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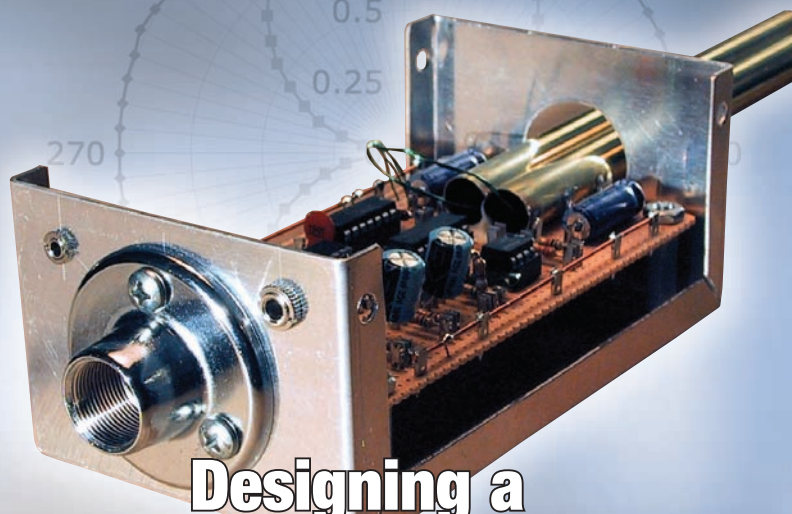
June 2009

US \$7.00/Canada \$10.00

HEADPHONE AMP UPGRADE

How to Achieve
A Better Preamp

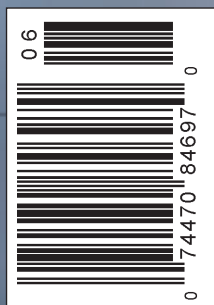
Uncovering a
Class D Amp Secret



Designing a
Studio-Quality Mike



Building a **BETTER**
Single-Ended Transformer



New from **Stanton**, the T.92 USB turntable features a high-torque drive motor, an S-shaped tonearm, and built-in DSP with key lock. Another newly upgraded DJ deck, the T.55, includes

a straight tonearm. Both are built on a rugged and sturdy base, and come



equipped with a slipmat, dustcover, and Stanton's 500.V3 cartridge. For more information, go to www.stantondj.com.

For its third annual meeting, the **Lone Star Audio Fest 2009** will take place in Dallas, Tex. at the Embassy Suites Dallas Park Central June 19-21. There is no registration cost for exhibitors, and no admission price, with the only participant prices being to the hotel. To make reservations by phone call 972-234-3300 and tell them you are coming to the Lone Star Audio Fest,

or go to <http://embassysuites.hilton.com/en/es/groups/personalized/DALTXES-LSA-20090616/index.jhtml> to register online.

The AMP-15, from **TERK Technologies** (www.terk.com), is a high-performance 15dB amplifier designed to improve the signal strength needed for broadband cable and antenna reception, and guarantees quality signal levels for longer cable runs. Supporting two-way communication required for broadband cable devices, the AMP-15 includes a power adapter and RG-59 connection cable.

TDL Technology's new Model 603A Tuneable Stereo Rumble Filter removes low-frequency noise from the turntable and records without adding any audible difference in the music. Operating from a 24V DC "wall wart," the 603A is tunable in 5Hz steps from 15 to 65Hz, has input resistance of 47k Ω , and

features two two-pole Butterworth high-pass filters (-12dB/octave).

For more, please go to www.tdl-tech.com.



The Audio Engineering Society (AES) 36th International Conference, "Automotive Audio—Sound in Motion" will take place from June 2-4, at the Hyatt Regency in Dearborn, Mich. The program will cover the latest technology and techniques used in the industry such as active noise cancellation, loudspeakers, voice recognition, and more. This presents an opportunity for the automotive audio community to share experiences and define in-car audio for the next generation. Those interested in sponsorship or exhibiting should contact 36th_sponsorship@aes.org.

Phiaton (www.phiaton.com) has unveiled the PS 300 NC noise-canceling

D42-1

DIY Kits

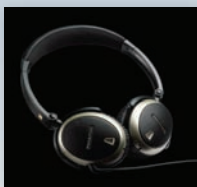
2 way design

--- w passive radiators

Message in a Bottle

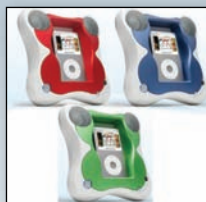
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headphones, which play even if the noise canceling battery dies and features a mute button. Wearers of the PS 300 feel the ear pads contract, while pressure is exerted on the front of the speakers. Also featured with these headphones is an adaptive acoustic impedance control, 40mm electrodynamic drivers, 98dB sensitivity, and a maximum input power of 1000mW.



Designed to offer a quick, easy way for users to power up their circuits, SchmartModules are six single voltage regulator devices from **SchmartBoard**. The company's prototyping system is used to hand solder components onto SchmartBoards, and these new power modules help ease the process even further. These Regulated Power Modules come in 1.5, 1.8, 2.5, 3.5, 5, and 9V, and suggested retail price is \$15. For more information, go to www.schmartboard.com.

The Butterfly, from **Vestallife**, is an iPod speaker dock which includes three interchangeable colored faceplates. It features 3.5mm Aux input, LED volume indicators, and dock connectors for any iPod sizes. To learn more, please visit www.vesta-life.com.



Paradigm Electronics Inc. introduces the SUB 12, a 12" driver which includes mineral-filled co-polymer polypropylene cones with Resonance Control Ribs, dual voice-coil designs, and proprietary Airflow Ventilation System cooling. The SUB 15 is the 15" model with all these features; both also include switched mode power supplies, high-power transformers, noise suppression networks, glass/epoxy circuit boards, and advanced short-circuit protection. Low frequency extension in the SUB 15 is 12Hz, while the SUB 12 has 16Hz. For more, go to www.paradigm.com.

The DP-200USB, from **Denon Electronics**, is a fully automatic turntable that allows users to easily convert LPs to modern MP3 digital audio files. Featuring a see-through flip-up cover and built from high-quality aluminum, the DP-200USB includes an internal MP3 encoder and a front panel USB port. The PC software contains Audio Waveform Recognition as well as Auto Track Divide. To find out more, visit www.dm-holdings.com.

The Sonus is a pocket-sized sound level meter range which includes a combined sound- and dose-meter sporting dual-frequency measurement, data-logging to a sizeable memory, and measurements to support noise at work and environmental applications.



These new meters feature three Class 1 and three Class 2 instruments to the IEC61672 Standard, and two dosimeters. Measurement range is 30-140dB, and parameters include Leq, Lmax, LCpeak, and Lepd. For more, visit www.scantekinc.com. *ax*



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Beyond G4OEP: Improving a Flea-Powered Headphone Amp

This headphone amp upgrade adds four relay-selectable inputs, a numeric display, a phono-preamplifier stage, and an auxiliary output.



PHOTO 1: The G4OEP.

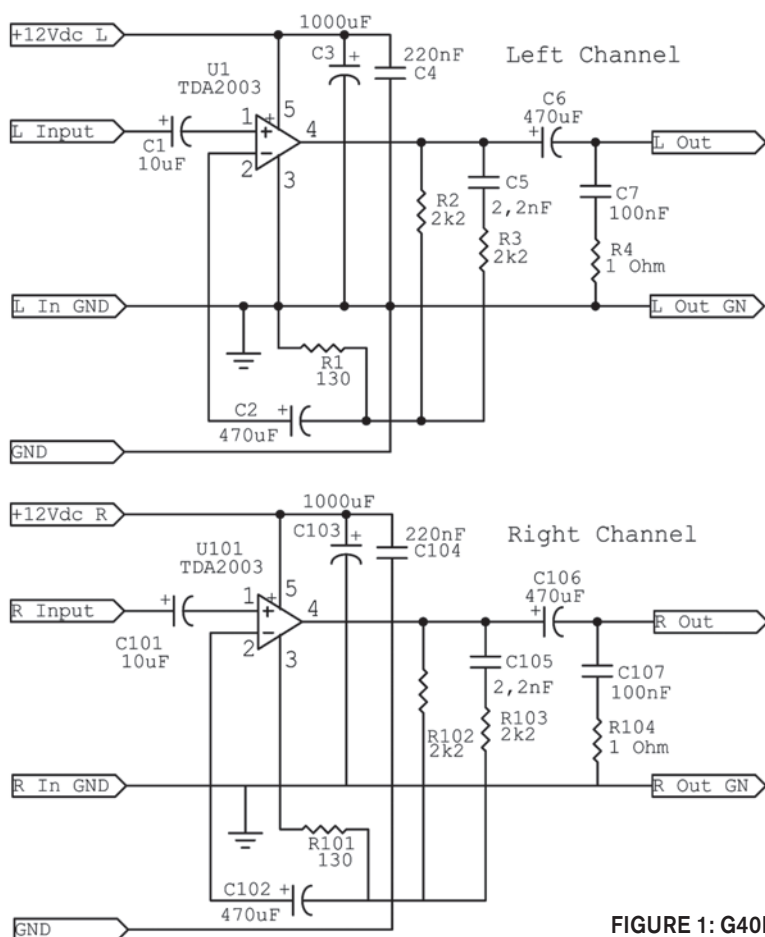


FIGURE 1: G4OEP.

First of all, I wish to thank Aren van Waarde (and Dr. Smith, a.k.a. G4OEP) for his article on the G4OEP (March '08, p. 36). Yes, I accepted Mr. van Waarde's challenge and built the headphone amp. After powering up the circuit, I was amazed what I heard!

This amp is so simple and clean—no magic—with a straightforward circuit (a little bit different from TDA2003 manufacturer suggested topology) that works very well.

I must confess that the circuit did not hold my attention at first look. After two weeks, I decided to search for the parts in my junk boxes and put everything together. The components are easy to find, with neither one “white-fly” nor esoteric “audio grade” component. In a few hours of heat and some smoke, the circuit was working very well, to my surprise. In fact, this simple circuit surpassed my expectations.

So, I decided to add the following improvements:

- Multiple selectable inputs
- Phono (MC) preamplifier with RIAA passive network
- Volume control
- Visual indication to selected input (a LED seven-segment subminiature display)
- Auxiliary output (to feed my single-ended “A-class” 47 triode mode amplifier)
- A reasonably fine-looking cabinet

CIRCUIT FEATURES

You can analyze these improvements separately in five parts:

1. The G4OEP, a TDA2003-based low-gain amplifier (see the *aX* article for more details)

2. Power supplies: two separate single +12V DC to each G4OEP and one dual $\pm 12V$ DC for the phono stage
3. The phono stage itself
4. The MUX and digital section
5. The cabinet

G4OEP

This is the same circuit as in the *aX* article. But, due to the difficulty of finding R1/R101 (130Ω), I mounted a 270Ω in parallel (135Ω), which worked fine. Basically it's a TDA car audio power amplifier operating at low gain so the quiescent point leads the amp to work at "A-class" in a very linear region of its transference curve. See **Fig. 1** and **Table 1**.

POWER SUPPLIES

As I said before, I used three power supplies. Two of them are identical (**Table 2** and **Fig. 2**), and a symmetrical (positive and negative) one for the phono stage (**Table 3** and **Fig. 3**). These small transformers rated to 500mA are so simple, like all the circuitry discussed here.

For the positive ones I used transformers rated to 110V AC/15V AC, single secondary coil. For the phono stage power supply, I used an 110V AC/15+15V AC center-tapped. In the symmetrical power supply for the phono stage, all transformers are in parallel with the mains, S0 and F1 circuit.

PHONO STAGE

I have found many interesting circuits at the ESP website. While searching for an RIAA passive network preamp, I encountered Mr. Elliot's web pages, which included an interesting topology that I changed by substituting the original Burr-Brown chips to the common TL074 (Texas Instruments). My objective was to simplify the original cir-

TABLE 1: CIRCUIT PARTS LIST

Component	Attributes	Designation
$10\mu F \times 25V$	Polyester	C1, C101
$470\mu F \times 16V$	Electrolytic	C2, C102, C6, C106
$1000\mu F \times 16V$	Electrolytic	C3, C103
$220nF \times 100V$	Polyester	C4, C104
$2.2nF \times 100V$	Polyester	C5, C105
$100nF \times 100V$	Polyester	C7, C107
TDA2003	Power car audio chip	U1, U101
130Ω	Metal film $\frac{1}{4}W$	R1, R101
2k2	Metal film $\frac{1}{4}W$	R2, R102, R3, R103
1Ω	Metal film 1W	R4, R104

TABLE 2: POSITIVE POWER SUPPLY

Components	Attributes	Designation
$100nF \times 250V$	Polyester	C1-4, C6, C101-4, C106
$4700\mu F \times 25V$	Electrolytic	C5, C7-9, C105, C107-9
1N4007	Si rectifier diode	D1-8
7812	1A positive regulator	U1, U2
1A	Glass $5 \times 20mm$	F1
$5 \times 20mm$	Fuse-holder panel mount	F1
110/220V-0/15V $\times 7.5VA$	Single secondary	T1, T2
IEC male plug 3-pin	Panel mounting	V AC

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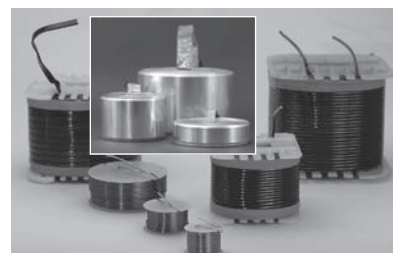
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TABLE 3: SYMMETRICAL POWER SUPPLY

Component	Attributes	Designation
100nF × 250V	Polyester	C1-4, C101-4
4700μF × 25V	Electrolytic	C5-7, C105-7
1N4007	Si rectifier diode	D1-4
78L12	400mW pos regulator	U1
79L12	400mW neg regulator	U2
110/220V 15/0/15V × 7.5VA	Center-tapped secondary	T1

TABLE 4: PHONO STAGE PARTS LIST

Components	Attributes	Designation
22μF × 16V	Electrolytic	C1, C3, C101, C103
22nF × 100V	Polyester	C2, C102
82nF × 100V	Polyester	C4, C104
1μF × 25V	Bipolarized	C5, C105
47k	1% metal film ¼W	R1, R101
2k2	1% metal film ¼W	R2, R102
4k7	1% metal film ¼W	R3, R103
180k	1% metal film ¼W	R4, R104
10k	1% metal film ¼W	R5, R105
2k7	1% metal film ¼W	R6, R106
100k	1% metal film ¼W	R7, R9, R107, R109
820Ω	1% metal film ¼W	R8, R108
TL074	Quad-operational amp	U1

cuitry that uses four single audio grade op amps from TL family (quad in a 14 DIP package).

The result was very good: No PCB layout, everything was mounted in a perfboard (pad pattern, instead of track pattern) with ground plane provided by aluminum foil (used for cooking and glued with glue-stick inside the cabinet and correct grounded in a single point). The schematics I used are in **Fig. 4**. See also **Table 4**.

At Mr. Elliot's website (<http://sound.westhost.com/project06.htm>), you can find more information about RIAA accuracy and other significant parameters. If you want to improve the phono pre-amplifier according to your needs, drop a note to *aX* and let us know what you did. The phono circuit has deep clean bass and affordable mids (voice is very clean), the highs are very honest, too. You will soon note the huge differences of recordings among your vinyl records.

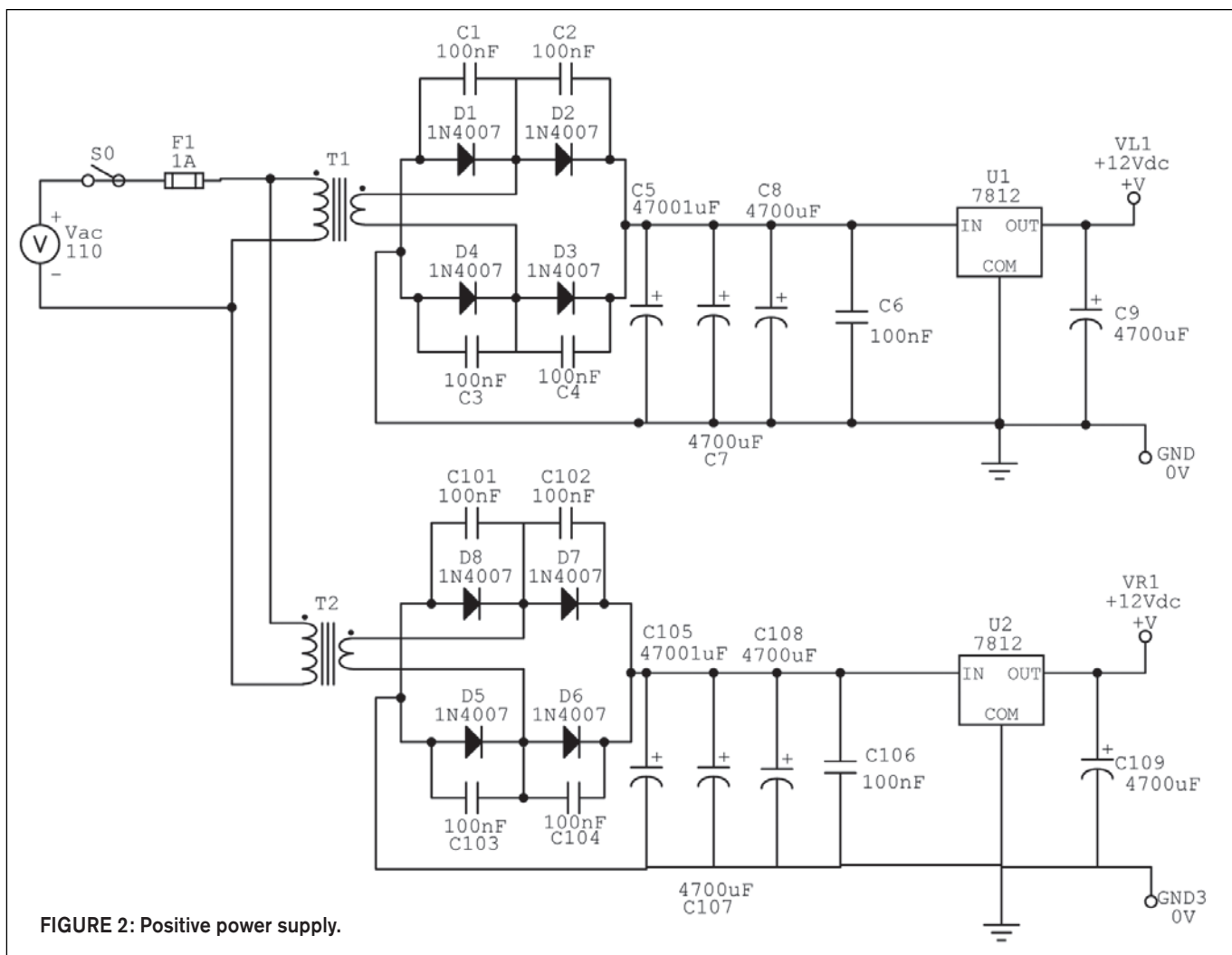


FIGURE 2: Positive power supply.

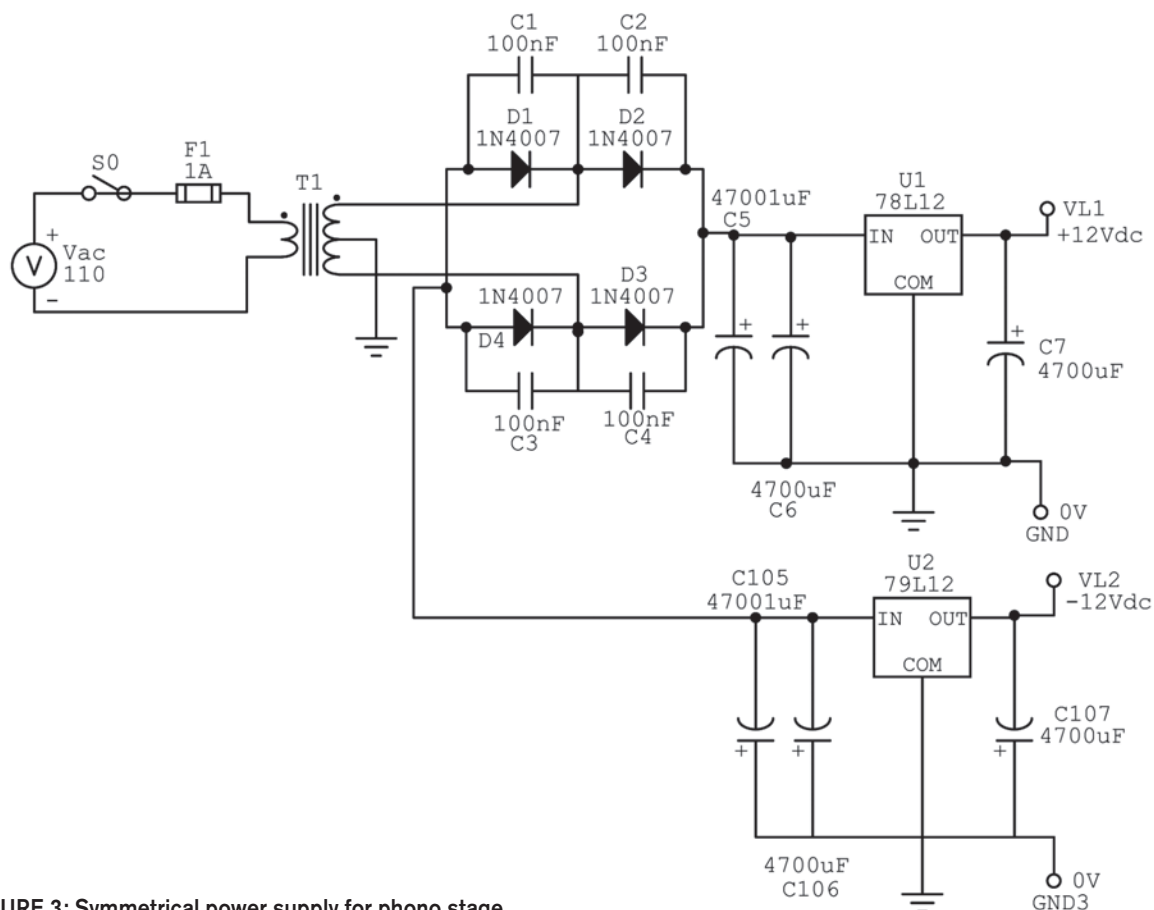


FIGURE 3: Symmetrical power supply for phono stage.

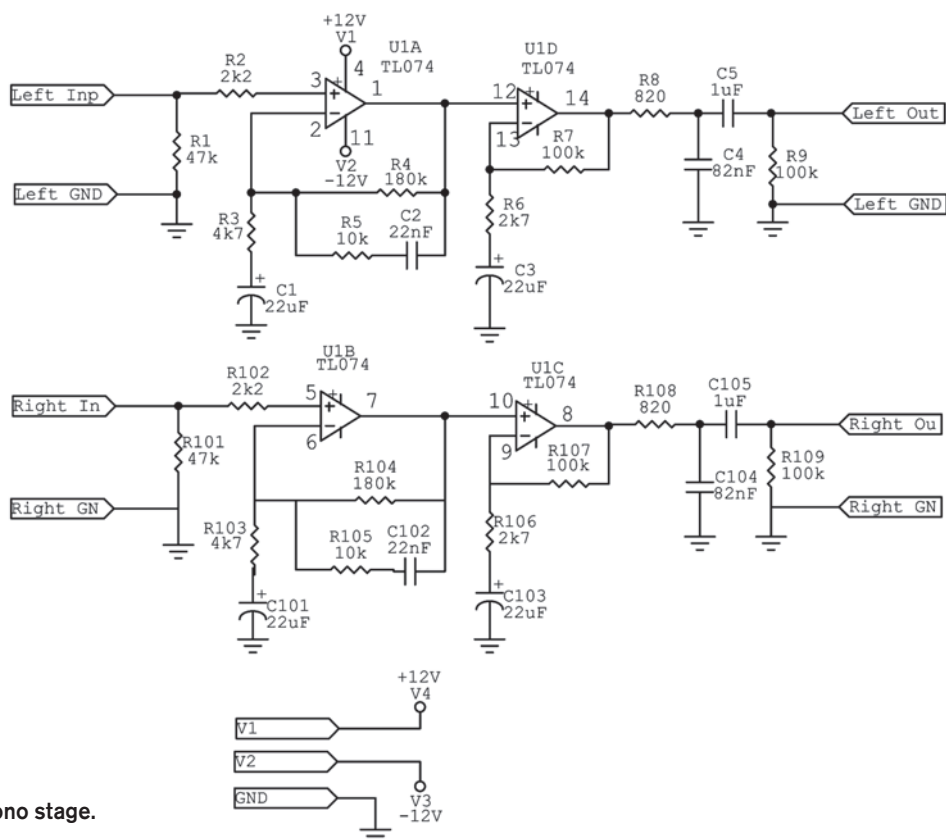


FIGURE 4: Phono stage.

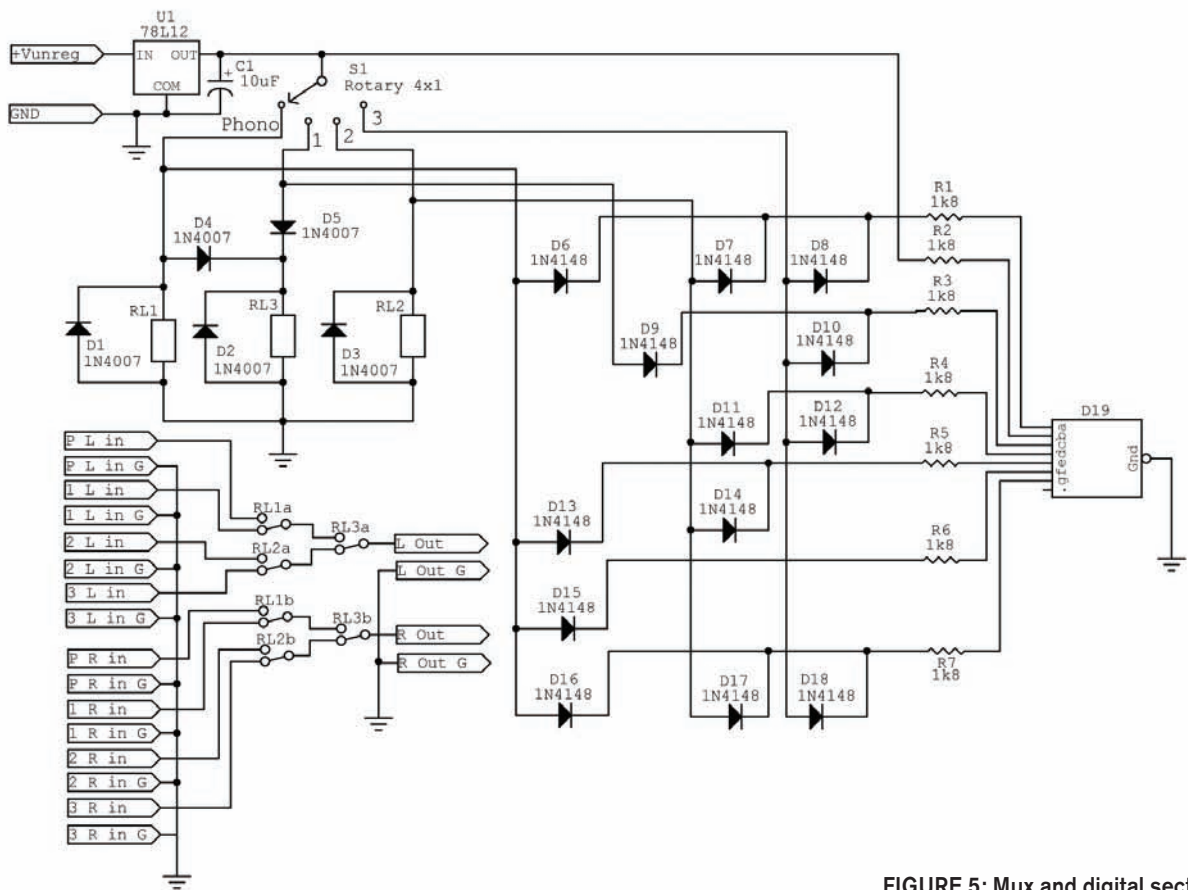


FIGURE 5: Mux and digital section.

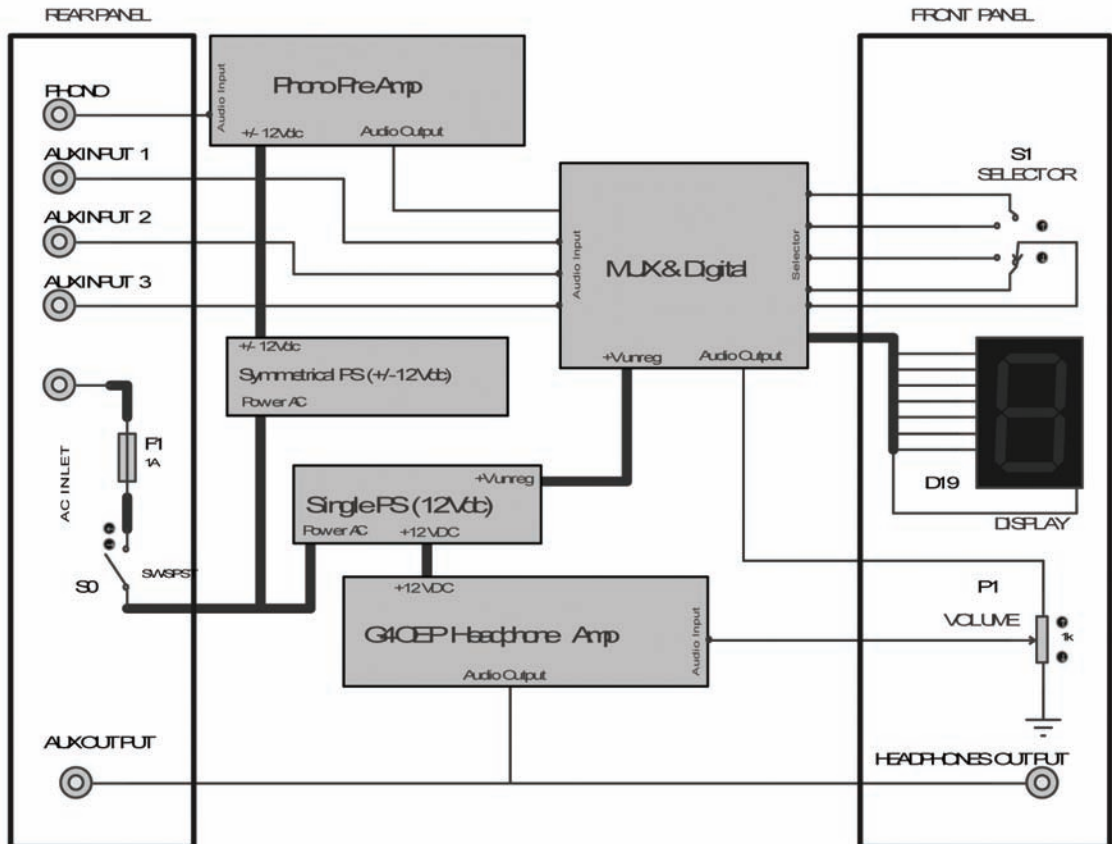


FIGURE 6: Connections diagram.

MUX AND DIGITAL SECTION

I needed to use my headphones with the following different sources: a Marantz CD-63SE CD player, a Pioneer VSX-4800 tuner, a Marantz SD-535 dual tape deck, and a turntable (Garrard B2-FG and/or a Thorens TD-166). So I decided to build a relay circuitry to do this and more. Small numbers around the selector switch were not an option, because I needed to see the selections from a distance. **Figure 5** and **Table 5** show the MUX and digital section.

I used three small sealed reversible relays with two sections NO-NC each, to do the job, commanded by the S1 rotary switch and the “OR logic” provided by some diodes in a simple and old-fashioned way. I used the same approach to command the D19 (seven-segment CC display). Depending on the S1 position, D19 displays “P” for phono input or “1,” “2,” or “3” for auxiliary direct inputs.

Note: If the switch is in the P (phono) position, all relays are displayed “off” in the schematics. This is to facilitate construction, regarding normally open (NO) and normally closed (NC) contacts, depending on the kind of relays you are using.

D19 is an LED seven-segment common-cathode (CC) device. I used a subminiature one (10mm × 7mm). As you can see, I used a LM78L12 TO-92 small regulator, because I supplied the power for this perfboard from unregulated +12V DC (in fact, 18V DC unregulated—taken before U1 or U2) from the G4OEP power supply. You can choose any one unregulated source to feed the MUX; left or right will work fine. I experienced no audible interference.

CONSTRUCTION

The cabinet is a simple MDF box with five faces, no bottom end, covered with Curupixa (pronounced Kooroopyshah, a Brazilian wood) darkened foil. I machined all holes in the MDF box from the inside before mounting. After machining, I covered the box and the wooden foil with contact glue (beware of poisonous vapors when using), let them dry for ten minutes, and carefully applied the foil over the box with a metallic, smooth, rounded mass to eliminate air bubbles between the MDF and the covering.

I then applied glue to a ground-plane made from aluminum cooking foil and covered the inner box surface with glue from a glue-stick. Some sanding work and carnauba-wax application guarantees a natural opaque finish.

I chose the chicken-head style knobs to give a vintage look to the G4OEP headphone amplifier. I also built a pair of acrylic round washers (30mm diameter, 2mm thickness, and 5mm in length) to put around the knobs, just for aesthetics; my intention was to illuminate with LEDs

from the backside when it is powered-up.

The small white square in **Photo 2** is the display, showing the selected input from the back-panel. Finally, a grounded 2mm aluminum plate covers the bottom side of the MDF box. Small rubber feet glued on this plate sustain everything. **Table 6** lists the final parts.

Figure 6 shows the system connections (one channel shown). I hope this little challenge encourages other *aX* readers to build their own units and share the results.

aX

TABLE 5: MUX AND DIGITAL SECTION PARTS LIST

Components	Attributes	Designation
10μF × 25V	Electrolytic	C1
1N4007	Si rectifier diode	D1-5
1N4148	Si signal diode	D6-D18
7-seg display cath comm	Subminiature red type	D19
78L12	400mW positive regulator	U1
1k8	metal film ¼W	R1-7
12V DC 2 reversible sealed	JX2RC2 Metaltex	RL1-3
Rotary switch	EV4X1	S1



PHOTO 2: Completed headphone amp.

TABLE 6: MISCELLANEOUS PARTS

Components	Attributes	Designation
100k	LOG dual potentiometer	P1
On-off switch	SPDT switch 220V AC at 1A	S0
Aluminum plate	150 × 260 × 2mm	
Aluminum foil	Half square meter	Ground plane
4 rubber feet	20 × 10mm	
RCA connector plate	4 dual female	4 selectable inputs
RCA connector plate	1 dual female	Auxiliary output
2 knobs	Chicken-head style/beige	Volume and Input selector
P-10 stereo jack output	Connector to headphones	
Power cord for IEC connector		
MDF 15mm thick box	190 × 95 × 270mm (DHW)	5 sides (without bottom side)
2 Acrylic washers	30mm diameter 2mm thick	

Removing the Six “Straight Wire” Kinks: Prologue to a Better Preamp

Here are some—perhaps forgotten—tips on preamp design.

I finally needed to replace my “pre-amp,” which was nothing more than a stepped attenuator and four jacks in a small aluminum frame. It just didn’t look right, and was a bit limiting when changing sources. Plugging and unplugging cables have just too many disadvantages.

It is often said that the best preamplifier is a “straight wire with gain.” But I do not think we should set our sights that low.

Electric circuit theory is a simplification of how circuits behave. Even in the limited scope of this theory, wire has three characteristics of concern: resistance, capacitance, and inductance. Many audio effects are not well covered by this approach—for example, what happens when a semi-conducting crystal forms in a conductor—but circuit theory is a good starting place.

FACTORS

It may be simplistic to assume that you want the lowest resistance possible without considering the actual use. The source impedance of most consumer audio gear is about 100Ω at low frequencies rising at higher frequencies, designed to be loaded by at least a $10,000\Omega$ load. When you have resistance in the “straight wire” conductor, there is a loss of signal level. Three feet of 24-gauge wire loaded by $10,000\Omega$ has a loss of about 0.0000779dB. Increasing the wire to 10 gauge drops the loss to 0.00000258dB. In audio there are gifted people who do hear more than others, but not in this case.

The second concern is capacitance loading the signal source. The 3’ cable loads the source causing a loss of about 0.011dB for low source impedance to 1.16dB with a rising to high source im-

pedance. The phase shift may run from just under 1° to more than 7° at the upper frequencies.

In the worst case this is something a critical listener could hear. When you consider that the signal probably goes through two or more cables (they may be longer than 3’), the signal source probably does not drive capacitive loads as nicely as pure resistive loads, and the capacitance may even be nonlinear, then you begin to see where one “straight wire” problem lies.

The issue of inductance is a bit different. Resistive losses do not vary with frequency. Capacitive losses increase with frequency, as do series inductive losses. The difference is that capacitive losses load the source, while inductive losses raise the load impedance. Inductive losses are also greater at high frequencies. In the 3’ cable you see a loss of about 0.00785dB at high frequencies compared to 0.0000000785dB at the lowest frequencies. Is it possible some could hear this? Probably not!

MORE CONCERNS

The next problem of the “straight wire” are the six sources of noise it is susceptible to: radio frequency interference (RFI), capacitive coupling, magnetic coupling (or mutual inductance), common source impedance, vibration, and moisture.

I treat RFI as a separate source because it is usually from a far-off source. In reality it is a combination of capacitive coupling and magnetic coupling. The cure is to shield the wire and gear, even though that adds capacitance and signal loss. Shields also have the problem known as skin effect—the lower the frequency of the RFI, the farther it can penetrate the shield. RFI often shows up

as added noise or grit to the sound.

Capacitive coupling of noise can be treated by field theory or approximated by circuit theory. For the circuit theory approach, assume there is a small capacitor hidden in the circuit that couples the “straight wire” to the noise source. You can also model a capacitor to ground from the conductor and even allow for some attenuation from a shield.

Magnetic coupling depends on the area of the conductor and strength of the field. Moving the power transformer away from sensitive areas is always a good idea. Making circuit paths shorter also provides benefits. If the “straight wire” is a real cable grounded at both ends, it is almost certain that you will have a magnetically induced voltage.

Common source impedance is why lights flicker when a motor starts. When the desired signal shares a signal path with something else, the other signal causes a voltage drop in that path. If the other signal varies—say, from noise or even signal—that variation and its companion voltage drop shows up in the signal path.

Vibration is always present. Insulated wires will have triboelectric noise. Capacitors will react by modulating the spacing between electrodes and hence their capacitance, resulting in an AC voltage from any bias placed across them. Inductors may have movement in their windings, causing value shifts. Even some resistors may generate noise from movement.

Moisture, of course, can cause corrosion. Whenever two different metals touch and moisture is present, you produce a small noisy battery. Many folks find always leaving their gear on makes it sound better. Some of this is due to heat reducing the moisture in compo-

nents such as resistors and capacitors. Both are affected to a small degree by moisture. Of course, the biggest problem can occur in switches, where the moisture may cause oxidation of the contacts. Sulfur or other compounds in the air make this even worse.

The final issue in the "straight wire with gain" is gain. If you are using a CD player or tuner as your music source, both of these are capable of delivering 1V RMS. Some high-powered professional amplifiers will produce full power with less than that. Even a low gain (26dB) power amplifier will deliver 50W with that input. So for many cases you do not need gain—you need loss! Of course, if you have inefficient loudspeakers or perhaps want tone control, then you may want some gain.

SHIELD SOLUTIONS

It is misleading to think of gain as only voltage amplification. You may wish to increase the current available to the load. This helps to reduce the effect of capacitance or other loads inside the power amplifiers or connecting cables.

It turns out there is a nice simple solution that will almost completely eliminate capacitive loading of interconnecting cables, increase the effective attenuation of the shield, and add almost no noise of its own.

That solution is the driven shield. Instead of surrounding the signal carrying wire with a grounded shield, use one that carries the same signal! Now the capacitance is not to ground but to the same voltage. The insulation can be nonlinear, because there is no voltage across it, no nonlinearity. The inductance of the wire is very slightly reduced by solenoid effect and some reverse capacitive coupling, although this is not really of practical value.

For RFI the interference now must travel through an extra isolated shield, so this one change multiplies the loss of a single shield. Same deal for any capacitively coupled noise.

To get even trickier, you can now cover this shield with a second insulated shield, which will prevent the signal from inside leaking out and contaminating the ground or return reference signal. If you twist a single conductor wire around this second shield and use this

as the ground reference, this will help to reduce magnetic coupling. You get some of the advantages of a balanced line in the single-ended cable.

Of course, to keep the RFI out of the ground, you should add a third shield. To minimize the undesired energy, the second and third shields are grounded only at the driven end. This prevents mutual impedance of the shielding system from entering the signal path.

You can add a buffer amplifier at your source to power the driven shield. Of course, if you do not want to modify your existing gear, an easier way is to put all the buffers in a new preamp. The buffers need to be decent, but not great amplifiers. After all, they will be attenuated by the amount that used to be adequate just for noise. This is a good place for amplifiers such as the TL072 or any FET input amplifier, because they are less susceptible to RFI. Of course, you probably can even get away with a golden oldie such as an NE1558. I mixed them up to see whether I could hear a difference.

In my design I used several small cards to become input or output circuits (Fig. 1). The sending source can be a CD player, DVD, tuner, and so on.

I designed two versions of the buffer card. One is a send card designed to bridge the output and provide a modest current signal to the first active shield. It also tries to make sure all common mode currents are accounted for properly. It does this by grounding the source of the active shield's current at the source. The return signal wire, which was the ground and shield in a normal cable, is now just a wire going to the common of both connectors. The final outer third shield is there to try and scare RFI away.

The send card can take either a direct wired input or a signal from an optional RCA jack. The output is on a stereo ¼" jack with a case grounded binding post for the outer shield. The connecting cable has a matching ¼" stereo phone plug and a wire lead on one end and a standard RCA on the other.

The receive version of the card works by making it easier to literally suck more signal from the source. It requires a special cord that has an RCA plug for the input and a phone plug for the output. The received center conductor signal



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AP-016



\$94.50

The AP-016 is available in three versions. The AP-016A crossover point is at 2KHz, the AP-016B is at 2.7KHz and the AP-016C is at 3.5KHz. All of them are 4th order Linkwitz-Riley crossover.

Specifications:

HF Power Output: 30Wrms

LF Power Output: 80Wrms

THD: 0.03%

S/N ratio @ rated W: 90db

Input sensitivity: 1V

Input impedance: 22Kohms

4th order X-over: 2KHz AP-016A

2.7KHz AP-016B

3.5KHz AP-016C

Weight: 2.6Kgs (5.7lbs)

Dimensions W x H x D:

W: 137mm (5.4")

H: 218mm (8.6")

D: 81mm (3.2")

Cut-Out W x H:

W: 108mm (4.25")

H: 190mm (7.5")

AC Voltage: 115V / 230V

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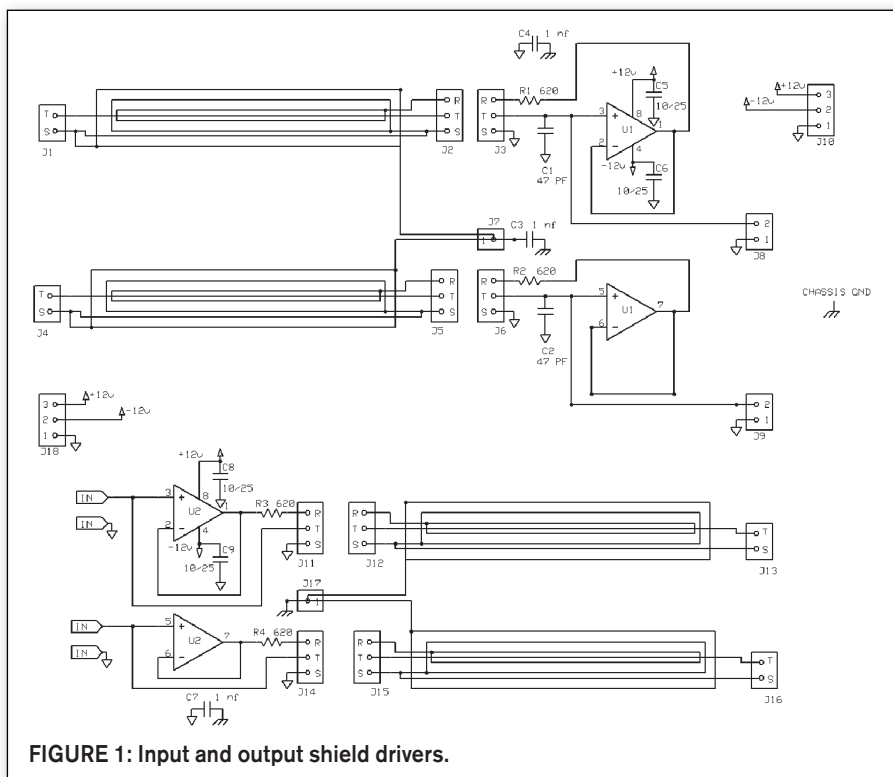


FIGURE 1: Input and output shield drivers.

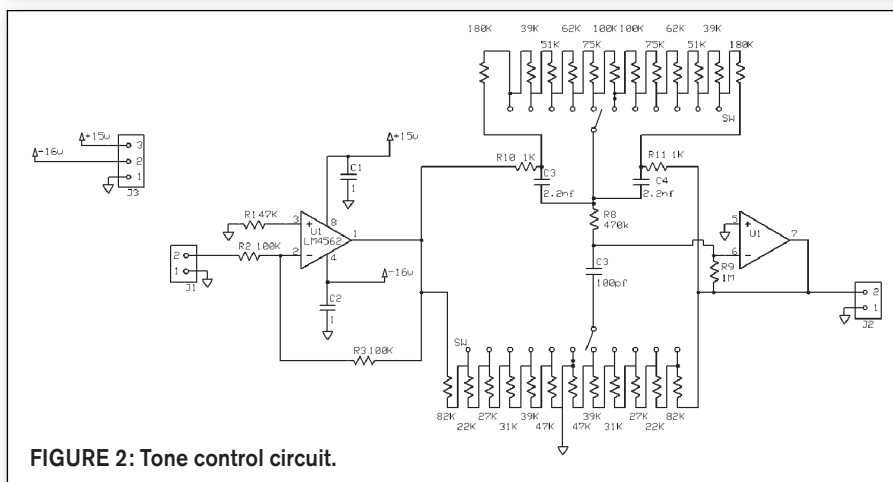


FIGURE 2: Tone control circuit.

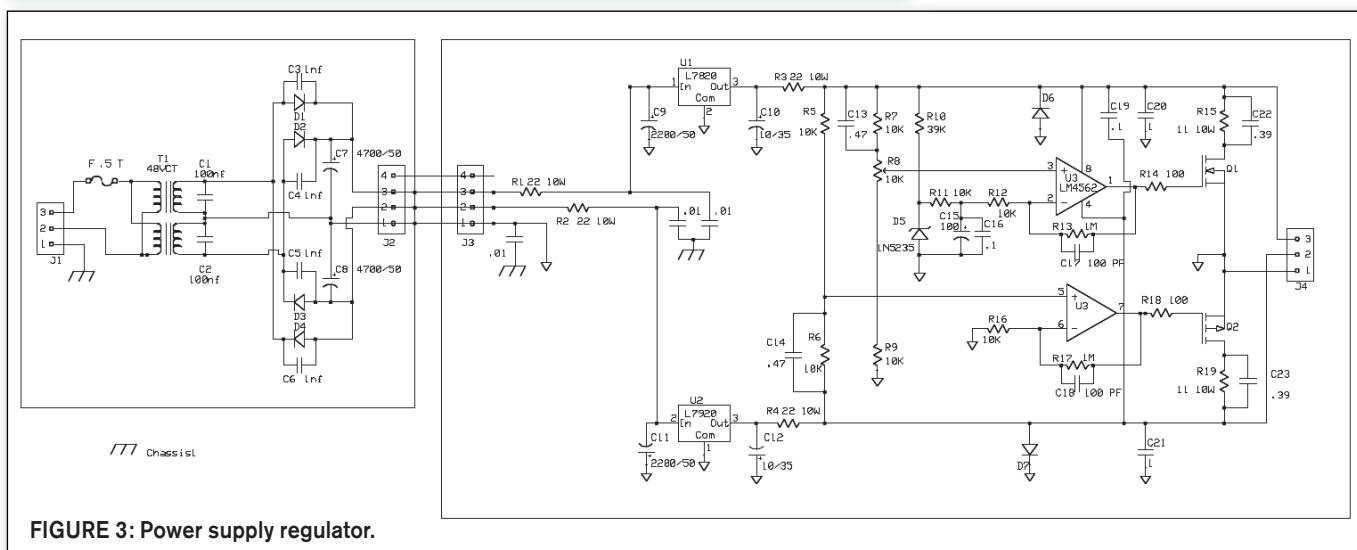


FIGURE 3: Power supply regulator.

is buffered and sent backwards on the active shield. The next shield is now grounded at the receive end, where the induced current originated. The outer shield is connected to the send side because any noise from an outside source is best dealt with from the stronger signal side. However, the receive end of this shield is connected to the case through a small capacitor.

The capacitor to the case keeps any RFI shielded, or at least all of the components at the same potential inside a Faraday cage. There is also a capacitor (C4) from the buffer common to the case.

PARTS SELECTION

One problem quickly showed up. It seems that some designers have tweaked their designs to work into a capacitive load! Some outputs are of such high impedance that the shield buffer acts as positive feedback. So placing a small capacitor 47 to 330pF silver mica (C1 and C2) between the signal lead and buffer/preamp common when needed will restore the source signal to flat.

A second problem is that a rail-to-rail op amp does not want to work into a capacitive load, so don't use one.

I used gold-plated Neutrik NRJ6HF-AU phone jacks for the 1/4" phone jacks. These come without mounting nuts, so be sure to order some. They fit nicely into a 7/16" hole. Neutrik also makes gold-plated NP3Xb phone plugs. I did not use anything fancy for the third shield binding posts.

For the RCA connectors I tried three

different types: all gold-plated audio-ophile types. You can pick your favorite here; just be sure it can handle a large diameter cable.

To construct some of the cables I used silver or silver-plated stranded Teflon insulated wire. For others I cheated and started with a piece of shielded audio cable cut from a decent premade RCA-RCA cord. The quality of the signal conductors does affect the final product. I have found that thinner better metals do have an advantage.

The shield braid used to be a common item. I could only find it in the larger size at McMaster Carr (part #5537K26 and 5537K14).

I used a third-hand jig to hold the connector and cable while soldering. I did not want any motion while the solder cooled because theory suggests that this promotes crystal formation.

After soldering the RCA end center silver 22-gauge conductor, I slipped over 1/8" heat shrink, cooked it, and then placed the rear cover. Next to go on was a thin shield braid that did not quite extend into the cover. I added a small jumper connected to this at the non RCA end. Tape kept it in place so it would not short to the center conductor.

A layer of 1/2" heat shrink that did go into the cover was next. More thin braid went over this, but again not into the cover. The jumper went to the RCA end for the send cables, the far end for the others. I wrapped a silver or silver-plated insulated wire around this and soldered it to the shell terminal. If this is a send cable, you need to connect the jumper and a short bit of uninsulated wire. The next piece of 1/2" heat shrink covered this.

I connected the phone plug to the other end. I needed to cut the strain relief short and ream it a bit to fit the cable. The larger shield braid goes over the completed connectors, and then trimmed and soldered to the appropriate jumpers. A bit of tape held this in place and Techflex tubing in red or black covered the outer shield. This was held in place at each end by more 3/4" heat shrink tubing that also covered the rear shells of the connectors.

It was easy to push the braid and Techflex over the cable, but the heat shrink sometimes was difficult. It was easier to push a stiff wire through the

heat shrink and use it to pull the cable through. On long runs you can use a brass wire and solder it to the silver wire. After you clean the connectors with De-oxit, the cables are complete.

Of course, if you are using audio cable for the core, you can skip a step.

I positioned the send and receive modules so that two receive and one send fit on an Expresspcb mini-card. There is nothing special about the cards, which do not handle high-quality signals. I did isolate their power supplies

from the audio circuits by using two 10Ω resistors for each card.

You can use the cards and cables as standalone devices. I used one as a guitar cable buffer, which resulted in noticeably better high end. I also chose to build these cards into my new preamp.

In my preamp the signal lead from each input card goes to a selector switch via a piece of un-insulated dead soft silver wire, which is attached to nylon standoffs to minimize vibration-induced signals. I used a 12-pole double throw

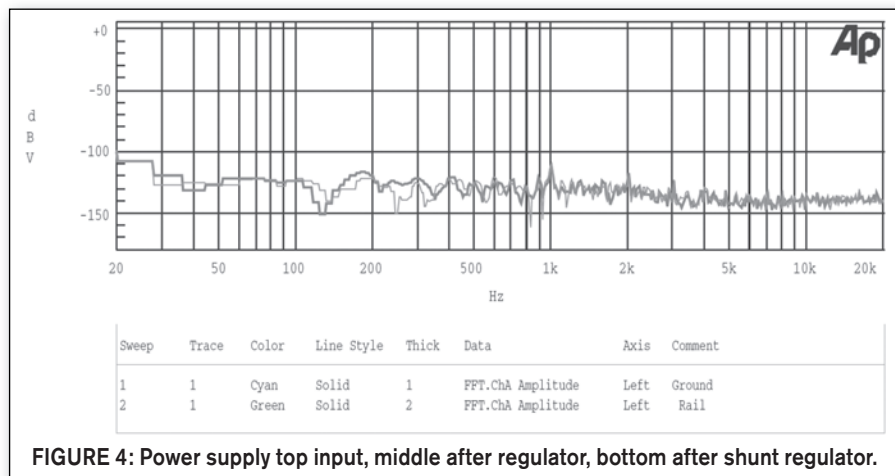


FIGURE 4: Power supply top input, middle after regulator, bottom after shunt regulator.

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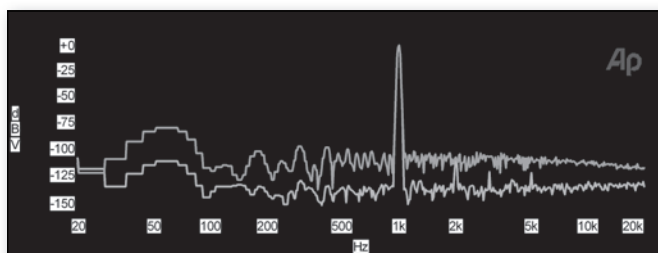


FIGURE 5: Not so good DIY cable top, bottom preamp and cables. Note HF rolloff and noise level hides THD inherent in test equipment.

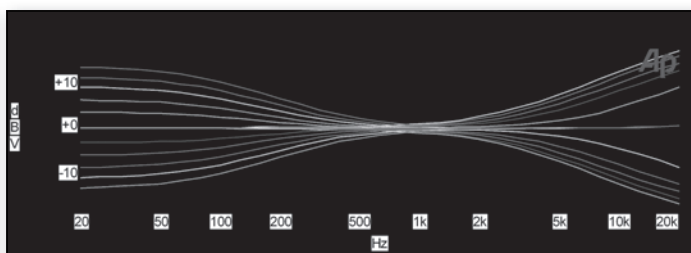


FIGURE 6: Preliminary tone control curves.

(12PDT) rotary switch for the selector. It is a military surplus silver-plated large diameter ceramic rotary switch. I treated the contacts with Deoxit. The signal return or “ground” leads from each input go directly to the output send buffer card, where I establish my common ground reference point. A better point would be at the return end of the attenuator, but I did not figure that out until I finished.

The output of the selector switch goes to a stepped attenuator. I had a leftover 24P4T Alps switch made for this purpose. I made PC cards to hold 1206 size 1/3W surface mount resistors, a size that can be soldered without magnification. I used 94 of the IRC PFC-W1206LF-03-XXXX-B tantalum nitride 0.1% 25 ppm/C non-magnetic resistors to make the attenuator. I selected these because as high-end audio parts go, these are reasonably priced. I like the 1206 size because they are easier to solder. My attenuator has 10K input impedance and varies on the output. I learned from my previous “preamp” that most of my listening is done about 20dB down, so I jiggled my step sizes to have more resolution in that range.

After doing all this, I realized next

time I will buy an attenuator already assembled, which is easier, works the same or a bit better, and actually costs less!

The output of the attenuator originally went to the output card. An outboard $\pm 36V$ DC power supply fed an internal shunt regulator to power the buffer cards. A direct run went to each card establishing a second common ground at the power supply, reducing any mutual impedance signals. A single connection coupled the output card to the power supply.

I used a shunt regulator modified to have more AC response; in this use with a known load and good shielding, RFI should not be a problem. Measurements showed the shunt had more than 20dB additional noise attenuation than with just the regulator. The power transformer, bridge rectifier, and first filter caps are in a small steel box 6' away from the rest of the gear.

The downside of this shunt regulator is that it draws a higher startup current than a series regulator. So it's best to use a time delay fuse.

TONE CONTROL CIRCUIT

You need a tone control to complete the preamp. The tone control was actually

the trickiest part. In the late 70s, when mixing boards were becoming common, the tone control issue re-emerged. Initially it was more a question of how to make an effective low-cost tone control, but once the knockout punch was published by P. J. Baxandall in 1952, there was only quibbling about variations. By the 70s, with cheap parts, ICs, and biquad or other new filter topologies, the issue became how should the knob change the sound?

The problem now was not so much designing the circuit, but getting the “feel” right. When you give the bass knob a quarter turn, you expect certain changes. More or less change is awkward. As loudspeakers improved, the apparent effect of the control was greater.

What was once a fine treble control now is too harsh on a modern tweeter. A bass control that was fine on a tight suspension woofer becomes tubby on many around today.

The standard Baxandall circuit for a two-knob tone control is three capacitors, two variable resistors, and a few fixed resistors. To get smooth action the controls needed to be linear. A later takeoff used log pots and different capacitor values to get a better feel. The

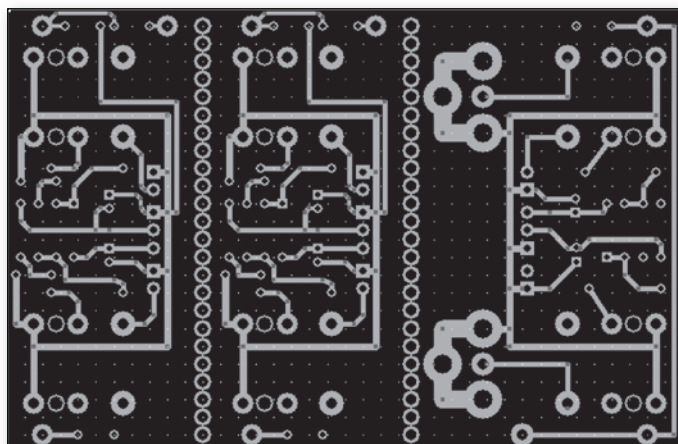


FIGURE 7: Bottom PC layout.

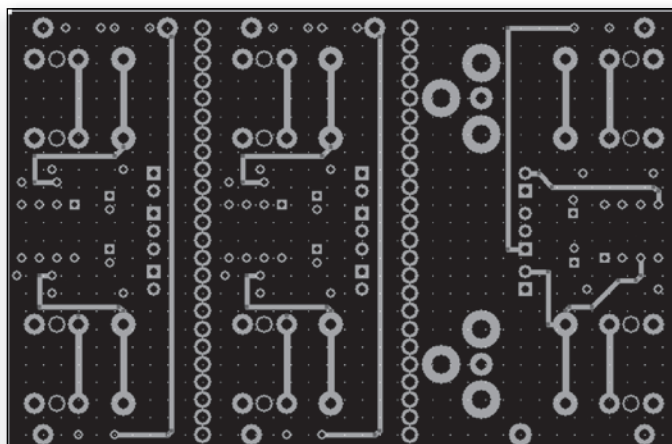


FIGURE 8: Top copper layer.

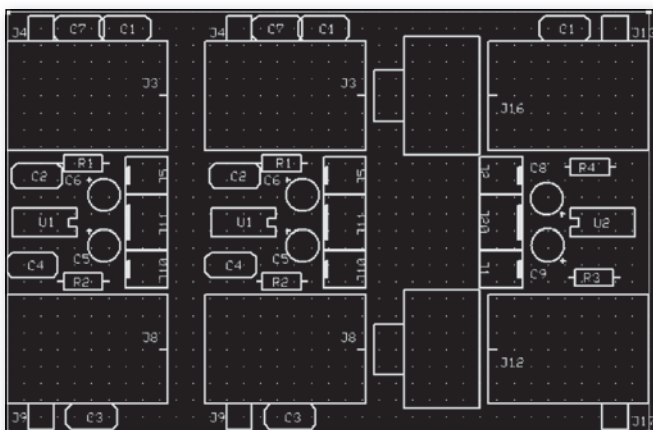


FIGURE 9: PC cards silkscreen layer.

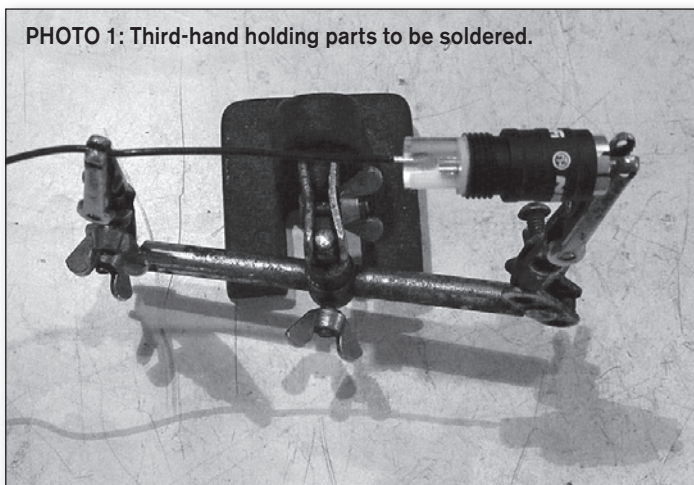


PHOTO 1: Third-hand holding parts to be soldered.

disadvantage of the Baxandall was that it required a center-tapped pot and a pentode to provide the negative feedback.

Today negative feedback is not a novel application. Stereo, however, has placed a few more demands on a tone control. A stereo potentiometer with good tracking, high resistance, and a center tap is really hard to find. Using switches and resistors gives me a great advantage. I can select the resistor values to give me logarithmic control and still use matching capacitors.

I used regular carbon film resistors for the control switches. Once my crew and I have used the control a bit, it will be practical to home in on the useful range and pick exact values. Then I will change to better switches and resistors. The circuit I used is deliberately very close to the original design and is shown in Fig. 2.

I used an LM4562 op amp, silver mica capacitors, and precision resistors for the tone control cards because I need to only

change the switches later. The PC cards for the tone control are stacked on top of each other to maintain a similar level of wire lead path length and vibration level.

To keep out vibration I made some feet out of two aluminum discs and two different foam rubbers. Both are quite soft—40 00 durometer for one and a bit firmer for the other. The impedance mismatch helps to reduce vibration. I mounted the feet to a sub-plate $\frac{1}{2}$ " deep and added a bit of foam to seal this chamber to the chassis. I used 6-32 screws with pieces of auto store $\frac{1}{8}$ " ID rubber hose as flexible spacers to attach the plate.

I punched a 1" matrix of holes in the bottom of the chassis to make mounting components easy. The sub-plate now sealed the chassis. Inside was silica gel to keep moisture down and as a bonus the loose fill absorbs vibrations. The final environmental control was a strip of silver tarnish paper to remove sulfur and keep the silver shiny new.

RESULTS

Measurements of the preamp showed that it was actually better than one web DIY cable design. It was as good as a jumper cable across the test equipment for most of the audio band. There was a higher level of 60Hz hum than from a short jumper. This was interesting because the fundamental of the power supply was 120Hz, which was 105dB down at the output of the regulator. So I suspect this is due to the longer cables and greater area in the possible magnetic field.

When I first turned on the new preamp, I heard a noticeable buzz in the sound. Then I realized the preamp was set to tuner for a sports-talk station. The buzz was just normal for them! On everything else the system seems dead quiet.

So it is possible to do better than a "straight wire with gain." You can have less noise, lower losses, higher output current, selection ability, gain, and tone controls. *aX*

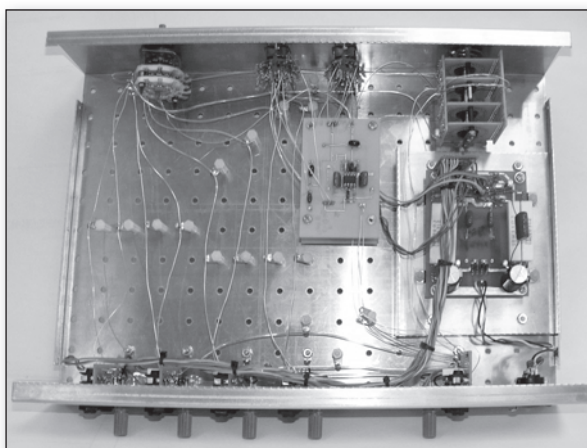


PHOTO 2: Chassis wiring.



PHOTO 3: Bottom pan with gasket and silica gel, foot parts, and rubber tubing spacers.

Three Novel Transformers for SE Amps

Introducing the next generation of output transformers.

For many reasons, I am convinced that in the sound reproduction amplifiers field the single-ended topology, combined with triodes, produces a better sound than any other. Many readers will agree with this statement, others will not. I must specify that most of those who support the push-pull layout are motivated by the power output they can get from this kind of amp, which can easily reach 50, 70, 100 or more watts with just a pair of strong output valves per channel. On the other hand, the single-ended (SE) units, in order to deliver such power, require very large and expensive output transformers.

This explains why SE amps in excess of 20W are not the rule. The cause is well known: the idle DC (direct current) at which the valves operate in their still condition in a SE layout.

When designing a standard SE transformer, you must consider that part of its core magnetic headroom (that depends on the size: cross-sectional area and length of the magnetic path in which the flux travels, as well as the quality of the ferrous material) will be "occupied" by the magnetization due to this direct current. Enough availability must be reserved in the core for the alternating current signals to prevent an excess of flux density to cause its saturation, in which case the distortion of the sound increases.

Many well-known scientists have spent some time trying to find a solution to the SE transformer's drawbacks, starting with the non-productive DC field, but their suggestions were not considered justified to re-

place the gapped¹ system that remained the best alternative to the push-pull. Being deeply interested in using the SE topology, but with more power, I resumed the search to find a solution to this unacceptable lack of efficiency.

SC-OPT (SELF-COMPENSATED OUTPUT TRANSFORMER)

Who hasn't thought about a system based on an extra winding generating a magnetic field of the same strength as the primary's, but in the opposite direction, in order to accomplish a complete and automatic offset of the DC fields? It is certainly the first idea that comes to mind, if you overlook the fact that if this works for the direct current, the same offset also applies to an AC signal applied to the primary. Consequently, no voltage would be transferred to the

secondary² (Fig. 1).

To save the AC of the primary, you need to suppress the AC voltage that fights against it, in the opposed winding (that I will name the "tertiary"). A suitable capacitor shunting the tertiary can do the job, for it hits only the AC, not the DC, but you soon realize that the primary's inductance is severely affected, as you will see later. Consequently, if, on one hand, it is quite easy to achieve the DC offset, and short-circuit the opposed AC potential, another problem arises, and you must restore the primary's inductance that has collapsed, due to this short circuit.

You know, however, that the reduction of inductance is greater if the coefficient of coupling between the opposed windings is high³. This coefficient, in practice, ranges from very close to 1 (if the coupling is tight) and falls below 0.9 (if the coupling is loose). It becomes obvious that if you want to limit the damages of the tertiary's AC short-circuiting solution, you must reduce the coupling between, on one side, the tertiary and, on the other side, the primary with its secondary(ies). Irrespective of the coefficient of coupling between the main opposed windings, tertiary and primary, the latter must have an extremely high coupling coefficient with its secondary, which means a good winding geometry and interleaving.

I started bench experiments on this basis and included the physical separation of the conflicting windings, using suitable C or UI cores. But the real solution came when, encircling the tertiary's bobbin with a few turns of magnetic band on

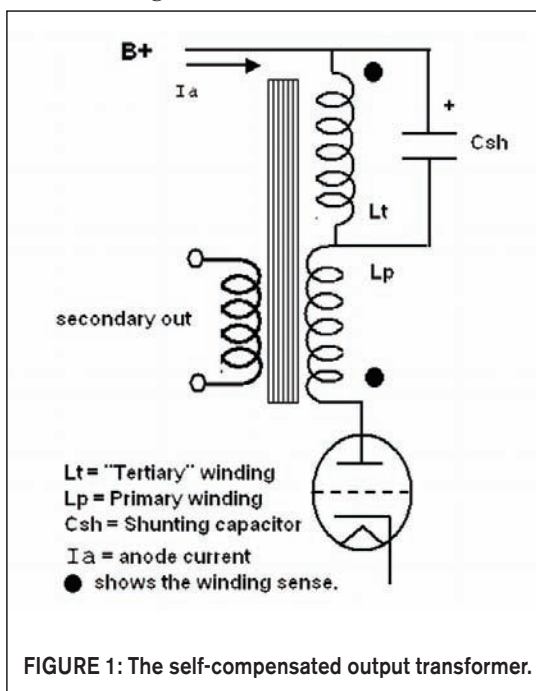


FIGURE 1: The self-compensated output transformer.

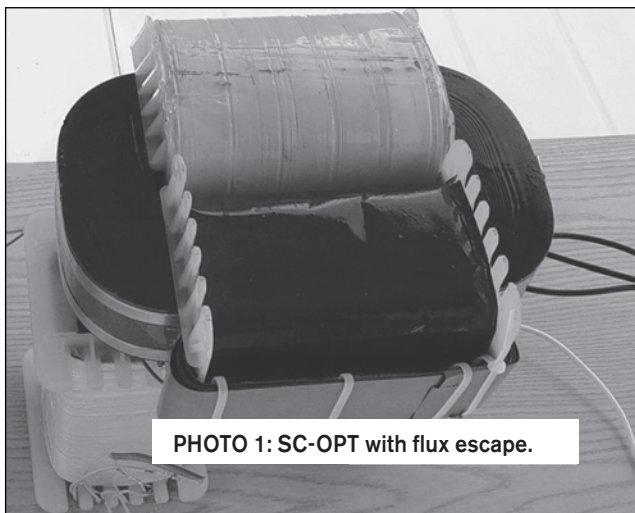
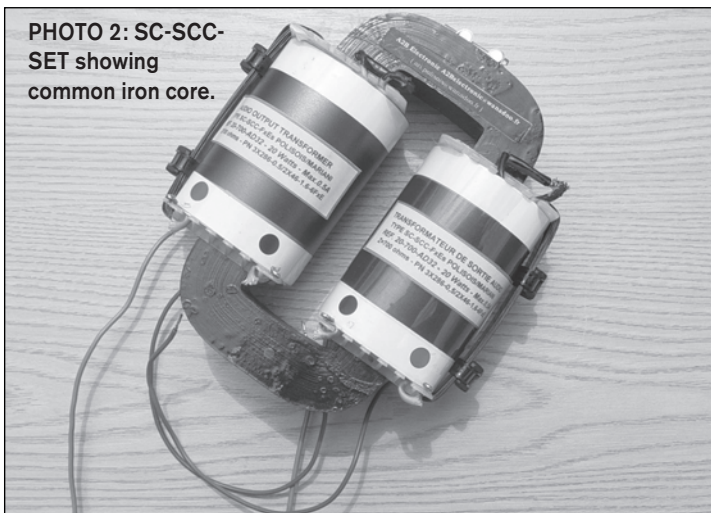


PHOTO 1: SC-OPT with flux escape.

PHOTO 2: SC-SCC-SET showing common iron core.



the outer side, I noticed a significant reduction in the loss of inductance of the primary. **Photo 1** shows the “magnetic flux escape.” This device can be considered an auxiliary magnetic core that absorbs part of the magnetic lines, thus reducing the effective number of lines that the tertiary and primary have in common (and, consequently, the coupling factor).

As an example, I consider the case of an output transformer built for a 6C33C-B valve, the powerful tube with about 100Ω internal resistance. Initial primary inductance (measured, without load, with an inductance meter) = $4.8H$ (1200 turns – core area $7.6cm^2$ – magnetic length $0.43m$). This inductance, with the tertiary (same number of turns) shorted, falls to $0.48H$ (1/10 of the initial). I calculated the coupling coefficient to be 0.95.

The $-3dB$ point at low frequencies results (theoretically) in $35Hz^4$. By adding six turns of a $0.35mm$ thick magnetic band around the tertiary bobbin, the primary inductance was restored (due to the magnetic lines leaking from the tertiary field in the main core to this additional external magnetic circuit) to over $1H$.

In spite of an improved bass response, the advantages I was seeking (reduced bulkiness and weight, compared to a standard SE output transformer) had not been fully reached, and, from an efficiency point of view, you could not consider the unit able to compete successfully against a push-pull transformer. Nevertheless, the fact that a SE amp using low internal resistance valves,

when equipped with an SC-OPT, delivers much more power than with a gapped transformer of the same size, is a real, substantial improvement.

A pleasant surprise was the discovery that the sound it produces is extremely clean and airy, as many listeners noted. However, no scientific explanation was found for this sonic feature. This unit was presented at the AES 118th convention (Spain, May 2005, paper 6346).

Despite its use, for best results, of low internal resistance valves, this special SC-OPT features several advantages with respect to sound quality⁵.

An amp I listen to very often at home is fitted with two paralleled 2A3s per channel (equal to 400Ω internal resistance). It uses a self-compensated output transformer and even if its bass range extends to only $46Hz$, at $-3dB$, it is my preference, because of the unequalled

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SC-SCC-SET

In May 2006, engineer Giovanni Mariani, technical director of Graaf, Modena, Italy, specializing in output transformer-less (OTL) amps, and I presented a paper describing another novel transformer, the SC-SCC-SET (split core, stereo common circuit, single-ended transformer) at the 120th AES Convention held in Paris, France (paper 6831).

Considering the fact that a loose coupling in the same magnetic core allowed some freedom for the primary winding to work, producing an ideal separation with the help of the opposed DC fields, both the right and left channels (two primaries with their closely coupled, interleaved secondaries) were fitted on the same common iron core (**Photo 2**). This solution required a phase inverter, as with the push-pull transformers.

The DC offset was also ensured, as in the p-p OPT, by feeding the left and right primary coils, having opposed winding senses, with the same amount of anode DC current. Obviously, there was no self-offsetting feature, and the balancing of the DC fields must be accomplished with a suitable, manual device.

The advantages were:

- extended bass range (as low as 5Hz at -3dB) with a very good high-frequency limit (**Fig. 2**)
- saving space, weight, and cost, with regard to the former SE transformers⁶⁻⁷
- getting as much power as the push-pull configurations, but with the properties of the SE layout (sensibil-

ity, lower content of odd harmonics, simplicity of the circuits, and so on).

This transformer was also designated as “double-ended,” as opposed to the classic SE transformer.

Its disadvantage, however, is that some [crosstalk]⁶ is present, but limited to the low frequencies (thus not affecting the stereophonic effect). It is practically unnoticeable, if the loudspeakers are three or more meters apart.

As stated, the two primaries (left and right channels) in this setup produce opposed fields (as in a push-pull transformer). The DC magnetic fields produced cancel each other, provided the idle currents are of the same intensity (but, as mentioned, this is not automatically ensured), whereas the AC ones, thanks to the opposed phases, do not interfere with each other.

How does the sound of this OPT compare with that of an SC-OPT? What strikes me immediately is its richness, due to the healthy presence of the bass range. At ease with a jazz quartet as well as with a philharmonic orchestra, it does not miss any detail and it handles powerful transients effortlessly.

The sonic scene is filled and uniform and, depending on the recording approach and orchestras, can give the impression that the instruments are there in front of you, playing next to each other, but distinctly. The sound does not “jump” from one place to the other. The instruments can be heard as in a natural, wide angle space. People who are unaware of the output transformer’s secret notice a difference with the characteristics of the sounds they are used to, but they cannot describe the differ-

ence, still being positively impressed. No doubt that an SC-SCC OPT is worth a try, not only for the special sound, but also for its advantages (lower price, less than half of the two SE transformers it replaces, improved simplicity, and efficiency)⁶⁻⁷.

The curve in **Fig. 2** pertains to a prototype of an amp that has been duplicated eight times by eight young students (13- to 15-years old) in Italy (**Photo 3**), under the supervision of my friend Antonio Tucci⁸. This result proves that, with this transformer, the SE topology compares favorably to the push-pull, still keeping its appreciable character.

4 × SC-SCC-SET DEVICE

After my initial enthusiasm, a doubt crept into my mind. Was it possible to achieve further improvements?

Consider **Fig. 3** and **Photo 4**. As explained in the AES 122nd Convention Paper #7151 (Vienna, Austria, May 2007), the number of primary/secondary windings’ bobbins is increased (two for each channel in the basic setup, with the possibility of more), connected in such a way that the currents that cross them lead to an automatic compensation of the magnetism, reducing the same very close to zero, independently of the idle DC intensities flowing to the valves.

The advantage of this layout is that, with proper connections, it does not require an external direct current balancing device—everything is done automatically *inside the transformer itself*. It also eliminates the need for matching parallel valves.

Figure 4 shows that, in every branch, the direction of the DC flow is the same,

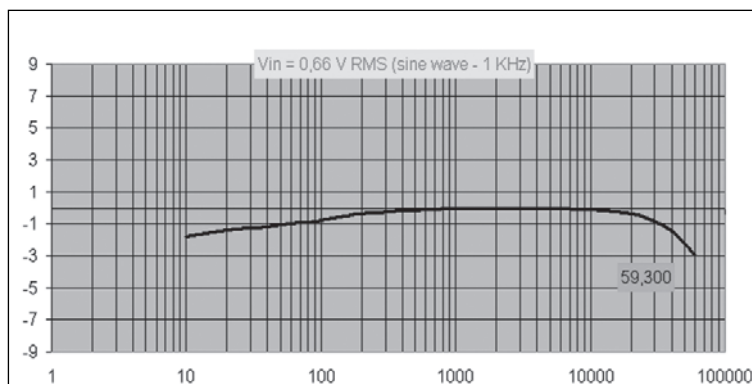


FIGURE 2: Frequency range of the SC-SCC-SET.

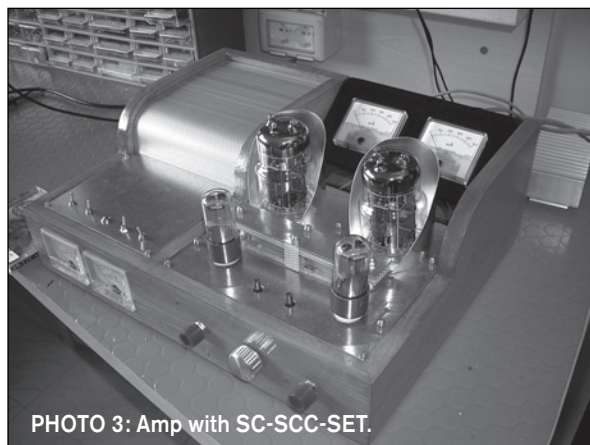
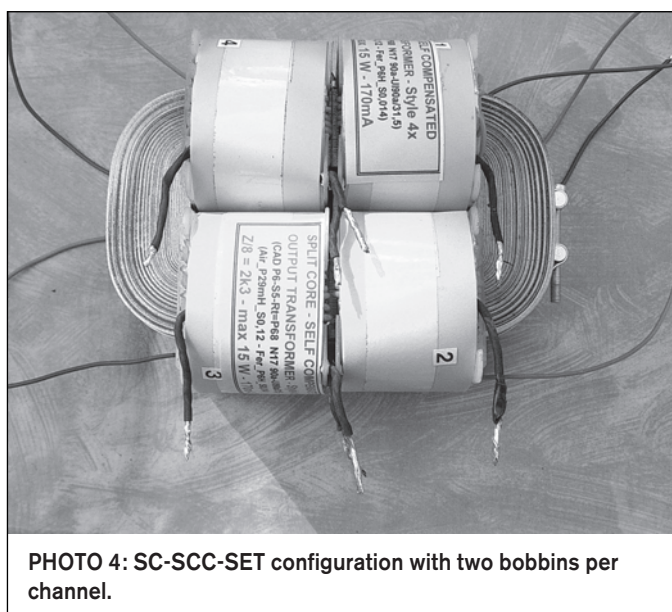
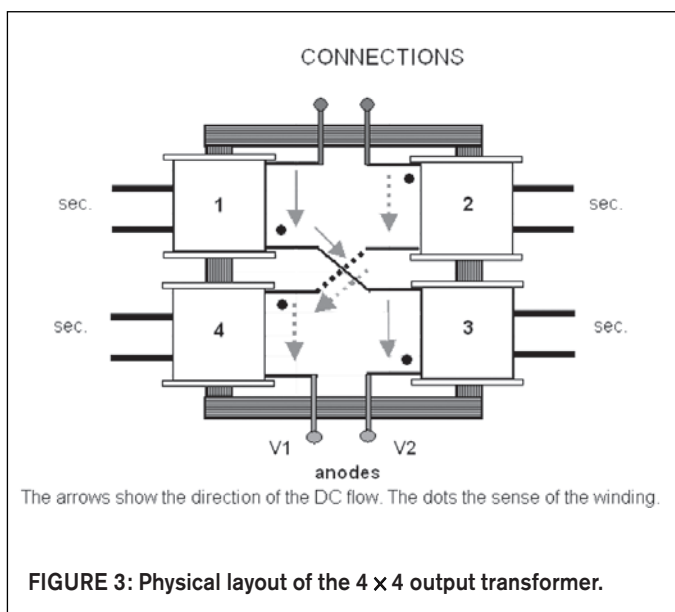


PHOTO 3: Amp with SC-SCC-SET.



but the winding sense of each primary is opposed to the other and that the primary coil's terminals are connected in a crossed way, from one branch to the other.

The power supply current flows in the primaries 1 and 3, connected in series, but on different branches of the magnetic circuit, until it reaches the anode of valve 2. Primaries 2 and 4 are in the same situation as the other valve. Four fields are produced by the DC, that, under certain conditions, fight each other. As a result, the overall magnetization is close to zero.

The simplest case is when all windings have the same number of turns (N1-N2-N3-N4) as in cases 1 to 5 in Fig. 4. No matter how strong or weak the anode currents are—their relative levels—the cancellation of the magnetization takes place.

In cases 6 to 9 of Fig. 4, as a result of the four fields strength (aiding and opposed) produced by the windings operating in the magnetic path, whose algebraic sum is zero⁹, the magnetization of the core is also cancelled.

Some other alternatives—with the number of turns of the primaries equivalent in pairs—also produce the same result (cases 6 to 9). However, some combinations do not work, such as cases 10 and 11. Thus, the DC offset can be ensured in the favorable cases.

With regard to the alternating current, due to the fact that valve 1 and valve 2 are modulated with signals 180°

apart, the primary windings concerned operating in series aiding condition are not antagonist and they work freely. An important point is that if *one transformer and a pair of valves (or more) are used for each channel, no [crosstalk] takes place*. The 4x-SC-SCC is currently undergoing extensive tests in order to check that the

expectations are fulfilled also as regards the sound quality.

ONE OPT USED FOR BOTH CHANNELS

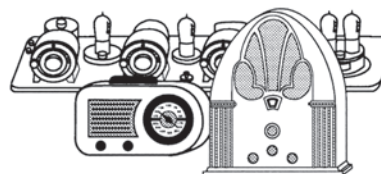
In this case, the [crosstalk] is present, but as with the SC-SCC, it concerns only the low frequencies with a negli-

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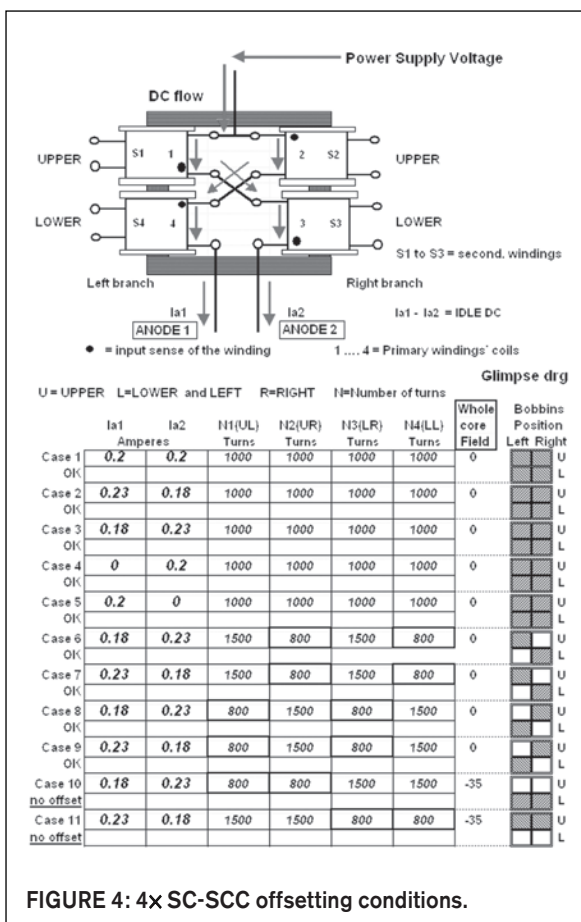


FIGURE 4: 4x SC-SCC offsetting conditions.

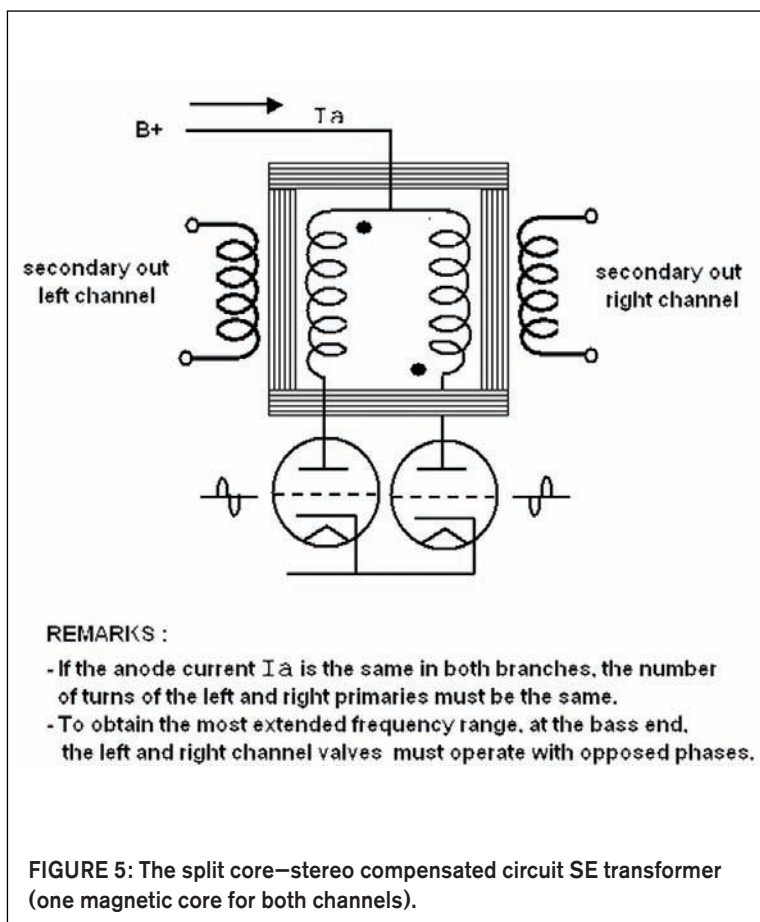


FIGURE 5: The split core—stereo compensated circuit SE transformer (one magnetic core for both channels).

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gible impact on the stereophonic effect. However, it would be possible to control the [crosstalk] by shunting the upper blocks 1 and 2 with suitable capacitors, for instance, as is done in the SC-OPT to the degree (depending upon the value of the capacitor) that it does not cause a collapse of the other primaries' inductances.

With this solution the [crosstalk] is moved to lower frequencies. Readers with comments or questions about the behavior of the 4X SC-SCC are invited to e-mail Ari.polisois@wanadoo.fr, P.touzelet@orange.fr, lpolisois@yahoo.it, or, preferably, A2Belectronic@wanadoo.fr.

UPCOMING DEVELOPMENTS

Tests are being conducted on bench units to utilize fully and advantageously the principles discovered with the former versions (SC-OPT, SC-SCC-SET, and 4x SC-SCC) in order to:

- Reduce the [crosstalk] (when the 4x is used for *both* channels on a common core). As mentioned, in a layout using *one 4x transformer per channel*, with parallel valves, obviously, there is

no [crosstalk]. This corresponds to a push-pull having SE sound character, but it does not require an auxiliary offsetting circuit.

- b) Extend the high-frequency range beyond the current limits, the goals being 5-10Hz and 70-80kHz at -3dB.
- c) Obtain the offset and balancing conditions automatically, without the use of *any external device*.
- d) Improve the simplicity of the single- or double-ended amps by integrating the loudspeakers' crossover units into the output transformer, for instance.

This feature, which took years of research, is practically concluded and will be the subject of a presentation at a coming AES convention.

Today, the SE topology has a new life and the future is promising. I have been striving to give DIY audiophiles a wider choice of transformers (output or interstage) they can fit to their amps, according to their objectives. On behalf of engineers Mariani, Touzelet, and myself, many thanks for your interest in this field. Please do not hesitate to contact us with any questions or comments. *aX*

REFERENCES

1. Why a gap in a SE OPT (or choke)? Just because you must avoid the saturation of the core due to an excess of permeability, which is an improvement on the core magnetization capacity obtained by inserting a ferrous or similar material inside the windings. However, permeability has a particular characteristic that you might consider a defect: It changes with the field strength, and when the latter is relatively too strong, it starts lowering, due to the saturation of the core. When this happens in an audio transformer, the sound becomes greatly distorted.

2. In practice, this is absolutely true as far as the low-frequency end is concerned, but you still get a weak one on the high-frequency side.

3. In the case of opposing coils in series, such as the primary and tertiary of the SC-OPT, the formula that applies is: L_t (total inductance) = $L_p + L_t - 2M$, where L_p is the primary inductance, L_t the tertiary's, and M the resulting mutual inductance according to their coefficient of coupling, calculated as follows: $M = k\sqrt{(L_p \cdot L_t)}$, where k is the coupling factor (equal to 1, if the number of lines of force passing through one

coil is exactly the same as those passing in the other, which is practically impossible) to less than 1.

4. Because of the load and other actual working conditions, the effective inductance is different.

5. There are many valves having a low internal resistance (6C33 - 6336 - 6080 - 6550 - KT, and so on, triode connected) that can eventually be reduced by putting several tubes in parallel. Until now used for other applications, such as voltage regulators, this family of valves comes, luckily, back to active life. Most of them do not require extremely high anode voltages.

6. This novel device was presented at the AES 120th convention, held in Paris in May 2006 (paper 6831). More details on the transposing of the low frequencies are given in the above paper.

7. To better illustrate this result, the OPT (bi-channel) in **Photo 2** replaces two OPTs 50% bigger in size.

8. Thanks to Antonio, these kids have discovered the sound of a valve amp and some of them will surely become *audioXpress* readers.

9. The electric fields strength of each winding, (producing aiding or opposing magnetism) result from the product: $-N$ (number of turns) $\times I_a$ (DC anode current).

ACKNOWLEDGMENTS

The 4x SC-SCC transformer is the latest step in a program that started about ten years ago to improve the sound of valve audio amplifiers.

With the invaluable support of engineers Menno van der Veen and Pierre Touzelet as well as, most recently, Giovanni Mariani of Graaf, Italy, sensible improvements have been achieved that led to the application of several patents, five of which have been granted by the French INPI (Institut National de la Propriété Industrielle).

I am also grateful to Edward T. Dell, Jr., Editor of *Glass Audio* and *audioXpress* magazines, for publishing my first article on the subject ("A Universal Phase Splitter," p. 32, *GA* 1/00) and subsequent ones, and also to the staff of the Italian magazine *Costruire HI FI*, for printing several other articles, and, recently, also to Elektuur and Elektor Deutschland of the international periodical of electronics, *Elektor*. Finally, I would like to mention engineer Menno van der Veen (worldwide known for his works in the audio reproduction field), whose support has been of the utmost importance.

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The Pragmatic Class D Amplifier

With the "best-kept secret" in audiodom, you can construct your own Class D amp.

Class D amplifiers are gaining in popularity these days. Hypex and Tripath offer great products for DIY enthusiasts. Class D has some advantages over class A, AB, B, or tubes—one of which is its high efficiency. At normal output power there is no need for heatsinks, simply because heat dissipation is almost nonexistent. At higher output only a rather small heatsink is required. Another advantage is the amp's physical size and weight.

One of the very early class D designs was the Sinclair X-10 amplifier, which was sold in the early 60s as both a kit and as a complete product. This power amplifier yielded 10W of output power but, in reality, produced only a couple of watts. Those who successfully built the kit reported that the amplifier radiated like a radio transmitter.

In the 80s I repaired high-end audio for a specialized audio shop. Many rare amps passed through my hands, including the heavily-built Sony TA-N88 audio power amplifier. This model was one of the first commercial amplifiers to use class D pulse width modulation (PWM) technology, along with the use of a switching power supply.

At that time I did not really pay attention to this. My interests in class D started in 2000 after reading an article on Tripath's class D technology. I even ordered some modules. However, for some reason, I never got around to using them in a project. Last year I listened to Tripath and Hypex UcD based class D amps and built a class D

amplifier using B&O's ICEPower technology based on the 200ASC board as used by Cary Audio and Jeff Rowland.

Searching eBay is often part of my daily routine, and one day I came across a tiny amplifier that was completely unknown to me. Even the company's name *Pragmatic* didn't ring a bell, but it piqued my interest nonetheless. The ad listed the following features: class D, low distortion of typical 0.01% to 0.05%, no ripple in the passband, and audiophile sound.

Searching for "pragmatic amplifier" in the Google search engine, I was disappointed by the number of related hits. Either this amp was a forgotten treasure or just not a good item, but I decided to take a risk anyway. I purchased three of the ten pairs of amplifiers the eBay auction offered. The final price was \$115 US—a good deal—and I later regretted that I did not buy all ten pairs.

THE PRAGMATIC CLASS D MODULE

The modules (**Photo 1**), measuring only $8.7 \times 5.1 \times 2.0\text{cm}$ (L $\times$ W $\times$ H), use

SMD components throughout. Because the modules use an EMI/RFI PCB enclosure, you cannot determine component locations or any device numbers. So the amplifier technique and circuit design remain a mystery.

Pragmatic Communications Systems has been selling these class D products since 1996. Prasanna Shah, who designed this class D module, says the following: "One reason for our anonymity is that we have not been able to spend money on magazine advertisements. We have invested money in technology and performance rather than advertisements. We have sold our amplifiers to lovers of high fidelity all around the globe (Sweden, Norway, Denmark, Switzerland, UK, Germany, the Netherlands, Belgium, Italy, France, Hong Kong, Singapore, Australia, USA), and most of the sales have been word-of-mouth referrals."

Shah continues: "The Pragmatic amplifiers have been tested against very expensive Mark Levinson amplifiers at Madrigal Audio Labs' listening station. . . at listening sound levels they could not tell the difference between the Pragmatic amplifier and Mark Levinson amplifiers, except that the Mark Levinson amplifiers were an order of magnitude more expensive, weighed more than 100 times more, and dissipated tremendous amount of heat energy into the room."

THE PRACTICAL REALIZATION

These tiny amps are very easy to use. Each amp comes as a complete working module, so

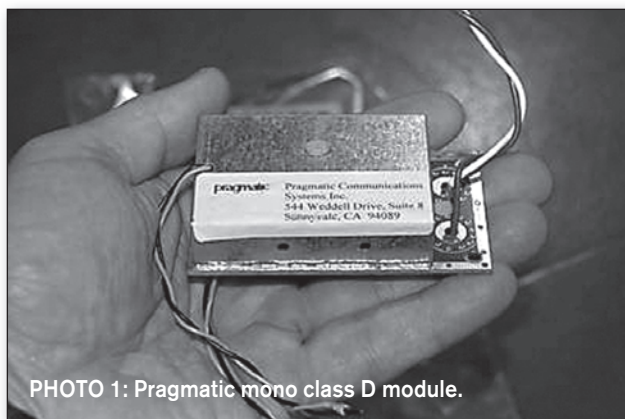


PHOTO 1: Pragmatic mono class D module.

there is no need for component soldering or stuffing a circuit board using SMD devices and no need for adjustments.

Each amp comes with three pairs of wires: power supply (red and black), line input (orange and black), and a speaker pair (positive output marked on PCB). All you need is to connect the 12V DC PSU (at least 5A per channel) and a fine line input signal (plus, of course, speakers) to have a full power amp.

I first set up a laboratory power supply and two Pragmatic modules, with no enclosure—just some wires laying around on the floor. All worked without hum or hiss.

I built three amplifiers. Amp One (Photos 2–4) uses an ABS enclosure and a Dell SX280 miniPC 12V DC switching power supply, both of which I bought for not too much at a radio flea market. The standby function of the PSU is used for switching on and

off, with the speakers not suffering from any dangerous pops or clicks. I built this amp for a total cost of less than 70 euros (about \$65 US).

Because the first amplifier performed so well, I built another, better-looking, amp. Amp Two (Photos 5–7) uses a switching power supply by XP Power (Farnell 1337158), which was a leftover from a former project. A PSU by Mean Well, such as the SP-150-12 or SP-100-12, would be a

better choice, pricewise, but the XP PSU fits the used enclosure perfectly, and the specs are great.

Amp Two uses a Sansui VU meter, which I also bought at a radio flea market. The VU meters use a dedicated DIY circuit board by JLM Audio Australia, and these are directly mounted on the VU meters. This circuitry is powered by the 12V DC from the switching power supply. The 12V scale lighting uses a flexible white LED strip.

Because I still had one amplifier pair left, I built a third amplifier (Photos 8–10). This amp uses a conventional power supply with a toroidal mains transformer and a voltage regulator; I used one single 120VA transformer and a LM338 voltage regulator for each channel. The LM338 comes in TO-3 housing and is good for 5A.

I also purchased all parts for this amp at a radio flea market. The LM338 goes for 1.50 euros and the 120VA mains transformer for 20 euros. I bought the ABS enclosure for just 15 euros.

Use of a switching power supply is interesting, not only because of its price and compact size, but also because this kind of power supply works on a similar principle as that of a class D (or pulse width modulation) ampli-



PHOTO 2: Prototype Amp One.

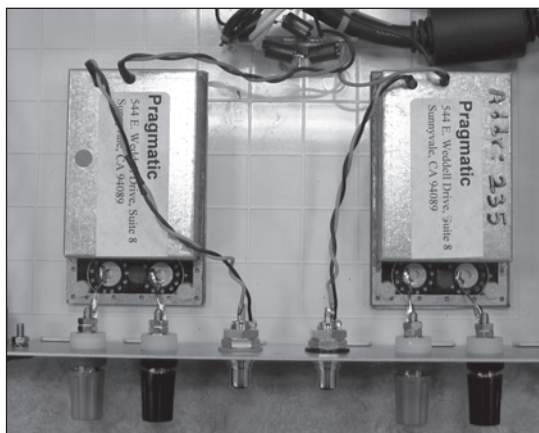


PHOTO 3: Overview inside Amp One.

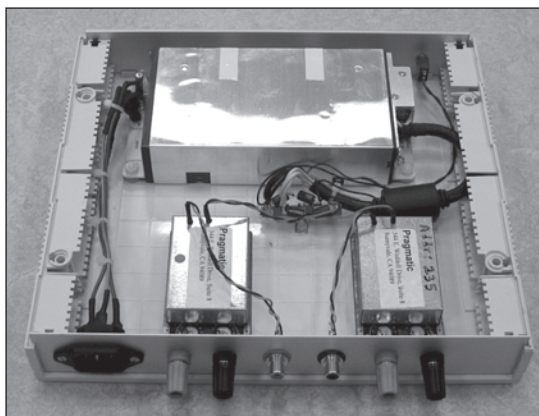


PHOTO 4: Amp One.

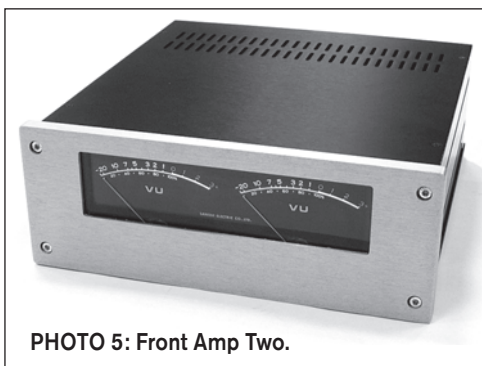


PHOTO 5: Front Amp Two.

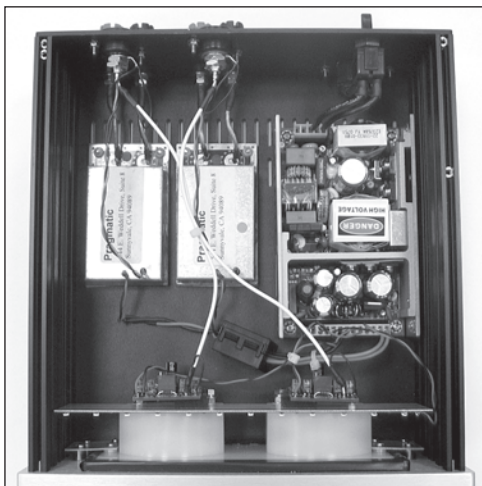


PHOTO 6: Amp Two overview.

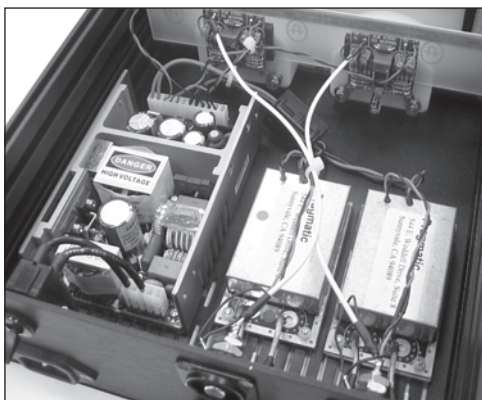


PHOTO 7: Amp Two inside overview.

fier except that the reference voltage of a PWM amplifier is the varying audio signal instead of a fixed voltage as in a switching power supply.

TECHNICAL ASPECTS

Its monoblock design provides extremely clean state-of-the-art power. The SIWA-248 unit accepts any volume-controlled line level input and uses a 12V-15V DC power supply. It has continuous power of 24W and a peak power of 60W. Advanced protection circuitry guards against system damage from overload, overheating, or short circuits.

COMPANY SPECS

Power: 12-15V DC
Continuous power: 24W per channel
Peak Power: 60W
Signal-to-noise ratio: >100dB
Dynamic range: >100dB
Frequency response: 20Hz to 20kHz

Applications: Features and Benefits

- Compact monoblock amplifier
- Residential or commercial audio amplification compact design
- High efficiency
- High clock frequency
- Very linear gain and phase character-

istics

- Flat frequency response from 10Hz to 50kHz
- No ripple in the passband
- Very high signal-to-noise ratio
- Very high dynamic range
- 24W (peak 60W)
- Only 12V DC power required
- Module works with 4 and 8Ω loudspeakers
- Very low and uniform distortion on all frequencies and very high signal-to-noise ratio and dynamic range.

SOUND IMPRESSIONS

To evaluate the sound performance, I used a large Tannoy D700 and a small-size Dynaudio. The Dynaudio is based on a T330D Esotar tweeter and 17WXL04 low mid unit, both by Dynaudio. At first you notice the ease with which music is reproduced. Sound

TABLE 1

How to connect the Pragmatic module

power supply	+12V DC = red.....	ground = black
line input	signal in = orange.....	ground = black
speakers	- output not marked on PCB.....	+ output is marked on PCB

Please note: the amplifier circuitry is designed in such a way that you can *not* connect the negative loudspeaker output to ground.

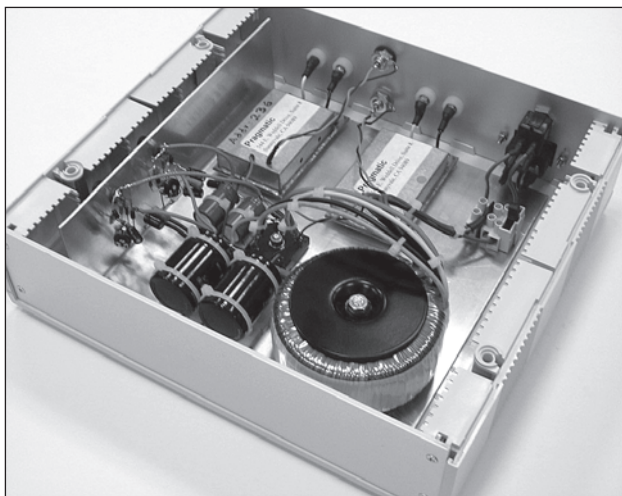


PHOTO 8: Front Amp Three.

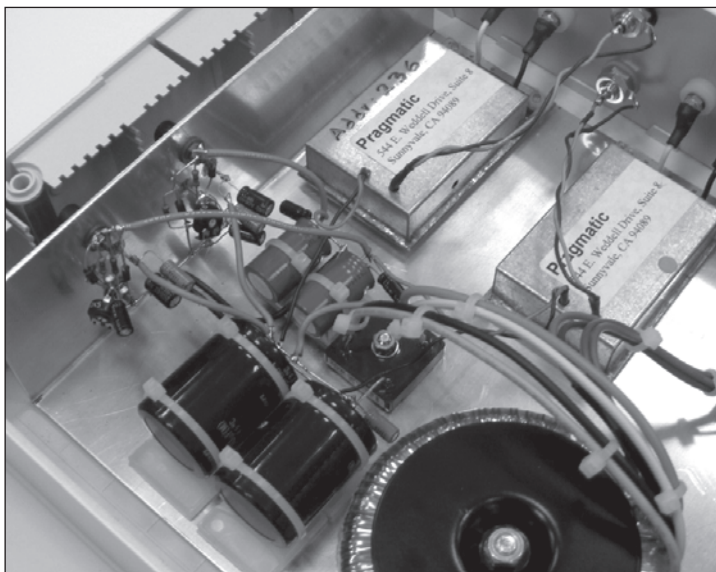


PHOTO 10: Amp Three inside.

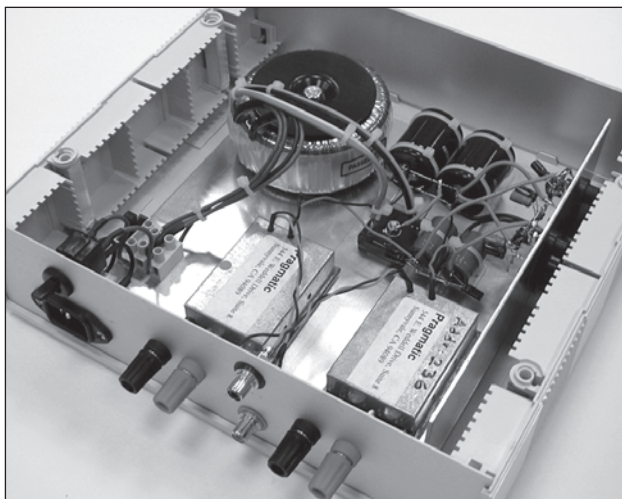


PHOTO 9: Amp Three overview.



PHOTO 11: The Pragmatic SAA-224 stereo standalone amplifier.

is very clean, dynamic, and precise with a beautiful soundstage. The Dynaudio's 4Ω load and 84dB sensitivity are no problem for these low power modules; however, the Tannoy's 94dB sensitivity allows *headbanging* levels.

Big Band sound, such as in Gerald Wilson's *In My Time* CD, drives both speakers to a very high level. Music is reproduced with great authority and ease, rhythm and pace are excellent, and it is also free of any distortion artifacts. Also, at low and moderate levels all music can be felt and heard; its dynamic behavior is excellent.

Chet Baker's *Great Last Concert* (Enja) sounds impressive, with very good tonal balance and great atmosphere/ambience. Piano solos such as classical pianist Arcadi Volodos' *Live in Carnegie Hall* (Sony) are wonderfully reproduced; the Rachmaninoff pieces are extraordinary and captured by this amp in the best way. Audience, the bench, recording/hall noises/ambience—this amp offers high-resolution sound with impact and drive.

Another great piano solo recording is Marc-André Hamelin's Nikolaj Medtner four-CD box (Hyperion). This wonderful recording by Tony Faulkner is *real*, and, with these Pragmatic amplifiers, the Dynaudios fade away. The low and high units melt together as the left and right speaker and results in a three-dimensional sound.

The music of Russian composer Nikolaj Medtner results in a wonderful journey. Natalie Cole's voice on *Take a Look* is full-bodied with lots of air around it; the differences in recording of the different tracks can easily be detected. Fusion jazz such as Lee Ritenour's *Wes Bound* (GRP) will make you smile. Such big sound from such a tiny amplifier. Also noticeable is that the higher frequencies don't show any rolloff. The Keith Jarrett Trio *Live at the Blue Note* six-CD set is a real treasure; I guess these Pragmatic modules already knew this.

Yes, I am aware that I am overly enthusiastic. These modules are really fun and capable of reproducing music in all its pros and cons.

EPILOGUE

These interesting Pragmatic modules are very quiet, even in my simple test

setup. Musicality is enormous; they can rock and shake. Intimacy, imaging, soundstaging, and high resolution are its strong points. I guess these modules are probably the best-kept secret in audiophile circles. And remember: this class D design dates from 1996, which is remarkable because class D back then was not in the picture nor as developed as today.

Just before publishing deadline, I found out that the Pragmatic Company never sold these modules separately, so

the modules I bought on eBay were likely taken from a Pragmatic SIWA in-wall configuration. However, the Pragmatic Company sells modules as standard products. Check out their website for the standalone amplifiers such as the mono MBA-124 and the stereo SAA-224 amplifier (**Photo 11**), as well as in-wall configurations such as the SIWA-224; the first two are beautifully made standalone units with external power supplies. A 12-channel version is also available. **ax**

ABOUT THE DESIGNER

Prasanna Shah is the Chief Pragmatist and Thinker at Pragmatic Communications Systems, Inc. He earned his M.S. degree in Electrical Engineering from Stanford University and B.S. degree in Electrical Engineering from Santa Clara University, Santa Clara, Calif. He is a member of Tau Beta Pi and Eta Kappa Nu Electrical Engineering Honor Societies. He has almost three decades of experience working on semiconductors for analog and digital audio, video, cellular telephones, wireless communications, fiber optic communications, data communications, data acquisition, and industrial controls in the Silicon Valley. He holds one patent on power line communications, several patents on audio, video, and wireless multimedia systems, and many more patents pending in the area of industrial automation, health care patient interaction, and power generation/energy storage technologies. An avid audiophile and music lover, in his free time he likes to apply his pragmatic thinking to solve complex problems.

ABOUT THE COMPANY

Pragmatic Communications Systems, Inc. (www.wireless-experts.com and www.pragmatic1.com) is a Silicon Valley company established in 1994 to design, develop, and produce a variety of innovative products for wireless audio, video, and data communications. The majority of the company's products until 1997 were primarily designed for industrial and commer-

cial applications. The success of these products has led to a strong and dynamic technology base, which has been applied to a new generation of products such as CATSTM and others such as the DMSTM, TrueMusicTM wireless hi-fi system and TrueViewTM wireless video system. The company's philosophy is to use its technologies to solve problems that have gone unresolved for decades.

Further products include high-performance, low-profile line-array speakers, high-efficiency audio amplifiers (monoblock, stereo, multi-channel, in-wall), high-performance TV/CATV/FM tuners, category 5 (CAT5) audio/video transmission and cross-point matrix switchers, hard disk music servers, ambient audio levelling systems, microphones, and IR detectors.

PADSTM—Multi-room audio distribution system on a single category 5 wire.

DMSTM—Multi-source, multi-zone distributed music system on a single category 5 wire.

DAVESTM 4.43—Distributed Audio Video Entertainment System.

CATSTM—Multi-source, multi-zone DVD quality audio/video signal distribution and IR control on a single category 5 wire.

CATSTM HD—High definition audio/video on category 5 wire.

The Class D tiny power amplifiers are part of the so-called Hide-AwayTM Series (SIWA).

All are trademarks of the Pragmatic Company. ■

Time-Adjusted Passive Crossover Systems

This Australian study on passive crossovers is an interesting adjunct to Koonce's work published last summer.

I would like to thank G.R. Koonce for his very informative articles¹ on linear phase passive crossovers. In the past 18 months, John² and I have spent some time analyzing passive time-adjusted crossovers proposed by Yamanaka and Baekgaard³; both require a “filler” speaker. Koonce's investigation concentrates on the acoustic characteristics of loudspeaker systems with linear phase crossovers. In contrast, we study the non-ideal electrical properties of filter components and speaker impedance, using PSpice (student version 9.1) to evaluate the worst-case conditions of various time-adjusted crossovers. This study appears to stand as a supplement to Koonce's excellent articles.

We use formulas by Yamanaka and Baekgaard to determine crossover components, which are accurate to three digits, and allow them to have tolerance. We assume that speakers are matched with zobel and/or notch networks such that the impedance is 8Ω and resistive. Again, we allow a tolerance to be added to individual speakers.

As a comparison, we also include a first-order system, using worst-case and sensitivity analysis to evaluate the performance of the different designs in both the time domain and frequency domain. In frequency analysis we use the phase of the summed output as the output variable. In time-domain analysis the magnitude is the output variable.

We use the following conditions/assumptions:

- Nominal crossover component values are calculated to the third significant figure with all speakers made equivalent to an 8Ω resistor.
- Component tolerance and speaker tolerance are assigned to be +1% or +10%.
- Crossover frequency is fixed at 2kHz in all cases.
- Acoustic properties of speakers are ideal; for example, all speakers have

the same acoustic center, and each acts like a perfect point source.

- In frequency-domain analysis both the magnitude and phase responses are considered. The worst case is derived when the phase plot produces the biggest discrepancy from the nominal.
- In time-domain analysis a 500Hz square wave is used as the test signal. The worst-case is derived when the magnitude produces the biggest discrepancy from the nominal.
- Component values are similar but not

identical to those in reference 2.

- PSpice does not use a true optimization algorithm. However, worst-case simulation results should show that the *actual* worst-case, if different, could only be worse.

RESULTS

Figures 1A-1C show the (worst-case) frequency response (Bode plots) of the second-order Yamanaka system with various component tolerances. In an ideal world the magnitude plot is a horizontal line of 0dB and the phase plot is another

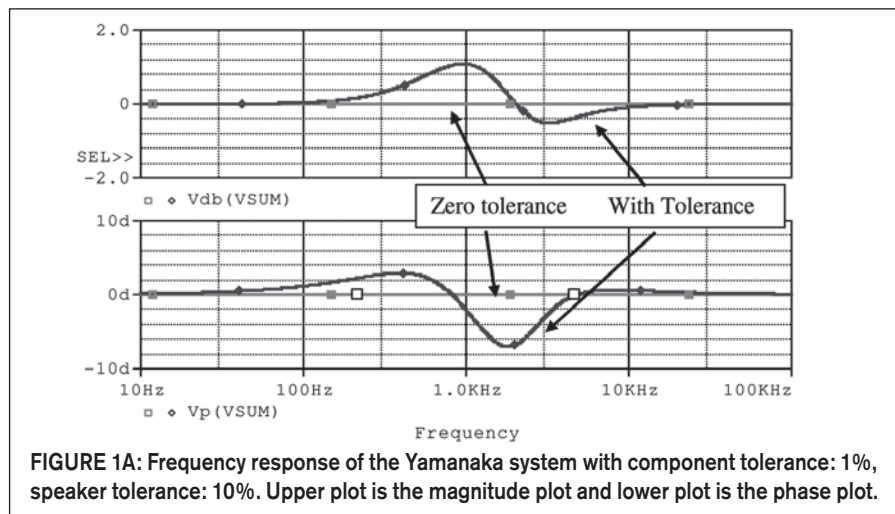


FIGURE 1A: Frequency response of the Yamanaka system with component tolerance: 1%, speaker tolerance: 10%. Upper plot is the magnitude plot and lower plot is the phase plot.

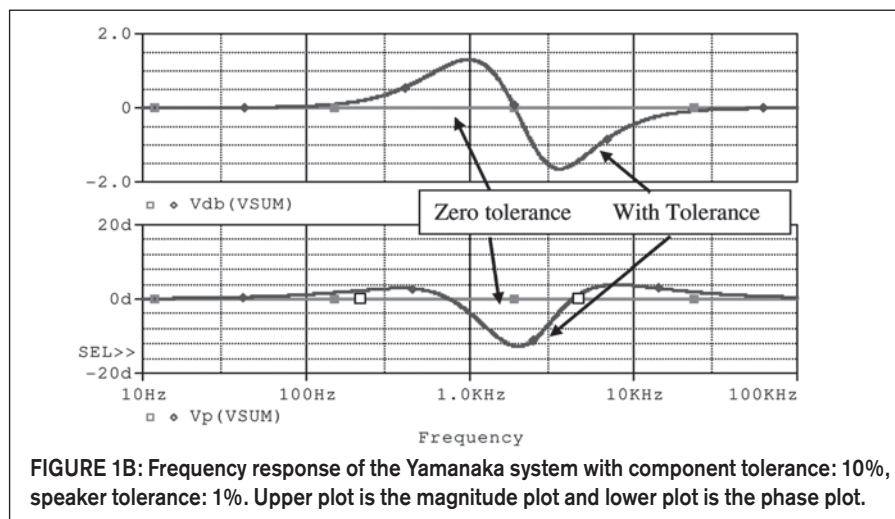


FIGURE 1B: Frequency response of the Yamanaka system with component tolerance: 10%, speaker tolerance: 1%. Upper plot is the magnitude plot and lower plot is the phase plot.

horizontal line with 0°. **Figure 2** shows the response from a first-order system. **Figure 3** shows the results from a third-order Baekgaard system. All are affected by component tolerance with the third-order Baekgaard system being the worst.

Another set of worst-case results can be generated when the magnitude response dips instead of peaks. **Figures 4A and 4B** show the (worst-case) time response of the Yamanaka system with two different tolerance settings. In an ideal case the original 1V 500Hz square wave can be reproduced. **Figure 5** shows the relatively poor time-response from the Baekgaard system. For the second-order system, component and speaker tolerance must be kept within 10% if frequency response abnormalities are kept within $\pm 2\text{dB}$ and $\pm 20^\circ$.

CONCLUSION

From the above samples and John's work², I can conclude that

- Speaker units must be tamed to be a constant resistance, probably within 10% tolerance or better.
- Crossover components must be selected (almost) exactly according to the designed values and with the best tolerance. The importance of tolerance increases with system order.
- The third-order Baekgaard system is around two times worse than the second-order Yamanaka system with the same component/speaker tolerance.
- The second-order Yamanaka system is close to the first-order system with component/speaker tolerance of 10% if the former has 1% crossover components.
- The first-order system is not immune to component and speaker tolerance. It is more tolerant. However, the shadow sloping of the filters is not good for most speakers.
- The third-order Baekgaard system demands very tight component tolerance

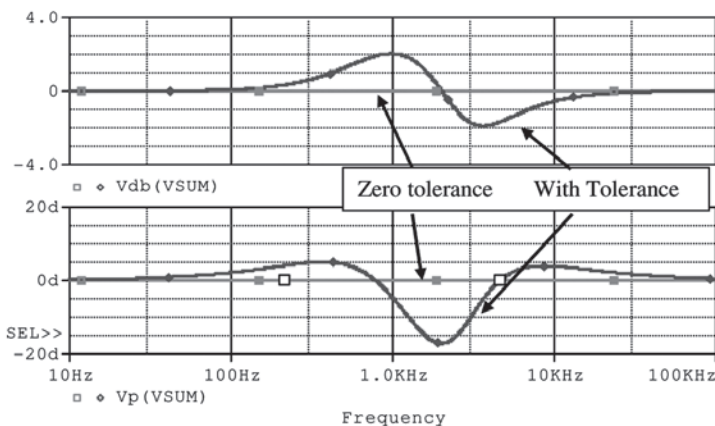


FIGURE 1C: Frequency response of the Yamanaka system with component tolerance: 10%, speaker tolerance: 10%. Upper plot is the magnitude plot and lower plot is the phase plot.

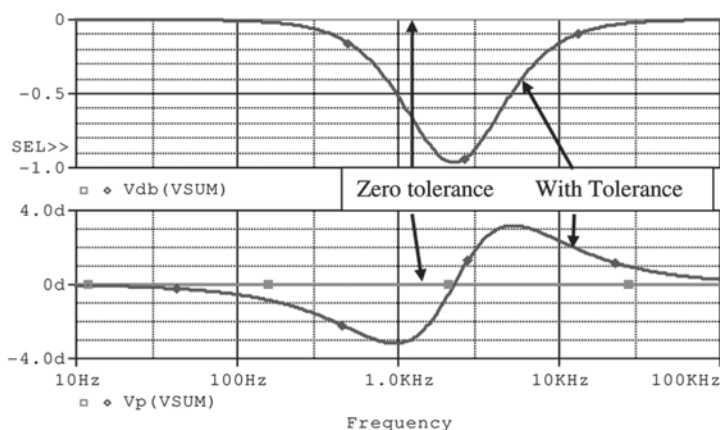


FIGURE 2: Frequency response of the first-order system with component tolerance: 10%, speaker tolerance: 10%. Upper plot is the magnitude plot and lower plot is the phase plot.



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and does not appear to be practical.

- The second-order Yamanaka system seems to be the only viable solution. (This conclusion is quite different from Koonce's listening experience.)

The performance of a complete speaker system is undoubtedly affected by non-ideal acoustic and electrical conditions, so it appears that a perfectly time-adjusted passive loudspeaker system is unlikely to be practical in the real world. Perhaps an active system, and possibly with the aid of digital techniques⁴, is the only way to move closer toward this objective. *aX*

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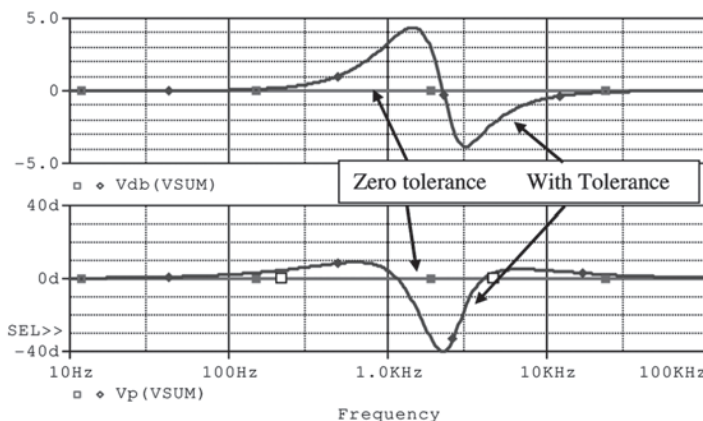


FIGURE 3: Frequency response of the Baekgaard system with component tolerance: 10%, speaker tolerance: 10%. Upper plot is the magnitude plot and lower plot is the phase plot.

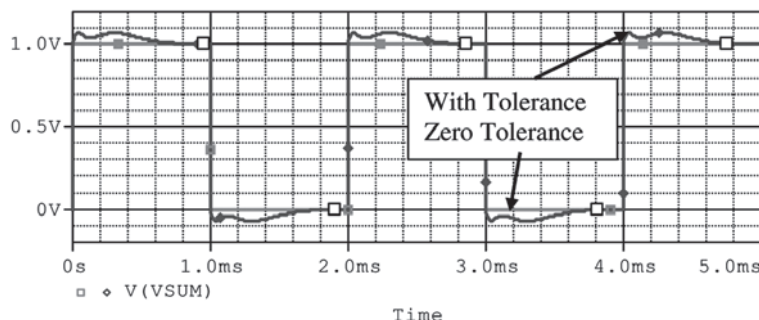


FIGURE 4A: Time response of the Yamanaka system with component tolerance: 1%, speaker tolerance: 10%.

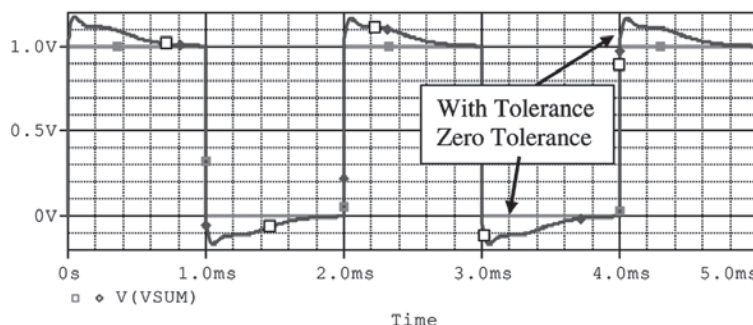


FIGURE 4B: Time response of the Yamanaka system with component tolerance: 10%, speaker tolerance: 10%.

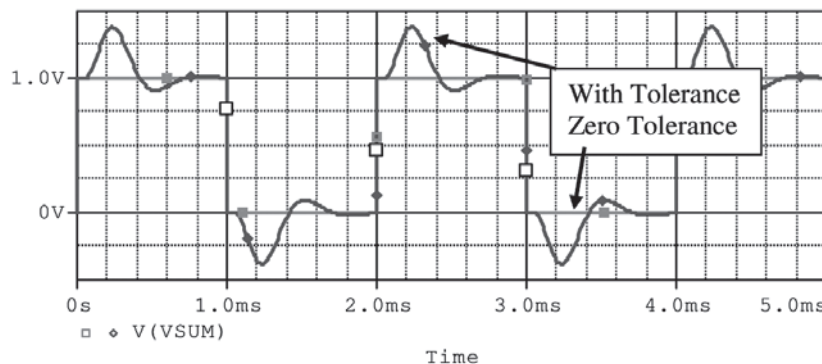


FIGURE 5: Time response of the Baekgaard system with component tolerance: 10%, speaker tolerance: 10%.

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Design and Measurement Of a Dipole Microphone

Tone burst techniques, gated measurement, analog computing, and imaginary numbers.

I've been interested in microphones for some time, especially those used to record music in concert and studio settings. More recently, however, I've become especially interested in how various microphones differ, how to measure and compare those differences, and what design trade-offs go into the development of a new microphone. This article examines a particular kind of microphone, the acoustic dipole or "gradient" microphone and its design parameters¹.

First I'll describe its operation mathematically in enough detail to predict frequency and polar response characteristics in a spreadsheet. Next, I'll discuss the techniques needed to obtain accurate measurements of microphone frequency and polar response, using a computer sound card and some simple software programs. The spreadsheet and computer source code are posted online (www.audioXpress.com) for readers to download. Finally, I'll show the construction of a prototype dipole microphone using two omnidirectional electret condenser microphone cartridges.

My goal here is not to produce a design that's better or different from existing designs, but rather to gain a more complete understanding of what goes into microphone design and measurement. This way, we can all be a little more confident about making the best use of the microphones that are available to us.

MICROPHONE TYPES

The simplest microphone is the omnidirectional pressure microphone, which senses instantaneous pressure at a point in space. Pressure doesn't have a direction, so an ideal pressure microphone responds equally to sounds from all di-

rections. Sounds arriving simultaneously from various directions simply add up to a single pressure value at the location of the microphone. Note, however, that the addition is instantaneous, so phase matters when doing the summation. Two sounds at the same frequency could add up constructively or destructively, depending on their phase relationship.

Now suppose you want to design a microphone that picks up sounds along a particular axis, while ignoring sounds along some other axis. A simple technique is to place two pressure microphones of equal sensitivity near each other, and then obtain the difference between the two signals. Sound waves traveling along the axis of the dipole (that is, along a line through the acoustic centers of both microphones) will arrive at slightly different times. While the amplitudes will be roughly the same, there will be a phase shift between the two signals so they don't completely cancel. This, then, is the "on-axis" signal of the directional microphone.

Sound waves arriving from any direction in a plane perpendicular to the dipole axis will arrive at both capsules at the same time, so there will be no phase shift between them, allowing for complete cancellation of the signal. This is the "off-axis" response, for what you now see is a bidirectional microphone, having a figure-eight polar response pattern. The combined dipole microphone now responds to the gradient of pressure instead of pressure itself, and has a preferred axis for pickup.

But what if you want to pick up sounds arriving from one direction along the dipole axis while rejecting sounds arriving from the opposite direction? It turns out that the dipole gradient micro-

phone can distinguish front from back, in the sense that sound from the front gives a gradient signal which is in phase with the omni response, while sound from the back gives a gradient signal which is opposite in phase to the omni response. Assuming the omni and on-axis gradient responses have the same sensitivity, they may be added together to obtain a response that is twice the omni response on-axis from the front, equal to the omni response from the side, and null (complete cancellation) along the dipole axis from the rear.

The dipole microphone is now giving what is commonly called a "first-order cardioid" polar response pattern, preferring sounds from one direction while rejecting sounds from the opposite direction. This microphone type is widely used—both in studios and in live recording—because it allows direct sound from a performer or instrument to be emphasized, while reducing ambient and reverberant sounds².

Most commercial microphones having a first-order cardioid response actually have just one diaphragm and one electrical output. The gradient is obtained by having two ports to accept sound, which is then routed to the front and back of the diaphragm with appropriate time delays and other acoustic elements. This is a form of analog computer, implemented with air passages. The best way to see how this works is to take one apart, although you may ruin it in the process.

For my work here, I'll use two separate omnidirectional capsules and then develop the various polar patterns using an analog computer implemented with gain stages, adders, and an integrator. Clearly, the computations could also be implemented in a digital computer, but real-

time digital signal processing is beyond the scope of this article³.

MATHEMATICAL DESCRIPTION

Now let's work out the math needed to describe the polar and frequency response for the dipole microphone. First you need some assumptions, so things don't get completely out of hand. I'll assume the dipole is made of two ideal pressure microphones having identical sensitivity, and flat frequency and phase response over the range of interest. I'll limit range of interest to 100Hz to 10kHz, for reasons that will become clear as we go along. I'll assume free-field conditions, considering only the direct sound from an ideal point source, with no reflections or reverberations, and, of course, no ambient noise.

Sounds from the ideal point source will propagate as spherical waves, so assuming no transmission loss in the medium, acoustic pressure will fall as the inverse of distance from the source. Acoustic power, of course, falls as the inverse of distance squared, but the microphone responds only to pressure, not power, thus the $1/r$ dependence. The central axis of the dipole will, for my purposes, always remain in a plane that includes the point source.

Now I write the wave equation giving instantaneous pressure p at distance x and time t , for a steady-state sound of frequency ω in units of radians per second⁴:

$$p(x, t) = \frac{x_0}{x} e^{i(kx - \omega t)} \quad (1)$$

Here e is the base of natural logarithms (2.71828...), and k is the wave number $2\pi/\lambda$, where λ is the wavelength of sound at the given frequency. I define

a reference distance x_0 at which pressure is arbitrarily set to unity. This will be the distance from the point source to the acoustic center of the dipole. Because wavelength is just c/f , where c is the speed of sound and f is the frequency in units of Hz, you can now write the wave equation as⁵:

$$p(x, t) = \frac{x_0}{x} e^{i2\pi f(x/c - t)} \quad (2)$$

You've probably noticed that I'm using complex notation here, where i is the square root of -1 . It may seem like an unnecessary burden, but it actually helps to keep track of phase when you do the calculations in Excel or other math programs. You just need to be sure to let Excel know that these are complex numbers. Note that some math programs expect "j" instead of "i" when entering imaginary values.

Now consider an angle θ between the axis of the dipole and a line from the point source to the center of the dipole, where d is the acoustic distance between the two microphone capsules in the dipole (Fig. 1). By the law of cosines, the distance from the source to each microphone capsule is given as⁶:

$$x_A = \sqrt{x_0^2 + \left(\frac{d}{2}\right)^2 - 2x_0\left(\frac{d}{2}\right)\cos\theta} \quad (3a)$$

$$x_B = \sqrt{x_0^2 + \left(\frac{d}{2}\right)^2 + 2x_0\left(\frac{d}{2}\right)\cos\theta} \quad (3b)$$

Here I've labeled the distance from source to microphone as x_A for the front capsule and x_B for the rear capsule. For

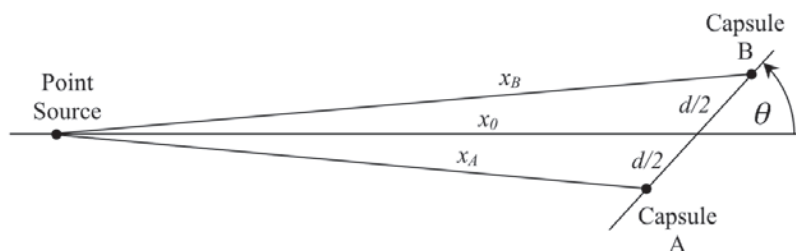


FIGURE 1: Dipole microphone design features two omnidirectional capsules responding here to spherical waves emanating from a point source. Distance from source to capsule A is x_A , and from source to capsule B is x_B . Acoustic center of the dipole is at distance x_0 . Length of the dipole d is the acoustic distance between capsules A and B.



PHOTO 1: Auratone 5C Super Sound Cube loudspeaker mounted on top of a tripod-style mike stand. Dipole microphone prototype is supported by a mike stand sitting on a cheap turntable. Speaker and microphone are positioned 1m apart, midway between floor and ceiling, near the middle of the room.

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large x_0 (that is, $x_0 \gg d/2$), this can be simplified to $x_0 \pm (d/2) \cos\theta$, but let's keep the full form for now because I will be considering small x_0 (that is, x_0 comparable to $d/2$) when I look into the proximity effect.

Now you can construct a spreadsheet to calculate the omni, gradient, and cardioid polar response patterns for an acoustic sine wave. I'll begin with some columns for given values, including the speed of sound (345m/sec assuming 22° C ambient temperature), distance from source to dipole (1m), length of the dipole (0.01m), frequency of the incident sound wave (1000Hz), and angle of dipole axis with respect to the source (0°). I've used the "Fill Down" command to copy the given values into 72 consecutive rows, except for angle, which I set to run from 0° to 355° in 5° steps. I placed a column alongside the angle which converts from degrees to radians using the Excel built-in function RADIANS(), because many Excel functions assume radians as input. Now you calculate the distance from source to microphone A using equation (3a), and from source to microphone B using equation (3b).

All calculations up to this point have been real numbers, but now I will switch over to complex math to obtain the response to the source signal using equation (2) for each capsule, replacing x with the value of x_A or x_B as needed for each capsule. Because I've assumed steady-state conditions, the phase relationship between the two capsules will be the same at any time t , so you can set $t = 0$ to simplify the calculations. You then take the complex sum and complex difference of the signals from capsules A and B. The absolute value (radius in the complex plane) is the output level in units of pressure where the reference level is unity, while the argument (phase angle in the complex plane) is the output phase in radians. A radar plot of the difference output level should show a figure-eight pattern, while the sum output level should be a circle.

Use the Excel functions IMABS() and IMARGUMENT() to obtain real-valued numbers for plotting. If Excel gives a "#NAME?" error when you try to use complex math functions, check to be sure that you have the Analysis ToolPak enabled, under "Tools Add-Ins..."

But now you notice that at 1kHz, the difference (gradient) output level is much lower than the sum (omni) output level, and as you change frequency you find that the difference output level has a large frequency dependence, increasing along with frequency. What happened was that when I calculated the difference between the microphone A and microphone B signals, I effectively took the differential Δp , which caused a factor of $i(2\pi f/c)\Delta x$ to pop out, where Δx is just the dipole length d . This factor causes the frequency dependence along with a 90° phase shift.

To make the on-axis difference signal look like the omni signal, you must divide out a factor of $i(2\pi f/c)\Delta x$. Luckily, you can integrate the gradient signal with respect to time to remove $i2\pi f$, and $\Delta x/c$ is just a constant gain factor equal to the dipole length divided by the speed of sound. So add a new column to the spreadsheet where these factors have been divided out of the complex difference response.

With this change the gradient response is now uniform over frequency, but still off by a factor of 2, due to the sum signal including the signal from two capsules. So you divide the complex sum response by 2 (think of it as an average), and can now compare the amplitudes of the omni and gradient responses on an equal basis. Alert readers will have noticed that x appears twice in equation (2), but when I integrated the difference signal with respect to time, I accounted for only one of the x 's. The presence of this second x is actually the origin of the proximity effect in gradient microphones.

The final step is to show the cardioid response in the polar plot. This is done by taking the complex average of the omni and gradient responses and plotting the absolute value. The result should be a familiar heart-shaped pattern, but it's pointed in the wrong direction!

It turns out that I overlooked a minus sign when obtaining the time integral of the gradient response. Reinstating the minus sign will point the cardioid microphone in the right direction. So you can obtain both a front-facing and rear-facing cardioid response from the same omni and gradient signals (Fig. 2).

By the way, an important characteris-

tic of first-order cardioid microphones is that the response at 90° and 270° should be exactly 50%, or -6dB when plotted logarithmically. You now have a toy model that reveals many of the characteristics of real microphones, as frequency and distance from the source vary. If you are using Microsoft Excel, you can download a fully populated spreadsheet of this model (www.audioXpress.com).

MEASUREMENT TECHNIQUES

To test the model spreadsheet, I'll describe the construction of a prototype dipole microphone. But first, you need a set of measurement techniques in order to evaluate the prototype. According to the assumptions that went into the spreadsheet, you want to measure the free-field steady-state response to spherical sound waves arriving from a point source, in the frequency range from 100Hz to 10kHz. If you are already set up to measure microphone polar and frequency response, you can skim over this part of the article. For readers who are interested in developing their own setup for microphone measurements, however, I'll give the complete details.

To achieve free-field conditions is a bit of work, but well worth the effort. Building an anechoic room is one option, but way beyond reach for most of us. If you want to be able to stand up inside your anechoic chamber, you'll need 2m clearance vertically, plus 2m above and 2m below for the sound-absorbing wedges, requiring a space of at least 6m in height for construction. This would be for a room with a cutoff frequency

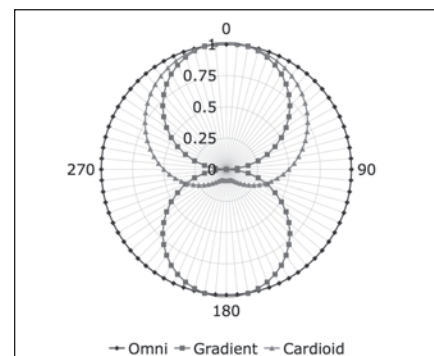


FIGURE 2A: Calculated polar plot of 300Hz response with dipole length of 10mm. Signals shown include omni, gradient, and cardioid. Distance from source is set to 1m, so proximity effect is minimal. Radial axis is linear.

around 86Hz, where 2m is a half-wavelength. Below cutoff, the chamber reverts to pressure mode and is no longer anechoic. Readers living in a mild climate might consider using an open field or empty parking lot, especially if they can dig a small pit for a loudspeaker to be mounted so the baffle is even with the ground, pointing straight up in a 2π (hemispherical) configuration.

The microphone under test can then be suspended directly above the loudspeaker, pointing straight down. Note, however, that you'll need to accept whatever ambient noise is present in the environment. Also, the speed of sound has a rather strong temperature dependence, varying almost 4% over the temperature range from 0 to 20° C.

I actually chose to obtain free-field conditions in my living room with a tone-burst source and gated measurement system⁷ using wave files and the sound hardware of a laptop computer. My living room ceiling has a height of 2.43m. With loudspeaker and microphone suspended 1m apart midway between floor and ceiling, the direct sound arrives at 2.9msec after launch, while the first echoes arrive at 7.6msec (**Fig. 3**). That gives me 4.7msec of direct sound with no echoes, satisfying the requirements for a free-field measurement. I can fit at least one complete cycle of a free-field measurement at frequencies of 212Hz and above.

When I measure frequencies below 212Hz, at least some of the first echo from floor and ceiling is included in

the measurement. If loudspeaker and microphone are moved closer, to 0.3m separation, then I have 6.2msec in the free-field regime, allowing good measurements down to 161Hz. Measurements below cutoff don't blow up, but rather begin to include more and more sound from directions other than the source with falling frequency. While it's definitely not steady-state, you will see that this technique gives results that are easily good enough to justify the effort in obtaining them.

All test signals were calculated and written directly to wave files using a simple command-line utility program called "tbq.exe," written for the purpose in C++. This technique of using offline signal generation and analysis has worked out very well for me, and seems to have no real disadvantages when compared to more sophisticated graphical user interface (GUI) programs that make their measurements in real time. Complete source code is available for readers to download (www.audioXpress.com).

PROTOTYPE CONSTRUCTION

Now you can build a prototype dipole microphone, to see whether the characteristics predicted in the model spreadsheet are meaningful. The first thing you must decide on is the length of the dipole. As you can see from the spreadsheet model, there are nulls (complete cancellation) in the on-axis omni response whenever the dipole length is equal to odd multiples of a half wavelength. For instance, a dipole length of

10cm gives nulls at 1725Hz, 5175Hz, 8625Hz, and so on. I've used 10mm diameter capsules for my prototype, so the closest I can easily place the acoustic centers of the capsules is 10mm, giving the first null at 17.25kHz. This helps to explain why, especially for large diameter microphones, it may be better to avoid gradient behavior at high frequencies.

The basic requirement is to sample pressure at two nearby points in space. To do this I used a pair of Panasonic omnidirectional electret condenser microphone cartridges, part #WM-034BY, which I had on hand. It seems that this part number has become obsolete, but the cartridges themselves are completely generic, and can be easily substituted without affecting performance. I chose to mount the microphone capsules facing upwards in a probe configuration (**Photo 2A**). This allowed me to put the electronics beneath the microphone, away from the line-of-sight for incident sounds, and made the microphone nearly symmetric about the vertical axis for polar plots.

You may ask why I didn't face the capsules forwards, or perhaps one forward and the other one backwards. I actually considered all of these configurations, but didn't try them because I had in mind to make a bidirectional microphone that would allow me to develop both front-facing and rear-facing cardioid patterns from the same signals. I also considered the fact that gradient microphones are notoriously sensitive to handling noise and mechanical vibra-

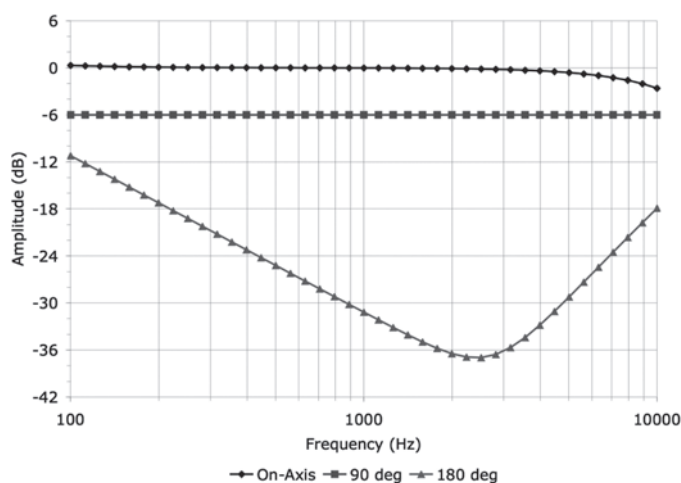


FIGURE 2B: Calculated frequency response of microphone with dipole length of 10mm. On-axis, 90°, and 180° plots are shown. Vertical axis is in dB.

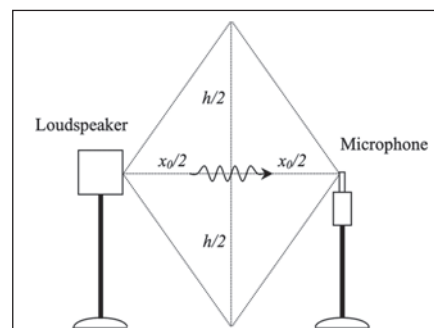


FIGURE 3: Direct sound arrives at the microphone at time $t = x_0/c$, where x_0 is the distance from source to microphone, and c is the speed of sound in air. First echo arrives a bit later, after being reflected off floor and ceiling at height h . Free-field conditions are obtained during the interval between direct sound and first echo⁷.

tion. Facing both capsules in the same direction should minimize the difference signal between them when both are subjected to the same vibrations⁸.

As shown in **Photo 2B**, two brass tubes hold the capsules up above an electronics

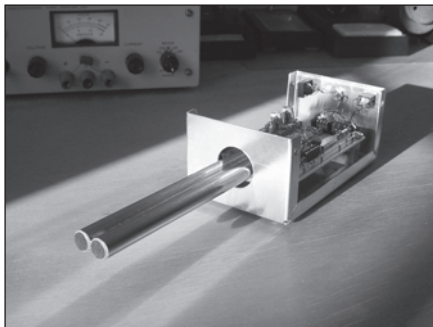


PHOTO 2A: Two electret microphone cartridges are mounted side-by-side in a probe configuration above an electronics package. Thin wall brass tubes are tack-soldered to each other, and rigidly mounted to the perfboard circuit assembly.

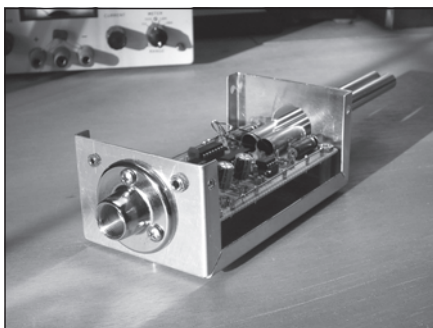


PHOTO 2B: Threaded mike stand adapter is aligned with the acoustic center of the dipole. Power supply input and line level output jacks are also mounted on the bottom end of the project box.

package contained in an aluminum project box. The tubes and project box are all tied to ground for proper shielding. I obtained a threaded mike stand adapter from Atlas Sound which I mounted on the bottom of the project box, centered along the vertical axis of the microphone. Although not shown, you can easily fit an inexpensive foam windscreen around the capsules if needed.

I provided two jacks at the bottom of the project box, one for a 9 to 12V DC power input, and one for a stereo line level output to match my computer's sound card input. The two capsules should be rigidly mounted with respect to each other, and should have mechanical resonances well damped. In my prototype, I found that the wires connecting to the capsules had a strong resonance in the low audio range, so I stuffed the brass tubes with tissue paper to keep the wires quiet. I also experimented with using a pyramid, glued from foam-core board, to blend the surfaces of the project box to the probe tubes, but wound up not using it for the measurements in this article.

The electronics package implements the analog computer I designed while coding the spreadsheet earlier in this article. It consists of gain stages, adders, and an integrator, as indicated in the block diagram of **Fig. 5**. A detailed schematic is given in **Fig. 6**. The completed assembly is in **Photo 3**.

I mounted all parts on a piece of 0.1" grid perfboard using DIP sockets and 0.042" perf-clips. Wiring was done point-to-point using a "wrap and burn"

technique with enameled magnet wire. A set of jumpers JMP1 allows me to reconfigure the outputs between measurements.

Instead of providing trim pots for balancing gains, I decided to mount fixed resistors in parallel to accomplish the same thing. I used packaged resistor networks RN1 and RN2 for the 15k Ω resistors, because their values should track well over time and temperature. Coupling capacitors C1, C2, C4, and C5 in the circuit path are all sized to put the low-frequency cutoff well below the lowest frequency of interest. The integrator capacitor C3 is sized to roughly match the gradient and omni signal levels. R5 limits the DC gain of the integrator, and can be used to adjust low-frequency response of the gradient signal, and thus tune the proximity effect in the cardioid outputs.

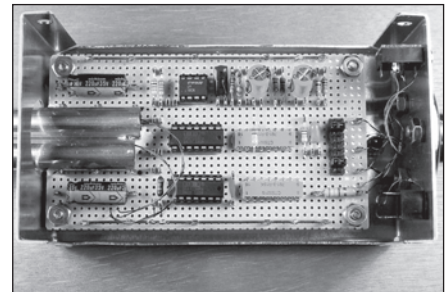


PHOTO 3: Analog computer is assembled onto a piece of perfboard with 0.1" hole grid. Power supply and ground bus wires are at top and bottom. Circuit adjustments are made by soldering in optional resistors. Jumper block at right allows various signals to be monitored.

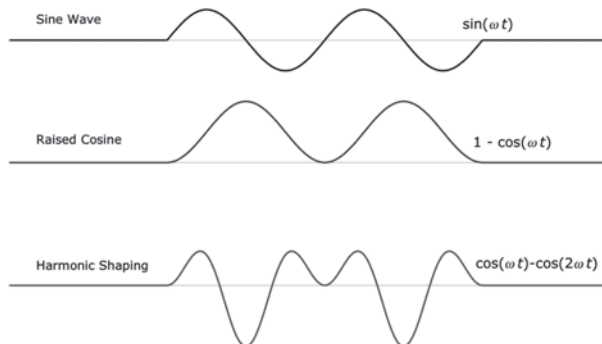


FIGURE 4: Simple sine tone burst shown above has discontinuous slope at the start and end. Raised cosine has continuous slope, but gives a DC bump during each tone burst. Adding in the second harmonic with equal amplitude and opposite polarity corrects the DC bump, and gives a continuous band-limited function that can be repeated as necessary to fill a tone burst. Matched filter removes the harmonic during analysis.

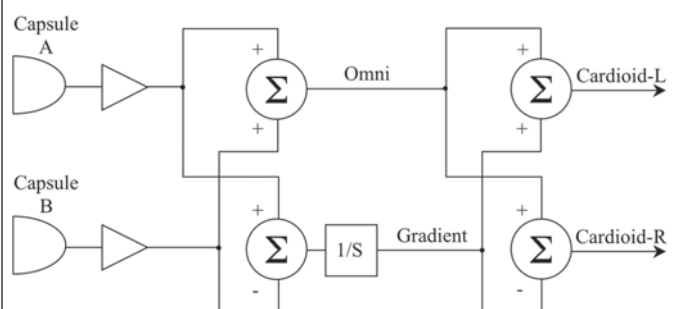
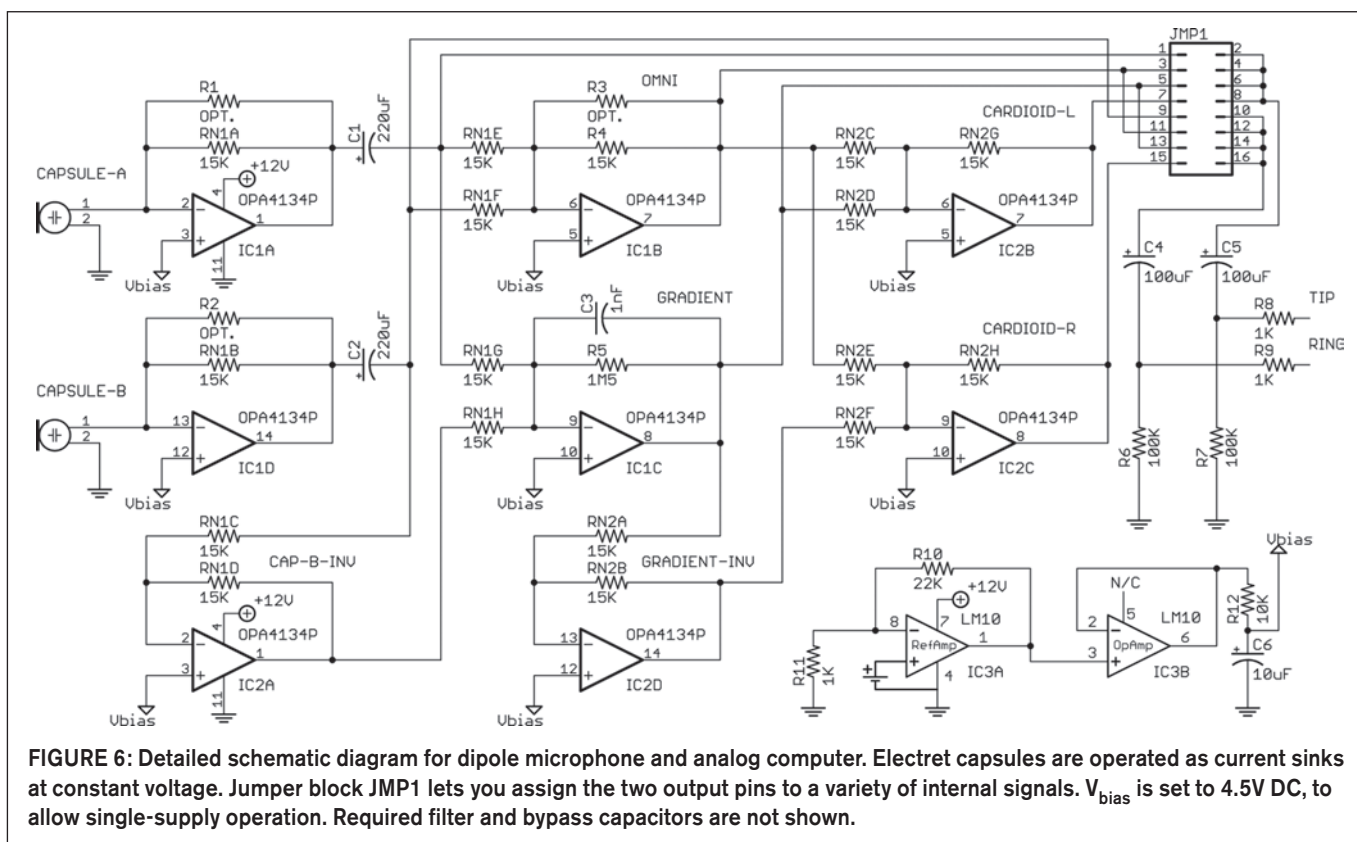


FIGURE 5: Block diagram of dipole microphone with analog computer. Gradient and cardioid signals are derived using sum, difference, and integrator blocks¹. Gain stage for each capsule converts current to voltage.



I used two Burr-Brown OPA4134 quad op amp packages at IC1 and IC2 for the analog computer, mainly so I could become familiar with their characteristics. I found them to have low noise, moderate current consumption, and they are very fast. Layout and decoupling are important, as is an oscilloscope that can see oscillations well beyond 20MHz. Readers may prefer to use TL074 or other unity-gain-compensated op amps in this application. I chose an LM10CN integrated op amp and voltage reference for IC3 to set the bias voltage for single-ended power supply operation.

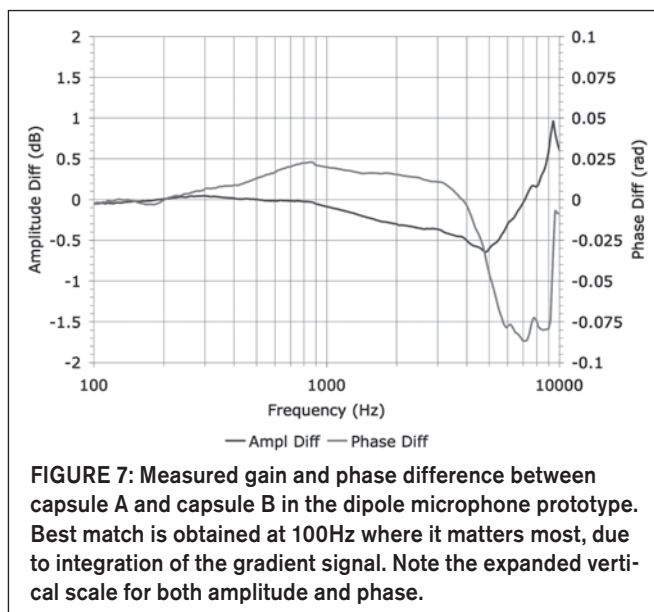
Instead of just buying two electret microphone capsules and hoping they would match, I actually bought ten capsules (they're not expensive), measured the on-axis phase and frequency response of each, and then chose the two that had the best phase match at 100Hz, while maintaining a similar frequency response all the way up to 10kHz. I actually measured the capsules in pairs, so I could look at their

relative phase response instead of the absolute response, and thus subtract out the absolute distance from speaker to microphone. The assumption here is that I've precisely set the two capsules equidistant from the loudspeaker. Note that at 100Hz, 1 milli-radian (mrad) of phase difference is just 0.55mm!

The next step is to match the sensitivity of the two microphone cartridges over as broad a range of frequencies as

possible. I did this by first measuring the amplitude difference versus frequency for the two capsules when connected to the preamps as shown in **Fig. 7**. I then calculated the gain change required to match the outputs at 1kHz and soldered in the appropriate resistors at optional locations R1 and R2. The gain and phase differences between the two capsules after this adjustment are shown in **Fig. 7**, measured using the harmonically shaped tone burst technique.

You can now measure the omni and gradient responses, both in frequency and polar domain. The results look encouraging (**Fig. 8**) with almost 30dB of side rejection in the gradient response at 300Hz. It seems that the integrator design was correct, because the omni and gradient on-axis frequency responses are a good match. To obtain the correct cardioid response, you need to precisely match sensitivity of the omni and gradient responses for sounds coming from the rear, the direction of greatest rejection in the cardioid pattern.



It turns out that the gradient polar response can be asymmetrical due to a small mismatch in phase between capsules A and B, so you may need to adjust the “left” and “right” cardioid balance separately. Gross adjustments can be made by changing the value of integrating capacitor C3. Fine adjustments can then be made by adding re-

sistors at optional R3, or in series with R4 to optimize the left cardioid pattern. The goal is to obtain the best null for rear rejection over the widest range of frequencies.

If your gradient polar response is symmetrical, then the right cardioid response should also be optimum at this point. In my case, however, the gradient response

was about 1dB different from front to back, so I adjusted the gradient signal in the right cardioid by adding some resistance in parallel with RN2A or RN2B around amplifier IC2D. Now I have two cardioid responses, one facing front and the other facing back, with rear rejection of 20dB or better in the range from 200Hz to 10kHz (Fig. 9).

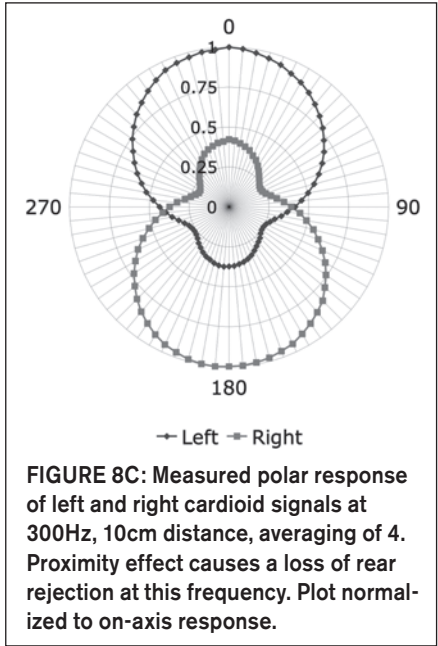
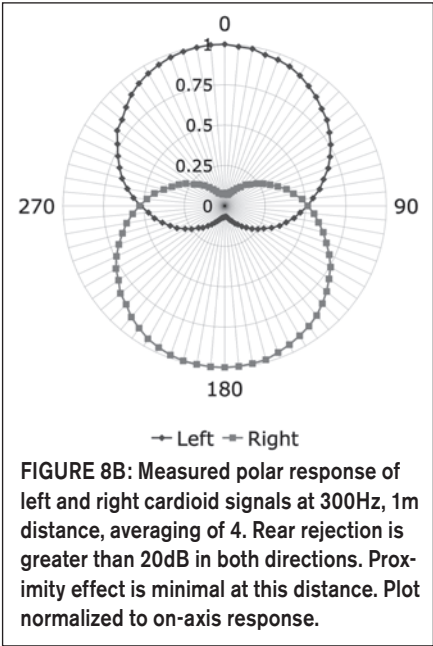
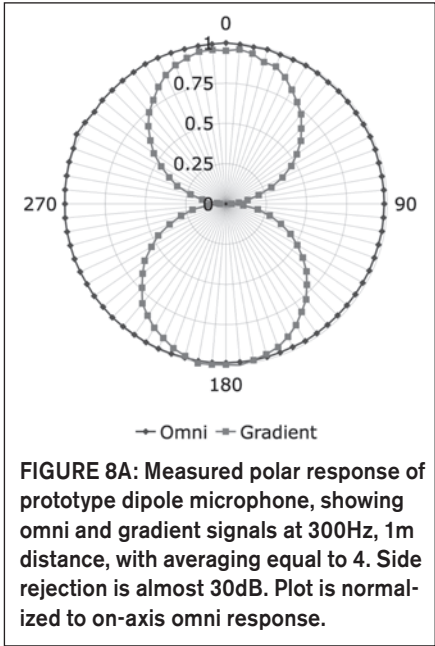
You may be wondering whether the sum and difference operations implemented in the analog computer discard any information. It turns out that they don't, in the sense that given the left and right cardioid signals, you can recover the original omni and gradient signals by applying another matrix operation which is the inverse of the sum and difference operation. Expressed as a matrix, it turns out that the sum and difference operator is self-inverse, times a factor of 2.

$$\begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} = \begin{bmatrix} 2 & 0 \\ 0 & 2 \end{bmatrix} \quad (5)$$

This same matrix operator appears in other places in audio, for instance, in stereo modulation of broadcast signals, and in mid-side stereo microphones.

At this point, I've barely scratched the surface of what can be done with directional microphones. I've assumed flat response everywhere, except for rolling off the integrator on the low end. Obviously, much more work can be done with equalization and phase matching here.

TABLE 1 Dipole Microphone Prototype Parts List			
Designator	Part Number	Description	Manufacturer
		0.1μF 50V ceramic disk bypass capacitors	any
		suitable project box	
4 pcs.		#6-32 3/4" metal standoffs	
2 pcs.	137	7/16" dia. brass tubing	K&S Engrng.
	169P44	0.1" grid perfboard	Vector
	AD11B	5/8"-27 mike stand flange	Atlas Sound
	T42	0.042" perf-clip terminals	Vector
2 pcs.		14-pin DIP socket	any
		8-pin DIP socket	any
C1,C2		220μF 35V electrolytic capacitor	any
C3		1nF mylar film capacitor	any
C4,C5		100μF 35V electrolytic capacitor	any
C6		10μF 16V tantalum capacitor	any
Capsule A,			
Capsule B	WM-034BY	Omnidirectional Electret Condenser Microphone Cartridge	Panasonic
IC1,IC2	OPA4134PA	High-performance quad op amp	Burr-Brown
IC3	LM10CN	Op amp and voltage reference	National Semi
JMP1		8-position 2-row header	any
line out		3.5mm 3-circuit phone jack	any
power in		2.5mm 2-circuit phone jack	any
R1,R2,R3		Optional gain trim resistors	any
R4		15kΩ 1/4W resistor	any
R5		1.5MΩ 1/4W resistor	any
R6,R7		100kΩ 1/4W resistor	any
R8,R9,R11		1kΩ 1/4W resistor	any
R10		22kΩ 1/4W resistor	any
R12		10kΩ 1/4W resistor	any
RN1,RN2	761-3-R15K	15kΩ DIP resistor network	CTS
Substitutions are encouraged, in the spirit of promoting research and development.			



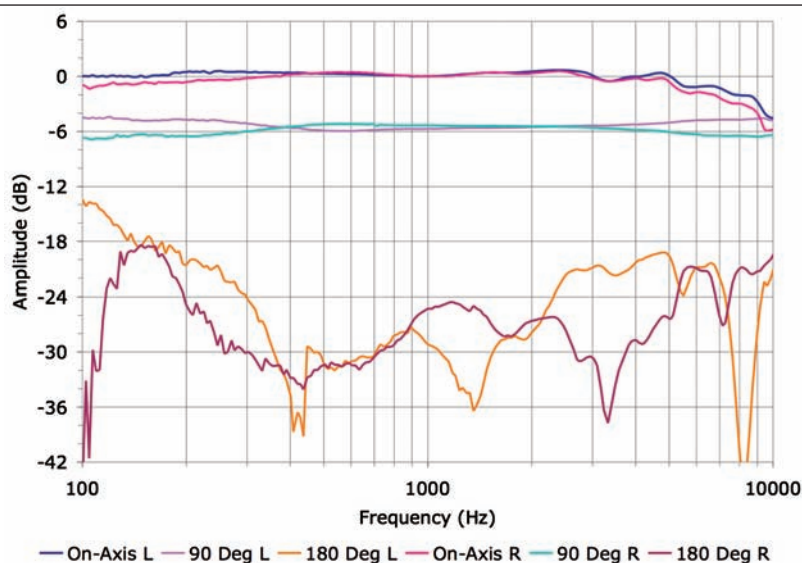


FIGURE 9: Measured frequency response of left and right cardioid signals measured at 1m with averaging equal to 4. Left and right curves are superimposed, with dipole axis at 0°, 90°, and 180° with respect to the source. All curves are normalized to the 90° response of the rear capsule. Side response is down 6dB, as it should be. Rear rejection is 20dB or better from 200Hz to 10kHz from either side, which is quite respectable.

You could add more capsules to obtain higher order polar patterns or to simulate larger diaphragm microphone designs. The electret capsules weren't modified for improved headroom, as reported by some authors. I haven't used any acoustical elements in this design, but you could certainly do that as well.

CONCLUSION

So what does a dipole microphone sound like in actual use? To find out, I recorded a rehearsal of some musician friends. The left- and right-facing cardioid patterns cleanly divided the group into two bunches, with a few stragglers in between. The bidirectional cardioid pattern could be useful in cases where the only place you can put a microphone is in the middle of the stage. Note that the sum of the left and right signals is just the omni response. This really is a special-purpose microphone as designed, but it holds out lots of promise for future development.

Mechanical noise, coupled from the floor through the mike stand, is tolerable but needs improvement. Signal amplitude is awkwardly positioned between microphone level and line level. Some of the outputs are inverted in terms of absolute phase. Self-noise out of the analog computer is acceptable but needs work. Wind and pop noise sensitivity are high as expected, and readily improved with a generic open-cell foam windscreen.

Proximity effect is less than I've experienced with other microphones, and may take some getting used to. Now I can begin the real work of correlating measured results with listening performance, for both purchased microphones, and those of my own design. For more on the test methods used here, go to www.audioXpress.com.

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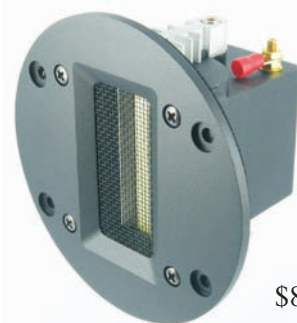
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CORRECTION

More legible figures for the published article T-reg: a High-Voltage Regulator for Tube Amps are available on the author's website: www.linearaudio.nl. The figures as published in the magazine (April '09, p. 24) are correct.—Eds.

TESTING IMPEDANCE

I thank Mr. Yaniger for a cordial exchange of e-mails and especially for his willingness to test my 12AX7 suggestion, which did not pan out (Xpress Mail, April '09, "The Impasse Preamp," p. 42). His original 6SN7 was the better gain stage. Unfortunately, we have

not been able to agree on his statement that the total impedances at the phase splitter's (PS) plate (P) and cathode (K) are identical if the P and K load impedances are also identical.

Mr. Yaniger's proposed test of output impedance requires that identical test loads be present simultaneously at both the P and K. No justification is given for this requirement, which is not found in any engineering textbook that I have ever seen. The standard prescription for testing output impedance involves driving the input of the circuit under test (C.U.T.) with a signal. The unloaded voltage is measured at the output node

under test.

Next, a short circuit to ground at the frequency of the test signal is applied to that node, and the current into the short is measured. The output impedance is defined as the voltage divided by the current. Never is any mention made of placing additional test loads at any other node in the C.U.T. Nor is there any discussion of testing certain kinds of circuits differently, whether those circuits are balanced, symmetrical, have differential outputs, or have any other particular characteristic.

Certainly, shorting an output will lead to operation in a manner quite different

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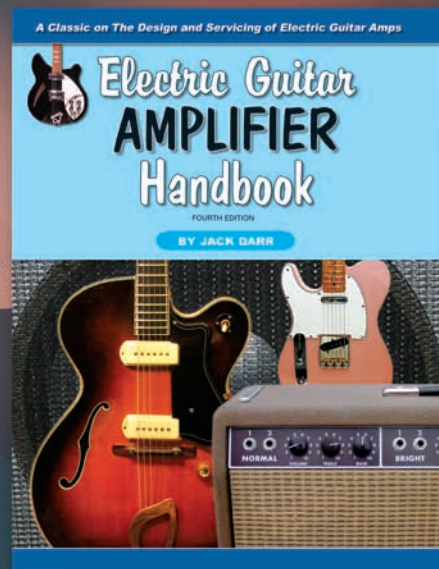
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from what was intended by the designer, but this is clearly a legitimate part of an impedance measurement. Clearly, adding test loads to multiple circuit nodes will yield results different from the classical impedance test. Such a change in an established definition requires a hearty justification, especially since the results are at odds with the standard measurement.

Mr. Yaniger mentions the Jones text, which offers two arguments in favor of his position. The first argument states without proof or derivation that feedback in the PS *reduces* the P impedance. It is worthwhile noting that pages 312-313 (and 329-330 in more detail) of the 4th edition of the *Radiotron Designer's Handbook* (RDH4) provide a derivation which shows that the feedback through the cathode load *increases* the P impedance. There are numerous web references (check out, for instance, www.rocw.raifoundation.org/computing/BCA/electroniccircuit/lecture-notes/lecture-26.pdf) which agree with this conclusion as do my own derivations, which I'd be happy to share with anyone interested.

This wealth of contradictory derivations should at least make the text's claim a contentious one requiring a justification. Strangely enough, the text itself on page 404 states that the impedance looking into the plate of a triode is $r_p + (1 + \mu)R_k$. This would clearly be *reduced* rather than increased (as later claimed) if R_k were bypassed and cathode feedback eliminated. It is unclear to me how these contradictory statements in the text itself can be reconciled.

The second claim of the text is that since the frequency responses are identical at the P and K in a triode with identical P and K loads, "the output resistances must also be equal." But while this observation is true, the conclusion is demonstrably false.

Suppose I drove a load Z from a voltage source with a low impedance r . A current $i(t)$ would flow from the source through r and Z , and a voltage $v(t) = i(t)Z$ would arise across Z . Now suppose I drove the same load Z with a current source of a high impedance R . If I wished $v(t)$ to arise across Z again, I'd set my current source to supply $i(t) + v(t)/R$.

The total impedance of these circuits at the load is the parallel combination of the source impedance and Z . Because the source impedances differ in the two circuits, the total impedances differ even though the same voltage arises across each Z . So equal voltages across identical Z s in the PS clearly do not imply equal total impedances. Equal voltages are a simple consequence of the facts that the same current flows through the P load, the triode P and K and the K load, and that the P and K loads are identical, as noted in the *RDH4* reference. Ohm's law takes care of the rest.

A straightforward, classical, by-the-book measurement or derivation of the impedances at the P and K in the PS with identical P and K loads shows those impedances in general to be quite different—not identical.

I stated in my prior letter that it would be possible to eliminate the impedance imbalance of the PS by adding a resistor in series with the cathode output. Mr. Yaniger correctly pointed out that this change alone would cause a voltage imbalance at power ampli-

fier inputs with finite impedances. I was remiss in not having pointed out that there would need to be a compensatory change in the rest of the circuit to address that imbalance.

It turns out that you can modify the PS to have both balanced output signals and balanced drive impedances for a selected power amplifier input resistance. You need to increase slightly the value of the cathode bias resistor when you add the series resistor. Unfortunately, the gain balance then becomes worse than that of the standard PS the moment you add a common value of capacitance across the power amplifier inputs. Not exactly a step forward. Again, if anyone is interested, I'd be happy to share the details.

Christopher Paul
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STRANGE MIDPOINT

I studied the solid-state piece by Ian Hilton in the April 2009 issue of *audioXpress*, "Keeping the Hobby Alive" (p. 38), with interest. I am sure that

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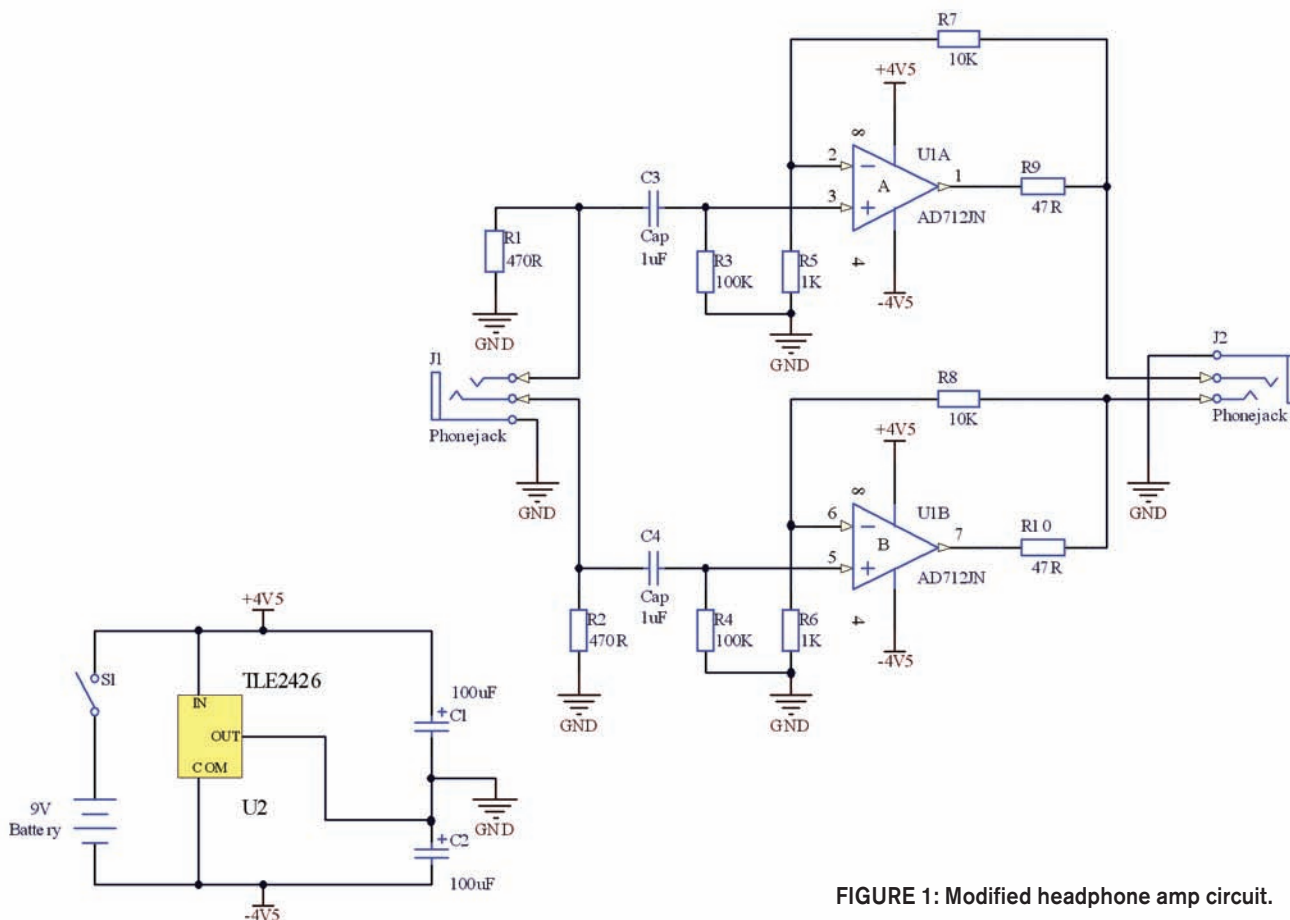


FIGURE 1: Modified headphone amp circuit.

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the solution for this headphone-circuit sounds great, but, in my opinion, the emitter-follower midpoint circuit does not work. First of all, there is nothing to define the midpoint voltage. I suggest that what really does the trick is the common mode and the power-supply rejection ratios of the op amps. I believe the circuit would work better if you entirely removed the 2N3904 Q1 transistor and substituted a connection between the junction of R1 and R2 and the junction of C1 and C2, and call that GND.

As it is, the midpoint is not a midpoint at all; it is some voltage in which Q1 just barely—perhaps—conducts. There is nothing to define that voltage if the midpoint is more positive than -0.6V. In other words, if the midpoint is anywhere in the positive direction above -0.6V of the middle, Q1 is completely off and the midpoint is undefined.

I suggest that what really does the job of defining an AC ground are the two 100 μ F capacitors; the transistor does nothing.

Hans J. Weedon
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Ian Hilton responds:

Thanks for pointing that out, Hans. You are right, of course. The emitter-follower scheme will only work if the majority of current is drawn from the pseudo -4.5V rail. In my defense, it was just a quick lash-up and not fully thought out, but it still sounds good.

A better scheme is to have a push-pull NPN-PNP pair driving the pseudo ground. Incidentally, Quad uses such a scheme (push-pull pair) to generate a pseudo ground in their Quad 306 power amp.

Further to the constructive comments from Hans Weedon concerning the pseudo ground circuit in the simple headphone amp, I have revisited the design and amended it (Fig. 1). The pseudo ground is now generated using U2, a TLE2426, a device specifically designed for that purpose. Further amendments are the inclusion of 470R input resistors. These were required with a specific MP3 player that did not like a high impedance load and commenced oscillating.

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Ed Simon ("Removing the Six "Straight Wire" Kinks: Prologue to a Better Preamp," p. 14) received his B.S.E.E. at Carnegie-Mellon University. He has installed over 500 sound systems at venues including Jacob's Field, Cleveland, Ohio; MCI Center, Washington D.C.; Museum of Modern Art Restaurants, New York; The Coliseum, Nashville, Tenn.; The Forum, Los Angeles; Fisher Cats Stadium, Manchester, N.H.

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Manu Katché|Playground|ECM 2016

Manu Katché has drummed with Jan Garbarek, Peter Gabriel, Sting, Joni Mitchell, Joan Armatrading, Youssou N'Dour, and scores more on more than 200 albums. His "Neighbourhood" album was a key 2005 release, and here he has retained pianist Marcin Wasilewski and bassist Slawomir Kurkiewicz (from Tomasz Stanko's band). But where "Neighbourhood" gloried in the combined brilliance of Stanko and Garbarek, trumpeter Mathias Eick and saxophonist Trygve Seim don't put such a firm stamp of individuality on the music. Nor are Katché's compositions as strong this time, although they do gradually insinuate themselves. Understated grooves and textural subtlety are the mainstays, and given that the man with one of the planet's most beautiful-sounding drum-kits has been well recorded, the aesthetics are not in question. You just feel there was another level that Playground could have reached.

Mike Nock Project|Meeting of the Waters|Jazzhead HEAD086

Despite being made by a marvelous nonet, this CD's ultimate glory is three minutes of more or less solo piano on a heart-rending little lullaby called *Sho's Cradle Song*. The spiky styles of trumpeter Eugene Ball and alto saxophonist Tim Wilson contrast with the more fleshy approaches of trombonist James Greening and tenor saxophonist Julien Wilson. Meanwhile, Dale Gorfinkel's vibraphone and Stephen Magnusson's guitar blur the line between horns and rhythm section. Bassist Brett Hirst solos pensively over the gentle wash of *Cyboréal* (as does Nock), and robustly on the title track. *Mysteries* has the vibraphone and guitar clouding the ensemble in a shimmering heat-haze, after which a woolly, languid tenor solo occasionally arches up to a plaintive high note, around which Magnusson instantly paints a tiny aura.



Judy Bailey Trio|Pendulum|Birdland BL017

Those who still dabble in "generic post-bop jazz" often are merely echoing a style, rather than molding substance. Pianist Judy Bailey, however, has been steeped in this music for half a century, and deploys an effervescent lightness of heart with a commensurate lightness of touch (especially notable here on *Night and Day*). This album, with bass-

ist Craig Scott and drummer Tim Firth, is centered around breezy standards. After an overly fussy arrangement of *Love Walked In*, Bailey deftly slides into an improvisation encrusted with glistening, gem-like ideas. She weights the notes with aching precision on the delicate *Flamingo*, and lilts and skips on the $\frac{3}{4}$ title track. Scott develops a compelling narrative in his solo on *You and the Night and the Music*, while Firth swings throughout and solos melodically.



Maria Schneider Orchestra|Sky Blue|ArtistShare AS0065

Maria Schneider eschews the tired cliché of big-band writing. She treats her orchestra as exactly that, so the trumpet section, for instance, is not a block of sound, but a set of possibilities. Nor does she have a chain of solos over a given backdrop. Rather, each solo is a true extension of the composition, specifically colored by the orchestrations. *The Pretty*

Road has Ingrid Jensen's digital delay-treated trumpet soloing elegiacally against ever-changing water-color washes from the orchestra. It's such a glorious piece of music that the tune it resolves into seems bland by comparison. Any big band is only as good as its rhythm section, and guitarist Ben Monder, pianist Frank Kimbrough, bassist Jay Anderson, and drummer Clarence Penn provide a mighty foundation for this kaleidoscope of beauty.

Enrico Rava/Stefano Bollani|The Third Man|ECM 2020

Although the expectations of rapturous lyricism from trumpeter Enrico Rava and pianist Stefano Bollani were not disturbed, this album still surprised, as though hearing music for the first time, so each note glistened, newmade and just thought of. The source of this surprise is not how brilliantly the pair play, but where the music takes these old colleagues (a generation apart), between whom the music falls as naturally as a waterfall. Perhaps that is the source of the album title (and an enthralling free improvisation), *The Third Man*: the phantom player who emerges as a combination of the two. The piano is supremely well-recorded, and needs to be to capture all the subtlety of Bollani's playing as he delicately cross-hatches the sadness and exultation spilling from that silvertined trumpet.



Albert Mangelsdorff Quintett|Folk Mond & Flower Dream|Nekta/Planet 68.861

Inside a few notes a brilliant wind player tells you just as much about himself as a good singer. The late Albert Mangelsdorff was already one of the world's great trombonists when this was recorded in 1967. The German virtuoso had not yet developed multiphonics (the art of playing more than one note at once) and was five years away from launching his famed unaccompanied concerts. Already in place, however, was a powerful sense of the warmth of the man just bursting out of the bell of the trombone. His sound is as plump and inviting as a ripe mango, and finds perfect foils in his Quintett's two slicing saxophonists (the other instruments being bass and drums). His lively original tunes are adventurous, but never anarchic: Mangelsdorff would rather charm than bully you.