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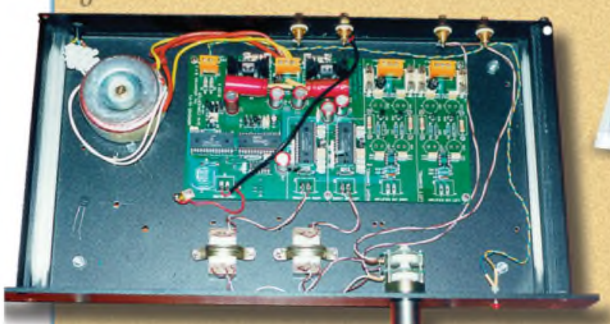

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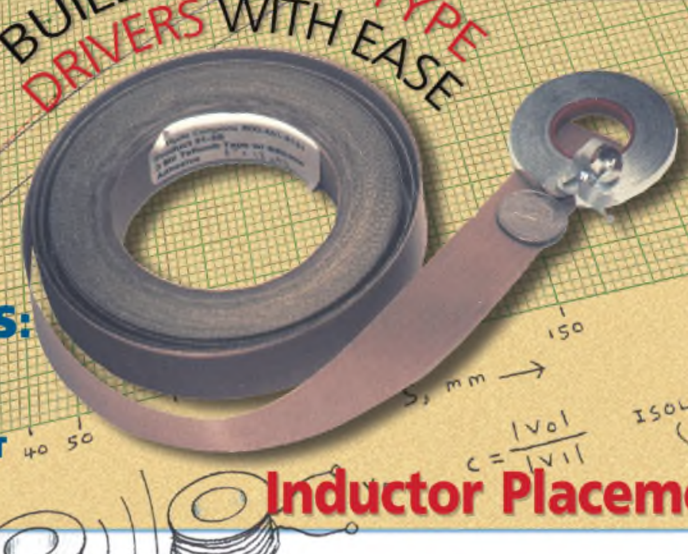

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Editorial

Is Music Property?

By Edward T. Dell, Jr.

Years ago, wandering in the street shops in Taipei, Taiwan, I was astonished to find sidewalk bins full of hundreds of classical LP recordings with EMI, London, RCA, Columbia, and Deutsche Grammophon labels, in packaging mimicking those in the bins of Sam Goody in New York City. On closer examination, however, you could see that the printing wasn't really crisp, and the insides of the records were a little strange. At that time Taiwan still did not recognize the Berne Copyright Convention and did little to control piracy of this type.

The news channels recently have been full of comment on the court case under consideration brought by A&M Records against Napster, Inc. To get a little historical background, we need to go back to Gutenberg, who put together the first technology in the West that allowed technological reproduction of copies of written documents: printing. Before that time, when books were handmade copies, no one seems to have had an idea that intellectual products had value.

Making copies of items has become a very large business, and the means of copying seem to get more sophisticated by the week. The copyright laws of the world have a fairly long history. They have been enacted and are enforced in most of the countries of the world. There's even an international agency in Switzerland for registration of world copyrights, usually noted with the © symbol. Copyright laws protect the rights of those who compose or perform music—as well as other original work which humans produce.

A couple of years ago a teenager decided that the Internet was a great way to distribute music to enthusiasts by setting up a company which essentially allows individuals to post performances of music in a central place which is available to others for download. I think the 19-year-old who put this idea together named it appropriately. A service for kids = kidnapster.

The Consumer Electronics Association, representing companies who annually sell $70 billion worth of consumer electronics, believes that the Ninth Circuit Court of Appeals' decision the other day "ignored basic principles of copyright infringement and fair use established in the U.S. Supreme Court's Betamax decision."

The CEA believes that this will stifle technological innovation. It seems to me this is talk straight out of *Animal Farm*. There are certain other bright kids who have developed new technologies for breaking into central bank computers and appropriating credit cards. I am sure the police are not considering the question of stifling new technology in such cases.

In the Betamax decision the court allowed individuals to make copies for their own use. In such a case someone had bought a copy of the document, and therefore had paid for its use, at least one use. But VCRs don't work well if copies of copies of copies are made. In the Napster case, no one apparently needs to pay any owner for more than one usage—with 50 million users welcome to make perfect copies.

The CEA further complains that the "play" buttons on our machines will become "pay" buttons by this new ruling. Well, seems to me that they are already "pay" buttons. And they should be—they always have been. This is more "newspeak" from special interests.

This is a much larger battle than this controversy. Technology for making copies is raising the issue in dozens of situations. Those who produce movies and music are attempting to watermark recordings in order to detect piracy. Purists are complaining that this degrades the content. That battle is ongoing with no real resolution in sight.

Classrooms have been appropriating copyrighted information in journals, books, and other data forms as part of teaching. An institution exists to help manage that usage fairly, and the academic institution pays a small, pro-rated fee for such use. In the past, organizations such as ASCAP were founded to collect royalty fees on broadcast and other uses of music compositions and performances.

There are, of course, systems of philosophical belief which consider property ownership as theft. Many societies have existed—and a few primitive ones still exist—which subscribe to that belief. The Manhattan tribes who "sold" the island for beads must have had a different view of the nature of ownership than the Dutch who "paid" for it. We are, however, in a capitalist society, whatever we may think about it. The battle is between people who consider "hardware" real and "software" or "thinkware" unreal, on the one hand, and those who think otherwise.

Even the people at Napster now seem to believe (with some legal nudging) that somehow a way must be found to pay the piper. They have been offered help by the giant Bertelsman, who owns a lot of publishing companies as well as BMG Music, EMI, RCA, and will probably soon acquire others. Those who like Napster's ability to publicize new music should give the company permission to showcase their work. No one will have any problem with that use of the technology.

Whatever happens, those of us who love music will do well to find some kind of value which we believe should be placed on the work of artists who compose and perform it. CEA wants us to write our congressional representatives to support Napster. I think any of us who may think otherwise had better make our views known to Washington's legislators as well.—**E.T.D.** ❖

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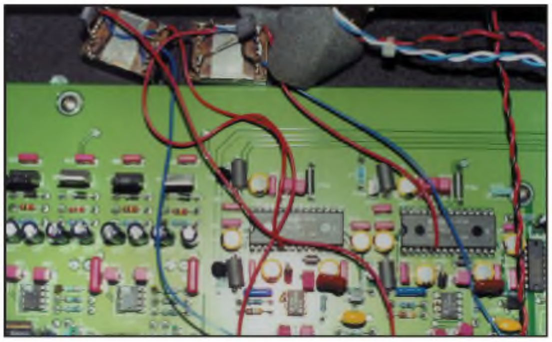

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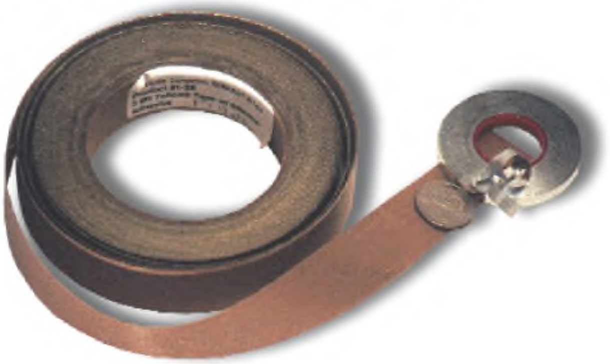

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The peculiar evil of silencing the expression of an opinion is, that it is robbing the human race; posterity as well as the existing generation; those who dissent from the opinion, still more than those who hold it.

JOHN STUART MILL

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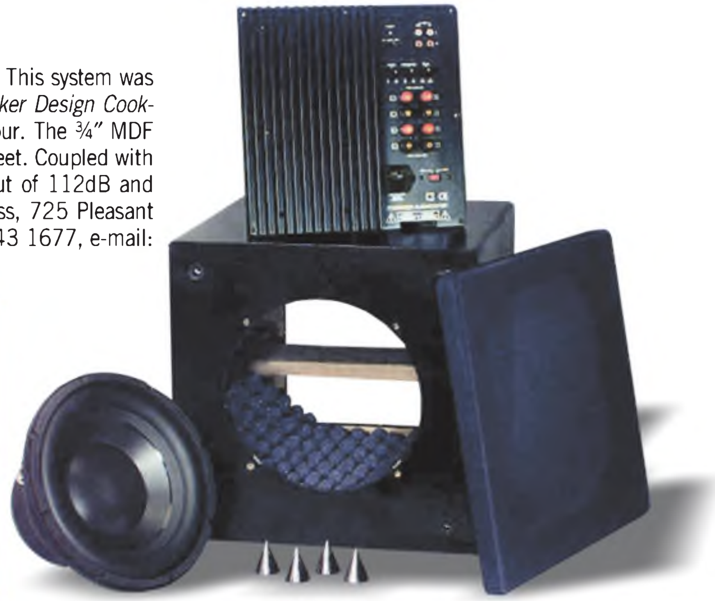
Audio News

TITANIC SERIES SUBWOOFER SYSTEM

Parts Express introduces the Titanic Series 10″ Powered Subwoofer System. This system was designed by Vance Dickason and is featured in the Sixth Edition *Loudspeaker Design Cookbook*. It comes in kit form and can easily be assembled in less than one hour. The ¾″ MDF cabinet is finished in textured black enamel and has "black chrome" spike feet. Coupled with the Parts Express 250W amplifier, this system produces a maximum output of 112dB and yields an f_3 of around 24Hz in small- to medium-sized rooms. Parts Express, 725 Pleasant Valley Drive, Springboro, OH 45066-1158, (513) 743 3000, FAX (513) 743 1677, e-mail: sales@partsexpress.com.



NEW SONIFLEX REDBOXES

Throughout this year, Soniflex is launching several new additions to their Redbox range of analog and digital audio interface boxes. New Redboxes include three stereo matching amplifiers for interfacing domestic or semi-professional unbalanced equipment, such as a CD player, to professional balanced line levels: the RB-UL1 single, and the RB-UL2 dual and RB-UL4 quad unbalanced to balanced converters. Soniflex Ltd., 61 Station Road, Irthlingborough, Northants, NN9 5QE, United Kingdom, +44 (0)1933 650 700, FAX +44 (0)1933 650 726, e-mail: sales@soniflex.co.uk, internet: http://www.soniflex.co.uk.

B+K TEST BENCH DMMS

B+K Precision Corporation introduced a new family of multifunction DMMs, Models 388B, 389A, 390A, and 391A. These meters include component test capabilities, resistance diode test and capacitance and measure frequency, temperature, and logic indicator. The Models 389A and 390A feature increased capacitance capabilities and expanded frequency measurement up to 40MHz. The Model 390A also offers an IR-RS-232 interface and comes complete with interface cable and software, while the Model 391A features true RMS capability. These units measure 3.5″ wide by 1.57″ deep by 7.8″ high, and weigh 11.3 oz. They come complete with 9V battery, test leads, and instruction manual; the Model 390A also comes with a thermocouple probe, IR-RS-232 cable, and software. B+K Precision Corporation, 1031 Segovia Circle, Placentia, CA 92870-7137, (714) 237-9220, FAX (714) 237-9214, www.bkprecision.com.

CANTARES SURROUND-SOUND DECODER

Cantares introduces the SSP-1 Surround Sound Decoder, stereo enhancer, and subwoofer crossover. This all-analog unit is designed for both consumer and professional/institutional use. The SSP-1 provides home-theater functions, without the intrusion of logic circuitry, and can integrate with 5.1 components, if desired. It supports three to six speakers, as well as a subwoofer and an overhead. Soundfield microphone users can make use of the B-format output and can produce a 5.1 signal from it, monitor B-format in preparation for 5.1, UHJ, or other encoding, and make surround versions of stereo or quad material when a multi-track master is not available. Cantares, 74 George St., Waterloo, Ont., N2J 1K7, Canada, (519) 744-9395, FAX (519) 744-7129, mdunn@cantares.on.ca, http://www.cantares.on.ca/.

MCINTOSH LABORATORY VACUUM TUBE AMPLIFIER

McIntosh Laboratory introduces the MC2102 vacuum tube amplifier, the second entry in McIntosh's new 2000 series of tube amplifiers and preamplifiers. Components in the 2000 series will be limited editions and share key characteristics; all will be designed and engineered by McIntosh co-founder and former President, Sidney Corderman. The MC2102 produces 100W in stereo or 200W in mono, accepts loads of 2, 4, or 8Ω, and has common mode rejection greater than 60dB at mid frequencies. Each channel uses eight tubes, and the unit features the unity coupled output transformer, which uses two bifilar wound primaries. The MC2102 also features a remote power control, gold-plated input and output jacks, 100dB "A" weighted signal-to-noise ratio, .8V sensitivity, and a wideband damping factor greater than 18. J.B. Stanton Communications, (860) 542-1234, FAX (860) 542-0005.

The Virtual Crossover, Part 1

The first of this three-part series introduces you to the theory behind computer modeling of both active and passive crossovers.

By Richard Mains

The final stage in all the speaker building projects I have attempted is always tweaking the crossover. Measurements and computer analysis can get close to the optimum design, but I still end up trying many different component values and circuit configurations before I am satisfied with the results. This can be a very tedious and time-consuming process.

If you are working with passive crossovers, it may require winding yet another coil or finding the right capacitors to parallel in order to try out a crossover modification. If you use active crossovers, you may need to unsolder and resolder components, and if you have made a crossover board, it may be nearly impossible to try different circuit topologies. As a result, it is often necessary to settle for less than optimum results, simply because trying everything would take a prohibitive amount of time and effort.

COMPUTER MODELING

Somewhere in the midst of one such process, it occurred to me how much easier it would be if you could simply model the crossover circuit on the computer and somehow apply the results directly to the speakers, without needing to construct a real circuit. The more I thought about this idea, the more I was convinced it could be done, particularly with the availability of powerful floating-point DSP (digital-signal processing) chips.

In my electrical engineering work, I have had considerable experience in computer modeling of circuits, so the idea of simulating the crossover circuit was not foreign to me; however, the much more difficult tasks would be to carry out the simulation in real time, and to somehow apply the results to the speakers with acceptably low distortion.

In addition to saving a lot of time that would otherwise be spent modifying real circuit components, there are perhaps even more compelling reasons to use DSP techniques to implement the crossover. You can accomplish things in the digital domain that are not possible using real, analog circuits. If you have worked with analog filters higher than first-order, you know that they invariably introduce phase shifts; if a crossover uses high-order analog filters, it is nearly impossible to obtain accurate transient response from the resulting speaker system.

For example, a square pulse supplied to the speakers will appear as quite a different waveform at the output. However, by using digital techniques it is straightforward to implement filters with the same high-order amplitude characteristics, but with zero net phase shift.

VIRTUAL CROSSOVER

Another benefit of simulating the crossover circuit is that you can use component values that are otherwise not easily realizable. If your design calls for a 4.0H inductor, for example, it's no problem; just type the value into the simulation—no need to actually wind such a coil. Also, the digital domain allows you to introduce arbitrary delays between drivers, eliminating problems of driver offsets. As I thought about these advantages, it became apparent that while you could use such a device as a temporary stage to optimize the real crossover, the most desirable approach might be to replace the real crossover entirely.

In Part 1 of this article, I will explain the basic theory that I used to implement a device for modeling both active and passive crossovers. In addition to playing music, the device I built can also carry out acoustic measurements, so that you can observe the change in speaker response as you vary parameters in the crossover circuit simulation.

In Parts 2 and 3, I will present circuit diagrams and construction details, and will show an application of this device in designing a crossover for a Roger Sanders-type hybrid electrostatic/transmission-line speaker system[1–3] that yields very good transient response, which I was never able to obtain with active crossovers I constructed previously for this system using op amps.

When I described my proposed device to a good friend of mine, Ed Braytenbah, he came up with a name for it that I found to be quite appropriate: the "Virtual Crossover."

SHANNON'S SAMPLING THEOREM[4–6]

One of the most basic and important results from digital signal processing theory is Shannon's Sampling Theorem. I

ABOUT THE AUTHOR

Richard Mains obtained a Bachelor of Science degree in electrical engineering and a Master of Science degree from the Ohio State University in 1972 and 1974, and a Ph.D. in electrical engineering from the University of Michigan in 1979. He worked for several years in the area of high-frequency semiconductor device modeling at the University of Michigan. Since the early 1990s, he has worked in the area of power electronics circuit design and microprocessor programming at McCleer Power, Inc., located in Jackson, Mich. His interests have included working on several speaker systems, developing measurement systems for loudspeakers, and designing power amplifiers and other electronics for music reproduction.

find it to be a remarkable result, truly a case where things work out better than expected.

The basic problem is that to do what I propose, I must first discretize the analog audio signal before it can be processed using digital techniques. The concern then is, if I throw away all of the analog signal except the values at a set of evenly spaced discrete samples, how much information is lost? Even without processing the signal at all, but just converting it right back to analog form, how much would it be degraded just due to the sampling process?

The theorem states that if the analog signal you are sampling is band limited, i.e., if it has zero frequency content for $f > f_c$, then the sampling process itself will lose absolutely no information, provided the sampling rate is greater than $2f_c$. Consider the case where the analog signal is a single-frequency sine wave; then the theorem states that if there are at least three samples within each period of the sine wave, you lose no information about the original analog signal.

I find this remarkable, because three samples in a sine wave period yield a very coarse representation of the signal. Nevertheless, it is possible to recover the original sine wave exactly, just from these rather coarse samples.

To recover the analog waveform from its samples, you use an interpolating function to "fill in" the gaps between them. Let x(t) be the original analog signal that was sampled, and let x_n be the discrete samples $x(n\Delta t)$. Then, provided that x(t) has zero frequency content for $f \geq 1/(2\Delta t)$, the continuous function x(t) may be recovered exactly at any time t from the set of samples x_n as follows:

$$x(t) = \sum_{n=-\infty}^{+\infty} x_n \left\{ \frac{\sin(\pi(t - n\Delta t)/\Delta t)}{\pi(t - n\Delta t)/\Delta t} \right\} \qquad (1)$$

Equation 1 essentially weighs each sample with a sin(x)/x function, which is the impulse response of an ideal low-pass filter.

Although the theorem says that you can recover an analog signal completely from its samples, in practice you can only approach this ideal. That is because, first of all, the sampled signal may not in general be completely band limited, although it may have very little frequency content above some cutoff frequency f_c; and secondly, because the analog-to-digital converter used to obtain the set of samples x_n will have only finite precision. However, you can come as close as you wish to the complete reconstruction of an analog signal, depending upon how carefully you address these issues.

CIRCUIT SIMULATION AND THE IMPULSE RESPONSE

Once the Sampling Theorem reassured me that I could switch back and forth between the digital and analog domains with little signal degradation, the next step was to decide upon a method to simulate the crossover circuits. I have had much experience solving the differential equations for circuits using finite difference and finite element techniques, but these methods require a time step Δt that is much smaller than the time constants inherent in the circuit response, and they are computationally intensive.

To model a crossover circuit, you must carry out the simulation in real time, because you need a constant supply of updated results to drive the speakers. From the previous discussion of the Sampling Theorem, if the input signal you are sampling is band-limited to 20kHz, then you must sample it at a frequency greater than 40kHz, and you must also provide simulation results at the same rate, i.e., at least every 25µs.

Probably the time step in the simulation would need to be much smaller than 25µs, so it would be necessary to solve the equations several times for each result fed to the speakers. Even using the fastest computers available today, this would be difficult to accomplish.

An alternative method that is much better suited to this application is to use the impulse response of the crossover circuit. According to linear circuit theory, you can completely determine the output of a circuit if you know its impulse response and the input to the circuit.[7] Therefore if you can determine the impulse response before carrying out the simulation, all you need do is store it in the computer and process it together with the input to the circuit.

The next question is, how do you obtain the crossover-circuit impulse response? An ideal impulse is defined as an infinitely high and infinitely narrow pulse, such that the area under the pulse is 1. Obviously, it is not possible to use such a pulse waveform in an actual calculation. Also, you need to be concerned about violating the sampling theorem—an ideal impulse function is not band-limited; in fact, it contains a uniform distribution of all frequencies. If you calculate the impulse response using such a function, you will generally not be able to sample the

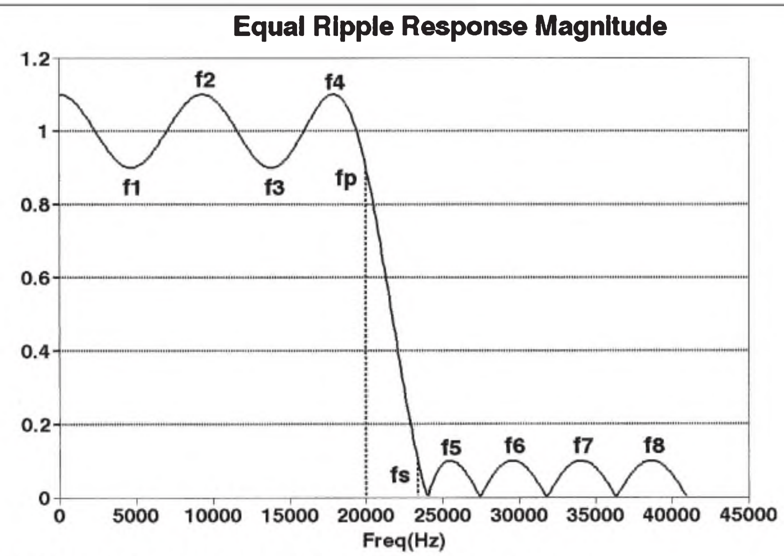


FIGURE 1: Illustration of equal-ripple low-pass filter response.

result at a sufficiently high rate to avoid loss of information.

BAND-LIMITED IMPULSE

My solution to this problem was to use a band-limited impulse to calculate the circuit-impulse response. What is required is an impulse that has a nearly uniform frequency distribution up to some cutoff frequency, but that then rapidly cuts off and has high attenuation for frequencies at and above one-half the sampling frequency. This type of frequency characteristic is most easily obtained using equal-ripple type filters.

Figure 1 shows the general shape of the frequency-response magnitude for an equal-ripple, low-pass filter. The filter in *Fig. 1* was designed to have passband and stopband ripple of 0.1, which of course is much larger than you would use in practice; I made this choice only as an illustration, so that the ripple would be clearly visible.

The filter in *Fig. 1* passes all frequencies up to the passband edge f_p nearly without attenuation, but above f_p (20kHz in this case), the filter cuts off rapidly. Beginning at the stopband edge f_s, the filter attenuates all higher frequencies. The frequencies labeled f1 through f8 in *Fig. 1* are the response extrema frequencies, which are important in the design method for this type of filter.

To design an equal-ripple, low-pass filter, iterative computer techniques must be used. I wrote a program, called ERLPF, to design this type of filter, and would be glad to share it with anyone interested. I will post this program, along with others discussed in this article, on my website.

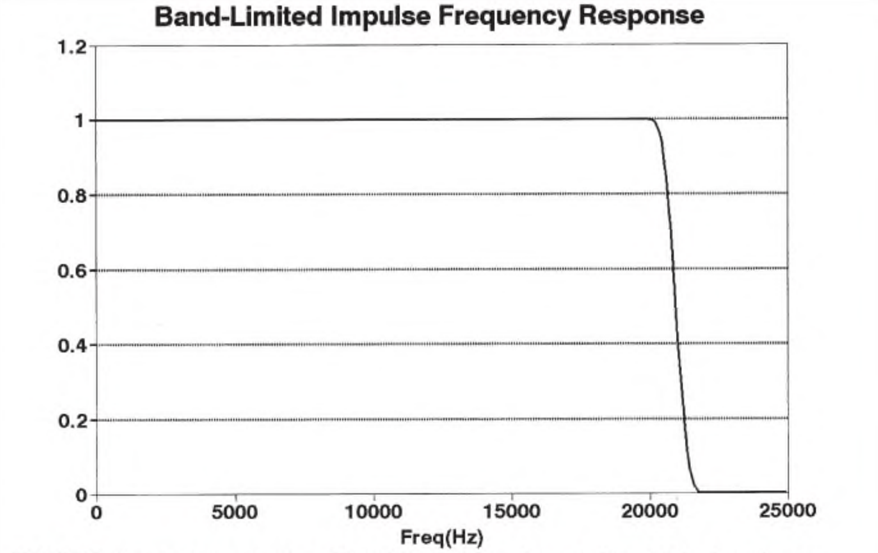


FIGURE 2: Frequency response of band-limited impulse used to obtain impulse response of crossover circuits ($f_p = 20$kHz, $f_s = 21.9$kHz). B-1773-2

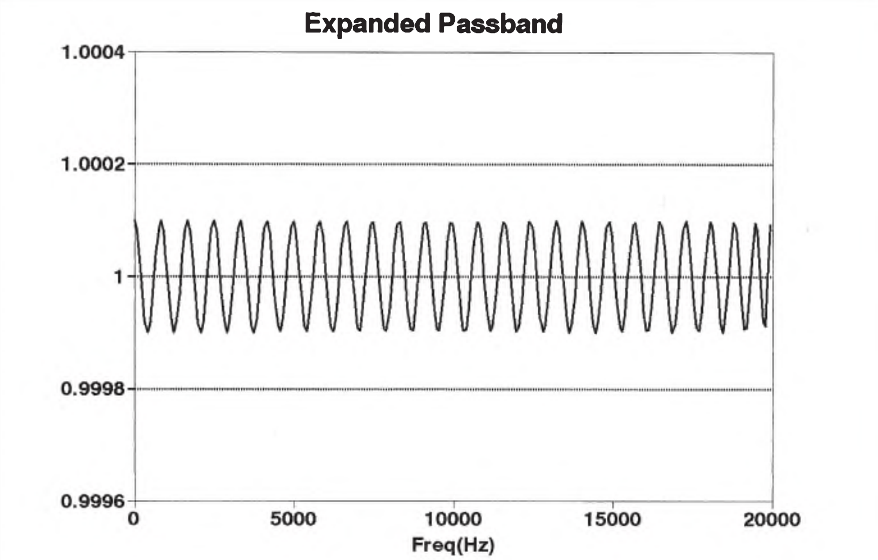


FIGURE 3: Expanded passband of *Fig. 2*, showing the 10^{-4} passband ripple. B-1773-3

FILTER DESIGN

The basic design procedure for the filter type in *Fig. 1* is described in Oppenheim and Schaffer[4], pp. 259–260. I will provide only an outline of the method here. The first step is to recognize that the response of the type in *Fig. 1* can be expressed as follows:

$$X(\omega) = x_0 + 2\sum_{i=1}^{M-1} x_i \cos(i\omega)\,, \qquad (2)$$

where $X(\omega)$ is the frequency response in terms of $\omega = 2\pi f$, x_i are samples of the time-domain waveform at times given by $i\Delta t$, and where it is assumed that the time-domain waveform is symmetric about zero (i.e., $x_{-i} = x_i$), so that the frequency response $X(\omega)$ is a real function. In equation 2, M is the total number of extremum points in the normalized frequency range $0 \le \omega \le \pi$, including the endpoints.

Further note that equation 2 can be rewritten as follows:

$$X(\omega) = \sum_{i=0}^{M-1} a_i (\cos(\omega))^i\,, \qquad (3)$$

so that you can express the frequency response as a polynomial in cos(ω). The reason for an iterative procedure is that the values of the extremum frequencies (f1 through f8 in *Fig. 1*) are not known beforehand.

I start out with an initial guess, assuming that the extremum frequencies are uniformly distributed throughout the frequency range. Using this initial guess, I construct a polynomial as given in equation 3, and then I find the actual extremum frequencies of this polynomial, which differ from the initial guess. These updated values are used as a new initial guess for the extremum frequencies, and the process is repeated until convergence is obtained.

I used this method to generate a

band-limited impulse to use in calculating the impulse response of crossover circuits. Rather than the coarse ripple value of 10^{-1} shown in *Fig. 1*, I specified a ripple value of 10^{-4} and a passband edge f_p of 20kHz for the impulse. *Figure 2* shows the result obtained. The stopband edge f_s, which is a result of the program, turned out to be 21.9kHz for this case. The 10^{-4} ripple is too small to be seen on the plot in *Fig. 2*; *Fig. 3* shows an expanded view of the passband, where the ripple is clearly visible.

I chose the passband edge of 20kHz specifically to pass music signals up to this frequency without attenuation. It turns out that my system uses a sampling frequency of 48kHz, so according to the sampling theorem, I must band-limit the signals to at least 24kHz; the stopband frequency of 21.9kHz is therefore adequate for a 48kHz sampling frequency.

SYMMETRIC WAVEFORM

Figure 4 shows the time waveform that corresponds to the frequency response given in *Fig. 2*. Note that the impulse waveform is symmetric about its peak value. Time domain responses of digital filters are often designed to be symmetric about t = 0, because the frequency response of such a symmetric waveform has zero phase shift.

To make the response causal (i.e., so that the waveform does not begin for t < 0), it is then shifted up in time until the response near t = 0 is negligibly small. The shifting introduces a linear phase term in the frequency domain, but by suitable time delaying of all the waveforms from the different crossovers, it is still possible to end up with zero net phase shift in the resulting output.

You may object to the ringing that is evident in the impulse waveform of *Fig. 4*; it is characteristic of the time responses of sharp cutoff filters. The ringing is within the transition and stopbands of the frequency response in *Fig. 2*; i.e., it is above 20kHz. As long as the input audio signal is truly band-limited to 20kHz, there will be no ringing added at the output of the circuit. However, if the input signal is not band-limited to 20kHz, the impulse function may need to be redesigned. There is nothing sacred about my choice of band-limited impulse function; you could implement another design using the ERLPF program I provide on my website.

CALCULATING THE IMPULSE RESPONSE

The next step is to apply the waveform given in *Fig. 4* at the input of the desired crossover circuit(s), and to calculate the output; this result will be the band-limited impulse response that will characterize the crossover and allow calculation of the crossover output in real time.

To calculate the band-limited impulse response of the crossover, I use the state-variable technique.[8] In this method, the variables of the simulation are chosen to be the voltages across capacitors and the currents through inductors. One advantage of this choice is that it leads to a coupled set of first-order differential equations to solve.

It is also fairly straightforward to program this method to handle general circuit configurations. I will not go into the details of programming this method here, but will limit the discussion to a simple example of how you can calculate the band-limited impulse response of the low-pass RC circuit shown in *Fig. 5*. In this figure, V is a voltage generator that applies the impulse, and you measure the output of the circuit across capacitor C.

Forgetting about band-limiting for the moment, and assuming that V applies an infinite impulse of area equal to 1, it is simple to determine what the ideal impulse response of this circuit will be. When the impulse is applied, the voltage across the series R-C combination is very large, and the voltage across the capacitor will be negligible, since the capacitor will take some time to charge up. During this interval, therefore, a nearly constant current flows through the circuit, given simply by V/R.

Since the area of the impulse is 1, we know that $V \times \Delta t_i = 1$, where Δt_i is the time duration of the impulse. Therefore, at the end of the impulse, the charge built up on the capacitor will be 1/R coulombs. Since $C = Q_c/V_c$, where Q_c is the charge on the capacitor and V_c is the voltage across the capacitor, at the end of the impulse the voltage across the capacitor is given by 1/RC.

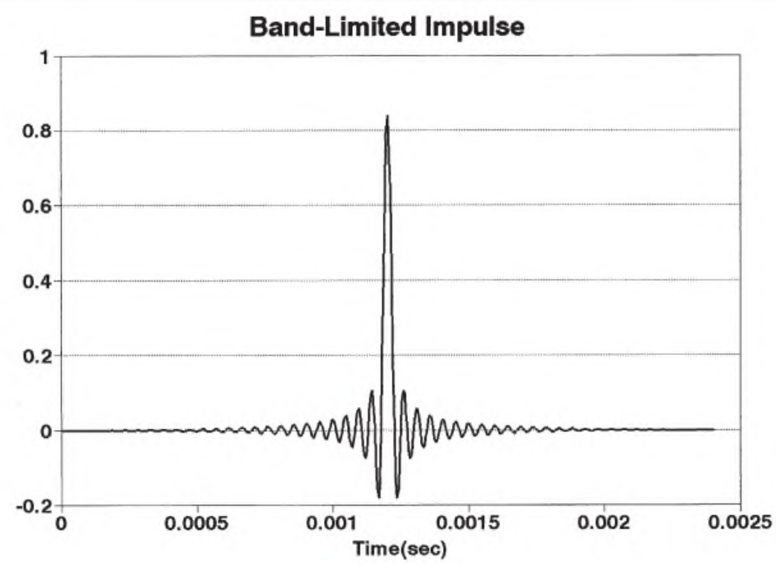


FIGURE 4: Band-limited impulse used to obtain impulse response of crossover circuits (shown at a finer time-sampling rate than is actually used in the simulations). B-1773-4

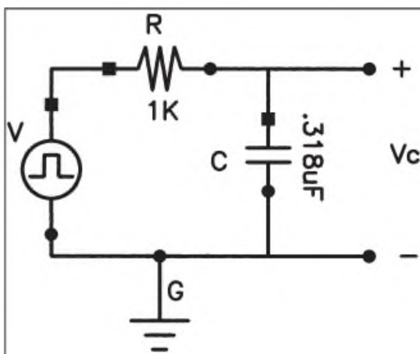


FIGURE 5: First-order low-pass RC circuit used to illustrate impulse response calculations. B-1773-5

When the impulse generator V switches off, this voltage starts to decay through resistor R, with a time constant given by RC. The top trace in *Fig. 6* shows this ideal impulse response for the RC circuit of *Fig. 5*. The time constant of the decay after the impulse is applied is .318ms, and the corner frequency for this low-pass filter is 500Hz.

CALCULATION METHOD

How do you calculate the band-limited impulse response of this RC circuit? As indicated previously, the simulation variable for the circuit is the voltage across the single capacitor, V_c. In this case, it is easy to write the first-order differential equation for V_c. First note the following expression for the current in this circuit:

$$i = C\frac{dV_c(t)}{dt} = \frac{V(t) - V_c(t)}{R} \quad , \qquad (4)$$

where V(t) is the impulse function. You can rearrange equation 4 to give the time variation of $V_c(t)$:

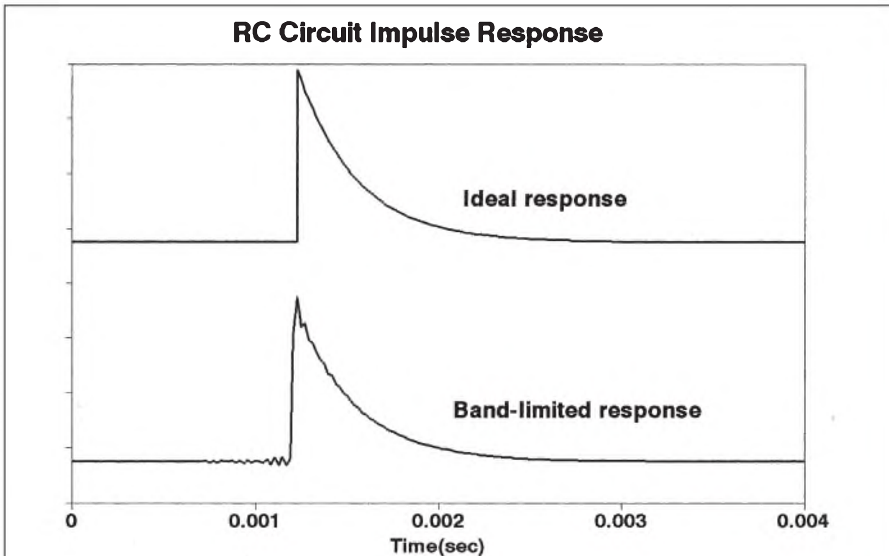


FIGURE 6: Ideal impulse response (top trace) and calculated band-limited impulse response (bottom trace) for the circuit of *Fig. 5*.

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$$\frac{dV_c(t)}{dt} = \frac{V(t) - V_c(t)}{RC} \quad . \qquad (5)$$

To calculate $V_c(t)$, you need to discretize equation 5. I use a finite difference discretization called the implicit method, which means that the terms on the right-hand side of equation 5 are taken to be at the future time; this leads to a very stable formulation. Using this method, if you know the value for the state variable $V_c(t)$ at time t, you can obtain the value at t + Δt by

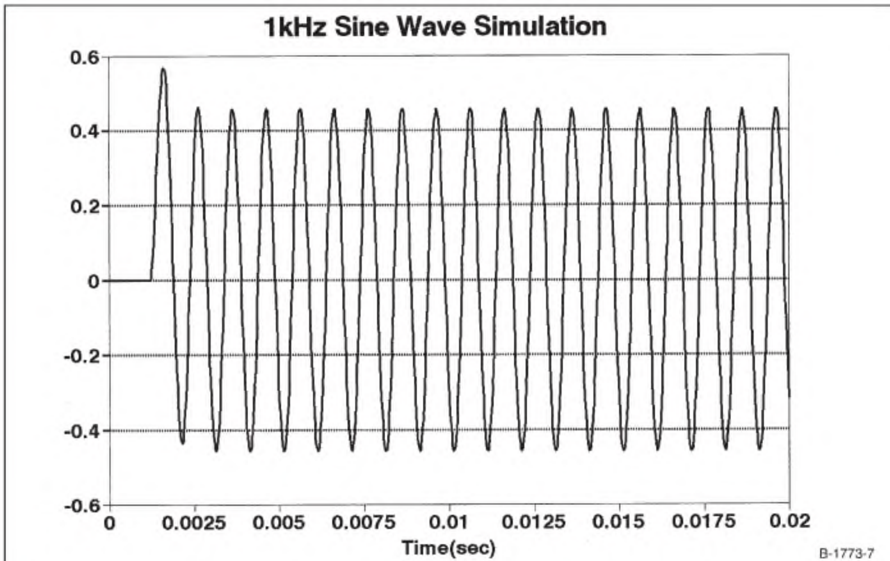

FIGURE 7: Simulation resulting from passing a unit amplitude, 1kHz sine wave through the RC circuit of *Fig. 5*, using the band-limited impulse response (lower trace in *Fig. 6*).

solving the following equation:

$$\frac{V_C(t+\Delta t)-V_C(t)}{\Delta t}=\frac{V(t+\Delta t)-V_C(t+\Delta t)}{RC}\;; \quad (6)$$

or, solving for $V_c(t+\Delta t)$:

$$V_C(t+\Delta t)=\frac{\Delta t V(t+\Delta t)+RCV_C(t)}{\Delta t+RC}\,. \quad (7)$$

I have a program called IMPRSP (impulse response) that calculates the equations of this form for crossover circuits. I will post it on my website, and you are welcome to use it. For a general circuit with more than one capacitor or inductor, rather than a single equation, you need to solve a set of coupled differential equations, which requires a matrix solution.

I have carried out this calculation using the waveform in *Fig. 4* as V(t); the lower trace in *Fig. 6* shows the result of this calculation. You can see that the form is similar to the ideal impulse response, but the band-limited impulse response smears out the abrupt transition present in the ideal response.

You need do this calculation only once before modeling the crossover for either playing music or carrying out acoustic measurements. I use a very small time step for the calculation, much smaller than that given by the 48kHz sampling rate, in order to obtain an accurate impulse response. I save the results only at the sampling rate, however.

THE FINAL STEP—CALCULATING CIRCUIT OUTPUT

The lower waveform in *Fig. 6* contains the information needed to model the crossover circuit in real time; you can think of it as the "signature" of the crossover. Say you wish to use this impulse response to play music through the crossover. Let $V_a(t)$ be the analog input signal from the audio source, which might be the output of a CD player, for example. According to linear network theory, the output from the crossover resulting from $V_a(t)$ may be obtained exactly from the following equation, which is called the convolution integral:

$$V_{out}(t)=\int_{-\infty}^{+\infty} V_a(\tau)h_{imp}(t-\tau)d\tau, \quad (8)$$

where I have used $h_{imp}(t)$ to indicate the circuit impulse response at time t, i.e., the waveform such as the lower response in *Fig. 6*.

Of course, you don't calculate this integral to determine the output voltage $V_{out}(t)$, but rather the following discrete version of it:

$$V_{out}(n)=\sum_{i=0}^{N} h_{imp}(i)V_a(n-i), \quad (9)$$

RESOURCES
The author has posted the software that was used in the development of the Virtual Crossover on his website, which can be found at www.usol.com/~rkm/audio. You may direct questions regarding the software or these articles to his e-mail address at rkm@usol.com.

where the integer arguments in parentheses denote the sample of the function at a particular time; for example, $V_{out}(n)$ is the output voltage at time $n\Delta t$.

Those of you familiar with digital filter theory will recognize equation 9 as the expression for the output of an FIR (finite impulse response) filter. You assume the impulse response h_{imp} to have only N + 1 nonzero values, from $h_{imp}(0)$ to $h_{imp}(N)$; in practice, this is done by truncating the waveform in *Fig. 6* at points where the response has decayed to a small value.

This truncation is necessary, because the number of computations you must perform within each sampling interval is proportional to N + 1, and since computing resources are finite, you must make a judicious choice as to the length of the discrete impulse response used to characterize the crossover circuit. (I should also mention that the discrete impulse response values $h_{imp}(i)$ in equation 9 must be weighted by the sampling time step in order to obtain correct results.)

Figure 7 shows the calculated output of the RC circuit in *Fig. 5* when a unit-amplitude, 1kHz sine wave is presented at the input. More specifically, for this calculation $V_a(t) = \sin(2\pi \times 10^3)$, and $h_{imp}(t)$ is taken as the lower trace in *Fig. 6*. The result $V_{out}(t)$ is displayed in *Fig. 7*. Note that since the corner frequency of this RC filter is 500Hz, you can expect the filter ideally to attenuate the 1kHz signal by a factor of .447, which is very close to the result obtained in *Fig. 7*.

CONCLUSION

So far, I have attempted to outline the theory I used to implement the Virtual Crossover. I did not wish to present so much detail that the discussion would become too tedious, but at the same time my intention was to give a clear idea of the theoretical considerations involved in my implementation of this device. I hope I have struck a good balance between the two. In Parts 2 and 3, I will present more practical information, details about the hardware implementation of the crossover, and some results as applied to a hybrid electrostatic/transmission line speaker system. ❖

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A Hybrid Tube/MOSFET SE Amp

This Class A, single-ended hybrid power amp combines the best of both worlds: the warm sound of tubes and the technological advances in today's power-MOSFET devices. **By Generoso Cozza**

For many years, power amplifiers used only vacuum tubes, and today's modern amplifiers use transistors almost exclusively. Tube amplifiers operate on the same principles as transistor amplifiers, but the internal construction may be considerably different. Generally tubes are devices that operate at high voltage and supply low current. In contrast, transistors operate at low voltage, but can supply high current. Also, tube amplifiers tend to dissipate a lot of energy in the form of heat, and in general are not very efficient.

One of the most striking differences between tube and transistor amplifiers is the tube amp's need of an output transformer. Because of the high output impedance of a tube circuit, it generally requires a transformer to properly deliver power to the loudspeaker. High-quality audio-output transformers are not only difficult to design, but tend to be large, heavy, and expensive. On the other hand, a transistor amp does not require an output transformer, and therefore tends to be more efficient, smaller, and less in need of periodic replacement.

Many people believe that the sonic quality of tube amplifiers can be both superior and possessed of a unique character. What is certain is that there are sonic differences between tube and transistor amps. I sincerely appreciate both worlds, and have had the opportunity to hear wonderfully sounding systems using both technologies.

COMBINED OPERATION

My intention in designing this hybrid amplifier (*Fig. 1*) was to combine the best attributes of both tube and transistor technologies. Tubes offer full and faithful sound reproduction, rich detail, brilliant clarity, and accurate tracking of complex waveforms. They are also better at reproducing deep bass and extended, sweet, natural highs. Transistors are able to drive even difficult speakers while providing authoritative bass performance.

In the hybrid amplifier, the magical midrange, the soundstage size, the air and overall musicality of a tube input stage passes directly to a low-distortion solid-state output stage that retains much of the good tube qualities, yet provides a better interface to modern loudspeakers.

THE HYBRID CIRCUIT

The circuit (*Fig. 2*) is a simple design that incorporates interesting ideas such as Erno Borbely's low-voltage tube operation[1] and Reinhard Hoffmann's Zen output stage with differential power supply.[2]

This hybrid amplifier is a two-stage, DC-coupled, single-ended Class A amp, capable of delivering around 30W in an 8Ω load, or 15W in a 4Ω load. You can easily increase the output power by paralleling more output MOSFET devices with its associated current source. Such parallel devices will increase the damping factor and lower the dependence on the load impedance.

A stereo amplifier with two output MOSFET devices per channel will provide more than 50+50W of pure Class A usable power to 6–8Ω loads. Due to its Class A operation, under such conditions the stereo amplifier will dissipate more than 300W, so you must use appropriate heatsinks (with at least a thermal resistance of 0.2°C/W) and a suitable well-ventilated enclosure.

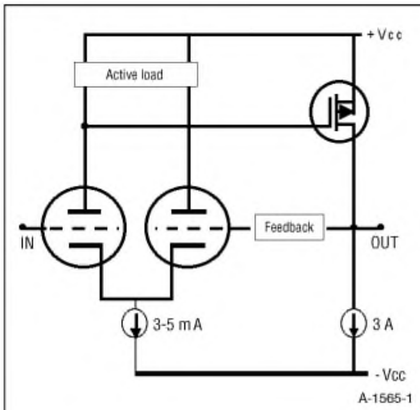

FIGURE 1: A simplified diagram of the hybrid amp.

The input stage is based on a 6DJ8/ECC88 dual-triode tube in a differential-amplifier configuration. I selected the 6DJ8 for its linearity and good operation at a 35–40V anode voltage. For the 6DJ8/6922/ECC88/E88CC, mu is constant within 20% from 0.4mA to at least 6mA, and the trend continues to flatten to 15mA. I chose an operating current of 3–5mA for each half of the stage, and a plate voltage of 35–40V to keep the dissipation well below the rated value of 1.8W. You can achieve almost all the 6DJ8's virtues at 5mA or lower.

The cathode current is supplied by constant-current source Q3, while Q1 and Q2 form an active load or current mirror. The active load forces the anode/cathode of both triode currents to be nearly equal, which provides excellent cancellation of the second-harmonic distortion, and contributes both to linearize the operation and increase the common-mode rejection and slew rate.

With P3, it is possible to adjust the bias current from 1 to approximately 7mA, and P1 controls the output offset voltage that you must adjust close to 0V.

OUTPUT STAGE

The output stage is composed of one or more P-channel MOSFETs in a single-ended, Class A, common-source configuration similar to the Nelson Pass Zen amp (for more details see http://www.passlabs.com/zenamp.htm). Its drain current is supplied by constant-current source Q4, which develops a 3A idle current using the specified value of R14. You can experiment with different values for the idle current by changing R14 in the formula $Id = (V_z - V_{gs})/R14 = 0.9/R14$. When defining different idle currents, you must consider the maximum ratings of the MOSFET output devices. In general, the Class A stage must carry a current of at least 50% more than the load will draw.

The overall closed-loop gain of the amplifier is around 20, and it depends on the values of R8 and R9. In this way, a 1V input signal will drive the amp to full power, so the output level of a typical CD player is sufficient to drive the amp. You may calculate a different gain by using the following formula: $A_v = 1 + (R9/R8)$.

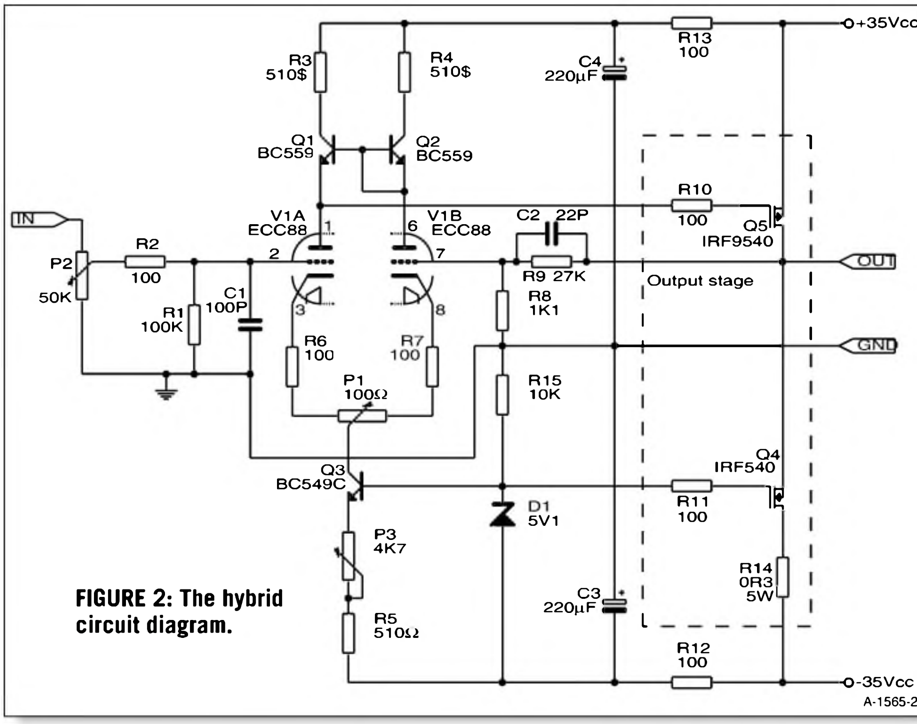

FIGURE 2: The hybrid circuit diagram.

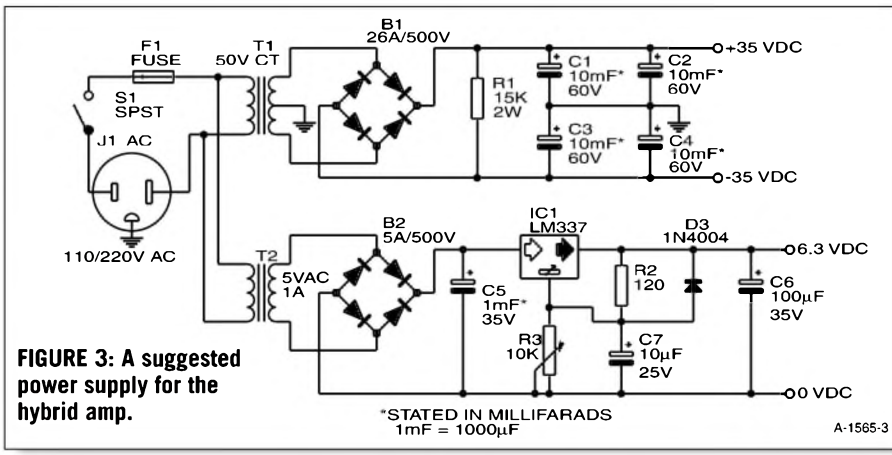

FIGURE 3: A suggested power supply for the hybrid amp.

A tested PCB that you can use to build this amp is available in Ivex WinBoard format. For a free copy of the file, please send an e-mail to generoso_cozza@hotmail.com. In this PCB, the tube and the MOSFET devices are mounted on the solder side.

THE POWER SUPPLY

Each channel of the hybrid amp requires a power supply capable of delivering 35+35V DC/6A for the main amplifier, and a regulated 6.3V DC/0.5A power supply for the tube heaters. The main amplifier power supply must include 20A or more diode/bridge rectifiers to support the constant 3A or greater current. *Figure 3* shows the schematic of a power supply for this amp.

THE RESULTS

This hybrid amp offers a flat response over the entire audio-frequency range. Even with inefficient speakers, you can appreciate its clarity and detailed musicality, especially when a CD player is directly connected to it. With one output device, the amp provides up to 20W with a THD under 1%, but it will work better with a second output stage in parallel.

I have had the opportunity to appreciate some of the best Class A amplifiers in the market, and I believe this hybrid emanates the same fragrance and sensation of freshness when you're listening to quality music materials. ❖

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A Transient-Perfect Second-Order Passive Crossover

This article develops a set of design rules for constructing transient-perfect second-order crossovers for two-way speaker systems.

By John Kreskovsky

The ideal loudspeaker should produce as its output a time-dependent acoustic waveform that is identical to the input electrical waveform. Theoretically, such a loudspeaker would necessarily possess flat, infinite bandwidth, and would be either a minimum-phase device or introduce a phase shift corresponding to a constant time delay. That is to say, the speaker would be transient perfect.

Because of the limited bandwidth of sound reproducing devices—be they dynamic drivers, planar magnetic or electrostatic panels, or ribbon devices—this ideal speaker is unattainable. The closest approximation is a system that has the broadest range of flat frequency response with either of the phase characteristics given above. Unfortunately, with very few exceptions, the best that today's technology has to offer from single drivers is a response that extends over seven or, in extreme cases, eight of the ten octave audio bands, 20Hz to 20kHz.

When dealing with conventional dynamic drivers, this range is typically more limited. For example, high-quality, 28mm dome tweeters are limited, at best, to a useful range extending over the top four octaves (1kHz to 20kHz). Similarly, a high-quality, 13cm midwoofer is limited to a useful range of about six octaves (50Hz to 3.5kHz). Thus, speaker builders are forced to compromise and employ multiple drivers to cover the audio band.

This use of multiple drivers necessitates inserting some kind of electrical network to blend the acoustic output of the drivers into a coherent sound field. When you consider the radiation characteristics of the drivers employed and the interference patterns that arise from the use of multiple drivers, it quickly becomes apparent that you can achieve the ideal response only over a very limited position in space. However, this should not deter you from pursuing the ideal of a transient-perfect design.

REVIEWING PREVIOUS WORK

It is a common misconception that you can obtain the desired transient-perfect response only with a first-order crossover using a Butterworth alignment. Small[1] showed that you can easily develop transfer functions for so-called constant-voltage crossovers—with nonsymmetrical high-pass (HP) and low-pass (LP) filter sections, that have flat-summed amplitude and phase responses. Given that $G_L(s)$ and $G_H(s)$ are the transfer functions of the LP and HP sections, respectively, then the desired result may be expressed as

$$G_L(s) + G_H(s) = 1 \qquad (1)$$

Equation (1) is a vector equation involving both amplitude and phase. It follows that for any choice of HP or LP section, you can obtain the complementary filter section from (1). For example, if you were to choose the filter characteristic for the HP section, the LP section transfer function would be given by

$$G_L(s) = 1 - G_H(s) \qquad (2)$$

The problem with using this approach is that while the summed response of the HP and LP filter sections yields a transient-perfect response, the transfer function of the complementary filter ultimately has a 6dB-per-octave rolloff.[1] Additionally, Small indicated that, other than in the first-order case, you cannot implement these constant-voltage crossovers using driver-terminated passive networks.

Figure 1 shows the HP and LP filter-section response curves derived from equation (2) when a 1kHz, third-order Butterworth filter is selected for the low-pass section. The decibel scale is 5dB per division in *Fig. 1*. Observe that the HP response has a peak of 4dB at the LP –3dB point and that the HP section is down 3dB at approximately 350Hz. *Figure 2* shows a similar result when the HP filter is a 1kHz, third-order

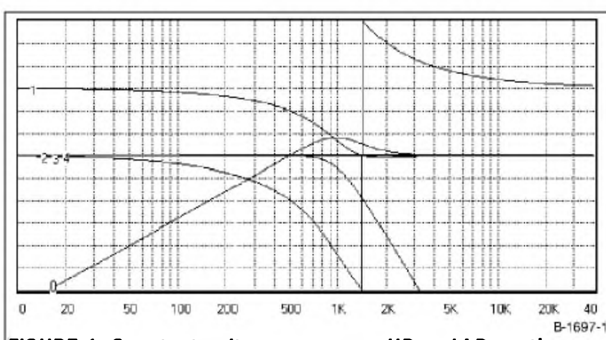

FIGURE 1: Constant-voltage crossover HP and LP sections using a third-order Butterworth low-pass filter; 0—HP amplitude response, 1—HP phase response; scale: 5dB/division.

ABOUT THE AUTHOR

John Kreskovsky received an advanced degree in Mechanical Engineering from Pennsylvania State University in 1973. After graduation, he performed research in the field of fluid dynamics and CFD. In 1980 he became active in the fields of semiconductor-device physics, modeling, and materials research. He has been involved with speaker design as a hobby since the late 1960s, and he developed his own crossover-design software in the early '80s. Retired in 1997 at the age of 50 to pursue other interests, he retains a consulting position in the company he helped found, Scientific Research Associates.

Butterworth filter. In both cases, and in all others to follow, both filter sections are connected with normal polarity.

SYMMETRIC FILTER FUNCTIONS

Small also showed it was possible to derive constant-voltage crossovers with symmetric filter characteristics. He provided an example using filters with second- and third-order asymptotic slopes. Again, these filters have unconventional transfer functions and are not suited for implementation with passive networks. Some 14 years later, Lipshitz and Vanderkooy[2] revisited the use of higher-order filter functions—in particular, symmetric ones—in an effort to produce linear-phase high-slope crossover networks.

Contrary to Small, Lipshitz and Vanderkooy examined standard higher-order filter alignments, including the Butterworth, Bessel, and Linkwitz-Riley (L-R). They showed that by overlapping the HP and LP sections of these symmetric filters, you could reduce the summed response to a minimum-phase response. However, the amplitude response of these overlapped filters was not flat.

The object was then to equalize the response using an additional minimum-phase circuit, finally resulting in a flat, minimum-phase response. Mathematically, you may express this as

$$G_{eq}(s) \times (G_L(s) + G_H(s)) = 1 \qquad (3),$$

and the required equalization response

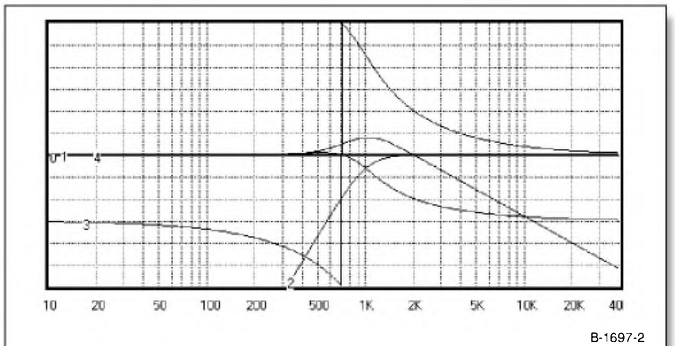

FIGURE 2: Constant-voltage crossover HP and LP sections using a third-order Butterworth high-pass filter; scale: 5dB/division.

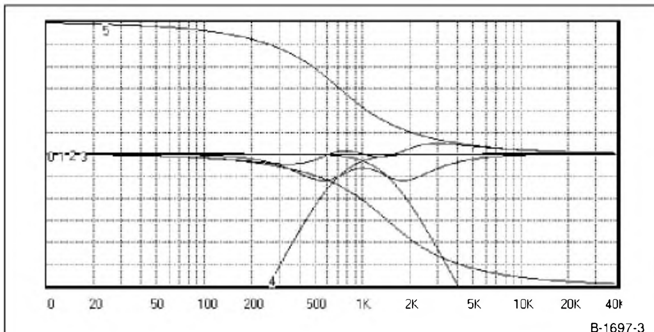

FIGURE 3: Summed second-order overlapped Butterworth crossover at 1kHz following Lipshitz and Vanderkooy;[2] overlap parameter = 2; scale: 3dB/division.

then becomes

$$G_{eq}(s) = 1/(G_L(s) + G_H(s)) \qquad (4),$$

where $G_{eq}(s)$ is the transfer function of the equalization network.

Lipshitz and Vanderkooy concluded that crossovers employing filters of higher than third-order required unreasonable equalization, and were therefore of limited use. They also provided the criteria for the minimum overlap required to render the summed-response minimum phase, but they did not indicate how equalization circuits for any order of crossover could be developed.

Following Lipshitz and Vanderkooy, for a given crossover frequency, fo, the corner frequencies of the HP and LP sections are related to the overlap parameter γ as

$$F_H = f_o/\sqrt{\gamma} \qquad (5)$$

$$F_L = f_o \times \sqrt{\gamma} \qquad (6)$$

Based on the relationships of equations (5) and (6), *Fig. 3* shows the summed amplitude and phase response for a 1kHz crossover derived from second-order Butterworth HP and LP sections with an overlap parameter of 2. Note the peak in the amplitude response at the crossover frequency, 1kHz, and the dips to either side, which give rise to multiple inflection points before the response goes to flatband below 100Hz and above 10kHz. Also note the phase response, which is indicative of a minimum-phase system in accordance with Bode's theorem.[3] It also exhibits a double-crested behavior, with multiple inflection points.

In contrast, *Fig. 4* shows the result when second-order Linkwitz-Riley filter sections are used in place of the Butterworth sections. With these filter sections, the summed amplitude response is smooth, with a single dip at the 1kHz crossover point. The phase response also exhibits a smooth functional shape, with a single inflexion point at the crossover frequency. As will become apparent, I found this observation to be an important one.

Finally, *Fig. 5* shows the result when using third-order Butterworth filter sections with an overlap parameter of 4. The response resembles that of the second-order Butterworth system, but with much deeper dips in the amplitude response and more significant phase variation. However, the phase response is still minimum phase.

MY PRESENT APPROACH

My approach to developing a transient-perfect crossover is an extension of the work of Lipshitz and Vanderkooy, limited to the use of second-order filters. However, rather than require the filter transfer functions to be those of traditional filters, I have left the specific alignment of the HP and LP filters as variables in the overall design. The alignment is to be determined, in conjunction with the alignment of a specified equalization circuit, in an effort to achieve flat amplitude and phase response of the final system.

As you will see, the result I found is a class of equalized, second-order crossover networks for which the overlap and filter- and equalization-section Qs are optimum for the required level of equalization. I have developed the filters with the intention of implementing the equalization through the use of pas-

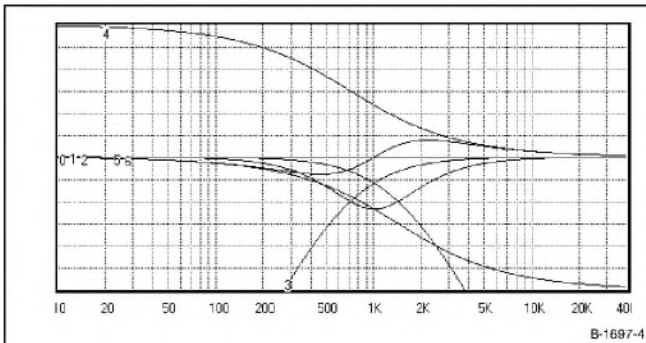

FIGURE 4: Summed second-order overlapped Linkwitz-Riley crossover at 1kHz following Lipshitz and Vanderkooy;[2] overlap parameter = 2; scale: 3dB/division.

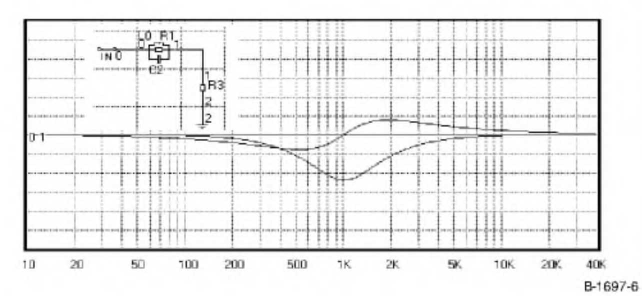

FIGURE 6: Trap circuit having a response similar to overlapped L-R crossover; scale: 3dB/division.

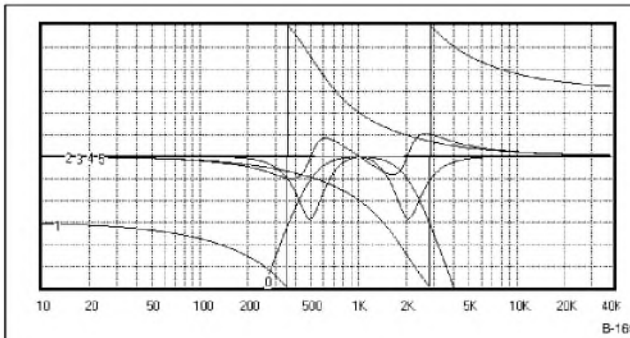

FIGURE 5: Summed third-order overlapped Butterworth crossover at 1kHz following Lipshitz and Vanderkooy;[2] overlap parameter = 4; scale: 3dB/division.

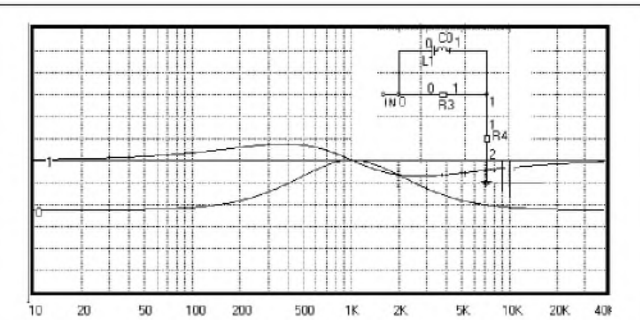

FIGURE 7: Passive equalization circuit for overlapped second-order Linkwitz-Riley crossover; scale: 3dB/division.

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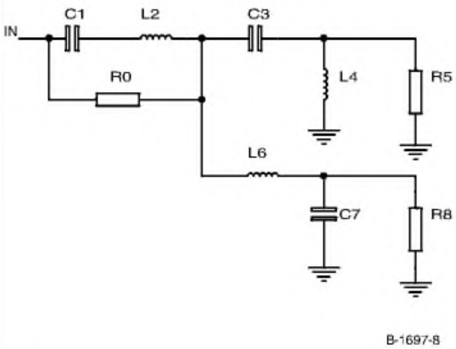

FIGURE 8: Second-order crossover and equalization used in CALSOD optimization.

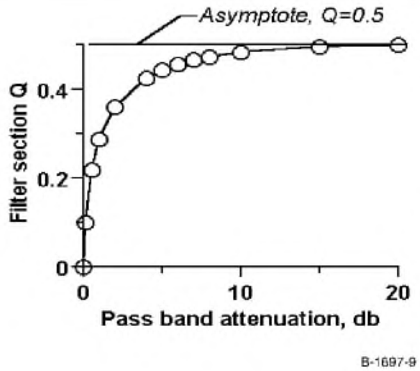

FIGURE 9: Filter section Q versus passband attenuation.

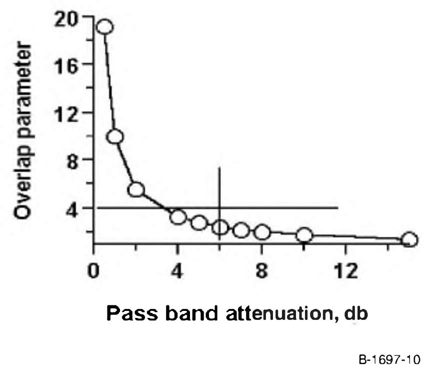

FIGURE 10: Overlap parameter versus passband attenuation.

sive circuitry. However, you could substitute an active equalization circuit with the same transfer function.

I observed that the summed response of the Linkwitz-Riley overlapped crossover shown in *Fig. 4* looked very similar to that of an RLC trap circuit. Such a circuit and its response are shown in *Fig. 6*. This similarity led me to consider a passive equalization circuit as shown in *Fig. 7*, where R4 represents the load due to the crossover. Using the Linkwitz-Riley HP and LP filter sections with $\gamma = 2$, the required equalization is in excess of 6dB, which I considered excessive.

However, to determine whether the approach was feasible, I set up the circuit shown in *Fig. 8* in CALSOD[4] and optimized the components for the equalization circuit. I set the target response level at the minimum in the unequalized response, which occurs at the crossover frequency (*Fig. 4*). I held the filter functions fixed at the Linkwitz-Riley values and performed the optimization. The results were encouraging. The frequency response was flat to within ±0.25dB with the corresponding minimum phase response.

Failure to achieve a perfectly flat response indicated that the shape of the response of the summed, unequalized filters did not correspond exactly to that obtainable with the resonant RLC equalization network. That is to say, $G_{eq}(s)$, as determined from equation (4), did not correspond to a functional form obtainable from an RLC circuit.

To determine whether a more accurate response could be obtained with the simple RLC equalization circuit, I repeated the optimization with all circuit variables included in the procedure. This was successful, yielding a perfectly flat amplitude response with zero phase shift—the perfect second-order, symmetric, passively implemented crossover.

Analysis of the circuit components showed that the overlap had changed somewhat, as did the filter-section Q. However, the only significant negative factor was the rather excessive power dissipation that occurred in what would normally be considered the flatband, or passband, regions of the unequalized HP and LP filter sections.

The observation that a slight alteration of the filter-section Q and overlap produced a perfectly flat summed, equalized response also suggested to me the possibility of determining specific values of the filter Q and overlap for different levels of attenuation. The attenuation corresponds to the depth of the dip in the summed, unequalized amplitude response at the crossover frequency. I will next discuss the determination of the relationships between this attenuation and the filter section Q_f, the overlap, γ, and Q_{eq} of the equalization circuit.

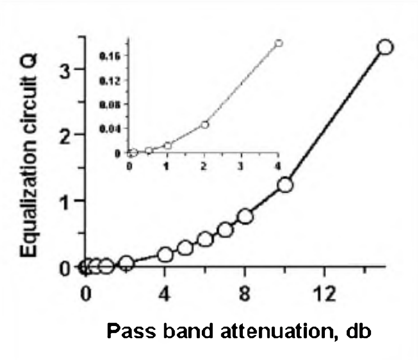

FIGURE 11: Q of required equalization circuit.

EXCESSIVE ATTENUATION

I next considered the problem of the excessive power dissipation, or attenuation, required to passively equalize the crossover. Since the equalization circuit is rather simple, it would not be difficult to construct and place an active equalizer between the preamplifier and amplifier. Such a system would have the added benefits that it would ultimately allow for easy biwiring or biamplifying, enabling the HP and LP sections to default to relatively standard-type filter functions closely approximating the Linkwitz-Riley characteristic.

In any case, I thought there would clearly be a benefit to a purely passive system, with active equalization viewed as an available option. In view of the fact that many of today's commercial speakers operate in the range of 82 to 90dB/W/m—or perhaps better stated as 82 to 90dB/2.83V/m—it would seem that if the equalization could be brought into the 4 or 5dB range, the passive approach would be acceptable. With today's higher-quality, higher-sensitivity drivers, I anticipated that it would be possible to construct systems with sensitivities in the 84 to 86dB/W/m range. While perhaps not the most efficient, I consider this to be a reasonable tradeoff.

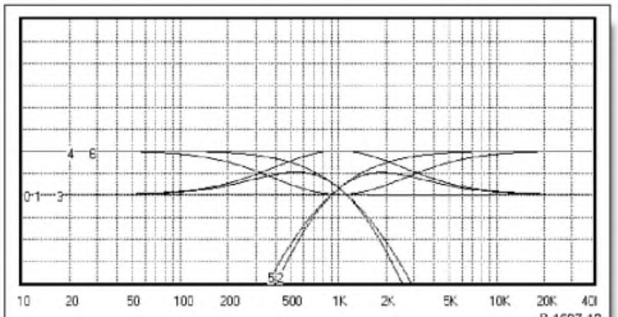

FIGURE 12: Amplitude response of HP and LP filter sections with and without equalization; summed response with and without equalization; and the equalization response; scale: 2dB/division.

In an effort to determine whether such systems could be developed with passive equalization, I made a series of calculations with CALSOD, setting different target attenuation levels for the equalization circuit. The results were successful and quite interesting. In every case the optimization procedure found solutions for symmetric HP and LP filter functions that yielded perfectly flat response and zero-phase shift to within the limits set for convergence.

The results are summarized in *Figs. 9–11*, where I show the filter section Q_f, the overlap parameter, γ, and the Q of the required equalization circuit, Q_{eq}, as functions of the required passband attenuation. In the extreme, as the attenuation goes to 0dB, the filter section Q goes to zero, and the overlap parameter goes to infinity. The interpretation of this is that the filters reduce to the standard first-order Butterworth alignment, with a crossover frequency of f_o at this point.

At the other extreme, as the attenuation approaches infinity, the filter section Q goes to 0.5, and the overlap parameter is unity. This is the standard Linkwitz-Riley crossover with the driv-

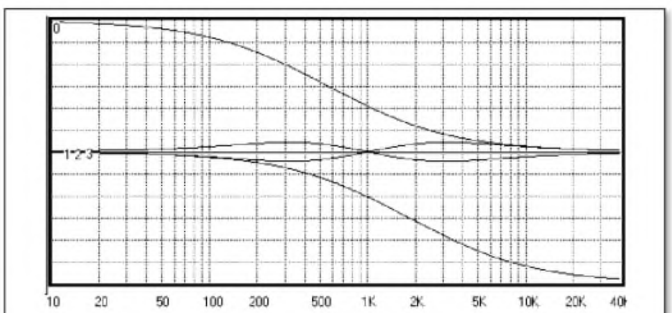

FIGURE 13: Phase response of unequalized HP and LP sections; summed unequalized phase; and phase of equalization circuit; scale: 30° per division.

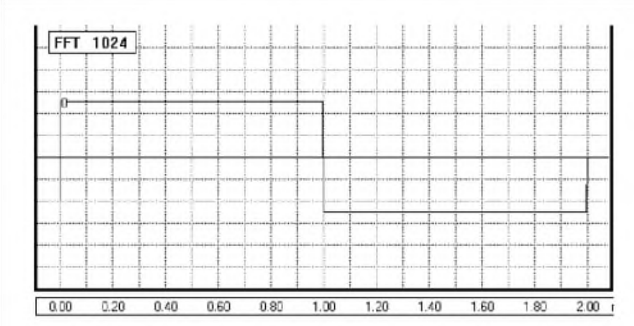

FIGURE 14: Simulated response to a 500Hz square-wave input.

ers wired in phase, and is the limiting minimum-phase configuration. An equalization circuit with a Q of infinity and infinite flatband attenuation would render this limiting case flat. A minimum-phase crossover with a flat response of zero amplitude is of no practical use, yet it is an important result since it shows that filter sections with Q greater than 0.5 cannot be overlapped and equalized in a manner that will yield flat response with a simple second-order RLC equalization circuit.

AVOIDING EXCESSIVE EQUALIZATION

Not only because of the loss of power when using passive equalization, but also because of the excessive and steep phase variations that appear in the individual filter responses under such conditions, you should avoid excessive equalization used with low overlap. While the summed system response is theoretically flat on the design axis in any case, such rapid phase variations will adversely affect the off-axis response of a multiple-driver system. Similarly, excessive overlap defeats the intent of the design, namely to limit the operational bandwidth of the HP and LP sections relative to that of a first-order system.

I recommend that acceptable solutions are in the area of 6dB attenuation or less, with an overlap parameter of less than 4. The boxed region in *Fig. 10* indicates this range. For example, at 4dB attenuation, the filter section Q is approximately 0.425, the overlap parameter is about 3.25, and the Q of the

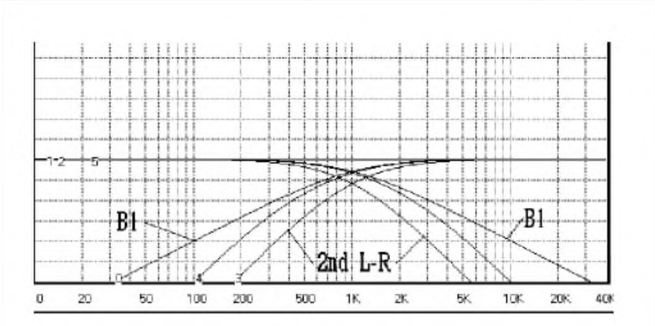

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FIGURE 15: Comparison of second-order L-R, first-order Butterworth, and the author's HP and LP filter amplitude responses; scale: 5dB/division.

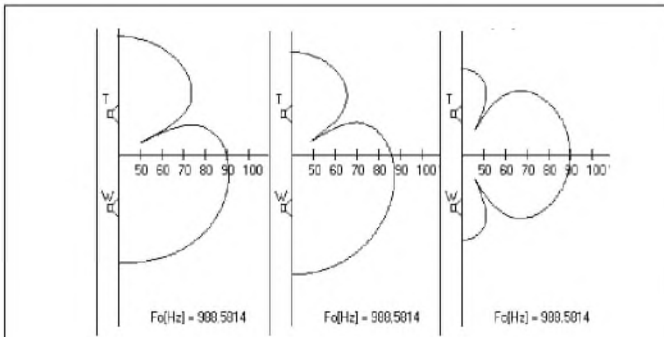

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FIGURE 16: Theoretical vertical polar response at 1kHz for first-order Butterworth (left), the author's second-order overlapped (center), and second-order L-R (right) crossovers.

required equalization circuit is approximately 0.18. This represents a reasonable design point.

Choosing a crossover frequency of 1kHz, the HP and LP corner frequencies are given by equations (5) and (6) as

$$F_H = 1000/\sqrt{\gamma} \approx 555\text{Hz} \qquad (7)$$

$$F_L = 1000 \times \sqrt{\gamma} \approx 1803\text{Hz} \qquad (8)$$

The capacitor and inductor values for the filter sections are obtained from the filter Q_f and the load resistance R by

$$C_f = Q_f/(2\pi f \times R) \qquad (9)$$

and

$$L_f = R/(2\pi f \times Qf) \qquad (10),$$

and the required resistance for the equalization circuit is given by

$$R_{eq} = R(1 - \log^{-1}[-A/20])/\log^{-1}[-A/20] \qquad (11),$$

where R is again the load resistance and A is the desired attenuation, 4dB in this case. Finally, the capacitor and inductor values for the equalization circuit are given as

$$C_{eq} = 1/Q_f(2\pi f_o \times R) \qquad (12)$$

$$L_{eq} = R \times Q_f/2\pi f_o \qquad (13)$$

Here, R is once again the load resistance, and f_o is the crossover frequency.

CROSSOVER SIMULATION

I performed a simulation of a crossover based on the selected crossover frequency of 1kHz and a 4dB attenuation, with the circuit components given by equations 9–13 (*Figs. 12–14*). *Figure 12* shows the simulated HP and LP filter amplitude response with and without equalization, the summed response with and without equalization, and the response of the equalization circuit alone.

Examination of the figure shows the smooth rolloff of the unequalized HP and LP sections, typical of a second-order filter with a Q near the critically damped value. Also note that the equalized filter responses show peaks of approximately 2dB at the corner frequencies for the HP and LP sections, respectively. Finally, note the symmetry of the equalizer response and the summed, unequalized filter response about the –2dB point.

Figure 13 shows the phase response of the unequalized HP and LP sections, along with that of the summed, unequalized response and the phase of the equalization circuit. Note that because of the overlap, the phase difference between the HP and LP sections at the crossover point is approximately 120° as opposed to the typical 180° of an unoverlapped second-order crossover. In addition, the phase of the summed response shows only a little more than ±15° phase shift prior to applying the equalization. As expected, the phase response of the equalization circuit is symmetric to the summed HP+LP response about the zero axes.

Finally, the simulated response to a 500Hz square wave is shown in *Fig. 14*. The 500Hz frequency places the fundamental frequency of the square wave below the crossover point, with the harmonics above it. As expected for a filter with no phase shift and flat amplitude response, the reproduction is perfect, with the exception of the reduction in amplitude from the application of passive equalization. The time scale in the figure is msec.

A WORD ABOUT OVERLAP

One of the concerns about this type of crossover, or any other in which the HP and LP sections are overlapped when used in the design of a loudspeaker, is that such a filter requires using drivers with extended bandwidth. The question arises as to just how much overlap is required. Perhaps the easiest way to answer this is by way of example.

In *Fig. 15* I have compared the response of the unequalized HP and LP sections used in the previous example with the conventional L-R second-order filter and with the standard first-order Butterworth filter. Recall that for the current class of filters, in the limit of infinite overlap, the crossover reduces to the first-order Butterworth and the L-R[2] filter represents the limit of minimum overlap. Thus, it is apparent from *Fig. 15* that, regardless of the degree of overlap, the rolloff of the HP and LP sections of the current crossover will always lie between that of the first-order Butterworth and the second-order L-R crossover.

VERTICAL POLAR RESPONSE

Another issue with overlapped higher-order crossovers is the effect of the overlap on the vertical polar response (VPR). I investigated this by examining the VPR of a theoretical two-way system, using Sound Easy[5], with the woofer/tweeter spacing set at 12cm. The result is compared in *Fig. 16* to that for geometrically similar systems employing first-order Butterworth and second-order L-R crossovers. The crossover frequency for the systems is 1kHz. I computed the polar response as closely to this frequency as the software would allow, 988Hz.

Note that the polar response for the first-order and the present overlapped crossover is quite similar. This you would expect since, as *Fig. 15* shows, in the crossover region the present design follows the first-order Butterworth response very closely. The L-R crossover exhibits its classical symmetric response. While there has been, and continues to be, opposition to the use of first-order crossovers based on the polar-response characteristics, I believe a well-thought-out MTM design can go a long way towards alleviating the polar-response problems.

A QUICK EXPERIMENT

To provide experimental verification that such results can be obtained in

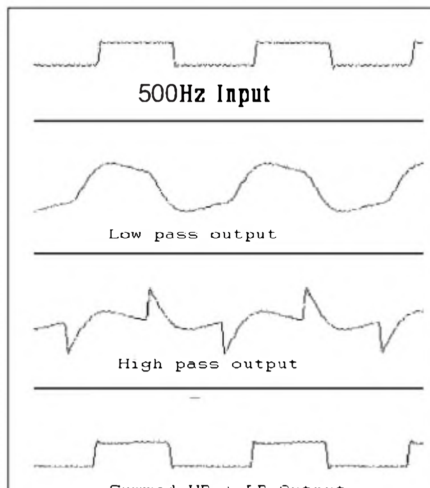

FIGURE 17: Experimental result for a 500Hz square wave.

practice, I built a 1kHz crossover on a breadboard and tested it. I made attempts to tune the network exactly, since I used spare parts. *Figure 17* shows the input 500Hz square wave, the output of the HP and the LP sections, and the summed response as measured using the IMP6 as the data-acquiring system. Please note that the response curves are not synchronized; that is, the cycle does not start at the same point on each curve. Nevertheless, it is clear that theory and practice are in close agreement. The summed output is an excellent representation of the input.

APPLICATION TO REAL LOUDSPEAKERS

Application of this crossover design to real loudspeakers is complicated by several issues. Easily addressed are the facts that real drivers mounted in a box or on a baffle have complex, frequency-dependent impedance functions and do not necessarily have flat amplitude response. In addition, it is unlikely that both the woofer and the tweeter would be of the same impedance, but if so, you can also overcome this fairly directly.

However, the low-frequency rolloff of the woofer/box alignment introduces phase shifts below 100 or 200Hz, which you can not readily deal with by using passive networks. This suggests using sealed-box alignments because of the lower phase shifts they introduce at low frequencies.

Accepting this limitation, the first step in the design procedure is to choose drivers with a suitable overlap in useful frequency range. Once you have specified suitable drivers and set the overlap, you can determine the required filter section and equalizer Q values from the design curves in *Figs. 9* and *11*.

The three necessary components, the HP, LP, and equalization section, are all straightforward second-order systems. Just as in the case of the design of a second L-R crossover, you must design the HP and LP sections to yield acoustic response functions that match those of the specific filter function. That is to say, it is the acoustic output of the drivers, measured at the design point, that must match the amplitude and phase response of the corresponding filter transfer function, not the electrical signal applied at the driver terminals.

Provided that the networks used in developing the acoustic filter response are minimum-phase networks, the acoustic response of the driver plus the filter will also be minimum phase owing to the nature of loudspeaker drivers. If you possess the skills to design filters yielding conventional acoustic HP and LP transfer functions—such as Linkwitz-Riley, Bessel, or Butterworth—then the design of the required HP and LP sections of this crossover should present no great challenge. Be careful, however, just as with conventional filters, to ensure that the acoustic centers of the drivers are correctly aligned. Otherwise, the summed driver response will not exhibit the correct amplitude and phase response.

A MORE DIFFICULT CHALLENGE

Development of the passive equalization circuit, however, can be a complex task. Ideally, the impedance response for the summed network must be compensated to yield, as closely as possible, constant impedance. While not always a simple task, you can generally do this through the addition of various Zobels and RLC shunts across the speaker terminals. Because of the double-peak nature of the woofer impedance in vented boxes, impedance compensation for such systems may prove difficult. For that reason, my approach of passive equalization is again better suited for sealed-box systems.

In cases of significant tweeter/woofer impedance mismatches, it is also possible to compensate the tweeter and woofer networks separately, and then develop separate equalization circuits for each. Whichever approach you take, here is another area in which the power of optimization programs such as CALSOD can be brought to bear. With the filter components known, you can hold them fixed and apply the full power of the optimization procedure to the equalization circuit. Of course, it is always possible to circumvent the complexity of passive equalization and default to an active equalization circuit, the design of which reduces to not much more than a textbook exercise.

The series-attenuation resistor in the equalization circuit must be taken into account in the design of the woofer enclosure. This resistor will result in an increase in the value of Q_{ES} of the driver, and will be reflected in an increased value for Q_{TS}. The net result will be the need for a greater box volume for correct woofer alignment. And while there is a loss in sensitivity because of the presence of this resistor, the aforementioned effects on the driver parameters will result in a lower system f_3 for a correctly aligned woofer. ❖

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A 211 SE Triode Amp

Attention, all first-time (or wanna be) builders of single-ended directly heated triode amps. This project's for you.

By Nick Soudas

I grew up listening to solid-state equipment, having been born too late to catch the tube wave. During my formative years I read about vacuum tubes, but always dismissed their use in audio, since the mighty transistor was far superior. After building a 2A3 single-ended (SE) amplifier and retrofitting a set of 300Bs in place so I could compare the two types of output tubes, I was amazed at what a SE directly-heated triode (DHT) amplifier could sound like.

I can now honestly say that I prefer vacuum tube amps (although I like tinkering with all types of audio gear). After the 2A3/300B project I wished to take the SE DHT concept to the next level and use a different output tube. As with most hobbies (or electronics, for that matter), it's not difficult to find the next level.

OUTPUT TUBE

While surfing the Internet and reading various periodicals, I noticed "big power triodes," namely, the 211/845 family. Even though Svetlana has a current line of power triodes (572 and 811 series), I was curious about the older American triodes. Supposedly, for the ultimate SE DHT sound, these were the tubes to use.

ABOUT THE AUTHOR

Nick Soudas has worked in the mobile electronics field since the early '80s (then in its infancy). He received formal electronics training in the Army and is still active in the Reserve. Currently, he is employed by Jensen Mobile Electronics as part of an electronics design team, designing, testing, and evaluating current/new products/technologies. His main focus is amplifiers and signal processor design. When not tinkering with electronics, he likes to build and fly remote-control airplanes. He can be reached at nsoudas@mindspring.com.

The 211 was originally designed in the early '20s, deriving its heritage from an earlier tube used during WWI in Navy radio transmitters. If you compare the plate characteristics of DHTs, such as the 2A3 or the 300B, the 211/845 family is

PHOTO 1: Front view of amp showing rack-mount chassis.

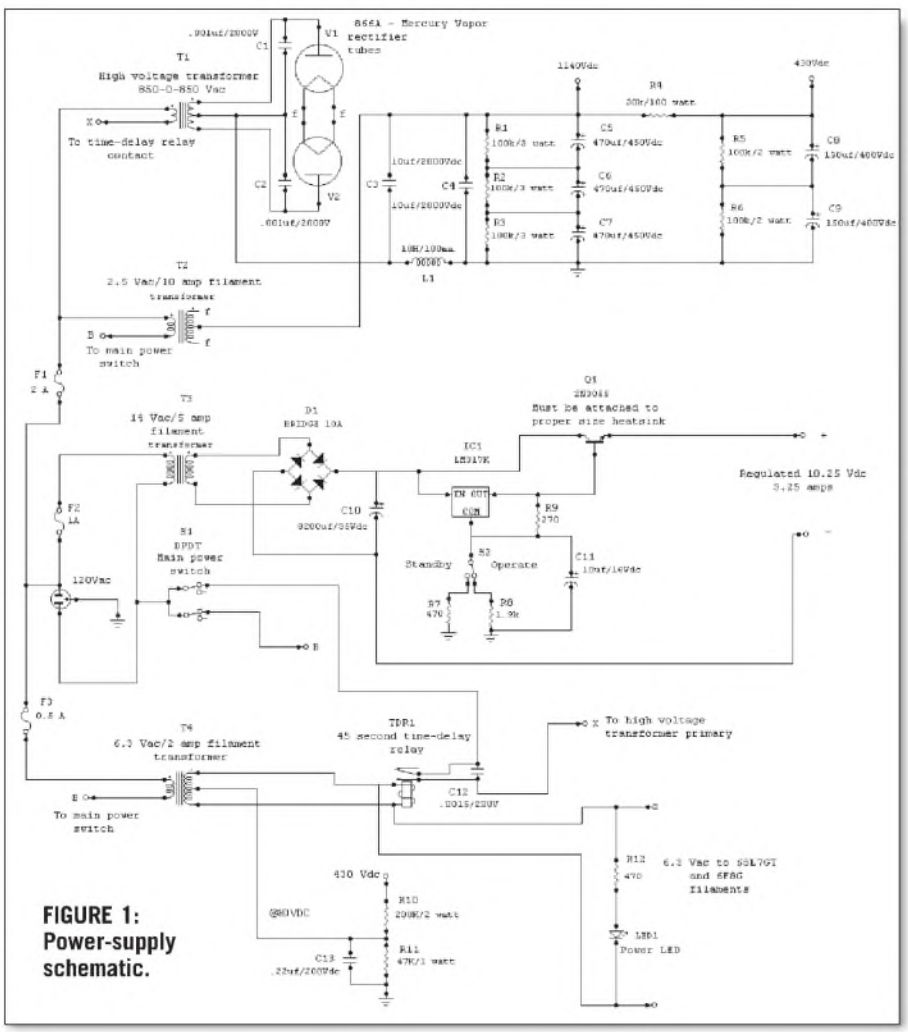

FIGURE 1: Power-supply schematic.

TABLE 1
POWER SUPPLY LIST (ONE CHANNEL)

HIGH VOLTAGE SECTION	
V1, V2	866A, mercury vapor rectifier tubes (AES)
TDR1	Amperite normally open (NO) 45 second delay, 6.3V, Amperite Part # 6NO45 (AES)
T1	Hammond plate transformer, 890-0-890, 175mA DC, Hammond catalog number 726 (AES)
T2	Hammond filament transformer, 2.5V AC, 10A, center-tapped, Hammond catalog number 166S2 (AES)
T4	Hammond filament transformer, catalog number 166L6, 6.3V AC, 2A, center-tapped (AES)
L1	Hammond choke, catalog number 158M, 10H, 100mA, 262Ω (AES)
S1	Double-pole, double-throw 120V AC/10A (RS, SP)
F1	2A, 250V, fast blow (RS)
C1, C2	.001µF/2000V DC, ceramic disc (AES), part number C-D001-6K
C3, C4	10µF/2000V DC, oil-filled (SP)
C5–C7	470µF/450V DC, aluminum electrolytic (SP)
C8, C9	150µF/400V DC, aluminum electrolytic (SP)
C12	.0015µF/200V DC, ceramic disc (AES, RS, SP)
R1–R3	100k/3W, carbon composition (SP)
R4	30k/100W (minimum), wire wound (SP)
R5, R6	100k/2W, carbon composition (SP)
**	20k/200W, wire wound, with sliding tap. Optional, see text (SP)
**	1.25k/50W, wire wound. Optional, see text (SP)
2	4-pin sockets for 866As, P-ST4-312 (AES) Plate caps for 866As, P-SG115 (AES)
5′	18ga high-voltage wire, 2000V or higher (SP)
211 FILAMENT POWER SUPPLY	
T3	Hammond filament transformer, 14V AC, 6A, center-tap not used. Hammond catalog number 166Q14 (AES)
IC1	LM317, adjustable regulator (RS, SP)
Q1	2N3055 NPN power transistor or equivalent (RS, SP)
D1	Bridge rectifier, 10A or higher (RS)
F2	1A, 250V, fast blow (RS)
S2	Single-pole, double-throw 120V AC/3A (RS)
C10	8200µF/35V DC, aluminum electrolytic (RS, SP)
C11	10µF/16V DC, aluminum electrolytic (RS)
R7	470Ω/.25W, carbon or metal film (RS, SP)
R8	1.9k (1.8k + 100)/.25W, carbon or metal film (RS, SP)
R9	270Ω/.25W, carbon or metal film (RS, SP)
**	Heatsink for Q1, pass transistor (SP). Optional, if using an aluminum chassis you can mount the transistor directly to the chassis using a silicone thermal pad to isolate the collector from chassis ground.
INPUT/DRIVER FILAMENT POWER SUPPLY	
T4	6.3V AC, shared with high-voltage supply
R10	200k/2W, carbon composition (SP)
R11	47k/1W, carbon composition (SP)
R12	470Ω/.25W, carbon film/composition (RS, SP)
F3	0.5A, 250, fast blow (RS)
C13	.22µF/200V DC, ceramic disc, orange drop (RS, SP)
MISCELLANEOUS (COMMON TO ALL POWER-SUPPLY SECTIONS)	
1	IEC 18/3 power cord (AES, SP)
1	IEC chassis/panel mount power connector (AES, SP)
3	Chassis/panel mount fuse holders (RS, AES, SP) 5 lug terminal strips (AES, SP)
5′	18ga 600V general-purpose wire (AES, SP)
5′	14ga ground wire, solid core, no enamel or varnish (SP, HD)
LED1	Standard "RED" (or any color) panel mount LED (RS) 120V AC box fan, optional (RS, SP)

much more linear. Of course, this linearity has a price: a high plate voltage.

I had three goals in mind when I started this project. First, I wanted to design a direct-coupled input/driver circuit that would provide the proper drive voltage to the grid of the 211. Second, the circuit had to be simple (for sonic reasons), but able to supply the full drive voltage to the 211 grid. Third, I wanted to keep construction simple (for building purposes), to encourage other do-it-yourselfers to try a 211-based amplifier.

I came across a few designs using 211 and 845 transmitting tubes. Some of the designs used choke loaded plates, while others used transformer coupling between voltage gain stages. All the designs had at least two gain stages. I found this unusual since the 211 grid needed only about 150V p-p, which is moderate.

After tossing around the idea for a while, I decided to use one of my favorite circuits—a voltage amplifier direct-coupled to a cathode follower. I chose to use new Chinese 211 tubes to test my design, and then retrofit some NOS American 211s (*Photo 1*).

PHOTO 2: High voltage/filament power-supply section for the 6SL7 and 6F8G. Time-delay relay (upper-left corner). Part of amplifier section (top-right corner).

INPUT/DRIVER STAGE

I have always preferred octal preamp tubes instead of the miniature glass type. Every project I have built in the past few years has used at least one 6SN7 or 6SL7. The plate curves for these octal tubes are very linear. Initially, I wished to use the 6SL7 in a shunt regulated push-pull (SRPP) configuration to get maximum gain, but since the heater-to-cathode voltage is only 90V, I was unable to use the SRPP topology with a high plate voltage. Typically, a SRPP stage will have half of the plate voltage on the upper cathode.

To see why I was unable to use the SRPP stage, assume a plate voltage of 430V. If both cathode resistors are the same (1kΩ), then I should have about 215V on the upper cathode with reference to ground. If I bias the heater on the 6SL7 to 90V (maximum rating according to data), then I still have a 125V difference between the upper cathode and the heater.

Although the 6SL7 could probably handle the voltage difference, I was concerned about not stressing the tube to ensure that the circuit operated well within ratings for a long, trouble-free life. Since I wished to use the 6SL7, I pondered other ways that I might use it. I finally decided to parallel both triode sections of the 6SL7 to lower the plate resistance, and to get a better noise figure.

While experimenting with the plate curves for the 6SL7, I discovered that, with a plate voltage of 435V, I could easily get about 150V p-p. The input/driver stage load line exhibited .32% THD at 13V RMS output. This voltage output coincided with 1W output to the speaker. The input stage distortion could be lower, but later I'll explain how I reduced it to extremely low levels. The output stage clips before the input/driver circuit, which, in my opinion, is the way an amp should behave.

While paralleling both sections offers some advantages, there is also a disadvantage; the shunt input capacitance doubles too. I was concerned about this at first, but after some frequency-response measurements of the input/driver stage, the bandwidth seemed in order. The actual input/driver circuit (not connected to the 211) swings 153V p-p (54V RMS) with a bandwidth from 10Hz (down −0.53dB) to 90kHz (down −3.0dB).

The 6SL7 has some cousins, which you should not overlook, the 12SL7, 7F7, and 14F7. Sylvania manufactured the 7F7 and 14F7 to gain some market share from RCA, who produced the octal-type tubes. Except for the different type of base, the Sylvania tubes are identical electrically to the RCA octals.

Instead of using an octal base, Sylvania had a Loktal base; that is, an octal base with a lock-in feature. The Loktals had eight pins, but they were much smaller than the octal type. Any of these 6SL7 derivatives will produce excellent sonic results.

The cathode-follower section is direct-coupled to the input stage for low phase shift and stability. A 6J5, 6C5, 6P5G, or a 6SN7 would have worked great in this section, but I preferred to use a shoulder-type tube for aesthetic

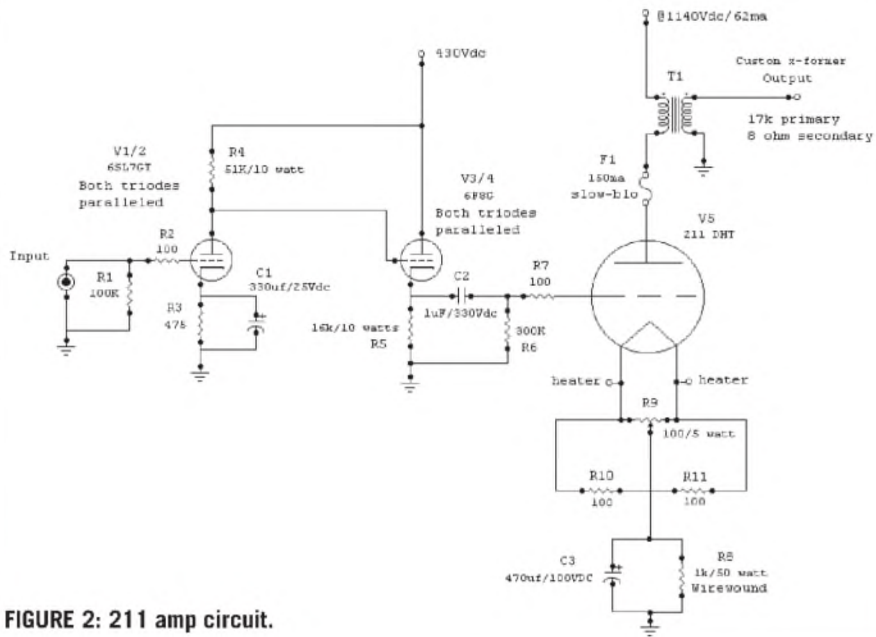

FIGURE 2: 211 amp circuit.

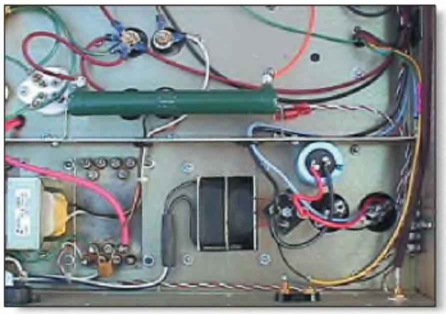

PHOTO 3: 211 filament power supply. Transformer, rectifier, and filter capacitor (bottom right).

reasons. The 6F8G is very similar to the 6SN7 family of tubes; it has a shoulder-type bulb (which I think is more nostalgic-looking), and its internal capacitance is slightly different. Plus, I wished to parallel the triode sections (for more driving current) to ensure that the 211 grid capacitance would not load down the cathode follower.

POWER SUPPLY

The power supply for this amplifier (*Fig. 1*) may seem very complex at first, but, if you break it down into three basic sections, it does not seem that daunting:

1. Main B+ high voltage/rectifier filaments/time delay
2. 211 filament
3. Input/driver filament

Table 1 shows the power-supply parts list. Ironically, although the amplifier circuit is very simple, the power-supply turned out to be a monster. It occupies two-thirds of my chassis (*Photo 2*).

This was my first time building a power supply of this type and magnitude. While the voltages in the power supply are quite high, do not be discouraged, just take a little extra time and be extremely careful when working around the power-supply section. The 211/845 family of power tubes can operate with 1250V on the plate. Anything over 1000V is enough to improve the tubes' sound, in my opinion.

The tube is rated at 75W, which seems rather low given the size of the plate. Early 211s had a stamped plate, and later ones utilized a graphite-type plate. The Chinese tubes were modeled after the later American tubes, so they

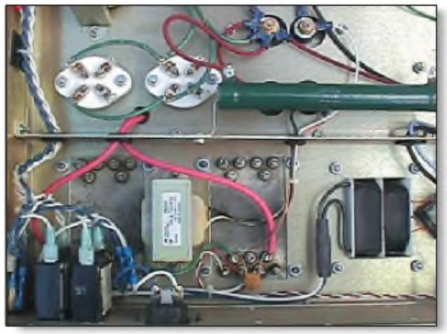

PHOTO 4: Main high-voltage transformers and power-supply section. Mercury vapor tubes/sockets (upper-left corner). Voltage divider resistor for input tube section (big resistor in center). 211 filament transformer (bottom right).

all have (carbon) graphite plates. The older American tubes (with graphite plates) can probably handle a 100W plate dissipation, but I doubt the Chinese clones can.

Running the tubes at 1000V/60mA produces a power dissipation of 60W, which is less than the rating, so tube life should be extended. With this combination of voltage and current, I realize around 14.5W/channel from my amplifier at 3% THD. I used a "pi" input filter directly after the 866As.

The first capacitor is a 10µF/2000V DC oil capacitor, followed by a Hammond choke rated at 10H/100mA, and then another 10µF/2000V DC oil capacitor. Initially this input filter yielded

PHOTO 5: Front view of amp showing front-panel controls. Power indicator, far left. Switch next to power indicator is the main power switch. Far-right switch is the 211 filament "operate/standby" switch. Right-hand side of panel, contains an optional current meter to monitor 211 output tube current.

good results, but a little hum was noticeable on the output, so I added additional capacitance in parallel to the second oil capacitor. Since I couldn't find a high capacity/high voltage capacitor locally, nor could I afford an adequate oil capacitor, I reluctantly used aluminum electrolytic capacitors...after convincing myself that it is perfectly acceptable to mix and match capacitors (for that sonic flavor).

The Hammond choke has a voltage rating of 400V DC. Not wishing to risk voltage breakdown of the choke, I used "negative lead" filtering. This is very common with higher voltage power supplies. My B+ high-voltage ended up

TABLE 2
AMPLIFIER PARTS LIST (ONE CHANNEL)

V1, V2	6SL7GT, NOS, both triodes paralleled (AES)
V3, V4	6F8G, NOS, both triodes paralleled (AES)
V5	211, NOS/Chinese (TS, TW)
T1	Custom output transformer 17.1k primary/8Ω secondary (CL)
F1	.15A slo-blo fuse (RS)
C1	330µF/25V DC, aluminum electrolytic (RS)
C2	1µF/330V AC, motor, oil-filled capacitor (SP)
C3	470µF/100V DC, aluminum electrolytic (SP)
R1	100k/.25W, 1% metal/5% carbon film (RS, SP)
R2, R7	100Ω/.25W, 1% metal/5% carbon film (RS, SP)
R3	475Ω/.25W, 1% metal/5% carbon film (RS, SP)
R4	51k/10W (minimum), 1% wire wound (SP)
R5	16k/10W (minimum), 1% wire wound (SP)
R6	300k/.25W, 1% metal/5% carbon film (RS, SP)
R8	1k/50W (minimum), 1% wire wound (SP)
R9	100Ω variable/5W, 5% wire wound (SP)
R10, R11	100Ω/3W, 5% wire wound, metal oxide (SP)
MISCELLANEOUS	
1	Chassis/panel mount fuse holder (RS, SP)
1	Chassis/panel mount RCA connector (RS, SP)
1	5-way binding post, speaker output (AES, SP)
1	Grid cap for 6F8G, P-SG116 (AES)
1	4-pin socket for 211, P-ST4-192 (AES)
3	8-pin sockets for 6SL7, 6F8G, and time-delay relay, P-ST8-311 (AES)
3	5 lug terminal strips (AES, SP)

TABLE 3
INPUT/DRIVER STAGE

0.28V RMS input
1W output, at speaker
8Ω resistive load, at speaker
0.32% THD, at cathode-follower output
13V RMS, at cathode-follower output

0.95V RMS input
10W output, at speaker
8Ω resistive load, at speaker
1.15% THD, at cathode-follower output
42.5V RMS, at cathode-follower output

TABLE 4
THE COMPLETE 211 AMPLIFIER

1W output, at speaker
8Ω resistive load, at speaker
0.085% THD, at speaker output—second harmonic

3W output, at speaker
8Ω resistive load, at speaker
0.15% THD, at speaker output—second harmonic

5W output, at speaker
8Ω resistive load, at speaker
0.21% THD, at speaker output—second harmonic

10W output, at speaker
8Ω resistive load, at speaker
0.46% THD, at speaker output—second harmonic

12W output, at speaker
8Ω resistive load, at speaker
0.92% THD, at speaker output—second and third harmonics

14W output, at speaker
8Ω resistive load, at speaker
2.75% THD, at speaker output—second and third harmonics

TABLE 5
FREQUENCY RESPONSE

1W, reference power—8Ω load
Frequency response—20Hz–20kHz. Overall it was extremely flat, except for the slight dipping at the frequency extremes.
Down −1.1dB at 20Hz
Down −1.2dB at 20kHz
Down −3dB at 11Hz
Down −3dB at 37kHz

TABLE 6
TEST POINTS

R4—Load side of this resistor should be about 430V DC. Do not let the voltage at this point exceed 450V DC.
Center tap of T2—1140V DC. Do not let the voltage at this point exceed 1160V DC.
211 filament—Standby mode, 3.4V DC/operate mode, 9.5V DC
R10, R11 junction—82V DC. Do not let this voltage exceed 90V DC.
V1, V2—Cathode, 1.9V DC/plate, 244V DC
V3, V4—Cathode, 244V DC/plate 249V DC
V5—Cathode, 60V DC/plate, 1065V DC

being 1140V DC, with a current draw from the amplifier of about 85mA. The *Radio Amateurs Handbook*, published by the American Radio Relay League (ARRL), 35th edition—or any edition around the late '50s or early '60s—contains much information regarding tube power supplies.

After obtaining all the power-supply components, I temporarily connected everything on a piece of wood (18″ × 18″ piece of pine) to test, measure, and work out any bugs in my paper design. You must be careful to choose wire with the correct voltage rating in the high voltage section. I used wire with a rating of 3,000V because this is what I had on hand, but you can use anything with a rating of 1,500V or higher.

The secondary of the high-voltage transformer needs to be 900-0-900/200mA, or close to this, for a full-wave capacitor input supply. Power-supply design is as hotly debated as amplifier design itself. I believe that if you are going to go through the trouble of designing and building a tube amplifier around modern components (or even NOS parts, for that matter), a tube-based power supply is the only logical choice. It is not as simple or inexpensive as a solid-state supply, but I think the extra money and effort are worth the sonic benefits.

The 211 filament power supply (*Photo 3*) is regulated DC. An LM317 adjustable IC regulator drives an external NPN pass transistor. The external pass transistor is necessary because the LM317 has a current rating of only 1.5A. The 211 tube requires 10.25V/3.25A AC or DC on the filaments.

When I built my 2A3 amplifiers, I initially used AC on all the filaments and was surprised to find hardly a trace of hum on the output. When I retrofitted the 300Bs, I decided to use regulated DC on the filaments because I did not have a 5V AC filament transformer. With DC on the output tube filament only, and AC on the input section, the slight hum previously present was now gone.

In keeping with my simple design philosophy, and having had good results in the past, I used AC on the input/driver filaments. 211 tubes that operate continuously seem to have a lower mortality rate than those frequently turned on and off. Apparently, this is due to the high in-rush current of the tube filament (and consequently, the rapid heating and cooling of the filament when it is turned on/off frequently contributes to the filament literally "breaking" in half as if bending a piece of metal back and forth until it breaks).

I decided to incorporate a standby switch for the 211 filament circuit. It has two operating modes: standby and operate. In the "standby" mode, the filament is just barely glowing orange. When switched to "operate," the filament gets full operating voltage.

MERCURY VAPOR RECTIFIERS

Being limited by the limited number of available tube rectifiers that could handle high voltage, I finally settled on using 866A mercury vapor (MV) rectifiers (*Photo 4*). The 866As can handle up to 10,000V and fairly high current surges (briefly). Mercury vapor types must be treated differently than other tube-based rectifiers.

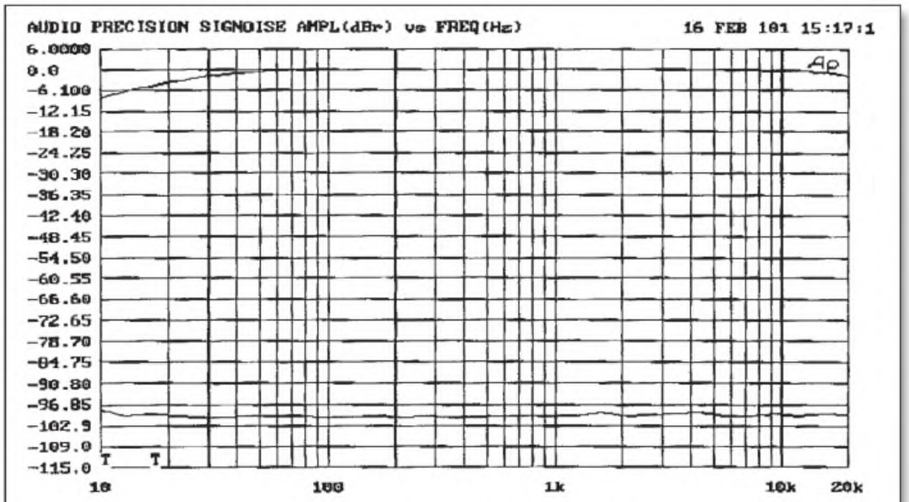

FIGURE 3: Signal-to-noise measurement.

If they have been turned upside down or the mercury inside has been spread all over the inside of the bulb (if you mail-ordered your tubes do not omit this step), you may wish to connect filament voltage (no plate voltage) for about 10-15 minutes and let the mercury vaporize inside the tube. Afterward, make sure the tube is supported in an upright position until you can insert it into the circuit. This precautionary step will minimize arcing and flashover. Also, the MV tubes must be "pre-heated" before they can be used in a circuit.

Typically, when the tube is upright, a small amount of mercury will collect at the bottom of the tube base. A "preheat" time is required to vaporize the mercury (raise the temperature of the condensed mercury) before the tube can start to conduct current. If the temperature inside your home is at least 70°, then a 45–60 second warm-up time should be sufficient.

I used 45-second Amperite "normally open" (NO) time-delay tubes in the power supply with no problems. Of course, these times are temperature dependent; if the tube is extremely cold a longer pre-heat time will be required to "vaporize" the mercury inside the bulb. Since the temperature inside most homes is fairly stable, it would be safe to assume 45–60 second pre-heat time to be adequate for most projects. If extreme temperatures are encountered, the RCA data sheet for an 866A has good information about operating temperatures.

The 866A's filaments require 2.5V/5A AC for operation. A 2.5V/10A center-tapped transformer is needed for the two 866As' rectifiers. This is a common transformer, so you should not have any trouble locating one for this project.

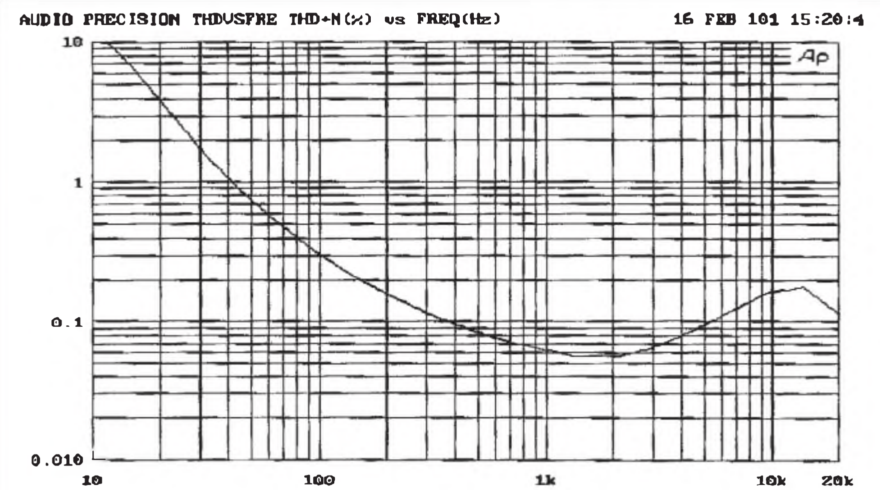

FIGURE 4: THD vs. frequency.

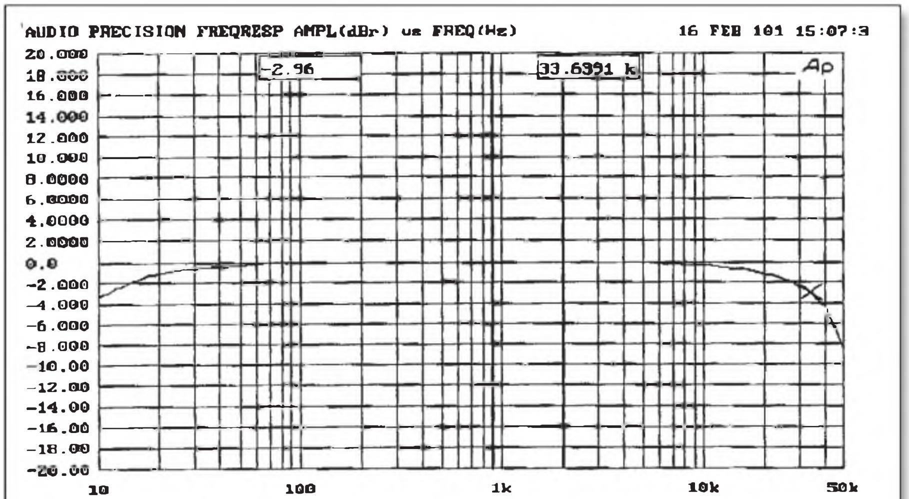

FIGURE 5: Frequency response.

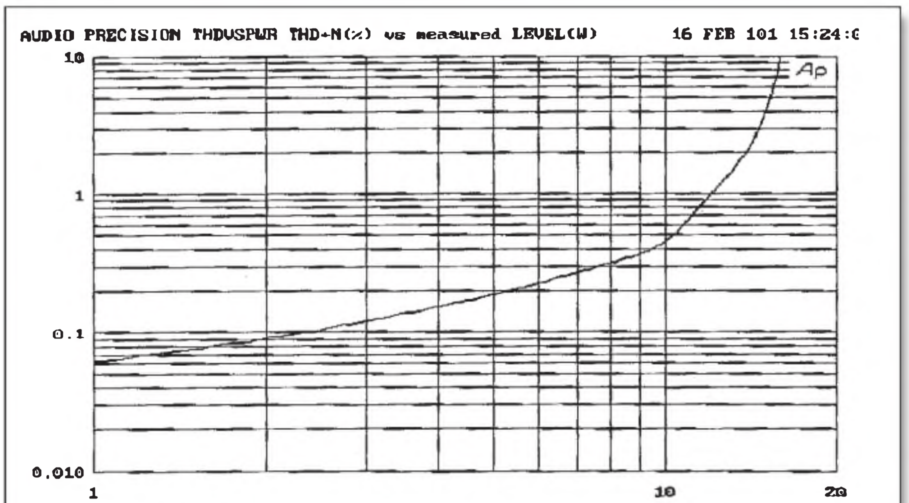

FIGURE 6: THD vs. power.

THE COMPLETE AMPLIFIER

After powering the amp up for the first time (*Photo 5*), I made measurements to verify operating points and such. (See the amp circuit in *Fig.* 2 and the parts list in *Table* 2.) I used the following test equipment:

Fluke 87 DMM
Old GE analog 2kV DC panel meter
HP 334A Distortion Analyzer
Instek 20MHz Dual trace O-scope
Sound Technology low distortion 1400A Oscillator
B&K 80MHz frequency counter

Table 3 shows measurements for the input/driver stage; *Table 4* for the complete amplifier; *Table 5* for the frequency response; and *Table 6* for test points. *Figures 3–6* confirm measurements for power, distortion, and frequency.

OUTPUT TRANSFORMER

Since the 211 has a high plate resistance, I knew that my selection of output transformers was going to be limited. Most, if not all, of the commercial transformers for 211-type amplifiers are about 10kΩ primary impedance. I plotted a 10kΩ load line for the 211, and it looked pretty good.

Next I plotted a 12kΩ load line, and it looked even better. I finally settled on a primary load impedance of 16k for the 211, which yielded very low distortion. As the primary load impedance went up, distortion came down, and power output decreased only slightly.

After crunching all the numbers I

asked myself "Where am I going to get a transformer with that high a primary impedance (not to mention good frequency response)?"

While looking on the internet for some NOS 211 tubes, I came across a website called Chimera Laboratories (ChimeraLabs.com). I called and asked about the selection of NOS tubes, and, in the course of our conversation, my homebrew 211 amp came up. Dennis Boyle (proprietor of Chimera) mentioned that he had custom transformers available for the 211 tube. He said most of the commercial ones had too low a primary impedance for the 211. Dennis said the optimum plate load for a 211 was approximately 17kΩ—not far off from what I had calculated.

I ordered a set of transformers from Dennis, and about eight weeks later I was in business (a long eight weeks).

Listening to these new transformers was amazing. I really had doubts about the high primary impedance, but they sounded wonderful. Next came time to collect more data from my test bench.

As you read earlier, the distortion and frequency response are respectable. The reduction in distortion, when both input and output stages are cascaded, is a result of second-order cancellation within the amplifier. It would seem that the 6SL7 and the 211 are a good match for each other, with the load lines chosen for each stage.

According to Dennis, the output transformers have the capability (with the proper input/driver circuit) of going up to around 70kHz. I might try a new input/driver circuit in the future, but for now I'm going to listen and enjoy some music for a while.

CONSTRUCTION

The main chassis for my amplifiers are modified 19″ rack-mount computer monitor enclosures (surplus enclosures with a 17″ × 17″ × 3″ footprint). If you are going to build mono amplifiers, I suggest you use this size chassis, which will allow you to position the power-supply section well away from the amplifier section so electromagnetic fields will not couple and interfere with each section (*Photo 6*). If you decide to build a stereo amplifier, I suggest building the power-supply and the amplifier sections on separate chassis.

I suggest this arrangement mainly for convenience. This amplifier gains weight very quickly, and you really won't care to haul it very far to your workbench. If you're building a stereo version, I still suggest at least a 17 × 17 × 3 chassis.

PART SOURCES

RS—Radio Shack

AES—Antique Electronic Supply, 6221 S. Maple Ave., Tempe, AZ 85283, 480-820-5411, website: www.tubesandmore.com

SP—Skycraft Parts and Surplus, 2245 W. Fairbanks Ave., Winter Park, FL 32789, 407-628-5634, website: www.skycraftsurplus.com

CL—Chimera Laboratories, Dennis Boyle, 1707 South Ervay, Dallas, TX 75215, 214-428-3901, Fax: 214-426-6605, e-mail: CL-Sales@ChimeraLabs.com or CLInfo@ChimeraLabs.com, website: www.chimeralabs.com/~CL/

HD—Home Depot or local hardware store

TS—thetubestore.com, 877-570-0979, website: www.thetubestore.com

TW—Tube World, 2712 Superior Ave., Sheboygan, WI 53081, 920-331-3126, website: www.tubeworld.com

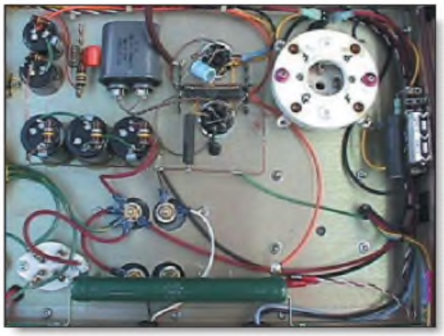

PHOTO 6: Main amplifier section and part of the power supply. 211 tube (top right corner). 6SL7 and 6F8G input section (top center). Bias resistor and bypass capacitor mounted to side of chassis (right center).

Build the power supply first. Start with the high-voltage section, which is the most critical part of the amplifier. The plate voltages specified on the schematic should not deviate more than ±20V. A lower voltage will limit your maximum output power. A higher voltage will cause some of the components to operate closer to their design limit.

After assembling the high-voltage section (T1, T2, V1, V2, C1, C2, C3, C4, L1), connect a 20k/200W load resistor with a sliding tap (centered on the resistor) from the output of the supply to ground. Do not connect the rest of the power-supply components at this time.

Apply power to T2 for only about one minute. After one minute connect T1. The mercury vapor tubes should glow a purple/blue (to see the glow you must be in low light conditions). The mercury vapor tubes will only glow if there is a load connected.

If you hear any “loud buzzes or pops,” disconnect power immediately. If you have a digital meter that can measure DC voltages higher than 1300V DC, then skip the following section and proceed to measure the output voltage.

For those of you with meters that go only to 1000V DC, proceed with the following. With the sliding tap on the power resistor centered on the resistor, measure the voltage from *ground to the tap only*, then from the *tap to the power-supply output*. Do not measure the full output voltage across your meter, or it may be the last voltage it measures.

Add these two voltages together for total output voltage, which should be between 1150 and 1250V DC. If it is within ±20V of 1150, you are O.K. If it is closer to 1250, then you will need to have to add a power resistor to the center-tap of T2.

Disconnect the center-tap of T2 from C3. Insert a 1250Ω/50W wire-wound resistor between the center-tap and C3. This resistor should bring your output voltage down closer to the 1150 range. A little patience and experimenting will be required here, so take your time.

When the output voltage is in the 1150 range, add the rest of the high-voltage components to the supply, making sure to discharge the capacitor through a suitable resistor before proceeding. With the main B+ voltage correct, everything else will fall into place. This is the most critical part of this project. The filament sections are fairly straightforward.

The main DPDT power switch (S1) performs two functions. It applies power directly to T1, while T2 is time-delayed for 45 seconds through TDR1. After 45 seconds TDR1 closes and sends power to T1, thus energizing the plates of the 866As. When S1 is opened it disconnects T1 and T2 immediately.

The amplifier section is very simple, but effective. The only adjustment is in the 211 filament circuit. Measure from ground to each side of the 211 filament with your meter. Adjust R9 until the voltage on the filament is nearly equal (+5V, 0, −5V). The bias on the 211 cathode should be around 60V DC. Since there is a 1k resistor in the cathode circuit, whatever voltage you measure across the resistor will also be the current through the resistor and through the output tube.

As an option, you can install a 120V AC fan on top of or underneath the chassis to provide some means of ventilation (a small box fan similar to one in a computer).

The 1k/50W 211 bias resistor and the 30k/100W voltage divider resistor become fairly warm. Providing ventilation will enhance long-term reliability. Do not connect a 12V DC fan to the 211 filament circuit; this could inject noise into it. These small DC fans have switching power supplies inside them, which generate all kinds of trash that must be filtered out. ❖

Inductor Coupling

This article measures signal coupling (crosstalk) for various orientations and spacings of typical air, ferrite, and iron-core crossover inductors, and discusses the effects of such coupling. **By Dennis Colin**

It is standard practice to orient crossover coils at 90° and space them as far apart as possible, but some questions arise, such as the following:

1. How much spacing is enough?

2. What are the effects of positional and angular offsets?

3. Does 90° orientation ensure minimum coupling?

PRINCIPLES OF MAGNETIC COUPLING

If magnetic flux from one coil (L1) is intercepted by that of another coil (L2) in such a way that the former has a net component aligned with the magnetic axis of L2 (usually its centerline), then a voltage will be induced in L2 proportional to the rate of change of this flux (and of the current in L1 producing it). A constant AC voltage across L1 produces a current (and flux) inversely proportional to frequency (if DC resistance is negligible). This results in the induced voltage in L2 being constant with frequency, if L2 is not loaded (open-circuit induced voltage).

Figure 1 is a graphic representation of the magnetic field of an air coil of typical geometry. The "lines of force" are actually cross-sections of symbolic three-dimensional "doughnuts" symmetrically surrounding the coil—symbolic, of course, because the actual field is continuous in 3-D space, at least down to the quantum scale.

Of interest is the orientation of another ("secondary") coil to the energized one ("primary"), shown radiating flux. Note that the coils labeled "A" are oriented for maximum coupling; the coil's magnetic axis is aligned with the field "lines." Conversely, the coils labeled "B" have (ideally) zero coupling; their axes are perpendicular to the field lines.

Also note that for coils not in the same plane as the primary (only B4 is so in *Fig. 1*), 90° orientation of coil axes does *not* result in nulling of coupling (coil A1, for example). And coil B1 is parallel to the primary, yet has (ideally) zero coupling.

ABOUT THE AUTHOR

Dennis P. Colin graduated with a BSEE from the University of Lowell (MA) and is currently an Analog Circuit Design Consultant for microwave radios. Previously a band keyboardist and recording engineer, he has been published in the *Journal* of the Audio Engineering Society. He demonstrated the audibility of phase distortion at Boston Audio Society and also designed the "Omni-Focus" speaker bi-polar coincidental with phase-linear first-order crossover, the ARP 2600 analog music synthesizer, the 1kW bi-amp and PWM supply at A/D/S, and Class D amps.

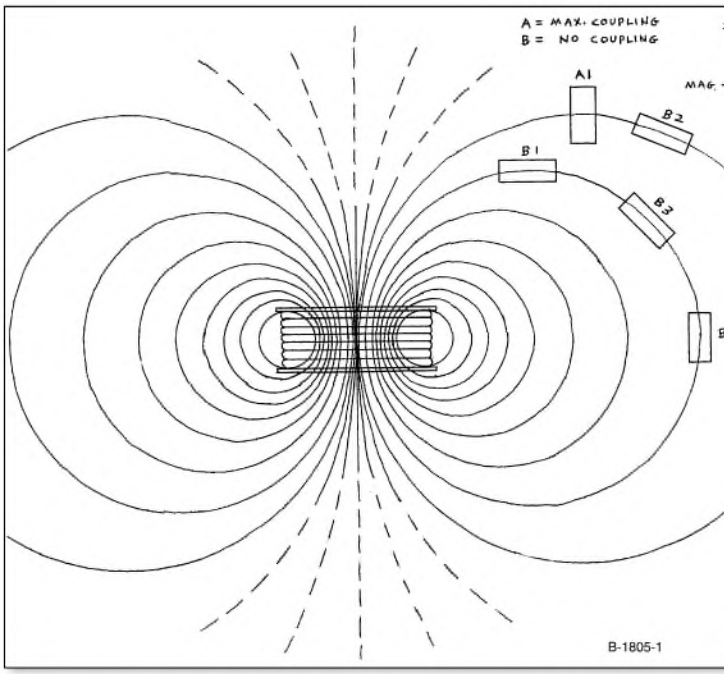

FIGURE 1: Magnetic field of air coil—side view.

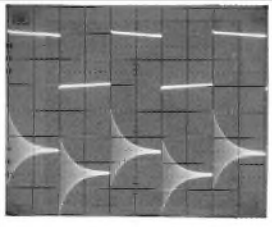

PHOTO 1: R_L = open. Low damping of 312kHz resonance due to 153pF C (103pF coil + 50pF cable).

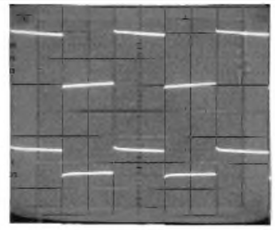

PHOTO 2: $R_L = 1.8K\Omega$. BW –3dB ≈ 200Hz – 300kHz. LF

$$F\ -3dB = \frac{2.25\Omega}{2\pi L}$$

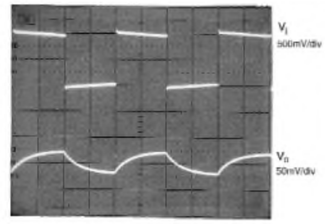

PHOTO 3: $R_L = 50\Omega$. Overdamped. HF

$$F\ -3dB = 4.7kHz = \frac{R_L}{2\pi L}$$

Most of the following coupling measurements use worst-case parallel-aligned orientation of two identical coils, to show the maximum coupling likely for a given coil-pair spacing.

COUPLING COEFFICIENT

Coupling coefficient "C" is defined here as the ratio of the open-circuit voltage (V_0) induced in coil L2, to the voltage (V_1) across the driving, or primary, coil L1 (*Fig. 2*).

A more fundamental definition of C is the fraction of total flux emitted by L1 not only intercepted by, but "utilized" by L2; you could call this the "mutual magnetic alignment factor." By this definition, C will equal $\frac{V_0}{V_1}$ except for loading losses.

Coupling in dB = 20 log C; conversely, isolation in dB is simply its negative, –20 log C. (Isolation, a positive dB value, is used for convenience in *Fig. 3*.)

A very efficient bifilar-wound iron-core transformer can have more than 95% voltage coupling; C > 0.95; isolation, or loss, is less than 0.5dB. But of course, in crossover filters, the lowest possible coupling is desirable. (Coupled inductors can be a useful part of filter design, but they can involve very complicated analysis.)

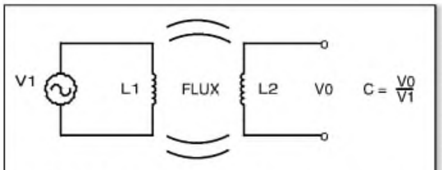

FIGURE 2: Definition of C. B-1805-2

THE EFFECT OF COUPLING IN CROSSOVERS

For the typical low values of coupling, the induced signal V_0 appears as a voltage source in series with the inductance of L2 (*Fig. 4*). The load this sees (other crossover components and driver loading) generally attenuates V_0. Only with a high-Q capacitive load can V_{LOAD} exceed V_0, but this is unlikely; high-Q reso-

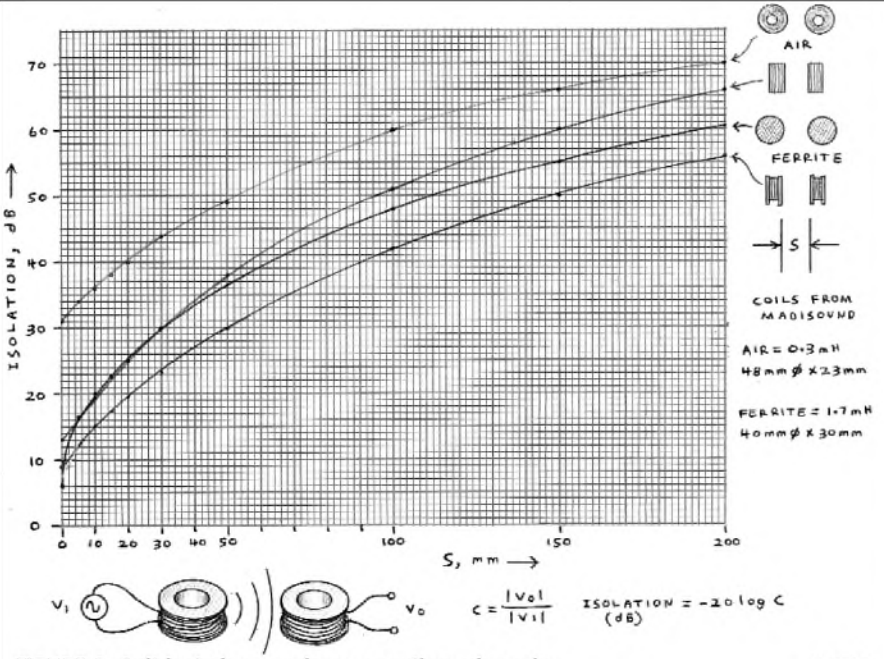

FIGURE 3: Coil isolation, maximum coupling orientation.

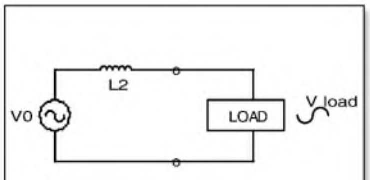

FIGURE 4: Loading of coupled output.

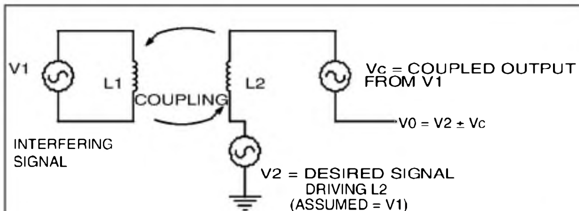

FIGURE 5: Induced error.

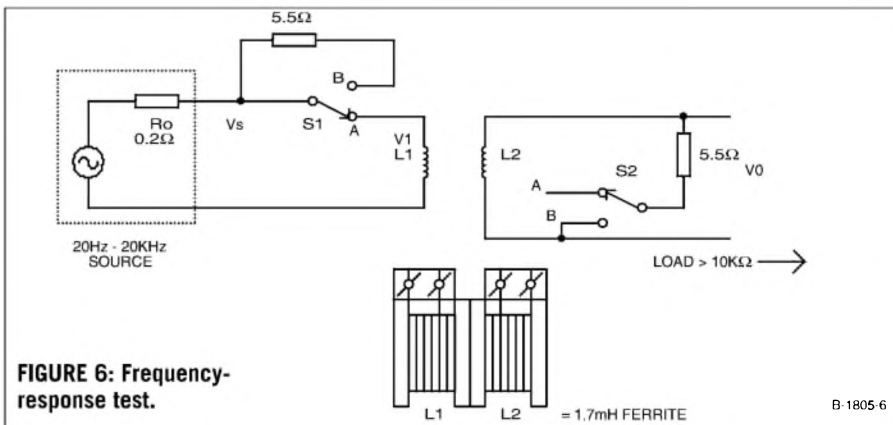

FIGURE 6: Frequency-response test.

nances are most undesirable, of course. So generally, the unloaded induced voltage V_0 is a conservative measure of maximum undesired crosstalk signal.

SIMPLIFYING ASSUMPTIONS

For the sake of simplicity, I made the following assumptions:

1. Coil parasitics (R, C, and core losses) are ignored.
2. The desired signal intentionally driving L2 is assumed to be equal to that feeding L1; typically they will be of similar levels.

3. The two coils are assumed to be identical.

DEFINITION OF INDUCED ERROR (E_{P-P})

If the two coupled coils are part of the same frequency channel, the coupled signal will probably cause frequency response errors. Induced error (e_{p-p}) is defined as the pk-pk dB variation in signal level that's possible with the coupled signal being either in phase or inverted with respect to the desired signal (*Fig. 5*).

Coupling coefficient

$$C = \frac{V_C}{V_1} = \frac{V_C}{V_2},\ \text{so } V_C = CV_2.$$

Output $V_0 = V_2 \pm V_C = V_2 \pm CV_2 = V_2 (1 \pm C)$,

so $1 \pm C = \frac{V_0}{V_2}$, the ratio of actual to desired signal.

In dB, this is 20 log (1 ± C); worse case variation in dB is defined as $e_{p-p} = 20 \log (1 + C) - 20 \log (1 - C)$.

The definition applies only to crosstalk within a same-frequency channel, such as a midrange crossover section. For inter-channel crosstalk (say, a woofer coil "talking" to a tweeter coil), the coupling coefficient C is the only relevant measure, indicating, for example, how much woofer signal could reach the tweeter in addition to (or subtraction from, depending on phase) the normal crossover overlap.

FREQUENCY RESPONSE TESTS

I used the setup of *Fig. 6* to verify that even loosely coupled coils are capable of flat, wideband response. I mounted a pair of 1.7mH ferrite-bobbin "Sledgehammer" inductors face-to-face; while this may not seem like "loose" coupling, the measured C was only about 0.35, or −9dB. In *Fig. 7*, trace 1 is the input V_S (not a perfect voltage source, but its output impedance was low enough to provide a constant voltage to L1 above about 50Hz).

Note that trace 2 is similarly flat; this is the output V_0 with low source impedance (S1 in position A) and high load impedance (S2 in position A). Trace 3 is with 5.5Ω (plus source 0.2Ω) driving impedance; the LF rolloff of approximately 500Hz corresponds to the high-pass filter formed by 5.7Ω and 1.7mH. Of interest is that much below the rolloff frequency, L1 is driven by a constant current, which means constant magnetic flux. Since V_0 is proportional to the flux's rate of change, V_0 increases at 6dB/octave with frequency.

Conversely, applying a 5.5Ω load to the secondary (with a low source-impedance primary drive) results in trace 4, a HF rolloff, also at approximately 500Hz. A tightly coupled transformer (low leakage inductance) wouldn't do this, but with loose coupling, the leakage inductance is nearly the value of L2, 1.7mH. This then appears in series with the coupled output voltage, forming a low-pass filter with the 5.5Ω load.

Finally, trace 5 shows the bandpass response when both source and load impedances are about 5.5Ω. This corresponds to typical crossover use, where all inductors are moderately loaded.

This test establishes that (1) the frequency of maximum coupling depends on coil-loading impedance, and (2) if actual coil voltages V_1 and V_2 are measured with no output loading, response is very flat, so you can measure coupling C at any frequency well within the passband determined by coil and test-equipment parasitics.

PULSE RESPONSE TESTS

Figure 8 and *Photos 1-4* show 5kHz square wave responses with various source and load impedances on a coupled pair of 1.7mH ferrite coils spaced 40mm apart. The purpose of these tests is to show the previous loading effects in the time domain. *Photo 2* also demonstrates the good square-wave response capability of a transformer if it's properly driven and loaded, a consideration of which vacuum-tube power-amp designers are well aware.

TEST SETUP

Figure 9 shows the test setup. I made measurements on two types of coil pairs, both from Madisound:

1. 0.3mH "Sidewinder" air coils, 48mm OD × 23mm.
2. 1.7mH "Sledgehammer" ferrite-bobbin coils, 40mm OD × 30mm.
3. 5mH "Sledgehammer" steel-laminate coils (*Fig. 10*).

The voltage V_1 applied to L1 was held constant. I used 26kHz with the 0.3mH

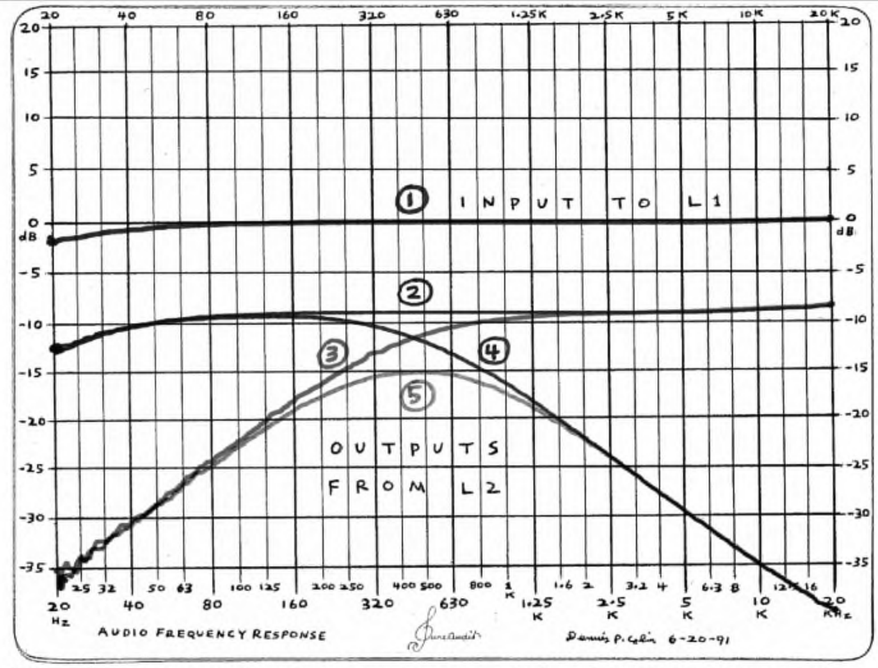

FIGURE 7: Coupled-coil frequency responses, face-to-face 1.7mH "Sledgehammer" ferrite inductors.

B-1805-7

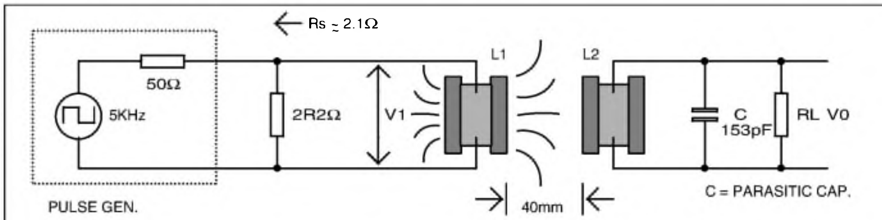

FIGURE 8: Pulse-response test. L1, L2 = 1.7mH ferrite (Madisound). R_{DC} = 0.15Ω.

B-1805-8

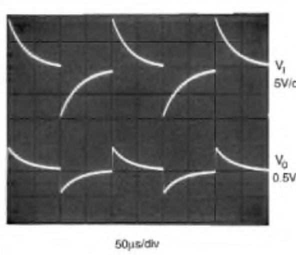

PHOTO 4: Same as *Photo 2* except R_S = 50Ω. LF

$$F\ -3dB = \frac{50\Omega}{2\pi(1.7mH)} = 4.7kHz$$

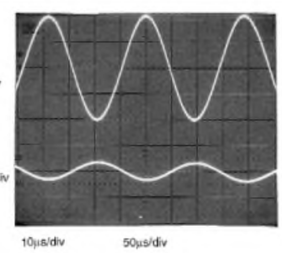

PHOTO 5: L1, L2 = 0.3mH, horizontal. S = 30mm, F = 26kHz. Top = 5V/div, bottom = 200mV/div.

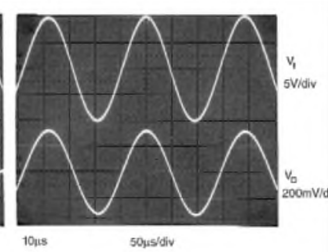

PHOTO 6: Same as *Photo 5* except coils vertical.

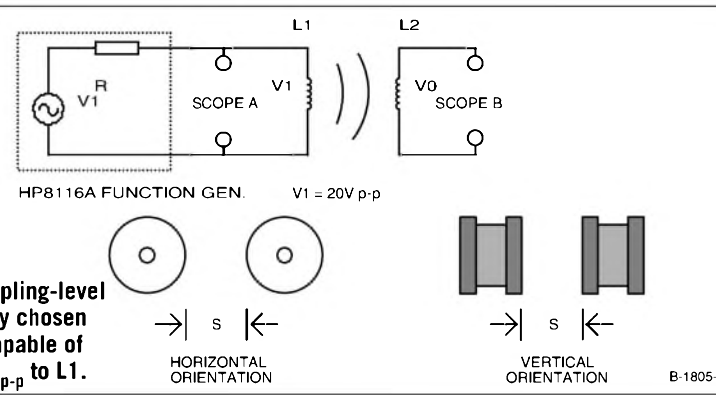

FIGURE 9: Coupling-level test. Frequency chosen is minimum capable of supplying $20V_{p-p}$ to L1.

B-1805-9

TABLE 1
AIR COIL COUPLING

HORIZONTAL

S MM	D MM	$\frac{d}{D}$	V_0	$C = \frac{V_0}{V_1}$	C DB	$E_{p\text{-}p}$ DB
0	48	1.09	580mV	0.029	–31	0.50
5	53	1.20	420mV	0.021	–34	0.36
10	58	1.32	320mV	0.016	–36	0.28
15	63	1.43	250mV	0.0125	–38	0.22
20	68	1.55	200mV	0.010	–40	0.17
30	78	1.77	132mV	0.0066	–44	0.11
50	98	2.23	68mV	0.0034	–49	0.058
100	148	3.36	21mV	0.0011	–60	0.018
150	198	4.50	9.5mV	0.0005	–66	0.008
200	248	5.64	6mV	0.0003	–70	0.005
250	298	6.77	4mV	0.0002	–74	0.003

VERTICAL

S MM	D MM	$\frac{d}{D}$	V_0	$C = \frac{V_0}{V_1}$	C DB	$E_{p\text{-}p}$ DB
0	23	0.52	4.4V	0.22	–13.1	3.9
5	28	0.64	3.1V	0.155	–16.2	2.7
10	33	0.75	2.15V	0.108	–19.4	1.9
15	38	0.86	1.5V	0.075	–22.5	1.3
20	43	0.98	1.1V	0.055	–25	1.0
30	53	1.20	640mV	0.032	–30	0.56
50	73	1.66	250mV	0.0125	–38	0.22
100	123	2.80	57mV	0.0029	–51	0.049
150	173	3.93	21mV	0.0011	–60	0.018
200	223	5.07	10mV	0.0005	–66	0.009
250	273	6.20	5mV	0.0003	–72	0.004

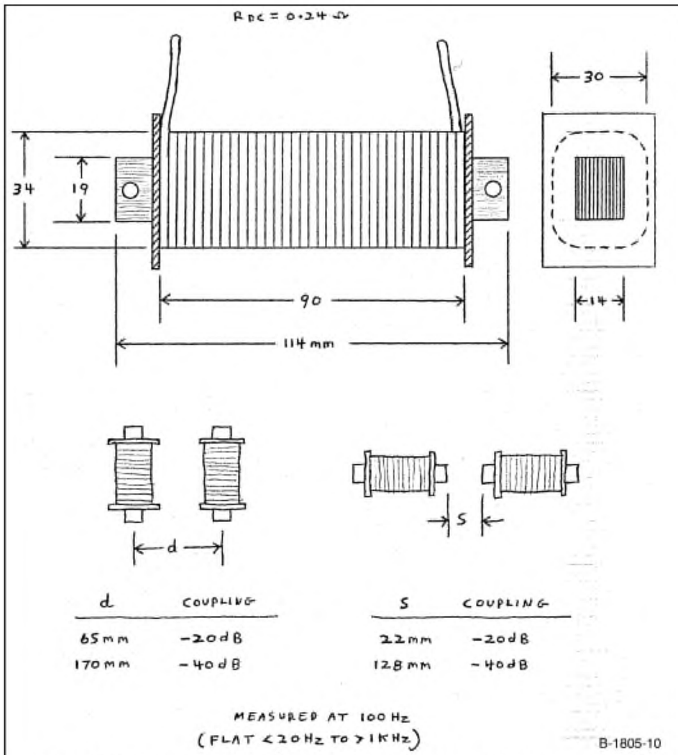

FIGURE 10: 5.0mH "Sledgehammer" steel-laminate inductor.

air coils, 4kHz with the 1.7mH ferrites, and 100Hz with the 5mH coils. I chose these frequencies simply for measurement convenience.

COUPLING RESULTS

Table 1 shows the results for the 0.3mH air coils with maximum-coupling parallel orientations, both horizontal and vertical (*Photos 5* and *6*). Some observations:

1. Vertical orientation produced more coupling than horizontal for a given spacing, particularly when close in relation to coil diameter.

2. With horizontal orientation, a spacing of only 20mm was needed to produce 40dB isolation and an induced error $e_{p\text{-}p}$ of only 0.17dB—less than a ±0.1dB error.

3. With vertical orientation, the corresponding spacing needed was about 56mm.

4. For spacings greater than one coil diameter, coupling (C) decreases approximately as the inverse cube of d, the center-to-center coil separation.

Table 2 shows the results for parallel orientations of the 1.7mH ferrite coils. Note that while the air coils needed only 20mm (horizontal) and 56mm (vertical) spacing for 40dB isolation, the ferrite coils needed 63mm and 91mm, respectively. This is, of course, due to the magnetic-field concentration of the ferrite cores.

Figure 3 is a graph of isolation in dB (–20 log C) versus spacing distance (S) for horizontal and vertical orientations of both the air and ferrite coils. Note that about 40dB isolation results in an induced error ($e_{p\text{-}p}$) of 0.2dB, or ±0.1dB. Isolation (plotted vertically) is simply the negative of coupling in dB.

Figure 10 shows a 5.0mH "Sledgehammer" steel-laminate inductor from Madisound, and the (worst-case) parallel-aligned orientation spacings required for couplings of –20dB and –40dB.

TABLE 2
FERRITE COIL COUPLING

	HORIZONTAL		VERTICAL			
S MM	V_0	C DB	$E_{p\text{-}p}$ DB	V_0	C DB	$E_{p\text{-}p}$ DB
0	10V	–6.0	9.5	7.1V	–9.0	6.4
5	3.0V	–16.5	2.6	4.9V	–12.2	4.3
10	2.05V	–19.8	1.8	3.5V	–15.1	3.1
15	1.45V	–22.8	1.3	2.7V	–17.4	2.4
20	1.1V	–25	0.96	2.1V	–19.6	1.8
30	650mV	–30	0.56	1.35V	–23.4	1.2
50	300mV	–36.5	0.26	660mV	–30	0.57
100	80mV	–48	0.069	165mV	–42	0.14
150	36mV	–55	0.031	64mV	–50	0.056
200	19mV	–60.5	0.016	32mV	–56	0.027
250	12.5mV	–64	0.011	18mV	–61	0.016
300	9mV	–67	0.008	12mV	–64	0.010

TABLE 3
NORMALIZED COMPARISON

COIL TYPE	D_H	D_V	MAX. IND. DIM.	A_H	A_V
0.3mH air	64mm	77mm	44mm	1.45	1.75
1.7mH ferrite bobbin	103mm	121mm	40mm	2.58	3.03
5.0mH steel laminate	170mm	242mm	114mm	1.49	2.12

NORMALIZED SPACING-TO-SIZE COMPARISON

In *Table 3*, coil center-to-center distance (d) is used as the common measure; d_H is for "horizontal" orientation (magnetic axes parallel), and d_V is for "vertical" (magnetic axes aligned); both d_H and d_V are the separations required for 40dB unloaded isolation. "Maximum Inductor Dimension" is that of the windings in the air coils, but of the cores in the other two types. The terms a_H and a_V are forms of normalized space-to-size ratios:

$$a_{H,v} = \frac{d_{H,v}}{\text{Maximum Inductor Dimension}}$$

You can regard a_H and a_V as the magnetic "reach" of a coil—at what distance, in terms of its own size units, can the coil couple −40dB to another identical coil?

As expected, the magnetic-core types extend flux farther (relative to size) than the air coils. But why does the ferrite bobbin reach farther than the steel laminate, even though the latter likely has higher permeability? Well, the steel core is long and thin, while the ferrite is nearly square in aspect ratio. So scaled to the same maximum dimension, the ferrite bobbin appears thicker—more core volume per normalized space.

But regardless of theory, the quantities a_H and a_V are simply a practical measure of the −40dB coupling extension of an inductor, relative to its own maximum size dimension.

To explore magnetic interactions further, it would be very useful to actually see the field distribution. You can accomplish this with either $50,000 worth of software, or a bunch of iron filings and a battery. (Guess which method Joseph Henry and other pioneers used?)

OFFSET AND NONPARALLEL ORIENTATIONS

Figures 11–21 show coupling in dB for a variety of configurations, some with optimum coupling/nulling symmetry. *Figure 16* is not in the latter category, but is a side view of two coils mounted flush on the same plane, such as a PC-board surface. Even though the coils' axes are situated at 90° to one another, the coils are not magnetically symmetrical, so coupling isn't nulled out. *Figure 22* and the next section describe why.

MISCELLANEOUS CONSIDERATIONS

Perpendicularity of the coil axes does not necessarily cancel coupling,

(to page 50)

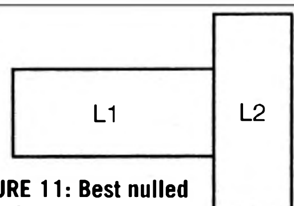

FIGURE 11: Best nulled coupling equals −86dB. B-1805-11

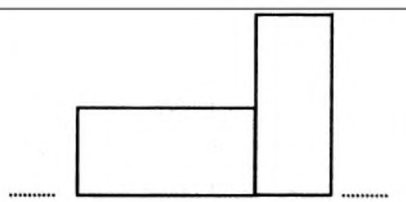

FIGURE 12: −26.5dB. B-1805-12

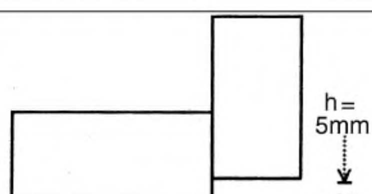

FIGURE 13: −25dB, maximum @ h = 5mm. B-1805-13

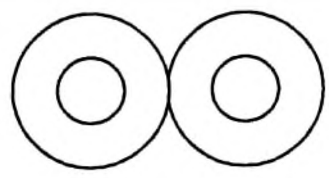

FIGURE 14: −30.5dB. B-1805-14

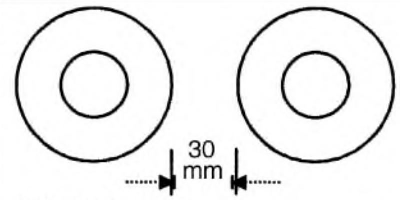

FIGURE 15: −44dB. B-1805-15

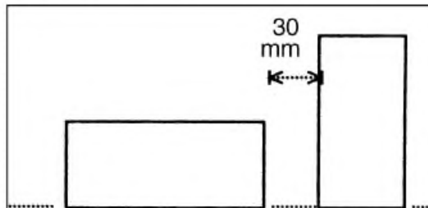

FIGURE 16: −45dB. B-1805-16

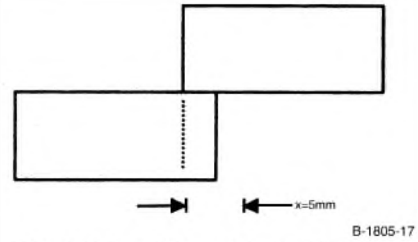

FIGURE 17: –54dB, minimum @ x = 5mm.

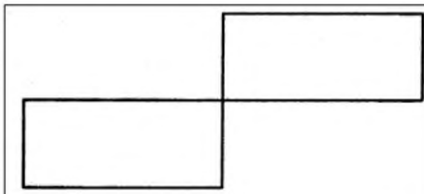

FIGURE 18: –45dB. B-1805-18

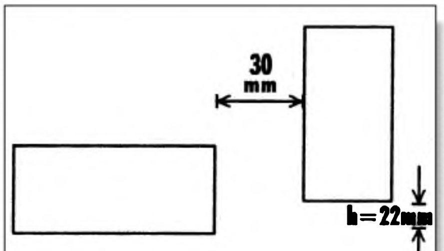

FIGURE 19: –41dB, maximum @ h = 22mm. B-1805-19

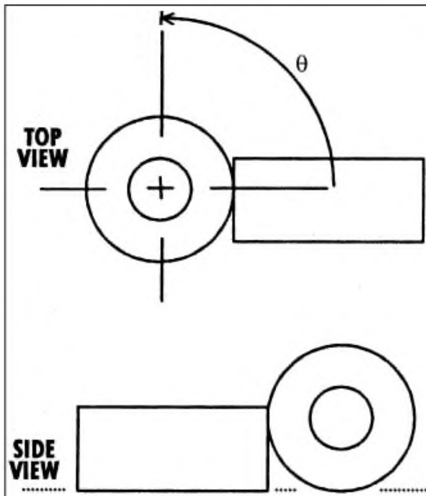

FIGURE 20: –80dB, θ = 90°. B-1805-20

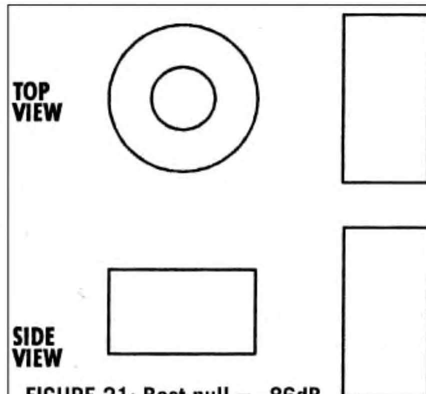

FIGURE 21: Best null = –86dB.

(from page 47)

for example, horizontal and vertical coils on a circuit board (*Figs. 12*, *16*, and *22*). The 0.3mH air coils positioned 30mm apart (*Fig. 16*) with –45dB coupling produce results only 1dB less than with both coils horizontal with the same separation (*Fig. 15*). With the coils touching and oriented as in *Fig. 12*, coupling was –26dB. But raising L1 up until centered with L2's axis, and carefully adjusting for the null, resulted in a coupling of only –86dB (*Fig. 11*).

In *Fig. 11*, the axis of one coil (L2) is centered within the plane of the other (L1), and intersects L1's axis at 90°. While these two conditions (ideally) produce complete cancellation of coupling, this is not the greatest degree of symmetry possible. This configuration, with one degree of symmetry, is very sensitive to positional and angular offset. A 3° departure from perpendicularity of the true magnetic axes (not so symmetrical in real-world coils) resulted in 20dB more coupling than the best null. (To keep this in perspective, however, even with 3° error, and with coils touching, the coupling was only –60dB.)

But for the lowest coupling, you should use the configuration of *Fig. 23*: it has 2 degrees of symmetry, the greatest possible with cylindrical coils (solenoids). Here, each coil's axis is both

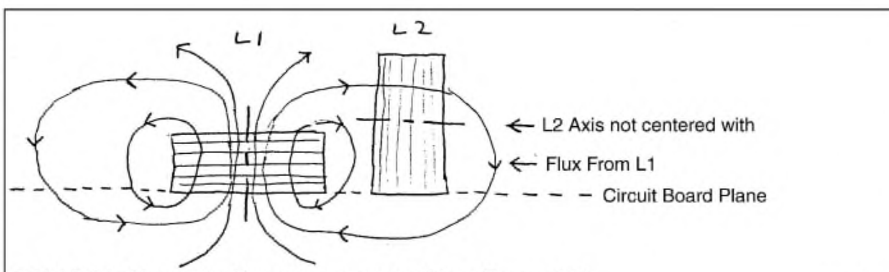

FIGURE 22: Magnetic dissymmetry even with coil axes 90°. B-1805-22

perpendicular to that of the other, and centered in the plane of the other. Each coil can be offset along its axis, or rotated in the plane of the other coil, without (ideally) causing any coupling.

THREE OR MORE COILS

One easily implemented configuration for three coils is a straight line, with three mutually perpendicular axis orientations and the horizontal coil elevated so its central plane is at the same height as the vertical coils' axes (*Fig. 24*).

For four or more coils, this alternating pattern can be continued. Then like-oriented coils (every third one) will be spaced far enough apart so coupling will be very low. But with four or more coils, not all of them can be perpendicular, so the best strategy is simply separation, using all available board space, and orienting the closest coil pairs at 90°.

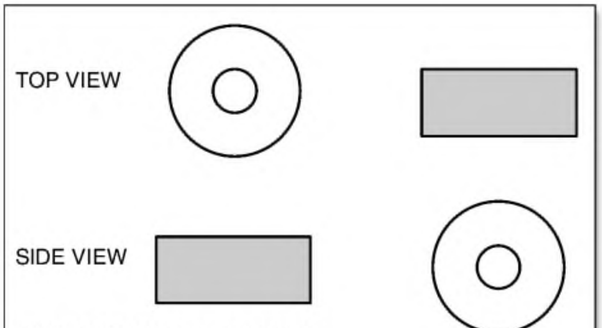

FIGURE 23: Most symmetrical arrangement of two coils.

With the air coils studied here, a horizontally oriented (parallel magnetic axis) pair has 60dB worst-case (unloaded) isolation when spaced apart at two coil diameters. The same-channel induced-response error is less than ±0.01dB, and interchannel crosstalk with normal loading is probably less than −70dB.

SCALING AND EXTRAPOLATION

While only three types of inductors were measured here, the following considerations apply to other inductors:

1. For a coil of the same type and winding proportions as one of those studied here, simple scaling applies; that is, a given ratio of spacing to, say, coil diameter produces the same coupling regardless of size.

2. For different geometries within a given type (air, ferrite bobbin, and so on), using the ratio of spacing to the largest coil dimension, be it length, diameter, or both if equal, will give at least "ballpark-similar" coupling results.

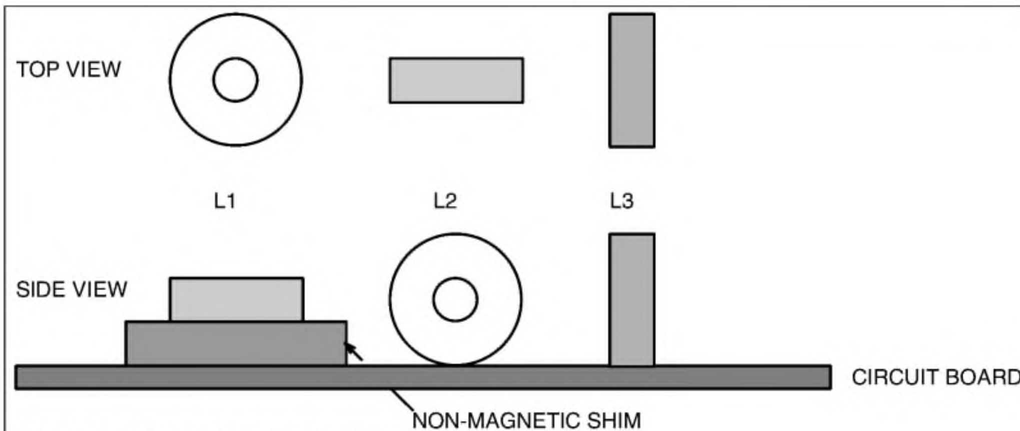

FIGURE 24: Good orientation for three coils.

I hope this article has presented a sufficient quantity and clarity of data and relevant theory to be of good use to those laying out crossover boards or any circuit with multiple inductors. ❖

Exploring the DAC Chip World

This author surveys DAC chips to determine a simple solution to improve performance. **By Andrea Ciuffoli**

My preferred hobby is to design and build audio amplifiers and loudspeakers, and every time I start a new project, my aim is to enhance performance over the previous projects. But it is difficult to increase an amplifier's sonic quality if the source is not good. The source is important in the final result, and there should be no compromise in a high-end system.

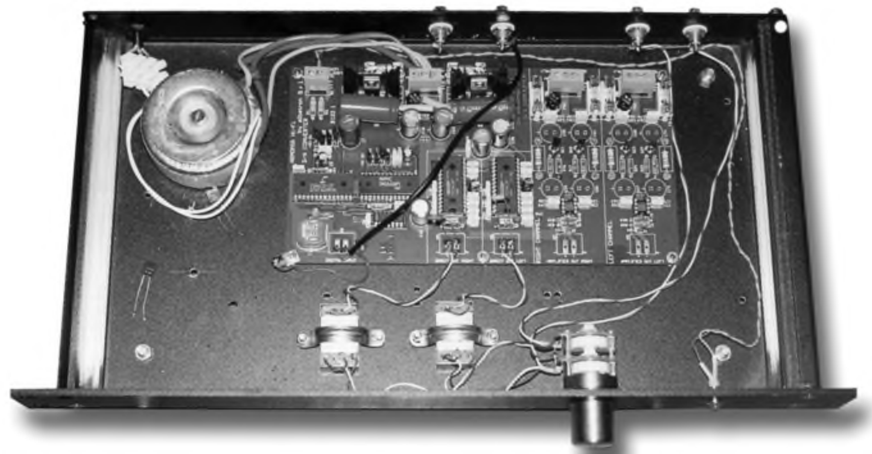

PHOTO 1: Layout of the DAC with Armonia hi-fi board.

IMPROVING DAC PERFORMANCE

The DACs normally included in commercial CD players (even those of significant price) have rather low performance because the common use of high feedback operational amplifiers "destroys" the sound through compression effects and poor soundstage stability.

I found a good solution to this problem in Stefano Perugini's design concept ("A 24-Bit DAC," *GA* 6/98, p. 1). In that article, he proposed to use a transformer as the DAC's output stage to clear the signal without active components. I have undertaken many tests with this solution, initially using the Perugini transformers, and later, the Lundahl models.

My idea was to explore the entire world of DAC chips to find the best. Building a DAC board is not easy, because you must observe the rules about ground and flow, and a poorly designed board can compromise the final sonic result. Looking at the Crystal Internet website, I found evaluation board CDB4390, which is a complete DAC with optical and coax digital inputs and analog outputs, so you can test the transformer output stage by changing only five simple connections.

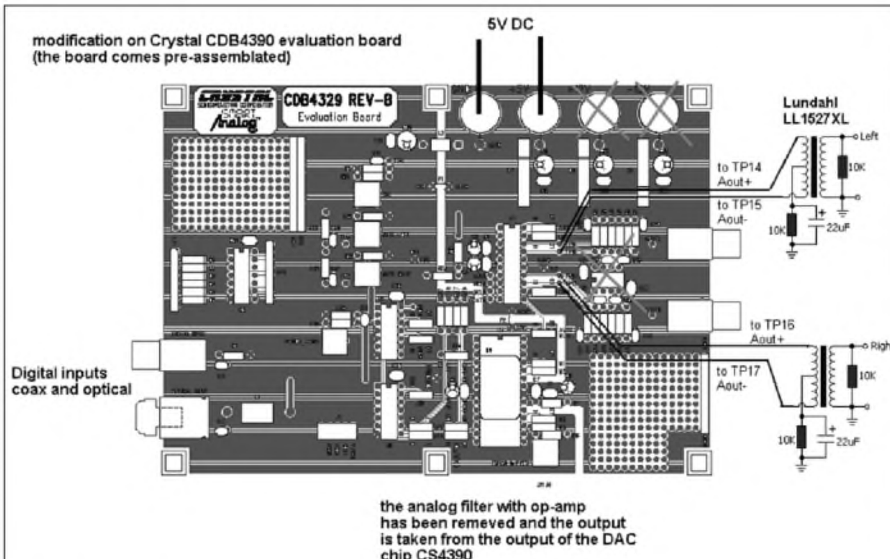

FIGURE 1: Modification of the Crystal evaluation board.

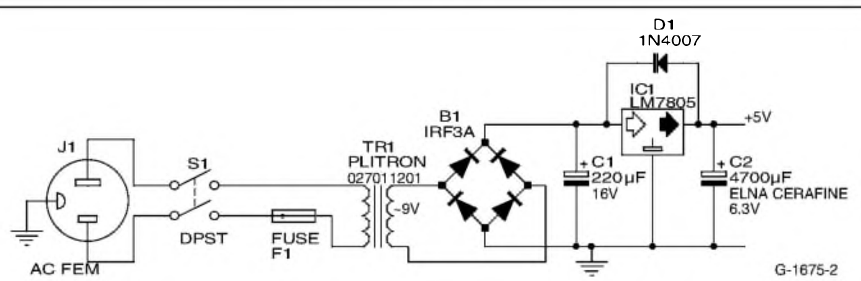

FIGURE 2: Power supply of the Crystal evaluation board.

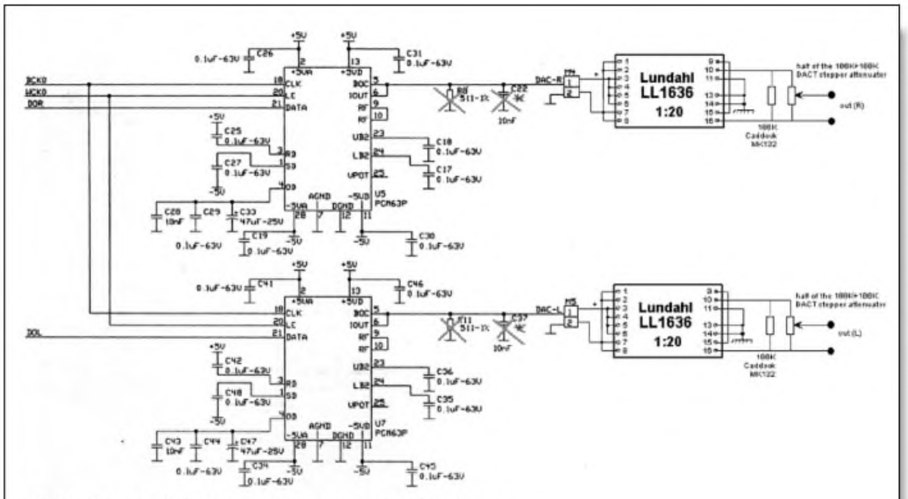

FIGURE 3: Modification of the Armonia hi-fi board.

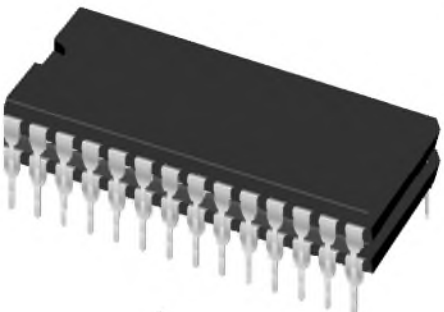
PHOTO 2: Design of the piggyback configuration.

The result is impressive, especially with the Lundahl LL1527XL transformers connected as shown in *Fig. 1*, and if compared to my CDT Musical Fidelity. *Figure 2* shows the power supply to use with this configuration. (Just a little note about the CDT Musical Fidelity: a good improvement is possible if you skip the output-tube stage and get the analog signal from the DAC board.)

SEEKING EVALUATION BOARDS

To test the Burr-Brown chips, I searched for an evaluation board on their website and found the unique and incredible EMV1702. This uses four DAC chips per channel and optoisolators for the digital/analog section, but it is very expensive.

After many months of further searching, I found a board produced also as a kit by the Italian company Armonia Hi-Fi. It has a very low price and uses the Burr-Brown PCM63, considered by many the best DAC chip.

The PCM63, as do all top-line Burr-Brown chips, has an output current instead of a voltage, so you need an I/V converter connected to the output of the chip. Obviously, the simplest I/V converter is a resistor, and I have found that the best soundstage uses the range of 100–125Ω (use only first-class resistors, such as the Caddock MK132).

With this low-value resistor, the output signal voltage is very low, so to increase it you need to use a transformer with a high turn ratio of about 1:20. For this I have used the Lundahl LL1636, with an amorphous core that doesn't store energy (unlike conventional Mumetal cores) and gives better bass frequency. To use the LL1636, skip all resistors and capacitors on the DAC-chip output (R8, R11, C22, and C37) and connect it directly to the primary of the transformer in the 1:20 configuration. On output, connect a 100k MK132 Caddock resistor and a stereo 100k+100k CT1 DACT stepper attenuator (*Fig. 3* and *Photo 1*).

(Instead of the normal brands, all my new designs use the CT1 or the CT2 DACT stepper attenuators, since they have higher sonic detail and lower coloration.)

MODIFICATIONS

The last chips I tested were Analog Device's very good AD1860s. In this case I didn't find an evaluation board or a kit, so I sought an inexpensive product to modify. I chose the D18 model produced by Monarchy Audio, which is a very interesting company that also makes many Class A amplifiers without feedback. The D18 produced very clear analog signal with the normal output stage.

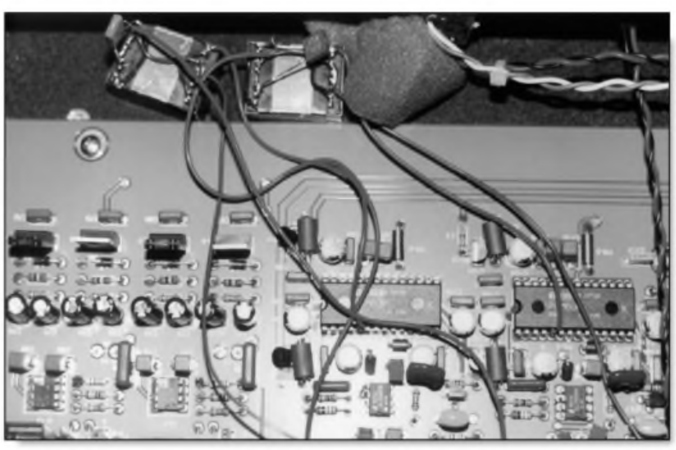
PHOTO 3: Layout of the DAC with the Monarchy Audio.

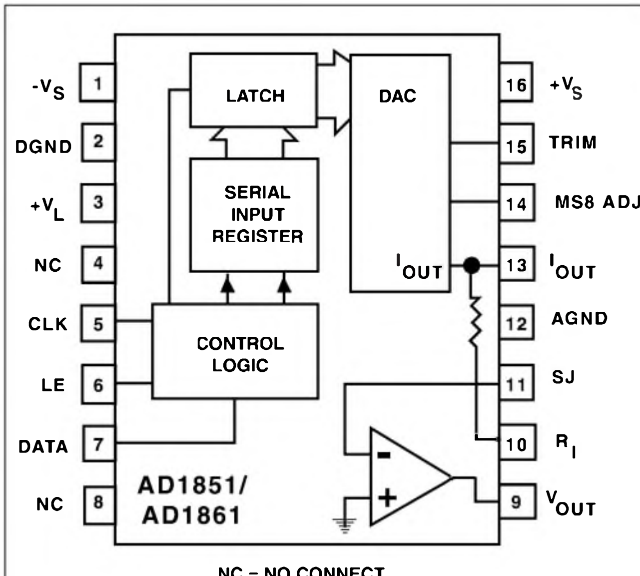

FIGURE 4: Pins of the AD1860/AD1861 chips.

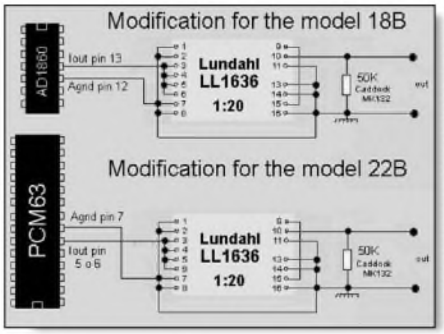

FIGURE 5: Model 18B (top) and Model 22B modifications.

The AD1860 DAC chips are used with the new piggyback technology that doubles up the D/A converter chips for even more analog sound and higher output drive (*Photo 2*). With this DAC I have also used the Lundahl LL1638 with the turn ratio of 1:20, because two AD1860s in the piggyback configuration give the same output current as one PCM63 (*Photo 3* and *Figs. 4* and *5*).

I have also tested the Monarchy Audio model 22B using the PCM63K Burr-Brown chips, a selected version of the PCM63. The sonic results of these four DACs, modified as described, are very impressive, and the choice is difficult. ❖

WEBSITE REFERENCES

My website:
http://www.audiodesignguide.com

Stefano Perugini's website:
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http://www.crystal.com/

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http://www.areacom.it/biz/armonia/

Lundahl transformers:
http://www.lundahl.se/

Monarchy Audio:
http://www.monarchyaudio.com/

DACT stepper attenuators:
http://www.DACT.com

Ribbons Made Easy

This article describes the construction of a ribbon speaker for use at mid/high frequencies. Design goals were high acoustical quality, ease of construction, and adjustablity. **By Justus V. Verhagen**

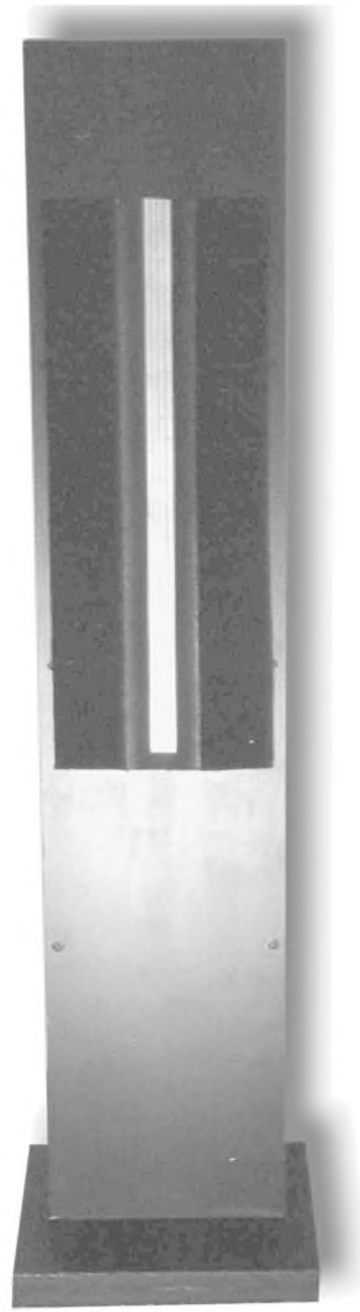

PHOTO 1: Front view of ribbon speaker, finished with flat black paint.

Ever since Ole Thofte[1] presented his wonderful ribbon design in *Speaker Builder*, I have been addicted to building and listening to ribbon or ribbon-like speakers. I have owned Apogees for many years and love their detailed imaging and sense of spaciousness.

There are a few limitations to Thofte's design that I thought would be helpful to overcome. First of all, his design doesn't allow for easy and accurate adjustability of the size of the magnet gap. Second, it lacks attention to detail in the design that would allow for optimal performance. In my current design, I have tried to solve these limitations by using thick angle irons to hold the magnets, a parallel-notch shaping circuit, and two methods of reducing diffraction.

MAKING THE FRAME

The frame (*Figs. 1* and *2*) consists of four layers: the back plate (shown on the left) with two narrow strips of ¾″ MDF glued to it, and the front plate, which is screwed against it. (See the parts list in *Table 1*.) Also shown is the location of the angle irons, which are clamped at three points against the back plate. These very stiff bars don't need additional support. Look for the straightest ones. If you're able to cut the front and back plate at 45°, do so to reduce diffraction. I wasn't able to (*Photo 1*).

Basically, you start by gluing the long strips of MDF against the back plate and clamping them overnight. *Figure 2* shows the location of the ribbon clamps (6″ × 6″ MDF), also glued against the back plate. I added a similarly sized thin rubber sheet to them to increase friction. Also cut out two 2″ × 6″ pieces of thin wood to act as the other half of the clamps. I simply used screws for fastening these, drilling in the small pieces two holes that were wider than the screws, so the pieces would be able to move freely to and from the ribbon.

Next cut out the slits for the angle irons, and mount them with nuts, washers, and bolts at a distance of around 3⅛″ from each other (assuming your magnets are 1″ wide). You cut the slits so you can move the angle irons toward and away from each other. This is important, since it appears that you can reduce "cavity resonances" by minimizing the air gap between ribbon and magnets. Next, attach your speaker-cable binding posts. I used what was lying around, but prefer the simple bolt type.

Now it's time to make the stand. Glue two MDF plates of about 30″ × 30″ together and clamp them overnight. Predrill the appropriate holes for some angles to connect the frame with its feet.

I bought 50 1″ × 2″ (magnetized over w) ceramic type #5 strontium-ferrite magnets per speaker. These provide the most magnetic flux per dollar. I have been told that one magnet of 1″ width has a magnetic-field strength about 10–30% stronger than two of ½″ placed on each other, which is why I bought the wide ones. Place the magnets in the center of the angle iron.

MAKING THE RIBBON

Ribbon speakers are basically formed of a strip of a thin conducting material suspended in a magnetic field. Usually the conductor consists of several thin parallel strips of aluminum foil glued onto a polyimide (Kapton®) or the less heat-resistant polyester (Mylar®) carrier. Alternatively, you can etch the conductors out of a preexisting laminate of a plastic and aluminum foil. Whereas the former materials are much easier to obtain and produce, the latter may be lighter and more accurately spaced.

In my design, I use a new type of carrier material, namely Teflon®-coated fiberglass (see *Photo 2* and *Table 1*). As far as I know, there's no particular acoustical reason to employ either Kapton or Mylar, apart from the fact that they are widely available as very thin tape and can come with an adhesive layer.

I have read on several occasions that Teflon has a high degree of internal damping (internal loss). For example, Oskar Heil's Air Motion Transformers employ Teflon for this reason. Other materials with high internal loss are polyethylene and PVC.

Also, Nieuwendijk[2] employed fiberglass to reduce the amount of distortion in his planar tweeter design, presumably by reducing the resonant modes of the less than uniformly driven planar membrane. This may reduce ringing of the ribbon, and hence the "metallic glare" possibly associated with ribbon sound. Nieuwendijk established that the amount of distortion decreases when more surface area of the carrier material is actually being driven. This may be a predominant source of distortion in nonuniformly driven speakers such as the Magnepan panels or Daniel Patten's[3] nice push-pull speakers.

Here, a large diaphragm is driven by a wire conductor of much smaller surface area. Thus, try to cover at least ~75% of the tape you use with aluminum foil (or whatever conductor you use). I ordered a 1″ wide roll of C.S. Hyde's thinnest tape of this kind (3 mil backing + 2 mil silicone adhesive; 1 mil = 0.001″), for these two considerations sounded promising. Note that this does make a rather thick tape. Order the same without the adhesive if you can obtain aluminum foil with good adhesive on it. Plain 1-mil Kapton is also available and would increase efficiency of the speaker.

THE CONDUCTOR

There are several choices for the conductor. Figuring all of them out took

TABLE 1
PARTS LIST (FOR TWO SPEAKERS)

ITEM	AMOUNT NEEDED	TOTAL COST	SUPPLIER
MDF (¾″):		$35	
10.5 × 50.5″	4		
1.5 × 50.5″	8		
1.5 × 7.25″	4		
2.5 × 2.5″	4		
13 × 13″	4		
1.5 × 36″ angle iron (3/16″ thick)	4	$20	
0.75 × 1 × 2″ magnets	100	$186	Bunting Magnets
0.25 × 1.5″ bolts, nuts, and washers	12, 24, 12	$6	
¾″ threaded rod, nuts	1, 8	$10	
#8 × 1.5″ sheet metal screws	100	$5	
Teflon fiberglass tape (1″ wide; 3+2 mil thick)	18 yards	$15	C.S. Hyde Co.
cue tape (0.3 mil thick)	150′	$20	
foil block connectors	16	$4	
ring terminals	20	$3	
hookup wire; binding post			
Total:		$304 (about $150 per side)	

most of the time for this project. You can take thin household aluminum foil, cut it in narrow strips, and glue it on the tape. The easiest way to do this is by using a "rotary knife," as sold by fabric stores (kind suggestion of Lief Aden and John Whittaker). Cut the aluminum on a garbage bag or other soft, even surface. Do not get the strong aluminum foil, since this tends to be even thicker. Standard foil is around 1 mil thick, which necessitates a long and narrow total length of it in order to meet about 2.5W DC resistance (which translates into 2.5–3W of AC impedance).

In order to calculate this, you can use the following formula:

$L = R \times m^2/27 \times 10^{-9}$, where L = total length necessary, R = resistance desired (ohms), and m^2 = the cross section area of the foil.

For example, if your foil is 25 microns thick and 5mm wide, then to reach an impedance of 4Ω, you will need a length of 18.5m. The value 27×10^{-9} is the resistivity of aluminum (in ohms/m). The desired resistance = $27 \times 10^{-9} \times$ length (L)/cross section (m^2). The thickness of the foil (L) = length (L) $\times 27 \times 10^{-9}$/resistance (ohms) × width (m). (To calculate the width, exchange thickness with the width.) Note that you can also verify your calculations with a weight measurement by cutting a strip of foil of one width and length:

thickness of aluminum foil (m) = (weight (gram)/2.7×10^6)/length (m) × width (m)

For example, I cut a piece of Reynolds aluminum foil 10mm wide and 1m in length. It weighed 0.513g. This results in a thickness of 19μm. Using the formula for resistance, I calculated the resistance to be 0.14Ω. The measured resistance was 0.15Ω. Thus, the measurements and calculations are consistent with each other.

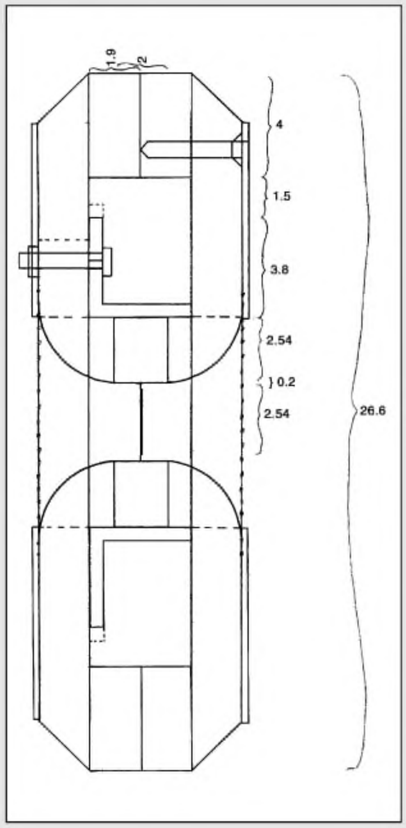

FIGURE 1: Top view of the ribbon speaker. Note that the angle irons (the L-shaped structures) must fit between the top and back plate.

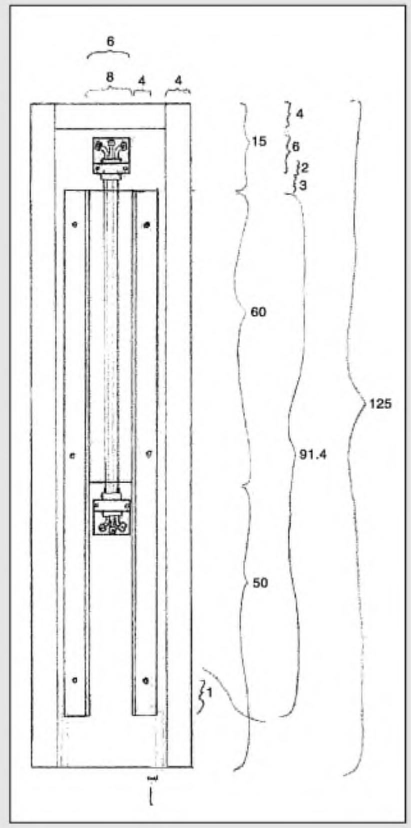

FIGURE 2: Frontal view of the back plate (without stands).

PHOTO 2: The 1″ wide Teflon-fiberglass tape and the sensing aluminum tape. The quarter provides a sense of the dimensions.

ANOTHER CHOICE

Another option is to get a film and foil capacitor (non-metalized). These consist of separable layers of a plastic dielectric and either aluminum or tinfoil. Tinfoil doesn't work, since it tears very easily. Get the aluminum one. It is exceptionally light because it's so thin (about 0.23 mil or 6.5μm thick) and has no adhesive. It results in a high resistance/length of foil, reducing the amount of cutting you need to do. I will try this in combination with the fiberglass/Teflon tape carrier.

Make sure not to buy a cap that has been coated with any resin; you need easy access to the foil. Just cut off the ends of a radial cap (around 0.1μF will do) with a sharp knife so that you're left with a foil of the desired width. This method achieves high accuracy.

The last option (which I took) is to buy a roll of "sensing tape" (or "cue tape" or "sensing foil") from Sound Investment Co. or Media Technology Source. This material is used in cinemas, stuck to the celluloid to trigger

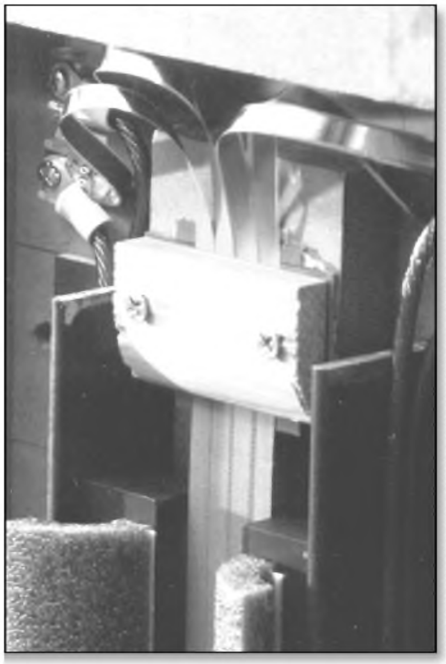

PHOTO 3: The ribbon clamp (center) and the termination of the foil using foil block connectors (used in security systems). Note how the ring terminals are easily attached onto these blocks. The lower part shows the quarter-round cylinder copper-pipe insulating foam used for decreasing diffraction.

preprogrammed events (lights dimmed, curtains opened, and so on). Mine (150′, 0.22″ wide) turned out to be exceptionally strong, probably due to its rather thick adhesive layer (*Photo 2*). The aluminum layer's thickness is around 8.6mm, and that of the adhesive layer is probably another 10–20μm.

MAKING THE RIBBON

First cut off about 48″ of the fiberglass/Teflon tape. Next, tape it, slightly stretched, with the adhesive side up, to a flat surface. Next, fasten a strip of sensing foil as close to each of the tape's edges as possible. Do this slowly, putting only a little tension on the sensing foil. Now, you need to place two more conductors on the tape. Distribute them on the tape as evenly as possible. Make sure the aluminum is fixed completely to the tape, and cut off 44″ of it. Make sure not to bend the tape much. Doing so may detach small pieces of the foil.

Now lead the foil through the ribbon clamps, with the foil side towards the back plate. Make cuts into the ribbon so that you can lead each conductor to its foil block connector (*Photo 3*). Now hook all the conductors in series by using speaker cable terminated with ring terminals (*Photo 4*). Finally, hook up the two ends to the binding posts of your choice.

Note that the polarity needs to be correct: hook up a 1V battery so that the ribbon will move forward (away from the back plate). The side of the ribbon attached to the battery's positive pole should be hooked up at the positive pole of the binding post. When all these conductors are connected in series, the resistance should be around 2.5Ω.

Make sure not to put any vertical tension on your ribbon as you clamp it: its center should be able to move back and forth about ¼″. Otherwise, you'll create a spring system with nasty resonant modes. I used some foam in between the ribbon and the clamp to allow for the ribbon's easy and accurate placement.

FINISHING TOUCHES

Besides the usual paint job, I decided to try to minimize diffraction. First of all, I wished to have a smooth, round surface between the edges of the magnets and the front and back plate.

Buy four gray foam insulating tubes for copper-pipe. Lay them on a table and get a book that is about half their thickness. Now, place a long, thin hobby knife crossways on the book so that it sticks out about half an inch. Using the book as a guide, slowly move the tube against the protruding knife so that it cuts into one of the tube's sides. In this way, cut the tube into four more or less equally thick quarter-round cylinders. Next make sure they fit well in the magnet-front plate gap, and that they make a smooth transition with them.

Now prime the foam and paint it. Next put a piece of double-sided sticky tape on every other magnet. Finally, put two drops of hot glue on each piece of

PHOTO 4: The back plate of an assembled ribbon speaker. Note the two bolts and nuts below the lowest ribbon clamp for closing the magnet circuit.

PHOTO 5: The back view of a finished ribbon speaker. Note the horizontal slits that allow the angle irons to move relative to each other.

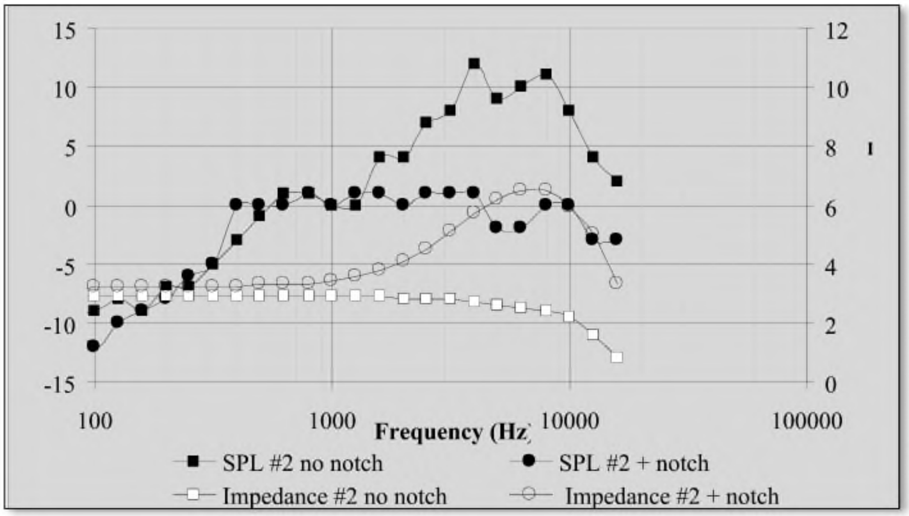

FIGURE 3: Frequency and impedance curves of ribbon speaker #2 with and without the parallel notch filter.

tape and press the insulation foam onto it. It should make a fairly strong bond (*Photo 3*); the foam actually melts a bit.

Second, I wished to damp the front plates' reflections. For this I used velvet, gluing it on top of the primed and painted speaker (*Photo 1*).

Lastly, I tried to increase the efficiency of the speakers by closing the magnetic path. As shown in *Photo 4*, I cut about 2″ (it should just fit between the angle irons) from a threaded ¾″ rod, and screwed a large nut onto each end. I then placed them beneath the lowest ribbon clamp, and clamped them between the two angle irons, thus closing the magnetic circuit. I haven't yet measured the difference in efficiency.

HOW DO THEY SOUND AND MEASURE?

I did the measurements in my room with a Radio Shack SPL meter. I corrected for its sensitivity, and it was C-weighted. I used third-octave test tones. In *Fig. 3*, speaker #2 (solid squares) shows that the response gradually increases from 200–600Hz, reaches a plateau, and finally rises to a 10dB peak at around 6kHz. The SPL then drops off to 0dB at around 20kHz. Rotating the speaker to 20° off-axis reduced this peak by only 2dB. Indeed, this is what all ribbons do.

See, for example, *Fig. 2* of Richard Painter's ribbon speaker[4]: the SPL at 1kHz is 6dB below the peak at 16kHz, with a clear rising slope in between. Think about what could cause it. The width of the ribbon is only 1″, so you'd expect bundling to start occurring around 7kHz (340/0.025 × 2). However, the SPL falls off from that frequency on, so this doesn't help.

Alternatively it could be due to the limited height of the ribbon (around 26″). You could induce bundling by gradually increasing the limited vertical dispersion. You could expect this to start at 260Hz on (340/0.65 × 2), but not to play much of a role at more than two octaves above this frequency (1kHz). It may also have to do with "front-back cancellation." Because of the small size of the frame, the airwaves can travel and meet on the sides to cancel each other out (hence the "figure 8" radiating pattern of dipoles—they are 180° out of phase with each other).

The path length is about 8″, corresponding to a frequency of 2kHz, which is approximately where the slope starts. Martin Colloms suggested that this is the basis of the steady slope. If anyone has better ideas, I'd like to hear them.

No matter what the physical cause is, I tried two electronic techniques to compensate for its response. The simpler one, a contour filter, failed. This type of filter (a resistor and capacitor hooked up in parallel, but as a whole in series with the ribbon) increases the impedance of the system with increasing frequency.

I was unable to remove a strong peak at around 3kHz without severely reducing the output at the highest frequencies (since these are close to 0dB). My Apogee Duettas, however, do employ this circuitry fairly effectively (*Fig. 4*), using a 0.2mH inductor paralleled with a 5Ω resistor.

BEST APPROACH

For my ribbons, a parallel notch filter is the best approach: it allows you to select the frequency band to be shaped and the center frequency at which the decrease in amplitude will be maximal. It consists of a resistor, capacitor, and inductor in parallel with each other, but as a whole in series with the ribbon. Using the following formulas[6], I calculated capacitance, inductance, and resistance:

C (in farads) = 0.03/f

L (in henries)= $0.023/f^2 \times C$

R (in ohms)= 1/6.28 × C × B

where B is the frequency bandwidth (highest frequency affected minus lowest)

My results were: C = 5μF; L = 0.13μH; R = 2.3Ω. I suggest, though, that you lower C and proportionately increase L when the frequency bandwidth spans more than an octave. Since in my case B was about 18kHz (20k–2k) (three octaves), I decided to try 3μF and 0.2μH (note that their product is about the same).

Measuring the SPL with various resistors, I found that at 4Ω the response seemed most flat (*Fig. 3*, solid circles). I used a 0.17μH air-core inductor (0.7Ω DC) and a 3.1μF capacitor (metalized polypropylene bypassed with a film and foil polypropylene). In fact, it's flat (±3dB) from 350Hz to 20kHz! Note how nicely the impedance with the parallel notch filter parallels the SPL of the uncorrected speaker. Although this

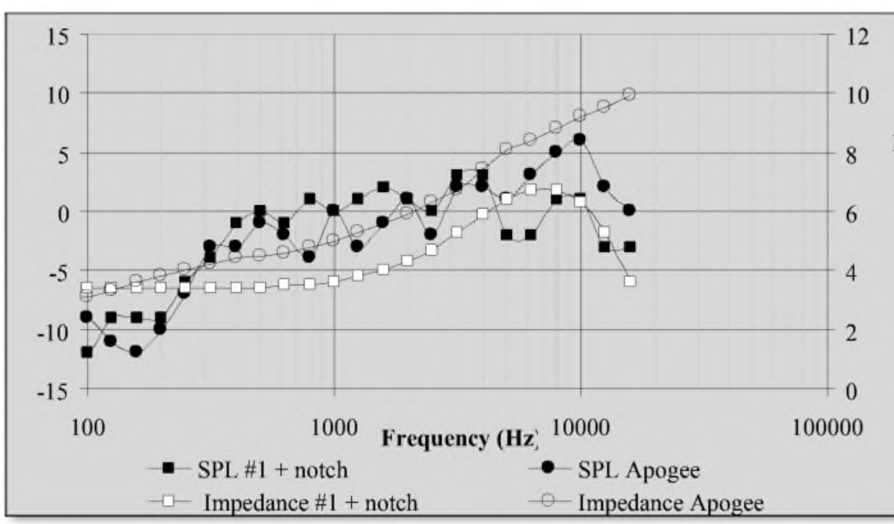

FIGURE 4: Frequency and impedance curves of ribbon speaker #1 and an Apogee Duetta MKII ribbon.

makes the speaker less efficient (it wastes power in the network), it makes it a more suitable impedance match to most power amplifiers (at least 4Ω from around 1kHz on).

Note the similarity of the SPL curves between the Apogee and speaker #1 (*Fig. 4*). The Apogee seems to peak at 6dB at 10kHz, whereas such peaks do not appear in either of my ribbons. In fact, it seems that the Apogee's SPL increases with about 1.3dB/octave from 300Hz on. Maybe its contour circuitry should have been slightly steeper.

When I placed them next to the Duettas and drove them (using an active crossover at 500Hz driving my Dynaco MKIIIs with EL-34s in triode mode), in contrast to the Duetta ribbons, the sound was smooth, detailed, and well imaged. It was very similar to the Apogee's, but had more of the ease of a soft-dome tweeter. I think this is attributable to the damping characteristics of the Teflon/fiberglass tape I used. They are also very reliable. I have played them for more than 100 hours, laid them on their sides, and so on, and the ribbons seem indestructible. Mind you, I also have two cats!

SOME SUGGESTIONS

This speaker sounds very good (certainly for the money) and is easily adjustable. You could experiment much more with it by, for example:

1. making other ribbon sizes/materials (making them lighter);
2. investigating the ribbon-magnet gap effects;
3. trying corrugation of the ribbon (won't work with the fiberglass, since they'll break);
4. equalizing the SPL at line level instead of passively after the power amplifier, which will increase the speaker's efficiency.

I hope you enjoyed the project and the sound. I am curious about your experience. ❖

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Bunting Magnets, 500 S. Spencer Ave., P.O. Box 468, Newton, KS 67114-0468, (316)284-2020, FAX (316)283-4975, http://www.buntingmagnets.com/bunting/home.htm. (magnet type MA-1065)

C.S. Hyde Company, Inc., 461 Park Ave., Suite 300, Lake Villa, IL, 60046, 1-800-441-8063/(847)265-6903, FAX (800)441-8063, http://www.cshyde.com. (tape type 21-38)

Sound Investment Co., 2688 Peachtree Square, Doraville, GA, 30360, (770)458-1679 or 1-800-659-TAPE (8273), http://www.tapewarehouse.com. (tape type #9199A)

Media Technology Source, 10501 Florida Ave. South, Minneapolis, MN, 55438, (612)829-0161, FAX 612-829-0166, http://www.mediatechnologysource.com/home.asp.

Old Colony Sound Lab, PO Box 876, Peterborough, NH 03458, (603)924-9464, FAX 603-924-9467, http://www.audioXpress.com.

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Low Down Power, Part 1

Reviewed by Thomas Perazella

PHOTO 1: Apex Junior front.

Here's a roundup cf four popular dedicated subwocfer amps: the Apex Junior, Marchand PM31, and the Parts Express SW250A and 300-794.

Do you remember that old phrase, "How are you going to keep them down on the farm after they've seen Paree"? The corollary in the audio world should be, "How are you going to listen to music with just your old speakers after you've heard a good subwoofer?"

There is no doubt that clean, deep bass is the foundation of music. Building a good subwoofer is one of the most satisfying segments of the speaker-building hobby. Unlike times past, you now have a wonderful selection of drivers with parameters suited to building a good sub, and at prices that when corrected for inflation are real bargains.

The other three main components of a sub—the enclosure, crossover, and amplifier—have also been going through an evolutionary process. One of the most interesting areas of change has been in the amplification used for subs, and specifically the development of dedicated subwoofer amplifiers.

SUB AMPS

The subwoofer amplifier can actually be defined with a range of parameters. There is no set list of functions for all amps. One common characteristic they all have is a design for mounting directly on the sub enclosure through a hole that puts most of their workings inside the enclosure with connections and controls on the outside.

This is a great advantage, as no space external to the sub is needed for the amplification function. Wiring is easier as only a single line-level cable is necessary from the signal source, although most sub amps also have speaker-level inputs as a convenient function in case line-level outputs are not available. The only other thing needed is a nearby AC power outlet.

Power output from these sub amps ranges from a few watts to hundreds of watts. Most can drive loads to 4Ω, with some capable of going lower. Other common features that can greatly add to convenience include:

1. Input volume level control.
2. Signal feedthrough for line- and speaker-level signals.
3. Line-level low-pass filters with a wide range of adjustment.
4. Phase adjustment.
5. Speaker-level high-pass filters.
6. Standby mode with signal activated auto on.
7. Bass boost.

Some also feature remote controls, balanced inputs, energy-saving class-D circuitry, and defeatable low-pass filters. The best feature is often a very competitive price compared to standard amplification when you take all features into account.

POWER REQUIREMENTS

How can you decide whether a sub amp is the solution to your problem? If convenience and ease of use are high on your list, this is the place to look. If very high power and flexibility are at the top, the decision may become harder. First you must dispel an old myth that says you need a huge amount of power to get deep bass.

The amount of power you need depends on several factors:

PHOTO 2: Marchand PM31 front.

1. What SPLs are you trying to achieve (or, to put it another way, how understanding are your significant other and/or neighbors)?

2. How large is your room?

3. How low do you wish to go?

4. What tuning will you use—sealed box, vented box, passive radiator, transmission line, horn loaded (hope you have a big room or like to live inside a horn throat if you plan to go low with a horn), and so forth?

5. How efficient are the drivers you plan to use?

6. How transparent is your room to low bass; that is, how much room gain can you expect?

7. How big can the enclosure be?

The amount of power you can actually use, however, is primarily dictated by how low you wish to go. How can this be? The traditional wisdom is that the lower you go, the more power you need. Unless you are working in a very small sealed box, the chances are quite good that with a typical driver, if you wish to go very low you will probably run out of excursion before you run out of power handling.

If you don't believe this, try a simple experiment. Using a variable DC supply, measure the cone excursion versus the DC voltage applied to the speaker terminals of a typical woofer. When you do this, be very careful to apply a DC voltage to the driver for only very short periods of time. It does not take much DC to burn out a driver, since the coil is not moving and there is no cooling air moving across it.

You will find that it doesn't take much voltage to drive the cone to the end of its linear travel range. The same is true at low music frequencies. Remember that if you are operating a linear driver above resonance, as you decrease the frequency being reproduced by one octave, you increase the cone travel by a factor of four. It does not take much power at low frequencies to drive most woofers to their excursion limits.

Having heaps of power at very low frequencies, especially in ported designs, usually only results in driving the voice coil harder against the back plate (accompanied by a loud noise) or perhaps right out of the gap. From this, you can extrapolate that you will need a lot of volume displacement (excursion X moving area) to get a lot of clean low frequency bass. Because of linear excursion limits, this generally means multiple drivers. There are exceptions to this, but they are generally designs that are beyond the reach of most hobbyists.

FREQUENCY RANGE

This brings us to another crossroad. What is the definition of a subwoofer? Some would claim that a subwoofer should reproduce frequencies as high as 150Hz. In fact, you will find some of the sub amps have crossovers that can go that high. The other school of thought is that a subwoofer is what its name implies, a device to work below a woofer, generally in the range below 60Hz.

My beliefs definitely fall into the latter category. I won't go into the pros and cons of those two philosophies here. However, you should be aware that if you venture into the over 50–60Hz range with a sub, the driver that bottomed with a loud "thunk" at 20Hz will probably be able to handle the same power at those higher frequencies without excursion becoming a factor. In those ranges, you can probably get higher SPLs without resorting to multiple drivers, but will need more power. That is where the big amps come in.

For this study, I chose four sub amps that—although by no means all-inclusive—give a good indication of what to expect in terms of features, price, and performance when you go shopping. Originally, one of the four I tested had remote-volume control, but it is not cur-

TABLE 1
SPECIFICATIONS

	APEX JUNIOR	MARCHAND PM31	PARTS EXPRESS SWS250A	PARTS EXPRESS 300-794
Power—8Ω	100W	150W	Not specified	180W
Power—4Ω	130W	150W	250W	272W
Height	9 1[illegible]/16″	11″	10″	10 [illegible]/16″
Width	9 1[illegible]/16″	8½″	10″	10 [illegible]/8″
Depth behind	3 [illegible]/8″	5 [illegible]/8″	2½″	3¼″
Depth in front	1½″	1¾″	1″	1″
Circuit type	Not specified	Class AB	Class D	Class AB
Power input	115V 60Hz	120V 60Hz	115V 60Hz	115/230V 50/60Hz
Power cord	Fixed 5½′ cond	6′ 3 cond IEC320	Fixed 5½′ 2 cond	6′ 2 cond removable
AC input filtering	Unknown	L/C filter network	Ferrite core choke	No
Fuse location/size	Front panel/4A	Front/4A breaker	Internal/4A	Front/4A
Exposed power	No/warning label	Yes/no warning	Yes/warning label	No/no warning
Auto power on	Yes	No	Yes	Yes
Line level input	L&R RCA jacks—left jack for mono	XLR socket w/RCA adapter cord	L&R RCA jacks—left jack for mono	L&R RCA jacks— not marked for mono
Line feed through	L&R RCA jacks	None	L&R RCA jacks	L&R RCA jacks
Speaker level input	L&R spring clips—left for mono	None	L&R binding posts ¾″ spacing	L&R binding posts 1″ spacing
Volume control	Rotary marked MIN to MAX	Rotary marked 1 to 10	Rotary marked MIN to MAX	Rotary marked MIN to MAX
Crossover	Continuous rotary marked at 60, 80, and 125Hz	None	Continuous rotary marked at 40 and 160Hz	Continuous marked at 40 and 160Hz
Bass boost	None	None	6dB fixed	6dB fixed
Line level output	Direct	None	None	High pass −13dB at 44Hz, 0dB at 88Hz, +8dB at 500Hz
Speaker level output	Spring clips direct	None	L&R binding w/220µF cap for 6dB/octave filter	L&R binding w/150µF cap for 6dB/ octave filter
Woofer output	Fixed wires 13.5″ w/0.187 female quick disconnects	Fixed wires 26″ non terminal	Fixed wires 22″ w/0.187 female quick disconnects	Fixed wires 30″ w/0.250 quick disconnects
Phase control	Push button switch 0 & 180°	Heavy duty toggle switch 0 to 180°	Rocker switch 0 & 180°	Continuous rotary 0–180°
Crossover range	60 to 125Hz	None	40–160Hz	40–160Hz
Panel indicators	Standby/on LED red—standby green—on	None	Signal tracking (auto on) green LED	Standby/on LED red—standby green—on
Remote control	None	None	None	None
Mounting method	12 holes with attached rubber gasket	8 holes with no gasket supplied	8 holes with no gasket supplied	8 holes with attached rubber gasket

rently being distributed in the US, so I replaced that amp with the new Parts Express 300-794. As a comparison, I tested a standard, high-quality, stereo separate amp using the same procedures, including listening tests. This may help you decide the best route to take. [Two vendors did not choose to participate in this review.—Editor.]

THE CONTENDERS

I chose the following four amps, in alphabetical sequence by manufacturer/distributor: the Apex Junior, the Marchand PM31, the Parts Express SWS250A, and the Parts Express 300-794. As a comparison, I included a "standard" stereo amp, the excellent-value Audiosource Amp Three, to give a reference as to what to expect in terms of performance when taking the "classical" route. Manufacturer specifications are shown in *Table 1*.

APEX JUNIOR

The Apex Junior is the smallest, lowest powered, and least expensive amp of this group. When I first saw it I thought to myself, "Well, what can you expect for less than $90." What I discovered during testing and listening is that the Apex gives a lot more than you would expect. But, more on that later.

It is actually quite nicely featured, with a built-in continuously adjustable crossover, phase reversal switch, level control, both line- and speaker-level inputs, auto on, and both line- and speaker-level outputs. Cost-saving measures are apparent in the spring-loaded speaker-level connections and the fixed AC power cord. To be honest, I'm not dismayed by the speaker-level connections, which appeared to be functional, and you are better off using line-level inputs if possible.

My first impression was, "Not too shabby, especially considering the price." The back is sealed on this unit, so I was unable to photograph the innards. A problem for me, but a definite safety plus for anyone using the amp. Street price at the time of the article was $89.95. *Photo 1* shows the front-panel layout.

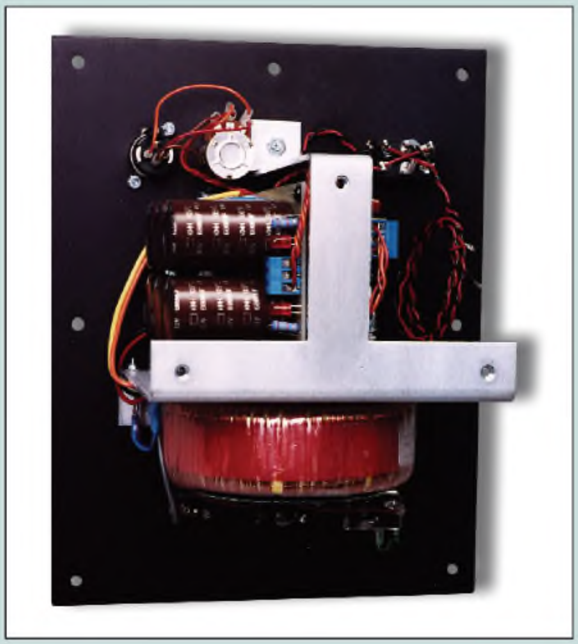

PHOTO 3: Marchand PM31 rear.

PHOTO 4: Parts Express SWS 250 front.

MARCHAND PM31

This piece just screams out "industrial solid." Fancy-schmancy gewgaws are not the province of this amp. From the removable three-conductor power cord to the metal XLR connector and through all the toggle switches and circuit breakers, the design and construction of this amp says "no short-cuts." *Photo 2* shows the front-panel layout.

A look at the back of the panel is equally impressive (*Photo 3*). The design and layout of the circuit boards seem first rate. The power transformer is a huge toroidal design and is followed by a large bank of filter capacitors. The money spent in supply design shows up in the later measurements.

The PM31 is also different from all the others in that it includes no crossover. This is not surprising because Marchand is in the business of making very high quality, very flexible external crossovers, and the unit is designed to complement those crossovers. As such, it is really a full bandwidth amp, making it not only suitable as a sub amp, but also as an internal amp for a self-powered full-range speaker utilizing passive crossovers.

The one caveat immediately evident is that line power is available to the touch at several points on the back of the amp. This is no problem once the amp is safely snuggled up inside the speaker enclosure, but can be the source of a nasty, if not lethal, shock if you plug the unit in before the rear is covered up. In addition, there is not even a warning label to that effect, which is quite surprising in today's world where major lawsuits routinely arise over spilled hot coffee, if not yet over spilled milk. Price at the time of this article is $295 in kit form and $595 assembled.

FARTS EXPRESS SWS250A

What's different about this amp? Well, both the power-supply and amplifier output sections are operated in a switched mode. This results in lighter weight (important in a portable design, but probably not in a sub that sits in a corner), higher efficiency, and lower power dissipation.

Features include a built-in adjustable crossover, volume control, phase reversal switch, and binding posts for the speaker-level connections. Power is provided by a fixed two-conductor cord, and the amp has an auto-on provision. Noticeably absent are line-level pass-through jacks. *Photo 4* shows the front-panel layout.

A view of the back (*Photo 5*) shows a circuit layout more reminiscent of a computer motherboard than a traditional amp. There are no big transformers or large banks of capacitors. The size of those components for a given power capability are proportional to the line frequency.

That's why large airplanes use 440Hz AC instead of 50 or 60Hz. The transformers and filter capacitors would add quite a bit of excess weight to the airplane if the AC frequency were lower. Computer supplies and this amp take that concept a step further and convert the AC line power to DC with no transformer, then use a very high frequency switch to convert it back to AC where a much smaller transformer, rectifiers, and capacitors can re-convert it to DC. Sounds ideal.

What are the drawbacks? High frequency noise. Although the size decreases with frequency, the radiated noise increases. Good circuit layout and shielding are a must to prevent producing an unwanted RF noise generator.

The output section of the amp also operates in a switching fashion. A normal amp has output devices that vary the voltage to the speaker by acting as variable resistors, dividing the supply voltage across themselves and the speaker in proportion to the input signal. That results in a lot of power dissipation across the devices, be they transistors or tubes. In a switching amp, the transistors are either on or off.

How can you get a smooth waveform from a switching device? Switch at very high frequencies, and vary the time on versus off. Then put a filter on the output to smooth out the switching transients and, *voilà*, an analog waveform appears. Since the transistors spend most of their time fully on or off where power dissipation is minimal, the efficiency can be much higher. Less power going to heat means less power used.

Drawbacks? Noise and stability. Again, good circuit design and filtering are in order. Like the Marchand, this amp also had exposed line voltage on the back but did have a warning label. You still need to exercise care until the amp is sealed in the cabinet. Street price was $249.80.

PHOTO 5: Parts Express SWS 250 rear.

PHOTO 6: Parts Express 300-794.

FARTS EXPRESS 300-794

What's with this 300-794? No, it's not the name or model number. Neither of those is evident on this amp. Actually, it is the ordering code, but since I couldn't find another identifier, I'll use the code. Like the SWS250A from Parts Express, this amp is advertised as putting out 250W into a 4Ω load. The specification sheets claims 272W into 4Ω.

This amp has many features (*Photo 6*). RCA jacks are provided for both line and line out signals. Binding posts on ¾″ centers do the same duty for speaker-level signals. There are also continuously variable controls for volume, crossover frequency, and phase. More about the pros and cons of the continuous phase adjustment later.

AC power is connected through a removable three-wire power cord into a standard IEC320 socket with internal fuse and spare fuse. A switch is provided for 120V or 240V operation.

The back of the amp shows a large, standard EI-core-type power transformer, two relatively large filter capacitors, and two circuit boards. The AC power connections are all covered by shrink tubing to prevent a shock hazard.

When you plug the amp in, the power transformer is always energized, even if the power switch is in the off position. The switch only controls the power out section of the amp. Unlike other amps that have a passive line-level pass-through, input line signals in this amp go through a circuit that provides a high-pass function on their journey to the output jacks as long as it is plugged in. Street price was $225.80.

AUDIOSOURCE AMP THREE

This amp has become one of my favorites for best bang for the buck in a basic stereo power amp. It features a rating of 150W/ch into 8Ω and 400W in bridged mono. It is small, rugged, and performs well. In an upcoming project, I will be using three of them instead of a single super amp. The front panel has two volume controls, an illuminated pushbutton power switch, and two illuminated pushbutton switches for selection of two sets of speakers.

The rear has a switch for manual or auto on, two sets of line jacks for line in and out, a stereo/bridged mono switch, and two sets of binding posts on ¾″ spacing for two sets of stereo speakers. Power is provided by a detachable three-wire AC cord. A fuse holder is provided for input power, but four fuses are located inside the unit for DC power bus protection.

Inside, it has a large toroidal power transformer, four large filter caps, and relay switching of the speaker outputs which doubles as speaker protection against faults or DC offset. Street price was $350.

Next month, Part 2 covers system performance of the four units and includes listening tests. ❖

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FAX 818-248-0490
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URL www.audioc.com

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HSU Research
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Phone 714-666-9260
FAX 714-666-9261
URL www.hsuresearch.com

Marchand Electronics, Inc.
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Webster, NY 14580
Phone 716-872-0980
FAX 716-872-1960
URL www.marchandelec.com

Parts Express
725 Pleasant Valley Dr.
Springboro, OH 45066-1158
Phone 800-338-0531
FAX 513-743-1677
URL www.partsexpress.com

Product Review

MJ Magazine's LP Test Disk 3

By Gary Galo

MJ Technical Disk vol. 3. *MJLP-1001/1002 (2 LPs). MJ Records, Seibundo Shinkosha Publishing Co., Ltd., Japan. Available from Old Colony Sound Lab, PO Box 876, Peterborough, NH 03458, (603) 924-6371, FAX (603) 924-9467, e-mail custserv@audioXpress.com, www.audioXpress.com (part number LFMJ1, $99.99, plus s/h).*

MJ Records of Japan has produced a test LP set that invites comparison to the *Hi-Fi News and Record Review Test Record*, HFN 001, which I reviewed in *AE* 2/98. Most recordings in the MJ set were made by Nippon Columbia, Ltd. (Denon). Although the actual disc cutting was done in Japan, the lacquer masters were sent to the US for plating and pressing by Record Technology, Inc. The pressings are made on 180-gram vinyl, and are impeccable.

Although the jacket notes and original booklet are in Japanese, Old Colony Sound Lab has included a second booklet with a complete English translation of all notes and photo captions. The translation is not without problems (more on this later).

MUSIC RECORDINGS

MJ Technical Disk vol. 3 consists of two discs, one containing music, and the other a battery of test signals for your LP playback system. The three works on the music LP, Disc #1, each feature harpist Mai Takematsu. Side 1 contains Handel's *Concerto for Harp and Orchestra in B-flat, Op. 4, No. 6*, with Ms. Takematsu ably accompanied by the Jean-François Paillard Chamber Orchestra conducted by M. Paillard himself. Readers collecting classical records back in the 1960s and '70s will recall many outstanding recordings by the Paillard ensemble recorded in France by Erato, and issued in the US by The Musical Heritage Society.

The Handel recording was made in Paris at Opus Systems Studio, and engineered by Mr. T. Shiozawa. Eight microphones were used for this recording, the main pickup consisting of a pair of B&K 4006 spaced omnis, plus a pair of vintage Neumann U-47 mikes as outriggers (the U-47 is the mike used by Mercury in their first Living Presence recordings).

A pair of Schoepps 62HU microphones was used for the solo harp, and two Schoepps CMC-5 accent mikes were also used on the harpsichord and the flutes. The mixing console was an SAJE MkII, with additional microphone preamps manufactured by Studer and Avalon. The analog tape recorder was a ½″ Studer A-820 running at 38cps (15 ips), with outboard Dolby SR noise-reduction units.

Side 2 of the music LP consists of Ravel's *Introduction and Allegro* (for harp, flute, clarinet, and string quartet), plus a solo harp performance of Debussy's well-known piano work *Clair de Lune*. These recordings, also engineered by Mr. Shiozawa, were made in Tokyo at Columbia's Studio No. 1. Six microphones were used for the Ravel, including the B&K 4006 mikes for the main pickup, with Neumann U-67 Revivals as outriggers, and the Schoepps 62HU mikes on the harp.

The Debussy was recorded with the Schoepps pair on the harp. An SSL-9000 mixing console was used, with GML-8300 microphone preamps, along with the ½″ Studer A-820 and Dolby ST noise reduction. Digital reverberation was added to both the Debussy and Ravel recordings.

The Handel recording appears to have also had artificial reverb added, though no mention is made of this. Although the notes describe a dead recording venue with a low ceiling and carpeted floor, this recording is extremely reverberant—too much so for my taste. Instrumental timbres are quite realistic, but the stereo image is rather vague, and the harpsichord is difficult to hear. Many Japanese classical recordings are overly bright in tonal balance. This recording is not, and the string sound is especially warm and sweet.

I much preferred the Ravel and Debussy side. In the former, the soundstage is rendered with greater precision than in the Handel, and the digital reverb has been added more judiciously. I still find the recording too reverberant, though, and the reverb tails do not sound as though they are from the same acoustic space as the performers. There is a realistic warmth and immediacy to the reproductions of the individual instrumental timbres, but most analog purists will balk at the idea of using digital reverberation on an LP.

For a superior recording and performance of the Ravel, I suggest tracking down a used L'Oiseau-Lyre LP (SOL 60048, pressed in the UK by Decca) of the fabulous Melos Ensemble recording of this work, featuring harpist Osian Ellis. This recording was made in

1961 in Walthamstow Assembly Hall by Decca's Kenneth Wilkinson, one of the 20th century's foremost classical recording engineers. A fine CD edition appeared in 1997 on Decca's *The Classic Sound* reissue series (452 891-2).

TEST TONES

Disc #2 contains a battery of test tones for turntable, arm, and cartridge evaluation. Bands 1–4 are 1kHz reference signals for level calibration, left, right, left plus right (mono lateral), and left minus right (mono vertical). The first two can be used for crosstalk evaluations. If a level meter is available, bands 3 and 4 should produce the same level reading as 1 and 2. If your arm and cartridge are in top shape, nearly complete cancellation should occur on band 4 if the mono switch is activated.

HFN 001 doesn't duplicate all of these tests, but its voice phasing and pink-noise channel balance tests are a reasonable substitute for the first three. You can use the left- and right-channel pink-noise test to evaluate audible and measurable wideband crosstalk, and use the voice phasing test in conjunction with the mono switch for checking cancellation of out-of-phase information. HFN 001's cartridge azimuth test is similar to band 4, consisting of a vertical (L-R) 300Hz tone, and can also be used with the mono switch for out-of-phase cancellation testing.

Bands 5–14 contain spot frequency tones for frequency-response testing. Following a 1kHz reference tone, there are nine tones at one-octave intervals from 16kHz down to 63Hz. These have been recorded with the RIAA curve. You will need an accurate RIAA phono preamp and dB meter for these tests. HFN 001 does not contain this test.

Bands 15 and 16 contain 300Hz test tones for stylus force evaluation (incorrectly called "stylus pressure" in the booklet). As the instructions note, you must make sure that your anti-skate adjustment is correct for the stylus force you are using before proceeding (more on the anti-skate test later).

Band 15 is recorded at 50µm lateral, and band 16 at 70µm lateral. Band 15 should play cleanly on any properly operating playback system, but band 16, which is 3dB higher, may stump some lower-compliance audiophile cartridges. HFN 001 has three 300Hz bands for tracking ability—at the beginning, middle, and end of Side 2—which is more useful for evaluating the combined tracking force/anti-skate setting at a variety of points along the record surface.

Band 17 is a mechanical impedance test for the cartridge. For those mathematically inclined, formulas are given for calculating the cartridge's mechanical impedance and compliance. The entire band consists of a 100Hz tone. You must adjust the tracking force and note the force at which mistracking occurs in order to do the math required for this test. HFN 001 has no such test, but this one is likely to befuddle the non-engineers in the audiophile community.

Band 18 is a wow and flutter test using a 3150Hz sine wave. Although the effects of wow and flutter are audible with this signal, precise measurements are best made with appropriate test equipment. HFN 001 has no such test.

Bands 19 and 20 contain intermodulation distortion (IM) tests. Band 15 has 5kHz and 5.4kHz tones, and band 16 has tones at 8kHz and 8.4kHz. In both cases, the levels of sum and difference products may be audible, but test equipment capable of IM measurements will be necessary for precise evaluation.

Note that conventional two-tone SMPTE IM measurement systems, such as the Sound Technology 1700B, will not support the frequencies used on these bands. You will need a sophisticated measurement system, such as the Audio Precision System 1 or System 2, in order for these tests to be useful. HFN 001 does not contain IM tests.

ANTI-SKATE

Band 21 is a blank (grooveless) zone for anti-skating adjustment. The translator must have been unfamiliar with audio terminology, since the English translation refers to the anti-skate adjustment as the "inside force canceller adjustment." This is a test I've used for many years. Simply adjust the anti-skate until the arm remains stationary on the grooveless record.

However, HFN 001 contains four bands of 300Hz test tones, at four levels, for anti-skate adjustment. This is a more effective test than the grooveless record, since there are dynamic effects that take place when the stylus is modulated with recorded signal. The grooveless disc will get you close, but HFN 001 offers a superior method for fine-tuning. Note that HFN 001 has rather wide, grooveless spaces between the various bands on the record, so you can use these spaces for coarse adjustment.

Bands 22–27 are crosstalk check signals at 300Hz, 1kHz, and 5kHz for testing left-to-right and right-to-left crosstalk. You can use these without test gear, but you'll need a dB meter for a precise evaluation. HFN 001 does not duplicate these tests (see my previous comments on pink noise). Bands 28–31 contain phase test signals at 315Hz and 1kHz.

In each of these bands, the sound image begins in phase between the two channels (centered mono). As the phase sweep proceeds, the sound image moves toward the right channel, then back and forth between left and right, then out of phase, and finally centered mono again. If your system and room are in good shape, you should be able to precisely localize the audio signal as it makes its passes through the various portions of these tests. This is a much more sophisticated test than the simple voice phase check included on HFN 001.

Bands 32–35 contain reference tones for tuning instruments. In these four bands, a reference "A" above middle "C" is given at the international standard of 440Hz, plus 442Hz, 444Hz, and finally at 415.3Hz (used by performers who specialize in Baroque literature on period instruments). These tones are not duplicated on HFN 001, but they will probably not be useful to most audiophiles.

ARM/CARTRIDGE RESONANCE

Side 2 of disc 2 is devoted to arm/cartridge resonance testing. Again, the translator fails to use the proper English terminology, with the word "toner" used instead of "tone arm". This side is titled "Toner Adjusting Signal," with the caution that it is important that "the toner match the cartridge." There are no voice announcements on this side (they would have to be in English to be useful, anyway).

After a long lead-in groove, you hear a 1kHz pilot tone for 24 seconds. Immediately following the pilot tone, a sweep from 3Hz to 100Hz begins. In order to use this band without test equipment, you are instructed to start a stopwatch as soon as the 1kHz pilot is over.

When the arm/cartridge resonance point is reached, and the tonearm visibly vibrates, stop the watch. Using the supplied timing chart, you can determine at what frequency the resonance occurred. This is a cumbersome and inexact process.

You can also use this band with a B&K 2305 chart recorder, but there is confusion in the translation of the instructions. These suggest using the 1kHz pilot tone as a means of synchronizing the timing of the 2305 recorder and state that you should "read the scale as 1/10 on the recording chart," so that the pilot signal occupies the space between 10Hz and 30Hz. But, the supplied sample chart requires no such compensation, since it is calibrated from 1 to 100Hz. Therefore, the pilot signal should occupy the space between 1Hz and 3Hz, with the recorder timing set so this takes exactly 24 seconds.

Band 2 of disc 2 contains 13 ultra-low frequency spot signals at 1Hz intervals from 3Hz to 15Hz. Each low frequency tone is recorded in the left channel, accompanied by a 1kHz pilot tone in the right channel. When the arm/cartridge resonance point is reached, the low frequency tone will cause an audible warble in the 1kHz tone. There are three seconds of silence between each spot signal, but no voice announcements.

In order to keep track of the frequency, you must count the number of times that the 1kHz pilot tone appears, and determine the low frequency spot from your count. As an example, your stylus has arrived at the 6Hz spot tone the fourth time you hear the 1kHz pilot tone. I found this entire side extremely frustrating to use. HFN 001, with its English frequency announcements, is much easier. In addition, HFN 001 has two separate sweeps, one for vertical resonance, and the other for lateral.

One further complication is that the

tests on side 2 of the MJ disc were recorded with a 250Hz bass turnover frequency. MJ supplies two schematics, one a complete op-amp-based RIAA circuit with a 250Hz turnover (but with no advice on which op-amp to use), and the other a compensation circuit to re-equalize an RIAA preamp (which has a bass turnover of 500Hz). These schematics are supplied without explanation, and the re-equalizer, in particular, will probably confuse the technical novice. The entire HFN 001 record was recorded with the standard RIAA curve. Considering this factor, the ease of use, and the inclusion of both lateral and vertical test, the HFN test disc wins the arm/cartridge resonance tests hands down.

In addition to the translation problems already mentioned, there are other instances in which the English is awkward, such as the subtitle "Analog Check LP in CD Age." The word "check" is used consistently where "test" would be more appropriate. Singular is frequently used where plural is more appropriate. As an example, the ten bands used for the frequency response test are called "Spot Frequency Test Signal..." rather than "Signals." This translation should have been proofed by a technically adept person whose native language is English.

WORTH THE PRICE?

At $99.95 US, *MJ Technical Disc vol. 3* is an expensive set of test LPs. With the *Hi-Fi News and Record Review* disc selling for only $29.95 from Old Colony, audiophiles will naturally wish to know whether the MJ set is worth the extra $70. The MJ collection will be most useful to those who have the test equipment to take advantage of some of the more sophisticated tests, such as IM distortion, wow and flutter, frequency response, and crosstalk.

The *Hi-Fi News* LP is still my first choice among test LPs, allowing you to perform all critical tonearm and cartridge adjustments easily and accurately, without test equipment. I find the arm/cartridge resonance tests on side 2 of MJ disc #2 so cumbersome to use that I do not plan to ever return to them. HFN 001, with voice announcements, is much easier to use, and the separate lateral and vertical tests make it even more thorough.

HFN 001 also has more precise anti-skate tests, and spreading the tracking ability tests over the beginning, middle, and end of the side makes the *Hi-Fi News* LP superior for this test, as well. The first band on HFN 001 consists of a left- and right-channel voice identification, a fitting place for any battery of tests to begin. The MJ LP has no such channel ID (unannounced test tones make channel identification more difficult).

I was also less than enthralled with the classical music recordings on MJ disc #1, but some listeners will, undoubtedly, disagree. If you have the test equipment to take advantage of the specialized tests on the MJ set, then it may be worth the cost. For most audiophiles, the *Hi-Fi News* LP is a better bet for a lot less money. ❖

Note: A new revision of the anomalies in the text translation referred to by Mr. Galo has been made in the hope that the English version of it will be clearer.

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SERIES COS

Angel Luis Rivera shares with SB *readers G.R. Koonce's response regarding series crossover networks*—Eds.

I saw your letter in *SB* 5/00 about series COs (crossovers) and thought I would e-mail you. Note that the CO shown in David Weems' *Great Sound Stereo Speaker Manual*, Fig. 8-6(B) (page 146), is not a true series three-way CO, but a combination series-parallel CO. It is correctly labeled in the second edition of this book, which we co-authored just this year. This same CO schematic appears as Fig. 6-5(B) (page 83) of *Designing, Building, and Testing Your Own Speaker System*.

The series CO looks good in theory, but is a bear in practice. The first-order is a "constant-voltage" CO, as the sum of its outputs always matches the input, obvious by inspection as the two drivers are in series across the input. Thus with "ideal" drivers the acoustic output would always match the electrical input (in magnitude and phase) even if you miss the CO parts by a bit. This may have been the CO that Mr. Fried [*see Rivera letter*] was discussing. I have seen the first-order series CO used in commercial speakers. Higher-order series COs do not have this constant-voltage characteristic, and I have never seen a higher-order series CO used, but I'm sure someone has done so.

The problems in practice with the series CO are:

1. Tough requirements on the coils: It is harder to get coils near "ideal" than it is capacitors. You worry about the losses in all coils, but for coils in series with the woofer you must also worry about the resistance and linearity with large currents. With a parallel CO the woofer coils affect only the woofer and you don't worry about their high-frequency performance as long as they keep blocking the current to the woofer.

Look at the first-order series CO schematic; the woofer coil is right across the tweeter. Thus not only must it have low resistance, high current capacity, and moderate losses at woofer frequencies, but it must also under these same conditions have low losses and high linearity at the tweeter frequencies. This just about rules out any cored coil (a big cored coil will have high losses) and means very large air-core coils to get the resistance down. Any low-frequency voltage developed across the woofer coil due to resistance appears right across the tweeter terminals and could lead to distortion. With a second- or higher-order series CO this voltage would be blocked by the series capacitor.

2. Sensitivity to load variations: All CO sections are sensitive to load variations, and thus Zobels are generally used to try to make the driver look closer to a resistor. With a parallel CO the load variations of a driver will affect only the one CO section. With the series CO any driver impedance variation will affect the response of all the drivers. Here again the first-order tends to "correct" for this, but the higher-order will not.

3. Interaction between HP & LP: As you noted in your letter, when one CO component value is changed it affects the response of all drivers. You must thus change all values to correct for this. This leads to the following problem.

4. About impossible to design for real driver loads: I have found that when you design a CO via book equations and then test the system with real drivers, you do not get what the books predict. This is true even if you use Zobels. One major problem is that the "book" equations do not take into account the offset in location of the acoustic phase centers of the two drivers.

For example, the books say that for a second-order CO (series or parallel) you must hook the tweeter inverted polarity. I find in testing that about half the time you need the tweeter normal polarity with the second-order CO. Note that the first-order CO (either series or parallel) with both drivers normal polarity is the most sensitive to acoustic center offset. This was covered in *SB* a couple of years ago.

I have found that the only way to develop a CO that works is via listening and playing, testing, or computer modeling. Listening and playing takes months, testing takes weeks, modeling only hours. Several years ago when I was developing my DOS-based CO modeling program based on discussions with David Weems, I included the ability to do series COs.

In several years of using the program I have successfully developed only one series CO! That was a first-order and I did it by developing the parallel CO and then converting it to a series CO. David was going to try the series first-order versus a parallel first-order versus a parallel second-order for one project for our book, but we ran out of time so that project was dropped. I don't think he has yet gotten back to work on that system. I am interested in what he finds in terms of "sound" with the various CO designs.

As you gain experience with modeling parallel COs, you learn what component to change to vary the total system acoustic response and can rather quickly get to an acceptable design (much of the time). With the series CO this is not true; each component you change means you must go back and change all the other values! You quickly become confused and give up.

By the way, if you are interested in playing with CO modeling, the DOS-based modeling program ships with our book along with the needed driver data files for several drivers. After my experiences with trying to design series COs via modeling and discussions with David, I omitted that capability when I re-wrote the modeling program for Windows.

In summary I was at one time very interested in series CO designs. After trying to develop some I have lost interest. If you are going to play with them, I recommend selecting some drivers that will allow a first-order, which has the best theoretical advantage and you are less likely to go nuts playing with it. I would also stay with a good quality aircore coil. Hope these experiences are of some value to you.

I am including third-order series CO equations, which I just found. Seems back in 2/88 I wrote a CO design program that designed the Butterworth third-order series CO. Referring to the schematics published with your letter, Fig. 2 (A) the third-order series: the top L to woofer is L1, the woofer input cap is C1, and the cap across the woofer is C2. For the tweeter the cap is C3, the input L is L2, and the L across the tweeter is L3. Rw is the woofer resistance, Rt is the tweeter resistance, fc is the CO frequency, and Wc = 2π fc.

For the woofer:

L1 = 4 Rw/(3 Wc) = 0.212 Rw/fc
C1 = 3/(2 Rw Wc) = 0.239/(Rw fc)
C2 = 1/(2 Rw Wc) = 0.0796/(Rw fc)

For the tweeter:

L2 = 2 Rt/(3 Wc) = 0.106 Rt/fc
L3 = 2 Rt/(Wc) = 0.318 Rt/fc
C3 = 3/(4 Rt Wc) = 0.1194/(Rt fc)

CHEATER PLUG CONTROVERSY

After reading the manufacturer's comments from C.C. Poon, President, Monarchy Audio (*AE* 5/00, p. 33), I am absolutely outraged by his last paragraph, which reads, "For users who experience hum problems, a simple solution is to use a 'cheater plug,' available from most hardware stores. The safety issue will not be compromised."

Not only is his statement absolutely false, it is symptomatic of widespread ignorance of basic engineering principles among high-end audio manufacturers. I've heard such recommendations voiced by representatives of many other (well-known) manufacturers, but I never thought I'd actually see it in *print*. The "cheater plug" he refers to is more properly called a 3-to-2-prong adapter. They are intended to provide a safety ground for equipment that requires one (i.e., was supplied with a three-prong plug), but must be used with an old two-contact outlet!

Use of such adapters to disconnect the safety ground is not only extremely dangerous, but it also violates the spirit of National Electrical Code and Underwriters Laboratory safety testing. For a manufacturer, it also produces legal liability for subsequent shock, electrocution, or fire. Bear in mind that, in the absence of a safety ground connection, the interconnecting audio cables in a system will carry lethal voltages throughout the entire system in the event that just one piece of equipment fails.

The only exception to a requirement for safety grounding is equipment originally supplied with a two-prong plug. It incorporates internal design features, such as double-insulation or overtemperature one-shots, which ensure safety even under conditions of internal component failures.

The most effective cure for "ground loop" hum and buzz problems is also safe. It is to install an isolation transformer on the signal line(s) at the input of the amplifier. Products such as Jensen's ISO-MAX isolators are designed specifically for the purpose.

Bill Whitlock
Bill.whitlock@verizon.com

C.C. Poon responds:

I am surprised by the scathing attack on our product by Mr. Whitlock without his inspecting the product. Let me state up-front:
It is perfectly safe to use a cheater plug to convert all Monarchy amplifiers from three-pin to two-pin AC sockets.

Inside the Monarchy amps, the hot and neutral wires of the incoming AC stay in the primary winding of the toroid transformer, which is designed to stand up to 3550V for at least one minute between the primary and secondary windings. There is no other connection—through a bleed resistor, or a capacitor, or whatever—that might produce a path for the incoming AC voltage to the chassis. Using a two-prong plug adapter, all Monarchy products are as safe as any other equipment originally supplied with a two-prong plug.

We use a standard three-pin IEC connector on our products to make it easy to adapt to other world standards, just like all the PCs. We could very well have used a two-wire plug.

Nor would the use of an isolation transformer totally remove the ground noise, if the ground (third pin) is still intact. Jensen makes excellent transformers, but for about the same investment an AC isolation transformer would not only eliminate ground noise, but present a much higher level of safety than using a ground wire.

While using a cheater plug does not compromise the safety of Monarchy Audio products, it probably may not apply to other products, and is therefore not recommended.

GO FIGURE

I just read with great interest Louis C. McClure's article, "Determining Optimum Box Dimensions" (*SB* 2/00, p. 42). I implemented the solution using the Solver in a Microsoft Excel spreadsheet (*Fig. 1*).

Gary Manigian
Martinsville, N.J.

Louis C. McClure responds:

I appreciate Mr. Manigian's interest in my article and his taking time to run the calculations on the Microsoft Excel 97/2000 program.

Mr. Manigian's results were very close to the dimensions that I stated in the article, although there are slight differences that I do not thoroughly understand. I re-ran my calculations as presented in the article for the same volume (5530 in³), and the results are as follows:

MR. MANIGIAN'S RESULTS:	MY RESULTS:
Depth: 10.929″	10.75563″
Width: 17.684″	17.92605″
Height: 28.615″?	28.68168″
H × W × D: 5530.3763	H × W × D = 5529.9989

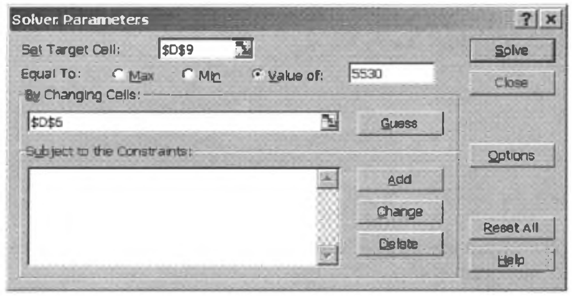
(Note: I rounded the figures off to the fifth place, using a scientific calculator. Multiplying the depth × the width × the height in Mr. Manigian's results produced a volume of 5530.376 in³. Multiplying the same in my calculations resulted in a volume of 5529.9989 in³.

This very slight difference is not important; however, I wonder why there should be a difference in results using my method and Mr. Manigian's method. I do not have the Microsoft Excel 97/2000 program that Mr. Manigian used, and therefore was unable to run the computer program. Instead, I used a standard (Casio Super FX) scientific calculator. Thank goodness I did not need to use my trusty old slide rule—or there would have been a much greater difference between his results and mine!

For fun, I calculated that his results were approximately .00005714% over, while mine were .0000198% under! I ran my calculations out to five decimal places, whereas he ran his to three decimal places. However, I think both methods are close enough. This was an interesting comparison between results obtained using a computer program and a scientific calculator.

Side	Dimension(in)
Depth	10.929
Width	17.684
Height	28.615
Box Volume	5530.000

1) Run Solver (From menu, Tools, Solver)
2) Set parameters as shown below:



Set Target Cell: D9, the total box volume that you are solving for.
Equal To: Value of: 5530, the value for the box volume in the example.
By Changing Cells: D6, the depth of the box.
Click the Solve button and accept the value

FIGURE 1: Solving for box dimensions.

POWER-SUPPLY MOD

I think I have a very plausible explanation for the features in the PP-1 power supply ("A Simple PP-1 Mod," *AE* 6/00, p. 14) that don't seem to make sense (the regulator that is used as a current source and the two Zener diodes that, if anything, could serve to limit voltage surges).

Try replacing the two 17V Zener diodes (D1 and D2) in the original schematic with 7.5V Zener diodes. After the change, you will have a Zener-regulated power supply, with the Zeners fed by a current source, which makes very good sense. This supply will provide very cleanly regulated >±7.5V DC at low current. If the load current is relatively constant, values can be calculated that result in minimal power dissipation by the Zener diodes (*Fig. 2*).

It seems obvious to me that this was the intent of the original design and that the circuit was changed afterwards, either inadvertently or due to some problem—maybe heat dissipation in the Zeners. I don't know the current draw of the preamp itself, but, as long as the chosen Zener diodes can dissipate the required power, this certainly looks like a very simple and worthwhile modification. (I do not

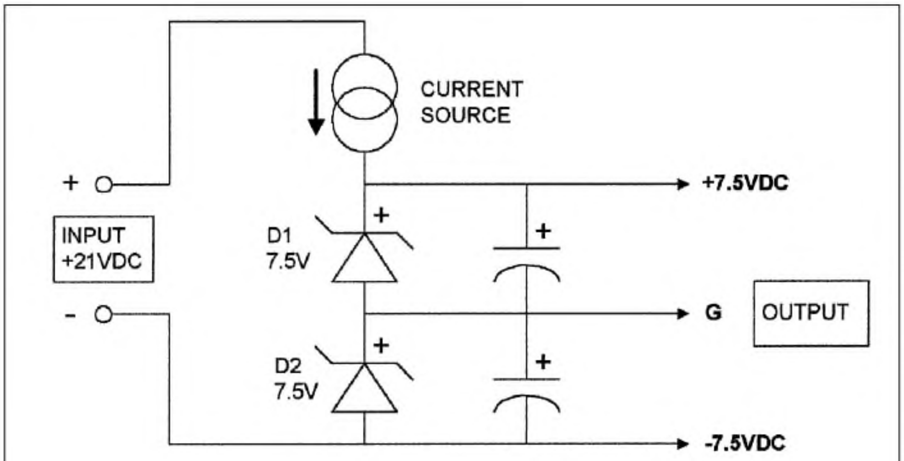


FIGURE 2: Simplified schematic (after change).

own one of these and haven't tried this, so this information is deduced from the schematics.)

Keith Levkoff
kLevkoff@panix.com

Gary Galo responds:

Mr. Levkoff offers a possible explanation of the NAD PP-1 power supply, but I doubt that this is what NAD had in mind. If a Zener-regulated power supply is what they had intended, then the Zener diodes they selected have the wrong value. As Mr. Levkoff points out, these Zeners must be changed to 7.5V in order for the circuit to operate as a Zener-regulated supply.

But I do not consider this to be a satisfactory solution. In my opinion, "Zener-regulated" is an oxymoron, since this type of supply does not regulate at all. It simply provides a voltage reference.

Ripple rejection is also inadequate, since the 317 configured as a current limiter has minimal line rejection. The only line rejection provided by the NAD circuit is that of the post-317 R/C networks.

A true regulator monitors its own output, and provides dynamic correction using feedback. The regulator circuit I suggest in Fig. 1 offers dynamic performance and line rejection far superior to the series resistor/Zener diode topology.

SATISFIED CUSTOMER

In today's business world, you don't often run into an outstanding outfit. I recently had a very pleasant experience with the good folks at Audio Advancements, a high-end audio firm run by Mr. Hart Huschens, and believe that such business ethics deserve acknowledgment.

I wished to experience pure triode, class-A tube sound. High power single-ended triode amps cost megabucks, so I decided the affordable way to get that sound was with a headphone amp. Soon I ran across Audio Advancements and their Ear-Max unit, a neat little three-tube, all triode, all-class-A unit. After a few days of use, I can testify that it is one of the smoothest, sweetest-sounding tube amps I have ever experienced. But, that's not the reason for this letter.

The first unit I received had a minor problem. I thought, "Uh-oh, now I must go through the return-and-wait hassle." I contacted Mr. Huschens and got a surprise. His response was, "That shouldn't have happened—I will get another unit to you via next-day air, and you can ship the defective one back to me in the same box, at my expense." And he did just that.

If you are in the market for high-end turntable components, headphones, or headphone amplifiers, be sure and consider Audio Advancements. People who conduct their business with such high standards deserve to be rewarded.

William A. Shappley
Monteagle, Tenn.

TREBLE FILTER UPDATE

This is in reference to L. Mirabel's interesting article, "A Filter for Treble Distortion in Recordings" (*AE* 6/00, pp. 6–9).

Connection dots are missing in Fig. 3, at the junction of R1/C2/R2/D2 in both supply regulators. Figure 7 is not really the printed circuit layout, as indicated in the text—it is a stuffing guide. Anyone wishing to make a PC board must first invert the layout shown. This was a bit difficult to ascertain since the stuffing-guide component designations didn't correlate to those in his schematics.

I was curious about Mr. Mirabel's preference for two single amplifiers over a twin for "quality considerations," and was hoping he could elaborate. If compensation is needed, of course the single amplifier is mandatory, but he did not use compensation (pins 1 and 5) on any of the TL071s. The specs for the TL071 single and the TL072 dual amplifiers are identical in the Texas Instruments Data Book. Even the quad TL074 is relaxed only in

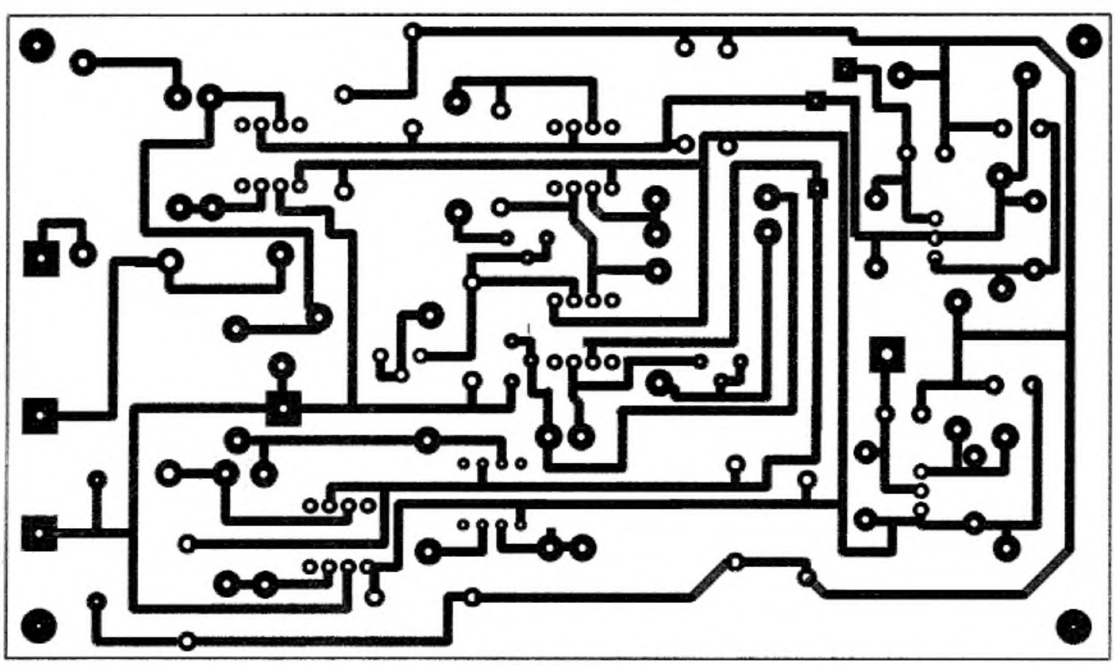


FIGURE 3: Treble filter circuit board layout.

the input offset voltage: 15mV versus 9mV for the single/dual. All the multiple amplifiers have 120dB of crosstalk attenuation, so that would not appear to be a quality issue.

Looking at the specs for the much higher performance Burr-Brown OPA 604 and OPA 2604 (single and dual) he refers to on page 7, the 2604 has 100pA input bias current versus 50pA for the 604. Current noise density is 6fA/√Hz versus 4fA/√Hz for the 604. Neither difference is significant. The 2604 has 142dB of channel separation.

I was also curious as to why he had oscillation problems with the NE5534, OPA 604, and OPA 2604 (see p. 7 under "Components"). The OPA chips are both JFET input op amps (as is the TL071) and are unity-gain stable, but the bipolar input NE5534 is not, and would require an external compensation cap for voltage gains less than five. While there are no decoupling caps shown in Fig. 1, there appear to be decoupling caps at the supply pins of each IC (the "Cx"s in Fig. 7).

Nothing in his design or PC board layout indicates a propensity toward instability. (Note that if you build the circuit as-is, be sure to use the TL071 single and not the TL070. The latter is not unity-gain stable, and would require an external compensation cap.)

One final consideration based on years of doing mean time between failure (MTBF) reliability calculations: Multi-turn trimpots have two advantages over single-turn types. First, the lead screw mechanism uses a slip clutch at each end, making it impossible to overpower the end stop and damage the pot.

Second, the small adjust screw opening in the case (sealed by an O-ring in the better pots) keeps dirt and contaminants out of the wiper/element. However, as Mr. Mirabel pointed out, it does indeed take a fair amount of time to crank in a desired setting. Since there is no wiper position indicator, presetting requires an ohmmeter, usually before you solder the trimpot in place. Also, there are no audio taper multiturn pots to my knowledge.

My letter is certainly not meant to be a critique of Mr. Mirabel's inventive article. I'm just interested in understanding his design philosophy a bit better.

Charles Hansen
Ocean, N.J.

L. Mirabel responds:

First of all I wish to thank Mr. Hansen for his close and helpful reading of my article.

1) Yes, a dot is missing in Fig. 3 (supply regulators) at the junction of R1/C2/R2/D2.

2) Figure 7 is indeed a stuffing guide. The printed circuit negative was not included in the article. Both the printed circuit and the stuffing guide have not been modified to conform with the final version as shown on the schematic. Neither has been fully implemented, and Fig. 3 *should be considered a draft only.*

The component numbers were taken from the first version of this article. Correction of the discrepancies follows (below).

Note: Pot 4 was not included in the stuffing guide and the printed circuit. It is indispensable, though. For its position and connections, follow the schematic. In the text on page 9, second and third column, substitute R3 for R1, R11 for R7, and P4 for R7A.

AS IN FIG. 3		AS IN FIG. 1	FIG. 3		FIG. 1	
C IN	should be	C1	For RF	READ	R1	
C1	"	C3	R IN	"	R2	
C OUT	"	C4	R1	"	R3	R1A = P1
			R2	"	R4	R2A = P2
			R3	"	R5	R3A = P3
			R5	"	R6	
			R4	"	R7	
			R6	"	R8	
			R8	"	R10	
			R7	"	R11	R7A = P4
			R11	"	R12, R13	

3) I agree that in all likelihood twins would do just as well as two single amps. In fact that is what I have been using in my own implementation, wiring point to point. I did not have an actual printed circuit available but the idea was that two single units might be easier to fit in. The "accuracy" preference was expressed by some writers on the subject but as you state is not a major consideration.

As for the NE5534s and Burr-Brown 604 and 2604 oscillation was a constant problem as soon as amplification or notching were required. Both behaved well at unity gain. Again point to point wiring may have been part of the problem as I said I would welcome others' experience. I had decoupling caps at each chip. Burr-Brown had no suggestions.

4) I found turning multi-turn pots an impossible adjusting chore. At times I needed to adjust four or more pots at a time. Patience would fail me. In addition, when turning and turning it is very hard to determine where you are.

5) Finally, an important warning and an apology: Somewhere along the way I misread the power-supply readings in the Fig. 1 schematic and attached V- to ICs' pin number 4 and V+ to pin number 7. This, of course, is completely wrong—the true positions should be V- to pin number 7 and V+ to pin number 4.

DIRECTIONAL GAIN

I enjoyed G. R. Koonce's article ("Dual-Driver Confusion," *SB* 5/00), but found myself agreeing with many of Bill Fitzmaurice's comments in his letter "DB or not DB" (*SB* 7/00, p. 38). When someone shows me a system that appears to have more power coming out than going in, it trips the same alarm bells that tripped for Bill Fitzmaurice. However, after giving it some thought, I have determined that this is not what G. R. Koonce is saying.

Adding a second voltage source to the resistor network doubles the input power but quadruples power at the load (R3). This can happen because less power is used by R1 and R2 when the voltage increases at R3. Taken as a model, the resistor network suggests that somehow the drivers become more efficient when they work together. After failing to think of a way for the drivers to assist each other, I was about to quit trying to solve this puzzle when a different explanation struck me.

From the little radio work I have done, I had seen this kind of thing before. Antennas have gain by concentrating the field they radiate in a particular direction. I've worked with two antennas connected in what is called a "phased array," which reminds me very much of dual drivers. Both antennas are driven with just the right phase so that the radiated power in the forward direction adds while decreasing in other directions. If you measure the field in front of one an-

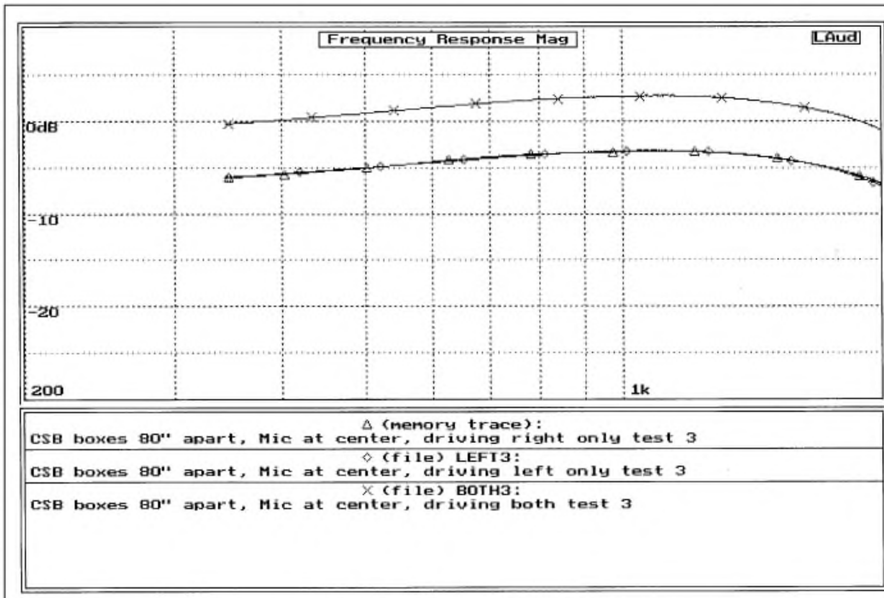


FIGURE 4: Testing dual-driver response.

tenna and measure again when the second antenna is added, you see the power at that point almost double with no increase in power from the transmitter.

With this in mind, I reread G. R. Koonce's article. It is quite clear from his evidence that the "increase in efficiency" is directional and occurs at a focal point on-axis. He states that the effect is due to superposition. Both sound and radio waves obey the superposition principle, so I would expect that what happens with one, happens with the other. If this is true, then the 6dB change at the test mike is the result of a 3dB increase in input power plus 3dB caused by directional gain.

I wrote to G. R. Koonce to ask his opinion. In reply he sent me a plot (*Fig. 4*) showing the test results from two small speaker systems facing each other while 80″ apart. The test mike is centered between the speakers so that each system produces the same response. Quasi-anechoic measurements were taken from 350Hz to 2kHz. The range was restricted because of echoes from the test area at other frequencies.

When both systems are driven, the whole response increases by 6dB. Given the 80″ separation between the speaker systems, it is hard to see how interaction between drivers could improve driver efficiency. I think that this shows that the dual driver 3dB "efficiency increase" happens outside the box and not in it.

I wish that we would stop calling the dual driver phenomenon an "efficiency increase." It misleads people such as myself into looking for a total power gain. I think that the radio term "directional gain" is more descriptive and implies that what you gain in one direction you lose in others.

Pat McChesney
Vancouver, Wash.

SOFTWARE SEARCH

Do you have or know of a crossover design program that imports files from MLSSA, allows steeper than fourth-order targets, and is easier to use than Filtershop/Leap?

Marc W. McCalmont
MACH 1 Acoustics

G. R. Koonce responds:

I really have little knowledge of what software for crossover design is available on the market. I do know that I have not written anything that meets Mr. McCalmont's requirements. I e-mailed the requirements to Ingemar Johansson, the creator of the LspCAD program, and his response follows: "Both versions of LspCAD allow up to eighth-order targets; moreover, the professional version allows you to import a user-defined target. Import from MLSSA (text file format with freq-mag-phase) should be no problem. I still have no minimum phase transformation feature, so I rely on the phase data given by the measurement software. Thanks for spending time on this. I have tried to put a thorough description of LspCAD on my homepage. The direct link to the LspCAD stuff is http://hem.passagen.se/ijdata/lspcad.html."

It is difficult to say how easy to use you'll find a given software product. Various versions of LspCAD can meet the technical requirements of Mr. McCalmont, who may wish to view Mr. Johansson's homepage to select which version is best for his application.

TRANSFORMERS AS CHOKES

In reference to Neal Haight's "Miniature Transformers As Chokes" (*GA* 6/00, p. 63), the major difference between transformers and chokes (other than the multiple windings in a transformer) is that the choke has an air gap to prevent saturation due to the DC current through it. A transformer is not designed to work with DC in its primary winding, and has as its mission the transfer of the maximum amount of AC volt-amperes from primary to secondary.

A magnetic-core filter choke is designed to smooth the output ripple in a DC power supply. It carries all the DC current, and so must be designed with an air gap to prevent saturation. Once a choke saturates, it becomes a resistor whose resistance is that of the wire making up the choke winding. The filter choke thus must be made relatively large to optimize the DC energy stored in the core per unit volume.

This is not to say that a small transformer won't work over a limited range of DC current. However, before committing to the extra space it takes, first check the ripple voltage across the second power-supply cap. Then substitute a resistor of the same value as the DC resistance of the transformer winding. If the ripple is not significantly improved with the transformer, it cannot act as an effective choke.

Furthermore, the copper wire in a choke or transformer has a positive temperature coefficient (tempco) of 0.4%/C°, which will cause the voltage drop to increase as the winding heats up with higher currents and ambient temperatures. A wirewound resistor has a very small tempco, so its resistance is unaffected by temperature.

Chuck Hansen
Ocean, N.J.

Neal Haight responds:

I would like to thank Charles Hansen for taking the time to properly educate me, and all others concerned here.

As Mr. Hansen points out, a miniature power transformer used as a choke will have limited effect, at best, for ripple reduction. In fact, based on the information that he has presented, I recommend that this scheme be used only in a phono preamp or line amp to ensure its effectiveness. However limited it may be, you would do well to limit its use to only preamp circuits, or in chassis where a conventional filter choke just won't fit.

To repeat what I stated in the original article, do not use this method in any type of audio power output circuit. It's also important to note that if adequate filter capacitors are used, you might not even need a choke at all.

HELPFUL FEEDBACK

I may have an answer to Bill Eckle's phase matching problem ("Building an Altec 816 System," *SB* 8/00, p. 30). In order for a null to appear when the high frequency horn is connected electrically out of phase, there must be significant output from both the woofers and the high frequency horn *at the same frequencies*.

I do not believe the Altec N-500-G crossovers have automatic phase correction (although I could be wrong). Certainly the crossover components will affect the phase and could make it difficult to find a well-defined null. Also, the addition of the crossover reduced the amount of frequency overlap of the drivers to the point where there was not much cancellation. Once the crossover was removed, the frequency overlap increased and null appeared.

In any case, Bill has shown that acoustic phasing should be done without the crossover. Thanks for the opportunity to comment.

Jim Forte
Forte Acoustics, Inc.

William Eckle responds:

I thank Mr. Jim Forte for his explanation of my difficulties in achieving phase matching. I have since delivered the cabinets to the owner, who is in the process of designing and building a custom crossover for his system. I was attempting to give him a starting point for his crossover design by finding the correct acoustic phase *position. Thanks again to Mr. Forte for his logical explanation.*

MIDRANGE STUDY

I really enjoyed Jim Moriyasu's two-part article, "A Study in Midrange Enclosures" (*SB* 7/00 and 8/00). He selected several common shapes, some of which stir strong emotions from their advocates or detractors. One of the strengths of the articles is Mr. Moriyasu's willingness to try the good and the bad and let the results speak for themselves.

He concludes, "size of the enclosure is more critical than shape," which is very true. He assumes stuffing the cabinet has no effect on box Q_{TC}. Stuffing causes a heterogeneous blob that perverts the speed of sound. It also affects the "density" of material within the box. A stuffed enclosure is no longer an air enclosure. Stuffing alters Q_{TC}.

He says the smaller enclosures have a .707 Butterworth alignment, which I assume is in the unstuffed state. If Photo 10 demonstrates his usual stuffing strategy, he appears to use polyester packed well over 0.5#/ft^3. Stuffing certainly raised their Q_{TC} over 1.50. It is hard to tell without knowing the packing density and material he used.

He also suggests underdamped boxes with Q_{TC} of .500 are better. His larger boxes, when stuffed, are closer to the Butterworth alignment, but I am sure the enclosures are actually overdamped.

My point is that larger boxes are required as soon as the lightest stuffing is introduced. Larger boxes are needed, but for a different reason.

Rick Schultz
Rschultz@familyconnect.com

James Moriyasu responds:

Thanks for your comments and observations. I must admit I'm still puzzling over some aspects of the study, so your input is welcome.

The Q_{TC} measurements are with the damping material in place. You are right that I was a little sloppy with the amount of damping and did not adhere strictly to 1 lb per ft^3. However, Q_{TC} numbers were pretty consistent with volume. The smaller enclosures were just under .707, while the larger enclosures, except for the anechoic, were under .500.

Perhaps, you could comment on the "residual effect" below 500Hz, which tends to be 1–2dB higher than the unenclosed midrange for the smaller enclosures and around 0.25dB for the larger enclosures. I suppose it is due to the stiffness of the air in the smaller enclosures changing the mechanical Q. These results have also raised the question of what subjective benefits would come from building a sealed box subwoofer with a Q_{TC} of .500 or lower. It would be interesting to compare that alignment to an enclosure with a Q_{TC} of .707.

POWER PARAMETERS

I enjoyed Charles Hansen's review of Jensen's ISO-MAX products (*AE* 6/00, p. 28). I have one question about his suite of hardware. Was all the interconnected equipment powered by one single circuit or by two or more circuits?

Ray Segura
New Orleans, La.

Charles Hansen responds:

I have all my audio equipment on one 120V AC 20A dedicated circuit, and all the equip-

ment is plugged into four Hubbell hospital-grade (orange) duplex outlets mounted in one steel four-gang electrical box.

HELP WANTED

I have been searching everywhere for information/parameters on a pair of 6″ model LW 6004 PMR mid/woofers. Its box states, "Honorably designed in France for Dynavox." I have looked on the web and found a site for Dynavox Electronics, but found nothing like these in their speaker section.

They have machined 5mm-thick mounting rings, a large 75mm voice coil, flat spiders that are vented, large magnet structures with a vented pole piece, and rubber surrounds. They are rated as 8Ω and look similar to Dynaudio cones.

Any information you can supply would be greatly appreciated.

Edward Lewis
edntamlewis@email.msn.com

I have been a subscriber to *The Audio Amateur*, (then *Audio Electronics*, now *audioXpress*), since the early to mid-1970s.

I have obtained one of the late Dan Meyer's SWTPC 275 amplifiers (I've built several of his amps, along with several projects from Old Colony Sound Lab, but I didn't build this one). I would like to know some possible sources of a circuit diagram, preferably with component values. I don't remember this particular amplifier being written up in any periodicals at the time...I think I must have read them all! It puts out about 70W into 8Ω, and, on examination of the circuit board, it appears that it may have a complementary differential amplifier input, like Meyer's Tiger .01 and Tigersaurus, and many others since.

If anyone can steer me in the right direction, I would appreciate it.

Glen L. Orr
glenlorr@essex1.com

Back around 1993 I built two pairs of speakers from a design published in *Speaker Builder*. The author later provided positive feedback regarding the addition of Norsorex gaskets around the drivers. I found a 1992 issue in which Polydax advertised these gaskets in various sizes. They seem to be defunct now. Do you know the trade name for this material, or can you point me toward a source of similar gaskets?

Fred Trusell
Colorado Springs, Colo.

*The company name is now Audax of America at www.audax.com.—**Ed.***

I was told recently by an audio expert that transistor amps were much better than tube amps. The tube amps—only by limitation of frequency and distortion—were able to mask the sound of a transistor-type amplifier, and so made the sound appear to be more mellow. What is the truth?

Rusyl Hillstrom
PO Box 325
Scobey, MT 59263

Calling ZIP Codes 228-, 229-, 244—

I'm looking for someone to collaborate on the occasional loudspeaker project with me. I live in a less-than-urban part of Virginia and have long had hopes of running into another hobbyist, but no such luck. So if you're a compulsive loudspeaker (or valve) builder, especially if you're into the more involved testing aspects, drop me a note.

Tom Yeago
PO Box 713
Staunton, VA 24402

I have a 12″ audiopipe sub, and need to construct an enclosure for a certain Q_{TC}, but the manufacturer only supplied these parameters: frequency response = 20Hz–700Hz, SPL = dB, f_S = 25Hz, and maximum power = 750W. However, to construct an enclosure for a given Q_{TC}, you need the following parameters: f_S, Q_{TS}, and V_{AS}. Can anyone offer suggestions on how to calculate the Q_{TS} and V_{AS} using the parameters that the manufacturer supplied?

Andrew Hunter
Andrewcty@yahoo.com ❖

New Chips on the Block

Apogee DDX-2000/2060 Direct Digital Amplification Chipset

By Charles Hansen

Apogee's Direct Digital Amplification (DDX®) is a high-efficiency, all-digital amplifier technology designed to meet the needs of digital audio systems ranging from PC multimedia to home-theater systems. Apogee provides DDX amplification solutions as semiconductor products and OEM board products, and through technology licensing. The DDX-2000/2060 amplifier chipset consists of the DDX-2000 controller and DDX-2060 power device and utilizes Apogee's DDX technology to provide over 30W per channel of high-quality audio power without the need for an external heatsink.

The DDX chipset directly interfaces with digital audio sources such as DVD and MP3 players. Since DDX completely eliminates the requirement for a digital to analog converter (DAC) and external heatsink, it enables manufacturers to develop compact state-of-the-art products that deliver digital sound production from source to speaker.

TABLE 1
DDX-2000/2060 AMPLIFIER PERFORMANCE

PARAMETER	PERFORMANCE
Power output (W, RL = 8Ω)	
THD = 1%	42
THD = 10%	52
Efficiency (%)	
1W	50
30% FS	84
FS	89
THD+N (%)	
1W, 1kHz (typical)	0.08
Max (20Hz–20kHz)	0.20
IMD (%, 19kHz + 20kHz, 1:1 IHF)	0.13
SNR (dB A-weighted)	93
Frequency response	20Hz–20kHz
PSRR/SVR (dB, Vr = 0.5V, fr = 100Hz)	60
Crosstalk (dB, 0dB = 1W, f = 1kHz)	70

The DDX-2000/2060 chipset is ideal for products such as digital speakers, PC sound cards, MP3 playback devices, surround sound systems, or any application requiring front-to-back digital sound reproduction. In addition, the high-efficiency design is perfect for portable devices in which extended battery life is important.

Apogee's patented Direct Digital Amplification (DDX) technology uses advanced signal processing to convert digital audio data into Apogee's damped ternary or three-state modulation. This digital output controls the switching of high-efficiency power transistors to produce high-quality audio power.

Ordinary switching amplifier designs use analog-controlled two-state modulation, which continually modulates the power to the speaker. In contrast, DDX's damped ternary modulation is digitally controlled and connects the speaker to the power supply only when an output is needed. When no output is required, the speaker is connected to ground or in the damped state. This approach reduces power consumption by up to 20% over two-state switching amplifier designs, and more than 300% over conventional analog amplifiers when producing music signals. In addition, it eliminates amplifier idle noise and "pops and clicks" during power-on and power-off.

The DDX-2000/2060 stereo amplifier chipset interfaces directly with digital audio sources and provides over 30W per channel audio power into an 8Ω load at very low distortion. The DDX-2000 controller features digital volume control, automatic muting, support for multiple audio serial interfaces, and specialized digital processing to reduce distortion associated with signal clipping. The DDX-2060 power device's surface-mount package eliminates the need for an external heatsink. Over 88% efficient, the power device includes a power-down mode and short-circuit, over-voltage, and thermal-protection circuitry with automatic recovery. The chipset can also be configured in a mono mode to provide over 60W into a 4Ω load.

PRICE AND DELIVERY

The DDX-2000/2060 chipset is priced at $6.98 in 1000-piece quantities. Apogee has also released the EB-2060x evaluation board, which enables manufacturers to quickly hear and evaluate the advantages of the DDX-2000/2060 chipset. Mono and stereo versions of the board are available immediately for $198. For further information, contact:

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TI Digital Audio System Chip Set

By Charles Hansen

Texas Instruments has announced a fourth-generation DSP chip and the industry's first complete, fully available digital audio system. This true digital audio amplifier keeps the audio signal in a digital format from the source to the speaker. By combining TI's fully Digital Audio Amplifier with its DSP technology, this system has solved the "noise and hum" problems associated with today's conventional units. Besides being ideal for home-theater-in-a-box (HTIB) systems, TI's digital-audio technology also helps bring the highest sound quality to other consumer and automotive entertainment devices, such as digital headsets, PC and car audio systems, and traditional home audio mini-stereo systems.

Based on hardware and software components that are all available today from TI, the heart of the system is TI's DSP technology. When applications require decoding or additional audio processing, this system uses a programmable TMS320DA250™ DSP and strong portfolio of algorithms to provide digital audio decoding at the speaker. This enables the host system to send digital data directly to the speakers from a CD, DVD, TV, game console, the Internet, or other sources. By having a digital connection, designers avoid the power and signal losses associated with transmitting analog signals in today's audio systems.

The DSP also provides audio processing for industry-standard formats, such as Dolby Digital 5.1, 3-D Surround Sound, DTS, MP3, Qdesign QDX, DRM technologies, and Advanced Audio Coding (AAC). Based on the TMS320C55x battery core, it extends battery life up to 70% longer than other DSPs. The DA250 offers the first dual multiply and accumulate chip (MAC) on a DSP for Internet Audio, and the first chip for portable devices to embed USB capabilities. This DSP supports Secure Digital (SD), Memory Stick, Compact Flash, Smart Media, and Multi-Media Card (MMC) devices.

For applications where high-quality post-processing is required, TI has developed the TAS300x family of dedicated digital-audio processors that run advanced on-chip filtering algorithms, allowing on-the-fly configurable parametric equalization and speaker-response correction. Designers can currently use the TAS3001 to perform these tasks together with dynamic range compression/expansion and volume, bass, and treble control. TI also offers the new

TAS3002, which is a digital audio processor that performs the same functions as the TAS3001, but features an integrated 24-bit codec to help reduce board-space requirements and lower design complexity.

For interfacing with the PC, designers can use the TUSB3200 multichannel USB Streaming Controller that allows digital audio systems to connect bidirectionally to host systems via the USB used in almost all PCs and Macs. In addition, this technology supports the S/PDIF digital interfacing standard used in many consumer systems with the DIR1701 S/PDIF receiver. Scheduled to be available in the first quarter of 2001, this state-of-the-art device from TI's Burr-Brown product line allows the stable recovery of digital audio data from all S/PDIF sources, such as DVD, CD, and MD players.

TRUE DIGITAL AMPLIFICATION IS HERE TODAY!

The final piece of the digital audio solution is TI's new true digital audio amplifier, the TAS5xxx chipset family based on the equibit™ technology from Toccata Technology, a company that TI acquired in March 2000. Consisting of a TAS5000 pulse-width modulator (PWM) and two TAS5100 H-bridges, this surface-mount-technology (SMT) chipset drives speakers digitally for greater power efficiency. The TAS5000 digital modulator takes I^2S, 44.1-96kHz, 16- to 24-bit data streams and processes the incoming Pulse Code Modulation (PCM) data into PWM digital data using state-of-the-art modulation techniques. The TAS5100 digital power stage is fed by the digital PWM signal directly from the TAS5000. The TAS5100 then converts the PWM signal into the desired digital power level at the speaker terminals, and the conversion into an analog signal takes place either in the speaker coil itself or is done by a simple LC filter.

Other key benefits include surface mounting with no need for heatsinks or other cooling devices, even when supplying 25W per channel (continuous) digital sound power. By eliminating bulky heatsinks, the TAS5xxx chipset greatly reduces the form factor size and power consumption by up to a factor of ten.

In addition, a key challenge in using digital audio amplifiers is the power management design. TI has a broad portfolio of high-performance power ICs that help designers address their power requirements while offering the necessary support to help speed time-to-market.

APPLICATIONS IN ALL AUDIO AREAS

Many types of audio applications can benefit from the high quality, power efficiency, and space savings in TI's all-digital audio solution. These include consumer entertainment systems, such as HTIB, DVD receivers and mini/micro systems; gaming and entertainment applications; business applications, such as speakerphones and multimedia conferencing systems; and digital headsets for all application areas. In addition, TI's all-digital audio solution will enhance automotive entertainment, reducing space and heat dissipation for manufacturers, while allowing consumers to enjoy the experience of high-quality multichannel digital audio in their vehicles.

AVAILABLE TODAY

TI's complete package is supported by thorough design documentation and test results documenting FCC electromagnetic compatibility (EMC). Audio designers can order—in high volumes only—directly from TI at the following prices:

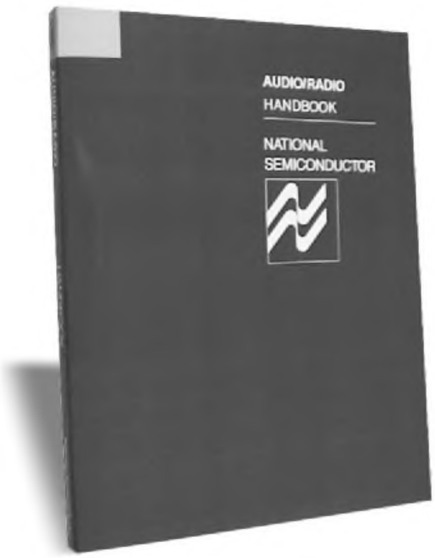
TMS320DA250 family of DSPs (qty. 250,000), $10
TAS3001 digital audio processor (qty. 50,000), $1.35
TAS3002 digital audio processor with integrated codec (qty. 50,000), $2.98
Digital Audio Amplifier chipset; TAS5000 and two TAS5100s (qty. 50,000), $8.95
TUSB3200 USB interface chip (qty. 50,000), $4.89
DIR1701 S/PDIF receiver (qty. 50,000), $2.73

For more information, see www.ti.com/sc/digitalaudio and www.ti.com/sc/msds5397u. Also, contact Texas Instruments Semiconductor Group, 12500 TI Blvd., Dallas, TX 75243, (800) 336-5236. ❖

Book Review

National Semiconductor Audio/Radio Handbook

Reviewed by Charles Hansen



National Semiconductor Audio/Radio Handbook *reprint, $14.95. Old Colony Sound Lab, 305 Union St., PO Box 876, Peterborough, NH 03458-0876, 603-924-6371, FAX 603-924-9467, E-mail custserv@audioXpress.com.*

As many of you know, Publisher Ed Dell has a single-minded passion to preserve and reprint those important published works particular to the field of audio. Few people are aware of the *National Semiconductor 1977 Audio Handbook*. I had a copy and suggested to Ed that it might make a good reprint candidate. I touched base with Bob Pease, staff scientist at NSC, to find whom to contact to secure the rights to reproduce the *Handbook*.

UPDATE

NSC had difficulty finding a clean copy of the *1977 Audio Handbook*, so I sent Ed my well-used dog-eared copy. While a number of pages had transparent-tape tabs on them for quick reference, I did manage to erase all the penciled notes. Fortunately, Sharon Nowling at NSC discovered that the *Handbook* had been updated in 1980, and they had a pristine copy for the camera.

As with the 1977 version, the *Handbook* has as its primary emphasis the application of operational amplifiers to audio circuits. The radio section existed in the '77 issue and "Radio" was added to the book title in 1980. Thus the 1980 edition consists mainly of minor updates and corrections, and a few (then) new ICs.

Chapter One is the introduction, which explains the scope of the *Handbook* and explains in detail the IC parameters applied to audio: slew rate, open-loop gain, bandwidth, gain-bandwidth, noise, total harmonic distortion (THD), supply voltage, and power-supply ripple rejection.

PREAMPS

Chapter Two covers the design of preamplifiers. The first few sections cover ground loops, single-point grounding, and supply bypassing, with additional stability tips. A section on noise (thermal noise, noise bandwidth, 1/f noise, and "popcorn" noise) is included. Noise modeling discusses the noise sources within an op amp and explains voltage and current noise specifications (e_n and i_n) in both RMS terms and the familiar $nV/\sqrt{Hz}$ and $pA/\sqrt{Hz}$ parameters used to determine noise over a specified bandwidth. This section also covers the effects of ideal and practical feedback on noise, with equations and real-world circuit examples.

The *Handbook* includes a section on mitigating the problem of radio frequency interference (RFI), to which many early phono circuits were susceptible. This chapter also contains a section on noise-measurement techniques, signal-to-noise (S/N) ratio, and the use of weighting filters. National produced a line of low-noise audio IC preamps in the '70s, so the application of the LM381, LM382, LM387, and LM1303 are covered in great detail in this chapter.

The next topic is the phono preamp. The authors provide a detailed explanation of the vinyl stereo recording process and RIAA playback equalization (EQ). They even include historical information on low-fi ceramic and crystal cartridges. While constant velocity moving coil and moving-magnet cartridges require equalization components, these high-output constant amplitude transducers needed neither a preamp nor EQ circuits.

The section on tape preamps covers the magnetic tape recording/reproduction process and the NAB tape playback EQ curve. The section provides graphs of the effects of tape speed and head gap on frequency response, and describes how to calculate gap loss.

Microphone preamplifiers are classified in the book as high-impedance and low-impedance types. The low-impedance circuit designs are presented for transformer-coupled and transformerless, balanced and unbalanced.

The next section covers tone controls, both passive and active, with a detailed explanation of the feedback response effects of active bass, treble, and midrange tone control circuits. The section shows one design for a loudness contour circuit, used to compensate for the logarithmic nature of human hearing. This circuit compensates for the pronounced loss of bass response and slight loss of treble response as the volume level decreases.

Early audio equipment also had controls to help reduce the annoyances of scratched records and the poor bearing designs of inexpensive record changers. The low-pass scratch filter removed high-frequency noise and ticks from worn records (along with much of the treble information). The high-pass rumble filter removed low-frequency noise due to turntable support bearings and AC line hum (along with the extended bass information). These circuits are presented here, along with a bandpass speech filter (300Hz–3kHz).

Next in the preamp chapter is the bandpass active filter and its application to the graphic or octave equalizer. A pink-noise generator circuit is provided, to make the most of setting up the equalizer in a listening room.

The pro audio field is covered in the form of mixer, panning, and noiseless switching circuits. A section discusses current-boosted op amps, showing both discrete transistor and the current booster LH0002 IC amplifier to drive low-impedance audio lines.

RADIO

Chapter Three is the radio section, covering amplitude modulation (AM), frequency modulation (FM), and FM stereo multiplexing. AM radio requires that the antenna RF signal be converted to a useful voltage with ferrite loop or capacitive auto antennas. Since NSC was in the radio chip business, the book introduces the LM3820 AM receiver IC and explains RF amplification, intermediate frequency (IF) conversion, and automatic gain control (AGC).

The FM radio section discusses IF amplifiers and FM detection and introduces the LM3089 and LM3189 FM IF/detector ICs and the LM1310 and LM1800 stereo demodulator ICs. These integrated circuits provide IF amplification, FM stereo quadrature detection, automatic frequency control (AFC), and audio and mute control amplifiers. The section gives the reader application schematics as well as RF PC board layout tips. Lastly, the subject of stereo with mono blend for improved S/N ratio is introduced along with the LM4500A and LM1870 FM blend demodulator/decoder ICs.

POWER DEVICES

Chapter Four is concerned with power amplification, with the main emphasis on NSC's power ICs. This was well before the LM3875 56W Overture® power amp IC was introduced, so output power is limited to less than 10W (at 10% THD). Topics covered in this chapter include frequency response, THD, slew rate, crossover distortion, output stage topology, bootstrapping, and output stage protection circuitry. The chapter also includes nice design tips on layout, ground loops, supply bypassing, and stability.

The ICs presented are dual and mono DIP amplifiers in the LM38x series, and the LM1877, LM2000, and LM2877. A section shows how to use the dual amplifier chips in mono bridge configuration to provide higher power to a floating load. This chapter also contains a number of non-hi-fi applications, such as a Wien bridge power oscillator, square-wave oscillator, intercom, power converter, and motor controller circuits.

Boosted power amps with discrete transistor output stages (up to 100W) are also covered in some detail. The subject of safe operating area (SOA) for the output devices and thermal runaway is introduced, since, unlike the output stages of the IC, the discrete output devices are not self-protected.

Finally, no discussion of power devices is complete without heatsinking, so the *Handbook* includes a nice section on this topic. The authors show you how to model heat flow and how to determine the thermal resistances of devices, packages, heatsinks, insulators, and thermal greases. They include procedures for selecting and using heatsinks, simple rules for their application, and lots of information on heatsink surface finishes and materials. Finally, a section discusses how to use PC board foil as an effective heatsink.

FLOOBYDUST

Chapter Five is called "Floobydust," a topic familiar to anyone who has read Bob Pease's books. This mixed bag of miscellaneous topics not covered elsewhere in the *Handbook* covers biamplification, active crossovers, reverb, phase shifters, tremolo, acoustic guitar pickup preamps, transconductance amps, and noise-reduction circuits.

In the last chapter, the Appendices, section A1 discusses audio power-supply design in great detail. You are guided through diode and transformer selection, capacitor and L-C filter design, capacitor selection, transient protection, linear regulators, and voltage doublers. A number of very useful nomograms shows peak diode power dissipation versus RMS ratings, covering various half-wave and full-wave rectification methods.

Section A2 covers decibel conversion, and A3 shows wye-delta network transformation techniques for circuit analysis. Section A4 discusses the standard amplifier building-blocks (non-inverting, inverting, inverting-summing, buffer, and difference).

Section A5 reviews noise basics and analyzes phono cartridge noise and impedance. Section A6 is an application selection chart for NSC's line of FET and bipolar input audio ICs (circa 1980). Section A7 discusses feedback resistors and amplifier noise.

Section A8 talks about reliability and operating temperatures, and Section 9 is the eight-page audio/radio glossary. A complete index follows, with several blank pages for taking notes.

CONCLUSIONS

The *National Semiconductor 1980 Audic/Radio Handbook* offers a surprisingly thorough treatment of audio technology as it was 20 or more years ago. You won't find anything here on digital audio, CDs, MP3, or DAT. The topics that are covered are done well, with both practical overviews and all the mathematical detail you need to design high-quality circuits. ❖

Glass Shard

Another Subminiature Preamp

Eric Barbour's recent article, "A Pocket Preamp,"[1] was very engaging. For several years I have been interested in the smaller, subminiature tubes Barbour discussed from a cost, availability and familiarity point of view.

From my background in nuclear and satellite electronics, I have previously been exposed to, and worked with these subminiature tubes and have found them to be very reliable, low microphonic, and fairly low in noise. I hadn't thought of them with respect to sound until about a year and a half ago when the company I work for cleared out some of their old inventory of discontinued parts. I acquired about a hundred "on the cheap" and have constructed a most satisfying line-stage preamplifier.

DESIGN INFLUENCES

My preamp is based on a differential input stage direct coupled to a cathode-follower output stage (*Fig. 1*). This is not very fancy, since I like to keep my layouts clean and simple. *Table 1* lists the component values.

My line-stage design evolved from a circuit[2] I discovered while reading some older issues of IRE Transactions on Audio. At the time, I was developing a phonograph preamplifier and was looking for a good output-stage buffer. I toyed with several tube operational-amplifier-based output stages,[3] but desired something a bit simpler which required lower output-voltage supplies.

I read two articles by J. Ross MacDonald[2,4] about augmented cathode-follower amplifiers, which were typically used in equalization circuits or low-noise preamplifier circuits, and decided to pursue this approach. I have also read several articles about cathode followers[5,6,7,8] and their attendant variations such as the White Follower[9,10], and tried several of them. I liked MacDonald's approach for its simplicity in parts count and overall sound. In its most basic form, it is composed of a differential pair followed by a cathode follower.

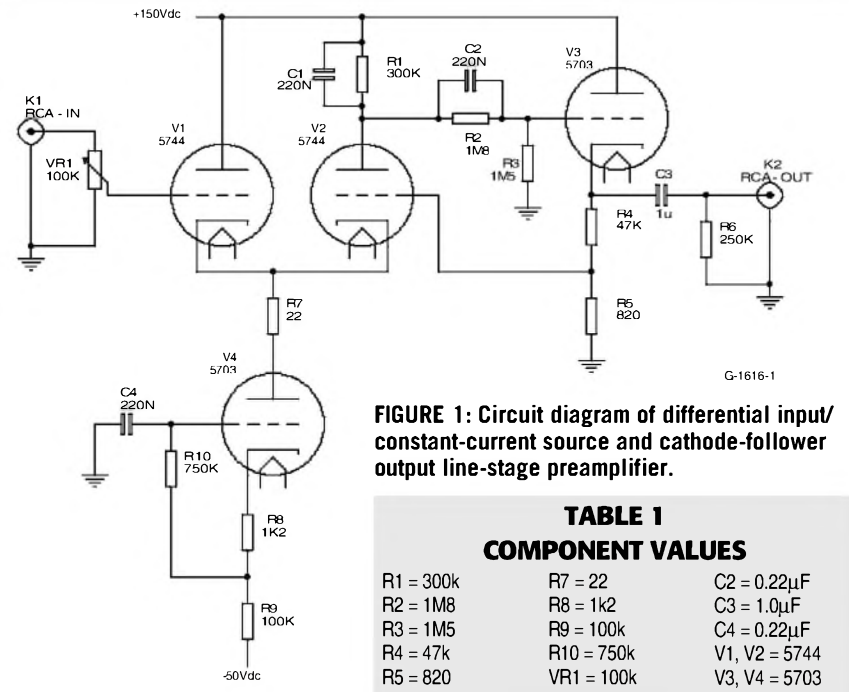


FIGURE 1: Circuit diagram of differential input/ constant-current source and cathode-follower output line-stage preamplifier.

TABLE 1
COMPONENT VALUES

R1 = 300k	R7 = 22	C2 = 0.22µF
R2 = 1M8	R8 = 1k2	C3 = 1.0µF
R3 = 1M5	R9 = 100k	C4 = 0.22µF
R4 = 47k	R10 = 750k	V1, V2 = 5744
R5 = 820	VR1 = 100k	V3, V4 = 5703
R6 = 250k	C1 = 0.22µF	

In my application, I selected the 5744, mu = 70; and 5703, mu = 20, subminiature "JAN" grade tubes, manufactured by Raytheon. These tubes conveniently have flying-lead connections, which makes it very simple to plug into a standard breadboard. I hooked up a test circuit on the bench and tweaked it for a couple of days. I was able to obtain a reasonable amount of gain, 25dB, at a decent total harmonic distortion, <0.1%. My test-bench listening test produced no overall harshness or mushy lows, so I decided to build it.

PREAMP CONSTRUCTION

After the Barbour article in *Glass Audio*,[1] I changed my plans a bit and decided to make it smaller. My final construction was on ⅛″ phenolic board, drilled and tapped for double-turret-style terminals (Keystone part no. 1587-2). I have settled on this manufacturing process over several years of designing and building electronics. These types of terminals are more robust and easier to change, and enable easier last-minute revisions, than printed circuit boards. Best of all, they require no lifted pads and no messy chemicals. Additionally, it is conceptually easier to trace out where all your wires go. The only down side is that you must be careful with your layout as the wiring is point-to-point, and signals have a way of ending up where they shouldn't. I have generally found this not to be a problem at audio frequencies. Of more importance is adequate shielding from all the excess RF noise generated by computers, radios, and other electronic devices.

The first stage of the line-stage pre-

TEST EQUIPMENT USED

1. Hewlett-Packard HP35670A Dynamic Signal Analyzer (THD)
2. Tektronix TDS210 Oscilloscope (circuit breadboarding, square wave and THD)
3. Wavetek 111 Signal Generator, 50Ω output impedance terminated at preamplifier with 50Ω (square wave)
4. Neutrik Minirator MR1 Audio Generator, 200Ω output impedance (circuit breadboarding)
5. Hewlett-Packard HP6209B, HP6206B, HP6214C power supplies (circuit breadboarding)
6. Final power supplies are in progress

amplifier is formed by a differential input[6,7,11] composed of two 5744 subminiature tubes, V1 and V2, which have a relatively high mu of 70. Gain is controlled by the load resistor, R1, shunted by a high-frequency bypass capacitor, C1, to prevent any excessive high-frequency ringing.

I should note that I didn't have a ringing problem while using the test board due to the tubes' leads and the test board's capacitance. I noticed it when conducting final tests using square-wave[12,13] input test signals. The ringing is actually a large overshoot.

The C1 capacitor reduces the overshoot and also slightly reduces the amount of third harmonic signal component, while it increases the second harmonic. Tube V4 provides a constant-current source cathode load to the differential pair. I set the combined current for V1 and V2 at approximately 1mA.

The output from the differential pair goes to a standard cathode follower, with a tap returning a small amount of feedback to the first stage's differential pair inverting input. Gain, as well as distortion, is highly dependent on where this is selected. Resistors R2 and R3 set up the bias condition for the cathode follower and capacitor C3 isolates/prevents the static DC from being applied to the main amplifier.

Overall, at full gain of 24dB, THD over the first three harmonics measured less than 0.1% from 20Hz to 15kHz. At 1kHz, THD was approximately 0.054%. At a gain of 3dB, 1kHz THD rose slightly to 0.063%. Gain values were all measured with reference to a 100mV RMS input sine wave.

NOTES

1. Allow heaters time to warm up before applying high voltage to the circuit.
2. A large, 50V pk, slow voltage spike occurs when high voltage is applied to the circuit due to charging of the output coupling capacitor. Take appropriate precautions before connecting the unit to your power amplifier. ❖

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Test Tracks

Here are my favorite audio test tracks for evaluating various audio systems.

1. Blues Traveler, *Four*, A&M Records, 31454 0265 2. This is one of the best rock-and-roll albums of the 1990s. The first track, "Run-Around," explodes with a powerful but smooth bass line and clean, steady persuasion. The tambourine and guitar-strumming really stand out on a detailed playback. The signature harmonica playing of bandleader John Popper—particularly on "Fallible," "Crash and Burn," and "Hook"—is exciting and really pushes the high-end response of speakers.

2. Joni Mitchell, *Blue*, DCC Compact Classics, GZS-1132. This is a remastered version of the 1971 reprise album pressed on a so-called gold CD. (I remain neutral on the gold CD question, but am quite happy with the remaster.) This is my favorite female vocal CD.

Mitchell's voice varies from a trebly soprano to a sultry alto, but is always warm and comforting. Many of the tracks—particularly "Carey" and "California"—feature acoustic guitars and dulcimer to form a surprisingly large soundstage. For each improvement I have made to my stereo, the track "This Flight Tonight" has become more ominous-sounding.

3. Blue Merles, *Music of Known Hooligans*, Second Fret MKH2NDF. The Blue Merles are a jazz/bluegrass ensemble, and *Hooligans* demonstrates their great picking and fiddling. This is the first album I go to for evaluating detail. The percussive quality of the string instruments really comes out. The sounds of picks and fingers touching the strings on the mandolin, guitar, and banjo are natural and clean.

The fifth track, "Take 5," is a prime example of the natural, but not rough, recording. Each instrument is clearly placed on a small café-like soundstage. The breathing of the guitar player is faintly heard in the background.

4. Paul Simon, *Graceland*, Warner Brothers, 9 46430-2. Even ignoring the cultural significance of this album, *Graceland* is one of the most important and influential projects in modern history. The songs, performance, and production are all superb. Most of the songs have a warm, melodic electric bass that is superimposed by complex percussion, sweet guitar, and the well-known vocal style of Paul Simon. These songs highlight the ability to reproduce music in detail.

Every song on the album is a stage for male vocal reproduction. My experience in listening to this CD is that the better the system, the more individual instruments and voices can be clearly differentiated and placed on the sound stage.

5. The Beatles, *Abbey Road*, EMI CDP 7 464462. This is a prime example of how "vintage" recordings, with all their sonic negatives (noise, distortion, and so forth), can still be superior to those made in multi-million-dollar digital-recording studios. This is an album that I prefer to listen to from start to finish, because it is a continuous ride of ebbing and flowing texture, sound-staging, and all-around fantastic sound. It is difficult to be analytical about any of the individual songs, which flawlessly flow from one to the next. The richness of some of these songs, such as the bass line in the opener "Come Together," can be overwhelming.

There are also many individually unique electric and acoustic guitar sounds, each projecting a unique impression from the speakers. The eighth track, "Because," is one of the most sonically interesting recordings in my library. It is a mixture of synthesized harpsichord, horn, and other sounds, with piles of psychedelic choral voices overlaid. This album is full of "experimental" recording techniques that are used with restraint and taste. One example is the graceful left-right panning of guitar on "Sun King."

6. Claudio Monteverdi, Stepen Stubbs, Cond. *Madrigali Concertati: Tragicomdedia*, Teldec 4509-91971-2. Stubbs does a beautiful job in recreating the madrigal music of Monteverdi. The album is performed by a small male vocal ensemble accompanied by a minimal arrangement of period (16^{th}–17^{th} century) instruments. The performance space can be felt in the very natural reverberation of the voices.

The crispness of the Italian syllables produces a sense of reality of the singers. It also helps that I do not understand Italian, so I can appreciate this strictly on the sound, not on the meaning of the words! (The liner notes are translated into English, but for some reason the translation is left only in German.)

7. Erich Kunzel, Cond., Cincinnati Pops Orchestra, *The Great Fantasy Adventure Album*, Telarc CD-80342. This is a CD designed to show off home-theater sound systems. I do not have a surround-sound system, but still appreciate the grandeur of these movie-score recordings. The CD also boasts 20-bit surround-sound recording (of which I cannot currently take advantage).

The pieces all have enormous dynamic range and sound great played loudly. Horns and tympani in "*Hook*: Main Title Theme," and others, swell and punch. The choir in "*The Abyss*: End Titles" is so moving, that you will want to see the end of the movie from which it came. The ominous music and rumbling at the beginning of the eighteenth track, "*The Hunt for the Red October*: Hymn to Red October," really gives you the feeling that you are in a doomed submarine. This is then followed by great Russian choral music.

This disc is also peppered with some movie sounds (dinosaurs, shooting arrows, and so on) that may not be musically significant, but are kind of fun. Overall, this is a great album that pays tribute to the composers of music scores, including John Williams, Danny Elfman, Mark Knopfler, and many others. The music stands on its own without the movies they accompany. ❖

Stuart Rubin
Philadelphia, Pa.

Classifieds

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The LP track carries both stereo channels vs. alternating pits on the CD. The LP remains superior with "tight coupling for audio," $15; using an "OTL amplifier circuit," $35; driving efficient "Bass Horn" plans, $50. Visa, MC. ***STEEN AUDIO LAB***, PO Box 2185, Vancouver, BC V6B 3V7, CANADA.

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This rare schematic diagram of a Marantz Audio Consolette control preamplifier-equalizer is courtesy of C. Victor Campos. ❖