A New Approach to FIR Digital Filters with Fewer Multipliers and Reduced Sensitivity

JOHN W. ADAMS, MEMBER, IEEE, AND ALAN N. WILLSON, JR., FELLOW IEEE

Abstract—A new approach to the design of efficient finite impulse response (FIR) digital filters is presented. The essence of the proposed method is to decompose the design problem into two parts: the realization of an efficient prefilter and the design of the corresponding amplitude equalizer. It is shown that this method can provide benefits in three areas: reduced computational complexity, reduced sensitivity to coefficient quantization, and reduced roundoff noise.

I. Introduction

A FUNDAMENTAL goal in digital filter design is to minimize the computational complexity of the filter realization. It is virtually always desirable to minimize the quantities M, b, and A, where M denotes the number of multipliers, b denotes the number of bits used for the multiplier coefficients, and A is the number of adders. The relative importance of the above quantities depends on the specific application, so that various complexity measures are in use. Typical complexity measures are: Mb, M, and $M+A$.

Mb is often used to characterize the complexity of digital filters implemented with special purpose hardware. Here multipliers are the slowest and most expensive components, and their cost depends on the number of bits. For the general purpose computer implementation, M by itself is an appropriate measure because the wordsize is usually more than adequate for digital filtering (especially when floating point arithmetic is used, as is common with general purpose computers). The measure $M+A$ is appropriate in the context of digital filters implemented in programmable signal processors with a pipelined architecture (which are becoming common in modern radar and sonar systems). Due to the pipelining, multiplication and addition typically execute equally fast so that $M+A$ is a reasonable digital filter complexity measure.

Our objective in this paper is to present a novel approach to linear phase finite impulse response (FIR) digital filter design which yields filters that have reduced computational complexity (according to all three of the measures discussed in the above) when compared to conventional filters. The sensitivity of the frequency response to coefficient quantization is also reduced, along with the quantization noise generated by the filter.

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J. W. Adams is with the Radar Systems Group of the Hughes Aircraft Company, El Segundo, CA 90009.

A. N. Willson, Jr. is with the Department of Electrical Engineering, University of California, Los Angeles, CA 90024.

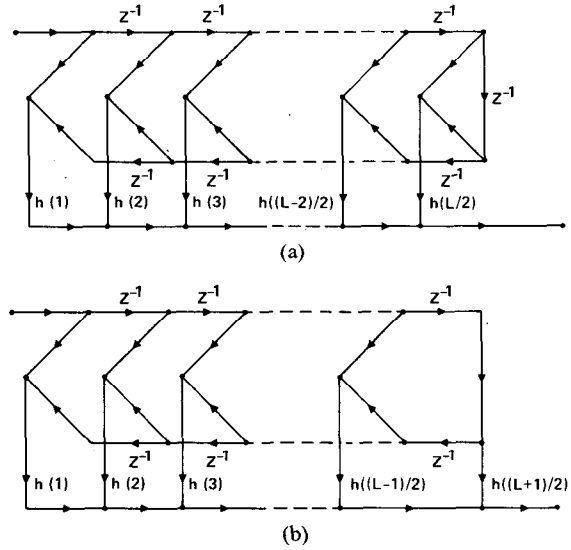


Fig. 1. Linear phase FIR digital filter structure. (a) Even length. (b) Odd length.

II. Review of the Conventional Approach to FIR Digital Filter Design

The optimal (in the minimax sense) FIR digital filter is defined as the filter for which the maximum weighted error in approximating a desired amplitude response function is minimized. For the lowpass case the optimal filter is characterized by the following set of parameters:

L	length of the impulse response
ω_P	passband edge frequency
ω_s	stopband edge frequency
DB_p	passband ripple in decibels
DB_s	stopband attenuation in decibels.

The standard form of the linear phase FIR digital filter structure is shown in Fig. 1. This structure takes advantage of the symmetry of the impulse response, so that the number of multipliers is approximately half the filter length.

$$\text{Number of multipliers} = \begin{cases} L/2, & \text{for even } L \\ (L+1)/2, & \text{for odd } L. \end{cases}$$

In the conventional approach to FIR digital filter design, one set of filter coefficients, $\{h(n)\}$, is designed to meet the overall amplitude response specifications. The $h(n)$ are typically computed by using the techniques developed in [1]–[6]. This minimizes the length of the impulse response,

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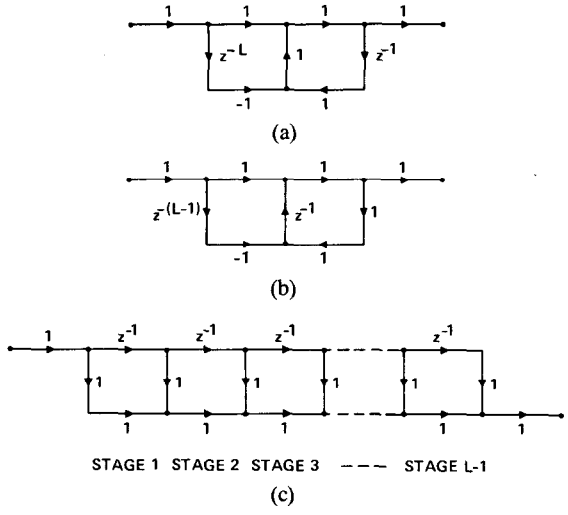


Fig. 2. (a) "Recursive realization of a running sum," as in [11]. (b) Equivalent to (a), but requires one less delay. (c) Direct-form implementation of the RRS.

but does not necessarily minimize the computational complexity.

III. Proposed Filter Design Approach

The essence of the proposed method is to decompose the design problem into two parts:

(1) Extract an efficient "prefilter" design from the specifications. The prefilter should have the best possible frequency response, with a minimal number of multipliers and adders.
(2) Cascade the prefilter with an "amplitude equalizer" to achieve the desired overall result.

This separation of the design problem into two parts allows the filter designer to concentrate on the computational complexity issue within the simplified context of the prefilter network.

In general, the number of possibilities for digital prefilter structures is unlimited. However, in order to have a specific example in this paper for illustrating the prefilter–equalizer design approach, we shall focus on one particular variety of prefilter, as shown in Fig. 2(a).[1] This structure is referred to as the "recursive realization of a running sum" in [11]. Fig. 2(b) shows an alternate structure which requires one less delay element.[2] In the interest of notational convenience, we shall refer to the "recursive running sum" filter network as the "RRS."

The RRS is a very simple and efficient filter structure. It requires only two adders and no multipliers at all, regardless of the filter length. (The length of the impulse response L is equal to the number of delays for the network given in Fig. 2(b).) The mathematically equivalent direct-form implementation of the RRS is shown in Fig. 2(c). The RRS is just an FIR digital filter with unity coefficients. However, it yields a "very respectable" (considering its simplicity) lowpass response which may be tuned by adjusting the filter length.

The prefilter should be chosen on the basis of relieving the amplitude equalizer from making a sharp transition from the passband to the stopband. The prefilter is intended to increase the effective transition bandwidth of the equalizer and thereby minimize its order. Approximate FIR digital filter design equations are derived in [12] and [13]; they show that a filter's length is inversely related to its transition bandwidth.

The frequency response of an RRS with length L is given by:[3]

$$P(e^{j\omega}) = \frac{\sin(\omega L/2)}{\sin(\omega/2)} e^{-j\omega(L-1)/2}.$$

The first null occurs at: $\omega_{\text{null}} = 2\pi/L$. All of the nulls of the prefilter's frequency response must of course lie in the stopband, which implies that: $L < 2\pi/\omega_s$. This constraint may be refined by noticing that the transition band of the equalizer can be effectively widened if the first null of the prefilter's response is placed just slightly above ω_s. This causes the prefilter to provide a large amount of attenuation near the stopband edge, so that the equalizer is not required to work very hard until the frequency of the prefilter's first sidelobe. According to the above discussion, the length of the RRS prefilter should be chosen as follows:

$$L_p = \text{ISLT}\{2\pi/\omega_s\}$$

where

$$L_p = \text{RRS prefilter length}$$

$$\text{ISLT}\{\cdot\} = \text{"Integer Slightly Less Than"}.$$

For a specific example, we shall consider the following set of lowpass filter requirements: $\omega_p = 0.042\pi$ rad/sample, $\omega_s = 0.14\pi$ rad/sample, $DB_p = 0.2$ dB, $DB_s = 35$ dB. (ω_p and ω_s are the same as used for the examples in [10].) In this case, $2\pi/\omega_s = 14.29$ so that $L_p = 14$ and $L_p = 13$ are promising candidates. The next step is to design an equalizer with minimum length for each prefilter candidate, such that the product of the prefilter and equalizer frequency responses meets the overall specifications. Appendix B describes how the Parks–McClellan computer program [4] can be modified to design the optimal equalizer. (The procedure outlined in Appendix B is sufficiently general to allow the design of the optimal equalizer for an arbitrary prefilter structure, not just the RRS.) Equalizers with lengths of 24 and 26 are required for RRS's with lengths 14 and 13, respectively. Clearly, the length 24 equalizer (in conjunction with the RRS of length 14) should be used. The filter coefficients for the equalizer are given in Appendix C.

[1] More sophisticated prefilter structures are discussed in [7] and will be presented in [8]. They include: prefilters based on an extension to the filter sharpening method of Kaiser and Hamming [9], prefilters derived from the method of Bateman and Liu [10], and highpass and bandpass prefilters. Also, prefilters composed of a simple cascade of RRS's are considered.

[2] This particular structure for the RRS was suggested by Prof. H. J. Orchard.

[3] An RRS of length L has an intrinsic dc gain of L. Various methods for implementing the RRS such that its dc gain is normalized to unity are discussed in Appendix A.

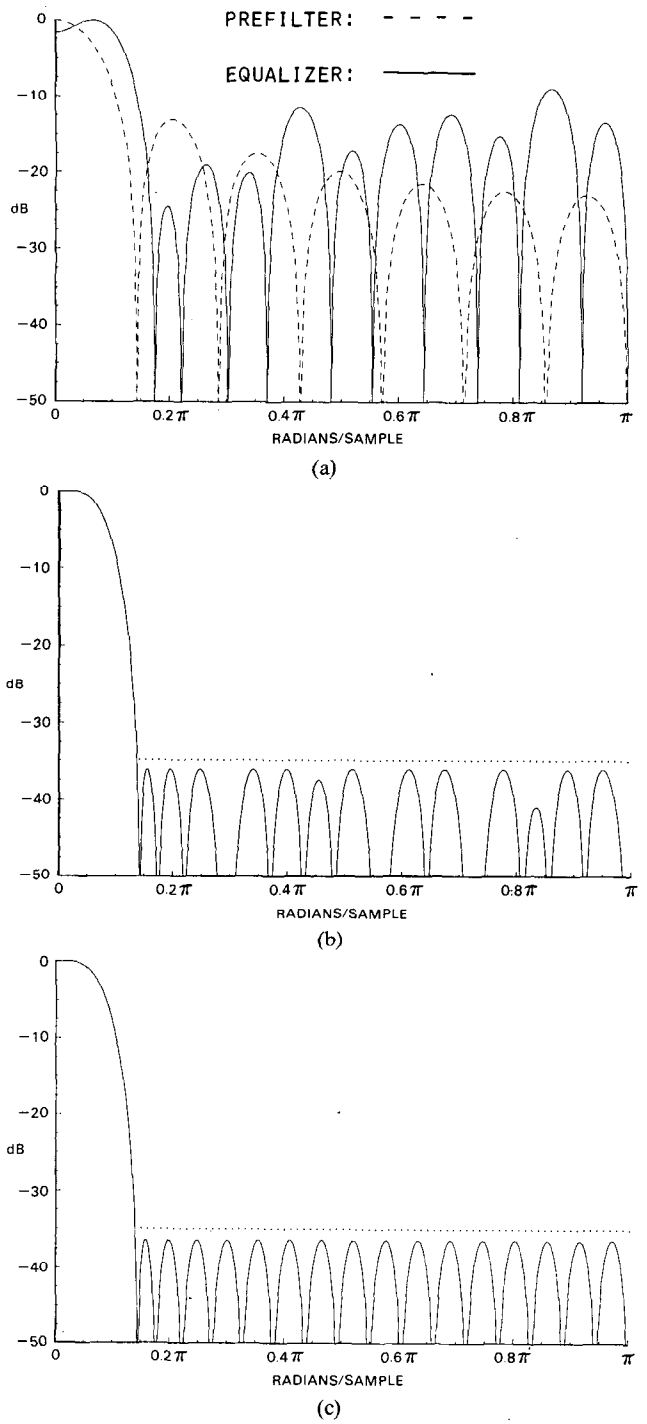


Fig. 3. (a) Individual amplitude responses for the prefilter and the equalizer. (b) Prefilter–equalizer cascade amplitude response. (c) Amplitude response of the conventional filter.

Fig. 3(a) shows overlays of the individual RRS prefilter and equalizer amplitude responses. Notice that not only does the prefilter allow a wider transition band for the equalizer, but also the stopband attenuation requirements on the equalizer are relaxed for those frequencies in the vicinity of nulls on the prefilter response. Fig. 3(b) provides the overall amplitude response of the prefilter–equalizer cascade. As a basis for comparison, a conventional filter was designed to meet the same specifications. The required length was 36 (35 delays), substantially more than the length of the equalizer. (The filter coefficients are given in Appendix C.) Fig. 3(c) shows the amplitude response of the conventional filter. A summary of the hardware requirements for the prefilter–equalizer cascade and the conventional filter is given in Table I. (For Table I it is assumed that both the conventional filter and the equalizer are implemented with the standard linear phase structure shown in Fig. 1.) In this example, the conventional filter uses 5.7 percent fewer delays, but 40 percent more adders and 50 percent more multipliers, than the prefilter–equalizer cascade. Clearly, the proposed filter design method has provided a very significant savings.

TABLE I
HARDWARE REQUIREMENTS SUMMARY FOR THE EXAMPLE

	DELAYS	ADDERS	MULTIPLIERS
PREFILTER	14	2	0
EQUALIZER	23	23	12
TOTAL	37	25	12
CONVENTIONAL FILTER	35	35	18

Quantization Noise

For practical purposes, a statistical model for the quantization noise contributed by each multiplier is often used. The quantization stepsize is commonly denoted as Q and is given by $Q = 2^{-(b-1)}$, with b denoting the number of bits used in the fixed-point arithmetic. The quantization error contributed by an individual multiplier is treated as noise that is uniformly distributed and having a variance of $Q^2/12$. For an FIR digital filter with M multipliers, the variance of the total quantization noise at the output is simply $MQ^2/12$. For the previous example, the conventional filter required 50 percent more multipliers than the equalizer and will consequently suffer from 50 percent more quantization noise. The quantization noise produced by the RRS depends on the particular implementation (it can be made to be negligible) and this is discussed in Appendix A.

Coefficient Quantization

Fig. 4(a)–4(c) illustrate the effects of coefficient quantization on the amplitude responses of the proposed and conventional filters. The proposed filter clearly exhibits superior performance with respect to coefficient-quantization sensitivity in this case. Moreover, there are some fundamental reasons why this should be true in general. A general analysis of the sensitivity of the frequency response to coefficient quantization is developed in the next section.

IV. Sensitivity of the Frequency Response to Coefficient Quantization

In practice the filter's multiplier coefficients must be represented by finite-length computer words. Typically each

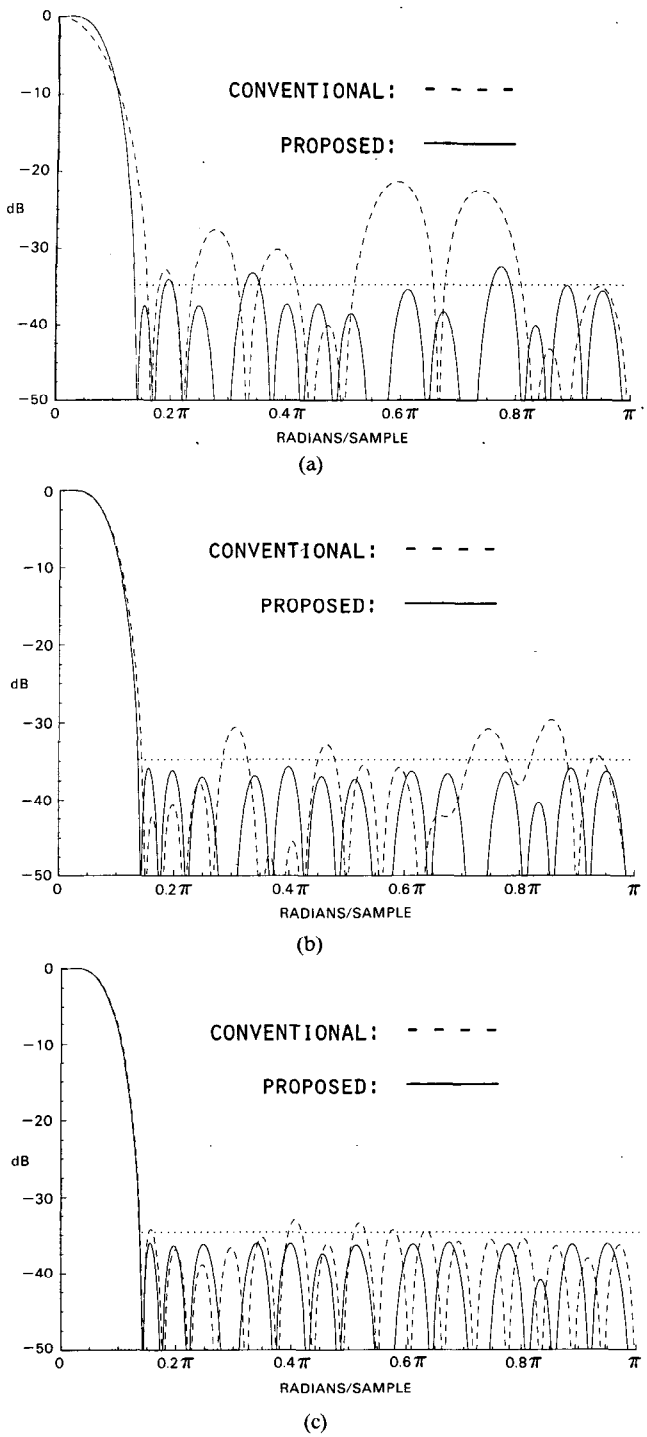


Fig. 4. Amplitude responses of the proposed and conventional filters with coefficients quantized to: (a) 6-bits, (b) 8-bits, (c) 10-bits.

ideal coefficient value is rounded to the nearest representable number, so that the coefficient-quantization error incurred is no more than half the quantization stepsize. We let $h(n)$ and $h'(n)$ denote the infinite precision and finite precision coefficients, respectively. Then the quantization error $e(n)$ is given by: $e(n) = h'(n) - h(n)$. We let $H(e^{j\omega})$, $H'(e^{j\omega})$, and $E(e^{j\omega})$ denote the Fourier transforms of the sequences $\{h(n)\}$, $\{h'(n)\}$, and $\{e(n)\}$, respectively. As the Fourier transformation is a linear operator, then we also have: $E(e^{j\omega}) = H'(e^{j\omega}) - H(e^{j\omega})$. Assuming that the $e(n)$

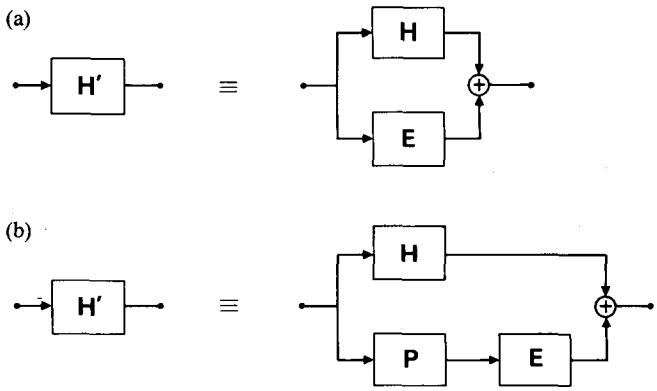


Fig. 5. Representation of a filter with quantized coefficients as the parallel connection of the original filter with ideal coefficients and an "error filter." (a) Conventional filter. (b) Prefilter–equalizer network.

are uncorrelated, it follows that $E(e^{j\omega})$ should be an approximately uniform and noise-like spectrum.

The preceding discussion was given in the context of conventional FIR digital filters. Now we shall consider the prefilter–equalizer cascade. We let $P(e^{j\omega})$ and $Q(e^{j\omega})$ define the prefilter and equalizer frequency responses, respectively. Also, $H(e^{j\omega})$ represents the frequency response of the cascade, so that: $H(e^{j\omega}) = P(e^{j\omega})Q(e^{j\omega})$. The frequency responses corresponding to the quantized coefficients will be denoted $H'(e^{j\omega})$, $P'(e^{j\omega})$, and $Q'(e^{j\omega})$. The quantization error spectrum for the equalizer will be defined as $E(e^{j\omega})$, so that: $Q'(e^{j\omega}) = Q(e^{j\omega}) + E(e^{j\omega})$. For prefilters such as the RRS, which are immune to coefficient quantization, we have: $P'(e^{j\omega}) = P(e^{j\omega})$. We may express $H'(e^{j\omega})$ as a sum of $H(e^{j\omega})$ and an error spectrum as follows. (Notice that the dependence on $e^{j\omega}$ is suppressed in the following for notational convenience.)

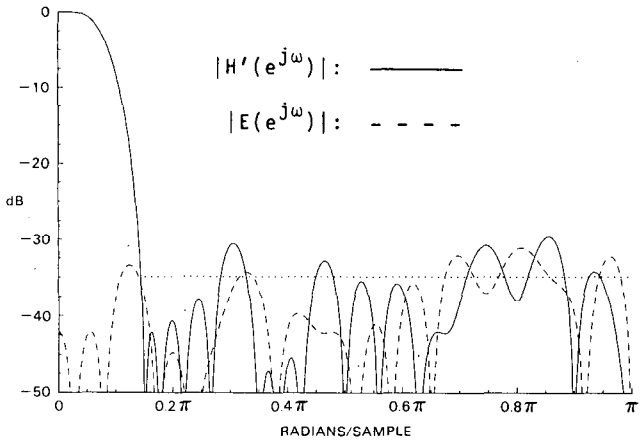
$$H' = P'Q' = PQ' = P(Q + E) = PQ + PE$$

$$H' = H + PE.$$

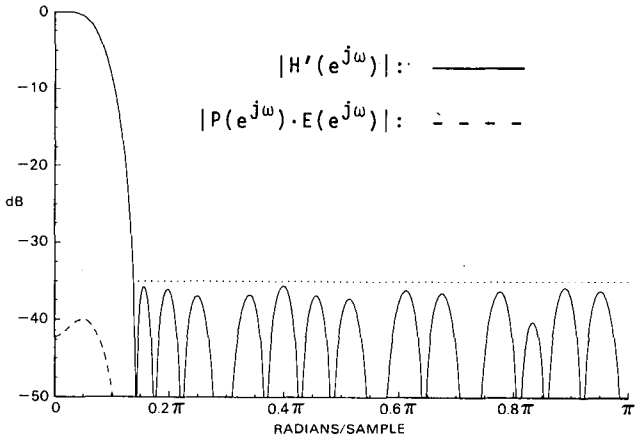
Clearly, the prefilter attenuates the equalizer's quantization error spectrum for frequencies in the stopband. Fig. 5 shows block diagram representations of the above relations for the conventional, and the prefilter–equalizer cascade filters. Fig. 6(a) shows plots of $|H'(e^{j\omega})|$ and $|E(e^{j\omega})|$ for the conventional filter example discussed in Section III, with 8-bit coefficients. Similarly, Fig. 6(b) illustrates $|H'(e^{j\omega})|$ and $|P(e^{j\omega})E(e^{j\omega})|$ for the prefilter–equalizer cascade version of this example, again using 8-bit coefficients. The attenuation of the coefficient-quantization error spectrum is clearly evident in the prefilter–equalizer cascade for frequencies in the stopband.

By employing Parseval's theorem it is easy to demonstrate that the total coefficient-quantization error energy at all frequencies, including those in the passband, is expected to be less in the prefilter–equalizer cascade than in the conventional filter. Parseval's theorem relates the total energy in the error spectrum to the energy in the coefficient-quantization error sequence as follows.

$$\frac{1}{2\pi}\int_{-\pi}^{\pi} |E(e^{j\omega})|^2 \, d\omega = \sum_{n=1}^{L} e^2(n).$$



(a)



(b)

Fig. 6. Amplitude response of the filter with quantized coefficients and the "error spectrum." (a) Conventional filter with 8-bit coefficients. (b) Prefilter–equalizer network with 8-bit coefficients. (Notice that the level of the error spectrum is below −50 dB for frequencies in the stopband so that the effect on the overall amplitude response is very small.)

Assuming that the $\{e(n)\}$ are uniformly distributed, the expected value of the total error energy is given by

$$\mathcal{E}\left\{\frac{1}{2\pi}\int_{-\pi}^{\pi}|E(e^{j\omega})|^2\,d\omega\right\} = \begin{cases} MQ^2/6, & \text{for even } M \\ MQ^2/6 - Q^2/12, & \text{for odd } M \end{cases}$$

where

$$\mathcal{E}\{\cdot\} = \text{expectation operator.}$$

According to the above, the total energy expected in the coefficient-quantization error spectrum is proportional to the number of multipliers. Since the equalizer has a reduced number of multipliers compared to the conventional filter, the expected total error energy will also be reduced. Furthermore, as mentioned previously, the prefilter attenuates the coefficient-quantization error energy in the stopband, providing an additional improvement factor. For these reasons, the frequency response of the prefilter–equalizer cascade should be much less sensitive to quantization of the filter coefficients.

V. Limitations of the Method

As discussed in Appendix B, there exists a unique optimal equalizer for use in conjunction with any given prefilter. Furthermore, the Parks–McClellan computer program can easily be modified to automatically design the optimal equalizer, so that for practical purposes the design of the equalizer is trivial. In contrast, the design of an efficient prefilter is in general not as easy.

The RRS can be a very efficient prefilter for applications where the desired filter response is reasonably close to the frequency response of the corresponding RRS. Otherwise, a different type of prefilter should be used. In particular, the simple RRS is not suitable for wide-band or bandpass filtering applications. It would be desirable for the filter designer to have a "catalog" of efficient prefilter structures for various applications, but a complete catalog certainly does not exist. In particular, the authors are not aware of any prefilter structures that would be efficient for wide-band applications. More advanced (compared to the RRS) prefilter structures are discussed in [7] and will be presented in [8], but there is still a need for a more complete set of designs.

VI. Conclusion

Historically, equalization has been necessary in certain situations due to imperfections in various parts of a system. For example, the filter designer may have been forced to compensate for imperfect amplifiers, transmission lines, etc. Here, however, we propose to *deliberately* use the equalization concept to deal with the computational complexity issue in digital filters. We have shown that at the expense of a minor increase in the number of delays, a major reduction in the number of multipliers and adders can be obtained. Moreover, the sensitivity of the frequency response to coefficient quantization is reduced so that fewer multiplier-coefficient bits are needed.

The prefilter–equalizer approach to minimizing digital filter complexity is extensive in scope and cannot be treated exhaustively in a single publication. Here we have considered only linear phase FIR digital filters; clearly the method could also be applied to IIR filters and this suggests an area for further research. Also, the number of possibilities for efficient digital prefilter structures is certainly unlimited.

Appendix A

As for any digital filter, the details of the implementation of the RRS must depend on the specific digital signal processing environment in which it is used. In particular, the scaling of the gain of the RRS would normally depend on many factors, including: the desired overall system gain, the nature of the signal being filtered (its spectral composition and amplitude), and also various hardware constraints (for example, scaling is usually not even an issue for digital filters implemented in a general purpose computer with floating-point arithmetic).

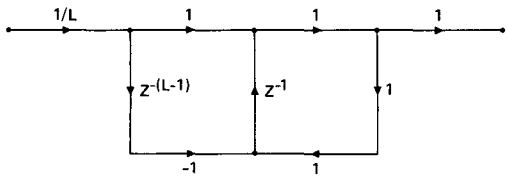


Fig. 7. A simple approach for scaling the dc gain of the RRS to unity.

The intrinsic dc gain of an RRS with length L is simply L. If it is desired to normalize the dc gain to unity then there are various options for doing this, depending on the hardware environment. First, we shall consider the simplest approach, which is to scale the input signal by the factor $1/L$, as shown in Fig. 7. (Scaling the input signal instead of the output is done to prevent overflow within the filter.) This introduces a noise source at the input to the RRS with a variance of $Q^2/12$, where Q denotes the quantization stepsize. The noise variance at the output is $LQ^2/12$, and the noise power spectral density is shaped by the frequency response of the RRS. An alternate scaling strategy is discussed in the following which allows the "$Q^2/12$ noise" to be injected at the output of the RRS instead of the input.

The RRS is very efficient because it requires only 2 adders, regardless of its order. In the hardware realization of the RRS-equalizer network, the overall cost of the filter would be only slightly increased if a few additional bits were used for the 2 adders in the RRS. (A comparatively large increase in the expense would be expected if additional bits were used for all of the multipliers and adders in the equalizer.) These extra "overflow bits" could be used to allow the signal level to build up in the two RRS adders without danger of overflow, thus eliminating the need to scale the input signal. For example, four additional bits could be used in a length sixteen RRS to guarantee against overflow, and a dc gain of unity could be obtained by reading out (or "masking") the data from all but the four least significant bits (LSB's). For this configuration, the quantization noise at the output of the RRS has a variance of $Q^2/12$. If the RRS's length is not exactly a power of 2, say $L=14$, then the above procedure can be used by rounding up to the nearest power of 2. This will produce gains between 0.5 and 1.0. In the case of $L=14$, four additional bits would be used for the two adders in the RRS, and the output data would be masked from all but the four LSB's. This provides a gain of $14/16=0.875$. The factor of 0.875 would normally be compensated by adjusting the scaling of some other signal processing operation in the system (for example, the equalizer), rather than using a separate multiplier.

In many digital signal processing systems, the word size in the signal processor is larger than the word size for the A/D converter which provides the input data. This is done to minimize the cost of the A/D converter and yet have a sufficient word length in the processor to allow high-order FFT's to be performed. For example, a signal processor with a 16-bit word length may be used with an 8-bit A/D converter. If an RRS-equalizer network was used at the front-end of such a system, the 8 bits from the A/D converter could be loaded into the 8 LSB's of the 16-bit processor words, and no scaling at all would be needed for the RRS. Since only additions are performed in the RRS, no quantization noise would be produced.

Appendix B

The Parks–McClellan computer program [4] calculates the optimal filter coefficients on the basis of minimizing the maximum value of the weighted error function, $\Delta(e^{j\omega})$. (There exists a unique optimal solution to the problem, as discussed in [4].)

$$\Delta(e^{j\omega}) = |W(e^{j\omega})\cdot\{G(e^{j\omega}) - G_d(e^{j\omega})\}|$$

where

$G_d(e^{j\omega})$ desired gain function
$G(e^{j\omega})$ actual gain function
$W(e^{j\omega})$ relative weighting or "cost" function.

For lowpass filters the unmodified Parks–McClellan computer program [4] uses $G_d(e^{j\omega})$ and $W(e^{j\omega})$ as follows:

$$G_d(e^{j\omega}) = \begin{cases} 1, & 0 \leqslant \omega \leqslant \omega_p \\ 0, & \omega_s \leqslant \omega \leqslant \pi \end{cases}$$

$$W(e^{j\omega}) = \begin{cases} 1, & 0 \leqslant \omega \leqslant \omega_p \\ K, & \omega_s \leqslant \omega \leqslant \pi. \end{cases}$$

Now we will show that the Parks–McClellan algorithm can be modified to allow the design of the optimal equalizer for use in conjunction with any given prefilter. Let $\tilde{P}(e^{j\omega})$ and $\tilde{Q}(e^{j\omega})$ denote the prefilter and equalizer gain responses, respectively. The product, $\tilde{P}(e^{j\omega})\cdot\tilde{Q}(e^{j\omega})$, is the actual overall gain function, but only $\tilde{Q}(e^{j\omega})$ is subject to optimization. The design problem may be reformulated in the terms of the Parks–McClellan algorithm as follows:

$$\Delta(e^{j\omega}) = |W(e^{j\omega})\cdot\{\tilde{P}(e^{j\omega})\cdot\tilde{Q}(e^{j\omega}) - G_d(e^{j\omega})\}|$$

$$\Delta(e^{j\omega}) = |W(e^{j\omega})\cdot\tilde{P}(e^{j\omega})\cdot\{\tilde{Q}(e^{j\omega}) - G_d(e^{j\omega})/\tilde{P}(e^{j\omega})\}|$$

$$\Delta(e^{j\omega}) = |\tilde{W}(e^{j\omega})\cdot\{\tilde{G}(e^{j\omega}) - \tilde{G}_d(e^{j\omega})\}|$$

where

$$\tilde{G}_d(e^{j\omega}) = \begin{cases} 1/\tilde{P}(e^{j\omega}), & 0 \leqslant \omega \leqslant \omega_p \\ 0, & \omega_s \leqslant \omega \leqslant \pi \end{cases}$$

$$\tilde{W}(e^{j\omega}) = \begin{cases} |\tilde{P}(e^{j\omega})|, & 0 \leqslant \omega \leqslant \omega_p \\ K|\tilde{P}(e^{j\omega})|, & \omega_s \leqslant \omega \leqslant \pi \end{cases}$$

$$\tilde{G}(e^{j\omega}) = \tilde{Q}(e^{j\omega}), \qquad 0 \leqslant \omega \leqslant \pi.$$

The weighting function, $\tilde{W}(e^{j\omega})$, is required to be strictly positive. However, for frequencies in the stopband it is possible for $\tilde{P}(e^{j\omega})$ to become identically zero. (We note that a division by the weighting function occurs in the Parks–McClellan computer program.) A convenient remedy is for the program to automatically limit $\tilde{W}(e^{j\omega})$ to be greater than some small positive number, ε. (For example, $\varepsilon = 10^{-6}$.) It has been observed that the filter design does

not depend on the exact value chosen for ε, provided that it is reasonably small. (An alternate solution, which is slightly more complicated, is for the program to automatically specify "don't care" intervals in the vicinity of nulls on the prefilter's frequency response.)

The above discussion was given in the particular context of lowpass filter design for the sake of concreteness. However, the same approach can of course also be used for the bandpass and highpass cases.

Appendix C

The coefficients for the filters discussed in Section II are given below. For consistency, both the equalizer and the conventional filter were scaled to have a dc gain of unity.

Equalizer	Conventional Filter
−0.12163029	−0.01084545
0.02319508	−0.00733512
−0.02155842	−0.00876169
−0.00969477	−0.00942935
−0.00605022	−0.00898000
0.01019741	−0.00704726
0.05381588	−0.00336417
0.07186058	0.00222462
0.11563688	0.00969186
0.08261736	0.01891595
0.14487070	0.02952445
0.15673977	0.04103817
	0.05282516
	0.06417377
	0.07436218
	0.08270440
	0.08861664
	0.09168579

References

[1] T. W. Parks and J. H. McClellan, "Chebyshev approximation for nonrecursive digital filters with linear phase," *IEEE Trans. Circuit Theory*, vol. CT-19, pp. 189–194, Mar. 1972.

[2] ——, "A program for the design of linear phase finite impulse response digital filters," *IEEE Trans. Audio Electroacoust.*, vol. AU-20, pp. 195–199, Aug. 1972.

[3] J. H. McClellan and T. W. Parks, "A unified approach to the design of optimum FIR linear phase digital filters," *IEEE Trans. Circuit Theory*, vol. CT-20, pp. 697–701, Nov. 1973.

[4] J. H. McClellan, T. W. Parks, and L. R. Rabiner, "A computer program for designing optimum FIR linear phase filters," *IEEE Trans. Audio Electroacoust.*, vol. AU-21, pp. 506–526, Dec. 1973.

[5] L. R. Rabiner, J. H. McClellan, and T. W. Parks, "FIR digital filter design techniques using weighted Chebyshev approximation," *Proc. IEEE*, vol. 63, pp. 595–610, Apr. 1975.

[6] L. R. Rabiner, "Approximate design relationships for lowpass FIR digital filters," *IEEE Trans. Audio Electroacoust.*, vol. AU-21, pp. 456–460, Oct. 1973.

[7] J. W. Adams, "New approaches to finite impulse response digital filter design," Dept. of Electrical Engineering, UCLA, Los Angeles, CA, May 1982.

[8] J. W. Adams and A. N. Willson, "Some efficient digital prefilter structures," to be published.

[9] J. F. Kaiser and R. W. Hamming, "Sharpening the response of a symmetric nonrecursive filter by multiple use of the same filter," *IEEE Trans. Acoustics, Speech, Signal Processing*, vol. ASSP-25, pp. 415–422, Oct. 1977.

[10] M. R. Bateman and B. Liu, "An approach to programmable CTD filters using coefficients 0, +1, and −1," *IEEE Trans. Circuits Syst.*, vol. CAS-27, pp. 451–456, June 1980.

[11] L. R. Rabiner and B. Gold, *Theory and Application of Digital Signal Processing*, p. 631, Englewood Cliffs, NJ: Prentice Hall, 1975.

[12] O. Herrman and L. R. Rabiner, "Practical design rules for optimum finite impulse response lowpass digital filters," *Bell Syst. Tech. J.*, vol. 52, pp. 769–799, July 1973.

[13] J. F. Kaiser, "Nonrecursive digital filter design using the I_0-sinh window function," in *Proc. 1974 Int. Symp. Circuits and Systems*, pp. 20–23, Apr. 1974.



John W. Adams (S'75–M'81) was born in Santa Monica, CA, on February 8, 1954. He received the B.S., M.S., Engr., and Ph.D. degrees in electrical engineering from the University of California, Los Angeles, in 1976, 1976, 1978, and 1982, respectively.

In 1978 he joined the Radar Systems Group of the Hughes Aircraft Company, where he now holds the position of Senior Staff Engineer in the Systems Engineering Department. His current research interests are in the areas of digital signal processing and synthetic aperture radar.

Dr. Adams is a member of Phi Beta Kappa, Tau Beta Pi and Sigma Xi.



Alan N. Willson, Jr. (S'66–M'67–SM'73–F'78) was born in Baltimore, MD, on October 16, 1939. He received the B.E.E. degree from the Georgia Institute of Technology, Atlanta, GA, in 1961, and the M.S. and Ph.D. degrees from Syracuse University, Syracuse, NY, in 1965 and 1967, respectively.

From 1961 to 1964 he was with the IBM Corporation, in Poughkeepsie, NY. He was an Instructor in Electrical Engineering at Syracuse University from 1965 to 1967. From 1967 to 1973 he was a Member of the Technical Staff at Bell Laboratories, Murray Hill, NJ. Since 1973 he has been on the faculty of the University of California, Los Angeles, where he is now Professor of Engineering and Applied Science, in the Electrical Engineering Department. In addition, he served the UCLA School of Engineering and Applied Science as Assistant Dean for Graduate Studies, from 1977 through 1981. He has been engaged in research concerning the stability of distributed circuits, properties of nonlinear networks, theory of active circuits, digital signal processing, and analog circuit fault diagnosis. He is editor of the book *Nonlinear Networks: Theory and Analysis*, IEEE Press, 1974.

Dr. Willson is a member of Eta Kappa Nu, Sigma Xi, Tau Beta Pi, the Society for Industrial and Applied Mathematics, and the American Society for Engineering Education. From June 1977 to June 1979 he served as Editor of the IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS, and during 1980 he was Vice-President of the IEEE Circuits and Systems Society. He now holds the office of President-Elect of the IEEE Circuits and Systems Society. He is the recipient of the 1978 Guillemin–Cauer Award of the IEEE Circuits and Systems Society, for co-authoring the best paper published in their TRANSACTIONS during the previous year. He is the recipient of the 1982 George Westinghouse Award of the ASEE, and the 1982 Distinguished Faculty Award of the Engineering Alumni Association.