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I. Introduction

For the Software Defined Radio (SDR) [Mitola J., 2000], we require a technology which will cater to multi-band, multi-standard and multi-service. This requires systems which are reconfigurable and reprogrammable by software. From the modulated input signal, depending on the communication standards, sub-bands with non-uniform bandwidths are to be extracted by the channel filters in the channelizer in a SDR receiver [Hentschel T et.al, 1999]. The channelizer operates at the highest sampling rate and hence requires low power and low complexity filters. Reconfigurability is an important requirement in an SDR handset. To facilitate this, a reconfigurable structure based on FRM [Lim Y.C., 1986] had been proposed by Smitha K.G. et. al. (2011, 2008) and Mahesh R. et. al (2011, 2007), which offers complexity reduction over the conventional per channel (PC) approach [Hentschel T, 2002]. In the PC approach, a separate filter is used for each communication standard. The low complexity reconfigurable architecture proposed by [Smitha K.G. et. al. (2011, 2008) and Mahesh R. et. al (2011, 2007) is to realize channels with non-uniform bandwidths and is based on the frequency response masking (FRM) method. The same prototype filter but separate interpolation factors and masking filters are used to derive different non-uniform bandwidth channels from the wideband input. So this is a very useful method to get non-uniform sharp transition-width channels and can be extended to realize sharp non-uniform filter banks with low complexity. The complexity of these reconfigurable filter banks can still be reduced by employing multistage masking instead of single stage masking whenever the order of the masking filters is high as proposed by Smitha K.G. et. al. (2011) and Mahesh R. et. al (2007). The problem with this method is that a high order prototype filter also increases the overall complexity. To address this problem, we use a modified architecture in which, if the order of the prototype filter estimated using Bellanger's equation [Bellanger M., 1981] is very high, we can realize the prototype filter also as an FRM filter [Lim Y.C., 1986]. Design example shows that this method offers 98.77% complexity reduction when compared to the PC approach and 53.01% complexity reduction when compared to the approach proposed by Smitha K.G. et. al. (2011) and Mahesh R. et. al (2007). By suitably combining appropriate low pass channels, we can get sharp transition-width band-pass channels also.

The filters designed using Parks-McClellan approach [Parks T. et.al, 1972, James H. et.al, 2005] result in filter coefficients with infinite precision. But for hardware implementation, they have to be represented using finite number of bits. The signed power-of-two (SPT) system [Yong Ching Lim D.L. et.al, 1999] is a representation in which the multiplier coefficients are represented using few non-zero bits. Thus the multiplication operation can be done using a series of shift and add operations. FIR filters with SPT representation is named as multiplier-less digital filters. Canonic Signed Digit (CSD) representation [Reid M. Hewitt et. al, 2000] is a special case of the SPT system. If we restrict the number of non-zero bits in the representation of the filter coefficients, the number of non-zero partial products needed to realize the filter will be reduced and hence the switching activity. Since the

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power consumption is directly proportional to the switching activity, power consumption also reduces.

But there will be degradation in the magnitude response of the filter due to the restricted number of SPT terms in the representation of the filter coefficients. Degradation in the frequency response due to the truncated FIR filter coefficients can be improved using suitable optimization techniques resulting in low power and low complexity FRM based multiplier-less reconfigurable non-uniform channels. Genetic algorithm for optimizing the CSD represented FRM filter is reported in [Patrick Mercier et.al, 2007], [Yu Y J and Y.C.Lim, 2002] and [Kilambi S. and B. Nowrouzian, 2006]. In this paper, we use ternary coded genetic algorithm [Manoj V.J. and E. Elias, 2009, 2009] for improving the performance of the multiple channel CSD represented filter architecture.

The paper is organized as follows. Section II gives an overview of the single stage and multi stage FRM method. Section III gives an insight into the reconfigurable architecture based on FRM. Section IV explains the CSD technique. Section V gives an overview about genetic algorithm based optimization. A design example and MATLAB simulation results are presented in section VI. Section VII compares the number of multipliers in the proposed method with the existing methods. Section VIII illustrates the method of obtaining band pass channels from low pass channels. Section IX concludes the paper.

II. Overview of Frm Approach

a) Single Stage FRM

Let $H(z)$ be the transfer function of the desired FIR low pass filter with pass band and stop band edge frequencies f_p and f_s respectively. In the FRM technique, the overall sharp transition width filter is composed of many sub-filters of wide transition width. If $H_a(z)$ represents the transfer function of a low pass linear phase filter, its complementary filter $H_c(z)$ can be expressed as given below

$$H_c(z) = z^{-(N-1)/2} - H_a(z) \qquad (1)$$

If $H_a(z)$ and $H_c(z)$ are interpolated with a factor M, $H_a(z^M)$ and $H_c(z^M)$ are obtained, whose transition-width is 1/M times the transition-width of $H_a(z)$ i.e. $(f_{as}-f_{ap})/M$. The filters $H_a(z^M)$ and $H_c(z^M)$ are cascaded to the masking filters $H_{Ma}(z)$ and $H_{Mc}(z)$ respectively, which suppress the unwanted images of $H_a(z^M)$ and $H_c(z^M)$. Thus, the transfer function of the overall FIR filter $H(z)$ [Lim Y.C., 1986] is given by

$$H(z) = H_a(z^M)\, H_{Ma}(z) + H_c(z^M)\, H_{Mc}(z) \qquad (2)$$

The structure of the FRM FIR filter is given in Fig. 1 [Lim Y.C., 1986].

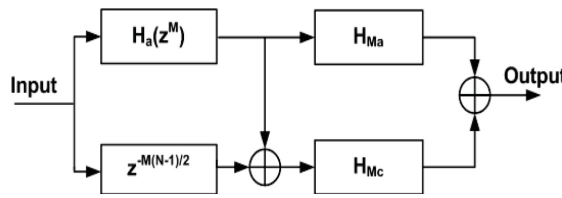

Fig.1 : Basic FRM Filter Architecture

The transition width of the overall filter H(z) is 1/M times the transition-width of $H_a(z)$ i.e. $(f_{as} - f_{ap})/M$. The design steps for the sub-filters are given below

$$m = \lfloor f_pM \rfloor \quad f_{ap} = f_pM - m \quad f_{as} = f_sM - m \qquad (3)$$

$$f_{map} = f_p \qquad f_{mas} = (m+1-f_{as})/M$$
$$f_{mcp} = (m-f_{ap})/M \qquad f_{mcs} = f_s \qquad (4)$$

Where $\lfloor x \rfloor$ denotes the largest integer less than x, M is the interpolating factor, f_p and f_s respectively are the pass band and stop band frequencies of the final filter $H(z)$. f_{ap} and f_{as} are the pass band and stop band frequencies respectively of the prototype filter $H_a(z)$. f_{map} and f_{mcp} are the pass band frequencies and f_{mas} and f_{mcs} are the stop band frequencies of the masking filters $H_{Ma}(z)$ and $H_{Mc}(z)$ respectively. The frequency responses of each filter are given in Fig. 2.

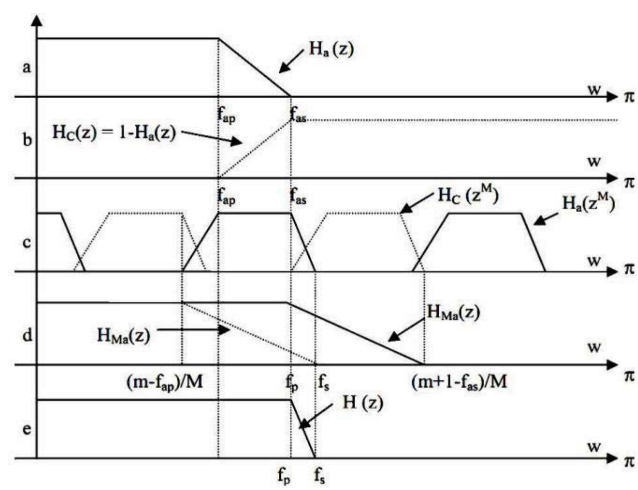

Fig.2 : Frequency response illustration of FRM approach

b) Multistage FRM

If we require a prototype filter with sharp transition width, its order becomes very high and this increases the overall filter complexity. The length of any filter can be estimated using Bellanger's equation [M. Bellanger, 1981] as given below.

$$N = (-2\log(10\delta_1\delta_2)/3\Delta f) - 1 \qquad (5)$$

Where δ_1 and δ_2 are the peak pass-band and stop-band ripple magnitudes respectively, and Δf is the normalized transition-bandwidth. If the length of the prototype filter estimated using the above equation is high, multi-stage FRM filter implementation can be used [Lim Y.C., 1986], [Lim Y. C and Lian Y, 1993], [Tapio Saramaki and Yong Ching Lim, 2003] [Yli- Kaakinen J. et.al., 2011, 2004] instead of single stage FRM filter. Thus the complexity of the overall filter can be reduced. The structure of a two stage FRM filter is given in Fig. 3.

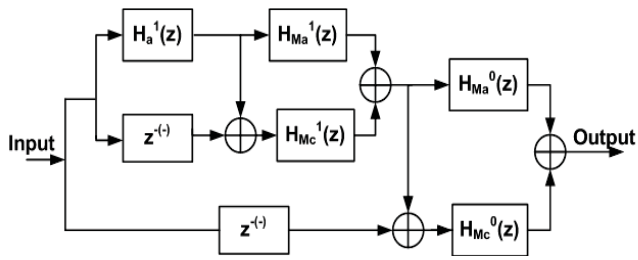

Fig.3 : Structure of two stage FRM filter

III. Reconfigurable Filters Based on Frm

a) Reconfigurable filters with single stage masking

The FRM approach can be extended to more number of channels to form the reconfigurable non-uniform bandwidth channels [Smitha K.G. et. al., 2011, 2008 and Mahesh R. et. al, 2011, 2007]. In this technique, the reconfigurability is achieved using the same prototype filter for all the channels. Different interpolation factors and masking filters are used to derive the different channels. For example, suppose we need two channels with pass band frequencies f_{p1} and f_{p2} and stop band frequencies f_{s1} and f_{s2} respectively. Let f_{ap} and f_{as} be the pass band and stop band frequencies respectively of the prototype filter which is the same for both channels. The prototype filter with specifications f_{ap} and f_{as} and interpolation factors M_1 and M_2 for channels 1 and 2 respectively can be designed by iterating the equation given below for different values of M_1 and M_2 [Smitha K.G. et. al., 2011, 2008 and Mahesh R. et. al, 2011, 2007].

$$f_{ap} = f_{p1}M_1 - \lfloor f_{p1}M_1 \rfloor = f_{p2}M_2 - \lfloor f_{p2}M_2 \rfloor$$

$$f_{as} = f_{s1}M_1 - \lfloor f_{s1}M_1 \rfloor = f_{s2}M_2 - \lfloor f_{s2}M_2 \rfloor \qquad (6)$$

The masking filters for each channel can be designed using the equations given in (4). The structure of the two channel reconfigurable filters is given in Fig 4.

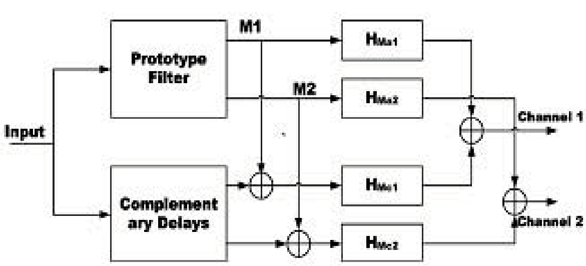

Fig. 4 : Two-channel single-stage architecture

H_{Ma1} and H_{Mc1} are the masking filters for channel 1 and H_{Ma2} and H_{Mc2} are the masking filters for channel 2. M_1 and M_2 are the interpolation factors for channel 1 and channel 2 respectively.

b) Reconfigurable filters with multi-stage masking

If the interpolation factor is high, the complexity of the masking filters also becomes high. This can be reduced using multi-stage masking instead of single stage masking. In multi-stage masking, the interpolation factor is factorized and masking is implemented in multiple stages where, in each stage masking filters are implemented for lower interpolation values which are factors of the overall interpolation factor [Smitha K.G. et. al., 2011 and Mahesh R. et. al, 2007]. The architecture of the reconfigurable filters using two stage masking is shown in Fig 5. Here, the prototype filter response is interpolated by factors M_1 and M_2 respectively. H_{Ma1} and H_{Mc1} are the masking filters designed for interpolation factor M_1. If M_2 can be factorized as $M_{21}*M_{22}$, then masking can be implemented in two stages where M_{21} and M_{22} are the interpolation factors of the first and second stage respectively. H_{Ma21} and H_{Mc21} are the first stage masking filters designed using equations (4) for interpolation factor M_{21} and they are interpolated with value M_{22}. H_{Ma22} and H_{Mc22} are the second stage masking filters designed for interpolation factor M_{22}.

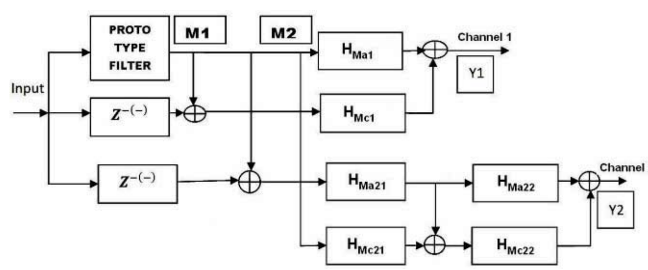

Fig. 5 : Architecture of two stage masking

c) Proposed Method - Reconfigurable filters with multistage FRM and multistage masking

In the architecture given in section IIIB, the problem with higher order prototype filter is not addressed. If the order of the prototype filter is high, the overall complexity of the multiple channel filter structure also will be high. In the proposed method, we modify the architecture in section IIIB by implementing the prototype filter using FRM as discussed in section IIB. This is shown in Fig. 6. Using this method, sharp transition-width channels with very low complexity can be obtained. Also since the number of filters is increased, we get improved pass band and stop band characteristics. By selecting appropriate low pass channels for subtraction, sharp transition-width band-pass channels with very low complexity can also be obtained. This is illustrated in section VIII.

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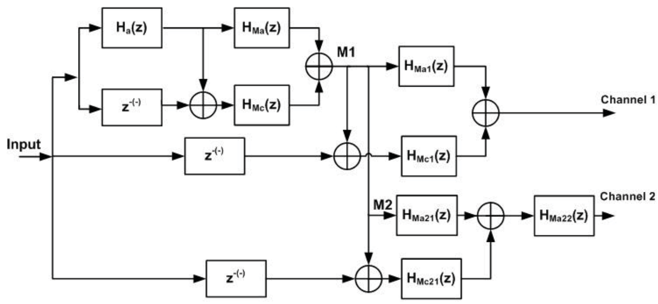


Fig. 6 : Reconfigurable non-uniform bandwidth filters with multistage masking and multistage FRM

IV. Canonic Signed Digit Representation

Any FIR filter can be represented [Zhangwen Tang et.al, 2002] as

$$y(n) = \sum_{k=0}^{N-1} h(k)x(n-k) \quad (7)$$

Where N is the length of the FIR filter, h(k) are the filter coefficients and x(n) is the input signal. The FIR filter implementation consists of multiplications, which are realized by shifters and adders. For the multiplication of the two N-bit numbers represented in the 2's complement form, in the worst case, N shifters and N-1 adders are needed. The number of non-zero partial product additions is determined by the number of non-zero bits in the filter coefficient. As the number of non-zero bits is reduced, the partial product additions are also reduced. CSD representation is a unique representation of the filter coefficients with minimum number of non-zero bits [Reid M. Hewitt et. al, 2000 and Zhangwen Tang et. al, 2002]. A fractional number q is represented in CSD format as

$$q = \sum_{i=1}^{W} c_i 2^{R-i} \quad (8)$$

Where $c_i = \{-1, 1, 0\}$ and W is the word length of the CSD number. Since, this encoding uses -1, 0 and 1 digits, it is called ternary coding. No adjacent digits in the CSD representation can be non-zero i.e. $c_i * c_{i-1} = 0$, where c_i is the i^{th} digit in the CSD representation. The maximum number of non-zero digits in the CSD representation of an n-bit number is floor(n/2), compared with n bits in the 2's complement representation. The number of adders/subtractors needed to realize CSD represented filter coefficient will be floor(n/2)-1.

V. Genetic Algorithm

Several optimization methods are proposed for the optimization of infinite precision FRM filter in which either separate optimization of sub-filters is done or joint optimization of the filters is done. The sub-filters designed using linear programming in the paper [Lim Y. C., 1986] reduced the error in the pass band and stop band of the overall FRM filter. Another approach to reduce the error in the pass band and stop band of the overall FRM filter is to design the sub-filters using Remez algorithm as proposed by Tapio Saramaki and Yong Ching Lim, 2003.

The optimization of the FRM filter in the discrete space is a complicated process and so efficient non-linear optimization techniques need to be used. The classical gradient based optimization techniques cannot be directly applied to this problem because, here the search space consists of integers. In this context, meta-heuristic algorithm is a good optimization tool as the proper selection of the parameters with respect to a particular design problem can bring forth global solution.

Genetic algorithms (GA) have been established as a good alternative for the optimization of multimodal, multi dimensional problems. This is a population based evolutionary algorithm where, in each iteration, candidate solutions are generated using genetic operations like reproduction, crossover and mutation. The use of GA for the optimization of the frequency responses of the CSD represented filters is employed by Yu Y.J. and .C. Lim, 2002, Samadi P. and M. Ahmad, 2007, [Fuller A. et.al, 1999 and Uppalapati H. et.al., 2005. Genetic algorithm for the optimization of FRM filter in the SPT space is used by Patrick Mercier et.al, 2007, Yu Y J and Y.C.Lim, 2002 and Kilambi S. and B. Nowrouzian, 2006.

a) Encoding of the optimization variables

Once the infinite precision filters are designed, the coefficients of these filters have to be converted to the CSD representation. As we have discussed in section IV, one way to encode filters in the CSD space is to represent them using ternary coding as proposed by Manoj V.J. and E. Elias, 2009. A look up table is created which has four fields. The four fields are index, CSD numbers, decimal equivalents and the number of SPT terms. In our work, the maximum allowed precision is 12 bits.

When crossover and mutation operations are performed on these ternary coded coefficients, the canonical property of the CSD representation of the coefficients may be lost. To ensure the canonical property of the filter coefficient representation, many restoration algorithms are proposed [Fuller A. et.al, 1999, Uppalapati H. et.al., 2005]. But these restoration algorithms increase the computational complexity of GA. A ternary coding based GA and a simple modified restoration algorithm is proposed by Manoj V.J. and E. Elias, 2009. We use this GA and restoration algorithm in our work.

b) Objective function for the design of the CSD coefficients of filters

When the filter coefficients are rounded to the nearest CSD number with restricted number of SPT terms, the stop band and pass band properties get degraded. So the filter coefficients in the CSD format have to be optimized to improve the pass band and stop band characteristics. In this paper, we use genetic algorithm based optimization technique to improve these characteristics. In our design, the CSD filter coefficients can use any number of SPT terms. But the total number of SPT terms used is restricted. This method gives more flexibility over restricting each coefficient with fixed number of SPT terms [Manoj V.J. and E. Elias, 2009].

The pass band and stop band characteristics are taken care of by minimizing the pass band ripple and maximizing the stop band attenuation. So, the objective function used in our work is given by

Minimize

$$\text{Max}\,\{\,|\,(|H(e^{j\omega})|\;-\;1)\,|\,,\;\omega < \omega_p \quad |H(e^{j\omega})|\,,\quad \omega > \omega_s\} \tag{9}$$

ω_p and ω_s are the pass band and stop band cut off frequencies respectively. So the objective function is to optimize the above values under the constraint that the total number of SPT terms used is restricted.

c) Optimization of multiple channel filter architecture using ternary coded GA

Since all the filters in our architecture have linear phase property, only the first half of the CSD represented filter coefficients are optimized and the other half of the filter coefficients are obtained using linear phase property. The genetic algorithms for optimizing the multiple channel filter architecture have two phases. Since we have multiple channels in our architecture, each channel has to be optimized separately. Also, since all the channels are derived from the same prototype filter outputs, we have to separately optimize the masking filters of each channel and the prototype filter. So, in the first phase, the prototype filter is optimized and in the second phase the masking filters of each channel are optimized.

Phase 1: Optimization of prototype filter: The basic steps in GA to optimize the filters are given in Fig 7 and are explained below [Randy L.H. and Sue E.H, 2004].

a) Initialization: The initial chromosome is generated by concatenating the first half of the continuous filter coefficients of the prototype filter. If the prototype filter is implemented by another FRM filter, the initial chromosome is generated by concatenating the first half of the continuous filter coefficients of all the sub-filters constituting the FRM prototype filter.

This initial chromosome has been rounded to the nearest CSD representation with maximum number of non-zero bits, which is chosen as 6 in this work. By changing the initial chromosome by random perturbations, a population pool of N-1 chromosomes is generated, where N is the population size. The coefficients of all these N-1 filters are then converted to the CSD representation with restricted number of non-zero bits. These N-1 perturbed chromosomes along with the initial non-perturbed chromosome will make the initial population of size N.

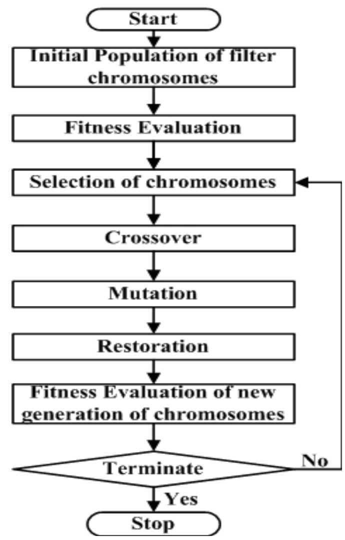


Fig. 7 : GA design flow for filter optimization

b) Fitness Evaluation of initial population : Each chromosome in the population is evaluated using the objective function given in (9) and they are ranked.

c) Selection : The best chromosomes based on ranking are selected from the population and they form the mating pool. In this paper, Roulette wheel rank selection [Randy L.H. and Sue E.H, 2004] is used for selecting the mates for cross over and reproduction.

d) Crossover : Offspring for the next generation are generated by exchanging the genes of parents. We use two-point crossover in our work, in which the genes are swapped between the parents between the two selected points.

e) Mutation : Some of the best solutions from the current population are propagated to the next generation without any change and the remaining chromosomes are mutated with some new information and propagated to the next generation. Due to cross over and mutation, the canonic property of the filter coefficients may be violated. The canonic property is retained using a simple restoration algorithm in which when two consecutive non-zero terms come, one of them is made zero [Manoj V.J. and E. Elias 2009].

f) Fitness Evaluation of New Population: Each chromosome in the new population is evaluated using the objective function given in (9) and they are ranked.

The steps from c to f are repeated until a maximum number of iterations are reached and then GA is terminated. When it is terminated, the best chromosome is taken from the population and is decoded to get the optimum CSD represented filter. The above steps have to be repeated for the prototype filter and masking filters separately.

Phase 2: Optimization of masking filters of each channel: After the prototype filter is optimized, the masking filters of each channel need to be optimized separately using the same objective function given in (9). In this phase, we adopt the joint optimization of the masking filters in which the initial chromosome is generated by concatenating the first half of the continuous filter coefficients of all the masking filters of a channel. The steps of GA algorithm are repeated until the maximum numbers of iterations are reached. This same optimization process has to be repeated for the masking filters of all the channels.

VI. Design Example and Simulation Results

The design example shown here compares the proposed method discussed in section IIIC with the methods discussed in section IIIA and IIIB. In this example, two channels of non-uniform bandwidth are designed for CDMA (Code Division Multiple Access) and PHS (Personal Handy Phone System) standards. The specifications for each channel are given below.

First Channel (CDMA)

Pass band (PB) frequency: 1250 kHz

Maximum pass band ripple: 0.1 dB

Stop band (SB) frequency: 1251 kHz

Minimum stop band attenuation: 40 dB

Sampling Rate: 6 MHz

Normalized PB frequency, f_{p1} = 1250/6000 = 0.2083

Normalized SB frequency, f_{s1} = 1251/6000 = 0.2085

Second Channel (PHS)

Pass band frequency: 300 kHz

Maximum pass band ripple: 0.1 dB

Stop band frequency: 301 kHz

Minimum stop band attenuation: 40 dB

Sampling Rate: 6 MHz

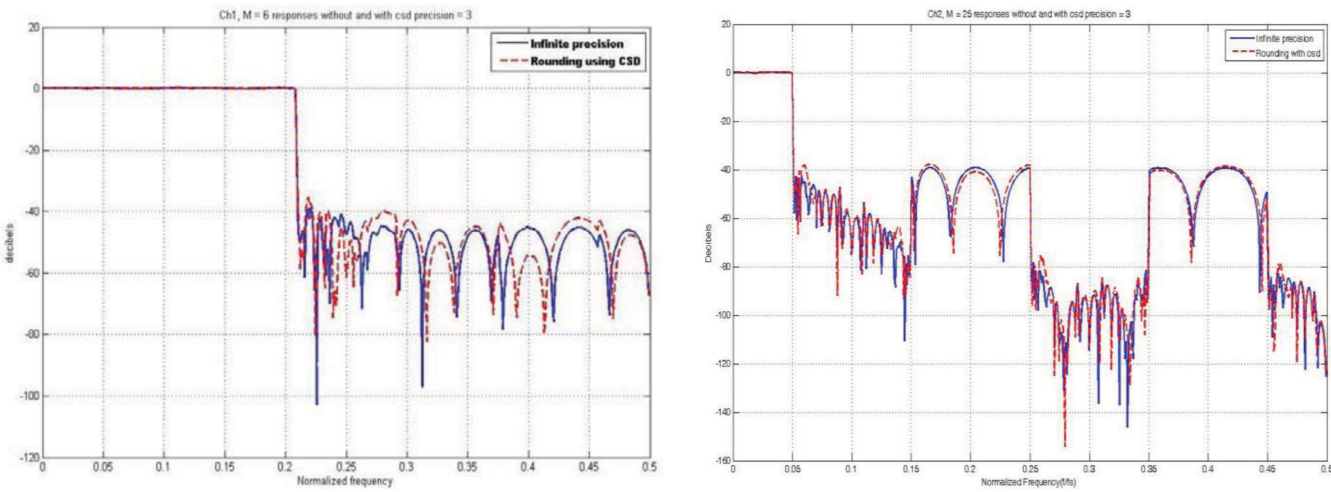
Normalized PB frequency, f_{p2} = 300/6000 = 0.05

Normalized SB frequency, f_{s2} = 301/6000 = 0.0502

The architecture for this non-uniform bandwidth two channel implementation is shown in Fig. 6. Using equation (6) and iterating for different values of M_1 and M_2, we obtain $M_1 = 6$, $M_2 = 25$, $f_{ap} = 0.25$ and $f_{as} = 0.255$. When these values are substituted in Bellanger's equation given in (5), the length of the prototype filter is obtained as 391 and the number of multipliers needed is 196. Since the length of the prototype filter is high, it is implemented as FRM filter and the number of multipliers needed for FRM implementation is then found to be 60. The CDMA filter response is implemented using single stage masking since its interpolation factor is small (M_1 = 6). For PHS filter response, the interpolation factor is 25 and can be factorized into 5 and 5. So, PHS filter is implemented using two stages masking with interpolation factor 5 in each stage. The length of the masking and masking complementary filters H_{Ma1} and H_{Mc1} of channel 1 is 24 and 23 respectively. The length of the first stage masking and masking complementary filters H_{Ma21} and H_{Mc21} of channel 2 is 20 and 19 respectively. The length of the second stage masking filter H_{Ma22} of channel 2 is 20. The total number of multipliers needed for the proposed method is 117 compared to the method proposed by Smitha K.G. et. al. (2011) and Mahesh R. et. al (2007) where 249 multipliers are needed. Thus, there is a 53.01% reduction in the number of multipliers. The peak pass band ripple and minimum stop band attenuation obtained for channel 1 and channel 2 when the prototype filter was implemented directly using Parks - McClellan method as in Fig. 5 and when it was implemented using FRM method as in Fig. 6, are shown in Table 1. From this table, it is seen that when the prototype filter was implemented using FRM method, the stop band characteristics are also improved compared to the direct implementation using Parks-McClellan method.

Table 1 : Performance comparison

Methods	Channel 1		Channel 2	
	Max. Pass Band Ripple (dB)	Min. Stop Band Attenuation (dB)	Max. Pass Band Ripple (dB)	Min. Stop Band Attenuation (dB)
Reconfigurable two channel filters with two stage masking and direct implementation of prototype filter	0.213	37.28	0.2268	37.98
Reconfigurable two channel filters with two stage masking and FRM prototype filter	0.178	38.66	0.2306	39.14



(a) Channel 1 (b) Channel 2

Fig. 8 : Frequency response of channel 1 (M_1 = 6) and channel 2 (M_2 = 25) using infinite precision filters and CSD represented filters with maximum 3 SPT terms

Next, the filters are represented using CSD with restricted number of bits. The restriction in the number of non-zero bits in the CSD representation is varied from 2 to 6 and the performance is compared. The frequency responses of the infinite precision filters and CSD represented filters with maximum three SPT terms, for channel 1 and channel 2 are shown in Fig 8a and 8b respectively.

In Table 2, the peak pass band ripple and minimum stop band attenuation of the 2 channels for the infinite precision filters and CSD represented filters using different number of non-zero bits are shown. From Table 2, it can be seen that when the number of non-zero bits for the CSD representation is restricted to 3, for the first channel, the peak pass band ripple is increased by 0.0215 dB and minimum stop band attenuation is degraded by 3.22 dB. For the second channel, the stop band attenuation is degraded by 1.33 dB compared to the infinite precision filter response. So we have to optimize the CSD represented filter responses to get better responses for channel 1 and 2. Here, we have employed GA optimization to improve the pass band and stop band responses.

For the multiple channel filter architecture, optimization includes two phases. In the first phase, the coefficients of the prototype filter are optimized which is implemented as an FRM filter. The FRM prototype filter is made up of sub-filters and each of these continuous filter coefficients has linear phase characteristics. So for optimizing the coefficients of a prototype filter, the genes corresponding to the first 30 genes of the FRM sub-filter, the first 16 genes of the masking filter and the first 16 genes of the masking complementary filter are concatenated, to generate the chromosomes in the initial population. The maximum allowed number of SPT terms is taken as 186 so that the average number of SPT terms in a filter coefficient is 3. The different parameters used for the optimization of the prototype filter are given below:

Number of Iterations = 500

Population Size = 50

Table 2 : Performance comparison when max. no: of SPT terms are varied

	Channel 1		Channel 2	
Coefficients	**Max. Pass Band Ripple(dB)**	**Min. Stop Band Attenuation (dB)**	**Max. Pass Band Ripple(dB)**	**Min. Stop Band Attenuation (dB)**
With Infinite precision	0.178	38.66	0.2306	39.14
Number of non-zero bits = 2	0.2965	32.07	-0.4419	28.24
Number of non-zero bits = 3	0.1995	35.44	-0.2144	37.81
Number of non-zero bits = 4	0.1944	39.2	0.2254	38.44
Number of non-zero bits = 5,6	0.1943	39.58	0.2171	38.76

Number of population members that survive each generation = 1

Mutation Rate = 0.02

Number of the best population which is kept without change during mutation (Elite Count) = 10

In the second phase, the coefficients of the masking filters of each channel are optimized. Each of these continuous filter coefficients has linear phase characteristics. Therefore, for optimizing the coefficients of the masking filters of channel 1, the genes corresponding to the first 12 genes of the masking filter and the first 12 genes of the masking complementary filter are concatenated to generate the chromosomes in the initial population. The maximum allowed number of SPT terms is taken as 72 so that the average number of SPT terms in a filter coefficient is 3. Similarly for optimizing the coefficients of the masking filters of channel 2, the genes corresponding to the first 10 genes of the first stage masking filter, the first 10 genes of the first stage masking complementary filter and the first 10 genes of the second stage masking filter are concatenated to generate the chromosomes in the initial population. The maximum allowed number of SPT terms are taken as 90 so that the average number of SPT terms in a filter coefficient is 3. The different parameters used for the optimization of the masking filters of both channels are given below:

Number of Iterations = 500

Population Size = 50

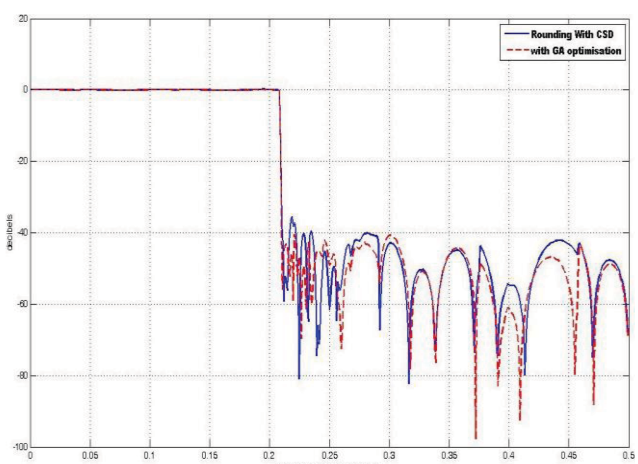
Number of population members that survive each generation = 5

Mutation Rate = 0.2

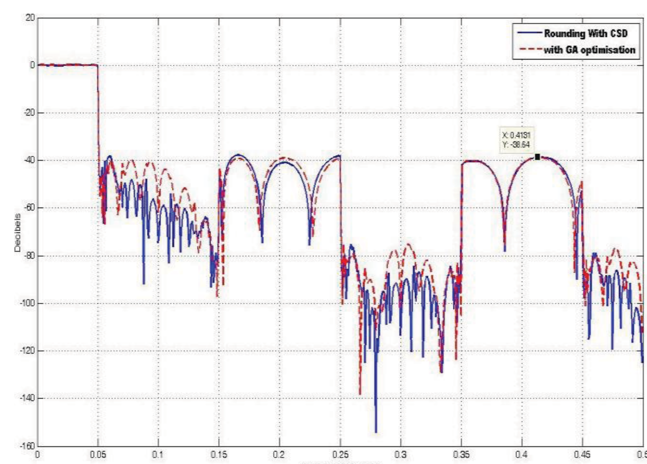
Number of the best population which is kept without change during mutation (Elite Count) = 10

The frequency responses of channel 1 and channel 2 after GA optimization are given in Fig 9a and 9b respectively.

The peak pass band ripple and minimum stop band attenuation obtained for channel 1 and channel 2 outputs after GA optimization of the prototype filter and the masking filters with maximum 3 SPT terms are given in Table 3. From the table it is clear that, when we employed optimization, the pass band responses of channel 1 and channel 2 are improved by 0.0233 dB and 0.0723 dB respectively and the stop band responses of channel 1 and channel 2 are improved by 4.92 dB and 0.83 dB respectively.



(a) Channel 1



(b) Channel 2

Fig. 9 : Frequency response of channel 1 (M_1 = 6) and channel 2 (M_2 = 25)with CSD represented filters with maximum 3 SPT terms before and after GA optimization

Table 3 : Performance comparison of CSD represented filters with maximum 3 SPT terms before and after GA optimization

.	Channel 1		Channel 2	
Coefficients	**Max. Pass Band Ripple(dB)**	**Min. Stop Band Attenuation (dB)**	**Peak Pass Band Ripple(dB)**	**Min. Stop Band Attenuation (dB)**
With Infinite precision	0.178	38.66	0.2306	39.14
With CSD representation	0.1995	35.44	-0.2144	37.81
With CSD representation and GA	0.1762	40.36	-0.1421	38.64

If the number of non-zero bits is reduced to 2, further reduction in complexity is obtained. But from Table 2, it can be seen that when the number of non-zero bits for the CSD representation is restricted to 2, the peak pass band ripple is increased by 0.1185 dB and minimum stop band attenuation is degraded by 6.59 dB for the first channel. For the second channel, the peak pass band ripple is increased by 0.2113 dB and the minimum stop band attenuation is degraded by 10.9 dB compared with the infinite precision filter responses. The different parameters used for the optimization of the prototype filter are given below:

Number of Iterations = 500

Population Size = 50

Number of population members that survive each generation = 1

Mutation Rate = 0.02

Number of the best population which is kept without change during mutation (Elite Count) = 5

The different parameters used for the optimization of the masking filters of channel 1 and channel 2 are given below:

Number of Iterations = 500

Population Size = 50

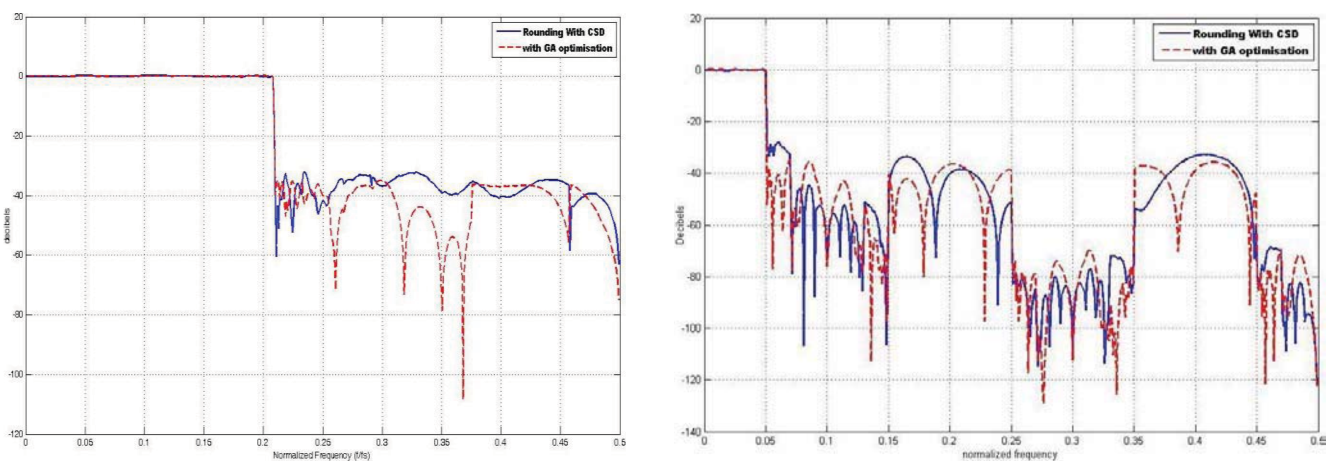
Number of population members that survive each generation = 15

Mutation Rate = 0.1

Number of the best population which is kept without change during mutation (Elite Count) = 20

The frequency responses of channel 1 and channel 2 after GA optimization are given in Fig 10a and 10b respectively.

The peak pass band ripple and minimum stop band attenuation achieved for channel 1 and channel 2 outputs after GA optimization of the prototype filter and masking filters are given in Table 4. From the table it is clear that, the pass band responses of channel 1 and channel 2 are improved by 0.0832 dB and 0.2806 dB respectively and stop band response of channel 1 and channel 2 are improved by 2.6 dB and 6.34 dB respectively, after optimization.



(a) Channel 1 (b) Channel 2

Fig. 10 : Frequency response of channel 1 ($M_1 = 6$) and channel 2($M_2 = 25$)with CSD represented filters with maximum 2 SPT terms before and after GA optimization

Table 4 : Performance comparison of CSD represented filters with maximum 2 SPT terms before and after GA optimization

	Channel 1		Channel 2	
Coefficients	**Peak Pass Band Ripple(dB)**	**Minimum Stop Band Attenuation (dB)**	**Peak Pass Band Ripple(dB)**	**Minimum Stop Band Attenuation (dB)**
With Infinite precision	0.178	38.66	0.2306	39.14
With CSD representation	0.2965	32.07	-0.4419	28.24
With CSD representation and GA optimization	0.2133	34.67	0.1613	34.58

VII. Complexity Comparision

Here, we compare the number of multipliers used to design channel 1 and channel 2 filters. The number of multipliers used to design any filter is given by

$$f(N) = (N + 1)/2 \text{ if N is odd}$$
$$(N/2) + 1 \text{ If N is even}$$

Where f(N) denotes the total number of multipliers needed for the implementation of an FIR filter with order N. If N_a is the order of the prototype filter $H_a(z)$, N_{Ma} is the order of the masking filter $H_{Ma}(z)$, and N_{mc} is the order of the masking complementary filter $H_{Mc}(z)$, the number of multiplications, π, required to implement the overall FIR filter is given by

$$\pi = f(N_a) + f(N_{ma}) + f(N_{mb})$$

In Table 5, the complexity comparison of the proposed method with the existing method in terms of multipliers is shown. We can see that the proposed method offers a 53.01% reduction over the method discussed in section IIIB.

VIII. Band-pass Channels From Low Pass Channels

In this section, the method of extracting band pass channels from the designed low pass channels is shown. To illustrate this, we have modified the structure shown in Fig. 6 to a three channel structure by adding one more channel with the following specifications. The frequency responses of all the three low pass channels are given in Fig 11(a-c).

Third Channel

Pass band frequency: 900 kHz

Maximum pass band ripple: 0.1 dB

Stop band frequency: 901 kHz

Minimum stop band attenuation: 40 dB

Sampling Rate: 6 MHz

Normalized PB frequency, fp2 = 900/6000 = 0.15

Normalized SB frequency, fs2 = 901/6000 = 0.1502

The band-pass responses are shown in Fig 11(d-f). The band-pass response shown in Fig. 11d is derived by combining the first and second channel low pass responses. The band-pass response shown in Fig. 11e is obtained by combining the first and third low pass channel responses. The band-pass response shown in Fig. 11f is obtained by combining the second and third channel responses.

Table 5 : Complexity comparison of various methods with the proposed method for two channel filter architecture

No:	Method	Channels	No. Of Multipliers				% saving in multipliers
			H_a	H_{Ma}	H_{Mc}	Total	
1	PC method	(First channel)		4751		9502	
		(Second channel)		4751			
2	PC method using FRM Filters	(First channel)	135	88	65	573	93.97%
		(Second channel)	143	70	72		
3	Reconfigurable two channel filters with single-stage masking	(First channel)	192	13	12	314	96.7% w.r.t 1 45.2% w.r.t 2
		(Second channel)		49	48		
4	Reconfigurable two channel filters using two stage masking for second channel filter	(First channel)	192	13	12	249	97.38% w.r.t 1 56.54% w.r.t 2 20.7% w.r.t 3
		(Second channel first stage masking)		11	10		
		(Second channel second stage masking)		11			
5	Reconfigurable two channel filters with two stage masking and FRM prototype filter	(First channel)	60	13	12	117	98.77% w.r.t 1 79.58% w.r.t 2 62.74% w.r.t 3 53.01% w.r.t 4
		(Second channel first stage masking)		11	10		
		(Second channel second stage masking)		11			

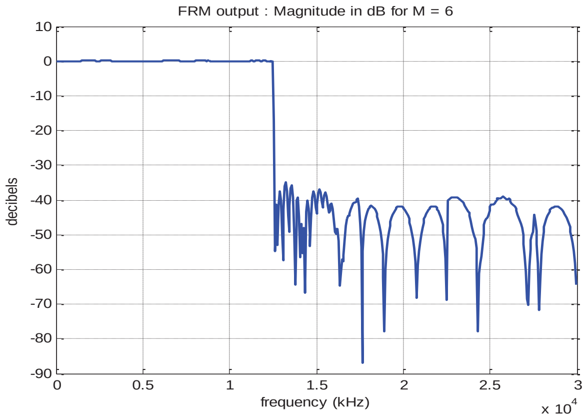


Fig.11a : First low pass channel response

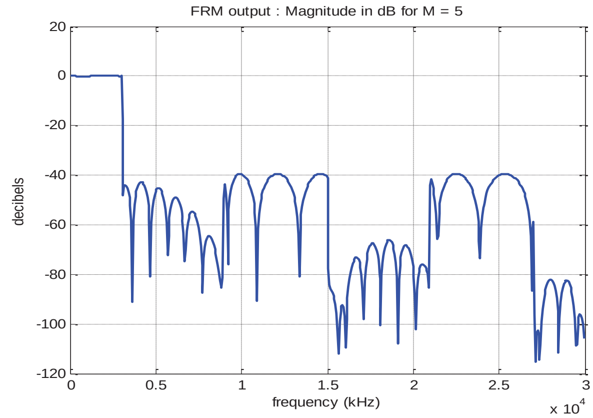


Fig.11b : Second low pass channel response

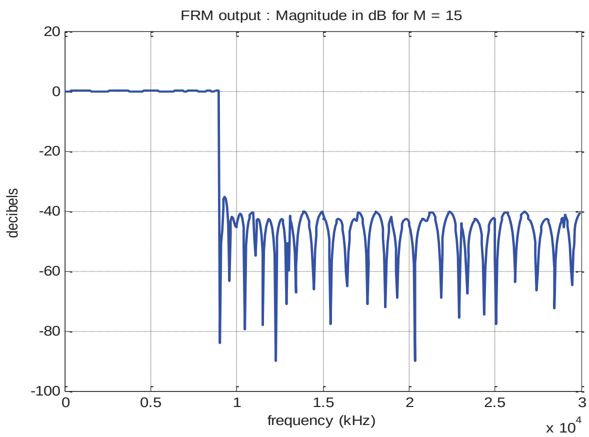


Fig.11c : Third low pass channel response

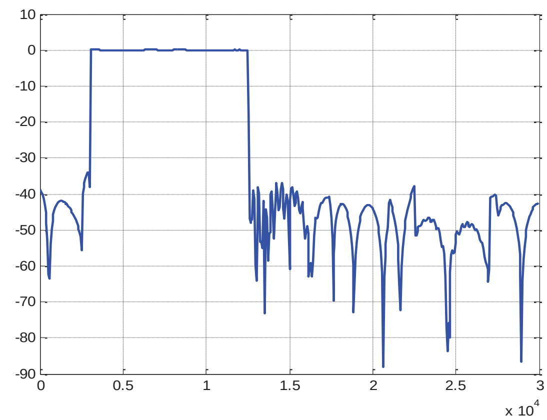


Fig.11d : First Band-pass response

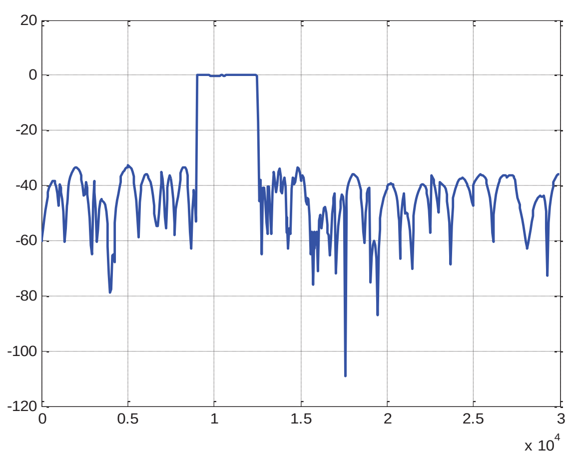


Fig.11e : Second Band-pass response

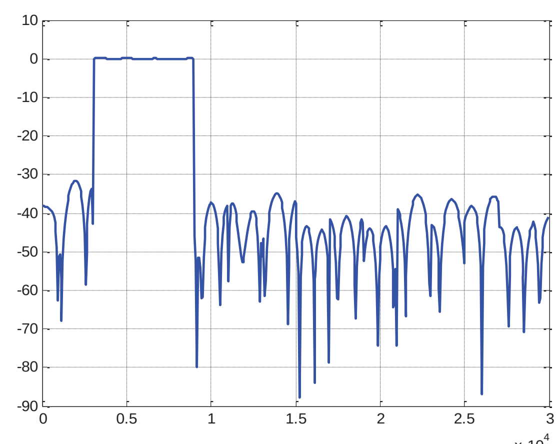


Fig.11f : Third Band-pass response

IX. Conclusion

The existing architecture with low power, low complexity and reconfigurability for software defined radio is modified in this paper. The prototype filter in the FRM structure is replaced by another FRM structure. The filter coefficients are represented in the signed power of two spaces, where CSD representation is employed. Thus we get multiple channel architecture with very low complexity and low power which is ready for hardware implementation. Since the design of the filter in the discrete space degrades the performance, the response is optimized by a modified genetic algorithm which results in near optimal solutions. This leads to the implementation of low power, low complexity multiplier-less, reconfigurable, non-uniform channel filters. Since different channels correspond to different communication standards, different objective functions and different optimization techniques may be used which may lead to better performance. This architecture can be extended to more number of channels also.

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