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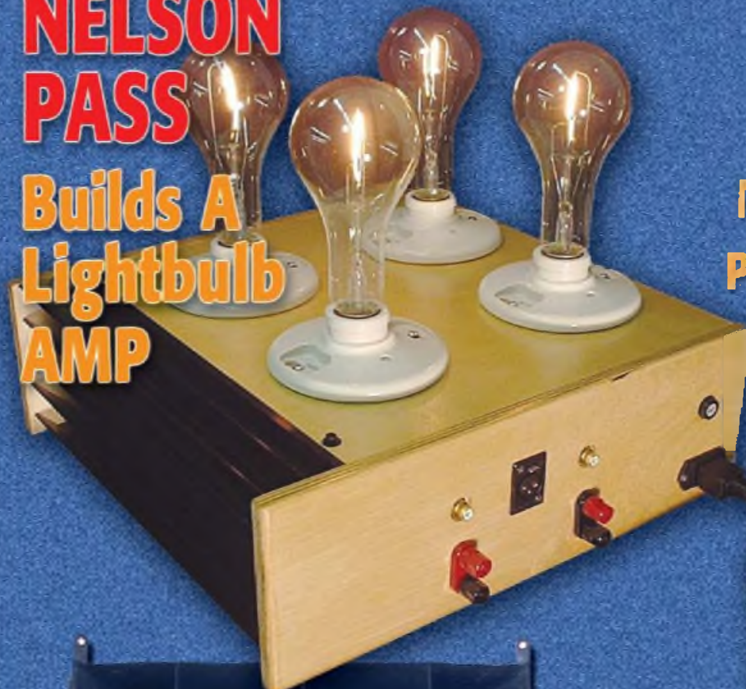

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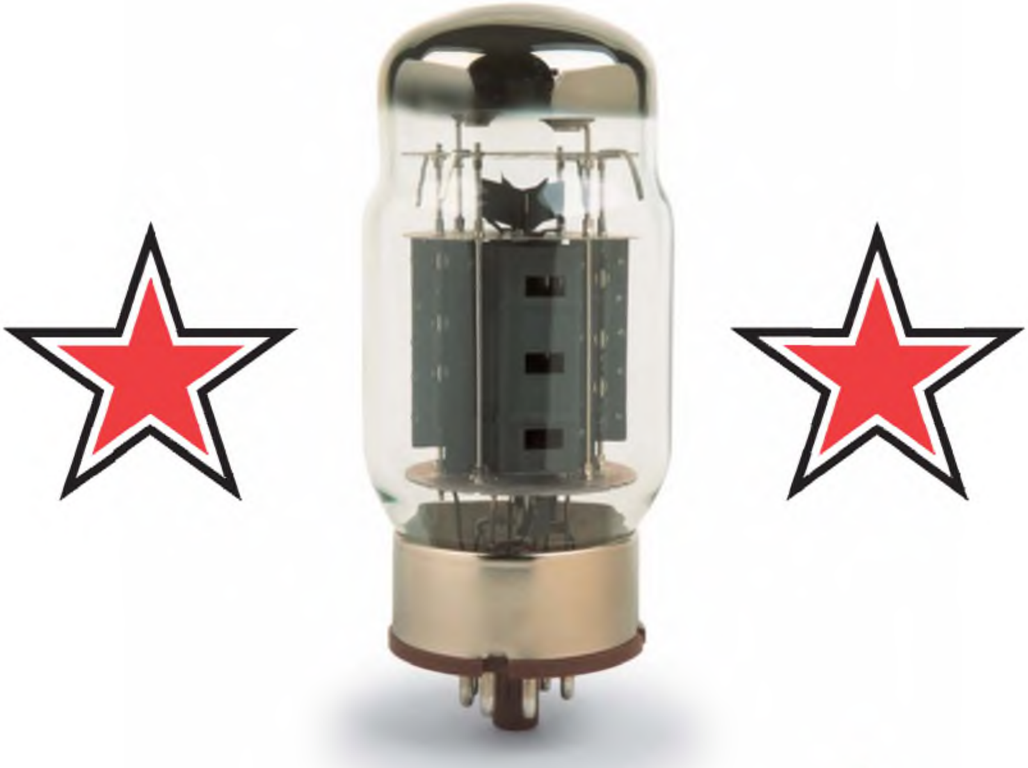

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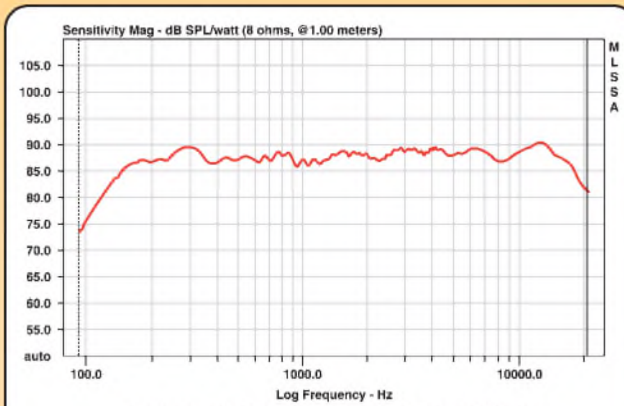

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Neo3

Neo3FP

Neo8

PE Part #	Mfg #	Power Handling	Frequency Response	Dimensions (L x W x D)
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264-720	Neo3FP	20W	2,000-20,000	5-1/8" diameter x 3/4" deep
264-712	Neo8	50W	200-20,000	7-7/8" x 3-1/2" x 1/2"

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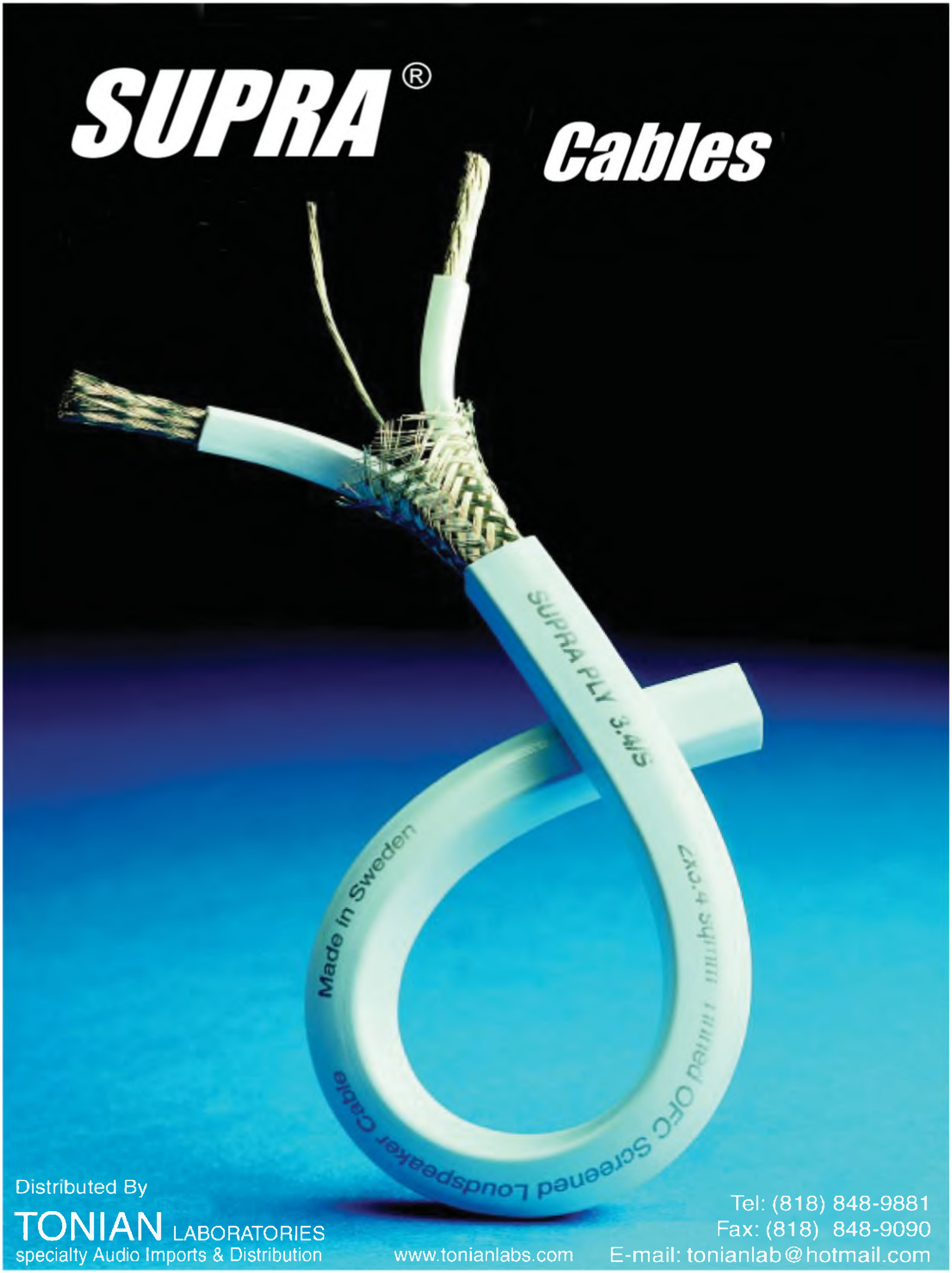
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Editorial

audioXpress: The First Year

Our new, combined periodical, *audioXpress,* became one year old with publication of the December issue. It seems appropriate to take some stock of what has happened during the months since we launched this publication dedicated to hands-on audio technology.

In one sense, *audioXpress* is a return to our roots in *Audio Amateur*, whose first decade covered audio hardware of all kinds. Only with the launch of *Speaker Builder* in 1980 did the magazines begin to become more specialized. *Glass Audio* in 1988 was a response to the growing interest evident in many manuscripts about the renaissance of tube audio.

Our goals for *aX*, resulting from a year of planning, were a monthly which was balanced in presenting all three technologies: solid state, loudspeakers, and vacuum tubes. Article ratios were 2, 3, 2 for each issue, so that over the course of the year we intended to publish 24, 36, 24 articles, respectively, for the three interest categories.

We redesigned the magazine, upgraded the paper, used more color, and set a 100-page per issue target, which we maintained for the first ten issues. We expected circulation to go to somewhere between sixteen and seventeen thousand—although we had no guarantees from our news dealers, for whom this was a "new," untested magazine. I am glad to say we met that target.

In advertising, we had great hopes that advertising clients would stay with us from the three previous titles. If they had done so, we ought to have had some fifty pages of advertising per issue. On average we have been running 27 to 30 per issue. Now the facts of life about ad ratios in monthly periodicals *require* a level between forty and fifty percent. We raised ad rates, fairly modestly, because we were delivering almost twice the circulation level of any of our previous magazines.

Gross advertising revenue for *audioXpress* for 2001 was higher than it was for the three previous titles. This despite some problems we have experienced with sales agencies. I am very grateful to those advertisers who recognized that we were offering an enhanced market for their products and have taken advantage of this. I can understand the hard business decisions others have had to make by keeping budgeted amounts level, and going for smaller space contracts.

The new magazine's production is more expensive than the three previously published titles. Mailing and shipping costs are up significantly, and will rise significantly again in 2002.

It is of primary importance for every reader to absolutely realize that his or her response to the advertisers is crucial to the health of this avocation. Without the advertisers and the products they offer, the audio components we build or buy for music reproduction will not be available. And the magazine will not exist.

Audio reproduction is undergoing massive changes these days. Audio is now being regarded as just one ingredient in communication media: video and home theater, computers and internet music as well as music storage and editing. Some are prophesying that "stereo is dead."

Does a magazine committed to high-quality sound allied to craftsmanship still have a place in the world? We at Audio Amateur Inc., believe that it does. It may become an "old" technology, but the interest in vacuum tube technology and the continuing popularity of long-play disks confirm that older ways of doing things have a continuing life in surprising ways.

We believe that you—in being part of life for the first year of *audioXpress*—have confirmed that craft audio is a vibrant and encouraging success. Although for some, the change was a disappointment, and they have dropped subscriptions. Those have been replaced, and others attracted, so that in each and every month we registered net growth in circulation.

Despite the general downturn in business, we are seeing a small growth in advertising month by month. Given such economic facts, this is undoubtedly a success story. I am grateful to you as subscribers for your response. I thank our advertisers for their participation and believe they are benefiting from the change.

I thank our wonderful staff who have given generously of their intelligence and remarkable skills to make *aX* a reality month by month. It would be impossible to overstate the contribution authors have made to the publication. I am just amazed, pleased, and delighted day after day to receive the excellent offerings of the authors whose work fills our pages.

The year 2002 brings some new adventures for the company. We instituted a subscription rate increase from $29.95 to $34.95. This makes the subscription price less than 50% of the cover price, just under $3.00 per month. The change went into effect January 1, 2002. In addition to our two annuals, *The World Tube Directory* and *The Loudspeaker Industry Sourcebook*, we plan a special for tube lovers: *A Glass Audio Special.* This will feature twelve articles on tube designs and will be available in mid-April.

At the end of 2002, we intend to publish a new, comprehensive equipment directory, covering loudspeakers, wire, crossovers, and equalizers. In 2003, we hope to publish Part 2, covering all other audio hardware equipment offerings.

We are looking forward toward a new year and fresh opportunities for audiophiles and for continued improvement in the quality of reproduced sound—of every kind. I hope you are as encouraged as we are about what is becoming possible in this changing environment. One thing is certain: our love of music is immortal and our commitment to keeping it as beautiful and untainted as possible is strong and resolute.**—E.T.D.** ❖

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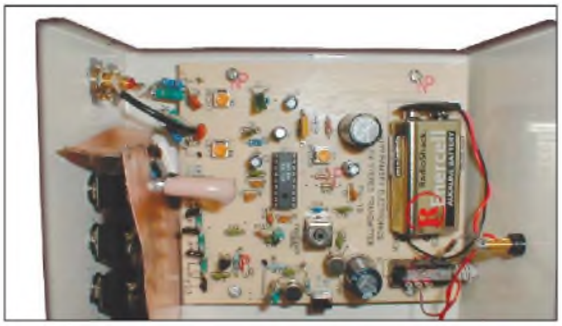

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solid state | tubes | speakers

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The peculiar evil of silencing the expression of an opinion is, that it is robbing the human race; posterity as well as the existing generation; those who dissent from the opinion, still more than those who hold it.

JOHN STUART MILL

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Parts Express Announcements

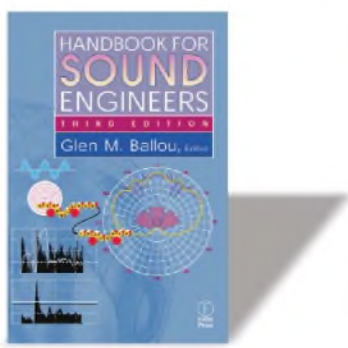
Peerless Denmark appointed Parts Express as a stocking distributor. Parts Express also stocks products from Peerless Denmark's sister companies, Vifa and Scan-Speak, under the Danish Sound Technology A/S umbrella. Parts Express also announces the release of their free comprehensive, 316-page catalog, offering loudspeaker drivers for home and automotive applications, as well as home theater and home automation products, speaker design software, stage lighting, pro-sound telephone products, wire, connectors, and instructional books. For more contact Parts Express, 725 Pleasant Valley Drive, Springboro, OH 45066-0611, 1-800-338-0531, www.partsexpress.com.

Onkyo DW-S500

Onkyo introduces the DW-S500, a high-end desktop CD receiver with slimline speakers and USB input for connection to a home computer. This system, with the Internet streaming audio and MP3 playback capabilities of a PC, features an extruded aluminum chassis, iridescent blue display, and backlit vertical CD player with a motorized door. The CR-N1 Desktop CD receiver is designed for ease of use, with FM auto-scan tuning, 30 AM/FM presets, a full-function remote, two-mode acoustic presence, optical digital output, output for the addition of an optional subwoofer, and output jacks for connection to other analog components, such as TV and VCR. The included Onkyo D-N1 slimline loudspeakers feature a 3.5″ Onkyo Microfiber (OMF) cone woofer and 1″ soft dome tweeter with a powerful neodymium magnet. For more contact Onkyo USA, 18 Park Way, Upper Saddle River, NJ 07450, 201-785-2600, www.onkyousa.com.

Handbook for Sound Engineers

Focal Press announces the release of the *Handbook for Sound Engineers* (Third Edition). This comprehensive reference is available for audio engineers and is written by many of the top professionals in the field, including Glen M. Ballou, Ken Pohlmann, Davis Miles Huber, Dr. Eugene Patronis, Bill Whitlock, Pat Brown, Ray Rayburn, and Dr. Wolfgang Ahnert. The audio industry has seen many changes since the previous edition, particularly in the digital area, all of which are included in this edition. For more, contact Focal Press (an imprint of Butterworth-Heinemann), 225 Wildwood Avenue, Woburn, MA 01801-2041, 781-904-2500, FAX 781-904-2640, www.focalpress.com.



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The Imerge SoundServer S1000 hard disk audio server enables users to store and organize entire music collections and play them back in multiple rooms. Imerge recently expanded its range of S1000 hard disk-based audio servers with the addition of three new models; the S1000-80 features a storage capacity of 80GB (up to 1300 hours of music), and the same storage capacity is available in the other two new models, the S1002-80 and S1003-80, which have two and three discrete outputs, respectively. Also, a software upgrade to be released in the near future will add the ability to access XiVA-Net™, a new Web-based entertainment service for delivering content onto a TV screen through SoundServer. For more, contact Imerge Ltd, +44 (0)1954 783 600, FAX +44 (0)1954 783 601. For more information on XiVA, visit www.xiva.com or www.xiva-net.com.

AVS-632

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CE Distribution Catalog

CE Distribution announces the release of its new 56-page wholesale dealer catalog. CE Distribution serves the guitar and audio markets and offers a full line of parts to dealers and OEMs for replacement and reproduction. New catalog items include replacement parts for Ampeg, Dunlop, and Gibson, as well as tools, effect pedals, expanded offerings of vacuum tubes, sockets, capacitors, and so on. Featured items are Jensen speakers, including the new Mod Series speakers, JJ and Svetlana vacuum tubes, transformers, and other electronic parts, as well as a large assortment of books. For more, contact CE Distribution, 480-755-4712, FAX 480-820-4643, e-mail info@cedist.com, www.cedist.com.

Zen Variations
Part 1: Zen-lightenment

This first in a series on Zen amp design will shed some light on simple amp performance, aided by light bulbs.

By Nelson Pass

PHOTO 1: Light bulb amp.

As an exploration into the potential performance of a very simple amplifier, the Zen amplifier has succeeded in creating notoriety and some controversy over the last eight years. More important, its novel and simple construction appears to have encouraged a large number of do-it-yourselfers to take up a soldering iron and jump in.

ZEN-TRODUCTION

Having only a single gain device, the design's name is a pun on the Zen Koan, "What is the sound of one hand clapping?", but the point is quite serious. High-quality sound can be obtained with simple and accessible circuits. Conversely, it is quite easy to design a complex circuit which sounds subjectively lifeless or even irritating.

This is part 1 of the Zen Variations, and each part will illustrate one of the many ways to build a single-stage audio amplifier. There are many possibilities here; I recently counted several hundred permutations. After considerable meditation, I winnowed these down to approximately 30 interesting and nontrivial examples, which we will explore one at a time in no particular order.

THE ORIGINAL ZEN AMPS

Figure 1 is the simplified schematic of the original Zen amp. Here a single gain device, a power MOSFET, operates in common source mode, where the input comes into the Gate of the device and the output, which drives the loudspeaker, comes out of the Drain. The MOSFET is biased by a constant-current source from the positive supply, and a small network of resistors and capacitors sets up the operating voltages and provides feedback. You'll find the entire article in *Audio Amateur* 2/94, with a revision in 3/94. It is also available on-line at www.passdiy.com.

I followed up the Zen with Son of Zen (*Audio Amateur* 2/97), whose unsimplified circuit is in *Fig. 2*. Here a single gain stage is formed by a differential pair of identical devices, allowing greater simplification and the removal of coupling capacitors and negative feedback. Both these designs also inspired preamplifying circuits based on the same topologies, giving Bride of Zen (*Audio Amateur* 4/94) and Bride of Son of Zen (*Audio Electronics* 5/97).

LET THERE BE LIGHT

Some of the DIYers who built the Zen amp were put off by the complexity of having a constant-current source comprised of two additional transistors and several resistors. However, you can replace a constant-current source with a high-power resistor in the Zen amp, if you are willing to use a higher supply voltage and a resistor with a high enough value to simulate a constant-current source for practical purposes.

At the same time, others complained of the difficulty of obtaining the high-power resistors required for the Son of Zen. I received a few such complaints, which eventually became a source of inspiration.

And so the light came on and I had a bright idea: A light bulb is a power resistor which is conveniently obtained and which can dissipate large amounts of power without a heatsink. So what kind of light bulb might be appropriate?

I bought all sorts of samples of light bulbs. A little measurement revealed that common household incandescent bulbs are quite useful for this purpose. *Figure 3* shows the current versus voltage for a couple of 120V lamps—one rated at 150W, and the other at 300W. The resistance (in ohms) of each of these at any point is simply the volts divided by the amps.

SCHEMATIC LITE

Figure 4 shows the simplified schematic of a Zen amp using a light bulb instead of a constant-current source. The input goes into the Gate and comes out the Drain. The two resistors R1 and R2 set up the bias voltage for the Gate, which must operate at approximately 4V above the Source, which is grounded. The two resistors also provide feedback for gain control and distortion reduction.

The Drain voltage of this circuit is the operating Gate voltage (Vgs) times (R1+R2)/R1. In this case with R2 = 2 × R1, and Vgs of 4V, the Drain voltage is set at 12V DC. This ratio of 2:1 of resistance also sets the gain of the circuit, which is approximately R2/R1, or 2X, which is 6dB.

This is a perfectly workable circuit with a 200 or 300W light bulb and a clean 40–80V supply. It has a couple of small limitations: It requires a signal source which is direct coupled and which can sink about 4mA, and a load which doesn't mind 12V of DC. These conditions can be met in quite a few cases, but are not to be counted on for most systems.

Another limitation is that the AC gain setting is the same as the DC bias point, and it is often convenient to be able to separate these two settings.

FURTHER SCHEMATIC ILLUMINATION

Figure 5 shows a real schematic which remedies these shortcomings. The supply voltage is made quieter by inductor L1 and capacitor C3, which remove noise from whatever supply is represented by V+. C2 capacitively couples the input so that you don't need to worry about the DC characteristic of the signal source. C1 capacitively couples the output to the speaker so the woofer cone doesn't fall out onto the floor. R1 and R2 are still there, and they set the DC bias point of the MOSFET Q1.

A potentiometer P1 (Digikey 381N252-ND) and two new resistors R3 and R4 set the AC gain of the amplifier, which is variable from approximately –20 to +20dB referenced to the input voltage. The feedback loop represented by this potentiometer encloses the input and output capacitors of the amplifier and imposes correction on distortion they introduce, in addition to distortion offered by the MOSFET, the light bulb, and noise from the power supply. R3 and R4 have been introduced to limit the lowest possible impedance that can be seen by the Gate of Q1 to avoid parasitic oscillation.

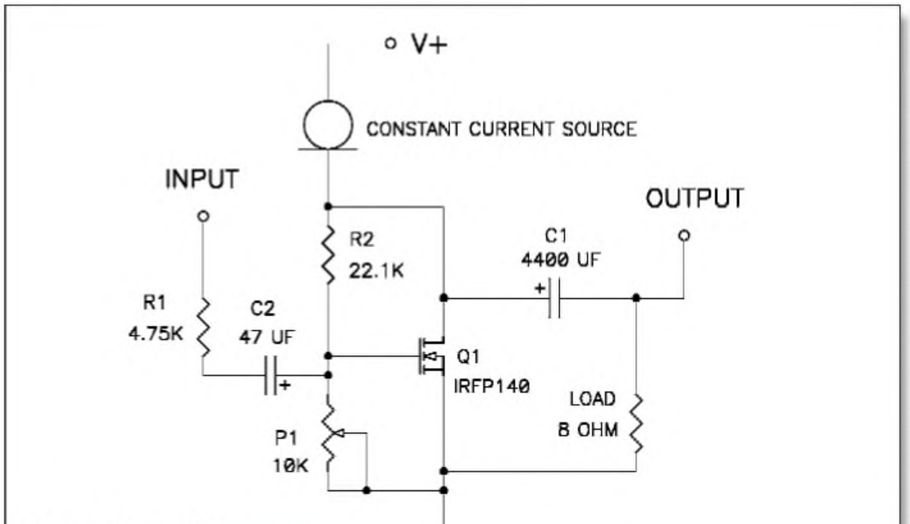


FIGURE 1: Simplified Zen amp.

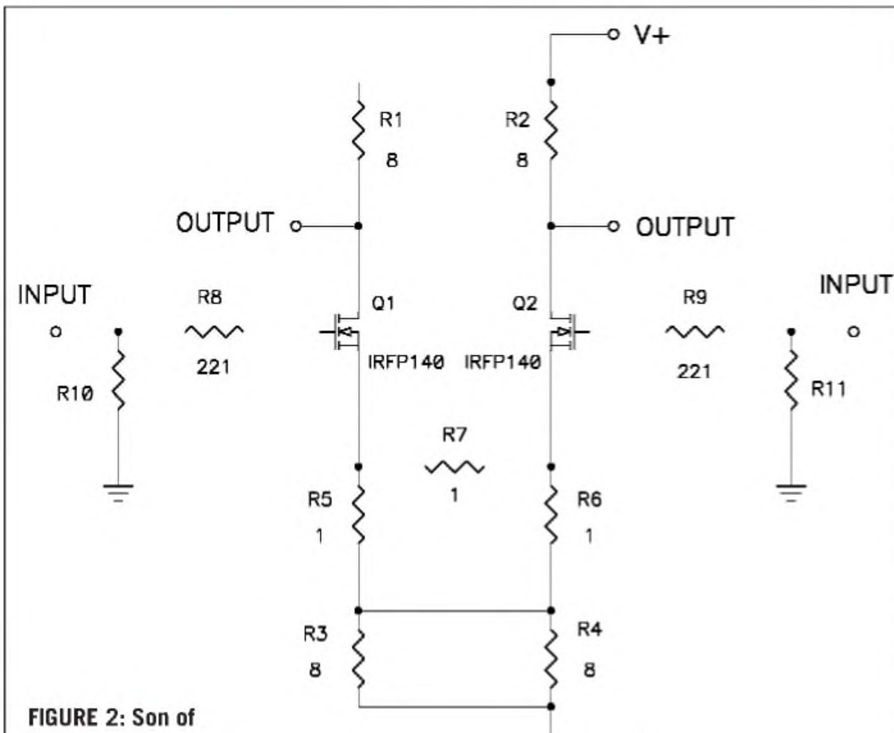


FIGURE 2: Son of Zen.

For you rabid audiophiles, the electrolytic capacitors are bypassed with 3µF film types. I used 3µF Axons from Orca Designs (www.orcadesign.com), but you can get comparable Panasonic parts from Digikey (www.digikey.com).

None of the values here are critical. The electrolytic capacitors should be voltage rated at the supply voltage, which in this project can vary from 40–80V DC. All resistors are ¼W. All the parts are available from the Digikey catalog except for the incandescent lamp and inductor L1. For L1, I used either MCM #50-1080 (www.mcmelectronics.com) or the ERSE 4.0mH/14 gauge (www.zalytron.com). I tested both up to 6A DC current without seeing loss of inductance, and I believe the ERSE will do more.

Figure 5 shows the use of a 300W Sylvania bulb. You will later see two such bulbs in parallel for more bias current. Depending on the supply voltage and the dissipation capability of the heatsinks, you can place more bulbs in parallel to get the desired bias current. Can't find a 300W bulb? Two 150W bulbs in parallel behave virtually the same, and two 200W bulbs are perfectly workable for a little more bias current.

The light bulbs were set in standard Leviton sockets found at the local hardware store. The clear 300W bulbs came from the McMaster-Carr catalog (www.mcmaster.com). I believe you can get the sockets there also.

I have specified the IRFP240 MOSFET transistor in *Fig. 5*, but a wide range of similar N channel devices will work fine. The best performance is obtained with the IRFP040 or IRFP044, but unfortunately they seem very difficult to get. An in-between alternative is the IRFP140. When substituting other MOSFETs, the Gate to Source pin operating voltage becomes an important consideration. You will want a Drain voltage of about 12–15V, and if the Vgs of the MOSFET chosen is very different from about 4V, you will need to adjust the value of R1. Decreasing R1 will raise the Drain voltage, and of course increasing R1 will lower the Drain voltage.

MOSFETs are sensitive to damage with high Gate voltages, most specifically static electricity. Always take modest care to avoid static discharge when handling the parts. Personally, I take almost no precautions, and I hardly ever have a problem, so don't become too worried about it, particularly once you wire the transistor into the circuit.

Unlike some previous projects with MOSFETs, there are no zener protection diodes on the input of the amplifier. This again means that you should use some modest care when attaching an input cable, but this simply means touching the ground connections first. If both the source and amp are earth grounded via the wall outlet, you don't even need to think about it.

You may ask, Is there a PC board layout for this project? No. The small quantity and odd nature of the parts lends itself better to point-to-point construction, and that is how I have built several of these. Keep the wire lengths down around 6″ or less and terminate all the grounds of *Fig. 5* at one point. If you build an unregulated supply for V+ and use a large electrolytic capacitor for its filter, then connect all of its ground connections at a separate single point.

Connect these two ground points, the main supply and *Fig. 5*, with a fat piece of wire. In this manner, the large current pulses going through the main electrolytic of the unregulated supply will not pollute the nice clean signal grounding point of *Fig. 5*. Later we will look at a sample unregulated supply.

SPOTLIGHT ON PERFORMANCE

Figure 6 shows the total harmonic distortion plus noise of the circuit of *Fig. 5* versus output power into 8Ω. The gain has been set at 10dB, and the supply voltage has been varied from 40–80V. The lowest distortion accompanies the highest voltage, and this effect is due to the higher bias current going through the MOSFET. In general, the higher the bias current, the lower the distortion.

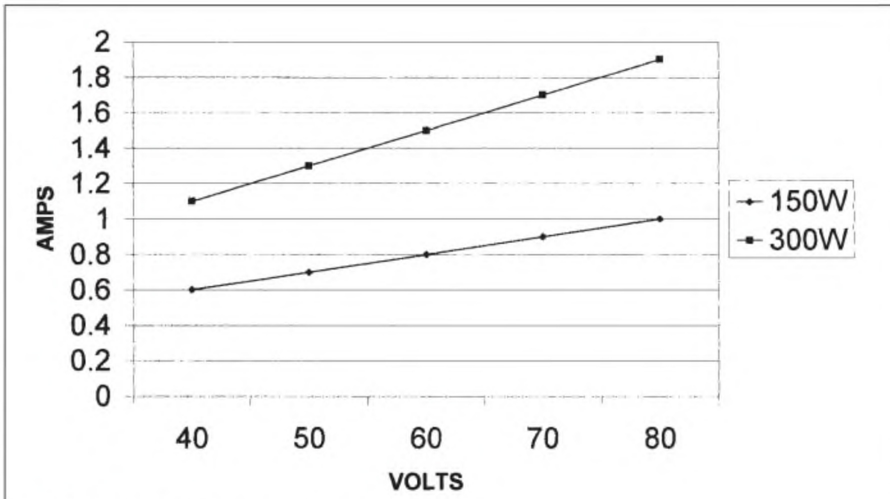


FIGURE 3: Light bulb current draw.

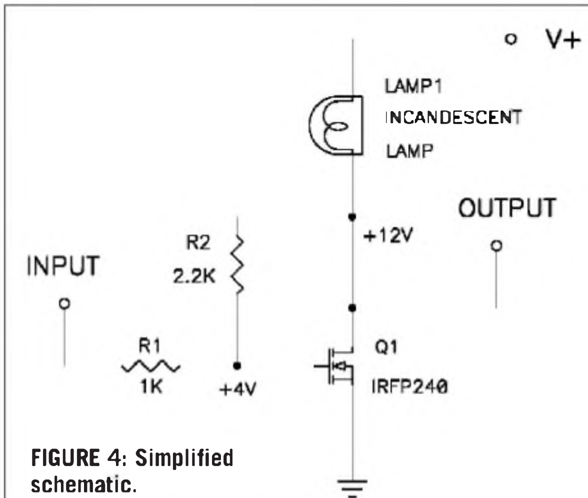


FIGURE 4: Simplified schematic.

Figure 7 shows this same amplifier with distortion versus frequency and the supply voltage fixed at 60V and an output level of 1W. This is quite good, as the distortion rises only slightly at 10Hz and 20kHz.

Figure 8 shows the distortion as in *Fig. 6*, but with a 16Ω load. For a given output voltage (not wattage), notice that the distortion is nearly half. This is expected, as the dominant source of distortion is variation in current through the gain device.

Keep in mind that the Zen amp is a polarity inverting circuit. To keep proper signal polarity, the speaker + terminal goes to ground.

EVEN MORE INCANDESCENCE

It is practical to increase the current through the MOSFET—and improve the performance—by paralleling lamps. *Figure 9* shows the harmonic distortion of the circuit of *Fig. 5* with twin 300W lamps in parallel. The distortion drops substantially and there is more power.

Figure 10 shows a further improvement available with a different MOSFET, the hard-to-find IRFP040, where the distortion is seen to drop by about one-third.

Figure 11 documents three distortion versus frequency curves for the IRFP240 (top curve at 1kHz), the IRFP040 (bottom curve), and matched parallel IRF240s (middle). Here you see that the IRFP040 is clearly the best choice, but note that its distortion at the highest frequencies is about the same as the IRFP240. The source of distortion rise at high frequencies is the nonlinearity in the input capacitance of the MOSFETs. If you parallel devices you will see the greater capacitance and the kind of effect illustrated in this curve, where the distortion is lower at lower frequencies, but higher at the top end.

This phenomenon tells something about the choices of MOSFETs for this circuit. Comparing devices such as IRF230, 240, and 250, you'll see current

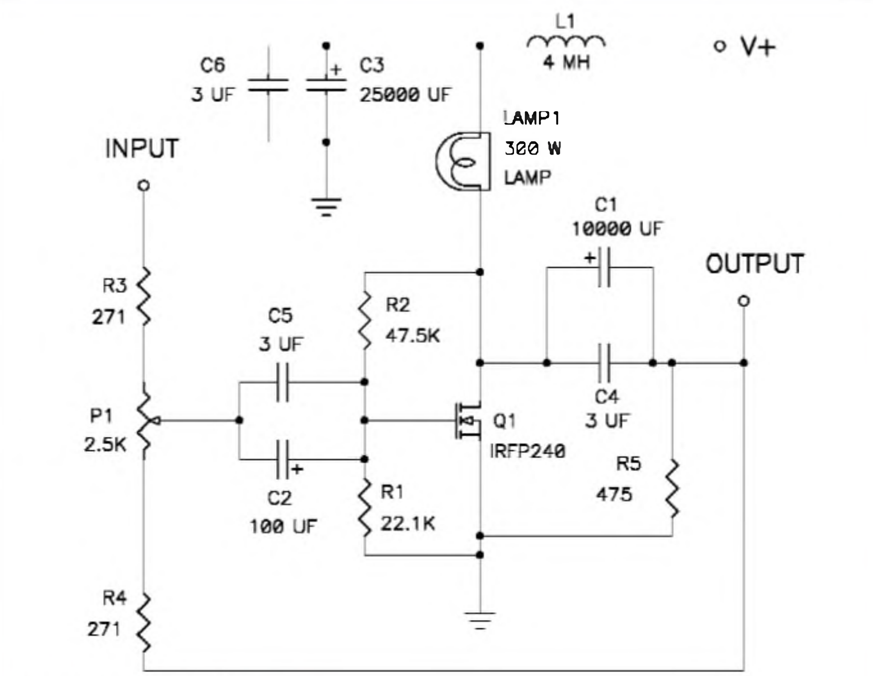


FIGURE 5: Actual schematic.

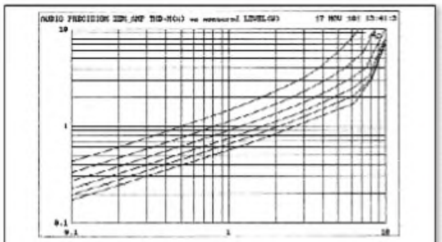


FIGURE 6: THD versus output/40–80V/8Ω.

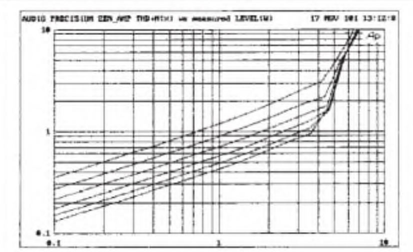


FIGURE 8: THD versus output/40–80V/16Ω.

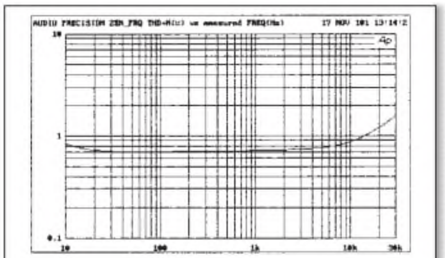


FIGURE 7: THD versus frequency/60V/8Ω.

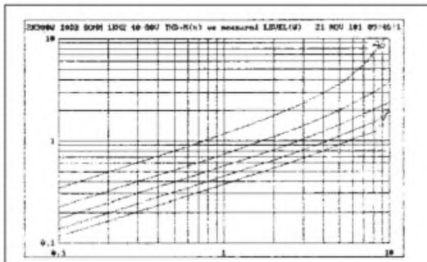


FIGURE 9: THD versus output/40–80V/dual bulbs.

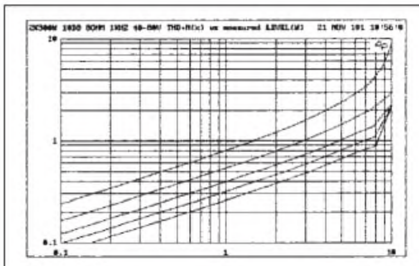


FIGURE 10: THD versus output/40–80V/IRFP040.

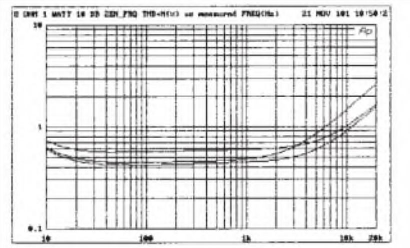


FIGURE 11: THD harmonic distortion versus frequency/60V (see text).

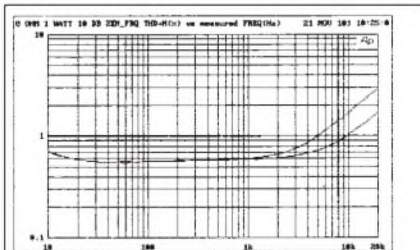


FIGURE 12: Distortion versus frequency versus loop impedance.

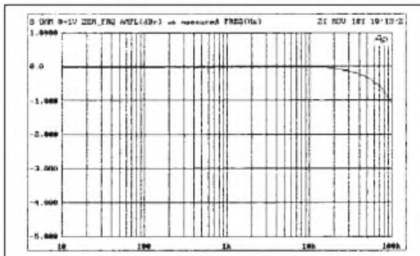


FIGURE 13: Frequency response.

ratings of 9, 18, and 30A, as well as capacitances of 600pF, 1,300pF, and 2,600pF, respectively. These figures are proportional to chip size, and from actual testing you'll get similar performance from two IRF230s in parallel as compared to a single IRF240. Also, the IRFP250 behaves similarly to a pair of IRF240s.

In general, the bigger the chip, the better performance through the low and mid frequencies, but the higher distortion at the top end. This distortion is affected by the impedance of the network driving the gate of the device, and as this impedance goes up, so does the distortion at high frequencies.

Figure 12 illustrates this with comparative performance with distortion curves for a 2.5kΩ potentiometer value versus 5kΩ value. The curves diverge a bit above 2kHz, and at 20kHz the distortion of the 5kΩ pot is nearly twice.

The lower voltage chips of the same type will give lower distortion, apparently due to their lesser sensitivity to voltage fluctuations on the Drain. They are generally the preferred parts.

For some unknown reason, the Zen amplifiers have been criticized as having low bandwidth, but I have not seen a case where this has been true. *Figure 13* documents the frequency response of this Zen amp, showing –1dB at 100kHz.

The damping factor of the Zen amp is quite low, reflecting the paucity of neg-

ative feedback in the design. With the IRFP240 devices, the damping factor was about eight referenced to 8Ω, and about 12 with the IRFP040s.

The circuit of *Fig. 5* has a substantial turn-on thump as the voltage across the MOSFET goes from 0 to 15V, and some of your more sensitive speakers might take offense. *Figure 14* shows a simple turn-on delay circuit, which will short the output of the amp to ground for a few seconds. The relay in this case is #Z789-ND from Digikey with the switch contacts wired normally closed. The transistor is a generic NPN device rated at 1A and 100V. When the power supply is applied, the capacitor in the circuit slowly charges up until it turns on the transistor, which opens the relay switch.

You will notice a resistor in series with the relay coil to adjust the voltage across the coil to its rated voltage versus the supply voltage. If you go with the 80V supply, this resistor will probably need to be 5W, but with 60V you can use 3W. Alternatively, you can simply use a switch to short the output to ground manually.

Two other solutions: A lower value of output capacitor, or a power supply which turns on slowly. The circuit does not have a substantial turn-off thump, and so there is no need to address that situation.

LIGHTLY BALANCED

You can operate the Zen amp in balanced mode if you have a balanced source, such as the output of the Bride of the Son of Zen (*Audio Electronics* 5/97). *Figure 15* shows how you can accomplish this using two identical channels sharing a common supply. (If you are employing both separate Ch 1 and Ch 2 of *Fig. 16*, you need to attach the V+ and ground connections of the two channels to each other.) In the stereo amplifiers I built, I put an XLR connector on the amplifier and wired the ground to pin 1, and the two inputs to pin 2(+) and 3(–) for use as a balanced input.

The loudspeaker is attached across the outputs. The advantages to this circuit are mainly cancellation of second harmonic distortion and common rejection of any noise from the supply.

By the way, you will probably not need the turn-on thump circuit with bal-

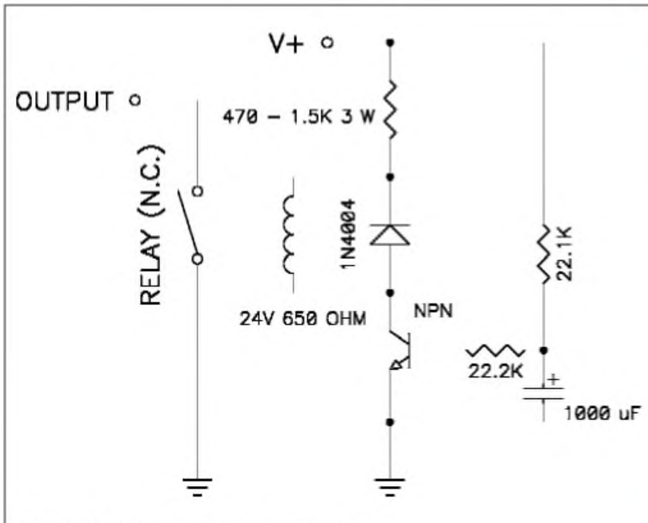
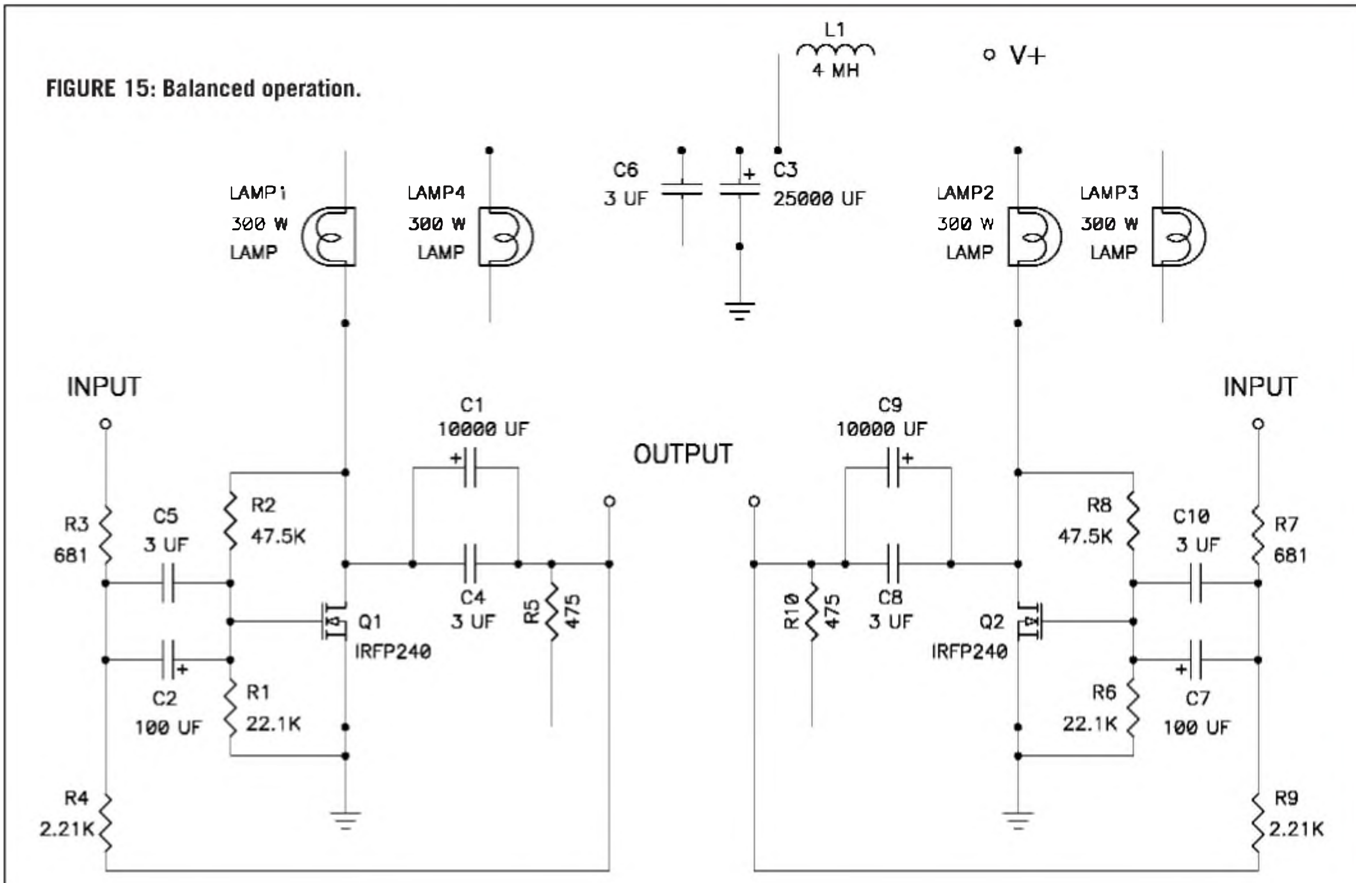


FIGURE 14: Turn-on thump circuit.

FIGURE 15: Balanced operation.



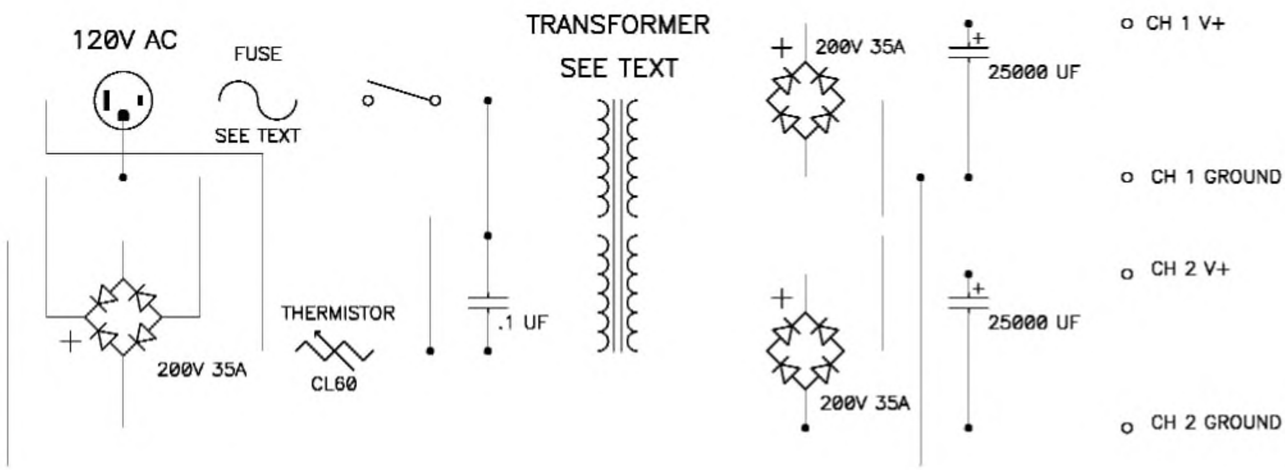


FIGURE 16: Power supply.

anced operation, but it won't hurt. Another potential benefit to balanced operation is the elimination of the output capacitors, as illustrated by the circuit of *Fig. 17*. While the 15V or so appears at the loudspeaker, it is the same on both sides, and thus is zero from the speaker's standpoint. It is important to have these two DC voltages equal, and if you match Q1 and Q2 well, the offset is about 30mV or so. Lacking that fine degree of matching, I have provided potentiometers P1 and P2, which allow fine adjustment of the DC value.

The performance of this circuit with a single pair of 300W bulbs is shown in *Figs. 18* and *19*. Here you can see the fairly dramatic distortion reduction due to cancellation of second harmonic.

Put two 300W bulbs on each side, and the improvement becomes more pronounced, as seen in *Figs. 20* and *21*. With an 80V rail the amplifier now makes it up to 20W. *Figure 22* shows the frequency response of this circuit, which is down 2dB at 100kHz.

I built some with three 300W bulbs in parallel also: BWAAA-HAHAHAA!

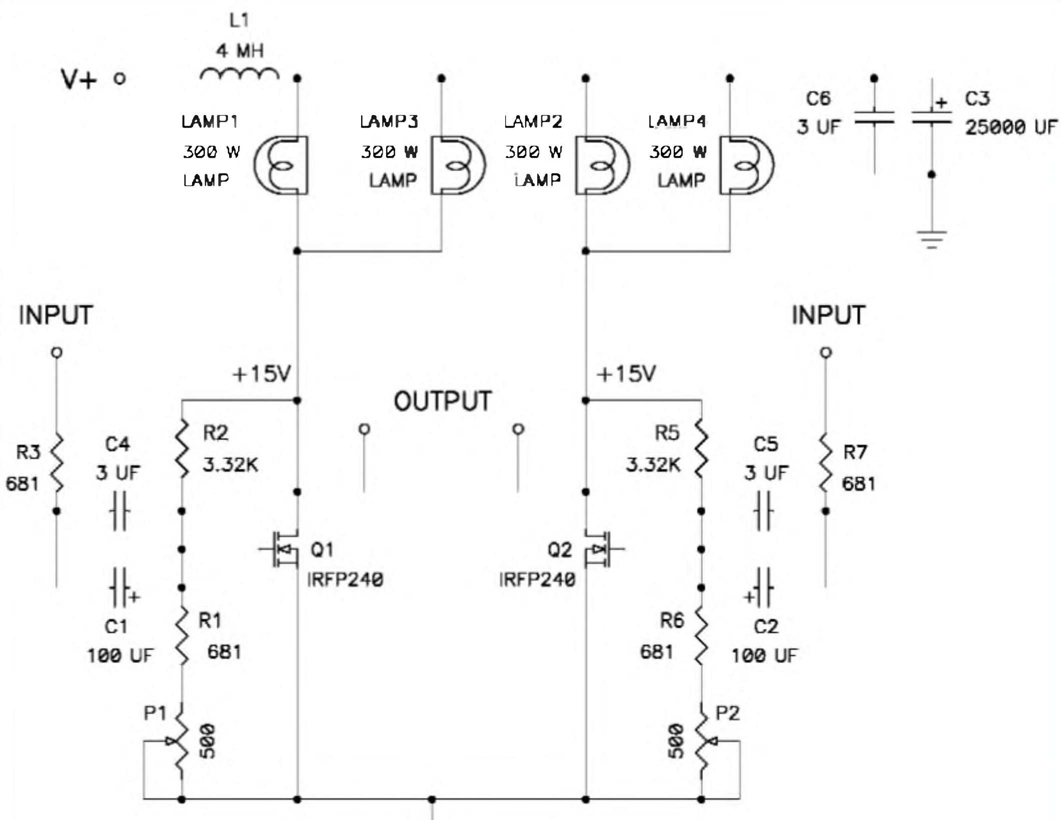


FIGURE 17: Balanced operation without output caps.

ENERGY CONVERTER FOR PHOTONIC EMISSION BIAS BULBS

Figure 16 shows an example of the unregulated supply. This is a standard sort of circuit, and the caveats and comments from the other Zen amps apply here. Several of the choices in this circuit will depend on the supply voltage and the number and type of light bulbs you will be using.

The fuse should be a slow-blow type set as low as possible; that is, one value higher than a value which pops often. Start at 2A and go up. The on/off switch should be a heavy-duty type, rated at 15A or better for reliable operation.

I have placed a thermistor in the circuit (Digikey KC006L-ND) which reduces inrush current. The .1µF capacitor across the line must be line voltage rated, and this one is Digikey P4603-ND. Both the thermistor and line capacitor are optional, as is the power switch. The fuse is not optional.

The bridge rectifiers are simply generic high-current types. I recommend 200V 35A types, but I know that a certain number of you are going to run out and get exotic Schottky or Ultra-High-Speed-Recovery or Ultra-Soft-Recovery types. Feel free to do so, and consider some nice RC snubbers across the pins while you're at it. You will need to heatsink these bridges, by the

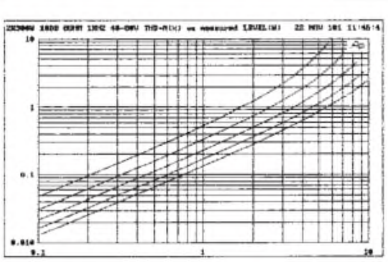


FIGURE 18: THD versus output/40–80V/balanced operation.

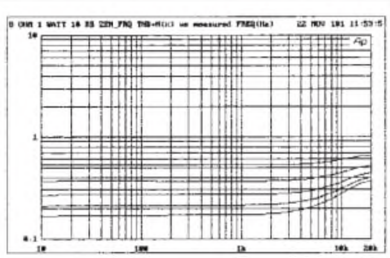


FIGURE 19: THD versus frequency/40–80V/balanced operation.

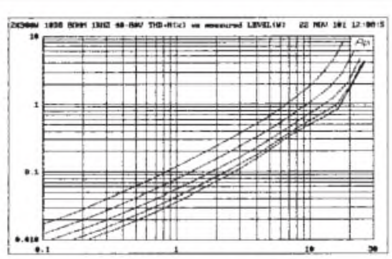


FIGURE 20: THD versus output/40–80V/balanced operation/2 bulbs.

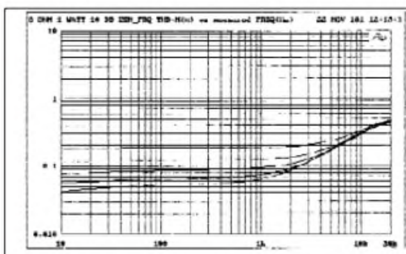


FIGURE 21: THD versus frequency/40–80V/balanced operation/2 bulbs.

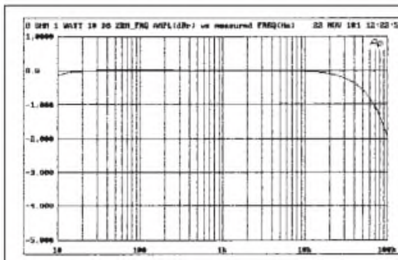


FIGURE 22: Frequency response/balanced operation.

PHOTO 2: Rear view.

way. I simply mounted mine on the same sink as the output transistors.

The power-supply capacitors are whatever big cans you want to get that are rated at the voltages you will be using. Big is good, but 1F is too good.

You will notice in *Fig. 16* that I have shown a supply for two-channel operation which isolates the secondary supplies from each other. This is a nice touch, but not essential, and if you are planning on operating the two channels balanced, you will need to connect the V+ and ground connections together at the main capacitors.

As shown, the two supplies are ground isolated, which helps eliminate any ground loops. A connection to earth ground at the wall outlet is provided for safety and goes through a power rectifier bridge. This bridge will conduct at voltages above .6V, providing shock protection while raising a small voltage barrier to noise.

The transformer is a more tricky choice. Many of the commercial stock devices from Avel or Plitron or Toroid correspond to this configuration, but you must select the voltage and power rating. The voltage is easy. Figure out what DC voltage you want for the supply rail and select a secondary winding which is AC rated at 75–80% of that value.

For figuring out the VA rating, use *Fig. 3* to determine the current drawn by a single bulb for the DC voltage you have chosen (interpolate for a 200W bulb). Now multiply this current by the voltage and then by the total number of light bulbs and then by two. This is the VA rating of the transformer. The factor of two times the actual power draw of the circuit covers the "power factor" of the AC to DC conversion and allows a little more for margin. You are encouraged to consider this as a minimum figure.

For example, consider using a 60V supply with two channels having two 300W bulbs each. At 60V the 300W bulb draws 1.5A:

$$\text{VA rating} = 1.5\text{A} \times 60\text{V} \times 4 \text{ bulbs} \times 2$$
$$\text{VA rating} = 720\text{W. Minimum.}$$

If this is too high, use two transformers, one for each bridge.

As always, remember that you are dealing with potentially lethal voltages. If you don't know what you are doing and are unwilling to take personal responsibility for the results, get some technically competent help.

You are welcome to consider using batteries to power this circuit. This can be accomplished by stacking them to get the appropriate voltage. If you take this approach, you can scrap *Fig. 16* altogether, and also you can eliminate the inductor L1 of the main circuit. Use whatever capacitor to ground for filtering that you like, but I recommend at least the 3µF film cap.

CONCLUSION: SOME LUCID OBSERVATIONS

Photos 1 and *2* are of the actual amplifier. I made the chassis out of baltic birch plywood and finished it with a couple coats of lacquer.

This amp sounds pretty much like the earlier Zen amp. Over the years numerous challengers have opined that their favorite amplifier, (fill in the blank), sounds much better. Probably does.

As I explained first time around, this is an exploration into simplicity, and I make no claim that the Zens are the best-sounding amplifiers you can make. They do sound pretty good though, and they are a lot of fun.

If you need an argument, I'll tell you that Roger Corman and Stanley Kubrick are the greatest film auteurs of the 20th century. ❖

Special thanks to Desmond Harrington for layout and production, and Karen Douglass for the photos.

Building a Highly Analytical Horn System

Combine the Altec 805B horn with the 288-8K driver and ES subwoofer for outstanding sound. **By Louis W. Erath**

I use a high-quality set of loudspeakers (*Fig. 1*) in my stereo system, which I designed and built 25 years ago. I recently acquired a pair of Sennheiser HD600 headphones, and I have been comparing the sound of the loudspeakers with the sound of the headphones. The loudspeakers compare favorably with the headphones, except that the latter exhibit more fine detail, such as the rattle of the snares on the snare drum or the harmonic texture of horns playing together.

My theory about why this happens is that the sound from the headphones is 100% direct to the ears, while that of the loudspeakers to my listening position (12′ away) is perhaps 50% direct, with the balance reflected from the floor and walls. I don't think my 11′ ceiling contributes much in the way of early reflections, which are delayed 2-5ms, and combine with the direct sound to cover its fine texture. To prove this, I placed my loudspeakers 3′ apart facing each other and listened with my head between them at an SPL (sound pressure level) of about 80dB. Voila! Most of the fine structure was there.

So, what are my options for reducing the early reflections? The floor already has a thick carpet and pad on it, and to place sound-absorbing panels along the walls would interfere with the normal use of the room—maybe loudspeakers with some degree of directivity? Funny that I should think of that, because I just happen to have some Altec 805B horns and 288-8K drivers that have a directive 3dB contour of 80°H (horizontal) and 40°V (vertical). Since these are not easy to come by, I will tell you how I got them.

Last year, I acquired a large number of 288-8K drivers from a commercial sound company in Houston, Tex. Gerald Ognibene, of Marquis Electronics here in Abbeville, helped me dispose of them through sales and barter. In the process, we acquired a variety of multicellular horns. We kept a pair of drivers for my use, a pair for Gerald, and one spare.

Marquis Electronics custom-builds loudspeakers using my ES woofer, and we decided to jointly do this project. Gerald would build the loudspeakers and I would develop and build the crossover networks and electronics. And while I did not expect it, I needed to finish building a vacuum tube amplifier as well.

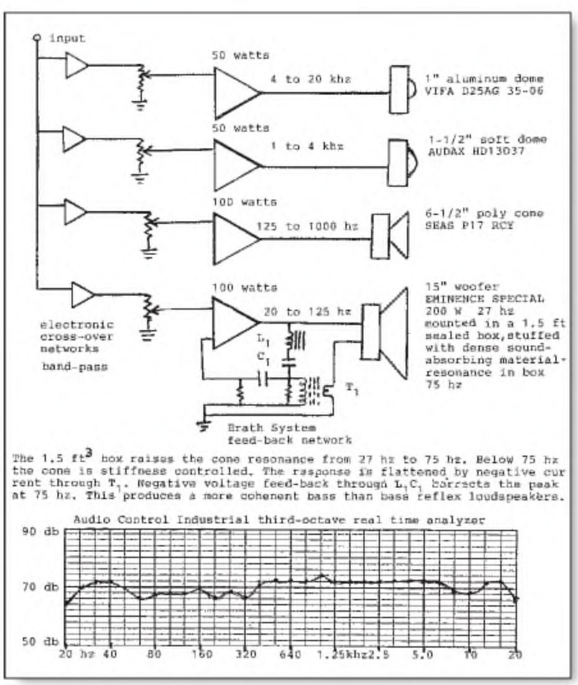


FIGURE 1: Author's existing sound system.

PHOTO 1: Finished horn system.

PREPARING THE HORNS

The horns were cleaned and the sheet metal was straightened and refinished to look like new. I measured a free-field response on the horn and driver to gain the information needed to design the high-pass crossover network. I used a B&K (Bruel & Kjaer) condenser microphone that is flat from 20Hz to 100kHz and a General Radio curve plotter.

ABOUT THE AUTHOR

Louis W. Erath studied physics, vacuum tube theory, and radio engineering at Louisiana State University. For 40 years he was involved in electronics for geophysical prospecting for oil in Houston, Tex., where he is currently a consultant. He also worked on naval ordnance at the NOL in Washington, D.C. and the Bell Labs in Summit, N.J. He holds over 100 US patents, is a Life Fellow in the IEEE, and is involved in amateur radio (W5BM).

After fabricating a preliminary network, I assembled the horns, drivers, and woofers as shown in *Fig. 2. Photo 1* shows the completed system; *Photo 2* pictures the unit with the grille removed, and *Photo 3* is the back of the system. I conducted the preliminary listening tests using an amplifier claiming 125W/channel at 8Ω with no more than .005% THD (total harmonic distortion) to drive the horns. I used a pair of 60W amplifiers to drive the woofers, which worked fine, but the sound from the horns was very bad! I borrowed another "good quality" transistor amplifier from Gerald, but it sounded even worse.

PHOTO 2: System with grille removed.

I knew what the problem was. I had measured the SPL at 80dB in the listening area using pink noise from an Audio Control analyzer. With both speakers operating, I found that it required only .1V across the voice coil of the 288-8K drivers as measured on an HP (Hewlett-Packard) RMS AC voltmeter. If power = E^2/R, then $.1 \times .1/8 = .00125$W, or 1.25mW.

I realized that the horns were like an acoustical microscope, or at least an acoustical milliscope, looking back at the amplifier's output characteristic. What the horn drivers saw was a ragged crossover region down at the milliwatt level. Feedback cannot completely cover one transistor switching off while the other transistor switches on; the crossover region will never be completely smooth. What was needed was an amplifier without a crossover at all!

The quickest way for me to achieve this was to finish a 75%-complete class A vacuum tube amplifier on my workbench. In a few hours, I had a 25W/channel class A SET (single-ended tetrode) amplifier with a measured .1% THD at 8W into 8Ω and .03% at 1W. I suspect that I could do as well or better using class A MOSFETs (metal-oxide substrate field-effect transistor), but that will come later.

After I installed the vacuum tube amplifier, the horns sounded great, and I proceeded to listen and tweak the crossover networks for the next several days. The component values I am currently using are those shown. I put in the 10μH coil and the 10μF capacitor, shunted by the 10Ω resistor, to correct

PHOTO 3: Rear of unit.

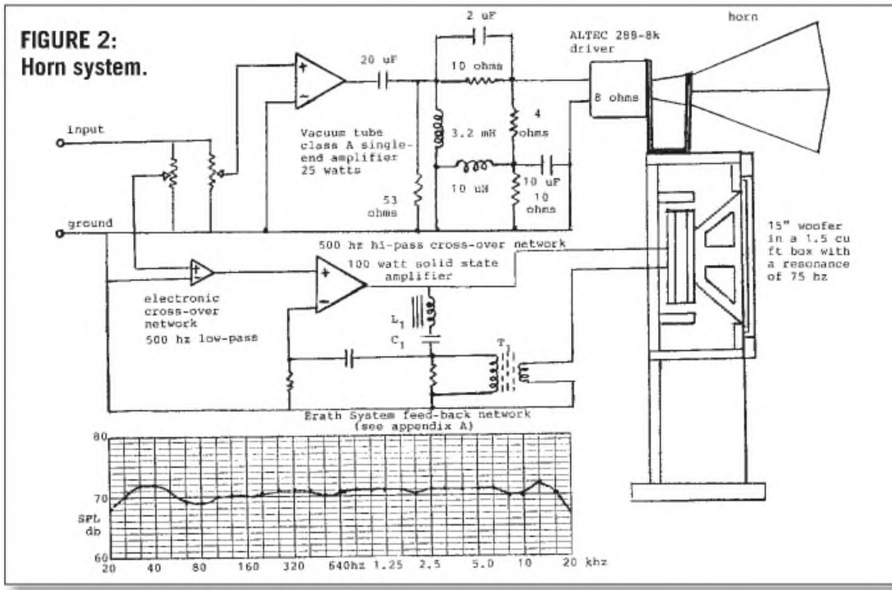
for a dip at 12.5kHz. The peak shown at that frequency is the result. You could reduce the 10Ω resistor to flatten this peak, but it sounds good as it is.

THE ES WOOFER

The ES woofer is not new! I developed it in the '60s and was issued a US patent on it in 1969. My design objectives were to develop a woofer having ample output in the frequency range from 20Hz to 500Hz for the average room, and to produce a more coherent sound than the typical "resonance assisted" woofer.

I had CTS (Chicago Telephone Supply) build a number of 15″ 200W woofers with a voice coil substantially longer than the pole-piece face and with no

FIGURE 2: Horn system.



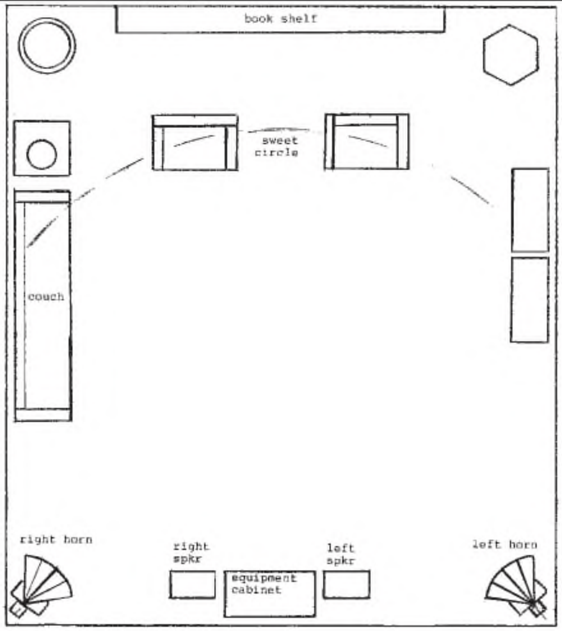


FIGURE 3: The listening room (15′ × 17′ × 11′).

cone breakup below 1kHz. I mounted this 27Hz woofer in a sealed box designed to increase its cone resonance to 75Hz. This requires a 9:1 increase in stiffness. The stiffness of the air behind the cone is very linear compared to that of the suspension of the cone and voice coil. This linear stiffness swamps out the third harmonic distortion produced by the cone suspension for large cone excursions.

In *Fig. 2*, L_1C_1 is adjusted to eliminate the resonance peak at 75Hz and it also happens to minimize any third harmonic distortion generated by large voice coil excursions in the 20 to 30Hz region. Below 75Hz, the movement of the cone is stiffness-controlled and the frequency response is flattened by the negative current feedback through T_1. The ES (Erath System) woofer has a good tight bass sound without the use of a subwoofer, keeping the low bass coherent and out of the "mud."

The term, coherent, as I use it here, is measured on a scale of 0 to 1 and means that if the waveform of the sound is exactly the same as that of the electrical signal producing it, the coherence is one. If they are totally different, the coherence is zero.

ANALOGY OF THE RADIATION OF SOUND TO THE RADIATION OF LIGHT

Loudspeakers which use direct radiator elements, such as in my existing sound system (*Fig. 1*), essentially radiate sound into a hemisphere, in all directions, much as a light bulb mounted on the surface of a flat vertical panel. The horn system (*Fig. 2*) does not radiate in that sense, but projects sound to the listener, similar to eight flashlights mounted like the eight horns in the Altec 805B. Very little of the light or sound is directed to the floor, walls, or ceiling, but the listener sitting in the "sweet circle" is fully "illuminated."

This directivity produces such a strong soundstage that the speakers may be placed in the corners, 15′ apart without making a "hole in the middle." When playing mono CDs, the sound seems to come from the equipment cabinet (*Fig. 3*). The transient response is excellent and the detail in the sound structure is outstanding. ❖

Big Mike and the Jimmy, Part 2

Last month, you were introduced to Big Mike, a tube-based microphone preamp design. This month we meet his power supply. **By Paul J. Stamler**

PHOTO 3: Power supply for Big Mike preamp.

Let's look at the bill of particulars for a high-quality power supply.

- The supply, obviously, must provide adequate voltage and current to the active stages. In a tubed circuit, this means plate voltage, filament voltage(s), and, for a mike preamp, voltage for the phantom-power circuit.
- The supply (or supplies) must have a low enough impedance—at all frequencies within the tubes' operating range—that the voltage does not sag when current is drawn. Sagging supplies can cause audio stages to sound muddy or distort; they can also produce crosstalk between stages or channels. In extreme cases, they can make the audio circuits unstable.
- The supply must keep AC line garbage from leaking into the audio circuits. This can include radio frequency interference (RFI) from neighboring broadcasting stations, CB and ham radios, taxi dispatchers, or wireless appliances. It can also include hash from fluorescent lights and SCR dimmers, digital junk from computers or digital audio gear, and sudden jumps when the fridge turns on. Line garbage can, of course, show up in the audio—you've all heard radio stations or CB calls in phono circuits or guitarists' stomp boxes.

More insidiously, RF signals can intermodulate with audio to produce distortion—less likely with tubes, but it still can happen.[5] You should also remember that one of the worst sources of RF garbage can be a power supply's own rectifier diodes, as they switch on and off.

OVERKILL

In designing high-end equipment, I think power-supply overkill is appropriate. I remember reading an interview with one of Krell's design engineers. He said, to paraphrase, that every improvement in the supply produced an improvement in the audio quality—they hadn't yet reached a point where that was no longer true.

I built the raw supply outboard (*Photo 3*). This meant that I wouldn't need to worry about radiated hum fields in the case with my input circuits. It also let me drive several (separate) preamp units from a single supply, and left plenty of breathing room in the preamp cases themselves.

As readers of my first preamp article will remember, I've become a disciple of Ben Duncan; I design power supplies to minimize the possibility of RFI leaking into the audio circuits.[6] The family of rectifier diodes I use has been shown to produce significantly less switching noise than their standard-issue counterparts.[7] Input resistors help damp resonances in the filter cap circuits.

I did some lab work on the audiophile designer's common practice of bypassing large electrolytic capacitors with polypropylene film caps. It's common, but it turns out not to be simple. I measured the impedance of several popular electrolytics, and from this determined the equivalent circuits (*Fig. 7* and *Table 1*). Making the close-to-valid assumption that the polypropylene bypass caps I use are almost pure capacitance up to, say, 1MHz, I ran PC-ECAP simulations of various electrolytic-plus-bypass combinations.

The results surprised me: The inductance of the electrolytic, when combined with the capacitance of the polypropylene cap, produced sharp resonances. To damp these down, I needed a Zobel network to flatten out the electrolytic's high-frequency impedance; a polypropylene bypass then rolled the top off nicely.

This seemed like a hassle. Luckily, one capacitor—the 100µF 450V Panasonic I chose for the output cap in this supply—had low enough inductance and adequate damping that a Zobel wasn't needed. With this cap, and only this cap, a simple 470nF polypropylene was fine. When I needed more capaci-

TABLE 1
CAPACITORS, INDUCTANCES, AND ZOBELS

CAPACITOR (µF/DCWV)	FREQ. OF LOWEST Z (kHz)	EQUIVALENT SERIES R (MILLIOHMS)	INDUCTANCE (nH)	ZOBEL CAPACITOR (µF)
470/450 Nichicon	19	140	150	7.7
470/450 Panasonic	33	83	49	7.2
100/450 Panasonic	330	338	2.32	.021

Zobel resistor is equal to equivalent series resistance; Zobel capacitor is calculated as :

$$C = \frac{\text{inductance}}{(\text{equivalent series resistance})^2}$$

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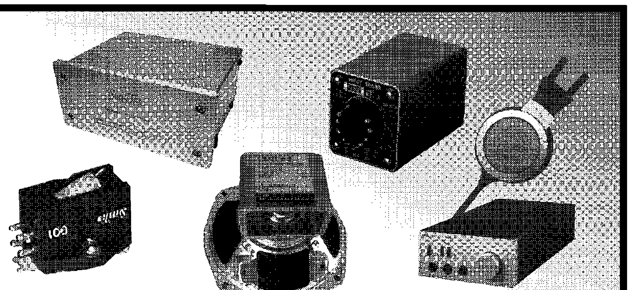

■Cartridge

Model	Price (US $)	Postage(Air Economy)
Denon DL-102 (MONO)	150	Area I $18: China, Korea, Hong Kong, Taiwan
Denon DL-103 (STEREO)	200	Area II $22: Singapore, Malaysia, Indonesia
Denon DL-103R (STEREO)	250	Area III $27: North America, Oceania, Europe
Shelter Model 501 II (CROWN JEWEL REFERENCE)	650	Area IV $34: Africa, South America
Shelter Model 901 (CROWN JEWEL SE)	1,300	

These Area I ~ IV are for all products except book.

■Japanease Audio Book

Postage $15

Title		Price(US $)
Attractive Tube Amps Vol. 1&2	(Isamu Asano)	30 each
Top-Sounding Vintage Power Tubes	(Stereo Sound)	30
The Joy of Vintage Tube Amps	(Tadaatsu Atarashi)	30
MJ Selected 300B Amps	(MJ)	30
Audio Tubes	(Hisashi Ohtsuka)	35
SE Amps by Transmitting Tubes	(Kouichi Shishido)	50
Classic Valve	(Hisashi Ohtsuka)	40
20TH CENTURY OF AUDIO	(Stereo Sound)	30
The Remembrance of Sound Post	(Susumu Sakuma)	30
Tube Amp Craft Guide	(MJ)	30
FOSTEX CRAFT HAND BOOK	(FOSTEX)	20

■MC STEP UP TRANS

Model	Specifications			Price (US$)	Postage**
	Pri.Imp(Ω)	Sec.Imp(kΩ)	Response		
Shelter Model 411	3~15	47	20Hz~50kHz	980	Area I $25
Jensen JE-34K-DX	3	47	20Hz~20kHz	550	Area II $30
Peerless 4722	38	50	20Hz~20kHz	300	Area III $40, Area IV $50

**Air Economy

■Speaker

Model	Specifications					Price*	Postage** (US $)			
	D(cm)	Ω	Response	db	w	(US $)	I	II	III	IV
Diatone P-610MB	16	8	45Hz~20kHz	90	7	360	30	40	50	66
Fostex FE208Σ	20	8	45Hz~20kHz	96.5	100	296	62	74	120	156
Fostex FE168Σ	16	8	60Hz~20kHz	94	80	236	42	50	73	98
Onken OS5000T	—	8	7kHz~25kHz	105	2.5	4,000	70	84	133	181
ALE 1710 Tweeter	8	16	6kHz~	118	10	3,380	85	110	170	230

*Price is for a pair **Air Economy

■STAX

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

■TANGO TRANS(28 models are available now)

Model	Specifications				Price (US $)	Postage** (US $)				
	W	Pri.Imp(kΩ)	Freq Response	Application		I	II	III	IV	
XE-20S (SE OPT)	20	2.5, 3.5, 5	20Hz~90kHz	300B,50,2A3	396	47	56	84	113	Price is for a Pair
U-808 (SE OPT)	25	2, 2.5, 3.5, 5	20Hz~65kHz	6L6,50,2A3	242	42	50	73	98	
XE-60-5 (PP OPT)	60	5	4Hz~80kHz	300B,KT-88,EL34	620	62	74	115	156	
FX-40-5 (PP OPT)	40	5	4Hz~80kHz	2A3,EL34,6L6	320	47	56	84	113	
FC-30-3.5S (SE OPT) [XE-60-3.5S]	30	3.5	20Hz~100kHz	300B,50,PX-25	620	62	74	115	156	
FC-30-10S (SE OPT) [XE-60-10SNF]	30	1.0	30Hz~50kHz	211,845	620	62	74	115	156	
NC-14 (Interstage)	—	[1+1 : 1+1] 5	25Hz~40kHz	[30mA] 6V6 (T)	264	30	40	50	70	
NC-16 (Interstage)	—	[1+1 : 2+2] 7	25Hz~20kHz	[15mA] 6SN7	264	30	40	50	70	

** Air Economy

■TAMURA TRANS(All models are available)

F-7002 (Permalloy)	10	3.5	15Hz~50kHz	300B,50	740	60	70	110	145	Price is for a Pair
F-7003 (Permalloy)	10	5	15Hz~50kHz	300B,50	760	60	70	110	145	
F-2013	40	10	20Hz~50kHz	211,242	730	70	84	133	181	
F-5002 (Amorphous)	8	3	10Hz~100kHz	300B,2A3	1276	65	80	120	160	

** Air Economy

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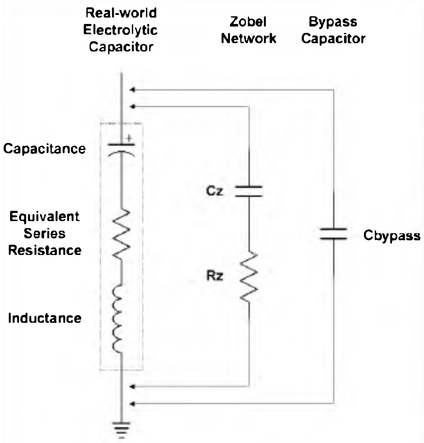


FIGURE 7: Real-world electrolytic capacitor, showing parasitic components, and the Zobel network and bypass capacitor used to negate their effects.

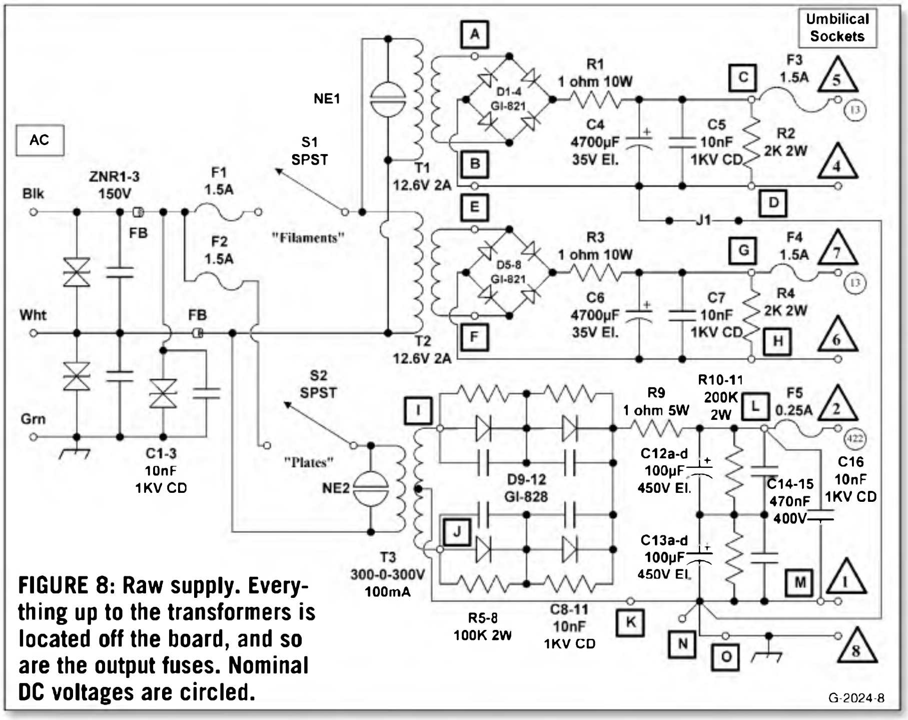


FIGURE 8: Raw supply. Everything up to the transformers is located off the board, and so are the output fuses. Nominal DC voltages are circled.

tance, I connected more 100µF caps in parallel.

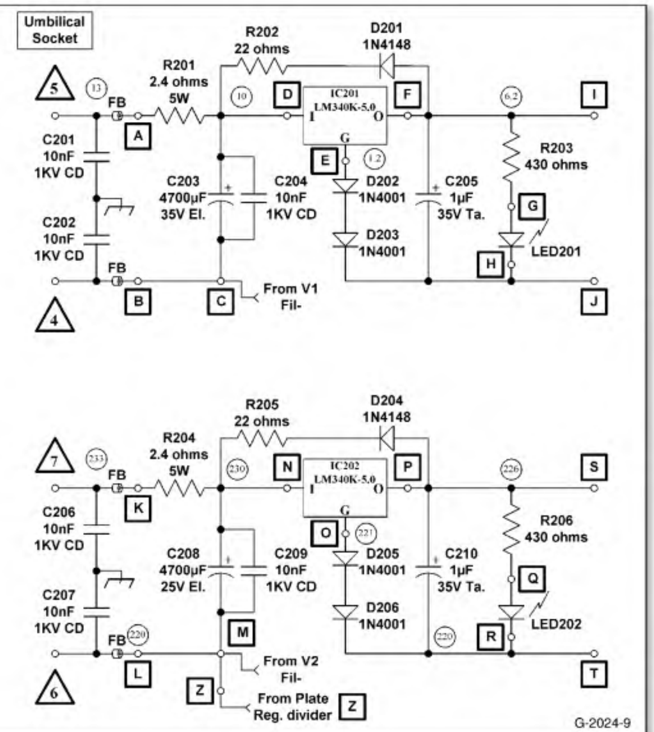
SEPARATE CABLES

In the raw supply design (*Fig. 8*), the output voltages may seem unnecessarily high, given the final working voltages. This is because I wanted the regulators to keep regulating even when the line voltage drops 10%, which is common during St. Louis summers. There's RFI and spike filtering at the AC input, but this preamp will still benefit from being plugged into a good line filter and surge protector—I use the Monster Cable HT800.

I used two filament supplies—one grounded, for the input tubes, the other floating for the cathode followers. They share a power switch, while the plate transformer has a separate switch. This lets you turn on the filaments, allow a few minutes for them to warm up, then switch on the plates, preventing cathode stripping. (To be truthful, I've never observed anything I could blame on cathode stripping in a small-signal amplifying tube. Power tubes, yes, but not preamp or driver tubes. Still, it's worth playing it safe, given the price of tubes these days.) You can, if you like, design a fancy time-delayed switch-on circuit, but the manual method works just as well, and is much cheaper.

The raw supply box connects to each preamp via an umbilicus. In the solid-state preamp, I used a shielded multi-conductor cable to carry the supply

FIGURE 9: Filament regulators. C201–202, C206–207, IC201–202, and LED201–202 are mounted off-board (the ICs on HS201–202). Unless otherwise stated, all resistors are 1W metal oxide.



voltages. For the tube preamp, however, I ran into a snag.

The plate supply can run as high as 460V, which is way higher than the rated DC voltages for most multi-conductor cables. After some searching, I found the ideal solution: Canare Star-Quad microphone cable, with excellent shielding and a 500V rating. I ran one length of Canare and an equal length of shielded multi-conductor cable (carrying raw voltages for the filament supplies) to each preamp.

Plugs were a problem. I hesitated to use XLR-type connectors with such high voltages. Ideally I would use Jones plugs, which are easy to solder, and extremely rugged. They mount in square or rectangular holes, though, and my tendinitis told me not to get involved with nibbling tools, files, and so forth.

So I settled on octal plugs and sockets—easily available and reasonably rugged, though the plugs are a bear to solder. There are alternatives. You might, for example, decide to use inline Jones plugs and sockets.

IF IT MOVES, REGULATE IT

Figure 9 shows the filament regulator circuits. Each uses an additional stage of RC filtering to feed the ICs, which are standard 5V 3-terminal devices. Each IC regulator has two diodes in series with the ground lead, raising the output to 6.2V.

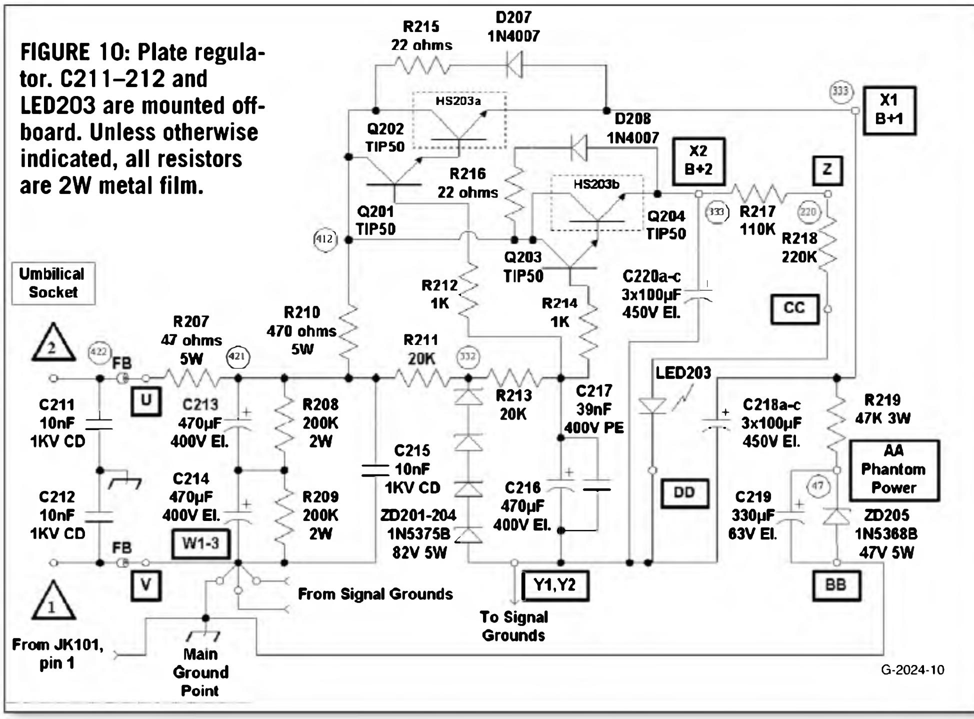
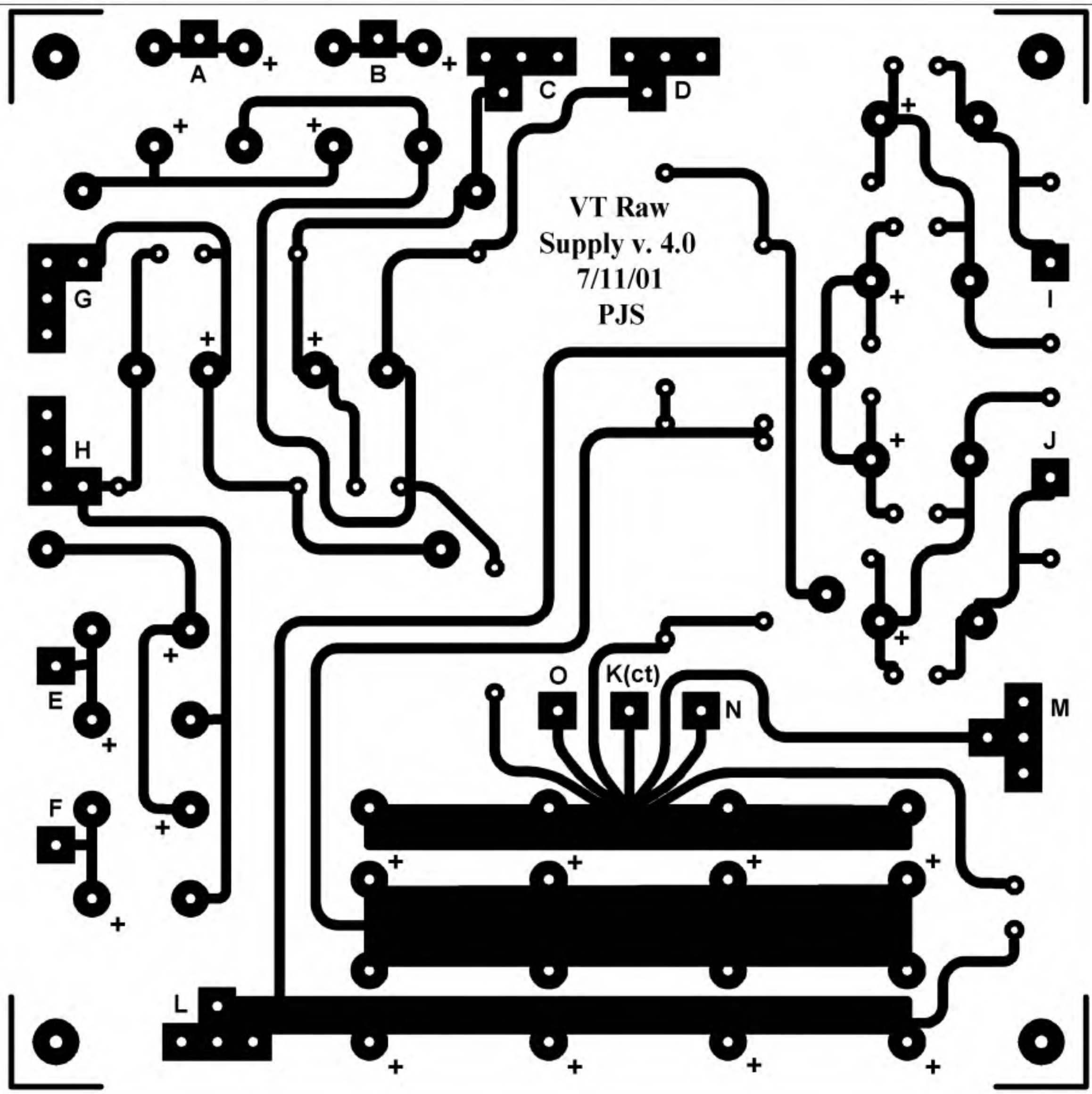


FIGURE 10: Plate regulator. C211–212 and LED203 are mounted off-board. Unless otherwise indicated, all resistors are 2W metal film.

As I mentioned before, one of the circuits is grounded (back in the raw supply box); the other is floated at +220V by a voltage divider in the plate supply. The resistor and diode running from the outputs back to the inputs serve to drain the output capacitors during turnoff. In keeping with Ben Duncan's suggestion, the diodes are low-capacitance glass signal devices to keep stray RFI from detouring around the regulator into the filaments. The resistors limit current to avoid frying the diodes.

I mounted the filament regulators and their heatsinks on the back of the chassis to keep their heat outside the box. Because one regulator's case floats at an unpleasantly high voltage, I had to use a plastic

FIGURE 11: Raw supply board—foil side.



cover for the device. Those aren't easy to find, but Mouser had them.

A LOT ON MY PLATE

Figure 10 shows the plate regulators–one for the input stage, one for the output. Again, I began with an RC filter stage, reducing hum and (one hopes) RFI.

I had the devil's own time with regulators. I tried using a Last PAS regulator, but for reasons I don't understand, I could not get the device to work stably. I also tried an LM-11C circuit (from National Semiconductor's application notes), but it wouldn't regulate.

I finally settled on the old reliable zener-fed series regulator, using a pair of TIP-50 transistors in Darlington configuration for lower output impedance. (Oddly enough, when I tried using a triple-Darlington, the impedance was

FIGURE 12: Raw supply board—component side. There is one jumper, J1.

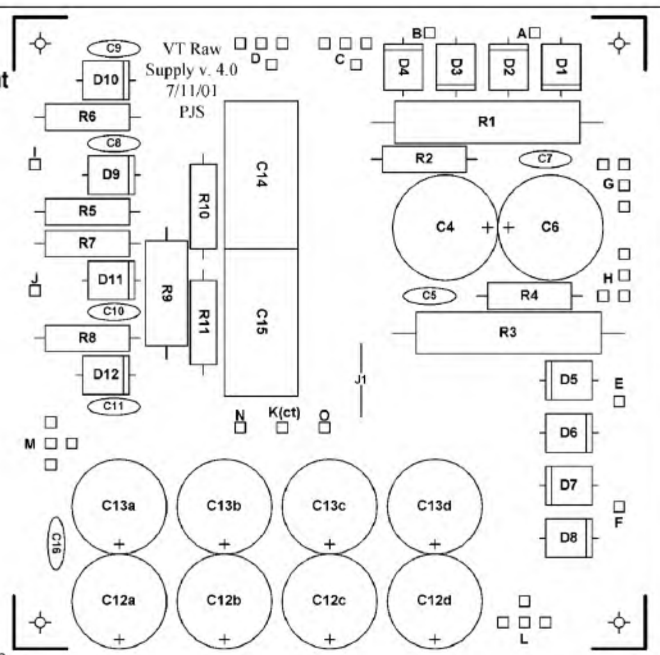


FIGURE 13: Regulator board—foil side.

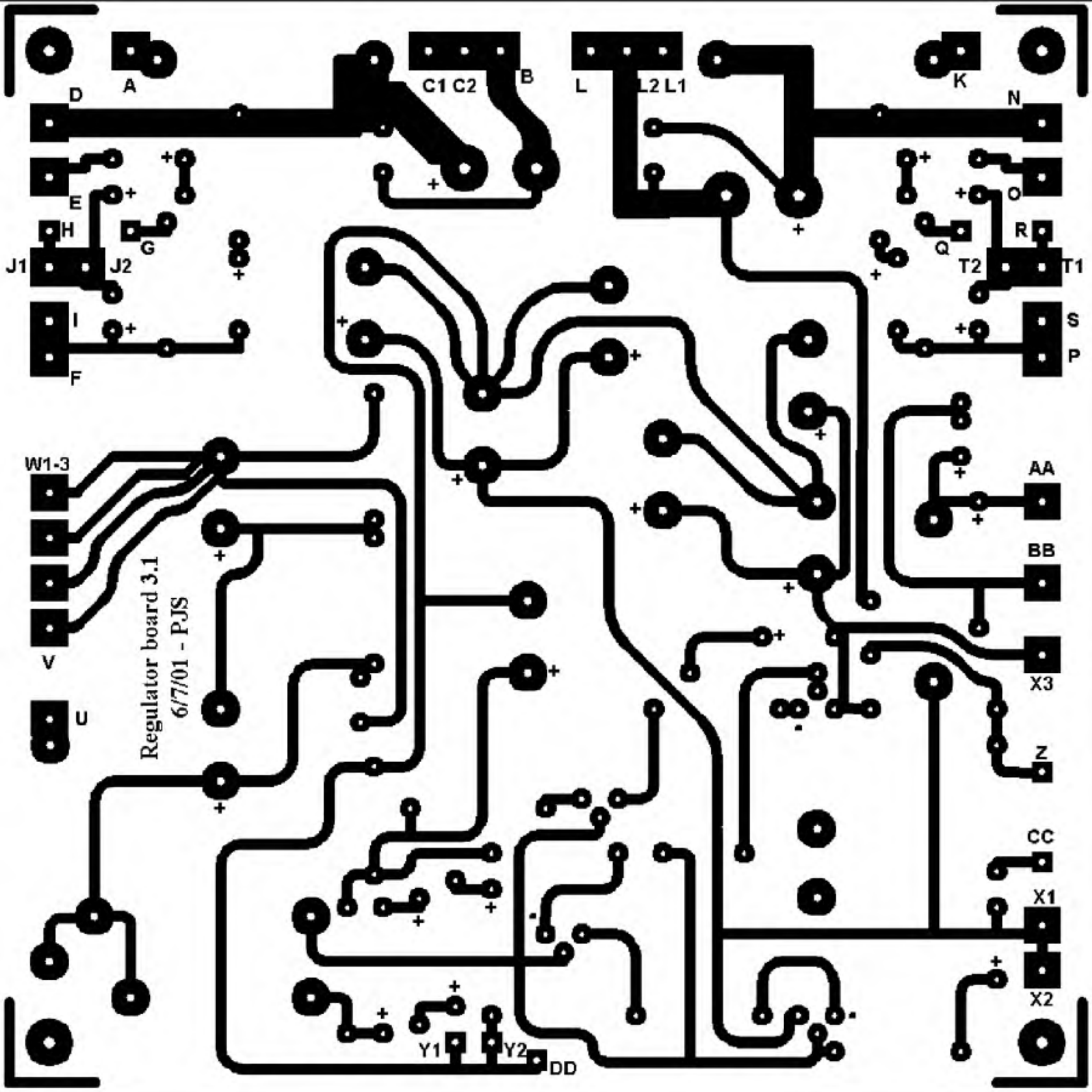


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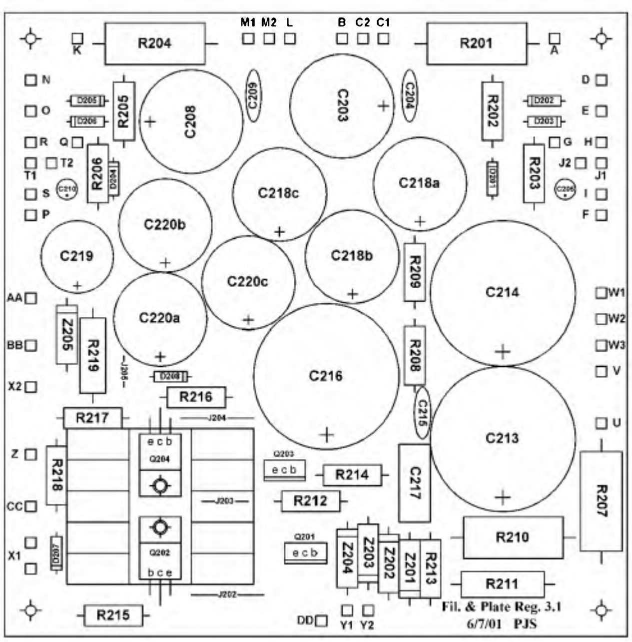
FIGURE 14: Regulator board layout—component side. There are four jumpers (J202–205) on the board.



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higher than with two—can anyone out there explain that?) The zener string establishes the reference voltage for two pairs of pass transistors (I calculated that crosstalk between them would be negligible).

I added a few wrinkles to the standard series regulator design, including decoupling the zener diodes' bypass capacitor to improve line rejection and reduce zener noise. If I was really ambitious, I might try feeding the zener string with a constant-current source, but it works fine and sounds excellent as is. The zeners are 5W units, for reliability.

Each output capacitor is a composite of three 100µF/450V Panasonic electrolytics and a 470nF polypropylene. Above about 600Hz, most of the regulator's low output impedance comes from

(to page 68)

Load-Invariant Power Amplifiers, Part 1

This noted author continues in his efforts to help us understand distortion. **By Douglas Self**

Audio power amplifiers always—without exception—give more distortion with heavier loading. The lower the load impedance, the worse the nonlinearity. Loudspeaker impedance curves are one of the least standard occurrences in hi-fi; dips to 4Ω or less are common, and so must induce increased amplifier distortion.

I decided to see whether this increase in distortion could be eliminated—or at any rate, reduced in its severity—in bipolar amplifiers, so I needed to understand exactly why it happens. Crossover distortion is known to increase with loading, but in bipolar transistor amplifiers there is an extra distortion mechanism that becomes more important and eventually overwhelms the crossover products as load impedance falls and the output stage currents increase. My initial target was a load-invariant amplifier—i.e., one with the same THD at 4Ω loading as at 8Ω. I have not yet built an amplifier that is totally load-invariant, but you'll see that I came pretty close.

Another way to tackle this issue is to urge loudspeaker designers to strive for flatter impedance curves than we currently get, but you must accept that electronic problems are much easier to solve than electromechanical ones, so it is reasonable for amplifiers to accommodate themselves to loudspeakers rather than the other way around. Thus, an amplifier must be able to cope gracefully with impedance dips to 4Ω or lower. Such dips tend to be localized in frequency, so normal music is not going to dwell in them for long, but they still need to be handled properly.

ABOUT THE AUTHOR

Douglas Self obtained a degree in engineering at Corpus Christi College, Cambridge, and went on to study experimental psychology and psychoacoustics at Sussex University. He has worked in both the professional audio and hifi fields for quite some time now. He lives in Forest Gate, in East London, and maintains a personal website at dself.demon.co.uk.

THE BLAMELESS AMPLIFIER CONCEPT

A Blameless Class-B power amplifier is defined (by me) as one in which all the distortion mechanisms shown in *Table 1* have been eliminated or reduced to below the noise floor, except for the intractable Distortion 3 in its three subcategories. I explained these distortion mechanisms in my series of three articles in *Audio Electronics.*[1] I have also produced a book or two which cover this subject in much greater detail than magazine articles permit[2,3] and contain all the references from here if you can't get hold of the original journals.

I use the word "Blameless" to emphasize that the high performance of such amplifiers is due to avoiding a series of common errors rather than stemming from brilliantly innovative circuit design. A Blameless amplifier design gives a THD performance into 8Ω that depends very little on variable transistor characteristics such as beta, because at this load impedance the output stage nonlinearity is almost all crossover distortion, which is essentially a voltage-domain effect, and not affected by beta. For minimal crossover distortion the quantity to be optimally set is Vq, the voltage across *both* the two output emitter resistors Re; the actual value of the resulting quiescent current Iq is incidental.[4] Mercifully, in Class-B the same Vq remains optimal whatever the load impedance; if it varied the extra complications would be serious.

The value of a Blameless amplifier is not just excellent linearity; more to the point, its clean operation allows intruding distortion mechanisms to be seen against a clear background.

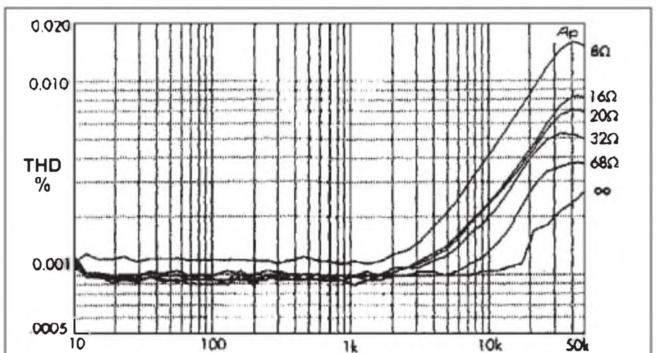

FIGURE 1: The crossover distortion from a Blameless amplifier increases as load resistance falls to 8Ω. This is not due to LSN. (All THD plots are taken at 80kHz bandwidth, RMS sensing.)

TABLE 1

NO.	MECHANISM	CATEGORY	SENSITIVE
1	Input Vin/Iout nonlinearity	Inherent	No
2	VAS lin\Vout nonlinearity	Inherent	Yes?
3	Output stage distortions:		
a	Large-signal nonlinearity	Inherent	Yes
b	Crossover distortion	Inherent	No
c	Switchoff distortion	Inherent	Yes
4	Nonlinear VAS loading	Inherent	Yes
5	Rail decouple grounding	Topological	No
6	Rail current induction	Topological	No
7	Error in NFB takeoff-point	Topological	No
8	Feedback cap distortion	Inherent	Yes

OUTPUT LOADING AND DISTORTION

As the load impedance of a Blameless Class-B amplifier is decreased from infinity to 4Ω, distortion increases in an intriguing manner. Unloaded, the THD is not much greater than that from the AP test oscillator, but with loading crossover distortion increases steadily (*Fig. 1*). This increase is due to the fundamental nature of Class-B, and is not the effect we are stalking here.

When the load impedance falls below about 8Ω, a new distortion begins to appear, dominating the crossover nonlinearities. It is low-order, essentially third-harmonic, and very easily distinguished from crossover in a THD residual. In *Fig. 2* the upper (4Ω) THD trace is consistently twice that for 8Ω, once it clears the noise floor, and most of the difference is due to this new distortion.

In my previous writings, I have labeled this effect as Distortion 3a, or Large Signal Nonlinearity. "Large" refers to currents rather than voltages. Hereafter I just call it LSN. Unlike crossover (Distortion 3b), the amount of LSN produced is very dependent on device characteristics.[5] The distortion residual is essentially third-order due to the symmetric and compressive nature of the output stage gain characteristic, but its appearance on a scope may be complicated by different amounts of nonlinearity in the upper and lower output stage halves.

LSN occurs in both emitter-follower (EF) and complementary feedback pair (CFP) output configurations; this part of the article concentrates on the CFP, as in *Fig. 3*. *Figure 4* shows the incremental gain of a SPICE-simulated CFP output stage for 8 and 4Ω; the lower 4Ω trace has a greater downward curvature; i.e., a greater falloff of gain with increasing current. Simulated EF behavior shows a similar increased droop with 4Ω loads.

The typical result of 4Ω amplifier loading was shown in *Fig. 2* for the relatively modern MJ15024/25 complementary pair from Motorola. *Figure 5* shows the same test on one of the oldest silicon complementary pairs, the 2N3055/2955. The 8Ω distortion is similar for the different devices, but the 4Ω THD for the venerable 2N3055/2955 is three times higher rather than twice. Clearly, device characteristics do affect LSN.

Such experiments with ancient transistors give useful perspective to the Blameless amplifier concept—from the types tried so far you can say that even with old silicon, Blameless performance should not exceed 0.001% THD at 1kHz and 0.006% at 10kHz, into 8Ω. It is a sobering thought that all the components existed to build sub-0.001% THD amplifiers in the late '60s, had we but known how to do it.

Low-impedance loads obviously have other implications beyond worsening the THD. Long-term 4Ω operation demands significantly more heatsinking and power-supply capacity if reliability is to be maintained. For economic reasons the high peak/average ratio of music is usually fully exploited in commercial equipment, though this can cause real problems on extended tests, such as the old FTC 40%-power-for-an-hour preconditioning procedure.

The main subject of this article is the extra distortion generated in the output

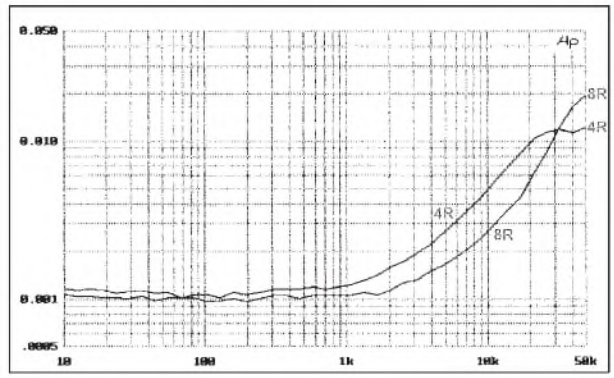

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FIGURE 2: The distortion roughly doubles due to LSN as loading increases from 8 to 4Ω, using modern output devices such as MJ15024/25. (NB: The 4Ω THD is less than the 8Ω above 30kHz solely due to fortuitous cancellation of harmonics.)

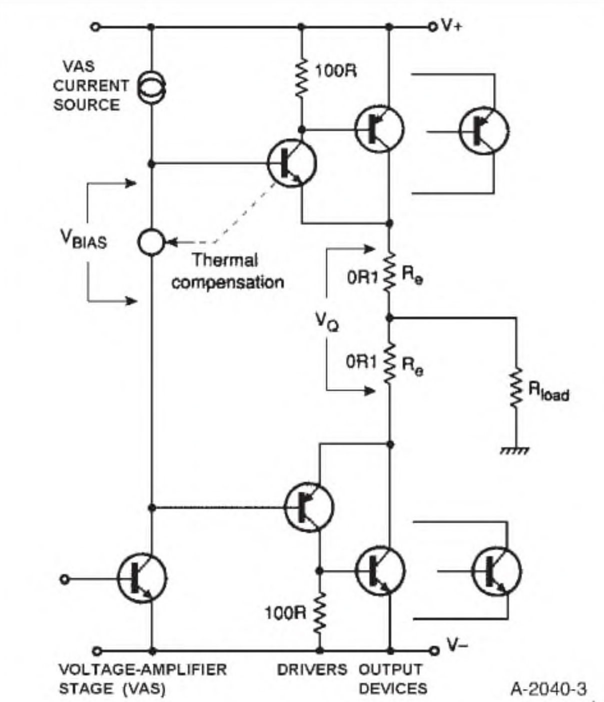

FIGURE 3: The standard CFP output stage, also showing how extra output devices will be added in parallel.

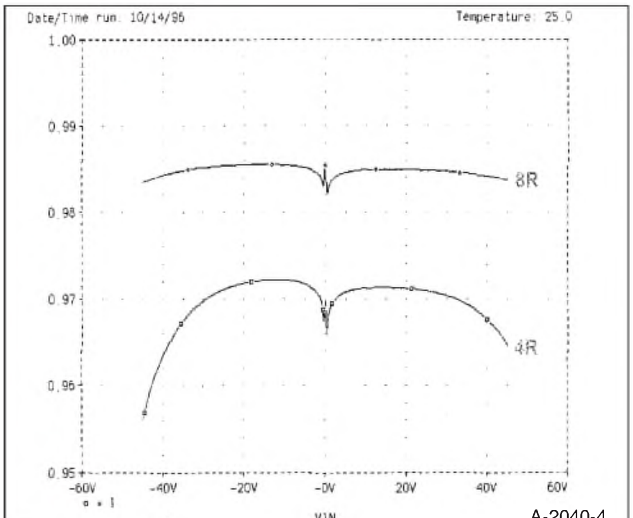

FIGURE 4: The incremental gain of a standard CFP output stage. The 4Ω trace not only shows lower general gain but droops much more as the gain falls off at higher collector currents.

stage itself by increased loading, but there are other ways in which the increased currents flowing may degrade the total amplifier distortion. In *Table 1*, Distortions 1, 2, and 8 are unaffected by output stage conditions. Distortion 4 might be expected to increase, as the increased loading on the output stage is reflected in increased current taken from the voltage-amplifier-stage (VAS).[6] However, both the beta-enhanced EF and buffered-cascode VAS configurations cope effectively down to 3Ω loads at moderate powers, without introducing extra distortion. At high powers (say greater than 80W/8Ω), it may be necessary to use output triples to reduce the VAS loading; more on this in Part 2 of this article.

The greater supply currents drawn could increase supply rail ripple, which will worsen Distortion 5 if it has not been thoroughly suppressed by RC filtering of bias supplies, and so forth.[7] On the other hand, if the supply reservoir capacitance has been increased to permit greater power delivery, ripple will be reduced again, so this tends to cancel out.

Distortion 6 may be more difficult to eliminate, as the half-wave currents flowing in the output circuitry are twice as large, with no counteracting mechanism. Distortion 7, if present, will be similarly worsened due to the increased currents flowing in the output stage wiring resistances.

Of these mechanisms, Distortion 4 is inherent in the circuit configuration (though not a problem in practice), while 5, 6, and 7 are topological, in that they depend on the spatial and geometrical arrangements of components and wiring. The latter three can therefore be completely eliminated in both theory and practice. This leaves only the LSN component of Distortion 3 to grapple with, so things look encouraging.

THE LOAD-INVARIANT CONCEPT

Ideally, the extra distortion component LSN would not exist, and an amplifier would give no more distortion into 4Ω than 8. I would call that "load-invariant to 4Ω." The loading qualification is required because the lower the impedance, the greater the difficulties in aspiring to load-invariance. I am assuming that you start out with an amplifier that is Blameless at 8Ω; it would be logically correct but fairly pointless to apply the term "load-invariant" to an ill-conceived amplifier delivering 1% THD into both 8 and 4Ω.

THE LSN MECHANISM

LSN is clearly a current-domain effect, dependent on the magnitude of the signal currents flowing in drivers and output devices, as the voltage conditions are unchanged. A major consequence of increased collector current is a falloff in current gain (beta). The beta falloff is ultimately due to what are called high-level injection effects in the transistor. However, this is not the place to get into semiconductor physics; suffice it to say that these effects vary with device type, so device characteristics now really do affect performance.

Moving from an 8Ω to a 4Ω load doubles the output device currents, but this does not in itself generate significant extra distortion. The crucial factor appears to be that the current drawn from the drivers by the output devices *more* than doubles, due to beta falloff in the output devices with increasing collector current. It is this *extra* increase of current due to beta-droop that causes almost all the additional distortion. The exact details of how this works are not completely clear, but seems to be because the "extra current" due to beta falloff is particularly nonlinear with respect to output voltage, and this nonlinear extra current combines with driver nonlinearity in a particularly pernicious way.

Surprisingly, the output devices do not contribute increased distortion with increased loading—it all comes from the drivers.

There is good SPICE-simulator evidence that this explanation is correct, and LSN is entirely due to the beta-droop causing extra current to be drawn from the drivers, degrading their linearity. Here is the evidence:

- Simulated output stages built from output transistors modified to have no beta-droop at all (simply by increasing the SPICE model parameter IKF) show no LSN. This shows that it is specifically the extra current drawn due to beta-droop that causes the trouble.
- Simulated output devices driven with zero-impedance voltage sources instead of transistor drivers show no LSN. This shows that LSN does *not* occur in the output transistors themselves, but in the drivers.
- If you look at output stage distortion as an error voltage between input and output, the double emitter-follower (EF) stage error is: error = (driver Vbe + output Vbe + Re drop).

A simulated EF output stage with the usual drivers shows that of these three terms, it is primarily nonlinearity in the driver Vbe that increases as the load resistance reduces, while the output Vbe nonlinearity is hardly changed. The voltage drop across Re is essentially linear, as it is in series with the load.

Understanding the mechanism of LSN leads to some solutions. To put it succinctly: "Increased driver nonlinearity is caused by beta-droop, which is caused by increased output device collector current." There are three points of attack here.

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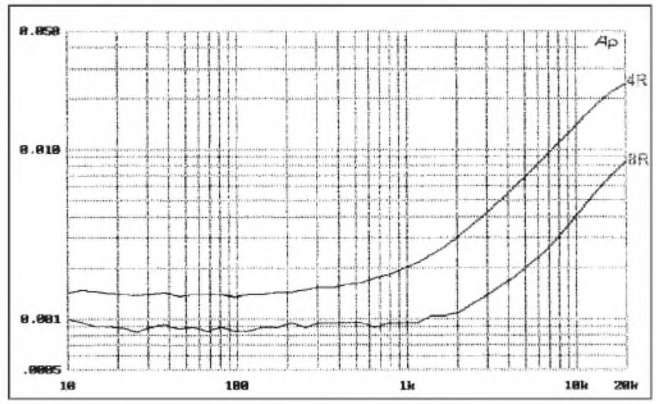
First, increase the linearity of the driver system; this can't be done if it's a single-transistor, as is usually the case.

Secondly, the output device types can be selected for the least beta droop. This is very device-type dependent, so it's worth having a good look around for promising types.

Third, the per-device collector current can be reduced by using parallel output devices. This in turn will reduce the beta-droop.

Part 2 will provide practical measurements to confirm the simulation-based reasoning presented here, and includes a practical circuit design for a load-invariant power amplifier that will give beautiful linearity even in those impedance dips. ❖



A-2040-5

FIGURE 5: 4Ω distortion is three times greater than 8Ω for the venerable 2N3055/2955 output devices. Compare *Fig. 2*.

FM Stereo Signal Generator

Here's a circuit to generate a frequency-modulated stereo RF signal to align and test FM receivers. **By Charles Hansen**

This circuit is built around the $35 Ramsey Electronics FM10A FM Stereo Transmitter kit, modified here for better performance, and with several additional features. Total cost of parts is about $78.

FM STEREO BROADCASTING OVERVIEW

The FM stereo modulation method was devised after FM monaural had already been established, and it was required to be backward-compatible with existing FM mono receivers. As with black-and-white and color TV, the backwards compatibility came at the cost of additional complexity, and lowered the stereo signal quality.

To broadcast a stereo signal that is compatible with an FM mono receiver, you need to sum the left and right channels to mono, then use this L+R baseband 20–15kHz audio signal to frequency-modulate the FM transmitter. Then you add a 38kHz subcarrier, which is modulated by the difference between the left and right audio channels (the L-R signal). Finally, add a 19kHz pilot tone to activate the receiver's stereo decoder circuitry.

The 38kHz signal is a suppressed carrier wave that is amplitude modulated (AM) by the L-R signal. The receiver demodulates the 38kHz L-R signal and the baseband L+R signal and sends them into a stereo matrix decoder that restores them to the original L and R stereo audio signals. This decoding process generates a fair amount of noise.

In addition, at 100% stereo modulation, the 19kHz tone comprises only 10% of the modulation signal, with the left and right audio comprising 40% each. This requires greater RF sensitivity on the part of the FM receiver to produce a quiet stereo signal. Many FM receivers use a blend circuit to gradually return the audio to the quieter mono mode when the signal strength is reduced.

PHOTO 1: Signal generator front view.

A 75μs (13.33kHz) pre-emphasis filter is used to boost the higher audio frequencies before they enter the transmitter modulator. A 75μs de-emphasis circuit in the receiver restores the audio response to that of the source material. This improves the signal-to-noise ratio, since most of the matrix decoder noise is concentrated in the higher frequencies. (A 50μs pre-emphasis curve is used in Europe, and the Ramsey kit comes with components for either curve.)

There are two other sub-carriers used with the FM broadcast system. One is the subsidiary communications authorization (SCA), the paid-subscription commercial-free "elevator music." This system uses a narrowband 67kHz FM subcarrier with an audio bandwidth of 7kHz, transmitted with just 10% of the amplitude of the stereo signal. Although it is not illegal to own an SCA decoder for noncommercial use, FM tuner circuits generally have filters to reject the 67kHz SCA subcarrier.

The second system is RDS, the radio data system that allows text or other information to piggyback on the standard

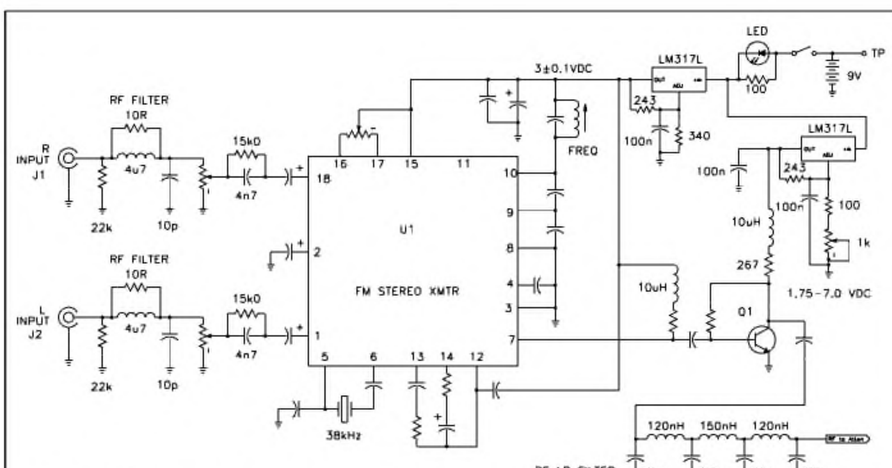


FIGURE 1: Stereo generator schematic.

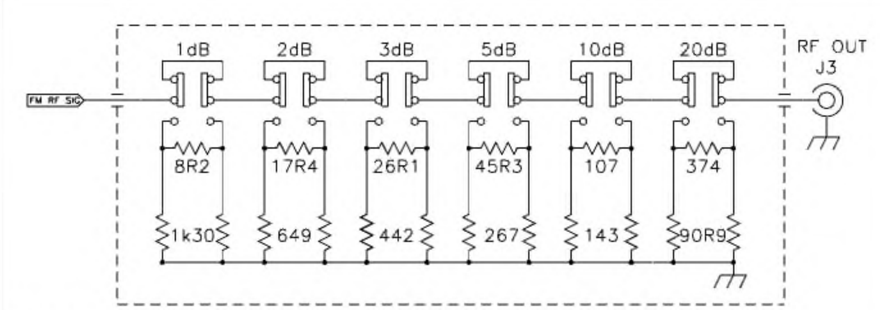


FIGURE 2: RF attenuator schematic.

FM radio signal. It uses a 57kHz subcarrier (triple the 19kHz stereo pilot signal) that is divided by 48 in the receiver to recover 1.1875kHz NRZ (non-return to zero) digital data. RDS allows you to view station call letters on a digital display in RDS-compatible receivers. [Some RDS stations also include the correct time.—Ed.]

HOW IT WORKS

The schematic diagram for the FM stereo generator is shown in *Fig. 1*, and the modification parts list is in *Table 1*. Values for the standard FM10A kit have been omitted for clarity, with the modified part values shown. These modifications are based on years of experience (by others than myself). The modification design notes are available on the web[1]. Of course, this modification will void the FM10A warranty.

Stereo audio signals are input to J1 and J2, and pass through RF filters to block the RF signal from getting back to the external audio source. Trimpots are used to set the audio level into U1, the FM stereo transmitter chip[2]. US standard 75µs pre-emphasis networks, 15k in parallel with 4n7, precede the aluminum decoupling caps, which input the audio into U1 pins 1 and 18. The pot at pins 16 and 17 is used for the audio balance adjustment.

The early version of the FM10 was much maligned for both its frequency drift, which prevented modern digitally tuned FM receivers from locking onto its transmitter frequency, and for the poor stability of the 19kHz stereo pilot signal R-C oscillator network. The FM10A kit addresses many of these problems and uses NP0 ceramic capacitors in the tank circuit and a 38kHz crystal oscillator. I have not experienced any drift off the 98.1MHz FM frequency with my signal generator.

The 19kHz stereo pilot signal and 38kHz suppressed subcarrier are generated by the 38kHz crystal at pins 5 and 6. Audio input above 15kHz is steeply rolled-off in the chip so that no 19kHz component exists that would confuse the receiver's pilot signal detector. The L-C tank circuit connected to U1 pin 10 adjusts the FM "broadcast" frequency.

Three values of NP0 ceramic capacitor are supplied with the FM10A. They allow you to select one of three 10MHz tuning ranges within the 88–108MHz FM band, using the ferrite slug of the inductor for tuning. It helps to mount the NP0 cap on the back of the PC board directly to the inductor PC board pins. This greatly reduces the parasitic inductance of the capacitor leads and improves stability.

The modulated RF output at U1 pin 7 is sent to the base of Q1, an RF NPN transistor, for amplification. This boosts the signal for its intended purpose as a low-power FM transmitter. It also buffers the chip output to the 75Ω impedance of FM radio. I added a seven-element RF low-pass (LP) filter[1] after the collector output of Q1 to remove the spurious high-order harmonics, which can cause radio interference from the RF output. I also deleted one of the 1nF RF output coupling caps, which was designed to couple the FM10A to an onboard whip antenna. The antenna and a plastic case are another available kit, which you do not need.

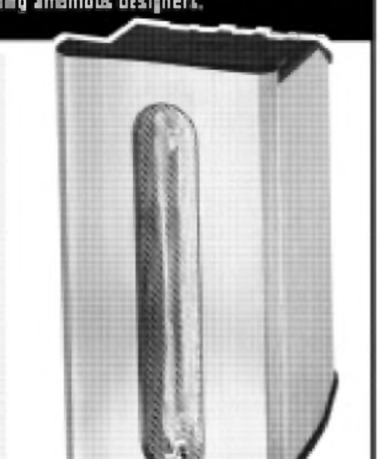


PHOTO 2: Signal generator rear view.

KIT MODIFICATIONS

The FM10A kit still uses three silicon diodes in series to provide 1.8V DC Vcc to U1 pin 15, another sore-point with its users. This power-supply voltage depends on the forward voltage drop of the silicon diodes and varies considerably with temperature. I replaced the three diodes with an LM317L adjustable regulator to bring the chip supply up to 2V DC (2.5V DC is the absolute maximum rating).

The higher voltage also provides more headroom for the audio signal level, which I run at 40–60mV input to J1 and J2 with the input trimpots turned down about two thirds. Higher audio input signals tend to cause ringing and overmodulation above 10kHz, which causes distortion in the receiver. The 10mH choke in series with the bias resistor at the base of Q1 helps keep the RF out of the power supply.

I added another LM317L, with a 1k adjust pot, to vary the collector voltage to Q1 from 1.75 to 7V DC. This allows me to adjust the RF output to the desired level and minimize audio distortion. Another 10µH choke blocks RF from getting to the variable DC supply.

You must mount the modification components to the board with minimum lead lengths. You need to do all connections to ground at the PC board ground plane. Drill small (#58 drill) holes, remove the green epoxy solder mask coating, and solder the leads to the ground plane.

I also drilled holes for some of the other component leads to make connections shorter. I drilled in areas without copper where possible. In some cases I needed to drill through the ground plane, so I isolated the copper from ground using a Vector P138A pad cutter, available from Mouser as P/N 574-P138A.

The total current drawn from the 9V DC battery is about 30mA, so I put the LED power-on indicator in series with the battery so I would not waste current. The 100Ω resistor across the LED carries about two-thirds of the total supply current, so the LED is not overloaded. A recessed pin jack on the rear panel lets me check the battery without opening the case.

The RF level at the output of Q1 is still well above that needed for a "closed-loop" FM tuner test setup. With 3V DC to the collector of Q1, my FM10A board puts out about 30mV of RF. This is 10µW into 75Ω, or 100dBf (dB above one femtowatt, one fW being 10^{-15}W). This will easily overload most FM tuner front ends. FM mono sensitivity (for 65dB quieting) is in the 10–15dBf range.

In order to reduce the RF to a useful level, I built an RF attenuator into the chassis and shielded it from the FM10A board with sheet copper. I used eight switches in an attempt to provide a total of 80dBf attenuation with 1dBf resolution. This proved to be wildly optimistic on my part. At best, I can reduce the RF level by about 20dBf, to 80dBf or 100nW into 75Ω.

TABLE 1
MODIFICATION PARTS LIST

VALUE	DESCRIPTION	VENDOR	PART NO.	QTY
10pF 200V	Ceramic X7R	Mouser	21RX510	2
4n7 100V 2%	Polyester film	Mouser	140-PF2A472F	2
33pF 300V 5%	Silver mica	Mouser	5982-5-300V33	2
56pF 500V 5%	Silver mica	Mouser	5982-10-500V56	2
	BNC, 75Ω bulkhead or type "F" bulkhead	Mouser	523-31-221-75RFX or, 161-0062	1
	Phono jacks, gold (pair)	RS	274-852	1
	Tip jack, red	Mouser	530-105-0802-1	1
10µH	Inductor, molded	Mouser	70-IMS5-10	2
120nH	Inductor, molded	Mouser	70-IM2-.12	2
150nH	Inductor, molded	Mouser	70-IM2-.15	1
4µ7	Inductor, molded	Mouser	70-IMS5-4.7	2
	LED clip	Mouser	593-CLP125	1
	Red LED, T-1 ¾	Mouser	351-5110	1
100 5%	Carbon film	Mouser	29SJ250-100	1
22k 5%	Carbon film	Mouser	29SJ250-22k	1
1k	Cermet 10T trimpot	DK	72-T93YA-value	1
107 1%	Metal film	DK	107XBK-ND	1
143 1%	Metal film	DK	143XBK-ND	2
15k0 1%	Metal film	DK	15.0KXBK-ND	2
17R4 1%	Metal film	DK	17.4XBK-ND	1
1k30 1%	Metal film	DK	1.30kXBK-ND	2
243 1%	Metal film	DK	XBK-ND	2
267 1%	Metal film	DK	267XBK-ND	2
26R1 1%	Metal film	DK	26.1XBK-ND	1
147 1%	Metal film	DK	147XBK-ND	2
374 1%	Metal film	DK	374XBK-ND	1
442 1%	Metal film	DK	442XBK-ND	2
45R3 1%	Metal film	DK	45.3XBK-ND	1
649 1%	Metal film	DK	649XBK-ND	2
90R9 1%	Metal film	DK	90.9XBK-ND	2
10R 5%	Metal oxide	DK	10W-1-ND	1
8R2 5%	Metal oxide	DK	8.2W-1-ND	1
DPDT 6A	Mini toggle	PE	060-338	6
LM317L	Reg, +Adj, TO5	DK	LM317LZ-ND	2
	9V battery	any		1
	Chassis	RS	270-274	1
	FM transmitter kit	Ramsey	FM10A	1

Drafting appliqué film (Letraset *Letracopy Creative* or Chartpak *DAF8*), available at art or drafting supply houses.

Metal standoffs, hookup wire, shielded audio cable, 75Ω coax cable, sheet copper or steel (preferred), solder, hardware, and so forth.

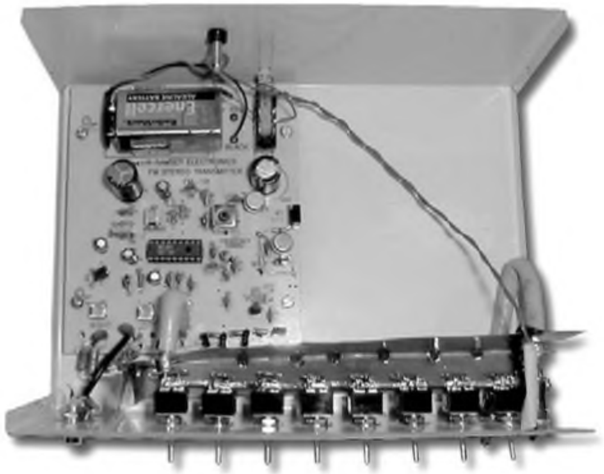


PHOTO 3: Signal generator interior view.

Figure 2 shows the schematic for a six-switch attenuator, for a still-optimistic total of 40dBf. Just change the resistor values in any of the π-filter sections to change the attenuation for that switch. Solder the resistors going to ground directly to the copper sheet. Connect the shield to the PC board ground using solder wick, which makes a great braided cable.

I also recommend making the 75Ω BNC connection directly to the last switch, rather than connecting it via coax as I did in my prototype. You can also use a video type “F” bulkhead connector instead of the 75Ω BNC I used (see parts list).

I use an external 0–70dB RF attenuator to get the signal level at the tuner down to 10dBf. Apparently, despite the copper shielding I used, there is enough leakage into the attenuator area so that it acts as an antenna. Copper is great against electric E-fields, but the close proximity magnetic H-field requires a ferromagnetic material for effective attenuation.

By contrast, my RF coax attenuator unit has a steel chassis with overlaps to prevent a slot antenna effect. The individual slide switches have mu-metal shields between them, effectively isolating each π section of the filter from the other sections. The center conductors of the 75Ω BNC connectors at each end are directly soldered to the adjacent switch poles. This is good basic RF engineering at work.

Since you will be starting from scratch, I think the best solution is to use a fixed 75Ω π-section 20dB attenuator in the signal generator chassis and build (or buy) an external RF attenuator in a separate metal box, with 75Ω RF input/output connectors[3].

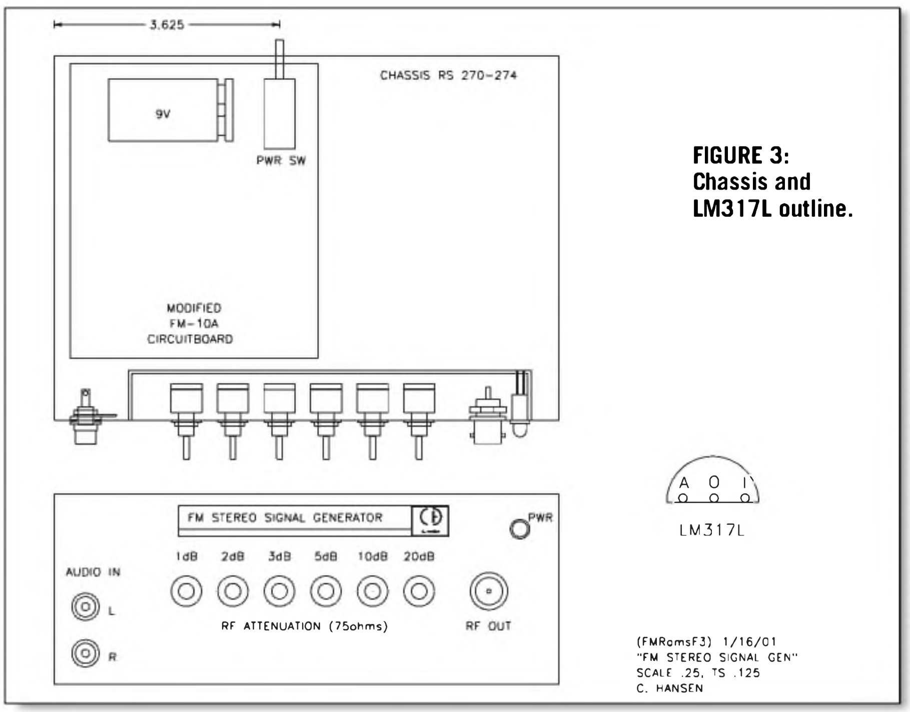


FIGURE 3: Chassis and LM317L outline.

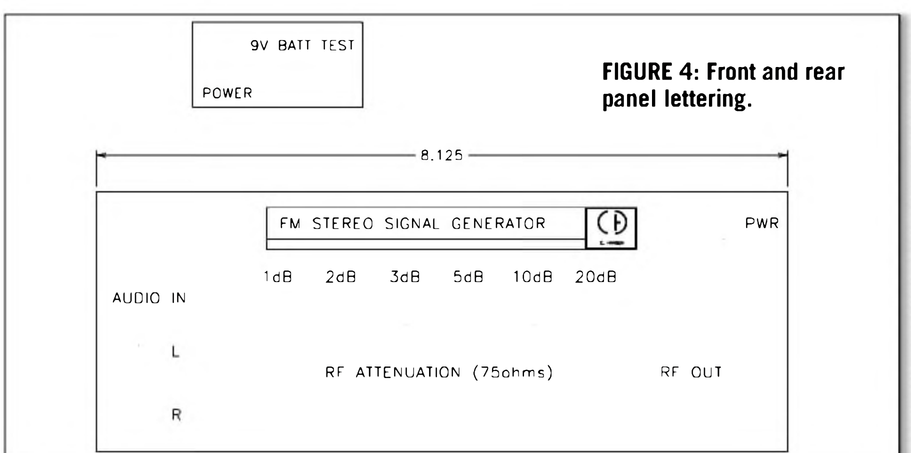


FIGURE 4: Front and rear panel lettering.

CONSTRUCTION

Photo 1 shows the front view of my prototype, with the rear view in *Photo 2*. I used a Radio Shack 270-274 steel chassis for this project. *Figure 3* shows the top and front views of the FM stereo generator chassis layout and the outline for the LM317L IC. Note the differences from the prototype photos, for the reasons described previously.

Figure 4 shows the panel lettering designations for the front panel. A full-size copy of this lettering was made on drafting appliqué film, which is an adhesive-backed transparent plastic (see note after the parts list). Apply the lettered film to the front of the unit in the proper locations. Cut through the film at the drilled holes and then spray it with two coats of clear polyurethane for protection.

You must accurately locate the hole for the power-switch actuator, since the switch is soldered to the PC board. Install the metal standoffs on the board and position the switch in its proper location on the board. Mark the inside of the rear panel at the actuator and drill a 1/16″ hole.

Be sure the switch actuator is in the on position. Locate the board so the actuator protrudes through the chassis in its on position and mark the locations for the standoff holes in the chassis bottom. Drill a hole for the battery test pin jack, in line with the rear panel label.

Photo 3 shows the interior of my prototype signal generator. I mounted the FM10A PC board (*Photo 4*) on four tapped aluminum spacers, so it makes a good RF connection to the chassis. Also mount the audio input jacks and RF output jack directly to the chassis.

Single-point star grounding is the rule with audio equipment, but RF grounds need to be short and direct to avoid parasitic inductance and wire lengths that can act as antennas. Be sure to scrape the paint off the chassis wherever a ground connection is required, and where the two chassis halves meet at the cover mounting screws.

Use shielded wire from the phono jacks to the circuit board and 75Ω coax from the board to the internal attenuator (if you decide to build one). Wire the RF output connector directly to the last attenuator switch. You will need to shield the attenuator from the transmitter board. I used copper foil, but sheet steel is much more effective.

TEST AND CALIBRATION

To perfectly adjust the FM stereo generator (Method 1), you need some RF test equipment: an RF millivoltmeter or 75Ω RF power meter, an RF frequency counter, and a general-purpose oscilloscope. However, you can do a decent job (Method 2, later) with a high-performance oscilloscope and a modern (digitally synthesized) FM tuner. My NAD tuner has 50kHz tuning increments, which make it easy to precisely tune the stereo generator. The scope must be capable of at least 150MHz; more is better (discussed later).

You must make all connections with 75Ω coax cable. Cable TV also uses this impedance and video "F" connectors (probably also used on your FM receiver). You can use cable-TV coax with an F-to-BNC adapter (Radio Shack P/N 278-251) at the FM stereo generator, or build your generator with the bulkhead "F" connector in the parts list.

If you find there is no RF output—or some other problem—the FM10A kit instructions include a pretty good set of troubleshooting steps. Since you have modified the PC board, you may not have access to warranty claims. You can't get the BA1404 IC just anywhere. I did find a couple of other surplus sources, but there are no more BA1404s other than those that Rohm made up until 1994.

METHOD 1

Turn off all the internal attenuator switches. Connect the FM stereo generator to the RF frequency counter and RF meter and adjust the L-C tank coil slug to the proper frequency. Remember that US FM stations are assigned to the odd 100kHz spacings (98.1MHz is valid, 98.2MHz is not). You need to load the stereo generator with a 75Ω terminator (in the RF probe or at the counter). A good BNC terminator is the Mouser 523-46650-75RFX.

To adjust the RF power, use the 1k pot in the LM317L regulator circuit that feeds the collector of Q1. I found I needed a minimum of 3V DC for reasonably linear RF modulation (<2% THD at the receiver). To make life easier, adjust the RF to a power level that equates to a multiple of 5dBf output (*Table 2*). You

TABLE 2
DBF FOR RF REM VOLTAGE/POWER INTO 75Ω

DBF	RF VOLTAGE	RF POWER
100	27.4mV	10µW
95	15.4mV	3.16µW
90	8.66mV	1µW
85	4.87mV	316nW
80	2.74mV	100nW
75	1.54mV	31.6nW
70	866µV	10nW
65	487µV	3.16nW
60	274µV	1nW

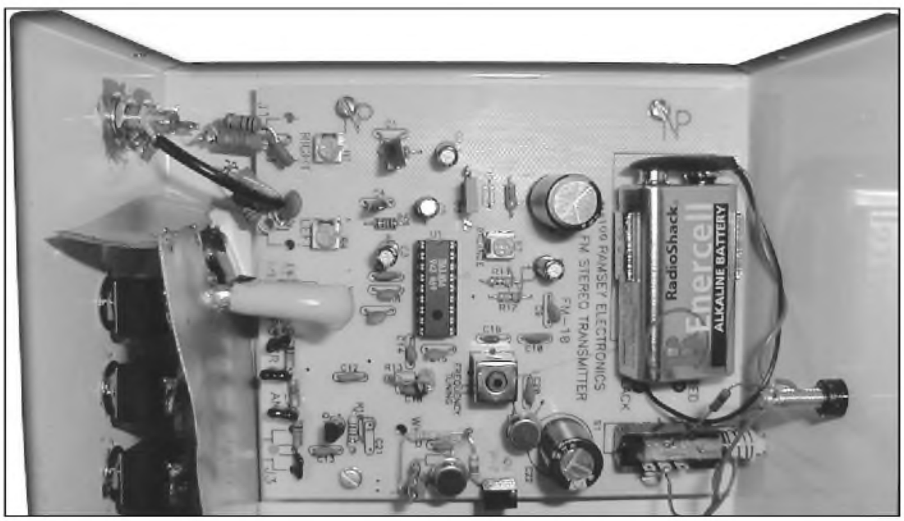


PHOTO 4: Modified FM10A circuit board close-up.

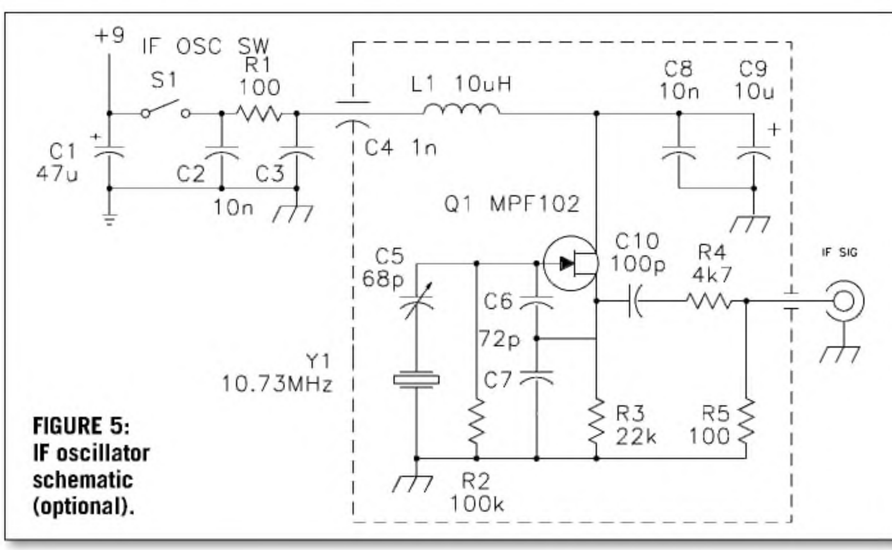


FIGURE 5: IF oscillator schematic (optional).

will probably not be able to get below 80dBf. Full FM quieting is usually specified at 65dBf, so some form of RF attenuation is required. The base level of 0dBf is equal to 0.274μV or 1fW.

Now turn on each of the internal RF attenuator switches and measure the RF power reduction as a result of adding each RF π-filter. This will allow you to make an accurate attenuation chart for your FM signal generator.

METHOD 2

Turn on all the internal RF attenuator switches. Connect the FM signal generator to the 75Ω input of an FM tuner, and then to a high-frequency oscilloscope using a Mouser 532-31-208-RFX BNC T-adapter. (This is a 50Ω device, rather than the desired 75Ω, but it should work okay since the tuner terminates the generator properly.) If your tuner has only 300Ω input terminals, use a 75Ω–300Ω balun transformer between the generator and the tuner.

The FM10A frequency-adjustment slug provides pretty good resolution, and you can use the tuner scan button to find the frequency where the FM stereo generator is actually tuned. Once you are close to your desired operating frequency, pre-tune the FM tuner to that frequency and adjust the L-C tank coil slug until the FM tuner locks onto the stereo signal. Adjust the tuner in each direction, to the points where the FM stereo indicator goes out, and then point it in the center of these two positions. Many tuners have a meter or bar graph tuning/signal strength display, which makes life easier. You may need an external RF attenuator if the tuner overloads.

You can view the FM10A RF voltage output on a high-frequency oscilloscope and adjust the collector supply for Q1. Remember two things. First, the scope is presenting the peak-to-peak RF voltage, so you need to divide this by 2.828 to get the RMS value in *Table 2*. Second, depending on the scope's frequency rating, the level displayed on the scope trace may be less than the actual RF output. If you are using a 100MHz scope, and the FM10A is set to 100MHz, the actual RF voltage may be 1.414 times higher, since the 100MHz rating of the scope is probably its –3dB response point. If you are lucky, there will be a response-versus-frequency chart in your scope manual.

Once you have made the RF adjustments, turn the two 1k audio trimpots about two-thirds counterclockwise and connect an audio oscillator to the two audio input jacks with a Y-adapter (Radio Shack P/N 42-2438). Keep the signal very low (20mV RMS or so at 1kHz) and view the audio output at the line output jacks of the FM tuner or receiver. Increase the audio signal level until the signal clips, or becomes "splatty" with spikes or spurious oscillation.

Now change to 15kHz audio, which is the upper bandwidth limit for FM radio. Decrease the audio level so you have a clean sine wave. A high-quality FM tuner will have flat (±1dB) frequency response from 30Hz to 15kHz. When the audio level is properly adjusted, the FM10A is capable of linear RF modulation within this range.

Now, gradually decrease the RF attenuation, keeping an eye on the 15kHz sine wave at the output of the tuner. When the sine wave becomes clipped, you have reached the maximum RF input level for the tuner. Most quality tuners can accept 80dBf or more. Auto radios, which have an additional RF amplifier stage to cope with constantly varying FM reception, usually go into overload below this point.

If you have an audio distortion meter, you can adjust the RF level with the Q1 supply trimpot for minimum 1kHz distortion at the tuner's line outputs.

USING THE FM STEREO GENERATOR

I use the FM stereo generator for making sensitivity and frequency response measurements on FM receivers. I found that the FM10A chip's modulator distortion exceeds 1%, so it is not suitable for audio distortion measurements. Really good FM tuners have distortion levels below 0.3% for stereo reception. It is also useful for measuring stereo separation where you can turn off the tuner's stereo-mono blend circuit.

I preset my FM stereo generator at 100nW (80dBf) with all the internal attenuation switches on. As I stated earlier, I use an external 0–70dB RF attenuator to reduce the RF level to the 65dB audio quieting points in mono and stereo and to find the minimum useable FM signal level (sometimes this is less than the fully attenuated 10dBf level of my test setup).

Figure 5 shows an optional 10.7MHz IF (intermediate frequency) marker oscillator that you can use to align the IF strip in older FM receivers having IF bandpass tank circuits with adjustable slugs. Some of the finest FM tuners from the vacuum-tube era use this method. This Colpitts oscillator is independent of the FM stereo generator and can be built in a separate enclosure.

C6 and C7 should be silver mica or NP0 ceramic caps. C4 is an RF feedthrough cap. C5 is used to tune the crystal to 10.7MHz. The *ARRL Handbook* has many interesting and pertinent RF circuit designs[4]. ❖

SOURCES

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

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

Parts Express
1-800-338-0531
www.partsexpress.com

Radio Shack
Local stores, or
Radio Shack Unlimited (RSU)
1-800-THE-SHACK
www.tandy.com

Ramsey Electronics
793 Canning Parkway
Victor, NY 14564
1-800-446-2295
www.ramseykits.com

REFERENCES

1. High Fidelity FM Stereo Modulator Circuit Radio Free-Association, (C) G. Forrest Cook, July 6, 1998; revised Apr. 14, 1999, June 21, 1999. http://www.solorb.com/elect. This site has many links to FM stereo transmitters, modifications, and other related circuits.

2. The Rohm BA1404 FM stereo transmitter chip is no longer in production, but Ramsey Electronics uses it in several transmitter kits and seems to have a significant supply.

3. "An RF Step Attenuator," D. Bramwell, *QST* Magazine, 6/95, pp. 33–34. This article presents a 50Ω attenuator. Just multiply the resistor values by 1.5 for 75Ω, or use the values in *Fig. 2* of this article.

4. *The ARRL Handbook for Radio Amateurs*, Amateur Radio Relay League, 1-888-277-5289, pubsales@arrl.org.

e-Bass: An Application of Phase Linear Sub-Bass Equalization

Reprinted with permission from *Electronics World*, February 2001.

By Graham Maynard

Hi-fi, stereo, semiconductors, digital, each of these milestone audio developments has contributed towards the accuracy, miniaturization, and power of affordable music reproduction systems. However, and in spite of such technical developments, today's low distortion tuners, CDs, and 20-20 flat amplifiers cannot ensure that we will enjoy "fully realistic" sound reproduction. This is because their audio outputs are still used to drive permanent-magnet loudspeakers in much the same way as they did during the twilight years of the classic valve era.

I have no wish to denigrate any quality audio system—whether solid or hollow state. Most can generally be adjusted to perform well, but no matter how low their distortion, they cannot properly replicate lower bass sound waves with a flat 20Hz to 20kHz hi-fi response. This is because

- loudspeaker transduction efficiency falls with reducing frequency
- socially acceptable playback levels make some low-frequency boost necessary to compensate for the natural fall-off in human hearing abilities.

EQUALIZER EFFECTS

You can use graphic equalizers and electronic crossovers feeding independently adjustable multi-channel amplifiers that drive separate bass, mid, and treble loudspeakers to construct the most impressive of sound systems. Yet no matter how you adjust the controls to counter loudspeaker and hearing deficiencies, fully realistic reproduction remains frustratingly elusive. This lack of fidelity is due to the frequency-selective-amplitude corrections for the natural rolloffs that occur within the audio spectrum. These corrections lead to our hearing becoming subjected to frequency-variable phase changes and time delays that affect waveform coherence and transient delivery.

Phase changes are why, for example, a little necessary bass boost from a Baxandall tone control or equalizer slider can sound okay. However, the greater amounts of amplitude correction that we might ultimately wish to correctly apply at lower listening levels actually sound somewhat off-putting. A "loudness" control is little better, and if one is used, in addition, to bass boost, then the sound becomes atrociously soft and boomy.

Purists might use headphones, or stick to a flat amplifier response that they know does not alter signal phase or waveform coherence. Then, after selecting the best loudspeaker equipment that they can afford, they simply accept the natural low-frequency rolloff that they can do nothing about.

Fortunately for experienced ears, flat amplification has little effect on the reproduction of choral works and the musicality of classical string and wind instruments. Yet, while it is the music and not the sound system that we listen to, it occasionally becomes obvious that some recordings have been made via studio equipment that had itself not been running fully flat!

REALISM

For most listeners, however, flat amplification and conventional loudspeakers will lead to a considerable loss of realism at playback. This is especially so when attempting to reproduce electronic and percussion instruments, full orchestra and open air performances, or the deep lows "ambience" of larger auditoria where longer wavelengths have had the freedom to develop a "spatial" sound that has been "characterized" by the building itself. We might not be able to hear 10Hz, but infrasound beats and standing-wave sensations are physically observed when we attend live classical and pop music performances. These become essential components of our learned experience.

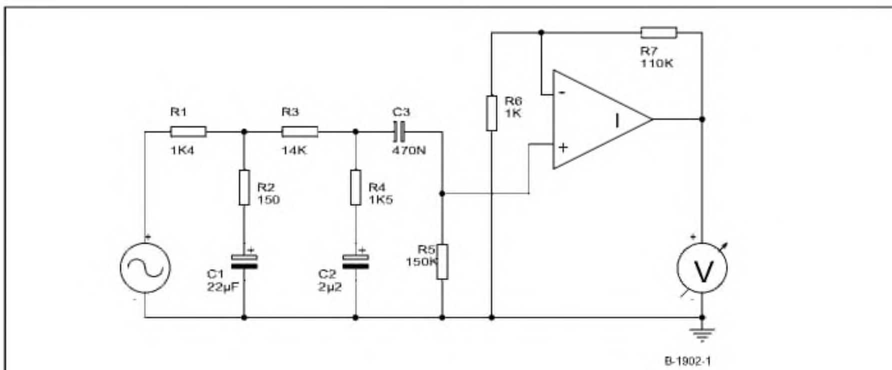


FIGURE 1: Using a small cabinet, incorporating a rigid-coned speaker and mounted against a wall, offers many benefits over traditional bass enclosures. The drawback is that such a speaker needs considerably more power as frequency falls. This filter provides boost rising to 20dB at 15Hz.

When playback is carrying the subliminal harmonic and mixed signatures of such deep lows but their body-vibrating fundamentals are missing, then we cognitively realize that reproduction lacks the full set of characteristics that are essential to match our known experience, and thus to produce the illusion of "being there." Good recording equipment does transfer infrasound frequencies to masters. When these are not filtered out prior to the disc printing run, then good CD players and systems can be made to reproduce them. This might be a tiny part of the sensory spectrum, and it is not always recorded, but it does remain very important: If you believe that the bass currently reproduced by conventional audio systems is not seriously compromised, then take a listen to known good music test tracks when played back through quality headphones.

EXISTING BASS EQUALIZERS

Many active bass-equalizer circuits have already appeared in electronics publications. Most are additional inline stages, though some use the power amplifier's negative feedback loop. Those that start to roll off around 20Hz, or which incorporate a tuned-feedback loop to achieve their boost, make me cringe as I imagine the nature of the bass sound that would be reproduced.

With any frequency-selective amplitude adjustment or turnover, whether it be passive, electronic, or loudspeaker in origin, there is always a phase change about and beyond its turnover frequency. If signal phase changes occur in the 15 to 100Hz range, then low bass sound waves that initially had a coincidental starting point can recombine differently at playback and make an instrument sound as if it has been physically altered. Occasionally, combinations of loudspeaker and equalization characteristics so badly affect the sound that fundamental tones can run into quadrature with their own range of reproduced harmonics. Alternatively, they can become out-of-phase at frequencies that are not too far above and below what becomes an effective resonant "Q" like turnover well within the audio spectrum.

When an amplifier or a loudspeaker amplitude response starts to roll off at, say, 20Hz, it is generally capable of good reproduction above 40Hz. However, even though normal musical sounds seldom have fundamentals below 30Hz, reproduction can still be shown to be phase shifted, and thus harmonically distorted. This is because a small deficit in the amplitude of a low-frequency fundamental introduces a larger level of phase shift with respect to its entire family of harmonic overtones than would be the case if only a few of the upper orders had been similarly altered.

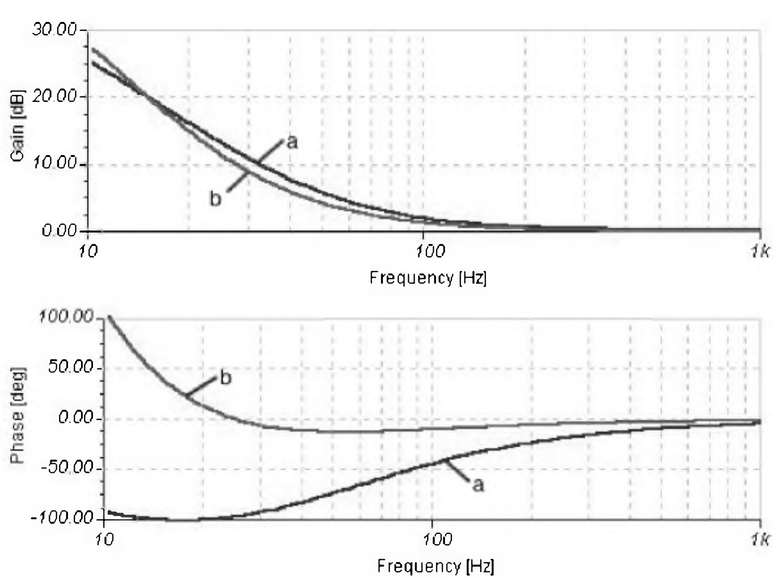


FIGURE 2: Sub-bass amplitude and phase characteristics for the *Fig. 1*, and "e-bass" circuits. The a traces result from *Fig. 1*, while the b traces are the e-bass circuit output characteristics.

It is true that crossover networks also cause phase changes. Usually though, the aim here is to counter the reproduced shift in one driven channel with an equal but opposite effect in its continuing partner.

WHY IS PHASE IMPORTANT?

At the low-frequency limit of human hearing, where no other driver takes over, such phase distortions generate what we commonly term as a loudspeaker or system footprint. It has been said that we are less capable of noticing distortion at low frequencies, but I wonder what the reference has been. I also wonder whether this conclusion has been drawn while listening via conventional loudspeakers and amplifiers that start to roll off at 20Hz.

When phase changes occur, a range of frequency-related time delays develops. Previously accompanying wave components become separated at playback; i.e., they become displaced in time. The resultant sound quite literally loses its coherence. Occasionally, the harmonic "voice" of a bass instrument can be perceived before the peak of the first half cycle of its fundamental note. It is as if the bass loudspeaker cone is operating from farther and farther away with increasingly lower frequency.

There is thus a most definite need to maintain an overall flat phase response at all audible frequencies. Also, the associated amplitude linearity needs to extend as far below hearing as is stably possible, without risking equipment reliability. Filters that alter amplitude introduce phase-related effects that extend out to two or three times, also to a half or a third of their nominal turnover frequency. Because of this, I personally consider that an amplitude rolloff should not be allowed above 5 to 7Hz for amplifiers and loudspeakers that are expected to perform linearly down to the lower limit of human hearing.

From a constructional viewpoint, minimizing bass driver and cabinet de-

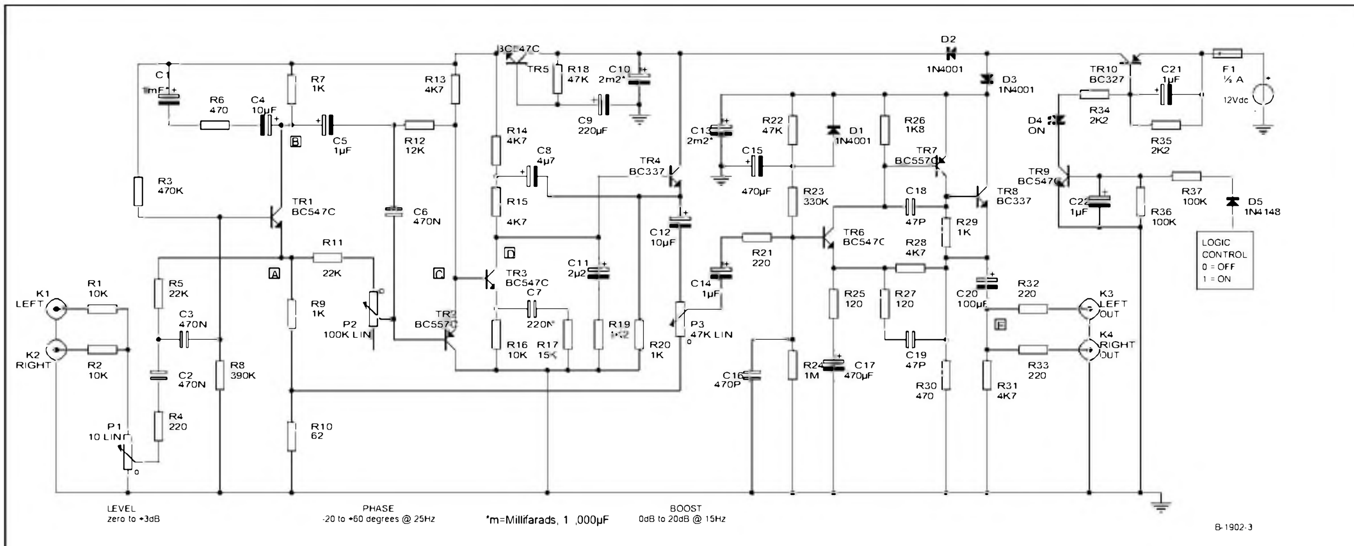


FIGURE 3: Although more complex than the filter of *Fig. 1*, this "e-bass" compensation circuit provides a much more faithful phase response, resulting in much-enhanced fidelity at sub-100Hz frequencies. A 5Hz input filter removes subsonic effects of CD gain and level settlement data that shouldn't occur—but sometimes does.

rived phase shifts below 100Hz then precludes the use of baffles. They also preclude large, ported or labyrinth cabinets and low-frequency auxiliary bass radiators that look like heavily coned loudspeakers without magnets. These preclusions apply even when any of these options is driven via a specially designed active equalizer element that maintains an overall flat audio response via that particular system. This is because air column and mechanical resonances become independent of signal input once they have been exited.

Resonant displacement can be controlled only via physical damping and/or motional feedback. Yet the solution is simple.

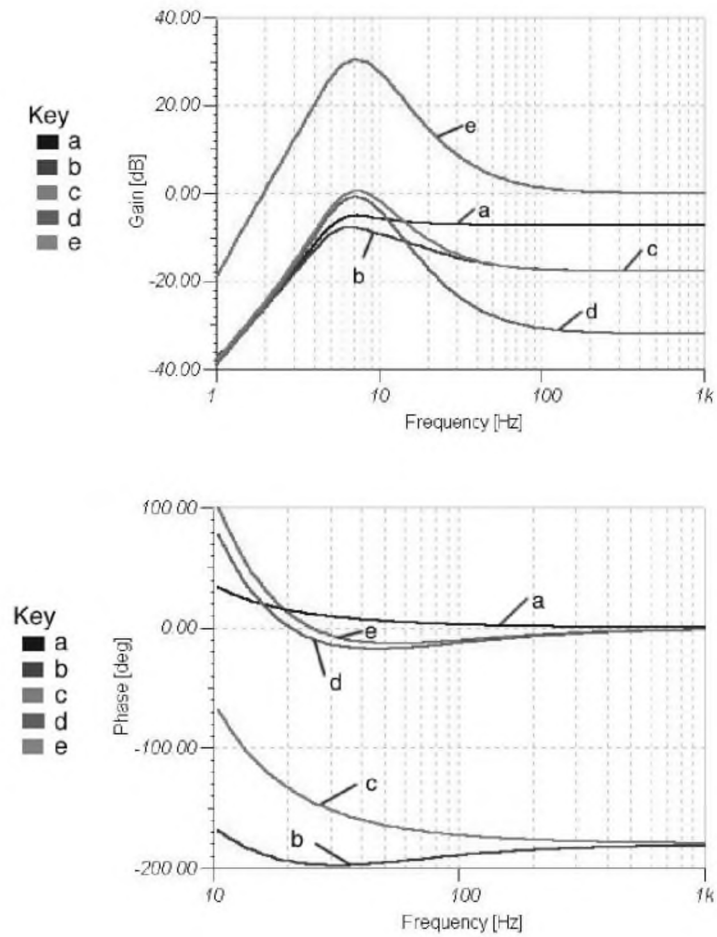


FIGURE 4: Signal amplitude and phase characteristics at points within the "e-bass" equalizer circuit.

SOLUTION TO PHASE PROBLEM

Take a rigidly coned long-throw driver motor, mounted within or against a small airtight enclosure. When crossed over and used resistively only at those frequencies that are below the new relatively high resonant frequency, it will be more phase linear, more reliable, and more compact than the conventional cabinets that we have come to accept. However, a power transient induced resonance within such a compact cabinet that is fed by a poorly damping L-C crossover can be unpleasantly distinct when compared to larger and conventional cabinets. As a result, an electronic crossover with a directly damping amplifier to voice coil connection becomes essential.

Also, the use of any resistor-capacitor type signal equalization or an electronically resonant circuit that generates necessary sub-bass boost then merely introduces other separate but additional phase changes and/or a rumbly single note overhang effect. These distortions are as if the driver itself had been made very large, been mounted in a much larger enclosure, or had itself become under-damped. Thus the mechanical phase problems due to a loudspeaker being mounted in a conventional cabinet could erroneously be replaced by newly introduced electronic ones that sound equally undesirable. You simply cannot reproduce a recording of clean and tight sounding bass on any system that is not itself clean and tightly damped in every regard; that means electronically, electrodynamically, and physically.

EQUALIZATION

Of course, this suggested compact loudspeaker solution is also seriously inefficient at generating sub-bass frequency sound waves. A considerable boost to signal amplitude becomes necessary at 15Hz with respect to 100Hz; (approximately 20dB). At 15Hz, this 20dB boost then places serious demands upon amplifiers and drivers. A handling requirement of 100W at 15Hz becomes necessary to complement 1W at 1kHz.

Figure 1 shows a passive multi-stage resistor-capacitor filter network capable of providing the necessary boost. Its simulated amplitude response is trace a in the upper half of *Fig. 2*. Also note, however, that the phase response examination in the lower half of *Fig. 2* reveals a considerable turnover lag at all audible bass frequencies. You will observe this same phase lag on any normal combination of active or passive equalizer circuits that act with respect to ground to provide the same slope of sub-bass increase. This degree of phase shift runs bass instrument fundamentals into quadrature with the rest of their characterizing harmonics, and renders boosted sub-bass reproduction woolly beyond tolerance.

During the summer of '98, I devised and test-auditioned many different circuit architectures that were each individually capable of generating the same 20dB of amplitude boost at 15Hz. All were measured to check that their equalization mirrored, and therefore compensated for, the natural rolloff of a dynamic loudspeaker working against a compact airtight enclosure. This meant a response that would hold 12dB per octave down to 10Hz.

In spite of their identical audio amplitude responses, each circuit sounded different. Some were grossly unacceptable due to phase distortion. Then I realized that my dual-trace oscilloscope could be used to display the output waveform against signal generator input. This would allow me to observe any phase shifts that accompanied gain directly.

Eventually, I achieved a correct low-frequency amplitude boost characteristic. It always delivered output in the correct sense per sine-wave half cycle.

This means that it delivered zero crossing AC voltage input and output transitions that were in phase at all audible frequencies. My desire was to deliberately counter the more usual trace-a like bass frequency lag that I had already heard, measured, and so disliked, resulting from filters similar to the one shown in *Fig. 1*. Much experimentation and repeated trials led to the development of my final circuit, *Fig. 3*. The b traces in *Fig. 2* show the much flatter phase response achieved by my similarly boosted e-bass circuitry.

THE CIRCUIT

In *Fig. 3*, TR1 forms a low-Q 5Hz input filter. (*Table 1* is the parts list for this circuit.) This is necessary to reduce infrasound playback fluctuations. These can be caused by digital-audio bias and gain level settlement data that should not be—but nevertheless occasionally is—imprinted upon some CDs. This filter has a negligible effect on amplitude at audio frequencies, though it does introduce the unavoidably increasing phase lead that comes in below 15Hz. See trace a in *Fig. 4*.

All mid-high audio frequencies are twice phase reversed—first at TR1 collector (trace b), and then again at TR3 collector (trace d). TR3 output, buffered by TR4, is thus shifted back to being fully in phase with TR1 base input, yet without delay.

Bass frequencies are separately taken from the emitter of TR1 (trace a) with respect to the 180° phase shifted signal at the collector of TR1 (trace b). The sub-bass frequencies at TR2 emitter (trace c) thus develop an intermediate and frequency-dependent phase lead with respect to the higher frequencies at TR1 collector. This intermediate lead counters the *Fig. 2* trace a like lag that is caused by the passive resistor-capacitor high cut and thus sub-bass boosting networks acting on both TR1 and TR3 collectors with respect to ground.

The outcome at the emitter of TR4 is that all audio frequencies now remain in phase with the input. Bass frequencies have received a 12dB/octave characteristic boost through +20dB at 15Hz, with respect to 100Hz and above.

Audio output at TR4 emitter is fully in phase with the signal at input. As a result, a simple potentiometer arrangement between it and the first stage, VR3, is all that is necessary to tap off any desired level of phase-linear sub-bass boost between flat and +20dB at 15Hz with respect to 100Hz. The greater the level of boost, then obviously the more that potentiometer VR1 might need to be backed off to prevent sub-bass power-amplifier overdriving.

During listening tests, it became obvious to me that some CDs had been produced using equipment that had not maintained a flat phase response at sub-bass frequencies. Additionally, some unused equalizer phase shifting ability remained available between TR1 and TR2, so I introduced the third potentiometer.

TABLE 1
E-BASS CIRCUIT PARTS LIST

REFERENCE	PART
C1	1mF
C5, C14, C21, C22	1µF
C2, C3, C6	470N
C4, C12	10µF
C7	220N
C8	4µ7
C9	220µF
C10, C13	2m2
C11	2µ2
C15, C17	470µF
C16	470P
C18, C19	47P
C20	100µF
D1, D2, D3	1N4001
D4	LED ON
D5	1N4148
F1	½A
K1, K3	In & out left
K2, K4	In & out right
P1	10Ω LIN
P2	100k LIN
P3	47k LIN
R1, R2, R16	10k
R3	470k
R4, R21, R32, R33	220
R5, R11	22k
R6, R30	470
R7, R9, R20, R29	1k
R8	390k
R10	62
R12	12k
R13, R14, R15, R28, R31	4k7
R17	15k
R18, R22	47k
R19	1k2
R23	330k
R24	1M
R25, R27	120
R26	1k8
R34, R35	2k2
R36, R37	100k
TR1, TR3, TR5, TR6, TR9	BC547C
TR2, TR7	BC557C
TR4, TR8	BC337
TR10	BC327

This potentiometer could prove useful in compensating for other undesirable loudspeaker and crossover shifts.

I chose a component value that would allow a fully variable, but independent, sub-bass shift of between +60 and –20° at 25Hz. Thus phase was adjustable relative to unequalized audio that was being reproduced separately by the main bass-mid-treble component drivers. The result was a uniquely useful and user-controllable range of phase variation!

Initially, up to –60° had been available, but just as with ordinary bass controls, this amount of lag never sounded acceptable. Zero degrees is dead center on the 100kΩ pot, and this turns out to be the most common position set by ear.

TR4 stably bootstraps TR3 down to 7Hz. This optimizes stage gain without any need for low impedance components at TR3 emitter. TR6, 7, and 8 form a simple two-stage, self biasing 2V RMS output amplifier, which counters equalization circuit losses when running flat.

POWER SUPPLY

TR5 isolates the equalizer filter from supply-rail ripples. Thus the entire circuit eliminates loudspeaker thumping at power-up, which is something that op-amp circuitry cannot easily do. TR9 and 10 form an electronic on-off switch, thereby enabling remote DC control from a preamp or head unit, but these controlling transistors and their associated components can be omitted. If you decide to do so, diodes D2 and 3 will need to be powered directly from a main amplifier supply or via a separately switched rail.

This class-A (constant-current) circuit requires a single, positive, 11V to 16V (25mA) DC supply. Fed via D2 and 3, it is tolerant of momentary psu interruptions, ripples, and power-audio induced supply rail voltage dips when running from a common supply or car battery. Actually, D2 and D3 can be connected to a raw AC supply. Where the equalizer must be installed at the end of long cable runs, a separately powered DC to DC converter might be considered. This would be inserted between the diodes plus the signal earth line, and TR9 plus the power earth line. Adding a converter in this way would isolate audio circuitry from DC rails. I have not experienced any supply rail or common earth problems with prototypes, mainly due to the sharp rolloff below 5Hz.

REMOTE CONTROL

During the design phase, I connected VR1, 2, and 3 so they could be controlled remotely from the listening position. I used a six-meter screened multicore cable for this. No instability, interference, or induced hum pickup problems were encountered.

My thanks to Andy Collinson at www.mitedu.freeserve.co.uk for the use of his TINA simulator program to generate the amplitude and phase graphs directly from my final circuit. His simulations confirm exactly my own longhand measurements made on 1998 prototypes. In *Fig. 4*, amplitude trace e shows a sharp turnover between the 12dB/oct boost and the 20dB/oct cut below 5Hz; this sharpness is achieved over several stages and thus is not resonant. Also note that trace e shows the overall phase deviation between input and output not exceeding ±13° down to 20Hz—figures that hold for the zero boost trace a phase line: Can any conventionally cabinetted loudspeaker do better?

REALIZATION

The most significant test bed for this equalizer has been my son's 50Hz electronically crossed over sub-bass channel, feeding a Rockford Fosgate (USA) 100+100W 4Ω amplifier that maintains output down to 5Hz. This drives a Rockford Fosgate 15″ diameter double 4Ω voice coil bass woofer, sealed inside and thus occupying some of the volume of a 2.3ft^3 sealed enclosure.

With one amplifier channel per voice coil, I observed a much superior cone control, and the combination is capable of displacing up to one liter of air per half cycle; that is two liters peak to peak! As this amplifier is capable of handling down to 1Ω per channel, I tried it in a bridged configuration with the voice coils in parallel to form a 2Ω load. The bridged setup functioned exactly as expected; unnecessarily powerful with audible degradation in cone driving accuracy and damping.

An equalized power of 200W cannot set any SPL records when driving a compact sub-bass loudspeaker, but it is notably more realistic. The low frequencies of beating bowed cellos properly resound, and kick drums cleanly and tonefully kick air rather than thud-boomphing as if not a drum at all. Also, reproductions from CD render more life-like stage and theater hall acoustic ambiences. This aspect is hard to describe, but it produces the illusion that your hearing is extended far beyond loudspeakers and the listening room. Open-air recordings actually become more “open,” with distant sounds clearly sounding distant because they are not modified by those rumbly bass loudspeaker colorations that we have become so used to, and which we can hear being differently exited on different systems as a result of air disturbance on a source microphone.

POSITIVE CLIPPING

Another positive attribute of the separately powered sub-bass channel is that when a sub-bass amplifier runs into clipping, the rest of the audio system remains unaffected. Thus, momentary sub-bass overdriving does not distort other bass-mid-treble output. With a clean clipping sub-bass amplifier it can be determined that an insufficiency has occurred without it having had any mind-catchingly deleterious effect upon listening pleasure. This is simply not the case with an inadvertently clipping main or whole bass amplifier, the effects of which can totally capture your attention and ruin your enjoyment for a long time after the momentary distortion has occurred.

Although this equalizer has proven effective with long-throw bass drivers as small as 8″ in diameter, I trust that anyone who seriously attempts to produce realistic results will use a minimum single 12″ or twin 10″ drivers, one facing inwards the other outwards and wired oppositely. Such sub-bass speakers should possess a genuinely continuous power rating of at least 100W RMS. Also, the driving amplifier should clip cleanly and itself be capable of a continuous minimum 100W RMS output. Its amplitude response should hold flat down to at least 5Hz so that its phase response will remain flat through the audible spectrum.

7HZ TURNOVER

To ensure that reproduction phase changes do not occur at normally audible frequencies, this equalizer's low-frequency boost has been set to turn over at 7Hz. While this is similar to a normally specified half-power point, it is also the frequency at which the overall loudspeaker drive is allowed to fall away from a 12dB/oct boost characteristic. In other words, 7Hz is the frequency at which the equalizer and amplifier are no longer expected to provide increasing output with falling frequency. Fortunately, there are few genuinely musical signals that require power below 30Hz, so that a rising power output is seldom necessary down to 7Hz. The 100W at 15Hz to match 1W at 1kHz really should be more than sufficient for loud sub-bass reproduction, even with modern synthesized instruments.

Nor should we forget that all of these turnovers—whether from flat to boost or cut, or from boost or cut to flat—alter the phase response for a factor of two to three beyond any nominal turnover frequency. Thus the 7Hz rolloff introduces a phase advance that adds to the input filter below 17Hz. Had the gain characteristic merely been allowed to satisfy the accepted 20Hz "hi-fi" specification limit, then all audible sub-bass frequencies could have become phase distorted.

I tested this circuit for several months with a 2.5Hz input filter turnover, and 5Hz for the boost rolloff. The phase integrity was within ±20° down to 15Hz before the high-pass leading phase characteristics came in, but the resulting higher gain at infrasound frequencies then overemphasized previously unnoticed CD flutter errors. Although inaudible, these errors did pressurize the listening room and introduce an unnecessary risk of additional nonlinearity product generation.

Turnover points of 5 and 7Hz were eventually deemed the sensible compromise between what was possible with good CDs and what was realistically acceptable and practical on an everyday basis.

With 100 to 200W of amplifier drive and an equivalently RMS rated low-impedance sub-bass motor that is properly sealed against a compact and adequately rigid airtight enclosure, the cone simply cannot over-travel due to cabinet air incompressibility. Here I must emphasize that cone stiffness and cabinet rigidity are essential. Normal driver cones or cabinets having sides that flex when you stand on them can leave a sub-bass system not necessarily vulnerable to, but capable of, occasionally producing a most unpleasant and whip-cracking-like release of sound energy that can be induced by on-going voice-coil drive attempting to generate air spring pressures that cause reverse flexures in cone/surround and/or cabinet side walls. Do not use a woofer for sub-bass if its entire cone does not move as a piston when it is pushed at a single point just inside the surrounding edge suspension; the cone must not be flexible.

It is also worth noting that loudspeaker manufacturers have reduced bass driver coil impedances considerably, thereby enabling longer-wavelength air displacements to be generated without risking power amplifier output devices to failure with higher voltage supply rails challenging their SOAR limitations. Speakers with dual voice coils as low as 1.4Ω have already been manufactured in the US.

To reveal the qualities and weaknesses of your own system, I can recommend the highly competent Sheffield Labs Audiophile Reference test tracks that are supplied with recording plans in booklet form. These are compiled on the International Auto Sound Challenge Association competition CDs, also from the US. Check out the specialized links to manufacturers at www.iasca.com.

For sub-bass loudspeaker construction, a simple starting point is to mount the chosen woofer externally against (not inside) an already sealed, thick-walled cube, which should be lined and half loosely filled with natural fiber wadding. It should have internal side dimensions that equal the chassis' outside diameter. Of course, one internal side dimension could be doubled if the depth is halved and the wadding is placed in line with the driver cone.

There are many "per chassis" Thiele/Small parameter calculations that could be made, and I would not wish to dissuade constructors from satisfying themselves in these regards. However, calculated results could lead to the fabrication of an enclosure that is too large and thus not resistive enough for the 12dB/octave characteristic to be safely and smoothly applied. When used singly, loudspeakers that are intended for sub-bass reproduction actually run more cleanly if they are externally mounted onto an already constructed cabinet. This is because air flow through a tapered magnet vent is better able to cool the voice coil. As a bonus, harmonics that develop at the center of the cone are less able to radiate directly. It is as if the crossover has an additional pole at no extra cost.

COMPONENT CHOICES

I did not use any especially close tolerance components when making and testing prototypes. Resistors were sub miniature 2%, and all miniature capacitors up to 10µF were individually LCR metered to use only those that were within a 10% tolerance. For neatness I used multiples of miniature 470µF radials to make up the larger electrolytic values. Initial DC testing was completed by checking that approximately half of the supply voltage developed at the collectors of TR4 and TR8 after DC bias levels had been allowed to settle.

As all audio systems are different, the sub-bass crossover frequency will need to be determined by ear while simultaneously adjusting the equalizer potentiometers and any existing tone controls to level out an overall bass sound. This is not as difficult as you might imagine, and good results can be quickly achieved, but don't wrongly use maximum or impressive levels of boost, and don't use too high a crossover frequency.

BEING REALISTIC

While this circuit maintains a tolerably flat phase response at all audio frequencies, it is no better at generating a low-frequency amplitude boost at the instant it occurs in real time than are any of the other frequency selective amplitude adjusting networks already in use. Here I am referring to the ineffectiveness of capacitors below a turnover frequency and thus to the leading phase response that normally comes in below 30Hz. Thankfully our brains appear to process waveforms on an averaging basis, and when audible amplitude and phase responses are correct, we are much less distracted by the relatively few transient induced errors that here are now below the range of human hearing. I am just being entirely open here because most listeners are amazed by this equalizer's improvement to sub-bass sound when the significant phase shifts occur below 15Hz. Either we put up with larger loudspeakers and their audible weaknesses, or we accept equalized audio with occasional infrasound phase errors that are much less noticeable.

An overall flat amplitude with flat phase sub-bass response actually sounds different from what is currently accepted as being "normal" on conventional systems! Of course, VR1, 2, and 3 can be used to adjust the equalizer output towards a more direct and flat response if some source material already seems to be bass heavy or phase shifted. Medium boost with a leading phase can be applied if hardcore "dance" playback is required.

Some discs sound disappointing. This led me to ask what types of loudspeakers are used in some of the production suites that monitor our prospective purchases. I wonder whether producers watch out for or observe the flutter errors that become permanently imprinted upon their CD stamping masters after all dynamic and digital compression procedures have been completed, but before they go into production.

These low-frequency flutterings are to CDs what recorded rumbles were to some old vinyl discs, only here you can see them inaudibly shifting the bass driver cone. Take your grilles off and watch the bass cone move on some tracks. Actually, cone movement can occasionally be observed due to level compression or badly biased circuitry reacting to voice and solo instrument tones when no bass is being reproduced, though at least the vast majority of CDs do play satisfactorily.

FINALLY

As far as I am aware, this is the first "phase linear" sub-bass equalizer circuit to be constructed and published. It was born of my unwillingness to accept bass phase distortion on everyday systems, and it was completed through repeated long-term listening tests, with a determined empiricism and an enthusiastic disregard for any time that was necessary for development or evaluation. I have not written about the many designs that were completed, discarded, or superseded, or those that produced the same results but needed 1% capacitors, or which were less stable when driven with transients, and so forth.

After hearing the cleanly extended response and the clear increase in sound "dimension" that is achieved via a loudspeaker cabinet similar in size to those that they already install, many car entertainment enthusiasts asked for this equalizer circuit to be made for them, or for its circuit diagram to be supplied. They are amazed at the much greater level of low-frequency output that is correctly reproduced.

Where an ICE enthusiast can and does experience pocketbook deprivation due to the destruction of a driver that has been housed in a ported enclosure and driven with modern pop music tones that are below loudspeaker resonance, he or she can be reassured of reliability with greater sub-bass output when the same driver is used with in-line audio equalization and mounted against an airtight cabinet, or with tuning ports sealed over. Separate black box sub-bass equalization units are already available to ICE installers, but I have not yet observed phase linearity.

Like other *Electronics World* readers through the decades, I have had many opportunities to read about amplitude and phase relationships in audio articles. Soon after embarking upon this project, I was encouraged by the appearance of John Watkinson's "Speakers' Corner" series in *Electronics World*. His writing sustained my determination to produce a circuit that would not generate the phase shifts that are normally introduced by amplitude equalization when driving a compact and resistively loaded low-frequency loudspeaker, which is otherwise much more phase linear in its own right. Thus I regard his articles as being essential reading for anyone who has an interest in this form of bass reproduction; especially the October 1998 notes.

It is also worth noting that directly driven resistive loudspeaker enclosures that operate below their resonant frequency do not hit amplifier output stages with the same momentary and transistor popping dynamic impedance dips that conventional loudspeakers systems do.

I am not claiming that this equalizer is the best or only way forward, but it does genuinely promote our striving for ever more realistic sound reproduction. I look forward to seeing other phase correction related circuits or letters appearing in future *EW* and *aX* pages.

The circuit is straightforward and inexpensive, and is simple to insert between an electronic crossover and a sub-bass amplifier. It may even be used as the basis for an additional and standalone central sub-bass channel for use with already existing stereo systems, so that the original system might correctly be run flat.

Good-sounding cinema, disco, and professional loudspeakers need no longer be so large, and manufacturers could easily integrate an equalizer with an electronic crossover and sub-bass driving amplifier, to make 20Hz from 1ft^3 the domestic norm. Solid state sub-bass can also be an excellent partner for tube amplified hi-fi. Many possibilities exist!

Phase linearly equalized sub-bass channels inconspicuously add depth to the sound stage in a surprising way that cannot be appreciated until experienced, whereupon all "normal" systems seem lacking, even when good and expensive reproducing equipment is already in use. We cannot miss what we have not heard from our own CDs, but that is not a reason for doing nothing. Modern developments in amplifier and loudspeaker technologies already make the "e-bass" approach a practical possibility. So come on—get out your soldering iron. ❖

AUTHOR POSTSCRIPT

Feedback from my *EW* article came only from readers who did not actually build and audition this equalizer. They suggested that the best approach for sub-bass reproduction was by using bass lift with conventional drivers in well-damped cabinets. However, conventional drivers need gigantic enclosures for sub-bass, and this leaves them vulnerable to over-excursion; cabinet damping does not get rid of driver resonance; and bass lift muddies the sub-bass phase response.

In order to try this circuit, there is no reason why an evaluation prototype could not be built in an evening using brass pins pushed into thick corrugated cardboard, mimicking the breadboard construction of early 1920s tube radios. Test it using a 12V battery without the DC switching, and also make sure that the sub-bass channel crossover is sharp and below the driver-enclosure resonance, so that the compact combination cannot become excited by normal bass drive.

The circuit is not patented, though I can be contacted at graham.maynard1@virgin.net. Good luck with the construction, and do please let me know of your results.—**GM**

Harmony and Distortion, Part 2

Reprinted with permission from L'Audiophile, *Nov. 1981, pp. 41–55. Translated from the French by Suzanne Scara.* **By Jean Hiraga**

The different propositions regarding dissonance mentioned in Part 1 demonstrate that music contains a considerable number of pure frequencies which are knowledgeably combined and that the slightest alteration in the harmonics of a given frequency can destroy the sound harmony. Moreover, the ear is an extraordinarily sensitive receiver. Despite its size, thickness, and mass, a human eardrum vibrating to an amplitude of less than 1/10 of the diameter of a hydrogen molecule is sufficient for a person to perceive a sigh, a breath, or the sound of a delicate handkerchief falling. We are also aware of the ear's extraordinary acuity concerning the perception of frequencies and of the respective pitch of numerous frequencies emitted at the same time.

In the course of the past several decades, we have been able to note an improvement—either apparent or real—in the diverse components constituting systems of high-fidelity reproduction. After the era of triode vacuum tubes, there was a progression from pentode tubes to germanium transistors to silicon transistors to integrated circuits. All the same, in spite of the very clear improvement in electronic reproductive performances, frequency response, and the different forms of distortion, we very quickly took notice of the ear's astonishing acuity regarding the perception of nonlinear phenomena, however miniscule.

For amplifiers or preamplifiers, it has been recognized since at least 1975 (*Revue du Son*, Dec. 1975, Spectral Analysis of Distortion by Harmonics) that the contents of distortion were much greater than its total value, evaluated by percentage. Later on, several researchers, including Matti Otala, discovered the troubling effects of the active charge represented by the loudspeaker, on one hand, and on the other, the different problems of stability and dynamic linearity brought on by the negative feedback loops. Moreover, this author has often spoken of the distortion termed "soft" or "hard," the harmonic residue being either the predominant harmonic 2 (soft distortion) or a significant number of harmonics, including uneven harmonics (hard distortion), as well as of their influences on the sound quality.

The contents of harmonic distortion—weak as it may be—can directly influence the sound quality, or timbre. On the other hand, the measurements taken are most often carried out with the help of pure and fixed frequencies. In practice, the conditions are much more severe because millions of pure frequencies, rather than just a single

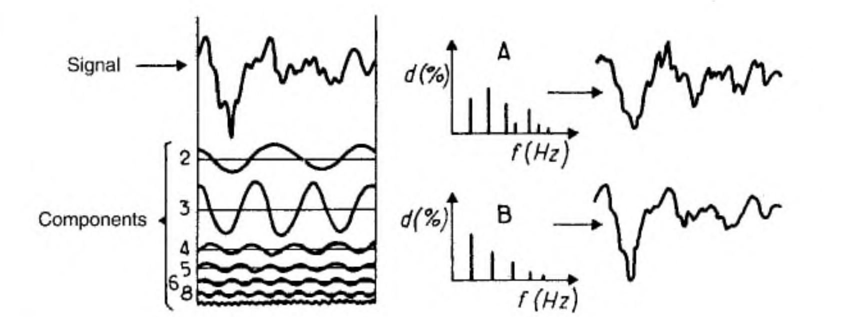


FIGURE 9: A complex signal (for example, a musical audio signal) subjected to an amplifier having an irregular distortion characteristic (A) or a regular distortion characteristic (B). You observe that the harmonics of high order, although very weak in level, combine to form an envelope of the amplified signal. In this case an irregular distortion spectrum distorts the original signal in a more or less pronounced fashion, whereas a "soft" distortion characteristic respects the envelope of the original signal, and hence the timbre of the instrument. A-hirag-9

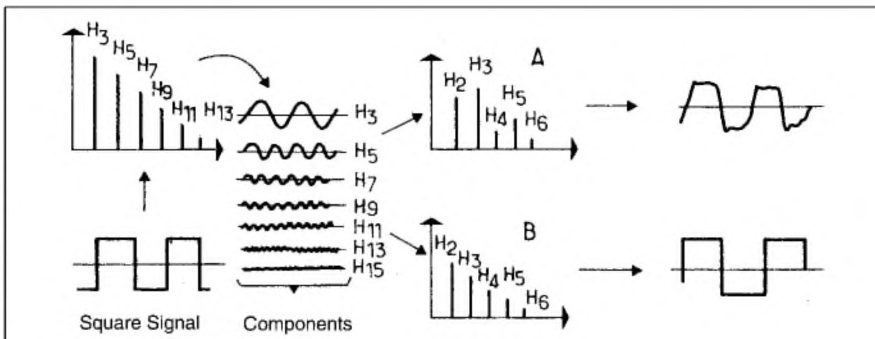


FIGURE 10: A square signal, whose spectral analysis is shown, subjected to two amplifiers—one having a "hard" and irregular distortion characteristic, the other "soft" and regular. Note that for B ("soft" distortion) the signal is not distorted. A-hirag-10

pure frequency, are involved, each having its own well-determined characteristics of frequency, frequency stability, phase, level, and level variation. If simple analytical measurements of harmonic distortion already show a badly graduated spectrum, unstable in terms of output level, frequency, and outgoing charge (a loudspeaker, for example), certainly the repercussions are going to be very clear regarding the auditory result.

Figure 9 shows a complex signal to which has been added a certain type of distortion (A). In B, note the addition of "soft" distortion, made by a regularly graduated spectrum. Observe that in B the resulting envelope of the original signal is only slightly affected. In A, where there is a weak rate of distortion that is nonetheless "hard" and unstable in transitory terms, the spectral envelope will be modified and the timbre will be unstable at the time of its temporal evolution.

Figure 10 shows a square signal to which is added either a hard distortion (A) or a soft and well-graduated distortion (B). In the latter, you still observe a more perfect respect for the original curve.

The spectral analysis of the rate of harmonic distortion is thus a very important concept, which bears a strict relationship to auditory quality. This is not, of course, the only criterion to consider, as this measurement is still too simple. Yet, for many amplifiers, this distortion spectrum is a faithful representation of the subjective quality.

Figure 11 shows several examples of this, with measurements carried out in 1975 by the Japanese firm NF Circuit on behalf of the magazine *Radio Gijutsu*. The different points considered up till now concerning the "harmony of harmonics" emitted simultaneously are sufficient to show just how important this characteristic is. With respect to "soft" distortion of "well-graduated" harmonic spectra, the minimally troubling effects of this form of distortion—which could be called "linear" distortion, or even more appropriately "well-tempered distortion"—can be very easily proven by the theories of Wegel and Lane, which appeared as early as 1924 in *Physical Review* under the title "The Theory of Auditory Masking of One Pure Tone by Another Tone, and its Probable Relation to the Dynamics of the Inner Ear."

These research efforts have been resumed several times and essentially concern the masking effects of pure sounds on each other and of complex sounds on each other. Since then, we have been able to establish definitive "standards" of the ear's characteristics when listening to pure sounds slightly "dirtied" by harmonics (harmonic distortion).

One of the most important, vis-à-vis the famous "beautiful gradations" of "soft" distortion, is certainly the masking effect—the exact values of each harmonic's level in a given condition—giving the ear the illusion of hearing nothing but a pure sound. The "beautiful gradation," contrary to what you might think, does not add to the sound, like a sort of "tail" made up of numerous harmonics, embellishing the sound. *Figure 10*, for example, shows that an irregular envelope of distortion gradation—even very weak in level—is enough to modify the envelope of the original signal.

DISTORTION AND MASKING EFFECT

With successive masking effects, it is possible—despite the presence of a fundamental note debased by harmonics (harmonic distortion)—to find a scheme in the level of each of these harmonics which produces by the effect of harmony, of the fusion of sounds, on one hand, and by multiple masking effects on the other, the illusion of hearing only a pure sinusoid. In practice, listening to a pure sinusoid is not very easy, and a pure sound is often considered as being, subjectively, a sound seemingly of rather average purity. This seems wrong.

In fact, for the vast majority of loudspeakers, even those whose linearity and distortion rates appear particularly exemplary, we have noticed that the sinusoid injected into the moving coil terminals does not seem to produce a perfectly pure sound. To this extent, we can already detect an instability in frequency and level, which will have repercussions such as those indicated above. The distortion measurement gives a spectral purity on the order of

only 0.5 to 1%, and the harmonic contents are never perfectly regular and stable (above all, dynamically). But in fact a sinusoidal sound is much more pleasant and "sweeter" than what you are in the habit of hearing through the membranes of current loudspeakers.

Whereas voice produces a very specific sound spectrum—an "imprint" that is easily revealed by sonogram (sound imprint)—a "whistling" sound of rather low frequency produces a sinusoid of rather high purity, though mixed with some white noise and affected by slight instabilities (frequency, level). Nonetheless, this sound seems purer than that reproduced by a loudspeaker and only a few rare compression chambers (sometimes not loaded by the horn) are capable of reproducing a sinusoid of great subjective purity. With such loudspeakers, a scanning in frequency produces a sound effect that is quite different from that of the current type of membrane-equipped loudspeaker: the sound is mellower, more flowing, more melodious, more whistled, and especially more pleasant. On this point, an experiment is much more effective than words to express these subjective impressions.

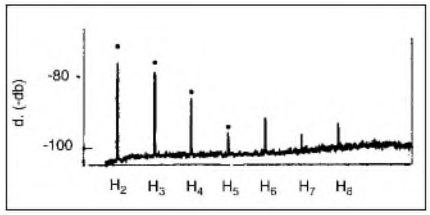
As early as 1930, Wegel and Lane were able to determine with great precision the respective levels of each harmonic able to yield an auditory result that—by successive masking effects and multiple harmony effects—gave the illusion of listening to only a pure sinusoid rather than a sound affected by harmonic distortion. For an acoustic level of 76dB SPL and a fundamental note of 400Hz, Wegel and Lane were able to establish that the harmonics required a level of:

- 800Hz (harmonic 2) 61dB
- 1,200Hz (harmonic 3) 58dB
- 1,600Hz (harmonic 4) 55dB
- 2,000Hz (harmonic 5) 50dB

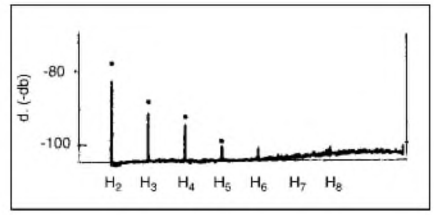
(SPL: Sound Pressure Level)

However, the measuring devices of the time were not of a sufficient purity to go any higher in the harmonic order.

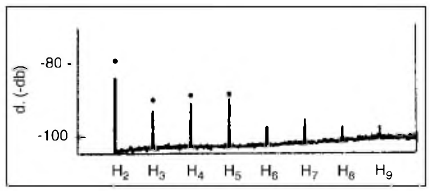
It was only toward 1960 that you were able to go up to high harmonic orders, such as the 20th harmonic. For the harmonics between orders 2 and 5 in



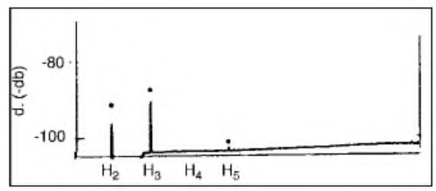
A 2 × 30W amplifier, moderately priced but having good musicality. Note the presence of the principal harmonics, the presence of H_6 and H_8, predominating slightly in comparison to the harmonics of odd-numbered order.



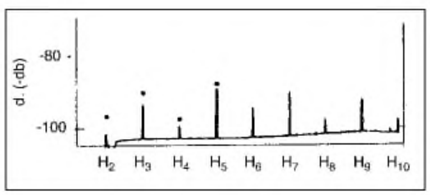
A 2 × 40W amplifier, moderately priced but having excellent musicality. Note the slight predominance of the even-numbered harmonics over the odd-numbered harmonics.



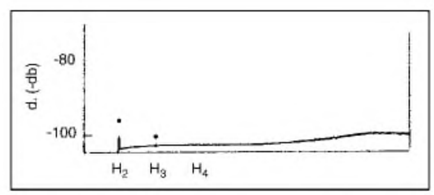
A 2 × 60W amplifier, of fairly good musicality, but having a sound that is a bit too "sparkling," seeming to lack low-middle register. Note the predominance of H_7 and H_9, odd-numbered harmonics.



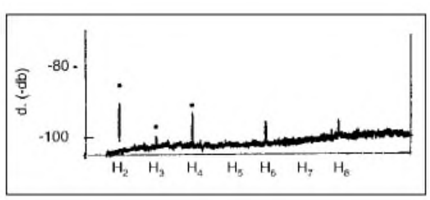
A high-powered amplifier (400W) characterized by a sound that is dynamic, but very harsh and lacking in definition. Note the troubling predominance of H_3 (highly perceptible to the ear), the absence of H_4 (circuit of partial negative feedback in the power stage), and the absence of harmonics of high order (high overall rate of feedback).



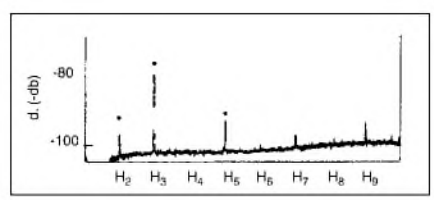
A high-powered amplifier (2 × 300W) operating in Class B. Note the absence of harmonic 2 and the predominance of the harmonics of odd-numbered order. The sound, a little more defined than that of the 400W amplifier, seems subjectively as if "shifted" toward the top of the spectrum. In Class B, the origin of these flaws seems to be the presence of several zener diodes and imperfect complementary transistors.



A Class-A amplifier of 2 × 50W. Renowned for its musicality and its exemplary definition. Note the regular gradation, in which the top part is masked by the background noise (−100dB).



A 2 × 40W amplifier. The very distinctive and symmetrical circuit has a tendency to absorb the harmonics of odd-numbered order. The sound produced, pleasant and sufficiently defined, seems slightly "rounded" and is a little reminiscent of certain tube amplifiers.



A 2 × 200W amplifier. The sound is especially harsh and unpleasant, "metallic." Note the effect of the "self-compensating" distortion circuit canceling out only the harmonics of even-numbered order.

FIGURE 11: Harmonic distortion characteristics of different transistorized amplifiers. Effect of the distortion spectrum on the timbre produced. (N.F. Circuit Laboratory, on behalf of the Japanese magazine *Radio Gijutsu*, June 1975, "55 new amplifiers tested.")

referring to the level of 400Hz (88dB in current terms, 80dB SL, 76dB SPL), you obtained these levels of harmonic orders: –22dB, –26dB, –30dB, –37dB, and distortion rates of 0.08% (H2), 0.05% (H3), 0.03% (H4), 0.014% (H5), respectively. For the higher orders, the levels obtained were extremely weak, approaching values of 0.0008% for the highest harmonics.

All of this clearly explains why the current amplifiers, as perfect as they seem, still produce too much distortion, and produce a form of distortion that is too unstable in terms of level, frequency, and the transitory variations of each, such that it does not go undetected by the ear.

Figure 12 graphs the ideal spectrum of distortion, which goes undetected by the ear. You should note that certain triode tubes—and certain directly-heated triode tube assemblies such as tube assemblies of 300 B, PX 4, TM 100, VT 52—enable you to obtain a spectrum remarkably close in appearance. If you think you have succeeded—by means of other techniques and assemblies, including those with transistors—that only remains valid on a static level. On a dynamic level, this spectrum changes and easily becomes unstable. As for the feedback, you must recognize that apart from providing several advantages (a broadening of the bandwidth, a reduction in the rate of harmonic distortion), its negative effects are manifold:

- dynamic instability (change in distortion characteristics with level and frequency)
- "hard" distortion
- "hard" saturation (peak of the sinusoid is "truncated" and not "rounded")
- decrease of phase margin (the presence of active charge of the loudspeaker)
- absorption of high harmonic orders

For this last defect, *Fig. 13*, graph A, shows the distortion spectrum of a directly-heated triode tube amplifier (PX 4, Marconi) without negative feedback. In graph B, the application of the feedback absorbs the harmonics of high order while reducing the distortion rate. This explains the subjective impression of various "deficiencies" on a musical level: the sound is less "sparkling," less "airy," there is less perceptible sound from the violin bow (a type of white noise), and so forth.

A recent experiment has been conclusive about this point. It consists of using a generator/synthesizer of very high purity (which is quite rare), capable of simultaneously supplying more than 20 sinusoids that can be locked in phase and in level. The system must

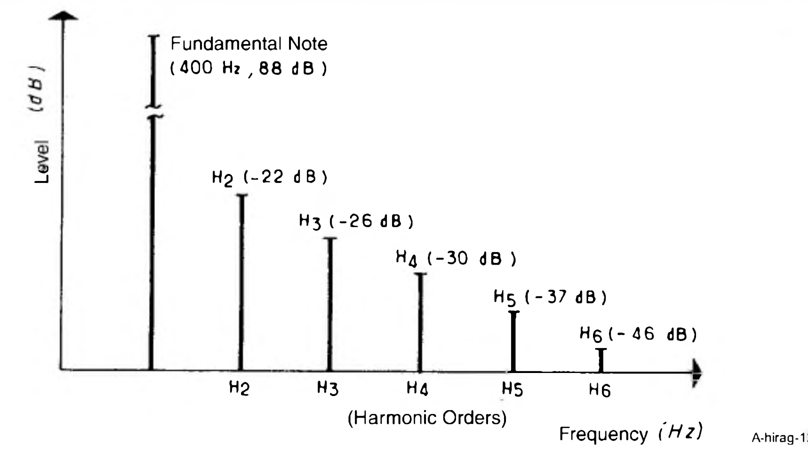


FIGURE 12: The characteristic of a pure sound debased by harmonics, giving the auditory illusion, by harmonic effect and by multiple masking effects, of hearing nothing other than a pure sound. This characteristic had been defined around 1930 by Wegel and Lane, and has, since 1960, been reexamined by Kuriyakawa, Kaméoka (Toshiba Laboratories, Japan). To be truly "transparent" on a subjective level, this characteristic must be not only static, but also dynamic; that is, it must be stable in terms of the frequency, output level, and charge—conditions that are quite difficult to obtain in practice.

also have an exceptional signal/noise ratio if you want to "imitate" a musical signal (most harmonics each represent only 0.01 to 0.5% of the total energy).

Figure 14 shows this signal which, once composed and locked in phase and in level, passes through a level-controller of a varying level (an electronic level-controller). This signal, sent to an amplifier and amplified by it, is then analyzed and put in memory. This method, complex and difficult to carry out, because it requires a particularly efficient generator/synthesizer, is the only one capable of proving that in almost every case a feedback circuit absorbs the harmonics of high order and with a very low level of amplified signal.

On a subjective level, this measurement proves that the thousands of connoisseurs of vacuum-tube amplifiers (the majority of Japanese connoisseurs) must not be wrong if they prefer simple circuits, without negative feedback, and triode output tubes in directly-heated triode tube assembly. *Figure 11* shows the essential causes of distortion or instability for certain distortion spectra.

CONCLUSION

It seems important to point out that turntables, player arms, and loudspeakers can easily cause defects that will have repercussions, not on the signal/noise ratio, but on the harmony of the sounds, on possible effects of dissonance. For these units, certain mechanical resonances of elevated amplitude have the tendency to "advance," to "shift" certain frequencies of the audio signal, exactly like a "sympathetic" resonance of piano strings. Certain turntable manufacturers have thus found it easy to prove in live demonstrations that a record (carefully selected) such as a slow-tempo piano and guitar duet could, on certain turntables (from rival manufacturers, of course), give the very clear impression of a "beating" during a piano/guitar chord, a beating of very low but clearly perceptible frequency. On the other hand, the turntable in question would deliver a chord that was "well-sustained," up until the sound's very extinction.

This brings us back to Fig. 7, where you see that some pure sounds debased by numerous harmonics become partic-

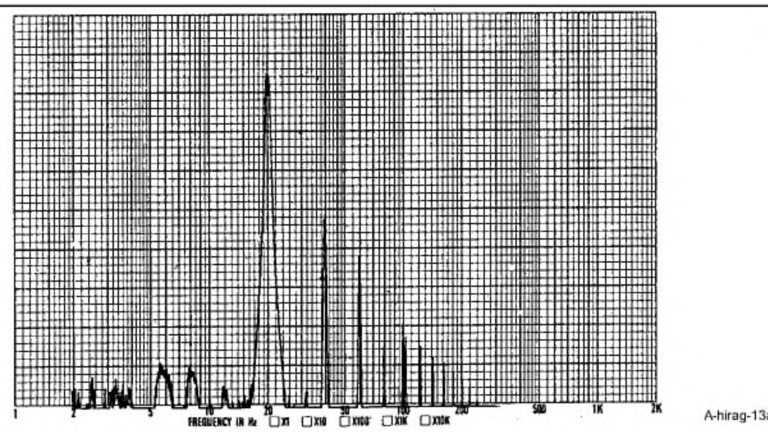


FIGURE 13(A): An amplifier with a single output stage, PX 4 tube (Marconi) producing 5W of power. The very simple circuit does not utilize negative feedback. The triode output tube, especially good, produces, in spite of a rather significant distortion rate (greater than 1%), a subjective impression of distortion that is virtually nil and, above all, a very high level of definition. Note the exceptional regularity of the harmonic spectrum, extending all the way to harmonic 13. This characteristic should not be confused with a distortion curve near the saturation point producing a high rate of odd-numbered harmonics.

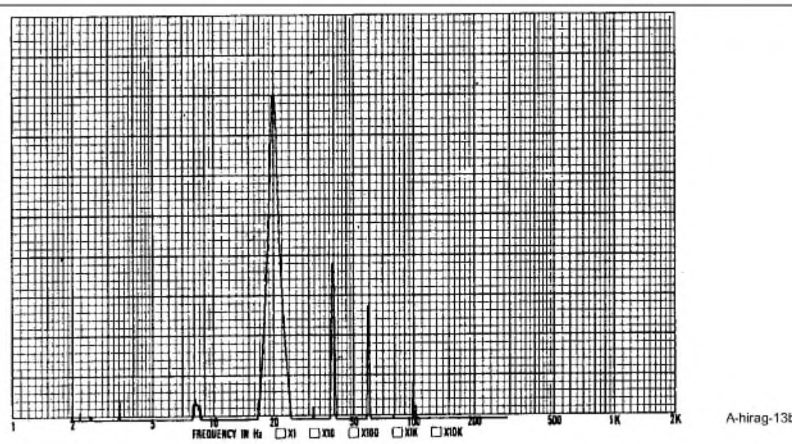


FIGURE 13(B): The same amplifier as in *Fig. 13(A)*, but to which has been applied a rate of negative feedback of 6dB, which is, in effect, very little. Although the distortion rate has decreased, and although the bandwidth is slightly widened, one notes an apparent "compression" of the sound, a lack of aeration, of refinement (in the high notes). Note that the harmonics of high order have disappeared, absorbed by the feedback loop.

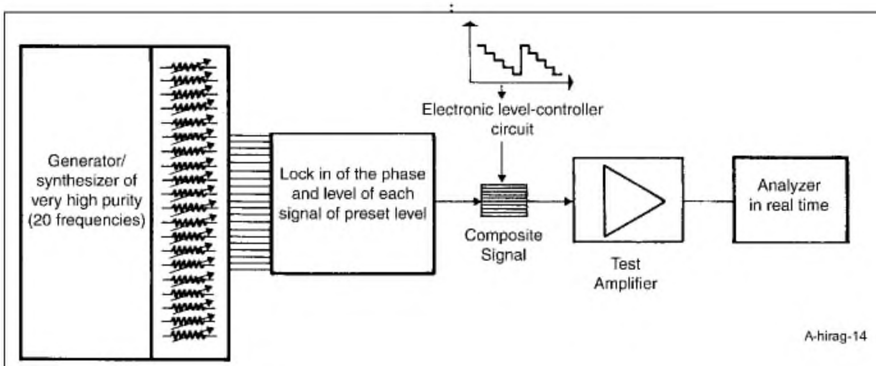


FIGURE 14: Measurement technique illustrating the behavior of an amplifier circuit receiving a complex audio signal, yet precisely predetermined (the composition of this signal). The analyzer enables a precise comparison of the level of each harmonic versus the original signal. This method of measurement, which requires the use of a generator/synthesizer of very high spectral purity, has permitted the observation that the majority of negative feedback loops has an unfortunate tendency to absorb the high harmonic orders, with a low level of applied audio signal.

ularly difficult to harmonize, and that even some very weak rates of distortion are, nonetheless, sufficient to cause the defects described in this article.

To conclude, you may think that if very substantial progress has been made in the audio field in the last years, the work that remains to be done seems even more difficult. It involves perfectly reproducing all sounds, including what is yet almost essential—that is, all the bits of micro-information that are necessary to respect the timbre of a musical instrument and to reproduce the ambience of a concert hall, the effect of space or direction, and all of the "emotion" that is contained in music. The characteristics of static and dynamic nonlinearity of the different components on one hand, and the degree of working stability of these diverse components on the other, are the main phenomena that hinder the perfect reproduction of sounds and of music.

In 1981, it remains virtually impossible to mass-produce violins with the sound quality of the best Stradivarius. And without doubt it is not only on account of snobbism that certain artists will not hesitate to pay a veritable fortune to acquire a beautiful instrument of this vintage.

Now the belief that it must be possible to perfectly reproduce the timbre of such instruments with the current sound-reproduction components is perhaps a fundamental mistake, a grand illusion. ❖

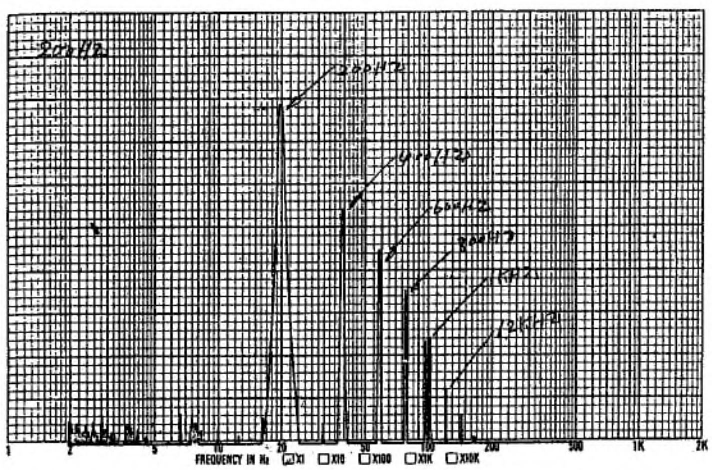


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FIGURE 15: A single-stage amplifier with a high-powered triode output tube, with high linearity, of reference 300 B. Note the exemplary distortion spectrum curve.

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Product Review

The Origin Live Turntable Upgrade Kit

By Edward T. Dell

When your Linn LP12 Turntable system develops a sluggish motion at turn-on and requires a push to get it up to speed, you know that some remedy is called for. A small ad in this periodical some time back caught my eye. I e-mailed the people at a small British company calling itself Origin Live (origin.live@virgin.net) and received a prompt answer in the form of a half-page of text about their kit for upgrading the Linn LP12 as well as Rega, Rock, and Systemdek turntables.

KIT CONTENTS

The kit includes five-and-one-half pages of instructions, including several clear color photos and a heavier sheet containing a cut-out stroboscopic disk for setting speeds (33⅓ and 45 rpm). It also includes a replacement motor and mounting plate plus screws, a small speed control board, a wired speed change switch and mounting plate, and a power supply (*Photo 1*). The Linn bottom cover and rubber feet have been removed, of course.

The standard power supply for the Origin upgrade is a nice black plug-in unit with convenient color-coded small polarized connectors which match leads from the circuit card. The supply plugs into any convenient outlet but does not have polarized prongs. The adapter is 2⅞″ W × 2½″ H × 4″ L and produces 300mA at 8V DC. I had ordered the "special optional power supply" for this kit (shown in *Photo 1),* which adds considerably to the cost, but since it was billed as "extremely advanced" and is said to offer "...improved transparency, dynamics and deeper bass," I thought the quality of the turntable justified the added expense.

Fortunately, Origin included the standard supply with the shipment, as well, since I had indicated the kit was to be the subject of a review. I say fortunately since the "special optional" supply promptly blew the wall socket circuit breaker immediately on plugging it in. Fortunately, no damage to the control circuit ensued. The standard power supply worked exceptionally well with no discernible noise or hum. When I remarked on this surprising failure, the company reported that my experience was unique, and no other failures have occurred.

DISMANTLING THE LINN

The Linn LP12 is beautifully made as befits this world-class beauty, which deserves its reputation. However, the motor and its extensive circuitry had simply ceased to do their proper job. In order to work on the unit conveniently and without damage to the tonearm or the platter, I C-clamped two plywood scraps to the wood turntable base, which held the unit in a vertical position for easy access. *Photo 2* shows the arrangement, which worked extremely well.

The original motor, shown at the upper left of *Photo 2,* is easily removed by loosening two bolts and the associated wedge-type washers which hold it in place. The wiring to the large circuit card is removed simply by unscrewing fixing screws in a terminal. The instructions suggest that the modifier make a record of the colors of the wires and the numbered terminals from which they are removed, just in case the user wishes to return the unit to its original assembly. I put all the parts removed into a plastic bag with notes on how to re-install the original parts.

The on/off switch, located in the upper right corner of *Photo 2*, is easily removed, as well. I disconnected the AC line cord to the Linn electronics and packed it away with the other parts. Origin Live suggests that you leave the Linn circuit card in place. When all the unnecessary parts are removed (*Photo 3*), the unit is ready for installation of the new motor and control board.

PHOTO 1: The Origin upgrade kit of parts, showing the special optional power supply, which comes with a bridge rectifier with flying leads, presumably to be mounted on the base wall. The turntable is in the background set up for modification, held by C-clamps and plywood scraps.

INSTALLATION

I turned the unit end-for-end so that the switch and motor mounting openings were nearest the worktable. *Photo 4* shows the speed/on-off switch in place at the left. The switch is assembled and already attached to the circuit card in the assembled version of the upgrade kit.

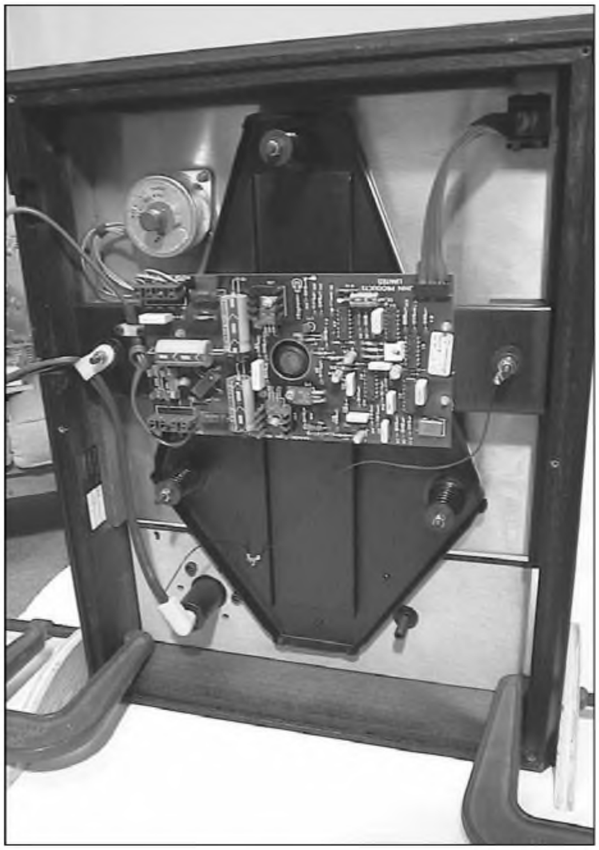


PHOTO 2: The turntable underchassis before modification.

Two brushed stainless-steel plates are provided to fit over the original hole in the main plate of the turntable. A standard lock washer and nut supplied with the switch are used to hold the plates and switch in place. The plastic spindle of the switch must be shortened to a proper length before mounting. The attractive black-anodized knob is then affixed with a small hex-wrench (supplied).

The motor comes mounted on a small plate with a round hole at one end and a slotted one at the other. The latter allows for belt-tension adjustment. The original mounting screws are used to install the new motor. The arrangement of washers, machine screws, and lock washers was the only unclear procedure in the supporting instructions. I finally managed to install the motor without too much difficulty.

The red and black wires from the controller board must be soldered to the two small lugs on the Dutch-manufactured DC motor. This is a bit tricky since the wires are a tight fit for the small lugs, and the plus and minus markings on the motor case are rather faint stampings into the metal.

PHOTO 3: The turntable after removal of the Linn motor and power switch.

Origin advises you to leave mounting the controller board to the very last step. The reason is the need to set the 10-turn speed control pots on the board. The unit must be unclamped and returned to its normal position, with the board outside the base, in a good light.

You attach the power line from the remote supply, carefully insulating the connectors with tape. The power wires are visible in the lower right of *Photo 4*, threaded out through the notch previously occupied by the Linn AC cord. A top view of the motor in place is shown in *Photo 5*.

ADJUSTMENTS

After the table is returned to its normal position, you re-install the outer platter and its pad, turn the unit on, and cut out the strobe disc. After that you spend a half-hour looking for a fluorescent lamp to check the proper strobe circle for speed adjustment. Four patterns cover the two speeds at 50 and 60Hz line frequencies.

The strobe instructions are very clear and helpful. You need a small screwdriver to adjust the pots. It is a good idea to let the motor run for an hour or two and re-check the settings from time to time. My unit seemed to settle down nicely and has remained accurately set to the two speeds.

Mounting the speed control board on the inner wall of the base is not easy. Only one hole is provided. I found a suitable wood screw and a small grommet to keep the board slightly off the side wall and carefully drilled a small pilot hole. The uneasiness in the process was the fear of damaging components, but all went well.

If the wires to the control had been somewhat longer, I might have mounted the board on the back outside the base, just to make any later speed adjustment easier. If adjustment proves to be needed often, I may replace or splice the switch wires to enable such a modification.

All in all I would pronounce the modification a very satisfying success, and the turntable performs just as it should, as good as new.

The kit is supplied by Origin Live, 87 Chessel Crescent, Bitterne, Southampton, SO19 4BT UK, 011 2380 442183, Fax 011 2380 398905, e-mail originlive@virgin.net. The kit may now be purchased only with an assembled regulator board for £222.98, with air post adding an additional £27.00, for two-week delivery. The special optional power supply is an added £148.94. In ordering, specify which turntable from their list you will be modifying. ❖

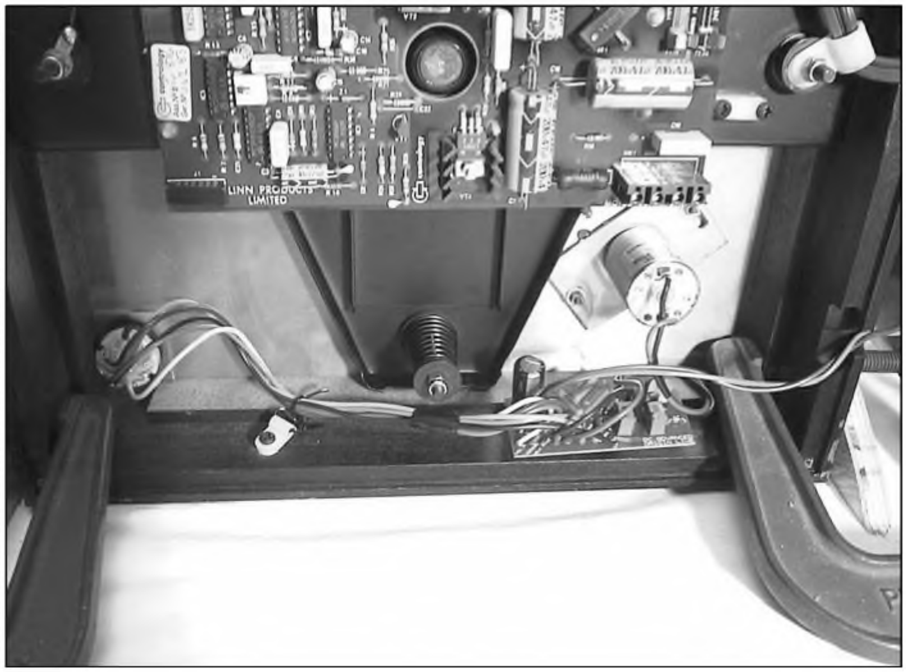


PHOTO 4: Completed modification showing speed/power switch at left, the speed control board, and the motor. The DC power wires are arranged to exit the base at the right.

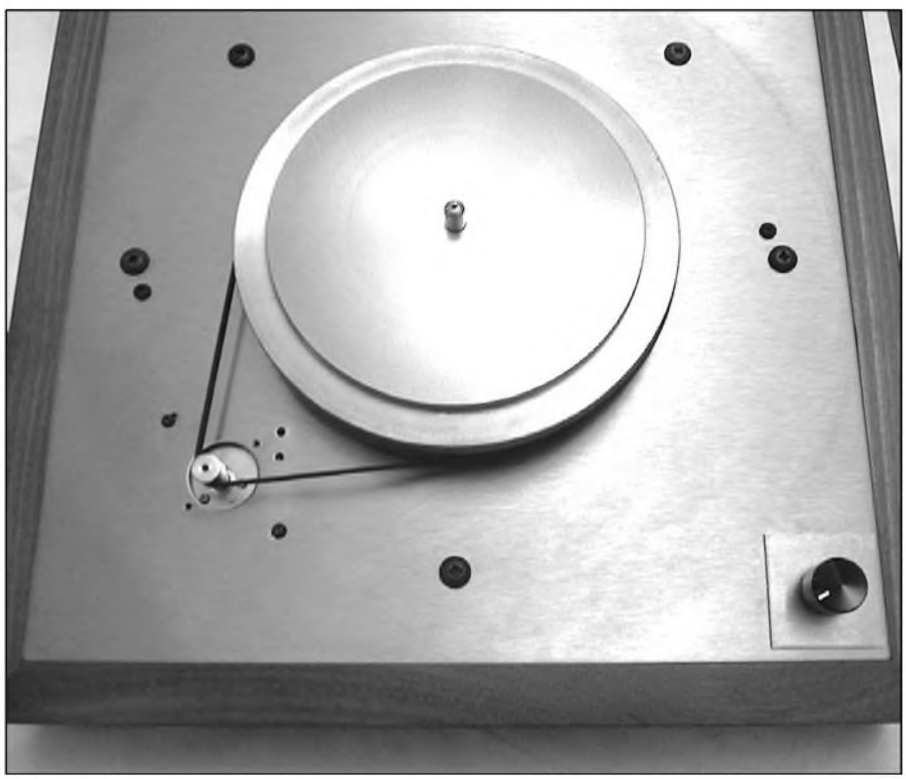


PHOTO 5: Top view of the turntable prior to replacing the main platter, with the new motor and control switch in place.

Book Review

Electronic Circuit Action Series—Amplifier Circuits

Reviewed by Charles Hansen

Electronic Circuit Action Series–Amplifier Circuits CD-ROM, Thomas M. Adams, Howard Sams, 2nd Edition 1966, 138 pdf pages. Reissued by Old Colony Sound Lab, PO Box 876, Peterborough, NH 03458, 603-924-9464, Fax 603-924-9467, E-mail: custserv@audioXpress.com ISBN, $16.95 US.

This particular title was just one in the *Electronic Circuit Action Series*. Capt. Adams (USN, retired) used a unique approach and analysis in describing vacuum tube circuit actions in his textbooks. He identified every electron at work in each circuit, then wrote a detailed discussion of the movement of the electron/current, as well as the significance of the electron flows.

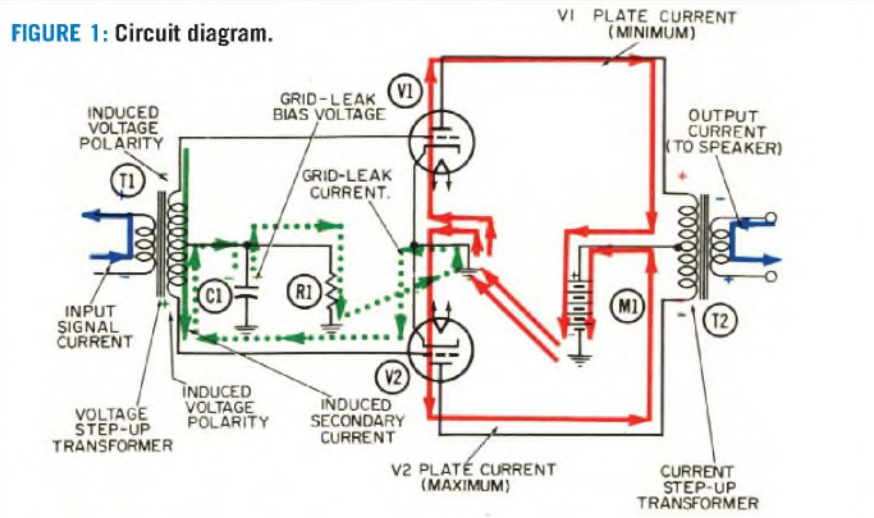
The circuit diagram (*Fig. 1*) treats electrons as "moving parts," whose multi-color visualization greatly enhances the understanding of the circuit operations. The author uses electron flow rather than conventional current, which he describes as "electrons which travel from cathode to plate become plate current." All waveforms are drawn in time domain.

The circuits are general in nature, and applicable to public address, radio, TV, record player, and tape recorder audio amplifiers of the day. The amplifier circuits in this CD-ROM reprint are classified into three groups:

Audio-frequency voltage amplifiers
Audio-frequency power amplifiers
Radio-frequency voltage amplifiers

The circuit descriptions include R-C, impedance, transformer coupling, and positive and negative feedback. Conventions of the day were preserved. Frequency is in cps (cycles per second) rather than hertz, and micro-micro farads (mmF) are used instead of picofarads (pF). Each chapter is followed, textbook fashion, by review questions.

FIGURE 1: Circuit diagram.



The textbook pages are in Adobe pdf format, and the Adobe Reader is included on the CD-ROM. Full pages are easier to read on larger monitor screens since the pages are vertical and computer monitors have a longer horizontal dimension. You can use the Adobe view, zoom and view, and fit modes to obtain the text size you desire. I found it best to print the table of contents and index pages so I could go directly to specific topics using Document, Go to Page.

CHAPTER 1—BASIC VACUUM-TUBE ACTIONS

This covers diode, triode, tetrode, and pentode tubes. The author gives plate curves for the 2X2A diode and the 6BN4 triode. This chapter covers the basic concepts of grid action, gain (μ), transconductance (g_m), and plate resistance (r_p).

CHAPTER 2—R-C COUPLED AF VOLTAGE AMPLIFIERS

Here we have circuit descriptions of R-C coupled audio amplifiers, electron "current" flows, the cathode circuit (with filter actions and time constants, a topic common to all the cathode circuit descriptions in each chapter), waveform analysis, frequency response (with LF and HF limitations), voltage and current feedback, amplitude distortion, and phase distortion.

CHAPTER 3—TRANSFORMER-COUPLED AF VOLTAGE AMPLIFIERS

The author describes transformer action, frequency response (again with LF and HF limitations), frequency distortion, and negative voltage feedback. Cathode filter action is included in the feedback section.

CHAPTER 4—AUDIO FREQUENCY POWER AMPLIFIERS

Transformer-coupled power amplifier circuits include both input and output transformers, beginning with a description of permanent magnet speaker operation. The single-ended Class-A amplifier stage is described first, followed by push-pull Class-AB amplifiers, with their "advantages" (second harmonic cancellation, smaller output transformer, smaller power supply filter caps). The push-pull circuits' descriptions evaluate "grid-leak bias" and cathode bias. The final section presents the Class-B push-pull circuit.

CHAPTER 5—AUDIO PHASE-INVERSION CIRCUITS

Since the push-pull amplifier requires a phase inverter, two types of tube phase inverter circuits are presented: the single-triode cathode follower (split-load phase splitter), and the two-triode inverter (cathode-coupled phase splitter). Both use electrolytic bypass capacitors around part or all of the cathode resistance. The latter circuit has its inverting tube's grid connected to the grid circuit of the output tubes rather than being returned to ground through a capacitor.

CHAPTER 6—IMPEDANCE-COUPLED RF VOLTAGE AMPLIFIERS

These last two chapters depart from audio amplification, and present RF pentode circuits. The purpose of this textbook was, after all, to discuss vacuum tube amplifiers in all forms. The sections of this chapter are untuned impedance-coupled amplifier, tuned impedance-coupled amplifier, RF alternating currents, filtering currents (screen-plate positive feedback and oscillation are covered here), and unidirectional currents. The latter refers to the plate and screen current in a tank oscillator circuit, and "current," as always, refers to electron flow rather than conventional current.

CHAPTER 7—TRANSFORMER-COUPLED RF VOLTAGE AMPLIFIERS

The sections of this chapter are untuned-primary-tuned-secondary coupling, tuned-primary-tuned-secondary coupling, the bandpass amplifier (this circuit is a tuned circuit with damping resistors to lower the Q, and thus spread the bandwidth out), and a discussion of series versus parallel impedance in tank circuits.

The two brief pages of the index are not really comprehensive but, in conjunction with the table of contents, will get you to any topic on the CD-ROM.

CONCLUSION

I really enjoy historical electronic textbooks. I found this one to be a fascinating and technically valuable book, with some unusual vacuum tube circuits. The CD-ROM format is not as convenient as a paper book, but it was the only practical way to preserve and offer this book to you. ❖

Xpress Mail

SUPERFLUOUS TUBES

The articles that Bill Chater has written for your magazines over the years have intrigued me, although I haven't always cottoned to his overall circuit approaches. My preferences for audio circuits lean in a different direction—toward pure tube topologies rather than solid-state or hybrid circuits. I have, nevertheless, benefited from reading Mr. Chater's prolific and imaginative circuit ideas, and have admired the high levels of diligence he must possess to make these complex topologies actually work! Having just read his most recent article in the Oct. '01 issue of *audioXpress* ("The Trouble with Screen Grids," p. 24), I have a couple of observations and questions.

The major point of his article was to disclose a technique for removing the deleterious effects of screen current variations in output beam tubes, with the goal of reducing distortion—a result he effectively proved. By adding back the screen current to the cathode using an op-amp-based feedback circuit, plate current is made more nearly constant, or at least is more tightly slaved to the current demands of the driven MOSFET at the bottom of this cascode.

I designed a similar circuit for a different application a while back. I needed a high-voltage DC current source to pull current out of another tube's plate circuit, without materially reducing the impedance of that node. While a pentode or beam tube will show relatively high plate impedance (which is the basic measure of a current source), as plate voltage swings low with signal variation, the screen's current will increase, stealing it from the plate.

This effect explains the downward slopes at the left of characteristic plate curves for a pentode. A beam tube aims to be better in this regard, but eventually its curves droop as well. The plate impedance, represented by the inverse slope of these curves, falls precipitously at these lower voltages.

In my application, I employed double current mirrors to "measure" the screen current and to add back to (or sink from) the cathode that same current value. I also connected a low voltage current reference to the cathode to complete the cascode circuit. As expected, the plate curves flattened all the way down to the "diode line."

The dynamic plate impedance was markedly increased, as desired for my application. In Mr. Chater's application for his audio output circuit, the effective plate impedance is bound to be enormous, not only because of the screen current compensation, but also because the beam tube sits on top of the drain of a MOSFET, effectively multiplying that device's large drain impedance through the cascode tube. What you have here is a push-pull pair of two extraordinarily high impedance current sources driving a 1239Ω transformer primary. This scenario seems to be a far-from-ideal impedance match.

The excellent Plitron PAT-4008 transformer was intended to be driven from low impedance triode tubes. In Mr. Chater's application—at least in the open-loop case—the usual bass rolloff ought to occur at a higher frequency than usual due to the high driving impedance interacting with the primary inductance, not to mention possible transformer distortion and high-frequency response effects. And the output impedance from the secondary to the speaker must be unusually high (implying a very low "damping factor").

Mr. Chater's circuit uses negative feedback, which will tend to improve these concerns, but by how much? Would Mr. Chater care to share any measurements he has made of the frequency response and output impedance, both in the open-loop and closed-loop cases?

Finally, given these concerns and Mr. Chater's willingness to use op-amps liberally, I wonder why he even bothered with output tubes. Why not replace the output stage with MOSFETs or bipolar power transistors, and skip the expensive output transformers, the tubes, and the high-voltage supplies? I accept the design for its thought-experiment value, but I must admit to wondering whether it makes sense for someone just wanting to build an amp that plays music beautifully.

I do not want my observations to be taken as an attack on Mr. Chater or his clever circuit design; rather my comments are made in the spirit of an open exchange of ideas about our shared hobby, where each of us can teach and learn from one another.

Brian Beck
Melbourne Beach, Fla.

Bill Chater responds:

Mr. Beck's cordial comments are appreciat-

ed, and I can assure him I am not offended by his attempts to share experiences.

The subject of output impedance has often been mentioned in connection with the cascode amplifier, as the essential current-source character of the cascode output stage is bound to produce this effect. It does this even with a triode output tube, let alone when using a beam-power tetrode.

In one of his letters to the editor (GA 1/98, page 39), Menno van der Veen made the observation that this circuit was a current-source. In response, in my article on Cascode Power Amplifier Circuits, Part I (GA 5/98), I included the idea of using a local feedback connection around the output stage, so as to control the output stage impedance. I described the conversion of the circuit from pentode to triode action in that article. The output stage then exhibited an effective plate resistance of 5700Ω as far as the output transformer was concerned. This makes the stage attain a more satisfactory match of tubes to OPT, if you choose to achieve that adjustment.

In Part II of the same article (GA 6/98), I supplied more complete design details of the cascode circuit. In that circuit, which used 6C33 output triodes, the output transformer was a Plitron FAT4128-00. This unit, which is a member of the Plitron family of OPTs, was designed especially for use with a current-source type of output stage. Mr. Van der Veen has told me that he deliberately adjusted the OPT winding parameters so as to make the transformer have essentially the same damping and bandwidth when driven from low, normal, or large output impedances. Such adjustments are possible, evidently, and tend to suggest that the toroid line, at least, can be made to accommodate these variations.

As to the gain and bandwidth of the design, as well as the output impedance, there is a plot of these characteristics in Part II (GA 6/98, page 48). This data applies to the circuit offered in Part II, which is not the same as that of Part I, because of the introduction of a discrete version of the driver stage. Still, the circuit bandwidths and gain are designed to provide identical results. Thus, you can apply the data shown in the Part II article to the Screen Grid article with small error.

In particular, there is a sidebar (GA 6/98, page 50) that describes the analysis for gain and output impedance for such an amplifier topology. With feedback, the amp has a damping factor of about 13 over the audio band. (I am sorry the figure 35 detail is difficult to read, and the scale of ohms is missing. The plot shows an essentially constant 0.6Ω Z_{out} versus frequency, which rises to 0.7Ω at 20Hz and at 50kHz.)

In Mr. Beck's letter, he suggests the tubes are actually almost superfluous. I quite agree! In fact, I made such an amp version, using larger power IGFETs in cascode with the driver units, instead of tubes.

This has some interesting advantages: The power supply voltage is more efficiently used, because the minimum tube drop of around 100V now can be developed across the OPT primary. This results in more output power. Another advantage is the removal of the tube's inaccuracies from the signal path, as well as its warm-up time. The arrangement does indeed make a perfectly acceptable amplifier design. So, why did I choose to publish the circuit with the tubes?

The answer is that there is a hidden flaw *in the all-semiconductor approach. This flaw announced itself with a meltdown in the output devices. It turned out that here was, unintentionally of course, a turn-on transient in the preamp connected to the amplifier's input. This particular preamp had the nasty turn-on habit of outputting a long time-constant charging event in its output wiring.*

The result was a long offset from ground in the DC level at the amp's input. This DC pulse got through the amp's input R-C with sufficient amplitude to cause the amp to drive an ever-rising current into one of its output transistors. The DC impedance of the OPT being only the winding resistance, there was no reasonable limit to the resulting current in the output transistor. The meltdown was quicker than the fuse.

After puzzling over this phenomenon, I decided the tube wasn't such a bad idea after all. It was good insurance against these meltdowns, and probably better than the effects of a servo riding herd on the input. Perhaps there is a reader who can suggest a safe solution to this problem, and breathe life back into the all-semi version?

CLASSIC DESIGNS

I thoroughly enjoyed John Stewart's article ("The Circlotron Amp") in Aug '01 *aX*. It brought back memories from those years, long ago, when I started to become interested in DIY audio. I have actually built a few of these amps, which at the time we called parallel push-pull (PPP) in contrast with the "normal" Series Push-Pull (SPP) configuration, for obvious reasons.

At that time, PPP was pushed (no pun intended) by Philips, which, in fact, designed a speaker to go with this amp—the 9710-series speaker. It was a wideband driver with a nominal impedance of 800Ω! (later they also produced a 400Ω version).

You can see in *Fig. 1* that the two DC currents through the Za loads are opposite and (if well designed) equal in magnitude.

Therefore, the two loads can be connected. The benefit of the 800Ω driver was that no output transformer was needed, only a center-tapped choke for the sole purpose of providing a ground reference. We used tiny little chokes—10 or 20H—for a DC current, I think, of some 50mA (*Fig. 2*). Very cheap.

In his article John refers to bootstrapping the driver stage so that it can provide the required high drive voltage, which is needed because each output half, in fact, receives negative feedback at its cathode of half the output voltage. However, the anode voltage to the driver tubes also is in series with half the output voltage.

This output voltage will divide itself over the driver tube's Ra and Ri, and the part over Ri contributes to the drive voltage to the output tubes. Isn't it a neat and elegant circuit?

To use this amp with a normal 8Ω driver, you would provide an "8Ω tap" at one side of the choke as shown. This part of the winding would need heavier gauge wire, of course, but it would be an auto-transformer rather than an isolating transformer. This would still make it easier and less expensive for the same performance. I still have some winding data for such an autotransformer for 2× EL84 and 2× EL34 if anyone is interested.

This configuration will also provide more output power for the same supply voltage, because there is no DC loss in the output transformer. In a SPP, each half of the output transformer carries the DC for one output tube, which decreases the supply at the anode. In the PPP, there is zero net DC through the choke and the full anode supply is available at the tube.

Jan Didden
diddenj@wxs.nl

John L. Stewart responds:

Thank you for your letter in regard to my article in audioXpress. *It is responses like yours that keep this activity alive and far from boring. There are still many challenges in vacuum tube design.*

I suspect from your comments that you are about the same age as I. Also by your e-mail address I was able to determine that you live in Holland. Many years ago while working in the development lab, we used many fine products developed in Holland by Philips. Their tubes and semiconductors were a great help in completing our various projects.

Like many at that time, we were on the *mailing list for the "Philigram," an excellent technical journal published by Philips. It kept us up to date with the latest new products and developments at Philips. I, too, recall reading about the Philips development of circuitry and loudspeakers that would allow a vacuum tube amplifier to drive a loudspeaker directly without the help of an output transformer.*

For some reason, I had always thought they had used the so-called "series push-pull" (Fig. 3), *or totem-pole arrangement developed by Futterman, perhaps to get around the patent on the Circlotron configuration. It would be interesting to hear from anyone out there who might have the real story on what happened.*

About the time these circuits and loudspeakers were showing up, another factor sealed their doom, at least for a time. The first successful germanium power transistors were on the market. In North America these were the complementary pair, 2N68 and 2N95, I believe. I recall Philips followed close behind with their OC series. I have some OC26s and OC28s still in their containers in my junk box. Are there any collectors out there?

Another problem for either version of these low-impedance tube circuits would be the difficulty of building a reliable matching high-impedance loudspeaker. Speaker building is difficult enough without having the complication of high impedance (for a speaker). For the Circlotron (parallel push-pull), the need for two completely independent and floating high-voltage power supplies didn't help much either.

I did try a version of the Futterman Amp (series push-pull) at that time. It successfully drove a 16Ω loudspeaker, although I don't recall how well. My testing of amplifiers in those days was nowhere as sophisticated as it is now. You can read about that circuit beginning on page 10 of Volume 3 of the Audio Anthologies, *available from Old Colony Sound Lab, PO Box 876, Peterborough, NH 03458, 888-924-9465, FAX 603-924-9467, e-mail custserv@audioXpress.com.*

It used 40dB of negative voltage feedback to get things under control. That was needed since the driving circuitry was not properly designed so that the top and bottom parts of the output stage would get the unequal drive required to get a symmetrical output.

As you can see in the SPP

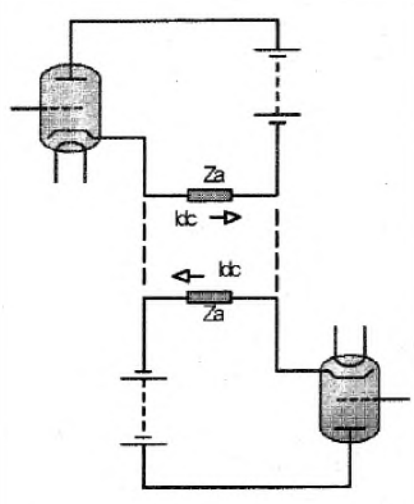


FIGURE 1: Parallel push-pull amp speaker configuration.

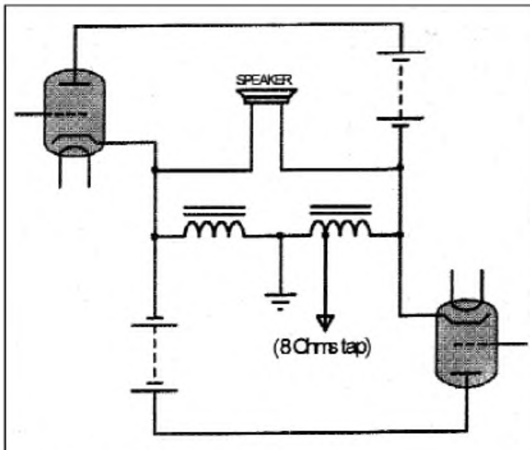


FIGURE 2: Speaker design.

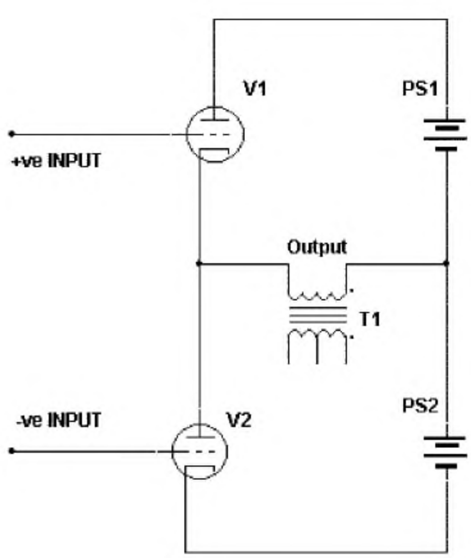


FIGURE 3: Series push-pull connection.

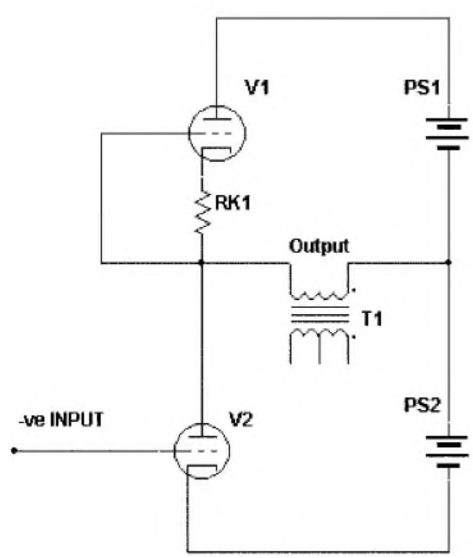


FIGURE 4: Series push-pull with single-ended input.

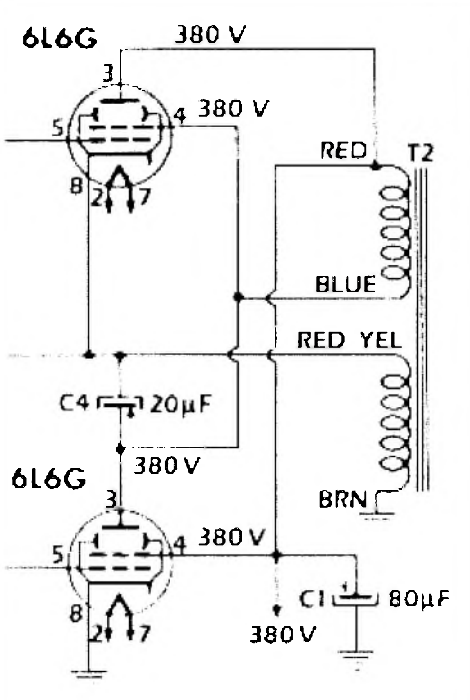


FIGURE 5: Single-ended push-pull amplifier circuit.

schematic, the top tube V1 is connected as a cathode follower and has no gain. The bottom tube V2 may have gain depending on its plate impedance in relation to the load. With equal drive you will usually get an asymmetric output.

One of the desirable properties of the Circlotron (PPP) is that the output stage is completely symmetrical. The Futterman (SPF) is not. That's what is called 20/20 hindsight. Everyone has some!

Another good reference to the Futterman Amp (SPF) is found beginning on page 60 of Audio Anthologies, *Volume 5. That article is by one of the all-time best audio gurus, Norman Crowhurst.*

For anyone who is interested in delving even further into the mysteries of the series push-pull amp, see the article by Peter Sulzer in Volume 2 of the Audio Anthologies. *It begins on page 25 and is an excellent overview of various output configurations (but he missed or ignored the Circlotron).*

I recommend you avoid the topology shown in Figure 2 (c) of that article. In that circuit only the lower grid is driven (refer to Fig. 4*). The upper grid depends on voltage developed across a resistor between the tubes. That way the upper tube manages to amplify any distortion produced by the lower tube as well as generate its own!*

Output balance and distortion depend very much on parts, and it still needs two high-voltage supplies. Aging will do terrible things to its performance. It looks and works similarly to the famous mu-follower. However, it is not recommended for a power output stage.

A better alternative is shown in Fig. 2 (b) of the same article. It was also briefly mentioned in Glass Audio *3/95, p. 54. I have taken the liberty of reproducing it here as* Fig. 5. *It is a very clever rearrangement of the various parts so that only one high-voltage power supply is needed. I did some preliminary work on that in 1995 but ran out of time. I intend to have another look at it soon. Most of what I will need is here and ready to go.*

Finally, around 1960 I did build a series push-pull amp to drive a synchronous motor (not a loudspeaker). It uses 20 (count them) 6AQ5s and needs plus and minus 300V power supplies. It has a bootstrap for the driver of the upper row of tubes. The same bootstrap is used to run the upper row of screen grids. That way the 6AQ5s run in the beam tetrode mode. The output looked quite good, but again I have no figures.

It was able to generate more than 100W with ease. It used negative current feedback from the load. That way the amp looked like a current source to the motor which formed the load. I could then control the motor speed by simply varying the frequency of the input to the amplifier. Since the motor impedance rose as the frequency was increased, it needed to have the applied voltage rise simultaneously. Hard to believe, but I still have that amp all covered in dust! I may get it running again!

As you have pointed out, another property of the parallel push-pull amp is that there is little or no DC current flowing in the output circuit so that you will avoid some power losses. However, if you were to use an output transformer here as many do, you would still have both copper and core losses in the transformer due to the audio current flow. You could have the output wound as an autotransformer, but as you have also pointed out, the loudspeaker part of winding would need to be of larger gauge wire.

My own preference is to have all of the leads brought out, which allows for more flexibility. Some regulating authorities require that an electrically isolated winding be provided for the loudspeaker. You would see this in some

PHOTO 1: UL-approved test leads from Circuit Specialists.

PHOTO 2: UL-listed multimeter costing $29.

public-address systems. You will also find it in the famous McIntosh amplifiers.

DMM TEST REPORT

In the October issue Gary Galo reviewed a $19 DMM ("Product Review," p. 78)! No mention of it meeting even the minimal UL listing requirements for safety. If the test leads and meter are not UL listed to Class 1, 2, or 3 environments, why on earth would anybody risk their life using a dangerous piece of test equipment?

Have you ever seen what a meter can do if it is not properly protected for all forms of misuse? Such as leaving the meter on the ohm scale and inadvertently reading, say, a line voltage? It happens. A quality piece of test equipment will save your life—such as products from FLUKE, Simpson, HP, Tektronics.

A $19 piece of test equipment is almost always not UL listed, nor would it meet the most minimum standards for safety. Recommending such a tool is a grave disservice.

Please read up on UL safety requirements for test equipment. I wouldn't be caught using a test tool that does not meet IEC and UL standards. It's your life; if it's worth only $19, so be it. Do some research on UL listing standards for test equipment. PLEASE! That article is enough to almost make me cancel my subscription.

Carl Engebretsen
Hamilton Square, N.J.

Gary Galo responds:

Mr. Engebretsen has assumed that just because a product is inexpensive, it is by definition unsafe. All test leads I have received from Circuit Specialists are UL listed. Photo 1 *shows the test leads supplied with four meters I have purchased from them, and all bear the same UL logo. The $29.95 CSI9903 multimeter is also UL listed.* Photo 2 *shows the rear panel of the meter.*

The $19 9300GP does not bear the UL designation, nor does their $29 Techmeter (which was not part of my review), so I e-mailed Circuit Specialists and inquired as to whether or not these meters meet UL safety requirements. Here is the response I received from John Ruff, General Manager of the company:

"In a general sense, they do. They are manufactured in the same factory with the same basic safety features as the CSI9903. However, UL is a private testing company. We are not able to say that they meet UL standards unless we submit the meters to UL for extensive testing. This is not practical for a low-cost DMM. (Note: The reason the CSI9903 was submitted is because it was originally designed and marketed by Sears in the Craftsman line. With Sears and the huge quantity they purchased, the manufacturer was able to submit the item for UL approval)."

Mr. Engebretsen is mistaken when he says that a $19 meter is almost always not UL listed, and can't meet basic safety requirements. Radio Shack sells a digital multimeter for $17.99 that is UL listed (#22-810), and MCM Electronics has a UL-listed DMM for $9.95 (#72-5070).

I have seen products bearing the UL logo that I would not use. Sears sells a UL-listed DMM for $19.95 (#82015). However, the banana plugs on the UL-listed test leads are not fully safety-shrouded—about ½" of bare banana plug is exposed on both test leads. This is true on all of the in-store samples I looked at, and can also be seen by enlarging the image of this meter on the Sears web site (www.sears.com; search for "digital multimeter"). What if either test lead is accidentally unplugged from the meter as you are about to read the AC line voltage? How did these test leads get a UL safety rating?

If I were to purchase this meter, I would not use it without replacing the test leads. Radio Shack sells test leads with fully-shrouded banana plugs for $4.99 (#278-708). These test leads are not UL-listed but, in my opinion, they are safer than the Sears leads bearing the UL logo. All of the test leads on the Circuit Specialists meters have full safety shrouding on the banana plugs.

I'm sure that Mr. Engebretsen is well aware that static discharge can damage sensitive digital circuitry. This does not mean that the product is unsafe. Drawing a spark when connecting the negative test lead (which is what happened with one of my 9300GP meters) probably won't blow the meter's fuse, but it was certainly enough to render mine nonfunctional.

A friend of mine is a technician—and an outstanding one, at that—for a local audio/video store. He tells me that he frequently repairs name-brand, major-manufacturer, UL-listed televisions that have failed due to a lightning strike. He says it is common to find such sets with destroyed transistors in the switching power supply, but with the line fuses intact!

Circuit Specialists has sold over 16,000 of the 9300GP meters, and I am not aware of any safety-related issues with this product. It is indeed sad for American jobs, but labor rates in China allow them to make dirt-cheap DMMs that are safe. I am satisfied that the 9300GP is safe. Nonetheless, if readers are uncomfortable with a product unless it bears the UL stamp, then I recommend spending another $11 for the CSI9903. As I stated in the review, the CSI9903 has to be the best value in test equipment from any supplier.

There is no question that test equipment from the name manufacturers mentioned by Mr. Engebretsen—and I would add Wavetek, Triplett, and B+K Precision to his list—offer superior construction and long-term durability for the professional user. Many electronics suppliers also offer more expensive "house brand" test equipment that competes favorably with the higher-priced national brands in this regard. The higher-priced offerings in MCM Electronics' Tenma line are a case in point (also sold by Newark Electronics). But, there is now affordable—and safe—test equipment within the reach of electronics hobbyists, and that is why I reviewed these products.

SERIES CROSSOVERS

I am very complimented by your reference ("Xpress Mail," Aug. '01, p. 76) to the letter that mentions my name in the May 2001 issue of *audioXpress* (p. 74). However, the real credit for the letter goes to G.R. Koonce. I was merely a conduit, passing on to the readership the wisdom that G.R. had been so kind to share with me.

I very much appreciate your further thoughts on the subject as well. I am very much fascinated by the series crossover network. As it turns out, not too long ago, I was reflecting on how all the schematics and writings I've ever seen on the matter of series crossover networks never included discussions of Zobels and other response-shaping networks that could be added to series networks. As you know, these networks abound in parallel crossovers.

I have been modifying a pair of speakers for a friend and decided that I would use series networks for the crossovers and incorporate Zobels to see what, if any, advantage they would offer. You can't imagine my joy at seeing that someone of your caliber had considered, and actually already implemented, the same. I believe I am in very good company, indeed.

I wish that you had depicted in schematic form the "cascaded series first-order sections" you refer to in your letter. Would you kindly submit to *audioXpress* schematics of the same for 2-way and 3-way systems? I would love to see what these look like, and I'm certain the readership would too!

Angel Luis Rivera
alrivera@tampabay.rr.com

Victor Staggs responds:

Unfortunately, it is no longer practical for me to provide readers with my series crossover network design using cascaded first-order sections. The woofer was the CTS 12W54C made by CTS in Paducah, Ky. It is long out of manufacture, and I don't believe that there is a comparable replacement. It was a large-magnet heavy-cone design resembling a 12" JBL woofer in its functioning.

The filler driver was a Pioneer 5¼" woofer-midrange, now long out of manufacture (I designed this system in the middle 1970s). The 2" dome midrange was the Philips AD0211/SQ8, also long out of manufacture, used as an example in my article about measuring and modeling loudspeaker voice-coil impedance[1]. Finally, the tweeter was the venerable Electro-Voice T35B horn tweeter, used as an anonymous example by Richard Heyser in his monumental AES paper about measuring loudspeaker phase[2], and now fetching upwards of $250 for a good used 8Ω pair on eBay, exclusive of the dividing networks.

I would need to start all over again assembling the drivers for a more modern rendition of the design. It took a couple of years to produce a finalized version of the first pair, and I don't have that kind of spare time now for making a new version of this concept. So, I must leave it up to the imagination of the readers to proceed on their own if they wish. ❖

REFERENCES

1) Victor Staggs, "Exploring Loudspeaker Impedance," *Speaker Builder* 5/94.

2) Richard C. Heyser, "Loudspeaker Phase Characteristics and Time Delay Distortion: Part 1," JAES, January 1969, Figure 8; also in "Loudspeakers—An anthology of articles on loudspeakers from the pages of the *Journal* of the Audio Engineering Society—Vol. 1–Vol. 25 (1953–1977)."

Audio Aid

The Self-Balancing Phase Inverter

I have an Eico HF-86 14W stereo power amplifier that utilizes the common split-load phase-inverter circuit to drive the push-pull output stage. Refer to the schematic of this circuit in *GA* 4/99 (p. 68). The first stage of the 12DW7 dual triode tube is the voltage-amplifier, which drives the splitter where the bi-phase outputs are taken from the plate and cathode.*

I was not happy with the phase-splitter circuit for two major, critical reasons: The output impedances for the two signals are dramatically different.[2] And the frequency responses of the two outputs are not the same, either.[3]

In an ideal phase-splitter the two outputs would be better balanced in impedance and frequency response if they were both taken from plate circuits. This is where some kind of a cathode-coupled circuit with the combination common-cathode stage and grounded-grid stage would be best. But this topology has the disadvantage of one-half the gain, and I did not wish to add another tube stage to compensate. My goal for this project was a circuit that features the same gain as the original voltage-amplifier with the same negative-feedback input arrangement, but with the two outputs taken from two plates, like a cathode-coupled circuit.

SELF-BALANCING SOLUTION

I came up with a variation of the "floating paraphase" inverter circuit, sometimes also called a "self-balancing paraphase" inverter.[4] This circuit (*Fig. 1*) allows the original style for the input of the negative-feedback signal to remain, and also retains the same plate and cathode-biasing scheme.

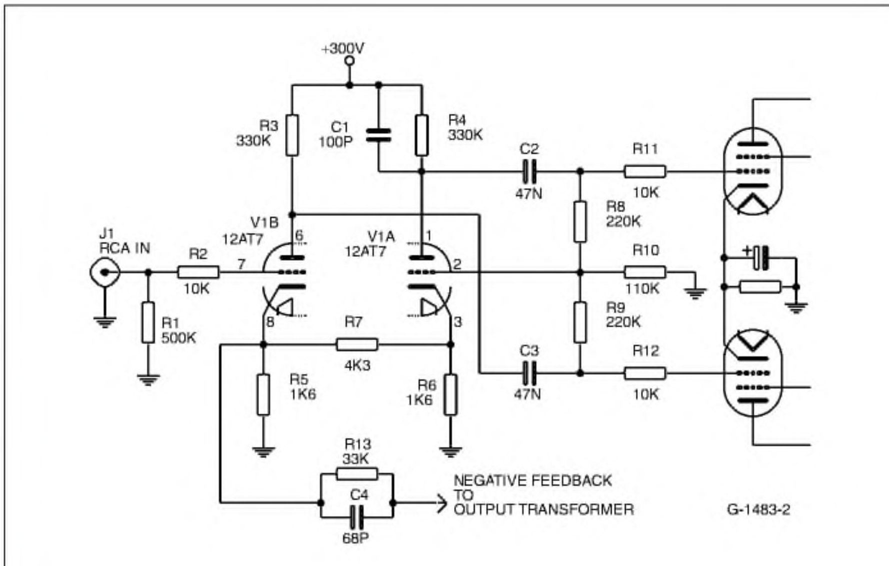


FIGURE 1: The new self-balancing inverter has better high-end performance characteristics than the old split-load inverter.

I finally settled on using a 12AT7 for this design because it matched the original gain and output power and had the lowest distortion at the speaker output. Using the old 12DW7 was out of the question due to the two different mu sections. I tried 12AU7, but the gain was less than desired, and then I used a 12AX7, but found that although the gain was good, there was intolerable instability and distortion.

The outputs of both sections of this paraphase inverter should be quite balanced as far as impedance and frequency response, and the two output voltages are very close. This topology is a self-balancing type, which means that any difference in the gain of the individual triodes is effectively neutralized so that equal output voltages are produced, as long as R8 and R9 do not vary. This feature comes in handy if the tube changes with age, or if you replace the tube with others of different gains.[5] The resistors I used were ½W ±2% metal films, the two coupling capacitors were 600WV DC Mylars, and the by-passes were mica because they were easily available, but any of those popular "poly-plastic" caps would be fine.

The secret of circuit operation involves the just-mentioned voltage-divider resistors, R8, R9, and R10, which also double as the grid-bias resistors for the power output tubes. For the circuit's exact operation, see reference 3. The selection process for the three resistors was arbitrary because I could not find any guidelines governing value selection. I also did not experiment with R8, R9, and R10 for optimal use, because from the circuits I examined, the values did not appear to be very critical.[6]

* The 12DW7 was a tube made especially for Eico; one section "is a medium-mu triode of the 12AX7 type and a low-mu triode of the 12AU7 type." Eico said, "the 12DW7 conveniently provides significantly better performance characteristics in the familiar voltage amplifier and split-load phase inverter of the Williamson-type circuit."[1] Although the manual did not specify, I assume the gain stage is the medium-mu and the splitter is the low-mu section.

I have seen some designs that use separate voltage-divider resistors and grid-bias resistors, but I figured I could make it more efficient by combining their operation. My only goal here was to make the grid-bias resistance to ground for each side the same as the original circuit (R8 + R10 and R9 + R10 should equal 330kW).

I added a unique feature to this circuit, positive feedback.[7] The resistor, R7, between the two cathodes provides a small amount of positive feedback that increases the gain of the triodes. This allows R8 to equal R9 for equal output voltages.

The feedback resistor is actually optional. If it is not used, then you will still obtain excellent results, but the value of R9 would need to be around 20% less than R8. One final note, if the value of R7 is too small, then oscillation will occur, so, some fine-tuning of its value will be required.

TESTING

With the new circuit and amplifier up and running, I applied a square wave and noted some minor anomalies. With a 1kHz input, the 4W output, with a 4W load, looked just as good and square as did the original circuit output. However, there were some spikes on the rising edges of the output square wave and some ringing at the beginning of the leading edge. But I cleaned this up by adding small capacitors across the plate resistor of the second triode. This fix also cured similar problems on the output signal at 20kHz.

The output at 20kHz looked better than before, due to the above-mentioned capacitor, with squarer corners, less ripple on the tops and bottoms, no spikes or ringing, and less slewing on the rising and falling edges. The phase-shift of the output compared to the input was also a little better.

The output at 100Hz, however, showed a severe case of deficient low-frequency response on one of the channels. After troubleshooting, I discovered an output tube was bad, which I confirmed on a tube tester. A new tube fixed the problem.

Another problem I found with the other channel was a distorted output on the 8Ω tap. It showed a small stair-step on the square wave half way up and down the rising and falling edges. I suspect a shorted turn in the transformer secondary, but I'll just have to live with it for now. The amplifier has 4, 8, 16, and 32Ω taps on the transformer.

The test on the other taps, properly loaded, looked fine with no steps, and the other channel looked fine on all taps, too. The funny thing about these two problems is that I never would have found out about them unless I had done this frequency response test, because with music, everything sounded normal.

Sine-wave testing showed perfectly beautiful sine waves on the output at all frequencies. Unfortunately, I do not have an harmonic distortion meter, so I cannot supply any THD data. Listening tests validated my bench testing. The power output was very close as before, and not only did the bass sound good, but the highs were a tad more crisp and clean.

Although this modification showed characteristics in the low- and medium-frequency response similar to the original split-load circuit, it demonstrated a quantum improvement in the high end frequency response. I am very pleased with this project because with just a few minor changes to the split-load circuit, the self-balancing paraphase circuit turned out to have much better performance. ❖

Michael Kornacker
Palatine, IL

REFERENCES

1. *Eico HF-86 Instruction Manual*. Electronic Instrument Co., Inc., Long Island, NY.

2. Bruce Rozenblit, *Beginner's Guide to Tube Audio Design*, Audio Amateur Press, Peterborough, NH 1997, p. 35.

3. Daniel P. Norman, "Understanding Tube Phase Inverters," *GA* 1/90, p. 14.

4. *Amplifier Handbook*, Richard F. Shea, Ed, McGraw-Hill Book Co., NY., 1966, p. 17–73.

5. Ibid. p. 42.

6. Gray, Truman S., *Applied Electronics*, 2nd edition, MIT Press, Massachusetts Institute of Technology, Cambridge, MA 1965, p. 532.

7. *Radiotron Designer's Handbook*, F. Langford-Smith, Ed., Wireless Press, RCA 4th Ed., 1953, p. 352–355.

Big Mike & the Jimmy
from page 27

the caps, rather than the active devices, so it's important that the composite impedance be smooth, with no peaks.

A pair of resistors in series (R217-R218) forms the current-limiting resistor for the plate regulators' pilot LED; the junction between them is also the float point for the second filament supply. R219 feeds the 47V zener diode that regulates the phantom supply. It's not a particularly elegant regulator, but I've decided that a fancier one is superfluous, as every phantom-powered microphone in my experience re-regulates its supply anyway, using a zener.

R201, R204, R207, R210, R211, and R219 dissipate a good deal of heat, as do the various zener diodes. It's a good idea to mount them about ½" above the circuit board, giving air a chance to circulate and cool them. Q202 and Q204 are mounted on a common heatsink, and remain relatively cool.

Figures 11 and *12* show the circuit board design and parts layout for the raw supply. *Figures 13* and *14* show design and layout for the regulator board.

It's worth noting here that this raw supply and regulator form a useful general-purpose power supply for high-quality tubed equipment; you get two filament supplies (one floating) and two regulated plate supplies, plus a supplemental zener-regulated supply for less-critical circuits. With a little modification, you could use this supply to drive most tubed preamps, line amp or signal processing circuits, or the input and driver stages of a power amp.

Next month I'll look at construction and possible configurations of the final project. ❖

REFERENCES

5. Paul Miller, "Resonances and Repercussions," *Hi-Fi News & Record Review* 34:6 (June 1989), p. 35.

6. Ben Duncan, "A State-of-the-Art Preamplifier: AMP-02, Part 2," *Hi-Fi News & Record Review* 34:11 (Nov. 1989), p. 45.

7. Rick Miller, "Measured RFI Differences Between Rectifier Diodes in Simple Capacitor-Input Power Supplies," *The Audio Amateur* 1/94, p. 26.

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I do most of my listening on three audio systems I have set up at this time. One is my 6550 single-ended amp directly driven by a Dynaco CDV-1 vacuum tube CD player. The second has my "little amp," a 12V6 push-pull, directly driven by a McCormack CD transport SST-1 with a McCormack DAC-1 converter. The third, my favorite, is equipped with a Mac 70 project amp controlled by a VTL PR-1 preamp with analog inputs from a Linn turntable and a very POOGEd Magnavox CD player.

All three systems have their own special sonic signature, and each has its own unique way of dealing with a particular recording. The results are always very enjoyable, and the following recordings are the ones that I use to evaluate the performance of these or any audio system I may be building or modifying.

1. Charlotte Church, *Enchantment*, Columbia CD CK897120, track #8, "From My First Moment."
This album contains some fine work by a young soprano who seems to possess the voice of an angel. I hope she's around for a long time to bring us more beautiful music. This album is a fine recording with a medium-wide soundstage. Her vocal on this track will demand a lot from your midrange drivers.

It really sounds magical on a single-ended amp and will almost bring tears to your eyes. If a system can't convey the delicate beauty of this lady's voice, it is in need of some adjustments. This album is recorded, edited, and mixed by Bill Whittington, who did a fantastic job.

2. Jane Monheit, *Never Never Land*, N-Coded Music CD NC-4209-2, track #2, "Detour Ahead."
If you love smooth jazz, performed by a young lady with a great voice, this album is for you. This new jazz artist, combined with some experienced jazz musicians, will give you your money's worth and will also give you a great soundstage to test your system. The saxophone, bass, and piano are rendered in a superb fashion on this track and the microphone placement is just right. My hat comes off to engineer Tom Shick and mastering pro Gene Paul for the great job they did here. Gentlemen, and lady, take a bow!

3. Stan Getz, Joao Gilberto, *Getz/ Gilberto*, Verve CD 314521414-2, track #1, "The Girl From Ipanema."
Originally recorded in March 1963, this jazz album dominated the charts and was inducted into the Hall of Fame for Recordings by the National Academy of Recording Arts and Sciences. And rightly so! If there is a better studio recording of saxophone, guitar, and vocals, I would love to hear it. This reissue, mastered by Chris Herles, is among the best to date. The original analog sources were very carefully restored to obtain the finest clarity. This track is a perfect example of that restored clarity. This transfer to CD is so well done that it produces a great test for instrument placement and separation. Bravo!

4. Three Blind Mice, *The Misty Mood of Three Blind Mice*, TBM CD GSCD006, track #8, "Alone Together."
This track is one of the ultimate tests for the accurate string bass reproduction of an audio system. The great recording techniques of Mr. Yoshi Kannari are very obvious here as he captures the sound of the plucked bass strings with loads of air around them! As an audiophile and a jazz fan, he knows what the quality of the final sound should be. This one will give your woofers a real workout!

5. Dave Brubeck Quartet, *Dave Brubeck's Greatest Hits*, CBS LP 62910, track #1, "Take Five."
A wonderful recording of one of the most famous jazz quartets in history. Great for testing the accuracy of any LP playback system. The music itself is jazz perfection with some very unusual rhythms. What more can I say?

6. Royal Philharmonic Orchestra, Bizet and Tchaikovsky, Chesky LP CR7, track #2, "Andante (Adagio)."
This is a very good recording of the Royal Philharmonic in the Walthamstow Town Hall near London. You can almost see conductor Charles Munch in action as this group of musicians sets a very wide soundstage with the perfect execution of this piece. This selection is a very good test of the tracking accuracy of any tonearm and cartridge combo. Proper tracking will reveal the violins in all of their splendor. I still have a hard time believing that this album was actually recorded in April 1963!

7. Cincinnati Pops Orchestra, Grofe-Grand Canyon Suite, Telarc CD 8006, track #12, "Cloudburst."
Excellent full digital recording of the Cincinnati Pops. Hold on to your subwoofers on this track and the one just before it. And please heed the warnings contained in the album notes! This digital recording of real thunderstorms can devour any woofer which is overdriven and has a weak suspension system! Don't listen to this one with headphones! I like my Grado RS-1s (and my ears) too much to even try it.

As audiophiles, we expect a lot from our audio components. As audio hobbyists, we realize that part of what we hear depends on the effort we put into building and modifying our equipment. With even a small amount of hands-on time, we can give our preamps, amplifiers, and speakers a special and personal vocal signature. (Think POOGE.)

I thoroughly enjoy all of the time and effort I invest in my own audio systems, especially when it brings an even greater enjoyment to the music I am listening to. So, fellow audio hobbyists, rejoice in the knowledge that we are among the very few that can really have it our own way. ❖

Rick Spencer
Clovis, Calif.