Efficient Filter Design for Loudspeaker Equalization

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The advent of digital storage of audio signals and the availability of high-speed digital signal processing devices facilitate the implementation of high-order filter functions for loudspeaker equalization. A method is presented for generating a digital model of a loudspeaker system from which an efficient compensation filter using an IIR structure is derived. Application of this technique allows the relative attributes of FIR and IIR structures to be distinguished and an equalization filter to be realized which provides simultaneous magnitude and phase equalization.

0 INTRODUCTION

From the conception of the Compact Disc (CD) player in 1982, digital signal processing (DSP) has had an ever-increasing role to play in audio systems. The digital serial output port that is now commonplace on CD players greatly facilitates the application of DSP to the many processes that take place in audio systems and, indeed, permits additional processing, which would otherwise be costly or impractical with analog techniques.

Loudspeaker systems, which are the most nonideal components of an audio system, potentially stand to benefit the most from the powerful processing capabilities available with modern digital signal processors. Crossover networks, traditionally, have been the source of both practical and theoretical consternation. For some of the reasons now discussed, these would appear to be a good starting point on the road to loudspeaker system improvement. A problem encountered with analog crossover networks arises from the sensitivity of the loudspeaker response to component tolerances. Digitizing the crossover, therefore, reduces these effects, as the component values are now simply coefficients in a digital signal processor. (Tolerances are now in the form of quantized coefficients.) Another undesirable feature inherent in analog crossovers, discussed by Lipshitz and Vanderkooy [1], is the tradeoff between overall phase linearity and polar response. Bews [2], in a Ph.D. thesis, proposes the use of FIR filters for crossover networks. The linear-phase property obtainable with FIR filters enables the high- and low-pass filters to remain in phase, with the combined output also being linear phase. Hence improved lobing error and transient response are achieved. A further advantage of digital crossovers, discussed in [2], is the ability to incorporate drive unit compensation which, similarly, improves the polar and transient responses of the system.

These advantages may well result in the use of digital crossovers in the high end of the audio market or studio applications. Digital filters, however, have to be part of an active loudspeaker system which requires separate power amplifiers for each of the drivers. This aspect will probably price digital crossover systems outside much of the domestic market. An alternative strategy, therefore, is to use a passive loudspeaker system in conjunction with an outboard digital equalizer. The availability of digital outputs from CD players and outboard digital-to-analog converters (DACs) makes the concept of a standard digital equalizer structure, with coefficients matched to commercially available loudspeaker systems, a most versatile and attractive proposition. An outboard equalizer further offers the ability to compensate, at least in part, for cabinet errors such as diffraction at the cabinet edges and internal reflections from within the cabinet. Thus a complete compensation system, taking into account crossover, driver, and cabinet errors, is feasible. Being flexible in nature, the specific equalization task can be tailored either to the loudspeaker systems or to their environment. For ex-

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ample, one may choose to ignore the more directional aberrations, giving a greater "depth of field" to the equalization, though a less focused one.

Owing to the potential of digital equalizers in acoustic applications, the subject has recently received much attention, a sample of which is given in [3]–[5]. This paper proposes an efficient method of designing the equalizer. Sec. 1 briefly considers some of the issues surrounding loudspeaker equalization, such as what form of equalization provides the best perceived performance. Sec. 2 is a brief discourse on current equalizer designs and surmises that the total reliance on FIR filters places a heavy demand on existing digital signal processors. This could be alleviated with the use of IIR filters. In Sec. 3, an equalizer system derived from an IIR model of a loudspeaker system is described and offered as a more efficient solution to the equalizer problem. Sec. 4 gives applications of this design strategy, including a discussion on its use in the evaluation of the subjective importance of phase distortion.

1 EQUALIZER DESIGN CONSIDERATIONS

Two important considerations need to be taken into account before design is commenced, 1) what is the desired loudspeaker system response? and 2) what needs to be equalized?

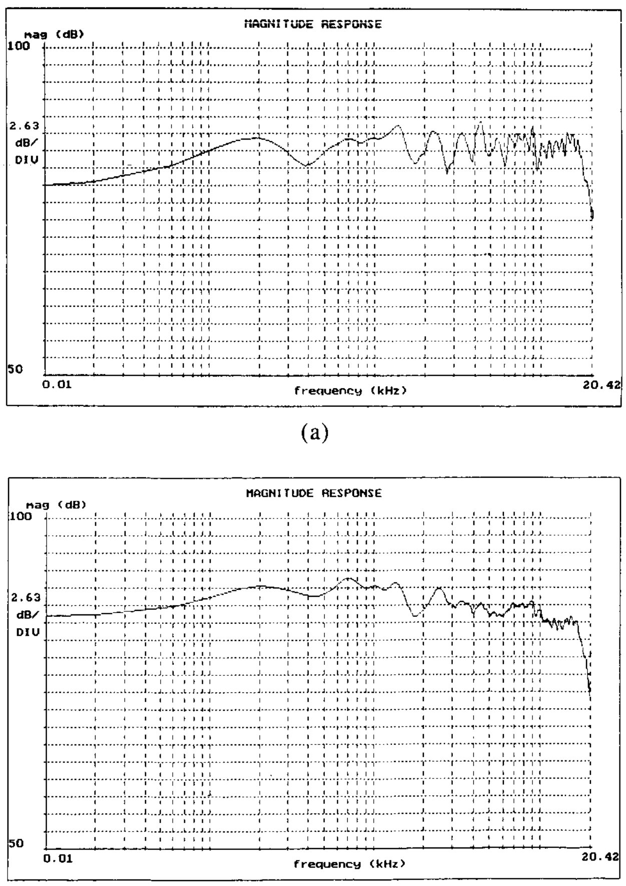
As well as being of fundamental concern to both loudspeaker and equalizer design, these two points touch on precarious psychoacoustic issues. The subject of this paper is aimed purely at equalizing an individual loudspeaker system's response; no attempt will be made to equalize for room effects. The concept of a perfect loudspeaker system is a matter for debate, but for the purpose of this paper, an ideal loudspeaker system will be considered as having flat magnitude and phase responses, which remain constant over all forward space (that is, a flat polar response). This corresponds to a system that instantly produces an impulse over the entire listening space (when driven by an impulse), which is an unrealistic aiming point for an equalization scheme for two reasons.

1) There must be some delay, as physical delays are incurred in the signal transmission path which cannot be equalized. (It is not possible to generate negative time.)

2) The polar response is a function of multiway crossover networks and the physical construction of the loudspeaker system. Hence no amount of equalization can achieve this requirement in a simple stereo system.

A better approach is to aim for a flat-magnitude response with a pure delay (linear phase) over a finite listening space (or a function thereof). With this concept of a realistic loudspeaker system's response, the practical target requirements of the equalizer need to be clarified.

First, consideration must be given to the measurements on which the equalizer should be based. The on-axis measurement appears to be an obvious choice. But bearing in mind the requirement for equalization over a listening space, the value of correcting small aberrations that occur on the on-axis response and do not occur on the off-axis responses seems suspect. Indeed, this form of equalization may prove detrimental to off-axis responses. An alternative strategy is to use an averaged measurement formed from a number of responses taken over the finite listening space. Fig. 1 shows the responses of a three-way loudspeaker system measured on axis and averaged over a a number of points in space. The averaged response is formed from nine responses measured on a grid spanning ± 15° vertically and ± 30° horizontally. (All measurements are taken at 1 m from the assumed acoustic center of the loudspeaker.) Note that both of the responses shown are, effectively, the anechoic response of the loudspeaker, as they are derived from an impulse measurement which has been truncated prior to the first room reflection. A number of interesting features are revealed by a comparison of the two plots. First, the average measurement is significantly smoother than the axial response. This indicates that a number of the ripples in the magnitude response are directional (probably caused by diffraction effects), and therefore do not want specific equalization. If the loudspeaker system response is assumed to be of finite order (that is, it can be modeled by a polynomial of finite order),



(a)

(b)

Fig. 1. Measured loudspeaker magnitude responses. (a) On-axis response. (b) Spatially averaged response.

a smoothing function in the frequency domain can be represented as an effective reduction in the order of the loudspeaker system response. Thus a practical advantage is gained with the averaged response, as the equalizer based on the inverse response will be correspondingly of lower order. Referring back to Fig. 1, the averaged response exhibits a downward tilting response relative to the axial measurement. The downward tilt is due to the increasingly directional higher frequency response, a trait common to most loudspeaker systems. An equalizer based on the averaged measurement is therefore likely to produce a perceived high-frequency emphasis effect that will be audibly objectionable.

It is the authors' belief that a better solution involves some form of compromise between the axial- and averaged-response-based equalization schemes. For example, the average is formed from a set of weighted measurements, that is, the on-axis measurement is normalized to unity, with a reduction in weight as the angle off axis is increased. Alternatively, an averaged measurement can be used with compensation for the high-frequency rolloff. These ideas are largely conjecture at the moment, and before any definitive conclusions can be made on this subject, more work is required on the subjective effects of the various equalization schemes. It may be that different loudspeaker systems require differing equalization strategies. For example, the response exhibited by asymmetrical constructions would not necessarily benefit from the same equalization strategy as that required by symmetrical constructions.

A point regarding low-frequency equalization is worth mentioning at this stage. Unless the loudspeaker system in question is specifically designed to accept bass extension, its practice is not recommended from either digital-implementation or high-fidelity contexts. From the digital perspective, a boost in magnitude of any frequency band demands an increase in signal dynamic range at the output of the digital filter. As all current digital systems work on a fixed-word-length architecture (even if the digital processor has floating-point capability, the signal ultimately has to be presented to a finite-precision DAC), a gain boost in one frequency band actually translates to attenuation in all other bands. Consequently the digital filter will exhibit a reduced dynamic range and possibly suffer from other undesirable distortion artifacts caused by the signal requantization. From a mechanical perspective, bass extension will extend the diaphragm displacement, overdriving the loudspeaker and causing excessive nonlinear distortion. Because of these considerations, low-frequency equalization will not generally be considered a function of the digital equalizer. For the equalizer design process used in this paper, in order to prevent the optimization scheme from attempting low-frequency equalization, the signal presented to the algorithm must be preequalized. This is achieved by deriving an equalizer based on the low-frequency electroacoustic model of a loudspeaker system. Such equalization techniques are discussed elsewhere (see, for example, Greiner [6]), and therefore will not be considered further here. Note that this equalization is performed in software at a "postprocessing" stage, and not in hardware, prior to the loudspeaker measurement. Similar consideration must be given to the high-frequency rolloff of the loudspeaker. Fortunately in non-oversampled systems, there is usually no need to pay specific attention to this point, as the response of the digital filter is bounded by the Nyquist frequency. This is generally below the frequency at which tweeters begin to fall off.

2 CURRENT DIGITAL EQUALIZATION SCHEMES

The use of FIR filters in digital equalization schemes seems a logical choice. Briefly, functions with arbitrary magnitude and phase are readily generated. This is particularly useful for equalization schemes where the equalizer is determined by an impulse response derived from the complementary frequency response of the loudspeaker system (taking into account the low-frequency equalization). Further, FIR filters are well suited to optimization techniques, giving simplified design tools and also the possibility of adaptive equalizers. For a more enlightening description of FIR filter design and implementation, the reader is referred to one of the numerous books on the subject, such as [7], [8].

Jensen [3] adopts a deterministic approach where the system is made up of three filters, each covering a fraction of the audio band. The lower frequency band filters operate on subsampled data which, after filtering, are then oversampled back to the original sampling frequency and combined with the higher frequency bands to form the broad-band equalizer. The decimation and interpolation process is necessary to provide sufficient resolution of equalization over the entire frequency range, which cannot be achieved with a simple system using existing digital signal processors, as the number of coefficients is prohibitive.

A different approach, adopted by Mourjopoulos [4], for example, uses optimization procedures such as the least mean squares (LMS) algorithm. In [4], Mourjopoulos describes the equalization of a system response as a deconvolution process. Here the error function, used to optimally generate the equalizer coefficients, is formed from the difference between the desired output and the convolution of an FIR filter (the equalizer) with the system's response. Typically, in acoustic applications, direct implementation of this method is confined to low-sampling-rate systems or otherwise restricted bandwidth systems. This is because the number of coefficients required to equalize low-frequency anomalies or high-Q resonances is too great for existing digital signal processors.

The limiting factor in all FIR-based equalizers is that the amount of processing involved is directly related to the impulse duration required by the equalization task. The application of IIR filters is therefore attractive where the extended impulse response possible is better suited to the equalization requirements. The following section describes an IIR approach to the equalization problem.

3 EQUALIZER DESIGN

Although it may be possible to design an IIR equalizer analytically, the complicated nature of both the filter type and the functions to be equalized makes this approach impractical. An optimization approach, therefore, appears preferable and is the method used here. The mixed-phase nature exhibited by most loudspeaker systems creates severe problems for IIR equalization schemes. The presence of zeros in the right-hand side of the s plane, in mixed-phase signals, leads to an acausal equalizer response. This implies that any "true" IIR equalizer would be unstable. (It would have poles in the right-hand side of the s plane.) It is therefore not possible to use the direct deconvolution approach adopted in [4], [5]. Thus an alternative strategy is needed.

As the response of all loudspeaker systems (one would hope) is both causal and stable, another strategy is to form a model of the loudspeaker response. Using the coefficients obtained from the model, the minimum- and excess-phase components can be isolated and dealt with separately. (The excess-phase component is the objectionable factor, demanding the acausal equalizer response.) The design of an equalizer using this approach requires four steps:

1) Loudspeaker system modeling using an IIR filter
2) Separation of minimum- and excess-phase components
3) Formation of minimum-phase equalizer (magnitude equalizer)
4) Formation of excess-phase equalizer (all-pass equalizer).

3.1 Loudspeaker System Modeling

This section describes a technique which generates coefficients of an IIR model using the LMS algorithm. The LMS algorithm is particularly well suited to FIR structures (tapped delay lines) where the correlation cancellation loop (CCL), described by Morgan and Craig [9], is readily employed. The stability complications encountered in IIR filters generally require more advanced algorithms, such as SHARF by Larimore et al. [10]. However, if, as in the case of system modeling, the system, and therefore the model also, is driven only by an impulse, the modeling process has a number of simplifications. The process can now be decomposed into two parts: the time period when the feedforward coefficients have no effect on the model output, and the period when both feedback and feedforward coefficients contribute to the output. Thus simpler optimization, combined with deterministic approaches, can be used to find the feedback and feedforward coefficients, respectively. This aspect is illustrated by the following.

Consider the direct from 1 IIR structure of Fig. 2. The output $y(n)$, when the filter is driven by $x(n)$, is

$$y(n) = \sum_{i=0}^{N} x(n-i)b_i + \sum_{j=1}^{M} y(n-j)a_j \ . \tag{1}$$

If the filter is driven by $\delta(n)$,

$$y(n) = \sum_{i=0}^{N} \delta(n-i)b_i + \sum_{j=1}^{M} y(n-j)a_j \ . \tag{2}$$

For $n > N$,

$$\sum_{i=0}^{N} \delta(n-i)b_i = 0 \ . \tag{3}$$

Hence the output of the filter will be

$$y(n) = \sum_{j=1}^{M} y(n-j)a_j \ . \tag{4}$$

If $a_0 = 0$ is assumed, then

$$y(n) = \sum_{j=0}^{M} y(n-j)a_j \tag{5}$$

which is the response of an FIR filter excited by its own previous outputs. Thus it would appear that the feedback coefficients can be found using standard Wiener estimation techniques. The function to be minimized, giving the optimum coefficient set, in the LMS sense, is

$$I = \sum_{n=N+1}^{L-1} e^2(n) = \sum_{n=N+1}^{L-1} [y(n) - \hat{y}(n)]^2 \tag{6}$$

$$I = \sum_{n=N+1}^{L-1} \left[y(n) - \sum_{j=1}^{M} \hat{y}(n-j)a_j \right]^2 \tag{7}$$

where $y(n)$ is the system response, $\hat{y}$ is the estimated

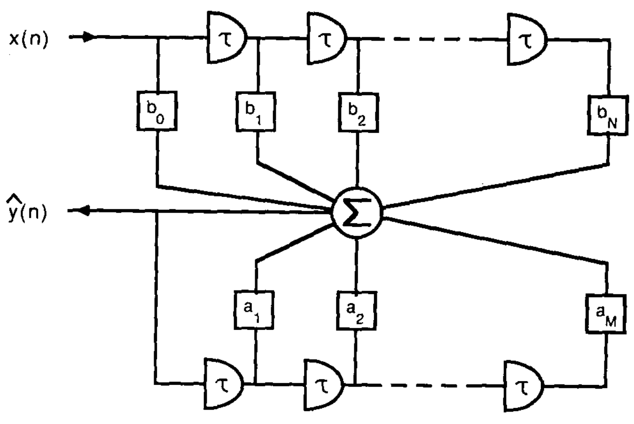


Fig. 2. Direct form 1 IIR filter structure.

response, and $e(n)$ is the error between the desired and the estimated responses. L is the length of the impulse response over which the optimization is being performed. Note that the function given in Eq. (7) differs from the standard LMS problem, as the signal driving the filter is not known a priori. Therefore a direct solution to the normal equations[1] is not possible. Instead, the approach used to minimize the function in Eq. (7) uses a gradient search (or steepest descent) algorithm. The gradient search method attempts to approximate the gradient of the error function by iteratively searching along the line of steepest descent (given by an estimate of the current error gradient function). The filter coefficient values are found by iteratively evaluating the function

$$A(n + 1) = A(n) - \mu \frac{\partial e^2(n)}{\partial A} \tag{8}$$

where A denotes the coefficient vector and μ is a damping factor. Substituting the error function $e^2(n)$ from Eq. (7) into Eq. (8) leads to the simple coefficient update formula

$$a_j(n + 1) = a_j(n) + 2\mu e(n)\hat{y}(n - j) \tag{9}$$

from which we see that no a priori knowledge is required of the filter input, as only the current error value and previous filter outputs are required to form an estimate of the error gradient.

At this stage it is worth discussing a couple of points about the algorithm. It is possible to find a direct solution (as opposed to the iterative gradient search method) by making a slight modification to Eq. (7). If the modeling filter is driven by the known system response, instead of the estimated response, the optimization becomes the standard solution of the LMS problem. This will indeed, give an optimum solution over the period for which it has been optimized. However, there is no control of the response of the filter beyond this point. This procedure has been attempted, and has been found to work satisfactorily, in cases where the system response is well behaved and of finite order. When applied to acoustic measurements, the modeling algorithm nearly always produced an unstable filter. We return now to the gradient search algorithm given in this paper. The coefficient updates are made based on an estimate of the error gradient. If, at one point, an estimate is generated which produces an unstable filter, the error term will begin to diverge. This divergence effectively indicates that the last update was not made in the direction of steepest descent and, therefore, will force a reversal in the search direction, bringing the model back into a stable condition. Thus provided the initial coefficients give a stable filter (generally the feedback coefficients are set to zero), the model should remain in a stable condition. If the modeling process does continue to diverge, a smaller damping factor μ should be tried. To the authors' knowledge, there is no way to optimally determine the damping factor, as the input to the modeling filter is not known a priori. To date trial-and-error methods (and some educated guesses gained with experience) have been used to determine a suitable damping factor. It should be noted that the smaller the damping factor, the more accurate the final model will be. As this is not a real-time application, the reduction in the algorithm's convergence rate is not a crucial factor, while its stability is. Note also that the model should be optimized over a time period where the training signal is clearly converging. This will pull the model into a similarly convergent pattern.

Let us now determine the feedforward coefficients. Having found the feedback coefficients, a simple method for obtaining the feedforward coefficients is as follows:

Consider time $n = 0$

$$y(0) = b_0\delta(0)$$

$$>> b_0 = y(0)$$

then at time $n = 1$

$$y(1) = b_1\delta(n) + a_1y(0)$$

$$>> b_1 = y(1) - a_1y(0)$$

and so on, until time = N,

$$b_N = y(N) - a_1y(N - 1) - a_2y(N - 2) - \cdots - a_Ny(0) \; . \tag{10}$$

Eq. (10) is a general equation, which is used to find all of the feedforward coefficients from a knowledge of the impulse response and the previously found feedback coefficients.

The procedures described for finding the feedforward and feedback coefficients are used in a program which calculates a specified-order model from a given system impulse response. Fig. 3(a) shows the frequency and time domain responses of a commercial loudspeaker system and Fig. 3(b) shows the corresponding responses of a 40th-order model. The order of the model is arbitrary. However, the accuracy of the model is directly related to its order, which is seen in both the residual mean squared error and the response plots. Fig. 3(c) shows the responses of a 55th-order model, where the improved accuracy is readily observed. An interesting feature of this modeling procedure is that, although it is optimized in the time domain, the model tends to lock onto the most dominant resonances that are apparent in the frequency domain. This is observed in Fig. 3(a) and (b), where the lower order model follows a smoothed version of the loudspeaker response. The higher order model bears the same general characteristic, but has

[1] The details of LMS techniques will not be covered here. For the interested reader there are a number of texts available on the subject. See, for example, [11].

also managed to pick up some finer detail in the frequency response. An explanation for this phenomenon comes from noting that the response of an IIR filter can be formed from the summation of geometric series. Consider the frequency response of an IIR filter, given by the z transform of Eq. (2),

$$H(z) = \frac{\sum_{i=0}^{N} b_i z^{-i}}{1 - \sum_{j=1}^{M} a_j z^{-j}} . \tag{11}$$

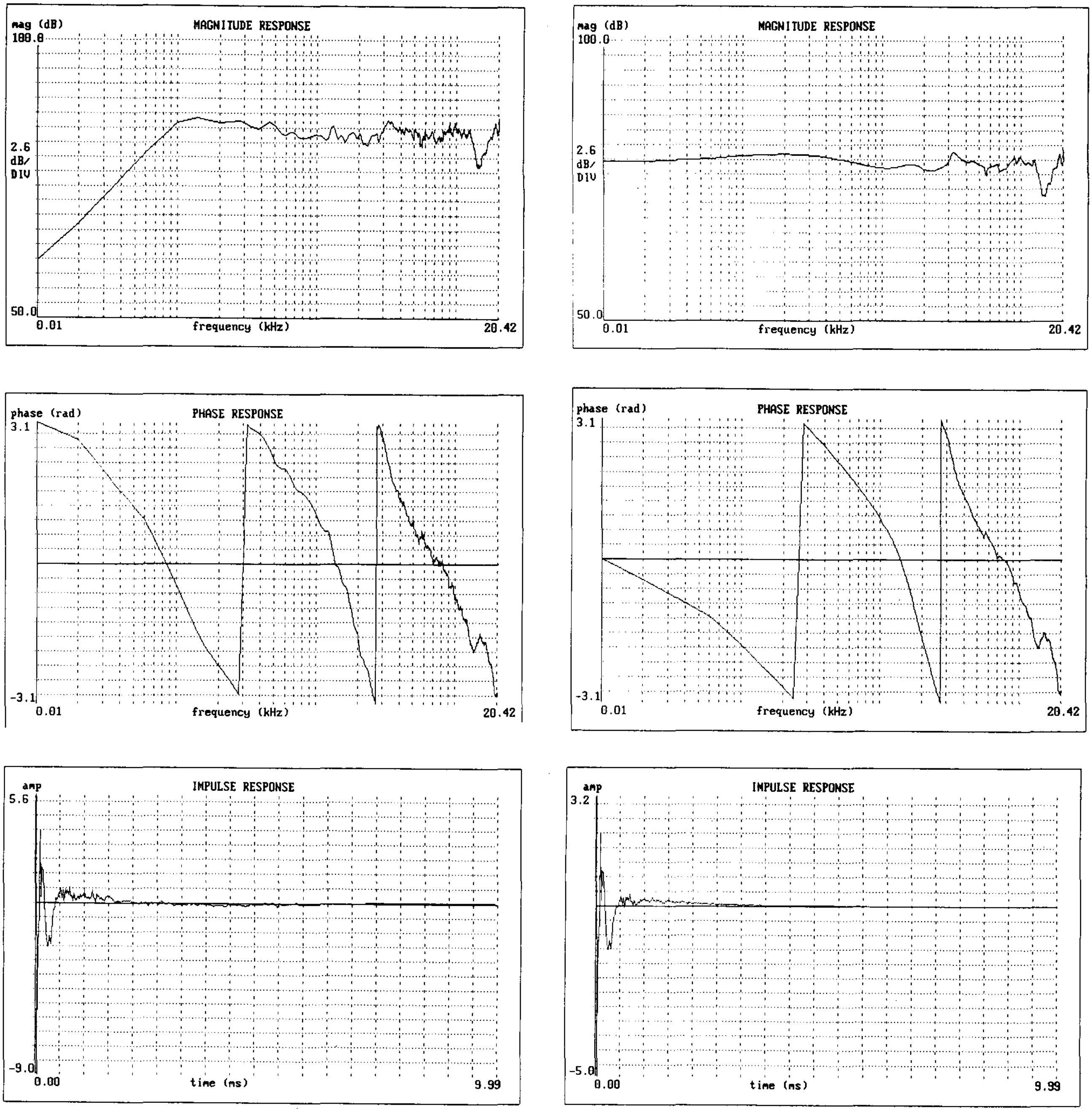
This can be split into its partial fraction expanson (for the simple case of no repetitive poles), giving

$$H(z) = z \sum_{i=1}^{M} \frac{R_i}{z - \alpha_i} + R_{M+1} \tag{12}$$

where α_i are the filter poles and R_i are the residues found from the filter zeros. Eq. (12) transforms into the summation of geometric series in the time domain, giving

$$h(n) = \sum_{i=1}^{M} R_i \alpha_i^n + R_{M+1} \delta(n) . \tag{13}$$

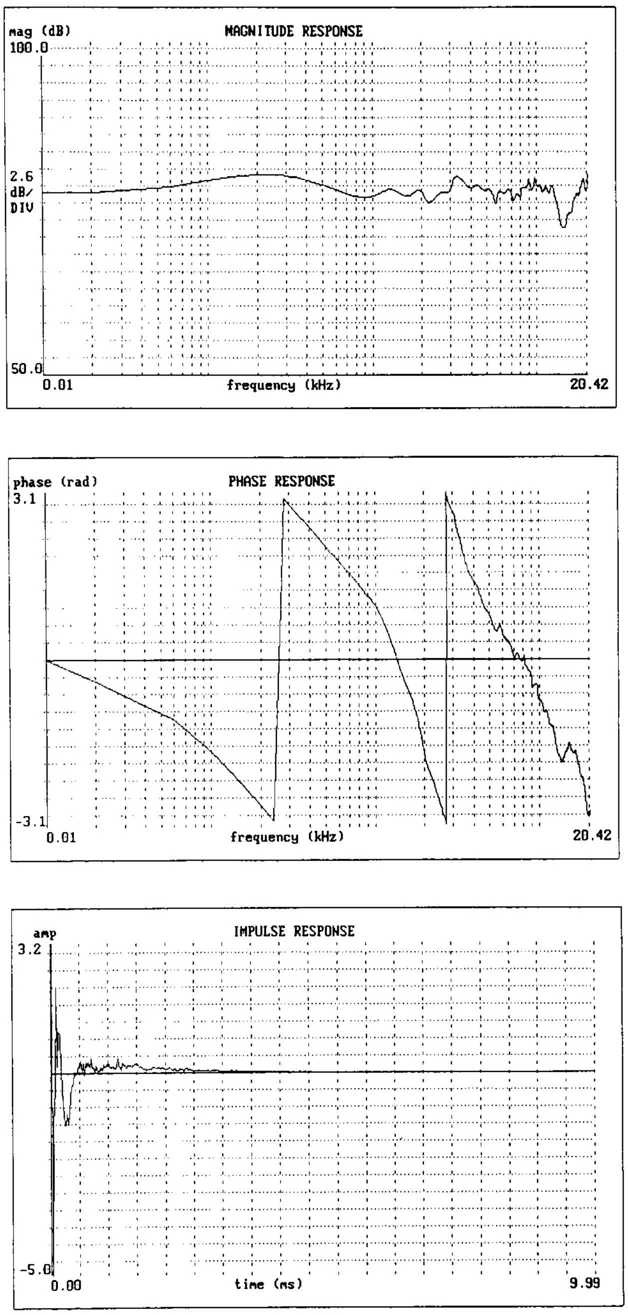
Eq. (13) shows how the pole locations in the frequency



(a) (b)

Fig. 3. Loudspeaker system responses. (a) Measured. (b) 40th-order model. (c) 55th-order model.

domain are directly related to the time domain response of the IIR filter. The correlation cancellation loop [Eq. (9)] works by locking onto correlations within the signal. It is therefore likely that the most dominant (or high-Q) resonances will be the first to be picked up by the modeling process. This accounts for the lower order model's ability to extract the main features of the loudspeaker response and discard the finer details first. While this aspect of the modeling process is in most cases desirable, from a subjective perspective, one must be aware that all-pass responses are also capable of exhibiting very high Q resonances. All-pass responses, if not inaudible, are certainly less audibly significant than most amplitude anomalies. Yet, the model will lock onto these resonances in preference to weaker amplitude resonances.



(c)

Fig. 3. continued.

3.2 Separation of Minimum- and Excess-Phase Components

Separation of the minimum and excess-phase components requires the factorization of a z-domain polynomial into its poles and zeros. In the z domain a mixed-phase function is one in which zeros are present outside the unit circle (excess-phase zeros). For example, Fig. 4(a) shows a pole–zero plot of a typical mixed-phase function in the z domain. A mixed-phase function can be represented by all-pass and minimum-phase functions in cascade, as follows:

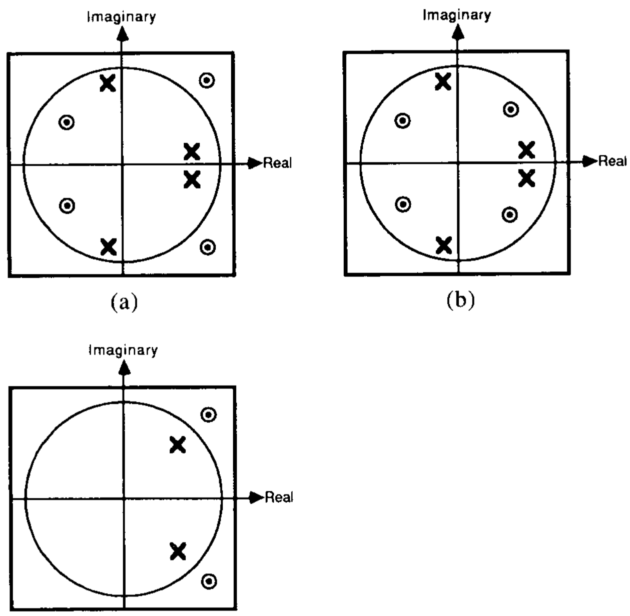
$$H(z) = \frac{(z - n_{\mathrm{m}})(z - n_{\mathrm{e}})}{(z - d_1)(z - d_2)}$$

$$H(z) = \frac{(z - n_{\mathrm{m}})(z n_{\mathrm{e}}^{*} - 1)}{(z - d_1)(z - d_2)} \frac{z - n_{\mathrm{e}}}{z n_{\mathrm{e}}^{*} - 1} \tag{14}$$

$$H(z) = H_{\min}(z) H_{\mathrm{ap}}(z)$$

where n_{m} and n_{e} denote roots inside and outside the unit circle, respectively, and * denotes the complex conjugate.

In essence, the minimum-phase function is formed simply by replacing the excess-phase zeros by their reflection about the unit circle. Similarly, the all-pass function is formed from the excess-phase zeros and poles, which are the excess-phase zeros reflected about the unit circle. To demonstrate this graphically, the pole–zero plots of the minimum-phase and all-pass functions are given in Fig. 4(b) and (c), respectively.



(c)

Fig. 4. Pole and zero plots. (a) Fourth-order mixed-phase function. (b) Corresponding fourth-order minimum-phase function. (c) Corresponding second-order all-pass function.

From Eq. (14) the minimum-phase function contains all its zeros inside the unit circle and is therefore invertible (the criterion for stability in the z domain being that all poles must be contained within the unit circle), while the all-pass function contains all its zeros outside the unit circle and is therefore noninvertible. Now an all-pass function, as its name suggests, passes all frequencies with unity gain. Therefore the minimum-phase function contains all the magnitude characteristics of the loudspeaker system response. Magnitude and minimum-phase equalization can, therefore, be performed by the inverse minimum-phase function, while equalization of the resulting all-pass response requires an additional equalizer based on the all-pass function.

3.3 Formation of Minimum- and Excess-Phase Equalizers

The minimum-phase equalizer is the reciprocal of the minimum-phase function found in the Sec. 3.2. The polynomial in the z domain is converted into the filter coefficients using the relationship observed between Eqs. (1) and (11). This produces an IIR filter capable of equalizing all magnitude effects over the entire frequency range. Application of this type of equalizer is discussed in the following section.

The excess-phase equalization is considerably more problematical. As was shown in preceding sections, an IIR equalizer based directly on the inverse function is not possible. Looking at the problem from a different perspective, the all-pass characteristic is a result of pure phase or delay effects, which can be equalized with the time-reversed version of the phase or delay effects. Thus,

$$e^{j\omega\sigma}e^{-j\omega\sigma} = 1 \qquad (15)$$

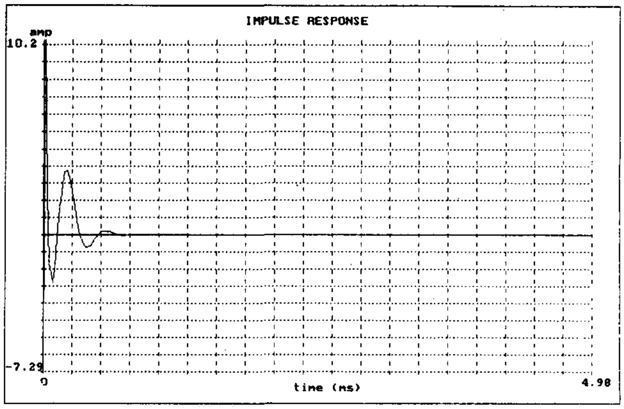
where ω is the frequency variable and σ is the frequency-dependent phase function. Although negative time is unrealizable, it is possible to introduce an overall delay. Hence the phase equalization is affected as follows:

$$e^{j\omega\sigma}e^{j\omega(\tau-\sigma)} = e^{j\omega\tau} \; .$$

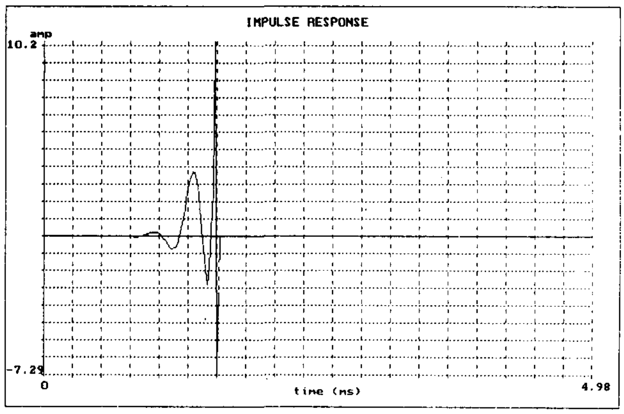
which is a pure delay or linear phase shift. Note that the time delay τ must be chosen such that $\tau - \sigma > 0$ for all ω.

Although approximations to a pure delay can be made with IIR filters, it seems more expedient to make use of FIR techniques where delays are inherent of the structure. Thus the phase equalizer is formed from the time-reversed, time-shifted, and time-windowed impulse response. The all-pass impulse response is derived from the all-pass function found in Sec. 3.2. Fig. 5(a) gives an example of an all-pass impulse response and Fig. 5(b) gives the corresponding impulse used to form the equalizer.

As an aside, a useful feature of the filter derivation process is that it allows a limited amount of control of the impulse duration requirements that are demanded by the excess-phase equalizer. A major concern with all FIR filters is the duration of the impulse response. Typically, some form of time window is applied to the impulse response in order to constrain its duration. Now, applying a time window to an all-pass response will result in the response no longer being all pass. Thus unless the filter is of sufficient length to completely accommodate the mixed-phase equalizer, only minimum-phase equalization should be attempted. This results in an all-pass response from the equalized system, which certainly is preferable to the amplitude distortions that would have been incurred had the excess-phase equalizer impulse response been prematurely truncated. Referring back to Eq. (14), we note that the all-pass function is formed from a set of matched pole–zero pairs. Elimination of any pole–zero pairs, from the set of pairs, will not affect the amplitude response (that is, an all-pass filter remains all pass). Using this property, one can discard any slowly decaying all-pass resonances from the all-pass function and use the remaining pairs to equalize the remaining phase distortion. In loudspeaker systems, the all-pass resonances that are likely to cause a problem will be with low-frequency crossovers. If the pole–zero pairs associated with this crossover region are discarded, in most cases an FIR filter of sufficient length can be derived to deal with the not so demanding, but more significant midband crossover distortion. To demonstrate the possibilities of this procedure, Fig. 6(a) shows a fifth-order all-pass



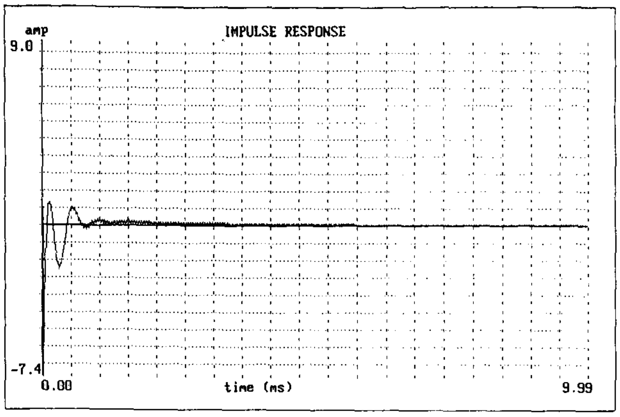
(a)



(b)

Fig. 5. Impulse responses. (a) All-pass function. (b) Corresponding equalizer function.

response obtained from a measured loudspeaker. Along with the all-pass response caused by a crossover network centered at 2.5 kHz, the all-pass function contains two slowly decaying resonances (one at approximately 100 Hz and the other at 20 kHz). The pole–zero pairs associated with these two troublesome resonances are removed, giving the responses shown in Fig. 6(b). The duration of the impulse is now reduced considerably. Using the reduced-order function to derive a 70th-order FIR equalization filter gives the quasi-excess-phase equalized response shown in Fig. 6(c). The midband phase distortion has been removed almost entirely



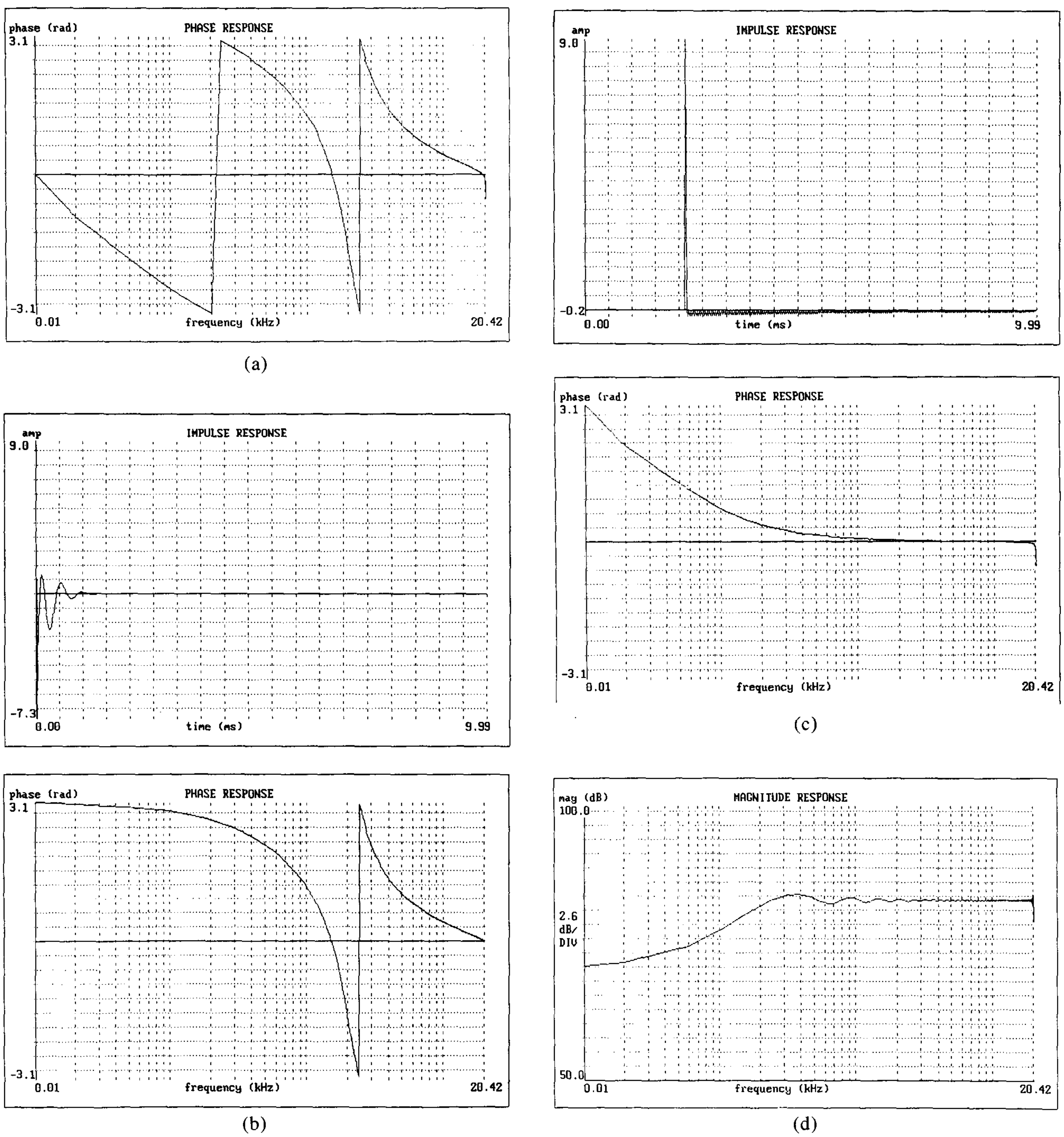


Fig. 6. Impulse and phase responses. (a) Fifth-order all-pass filter. (b) Third-order all-pass filter derived from (a). (c) Fifth-order all-pass network after equalization by reduced-order all-pass filter. (d) Magnitude response of all-pass equalizer derived from original fifth-order all-pass function truncated at 70 samples.

without compromising the amplitude response. For reflection, Fig. 6(d) shows the magnitude response of a 70th-order excess-phase equalizer derived from the original fifth-order all-pass function using a rectangular window. Note the severe amplitude distortion now incurred.

This completes the description of loudspeaker equalizer design. The remainder of this paper discusses various applications where this form of equalization may be of use.

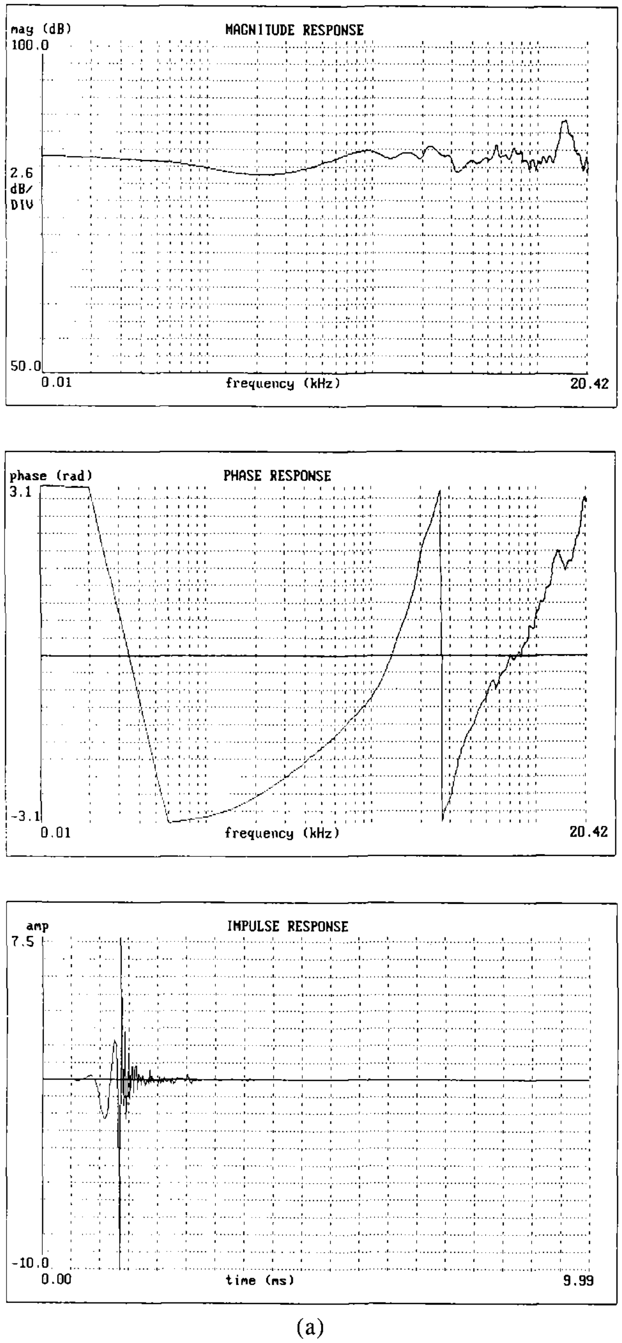
4 APPLICATIONS OF EQUALIZATION TECHNIQUE

The principal application is the equalization of a loudspeaker system response. Secondary issues, which will also be discussed, are its application to the subjective effects of phase distortion and to loudspeaker system measurement.

For the purpose of this paper, the equalizers demonstrated are derived from the on-axis response measurement. This is done here because of the ease of observing the effect of the equalizer on the loudspeaker system's response. This form of equalization also has application to loudspeaker system measurement and quality assessment, which will be discussed later. The data on which the equalizer is based are obtained from an impulse measurement technique similar to that of Berman and Fincham [12], which provides an impulse response in a digital form suitable for direct input into a computer. The frequency response measurements of a commercial loudspeaker system (obtained from the fast Fourier transform of the impulse response) are shown in Fig. 3(a) and the corresponding responses of a 55th-order model in Fig. 3(c). The equalizer responses, formed from the cascade of a 70th-order FIR filter (excess-phase equalizer) and the 55th-order IIR filter (minimum-phase and magnitude equalizer) are shown in Fig. 7(a). The measured equalized responses, given in Fig. 7(b), show significant improvement in the impulse response and, consequently, the frequency responses as well. The equalized responses presented in this paper were obtained using a real-time digital filtering system built around the TMS320C25 digital signal processor. The loudspeaker equalizer uses the AES/EBU digital interface and is therefore compatible with most two-box CD systems (CD transport and outboard DAC). The prototype digital equalizer is shown in Fig. 8.

The method of deriving an equalizer described in this paper enables some subjective evaluation of the loudspeaker's inherent phase distortion to be performed (as opposed to additionally imposed phase distortion). Sec. 3 discussed the implementation of separate equalizers, a magnitude with minimum-phase equalizer, and an excess-phase equalizer. If just the former is used, the resultant response is all pass, that is, only phase distortion is incurred. The minimum-phase (magnitude) equalized responses are given in Fig. 7(d). The magnitude responses of Fig. 7(d) and (b) are similar, while the phase responses and consequently the impulse responses are quite different. Thus the application of these equalizers will permit the subjective effects of the excess-phase distortion exhibited by the loudspeaker to be readily observed and compared to measured performance. Such experiments have been carried out and were reported elsewhere [13].

Equalization also has application in loudspeaker system measurement and quality assessment. One deficiency of the impulse measurement technique is the truncation of the impulse before the first reflection [12].



(a)

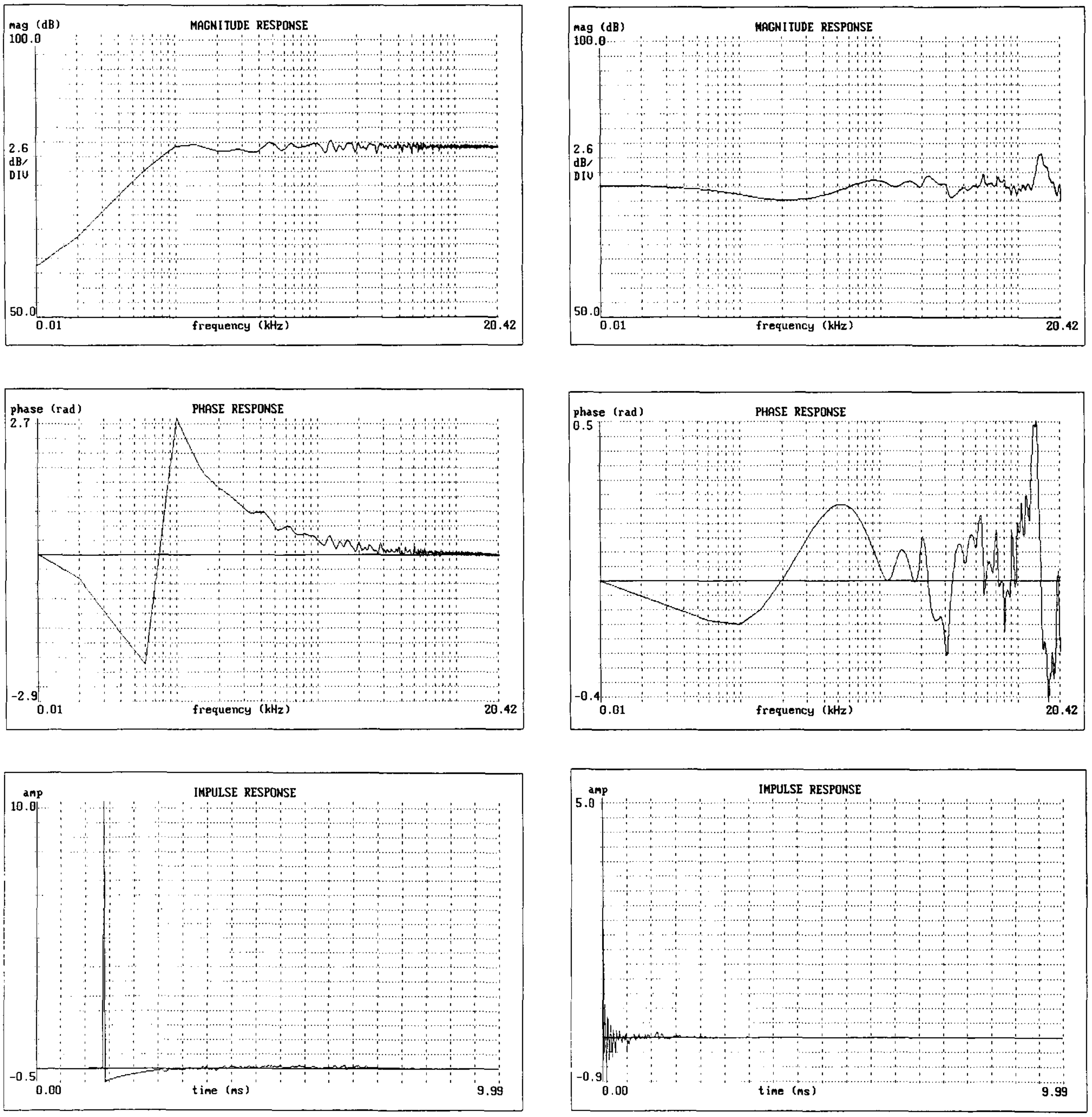
Fig. 7. (a) Magnitude and phase equalizer responses. (b) Measured loudspeaker responses after magnitude and phase equalization. (c) Magnitude and minimum-phase equalizer responses. (d) Measured loudspeaker responses after magnitude and minimum-phase equalization.

This leads to a loss of information, particularly in the low frequencies. In a follow-up paper to [12], Fincham [14] proposes the use of an equalizer to reduce the duration of the impulse tail. Hence a reduced amount of information is lost by the truncation. The techniques described in this paper are similarly applicable to the enhancement of acoustic measurements. Indeed, there may be some additional gain to be had by removing higher frequency artifacts from the measurement. These, too, may lead to an extended impulse response.

Another application of equalization is in loudspeaker system quality assessment where, briefly, if the loudspeaker system is equalized on axis, deterioration of the off-axis responses would show deficiencies of the system. Three-dimensional plots of the time or frequency responses versus measurement angle of the equalized loudspeaker would illustrate valuable information about the performance of a loudspeaker system.

5 CONCLUSION

An equalizer design technique which uses both FIR and IIR filters has been presented. The function of the IIR filter is to equalize the magnitude response of the loudspeaker. It is believed that, in most instances, an IIR structure will perform this task far more efficiently



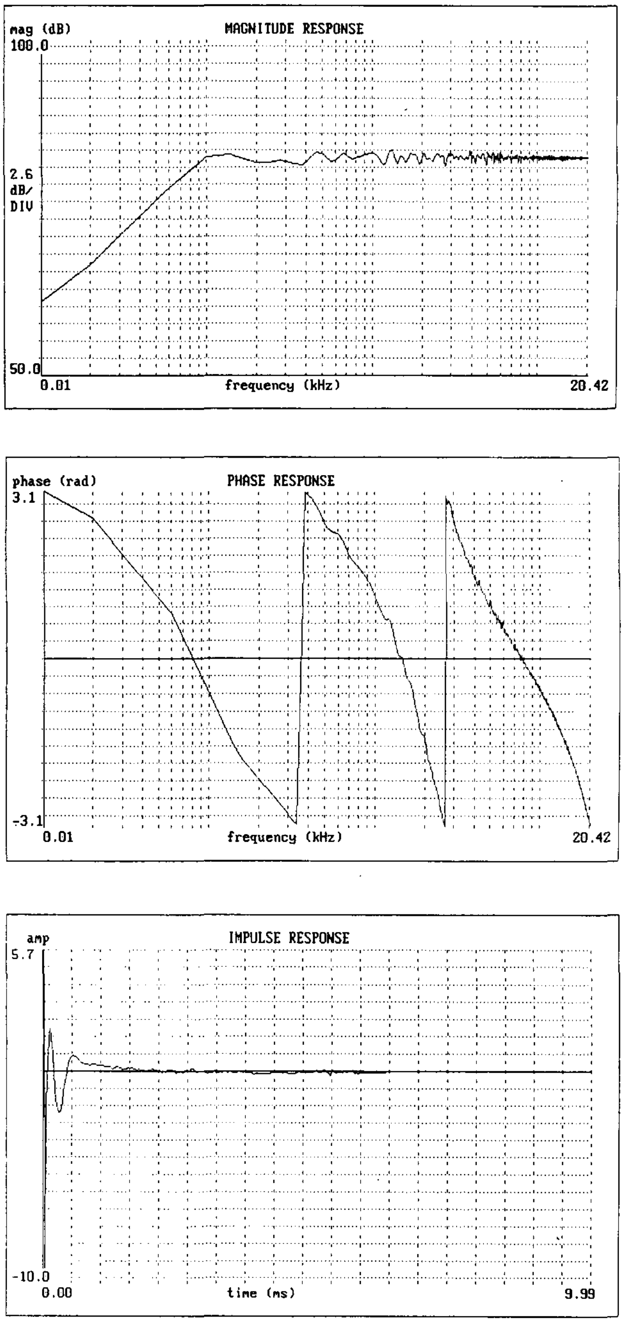
(b) (c)

Fig. 7. continued.

than what can be achieved using FIR structures. The FIR filter is used to deal with any acausal requirements of the equalizer. Measured responses, taken from a real-time digital equalizer, demonstrate significant improvements in the loudspeaker's linear transfer function. A method was introduced for constraining the impulse response of the FIR filter. This allows practical implementation of the excess-phase equalizer without compromising the amplitude response, although the amount of excess-phase equalization is now reduced.

A feature of the design process is the ability to separate the minimum- and excess-phase components of the system. Thus the audible effects of loudspeaker excess-phase distortion can be assessed and compared to measured responses. The subjectivity of phase distortion is a controversial topic, and this technique may provide valuable contributions toward it.

Finally, the application of the scheme to loudspeaker system measurement and quality assessment is suggested, which could prove useful in loudspeaker system design and manufacture.



(d)

Fig. 7. continued.

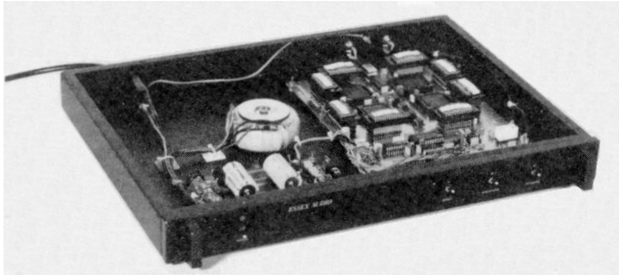


Fig. 8. Digital loudspeaker equalization system.

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Since his employment at Essex, he has established the Audio Research Group, where research on amplifier studies, digital signal processing, and loudspeaker systems has been undertaken. Since 1982 research into digital crossover systems has begun within the group and, more recently, oversampling and noise shaping investigated as a means of analog-to-digital/digital-to-analog conversion.

Dr. Hawksford has had several AES publications that include topics on error correction in amplifiers and oversampling techniques. His supplementary activities include writing articles for *Hi-Fi News* and designing commercial audio equipment. He is a member of the IEE, a chartered engineer, a fellow of the AES and of the Institute of Acoustics, and a member of the review board of the *AES Journal*. He is also a technical adviser for *HFN* and *RR*.