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Farrow-Structure-Based Reconfigurable Bandpass Linear-Phase FIR Filters for Integer Sampling Rate Conversion

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***Abstract*—A class of Farrow-structure-based reconfigurable bandpass FIR filters for integer sampling rate conversion is introduced. The converters are realized in terms of a number of fixed linear-phase FIR subfilters and two sets of reconfigurable multipliers that determine the passband location and the conversion factor, respectively. Both Mth-band and general FIR filters can be realized, and the filters work equally well for any integer factor and passband location. Design examples are included, demonstrating their efficiency compared with modulated regular filters. In addition, in contrast to regular filters, the proposed ones have considerably fewer filter coefficients that need to be determined in the filter design process.**

***Index Terms*—Bandpass converters, convex optimization, linear-phase FIR filters, low complexity, Mth-band filters, reconfigurable converters, sampling rate converters.**

I. Introduction

THIS brief deals with interpolation and decimation filters, which are fundamental signal processing blocks as sampling rate conversions are required in many different contexts [1]–[3]. Specifically, a class of Farrow-structure-based reconfigurable bandpass FIR filters for integer sampling rate conversion is introduced. Reconfigurable converters are required in, for example, emerging communication systems that are to support several different standards and operation modes [4]–[6]. A main feature of the proposed class of filters is that it facilitates efficient realization of reconfigurable conversion factors and passband locations. This is achieved by utilizing Farrow-structure-based reconfigurable fractional delay (FD) FIR filters [7]–[12] and cosine modulation [3]. This way, the converters are realized in terms of a number of fixed linear-phase FIR subfilters and two sets of reconfigurable multipliers corresponding to the modulation and FD multipliers, respectively. When changing the sampling rate conversion factor and the passband location, it suffices to alter the values of the multipliers in these two sets, respectively.

The bandpass filters introduced here are extensions of the low-pass filters introduced in [13], and like those filters, they can be used to realize both Mth-band and general interpolation and decimation filters. Another feature of this class of filters is that it is not restricted to certain conversion factors or passband locations but works equally well for any integer factor and location. This is in contrast with simpler schemes that may be efficient for some cases but fail in general [14], [15]. To exemplify, low-pass conversion by 2, 4, and 8 can be supported by a cascade of three low-complexity converters, each one converting by a factor of 2 or through the use of coefficient decimation [15]. It is, however, obvious that such techniques face problems in general, particularly in the bandpass case where the passband location may also vary. Furthermore, in contrast to regular filters, the proposed ones have considerably fewer filter coefficients that need to be determined in the filter design (optimization) process, which is beneficial from the convergence and design time points of view.

Following this introduction, Section II reviews the low-pass filters in [13], which are used as prototype filters for the bandpass filters derived in Section III. Section IV discusses the design and provides a design example, whereas Section V concludes this brief.

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II. Reconfigurable Low-Pass Filters

For conversion by an integer M, we write the transfer function $H_{\mathrm{LP}}(z)$ in polyphase form according to

$$H_{\mathrm{LP}}(z) = \sum_{m=0}^{M-1} z^{-m} H_m(z^M) \tag{1}$$

where $H_m(z), m = 0, 1, \ldots, M-1$, are the polyphase components [3]. The converters are efficiently realized using the corresponding polyphase structures. In the interpolator structure, the input signal is fed into the M polyphase components, producing M output signals $y_m(n)$, which are then interleaved to form the overall higher rate output as $y(m) = y_{\mathrm{mod}(m,M)}(\lfloor m/M \rfloor)$ [3]. The decimator structure is obtained from the corresponding interpolator structure using transposition [3]. The following text reviews how the polyphase components are formed in the low-pass filters in [13], which are used as prototype filters for the bandpass filters in Section III. For that purpose, the conversion factor in the low-pass filter is always even, which, therefore, is assumed below.

The polyphase components should ideally equal $z^{-(N/2-m)/M}/M$ for a unity-gain overall low-pass filter of order N and, thus, all-pass filters with FDs m/M [16]. In [13], the ideal all-pass responses are approximated using the Farrow structure, except for the zeroth component, which is approximated with a conventional filter. To be precise, $H_0(z)$ is an N_0th-order type-I linear-phase FIR filter, which also means that the overall order N is appropriately selected as $N = N_0 M$. The remaining polyphase components $H_m(z)$, $m = 1, 2, \ldots, M-1$, are realized using the Farrow structure, i.e.,

$$H_m(z) = \sum_{k=0}^{L} d_m^k G_k(z). \tag{2}$$

There are basically two different cases, depending on whether the order of the subfilters $G_k(z)$ is odd or even [13], but the basic technique is essentially the same in both cases. Due to the limited space, we concentrate on odd-order subfilters in this brief. In this case, $G_k(z)$ are type-II (symmetrical) and type-IV (antisymmetrical) linear-phase FIR filters for even and odd values of k, respectively, whereas the FD coefficient values d_m are selected as

$$d_m = -m/M + 1/2 \tag{3}$$

which means they exhibit antisymmetry according to $d_m = -d_{M-m}$. Furthermore, the order of $G_k(z)$, e.g., N_1, must be chosen as $N_1 = N_0 - 1$. The filter and parameter selections above ensure that the overall filter $H_{\mathrm{LP}}(z)$ is always a type-I linear-phase FIR filter. Furthermore, two different classes of filters are obtained depending on how the zeroth polyphase component $H_0(z)$ is selected. If $H_0(z)$ is a pure delay, the overall filter becomes an Mth-band filter as every Mth impulse response value then becomes zero,[1] except for the center tap that is equal to $1/M$ for a unity-gain filter. If $H_0(z)$ is instead a general type-I linear-phase FIR filter, general interpolation and decimation filters can be obtained.

A main advantage of using this class of filters is that it facilitates converters with different conversion factors using the same set of fixed subfilters $G_k(z)$. It thus suffices to change the set of FD coefficients d_m^k when the conversion factor is changed. In other words, this class of filters is efficient for reconfigurable converters. The following section extends this class to bandpass filters.

III. Reconfigurable Bandpass Filters

Assume now that we are to realize a bandpass filter for the integer conversion factor M and with a passband located (ideally) between $r\pi/M$ and $(r+1)\pi/M$, $r \in [0, M-1]$. To this end, we start with a linear-phase low-pass prototype filter, i.e., $P(z)$, belonging to the class of low-pass filters described in Section II, but for the conversion factor $2M$ and, consequently, with the (ideal) passband ranging from 0 to $\pi/(2M)$. The prototype filter $P(z)$ is then appropriately modulated (frequency shifted) to obtain the desired bandpass filter, as detailed below.

A. Transfer Function and Structure

The low-pass prototype filter transfer function can be written in polyphase form as

$$P(z) = \sum_{m=0}^{2M-1} z^{-m} P_m(z^{2M}) \tag{4}$$

[1]In general, it is not necessarily the zeroth polyphase component that is a pure delay in an Mth-band FIR filter, but for the proposed class of filters, this is assumed *a priori* due to the selection $N = N_0 M$.

where $P_0(z)$ is a regular N_0th-order type-I linear-phase FIR filter, whereas the remaining $P_m(z)$, $m = 1, 2, \ldots, 2M-1$, are given by

$$P_m(z) = \sum_{k=0}^{L} d_m^k G_k(z) \tag{5}$$

with d_m being given by (3), but with M replaced with $2M$, and with $G_k(z)$, as defined in Section II. The prototype filter $P(z)$ is then modulated by

$$\omega_r T = (r + 1/2) \times \pi/M \tag{6}$$

to the right and to the left and added to form the bandpass filter $H_{\mathrm{BP}}(z)$ with a passband at the desired location. As a consequence, the overall transfer function is formed as

$$H_{\mathrm{BP}}(z) = e^{j\Theta} P(ze^{j\omega_r T}) + e^{-j\Theta} P(ze^{-j\omega_r T}) \tag{7}$$

where

$$\Theta = \omega_r T \times N/2 \tag{8}$$

with N being the order of $P(z)$ and $H_{\mathrm{BP}}(z)$. The factors $e^{j\Theta}$ and $e^{-j\Theta}$ are included to make $H_{\mathrm{BP}}(z)$ a linear-phase filter as well. Inserting (4) into (7), one obtains

$$H_{\mathrm{BP}}(z) = \sum_{m=0}^{2M-1} W_m z^{-m} P_m(-z^{2M}) \tag{9}$$

where

$$W_m = 2\cos\left(\omega_r T(m - N/2)\right). \tag{10}$$

Writing $H_{\mathrm{BP}}(z)$ as in (1), one obtains the polyphase components from (9) as

$$H_m(z) = W_m P_m(-z^2) + z^{-1} W_{m+M} P_{m+M}(-z^2). \tag{11}$$

Inserting (5) into (11) finally yields

$$H_m(z) = W_m \sum_{k=0}^{L} d_m^k G_k(-z^2) + z^{-1} W_{m+M} \sum_{k=0}^{L} d_{m+M}^k G_k(-z^2). \tag{12}$$

Each polyphase component in the bandpass converter can thus be implemented by combining two polyphase branches in the corresponding low-pass converter after replacing z with $-z^2$ and after appropriate weighting and delay, as shown in Fig. 1. The first branch is a general type-I linear-phase FIR filter as it is not generated via the Farrow structure. It is given by

$$H_0(z) = 2P_0(-z^2) \tag{13}$$

which is because $W_M = 0$ (see Section III-C). As the corresponding decimator structure can readily be obtained by transposing the interpolator structure, it is not shown here.

Like the low-pass converter, the bandpass converter is thus realized in terms of the fixed part $G_k(-z^2)$ and a reconfigurable part. The main difference is that we have here two sets of

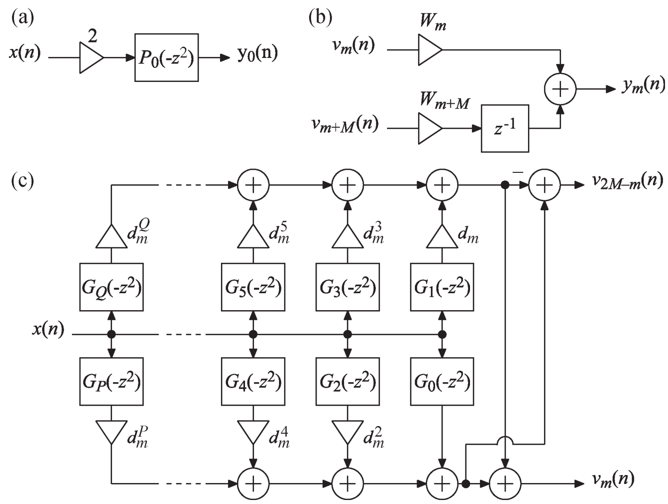

Fig. 1. Realization of the proposed reconfigurable bandpass interpolator in terms of fixed subfilters $P_0(z)$ and $G_k(-z^2)$ and reconfigurable modulation multipliers W_m and FD multipliers d_m^k. (a) Realization of the zeroth polyphase branch. (b) Realization of the mth polyphase branch for $m = 1, 2, \ldots, M-1$. (c) Realization of the polyphase branches m and $2M-m$ for $m = 1, 2, \ldots, M-1$ in the low-pass prototype interpolator, but with z replaced with $-z^2$, as required for the bandpass interpolator.

variable multipliers since modulation multipliers W_m are also required in addition to the FD multipliers d_m^k. When changing the sampling rate conversion factor, it suffices to alter the values of the multipliers in these two sets and, of course, the number of polyphase branches. It is also noted that the bandpass converter structure can be used for low-pass converters as well and is obtained for $r = 0$. This can be utilized for efficiently supporting both low- and bandpass converters as one otherwise needs two sets of fixed filters, namely, $G_k(z)$ for the low-pass converters and $G_k(-z^2)$ for the bandpass converters.

B. Impulse Response

The impulse response of the overall filter can be expressed in terms of the prototype filter impulse response as

$$h(n) = 2p(n)\cos\left(\omega_r T(n - N/2)\right). \qquad (14)$$

Furthermore, the impulse response values $p(2Mn)$ satisfy

$$p(2Mn) = p_0(2Mn), \qquad n = 0, 1, \ldots, N_1 + 1 \qquad (15)$$

where $p_0(2Mn)$ is the impulse response of the polyphase component $P_0(z^{2M})$ in (4), whereas $p(2Mn+m)$, $m = 1, 2, \ldots, 2M-1$, are the weighted sums of the impulse response values $g_k(n)$ according to

$$p(2Mn+m) = \sum_{k=0}^{L} d_m^k g_k(n), \qquad n = 0, 1, \ldots, N_1. \qquad (16)$$

In addition, recall that $p(n) = p(N-n)$ since it is an Nth-order type-I linear-phase FIR filter.

C. Mth-Band Property Preserved

If the prototype filter $P(z)$ is a unity-gain $2M$th-band filter, $P_0(z)$ is a pure delay, namely, $P_0(z^{2M}) = z^{-N/2}/(2M)$. It then follows from (13) that the bandpass filter $H_{\mathrm{BP}}(z)$ is an Mth-band filter. Equation (13) holds because $W_M = 0$, which is due to the fact that $N = 2MN_0$, and N_0 is an even integer. Indeed, we have from (6), (8), and (10) that

$$W_M = 2\cos\left((r + 1/2)\pi(1 - N_0)\right). \qquad (17)$$

Since r is an integer and N_0 is an even integer, it follows that $(r+1/2)(1-N_0)$ is an integer plus a half, which implies that W_M becomes zero.

D. Implementation Complexity

The number of multipliers and adders required, i.e., C_m and C_a, are given by

$$C_m = \sum_{i=1}^{4} C_{mi} \quad C_a = \sum_{i=1}^{4} C_{ai} \qquad (18)$$

where the indexes $i = 1, 2, 3$, and 4, respectively, correspond to the complexities emanating from the subfilter $P_0(-z^2)$; all subfilters $G_k(-z^2)$, $k = 0, 1, \ldots, L$; all FD multipliers d_m^k and d_{m+M}^k, $k = 1, 2, \ldots, L$, $m = 1, 2, \ldots, M-1$; and all modulation multipliers W_m and W_{m+M}, $m = 1, 2, \ldots, M-1$. Recalling that the orders of the filters $P_0(z)$ and $G_k(z)$ are $N_0 = N_1 + 1$ and N_1, noting that the substitution $z \rightarrow -z^2$ does not change the arithmetic complexity, and utilizing the coefficient symmetries of $P_0(z)$, $G_k(z)$, and d_m^k, one finds that

$$C_{m1} = \frac{N_1+3}{2} \qquad C_{m2} = \frac{(L+1)(N_1+1)}{2}$$
$$C_{m3} = L(M-1) \qquad C_{m4} = 2(M-1) \qquad (19)$$
$$C_{a1} = N_1 + 1 \qquad C_{a2} = N_1(L+1)$$
$$C_{a3} = (L+1)(M-1) \quad C_{a4} = M-1. \qquad (20)$$

One also finds that the number of delay elements, i.e., C_d, is

$$C_d = 2(N_1+1) + M - 1 \qquad (21)$$

as the delay elements can be shared between the subfilters. Thus, the term $2(N_1+1) = 2N_0$ emanates from the subfilters $P_0(-z^2)$ and $G_k(-z^2)$, whereas the term $M-1$ comes from the additional delays shown in Fig. 1(b). In the Mth-band case, $P_0(-z^2)$ is a pure delay in which case one has $C_{a1} = 0$ and $C_{m1} = 0$ ($C_{m1} = 1$) for interpolators (decimators) assuming energy preservation in the sampling rate process, which means that the passband gain in the interpolation case (decimation case) is M (unity). It is finally noted that all these figures indicate the operations required per input/output sample in the interpolator/decimator. Alternatively, one can instead compute the multiplication rates, i.e., R_m and R_a, which are the operations required per output/input sample and thus obtained as $R_m = C_m/M$ and $R_a = C_a/M$.

IV. Filter Design and Design Examples

The impulse response of the overall bandpass filter is a weighted linear combination of subfilter impulse responses, according to (14)–(16). For each conversion factor, the reconfigurable modulation and FD multipliers are fixed. The filter design problem can therefore be posed as a convex optimization problem, the solution of which is globally optimum. This holds true also when the filter is designed to support several different conversion factors and passband locations. Each pair of passband location and conversion factor, with the associated fixed modulation multipliers W_m and FD multipliers d_m^k, corresponds to one specification. However, since the only unknowns in the design are the filter coefficients of the subfilters $G_k(z)$, which are common to all different modes, one can design all filters simultaneously by solving one overall convex optimization problem. This has been explained in detail in [13] and will therefore not be repeated here. The only differences from [13] are that the passband and stopband regions are defined differently and that the impulse responses differ due to the cosine multiplications in (14).

It is also stressed that, in the filter design (optimization) process, the filter coefficients of $G_k(z)$ are variables (unknowns), whereas each pair of sets of multipliers W_m and d_m^k contains predetermined and, thus, fixed values. The number of free parameters to optimize is therefore few, which is advantageous from the convergence and design time points of view. Once the optimization is done, the coefficients of $G_k(z)$ are fixed in the implementation of Fig. 1, whereas the multipliers W_m and d_m^k are reconfigurable. It is then the change from one pair of sets of the reconfigurable multipliers to another that alters the sampling rate conversion factor and the passband location.

Example: Consider a bandpass filter that is to support all integer conversion factors between 2 and 10, and all passband locations, i.e., for each $M = M_i$, $i = 2, 3, \ldots, 10$, all passband locations from $r_i\pi/M$ to $(r_i + 1)\pi/M$, $r_i \in [0, M_i - 1]$, are to be supported. This means that 54 different converters in total are to be supported. It is further assumed that all the corresponding filters should have a stopband attenuation of at least 40 dB and that they be Mth-band filters with transition bands covering 40% of the ideal passband width, which, for the bandpass filters (low/high-pass filters), corresponds to 60% (80%) utilization of the ideal passband width. Using the proposed bandpass filter, the specifications are met with $L = 3$ and subfilters $G_k(z)$ of order $N_1 = 5$, resulting in some 44-dB attenuation. This means that all 54 different bandpass filters are simultaneously designed to meet their respective specification by optimizing only $(L + 1)(N_1 + 1)/2 = 12$ filter coefficients. Fig. 2 shows the magnitude responses for $M = 2$, 5, and 10. The different parameters are summarized in Table I.

As a comparison, we use regular Mth-band linear-phase FIR filters for the prototype filter together with the same modulation technique as that used in the proposed filters. The results are compiled in Table II.[2, 3] The overall complexity can again be divided in the same way as for the proposed filters, which is why the same notation is used. The regular filters do not contain any FD multipliers, which is why the corresponding figures for C_{m3} and C_{a3} are zero.

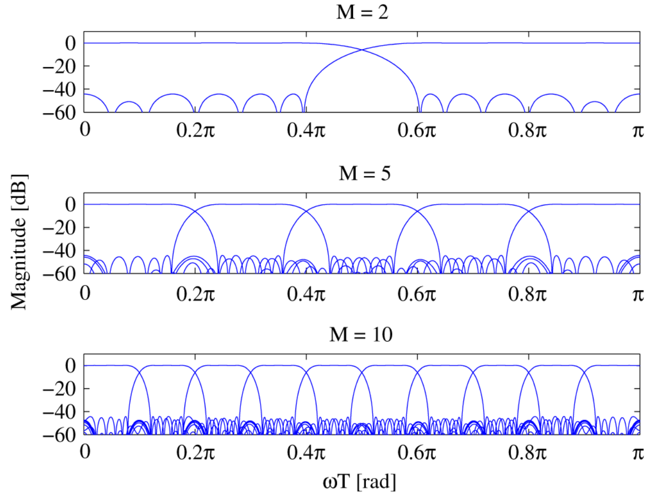

Fig. 2. Magnitude response of the proposed filter in the example for $M = 2$, 5, and 10.

It is shown from the tables that the proposed filters are more efficient for each value of M, except for $M = 2$ and possibly $M = 3$, depending on the cost of different operations. This is also explained by the different behavior with regard to the multiplication and addition rates (R_m and R_a) versus M, which is in line with the results for the basic low-pass filter [13]. Moreover, a main advantage of the proposed filters is that they support several different conversion factors and passband locations more economically. Using regular filters in this example, one needs to design nine separate and different filters and store their coefficients. The overall number of coefficients is the sum of the figures in the C_{m2} column in Table II, which should be compared with the sum of the figures in the C_{m3} column in Table I plus 12 (as the 12 filter coefficients in the C_{m2} column in Table I are the same for all M). The tables reveal that 508 coefficients need to be stored for the regular filters but only 147 for the proposed filters.[4] One should also take the modulation multipliers into consideration, corresponding to C_{m4}, but they are the same in both cases. Moreover, using the proposed filters, the FD multipliers, which correspond to C_{m3}, can alternatively be readily computed online through (3) when changing one filter to another. It is not necessary to store all of these coefficients, which may offer even further savings compared with the regular filters. A price to pay is a slight increase in filter order, delay, and number of delay elements, typically only some 10%, as seen when comparing the filter orders in the N and C_d columns in Tables I and II. Recall that the delay here is $N/2$.

V. Conclusion

This brief has introduced a class of reconfigurable bandpass FIR filters for integer sampling rate conversion. As illustrated

[2]The results in Table II have been derived by first designing an optimum half-band filter (for $M = 2$) and then, from the so-obtained result, computing the figures for the remaining M through scaling, taking into account that the order of an FIR filter is inversely proportional to the transition bandwidth and that one polyphase branch of Mth-band filters is always a pure delay. Designing all filters properly (in a similar way as for the proposed ones) may give slightly different results, but the figures in Table II are accurate enough for a qualitative comparison with those in Table I.

[3]When computing the number of multipliers in Table II, we have not taken into account that there exist symmetries between the coefficients. As opposed to the fixed converter case [17], it is not obvious that one can benefit from such symmetries in a reconfigurable converter.

[4]The complexity of the regular filters can be reduced using coefficient decimation [15] in which case the filters for $M = 2$, 3, 4, and 5 are obtained from the filters with $M = 8, 6, 8$, and 10, respectively. However, this only saves 116 coefficients, which means that 392 coefficients still need to be stored.

TABLE I
RESULTS OF THE EXAMPLE FOR THE PROPOSED FILTERS

M	N	C_d	C_{m1}	C_{m2}	C_{m3}	C_{m4}	C_m	R_m	C_{a1}	C_{a2}	C_{a3}	C_{a4}	C_a	R_a
2	24	13	0	12	3	2	17	8.50	0	20	4	1	25	12.50
3	36	14	0	12	6	4	22	7.33	0	20	8	2	30	10.00
4	48	15	0	12	9	6	27	6.75	0	20	12	3	35	8.75
5	60	16	0	12	12	8	32	6.40	0	20	16	4	40	8.00
6	72	17	0	12	15	10	37	6.17	0	20	20	5	45	7.50
7	84	18	0	12	18	12	42	6.00	0	20	24	6	50	7.14
8	96	19	0	12	21	14	47	5.88	0	20	28	7	55	6.88
9	108	20	0	12	24	16	52	5.78	0	20	32	8	60	6.67
10	120	21	0	12	27	18	57	5.7	0	20	36	9	65	6.50

TABLE II
RESULTS OF THE EXAMPLE FOR THE REGULAR FILTERS

M	N	C_d	C_{m1}	C_{m2}	C_{m3}	C_{m4}	C_m	R_m	C_{a1}	C_{a2}	C_{a3}	C_{a4}	C_a	R_a
2	22	12	0	12	0	2	14	7.00	0	11	0	1	12	6.00
3	34	13	0	24	0	4	28	9.33	0	22	0	2	24	8.00
4	44	14	0	34	0	6	40	10.00	0	31	0	3	34	8.50
5	56	15	0	46	0	8	54	10.80	0	42	0	4	46	9.20
6	66	16	0	56	0	10	66	11.00	0	51	0	5	56	9.33
7	78	17	0	68	0	12	80	11.43	0	62	0	6	68	9.71
8	88	18	0	78	0	14	92	11.50	0	71	0	7	78	9.75
9	100	19	0	90	0	16	106	11.78	0	82	0	8	90	10.00
10	110	20	0	100	0	18	118	11.80	0	91	0	9	100	10.00

through an example, this filter class features lower complexity and fewer coefficients compared with a set of modulated regular filters. It is noted, however, that the proposed filters are more attractive for applications with many modes, such as in the example. In special cases, with a few modes, it may be beneficial to use limited schemes, as mentioned in Section I, but they fail in general as opposed to the proposed ones that can handle any integer conversion factor and center frequency. It is also mentioned that there are other techniques for realizing variable single-rate bandpass filters [18], but they are not appropriate for the multimode sampling rate conversion considered here. Finally, it is noted that there are additional aspects to consider in order to minimize the complexity of the proposed filters, particularly the use of unequal-order subfilters and the setting of the smallest FD multipliers to zero. These issues are discussed in more detail in [13], which introduced the basic low-pass filters.

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