# A LOW COMPLEXITY RECONFIGURABLE CHANNEL FILTER BASED ON DECIMATION, INTERPOLATION AND FREQUENCY RESPONSE MASKING

*Sumedh Dhabu, Smitha K. G. and A. P. Vinod*

School of Computer Engineering, Nanyang Technological University,
Nanyang Avenue, Singapore 630798.
Email: sumedh1@e.ntu.edu.sg

**ABSTRACT**

The channel filters in multi-standard wireless communication receivers must be capable of extracting radio channels (frequency bands) of distinct bandwidths located at different center frequencies. In this paper we propose a method to realize low complexity reconfigurable channel filters based on decimation, interpolation and masking techniques. We show that our method provides approximately twice the number of channel filter frequency responses compared to the existing reconfigurable channel filter realization method. The resolution of center frequency locations of our channel filter responses is also twice that of the existing method. The enhancement in center frequency resolution and the increased number of channel filter frequency responses are obtained without any hardware overhead.

***Index Terms***— Reconfigurable FIR filter, low complexity channel filter

## 1. INTRODUCTION

The channel filter in the wireless communication receiver extracts the radio channel (frequency band) of interest from the wideband input signal. In a multi-standard communication receiver, the channel filter should be able to dynamically adapt its frequency response to the channel bandwidth and location (center frequency) corresponding to the current standard of operation [1].

Channel filter is generally a finite impulse response (FIR) filter because of its stability and linear phase. Different reconfigurable FIR filter realization approaches have been proposed in the literature. One approach is to use separate coefficient memories for different frequency responses as in case of intermediate frequency filters in [2]. The filter structure which has the maximum number of coefficients (taps) corresponding to the communication standard that has the most stringent adjacent channel attenuation specification amongst the various standards to be supported, is implemented. The filter coefficients of each standard are stored in separate coefficient memory. The coefficients are loaded from appropriate memory according to the desired frequency response to be obtained. In [3], canonic signed digit (CSD) based reconfigurable filter architecture was introduced in which a digit processing unit (DPU) is used as basic building block for filter structure. Each DPU consists of a register to hold control signals for that DPU, multiplexers, multiplier and shifter. Filter is implemented by cascading several DPUs and summing up all the outputs of DPUs. The drawback of this method is that it requires huge amount of hardware resources (multiplexers, registers and shifters). Programmable CSD based reconfigurable filter structure for high speed applications was introduced in [4]. The main aim of the work in [4] is to realize a high speed filter. This method requires more area and consumes more power compared to other CSD based methods. In [5], the reconfigurable multiplier block based filter architecture utilizing graph dependent (GD) algorithms was proposed. This architecture is suitable for slow speed applications as GD algorithms are sequential. The main drawback of all these methods is that the filter structure corresponding to the communication standard with most stringent specifications needs to be implemented as the basic filter and then other channel filter responses are obtained from scaled down version of the full-length basic filter. Also the reconfiguration time is high.

Another set of approaches for reconfigurable filter realization is based on fixed-coefficient filter realization [6] - [9]. The methods in [7] - [9] used a fixed-coefficient prototype filter (initial filter) to obtain various frequency responses by employing frequency response masking (FRM) and/or coefficient decimation technique. The FRM technique [6] was used to obtain sharp cutoff filters with low complexity. In FRM filters, the interpolated filter's bandwidth, $BW_{int}$, is given by $BW_{initial}$ x $1/M$, where $BW_{initial}$ is prototype filter's bandwidth and $M$ is the interpolation factor. In [8], reconfigurable filter is realized by performing coefficient decimation followed by interpolation and then extraction of the desired band by masking. Coefficient decimation results in increasing the passband width of the prototype filter by the decimation factor of $D$. The bandwidth of the filter after coefficient decimation, $BW_{dec}$, is given by $D$ x $BW_{initial}$, where $BW_{initial}$ is prototype filter's passband width. Interpolation produces multiband response and the desired band is extracted using appropriately designed masking filter. The bandwidth, $BW_{final}$, of the filter after performing coefficient decimation and interpolation is given by $BW_{initial}$ x $D/M$, where $BW_{initial}$ is the bandwidth of prototype filter, $D$ is coefficient decimation factor and $M$ is interpolation factor. In [9], this idea was extended to realize reconfigurable filter banks.

We observed following limitations in the reconfigurable filter realizations methods [7] – [9]. In [7], the bandwidth variations have only one degree of freedom, $M$. In [8], even though it has two degrees of freedom $D$ and $M$, a reconfigurable masking filter is used for extracting the desired response. Hence the masking filter needs to be reconfigured before new band can be extracted which increases the band extraction time. This reconfigurable filter approach becomes area inefficient when the number of interpolation and decimation values used increases. In [9], a generalized masking filter design approach is adopted but still it is fixed for particular interpolation value. The entire masking filter design needs to be changed if a new interpolation value is chosen.

In this paper we propose a filter realization method based on the modification of the method in [8]. Our modified filter has higher center frequency resolution and can produce more number of distinct channel filter responses than the filter designed using the method in [8], without any hardware overhead. A set of fixed-coefficient masking filters is used to extract all the possible bands generated by varying $M$ and $D$ values using the modified approach.

The rest of the paper is organized as follows. Section 2 presents the modified filter realization method. Section 3 presents the masking filter design. Section 4 presents the results and comparisons. Finally section 5 provides our conclusions.

## 2. PROPOSED METHOD

The block diagram of our modified filter realization method is shown in Fig 1. The step by step approach is presented as follows.

1. First, a lowpass filter (termed modal filter) is designed.
2. The coefficient decimation operation is performed on this modal filter to increase the bandwidth by factor of $D$. In the coefficient decimation technique, every $D^{th}$ coefficient of the modal filter is retained; discarding the coefficients in between.
3. In order to enhance the flexibility of the receiver for extracting different sets of bands (other than the bands obtained from the decimated and interpolated modal filter), inverse operation is performed on decimated modal filters. An inverse filter of the decimated modal filter is obtained when the sign of every alternate coefficient from the retained set of coefficients is reversed. This is done by reconfiguring the adder block corresponding to every alternate coefficient from the retained set of coefficients to work as subtractor. This can be done with the help of 'select' signal in the adder/subtractor block.
4. The resulting decimated/inverse decimated filters are further interpolated using different interpolation values. Interpolation produces multiband response and also reduces the bandwidth by factor of $M$. The bandwidth of these bands is $D/M$ x $BW_{initial}$, where $BW_{initial}$ is the modal filter's bandwidth. Interpolation of decimated/inverse decimated filter generates multiband response with bands centered at the odd multiples of $\pi/M$. Hence the center frequency resolution that can be achieved using our modified method ($\pi/M$) is twice that of the method ($2\pi/M$) in [8]. The number of distinct frequency bands obtained using our method is approximately twice that of the method in [8].
5. Individual channel filter responses are extracted using a set of fixed-coefficient masking filters.

Our method extends the design in [8] to include the higher interpolation factors, adds more flexibility by incorporating inverse operation as well as uses generalized set of fixed-coefficient masking filters for extracting the channel filter responses. The proposed method can be illustrated using following example. Consider a modal filter with the passband and stopband edges, $f_p = 0.07$ and $f_s = 0.1$, respectively. Coefficient decimation by factor of $D$ gives the decimated filter having with bandwidth $D$ x $BW_{initial}$. Interpolation of decimated modal filter by factor of $M$ produces total ($M$+1) bands having passband and stopband edges at ($2\pi K/M \pm D$ x $f_p/M$) and ($2\pi K/M \pm D$ x $f_s/M$) respectively, where $K$ = 0 to $M$. Note that out of all these bands, only ($M$+1 – ceil($M$/2)) bands are in the real part of frequency spectrum. We consider the interpolation factors from 3 to 8. For $M$ = 3 to 8, we get total 21 bands (computed using $\sum(M+1\text{-ceil}(M/2))$) by interpolating the decimated modal filter as shown in Fig. 2. Note that in this case $D$ = 1. Another set of 18 distinct bands (computed using $\sum\text{ceil}(M/2)$) is obtained by interpolating its inverse filter as shown in Fig. 3. Thus a total of 39 bands are obtained from the decimated modal filter and its inverse modal filter. We perform the decimation prior to the inverse operation and interpolation and hence we get such 39 bands for each of the

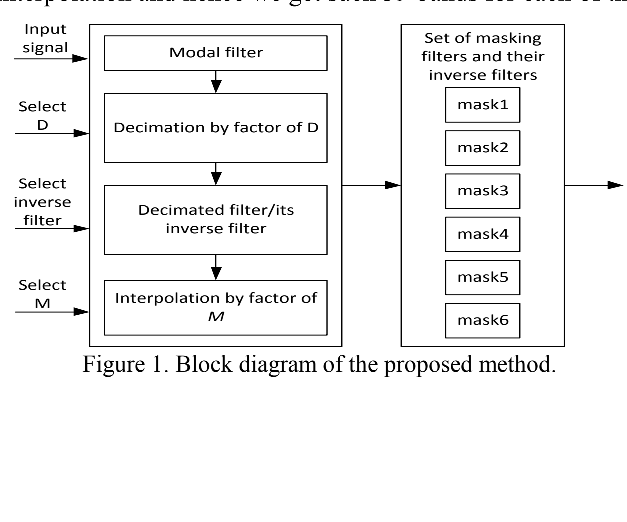


Figure 1. Block diagram of the proposed method.

decimated versions of modal filter. For $D$ = 1 to 4, we get 4x39 i.e. 156 bands. The interpolated responses of decimated modal filter and its inverse filter for $D$ = 4 are shown in Fig. 4 and 5 respectively. Note that only the real part of the frequency spectrum is shown in the figures in this paper. Also all the frequency values are normalized with respect to half the sampling frequency.

Using appropriate frequency response masking filters, each of these bands can be extracted and used as a channel filter. The generalized masking filter design is presented in next section.

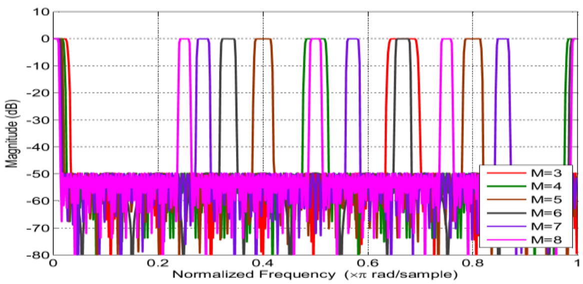


Figure 2. Interpolated modal filter ($D$ = 1) response for $M$ = 3 to 8.

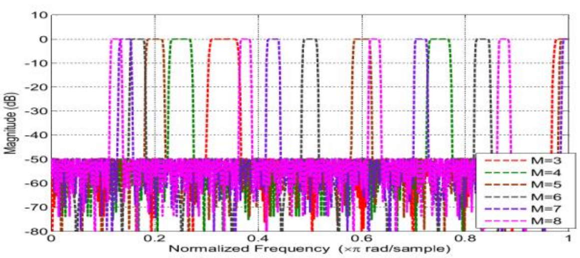


Figure 3. Interpolated inverse modal filter ($D$ = 1) response for $M$ = 3 to 8.

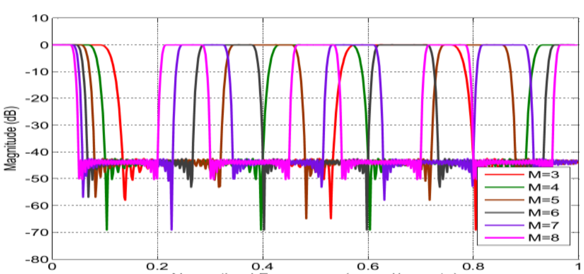


Figure 4. Interpolated responses of decimated modal filter ($D$ = 4) for $M$ = 3 to 8.

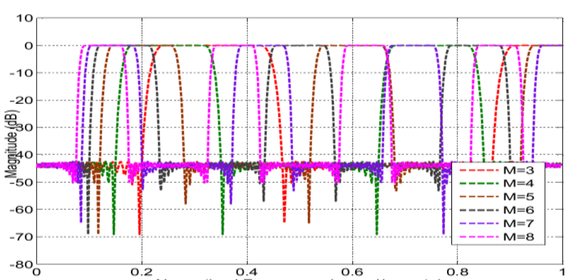


Figure 5. Interpolated responses of inverse of decimated modal filter ($D$ = 4) for $M$ = 3 to 8.

## 3. MASKING FILTER DESIGN

Masking filters are required in order to separate the bands obtained by interpolation. In the conventional FRM approach [6] one masking filter is used per band. This increases the hardware complexity when number of bands to be extracted is large. In [8], a reconfigurable masking filter is employed whose frequency response can be changed according to the band to be extracted. This method also increases the hardware complexity as well as extraction time because the masking filter needs to be reconfigured before new band can be extracted. In [9], a generalized masking filter design is used. But this design is specific to the chosen interpolation value and entire design needs to be changed if another interpolation factor is chosen.

Masking filter design is based on the stopband frequencies of the bands to be extracted and the bands to be rejected. The passband of a masking filter should fully cover all the bands to be extracted. Hence the stopband frequencies of the bands to be extracted determine the passband edges of the masking filter. The masking filter is designed in such a way that only one band of the multiband interpolated response is to be extracted and rest of the bands are to be rejected. Hence the stopband edges of the bands to be rejected determine the stopband edges of the masking filter. The stopband edges of the bands produced by interpolation depend upon the stopband edges of decimated modal filter as $2\pi K/M \pm D \times f_s/M$. As the bands produced for $D < D_{max}$ lie inside the bands produced for $D_{max}$, the masking filters are designed considering the stopband edges of the bands corresponding to maximum decimation factor $D_{max}$. The number of masking filters required and their complexity increases with increase in $D_{max}$. In our case we limit the value of $D_{max}$ to 4 to limit the number of masking filters and their complexity. The proposed method can be easily extended for $D_{max} > 4$.

We present the generalized masking filter design so that a set of fixed-coefficient masking filters can extract all the bands (for $M$ = 3 to 8 and $D$ = 1 to 4) which can be generated using our modified method. We require total 11 masking filters for extracting all the 156 bands. Out of these 11 masking filters, only 6 filters need to be actually implemented and the frequency responses of remaining 5 filters are obtained using the inverse operation on the implemented masking filters. These 11 masking filters are shown in Fig. 6 and are referred to as 'mask1' to 'mask11'. The responses of 'mask7' to 'mask11' are obtained as a result of inverse operation on 'mask1' to 'mask5' respectively and are shown with line style 'dash' in Fig. 6.

The masking filter design procedure is explained using the example of masking filter referred to as 'mask3'. The other masking filters are designed similarly. Fig. 7 shows the masking filter, the bands to be extracted and the nearest bands to be rejected by this masking filter. Note that the term 'original filter' refers to decimated modal filter ($D$ = 4)

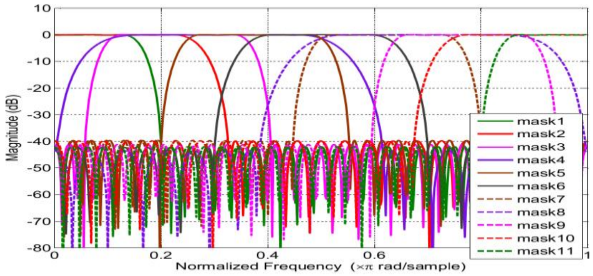


Figure 6. Masking filters.

and the term 'inverse filter' refers to inverse filter of decimated modal filter. '$f_{so}$' refers to stopband edge of original filter. The design procedure of the masking filter is as follows.

1. The desired bands to be extracted using the masking filter are decided initially. As shown in Fig. 7, 'mask3' extracts the second set of bands generated by $M$ = 7 and 8 for original filter and first set band generated by $M$ = 5 for inverse filter.
2. The passband of masking filter should fully cover all the bands. Hence the passband of 'mask3' should start at the rising stopband edge of leftmost band to be extracted and end with the falling stopband edge of rightmost band to be extracted.
3. Hence the rising passband edge ($f_{p1}$) of 'mask3' is equal to the rising stopband edge of band produced by $M$ = 5 for original filter (i.e. 1/5 – $f_{so}$/5) and the falling passband edge ($f_{p2}$) of 'mask3' is equal to falling stopband edge of band produced by $M$ = 7 on inverse filter (i.e. 2/7 + $f_s$/7).
4. The adjacent band produced along with any of the bands to be extracted needs to be rejected by this masking filter. In this case, the adjacent band produced for $M$ = 5 on inverse filter, the band produced for $M$ = 8 on original filter (shown in Fig. 7) and the band produced for $M$ = 7 on original filter (shown in Fig. 7) are to be rejected by 'mask3'.
5. Out of all these bands, the nearest band on the frequency axis decides the stopband edge of the masking filter. In this case, the lowpass band produced for $M$ = 7 on original filter determines the stopband edges of 'mask3' as shown in Fig. 7.
6. Hence the rising stopband edge of 'mask3' is $f_{so}$/7 and falling stopband edge of 'mask3' is $f_{p2} + f_{p1} - f_{so}/7$.

## 4. RESULTS AND COMPARISON

The modal filter in our design example has $f_p$ = 0.07 and $f_s$ = 0.1, stopband attenuation = -48dB, order = 170. The modal filter is overdesigned to get the final stopband attenuation of -40dB after decimation by 4, taking into account the stopband attenuation deterioration, which is an inherent problem of coefficient decimation technique. For $D$ = 1 to 4, we get 4 decimated filters from the modal filter. Inverse operation on each of the decimated version of modal filter gives another set of 4 filters. Interpolation by $M$ = 3 to 8 on

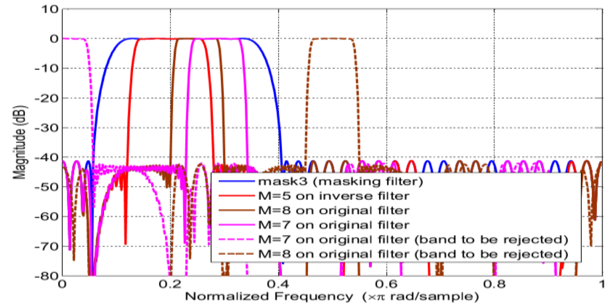


Figure 7. Frequency response of masking filter 'mask3', the bands to be extracted using 'mask3' and the band to be rejected by 'mask3'.

four decimated versions and their inverse filters results in 4 x [∑($M$+1-ceil($M$/2)) +∑ceil($M$/2)] = 156 bands. Hence using our modified method, we can obtain total 164 distinct channel filter responses (4 decimated filters + 4 inverse filters + 156 bands). The method in [8] can generate 22 channel filter responses (1 decimated filter response + 21 bands) for one decimation factor. Hence the total number of channel filter responses obtained in [8] is 4x22 i.e. 88. Note that approximately twice the number of channel filter responses are obtained in our modified method at no extra hardware cost. This is basically because we only need to change the 'select' signal to change the adder block to a subtractor for alternate coefficients. It can also be noted that the center frequency resolution of our modified approach is twice of that in case of method in [8].

The masking filters are designed in such a way as to extract the adjacent bands on the frequency axis not generated simultaneously i.e. not generated for same combination of $M$ value and decimated filter/its inverse filter. Hence only fewer masking filters are required to obtain a large number of channel filter responses in our method. Only 6 masking filters (total filter length of all masking filters = 318) are to be implemented to extract 156 bands that can be generated. As the masking filters are fixed-coefficient masking filters, no extra time is required for reconfiguration as opposed to the method employed in [8].

## 5. CONCLUSION

In this paper we have proposed a method to generate variable bandwidth and variable center frequency channel filter responses covering the entire spectrum. Channel filter responses centered at the multiples of $\pi/M$ are obtained. Without any increase in hardware complexity compared to [8], our modified method generates approximately twice the number of frequency responses that of [8]. A generalized and optimized design of masking filters is also presented.

## 6. REFERENCES


[1] Santhi K. R., Srivastava V. K., SenthilKumaran G and Butare A, "Goals of true broadband's wireless next wave (4G-5G),"

*IEEE 58$^{th}$ Vehicular Technology Conference*, Florida, USA, vol. 4, pp. 2317-2321, 2003.

[2] Minnis B. J. and Moore P. A., “A highly digitized multimode receiver architecture for 3G mobiles,” *IEEE Transactions on Vehicular Technology*, vol. 52, no. 3, pp. 637-653, May 2009.

[3] Kuan-Hung Chen and Tzi-Dar Chiueh, “A low-power digit-based reconfigurable FIR filter,” *IEEE Transactions on Circuits and Systems II: Express Briefs*, vol. 53, no. 8, pp. 617-621, Aug. 2006.

[4] Zhangwen Tang, Jie Zhang and Hao Min, “A high-speed, programmable, CSD coefficient FIR filter,” *IEEE Transaction on Consumer Electronics*, vol. 48, no. 4, pp. 834-837, Nov. 2002.

[5] Demirsoy S. S., Kale I. and Dempster A. G., “Efficient implementation of digital filters using novel reconfigurable multiplier blocks,” *Conference Record of the Thirty Eighth Asilomar Conference on Signals, Systems and Computers*, California, USA, vol. 1, pp. 461-464, 2004.

[6] Y. C. Lim, “Frequency-response masking approach for synthesis of sharp linear phase digital filters,” *IEEE Transactions on Circuits and Systems*, vol. 33, no. 4, pp. 357-364, Apr. 1986.

[7] Yong Lian and Ying Wei, “A computationally efficient non-uniform FIR digital filter bank for hearing aids,” *IEEE Transaction on Circuits and Systems I: Regular Papers*, vol. 52, no. 12, pp. 2754-2762, Dec. 2005.

[8] K. G. Smitha and A. P. Vinod, “A new low power reconfigurable decimation-interpolation and masking based filter architecture for channel adaptation in cognitive radio handsets,” *Physical Communication*, vol. 2, no.1, pp. 47-57, Mar. 2009.

[9] S. J. Darak and A. P. Vinod, “A low complexity reconfigurable non-uniform filter bank for channelization in multi-standard wireless communication receivers,” *Journal of Signal Processing Systems*, vol. 68, no. 1, pp. 95-111, Feb. 2011.