Digital Parametric Equalizer Design with Prescribed Nyquist-Frequency Gain*

SOPHOCLES J. ORFANIDIS, *AES Member*

Department of Electrical and Computer Engineering, Rutgers University, Piscataway, NJ 08855-0909, USA

A new type of second-order digital parametric equalizer is proposed whose frequency response matches closely that of its analog counterpart throughout the Nyquist interval and does not suffer from the prewarping effect of the bilinear transformation near the Nyquist frequency. Closed-form design equations and direct-form and lattice realizations are derived.

0 INTRODUCTION

Conventional bilinear-transformation-based methods of designing second-order digital parametric equalizers [1]–[11] result in frequency responses that fall off faster than the corresponding analog equalizers near the Nyquist frequency due to the prewarping nature of the bilinear transformation. This effect becomes particularly noticeable when the peak frequencies and widths are relatively high. Fig. 1 illustrates this effect.

In this paper we introduce an additional degree of freedom into the design, namely, the gain at the Nyquist frequency, and derive a new class of digital parametric equalizers that closely match their analog counterparts over the entire Nyquist interval and do not suffer from the prewarping effect of the bilinear transformation.

The design specifications are the quantities $\{f_s, f_0, \Delta f, G_0, G_1, G, G_B\}$, that is, the sampling rate f_s, the boost or cut peak frequency f_0, the bandwidth Δf, the reference gain G_0 at zero frequency, the gain G_1 at the Nyquist frequency $f_s/2$, the boost or cut peak gain G at f_0, and the bandwidth gain G_B (the level at which the bandwidth Δf is measured).

All previous methods of designing second-order equalizers assume $G_1 = G_0$ (usually set equal to unity). In these methods the bilinear transformation is used to transform an analog equalizer with equivalent specifications into the digital one. As remarked by Bristow-Johnson [9], all of these designs are essentially equivalent to each other, up to a different definition of the bandwidth Δf and bandwidth gain G_B. For the equivalent analog equalizer, the quantity $G_0 = G_1$ represents the gain at direct current and at infinity, with the latter being mapped onto the Nyquist frequency $f_s/2$ by the bilinear transformation.

In the method proposed here we allow G_1 to be different from G_0. In particular, we set G_1 equal to the gain an analog equalizer would have at $f_s/2$ if it were not bilinearly transformed. This condition on G_1, together with the requirements that the gain at direct current be G_0, that there be a peak maximum (or minimum) at f_0, that the peak gain be G, and that the bandwidth be Δf at level G_B, provide five constraints that fix uniquely the five coefficients of the second-order digital filter.

The resulting digital filter matches the corresponding analog filter as much as possible, given that there are only five parameters to adjust. The matching is exact at $f = 0, f_0, f_s/2$, and the two filters have the same bandwidth Δf. These design goals are illustrated in Fig. 2.

Thus such a digital equalizer can be used to better emulate the sound quality achieved by an analog equalizer. This is the main motivation of this paper. Moreover, setting $G_0 = 0$, we also obtain more realistic modeling of resonant filters of prescribed peaks and widths for use in music and speech synthesis applications.

In the following sections we summarize the conventional analog and digital equalizer designs, present the new design and some simulations, and discuss direct and lattice form realizations and the issue of bandwidth. We also give a small MATLAB function for the new design.

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1 CONVENTIONAL ANALOG AND DIGITAL EQUALIZERS

Here we review briefly the design of analog and digital equalizers, following the discussion of [11]. A second-order analog equalizer with gain G_0 at direct current and at infinity has the transfer function

$$H(s) = \frac{G_0 s^2 + Bs + G_0 \Omega_0^2}{s^2 + As + \Omega_0^2} \tag{1}$$

and the magnitude response

$$|H(\Omega)|^2 = \frac{G_0^2(\Omega^2 - \Omega_0^2)^2 + B^2\Omega^2}{(\Omega^2 - \Omega_0^2)^2 + A^2\Omega^2} \tag{2}$$

where $\Omega = 2\pi f$ is the physical frequency in radians per second and $\Omega_0 = 2\pi f_0$ the peak frequency. The filter coefficients A and B are fixed by the two requirements that the gain be G at Ω_0 and that the bandwidth be measured at level G_B. These requirements can be stated as follows:

$$|H(\Omega_0)|^2 = G^2, \qquad |H(\Omega)|^2 = G_B^2 \tag{3}$$

where the solutions of the second equation are the right and left band-edge frequencies, say Ω_2 and Ω_1. They satisfy the geometric-mean property,

$$\Omega_1 \Omega_2 = \Omega_0^2 . \tag{4}$$

Defining the bandwidth $\Delta\Omega = 2\pi\Delta f$ as the difference of the band-edge frequencies, $\Delta\Omega = \Omega_2 - \Omega_1$, the two conditions in Eq. (3) determine the filter coefficients as follows:

$$A = \sqrt{\frac{G_B^2 - G_0^2}{G^2 - G_B^2}}\,\Delta\Omega, \qquad B = GA . \tag{5}$$

The equalizer gain at a desired Nyquist frequency $f_s/2$ can be obtained by evaluating Eq. (2) at $\Omega_s = 2\pi(f_s/2) = \pi f_s$, giving

$$G_1^2 = \frac{G_0^2(\Omega_s^2 - \Omega_0^2)^2 + B^2\Omega_s^2}{(\Omega_s^2 - \Omega_0^2)^2 + A^2\Omega_s^2} . \tag{6}$$

A digital equalizer can be designed by applying the bilinear transformation to an equivalent analog filter of the form of Eq. (1). The bilinear transformation is defined here as

$$s = \frac{1 - z^{-1}}{1 + z^{-1}}, \qquad \Omega = \tan\left(\frac{\omega}{2}\right), \qquad \omega = \frac{2\pi f}{f_s} \tag{7}$$

where Ω is now the prewarped version of the physical frequency ω. The physical peak and bandwidth frequencies are in units of radians per sample,

$$\omega_0 = \frac{2\pi f_0}{f_s}, \qquad \Delta\omega = \frac{2\pi\Delta f}{f_s} . \tag{8}$$

The prewarped versions of the peak and band-edge frequencies are $\Omega_0 = \tan(\omega_0/2)$, $\Omega_1 = \tan(\omega_1/2)$, and $\Omega_2 = \tan(\omega_2/2)$. They satisfy the prewarped geometric-mean property,

$$\tan\left(\frac{\omega_1}{2}\right)\tan\left(\frac{\omega_2}{2}\right) = \tan^2\left(\frac{\omega_0}{2}\right) \tag{9}$$

and the following relationship between the physical bandwidth $\Delta\omega = \omega_2 - \omega_1$ and its prewarped version $\Delta\Omega = \Omega_2 - \Omega_1$:

$$\Delta\Omega = (1 + \Omega_0^2)\tan\left(\frac{\Delta\omega}{2}\right) . \tag{10}$$

Replacing s by its bilinear transformation in Eq. (1) gives, after some algebraic simplifications, the digital transfer function

$$H(z) = \frac{\left(\dfrac{G_0 + G\beta}{1 + \beta}\right) - 2\left(\dfrac{G_0 \cos\omega_0}{1 + \beta}\right)z^{-1} + \left(\dfrac{G_0 - G\beta}{1 + \beta}\right)z^{-2}}{1 - 2\left(\dfrac{\cos\omega_0}{1 + \beta}\right)z^{-1} + \left(\dfrac{1 - \beta}{1 + \beta}\right)z^{-2}} \tag{11}$$

where the parameter β is given by

$$\beta = \sqrt{\frac{G_B^2 - G_0^2}{G^2 - G_B^2}}\tan\left(\frac{\Delta\omega}{2}\right) . \tag{12}$$

By design the gain of this digital filter at the Nyquist frequency is equal to G_0, whereas that of a physical analog filter is G_1, as given by Eq. (6). This can be seen directly from Eq. (11) by setting $z = -1$, or from the equivalent analog filter by taking the limit of Eq. (1) as $s \to \infty$. Fig. 1 compares the conventional analog and digital equalizer designs.

2 DIGITAL EQUALIZER WITH PRESCRIBED NYQUIST-FREQUENCY GAIN

Because the bilinear transformation maps $z = -1$ onto $s = \infty$, in order to design a digital filter with prescribed Nyquist-frequency gain G_1, we may start by designing an equivalent analog filter whose gain at $s = \infty$ is G_1. The transfer function of such a filter is the modified form of Eq. (1),

$$H(s) = \frac{G_1 s^2 + Bs + G_0 W^2}{s^2 + As + W^2} . \tag{13}$$

It has gain G_1 at $s = \infty$, and G_0 at $s = 0$. Its magnitude response is

$$|H(\Omega)|^2 = \frac{(G_1\Omega^2 - G_0W^2)^2 + B^2\Omega^2}{(\Omega^2 - W^2)^2 + A^2\Omega^2}. \tag{14}$$

The parameter W is no longer equal to the peak frequency Ω_0, but is related to it. The filter coefficients A, B, W^2 can be determined by requiring the three conditions that $|H(\Omega)|^2$ have a maximum (or minimum) at Ω_0, that the peak gain be G, and that the band-edge frequencies be measured at level G_B,

$$\frac{\partial}{\partial\Omega^2}|H(\Omega_0)|^2 = 0, \qquad |H(\Omega_0)|^2 = G^2, \qquad |H(\Omega)|^2 = G_B^2. \tag{15}$$

The solutions of the third equation are the left and right band-edge frequencies Ω_1, Ω_2, which define the analog bandwidth as the difference $\Delta\Omega = \Omega_2 - \Omega_1$. Solving Eqs. (15) (see Appendix 1 for details) gives the filter coefficients

$$W^2 = \sqrt{\frac{G^2 - G_1^2}{G^2 - G_0^2}}\,\Omega_0^2, \qquad A = \sqrt{\frac{C + D}{|G^2 - G_B^2|}},$$

$$B = \sqrt{\frac{G^2C + G_B^2D}{|G^2 - G_B^2|}} \tag{16}$$

where C and D are given in terms of the center frequency Ω_0, bandwidth $\Delta\Omega$, and gains as follows:

$$C = (\Delta\Omega)^2\,|G_B^2 - G_1^2| - 2W^2\left[|G_B^2 - G_0G_1| - \sqrt{(G_B^2 - G_0^2)(G_B^2 - G_1^2)}\right]$$

$$D = 2W^2\left[|G^2 - G_0G_1| - \sqrt{(G^2 - G_0^2)(G^2 - G_1^2)}\right]. \tag{17}$$

Moreover, the band-edge frequencies satisfy the modified geometric-mean property,

$$\Omega_1\Omega_2 = \sqrt{\frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}}\,W^2 = \sqrt{\frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}}\sqrt{\frac{G^2 - G_1^2}{G^2 - G_0^2}}\,\Omega_0^2. \tag{18}$$

Eqs. (16) and (17) implement the design of a second-order analog filter of given peak frequency Ω_0, width $\Delta\Omega$, and prescribed zero frequency, high-frequency, peak, and bandwidth gains G_0, G_1, G, and G_B. Note that the absolute values in Eqs. (16) and (17) are needed only when designing a cut, as opposed to a boost.

The desired digital filter can be designed now by the bilinear transformation applied to this analog filter. To complete the design, the given physical frequency parameters ω_0 and $\Delta\omega$ of Eq. (8) must be mapped onto those of the equivalent analog filter. This can be done via the transformations (see Appendix 1)

$$\Omega_0 = \tan\left(\frac{\omega_0}{2}\right), \qquad \Delta\Omega = \left(1 + \sqrt{\frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}}\sqrt{\frac{G^2 - G_1^2}{G^2 - G_0^2}}\,\Omega_0^2\right)\tan\left(\frac{\Delta\omega}{2}\right). \tag{19}$$

Applying the bilinear transformation (7) to Eq. (13) gives rise to the digital filter transfer function

$$H(z) = \frac{\left(\dfrac{G_1 + G_0W^2 + B}{1 + W^2 + A}\right) - 2\left(\dfrac{G_1 - G_0W^2}{1 + W^2 + A}\right)z^{-1} + \left(\dfrac{G_1 + G_0W^2 - B}{1 + W^2 + A}\right)z^{-2}}{1 - 2\left(\dfrac{1 - W^2}{1 + W^2 + A}\right)z^{-1} + \left(\dfrac{1 + W^2 - A}{1 + W^2 + A}\right)z^{-2}}. \tag{20}$$

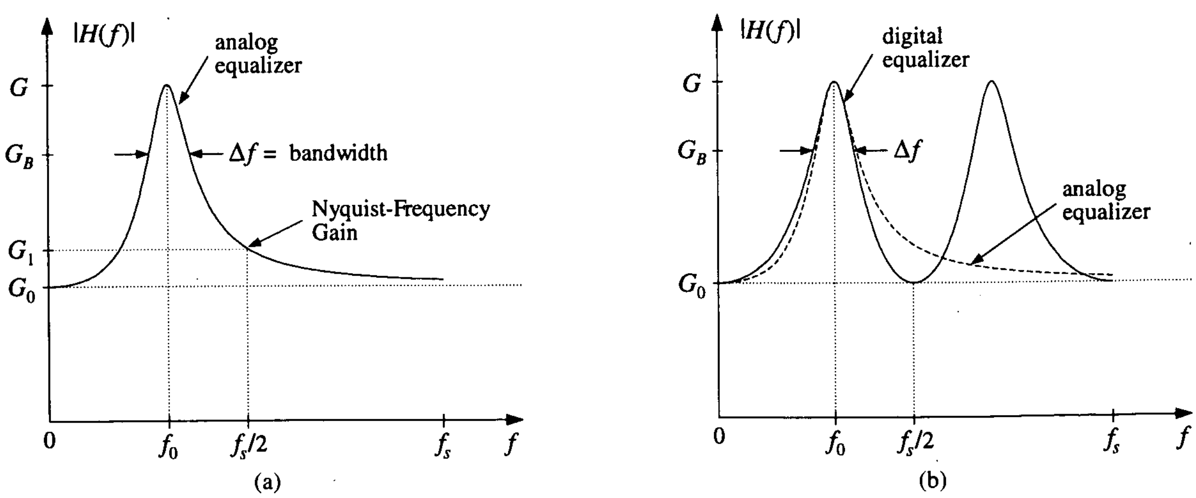


Fig. 1. Conventional analog and digital equalizers. Digital design has $G_1 = G_0$ at $f_s/2$.

To summarize, given the set of digital filter specifications $\{\omega_0, \Delta\omega, G_0, G_1, G, G_B\}$, use Eqs. (19) to calculate the prewarped analog frequencies. Then use Eqs. (16) and (17) to calculate the parameters $\{A, B, W^2\}$, from which the digital filter coefficients of Eq. (20) are determined.

In the special case, $G_1 = G_0$ we recover the results of Section 1. Indeed, we have $W = \Omega_0$, $D = 0$, and $C = (\Delta\Omega)^2|G_B^2 - G_0^2|$, and A, B reduce to Eq. (5). Similarly, Eq. (20) reduces to Eq. (11).

So far the Nyquist-frequency gain G_1 has been chosen arbitrarily. However, for the digital filter to match the corresponding (physical) analog filter as much as possible, the gain G_1 must match the analog filter's gain at $f_s/2$, as given by Eq. (6). Using Eq. (5), we can rewrite Eq. (6) in terms of the normalized digital frequencies ω_0 and $\Delta\omega$ of Eq. (8) as follows:

$$G_1^2 = \frac{G_0^2(\omega_0^2 - \pi^2)^2 + G^2\pi^2(\Delta\omega)^2 (G_B^2 - G_0^2)/(G^2 - G_B^2)}{(\omega_0^2 - \pi^2)^2 + \pi^2(\Delta\omega)^2 (G_B^2 - G_0^2)/(G^2 - G_B^2)} . \tag{21}$$

Fig. 2 compares the new digital equalizer with the conventional analog and digital designs. The overall design method contained in Eqs. (16)–(21) is implemented by the MATLAB function peq.m of Appendix 2.

For cascadable parametric equalizers, the dc reference gain must be set equal to unity, $G_0 = 1$. For resonator filters it must be set to zero, $G_0 = 0$, and the peak gain set to unity, $G = 1$. The Nyquist-frequency gain is still calculated by Eq. (21) (with $G_0 = 0$) and represents the gain an analog resonator has at $f_s/2$. Thus such a digital resonator filter can better emulate physical resonances. Digital notch filters emulating analog ones can also be designed by setting $G_0 = 1$ and $G = 0$.

3 REALIZATIONS

The digital filter of Eq. (20) was given in terms of the direct-form numerator and denominator coefficients, and can be written in the compact form

$$H(z) = \frac{B(z)}{A(z)} = \frac{b_0 + b_1z^{-1} + b_2z^{-2}}{1 + a_1z^{-1} + a_2z^{-2}} . \tag{22}$$

It can be realized in any of the standard direct-form realizations, such as direct forms I or II or transposed forms. The direct form II should perhaps be avoided since it requires special care to prevent internal overflows.

The filter can also be realized in its lattice or ladder form [12], [13], which has good numerical properties. For the conventional design, the Regalia–Mitra realization [5], [7], [8], [10] uses an all-pass filter realized in its lattice form and allows the independent control of the three parameters of center frequency ω_0, bandwidth $\Delta\omega$, and peak gain G.

In this section we discuss the lattice realization of Eq. (20) and find that it leads to a generalization of the Regalia–Mitra form. The lattice realization is built out of the lattice recursions of the denominator polynomial $A(z)$, that is, iterating up to order 2,

$$\begin{aligned} A_0(z) &= 1 \\ A_1(z) &= A_0(z) + k_1z^{-1}A_0^R(z) = 1 + k_1z^{-1} \\ A_2(z) &= A_1(z) + k_2z^{-1}A_1^R(z) = 1 + k_1(1 + k_2)z^{-1} \\ &\quad + k_2z^{-2} \end{aligned} \tag{23}$$

where the reversed polynomials are

$$\begin{aligned} A_0^R(z) &= 1 \\ A_1^R(z) &= k_1 + z^{-1} \\ A_2^R(z) &= k_2 + k_1(1 + k_2)z^{-1} + z^{-2} . \end{aligned} \tag{24}$$

Identifying $A_2(z)$ with the direct-form denominator $A(z)$

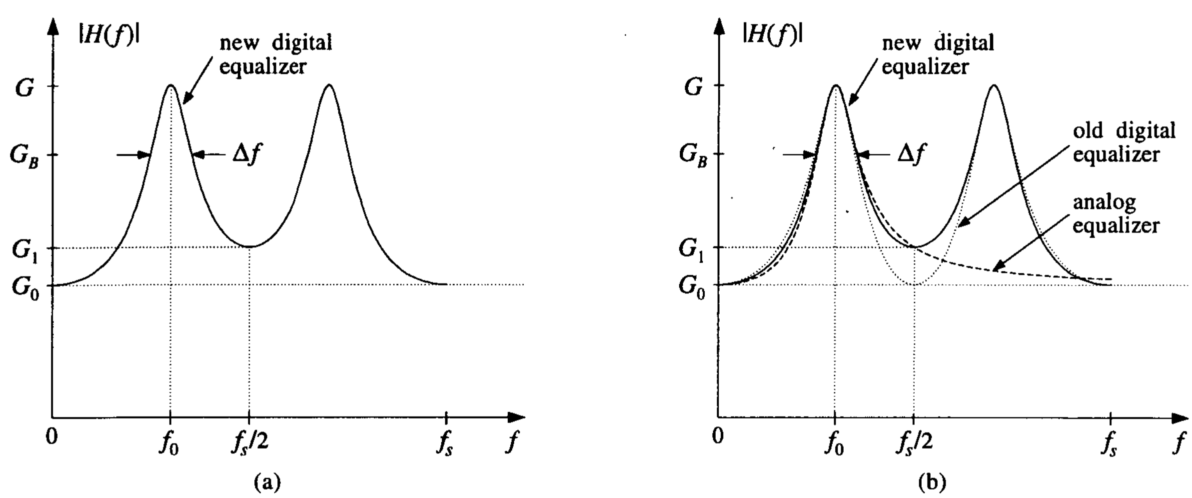


Fig. 2. New digital equalizer matches Nyquist-frequency gain of corresponding analog equalizer.

of Eq. (22) gives for the coefficients $k_2 = a_2$ and $k_1(1 + k_2) = a_1$, which may be solved for the reflection coefficients,

$$k_1 = \frac{a_1}{1 + a_2}, \qquad k_2 = a_2 . \tag{25}$$

Using the coefficients of Eq. (20), we find

$$k_1 = -\frac{1 - W^2}{1 + W^2}, \qquad k_2 = \frac{1 + W^2 - A}{1 + W^2 + A} . \tag{26}$$

We can now express the numerator polynomial $B(z)$ of Eq. (20) in terms of the reflection coefficients k_1 and k_2 in the following form:

$$\begin{aligned} B(z) = {} & \frac{1}{2}\overline{G}_0(1 + k_2)(1 + 2k_1z^{-1} + z^{-2}) \\ & + \frac{1}{2}\overline{G}(1 - k_2)(1 - z^{-2}) + \Delta G(k_1 + 2z^{-1} + k_1z^{-2}) \end{aligned} \tag{27}$$

where we defined the quantities

$$\begin{aligned} \overline{G}_0 &= \frac{1}{2}(G_0 + G_1) \\ \overline{G} &= \frac{B}{A} = \sqrt{\frac{G^2C + G_B^2D}{C + D}} \\ \Delta G &= -\frac{1}{4}(G_1 - G_0)(1 + k_2) . \end{aligned} \tag{28}$$

The three polynomial terms of $B(z)$ can be written in terms of the lattice polynomials $A_1(z)$, $A_2(z)$, and their reverse, as follows:

$$\begin{aligned} A_2(z) + A_2^R(z) &= (1 + k_2)(1 + 2k_1z^{-1} + z^{-2}) \\ A_2(z) - A_2^R(z) &= (1 - k_2)(1 - z^{-2}) \\ A_1^R(z) + z^{-1}A_1(z) &= k_1 + 2z^{-1} + k_1z^{-2} . \end{aligned} \tag{29}$$

Thus Eq. (27) can be written in the form

$$\begin{aligned} B(z) &= \frac{1}{2}\overline{G}_0[A_2(z) + A_2^R(z)] + \frac{1}{2}\overline{G}[A_2(z) - A_2^R(z)] + \Delta G[A_1^R(z) + z^{-1}A_1(z)] \\ &= \frac{1}{2}(\overline{G}_0 + \overline{G})A_2(z) + \frac{1}{2}(\overline{G}_0 - \overline{G})A_2^R(z) + \Delta G[A_1^R(z) + z^{-1}A_1(z)] . \end{aligned}$$

The transfer function of Eq. (20) will then be

$$H(z) = \frac{B(z)}{A(z)} = \frac{1}{2}(\overline{G}_0 + \overline{G}) + \frac{1}{2}(\overline{G}_0 - \overline{G})\frac{A_2^R(z)}{A_2(z)} + \Delta G\frac{A_1^R(z) + z^{-1}A_1(z)}{A_2(z)} . \tag{30}$$

A block diagram realization of Eq. (30) is shown in Fig. 3, where the all-pass transfer function $A_2^R(z)/A_2(z)$ has been realized in its lattice form. As a consequence of the lattice structure, it can be verified easily that the terms $A_2^R(z)/A_2(z)$, $A_1^R(z)/A_2(z)$, and $A_1(z)/A_2(z)$ are the transfer functions from the input x to the signals y_2, y_1, and x_1, respectively. It might appear strange that we introduced a third delay into the realization of this second-order filter. However, this was done for convenience in order to make use of the successive outputs of the lattice sections.

In the limiting case when $G_1 = G_0$, we have $\Delta G = 0$, and Eq. (30) and Fig. 3 reduce to the Regalia–Mitra realization for the conventional design of Eq. (11). Indeed, we have in this limit

$$\overline{G}_0 = G_0 , \qquad \overline{G} = G , \qquad k_1 = -\cos\omega_0 ,$$
$$k_2 = \frac{1 - \beta}{1 + \beta} \tag{31}$$

where β is given by Eq. (12). In the general case the realization coefficients k_1, k_2, $\overline{G}_0$, $\overline{G}$, and ΔG do not quite provide independent control of the equalizer's parameters. Thus the value of this lattice realization lies mostly in its numerical properties.

Strictly speaking, the conventional Regalia–Mitra realization with parameters given by Eq. (31) is not completely decoupled either, because β depends on both G and $\Delta\omega$, unless one defines G_B^2 as the weighted arithmetic mean of Eq. (36), as discussed in the next section.

An alternative realization—which is the standard lattice and ladder realization [12], [13]—can be obtained by expressing $B(z)$ as a linear combination of the three reverse filters $A_0^R(z)$, $A_1^R(z)$, and $A_2^R(z)$ in the form

$$B(z) = c_0A_0^R(z) + c_1A_1^R(z) + c_2A_2^R(z) \tag{32}$$

where the expansion coefficients can be obtained from the direct-form coefficients $\{b_0, b_1, b_2\}$ of $B(z)$ via the backward substitution

$$\begin{aligned} c_2 &= b_2 \\ c_1 &= b_1 - a_1c_2 \\ c_0 &= b_0 - k_1c_1 - k_2c_2 . \end{aligned} \tag{33}$$

Then the equalizer transfer function becomes

$$H(z) = \frac{B(z)}{A(z)} = c_0 \frac{A_0^R(z)}{A_2(z)} + c_1 \frac{A_1^R(z)}{A_2(z)} + c_2 \frac{A_2^R(z)}{A_2(z)} . \tag{34}$$

The transfer functions of the three terms are obtained at the lattice section output signals y_0, y_1, and y_2 of Fig. 3, which can then be combined linearly with the c coefficients. This structure also has good numerical properties, and it is straightforward to modify the MATLAB function **peq.m** to compute the coefficients $\{k_1, k_2, c_0, c_1, c_2\}$.

We note finally that the transfer function (20) is a minimum-phase transfer function, so that both $H(z)$ and its inverse $1/H(z)$ are stable and causal. This follows [12], [13] from the fact that the denominator reflection coefficients have magnitudes less than 1, $|k_1| \leq 1$, $|k_2| \leq 1$, and so do the numerator reflection coefficients, which are

$$k_{b_1} = \frac{b_1/b_0}{1 + b_2/b_0} = -\frac{G_1 - G_0W^2}{G_1 + G_0W^2},$$

$$k_{b_2} = \frac{b_2}{b_0} = -\frac{G_1 + G_0W^2 - B}{G_1 + G_0W^2 + B}.$$

4 BANDWIDTH

As discussed by Bristow-Johnson [9], there is considerable variation in the literature in the definition of bandwidth $\Delta\omega$ and bandwidth gain G_B. For example, one can define $\Delta\omega$ to be the difference of the band-edge frequencies in linear frequency units, or define it in octaves in log units.

As seen in Fig. 1, for a conventional digital equalizer, the gain G_B must always be defined to lie somewhere between the reference and the peak gains, that is,

$$\begin{aligned} &G_0 < G_B < G \qquad \text{(boost)} \\ &G_0 > G_B > G \qquad \text{(cut)} . \end{aligned} \tag{35}$$

For a boost one may define G_B to be 3 dB below the peak gain, $G_B^2 = G^2/2$, or take it to be 3 dB above the reference, $G_B^2 = 2G_0^2$ or define it as the arithmetic mean of the peak and reference gains, $G_B^2 = (G_0^2 + G^2)/2$, or as the geometric mean, $G_B^2 = G_0G$, which is the arithmetic mean of the gains in decibel scales. The 3-dB definitions are possible only if the boost gain G is itself greater than 3 dB, that is, $G^2 > 2G_0^2$.

For a cut one may take G_B to be 3 dB above the cut gain, $G_B^2 = 2G^2$, or 3 dB below the reference, $G_B^2 = G_0^2/2$, or use the arithmetic and geometric means. Again, the first two definitions are possible only if the cut gain is at least 3 dB below the reference, $G^2 < G_0^2/2$.

In the special cases of a resonator ($G_0 = 0$, $G = 1$) and a notch filter ($G_0 = 1$, $G = 0$) the 3-dB definitions of G_B are always possible, and in fact, they coincide with the arithmetic mean.

The arithmetic and geometric mean definitions are always possible for any value of the boost or cut gain G. They can be replaced by the more general weighted arithmetic or geometric means,

$$\begin{aligned} G_B^2 &= \alpha G_0^2 + (1 - \alpha)G^2 \\ G_B &= G_0^\alpha G^{1-\alpha} \end{aligned} \tag{36}$$

where $0 < \alpha < 1$. The conventional means have $\alpha = 1/2$. The weighted geometric mean is equivalent to a weighted arithmetic mean of the decibel gains.

The weighted geometric mean is attractive because a boost and a cut by equal and opposite gains in decibels cancel exactly [9]. (Their transfer functions are inverses of each other.) The arithmetic mean is attractive because it makes the conventional Regalia–Mitra realization truly independently controllable by the three equalizer parameters $\{\omega_0, \Delta\omega, G\}$.

Indeed, if G_B^2 is given by the weighted arithmetic mean of Eq. (36), then the square-root factor in the definition of β in Eq. (12) becomes independent of G,

$$\sqrt{\frac{G_B^2 - G_0^2}{G^2 - G_B^2}} = \sqrt{\frac{1 - \alpha}{\alpha}}$$

and it is equal to unity when $\alpha = 1/2$.

For an analog equalizer, as well as for the new digital design, it is evident from Fig. 2 that the peak gain G must be greater than the Nyquist-frequency gain G_1. Thus the minimum requirement for the choice of G_B is that it lie in the intervals

$$\begin{aligned} &G_0 < G_1 < G_B < G \qquad \text{(boost)} \\ &G_0 > G_1 > G_B > G \qquad \text{(cut)} . \end{aligned} \tag{37}$$

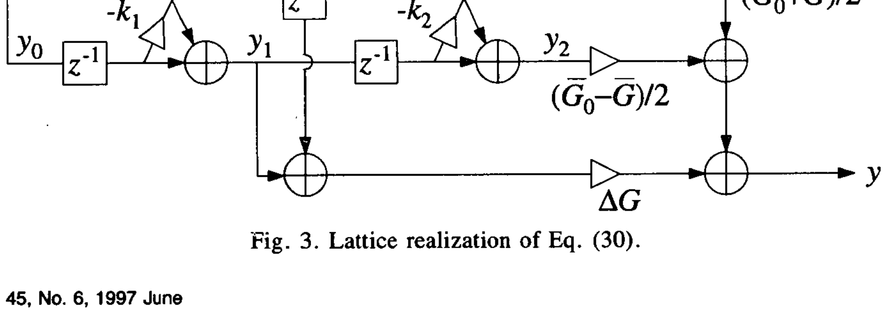


Fig. 3. Lattice realization of Eq. (30).

It follows from these inequalities that the arguments of all the square roots in Eqs. (16)–(21) are always positive, for either a boost or a cut.

The result that a boost and a cut of equal and opposite decibel gains cancel each other can be generalized to the new design as follows. Given a set of design gains $\{G_0, G_1, G_B, G\}$ for a boost, we can get a design set for a cut by the transformation

$$\{G_0, G_1, G_B, G\} \rightarrow \{G_0^{-1}, G_1^{-1}, G_B^{-1}, G^{-1}\}\,. \tag{38}$$

Indeed, if the gains $\{G_0, G_1, G_B, G\}$ satisfy the boost inequalities in Eq. (37), then the inverted gains $\{G_0^{-1}, G_1^{-1}, G_B^{-1}, G^{-1}\}$ will satisfy the cut inequalities.

For a unity reference gain ($G_0 = 1$), the transformation $G \rightarrow G^{-1}$ implies that the boost and cut will have equal and opposite peak gains in decibels. The transformation $G_B \rightarrow G_B^{-1}$ means that the bandwidth gain for the cut must be measured at a decibel level that is the negative of the bandwidth level of the boost.

Under the boost-to-cut transformation [Eq. (38)] and for fixed values of center frequency ω_0 and bandwidth $\Delta\omega$, it follows that the cut transfer function will be exactly the inverse of the boost,

$$H_{\text{cut}}(z) = \frac{1}{H_{\text{boost}}(z)}\,. \tag{39}$$

To see this, we note that under the transformation $\{G_0, G_B, G\} \rightarrow \{G_0^{-1}, G_B^{-1}, G^{-1}\}$ the Nyquist-frequency gain of Eq. (21) transforms according to $G_1 \rightarrow G_1^{-1}$. Moreover, the prewarped bandwidth $\Delta\Omega$ of Eq. (19) remains invariant. It follows from Eqs. (16) and (17) that the analog filter coefficients will transform as

$$W^2 \rightarrow G_0G_1^{-1}W^2\,, \qquad A \rightarrow G_1^{-1}B\,, \qquad B \rightarrow G_1^{-1}A\,.$$

These and Eq. (38) imply that the equivalent analog transfer function of Eq. (13) will map into its inverse,

$$\frac{G_1s^2 + Bs + G_0W^2}{s^2 + As + W^2} \rightarrow \frac{G_1^{-1}s^2 + (G_1^{-1}A)s + G_0^{-1}(G_0G_1^{-1}W^2)}{s^2 + (G_1^{-1}B)s + (G_0G_1^{-1}W^2)} = \frac{s^2 + As + W^2}{G_1s^2 + Bs + G_0W^2}$$

or, $H_{\text{cut}}(s) = 1/H_{\text{boost}}(s)$. Then the bilinear transformation implies Eq. (39).

Although the boost and cut levels G_B and G_B^{-1} are equal and opposite in decibels, they do not have to be measured at the arithmetic-mean decibel level, $G_B = \sqrt{G_0G} = \sqrt{G}$. (This may not even be possible if G is so small that $\sqrt{G} < G_1 < G$.) A better choice might be $G_B = \sqrt{G_1G}$.

Finally, we discuss the modifications to the design when the bandwidth is to be specified in octaves. If the band-edge frequencies are related by $\omega_2 = 2^{\Delta\gamma}\omega_1$, so that the bandwidth is $\Delta\gamma$ octaves, then for the design of a physical analog equalizer, the band-edge frequencies will lie symmetrically with respect to the center frequency ω_0 in a log-frequency scale, and their difference $\Delta\omega = \omega_2 - \omega_1$ can be expressed in terms of the octave width $\Delta\gamma$ as follows [9] (in units of radians per sample):

$$\Delta\omega = 2\omega_0 \sinh\left(\frac{\ln 2}{2}\Delta\gamma\right). \tag{40}$$

For the conventional digital equalizer design, Bristow-Johnson [9] suggests the following approximation for the prewarped difference $\Delta\Omega$, which effectively amounts to linearizing the bilinear transformation mapping $\ln \Omega = \ln \tan(\omega/2)$ about the center frequency $\Omega_0 = \tan(\omega_0/2)$:

$$\Delta\Omega = 2\Omega_0 \sinh\left(\frac{\omega_0}{\sin\omega_0}\,\frac{\ln 2}{2}\Delta\gamma\right). \tag{41}$$

Using Eqs. (10) and (12), the design parameter β can be expressed as

$$\beta = \sqrt{\frac{G_B^2 - G_0^2}{G^2 - G_B^2}}\,\sinh\left(\frac{\omega_0}{\sin\omega_0}\,\frac{\ln 2}{2}\Delta\gamma\right)\sin\omega_0\,. \tag{42}$$

The approximation (41) works well also for the new digital equalizer. The only difference is that now the quantity $\Delta\Omega$ of Eq. (17) must be calculated from Eq. (41) instead of Eq. (19), and Eq. (40) must be used in Eq. (21) to calculate G_1.

An alternative approach, which leads to an exact solution, is to specify the center frequency ω_0 and one (but not both) of the band-edge frequencies, say ω_2, and give it in linear or octave scales. For example, $\omega_2 = 2^{\gamma_2}\omega_0$ lies γ_2 octaves above ω_0. Then solve for the other band-edge frequency using the prewarped geometric-mean rule [Eq. (18) or Eq. (43)]. Then calculate $\Delta\omega = \omega_2 - \omega_1$, and proceed with Eqs. (16)–(21) to complete the design. Eq. (18) can be written in terms of the physical frequencies as follows:

$$\tan\left(\frac{\omega_1}{2}\right)\tan\left(\frac{\omega_2}{2}\right) = \sqrt{\frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}}\,\sqrt{\frac{G^2 - G_1^2}{G^2 - G_0^2}}\,\tan^2\left(\frac{\omega_0}{2}\right). \tag{43}$$

5 DESIGN EXAMPLES

Next we present some design examples. Fig. 4 shows an equalizer with a 12-dB boost (and a 12-dB cut), a peak frequency of $\omega_0 = 0.5\pi$ rad per sample, a bandwidth of $\Delta\omega = 0.2\pi$ rad per sample, and a reference dc gain of 0 dB. The bandwidth is measured at 3 dB below the peak, that is, at a level of 9 dB. Thus the design gains will be

$$G_0 = 10^{0/20} = 1\,, \qquad G = 10^{12/20} = 3.9811\,,$$

$$G_B = 10^{9/10} = 2.8184\,.$$

Fig. 4 shows the magnitude responses in decibels, that is, $20 \log_{10} |H(\omega)|$, of the new design of Eq. (20), the old design of Eq. (11), and the analog design of Eq. (1). The frequency axis extends over one complete Nyquist interval, $0 \leqslant \omega \leqslant 2\pi$. For the cut case, all gains are the inverses of these values (or the negatives in decibels), and the corresponding transfer functions become the inverses of the boost case, according to Eq. (39).

To clarify the design method, we discuss the steps leading to the transfer function of the boost case. Inserting the given values of $\{\omega_0, \Delta\omega, G_0, G, G_B\}$ in Eq. (21), we find $G_1 = 1.3685 = 2.7251$ dB. Then Eq. (19) gives $\Omega_0 = \tan(\omega_0/2) = 1$ and $\Delta\Omega = 0.6620$. Using Eqs. (16) and (17), we calculate the parameters $W^2 = 0.9702, C = 2.5005, D = 0.1446, A = 0.5784$, and $B = 2.2712$, which are used in Eq. (20) to give

$$H(z) = \frac{1.8088 - 0.3126z^{-1} + 0.0265z^{-2}}{1 - 0.0234z^{-1} + 0.5461z^{-2}}\,.$$

Alternatively, the same filter coefficients may be obtained by the following call to the MATLAB function peq.m of Appendix 2:

[b, a, G1] = peq(G0, G, GB, w0, Dw) .

By comparison, the conventional equalizer transfer function for this example is obtained by using Eq. (12) to calculate $\beta = 0.3045$ and then using Eq. (11) to give

$$H(z) = \frac{1.6959 - 0.1627z^{-2}}{1 + 0.5332z^{-2}}\,.$$

We note also that in these examples the analog filter's magnitude response [Eq. (2)], was calculated in terms of the normalized digital frequencies in the same fashion as Eq. (21), that is, after using the relationships of Eq. (5),

$$|H(\omega)|^2 = \frac{G_0^2(\omega^2 - \omega_0^2)^2 + G^2\omega^2(\Delta\omega)^2(G_B^2 - G_0^2)/(G^2 - G_B^2)}{(\omega^2 - \omega_0^2)^2 + \omega^2(\Delta\omega)^2(G_B^2 - G_0^2)/(G^2 - G_B^2)}\,.$$

Fig. 5 shows two more examples with the same specifications as the preceding except that the peak frequency is now $\omega_0 = 0.3\pi$ for Fig. 5(a) and $\omega_0 = 0.7\pi$ for Fig. 5(b). The corresponding Nyquist-frequency gains are 2.053 and 4.420 dB, respectively.

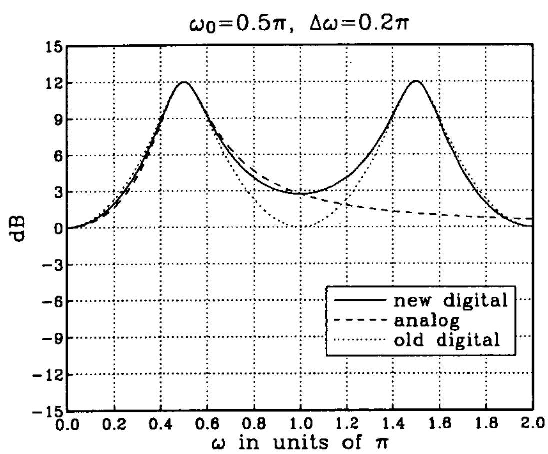
Fig. 6 illustrates the dependence of the Nyquist-frequency gain G_1 given by Eq. (21) on the design parameters $\{\omega_0, \Delta\omega, G\}$. Fig. 6(a) shows G_1 in decibels as a function of the bandwidth $\Delta\omega$, varying over the range $0.01\pi \leqslant \Delta\omega \leqslant 0.5\pi$. The peak frequency was fixed at $\omega_0 = 0.5\pi$, and the following three values of the peak gain were chosen: $G = 12, 9, 6$ dB. The bandwidth was always measured at 3 dB below each peak.

Fig. 6(b) shows the dependence of G_1 on the peak frequency ω_0, being varied over the interval $0.01\pi \leqslant \omega_0 \leqslant 0.95\pi$. The same three peak gains were used. The bandwidth was fixed at $\Delta\omega = 0.1\pi$ and measured at 3 dB below each peak.

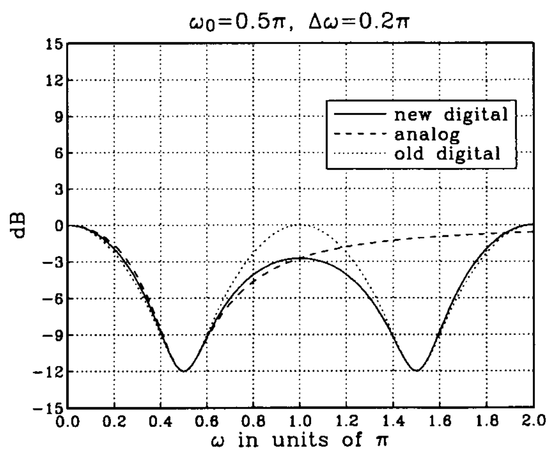
Figs. 7–9 illustrate the nature of the approximations of Eq. (41). The frequency axis is $\log_2(\omega/\pi)$, and is measured in octaves below the Nyquist frequency. In all cases the peak gains are 12 dB and the bandwidths are measured at 9 dB.

The designs of Fig. 7 have center frequency $\omega_0 = 2^{-1}\pi$ and octave widths $\Delta\gamma = 1$ and $\Delta\gamma = 0.5$. Fig. 8 has $\omega_0 = 2^{-0.5}\pi$ and octave widths $\Delta\gamma = 0.5$ and $\Delta\gamma = 0.25$. Fig. 9(a) has $\omega_0 = 2^{-0.25}\pi$, $\Delta\gamma = 0.25$, and Fig. 9(b) has $\omega_0 = 2^{-0.125}\pi$, $\Delta\gamma = 0.125$.

In all cases the analog design has symmetric bandwidth about the center frequency, and the new digital design attempts to follow the analog one as much as possible.



(a)



(b)

Fig. 4. 12-dB boost and cut at $\omega_0 = 0.5\pi$; bandwidth $\Delta\omega = 0.2\pi$ is measured at ± 9 dB.

6 DISCUSSION

The design method of this paper results in the most general type of second-order digital parametric equalizer, because the five filter coefficients are fixed uniquely by five different design constraints.

The method encompasses the conventional design as a special case. For low center frequencies and widths, the new method will be almost identical to the conventional method, because the Nyquist-frequency gain G_1 is almost equal to G_0. The differences of the two methods are felt only for high frequencies and widths. Fig. 6 gives an idea of how high is "high."

The method allows various ways of defining the band-

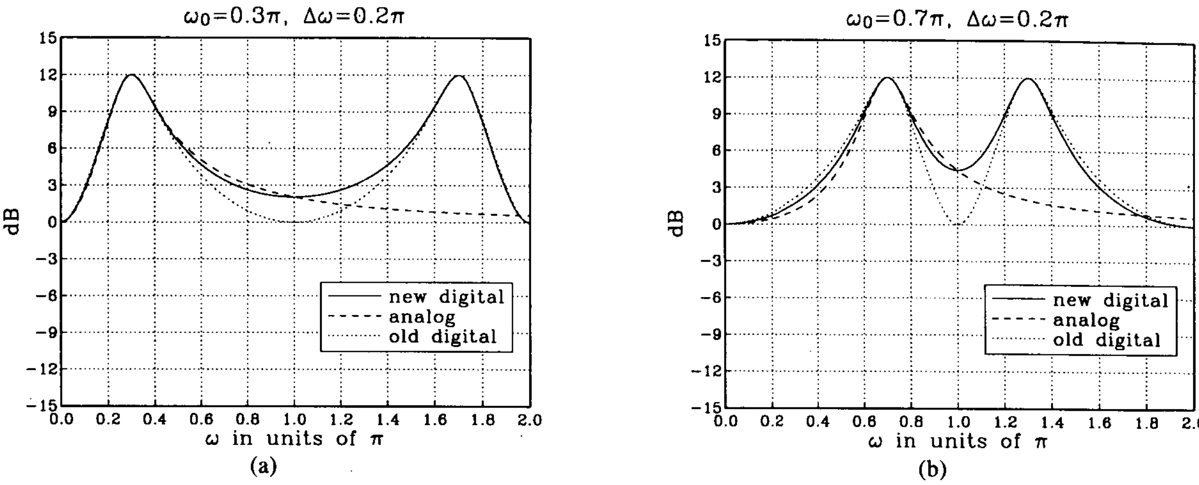


Fig. 5. 12-dB boost at $\omega_0 = 0.3\pi$ and 0.7π; bandwidth $\Delta\omega = 0.2\pi$ is measured at 9 dB.

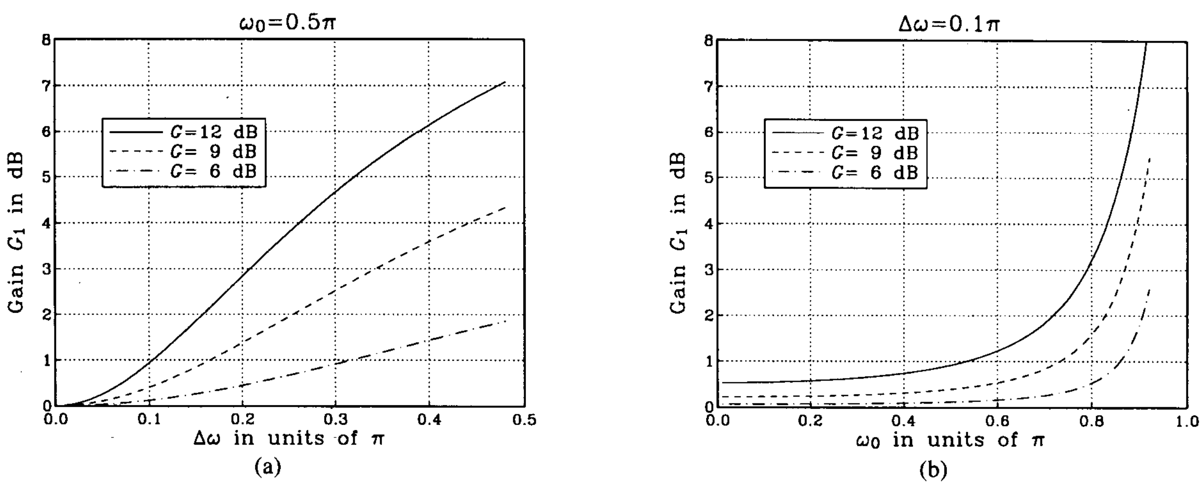


Fig. 6. Nyquist-frequency gain G_1 as a function of ω_0 and $\Delta\omega$.

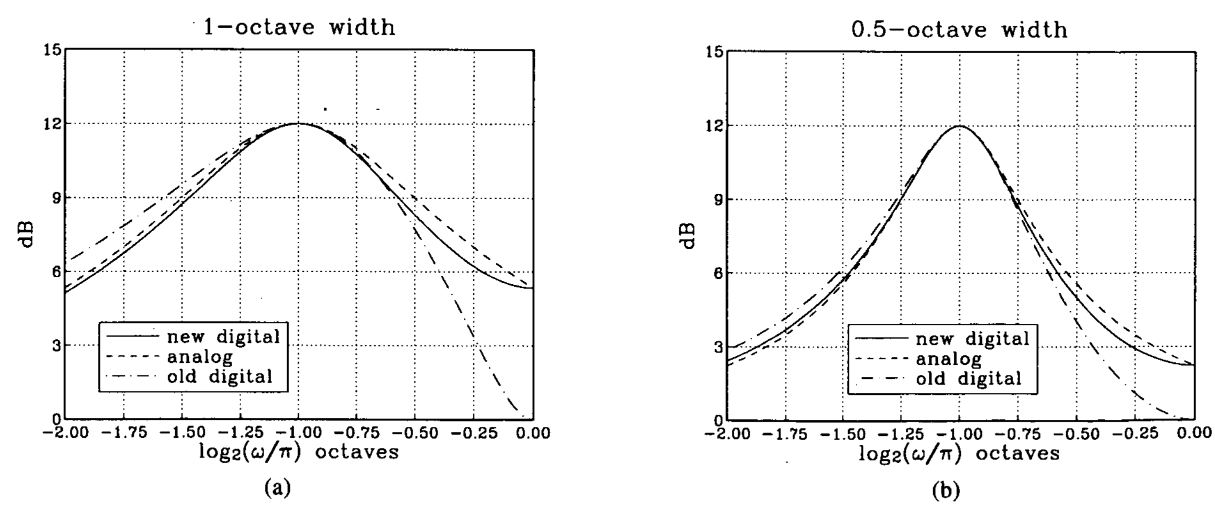


Fig. 7. Center frequency $\omega_0 = 2^{-1}\pi$; octave widths $\Delta\gamma = 1$ and 0.5.

width in linear and log frequency scales and of defining the bandwidth gain G_B. Given the wide variety of possibilities in choosing G_B, it is perhaps best to leave G_B as a free parameter to be chosen by the user, as long as it satisfies Eq. (37).

7 ACKNOWLEDGMENT

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8 REFERENCES

[1] K. Hirano, S. Nishimura, and S. Mitra, "Design of Digital Notch Filters," *IEEE Trans. Commun.*, vol. COM-22, p. 964 (1974).

[2] M. N. S. Swami and K. S. Thyagarajan, "Digital Bandpass and Bandstop Filters with Variable Center Frequency and Bandwidth," *Proc. IEEE*, vol. 64, p. 1632 (1976).

[3] J. A. Moorer, "The Manifold Joys of Conformal Mapping: Applications to Digital Filtering in the Studio," *J. Audio Eng. Soc.*, vol. 31, pp. 826–841 (1983 Nov.).

[4] S. A. White, "Design of a Digital Biquadratic Peaking or Notch Filter for Digital Audio Equalization," *J. Audio Eng. Soc. (Engineering Reports)*, vol. 34, pp. 479–483 (1986 June).

[5] P. A. Regalia and S. K. Mitra, "Tunable Digital Frequency Response Equalization Filters," *IEEE Trans. Acoust., Speech, Signal Process.*, vol. ASSP-35, p. 118 (1987).

[6] D. J. Shpak, "Analytical Design of Biquadratic Filter Sections for Parametric Filters," *J. Audio Eng. Soc.*, vol. 40, pp. 876–885 (1992 Nov.).

[7] D. C. Massie, "An Engineering Study of the Four-Multiply Normalized Ladder Filter," *J. Audio Eng. Soc.*, vol. 41, pp. 564–582 (1993 July/Aug.).

[8] F. Harris and E. Brooking, "A Versatile Parametric Filter Using Imbedded All-Pass Subfilter to Independently Adjust Bandwidth, Center Frequency, and Boost or Cut," presented at the 95th Convention of the Audio Engineering Society, *J. Audio Eng. Soc. (Abstracts)*, vol. 41, p. 1066 (1993 Dec.), preprint 3757.

[9] R. Bristow-Johnson, "The Equivalence of Various Methods of Computing Biquad Coefficients for Audio Parametric Equalizers," presented at the 97th Convention of the Audio Engineering Society, *J. Audio*

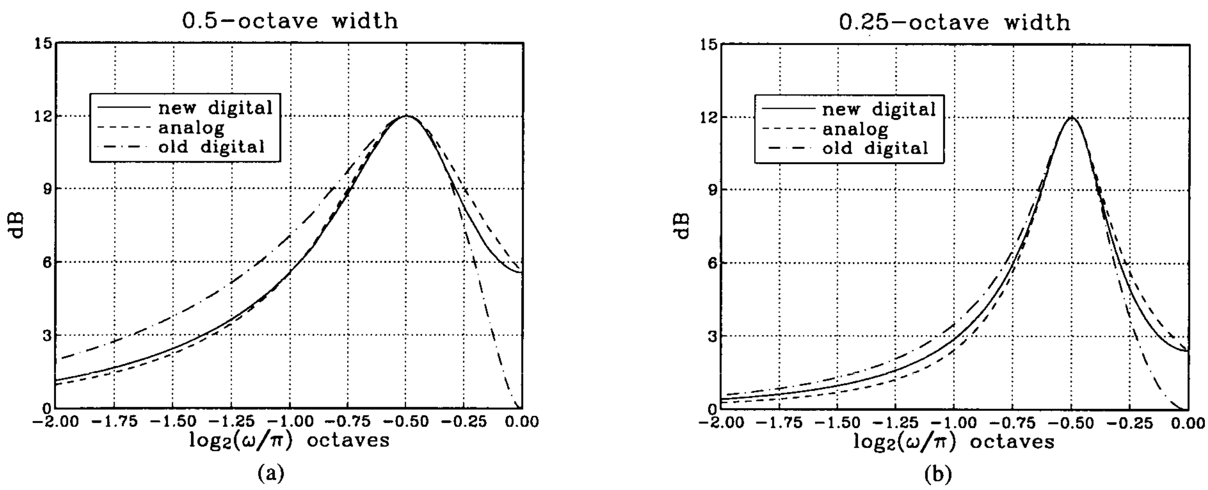

Fig. 8. Center frequency $\omega_0 = 2^{-0.5}\pi$; octave widths $\Delta\gamma = 0.5$ and 0.25.

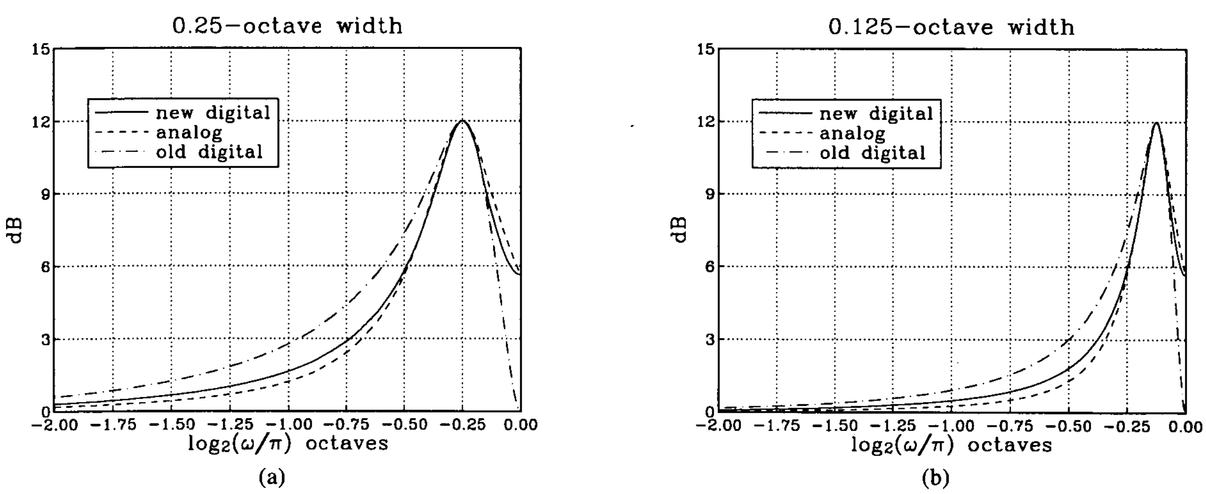

Fig. 9. (a) $\omega_0 = 2^{-0.25}\pi$, $\Delta\gamma = 0.25$. (b) $\omega_0 = 2^{-0.125}\pi$, $\Delta\gamma = 0.125$.

Eng. Soc. (Abstracts), vol. 42, pp. 1062–1063 (1994 Dec.), preprint 3906.

[10] U. Zölzer and T. Boltze, "Parametric Digital Filter Structures," presented at the 99th Convention of the Audio Engineering Society, *J. Audio Eng. Soc. (Abstracts)*, vol. 43, p. 1090 (1995 Dec.), preprint 4099.

[11] S. J. Orfanidis, *Introduction to Signal Processing* (Prentice-Hall, Upper Saddle River, NJ, 1996).

[12] A. H. Gray and J. D. Markel, "Digital Lattice and Ladder Filter Synthesis," *IEEE Trans. Audio Electroacoust.*, vol. AU-21, p. 491 (1973).

[13] A. V. Oppenheim and R. W. Schafer, *Discrete-Time Signal Processing* (Prentice-Hall, Upper Saddle River, NJ, 1989).

APPENDIX 1

Here we present some of the derivations of the design equations (16)–(21). The derivative of Eq. (14) can be written in the form

$$\frac{\partial}{\partial\Omega^2}|H(\Omega)|^2 = \frac{2G_1(G_1\Omega^2 - G_0W^2) + B^2 - |H(\Omega)|^2\,[2(\Omega^2 - W^2) + A^2]}{(\Omega^2 - W^2)^2 + A^2\Omega^2}$$

The first two conditions of Eq. (15) applied at $\Omega = \Omega_0$ give

$$2G_1(G_1\Omega_0^2 - G_0W^2) + B^2 - G^2[2(\Omega_0^2 - W^2) + A^2] = 0$$

$$\frac{(G_1\Omega_0^2 - G_0W^2)^2 + B^2\Omega_0^2}{(\Omega_0^2 - W^2)^2 + A^2\Omega_0^2} = G^2 .$$

Solving these equations for the quantities W^2 and $B^2 - G^2A^2$ gives Eq. (16) for W^2, and

$$B^2 - G^2A^2 = -2W^2[(G^2 - G_0G_1) - \sqrt{(G^2 - G_0^2)(G^2 - G_1^2)}] \equiv -D . \quad (44)$$

Next consider the band-edge condition

$$\frac{(G_1\Omega^2 - G_0W^2)^2 + B^2\Omega^2}{(\Omega^2 - W^2)^2 + A^2\Omega^2} = G_B^2 .$$

It can be written as the quartic equation

$$\Omega^4 - \left(\frac{G_B^2 - G_0G_1}{G_B^2 - G_1^2}2W^2 + \frac{B^2 - G_B^2A^2}{G_B^2 - G_1^2}\right)\Omega^2 + \frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}W^4 = 0 .$$

It follows that the two band-edge frequencies will satisfy

$$\Omega_1^2 + \Omega_2^2 = \frac{G_B^2 - G_0G_1}{G_B^2 - G_1^2}2W^2 + \frac{B^2 - G_B^2A^2}{G_B^2 - G_1^2}$$

$$\Omega_1\Omega_2 = \sqrt{\frac{G_B^2 - G_0^2}{G_B^2 - G_1^2}}W^2 \quad (45)$$

Using Eq. (45), we find for the difference $\Delta\Omega = \Omega_2 - \Omega_1$,

$$\Delta\Omega^2 = \Omega_2^2 + \Omega_1^2 - 2\Omega_2\Omega_1 = \frac{B^2 - G_B^2A^2}{G_B^2 - G_1^2} + \frac{G_B^2 - G_0G_1 - \sqrt{(G_B^2 - G_0^2)(G_B^2 - G_1^2)}}{G_B^2 - G_1^2}2W^2$$

which can be rewritten as

$$B^2 - G_B^2A^2 = (\Delta\Omega)^2(G_B^2 - G_1^2) - 2W^2[(G_B^2 - G_0G_1) - \sqrt{(G_B^2 - G_0^2)(G_B^2 - G_1^2)}] \equiv C . \quad (46)$$

Solving Eqs. (44) and (46) for A and B gives Eq. (16). Finally, we derive the prewarped bandwidth in Eq. (19). Using a trigonometric identity and the bilinear transformation, we have for the physical bandwidth difference $\Delta\omega = \omega_2 - \omega_1$,

$$\tan\left(\frac{\Delta\omega}{2}\right) = \tan\left(\frac{\omega_2 - \omega_1}{2}\right) = \frac{\tan\left(\frac{\omega_2}{2}\right) - \tan\left(\frac{\omega_1}{2}\right)}{1 + \tan\left(\frac{\omega_2}{2}\right)\tan\left(\frac{\omega_1}{2}\right)} = \frac{\Omega_2 - \Omega_1}{1 + \Omega_2\Omega_1} = \frac{\Delta\Omega}{1 + \Omega_2\Omega_1}$$

which leads to Eq. (19).

APPENDIX 2

The following MATLAB function, peq.m implements the design equations (16)–(21):

```
% peq.m - Parametric EQ with matching gain at
Nyquist frequency
%
% Usage: [b, a, G1] = peq(G0, G, GB, w0, Dw)
%
% G0 = reference gain at DC
% G = boost/cut gain
% GB = bandwidth gain
%
% w0 = center frequency in rad/sample
% Dw = bandwidth in rad/sample
%
% b = [b0, b1, b2] = numerator coefficients
% a = [1, a1, a2] = denominator coefficients
% G1 = Nyquist-frequency gain

function [b, a, G1] = peq(G0, G, GB, w0, Dw)

F = abs(G^2 - GB^2);
G00 = abs(G^2 - G0^2);
F00 = abs(GB^2 - G0^2);

num = G0^2 * (w0^2 - pi^2)^2 + G^2 * F00 *
pi^2 * Dw^2 / F;
den = (w0^2 - pi^2)2 + F00 * pi^2 * Dw^2 / F;

G1 = sqrt(num/den);

G01 = abs(G^2 - G0*G1);
G11 = abs(G^2 - G1^2);
F01 = abs(GB^2 - G0*G1);
F11 = abs(GB^2 - G1^2);

W2 = sqrt(G11 / G00) * tan(w0/2)^2;
DW = (1 + sqrt(F00 / F11) * W2) * tan(Dw/2);

C = F11 * DW^2 - 2 * W2 * (F01 - sqrt(F00 * F11));
D = 2 * W2 * (G01 - sqrt(G00 * G11));

A = sqrt((C + D) / F);
B = sqrt((G^2 * C + GB^2 * D) / F);

b = [(G1 + G0*W2 + B), -2*(G1 -G0*W2), (G1 -
B + G0*W2)] / (1 + W2 + A);
a = [1, [-2*(1 - W2), (1 + W2 - A)] / (1 + W2
+ A)];
```

Its inputs are the gains G_0, G, and G_B in absolute units, and the digital frequencies ω_0 and $\Delta\omega$ in units of radians per sample. Its outputs are the Nyquist-frequency gain G_1 given by Eq. (21), and the numerator and denominator coefficient vectors $\boldsymbol{b} = [b_0, b_1, b_2]$ and $\boldsymbol{a} = [1, a_1, a_2]$, defining the transfer function of Eq. (20) or Eq. (22).

THE AUTHOR

Sophocles J. Orfanidis is an associate professor at the Department of Electrical and Computer Engineering at Rutgers University. He received the B.S. degree from Miami University and the Ph.D. degree from Yale University, both in physics. Prior to joining Rutgers, he worked as a researcher at the Rockefeller University and New York University physics departments. His research interests are in adaptive signal and array processing, spectrum estimation, neural networks, and digital audio.

Dr. Orfanidis is the author of the books *Optimum Signal Processing* (McGraw-Hill) and *Introduction to Signal Processing* (Prentice Hall). He received the Rutgers College Parents Association "Teacher of the Year" award in 1990 and 1996.