
Frequency Response	20 to 20kHz ±0.5dB	10 to 41kHz ±0.5dB, 8Ω
Signal/Noise Ratio	>99dB, 50W 8Ω A-weighting >85dB, 1W 8Ω A-weighting	109dB, 50W 8Ω A-weighting 91dB, 1W 8Ω A-weighting
Noise	0.8mV	0.71mV unweighted
Gain	N/S	35dB 4Ω, 36dB 8Ω
Amplifier Output Impedance	N/S	0.12Ω 1kHz, 8Ω 0.13Ω 20kHz, 8Ω
FM Tuner Input Sensitivity	Mono: 16.1dBf Stereo: 36.1dBf	Mono: 16dBf (subjective) Stereo: 30dBf
FM Harmonic Distortion	Mono: 0.2% Stereo: 0.5%	Mono: 0.19% (400Hz) Stereo: 0.48% (400Hz)
Signal/Noise	Mono: 69dB Stereo: 64dB	Not Tested
Stereo Separation, 1kHz	40dB	42dB
Frequency Response	30Hz - 15kHz ±1.5dB	20Hz - 15kHz ±1.4dB
Composite Video Out (RCA jack, yellow)	1Vpp 75Ω negative sync	Not Tested
Component Video Out (3× RCA jacks)	Y (luminance), 1Vpp 75Ω, negative sync, C _B , C _R (color), 0.7Vpp	Not Tested
S-Video Output	Y, 1Vpp 75Ω, negative sync, C (color), 0.286Vpp 75Ω	Not Tested
Mini-DIN, 2× 4-pin	CVBS, 1Vpp 75Ω, negative sync, RGB, 0.63Vpp 75Ω	Not Tested
SCART video out (EUR only)		
Analog Audio Outputs	2V RMS, 1kHz 0.dB, 330Ω	Line: 1.94V RMS, 1kHz -0.26dB, 460Ω Subwoofer: 60Hz, 110Ω
Analog Audio Inputs	2V RMS, 1kHz 0dB, 330Ω	2V RMS, 1kHz 0dB, ≈90k
DVD signal	PAL/NTSC/AUTO	Not Tested
CD Audio Frequency Range	4Hz - 20kHz	4Hz - 20kHz ±0.28dB
Signal/Noise (audio)	>100dB (EIAJ)	Not Tested
Dynamic Range (audio)	>95dB (EIAJ)	Noise floor ≈ -80dB (CD)
Harmonic Distortion (audio)	0.008%	0.009% 1kHz 0dB



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INCREDIBLE SPEAKER

I've built the proposed speaker enclosure in Aren van Waarde's "A Simple Audio System" (March '05 aX). The supplier, Digitex, from Florence (Italy), where I ordered the two Fostex F103 sigma, was out of stock of this model but they suggested a *re-logoed* model remarketed in Switzerland—FE103S ACR, which, they claim, is nothing but the original Fostex.

I built the box according to the drawing but using a 25mm multilayer pine wood, damping the inside with two layers of different material: One was a 6mm-thick foam used in our house construction to acoustically isolate floors and, on top of it, a 5mm thick felt used underneath carpets.

I hooked up my excellent, evergreen, Heathkit AR 1500 amp to test it. Well, I was astonished by its sound. It is almost incredible how a 100mm speaker is capable of reproducing the bass. All this and more is well described in the article, but you've got to hear to believe it!

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boxes for lovers of good sound.

Bravo, Aren!

*Atto Rinaldo
Tambre, Italy*

Aren van Waarde responds:

Nice to hear that you have built Petite speakers and that they work well in your system. The ACR driver is indeed the original Fostex; I used ACR FE103S drivers myself, which I ordered from Spectrum Audio (Bremen, Germany).

However, your praise should go to Gérard Chrétien and the people from *L'Audiophile* in Paris, who originally published this column-shaped enclosure.

SATELLITE RADIO

In response to my letter regarding satellite radio, which appeared in the June '05 issue of *audioXpress* (p. 49), I received two replies. From them, I found that XM uses Advanced Audio Coding, which the National Association of Broadcasters says is better than Sirius' Perceptual Audio Coding. Both services use enhanced versions of these codecs, and they both cut off brick wall at about 15kHz. I have heard highly mixed reports as to which is better, although XM has over four times the music library size of Sirius. More information can be found on www.XM411.com and www.siriusbackstage.com.

Because the music on traditional AM and FM stations has become abysmal, I took the plunge at the beginning of 2005 and bought a Sirius Sportster receiver and took out a one-year subscription. Because the receiver came with a car kit and the home kit would be \$50 more, I cut the cigarette lighter plug off the provided car kit and connected the wires into an old 12V 7812 regulated video camera power supply. First off, the FM transmitter option adds considerable distortion; by all means use the direct audio output connection into your amplifier.

Even with this connection, upon hearing a song played on a Sirius program and then immediately playing a CD or record copy of that same song, the Sirius quality is more like that of a 45 rpm record than a CD. The news/talk/sports channels have an even more swirly sound due to their much lower sampling rates. I heard the Polk Audio XM

component tuner in a local audio salon, and noticed it also had some swirl in its sound.

On the positive side, satellite radio does provide a variety of music and other programming rarely found on traditional radio, and without all the commercials. The audio is better than the FM stations, which highly compress their audio, sometimes to utter clipping. However, the few still good FM stations are better than satellite. I wish to thank Scott and Peter who responded to my letter.

*Michael Kiley
Crestwood, Ill.*

SPEAKER DAMPING

I am writing in response to the article titled "Speaker Damping," by Cornelius Morton (Aug. '05). In the beginning of his article, Mr. Morton talks about adding resistance to reduce damping. This statement initially confused me because it is normal to see a loudspeaker's system Q go up with an increase of electrical source resistance. However, a few paragraphs further on, he stated that most drivers typically have a Q of 4 or more, and from this I think he must be referring to the mechanical Q (which is more usually described as Qms), and the resistance he mentioned must be mechanical resistance, not electrical resistance.

Intrigued to see where Mr. Morton's line of rhetoric was leading, I read on, only to realize that he did not appear to be making any distinctions between a driver's Qms, Qes, or Qts, all of which play an important role in the frequency response and system Q of a loudspeaker system, especially when the possibility of wide variations of an amplifier's output impedance are included.

Mr. Morton then goes on to describe Photo 1 as a near-field measurement (microphone positioned within 2" of the driver's cone) of a driver mounted in an over-stuffed sealed enclosure, which to me appears plausible. After this, he describes the same type of measurement on a loudspeaker that employs a ported enclosure as shown in his Photo 2. On examining Photo 2, I believe that the response was made without the contributing response of the port.

I should point out that the driver has a minima in its excursion at system resonance, but the port output will be at its

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peak. Indeed, the output from a microphone when making near-field measurements of the driver and port can easily differ by 20dB and more at resonance. I therefore strongly suspect that Photo 2 is demonstrating this phenomenon and thus gives a highly distorted view of this loudspeaker's response.

At the end of the article, he states that there is a phase shift of 90° from the voltage applied to the loudspeaker terminals, compared with the voltage appearing at the output of the microphone. The reason for this is most likely the group delay of the system at resonance, which is where the group delay approaches its maximum. As frequency goes up, the group delay will rapidly drop to a low value.

For those interested in this, I recommend checking the group delay of a 2nd order HP function with a -3dB response at, say, 30Hz. Such a filter should equate well with the acoustic response of a loudspeaker in a sealed enclosure with similar parameters. You should find a group delay of about 7.5mS at 30Hz and about 0.7mS at 100Hz.

After this, Mr. Morton describes two methods he used to measure the electrical impedance of drivers at their terminals. These methods are the constant-current and constant-voltage methods. He then states that the Qs of several drivers were calculated as listed in his Table 2 and provides the reader with the equation $Q = F_s/\text{Bandwidth}$; however, he does not define how bandwidth is measured. I would be grateful if Mr. Morton can define this for me.

I assume that because Mr. Morton started the paragraph by describing impedance measurements, that he determined the Q of these drivers from his measured impedance data and, furthermore, that because the impedance data changed between impedance measuring methods, he then declares the Q of the drivers changed as a function of the output resistance of the impedance measuring system.

Theoretically speaking, the impedance of any driver should not change, regardless of the measurement method used. If the impedance of a driver were made equivalent to an electrical network as described by Mr. Morton's Fig. 2, then for the impedance to change it would require one or more of the driver's equivalent components to have changed (i.e., like the moving mass or the suspension's compliance, and so on)!

Having said all this, I should emphasize that I am quite sure that Mr. Morton's measured data is quite accurate and I have no

intention of disputing this. However, it is my belief that Mr. Morton did not measure the two sets of data under exactly the same conditions; i.e., the voltage across the drivers terminals was not made the same for the two measurement methods, which was the major reason for his obtaining two conflicting sets of impedance data.

It is here that I should point out that it is indeed possible for equivalent component values in Mr. Morton's Fig. 2 to vary considerably depending upon the Cone's excursion, which in turn is dependent on the voltage applied to the driver's terminals. These variations have become well known in the loudspeaker industry with parameters such as the driver's free-air resonance (Fs) varying by as much as 27% when comparing low-level measurements with those of even modest cone excursions of 2mm peak. This data came from one of the many excellent papers published on the Klippel website <http://www.klippel.de/>, where many more examples of nonlinear functions are described. So the reason for the change in Q that Mr. Morton observed was, in my opinion, almost certainly due to the significant nonlinear components in the drivers he was measuring, brought about by different measurement conditions for the

two measurement methods.

*Peter L. Bath
San Gabriel, Calif.*

Cornelius Morton responds:

Thanks for your interest in my article. I will attempt to answer your questions in order.

First, concerning damping and resistance, I totally agree with your statement that adding source resistance decreases damping. However, I believe the exact wording of the article was "by inserting a fixed resistance into the system dissipating those lurking frequencies, damping the resonance." "System" indicates the driver, not the external source.

The effect of inserting a lossy element into the workings of the driver is to dissipate energy, both applied and stored. The stored energy is what you are after, but the lossy element loses applied energy as well as stored, so the end result is a lower Q (good) in conjunction with a loss of sensitivity (bad). It was a bit of a fad in the fifties and sixties to coat the surrounds and spiders with a viscous fluid that increased the friction generated by cone movement to increase damping; it also reduced the compliance of the surround and spider which lowered the resonant frequency of the driver. Figure 2 illustrates the equivalent electrical components of a typical cone driver.

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Your question of which "Q" I measured—Qms, Qes, or Qts—is certainly valid. I attempted to qualify that the measured Q was the measure of energy stored in the driver system per cycle against the energy lost in the system per cycle. The closest Thiele parameter to this is Qms.

Qes is the measure of control—losses—of the electrical system of the driver. Qts is a calculated parameter that relates to the most suitable enclosure type for the driver and is defined as $(1/Q_{ms} + 1/Q_{es})^{-1}$. The value of Qms will be typically higher than the measured Q of the driver due to a multiplier used in the Thiele measurements. The multiplier used is $ro^{1/2}$, where ro is defined as the ratio of voice coil current off resonance to the current at resonance or the impedance at resonance to the impedance off resonance.

Your concern for variations of amplifier output impedance variations could be valid for the constant voltage measurements when using tube or solid-state amplifiers without feedback. For the most part, modern solid-state amplifiers have a Zo of 0.05Ω or less, which is small compared to a 4 or 8Ω speaker. The constant-current measurements were made with a simulated Zo of 1kΩ, which, compared to the speaker, is, for all practical purposes, an open, as intended. The constant voltage measurements were made with

a simulated Zo of 0.24Ω, which represents a 3% error (a damping factor of 33) when measuring the impedance peak and a 4.2% error at the -3dB points. These values are certainly within the tolerances required for a valid representation of the effect of amplifier damping on driver Q.

Ported speakers *do* have a minimum in cone excursion at resonance, and—yes—the port adds to the sound levels at resonance. The problem is the resonance peak below and above the system resonance due to the overcoupling between the tuned port and the driver. If you examine the impedance and phase curves for a ported speaker, the low- and high-frequency peaks are quite evident, with the phase showing a maximum positive at the low-frequency 3dB point and a minimum negative at the high-frequency 3dB point of each peak. This is the signature of a resonance energy storage point that promotes the sonic signature of a ported system.

The impedance null and zero phase shift at the port/driver frequency indicates that there is no energy storage at that frequency. Sonically, the greater the amplitude of the lower and upper peaks, the more sonic coloration the system will exhibit. Note that cone excursion is at a maximum at the lower and upper peak frequencies.

The comment about phase shift or delay at resonance was tossed in for general information

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and to illustrate the difficulty of trying to time-adjust a subwoofer to the rest of the speaker systems.

Frequency, bandwidth, and Q measurements seem to cause you some problems, especially when derived from a constant-current drive and a constant voltage drive. There are two sources of impedance in a cone-type loudspeaker. One is the DC resistance of the voice coil and the second is the velocity of the voice coil through the magnetic field of the gap.

Figures 1 and 2 detail the electrical equivalences. This is the source of the impedance peak at resonance where the cone/voice coil velocity is highest. The moving voice coil generates a back (opposing) emf that is determined by the length of wire on the coil, the magnetic intensity through the gap, and the velocity of the voice coil within the gap.

As long as the coil remains within the constant magnetic field area of the gap and the coil length is unchanged, the determining function is the voice coil velocity. When performing constant-current measurements, the voice coil is effectively shunted with $1k\Omega Z_o$, which limits the current caused by the back emf to about 1mA per volt with a power dissipation through the 8Ω voice coil of about eight microwatts per volt.

When driven by the constant voltage drive, the voice coil is shunted by 0.24Ω , and the current caused by the back emf is now limited by the 8Ω DC resistance of the voice coil plus the $0.24\Omega Z_o$. Now the current approaches 121mA per volt of back emf and the power dissipated is approximately 117mW per volt. The result slows the voice coil velocity, reducing the effective impedance and damping the system. The only difference between the constant-current and the constant-voltage measurements is that the latter

allows the back emf to generate a current limited by the voice coil DC resistance, whereas the former essentially reduces currents generated by the back emf to zero.

To get a sense of the braking action involved, take your favorite bass driver with the voice coil disconnected and gently flex the cone; you will basically feel the stiffness of the suspension. Then do the same with a clip lead shorting out the voice coil terminals. The difference is the braking effect of the voice coil. The preceding ignores the effect of voice coil inductance, but in the frequency range below 100Hz this represents a fraction of an ohm and may safely be ignored.

To measure frequency and bandwidth, I used standard techniques. For the constant-current drive, I placed the voltmeter across the driver terminals and adjusted the drive for a reasonable level, about 20mV. I also adjusted the frequency of the drive to achieve a peak level and recorded the frequency, Fr. I reduced the drive level by 3dB, recorded the resulting voltage, and then returned the drive to the original level. I then adjusted the drive frequency to achieve the recorded -3dB level below and above Fr. These frequencies were recorded as Fl and Fh.

For the constant-voltage measurements, I read the voltage across the 0.24Ω resistor, set the initial off resonance value to 15mV, then adjusted the frequency to achieve a null reading and recorded Fr. I increased the drive by 3dB and recorded the voltage, and then reduced the drive to the original level and adjusted the frequency to achieve the recorded voltage below and above Fr. These frequencies were recorded as Fl and Fh.

The values for Fr measured with a constant-current drive and with a constant voltage drive

were typically within a cycle or two of each other for the same driver, which is an indication of good data. The low drive levels—30mW maximum at the voice coil terminals—ensured that the measurements were taken well within the linear portion of cone excursion.

A comment on your last paragraph: Simply put, you state that Q variations measured with a constant-current drive and a constant-voltage drive are due entirely to nonlinear variations in driver parameters induced by measurement differences. However, in your initial paragraph you state, "it is normal to see a loudspeaker's system Q go up with an increase of electrical source resistance." Yet here you are insisting that when the source resistance is decreased, the observed decrease in Q must be caused by measurement irregularities.

Because the data is consistent with loudspeaker theory and the test methods and results do not show the irregularities that would indicate bad data, I must disagree with your conclusions. For additional material on this subject, I refer you to *Elements of Acoustical Engineering*, second edition by Harry F. Olson, EE, Ph.D. D. Van Nostrand Co., 1947. This is an oldie but still a premiere reference for the field. Chapter 6 is devoted to direct radiating loudspeakers.

One of the major thrusts of this article was to explore the limits of speaker damping versus amplifier output impedance. The results indicate that the limiting factor is the DC resistance of the voice coil for amplifiers with a moderately high to high damping factor.

The second objective was to explore the availability of damping methods that lowered the system Q below that which is achievable by electrical damping alone. I believe that both objectives have been met. *ax*



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