Oversampling Filter Design in Noise-Shaping Digital-to-Analog Conversion

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Oversampling Filter Design in Noise-Shaping Digital-to-Analog Conversion*

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A number of oversampling filters in association with a heavily oversampled and noise-shaped digital-to-analog converter (DAC) are presented for high-resolution applications in professional and consumer digital audio environments. Computer simulations of oversampling filter, noise shaper, and DAC errors are described, together with results to demonstrate interactive performance dependence on filter nonideality.

0 INTRODUCTION

The performance of a digital audio system that is bounded by uniform source quantization and sampling should result only in band limitation of the input signal, together with an additive noise component. The realization of this target is dependent on identifying analog-to-digital (ADC) and digital-to-analog conversion (DAC) systems that can attain near ideal anti-aliasing and signal recovery filters in association with a virtually perfect quantization characteristic and the means of decorrelating signal and quantization distortion so as to mimic the characteristics of purely additive noise.

In conventional ADC and DAC systems a principal limitation is low-level linearity of the quantization characteristics [1], where the least significant bit can reveal both relative and absolute errors in the converter. This level displacement prevents the resolution potential of the digital channel from being achieved, where, as signals approach and enter the quantization noise floor, the nonlinearity produces waveform distortion, rather than the ideal, where a signal sinks into the quantization noise without observable correlation.

To achieve this performance target requires the identification of hardward that can attain the linearity potential of the digital format, implement effective signal quantization decorrelation, and decorrelate the distortion arising from hardware nonlinearities so as to form a noiselike residue.

In an earlier paper [2], [3] a DAC was presented that used a high oversampling ratio in association with a noise shaper where, for order $R_N = 4$, a wide dynamic range was demonstrated that in the absence of DAC nonidealities, extended well beyond the requirement of current digital formats. The fourth-order noise shaper was shown to exhibit a chaotic behavior as the output signal spanned approximately 16 quantization levels (that is, 4 bit), which resulted in the desirable quality of translating DAC imperfections to a noiselike residue.

To maximize the performance of this system, the 16 DAC reconstruction levels should adhere to an effective 16-bit precision, although adequate performance was still demonstrated if this figure was reduced to 12 bit by introducing a combination of random and systematic displacements of the DAC levels. However, since the DAC has only a limited amplitude range and each reconstruction level requires equal weight, the desired accuracy is relatively simple to achieve.

As an alternative to the parallel DAC architecture, a second study [4] has shown how the 16 output level code of an $R_N = 4$ noise shaper can be translated to a serial bit stream using codes of optimized low-frequency spectral form. This signal format offers a similar advantage to that of delta–sigma modulation [5]–[7], whereby the DAC reduces to a 1-bit gateway and exhibits an excellent tolerance of hardware imperfection. The advantage of combining a recursive noise shaper and a nonrecursive code converter is that a high-order loop

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can be achieved using equally weighted loop integrators of optimal gain–bandwidth product, whereas a similar 1-bit noise shaper requires special attention in the loop design to achieve a stable performance. Also, where serial bit rates greater than 100 MHz are anticipated, the two-stage system appears more practical in terms of hardware realization.

Irrespective of whether a parallel or a serial port is used for the noise shaper, both systems require an interpolation filter to upconvert audio data at 44.1 kHz (or 48 kHz) to about 5.64 MHz sampling, where the filter also achieves the signal recovery filter function used in more conventional systems. In this paper a number of interpolation filters are presented and, by means of computer simulation, the relationship between filter design and overall system performance is investigated. The filters can be modeled as either a single-stage or a multistage process. The two methods are compared with respect to computational rate and coefficient storage requirements. The results of the computer simulation, which include interpolation filter, noise shaper, and DAC errors, are presented using spectral analysis of computed data sequences.

1 OVERSAMPLING FILTER, NOISE SHAPER, AND DAC

The technique of noise shaping illustrated in Fig. 1 was described in an earlier paper [2], where an analysis of coding behavior was given. Noise shaping has also been the subject of numerous other studies [8]–[10]. In the present study the oversampling filter is considered in relationship to the noise shaper and DAC, as shown in Fig. 1(a). The function of the oversampling filter is to remove the spectral replications about integer multiples of the Nyquist sampling frequency, which become redundant when the sampling rate is increased by a factor L. The process is illustrated in Fig. 2 in both the time and the frequency domain. Once a signal is oversampled, the amplitude resolution of the output samples can be reduced and the requantization distortion located in the now redundant signal space created by oversampling. It is the function of the recursive noise shaper in association with the requantizer to shape the noise spectrum to fit the available signal space, yet to inflict minimum linear and nonlinear distortion on the 0–22.5-kHz audio band.

The oversampling and noise-shaping system is cascaded with a high-speed low-amplitude resolution DAC and low-order analog reconstruction filter to complete the conversion. Earlier studies [2], [3], [10] have already discussed both the inclusion of a nonideal DAC and the translation of DAC errors to a random process, though the oversampling filters, at this stage, were assumed perfect.

2 FILTER STRUCTURES

The approach taken in this paper is to review several filter design methods that are applicable to interpolation and then present simulated results based on a number of filter examples. However, detailed design and theory of each filter are not given as they are adequately described in the cited references.

The choice of filter considered is limited to nonrecursive structures as exact linear-phase and reduced-word-length effects are desirable characteristics for oversampling. Consequently an output sequence $y(n)$ is computed from the input sequence $x(n)$ and the discrete impulse response $h(k)$ using the discrete con-

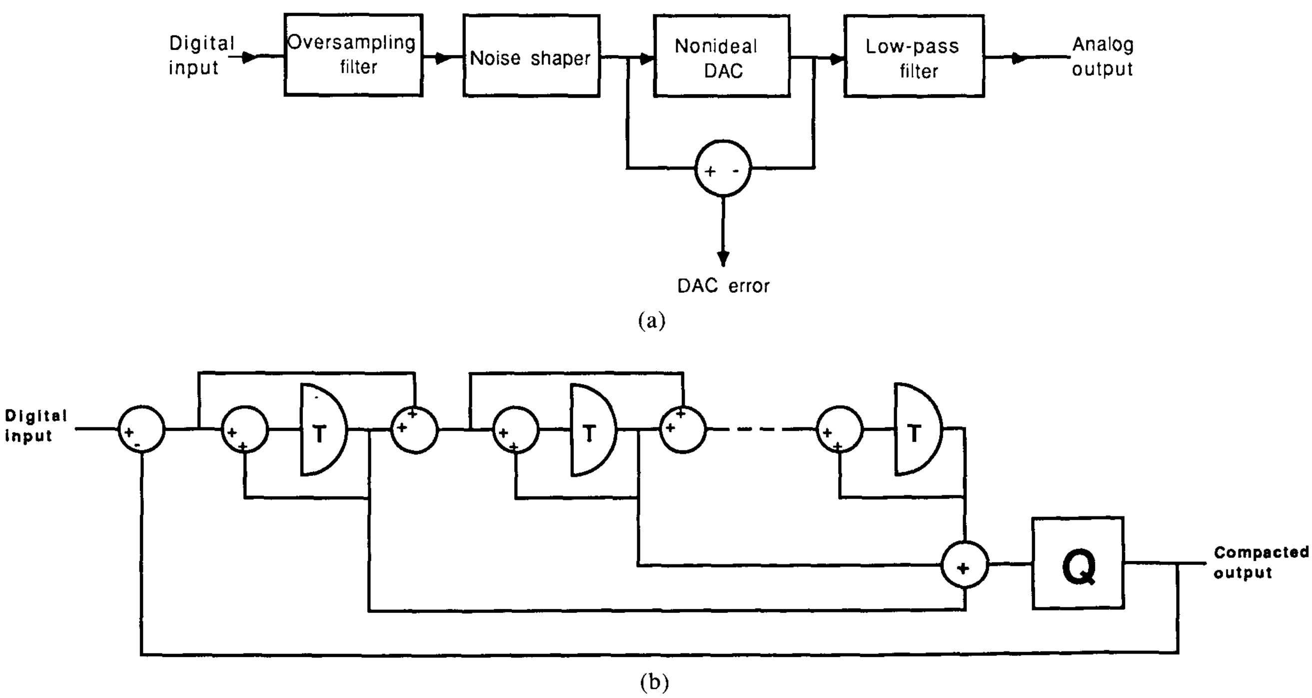

Fig. 1. (a) Oversampling and noise-shaping DAC with level compaction. (b) Nth-order linear-feedback noise-shaping coder.

volution operation

$$y(n) = \sum_{k=1}^{m} h(k) \cdot x(n - k) \ . \tag{1}$$

In the case of the oversampling filter, the arithmetic operations are performed at a rate Lf_s, where f_s is the Nyquist sampling rate of the input sequence in hertz. However, the nonrecursive structure enables the computation rate to be reduced by a factor L by identifying the redundant, zero multiplications, as shown in the signal graphs of Fig. 3 [12].

A range of design techniques are available to address filter design problems. The most prominent are as follows:

1) Equiripple FIR design [13]
2) FIR design with or without "don't care bands" [14]
3) Half-band filters [15]
4) Minimum-mean-square error design and the linear/Lagrange interpolator [16].

The widely selected optimal FIR design uses the Chebyshev approximation for the desired frequency response. The program of McClellan et al. [13] cal-

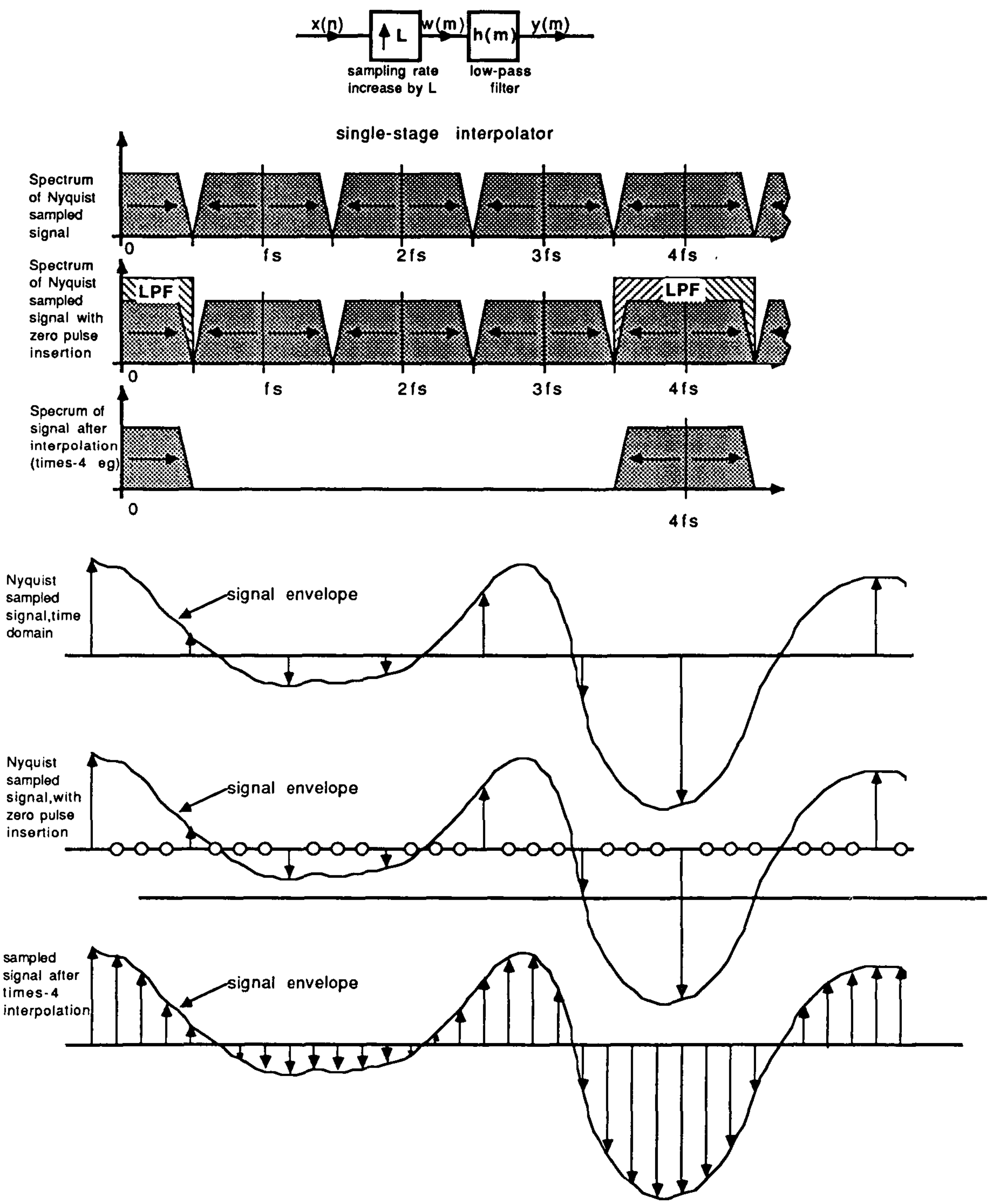

Fig. 2. Example of single-stage interpolation for $L = 4$; in both frequency and time domains.

culates the filter coefficients. A tolerance scheme, shown in Fig. 4, sets limits for the design that bounds passband amplitude ripple, stopband amplitude ripple attenuation, and transition region, together with the number of filter coefficients, which then form the program input parameters.

A significant reduction in the multiplication rate is achieved by partitioning the interpolation process into a number of intermediate interpolation stages of ratio L_i, where

$$L = \prod_{i=1}^{L} L_i \ . \qquad (2)$$

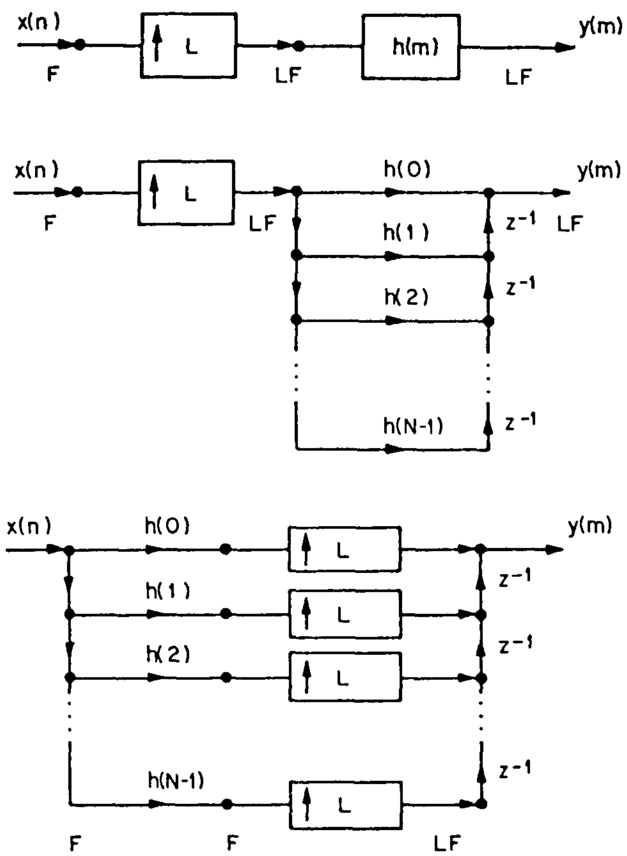

Fig. 3. Generation of efficient structure of 1-to-L interpolator.

Fig. 5 presents an illustrative example of ×4 interpolation implemented as two cascaded stages of ×2 interpolation. Inspection of the second interpolation filter reveals that, because of "don't care bands," the design can be relaxed and the broader transition band allows the number of filter coefficients to be reduced if, for example, the McClellan program is used. Multistage filters reduce both the overall computation and the storage requirements, where each stage is now an independent interpolation stage with simplified design. A method by Crochière and Rabiner [14] is to choose all stages with $L_i = 2$, which has the advantge of combining half-band filters, while another approach by the same authors [14] uses an optimization procedure which minimizes the total computation rate.

The half-band filter exploits the possible symmetry in the transition region where, for a transfer function of the form shown in Fig. 6, the impulse response can be shown to have every other coefficient equal to zero, which represents a 50% saving in multiplications. Half-band filters can be designed using the McClellan procedure, but the resulting coefficients do not exactly match the required constraint of symmetry. However, a new "design trick" [17], [18] facilitates the filter design by enabling only the nonzero $h(k)$ coefficients to be calculated.

A further design procedure makes use of time-domain optimization [14], where the signal error $y(m) - \hat{y}(m)$ of the target interpolated signal $\hat{y}(m)$ and the actual interpolated signal $y(m)$ is minimized. A computer program [16] uses this algorithm, which is based on a solution of linear equations. Finally, linear/Lagrange interpolators are filters with coefficients calculated by mathematical approximation, where, effectively, new

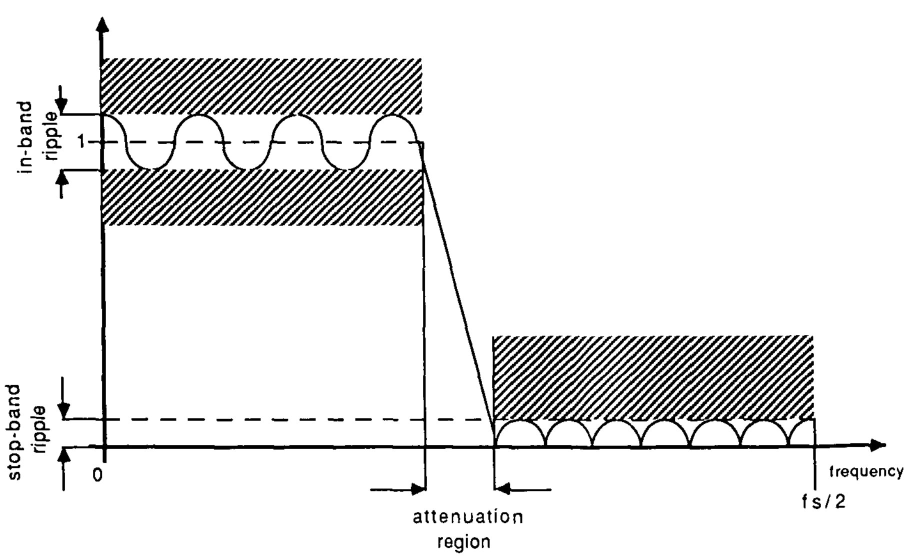

Fig. 4. Tolerance scheme for low-pass filter design.

samples are created by interpolating between two or more of the originals.

3 COMPUTER SIMULATION

The computer simulations follow earlier procedures [2], [3], [10] but now include the interpolation filter to enable the effect of nonideal interpolation to be calculated for a range of filter designs. The simulation uses 4096 sample points of a sine wave signal to calculate the operations of filtering, noise shaping, and DAC nonidealities (both static and dynamic errors). The following data and filter specification were selected:

Data:

Input signal frequency	11.025 kHz
Input signal amplitude	$\sqrt{2}$ units
Noise shaper requantization interval	1 unit
Noise shaper sampling frequency	5.6448 MHz
Number of sample points	4096
Number of requantization bits	4 bit
Noise shaper, loop order	4 integrators
Oversampling ratio	128
Number of input cycles shown by computation	8

Filter:

Passband ripple (0–20 kHz)	0.0001 dB
Transition band	20–24.1 kHz
Stopband (24.1 kHz)	−100 dB
Number of stages	2–6

4 RESULTS

4.1 FILTER PERFORMANCE

The multistage structure of an oversampling filter enables a decrease in the required memory and computation rate. A single-stage interpolator where $L = 128$ needs about 8223 coefficients to achieve an attenuation of -100 dB with a multiplication rate of 181 ×

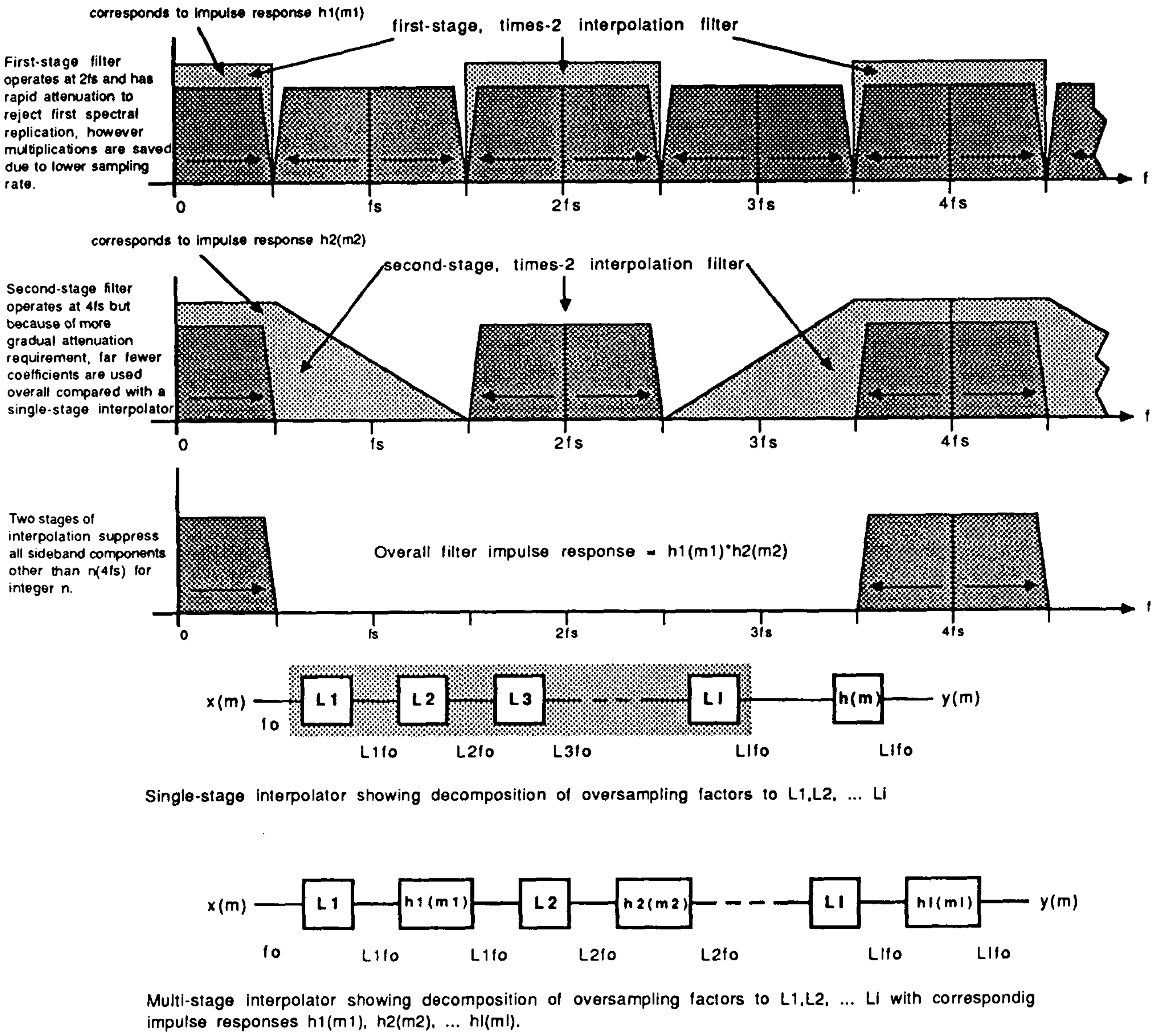

Fig. 5. Multistage interpolation with illustrative example of two-stage interpolation process.

10^6 multiplications per second. A two-stage filter reduces the coefficients to 508 with a corresponding 28 × 10^6 multiplications per second, which results from a widening of the transition band. Most of the gain for optimized computation has been achieved taking $L_i = 2$ for the first subfilter, although a half-band filter reduces the number of coefficients by 50%.

Further design shows that the largest decrease of computation rate occurs when going from one to two or three stages; further partitioning does not result in significant reductions. The digital interpolator design [16] creates a multistopband filter for $L > 2$ and achieves better results than the Lagrange interpolator. For $L_i = 2$, both techniques approach the result of half-band filters with even better results than the Remez algorithm for short filters and large transition bands ($N = 7, \ldots, 11$), such as for the last stages.

Tables 1 and 2 present a comparative overview of some design examples using multistage structures (2 to 6 stages) and Fig. 7 shows the example of ×4 interpolation presented as a series of frequency response plots.

4.2 Quantization of Filter Coefficients

The quantization of the filter coefficients $h(k)$ is simulated with both fixed-point and floating-point arithmetic, where the more optimum floating-point reveals frequency responses enhanced by up to 10 dB, as shown in Fig. 8. The reason lies in the greater dynamic range of floating-point arithmetic, which is particularly important for the first filter stage when the number of filter coefficients exceeds 100.

4.3 DAC Nonideality

4.3.1 Ideal DAC

The overall error spectrum for a fourth-order noise shaper using the simulated oversampled filters does not deviate significantly from the simulation for ideal

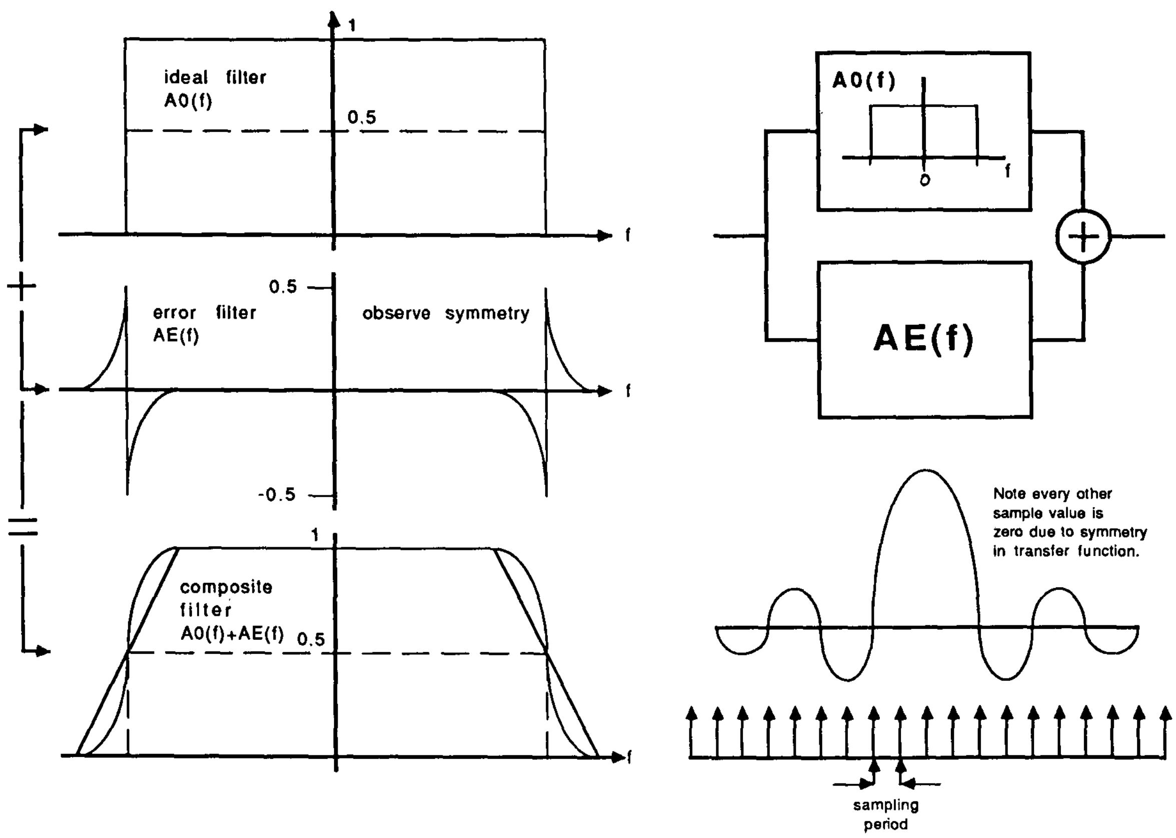

Fig. 6. Half-band filter showing symmetry in frequency response and corresponding alternate zeros in impulse response.

Table 1. Filters designed with multistage optimization structure.

Stages (L_j)	Coefficients	Memory	$R/10^6$ s	Passband (dB)	Stopband (dB)
2 (4, 32)	255, 253	255	28.05	−0.000162	−97.95
3 (2, 4, 16)	143, 49, 113	119	23.94	−0.000105	−108.47
3 (2, 8, 8)	143, 97, 53	113	25.00	0.000116	−105.61
4 (2, 2, 2, 16)	143, 25, 15, 113	115	24.30	0.000065	−108.46
4 (2, 2, 4, 8)	143, 25, 31, 53	93	24.65	−0.000110	−105.61
5 (2, 2, 2, 2, 8)	143, 25, 15, 15, 53	90	25.00	0.000066	−105.61
5 (2, 2, 2, 4, 4)	143, 25, 15, 31, 23	86	26.77	0.000113	−107.58
6 (2, 2, 2, 2, 2, 4)	135, 25, 15, 15, 15, 23	76	26.33	0.000097	−103.37
6 (2, 2, 2, 2, 2, 4)	143, 25, 15, 15, 15, 23	78	26.42	0.000067	−107.58

oversampling, where on average an in-band noise of about -130 dB is achieved.

4.3.2 Nonideal DAC

The inclusion of errors in the DAC has already been discussed [2], [3], where a range of errors including both static and dynamic mechanisms were modeled. Again, the example simulations shown in Figs. 9–11 demonstrate that the type of oversampling filter does not significantly affect the in-band noise level (-106 to -109 dB) sine degradation from the ideal DAC case is primarily attributable to DAC errors.

4.3.3 Filter Quantization Noise

A further noise source inherent to the oversampled system is associated with quantization or truncation of the signals within the FIR filters. To illustrate this mechanism, a four-stage filter was quantized with different resolutions for each stage as follows:

Stage 1 16 bit

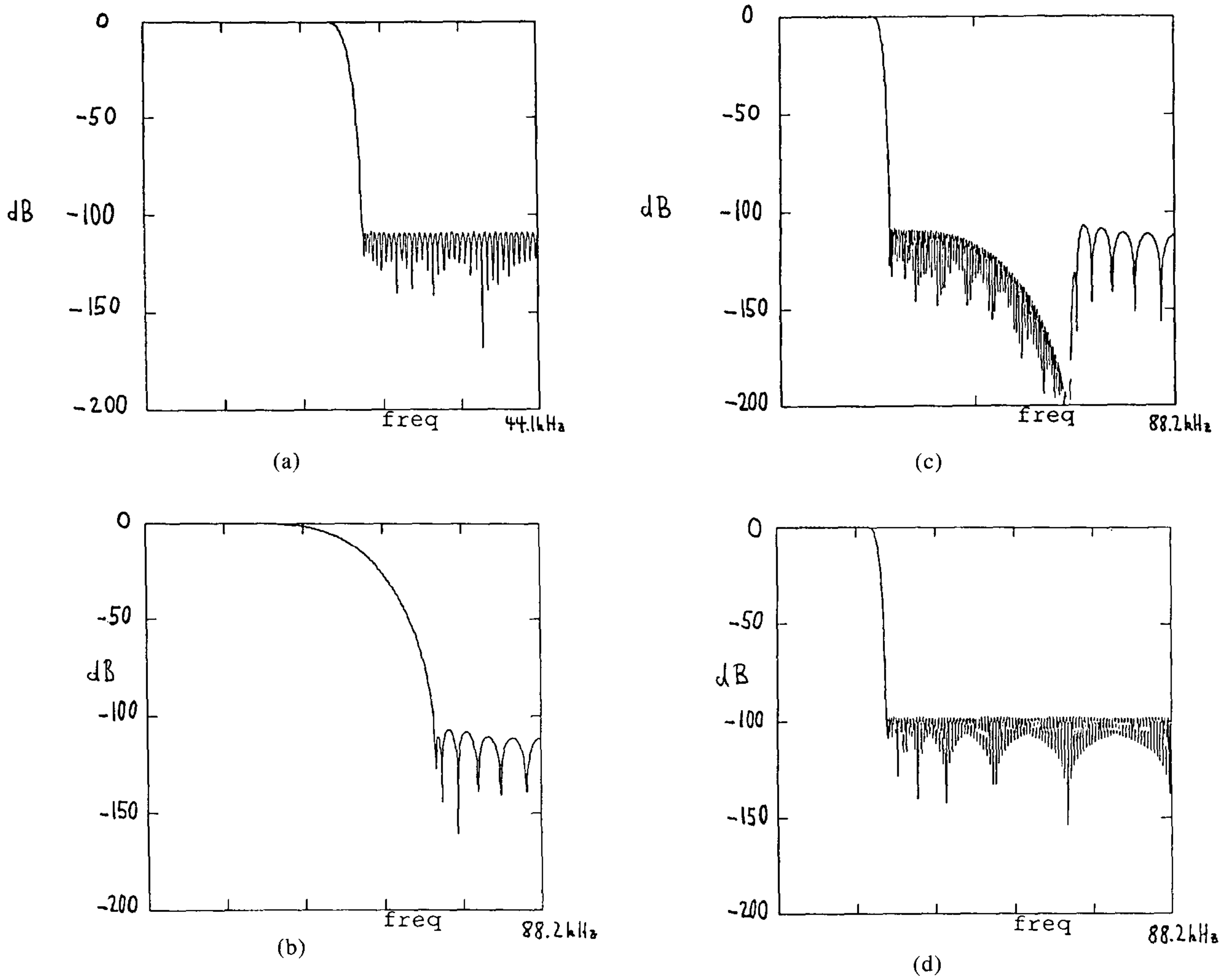

Fig. 7. Comparison of single-stage and two-stage interpolation structures. (a) 1-to-2 interpolation filter (143 coefficients). (b) 2-to-4 interpolation filter (23 coefficients). (c) Overall frequency response. (d) 1-to-4 interpolation filter (255 coefficients).

Table 2. Filters designed with halfband design procedure.

Stages (L_j)	Coefficients	Memory	$R/10^6$ s	Passband (dB)	Stopband (dB)
3 (2, 2, 32)	127, 19, 253	166	24.30	0.000259	−98.45
3 (2, 2, 32)	131, 23, 253	167	24.52	−0.000170	−100.92
3 (2, 2, 32)	139, 23, 253	170	24.61	−0.000101	−104.51
3 (2, 2, 32)	143, 27, 253	171	24.65	−0.000084	−104.51
4 (2, 2, 2, 16)	143, 27, 11, 113	106	23.15	0.000122	−108.44
4 (2, 2, 2, 16)	143, 27, 15, 113	107	23.33	0.000044	−108.44
4 (2, 2, 2, 16)	131, 23, 11, 113	103	22.93	0.000208	−100.92
4 (2, 2, 2, 16)	131, 23, 15, 113	103	23.14	0.000149	−100.92
5 (2, 2, 2, 2, 8)	143, 27, 11, 11, 53	80	23.51	0.000120	−105.61
5 (2, 2, 2, 2, 8)	143, 27, 15, 11, 53	81	23.68	0.000045	−105.61
5 (2, 2, 2, 2, 8)	143, 27, 11, 15, 53	80	23.86	0.000118	−105.61
5 (2, 2, 2, 2, 8)	143, 27, 15, 15, 53	81	24.03	−0.000043	−105.61
6 (2, 2, 2, 2, 2, 4)	143, 27, 15, 11, 7, 23	69	23.68	−0.000054	−107.59

Stage 2 14 bit
Stage 3 14 bit
Stage 4 12 bit.

The input signal $x(n)$ and the output signal $y(n)$ are quantized with the same bit number whereon the filters increase the noise level of the noise shaper (ideal DAC) from -130 dB average in-band noise to about -112 dB. For floating-point arithmetic the increase is less, as shown in Tables 3 and 4. However, when DAC errors are included, the difference in noise level with and without quantization is small; DAC nonidealities are again the dominant noise source.

5 CONCLUSION

This paper has demonstrated the feasibility of a digital oversampling filter that is compatible with the requirements of a fourth-order noise-shaping and 4-bit DAC in terms of inherent noise generation and signal recovery filtering.

Multistage structures reduce significantly both the computation rate and the required filter coefficient storage. Half-band filters are also efficient in this application, where attenuations of -100 dB can be achieved with less than 100 words of coefficient memory.

The filter simulations illustrate that filter-induced nonidealities overlap those of DAC errors but do not significantly degrade the coder performance, where the overall error spectrum of the system is virtually identical to that measured with an ideal oversampled input signal. The output noise level of the complete system lies well below -100 dB and can readily attain -110 dB, so enabling a signal coded to 18 bit to maintain its signal-to-noise ratio.

A useful characteristic of the process is the virtual elimination of low-level nonlinearity compared with a conventional DAC, where nonlinearity results from suboptimal weights, particularly with the least significant bits. The system imperfections instead manifest themselves as a noise residue, whereby signals, provided they are appropriately coded, can enter well into the system noise without nonlinear impairment. A measure of the effectiveness of this technique is to observe the noise-shaper noise floor in association with an ideal DAC, where for $\times 128$ oversampling and a fourth-order noise shaper, the output noise level is about -130 dB.

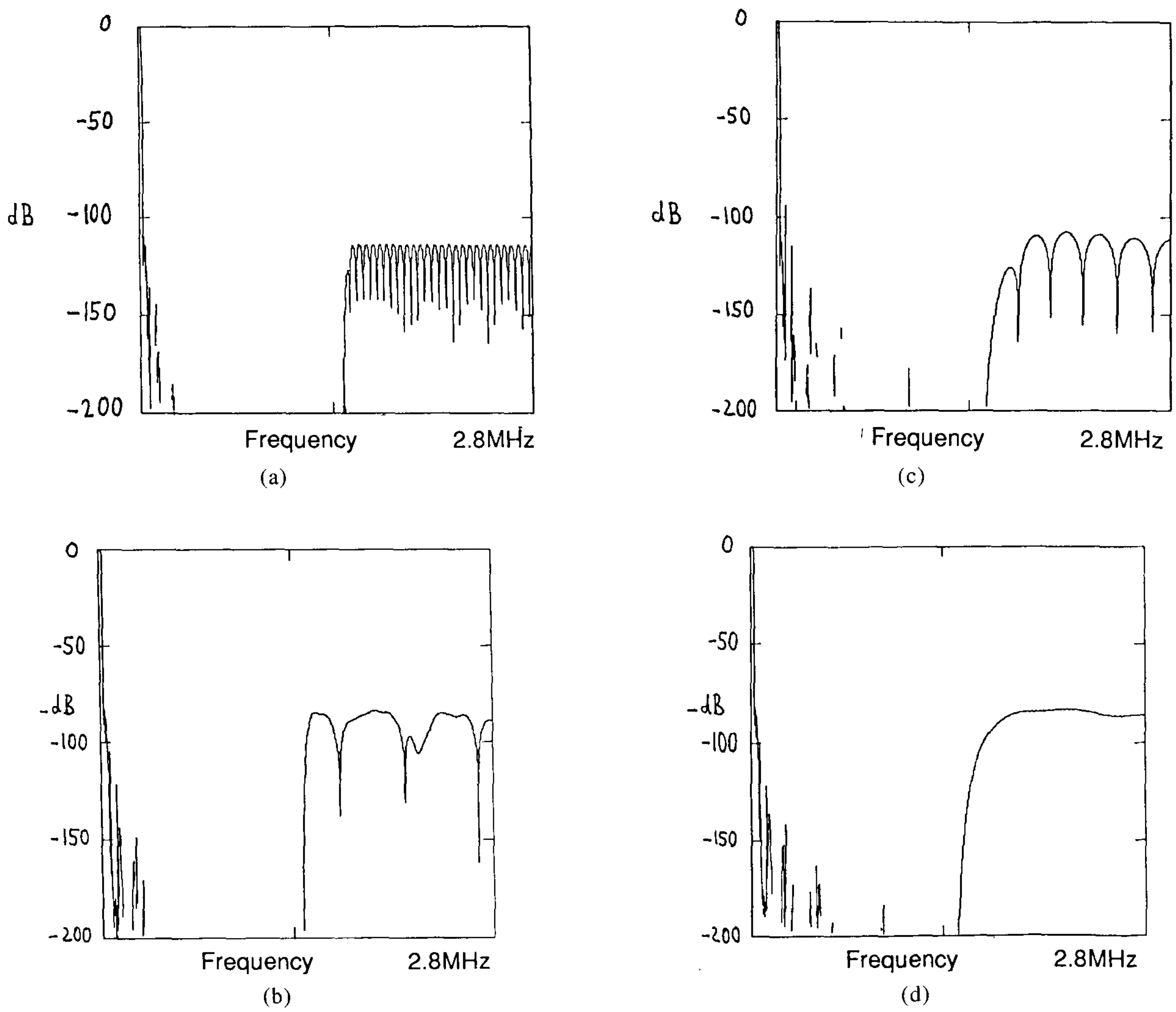

Fig. 8. Four-stage and six-stage $\times 128$ interpolation filters with and without quantization. (a) Four-stage filter. (b) Four-stage filter quantized with 16 bit. (c) Six-stage filter. (d) six-stage filter quantized with 16 bit.

Hence with an appropriate 4-bit DAC design using a precision-weighted network to achieve 16-bit accuracy, or with code conversion to a 1-bit serial format, resolution approaching 19–20 bit should be attainable together with a clean low-level signal characteristic, commensurate, of course, with optimal ADC and signal-distortion decorrelation.

The study of filter quantization effects supports the use of floating-point arithmetic for high-resolution systems, particularly as the coefficient range is high, where quantization can cause more noise in the higher frequencies.

The technique of high oversampling and noise shaping therefore addresses most of the requirements of a DAC system for high-performance digital audio systems. The oversampling filters, when partitioned to three or four cascaded stages, are practical and can be designed for exemplary in-band amplitude ripple and out-of-band attenuation and, when using floating-point arithmetic, can achieve a high dynamic range. The heavily oversampled noise shaper of order 4 does not appear to suffer significant impairment from slight ultrasonic components as a consequence of nonideal interpolation and can also achieve a wide dynamic range, particularly as the topology does not require multiplications. The main source of impairment is again the DAC, though this is aided by errors translating to noise and the relative ease of designing a converter of high linearity.

The heavily oversampled and noise-shaped DAC is seen as complementary to the ADC system of Adams [19], where oversampling and noise shaping simplifies analog circuitry, eliminates problems associated with the least significant bits of conventional DACs, and reduces sensitivity to sampling jitter significantly when allied with switched capacitor or similar jitter reduction techniques [7]. The data rate at 44.1 kHz then becomes only an information channel, where distortions related to the direct conversion of Nyquist samples are eliminated. Also the techniques are equally applicable where data-reduction algorithms are implemented. Indeed, with near-transparent conversion the true performance potential of such algorithms can be investigated without masking effects from suboptimal ADC and DAC systems.

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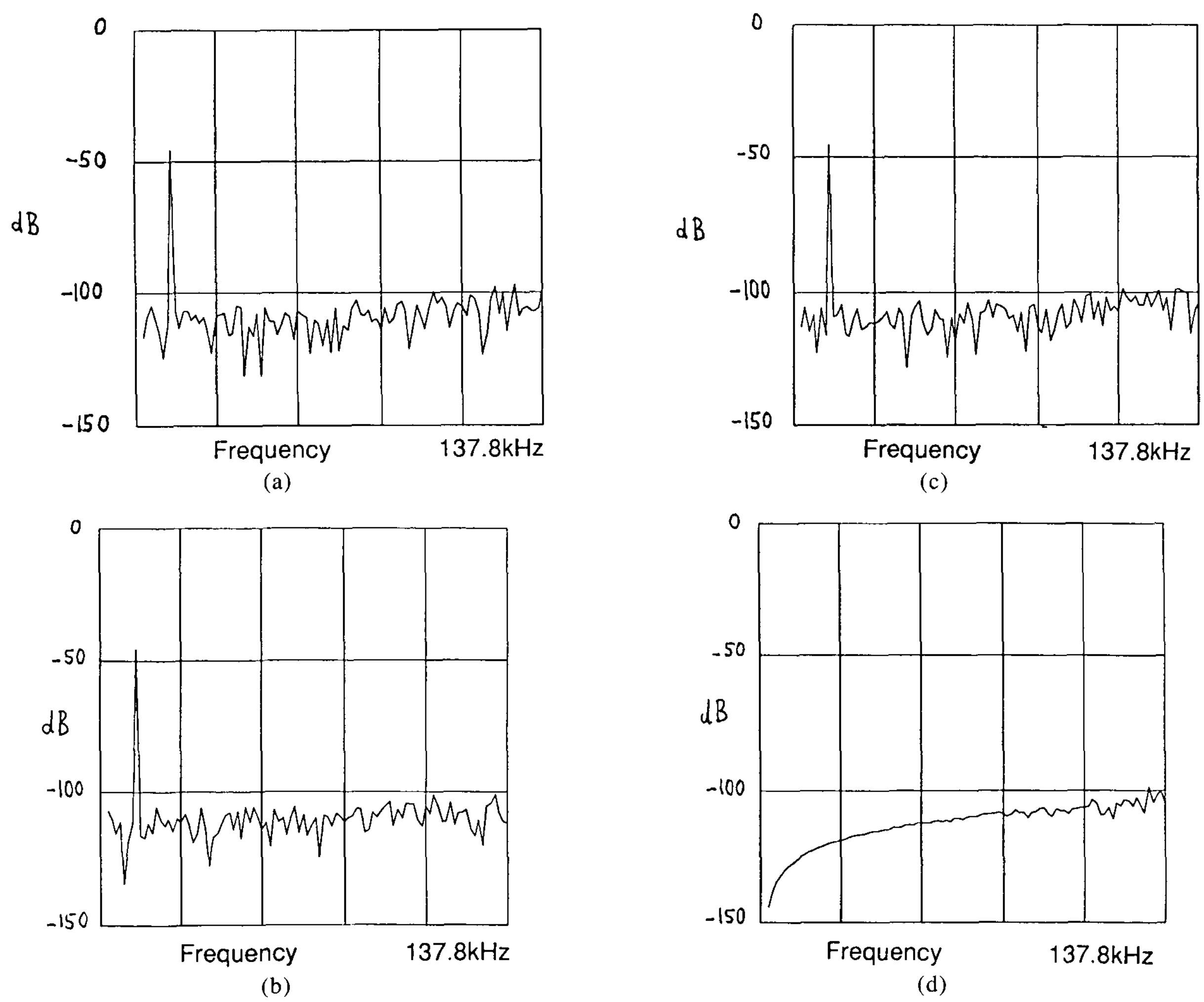

Fig. 9. Overall error (nonideal DAC), 0–137.8 kHz. (a) Filter 1. (b) Filter 2. (c) Filter 3. (d) Ideal input.

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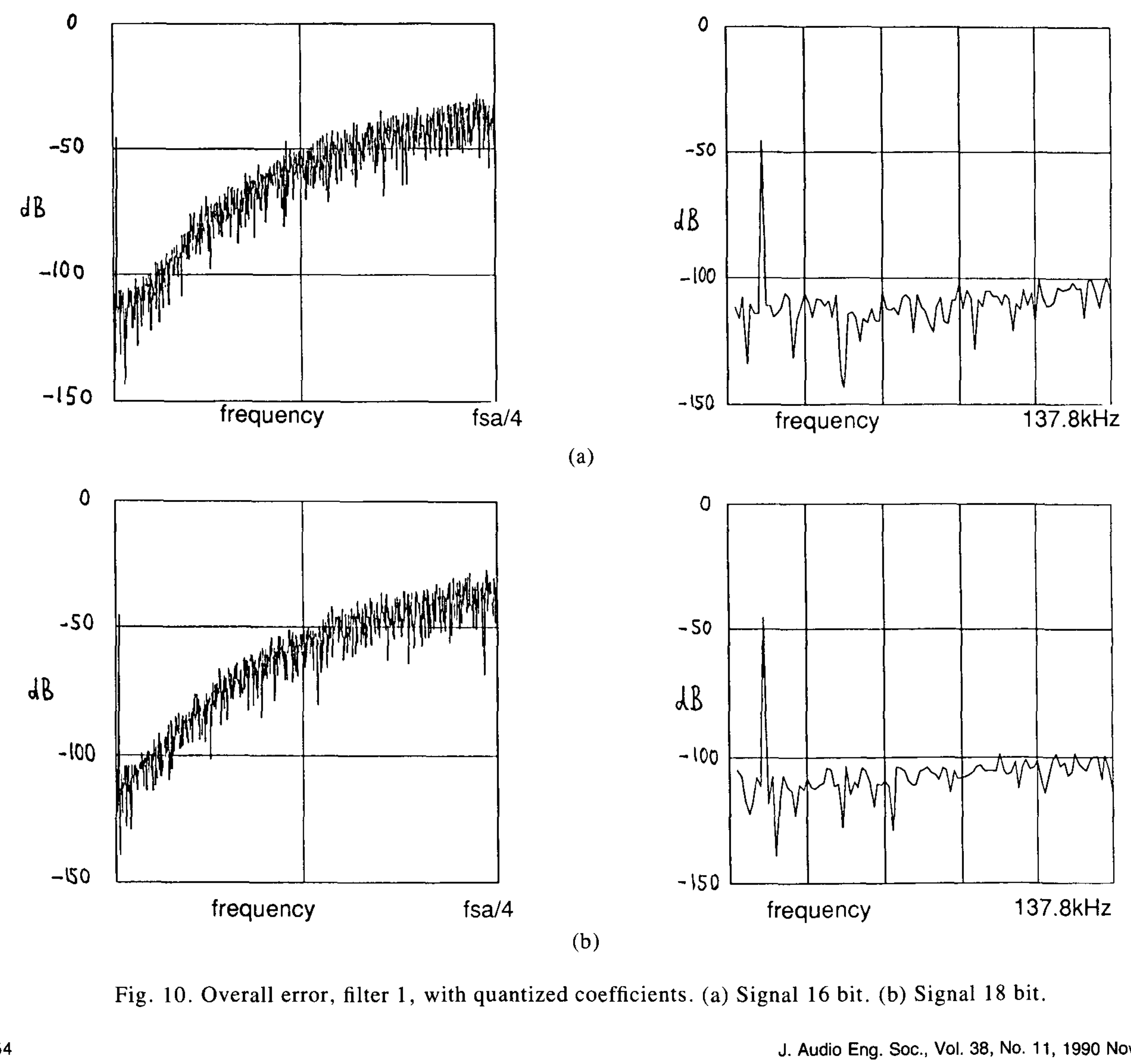

Fig. 10. Overall error, filter 1, with quantized coefficients. (a) Signal 16 bit. (b) Signal 18 bit.

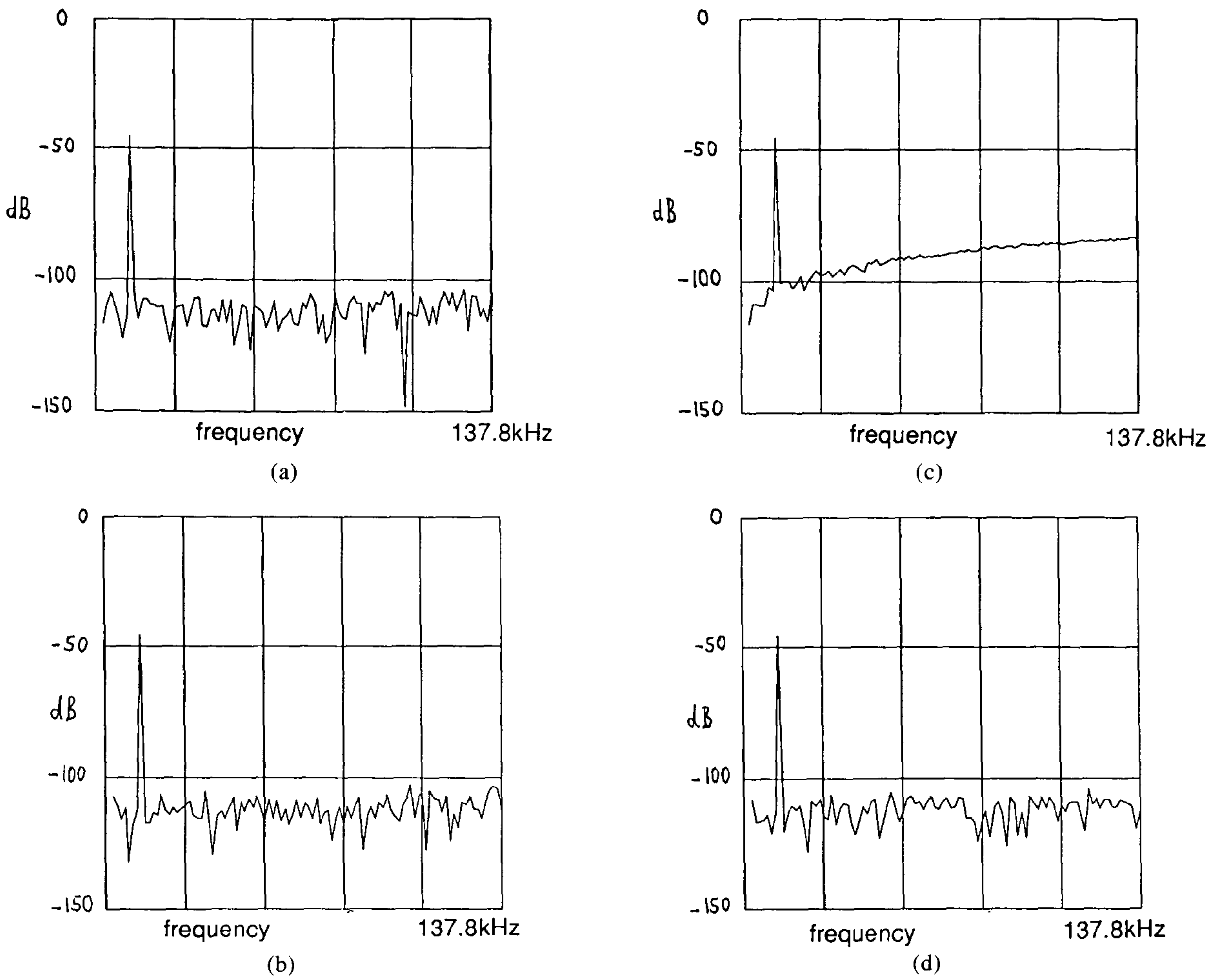

Fig. 11. DAC error, 0–137.8 kHz. (a) Filter 1. (b) Filter 2. (c) Filter 3. (d) Ideal input.

Table 3. Fixed-point arithmetic, average overall error, in decibels.

	Ideal DAC		Nonideal DAC	
Signal bit	0–137 kHz	0–20 kHz	0–137 kHz	0–20 kHz
16	−97.166	−112.652	−105.609	−106.231
17	−99.362	−114.778	−109.092	−109.367
18	−95.566	−111.028	−110.875	−109.152
19	−96.086	−111.536	−109.500	−108.465

Table 4. Floating-point arithmetic, average overall error, in decibels.

	Ideal DAC		Nonideal DAC	
Signal bit	0–137 kHz	0–20 kHz	0–137 kHz	0–20 kHz
16	−99.913	−115.381	−110.718	−108.645
17	−95.392	−110.836	−111.303	−108.370
18	−110.879	−126.401	−107.598	−109.872
19	−108.420	−123.901	−106.009	−110.188

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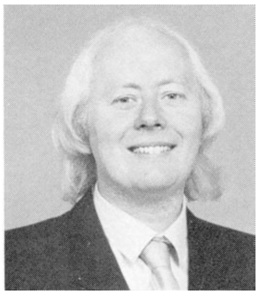

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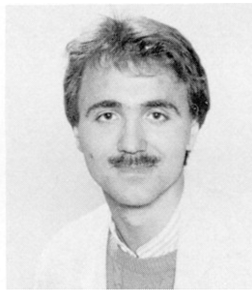

W. Wingerter

Malcolm Omar Hawksford is a reader in the Department of Electronic Systems Engineering at the University of Essex, where his principal interests are in the fields of electronic circuit design and audio engineering. Dr. Hawksford studied at the University of Aston in Birmingham and gained both a First Class Honors B.Sc. and Ph.D. The Ph.D. program was supported by a BBC Research Scholarship, where the field of study was the application of delta modulation to color television and the development of a time compression/time multiplex system for combining luminance and chrominance signals.

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*.

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Wolfgang Wingerter was born in Landau, FRG, in 1962. He studied electrical engineering from 1982 to 1988 at the Technical University of Karlsruhe in West Germany and took part in the Joint Electrical Engineering Degree Scheme within Europe at the University of Karlsruhe, ESIEE Paris (France), and the University of Essex, U.K. He was a member of the Audio Research Group at the Department of Electronics Systems Engineering, Essex University, during 1987/1988.

He is now a fellow of CERN/Geneva where he has worked in the Radio Frequency Group of the SL Division since 1988. His current research interests lie particularly in the areas of analog-to-digital/digital-to-analog conversion, signal processing, and fast data acquisition.