Designing $M$-Channel IIR Uniform DFT Filter Banks Using Interpolated IIR Technique

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Abstract—This paper introduces a method of designing an $M$-channel, causal, stable and perfect reconstruction (PR), infinite impulse response (IIR), parallel uniform discrete Fourier transform (DFT) filter banks. The basic idea behind this is based on an efficient technique of designing and implementing narrowband lowpass finite impulse response (FIR) filters, called the interpolated FIR (IFIR) technique. In this paper, we have extended the IFIR concept to the IIR filter, known as Interpolated IIR (IIIR), and applied the concept to IIR DFT filter banks. It has been shown that this technique is good for IIR DFT filter banks in polyphase oversampled case and also has a modular structure which is therefore very useful for very large scale integration (VLSI) implementation.

Index Terms—Computational complexity, Filter bank, Interpolated filter, IIR filter, Perfect reconstruction.

I. INTRODUCTION

The $M$-CHANNEL uniform DFT filter bank is a particular class of the $M$-channel subband coding scheme shown in Fig. 1, where the input signal is split into equal bands. Perfect reconstruction (PR) IIR filter banks are very attractive because of their potentially low system delays and better frequency response when compared to their FIR counterparts [1]. In practice, designing IIR DFT filter banks (FBs) is quite involved and generally has been restricted to only the two-channel case [2]-[4]. In fact, the difficulty in designing such filter banks lies in satisfying the complicated PR condition and the causality-stability requirements of the filters. Recently, some $M$-channel FBs have been presented [5], [6]. In the first method, the design includes a complicated stabilization procedure of the synthesis filters. In the second method, which is based on the geometrical progression expansion of the denominator of the prototype filter, there is no flexibility on the number of channels. In other words, the number of channels is a function of the prototype filter order.

In this paper, a method of designing a class of complex perfect reconstruction IIR DFT filter banks using uniform bandwidth, stable and causal polyphase components is proposed. The proposed procedure has the same transfer function for all the polyphase components of its prototype filter and therefore it is simpler to satisfy the PR condition and the causality-stability requirements. The present work deals with the design of analysis filters. Using a polyphase approach, the synthesis filters are automatically known. The main advantages of the proposed method are summarized below:

- These filter banks satisfy the PR and causality-stability conditions.
- This technique can be used to design filter banks for an arbitrary number of channels ($M$), which is a multiple of the decimator/expander factor ($D$).
- The analysis and synthesis filters are obtained easily from the prototype filter and their implementation is very efficient. Since they have modular structures and are highly parallel, they are ideal candidates for VLSI implementation.
- The optimization (to obtain the prototype filter) is done directly with a given set of specifications by the classical methods.

The paper starts with Section II, detailing some considerations on stable-causal, PR conditions. Section III explains a general design method with arbitrary frequency specifications for the prototype filter. By allowing the number of polyphase components of the prototype filter to be either a multiple of the decimator/expander factor $(D)$ or equal to the decimator/expander factor $(D)$ itself, we have two different structures, i.e., polyphase oversampled and critically sampled filter bank structures. This Section also presents the realizations of the parallel structures. Section IV gives several examples, which are helpful in understanding the proposed method. Finally, Section V summarizes the results.

II. UNIFORM DFT FILTER BANK ANALYSIS AND PERFECT RECONSTRUCTION CONDITIONS

The $M$-channel subband coding filter bank is shown in Fig. 1, where the input signal is split into different subbands. In the uniform DFT FB case, the analysis filters...
\[ H_k(z) \] are all frequency translated versions of a prototype lowpass filter. Thus, \( H_k(z) = H_0(ze^{-j\pi k/M}) \), \( k = 0, 1, ..., M-1 \), where \( H_0(z) \) is the prototype lowpass filter. Same relations also hold for synthesis filters. The reconstructed signal generally \( y(n) \) differs from \( x(n) \) due to distortions such as aliasing, imaging, amplitude and phase. This is due to the non-ideal filters. However, with a proper choice of the prototype filter, the system is free from all these distortions. Such a system satisfies the relation \( y(n) = c x(n-n_0) \), and is called a perfect reconstruction system, where \( c \) and \( n_0 \) are constants. Using the polyphase decompositions [1], we can express the analysis and synthesis filter banks of Fig. 1 as

\[
H_k(z) = \sum_{l=0}^{M-1} z^{-l} E_{kl}(z^M) \\
G_k(z) = \sum_{l=0}^{M-1} z^{-l} R_{kl}(z^M) \tag{1}
\]

where \( E_{kl}(z^M) \) and \( R_{kl}(z^M) \) are the polyphase elements of \( H_k(z) \) and \( G_k(z) \), type I and type II, respectively. More generally, it can be shown [1] that the system has PR property if and only if

\[
R(z)E(z) = cz^{-m_0} \begin{bmatrix} 0 & I_{M-r} \\ -1 & 0 \end{bmatrix} \tag{2}
\]

where \( E(z) = [E_{kl}(z)] \) and \( R(z) = [R_{kl}(z)] \) are the polyphase matrices, \( 0 \leq r \leq M-1 \), \( m_0 \) is an integer, \( c \neq 0 \) is a constant and \( I_r \) is the \( r \times r \) identity matrix.

Under this condition, the reconstructed signal is \( y(n) = c x(n-n_0) \), where \( n_0 = Mm_0 + r + M-1 \). In the uniform DFT FBs, polyphase matrices have the following expressions as derived in [1]:

\[
E(z) = W^* \begin{bmatrix} E_0(z) & 0 & ... & 0 \\ 0 & E_1(z) & ... & 0 \\ ... & ... & ... & ... \\ 0 & 0 & ... & E_{M-1}(z) \end{bmatrix} \tag{3}
\]

\[
R(z) = \begin{bmatrix} R_0(z) & 0 & ... & 0 \\ 0 & R_1(z) & ... & 0 \\ ... & ... & ... & ... \\ 0 & 0 & ... & R_{M-1}(z) \end{bmatrix} \tag{4}
\]

where \( W \) is the \( M \times M \) DFT matrix with elements \( e^{-j2\pi kl/M} \) (\( W^* \) is the transpose-conjugate of \( W \)); \( E_k(z) \) and \( R_k(z) \) are the polyphase components of the analysis and the synthesis prototype filters, respectively.

For simplicity, we consider the PR condition with \( r = 0 \), \( m_0 = 0 \) and \( c = 1.0 \), hence, the condition for PR could be written as: \( R(z)E(z) = I_M \), and by substituting (3) in (2), since the DFT and the IDFT cancel each other out, we will have:

\[
\begin{bmatrix} E_0(z) & 0 & ... & 0 \\ 0 & E_1(z) & ... & 0 \\ ... & ... & ... & ... \\ 0 & 0 & ... & E_{M-1}(z) \end{bmatrix} \begin{bmatrix} R_0(z) & 0 & ... & 0 \\ 0 & R_1(z) & ... & 0 \\ ... & ... & ... & ... \\ 0 & 0 & ... & R_{M-1}(z) \end{bmatrix} = \begin{bmatrix} 1 & 0 & ... & 0 \\ 0 & 1 & ... & 0 \\ ... & ... & ... & ... \\ 0 & 0 & ... & 1 \end{bmatrix}
\]

Then, it is sufficient that \( R_k(z) = 1/E_k(z) \), for \( k = 0, 1, ..., M-1 \). It is known that if the prototype analysis filter is causal and stable, then the components \( E_k(z) \) are also causal and stable [1]. In the synthesis part, the resulting components \( E_k(z) \) are also causal, but to be stable, it is required that \( E_k(z) \) is minimum phase for all \( k \). The minimum phase condition is fulfilled if all zeros of the system function \( E_k(z) \) lie within or on the unit circle [7]. However, \( E_k(z) \) is assumed to be strictly minimum phase (or strictly Hurwitz) if all zeros are within the unit circle [1]. Therefore, \( R_k(z) \) is stable.

III. PROPOSED M-CHANNEL PR IIR FILTER BANK

The proposed method is based on a structure previously presented by Neuvo et al. [8] as an efficient technique for the design and implementation of narrowband lowpass filter, called the interpolated FIR (IFIR) technique. In this paper, we have extended the IFIR concept to IIR filters and applied it for the design and implementation of stable-causal, IIR DFT FBs. We propose to design the prototype lowpass filter as the following structure:

\[
H_0(z) = F(z^D)/I(z). \tag{5}
\]

The first section of the filter \( F(z) \) is an IIR filter and is called the model filter. The second one, \( I(z) \), is an FIR filter and is called the interpolated or image suppressor filter. To explain the basic idea, consider Fig. 2(a), which shows the lowpass filter specifications (ripples are not shown). Let \( N \) denote the required filter order. Now, instead of meeting these specifications, suppose we try to meet the \( D \)-fold stretched specifications, i.e. Fig. 2(b). The stretched filter \( F(z) \) has a transition bandwidth of \( D\Delta f \), where \( \Delta f = f_s - f_p \). Since the order of FIR filter is proportional to the inverse of the transition bandwidth, the order decreases to \( N/D \). Hence, this is an effective approach to implement a narrowband lowpass filter with a significant savings in the number of arithmetic operations and is called IFIR technique [8]. In the current case (IIR filters), the order of a lowpass filter meeting the given passband and stopband specification is a function of the transition ratio \( \tan(\pi f_p)/\tan(\pi f_s) \), where \( f_p \) and \( f_s \) are the passband and stopband edge normalized frequencies, respectively [9]. According to \( \tan(\pi f_p)/\tan(\pi f_s) \), if we can keep the ratio \( f_p/f_s \) a constant, then the order of the
stretched filter does not change and hence the interpolation technique by itself is not useful for IIR filters. If \( F(z) \) is an IIR filter which meets the stretched specifications, the system which meets the original specifications is given by (5). The filter \( F(z^D) \) has \( D-1 \) unwanted passband images, in addition to the desired passband centered at \( f = 0 \), as shown in Fig. 2(c). The image suppressor filter \( I(z) \), which is an FIR lowpass filter, eliminates these unwanted passbands. The transition bandwidth of \( I(z) \) depends on \( f_p, f_s \) and \( D \); it is \( \Delta f = 1/D - f_s - f_p \). For the linear phase FIR lowpass filter meeting the specifications given in Fig. 2(d), it has an order of \( M_0 \approx \Delta f \left| \frac{\zeta(\delta_p, \delta_s)}{\Delta f} \right| \), where \( \zeta(\delta_p, \delta_s) \) is a function of the peak passband ripple \( \delta_p \) and peak stopband attenuation \( \delta_s \) [9]. A well-known formula that has been proposed for estimating the order of this type holds good for equiripple design (Kaiser’s formula) as well [9]:

\[
M_0 = \frac{10 \log_{10}(\Delta f \delta_s) + 13}{14.6 \Delta f} .
\]

Suppose we design a proper filter with \( I(z) \) and \( F(z) \), such that it meets all the desired specifications of the prototype filter \( H_0(z) \) and \( F(z) \) is a minimum phase filter. We rewrite (5) in the following fashion:

\[
H_0(z) = F(z^D) I(z) = F(z^D) \sum_{k=0}^{M-1} a_k z^{-k}
\]

Using the oversampled polyphase components representation as [6], [10]:

\[
H_0(z) = \sum_{k=0}^{M-1} z^{-k} E_k(z^D)
\]

This means that all analysis/synthesis polyphase components are equal to 1. The reconstructed signal could be written as [1], [6]:

\[
Y(z) = \sum_{k=0}^{M-1} z^{-(M-1-k)} \sum_{l=0}^{D-1} z^{-k} X(z e^{\pi i l / D}) e^{-\pi i 2 kl / D}
\]

Assuming \( M = n_0 D \) and knowing that \( \sum_{k=0}^{M-1} e^{-\pi i 2 kl / D} = M \delta(l) \) (where \( \delta(l) \) is the discrete-time impulse function), the FB output is:

\[
Y(z) = \frac{M}{D} z^{-(M-1)} X(z) = n_0 z^{-(M-1)} X(z).
\]

Since all \( E_k(z) \) filters are minimum phase (causal and

\[
E_k(z) = a_k F(z^l), \quad 0 \leq k \leq M - 1
\]

where \( \{ a_k \} \) is the coefficients of the FIR filter, \( I(z) \). As we have shown in (11), for PR, the number of channels \( M \), or the order of \( I(z) \) must be a multiple of the decimator/expander factor \( D \), where for \( M = D \), we get a polyphase critically sampled filter bank structure and for \( M = n_0 D \), a polyphase oversampled filter bank structure is resulted. Moreover, all the polyphase elements are the scaled versions of the same filter as shown in (12). From (4), the corresponding synthesis polyphase elements \( R_k(z) \) are IIR filters with:

\[
R_k(z) = \frac{1}{E_k(z)} = \frac{1}{a_k F(z^l)}, \quad k = 0, 1, ..., M - 1.
\]
stable), i.e., all the zeros and poles of \( E_k(z) \) are strictly inside the unit circle; all \( R_k(z) \) filters are also causal and stable. Therefore, using the classical IIR digital filter design techniques, we can design a perfect reconstruction IIR DFT filter bank with stable and causal polyphase elements by making use of the following procedure:

- Design a stable, causal (with \( F(z) \) minimum phase) filter consistent with the transfer function given in (5) as a prototype analysis filter with a given set of specifications.
- Find the coefficients \( \{ a_k \} \) of the FIR filter \( I(z) \) and set \( R_k(z) = \frac{1}{a_k F(z)} \).

All of the \( \{ a_k \} \) parameters should be non-zero. Otherwise, PR is not possible, since some synthesis filters with \( R_k(z) \) will have infinite coefficients. The important point is in the selection of a suitable order for the FIR filter \( I(z) \), which suppresses all the unwanted images. It can be seen from (8) and Fig. 2(d) that the required order of \( I(z) \) \( (M_0) \) is a function of the transition bandwidth \( (\Delta f = 1/(Df) - f_p - f_q) \) and the prototype filter specifications \( (\delta_p, \delta_q) \). On the other hand, the order of \( I(z) \) is also equal to the number of channels, which is fixed as \( M \). Hence, for a good suppression, we have to try to reduce the required order \( M_0 \) such that \( M_0 \leq M \). This could be done by increasing the transition bandwidth \( \Delta f \) in (8), decreasing decimator/expander factor \( (D) \), or reducing the number of unwanted passband images. As we have seen, for PR, the number of channels \( M \) must be a multiple of the decimator/expander factor \( D \) \( (M = n_D D) \).

Based on this fact, we can choose a suitable \( D \) (or \( n_D \)). In the polyphase critically sampled case (maximally decimated FB, \( M = D \) or \( n_D = 1 \)), we have the minimum transition bandwidth and the unwanted passband images are maximum. Hence, for the FIR filter, \( I(z) \) requires an order \( (M_0) \) for good suppression, which is generally greater than the existing order \( M \), and the proposed method is generally not suitable for the critically sampled case.

The polyphase structure of Fig. 4(a) shows FB structure in the oversampled case. Using the noble identities [1] for polyphase filters, the decimators and the expanders are brought after the filters \( F(z^D) \) and before the filters \( F(z^D)^{-1} \). Hence, a second realization is obtained as shown in Fig. 4(b). The structure shown in Fig. 4(b) is equivalent to placing an IIR filter with system function \( F(z^D) \) in front of a purely FIR DFT filter bank with taps \( a_k \). Consequently, the filter with the system function \( F(z^D)^{-1} \) required after the FIR DFT synthesis bank is shown in Fig. 4(b). An important property of the filter banks, obtained from the prototype filter as given in (5) and polyphase components in (12) and (13), with the realizations shown in Fig. 4, is that they are very simple and modular. This is due to the same polyphase components in the proposed approach when compared to the general cases shown in Fig. 3.

The computational complexity is determined by the number of multiplications per unit time required to implement the IIR DFT filter bank. In the realization shown in Fig. 4(a), if we use the IFFT algorithm, the analysis filters need \( N \log_2(M) \) complex multiplications for the DFT. Also, for given polyphase components of order \( N \), the multiplication of complex input values with real coefficients requires \( 2M(N + 2) \) real multiplications for the polyphase elements. The statement about the complexity in the analysis filter part is also valid for the synthesis part. Therefore, for the entire filter bank, the number of real multiplications per unit time is \( (3M \log_2(M) + 4M(2N + 2))/D \). The factor \( D \) appears in the denominator, due to the reduced rate when compared to the input rate. We have considered three real multiplications for the multiplication of two complex numbers [11]. In the second realization given in Fig. 4(b), the number of real multiplications amounts to \( (3M \log_2(M) + 4M + 4D(2ND + 1))/D \).
Fig. 4. Realizations of the proposed stable, causal IIR DFT FB: (a) first realization and (b) second realization.

The computational complexity ratio is defined as

$$CCR = \frac{(3M \log_2(M) + 4M + 4D(2ND + 1))}{(3M \log_2(2M) + 4M(2N + 2))}. \quad (14)$$

The computational complexity ratio shows that the complexity of the structure proposed in Fig. 4(b), when compared to Fig. 4(a), depends on the oversampled ratio $r_0 (M = r_0 D)$. In other words, increasing the oversampled ratio $r_0$ caused a reduction in the amount of complexity of Fig. 4(b) when compared to Fig. 4(a) and vice versa. For example, with $M = 32$ and $D = 4$, the computational complexity ratio is 0.62, which means that the complexity of the structure shown in Fig. 4(b) is 62% of the computation cost of the structure shown in Fig. 4(a). In comparison to the conventional FIR [12], the IIR DFT filter banks proposed in this paper have a reduced order, for the same number of channels. The overall cost is higher and filter bank delays are much longer for real-time processing, when compared to the delay in the proposed filter bank.

IV. DESIGN EXAMPLES

An algorithm for the design of IFIR digital filters with the overall system function given in (5) is presented in [8]. It can be summarized as follows:

- From the given filter stopband edge frequencies, calculate $D_{\text{max}} (D_{\text{max}} = \lceil \pi / \omega_s \rceil = \lceil 1/2f_s \rceil)$ and select a suitable value for $D$ ($D < D_{\text{max}}$). Selection of $D$ determines the position of unwanted repetitions of the passband.
- Design the interpolator $I(z)$ to attenuate these repetitions of the passband to or below the stopband level.
- Design the model filter $F(z)$. The band edge frequencies of the model filter are obtained by multiplying the edge frequencies of $F(z^D)$ by $D$.

Designing an $M$-channel filter bank merely involves designing the prototype filter $H_0(z)$. The prototype $H_0(z)$ has a normalized cut-off frequency, $f_n = 1/2M$. To design a proper prototype filter based on interpolated
technique, we have applied the aforementioned algorithm. In this case, the order of interpolated filter $I(z)$ is given, i.e., $M$ the number of channels. Therefore, to design the interpolator $I(z)$ to suppress unwanted repetitions of the passband to stopband level, we should select a proper decimator/ expander factor $D$ ($D_{\text{max}} = M$). Assuming $f_n = (f_p + f_s)/2$ and noting that $M = r_0 D$, the transition bandwidth of the interpolator filter is

$$\Delta_f = 1/D - f_s - f_p = 1/D - 2f_n = (r_0 - 1)/M. \quad (15)$$

Substituting (15) in (6) and knowing that $M_0 \leq M$, we get:

$$r_0 \geq \frac{10 \log_{10}(\delta_p \delta_s) + 13}{14.6} + 1 \quad (16)$$

where $\delta_p$ and $\delta_s$ are the passband ripple and stopband attenuations for the prototype filter, respectively. Since we use FFT/IFFT algorithm, the number of the channel and hence $r_0$ should be of radix-2. Therefore, $r_0$ calculated in (16) should be rounded off to the nearest integer power of 2. The design of the interpolator filter $I(z)$ can be done using any FIR filter design software.

The next step is to design a lowpass IIR filter or model filter $F(z)$ with $D$-fold of edge frequencies of the prototype filter. The design of the model filter can be done using any IIR filter design program and its order depends on the desired specification of prototype filter, i.e., passband ripple, stopband attenuation and edge frequencies. The simulation of the proposed polyphase filters used in the realizations mentioned above in Section III is carried out in Matlab®. The simulation results are discussed in detail in the following examples:

**Example 4.1**

Let us consider a 16-channel perfect reconstruction IIR filter bank with the normalized cut-off frequency $1/32$, the passband ripple of $\delta_p = 0.05$ dB and the stopband attenuation of $\delta_s = 40$ dB. To design a proper prototype filter based on the aforementioned method, we should design the interpolator filter. Using (16), the oversampled ratio obtained is $r_0 \geq 2.84$. Hence, we select $r_0 = 4$ and decimator/expander factor is $D = 4$. Therefore, the interpolator filter is a lowpass FIR filter with order 15, transition bandwidth $\Delta_f = 3/16$ and the passband ripple and stopband attenuation as mentioned above. This filter is implemented using the classical method of windowed linear-phase FIR digital filter design [9], to which the Kaiser window is applied. Hence, the numerator coefficients $a_k$ will be symmetric, i.e., $a_k = a_{M-1-k}$. The next step is to design the model filter $F(z)$, with the aforementioned specifications and normalized frequency of $Df_n$. The filter’s order is calculated based on the given specifications. In this case, a 4th order Elliptic filter satisfies the above requirements. Fig. 5 shows the frequency response of the analysis filter designed by the proposed method in the polyphase oversampled case, for $M = 16$ and $D = 4$. It can be seen that the stopband attenuation and all the unwanted images in the stopband are suppressed very well.

**Example 4.2**

Let us design a 32-channel DFT filter bank with the normalized cut-off frequency of $f_n = 1/64$, the passband ripple and stopband attenuation of $\delta_p = 0.025$ dB and $\delta_s = 40$ dB, respectively. From (16), the oversampled ratio is $r_0 \geq 2.85$ and when we select $r_0 = 4$, the decimator/expander factor is $D = 8$. In this case, the order of the interpolator filter $I(z)$ is 31. Again the Kaiser window is applied to get a linear-phase FIR filter. Based on the desired specifications and the normalized frequency of $Df_n$, the model filter is designed as a 4th order Elliptic filter. Then, the overall system function of prototype filter is given by (5). Fig. 6 shows the frequency response of the analysis filter designed by the proposed method in the polyphase oversampled case, for $M = 32$ and $D = 8$.

**Example 4.3**

Let us design a prototype filter for a 64-channel DFT
filter bank with the normalized cut-off frequency of $f_n = 1/128$, the passband ripple and the stopband attenuation of $\delta_p = 0.1$ dB and $\delta_s = 60$ dB, respectively. By using the aforementioned approach, a decimator/expander factor $D = 8$ is needed to suppress unwanted repetitions of the passband to the stopband level. In this case, the order of the interpolator filter $I(z)$ is set to 63 and the Kaiser window is applied. The model filter is designed using a 5th order Elliptic filter. Then, the overall system function of the prototype filter is the multiplication of interpolated filter and the model filter as given in (5). Fig. 7 shows the frequency response of the interpolated, model and prototype filter designed by the proposed method. The analysis filters are the translated versions of the prototype filter.

Example 4.4

Fig. 8 shows the frequency response of some analysis filters for 128-channel DFT filter bank in the polyphase oversampled case, with $\delta_p = 0.05$ dB and $\delta_s = 45$ dB. In this case, the proposed decimator/expander factor is $D = 32$. It can be seen that we have got the desired specifications in the stopband and passband with this decimator/expander factor $D$. The order of model filter is set to 6 to satisfy the aforementioned specifications.

V. CONCLUSION

We have proposed an algorithm for designing an M-channel causal, stable and perfect reconstruction IIR DFT filter bank based on a structure known as interpolated IIR (IIIR) filter, which has been proposed earlier for the design of FIR filter (IFIR). The design procedure for an arbitrary channel with prototype filter specifications is included. Perfect reconstruction filter banks are always possible with stable and causal polyphase components. Two realizations with efficient implementations are discussed. Since the realizations are modular in structure, they are suitable for VLSI implementation. The proposed algorithm is very useful for stable, causal, IIR DFT filter banks in the polyphase oversampled case. Examples have been included to illustrate the performance of the design technique.

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REFERENCES

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