International Journal on Emerging Technologies **11**(3): 959-964(2020)

ISSN No. (Print): 0975-8364
ISSN No. (Online): 2249-3255

Design of Successive Approximation (SA) based Low Complexity Multiplier-less Coefficient Decimation Filter Bank for SDR Receivers

K. Baboji[1] and Sriadibhatla Sridevi[2]

[1]Assistant Professor (Jr.), Department of Micro and Nano Electronics,
Vellore Institute of Technology, Vellore-632014, Tamilnadu, India.
[2]Associate Professor, Department of Micro and Nano Electronics,
Vellore Institute of Technology, Vellore-632014, Tamilnadu, India.

(Corresponding author: Sriadibhatla Sridevi)
(Received 18 March 2020, Revised 16 May 2020, Accepted 19 May 2020)
(Published by Research Trend, Website: www.researchtrend.net)

ABSTRACT: Coefficient decimation filter banks (CDFBs) found its application in software defined radio (SDR) channelizers because of its advantages such as highly reconfigurable and low complexity (number of multiplications). Multiplier-less realization is highly preferable when the ultimate goals are implementation simplicity and processing speed. This paper presents a design of multiplier-less CDFB where coefficient space of filters belongs to sum of powers-of-two (SOPOT) form. To obtain, the SOPOT approximations of impulse response values of the CDFB sub filters, a modified version of vector successive approximation (SA) technique termed as Matching Pursuits Generalized BitPlanes (MPGBP) algorithm is utilized. Design examples show that the computational time and implementation complexity (number of adders) of the proposed CDFB is lower than those of its counter parts.

Keywords: CDFB design, Coefficient decimation, SA technique, SOPOT representation, SDR receivers.

Abbreviations: FB, filter bank; SDR, software defined radio; DFTFB, discrete Fourier transform FB;CDFB, coefficient decimation FB; SPT, signed powers-of-two; GA, genetic algorithm; FRM, frequency response masking; CSD, canonic signed digit; DBNS, double base number system; MDFTFB, modified DFTFB; CMFB, cosine modulation FB; SA, successive approximation; CDP-I, coefficient decimation process-I,CDP-II, coefficient decimation process-II; SOPOT, sum of powers-of-two; MPGBP, Matching Pursuits Generalized BitPlanes; CDMA, code division multiple access; ICGA, integer coded GA; ICDE, integer coded differential evolution; APBR, average pass band ripple; ASBA, average stop band attenuation; CC, continuous coefficients.

I. INTRODUCTION

FB found its applications including compression of images and data, trans-multiplexers, adaptive and bio-medical signal processing [13]. One such application of digital FBs is found in channelizers of SDR receivers [9,10]. Here the role of channelizers is to extract multiple channels (radio) of different communication standards from the wide band input signal for further processing where the channel bandwidth of each communication standard is not the same. Hence, channelizers must be reconfigurable in order to support various communication standards.

The usage of Per-channel approach for this purpose has the limitation that the channelizer's complexity increases linearly with an increase in the number of channels that needs to be received simultaneously [1]. A much better alternative to the Per-channel approach is DFTFB [14]. DFTFB comprises a single low pass filter and DFT [3]. The complexity of the DFTFB does not depend on how many channels to receive. DFTFB is a modulated FB consisting of band pass filters of equal bandwidth. Hence it is unable to separate channels of various communication standards. A more flexible low complexity FB which can be used as a better substitution to DFTFB is CDFB [4]. CDFB has advantages such as absolute control on centre frequency locations of the generated pass bands and the pass band widths.

A further reduction in the FB's complexity can be accomplished by confining its filter coefficients to SPT representations. The corresponding hardware implementations are belonging to class of multiplier-less FBs where multiplication operations are performed by shifting elements and addition operators. Many multiplier-less filters and FBs are discussed in [8,2,11]. The paper in [8] uses the genetic GA to optimize the coefficient spaces of multiplier-less FRM filter over CSD and DBNS. In [2, 11], the design of FRM based MDFTFB and CMFB using meta-heuristic algorithms is presented. The filters in DFTFB and CMFBs can be obtained from a single model filter by using exponential and cosine modulations respectively. Hence, in these FBs, the optimization of prototype filter using evolutionary algorithms suffices to design the respective FB. However, the design of multiplier-less FBs using evolutionary algorithms has limitations such as the generated solutions may not be the optimum solutions and the design time is more.

This paper deals with the design of multiplier-less CDFB employing SA of vectors. The design time and implementation complexity (number of adders) of the proposed FB is lower over its counter parts.

The paper is arranged the following way. Section I discusses the review of CDFB for SDR receivers. Section II presents the proposed multiplier-less CDFB using vector SA method.

The results analyses are discussed in Section III. Section IV explains the conclusion.

II. REVIEW OF CDFB FOR SDR RECEIVERS

In channel extraction using CDFB, channels of multiple communication standards are extracted with the help of two processes namely CDP-I and CDP–II [4]. CDP-I begins with designing an N –tap low pass FIR filter termed as model filter. Afterwards, the model filter's M^{th} coefficient is unchanged and the remaining coefficients are replaced by zeros. This process results in a decimated model filter with frequency response consisting of model filter pass band images at integer multiples of $2\pi / M$.

Let $f(n)$ represent the initial set of coefficients, and $f'(n)$ is the new coefficient set formed by replacing the initial coefficients with zeros other than the M^{th} (CDP–I process). If $F(e^{j\omega})$ and $F'(e^{j\omega})$ are the Fourier transforms of $f(n)$ and $f'(n)$ respectively then $F'(e^{j\omega})$ is given by [4]

$$F'\left(e^{j\omega}\right)=\frac{1}{M}\sum_{r=0}^{M-1}F\left(e^{j\left(\omega-2\pi r/M\right)}\right) \quad (1)$$

It can be observed from Eq. (1) that the magnitude response is scaled by M and at integer multiples of $2\pi/M$, the replicas of original spectrum is produced. Hence, it is necessary to scaling up the output of the filter by M to recover the original signal. It is also evident from Eq. (1) that each value of M gives a distinct multi-band response. One can obtain the required frequency bands by subtracting one multi-band response from another multi-band response or exploiting appropriate frequency masking filters.

In CDP–II, by grouping every M^{th} coefficient of the model filter and removing in-between coefficients, a frequency response which is a decimated version of original frequency response (model filter frequency response) can be obtained. In decimated frequency response, the pass band width is M times the original pass band width. By consecutively applying CDP–II and CDP-I operations on model filter, it is possible to extract the channels whose bandwidth is M times of model filter band width.

The generalized architecture of CDFB is illustrated in Fig. 1. The output of the model filter is shown as y_1 . y_2 shows the output of the CDP-I for $M = 2$, y_3 shows the output of the CDP-I for $M = 3$ and so on. As shown in Fig. 1, y_M can be obtained by keeping the model filter's M^{th} coefficient unaltered and replacing the remaining model filter's coefficients with zeros. $y_2 - y_1$ and $y_3 - y_1$ gives the outputs which are obtained by spectral subtraction of two multi-band responses. Filters H_{M1}, and H_{M2} are frequency masking filters which extracts specific channels from input y_{3c} where y_{3c} gives the output of the complementary filter. The CDFB's complete design procedure is given as sub-section 2.3 in [4].

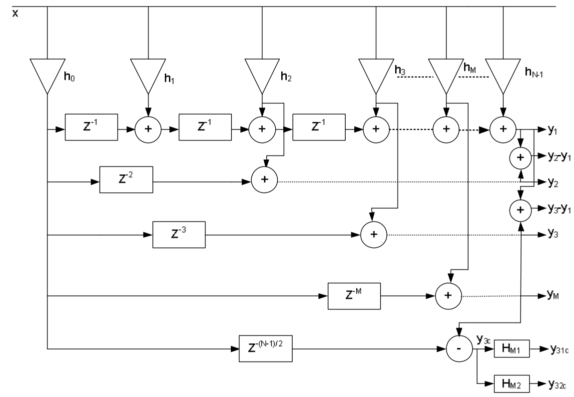

Fig. 1. The generalized architecture of CDFB [4].

Multiplier-Less Design of CDFB using SA Method: Here we provide a vector SA method for designing the sub-filters' (model and masking filters) in CDFB whose coefficient space is SOPOT [12]. If CDFBs have filters whose coefficient representations belong to such a class are known as multiplier-less CDFBs given that the multiplications operations are performed with addition operations and shifting operations. The design process begins with obtaining the N^{th} order Parks-McClellan prototype designs $h(n)$ for a given sub-filters' specifications which are represented as

$$h = \left[h(0)h(1)h(2)\cdots h(N)\right]^T \in R^N \quad (2)$$

Afterwards, the SOPOT approximations of the impulse response values are obtained using a modified MPGBP algorithm as shown below. Here the impulse response values (coefficients) are approximated successively using a dictionary codewords. Each codeword is a N -dimensional vector whose components take only values 0, +1 and -1. These vectors are weighted with power of two terms when approximating an impulse response.

The dictionary defined for SOPOT approximation of impulse response values is given by $D = \{\pm l_1, \pm l_2, \pm l_3 ......\pm l_Q\}$ in which each codeword $\pm l_j \in R^N$ has P components of unity magnitude and $N+1-P$ components of zero magnitude i.e., codewords are permutations of P ± 1 's and $(N+1-P)$ 0's. The residue vector v in modified MPGBP algorithm is initialized with vector h .

The modified MPGBP algorithm. [12]

1. Initiate $k = 1, v_1 = h$.
2. Repeat the process still a certain stop criterion is obtained
   a) Choose $i_k \in \{1,2,3,4\cdots Q\}$ in such a manner that
   $$\langle v_k, l_{i_k}\rangle = \max_{1\le i\le Q}\{\langle v_k, l_i\rangle\}$$
   b) Select
   $$p_k = \left\lceil -\log_2\left(\frac{4\langle v_k, l_{i_k}\rangle}{3P}\right)\right\rceil$$

c) Replace

$$v_{k+1} \leftarrow v_k - (1/2)^{p_k} l_{i_k}$$

d) $k \leftarrow k+1$

3. Stop

where $\langle v_k, l_{i_k} \rangle$ = Inner product of v_k and l_{i_k} vectors and $\lceil y \rceil$ is the ceil value of *y*.

Following K steps, the MPGBP algorithm provides a vector h approximation and is given by [12]

$$h^{(k)} = \sum_{k=1}^{K} 2^{-p_k} l_{i_k} \quad (3)$$

For a component n of *h*, its SOPOT approximation is given by

$$h^{(k)}(n) = \sum_{k=1}^{K} 2^{-p_k} l_{i_k}(n) \text{ for } n \in [0, N-1] \quad (4)$$

To obtain the closest codeword l_{i_k} in D to v_k in Step. 2(a) of modified MPGBP algorithm, a two-step procedure is used. In the first step, the coordinates of the current residue v_k are arranged in decreasing order of magnitude and the index values of the P largest ones are stored. In the next step, by setting P coordinates whose indexes are stored previously, to +1 if the coordinate is positive, and -1 if the coordinate is negative, the closest codeword l_{i_k} can be obtained. The other coordinates will be set to zero.

Here the stop criteria of the algorithm are taken as the maximum number of MPGBP steps. The value of K gives number of MPGBP steps for which the approximation error $\|h - h^{(k)}\|$ to be bounded by ε and is given by [12]

$$K \geq \begin{cases} \dfrac{\log\left(\dfrac{\varepsilon}{\|h\|}\right)}{\log\sqrt{1 - \dfrac{8}{9}\dfrac{P}{N}}} & \text{for} \quad P < \sqrt{N} \\[2ex] \dfrac{\log\left(\dfrac{\varepsilon}{\|\mathrm{h}\|}\right)}{\log\sqrt{1 - \dfrac{8}{9}\dfrac{1}{P}}} & \text{for} \quad P \geq \sqrt{N} \end{cases} \quad (5)$$

Here ε = upper bound on approximation error $(\|h - h^{(k)}\|)$ = norm of vector *h*.

The SOPOT filter designs obtained from the output of the MPGBP algorithm may not be the minimal design i.e., the design with minimum number of terms, as there may be SOPOT representations of the filter coefficients whose powers-of-two terms may be combined into a different power of two term $(2^{-k} + 2^{-k} = 2^{-k+1})$. Hence minimal representations of the coefficients can be obtained by using an algorithm named as Common-terms reduction algorithm as shown below. The algorithm begins with the $h^{(k)}(n)$ coefficient after k^{th} step from the MPGBP algorithm's output. Here the minimal representation of the coefficient is given by $\sum_m 2^{a_m}$.

Common-Terms Reduction Algorithm. [12] For each $h^{(k)}(n)$, $n \in [0, N-1]$

1. Set $b = h^{(k)}(n)$, $m = 0$
2. Repeat until $b = 0$

(a) Increment *m*

(b) Obtain $a_m = \lfloor \log_2 b \rfloor$

(c) Replace $b \leftarrow b - 2^{a_m}$

Here $\lfloor y \rfloor$ is the floor value of *y*.

For a given filter specification, the best approximation with the proposed method can be obtained by testing prototype designs of various orders of N, beginning with least order, required to meet the prescribed specifications and with different values of P. To achieve this, the steps of the MPGBP algorithm is repeated over various N values with different P values. After each MPGBP step (k), the frequency response of the approximated design $(h^{(k)})$ is determined and tested for its validity by comparing it with the given frequency specifications and if it satisfies the specifications, $h^{(k)}$ is taken as the valid SOPOT design. Afterwards, one may look for the best design which meets the prescribed specifications and provides the minimum implementation complexity (number of adders).

Thus the best SOPOT approximations of sub-filters' are obtained as discussed above and finally these designs are used to obtain the overall multiplier-less CDFB whose implementation complexity (number of adders) is low.

III. SIMULATION RESULTS

All simulations in the proposed work were carried out on Intel(R) core(TM) i3–3110M CPU running at 2.30 GHz with the help of MATLAB R2017b.

Design example. Here we considered the design example as multiplier-less design of 2-channel CDFB which extracts two CDMA channels as given in [5]. Assume that the input signal has two CDMA bands. Let the CDMA signals vary between 5000 KHz to 6250 KHz and between 8000 KHz to 9250 KHz (two 1250 KHz CDMA channels each). The chosen sampling rate is 20 MHz. It is also stated that pass band ripple and stop band attenuation requirements of channels 1 and 2 as 0.1dB and -40dB respectively.

The frequency specifications of the model filter F_{a1} designed for the CDMA response is $f_p = 0.0625$ and $f_s = 0.08$ (Frequencies for half of the sampling rate are normalized). From Eq.(1), the value of M is obtained as 7. By applying CDP-I operation on model filter F_{a1} with $M = 7$, the frequency response as shown in Fig. 2 can be obtained. Appropriate masking filters F_{Ma1} and F_{Ma2} can therefore be used to separate the CDMA channels (third and fourth band as shown in Fig. 2). The frequency specifications of the required masking filters F_{Ma1} and F_{Ma2} are shown in Tables 1 and 2 respectively.

The 2-channel CDFB architecture is illustrated in Fig. 3.

For the multiplier-less implementation of the CDFB for extracting CDMA channels, the required SOPOT representations of the coefficients of the filters F_{a1}, F_{Ma1} and F_{Ma2} are obtained using SA method. Since the designed filters are having linear phase property, the half of the coefficients of the filters are used in the formation of an initial residue vector in SA method which in turn reduces the dimensionality of the residue vector; hence the computation time is low. In the beginning, the minimum lengths of the Parks-McClellan prototype designs of the filters F_{a1}, F_{Ma1} and F_{Ma2} are determined.

Table 1: Frequency specifications of the masking filters F_{Ma1}.

Masking filter F_{Ma1} frequency specifications	Values
f_{ma1s1}	0.3482
f_{ma1p1}	0.5089
f_{ma1p2}	0.6339
f_{ma1s2}	0.7946

Table 2: Frequency specifications of the masking filters F_{Ma2}.

Masking filter F_{Ma2} frequency specifications	Values
f_{ma2p}	0.6514
f_{ma2s}	0.7946

Here we considered the initial lengths of the filters F_{a1}, F_{Ma1} and F_{Ma2} as 320, 35 and 31 respectively. Since CDP-I operation is performed on model filter (of length 320) with $M = 7$, the required number of multiplications becomes $\left\lceil \frac{320}{7} \right\rceil$. The SA method is repeated for various values of N staring from minimum length with different P values in order to search for the minimum implementation complexity. Considering the better performance of 2-channel multiplier-less CDFB in terms of the ripple specifications and adder complexity, the P and N values of the selected SA designs of the filters F_{a1}, F_{Ma1} and F_{Ma2} which are used in the implementation of the overall CDFB are reported in Table 3.

The magnitude responses of the SA designs of the filters F_{a1}, F_{Ma1} and F_{Ma2} are plotted in Fig. 4 where as that of channels 1 and 2 are shown in Fig. 5 and Fig. 6 respectively. Here the frequency response of the channels is plotted using proposed SA method against continuous coefficient method.

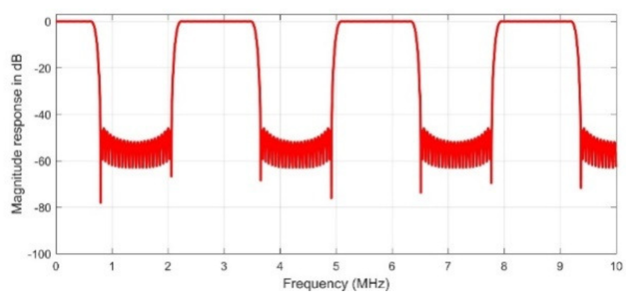

Fig. 2. The magnitude response of the model filter with $M = 7$.

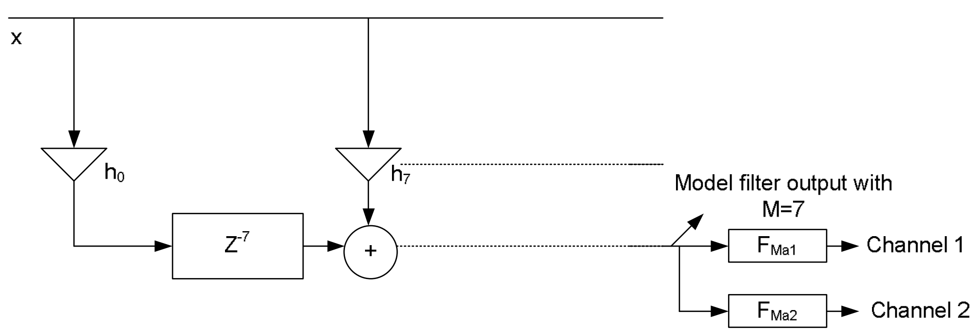

Fig. 3. The architecture of 2-channel CDFB [4].

Table 3: Selected *P* and *N* values of the filters (F_{a1}, F_{Ma1} and F_{Ma2}).

Filters used	Order (N)	P	Number of SPT terms
Model filter F_{a1}	326	1	103
Masking filter F_{Ma1}	35	1	31
Masking filter F_{Ma2}	31	3	31

For the comparison purpose, we designed various multiplier-less CDFB designs using Direct CSD truncated design, ICGA [6], ICDE [7]. In direct CSD truncation method, the model (F_{a1}) and masking filters F_{Ma1} and F_{Ma2} are truncated with 16, 11 and 10 bit CSD values respectively where one bit is used to represent the integer value and the remaining bits used to represent the fraction value. In multiplier-less CDFB designs using ICGA [6] and ICDE [7], the direct CSD truncated values are taken as the initial solution and the solution is further optimized with certain procedural steps.

The comparison description of APBR and ASBA values of the channels 1 and 2 are shown in Tables 4 and 5 respectively. It can be seen that the APBR and ASBA values of channels offered by the proposed design are closer to those values of channels designed using CC and offering better values than the rest of the compared designs.

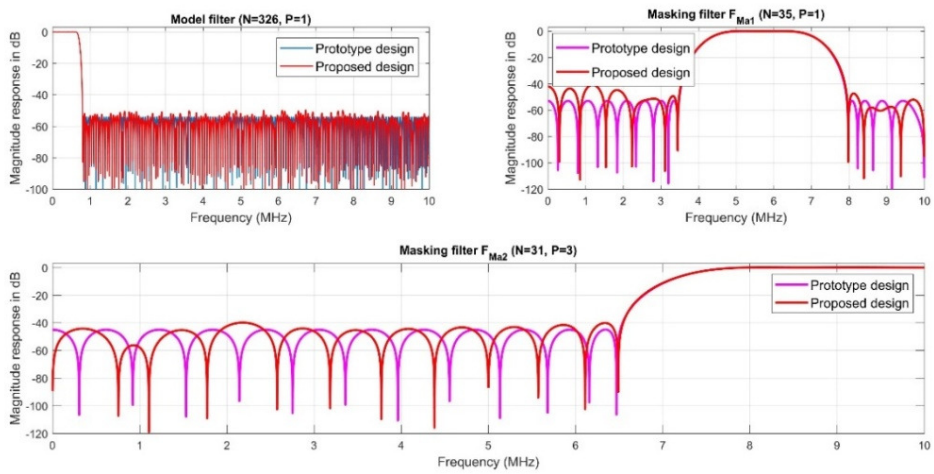

Fig. 4. The magnitude response of the model and masking filters.

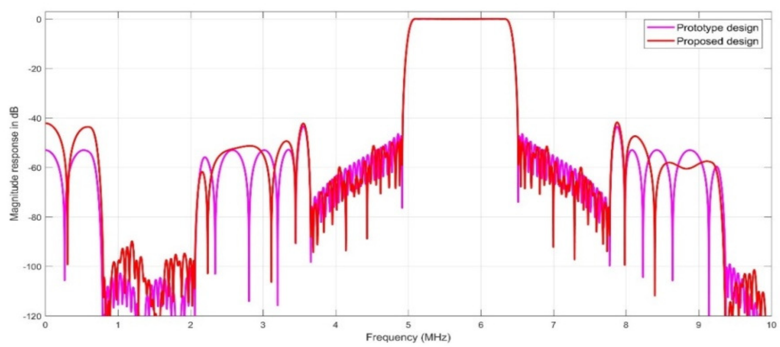

Fig. 5. The magnitude response of CDMA channel 1.

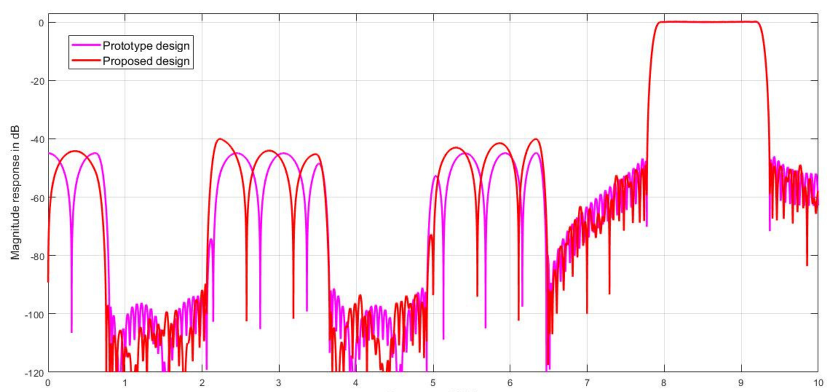

Fig. 6. The magnitude response of CDMA channel 2.

Table 4: Comparison summary of the APBR and ASBA of CDMA channel 1.

Optimization methods	ASBA1 (dB)	APBR (dB)	ASBA2 (dB)
CC	40.8946	0.1012	40.5956
CSD	37.7396	0.1476	40.9203
ICGA [6]	36.9769	0.1026	38.6724
ICDE [7]	36.9614	0.1377	40.1708
Proposed method	42.1352	0.0843	41.7655

Table 5: Comparison summary of the APBR and ASBA of CDMA channel 2.

Optimization methods	ASBA1 (dB)	APBR (dB)	ASBA2 (dB)
CC	41.2194	0.1437	40.8802
CSD	39.7724	0.2213	40.6147
ICGA [6]	38.9600	0.2363	38.5718
ICDE [7]	38.4175	0.1833	40.2656
Proposed method	40.0496	0.1300	48.3283

Table 6 shows the runtime of 2-channel multiplier-less CDFB design. Here runtime is obtained as the sum of the runtimes of the individual filter designs F_{a1}, F_{Ma1} and F_{Ma2}. The runtime of the proposed design is lower than the designs using costly optimization techniques like ICGA [6] and ICDE [7].

Table 6: Comparison overview of 2–channel CDFB run time.

Optimization methods	Run time (seconds)
ICGA [6]	423.11
ICDE [7]	315.69
Proposed method	77.87

Table 7 provides a description of the adder complexity of the proposed CDFB design for CDMA channels 1 and 2.

Table 7: Comparison summary of the complexity of the proposed FB.

Methods used	No. of SPTS	Adders due to SPTs	No. of structural adders	Total no. of adders
CSD	190	111	107	218
ICGA [6]	179	101	107	208
ICDE [7]	169	93	101	194
Proposed method	165	42	106	148

Here the no. of adders due to SPTs is computed as 42 while considering possible common sub-expressions among the coefficients. Considering zero valued coefficients, the no. of structural adders of the proposed FB is computed as 106 and eventually the total number of adders is obtained by summing the structural adders and adders due to SPT terms. The proposed design utilizes less number of adders over other methods as can be seen in Table 7.

IV. CONCLUSION

This paper presented multiplier-less design of low complexity (number of multiplications), highly reconfigurable CDFB. Here, the impulse response values of CDFB sub-filters are approximated by the SOPOT forms employing a modified SA method, known as the MPGBP algorithm. The design of the proposed multiplier-less CDFB is compared with various other multiplier-less CDFB designs using Direct CSD truncated method, ICGA and ICDE. The results showed that the proposed FB offers superior performance in terms of computational time and implementation complexity (number of adders) over existing ones.

V. FUTURE SCOPE

The possible further research can be obtaining multiplier-less MDFTFBs using vector SA method.

Conflict of Interest. No.

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How to cite this article: Baboji, K. and Sridevi, S. (2020). Design of Successive Approximation (SA) based Low Complexity Multiplier-less Coefficient Decimation Filter Bank for SDR Receivers. *International Journal on Emerging Technologies*, *11*(3): 959–964.