Number 12 Volume 22 December 2016 Journal of Engineering

A Hybrid Coefficient Decimation- Interpolation Based Reconfigurable Low Complexity Filter Bank for Cognitive Radio

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ABSTRACT

Non uniform channelization is a crucial task in cognitive radio receivers for obtaining separate channels from the digitized wideband input signal at different intervals of time. The two main requirements in the channelizer are reconfigurability and low complexity. In this paper, a reconfigurable architecture based on a combination of Improved Coefficient Decimation Method (ICDM) and Coefficient Interpolation Method (CIM) is proposed. The proposed Hybrid Coefficient Decimation-Interpolation Method (HCDIM) based filter bank (FB) is able to realize the same number of channels realized using (ICDM) but with a maximum decimation factor divided by the interpolation factor (L), which leads to less deterioration in stop band attenuation (SA). The proposed architecture is able to realize a greater number of sub-bands locations. The proposed (HCDIM) based (FB) shows an inherent low complexity offered by the (CIM) technique when compared with the alternative FBs. The reduction in the number of multiplications is by 50.77% compared with ICDM in non-uniform channelization, while the reduction in the number of multiplications is about 59.64% over the discrete Fourier transform (DFTFB) and 31.19% over ICDM based FB in uniform channelization.

Key words: filter bank (FB), hybrid coefficient decimation-interpolation method (HCDIM), coefficient interpolation method (CIM), low complexity.

مجموعة مرشحات قائمة على هجين من معامل الهلاك ومعامل الاستيفاء المعاد تشكيلها والمنخفضة التعقيد

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الخلاصة

تشاطر القنوات الغير منتظم مهمة حرجة في اجهزة الاستلام الراديوية المعرفية للحصول على قنوات منفصلة من اشارة ادخال رقمية الاتساع في فترات مختلفة من الزمن. المتطلبين الرئيسيين في شاطر القنوات هي اعادة التشكيل وانخفاض التعقيد. هذا البحث يقترح بنية قابلة لاعادة الهيكلية فيها الجمع بين طريقة معامل الهلاك المتطورة وطريقة معامل الاستيفاء. هذه الطريقة المختلطة المقترحة قادرة على تحقيق نفس العدد من القنوات التي تتحقق باستخدام طريقة معامل الهلاك المتطورة ولكن باستخدام معامل هلاك مقسوما على معامل الاستيفاء، الامر الذي يؤدي الى تدهور اقل في توهين حزمة التوقف. الهيكل المقترح قادر على تحقيق عدد اكبر من الحزم الفرعية. مجموعة المرشحات القائمه على اساس الهيكل المقترح تظهر تعقيد اقل وانخفاض في عدد عمليات الضرب نتيجة لاستخدام طريقة معامل الاستيفاء فيه مقارنة بباقي المرشحات .

Cognitive Radio (CR) can offer competent utilization of the radio, electromagnetic spectrum. The Simple principle of CR is detecting the spectral occupancy over a wide frequency range permitting unlicensed users (called secondary users) to have adaptable access of the available frequency bands prearranged to licensed users (called primary users), **Ambede, et al., 2012**. The role of CR is to use those unused spectrum allocations when the primary (i.e., licensed) users are not present without additional license, by approving a concept of dynamic spectrum resource managing, **Powell, et al., 2003**. The ability of CR to precisely detect the spectrum usage status over a wide frequency range attending various wireless communication standards ensures its successful use **Park, et al., 2009**. In a usual CR, the non-uniform channels of various wireless standards simultaneously coincide in the input signal. The spectrum sensing block need to be used to observe these channels accurately and the distinct channels must to be extracted using the channelizer, so demanding the capability of implementing multi-standard channelization. Digital FBs show a significant role in channelization and spectrum sensing in the CRs. The frequency range of wideband input is divided into non-uniform or uniform sub-bands using FBs in FB-based spectrum sensing, and the existence of signals is then sensed using other techniques for example energy detection **Ambede, et al., 2015**. A low complexity reconfigurable Finite Impulse Response (FIR) filters were proposed in, **Vinod,** and **Mahesh, 2008,** using coefficient decimation method (CDM). Variable frequency responses can be generated using CDM by working upon stable filter coefficients using two coefficient decimation operations, one to change the pass-band width of the modal filter (named CDM-II) and another to produce multi-band frequency responses (named CDM-I). A real valued design algorithm for oversampled FIR FB of five channels even order filter banks is proposed in, **Chougule,** and **Patil**, **2011**. The frequency bands and order of filter are selected which cause noise removal and reduction in amplitude distortion and in-band aliasing. This system is suitable for any image format. A Modified Coefficient Decimation Method (MCDM) was proposed in**, Vinod, et al., 2012,** to obtain reconfigurable FIR filters with improved frequency response flexibility and twice the center frequency resolution when related to the conventional CDM. A higher degree of reconfigurability can be provided using MCDM and the FIR filters obtained show less stop-band attenuation deterioration when compared with those obtained using conventional CDM, and have a lower complexity due to lower order modal filter in the MCDM. The modified CDM (MCDM-II) was combined with the conventional (CDM-II) in, **Ambede, et al., 2012**, to design a new channel filter, the new method termed as improved coefficient decimation method II (ICDM-II), in which the channel filter has a considerably lower complexity when compared to other channel filters based on CDM**, Mahesh,** and **Vinod, 2011.** A combination of MCDM-I and the conventional CDM-I is presented in, **Ambede, et al., 2014,** to employ uniform FB which is termed as the Improved coefficient decimation method I (ICDM-I). After performing ICDM-I operations on the modal filter, the desired sub-bands can be extracted from the multi-band frequency responses obtained, using low order wide transition bandwidth (TBW) frequency response masking (FRM) filters. The ICDM-I, ICDM-II and (FRM) technique are combined in, **Ambede, et al., 2014,** and named as an improved coefficient decimation method (ICDM). The ICDM eliminates the least common divisor LCM constraint in, **Ambede, et al., 2012,** as the compensation in group delay is not required because the filters resulting after performing ICDM-I operations have the similar filter order. But the transition bandwidth (TBW) is the minimum standard TBW divided by the decimation factor corresponds to that standard, which leads to a narrow TBW and higher order modal filter with increasing value of decimation factor. In this paper, a hybrid combination of ICDM and the CIM digital FB is proposed to serve uniform and non-uniform channelization. Using the proposed architecture (named HCDIM); the desired sub-bands extracted using Frequency Response asking (FRM) technique can be obtained. The procedure starts by inserting zero coefficients between every two coefficients of the modal filter and then applying the CDM on the resulting filter. In the proposed HCDIM-based FB, the overall sharp transition-band filter can be composed using wider transition-band modal filter by using the CIM technique. Hence, it offers inherent low complexity features. The number of the resulting sub-bands represents the product of the interpolation factor and the decimation factor; i.e. a particular sub-band can be obtained using smaller decimation factor than that used in ICDM method. The deterioration in the stop-band attenuation increases with increasing the decimation factor. As the decimation factor is reduced in the proposed approach, the stop-band deterioration is reduced too, i.e. less filter order. The HCDIM-based FB also has a higher flexibility of $\pi/(M \times L)$ in terms of the possible number and locations of its sub-bands when compared with the other FBs in the literature.

Outline of the Paper: In Section 2, a revision of coefficient decimation, the improved coefficient decimation method and an explanation for the coefficient interpolation method is presented. In Section 3, the proposed HCDIM-based FB is presented with the design steps of the proposed FB for various channelization scenario situations. The design examples are involved in section 4 as well as the complexity comparison of the HCDIM-based FBs designed using our method with that of the ICDM-FB designed for the same required conditions. A performance comparison in terms of pass-band ripple and stop-band attenuation of final filter and the flexibility of the proposed FB is presented in the same section. The conclusion is given in Section 5.

2 COEFFICIENT DECIMATION, IMPROVED COEFFICIENT DECIMATION METHOD AND COEFFICIENT INTERPOLATION METHOD

In the conventional CDM, **Vinod,** and **Mahesh, 2008,** if the coefficients of a low pass modal filter are decimated by a decimation factor *M*, i.e., reserving every *M*th coefficient and substituting the others by zeros, the resulting FIR filter will be a multi-band uniform sub-band bandwidth (BW) frequency response. The resultant center frequency locations of the sub-bands are given by $2\pi k/M$, where k is an integer ranging from 0 to (M - 1). If $H(e^{jw})$ represents the Fourier transform of the modal filter coefficients, then the Fourier transform of the resulting filter coefficients is given by:

$$H(e^{j\omega}) = \frac{1}{M}\sum_{k=0}^{M-1} H(e^{j(\omega-\frac{2\pi k}{M})}) \qquad (1)$$

This process is called CDM-I. After performing CDM-I by decimation factor *M*, if all the reserved coefficients in the resulting filter are collected together by removing the intermittent zeros, a low pass frequency response is obtained with its pass band and transition band widths *M* times that of the modal filter. This operation is called CDM-II.

When the coefficients of the modal filter are decimated by *M*, every *M*th coefficient is reserved and the sign of every alternative reserved coefficient is reversed. All the other filter coefficients are substituted by zeros. This operation is called the modified coefficient decimation method I (MCDM-I), **Ambede, et al., 2012**, and gives an FIR filter with a multiband frequency response with center frequencies of the sub-bands given by $(2k+1)\pi/M$. The Fourier transform of the resulting filter coefficients is represented by:

$$H(e^{j\omega}) = \frac{1}{M}\sum_{k=0}^{M-1} H(e^{j\left(\omega-\frac{\pi(2k+1)}{M}\right)}) \quad (2)$$

A high pass frequency response of a pass band and TBWs M times that of the modal filter can be obtained after grouping all the reserved coefficients in the MCDM-I together and removing the intermittent zeros. This operation is termed as MCDM-II.

It can be noted from Eq. (1) that the center frequency locations of the achievable sub-bands after performing CDM-I operations are even multiples of π/M, while the center frequency locations from Eq. (2) are odd multiples of π/M using MCDM-I operations. The ICDM-I is a combination of CDM-I and MCDM-I, and the ICDM-II is a combination of CDM-II and MCDM-II. A combination of ICDM-I and ICDM-II is performed in ICDM, **Ambede, et al., 2014**. In all ICDM operations, the order of the desired modal FIR filter (*N)* can be obtained using, **Bellanger, 1982**:

$$N = -\frac{4log_{10}(10 * \delta p * \delta s)}{3(f_s - f_p)} - 1 \quad (3)$$

where f_p is the desired pass band frequency and f_s is the desired stop band frequency (normalized in the range 0–1, with 1 corresponding to the Nyquist frequency) and δp is the desired pass band peak ripple and δs is the desired stop band peak ripple.

It can be noted that the increase in the value of M causes a deterioration in stop band attenuation (SA) of the resulting filters, **Vinod,** and **Mahesh, 2008,** and its mathematical expression is:

$$\delta_{s(mod)} = \frac{\delta_{s(final)}}{M} \quad (4)$$

where $\delta_{s(mod)}$ is the SA of the modal filter, and $\delta_{s(final)}$ is the SA of the resulting filter after performing a CDM and MCDM operations by M. The deterioration in SA can be overcome by overdesigning the modal filter. If a CDM is performed by a factor of M to a filter, the SA of the resulting filter can be kept within desired value δs from Eq. (3) and Eq. (4), and the minimum order of the overdesigned modal filter can be calculated using:

$$N = \left[-\frac{4log_{10}(10 * \delta p * \delta s)}{3(f_s - f_p)} - 1\right] + \frac{4log_{10}M}{3(f_s - f_p)} \quad (5)$$

The second term in Eq. (5) represents the rise in order of the overdesigned modal filter essential to compensate the SA deterioration happened after performing coefficient decimation method by M.

On the other hand, the Coefficient Interpolation Method (CIM) is an efficient technique to synthesize FIR filters with sharp transition bands, using wide transition-band filter with low complexity since the resulting filter will have many sparse coefficients, **Mahesh,** and **Vinod, 2007**. Suppose a low-pass modal filter $H_a(z)$ of order *N*, its frequency response can be shown in **Fig. 1** (a). If (L -1) zeros are added between every two coefficients of $H_a(z)$, which is equivalent to replacing each delay element of $H_a(z)$ by L delays, then the resulting L subbands of $H_a(z^L)$ are factor of L narrower than that of $H_a(z)$ as shown in Fig 1.b, where L is the interpolation factor. The resulting L multiband will lie on $(2\pi k)$ where ($k = 0,1, \ldots\ldots L$ -1). If the pass band edge and the stop band edge of the modal filter is f_p and f_s respectively as can be shown in **Fig. 1** (a), then the pass band edge, and the stop band edge of the resulting sub-band after performing (CIM) is $\frac{f_p}{L}, \frac{f_s}{L}$ respectively as shown in **Fig. 1** (b).

3 PROPOSED FILTER BANK

In this paper, a method to realize reconfigurable, low complexity filter bank based on combined decimation, interpolation and frequency response masking is proposed. The proposed architecture consists of three stages. The first stage is the design of modal filter. The second stage is interpolation by the L factor. The third stage is decimation by the M factor, resulting in an $(M \times L)$ sub-bands. The desired channels are extracted by using suitable masking filters to mask the unwanted channels. The complexities of the masking filters are low, as they have large TBWs. The principle of (FRM) was originally introduced in, **Lim, 1986**. The (HCDIM) proposed in this paper offers two grades of freedom, M and L. Therefore, this method pointedly improves the filter architecture flexibility and reduces the coefficient decimation factor M that used to extract the same channels number that extracted in ICDM. The number of extracted channels in the proposed method is (M×L).

The HCDIM operations are further clarified with the help of a descriptive example. Consider the modal filter in **Fig. 2** (a) that has a pass band frequency of 0.12, and a stop band frequency of 0.132. The peak pass band ripple is selected as 0.1 dB, whereas, the stop band attenuation of the filter is selected as -50dB. **Fig. 2** (b) represents the modal filter frequency response after interpolation by a factor of (L =3) i.e. by inserting (L -1= 2) zeros between modal filter coefficients. It can be noted from **Fig. 2** (b) that the resulting filer has a transition band width narrower by a factor of 3 than that of modal filter. **Figures 2** (c) to **2** (g) show a few of the frequency bands that can be achieved after performing CDM on the interpolated modal filter. **Figure 2** (c) represents the frequency response achieved after performing MCDM-II on the frequency response of **Fig. 2** (b), using M = 4. **Fig. 2** (d) represents the frequency response achieved when performing CDM-I on a modal filter interpolated by L = 2, using M = 3. **Fig. 2** (e) represents the frequency response resulted after performing both MCDM-I and CDM-I on the frequency response of **Fig. 2** (b), using M = 2. **Fig. 2** (f) represents the frequency response achieved after performing both MCDM-I and CDM-I on the frequency response of **Fig. 2** (b) using M=3. **Fig. 2** (g) represents the frequency response achieved after performing both MCDM-I and CDM-I on the frequency response of **Fig. 2** (b) using M=4. It can be noted that the number of resulting sub-bands in **Figure 2** (d) to **2** (g) are 6, 6, 9, 12 respectively, which represent the product of the decimation factor and the interpolation factor. It can be observed that the number of distinct sub-bands is increased and the flexibility of the location of center frequency is enhanced. As can be seen in **Fig. 2** (e), (f) and (g), the frequency responses obtained for CDM and MCDM are mirror images of each other. The modal filter used in this example has a normalized stop band frequency f_s = 0.132. Therefore, M values greater than $\lfloor 1/0.132 \rfloor$ = 7 will cause aliasing and cannot be used in the coefficient decimation operations. The maximum decimation factor allowed in coefficient decimation operations after performing CIM ($L = 3$) in HCDIM approach is $\lfloor 1/0.044 \rfloor$ = 22. According to formula advanced by Bellanger Eq. (3), the order of the modal filter is (457) and the over designed filter order is (524). The resulting filter after interpolation has specifications of (f_p = 0.04 and f_s = 0.044) as shown in **Fig. 2** (b). A filter of order (1374) should be used to obtain such sharp TBW, and the over designed filter order is (1441). Thus, this type of filter bank offers an inherent reduction in the complexity as can be further explained in multiplication complexity comparison in the design example.

To achieve the proposed FB for different applications, the design steps which are similar to that described in**, Ambede, et al., 2014** are used with some modifications in steps three, four and six concerned with the use of CIM in the proposed FB. Let the different communication standards be sampled at sampling frequency f_{sam}. Their bandwidths are BW_1, BW_2, . . . , BW_m, where m represents the number of standards. The TBW specifications are TBW_1 , TBW_2, . . . , TBW_m. The desired passband peak ripple specifications and stopband attenuation specifications for the channels of m standards are $\delta_{p1}, \delta_{p2}, \dots, \delta_{pm}$ and $\delta_{s1}, \delta_{s2}, \dots, \delta_{sm}$, respectively. The steps for extracting m different communication standards simultaneously using the proposed HCDIM-based FB are as follows:

Step one: All channel BW and TBW specifications should be normalized to $f_{sam}/2$.

Step two: Divide the pass-BW of each standard by 2.

Step three: Recognize the interpolation factor L required to perform CIM on the modal filter for obtaining suitable sharp transition bandwidth.

Step four: Calculate the modal filter's pass band width (BW_{mod}) as the greatest common divisor (GCD) of the pass band widths obtained in step two multiplied by L.

$$BW_{mod} = \text{GCD}\ \{\frac{BW_1}{2}, \frac{BW_2}{2}, \ . . . , \frac{BW_m}{2}\} * L \tag{6}$$

Step five: Specify the decimation factors required for performing ICDM-II operations on the modal filter or the interpolated modal filter to get low and high pass frequency responses corresponding to the channel bandwidths of the m standards. These values is identified as $D_1, D_2, \dots, D_m$.

Step six: Calculate the new TBW of each standard by dividing the TBW of each standard by its corresponding D. The TBW of the modal filter (TBW_{mod}) is the minimum of the computed values multiplied by L.

$$TBW_{mod} = \min\ \{\frac{TBW_1 * L}{D_1}, \frac{TBW_2 * L}{D_2}, . . . , \frac{TBW_m * L}{D_m}\} * L \tag{7}$$

Step seven: The set of decimation factor value used in HCDIM-I operations should be identified to obtain frequency responses which are used to extract the wanted channels in the filter bank. The value of maximum decimation factor is indicated as M_{max}.

Step eight: The stop band attenuation of the modal filter δ_{smod} is the minimum of the new SA for each standard:

$$\delta_{smod} = \min\ \{\frac{\delta_{s1}}{D_1 \times M_{1max}}, \frac{\delta_{s2}}{D_2 \times M_{2max}}, . . . ,, \frac{\delta_{sm}}{D_m \times M_{mmax}}\} \tag{8}$$

The peak pass band ripple of the modal filter (δ_p) is the minimum of the m standards.

Step nine: The modal filter order corresponding to the obtained specifications is calculated using Eq. (3). The HCDIM processes are performed on the modal filter with the specified values of interpolation and decimation factor to obtain the corresponding frequency responses. The design examples in the next section will make the above steps applicable.

4 DESIGN EXAMPLES

Two design examples are presented here to compare the proposed methodology with the previously used approaches. The modal filter and the masking filters in both design examples are designed using equiripple (Parks-McClellan algorithm) transposed direct-form FIR filter technique. The maximum error between desired frequency response and actual frequency response is minimized in equiripple technique by spreading the approximation error uniformly over each band, **Kumer, et al., 2015**. The transposed direct-form FIR filter structure is that exploits the property of filter coefficients symmetry. This type of implementation is preferred because extra shift register for input signal is not needed and extra pipeline stage for the adder of the products to achieve high throughput is not needed.

4.1 Fixed Channel Stacking Channelization

The proposed HCDIM-based FB is used in this section, to extract uniform sub-bands using a design example to compare with other FBs used which is the same design example used in the ICDM and DFTFB-FB that consists of 8-channel whose output frequency response is as shown in **Fig. 3**. The prototype filter has a frequency edges similar to those of sub-band one (sb1). The pass band frequency is $f_p = 0.1125$, and the normalized stop band frequency is $f_s =$ 0.1375. The desired pass band peak ripple is 0.1 and the desired stop band peak ripple is −40 dB. The order of the prototype filter is 160 according to Eq. (3). To obtain the desired 8-point DFTFB, the prototype filter has to be followed by an eight-point IDFT process. In the proposed HCDIM based FB, the pass band and the stop band frequencies of the modal filter are different from those of the prototype filter used in DFTFB and the modal filters used in ICDM designs. As the CIM is used in the proposed approach so the pass band and stop band edges of the modal filter are of factor L wider than that of the prototype filter and the modal filters used in the former methods designs. In order to extract the five sub-bands in **Fig. 3**, the modal filter pass band and stop band frequencies should be multiplied by L.

Let the interpolation factor L be 2, the corresponding pass band and stop band frequencies are ($f_p = 0.1125\times2= 0.225$) and ($f_s = 0.1375\times2= 0.275$) respectively. Using the same peak pass band ripple and stop band ripple used in DFTFB, CDM and ICDM of 0.1dB and -40dB, respectively. The order of the new modal filter is (80) using Eq. (3). The maximum decimation factor elaborated in the HCDIM design is 4 and the equivalent overdesigned order of the modal filter using (5) is 97. Different stages of the HCDIM-FB are shown in **Fig. 4** and the associated frequency responses of the output. **Fig. 4** (a) represents the frequency response of the modal filter. **Fig. 4** (b) represents the frequency response of the modal filter after performing CIM on the modal filter by a factor of $L = 2$. Sb1 can be extracted after masking the frequency response of **Fig. 4** (b) by a wide transition band width low order (N=4) low pass masking filter 1 (MF1) (f_p = 0.1125, f_s = 0.8625). Sb5 can be obtained by masking the same frequency response by a wide transition band width low order (N=4) high pass masking filter (MF5) (f_p =0.8875, f_s = 0.1375). **Fig. 4** (c) represents the frequency response resulted when MCDM-I is performed on the interpolated modal filter with $M = 2$, which gives the desired band sb3. Sb2 and sb4 are obtained from **Fig. 4** (d), which represents the frequency response after performing MCDM-I using $M =$ 4 on the response of **Fig. 4** (b). Sb2 and sb4 can be extracted by using wide TBW low order (N=18) masking filter 2 (MF2) (f_p = 0.3875, f_s = 0.6125) and masking filter 4 (MF4) (f_p = 0.6125, f_s = 0.3825). Thus, the five desired sub-bands can be obtained using the proposed HCDIM-FB. The maximum decimation factor elaborated in the design of the proposed HCDIM-FB is 4. Sb2 and sb4 can be separated by the same masking filters used in CDFB and ICDM designs should be used for reasonable comparison. The masking filters used to extract sb1 and sb2 are particular in the proposed FB only. In this design example, the order of the modal filter can be further reduced if a higher L is used. For example, let $L = 4$, the frequency specification of the modal filter will be (f_p =0.1125×4= 0.45) and (f_s =0.1375×4= 0.55). The frequency response of such filter is shown in **Fig. 5** (a). **Fig. 5** (b) represents the frequency response of the modal filter after performing CIM, using $L = 4$, which is similar to the frequency response obtained after performing CDM-I on the interpolated modal filter using M=4. The three low-order masking filters (as shown in **Fig. 5** (b)) are employed to extract the three sub-bands, MF1 (f_p = 0.1375, f_s = 0.625), MF3 ($f_{p1} = 0.3625, f_{s1} = 0.0.1375, f_{p2} = 0.6375$ and $f_{s2} =$ 0.8625) and MF5 (f_p = 0.8625, f_s = 0.6375). **Fig. 5** (c) represents the frequency response obtained after performing MCDM-I, using M=4 on the interpolated modal filter which represents sb2 and sb4. Two masking filters (as shown in **Fig. 5** (c)) similar to those used in the previous example are required to extract sb2 and sb4. It can be noted that sb1, sb3 and sb5 can be extracted using appropriate masking filters after performing CIM, using $L = 4$ without using the decimation method. But this means that upsampling operation is not followed by downsampling operation which leads to increasing sampling frequency. The modal filter order used in this example is 39 according to Eq. (3). The maximum decimation factor elaborated in the HCDIM design in this example is 4 and the equivalent overdesigned order of the modal filter using Eq. (5) is 41. In contrast to just 1 maximum decimation factor involved in modal filter of **Fig. 5** (a), the involved maximum decimation factor is 7, using the HCDIM design technique. It can be noted that using a higher value of L ($L = 4$), additional masking filter to extract sb3 (MF3), and higher order masking filters to extract sb1 and sb5 are required. Thus, using larger values of L, narrower TBWs masking filters need to be designed, which may increase the complexity and lead to inefficient implementation.

The complexity of the proposed FB (which depends on the number of multiplication operations involved) and various FBs designed to serve uniform channelization, is summarized in Tab. 1. The length of N order FIR filter (represented as l) can be calculated as $l = N + 1$, **Proakis, and Manolakis, 2007**. The total multiplications number in the DFTFB is the sum of the prototype filter length (l_p), and the multiplications number needed for an 8-point fast Fourier transform (FFT) computation ($Slog_2S$ multiplications for S-point FFT, **Proakis, and Manolakis, 2007**, used in S-channel DFTFB, **Vaidyanathan, 1990**). The modal filter and masking filters used in both these FBs are implemented with the transposed direct-form FIR filter structure, exploiting the coefficients symmetry property. Let the modal filter and masking filter lengths, be denoted as l_{Mod} and l_{Mas}, respectively. Using the transposed direct-form FIR filter structure, the number of multiplications required to modal filter and masking filter implementation are ($l_{Mod}/2$) and ($l_{Mas}/2$), respectively, **Vaidyanathan, 1990**. The complexities of multiplication ICDM and HCDIM-based FB are lower than that of the DFTFB because of using transposed direct-form FIR filter structure for implementation of the modal filter. It can be noted from the design example described in this section that as the proposed FB employs both CDM and CIM operations, the modal filter essential in the proposed HCDIM-based FB has a lower order than the modal filter essential in the ICDM-based FB. From **Tab. 1**, it can be noted that the proposed HCDIM based FB, using $L = 2$, offers a reduction in multiplication complexity of 59.46% over DFTFB, and 31.19% over ICDM based FB, while the reduction in complexity using $L = 4$ is about 63.24% over DFTFB and 37.61% over ICDM based FB. It can be observed that in**, Ambede, et al., 2015**, the maximum value of the decimation factor involved is less than that used in the proposed HCDIM-FB due to the use of complementary delays approach to perform sb2 and sb4. If this approach is applied in our proposed design method, the corresponding over designed modal filter order will be reduced and the percentage reduction in multiplication will be enhanced. Thus, for the same modal filter specifications, the worst SA value observed in the ICDM based design is half of that detected in the HCDIM based approach having the same TBW.

4.2 Multi-Standard Channelization

In this section, the capability of the proposed HCDIM-based FB to perform multi-standard channelization is confirmed and demonstrated. The frequency response of an input spectrum is shown in **Fig. 6**. In this figure, four Bluetooth (BT) channels, one Zigbee channel, and one wideband code division multiple access (WCDMA) channel are simultaneously existing at different locations in the wideband frequency range.

The channel bandwidths of BT, Zigbee and WCDMA standards are 1, 4, and 5 MHz, respectively, and their corresponding transition bandwidth specifications are chosen as 50, 200, and 500 KHz, respectively. The sampling frequency is chosen as 40 MHz. The desired pass band and stop band peak ripple specifications for Zigbee and WCDMA channels are 0.1 and −40 dB, and for BT channels 0.1 and −55 dB, respectively. According to design steps described in section 5, an HCDIM-based FB can be used to obtain the different channels shown in **Fig 6**. According to step one, the normalized channel BWs of BT, Zigbee, and WCDMA are computed to be 0.05, 0.2, and 0.25, respectively. Let the interpolation factor (L) be 2. The GCD of {0.05, 0.2, 0.25} is obtained as 0.05 which represents the pass-BW of the modal filter according to Eq. (6) in step four. The decimation factor values required to achieve the low pass and high pass frequency responses with their pass bandwidths corresponding to the channel BWs of the three standards are $D1 = 1$, $D2 = 2$, and $D3 = 5$, according to step 5. The TBW of modal filter is then calculated to be 0.005 using Eq. (7) in step six. The pass band and stop band edge frequency specifications of the modal filter are chosen as $f_p = 0.045$ and $f_s = 0.05$, respectively. The three sets of decimation factor value to be used in HCDIM operations to obtain the sub-bands corresponding to the three standards are {5}, {2}, and {1}, respectively, according to step seven. The pass band and stop band attenuation specifications are obtained according to step eight as 0.1 dB and -60 dB, respectively. The modal filter is designed with the calculated specifications values and the equivalent filter order is 1396 using Eq. (3). Suitable HCDIM operations are performed on the designed modal filter using the recognized values of decimation factor. The frequency responses of each stage in the HCDIM-based FB is shown in **Fig. 7**. **Fig. 7** (a) represents the frequency response of the modal filter after performing CIM, using $L = 2$. Then CDM-I is performed on the interpolated modal filter using $M = 5$ to obtain the frequency sub-band corresponding to the BT2 as shown in **Fig. 7** (b). To extract channel BT2 from the multiband frequency response obtained, a masking filter (MF2) of order 39, designed according to Eq. (3). To extract BT1, BT3, and BT4 channels, perform MCDM-I using $M = 5$ on the modal filter, as shown in **Fig. 7** (c). Three masking filters of the low order 39 (MF1, MF3 and MF4) are used to extract the three channels. The WCDMA channel is extracted using the frequency response obtained after performing MCDM-II on the modal filter using $M = 5$ as shown in **Fig. 7** (d). To obtain Zigbee sub-band channel, perform CDM-II on the modal filter using $M = 2$ as shown in **Fig. 7** (e), then perform MCDM-I using $M = 2$ on the resulting filter to get Zigbee sub-band as shown in **Fig. 7** (f). Hence, all the BT, Zigbee and WCDMA standards channels which are simultaneously existing in the input signal can be extracted using the proposed FB. A block diagram that summarizes various stages in HCDIM-based FB is shown in **Fig. 8**.

Since the bandwidths of Zigbee and WCDMA standards are not integer multiples of each other, the multiple FBs need to be designed in both the MPRB and the CMFB approaches to extract the different frequency channels in **Fig. 6, Ambede, et al., 2014**.

It can be noted that if ICDM-FB is used to extract the different frequency channels in **Fig. 6**, the channel BT2 can be obtained after performing CDM-I using $M = 10$. The rest BT channels can be obtained after performing MCDM-I using $M = 10$. All BT channels can be extracted by means of the proposed HCDIM-based FB using CDM-I and MCDM-I at $M = 5$, which is half the M value required in the ICDM-FB case. In this design example, the order of the modal filter can be further reduced if a higher L is used. For example, let $L = 5$, the frequency specification of the modal filter will be ($f_p = 0.0225\times5 = 0.1125$) and ($f_s = 0.025\times5 = 0.125$). The frequency response of such filter is shown in **Fig. 9** (a). **Fig. 9** (b) represents the frequency response of the modal filter after performing CIM, using $L = 5$. To extract BT channels, perform CDM-I, using M=2 to obtain a frequency response similar to that of **Fig. 7** (b), then perform MCDM-I, using M=2, to obtain a frequency response similar to that of **Fig. 7** (c). To extract the Zigbee channel, an MCDM-I on the modal filter, using M=2, is performed. Perform CIM on the modal filter, using L =2 and then perform CDM-II, using M=5 to extract WCDMA channel. According to design steps in Section 5, the stop band attenuation is -60, and the corresponding modal filter order is 531.

It can be noted that while the order of a modal filter is 1396, 531, using $L = 2$ and 5 respectively, in the proposed HCDIM-based FB (where the maximum necessary value of $M = 5$ using $L = 2,5$ respectively, and TBW of modal filter is wider by a factor of L). The modal filter order required in the ICDM-based FB is 2928 (wherein the maximum required value of $M = 10$). A summary of the number of multiplications required to implement the HCDIM-FB and ICDM that are designed for obtaining the different standards of Fig. 7, is presented in Tab. 2. It can be noted that proposed HCDIM-based FB offers a multiplication complexity reduction (using $L = 2$) of 44.40% over ICDM. The multiplication complexity reduction (using $L = 5$) is about 77.61% over ICDM. The proposed HCDIM-based FB achieves a lower complexity than the ICDM because of using both coefficient interpolation and decimation methods and the masking filters involved, that result in modal filter of lower order.

It can be noted from the frequency responses obtained using HCDIM technique, that the increase in the value of M deteriorates the stop band attenuation (SA) of the filters obtained after performing coefficient decimation method. This is an intrinsic disadvantage of CDM and is existing in HCDIM too. The mathematical expression of the deterioration in SA can be given by Eq. (4). The SA deterioration problem is overcome by overdesigning the modal filter given in Eq. (5). The design steps described in section 5 have taken into account the SA deterioration problem too. This problem occurs when CDM is used only, so the use of CIM in the proposed FB has no action on the SA deterioration. The maximum required value of M is 10 using ICDM to extract channels of **Fig. 7** and the corresponding SA is -65dB. Using the proposed HCDIM approach to extract channels of **Fig. 6**, a maximum required value of M is 5, and the corresponding value of SA is -60dB according to Eq. (8). It is clear that the deterioration in SA means the larger filter order and the decreasing of this deterioration is an advantage. It can be noted in **Fig. 7** (f) that there seems to be a deterioration in the pass band magnitude in sb3 since the resulting response obtained after performing CDM-II, using M=2 is scaled by 2, then scaled by 2 after performing MCDM-I, using M=2. The resulting response needs to be scaled by 2 again to have a 0 dB magnitude. Hence, in all HCDIM operations as well as ICDM operations, the pass band ripple does not alter and remains constant (0.1dB) after scaling the resulting response by the appropriate M.

It can be noted that the resolution of center frequency in the resultant multiband frequency responses in ICDM operations, is π/M. While possible center frequency sub-bands locations of $2\pi/M$ are achievable in the M-channel DFTFB. Using HCDIM based FB, the resolution of center frequency in the resultant multiband frequency responses is $\pi/(M \times L)$. Reconfigurable CD and FRM methods offer only one degree of freedom (M and L, respectively) to change the location of center frequency and BWs of channels, whereas the HCDIM based FB offers two degrees of freedom, M and L, which can be changed individually. Hence, the proposed method considerably improves the flexibility of the filter architecture to adapt to the channel spacing of different communication standards.

5 CONCLUSIONS

Reconfigurable filter bank architecture for dynamic channel adaptation for a CR terminal, based on decimation, interpolation and frequency response masking is proposed in this paper, and termed as HCDIM based FB and used for non-uniform and uniform channelization. The proposed architecture is a flexible alternative to other kinds of FBs since the resulting filter is an $(M \times L)$ sub-bands and has a center frequency resolution of $\pi/(M \times L)$ in the resultant multiband frequency responses. Also, if the same modal filter is used to obtain ICDM based channel filter, the worst case transition bandwidth and stop band attenuation values detected in the HCDIM based design are enhanced. Complexity analysis regarding the design examples shown clearly specify that the proposed architecture offers a better filter length saving compared to that of the other methods. The reduction in the number of multiplications is about 31.19% over ICDM based FB in uniform channelization and 50.77% in non-uniform channelization. The complexity of the modal filter can be further reduced if a higher interpolation factor is used. Thus, the proposed HCDIM-based FB is highly appropriate for use in applications of resource constrained such as portable CR handsets because of its significant advantages in terms of flexibility, complexity, and resource utilization over the other FBs.

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ω Frequency, rad/sec..

δ_p Passband ripple, dB.

δ_s Stopband attenuation, dB.

f_p Normalized passband frequency.

f_S Normalized stopband frequency.

Table 1. Comparison of multiplication complexity: design example of uniform channelization

	DFTFB	ICDM	Proposed HCDIM based FB	
			$L = 2$	$L = 4$
Prototype/modal filter length (l_p/l_{Mod})	l_p=161	l_{Mod}= 177	l_{Mod}=97	l_{Mod}=42
Masking filter length (l_{Mas})	–	19∗2=38	19*2+5*2 =48	19∗2+17∗2+17= 89
No. of multiplications = ([$\lceil l_{Mod}/2\rceil$]+[$\lceil l_{Mas}/2\rceil$])	161	{$\lceil 177/2\rceil$+ (2∗ $\lceil 19/2\rceil$)} =109	{$\lceil 97/2\rceil$ +(2*$\lceil 19/2\rceil$+2*$\lceil 5/2\rceil$)}=75	$\{\lceil 42/2\rceil + (2 * \left\lceil\frac{19}{2}\right\rceil + 3 * \left\lceil\frac{17}{2}\right\rceil)\} = 68$

No. of multiplications for S-point DFT (S=8)	$Slog_2$S =8$log_2$8 =24	_	_	-
Total no. of multiplications	185	109	75	68

Table 2. Multiplication complexity comparison: non-uniform channelization design example

	ICDM	Proposed HCDIM based FB	
		$L = 2$	$L = 5$
Modal filter length (l_{Mod})	2929	1397	532
Masking filter length (l_{Mas})	40*4=160	40*4=160	40*4=160
No. of multiplications =($\lceil l_{Mod}/2\rceil$+$\lceil l_{Mas}/2\rceil$)	{$\lceil 2929/2\rceil$ +(4*$\lceil 40/2\rceil$)} =1545	{$\lceil 1397/2\rceil$ +(4*$\lceil 40/2\rceil$)} =859	{$\lceil 532/2\rceil$ +(4*$\lceil 40/2\rceil$)} =346
Total no. of multiplications	1545	859	346

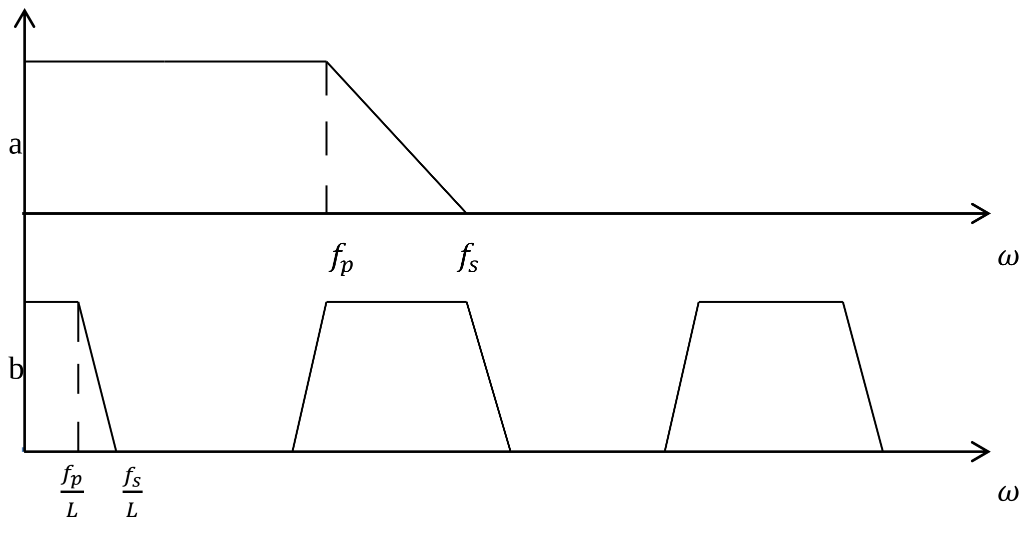


Figure 1 Coefficient interpolation method by a factor of L.

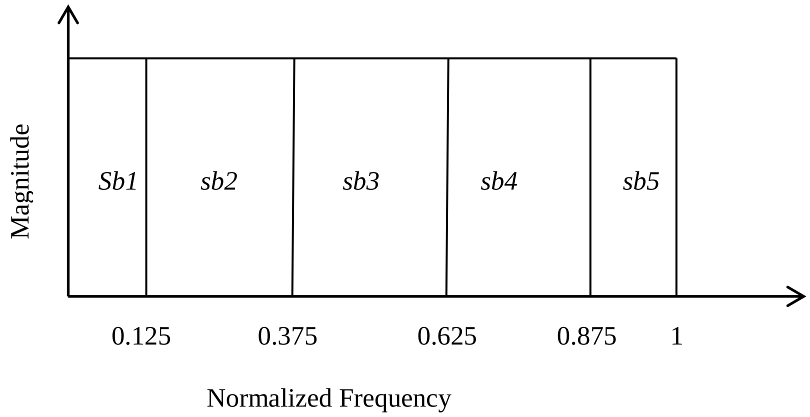
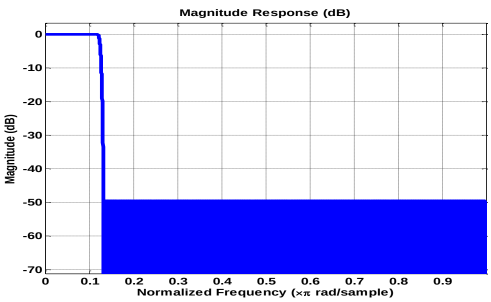
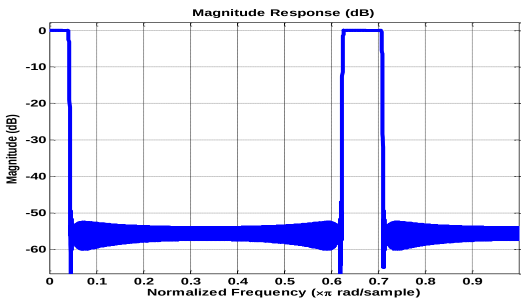


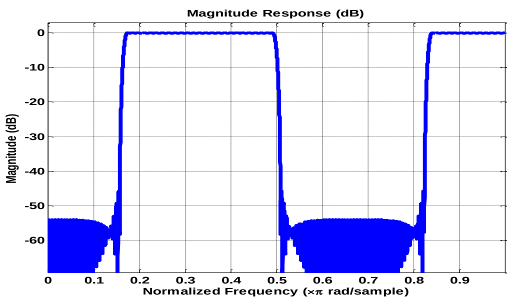
Figure 3 Frequency response of uniform design example



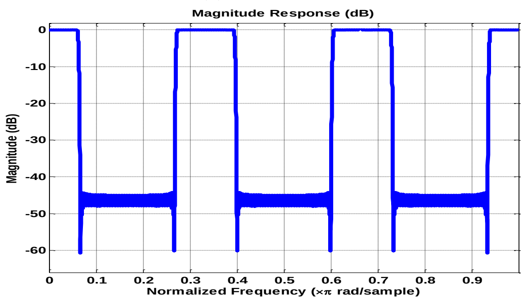
Modal filter



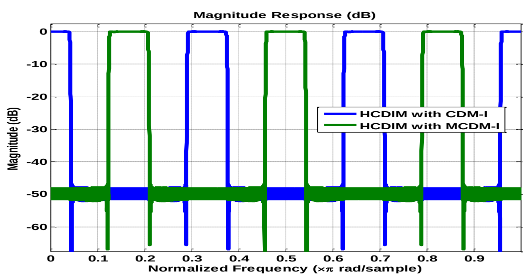
(b) CIM, L=3



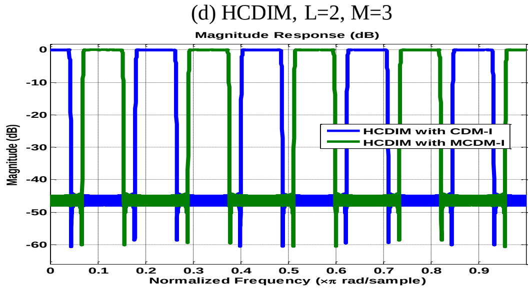
(c) HCDIM, L=3, M=4



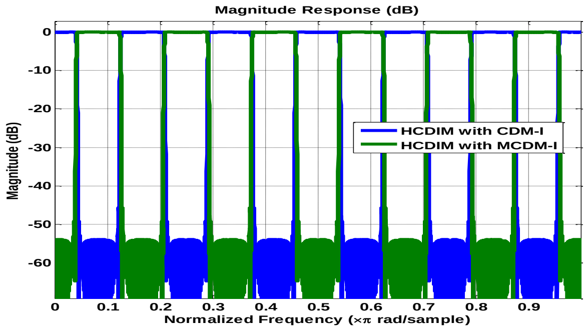
(d) HCDIM, L=2, M=3



(e)HCDIM, L=3, M=2.

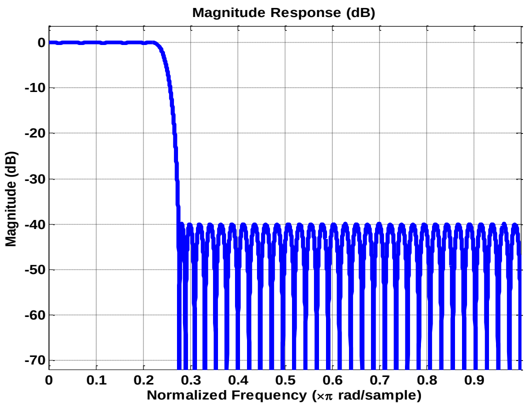


(f) HCDIM, L=3, M=3

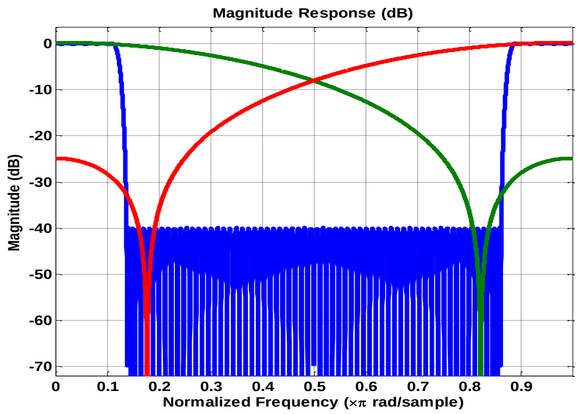


(g) HCDIM, L=3, M=4

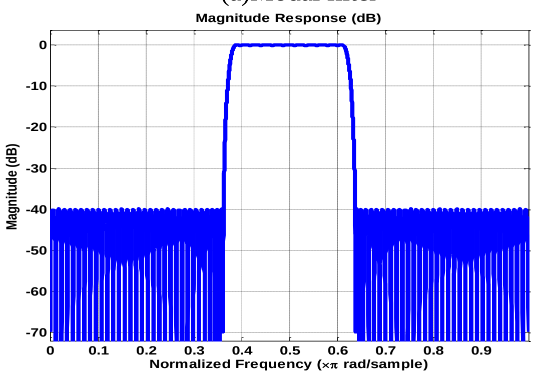
Figure 2 HCDIM operations on modal filter having f_p = 0.12 and f_s = 0.132



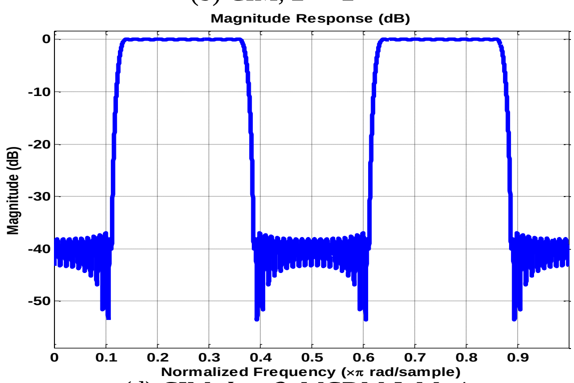
(a)Modal filter



(b) CIM, $L = 2$

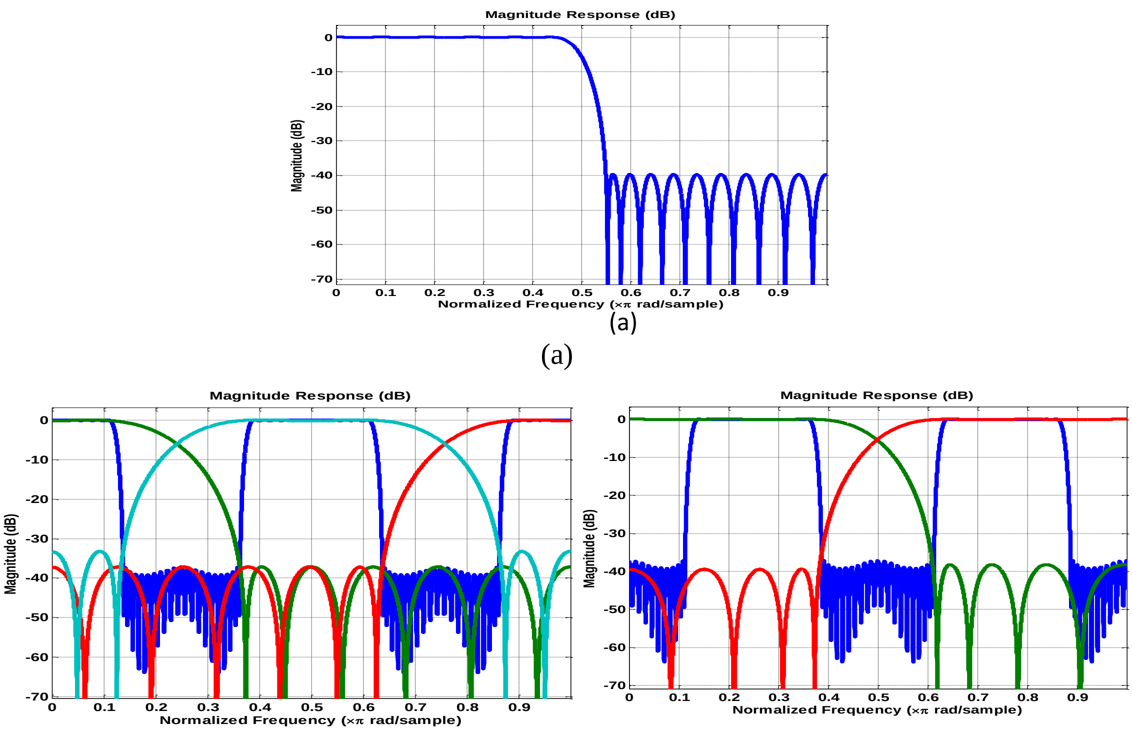


(c) CIM, $L = 2$, MCDM-I, M=2



(d) CIM, $L = 2$, MCDM-I, M=4

Figure 4 HCDIM-based FB for uniform channelization



(a)

(b) (c)

Figure 5 (a) Frequency response of modal filter (b) Frequency response of modal filter after performing CIM, using L=4 and CDM-I, using M=4 and appropriate masking filters (c) Frequency response of modal filter after performing CIM, using L=4 and MCDM-I, using M=4 and appropriate masking filters .

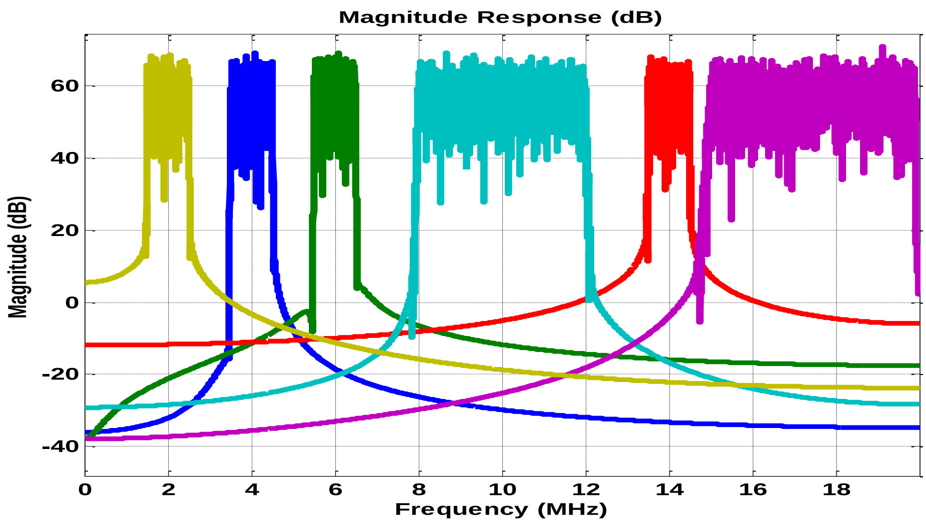
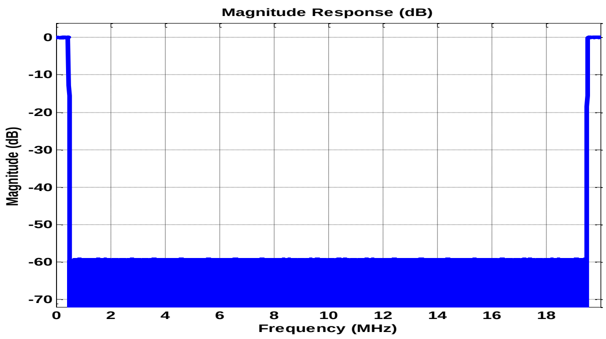
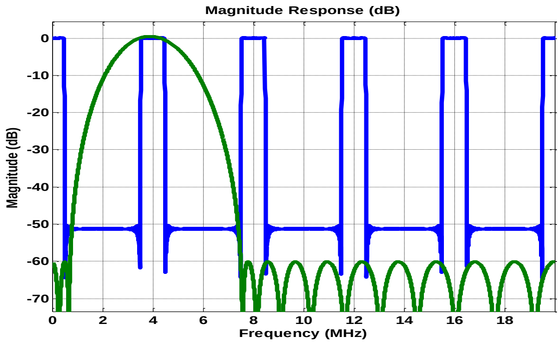


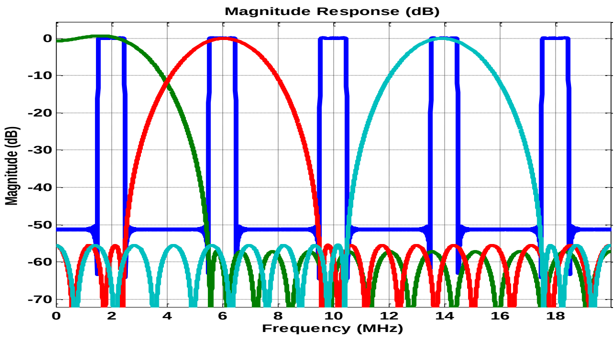
Figure 6 Design example of multi-standard channelization



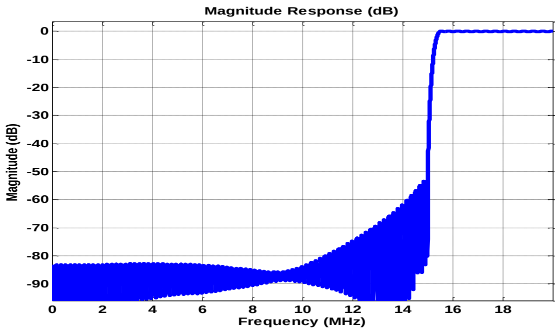
(a) CIM, $L = 2$



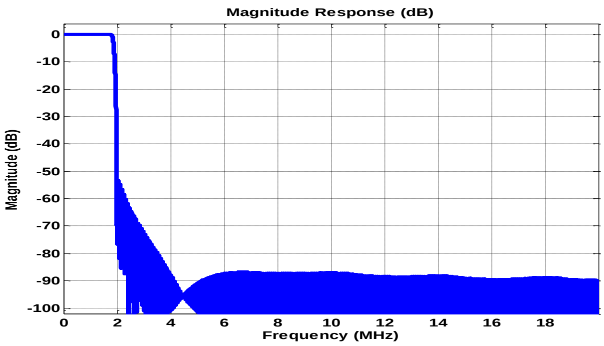
(b) CIM, $L = 2$, CDM-I, $M = 5$



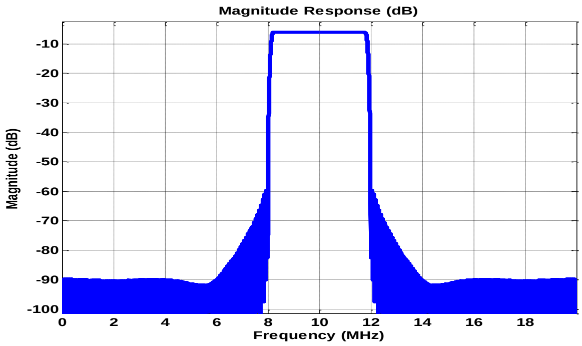
(c) CIM, $L = 2$, MCDM-I, $M = 5$



(d) MCDM-II, $M = 5$



(e) CDM-II, $M = 2$



(f) CDM-II, $M = 2$, MCDM-I, $M = 2$

Figure 7 HCDIM based FB for non-uniform channelization using $L = 2$

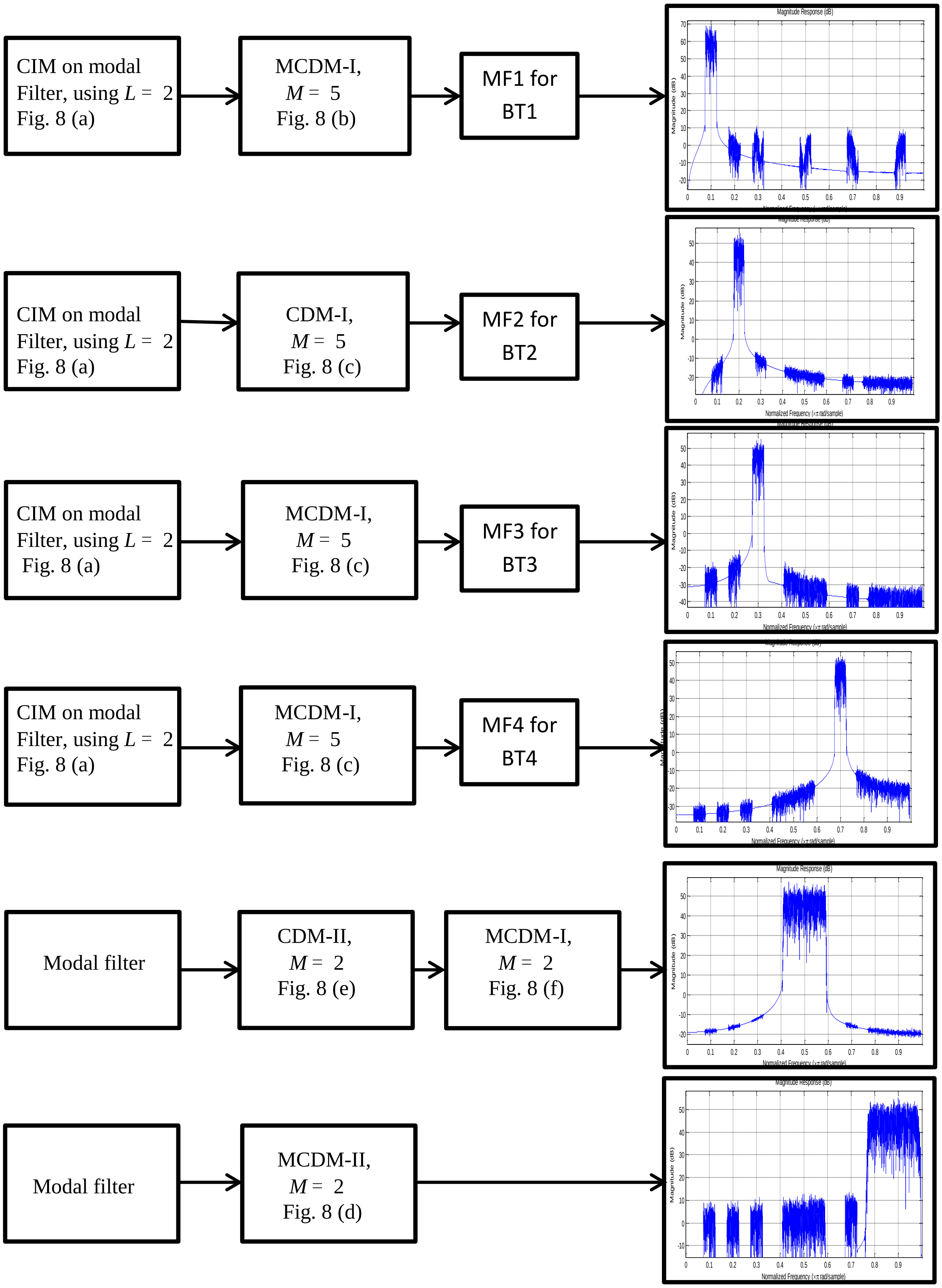
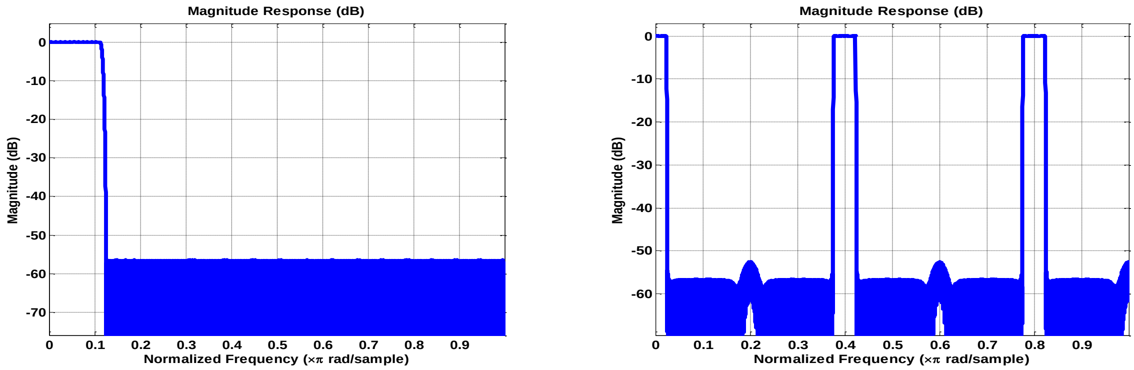


Figure 8 Block diagram of HCDIM: non-uniform channelization (using $L = 2$)



(a) (b)

Figure. 9 (a) Frequency response of modal filter (b) Frequency response of modal filter after performing CIM, using $L = 5$