

A Dual-Channel MLS-Based Test System for Hearing-Aid Characterization*

T. SCHNEIDER, AES Member, AND D. G. JAMIESON

*Hearing Health Care Research Unit, Department of Communicative Disorders,
University of Western Ontario, London, Ont., N6G 1H1, Canada*

A dual-channel maximum-length-sequence test system has been implemented to characterize the electroacoustic performance of hearing aids. The system applies a frequency-shaped maximum-length sequence to a hearing aid as an acoustic stimulus and measures the acoustic response of the aid in an IEC standard occluded-ear simulator. Measurements are completed in one-third the time required for noise-based measurements, and the responses measured by the system compare favorably to those measured via conventional noise-based methods.

0 INTRODUCTION

This paper describes the design and implementation of a maximum-length-sequence (MLS)-based test system for the evaluation of hearing aids and similar devices. Previous publications have described applications for MLSs in loudspeaker testing and the analysis of room acoustics [1], [2]. Only recently has the possibility of applying MLSs to characterize hearing aids been discussed [3], [4]. No complete reports of successful applications have been published.

Most conventional hearing-aid testing systems use swept tones as input signals. An American National Standards Institute (ANSI) standard, S3.22 [5], details the set of swept-tone tests commonly used to characterize hearing aids in North America; two International Electrotechnical Commission (IEC) standards, Publications 118-0 and 118-2 [6], [7], detail the comparable tests that are widely used in Europe. While these tests are simple to implement, they can be misleading with respect to how many modern hearing aids will behave in typical use [8], [9]. For example, swept-tone testing of an aid that uses automatic gain control (AGC) often produces a relative artificial accentuation of low-frequency gain known as "blooming" [8]. Blooming occurs because many AGC aids have frequency-sensitive AGC circuits that require more energy at low frequencies for AGC activation. Because a pure-tone test presents only one frequency at a time, the frequency response measured in this way is not indicative of the frequency response when the hearing aid is exposed to complex sounds.

An approximation of the characteristics of the hearing aid in actual use can be obtained by using a signal which is acoustically similar to "real-world" input signals. Normally the input signal of interest is speech, possibly in the presence of an interfering noise background. For this reason there has been a recent trend toward using broad-band "speechlike" signals to test hearing aids [8]–[10]. A new ANSI standard, S3.42 [11], uses a speech-shaped noise test signal and fast Fourier transform (FFT)-based analysis for frequency response measurement. This approach avoids "blooming" and gives a frequency response measurement that is a better approximation of a hearing aid's response to a "real-world" stimulus.

It has also been suggested to the ANSI standards committee that MLS-based testing be considered as a candidate for hearing-aid measurement [3]. This scheme uses an MLS to excite a linear system and obtain an impulse response. The frequency response can then be

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found by taking the discrete Fourier transform of the impulse response.

This method has several advantages over traditional noise-based test methods. In the absence of external interference (noise), the frequency response can be computed accurately with no averaging. Noise-based measurements require averaging to obtain a good estimate of the input and output (or cross) power spectra: as more records are averaged, estimates of the power spectra improve. In the presence of noise, time-domain averaging can be used with MLS-based testing to improve the signal-to-noise ratio (SNR), but this is not required.

We have also found MLS-based testing to be faster than noise-based measures. Using the fast Hadamard transform (FHT), the computation of the impulse response is completed very rapidly (faster than an FFT) and only a single FFT is required to obtain the frequency response. Noise-based measurements require the computation of many windowed, overlapping FFTs that are averaged to obtain a power spectrum.

The power spectrum of an MLS is white. A procedure similar to that used to compute the impulse response can also be used to filter an MLS. Thus arbitrarily shaped (such as speech-shaped) MLSs can be quickly computed and applied to a hearing aid.

The remaining sections of this paper present the basic theory behind MLS-based testing and describe the design and implementation of our dual-channel system. Test results for two hearing aids are also presented.

1 BACKGROUND

MLSs have been used in system analysis since the early 1970s [12]. Only a summary of important results is presented here; the theory has been presented fully in previous papers [1], [13], [14].

1.1 Basic Theory

A discrete linear time-invariant (DLTI) system can be completely characterized by its unit sample or impulse response (IR), $h[n]$. The unit sample sequence $\delta[n]$ is defined as

$$\delta[n] = \begin{cases} 1, & \text{for } n = 0 \\ 0, & \text{otherwise} \end{cases} \quad (1)$$

Let $x[n]$ be the input to a DLTI system that has IR $h[n]$. The output $y[n]$ (Fig. 1) is given by

$$y[n] = x[n] * h[n] \equiv \sum_{k=-\infty}^{+\infty} x[k]h[n-k] \quad (2)$$

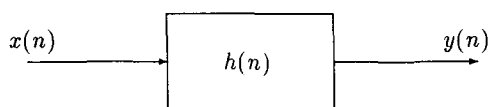


Fig. 1. Linear-system block diagram.

where $h[n]$ is the output of the system for the input $\delta[n]$ and $*$ indicates convolution.

An MLS is a binary, periodic, pseudorandom sequence of length $L = 2^N - 1$, where N is an integer ($N > 0$) known as the MLS order. MLSs are generated by shift registers with modulo-two feedback taps [15]. Because MLSs are periodic, we must necessarily measure the periodic impulse response (PIR) of a DLTI system. Note that the IR can be considered as a PIR with a period approaching infinity. The periodic unit sample is defined as

$$\delta'[n] = \begin{cases} 1, & n \bmod L = 0 \\ 0, & \text{otherwise} \end{cases} \quad (3)$$

that is, $\delta'[n]$ equals 1 every L samples; otherwise it is 0. For a periodic input, a DLTI system output is computed via circular convolution (indicated by $\star$),

$$y'[n] = x'[n] \star h'[n] \quad (4)$$

where the prime (') indicates a periodic sequence. The indices for all periodic sequences are computed modulo L .

For measurement purposes, a PIR and an IR closely approximate each other over the first L samples, provided that the IR decays to a negligible value over L samples. This stops the PIR from wrapping back on itself and causing an error known as *time aliasing* [13]. Thus the duration of the MLS must be long enough so that the IR decays to a negligible value.

For measurement of an analog system, the usual binary values of an MLS use the mapping $\{0, 1\} \rightarrow \{+V_0, -V_0\}$ to produce a sequence that is symmetric about zero. A circular autocorrelation, denoted by Φ , is defined as

$$\Omega_{ss}[n] = s'[n]\Phi s'[n] \equiv \frac{1}{L+1} \sum_{k=0}^{L-1} s'[k]s'[n+k] \quad (5)$$

(Note that the autocorrelation is normalized by $L+1$ instead of by L .) If $s'[n]$ is a symmetric MLS, then

$$\Omega_{ss}[n] = \begin{cases} \frac{L}{L+1}, & n = 0 \\ \frac{-1}{L+1}, & 0 < n < L \end{cases} \quad (6)$$

This can also be written as

$$\Omega_{ss}[n] = \delta'[n] - \frac{1}{L+1} \quad (7)$$

Thus a symmetric MLS has the special property that its autocorrelation is *approximately* a periodic unit

sample.

If an MLS is applied to a DLTI system with PIR $h'[n]$, the system output $y'[n]$ is

$$y'[n] = s'[n] \star h'[n] = \sum_{k=0}^{L-1} s'[k]h'[n-k] \quad (8)$$

To compute the PIR for the system, the system output and input are cross-correlated,

$$\begin{aligned} \Omega_{xy}[n] &= s'[n]\Phi y'[n] \\ &= s'[n]\Phi (s'[n] \star h'[n]) \\ &= \Omega_{ss}[n] \star h'[n] \\ &= \left(\delta'[n] - \frac{1}{L+1} \right) \star h'[n] \\ &\approx h'[n] - \frac{1}{L} \sum_{k=0}^{L-1} h'[n] \quad \text{for large } L \end{aligned} \quad (9)$$

that is, $\Omega_{xy}[n]$ is the impulse response with the dc component removed.

The frequency response can be computed by taking the DFT of the PIR. Note that the computed PIR is always $L = 2^N - 1$ samples long. Thus the PIR must be zero-padded with one sample if an FFT is used. If the PIR decays out quickly, the computational load can be reduced by truncating and windowing the PIR so that only the "useful" portion is transformed. This reduces the effects of noise and distortion [13] on the frequency response. To improve the SNR of the MLS measurement, the system output can be averaged (in the time domain) over a number of MLS periods before the PIR is computed via the FHT cross correlation. This operation is called preaveraging.

1.2 Fast Hadamard Transform and Fast Convolution

The number of operations required for direct computation of the input-output cross correlation is $L(L-1)$ for an MLS of length L . This can be reduced significantly by utilizing the FHT [14]. The cross-correlation computation [from Eq. (9)],

$$\Omega_{xy}[n] = \frac{1}{L+1} \sum_{k=0}^{L-1} s'[k]y'[n+k] \quad (10)$$

can be written in vector notation as

$$\Omega_{xy} = \left(\frac{1}{L+1} \right) SY \quad (11)$$

where S is a right-circulant matrix containing circularly shifted versions of the MLS. Notice that only additions and subtractions are required to compute the matrix

multiplication since S contains only $\{-1, +1\}$.

The FHT uses a successive doubling approach (similar to the FFT) to reduce the number of operations to approximately $2.5 L \log_2(L)$ for an MLS of length L . To compute the cross correlation quickly, the noise matrix S is converted to a Sylvester-type Hadamard matrix [14]. The matrix multiplication is then computed via the FHT and the conversion is "undone." In matrix notation this sequence of operations is

$$\Omega_{xy} = \left(\frac{1}{L+1} \right) P_2 S_2 H_{L+1} S_1 P_1 Y \quad (12)$$

where $P_2 S_2 H_{L+1} S_1 P_1 = S$.

The derivation of the selection matrices (S_1, S_2) and the permutation matrices (P_1, P_2) is covered by Borish and Angell [14]. Borish also presents a method of computing the permutations in real time using a set of shift registers [16]. More recently Xiang [17] derived a permutation method that requires the calculation of only one permutation matrix. Our implementation computes the permutation matrices directly from the MLS when it is generated and stores them on disk.

An algorithm for quickly convolving an MLS with a vector of FIR filter coefficients F is very similar to the fast convolution algorithm [18]. Let the filter output be Y . Then

$$Y = S^T F = P_1^T G_L P_2^T F \quad (13)$$

where ^{T} indicates transpose, $G_L = S_2 H_{L+1} S_1$ is symmetric, and S^T contains circularly shifted versions of the MLS. Thus the only changes are to apply transposed versions of the permutation in the reverse order without scaling. This algorithm allows very high-order filters (order = MLS length) to be computed rapidly, so that a frequency-shaped MLS can be used as a stimulus.

1.3 MLS Crest Factor

A new ANSI standard (S3.42) for noise-based measurements requires that the crest factor of the test signal be $12 \text{ dB} \pm 3 \text{ dB}$. Rife and Vanderkooy [13] report that when an MLS passes through a linear system having an impulse response longer than a few samples in duration, a quasi-Gaussian probability density function with a nominal crest factor of 11 dB is produced. In our system the MLS passes through a number of filters and other equipment before it is applied to the hearing aid. Thus the crest factor of the acoustic MLS stimulus will comply with ANSI S3.42.

To confirm this result, two crest-factor measurements were made on the shaped and flattened MLS (see Section 2.2.4). The crest factor of the signal recorded from the reference microphone was 14.6 dB; the crest factor of the MLS stored on disk was 10.2 dB.

2 SYSTEM DESIGN

To make the test system practical for a clinical setting, we used readily available low-cost test equipment. The

system was designed with flexibility in mind so that future tests can be developed rapidly.

MLS-based frequency response measurements apply an acoustic MLS to a hearing aid at a number of input sound pressure levels (SPLs). By testing at different input SPLs we can observe the effect of the input level on the frequency response and determine the degree of distortion at each input level. For practical measurement times (such as a few seconds), asynchronous noise-based measurements will have "jagged" frequency response curves if the hearing-aid output is distorted. Averaging many FFT records will smooth the responses, but this leads to excessively long measurement times. Similarly, distortion causes MLS-based frequency response curves (which are synchronous measurements) to appear "jagged" or "lumpy" [13], [19].

Testing is done at two hearing-aid gain settings: at the reference test position (RTP) and at full-on gain (FOG). The FOG is set by adjusting all controls of the aid to the position in which they will produce the maximum output SPL. The RTP is set by measuring the output SPL in the FOG condition for 90-dB SPL input, then reducing the input to 60-dB SPL and adjusting the gain of the aid so that the output SPL drops by 17 dB [11].

Measurements step the input level from 90- to 50-dB SPL in 10-dB steps. Frequency responses for all input SPLs are plotted on the same graph. For aids with compression ratio adjustment or tone controls, a number of measurements are usually made at typical control settings.

For all of the tests presented in this paper, the frequency spectrum of the MLS sequence is modified so that it is shaped as shown in Fig. 2. This spectrum approximates the long-term average spectral characteristics of speech [11], [20].

2.1 Hardware

The test system hardware components are shown in Fig. 3. The DSP56 (digital signal processor board) and QDA2 (digital playback board) both reside in a 33-MHz 80486-based EISA PC that is used for data collection and computations. The DSP56 samples both

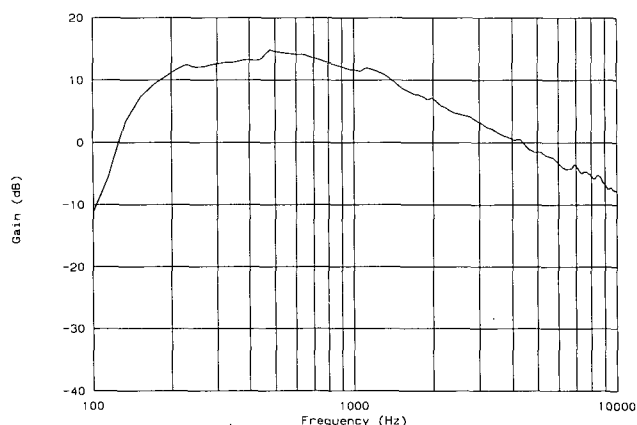


Fig. 2. Speech-shaped MLS spectrum.

channels simultaneously and records data directly to disk. The QDA2 plays stimuli that have been uploaded to on-board memory. To ensure that record and playback operations are tightly synchronized, the DSP56 uses hardware triggering to start the QDA2 playback. Both devices share the same sample clock (sample rate 39.0625 kHz).

A programmable output attenuator (PA_OUT) is used to adjust the SPL in the portable anechoic chamber. To reconstruct the output signal after the digital-to-analog converter, an 8th-order elliptic low-pass filter (LPF) with a cutoff frequency of 12 kHz is used. The output from this filter feeds the power amplifier that drives the loudspeaker in the anechoic chamber.

Two programmable input attenuators (PA_REF and PA_CPL) and a dual-channel programmable preamplifier/high-pass filter are used to scale the input level into range for the DSP56 analog-to-digital converters. The high-pass filters ($f_c = 80$ Hz) remove low-frequency room (ventilation) noise.

A Bruel & Kjaer type 4134 pressure microphone is used on each input channel. The reference microphone and the hearing aid (that is, the device under test) are located opposite each another inside the chamber (equidistant from the center of the chamber) in a symmetric sound field. The acoustic output of the hearing aid is connected to an IEC standard occluded-ear simulator [21] to approximate the "acoustic environment" of an ear. The coupler channel microphone is inside the occluded-ear simulator.

Type 4134 microphones typically have a sensitivity of 11.2 mV/Pa (0-dB SPL referred to 20 μ Pa). Thus the output voltage from the microphone is given by

$$V_0 = (11.2 \times 10^3)(20 \times 10^{-6})10^{\text{SPL}/20} \quad (14)$$

For 130-dB SPL (the highest output SPL) the output voltage is 708.3 mV; for 50-dB SPL (the lowest input SPL) the output voltage is 0.22 μ V. High-input and output SPLs will produce microphone output voltages that are high enough to avoid noise problems. However, for low-input SPLs, electrical noise may be a problem. Also, unless the SNR is improved by preaveraging MLS periods (see Section 1.1), ambient room noise and electrical noise may cause measurement difficulties at lower SPLs.

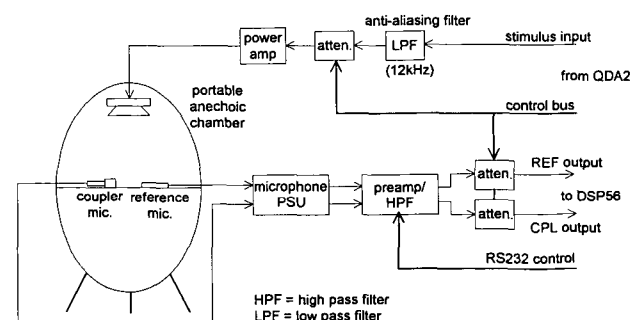


Fig. 3. Test-system hardware components.

2.2 Software

The software is divided into three major components: level setting, data collection, and postprocessing. The level-setting portion of the program calibrates the output SPL in the anechoic chamber and adjusts the input attenuators to best utilize the input dynamic range. Samples from both channels are recorded directly to disk in the data collection phase of operation. Following data collection, the data from each channel are post-processed and the frequency response is computed.

Test parameters are entered on the command line. If no parameters are specified, the following default parameters are used: starting SPL = 90 dB, final SPL = 50 dB, SPL step size = 10 dB, MLS order = 14, test with FOG, and no frequency response smoothing.

2.2.1 Level Setting

1) *Output SPL*. The program first sets the output SPL by successively adjusting the reference input channel gain and PA_OUT until the target output SPL is achieved and the reference channel (REF) peak input level is within the desired range.

The output SPL is measured in time domain using an rms measurement with a time constant of 1 s. The SPL measurement is computed for four time constants and then averaged over the last time constant. After each computation of the output SPL, the output attenuation is adjusted to achieve the target output SPL. If the REF channel input level is out of range, the REF channel input gain is set [see 2) Input Levels]. Once the output SPL is within ± 0.5 dB of the desired value, the procedure is halted.

Once a target output SPL setting is reached, the coupler channel (CPL) input level is set using the same procedure that is used for the REF channel. The CPL channel SPL is then measured and stored.

2) *Input Levels*. To best utilize the input-channel dynamic range, both DSP56 input channel gains are automatically adjusted to avoid overload while the peak level is set within a desired window. The overload threshold (OL) is set 3.0 dB below the input saturation level (SL) to allow for a margin of error (Fig. 4). A desired level threshold (DL) is set 6.0 dB below the OL point.

In operation, the input is recorded for one period of the MLS. The gain required to raise the peak level to

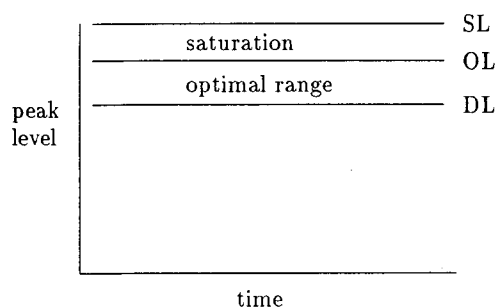


Fig. 4. Input-channel level-setting thresholds.

at least the OL point is computed and applied to the signal. (The input preamplifiers adjust in 10-dB steps.) Then the input attenuator (0.1-dB steps) is adjusted using a binary search to find the attenuation where no more than one sample exceeds OL and at least five samples lie between DL and OL. If the two attenuation end points used for the binary search converge and no samples are in range, the input calibration is halted. This condition is then reported and the frequency response measurement starts. The same procedure is used to adjust both input-channel (REF and CPL) gains.

To reduce the effects of equipment and room noise, the output SPL is set at the highest SPL (typically 90 dB), and the output attenuator is used to step to the other SPLs required for the test. The reference channel SPL is not recomputed for each step.

2.2.2 Data Collection

Once both input levels are set and the output SPL is at the target value, data collection begins. The DSP56 records both channels to disk simultaneously in an interleaved stereo data file containing 16-bit samples. To eliminate transient "start-up" effects, the first period from each channel is discarded. Additional samples are also recorded to account for the group delay through the DSP56 ADC16 analog-to-digital converters [22]. An equivalent number of samples is discarded from the start of the data file for each channel. The interleaved file is then split into two separate files (one for each channel), and the remaining periods are averaged.

2.2.3 Postprocessing

The average for each channel is cross-correlated using the FHT-based algorithm described in Section 1.2. The permutation matrices required for the cross correlation are read from disk. They are generated in advance by a separate program that also generates the MLS and stores it on disk.

The resulting impulse response for each channel (REF and CPL) is truncated from 16383 to 2048 samples, windowed with a Hanning window, and transformed with an FFT. This yields the frequency response for each channel. The magnitude of the response for each channel is expressed in decibels and is corrected by the input channel gains and a scaling factor (see Section 2.2.4). The frequency response of the hearing aid is then

$$H_{\text{aid}}(f) = H_{\text{CPL}}(f) - H_{\text{REF}}(f) - H_{\text{diff}}(f) \text{ dB} \quad (15)$$

where $H_{\text{diff}}(f)$ is the frequency response difference between channels that is computed during calibration (see Section 2.2.4) and saved on disk.

2.2.4 Calibration

Numerous errors can adversely affect measurements made with the test system. Effects of room and equipment noise can be reduced by preaveraging, provided that the measured responses are coherent and the noise

is incoherent with respect to the MLS test signal [1]. The order of the MLS also influences the noise and distortion rejection of an MLS measurement. For longer MLS periods and a constant analysis window size, more distortion and noise are rejected as the MLS order is increased because of the phase randomization and energy conservation properties possessed by MLS measurements [23]. Originally we opted for a fixed 14th-order MLS (and preaveraging) in our design because orders higher than this required more memory than could be allocated and accessed quickly by our DOS-based processing software. (We have since overcome this constraint by switching to a protected-mode compiler.) Microphone placement will introduce significant intensity errors only at high frequencies. These errors can be reduced through careful calibration and consistent microphone placement.

Input SPL measurements on each channel are calibrated at 250 Hz to within ± 0.1 dB using a pistonphone microphone calibrator. A calibration program writes a calibration file to disk that contains the microphone sensitivity (mV/Pa) entered by the user and a scaling factor that accounts for the gain of the analog input circuitry on the DSP56 [24], the input channel components, and the analog-to-digital conversion process.

The frequency response errors from the reference, coupler, and output channels are corrected by "flattening" the reference channel frequency response and removing the difference in frequency response between the two input channels. The REF channel frequency response error is flattened by first measuring the frequency response of the reference channel with an unshaped MLS. Then an inverse frequency response magnitude between 100 Hz and 10 kHz is computed and multiplied by the speech-shaping filter response. A 2048th-order FIR filter that approximates the combination of these two responses is designed (using a least-squares frequency sampling approach [25]) and applied to the MLS. The filtered MLS is written to disk. This process flattens the REF channel frequency response and applies the desired frequency shaping. The phase response is not corrected because we do not make phase response measurements on hearing aids.

The difference in frequency response between the REF and CPL channels, $H_{\text{diff}}(f)$, is found by making a dual-channel frequency response measurement without the coupler installed. This difference is stored and subtracted from the computed response each time the frequency response of a hearing aid is computed. The combination of these two frequency response computations (flattening the reference channel and subtracting the difference between channels), flattens both channels, shapes the MLS, and delivers the desired acoustic input signal to the hearing aid.

3 TEST RESULTS

The selection of measurement parameters for the tests presented here involved a number of tradeoffs. The sampling rate was fixed at 39 062.5 kHz by hardware

constraints. In general, hearing-aid responses do not have significant gain below 200 Hz or above 10 kHz (in fact, significant gain is rare above 6 kHz). Thus the analysis bandwidth was limited to 200 Hz to 10 kHz.

For MLS-based frequency response measurements, the sequence length, number of periods to average, and FFT size must be chosen. The duration of the MLS must be longer than the longest IR to be measured. Preliminary tests on a number of hearing aids showed that a 14th-order MLS (duration 419 ms) was more than long enough to eliminate time aliasing. A 14th-order MLS sequence fits into a 64-kbyte array if single-precision floating-point (4-byte) values are used. (There was a 64-kilobyte array size limitation with the original DOS real-mode version of our software.) Averaging eight periods in the time domain provides reasonably fast execution; it improves the measurement SNR by 9 dB.

The FFT size is a tradeoff between execution time, frequency resolution, low-frequency error, and noise/distortion rejection [13]. A 2048-point FFT was found to be long enough to encompass the decay of the linear portion of the PIR. This size of FFT provides fast execution together with a 19-Hz frequency resolution, which is more than adequate. The use of a 14th-order MLS and a 2048-point analysis window provides an eightfold (18-dB) reduction of distortion and noise (assuming that noise and distortion are uniformly distributed over the entire PIR period).

3.1 Calibration

3.1.1 SPL Errors

Input-channel SPL measurement error and output-channel SPL setting error were measured by requesting an SPL from the test system and recording the SPL measured by the system. The effects of ambient noise were minimized by placing the portable anechoic chamber within a double-walled sound-attenuating room. A reference SPL measurement was also made at the same time, using a Bruel & Kjaer type 2035 spectrum analyzer. A wide-band (white-noise) test signal was used at 90- to 50-dB SPL input level (5-dB SPL steps). The measurement bandwidth was 80 Hz to 20 kHz.

The worst-case errors occurred at the lowest SPLs and are a result of ventilation and equipment noise (measured as 51-dB SPL). Below 55-dB SPL, the worst-case total error is 4.2 dB. This error is due to line-frequency harmonics in the input-channel components and low-frequency ventilation noise. For 60-dB SPL input level or greater, the worst-case total error was 0.8-dB SPL. This is within the error allowed by ANSI S3.22 [5].

3.1.2 Frequency Response Errors

Fig. 5 shows the unflattened and flattened responses for the reference channel at 90-dB SPL input level. Preliminary testing showed that it was necessary to compensate for the group delay through the shaping

and flattening filter. Thus when a frequency response is computed, the FFT is taken $K/2$ samples from the start of the data, where K is the filter order. The response is flat to within ± 2 dB from 200 Hz to 10 kHz. Larger errors occur at lower frequencies because the system frequency response is undersampled (by the filter realization) and is not accurately represented by the small number of samples used. At high frequencies more samples are used and the match is better.

Fig. 6 shows the frequency response of the test system before (uncompensated) and after (compensated) the difference in channel frequency responses is removed (at 90-dB SPL). The uncompensated response shows the largest error at higher frequencies where the wavelength of the input signal approaches the dimensions of the microphones. The compensated response is flat to within ± 0.2 dB from 200 Hz to 10 kHz. Thus the total error (worst case) in the frequency response is less than ± 3 dB from 200 Hz to 10 kHz, which is within the ANSI S3.42 specification [11].

3.2 Hearing-Aid Test Results

To verify the operation of the test system and to examine the performance of MLS-based measurements, two prototype hearing aids, a linear aid with output limiting and an AGC aid, were tested. The MLS-based

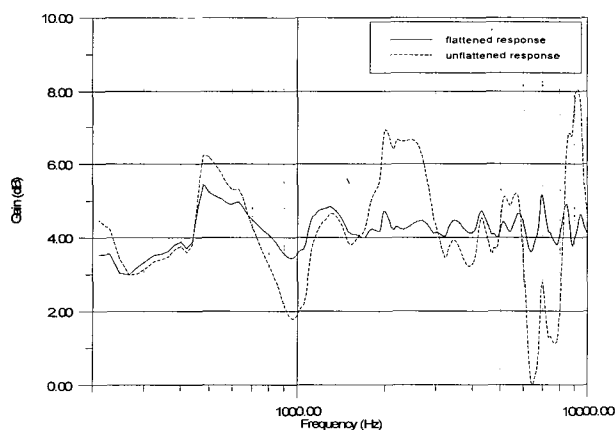


Fig. 5. Unflattened and flattened reference-channel frequency response.

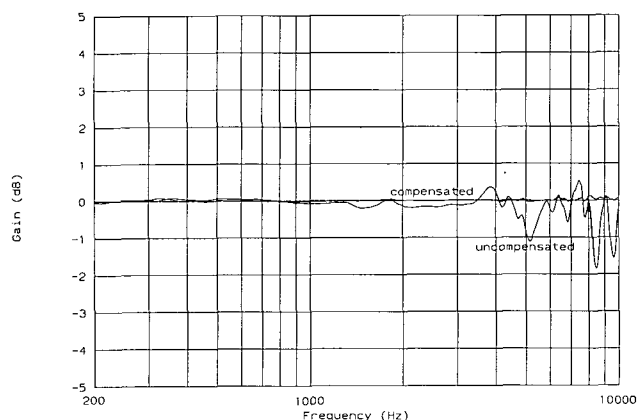


Fig. 6. Compensated and uncompensated test-system frequency response.

test results for each aid are compared to results obtained using a noise-based method on our test system. The frequency response for the noise-based method is computed as

$$H(f) = \frac{S_{xy}(f)}{S_{xx}(f)} \quad (16)$$

where $S_{xx}(f)$ is the input autospectrum and $S_{xy}(f)$ is the input-output cross spectrum. The cross-spectrum method [Eq. (16)] gives the best linear approximation to the measured response. Speech-shaped noise is used as the input stimulus. The test conditions employed by Kates [10] were used: 1.85 s of noise are recorded and analyzed by 2048-point FFTs with 50% overlapping Hanning windows. All frequency responses are shown without smoothing. The noise-based measures were calibrated to specifications that are at least equal to those of the MLS-based measurements.

3.2.1 Linear Hearing Aid

A linear hearing aid with output limiting was tested first. This aid has four controls: high tone (HT), low tone (LT), volume (VOL), and maximum power output (MPO) (that is, limiting). For the tests the HT and LT were both set to maximum. The MPO control was also set to maximum. The volume control was adjusted to RTP. The frequency response plots for 90- to 50-dB SPL input (10-dB SPL steps) are shown in Figs. 7 (MLS) and 8 (noise).

The frequency responses indicate that the hearing aid is output limiting for 80- and 90-dB SPL input levels. For these input levels the rough frequency response curves indicate that the aid is also distorting somewhat. (Although rough frequency response curves are not strictly an indication of distortion, it has been our experience that when a hearing aid exhibits smooth frequency responses at low input SPLs and the curves get progressively rougher as the input SPL increases, the roughness is a result of distortion.) For lower input SPLs (50–70-dB SPL inclusive), the aid is not limiting. The smoother frequency response curves indicate that the aid is operating with much lower distortion at these levels.

For all input SPLs the noise and MLS-based results compare favorably. Above 500 Hz, for each input SPL, all responses are within 1 dB of each other. For the 50- and 60-dB SPL input levels, the frequency responses are identical across the entire analysis bandwidth of 200 Hz to 10 kHz. For 70-dB SPL input level the MLS frequency response is more jagged than the noise response. This indicates that the MLS-based measurement is slightly more sensitive to distortion than the noise measurement. The smoothness of the frequency response measured by the cross-spectrum noise method is probably a result of the cross-spectrum method used in the frequency response calculation and the number of averages used in the measurement.

Some small differences can be observed between the noise and MLS frequency responses at low frequencies

for high input SPLs. For 90-dB SPL input level, below 300 Hz, the MLS response had approximately 2.5 dB more gain than the noise response. For 70- and 80-dB SPL input level, the MLS response has around 2.0 dB less gain than the noise response at low frequencies. These differences are probably a result of two competing phenomena. When an MLS-based method is used to analyze a nonlinear system, an increase in low-frequency gain is often observed [13]. To reduce the influence of this nonlinearity on the response, a longer MLS is used and the frequency response is computed from a truncated version of the impulse response. However, truncating the impulse response also reduces the observed low-frequency gain of the system because low-frequency impulse response components (which take longer to decay in amplitude) are eliminated from the frequency response computation. This error could be reduced by increasing the size of the FFT. However, this would increase the low-frequency error caused by the nonlinear characteristics of the hearing aid.

3.2.2 AGC Hearing Aid

The second aid tested was an AGC hearing aid. This aid has four controls: high tone (HT), low tone (LT), volume (VOL), and threshold (TK). The TK control adjusts the AGC threshold (knee point). The aid also

has an FFR switch that selects a flat frequency response or a level-dependent frequency response that provides a "treble increase at low levels" (TILL) response. For the tests, both tone controls were set to their maxima and the TK control was set at its midpoint. The FFR switch was put into the TILL position and the volume was adjusted for RTP. The frequency response plots for 90- to 50-dB SPL input level in 10-dB steps are shown in Figs. 9 (MLS) and 10 (noise).

Both sets of responses show that the hearing aid is always compressing the output signal, with less compression occurring as the input level decreases. For lower input SPLs, the TILL frequency response is clearly visible—the low-frequency gain (<300 Hz) remains relatively constant as the input level decreases, while the high-frequency gain (>1 kHz) increases. For 80- and 90-dB SPL input levels, both response measurements are jagged. Similar to the measurements for the linear hearing aid, an increasing roughness in the frequency response with input SPL indicates that the hearing aid is distorting for higher input SPLs.

For all input SPLs the curves are nearly identical. For the higher input SPLs (70–90-dB SPL inclusive) there is a small difference in the low-frequency gain. As discussed, this is probably a result of truncating the impulse response.

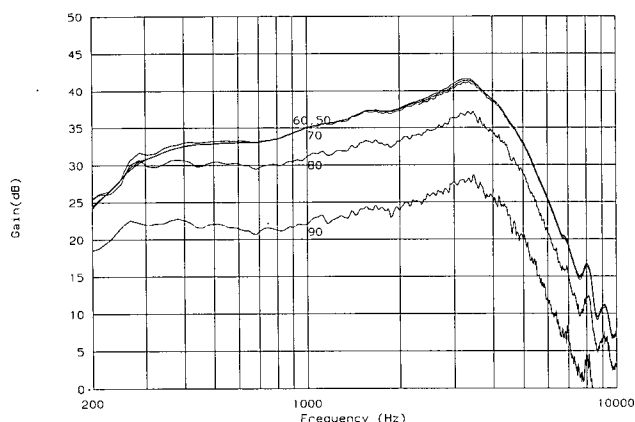


Fig. 7. Linear hearing-aid MLS frequency response (for 50–90-dB SPL input level).

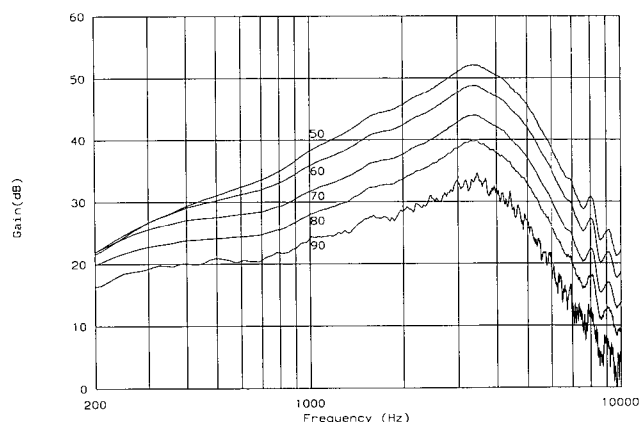


Fig. 9. AGC hearing-aid MLS frequency response (for 50–90-dB SPL input level).

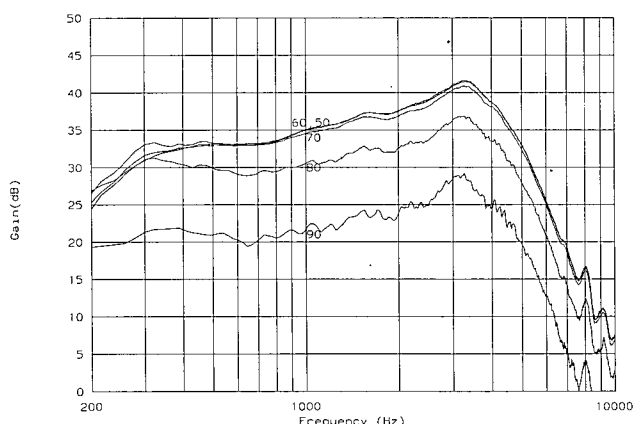


Fig. 8. Linear hearing-aid noise frequency response (for 50–90-dB SPL input level).

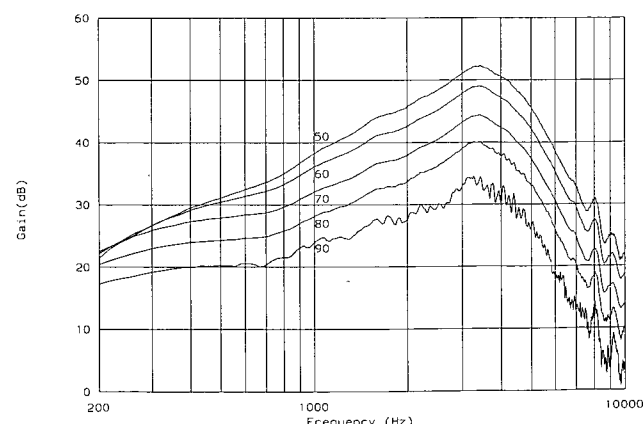


Fig. 10. AGC hearing-aid noise frequency response (for 50–90-dB SPL input level).

3.2.3 Comparison of Execution Time

A significant advantage of MLS-based over noise-based measurements is execution time. Hearing-aid measurements are often made in a clinic while the patients wait for their hearing aids. Thus faster measurements are desirable.

The execution times for computation and data collection were measured for each test presented. Each SPL step of the noise-based measurement executed in 27 s, each MLS measurement step took 9 s. Thus on our test system the MLS-based measurement is three times faster than the noise-based measurement.

This observed speed increase is a result of the reduced computation required by the MLS-based measurements. For the tests presented, the MLS measurement requires averaging eight periods (each 16 383 samples long) in the time domain, the computation of one FHT, and the computation of one windowed FFT for each channel. The noise-based measurement requires the computation of

$$2 \left[\frac{t_s \times \text{sample rate}}{\text{FFT size}} \right] - 1 \approx 71 \quad (17)$$

windowed FFTs, which are averaged to obtain a power spectrum for each channel (t_s is the analysis duration.)

By assuming that a 16 384-point FFT requires the same computation as a 16 383 point FHT [in fact, an $(N - 1)$ -point FHT requires less computation than an N -point FFT because $\sin(\)$ and $\cos(\)$ twiddle factors are not computed and no multiplications are required for an FHT], an upper bound on the computations required for a MLS measurement can be found to be less than or equal to the computations required to compute ten 2048-point FFTs. The MLS computations require no more than

$$\frac{\text{one 16 383-point FHT} + \text{one 2048-point FFT}}{\text{seventy-one 2048-point FFTs}} = \frac{10 + 1}{71} \approx 0.15 \quad (18)$$

times the computation of the noise-based method.

Thus despite the fact that the MLS-based measurement data collection takes almost twice (8×419 ms) as long as the noise-based measurement data collection (1.85 s), the MLS measurement executes in one-third the time because of the reduced computational load. The larger computational load of the noise-based measurement is a result of the large number of FFTs that are computed.

The noise-based measurements could be made faster by averaging fewer FFTs and one-third or one-tenth-octave smoothing the resulting frequency response curves to reduce the variance of the measurement. However, smoothing would reduce the frequency resolution of the measurement. More importantly, aver-

aging fewer FFTs will result in a lower SNR for the measurement. In noise-based measurements (which are statistical in nature), averaging multiple FFTs reduces the variance of the measurement and improves the SNR. If the data collection is asynchronous with the input signal, averaging will also reduce the effects of distortion on the measurement. For MLS-based measurements (which are deterministic) averaging is used only for SNR improvement. Distortion rejection is achieved by using an analysis window that is shorter than the PIR duration [13].

A further consideration is the comparison of execution times when a coherence measurement is desired. Magnitude-squared coherence (MSC) is a measurement that expresses the degree to which the output of a system is linearly related to the system input [26]:

$$|\gamma(f)|^2 = \frac{|S_{xy}(f)|^2}{S_{xx}(f)S_{yy}(f)} \quad (19)$$

where $S_{xy}(f)$ is the cross-spectral density between input and output and $S_{xx}(f)$ and $S_{yy}(f)$ are the auto-spectral densities of the input and output, respectively. MSC has received some attention as a distortion measurement for use with hearing aids [27]–[29].

With a noise-based measurement, only the auto-spectrum of the output signal must be computed to obtain the spectral densities required to compute the MSC; the input auto- and input-output cross spectra are already computed for the frequency response measurement. An MLS MSC measurement that is analogous to a noise-based MSC measurement can also be computed as the “frequency decomposed ratio of the energy in the analysis window to the energy in the PIR as a whole” [13]. Rife and Vanderkooy present a convincing argument indicating that this measurement has the same meaning as the conventional (statistical) definition of MSC. Since the energy in the analysis window is computed for the frequency response measurement, only the energy in the remainder of the PIR must be computed. Computing the PIR energy outside the analysis window requires roughly the same amount of computation needed to compute the output signal autospectrum in the noise-based coherence measurement. We are currently investigating the application of MLS MSC measurements to hearing aids.

4 SUMMARY AND CONCLUSIONS

In conventional hearing-aid test procedures, pure tones are used to measure the frequency response. When applied to hearing aids that incorporate nonlinear or frequency-selective processing, these tests can sometimes give misleading results. A new ANSI standard [11] addresses this problem by using a speech-shaped noise stimulus with FFT-based analysis. We concur with others who propose that MLS-based methods should be considered as candidates for this measurement application because of their increased speed and their flexibility.

We have presented the design and implementation of a dual-channel MLS-based test system. Our test system applies a flattened and speech-shaped acoustical MLS to a hearing aid and computes a frequency response three times faster than a noise-based measurement running on the same test system (using the test parameters employed by Kates [10]).

Test results for two hearing aids, a single-channel AGC aid and a linear aid with output limiting, were presented. Each hearing aid was tested using both MLS-based and noise-based methods. Aside from small low-frequency errors, which occurred only at high-input SPLs, the responses measured by the two methods were identical. These low-frequency errors are probably caused by truncation of the impulse response and the nonlinear characteristics of the hearing aids at high-input SPLs. Unlike some broad-band test schemes, MLS-based measurements do not break down when measuring hearing aids that distort the output signal.

Using MLS-based measurements, the frequency response of a hearing aid to a frequency-shaped broad-band input stimulus can be computed rapidly. If the frequency response is computed from a truncated version of the impulse response, distortion and noise are rejected in a manner similar to that of the cross-spectrum method used for noise-based testing. For tests conducted in a noisy environment, MLS-based measurements provide good rejection of transient noises. The SNR can be improved by averaging MLS periods in the time domain before the FHT is computed.

An algorithm for rapidly filtering a MLS with a high-order filter (order $\leq$ MLS length) also exists. Using this method, arbitrarily frequency-shaped MLSs can be computed quickly and applied to hearing aids. This procedure was used in our test system to flatten and speech-shape the MLS applied to the hearing aid under test. A filtered MLS (through any nontrivial filter) has a crest factor that is approximately 12 dB, the crest factor specified by ANSI S3.42.

Recent work by Vanderkooy [19] on the characteristics of MLS measurements, where the system under test exhibits various types of distortion, indicate that most forms of high-order distortion are uniformly distributed throughout the PIR. Distortion introduced by slowly time-varying systems is also uniformly distributed throughout the PIR. This result is important for the measurement of AGC hearing aids whose gain may vary slightly in response to a broad-band input signal. Because of the time-domain separation of linear and nonlinear components in the PIR, the influence of these forms of distortion on measured frequency responses can be reduced by truncating the impulse response.

Low-order distortions, such as slew limiting and crossover distortion, lead to a nonuniform distortion "finger print" in the PIR that depends on the MLS that is used and the device under test. Vanderkooy suggests that by using different but equal-order MLSs, distortion introduced by these phenomena can be "identified and removed through appropriate editing." The distortion rejection provided by MLS-based measurements is

certainly a desirable feature that may prove to be an interesting topic for future study.

In conclusion, we feel that dual-channel MLS-based measurements show promise for further application in hearing-aid measurement. Since time-domain averaging improves the SNR of an MLS measurement, it is possible to make frequency response measurements with a low-level MLS while biasing the hearing aid with a broad-band test signal (provided that the impulse responses are coherent and the bias signal is incoherent with the MLS). Also, MLS-based coherence measurements can be applied to hearing aids for distortion measurement. We are currently investigating these and other applications of MLS-based measurements for hearing-aid characterization.

5 ACKNOWLEDGMENT

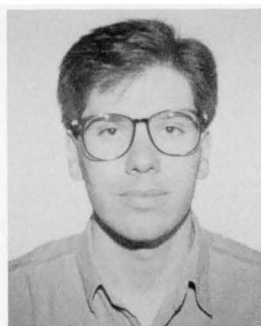
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THE AUTHORS



T. Schneider

Todd Schneider received a B.A.Sc. (1989) and a M.A.Sc. (1991) both in electrical engineering from the University of Waterloo in Waterloo, Ontario, Canada. He has been employed as a research engineer at the Hearing Health Care Research Unit at the University of Western Ontario in London, Ontario, since mid-1991. He is currently engaged in research on improved hearing aid test methods and real-time digital signal processing. Mr. Schneider is a member of the IEEE and AES.



D. Jamieson

Don Jamieson is Professor of Communicative Disorders, Otolaryngology and Psychology at the University of Western Ontario, London, Ontario, Canada, and Director of Ontario's Hearing Health Care Research Unit (HHCRU). He also leads the Hearing Team of the Ontario Rehabilitation Technology Consortium. His research program focuses on speech perception, hearing, and rehabilitation for these domains. This work includes directing the CSRE (Canadian Speech Research Environment) Project, which created the general-pur-

pose PC software system for speech editing, analysis, synthesis, and testing, now being used by researchers in 12 countries.

Dr. Jamieson received a B.A. in developmental psychology from the University of British Columbia in 1972, and M.Sc. and Ph.D. degrees in Experimental and Mathematical Psychology from Carleton University in 1974 and 1976, respectively. He was a faculty mem-

ber at the University of Calgary, Alberta, Canada, from 1976 to 1987, serving as Head of the Department of Psychology there from 1981 to 1985. From 1985 to 1987 he was a visiting fellow at the Waisman Centre, University of Wisconsin (Madison) in the U.S. He moved to the University of Western Ontario in 1987, chairing the Department of Communicative Disorders until 1992.