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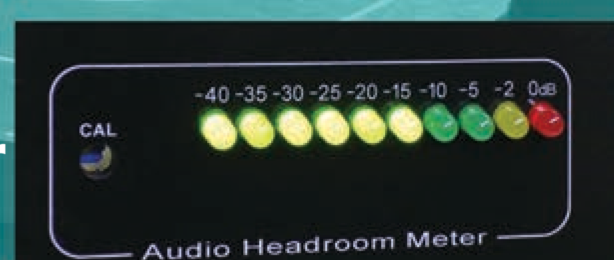

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A Bridge to Better Audio?

This issue of *audioXpress* was in the editing stages when we returned from two intensive days of the ALMA International Symposium & Expo (AISE) followed by CES 2016 in Las Vegas, NV.

AISE 2016 offered an excellent mix of presentations on new generation audio drivers, speaker component testing and quality control, DSP tools, and great product presentations, especially in the test and measurement area, which is the focus of our March issue.

The event was right on target regarding the main trends and topics for anyone working in the audio industry. And that's precisely what motivated me to write about high-resolution audio (again). "Hi Res Audio: A Bridge to Build or a Bridge Too Far?" was also the main theme of AISE 2016 and it was debated from all possible perspectives.

As reflected in some of the articles we selected for this issue—and directly related to some of the AISE 2016 presentations—the main take-away is the understanding that this high-resolution audio debate represents an important opportunity for consumers and the industry in a crucial transitional moment.

Several AISE sessions also served to demonstrate that the audio engineering community acknowledges we need better test and measurement tools and a better understanding of audio reproduction to truly leverage that additional level of information in the digital domain.

Steve Temme (founder and President of Listen, Inc.) provided an overview of the challenges faced by equipment manufacturers in his discussions covering "Correlation of Hearing and Measurement at Very High Frequencies" and "Challenges of High Resolution Measurements." He "defended the measurable facts," such as that "to achieve a flat frequency response (magnitude & phase) out to 20 kHz requires a bandwidth beyond 20 kHz; that loudspeakers and microphones that measure beyond 20 kHz can exhibit trade-offs in performance (e.g., audible intermodulation distortion); and that his own experiments with ABX listening comparisons have made him want to find out more about those things that we are not measuring, such has headphone response at very high frequencies.

Like G.R.A.S and Audio Precision, Listen also detailed how to gather objective response data for several commercially available headphones and the need to "develop a robust process for evaluating and appropriately equalizing headphone responses to a psycho-acoustically valid target and to understand the constraints."

Finally, the last minute addition to the AISE 2016 program of a presentation by Bob Stuart—Meridian's founder and also the creator of Master Quality Authenticated (MQA) technology, now converted into a separate company—delivered a much needed cutting-edge perspective to the topic.

In a two-hour seminar and subsequent contributions to the High-Resolution Audio Forum debate, Stuart explained why we should reevaluate our perspectives by reminding us about important progress achieved in the last 10 to 15 years in the understanding of the hearing mechanism and the neural processing of sound, as well in the fields of psychoacoustics and perception.

In his work on MQA, Stuart proposes that we need to completely rethink digital audio and sampling techniques and the definition of high-resolution audio in itself. In my opinion, the MQA approach introduces some refreshing concepts, including the notion that no vital information to the recorded signal should be removed just because we are not able to measure it or understand it (yet). We highly recommend his published work on the subject, in particular where he argues that sampling to digital should not involve discarding frequencies outside the conventional hearing threshold, which causes loss of critical information in the time domain. Stuart reminds us that all the information contained in the original signal provides vital references on locations (position and distance) and environmental acoustic information— "temporal resolution of microstructure in sound being analogous to spatial resolution in vision"—making a case for preserving "temporal structure at a finer scale for our ears to take advantage of the original sound."

More than simply a theory, this was successfully implemented in its MQA technology and music distribution format with the additional benefits of a clever lossless compression scheme, offering convenience without compromising quality for the first time. Those are enough reasons to declare MQA as "Audio Technology of The Year 2016."

João Martins
Editor-in-Chief

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COLUMNISTS

Vance Dickason has been working as a professional in the loudspeaker industry since 1974. He is the author of *Loudspeaker Design Cookbook*—which is now in its seventh edition and published in English, French, German, Dutch, Italian, Spanish, and Portuguese—and *The Loudspeaker Recipes*. Vance is the editor of *Voice Coil: The Periodical for the Loudspeaker Industry*, a monthly publication. Although he has been involved with publishing throughout his career, he still works as an engineering consultant for a number of loudspeaker manufacturers.

Dr. Richard Honeycutt fell in love with acoustics when his father brought home a copy of Leo Beranek's landmark text on the subject while Richard was in the ninth grade. Richard is a member of the North Carolina chapter of the Acoustical Society of America. Richard has his own business involving musical instruments and sound systems. He has been an active acoustics consultant since he received his PhD in electroacoustics from the Union Institute in 2004. Richard's work includes architectural acoustics, sound system design, and community noise analysis.

Mike Klasco is the president of Menlo Scientific, a consulting firm for the loudspeaker industry, located in Richmond, CA. He is the organizer of the Loudspeaker University seminars for speaker engineers. Mike specializes in materials and fabrication techniques to enhance speaker performance.

Steve Tatarunis has been active in the loudspeaker industry since the late 1970s. His areas of interest include product development and test engineering. He is currently a support engineer at Listen, in Boston, MA, where he provides front-line technical support to the SoundCheck test system's global user base.

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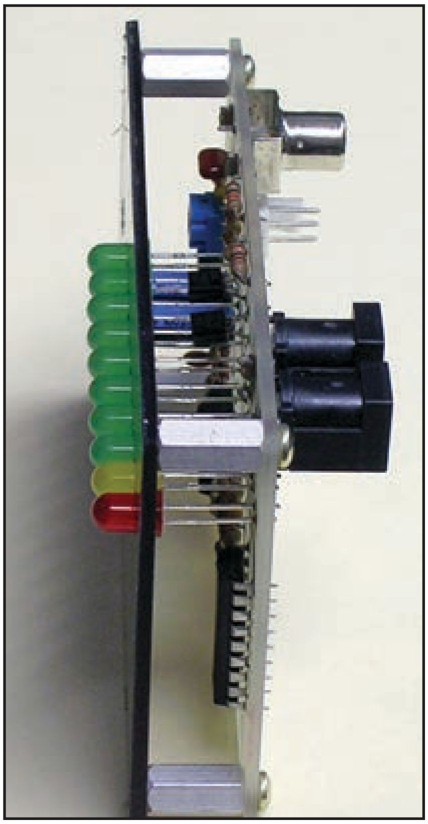
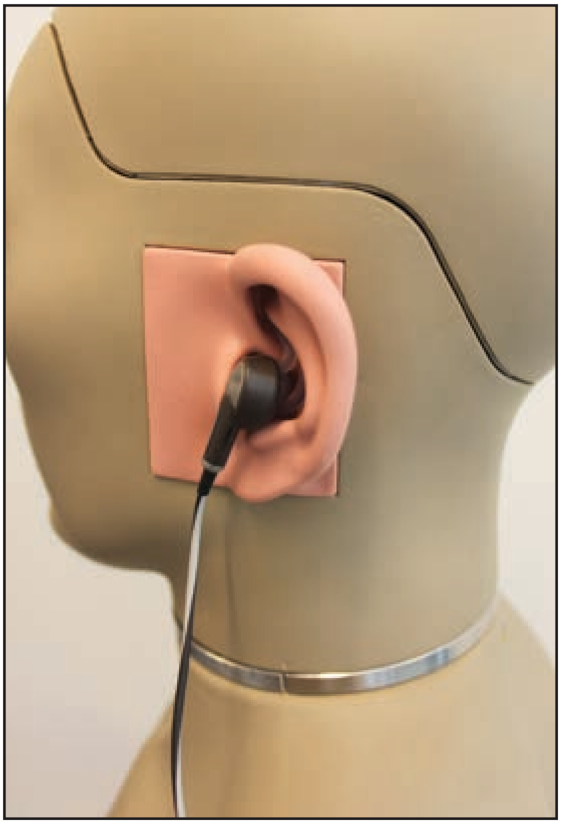


Special Focus

Volume 47 – No. 3 March 2016



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Playing with Room Correction

Because immersive audio formats and sound technology is sufficiently new, my last few articles have not had much DIY information. Experimenting with correcting a room's acoustics with some computer software provides the opportunity for some "hands-on" experience.

Photo 1: The ARC System 2 consists of a calibrated measurement microphone and a software disc with its serial number, which arrives in a single box. IK Multimedia recommends downloading the latest software versions from its website.

By
Ron Tipton
(United States)

Virtually all recent Audio Video Receivers (AVRs) have some form of built-in room correction. There are several different systems, depending on your brand of AVR, but they all have similar setups. A furnished microphone is placed at the primary listening position, or a number of different locations, and test tones are played through each of the loudspeakers. The resulting data is used to arrive at an average equalization to smooth the frequency (and usually the time delay) response of your system. The data is also used to attempt to adjust the relative levels of each of your speakers so all of them have equal loudness. It's this "equal loudness" attempt that can cause problems, but more on this later.

Some audio experts question automatic correction. For example, Dr. Floyd Toole writes in his book, *Sound Reproduction, Loudspeakers and Rooms* (Focal Press, 2008), "…all aspects of room acoustics are not targets for 'treatment.' It would seem to be a case of identifying those aspects that we can, even should, leave alone and focusing our attention on those aspects that most directly interact with important aspects of sound reproduction."

It is also common knowledge that our hearing adapts to our environment as much as is possible. So we might conclude that trying to compensate for all aspects of a room's acoustics may actually make the sound less natural.

A recent poll of AVR auto-correction users (www.avsforum.com) shows mixed results. Some say they would not be without it, while others say they never use it. There are also comments that indicate the program content (e.g., movies and the type of soundtrack or music plays an important part in deciding whether to use auto-correction). With all this said, I will mention my results with the AccuEQ Room Calibration system (Integra, using technology developed by Audyssey) in my Onkyo TX-SR444 AVR.

The microphone must be placed on a tripod or another steady support at the main listening position with its connector plugged into the front-panel jack. Then, select Configuration on the Main Menu. It asks many questions about the speaker system and some are difficult to answer because the available choices don't really fit. For example, there are just two choices for the front speaker size: small (less than 6.5" cone diameter) or large.

My front speakers are 3' tall Woofer-Midrange-Tweeter-Midrange Woofer (WMTMW) three-ways so "large" is not quite correct. Other questions are equally ambiguous and the automatic level calibration does not suffice. The chief difficulty lies in using test tones for this calibration, but what else can you use? Surround soundtracks are variable. What is too quiet from the surround speakers in one movie is too loud in another. Constant playing with the Setup screens is a nuisance so I prefer individual volume controls (knobs) for each speaker, or at least, each speaker pair—front Right (R) and Left (L) surround R & L, height R & L, and so forth. So, have I shot auto-correction full of holes? Maybe not.

Advanced Room Correction System (ARC 2)

The Advanced Room Correction System ARC 2 package (see **Photo 1**) is marketed as a correction system for the monitoring room in a recording studio, but it works just as well in your listening room. The current version is for two-channel stereo only, but I have discovered that correcting just the left and right front loudspeakers makes a noticeable improvement to a 5.1 or higher surround system. ARC 2 uses Audyssey's patented MultEQ XT32 technology, which corrects both frequency and time. (A multichannel version is said to be "coming" but no date is quoted.)

There are two software modules—one for making the room measurements and the other for applying the correction, both use your computer's sound card. The package includes a calibrated microphone for room measurement. It requires a preamplifier, which is not supplied. I used a TDL Technology Model 411 and I needed 40 dB gain to get a useable signal. The measurement software tells you when the microphone level is "OK."

Up to 16 measurements can be made from different room positions and at least seven are required. The first measurement number should be taken at the center of the area to be corrected. The User Guide (included in the Supplementary Material) shows suggested microphone positions for a recording studio's monitoring room, but these can be easily adapted to your room. When the measurement is complete, a correction file is generated for both stereo channels. Typically, studio monitors are near-field so if your listening room is large, you may have poor results. However, the microphone placement diagram on page 33 of the User Guide shows the equalized area extended to the so-called "Client's Couch," which is shown well behind the mixing board. And, the diagram on page 34 shows a 10-chair home theater.

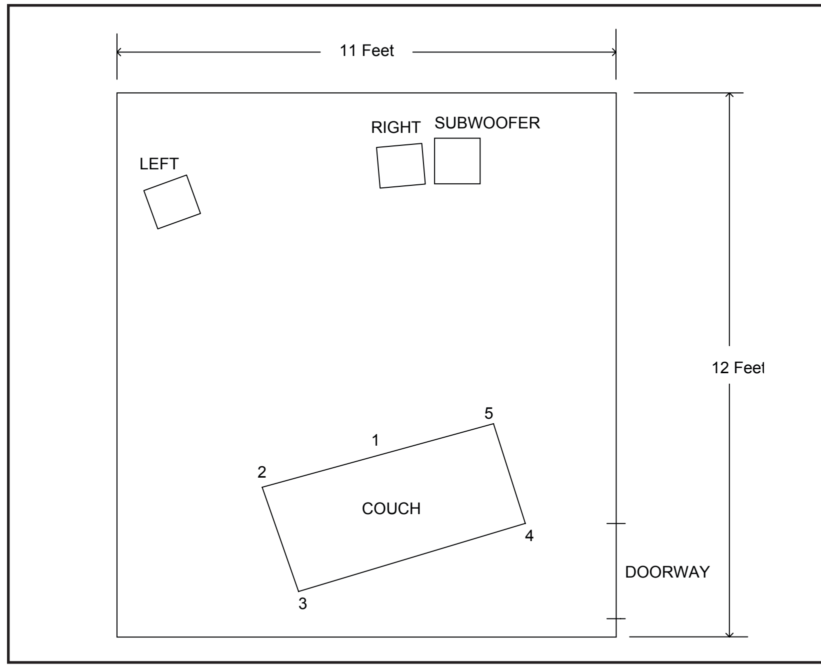

Figure 1: This is a sketch of my listening room with the couch as the area to be equalized. The other equipment and furniture is not shown. The floor is carpeted. The sloped ceiling is moderately non-reflecting, but the walls are acoustically untreated.

I made nine measurements around the 5' couch shown in **Figure 1**. Five of them at the numbered locations shown with four more in between. ARC-2 does not show the measurements, but they are used to generate the correction file with an arc2 extension.

The correction module (see **Photo 2**) is a DAW plug-in—VST, RTAS, or AAX for Windows and Audio Units for Mac. I tried two inexpensive DAWs—Cantabile version 2 and Reaper version 5, and both worked as expected.

Rather than make before-and-after room measurements, I just listened to several selections of stereo music. The Correction On/Off button on the

Photo 2: This screenshot shows the ARC 2 correction plug-in with the uncorrected and corrected frequency response curves for the left and right channels. The name of the measured (uncorrected) data file is listed on the left side under Measurement. Different files can be loaded by clicking the white down arrow to the right of the file name.

Photo 3: The miniDSP 2x4 module measures 3″ × 4″ by about an 1″ thick. It is furnished with a USB cable for programming, which also supplies power. After it is programmed, the computer can be disconnected. However, a 5 to 24 VDC supply with a 2.5 mm female connector, center pin positive, is needed. The power supply is not furnished.

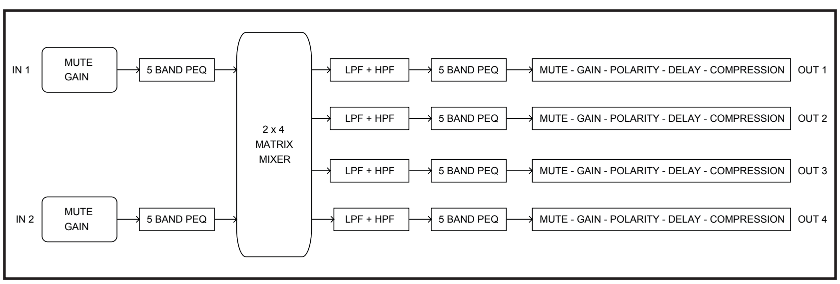

Figure 2: The block diagram of the miniDSP 2x4 Advanced plug-in shows its two inputs and four outputs. Each of the blocks can be programmed by clicking on it.

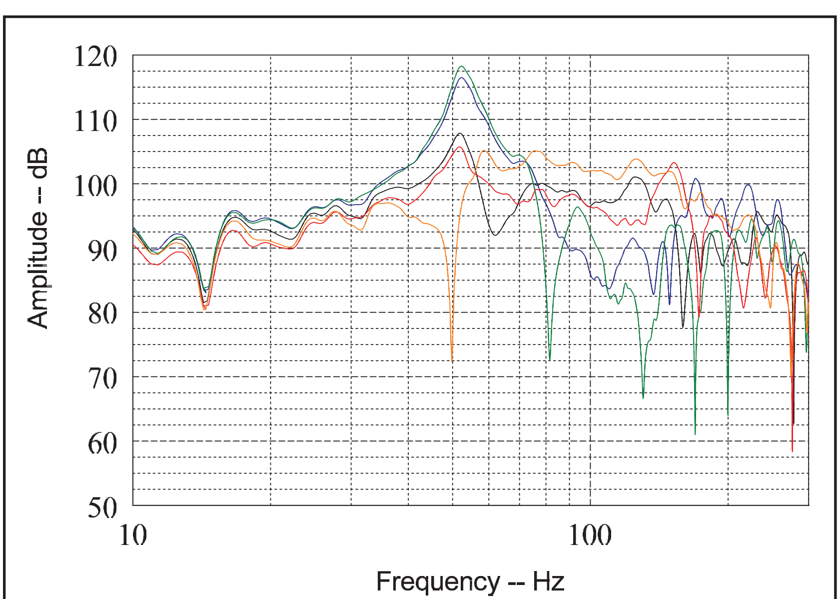

Figure 3: This graph shows the measured frequency response taken at the numbered locations shown in Figure 1. Each position corresponds to a line color: 1 is black, 2 is blue, 3 is green, 4 is red, and 5 is orange.

plug-in makes it easy to evaluate any improvement. The music seemed clearer and more focused with the correction on so I'm keeping it in my sound system.

Using the miniDSP

The miniDSP 2x4 is a small, versatile hardware processor (see **Photo 3**). It costs $105 plus shipping from Hong Kong, which is very fast. One of its applications is a parametric EQ after being programmed with digital filter coefficients. The hardware has been closely teamed with the free Room EQ Wizard (REW) software, which makes the room measurements and computes the filter coefficients. You also must purchase a miniDSP software plug-in ($10), which reads the REW coefficient file and does the programming. Several miniDSP plug-ins are available, but I found the 2x4 Advanced to be the most useful. **Figure 2** shows its block diagram.

This device has two inputs (standard stereo) and four outputs. Using the software, you will notice that it has a programmable five-band input EQ as well as a five-band PEQ for each output. I'll discuss the possible uses of four outputs later.

The REW software can be set up for a full-range frequency scan (10 Hz to 24 kHz) to EQ your stereo loudspeakers (or front left and right surround speakers). Each speaker can be separately measured. The two coefficient files can be programmed into two of the four outputs of the 2x4 Advanced plug-in.

As an exercise, I measured both stereo speakers at the same time at five locations in my listening room (see the numbered locations shown in **Figure 1**). Next, I computed the average response and the predicted frequency response after correction. I've included this data in the Supplementary Material as REW-FullRange.pdf.

The REW software can also be set for a 10-to-300-Hz scan for equalizing a subwoofer, which is the focus I chose for this discussion. The subwoofer response is typically terrible, primarily because the low-frequency wavelengths are very long compared to listening room dimensions, which cause multiple summations and cancellations.

There are several schools of thought as to how to flatten the subwoofer frequency response (e.g., acoustic treatment of the room, multiple subwoofers, and EQ). Several experts favor multiple subwoofers (typically four). Earl Geddes (www.gedlee.com) makes a good argument for this so I've included two of his articles in the Supplementary Material.

I don't have a listening room large enough for four subwoofers so my choice was limited to equalization. **Figure 3** shows the 10-to-300-Hz scans

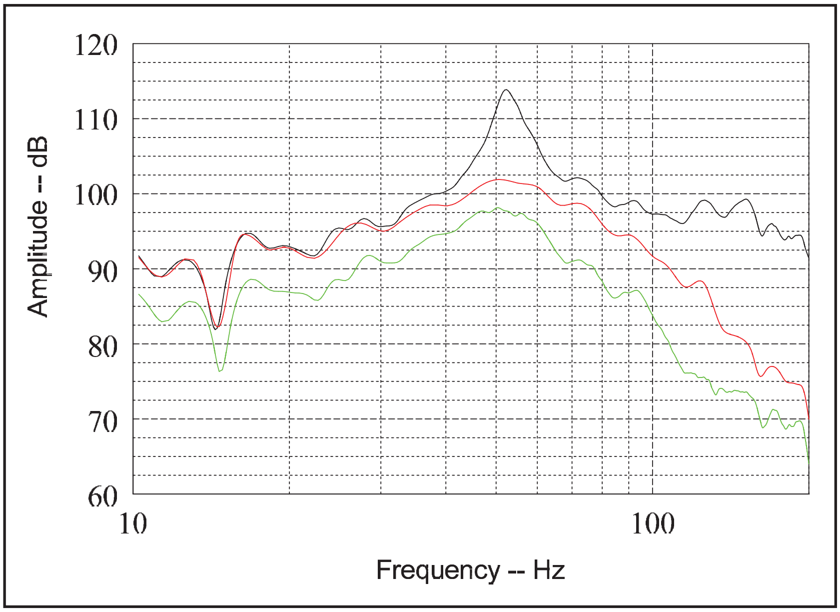

Figure 4: This subwoofer comparison graph shows the unequalized average response in black, the equalization target in red, and the measured equalized average response in green. Without regard to amplitude level, the equalized response closely matches the target.

for the same five microphone locations I used for the full-range measurements.

Before going to the EQ results, I obtained using the miniDSP, let me mention that although multiple subwoofers will inherently flatten the modal response, equalizing them will be even better. If you do have room for four subwoofers, you can separately EQ each one and still drive them with the same mono input. Independently measure each one from several microphone positions, compute an average, and generate a coefficient file. Each of the miniDSP's four outputs can be programmed with a different EQ file using the 2x4 Advanced plug-in. This would work equally well for just two or three subwoofers.

To generate the filter coefficients, open the averaged file in REW and click the EQ icon on the right-side of the top menu bar. The instructions for generating the filter coefficient file is nicely explained in the miniDSP application note, "Auto-EQ tuning with REW," which is included in the Supplementary Material. This application note also explains how to import the coefficient file into the 2x4 Advanced plug-in and program the miniDSP. The EQ Window, EQ Filters, and Equalizer Selection chapters of the REW User Guide will also be helpful (see the REWHelp.pdf).

You can move from item to item by clicking on the desired block in the initial plug-in screen, which is the same as the one shown in **Figure 2**. My single subwoofer has one mono input so I set the mixer selection as IN 1 to OUT 1 and muted everything else. I loaded the filter file into the input five-band PEQ, but the five-band PEQ in the OUT 1 line would work equally well.

Figure 4 shows the comparison between the initial (unequalized) averaged file, the equalization target, and the final equalized averaged file. The bass sounds much smoother so I am keeping the miniDSP in my sound system. The large peak at about 52 Hz was flattened. I decided to ignore the dip at about 16 Hz because it's the peaks that cause the most problems.

miniDSP designs and sells a complete range of digital processing solutions for home-theater, including solutions for commercial audio with balanced I/O. miniDSP also sells the Dirac Series of high-resolution audio processors with Dirac Live, an advanced hardware/software solution for digital room correction (DRC). Other auto-correction products are available and a resource list is included in the Supplementary Material as Automated-Room-Correction-Explained.pdf.

Author's Note: Logically, this month's article should be on the third objected-based surround system, DTS:X. However, there is not yet much to write about. It was introduced in the spring of 2015 as a commercial motion picture theater system and a home system, but affordable integrated receivers supporting the format are just now supposed to be available. As of December 2015, only three Blu-ray movies with a DTS:X soundtrack have been released, Ex Machina, American Ultra, *and* The Last Witch Hunter. *I'll get back to DTS:X in a future article when there is more to write about.*

Project Files

To download additional material and files, visit http://audioxpress.com/page/audioXpress-Supplementary-Material.html.

Resources

AVS Forum, AVR Room Correction Poll, www.avsforum.com/forum/90-receivers-amps-processors/1111941-avr-room-correction-poll.html.

Dr. Earl Geddes, http://gedlee.azurewebsites.net

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Audyssey Laboratories, Inc. | www.audyssey.com

Cantabile 2.0 Solo DAW
Cantabile | www.cantabilesoftware.com

Reaper DAW
Cockos, Inc. | www.reaper.fm

DIRAC Digital Room Correction
Dirac Research AB | www.dirac.com

ARC System 2
IK Multimedia Production srl | www.ikmultimedia.com

Integra AccuEQ
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miniDSP hardware and software plug-in
miniDSP, Ltd. | www.minidsp.com

Onkyo TX-SR444
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Room EQ Wizard
REW | www.roomeqwizard.com

Microphone Preamp Model 411
TDL Technology | www.tdl-tech.com/data411.htm

About the Author

Ron Tipton has degrees in electrical engineering from New Mexico State University and is retired from an engineering position at the White Sands Missile Range. In 1957, he started Testronic Development Laboratory (now TDL Technology) to develop audio electronics. He is still the TDL president and principal designer.

Audio Praxis

One Model to Rule Them All

A Unified Model for Grid-Excited, Single-Triode Circuits

What if I told you that there was one model that could accurately predict the gains, the frequency responses, the impedances, and the power supply rejections (the "Four Horsemen") of three separate circuits: the cathode follower, the cathodyne/concertina/phase splitter, and the voltage gain stage (the "Big Three")? Would you be interested? I hope so, because that's what this article is about. These three circuits differ only in component values and bias voltages, so we should expect some commonality here.

By
Christopher Paul
(United States)

Why seek a unified model? Well, physicists pine for a Unified Field Theory, or a "Theory of Everything." This would illuminate the similarities between things that might otherwise have been thought to be quite different, and that would aid in a better understanding of each one. For similar reasons, we might welcome a single model for multiple circuits.

The Unified Model addresses triodes with arbitrary grid voltage, supply noise, and plate and cathode loads Z_p and Z_k (see **Figure 1**). We use Zs rather than Rs here, because the loads may have capacitive or inductive as well as resistive portions.

To keep things simple, we will restrict ourselves to no grid current and to small signal AC linear operation, and we'll ignore parasitic capacitances. Still, we'll cover a pretty useful range of operation. The goal will be to develop separate models for the plate and cathode circuits (see **Figure 1c**), which will offer clearer insights into circuit operation.

For any of the Big Three, the selection of Z_p and Z_k and the DC voltages E_G and E_B establishes the DC bias of its triode. For a given triode, DC bias determines its amplification factor μ and plate resistance r_p. With a good datasheet, you can read or interpolate these from the curves provided. Alternatively, you can measure them using well-established methods.

We can model the small signal operation of the circuit shown in **Figure 1a** by replacing its triode with the voltage source $\mu \times (e_k - e_g)$ and plate resistance r_p as shown in **Figure 1b**.

Figure 1b also dispenses with the DC voltages, which, having helped to establish triode bias, are no longer useful in a small signal AC analysis. Unfortunately, this does not make us much wiser about any of the Four Horsemen (the gains, the frequency responses, the impedances, and

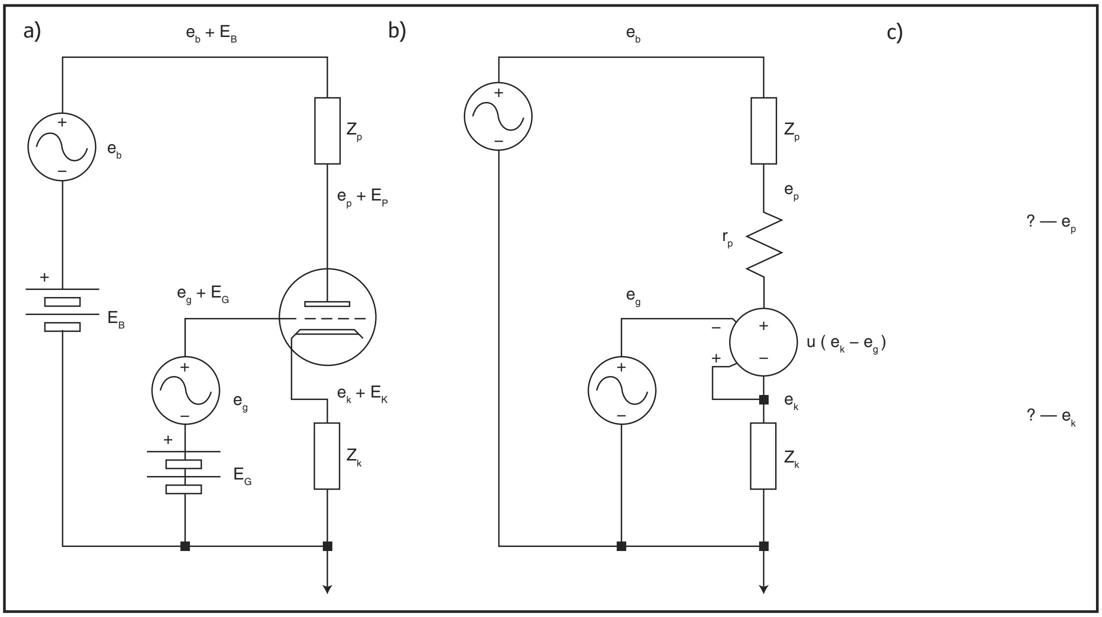

Figure 1: (a) Appropriate selections of values for E_G, Z_p and Z_k can configure this circuit as a cathode follower, phase splitter, or a voltage gain stage. (b) This is an AC small signal model of the circuit shown in (a). (c) Here are separate models of the plate and the cathode circuits to be developed.

the power supply rejections. For that, we have to decompose **Figure 1b** into two separate models—one for the plate circuit and one for the cathode.

Fortunately, we can call upon Kirchhoff's Current Law, which states that the sum of all currents leaving any node in a circuit is zero (arriving currents are considered to be negatives of the departing ones). Two equations result when we apply Kirchhoff's Current Law at the plate and at the cathode:

$$\frac{e_p - e_b}{Z_p} + \frac{e_p - \left[e_k + \mu \times \left(e_k - e_g\right)\right]}{r_p} = 0 \qquad \text{[1] (plate)}$$

and:

$$\frac{e_k}{Z_k} + \frac{\left[e_k + \mu \times \left(e_k - e_g\right)\right] - e_p}{r_p} = 0 \qquad \text{[2] (cathode)}$$

We can solve these equations for the two unknowns, e_p and e_k. Since most *audioXpress* readers don't salivate at the prospect of algebra, allow me to spare you the pain. The results are:

$$e_p = \; < e_g \times (-m) > \times \left[\frac{Z_p}{Z_p + \{z_p\}}\right] + < e_b > \times \left[\frac{\{z_p\}}{Z_p + \{z_p\}}\right] \qquad [3]$$

where:

$$z_p = r_p + (\mu + 1) \times Z_k$$

and:

$$e_k = \; < \frac{e_g \times \mu}{1+\mu} + \frac{e_b}{1+\mu} > \times \left[\frac{Z_k}{Z_k + \{z_k\}}\right] \qquad [4]$$

where:

$$z_k = \frac{Z_p + r_p}{\mu + 1}$$

There's an easy way to check equations [3] and [4]. Just make up a bunch of random numbers for the parameters on their right-hand sides and calculate e_p and e_k. Plug all of these numbers back into equations [1] and [2] and verify that you actually have equalities. Repeat until you're convinced. This is easily done with a spreadsheet,

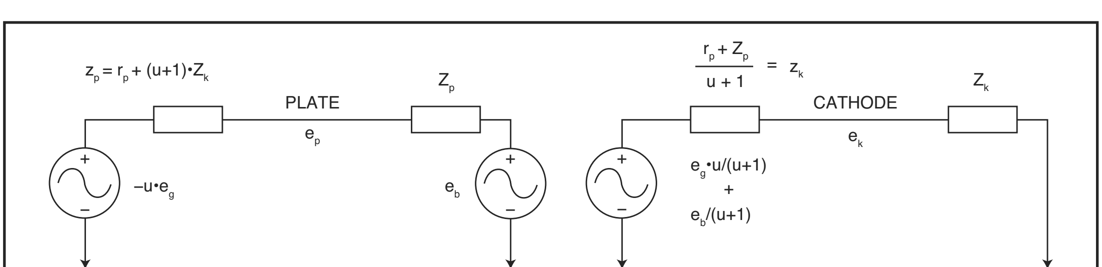

Figure 2: The Unified Model shows the values of three of the Four Horsemen by simple inspection. The fourth, the frequency responses in cases where Z_p and Z_k are partially capacitive and/or inductive, is also accounted for but requires some computation.

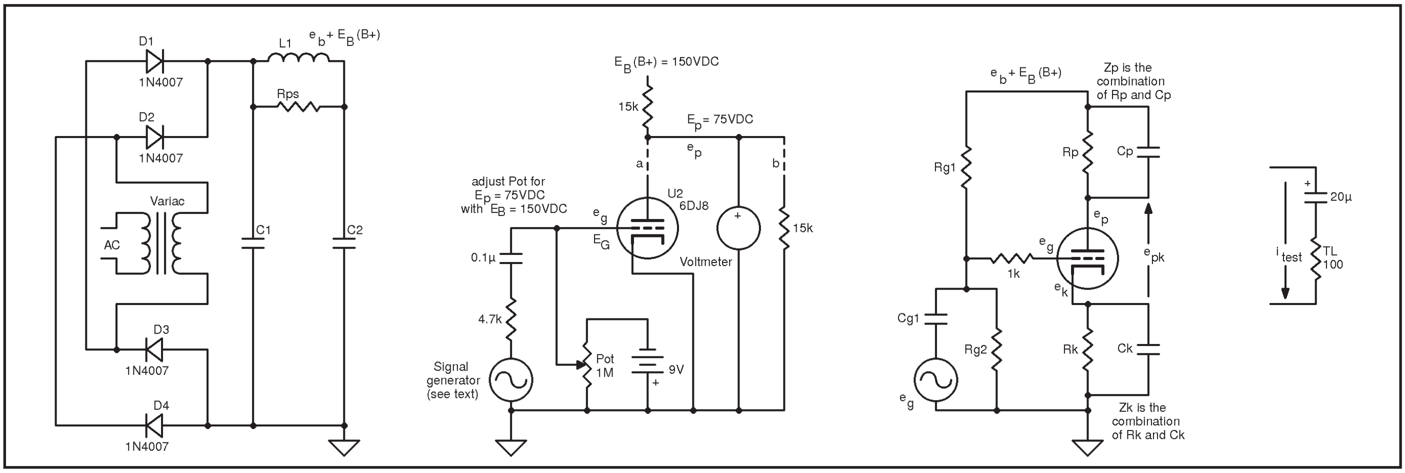

Figure 3: The Test Bed schematics show the equipment used to test our theory. Starting from the left, we see the power supply, the triode test bed, the Test Circuit for the Big Three, and the test load with resistance TL.

	Cathode Follower	**Phase Splitter**	**Voltage Gain Stage**
Values Used for the Test Circuit in All Tests			
R_{g1}	868 kΩ	868 kΩ	not used
R_{g2}	834 kΩ	422 kΩ	not used
C_{g1}	1 µF	1 µF	shorted
R_p	0 Ω	15 kΩ	15 kΩ
R_k	15 kΩ	15 kΩ	302 Ω
Values Used in Test Circuit for Bandwidth Tests Only (not used for other tests)			
C_p	not used	0.0998 µF	0.001 µF
C_k	0.1018 µF	0.1018 µF	0.1018 µF
Values for the Power Supply (Clean), All Except Power Supply Rejection Tests			
E_B (B+)	150 VDC	225 VDC	152 VDC
C_1	330 µF	330 µF	330 µF
C_2	220 µF	220 µF	220 µF
L_1	7 H	7 H	7 H
R_{ps}	10 kΩ	10 kΩ	10 kΩ
Values for the Power Supply (Dirty), Power Supply Rejection Tests Only			
E_B (B+)	150 VDC	225 VDC	152 VDC
C_1	15 µF	15 µF	15 µF
C_2	not used (open)	not used (open)	not used (open)
L_1	0 H (short)	0 H (short)	0 H (short)
R_{ps}	not used (open)	not used (open)	not used (open)
Triode Parameters			
Amplification factor mu (µ)	29.74	Plate resistance r_p	4134 Ω
Test Load Resistance (TL)			
100 Ω			

Table 1: The Test Bed component values, plate supply voltages, and triode parameters are provided.

Table 2: These are the measurements and calculations of voltage gains.

	From Measurements			Calculations of U Model Expressions	
	e_g =	e_p =	e_k =	e_p from eq [3], e_b = 0	e_k from eq [4], e_b = 0
Cathode follower	0.0704	----	0.0675	----	0.0675
Phase splitter	0.5110	-0.4710	0.4720	-0.4747	0.4747
Voltage gain stage	0.0415	-0.6680	0.0135	-0.6515	0.0131

but you can do it without one too. (Some of us call the latter method "plug 'n' chug".)

Equations [3] and [4] describe circuits in which sources of voltages <> and impedances { } drive the loads Z_p and Z_k (see **Figure 2**). Note the voltage divider forms appearing within the [ ] terms. These equations and circuits constitute what I call the Unified (U) Model. They contain the same information, just in different forms.

The U Model has several interesting and useful characteristics. For one thing, it arrives at circuit impedances without employing any test loads! For another, when the loads Z_p and Z_k are resistive only, gains, single-ended impedances, and power supply rejection ratios are all immediately obvious. This helps to provide an intuitive feel for how these parameters vary with component values and circuit type.

The algebra becomes more complicated when the loads contain capacitors and/or inductors. In that case, you can still approximate the impedance of the load at a given frequency and apply it directly to the model. But for more accurate results, it would be better to employ either steady state frequency response analysis (something I do not address in this article), or to just put the U Model in a simulator.

Let's gain some confidence in this model through a series of bench measurements. We will be confirming that the model can predict the Four Horsemen for one example each of the Big Three. The equipment used in the tests is shown in **Figure 3**, and many of its components are found in **Table 1**. The triode heater is powered by a regulated 6.3 VDC power supply connected to ground. All resistors are 1% except for R_{ps} in the power supply and the triode grid stopper resistor, which are 5%. Test frequencies are 1 kHz, except for the additional tests at 10 Hz and 20 kHz in the frequency response section. Voltages are root mean square (RMS) unless otherwise stated.

Triode Small Signal Parameters

First, we measure the triode's small signal parameters to be used in the tests (see **Figure 3**). Configure the power supply with the "Clean" components listed in **Table 1** and connect it to the triode test bed. Leave the triode test bed signal generator off and set the pot so that E_G is −3 VDC.

Make strap b, leaving strap a open. Now adjust the power supply voltage E_B with the variac so that the voltage E_P is 75 VDC. Disconnect strap b and connect strap a. Adjust the potentiometer so that E_P is once again 75 VDC. At this plate voltage, the triode current is 5 mA. These shall be the triode DC plate-cathode voltage and current in all of the Big Three tests. Now set the signal

From Measurements							
Test Frequency		10 Hz		1,000 Hz		20,000 Hz	
	e_g =	e_p =	e_k =	e_p =	e_k =	e_p =	e_k =
Cathode follower	0.03704	----	0.03556	----	0.03484	----	0.01761
Phase splitter	0.04144	0.0394	0.03944	0.03812	0.03888	0.01958	0.01938
Voltage gain stage	0.04156	0.665	0.01345	0.663	0.01332	0.718	0.00806
Calculations of U Model Expressions (Equations 3 and 4)							
Test Frequency		10 Hz		1,000 Hz		20,000 Hz	
	e_g =	e_p =	e_k =	e_p =	e_k =	e_p =	e_k =
Cathode follower	0.03704	----	0.0355	----	0.0354	----	0.0180
Phase splitter	0.04144	0.0385	0.0385	0.0391	0.0383	0.0203	0.0199
Voltage gain stage	0.04156	0.6524	0.0131	0.6563	0.0131	0.7456	0.008042

Table 3: These are the measurements and calculations of frequency response.

	From Measurements			Calculations of U Model Expressions	
	e_b =	e_p =	e_k =	e_p from eq [3], e_g = 0	e_k from eq [4], e_g = 0
Cathode follower	2.4900	----	0.0808	----	0.0803
Phase splitter	2.7200	2.6400	0.0824	2.6350	0.0850
Voltage gain stage	2.4800	1.1680	0.0270	1.1709	0.0264

Table 4: Here are the measurements and calculations of power supply rejection. The voltages are peak to peak.

generator to 1 kHz and 41.56 mV. (The gain knob is very touchy, so I just got close to my 40 mV target.) The AC voltage e_p at the plate is 969 mV. Now connect the Test Load. The drop across the 100 Ω test load resistor is 29.00 mV. The two relevant equations are:

$$\frac{\mu \times 15\text{k}\Omega}{15\text{k}\Omega + r_p} = \frac{969}{41.56}$$

and:

$$\frac{\mu \times L}{r_p + L} = \frac{29.00}{41.56}$$

where L is the parallel combination of 100 Ω and 15 kΩ and is equal to 99.34 Ω. (The impedance of the test load capacitor is negligible in this calculation.) Solving, we get μ = 29.74 and $r_p = \mu/g_m$ = 4,134 Ω.

This completes our tests using the Triode Test Bed. Power down the plate supply and disconnect it.

Measuring the Big Three's Four Horsemen

Now we assemble the "Test Circuit for the Big Three" shown in **Figure 3** for each of the of the Big Three and connect it to the power supply, using the appropriate component values and supply voltages listed in **Table 1**. We'll be recording in various tables the values of e_b, e_g, e_p, and e_k along with currents i_{test} through the Test Load (TL) when we measure impedances. We'll use these along with appropriate component and parameter values from **Table 1**, inserting them into the Unified equations and comparing the results with some simple calculations made from our measurements.

Gain

Gain tests are straightforward. The "Clean" version of the power supply is retained to keep e_b virtually zero. C_p and C_k are not used in this test. e_g is set to the levels listed in **Table 2**, where

	Measurements								
	e_g =	e_{pk} (plate-cathode) =	i_{pk} (plate-cathode) =	e_g =	e_p =	i_p =	e_g =	e_k =	i_k =
Cathode follower	----	----	----	----	----	----	0.03712	0.03512	0.00015
Phase splitter	0.03024	0.10648	0.0003038	4.624	4.248	0.0003	4.624	0.207	0.00031
Voltage gain stage	----	----	----	0.04156	0.658	9.4E-05	0.04156	0.01326	4.4E-05

	Calculations Using Measurements		
	e_{pk}/i_{pk} - TL =	e_p/i_p - TL =	e_k/i_k - TL =
Cathode follower	----	----	128.65
Phase splitter	250.49	14174.19	573.83
Voltage gain stage	----	6885.14	198.65

	Calculations of U Model Expressions		
	Derived from U model, $2 \times (r_p/(\mu+2)) \times (Z_p+Z_k) / (2 \times r_p/(\mu+2)+Z_k+Z_p)$ =	From Figure 2, $Z_p \parallel z_p$ =	From Figure 2 $Z_k \parallel z_k$ =
Cathode follower	----	----	133.29
Phase Splitter	258.25	14531.48	597.65
Voltage gain Stage	----	7082.34	203.34

Table 5. Here are the measurements and calculations of the impedances.

the measurements of e_p and e_k are also recorded. Gain ratios are compared to the results using the Unified equation calculations.

Frequency Response

This test is similar to that for gain except that C_p and C_k are populated as listed in **Table 1**. The circuits are tested at frequencies of 10 Hz, 1,000 Hz, and 20,000 Hz. The Unified equations are again employed for comparison, but evaluations are more complex because of the frequency-dependent effects of C_k and C_p. The data and results are found in **Table 3**.

Power Supply Rejection

Here, e_g is set to zero, and the "Dirty" version of the supply specified in **Table 1** is used. Now the reason for the complex R_{g1}-R_{g2}-C_{g1} supply noise filter network becomes apparent. We want the grid to be noise free.

Anything getting in through the grid would be due to whatever drives our test circuit rather than the test circuit itself. Supply noise is applied to the test circuit only through Z_p. C_p and C_k are not used in this test. The data and results are shown in **Table 4**.

Impedance

Perhaps none of the other parameter measurements is as controversial as that of impedance. To make certain I am correctly measuring impedances, I employ Thevenin's Theorem as stated by the Institute of Electrical and Electronics Engineers (IEEE) in *The New IEEE Standard Dictionary of Technical Terms*:

"Thevenin's Theorem states that the current that will flow through an impedance Z′, when connected to any two terminals of a linear network between which there previously existed a voltage E and an impedance Z, is equal to the voltage E divided by the sum of Z and Z′."

This means that I can apply a test load (TL) between any two terminals of any linear circuit to help measure the impedance Z between them. The capacitive impedance of Test Load (Z′) is effectively zero for the impedances we'll be measuring.

I can measure E with no load and divide it (with the TL connected) by the measured test load current I_{test}. Thevenin says this ratio is "the sum of Z and Z′." From this I can subtract Z′ (the resistive, TL portion) to determine the impedance Z of the circuit between the terminals.

We go back to the "Clean" power supply with $e_b \approx 0$. C_p and C_k are not used. e_g is set to the levels shown in **Table 5**. We wish to determine three impedances: the ground-referenced ones at the plate and cathode and the one between the plate and the cathode.

We start by measuring the voltages between the three unloaded node pairs, and then the currents through the TL with it connected to one pair at a time. We then apply Thevenin. The impedances are z = (unloaded voltage between the node pair)/(current through TL) − TL.

From **Figure 2**, we see that the U Model shows the impedance of the source driving Z_p is $r_p + (1 + \mu)\times Z_k$, and that driving Z_k is:

$$\frac{r_p + Z_p}{1+\mu}$$

The only impedance that is not obvious is that between the plate and cathode. A discussion of how this is arrived at is provided in the following section. The U Model's answer is that:

$$z_{pk} = \frac{2\times Z\times\left(\frac{r_p}{\mu+2}\right)}{Z+\frac{r_p}{\mu+2}}$$

where $Z = Z_p = Z_k$. The results for all impedance tests can be found in **Table 5**.

A comparison of the results from the Unified equations and the measurement calculations

Resources

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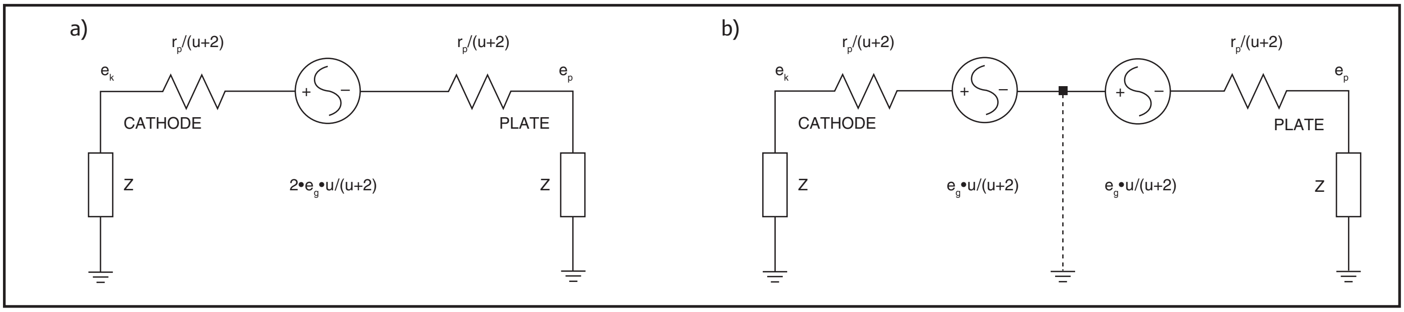

Figure 4: With (a) or without (b) the dotted line connection to ground, the corresponding signals in these two a-S models are indistinguishable.

shows differences of less than 4%, with most considerably less than that.

The Miracle of the Match

As long as the grid current is zero, the magnitudes of the plate and the cathode currents are identical. If we set the supply noise e_b to zero, then when $Z_p = Z_k = Z$ as with the phase splitter, Ohm's Law demands that e_p and e_k have the same magnitude (here, they are negatives of one another.)

Ohm alone is sufficient to explain this, and it neither requires nor demands any particular overall impedances in the circuit. Question: Under these restrictions, can we simplify the U Model? Let's see where Thevenin gets us.

Removing the load on the triode by allowing $Z_p = Z_k = Z$ to increase to infinity, equations [3] and [4] tell us that the equivalent triode voltage source between the plate and the cathode is:

$$e_{ss} = 2 \times e_g \times \frac{\mu}{\mu + 2}$$

If we short the plate to the cathode by setting the Zs to zero, **Figure 2** shows us the current i_{sc} out of the cathode equals the current into the plate:

$$\frac{\mu \times e_g}{r_p}$$

Thevenin tells us the impedance of this source is:

$$\frac{e_{ss}}{i_{sc}} = \frac{2 \times r_p}{\mu + 2}$$

We can split this into equal halves, put one on each side of the source, and add the loads back, all as shown in **Figure 4a**. The circuit components are symmetric about its center, but its voltages are negatives of one another. So we dub the model anti-symmetric, or a-S.

The impedance z_{pk} of the entire circuit between the plate and cathode can be seen from inspection, being equal to the parallel combination of:

$$\frac{2 \times r_p}{\mu + 2}$$

and Z + Z and was given earlier.

We can also split the voltage source of **Figure 4a** into equal halves in series (see **Figure 4b**). Its a-S nature tells us that the voltage at the sources' junction is zero. This means that we can ground this point without disturbing anything. The corresponding voltages and currents in the two circuits are indistinguishable.

Long-time aficionados of the phase splitter/cathodyne/concertina will recognize them. Since the loads can include capacitances, inductances, and resistances, these models will be able to predict frequency responses too. So we've succeeded from the points of view of grid voltage-induced gains and frequency responses, and the plate-to-cathode impedance. But do these models have any limitations?

Consider the power supply noise, which can be minimized but never eliminated. If the a-S models fully reflect how the phase splitter works, then simply adding the noise source e_b between ground and Z_p should enable them to predict power supply rejections (PSRs).

Contrapositively, if this doesn't work, then the models are not the complete and accurate representations we had hoped for. To proceed, we set e_g to zero, because signals on the grid come from the circuit driving the phase splitter, not the phase splitter itself.

To evaluate, let's use the phase Splitter circuit values found in Table 1 for the circuit we measured. We get about 1/2 = 50% for e_p/e_b and e_k/e_b in **Figure 4a** since typically:

$$\frac{r_p}{\mu+2} << Z$$

But we get exactly:

$$\frac{\frac{r_p}{\mu+2}}{Z_p+\frac{r_p}{\mu+2}} \approx 8.6\% \text{ and } 0\%$$

respectively, for these ratios in **Figure 4b** (with the dotted line to ground).

In addition, **Figure 4a** says that the ground-referenced plate and cathode impedances are about Z/2 ≈ 7.5 kΩ while **Figure 4b** says they are about:

$$\frac{r_p}{\mu+2} \approx 130\Omega$$

Surprisingly, the models contradict one another!

Let's compare these calculations to the test measurements in **Table 4**. The attenuation factor of e_p/e_b is about 97%, and that of e_k/e_b ≈ 3.0% (vs. the 97% and 3.1% predicted by the Unified Model.)

For power supply rejection ratio (PSRR), we can think of the plate signal as the output of a voltage divider with an upper branch of Z_p and a lower of z_{pgnd}, the impedance between the plate and ground. Applying the voltage divider rule, 97% implies that:

$$z_{pgnd} = 0.97 \times \frac{Z_p}{1-0.97} = 485\text{k}\Omega$$

(The Unified Model predicts r_p + (1+μ)×Z_k = 469 kΩ.) The test measurements and the Unified model are in close agreement. But the a-S models disagree significantly with them and among themselves, and badly misjudge both power supply rejections and the plate to ground impedance.

What does this mean? First of all, the a-S models still properly predict those portions of the voltages that arise from the grid signal. And, they are right about the circuit's plate to cathode impedance.

But, getting the other parameters wrong demonstrates that they can't determine PSR and mislead about load-driver impedances. In a way, it's a shame that they get some things right. It has led some folks to infer that they get everything right.

The Unified Model, however, with its very different plate and cathode load-drive impedances of r_p + (1+μ)×Z_k and:

$$\frac{r_p+Z_p}{1+\mu}$$

respectively, properly predicts all of the Four Horsemen for each of the Big Three.

What if the Thevenin equivalent of the a-S models' triode source had considered e_b? Well, in that case, we could not have connected it to matched and grounded loads. This is because as noted before, the same current flows through both loads, demanding voltages of identical magnitudes at the plate and cathode. And as we have measured, the power supply noises at these two electrodes are very different.

Summing Up

A Unified Model has been presented which provides for a common, intuitive understanding of how single-triode cathode followers, phase splitters and voltage gain stages work. It was developed by replacing the triode with one of its widely accepted small signal models, applying Kirchhoff's Current Laws to the plate and cathode, and solving for the plate and cathode voltages. The resulting equations led to circuit models from the perspectives of the plate and cathode.

This U Model accurately predicts the Four Horsemen of linear parameters for all of the Big Three circuits. Bench test results of examples of the three distinct circuit types have been shown to be in close accord with the Unified Model's predictions.

The author welcomes comments sent to www.diyaudio.com/forums/tubes-valves/189153-phase-splitter-issue-111.html.

About the Author

Christopher Paul was born in the US the year before the Nobel prize was awarded for the invention of the transistor, the development of which he claims no credit for. His love of music from an early age led to some puzzling experiments with musical instrument amplifiers, and from there to an interest in electronics. Chris finds it interesting to delve into relatively simple circuits and to investigate and write about aspects of their performance that are little known or appreciated. He enjoys deriving equations that can be used to evaluate and design them, although he admits to being in therapy for this. Chris has worked for companies whose products are in the fields of communications, electronic article surveillance, and enterprise hand-held computers. He holds two M.Sc. degrees in electrical engineering from Brooklyn Polytechnic (now part of New York) University.

Calculate Reverberation Time Using Statistical Methods

Last month, we discussed how to measure reverberation time. This article explores various statistical methods that can be used to calculate reverberation time (RT).

By
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Today's acoustician has almost unbelievable modeling tools at his/her disposal, as we have discussed in some depth in previous Sound Control articles. But there are times when virtually all professionals in the field revert to the statistical methods. There are also many sound professionals who need to predict a room's Reverberation Time (RT), but cannot justify or afford the cost of acoustical modeling software. And there is even some well-known acoustical software that uses only statistical methods for RT prediction. All ray-tracing acoustical modeling software includes statistical predictions to provide a point of comparison.

With these things in mind, we should understand the various statistical approaches used in acoustics. To simplify our discussion, we will omit air absorption from consideration. Although important in large venues, especially at high frequencies, absorption of sound by air is easy to conceptualize and to predict.

Sabine's Equation

The first equation for calculating RT was developed by Wallace Clement Sabine as part of his work involving extensive measurements at Harvard University. Sabine assumed that the sound field in the room is perfectly diffuse, meaning that the reverberant sound direction is completely random, as well as being the same at all points in the room at any particular time. This implies all surfaces have the same uniform absorptivity, and all sound sources being centered in the room and nondirectional. Also, the room shape must be such as to preserve complete diffuseness as the sound reflects multiple times: the shape must be "mixing." The Sabine equation using SI units is:

$$RT_{sab} = \frac{0.161V}{S\bar{\alpha}_{sab}}$$

where V = the volume of the room in m^3,

S = the total surface area of the room boundaries and all internal objects, and

$\bar{\alpha}_{sab}$ = the area-weighted average absorption coefficient of all surfaces in the room.

The denominator of the Sabine equation for a room containing n separate surfaces can be calculated as:

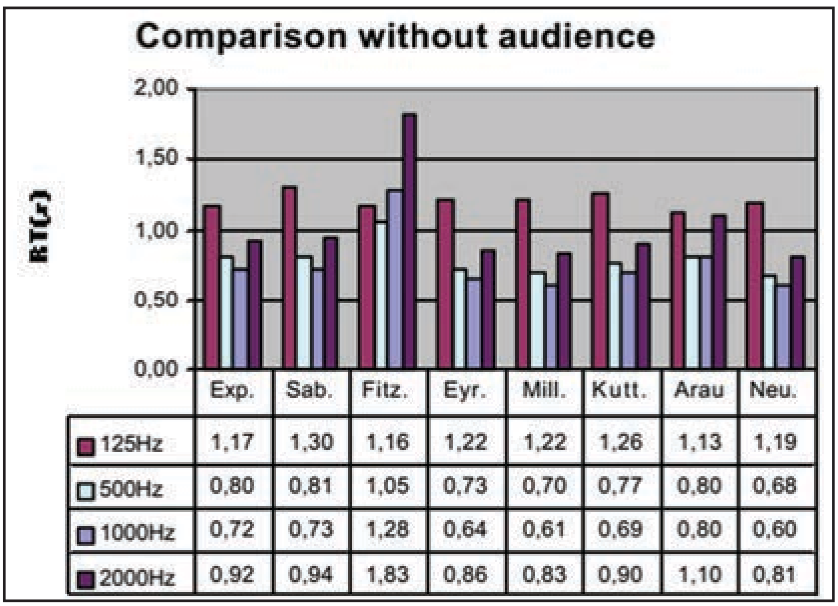
$$S\bar{\alpha}_{sab} = \sum_{i=0}^{n} s_i \alpha_i$$

where s_i and α_i are the area and Sabine absorption coefficients, respectively, corresponding to the different surface materials used in the room.

The Sabine absorption coefficient α_{sab} has often been erroneously described as the decimal fraction of incident energy absorbed by a surface. As Leo Beranek pointed out, this description is nonsensical, since in a room having all surfaces completely absorptive, $\bar{\alpha}_{sab}$ would equal unity, leading to the result RT=0.163V/S; whereas, from basic physics we know that a room with 100% absorption (no reflected energy) would have an RT

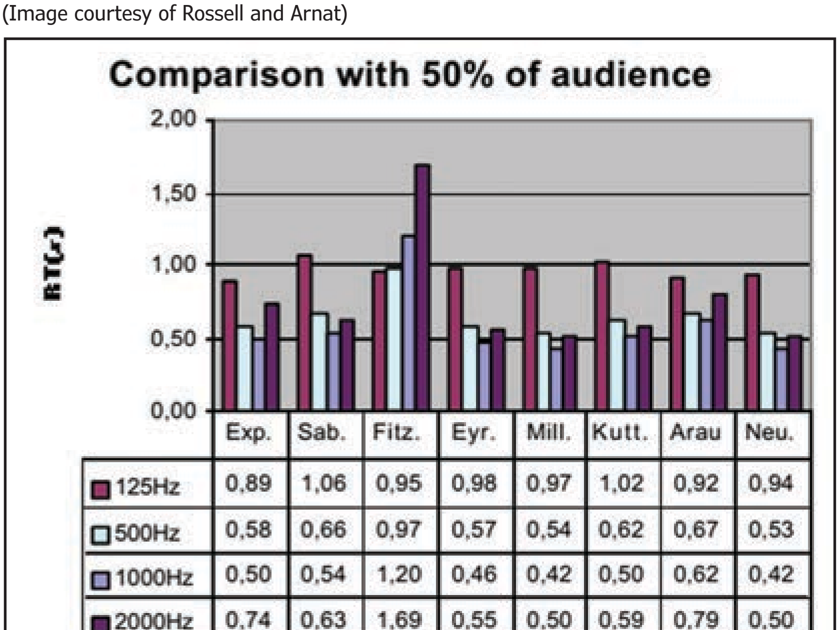
of zero. Obtaining an RT of zero using the Sabine equation would require $S\bar{\alpha}$ to be infinite.

A more precise, though more cumbersome, definition of α_{sab} would be that it is an attenuation constant which, when used in the Sabine equation, yields the RT for a room meeting the Sabine equation's assumptions. This definition implicitly points out that the measured value of α_{sab} for any material depends upon the measurement being made using the Sabine method, as per most international standards.



	Exp.	Sab.	Fitz.	Eyr.	Mill.	Kutt.	Arau	Neu.
125Hz	1,17	1,30	1,16	1,22	1,22	1,26	1,13	1,19
500Hz	0,80	0,81	1,05	0,73	0,70	0,77	0,80	0,68
1000Hz	0,72	0,73	1,28	0,64	0,61	0,69	0,80	0,60
2000Hz	0,92	0,94	1,83	0,86	0,83	0,90	1,10	0,81

Figure 1: The measured and calculated RT are compared, with the room unoccupied. (Image courtesy of Rossell and Arnat)



	Exp.	Sab.	Fitz.	Eyr.	Mill.	Kutt.	Arau	Neu.
125Hz	0,89	1,06	0,95	0,98	0,97	1,02	0,92	0,94
500Hz	0,58	0,66	0,97	0,57	0,54	0,62	0,67	0,53
1000Hz	0,50	0,54	1,20	0,46	0,42	0,50	0,62	0,42
2000Hz	0,74	0,63	1,69	0,55	0,50	0,59	0,79	0,50

Figure 2: The measured and calculated RT are compared, with the room 50% occupied. (Image courtesy of Rossell and Arnat)

The failure of the Sabine equation to predict an RT of zero for a completely absorptive space leads to inaccuracies when the Sabine approach is used for highly absorptive spaces. The identification of such spaces is often given as a space in which $\bar{\alpha}$ is less than some rule-of-thumb value, usually in the range of 0.2 to 0.5.

In an effort to improve RT predictions beyond the results achieved using the Sabine equation, Henry Eyring published a paper in 1930 in which he introduced a new method for calculating RT. Rather than assume, as Sabine had, that as a sound wave travels around a room, it encounters absorptive surfaces sequentially, the Eyring approach (sometimes called "Norris-Eyring") assumes that all surfaces are struck simultaneously by "representative particles" making up the initial sound wave, thereby losing part of their energy depending upon the $\bar{\alpha}$. Then, each of the reflective "particles" travels a distance called the "mean free path" (average distance between collisions of a "particle" with a surface) upon which they all impact another average surface. Thus, the energy in the sound wave decreases in a stepwise fashion, with each "particle" losing the same amount of energy with each collision. (Neither Sabine's assumption nor Eyring's assumption is ever actually fulfilled in real rooms.) Eyring's equation is:

$$RT_E = \frac{0.161V}{S\bar{\alpha}_E} = \frac{0.161V}{-S\left[\ln\left(1-\bar{\alpha}_{sab}\right)\right]}$$

Note that the Eyring equation requires a different absorption coefficient α_E, which can be derived from α_{sab}. The Eyring absorption coefficient can never exceed unity. Many acousticians prefer the Eyring equation for rooms having a low RT; that is, a high $\bar{\alpha}_{sab}$.

W. B. Joyce developed an exact RT equation with no restrictions on room shape, distribution of absorption, or nature of surface irregularities. The calculations must be performed numerically, and even so, are quite difficult except for the simplest of cases.

To more accurately account for the effect of a variety of absorption coefficients, G. Millington and W. J. Sette introduced yet another formulation for the average absorption coefficient:

$$\bar{\alpha}_{MS} = -\sum_{i=0}^{n} s_i\left[\ln\left(1-\alpha_i\right)\right]$$

Where α_i = the Sabine absorption coefficient of each material.

Heinrich Kuttruff corrected two sources of possible error in the Eyring equation by replacing arithmetic averaging by probability distributions, culminating in his equation for the average absorption coefficient:

$$\bar{\alpha}_K = \left[-\ln\left(1-\bar{\alpha}_{sab}\right)\times 1+\frac{\gamma^2}{2}\ln\left(1-\bar{\alpha}_{sab}\right)\right]$$

Where γ = the variance of the probability as a function of the mean free path.

Although some of these approaches allow for a variety of absorption coefficients, none can take into account the effect of unequal distribution of absorption on boundary pairs: side walls, end walls, and ceiling/floor. Daniel Fitzroy introduced an equation that explicitly includes these effects:

$$RT_F = \left(\frac{s_x}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ex}}\right)+\left(\frac{s_y}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ey}}\right)+\left(\frac{s_z}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ez}}\right)$$

Where s_x, s_y, and s_z are the total areas of surfaces normal to the x, y, and z directions, respectively, and $\bar{\alpha}_{Ex}$, $\bar{\alpha}_{Ey}$, and $\bar{\alpha}_{Ez}$ are the corresponding Eyring average absorption coefficients.

Higini Arau-Puchades proposed that rather than an arithmetic average of the x, y, and z "partial RT" values, a geometrical average should be used:

$$RT_{AP} = \left(\frac{s_x}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ex}}\right)^{s_x/S}\left(\frac{s_y}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ey}}\right)^{s_y/S}\left(\frac{s_z}{S}\right)\left(\frac{0.161V}{S\bar{\alpha}_{Ez}}\right)^{s_z/S}$$

Arau-Puchades presented this equation for the "middle" decay; for "early" and "late" decays, modifications are made.

Reinhard Neubauer modified Fitzroy's approach to yield separate predictions of a room's vertical surfaces (walls) and horizontal surfaces (floor and ceiling). This is especially useful in auditoria, where often most of the absorption is due to the audience which is located on the floor surface. Neubauer's equation is:

$$RT_N = \left(\frac{0.32V}{S^2}\right)\left[\frac{h\left(l+w\right)}{\bar{\alpha}_{WW}}+\frac{\text{lw}}{\bar{\alpha}_{CF}}\right]$$

where:

$$\bar{\alpha}_{WW} = -\ln\left(1-\bar{\alpha}_{sab}\right)+\left[\frac{\left(1-\bar{\alpha}_{WW\,sab}\right)\left(\bar{\alpha}_{sab}-\bar{\alpha}_{WW\,sab}\right)S_{WW}{}^2}{\left(1-\bar{\alpha}_{sab}\right)^2 S^2}\right]$$

and:

$$\bar{\alpha}_{WW} = -\ln\left(1-\bar{\alpha}_{sab}\right)+\left[\frac{\left(1-\bar{\alpha}_{CF\,sab}\right)\left(\bar{\alpha}_{sab}-\bar{\alpha}_{CF\,sab}\right)S_{CF}{}^2}{\left(1-\bar{\alpha}_{sab}\right)^2 S^2}\right]$$

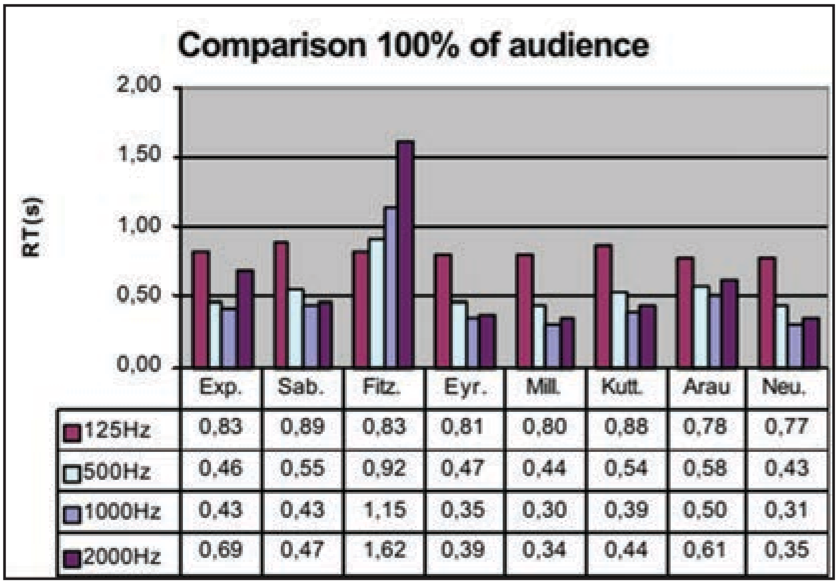
The subscripts WW and CF represent average absorption coefficients or surface areas of the side and end walls, and the ceiling and floor, respectively.

Stefan Drechsler developed an anisotropic reverberation model (ARM), a compromise between simpler analytic models and the more accurate but more costly ray tracing models. The ARM leads to a system of linear ordinary differential equations which are most readily solved by computer.

Ivana Rossell and Isabel Arnet of the Universitat Ramon Llull in Barcelona compared the predictions obtained by using these different approaches to calculate RT of a classroom with 0%, 50%, and 100% occupancy, using absorption data for occupied and unoccupied chairs that they measured in the university's reverberation chamber.

Figures 1–3 show the results of the various calculations, compared with actual measurements, for four octave bands. The measured RT is donoted as Exp. Notice the variation with frequency as to which method gives most accurate results. These figures are from Rossell and Arnet's paper (see Resources).

Figure 4 shows a comparison of error in graphical form. Arau's approach performs well; in fact, all of the approaches tested give reasonable



	Exp.	Sab.	Fitz.	Eyr.	Mill.	Kutt.	Arau	Neu.
125Hz	0,83	0,89	0,83	0,81	0,80	0,88	0,78	0,77
500Hz	0,46	0,55	0,92	0,47	0,44	0,54	0,58	0,43
1000Hz	0,43	0,43	1,15	0,35	0,30	0,39	0,50	0,31
2000Hz	0,69	0,47	1,62	0,39	0,34	0,44	0,61	0,35

Figure 3: The measured and calculated RT are compared, with the room 100% occupied. (Image courtesy of Rossell and Arnat)

results, except for Fitzroy's, which is only accurate for the 125-Hz band. It is important to recognize that these comparisons are only valid for the actual room that was studied, for reasons that will become clearer in the following paragraphs.

All methods for calculating reverberation are built upon certain assumptions, as we have discussed. Both the extent and the manner in which real rooms deviate from the assumed criteria vary from room to room. Thus, the fit between calculated and measured RT using any particular method is room-dependent. A long-standing problem in architectural acoustics is the deviation in measured absorption values when the same material is measured in different laboratories. For years, this difference has been attributed to "edge effect," a putative but not rigorously defined behavior related to diffraction. The ongoing research on edge effect will be discussed in a future column, but for now, it is enough to point out that:

(1) Absorption coefficients exceeding unity are real, and not necessarily a consequence of differences among labs, or errors in testing

(2) The measured absorption coefficient of any material is correct only for the particular piece (including size and shape) of material measured, when it is used in the particular location in the particular room in which the measurement was performed.

This is a matter of some concern, since the accuracy of the absorption coefficient we use ultimately limits the accuracy of both calculation and ray tracing methods.

The most efficient way to calculate RT of a room is to set up a spreadsheet containing the octave-band absorption coefficients of the materials used in the room. The equations for the preferred calculation method are entered in some cells; the chosen RT equation is used to calculate octave-band RTs; and then the octave-band results are averaged as desired. For a general-purpose spreadsheet that will be useful for different rooms in the future, the octave-band absorption data can be entered into a

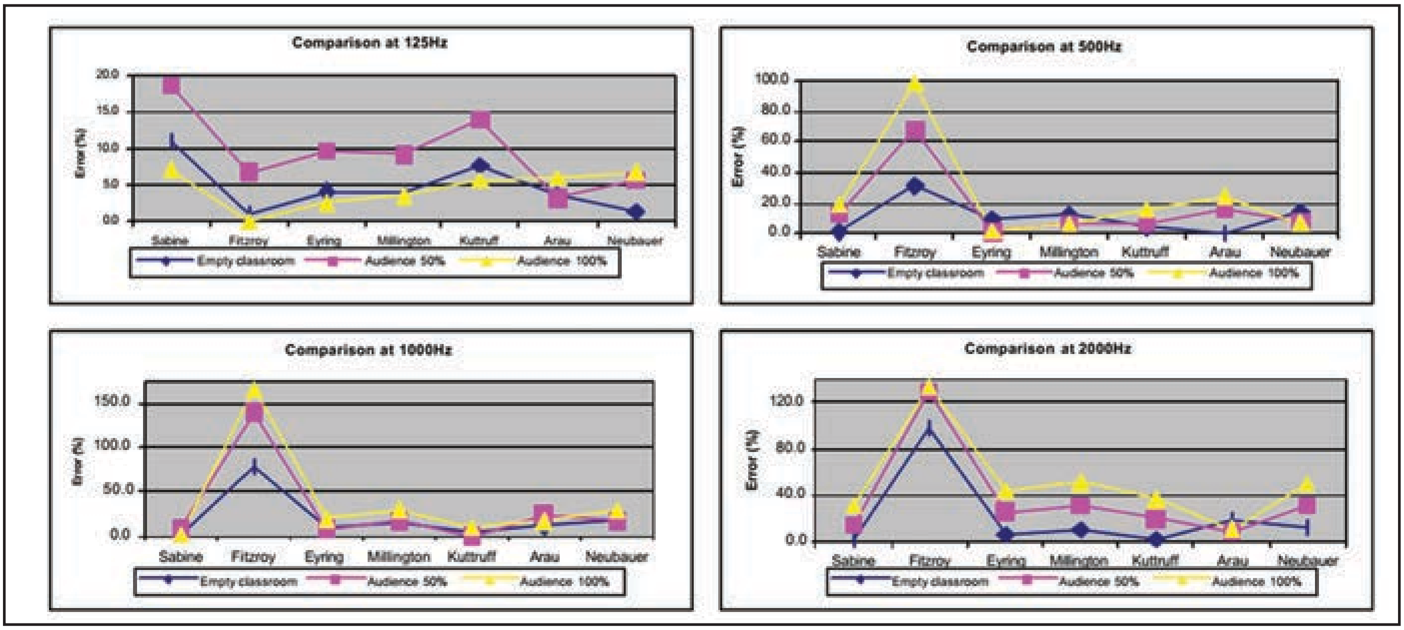


Figure 4: The prediction errors for the various methods are compared. (Image courtesy of Rossell and Arnat)

database from which the calculation spreadsheet can draw. This data is available from numerous sources, including textbooks, technical journal papers, and many Internet pages. Preparing such a comprehensive spreadsheet with a database is time-consuming, but most acoustical consultants have found that ultimately the time investment is more than repaid by the convenience of making RT calculations for rooms whose complexity and/or budget do not justify computer modeling. The spreadsheet approach also speeds the task of running instant “what-if” scenarios involving material changes. Having chosen a particular method of calculation, the acoustician will develop a feel for the kinds of rooms in which the calculated results are probably trustworthy, or may under- or over-predict the RT value. Incidentally, developing this feel for accuracy (which necessitates making measurements of the RT in the finished room) is also important as a sanity check for computer models. For while modeling software may be extremely good, either poor absorption data or faulty modeling practices can torpedo the accuracy of the results.

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Headsets

Testing Headphones and Earphones

The G.R.A.S. 43BB Low-Noise Ear Simulator

Photo 1: The complete G.R.A.S. 43BB solution is supplied in its own case, complete with the G.R.A.S. RA0234 low-noise ear simulator, G.R.A.S. 40AH 1/2" high-sensitivity extension polarized pressure microphone, G.R.A.S. 26HG 1/4" preamplifier with 3-m integrated cable, G.R.A.S. 26HT gain and filter unit for 40HT, and G.R.A.S. 12HF one-channel power module for low-noise systems.

Readers know that when testing speakers, at least one approach is to position the speaker in an anechoic chamber, point the microphone at the speaker, and run the test sequence with an acoustic analyzer. But how about headphones, earphones, and hearing aid devices? In this review, we focus on the new G.R.A.S. 43BB Low-Noise Ear Simulator System.

By
Mike Klasco
(United States)

The human ear has various chambers and represents a complex acoustic load. In non-human testing an acoustic coupler is used for simulating the acoustic load of a human ear canal to more accurately attain realistic test data. Over the years there have been several attempts at developing a practical ear simulator coupler standard, eventually leading to the International Electrotechnical Commission (IEC) 60318 standard.

Today, the IEC 60318 standard is most commonly used for measuring "high-frequency" audiometric headphone responses and for low-cost telephone handsets and general consumer circum/supra-aural headsets. The IEC 60711 standard is used for insert-type ear phones. This coupler (711) became an IEC standard in 1981 and is standardized up to 10 kHz. The 318 is standardized to 8 kHz.

These solutions are available from G.R.A.S. Sound & Vibration A/S, Brüel & Kjær Sound & Vibration Measurement A/S, Larson Davis, CRY and other acoustic test instrument vendors. The 711 was mostly conceived for telephone transmission quality, and calibration is limited to 10 kHz, which is not adequate for the wideband standards coming for the next generation of smartphones, let alone audiophile grade earphones and headphones.

Today, when measuring premium headphones with extended top-end response (e.g., the Sennheiser HD800 claims response to 44 kHz) and when benchmarking against these flagship headphones, the response above 10 kHz is often measured with the headphone driver removed and mounted on a baffle with a precision wideband microphone in the near field. When testing, all you can do is keep the test setup the same and check the relative extended range response between products. But, what is linear to the ear's perception is anyone's guess—not exactly precision data!

G.R.A.S. RA0045

The G.R.A.S. Type RA0045 ear simulator is an example of a 711 standard coupler for making acoustic measurements on earphones coupled to the human ear by ear inserts (e.g., tubes, ear molds, or ear tips). It is delivered with a built-in G.R.A.S. 0.5" pressure microphone Type 40AG and an individual calibration chart for the coupler-microphone combination (see **Photo 1**). The 711 coupler is the heart of the G.R.A.S. Artificial Ear Types (the 43AC, the 43AG, etc.) and is found in the KEMAR head commonly used for headphone testing.

The problem with the 711 standard is this design results in a high Q peak centered about 13.5 kHz, ±1 kHz and the existing 711 coupler's noise floor (mic plus preamp noise) are not as quiet as you might like for active noise control (ANC) work or testing super low distortion products.

After 35 years of the 711 standard, we now have a significant refinement with the new G.R.A.S. 43BB (see **Photo 2**). The G.R.A.S. 43BB enables realistic testing out to 20 kHz and has noise floor so low that ANC residual noise and ultra-low distortion testing are now practical.

Noise Floor

Having the noise floor be below the threshold of human hearing is obviously desirable for designing and testing world class headphones. It is also desirable for those who want to design consumer headphones with active circuitry (i.e., ANC, DSP, Bluetooth, and wireless) enabling validation of much of the electronics.

The 43BB consists of the standardized IEC 60318-4 ear simulator and the G.R.A.S. 40HT low-noise microphone system. The 43BB is based on the standard 711 ear simulator and the 40HT system. There's a fair bit of engineering in that mix to blend the two together. The G.R.A.S. 43BB's low noise floor—below 10.5 dBA—can measure sound levels below or close to the threshold of human hearing (see **Figure 1**).

For comparison, a standard IEC 60318-4 (711) ear simulator with a G.R.A.S 40AG 0.5" microphone has a noise floor at 24.2 dB(A), not quite as good as some ANC headphones. Yet, you can still hear the microphone's residual noise and the microphone preamp noise of these headphones in a quiet hotel room. So one of the motivations to develop the 43BB was to meet the audio industry's need for R&D testing of in-ear headphones with ANC.

Photo 2: The new G.R.A.S. 43BB solution enables realistic testing out to 20 kHz and has noise floor so low that ANC residual noise and ultra-low distortion testing are now practical.

Because it can measure down to and below the threshold of human hearing, it can measure the influence of the electronics on the audio qualities of the earphones (see **Figure 1**).

The other goal for the 43BB is that it can be used for measurements above 10 kHz, because its resonances above that frequency are well-controlled, with low Q, and repeatable, making it much improved over the existing standardized artificial ear. Below 10 kHz, the frequency response is identical to that of a standard IEC 60318-4 ear simulator. Above 10 kHz, the differences in

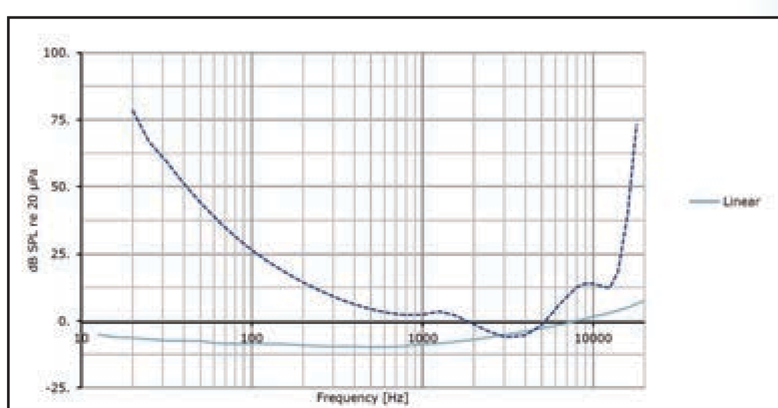


Figure 1: The challenge is hearing threshold vs. test equipment. Having the noise floor below the threshold of human hearing is obviously desirable for designing and testing world-class headphones. The noise floor (solid curve) is typically below the threshold of human hearing (dashed curve) as it is defined in "ISO 389-7:2005 Acoustics - Reference zero for the calibration of audiometric equipment (Part 7): Reference threshold of hearing under free-field and diffuse-field listening conditions."

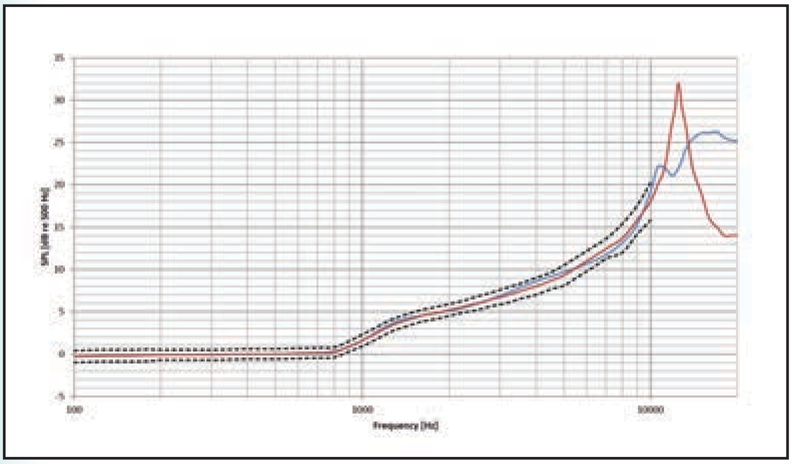


Figure 2: The typical frequency response for 43BB (solid blue) is compared to the IEC tolerances (dashed black), and the ideal IEC 60318-4 frequency response (red curve).

the microphone diaphragm impedance results in substantial differences.

The standard coupler has a high-Q resonance around 14 kHz related to the length of the ear canal and the diaphragm impedance. This sharp resonance does contribute to large measurement uncertainties. The IEC standard states, "Due to resonances in the acoustic transfer impedance of the occluded-ear simulator above 10 kHz, high measurement uncertainties, (e.g., in the order of 10 dB), can occur in earphone responses." **Figure 2** shows a typical 43BB frequency response compared to IEC standards.

The single high-Q resonance is replaced by two balanced and more damped resonances. One is related to the length of the coupler cavity, which is the same as for the standard coupler. The other is related to the undamped diaphragm resonance of the low noise microphone. The much lower Q of the two high-frequency resonances means the low-noise ear simulator can be used above 10 kHz with good repeatability.

With the G.R.A.S.'s low noise microphone inside the ear simulator, the microphone's high-frequency performance is tuned to coincide with the coupler's naturally occurred Lamda/2 resonance (that is the 14 kHz peak). The result is a much smoother "double bump" as opposed to the violent high-Q resonance of the 711.

If you look at the IEC 60318 standard, it says to not trust your measurements above 10 kHz, and G.R.A.S. has conquered this limitation. Its low noise floor and usability above 10 kHz means

A New Soft Ear

The new G.R.A.S. KB5000 right pinna with anthropometric concha and ear canal for KEMAR is made of soft silicone (35 Shore OO hardness). The external shape of the pinna is identical to that of the standardized KEMAR pinna, but the concha and the canal have been modified so that they closely mimic the properties of a real human ear. The ear canal has been extended and is an integral part of the pinna that now seals directly against the ear simulator. As with the human ear, the ear canal now has the first and second bend, and the interface with the concha is oval.

This shape makes it possible to achieve good fit and sealing with anatomically shaped ear-bud headphones and in-ear hearing protectors. Controlling the insertion depth is easy, leading to better insertion consistency and improved repeatability of measurements.

The outer pinna has the same shape as the standardized pinna, but the flexibility has been improved to better mimic the way the human ear collapses when supra-aural and circum-aural earphones are mounted. When measuring the frequency response of these types of headphones, more reliable and repeatable measurements can be achieved because of the improved collapsability of the pinna.

In addition to the traditional push mounting from the outside, the pinna is secured with two screws from the inside of KEMAR's head. These two screws ensure that the pinna is held firmly in place. Therefore, it seals perfectly against the ear simulator and the head, and it is possible to repeatedly mount and dismount DUTs without compromising the seal.

At first sight, there is virtually no difference between the standardized KEMAR pinna and the new G.R.A.S. KB5000 right pinna. Touch will reveal the new pinna to be softer since it is made of soft silicone, allowing it to be pressed against the head, just as with a real ear.

Headsets

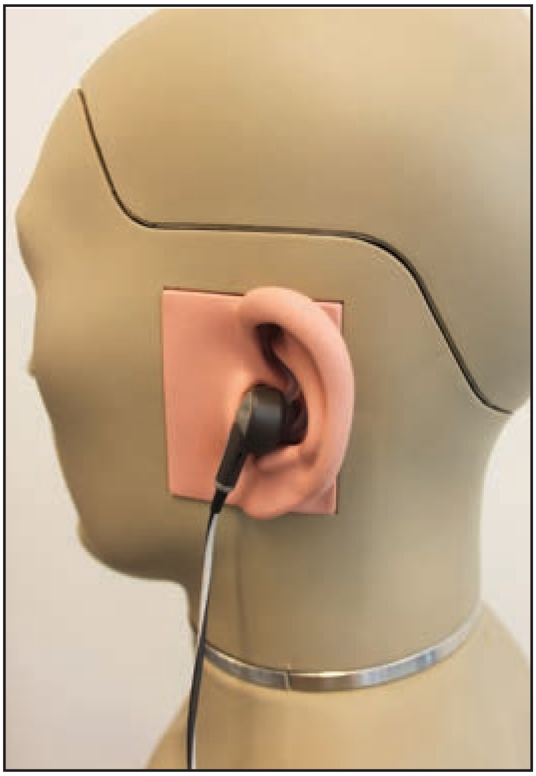


Photo 3: The new G.R.A.S. KB5000 right pinna with anthropometric concha and ear canal for KEMAR is made of soft silicone. For products going into the concha and ear canal, the new pinna with the anthropometric ear canal offers a much more natural and realistic fit.

that measurement results will have a strong correlation with the subjective feedback from test persons and users.

A Tool for Better Sound Quality

The system's low inherent noise also means that the device under test's (DUT) total harmonic distortion (THD), Rub & Buzz, and related perceptual distortions at very low levels can be investigated. This is intrinsically important in earphones and headphones. When a speaker system is listened to, much of the sound is reflected, and while this may provide some naturalness, the pristine clarity of the drivers is perhaps less critical. The designer may get away with at least a bit less than perfection. The typical listening room's ambient noise rarely reaches down to 25 dBA, but in the same room the typical closed-back headphone or earphone provides another 10 dB+ dBA passive attenuation to the ambient noise. So headphones and earphones

The KEMAR

The ear simulators can be incorporated into a Head-and-Torso Simulator (HATS) such as KEMAR, which lends more realism to in-situ anthropomorphic testing. KEMAR is a model of a human head and torso, and has been extensively used for the last 35 years to study the interaction between the human head and torso and sound fields.

The KEMAR head-and-torso simulator for hearing aid tests, ear- and head- phone tests, and sound quality recordings was developed by Knowles Electronics in 1972 and acquired by G.R.A.S. Sound & Vibration in 2005. It is the origin of all other head-and-torso simulators, and thus, the industry standard for in-situ anthropomorphic testing of all kinds of hearing instruments and head- and earphones. The data has been thoroughly documented in numerous studies, some of which have been collected in the book *Manikin Measurement* by Mahlon D. Burkhard (February 1978).

Introduced in 2013, the new generation KEMAR is available with and without a mouth simulator and fully backward acoustically compatible with earlier KEMAR models. It meets the requirements of American National Standards Institute (ANSI) S3.36 and International Electrotechnical Commission (IEC) 60318-7. KEMAR and can also be configured with more sizes of standardized pinna simulators and various 0.5" and 0.25" pressure microphones for binaural recordings. KEMAR accommodates for LEMO as well as CCP pre-amplifiers which are all electrically accessible from the connector panel on the back.

The new generation KEMAR is available with and without a mouth simulator and fully backwards acoustically compatible with earlier KEMAR models.

provide very near field direct sound and lower ambient noise, enabling the driver characteristics to be perceived with more detail.

Headphone driver microspeakers often use a film diaphragm, mostly commonly polyethylene terephthalate (PET). More than 60% of the contributing sound is the "surround," which flexes and makes break-up noises doing so. There are better materials, but measuring these quantitatively has been elusive.

The 43BB is a powerful tool in conjunction with a sophisticated acoustic analyzer for evaluating damped layer laminated diaphragms, sputtered coatings for diaphragms, treating lead-out wires that otherwise might be whipping, ferrofluid for voice coil anti-rocking, damping, and other refinements and separating wisdom from witchcraft in analyzing remedial measures. The 43BB's low noise floor and usability above 10 kHz means that measurement results will have a strong correlation with the subjective feedback from test persons and users.

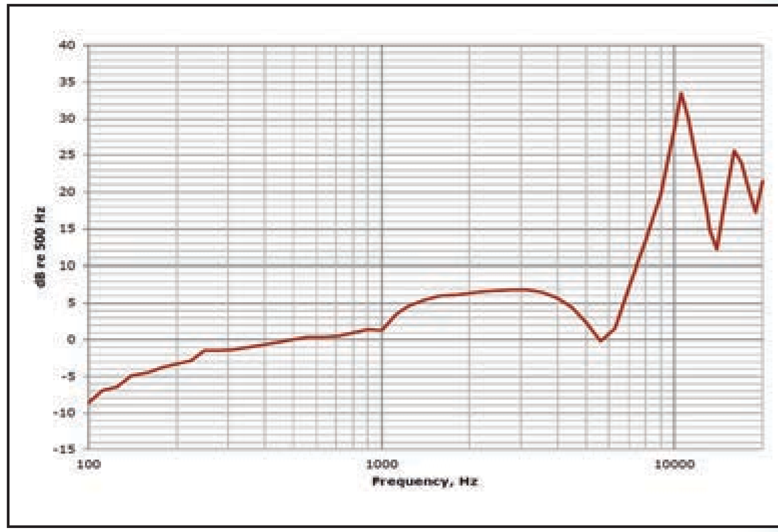


Figure 3: The typical acoustical input impedance of the RA0234 is shown from the ear entrance point. The input impedance is defined as the ratio of the sound pressure at the ear entrance point to the corresponding particle velocity. The sound pressure is measured with a probe microphone while a constant particle velocity is maintained via a high acoustic impedance sound source.

New Outer Ear for Improved Fit, Placement, and Seal

Testing ANC headphones and earphones typically focuses on low-end signal-to-noise ratios (SNRs). The big issue is both consistency and accuracy of the DUT's fit to the coupler with realistic and repeatable results. Fit, placement, and seal (isolation) are significant aspects. Precision low-frequency response and super isolation are the benefit of the new pinna, concha, and canal of this new artificial ear simulator.

The main challenges with the standardized pinnae are the cylindrical metal ear canals and the rigidity of the pinnae in the Z-plane (toward the head). For products going into the concha and ear canal, the new G.R.A.S. KB5000 pinnae with the anthropometric ear canal offers a much more natural and realistic fit (see **Photo 2**). It greatly improves repeatability and enables engineers to test final version DUTs (with the ear tips fitted) more accurately in a much shorter amount of time. And because you have greater fit and more realistic seal, you are able to better characterize ANC (if applicable) and low-frequency performance as well as low-end passive isolation.

For circumaural products, the pinnae's angle of protrusion is an accurate and troublesome feature. Because the pinnae were originally designed to hold a behind-the-ear (BTE) hearing aid, there was not much thought as to how they would collapse against the head under a force from a telephone handset or headphones. As a result, slimmer designed headphones are sometimes lifted off the side of the head and not correctly sealed.

No proper seal, means there will not be a good indication of ANC or bass performance. This is remedied by removing material behind the pinnae, so the actual shape stays the same (as does the HRTF), but the pinnae are now allowed to compress against the head, just as with a real ear (see **Photo 3**).

Physical Implementation and Upgrading Acoustic Test Jigs

The 43BB can be mounted in a monaural setup inside a KEMAR head (e.g., the 45BB and the 45BC) or a G.R.A.S. 45CA Hearing-Protector

G.R.A.S. Sound & Vibration Legendary Founder Turns 90

Distinguished acoustic engineer Gunnar Rasmussen, founded G.R.A.S. Sound & Vibration in 1994, and on his 90th birthday continues to inspire with his extraordinary vision and guiding principles for the Danish measurement microphones specialist.

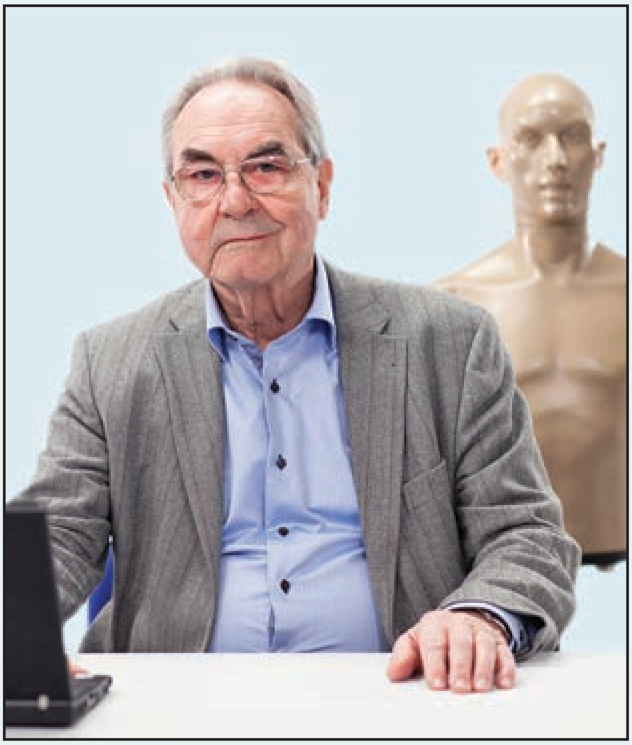
As an engineering student Rasmussen already showed great promise. Shortly after his graduation, he was hired by the Danish audio pioneers Brüel & Kjær. In the 1950s, the company sent Rasmussen to the United States to promote products and manage the after-sales service. The inquisitive microphone specialist also took the opportunity to get acquainted with America's futuristic technologies and to forge ties with a wide range of industries.

The fast-growing American automotive, aerospace, and consumer industries in the 1950s had a continuous demand for more and better acoustic measurement microphones. Through his intense involvement in several different acoustical measurement projects, Rasmussen realized the microphones available at that time needed some serious improvements. After returning to Denmark, he set out to develop a new generation of microphones, using the most advanced materials and assembly techniques. The outcome was a new, revolutionary, and simple design, which for the past 60 years has been preferred for all types of professional acoustic measurements and has been internationally standardized.

At the age of 67, when most people are heading for retirement, Rasmussen left Brüel & Kjær and founded G.R.A.S. Sound & Vibration. What began as a small family business, run by Rasmussen and his wife Hanna Hertz, quickly grew into a respected and leading manufacturer of measurement microphones, turning many of its founder's so-far unexploited ideas and inventions into commercial successes. Today, the company offers more than 100 different microphones types, employs 90 people, and has offices or affiliates in approximately 40 countries worldwide.

The couple's sons, Per Rasmussen and Peter Wulf-Andersen, play key R&D roles at G.R.A.S. Sound & Vibration, while Lars Kjærgaard, who joined the company in 2011, has been serving as CEO since 2014. As Chairman, Rasmussen continues to inspire those around him with his extraordinary vision and guiding principles for the company.

He was appointed as chairman of the ISO committee, made a Fellow of the Acoustical Society of America, and he received the Danish Design Awards for his design of measurement microphones and sound level meters. The famous conductor Leonard Bernstein turned to Rasmussen for help producing a better recording microphone for classical music. Rasmussen earned a CETIM award and the International Electrotechnical Commission (IEC) Award. In 2008, he was presented with the prestigious Lifetime Achievements in Acoustics award by the European Acoustics Association. However, Gunnar Rasmussen does not dwell much on accolades. On the contrary, he seems to be looking ahead to solve the next problem. For more information, visit www.gras.dk.



Acoustic engineer Gunnar Rasmussen, founded G.R.A.S. Sound & Vibration in 1994, and on his 90th birthday continues to be one of the leading Danish measurement microphone specialists.

Test Fixture. Because of the length of the 26HG 0.25" preamplifier, it is not possible to use the 43BB in a binaural setup inside a G.R.A.S. test fixture. Although the company recommends the new anthropometric pinna with integrated ear canal (KB5000), there is no need to toss your existing test fixtures. You can mount the 43BB in a KEMAR (single channel) if you use a standard ear canal extension (e.g., the G.R.A.S RA0237 Straight Ear Canal Extension Kit for KEMAR). If mounted in the G.R.A.S. 45CA, a G.R.A.S RA0172 Pinna Holder Kit can be used (see **Figure 3**).

The 43BB is based on the standard 711 ear simulator and the 40HT system so the 43BB can be fitted with a right angle adapter that will allow for 43AG implementation and binaural fit in a KEMAR. Binaural fit will however not be possible in the 45CA.

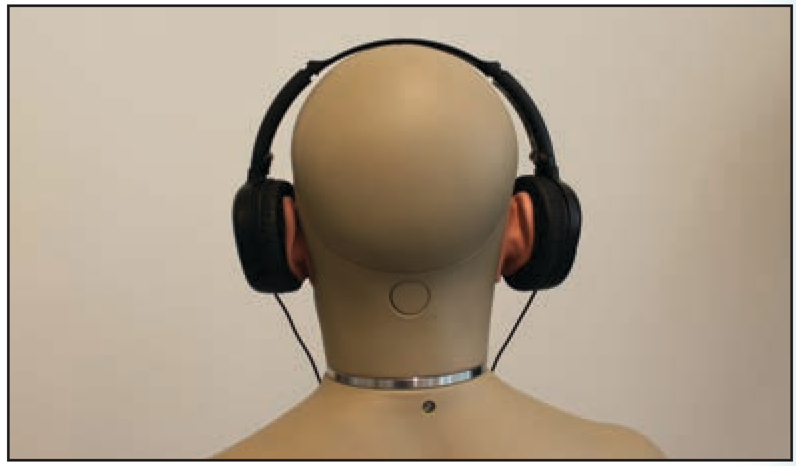


Photo 4: Because the pinnae were originally designed to hold a behind the ear (BTE) hearing aid, there was not much thought as to how they would collapse against the head under a force from a telephone handset or headphones. The new G.R.A.S. KB5000 pinna allows compression against the head, just as with a real ear.

Overall Opinion

This refined next-generation ear simulator was a project designed to improve fit and placement of in-the-ear, on-the-ear, and over-the-ear products. The repeatability of low-frequency response has been achieved along with stability of high-frequency response and greater dynamic range in the low end. The ability to use this new coupler as a laboratory tool for quantitative evaluation of exploratory techniques to refine high-grade earphones and headphones is exciting as well as a product review tool delivering higher correlation with subjective feedback while maintaining the aspects that are viable of the familiar 711 coupler.

I Can Hear It. Why Can't I Measure It?

In many cases, the reason given as to why a particular audible "distortion" cannot be measured is that the measurement equipment doesn't have sufficient signal-processing capabilities, such as measurement bandwidth or Fast Fourier Transform (FFT) resolution. In this article, we will explore an overlooked contributor—the instrumentation noise floor, which can have a significant impact on what can be measured, especially with low sound pressure level (SPL) signals.

By
Dan Foley
(United States)

Prior to headphones and insert earphones (earbuds) becoming the common way to listen to music, background noise from household appliances or heating and air conditioning fans in a typical listening environment could potentially mask playback system distortion, such as higher-order harmonics of AC line hum or switching amplifier intermodulation distortion (IMD). Typical suburban homes have background noise levels of 30 to 38 dBA, with a noise spectrum corresponding to the Noise Criterion (NC) 30 curve (see **Figure 1**). Apartment noise levels are even higher.

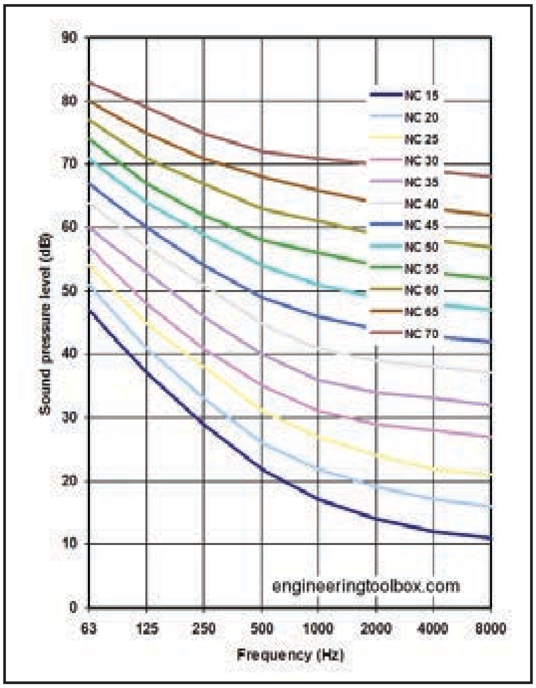


Figure 1: This is the noise criterion (NC) curve for a typical suburban home.[1]

To confirm this, I measured the background noise in my living room and positioned a G.R.A.S. 0.5" microphone (Model 46AE) on a tripod at ear height, and used an Audio Precision APx515 analyzer to measure the corresponding background noise, using a 64k FFT with eight linear averages. The APx Octave Band Utility was used to synthesize an octave-band spectrum. The overall sound pressure level was 38 dBA and the measured ambient noise spectrum just exceeds the NC-30 noise criterion curve by about 1 dB (see **Figure 2**).

However when one listens to music using circumaural headphones or insert earphones, the listening environment background can be attenuated by 20 dB or more, dramatically reducing the masking that was present due to background noise (see **Figure 3**). As a result, the listener's perception during very quiet passages more closely follows the threshold of the hearing curve, shown as a red line in **Figure 4**, bringing sensitivity to a sound pressure level (SPL) below 0 dB (20 micropascals). From a headphone/headphone preamplifier system perspective, this means that distortion artifacts could be as low as 0 dBSPL and still be audible.

Impact On Measurements

So how does all of this impact measurements? Every part of the measurement signal path introduces noise, beginning with the microphone capsule and its corresponding preamplifier. The IEC 711 couplers that are used in Head-and-Torso Simulators (HATS) test manikins and equivalent products typically incorporate a 0.5″ microphone with a 12.5 mV/Pa sensitivity or –38 dB referenced to 1 pascal.

Given that 1 pascal is equivalent to 94 dBSPL, the lowest SPL one can potentially hear around 3.2 kHz is –5 dBSPL. At 94 dBSPL the voltage from the microphone is 12.5 mV. At –5 dBSPL, the voltage level will be 99 dB below this value, which is 141 nanovolts!

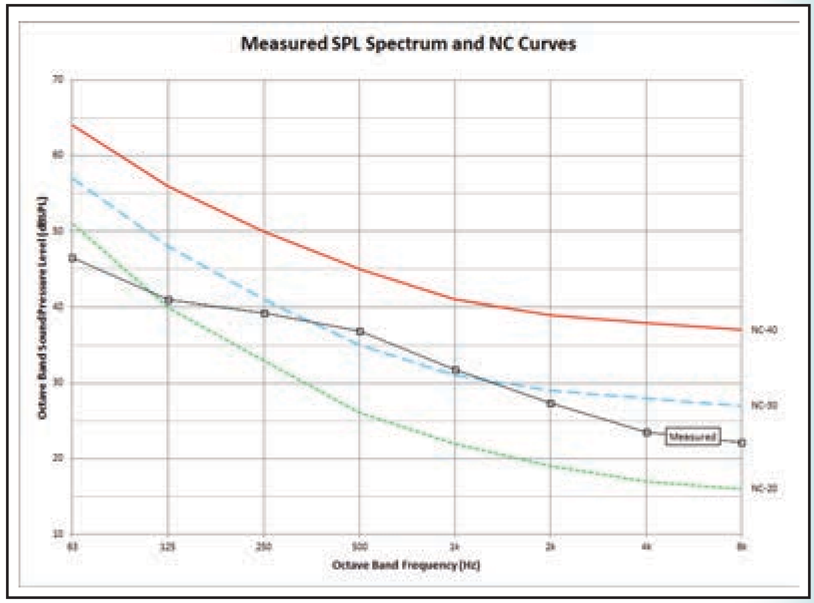


Figure 2: The background sound pressure level (SPL) for my living room is shown.

Instrumentation noise should be 10 dB below the level one wants to measure. In this case, the noise level in this frequency range would need to be 47 nanovolts, but it does not mean the instrumentation noise floor over the entire audio band needs to be 47 nanovolts. But, the lower the

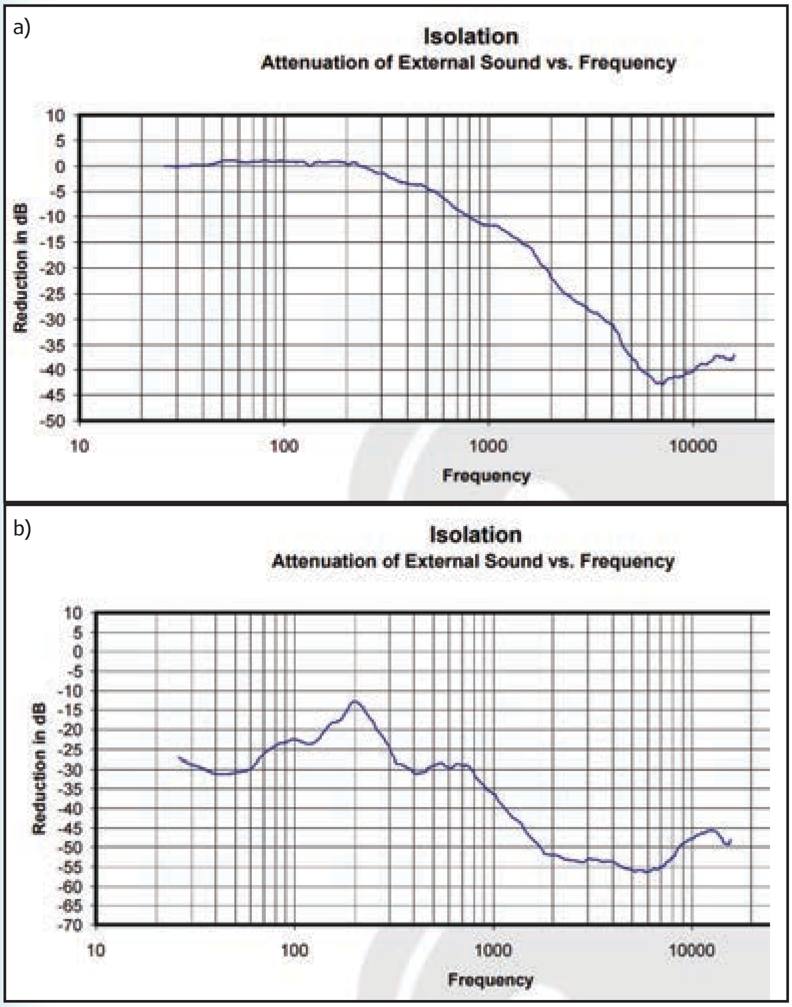


Figure 3: Attenuation graphs are displayed for Circumaural[2] (a) and Insert Earphones[3] (b).

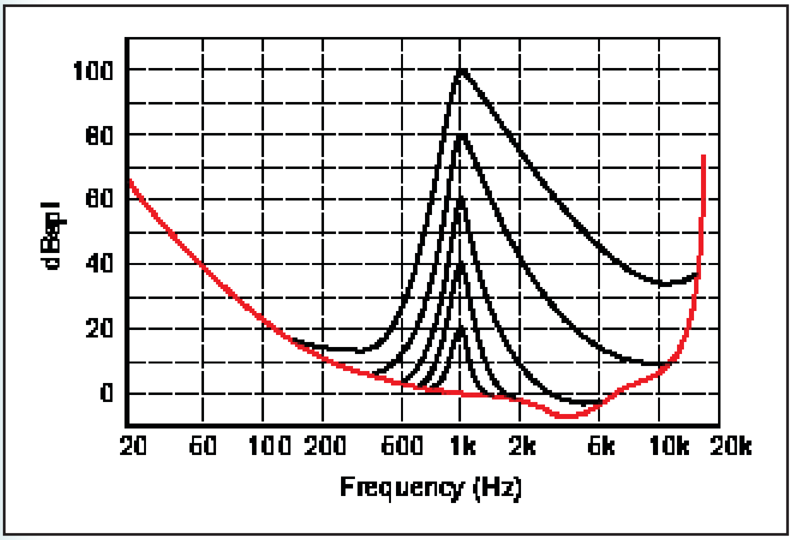


Figure 4: Masking for 1 kHz narrowband noise is shown at various sound pressure levels[4].

noise floor is from 20 Hz to 20 kHz, the better you can measure these extremely low voltages.

Figure 5 shows how the instrumentation noise floor can mask the presence of audible signals in the 2-to-5-kHz range. Two pieces of test equipment are compared: the Audio Precision APx515 analyzer and a high-end sound card representative of those often recommended as a data acquisition front-end.

The measurements were obtained using the Audio Precision APx555's analog generator, which has typical THD+N of –120 dB or better from 20 Hz to 20 kHz. A 1 mVRMS sine wave at 1 kHz was used as the stimulus for both the ADC of the APx515 and the sound card interface. A 1 mV level was chosen because it represents 72 dBSPL when using a 12.5 mV/Pa microphone. For both the APx515 and the sound card interface, 20 linearly-averaged 64k FFTs were used to generate the noise-floor spectrum.

The black curve represents the threshold of hearing. The red curve shows the noise floor of the sound card. The blue curve is that of the APx515. It can be clearly seen that when using the sound card interface, sound energy that is audible in the 2-to-5-kHz region may not be measurable. One can increase the FFT buffer size, which in turn will reduce the noise energy in each FFT bin, but this will increase test time.

Results

The ear—especially when little or no acoustic masking is present—is an extremely sensitive "microphone." Distortion artifacts, such as intermodulation distortion (IMD) and/or pulsating noises (e.g., GSM interference) may be audible even though the actual sound pressure is below 0 dBSPL.

Unless the test equipment (in particular the analyzer front-end) has a very low noise floor, there is a risk of not being able to objectively

About the Author

Dan Foley has been with Audio Precision since the middle of 2011. He is an experienced sales engineer with a background that includes 12 years as a sales engineer and US Manager of the Environmental Noise Group for Brüel & Kjær, six years as VP of Sales at Listen, Inc., and five years at Bose Corp. He is an expert in the field of acoustics and has extensive experience with audio-related standards committees, including those associated with the The Institute of Electrical and Electronics Engineers (IEEE), the Audio Engineering Society (AES), and Bluetooth Special Interest Group (SIG) organizations.

quantify (or develop a proper engineering solution for) distortion that is clearly audible in earphones.

References

[1] "Noise Criterion," The Engineering ToolBox, www.engineeringtoolbox.com/nc-noise-criterion-d_725.html.

[2] Audeze LCD-XC Headphone Measurements, Inner|Fidelity, www.innerfidelity.com/images/AudezeLCDXC.pdf.

[3] Shure SE535 Headphone Measurements, Inner|Fidelity, www.innerfidelity.com/images/ShureSE535.pdf.

[4] G. Martin, *Introduction to Sound Recording*, 2011, www.tonmeister.ca/main/textbook/intro_to_sound_recordingch6.html.

Sources

APx515 analyzer
Audio Precision | www.ap.com

Model 46AE 0.5" microphone
G.R.A.S. Sound & Vibration A/S | www.gras.dk

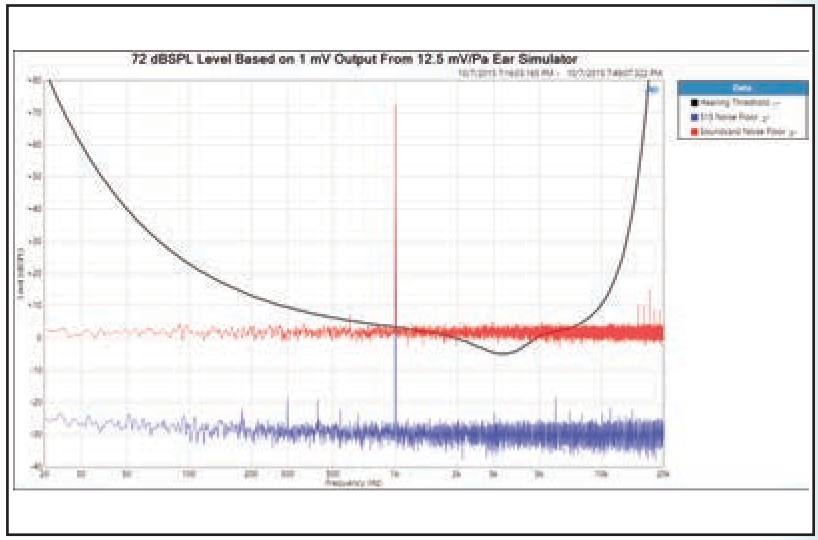
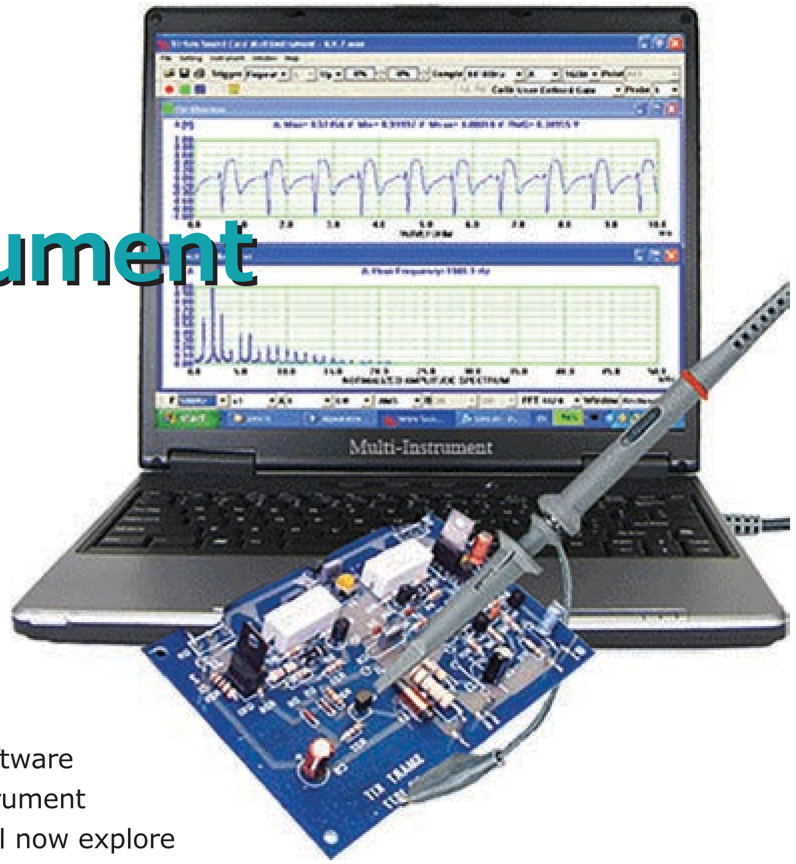


Figure 5: The APx515 noise floor is compared to the sound card noise floor.

The Virtins Multi-Instrument Software



In "Sound Cards for Data Acquisition in Audio Measurements (Part 3): Selecting the Right Hardware," (*audioXpress*, August 2015), I gave a quick overview of some of the software packages that enable you to use your sound card as a flexible data acquisition and measurement system. Among the software packages I mentioned was Multi-Instrument from Virtins Technology, which we will now explore in more detail.

By
Stuart Yaniger
(United States)

A recurring theme in my sound card article series (*audioXpress*, June 2015 to December 2015) was the software trade-off between versatility and feature set vs. complexity and learning curve. Inevitably, a versatile and powerful software package will be complex to use, and almost as inevitably, the user interface will be arcane and confusing. Simple and common features will be tucked away in unexpected places, the manual will have been written by a non-English speaking engineer, and tasks that should be smooth and rapid can take hours to get working the first time.

As time passes, the intrepid user will learn through repetition and hair-tearing and soon will be able to subconsciously operate the software, but the proficiency and familiarity will not be painless. Learning curves for complex software never resemble shallow ramps but always seem to be "Coyote and Road Runner" cliffs.

Among the software packages I mentioned was Multi-Instrument from Virtins Technology. It was complex. It was versatile. It was powerful. It drove me crazy. But in my usual prescient manner, I said, "I bet that a year from now, after I've gotten a greater comfort level, this will be my go-to choice for software." And, I confess that this has indeed come to pass.

The Software

Virtins (a portmanteau of "Virtual Instrument") is a Singapore-based company that sells a range of test and measurement software and hardware. Its principal products include USB-based digital storage oscilloscopes (DSOs), acoustic test and measurement systems, and vibration analysis systems. All of these products are comprised of a specialized hardware ADC/DAC module and a software system. The software system is also sold separately, suitable to be interfaced with a computer sound card, and is this article's focus.

On my axis of complexity vs. versatility, the Virtins software lands far on the upper right. The array of features and flexibility is mind-boggling, but trying to get a comfort level with routine use takes some work, time, and struggle. No short review could possibly be comprehensive, but I'll try to point out some useful features that you're not going to find in competitive products.

The focus of this review will be a few (!) of the features that distinguish this software from other packages. All of the "normal" functions (e.g., distortion analysis, distortion vs. frequency, distortion vs. level, frequency response, etc.) common to other software are present. So, in the interest of space, I'll glide by them relatively quickly, merely pointing out some extra capabilities within those functions that are useful and unusual.

The Software Organization

Virtins divides the test and measurement process into three layers: sensor, hardware, and software.

The sensor layer is the physical methods of acquiring a signal. It could be as simple as clip leads directly connecting a line level electrical signal to the sound card or something more complex such as a differential amplifier, a microphone and a preamp, or an accelerometer with signal conditioner. The output of this layer is a line level voltage.

The hardware layer is the data acquisition system, either a dedicated Virtins module, a National Instruments DAQmx card, or your sound card. Either way, there's an analog-to-digital converter (ADC), and in most cases, a digital-to-analog converter (DAC) as well.

The software layer controls the acquisition and the output, then processes and displays the acquired data. Within the software layer, the Multi-Instrument program is set up as independent instrument modules, which can run and be displayed in any combination. This is analogous to using voltmeters, oscilloscopes, frequency meters, phase meters, distortion analyzers, and spectrum analyzers as stand-alone instruments, but with a much lower cost and much less dedicated bench space.

My Complications Have Complications

We get our first taste of complexity when we whip out our credit card. Virtins offers the Multi-Instrument in various level packages, with different optional modules, at varied prices. There's six basic options, but the mixing and matching of package levels and modules is complex enough to require two lengthy tables on its website.

I'll try to simplify your life: My recommendation is to spend the $200 it takes to buy the Pro version, mostly because it is the only level that allows use of Audio Streaming Input Output (ASIO) drivers rather than Windows Multimedia Extension (MME). There will be features that you'll never use or don't care about (it reminds me of having to buy some special rims on a car to get an automatic transmission), but I guarantee that you'll find some uses you hadn't previously imagined you needed, but suddenly can't live without. Admittedly, this is more money than an extremely capable package such as ARTA costs, but it is considerably less than professional packages (e.g., Labview or MatLab).

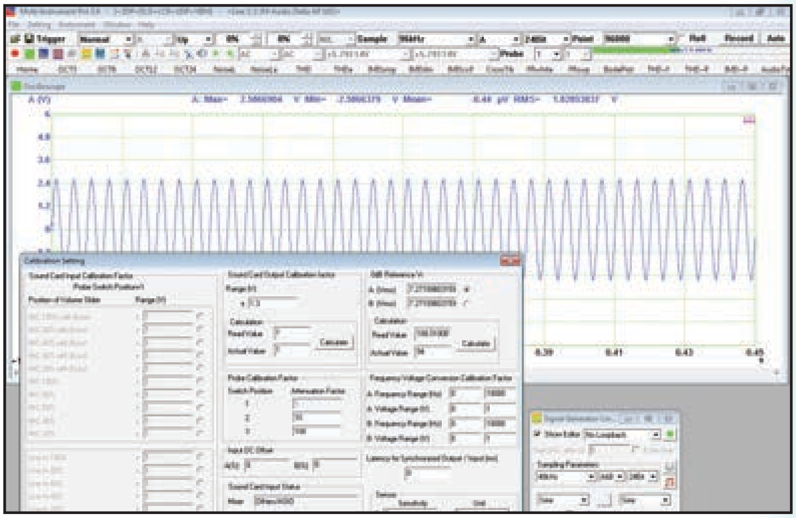


Figure 1: The Calibration window is used with an external voltmeter to calibrate input and output levels.

My Setup

As with most of the measurements in the sound card series, my test bed is an Hewlett-Packard desktop with an M-Audio Delta Audiophile 192 PCI sound card. I made up my own test leads, but Virtins sells various types as well.

Calibration

This step is actually pretty easy, but it will require some other calibrated meter (I used my trusty Fluke DVM) and a signal source (which can be the sound card's output with the Signal Generator module running). Open the Calibration window found in the Settings menu overlaying the Oscilloscope module screen (see **Figure 1**). Set the signal source to a voltage high enough to register as 80% to 90% on the sound level bars in the Oscilloscope module screen. The Oscilloscope screen will have a measured RMS voltage at the top of the display (shown as 1.8285 V in **Figure 1**). Enter this number in the Calculation box near the lower left corner of the Calibration screen.

Now, measure the actual value with your voltmeter and enter the measurement result in the Calculation box. Click "OK" and you're done. Use a test signal low enough in frequency that the AC reading on your voltmeter is accurate. If you're not sure, 100 Hz will be safe.

Signal Generator

The signal generator panel is small but busy (see **Figure 2**). Available signals include the usual sine, sawtooth, rectangle with variable duty cycle, white noise, pink noise, MLS, impulse, and step function. Additionally, the generator can provide a musical scale progression, as well as frequency sweeps for periodic signals. You can choose different signals independently for each channel, which can be useful in more complex measurements. It should be noted that the white noise and the pink noise are "true" continuous random signals rather than a limited random sequence, which is looped.

You can program your own custom signal as well, most simply with the Multitone option. In Multitone mode, you can specify various signals and add them together, outputting the sum into a selected output channel. For example, if you want to see the effect of adding low-level pink noise to a 1 kHz tone, or create a comb of different frequency, amplitude, and wave shapes, it's a simple matter of choosing the component tone parameters (wave shape, amplitude, and phase) and adding it to the desired channel via a user interface, which pops up when you select the Multitone option on the signal generator panel (see **Figure 3**). It's done iteratively. Choose the parameters of the signal's component on the panel's left side and hit the arrow in the middle to determine to which channel it's sent.

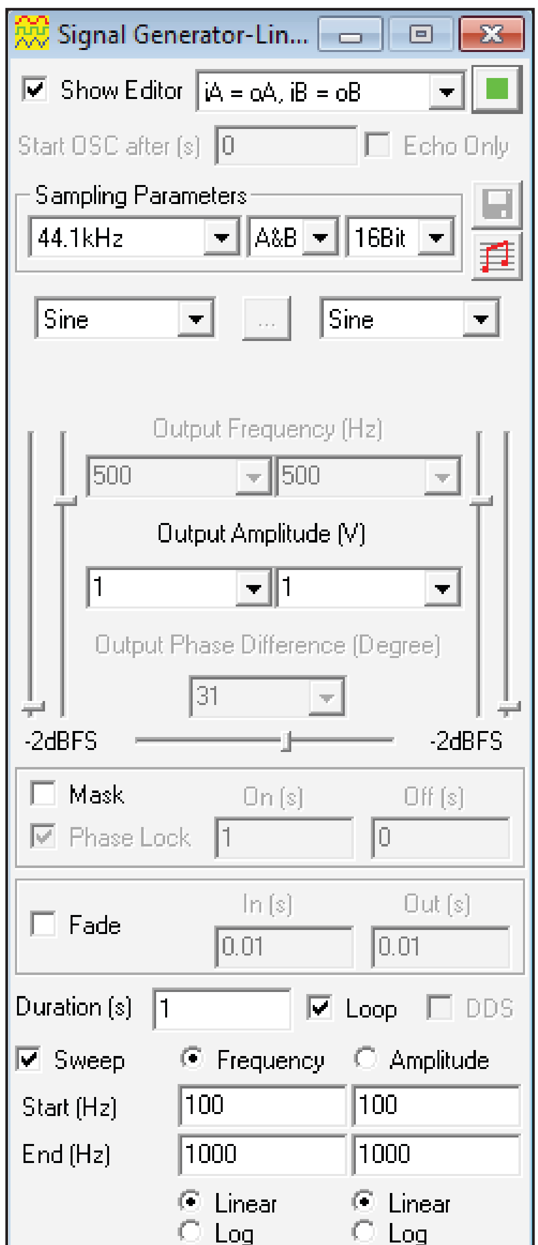


Figure 2: The Signal Generator window allows control of the sound card outputs with standard or custom waveforms.

One interesting feature in the panel is Mask. The Mask will zero out selected portions of the test waveform, selectable by the On and Off time parameters (e.g., when creating tone bursts). Let's say you want 10 cycles of a 1 kHz burst repeated one per second. Set the record length for 1 second, then set the mask On to be 0.01 second, Off to be 0.99 seconds. You now have your tone burst.

The generator can also store, then read and output selected wav files. This can be handy to put together mixed waveforms (e.g., a sine tone burst followed by an impulse) or for using repetitive sections of a music selection as a test signal. Included in the package are a couple dozen standard test signals (e.g., various multi-tone signals for performing different intermodulation standard measurements).

Another neat feature is the option for "No Spectral Leakage." If you click this option, the frequency is automatically adjusted to be centered in the nearest frequency bin, reducing the "skirts" at the base of the fundamental peak, which are a mathematical artifact of the sampling and transform process. This can be useful if you're trying to resolve small peaks at frequencies close to the fundamental (e.g., power supply modulation).

The signal generator panel is also where a non-related but useful set of options are hidden—a virtual patch panel for signals. This appears at the top of the generator screen. Besides the expected options (e.g., sending signals to different combinations of channels, stereo, and mono), you also have the option of completely bypassing the sound card and doing a loopback in software at the Windows mixer level. That can be a useful way of learning how to use some of the Multi-Instrument features, and as Virtins points out, it can also be used for learning about various aspects of digital signal processing (DSP).

Multimeter

The most basic function in Multi-Instrument is called a Multimeter. It can measure frequency, voltage, level (in dBU, dBV, dB FS, and A, B, or C weighted), duty cycle, and counts. **Figure 4** shows the Multimeter window and selectable options embedded in an oscilloscope window. In that screenshot, the Multimeter is reading the frequency of an applied square wave. I clicked on the menu pulldown to show the various other configurations available.

Derived Data Point Meter

This feature is basically the Multimeter on

About the Author

Stuart Yaniger has been designing and building audio equipment for nearly half a century, and currently works as a technical director for a large industrial company. His professional research interests have spanned theoretical physics, electronics, chemistry, spectroscopy, aerospace, biology, and sensory science. One day, he will figure out what he would like to be when he grows up.

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Celebrating
33 YEARS
1983-2016
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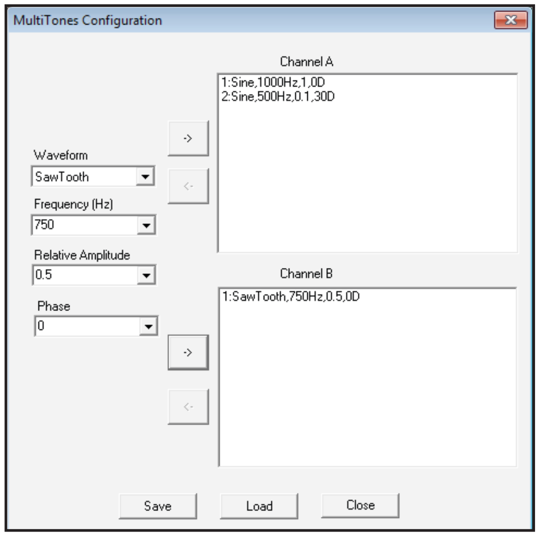

Figure 3: The Multitone window is used to build up complex signals from selected components.

Figure 4: The Multimeter is displayed overlaying the Oscilloscope screen, in this case showing the frequency of the square wave. The menu for other Multimeter options is also shown.

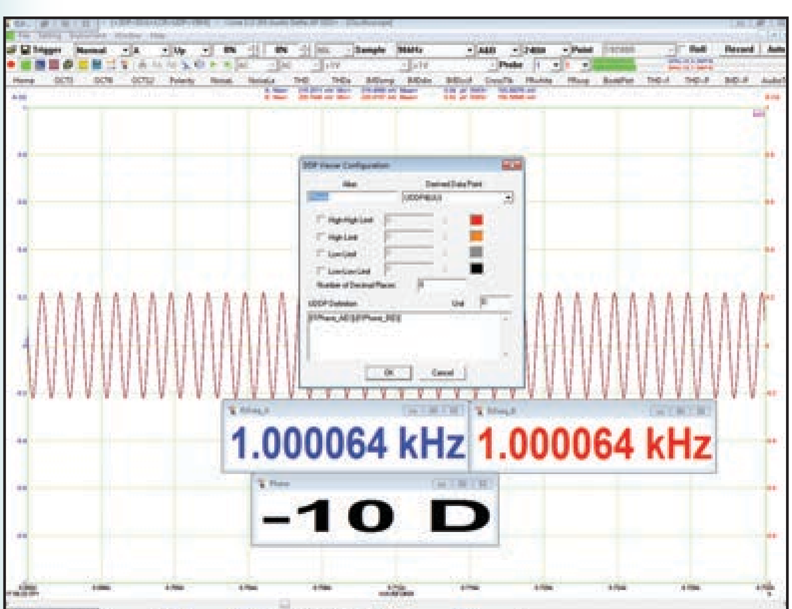

Figure 5: The Derived Data Point (DDP) option provides an easily visible display of a user-defined parameter, in this case the phase shift between two sine waves.

steroids. Besides reading and displaying the same sort of variables as the Multimeter, the Derived Data Point (DDP) meter adds in a few more—actually about 150 more! The displayed variables are updated with each acquired frame, so they are essentially real time. Up to 16 different DDP measurements can be simultaneously displayed.

The derived part of the name refers to the ability to set up windows that can calculate custom measurements using a relatively simple syntax. **Figure 5** shows an example.

I input a sine wave with a 10° phase shift between Channel A and Channel B (shown on the Oscilloscope screen). Using DDP, I chose two phase measurements, one for each channel. The phase shift is then the difference in phase between the two channels, so this calculation was entered into the User Defined Data Pont (UDDP) dialog box, the output assigned an alias (Phase), and the result displayed as a phase measurement. The user can define up to six different UDDPs.

Another DDP feature is its ability to set high and low limits giving visual warnings if a signal wanders outside of a user-specified range. This could be useful if the Multi-Instrument is used to monitor a process. It can also be helpful in troubleshooting situations, where you want to be warned if an over-voltage or over-current is being approached.

Oscilloscope

The oscilloscope module is pretty much what you'd think—a Digital Storage Oscilloscope (DSO), with bandwidth limited by the sound card's sampling rate and input range. (Translation: It is not a substitute for a "real" oscilloscope.) As usual with DSOs, the sweep can be set as automatic or triggered, with level, slope, and delay all adjustable. The screen can display the usual Channel A, Channel B, and both simultaneously. It can also be set to display the sum, difference, and the product of the two channels. Curiously, there's no option for ratio, but I'll bet this shows up as a feature at some future time.

The oscilloscope module can record the output into a file on a continuous basis, with a 2 GB size limit. I haven't tried using this feature as a digital recorder yet, but that will be coming soon! Where this capability differs from other systems is that the oscilloscope output can also be filtered using user-specified finite impulse response (FIR), infinite impulse response (IIR), or running Fast Fourier Transform (FFT). Built-in DSP includes low pass, high pass, band pass, and band stop filters.

A simple and obvious feature that seems to be missing in other software packages is X-Y plotting, that is, a plot of the magnitudes of the A channel vs. the B channel during signal excitation. In the old analog oscilloscope days, phase and frequency measurements were often made using Lissajous figures (see Resources). Real-time Lissajous capability is selectable in the display options on the bottom of the screen (denoted as A|B).

Figure 6 shows a screenshot of the Lissajous pattern of 300 Hz and 500 Hz sine waves applied to Channel A and B, respectively. The trace can be used to determine frequency ratios and relative phase, as well as phase relationships between stereo channels. If a musical selection is used as a test signal, with Left fed to Channel A and Right fed to Channel B, a mono signal will result in a diagonal line. Stereo information will appear as a "fuzzing" of the line.

I was fascinated by this feature, but a thought crossed my mind. Since we have a continuous display (or can do a one-shot display with infinite persistence), the Lissajous feature could be useful in looking at transfer functions (i.e., input vs. output).

To test my idea, I rigged up a simple undegenerated common-source field-effect transistor (FET) stage with a gain of 3, then applied a 500 Hz sine wave to it. The A channel was connected to the applied signal, the B channel was connected to the FET drain (output).

Figure 7 shows the resulting trace. You can clearly see the square-law transfer function of the FET. This could be a nice way of visualizing nonlinearity in real time and possibly be the core of a curve tracer, with the addition of some fairly simple external electronics (sound cards cannot generally output DC). With a transducer attached to a speaker cone and the generator set to give an amplitude or frequency ramp, the function could also allow you to visualize nonlinearity of the speaker with increasing drive.

Spectrum Analyzer

As with most other measurement software packages, Multi-Instrument enables real-time amplitude and phase spectra to be calculated from the time domain signal using a FFT. There are a few enhancements, though. First, the number of window functions is vastly greater than in other measurement packages (55, count 'em). If none of those suits your application, you can create your own. These are read into Multi-Instrument as wav files.

Another bit of extra flexibility is the FFT size. Multi-Instrument has the capability of a maximum of 4,194,304 data points per frame, which allows extremely fine frequency resolution, the trade off being acquisition and computation time.

There are pre-packaged test plans to acquire spectra using various total harmonic distortion (THD) and intermodulation distortion (IMD) test signals.

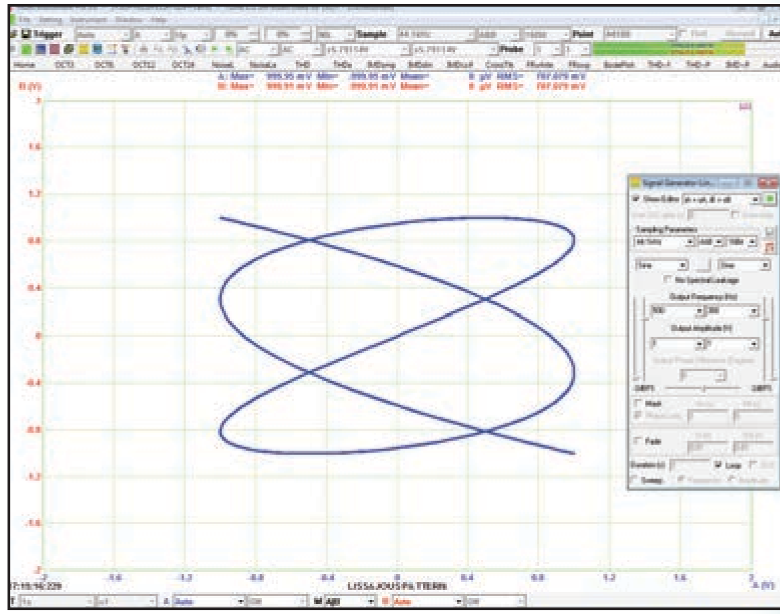

Figure 6: The Lissajous pattern of a 500 Hz and 300 Hz sine wave is shown. It was acquired using the A|B option on the bottom of the screen.

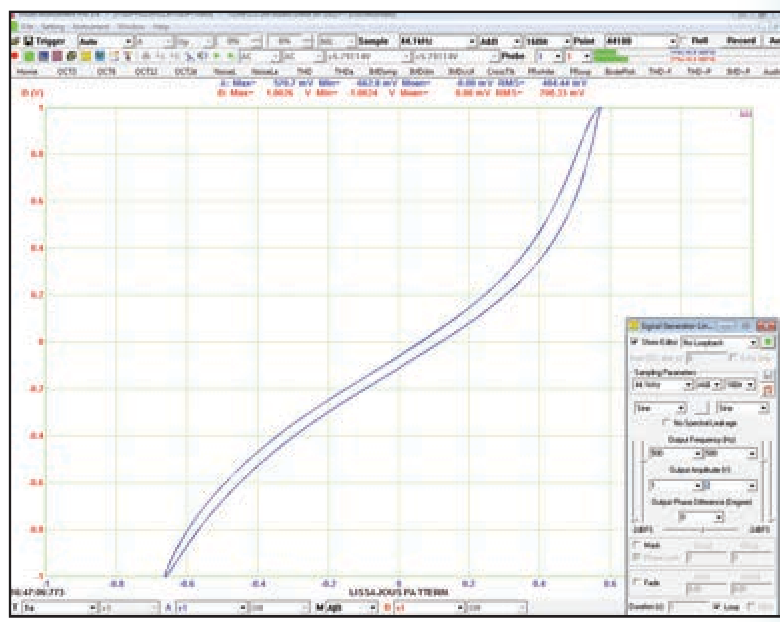

Figure 7: The Lissajous option can be used to derive transfer functions. In this example, the quadratic transfer function of a FET amplifier is shown.

Resources

"Lissajous Curves," *Wikipedia*,
https://en.wikipedia.org/wiki/Lissajous_curve.

"Measuring Phase Shift," YouTube,
https://www.youtube.com/watch?v=L78tdU9wPVc.

"Overview of Multi-Instrument," YouTube,
www.youtube.com/watch?v=msx5envAgGw.

"Using Cross-Correlation to Measure the Speed of Sound," YouTube,
www.youtube.com/watch?v=7u1nSD0RlwY.

"Using Cross-Correlation to Measure Time Delay," YouTube,
www.youtube.com/watch?v=L6YJqhbsuFY.

There's also selectable options to acquire THD or IMD vs. power and THD vs. frequency, which are basic measurements for power amplifier characterization. **Figure 8** shows an example of an IM measurement with 19 and 20 kHz tones. Note that besides the spectrum display, there are DDP boxes to give you the raw numbers, which is a handy feature in noise measurement. Generally, one has to calculate noise from the baseline using a summation of noise voltages in bins. This can be a tedious calculation if the noise floor isn't flat. Multi-Instrument does this calculation with a menu option (NoiseL) and displays the result as a dotted line overlaying the noise floor (see **Figure 9**).

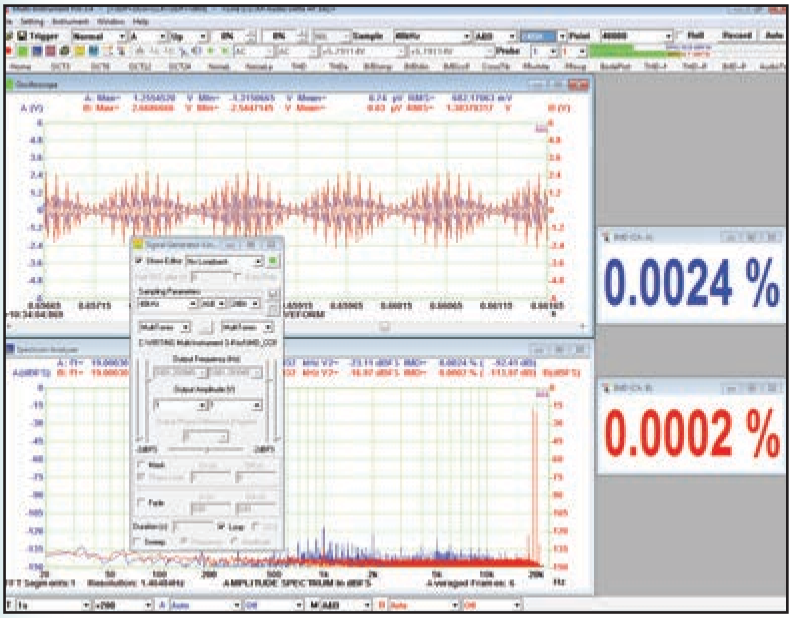

Figure 8: The distortion measurement options on the upper menu bar select test signals and display both spectra and numerical results.

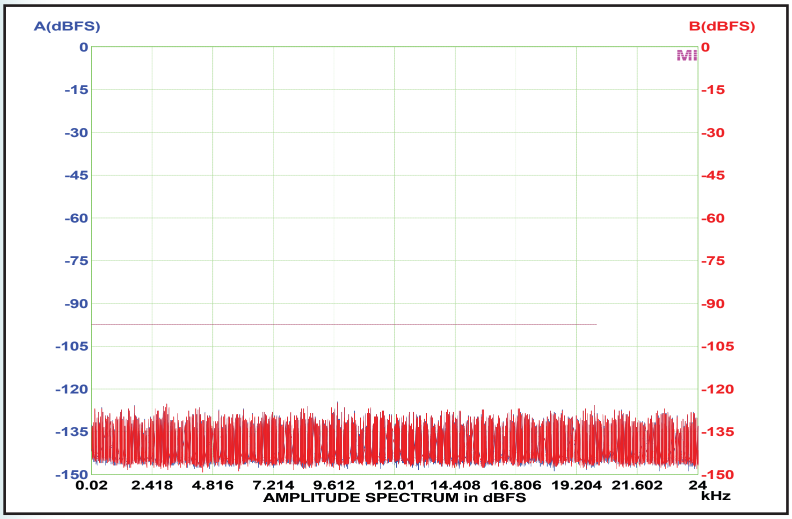

Figure 9: Noise measurements can be acquired without further calculation. The noise level is indicated by the dotted line (in this example, at about –95 dB).

Other functions include frequency response (measured by either a swept tone or with white noise), cross-talk, and transfer function (acquired as a Bode plot or Gain and Phase plot).

A useful and uncommon feature is the ability to measure correlation. The correlation between two signals is how closely they relate to one another as one of the functions is shifted through all possible values of time. If the two signals are applied to Channels A and B, the correlation R is defined as:

$$R(\tau) = \int_{-\infty}^{+\infty} A(t)\,B(t+\tau)\,\mathrm{d}t$$

If Channels A and B are presented with the same signal, R is called autocorrelation. If they are different, R is called cross-correlation. The most useful thing about correlation functions is that they can be used to determine time delays between a stimulus and a response. In a simple case, an impulse is fed to a linear system and the output is measured. The delay is easily determined by comparing the timing of the output and the input signals, which can be done by eye.

But what if the stimulus is a segment of white noise? Or the output signal is distorted, processed, or mixed with noise? Eyeballing is problematic. This is where the cross-correlation function can come in handy. It is evident that as the time delay between the signals at A and B approaches zero, the correlation will reach a maximum (see **Figure 10**). The specific example is a tone burst, but it could just as easily be a pseudorandom signal or an impulse buried in noise.

As a fun exercise (and a learning experience), Virtins has an instructional video showing how to use this function to measure the speed of sound, using white noise (see Resources).

Device Test Plan

If the user wants to configure Multi-Instrument to automatically run a series of tests, the Device Test Plan function can be used to program the desired test signals, the measurements, and the Pass/Fail criteria. The capability here is high enough that an entire article could be written on using this feature to create a virtual Automatic Test Equipment function. Multi-Instrument can also be used to control external test equipment using standard protocols such as ActiveX and Virtins's proprietary application program interfaces (APIs). This will primarily be used in professional quality assurance operations, but it could also come in handy for the amateur who wants to match or sort transistors or FETs for multiple variables (e.g., I_{dss} and g_m).

The software includes about 100 pre-written Device Test Plans that cover a range of common measurements. These can be run individually or as a specified series.

Summing Up

Multi-Instrument is a stunningly versatile and capable software package. It is far more complex and difficult to use than most hobbyist-oriented software (e.g., ARTA or AudioTester) and costs significantly more. But the price is reasonable compared to professional packages, and the capabilities far exceed the simpler software. The simpler and less expensive software will likely take care of 80% of your needs, but once you have climbed the learning curve for Multi-Instrument, your measurement power will be an order of magnitude greater.

I will confess that writing this review was delayed again and again as I'd discover a new feature or a new way to use existing ones. I still think that I haven't used more than 5% of Multi-Instrument's capability. My procrastination was aided by an excellent set of application videos that Virtins has produced for its YouTube channel.

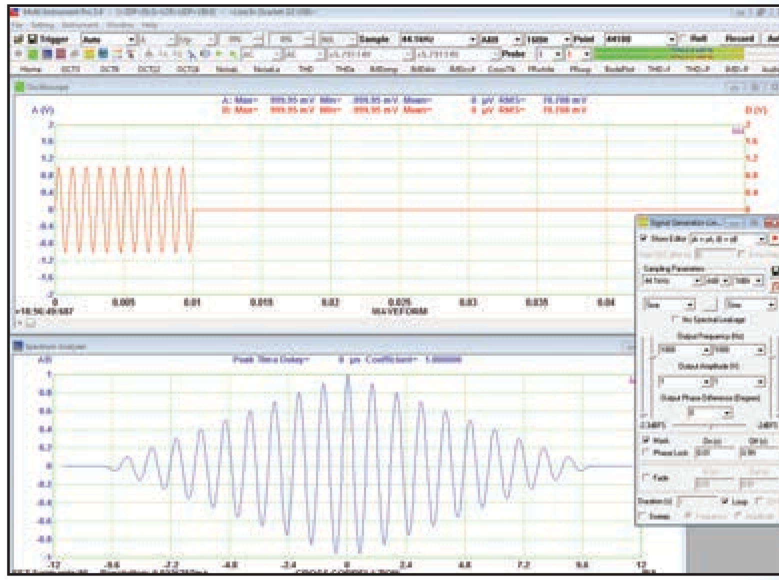

Figure 10: Correlation functions show how waveforms (in this case, a tone burst) relate as a function of time shift of one of the waveforms. Note that the maximum value for this example is at a zero time shift.

The videos explain basic concepts and walk you through science-fair-level demos (e.g., using cross-correlation functions to measure the speed of sound), which are good ways of learning basic measurement methods. They are also useful for discovering more of the software capabilities than are first evident. The production values of the videos are good, but the "Mr. Roboto" synthetic voice-overs are rather jarring. Overall, this software is highly recommended.

Questions & Answers

NTi Defines New Technologies in Audio

An Interview with Philipp Schwizer, CEO of NTi Audio

By
Shannon Becker
(United States)

Philipp Schwizer is the driving force behind NTi Audio's continued growth in the test and measurement audio industry.

Editor's Note: Our March issue has a Test & Measurement focus so we decided to learn more about one of the leading manufacturers of test and measurement solutions for acoustics, audio and vibration applications. With headquarters in Schaan, Liechtenstein, the company specializes in fast end-of-line audio testing for manufacturing quality control purposes, provides solutions for testing public address systems in safety-critical environments, and also produces handheld audio analyzers and generators aimed at the professional audio industry.

SHANNON BECKER: NTi stands for New Technologies in Audio. Tell us about the company.

PHILIPP SCHWIZER: NTi Audio was born in the year 2000 through a management buy-out from the connector company Neutrik. The then Test & Measurement division manager and I had a clear plan of how to further grow that business in exciting directions. As you can imagine the sales approach and sales channels of complex technology is quite different from selling connectors. That led to the decision to separate these two business divisions. I bought the company with some partners. At that time we had phenomenal success with the Rapid-Test platform for the mobile phone industry. It was the most dedicated and fastest transducer tester you could find on the market. In parallel, we had the first generation of handheld signal generators and analyzers in the queue.

These two major product groups still drive the company. On one end we serve the audio industry with powerful hand-held instruments, offering excellent technical specifications, paired with versatile functionality but still easy to operate. The other product family serves industrial quality control applications where the integration of acoustic and audio tests with measurement speed into an automated production line is key. We work worldwide with partner companies, subsidiaries, and more than 55 independent distribution partners to provide both technical, support, and sales expertise covering the entire world.

SHANNON: Tell us a little about your background and how you became involved with NTi Audio (www.nti-audio.com).

PHILIPP: I'm a Swiss citizen living in the tiny country of Liechtenstein where our company is also located. I was already passionate about audio when I was a teenager—building my own speaker systems and amplifiers and repairing audio equipment for my colleagues. I obtained my engineering degree in Electronics and signal processing in 1986 and joined the audio world in 1989 at Neutrik. While there, I developed audio processors for a while but soon changed to product management and later to marketing and sales.

Before the buyout, I was Director of Sales and Marketing at Neutrik. My heart was yearning to optimize the sales structure for the Test & Measurement products. That desire led to the birth of NTi Audio.

SHANNON: What is your role within the company?

PHILIPP: I'm the CEO of the company and also I head up the sales & marketing processes. I take care of finance and play a part in all the strategic product planning processes. But our success is by no means a one-man-show. I have excellent product management teams, technology architects, and a group of very senior audio design engineers around me.

SHANNON: What are some of the challenges that NTi Audio faces in this highly competitive audio market?

PHILIPP: The biggest challenges are the price/value propositions and the technology trends. The price/value proposition hits us because decent technology carries a price if you want to achieve high specifications. At the same time, there are gadgets and apps around that look very similar and seem to do the job to a certain extent. Thus, a fair percentage of particularly the younger clients believe and expect that our products should be available for a couple of dollars. We have to educate these clients by explaining the difference and why quality is necessary.

SHANNON: Tell us what makes NTi Audio different.

PHILIPP: NTi Audio's products cover the client's needs! We listen very carefully to the market and provide solutions that do the job. On top of that, we present somewhat complex technology in as simple a way as possible and carefully optimize the user interface so that the operator can intuitively

The XL2 Analyzer offers a unique combination of a state-of-the-art Sound Level Meter, a comprehensive Acoustic Analyzer, and a powerful Audio Analyzer.

control our instruments. The art here is to reduce the number of parameters, while at the same time providing full functionality, rather than trying to impress with a long list of settings that are hard to understand.

SHANNON: How has the company and its products evolved to keep up with rapidly changing technology?

PHILIPP: All our engineers are technology scouts and we encourage them to play with technology and be curious. We also maintain a valuable network with universities, technology companies, and other talents in the industry.

SHANNON: Tell us about NTi Audio's products.

PHILIPP: The NTi Audio Product range can be broken down into two main categories. The first is the hand-held instruments family covering signal generators, Acoustic and Audio Analyzers for analog and for digital audio. The signal generators MR-PRO and the Audio & Acoustics Analyzer XL2 are the two most prominent members of this EXEL group. The second product category is the modular FLEXUS family with production analyzers for quality control on a production line of audio equipment (e.g., amplifiers, mixers, processors etc.), and acoustical components (e.g., transducers, microphones, and speakers). The systems can be scaled from a basic two-channel analyzer to a large multi-channel analyzer or a speaker tester with all the components of a power amplifier, impedance shunt and signal conditioning on board so that the only external wiring is the cable to the Device Under Test and the microphone inputs.

SHANNON: What do you feel is NTi Audio's most successful product to date and why do you think it is successful?

PHILIPP: That is clearly the XL2 Audio & Acoustic Analyzer. A key factor for its success is the wide application functionality that protects your initial investment and gives you the best value for the money invested.

With numerous firmware updates that are all free for the users, we have proven that value of the instruments even increases over time. Also the functionality of the XL2 can be extended at any time with no hardware changes by simply adding firmware options (e.g., for measuring speech intelligibility, remote control, pass/fail testing, and cinema calibrations).

SHANNON: What's next for NTi Audio?

PHILIPP: That's a very good question yet very hard to answer. We have several ongoing projects but only a few I'm able to talk about. Next to come will be the network integration of the XL2 Analyzer. There is an increasing demand for placing the analyzer at remote locations and being able to review and access the measurement data from a distant device.

The upcoming NetBox product will enable the XL2 to deploy its measurement data to a cloud server and the NoiseScout portal allows any browser-enabled viewing device to access and view the data without interruption to the measurement process. That is a very interesting technology that carries a lot of potential also for future products.

The Flexus FX 100 Analyzer is a unique professional audio analyzer that adapts to any application. It also offers the simplest system integration for the fastest testing of audio and acoustic devices, providing comprehensive measurements including Passed/Failed tests in under 1 second.

TEST & MEASUREMENT FOCUS

You Can DIY!

Build an Audio Headroom Meter

Photo 1: The Audio Headroom Meter is designed to set microphone preamp gain.

How do you determine the headroom requirements for an analog audio signal? The answer depends in large part on how the audio signal level is defined and metered. Meters may have absolute peak, quasi-peak, or average responses to program material. Each type presents a specific technical or artistic advantage, and all see common use in audio recording and transmission systems. But, this article isn't all about metering. Rather, it introduces a simple and inexpensive visual presentation of headroom utilization.

By
James B. Wood
(United States)

Most of our audio entertainment arrives as recordings or broadcasts. These are delivered with predefined levels. We don't have to worry about a downloaded MP3 or a CD album exceeding "Zero Decibel Full Scale" (0 dBFS), or an FM station blasting much beyond 100% modulation (±75 kHz carrier deviation).

Those involved in recording or broadcasting live performances, on the other hand, have to cope with a virtually unlimited dynamic range. A given microphone may have an output of microvolts for a whisper, or perhaps as much as a volt when it's an inch off a drum head.

This Headroom Meter project (see **Photo 1**) is a spin-off from a microphone preamplifier (mic-pre) design that needed some sort of "sweet spot" indicator for trimming its gain under different live-sound pickup conditions.

Recording studio consoles or mixing desks generally have a "CLIP" indicator LED near the top of each channel strip. A gain-trim pot near this LED helps the session engineer independently adjust input stage sensitivity for each input assignment. During set up for a session, these pots will be set to keep each preamp out of clipping for the specific instrument or vocal assigned to that input channel.

What that single LED fails to show is how well the input stage's dynamic range is utilized. It would be simple to reduce input stage gain by a generous margin to avoid clipping under any circumstance. But as amplification is increased downstream to make up for low levels, noise from those input stages could become significant.

Although the main objective of this project was finding a better means of setting microphone preamp gain, the circuit described here is useful for keeping an eye on any analog signal path.

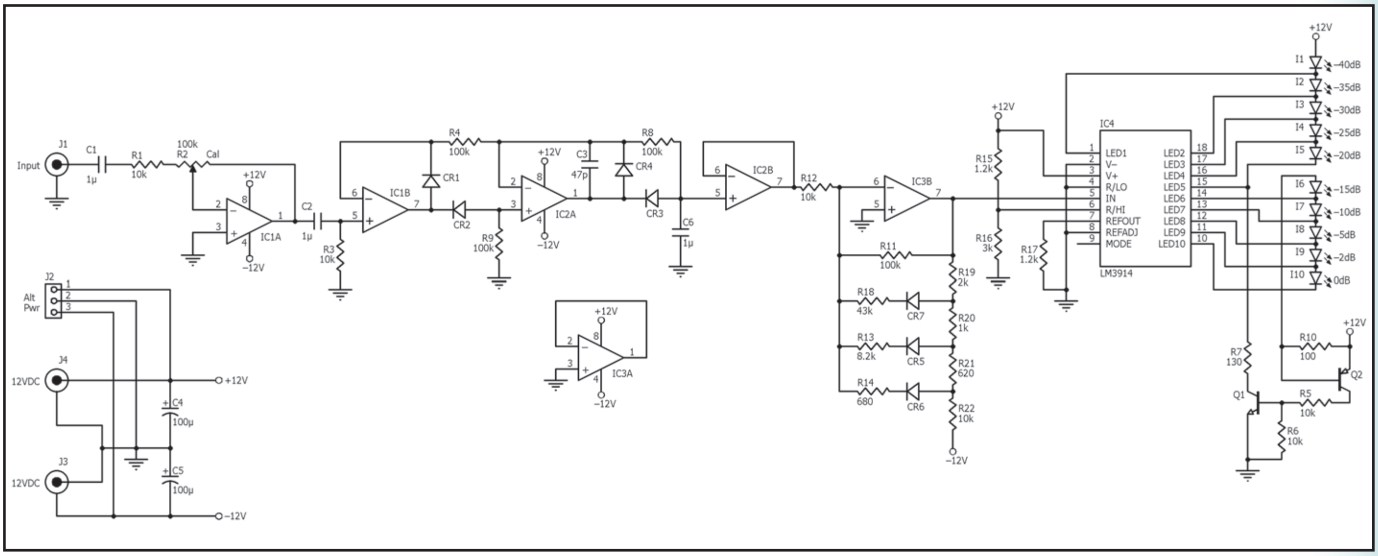

Figure 1: Here is the schematic for the Headroom Meter.

Choosing Analog or Digital

Someone with code-writing smarts could easily bang out a metering circuit using a PIC (elementary microcontroller) and a serially addressed display driver chip. But I decided to to keep things analog and simple enough for someone with no programming skills to construct the circuit from readily available parts.

The display is a string of 10 LEDs. The top of the scale, marked 0 dB, represents the clipping point of the audio path. Clipping may pertain to an amplifier, an ADC, or whatever else might impose a constraint on the absolute peak ceiling level of the input signal. The second LED is at −2 dB, the third at −5 dB, and the rest at 5 dB intervals down to −40 dB.

The Input Stage

IC1A is a stage with adjustable gain-to-scale measurement sensitivity to the clipping point of the signal to be metered (see **Figure 1**). CAL potentiometer R2 is adjusted to match the 0 dB LED trigger level to a clipping point anywhere between a little less than 1 V up to tens of volts of peak amplitude.

"Precision" Rectification

Circuits for LED level meters have appeared in print and on the Internet for years, but a surprising number of these have ignored the importance of full-wave rectification of the audio program signal. Sine, square, and triangle waves are often used as test signals in audio applications, but these electronically generated tones are inherently symmetrical. In the real world of speech and music, waveforms may be grossly asymmetrical (see **Figure 2**).

The data shown in **Figure 2** graphs a person singing "Ooooo" into a microphone. Note that the peak excursions below the baseline are nearly twice the amplitude of the positive-going peaks.

Depending on how the microphone happens to be phased, a meter with a single-diode, half-wave rectifier that captures only the positive peaks would register 6 dB low for this speech sample. Negative peaks might take an amplifier into clipping in this example, while the meter would display a level comfortably below 0 dB.

Rectification of both negative-going and positive-going voltage excursions is essential. A full-wave rectifier has an inherent logical OR function that yields an output based on the higher

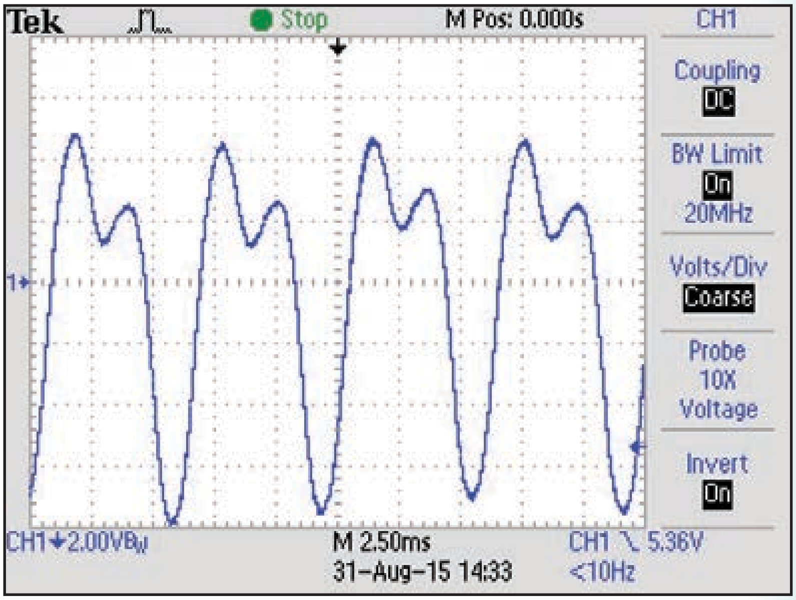

Figure 2: This typical speech waveform measures a person singing "Oooo" into a microphone.

of the positive or negative value, not their sum. A single diode, or even two in a voltage-doubler configuration, just won't do.

What's more, a typical silicon diode just begins to conduct at about 0.4 V. Associating this sloppy and temperature-sensitive threshold level with –40 dB on the meter scale means that program peaks would need to reach 40 V to indicate 0 dB at full scale.

In our circuit, IC1B and IC2A form a full-wave "precision" peak-detecting rectifier. What makes this rectifier precise is that all diodes are within the feedback loop of one stage or the other. This entirely negates the turn-on voltage of the diodes, making them near-perfect rectifiers. A rectifier of this type is quite linear over a range of at least 60 dB.

This circuit differs from the traditional precision rectifiers depicted in many textbooks and app notes, which are generally averaging rectifiers. The somewhat obscure circuit used here requires some careful analysis to completely understand it, but it does give an output based on the input zero-to-peak values regardless of polarity.

The rectifier generates a negative DC output for reasons that become clear later and has unity gain. A ±1 V peak input (2 V_{PP}) gives a –1 VDC output.

LED	VDC from Rectifier	VDC to LM3914
0 dB	8	8
–2 dB	6.35	7.2
–5 dB	4.5	6.4
–10 db	2.53	5.6
–15 dB	1.42	4.8
–20 dB	0.8	4
–25 dB	0.45	3.2
–30 dB	0.25	2.4
–35 dB	0.14	1.6
–40 dB	0.08	0.8

Table 1: This is the DC output from the precision rectifier for each of the 10 metering steps.

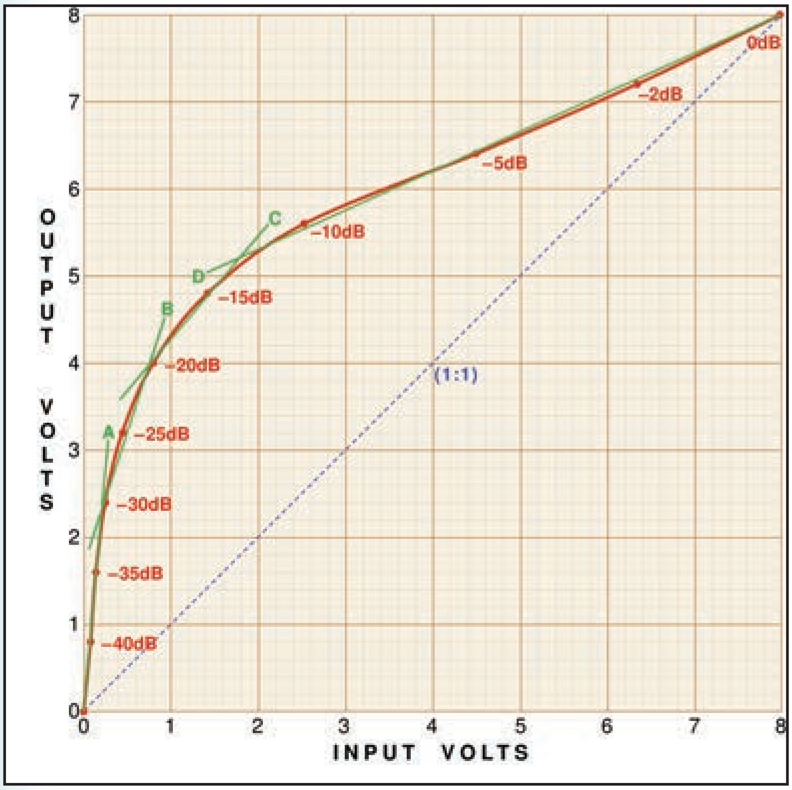

Figure 3: The transfer function graphically plots the 10 metering steps shown in Table 1.

Quasi-Log Transfer Function

The Headroom Meter uses the popular LM3914 display driver chip to light the 10 LEDs. Since the LM3914 lights the LEDs as a linear function of input voltage, and our precision rectifier has a similar voltage-linear in/out relationship, we need some form of level translator to give us a meter display that is decibel linear. This is afforded by the "quasi-log" converter stage, IC3B.

Table 1 shows the DC output from the precision rectifier for each of the 10 metering steps. Next to these figures is the DC level that the LM3914 must receive to light the associated LED. The 8-V trigger point for the full-scale 0 dB LED at the top is an arbitrary value chosen to ensure sufficient headroom within the metering circuit itself.

The two voltage columns have a logarithmic relationship between –40 dB and –5 dB. Expanding the scale a bit near the top gives better resolution near the clipping point. **Figure 3** graphically plots this relationship. The red line describes the transfer function with red dots marking each metering step.

This transfer function was initially hand-plotted and then "segmented" to approximate the curve with the four straight green lines A, B, C, and D. This approximation evolved from several experimental plots, along with an assessment of the deviation error in each case. For a meter limited to 10 display values, these four segments proved a good fit.

The intersections of the segments define break points for the gain of the quasi-log amplifier. The slope of the line represents amplifier gain over the span of that segment.

IC3B is the "quasi-logging" stage, an inverting configuration that delivers a positive voltage to IC4, the LM3914, from the negative-going rectifier output. With no input, the output of IC3B rests at 0 V (ground). The three break-point diodes—CR7, CR5, and CR6—are back-biased by current from the negative supply through R22, essentially removing the diodes from the circuit. The gain of this stage is entirely determined by the ratio between R12 and R11, a gain of 10, which matches segment A.

As the rectifier output goes more negative, the output of IC3B is driven increasingly positive. When the output of IC3 reaches about +2.4 V, CR7 begins to conduct. This places R18 in parallel with R11 and reduces gain of the stage to match the slope of segment B. At +4 V, CR5 brings R13 into the picture and gain follows segment C. CR6 switches in R14 at +5.2 V and gain follows segment D the rest of the way up.

Errors in this conversion process are greatest where the curve is the steepest, at the bottom of the scale. The −40 dB will be spot-on, but the next two steps might be as much as 1.5 dB off. This should not pose a problem at the bottom of the meter scale,

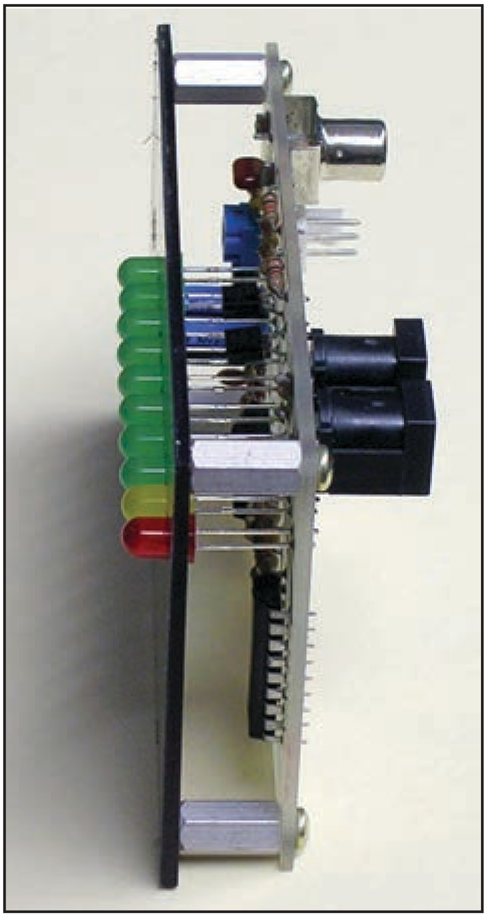

Photo 2: All the circuitry is contained on a small board that forms a "sandwich" with the front panel.

and accuracy is quite good from –20 dB up.

The Display Driver

The venerable LM3914 is a handy part for this type of project. It has an internal precision voltage divider string that determines accurate turn-on points for the 10 display LEDs. Ten individual voltage comparators monitor the input and deliver a constant current to each LED as it turns on. Very few outboard parts are required.

In this circuit, the bottom of the internal divider string is grounded. The top is fed from the junction of R15 and R16, a point resting at +8 V. This means the LEDs progressively turn on for each 0.8 V incremental increase of the input to the LM3914. These are the levels noted in right-hand column of **Table 1**. R17 establishes the constant brightness current through any and all LEDs, about 10 mA in this case.

The LM3914 is able to operate in either a "bar" or a "dot" mode, depending on whether pin 9 is grounded or left floating. In the dot mode, only one LED is ever illuminated at any time. The bar mode gives a more traditional "thermometer" level display, with LEDs sequentially lighting in a string from the bottom up. The downside to the bar mode is that each LED draws 10 mA, imposing a 0.1 A load on the supply when the last LED finally illuminates.

In its original 1980s app note for the LM3914, National Semiconductor provided a clever power-saving trick. Instead of grounding pin 9 for the bar mode, the pin is left floating, as if for the dot mode, but the LEDs are wired in series rather than directly taking each to the positive supply rail. This gives a bar display by default, and with just a 10 mA total supply load, even when all 10 LEDs are illuminated. However, this does require a power supply of more than 20 V to overcome the voltage drop of 10 LEDs in series.

The Headroom Meter uses a bipolar 12 V supply, so only +12 V is available to light LEDs. But we can still use the power-saving trick by breaking the series-connected LEDs into two strings.

From –40 dB to –15 dB, LEDs I1 through I5 constitute a first series string and operate in National Instrument's clever power-saving mode, drawing a constant 10 mA directly from the +12 V supply. In the second string, the anode of I6 does not go directly to the +12 V rail. Instead, part of the current for this string is drawn through the base/emitter junction of transistor Q2. When any one of these second string LEDs is illuminated, Q2 conducts, turning on Q1, which then independently lights the LM3914's entire first string. This means that each string draws 10 mA, a total supply load of just 20 mA when all the LEDs are on.

Photo 3: Punch holes in the blank panel using the template.

Power Supply

Two regulated 12-V switch mode "wall-warts" power the Headroom Meter circuitry. Admittedly, this is not the most elegant means of conjuring up a bipolar supply, but it's simple and quick. A single-chip switching regulator could develop –12 V from the +12 V input, but this simpler solution uses less circuit board real estate.

Nevertheless, a 3-pin header strip on the circuit board offers an alternate input for ±12 V regulated DC, power that could be borrowed from existing equipment. For multiple Headroom Meters, this connector makes it easy to daisy-chain several boards. Although the supply voltage is not all that critical for meter accuracy, these supplies do need to be regulated to 12 V, ±0.5 V.

Construction

All circuitry is contained on a small circuit board that forms a "sandwich" with the front panel (see **Photo 2**). Artwork for the circuit board is supplied as a 1:1 image and PCB Gerber files are available as a download.

Four 4-40 threaded spacers secure the board to the back of the panel, which may be cut to mount in a project box or integrated with an existing panel. You can make the panel larger or trim it down to fit a specific opening.

The front panel can be nearly any handy substance. I used 1/16″ sheet phenolic, but fiberglass circuit board material or thin hardboard (Masonite) are also easy to use. The panel drilling template, and the overlay artwork can be reproduced 1:1 on an office copier.

A high-resolution PDF has been included in the download file (see the Supplementary Material section on the *audioXpress* website). For the best reproduction of the black overlay, take the PDF file to a FedEx Office store and have them print it onto glossy card stock. You can apply a clear plastic laminate to the front for extra durability. Cut away the template and stick it down to your blank panel as a drilling guide. Take care to accurately center-punch the hole targets (see **Photo 3**). Then, drill all the holes with a 1/16″ drill (see **Photo 4**).

Open up the CAL and LED holes to 0.2″ with a 13/64″ or a #7 drill. Enlarge the four mounting holes to 1/8″. Countersink these four if you want to hide the screw heads under the overlay as I

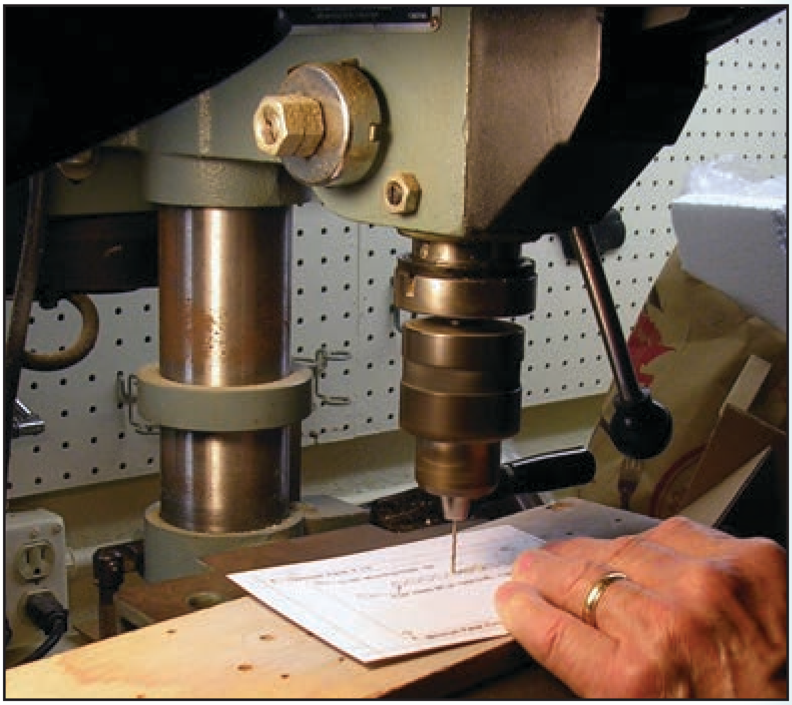

Photo 4: Drill the hols with a 1/16″ drill.

did (see **Photo 1** and **Photo 2**).

Drill the panel and then stick-down the overlay. 3M's "77" spray adhesive works well for such jobs. Once the overlay is in place, use a sharp X-Acto knife to open the holes and clean up the panel edges.

The Parts List and a Parts Locator drawing, along with the the PCB layout, the panel drilling template, and the overlay artwork are included in the Supplementary Material. All the components are through-hole parts. The connectors are soldered to the back of the board (see **Photo 2**).

Install solder and clip the components, observing polarity of the filter capacitors and all rectifier diodes. Insert the LEDs into their holes with the flat side (cathode) to the square pad on the left. Leave the leads sticking out the back at full length and don't solder these until the sandwich is assembled. At that point, push the LEDs through the holes in the front panel and solder them in.

Using the Headroom Meter

This is a single-channel device, so you'll need a pair for stereo. It would be quite possible to modify the circuit to indicate the higher channel of a stereo pair, but that involves a more complex rectifier block, outside the scope of this project.

Meter response to program peaks is near-instantaneous, and fallback is also pretty quick at around 100 ms. These ballistics ensure that even the briefest overload will be displayed, yet the meter will give a more "syllabic" representation of overall program dynamics. The meter's fast action enables you to visually ballpark the severity of any

About the Author

James Wood is Founder and Board Chariman of Inovonics, Inc., a 43-year manufacturer of audio and RF products for broadcast. Jim has been active in audio, radio frequency, and instrumentation circuit design and construction for more than 60 years. He holds several patents in the field and is a Life Member of the Audio Engineering Society (AES) and a Life Senior Member of The Institute of Electrical and Electronics Engineers (IEEE). A perennial student of scientific history, Jim takes particular delight in developing practical applications for arcane technologies.

full-scale clipping. The occasional quick flash of the 0 dB LED may suggest that the clipping will be more or less inaudible, while a sustained and more intense flash would definitely be cause for concern.

The input impedance of the Headroom Meter is 10 kΩ or more, which should have negligible loading effect on any respectable audio feed. Simply bridge the Headroom Meter across your system input using an RCA "Y" cable.

Input monitoring is recommended over monitoring a device's output. The clipping point of a power amplifier, for example, might be subject to AC mains voltage variations. At low-line your amplifier might not reach the clipping point you originally set the meter to show.

You'll need to apply a sine wave test tone to the system and have some means of telling when your preamp or amplifier (or whatever) goes into clipping. You'll certainly be able to view clipping on a scope at the output of your equipment, or you can trust your ears. Ears are most sensitive at about 3 kHz, so apply a 1 kHz tone and listen for the point at which it suddenly becomes "edgy."

At the clipping point, stick a small screwdriver through the CAL hole in the front panel and adjust R2 to so that the 0 dB LED glows at full brilliance. That's it. You should be set to go.

Project Files

To download additional material and files, visit http://audioxpress.com/page/audioXpress-Supplementary-Material.html.

Resource

"LM3914 Dot/Bar Display Driver," National Semiconductor, February 1995, http://pdf.datasheetcatalog.com/datasheet/nationalsemiconductor/DS007970.PDF

Sources

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1/16" Sheet phenolic
W.W. Grainger | www.grainger.com

Amplitrex AT1000 Tube Tester

We recently discussed the varieties of tube testers that have been used during the long history of hollow-state electronics. These units included continuity testers, tube checkers, emission testers, mutual conductance testers, dynamic conductance testers, and DC and AC parametric testers. In this article, we will focus on the Amplitrex AT1000 Tube Tester.

Photo 1: The AT1000 provides comprehensive test functions for electron tubes.

By
Richard Honeycutt
(United States)

In my article "Testing Tubes," (*audioXpress*, February 2016), I pointed out that each type has advantages and disadvantages, but the most reliable test for a tube's proper operation in a given circuit involves substituting a known-good tube. For matching and qualification of tubes by manufacturers or companies selling tubes, a parametric tester provides the most precise results, but it comes at a high expense in terms of acquisition cost and operator time.

Photo 2: The user selects the tube type to be tested in this screen. (Photo courtesy of Amplitrex)

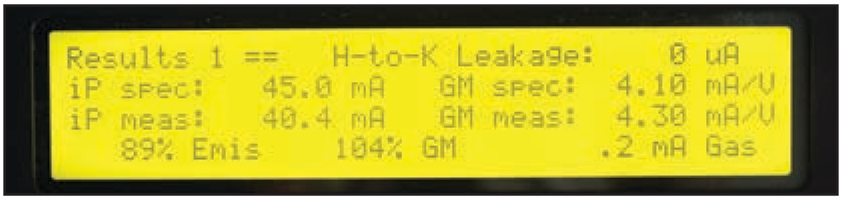


Photo 3: The test results are displayed on the LCD screen. (Photo courtesy of Amplitrex)

The AT1000 Electron Tube Test System manufactured by Amplitrex Audio Products, LLC (see **Photo 1**) couples the convenience of a simple emission tester with the precision of an AC parametric tester. The unit tests hundreds of popular vacuum tubes for emission, transconductance, and gas/leakage. It can also be used to test pentodes and tetrodes in triode mode. The AT1000 comes with a one-year warranty.

The Amplitrex AT1000's Features

Unlike testers of the past that required the user to determine when the tube had warmed up enough for testing, the AT1000 provides both timed and manual test modes. Fixed or auto bias can be used. Power tubes having defects that lead to thermal runaway/hot-plating can be detected.

A USB connection and accompanying software are included, enabling the user to create characteristic curves of tubes, and to set up test conditions for tubes not already included in the database. (See Resources for a list of the tube-type database.) The power supplies are regulated and overcurrent-protected. It can test tubes having 4-, 5-, or 6-pin bases, as well as octal and 7- and 9-pin miniature types, including those using plate or grid caps. Connector cables for these tubes are included. Socket adapters and socket savers are available for purchase as well, enabling the user to test almost any tube, including loctal types. The built-in sockets are of excellent quality.

Unlike many computer-enabled devices, the AT1000 comes with operator manuals for the tester itself and its accompanying software in a digital pdf format.

For most tests, the AT1000 can function as a stand-alone device. In this mode, the LCD screen and five tactile-feedback navigation keys provide the user interface. The backlit LCD display contains four lines of 40 characters each. Instructions on the display guide the user through mode, tube, and socket selections. Test setup information and the test results are also displayed on the LCD screen.

The operator can listen to a tube's output voltage through a standard 0.25" TRS stereo jack. This enables users to check for excess noise, microphonics, or popping caused by gas in the tube. The jack can feed headphones or a speaker, an oscilloscope, or AC voltmeter. The signal level at the audio-out jack is controlled by a volume control potentiometer.

The AT1000's Capabilities

When the AT1000 is first powered on, the screen shown in **Photo 2** appears, and the operator selects the type to be tested by using the arrow and ENTER keys. Next, a screen prompt indicates which socket on the AT1000 to use. After inserting the tube into the socket, the user presses ENTER again. If the tube requires a socket adapter, a screen will appear indicating that an adapter is required and which socket to use. If the tube being tested has a plate or grid cap, another screen will appear advising which banana jack (plate or grid) to plug the cap/lead into. The AT1000 is even capable of testing a tube that has both plate and grid caps.

At this point, the test will automatically begin. The heater is ramped up to its specified voltage, and a timer counts down. When the counter reaches zero, the warmup phase ends, the heater-to-cathode leakage test is done, and then B+ is applied to the tube, as indicated on the LCD screen. The tube is biased into conduction in one of two modes, as chosen in the system menus. In Auto Bias Mode, the grid voltage is automatically adjusted downward until the specified plate current is reached. In Fixed Bias Mode, the specified bias voltage is applied to the grid, and the plate current will be whatever results at this bias setting.

Once the plate current is flowing, the unit will automatically test the transconductance,

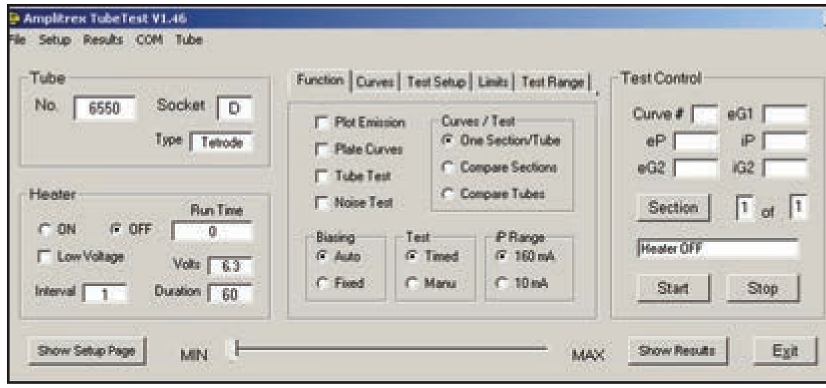


Photo 4: This is the opening screen for the Tube Test software.

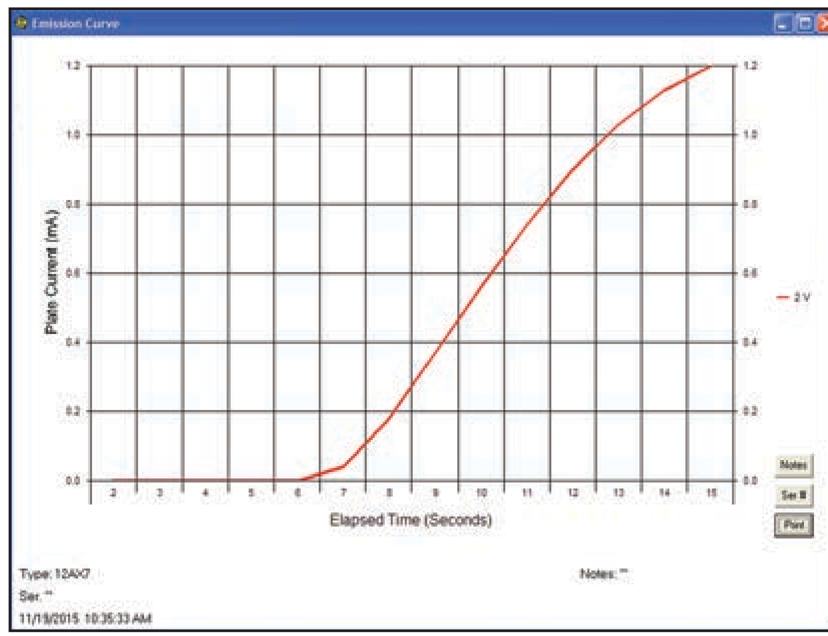


Photo 5: The emission vs. time of a 12AX7 tube is displayed.

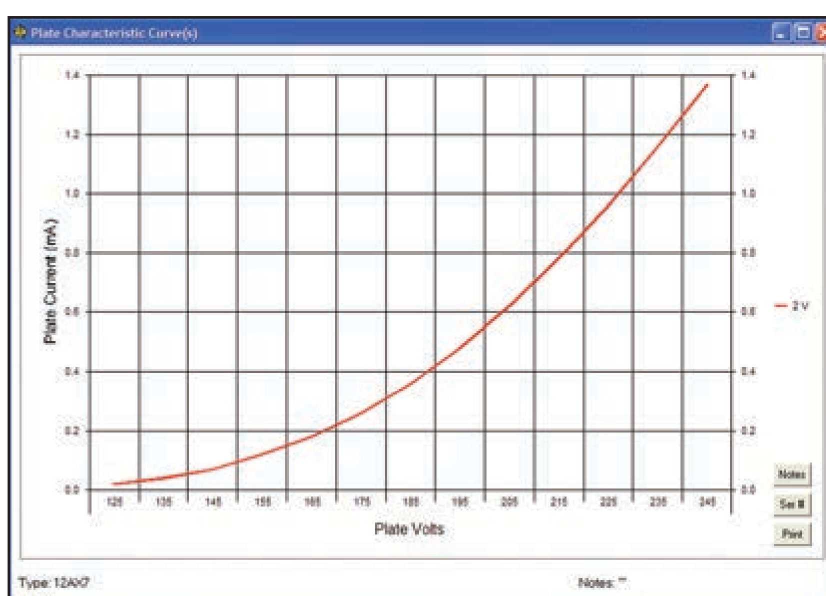
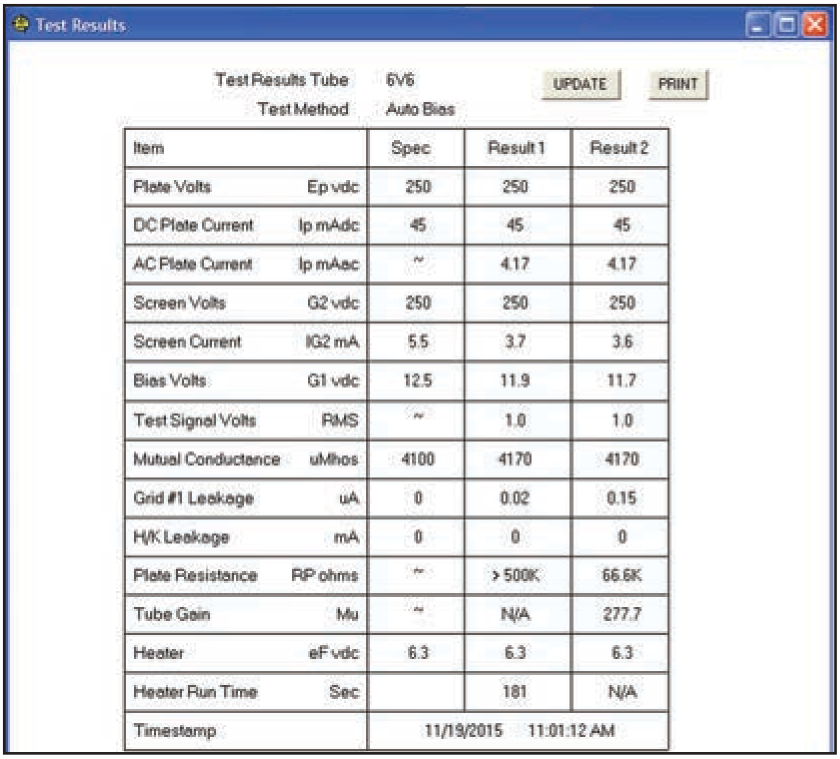


Photo 6: The software displays the plate curve measured from a 12AX7 tube.

if the AT1000 is in Timed Test Mode. If the unit is configured for Manual Test Mode, the user is prompted by the LCD display to "Press ENTER to continue." This enables the user to wait until plate current stabilizes, if desired, before continuing with the Transconductance Test. (The user can change the unit between Timed and Manual Test Modes by using the system menus.) A 1 kHz sine wave is applied to the grid of the tube. The AC plate current is measured, and the tube's transconductance is calculated and displayed.


Test Results

Test Results Tube 6V6

Test Method Auto Bias

UPDATE PRINT

Item		Spec	Result 1	Result 2
Plate Volts	Ep vdc	250	250	250
DC Plate Current	Ip mAdc	45	45	45
AC Plate Current	Ip mAac	~	4.17	4.17
Screen Volts	G2 vdc	250	250	250
Screen Current	IG2 mA	5.5	3.7	3.6
Bias Volts	G1 vdc	12.5	11.9	11.7
Test Signal Volts	RMS	~	1.0	1.0
Mutual Conductance	uMhos	4100	4170	4170
Grid #1 Leakage	uA	0	0.02	0.15
H/K Leakage	mA	0	0	0
Plate Resistance	RP ohms	~	> 500K	66.6K
Tube Gain	Mu	~	N/A	277.7
Heater	eF vdc	6.3	6.3	6.3
Heater Run Time	Sec		181	N/A
Timestamp		11/19/2015 11:01:12 AM		

Photo 7: Two 6V6 tubes are compared.

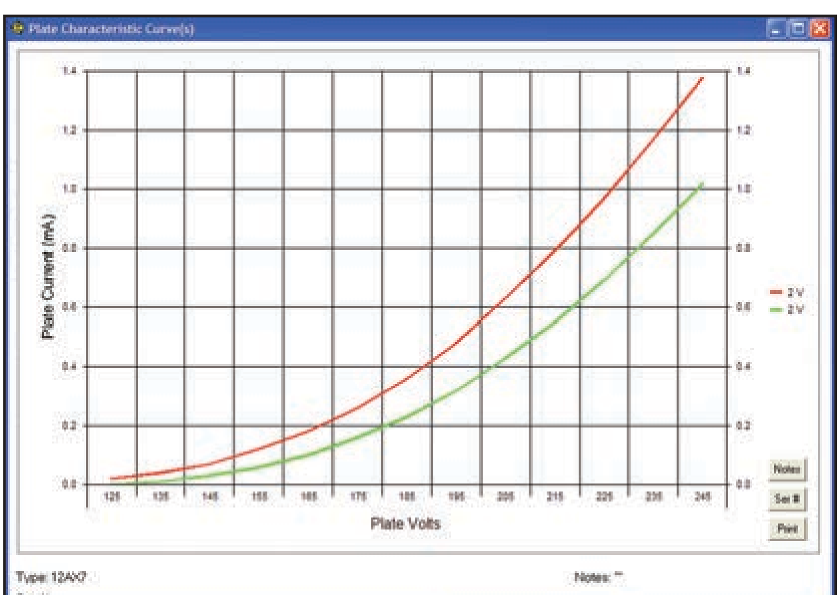


Photo 8: The two triode sections of a 12AX7 tube are compared.

Next the Gas Test begins. For the gas test, two plate current measurements are made, using different grid circuit resistances. The difference in these two readings becomes the Gas Result Value, and indicates the effect of the tube's gas contents (or other source of grid leakage).

After the gas test concludes, B+ is removed from the tube and the Test Results Screen is displayed (see **Photo 3**).

If there is more than one section in the tube under test, the other section(s) are tested before the results are shown. Each section of a multiple-section tube is tested as if it were a separate tube, and the results are shown separately for each section.

Any time B+ is applied to the tube, the AT1000 protects itself by shutting down if a short or overload condition occurs. In this case, the user is notified by a message that a short or overload has been detected.

With tubes that contain diodes or diode sections, such as the 6AL5 and the 6T8, the Diode Results screen will be displayed. In addition to the Heater-to-Cathode Leakage measurement that appears on the top line, a reading for forward and reverse current is shown. For full-wave rectifiers (e.g., the 5U4, the 5Y3, and the GZ34), a specialized results screen is displayed. It shows the measured forward and reverse voltage during the test, for both sections.

Electron Ray Indicator tubes (e.g., the 6E5 and the EM84), can be tested on the AT1000. In these cases, a special test is run on the display section of the tube. During the test, the tube's fluorescent screen will light up, and the shadow will move as the AT1000 varies the Ray Control Electrode voltage.

Vacuum tubes can be matched for use in balanced circuits, such as push-pull output or fully-differential preamplifier circuits. Single-section tubes can be matched to each other, and the sections of dual section tubes can be compared both within and between tubes. Tubes can be matched by emission, transconductance, or both.

The Software

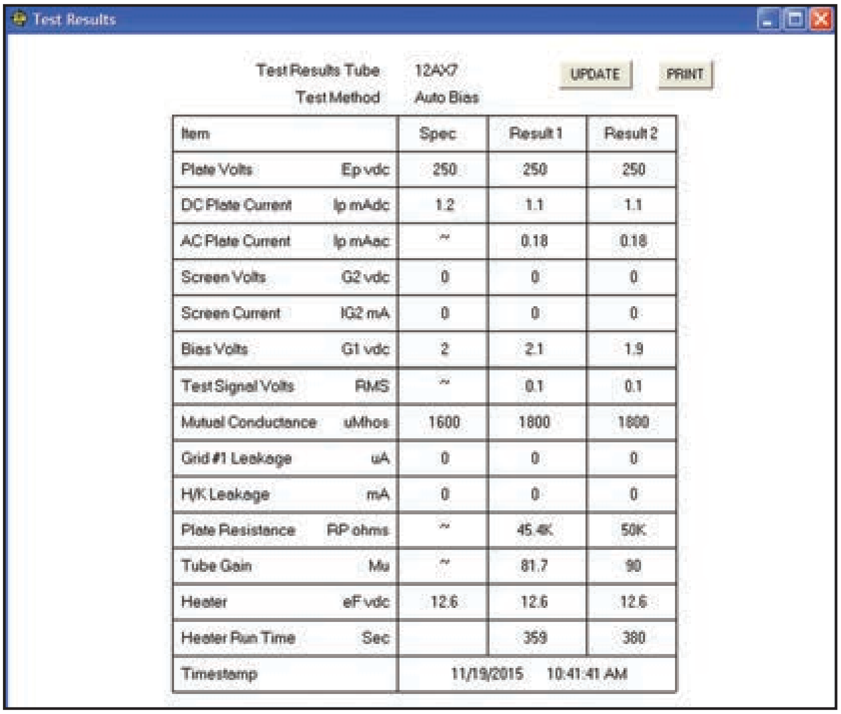
The Amplitrex Tube Test software is supplied with the AT1000. It provides a variety of computer-controlled tests for triode, tetrode, beam-power and pentode tubes.

After the software has booted, it displays the screen shown in **Photo 4**. The software can measure emission vs. time during warmup, as shown in **Photo 5**; the plate characteristic curve, as shown in **Photo 6**; tube-to-tube comparisons, as shown in **Photo 7**; section-to-section comparisons within the same tube, as shown in **Photo 8**; or complete test results (see **Photo 9**).

Overall Impressions

The Amplitrex AT1000 is a comprehensive tube-testing solution for serious hobbyists, technicians repairing hollow-state equipment, or sellers of hollow-state components. The price is not low, but compared to the cost of any device even remotely comparable several decades ago (not even considering inflation), it provides immense value for the money. There are a few other new tube testers on the market today, but most are limited to tubes commonly used in musical electronics, and none performs the variety of tests provided by the AT1000.

Contained in a heavy-duty carrying case, the AT1000 is rugged. Yet, even counting the weight of a laptop computer to allow extended functionality, it is still lighter than a vintage Hickok Cardmatic. More information about the Amplitrex AT1000 is available at www.amplitrex.com.


Test Results

Test Results Tube 12AX7 UPDATE PRINT

Test Method Auto Bias

Item		Spec	Result 1	Result 2
Plate Volts	Ep vdc	250	250	250
DC Plate Current	Ip mAdc	1.2	1.1	1.1
AC Plate Current	Ip mAac	~	0.18	0.18
Screen Volts	G2 vdc	0	0	0
Screen Current	IG2 mA	0	0	0
Bias Volts	G1 vdc	2	2.1	1.9
Test Signal Volts	RMS	~	0.1	0.1
Mutual Conductance	uMhos	1600	1800	1800
Grid #1 Leakage	uA	0	0	0
H/K Leakage	mA	0	0	0
Plate Resistance	RP ohms	~	45.4K	50K
Tube Gain	Mu	~	81.7	90
Heater	eF vdc	12.6	12.6	12.6
Heater Run Time	Sec		359	380
Timestamp		11/19/2015 10:41:41 AM		

Photo 9: The complete tube-test results are displayed.

Resources

"A List of Preloaded Tubes," Amplitrex Audio Products, LLC, http://amplitrex.com/list-of-preloaded-tubes.

R. Honeycutt, "Testing Tubes," *audioXpress*, February 2016.

Industry Calendar

Here are a few places where you might find a copy of *audioXpress* and possibly meet one of our authors and staff members:

February 29–March 3, 2016
Prolight + Sound Guangzhou 2016
China Import and Export Fair Complex, Guangzhou, China
https://prolight-sound-guangzhou.hk.messefrankfurt.com/guangzhou/en/visitors/welcome.html

China's largest show for the professional audio and lighting industries, the Guangzhou edition of Prolight+Sound is strongly focused on entertainment technologies. Visitors will find more lighting and rigging equipment than actual audio solutions. Still, compared to Prolight+Sound Shanghai in October, the Guangzhou show is growing faster and allowing for excellent business opportunities for professional audio companies.

March 15–17, 2016
electronica China 15th International Trade Fair for Electronic Components, Systems, and Applications
Shanghai New International Expo Centre, China
http://electronica-china.com

Since 2002, electronica China has transformed into a leading show for electronic components, systems and applications in China. electronica China takes place in conjunction with productronica China, Laser World of Photonics China, and SEMICON China, creating attractive synergies and added value for visitors.

March 18–20, 2016
Salon Image & Son
Hotel Bonaventure, Montréal, QC, Canada
www.chestergroup.org/salonsonimage/2016

The Montréal Salon Son & Image Expo is one of the longest running shows of its kind worldwide. Visitors can compare several audio brands, ranging from top of the range headphones to extraordinary loudspeakers while sitting in self-contained listening suites. Hundreds of audio brands and a unique environment have helped turn this show in a solid proposition for all high-end and consumer audio professionals.

March 19–20, 2016
CanJam SoCal 2016
Westin South Coast Plaza, Costa Mesa, CA
www.head-fi.org

An established event for one of the fastest growing segments of the audio industry, CanJam is dedicated to everything that goes on and around our ears, not just headphones and in-ears. The CanJam has expanded its focus and also features the audio players, the preamps, and the electronics that complete the experience. In March, 2016 more than 1,000 audio enthusiasts are expected to gather in Southern California for this headphone and personal audio expo. CanJam is expanding worldwide, but the Orange County edition is a must-attend.