

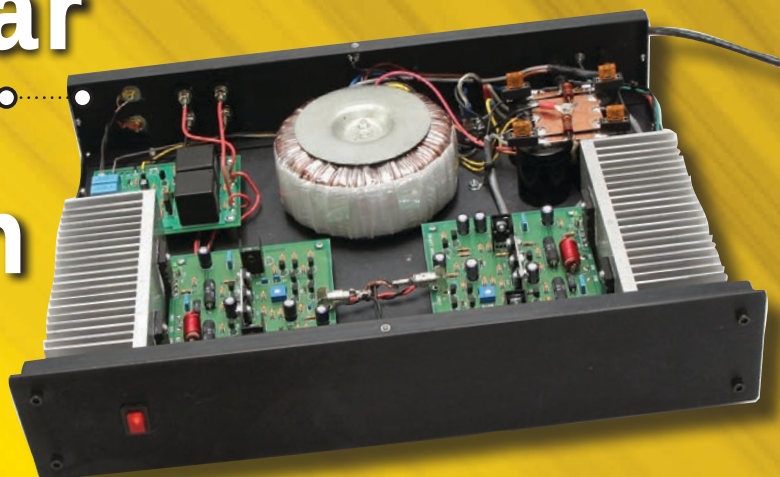
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Advancing the Evolution
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AUGUST 2012

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50-W Bipolar Transistor Amp Design



Loudspeaker Magnetics & Rare Earth Materials



DIY Pre-Amplifier Interface ("PI")

PLUS

- Defining Nonlinear Distortion
- A Small Mid-Range Driver Put to the Test
- Interview: Professional Vintage Radio Restoration
- New Products: A 1950s-Era Audio Tape is Reissued, Audio-Maker Opens Office in India, and a New Voltage Regulator



Audio Past, Present, & Future

If this is the first time you're reading *audioXpress*, you'll likely think it is a part system DIY and part audio history magazine. Well, in a way, you're correct. Our authors cover topics as diverse as the history of rare earth materials, vintage parts, circuit analysis, and the design of a 50-BJT stereo amplifier. There's something in this issue for everyone: from a tube fanatic to a speaker designer to a solid-state enthusiast.

Mike Klasco and Steve Tatarunis start the issue with a look at the history and politics of rare earth materials (p. 8). They successfully put the topic of loudspeaker magnetics into technical and historical perspective.

Robert Nance Dee presents his "PI" (pre-amplifier interface) project (p. 12). "A pre-amplifier interface enables us to control volume and inputs affording the pre-amplifier itself the sole task of amplification," he explains.

Turn to page 18 for Bill Reeve's analysis of the capacitance multiplier. He covers analog circuit design, I/O impedance, and more.

Glass audio aficionado Richard Honeycutt addresses tube sound on page 24. He looks at distortion and the effects of nonlinearity in a system. In contrast, Sam Dulaney focuses on solid state (p. 28). He writes: "These days, the DIYer is limited to designs using vacuum tubes or an esoteric topology... The 50-W bipolar transistor amplifier described here yields very good performance." On page 34, radio restoration pro Mike Urban weighs in on the tube-vs.-solid state debate as well.

Vance Dickason wraps up with the issue with the data from a test he performed on a 5.25" driver. Perhaps the results will encourage you to try Scan-Speak's 15M/4624G00 in the near future.

Regards,

C. J. Abate
editor@audioxpress.com

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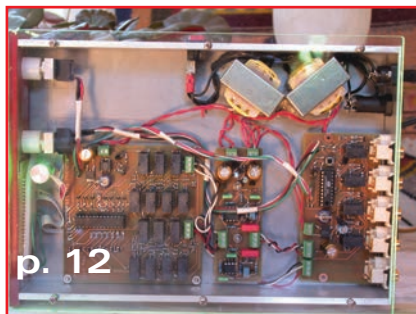
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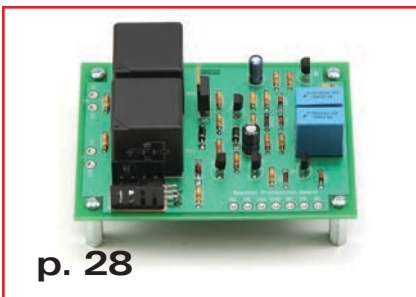
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Loudspeaker Magnetics

Politics, History, and Some Engineering

An in-depth look at magnetics and the state of rare earth materials

Speaker magnets are getting a lot of press lately as the pricing of neodymium ("neo") skyrockets and exports of it from China are tightly regulated. Neo is found in mainstream applications from automotive starter motors to hybrid electric car motors. This month's "Speakers—Parts is Parts" is more a review of the politics and history of rare earth materials than technical talk. Who would have thought the weight, distortion characteristics, and sensitivity of your loudspeakers would be influenced by Chinese politicians or African rebel leaders?

ELECTROMAGNETS

Let's start at the beginning. Early speakers were manufactured long before practical permanent magnetic materials became available. These first loudspeakers used electromagnets because adequate permanent "magnet metals" were generally not obtainable at a reasonable price. The speaker magnetic system "borrowed" power from the amplifier's power supply to create a magnetic field for the speaker's voice coil. Separate from the speaker voice coil was the field coil, which was energized by current through a second pair of connections to the speaker. This winding in the speaker's electromagnet usually served a dual role, acting as a choke coil filter for the power supply of the radio amplifier to which the loudspeaker was connected. AC ripple in the current was attenuated by the action of passing through the choke coil. An example of this type of speaker is the vintage Jensen A12 (see **Photo 1**).

ALNICO

While permanent magnet speakers existed before World War II, the complete transition of the speaker industry

to alnico magnets followed in the late 1940s. Consumer and industry products got back into production after the war. Alnico—first developed in Japan before the war—is a family of iron alloys which, in addition to iron, are composed of aluminum (Al), nickel (Ni), and cobalt (al-ni-co). Alnico alloys are ferromagnetic, with a high coercivity (resistance to loss of magnetism). It is used to make permanent magnets.

The source of most of the world's nickel and cobalt can be found in central Africa, in what was once a Belgian colony. The Belgium Congo was established in 1908, and the end of colonial rule was in 1960. Perhaps this gift of natural resources was more a curse—as over 10 million people from the Congo have died during its more than 50-year history of unrest. Called the Republic of Zaire between 1971 and 1997, today, it is known as the Democratic Republic of the Congo (DRC).

Up until the late 1970s, most speaker manufacturers used alnico magnets, and only inexpensive speakers used "mud" magnets (ferrite). The switch to ferrite in the late 1970s was in response to the unstable and shrinking supply of alnico. A civil war in Zaire ended the supply of commercial cobalt needed to create alnico. Speaker manufacturers were forced to switch to ferrite. At the time, I had a factory that made speaker systems for discos and one of the speaker vendors added a "mercenary" surcharge to its pricing because the magnet vendor hired soldiers to protect the shipments of cobalt and nickel from the mines to safe territory. Alnico availability returned after the civil war, but at substantially higher prices. Except for a few modern-day boutique applications (e.g., compression drivers and guitar speakers) alnico had faded from the loudspeaker scene by the early 1980s.

Photo 2 shows an example of an alnico guitar speaker.

SAMARIUM COBALT

A footnote in speaker magnetics is the Samarium Cobalt (SmCo) magnet. Aside from its use in alnico, cobalt is a key ingredient in SmCo permanent magnets. SmCo planar ribbon speakers were introduced in the 1970s. You may remember the Infinity EMIT tweeters



Photo 1: A vintage Jensen A12 electro-magnet speaker



Photo 2: An Eminence alnico guitar speaker (Courtesy of Eminence Speaker, LLC)



Photo 3: EMIT Tweeter Planar film diaphragm printed circuit tweeters often used SmCo magnets for enhanced performance.

that were popular near the end of that decade (see **Photo 3**). The magnetic field strength of ferrite planar speakers was too low. The samarium cobalt brought this up, enabling a reduction in the thickness of the magnet, which moved certain resonances of the ferrite planars out of the audible range. Most planar ribbons shifted to neo years ago, but with its pricing going through the roof, all alternatives are being revisited. Analog and Digital Systems (ADS)

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also utilized SmCo in the tweeters of their flagship L1530 and L2030 reference monitors in the 1980s.

THE SHIFT FROM ALNICO TO FERRITE

By the end of the 1970s, alnico was no longer an option and speaker designs had to change drastically. Alnico structures tended to be optimum with the magnet disc inside the voice coil. While ferrite structures are better served with large surface area contact between the top and back plates and the magnet donut outside the coil. Sound quality differences were caused by a combination of different distortion flux modulation mechanisms of the magnetic structures as well as differences in the air cavities around and behind the voice coil.

At this time, amplifier power—from home stereo receivers to commercial sound disco, and concert sound systems—was multiplying, and in at least some ways, ferrite was seen as an advantage. The huge, clumsy ferrite steel magnetic structures provided heatsinking of the voice coil heat, especially when vented pole pieces and other cooling schemes were used.

Alnico magnets don't like the shock as this can cause (mostly temporary) loss of some magnet strength. The shock could be caused by bouncing around or being hit with high-current transient signals. Flux modulation was another issue with ferrite designs, but copper caps and shorting rings on the pole piece can counteract and reduce intermodulation distortion.

With the ferrite magnet sandwiched, but not enclosed by the steel return structure, there was a great deal of stray flux. It is not such an issue now with LCD flat screens, but in the 1980s the CRT TV was king. This type of television required any speaker near it to have some shielding. For shielding the addition of a "bucking magnet," and often an outer "can" was added. Weight, depth, and loss of useful enclosure internal volume resulted.

NEODYMIUM & MAGNETS

Neo magnets are a compound, neodymium-iron-boron (NiB). It was first discovered in the late 1800s. In the early 1980s, practical methods of

processing neo were developed by General Motors, Magnequench, and Japan's Sumitomo Special Metals. Neo professional speakers (compression drivers) were introduced toward the end of the 1980s by ElectroVoice, and soon followed by JBL. Over the years, the performance of neo continued to increase in magnetic density, higher temperature tolerance, and improved corrosion resistance, while pricing remained stable.

There are multiple performance benefits of neo magnets for loudspeaker transducers. The most obvious is the reduced weight relative to conventional ferrite-based magnets. Other benefits for the transducer designer include stronger loudspeaker motors, which can translate to higher sensitivity, and physically smaller magnets, which allow for increased flexibility in placement. Also, most neo designs are inherently magnetically shielded, not so much of an issue today, but it was a welcome development during the swan song of the CRT television.

MINING, PROCESSING & GLOBAL POLITICS

Rare earths are extracted from mineral ores and much processing is required to reach the finished product. From 1965 to 1995, most of the global rare earth metal supply came from the Mountain Pass Mine in California. Most of these rare earths were used for color TV CRTs. The mine closed in 2002 because CRT production evaporated and the Chinese ramped up the production of neo. After the CES in Las Vegas, NV, this past January, I drove to Los Angeles, CA, on Highway 15, and you can see the Mountain Pass Mine from the road.

The escalating price of China's neo over the last few years has provided opportunities for many shuttered mines to get back into operation. Neo pricing peaked during the summer of 2011, and while pricing has softened somewhat, it is still more than three times the price it was in June 2010. Supply continues to barely outpace demand (see **Figure 1**).

The Mountain Pass rare earth mine is one of the mines ramping up for business again. The strategy for Mountain Pass is soup to nuts (i.e., from ore extraction, to separation, and magnet fabrication). Despite these new sources



Figure 1: Neo pricing June 2010–May 2012 (Courtesy of Asian Metal, Inc.)

preparing to enter the market in the next year or two, stable rare earth supply remains unclear. The Chinese may continue to place export restrictions on their rare earth resources, depending on supply-and-demand and other competitive issues. Don't forget, the speaker industry uses the least of the world's rare earth consumption with electric motors (e.g., starter motors and electric hybrid motors) competing for production supplies of neo.

Although neo magnets can be fabricated in various ways, the speaker industry buys sintered neo magnets. While some speaker assemblers buy uncharged neo, assemble these into magnetic structures, and charge (magnetize) the structure, larger neo magnetic structures require extremely powerful charges and some speaker assemblies prefer to buy pre-charged neo magnets.

WHAT'S NEXT?

Neo pricing is like gas pricing. There is a big price jump followed by a small price cut, then another big jump followed by a small cut. Death by 1,000 cuts. The U.S. government has complained to the World Trade Organization (WTO) about China monopolizing and price-gouging resources that western manufacturers need to make many tech products. While this may be true, frankly we have been outsmarted, and we are now paying the price.

But not all the credit goes to the Chinese, our own outdated mining regulations and lack of development incentives have deterred past investment in U.S.-based mining operations for rare

earth minerals. And while our complaints to the WTO look like we have been singled out, the reality is that Chinese speaker factories are also being hit with huge increases in neo prices. The difference since mid-2010 is that there has been an increasingly large delta between domestic neo pricing in China and its neo export pricing—with export prices as much as one-third higher.

Will multiple global sources for neo ore, separation, and finished magnets curb the price gouging? Ferrite will be okay for stationary speaker products (e.g., soundbars, in-wall speakers, and home theaters) as long as a bit more depth can be accepted. However, ferrite is not an option for portable professional speakers, microspeakers for laptops, headphone drivers, and other applications where performance, size, and weight count. What about alnico as an alternative to neo?

While the Zaire civil war is distant history, the present DRC remains unstable. The Lord's Resistance Army, based in adjacent Uganda, operates across the border in the DRC. Added to the mining supply instability in the DRC, we still have China's hand in the pricing. Over 50% of the world's cobalt reserves are in the DRC, but less than 5% of cobalt is refined there. Most of it goes to China for refinement. China remains the largest exporter of cobalt to the United States. Cobalt has numerous uses—including superalloys and catalysts—but its growth in battery applications has outpaced all other end uses and now accounts for 27% of overall consumption, compared to only 11% in 2002.

The positive electrodes in lithium batteries are typically lithium cobalt oxide. The infrastructure for alnico is long gone, with tooling for the pot structures not even a faint memory. If neo pricing continues to rise, and rationing and increasing demand for neo in motors puts a chokehold on the speaker industry, the road blocks to getting alnico back into the supply chain will start to look more like speed bumps. *ax*

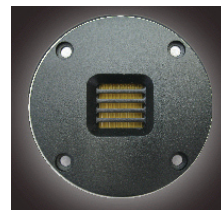


THE WINGS OF MUSIC

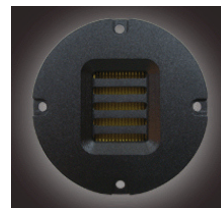
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The "PI" Design

A DIY Pre-Amplifier Interface

Amp design provides volume and balance control

My initial goal in this design was to control the level of a circuit with as little intervention as possible. After building the first prototype and evaluating it for several months (see *audioXpress*, September 2011), I decided to expand the number of volume levels and incorporate a balance control into the microprocessor algorithm. But, I'm getting ahead of myself.

This project is a pre-amplifier interface appropriately called "PI." The first issue we should address is nomenclature, there is no such thing as a "passive pre-amplifier." You simply cannot have the word "passive" and "amplifier" connected.

Amplifiers require power and power is active. A pre-amplifier interface enables us to control volume and inputs affording the pre-amplifier itself the sole task of amplification. This interface allows us to control volume, balance, and inputs. It's handy. If any of you are at all like me, which I suspect is the case, you have built, or you have gone through the process of building amplifiers (and will build many more). With this interface, you will never need another volume or balance control. In fact, you will be able to connect it to any amp you have already constructed. Or use it to connect your MP3 player directly to a power amp if the player has enough output level.

I could have traveled traditional routes, built expensive potentiometers (pots) with gold contacts, numerous resistors, multiple gangs and switches, and so forth, but I have always seen this as a flawed, albeit, brute force method lacking elegance. It's basically a poor way to address potentiometer design.

I played with digital pots, but they introduced semiconductors and added impedances into the circuit, even the PGA2311. They are also limited by how

much voltage, both positive and negative, they will accept and the distinct number of levels that are available. I decided to take out a clean sheet of paper and start from scratch to see if I could make a pot that rivaled present designs, while maintaining a reasonable cost. This is the result of that inquiry and months of exhaustive work testing, modifying, and listening. **Photo 1** shows the finished design. **Photo 2** shows how I laid out the different boards.

I wanted flexibility, that is, I wanted to be able to use it in audio or any other design. I also wanted to be able to manipulate the value of the pot with just a few components. It must have a high degree of accuracy along with flexibility. Next, I wanted it to have as few, if any, moving parts and a long life that didn't require any maintenance or lubrication. I also wanted it to work as well in 10 or more years as it did the day it was built. I needed to continuously know what level the pot was at so I could return to a specific point or listening level at any time. Lastly,

I wanted to interject as little into the circuit as physically possible (see **Figure 1**).

THE POWER OF 2

For the pot circuit of this design, I used Atmel's ATtiny48 8-bit microcontroller programmed in assembly, a lot of assembly! It allowed me to control the LCD display and the output to the relays. It also enabled me to use a small capacitor 0.33 F to save my setting with the power off so my circuit would not draw, nor require, any external power in the off condition.

The design uses binary coded decimal (BCD) to select any level from 0 to 127 (2^7) by manipulating seven double-pole relays (see **Table 1**). This has the advantage of requiring only 24 total resistors for two channels. Of the 24 resistors only 12 are ever in the circuit at any one level for our 127 (0 = true mute) levels and two channels. BCD is reflected in program as hexadecimal so no manipulation of the data is required for the outputs from the registers. Again, this can be seen from **Table 1**. It is important to understand what the values mean. This is a linear pot and each of the 128 selections gives a voltage ratio, e.g., let's say that we have our control set



Photo 1: The completed pre-amplifier interface

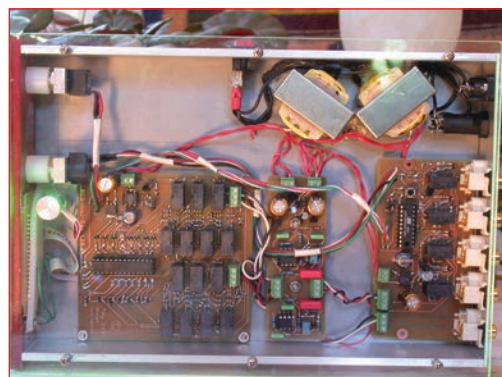


Photo 2: An inside view of the different boards

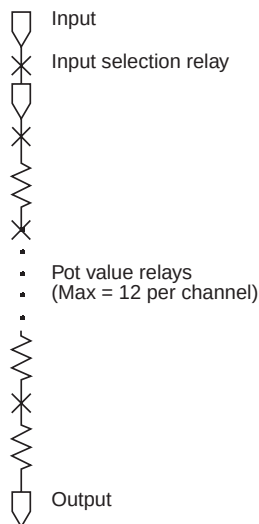


Figure 1: The pot circuit

at 32: $32/128 = 0.25$, meaning one-fourth of the original voltage (signal) is reflected in the output. I don't mind a linear pot as opposed to a logarithmic pot. After several months of living with PI, I have become accustomed to simply

setting the value of the display to my comfortable listening level. What I really like is the ability to return to an exact level anytime I want and having 128 positions means that I never hear distinct volume transitions when changing levels. The pot acts more like a continuous carbon pot without the degradation, wear, and circuit influence of carbon in my circuit. There is no carbon at any point throughout the circuit making this control system the least influential configuration possible at any cost.

The relays were chosen with special care as current draw, contact isolation, and relay life needed to be examined for the design to work well. The relays draw 6 mA each, for stereo there are 14 relays, seven per channel. We can also build one channel for mono by omitting seven relays.

The relays are hermetically sealed, washable, and have gold-plated contacts. Their cost is also low—14 relays cost under \$40 at the time of this writing. I spoke with the Tyco engineers who make the

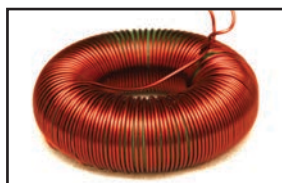
relay and the meantime between failures (MTBF) is in the millions at the current and voltages the audio circuits require. Using relays— not MOSFETs or other electronic devices—means that we have no positive or negative voltage limitations. The only thing our controlled circuit sees is the combinations of resistors...very good resistors! It is also important to realize that nowhere in this circuit is any switch exposed to outside air, thus, oxidation is not a factor.

For the resistors, I used Panasonic SMD low-noise, 0.1%, 0.125-W metal film devices. With the pot at maximum, 127 in my application, the total resistance error was 6 Ω for the 12+K pot I built, the linearity, as one might expect is extraordinary. To make up the values shown in **Table 1** only requires three

Resistance values						
	32	16	8	4	2	1
Binary	1	0	1	0	1	0
BCD	42					
Hexademcimal	2		a			

Table 1: The design uses a binary coded decimal

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For example:

- The total resistor value equals 127 Ω or 1.27 k Ω , 12.7 k Ω , 127 k Ω , or 1.27 M Ω .

To keep trouble-free operation at a low cost, a 64 step-optical encoder is used to switch the relays through the microcontroller. Operating as a stereo volume control, there is no sense of “stepping” as the volume is increased

product. I did not add remote control to this circuit but the design, by nature, lends itself well to this addition.

We have another benefit by controlling our circuit with a microprocessor, we can control how we apply the balance. A conventional balance control decreases the volume of one channel while the other channel remains unchanged. For this design, I wrote the algorithm to increase the level of one channel and decrease the level of the other. Thus, our overall volume remains constant, and we don't have to re-adjust the volume after changing the balance. As with the volume control section, you always know exactly where you are with both balance and volume. Pushing the balance button brings up the balance screen. With both channels at the same level, the screen reads " $L = R$," changing the volume control with the balance button in allows us to change the level of the left or right channel, depending



For this control, the ATtiny48 does all the work. The display mounts on the main board or through a ribbon tape. I did not isolate the supply from the resistor boards as noise measurements are below what my instruments can sense. The resistors are not tied into any part of the circuit only the sealed relay contacts. There is a jumper to tie the low leg of the pot to ground, which I utilized in my stereo control. But in all respects, the resistors are autonomous. There is a place to add a resistor in series with the respective channel. This is not for balancing both channels, none is needed. I use this resistor to control the level of the pre-amp. Simply jump across the resistor if you do not need it. By adding resistance here, the builder

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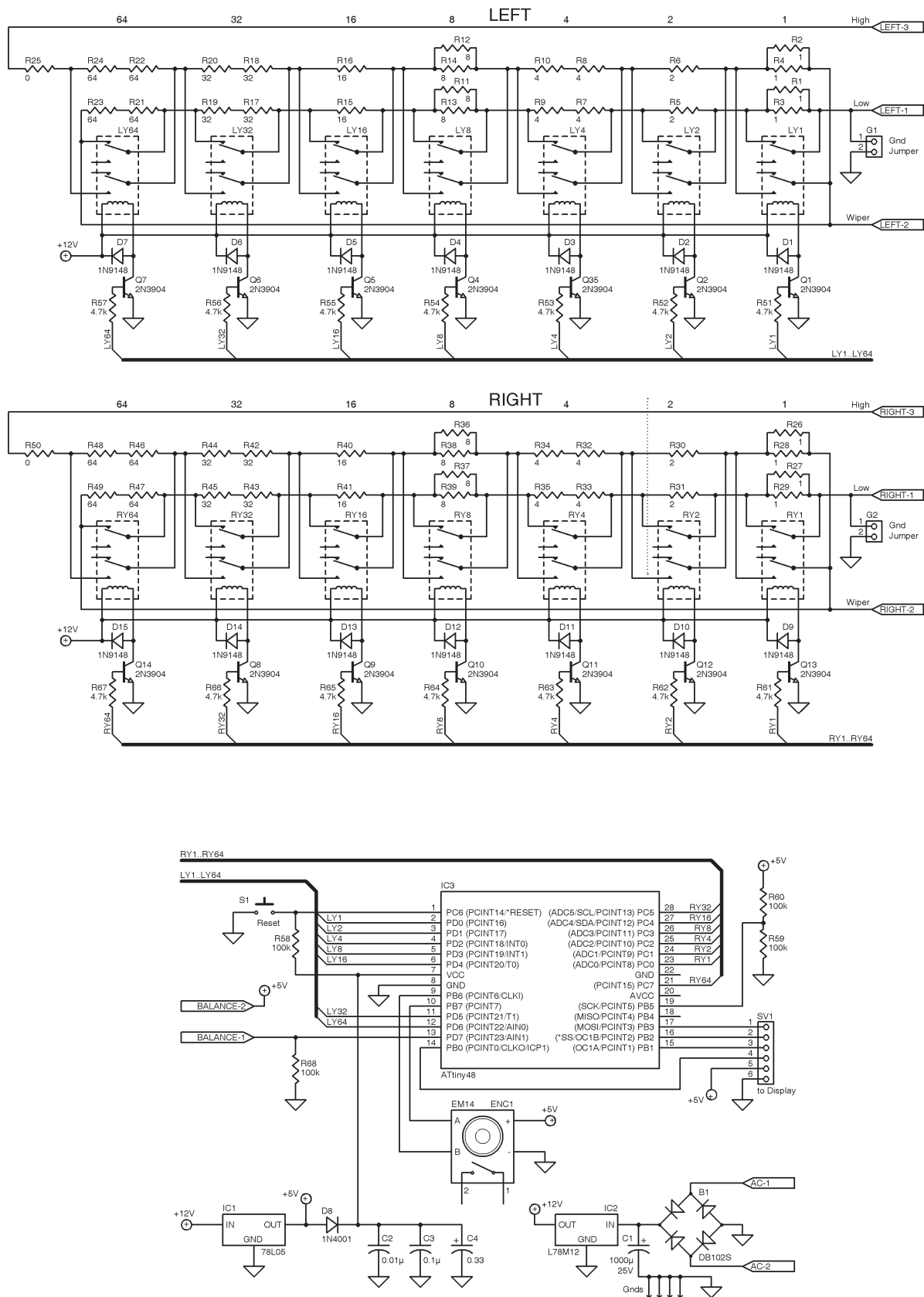


Figure 3: The schematics for volume/balance control

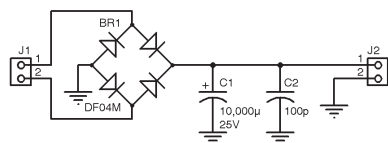


Figure 4: A schematic of the independent power supply

can reduce the maximum gain the pre-amp puts out. For example, I added a 2-k Ω resistor to my 12.6-k Ω pot to bring my gain down from 21.58 db to 20 db.

When the power to the BCD pot is off current, draw from the circuit is less than 140 nA, the 0.33-F capacitor will run for weeks holding our last level before going below the 1.8 V at which this chip is capable of running.

Lastly, the display reads a relative level, 31 can be 31 Ω or 310 k Ω , I did not see any need to set decibel output indications. My thinking is that volume is set to a specific level, and what is paramount is knowing and returning to

that level (see **Figure 2**).

SOFTWARE & OPERATION

The software was written in Atmel's Studio 4 in assembly. It starts up with an initialization of the registers driving PortC and PortD, the relay ports. The initial output is "00" (mute) the lowest setting. When the circuit is powered up for the first time it reads "-OK-" to show that the microcontroller and the display are correct. On normal turn on of the circuit, after this point, the display and relays go to the last set level (the 0.33- μ F capacitor now holds the last volume and balance position). If no other input pulses come in from the encoder the microcontroller goes to "wait" for a change in the encoder state and proceeds to update the outputs on PortD and the LCD. When the circuit is powered down, pin 19 (PortB,5), low sends the program to sleep (power down mode). In this mode, the internal 8-MHz oscillator turns off and the last output values are saved. Wake up is from a pin change (PortB,5 again), going high. This signals the program to

wait for a change from the encoder (pin 10, PortB,7), and the cycles repeat.

The output (PortD) drives the seven 2N3904 NPN transistors through seven 4.7-k Ω resistors and the transistors in turn ground the 12-V polarized relays negative lead (pin 16).

When the power to the circuit is turned off the 1N4001 diode on pin 10 of the microcontroller isolates it and the 0.33-F capacitor from the LCD and the relays. In this configuration, the only power the capacitor sees is that of the microcontroller in sleep mode. On wake up, it is recharged to 4.3 V (5 V minus the 0.7-V insertion loss from the 1N4001 diode). Do not change this diode to a Schottky diode as they have excessive leakage and the voltage drop benefit will not make up for the extra load on the 0.33-F capacitor (see **Figure 3**).

INPUT / OUTPUT BOARD

I have included independent power supplies on both the volume boards and the input/output board. This affords the builder the option of building either or both circuits, as each functions independently of the other (see **Figure 4**).

The are two things I would like to mention about the input board. First is how I use LEDs to indicate the chosen channel. I like this much better than numbers, figures, or whatever on most input circuits. Day or night it is simply a matter of looking at the interface to see which input the listener is using. The LED colors are at the discretion of the builder. Just make sure that none of the LEDs draw over 10 mA when hooked to the microcontroller. Each LED can be adjusted with its own resistor. I have included a resistor in series with the 5-V supply, which can also be changed to suit each builders specific needs.

Two, I only used one relay to isolate each input, ideally two relays will isolate both the signal and ground leads completely removing any path from any input to the output (see **Figure 5**).

CONSTRUCTION

Building the circuit is straight forward with a mix of SMD and thru-hole components. The only components not on the board are the LCD that can be



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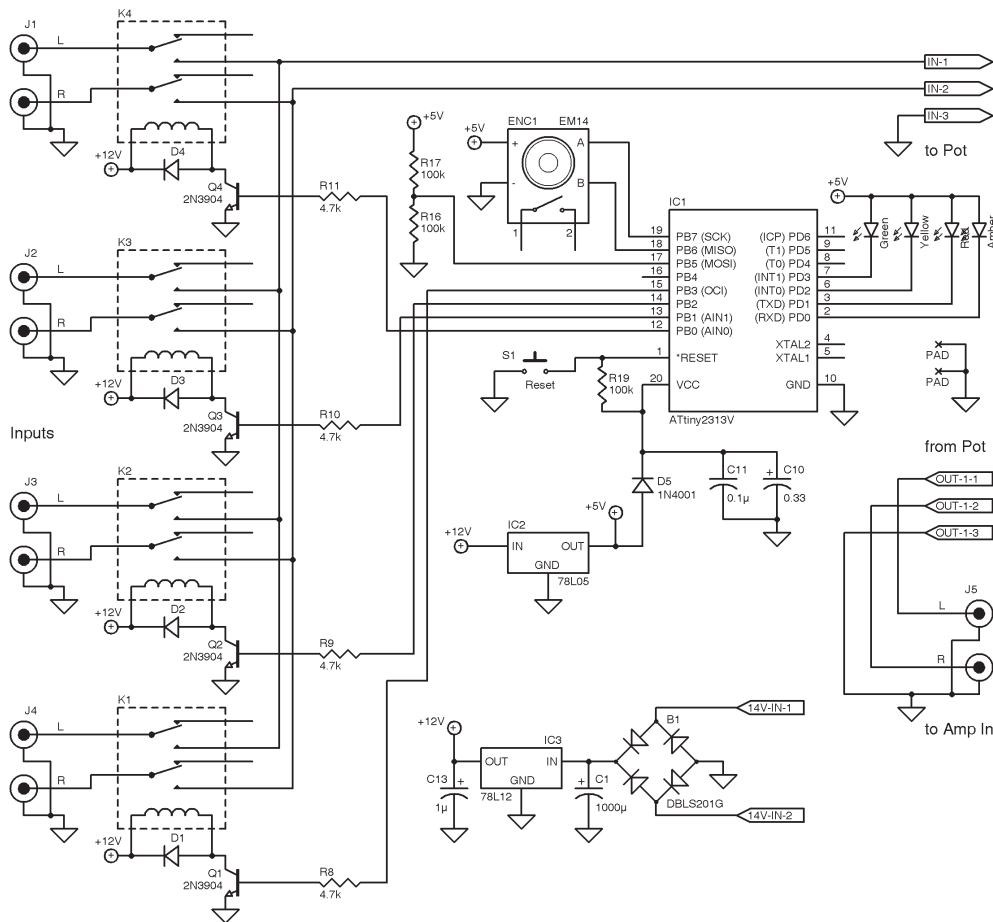


Figure 5: The input-output schematic

attached with a single six-pin 90° connector, and the encoder, which is wired through four leads to the board. The LCD can also be connected with a wire harness and extended to a front panel, if so desired. In my case, I attached the display to the main board and the encoder with wires (there are four). Build the main board first and test the voltages before inserting the encoder or the microcontroller. Pin 7 of the 28-pin IC socket should have about 4.3 V and the encoder should be just about 5 V. The positive side of the relay (pin 1) should have 12 V. (Again, these are polarized relays, they won't operate

with B+ attached to pin 16.) The LCD should also be attached after the initial testing and have 5 V on pin 1. Power for the circuit can be from a 12-V transformer at 300 mA or 15 VDC. This is handy to allow tapping off audio power when used as a stereo pot.

The display is a much nicer dot-matrix EA-DOGM-081 driven serially and not a basic seven-segment display that the original pot employed.

DESIGN PROVIDES GREAT SOUND

This design turned out exceptionally well. It works with no glitches, and its accuracy is far better than an audio circuit could ever need. I like the fact that it acts as a smooth transition when passing from one level to the next. I have never understood why detents were ever put into audio pots, especially with many only having 24 feeble levels. I always seemed to be listening somewhere between two detents, and multiple switch pots are no

better. This pot—with 127 levels and precision metal film resistors—is hard to beat for sound or facility. I hope you enjoy building and utilizing it as much as I enjoyed designing it! *aX*

Editor's note: Robert Nance Dee provided supplemental information for the project. The information, including additional figures and artwork, is available at www.audioamateur.com.

SOURCES

Agilent 34410A digital meter,
Agilent 33220A AWG
Agilent | www.agilent.com

ATtiny48 microcontroller
Atmel | www.atmel.com

SPICE
Labcenter Electronics ISIS | www.labcenter.com

Tektronix TDS 2012 Scope
Tektronix | www.tek.com

Specifications	
Noise	< 1 mV
Linearity	> 0.1%
Maximum error	< 0.1%
Current	110 mA, peak 2 channels
dB range	approx. 60 dB
Main board	3.935" × 3.945"

The Capacitance Multiplier

Making sense of a misunderstood circuit

The circuit in **Figure 1** is mentioned in “analog circuit cookbooks” and websites as a “capacitance multiplier” and is useful as a power supply low-pass filter (see **Figure 1**). It is commonly said that, like Miller capacitance, the capacitor’s value in this circuit is “multiplied” by the transistor gain (β) to lower the filter’s pole frequency. The transistor-enhanced filter does that and more. Here is an explanation of how the transistor lowers the filter’s insertion loss and pole frequency.

ANALOG CIRCUIT DESIGN

The *raison d’être* for this circuit is that the voltage drop and power dissipation across the resistor in a simple RC low-pass filter is too great for most low-voltage supplies (see **Figure 2**). A Zener diode is often added to the transistor-enhanced low-pass filter in parallel to the capacitor to regulate the output voltage. But for the time being, let’s examine the filters themselves.

Figure 3 shows both circuits used as power supply filters. We can understand how both these filters behave by calculating their voltage transfer functions, input impedances, and output impedances.

At this point, I’d like to stress the importance of putting pencil to paper when designing analog circuits. SPICE simulations and hardware breadboards are excellent verification methods, but a derived equation gives you two important things: performance sensitivity to component values and an intuitive understanding of how the circuit works. The goal of hand analysis should not be to determine exact answers (because this is impossible), but to derive approximate solutions that provide insight into how the circuit works. These

simple circuits are good practice for developing your analysis technique. In the following derivations, I’m assuming you know the basics of transistor small-signal analysis and the concept of Laplace transforms.

VOLTAGE TRANSFER FUNCTION

To calculate a voltage transfer function, maintain all the DC bias voltages and currents, apply an AC test voltage (V_{TEST}) to the circuit’s input and calculate the circuit’s AC output voltage (V_{OUT}). The voltage transfer function, which shows how the AC output voltage is a function of the AC input voltage, is a function of frequency and equals $V_{\text{OUT}}/V_{\text{IN}}$.

Because we are working in the frequency domain, we use Laplace transform versions of impedances and transfer functions; the Laplace variable(s) represents frequency dependence.

Figure 4 shows the circuit setups for calculating voltage transfer functions for both the RC filter (see **Figure 4a**) and the transistor-enhanced RC filter (see **Figure 4b**). Note that for these calculations we are assuming the power source’s output impedance is negligibly small relative to the filter and load impedances.

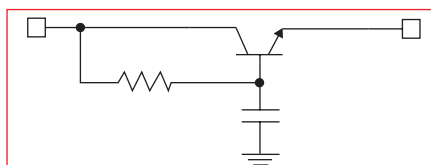


Figure 1: The “capacitance multiplier” circuit

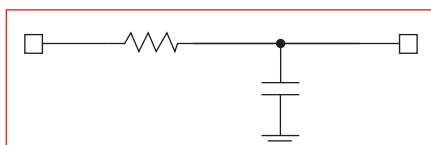


Figure 2: The simple low-pass RC filter

For the RC filter, we can use the voltage divider equation to calculate $V_{\text{OUT}}/V_{\text{IN}}$.

$$\frac{V_{\text{OUT}}}{V_{\text{IN}}} = \frac{\left(R_{\text{LOAD}} \parallel \frac{1}{Cs} \right)}{R + \left(R_{\text{LOAD}} \parallel \frac{1}{Cs} \right)}$$

Some algebraic manipulation puts the transfer function in “pole/zero form:”

$$\frac{V_{\text{OUT}}}{V_{\text{IN}}} = \frac{1}{RC} \left[\frac{1}{s + \frac{1}{(R \parallel R_{\text{LOAD}})C}} \right]$$

From this equation, we can see that the filter’s pole frequency (in Hertz) is:

$$\frac{1}{2\pi (R \parallel R_{\text{LOAD}})C}$$

At low frequencies ($s \rightarrow 0$), the capacitor is open circuit and the resistors form a voltage divider. Thus:

$$V_{\text{OUT}} = V_{\text{IN}} \left[\frac{R_{\text{LOAD}}}{R + R_{\text{LOAD}}} \right]$$

At high frequencies ($s \rightarrow \infty$), V_{OUT} goes toward zero because the capacitor shorts

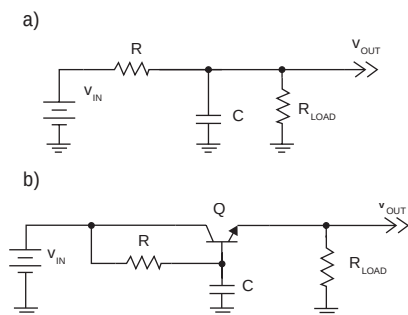


Figure 3a: The circuits used as power supply filters (a) the low-pass RC circuit; (b) the RC circuit enhanced with a transistor.

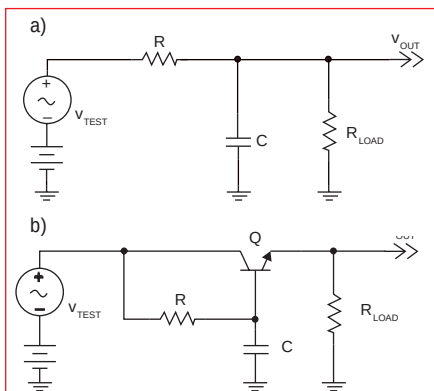


Figure 4a: These are circuit setups for calculating voltage transfer functions for the RC filter. b: These are circuit setups for calculating voltage transfer functions for the transistor enhanced circuit

across the load. Graphically, the magnitude (in decibel ohms) of this transfer function is shown in **Figure 5**.

We can use a simple bipolar transistor small-signal model (described in many texts), as shown in **Figure 6**, to calculate an approximation of the voltage transfer function for the transistor-enhanced circuit in **Figure 4b**.

Figure 7 shows the small signal circuit setup for calculating the approximate voltage transfer function of the transistor-enhanced low-pass circuit in **Figure 4b**. This voltage transfer function can be obtained by setting up and simultaneously solving two equations. The first is that the voltage V_{OUT} is equal to the currents flowing through the load times the load resistance.

$$V_{OUT} = R_{LOAD} \left[g_M V_X + \frac{V_X}{r_r} \right]$$

Knowing that $r_r \frac{\beta}{g_M}$, assuming $\left(\frac{\beta}{\beta + 1} \approx 1 \right)$, and solving for V_X yields:

$$V_X \approx \frac{V_{OUT}}{(R_{LOAD}) g_M}$$

The second equation is obtained by summing currents using Kirchoff's current law at the transistor's base.

$$\frac{V_{IN} - V_X - V_{OUT}}{R} = (V_X + V_O) C s + \frac{V_X}{r_r}$$

Substituting the first equation into the second, solving for V_{OUT}/V_{IN} , and neglecting small terms yields:

$$\frac{V_{OUT}}{V_{IN}} \approx \frac{1}{RC} \left[\frac{1}{s + \frac{1}{(R \parallel \beta R_L) C}} \right]$$

This equation is an excellent approximation of the actual voltage transfer function, if you have a reasonable idea of your transistor's β .

From this transfer function, we can see that the transistor-enhanced filter's pole frequency is:

$$\frac{1}{2\pi (R \parallel \beta R_L) C}$$

At low frequencies ($s \rightarrow 0$)

$$(s \rightarrow 0) \frac{V_{OUT}}{V_{IN}} = \left[\frac{\beta(R_{LOAD})}{R + \beta(R_{LOAD})} \right]$$

At high frequencies ($s \rightarrow \infty$) and V_{OUT} goes toward zero.

Figure 8 shows graphically the difference between voltage transfer functions for a simple RC filter and a transistor-enhanced circuit's transfer function with the same component values ($R = 1,000 \Omega$, $C = 0.22 \mu F$,

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$R_L = 47 \Omega$). Comparing the two transfer functions, we can see that the benefits the transistor-enhanced filter offers (a lower pole frequency, which results in better ripple rejection, and lower low-frequency insertion loss, which results in a lower power dissipation and lower drop out voltage) are more effective than equivalently multiplying the capacitance by the transistor's β . The reduced low-frequency insertion loss is equivalent to dividing the filter resistance (R) by β , and the lower pole frequency is equivalent to simultaneously multiplying the capacitance (C) by β and dividing the filter resistance (R) by β .

INPUT IMPEDANCE

To calculate a circuit's input impedance, leave the circuit output open and apply an AC test current (I_{TEST}) or test voltage (V_{TEST}) to the circuit's input and calculate the circuit's AC input voltage (V_{IN}), or input current (I_{IN}), respectively. The input impedance is V_{IN}/I_{TEST} , as a function of frequency.

By inspection, the RC filter (see **Figure 4a**) input impedance is:

$$Z_{IN} = R + \left(R_L \parallel \frac{1}{Cs} \right)$$

Working out the algebra yields:

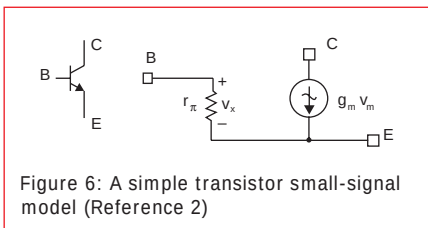


Figure 6: A simple transistor small-signal model (Reference 2)

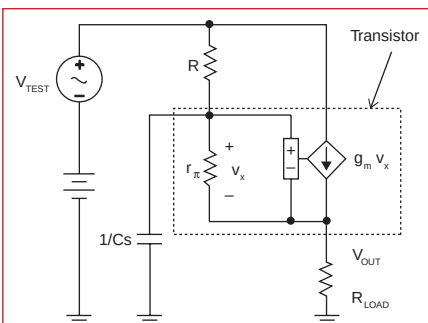


Figure 7: Small signal circuit for calculating approximate voltage transfer function for the transistor-enhanced circuit shown in Figure 4b

$$Z_{IN} = R \left[\frac{s + \frac{1}{(R \parallel R_L)C}}{s + \frac{1}{R_L C}} \right]$$

This impedance has a zero, at a frequency of:

$$\left[\frac{1}{2\pi (R \parallel R_L) C} \right]$$

and a pole at a frequency of:

$$\left[\frac{1}{R_L C} \right]$$

At low frequency ($s \rightarrow 0$), when the capacitor is high impedance, the input impedance becomes $R + R_L$ and at high frequencies ($s \rightarrow \infty$), when the capacitor is a short-circuit, the impedance is R .

Figure 9 shows the circuit schematic and the simplified small signal model used to calculate the transistor-enhanced filter's input impedance. In this case we'll apply an AC test voltage (I_{TEST}) to the circuit's input and calculate the AC input current (I_{IN}).

To find the input impedance, we need to solve two Kirchhoff voltage law equations. The first is that the voltage across the capacitor (equal to the current flowing through the capacitor times the capacitor's impedance) plus the voltage across the resistor (equal to the current flowing through the resistor times its resistance) equals the test voltage.

$$V_{TEST} = \frac{1}{Cs} [i_{IN} - i_x (1 + \beta)] + R [i_{IN} - \beta i_x]$$

The second equation is that the voltage developed across the capacitor is equal to the voltage developed across the load and r_π .

$$\frac{1}{Cs} [i_{IN} - i_x (1 + \beta)] = R_L [i_x (1 + \beta)] + i_x r_\pi$$

This equation can be solved for I_x as a function of I_{IN} .

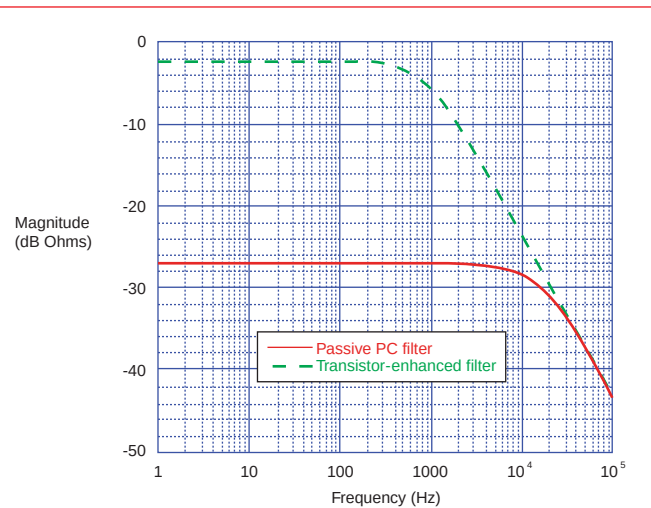


Figure 8: A comparison between voltage transfer function magnitudes of a simple RC filter and a transistor-enhanced filter with the same component values ($R = 1,000 \Omega$, $C = 0.22 \mu F$, $R_L = 47 \Omega$). The transistor collector current is assumed to be 100 mA and β is assumed to be about 50.

$$i_x = i_{in} \left[\frac{1}{Cs \left\{ (1 + \beta) \left(R + \frac{1}{Cs} \right) + r_\pi \right\}} \right]$$

This equation can be substituted into the first equation, and (after some algebra) solved for the input

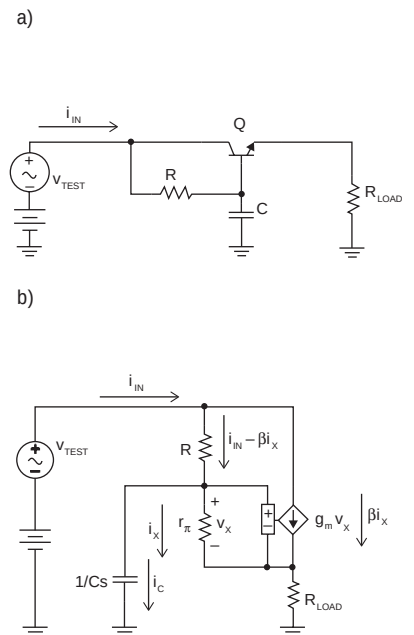


Figure 9: The circuit setup (a) and the simplified small-signal model (b). Both are used to calculate the transistor-enhanced filter's input impedance.

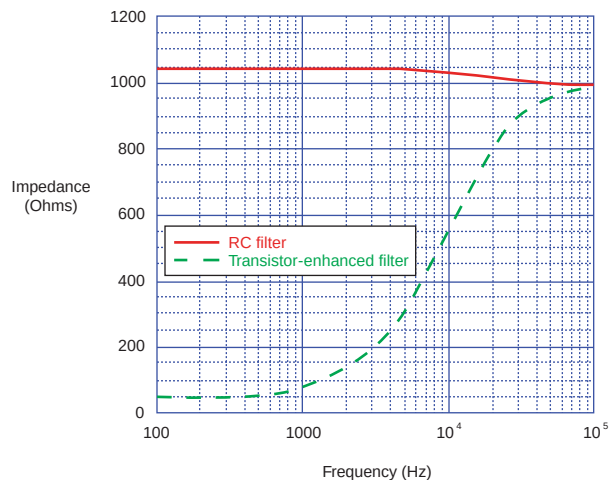


Figure 10: A comparison between the approximate input impedance magnitude of a simple RC filter and a transistor-enhanced filter. For both circuits, $R = 1 \text{ k}\Omega$, $R_L = 47 \text{ }\Omega$, and $C = 0.22 \text{ }\mu\text{F}$. The transistor collector current is assumed to be 100 mA. β is assumed to be approximately 50.

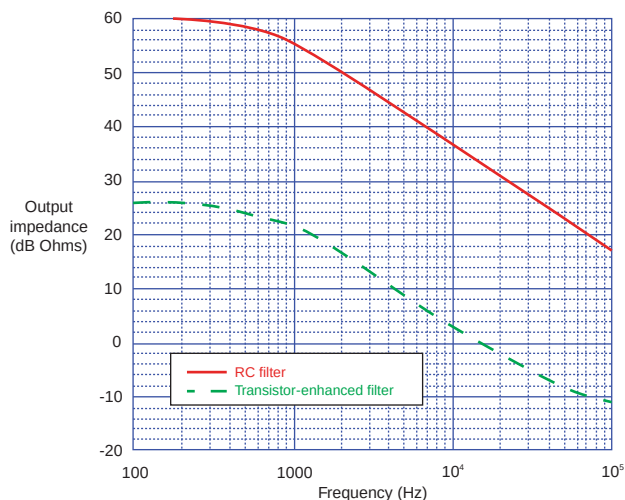


Figure 11: Comparison between the output impedance magnitude of a simple RC filter and a transistor-enhanced filter. For both circuits, $R = 1 \text{ k}\Omega$, $R_L = 47 \text{ }\Omega$, and $C = 0.22 \text{ }\mu\text{F}$. The transistor collector current is assumed to be 100 mA. β is assumed to be about 50.

impedance ($V_{\text{TEST}}/I_{\text{IN}}$). If we assume $\beta R_L \gg r_{\pi}$, which is a reasonable assumption for a power supply filter because the current through the transistor should be relatively large (and thus r_n should be relatively small), the input impedance is approximately equal to:

$$Z_{\text{IN}} \cong R \left[\frac{s + \frac{1}{RC}}{s + \frac{1}{R_L C}} \right]$$

Comparing this to the input impedance of the simple RC filter shows that, although at high frequencies ($s \rightarrow \infty$) the input impedance of both circuits approaches R , at low frequencies ($s \rightarrow 0$) the passive RC filter's input impedance approaches $R + R_L$ while the transistor-enhanced filter's input impedance approaches R_L . When using the passive RC filter, at low frequencies the power source has to feed the filter resistance (R) in series with the load, but, when presented with the transistor-enhanced filter, the power source feeds only the load at low frequencies. In addition, the transistor-enhanced circuit presents a high impedance to higher-frequency power line ripple.

Figure 10 is a comparison plot showing the input impedance magnitudes of a simple RC filter and a

transistor-enhanced filter. For both circuits, $R = 1,000 \text{ }\Omega$, $R_L = 47 \text{ }\Omega$, and $C = 0.22 \text{ }\mu\text{F}$.

OUTPUT IMPEDANCE

By inspection, the output impedance of the passive RC filter is:

$$R \parallel \left(\frac{1}{Cs} \right) \quad \text{or} \quad Z_{\text{OUT}} = \frac{1}{C} \left[\frac{1}{s + \frac{1}{RC}} \right]$$

Note that the load resistance is not included in the output impedance calculation.

To calculate the transistor-enhanced filter's output impedance, ground the circuit input, apply an AC test current (I_{TEST}) or test voltage (V_{TEST}) to the circuit's output and calculate the circuit's AC output voltage (V_{OUT}), or output current (I_{OUT}), respectively. The output impedance is $V_{\text{OUT}}/I_{\text{TEST}}$, as a function of frequency.

Using the same small-signal transistor model as before, the transistor-enhanced filter's output impedance can be found to be approximately.

$$Z_{\text{OUT}} \cong \frac{1}{g_m} \left[\frac{s + \frac{1}{(R \parallel r_{\pi}) C}}{s + \frac{1}{RC}} \right]$$

This is approximately equal to:

$$Z_{\text{OUT}} \cong \frac{1}{g_m} + \left[\frac{Z_{\text{OUT of passive RC filter}}}{1 + \beta} \right]$$

Figure 11 shows a comparison plot between the output impedance magnitude (in decibel ohms) of a simple RC filter and a transistor-enhanced filter.

Note that the transistor-enhanced filter reduces the output impedance by β , and at high frequencies, the transistor-enhanced filter's output impedance approached $1/g_m$. When using the transistor-enhanced filter, be sure to bypass its output with capacitance to lower its high-frequency output impedance.

In summary, we can see that the transistor-enhanced filter's benefits are more effective than equivalently multiplying the capacitance by the transistor's β . Adding the transistor provides two advantages over the passive RC filter. It lowers the insertion loss, and thus, there is less voltage drop and lower power dissipation across the filter. It also lowers the pole frequency, which is equivalent to simultaneously multiplying the capacitance by β and dividing the filter resistance by β , for better ripple rejection. **aX**

The Tube Sound (Part 1)

Understanding Nonlinear Distortion

A look at the aural differences arising from distortion



This oft-told story is no less entertaining for the possibility that it may not be strictly historical. It is said that an enterprising speaker designer at Oxford Speaker Co. proudly brought a new speaker sample to Leo Fender for use in his guitar amplifiers. Fender tested the speaker, and then informed the engineer that it was too clean: he would have to put some more distortion in it for it to be satisfactory.

Usually, adding distortion is not the problem; rather, it is how to remove more distortion. First, let's clarify our terms. By *distortion*, I am referring to the results of nonlinearities in the transfer function of a component in the signal chain: in this case, an amplifier. Some people define distortion as any change, other than amplitude, between the input and output waveforms of a device. This definition would also include the effects of frequency and phase-response, which are primarily linear effects. This column will discuss only nonlinear effects.

NONLINEARITY

Second, let's review the effects of nonlinearity in a system. Any form of nonlinearity will have two effects: it will cause harmonics (multiples) of the input frequencies to appear at the output; and it will cause modulation products (sum and difference) to appear. The former effect is often referred to as *harmonic distortion*. The latter is referred to as *modulation* (or *inter-modulation*) *distortion*. The specific extraneous frequencies that will be produced by a nonlinearity depend upon the nature of the nonlinearity. In other words, some forms of nonlinearity will cause proportionately more harmonic

distortion, while other forms will cause more modulation distortion. Also, the specific harmonic or modulation products created depend upon the form of the nonlinearity. For example, overdriving some amplifiers will produce strong higher harmonics (e.g., 10th, 12th, etc.) while overdriving others will produce mainly low harmonics (e.g., 2nd, 3rd, and 4th).

MODULATION DISTORTION

In general, modulation distortion is more audible and more objectionable than harmonic distortion. The reason for this is that adding harmonics to a tone merely alters its timbre, while adding sum/difference frequencies (which are likely not harmonically related to any of the program material) is like adding wrong notes in the orchestra. To be more specific, a 440-Hz tone (A above middle-C) composed of a single frequency sounds very pure, and when graphed appears as a sine wave. A triangle wave sounds a bit more razzy than a sine wave, and it is found to consist of a fundamental frequency, such as 440 Hz, and all odd harmonics, with amplitudes in inverse proportion to the square of the harmonic number. For example, a 440-Hz triangle wave having a 1-V fundamental would also contain a 1/9-V third harmonic (1,320 Hz), a 1/25-V fifth harmonic (2,200 Hz), and so on. A square wave, which sounds even more razzy than a triangle wave, has the same harmonics, but with

amplitudes inversely proportional to the harmonic number. In our example, this would equate to a 1/3-V third harmonic, 1/5-V fifth harmonic, etc. On the other hand, if two instruments are playing a middle-C (261.63 Hz) and an E above middle-C (329.63 Hz), respectively, the products of inter-modulation distortion would be the sum (591.26 Hz) and difference (68 Hz) frequencies, which correspond to something between a D and an E-flat, and something between a C and a D-flat, respectively: not only wrong notes, but out-of-tune wrong notes.

TIMBRE

In terms of timbre, a flute playing E a bit more than two octaves above middle-C almost produces a sine wave, with just a bit of second



Photo 1: A trumpet note

harmonic. A clarinet produces both odd and even harmonics, with the odd ones having a higher amplitude. A trumpet and a cornet also produce both odd and even harmonics, but with the amplitudes of the odd and even ones more nearly the same. **Photo 1** and **Photo 2** show the waveforms of a trumpet and a cornet playing notes very close in pitch. The primary difference in the construction of the instruments is that the trumpet has a cylindrical profile for most of its length, while a cornet has a conical profile. Thus, the harmonic structure is different (the cornet's fundamental is stronger in relation to the harmonics), as can be seen in the photos, and as can be heard as timbre differences. Thus, an amplifier having large amounts of harmonic distortion could make a clarinet or a cornet sound more like a trumpet.

PERCEPTION OF DISTORTION

Distortion has usually been specified as a percentage of the total output signal. Many experiments have been done to determine the minimum percentage of harmonic and modulation distortion that would be audible. The wide variation of results testifies to the fact that the experimenters were asking the wrong question. In fact, the minimum audible amount of distortion is dependent upon the frequency range, SPL, and complexity of the source material. Distortion can be perceived at much lower levels if the primary signal is a pure tone rather than if it is complex music. This fact makes it difficult not to laugh impolitely in the presence of self-proclaimed audiophiles who put on a heavy metal music CD to demonstrate their sound system.

FREQUENCY RESPONSE GRAPHS

High-end audio equipment manufacturers have traditionally included frequency response graphs in their data sheets. They have often responded to the difficulty of relating distortion specifications to subjective response by printing the distortion figures for electronic devices, in which the distortion may well be

under 0.01%. And, they omit distortion figures for speakers, whose harmonic distortion is seldom under 1% for the useful range of the speaker when the speaker is used at realistic listening levels. Yet, many experiments prove that the ear accommodates to most frequency response variations, while nonlinear distortion actually becomes more annoying with time, contributing to listener fatigue. The greater tolerance of the ear to distortion produced by speakers, compared to that produced by amplifiers, is a clue that distortion percentage does not tell the whole story.

In all natural musical instruments, the amplitude of the harmonic components decreases as harmonic number increases, although many instruments

exhibit a reduced level of fundamental, as seen in **Photo 1** for the trumpet. The relative levels of even- and



Photo 2: A cornet note

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odd-order harmonics varies, depending upon the instrument. Thus, while there has been argument about whether even- or odd-order harmonics are less objectionable (some even say, more desirable), there is universal agreement that any deviation from the more-or-less regular decrease of amplitude with increasing harmonic number, is particularly annoying. In an earlier column, mention was made that some 1930s motion-picture companies preferred single-ended triode power amplifiers rather than push-pull ones. This was because the early push-pull amplifiers suffered from crossover distortion, which produced unnaturally large amounts of higher-order harmonic distortion, as well as more intermodulation distortion. After a few years, Class AB biasing was introduced, greatly reducing this problem, and push-pull amplifiers gained almost universal acceptance.

SOLID-STATE AMPLIFIERS

When the first solid-state amplifiers were introduced in the mid-20th century, history repeated itself with a vengeance. Not only did many of these new amplifiers have crossover distortion issues, but the crossover distortion was of a nature that produces a far higher proportion of annoying high-order harmonics. After a few years, new circuitry was developed to eliminate or greatly reduce crossover distortion in solid-state amplifiers, but then the onset of linear integrated circuits was upon us. Most of those early linear ICs had a high enough gain-bandwidth product (GBW) at low-signal levels to maintain a good frequency response, but researchers, notably Matti Ojala of Finland, determined that the most common linear ICs of the day had such a low GBW at high-output levels that they could not make the negative feedback signal keep up with the signal being amplified. This phenomenon caused *slewing-induced distortion* (SID), also called *transient intermodulation distortion* (TIM), a distortion that was difficult to measure but easy to hear when the amplifier was

operated at high output levels. Like crossover distortion, SID produced unnaturally high levels of high-order harmonics as well as annoying levels of modulation distortion. Subsequent improvement in the *slew rate* (number of volts that an IC's output could change in one microsecond) has virtually eliminated this source of distortion in well-designed solid-state amplifiers.

THE GREAT DEBATE

This brings our background up-to-date. However, in spite of the very significant advances in amplifier technology, many people are convinced that they can hear differences between solid-state and hollow-state amplifiers. This argument has been dubbed in more than one article as "the great debate." It was discussed as recently as October 2011 in a paper in the *Journal of the Audio Engineering Society*.

The debate can be addressed in four parts:

- Do well-designed amplifiers (solid-state versus tube, or even different designs of either) differ audibly when operated within their linear (not overdriven) ranges?
- Do they differ audibly when overdriven?
- Are amplifiers routinely overdriven in normal use?
- And what components of these differences are more desirable than their counterparts?

Reasonable agreement can be found in the answers to the second and third questions:

- Different amplifiers produce different distortion signatures when overdriven, and greater propor-

tions of intermodulation distortion and/or of high-order harmonic distortion are definitely audible.

- Musical instrument amplifiers are indeed routinely overdriven. Amplifiers used for home music listening and/or home theater may be routinely overdriven, depending upon the amplifier, speakers, and listening habits of the user. Amplifiers used in good professional systems are seldom overdriven, as they are usually protected by compressors or limiters. Although compressors and limiters themselves can be audibly detectable in use, they do prevent the amplifiers from being overdriven if the system is properly set up.

The measurable differences in distortion products of different types of amplifiers will be discussed in a future column, as will the question of audible differences among amplifiers that are not overdriven, and the question of whether some forms of distortion may be perceptually pleasing will be addressed as well.

OVERDRIVING AMPLIFIERS

We can elaborate on the question of overdriving home-music and home-theater amplifiers. A typical speaker used in such a system probably produces about 85 to 90 dB_{SPL} at a distance of 1 m when driven with 1 W. If two speakers are reproducing the same waveform, in phase, in the same space, and are at least a few feet apart, the total SPL will be about 3 dB more. In many cinemas, 85 dB_{SPL} is the target average sound pressure level. Let us assume the same target level for your living room or home theater. Now we must correct for the fact that you are probably listening from farther away than



Photo 3: A trumpet playing part of a Haydn Concerto

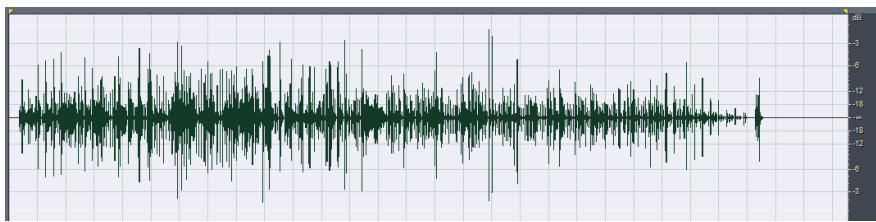


Photo 4: An African drum track

1 m. If we assume a listening distance of 3 m, the level would be reduced by about 10 dB in anechoic or outdoor conditions, or probably about 6 dB in a living room. Next, the imponderable: crest factor. **Photo 3** and **Photo 4** show waveforms from a Haydn trumpet concerto and an African drum selection, respectively.

Notice that the highest peaks of the waveforms are considerably above the average levels: about 9 dB for the Haydn and 18 dB for the drums. The ratio of the peak amplitude to the average amplitude is known as the "crest factor." Crest factor is a real wild card in home audio: some CDs are compressed so that the crest factor is near 0 dB, and in some ac-

tion movies with loud gunshots and explosions, it can be quite high. If we use 20 dB as a generous estimate for crest factor, then to produce 85 dB at a listener position 3 m from the speakers, using a speaker having a power sensitivity of 85 dB at 1 W at 1 m, we will need a power amplifier producing $6 \text{ dB} + 20 \text{ dB} = 26 \text{ dB}$ more than 1 Ω , or 400 W, if we are to avoid clipping on peaks. While many home-audio amplifiers can produce 400 W, most cannot, so peak clipping can be expected unless:

- The system is used mostly for pipe organ music or heavily-compressed pop music

- The speakers are unusually efficient
- The usual listening level is low

My late friend Paul Klipsch enjoyed listening to classical music on Klipschorns (112 dB at 1 W at 1 m) using a Brook amplifier with a push-pull triode output stage producing a maximum 10 W output. This system could produce 90 dB_{SPL} at 1 m, allowing for a 20 dB crest factor. This corresponds approximately to the highest average level normally encountered within the orchestra.^[1] Notice the effect of speaker efficiency!

Having laid the groundwork for understanding aural differences arising from distortion, we will continue next month by looking at the output waveforms of various amplifiers operated with slight clipping. *aX*

RESOURCES

[1] Royster, J.D., Royster, L.H., and Killion, M., "Sound Exposures and Hearing Thresholds of Symphony Orchestra Musicians," *Journal Of The Acoustical Society Of America*, June 1991.

MEMBER PROFILE

Education: BSEE, Washington University, 1962

Most Recent Purchase: Ken bought two Lepai TA 2020 Class C amplifiers from Parts Express. He said the audio performance of the amps is great for the money.



Ken Bird holds a 12" woofer

Occupation: Retired

Member Status: Ken has been a subscriber to *audioXpress* since 1989.

Affiliations: He is a member of the ARRL-Ham Call. His amateur radio (ham) call sign is W0KLB.

Audio

Interests: Ken enjoys speaker

and tube amplifier building, the Ham radio, and restoration of old radios.

Current Audio Projects: Ken restored a 1983 JVC RS33 receiver and two Altec Series 9 speakers for a friend. He is currently working on a three-way speaker system with dual 8" woofers in an unusual cabinet design called the Cathedrals. He said he might write an article for *audioXpress* about this project.

Dream System: Ken's dream system consists of two of the original 1953 Electro Voice Patrian enclosures, and they would be driven by a McIntosh MC275 amplifier. He added that he also needs to dream for a room that will be big enough to accommodate them.

Member Name: Kenneth L. Bird

Location: Lincoln, MO

50-W BJT Stereo Amplifier

An easy-to-build amp that generates quality sound

It would seem when it comes to published audio power amplifier projects these days, the DIYer is limited to designs using vacuum tubes or an esoteric topology. This is unfortunate since the 50-W bipolar transistor amplifier described here yields very good performance, sounds good, and is easy to build (see **Photo 1**).

Two amplifier boards with heat-sinks, a power supply, appropriate speaker protection circuitry, and an enclosure are all that would be needed to create a state-of-the-art stereo amplifier.

I built this amplifier so that I could listen to music while working and exercising in my basement lab. While



Photo 1: The amplifier in its enclosure

50 W is overkill for my application—10 W is probably more than enough—it should be a convenient level for many others. Although my ears are nowhere near golden, as an engineer I wanted performance numbers which would imply that I had a wire with gain, (i.e., no audible distortion, noise, or any other coloration or artifacts). Any processing that I may wish to do, although I haven't, would take place before this amplifier. The amplifier itself and the speaker protection circuit were built on PCBs that I designed. The power supply circuitry is discrete (see **Photo 2** and **Photo 3**).

POWER AMP CIRCUIT

The amplifier is a three-stage design, along the lines as those explored and documented by Self and Codell.

[1] [2] The first stage has a differential input which, when properly balanced, cancels even order input stage distortion products. The differential input also allows for a convenient global feedback point. The input stage is followed by the voltage amplifier stage which amplifies the input stage signal and can swing nearly rail-to-rail. To maximize the output swing capability, this stage is operated against the negative power supply rail. The output stage is a current amplifier with near unity voltage gain to drive the low impedance of the speaker load. The global feedback network routes a portion of the output signal to the input stage to reduce distortion, set the overall system gain, and to lower the amplifier's output resistance. In short, this amplifier is effectively a very low-distortion, 50-W op-amp.

Referring to **Figure 1**, Q1 and Q2 are the differential input transistors. The input signal is applied to Q1 while the feedback signal is applied to Q2. R2 and R3 apply local feedback to linearize the hyperbolic tangent transfer characteristics of this stage. Q7, along with Q8, form a current source for the differential pair. A current source, as opposed to a simple resistor, provides a high degree of isolation from the positive supply rail. The value of the current source sets the

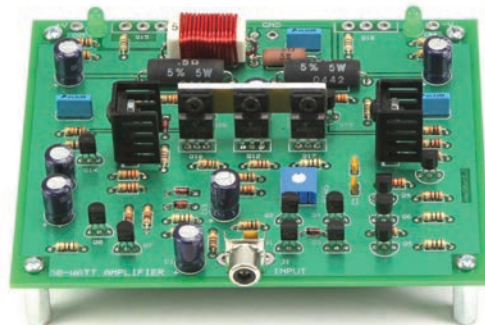


Photo 2: The power amplifier without final devices and heatsink

quiescent bias point for the differential pair and is approximately $Q7's V_{be}/100\ \Omega$, which in this design is about 6.5 mA. This moderately high value of bias helps keep the slew rate up since this input stage has to charge and discharge the feedback compensation capacitors which are in the following stage.

The collectors of Q1 and Q2 drive the emitters of Q3 and Q4 to turn the input transistors into cascode pairs. The cascode pairs act similar to a simple common emitter amplifier stage except that its feedback is very much lower and its output resistance is much higher. I found

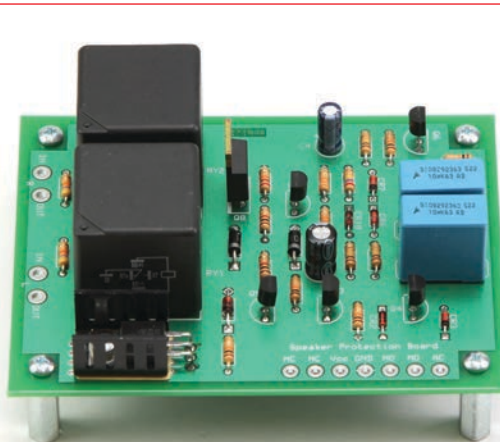


Photo 3: The speaker protection board

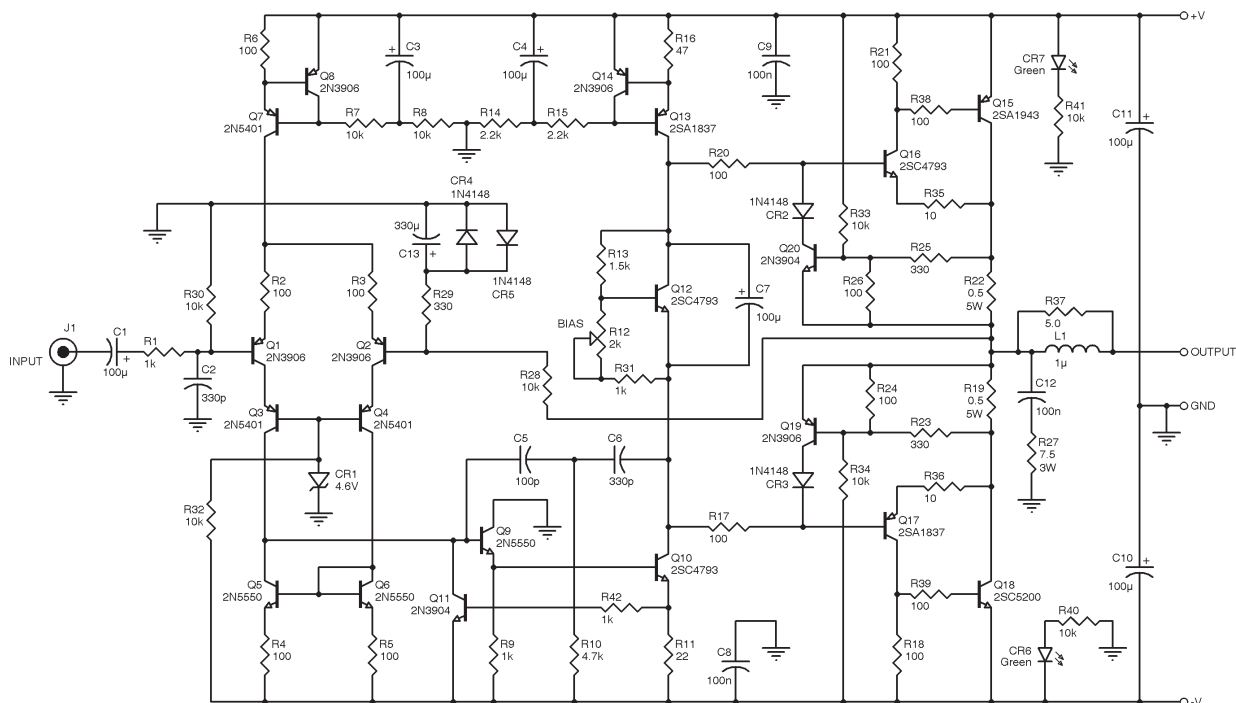


Figure 1: PA board schematic

that cascoding was needed to provide isolation from the negative supply rail. Without it, the noise and distortion floor was dominated by negative supply rail hum, probably inaudible, but an irritant to me. The load for the cascode input pairs is a differential to single-ended converter formed around current mirror transistors Q5 and Q6. The current mirror provides a high-impedance load to the input stage while helping keep the input stage balanced. R4 and R5 improve the matching of the current mirror by masking any I_C versus V_{be} variations between Q5 and Q6.

Q9 and Q10 together provide the voltage amplification required for the amplifier. The signal at Q10's collector can swing nearly rail-to-rail to drive the output stage. The frequency compensation, required for stability, is formed around C5, C6, and R10. These components cause the open-loop (feedback removed) frequency response to begin rolling off at a higher frequency than the more common single-capacitor dominant-pole compensation method. Even though the rolloff is at a faster rate until a fairly

high frequency is reached, the result is closed-loop stability with greater open-loop bandwidth. The end result is reduced distortion at the higher audio frequencies. To maintain high gain, the load for Q10 is an active current source using Q13 and associated components. The idle current in this stage is approximately 13 mA. To provide bias for the complementary output stages, an adjustable, temperature-stable voltage drop is created between the top and bottom halves of the output stages by Q12 and Q12's biasing resistors.

The output stage is built as two, mirror-imaged, complementary feedback sections. Each section consists of a transistor (Q16/Q17), which acts as a voltage-follower to the load and a complementary transistor (Q15/Q18), which senses the current of the voltage-follower and supplies the bulk of the total load current. Q19 and Q20, along with output current and voltage current sensing resistors, act to keep Q15 and Q18 operating within their safe operating area. Global feedback is provided by R28 and R29. With the values shown, the closed

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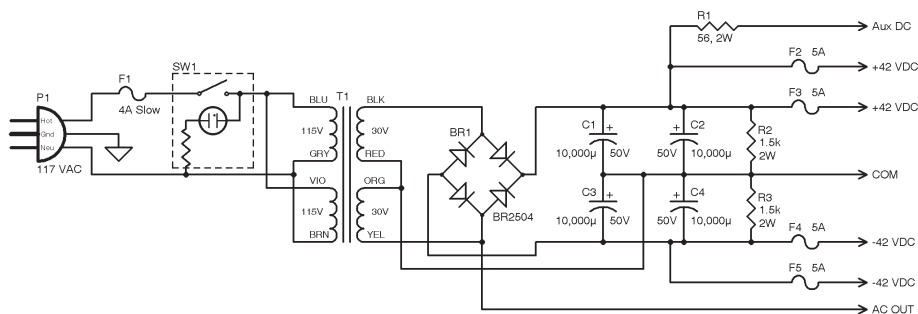


Figure 2: Power supply schematic

loop voltage gain is 31 (30 dB). Capacitor C13 reduces the closed loop gain to unity at DC so that any DC offsets between the two transistors in the differential amplifier stage do not get amplified. Since no low-frequency servo is used, there will be a small DC voltage (typically millivolts) at the speaker terminals caused by the base bias current of Q1 flowing through R30.

PERIPHERAL CIRCUITRY

For an overall product, a power supply is necessary. My supply is a simple

full-wave bridge rectifier followed by reservoir capacitors (see **Figure 2**). The total capacitance is not as large as it could be, but the power supply rejection of the amplifier is such that more is not needed. The transformer may be overkill on its kVA rating. It doesn't get the least bit warm in operation, but its turn-on surge is noticeable. A smaller transformer with its lower impedance could probably be used without ill effect. For my safety when troubleshooting, bleeder resistors were placed across the supply rail outputs to bleed the voltage

off after it was powered down. Both supply rails are fused with automotive-type DC fuses.

Although my speakers are not horribly expensive, they are pricier than the amplifier itself. Therefore, I designed and added a speaker protection board (see **Figure 3**). This board is universal enough to be used on any amplifier up to 100 W or more and which can provide a DC voltage greater than 30 V. A voltage regulator formed around Zener diode CR1 and transistors Q1

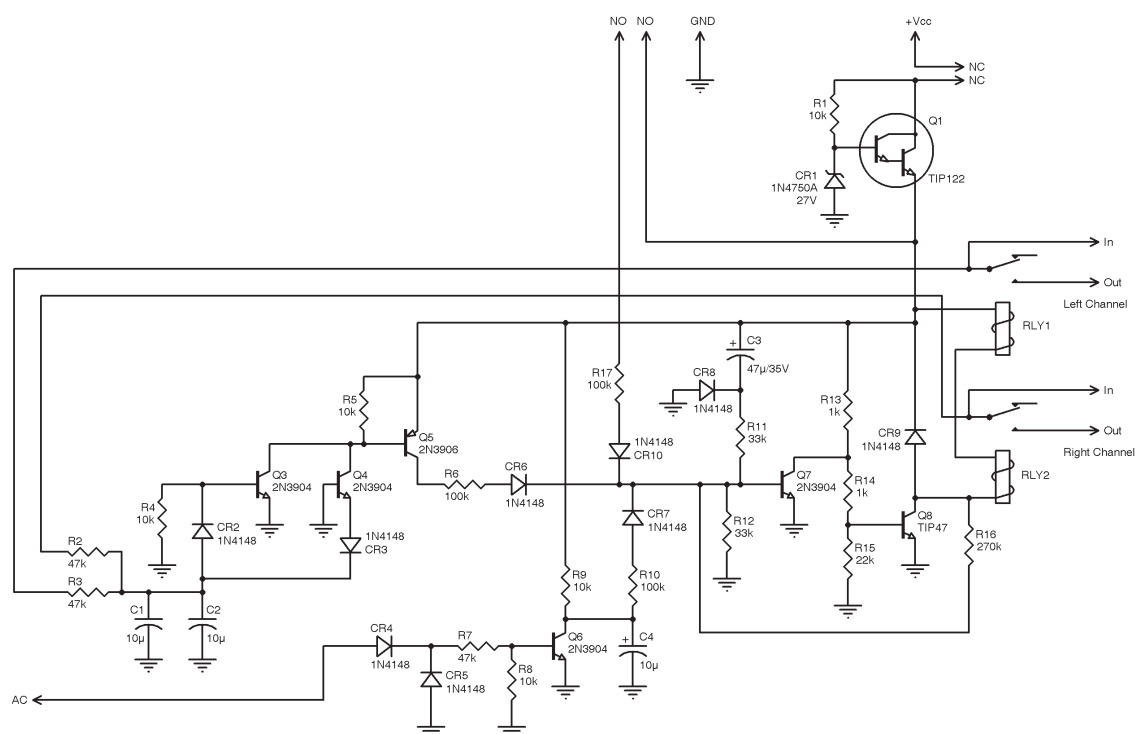


Figure 3: Speaker protection board schematic

Q5 is included to sense the presence of a DC component on the output waveform and turn off the speaker relays if DC is detected. Note the inclusion of R4. For positive DC levels, the base of a transistor is driven while for negative DC levels an emitter is driven. This would ordinarily result in a different sense point for positive versus negative DC components. This subtle point seems to be missed in many protection circuits that I have seen. R4's inclusion makes the sense points nearly equal. Also, missing in many speaker protection circuits is the use of hysteresis in the relay drive circuitry. R16 was added to make the relays switch on and off decisively with a snap action. Although not used in this design, the speaker protection board has provision for either normally-open or normally-closed thermal switches or other sensors.

Because of the modest power level of this amplifier and the fact that it most likely will never be operated at elevated temperatures, I decided not to include a thermal switch on the heatsinks, but there is provision for them on the protection board. Above a V_{CC} of about 35 V or so, the power dissipation in the voltage regulator can become excessive, even with a heat sink. A resistor of approximately 8 Ω per volt above 35 V should be connected in series with the V_{CC} input to reduce the voltage going to the board. I added a 56- Ω , 2-W resistor over on the power supply area to reduce the 42 V from the power supply to 35 V for the speaker protection board.

CONSTRUCTION & OPERATION

The power amplifiers themselves are constructed on a double-sided PCB that I laid out. Since I'm an RF engineer by profession, I naturally used some of my RF layout techniques for this project. The top (component) side of the board is an almost totally solid ground plane. All but one of the circuit traces are on the bottom side. Although I use SMD technology at work, for this project I chose leaded components for easy troubleshooting and replication by the average DIYer.

Even though it cost more, I used solder mask on both sides and silk screening on the top. The one thing I learned after the first layout is that grounding is critical, even with a solid topside ground plane. Even though there are only three ground points on the board—input, output, and power—the input connector had to be placed

along the board centerline rather than an edge. I was most likely being overly critical, but I could measure a difference so I moved the connector for the next pass at the layout.

In my design, the final output transistors, Q15 and Q18, are mounted on a heatsink large enough to handle rated power, 50-W average with

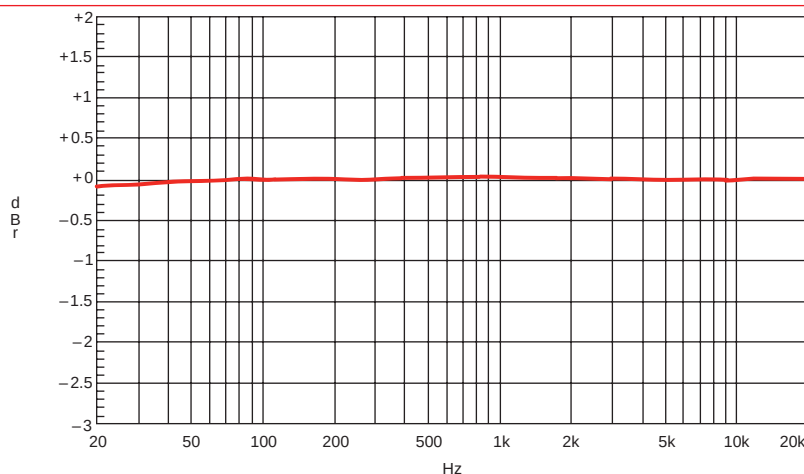


Figure 4: The frequency response

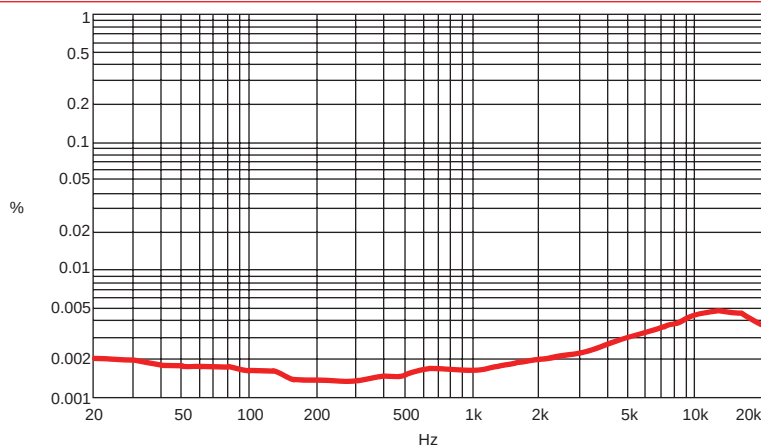


Figure 5: THD versus frequency at 50 W

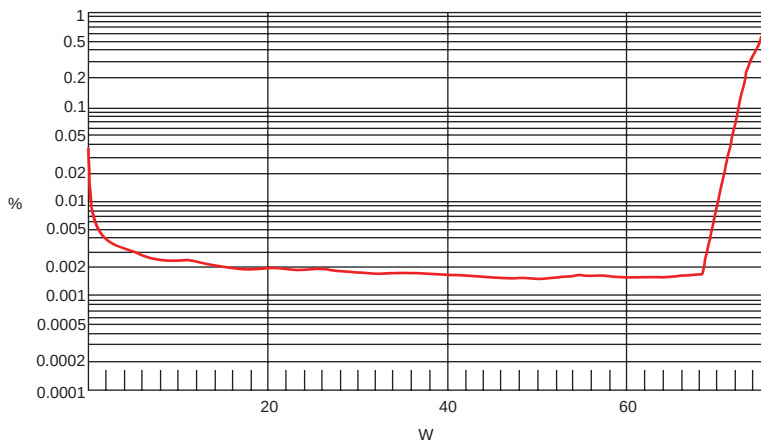


Figure 6: THD versus power at 1 kHz

Specifications	
Output power	50 W minimum
Gain	30 dB
Frequency response	± 1 dB, 20 Hz to 20 kHz, ref. 1 kHz
Total harmonic distortion	0.002%, 1 kHz, 50 W
SMPTE DIN IMD	
1:1	0.010%, 50 W
4:1	0.003%, 50 W
Noise (22 kHz bandwidth)	-100 dB, ref. 50 W
Damping factor	150
Input resistance	10 k Ω

a sine wave out of clipping, at room temperature without getting too hot. Although the output transistors don't need to be temperature compensated, the drivers, Q16 and Q17, do. The compensation is provided by mounting Q16 and Q17 on a common heat spreader with Q17. There's minimal heat to dissipate, but an effort should be made to keep the temperatures of all three transistors the same. Q10 and Q13 both dissipate over a quarter

of a watt and each must be mounted to individual heatsinks. The differential input transistors Q1 and Q2 should, ideally, be kept at the same temperature to minimize changes in the output DC offset, but I didn't make a special effort to do so except to keep them close together on the PCB. Testing with a distortion analyzer shows that the lowest distortion is achieved when the bias potentiometer, R12, is set such that the voltage across

the output resistors R19 or R22 is 50 mV. The voltage on the two resistors should be the same within the tolerance of the resistor. In reality, the setting of bias is not especially critical. The distortion only changes by 0.002% or so when going from minimum to maximum pot setting. In lieu of a distortion analyzer, setting the bias pot to mid-rotation is more than adequate.

PERFORMANCE

The performance of the overall amplifier in its enclosure and with its peripheral circuitry is shown in **Figure 4**, **Figure 5**, and **Figure 6**, and summarized in **Table 1**. At rated power and 1 kHz, THD is close to 0.002%. The limit of my Audio Precision System One is not much lower than that. Higher audio frequencies, 5 kHz or so and above, show higher distortion due to the open loop gain decreasing with frequency. Lower power levels show slightly higher distortion due to the noise floor of the amplifier and measuring equipment (see **Figure 6**). The rolloff in distortion in **Figure 5** is due to the 30-kHz low-pass filter of the audio analyzer. In short, this amplifier's numbers are impressive. I believe that with all of the measuring capability we now have an amplifier, which yields good numbers, will sound good as well. The distortion of the stand-alone amplifier is closer to 0.001% due to the ground loops and coupling to the power supply that occur when mounted in the enclosure.

HINTS AND KINKS

In the course of building this amplifier, I've learned a few things, some the hard way. Even when I knew better, in my haste I made some mistakes. From my experiences, I can offer the following advice:

Don't even think of hooking up a set of speakers to a direct-coupled amplifier until, and unless, a speaker protection circuit is in place. This circuit must disconnect the speakers when DC is detected. Almost all failures of the amplifier will drive the output to a supply rail. Think what that could do to a speaker and what it would sound like. Along this line,



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buy an audio dummy load and use it. (Parts Express sells them.) Make sure you are happy with the amplifier and have a speaker protection system in place before connecting your expensive speakers to the amplifier. Even inexpensive speakers are costly when compared to the cost of a protection board.

Put bleeder resistors across the reservoir capacitors and allow the capacitors to completely discharge before connecting any grounded instrument, like a soldering iron tip, to the supply rails. Even when the supply rails are fused, traces can burn up and sparks fly before the fuse blows. I added power rail LEDs on the amplifier PCBs to show the presence of supply voltage on the boards.

The wires from the output to the speaker terminals should be heavy gauge and twisted. To minimize a magnetic ground loop from the transformer to the inputs when both channels are connected, I had to twist the cables from the chassis input connectors to the amplifier input connectors. Also, I connected the grounds of the amplifier input connector together with a short piece of wire.

Buy extra semiconductors, you'll need them. Fortunately, the semiconductors I used, including the output devices, are not very expensive. Interestingly, the output devices had the lowest failure rate during development. Do keep them heatsunk, however.

Don't forget the insulating hardware on the regulator pass transistor/heatsink on the speaker protection board. On all parts connected to heatsinks, use thermal compound, but don't overdo it.

If using a PC board, buy a (shameless plug) Hakko 808 desoldering tool. It'll be the best \$180 you'll ever spend.

Don't hook the power supply rails up backwards! If you do, plan on replacing every semiconductor and polarized capacitor on the board and also some resistors. Fuses won't protect the board, although they will blow. In fact, rail fuses don't protect from much of anything except from fire. Consider protection diodes on

the supply rails, at least until the project is finished.

If smoke pours out of R27, the amplifier is oscillating at a high frequency or the input is being driven by a high frequency source. My speaker protection board does not detect oscillation, although the output inductor may reduce its impact on the tweeters. In any case, determine the cause and get it fixed before reconnecting the speakers.

The grounding location and wire routing has by far the most effect on distortion and residual hum, both of which show up in a distortion measurement. The power amplifier board by itself yields exceptional performance. Maintaining that kind of performance in a case with the power supply and other circuitry is not at all trivial and took the majority of my time on this project. My enclosure was black, anodized aluminum and its surface is an insulator. This insulated the separate panels from each other. However, a monolithic enclosure may require a completely different grounding system to yield good results.

And finally, have fun and keep the DIY spirit alive and well. Next up for me is to design an S/PDIF DAC and a good pre-amplifier. *aX*

REFERENCES

[1] Self, D., *Audio Power Amplifier Design Handbook*, 5th edition, Focal Press, Oxford, 2009.

[2] Codell, B., *Designing Audio Power Amplifiers*, McGraw-Hill, New York, 2011.

SOURCES

Resistors, capacitors, and semiconductors

Digi-Key Corp. (distributor)
www.digikey.com

Power transformer, dummy load, and bridge rectifier

Parts Express
www.parts-express.com

Enclosure and heatsink for finals

Par-Metal Products, Inc.
www.par-metal.com

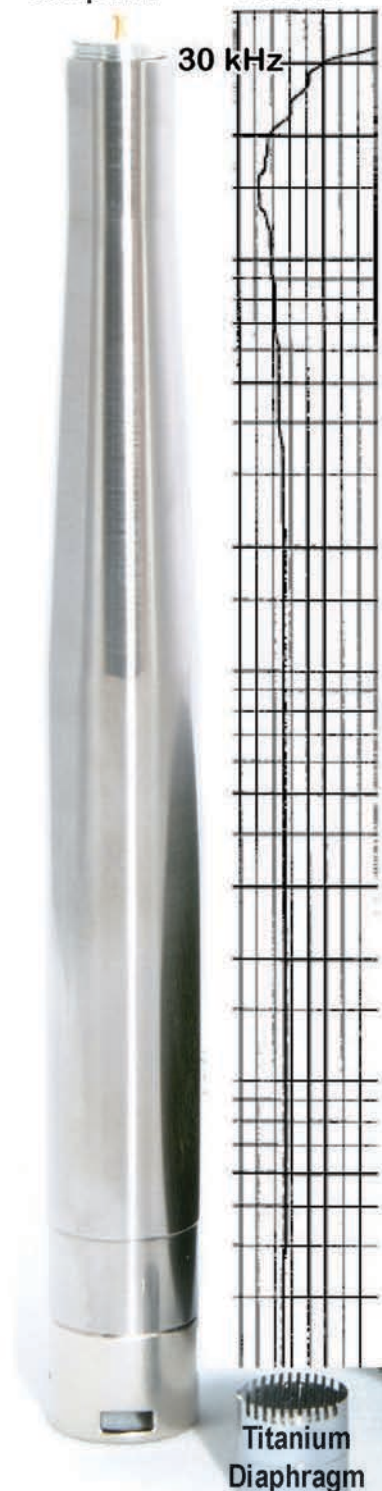
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SHANNON BECKER: Where do you live now?

MIKE URBAN: Windsor, CT

SHANNON: How did you become interested in electronics?

MIKE: At a young age (early 1960s), I received the Rocket Radio crystal set as a gift.

I clipped it onto an old house radiator pipe and listened to AM radio for free—no batteries needed. It was amazing trying to figure out how that worked! Later in life (mid-1970s), I found a large table radio for \$25 at a garage sale. It was a beautiful Grundig international radio with AM, shortwave, and FM. It was housed in a gorgeous blond wood cabinet with a very large, lighted glass dial, vacuum tubes, and awesome sound coming from two large speakers. Here, I discovered the superior quality of tube sound amplification. The radio sounded as real as live music which had a warm, rich, full-bodied tone.

SHANNON: What first attracted you to radios?

MIKE: I have always had a particular interest in radio, which provided hours of entertainment listening to local sports broadcasts—this was before cable TV became popular (early 1970s)—as well as, tuning to programs coming in via shortwave radio from all over the world. After graduating from high school, I was trying to figure out what to do with my life. I ventured into a few electronics classes at a local community college. I liked it enough to go further, earning a four-year engineering degree and obtaining a FCC HAM radio license.

SHANNON: You now own your own vintage radio business, Urban Radio and Vintage Hi-Fi. What made you venture out on your own?

MIKE: After graduating from college, I embarked on a 20-year corporate career developing computer and audio/video products for General Electric, Olivetti, Apple Computer, Quantel, Avid, and Barco. Overtime, with too many corporate downsizings and mergers, I needed to get control of my life and with a very supportive spouse, decided the time was right to do something that I truly enjoy, combining my passion for audio with my technical skills. Vintage electronics appeared very collectible with the invention of eBay (1990s), so I went in search of garage and attic treasures. Later, I found working at radio/TV repair shops while getting through college had provided a second benefit in life. Special skills were needed to get many of these electronic items working. Now that these items are very desirable to own and listen to, they need to be repaired or restored to their former glory. The time was right to start advertising, therefore, www.urban-antiqueradio.com came to be (see **Photo 1**).

SHANNON: Tell us about your work. Is there an actual storefront or is most of your business done via the Internet?

MIKE: Starting in a one-bedroom apartment, the business started to grow and the landlord finally had enough (as I took over his basement). So it was necessary to find another location. We then settled into a commercial storefront (for about eight years) and while the economy had its ups and downs decided to transition from retail and repair to mostly repair/restoration work. There were many customers in search of a repair shop, either for sentimental reasons (their heirloom item) or for the sheer pleasure of listening to quality audio. Again, the business grew to the

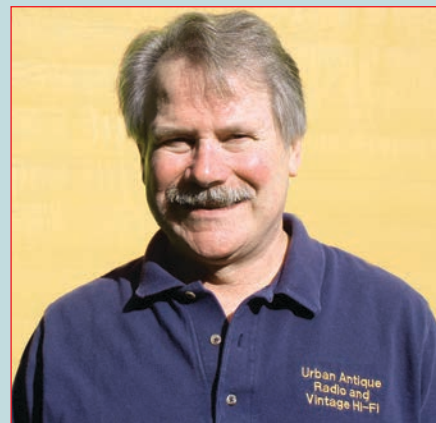


Photo 1: Mike Urban, owner of Urban Antique Radio and Vintage Hi-Fi

point of needing more space as all (or almost all) of the repair shops went out of business due to a “dying art” or to many new items becoming disposable (see **Photo 2**). So, my wife and I went in search of an antique home, as we have become preservationists! Settling in Windsor, CT (the Hartford area) seemed obvious as many customers travel between New York and Boston, plus there is a great radio museum here (more on that later). As for more distant customers, we get those items shipped to us for repair.

SHANNON: Where do you find vintage radios?

MIKE: There are many radio swap meets, HAM swap meets, specialized auctions, where you can find almost anything in the vintage electronic category. The best resource is the Internet. For example, to find upcoming events, try the Antique Radio Classified (www.antiqueradio.com) or google “ham radio shows hamfests”. There seems to be one or more event on every weekend (somewhere in the country) usually during the better weather months (some are outdoor tailgating events). Then, there are the car shows, too. We repair a lot of vintage car radios. We find collectors of one category often



Photo 2: Mike Urban's workbench



Photo 3: The AK84

branch out into another category!

SHANNON: How many radios do you have in your personal collection?

MIKE: Too many. I am always finding something interesting. Each one is like a work of art (see **Photo 3**). Then, it's time to thin out as we can only listen to a few at a time and I don't want to move again!

SHANNON: What has been your most exciting vintage radio find to date?

MIKE: The thing about collecting (it is addictive, too—I should start ARCA, Antique Radio Collecting Anonymous—just kidding but there should be one!) is finding yet another item you haven't seen before or finding an item in pristine condition. This keeps us excited for the next time, as to what may be found. As for my greatest find, actually it was a very rare pre-war TV (a full-length story can be found on my web site: www.urban-antiqueradio.com). Currently, my passion for vintage Hi-Fi has grown above and beyond radio. Finding the ultimate listening experience is never ending!

SHANNON: I understand there is a vintage radio museum in Windsor, CT. Is it open to the public? And how did it come to be?

MIKE: There is a Vintage Radio & Com-

munications Museum here in Windsor, CT (www.vrcmct.org). It is open to the public on weekends (and they are in search of more volunteers, too), has regular swap meets, and has a long history going back to 1990. It might be best to visit the website to get all the details.

SHANNON: What do you find to be the most challenging when it comes to refurbishing antique radios? What is the most rewarding?

MIKE: Parts! On the inside, you will find electronic, electro-mechanical, or

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all mechanical parts, and on the outside, you will find cabinet details, wood, metal, plastic, different knobs, bezels, speaker grille cloth, and so on. There are also many parts available but there are many custom parts for any given repair item (transformers, coils, controls, knobs, etc.). Just like a car, manufacturers would change the design (inside and out) every year to entice customers to purchase the next new thing.

Multiply that with decades of brands and models makes far too many to count. So we have to go in search of old stock (new and used), find parts sets, find reproduction parts, or rebuild/recreate the part. The most rewarding aspect is the finished product. Bringing it back to life is a grand accomplishment, especially wood cabinet radios made with special veneer designs of birds-eye maple or walnut burl. Many were fine pieces of furniture, too.

SHANNON: Do you have any advice for audioXpress readers who are considering audio technology improvements to their own antique radios?

MIKE: Beware of trying to repair them yourself. Seek out a professional. Inside there are high voltages that can be dangerous to an untrained technician, some lethal. Plugging into 120 VAC (turning it on after being in storage for many years or decades) can damage some parts causing smoke, and broken power cords can be a shock!

SHANNON: There has been a long-running debate between tube and solid-state amplifier aficionados? Do you have a personal preference, and if so, why?

MIKE: Tubes rule! They're hot! Seriously though, I have performed many A/B listening tests between tube and solid-state components, and I always prefer the sound of tubes (see **Photo 4**).

Some solid-state components do come very, very, close but I always go back to vacuum tubes. They sound warm, and they look glowingly warm, too. Ask any guitar player—they want a tube amp. Also, look at the prices paid on eBay for vintage electronic tube gear. Beyond the technical debate, the best analogy I have



Photo 4: A vintage Hi-Fi unit on display at Urban Antique Radio and Vintage Hi-Fi

is the comparison between video and film, sharp and contrasty to warm and smooth.

SHANNON: What projects are you currently working on?

MIKE: I just finished a Sparton 15-tube radio console (see **Photo 5**). Just when you've think you've seen and heard it all. Another amazing radio from 1938. Borders on early Hi-Fi with its 15-tubes, separate power chassis (2A3 pair) and big 15" speaker. It has a beautiful mirrored dial and wood cabinet, too! I use to think Zenith was the best in this timeframe but there are others out there, Capehart, and E.H. Scott, to name a few. Even after 14 years (full time), I still haven't seen everything!



Photo 5: Sparton

SHANNON: A lot has changed in audio technology during the past few decades. What changes do you consider positive? Any negative?

MIKE: I used to think digital was the way to go. Then, I listened to more and more analog. It changed my perception. There is no comparison, especially compressed digital audio. It does have its place in portability (e.g., iPods, CDs, etc.) but when listening at home to a loud-speaker stereo system, you hear the dif-

ference. Vinyl has made a comeback for that very reason. I even get customers who swear by reel-to-reel tape. The ear is analog! And, I don't need surround sound with umpteen speakers (I will have many home theater buffs mad at me—I do like it for action movies, though) but for classic mono jazz recordings, early Beatles, or many stereo recordings (on vinyl)—two ears, two speakers, and two channels works for me. Another negative is ultra loud, subsonic distortion, thumping out of car audio systems. I feel like I am being tortured. Anybody else? Finally, what I especially dislike is the disposable nature of today's electronics in the mass market. Analog was replaced with digital, custom integrated circuits replaced component repair, and the build quality has all but disappeared.

Now on a positive note, vinyl pressings are coming out for new recordings and some of the old classics have been reissued, new production vacuum tubes for the audio market are being manufactured (Russia and China), and good old U.S.-made, quality vintage Hi-Fi, radios, HAM gear, and so forth can be found at very affordable prices. Add in restoration and you have something special that will last another lifetime, at a fraction it would cost to manufacture that same item today. Vintage lives on! **aX**

Scan-Speak's 15M / 4624G00

A 5.25" driver is put to the test

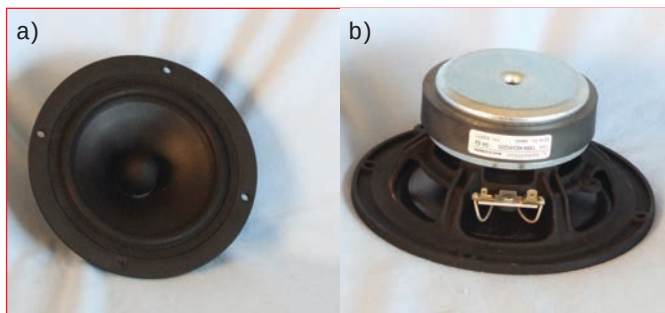
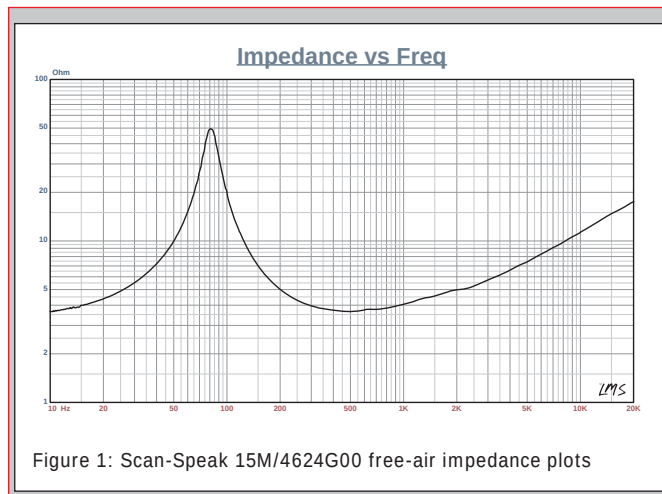


Photo 1: A top (a) and bottom (b) view of the 15M/4624G00 mid-range driver

The 5.25"-diameter 15M/4624G00 driver represents the mid-range addition to the relatively recent and cost-efficient Scan-Speak Discovery line (see **Photo 1**). Small-diameter mid-range drivers used in three-way designs have always been an important product for Scan-Speak. The 15M is built on a proprietary six-spoke cast aluminum frame that minimizes reflections off the frame back into the cone. Significant cooling is provided by six 23 × 6 mm "windows" below the spider mounting shelf that enable air to flow across the frontplate and around the voice coil on forward excursions. Powering this 5.25" device is a conventional 15-mm thick 85-mm diameter ferrite magnet, sandwiched between the polished frontplate and rear T-yoke. The T-yoke also has a 7-mm diameter pole vent. Other features include a NRSC fiber glass curvilinear cone with a lightdamping coating on the back, rubber coated 1"-diameter phase attached to the pole piece, a very low-damping black foam surround, 3"-diameter black flat cloth spider, 1" (25 mm) diameter voice coil (aluminum former wound with round copper wire), an aluminum shorting ring (Faraday Shield), and gold-plated terminals.

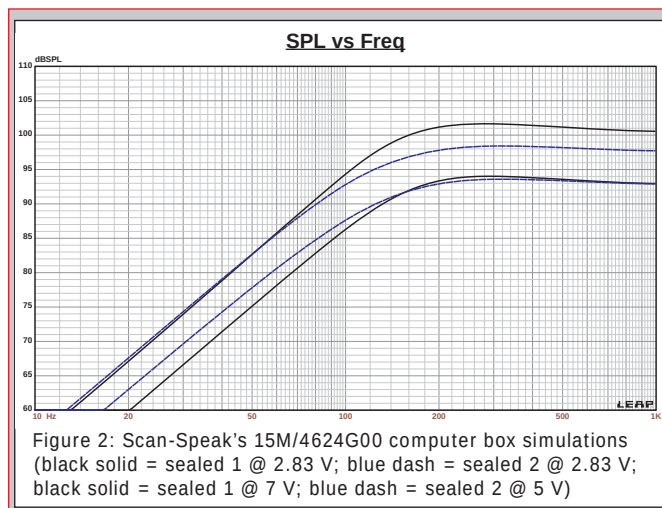
Testing commenced with the driver clamped to a rigid test fixture in free-air, voltage, and current sweeps taken at 0.3 V, 1 V, 3 V, and 6 V. Since this is a small diameter driver with the short Xmax of a mid-range (1.5 mm), I was surprised that the 6-V data was not too nonlinear for LEAP 5 to curve fit. The eight 550-point stepped sine wave sweeps for each 15M mid-range sample were post-processed and the voltage curves were divided by the current curves (admittance) to create impedance curves. The phase was added using the LMS hyperbolic phase calculation method and, along with the accompanying voltage curves, uploaded to the LEAP 5 Enclosure Shop software. In addition to the LEAP 5 LTD model results, I also created a LEAP 4 TSL model set of parameters

using just the 1-V free-air curves. The final data set, which includes the multiple voltage impedance curves for the LTD model (see **Figure 1** for the 1-V free-air



	TSL model		LTD model		Factory
	sample 1	sample 2	sample 1	sample 2	
Fs	80.5 Hz	75.3 Hz	77.2 Hz	72.4 Hz	100 Hz
Revc	3.14	3.14	3.14	3.14	3.2
Sd	0.0080	0.0080	0.0080	0.0080	0.0080
Qms	5.83	5.57	5.67	6.03	5.62
Qes	0.38	0.34	0.40	0.39	0.47
Qts	0.36	0.33	0.38	0.37	0.43
Vas	5.7 ltr	6.6 ltr	6.3 ltr	7.2 ltr	3.7 ltr
SPL 2.83 V	90.8 dB	91.0 dB	90.4 dB	90.3 dB	92.4 dB
Xmax	1.5 mm	1.5 mm	1.5 mm	1.5 mm	1.5 mm

Table 1: The Scan-Speak 15M/4624G00 driver



impedance curve) and the 1-V impedance curve for the TSL model, were selected and the parameters were created to perform the computer box simulations. **Table 1** compares the LEAP 5 LTD and TSL data and factory parameters for both of Scan-Speak 4" samples.

Unfortunately, the LEAP parameter calculation results for the Scan 15M mid-range didn't correlate well with the factory data. The factory data showed

a significantly higher F_s , a lower V_{as} , and higher sensitivity. I've found that published data isn't always updated throughout production, so sometimes this happens. Given the overall level of engineering integrity with Scan-Speak, I'm not too concerned about this. That said, I proceeded to set up computer enclosure simulations using the LEAP LTD parameters for Sample 1. I programmed in two sealed enclosures, one Butterworth and one Bessel response. For the first

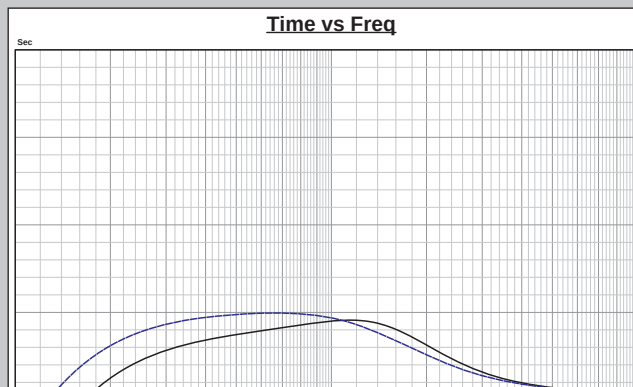


Figure 3: Group delay curves for the 2.83-V curves in Figure 2

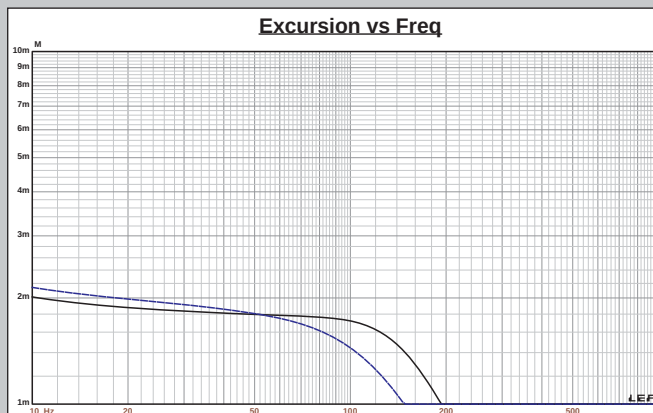


Figure 4: Cone excursion curves for the 7/5 curves in Figure 2

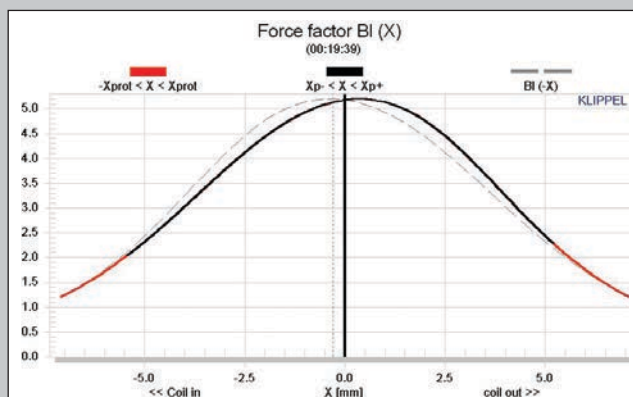


Figure 5: Klippel analyzer BI(X) curve for the 15M/4624G00

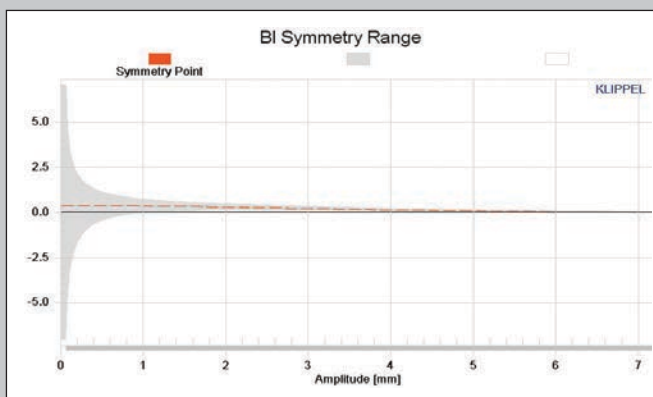


Figure 6: Klippel analyzer BI symmetry range curve for the 15M/4624G00

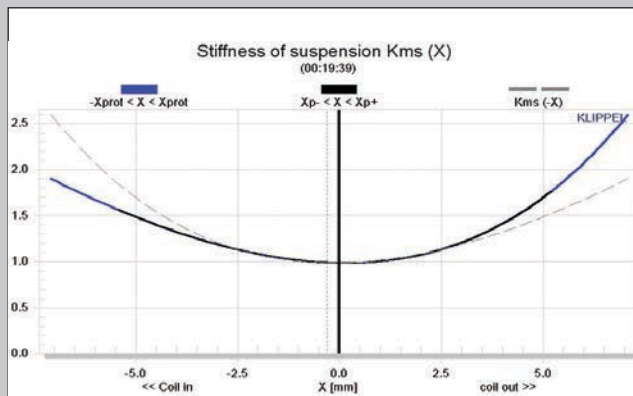


Figure 7: Klippel analyzer mechanical stiffness of suspension $K_{ms}(X)$ curve for the Scan-Speak 15M/4625G00

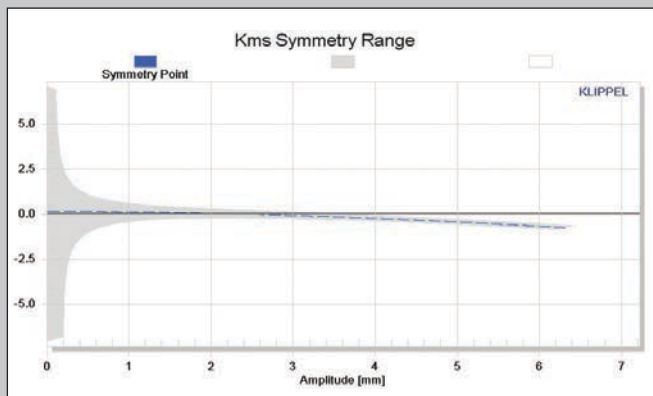


Figure 8: Klippel analyzer K_{ms} symmetry range curve for the Scan-Speak 15M/4625G00

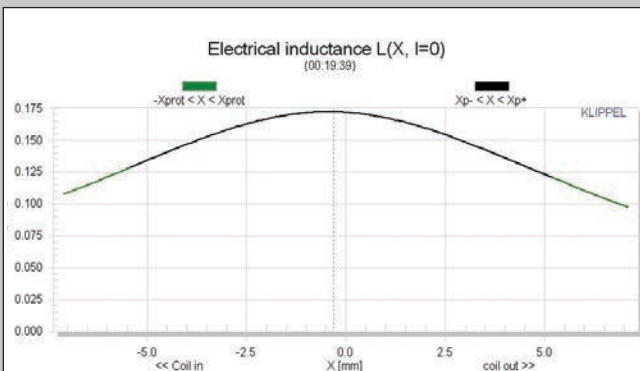


Figure 9: Klippel analyzer Le(X) curve for the Scan-Speak 15M/4625G00

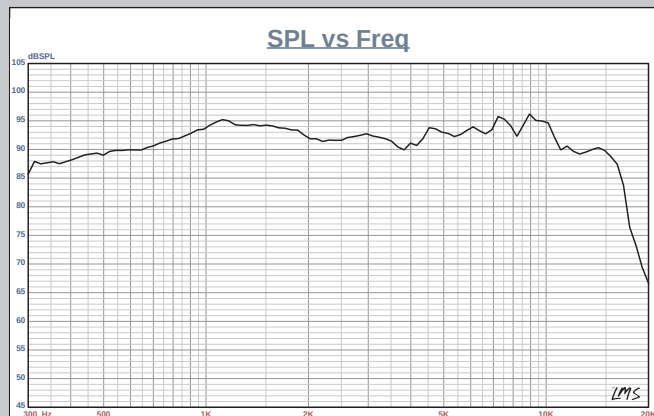


Figure 10: The Scan-Speak 15M/4625G00 on-axis frequency response

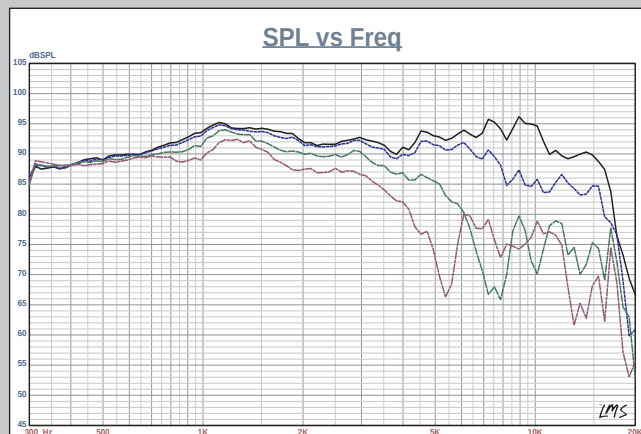


Figure 11: The Scan-Speak 15M/4625G00 on- and off-axis frequency response (0° = black, 15° = blue, 30° = green, 45° = purple)

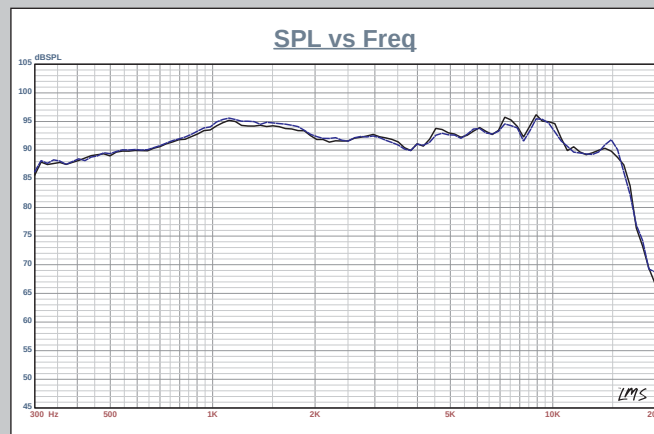


Figure 12: The Scan-Speak 15M/4625G00 two-sample SPL comparison

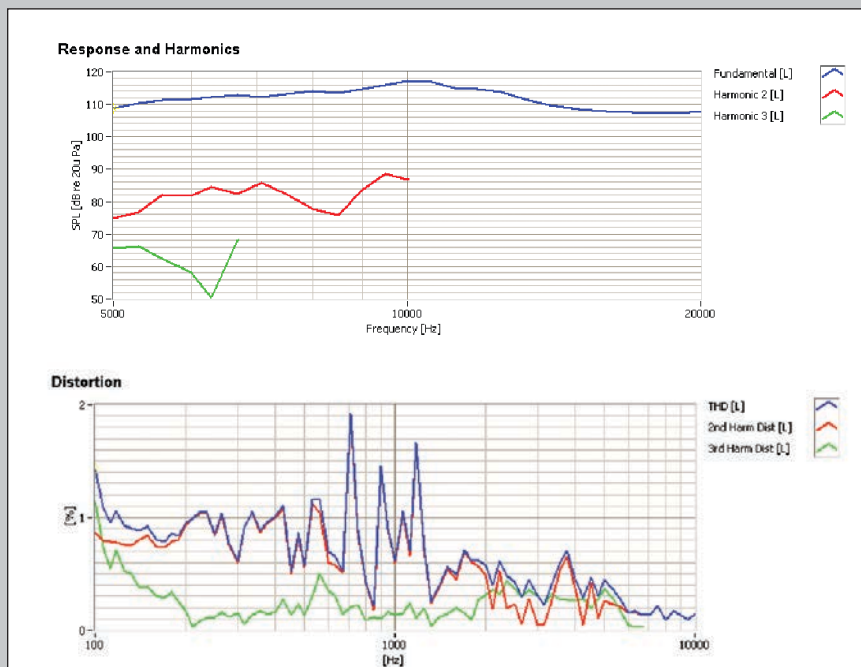


Figure 13: The Scan-Speak 15M/4625G00 SoundCheck distortion plot

closed-box Butterworth simulation, I used a 131-ci enclosure with 50% fiberglass fill material, and for the second sealed box, a larger volume of 238 ci with the same 50% fiberglass fill material.

Figure 2 displays the results for the Scan-Speak 15M/4624G00 in the two sealed boxes at 2.83 V and at a voltage level sufficiently high enough to increase cone excursion to $X_{max} + 15\%$ (1.7 mm for the 15M). This resulted in a $F_3 = 144$ Hz with a box/driver Q_{tc} of 0.65 for the 131-ci closed box design and a -3 dB = 117 Hz and a $Q_{tc} = 0.55$ for the 238-ci sealed box simulation. I conducted the two sealed simulations because the Butterworth enclosure was almost too small to be practical, plus making the enclosure larger lowers the resonance enough to make designing a high-pass network less of a problem.

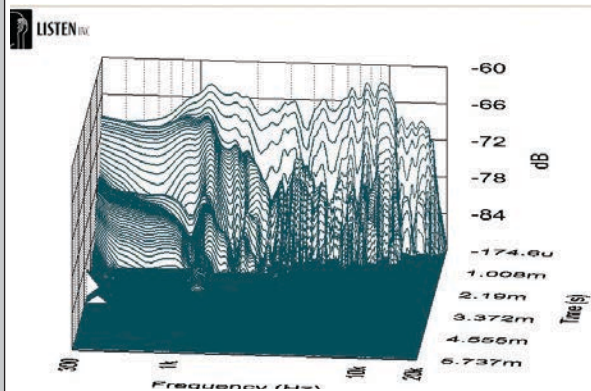


Figure 14: The Scan-Speak 15M/4625G00 SoundCheck CSD waterfall plot

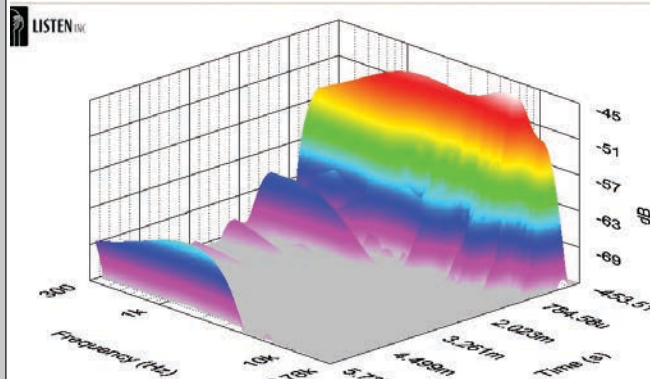


Figure 15: The Scan-Speak 15M/4625G00 SoundCheck Wigner-Ville plot

Increasing the voltage input to the simulations until the maximum linear cone excursion was reached generated a surprising (for a mid-range with no high-pass filter) 102 dB at 7 V for the smaller sealed enclosure simulation and 98.5 dB with a 5-V input level for the larger sealed enclosure. See **Figure 3** and **Figure 4** for the 2.83-V group delay curves and the 7/5-V excursion curves. Very reasonable performance for a 5.25" mid-range given this was "wide open" without the high-pass filter that will always be used with this product.

Klippel analysis for the Scan 15M mid-range driver produced the BI(X), Kms(X), BI, and Kms symmetry range plots given in **Figures 5–8**. The BI(X) curve for the 15M (see **Figure 5**) is somewhat narrow and symmetrical, typical of short Xmax drivers, with a trivial coil-out (outward) offset. Looking at the BI symmetry range curve in **Figure 6**, there is a 0.4-mm coil-out (forward) offset that goes to 0.3 mm at the physical Xmax position (1.5 mm for the 15M), so very nice. **Figure 7** and **Figure 8** give the Kms(X) and Kms symmetry range curves for the 5.25" mid-range. The Kms(X) curve is even more symmetrical. Figure 23 shows a 0.15-mm coil-out offset at the rest position that decreases to 0.09 mm at the physical Xmax of the driver. Displacement-limiting numbers calculated by the Klippel analyzer for the Scan-Speak mid-range were XBI @ 82% BI = 2.3 mm and for XC @ 75% Cms minimum was 3.6 mm, which means that for this 5.25" mid-range driver, the BI offset is the most limiting factor for a prescribed distortion level of 10%, but both numbers are significantly beyond Xmax.

Figure 9 gives the inductance curves L(X) for the 15M/4624G00, which shows the inductance decreases as voice coil travels inward covering more of the pole piece, typical of a driver with a shorting ring installed—it is aluminum in the case of the 15M mid-range. The inductance swing from Xmax forward to

Xmax rearward is about 0.16-to-0.17-mH inductance, which is excellent.

With the Klippel testing complete, I mounted the 15M mid-range in an enclosure that had a 15" × 5" baffle filled with foam damping material and proceeded to measure the driver frequency response both on- and off-axis from 300 Hz to 40 kHz at 2.83 V/1 m using a 100-point gated sine wave sweep. **Figure 10** shows the 15M's on-axis response, resulting in a very flat rising response that is ± 1.68 dB from 300 Hz to 4.5 kHz with a small peak just before the low-pass roll off. **Figure 11** has the on- and off-axis frequency response at 0°, 15°, 30° and 45°. A 3-to-4.5-kHz crossover frequency would be appropriate for this Scan-Speak small woofer because -3 dB at 30° with respect to the on-axis curve occurs at 4.5 kHz. **Figure 12** gives the two-sample SPL comparisons for the 5.25" 15M driver, showing a good match within the 15M's operating range.

For the last group of testing, I again fired up the SoundCheck analyzer, the SCM microphone, and power supply to measure distortion and generate time frequency plots. Setting up for the distortion measurement consisted of mounting the woofer rigidly in free air, and setting the SPL to 94 dB at 1 m (1.16 V) using a pink noise stimulus. (Two of SoundCheck's utilities are a software generator and SPL meter.) The distortion was measured with the SCM microphone placed 10 cm from the dust cap. This produced the distortion curves shown in **Figure 13**.

For the last test on the Scan 15M, I used the SoundCheck analyzer to get a 2.83-V/1-m impulse response for this driver. I then imported the data into Listen, Inc.'s SoundMap Time/Frequency software. The resulting CSD waterfall plot is shown in **Figure 14** and the Wigner-Ville (used for its better low-frequency performance) plot is shown in **Figure 15**. Visit the Scan-Speak website at www.scan-speak.dk for more information. *aX*