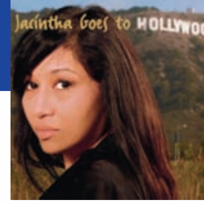
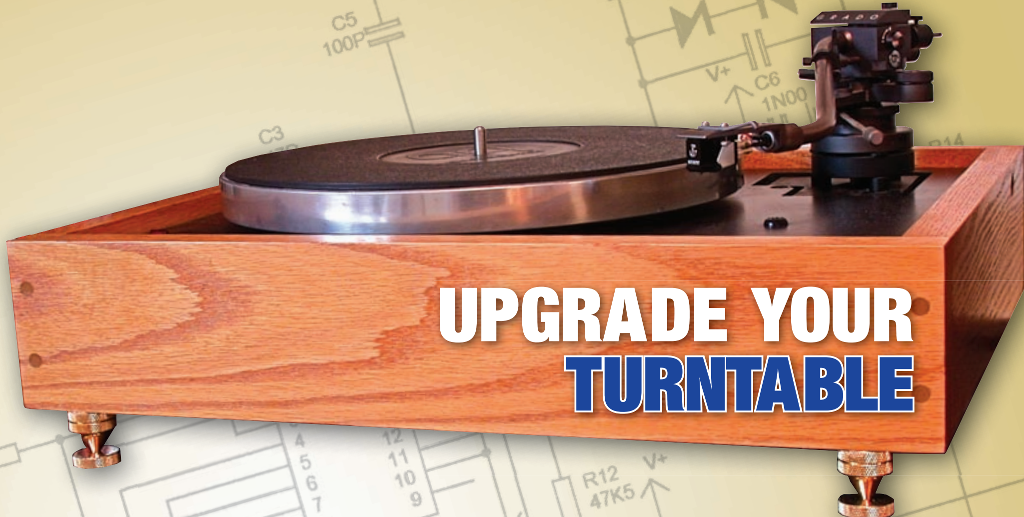


July 2008
US $7.00/Canada $10.00

audioXPRESS

Tube, Solid State,
Loudspeaker Technology

Versatile Filter For Room/Speaker Analysis



UPGRADE YOUR TURNTABLE

IMPROVED HYBRID AMP DRIVER CIRCUIT

BUILDING A PHASE-LINEAR SPEAKER

MEET SYSTEMS DESIGNER EDUARDO DE LIMA

IN SEARCH OF VINTAGE AUDIO TREASURES

A SIMPLE WAY TO FIGURE TUBE DISTORTION

07
0 74470 84697 0

By Dennis Colin

A Multimode Filter/Resonator for the Sound Strobe and Other Audio Applications

This highpass/bandpass/lowpass filter has a tunable fo (20Hz-20kHz) and a wide Q range (0.5-100) for audio filtering, resonant tone pulse generation, auditory music analysis, measurements, and more.

The Sound Strobe (*aX* March '06, and reviewed by Ed Simon in *aX* April '07) generates six selectable-spectrum (but wideband) pulses that are very revealing of speaker transient precision and in-room sound clarity. The tunable, variable resonance filter I describe here allows focusing the pulses' spectrum into a bandwidth (BW) ranging from 2.5 octaves to less than 1/48 octave (1/4 semitone). Then, with octave switching or a 20Hz – 20kHz sweep control of fo, you can hear transient response, frequency response smoothness, image focus, and room acoustics effects, as a function of frequency. (fo is the bandpass output's center frequency.)

With high Q (resonance) settings, the Sound Strobe pulses produce, at the bandpass (BP) output, an exponentially-decaying sinusoidal tone burst, at the filter's fo. These start with a clean and precise onset, producing a sound that resembles musical tone transients. Because of this, you hear the characteristics of speakers and room colorations with recognizable correlation to musical reproduction quality.

"AUTO Q" MODE

In addition to a manual mode (where you can adjust Q from 0.5 to 30), there's an "Auto" mode. Here, the Q increases with the fo setting, ranging from 7.3 at 20Hz to over 100 at 20kHz. Used with the Sound Strobe's "25ms exponential" pulse (a full-band step pulse rolled off below 6Hz), the filter's BP output has sufficient resonant decay time to hear a distinct tonality, while being short enough to hear transient-response details. As you tune fo, the pulse sound has these characteristics:

1. fo range of 20-50Hz: resembles bass drums of various sizes.
2. 40-300Hz: the fundamental tones of a plucked string bass.
3. 300Hz-1kHz: chimes and bell transients, somewhat guitar-like.
4. 1kHz-3kHz: the metallic transients of small bells, xylophones.
5. Above 3kHz: high-pitched transients such as small triangles.

As you tune fo over the full audio band, a multitude of speaker and room anomalies can be revealed. Also, slowly sweeping fo allows you to hear the sound system's coherence and smoothness (or lack thereof) of frequency response. And, as with the raw Sound Strobe pulses, you can easily discriminate the direct speaker sound from room effects. But with this filter, you can "hone in" on any specific frequency.

A switchable +1.82dB/octave slope equalizer (EQ), when used with the Sound Strobe 25ms exponential pulse and the filter's "Auto" Q mode, maintains a constant energy per pulse regardless of fo.

These precise-transient exponential tone bursts, as their fo is swept, will reveal many details regarding overall speaker sound clarity, timing precision, and spatial image focus.

OTHER FILTER USES

1. Auditory spectral analysis: Playing music through the filter (BP output) can be informative in several ways. With the Q (manual mode) set to about 4, the -3dB BW is about 1/3 octave (4.31 is the exact Q value). This is useful to explore the frequency range of various musical sources; with higher Q settings you can select individual instrument tone harmonics.

   But perhaps the greatest use is in identifying the frequency ranges where there's some coloration or distortion that you've previously noticed. For this purpose, it's useful to feed one of the stereo channels through the filter, so you can tune the fo while also hearing the full-band sound from the other channel.
2. You can use the highpass (HP) and lowpass (LP) outputs with music to hear the effects of bandwidth restriction, and to determine the bass depth and HF extension in a particular piece of music. You should set the Q to 0.7, because this produces a Butterworth rolloff (maximally flat without peaking) of 12dB/octave.
3. Noise measurements: The HP and LP filtering can be useful in determining the noise voltage above or below some chosen frequency. The HP output is also useful to attenuate AC line hum.
4. Notch filter: A notch (about 40dB deep) response is easily generated by summing the LP and HP outputs with two 1k resistors. A low Q produces a wide notch, and higher Q values make the notch narrower. The passband voltage gain will be 0.5.
5. Any general-purpose applications where tunable filtering is needed.
6. Spectrum analysis: You can use the BP output with an AC voltmeter to measure audio waveform levels in tunable frequency bands.
7. Crossover design: Using the LP and HP outputs, with one inverted and Q set to 0.5, lets you experiment with CO frequency with a tunable second-order CO.

STATE VARIABLE FILTER BACKGROUND

In 1971, while working at ARP Instruments (where I designed the ARP 2600 synthesizer, made known by Stevie Wonder), I presented a paper at the Audio Engineering Society on "The Electrical Design and Musical Applications of a Voltage-Controlled Filter/Resonator." The circuit I designed became the ARP Module 1047 Multimode Filter/Resonator (**Photo 1**). This made the variable tonality effect in the rhythmic background of The Who song entitled "Won't Get Fooled Again."

Fast-forward 36 years: **Photos 2** and **3** show the prototype of the circuit described in this article. I haven't designed it into an enclosure yet; I'm waiting to see whether there's enough interest. See **Fig. 12** for a recommended control component layout.

PHOTO 1: ARP Module 1047 (from 1971).

TOPOLOGY, FREQUENCY, AND SQUARE WAVE RESPONSES

The topology and Laplace (frequency domain) analysis of the state-variable filter (also known as the "Biquad") are shown in **Fig. 1**. **Figure 2** shows the frequency responses on dB versus log frequency scales. The principal advantages of this topology are (1) you can obtain a very high Q (up to 1000) with stability, (2) the latter doesn't depend upon precise component value matching, and (3) you can achieve a very wide fo sweep range (over 1000:1) while maintaining response shapes and unity-gain passband regions.

In **Fig. 1**, the elements labeled "-a" and "b" can be voltage controlled attenuators (VCAs) as in the design. The "a" elements must track, but can be either both inverting or both non-inverting. Inverting "a" elements provide non-inverting BP and LP outputs. Using a dual log pot for the "a" elements can provide about a 100:1 fo sweep, and both "a" and "b" can be fixed resistors for fixed fo and Q applications.

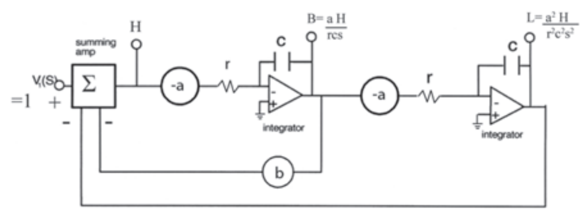


FIGURE 1: Laplace analysis of state variable filter.

$$B = \frac{aH}{rcS} \,.\, L = \frac{a^2H}{r^2c^2S^2} \,.\, H = 1 - \frac{abH}{rcS} - \frac{a^2H}{r^2c^2S^2}\,,\; H - 1 + \frac{abH}{rcs} + \frac{a^2H}{r^2c^2S^2} = 0 \qquad S^2r^2c^2H + SabrcH + a^2H = S^2r^2c^2.$$

$$H = \frac{S\frac{a}{rc}}{s^2 + s\frac{ab}{rc} + \left(\frac{a}{rc}\right)^2}$$

Highpass response $H = \dfrac{s^2}{s^2 + S\frac{ab}{rc} + \left(\frac{a}{rc}\right)^2}$

Bandpass response $B = \dfrac{S^2r^2c^2}{S^2r^2c^2 + Sabrc + a^2}$

Lowpass response $L = \dfrac{\left(\frac{a}{rc}\right)^2}{s^2 + s\frac{ab}{rc} + \left(\frac{a}{rc}\right)^2}$. $fo = \dfrac{a}{2\pi rc}$.

$Q = \dfrac{1}{b}$.

@f0 = fo, H = B = L = Q.

Figure 3 shows all three outputs' responses to a 1kHz square wave, with Q values of 0.5 (non-resonant, left column) and 1.0 (slightly resonant, right column). Filter fo is set to 5kHz. The LP output is DC coupled, and at fo all three outputs have a voltage gain that's numerically (not dB) equal to Q.

FILTER RESPONSES TO SOUND STROBE "PINK PULSE"

The "pink pulse" (**Fig. 4**) is so-called because, like pink noise, its spectrum is flat on a per-octave, per third octave (and so on) basis. The filter's BP response, for a given Q, also maintains a constant number of octaves (or fraction thereof) bandwidth regardless of fo. Therefore, the BP output from a pink pulse will maintain a constant energy per pulse as fo is changed. But with a constant Q setting, the pulse decay time is inversely proportional to fo.

To maintain constant pulse energy versus fo, the peak output voltage will increase proportionally to the square root of fo. With a step pulse filter input, the peak output voltage is constant with fo, so the decreasing decay time versus fo causes the pulse energy to decrease, inversely proportional to √fo; that is, at -3dB/octave. But this is expected, because the spectrum of a step pulse, while downsloping at -6dB/octave on a per-Hz analyzed basis, has a -3dB/octave downslope on a per-octave basis. The latter, as

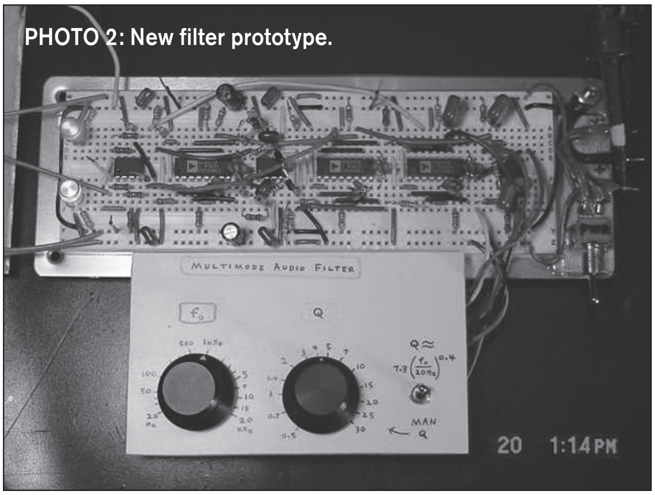


PHOTO 2: New filter prototype.

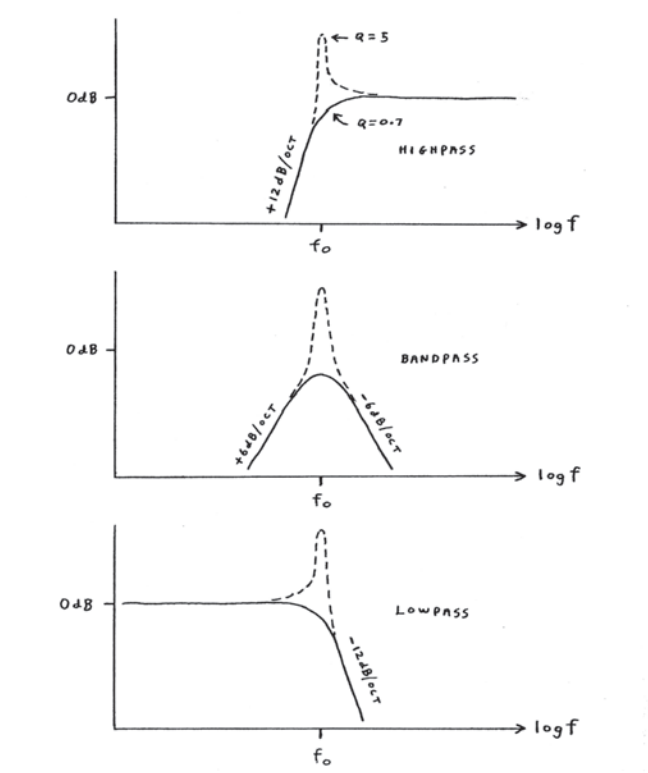


FIGURE 2: Frequency responses of multimode filter.

explained in the Sound Strobe article, is the way we hear: we analyze spectra on a per-octave (or fraction) basis.

SOUND STROBE PULSE WITH HIGH-Q RESONANCE

In **Fig. 4**, the Q is a high value of 59, with fo = 5kHz. Note the long resonant decay in the left waveform of **Fig. 4B**. The sinusoidal decay is a pure exponential; that is, a constant number of dB per time unit (linear dB/ms slope). This waveform takes 4.32ms to decay 10dB, and 25.9ms (130 waveform cycles) to decay 60dB.

This waveform is useful for hearing room acoustics properties; it's resonant enough to produce a distinct tonality (the spectral BW is 0.0245 octave, or 0.391 semitone), but the 60dB decay time of 25.9ms is much shorter than the reverberation time of any room except an anechoic chamber.

The right photo in **Fig. 4B** is the same waveform expanded to a 0.1ms/div time scale. Note the clean attack transient starting at the sinusoidal zero-voltage point. (The slight starting curvature is from the input pink pulse's finite rise time of about 13µs.)

PHOTO 3: Prototype controls.

TRANSIENT SOUND VARIATIONS

The resonant tone decay is the same at all three filter outputs (except for the +90° HP and –90° LP phase shifts regarding the BP out). But the transient "impact" sounds are quite different. The HP output (**Fig. 4A**) passes the full HF content of the pulse, resulting in a much sharper transient sound (at the 5kHz fo shown, the sound starts with a metallic "ting").

Conversely, the LP output transient is relatively "muted," but with significant LF content (manifested in **Fig. 4C** by the first half-cycle of higher amplitude than the others).

Figure 5 shows the most extreme range of HF and LF transient spectrum emphasis available. In **Fig. 5A**, the Sound Strobe impulse waveform is used with the filter's HP output. The impulse spectrum (flat per-Hz but upsloping +3dB/octave on the ear's per-octave analysis) has its strong HF content passed unattenuated by the HP output.

Figure 5B shows the opposite available extreme: The Sound Strobe's 252µs exponential pulse is used. This is a step pulse rolled off at 6dB/octave below 632Hz. The longer pulses (830µs and 25ms exponential) have more extended LF spectra, but I used the 252µs pulse here so its decay could be seen on the fast (0.1ms/div) time scale. The LP filter output is shown; note the strong LF content as the resonant sine cycles have

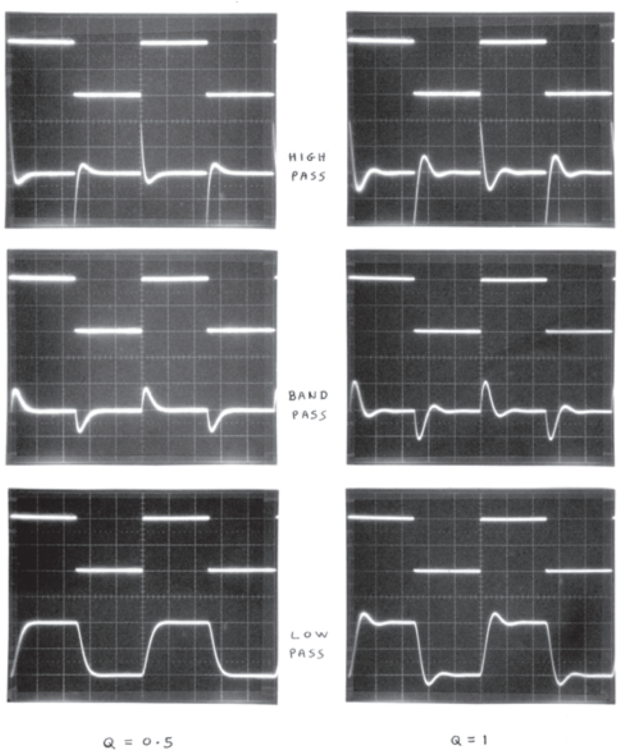


FIGURE 3: 1kHz square wave responses. fo = 5kHz, Q = 0.5 and 1. 1V/div, 0.2mS/div.

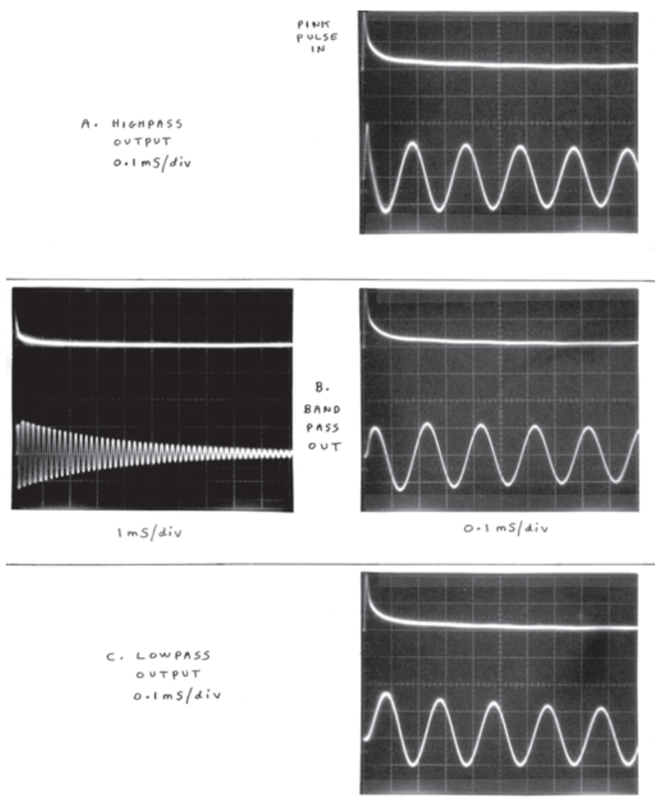


FIGURE 4: Sound Strobe pink pulse through filter. fo = 5kHz, Q = 59; 1V/div.

an initial offset that follows the input pulse decay. This type of waveform is useful for hearing a speaker's full-band transient coherence in the presence of a tunable, musical-sounding HF resonant decaying tone.

GENERAL PROPERTIES OF FILTERED TRANSIENT PULSES

Figure 6 shows a pink pulse through the BP output, with fo = 1kHz and a Q of 4.318, which corresponds (with a BP filter having ±6dB/octave asymptotic slopes) to a 1/3 octave BW. Also shown are equations for decay times, for various decay levels, fo, and Q values.

Only a handful of pulse outputs have been shown. But consider that with the Sound Strobe and this filter, the range of combinations includes six input pulse shapes, three filter outputs, a filter fo range of 20Hz – 20kHz, and a Q range of 0.5 to 30 (manual), 7.3 to over 100 (auto tracking mode). Even assuming that significant differences are represented by 1/3 octave fo steps and 2:1 Q changes, the number of combinations would be 1584. But I forgot to include

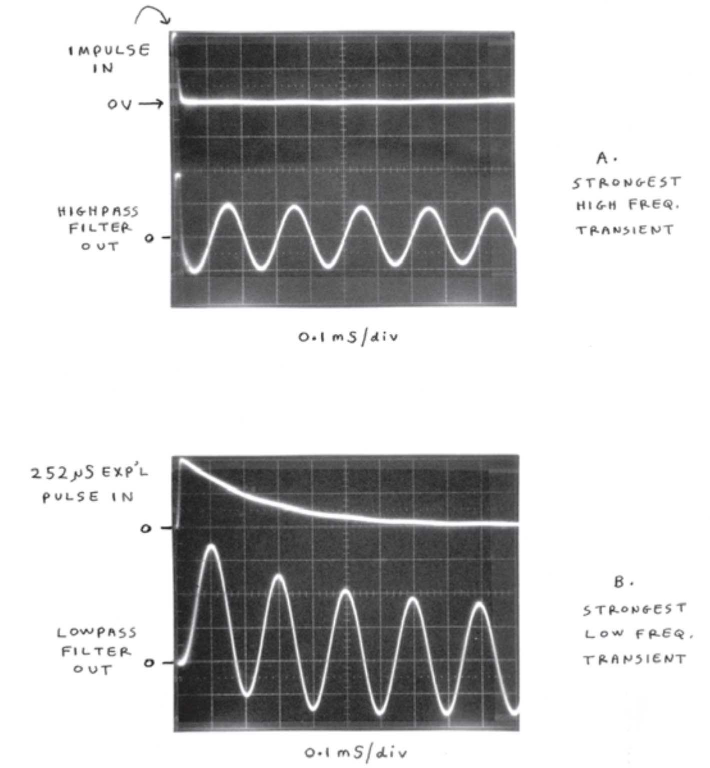


FIGURE 5: Sound Strobe pulse/filter combinations for strong HF and LF transients at start of resonant tone decay. Filter fo = 5kHz, Q = 59; 1V/div.

the Sound Strobe's pulse repetition frequency range of 0.5-41.2Hz!

THE +1.8DB/OCTAVE SLOPE EQ

For the "Auto" Q mode, I experimentally determined the best-sounding amount of Q tracking with fo. Roughly a compromise between constant Q and constant decay time (which would be a Q proportional to √fo), the relation turned out to be approximately Q = 7.3(fo/20Hz)$^{0.4}$. The pulses have enough resonant sustain to resemble musical tone transients, but with short enough decay times to reveal speaker anomalies in transient precision. Also, room effects are heard, distinguishable from the direct speaker output.

The Auto Q mode is meant to be used with the Sound Strobe 25ms exponential pulse, of frequency below about 4Hz (or a square wave below about 2Hz, with 2.3Vpp maximum amplitude, that of the Sound Strobe). These wideband step pulses produce a constant peak pulse voltage versus fo from the filter's BP output. (This is also true at the other outputs and regardless of Q, manual, or auto.)

Due to the decay time shortening as fo increases, the energy per pulse (and therefore the RMS voltage for a given pulse repetition frequency, or PRF) decreases with increasing fo. The pulse energy decreases with a slope of about -1.8dB/octave (18dB decrease from 20Hz to 20kHz fo).

The slope EQ, when switched in with S1, compensates for this when a constant pulse energy is desired, such as for measuring the reverberant power/frequency distribution in a room. But there's a disadvantage: As **Table 1** shows, the peak BP output pulse voltage increases by 19dB from 20Hz to 20kHz fo (0.50V to 4.59V with a 2.0Vpp Sound Strobe 25ms exponential pulse at rates below 2Hz). This can clip amplifiers and possibly blow tweeters.

However, for general speaker/room evaluations, you should use the constant-peak output with the EQ off (S1 set to "FLAT"). The pulse loudness versus fo sounds reasonably balanced.

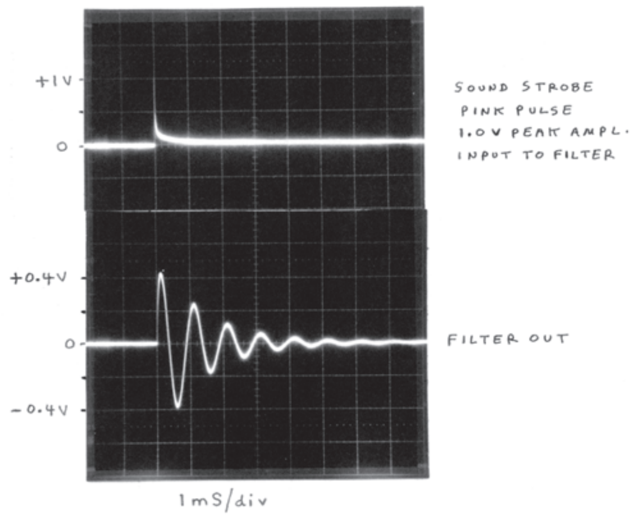


FIGURE 6: Sound Strobe pink pulse through 1/3 octave 1st order bandpass filter, fo = 1.00kHz, Q = 4.318, AV (fo) = 4.318, BW -3dB = 890.9 – 1122.5Hz.

Time	Decay to:
0ms	0dB
1.375ms*	-8.69dB
1.583ms	-10dB
3.165ms	-20dB
9.495ms	-60dB

*** Exponential decay time constant**
For any filter fo (1kHz shown above): constant pulse energy, T (– 10) = 10dB decay time = 1583/fo ms = Q/2pifolog10e = 0.36647Q/fo
Vp = 0.42V √fo/1000, BW – 3dB = fo/Q

FILTER CIRCUIT

The circuit (**Fig. 8**) follows the topology of **Fig. 1**. U1 is the input and feedback (FB) summing amp, and U2A, U2B are the integrators. VCA1 and VCA2 are the "-a" elements (controlling fo), while VCA3 is the "b" element (controlling Q). Note (from **Fig. 1**) that Q = 1/b, so if the gain of b is zero, the Q is infinite; that is, you have an oscillator. This is because without b, there are two integrators cascaded in a FB loop. At DC the FB is inverting, but for any AC signal the integrators have a phase shift of –90° (-180° for the two cascaded). So at the frequency (fo) where the loop gain (determined by a^2 and the value of C6, C12) is unity, the net positive FB produces oscillation.

This is the principle I used in "A Wide-Range Audio Sweep Oscillator" (*aX* Feb. '07). The operational transconductance amps I used there (the RCA/Harris CA3280AE) are no longer available. Compared to those, the Analog Devices SSM2018TPZ units I used in the present filter for voltage controlled amplifier/attenuator (VCA) devices have some disadvantages and advantages.

Disadvantages:

1. Less BW, about 250kHz instead of 3MHz for the CA3280AE (the latter at maximum gain where high BW is needed because fo is proportional to VCA gain). The SSM2018's 250kHz rolloff produces a phase lag (about –5°) at 20kHz. This would increase the integrators' desired –90° phase, which would make the filter oscillate at high intended Q values. C5 and C11 provide a phase lead, compensating this effect.
2. The SSM2018 has a low-Z voltage source output, while the CA3280 has (or rather, had) a high-Z current

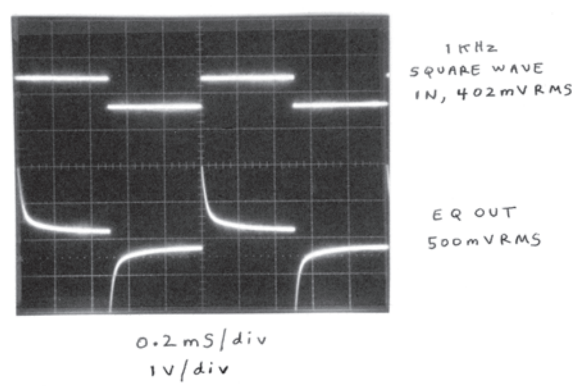


FIGURE 7: 1kHz square wave response of slope EQ. Average slope = +1.82dB/octave, 20Hz – 20kHz.

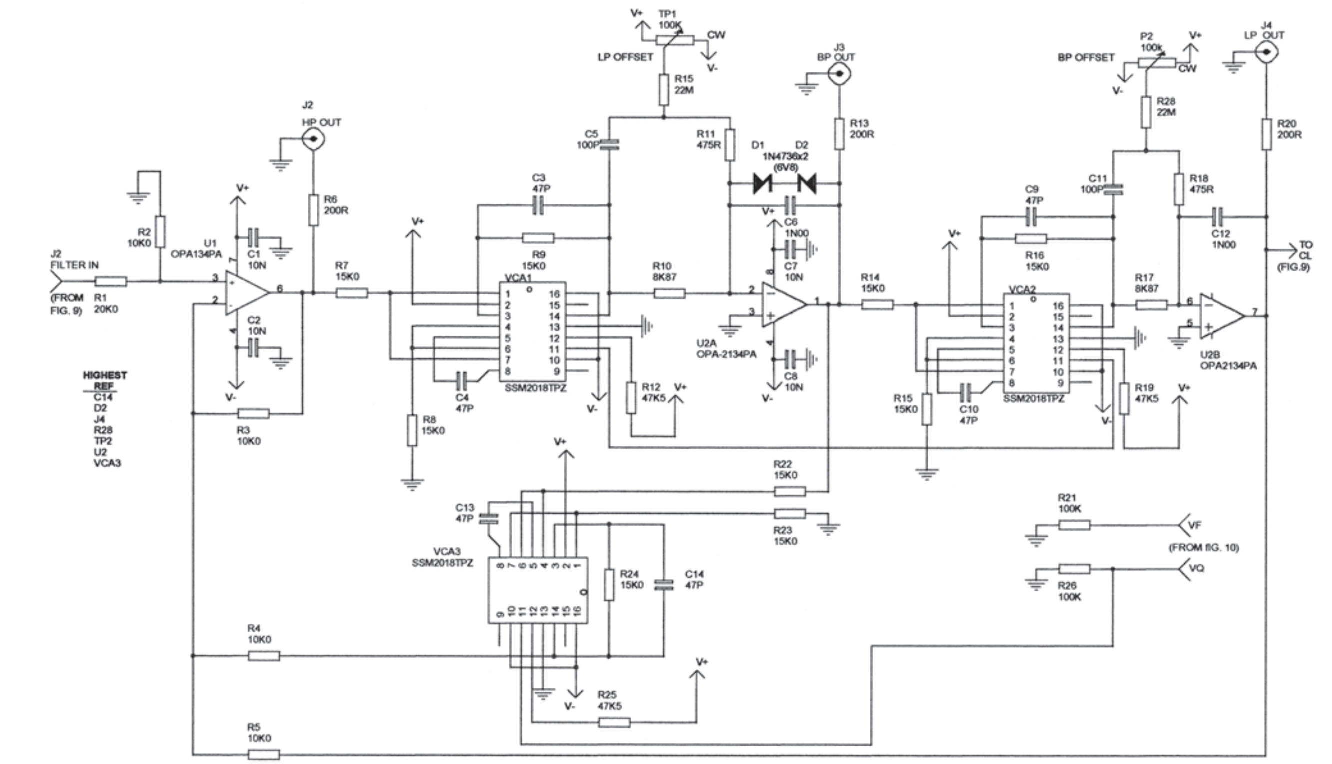


FIGURE 8: Multimode active filter.

source output. With the latter, DC offsets decrease in proportion to the controlled signal current; also, the integrator op amps' offset voltages don't see a low Z from the VCA. But with the SSM2018's low-Z output, plus the device's own output offset (1mV typical, 15mV maximum), there's a much more potential offset problem.

Because the integrators are open-loop at DC, it's only the filter's overall negative FB that keeps offsets in check. But at low fo values, the VCAs (1 and 2) are at high attenuation (fo is 17.7kHz when the VCA is at unity gain). At 20Hz, the two VCAs combined have 118dB of attenuation (a voltage loss of 787,000:1). So there's much less overall negative FB to control the integrators' output DC offset. TP1 and TP2 are needed to trim the filter for acceptably low offset at low fo (to be described later).

Advantages:

1. The SSM2018 has much lower distortion (0.01%) and higher gain-control precision, and a 140dB gain-control range.
2. It doesn't need (as the CA3280 did) an external exponential converter; the VCA has a very linear dB versus control voltage relation (at pin 11).

Q CONTROL

VCA3 (the "b" element in **Fig. 1**) introduces a second negative FB loop, this one around only the first integrator (U2A). This FB produces damping in proportion to the gain of VCA3. That's why the Q is inversely proportional to VCA3's gain. The gain of VCA3 varies from 2 (+6.0dB) for a Q of 0.5, to about 0.0085 (-41.4dB) for the maximum Q of 118 (at 20kHz in "Auto" Q mode).

The zeners (D1, D2) prevent high signal voltages from overdriving the VCA.

SLOPE EQ AND CLIP LIGHT

Figure 9 shows the simple, switchable RC circuit that provides the +1.82dB octave upslope, reasonably uniform over the audio band. The clipping indicator light (LED1) is useful when the filter Q is high. High Q is mainly intended for very low frequency (0.5-2Hz) Sound Strobe pulses, where you want to hear individual resonant-tone pulse responses of rooms or speakers.

However, the Sound Strobe pulse frequency goes up to 41.2Hz (the low open-E string on a 4-string bass). The filter, with high Q, can tune individual harmonics out of Sound Strobe pulses above about 5Hz, but the filter's gain is numerically equal to Q. Therefore, the filter outputs can easily clip in such applications. LED1 allows you to decrease the Sound Strobe's (or other source's) signal amplitude, when necessary to avoid filter clipping.

TABLE 1: BP output pulse characteristics, "Auto" Q mode. Source is Sound Strobe 25ms exponential pulse, 1Hz rate, 2.0Vpp amplitude. T-20 is 20dB decay time, N-20 is number of waveform cycles to 20dB decay.

			EQ on		EQ off			
fo	Q	EQ gain dB	pulse energy dB re 1kHz	peak volts	pulse energy dB re 1kHz	peak volts	T-20 ms	N-20 cycles
20Hz	7.29	-10.54	-0.25	0.50	+9.64	1.66	267	5.3
50	11.44	-8.60	-0.33	0.61	+7.62	1.69	168	8.4
100	15.83	-6.47	+0.20	0.75	+6.02	1.74	116	11.6
200	21.51	-4.93	+0.06	0.96	+4.34	1.77	78.8	15.8
500	31.07	-2.98	-0.37	1.25	+1.96	1.77	45.5	22.8
1kHz	39.61	-0.65	0.00	1.66	0.00	1.81	29.0	29.0
2	48.5	+1.80	+0.32	2.30	-2.13	1.81	17.8	35.5
5	59.1	+4.85	+0.25	3.34	-5.25	1.81	8.66	43.3
10	71.7	+6.73	-0.04	4.13	-7.42	1.81	5.26	52.6
15	86.0	+7.36	-0.38	4.38	-8.39	1.81	4.20	63.0
20kHz	117.7	+7.62	-0.01	4.59	-8.28	1.81	4.31	86.3

EQ gain is 0dB at 1.191kHz

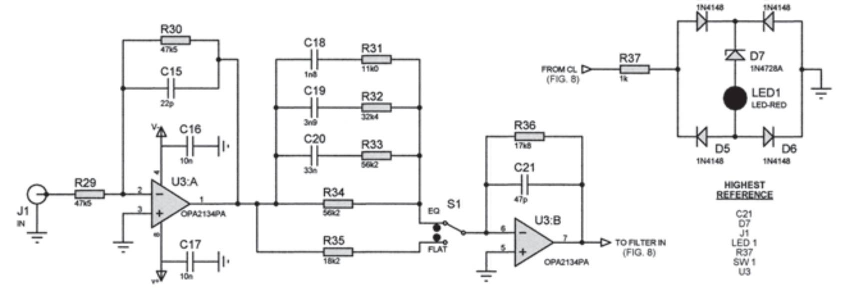


FIGURE 9: EQ and clip light.

CONTROL CIRCUIT

First, notice that this circuit (**Fig. 10**) uses voltages labeled "+5V PTAT" and "-5V PTAT." "PTAT" stands for "Proportional To Absolute Temperature" (degrees Kelvin). This will be described later; for now, suffice it to say that this temperature compensates for the VCAs' gain-control slope. For constant gain (or attenuation), the VCA control voltage applied must be PTAT.

Switch S2 selects fo (when S3 is in the "oct ±1 oct" position) in octave steps, from 40Hz to 10kHz. Control pot P1 adds a ±1 octave sweep range, so S2 and P1 can control fo from 20Hz to 20kHz.

When S3 is in the "Full Δfo sweep" position, S2 has no effect, but P1 now tunes fo over the full 20Hz – 20kHz range.

PTAT CIRCUIT

The VCAs have a control port sensitivity specification of -30mV/dB at +25° C; I measured -31.7mV/dB. But this is proportional to absolute temperature (PTAT). Therefore, to obtain temperature independence of the VCAs' gains (and thus the filter's fo and Q), the control voltages must be PTAT.

You could accomplish this by using a +3500 ppm/° C resistor (close enough to the ideal 3354 ppm/° C tempco), such as the model PT146 1k unit from Precision Resistor Co. of Largo, Fla. that I used in my 6-channel volume/balance control (*aX* Feb. '03), which uses the same VCAs. But these cost $5 each, the filter would need two (fo and Q), and I know of only this one source.

But there's another way: Make the fo and Q control and bias DC voltages PTAT. With the use of external fo and/or Q control signals, the above TC resistor would be needed. Here, though, the internal control pots and switches can all be supplied from PTAT DC references.

An excellent source for this is the plain old ubiquitous 1N4148 diode. If biased from a constant-current source, and the diode drop is subtracted from an appropriate constant voltage, the resulting voltage can be very close to PTAT, at least over a 0° to +50° C range (+32° to +122° F). The subtraction makes the tempco positive.

I measured the 0° to +50° C tempco of the voltage across three 1N4148 samples, at 1mA. The tempcos were within 1.1% of -2.1433mV/° C. The diode drop average was 611.38mV; highest and lowest were 616.8mV and 605.8mV.

To obtain a PTAT voltage, the output should be the product of the tempco magnitude and 298.15° K (equal to +25° C), which is 639.02mV. Then the DC reference should be the sum of this plus the typical 1N4148 drop of 611.38mV (at 1mA); this is 1.2504V ideally (the exact value isn't that critical).

By a strange Colincidence (term invented by my father, Robert Colin), the internal reference voltage in an LM317 regulator is very close to 1.25V. (Does National Semi know something I don't?

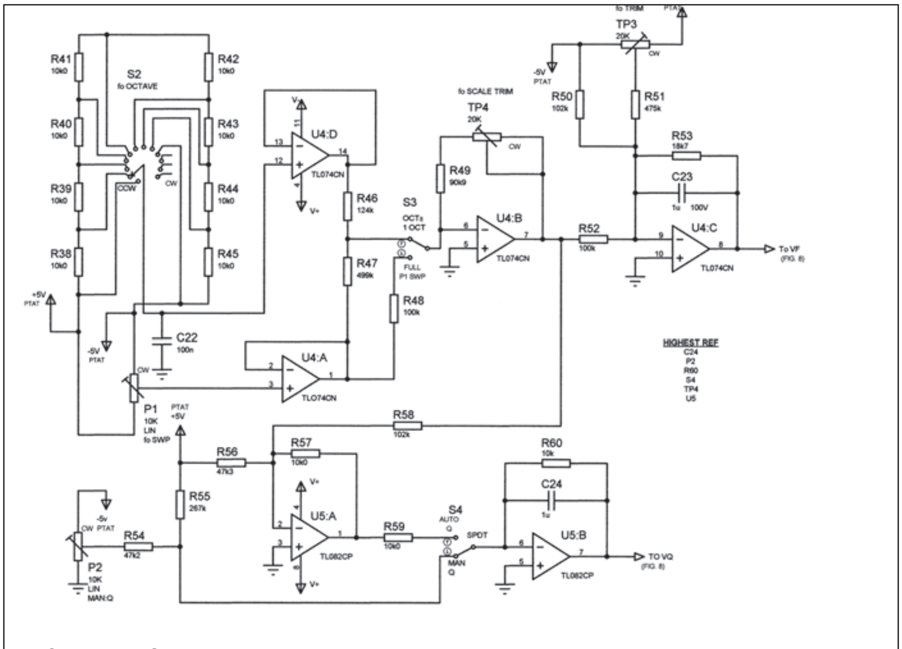


FIGURE 10: Controls.

Of course they do, they're experts.)

The circuit of **Fig. 11** takes advantage of this. The output of the LM317 is always 1.25V (typical VREF) above the ADJ terminal. So the diode, biased negatively regarding ground, has its voltage drop subtracted from the 1.25V VREF; and without the usual external feedback needed, the LM317 output provides a very stable voltage of about +639mV at +25° C that's proportional to absolute (° K) temperature. The 100Ω load resistor ensures that the LM317's minimum load current (3.5mA typical, 5mA maximum) is drawn.

POWER SUPPLY

The circuit in **Fig. 11** is similar to that in the Sound Strobe. The rechargeable 9V batteries can be either NiCd (such as Mouser 639-N6PT) with 110mAH capacity, or the newer NiMH types (such as Mouser 573-GP17R8H) with 170mAH capacity. The latter type is recommended, both for its higher capacity (which will power the filter for about 2.4 hours) and its higher minimum voltage (8.4V as opposed to 7.2V for the NiCd and some other "9V" NiMH batteries).

If you prefer, you can power the filter from any ground-referenced supply of ±9V to ±15V.

TRIMMING

A. Power Supply Test:

1. With the 24V wall supply connected and S5 set to "AC," LED2 ("AC") and LED3 ("PWR OK") should light. Measure the V^+ and V^- voltages regarding ground; they should be in the range of ±9V to ±11V.
2. Measure the +5V PTAT voltages. They should be in the range of ±4.3V to ±5.7V, depending on the forward drop of D13 and ambient temperature.

B. Offset Trimming:

1. Set TP1, 2, 3, 4 to their approximate midpoints.
2. Set S2 to the 630Hz position (the switch's midpoint), P1 to its approximate midpoint, and S3 to the "OCT ± OCT" position. Set P2 (the manual Q pot) fully CW (clockwise), and S4 to "manual Q."
3. Measure the DC voltages at the BP and LP outputs (re ground). If two meters are available, connect one to each output. (Accuracy isn't needed; the cheapest meters that can resolve down to 50mV DC will do.)
4. Lower the "fo octave" setting (S2) one step at a time. At some point the offsets will increase significantly. If using only one meter, alternately observe the BP and LP offsets.
5. Lower the "fo octave" setting as many steps as possible without either offset exceeding ±2V.
6. Adjust TP1 to lower the LP offset as much as possible (it might drift around, but you should be able to null it within ±50mV).
7. Adjust TP2 to null the BP offset.

Note: It might seem strange that each trimpot affects not the op amp integrator that it feeds, but rather the opposite one. This is because the integrators' offsets are stabilized not by local op amp FB (there isn't any at DC), but by the overall loop. Each integrator receives DC feedback from the other one's offset.

8. Lower the fo setting further, and repeat steps 6 and 7.
9. Continue this until the fo setting is 20Hz (S2 fully CCW (counterclockwise) and P1 fully CCW). Then null the offsets as much as possible.

C. Frequency Trimming:

Note: This requires a 20Hz – 20kHz sine wave generator and an oscilloscope (preferably) or an AC voltmeter capable of responding to the 20Hz – 20kHz range (voltage accuracy and true RMS response are not necessary).

1. Connect the audio generator to the filter input. If a dual-channel scope

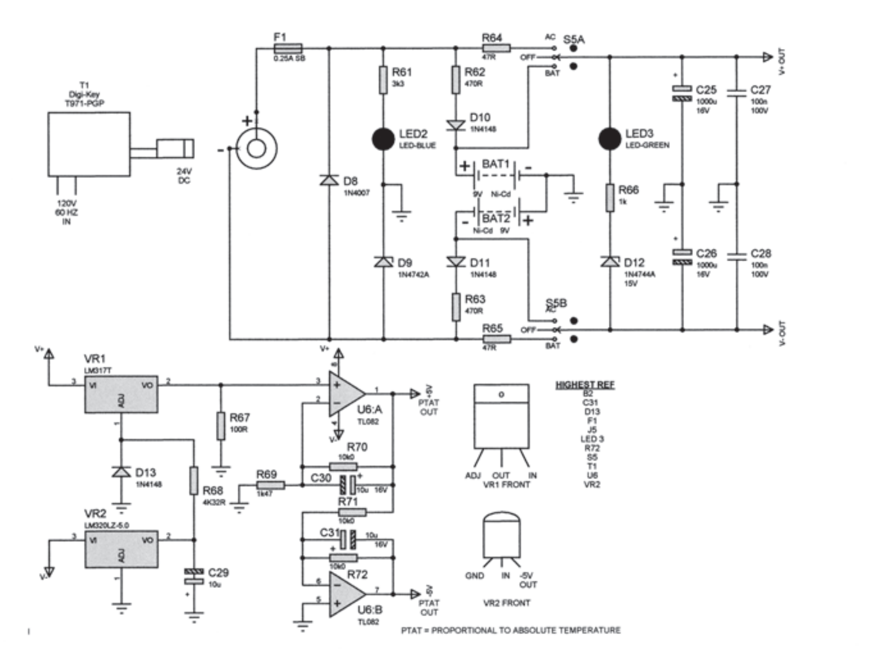


FIGURE 11: Power supply.

TABLE 2: Filter data (with slope EQ switched to "Flat" position).

Gain and Frequency Response	
LP out:	DC coupled, gain = 0dB for f <<fo, gain = Q at fo, rolloff = -12dB/octave for f >>fo.
BP out:	gain = Q at fo, rolloffs = ±6dB/octave for f far from f.
HP out:	gain = Q at fo, rolloff = +12dB/octave for f<<fo, HF response -3dB at 3MHz.
Maximum output = ±6.8V peak, 4.8V RMS	
Zin = 47k5, Zout = 200Ω	
THD ≈ 0.01%	
fo range = 20Hz - 20kHz	
Q range = 0.5 - 30 (manual), 7.3 - 118 (Auto)	
Noise Out, RMS	
fo = 1kHz, Q = 1: 99µV (20Hz - 20kHz), 62µV ("A" weighted)	
fo = 20Hz, Q = 1: 875µV (20Hz - 20kHz), 78µV ("A" weighted)	
fo = 1kHz, Q = 30: 250µV (20Hz - 20kHz), 230µV ("A" weighted)	
Output DC offset: Trimmable to ±100mV	
Power Supply	
24V DC, approximately 70mA, ground isolated. May also use ground-reference supply of ±9V to ±15V, approximately ±70mA.	

is available, connect the generator also to channel 1. Trigger the scope either from channel 1 internally, or (preferable) externally from the generator's trigger output, if it has one. Connect the scope's channel 2 to the filter's BP output. Set the generator's output amplitude to about 0.5V RMS (±0.7Vpp on the scope), and frequency to 630Hz.

2. Not required, but desirable, is to monitor the source frequency with a counter, unless the generator's frequency calibration is accurate to about ±2%.
3. Switch the slope EQ to "Flat," S2 to its midpoint (630Hz), S3 to "Oct ± 1 oct," and S4 to "manual Q." Set P1 to its midpoint, and P2 (manual Q) also to its midpoint (Q about 5).
4. Adjust TP3 (fo trim) for a maximum output from the BP out, on channel 2 of the scope (or AC meter if a scope isn't available). The maximum BP output voltage should be about 5× the input voltage. Note: the most accurate and sensitive way of tuning is for zero phase shift between the input and output waveforms. And the best way to see this is to switch the scope mode to "X-Y" if available; then at zero relative phase the ellipse will collapse to a straight tilted line.
5. Change the generator frequency to 20Hz, and set the filter fo controls to 20Hz by rotating both S2 and P1 fully CCW.
6. Adjust TP4 (fo scale trim) for maximum filter output and, if possible, zero input/output phase (as in step 4).
7. Set the generator frequency to 20kHz, and the fo controls also to 20kHz (S2 and P1 fully CCW).
8. Adjust TP4 for tuning as in steps 4 and 6.
9. For the most accurate full-band tuning, alternately repeat steps 6 and 7, while periodically adjusting TP3 if necessary to offset both the 20Hz and the 20kHz settings. This can be somewhat laborious, but is needed only if the highest accuracy is desired. For most applications, the results of step 8 are acceptable.

HIGHEST Q STABILITY

The loop phase-compensating caps C5 and C11 have a strong effect on stability at 20kHz and maximum Q (which occurs in the "Auto" Q mode). To test this stability:

1. Connect a 50Hz (approximately) 1Vpp square wave to the filter, and monitor the LP output on a scope.
2. Set fo to 2.5kHz. Set the Q mode to "manual," and set P2 (manual Q) fully CCW (Q = 0.5). The output should be a LP filtered square wave, with no overshoot.
3. Turn up the Q control slowly to fully CCW. As you do, decaying ringing cycles should appear.
4. Switch S4 to "Auto" Q mode. At fo settings above 1kHz (such as the 2.5kHz it's set to now), the Q should increase when switching S4 from "manual" to "Auto." This should make the ringing's decay longer.
5. Set P1 to about two-thirds of the way up, and switch S3 to "Full P1 Sweep."
6. Slowly turn up P1. When it's up fully CW, the filter might break into continuous oscillation at 20kHz. If this happens, C5 needs to be increased. Increasing it from 100pF to 120pF should be enough. Increasing it too much will reduce the maximum available Q.
7. The ideal is to have a value of C5 (C11 has the same effect) that produces a long ringing decay at 20kHz fo, without oscillating. C5 can be a trimmer cap with a minimum value of 80pF or less, and a maximum value of at least 150pF.

Table 3 is the parts list. **Figure 12** shows my suggested enclosure panel component layout. Please let *audioXpress* know

SPEAKER CROSSOVER COMPARISON EXAMPLE

I found the filtered Sound Strobe pulses to be very revealing of a speaker CO alignment. I could easily hear the frequency delay dispersion (see my article by that title in *aX* 11/07) of a woofer/mid CO that had very flat amplitude/frequency response, the Swans M1/Scan-Speak sub unit ("Stereo Subs for the Swans M1," *aX* Sept. '05).

The raw Sound Strobe pulses, particularly the LF, 25ms exponential, and 830µs exponential pulses, clearly revealed the 220Hz CO's transient delay dispersion. But filtering the 25ms exponential pulse with the BP filter fo at 220Hz made the delay dispersion more audible. I compared the Swans system to a full-range Seas coincident driver in a sealed box. With a filter Q of 30, the Swans made an audible separation between the transient attack and the resonant tone. But with the Seas, the sound was coherent; I could picture an actual drum skin sound.

Interestingly, there were many positions in the room where I could hear a standing wave null, almost removing the pulse's 220Hz tonality, with both speakers. However, I could still hear the superior "live drum-like" transient response of the full-range Seas, *regardless of room effects*.

This correlates with recorded drums and string bass sounds, but I must have to listen for the Swans/sub delay-smearing effect. With the filtered Sound Strobe pulses, it's immediately recognizable.

Figure 13 shows an electronic representation of this effect: The "A" waveform is the 25ms exponential pulse, filtered with fo = 232Hz and Q = 0.65; that Q produced the cleanest pulse without significant overshoot.

Waveform "B" is this pulse through a three-stage nonresonant allpass network, with its fo (center of its phase/frequency graph on a log frequency scale) at 232Hz, the same as the BP filter's fo. The allpass network's group delay ranges from 4.12ms for frequencies much lower than 232Hz, to 2.06ms at 232Hz, and approaches zero at frequencies much above 232Hz.

This allpass response is similar to a fourth-order CO that (like the allpass) has a flat amplitude/frequency response. Note how the clean transient "A" waveform is smeared with frequency selective delay. The original waveform's peak is delayed about 3.4mS, and also reduced in amplitude. *Unlike some measurements, this effect is as audible as it looks*.

whether you would find this filter useful. If there's enough interest, a kit and assembled product may become available. **Table 4** shows bandwidth versus Q data.

CONCLUSION

The wide range of transient pulses shown in this article can be very useful test signals for directly hearing speaker time misadjustment, frequency response anomalies, and spatial image coherence; and also the effects of room acoustics. The principal features are:

1. The raw Sound Strobe pulses are best for hearing wideband transient response and overall freedom from (or degree of) speaker coloration, and the smoothness of room acoustics (freedom from discrete echoes).
2. The filtered pulses allow frequency tuning of all the audible effects described. The Q control allows a wide range of resonant tonality (including none) to be selected.
3. The precise-transient resonant tone pulses, because of their resemblance to natural acoustic sounds, provide

TABLE 3: Parts List

Ref.	Description	Size	Mfr	Mfr/P/N	Dist.	Dist. P/N
B1, 2	battery, 9V, NiMH				M	573-GP17R8H
C15	cap, 22pF cer.					
C3, 4, 9, 10, 13, 14, 21	cap, 47pF, cer.					
C5	cap, 100pF cer.					
C11	cap, 120pF cer.					
C6, 12	cap, film, 1nF, 100V				DK	PS1102J
C18	cap, 1n8, 100V				DK	PS1182J
C19	cap, 3n9, 100V				DK	PS1392J
C1, 2, 7, 8, 16, 17	cap, 10nF, 100V				DK	PS1103J
C20	cap, 33nF, 100V				DK	PS1333J
C22, 27, 28	cap, 100nF, 100V				DK	EF1104
C23, 24	cap, 1µF, 100V				DK	EF1105
C29-31	cap, 10µF, 16V				DK	493-1279
C25, 26	cap, 1000µF, 16V				DK	493-1287
D1, 2	zener, 6.8V	1W		1N4736A-T	DK	1N4736ADICT
D3-6, 10, 11, 13	diode			1N4148		
D7	zener, 3.3V	1W		1N4728A-T	DK	1N4728ADICT
D8	diode			1N4007-T	DK	1N4007ADICT
D9	zener, 12V	1W		1N4742A-T	DK	1N4742ADICT
D12	zener, 15V	1W		1N4744A-T	DK	1N4744ADICT
F1	fuse, 0.25A, SB					
J1-4	jack, RCA				PE	091-1120
J5	jack, DC in, isolated				M	163-4303-EX
LED1	LED, red				DK	67-1612
LED2	LED, blue				DK	P466
LED3	LED, green				DK	516-1358
P1, 2	pot, 10k lin				DK	RV4N103C
R64, 65	Res, 5%, 47R	¼W				
R67	Res, 100R	¼W				
R62, 63	Res, 470R	¼W				
R37, 66	Res, 1k0	¼W				
R61	Res, 3k3	¼W				
R27, 28	Res, 22M	¼W				
R6, 13, 20	Res, 1%MF, 200R	¼W				
R11, 18	Res, 475R	¼W				
R69	Res, 1k47	¼W				
R68	Res, 4k32	¼W				
R10, 17	Res, 8k87	¼W				
R2-5, R38-45, R57, 59, 60, 70-72	Res, 1%, 10k0	¼W				
R31	Res, 11k0	¼W				
R7-9, 14-16, 22-24	Res, 15k0	¼W				
R35, 36	Res, 17k8	¼W				
R53	Res, 18k7	¼W				
R1	Res, 20k0	¼W				
R32	Res, 32k4	¼W				
R54	Res, 42k2	¼W				
R12, 19, 25, 29, 30, 56	Res, 47k5	¼W				
R33, 34	Res, 56k2	¼W				
R49	Res, 90k9	¼W				
R21, 26, 48, 52	Res, 100k	¼W				
R50, 58	Res, 102k	¼W				
R46	Res, 124k	¼W				
R55	Res, 267k	¼W				
R51	Res, 475k	¼W				
R47	Res, 499k	¼W				
S1, 3, 4	Switch, SPDT				DK	
S5	Switch, DPDT, Ctr, Off				DK	
S2	Switch, rotary, 12 pos.				M	105-SR2611F-26-21RS
T1	Wall transformer, 24V DC	400mA			DK	T971-P6P
TP3, 4	Trimpot, 20k				DK	32967-203
TP1, 2	Trimpot, 100k				DK	3296Y-104
U1	op amp		TI	OPA134PA		
U2, 3	op amp, dual		TI	OPA2134PA		
U4	op amp, Quad		TI	TL074CN		
U5, 6	op amp, dual		TI	TL082CP		
VCA1-3	voltage control amp		AD	SSM2018TP2		
VR1	Reg., Adj. pos			LM317T		
VR2	Reg., -5V			LM320LZ-5.0		
	LED Lens, red				DK	L30001
	LED lens, blue				DK	L30006
	LED lens, green				DK	L30005
	Knob, 0.925″ diameter				DK	226-4012
	Knob, 0.750″ diameter				DK	226-4011
	Battery holder				DK	BH9V-W
	Fuse holder				DK	F006

DK = Digi-Key, PE = Parts Express, M = Mouser good correlation with musical reproduction quality, but with very high sensitivity and repeatability. *aX*

REFERENCES

1. Dennis Colin, "The Sound Strobe," *aX* March '06.

2. Ed Simon, "Product Review: The Sound Strobe," *aX* April '07.

3. Dennis Colin, "A Wide-Range Audio Sweep Oscillator," *aX* Feb. '07.

TABLE 4: Normalized and octave bandwidths as a function of Q

Q	Lower* f - 3dB	Upper* f - 3dB	-3dB BW*	-3dB BW octaves
0.5	0.4142	2.4142	2.0000	2.5431
0.7071	0.5176	1.9319	1.4142	1.9000
1	0.6180	1.6180	1.0000	1.3885
1.4142	0.7071	1.4142	0.7071	1.0000
2	0.7808	1.2808	0.5000	0.7140
2.8710	0.8409	1.1892	0.3483	½
4.3185	0.8909	1.1225	0.2316	⅓
8.6514	0.9439	1.0595	0.1156	⅙
14.424	0.9659	1.0353	0.0693	⅒
30	0.9835	1.0168	0.0333	0.04809

*Normalized to fo = 1Hz

1. Note that for Q = 1, the upper and lower -3dB frequencies are 1.618034 and 0.618034. These are the "Golden Ratios," exactly equal to ½ (√5 ±1).

2. The normalized -3dB frequencies (f – 3dB/fo) are:

$$\frac{1}{2}\left[\sqrt{\left(\frac{1}{Q^2}+4\right)}\pm\frac{1}{Q}\right]$$

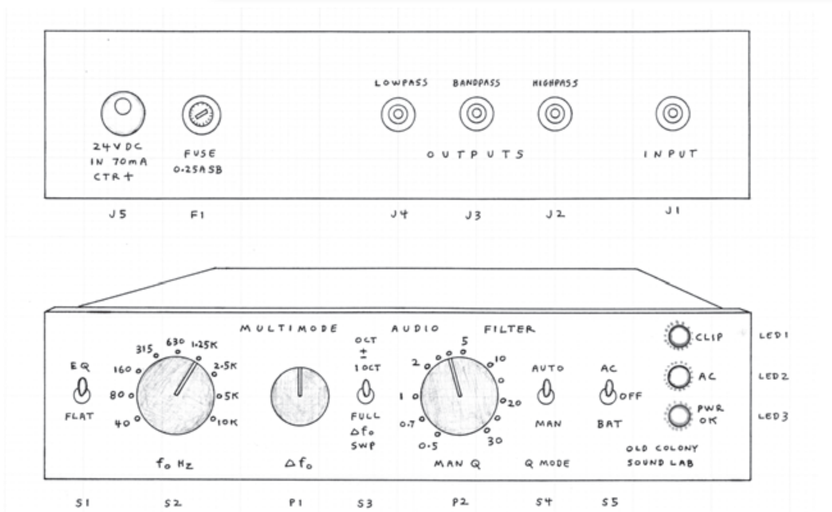


FIGURE 12: Suggested enclosure panel component layout, based on front panel dimensions = 9.0″ × 2.1″.

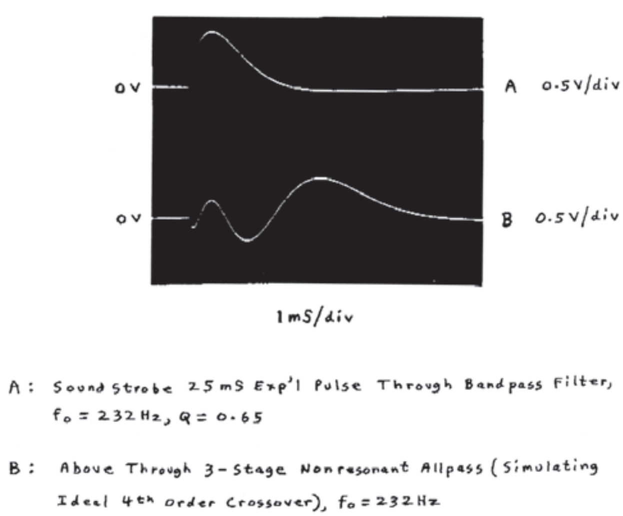


FIGURE 13: Simulated woofer/mid crossover effect on filtered Sound Strobe pulse.

A Belt-Drive Turntable Upgrade

Give your turntable a new lease on life with these tonearm re-wiring tips.

Back in the mid-1980s, I published three articles in this magazine's predecessor describing a belt-driven turntable based on the old AR-XA drive system. The first part described a new base for the turntable, and modifications to the original AR-XA sub-chassis to reduce resonance problems and accommodate the Mayware Formula 4 and Grace 747 tonearms[1]. The second installment described an electronic speed control designed to provide a low-distortion sine wave to the turntable motor. A variable-frequency oscillator allowed pitch adjustment on synchronous motor turntables, and 45 rpm speed without moving the belt to the larger pulley[2].

The third part involved replacement of the modified AR sub-chassis with Merrill Audio's acrylic sub-chassis, and the installation of Joseph Grado's now legendary Signature ToneArm[3]. Other enhancements included a high-torque Merrill synchronous motor, the Merrill replacement spindle, Merrill lead coating for the outer platter, Merrill platter balancing, heavier springs for the sub-chassis, and hum shielding necessary in order to use the turntable with the unshielded Grado Signature 10MR phono cartridge. This excellent cartridge was subsequently upgraded to the outstanding Grado Signature XTZII. The third article included a sidebar by George Merrill on sub-chassis design and hum shielding. All three articles have been reprinted in *The LP is Back!*[4]

Although I own over 6000 CDs, I continue to buy and play LPs, and am still actively interested in analog playback (including a large collection of 78 rpm records). The Galo/Merrill/Grado turntable has served me well for about 20 years, but it was time to either replace the stylus in my Grado XTZII or buy a new cartridge. After John Grado informed me that the XTZII stylus was still available, I decided to obtain a replacement rather than purchase a new cartridge (John is the nephew of Joseph Grado and the President of Grado Laboratories).

The XTZII, like the other Grado cartridges, is a moving iron design, and its low coil inductance and resistance (9mH; 70Ω), plus an ultra-efficient generator system, provides many of the advantages of moving coil designs without some of the liabilities. The Grado Signature ToneArm remains a stellar performer, and an ideal match for the Grado cartridges. We fortunate few were able to purchase one in the mid 1980s—during the short time it was available—for the bargain price of $485. If you're looking for a new cartridge, Grado has a full line covering all price ranges. You can order Grado cartridges and replacement stylii from Audio Advisor, Needle Doctor, and LP Gear.

SUSPENSION ISSUES

One problem with belt-driven turntables using three-spring suspensions concerns the tonearm interconnect cable. It's difficult to dress the tonearm cable in a way that doesn't interfere with the turntable suspension. In "The Belt-Driven Turntable Revisited," I suggested fastening the Grado cable to the underside of the turntable, on the end opposite the tonearm.

I also provided plans for constructing a rectangular "sub-base," which raised the turntable to allow the cable to droop without pulling on the turntable suspension or touching the surface below. Though this system worked adequately, it has always been rather cumbersome. But, the Grado tonearm cable was heavy enough that I really didn't have much choice (Grado used a Mogami Neglex 2534-style balanced quad cable, but Joe Grado told me that the copper was slow-drawn to his specifications).

Grado's tonearm cable was a fine performer in 1985, but there are many superior cables available today. My favorite interconnect is D.H. Labs' Air Matrix cable fitted with their RCA-HC Ultimate RCA Plugs. Like many high-performance interconnects, this cable is much too heavy to be connected directly to a tonearm, so I decided on a two-stage approach to the problem.

Cardas makes a 5-pin DIN-style tonearm base connector and a 4-conductor, 33 AWG shielded tonearm interconnect that is thin and extremely flexible. I use the Cardas interconnect to link the 5-pin tonearm connector to a pair of D.H. Labs CM-R1 RCA jacks mounted on the rear of the turntable base. The Cardas wire has no effect on the turn-

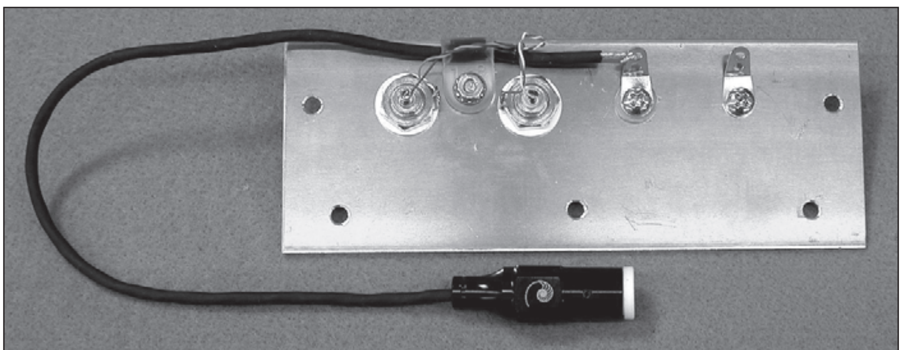


PHOTO 1: The assembled connector plate. New RCA jacks and ground posts are mounted on a piece of 2″ brass stock. The ground post closest to the RCA jacks is insulated from the brass plate.

table suspension. You can then use the interconnect of your choice—no matter how stiff or heavy—between the RCA jacks and your preamp's phono input. Although my instructions are tailored to the Galo/Merrill/Grado turntable, this approach is suitable for many other turntable designs, with and without a spring-suspended sub-chassis.

CONNECTOR PLATE

I mounted the new RCA jacks and ground posts on a brass plate, cut from a piece of 2″ wide brass stock, which I purchased in the hobby section of my local hardware store (**Photo 1**). Cut the brass stock 5 3/8″ long and drill the plate for the two RCA jacks (a 3/8″ hole is just right for the insulating washers supplied with the D.H. Labs jacks), two ground posts, plus a hole for a 4-40 machine screw, which will hold a nylon cable clamp between the two RCA jacks. You should also drill five mounting holes 5/16″ from the edge of the plate.

Mount the connectors and ground posts as shown in **Photos 1** and **2**. The ground posts are made with 6-32 × 5/8″ machine screws, #6 ground lugs, lock washers, hex nuts, and knurled thumbnuts. One ground post should be insulated from the brass with a nylon shoulder washer and nylon flat washer and mounted closest to the RCA jacks. The remaining ground post should make electrical contact with the brass plate.

Prepare a 12″ length of the Cardas tonearm cable by stripping about 1½″ of the outer insulation from one end and about ¾″ from the other. Carefully unravel the braided shield on each end, twist the strands together, and tin the ends. The Cardas center conductors are grouped into two twisted pairs. *Carefully* strip ¼″ of the Teflon insulation from the end of each conductor, twist the fine strands together, and tin them. Make the white/blue pair left signal and ground, and red/green pair right signal and ground.

Figure 1 is the wiring diagram. Solder the end of the Cardas cable—with ¾″ of insulation removed—to the Cardas DIN Phono Connector (the view in **Fig. 1** is facing the solder terminals on the Cardas connector). Solder the other end of the Cardas cable to the connector plate assembly. The shield should be soldered to the ground post that is insulated from the brass plate. Use a short length of heat-shrink tubing over the shield, and another short piece over the entire cable. Make sure that the cable clamp holds the cable firmly in place, and that there is no strain on the thin center conductors. After you've finished, use an ohmmeter to verify that the signal and ground connections match those of the original Grado tonearm cable.

DISASSEMBLY

You must disassemble the turntable in order to install the new tonearm wiring. Remove the outer platter, the belt, the counterweights, and head shell from the Grado tonearm. Then, remove the

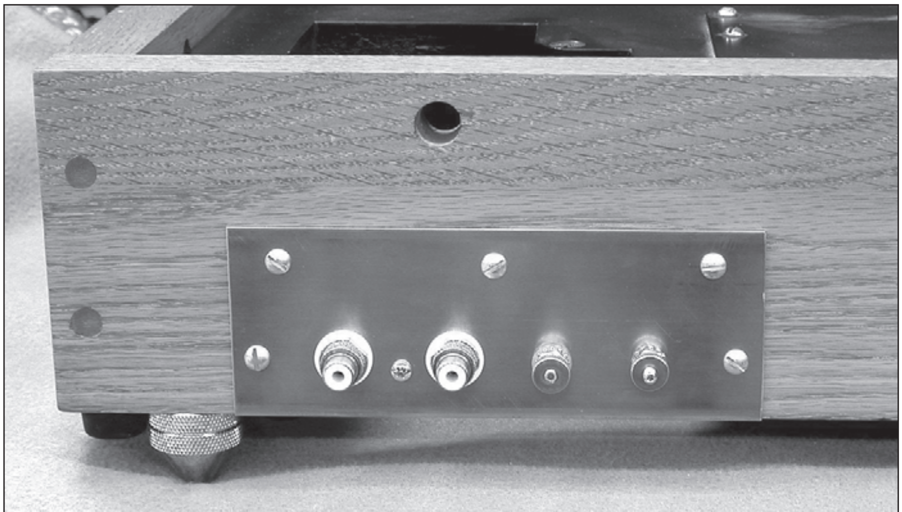


PHOTO 2: The connector plate mounted on the rear of the turntable base. Nearly any interconnect cable, no matter how heavy, can be used to connect the turntable to the preamp.

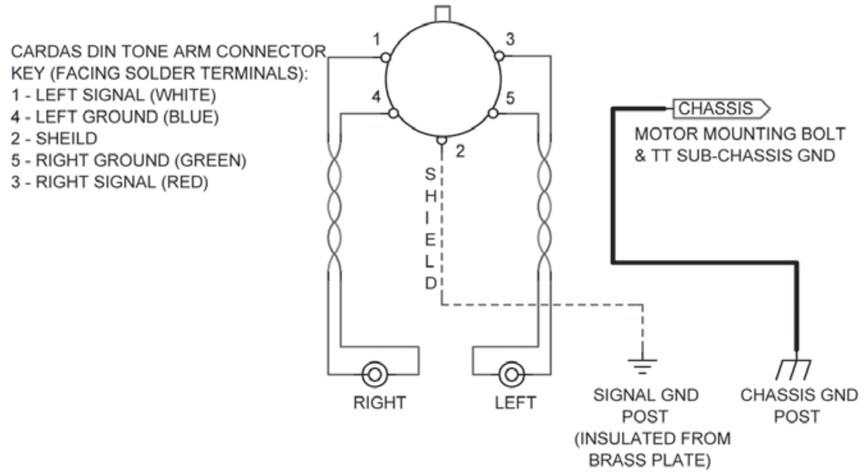


FIGURE 1: Wiring diagram for the belt-driven turntable upgrade. Separate chassis and signal ground posts accommodate a variety of installation requirements.

arm from its base, using the Grado instruction booklet as a guide (you don't need to remove the Grado arm base; it can remain fastened to the Merrill sub-chassis).

Remove the inner platter and wipe any oil remaining on the spindle with a soft, lint-free cloth. Store in a safe place. Disconnect the Merrill sub-chassis ground wire from the motor mounting bolt and remove the sub-chassis. Keep the sub-chassis upright so you don't spill any oil from the spindle well, and store the sub-chassis in a safe place, covering the spindle hole to keep out dirt.

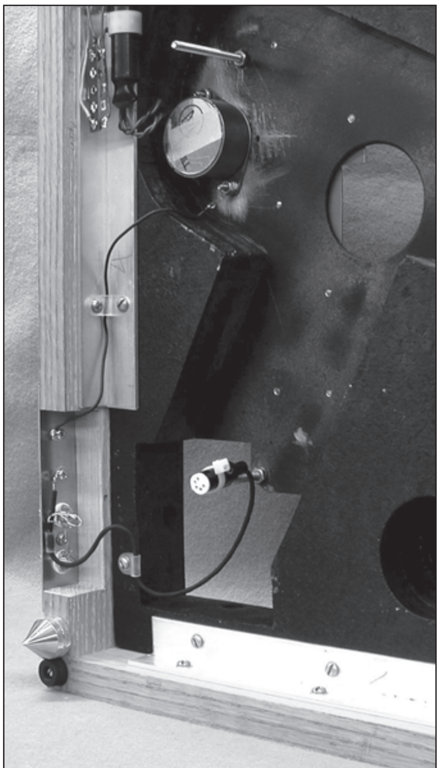


PHOTO 3: Bottom view of the turntable base with the new connector plate installed. The rear aluminum angle bracket is shortened to make room for the connector plate cutout.

As you face the rear of the turntable, the left edge of the brass plate will be 1 7/8″ from the edge of the oak turntable base (**Photo 2**). Remove the rear piece of aluminum angle stock that supports the top plate assembly (**Photo 3**). Make a rectangular cutout—about 1 3/8″ × 4 1/8″—in the back of the turntable base (**Photo 4**). Cut the aluminum angle piece so it doesn't extend beyond the rectangular cutout, and re-install. Include a 1/8″ cable clamp as shown in **Photo 3**.

Drill pilot holes for mounting the assembled brass plate, and mount with five #6 × 5/8″ brass wood screws. Mount another 1/8″ nylon cable clamp for cable support, as shown in **Photo 3**. The cable should be able to slide in this clamp, allowing you to adjust the position of the cable as needed.

Temporarily secure the Cardas DIN connector to the sub-chassis suspension bolt with a nylon cable tie, to prevent strain on the cable and connector. Connect an 18 AWG ground wire from the motor mounting bolt to the ground post closest to the edge of the brass plate. Route this ground wire through the first cable clamp. Use a #6 ground lug on the end connected to the motor mounting bolt.

BRASS CONES

With the new tonearm wiring arrangement, the turntable no longer needs the old sub base. You can re-use the old rubber feet of the main turntable base, but I prefer to replace them with something more effective at dissipating energy away from the turntable. A variety of suitable metal cones are available, including some that are rather exotic and expensive.

Parts Express carries its own Dayton brand of isolation cones that work very well, and a package of four costs only $25. The cone itself is a two-piece, threaded arrangement that is height-adjustable, which is handy for turntable leveling. The cones come with flat receptacles, which prevent the sharp points from damaging the surface below, and

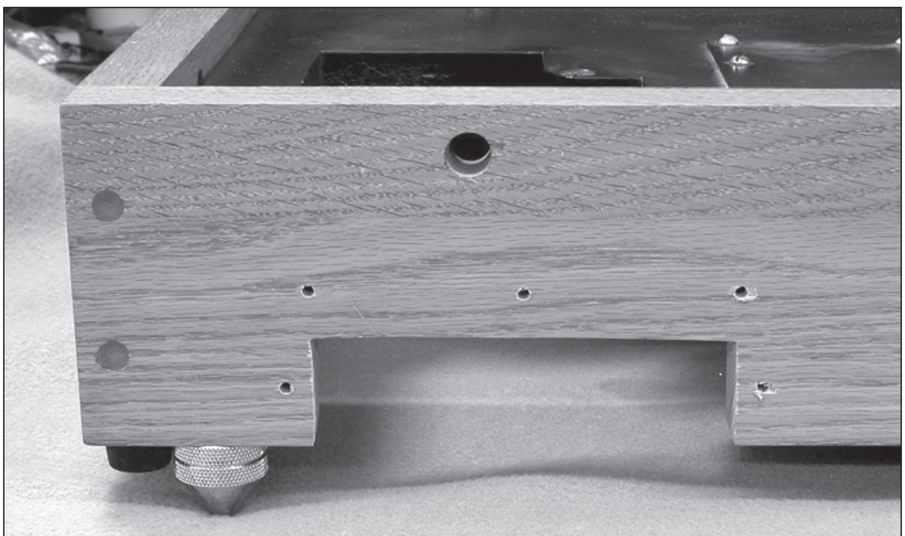


PHOTO 4: Rear view of the turntable base, showing the cutout and pilot holes for mounting the connector plate.

are available in either gold or chrome (gold is a better match for my oak turntable base).

The cones are also supplied with eight adhesive foam cushions, but I prefer to secure them with a *very thin* drop of silicone glue. I mounted the cones just inside the old rubber feet, as shown in the photos, because this was the best position for the top shelf on my equipment rack (I left the old rubber feet in place). If your turntable shelf is large enough, you can remove the rubber feet and install the cones in the four corners.

Long-time readers will note that I no longer use the turntable isolation shelf that I described back in *TAA* 1/87[5]. We moved to a new house in Sept. 2005, and my downstairs listening room has a concrete floor, so I have no need for it. If you have an unstable floor, I still recommend using some type of wall-mounted shelf.

After the brass cones are in position, let the weight of the turntable base hold them in place for 24 hours to allow the silicone glue to fully cure. Use the leveling feature of the cones to ensure that equal pressure is applied to all four. I also glued the receptacles to the shelf on my equipment rack. Use as little silicone glue as possible (the tiniest dab will be enough to hold the cones firmly in place), to avoid having a reactive coupling of the turntable to the cones and the shelf below.

REASSEMBLY

Remove the cable tie temporarily holding the Cardas connector to the sub-chassis mounting bolt. During reassembly, take care not to strain the cable and connector. Reassemble the turntable following the directions in "The Belt-Driven Turntable Revisited." Be sure to reconnect the ground wire from the Merrill sub-chassis to the motor mounting bolt.

You may need to add a small amount of Merrill Ultra High-Film Strength Turntable Oil to the spindle well (this is the oil originally supplied with the Merrill sub-chassis). Fortunately, George Merrill still has this oil in stock, if you need more. Contact him at George Merrill's Analog Emporium, or try Merrill-Scillia Research.

AR-XA drive belts stretch out over time and eventually need replacement. The best source is A-B Tech Services, a company supplying parts for old AR loudspeakers and turntables. The company is run by Alex Barsotti, a long-time employee of Acoustic Research. A-B Tech's parts are manufactured to original specifications, and are often less expensive than the substitutes sold by other dealers.

Replacement belts normally have a powdery white coating of talc already applied. If yours doesn't, be sure to talc the belt before re-installing, using pure talc, *not* talcum powder.

After you reinstall the Grado tonearm, plug the Cardas connector into the base of the arm. It will be a tight fit. The Cardas cable will form a small loop under the tonearm. Adjust the position of the Cardas cable in the cable clamp as needed. Follow the original Grado Signature ToneArm setup instructions to the letter.

Once you have re-assembled the turntable, connect it to your preamp with your choice of interconnect cables. As mentioned, I am extremely pleased with the performance of D.H. Labs' Air Matrix Interconnect Cables fitted with their RCA-HC Ultimate RCA Plugs, which you can buy from Audio Concepts, the well-known loudspeaker manufacturer. The Air Matrix cables are not sold in bulk form—only pre-assembled.

For those who wish to assemble their own interconnects, I recommend D.H. Labs' Pro Studio Interconnect Cable, plus the RCA-HC Ultimate RCA Plugs, both available from Parts Connexion. Parts Connexion also carries a D.H. Labs cable custom-manufactured for them, the BL-Ag, which consists of 23 AWG solid-core silver center conductors, a foamed Teflon tape-wrapped dielectric, plus a foil shield and drain wire. I have not tried it, but it's certainly worth investigating. The BL-Ag is too small for the Ultimate RCA Plugs—D.H. Labs recommends its RCA-2C for this cable. For a cost-effective solution, D.H. Labs' BL-1 fitted with RCA-2C connectors is still hard to beat for the money.

Normally, I recommend putting a jumper between the two ground posts, and running a single ground wire to the preamp. Make the ground and jumper wires with 18 AWG stranded, insulated wire and crimp-on spade connectors. My system is silent using this grounding arrangement.

However, some preamps or integrated amplifiers may yield lowest noise with separate ground wires from the tonearm and the turntable chassis ground posts. The two-ground post arrangement allows maximum flexibility for a variety of installation requirements.

The upgraded turntable is shown in **Photo 5**. The number of high-priced turntables on the market continues to amaze me, with some stretching well into the five-figure range. With the tonearm wiring upgrade, and a new stylus or cartridge, the Galo/Merrill/Grado turntable should continue to provide musically satisfying sound from your vinyl collection for years to come, without breaking your bank account. *aX*

PHOTO 5: The upgraded belt-driven turntable. Dayton adjustable brass cones replace the old turntable sub-base.

REFERENCES

1. Galo, Gary, "AR System Drives New Turntable," *Audio Amateur*, 3/85, pp. 27-38.
2. Galo, Gary, "An Electronic Speed Control," *Audio Amateur*, 1/86, pp. 7-13.
3. Galo, Gary, "The Belt-Driven Turntable Revisited," *Audio Amateur*, 3/88, pp. 24-30 (including the sidebar "Sub-chassis Design and Hum Shielding Considerations" by George Merrill).
4. *The LP Is Back!* Peterborough, NH, Audio Amateur Press, 1999 (available from www.audioxpress.com).
5. Galo, Gary, "A Turntable Isolation Shelf," *Audio Amateur*, 1/87, pp. 45-47 (Not reprinted in *The LP Is Back!*).

VENDORS

Grado Laboratories
4614 Seventh Ave.
Brooklyn, NY 11220
(718) 435-5340
www.gradolabs.com
info@gradolabs.com

Audio Advisor
3427 Kraft Ave., S.E.
Grand Rapids, MI 49512
1-800-942-0220
www.audioadvisor.com

Needle Doctor
419 14th Ave., SE
Minneapolis, MN 55414
800-229-0644 (orders only)
612-378-0543
www.needledoctor.com
info@needledoctor.com

LP Gear
2227 Double Tree Ave.
Henderson, NV 89052
www.lpgear.com
LPGear@yahoo.com

D.H. Labs
9638 NW 153rd Terrace
Alachua, FL 32615
386-418-0560
www.silversonic.com
dhlabs@silversonic.com

Parts Connexion
2885 Sherwood Heights Drive
Unit #72
Oakville, Ontario
Canada L6J 7H1
1-866-681-9602
www.partsconnexion.com
info@partsconnexion.com

Audio Concepts, Inc.
PO Box 26
Onalaska, WI 54650
608-786-4729
www.audioc.com
sales@audioc.com

Parts Express
725 Pleasant Valley Dr.
Springboro, OH 45066
800-338-0531
www.partsexpress.com
sales@parts-express.com

Mouser Electronics
1000 North Main Street
Mansfield, TX 76063
800-346-8673
www.mouser.com
sales@mouser.com

George Merrill's Analog Emporium
820 Herbert Rd. #109
Cordova TN 38018
901-751-3337
www.gmanalog.com
hifigem@gmanalog.com

Merrill-Scillia Research, LLC
PO Box 539
North Branford, CT 06471
901-619-3322
www.merrillaudio.com
info@merrillaudio.com

A-B Tech Services
800-225-9847
www.abtechservices.com
abtechservices@mindspring.com

Parts List

(1) piece 0.064″ thick brass stock, 2″ wide × 5 3/8″ long, K&S Engineering #14121-10249 or equivalent (local hardware store or hobby shop)
Cardas SDIN, 5-pin DIN tonearm connector, straight (Parts Connexion #53481)
Cardas 4 × 33 AWG tonearm wire with shield, 12″ (Parts Connexion #64312)
(1) pair DH Silver Sonic CM-R1 chassis RCA jack (Parts Connexion #68098)
(2) each 6/32 × 5/8″ machine screws, lock washers, nuts and knurled thumbnuts (local hardware store)
(1) each nylon shoulder washer and flat washer (local hardware store)
(3) #6 ground lugs (Mouser #534-7312)
(3) 1/8″ polypropylene cable clamps (Radio Shack #64-3028)
(5) #6 × 5/8″ brass wood screws (local hardware store)
1/8″ heat-shrink tubing (Radio Shack #278-1627)
Dayton ISO-4G (gold) or ISO-4C (chrome) Isolation Cones (Parts Express 240-720 or 240-721)
D.H. Labs Air Matrix Interconnect Cables (Audio Concepts)
or
D.H. Labs Pro Studio bulk interconnect cable (Parts Connexion #64786 or Audio Concepts) and RCA-HC Ultimate RCA Plugs (Parts Connexion #64434 or Audio Concepts)
or
D.H. Labs BL-Ag bulk interconnect cable (Parts Connexion #66070) and RCA-2C RCA Plugs (Audio Concepts)
or
D. H. Labs BL-1 interconnect cable (Audio Concepts) and RCA-2C RCA Plugs (Audio Concepts)

Miscellaneous: 18AWG ground wire; crimp-on spade connectors (Radio Shack #278-1220 & #64-3124)

By Chi C. Wong

An Input/Driver for a **Hybrid Power Amp**

Here's an interesting circuit for use in a hybrid amp up to 150W.

I have seen many hybrid amplifier designs in recent years. The topologies look similar: with an SRPP (or its brothers/sisters) as the input/driver and power MOSFETs connected in a typical push-pull configuration at the output[1, 2]. I have built a few hybrid amps with this basic arrangement over the past five years using the input/driver circuit described in reference 3. I am reasonably happy with the basic design; however, there are at least three limitations that I wish to improve.

1. Voltage gain. Essentially fixed by the valve, but I prefer more flexibility.
2. Output impedance. Usually on the order of around 1Ω (dependent on the output devices), which is a bit high for a solid-state design.
3. Total harmonic distortion (THD). On the order of 2-3% at rated power, which could be much better.

In order not to make the design too complex, I chose to use two triodes (one glass envelope) as in the SRPP version.

This article describes how to design the driver/input stage to improve these conditions. It is aimed more at experimenters than first-timers, devoting more space to design principles than practical implementations.

CHOOSING THE CIRCUIT AND TUBE

The SRPP and variants are popular, but are not necessarily the best choice in a hybrid amplifier. A typical common-cathode and cathode-follower pair may, arguably, be better[4]. Anyway, all these versions cannot improve the issues I mentioned.

A possible arrangement is to use two common-cathode stages in cascade. This configuration was popular 40 years ago in both power and preamp circuits, including the venerable Marantz 7. A two-stage common-cathode amp has too much gain and requires substantial global feedback to reduce it accordingly.

To some users, feedback in audio is sinful. I do not think that feedback is a bad idea provided that the open-loop behavior of the base amp is good. Feedback can make it even better.

I decided to build around a two-stage design based on the common-cathode configuration. With a 70W hybrid power amp in mind, I set some major design criteria (for the input and driver circuit):

- Voltage gain: 26-32dB
- Output resistance: 500Ω or less
- Bandwidth (-1dB): 50kHz or better
- Output voltage swing: 70V peak-peak minimum
- THD: less than 0.5% at rated output

Power MOSFETs have high input resistances; however, their inter-electrolyte capacitances are high, and, consequently, the driver must have low output impedance and a bias current of 10mA or more. Commonly used valves (tubes) such as 12AU7, 6922 are not suitable candidates unless many of them are connected in parallel. I prefer not to parallel active devices, whether for valves or power MOSFETs.

Fortunately, there are many valves on the market that can do the job. A suitable choice is the 5687 and substitutes (more on this later). The output im-

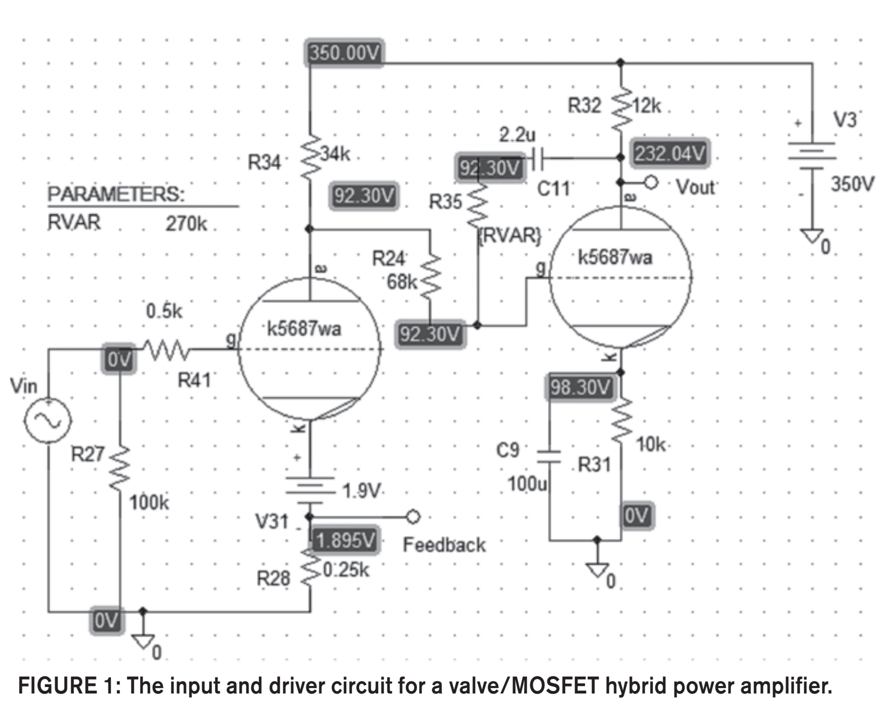


FIGURE 1: The input and driver circuit for a valve/MOSFET hybrid power amplifier.

pedance of a common-cathode stage is mainly determined by the valve's plate resistance r_p and is around 2.5kΩ for a 5687. Therefore, local feedback is required to bring down the output impedance. With the selected circuit and valve, I can now finalize the topology (**Fig. 1**).

STAGE DESIGN

Designing the component values is relatively simple. I first check the typical operation conditions of the valve (5687) and select the following (you can easily download datasheets from reference 5):

- $V_{PK} \approx$ 90V, $I_P \approx$ 7.5mA for the input valve;
- $V_{PK} \approx$ 120V, $I_P \approx$ 10mA for the driver valve.

The first stage, a typical common-cathode amplifier, is directly coupled to the second, and therefore V_P of the first becomes V_G of the second. V_G (of the driver) must be a few volts higher than the corresponding V_G. This helps to determine the supply voltage (around 350V). Simple Ohm's law then determines R31 and R32. The bias voltage of the input valve is around 3.8V. This is achieved by a green LED in series with a suitable resistor (=1.9V/7.5mA≈0.25k). Using Ohm's law again, I can find the value of R34. C9 is used to bypass R31. All relevant component values are shown in **Fig. 1**.

The second stage is a modified common-cathode amplifier. The R35 and R24 combo forms the feedback circuit. Academically, this is a "transimpedance" amp. Some useful design formulas (estimations only) are

- Stage gain (A_{vS}) ≈ R35/R24
- Output impedance ≈ $r_p(A_{vS}/\mu)$, where μ is the valve's amplification factor
- Input impedance ≈ R24 (worst-case)

This type of feedback, though not new, is uncommon in audio circuits. I think the major drawback is that the input impedance loads the preceding stage and hence adversely affects the gain and distortion. To increase R24 seems to be a good solution, but this will compromise the high-frequency response. I consider a suitable range of R24 is from 47k to 82k.

The second step is to fix the stage gain, which determines R35. In the circuit of **Fig. 1**, with R35 equal to 270k, the gains of the first and second stages are estimated to be 10 and 4, respectively. I can vary R35 between 120k to infinity (no feedback) as I wish. C11 should be increased if R35 goes below 120k.

SIMULATION AND MEASUREMENT RESULTS

Basic design equations do not accurately predict the amplifier's performance. A circuit simulator becomes extremely helpful in this department. I use the student version of PSPICE (freely available from www.orcad.com) to evaluate the amp's characteristics.

The DC bias voltages are shown in **Fig. 1**, while the plate current of the input and driver valves are 7.6mA and 9.8mA, respectively. **Figure 2** depicts the frequency response of the amp with

Table 1 Characteristics of the amplifier—from simulation.

R35	Gain (dB)	-1dB Bandwidth	Zout at 100Hz	THD at 1kHz and 35V peak output
120kΩ	24.5	135kHz	335Ω	
270kΩ	30.0	72kHz	493Ω	0.25%
infinity	40.2	20kHz	1.27kΩ	
120kΩ	*24.6*	*133kHz*	*291Ω*	
270kΩ	*31.60*	*67kHz*	*457Ω*	*0.13%*
infinity	*45.2*	*14kHz*	*1.75kΩ*	

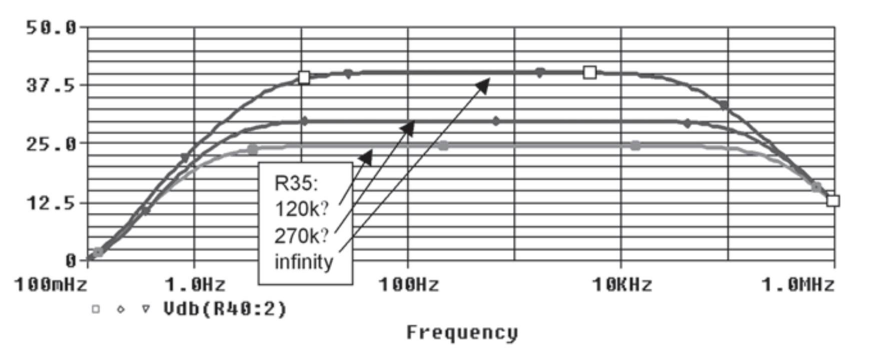


FIGURE 2: Frequency response of the amp. Vertical axis is the gain in dB.

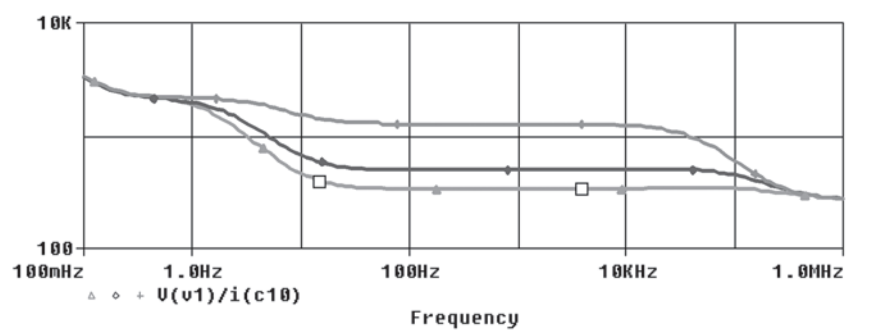


FIGURE 3: Output impedance of the amp. Vertical axis is in ohms.

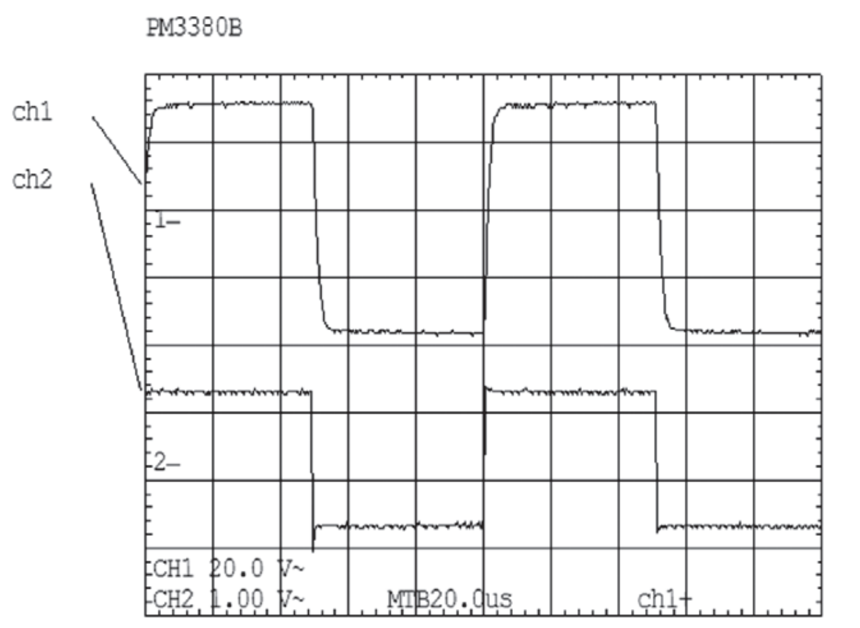


FIGURE 4: Square-wave response of the amp in *Fig. 1*. Ch2 is the input = 1V peak.

three different values of R35. **Figure 3** shows the effect of R35 on the output impedance. **Table 1** shows the details including a high current version (in italics) discussed in a later section.

A real-world test of the circuit revealed the following measurement results (with R35 = 270k):

- Gain: 35 (30.9dB)
- Bandwidth (-1dB): 50kHz
- THD+N (on a HP334A distortion analyzer) at 35V peak o/p: less than 0.24%, 0.1kHz to 10kHz

Figures 4 and **5** contain the square wave and triangular wave responses, respectively, both at 10kHz and around 35V peak. **Figure 6** illustrates that the amplifier is capable of generating a voltage of at least 100V peak-peak without clipping or obvious distortion. Note that noises superimposed on the waveforms are from the digital scope (Fluke PM3380B) itself. No noise can be detected in analog mode.

Some designers/authors do not want users to change the bias conditions or component values. I believe design should be flexible. Provided that some basic design principles are adhered to, you can always modify the circuit as you wish, and that is the fun part.

BIAS CONDITIONS

I have designed a higher current version with R31 and R32 equal to 4.7k and 5.6k, respectively, such that I_P of the driver valve becomes 18mA. Simulation results (**Table 1**) show that this is better in almost all aspects, including gain and distortion. However, I prefer not to tax the lifespan of the valve, although 5687 is capable of dissipating such a power. This is a good alternative if you want to run your valves hot.

You can also change the bias condition of the first valve, but it is better to keep its plate voltage to within 100V. The circuit should perform well with a supply voltage ranging from 310V to 360V. You will need the latter if the circuit is driving a 150W output stage.

VALVES

I use the 5687 tube because I have many of them in my collection. I have used E182CC (7119), which may be considered a direct replacement. Measurements show that the DC conditions are similar, while AC gain is slightly higher and with a better bandwidth:

Gain = 41 (33.2dB) and bandwidth (-1dB) = 55kHz.

ECC99 is another suitable candidate that you can use with an identical design; however, the pin layout is different (so be careful). 6H30 is a very powerful triode that will suit the need well, but the design must be modified.

My design uses the same triode type for the input and driver. Obviously, this is not necessary. You will open up even more choices if you remove this limitation. An essential requirement for the input valve is that it should have a low r_P (so 12AX7 and 6SL7 tubes are out).

LOCAL FEEDBACK VS. GLOBAL FEEDBACK

A push-pull power output stage has unity (or slightly lower) gain with a wide bandwidth and low distortion. Therefore, the input and driver tend to dominate the performance of the amplifier. You can now adjust R35 to your taste. The effect is summarized in **Table 2**. Reducing R35 increases local feedback and reduces gain. On the other hand, this widens the bandwidth, reducing the output impedance and distortion. You can always experiment with R35 ranging from 120k to open-circuit.

I prefer a relatively low output impedance from the driver, but some may suggest that the push-pull output stage is basically a source-follower circuit, and hence the input capacitance is not really an issue for the driver. Note that local feedback is not the same as global feedback; a source-follower uses 100% local feedback, for example.

Unfortunately, I cannot reduce the output impedance of the complete hybrid amplifier by toying around with the input/driver circuit alone. I must use global feedback to achieve this. The use of global feedback has been out of fashion in recent years. In my opinion, global feedback is not necessarily bad, provided that the open-loop behavior of the amp is good. In this case, the circuit is already better than an equivalent current-source driven SRPP[2, 3] and can work perfectly without any global feedback.

For the sake of completeness, **Fig. 7**

Table 2 Effect of feedback on amplifier performance.

R35	Feedback	Gain (Av_S)	Bandwidth	Output impedance	Distortion
↑	↓	↑	↓	↑	↑
↓	↑	↓	↑	↓	↓

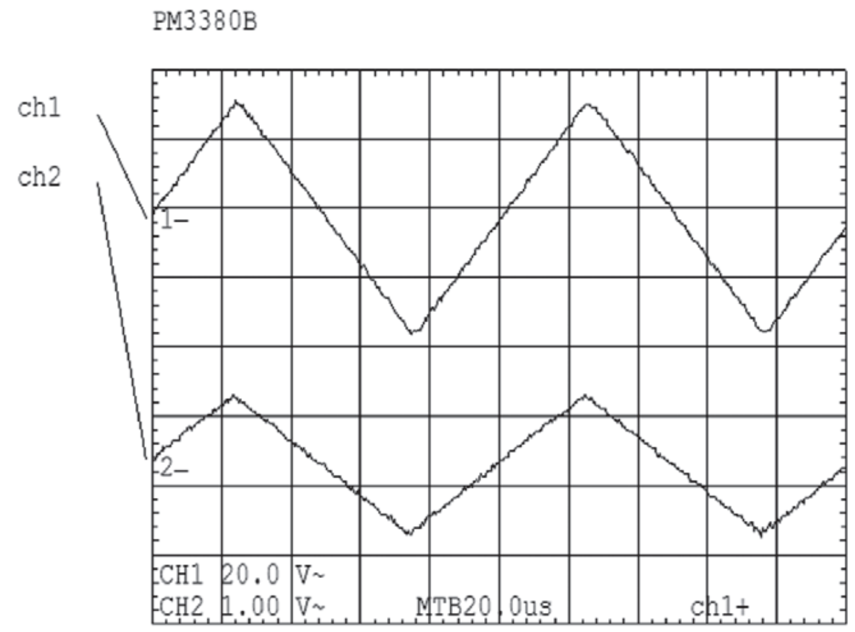


FIGURE 5: Triangular wave response of the amp in *Fig. 1*. Ch2 is the input = 1V peak.

shows the output stage with the associated bias circuit. I use a pair of Exicon MOSFETs ECF10P16/ECF10N16, and the rated output of the complete amplifier is 60W at 8Ω. I apply 7dB of feedback (via Rf with a parallel 100p capacitor—not shown) that reduces the output impedance to below 0.5Ω. The overall gain is 23dB with a (-1dB) bandwidth of over 100kHz. Distortion is even lower than that depicted in **Table 1**. I believe a small amount of feedback is beneficial. Of course, you can experiment to suit your taste. This is a non-inverting amp.

CONCLUSION

I have described the circuit using technical, objective, and measurable terms. . . until now, when I must face the inevitable question: *How does it sound?* I think it measures and sounds better than an equivalent SRPP-driven unit. It is clearer and more open.

After trying E182CC in place of 5687, I was left with the impression that the former is a bit more powerful, maybe even clearer; however, it may occasionally sound marginally harder. 5687 is easier to source and is cheaper, too. I have also tried to use 6DJ8 as the input and ECC99 as the driver, but somehow I am not particularly interested (subjectively) in this arrangement; perhaps you may find a better combination. The circuit is fairly flexible, so readers should feel free to experiment. *aX*

REFERENCES

1. "A Modular Hybrid Amp System," M. Ferretti, *audioXpress* 2/01.

2. "Amplifier War and Peace," S. Brown, *audioXpress* 3/07.

3. "Design a High Performance Valve-Based Line Amplifier", C.C. Wong, *audioXpress* 10/05.

4. "The Cathode Follower and Its Weaker Siblings," C. Paul, *audioXpress* 1/07.

5. www.r-type.org/pdfs/5687.pdf.

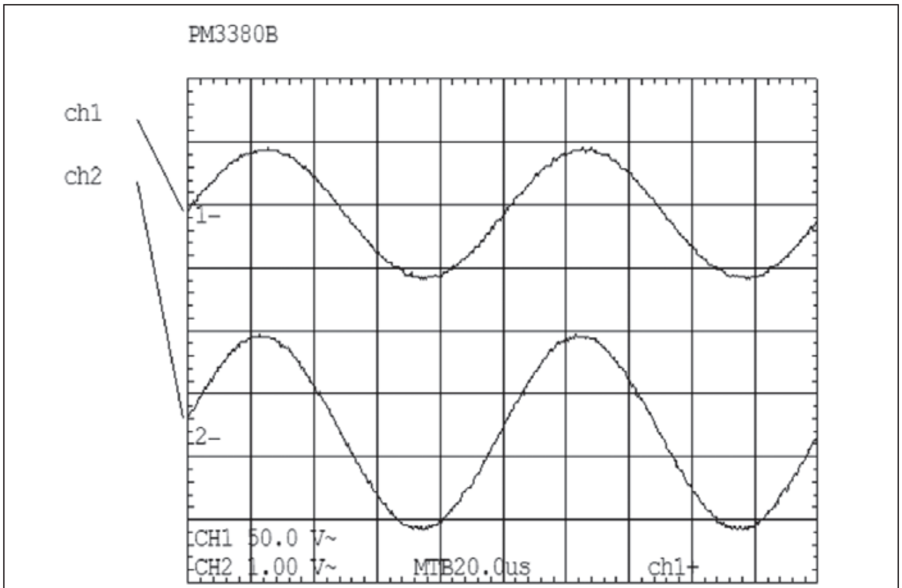


FIGURE 6: Undistorted output voltage at 100V peak-peak.

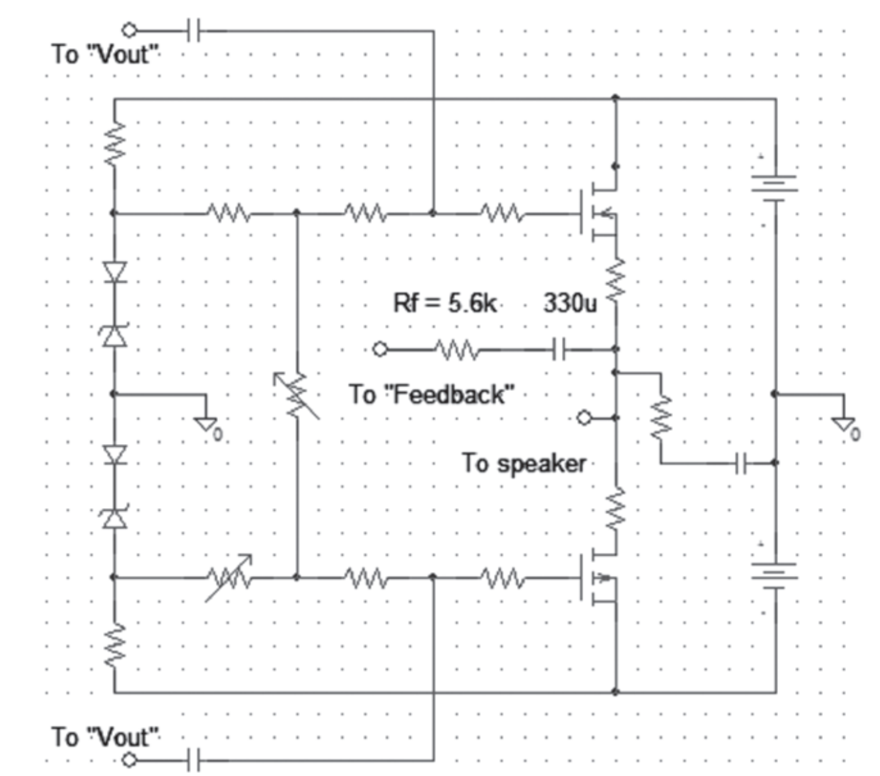


FIGURE 7: The power output stage.

Scrounging 101

Even today, vintage audio treasures abound. . . if you know where to look.

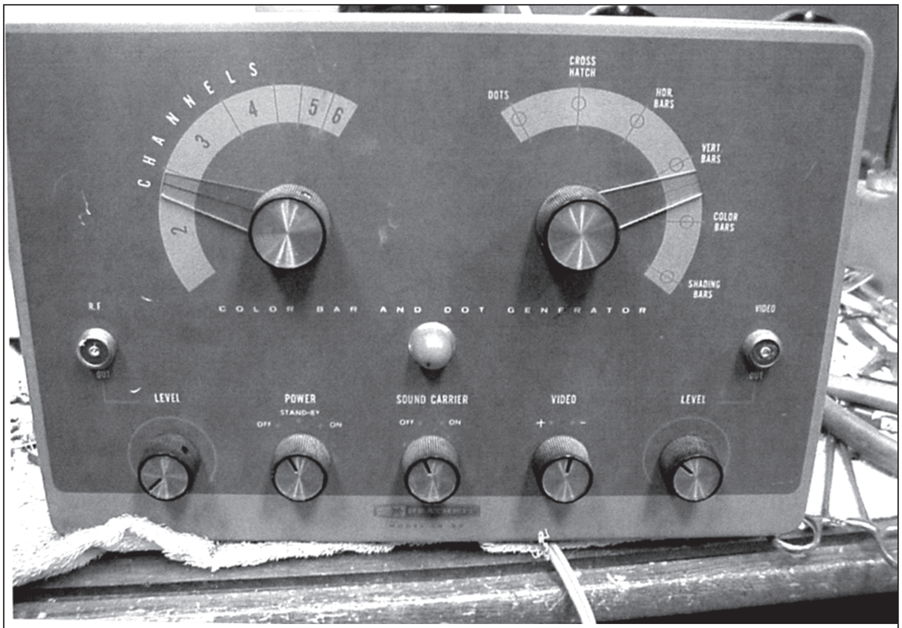


PHOTO 1: Heathkit IG 62 ripe for salvaging.

I was reminiscing with my friend, Larry, the other day about how we got started in the hobby of electronics. Many of you probably have similar stories. By the time I was 10 years old, I was hauling home TVs, radios, and any other electronic "junk" I found discarded by the side of the road on trash pickup days. I would spend hours dismantling, sorting, and organizing the parts I collected.

Even though I may date myself, I spent hours at the drug store testing tubes. Family friends would bring me stuff to "fix," and many times I was able to do so. Back in those days the wiring and layout were designed to be easy to service and pretty straightforward. I had cigar boxes full of resistors and capacitors, boxes of transformers and tubes, and even salvaged hook-up wire—a much simpler time.

Today, I still find my way to electronic and surplus stores to scrounge through old stuff looking for treasures. Recently, I bought an old piece of test equipment and when I started tearing it apart, it struck me that this is a very enjoyable part of my hobby and allows me to inexpensively build tube audio equipment. I thought I would share this with *audioXpress* readers. I am going to break this down and present the economic advantage as well as the ecology benefit to scrounging.

ONE MAN'S JUNK. . .

I found a Heathkit IG 62 Bar Dot Generator (**Photo 1**) at a surplus house and paid a princely sum of $15 for it. You can also find a huge amount of old unused test equipment at estate sales and online. Obviously, this generator is not of much use to anyone in today's world of digital and HD TVs. Upon removing the case, I was surprised at the number of tubes and parts in this unit—pretty hi tech for its day.

I am not saying that all the old stuff out there will yield the kind of bounty this did, but it is pretty typical of old tube test equipment. This piece of equipment was full of brand-new-looking Mullard tubes. Upon closer examination, I found that there were 4-12AU7s and 6-12AT7s, all Mullards! Pulling out the trusty tube tester, I proceeded to check them out. They all tested like new, and the sections were extremely well balanced. You really must love the old high-quality manufacturing standards of that day!

When I started testing the other tubes, I found the exact same thing, with the exception of the OA2 regulator tube, which was shorted. This explained the lack of a fuse in the fuse holder. (I would never plug in an old piece of test equipment to see whether it worked—as you know this can cause a lot of damage to components that haven't seen electricity in many years.) If this were an audio amplifier, I might bring it up slowly on my trusty Variac® to see whether it worked, before I rebuilt it or dismantled it for parts.

DISASSEMBLY

When you start your dismantling project, here are a few hints:

1. You can clip long leads close to their connect points and clean the terminals or sockets later (I will give you a couple of methods of doing this).
2. Unsolder short connections with a good 50W iron (I have an old one I use just for this—you can also use a Weller type 100/140W gun if you are careful).
3. You will need some sturdy needle-nose pliers and some hemostats/forceps/heatsinks.
4. Never try to reuse electrolytic capacitors unless you have a capacitor test meter and reformer (I built one from an older issue of *Glass Audio* and it

works great), although I reuse very few.
5. Don't try to salvage or reuse paper and wax caps, and unless you have a cap meter to check value and leakage, don't even try to reuse black or sealed type (more on this later).
6. Mica caps and most ceramics will be fine.

I usually start by clipping all the components I can and separating them into small plastic cups. Remove any transformers and chokes and label their leads, especially if they are not the universal color-coded ones. I also save all the vintage hardware by size in a small portioned clear plastic box.

I recently wrote an article for *audioXpress* about a vintage amp I built ("Upgrading a Classic Pilot Amp," Apr. '08, p. 8). Many of the components of this amp were salvaged items and it truly looks the vintage part.

Once this is done, you can begin removing the cut leads from the sockets, terminal strips, and controls (I do this before removing them from the chassis, to keep things stable). I do this by heating the connection with my big iron or gun, and when the solder is molten, grabbing the individual leads one at a time and removing them. This can be easy or hard depending on the original builder of the item. Some builders just looped the wire through the various terminals, so they're very easy to remove. Some people wrapped the leads around the terminals—even two to three times in some cases—and crimped them tightly. These can be difficult to remove, but with a little careful work it can be done.

My favorite tool for this is a heavy, short pair of stainless-steel surgical hemos, which have a very sharp rounded nose about 3/8″ long and about ¼″ wide. They are heavier than your run-of-the-mill hemos, and are high quality US made (I wish I could remember where I got them so I could buy several more). You can also use quality medium-size needle-nose pliers. You want something that has very accurately aligned jaws and sharp serrations. Take your time and don't make this a battle.

Larry has a very effective way of dismantling equipment, but it requires safety glasses and long sleeves and preferably gloves and a hat. He hooks up his air compressor to a blow nozzle with a pushbutton air valve, heats up the connection with his Weller gun, and blows a sharp blast of air at close range, while the solder is molten. This sprays solder everywhere (hence the protective wear), and so you need to do this in a shop or area where you can contain the residue (older solder contains lead). When he has finished blowing all the solder away, it is a real easy job to heat the joint and remove any wiring on the sockets and terminals.

Now you have a chassis with terminals and sockets splattered with solder. There are two ways to remove this. Larry places the chassis in a bead blasting cabinet in his shop and gives it a quick blast. This not only cleans up the solder mess, but also removes any corrosion that may be on the terminals. You need to remove any pots or rotary switches before you bead blast. If you don't, you will most likely ruin them (or at least make them real noisy). As I said, this is his method, it is dangerous, and you must wear eye and skin protection should you choose to use it.

SALVAGE SAVINGS

Once the solder is all removed (**Photo 2**) you can begin the mechanical salvage of the chassis. Remove all the hardware, inspect each socket and terminal, and keep what is good.

I have many large plastic bin cabinets and as I sort parts to these, I check them before I put them in. All the capacitors get tested for value and leakage and if they don't pass they are gone. I sort resistors, but don't check their value, because I routinely check them when I am building or repairing equipment. I cleaned switches and pots with Detoxit™ and tested them.

At this point you probably have a pretty bare chassis, case, and front panel. I usually keep these also. You can reuse the cases, the front panels, and the chassis for prototyping or actual construction. The vintage 6L6 amp I mentioned earlier used a chassis from an old telephone amp. Sometimes part of the fun is making use of something rather than

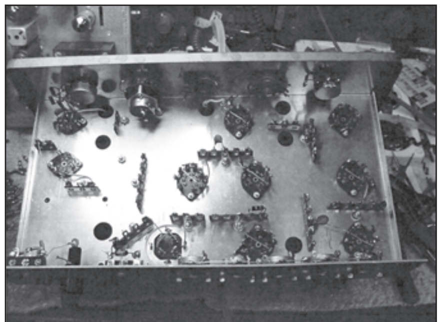


PHOTO 2: Solder removed.

PHOTO 3: Parts separation complete.

have it end up in a landfill.

You can see from **Photo 3** that there is a considerable pile of parts from this little salvage. Just for fun I went through them and got an approximate cost for new parts. (I used Antique Electronic Supply catalog for prices.) The results are:

-Power transformer (this has already been used to make the power supply for an amp project I will be submitting, using subminiature tubes)	$35
-Fuse holder	$1
-Line cord	$.50
-Strain relief	$.50
-6 Mullard 12AT7 tubes (I sold 5 of these on eBay)	$56
-4 Mullard 12AU7 tubes	$80
-1 6BQ7A tube	$2
-1 6CS6 tube	$3.15
-1 OA2 tube (bad)	
- 2 .015µF 400V 10% ceramic caps	$1
-3 0.5µF 200V Sprague Black Beauties	$3
-1 .25µF 400V Sprague Black Beauty	$1
-2 .01MF 600V Sprague Black Beauties	$1
-1 1 0.1µF 200V Sprague Black Beauty	$1
-3 10W power resistors	$1.85
- 3 Crystals and sockets	$3
-22 Mica caps 200-600V	$11
-2 4.7pF HiQ capacitors	$1
-25 ceramic disc capacitors	$4.50
-61 ½W carbon comp resistors	$15.25
-1 1W carbon comp resistor	$1.50
-1 2W carbon comp resistor	$1.50
-2 germanium diodes	$3.50
-5 nice 1¼″ knobs	$10
-2 nice 1½″ knobs	$4
-1 1MΩ pot	$2
-1 200Ω pot	$2
-4 wafer switches	$8
-2 jacks	$4
-1 fuse holder	$1
-1 6V pilot light assembly	$2.25
-16 terminal strips assorted sizes	$6
-10 9 pin wafer sockets	$20
-2 7 pin wafer sockets	$4
-1 9 pin molded socket	$2.25
-8 grommets	$2
-1 small variable cap	$10
-2 trimmer caps	$1
-5 ground lugs	$1
-4 section can cap (may or may not be good)	
-Aluminum chassis	
-Aluminum sub chassis (see note later)	
Total	$307.75

I paid $15 for the generator and salvaged $307.75 worth of parts from it. Even if you just take the original $15 investment and subtract from this the $56 I got for five Mullard 12AT7s, that is a profit of $41. Not too bad for getting a bunch of free parts.

RECYCLING

So the question is: Do I use these parts? You bet I do. Many of my vintage audio restorations require older carbon comp resistors. Pots and switches are always in demand. I already used the 6V pilot light in a guitar amp restoration. I also used the power transformer, fuse holder, power cord, some grommets, and terminal strips for a power supply for a pair of SE amplifiers using subminiature tubes.

As I said earlier, I use the chassis from this type of salvage for prototyping and even for finished projects. You can usually arrange your components to use factory holes and cover the sides or end panels with some wood. For example, **Photo 4** shows the sub-chassis after cutting some scrap pieces of maple, and putting it together with some glue and an air nailer. A little sandpaper, some spray clear finish, and in 10 minutes you have a perfectly good chassis for building a small tube amp project. (I am thinking a headphone amp.)

I am not suggesting that you go out and madly buy up old test equipment and try to make a living selling the pieces on eBay. What I am suggesting is that you can get a lot of really high-quality components, tubes, and transformers for other projects. This will not only save you money, but help the environment. *aX*

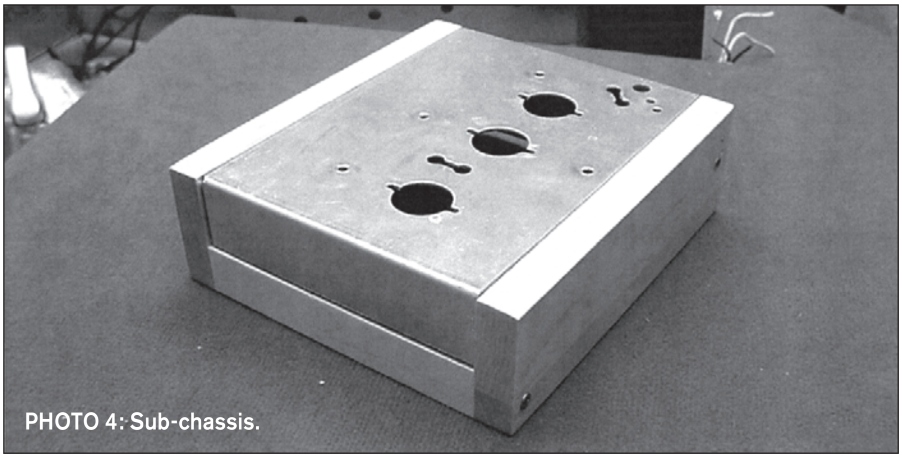


PHOTO 4: Sub-chassis.

Passive Crossover Linear Phase Speakers, Part 2

The author continues his in-depth study of using a passive crossover for linear-phase speakers.

I decided I wanted to try a linear-phase system. I intended to use a three-way with passive CO. This pretty much limited me to the first-order CO approach. I tried to develop a system with three drivers, but the results were poor, and I fortunately did not build such a system.

DECISIONS

As developed previously, the solution is a coaxial driver or a symmetrical array. Not aware of a tri-axial driver, I determined that the only viable answer was the symmetrical five-driver array—woofer over midrange over tweeter over midrange over woofer (WMTMW). George Augspurger informed me that the phase-linear monitors for studio work used arrays of five drivers in the WMTMW configuration. That decided it—I would build with the WMTMW approach.

I also knew that the acoustic origins would need to be aligned, meaning a curved or stepped front panel. Not knowing how to build a curved front panel or how to model the system with the drivers mounted at an angle, I chose a stepped front panel. To match the front panel, it would be nice to either curve or step the sides of the enclosure, but then I couldn't figure out how to apply the grille cloth. Thus I decided to go with straight sides on the enclosure and try to fight any diffraction/echo problems with damping material on the front panel.

Because the first-order CO requires such wide overlap in the woofer and midrange response, I figured a 6½″ woofer was the largest I could use. This meant the bass would be restricted. The remedy was to cross over to a subwoofer with an electronic CO. Thus the system would not be phase-linear over the full audio band, just from a couple hundred hertz up. The intent was to keep the frequency of crossing over to the subwoofer below 100Hz.

I developed a design with the three drivers having the responses that were shown back in Fig. 10. Fortunately, the tweeter is 4Ω and has the needed 6dB sensitivity advantage over the midrange driver. The woofer, with a slightly higher sensitivity than the midrange, could be used to compensate diffraction spreading loss at low frequency.

Figure 24 depicts the modeled performance of this system. The low-frequency rise is the diffraction-spreading-loss compensation. From about 200Hz up, the system is essentially linear phase on the tweeter axis. The polar magnitude response does not look bad, with few angles showing any higher magnitude than on the tweeter axis.

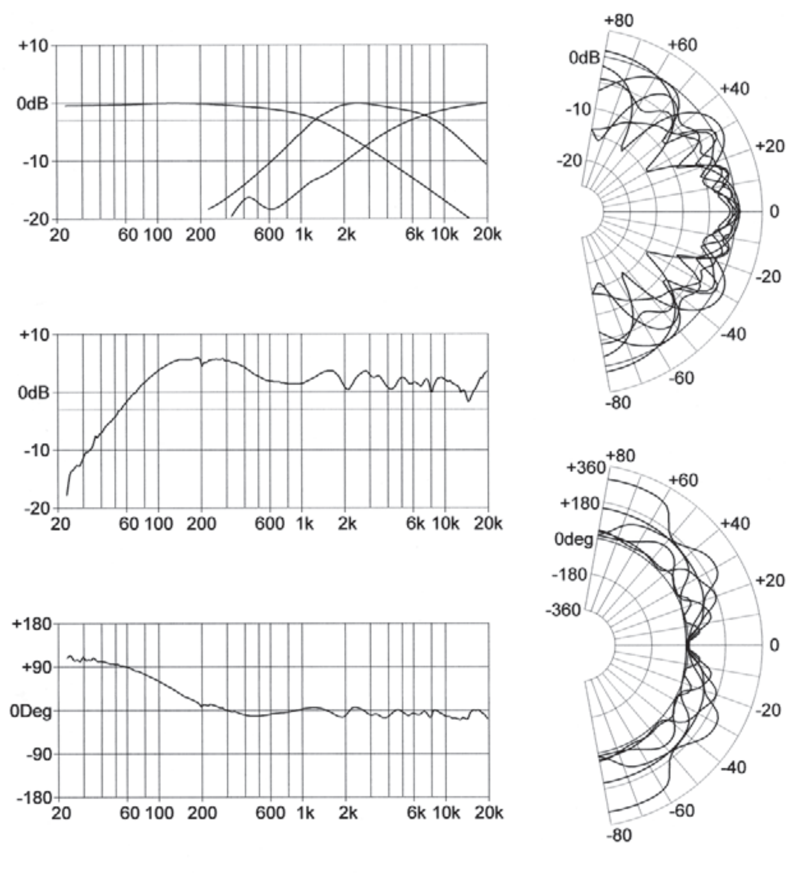


FIGURE 24: Modeled performance of three-way WMTMW system.

The system was designed to focus at a distance of 114″, which matches my listening position in the garage. This focus, along with the driver acoustic origin differences, determined the steps needed in the front panel. The driver CTC spacing was set at what I thought I could build keeping the angled steps reasonable.

Photo 1 is a front and back view of the basic enclosure showing the front panel steps. The angles between flush-mounted drivers are in the range of 20° to 24° to minimize any effects on the driver responses. The back of the chamber holding the tweeter is open and all driver wiring exits through this area.

The woofers and midranges are each in their own closed box (CB). The midrange CB is inside the woofer CB with its own removable back. The small midrange CBs have high-density fiberglass on the sides and back and are fully stuffed with polyester fiber. The woofer CBs have thick normal fiberglass on the sides, and I stuffed the volume behind the midrange CBs with slightly compressed fiberglass insulation. All chambers are stiffened via hardwood dowels that are damped with duct seal (dumdum).

Photo 2 shows the finished right-hand system for listening. The 8″ subwoofer covers the range 25 to 90Hz. Construction of this subwoofer using a TBSpeakers (Tang-Band) driver and flared port is covered in reference #11. The breadboard first-order CO (**Fig. 25**) sits on top of the subwoofer. Note that a zobel is used on all drivers and the series R-L-C network is included to limit the tweeter impedance rise at its resonance.

You can see in **Photo 2** that the front panel is covered with ½″ thick normal fiberglass insulation. The sidewalls in the midrange/tweeter region also have high-density fiberglass layers cut at about a 45° angle below the ½″ fiberglass layer. I had hoped this would be sufficient to limit enclosure edge diffraction problems with the stepped front panel. (Testing later indicated it may not be sufficient.)

I mounted the WMTMW columns on a sloped pedestal to keep them upright and to put a seated listener on the tweeter vertical axis. The two speakers face straightforward so you listen at an angle off the horizontal axis that I included in the design.

As discussed, both the woofers and midranges are in highly damped CB enclosures. The woofer Qtc measures just above 0.5 and the midrange Qtc in the 0.78 range. The midranges are thus flat down to about 100Hz, but would not be used to the lower limit of their range. My ability to test large enclosures is rather limited because of the size and congestion of my back room.

Avoiding echoes by quasi-anechoic testing, **Fig. 26** depicts the measured system response from 1kHz up. The systems are clearly as close to linear phase as the driver responses will allow. The wiggles in the tweeter response show I have more work to do in damping diffraction/echoes around the tweeter. This is the condition of the systems at the time of the listening sessions because I tested after the weather drove me out of the garage. I thought the strings were slightly strident, but I was using SACDs cut from old analog master tapes, so you never know whether the cause is your speakers or the recording. After testing, I suspect the speakers.

LISTENING TO A LINEAR-PHASE SYSTEM

Listening to the systems was a good experience. I liked them very much (naturally!). They formed a specific image that stayed behind the enclosures. The major attribute was the clarity and speed of transients, whether with a single instrument or an orchestra. I originally attributed this to the linear-phase characteristic.

George Augspurger told me that the linear-phase studio monitors sounded good, but he believed the reason was the careful component matching by the manufacturer rather than the linear-phase characteristic. To test this conjecture, I decided to listen to the systems with the linear-phase property destroyed. The first-order CO sums to nearly the same response with the midrange drivers normal polarity (phase linear) or with the midrange polarity inverted (not phase linear) (**Fig. 27**). To my surprise, and to that of other listeners who participated, the sound hardly changed. The systems were no longer phase linear, but were still fast and clean.

I then modeled and tried a second-order CO that matched the magnitude response of the original first-order CO as closely as I could (**Fig. 28**). This CO configuration requires the midranges to be wired with inverted polarity. The polar

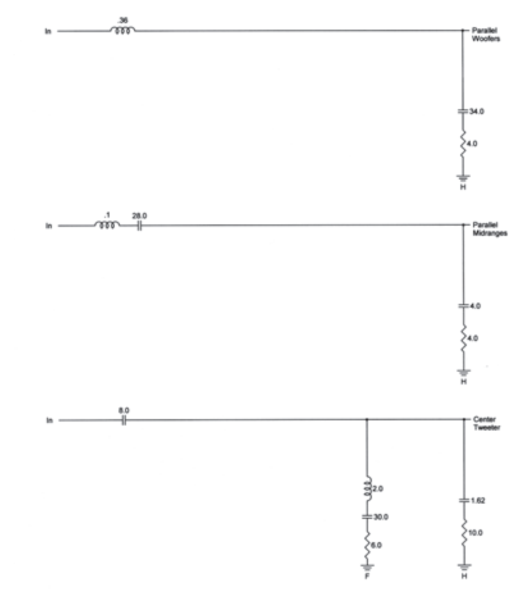


FIGURE 25: Final schematic for WMTMW system.

plots display a lot more nulls and phase reversals than with the first-order CO. Whether it is the strange polar plots or the components I used to implement the CO, the second-order CO did not sound quite as good as either version of the first-order CO. The sound was a little muted or congested mostly in the bass region.

Finally, I modeled and tried a third-order CO with all drivers positive polarity (**Fig. 29**). The sound of this CO was very good, and on some selections I thought it better than the first-order CO. Again, the systems were fast and clean without the phase linearity.

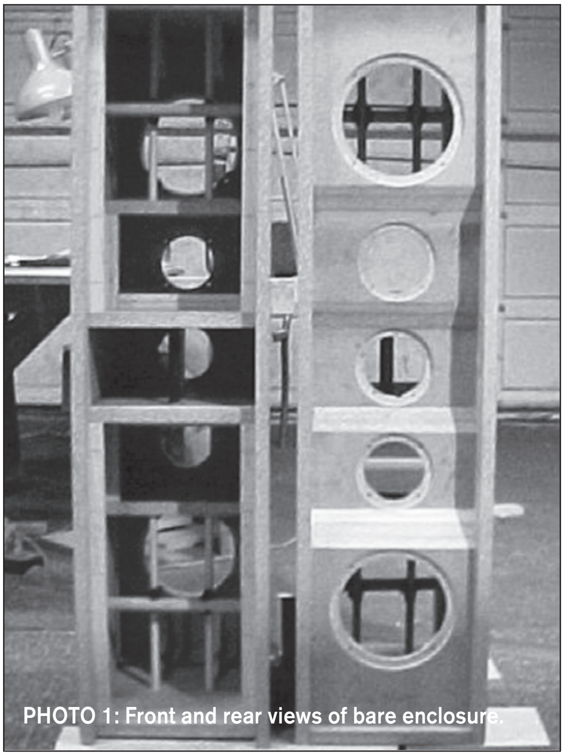


PHOTO 1: Front and rear views of bare enclosure.


PHOTO 2: Finished speaker with subwoofer and breadboard crossover.

DOES LINEAR PHASE MATTER?

My short answer is no. As long as you keep the phase change rate within reason, limiting the delay dispersion, the sound should be fine. However, the techniques you need to build a linear-phase system are worthwhile and should be used. I review these techniques later.

Colin[12, 13] and others have investigated the audibility of "delay dispersion," i.e., the fact that different frequencies are delayed different amounts of time. For the midrange/tweeter region, he believes up to about 1ms is tolerable. It is a complex topic, about which you should read in the original articles.

The point is that for a good-sounding system with quick, clear transients, you need to keep the phase shift at a minimum, not necessarily going for linear phase. Generally, a first-order CO with all drivers the same polarity is fine in terms of delay dispersion. Once you go to higher-order COs, the acceptability of the dispersion depends on the CO frequency and other variables. Crawford[14] investigates correction networks for CO delay dispersion.

While I found the effect of the various COs was subtle on listening to music, the same was not true when listening to a square wave. The square wave has a very "edgy" sound, and the first-order CO with all drivers the same polarity retained this sound better than the other COs. As Colin discusses in his articles, there may be certain musical instruments that display sonic change at lower values of delay dispersion.

The following factors are pretty much forced on you in building a linear-phase passive-crossover system and are generally good for all systems.

1. The symmetrical driver array is an important tool for any system. Colin[13] points out that the symmetry keeps the acoustic origin position of the system stable over frequency, removing an effect he believes is audible. To me, the main advantage is that it minimizes any lobing problem with a multi-driver system.

   To see just how powerful it is, I modeled the Yamanaka second-order fill driver CO used with a WMTMW array of ideal drivers (**Fig. 30**). The lobing problem so prominent with this three-driver system (Fig. 11) is greatly reduced. I then tried the Yamanaka

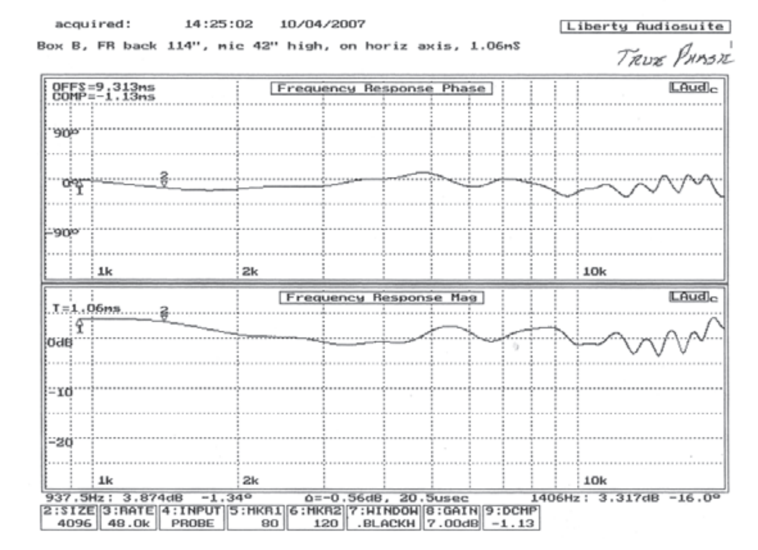


FIGURE 26: Measured performance of three-way WMTMW system.

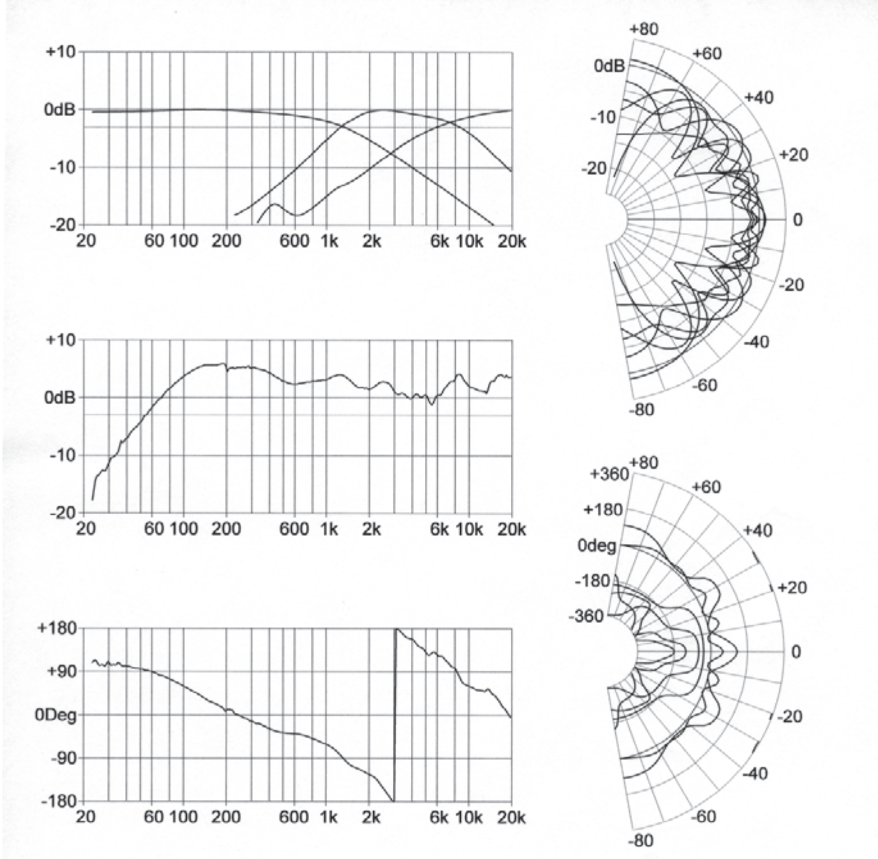


FIGURE 27: Modeled performance of WMTMW system with midrange polarity inverted.

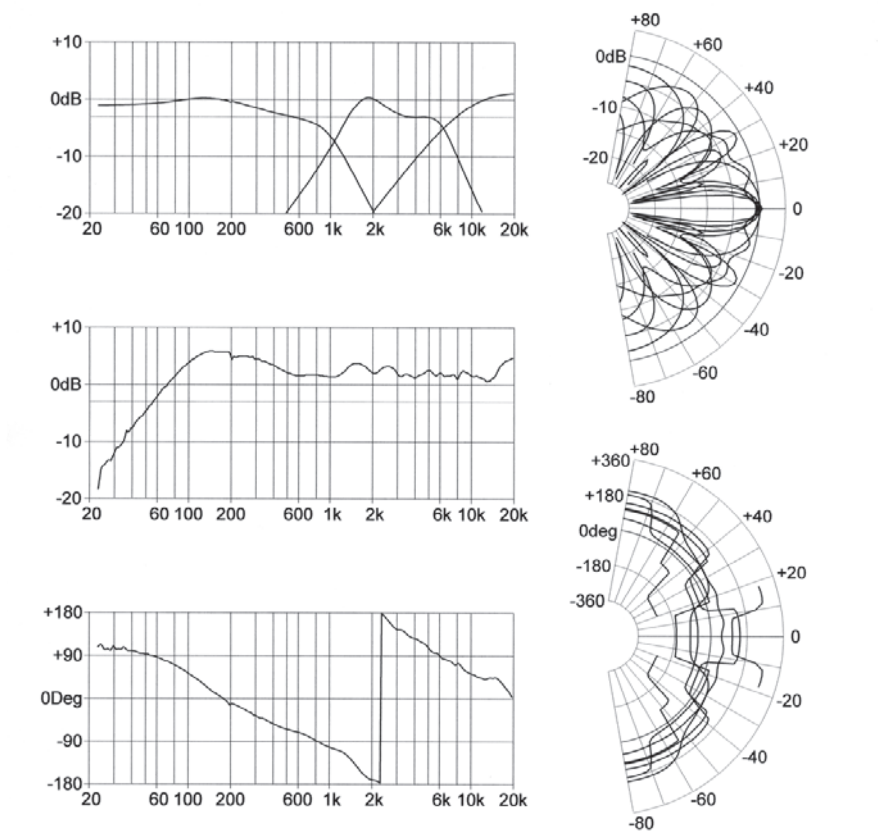


FIGURE 28: Modeled performance of WMTMW system with second-order crossover.

second-order CO on my five-driver columns (**Fig. 31**). I needed to change the front panel steps but I did not try to optimize the magnitude response, and still it looks like an acceptable system. The symmetrical array approach is very powerful, and you should certainly consider it if you plan to try any of the fill driver approaches. I'm of the mind that if you have a five-driver array, then build a three-way system!

One item to keep in mind is that a symmetrical array system will generally display low input impedance. I built my five-driver system with nominal 8Ω woofers and midranges and a 4Ω tweeter. With all the zobels and crossover, the input impedance was 3Ω from about 200Hz on up. I have never had much luck hooking up drivers in different enclosures in series, but if you can make that work, it would remove the requirement for the tweeter to have a 6dB higher sensitivity and yield a higher system input impedance, but require much larger inductors.

2. Aligning the acoustic origins of the

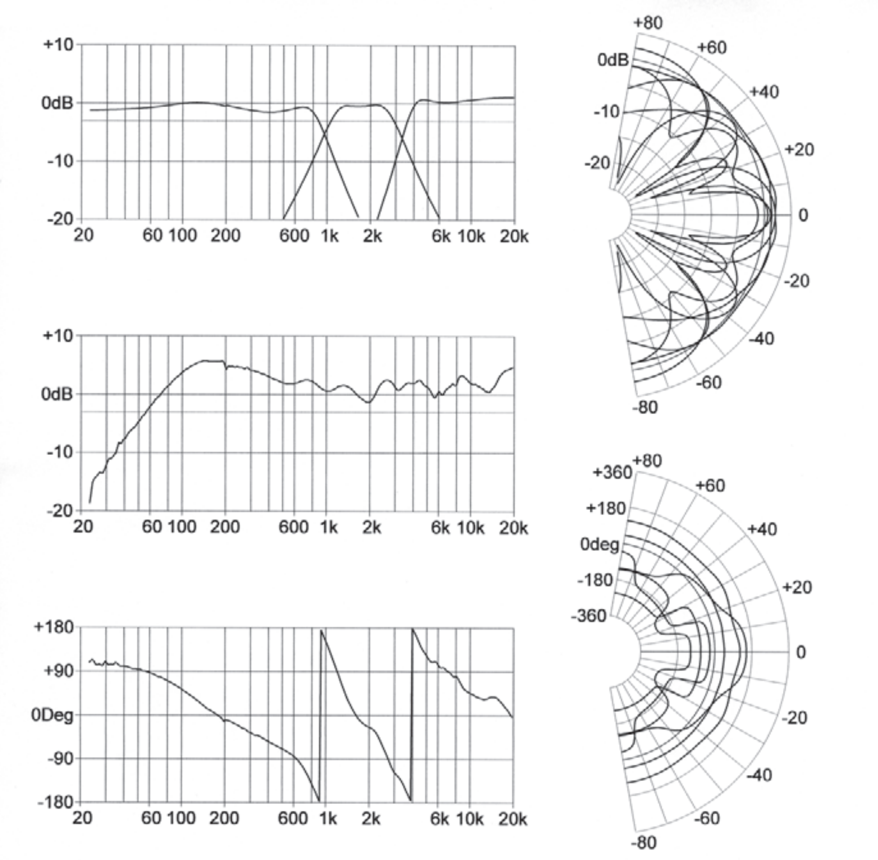


FIGURE 29: Modeled performance of WMTMW system with third-order crossover.

drivers is very important in obtaining linear phase, and I believe it may be useful for all systems. I hate that thought because such systems are so difficult to build. Also, any steps in the front panel cause diffraction/echo problems such as I am still fighting. If you are using only a driver pair, then acoustic origin alignment may be possible by simply tipping the front panel. This means you are listening to the drivers on an angle that will reduce the upper frequency limit on the woofer and show some rolloff in the tweeter response.

When you go over two drivers, you need to consider a curved or stepped front panel or mounting some of the larger drivers at an angle to move their acoustic origin forward. This angle mounting would, again, place a lower frequency limit on a driver and make system modeling more difficult. If it were not for the edge diffraction problem, I would simply build each driver in its own small enclosure so I could simply stack them up and slide them back and forth for acoustic origin alignment.

3. To build a linear-phase system dictated wide overlapping response ranges for the drivers. This, in turn, required physically small woofers. Such small drivers are inherently quicker than big drivers and are thus beneficial to transient response. In this day of subwoofers it probably makes sense to use smaller woofers and cross over to a subwoofer at some frequency below 100Hz.

SUMMARY

Passive CO approaches for building linear-phase systems are rather limited in number. This article examines the approaches I could locate. Most of them do not yield practical systems unless they are used with coaxial drivers or a symmetrical array of drivers. Building such a system also requires positioning the acoustic origins of the drivers, which makes construction rather difficult.

This work has supported those who believe that linear phase is not a requirement for a good-sounding system. It would appear that what is required is minimizing the system phase shift variation, i.e., the time dispersion.

The techniques needed to build a good linear-phase system, however, appear to hold merit for any system. The symmetrical array, acoustic origin positioning, and use of smaller, fast drivers can aid any system.

If you really wish to build a linear-phase system, I recommend an active CO and digital technology in which time delay can be implemented to make construction easier. Even with an active CO, you should not ignore any potential lobing problems.

ACKNOWLEDGMENT

I would like to thank George Augspurger for his numerous and useful comments throughout this odyssey into linear-phase systems. *aX*

REFERENCES

11. G.R. Koonce and R. O. Wright, "Vented 8″ Driver Subwoofer," *aX* 5/04, p. 18 and 6/04, p. 44.

12. Dennis Colin, "Waveform Phase Distortion," *SB* 1/97, p. 18.

13. Dennis Colin, "Frequency Delay Dispersion," *aX* 11/07, p. 16.

14. Dick Crawford, "The Phase Redeemer," *SB* 5/97, p. 34.

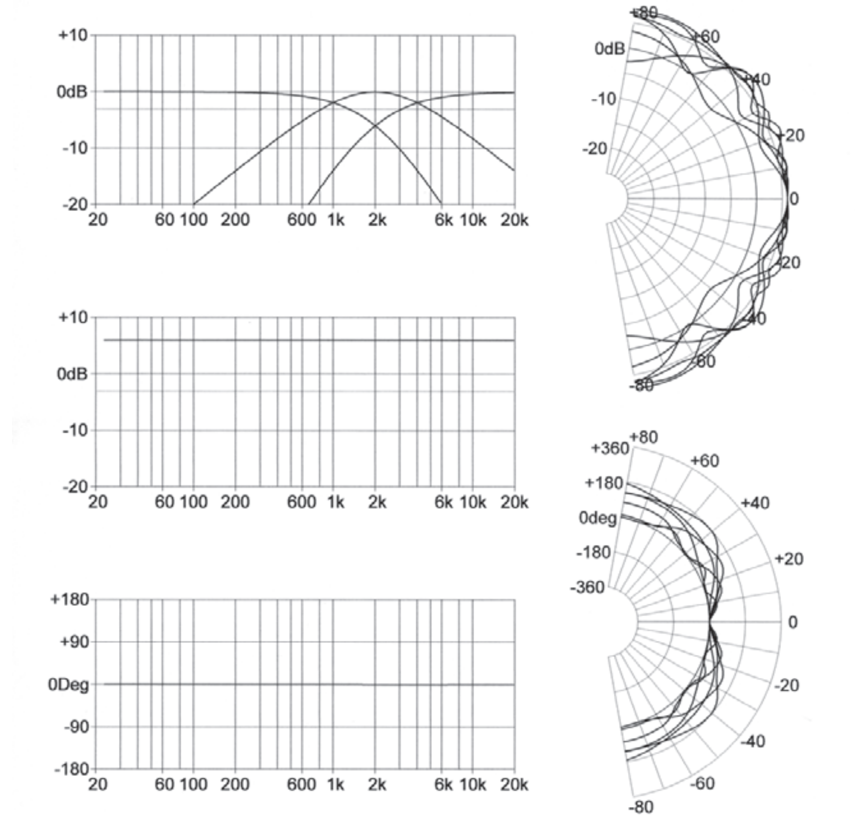


FIGURE 30: Ideal-driver WMTMW system with Yamanaka second-order fill-speaker crossover.

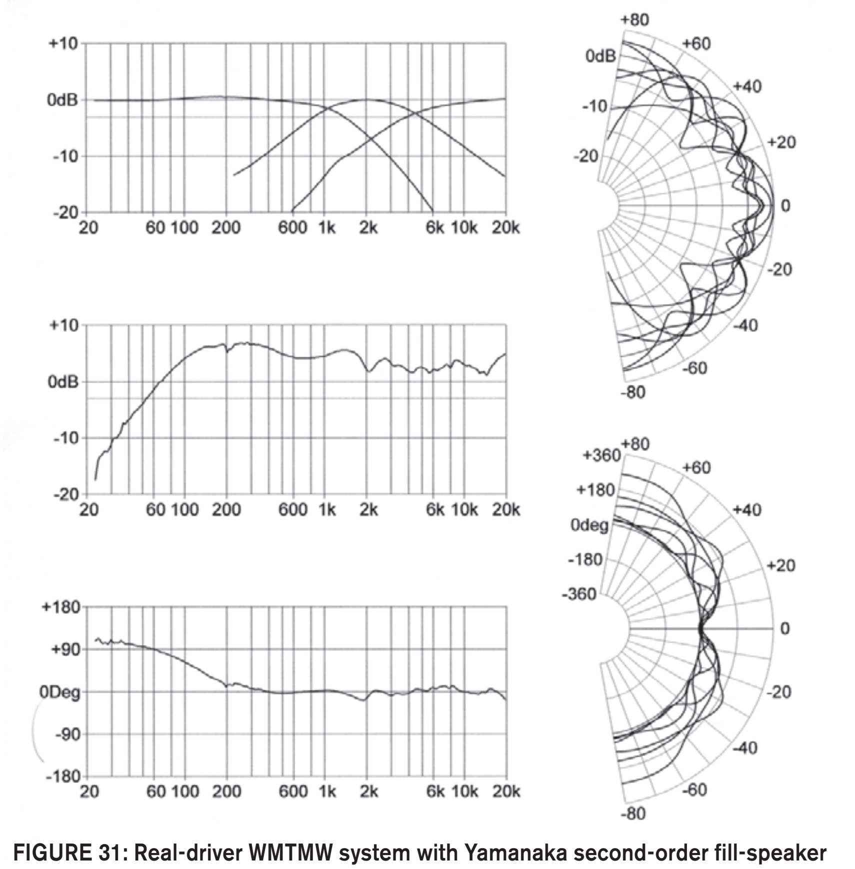


FIGURE 31: Real-driver WMTMW system with Yamanaka second-order fill-speaker crossover.

aX **Interviews** Eduardo de Lima

Eduardo de Lima, President/Owner of Audiopax, Brazil, was the keynote speaker at the European Triode Festival 2007, held at the end of November in The Netherlands (**Photo 1**). His designs are unique in the way he strives to produce synergy between the components of a stereo system. And his unusual creations have won many awards. Regular Contributor Jan Didden tracked him down to find out more about him and his design philosophy.

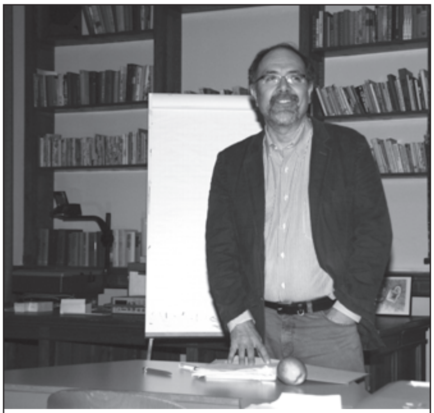


PHOTO 1: Eduardo de Lima explaining his concepts.

Jan Didden (JD): Mr. de Lima, *aX* readers know a bit about you from your article in *Glass Audio* in 1996. But before we go into your design philosophy, can you give us some background on you and your company?

Eduardo de Lima (EdL): Please call me Eduardo. Yes, of course. I think I built my first SE amp, a 300B design, more than 15 years ago. Just out of curiosity, really. I didn't think it would sound any good. It sounded much, much better than expected, but, to my consternation, it didn't make sense in view of the measurements. The THD was relatively large, and the output impedance of more than 3Ω, of course, ruined any chance of speaker damping. This very much intrigued me and set me on the path that I still travel today.

JD: So, why did it sound so good?

EdL: Well, one of the mistakes we often make is to look at the amp in isolation. But, of course, we don't listen to amps; we listen to speakers driven by that amp. The speaker has its own way to "voice" the musical signal. Actually, simple speakers and simple (SE) tube amps have many similar attributes. Both are simple in the sense that both produce simple harmonic distortion spectra. It's not often that speaker distortion and spectrum are quoted on a spec sheet, so I started to measure speakers in this respect. If you then compare that to a SE amp, you see in both cases a similar spread in predominantly low-order harmonics, with similar envelope.

JD: Are you saying that in the case of your amp, some of that amplifier distortion was cancelled by the speaker distortion?

EdL: Cancellation is too strong a word; I'd rather call it "compensation." You can imagine that the harmonic structure of both the amplifier and the speaker vary depending on program material, especially the frequency content and the level. Ideally, if the 2nd harmonics of the amp are in opposite phase of those of the speakers, they would cancel. If they are the same phase, they would reinforce each other. It is also a matter of statistics.

I have done much analysis on this, and as long as the phase difference is between 120 and 180°, the combined distortion is less than 1% of each of the two. So, statistically, you can expect a 33% chance that significant compensation occurs. Now, if you reverse the speaker connections, you even end up with a 67% chance of significant compensation! To look at it another way: If your amplifier THD is half of your speaker THD, you have almost a 90% chance that the resulting system THD is less than the original speaker distortion.

Engineers tend to discard effects that are not totally predictable, but these numbers are worthwhile. There is quite an audible difference between partial compensation and reinforcement!

JD: So that caused the surprisingly good sound of your first 300B. I guess you just got lucky with your speakers!

EdL: Absolutely. And I just mentioned harmonic distortion, but there is a similar effect from the amplifier output impedance. That has quite an influence on the speaker characteristics as well.

Once I found this out, I started to look more closely into that speaker, a home-made speaker with an unusual 24dB crossover between the woofer and the midrange that just by chance exhibited an almost totally resistive behavior. In other words, it looked largely like a resistive load; I mean much more so than normal complex speakers with a complex crossover.

I started to develop my own speakers with this in mind and I found that the effect can be predicted to some extent. It is not totally predictable, of course. Both amplifier and speaker behaviors under signal conditions are dynamic, and harmonic structures and phases vary constantly. But there is a lot of overlap, and by listening to the music, it is possible to find a "best match" that sounds best. And the differences are not subtle but clearly identifiable.

JD: Can you measure the effect? And how would you characterize the sound difference?

EdL: Yes, you can. The speaker measurement part is, of course, the most difficult. Repeated, consistent speaker measurements are very difficult to do.

But by using an anechoic chamber, you can make a measurement series within the same setup as comparative measurements.

The series of graphs in **Fig. 1** show the results of one such session. The differences between the speaker response with a "perfect" amplifier (**Fig. 1a**) and with a "complementary" amplifier (**Fig. 1c**) may not seem huge, but they are clearly audible. The audible difference mostly affects the stage width, instrument placement, and focus.

JD: There are very many different types of speakers out there; how can you design an amplifier that shows a worthwhile improvement in those different circumstances?

EdL: We design our amplifiers with adjustment controls that let the listener tune in to the best match between his particular speakers and our amp. I don't want to go too much into detail, because part of that is our intellectual property, of course. But one way to change the "character" of your SE amp is to use partial cathode feedback from a tap on the output transformer, the way Quad used to in the Quad II (although that was push-pull), as well as using the screen tap usually called ultralinear (**Fig. 2**).

Stockman has shown that a triode is just a pentode with feedback. By varying that feedback from both the cathode connection and the screen connection, you go from triode to pentode and anything in between. This way, you can have almost complete control over the amp's THD level and structure as well as phase.

In our ASTAT (Asymmetric Series Twin Amplifier Topology) amplifiers, we went so far as to construct it out of two amplifiers whose outputs are combined at the secondary of the output transformers, which themselves are slightly different. Each amplifier also has slightly different parameters, and you can adjust many of those while listening. The possibilities may seem overwhelming at first, but it is my experience that most people can zoom in on the best settings quite easily. The interesting thing is that most listeners come to almost the same settings for a particular installation. But the speed with which they zoom in varies significantly, with some listeners able to find the best setting very quickly.

JD: If I don't have speakers yet, what type should I buy to benefit from your amplifiers?

EdL: Well, of course, you can buy our speakers (smiling), but in general look for simple, not complex units. High-efficiency ones, with light cones, the Lowter types, are the best suited. They are simple mechanically and so have a simple nonlinearity character. Simple construction will give simple behavior. And I should also mention that you should do your adjustments at the level and with the type of music that you use most. The improvement will be very rewarding, as it has been for me for many years now.

JD: Eduardo, thank you very much for a most interesting discussion.

Further reading: www.audiopax.com/whitepapers *aX*

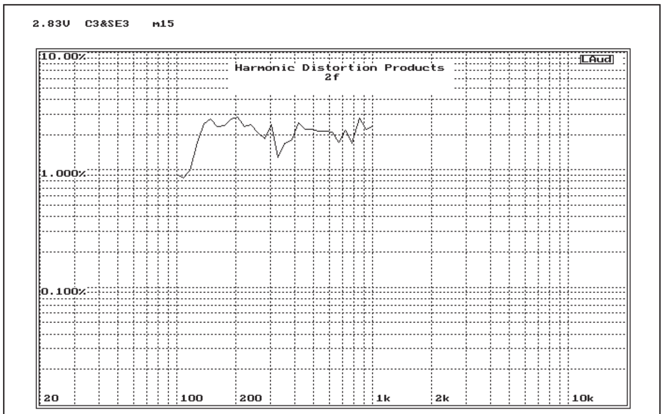


FIGURE 1a: Speaker output distortion when driven by a 0.05% THD amp.

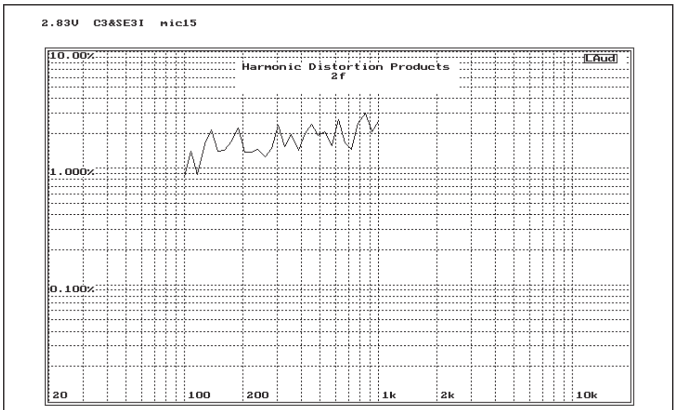


FIGURE 1c: Speaker output when reverse-connected to the SE amp.

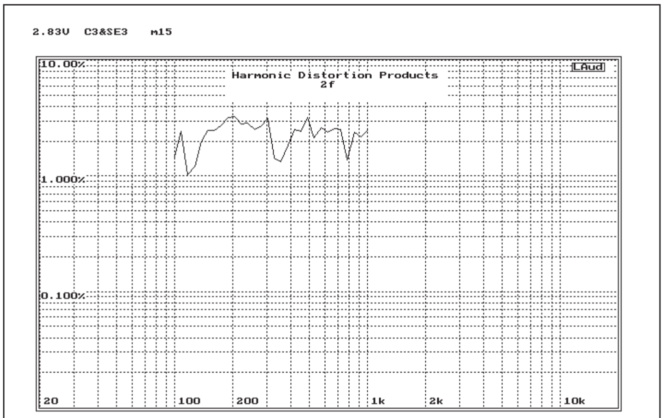


FIGURE 1b: Speaker output when driven from the SE amp.

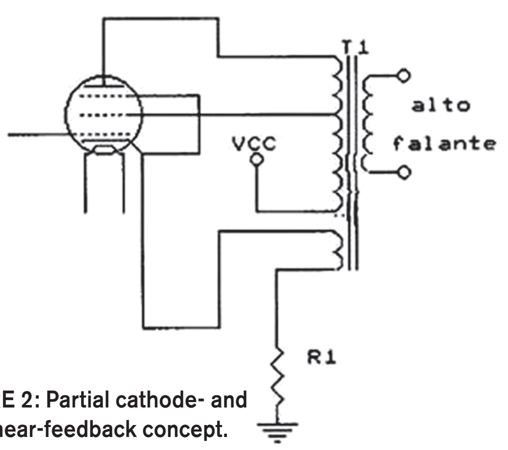


FIGURE 2: Partial cathode- and ultralinear-feedback concept.

A Graphical Evaluation of Harmonic Distortion in Valves

The author describes a graphical method to estimate the harmonic distortion produced by valves under predefined operating conditions.

The elegant method I describe here was often used in the past and was very popular because it required no sophisticated calculations. I also provide its mathematical justification and give an example to illustrate its use.

DESCRIPTION

Consider the anode characteristics of a given valve, as shown in **Fig. 1**, on which are added:

The bias point $P(Ip_0;Vak_0)$,

The load line R_{Load} passing through the bias point,

The points A and B on the load line, defined at the maximum sine input signal amplitude: $\pm e_0$,

The points C and D on the load line, defined at the half sine input signal amplitude: $\pm \frac{e_0}{2}$,

The point M on the load line, defined as the middle of AB.

You can then estimate the harmonic distortion using the following formulas:

$$\%H_2 = 100\left|\frac{MP}{AB}\right|$$

$$\%H_3 = 50\left|\frac{AC+BD-CD}{AD+BC}\right|$$

Higher order harmonic distortion can also be estimated using additional points on the load line. But, the accuracy of the graphical method decreases rapidly with the increasing order, making the method questionable.

The graphical method also applies to push-pull configuration of valves. In this case, curves to consider are composite characteristics.

JUSTIFICATION OF THE METHOD

For a given bias point P and a load line R_{load}, the anode current is only a function of the input signal: e.

You have: $(Ip_e - Ip_0) = F(e)$. You can approximate this function by the following Taylor's expansion:

$$F(e) = a_1 e + a_2 e^2 + a_3 e^3 + \cdots$$

which, in the case of a sine input signal, e = e_0 sinωt, can be rewritten as the following Fourier's series: *see below (1)*

To evaluate the second-order harmonic distortion, you limit the Taylor's expansion to the second order:

$$\%H_2 \simeq 100\left|\frac{a_2}{2a_1}e_0\right|$$

Coefficients a_1 and a_2 of the Taylor's expansion are determined using points A and B defined in the description of the method. Some simple calculations give:

$$\%H_2 = 100\left|\frac{\left(\frac{Ip_{e_0}+Ip_{-e_0}}{2}\right)-Ip_0}{\left(Ip_{e_0}-Ip_{-e_0}\right)}\right| = 100\left|\frac{MP}{AB}\right|$$

For the evaluation of the third-order harmonic distortion, you limit the Taylor's expansion to the third order:

$$\%H_3 \simeq 100\left|\frac{\frac{a_3}{4}e_0^2}{a_1+\frac{3}{4}a_3e_0^2}\right|$$

Coefficients a_1 and a_3 of the Taylor's expansion are determined using points A, B, C, and D defined in the description of the method. Here again, some simple calculations give: *see below (2)*

ILLUSTRATIVE EXAMPLE

Consider the output power pentode type 6550 shown in **Fig. 1**.

Operating conditions for this valve are

Operation	Class A
Configuration	single stage
Anode voltage	250V

(1)

$$F(e) = \left[\frac{a_2}{2}e^2+\cdots\right] + \left[a_1e+\frac{3}{4}a_3e^2+\cdots\right]\sin\omega t + \left[-\frac{a_2}{2}e^2+\cdots\right]\cos 2\omega t + \left[-\frac{a_3}{4}e^3+\cdots\right]\sin 3\omega t\cdots$$

(2)

$$\%H_3 = 50\left|\frac{\left(Ip_{e_0}-Ip_{\frac{e_0}{2}}\right)-\left(Ip_{-e_0}-Ip_{-\frac{e_0}{2}}\right)-\left(Ip_{\frac{e_0}{2}}-Ip_{-\frac{e_0}{2}}\right)}{\left(Ip_{e_0}-Ip_{-\frac{e_0}{2}}\right)-\left(Ip_{-e_0}-Ip_{\frac{e_0}{2}}\right)}\right| = 50\left|\frac{AC+BD-CD}{AD+BC}\right|$$

Grid n° 2 voltage	250V
Grid n° 1 voltage	-14V
Load resistance	1.5k

The harmonic distortion estimation for a grid swing voltage of $\pm 14\,V$ peak-to-peak is determined as follows:

Bias point P	
Grid n° 1 voltage	-14V
Anode current	140mA
Anode voltage	250V
Point A	
Grid n° 1 voltage	0V
Anode current	277mA
Anode voltage	47V
Point B	
Grid n° 1 voltage	-28V
Anode current	25mA
Anode voltage	422V
Point C	
Grid n° 1 voltage	-7V
Anode current	215mA
Anode voltage	140V
Point D	
Grid n° 1 voltage	-21V
Anode current	71mA
Anode voltage	355V

$$\%H2 = 100\left|\frac{\frac{(277+25)}{2}-140}{277-25}\right| = 4.4\%$$

see below (3)

CONCLUSION

As you can see, the graphical method introduced here is simple and accurate enough to estimate the harmonic distortion produced by valves under given operating conditions. Although this method was very popular in the past, it is nowadays on the point of being completely forgotten by designers who rely too much on modern software, hiding, in most cases, the beauty and elegance of clever solutions. aX

REFERENCE

TUNGSRAM Memento—Roger Crespin "Guide du Radio Electricien" 1942—Tome 2—pages 28 to 30.

(3)

$$\%H_3 = 50\left|\frac{(277-215)+(71-25)-(215-71)}{(277-71)+(215-25)}\right| = 9.1\%$$

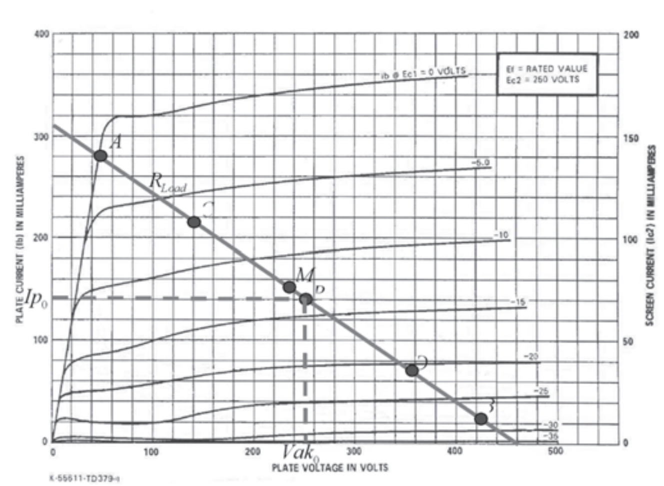


FIGURE 1: Average plate characteristics.

VENDORS

BILLINGTON
EXPORT LIMITED

All types of audio tubes. 300B 6DJ8
ECC81 ECC83 KT88 Mullard GEC
Minimum order £100.00
Discount for large quantity.
Billington UK. Tel (0)1403 784961.
Fax (0)1403 783519.
Email/website www.bel-tubes.co.uk

High Performance kits, Audiophile components
Custom designs, Custom Assembly
www.borbelyaudio.com
In Japan:
http://homepage3.nifty.com/sk-audio/

AudioClassics.com Buys - Sells - Trades - Repairs - Appraises McIntosh & other High End and Vintage Audio Equipment
800-321-2834

Vista-Audio, Radii, Audio Limits, Trafomatic.
Tube amplifiers, kits, custom transformers.

www.engineeringvista.com

HERE NOW. "The Restoration Preamp"TM (model 4010). For cylinders, 78s, 45s and early LPs.
www.tdl-tech.com (575-382-3173)

ALL ELECTRONICS
CORPORATION

Electronic and Electro-mechanical Devices, Parts and Supplies. Many unique items.

www.allelectronics.com

Free 96 page catalog
1-800-826-5432

WANTED

VINTAGE AUDIO & RECORDING EQUIPMENT
(DOMESTICALLY AND INTERNATIONALLY)

ERAS: 1910 - 1990

TYPES OF EQUIPMENT: AMPS, PREAMPS, TUBES, POWER SUPPLIES, SPEAKERS, HORNS, DRIVERS, MICROPHONES, FIELD COIL, CROSS-OVERS AND ANYTHING RELATED.

BRANDS: ALTEC, RCA, WESTERN ELECTRIC, TANNOY, MCINTOSH, MARANTZ, SIMPLEX AND MORE.

HISTORIC CINEMA PRESERVATION GROUP

800.353.8417 EXTENSION 1 - HISTORICCINEMAS@GMAIL.COM

WE WILL PICK-UP AND REMOVE EQUIPMENT, CASH PAID ON PICKUP

Ad Index

Yard Sale

FOR SALE

Speaker systems (pairs):
ADS L620 25″ × 14″ × 12″.
Advent Prodigy Tower II 28″ × 10″ × 9″.
CBS Reference 206L, black metal cabinets, damped, 11″ × 7.5″ × 8″.
Hundreds of classical LP records.
Call Jerry at 718-339-0435 or write P.O. Box 297076, Brooklyn, NY 11229

AMPEX 354 tube reel-to-reel, stereo, transport and 2 record/play amps in cabinet. Not functional but good cosmetic condition. In Minneapolis area. Buyer arranges pickup or shipping. $300/offer. Bill at 763-786-2692 or bwerner@mnnradio.com. Also have Omega D-2 photo enlarger, cheap.

WANTED

Record Cleaning Device at reasonable price. Please contact Angel Rivera at: Speakerman@excite.com

Original or copies of schematics, brochures, and advertisements on stereo equipment from NARDEUX. Made in Loches, France in the 1960s. Stereomatic 302-S vacuum tube amplifiers or solid-state power amps mono or stereo. Nardeux enthusiast or correspondence is welcome always.
John Badalamenti j.badalamenti@att.net.

Denon DP-47F turntable.
Lasonic TRC-911 radio boombox and single door cassette tape deck.
Onkyo Integra P-308 control amplifier.
M-508 100W per channel power amplifier.
Marantz DCC-850 BK single door cassette tape player.
Luther Wright
6404 Church Road
Petersburg, VA 23803

"Yard Sale" is published in each issue of *aX*. For guidelines on how subscribers can publish their free ad, see our website.

Super Fidelity

By John Sunier

Reprinted, with permission, from the November/December 2007 issue of *Australian HiFi* and *Home Theatre Technology*.

Pilgrimage | Michael Brecker | Heads Up SACD HUSA 9095
This historic SACD is the final recording by the pre-eminent jazz saxophonist and composer Michael Brecker, who died early last year after a long battle with MDS and leukemia. His dedication to the music kept him going for this last session of August 2006, and his advanced condition was kept from most of the leading players who joined him in the studio. Pat Metheny called it "...one of the great codas in modern music history. It's one of the most amazing, powerful, unbelievable things that I—and all who were there—have ever experienced or will ever see."

All nine compositions are from Michael Brecker. Most are not strong on diatonic melodies but maintain interest with frequent harmonic and rhythmic changes. *Half Moon Lane* seems to have the most unforgettable melody; Brecker's tone in the upper register is very smooth. *Tumbleweed* is a wildly tumbling extravaganza followed by the quiet introspective ballad *When Can I Kiss You Again*?

Tomorrow's Jazz Classics 2005 | Nagel Heyer Records | SACD NH 1022, [Distr. by Qualiton]
Sorry it took us so long to learn about this SACD, but now here it is and if you're a jazz fan with an SACD capability it shouldn't be missed. All 14 tracks are all first-rate, and all in 5.0 surround. Saxophonist Greg Osby does a fantastic solo on *Night Call*, Donald Harrison's quartet turns in a version of Miles' *So What* that's just as great as the original except in hi-res surround, a two-guitar/violin/doublebass quartet known as String Zone offers an original gypsy jazz number, one of today's best clarinettists—Ken Peplowski—shepherds his quintet through *With Every Breath I Take*, Trumpeter Brian Lynch dips into trad jazz for a swingin' *West End Blues*, and top bassist Frode Berg beautifully bows his way through his attractive original *Victor*. Most of these tracks made me want to have the entire CD they were taken from.

Sibelius: Symphony No. 2 | Pohjola's Daughter | LSO/Sir Colin Davis | MCH SACD LSO-0605
Sibelius' most popular symphony has received herein a performance and recording that will be difficult to surpass, in spite of the many competing versions. The first movement is the most Nordically cool of them all, it still has the catchy main theme which has become one of Sibelius' "greatest hits"—so much so that it showed up on a CD of "Hooked On the Classics" themes. Opening the disc with the quarter-hour-length Pohjola's Daughter is an appropriate move in that the work can be regarded as a very compact one-movement symphony—something like the composer's Seventh. Sibelius viewed it as his own Hero's Life, but the wizard-hero of the tale loses in the end his effort to seduce the daughter of the moon-god. The extremely subtle and quiet chamber-music writing in much of the score is more clearly heard and appreciated than on most standard-res recordings of the work.

Jacintha Goes to Hollywood | Groove Note | Stereo-only SACD GRV1040-3
Jacintha Abisheganaden delivers her English lyrics with such precision and lack of accent that it's easy to forget she's a well-known actress in Singapore. The tunes are all from hugely popular movies. No effort was made to include an obscure discovery or two. Jacintha's backing band is always top-flight and this one is no exception. Guitarist Anthony Wilson, who has a couple of killer albums on Groove Note, is a standout. Jacintha works very close to the mike for a sexy and intimate sound something like the jazz thrushes of the 1950s. She is the No. 1 jazz vocalist with many audiophiles due to her and Groove Notes' high technical standards—whether on SACD, standard CD or vinyl. Tracks include: *On Days Like These, Raindrops Keep Falling On My Head, Alfie, Windmills of Your Mind, California Dreaming, A Man And a Woman, Easy Living, Que Sera Sera, The Summer Knows*.

Masters and Commanders | Music from Seafaring Film Classics | Telarc Mch SACD-60682
This is the 84th album Kunzel and his Cincinnati forces have recorded for Telarc. Swashbuckling high-seas-type movies are not at the top of my must-see list, but I found the variety of hearty musical fare in this program more appealing than some of Kunzel's past collections. Some of the greatest composers for the silver screen are represented in these stirring themes, including Korngold, Elmer Bernstein, Franz Waxman and Miklos Rozsa. A nice variety is added to the selection by including some themes from films as far back as the 1940s. I'm afraid the scores by Klaus Badelt and Hans Zimmer for two of the current Pirates of the Caribbean series don't measure up, but all 18 tracks are performed with gusto and displayed in the best 5.0 surround you could want.

The Kinks | Low Budget | Konk/Koch Records Stereo-only* Hybrid SACD, VEL-SC-79817
Low Budget was the last really great pre-new wave rock album. Released in 1979, it shot to prominence thanks to the satirical track (W*ish I Could Fly Like) Superman*. The Kinks were one of the truly great live bands, but Ray Davies always felt that the public only remembered their classic songs, so with Low Budget, he abandoned the "concept" approach of the previous recent string of Kinks albums. The resulting disc offered several big hits along with numerous well-crafted rock songs with much of the wit and humor of classic Kinks output, and received critical praise and commercial success. Compared to my MFSL SACD version of the same disc I couldn't discern much difference other than a slight bass boost on the MFSL disc, so at half the price, I'd definitely go with the Koch version. The SACD layer trounces the Red Book layer in every respect. Highly recommended.

AD STANDARDS

The UK's Advertising Standards Authority exists to protect consumers from the lies and deceit that so many companies try to get away with in false advertising, something the audio industry is all too familiar with. The latest company to be put on notice for shady advertising is Russ Andrews Accessories that sells a multitude of audio equipment including speaker wire, power cables, and interconnects in the UK.

According to the ASA, Russ Andrews Accessories was challenged by a consumer on the following claims that showed up in their catalog.

1. "The key to success of our PowerKords is KIMBER's unique cable weave which has proven to dramatically reduce Radio Frequency Interference (RFI) already on the mains supply and to reject further pick up of RFI ...," because he believed the PowerKord cable would have little effect on conducted electromagnetic interference;
2. "... Distortion levels inside equipment is vastly reduced, letting you hear a sound that is vastly clearer and purer, more detailed and far more dynamic ...," because he believed the Signature PowerKord cable would have little effect on measurable distortion in hi-fi equipment, and
3. "... eliminate system sound fluctuation and help to create a super-quiet noise floor, allowing more believable dynamics, deeper bass and lower high frequency distortion ... Listen out for a quieter noise floor (expect more dynamic music and greater detail) and a much more cohesive musical sound ...," because he believed the advertised spike-protecting devices would have little effect on the noise floor in hi-fi equipment.

So what does the ASA do? Submit the claims to an expert for verification, of course. The expert determined that the claims were in fact unsubstantiated and that the research papers provided offered no supporting measurements and were not peer reviewed to give them any credibility.

In the end, the ASA told the advertiser to cut the bullshit and, "not use the claims again unless they could substantiate them with robust scientific evidence." Ouch.

Now how do we get an agency like that over here in the US?

Source: UK Advertising Standards Authority

Richard H. Campbell, PE, FAES, FASA, NCAC
Bang-Campbell Associates
26-G Chilmark Drive
East Falmouth, Mass.

Edward T. Dell, Jr. responds:

My longtime valued friend Dick Campbell is a highly talented engineer and teacher. His dedication to honesty and integrity where audio gear performance is concerned is second to none. Nothing arouses his righteous ire more than fanciful claims, unsupported by objective measurements, about wire especially. However, the brush he chose in his interesting letter about a British wire advertiser seems rather too broad. His instance cited is not a group of advertisers, although the claims are too often typical of what manufacturers' ad agencies include in their ads.

I think it is important to realize that almost all advertising of anything is now based on the consumer's feeling. Very few ads for cars extol measured technical details. They emphasize how owning the expensive vehicle makes you feel when you drive it, are seen in it, break the speed limit in it. I suspect wire buyers who choose gold-plated #10 gauge for connecting car speakers to 1kW amps are not thinking of measurements but what their girlfriend will think.

So, Richard, I think your letter is well taken, but most people know that hyperbole is a constant hazard. And it is fair to remember that not all cable ads are guilty—especially not by association.

AMP6 AMP

J. R. Laughlin's description of his AMP6 amplifier (March '08, p. 18) is solidly detailed as to layout and construction, but somewhat sketchy as to circuit operation.

The driver/phase inverter (6SL7) belongs to the class of so-called "floating" paraphase inverters. In particular, the author uses the "see-saw" configuration of M. G. Scroggie (*Wireless World*, July 1945). This and related circuits are treated in detail, with references, in F. Langford-Smith's *Radiotron Designer's Handbook*, 4th edition.

Mr. Laughlin's comment that section U2b of the 6SL7 is configured in a manner akin to that of a unity-gain inverting amplifier is well-taken and certainly not described in those terms by Langford-Smith. In fact, the term "floating," which refers to the voltage at the junction of R3 and R9 in Figure 7, would today be called a "virtual ground." If R3 = R9, then the junction of these resistors would approach zero AC volts to ground as the gain of U2b increases. (Because there is no DC current in these resistors, the DC voltage to ground is immaterial.) If U2b had the gain of a typical IC amplifier, there would be no need to adjust the ratio of the resistors for equal (but opposite) drives to the 6V6 grids. Using the equations given in the *Handbook* and assuming a gain of 40 for U2b, Mr. Laughlin's selections of 470K for R3 and R14 and 510K for R9 should result in drives balanced to about 1% or better and, of course, he allows for the use of a potentiometer to adjust for even closer balance.

Lawrence Wallcave
lwallc@yahoo.com

MAINS ANALYZER

I would like to compliment Iain McNeill on the excellent construction article ("A Mains Analyzer," Nov. '07, p. 8)—and an interesting choice of what to build. Personally, I never doubted exactly how bad the AC line was, but it is interesting to see that fact quantified.

There are, however, one or two of Iain's conclusions with which I disagree (and one or two I would add).

In a digital switching amplifier, contrary to what Iain suggests, the output tracks the input due to feedback, and the power supply is merely used to supply "raw packets of current" to what is essentially a reservoir—again under the control of feedback. All of the power is used in the form of high-frequency pulses, which are quite easily filtered and not terribly sensitive to individual variations. Most digital amplifier designs are, therefore, quite immune to supply variation or noise (luckily). [This may not be true of some older designs that operate as "linear converters" at relatively low frequencies.]

To me, the entire issue simply stresses that care must be taken when designing a power supply. With a power supply, you are basically starting with a 100% AC noise component, which is converted to (you hope) clean DC by the supply. Iain's findings stress the need for good power supply design, good circuit choice, and good matching between the two.

Solid-state amps should be designed with symmetrical supplies (noise which appears equally and out of phase on the plus and minus supplies is easier to reject). They should use sufficiently large filter capacitors and decent high-frequency bypass capacitors to remove the noise. Many poor designs scrimp on the filtering and rely on the PSRR of the circuitry alone—which should be avoided unless the PSRR is very high (and even then on general principles). Taking care to ensure that the supply bridges are connected symmetrically, or using a 1µF polypropylene bypass cap instead of a 0.01 mica, can make a big difference in supply noise, but often manufacturers or designers don't bother.

Sensitive circuits, particularly those with poor or no PSRR, should be supplied with regulated power—in both tube and solid-state equipment. If you're going to build that SET amplifier, you probably *should* regulate the output stage power supply. Yes, this will make the job harder, or it just might suggest that a single-ended amp isn't such a good idea after all. Your basic push-pull amplifier uses the output transformer to cancel out the noise on the output stage—so is a lot less fussy about supply noise—sounds like a better design to me. The driver stages, not being push-pull, should be given very well-filtered power or, better yet, regulated power. Because they don't draw much current, this isn't a hardship—and even an R-C-R-C-R-C filter will do a lot of noise suppression.

I would especially recommend Iain's article to anyone even considering purchasing one of those expensive line cords to attach his (or her) equipment to the mains. Ditto for fancy wall outlets. The very idea that anything is going to influence the way something sounds by somehow doing a better or worse job of passing on the crud that's coming out of the wall is pretty laughable (beyond a decent, thick, low impedance piece of copper to avoid voltage drop).

[If you're using a good power conditioner, the power going into it is the same crud from the wall, so a fancy cord or outlet won't help; the power it's sending out to your equipment should be pretty good already, so it doesn't *need* any help from a fancy tweak cord either.]

Keith Levkoff
kLevkoff@panix.com

Iain McNeill responds:

Thank you for those kind words and for your interest in my article. I took a look at the current offerings from TI, ADI, and Tripath and found that you are quite correct. The latest consumer-audio targeted products do have feedback loops and therefore can be expected to offer some degree of power supply rejection. My experience with class D was some time ago with custom chipsets for the wireless communications industry and had no such feedback; you could hear exactly what the power rails were doing. As always, buyer beware!

I completely agree with your comments regarding power supply design for audio. The biggest problem with power supply filtering is that the source impedance is so low, making it difficult to get significant attenuation. A typical 40W/channel stereo amplifier will have a transformer secondary resistance in the order of 0.3Ω (for 10% regulation), and so even the best electrolytics, with series resistances in the 25mΩ range, will give only around 20dB attenuation. Paralleling multiple capacitors helps, but is subject to the law of diminishing returns.

As you point out, it is important to use multiple types of caps because large electrolytic capacitors tend to have large parasitic inductance which increases their series impedance at high frequency. Personally, I would use the poly 1µF *and* the 0.01µF mica on top of the electrolytic reservoir caps and probably a 10pF as well for good measure. Using two of each type in both positive and negative rails would not be paranoid in my books! This is in addition to the liberal sprinkling of decoupling caps around the circuit. Obviously, if you can afford a little resistance in the supply lines, then things improve quickly and the R-C technique, as you describe, is very effective at isolating sensitive input stages from noisy supplies as well as each other.

I think the jury is still out on those fancy power cords. In some ways it is easier to filter on the line side of the power supply. On the 120V side you can have 9Ω of source resistance and still get 10% regulation for that 2 × 40W amp; now that capacitor is giving you 50dB of attenuation. Throw in a ferrite sheath (as some vendors claim) and you could potentially have some pretty decent broadband attenuation, albeit at the expense of some extra ground leakage current and some possibly significant heat dissipation. But as you say, this is what the power conditioner is doing anyway. I don't have any data to support any of this, but, theoretically, the approach does seem to have some validity.

OLD-TIME AMPLIFIER

I read the article by Mr. Brown ("Upgrading a Classic Pilot Amp," April '08, p. 8) with interest—very nicely done. The single grounding bus is how it was done in the 1940s and 50s, and how it should be done. One comment would be that lower hum can be achieved by mounting the mains transformer so that the windings are orthogonal to the plane of the chassis. It seems to have been done that way in the unit that Mr. Brown described. In addition to mounting the windings orthogonally to the chassis, I obtained additional hum reduction by connecting the mains transformer to the chassis in a single point.

I have studied the noise and fidelity of old amplifiers and discovered that electrolytic capacitors of old vintage have such a high and noisy leakage current that much rumble and "cooking" sound can result from those capacitors no matter where in the circuit they are used. It is a good high fidelity upgrade to replace those capacitors with a modern construction. Modern electrolytic capacitors are orders of magnitude better than old vintage ones. The manufacturers say that the old capacitors were not made of pure enough materials and will be noisy forever, if they last that long.

Cathode bypass caps should be of the modern variety as well. What I have found to improve the fidelity of an amplifier the most, however, is to use modern film capacitors for interstage coupling. Old paper and wax or paper and oil capacitors have significant leakage currents, which are very good low-frequency noise generators. An amplifier built with those capacitors will sound like someone is cooking water in a big pot when you place your ear to the speaker. You often get that same sound when you "recycle" old tube-sockets, especially the flat bakelite ones from cheap radios.

A trick we used in the 50s was to use carbon comp resistors of about five times the dissipation capability as the power-dissipation of the circuit itself. If, for instance, a resistor dissipated 0.5W, we would use a 2W resistor there.

From my experience, you can replace rectifier tubes with solid-state rectifier diodes as long as you place 2,200pF, 2kV capacitors across the diodes and a 10W 470Ω resistor in series with each diode. This circuit mimics the performance of a rectifier "bottle" so well that I doubt that anyone can tell the difference.

By all means put in a rectifier to make it look like you are doing the "right thing," but do not connect the rectifier heater to B+. Old mains transformers do not like the 300 or more volts on the rectifier heater for very long. So much moisture has become absorbed by the insulation over time that the electrolysis will destroy the transformer in a few months. At least that has been my experience.

There may be some cathode stripping from applying B+ before the cathodes in the rest of the circuit are hot, but that happened with 5U4 directly heated rectifier tubes as well, and I never experienced any problems with cathode stripping in my designs whether I used solid-state rectifiers or not.

Hans J Weedon
Salem, Mass.

SPEAKER PLACEMENT

Like many things in audio, Ambiophonic things are relative. Ambiophonics and RACE crosstalk cancellation—both in theory and in practice—work best with speakers that are relatively close together compared to the stereo triangle. Speaker angles from, say, 15 to 30° are optimum in my experience. There is no one value that is the golden angle because speakers, audiophile preferences or head sizes, and recordings vary.

It is easy, however, to find the optimum angle for your speakers and ears. Simply listen to just a left input signal and adjust RACE and the speaker angle until the signal is as far to the left as you want it to be. I like it at the extreme 90° position, which is where the theory says it should be if a signal is all left. But this is difficult to obtain for this particular test signal that has no interaural time difference, only the level difference. Then listen to an all-right signal and make sure things are the same for this case.

You will find that if the speakers are at the normal stereo angle of 60° apart, then the stage will not be very much wider than that but it still does something and it is better than stereo for most listeners. But the wide angle lessens the pinna advantage of the narrower angle and the sweet area is usually reduced, room response is harder to control, and so on. If you are an audiophile or appreciate high fidelity, I can't imagine you will be happy with compromised Ambiophonics.

In Roger Russell's reply (Jan. '08, p. 49) he writes about speaker spacing in feet, and so on. It is not the lineal distance between the speakers that is important; it is only the angle they subtend to the listener that is significant. I don't know about the SRS system, but in Ambiophonics as you move back along the center line away from the speakers, the width of the stage does not change very much. Normally, you cannot get far enough away from the speakers in the average room to lose the stage. If you get too close to the speakers, you begin to hear ordinary stereo. I have had six to ten people in a sloppy line all hearing a train go across the room.

Most of the adjustments in RACE (both in the TACT box and in the free software version), if not radically altered, are really inaudible to people with normal hearing. They are there to cope with unusual situations, but may be useful to audiophiles who can hear differences in cables and such, and are skilled in tweaking.

Ralph Glasgal
glasgal@ambiophonics.org

Roger Russell responds:

I hope, Mr. Glasgal, you were not offended by my example of listening at a 10′ distance. However, 10′ can be a typical listening distance for many people at home. Of course, speaker placement, listening position, and so on can be adjusted, but that was not the point of my reply. Although your reply has provided another opportunity for a sales presentation for your system, it in no way invalidates my example given in my previous reply.

You may find that crosstalk correction using my column speaker systems described in the November 2005 and July 2006 issues of *audioXpress* can provide better coverage. All of the drivers are in the same vertical line and each covers the entire frequency range. As a result, the excellent coherence of the sound can enhance the effectiveness of crosstalk correction. In addition, the sound will appear the same regardless of your listening height, and the columns project an expanded near-field.

Of course, in the final analysis, many "stereo" recordings are not made with only two microphones and we are limited in what we hear by various recording methods of miking, mixing, panning, and so on. Crosstalk correction is not a method of reconstructing what we might hear by attending a live concert. Instead, the correction adjustments can yield pleasing effects with many alternate realities, none of which may duplicate the original recording.

The best that can be done to eliminate stereo crosstalk is to use headphones, and that eliminates listening position, room effects, speaker directional characteristics, and so on. However, I think that most of us don't want to wear them. Perhaps binaural recordings could at least offer an improvement when played with crosstalk correction through two speakers.

CORRECTION

I have questions for Paul Loewenstein

on his article "A Dynamic Microphone Soundcard Amplifier (Feb. '08, p. 22)."

1. Under "Open Loop Gain": "3.5 – 1.2 = 1.3," not 2.3.

2. Then "gain is about 40 × 1.3 = 52." Where does the 40 come from? Perhaps it should be 40 × 2.3 = 92.

I suggest this calculation:

$r_e = \frac{26\Omega}{I_e}$, where I_e is in mA

$$I_e \cong I_c = \frac{Ring - V_{B2}}{R_2} = \frac{3.5 - 1.2}{5.6k} = 0.41mA$$

$$r_e = \frac{26\Omega}{0.41} = 63\Omega$$

$$A = G_{Q1} \cong \frac{R_2}{r_e} = \frac{Subk}{63\Omega} = 89 \text{ (close to 92)}$$

3. "Z_{in} = 29kΩ." Another way is: $Z_{in} = B_{re}$ = 270.63 = 17kΩ.

4. Closed loop gain equation is wrong. "Gain of 19." Analyzing Fig. 2, I get:

$$\frac{E_o}{E_i} = \frac{1}{\frac{1}{-A} - \frac{x}{-Az} - \frac{x}{-Ay} + \frac{x}{y}} \cong \frac{1}{\frac{1}{-A} + \frac{x}{y}} = \frac{\frac{y}{x}}{1 - \frac{y}{Ax}}$$

= 45, which makes sense. If $A \to \infty$

then gain = $\frac{\text{feedback}}{\text{input z}}$ as with an op amp, which would have a gain of: gain =

$$\frac{10k}{323} = 31.$$

So this circuit has less feedback and more gain.

I like this circuit. Straightforward.

Mick Trego
San Marcos, Calif.

Paul Loewenstein responds:

Wow! The first version of this article was circulated to a speech recognition user group in April 2003, was published before in *audioXpress* in September 2005, and you are the first to notice this elementary arithmetic error. Yes, 3.5 - 1.2 = 2.3. The open-loop gain calculation of 40 times the voltage across R2 corresponds to:

$$r_e = \frac{25\Omega}{I_e}$$

(with I_e in mA), which corresponds to a temperature of 290K rather than 300K. This makes little difference once the arithmetic error is corrected.

It would probably have been clearer had I gone back to first principles and calculated the grounded emitter gain and input impedance using your approach. However, once the arithmetic error is corrected and the temperature adjusted, the two approaches give identical results.

The error in the closed-loop gain equation is an error introduced in typesetting. The submitted equation (and the one published in September 2005) is:

$$\frac{-Ayz}{xy + yz + (1 + A)zx}.$$

The typeset version needs parentheses around the denominator because it is laid out as a single line fraction. I rechecked the submitted equation and I believe it to be correct. Your calculation of closed-loop gain appears to be incorrect; with the limited open-loop gain, the closed-loop gain should be less than the $A \to \infty$ gain of about 30.75.

My arithmetic error propagates to my understating open- and closed-loop gains and overstating harmonic distortion. I provide corrected figures here:

Open-loop gain 92
Grounded-emitter input impedance 16.4k
Closed-loop gain 23.2
Variation of grounded-emitter gain ±1%
Variation of emitter-follower gain ±0.4%
Variation of open-loop gain ±0.6%
Variation of closed-loop gain ±0.15%
Harmonic distortion 0.04%

The variation of gain and harmonic distortion assume ±1mV peak signal from an open-circuit microphone, giving ±23mV peak output. Harmonic distortion is directly proportional to the signal level.

Your correction has reduced the calculated harmonic distortion by a factor of 4, via a combination of reduced open-loop harmonic distortion and higher open-loop gain.

I can also take this opportunity to point out that the version published in September 2005 overstates harmonic distortion by a factor of 2 relative to the already overestimated gain variation. Thank you for your diligence in picking up this error. *aX*