Design of IIR Digital Differentiators Using Constrained Optimization

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Abstract—A new optimization method for the design of fullband and lowpass IIR digital differentiators is proposed. In the new method, the passband phase-response error is minimized under the constraint that the maximum passband amplitude-response relative error is below a prescribed level. For lowpass IIR differentiators, an additional constraint is introduced to limit the average squared amplitude response in the stopband, so as to minimize any high-frequency noise that may be present. Extensive experimental results are included which show that the differentiators designed using the proposed method have much smaller maximum phase-response error for the same passband amplitude-response error and stopband constraints when compared with several differentiators designed using state-of-the-art competing methods.

Index Terms—Digital differentiators, IIR filter design, design of filters by optimization

I. INTRODUCTION

Digital differentiators are used in various fields of signal processing such as in the design of compensators in control systems [1], extracting information about transients in biomedical signal processing [2]-[4], analyzing the signals in radar systems [5], and for edge detection in image processing [6]. Differentiators having perfect linear phase can be easily designed using FIR filters. However, in most applications perfectly linear phase is not required and differentiators having approximately linear phase are quite acceptable. In such applications, IIR differentiators are more attractive than FIR differentiators for two main reasons: Firstly, they can satisfy the given filter specifications with a much lower filter order thereby reducing the computational requirement or the complexity of hardware in a hardware implementation, and, secondly, they usually have a much smaller group delay thereby resulting in lower system delay.

The presence of the denominator polynomial in IIR filters renders their design more challenging than that of FIR filters because it results in highly nonlinear objective functions that require highly sophisticated optimization methods. As IIR filters lack the stability property of FIR filters, stability constraints must be incorporated in the design process to ensure that the filter is stable, which means constraining the poles to lie within the unit circle of the z plane.

Lowpass differentiators are appropriate when the signal of interest is at the low frequency end as they provide the advantage of reducing any high-frequency noise that may be present. In [7], [8], lowpass IIR differentiators have been designed by inverting the transfer function of lowpass integrators and then adjusting the denominator coefficients so that the poles lie within the unit circle. More recently in [9], two methods for designing lowpass IIR differentiators have been presented. In the first method, a fullband differentiator is cascaded with an appropriate lowpass filter, while in the second method the numerator is realized as a linear phase filter and the denominator is obtained using a constrained optimization method.

Earlier examples of fullband IIR differentiators have been provided in [10], although no method for their design is presented. In [11]-[15], fullband IIR differentiators are designed by taking the inverse of the transfer function of a fullband integrator and appropriately adjusting the denominator coefficients so that the poles lie within the unit circle. Then in [16], a sequential minimization procedure based on second-order factor updates was used, while in [17], an iterative quadratic programming approach with prescribed passband edge frequency is presented. The method in [17], however, uses a restrictive stability constraint that could affect the quality of the designs and, additionally, it requires that the group delay be specified. In [18] and [19], the differentiators are derived by taking an existing IIR differentiator and optimizing their pole-zero locations to improve the performance further.

In this paper, we propose a design method where the group-delay deviation with respect to the average group delay is minimized under the constraint that the maximum amplitude-response error be below a prescribed level. For lowpass IIR differentiators, we introduce an additional constraint to limit the average squared amplitude-response in the stopband, so as to minimize any high-frequency noise that may be present. By representing the filter in polar form, a non-restrictive stability constraint characterized by a set of linear inequality constraints can be incorporated in the optimization algorithm. The group delay is included as an optimization variable to achieve improved design specifications. Procedures for designing fullband and lowpass IIR differentiators are then described. Experimental results show that differentiators designed using the method have much smaller maximum phase-error for the same passband error and stopband constraint than several known state-of-the-art methods.

The paper is organized as follows. In Section II, we frame
the problem as an iterative constrained optimization problem. In Section III, we describe procedures for designing fullband and lowpass IIR differentiators. In Section IV, performance comparisons between filters designed using the proposed method and known methods are carried out. Conclusions are drawn in Section V.

II. THE OPTIMIZATION PROBLEM

In this section, we frame the problem at hand as an iterative constrained optimization problem by approximating each update as a linear approximation step as was done in [20] and [21]. To this end, we formulate the stability constraints, group-delay deviation, passband error, and stopband attenuation. Then, we incorporate the analytical results obtained within the framework of a constrained optimization problem.

A digital differentiator can be represented by the transfer function

\[ H(z) = H_0 \prod_{m=1}^{J} \frac{(z - r_a^{(1)} e^{j \theta_{am}})(z - r_a^{(2)} e^{-j \theta_{am}})}{(z - r_b^{(1)} e^{j \theta_{bm}})(z - r_b^{(2)} e^{-j \theta_{bm}})} \]  

(1)

where \( J \) is the number of differentiator sections, \( N = 2J \) is the differentiator order, and \( H_0 \) is a multiplier constant. An odd-order transfer function can be easily obtained by setting \( r_a^{(1)} \) and \( r_a^{(2)} \) to zero in the first section.

The ideal response of a causal differentiator is of the form [22]

\[ H_d(\omega) = j \omega e^{-j \tau \omega}, \quad 0 < |\omega| < \pi \]  

(2)

where \( \tau \) is the group delay. From (2), it is clear that at \( \omega = 0 \) the amplitude response is zero while the phase characteristic has a discontinuity of \( \pi \) and jumps between \( -\pi/2 \) and \( \pi/2 \) as frequency \( \omega \) switches between 0 and \( \omega_c \), respectively. Such a frequency response at \( \omega = 0 \) can be realized by placing a zero at \( z = 1 \) [10]. With this modification, the transfer function of the differentiator in (1) becomes

\[ H(c, z) = H_0(z - 1)(z - r_{a1}) \prod_{m=2}^{J} \frac{(z - r_a^{(1)} e^{j \theta_{am}})(z - r_a^{(2)} e^{-j \theta_{am}})}{(z - r_b^{(1)} e^{j \theta_{bm}})(z - r_b^{(2)} e^{-j \theta_{bm}})} \]  

(3)

where

\[ c = [r_{a1} \quad r_a^{(1)} \quad r_a^{(2)} \quad \theta_{b1} \quad r_{a2} \quad r_a^{(2)} \quad \theta_{a2} \quad r_b^{(1)} \quad r_b^{(2)} \quad \theta_{b2} \quad \cdots \quad r_a^{(1)} \quad r_a^{(2)} \quad \theta_{aj} \quad r_b^{(1)} \quad r_b^{(2)} \quad \theta_{bj} \quad H_0]^T \]  

(4)

To ensure that the differentiator is stable, the poles of the transfer function must lie within the unit circle [22]. If \( \epsilon_s \geq 0 \) is a stability margin of the pole radius from unity, and \( r_{b1}^{(1)}(k) \) and \( r_{b1}^{(2)}(k) \) are the corresponding values of \( r_{b1}^{(1)} \) and \( r_{b1}^{(2)} \) at the start of the \( k \)th iteration of the optimization, the stability conditions are given by

\[ |r_{b1}^{(1)}(k) + \delta_{b1}^{(1)}| \leq 1 - \epsilon_s \quad \forall \ m \in [1, J] \]
\[ |r_{b1}^{(2)}(k) + \delta_{b1}^{(2)}| \leq 1 - \epsilon_s \quad \forall \ m \in [1, J] \]  

(5)

where \( \delta_{b1}^{(1)} \) and \( \delta_{b1}^{(2)} \) are the corresponding updates for \( r_{b1}^{(1)}(k) \) and \( r_{b1}^{(2)}(k) \). Note that the conditions in (5) are convex inequality constraints and can, therefore, be incorporated within a convex optimization problem.

A. Group delay deviation

The group delay corresponding to the transfer function \( H(c, z) \) in (3) is given by

\[ \tau_g(c, \omega) = \alpha(1, 0, \omega) + \alpha(r_{a1}, 0, \omega) \]
\[ + \sum_{m=2}^{J} \left[ \alpha(r_{a1}^{(1)}, \theta_{am}, \omega) + \alpha(r_{a1}^{(2)}, -\theta_{am}, \omega) \right] \]
\[ - \sum_{m=1}^{J} \left[ \alpha(r_{b1}^{(1)}, \theta_{bm}, \omega) + \alpha(r_{b1}^{(2)}, -\theta_{bm}, \omega) \right] \]  

(6)

where

\[ \alpha(r, \theta, \omega) = \begin{cases} -1/2 & r = 1 \cos(\theta - \omega) - 1 \cos(\theta - \omega) + 1 \text{ otherwise} \end{cases} \]  

(7)

The group-delay deviation at frequency \( \omega \) is given by

\[ e_g(x, \omega) = \tau_g(c, e^{j \omega}) - \tau \]  

(8)

where

\[ x = [e^T, \tau]^T \]  

(9)

and \( \tau \) is the desired group delay which may be an optimization variable. To incorporate the \( L_p \) norm of the group-delay deviation, \( E_p^{(gd)}(k) \), in an iterative optimization problem we can approximate \( E_p^{(gd)}(k) \) for the \( k \)th iteration by a linear approximation given by [20]

\[ E_p^{(gd)}(k) \approx ||C_k \delta + d_k||_p \]  

(10)

where

\[ C_k = \left[ \begin{array}{cccc} k_g \nabla e_g(x_k, e^{j \omega})^T \\ \vdots \\ k_g \nabla e_g(x_k, e^{j \omega})^T \end{array} \right] \]  

(11)

\[ d_k = [d_1 \quad d_2 \quad \cdots \quad d_{N_p}]^T \]  

(12)

\[ d_i = k_g e_g(x_k, e^{j \omega}), \quad \omega_i \in \Psi_p \]  

(13)

\( x_k \) is the value of \( x \) in the \( k \)th iteration, \( \delta \) is the update to \( x_k \), \( k_g \) is a constant, and \( \Psi_p \) is the set of passband frequency sample points. The right-hand side of (10) is the right-hand side of (11) is the \( L_p \) norm of an affine function of \( \delta \) and, therefore, it is convex with respect to \( \delta \) [23].

B. Passband error

If \( H_d(\omega) \) is the desired frequency response of the differentiator in the passband and \( c_k \) is the value of vector \( c \) at the start of the \( k \)th iteration, a passband error function at frequency \( \omega \) can be defined as

\[ e_p(c_k, e^{j \omega}) = W(\omega)[|H(c_k, e^{j \omega})| - |H_d(\omega)|] \]
\[ = W(\omega)[|H(c_k, e^{j \omega})| - |H_d(\omega)|], \quad \omega \in \Psi_p \]  

(14)

Constant absolute or relative error may be required and \( W(\omega) \) can be chosen as unity or \( 1/|\omega| \) depending upon the
application. Note, however, that for constant absolute error, the relative error of the differentiator would tend to infinity as the frequency tends to zero; therefore, constant absolute error would not, typically, be of much practical interest and the design of differentiators with constant absolute error will not be considered further.

For the case of relative error, \( e_h(c_k, e^{j\omega}) \) can be expressed as

\[
e_h(c_k, e^{j\omega}) = |P(\omega) - 1|, \quad \omega \in \Psi_p
\]  

(15)

where

\[
P(\omega) = \frac{|H(c_k, e^{j\omega})|}{|\omega|} 
\]  

(16)

Function \( P(\omega) \) becomes indeterminate when \( \omega = 0 \). To circumvent this problem, we set \( z = e^{j\omega} \) and substitute (3) in (16) to obtain

\[
P(\omega) = \left| H_0(e^{j\omega} - r_{a1}) \right| - \omega \sin^2(\omega/2) + j 2 \, \omega \sin(\omega) \\
\sum_{m=2}^{J} \left( e^{j\omega} - r^{(1)}_{am} e^{j\theta_{am}} \right) (e^{j\omega} - r^{(2)}_{am} e^{-j\theta_{am}}) \\
\sum_{m=1}^{J} \left( e^{j\omega} - r^{(1)}_{bm} e^{j\theta_{bm}} \right) (e^{j\omega} - r^{(2)}_{bm} e^{-j\theta_{bm}})
\]

(17)

where

\[
sinc(x) = \begin{cases} 
  \frac{1}{x} & x \neq 0 \\
  1 & x = 0 
\end{cases}
\]

(18)

The modified form in (17) results in a deterministic value of \( P(\omega) \) at \( \omega = 0 \). Using the same approach as in Section II-A, the \( L_p \) norm of the passband relative error, \( e_h(c_k, e^{j\omega}) \), in (15) can be expressed in matrix form as

\[
E_p^{(pb)}(k) \approx \| D_k^{(pb)} \delta + f_k^{(pb)} \|_p 
\]

(19)

where

\[
D_k^{(pb)} = \begin{bmatrix} \kappa_{pb} \nabla e_h(c_k, e^{j\omega})^T & 0 \\ \vdots & \vdots \\ \kappa_{pb} \nabla e_h(c_k, e^{j\omega_{N_p}})^T & 0 \end{bmatrix}, \quad \omega_i \in \Psi_p
\]

(20)

\[
f_k^{(pb)} = \begin{bmatrix} f_1^{(pb)} \\ f_2^{(pb)} \\ \vdots \\ f_{N_p}^{(pb)} \end{bmatrix}
\]

(21)

\[
f_i^{(pb)} = \kappa_{pb} e_h(c_k, e^{j\omega})
\]

(22)

\[
\delta = \begin{bmatrix} \delta_i \\ \delta_r \end{bmatrix}
\]

(23)

where \( \delta_i \) is the vector update for \( c_k \), \( \delta_r \) is the scalar update for \( \tau \), and \( \kappa_{pb} \) is a constant. The elements of the last column of \( D_k^{(pb)} \) in (20) are all zeros since (19) is independent of \( \tau \).

C. Amplitude response in the stopband

The frequency response update for the differentiator at the \( k \)th iteration is given by

\[
H(c_k + \delta, e^{j\omega}) \approx H(c_k, e^{j\omega}) + \nabla H(c_k, e^{j\omega})^T \delta 
\]

(24)

In the stopband, the type of noise that may require attenuation may not always be white. If the spectrum of the noise in the stopband is known in advance, a weight \( W_s(\omega) \) can be incorporated in (24) so that more emphasis can be given by to frequency components with higher noise power; i.e.,

\[
W_s(\omega) H(c_k + \delta, e^{j\omega}) \approx W_s(\omega)[H(c_k, e^{j\omega}) + \nabla H(c_k, e^{j\omega})^T \delta]
\]

(25)

In such cases, \( W_s(\omega) \) can correspond to the normalized magnitude spectrum of the noise in the stopband. If the stopband noise is white, as assumed in all our experiments in Section IV, then \( W_s(\omega) \) is set to unity.

By using the same approach as in Section II-B, the \( L_p \) norm of the weighted frequency response in the stopband can be approximated as

\[
E_p^{(sb)}(k) \approx \| D_k^{(sb)} \delta + f_k^{(sb)} \|_p 
\]

(26)

where

\[
D_k^{(sb)} = \begin{bmatrix} \kappa_{sb} W_s(\omega_1) \nabla H(c_k, e^{j\omega_1})^T & 0 \\ \vdots & \vdots \\ \kappa_{sb} W_s(\omega_{N_s}) \nabla H(c_k, e^{j\omega_{N_s}})^T & 0 \end{bmatrix}, \quad \omega_i \in \Psi_s
\]

(27)

\[
f_k^{(sb)} = \begin{bmatrix} f_1^{(sb)} \\ f_2^{(sb)} \\ \vdots \\ f_{N_s}^{(sb)} \end{bmatrix}, \quad \omega_i \in \Psi_s
\]

(28)

\[
f_i^{(sb)} = \kappa_{sb} W_s(\omega) H(c_k, e^{j\omega})
\]

(29)

In the above equations, \( \Psi_s \) corresponds to the set of frequency points in the stopband and \( \kappa_{sb} \) is a constant.

D. Optimization problem

The optimization can be carried out by minimizing the group-delay deviation under the constraints that the passband error and stopband attenuation are within prescribed levels. The design of a lowpass differentiator can be obtained by solving the optimization problem

\[
\text{minimize} \quad E_p^{(pb)}(k) 
\]

subject to:

\[
E_p^{(pb)} \leq \Gamma_p \\
\text{passband error function} \leq \Gamma_p
\]

(30)

\[
\text{subject to: } \| D_k^{(sb)} \delta + f_k^{(sb)} \|_p \leq \Gamma_{sb} \\
\text{stopband gain} \leq \Gamma_{sb}
\]

(31)

where \( \delta \in \mathbb{R}^{6J-1} \) is the optimization variable. The optimum value of \( \delta \) is then used to update the optimizing parameters for the next iteration. Note that variables \( \delta_{bm}^{(1)} \) and \( \delta_{bm}^{(2)} \) are included within the vector \( \delta \).
In the design of IIR differentiators, the typical approach is to minimize the maximum passband amplitude-response error and maximum phase-response error. For the former, this can be done by making the value of $p$ large when computing the $L_p$ norm for the parameter in (19). However, for the latter it is more appropriate to take the $L_1$ norm of (10) since the group delay is a derivative of the phase and minimization of the $L_1$ norm of the group-delay error corresponds to better reduction of the maximum phase-response error than the minimization of the $L_\infty$ norm; this would be effective under the additional constraint that the average group-delay error in the passband is zero. For the stopband, the aim is to minimize the noise power, and therefore the $L_2$ norm is more appropriate. Furthermore, as in [20], we also include slack variable $\delta_{rlx}$ in the passband error constraint in case the initialization filter does not satisfy the maximum passband-error constraint. With these modifications, the problem in (31) becomes

$$\begin{align*}
\text{minimize} & \quad ||C_\delta \delta + d_k||_1 + V \delta_{rlx} \\
\text{subject to:} & \quad \text{sum} ||C_\delta \delta + d_k||_1 = 0 \\
& \quad ||D_k^{(p)} \delta + e_k^{(p)}||_\infty \leq \Gamma_p + \delta_{rlx} \\
& \quad ||D_k^{(b)} \delta + e_k^{(b)}||_2 \leq \Gamma_b \\
& \quad ||\delta||_2 \leq \Gamma_{small} + \delta_{rlx} \\
& \quad \delta_{rlx} \geq 0 \\
& \quad |r_{bn}^{(1)}(k) + \delta_{bn}^{(1)}| \leq 1 - \epsilon_s \quad \forall \ m \in [1, J] \\
& \quad |r_{bn}^{(2)}(k) + \delta_{bn}^{(2)}| \leq 1 - \epsilon_s \quad \forall \ m \in [1, J]
\end{align*}$$

where

$$\text{sum}[X] = \sum_i x_i$$

(33)

$\delta$ and $\delta_{rlx}$ are optimization variables, and $V$ is a positive weighing factor for the relaxation variable.

Note that as in [20], the group delay can be fixed to a prescribed value or optimized. However, in some applications it is desirable that the optimized group delay be small. In such cases, we can constrain the desired group delay $\tau$ in (9) to be below a prescribed upper bound $\Gamma_{gd}$. Such a constraint is given by

$$\tau \leq \Gamma_{gd} + \delta_{rlx}$$

(34)

where slack variable $\delta_{rlx}$ is also included if $\tau$ is greater than $\Gamma_{gd}$ during initialization. The minimization of the group-delay deviation instead of the phase-response error in (32) would result in a sign ambiguity in the final solution. This can be corrected simply by checking the sign of the final solution and multiplying the transfer function by $-1$ if the sign is reverse.

The optimization problem in (32) can be easily expressed as a second-order cone programming (SOCP) problem as in [21] and solved using efficient SOCP solvers such as the one available in the MATLAB SeDuMi optimization toolbox [24].

III. DESIGN OF DIGITAL DIFFERENTIATORS

In this section, we first describe a procedure for designing the lowest even- and odd-order filters that satisfy or nearly satisfy the amplitude-response constraints, which are then used for obtaining the initialization filters. We then describe the algorithm for the design of differentiators.

A. Lowest order filters satisfying the passband amplitude-response constraints

To find the lowest even- and odd-order filters that satisfy or nearly satisfy the amplitude response constraint, we modify the algorithm developed in [25] so that the absolute relative error is minimized instead of the squared amplitude-response error.

If we consider an IIR filter with the transfer function

$$H_m(z) = \frac{\sum_{i=0}^{m} b_i z^{-i}}{\sum_{j=0}^{n} a_j z^{-j}}$$

then setting $z = e^{j\omega}$ we can express the squared amplitude response as

$$\frac{N(\omega)}{D(\omega)} = |H_m(e^{j\omega})|^2 = H_m(e^{j\omega})H_m(e^{-j\omega})$$

$$p_m + \sum_{i=1}^{m} 2p_i \cos(\omega i)$$

$$q_n + \sum_{i=1}^{n} 2q_i \cos(\omega i)$$

where $p_m, \ldots, p_m$ and $q_n, \ldots, q_n$ are the numerator and denominator filter coefficients, respectively, of the product $H_m(z)H_m(z^{-1})$ such that $p_i = p_{-i}$ and $q_i = q_{-i}$. If $F_2(\omega)$ is the desired squared passband amplitude response of the differentiator, then the optimization algorithm in [25] can be used to find the filter coefficients that satisfy the constraints

$$\epsilon_l(\omega) \leq \left[ \frac{N(\omega)}{D(\omega)} - F_2(\omega) \right] \leq \epsilon_r(\omega), \quad \forall \ \omega \in \Psi_p$$

(37)

If $\delta_r$ is the maximum absolute relative error of the passband amplitude response, then

$$\frac{1}{|\omega|} \left[ \frac{N(\omega)}{D(\omega)} - \sqrt{F_2(\omega)} \right] \leq \delta_r, \quad \forall \ \omega \in \Psi_p$$

(38)

where $F_2(\omega) = \omega^2$. Now as shown in Appendix A, we can select $\epsilon_l(\omega)$ and $\epsilon_r(\omega)$ in (37) as

$$\epsilon_r(\omega) = (2\delta_r + \delta_r^2)\omega^2$$

(39)

$$\epsilon_l(\omega) = (2\delta_r - \delta_r^2)\omega^2$$

(40)

For the lowpass differentiator, an additional requirement is to limit the gain above the passband edge frequency so as to minimize any out-of-band high-frequency noise. One way to do this is to constrain the gain at $\omega = \pi$ to be below a certain threshold, that is,

$$\frac{N(\pi)}{D(\pi)} \leq \Gamma_p^2$$

(41)

where $\Gamma_p$ is the maximum allowable gain at $\omega = \pi$. Since the ideal gain of a fullband differentiator at $\omega = \pi$ is $\pi$, we can assume the upper limit for $\Gamma_p$ to be $\pi$. Consequently, the
optimization can be expressed as a linear programming (LP) problem given by

\[
\begin{align*}
\text{minimize} & \quad \nu \\
\text{subject to:} & \quad N(\omega) - D(\omega)[\omega^2 + \epsilon_r(\omega)] - \nu \leq 0 \\
& \quad -N(\omega) + D(\omega)[\omega^2 - \epsilon_l(\omega)] - \nu \leq 0 \\
& \quad N(\omega) \geq 0 \\
& \quad D(\omega) \geq 0 + \rho_s \\
& \quad N(\pi) - \Gamma_p^2 D(\pi) \leq 0 \\
& \quad \nu \geq 0
\end{align*}
\]

where \( \rho_s \) is a small positive constant used to ensure that the poles lie inside the unit circle. The above LP problem can be solved for \( \omega \in \Psi_p \) with \( \nu, p_i, \) and \( q_i \) as the optimization variables.

If the optimal value of \( \nu \) is close to zero, that is, \( \nu_{\text{opt}} \leq \epsilon_{\text{small}} \), then the solution would approximately satisfy the passband constraints and the next step is to recover the actual minimum-phase filter from the optimal values of \( p_i \) and \( q_i \). This is a straightforward step that can be carried out by using either spectral factorization [26] or a procedure described in [25].

For a fullband differentiator, the lowest filter order that would satisfy the passband constraint can be determined by means of the following procedure:

**Step 1:** Initialize the passband error, \( \delta_r \), and the passband sampling frequencies, \( \Psi_p \), to the prescribed values. Also set the starting filter order, \( M \), to 1, and \( \Gamma_p \) to a sufficiently large value greater than \( \pi \).

**Step 2:** For filter order \( M \), set \( m = n = M \) in (36) and solve the LP problem in (42).

**Step 3:** If the optimal value of \( \nu \) is close to zero (\( \nu_{\text{opt}} \leq \epsilon_{\text{small}} \)), the passband specification is satisfied; therefore, proceed to Step 4. Otherwise, set \( M = M + 1 \) and go to Step 2.

**Step 4:** The optimal values of \( p_i \) and \( q_i \) are used to obtained the lowest-order filter and the algorithm is terminated.

For the case of a lowpass differentiator, we use steps 1 to 4 above and then continue with the following additional steps:

**Step 5:** Without changing the filter order, find the smallest value of \( \Gamma_p \) between 0 and \( \pi \) that would satisfy the passband specification (i.e., \( \nu_{\text{opt}} \leq \epsilon_{\text{small}} \)) by solving the LP problem in (42) for different values of \( \Gamma_p \). This can be done by using a one-dimensional optimization procedure, such as the golden-section search [23], to find the optimum value of \( \Gamma_p \) between the bounds \([0, \pi]\); an accuracy of \( 10^{-2} \) is typically sufficient.

**Step 6:** The optimal values of \( p_i \) and \( q_i \) with the smallest value of \( \Gamma_p \) is then used to derive the lowest-order filter that would satisfy the passband error specification for the lowpass differentiator.

The next step is to find a second filter of order \( (M_{\text{low}} - 1) \) that has the smallest passband error, \( \delta_r \), while keeping \( \Gamma_p \) larger than \( \pi \). To find such a filter, we use the one-dimensional optimization procedure, as in Step 5 above, to derive a filter with the smallest value of \( \delta_r \) within the bounds \([0, 1]\) that satisfies \( \nu_{\text{opt}} \leq \epsilon_{\text{small}} \). However, if \( M_{\text{low}} = 1 \), which is the lowest possible order, the second filter is obtained by setting the filter order to 2 and then performing Steps 2 to 4 above for a fullband differentiator or Steps 2 to 6 for a lowpass differentiator.

The transfer functions of the two filters obtained, \( H_{\text{mag}}(z, M_1) \) and \( H_{\text{mag}}(z, M_2) \), are given by

\[
H_{\text{mag}}(z, M) = \prod_{i=1}^{M} \frac{z - r_{ai}e^{j\theta_i}}{z - r_{bi}e^{j\theta_i}}
\]

where

\[
\begin{align*}
M_1 &= M_{\text{low}} \\
M_2 &= \begin{cases} 
2 & \text{if } M_{\text{low}} = 1 \\
M_{\text{low}} - 1 & \text{otherwise}
\end{cases}
\end{align*}
\]

by letting \( M = M_1 \) or \( M_2 \).

**B. Initialization filters**

To obtain initialization filters of the desired filter orders, we add a number of biquadratic transfer functions to \( H_{\text{mag}}(z, M_1) \) and \( H_{\text{mag}}(z, M_2) \).

Two types of allpass transfer functions can be used. One possibility is to use

\[
H_{\text{ap}}^{(1)}(z, M_{\text{ap}}) = \begin{cases} 
1 & \text{if } M_{\text{ap}} = 0 \\
\prod_{i=1}^{M_{\text{ap}}} \frac{G_0}{G_0 - z^{-1}e^{j\theta_i}} & \text{otherwise}
\end{cases}
\]

where \( M_{\text{ap}} \) is the order of the transfer function,

\[
\theta_i = \frac{(i - 1)2\pi}{M_{\text{ap}}}
\]

and \( G_0 \) is a multiplier constant. The second possibility, \( H_{\text{ap}}^{(2)}(z, M_{\text{ap}}) \), can be obtained as follows: For an odd-order allpass transfer function, \( H_{\text{ap}}^{(2)}(z, M_{\text{ap}}) \) is obtained by rotating the pole-zero positions of \( H_{\text{ap}}^{(1)}(z, M_{\text{ap}}) \) by \( \pi \) radians in the \( z \) plane; on the other hand, for an even-order allpass transfer function, \( H_{\text{ap}}^{(2)}(z, M_{\text{ap}}) \) is obtained by rotating \( H_{\text{ap}}^{(1)}(z, M_{\text{ap}}) \) by \( \pi/2 \) radians either in the clockwise or counter-clockwise direction. Note that if the allpass transfer function is a multiple of 4, it can be easily shown that \( H_{\text{ap}}^{(1)}(z, M_{\text{ap}}) \) and \( H_{\text{ap}}^{(2)}(z, M_{\text{ap}}) \) are identical.

The transfer functions of the four initialization filters are given by

\[
\begin{align*}
H_{\text{init1}}(z) &= H_{\text{mag}}(z, M_1) \cdot H_{\text{ap}}^{(1)}(z, M_d - M_1) \\
H_{\text{init2}}(z) &= H_{\text{mag}}(z, M_1) \cdot H_{\text{ap}}^{(2)}(z, M_d - M_1) \\
H_{\text{init3}}(z) &= H_{\text{mag}}(z, M_2) \cdot H_{\text{ap}}^{(1)}(z, M_d - M_2) \\
H_{\text{init4}}(z) &= H_{\text{mag}}(z, M_2) \cdot H_{\text{ap}}^{(2)}(z, M_d - M_2)
\end{align*}
\]

where \( M_d \) is the differentiator order. Note that \( H_{\text{init1}}(z) \) and \( H_{\text{init2}}(z) \) are valid only if \( M_d \geq M_1 \), while \( H_{\text{init3}}(z) \) and \( H_{\text{init4}}(z) \) are valid only if \( M_d \geq M_2 \).
C. Passband phase-response error in the differentiator

If the average passband group delay of the differentiator is given by
\[ \bar{\tau} = \frac{1}{\omega_p} \int_{0}^{\omega_p} \tau_h(\omega) d\omega \] (49)
where \( \omega_p \) is the passband edge frequency and \( \tau_h(\omega) \) is the group delay of the differentiator, then the ideal phase response of the differentiator is given by
\[ \phi_{\text{ideal}}(\omega) = \frac{\pi}{2} - \omega \bar{\tau}, \quad \omega \in [0, 2\pi] \] (50)
The phase-response error is given by
\[ e_{\phi}(\omega) = \phi_h(\omega) - \phi_{\text{ideal}}(\omega), \quad e_{\phi}(\omega) \in [-\pi, \pi] \] (51)
where \( \phi_h(\omega) \) is the actual phase response of the differentiator. Consequently, the maximum peak-to-peak phase-response error, in degrees, is given by
\[ \xi_{\phi} = \frac{180}{\pi} \left[ \sup_{\omega} e_{\phi}(\omega) - \inf_{\omega} e_{\phi}(\omega) \right] \] (52)
Parameter \( \xi_{\phi} \) will be referred to as the maximum phase-response error hereafter.

D. Design procedure

The design of a digital differentiator that would satisfy prescribed specifications can be carried out using the following algorithm:

- **Step 1**: Compute the two lowest-order transfer functions, \( H_{\text{mag}}(z, M_1) \) and \( H_{\text{mag}}(z, M_2) \), using the procedure in Section III-A.
- **Step 2**: Set the desired differentiator order to \( M_d \) and compute the initialization filters in (48).
- **Step 3**: Solve the optimization problem in (32) for all the initialization filters derived in Step 2. For the fullband differentiator set \( \Gamma_{sb} = \infty \), while for the lowpass differentiator set it to the prescribed value.
- **Step 4**: Select the solution that has the smallest maximum phase-response error \( \xi_{\phi} \) and at the same time satisfies the passband error constraint; for the lowpass differentiator, the solution should also satisfy the stopband constraint.
- **Step 5**: If a solution is found that satisfies the phase-error specification in Step 4, stop. Otherwise, set \( M_d = M_d + 1 \) and go to Step 2.

E. Special case for differentiators with fixed group delay

In general, the average group delay of fullband differentiators with optimized group delay increases as the order of the differentiator is increased. The value of the average group delay usually follows that of the ideal fullband causal differentiator where the group delay is confined to \( \tau_n \) samples, where \( \tau_n \) is defined as
\[ \tau_n = 0.5 + n, \quad n \text{ is a nonnegative integer} \] (53)
In applications where the order of the differentiator is large, it may be desirable to have a differentiator with a smaller group delay at the expense of increased in amplitude- and phase-response errors. In such applications, a modified version of the design method in Section III-A can be used. Rather than finding the lowest-order filter that would satisfy the amplitude-response constraints, for differentiators with the smallest possible group delay we start from the opposite end by finding the prescribed highest and second-highest order filters that would satisfy the amplitude-response constraints, and then using the procedure in Section III-B we obtain the initialization filters. In this way, as the desired group delay is increased, the order of the filter that would satisfy the amplitude-response constraint is progressively decreased.

The same approach can be used for lowpass differentiators with fixed group delay.

F. Practical considerations

The frequency-dependent parameters are evaluated at frequency points that are sampled between \(-\pi\) and \(\pi\), such that the sample points between \(-\pi\) and 0 are the negative of the sample points between 0 and \(\pi\). To reduce the number of sample points and at the same time prevent spikes in the passband amplitude-response error function, the nonuniform variable sampling technique described in Chapter 16 of [22] can be used. Unlike the passband amplitude-response error, which is an \( L_{\infty} \) norm, the group delay and stopband errors are \( L_1 \) and \( L_2 \) norms, respectively, and hence the technique in [22] is not applicable. However, a uniform sampling can be used for these error functions.

The weight factor \( V \) for the relaxation parameter \( \delta_{rel} \) in (32) should not be too small, say, smaller than 100 as this could make the optimization algorithm unstable and prevent it from converging; at the same time, it should not be too large, say, larger than 10,000 as this can slow down the convergence. Values of \( V \) in the range 500 to 5000 were found to give good results.

To ensure that the optimization is not prematurely terminated, the termination condition is decided by monitoring the values of the objective function typically for the last 40 iterations as was done in [20].

IV. Experimental Results

In this section, we provide comparative experimental results to demonstrate the efficiency of the proposed method. Twelve design examples of various differentiator types are considered. Parameters \( \Gamma_{\text{small}} \) and \( V \) in (32) were set to 0.01 and 1000, respectively. The allpass transfer function, \( H_{\text{ap}}^{(1)}(z, M_{\text{ap}}) \), in (46) was initialized with \( r_k = 0.9 \). The default maximum pole radius was set to 0.98. A normalized sampling frequency of \( 2\pi \) was assumed in all design examples. The number of virtual and actual sample frequencies used in the nonuniform sampling technique [22] over the frequency range \(-\omega_p\) to \(\omega_p\) were 2000 and 68, respectively. Eight of the actual sample frequencies were uniformly distributed near the passband edge with a separation of \( 7.8 \times 10^{-4} \) rad/s between them. The group delay and stopband parameters, on the other hand, were uniformly sampled and evaluated using 800 uniformly sampled frequencies in the interval \([-\pi, \pi]\).

The amount of noise power in the stopband for the lowpass differentiator, assuming white Guassian noise, is proportional
Relative error in Tables VII, IX, XI, and XIII, respectively. The competing methods for Examples 1, 2, and 3 correspond to the third example in [10], the second example in [12], and the second example in [13], respectively. The required design specifications for these differentiators are given in Tables I, III, and V and the results obtained are summarized in Tables II, IV, and VI. The relative amplitude- and phase-response errors for Example 1 are plotted in Fig. 1. As can be seen in Fig. 1 and Tables II, IV, and VI, the IIR differentiators designed using the proposed method exhibit much smaller maximum phase-response error than the IIR differentiators designed using the competing method in [9].

### A. Examples 1, 2, and 3

The competing differentiators for Examples 1, 2, and 3 correspond to the third example in [10], the second example in [12], and the second example in [13], respectively. The required design specifications for these differentiators are given in Tables I, III, and V and the results obtained are summarized in Tables II, IV, and VI. The relative amplitude- and phase-response errors for Example 1 are plotted in Fig. 1. As can be seen in Fig. 1 and Tables II, IV, and VI, the IIR differentiators designed using the proposed method exhibit much smaller maximum phase-response error than the IIR differentiators designed using the competing method in [9].

### B. Examples 4, 5, 6, and 7

The design specifications for Examples 4 to 7 are given in Tables VII, IX, XI, and XIII, respectively. The competing differentiators for each of the examples correspond to the fourth, sixth, eighth, and thirteenth example in [9], respectively. Tables VIII, X, XII, and XIV and Fig. 2 show that the IIR differentiators designed using the proposed method have much smaller maximum phase-response error for practically the same passband relative amplitude-response error and average squared-amplitude response in the stopband, as the designs obtained with the competing method in [9].

### C. Examples 8, 9, and 10

In Example 8, we have designed fullband differentiators with fixed and optimized group delays using the proposed method and compared our designs with a competing differentiator taken from [19, eqn. (21)]. The design specifications are given in Table XV and the results obtained are summarized in Table XVI. From these results, we observe that differentiators designed with the proposed method have much smaller maximum phase-response error for practically the same passband relative amplitude-response error and average squared-amplitude response in the stopband. Note that the differentiator with optimized group delay has smaller phase-response error than the differentiator with fixed group delay but larger average group delay.

### Table I

**Fullband Differentiator Specifications for Example 1**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δr</td>
<td>0.0065</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

### Table II

**Design Results for Example 1 (Fullband Differentiator)**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [10]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>4</td>
<td>6</td>
<td>6</td>
</tr>
<tr>
<td>Max. rel. error, δr</td>
<td>0.006</td>
<td>0.0061</td>
<td>0.0065</td>
</tr>
<tr>
<td>Avg. group delay, φ</td>
<td>2.5</td>
<td>3.5</td>
<td>0.5</td>
</tr>
<tr>
<td>Max. phase error, ξφ</td>
<td>3.72</td>
<td>2.12</td>
<td>10.53</td>
</tr>
</tbody>
</table>

### Table III

**Fullband Differentiator Specifications for Example 2**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δr</td>
<td>0.055</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

### Table IV

**Design Results for Example 2 (Fullband Differentiator)**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [12]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>2</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>Max. rel. error, δr</td>
<td>0.05</td>
<td>0.05</td>
<td>0.055</td>
</tr>
<tr>
<td>Avg. group delay, φ</td>
<td>0.5</td>
<td>1.5</td>
<td>0.5</td>
</tr>
<tr>
<td>Max. phase error, ξφ</td>
<td>7.12</td>
<td>2.06</td>
<td>12.05</td>
</tr>
</tbody>
</table>

### Table V

**Fullband Differentiator Specifications for Example 3**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δr</td>
<td>0.035</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

### Table VI

**Design Results for Example 3 (Fullband Differentiator)**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [13]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>6</td>
<td>3</td>
<td>3</td>
</tr>
<tr>
<td>Max. rel. error, δr</td>
<td>0.031</td>
<td>0.031</td>
<td>0.0317</td>
</tr>
<tr>
<td>Avg. group delay, φ</td>
<td>0.5</td>
<td>1.5</td>
<td>0.5</td>
</tr>
<tr>
<td>Max. phase error, ξφ</td>
<td>8.26</td>
<td>3.26</td>
<td>12.05</td>
</tr>
</tbody>
</table>
TABLE VII
LOWPASS DIFFERENTIATOR SPECIFICATIONS FOR EXAMPLE 4

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δ_r</td>
<td>0.015</td>
</tr>
<tr>
<td>Maximum ASAR in SB</td>
<td>1.2</td>
</tr>
<tr>
<td>Passband edge, rad/s</td>
<td>0.7π</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband

TABLE VIII
DESIGN RESULTS FOR EXAMPLE 4 (LOWPASS DIFFERENTIATOR)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Max. rel. error, δ_r</td>
<td>0.01</td>
<td>0.0115</td>
</tr>
<tr>
<td>ASAR in SB, P_{sb}</td>
<td>1.09</td>
<td>1.184</td>
</tr>
<tr>
<td>Avg. group delay in PB, τ</td>
<td>2.02</td>
<td>1.24</td>
</tr>
<tr>
<td>Max. phase error, ξ_0</td>
<td>12</td>
<td>28.16</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband

Fig. 2. Plots of amplitude response, relative amplitude-response error and phase-response error for the proposed method (solid curves) and the method in [9] (dashed curves) for Example 4.

TABLE IX
LOWPASS DIFFERENTIATOR SPECIFICATIONS FOR EXAMPLE 5

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δ_r</td>
<td>0.016</td>
</tr>
<tr>
<td>Maximum ASAR in SB</td>
<td>0.45</td>
</tr>
<tr>
<td>Passband edge, rad/s</td>
<td>0.29π</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband

TABLE X
DESIGN RESULTS FOR EXAMPLE 5 (LOWPASS DIFFERENTIATOR)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>4</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Max. rel. error, δ_r</td>
<td>0.015</td>
<td>0.015</td>
<td>0.0155</td>
</tr>
<tr>
<td>ASAR in SB, P_{sb}</td>
<td>0.397</td>
<td>0.397</td>
<td>0.418</td>
</tr>
<tr>
<td>Avg. group delay in PB, τ</td>
<td>3.71</td>
<td>7.15</td>
<td>2.53</td>
</tr>
<tr>
<td>Max. phase error, ξ_0</td>
<td>1.52</td>
<td>0.30</td>
<td>8.26</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband; PB: passband

TABLE XI
LOWPASS DIFFERENTIATOR SPECIFICATIONS FOR EXAMPLE 6

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δ_r</td>
<td>0.04</td>
</tr>
<tr>
<td>Maximum ASAR in SB</td>
<td>0.55</td>
</tr>
<tr>
<td>Passband edge, rad/s</td>
<td>0.5π</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband

TABLE XII
DESIGN RESULTS FOR EXAMPLE 6 (LOWPASS DIFFERENTIATOR)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>3</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Max. rel. error, δ_r</td>
<td>0.035</td>
<td>0.035</td>
<td>0.036</td>
</tr>
<tr>
<td>ASAR in SB, P_{sb}</td>
<td>0.498</td>
<td>0.494</td>
<td>0.503</td>
</tr>
<tr>
<td>Avg. group delay in PB, τ</td>
<td>3.37</td>
<td>3.37</td>
<td>2.08</td>
</tr>
<tr>
<td>Max. phase error, ξ_0</td>
<td>1.71</td>
<td>0.0032</td>
<td>5.25</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband; PB: passband

TABLE XIII
LOWPASS DIFFERENTIATOR SPECIFICATIONS FOR EXAMPLE 7

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δ_r</td>
<td>0.07</td>
</tr>
<tr>
<td>Maximum ASAR in SB</td>
<td>0.95</td>
</tr>
<tr>
<td>Passband edge, rad/s</td>
<td>0.5π</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband

TABLE XIV
DESIGN RESULTS FOR EXAMPLE 7 (LOWPASS DIFFERENTIATOR)

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method 1</th>
<th>Proposed method 2</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>3</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Max. rel. error, δ_r</td>
<td>0.06</td>
<td>0.06</td>
<td>0.067</td>
</tr>
<tr>
<td>ASAR in SB, P_{sb}</td>
<td>0.939</td>
<td>0.939</td>
<td>0.944</td>
</tr>
<tr>
<td>Avg. group delay in PB, τ</td>
<td>2.31</td>
<td>4.46</td>
<td>1.66</td>
</tr>
<tr>
<td>Max. phase error, ξ_0</td>
<td>1.74</td>
<td>0.025</td>
<td>11.75</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband; PB: passband

TABLE XV
FULLBAND DIFFERENTIATOR SPECIFICATIONS FOR EXAMPLE 8

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, δ_r</td>
<td>0.15</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

TABLE XVI
DESIGN RESULTS FOR EXAMPLE 8 (FULLBAND DIFFERENTIATOR)

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>4</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Max. rel. error, δ_r</td>
<td>0.1</td>
<td>0.1</td>
<td>0.11</td>
</tr>
<tr>
<td>Avg. group delay, τ</td>
<td>3.5</td>
<td>0.5</td>
<td>0.5</td>
</tr>
<tr>
<td>Max. phase error, ξ_0</td>
<td>0.00056</td>
<td>7.44</td>
<td>11.27</td>
</tr>
</tbody>
</table>

OGD: optimized group delay; FGD: fixed group delay
In Examples 9 and 10, we compare lowpass differentiators where the group delay of the differentiator in the proposed method is constrained to be equal to or less than that in the competing design; this is done by incorporating the inequality constraint in (34) in the optimization problem in (32). To observe how the performance changes with and without the group-delay constraint, we have used the design specifications and competing differentiators in Examples 7 and 8 for Examples 9 and 10, respectively. The design results for the two examples are tabulated in Tables XVII and XVIII. The poles and zeros for our proposed designs are given in [27]. From the results, we observe that the proposed design method yields differentiators that have smaller phase-response errors and average group delay than the competing methods. Upon comparing the designs in Examples 9 and 10 obtained with our method with the corresponding designs in Examples 7 and 8 obtained with our method, we observe that the designs in Examples 9 and 10 offer lower group delay at the expense of increased phase-response error.

### D. Examples 11 and 12

In Examples 11 and 12, we compare the fullband and lowpass IIR differentiator designs with a corresponding optimal FIR design. The design specifications for these examples are given in Tables XIX and XXI, respectively. The optimization was carried out for various differentiator orders by varying the number of additional first-order filter sections. A differentiator order of \( N = 41 \) was required to satisfy the specifications in Table XIX for the fullband FIR differentiator and this was designed using the Remez Exchange algorithm described in Chapter 15 of [22]. On the other hand, the specification in Table XXI for the lowpass FIR differentiator required an order of \( N = 60 \) and the number of zeros at \(-1\) in the \( z \) plane set to \( K = 27 \). This was designed using the Selesnick-Type III design method [28]. The results obtained and the number of arithmetic operations per sampling period are presented in Tables XX and XXII. We have assumed a cascade realization of second-order sections both for the IIR and FIR differentiators. For the IIR differentiators, we have assumed a direct-canonic realization which would require a total of \( 2N + 1 \) multiplications, \( 2N \) additions, and \( N \) unit delays per sampling period where \( N \) is the differentiator order [22]. In the case of the FIR differentiator, \((N+1)/2\) multiplications, \( N \) additions, and \( N \) unit delays would be required per sampling period in view of the symmetry property of the transfer function coefficients in constant group-delay filters. From Tables XX and XXII, we observe a clear trade-off between filter complexity and group delay versus maximum phase-response error. It is apparent that the IIR differentiators offer a significant reduction in the number of arithmetic operations and system latency but at the cost of a nonzero phase-response error. For most applications, a perfectly linear-phase response is not required and a value of \( \xi_{\phi} \) in the range of 1 to 10, depending on the application, would be entirely acceptable. In such applications, a significantly more economical and efficient IIR design would be possible.

### E. Examples 13 to 18

In [27], we have included more comparisons of IIR fullband differentiators to demonstrate the effectiveness of our proposed method. The competing differentiators also include design examples from [11], [14], and [18].

### V. Conclusion

A method for the design of fullband and lowpass IIR digital differentiators that would satisfy prescribed speci-

---

**TABLE XVII**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method (CGD)</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Max. rel. error, ( \delta_r )</td>
<td>0.0143</td>
<td>0.0155</td>
</tr>
<tr>
<td>ASAR in SB, ( F_{as} )</td>
<td>0.396</td>
<td>0.418</td>
</tr>
<tr>
<td>Avg. group delay in PB, ( \tau )</td>
<td>1.76</td>
<td>2.53</td>
</tr>
<tr>
<td>Max. phase error, ( \xi_{\phi} )</td>
<td>5.66</td>
<td>8.26</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband; PB: passband; CGD: constrained group delay

---

**TABLE XVIII**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Proposed method (CGD)</th>
<th>Method in [9]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter order</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>Max. rel. error, ( \delta_r )</td>
<td>0.035</td>
<td>0.036</td>
</tr>
<tr>
<td>ASAR in SB, ( F_{as} )</td>
<td>0.498</td>
<td>0.503</td>
</tr>
<tr>
<td>Avg. group delay in PB, ( \tau )</td>
<td>1.452</td>
<td>2.08</td>
</tr>
<tr>
<td>Max. phase error, ( \xi_{\phi} )</td>
<td>2.63</td>
<td>5.25</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband; PB: passband; CGD: constrained group delay

---

**TABLE XIX**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, ( \delta_r )</td>
<td>0.0046</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
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</table>

**TABLE XX**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum rel. error, ( \delta_r )</td>
<td>0.0046</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
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</table>

**TABLE XXI**

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Values</th>
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</thead>
<tbody>
<tr>
<td>Maximum rel. error, ( \delta_r )</td>
<td>0.00095</td>
</tr>
<tr>
<td>Maximum ASAR in SB</td>
<td>0.65</td>
</tr>
<tr>
<td>Maximum pole radius</td>
<td>0.98</td>
</tr>
</tbody>
</table>

ASAR: average squared amplitude-response; SB: stopband
Substituting (57) in (37) and simplifying, we get
\[ e_s^2 \text{ amplitude-response error bounds} \]

\[ \text{A. Relationships between absolute-relative-error bounds and squared amplitude-response error bounds} \]

The squared amplitude-response error in (37) can also be expressed as
\[ \frac{N(\omega)}{D(\omega)} - F_d(\omega) = \omega^2 e_h^2(\omega) + 2\omega \sqrt{F_d(\omega)} e_h(\omega) \]  
(55)

where \( e_h(\omega) \) is the relative error which is given by
\[ e_h(\omega) = \frac{1}{|\omega|} \sqrt{\frac{N(\omega)}{D(\omega)} - \sqrt{F_d(\omega)}} \]
(56)

For a differentiator, \( F_d(\omega) = \omega^2 \) and hence (55) becomes
\[ \frac{N(\omega)}{D(\omega)} - F_d(\omega) = \omega^2 [e_h^2(\omega) + 2e_h(\omega)] \]
(57)

Substituting (57) in (37) and simplifying, we get
\[ |e_h^2(\omega) + 1| \leq 1 + \frac{\epsilon_r(\omega)}{\omega^2} \]
(58)

\[ |e_h^2(\omega) + 1| \geq 1 - \frac{\epsilon_l(\omega)}{\omega^2} \]
(59)

With the assumption that \( |e_h(\omega)| \ll 1 \), the term \( |e_h(\omega) + 1| \) is always positive. Consequently, (58) and (59) simplify to
\[ e_h(\omega) \leq \sqrt{1 + \frac{\epsilon_r(\omega)}{\omega^2}} - 1 \]  
(60)

\[ e_h(\omega) \geq \sqrt{1 - \frac{\epsilon_l(\omega)}{\omega^2}} - 1 \]  
(61)

If \( \delta_r \) is the maximum relative error, we have
\[ e_h(\omega) \leq \delta_r \]  
(62)

\[ e_h(\omega) \geq -\delta_r \]  
(63)

Equating (60) and (61) with (62) and (63), respectively, and simplifying, we get
\[ \epsilon_r(\omega) = (2\delta_r + \delta_r^2)\omega^2 \]  
(64)

\[ \epsilon_l(\omega) = (2\delta_l - \delta_l^2)\omega^2 \]  
(65)

REFERENCES
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