Design of low complexity variable digital filters and filter banks for software defined radio receivers

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**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**DESIGN OF LOW COMPLEXITY VARIABLE
DIGITAL FILTERS AND FILTER BANKS FOR
SOFTWARE DEFINED RADIO RECEIVERS**

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SCHOOL OF COMPUTER ENGINEERING

2016

Design of Low Complexity Variable Digital Filters and Filter Banks for Software Defined Radio Receivers

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SCHOOL OF COMPUTER ENGINEERING

A thesis submitted to the Nanyang Technological University

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I dedicate this work to my mother.

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Abstract

To seamlessly support the existing and upcoming wireless communication standards, the software defined radio (SDR) has been proposed as a solution. SDRs can process signals of different wireless communication standards by software reconfiguration of the same transceiver architecture, thereby reducing hardware resource utilization and associated costs. To perform multi-standard channelization, i.e., extraction of desired radio channels (frequency bands) from a wideband input signal, SDR receivers typically employ variable digital filters (VDFs) and filter banks that provide variable frequency responses by controlling a small set of parameters. VDFs and filter banks are also used in SDR based cognitive radios (CRs) to perform spectrum sensing (detection of presence/absence of licensed user signals in a wideband input frequency range) for dynamic spectrum access, to achieve opportunistic and efficient usage of the radio frequency spectrum. Realizing such VDFs and filter banks which showcase the desired attributes of high frequency response flexibility and low implementation complexity is a challenging task. To address this challenge, new algorithms and various architectures for designing VDFs and filter banks have been proposed in this thesis.

The first work presents design techniques to obtain variable frequency responses using the same set of prototype filter coefficients, along with the corresponding mathematical formulations. Two techniques are proposed – a modified coefficient decimation method (MCDM) and the improved coefficient decimation method (ICDM), the latter being a combination of the proposed MCDM and the conventional coefficient decimation method (CDM). Both the proposed techniques, i.e., MCDM and ICDM, involve selective usage of filter coefficients by performing operations such as replacing them by zeros and retaining/discarding them appropriately to obtain variable frequency responses. The ICDM is categorized into ICDM-I and ICDM-II, each comprising of two distinct coefficient decimation operations. The ICDM provides discrete control over the bandwidths and center frequencies of the subbands in the obtained frequency responses.

The second work presents the design of VDFs which can be used for channelization in SDR receivers. Three such VDFs, one each based on ICDM-I, ICDM-II and comprehensive ICDM are proposed and analysed. The comprehensive ICDM based VDF involves both ICDM-I and ICDM-II operations and features their constituent advantages while minimizing the effects of their individual limitations. The comprehensive ICDM based VDF can provide variable lowpass, highpass, bandpass, bandstop and multi-band frequency responses on-the-fly, by performing appropriate ICDM operations on the same set of prototype filter coefficients.

Similar to the design of VDFs based on ICDM-I, ICDM-II and comprehensive ICDM, the third work presents design of filter banks based on these three techniques. While the ICDM-I based filter bank can only be used for uniform channelization, the ICDM-II and comprehensive ICDM based filter banks can be used in SDR receivers for uniform as well as non-uniform channelization of signals corresponding to multiple wireless communication standards.

The fourth work presents the design of VDFs based on the combination of all pass transformation (APT) technique and the proposed ICDM. The proposed VDFs employ 1st order APT techniques along with the ICDM to provide variable lowpass, highpass, bandpass, bandstop frequency responses with unabridged control over the bandwidths and center frequencies of the constituent subbands. A spectrum sensing scheme employing VDFs based on the combination of APT and ICDM is proposed. Also, pipelined hardware implementation architectures for realizing high speed APT based VDFs are presented. These VDFs based on the combination of APT and ICDM can be used for realizing low complexity spectrum sensing schemes in SDR based CR receivers, wherein unabridged control over cutoff frequency is desired to obtain a fine sensing resolution over the entire Nyquist frequency range.

Publications

Journal Publications:

- J5. **Ambede A.**, Vinod A. P., “Design and implementation of high-speed all pass transformation based variable digital filters by breaking the dependency of operating frequency on filter order,” *IEEE Transactions on Very Large Scale Integration Systems (TVLSI)*, published online as Early Access Article in Oct. 2015.
- J4. **Ambede A.**, Shreejith S., Vinod A. P., Fahmy S., “Design and realization of variable digital filters for software defined radio channelizers using improved coefficient decimation method,” *IEEE Transactions on Circuits and Systems II (TCAS-II)*, vol. 63, no. 1, pp. 59-63, Jan. 2016.
- J3. **Ambede A.**, Smitha K. G., Vinod A. P., “Flexible low complexity uniform and non-uniform digital filter banks with high frequency resolution for multi-standard radios,” *IEEE Transactions on Very Large Scale Integration Systems (TVLSI)*, vol. 23, no. 4, pp. 631-641, Apr. 2015.
- J2. **Ambede A.**, Smitha K. G., Vinod A. P., “A low complexity uniform and non-uniform digital filter bank based on an improved coefficient decimation method for multi-standard communication channelizers,” *Circuits, Systems, and Signal Processing (CSSP)*, Springer, vol. 32, no. 6, pp. 2543-2557, Dec. 2013.
- J1. **Ambede A.**, Smitha K. G., Vinod A. P., “A new low complexity uniform filter bank based on the improved coefficient decimation method,” *Radioengineering*, vol. 22, no. 1, pp. 34-43, Apr. 2013.

Conference Publications:

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- C4. **Ambede A.**, Vinod A. P., “Design of low complexity variable digital filters using first order all pass transformation and improved coefficient decimation method,” in *Proceedings of 2nd IEEE International Conference on Devices, Circuits and Systems (ICDCS)*, pp. 1-5, Coimbatore, India, 6-8 Mar. 2014.
- C3. **Ambede A.**, Smitha K. G., Vinod A. P., “A new design method to obtain flexible and low complexity uniform filter banks,” in *Proceedings of 4th IEEE International Symposium on Electronic System Design (ISED)*, pp. 177 – 181, Singapore, 12-13 Dec. 2013.
- C2. **Ambede A.**, Smitha K. G., Vinod A. P., “An improved coefficient decimation based reconfigurable low complexity FIR channel filter for cognitive radios,” in *Proceedings of 12th IEEE International Symposium on Communications and Information Technologies (ISCIT)*, pp. 22-27, Gold Coast, Australia, 2-5 Oct. 2012.
- C1. **Ambede A.**, Smitha K. G., Vinod A. P., “A modified coefficient decimation method to realize low complexity FIR filters with enhanced frequency response flexibility and passband resolution,” in *Proceedings of 35th IEEE International Conference on Telecommunications and Signal Processing (TSP)*, pp. 658-661, Prague, Czech Republic, 3-4 Jul. 2012.

Contents

Acknowledgements.....	ii
Abstract.....	iii
Publications.....	v
Contents	vii
List of Figures	xi
List of Tables	xv
List of Abbreviations	xvi
List of Symbols	xviii
Chapter 1	1
Introduction.....	1
1.1 Introduction.....	1
1.2 Motivation.....	4
1.3 Objectives and Contributions.....	5
1.4 Outline.....	8
Chapter 2.....	9
Literature Review.....	9
2.1 Overview of Software Defined Radio.....	10
2.1.1 Software Defined Radio.....	10
2.1.2 Software Defined Radio Receiver Architecture	10
2.1.3 Digital Front End and Channelization	12
2.1.4 Spectrum Sensing in Software Defined Radio based Cognitive Radios.....	15
2.2 Variable Digital Filters for Software Defined Radio Receivers	16
2.2.1 FIR filters vs IIR filters.....	17
2.2.2 Programmable Filters.....	19
2.2.3 Fixed-coefficient Variable Digital Filters.....	21
2.3 Digital Filter Banks for Software Defined Radio Receivers.....	28
2.4 Summary	33
Chapter 3.....	34

Improved Coefficient Decimation Method	34
3.1 Introduction	34
3.2 Brief Review of Coefficient Decimation Method (CDM)	35
3.2.1 Illustrative Example	36
3.3 Proposed Modified Coefficient Decimation Method (MCDM)	38
3.3.1 Mathematical Formulation	38
3.3.2 Illustrative Example	41
3.4 Proposed Improved Coefficient Decimation Method (ICDM)	43
3.4.1 Mathematical Formulation	45
3.5 Summary	54
Chapter 4	55
Design of Flexible Low Complexity Variable Digital Filters based on Improved Coefficient Decimation Method	55
4.1 Introduction	55
4.2 Proposed ICDM-I based Variable Digital Filter	56
4.2.1 Hardware Implementation Architecture	56
4.2.2 Design Procedure	57
4.2.3 Design Example and Comparison	58
4.3 Proposed ICDM-II based Variable Digital Filter	63
4.3.1 Hardware Implementation Architecture	63
4.3.2 Design Procedure	64
4.3.3 Design Example and Comparison	65
4.4 Comprehensive ICDM based Variable Digital Filter	69
4.4.1 Hardware Implementation Architecture	70
4.4.2 Design Procedure	71
4.4.3 Illustrative Example	72
4.4.4 Channelization Design Example	76
4.4.5 FPGA Implementation Results	78
4.5 Summary	81
Chapter 5	82
Design of Uniform and Non-Uniform Filter Banks using Improved Coefficient Decimation Method	82

5.1	Introduction.....	82
5.2	Proposed ICDM-I based Filter Bank	83
5.2.1	Design Procedure	83
5.2.2	Design Example and Comparison.....	84
5.2.3	FPGA Implementation Results	89
5.3	Proposed ICDM-II based Filter Bank	91
5.3.1	Hardware Implementation Architecture	92
5.3.2	Design Procedure	93
5.3.3	Design Example and Comparison.....	94
5.4	Comprehensive ICDM based Filter Bank.....	101
5.4.1	Design Procedure	102
5.4.2	Design Example and Comparison.....	104
5.5	Summary	110
Chapter 6.....		112
Design of Variable Digital Filters based on All Pass Transformation and Improved Coefficient Decimation Method		112
6.1	Introduction.....	112
6.1.1	1 st order All Pass Transformation based Variable Digital Filters	113
6.1.2	2 nd order All Pass Transformation based Variable Digital Filters	115
6.2	Proposed ICDM and 1 st order All Pass Transformation based Variable Digital Filter.....	117
6.2.1	Hardware Implementation Architecture	117
6.2.2	Design Procedure	118
6.2.3	Design Example	120
6.3	Variable Digital Filter based Spectrum Sensing Scheme for Cognitive Radio Receivers.....	124
6.3.1	Hardware Implementation Architecture	125
6.3.2	Design Procedure	127
6.3.3	Design Example	129
6.4	Implementation of High Speed All Pass Transformation based Variable Digital Filters	131

6.4.1	Design Procedures for 1 st and 2 nd order All Pass Transformation based Variable Digital Filters	131
6.4.2	Design Example for 1 st and 2 nd order APT based VDFs	133
6.4.3	FPGA Implementation Results for 1 st and 2 nd order APT based VDFs.....	134
6.4.4	FPGA Implementation Results for APT-ICDM-I based VDF	138
6.5	Summary	140
Chapter 7.....		142
Conclusions and Future Work		142
7.1	Conclusions.....	142
7.2	Future Work	148
7.2.1	Improvement of Operating Speed of Proposed Filter Banks based on ICDM	148
7.2.2	Development of Spectrum Sensing Schemes using Proposed Variable Digital Filters and Filter Banks	149
7.2.3	Optimization based Design of Prototype Filters for Variable Digital Filters and Filter Banks based on the ICDM.....	150
7.2.4	Alternative Power Efficient Architectures to Realize the Proposed Variable Digital Filters and Filter Banks based on the ICDM	151
References.....		152

List of Figures

2.1	Block diagram of an ideal SDR receiver.....	11
2.2	Block diagram of a feasible SDR receiver.....	12
2.3	Transposed direct-form FIR filter implementation architecture.....	18
2.4	VDF using FRM technique: Block diagram.....	22
2.5	DFTFB: Block diagram.....	29
3.1(a)	Frequency response of the prototype filter.....	37
3.1(b)	Frequency response obtained by performing CDM-I using $M=2$	37
3.1(c)	Frequency response obtained by performing CDM-II using $M=2$	37
3.1(d)	Frequency response obtained by performing CDM-I using $M=3$	38
3.1(e)	Frequency response obtained by performing CDM-II using $M=3$	38
3.2(a)	Frequency response obtained by performing MCDM-I using $M=2$	42
3.2(b)	Frequency response obtained by performing MCDM-II using $M=2$	42
3.2(c)	Frequency response obtained by performing MCDM-I using $M=3$	42
3.2(d)	Frequency response obtained by performing MCDM-II using $M=3$	43
3.3(a)	Frequency response of the prototype filter.....	47
3.3(b)	Frequency response obtained by performing CDM-I using $M=3$	47
3.3(c)	Frequency response obtained by performing MCDM-I using $M=3$	48
3.3(d)	Frequency response obtained by performing CDM-II using $M=5$	48
3.3(e)	Frequency response obtained by performing MCDM-II using $M=5$	48
3.3(f)	Frequency response obtained by performing CDM-I using $M=6$	49
3.3(g)	Frequency response obtained by performing MCDM-I using $M=6$	49
3.4(a)	Frequency response of the prototype filter.....	51
3.4(b)	Frequency response obtained by performing CDM-II using $M=2$	51
3.4(c)	Frequency response obtained by performing CDM-II using $M=3$	52
3.4(d)	Frequency response obtained by performing MCDM-II using $M=1$	52

3.4(e)	Frequency response obtained by performing MCDM-II using $M=4$	52
3.4(f)	Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(a) from that in Figure 3.4(b).....	53
3.4(g)	Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(b) from that in Figure 3.4(c).....	53
3.4(h)	Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(d) from that in Figure 3.4(e)	53
4.1	Proposed ICDM-I based VDF: Hardware implementation architecture	56
4.2(a)	Frequency response of prototype filter	59
4.2(b)	Frequency response obtained by performing CDM-I using $M=2$	59
4.2(c)	Frequency response obtained by performing MCDM-I using $M=2$	60
4.2(d)	Frequency response obtained by performing CDM-I using $M=3$	60
4.2(e)	Frequency response obtained by performing MCDM-I using $M=3$	60
4.2(f)	Frequency response obtained by performing CDM-I using $M=4$	61
4.3	Proposed ICDM-II based VDF: Hardware implementation architecture.....	63
4.4	ICDM-II output frequency responses for $M=1$ to $M=5$	66
4.5(a)	Input signal containing Zigbee channel during (t_1-t_2) and Bluetooth channel during (t_3-t_4)	67
4.5(b)	ICDM-II frequency responses used for extracting the Zigbee and Bluetooth channels shown in Figure 4.5(a).....	67
4.6	Comprehensive ICDM based VDF: Hardware implementation architecture.....	70
4.7(a)	Frequency response of the prototype filter.....	73
4.7(b)	Lowpass frequency responses obtained by performing CDM-II using different M values.....	74
4.7(c)	Highpass frequency responses obtained by performing MCDM-II using different M values.....	74
4.7(d)	Bandpass frequency responses obtained by performing MCDM-I using $M_1=2$ and CDM-II using different M values.....	74

4.7(e)	Bandstop frequency responses obtained by performing CDM-I using $M_1=2$ and CDM-II using different M values.....	75
4.7(f)	Multi-band frequency responses obtained by performing CDM-I and CDM-II simultaneously, using non-identical M_1 and M_2	75
4.7(g)	Multi-band frequency responses obtained by performing MCDM-I and CDM-II simultaneously, using non-identical M_1 and M_2	75
4.8(a)	Input signal containing WCDMA, WiMAX and LTE channels during different time intervals.....	77
4.8(b)	Frequency responses used for extracting the three desired channels of WCDMA, WiMAX and LTE standards as shown in Figure 4.8(a)	77
4.9	Comprehensive ICDM based VDF: Pipelined hardware implementation architecture.....	79
5.1	Proposed ICDM-I based filter bank: Block diagram.....	84
5.2	Frequency response of 8-channel DFTFB from 0 to $f_{\text{samp}}/2$	85
5.3	Proposed ICDM-I based filter bank: Design example.....	86
5.4	Proposed ICDM-II based filter bank: Hardware implementation architecture.....	92
5.5	Proposed ICDM-II based filter bank: Block diagram.....	94
5.6	ICDM-II output frequency responses for $M=1$ to $M=6$	95
5.7(a)	Input signal containing two Zigbee channels.....	96
5.7(b)	ICDM-II frequency responses used for extracting the Zigbee channels shown in Figure 5.7(a)	96
5.8(a)	Input signal containing Bluetooth, WCDMA and Zigbee channels during the same time interval.....	98
5.8(b)	ICDM-II frequency responses used for extracting the Bluetooth, WCDMA and Zigbee channels shown in Figure 5.8(a)	98
5.9(a)	Input signal containing Bluetooth, Zigbee and WCDMA channels.....	105
5.9(b)	Comprehensive ICDM based filter bank: Channelization design example.....	107
6.1(a)	1 st order APT based VDF: Hardware implementation architecture.....	114

6.1(b)	1 st order all pass filter: Hardware implementation architecture.....	114
6.2(a)	2 nd order APT based VDF: Hardware implementation architecture.....	116
6.2(b)	2 nd order all pass filter: Hardware implementation architecture.....	116
6.3	APT-ICDM-I based VDF: Hardware implementation architecture.....	117
6.4(a)	Variable lowpass frequency responses obtained by performing 1 st order APT.....	121
6.4(b)	Variable highpass frequency responses obtained by performing MCDM-I with $M=1$ and 1 st order APT.....	122
6.4(c)	Variable bandpass frequency responses obtained by performing MCDM-I with $M=2$ and 1 st order APT.....	122
6.4(d)	Variable bandstop frequency responses obtained by performing CDM-I with $M=2$ and 1 st order APT.....	122
6.5(a)	Block diagram of proposed VDF based spectrum sensing scheme.....	126
6.5(b)	Hardware implementation architecture for VDF-I and VDF-II in the proposed scheme.....	126
6.6(a)	Input signal spectrum with three spectrum holes (SH1, SH2, SH3)	130
6.6(b)	Frequency response pairs obtained using VDF-I and VDF-II, which are used to detect the spectrum holes in Figure 6.6(a)	130
6.7(a)	Lowpass frequency responses obtained using VDF-I-B.....	134
6.7(b)	Bandpass frequency responses obtained using VDF-II-C.....	134
6.8(a)	1 st order APT based VDF: Pipelined hardware implementation architecture.....	136
6.8(b)	1 st order all pass filter: Pipelined hardware implementation architecture.....	136
6.9(a)	2 nd order APT based VDF: Pipelined hardware implementation architecture.....	136
6.9(b)	2 nd order all pass filter: Pipelined hardware implementation architecture.....	137
6.10	APT-ICDM-I based VDF: Pipelined hardware implementation architecture.....	139

List of Tables

2.1	Comparison of VDF design techniques.....	27
3.1	ICDM operations.....	44
4.1	Comparison of ICDM-I based VDF and CDM-I based VDF.....	62
4.2	<i>Decimation Selector</i> block operation in ICDM-II based VDF.....	64
4.3	ICDM-II based VDF: Multiplication complexity comparison.....	69
4.4	FPGA implementation results: Comprehensive ICDM based VDF designs	78
4.5	FPGA implementation results: Pipelined comprehensive ICDM based VDF designs.....	80
5.1	ICDM-I based filter bank: Multiplication complexity comparison.....	88
5.2	ICDM-I based filter bank: FPGA implementation results.....	90
5.3	ICDM-II based filter bank: Multiplication complexity comparison.....	100
5.4	Comprehensive ICDM based filter bank: Multiplication complexity comparison.....	109
6.1	APT-ICDM-I based VDF: Complexity comparison.....	123
6.2	FPGA implementation results: 1 st and 2 nd order APT based VDF designs.....	135
6.3	FPGA implementation results: Pipelined 1 st and 2 nd order APT based VDF designs.....	137
6.4	FPGA implementation results: Non-pipelined and pipelined APT-ICDM-I based VDF designs.....	139
7.1	Summary of the VDFs proposed in this thesis.....	145
7.2	Summary of the filter banks proposed in this thesis.....	147

List of Abbreviations

5G	5 th Generation mobile network technology
ADC	Analog to Digital Converter
APT	All Pass Transformation
ASIC	Application-Specific Integrated Circuit
BT	Bluetooth
CDFB	Coefficient Decimation Method based Filter Bank
CDM	Coefficient Decimation Method
CDMA	Code Division Multiple Access
CDM-I	Coefficient Decimation Method I
CDM-II	Coefficient Decimation Method II
CR	Cognitive Radio
DAC	Digital to Analog Converter
DFT	Discrete Fourier Transform
DFTFB	Discrete Fourier Transform based Filter Bank
DSP	Digital Signal Processor
FFB	Fast Filter Bank
FFT	Fast Fourier Transform
FIR	Finite Impulse Response
FPGA	Field Programmable Gate Array
FRM	Frequency Response Masking
GCD	Greatest Common Divisor
GFB	Goertzel algorithm based Filter Bank
GSM	Global System for Mobile communications
HSPA	High Speed Packet Access
ICDM	Improved Coefficient Decimation Method

ICDM-I	Improved Coefficient Decimation Method I
ICDM-II	Improved Coefficient Decimation Method II
IDFT	Inverse Discrete Fourier Transform
IIR	Infinite Impulse Response
LCM	Least Common Multiple
LTE	Long Term Evolution
MAC	Multiply-and-accumulate
MCDM	Modified Coefficient Decimation Method
MCDM-I	Modified Coefficient Decimation Method I
MCDM-II	Modified Coefficient Decimation Method II
MPRFB	Modulated Perfect Reconstruction Filter Bank
PDFB	Progressive Decimation Filter Bank
ReMB	Reconfigurable Multiplier Block
SB	Subband
SH	Spectrum Hole
SDR	Software Defined Radio
SPA	Spectral Parameter Approximation
VDF	Variable Digital Filter
WBAN	Wireless Body Area Network
WCDMA	Wideband Code Division Multiple Access
WiMAX	Worldwide Interoperability for Microwave Access
WLAN	Wireless Local Area Networks

List of Symbols

α	1 st order warping coefficient
β_2, β_3	2 nd order warping coefficients
δ_p	Peak-to-peak passband ripple
$\delta_{s(\text{prototype})}$	stopband attenuation specification of prototype filter
δ_s	Peak stopband ripple
$\delta_{s(\text{final})}$	stopband attenuation of filter obtained after coefficient decimation
$\delta_{p(\text{prototype})}$	Passband ripple specification for prototype filter
$\otimes$	Convolution operation
$A(z)$	Z-transform of 1 st order all pass filter
$B(z)$	Z-transform of 2 nd order all pass filter
$BW_{\text{prototype}}$	Passband width of prototype filter in comprehensive ICDM based filter bank
d	Difference between group delays of frequency responses that are spectrally subtracted
D	Decimation factor required in ICDM-II operations for obtaining bandwidths corresponding to different standards in comprehensive ICDM based filter bank
$D(k)$	Fourier series coefficients
$d_M(n)$	Periodic function (period $2M$) which performs coefficient retention/ sign reversal/ replacement by zeros in coefficient decimation
E	Energy value computed for a desired frequency band in the proposed VDF based spectrum sensing scheme
E_1	Energy value of filtered output signal of VDF-I in the proposed VDF based spectrum sensing scheme
E_2	Energy value of filtered output signal of VDF-II in the proposed VDF based spectrum sensing scheme

f_c	Cutoff frequency
f_{c1}	Cutoff frequency of the prototype filter
f_{c2}	Desired cutoff frequency to be obtained after 1 st order APT
f_{cl}	Lower cutoff frequency of desired frequency band
f_{cu}	Upper cutoff frequency of desired frequency band
f_p	Passband edge frequency
f_s	Stopband edge frequency
f_{samp}	Sampling frequency
$G(z)$	Z-transform of warped version of prototype filter
h	Filter coefficient
$h(n)$	Prototype filter coefficients
$h'(n)$	Modified filter coefficients after coefficient decimation
$H(e^{j\omega})$	Fourier transform of prototype filter coefficients
$H'(e^{j\omega})$	Fourier transform of modified filter coefficients after coefficient decimation
$H(z)$	Z-transform of prototype filter coefficients
$H_a(z^I)$	Interpolated $H_a(z)$
$H_a(z)$	Prototype filter in FRM
$H_c(z^I)$	Interpolated $H_c(z)$
$H_c(z)$	Complementary filter of $H_a(z)$
$H_i(z)$	Z-transform of inverse filter of $H(z)$
$H_{\text{ma}}(z)$	Masking filter for $H_a(z^I)$
$H_{\text{mc}}(z)$	Masking filter for $H_c(z^I)$
I	Interpolation factor
k	Integer ranging from 0 to $(M-1)$
L_{Mask}	Masking filter length in comprehensive ICDM based filter bank
L_{Mod}	Prototype filter length in comprehensive ICDM based filter bank
M	Coefficient decimation factor

M_1	Decimation factor for ICDM-I operations in comprehensive ICDM based VDF
M_2	Decimation factor for ICDM-II operations in comprehensive ICDM based VDF
$M_{\max}$	Maximum value of decimation factor
M_{out}	Multiplier value to regain original passband magnitude level after coefficient decimation in the comprehensive ICDM based VDF
N	Filter Order
N_{PDFB}	Order of the prototype filter required in PDFB
P	Integer valued ratio of bandwidth of coefficient decimation filter to that of the prototype filter in ICDM-II
R	Number of samples of $x(n)$
S	Select signal of multiplexer blocks
sel	Select signal of adder/subtractor blocks
$TBW_{\min}$	Minimum transition-band width specification
$TBW_{\text{prototype}}$	Transition-band width of prototype filter in comprehensive ICDM based filter bank
x	Input sample
$x(n)$	Digital signal whose energy is to be computed
$X(z)$	Input to a VDF/ filter bank
$Y(z)$	Output of a VDF/ filter bank
z^{-1}	Unit delay

Chapter 1

Introduction

1.1 Introduction

Mobile wireless communication is one of the constantly evolving and expanding fields of research in today's technology-driven world. Right from the basic text messaging services to the latest high speed internet data connectivity, mobile communication has seen tremendous growth over a short period of time. The advancements in mobile communication systems have had direct and significant impacts in the fields of healthcare, education, etc. Numerous mobile communication standards have been developed to support different services. For e.g., Global System for Mobile Communications (GSM) and Code Division Multiple Access (CDMA) support voice communication whereas standards such as Wireless Local Area Networks (WLAN), High Speed Packet Access (HSPA) and Long Term Evolution (LTE) provide data connectivity with different speeds and capacities. On the other hand, standards such as Bluetooth, ZigBee, etc. support communication between peripheral devices and the primary mobile terminals. In today's mobile communication systems, all these standards and the corresponding technologies gel together to provide a seamless user experience. Unlike the incremental improvements over previous generations that was observed during development of the four mobile cellular communication technologies thus far, the upcoming 5th generation (5G) mobile network technology is being developed as a

paradigm shift from the current state of the art [1-5]. 5G technology will showcase very high carrier frequencies with massive bandwidths, extreme base station and device densities, and unprecedented numbers of antennas [5]. The future mobile communication systems will be energy efficient, highly integrative and will feature high speed and spectrally efficient data, voice as well as peripheral device communications. Such advanced capabilities will directly affect and enhance the various technologies being used in multiple fields such as telecommunications, healthcare, wireless sensor networks in industries, etc. Thus, in the current as well as upcoming mobile communication systems, the ability to cater to multiple wireless communication standards is a necessary feature.

The radio transceivers that have been traditionally used in multi-standard communication scenarios are often implemented by having distinct receiver circuits for each of the different standards. This causes high resource utilization thus resulting in high power consumption and increased costs. To seamlessly support the existing and upcoming wireless communication standards, the software defined radio (SDR) has been proposed as a solution [6-8]. SDRs have the ability of software reconfiguration of their transceiver architecture which enables transmission/ reception of signals corresponding to multiple wireless communication standards. This ability of SDRs offers the advantages of low hardware resource utilization and thus reduced size and costs, while simultaneously catering to multiple wireless communication standards. The fundamental idea of SDR is to replace the conventional analog signal processing in radio transceivers with digital signal processing by placing the analog to digital converter (ADC) in receivers (digital to analog converters (DAC) in transmitters) as close to the antenna as possible. As SDR replaces analog signal processing with digital signal processing, it is possible to support multiple communication standards by mere software reconfiguration of the same hardware platform. The SDR technology is in fact being envisioned as an integral component of the next-generation 5G wireless communication networks [3-5].

The realization of wireless communication receivers in SDRs comprises of different tasks such as system design, radio frequency block design, intermediate frequency block design and baseband analog hardware design, digital hardware design, and software engineering [6-8]. Correspondingly, the major research topics involved in the realization of SDR receivers are -

- Analog front end design- consists of antenna and the analog filter handling the radio frequency signals.
- ADCs and DACs design - need of high speed, low power ADCs and DACs with large dynamic range for wideband communication signals.
- Digital front end design - consists of design of sample rate converters and reconfigurable channelizers that are used to extract multi-standard signals from the wideband input signals.

Efficient realization of SDRs thus involves the efficient implementation of different hardware architectures as well as corresponding software technologies.

Along with the challenges involved in the hardware and software development for supporting the advanced mobile communication systems, scarcity of the electromagnetic spectrum is another major challenge that is currently being faced [9]. The constantly growing number of both customers and services has led to huge demand for spectrum which is limited and non-growing. Spectrum utilization surveys have shown that the static spectrum allocation policies have led to heavily under-utilized spectrum [10, 11]. This is because licensed users are allotted dedicated parts of the spectrum and other users (licensed or unlicensed) are not allowed to use them even if they are unutilized by dedicated users. The SDR based cognitive radio (CR) has been proposed with the ability to achieve dynamic spectrum access based on intelligently acquired knowledge of the current spectral occupancy [12-14]. The CR has the ability to perform spectrum sensing, wherein it intelligently scans the spectrum to detect presence/ absence of licensed user signals. In case of absence of licensed user signals, the CR can help unlicensed users or other licensed users to adapt their transceiver characteristics to make opportunistic usage of the vacant spectrum. Thus, the CR technology can perform dynamic spectrum access and thereby achieve efficient spectrum utilization [15]. Advanced techniques are being researched and developed to realize CR based wireless communication systems [16].

Recent research proposes the CR as an enabling technology for achieving green radio networks [17-20]. The CRs can provide energy efficiency coupled with efficient spectrum usage thus supporting the fundamental requirements of green radio communications [18].

It is observed that dynamic spectrum management using CRs can offer up to 50% energy savings in radio networks [20]. Such significant energy savings coupled with the ability to enable dynamic spectrum management makes CRs highly feasible for current as well as future wireless communication applications.

Due to novel technologies and advanced capabilities involved in them, SDR and CR based mobile wireless communication systems form an exciting area of study and it offers numerous research opportunities.

1.2 Motivation

In the digital front end of a SDR, the analog signal processing is replaced as much as possible by digital signal processing [21]. The channelizer is the most computationally intensive part of the digital front end in SDR receivers as it needs to operate at the highest sampling rate. The channelizer is used to extract individual radio channels (frequency bands) present in the wideband input signal, from the output of the ADC. Variable digital filters (VDFs), i.e., filters whose output frequency responses can be modified by controlling a small set of parameters, and filter banks are typically used in the channelizer of SDR receivers to extract frequency channels corresponding to multiple wireless communication standards from a wideband input signal. In the SDR handsets, a VDF is used to extract a frequency channel corresponding to the wireless communication standard of operation in a given time interval, i.e., only a single channel needs to be extracted during a particular time interval. On the other hand, at the SDR base-station receivers, multiple channels corresponding to different wireless communication standards need to be simultaneously extracted using filter banks. The channels to be extracted can be of equal or unequal bandwidths and may be uniformly or non-uniformly distributed over the wideband input frequency range. Along with channelization, VDFs and filter banks are also used for spectrum sensing in SDR based CR receivers. In spectrum sensing in CR receivers, the presence and absence of licensed user signals is to be detected in the wideband input frequency range, to enable opportunistic spectrum access and allow the vacant frequency bands to be used by the unlicensed users. In filter bank based spectrum

sensing, filter banks are used to split the wideband frequency range into multiple frequency bands according to the desired resolution for user signal detection.

Various techniques have been proposed in literature to obtain VDFs as well as filter banks that can be used for channelization as well as spectrum sensing in wireless communication receivers. But these techniques and the corresponding implementation architectures need to be further enhanced and optimized to meet the stringent area, power and cost specifications of the channelizers in SDR receivers (especially in battery-operated, resource constrained portable radio handsets). Also, the conventional technique of achieving reconfigurability by switching the operation among distinct receiver blocks based on the current mode of operation is not an efficient approach from the perspective of resource utilization and power consumption in SDR receivers. Thus, designing VDFs and filter banks that provide high flexibility in obtaining the desired frequency bands along with low implementation complexity is a challenging task in the realization of SDR receivers. This forms the motivation behind the work presented in this thesis.

1.3 Objectives and Contributions

This thesis deals with the design of VDFs and filter banks which can be used for channelization and spectrum sensing in SDR receivers. VDFs can be used for extracting or sensing single channel at a time whereas filter banks can be used to sense or extract multiple channels simultaneously, corresponding to the same or different wireless communication standards. The VDFs and filter banks should be able to meet the stringent area, power and cost specifications of the channelizers in SDR receivers.

The objectives of the work presented in this thesis are -

- To propose a design methodology for obtaining variable frequency responses from a single prototype filter, using fewer number of variable parameters.
- To propose VDFs that can provide variable frequency responses which can be used for channelization as well as spectrum sensing in SDR receivers.

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- To propose filter banks for uniform as well as non-uniform multi-standard channelization in SDR receivers.
 - To achieve high frequency response flexibility in terms of bandwidths and locations of the subbands that can be obtained using the proposed VDFs and filter banks.
 - To achieve low complexity implementations of the proposed VDFs and filter banks.

The contributions of this thesis are summarized below -

i. Variable digital filters based on improved coefficient decimation method:

A modified coefficient decimation method (MCDM) consisting of two coefficient decimation operations MCDM-I and MCDM-II is proposed in Section 3.3 to obtain variable frequency responses using the same set of prototype filter coefficients [C1]. An improved coefficient decimation method (ICDM) is proposed in Section 3.4, which is a combination of the conventional coefficient decimation method (CDM) and the MCDM [J2, C2]. The ICDM provides variable lowpass, highpass, bandpass, bandstop and multi-band frequency responses using a single lowpass prototype filter. It consists of four different coefficient decimation operations namely CDM-I, CDM-II, MCDM-I and MCDM-II which are categorized into ICDM-I (combining CDM-I and MCDM-I) and ICDM-II (combining CDM-II and MCDM-II).

An ICDM-I based VDF is proposed in Section 4.2, which can be used in SDR handsets for extraction of individual channels of a single wireless communication standard. It performs ICDM-I operations to provide frequency responses having uniform bandwidth subbands with varying locations. Similarly, an ICDM-II based VDF is proposed in Section 4.3, which performs ICDM-II operations to obtain the desired subbands with varying bandwidths and locations [C2]. The ICDM-II based VDF can be used for extracting individual channels of multiple wireless communication standards present in the input signal during different time intervals. A VDF based on the comprehensive ICDM is proposed in Section 4.4 [J4]. It can perform all the ICDM operations (CDM-I, MCDM-I, CDM-II and MCDM-II) and provides frequency responses with varying subband bandwidths and locations, thus enabling multi-standard channelization. When

compared with VDFs based on the conventional CDM, the proposed VDFs based on the ICDM offer greater frequency response flexibility, lower implementation complexity and improved stopband attenuation performance.

ii. Filter banks based on improved coefficient decimation method:

An ICDM-I based filter bank is proposed in Section 5.2 for uniform channelization [J1, C3]. It performs ICDM-I operations to provide variable frequency responses from which the desired subbands can be obtained. The ICDM-I based filter bank has lower implementation complexity and twice the flexibility in terms of the possible number and locations of its subbands when compared with the discrete Fourier transform based filter bank and the conventional CDM based filter bank. An ICDM-II based filter bank is proposed in Section 5.3 for multi-standard channelization [J2]. It performs ICDM-II operations to provide variable frequency responses which can be used to obtain the desired uniform as well as non-uniform subbands. The ICDM-II based filter bank shows significant reduction in complexity and improvement in stopband and transition-band characteristics when compared with the conventional coefficient decimation method based progressive decimation filter bank. Another filter bank based on the comprehensive ICDM is proposed in Section 5.4 [J3]. It can perform all the operations of ICDM-I based filter bank as well as the ICDM-II based filter bank, while reducing the impact of design constraints involved in them. The comprehensive ICDM based filter bank is a low complexity alternative to the other relevant filter banks in literature and can be used for uniform as well as non-uniform channelization.

iii. Variable digital filters based on combination of all pass transformation and improved coefficient decimation method:

A VDF design technique based on the combination of 1st order all pass transformation (APT) and the ICDM-I is proposed in Section 6.2 [C4]. The proposed APT-ICDM-I based VDF provides variable lowpass, highpass, bandpass and bandstop frequency

responses on-the-fly, with unabridged control over the cutoff frequency in the entire Nyquist frequency range. It is a low complexity alternative to the other relevant VDFs in literature. In Section 6.3, an energy-detection based spectrum sensing scheme which employs low complexity VDFs based on the combination of 1st order APT and MCDM-I is proposed for SDR based CR receivers [C5]. The proposed scheme can be used to detect spectrum holes of varying bandwidths and locations in the entire Nyquist frequency range. In Section 6.4, modified pipelined hardware implementation architectures are proposed for all the different types of APT based VDFs, i.e., conventional 1st and 2nd order APT based VDFs as well as the proposed APT-ICDM-I based VDF [J5]. The proposed pipelined implementation architectures enable the realization of high speed APT based VDFs whose operating frequencies are independent of the prototype filter order.

1.4 Outline

This rest of this thesis is organized as follows - Chapter 2 presents a review of the basic signal processing blocks in a SDR receiver. A literature review of numerous VDFs and filter banks that can be used for channelization and spectrum sensing in SDR and CR receivers is provided along with the corresponding merits and demerits. Chapter 3 presents two coefficient decimation techniques – MCDM and ICDM which can be used to obtain variable frequency responses using the same set of prototype filter coefficients. Mathematical formulations for both the techniques are provided and the proposed coefficient decimation operations are illustrated using design examples. Based on the ICDM proposed in Chapter 3, VDFs and filter banks are presented in chapters 4 and 5 respectively. Hardware implementation architectures, generalized design procedures and design examples are provided for the all the different VDFs and filter banks proposed in chapters 4 and 5. Chapter 6 presents a VDF based on the combination of 1st order APT and the ICDM-I. An energy detector based spectrum sensing scheme that employs VDFs based on the combination of 1st order APT and MCDM-I is also presented in Chapter 6. Chapter 6 also presents pipelined hardware implementation architectures for realizing high speed APT based VDFs. Chapter 7 contains conclusions of the work presented in this thesis and some possible future work directions.

Chapter 2

Literature Review

This chapter presents a literature review of the relevant topics to provide a foundation for the various research contributions of this thesis.

An overview of software defined radios (SDRs) is presented in the initial part of the chapter along with the corresponding ideal and feasible SDR receiver architectures. A detailed description of the digital front end of SDR receivers is presented to emphasize the complexity of the channelization operation involved in it. The requirements and challenges involved in designing variable digital filters (VDFs) and filter banks for channelization are discussed. The spectrum sensing function of SDR based cognitive radios (CRs) is also introduced to explain the need and use of VDFs and filter banks for the same. A detailed literature review of the different VDFs and filter banks that can be used for channelization and spectrum sensing in SDR receivers is presented in the latter half of this chapter.

The main objective of this chapter is to provide a detailed review of the different VDF and filter bank design techniques available in literature that are relevant for SDR receivers, along with their merits and demerits.

2.1 Overview of Software Defined Radio

2.1.1 Software Defined Radio

The SDR can be defined as [21], “Software radio is a technology, thought to build flexible radio systems, multiservice, multi-standard, multi-band, reconfigurable and reprogrammable by software.” For realizing an SDR that satisfies this definition, many research issues need to be addressed. The two main tasks that need to be achieved are:

- To move (in transmitters as well as receivers) the border between analog and digital domains as much as possible towards radio frequency, by accomplishing the analog-to-digital and digital-to-analog wideband conversions as near as possible to the antenna.
- To replace application-specific integrated circuits (ASICs) with programmable digital signal processors (DSPs) or field programmable gate arrays (FPGAs) for baseband signal processing, in order to define as many radio functionalities as possible in software. The replacement of ASIC technology with DSP/FPGA paves the way for two possible horizons:
 - Software implementation of baseband functions, such as coding, modulation, channelization, equalization, and pulse shaping.
 - Programmability of the system to guarantee multi-standard operation.

2.1.2 Software Defined Radio Receiver Architecture

Figure 2.1 shows a block diagram of an ideal SDR receiver [21]. It consists of two main stages, radio frequency stage and the baseband stage as shown in Figure 2.1. The radio frequency stage consists of the antenna, anti-aliasing filter and the low noise amplifier. The anti-aliasing filter is used to prevent aliasing of the input signal during sampling by the analog to digital converter (ADC). The low noise amplifier is used to

Figure 2.1. Block diagram of an ideal SDR receiver.

amplify the input signal to the desired level. The baseband stage consists of the ADC and the digital signal processing blocks. The input signal is converted to the corresponding digital form using the ADC. The digital signal processing block then performs various tasks such as filtering, sample rate conversion, etc.

The practical realization of the ideal SDR receiver shown in Figure 2.1 faces some major challenges. The input signal to the ADC is a wideband radio frequency signal with a high dynamic range. Thus, to satisfy the Nyquist criteria of sampling the input signal with at least twice its radio frequency bandwidth range, the ADC needs to have a very high speed of operation, of the order of a few giga samples per second. Current state of the art ADCs do not support such a high speed of operation. For example, currently, the highest speed of operation available for a commercial ADC having a 16 bit resolution is 370 mega samples per second (TI ADC16DX370) [22], which is too low compared to the few giga samples per second required to process wideband input signals. Also, even if it is assumed that such high speed ADCs exist, the DSP platforms and the computational devices do not possess computational capacities that can handle such a high rate of inflow of data. Thus, it can be noted that it is not feasible to realize the ideal SDR receiver shown in Figure 2.1. It can be concluded that only partial band digitization is possible in a feasible SDR receiver in contrast to full band digitization desired for an ideal SDR. If the bandwidth of the receiver is limited, it is possible to devise a suitable antenna and to address the ADC sampling rate limitation as well as the computation load of the digital signal processing block. To achieve this bandwidth limitation, an intermediate frequency stage can be introduced between the radio frequency stage and the baseband stage of the

Figure 2.2. Block diagram of a feasible SDR receiver.

SDR receiver. The corresponding block diagram of a feasible SDR receiver is shown in Figure 2.2 [23]. The intermediate frequency stage consists of a mixer that down converts the frequency of the wideband input signal from radio frequency to intermediate frequency. This enables the use of currently available commercial ADCs at their corresponding speeds of operation.

The block diagram of the feasible SDR receiver shown in Figure 2.2 consists of two main parts, the analog front end and the digital front end. The analog front end consists of the low noise amplifier, mixer and anti-aliasing filters. It performs the tasks of radio frequency to intermediate frequency conversion and band limiting the wideband input signal to prevent aliasing. The block of the SDR receiver that comes immediately after the ADC is called as the digital front end [21]. The digital front end forms the most computationally demanding block of the receiver architecture as it operates at the highest sampling frequency. The digital front end is the main focus area of research in the work presented in this thesis. Its detailed description is given in the following section.

2.1.3 Digital Front End and Channelization

The digital front end of a SDR receiver consists of the following three main blocks as shown in Figure 2.2 [23] –

- Digital down conversion,
- Sample rate conversion,
- Channelization.

Digital down conversion forms a very important function in most communication systems. It is performed to bring the channel of interest from a higher frequency to the baseband. In the SDR receiver architecture shown in Figure 2.2, digital down conversion can be realized using relatively simple hardware as the basic channel selection task is performed in the analog domain itself in the intermediate frequency stage. When channels corresponding to multiple wireless communication standards are involved, the receiver has to handle the input signals at different sampling (symbol) rates. Correspondingly, the sample rate conversion can be adaptive to the different standards or it can be fixed to sample the input signal at a fixed sampling rate.

Channelization comprises all tasks necessary to select and extract the channel(s) of interest [24]. This includes channel filtering and de-spreading. Channel filtering is necessary to extract frequency-divided channels from the wideband input signal. Thus, the filters have to attenuate adjacent-channel interferers and have to meet the blocking characteristics of the current mode of operation. De-spreading is required for final user channel selection in the case of spread-spectrum systems such as Wideband Code Division Multiple Access (WCDMA). DSPs are not suitable to be used for channelization due to the high sampling rate requirements and hence dedicated hardware has to be designed for channelization.

Channelization is the process of extraction of single or more than one channels (frequency bands) of interest from the wideband input signal using VDFs (also known as channel filters) or filter banks. The channels to be extracted can have equal or unequal bandwidths and they may be uniformly or non-uniformly distributed in the input frequency range. In SDRs, the center frequencies of the desired channels are fixed. But in an ultimate SDR based cognitive radio (CR) [12-14], the center frequency locations of the desired channels can be variable depending on the corresponding spectrum occupancy scenario. This is because a CR has the ability to sense and detect the current spectrum

utilization, and change its behavioral and transmission characteristics dynamically. Thus, channel bandwidths as well as the center frequency locations of the desired channels are both variable parameters in CRs. Channelization is the first stage of digital signal processing where parameters such as bandwidth and sample-rate are changed according to the current mode of operation. Channelization is more precisely the down-conversion of the signal to baseband and channel filtering. For SDRs, it is desirable to realize as much of the channelization functionality as possible with digital signal processing. Along with reconfigurability to adapt to different wireless communication standards, area complexity and power consumption of the channelizer architecture should be as small as possible. The basic assumptions of channelization in SDR base-station receivers are [25]:

- The analog front end employs complex I-Q down conversion.
- The base-station should be able to receive multiple independent channels in parallel.
- All channels that are to be received can have the same bandwidth (i.e., belong to the same wireless communication standard) or can have different bandwidths (i.e. simultaneous reception of channels of different wireless communication standards).
- The bandwidth of the channels must be variable (tunable, i.e., software definition of channel bandwidth) to cater to different wireless communication standards.

The technical requirements for channelization are driven by the frequency allocation plans supported by the SDR. This can range from fixed carrier spacing with a constant radio frequency bandwidth per carrier channel as in cellular communications to variable carrier spacing and radio frequency bandwidth as in multi-standard satellite gateways. The processing devices for channelization must be capable of operating at high speed. Hence normally, front-end channelization is typically limited to FPGAs due to performance constraints in dealing with the wideband input, although back-end (baseband) processing which is performed on a per channel basis may incorporate digital signal processors or general purpose processors.

VDFs and filter banks are employed for channelization in SDR receivers to extract frequency channels corresponding to multiple wireless communication standards from a

wideband input signal. A VDF is employed in a SDR handset to extract a single channel of a relevant wireless communication standard during a particular time interval, whereas filter banks are employed in base-station receivers to simultaneously extract multiple channels of the same or different wireless communication standards. The channels to be extracted can thus be of equal or unequal bandwidths and may be uniformly or non-uniformly distributed over the wideband input frequency range. Designing VDFs and filter banks that exhibit high frequency response flexibility and high speed of operation with low implementation complexity is a challenging task.

2.1.4 Spectrum Sensing in Software Defined Radio based Cognitive Radios

SDR based CR has been proposed as a solution to target the opportunistic usage of the radio frequency spectrum [12-14]. The CRs have the ability to sense and detect the current spectrum utilization, and change their behavioral and transmission characteristics dynamically so as to achieve efficient spectrum access. An important function in CRs is spectrum sensing, wherein the presence and absence of signals of licensed users (called primary users) is to be detected in the wideband input frequency range in order to allow opportunistic access of the vacant frequency bands to the unlicensed users (called secondary users or cognitive radio users) [26].

In a typical CR, multiple radio channels corresponding to different wireless communication standards simultaneously coexist in the wideband input signal. These channels need to be accurately detected by the spectrum sensing block and the individual channels need to be extracted by the channelizer, thus necessitating the ability to perform multi-standard channelization. In CR receivers, along with channelization, VDFs and filter banks play an important role in spectrum sensing as well [27-29]. In filter bank based spectrum sensing, the wideband input frequency range is split into uniform or non-uniform subbands using filter banks and the presence of signals is then detected using techniques such as energy detection and cyclostationary feature detection. In [30], the use of filter banks for spectrum sensing applications is studied and it is shown that the filter

banks can be used for an efficient realization of energy detector based spectrum sensing. Consequently, numerous spectrum sensing schemes employing VDFs and filter banks have been proposed in literature for CR receivers [31-36]. For battery-operated portable CR receivers, VDFs and filter banks that have low hardware resource utilization, low power consumption and high frequency response flexibility are desired.

Along with channelization and spectrum sensing applications in SDR and CR receivers, VDFs and filter banks are widely used in audio engineering applications such as loudspeaker equalization, channelization in digital hearing aids [71-75]. In wireless body area networks (WBAN), multiple sensors are used to measure and transmit human body parameters wirelessly to portable WBAN terminals (or mobile phones) where they are recorded and analyzed [37-39]. Different sensors transmit their data using different wireless communication standards and frequency ranges. The WBAN terminal (or mobile phones) thus has to extract channels with variable bandwidths and locations corresponding to different sensors [40]. A CR based WBAN was proposed in [41], and it showed significant improvement in performance when compared with the non-CR based WBANs due to its ability to perform spectrum sensing. Thus, VDFs and filter banks with low complexity and high frequency response flexibility are also desired for channelization and spectrum sensing in WBANs.

In the subsequent sections, a review of the relevant VDFs and filter banks is presented.

2.2 Variable Digital Filters for Software Defined Radio Receivers

VDFs are employed as channel filters in the channelizer of a SDR receiver to extract frequency channels from a wideband input signal. As discussed in Section 2.1.3 for Channelization, the channels to be extracted may be of equal or unequal bandwidths and may be uniformly or non-uniformly distributed over the input frequency range. The channel filters have to attenuate adjacent-channel interferers and meet the blocking characteristics specified by the corresponding wireless communication standards. Due to

stringent adjacent channel attenuation specifications, higher-order VDFs that have narrow transition-bands are often required. As the order of the filter increases, the implementation complexity of the filter also increases thus increasing the area and power consumed. The problem of implementing digital filters with low area and power consumptions has therefore received significant attention in research.

2.2.1 FIR filters vs IIR filters

There are two types of digital filters - infinite impulse response (IIR) filters and finite impulse response (FIR) filters [42]. IIR filters are realized using recursive algorithms and they can offer sharper transition-band characteristics for a given number of coefficients when compared to FIR filters. A major disadvantage of IIR filters is that they are difficult to be reconfigured due to stability concerns. This is due to the finite word-length constraints, as new poles of the reconfigured filter may be located outside the unit circle, which will cause instability [42, 43]. On the other hand, FIR filters do not face the problem of instability as there are no poles involved in them. An advantage of FIR filters is that the group delay is constant, which provides the capability of obtaining a steep transition-band along with a linear phase response. However, for the same stopband and transition-band specifications, the complexity of a FIR filter is higher than its IIR counterpart due to the requirement of a higher filter order for the former than the latter. Summarizing, the FIR filters offer the following advantages over IIR filters:

- i. FIR filters are linear phase filters and therefore, they just delay the input signal, without causing any phase distortion.
- ii. FIR filters are more feasible for use than IIR filters in multi-rate digital signal processing applications. In case of decimation and interpolation based methods, the use of FIR filters allows omission of some calculations, thus enhancing the computational efficiency. On the other hand, if IIR filters are used, every output has to be individually calculated even if it is to be discarded. This is because of the feedback architecture of IIR filters wherein each output is fed back to the input, thus affecting the computation of the subsequent outputs.

- iii. FIR filters are less sensitive towards coefficient quantization when compared with IIR filters. The digital filters are implemented using finite-precision arithmetic, i.e., specific number of bits. The use of finite-precision arithmetic in IIR filters can lead to stability problems due to their feedback architecture. FIR filter implementation has no feedback involved, and thus they can be implemented with finite precision arithmetic without causing problems to stability.

Due to these advantages, FIR filters are widely used as prototype filters in the VDFs and filter banks used for channelization in wireless communication receivers. (A prototype filter is the basic filter which is subjected to appropriate mathematical operations by controlling a small set of parameters for obtaining variable frequency responses in a VDF or a filter bank.)

The transposed direct-form structure is widely used to implement FIR filters in hardware. Figure 2.3 shows a transposed direct-form structure for implementing a FIR filter of order N [42]. In Figure 2.3, $h_0, h_1, \dots, h_N$ denote the filter coefficients. For symmetric FIR filters, only half of the total number of coefficients need to be implemented if the transposed direct-form structure is used. Coefficient multiplication is the most computationally intensive operation in FIR filter implementations. Thus, as the coefficient multiplications are reduced by a factor of two, the transposed direct-form structure enables a significant reduction in the implementation complexity of FIR filters.

Numerous techniques have been proposed in literature for realizing VDFs that provide variable frequency responses using which the desired subbands can be obtained to extract channels corresponding to different wireless communication standards. These VDFs can be classified as –

Figure 2.3. Transposed direct-form FIR filter implementation architecture.

-
- i. *Programmable filters*: In these VDFs, the prototype filter coefficients are updated when the mode of operation changes, i.e., when the wireless communication standard whose channel is to be extracted changes from one standard to another standard.
 - ii. *Fixed-coefficient VDFs*: In these filters, variable frequency responses are obtained by keeping the set of prototype filter coefficients fixed and changing a small set of controlling parameters.

Different types of programmable filters and fixed-coefficient VDFs proposed in literature are reviewed in the subsequent sections.

2.2.2 Programmable Filters

The multiply-and-accumulate (MAC) operation based programmable filter [44] design is based on a fully programmable MAC based filter processor. The architecture of the filter processor consists of a data path with a single MAC unit, data and program memories, and a control unit [44]. The data path of the processor may include a 16 bit adder/subtractor, a multiplier and a 32-bit accumulator. The performance of the MAC based programmable filter is significantly affected by the delay of the data path, specifically the multiplier. A major disadvantage of the MAC based filter processor is the large hardware area and memory requirement for implementing the programmable filters.

In [45], a digit-based reconfigurable FIR filter implementation architecture is proposed. In this technique, the coefficient number and the number of non-zero digits in each coefficient are arbitrarily assigned and therefore the filter architecture is independent of the number of taps. In [45], the aim of the authors was to reduce the filter complexity by reducing coefficient precision without affecting the filter performance. But this technique has a high hardware area utilization and thus shows high power consumption.

A canonic signed digit based high speed reconfigurable FIR filter was proposed in [46]. The filter architecture proposed in [46] consists of a programmable canonic signed

digit based Booth encoding scheme and a partial product based Wallace adder tree. The final adder is a carry look ahead adder. This method achieves a high speed of operation. In [46], the only aim of the authors was to design a high speed reconfigurable filter and no consideration was given to the resource utilization and power consumption. Hence, similar to the digit-based reconfigurable filter proposed in [9], the canonic signed digit based reconfigurable filter also shows a high hardware area utilization along with high power consumption.

A reconfigurable multiplier block (ReMB) based programmable FIR filter was proposed in [47]. The ReMB consists of a coefficient memory and a general purpose multiplier. In the technique proposed in [47], the ReMB generates all the coefficient products and a multiplexer is used to select the required ones depending on the input. The authors showed that redundancy can be reduced by pushing the multiplexer deep into the multiplier block design. The resulting multiplier design can be more efficient in terms of area and computational complexity when compared with the general purpose general purpose multiplier combined with the coefficient memory. The hardware area utilization, power consumption and the operating speed of this technique are highly dependent on the filter length. In ReMB based implementation of higher order filters, high hardware area utilization and power consumption are observed along with a slow speed of operation. Thus, the programmable filter implementation technique in [47] is inappropriate for high order FIR filters.

A time-multiplexed multiple constant multiplication approach to implement programmable FIR filters was proposed in [48]. This technique takes coefficients as constants and uses graph dependence algorithms to reduce redundancy. But due to the sequential nature of graph dependence algorithms, long critical path lengths are observed which results in a low speed of operation. Also, similar to the ReMB based approach proposed in [47], area requirement of the filter architecture in [48] significantly increases with increase in the filter lengths.

The VDFs proposed in [44]-[48] provide variable frequency responses by updating the prototype filter coefficients, which thus necessitates the use of large coefficient memories. This significantly increases the complexity of these methods when higher order prototype

filters are required. Thus, the VDF realization techniques in [44]-[48] are appropriate for realizing only lower order filters. They are not suitable for realizing channel filters in wireless communication receivers, which usually require higher order filters due to the stringent adjacent channel attenuation specifications.

2.2.3 Fixed-coefficient Variable Digital Filters

The increase in complexity due to the use of variable coefficients can be reduced if the prototype filter coefficients are kept fixed and variable frequency responses are obtained by performing suitable mathematical operations on them using a small set of variable parameters, corresponding to the specifications of different wireless communication standards involved. This is the basic concept used in designing and implementing fixed-coefficient VDFs [49-69]. The fixed-coefficient VDFs can be classified as -

- i. *VDFs providing discrete control over cutoff frequency [49-56]:* These VDFs provide different types of variable frequency responses with discrete control over the bandwidths and locations of subbands. The resultant cutoff frequencies can be varied with a discrete control over the Nyquist frequency range by modifying the corresponding control parameters.
- ii. *VDFs providing unabridged control over cutoff frequency [57-69]:* These VDFs provide different types of variable frequency responses with unabridged control over the bandwidths and locations of subbands. The resultant cutoff frequencies can be varied with unabridged control over the Nyquist frequency range by modifying the corresponding control parameters.

2.2.3.1 Variable digital filters providing discrete control over cutoff frequency

In [49], a frequency response masking (FRM) technique was proposed to realize sharp transition-band FIR filters and was further studied and extended in [50-52]. In the FRM technique, the main idea is to use a set of lower order sub-filters to obtain frequency

Figure 2.4. VDF using FRM technique: Block diagram.

responses with sharp transition-bands. The block diagram of a VDF based on the FRM technique is shown in Figure 2.4. It consists of a lowpass prototype filter $H_a(z)$ having an odd filter order N . The complementary filter of the prototype filter can be obtained by subtracting the output of $H_a(z)$ from a suitably delayed version of the input. It can be represented as

$$H_c(z) = z^{-(N-1)/2} - H_a(z) \quad (2.1)$$

The prototype filter and the complementary filter are interpolated by a factor I to get the corresponding multi-band responses $H_a(z^I)$ and $H_c(z^I)$ respectively. From these interpolated frequency responses, the unwanted subbands are masked by cascading suitable low order frequency response masking filters $H_{ma}(z)$ and $H_{mc}(z)$ as shown in Figure 2.4. This operation gives the technique its name FRM. The output of the VDF shown in Figure 2.4 can be represented as

$$Y(z) = \{H_a(z^I) \cdot H_{ma}(z) + [z^{-I[(N-1)/2]} - H_a(z)] H_{mc}(z)\} X(z) \quad (2.2)$$

Different types of multi-band frequency responses are generated using different values of the interpolation factor I and appropriate masking filters are used to extract the desired subbands. By appropriate selection of the frequency response specifications (passband and stopband edge frequencies) of the prototype filter and the masking filters, a frequency response with a sharp transition-band width can be obtained. The FRM technique provides variable lowpass, highpass, bandpass and bandstop frequency responses using the same set of prototype filter coefficients. Thus, the implementation complexity of the

FRM technique is low compared to the programmable filter implementation techniques discussed in the previous section.

In the FRM technique, when high values of interpolation factor are used, the subbands in the resultant multi-band frequency responses are situated very close to each other. To extract the desired subbands individually, masking filters with narrow transition-band width and thus higher orders are required, or multiple wide transition-band width low order masking filters are cascaded. This leads to increase in the implementation complexity of the VDF in case of high order masking filters and increase the group delay values in case of cascaded masking filters.

Many approaches have been proposed to reduce the implementation complexity of FRM technique (elimination of a masking filter in [50], reduction in number of multipliers in [15], combining the FRM filter design with binary subexpression elimination technique [52]), but all these techniques suffer from the drawback of having large group delay values. Also, the FRM technique does not provide unabridged control over the bandwidth and locations of the subbands that can be obtained. It provides a discrete control over these parameters which is primarily governed by the values of the interpolation factor used. Lack of absolute control over the bandwidth and location of subbands and the design overhead involved in determining these parameters based on the interpolation factors, arithmetic operations and masking filters used are the limitations of the FRM technique.

In [53], a coefficient decimation method (CDM) was proposed to obtain low complexity and reconfigurable FIR filters. CDM involves the selective usage of filter coefficients by performing operations such as replacing them by zeros and retaining/discarding them appropriately to obtain variable frequency responses. The CDM consists of two coefficient decimation operations - coefficient decimation method I (CDM-I) and coefficient decimation method II (CDM-II) [90]. In the CDM-I operation, coefficients of a lowpass prototype filter are decimated by a factor M to obtain a multi-band frequency response having its subbands located at center frequencies given by even multiples of π/M , i.e., $2\pi k/M$ where k is an integer ranging from 0 to $(M-1)$. In the CDM-II operation, coefficient decimation of a lowpass prototype filter by M results in a lowpass frequency

response with its bandwidth M times that of the prototype filter. The resultant frequency responses after performing CDM operations can be subjected to complementary filter operation and appropriate arithmetic operations to obtain the desired subbands, followed by use of masking filters if needed. In [54], successive CDM-II operations are performed to obtain a channel filter that is used in a spectrum sensing scheme for mobile CR terminals.

It can be noted that similar to FRM, CDM provides discrete control over the bandwidth and locations of the subbands that can be obtained. The resultant cutoff frequencies and subband locations can be much easily determined in the CDM when compared with the FRM. This is because the locations and bandwidths are dependent on a single parameter M . The CDM based VDFs have the ability to extract non-uniform frequency channels corresponding to different wireless communication standards and are thus feasible for use in SDR and CR receivers. A detailed review of the CDM is presented in Chapter 3 of this thesis where modified and improved techniques to perform coefficient decimation are proposed.

Based on the combination of coefficient decimation, interpolation and FRM techniques, VDFs have been proposed in [55, 56] for channelization in SDR receivers. The VDF in [56] is an improved version of the VDF in [55] and provides a greater number of distinct frequency responses with an improved center frequency resolution. The VDFs proposed in [55, 56] retain the low implementation complexity characteristic of the above three design techniques, but they also suffer from the limitations of the respective design techniques in terms of the discrete control over bandwidths and subbands locations. Also, the extra design overhead involved in the VDFs in [55, 56] due to the combination of coefficient decimation, interpolation and FRM techniques leads to a complicated design procedure to obtain a VDF which can provide the desired subbands.

2.2.3.2 Variable digital filters providing unabridged control over cutoff frequency

The frequency transformation based VDFs were proposed in [57] and further studied and extended in [58-62]. In this design technique, a prototype filter is first implemented

using Taylor form. The sub-networks in the prototype filter are then replaced by appropriate sub-networks which can perform frequency transformations of the desired order. It is observed that cutoff frequency control on either side of the prototype filter's cutoff frequency is not possible using 1st order frequency transformation [57, 58]. Also, the transition-band width of the frequency response of the transformed filter is greater than that of the prototype filter. To overcome these drawbacks, 2nd order frequency transformation based VDF was proposed in [60].

The frequency transformation based VDFs proposed in [57-60] offer unabridged control over the cutoff frequency. But the frequency range in which the cutoff frequency can be varied is small. Also, to change the nature of the frequency response, the prototype filter coefficients need to be updated. To address these two design issues, two VDF design approaches were proposed in [61] and [62]. The VDF proposed in [61] combines the frequency transformation technique with the CDM. Variable lowpass, highpass, bandpass and bandstop frequency responses can be obtained using the VDF in [61] without updating the prototype filter coefficients. Similarly, the VDF in [62] combines frequency transformation technique with the interpolation technique to provide variable lowpass and bandpass frequency responses. The frequency range in which the cutoff frequency can be varied is greater in the VDFs proposed in [61, 62] when compared with the frequency transformation based VDFs proposed in [57-60], but it still isn't the entire Nyquist frequency range. Along with having a limited range of cutoff frequency variation, the frequency transformation based VDFs suffer from the drawback of having very high values of group delay.

Another VDF design technique which can provide unabridged control over the cutoff frequency is the spectral parameter approximation (SPA) technique. SPA based VDFs have been proposed in [63-66]. In this technique, the designed VDF is a weighted combination of fixed-coefficient FIR sub-filters. The sub-filters are designed offline with distinct cutoff frequencies and their weights are proportional to the spectral parameter. Different works propose the use of different optimization techniques such as minimax approximation, linear programming, least square, weighted least square, etc. to design the sub-filters required for realizing the desired SPA based VDF [63-66].

The SPA based VDFs provide unabridged control over the cutoff frequency using a single variable parameter. Also, the output frequency responses of SPA based VDFs have fixed transition-band width and low group delay values. But the biggest drawback of SPA based VDFs is their very high implementation complexity. The orders of the sub-filters used in the SPA based VDF depend on the desired cutoff frequency variation range. Also, for high order sub-filters, the corresponding filter coefficient values are very high. Due to this high implementation complexity problem, SPA based VDFs are usually employed for applications wherein cutoff frequency control is desired only over a small range of frequency.

Similar to the frequency transformation based VDFs [63-66], another transformation based VDF design approach is the all pass transformation (APT) based design technique [67-69]. The APT based VDF design technique was first proposed in [67]. In the APT based VDFs, every delay element in a prototype filter implementation architecture is replaced by an all pass filter structure of an appropriate order. The all pass filter coefficients are then modified on-the-fly to achieve the desired frequency responses with variable cutoff frequencies. APT based VDFs are also known as warped filters as the output frequency responses are warped versions of the prototype filter response.

Two types of APT based VDFs are typically used in various applications – 1st order APT based VDFs and 2nd order APT based VDFs [68]. Variable lowpass and highpass frequency responses with desired cutoff frequencies can thus be obtained using the same lowpass prototype filter in a 1st order APT based VDF. In case of bandpass and bandstop prototype filters, different values of the warping coefficient lead to modification in center frequencies as well as bandwidths of the resultant bandpass and bandstop frequency responses. This is because tuning of a single variable parameter allows control of only a single frequency edge in the resultant frequency responses. To obtain variable bandpass and bandstop frequency responses with independent and unabridged control over both their frequency edges, 2nd order APT can be performed which provides 2 tunable parameters that can be separately controlled. 1st and 2nd order APT based VDFs can be used together to obtain variable lowpass, highpass, bandpass and bandstop frequency responses from the same prototype filter. Based on the combination of CDM [53] and the APT technique, efficient realization of APT based VDFs was proposed in [69]. It was

shown that integration of the CDM and 1st order APT technique enhances the frequency response flexibility of the 1st order APT based VDFs.

APT based VDFs are widely used in audio signal processing applications such as loudspeaker equalization, spectral analysis using linear prediction, design of warped tunable adaptive filters, signal detection in broadband signals, hearing aids etc. [70-74]. Due to non-linear phase output frequency responses, APT based VDFs are not typically used for channelization in wireless communication receivers. But they can be used for applications such as energy detector based spectrum sensing in CR receivers.

The APT based VDFs offer unabridged control of the cutoff frequency over the entire Nyquist frequency range. Also, they involve lesser design overheads and achieve lower implementation complexities when compared with the frequency transformation based VDFs as well as SPA based VDFs. Thus, for applications such as spectrum sensing in CR receivers where fine sensing resolution is desired over the entire Nyquist frequency range, APT based VDFs are suitable candidates to be used.

Table 2.1 provides a brief summary of the different VDF design techniques proposed in literature which provide cutoff frequency control. When compared with the VDFs proposed in [49-56] which provide discrete control over cutoff frequency, the VDFs

Table 2.1. Comparison of VDF design techniques.

VDF design technique	Phase response	Group delay	Complexity	Cutoff frequency control over the Nyquist frequency range
Frequency response masking [49-52]	linear	very high	very low	discrete control over entire range
Coefficient decimation [53-56]	linear	low	low	discrete control over entire range
Frequency transformation [57-62]	linear	high	high	unabridged control but only over a limited section of the Nyquist range
Spectral parameter approximation [63-66]	linear	very low	very high	unabridged control over entire range, but used for limited section due to very high complexity
All pass transformation [67-69]	non-linear	non-linear	high	unabridged control over entire Nyquist range

proposed in [57-69] offer higher frequency response flexibilities by offering unabridged control over the cutoff frequency. On the other hand, VDFs which provide unabridged control over cutoff frequency have higher implementation complexities when compared with VDFs which provide discrete control over the cutoff frequency. The combination of both these types of VDFs to realize low complexity VDFs which can provide very high frequency response flexibility is an interesting area of research.

2.3 Digital Filter Banks for Software Defined Radio Receivers

When more than one frequency channels are to be simultaneously extracted from the wideband input signal, filter banks are employed for channelization. A filter bank can split the wideband input signal into multiple subbands at its output. When all the subbands have identical bandwidths, the filter bank is called as a uniform filter bank. On the other hand, if non-identical bandwidth subbands are achieved, the filter bank is called as a non-uniform filter bank. Numerous techniques have been proposed in literature to realize uniform as well as non-uniform filter banks.

In SDR and CR base-station receivers, filter banks can be employed for channelization as well as spectrum sensing applications. To cater to the different wireless communication standards, it is desired to have filter banks than can provide the desired subbands with varying bandwidths and locations. Along with the flexibility in achieving variable frequency responses that can used to obtain the desired subbands, low complexity implementations are desired for resource constrained SDR receivers.

Similar to the VDFs discussed in Section 2.2, a literature review of different filter banks which can be used in SDR receivers is presented in this section.

In the discrete Fourier transform (DFT) based filter bank (DFTFB), a single prototype filter is used to obtain multiple uniform subbands [75-77]. Figure 2.5 shows the block diagram of the DFTFB. In the DFTFB, a lowpass prototype filter in polyphase form is followed by the inverse discrete Fourier transform (IDFT) operation to obtain uniform subbands at the output. In polyphase DFTFB implementation, the decimator can be

Figure 2.5 DFTFB: Block diagram.

brought forward prior to the filtering operation using Noble identity and as a result the prototype filter operates at a lower sampling rate, reducing dynamic power consumption [77]. Due to use of a single prototype filter to obtain the desired subbands, the DFTFB has low implementation complexity even when a large number of channels are to be extracted. The DFTFB is widely used when multiple channels of the same wireless communication standard are to be extracted from the wideband input signal [75-77]. But it cannot perform non-uniform multi-standard channelization as it can provide only uniform bandwidth subbands. To change the nature of the subbands to extract signals of a different wireless communication standard, the entire filter bank needs to be re-designed with a different prototype filter. Also, the location of the subbands in a DFTFB is fixed with a center frequency resolution of $2\pi/M$ for an M -channel DFTFB, which results in a fixed channel stacking that limits the flexibility of the filter bank. The DFTFB therefore has a disadvantage of fixed channel stacking and an inability to perform multi-standard channelization.

To enable variable bandwidth subbands in the DFTFB, a filter bank design technique was proposed in [78] by incorporating the CDM-II operation in the prototype filter implementation. In this technique [78], CDM-II operation is used to change the bandwidth of the prototype filter's frequency response, thus eventually changing the

bandwidth of the subbands provided by the filter bank. Since the values of decimation factor are limited to discrete integer values, the filter bank proposed in [78] can provide only discrete control over the subband bandwidth. Also, the CDM-II operation has an inherent disadvantage of deterioration in stopband attenuation as well as transition-band width of the resultant frequency responses as the decimation factor value increases. To overcome this stopband attenuation and transition-band width deterioration problem, the prototype filter needs to be overdesigned with a higher filter order thus increasing the complexity of the filter bank. Also, similar to the conventional DFTFB, the filter bank in [78] also faces the issue of fixed channel stacking.

A Goertzel algorithm based filter bank (GFB) was proposed in [25] to overcome the fixed channel stacking problem of DFTFBs. In this technique, the Goertzel algorithm is used to compute the DFT and shift the frequency response of the prototype filter to any desired locations. Thus the fixed channel stacking problem is eliminated in the GFB. But similar to the DFTFB, the GFB too cannot provide non-uniform subbands. Thus, if multi-standard channelization is to be performed simultaneously using the GFB, separate filter banks are required to be designed corresponding to the different communication standards involved. Also, the GFB requires IIR filter for implementation of the GFB algorithm. Due to the non-linear phase property of the IIR filters and their stability issues in fixed-point implementations, the GFB is thus not a suitable candidate for channelization in SDR receivers.

The modulated perfect reconstruction filter bank (MPRFB) proposed in [79] can be used for multi-standard channelization. The MPRFB consists of an analysis section whose structure is similar to that of a DFTFB. This analysis section splits the wideband input signal into multiple narrow bandwidth subbands. The analysis section is followed by a synthesis section which can be used to combine an appropriate number of subbands to achieve the desired subband bandwidths and locations at the output. Thus, variable bandwidth subbands can be obtained using the MPRFB for non-uniform channelization. Thus the MPRFB overcomes the fixed channel stacking and uniform bandwidth limitation of the DFTFB. The drawback of the MPRFB is that its complexity is significantly higher than that of the DFTFB. It is approximately twice that of the DFTFB due to the presence of analysis and synthesis sections after the polyphase prototype filter, which perform

IDFT and DFT operations respectively. Also, the MPRFB can only extract channels of wireless communication standards whose channel bandwidths are integer multiples of each other. This is not always the case in multi-standard communication scenarios because the channel bandwidths of several standards are not related by integer factors. Therefore, the MPRFB is not suitable for such scenarios. Thus, due to its higher complexity and limited flexibility, the MPRFB is not an ideal candidate for multi-standard channelization in SDR receivers.

Tree structured filter banks, which allow splitting of the wideband input frequency range into a low-frequency and high-frequency bands successively at different stages are proposed in [80, 81]. The desired subband is obtained at an appropriate stage corresponding to the bandwidth of the channel to be extracted. The disadvantage of filter banks in [80, 81] is that they can only extract signals corresponding to those communication standards whose channel bandwidth are related by a factor-of-two. This constraint is embedded in these filter banks by the power-of-two subband stacking adopted in their architectures. Another disadvantage of the methods in [80, 81] is that, as the subband decomposition tree extends, the word length of the output increases linearly and finite precision multiplication would correspondingly introduce truncation error (round-off noise), which propagates along the tree. For tree structured filter banks designed with a large number of stages to achieve multiple subband bandwidths, large values of group delays and an increase in overall implementation complexity are the drawbacks involved.

If an FIR filter is interpolated by a factor I (i.e. each delay of an FIR filter is replaced by I delays), a $2I$ subband frequency response is obtained having the bandwidth and the transition-band width of each subband $1/I$ times that of the original FIR filter. These subbands can be extracted by selective frequency response masking, using suitably designed interpolated masking filters. Based on this approach, a fast filter bank (FFB) is proposed in [82] as a low complexity alternative to the DFTFB. The FFB can be used for the design of filter banks with very sharp transition-band widths. An advantage of the FFB is that its complexity does not increase significantly with the reduction in the transition-band width. Though the complexity of the FFB is significantly less than that of the DFTFB, it has the same limitations as that of the DFTFB, i.e., uniform subband

bandwidth and fixed subband center frequency locations [82, 83]. Also, the group delay of the FFB is larger than that of the DFTFB due to its multi-stage architecture. A multi-resolution FFB was proposed in [84] based on the idea of varying the interpolation values at every stage of the FFB, along with the number of stages involved. But this multi-resolution FFB also has the disadvantage of a high group delay similar to the original FFB, and only subbands that have their bandwidths related by a factor-of-two can be obtained.

Based on the FRM [49] and CDM [53], numerous different techniques have been proposed in literature to obtain uniform as well as non-uniform filter banks [85-90]. FRM based filter banks are proposed in [85] and [86]. In these filter bank design techniques, the desired uniform and non-uniform subbands in the filter bank are obtained by implementing appropriately designed FRM operations. Similar to the FRM based VDFs discussed in Section 2.2, the FRM based filter banks involve significant design efforts and have limited frequency response flexibility. In [87], a multi-stage coefficient decimation based filter bank was proposed that employs CDM-I, CDM-II and frequency responses masking filters for obtaining uniform filter banks with reconfigurable subband bandwidths. Another filter bank based on the combination of the FRM and CDM technique was proposed in [88]. In this technique, CDM-II operations are performed to achieve variable bandwidth frequency responses which are then subjected to appropriate FRM operations to obtain the desired subbands in the filter bank. In [89], a progressive decimation filter bank (PDFB) was proposed which performs multiple CDM-II operations followed by appropriate spectral subtraction of the resultant frequency responses to generate uniform as well as non-uniform subbands in the filter bank. The PDFB eliminates the reconfiguration overhead involved in the use of masking filters and needs a significantly lower design effort than the filter banks proposed in [85-88]. In [90], a CDM based filter bank (termed as CDFB) was proposed as a low complexity alternative to the other filter banks that can be used for uniform as well as non-uniform channelization in SDR receivers. In the CDFB, CDM-I as well as CDM-II operations are performed using appropriate decimation factors to obtain the resultant frequency responses from which the desired subbands can be obtained. The resultant frequency responses are then subjected to suitable spectral subtraction, complementary frequency response operation, and frequency

response masking filters to obtain uniform as well as non-uniform bandwidth subbands in the desired filter bank [90].

Due to their low complexities and high frequency response flexibilities, the CDM based uniform and non-uniform filter banks are suitable solutions to be considered for SDR receivers. This thesis presents new coefficient decimation techniques and proposes various VDFs and filter banks based on them for uniform as well as non-uniform channelization in SDR receivers. In subsequent chapters 3, 4, and 5, the proposed VDFs and filter banks are presented and compared with the relevant conventional CDM based VDFs and filter banks published in literature.

2.4 Summary

The basic signal processing blocks of SDR receivers were explained in this chapter. A detailed description of the digital front end of SDR receivers was provided along with the challenges involved in realizing the channelizer block, which is the most computationally intensive block in the digital front end. The use of VDFs and filter banks for channelization and spectrum sensing in SDR receivers was also discussed along with the corresponding requirements and challenges. A detailed literature review of the different VDFs and filter banks that can be used for channelization and spectrum sensing in SDR receivers was presented in the latter half of the chapter. Different VDFs and filter banks that can be used for uniform as well as non-uniform multi-standard channelization were reviewed. The design complexities of the different techniques and the issues involved in their realizations were also discussed. It can be concluded that the VDFs and filter banks need to be carefully designed and implemented to meet the stringent area, power and cost constraints of the channelizers in SDR receivers (especially battery-operated portable radio handsets). Also, there is significant scope for further enhancing the existing VDF and filter bank design techniques proposed in literature to make them suitable for use in SDR receivers.

Subsequent chapters will present the research contributions of this thesis.

Chapter 3

Improved Coefficient Decimation Method

3.1 Introduction

There are numerous techniques proposed in literature which provide variable frequency responses using the same set of prototype filter coefficients. Such techniques are widely used to realize variable digital filters (VDFs) and filter banks that exhibit high frequency response flexibility while ensuring low implementation complexity due to use of a single set of prototype filter coefficients. The coefficient decimation method (CDM) is one such technique that has been recently proposed. The CDM has been shown to be a low complexity alternative to other relevant techniques in literature, and efficient realizations of low complexity and flexible VDFs [53-56] as well as filter banks [87-90] have been proposed using the CDM. A new improved coefficient decimation method (ICDM) is proposed in this chapter. The ICDM is a combination of the conventional CDM and a newly proposed modified coefficient decimation method (MCDM). The ICDM can be used to obtain variable lowpass, highpass and multi-band frequency responses using a single fixed-coefficient lowpass prototype filter. The proposed ICDM provides twice the degree of resolution for center frequency selection when compared with the conventional CDM. The MCDM and the ICDM described in Sections 3.3 and 3.4 of this chapter have been published in [C1] and [J2, C2] respectively.

3.2 Brief Review of Coefficient Decimation Method (CDM)

In the conventional CDM, if the coefficients of a lowpass prototype filter are decimated by M , i.e., every M^{th} coefficient is retained and the others are replaced by zeros. This results in a filter with a multi-band uniform subband bandwidth frequency response. The center frequency locations of the subbands in the resultant frequency response are given by $2\pi k/M$, where k is an integer ranging from 0 to $(M-1)$. If $H(e^{j\omega})$ denotes the Fourier transform of the prototype filter coefficients, then the Fourier transform of the resulting filter's coefficients is given by

$$H'(e^{j\omega}) = \frac{1}{M} \sum_{k=0}^{M-1} H(e^{j(\omega - \frac{2\pi k}{M})}) \quad (3.1)$$

This operation is termed as coefficient decimation method I (abbreviated as CDM-I) and its mathematical derivation is given in [90]. For e.g., if $\{h_0, h_1, h_2, h_3, h_4, \dots, h_N\}$ represents the set of prototype filter coefficients, the resultant filter coefficients after performing CDM-I by $M=2$ are $\{h_0, 0, h_2, 0, h_4, \dots, h_N\}$. After performing the CDM-I operation by a decimation factor M , if all the retained coefficients in the resultant filter are grouped together by discarding the intermittent zeros, a lowpass frequency response is obtained with its passband and transition-band widths M times that of the prototype filter. This operation is termed as coefficient decimation method II (abbreviated as CDM-II) [90]. For the same set of prototype filter coefficients discussed above in the CDM-I example, the resultant filter coefficients after performing CDM-II by $M=2$ are $\{h_0, h_2, h_4, \dots, h_N\}$. The frequency response characteristics of the two CDM operations are summarized below -

- i. CDM-I: Provides variable multi-band frequency responses with a center frequency resolution of $2\pi/M$.
- ii. CDM-II: Provides variable lowpass frequency responses with cutoff frequencies that are M times that of the prototype filter.

3.2.1 Illustrative Example

The CDM-I and CDM-II operations are demonstrated in this section with the help of an illustrative example and the corresponding frequency responses are presented.

Figure 3.1(a) shows the frequency response of a lowpass prototype filter designed using the specifications – passband edge frequency (f_p) = 0.08, stopband edge frequency (f_s) = 0.1 (both edge frequencies normalized for the range 0 to 1, where 1 corresponds to half of sampling frequency), peak-to-peak passband ripple (δ_p) = 0.03 dB, and peak stopband ripple (δ_s) = -56 dB. The order of the prototype filter is 300. Note that all the prototype filters presented in the design/ illustrative examples in this thesis have been designed using the Matlab Filter Design and Analysis Tool [91], which uses the Parks - McClellan algorithm [92] to compute filter coefficients according to the desired frequency response specifications. Figures 3.1(b) and 3.1(c) show the frequency responses obtained after performing CDM-I and CDM-II operations respectively, using $M=2$, where M is the coefficient decimation factor. Similarly, Figures 3.1(d) and 3.1(e) show the frequency responses obtained by performing CDM-I and CDM-II operations respectively, using $M=3$.

The passband and transition-band widths of the subbands in the multi-band frequency responses obtained after performing CDM-I operation is same as that of the prototype filter. The passband and transition-band widths of the lowpass frequency responses obtained after performing CDM-II operations are M times that of the prototype filter. Also, the order of the filter obtained after performing CDM-I operation is same as that of the prototype filter (because every M^{th} coefficient is replaced by a zero), whereas the order of the filter obtained after performing CDM-II operation is $(1/M)$ times that of the prototype filter (because the zeros are removed and other coefficients are grouped together). An inherent drawback of CDM is the stopband attenuation deterioration with higher values of decimation factors (M values), which can also be observed when the frequency responses shown in Figures 3.1(b) - 3.1(e) are compared to the frequency response of the prototype filter shown in Figure 3.1 (a).

Figure 3.1(a). Frequency response of the prototype filter.

Figure 3.1(b). Frequency response obtained by performing CDM-I using $M=2$. Figure 3.1(c). Frequency response obtained by performing CDM-II using $M=2$.

Figure 3.1(d). Frequency response obtained by performing CDM-I using $M=3$.

Figure 3.1(e). Frequency response obtained by performing CDM-II using $M=3$.

3.3 Proposed Modified Coefficient Decimation Method (MCDM)

3.3.1 Mathematical Formulation

A modified coefficient decimation method (abbreviated as MCDM) is presented in this section. In the new coefficient decimation operation involved in the MCDM, if the coefficients of a lowpass prototype filter are decimated by a factor M , every M^{th} coefficient is retained and the sign of every alternate retained coefficient is reversed. All

other filter coefficients are replaced by zeros. This operation results in a filter with a multi-band uniform subband bandwidth frequency response with the center frequency locations of the subbands given by $(2k+1)\pi/M$, where k is an integer ranging from 0 to $(M-1)$. If $H(e^{j\omega})$ denotes the Fourier transform of the prototype filter coefficients, then the Fourier transform of the resulting filter's coefficients is given by

$$H'(e^{j\omega}) = \frac{1}{M} \sum_{k=0}^{M-1} H(e^{j(\omega - \frac{\pi(2k+1)}{M})}) \quad (3.2)$$

The mathematical derivation of this operation is given below.

Let the prototype filter coefficients be denoted by $h(n)$, and the modified coefficients be denoted by $h'(n)$. Let

$$h'(n) = h(n)d_M(n) \quad \text{for } n = 0, 1, 2, \dots \quad (3.3)$$

where $d_M(n)$ denotes a function that performs the operation of retaining every M^{th} filter coefficient and performing the appropriate sign changes. It can be represented as

$$\begin{aligned} d_M(n) &= 1 & \text{for } n = mM, m = 0, 2, 4, 6, \dots \\ &= -1 & \text{for } n = pM, p = 1, 3, 5, 7, \dots \\ &= 0 & \text{otherwise} \end{aligned} \quad (3.4)$$

From (3.4), it can be noted that $d_M(n)$ is a periodic function with a period $2M$. Its Fourier series expansion is given by

$$d_M(n) = \frac{1}{2M} \sum_{k=0}^{2M-1} D(k) e^{j\frac{2\pi kn}{2M}} \quad (3.5)$$

where $D(k)$ represents Fourier series coefficients which are given by

$$D(k) = \sum_{n=0}^{2M-1} d_M(n) e^{-j\frac{2\pi kn}{2M}} \quad (3.6)$$

From (3.4) and (3.6), it can be derived that $D(k) = 0, 2, 0, 2, 0, 2, \dots$ for any value of M , for $k = 0$ to $(2M-1)$. Using this result and substituting for $D(k)$, (3.5) is rewritten as

$$d_M(n) = \frac{1}{M} \sum_{k=0}^{M-1} e^{j \frac{\pi(2k+1)n}{M}} \quad (3.7)$$

The Fourier transform of the modified filter coefficients is given by

$$H'(e^{j\omega}) = \sum_{n=-\infty}^{\infty} h'(n) e^{-j\omega n} \quad (3.8)$$

Substituting (3.3) and (3.7) in (3.8),

$$H'(e^{j\omega}) = \sum_{n=-\infty}^{\infty} h(n) \left(\frac{1}{M} \sum_{k=0}^{M-1} e^{j \frac{\pi(2k+1)n}{M}} \right) e^{-j\omega n} \quad (3.9)$$

By interchanging the summations in (3.9),

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

$$\therefore H'(e^{j\omega}) = \frac{1}{M} \sum_{k=0}^{M-1} H(e^{j \left(\omega - \frac{\pi(2k+1)}{M} \right)}) \quad (3.10)$$

Thus from (3.10), it can be noted that frequency response of the modified coefficients is scaled by M and the original frequency spectrum is replicated at the locations $(2k+1)\pi/M$, where $k=0$ to $(M-1)$. This operation is termed as modified coefficient decimation method I (abbreviated as MCDM-I).

After performing MCDM-I by a decimation factor M , if all the retained coefficients in the resultant filter are grouped together by discarding the intermittent zeros, a highpass frequency response is obtained with its passband and transition-band widths M times that of the prototype filter. This operation is termed as modified coefficient decimation method II (abbreviated as MCDM-II). The MCDM-I and MCDM-II together constitute the proposed modified coefficient decimation method (MCDM) and their frequency response characteristics are summarized below -

-
- i. MCDM-I: Provides variable multi-band frequency responses with a center frequency resolution of $2\pi/M$, and
 - ii. MCDM-II: Provides variable highpass frequency responses with cutoff frequencies that are M times that of the prototype filter.

3.3.2 Illustrative Example

The MCDM-I and MCDM-II operations are demonstrated in this section with the help of an illustrative example. The same lowpass prototype filter that is described in the CDM illustrative example presented in Section 3.2.1 is used for performing MCDM operations. The frequency response of the prototype filter is as shown in Figure 3.1(a). Figures 3.2(a) and 3.2(b) show the frequency responses obtained after performing MCDM-I and MCDM-II operations on the prototype filter, using $M=2$. Similarly, Figures 3.1(c) and 3.1(d) show the frequency responses obtained by performing MCDM-I and MCDM-II operations respectively, using $M=3$.

Similar to the conventional CDM-I operation, passband and transition-band widths of the subbands in the multi-band frequency responses obtained after performing MCDM-I operations is same as that of the prototype filter. Also, the passband and transition-band widths of the highpass frequency responses obtained after performing MCDM-II operations are M times that of the prototype filter. The order of the filter obtained after performing MCDM-I operation is same as that of the prototype filter, whereas the order of the filter obtained after performing MCDM-II operation is $(1/M)$ times that of the prototype filter. Similar to the conventional CDM, it is observed that stopband attenuation performance of the frequency responses obtained after performing proposed MCDM operations deteriorates with higher values of decimation factors (M values). This can be noted from the frequency responses shown in Figures 3.2(a) - 3.2(d). This problem can be tackled by overdesigning the prototype filter taking into account the desired final frequency response specifications and maximum value of M . Still, the MCDM offers lower complexity than the conventional CDM.

Figure 3.2(a). Frequency response obtained by performing MCDM-I using $M=2$.

Figure 3.2(b). Frequency response obtained by performing MCDM-II using $M=2$.

Figure 3.2(c). Frequency response obtained by performing MCDM-I using $M=3$.

Figure. 3.2(d). Frequency response obtained by performing MCDM-II using $M=3$.

3.4 Proposed Improved Coefficient Decimation Method (ICDM)

A combination of the MCDM proposed in Section 3.3 and the conventional CDM [90] can be used to obtain digital filters with a greater frequency response flexibility and superior center frequency resolution when compared with either of them. The combined method presented in this section is termed as improved coefficient decimation method (abbreviated as ICDM). In the ICDM, the combination of CDM-I and MCDM-I operations is termed as improved coefficient decimation method I (abbreviated as ICDM-I), and the combination of CDM-II and MCDM-II operations is termed as improved coefficient decimation method II (abbreviated as ICDM-II) respectively. Thus the ICDM comprises of four different coefficient decimation operations -

- i. CDM-I and MCDM-I: Provide variable multi-band frequency responses.
- ii. CDM-II and MCDM-II: Provide variable lowpass and highpass frequency responses respectively.

The four different coefficient decimation operations categorized into two sub-divisions under the ICDM can be summarized as -

Table 3.1. ICDM operations.

Prototype filter coefficients		h_0	h_1	h_2	h_3	h_4	h_5	h_6	h_7	h_8	h_9	h_{10}	...	h_N
ICDM-I by $M = 2$	CDM-I by $M = 2$	h_0	0	h_2	0	h_4	0	h_6	0	h_8	0	h_{10}	...	
	MCDM-I by $M = 2$	h_0	0	$-h_2$	0	h_4	0	$-h_6$	0	h_8	0	$-h_{10}$	...	
ICDM-II by $M = 2$	CDM-II by $M = 2$	h_0	h_2	h_4	h_6	h_8	h_{10}	...						
	MCDM-II by $M = 2$	h_0	$-h_2$	h_4	$-h_6$	h_8	$-h_{10}$	...						
ICDM-I by $M = 3$	CDM-I by $M = 3$	h_0	0	0	h_3	0	0	h_6	0	0	h_9	0	...	
	MCDM-I by $M = 3$	h_0	0	0	$-h_3$	0	0	h_6	0	0	$-h_9$	0	...	
ICDM-II by $M = 3$	CDM-II by $M = 3$	h_0	h_3	h_6	h_9	...								
	MCDM-II by $M = 3$	h_0	$-h_3$	h_6	$-h_9$	...								

- i. ICDM-I = CDM-I + MCDM-I, and
- ii. ICDM-II = CDM-II + MCDM-II.

Corresponding to different ICDM operations and the values of M used, the prototype filter coefficients are appropriately retained, replaced by zeros or removed. These operations that are performed on the prototype filter coefficients are illustrated in Table 3.1. In Table 3.1, filter coefficients are represented as $h_0, h_1, h_2, \dots, h_N$ in the first row. Different ICDM operations and the corresponding resultant filter coefficients are shown in the subsequent rows.

From (3.1), it can be noted that center frequency locations that are even multiples of π/M , i.e., $\{0, 2\pi/M, 4\pi/M, 6\pi/M, \dots\}$ are achievable for subbands in the multi-band frequency responses obtained after performing CDM-I operations. Similarly, from (3.10), it can be noted that center frequency locations that are odd multiples of π/M , i.e., $\{\pi/M, 3\pi/M, 5\pi/M, 7\pi/M, \dots\}$ are achievable for subbands in the multi-band frequency responses obtained after performing MCDM-I operations. As ICDM-I is a combination of CDM-I and MCDM-I, center frequency resolution of the subbands that are obtained using ICDM-I operations is π/M , since locations that are both even (obtained using CDM-I) as

well as odd (obtained using MCDM-I) multiples of π/M are achievable for the center frequencies. Thus, when compared with the conventional CDM-I [90], the proposed ICDM-I offers twice the center frequency resolution. This means that for a particular value of the decimation factor used, the number of distinct subbands that can be obtained in the ICDM-I output frequency responses is twice that of the CDM-I. The superior center frequency resolution thus corresponds to an increased frequency response flexibility in the ICDM when compared with the CDM.

3.4.1 Mathematical Formulation

3.4.1.1 Overdesign of Prototype Filter

From the CDM and MCDM illustrative examples presented in Sections 3.2.1 and 3.3.2 respectively, it can be noted that both these methods suffer from the drawback of stopband attenuation deterioration with increasing value of the decimation factors. The proposed ICDM also has this demerit as it involves the CDM and MCDM operations. This stopband attenuation deterioration problem can be overcome by overdesigning the prototype filter. The corresponding mathematical formulation is presented below -

If f_p and f_s are the desired passband and stopband edge frequencies (normalized with respect to half of the sampling frequency), δ_p and δ_s are the desired passband and stopband peak ripple specifications, then the minimum value of order of the desired finite impulse response (FIR) filter (N) can be obtained using the formula [93],

$$N = -\frac{4 \log_{10}(10\delta_p\delta_s)}{3(f_s - f_p)} - 1 \quad (3.11)$$

The deterioration in stopband attenuation is mathematically given by,

$$\delta_{s(\text{prototype})} = \frac{\delta_{s(\text{final})}}{M} \quad (3.12)$$

where $\delta_{s(\text{prototype})}$ is the stopband attenuation of the prototype filter and is $\delta_{s(\text{final})}$ is the stopband attenuation of the filter obtained after coefficient decimation by M [55].

From (3.11) and (3.12), if coefficient decimation operation by a factor M is performed on the prototype filter and the stopband attenuation of the resulting frequency response is to be kept within a desired value δ_s , the minimum order of the overdesigned prototype filter is given by

$$N = -\frac{4\log_{10}[10\delta_p(\delta_s/M)]}{3(f_s - f_p)} - 1 \quad (3.13)$$

By rearranging the terms,

$$N = \left[-\frac{4\log_{10}(10\delta_p\delta_s)}{3(f_s - f_p)} - 1 \right] + \frac{4\log_{10}M}{3(f_s - f_p)} \quad (3.14)$$

The 2nd term on the right hand side of (3.14) is the increase in the order of the overdesigned prototype filter required to compensate the stopband attenuation deterioration after performing coefficient decimation by M .

The procedure of overdesigning the prototype filter to satisfy the desired stopband response of the final filter is explained below using a design example.

Let the most stringent frequency response specifications (desired frequency response of coefficient decimated filter) for the ICDM output frequency responses be $f_p = 0.08$, $f_s = 0.1$, $\delta_p = 0.03$ dB, and $\delta_s = -56$ dB. Assuming $M=6$ as the maximum decimation factor to be used, order of the required lowpass prototype filter is chosen as 352 using (3.14) and the corresponding filter coefficients are obtained. Figure 3.3(a) shows the frequency response of the prototype filter. Figures 3.3(b) and 3.3(c) show the frequency responses obtained after performing CDM-I and MCDM-I operations on the prototype filter shown in Figure 3.3(a), using $M=3$. Figures 3.3(d) and 3.3(e) show the frequency responses obtained by performing CDM-II and MCDM-II operations respectively, using

$M=5$. Similarly, Figures 3.3(d) and 3.3(e) show the frequency responses obtained by performing CDM-I and MCDM-I operations respectively, using $M=6$.

From Figures 3.3(a)-3.3(g), it can be noted that for all the frequency responses obtained using different ICDM operations, the stopband attenuation is larger than the desired specification of -56 dB. Thus in ICDM, the overdesign formulated in Section 3.4.1 can be effectively performed to obtain variable frequency responses adhering to the desired frequency response specifications.

Figure 3.3(a). Frequency response of the prototype filter.

Figure. 3.3(b). Frequency response obtained by performing CDM-I using $M=3$.

Figure. 3.3(c). Frequency response obtained by performing MCDM-I using $M=3$.

Figure. 3.3(d). Frequency response obtained by performing CDM-II using $M=5$.

Figure. 3.3(e). Frequency response obtained by performing MCDM-II using $M=5$.

Figure 3.3(f). Frequency response obtained by performing CDM-I using $M=6$.

Figure 3.3(g). Frequency response obtained by performing MCDM-I using $M=6$.

3.4.1.2 Spectral Subtraction and Group Delay Compensation

The spectral subtraction operation can be performed to obtain additional filter outputs using the filter outputs that are directly obtained after performing different ICDM operations. This operation is performed in the time domain in hardware, wherein the output samples of two filters obtained after performing suitable ICDM operations are subtracted to obtain the outputs corresponding to a filter with a distinct frequency response. Spectral subtraction of ICDM-II output samples can provide outputs corresponding to filters having uniform as well as non-uniform bandwidth frequency bands in their frequency responses. As discussed in Sections 3.2.1 and 3.3.2 for CDM-II

and MCDM-II operations respectively, the order of the resulting filters after ICDM-II operations varies due to the elimination of filter coefficients that are not retained in the coefficient decimation operations. As the group delays of filters with different filter orders are unequal, the group delays of the resulting filters after ICDM-II operations need to be compensated before performing spectral subtraction. If d is the difference between group delays of the two filters to be spectrally subtracted, group delay compensation can be done by adding d delay elements at the output of the filter having the lower group delay. The output samples of these filters are thereby aligned in time and can be subtracted from each other to obtain the outputs corresponding to the desired frequency response configurations. Group delay compensation is thus performed to facilitate spectral subtraction. But for performing group delay compensation to enable spectral subtraction, the group delay values of the corresponding filters have to be integer values. Thus, to ensure that the frequency responses obtained after performing ICDM-II operations can be spectrally subtracted, the order of the prototype filter should be chosen as a multiple of the least common multiple (LCM) of all the decimation factor values involved. This condition can be represented as

$$N = m \times [LCM\{M_1, M_2, M_3, \dots\}] \quad (3.15)$$

where $m = 2, 3, 4, \dots$ and $\{M_1, M_2, M_3, \dots\}$ represent the various decimation factor values that are used.

The spectral subtraction is further illustrated with the help of a design example. Figure 3.4(a) shows the frequency response of a lowpass prototype filter designed for the specifications $f_p = 0.12$, $f_s = 0.125$, $\delta_p = 0.1$ dB, $\delta_s = -73$ dB. The order of the prototype filter is computed as $N = 1680$ using (3.11) and (3.15). Figures 3.4(b) and 3.4(c) show the frequency responses obtained after performing CDM-II operations using $M=2$ and $M=3$ respectively. Similarly, Figures 3.4(d) and 3.4(e) show the frequency responses obtained after performing MCDM-II operations using $M=1$ and $M=4$ respectively. Figures 3.4(f) - 3.4(h) show a few of the frequency responses that are obtained by performing the spectral subtraction operation. The corresponding captions of figures 3.4(f) - 3.4(h) describe how these frequency responses are obtained by using appropriate frequency responses from the

ones shown in figures 3.4(a)-3.4(e). It can be noted that uniform frequency subbands having their bandwidths P times that of the prototype filter can be obtained by the spectral subtraction of the frequency responses obtained after performing ICDM-II operations successively, with the decimation factor (M) values separated by intervals of P (where P is an integer). For e.g., if uniform subbands with their bandwidths twice that of the prototype filter are to be obtained, ICDM-II operations are performed successively using $M = 2, 4, 6, 8, \dots$, and the resulting frequency responses are spectrally subtracted from each other. Non-uniform frequency subbands can be obtained by performing ICDM-II operations with M values selected according to the desired bandwidths and center frequency locations, and then spectrally subtracting the resultant frequency responses appropriately.

Figure 3.4(a). Frequency response of the prototype filter.

Figure 3.4(b). Frequency response obtained by performing CDM-II using $M=2$.

Figure 3.4(c). Frequency response obtained by performing CDM-II using $M=3$.

Figure 3.4(d). Frequency response obtained by performing MCDM-II using $M=1$.

Figure 3.4(e). Frequency response obtained by performing MCDM-II using $M=4$.

Figure 3.4(f). Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(a) from that in Figure 3.4(b).

Figure 3.4(g). Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(b) from that in Figure 3.4(c).

Figure 3.4(h). Frequency response obtained by spectral subtraction of frequency response shown in Figure 3.4(d) from that in Figure 3.4(e).

3.5 Summary

A modified coefficient decimation method (MCDM) that can be used to obtain variable frequency responses using the same set of lowpass prototype filter coefficients was proposed in this chapter. The MCDM consists of two coefficient decimation operations – MCDM-I and MCDM-II. If M is the coefficient decimation factor used, MCDM-I provides variable multi-band frequency responses with a center frequency resolution of $2\pi/M$, whereas MCDM-II provides variable highpass frequency responses with cutoff frequencies that are M times that of the prototype filter. A mathematical derivation to obtain the Fourier transform of modified filter coefficients after performing MCDM-I operation was presented. The derived mathematical expression can be used to determine the center frequency locations of subbands in the multi-band frequency responses obtained after performing MCDM-I operation using different M values.

An improved coefficient decimation method (ICDM), which is a combination of the conventional coefficient decimation method (CDM) and the MCDM was also proposed in this chapter. The ICDM comprises of four different coefficient decimation operations that are classified into two sub-divisions as ICDM-I and ICDM-II. ICDM-I operations can provide variable multi-band frequency responses, whereas ICDM-II operations can provide variable lowpass and highpass frequency responses. Center frequency resolution in the multi-band frequency responses obtained after performing ICDM-I operations is π/M , which is twice that of the conventional CDM-I. The proposed ICDM thus offers greater frequency response flexibility in terms of the number of distinct frequency responses that can be obtained, when compared with the conventional CDM. Design examples to illustrate the different types of frequency responses that can be obtained using the ICDM were presented in this chapter.

Due to the high frequency response flexibility offered by the proposed ICDM, it can be used to design VDFs and filter banks for channelization as well as spectrum sensing in software defined radio receivers. Following the same, different VDFs are presented in Chapter 4 along with the corresponding design procedures, implementation architectures, design examples and comparisons with other relevant techniques in literature.

Chapter 4

Design of Flexible Low Complexity Variable Digital Filters based on Improved Coefficient Decimation Method

4.1 Introduction

Variable digital filter (VDFs) are typically used in software defined radio (SDR) handsets to perform the channelization operation, i.e., extracting the desired radio channel from the wideband input signal. The VDFs that perform channelization are also called as channel filters. In resource constrained SDR handsets, it is desired to control the bandwidths and locations of subbands in the output frequency responses provided by a VDF while ensuring low complexity of implementation and high speed of operation. Thus, VDFs that offer high flexibility in terms of the number, location and bandwidth of the subbands in their output frequency responses are required as channel filters. This is because larger the number of variable frequency responses that can be obtained, larger is the number of wireless communication standards that can be supported by the SDR handset. Based on the improved coefficient decimation method (ICDM) proposed in Chapter 3, three VDFs which can be used as channel filters in SDR receivers are presented in this chapter. The next sections present the proposed VDFs in detail along

with their generalized design procedures, implementation architectures, design examples and comparisons. The ICDM-II based VDF and the comprehensive ICDM based VDF described in Sections 4.3 and 4.4 of this chapter have been published in [C2] and [J4] respectively.

4.2 Proposed ICDM-I based Variable Digital Filter

The improved coefficient decimation method I (ICDM-I) based VDF performs coefficient decimation method I (CDM-I) and modified coefficient decimation method I (MCDM-I) operations to provide variable frequency responses. The resultant frequency responses can be algebraically operated upon or suitably frequency response masked to obtain the subbands required to perform channelization in wireless communication receivers. The ICDM-I based VDF is presented in detail in the subsequent sections.

4.2.1 Hardware Implementation Architecture

Figure 4.1 shows the hardware implementation architecture for implementing the ICDM-I operations in the proposed ICDM-I based VDF.

Figure 4.1. Proposed ICDM-I based VDF: Hardware implementation architecture.

The architecture consists of 2:1 multiplexers for retaining a filter coefficient or replacing it by zero, according to the value of M used in the ICDM-I operation. For MCDM-I operation, the sign reversal of the corresponding coefficients is performed using the adder/subtractor (*add/sub*) blocks. This is valid because $(-h) \otimes x = -(h \otimes x)$, where h is a filter coefficient, x is the input and $\otimes$ represents the convolution operation. For performing CDM-I operation, *add/sub* blocks are used as adders. To obtain the frequency response of the prototype filter, all the coefficients are retained and the *add/sub* blocks are used as adders. The ‘*Decimation Selector*’ block is used to provide select signals of multiplexers (S) as well as those of the *add/sub* blocks (sel). If $S=1$ for a multiplexer corresponding to a particular filter coefficient, that coefficient is retained and when $S=0$, that coefficient is replaced by zero. A digital logic can be developed for this purpose using the low complexity decimation selector proposed in [94], which will reduce the hardware implementation complexity of the decimation selector circuit.

From (3.1) and (3.10), it can be noted that the resulting output after performing ICDM-I operations needs to be scaled by M to obtain the original magnitude response. This can be done by multiplying the output of the filter by M as shown in the Figure 4.1, using the associative property of convolution [42], i.e., $(M \times h) \otimes x = M \times (h \otimes x)$. An architecture can be derived from the one shown in Figure 4.1 to obtain multiple ICDM-I output responses in parallel. To individually obtain the desired frequency subbands, the multi-band ICDM-I output frequency responses can be subtracted from each other and appropriate masking filters can be applied to them or their complementary frequency responses obtained using (2.1).

4.2.2 Design Procedure

A design procedure to obtain the desired frequency responses using ICDM-I based VDF is given below.

Step-1: Design a lowpass prototype filter that meets the desired passband and stopband frequency response specifications using (3.14).

Step-2: Using (3.1) and (3.10), select the appropriate values of M that will produce multi-band frequency responses containing the desired frequency subbands.

Step-3: Perform appropriate ICDM-I operations using the identified M values to obtain frequency responses containing the desired subbands.

Step-4: Obtain the desired subbands individually by performing appropriate spectral subtraction of the ICDM-I output frequency responses or their complementary frequency responses. Low complexity frequency response masking filters can also be used to extract the desired subbands from the resultant multi-band frequency responses.

4.2.3 Design Example and Comparison

4.2.3.1 Design Example

The advantages of ICDM-I based VDF over the conventional CDM-I based VDF are discussed below with the help of a design example.

Figure 4.2(a) shows the frequency response of a lowpass prototype filter designed such that $f_p = 0.09$, $f_s = 0.1$, $\delta_p = 0.1$ dB, and $\delta_s = -55$ dB. The order of the prototype filter is $N = 600$. Figures 4.2(b) and 4.2(c) show the frequency responses obtained using the prototype filter coefficients by performing CDM-I and MCDM-I operations respectively, using $M=2$. Similarly, Figures 4.2(d) and 4.2(e) show the corresponding frequency responses for $M=3$. It can be noted from the figures for $M=3$, and also inferred from (3.1) and (3.10), that frequency responses obtained using odd values of M in the CDM-I and MCDM-I operations are inverses of each other. If $h(n)$ and $h_{\text{inv}}(n)$ denote the coefficients of a finite impulse response (FIR) filter of order N and its inverse filter respectively, the inverse filter coefficients can be represented as [42],

$$h_{\text{inv}}(n) = (-1)^{(n)} h(n) \quad (4.1)$$

where $n = 0, 1, 2, \dots, N$.

From (3.1) and (3.10), it can be noted that for even values of M , a desired frequency subband can be obtained after decimating the prototype filter by $M/2$ using ICDM-I operations instead of decimating by M using the CDM-I operation. The coefficient decimation factor required in ICDM-I is only half of that in the conventional CDM-I because of the superior center frequency resolution offered by ICDM-I (i.e., π/M) when compared with CDM-I (i.e., $2\pi/M$). This can be verified from the Figures 4.2(c) and 4.2(f) which show the frequency responses obtained after performing MCDM-I operation using $M=2$ and CDM-I operation using $M=4$ respectively, both of which constitute the subband centered at 0.5.

Figure 4.2(a). Frequency response of prototype filter.

Figure 4.2(b). Frequency response obtained by performing CDM-I using $M=2$.

Figure 4.2(c). Frequency response obtained by performing MCDM-I using $M=2$.

Figure 4.2(d). Frequency response obtained by performing CDM-I using $M=3$.

Figure 4.2(e). Frequency response obtained by performing MCDM-I using $M=3$.

Figure 4.2(f). Frequency response obtained by performing CDM-I using $M=4$.

4.2.3.2 Comparative Analysis

As discussed in Section 3.4.1, an inherent drawback of coefficient decimation (conventional CDM-I as well as proposed ICDM-I) is the stopband attenuation deterioration which increases with the value of the decimation factor. This deterioration can be observed in the frequency responses shown in Figures 4.2(a) - 4.2(f), wherein the frequency response in Figure 4.2(f) for $M=4$ shows a lower stopband attenuation than the frequency response of the prototype filter in Figure 4.2(a). If ICDM-I is performed, the desired center frequencies of subbands can be obtained using a smaller decimation factor whereas a larger decimation factor would be required to obtain the same center frequencies using the CDM-I. For example, a subband centered at 0.5 can be obtained using $M=2$ in the ICDM-I (Figure 4.2(c)) whereas $M=4$ is required in the CDM-I (Figure 4.2(f)) to obtain the same subband. Thus, as the decimation factors required in ICDM-I are smaller than those required in CDM-I, the stopband attenuation deterioration is also lesser in ICDM-I compared to CDM-I. Also, as smaller value of decimation means lesser stopband attenuation deterioration, the prototype filter of ICDM-I can be designed with a lower order than that required in the conventional CDM-I, to meet the same desired stopband attenuation specification for the final frequency responses to be obtained after coefficient decimation. This in turn will reduce the complexity and hardware resource utilization of the ICDM-I based VDF, compared to that of CDM-I based VDF. For the frequency response of CDM-I operation using $M=4$ shown in Figure 4.2(f), if the subband

centered at 0.5 needs to be separated, a frequency response masking filter has to be used or the frequency response in Figure 4.2(b) has to be subtracted from the one in Figure 4.2(f). Obtaining this subband separately using MCDM-I operation as shown in Figure 4.2(c) avoids the extra computation overhead of frequency response masking or spectral subtraction operations required in the conventional CDM-I. This will reduce the overall complexity of the ICDM-I based VDF when compared with the CDM-I based VDF.

Due to the improvement in center frequency resolution from $2\pi/M$ in the conventional CDM-I to π/M in the proposed ICDM-I, the flexibility in obtaining different subbands increases significantly. Also, certain subbands that cannot be obtained using the CDM-I operation can be obtained using the ICDM-I operations. For example, to obtain a subband centered at 0.1667 using the prototype filter in Figure 4.2(a), the theoretical value of M that is required in the conventional CDM-I is 12. However, it is not possible to use $M=12$ in CDM-I as values of M beyond $\lfloor (1/f_s) \rfloor$ (i.e., $M > 10$) would lead to aliasing. On the other hand, by using the MCDM-I operation provided in the proposed ICDM-I, the abovementioned band centered at 0.1667 can be obtained using $M = 6$ (which is less than the maximum permissible alias-free value M , i.e., 10). Therefore ICDM-I can be used to obtain significantly higher number of frequency subbands for channelization and spectrum sensing applications. Table 4.1 summarizes the advantages of ICDM-I based VDF when compared with the conventional CDM-I based VDF.

Table 4.1. Comparison of ICDM-I based VDF and CDM-I based VDF.

	CDM-I based VDF [53]	Proposed ICDM-I based VDF
Prototype filter order required	Larger	Smaller
No. of coefficient multiplications (Implementation complexity)	Larger	Smaller
Center frequency resolution of subbands (flexibility)	$2\pi/M$	π/M
Stopband attenuation performance (for same order prototype filter)	Smaller stopband attenuation	Larger stopband attenuation

4.3 Proposed ICDM-II based Variable Digital Filter

The ICDM-I based VDF proposed in Section 4.2 provides only uniform bandwidth subbands, which limits its application to single standard channelization. To perform multi-standard channelization, it is necessary to obtain subbands with varying bandwidths corresponding to different wireless communication standards that are supported by the receiver. To achieve this, a new VDF is proposed in this section which performs improved coefficient decimation method II (ICDM-II) operations to obtain variable frequency responses. The proposed VDF is termed as ICDM-II based VDF. Subsequent sections present the ICDM-II based VDF in detail and demonstrate its ability to perform multi-standard channelization.

4.3.1 Hardware Implementation Architecture

Figure 4.3 shows the hardware implementation architecture for implementing the ICDM-II operations in the proposed ICDM-II based VDF. It consists of 2:1 multiplexers (mux) that are used to select the filter coefficients according to the different M values being used in the ICDM-II operations being performed. The ‘*Decimation Selector*’ block drives the select lines (S) of the 2:1 multiplexers. If $S=1$ for a multiplexer corresponding to a particular filter coefficient, that coefficient is retained and when $S=0$, that coefficient is bypassed, i.e., eliminated. This operation of the *Decimation Selector* can be further

Figure 4.3. Proposed ICDM-II based VDF: Hardware implementation architecture.

Table 4.2. *Decimation Selector* block operation in ICDM-II based VDF.

Prototype filter coefficients	h_0	h_1	h_2	h_3	h_4	h_5	h_6	h_7	h_8	h_9	h_{10}	...	h_N
S values for ICDM-II by $M = 2$	1	0	1	0	1	0	1	0	1	0	1	...	
S values for ICDM-II by $M = 3$	1	0	0	1	0	0	1	0	0	1	0	...	
S values for ICDM-II by $M = 4$	1	0	0	0	1	0	0	0	1	0	0	...	

understood from Table 4.2 which presents the S values that are used corresponding to different decimation factors in the ICDM-II operations. Similar to the ICDM-I based VDF implementation architecture shown in Figure 4.1, adder/subtractor (*add/sub*) blocks are used to perform sign reversal of the alternate retained coefficients in the ICDM-II operations being performed in the proposed VDF. The select signal (*sel*) of the *add/sub* blocks is to be set to 1 (for subtraction) or 0 (for addition) according to the desired operation. This function is also performed by the *Decimation Selector* which drives the *add/sub* blocks according to the ICDM-II operation being performed and the corresponding M value being used in it. The resulting outputs after performing the ICDM-II operations need to be scaled by M to obtain the original magnitude response. This is done by multiplying the output of the VDF by M as shown.

4.3.2 Design Procedure

The steps involved in the realization of ICDM-II based VDF for multi-standard channelization scenarios are presented in the design procedure given below.

Step-1: Fix the passband width of the lowpass prototype filter with respect to the greatest common divisor (GCD) of the channel bandwidths of the different wireless communication standards involved.

Step-2: Identify the decimation factor (M) values required to obtain the desired frequency bands using appropriate ICDM-II operations, corresponding to the different wireless communication standards and their channel distribution in the wideband input

frequency range. Compute the least common multiple (LCM) of all the identified M values. Let the maximum decimation factor identified be $M_{\max}$.

Step-3: Fix the transition-band width of the lowpass prototype filter to be $1/ M_{\max}$ times that of the minimum transition-band width specification amongst those of the different wireless communication standards.

Step-4: Using (3.14) and (3.15), compute the order (N) of the prototype filter such that it is a multiple of the LCM calculated in *Step-2*. This ensures that the group delays of all the decimated filters are integer values. Also, to meet the desired stopband attenuation specifications, compute the value of N with respect to $M_{\max}$.

Step-5: Design the lowpass prototype filter using the specifications obtained in steps 1, 3, and 4 and obtain its coefficients. Realize the ICDM-II based VDF using the hardware implementation architecture shown in Figure 4.3.

Step-6: Perform appropriate ICDM-II operations on the prototype filter coefficients using the M values identified in *Step-2* and obtain the corresponding frequency responses.

Step-7: Obtain the desired frequency bands corresponding to the different standards by spectral subtraction of the frequency responses obtained in *Step-6*.

4.3.3 Design Example and Comparison

4.3.3.1 Design Example

In this section, the ability of the proposed ICDM-II based VDF to extract frequency channels of multiple wireless communication standards is illustrated with the help of a design example. Consider a Bluetooth/ Zigbee dual standard scenario where channels corresponding to these two standards are to be extracted. The channel bandwidths of Bluetooth and Zigbee standards are 1 MHz and 4 MHz respectively. The desired passband and stopband peak ripple specifications are chosen as 0.1 dB and -65 dB respectively for Bluetooth channels, 0.1 dB and -50 dB respectively for Zigbee channels. The transition-band width specifications are chosen as 200 kHz for Bluetooth channels

Figure 4.4. ICDM-II output frequency responses for $M=1$ to $M=5$.

and 300 kHz for the Zigbee channels. The sampling frequency is 20 MHz. Following *Step-1*, the passband width of the prototype filter is chosen as 0.1 for the normalized frequency range 0 to 1, where 1 corresponds to half of sampling frequency, i.e., 10 MHz. Considering the worst case (most stringent) stopband attenuation and transition-band width specifications from the ones desired, the specifications of the prototype filter are chosen as $f_p = 0.096$, $f_s = 0.1$, $\delta_p = 0.1$ dB, $\delta_s = -65$ dB. Following *Step-3*, decimation factor values required in the ICDM-II operations for obtaining frequency responses which can be used to obtain subbands for extracting channels of the two standards (Zigbee and Bluetooth) are identified as $\{1, 2, 3, 4, 5\}$. The maximum decimation factor involved is thus 5 and the LCM of the identified M values is 60. Using (3.14) and (3.15), the order of the prototype filter is calculated as 2160. Figure 4.4 shows the frequency responses obtained after performing ICDM-II operations on the prototype filter coefficients using $M=1$ to $M=5$. The frequency responses shown in Figure 4.4 are spectrally subtracted from each other to obtain uniform frequency bands for extracting 10 non-overlapping Bluetooth channels occurring at different time instances in the input signal. Zigbee channels having any of the frequency locations $\{(0-4 \text{ MHz}), (1-5 \text{ MHz}), (2-6 \text{ MHz}), (3-7 \text{ MHz}), (4-8 \text{ MHz}), (5-9 \text{ MHz}), (6-10 \text{ MHz})\}$, are extracted by performing spectral subtraction/addition of the ICDM-II output frequency responses for different combinations of M values. For example, output frequency responses with $M=1$ and $M=5$

Figure 4.5(a). Input signal containing Zigbee channel during (t_1-t_2) and Bluetooth channel during (t_3-t_4).

Figure 4.5(b). ICDM-II frequency responses used for extracting the Zigbee and Bluetooth channels shown in Figure 4.5(a).

can be used for extracting two non-simultaneously occurring Zigbee channels located at 1-5 MHz and 6-10 MHz in the input signal.

The VDF designed using ICDM-II is functionally verified by considering an input signal containing Bluetooth and Zigbee channels as shown in Figure 4.5(a). Note that the Zigbee and Bluetooth channels do not exist simultaneously. During the time interval t_1-t_2 , when the receiver is operating in Zigbee mode, one Zigbee channel is located between 0-4 MHz and at another time interval t_3-t_4 , when the receiver is operating in Bluetooth mode,

one Bluetooth channel is located between 9-10 MHz as shown in Figure 4.5(a). The ICDM-II output frequency responses that are used to obtain the frequency bands corresponding to this channel distribution are shown in Figure 4.5(b). The frequency band used to extract the Zigbee channel is the frequency response obtained after performing coefficient decimation method II (CDM-II) operation using $M=4$. The Bluetooth channel is extracted using the frequency band obtained as the output frequency response after performing modified coefficient decimation method II (MCDM-II) operation using $M=1$. The quality of channel extraction is assessed by computing the mean squared error of the extracted signal and the original signal. For both Zigbee and Bluetooth channels, the mean squared error values are very small (of the order of 10^{-8}) indicating that the distortion caused by the ICDM-II based VDF is very low.

4.3.3.2 Complexity Analysis

The complexity of a VDF implementation is mainly dependent on the order of the prototype filter and the corresponding number of coefficient multiplications required. For the ICDM-II based VDF in the Bluetooth/Zigbee channelization design example described in this section, the order of the prototype filter used is 2160. Exploiting the symmetric property of the filter coefficients, the prototype filter can be implemented using the transposed direct-form FIR filter structure. The total number of coefficient multiplications required for the prototype filter implementation is thus $\lceil 2161/2 \rceil = 1081$. The same set of prototype filter coefficients is used to obtain different frequency responses by performing ICDM-II operations. These resultant frequency responses are used to extract channels of the different communication standards and therefore only the prototype filter's implementation accounts for the total multiplication complexity involved. Only 5 decimation factor values ($M=1$ to $M=5$) are required in the proposed ICDM-II based VDF design for the Bluetooth/Zigbee channelization design example, whereas decimation factors from $M=1$ to $M=10$ are required in the conventional CDM-II based VDF implementation. The maximum decimation factor involved in the CDM-II based VDF is 10 and the corresponding LCM of all the required M values is 2520. Using (3.14) and (3.15), the order of the prototype filter required to satisfy the same stopband

Table 4.3. ICDM-II based VDF: Multiplication complexity comparison.

	Number of multiplications required
CDM-II based VDF	2521
Proposed ICDM-II based VDF	1081

attenuation and transition-band width specifications is obtained as 5040. Using transposed direct-form FIR filter structure for the prototype filter implementation, the total number of coefficient multiplications required in the CDM-II based VDF is $\lceil 5041/2 \rceil = 2521$. Thus, a 57.12% reduction in multiplication complexity is achieved using the proposed ICDM-II based VDF when compared with a CDM-II based VDF designed for the same frequency response specifications. Table 4.3 presents this multiplication complexity comparison. The maximum decimation factor involved in the proposed ICDM-II based VDF design is half of that required in the conventional CDM-II based VDF. Thus, for a given multi-standard scenario, if the ICDM-II based VDF and the CDM-II based VDF are designed using the same prototype filter, the worst stopband attenuation and maximum transition-band width values observed in the proposed ICDM-II based VDF are half of those observed in CDM-II based VDF.

4.4 Comprehensive ICDM based Variable Digital Filter

The ICDM-I based VDF proposed in Section 4.2 can only be used for channelization of signals corresponding to a single wireless communication standard as it provides frequency responses with uniform bandwidth subbands. On the other hand, the ICDM-II based VDF proposed in Section 4.3 can be used for multi-standard channelization as it can provide uniform as well as non-uniform bandwidth subbands. But the ICDM-II based VDF faces certain design constraints, as group delays of the resultant frequency responses after ICDM-II operations have to be integer valued to enable spectral subtraction. When a large number of decimation factors are involved, the order of the prototype filter satisfying the group delay constraint to ensure integer valued group delays would also be

larger. This would be a disadvantage when channels corresponding to large number different communication standards are to be extracted. In this section, a VDF is proposed which can perform all the ICDM operations – CDM-I, MCDDM-I, CDM-II as well as MCDDM-II to provide variable lowpass, highpass, bandpass, bandstop and multi-band frequency responses. The proposed VDF is termed as comprehensive ICDM based VDF. Due to its ability to perform all the ICDM operations, the comprehensive ICDM based VDF retains the advantages of both ICDM-I as well as ICDM-II based VDFs and minimizes the design and performance constraints associated with them. Subsequent sections present the comprehensive ICDM based VDF in detail along with design examples and the corresponding implementation results.

4.4.1 Hardware Implementation Architecture

Figure 4.6 shows the hardware implementation architecture for realizing the comprehensive ICDM based VDF, which can be used to perform ICDM-I as well as ICDM-II operations. It is a transposed direct-form FIR filter structure which is appropriately modified to enable ICDM operations on the filter coefficients.

Figure 4.6. Comprehensive ICDM based VDF: Hardware implementation architecture.

In Figure 4.6, the multipliers $h_0, h_1, \dots, h_{(N-1)}, h_N$ denote the prototype filter coefficients. Only half of the filter coefficients need to be implemented due to their symmetry property, which will reduce the number of coefficient multiplications by 50%. The multiplexers labeled as ‘mux I’ are used to perform ICDM-I operations, i.e., replacement of filter coefficients by zeros according to decimation factor M_1 . The multiplexers labeled as ‘mux II’ are used to perform ICDM-II operations, i.e., elimination of filter coefficients’ multiplication according to decimation factor M_2 . Similar to hardware implementation architectures of the ICDM-I and ICDM-II based VDFs discussed in previous sections of this chapter, adder/ subtractor (*add/ sub*) blocks are used to perform sign reversal of appropriate coefficients in the MCDM-I and MCDM-II operations. The default operation of *add/ sub* blocks is addition. When the sign of a filter coefficient is to be reversed in a MCDM-I or MCDM-II operation, the corresponding *add/ sub* block is set to perform subtraction by providing the select (*sel*) signal appropriately. It is to be noted that usage of *add/ sub* blocks to perform sign reversal instead of multiplying the filter coefficients by -1 is more efficient in terms of hardware resource utilization. The ‘Decimation Selector’ block is used to provide select signals of multiplexers *mux I* and *mux II* as well as those of the *add/sub* blocks. When $M_1=M_2$, the select signals of *mux I* and *mux II* corresponding to every filter coefficient are identical. The multiplier at the output of the implementation architecture is used to scale the filter output by M_{out} , to regain the original magnitude level of the passband after ICDM operations.

4.4.2 Design Procedure

A generalized design procedure to obtain different frequency responses using the proposed ICDM based VDF is presented in this section.

Step-1: Prototype filter design-

Compute the order of the required prototype filter using (3.14) by substituting $M = M_{\max}$, where $M_{\max}$ = maximum possible decimation factor (without aliasing), which can be calculated as $M_{\max} = \lfloor (1/f_s) \rfloor$.

Obtain the filter coefficients using a suitable filter design tool. Implement the filter coefficients using the proposed hardware implementation architecture shown in Figure 4.6. The following steps describe how different types of frequency responses can be obtained by performing appropriate ICDM operations using the implemented filter structure.

Step-2: Selection of coefficient decimation factor values-

Let M_1 be the decimation factor required to obtain a desired type of frequency response using an appropriate ICDM-I operation (CDM-I or MCDM-I). The CDM-II operation can be simultaneously performed to modify bandwidths of subbands in the default ICDM-I output response. Let M_2 be the decimation factor required in the CDM-II operation.

Step-3: Obtaining desired frequency responses-

To obtain a frequency response similar to that of an ICDM-I operation by M_1 and having subband bandwidths M_2 times that of the default ICDM-I output response, perform the following operations simultaneously:-

3(a). CDM-II using $M=M_2$ to achieve the desired bandwidth.

3(b). Appropriate ICDM-I operation (CDM-I or MCDM-I) using $M=M_1'=M_1 \times M_2$ to obtain desired frequency response.

3(c). Scaling the output of the filter using $M_{out} = M_1 \times M_2$.

Note: Variable lowpass and highpass frequency responses having bandwidths M_2 times that of the prototype filter can be obtained by setting $M_1=1$ and performing CDM-II and MCDM-II operations respectively, using $M=M_2$.

4.4.3 Illustrative Example

In this section, the flexibility of the comprehensive ICDM based VDF implementation architecture to obtain variable frequency responses is illustrated with the help of a design

example. Figure 4.7 (a) shows the frequency response of a lowpass prototype filter that is designed for the specifications $f_p = 0.07$, $f_s = 0.1$, $\delta_p = 0.05$ dB, $\delta_s = -40$ dB, $M_{\max} = 10$. Following *Step-1* of the design procedure, the order of the prototype filter is computed to be 180, using (3.14). Variable lowpass and highpass frequency responses can be obtained by setting $M_1=1$ and performing appropriate ICDM-II operations using $M_2 = \{1, 2, 3, \dots, 10\}$. Figure 4.7(b) shows different lowpass frequency responses obtained by performing CDM-II operation using $M_2 = \{2, 3, 6, 8\}$. Similarly, different highpass frequency responses obtained by performing MCDM-II operation using $M_2 = \{1, 2, 4, 5, 7\}$ are shown in Figure 4.7(c). *Step-3* of the design procedure is followed to perform appropriate ICDM-I and CDM-II operations simultaneously, to obtain distinct bandpass, bandstop and multi-band frequency responses. Figure 4.7(d) shows different bandpass frequency responses with passband center frequency 0.5, obtained by performing MCDM-I and CDM-II operations simultaneously using $\{M_1', M_2\} = \{2, 1\}, \{4, 2\}, \{8, 4\}$. Similarly, Figure 4.7(e) shows different bandstop frequency responses with stopband center frequency 0.5, obtained by performing CDM-I and CDM-II operations simultaneously using $\{M_1', M_2\} = \{2, 1\}, \{6, 3\}, \{8, 4\}$. Figure 4.7(f) shows different multi-band frequency responses obtained by performing CDM-I and CDM-II operations simultaneously using $\{M_1', M_2\} = \{3, 1\}, \{6, 2\}$. Similarly, Figure 4.7(g) shows different multi-band frequency responses obtained by performing MCDM-I and CDM-II operations simultaneously using $\{M_1', M_2\} = \{4, 1\}, \{8, 2\}$.

Figure 4.7(a). Frequency response of the prototype filter.

Figure 4.7(b). Lowpass frequency responses obtained by performing CDM-II using different M values.

Figure 4.7(c). Highpass frequency responses obtained by performing MCDM-II using different M values.

Figure 4.7(d). Bandpass frequency responses obtained by performing MCDM-I using $M_1=2$ and CDM-II using different M values.

Figure 4.7(e). Bandstop frequency responses obtained by performing CDM-I using $M_1=2$ and CDM-II using different M values.

Figure 4.7(f). Multi-band frequency responses obtained by performing CDM-I and CDM-II simultaneously, using non-identical M_1 and M_2 .

Figure 4.7(g). Multi-band frequency responses obtained by performing MCDM-I and CDM-II simultaneously, using non-identical M_1 and M_2 .

From Figures 4.7(a) - 4.7(g), it can be noted that variable lowpass, highpass, bandpass, bandstop and multi-band frequency responses can be obtained by performing appropriate ICDM operations using the proposed VDF hardware implementation architecture shown in Figure 4.6. These figures thus illustrate the flexibility of the comprehensive ICDM based VDF to provide variable frequency responses on-the-fly, by simply selecting the appropriate values of M_1 and M_2 to control the driving signals of the constituent multiplexers and adder/ subtractor blocks. To obtain larger number of distinct possible frequency responses, the implementation architecture shown in Figure 4.6 can be modified to include parallel branches, each performing different ICDM operations on the same set of prototype filter coefficients. The output frequency responses from the parallel branches can then be added/ subtracted from each other to retain the desired subbands and discard the unwanted ones. Complementary and interpolated frequency response operations can also be incorporated in the modified architecture by adding appropriate delay and multiplexer blocks. The different output frequency responses can then be algebraically operated upon or suitably frequency response masked to obtain various desired subband configurations.

4.4.4 Channelization Design Example

The ability of the proposed comprehensive ICDM based VDF to extract channels corresponding to different wireless communication standards is illustrated in this section. Consider a SDR channelization scenario wherein channels corresponding to three standards - Wideband Code Division Multiple Access (WCDMA), Worldwide Interoperability for Microwave Access (WiMAX) and Long Term Evolution (LTE) are to be extracted in different time intervals. Figure 4.8(a) shows one channel each of WCDMA, WiMAX and LTE with channel bandwidths 5 MHz, 10 MHz and 20 MHz respectively, present in the input signal at different time intervals. WCDMA signal is present during the time interval t_1 - t_2 , WiMAX during t_3 - t_4 and LTE during t_5 - t_6 . The sampling frequency is 100 MHz. Let the desired passband and stopband ripple specifications be 0.05 dB and -40 dB respectively. The design procedure given in Section 4.4.2 is used to obtain a comprehensive ICDM based VDF which can extract these three

desired channels from the input signal during their corresponding time intervals. The prototype filter is designed for the specifications $f_p = 0.09$, $f_s = 0.11$, $M_{\max} = 9$, with order computed to be 266 using (3.14). Figure 4.8(b) shows the frequency responses obtained after appropriate ICDM operations, which can be used to extract the desired channels from the input signal. WCDMA channel present during t_1 - t_2 can be extracted using the frequency response obtained after performing CDM-I operation using $M=1$ (it can be noted that this is the frequency response of the lowpass prototype filter itself). The

Figure 4.8(a). Input signal containing WCDMA, WiMAX and LTE channels during different time intervals.

Figure 4.8(b). Frequency responses used for extracting the three desired channels of WCDMA, WiMAX and LTE standards as shown in Figure 4.8(a).

WiMAX channel present during t_3-t_4 can be extracted using the frequency response obtained after performing MCDM-I operation using $M=2$. Similarly, the LTE channel present during t_5-t_6 can be extracted using the frequency response obtained after performing MCDM-II operation using $M=4$.

4.4.5 FPGA Implementation Results

In this section, hardware implementation results for different VDF designs of the comprehensive ICDM based VDF are presented. To study the trends in resource utilization, power consumption and operating frequency with respect to the prototype filter order, three comprehensive ICDM based VDFs: VDF-I, VDF-II and VDF-III were designed with lowpass prototype filters of orders 60, 120 and 180 respectively. The three VDFs were implemented on a Xilinx Virtex-6 (xc6vlx240t) FPGA based on the hardware implementation architecture shown in Figure 4.6. All the FPGA implementation results presented in this thesis have been obtained using Matlab, Simulink, Xilinx System Generator [95] and Xilinx ISE tools. Digital signal processor (DSP) blocks on FPGAs enable high throughput filter design, however, they are optimized for traditional transposed-form filter structures. High level descriptions can map such transposed-form filter structures efficiently. But while implementing complex and modified filter structures, it is difficult to take advantage of the performance benefits of DSP blocks and hence custom implementations are necessary [96]. Therefore, DSP block inference was disabled for all VDF implementations since the fixed coefficients and the proposed complex modified filter structure do not benefit from their use. Table 4.4 shows the resource utilizations and maximum operating frequencies of the three VDFs, along with

Table 4.4. FPGA implementation results: Comprehensive ICDM based VDF designs.

	VDF-I	VDF-II	VDF-III
No. of occupied slices	4280	7993	12276
Power (mW)	387.07	612.47	1070.74
Maximum frequency (MHz)	124.828	100.02	85.252

Figure 4.9. Comprehensive ICDM based VDF: Pipelined hardware implementation architecture.

the corresponding power estimates generated using post-place and route simulation data. In Table 4.4, the number of occupied slices denotes the hardware resource utilization. (Note that in FPGAs, a slice is a hardware resource which contains a set number of look-up tables (LUTs) and flip-flops. For example, in Xilinx Virtex 6 FPGA family, one slice contains four LUTs and eight flip-flops.) Along with the expected increase in resource utilization and power consumption, it is observed that increasing order of the prototype filter results in decrease in the maximum operating frequency. The maximum operating frequency decreases because as the prototype filter order increases, the critical path length increases with depth of the combinatorial computational path. If greater stopband attenuation and narrower transition-band widths are desired in the resultant frequency responses, higher order prototype filters are required which will lead to lower operating frequencies. Therefore, when the frequency response specifications are stringent, the dependence of operating frequency on the prototype filter order will be a major bottleneck if high operating frequencies are also desired. To eliminate this operating frequency bottleneck, a modified hardware implementation architecture for the comprehensive ICDM based VDF is shown in Figure 4.9. The modified architecture is extensively pipelined to break the combinatorial path, which is then balanced in the input data-path by

Table 4.5. FPGA implementation results: Pipelined comprehensive ICDM based VDF designs.

	VDF-I-P	VDF-II-P	VDF-III-P
No. of occupied slices	2985	5789	9188
Power (mW)	190.37	335.35	519.70
Maximum frequency (MHz)	157.853	158.078	157.754

adding delay-chains. The pipelined architecture, though increases the latency from input to output by N clock cycles (for N -order prototype filter), isolates the critical path and prevents the accumulation effect that is present in the original implementation architecture shown in Figure 4.6.

To compare the pipelined architecture (Figure 4.9) with the straightforward non-pipelined implementation (Figure 4.6), the three VDF designs were re-implemented using the pipelined architecture, and the same design options for the same FPGA device. Let the pipelined VDF designs be denoted as VDF-I-P, VDF-II-P and VDF-III-P with prototype filters of orders 60, 120 and 180 respectively. Table 4.5 shows the corresponding implementation results. Comparing the resource utilizations of different VDF designs shown in Tables 4.4 and 4.5, it can be observed that the pipelined designs achieve substantial reduction in the number of occupied slices, despite the increased number of registers used. The pipeline stages added in the designs allow the synthesis tool to optimize the multiplexers and related logic in the delay-chain, thus reducing the overall resource usage. This reduction in resources for the pipelined designs translates to approximately 50% reduction in power consumption, when compared with the non-pipelined implementations. To ensure a fair comparison, all the power estimates were generated based on signal activity rates determined by simulating the designs at an operating frequency of 50 MHz. From Tables 4.4 and 4.5, it can be observed that the pipelined designs achieve much higher operating frequencies than the corresponding non-pipelined designs, and the operating frequencies remain nearly constant across the different VDF configurations with varying prototype filter orders. The average maximum frequency achieved for the pipelined VDF designs was 157.895 MHz. The pipelined implementation architecture shown in Figure 4.9 thus makes the operating frequency of the comprehensive ICDM based VDF independent of the prototype filter order.

4.5 Summary

Based on the ICDM proposed in Chapter 3, different VDF design techniques were proposed in this chapter. An ICDM-I based VDF which provides variable frequency responses by performing the CDM-I and MCDM-I operations was proposed. The ICDM-I based VDF was shown to have greater frequency response flexibility, lower complexity and improved stopband attenuation performance when compared with the conventional CDM-I based VDF. The ICDM-I based VDF can be used in SDR handsets for extraction of individual channels of a single wireless communication standard. An ICDM-II based VDF which provides variable frequency responses by performing the CDM-II and MCDM-II operations was also proposed in this chapter. In the design example considered, the ICDM-II based VDF achieved 57.12% reduction in multiplication complexity when compared with the conventional CDM-II based VDF that was designed for the same frequency response specifications. Also, if the same prototype filter is used, the proposed ICDM-II based VDF offers superior stopband and transition-band characteristics when compared with the CDM-II based VDF. The ICDM-II based VDF can be used for extracting individual channels of multiple wireless communication standards present in the input signal during different time intervals.

A VDF based on the comprehensive ICDM was also proposed. It can perform all the ICDM operations (CDM-I, MCDM-I, CDM-II and MCDM-II) and provides frequency responses with varying subband bandwidths and locations, thus enabling multi-standard channelization. Low complexity FPGA implementations were achieved by pipelining the proposed hardware implementation architecture, which enabled significantly higher operating frequencies that are independent of the prototype filter order. The comprehensive ICDM based VDF can be used for low complexity multi-standard channelization in SDR handsets.

When compared with VDFs based on the conventional CDM, the proposed VDFs offer greater frequency response flexibility and lower implementation complexity. They are therefore compatible for use in SDR handsets, especially in resource constrained battery-operated portable radio handsets. Following the VDFs proposed in this chapter, different filter banks which employ the ICDM technique are presented in the next chapter.

Chapter 5

Design of Uniform and Non-Uniform Filter Banks using Improved Coefficient Decimation Method

5.1 Introduction

In software defined radio (SDR) base-station receivers, filter banks that are capable of providing uniform as well as non-uniform subbands are used for multi-standard channelization purposes. Such filter banks can simultaneously extract multiple channels corresponding to different wireless communication standards. It is desired to use filter banks which can provide maximum flexibility in terms of the number, location and bandwidth of the subbands. Also, in applications where the resources available are limited, low complexity filter banks with flexible frequency responses are desired. Three filter banks based on improved coefficient decimation method (ICDM) which can be used for channelization in SDR receivers are presented in this chapter. The subsequent sections present the proposed filter banks in detail along with their generalized design procedures, design examples and comparison with other relevant filter banks in literature. The ICDM-I based filter bank, ICDM-II based filter bank and the comprehensive ICDM based filter

bank described in Sections 5.2, 5.3 and 5.4 of this chapter have been published in [J1, C3], [J2] and [J3] respectively.

5.2 Proposed ICDM-I based Filter Bank

This section presents the first filter bank proposed in this chapter. It is based on the improved coefficient decimation method I (ICDM-I) which performs coefficient decimation method I (CDM-I) and modified coefficient decimation method I (MCDM-I) operations to provide variable frequency responses which can be used to obtain the desired subbands in the filter bank. In the proposed filter bank design method, frequency responses obtained using the ICDM-I are subjected to spectral subtraction, complementary filter operation and frequency response masking to obtain the desired subbands in the filter bank. The proposed filter bank is termed as ICDM-I based filter bank. Subsequent sections present the ICDM-I based filter bank in detail.

5.2.1 Design Procedure

Step-1: Choose the passband and stopband edge specifications of the lowpass prototype filter corresponding to the normalized bandwidths of the desired subbands.

Step-2: Using (3.1) and (3.10), select appropriate ICDM-I operations and the corresponding decimation factor values required for obtaining different frequency responses from which the desired subbands in the filter bank can be obtained. Let the maximum value of the decimation factor identified be $M_{\max}$.

Step-3: To meet the desired stopband attenuation specifications, use (3.14) to compute the order of the prototype filter corresponding to $M_{\max}$. Design the prototype filter and obtain its coefficients.

Step-4: Perform appropriate ICDM-I operations on the prototype filter using the M values identified in *Step-2* and obtain the corresponding frequency responses.

Figure 5.1. Proposed ICDM-I based filter bank: Block diagram.

Step-5: Obtain the desired uniform subbands in the filter bank by performing spectral subtraction of the appropriate frequency responses obtained in *Step-4* and their corresponding complementary frequency responses if required. Low complexity frequency response masking filters can also be employed if needed, for separating individual subbands from the multi-band frequency responses obtained in *Step-4*.

The design procedure of the ICDM-I based filter bank is summarized using a block diagram as shown in Figure 5.1.

5.2.2 Design Example and Comparison

This section presents the design example of a uniform filter bank, implementation results and comparison of the ICDM-I based filter bank with other relevant filter banks designed for the same frequency response specifications.

Consider the example of an 8-channel discrete Fourier transform (DFT) based filter bank (DFTFB) whose output frequency response consists of 5 real subbands in the frequency range 0 to $f_{\text{samp}}/2$, where f_{samp} is the sampling frequency, as shown in Figure 5.2. In Figure 5.2, the bandwidths of subbands SB1 and SB5 appear to be half of those of the other subbands because half of their bandwidths are located in the complex domain, i.e., the negative frequency range $-f_{\text{samp}}/2$ to 0.

Figure 5.2. Frequency response of 8-channel DFTFB from 0 to $f_{\text{samp}}/2$.

The prototype filter in the DFTFB is to be designed such that its frequency response matches with that of subband SB1 in Figure 5.2. Let the desired passband and stopband peak ripple specifications be 0.1 dB and -40 dB respectively. The normalized passband and stopband edge frequency specifications of the prototype filter are chosen as $f_p = 0.1125$, $f_s = 0.1375$ and the corresponding order of the prototype filter computed using (3.10) is 160. The prototype filter in polyphase form is followed by an 8-point inverse discrete Fourier transform (IDFT) operation to obtain the desired DFTFB having its subbands as shown in Figure 5.2.

The ICDM-I based filter bank is designed such that its subbands correspond to the five real subbands shown in Figure 5.2. The passband peak ripple and the passband and stopband edge frequency specifications of the prototype filter used in the ICDM-I based filter bank design are kept the same as those of the prototype filter used in the DFTFB design discussed above. The block diagram in Figure 5.3 shows the various stages in the designed ICDM-I based filter bank and the corresponding output frequency responses. The frequency response of the prototype filter itself gives the desired SB1. SB5 is obtained by spectrally subtracting the frequency response of the prototype filter from the frequency response obtained after performing CDM-I using $M=2$. The frequency response obtained after performing MCDM-I using $M=2$ gives the desired SB3. The complementary response of the frequency response obtained after performing MCDM-I using $M=2$ is obtained using complementary delays as shown in Figure 5.3. The resultant

frequency response after performing CDM-I using $M=2$ is then spectrally subtracted from this complementary response to obtain the desired SB2 and SB4 combined at the output. Two frequency response masking filters, each of order 18 designed using (3.10), are employed to separate the subbands SB2 and SB4. Thus, the five desired subbands are obtained using the ICDM-I based filter bank. It can be noted that the maximum decimation factor involved in the ICDM-I based filter bank design is 2. Also, the required order of the prototype filter used in the ICDM-I based filter bank is computed using (3.15) as 176.

Figure 5.3. Proposed ICDM-I based filter bank: Design example.

A coefficient decimation method (CDM) based filter bank (CDFB) is realized using the design procedure given in [90] to obtain the five real subbands shown in Figure 5.2. The peak passband ripple and the passband and stopband edge frequency specifications of the prototype filter in the CDFB are kept the same as that of the prototype filters used in the DFTFB and the ICDM-I based filter bank designs. CDM-I is performed on the prototype filter using $M=2$ and $M=4$ to obtain the corresponding output frequency responses. These frequency responses along with the prototype filter's frequency response are then algebraically operated upon to get the five desired subbands at the output [90]. The maximum decimation factor involved in the CDFB design is 4 and the corresponding order of the overdesigned prototype filter required to satisfy the desired stopband attenuation specification is 192, using (3.14). The masking filters used to separate SB2 and SB4 are kept the same as those used in the ICDM-I based filter bank design for fair comparison.

5.2.2.1 Complexity Analysis

The complexity of a filter bank is mainly dependent on the number of multiplication operations involved in its implementation. Table 5.1 shows the number of multiplications required to realize different filter banks to achieve the desired frequency response specifications in this design example.

The total number of multiplications required in the DFTFB implementation is the sum of the prototype filter length (i.e., filter order + 1) and the number of multiplications required for an 8-point IDFT computation. The radix-2 fast Fourier transform (FFT) algorithm was used for implementing the IDFT, and it requires $S \log_2 S$ multiplications for computing S -point FFT of a input signal [42, 43]. Thus, the number of multiplications involved in the DFTFB is $161 + 8 \log_2 8 = 161 + 24 = 185$. The Goertzel algorithm based filter bank (GFB) [25] shows a significantly higher multiplication complexity than the DFTFB due to the Goertzel algorithm used in it for DFT implementation. The transposed direct-form finite impulse response (FIR) filter structure which exploits the symmetry property of filter coefficients cannot be used in the DFTFB and GFB due to the polyphase

Table 5.1. ICDM-I based filter bank: Multiplication complexity comparison.

	DFTFB [75]	GFB [25]	CDFB [90]	Proposed ICDM-I based filter bank
Prototype filter length	161	161	193	177
Masking filter length	-	-	$19 \times 2 = 38$	$19 \times 2 = 38$
Number of multiplications for filter implementation	161	161	$\{\lceil 193/2 \rceil + (2 \times \lceil 19/2 \rceil)\}$ $= 117$	$\{\lceil 177/2 \rceil + (2 \times \lceil 19/2 \rceil)\}$ $= 109$
Number of multiplications for DFT computation	$8 \log_2 8$ $= 24$	8×161 $= 1288$	-	-
Total multiplications required in the filter bank	185	1449	117	109

form implementation of the prototype filter. The prototype filter and masking filters in both the CDFB and proposed ICDM-I based filter bank designs are implemented using the transposed direct-form FIR filter structure to exploit the symmetry property of their coefficients. Thus, the total number of multiplications required in CDFB is $\{\lceil 193/2 \rceil + (2 \times \lceil 19/2 \rceil)\} = 117$ and that in the ICDM-I based filter bank is $\{\lceil 177/2 \rceil + (2 \times \lceil 19/2 \rceil)\} = 109$. From Table 5.1, it can be noted that the proposed ICDM-I based filter bank offers multiplication complexity reductions of 41.08% over DFTFB, 92.48% over GFB [25] and 6.84% over the CDFB [90].

5.2.2.2 Flexibility Comparison

As discussed in Section 3.4, the center frequency resolution in the multi-band frequency responses obtained after performing ICDM-I operations is π/M , where M corresponds to the decimation factor used. Thus, the possible center frequency locations for the subbands are at integral multiples of π/M in the ICDM-I based filter bank, whereas the center frequencies are located at integral multiples of $2\pi/M$ for both CDFB [90] and

M -channel DFTFB. The proposed filter bank has twice the center frequency resolution for the location of its constituent subbands, when compared with the DFTFB and CDFB. Also, unlike in the case of DFTFB where the value of M is fixed due to the M -point IDFT implementation, different values of M corresponding to different decimation factors can be used in the ICDM-I based filter bank to obtain multiple sets of distinct center frequency locations with a resolution of π/M . Thus, the higher center frequency resolution provided by the ICDM-I used in the proposed filter bank enables it to achieve a larger number distinctly located subbands when compared with the DFTFB and CDFB [90].

In this design example, both the prototype filters used in the CDFB [90] and the ICDM-I based filter bank have their normalized stopband edge frequency $f_s = 0.1375$. Thus, values of M beyond $\lfloor 1/(0.1375) \rfloor = 7$ will lead to aliasing and cannot be used in the coefficient decimation operations [90]. It can be noted that the center frequency locations obtained by performing MCDM-I by $M = \{5, 6, 7\}$ cannot be obtained by performing CDM-I on these prototype filters, as values of $M = \{10, 12, 14\}$ are correspondingly required if CDM-I is to be performed [90]. The ability to perform both CDM-I as well as MCDM-I operations in the ICDM-I based filter bank results in a significantly more number of possible distinct center frequency locations for its constituent subbands when compared with the CDFB, which merely employs the CDM-I operation. Thus the proposed ICDM-I based filter bank has a significantly higher flexibility when compared to the DFTFB and CDFB in terms of both the number of subbands that can be obtained as well as their locations.

5.2.3 FPGA Implementation Results

Implementation results for various filter bank designs discussed in this design example are presented in this section. The DFTFB, CDFB [90] and the ICDM-I based filter bank designed for the desired specifications were implemented in Xilinx Virtex-5 (xc5v1x330) FPGA. The GFB design was not implemented in the FPGA because of its significantly higher complexity due to the larger number of multiplications required. Table 5.2 shows the hardware resource utilization details along with the power consumption and maximum

Table 5.2. ICDM-I based filter bank: FPGA implementation results.

	DFTFB [75]	CDFB [90]	Proposed filter bank
No. of occupied slices	20433	7032	5943
Power (mW)	499	268	206
Maximum frequency (MHz)	230.2	61.84	64.03

operating frequency values for the implemented designs. From Table 5.2, it can be noted that the ICDM-I based filter bank achieves 70.92% and 15.49% reductions in the number of occupied slices when compared with the DFTFB and CDFB [90] respectively. In the CDFB and ICDM-I based filter bank's implementation, the prototype filters as well as the masking filters are implemented using the transposed direct-form FIR filter structure, which requires only half of the total filter coefficients to be implemented. Thus, in spite of the use of masking filters and the oversized prototype filters, the CDFB and the ICDM-I based filter bank achieve significant reductions in the number of occupied slices when compared with the DFTFB. The transposed direct-form FIR filter structure cannot be used in the DFTFB due to the polyphase form implementation of the prototype filter. When compared with the CDFB, the lower order prototype filter used in the ICDM-I based filter bank and the corresponding fewer number of computational blocks result in a lower slice requirement for the latter. The proposed ICDM-I based filter bank offers 58.72% and 23.13% reductions in power consumption when compared with the DFTFB and CDFB [90] respectively. The DFTFB has the highest power consumption because of the significantly larger number of resources utilized in it. The ICDM-I based filter bank has the least power consumption as it has the minimum resource utilization.

In polyphase DFTFB, using noble identity [77], filtering can be moved after the decimation block, i.e., decimate the signal by M for M -channel DFTFB and then perform filtering. (Note: As described in [77], the 'noble identity' states that a system wherein a decimator is followed by a transfer function is equivalent to a system wherein the same transfer function in polyphase form is followed by a decimator.) This means that in DFTFB implementations, the filters can be operated at a lower sampling rate which will result in low dynamic power consumption. But this shifting of filtering block after

decimation is not possible in CDM and ICDM based design techniques, and thus the dynamic power consumption will be higher in them if compared with a modified DFTFB implementation which employs the noble identity. However, in nanoscale CMOS technology, static power dominates overall power consumption [97-100]. Also, static power consumption is directly proportional to hardware area utilization. Therefore, it is expected that the significant area reduction achieved in proposed ICDM based design techniques will reduce the static power, thus keeping overall power consumption lower than the other methods.

The maximum operating frequency values obtained after placement and routing in the FPGA show that the DFTFB can achieve highest operating frequency amongst the three filter banks, which is expected because of the fast Fourier transform (FFT) technique used in it for the DFT implementation. The ICDM-I based filter bank can achieve a 3.54% higher operating frequency than the CDFB due to the lower order prototype filter used in it, which ultimately results in a fewer number of computational blocks and delay elements required for its implementation.

The significant advantages in power consumption and resource utilization offered by the proposed ICDM-I based filter bank makes it a low complexity alternative to the other uniform filter banks, thereby making it compatible for use in SDR receivers where resource utilization and power consumption are critical parameters.

5.3 Proposed ICDM-II based Filter Bank

The ICDM-I based filter bank proposed in Section 5.2 provides only uniform bandwidth subbands which limits its application uniform channelization. To perform multi-standard channelization, it is necessary to obtain subbands with varying bandwidths corresponding to different wireless communication standards that are involved. To achieve this, a new filter bank is proposed in this section which performs improved coefficient decimation method II (ICDM-II) operations to obtain variable frequency responses which can be used to obtain uniform as well as non-uniform subbands. The

proposed filter bank is termed as ICDM-II based filter bank. Subsequent sections present the hardware implementation architecture and the generalized design procedure for the ICDM-II based filter bank. A design example with uniform as well as non-uniform channelization scenarios is also presented followed by comparison of the ICDM-II based filter bank with the conventional CDM based progressive decimation filter bank (PDFB).

5.3.1 Hardware Implementation Architecture

Figure 5.4 shows the proposed hardware implementation architecture for the ICDM-II based filter bank. It consists of adder/subtractor (*add/sub*) blocks that are used to perform the sign reversal of the corresponding coefficients in ICDM-II operations. The select signal (*sel*) of the *add/sub* blocks is to be set to 1 (for subtraction) or 0 (for addition) according to the desired operation. The *Decimation Selector* shown in Figure 5.4 performs this operation and drives the *add/sub* blocks appropriately.

Figure 5.4. Proposed ICDM-II based filter bank: Hardware implementation architecture.

The outputs of the ICDM-II operations for different M values are obtained at the different branches of the filter bank as shown in Figure 5.4. These outputs can then be spectrally subtracted to obtain the corresponding uniform or non-uniform subbands at the output. The resultant outputs obtained after performing any of the ICDM-II operations need to be scaled by M to obtain the original magnitude response. This is achieved by multiplying the outputs at the different branches by appropriate values of M as shown in architecture in Figure 5.4.

5.3.2 Design Procedure

The steps involved in the design of the ICDM-II based filter bank are given below.

Step-1: According to the channel bandwidths of different wireless communication standards and the desired frequency response specifications, compute the passband and stopband edge frequency specifications of the lowpass prototype filter. These specifications can be fixed with respect to the greatest common divisor (GCD) of the channel bandwidths of the communication standards involved.

Step-2: Using (3.1) and (3.10), select the decimation factor (M) values required to obtain the desired subbands in the filter bank using appropriate ICDM-II operations. Compute the least common multiple (LCM) of all the identified M values. Let the maximum decimation factor required be $M_{\max}$.

Step-3: For the desired subbands to be obtained in the ICDM-II based filter bank, if the minimum transition-band width specification is $TBW_{\min}$, then fix the transition-band width of the prototype filter to be $\frac{TBW_{\min}}{M_{\max}}$.

Step-4: Design the prototype filter using (3.14) and (3.15) such that its order (N) is a multiple of the LCM calculated in *Step-2*. This ensures that the group delays of all the decimated filters are integer values. Also, to meet the desired stopband attenuation specifications, compute the value of N with respect to $M_{\max}$.

Figure 5.5. Proposed ICDM-II based filter bank: Block diagram.

Step-5: Perform ICDM-II operations on the prototype filter with the M values identified in *Step-2* and obtain the corresponding frequency responses.

Step-6: Obtain the desired subbands in the filter bank by appropriate spectral subtraction of the frequency responses obtained in *Step-5*.

The design procedure of the ICDM-II based filter bank is summarized using a block diagram as shown in Figure 5.5.

5.3.3 Design Example and Comparison

In this section, the ability of the proposed ICDM-II based filter bank to simultaneously extract frequency channels of distinct bandwidths corresponding to multiple wireless communication standards is illustrated with the help of a design example. Consider a multi-standard communications scenario where Bluetooth, Zigbee and Wideband Code Division Multiple Access (WCDMA) channels are to be extracted using a filter bank. The channel bandwidths of Bluetooth, Zigbee and WCDMA are 1 MHz, 4 MHz and 5 MHz respectively. The sampling frequency is 24 MHz. The desired passband and stopband specifications are chosen as $\delta_p = 0.1$ dB and $\delta_s = -55$ dB for Bluetooth channels, $\delta_p = 0.1$ dB and $\delta_s = -45$ dB for Zigbee channels, and $\delta_p = 0.1$ dB, and $\delta_s = -50$ dB for WCDMA channels respectively. The transition-band width specifications are chosen as 240 kHz, 300 kHz and 400 kHz for Bluetooth, Zigbee and WCDMA respectively.

Figure 5.6. ICDM-II output frequency responses for $M=1$ to $M=6$.

Considering the worst case (most stringent) design specifications and using the proposed design procedure, the passband width of the prototype filter is fixed as $1/12 = 0.0833$ for the normalized frequency range 0 to 1 (1 corresponds to half of sampling frequency). The normalized passband and stopband edge frequencies of the prototype filter are chosen as $f_p = 0.0803$ and $f_s = 0.0833$ respectively. The different decimation factor values required are $\{1, 2, 3, 4, 5, 6\}$ and thus $M_{\max} = 6$. The LCM of these M values is 60. Following design *Step-4*, the prototype filter is designed with its filter order computed as 2400 using (3.14) and (3.15). After performing appropriate ICDM-II operations using the chosen M values, the corresponding frequency responses are obtained. These frequency responses are shown in Figure 5.6. The frequency responses shown in Figure 5.6 are spectrally subtracted from each other to obtain 12 adjacently located uniform subbands (The spectral subtraction operations can be performed as described in the illustrative example given in Section 3.4.1.2). These subbands can be used for extracting up to 12 non-overlapping Bluetooth channels simultaneously.

Consider the scenario where two Zigbee channels (ZB1 and ZB2) are simultaneously present in the wideband input signal with their spectral locations as shown in Figure 5.7(a). The Zigbee channel ZB1 located between 2-6 MHz is extracted using the subband obtained by the spectral subtraction of the coefficient decimation method II (CDM-II)

Figure 5.7(a). Input signal containing two Zigbee channels.

Figure 5.7(b). ICDM-II frequency responses used for extracting the Zigbee channels shown in Figure 5.7(a).

output responses for $M=2$ and $M=6$. The Zigbee channel ZB2 located between 8-12 MHz is extracted using the subband obtained as the modified coefficient decimation method II (MCDM-II) output response for $M=4$. The ICDM-II frequency responses used to obtain these subbands for simultaneously extracting the two Zigbee channels are shown in Figure 5.7(b).

The frequency responses shown in Figure 5.6 can be spectrally subtracted or added to obtain 9 uniform subbands with any of the following frequency locations $\{(0-4 \text{ MHz}), (1-5 \text{ MHz}), (2-6 \text{ MHz}), (3-7 \text{ MHz}), (4-8 \text{ MHz}), (5-9 \text{ MHz}), (6-10 \text{ MHz}), (7-11 \text{ MHz}), (8-12 \text{ MHz})\}$. These subbands can be used for extracting up to 3 non-overlapping Zigbee channels simultaneously, as discussed for the Zigbee channelization example shown in Figure 5.7. Similarly, 8 uniform subbands located at any of the frequency locations $\{(0-5 \text{ MHz}), (1-6 \text{ MHz}), (2-7 \text{ MHz}), (3-8 \text{ MHz}), (4-9 \text{ MHz}), (5-10 \text{ MHz}), (6-11 \text{ MHz}), (7-12 \text{ MHz})\}$ can be obtained by the spectral subtraction or addition of frequency responses shown in Figure 5.6. These subbands can be used for extracting up to 2 non-overlapping WCDMA channels simultaneously from the wideband input signal. For example, ICDM-II output responses with $M=1$ and $M=6$ can be used to obtain frequency subbands for simultaneously extracting two WCDMA channels located at 1-6 MHz and 6-11 MHz.

The ICDM-II based filter bank design is functionally verified for non-uniform multi-standard channelization by inputting a signal whose frequency spectrum is shown in Figure 5.8(a). Assume it is known that during one time interval, one WCDMA channel is located between 0-5 MHz, one Zigbee channel is located between 7-11 MHz and two Bluetooth channels BT1 and BT2 are located between 6-7 MHz and 11-12 MHz respectively. These distinct frequency channels corresponding to the three wireless communication standards WCDMA, Zigbee and Bluetooth are extracted using the frequency subbands obtained as discussed below.

The subbands corresponding to the Bluetooth signals are obtained by the spectral subtraction of the MCDM-II output responses for $M = 1$, $M = 5$ and $M = 6$. The CDM-II output response for $M = 5$ is used to obtain the frequency subband corresponding to the WCDMA signal. The Zigbee channel is extracted by the spectral subtraction of the MCDM-II output responses for $M=5$ and $M=1$. These frequency responses required to simultaneously obtain the subbands corresponding to WCDMA, Bluetooth, Zigbee channel distribution are shown in Figure 5.8(b).

Figure 5.8(a). Input signal containing Bluetooth, WCDMA and Zigbee channels during the same time interval.

Figure 5.8(b). ICDM-II frequency responses used for extracting the Bluetooth, WCDMA and Zigbee channels shown in Figure 5.8(a).

5.3.3.1 Complexity Analysis

The complexity of ICDM-II based filter bank and its comparison with other filter banks in terms of coefficient multiplications is presented here. For the ICDM-II based filter bank, the length of the prototype filter is 2401. The prototype filter is implemented using the transposed direct-form FIR filter structure and thus the total number of

coefficient multiplications required in the ICDM-II based filter bank implementation is $\lceil 2401/2 \rceil = 1201$.

The PDFB [89] performs CDM-II operations followed by spectral subtraction of the resultant frequency responses to obtain the desired subbands. Similar to the designed ICDM-II based filter bank, to obtain a flexible PDFB [89] implementation with the ability to extract Bluetooth, Zigbee and WCDMA channels located at variable locations, the decimation factors required are $\{1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11, 12\}$ and the maximum value of decimation factor required is thus 12. The LCM of these decimation factor values required in the PDFB is 27720. The corresponding order of the prototype filter required is calculated using (3.14) and (3.15), and is computed to be 55440. The corresponding number of coefficient multiplications required using the transposed direct-form implementation of the prototype filter is thus $\lceil 55441/2 \rceil = 27721$. When compared with a flexible PDFB designed for the same frequency response specifications, a 95.67% reduction in multiplication complexity is achieved using the ICDM-II based filter bank.

To obtain a PDFB [89] just for extracting the multi-standard frequency channels as shown in the channel distribution in Figure 5.8(a), the decimation factors required are $\{5, 6, 7, 11, 12\}$ and thus the maximum decimation factor involved is 12. The LCM of these required decimation factor values is 4620. Using (3.14) and (3.15), the order of the prototype filter required in this PDFB implementation is calculated as 9240. The corresponding number of coefficient multiplications required using the transposed direct-form implementation of the prototype filter is thus $\lceil 9241/2 \rceil = 4621$. Thus, a 74.01% reduction in complexity is achieved using the ICDM-II based filter bank when compared with the PDFB designed for the channel distribution shown in Figure 5.8(a).

The requirement of very high order prototype filter is a major drawback in the conventional CDM based PDFB in scenarios similar to the one considered in the design example, where higher decimation factor values are involved and several decimation factors are required. This drawback of PDFB is overcome in the proposed ICDM-II based filter bank because the maximum decimation factor value required in it is half of that used in the PDFB. Also, a substantially smaller number of distinct decimation factors are

required in the ICDM-II based filter bank than the PDFB. In general, if $M_{\max}$ represents the maximum decimation factor required in a flexible ICDM-II based filter bank implementation, the prototype filter design equations (3.14) and (3.15) that are to be used for a similar PDFB implementation are transformed as

$$N_{PDFB} \geq \left[-\frac{8\log_{10}(10\delta_p\delta_s)}{3(f_s - f_p)} - 1 \right] + \frac{8\log_{10}(2M_{\max})}{3(f_s - f_p)} \quad (5.1)$$

$$N_{PDFB} = m \times [LCM\{1, 2, 3, \dots, 2M_{\max}\}], \quad \text{where } m = 2, 3, 4, \dots \quad (5.2)$$

where N_{PDFB} represents the order of the prototype filter required in the PDFB. From (5.1) and (5.2), it can be noted that both these PDFB design equations can impose the requirement of very high order prototype filter when larger values of decimation factors are involved. Comparing these two equations with (3.14) and (3.15), it can be noted that the proposed ICDM-II based filter bank needs prototype filter of substantially lower order when compared with a PDFB designed for same desired specifications. This gives the ICDM-II based filter bank a major advantage over the PDFB in terms of multiplication complexity. For the design example described in this section, the multiplication complexity comparison of the ICDM-II based filter bank and the PDFB is presented in Table 5.3.

For a given set of specifications, if the ICDM-II based filter bank and PDFB are designed using the same prototype filter, the worst case stopband attenuation and

Table 5.3. ICDM-II based filter bank: Multiplication complexity comparison.

	Number of coefficient multiplications in the filter bank designed for channel distribution shown in Figure 5.8(a)	Number of coefficient multiplications in a flexible filter bank designed with variable subband locations
PDFB [89]	4621	27721
Proposed ICDM-II based filter bank	1201	1201

transition-band width values observed in ICDM-II based filter bank are half of that observed in PDFB. This is because the maximum decimation factor required in ICDM-II based filter bank implementation is half of that required in the PDFB. This can be justified using (3.1), (3.10) and (3.12). The proposed ICDM-II based filter bank thus shows significant advantages in terms of multiplication complexity, stopband and transition-band characteristics when compared with the conventional CDM based PDFB.

5.4 Comprehensive ICDM based Filter Bank

The ICDM-I based filter bank proposed in Section 5.2 can only be used for channelization of signals corresponding to a single wireless communication standard because the subbands have uniform bandwidth. To perform multi-standard channelization, an ICDM-II based filter bank was proposed in Section 5.3 which can provide subbands with varying bandwidths corresponding to different wireless communication standards that are involved. However, the ICDM-II based filter bank has two constraints involved in its design:

- i. Group delay constraint: The prototype filter has to be designed such that its order is a multiple of the LCM of the distinct decimation factors involved. This ensures that the resultant filters after ICDM-II operations have integer valued group delays, which is a necessary condition for performing spectral subtraction. If the desired number of subbands increases, the group delay constraint would impose the requirement of a higher order prototype filter, as the required number decimation factors and hence their LCM value will also be large in that case. This can significantly increase complexity of the ICDM-II based filter bank.
 - ii. Transition-band width constraint: In the ICDM-II operations, if the decimation factor is M , the of the lowpass and highpass filters obtained after coefficient decimation becomes M times that of the prototype filter. Therefore, in the ICDM-II based filter bank, the prototype filter has to be designed with a considerably narrower transition-band width which requires higher order filter so that all the
-

subbands obtained using ICDM-II operations have their transition-band widths within the desired specifications. If the required decimation factor values are large, this constraint will result in the requirement of a significantly high order prototype filter. This will increase the implementation complexity of the ICDM-II based filter bank, resulting in high hardware resource utilization and power consumption.

To overcome the uniform channelization limitation of the ICDM-I based filter bank as well as the design constraints involved in the ICDM-II based filter bank, a new filter bank is proposed in this section. The proposed filter bank employs the ICDM-I to obtain the desired subbands, while using the ICDM-II only to achieve variable bandwidths corresponding to different wireless communication standards. The proposed filter bank is termed as comprehensive ICDM based filter bank and it is presented in detail in the subsequent sections.

5.4.1 Design Procedure

The design procedure to realize the comprehensive ICDM based filter bank for various applications is presented in this section.

Consider a multi-standard channelization scenario where $S_1, S_2, \dots, S_n$ are n different communication standards whose channels are to be extracted from a wideband input signal sampled at a sampling frequency f_{samp} . Let $BW_1, BW_2, \dots, BW_n$ represent the channel bandwidth specifications and $TBW_1, TBW_2, \dots, TBW_n$ represent the transition-band width specifications corresponding to the n standards. Similarly, let $\delta_{p1}, \delta_{p2}, \dots, \delta_{pn}$ represent the passband peak ripple specifications and $\delta_{s1}, \delta_{s2}, \dots, \delta_{sn}$ represent the desired stopband attenuation specifications for the channels of the n standards. The steps to obtain a suitable comprehensive ICDM based filter bank to extract multiple channels of the n different communication standards simultaneously are described below.

Step-1: Normalize all the channel bandwidth and transition-band width specifications for the range 0 to 1, where 1 corresponds to $f_{\text{samp}}/2$.

Step-2: Calculate $BW_i' = BW_i / 2 \quad \forall \quad i = 1, 2, \dots, n$. Fix the passband width of the prototype filter ($BW_{\text{prototype}}$) as the greatest common divisor (GCD) of the computed values. This is represented as,

$$BW_{\text{prototype}} = \text{GCD} \{BW_1', BW_2', \dots, BW_n'\} \quad (5.3)$$

Step-3: The passband width of the prototype filter $BW_{\text{prototype}}$ is obtained using (5.3). Identify the decimation factor values required to perform ICDM-II operations on the prototype filter, for obtaining appropriate lowpass and highpass frequency responses with their passband widths corresponding to the channel bandwidths of n standards. Let these values be represented as D_i , where $i = 1, 2, \dots, n$.

Step-4: Calculate $TBW_i' = TBW_i / D_i \quad \forall \quad i = 1, 2, \dots, n$. Fix the transition-band width of the prototype filter ($TBW_{\text{prototype}}$) to be the minimum of the computed values. This is denoted as,

$$TBW_{\text{prototype}} = \min \{TBW_1', TBW_2', \dots, TBW_n'\} \quad (5.4)$$

where $\min \{x_1, x_2, \dots, x_n\}$ implies the smallest value among $\{x_1, x_2, \dots, x_n\}$.

Step-5: For each of the n standards and the corresponding D_i values identified in *Step-3*, use (3.1) and (3.10) to identify a set of decimation factor values $\{M_{i1}, M_{i2}, M_{i3}, \dots\}$ to be used in ICDM-I operations for obtaining frequency responses that can be used to obtain the desired subbands in the filter bank. Let the maximum value of decimation factor in each of the identified sets be denoted as $M_{i\max}$.

Step-6: Calculate $\delta_{si}' = \delta_{si} / (D_i \times M_{i\max}) \quad \forall \quad i = 1, 2, \dots, n$. Fix the stopband attenuation specification of the prototype filter ($\delta_{s(\text{prototype})}$) to be the minimum of the computed values. This is represented as,

$$\delta_{s(\text{prototype})} = \min \{ \delta_{s1}', \delta_{s2}', \dots, \delta_{sn}' \} \quad (5.5)$$

Step-7: Fix the passband peak ripple specification of the prototype filter ($\delta_{p(\text{prototype})}$) to be the minimum of those for the n standards involved. This is represented as,

$$\delta_{p(\text{prototype})} = \min \{ \delta_{p1}, \delta_{pn}, \dots, \delta_{pn} \} \quad (5.6)$$

Step-8: Use (3.14) to compute the order of the prototype filter corresponding to the specifications obtained in *Step-2*, *Step-4*, *Step-6*, and *Step-7*. Design the prototype filter and obtain its coefficients.

Step-9: Perform appropriate ICDM operations on the prototype filter with the different decimation factor values identified in *Step-3* and *Step-5* to obtain the corresponding frequency responses.

Step-10: Obtain the desired subbands by performing spectral subtraction of appropriate frequency responses obtained in *Step-9* and their corresponding complementary frequency responses if required. The complementary frequency response of a filter can be implemented by subtracting the output of the filter from an appropriately delayed version of the input signal.

Step-11: Design low order wide transition-band width frequency response masking filters if required, for obtaining individual subbands from the multi-band frequency responses obtained after performing ICDM-I operations. (Note: To ensure low complexity of the masking filters, choose the largest possible values for their transition-band widths according to the passband and stopband edge frequencies of the desired subband and its adjacent subbands in the corresponding multi-band frequency response. From (3.11), it can be noted that this results in lower order filters which ultimately reduces the complexity of the masking filters.)

5.4.2 Design Example and Comparison

The ability of the comprehensive ICDM based filter bank to perform uniform channelization and the corresponding performance characteristics are same as that of the ICDM-I based filter bank proposed in Section 5.2. This is because the comprehensive ICDM based filter bank can perform all the operations of ICDM-I based filter bank as well as the ICDM-II based filter bank, while reducing the impact of design constraints

involved in them. Therefore, a uniform channelization design example is not presented in this section as the results will be same as that of the design example presented in Section 5.2.2 for the ICDM-I based filter bank.

The ability of the comprehensive ICDM based filter bank to perform non-uniform channelization is functionally verified and illustrated in this section. Figure 5.9(a) shows the frequency spectrum of an input signal wherein four Bluetooth (BT) channels, one Zigbee channel and one WCDMA channel are simultaneously present at different locations in the wideband input frequency range.

The design of the proposed filter bank to extract each of the channels shown in Figure 5.9(a) simultaneously is demonstrated here. The channel bandwidths of Bluetooth, Zigbee and WCDMA standards are 1 MHz, 4 MHz and 5 MHz respectively. The transition-band width specifications for Bluetooth, Zigbee and WCDMA standards are chosen as 50 kHz, 200 kHz and 500 kHz respectively. The sampling frequency is 40 MHz. The desired passband and stopband peak ripple specifications are chosen as 0.1 dB and -40 dB for Zigbee and WCDMA channels, 0.1 dB and -55 dB for Bluetooth channels respectively.

The design procedure given in Section II-B is used to design a comprehensive ICDM based filter bank that can be used to extract the different channels shown in Figure 5.9(a).

Figure 5.9(a). Input signal containing Bluetooth, Zigbee and WCDMA channels.

Let S_1 =Bluetooth, S_2 = Zigbee and S_3 = WCDMA. Following *Step-1*, the normalized channel bandwidths of S_1 , S_2 and S_3 are computed to be 0.05, 0.2 and 0.25 respectively. The passband width of the prototype filter is chosen as 0.025, i.e., the GCD of {0.025, 0.1, 0.125}, using (5.3) in *Step-2*. Following *Step-3*, $D_1=1$, $D_2=4$, and $D_3=10$ are identified as the decimation factor values required to obtain the appropriate lowpass and highpass frequency responses with their passband widths corresponding to the channel bandwidths of the three standards. The transition-band width of prototype filter is then computed to be 0.0025 using (5.4) in *Step-4*. Corresponding to passband width and transition-band width values obtained using *Step-2* and *Step-4*, the passband and stopband edge frequency specifications of the prototype filter are chosen as $f_p = 0.0225$ and $f_s = 0.025$ respectively. Following *Step-5*, the three sets of decimation factor values to be used in ICDM-I operations to obtain the subbands corresponding to standards S_1 , S_2 , and S_3 are identified as {10}, {2}, and {1} respectively. Using (5.5), (5.6) in *Step-6* and *Step-7*, $\delta_{s(\text{prototype})} = -65$ dB and $\delta_{p(\text{prototype})} = 0.1$ dB are computed. Following *Step-8*, the prototype filter is designed with the computed values of its specifications and the corresponding filter order chosen as 2928 using (3.14). Following *Step-9*, appropriate ICDM operations are performed on the designed prototype filter using the identified decimation factor values and the corresponding output frequency responses are obtained. Following *Step-10* and *Step-11*, the desired subbands in the filter bank are derived from these output frequency responses as described below.

Figure 5.9(b) shows the various stages in the comprehensive ICDM based filter bank designed for extracting the Bluetooth, Zigbee and WCDMA channels shown in Figure 5.9(a). CDM-I operation is performed on the prototype filter using $M=10$ to obtain the frequency subband corresponding to the Bluetooth channel BT2 located between 3.5-4.5 MHz. A masking filter of order 39, designed using (3.11), is used to extract channel BT2 from the multi-band frequency response obtained as shown in Figure 5.9(b). To extract the other three Bluetooth channels BT1, BT3 and BT4, located between 1.5-2.5 MHz, 5.5-6.5 MHz and 13.5-14.5 MHz respectively, the frequency response obtained after performing MCDM-I operation using $M=10$ is used. The three channels are individually extracted using three low order masking filters each of order 39, designed using (3.11).

Figure 5.9(b). Comprehensive ICDM based filter bank: Channelization design example.

The WCDMA channel located between 15 MHz-20 MHz is extracted using the frequency response obtained after performing MCDM-II operation using $M=10$, as shown in Figure 5.9(b). To obtain frequency subband corresponding to the Zigbee channel located between 8 MHz-12 MHz, CDM-II operation is first performed on the prototype filter using $M=4$ so that the bandwidth corresponding to the Zigbee channel is obtained. Then MCDM-I operation is performed on the resultant filter using $M=2$ to obtain the subband which can be used to extract the Zigbee channel. Thus, all the different channels corresponding to three wireless communication standards that are simultaneously present in the input signal as shown in Figure 5.9(a) can be extracted using the comprehensive ICDM based filter bank.

The modulated perfect reconstruction filter bank (MPRFB) [79] can be used to extract the different frequency channels shown in Figure 5.9(a). However, multiple filter banks need to be designed in the MPRFB approach because the bandwidths of Zigbee and WCDMA standards are not integer multiples of each other. This could result in complex and inefficient implementations as different prototype filters along with their corresponding DFT as well as IDFT computation structures need to be separately designed and implemented.

From (3.1), it can be noted that if CDFB [90] is designed to extract the different frequency channels shown in Figure 5.9(a), CDM-I operation using $M=20$ needs to be performed to obtain a frequency response containing the subbands which can be used for extracting the channels BT1, BT3 and BT4. Having the ability to perform MCDM-I operation in the comprehensive ICDM based filter bank makes it possible to extract these channels using $M=10$, which is half of the value of M required in the CDFB [90] case. From (3.14), it can be noted that while a prototype filter of order 2928 is required in the comprehensive ICDM based filter bank (wherein the maximum required value of $M=10$), the order of the prototype filter required in the CDFB design is 3088 (wherein the maximum required value of $M=20$). The frequency response obtained after performing CDM-I operation using $M=20$ has twice the number of subbands than the frequency response obtained after performing MCDM-I operation using $M=10$. Thus, in the CDFB design, masking filters required to extract the channels BT1, BT3 and BT4 need to be designed with narrower transition-band widths than those required in the comprehensive ICDM based filter bank. From (3.1) and (3.10), it can be noted that three masking filters of order 119 each are required in the CDFB to extract the channels BT1, BT3 and BT4. To extract the channel BT2, the same masking filter of order 39 that is used in the comprehensive ICDM based filter bank can be employed in the CDFB.

5.4.2.1 Complexity Analysis

Table 5.4 summarizes the number of multiplications required for implementation of the comprehensive ICDM based filter bank and CDFB [90] that are designed for

Table 5.4. Comprehensive ICDM based filter bank: Multiplication complexity comparison.

	CDFB [90]	Proposed comprehensive ICDM based filter bank
Prototype filter length (L_{Mod})	3089	2929
Masking filter length (L_{Mask})	$(40) + (120 \times 3) = 400$	$40 \times 4 = 160$
No. of multiplications for filter implementation $= (\lceil L_{\text{Mod}} / 2 \rceil + \lceil L_{\text{Mask}} / 2 \rceil)$	$\{\lceil 3089 / 2 \rceil + \lceil 40 / 2 \rceil + (3 \times \lceil 120 / 2 \rceil)\}$ $= 1745$	$\{\lceil 2929 / 2 \rceil + (4 \times \lceil 40 / 2 \rceil)\}$ $= 1545$
Total no. of multiplications	1745	1545

extracting the different channels shown in Figure 5.9(a). From Table 5.4, it can be noted that comprehensive ICDM based filter bank offers a multiplication complexity reduction of 11.46% over CDFB. The comprehensive ICDM based filter bank has a lower complexity than the CDFB due to the lower orders of the prototype filter and the masking filters used in it.

As discussed in Section 5.2 for the ICDM-I based filter bank, the comprehensive ICDM based filter bank also offers twice the center frequency resolution for its constituent subbands and can extract a significantly larger number of distinctly located frequency channels when compared with the CDFB.

To design the comprehensive ICDM based filter bank for uniform channelization, the multi-band frequency responses obtained after performing ICDM-I operations are algebraically operated upon using appropriate spectral subtraction, complementary filter response operation and frequency response masking operations to obtain the desired subbands in the filter bank. In case of multi-standard channelization scenarios, appropriate ICDM-II operations are performed merely for achieving the distinct bandwidths corresponding to different communication standards. The ICDM-I operations are then employed to obtain the corresponding desired non-uniform subbands. Thus, unlike the use of ICDM-II operations to obtain the desired subbands in the ICDM-II based filter bank proposed in Section 5.3, the comprehensive ICDM based filter bank uses

ICDM-I operations to obtain the desired subbands and restricts the use of ICDM-II operations merely to obtain variable bandwidths. The use of ICDM-I operations to obtain the desired subbands eliminates the group delay constraint from the proposed filter bank design method, because no group delay compensation is required to be performed for spectrally subtracting the ICDM-I output frequency responses. This is because the filters obtained after performing ICDM-I operations have the same filter order and therefore same group delay as that of the prototype filter. The transition-band width constraint is present only in multi-standard channelization scenarios but it does not affect the complexity of the filter bank significantly as ICDM-II operations are merely used to obtain variable bandwidths corresponding to the different communication standards and not to obtain the desired subbands in the filter bank. This reduces the complexity of the comprehensive ICDM based filter bank when compared with the ICDM-II based filter bank and the PDFB [89].

5.5 Summary

Based on the ICDM proposed in Chapter 3, three filter banks were proposed in this chapter. An ICDM-I based filter bank was proposed for uniform channelization. It performs ICDM-I operations to provide variable frequency responses from which the desired uniform subbands can be obtained. The ICDM-I based filter bank has lower implementation complexity and twice the flexibility in terms of the possible number and locations of its subbands when compared with the discrete Fourier transform based filter bank (DFTFB) and the conventional CDM based filter bank (CDFB). In the design example considered, the ICDM-I based filter bank achieved 70.92% and 15.49% reductions in resource utilizations and 58.72% and 23.13% reductions in power consumptions when compared with the DFTFB and CDFB respectively.

An ICDM-II based filter bank was proposed for multi-standard channelization. It performs ICDM-II operations to obtain variable lowpass and highpass frequency responses, which are spectrally subtracted to obtain the desired uniform and non-uniform subbands. The ICDM-II based filter bank shows significant reduction in complexity and

improvement in stopband and transition-band characteristics when compared with the conventional CDM based progressive decimation filter bank (PDFB). In the design example considered, the ICDM-II based filter bank achieved 74.01% reduction in multiplication complexity when compared with a PDFB designed for a specific non-uniform multi-standard channel distribution. It also achieved 95.67% reduction in multiplication complexity when compared with a flexible PDFB designed for variable subband locations.

Another filter bank was proposed based on the comprehensive ICDM. It can perform all the operations of ICDM-I based filter bank as well as the ICDM-II based filter bank, while reducing the impact of design constraints involved in them. Design example showed that the comprehensive ICDM based filter bank is a low complexity alternative to the other relevant filter banks in literature and can be used for uniform as well as non-uniform channelization in SDR receivers.

When compared with other relevant filter banks in literature, the proposed filter banks offer the advantages of low complexity and high frequency response flexibility. They are compatible for use in SDR base-station receivers and can be used for uniform/ non-uniform channelization as well as spectrum sensing.

In the next chapter, a design technique based on the combination of all pass transformation (APT) and ICDM is proposed to obtain variable digital filters (VDFs) with unabridged control over the cutoff frequency. Based on the proposed VDF, a low complexity spectrum sensing scheme is also proposed. Pipelined implementation architectures for realizing high speed 1st and 2nd order APT based VDFs are also presented.

Chapter 6

Design of Variable Digital Filters based on All Pass Transformation and Improved Coefficient Decimation Method

6.1 Introduction

All pass transformation (APT) based variable digital filters (VDFs), also known as frequency warped VDFs, were proposed in [67]. In an APT based VDF, all pass filter structures of appropriate order (1st or 2nd) are used to replace delay elements in a prototype filter structure. The resultant filter can provide variable lowpass, highpass, bandpass and bandstop frequency responses with unabridged control over bandwidths and center frequencies in the entire Nyquist frequency range, without updating the prototype filter coefficients. The APT based VDFs have lower implementation complexity when compared with other relevant VDFs in literature which can provide unabridged control over cutoff frequencies. The improved coefficient decimation method (ICDM) based VDFs proposed in Chapter 4 have the advantages of low complexity and high frequency response flexibility but they offer only discrete control over the cutoff frequencies of subbands. If the APT technique and the ICDM are combined, it will be possible to realize VDFs with low complexity, high frequency response flexibility and unabridged control

over cutoff frequencies. These characteristics of APT and ICDM and their possible combination are the motivation behind the work presented in this chapter.

A VDF based on the combination of improved coefficient decimation method I (ICDM-I) and the 1st order APT technique is presented in Section 6.2. In Section 6.3, a spectrum sensing scheme is proposed for cognitive radio (CR) receivers, which employs VDFs based on the combination of 1st order APT and the modified coefficient decimation method I (MCDM-I). Section 6.4 presents generalized design procedures and pipelined hardware implementation architectures for realizing high speed APT based VDFs along with the corresponding implementation results. The works described in Sections 6.2, 6.3 and 6.4 of this chapter have been published in [C4], [C5] and [J5] respectively.

In the next section, a brief review of the 1st and 2nd order APT based VDF design techniques is presented along with their corresponding hardware implementation architectures.

6.1.1 1st order All Pass Transformation based Variable Digital Filters

In 1st order APT based VDFs, every delay element in a prototype filter architecture is replaced by a 1st order all pass filter structure [67]. If $H(z)$ is the z -domain representation of a prototype filter, $A(z)$ is the z -domain representation of 1st order all pass filter and $G(z)$ denotes the warped version of $H(z)$, then

$$G(z) = H(A(z)) \quad (6.1)$$

where $A(z) = \left[\frac{z^{-1} - \alpha}{1 - \alpha z^{-1}} \right]$, $|\alpha| < 1$ and is real.

α is called the 1st order warping coefficient. Figure 6.1(a) shows the 1st order APT based VDF implementation architecture. It is a transposed direct-form finite impulse response (FIR) filter implementation structure in which each delay element is replaced by $A(z)$. To exploit the symmetry property of prototype filter coefficients, only half of the coefficients are implemented as shown in Figure 6.1(a), thus reducing the implementation complexity

of the VDF. Figure 6.1(b) shows an implementation architecture of a 1st order all pass filter $A(z)$ [68]. The value of α determines the cutoff frequency in the resultant frequency response obtained using the implementation architecture shown in Figure 6.1(a). For a lowpass prototype filter, if $-1 < \alpha < 0$, the cutoff frequencies in the resultant frequency responses are higher than the original cutoff frequency. Similarly, for $0 < \alpha < 1$, the cutoff frequencies in the resultant frequency responses are lower than the original cutoff frequency. For $\alpha = 0$, the original frequency response of the prototype filter is obtained.

In 1st order APT based VDFs, variable highpass frequency responses can be obtained using a lowpass prototype filter by substituting every delay element in the filter structure

Figure 6.1(a). 1st order APT based VDF: Hardware implementation architecture.

Figure 6.1(b). 1st order all pass filter: Hardware implementation architecture.

by $-A(z)$. The flexibility to perform substitution of delay elements by either $A(z)$ or $-A(z)$ can be incorporated in the VDF implementation architecture using multiplier-multiplexer pairs as proposed in [69] or by using the lower complexity MCDM-I operation. Variable lowpass and highpass frequency responses with desired cutoff frequencies can thus be obtained using the same lowpass prototype filter in a 1st order APT based VDF. In case of bandpass and bandstop prototype filters, different values of α lead to modification of center frequencies as well as bandwidths of the resultant bandpass and bandstop frequency responses. This is because tuning of a single parameter α allows control of only a single frequency edge in the resultant frequency responses. To obtain variable bandpass and bandstop frequency responses with independent and unabridged control over both their frequency edges, 2nd order APT needs to be performed which provides 2 tunable parameters that can be separately controlled.

6.1.2 2nd order All Pass Transformation based Variable Digital Filters

In the 2nd order APT based VDFs, every delay element in a prototype filter architecture is replaced by a 2nd order all pass filter structure [67]. If $H(z)$ is the z -domain representation of a prototype filter, $B(z)$ is the z -domain representation of 2nd order all pass filter and $F(z)$ denotes the warped version of $H(z)$, then

$$F(z) = H(B(z)) \quad (6.2)$$

where $B(z) = \left[\frac{z^{-2} - \beta_2 z^{-1} + \beta_3}{1 - \beta_2 z^{-1} + \beta_3 z^{-2}} \right]$, $|\beta_3| < 1$, and both β_2 and β_3 are real.

β_2 and β_3 are called the 2nd order warping coefficients. Figure 6.2(a) shows the 2nd order APT based VDF implementation architecture. It is a transposed direct-form FIR filter implementation structure in which each delay element is replaced by $B(z)$. Figure 6.2(b) shows an implementation architecture of a 2nd order all pass filter $B(z)$ [68]. The values of β_2 and β_3 can be independently varied to determine cutoff frequencies in the frequency responses obtained using the implementation architecture shown in Figure 6.2(a).

Figure 6.2(a). 2nd order APT based VDF: Hardware implementation architecture.

Figure 6.2(b). 2nd order all pass filter: Hardware implementation architecture.

In the 2nd order APT based VDFs, variable lowpass, highpass, bandpass and bandstop frequency responses can all be obtained using a single lowpass prototype filter. Variable bandstop frequency responses can be obtained using a lowpass prototype filter by substituting every delay element in the filter structure by $-B(z)$. Unlike the 1st order APT based VDF, availability of 2 distinctly tunable parameters (β_2 and β_3) in the 2nd order APT based VDFs enables independent control of both frequency edges in the bandpass and bandstop frequency responses that are obtained. Thus any desired center frequency location and bandwidth can be achieved in the variable frequency responses obtained using 2nd order APT based VDFs.

6.2 Proposed ICDM and 1st order All Pass Transformation based Variable Digital Filter

In this section, a VDF based on the 1st order APT and the earlier proposed ICDM-I is presented. The proposed VDF is termed as APT-ICDM-I based VDF and it can provide variable lowpass, highpass, bandpass and bandstop frequency responses using a single lowpass prototype filter. In the proposed VDF, ICDM-I is used to generate the four types of frequency responses and frequency warping using 1st order APT is performed to achieve variable cutoff frequencies for the obtained responses. The nature of obtained frequency responses and their cutoff frequencies can be changed on-the-fly without updating the filter coefficients. With the help of a design example, it is shown that the APT-ICDM-I based VDF design has a significantly lower implementation complexity when compared with other relevant VDF design methods in literature. Subsequent sections present the APT-ICDM-I based VDF in detail.

6.2.1 Hardware Implementation Architecture

Figure 6.3 shows the hardware implementation architecture for the APT-ICDM-I based VDF. The prototype filter coefficients are implemented using transposed direct-form filter

Figure 6.3. APT-ICDM-I based VDF: Hardware implementation architecture.

structure, to exploit the symmetry property of the coefficients. The VDF architecture consists of 2:1 multiplexers (*mux*) that are used to retain the prototype filter coefficients or replace them by zeros while performing appropriate ICDM-I operations. The multiplexers are used only for performing ICDM-I operations by $M=2$, and hence are placed only at alternate filter coefficients in the implementation architecture. A simple digital logic is implemented in the ‘*Decimation Selector Logic*’ block to drive the select lines (*S*) of multiplexers appropriately (represented in Figure 6.3 using dashed lines). The adder/subtractor (*add/sub*) blocks are used to perform sign reversal of the alternate retained coefficients in an MCDM-I operation. Similar to the multiplexers, the *Decimation Selector Logic* drives the select lines (*sel*) of *add/sub* blocks appropriately. The resulting outputs after performing ICDM-I operations need to be scaled by M to obtain the original magnitude response. This is done by multiplying the output of the VDF by M as shown in Figure 6.3.

To perform frequency warping in the APT-ICDM-I based VDF, every delay element in the prototype filter architecture is replaced by a 1st order all pass filter structure. In Figure 6.3, these 1st order all pass filter structures are represented as $A(z)$. There are multiple possible architectures for implementing all pass filter structures [68]. The implementation architecture for $A(z)$ shown in Figure 6.1(b) is used in the proposed VDF.

6.2.2 Design Procedure

A generalized procedure to design the APT-ICDM-I based VDF and obtain the corresponding variable frequency responses is presented in this section. The corresponding mathematical formulae required to compute the various parameters are also provided.

Step-1: Design the prototype filter and obtain its coefficients. To meet the desired stopband attenuation specifications, use (3.14) to compute the minimum required order of the prototype filter, by substituting $M = 2$.

Step-2: Implement the prototype filter coefficients according to the hardware implementation architecture shown in Figure 6.3.

The following steps describe how different types of variable frequency responses can be obtained using the proposed VDF hardware implementation architecture -

Step-3(a): Variable lowpass frequency responses -

Variable lowpass frequency responses can be obtained by performing 1st order APT alone. ICDM operation is not required to obtain lowpass responses. Let f_{c1} be the normalized cutoff frequency of the prototype filter. Let f_{c2} be the desired normalized cutoff frequency of the lowpass frequency response to be obtained. The corresponding value of α that is required to achieve f_{c2} can be computed using the formula [63],

$$\alpha = \frac{\sin[(f_{c1} - f_{c2})\pi / 2]}{\sin[(f_{c1} + f_{c2})\pi / 2]} \quad (6.3)$$

Step-3(b): Variable highpass frequency responses -

Variable highpass frequency responses can be obtained by performing MCDM-I operation by $M=1$ and the 1st order APT. If f_{c2} is the desired cutoff frequency of the highpass frequency response to be obtained, the required value of α can be computed using the formula,

$$\alpha = \frac{\sin\{(1 - f_{c1}) - f_{c2}\}\pi / 2}{\sin\{(1 - f_{c1}) + f_{c2}\}\pi / 2} \quad (6.4)$$

Step-3(c): Variable bandpass frequency responses -

Variable bandpass frequency responses can be obtained by performing MCDM-I operation by $M=2$ and the 1st order APT. If f_{c2} is the desired upper cutoff frequency of the bandpass frequency response to be obtained, the required value of α can be computed using the formula,

$$\alpha = \frac{\sin\{(0.5 + f_{c1}) - f_{c2}\}\pi / 2}{\sin\{(0.5 + f_{c1}) + f_{c2}\}\pi / 2} \quad (6.5)$$

Alternatively, if f_{c2} is the desired lower cutoff frequency of the bandpass frequency response to be obtained, the required value of α can be computed using the formula,

$$\alpha = \frac{\sin\{[(0.5 - f_{c1}) - f_{c2}]\pi / 2\}}{\sin\{[(0.5 - f_{c1}) + f_{c2}]\pi / 2\}} \quad (6.6)$$

Step-3(d): Variable bandstop frequency responses -

Variable bandstop frequency responses can be obtained by performing coefficient decimation method I (CDM-I) operation by $M=1$ and the 1st order APT. If f_{c2} is desired lower cutoff frequency of the bandstop frequency response to be obtained, the required value of α can be computed using (6.5). If f_{c2} is the desired upper cutoff frequency of the bandstop frequency response to be obtained, the required value of α can be computed using (6.6).

6.2.3 Design Example

In this section, a design example is presented to illustrate the realization of the APT-ICDM-I based VDF using the design procedure given in Section 6.2.3. Let the maximum acceptable passband and stopband peak ripple specifications of the frequency responses to be obtained be 0.1 dB and -55 dB respectively. The normalized passband and stopband edge frequency specifications of the prototype filter are chosen as 0.1 and 0.14 respectively. Following *Step-1* of the design procedure, the prototype filter order is computed as 160 using (3.14) and its coefficients are obtained. Following *Step-2*, the prototype filter coefficients are implemented using the hardware implementation architecture shown in Figure 6.3, wherein the delay elements are replaced by 1st order all pass filter structures shown in Figure 6.1(b). Figures 6.4(a) - 6.4(d) show some of the different frequency responses obtained using the designed VDF. It is to be noted that a large number of distinct frequency responses can be obtained over the entire Nyquist frequency range but only some responses are shown here for illustration purpose. Figure 6.4(a) shows few of the different lowpass frequency responses obtained using appropriate values of α , as described in *Step-3(a)*. Figure 6.4(b) shows few of the different highpass

frequency responses obtained by performing MCDM-1 operation with $M=1$ and using appropriate values of α , as described in *Step-3(b)*. Similarly, Figures 6.4(c) and 6.4(d) show few of the different bandpass and bandstop frequency responses obtained by following *Step-3(c)* and *Step-3(d)* respectively (Note: In Figures 6.4(c) and 6.4(d), f_c denotes the upper cutoff frequency). Thus it can be noted that the proposed APT-ICDM-I based VDF can provide variable lowpass, highpass, bandpass and bandstop frequency responses having desired cutoff frequencies, by performing appropriate ICDM-I operations and using suitable values of α , which can be computed using (6.3) - (6.6). From Figures 6.4(a)-6.4(d), it can be observed that changes in the cutoff frequency in the resultant frequency responses are accompanied by changes in the transition bandwidth as well. The resultant transition bandwidth for a particular value of α (computed for a desired f_c setting) can be mathematically computed using the following equation which is derived by solving (6.3) for f_{c2} –

$$f_{c2} = \frac{2}{\pi} \arctan \left[\left(\frac{1-\alpha}{1+\alpha} \right) \tan \left(f_{c1} \frac{\pi}{2} \right) \right] \quad (6.7)$$

The passband and stopband edge frequencies of the prototype filter (say f_{p1} and f_{s1}) can be substituted in (6.7) to obtain the passband and stopband edge frequencies of the resultant frequency response (say f_{p2} and f_{s2}). The resultant transition bandwidth is the difference between the resultant edge frequencies f_{p2} and f_{s2} .

Figure 6.4(a). Variable lowpass frequency responses obtained by performing 1st order APT.

Figure 6.4(b). Variable highpass frequency responses obtained by performing MCDM-I with $M=1$ and 1st order APT.

Figure 6.4(c). Variable bandpass frequency responses obtained by performing MCDM-I with $M=2$ and 1st order APT.

Figure 6.4(d). Variable bandstop frequency responses obtained by performing CDM-I with $M=2$ and 1st order APT.

6.2.4.1 Complexity Analysis

In this section, the APT-ICDM-I based VDF is compared with other relevant VDFs which perform frequency warping using 1st order APT. The design approach wherein four parallel branches of different types of prototype filters are implemented, having 1st order all pass filter structures substituting delay elements in each [67], is termed as *Design-A*. Therefore, in *Design-A*, three 2:1 multiplexers are used to select the appropriate branch and obtain the desired frequency response at the output. The design approach containing 1st order APT based sections of the VDF architecture proposed in [69] is denoted as *Design-B*. Note that only variable lowpass and highpass frequency responses can be obtained using *Design-B*. To obtain variable bandpass or bandstop frequency responses, the VDF design approach in [69] employs 2nd order APT. The prototype filters in *Design-A* and *Design-B* are designed with the same frequency response specifications as the prototype filter in the proposed VDF, and their filter orders computed using (3.14) are 150 for each. Table 6.1 presents the complexity comparison of the three VDF designs in terms of the number of multipliers, 2:1 multiplexers and adders required for their implementation and the corresponding total gate-count for each. APT-ICDM-I based VDF offers 73.23% and 73.33% reductions in the required number of multipliers and adders respectively, when compared with *Design-A*. This is because *Design-A* requires a large number of multipliers and adders due to the parallel implementation of four types of prototype filter branches and the 1st order all pass filter structures involved in each branch. *Design-B* requires 55.37% higher number of multipliers than the APT-ICDM-I based VDF. This is due to use of one additional multiplier per filter coefficient to perform lowpass to highpass frequency response transformation in *Design-B*. These additional

Table 6.1. APT-ICDM-I based VDF: Complexity comparison.

	<i>Design-A</i> (VDF with 4 parallel branches)	<i>Design-B</i> (VDF in [69])	Proposed APT- ICDM-I based VDF
Multipliers	904	376	242
2:1 Multiplexers	3	150	40
Adders	1800	450	480
Total gate-count	1158259	457990	311180

multipliers are not required in the APT-ICDM-I based VDF due to the use of ICDM-I operations to generate the different types of frequency responses from the prototype filter. When compared with *Design-B*, the APT-ICDM-I based VDF offers 73.33% reduction in the number of 2:1 multiplexers required for its implementation. The APT-ICDM-I based VDF requires 6.67% more adders for its implementation when compared with *Design-B* due to the higher order prototype filter used in the former. However, this overhead is low compared to the substantial reductions in the number of multipliers and multiplexers offered by the APT-ICDM-I based VDF. This can be noted from the total gate-count values of the three designs. The number of NAND gates required for implementing a multiplier, a 2:1 multiplexer and an adder are computed. Using these values, total gate-counts are computed for the three VDF designs by summing the gate-counts of all the components required for their implementation. The proposed APT-ICDM-I based VDF achieves 73.13% and 32.06% reductions in the total gate-count when compared with *Design-A* and *Design-B* respectively.

6.3 Variable Digital Filter based Spectrum Sensing Scheme for Cognitive Radio Receivers

CRs have the ability to sense the current spectrum utilization and change their behavioral and transmission characteristics dynamically so as to achieve efficient spectrum usage [12]. Also, dynamic spectrum management using CRs can offer significant energy savings in radio networks [35]. Thus, significant energy savings coupled with the ability to perform dynamic spectrum management empowers CRs for applications such as air-ground communication systems [101, 102]. The current system of air-ground communication follows static spectrum allocation, which is a bottleneck for effective usage of the limited licensed spectrum allotted for air-ground communication [101]. The use of CRs for air-ground communication will not only help in solving the static spectrum allocation problem but also in achieving energy efficiency (in terms of transmitted power) in the battery-operated radio terminals. An important function in CR receivers is spectrum sensing, wherein spectrum holes are to be detected. Out of the

various techniques used for spectrum sensing in CR receivers, the energy detector based spectrum sensing has the least implementation complexity [27, 28]. It uses digital filters or filter banks to detect spectrum holes by comparing signal energy in the corresponding frequency bands with a threshold value. In ideal CR spectral scenarios, it is desired to detect spectrum holes of varying bandwidths and locations in the entire Nyquist frequency range. For this purpose, VDFs that provide frequency responses with unabridged control over cutoff frequencies can be used to realize frequency bands with varying bandwidths and locations. As discussed in Section 2.2.2, APT based VDFs can provide such variable frequency responses while also achieving low complexity implementations. Therefore, APT based VDFs are feasible candidates to realize low complexity spectrum sensing schemes with ability to detect spectrum holes of varying bandwidths and locations over the entire Nyquist frequency range.

A spectrum sensing scheme is proposed in this section for software defined radio (SDR) based CR receivers. The proposed spectrum sensing scheme employs VDFs based on 1st order APT and the MCDM-I to obtain variable frequency bands, followed by the energy detection technique to detect the presence/ absence of radio signals in them. The VDFs used in the proposed scheme are low complexity alternatives to the other relevant VDFs in literature. Subsequent sections present the proposed VDF based spectrum sensing scheme along with the corresponding hardware implementation architecture, design procedure and a design example.

6.3.1 Hardware Implementation Architecture

Figure 6.5(a) shows the block diagram of the proposed VDF based spectrum sensing scheme. The major components of the scheme are the two VDFs involved, VDF-I and VDF-II. Figure 6.5(b) shows the hardware implementation architecture for VDF-I and VDF-II. As shown in Figure 6.5(b), a single prototype filter is used for realizing both VDF-I and VDF-II. The prototype filter coefficients can be implemented using the transposed direct-form filter structure to exploit their symmetry property. The use of a single set of filter coefficients reduces the complexity by minimizing the number of multipliers required. VDF-I is implemented using the 1st order APT based VDF

implementation architecture shown in Figure 6.1. The VDF-I branch can provide lowpass frequency responses with variable cutoff frequencies at its output, by employing 1st order APT in its design. The hardware realization of VDF-II is shown in Figure 6.5(b). The lowpass/ highpass response toggle line is used to perform MCDM-I operation using $M=1$ when in highpass mode, thereby converting the resultant lowpass frequency responses after 1st order APT into corresponding highpass frequency responses. Based on the combination of 1st order APT and MCDM-I, the VDF-II branch can thus provide lowpass and highpass frequency responses with variable cutoff frequencies at its output. α_1 and α_2 represent the frequency warping coefficients required to perform 1st order APT in VDF-I and VDF-II respectively. The energy computation blocks shown in Figure 6.5(a) are used

Figure 6.5(a). Block diagram of proposed VDF based spectrum sensing scheme.

Figure 6.5(b). Hardware implementation architecture for VDF-I and VDF-II in the proposed scheme.

to compute energy values of the filtered output signals of VDF-I and VDF-II branches, which are termed as E_1 and E_2 respectively. The 2:1 multiplexer (*mux*) shown in Figure 6.5(a) is used to select the appropriate computed energy value, $(E_2 - E_1)$ or E_2 , to be passed on to the band occupancy decision block. The band occupancy decision block is used to compare the input energy value with a threshold to determine whether the desired frequency band is vacant or occupied.

6.3.2 Design Procedure

The design procedure to realize the VDF based spectrum sensing scheme is presented in this section.

Step-1: Prototype filter design –

Design the lowpass prototype filter to be employed in VDF-I and VDF-II (shown in Figures 6.5(a) and 6.5(b)) and obtain the corresponding filter coefficients. The minimum order of prototype filters that meets the desired frequency response specifications can be computed using (3.11).

Step-2: Implementation of proposed scheme –

Implement the proposed scheme as discussed in Section 6.3.1, using the block diagram and hardware implementation architecture shown in Figures 6.5(a) and 6.5(b) respectively.

Step-3(a): Energy computation for a frequency band from 0 to cutoff frequency f_c ($0 < f_c < 1$, where 1 corresponds to half of sampling frequency), i.e., a lowpass frequency response –

The VDF-II branch alone can be used in this case. Obtain the desired lowpass frequency response by keeping the toggle line in lowpass mode and achieve the desired cutoff frequency using an appropriate value of α_2 . The required value of α_2 to obtain the desired lowpass frequency response can be computed using (6.3).

Compute energy E_2 for the desired frequency band achieved using VDF-II and pass it on to the occupancy decision block by setting the *mux* select line appropriately. The energy E of a digital signal $x(n)$ of R samples can be computed using the formula [42]

$$E = \frac{1}{R} \sum_{n=0}^{R-1} |x(n)|^2 \quad (6.8)$$

Step-3(b): Energy computation for a frequency band from f_{cl} to f_{cu} (where f_{cl} and f_{cu} are normalized lower and upper cutoff frequencies respectively) such that $0 < f_{cl} < f_{cu} < 1$ –

In this case, obtain two distinct lowpass frequency responses by achieving f_{cl} in VDF-I and f_{cu} in VDF-II branches respectively, using (6.3). Perform energy computation for the two frequency bands (0 to f_{cl} and 0 to f_{cu} , $E_2 \geq E_1$) using (6.8). The difference between the computed energy values, i.e., $(E_2 - E_1)$ is the energy of the desired frequency band between f_{cl} and f_{cu} and can be passed on the occupancy decision block by setting the *mux* select line appropriately.

Step-3(c): Energy computation for a frequency band from f_c to 1 ($0 < f_c < 1$, where 1 corresponds to half of sampling frequency), i.e., a highpass frequency response –

Similar to the case discussed in *Step-3(a)*, the VDF-II branch alone can be used in this case. Obtain the desired highpass frequency response by keeping the toggle line in highpass mode and use an appropriate value of α_2 to achieve the desired cutoff frequency. The required value of α_2 to obtain the desired highpass frequency response can be computed using (6.4).

Using (6.8), compute energy E_2 for the desired frequency band achieved using VDF-II and pass it on to the occupancy decision block by setting the *mux* select line appropriately.

Step-4: Band occupancy decision –

Let E be the computed energy corresponding to the frequency band that is analyzed to find whether the band is occupied or vacant. The occupancy decision can be taken by comparing E with a pre-determined threshold value which is selected based on the

knowledge of noise variance [8]. If E is greater than or equal to the threshold, the band is occupied; else the band is vacant and can be used for the CR user's transmission. This comparison operation is performed by the band occupancy decision block shown in Figure 6.5(a).

6.3.3 Design Example

The functioning of the proposed spectrum scheme is illustrated in this section with the help of a CR based air-ground communication system design example. In current air-ground voice communication systems, the licensed band of 108-137 MHz is split into uniform sub-channels of 25 kHz bandwidth whose assignment is static and based on geographical areas and organizational structures [101]. With increasing demand for more voice channels, the 25 kHz channels are further being divided in 3 sub-channels of 8.33 kHz each, which is currently being practiced in Europe. Along with voice communication, transmission of data for applications such as aircraft communication addressing and reporting system also requires allocation of spectrum. The proposed spectrum sensing scheme can enable the identification and allocation of spectrum holes of varying bandwidths and locations for different communication applications. Thus by enabling CR based dynamic spectrum access, the proposed VDF based spectrum sensing scheme can help to overcome the static spectrum allocation problem in current air-ground communication systems, and achieve intelligent and efficient usage of the limited licensed spectrum.

Consider a spectrum sensing scenario wherein spectrum holes are to be detected in a wideband frequency range of 2.5 MHz from 108-110.5 MHz, within the 108-137 MHz air-ground communication frequency range. Figure 6.6(a) shows an input signal spectrum wherein three spectrum holes (SH1 of bandwidth 8.33 kHz, SH2 and SH3 of bandwidth 25 kHz) exist at different locations as shown. Figure 6.6(b) shows the three pairs of frequency responses obtained using VDF-I (provides f_{cl}) and VDF-II (provides f_{cu}) in the proposed scheme, which are used to detect the spectrum holes shown in Figure 6.6(a). The VDF-I and VDF-II outputs can be appropriately coupled to detect spectrum holes of

Figure 6.6(a). Input signal spectrum with three spectrum holes (SH1, SH2, SH3).

Figure 6.6(b). Frequency response pairs obtained using VDF-I and VDF-II, which are used to detect the spectrum holes in Figure 6.6(a).

varying bandwidth and location in the input frequency range. For the specifications $f_p = 0.4985$, $f_s = 0.5015$, $\delta_p = 0.1$ dB and $\delta_s = -40$ dB, a prototype filter is designed having its order $N = 1328$ computed using (3.11). The total number of multiplications required to implement corresponding filter coefficients and realize the VDF branches in the proposed scheme is 3321. For a conventional sensing scheme wherein two lowpass and one highpass prototype filter branches are implemented, the total number of multiplications required to implement the filter branches is 5314. It is 60% higher than the proposed

scheme mainly due to distinct sets of filter coefficients implemented in the lowpass and highpass filter branches. For a sensing scheme based on the VDF design approach proposed in [69], the total number of multiplications required to implement the filter branches is 4649. It is 40% higher than the proposed scheme due to use of one additional multiplier per filter coefficient to perform lowpass to highpass frequency response transformation. The cost of energy computation and spectrum hole detection is omitted from this comparison because it is equal in all the three discussed sensing schemes.

6.4 Implementation of High Speed All Pass Transformation based Variable Digital Filters

In wireless communication receivers where high bandwidth signals are processed, it is desired to employ high speed VDFs which can operate at high sampling frequencies. To achieve hardware realizations with high speed of operation, pipelined hardware implementation architectures for the conventional 1st and 2nd order APT based VDFs as well as the APT-ICDM-I based VDF proposed in Section 6.2 are presented in this section. First, generalized design procedures to realize 1st and 2nd order APT based VDFs are presented. Thereafter, FPGA implementation results of different designs of the 1st and 2nd order APT based VDFs and the proposed APT-ICDM-I based VDF using both non-pipelined as well as pipelined implementation architectures are presented and analyzed. It is shown that for all the different types of APT based VDFs considered, the proposed pipelined implementation architectures enable similar high operating frequencies that are independent of the prototype filter order.

6.4.1 Design Procedures for 1st and 2nd order All Pass Transformation based Variable Digital Filters

This section presents generalized design procedures that can be used to realize 1st and 2nd order APT based VDFs.

6.4.1.1 1st order All Pass Transformation based Variable Digital Filters

Step-1: Prototype filter design -

Design a prototype filter to meet the desired frequency response specifications and obtain its coefficients. The minimum required order N of the prototype filter can be computed using the (3.11).

Step-2: Implementation of VDF -

Implement the prototype filter coefficients using the hardware implementation architecture shown in Figure 6.1(a). Replace every delay element in the filter architecture by a 1st order all pass filter structure shown in Figure 6.1(b).

Step-3: Obtaining variable frequency responses -

Let f_{c1} be the cutoff frequency of the prototype filter. If f_{c2} is the cutoff frequency of the desired frequency response, the required value of α can be computed using (6.3).

As discussed in Section 6.1.1, variable lowpass, highpass, bandpass and bandstop frequency responses can be obtained using corresponding types of prototype filters.

6.4.1.2 2nd order All Pass Transformation based Variable Digital Filters

Step-1: Prototype filter design -

To meet the desired frequency response specifications, compute the required order of the prototype filter using (3.11) and obtain its coefficients.

Step-2: Implementation of VDF -

Implement the prototype filter coefficients using the hardware implementation architecture shown in Figure 6.2(a). Replace every delay element in the filter architecture by a 2nd order all pass filter structure shown in Figure 6.2(b).

Step-3: Obtaining variable frequency responses -

If f_{c1} is the cutoff frequency of a lowpass prototype filter, f_{cl} and f_{cu} are the lower and upper cutoff frequencies in the desired frequency response, following formulae show the computation of the required values of β_2 and β_3 [67],

$$k_1 = \frac{\cos[(f_{cu} + f_{cl})\pi / 2]}{\cos[(f_{cu} - f_{cl})\pi / 2]} \quad (6.9)$$

$$k_2 = \cot[(f_{cu} - f_{cl})\pi / 2] \times \tan[f_{c1}\pi / 2] \quad (6.10)$$

$$\beta_2 = \frac{2k_1k_2}{k_2 + 1} \quad (6.11)$$

$$\beta_3 = \frac{k_2 - 1}{k_2 + 1}. \quad (6.12)$$

To obtain a lowpass frequency response with cutoff frequency f_c using a 2nd order APT based VDF, compute the required values of β_2 and β_3 by substituting $f_{cl}=0$ and $f_{cu}=f_c$ in (6.9) and (6.10). Similarly, to obtain a highpass frequency response with cutoff frequency f_c , use $f_{cl}=f_c$ and $f_{cu}=1$ to compute the required values of β_2 and β_3 .

6.4.2 Design Example for 1st and 2nd order APT based VDFs

Let the desired passband and stopband peak ripple specifications be 0.05 dB and -45 dB respectively. Three half-band lowpass prototype filters of orders 24, 48 and 96 are designed using (3.11), with transition-band widths 0.2, 0.1 and 0.05 respectively. These prototype filters are subsequently used to design 1st and 2nd order APT based VDFs using the hardware implementation architectures shown in Figures 6.1 and 6.2. VDF-I-A, VDF-I-B and VDF-I-C are the 1st order APT based VDFs designed using prototype filters of orders 24, 48 and 96 respectively. Similarly, VDF-II-A, VDF-II-B and VDF-II-C are the 2nd order APT based VDFs designed using prototype filters of orders 24, 48 and 96 respectively. Figures 6.7(a) and 6.7(b) show few of the different frequency responses that can be obtained using the designed VDFs. Figure 6.7(a) shows few lowpass frequency

Figure 6.7(a). Lowpass frequency responses obtained using VDF-I-B.

Figure 6.7(b). Bandpass frequency responses obtained using VDF-II-C.

responses obtained using VDF-I-B, while Figure 6.7(b) shows different bandpass frequency responses obtained using VDF-II-C.

6.4.3 FPGA Implementation Results for 1st and 2nd order APT based VDFs

In this section, hardware implementation results of the 1st and 2nd order APT based VDF designs described in Section 6.4.2 are presented. The three 1st order APT based

Table 6.2. FPGA implementation results: 1st and 2nd order APT based VDF designs.

	VDF-I-A	VDF-I-B	VDF-I-C	VDF-II-A	VDF-II-B	VDF-II-C
No. of occupied slices	3146	6532	13183	6052	12596	25445
Power (W)	0.254	0.47	0.84	0.384	0.765	1.504
Minimum period (ns)	187.28	409.76	911.61	220.34	527.9	1039.47
Maximum frequency (MHz)	5.34	2.44	1.1	4.54	1.89	0.96

VDFs and three 2nd order APT based VDFs were implemented in Xilinx Virtex-6 (xc6vlx240t) FPGA using the hardware implementation architectures shown in Figures 6.1 and 6.2 respectively. Table 6.2 shows the corresponding implementation results that were obtained after placement and routing in the FPGA. It is observed that as the prototype filter order increases, the maximum operating frequencies of 1st as well as 2nd order APT based VDF designs decrease proportionately. This is due to increase in length of the resultant critical data paths. This dependence of operating frequency of the APT based VDFs on the prototype filter order can be a major bottleneck if high operating frequencies are desired along with stringent performance specifications, which necessitate the use of higher order prototype filters. To eliminate the operating frequency dependency on the prototype filter order, modified hardware implementation architectures for both 1st and 2nd order APT based VDFs are presented. Figures 6.8 and 6.9 show the modified architectures which comprise of additional delay elements when compared with the conventional architectures shown in Figures 6.1 and 6.2. The additional delay elements (registers) enable pipelining of the implementation architectures and ensure constant critical data paths which are independent of the prototype filter order. The three 1st order APT based VDFs and three 2nd order APT based VDFs were re-implemented in the FPGA using the pipelined architectures shown in Figures 6.8 and 6.9. The pipelined VDF implementations of 1st order APT based VDFs are denoted as VDF-I-A-P, VDF-I-B-P and VDF-I-C-P and those of the 2nd order APT based VDFs are denoted as VDF-II-A-P, VDF-II-B-P and VDF-II-C-P.

Figure 6.8(a). 1st order APT based VDF: Pipelined hardware implementation architecture.

Figure 6.8(b). 1st order all pass filter: Pipelined hardware implementation architecture.

Figure 6.9(a). 2nd order APT based VDF: Pipelined hardware implementation architecture.

Figure 6.9(b). 2nd order all pass filter: Pipelined hardware implementation architecture.

Table 6.3. FPGA implementation results: Pipelined 1st and 2nd order APT based VDF designs.

	VDF-I-A-P	VDF-I-B-P	VDF-I-C-P	VDF-II-A-P	VDF-II-B-P	VDF-II-C-P
No. of occupied slices	3254	6873	14023	6193	13019	27000
Power (W)	0.271	0.492	0.949	0.406	0.838	1.739
Minimum period (ns)	9.963	9.995	9.986	10.048	9.993	9.994
Maximum frequency (MHz)	100.37	100.05	100.14	99.52	100.07	100.06

Table 6.3 shows the re-implementation results obtained after placement and routing in the FPGA. For the pipelined hardware implementations of both 1st and 2nd order APT based VDF designs, an identical critical data path comprising of 2 adders and 1 multiplier is observed. It is independent of the different orders of prototype filters used and is fixed for all the implemented designs, which can also be verified from Figures 6.8 and 6.9. Identical critical data path length in the pipelined implementations results in almost equal maximum operating frequencies for all the six APT based VDF designs, as can be noted from Table 6.3. The average maximum frequency achieved for the APT based VDF designs is 100.04 MHz. Thus the maximum operating frequency of 1st as well as 2nd order

APT based VDFs can be independent of the prototype filter order if the pipelined hardware implementation architectures shown in Figures 6.8 and 6.9 are used.

From Tables 6.2 and 6.3, it can be noted that for both 1st and 2nd order APT based VDFs, the pipelined implementations show slightly higher values of number of occupied slices and the power consumed when compared with the non-pipelined implementations. This overhead is due to use of additional logic elements in the pipelined implementation architectures. This is an expected tradeoff between significant advantages for the maximum operating frequencies and marginal increase in number of occupied slices (and thus power) for the different VDF designs. The proposed pipelined hardware implementation architectures lead to a latency of $2N$ clock periods in the implementations of both 1st and 2nd order APT based VDFs. But as the pipelined VDF implementations achieve significantly high operating frequencies, this latency is negligible as it contributes to only a small initial delay increment for the arrival of the first output.

The pipelining presented in this section for conventional 1st and 2nd order APT based VDFs can also be applied to the APT-ICDM-I based VDF proposed in Section 6.2. The following section presents the corresponding results for the APT-ICDM-I based VDF.

6.4.4 FPGA Implementation Results for APT-ICDM-I based VDF

In this section, hardware implementation results for non-pipelined and pipelined versions of the APT-ICDM-I based VDF (proposed in Section 6.2) are presented. Figure 6.10 shows the pipelined hardware implementation architecture for the APT-ICDM-I based VDF, whose non-pipelined hardware implementation architecture was presented in Figure 6.3. The pipelined 1st order all pass filter structure shown in Figure 6.8(b) is used in the pipelined hardware implementation architecture for the APT-ICDM-I based VDF shown in Figure 6.10. Three lowpass prototype filters of orders 24, 48 and 96 are designed using (3.11), with transition-band widths 0.2, 0.1 and 0.05 respectively. These prototype filters are subsequently used to design three APT-ICDM-I based VDFs. Let VDF-A, VDF-B and VDF-C denote the three APT-ICDM-I based VDFs designed using prototype filters of orders 24, 48 and 96 respectively. The three APT-ICDM-I based

VDFs were implemented in Xilinx Virtex-6 (xc6vlx240t) FPGA using the non-pipelined as well as pipelined hardware implementation architectures shown in Figures 6.3 and 6.10 respectively. Table 6.4 shows the corresponding implementation results obtained after placement and routing in the FPGA. Note that in Table 6.4, the pipelined hardware implementations of VDF-A, VDF-B and VDF-C are denoted as VDF-A-P, VDF-B-P and VDF-C-P respectively.

Figure 6.10. APT-ICDM-I based VDF: Pipelined hardware implementation architecture.

Table 6.4. FPGA implementation results: Non-pipelined and pipelined APT-ICDM-I based VDF designs.

	VDF-A	VDF-A-P	VDF-B	VDF-B-P	VDF-C	VDF-C-P
No. of occupied slices	3645	3844	7235	7691	14107	15506
Power (W)	0.286	0.303	0.498	0.548	0.945	1.024
Minimum period (ns)	216.873	9.965	495.008	9.99	1049.554	9.979
Maximum frequency (MHz)	4.611	100.351	2.02	100.10	0.953	100.21

It can be noted from Tables 6.2, 6.3 and 6.4 that the pipelining results obtained for the proposed APT-ICDM-I based VDF are similar to those obtained for the conventional 1st and 2nd order APT based VDFs, and the trends in the different parameters (hardware resource and power consumption, operating frequency) follow similar patterns. An identical critical data path comprising of 2 adders and 1 multiplier is observed for all the pipelined implementations of the APT-ICDM-I based VDFs. It is independent of the different orders of prototype filters used and is fixed for all the implemented designs. The average maximum frequency achieved for the pipelined APT-ICDM-I based VDF designs is 100.22 MHz. It can thus be concluded that the maximum operating frequency of the APT-ICDM-I based VDF can be independent of the prototype filter order if the pipelined hardware implementation architecture shown in Figures 6.10 is used.

6.5 Summary

In this chapter, a VDF design technique based on 1st order APT and the ICDM-I was proposed. In the proposed APT-ICDM-I based VDF, the nature of obtained frequency responses (i.e., lowpass, highpass, bandpass and bandstop) and their cutoff frequencies can be changed on-the-fly without updating the filter coefficients. Design example showed that the APT-ICDM-I based VDF is a low complexity alternative to the other relevant VDF design approaches in literature.

Following the APT-ICDM-I based VDF's idea of combining coefficient decimation and APT, a VDF based spectrum sensing scheme was proposed for SDR based CR receivers. The proposed sensing scheme employs VDFs based on 1st order APT and the MCDM-I to obtain variable frequency bands, followed by the energy detection technique to detect the presence/ absence of radio signals in them. With a design example of an air-ground communication system, it was shown that VDFs employed in the proposed scheme are low complexity alternatives to the other relevant VDFs in literature. The proposed scheme is not limited to spectrum sensing in air-ground communication applications alone; it can be used in CR receivers for other wireless communication scenarios as well.

Generalized design procedures with non-pipelined as well as pipelined hardware implementation architectures were proposed for the conventional 1st and 2nd order APT based VDFs. Analysis of FPGA implementation results showed that proposed pipelined implementation architectures enabled high operating frequencies that were independent of the prototype filter order, for both 1st and 2nd order APT based VDF designs. Identical maximum operating frequencies were obtained for 1st as well as 2nd order APT based VDF designs, with an average value of 100.04 MHz. Similar analysis was also performed for the proposed APT-ICDM-I based VDF and the corresponding implementation results were presented. Using the proposed pipelined implementation architecture, an average maximum operating frequency of 100.22 MHz was obtained for different APT-ICDM-I based VDF designs. The proposed pipelined implementation architectures can be used for realizing APT based VDFs which can be used in high speed spectrum sensing applications.

Through the proposed VDF design and implementation techniques and the corresponding design examples presented in this chapter, it can be concluded that the APT technique which provides unabridged control over cutoff frequencies and the ICDM which enables low complexity and flexible VDF realizations can be effectively combined.

The next chapter contains the conclusions of the work presented in this thesis and some possible future work directions.

Chapter 7

Conclusions and Future Work

7.1 Conclusions

This thesis addressed the challenges involved in the realization of variable digital filters (VDFs) and filter banks for channelization and spectrum sensing in software defined radio (SDR) receivers. Channelization is the most computationally intensive task in the digital front end of SDR receivers. It involves digital filtering to extract desired frequency channels from a wideband input signal. Along with channelization, spectrum sensing is another task performed by SDR based cognitive radios (CRs) wherein VDFs and filter banks can be used to detect presence/ absence of licensed user signals to allow opportunistic access of vacant frequency bands to unlicensed users. VDFs and filter banks which provide high frequency response flexibility in terms of the bandwidths and locations of subbands along with low implementation complexities are desired in SDR receivers. Different VDFs based on programmable as well as fixed-coefficient implementation techniques were reviewed and the major design challenges involved in their realization were discussed in this thesis. Along with VDFs, different filter bank design techniques were also reviewed in this thesis and their advantages and disadvantages were summarized. It was observed that for use in SDR receivers, there is significant scope for improving the different VDFs and filter banks proposed in literature especially with respect to their implementation complexities.

A modified coefficient decimation method (MCDM) which can be used to obtain variable highpass and multi-band frequency responses using the same set of lowpass prototype filter coefficients was proposed. An improved coefficient decimation method (ICDM) was proposed as a combination of the conventional coefficient decimation method (CDM) and the proposed MCDM. It consists of four different coefficient decimation operations namely coefficient decimation method I (CDM-I), coefficient decimation method II (CDM-II), modified coefficient decimation method I (MCDM-I) and modified coefficient decimation method II (MCDM-II). These are categorized into improved coefficient decimation method I, i.e., ICDM-I, which combines CDM-I and MCDM-I to provide variable multi-band frequency responses; and improved coefficient decimation method II, i.e., ICDM-II, which combines CDM-II and MCDM-II to provide variable lowpass and highpass frequency responses. It was shown that the ICDM provides superior frequency response flexibility than the conventional CDM by providing twice the center frequency resolution, which leads to a greater number of distinct frequency responses that can be obtained. Design examples were presented to illustrate the different frequency responses that can be obtained using the coefficient decimation operations in the proposed MCDM and the ICDM.

Based on the ICDM, different VDFs were presented in this thesis along with the corresponding hardware implementation architectures and generalized design procedures. An ICDM-I based VDF was proposed, which can be used in SDR handsets for extraction of individual channels of a single wireless communication standard. It performs ICDM-I (CDM-I and MCDM-I) operations to provide frequency responses having uniform bandwidth subbands with varying locations. Similarly, an ICDM-II based VDF was proposed, which performs ICDM-II (CDM-II and MCDM-II) operations to obtain the desired subbands with varying bandwidths and locations. The ICDM-II based VDF can be used for extracting individual channels of multiple wireless communication standards present in the input signal during different time intervals. A VDF based on the comprehensive ICDM was also proposed. It can perform all the ICDM operations and provides frequency responses with varying subband bandwidths and locations, thus enabling multi-standard channelization. Implementation results for multiple comprehensive ICDM based VDF designs showed that by pipelining the proposed

hardware implementation architecture, high operating frequencies can be obtained that are independent of the prototype filter order. Design examples showed that when compared with VDFs based on the conventional CDM, the proposed VDFs based on the ICDM offer greater frequency response flexibility as well as lower implementation complexity.

Following the VDF design techniques, new filter banks were proposed in this thesis based on the ICDM. An ICDM-I based filter bank was proposed for uniform channelization. It performs ICDM-I operations to provide variable frequency responses from which the desired subbands can be obtained. The ICDM-I based filter bank has lower implementation complexity and twice the flexibility in terms of the possible number and locations of its subbands when compared with the discrete Fourier transform based filter bank and the conventional CDM based filter bank. An ICDM-II based filter bank was proposed for multi-standard channelization. It performs ICDM-II operations to provide variable frequency responses which can be used to obtain the desired uniform as well as non-uniform subbands. The ICDM-II based filter bank shows significant reduction in complexity and superior stopband and transition-band performance when compared with the conventional CDM based progressive decimation filter bank. Another filter bank was proposed based on the comprehensive ICDM. It can perform all the operations of ICDM-I based filter bank and ICDM-II based filter bank, while reducing the impact of design constraints involved in them. The filter bank based on comprehensive ICDM is a low complexity alternative to the other relevant filter banks in literature and can be used for uniform as well as non-uniform channelization.

To realize low complexity VDFs with unabridged control over cutoff frequency, the proposed ICDM was combined with the all pass transformation (APT) technique. A VDF design technique based on the combination of 1st order APT and the ICDM-I was proposed. The proposed APT-ICDM-I based VDF provides variable lowpass, highpass, bandpass and bandstop frequency responses on-the-fly, with unabridged control over the cutoff frequency in the entire Nyquist frequency range. Design example showed that it is a low complexity alternative to the other relevant VDFs in literature. Following the APT-ICDM-I based VDF, a VDF based spectrum sensing scheme was proposed for SDR based CR receivers. The proposed scheme employs VDFs based on the combination of 1st order APT and MCDM-I to obtain the desired frequency bands, followed by the energy

detection technique to detect the presence of radio signals in them. Design example showed that the VDFs employed in the proposed scheme are low complexity alternatives to the other relevant VDFs in literature. Generalized design procedures with non-pipelined as well as pipelined hardware implementation architectures were presented for 1st and 2nd order APT based VDFs. The proposed pipelined implementation architectures enabled high operating frequencies that were independent of the prototype filter order, for both 1st and 2nd order APT based VDFs. Similar pipelining results were also obtained for the proposed APT-ICDM-I based VDF. The proposed pipelined implementation architectures can be used for realizing APT based VDFs which can be used in high speed spectrum sensing applications.

It can be concluded that the significant advantages in complexity and frequency response flexibility that are offered by the VDFs and filter banks proposed in this thesis make them compatible for use in SDR receivers.

Table 7.1 summarizes the VDFs proposed in this thesis along with their advantages and limitations.

Table 7.1. Summary of the VDFs proposed in this thesis.

Proposed VDFs	Key Features, Advantages and Application	Limitation
ICDM-I based VDF	<ul style="list-style-type: none"> Multi-band frequency responses are obtained using CDM-I or MCDM-I operations. Desired frequency band is then obtained by appropriate spectral subtraction and frequency response masking. Provides greater frequency response flexibility, lower complexity and improved stopband attenuation performance when compared with the conventional CDM-I based VDF. Suitable for use in SDR handsets for extraction of a channel of interest of a single wireless communication standard. 	Provides frequency bands with uniform bandwidth only.
ICDM-II based	<ul style="list-style-type: none"> Desired frequency band is obtained by spectral subtraction of resultant frequency responses after 	When a large number of decimation factors

VDF	<p>appropriate CDM-II and MCDM-II operations.</p> <ul style="list-style-type: none"> • Provides frequency bands with variable bandwidths. • Provides lower complexity and improved stopband attenuation performance when compared with the conventional CDM-II based VDF. • Suitable for use in SDR handsets for extracting a channel of interest of different wireless communication standards during different time intervals. 	<p>are involved, the required order of the prototype filter to satisfy the group delay design constraint would also be large. This will lead to high implementation complexity.</p>
Comprehensive ICDM based VDF	<ul style="list-style-type: none"> • Desired frequency band location is obtained using ICDM-I operations while the desired bandwidth is obtained using ICDM-II operations. • Combines the advantages of ICDM-I based VDF and ICDM-II based VDF and reduces the impact of design constraints involved in them. When compared with ICDM-I based VDF, it has greater frequency response flexibility as it provides variable bandwidth frequency bands. When compared with ICDM-II based VDF, it has lower implementation complexity due to requirement of lower order prototype filters. • Suitable for use in SDR handsets for extracting a channel of interest of different wireless communication standards during different time intervals. 	<p>When a large number of distinct wireless communication standards are involved having widely varying possible channel locations, the design effort required will be greater than that required in the ICDM-II based VDF.</p>
APT-ICDM-I based VDF	<ul style="list-style-type: none"> • Based on the combination of 1st order APT and ICDM-I. • APT is used to obtain the desired cutoff frequencies while ICDM-I is used to obtain the desired type of frequency response. • Provides unabridged control over the cutoff frequency in the entire Nyquist frequency range. It is a low complexity alternative to the other relevant VDFs in literature that provide unabridged control over cutoff frequency. • Suitable for use in SDR based CR handsets to perform spectrum sensing for identifying spectrum holes of varying bandwidths and locations in the entire Nyquist frequency range. 	<p>The use of APT technique results in non-linear phase frequency responses at the output. The APT-ICDM-I based VDF is therefore not suitable for use in channelizers wherein linear phase frequency responses are desired.</p>

Table 7.2 summarizes the filter banks proposed in this thesis along with their advantages and limitations.

Table 7.2. Summary of the filter banks proposed in this thesis.

Proposed filter banks	Key Features, Advantages and Application	Limitation
ICDM-I based filter bank	<ul style="list-style-type: none"> Multi-band frequency responses are obtained using CDM-I or MCDM-I operations. Desired subbands are then obtained by appropriate complementary filter operations, spectral subtraction and frequency response masking. Provides lower implementation complexity and twice the flexibility in terms of the possible number and locations of its subbands when compared with the discrete Fourier transform based filter bank and the conventional CDM based filter bank. Suitable for use in SDR base-station receivers for extraction of channels of a single wireless communication standard. 	Provides subbands with uniform bandwidth only.
ICDM-II based filter bank	<ul style="list-style-type: none"> Desired subbands are obtained by spectral subtraction of resultant lowpass and highpass frequency responses after performing CDM-II and MCDM-II operations respectively. Provides subbands with non-uniform bandwidths. Provides significant reduction in complexity and improvement in stopband and transition-band characteristics when compared with the conventional CDM based progressive decimation filter bank. Suitable for multi-standard channelization in SDR base-station receivers. 	It involves a group delay constraint and a transition-band width constraint in its design. When large number of distinct subbands are desired, these constraints will result in requirement of high order prototype filters, thereby increasing the implementation complexity.

Comprehensive ICDM based filter bank	<ul style="list-style-type: none"> Desired subband locations are obtained using ICDM-I operations while the desired bandwidths are obtained using ICDM-II operations. Combines the advantages of ICDM-I based filter bank and ICDM-II based filter bank and reduces the impact of design constraints involved in them. Unlike the ICDM-I based filter bank, it can provide non-uniform bandwidth subbands. When compared with the ICDM-II based filter bank, it requires lower order prototype filter which reduces the implementation complexity. Suitable for low complexity multi-standard channelization in SDR base-station receivers. 	When a large number of uniform as well as non-uniform subbands are to be obtained, the design effort required will be greater than that required in the ICDM-II based filter bank.
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7.2 Future Work

Different VDFs and filter banks which can be used for channelization and spectrum sensing in SDR and CR receivers were proposed in this thesis. For these proposed VDFs and filter banks, following are some possible future work directions –

7.2.1 Improvement of Operating Speed of Proposed Filter Banks based on ICDM

Three different filter banks based on the proposed ICDM were presented in Chapter 5. These filter banks employ the ICDM to obtain variable frequency responses from which the desired subbands can be obtained. As observed for the ICDM-I based filter bank in Section 5.2, the use of ICDM enables low complexity realization of filter banks when compared with the other relevant filter bank design methods in literature. FPGA implementation results for a specific ICDM-I based filter bank design example were presented in Section 5.2.3. It can be noted from the corresponding Table 5.2 that although the ICDM-I based filter bank offered significant advantages over the DFTFB and the

CDFB in terms of hardware area utilization and power consumption, the maximum operating frequency that was achieved was significantly lower than the DFTFB. This can be a limiting factor when filter banks based on the ICDM are to be used for applications wherein high operating frequencies are desired. To address this challenge, pipelined hardware implementation architectures for the different filter banks designed using the ICDM can be developed. The pipelining can be performed similar to the pipelining of hardware implementation architecture proposed for the comprehensive ICDM based VDF, as presented in Section 4.4. Thus, a possible future work direction is to improve the operating speeds of the proposed filter banks based on ICDM. It will be interesting to study the improvement in operating frequency for different ICDM based filter banks designed for various uniform as well as non-uniform channelization design examples.

7.2.2 Development of Spectrum Sensing Schemes using Proposed Variable Digital Filters and Filter Banks

Along with channelization, spectrum sensing is an important application for VDFs and filter banks in SDR based CR receivers. Spectrum sensing can help achieve efficient usage of the scarce electromagnetic spectrum by enabling dynamic spectrum access. VDFs and filter banks can be used to split the wideband frequency range into desired frequency bands and then algorithms such as energy detection, cyclostationary feature detection can be applied to detect whether the bands are vacant or occupied. One such energy detector based spectrum sensing scheme which employs VDFs based on combination of 1st order APT and the MCDM-I proposed in Chapter 3 was presented in Chapter 6 of this thesis. The different VDFs and filter banks proposed in this thesis were mainly studied using design examples for channelization in SDR receivers. It will be interesting to study the proposed VDFs and filter banks by considering their use for spectrum sensing applications in CR receivers. Different detection algorithms can be combined with the proposed VDFs and filter banks to develop new spectrum sensing schemes, which will thus inherit the low implementation complexity and high frequency response flexibility features of the proposed VDFs and filter banks. Efficient techniques such as two-stage spectrum sensing (which involves a coarse detection stage followed by

a fine resolution scan to detect the presence and location of primary user signals) and co-operative sensing (wherein multiple CRs share their spectrum sensing observations to make occupancy decisions with higher accuracy and lesser overall complexity) can be used to test the proposed VDFs and filter banks and check their feasibility for the same. The developed spectrum sensing schemes, after testing and validating using simulations, can then be used to perform over-the-air tests with real world signals.

7.2.3 Optimization based Design of Prototype Filters for Variable Digital Filters and Filter Banks based on the ICDM

The ICDM was proposed in Chapter 3 of this thesis. This technique can be used to obtain variable frequency responses using a single set of prototype filter coefficients by performing operations such as retaining them, discarding them or replacing them by zeros based on an appropriate value of the decimation factor used. Different VDFs and filter banks based on the ICDM and showing high frequency response flexibility and low implementation complexity were proposed in Chapters 5 and 6 of this thesis. One drawback of the ICDM is that stopband attenuation in the resultant frequency responses deteriorates as the value of the decimation factor is increased. To address this problem, overdesign of the prototype filter was proposed along with the corresponding mathematical formulation in Section 3.4.1. But this resulted in the requirement of higher order prototype filters which increased the implementation complexity of the VDFs and filter banks designed using the ICDM.

It can be noted that the conventional CDM and the various CDM based VDFs and filter banks also faces the same stopband attenuation deterioration problem. To address this problem in the CDM, a linear programming based filter design technique was proposed in [103]. This optimization based technique considered the required decimation factors beforehand to generate additional constraints for the prototype filter design, unlike the conventional technique which uses just the basic filter design specifications to obtain the prototype filter coefficients. As a result, the prototype filters obtained using the optimization technique proposed in [103] showed better stopband attenuation

performance after CDM operations when compared with the ones designed using the conventional method and having identical filter orders. It will be interesting to apply such optimization based techniques to design prototype filters in the proposed VDFs and filter banks based on the ICDM. Different optimization techniques can be applied and the corresponding results can be analyzed to identify the potential techniques which can be used for designing prototype filters in the proposed ICDM. Thus, optimization based prototype filter design for VDFs and filter banks based on ICDM forms an exciting future work direction.

7.2.4 Alternative Power Efficient Architectures to Realize the Proposed Variable Digital Filters and Filter Banks based on the ICDM

For the different ICDM based VDFs and filter banks that are proposed in this thesis, hardware implementation architectures have also been provided. These architectures were designed with a goal to achieve low complexity implementations of the different VDFs and filter banks – mainly by minimizing the number of multipliers required, followed by minimizing the use of the other computational/ logic blocks such as adders/ multiplexers etc. It is to be noted that in all these architectures, while performing ICDM operations, even if a particular coefficient is to be replaced by zero or eliminated, the multiplication operation of the input signal and a filter coefficients is performed for all the coefficients. Thus, even if the architectures achieve low complexity implementations, there may be a wastage of power for certain settings when some filter coefficients are not being used. There will be other architectures to implement the ICDM based VDFs and filter banks in hardware wherein the multiplication operations can be avoided if a particular coefficient is not being used, which might provide savings in dynamic power consumption. But those architectures will need more number of computational and/ or logic blocks when compared with the architectures proposed in this thesis. Thus, it will lead to a higher hardware resource consumption and therefore higher static power consumption. The study of finding alternative architectures to implement the proposed ICDM based VDFs and filter banks in hardware in order to investigate the savings in dynamic power consumption versus the increase in the resource consumption forms an interesting future work.

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