Linear Phase Filter Bank Design with Unabridged Control over Bandwidth and Center Frequency of Subbands

Sumit J. Darak[1], Xiguang Wu[2], Jacques Palicot[2] and Honggang Zhang[1]*

[1]CominLabs, Université Européenne de Bretagne (UEB) & Supélec, CS 47601, Cesson-Sévigné Cedex, France
[2]Supélec/IETR, CS 47601, Cesson-Sévigné Cedex, France.
E-mail :{Sumit.Darak, Xiguang.Wu, Jacques.Palicot, Honggang.Zhang}@supelec.fr

Abstract

This paper presents a design of linear phase reconfigurable digital filter bank that offers unabridged control over the bandwidth as well as center frequency of each subband over entire Nyquist band. The proposed filter bank is designed by deftly integrating spectral parameter approximation (SPA) technique with the modified coefficient decimation method (MCDM) and abbreviated as SPA-MCDM-FB. A simple and efficient architecture with unabridged control over the bandwidth as well as center frequency of all subbands, lower group delay and substantial savings in gate counts over other filter banks make the SPA-MCDM-FB suitable for various wireless communication applications such as channelization and spectrum sensing in multi-standard Cognitive Radios.

1. Introduction

Digital filter banks are widely used in various wireless communication, audio and signal processing applications [1, 2]. For example, channelization in Software Defined Radios involves extraction of individual radio channel(s) from the wideband input signal. Owing to the limited reconfigurability of an analog filter compared to its digital counterpart, digital filter bank is responsible for the stringent channel selection task which involves the intra-standard channel bandwidth variations, e.g. 1.25 MHz – 20 MHz in the Long Term Evolution (LTE), as well as inter-standard channel bandwidth variations, e.g. 200 kHz in the Global System for Mobile Communications (GSM), 1.25 MHz in the Code Division Multiple Access (CDMA) etc [1]. In Cognitive Radios, digital filter bank is responsible for filtering task in channelization as well spectrum sensing. For these as well as other emerging applications, digital filter bank must have linear phase and should be dynamically reconfigurable i.e. it should have complete control over the number of subbands, their bandwidths as well as center frequencies without the need of hardware re-implementation. Since filtering is the one of the most computationally intensive and power consuming task, the filter bank should be hardware-efficient in terms of area, power and delay. Efficient realization of such filter bank architectures is a challenging task and objective of the work presented in this paper.

Among the existing filter banks, the modulation approach based discrete Fourier transform filter bank (DFTFB) is widely used due to its advantages such as lower power consumption and smaller group delay [2]. The DFTFB is a uniform filter bank which means that it cannot provide distinct bandwidth subbands with arbitrary center frequency. The interpolation approach based reconfigurable fast filter bank (RFFB) [3] is a low complexity alternative to the DFTFB [2] especially for applications requiring sharp transition bandwidth (TBW). Though the RFFB [3] allows fine control over the subband bandwidth, the center frequency of subbands is fixed and the group delay is very large. The modified coefficient decimation method based filter bank (MCDM-FB) [4] provides independent and individual control over the bandwidth and the center frequency of subbands. But only discrete control is possible using the MCDM-FB due to integer MCDM factors. In brief, existing filter banks [2-4] cannot provide an unabridged control over the bandwidth and center frequency of subbands. This is because, the bandwidth of all subbands changes concurrently when the cut-off frequency of the prototype filter is changed. Therefore, the resolution of these filter banks [2-4], which is an inverse function of the subband bandwidth, is very high. The higher the resolution, the higher is the area complexity and power consumption of filter bank. Also, significant reconfiguration efforts are needed to change the resolution of filter bank.

In this paper, a new design of linear phase reconfigurable filter bank is proposed. The proposed filter bank is based on the deft integration of the spectral parameter approximation based variable digital filter (SPA-VDF) [5] with the MCDM [4] and shall be referred to as SPA-MCDM-FB. The design and architecture of SPA-MCDM-FB are discussed in detail in Sections 2 and 3, respectively. The design example and complexity comparisons, given in Section 4, show that the SPA-MCDM-FB provides unabridged control over the bandwidth and center frequency of each subband without the need of hardware re-implementation and offers substantial savings in total gate count and group delay over other filter banks. Moreover, these savings increase further with the increase in the filter bank resolution (i.e., number of subbands). The application of the SPA-MCDM-FB for joint channelization and spectrum sensing application in CRs is also presented in Section 4. Section 5 has our conclusion.

978-1-4673-5225-3/14/$31.00 ©2014 IEEE

2. Proposed SPA-MCDM-FB Design

The design principle of the proposed SPA-MCDM-FB is simple and straightforward. The M-subband SPA-MCDM-FB is designed by obtaining (M-1) lowpass responses with cut-off frequencies equal to falling frequency band edges of subbands. Then, by subtracting the lowpass response whose cut-off frequency is equal to the rising edge cut-off frequency of subband from the lowpass response whose cut-off frequency is equal to its falling edge cut-off frequency, the subband with any desired bandwidth and the center frequency is obtained. The first step of design of the prototype tunable lowpass filter with unabridged and independent control over the cut-off frequency of multiple lowpass responses on entire Nyquist band is an important and challenging research problem. The constraints of fixed TBW as well as minimum area, power and delay requirements make it even more difficult.

A number of linear phase variable digital filter (VDF) designs such as frequency response masking (FRM), frequency transformation and SPA are available [3-6]. The FRM-VDF [6] have very low gate count complexity especially for sharp TBW. But, multiple responses each with the desired cut-off frequency are difficult to obtain due to multi-stage architecture consisting of the prototype and masking filters. Also, the group delay of FRM-VDF is very high. The second order frequency transformation based VDFs [3] have restricted cut-off frequency range, and architectural complexity makes it difficult to obtain multiple lowpass responses. The SPA-VDF [5] are designed using Farrow structure and have very low group delay and fewer number of variable multipliers. However, high gate count complexity and large dynamic range of the filter coefficients for wide cut-off frequency range may impose constraints when fixed-point implementation is desired. In this paper, a low complexity tunable lowpass filter is designed by combining SPA-VDF [5] and MCDM [4] and it is then extended to SPA-MCDM-FB.

All the frequency specifications are normalized with respect to half the sampling frequency. The tunable lowpass filter design is based on our observation that using the MCDM [4] and lowpass prototype filter with cut-off frequency, ω_c, three additional responses with cut-off frequencies of (π - ω_c), (0.5π - ω_c) and (0.5π + ω_c) can be obtained. The proposed approach is to replace the prototype filter with the N^{th} order lowpass prototype SPA-VDF [5], $H_\alpha\left(e^{i\omega_{cp\alpha}}\right)$, where $0.25\pi \le \omega_{cp\alpha} \le 0.5\pi$, with the TBW of TBW_d and controlling parameter, α. Finally, lowpass responses with ω_c over entire Nyquist band i.e. $\left\{\left(\frac{TBW_d}{2}\right)\pi \le \omega_c \le \left[1-\left(\frac{TBW_d}{2}\right)\right]\pi\right\}$ are obtained as:

1. $H_\alpha\left(e^{i\omega_{cp\alpha}}\right)$ provides lowpass responses in second quarter as shown in Fig. 1(a). It is denoted by $H^m_{\alpha 02}(e^{j\omega_c})$ where subscripts '0' and '2' represent D=0 and second quarter, respectively.
2. $H_\alpha\left(e^{i\omega_{cp\alpha}}\right)$ and MCDM with D =1 provide highpass responses, $H^m_{\alpha 1}(e^{i\omega_c})$, shown in Fig. 1(b). Then, lowpass responses in third quarter, $H^m_{\alpha 13}(e^{j\omega_c})$, shown in Fig. 2(c), are obtained by complementing $H^m_{\alpha 1}(z)$. Mathematically,

$$H^m_{\alpha 13}(e^{j\omega_c}) = \left(e^{-j\omega_{cp\alpha}\left(\frac{N-1}{2}\right)}\right) - H_\alpha\left(e^{j(\omega_{cp\alpha}-\pi)}\right) \quad (1)$$

3. $H_\alpha\left(e^{i\omega_{cp\alpha}}\right)$ and MCDM with D =2 provide bandpass response, $H^m_{\alpha 2}(e^{i\omega_c})$ and bandstop response, $H^m_{\alpha c2}(e^{i\omega_c})$ as shown in Fig. 1(d) and Fig. 1(e), respectively. Then, using masking filter, $H_m(e^{i\omega_c})$, shown in Fig. 1(f), lowpass responses in first quarter, $H^m_{\alpha 21}(e^{j\omega_c})$, are obtained and shown in Fig. 2(g). Mathematically,

$$H^m_{\alpha 21}(e^{j\omega_c}) = H^m_{\alpha c2}(e^{j\omega_c})H_m(e^{j0.5\pi}) \quad (2)$$

where

$$H^m_{\alpha c2}(e^{j\omega_c}) = \left\{\left(e^{-j\omega_{cp\alpha}\left(\frac{N-1}{2}\right)}\right) - \left[\frac{1}{2}\sum_{k=0}^{1} H_\alpha\left(e^{j\left(\omega_{cp\alpha}-\frac{\pi(2k+1)}{2}\right)}\right)\right]\right\} \quad (3)$$

4. Finally, the lowpass responses in fourth quarter, shown in Fig. 2(h), are obtained as,

$$H^m_{\alpha 24}(e^{j\omega_c}) = H^m_{\alpha 21}(e^{j\omega_c}) + H^m_{\alpha 2}(e^{j\omega_c})\left[e^{-j\omega_{cp\alpha}\left(\frac{N_m-1}{2}\right)}\right] \quad (4)$$

where

$$H^m_{\alpha 2}(e^{j\omega_c}) = \left\{\left[\tfrac{1}{2}\textstyle\sum_{k=0}^{1} H_\alpha\left(e^{j\left(\omega_{cp\alpha}-\frac{\pi(2k+1)}{2}\right)}\right)\right]\right\} \quad (5)$$

In this way, a new tunable lowpass prototype filter with unabridged control over the cut-off frequency on entire Nyquist band and fixed TBW of TBW_d is designed where TBW_d is the desired TBW of the SPA-MCDM-FB. Note that the proposed tunable prototype filter is not just crude combination of SPA and MCDM techniques. In fact, it is carefully designed by exploiting the architectural advantages of the Farrow structure based SPA-VDF [5] as well as exclusive multiband response capability of the MCDM [4]. Furthermore, the constraints of high gate count and large dynamic range of filter coefficients of the SPA-VDF [5] are no longer present in the proposed tunable lowpass filter since the cut-off frequency range of $H_\alpha\left(e^{i\omega_{cp\alpha}}\right)$ is four times smaller than that of [5].

3. SPA-MCDM-FB Architecture

The M-subband SPA-MCDM-FB architecture, shown in Fig. 2, consists of: 1) Prototype SPA-VDF, $H_\alpha(z)$, which has (L+1) fixed-coefficient sub-filters, $H_k(z)$, $0 \le k \le L$, each of order N, 2) M branches of variable multipliers, α ($0 \le \alpha \le 1$) and output logic units (OLU). The prototype $H_\alpha(z)$ provides variable lowpass responses with the TBW of TBW_d, $0.25\pi \le \omega_{cp\alpha} \le 0.5\pi$, the passband and stopband ripples of δ_p and δ_s, respectively. Since the MCDM with D = 2 leads to the deterioration in δ_s [4], δ_s of $H_\alpha(z)$ should be ($\delta_{sd}/2$) where δ_{sd} is the desired stopband ripple of the SPA-MCDM-FB. The control signals, *sel1_D* and *sel2_D*, select the MCDM factor D for the two branches of these sub-filters [4]. The more details of the MCDM implementation are given in [4]. The multiplexers in Fig. 2 are numbered as 1-4 and all multiplexers are controlled using single three bit control signal (not shown in Fig. 2 to maintain the clarity of figure) where two bits are reserved for 4:1 multiplexer numbered as 4 and remaining one bit controls other three 2:1 multiplexers. Each OLU consists of a fixed-coefficient masking filter, $H_m(z)$, to mask the higher frequency subband of $H^m_{\alpha c2}(z)$ as discussed in Section 2. The order, cut-off frequency and TBW of $H_m(z)$ are N_m ($N_m << N$), 0.5 and (0.5- 2*TBW_d), respectively. The numbers at the input of 4:1 multiplexer point to the quarter of Nyquist band of the corresponding lowpass responses. The Adder block receives (M-1) lowpass responses and provides M subbands with the desired bandwidths and center frequencies.

4. Design Example and Complexity Comparisons

Consider the design of M-subband filter bank with TBW_d = 0.2π, δ_{sd} = 50 dB and δ_{pd} = 0.1 dB. For these specifications, the SPA-MCDM-FB is designed with L = 5, N = 32 and N_m = 18. The illustrative frequency responses obtained using the SPA-MCDM-FB for M = 3 are shown in Fig. 3(a)-(d) and corresponding values of, α_1 and α_2, are {0.19, 0.21}, {0.19, 0.79}, {0.6, 0.39} and {0.6, 0.08}, respectively. It can be observed that the bandwidth and the center frequency of all subbands can be controlled individually and are not limited to any fixed range or set of values.

A 16x16 bit multiplier, 4:1 multiplexer, 2:1 multiplexer and 32 bit adder were synthesized on a TSMC 65*nm* process. The area in terms of total gate count is obtained by normalizing the respective cell area values obtained using Synopsys Design Compiler by that of a 2:1 NAND gate from the same library and summing the gate counts of all the components. The plot of total gate count vs. the number of subbands, M, is shown in Fig. 4. It indicates that the SPA-MCDM-FB offers substantial savings over other filter banks and these savings increases further as the value of M increases. Numerically, the SPA-MCDM-FB requires 85%, 79%, 51% and 48% lower gate counts than [4], [7], [3] and [5] respectively for M = 32. Also, the group delay of the SPA-MCDM-FB is lowest and equal to 25 compared to 500, 38 and 30 in [4], [3] and [5] respectively whereas the APT-FB [7] is a non-linear phase filter bank.

Next, the gate count complexity of the SPA-MCDM-FB and the DFTFB [2] for joint channelization and spectrum sensing application in CRs is compared. Consider a typical sparse input signal of bandwidth, B_{input} = 12 MHz, consisting of multiple radio channels of bandwidths ranging from 200 kHz to 5 MHz. The task of the digital filter bank in digital front-end of CRs is to simultaneously extract the channel(s) of interest (for channelization) as well as select the desired frequency band(s) (for spectrum sensing). For B_{input} = 12 MHz, the resolution, M, of the SPA-MCDM-FB is only 5 compared to 64 in case of the DFTFB [2]. This is because, the resolution of the DFTFB

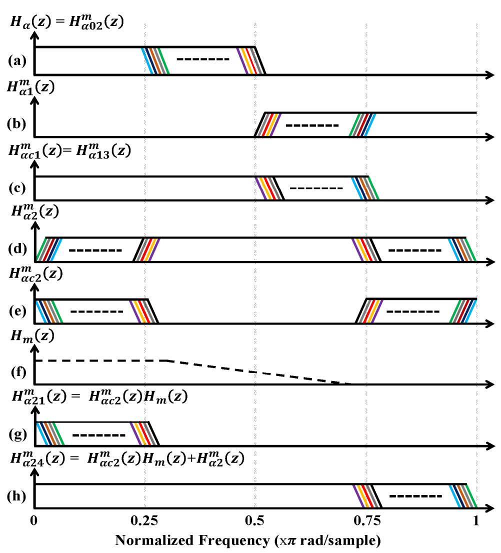

Fig. 1. Frequency responses of MCDM.

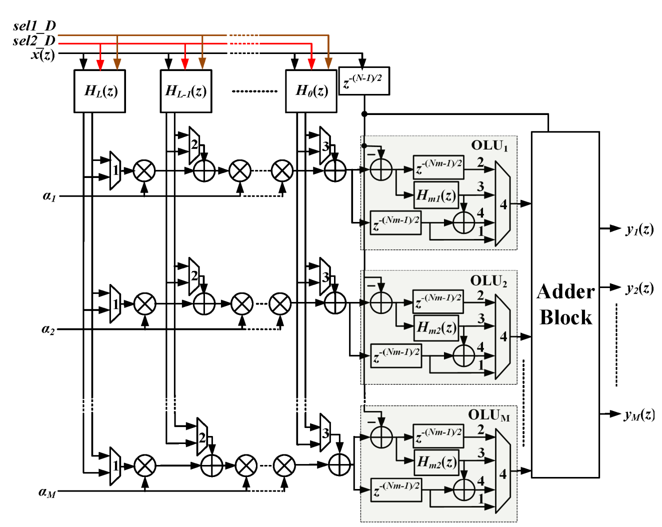

Fig. 2. Architecture of *M*-subband SPA-MCDM-FB.

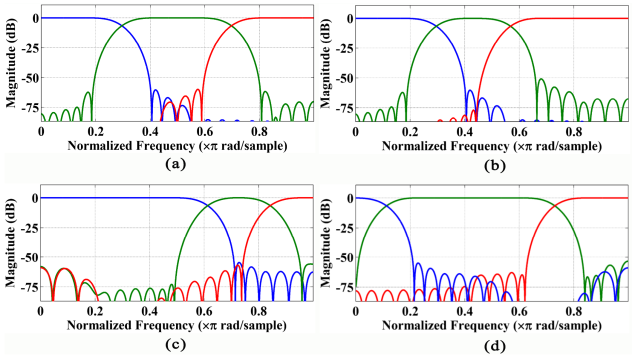

Fig. 3. Frequency responses of 3-subband SPA-MCDM-FB.

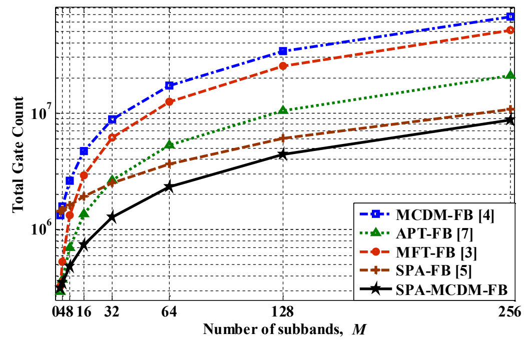

Fig. 4. Total gate count complexity vs. *M*.

is decided by the smallest channel bandwidth (i.e. 200 kHz) while resolution of the SPA-MCDM-FB is decided by the number of received channels (K) (i.e. 2 in this case). Thus, the gate count of the SPA-MCDM-FB is 381985 which is lower than the DFTFB's gate count of 819550 by 53%. Furthermore, if B_{input} is increased to 18 MHz, 90 subband DFTFB will be required which means further increase in the gate count complexity. On the other hand, gate count complexity of the SPA-MCDM-FB depends only on K and independent of B_{input}. In brief, the SPA-MCDM-FB is a low complexity alternative to the DFTFB for joint channelization and spectrum sensing in CRs when $K \leq 9$ and $K \leq 15$ which are sufficiently high for input signal with B_{input} of 12MHz and 18MHz, respectively.

5. Conclusion

In this paper, a linear phase reconfigurable filter bank designed by integrating modified coefficient decimation method with the spectral parameter approximation technique is proposed and it is termed as SPA-MCDM-FB. The design examples demonstrated that the SPA-MCDM-FB provides unabridged control over the bandwidth as well as the center frequency of each subband. It also offers substantial savings in total gate count and group delay over other filter banks. The future work will focus the efficient implementation of the SPA-MCDM-FB in FPGA via partial reconfiguration and integration with the decision making algorithms for spectrum sensing task.

6. Acknowledgment

This work has received a French state support granted to the CominLabs excellence laboratory and managed by the National Research Agency in the "Investing for the Future" program under reference No. ANR-10-LABX-07-01. The authors would also like to thank Region Bretagne, France, for its support of this work.

7. References

1. J. Wang, M. Ghosh, and K. Challapali, "Emerging Cognitive Radio Applications: A Survey," *IEEE Communications Magazine*, vol. 49, no. 3, pp. 74–81, 2011.

2. L. Pucker, "Channelization Techniques for Software Defined Radio," in *Proc. SDR Forum,* pp. 1-6, Orlando, Nov. 2003.

3. S. J. Darak, A. P. Vinod, K. G. Smitha and E. M-K. Lai, "Low Complexity Reconfigurable Fast Filter Bank for Multi-standard Wireless Receivers," *IEEE Transactions on VLSI Systems*, DOI: 10.1109/TVLSI.2013.2263813.

4. A. Ambede, K. G. Smitha and A. P. Vinod, "A New Low Complexity Uniform Filter Bank Based on the Improved Coefficient Decimation Method," *Radioengineering Journal,* vol. 22, no.1, pp. 34-43, April 2013.

5. S. S. Kidambi, "An efficient closed-form approach to the design of linear-phase FIR digital filters with variable-bandwidth characteristics," *Signal Processing* (*Elsevier*), vol. 86, no. 7, pp. 1656-1669, Oct 2005.

6.Y. J. Yu, "Design of Variable Bandedge FIR Filters with Extremely Large Bandedge Variation Range," *IEEE International Symposium on Circuits and Systems* (*ISCAS*), pp. 141-144, Rio de Janeiro, Brazil, May 2011.

7. S. J. Darak, A. P. Vinod and E. M-K. Lai, "Efficient implementation of Reconfigurable Warped Digital Filters with Variable Lowpass, Highpass, Bandpass and Bandstop Responses," *IEEE Transactions on VLSI Systems*, vol. 21, no. 6, pp. 1165-1169, July 2012.

8. A. Anandkumar, N. Michael, A. Tang, and A. Swami, "Distributed Algorithms for Learning and Cognitive Medium Access with Logarithmic Regret," *IEEE J. on Sel. Areas in Comms.*, vol. 29, no. 4, pp. 731-745, April 2011.