

FIR Filter Implementation in Frequency Domain For Wireless Communication System

A thesis submitted in partial fulfillment of the requirements for the award of degree of

**Master of Engineering
in
Electronics & Communication Engineering**

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CERTIFICATE

I hereby certify that the work which is being presented in this thesis entitled, "FIR Filter Implementation in Frequency Domain For Wireless Communication System", in partial fulfillment of the requirements for the award of degree of Master of Engineering in Electronics & Communication Engineering at Thapar University, Patiala, is an authentic record of my own work carried out under the supervision of Mr. Balwant Singh (Sr. Lecturer) and refers other researcher's work which are duly listed in the reference section.


The matter embodied in this thesis has not been submitted for the award of any other degree to any other university.

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ABSTRACT

Third Generation (3G) communication systems use Wideband Code Division Multiple Access (WCDMA) technique to transmit and receive data. The spectrum of a signal of the type used in WCDMA has adjacent channel side lobes. These side lobes introduce interference into users of the adjacent frequencies who could be of the same operator or could be of other operators. That is why, filtering in the transmitter should be used to overcome the effect of the adjacent channel side lobes. Numerous considerations led to the use of a FIR filter in the WCDMA transmitter, together with a matched filter in the receiver in a way to get a Nyquist filter as the combined filter response. The main property of Nyquist filters is that they result in zero Inter Symbol Interference (ISI) at the optimum sampling point for filtered data. This thesis presents a new method of implementing a high performance Finite Impulse Response (FIR) filter in the frequency domain that is suitable for third generation (3G) communication systems. The method discussed can be used for the implementation of low cost modules capable of performing filtering and Fast Fourier Transforms (FFT) for use with the wide variety of the wireless communication standards available nowadays. With the use of this method the new mobile equipments can support multiple standards in order to offer the most complete set of applications to the users like telephony, mobile internet and mobile television. A comparison between a classical time domain FIR implementation, and frequency domain FIR implementation, in terms of computational complexity, inter symbol interference rejection on the basis of average gain in immunity against ISI, Error Vector Magnitude (EVM) and peak distortion is presented. It is found that frequency domain implementation of FIR filter is better than time domain implementation of FIR filter in all respects. This advantage of FIR filter implementation in frequency domain makes it suitable for its use in 3G communication system.

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LIST OF ABBREVIATIONS USED IN THIS THESIS WORK

3G	3rd Generation
3GPP	3rd Generation partnership project
AICH	Acquisition indicator channel
AP-AICH	Access preamble acquisition indicator channel
ACLR	Adjacent channel leakage ratio
ADC	Analog to digital conversion
AFC	Automatic frequency control
AGC	Automatic gain control
AWGN	Additive white Gaussian noise
BB	Baseband
BCCH	Broadcast control channel
BCH	Broadcast channel
BER	Bit error ratio
bps	Bits per second
BPSK	Binary phase shift keying
BRP	Bit rate processing
CD/CA-ICH	Collision detection/channel assignment indicator channel
CSICH	CPCH status indicator channel
CCCH	Common control channel
CCTrCH	Coded composite transport channel
CDMA	Code division multiple access
CPICH	Common pilot channel
CR	Chip rate
CTCH	Common traffic channel
DAC	Digital to analog conversion
DC	Direct current
DCCH	Dedicated control channel
DCH	Dedicated channel
DCR	Direct conversion receiver
DL	Downlink

DPCCH	Dedicated physical control channel
DPCH	Dedicated physical channel
DPDCH	Dedicated physical data channel
DRP	Digital reference point
DSSS	Direct sequence spread spectrum
DTCH	Dedicated traffic channel
EDGE	Enhanced data rate for the GSM evolution
EVM	Error vector magnitude
FACH	Forward access channel
FDD	Frequency division duplexing
FFT	Fast fourier transforms
GPRS	General packet radio system
GSM	Global system for mobile communications
IFFT	Inverse fast fourier transform
IF	Intermediate frequency
Mcps	Mega chips per second
OFDM	Orthogonal frequency division multiplexing
P-CCPCH	Primary common control physical channel
P-CPICH	Primary common pilot channel
PA	Power amplifier
PCPCH	Physical common packet channel
P-CPICH	Page indicator channel
RF	Radio frequency
RACH	Random access channel
SF	Spreading factor
S-CCPCH	Secondary common control physical channel
TDD	Time division duplexing
TFC	Transport format combination
TFCI	Transport format combination indicator
UE	User equipment
UL	Uplink

UMTS	Universal mobile telecommunications system
UTRA	UMTS terrestrial radio access
UTRAN	UMTS terrestrial radio access network
WCDMA	Wideband code division multiple access

CHAPTER-1

INTRODUCTION

1.1 INTRODUCTION

Telecommunication sector is consistently trying to bring the improvement in wireless communication. So that it can enhance the service quality of the wireless communication with high performance and reduced cost. These efforts have led to different multiple access techniques in telecommunication standards like General Packet Radio Switching (GPRS), Universal Mobile Telephone System (UMTS) and Wireless Local Area Network (WLAN). All these standards require different multiple access techniques to access data, and consequently the different hardware. So it is a must, to extract similarities between multiple access techniques in order for a system to take advantage of the multiplicity of radio technologies .By extracting the similarities between different standards there would be no need to implement different hardware. So a system may take advantage of multiplicity of radio technologies. This is done by implementing the FIR filter in the frequency domain [1]. A number of scientists have discussed the implementation of FIR filters in the frequency domain since last two decades [2, 3, and 4]. The minimum requirements of communication systems based on WCDMA can be achieved by using a 64-tap filter (or sometimes less) [5]. Finite Impulse Response (FIR) Filters and Fast Fourier Transforms (FFT) have essential role in wireless communication systems. The researchers in [6] represented the importance of FIR filters in the WCDMA technique on which the UMTS standard is based. The use of FIR filter in the WCDMA transmitter is discussed in [7]. Orthogonal Frequency Division Multiplexing (OFDM) technique has effective role of FFT and IFFT [8]. WCDMA is based on the spread spectrum technique which has two common methods of spreading DS-Direct Sequence(used in WCDMA) and FH-Frequency Hopping[9]. Thus, present work discusses a new method of implementing the FIR filter in the frequency domain. This method enables the realization of a minimal size module capable of executing both the FIR and the FFT functions and increases the similarities between WCDMA and OFDM based communication systems. Then it is shown that by choosing frequency domain filter coefficients as the samples of the ideal frequency transfer function, it can boost the performance of such a filter mainly in terms of immunity against ISI.

1.2 NEED FOR FIR FILTERING IN FREQUENCY DOMAIN

The third Generation (3G) communication systems use WCDMA technique to transmit and receive data. The spectrum of a signal of the type used in WCDMA has adjacent channel side lobes. These side lobes introduce interference into users of the adjacent frequencies who could be of the same operator or could be of other operator's. That is why, filtering in the transmitter should be used to overcome the effect of the adjacent channel side lobes. Numerous considerations discussed in [7] led to the use of a FIR filter in the WCDMA transmitter, together with a matched filter in the receiver in a way to get a Nyquist filter as the combined filter response. Implementing the FIR filter in the frequency domain for the UMTS standard is actually feasible and even requires less computational complexity than a time domain realization. It also exposes a new and effective method to choose the best filter coefficients for the UMTS standard in order to increase performance of FIR filtering, namely its immunity against inter symbol interference.

1.3 OBJECTIVES OF THE THESIS

The main objective of the thesis is to implement FIR filtering in frequency domain for the UMTS standard and to obtain its advantages over time domain implementation of FIR filter. Third Generation 3G communication systems use WCDMA technique to transmit and receive data. The spectrum of a signal of the type used in WCDMA has adjacent channel side lobes. So in order to overcome the effect of the adjacent channel side lobes, numerous considerations led to the use of a FIR filter in the WCDMA transmitter. So the objectives of the thesis are:

- To implement two FIR filters one in time domain and other in frequency domain.
- To abstract the filter coefficients for time domain implementation of FIR filter by sampling the time domain representation of Root Raised Cosine (RRC) filter.
- To abstract the filter coefficients for frequency domain implementation of FIR filter by sampling the frequency domain representation of RRC filter.
- To compare the performances of the two filter implementation on the basis of :
 1. Computational complexity.
 2. Inter Symbol Interference (ISI) rejection based on:

- (a) Average gain in immunity against ISI.
- (b) Error Vector Magnitude (EVM).
- (c) Peak distortion.
- (d) Additive White Gaussian Noise (AWGN) rejection.

1.4 ORGANISATION OF THE THESIS

- **CHAPTER ONE:** This chapter describes the limitation of WCDMA spectrum, need for FIR filtering in frequency domain & objectives of the thesis.
- **CHAPTER TWO:** This chapter explores the basics of WCDMA, its principle of operation and the air interface for uplink and downlink.
- **CHAPTER THREE:** This chapter gives the practical design requirements for WCDMA transmitter and receiver.
- **CHAPTER FOUR:** This chapter defines the various types of filters and their design methodology.
- **CHAPTER FIVE:** This chapter gives design methodology to obtain the objective, simulation results and their performance evaluation by making the comparisons.
- **CHAPTER SIX:** In this chapter it is concluded that FIR filter implementation in frequency domain is better than time domain implementation of FIR filter after the performance evaluation.

CHAPTER-2

WIDEBAND CODE DIVISION MULTIPLE ACCESS

2.1 INTRODUCTION

Code Division Multiple Access (CDMA) is the technology chosen for third generation mobile communication standard (3G). CDMA is the wireless communication technology based on the principle of spread spectrum technique. All users in CDMA systems use the same carrier frequency and transmit simultaneously. To be able to distinguish between different users, CDMA uses code sequence which is unique for every user.

2.2 SPREAD SPECTRUM

In normal wireless data transmission system, the used bandwidth is more or less equal to user data rate. In wireless data transmission system based on spread spectrum, the used bandwidth is more than the user data rate. The result of this more used bandwidth is that the transmitted power is spread over a wide frequency band [9].

Common methods to achieve spread spectrum:

- DS-Direct Sequence(used in WCDMA)
- FH-Frequency Hopping

2.3 WCDMA

Wideband CDMA is an air interface technology based on artificially increasing the bandwidth of a signal by modulating each baseband symbol with a binary or quaternary signature of much higher rate than that of the original data symbol. Each sample of the signature is called a chip, and the number of chips that are modulated by a single data symbol is referred to as the Spreading Factor (SF) of the system [10]. The process of modulating a high rate signature with lower rate data symbols is called 'spreading', as the resulting bandwidth is spread by a factor of SF relative to that of the original signal. By raising the bandwidth of the signal, the potential diversity gain in dispersive propagation channels is increased. As the signal energy is spread over a larger bandwidth, the power spectral density of the transmitted signal is lower by a factor

of SF than that of the original data signal. At the receiver, the signal is sampled at the appropriate rate, and the received samples are correlated with the known signature code at the receiver. This correlation over SF chips results in significant in band noise and interference rejection, commonly labeled the 'processing gain' or 'spreading gain'. For a single path propagation channel, the processing gain is equal to the SF. This means that in the first approximation the overall performance of a CDMA system in Gaussian noise in a Line Of Sight (LOS) channel is equal to that of a non CDMA system employing direct modulation of a carrier. The benefits of Wideband CDMA (WCDMA) systems lie in the diversity gains that can be achieved over multipath propagation channels and the significant improvement of intercellular handovers. Also, WCDMA has resistance to narrowband jamming; the wideband signals can be hidden at very low power spectral density levels, which was of importance in the early military roots of CDMA technology. Using appropriate signature codes, multiple information streams can share the same signal bandwidth by using different orthogonal or near orthogonal codes to spread the different data symbols. In the downlink of a cellular CDMA system, different scrambling codes are used to separate the different nodes because they use the same frequency band, and different spreading codes are employed to discriminate the different users. In the uplink, the User Equipments (UE) use different scrambling codes and may also use different spreading codes. Reserved signatures are also used to transmit synchronization, signalling and broadcast channels alongside the user data.

- WCDMA is a wideband Direct-Sequence Code Division Multiple Access (DS-SS) system, i.e. user information bits are spread over a wide bandwidth by multiplying the user data with quasi-random bits (called chips) derived from CDMA spreading codes. In order to support very high bit rates (up to 2 Mbps), the use of a variable spreading factor and multimode connections is supported.
- The chip rate of 3.84 Mcps leads to a carrier bandwidth of approximately 5 MHz.
- WCDMA supports highly variable user data rates, in other words the concept of obtaining Bandwidth on Demand (BoD) is well supported. The user data rate is kept constant during each 10 ms frame. However, the data capacity among the users can change from frame to frame.
- WCDMA supports two basic modes of operation: Frequency Division Duplex (FDD) and Time Division Duplex (TDD). In the FDD mode, separate 5 MHz carrier frequencies are

used for the uplink and downlink respectively, whereas in TDD only one 5 MHz is time shared between the uplink and downlink.

2.4 SPREADING AND DISPREADING

The basic operations of spreading and despreading for a DS-SS are shown in the Figure 2.1 & Figure 2.2. User data is here assumed to be a BPSK-modulated bit sequence of rate R , the user data bits assuming the values of ± 1 . The spreading operation, shown in Figure 2.1, is the multiplication of each user data bit with a sequence of 8 code bits, called chips. It is assumed also for the BPSK spreading modulation. It is seen that the resulting spread data is at a rate of $8 \times R$ and has the same random (pseudo-noise-like) appearance as the spreading code. In this case it would say that it used a spreading factor of 8. This wideband signal would then be transmitted across a wireless channel to the receiving end. During despreading the spread user data/chip sequence is multiplied, bit duration by bit duration, with the very same 8 code chips as used during the spreading of these bits. As shown in Figure 2.2, the original user bit sequence has been recovered perfectly, provided there is also perfect synchronization between the spread user signal and the despreading code [11][12][13][14]. The increase of the signaling rate by a factor of 8 corresponds to a widening (by a factor of 8) of the occupied spectrum of the spread user data signal. Due to this virtue, CDMA systems are more generally called spread spectrum systems. Despreading restores a bandwidth proportional to bit sequence rate R for the signal. The basic operation of the correlation receiver for CDMA is shown in Figure 2.2. The upper half of the figure shows the reception of the desired signal. Figure 2.2 shows the despreading operation with a perfectly synchronized code. Then, the correlation receiver integrates (i.e. sums) the resulting products (data *code) for each user bit. The lower half of Figure 2.2 shows the effect of the despreading operation when applied to the CDMA signal of another user whose signal is assumed to have been spread with different spreading code. The result of multiplying the interfering signal with the own code and integrating the resulting products leads to interfering signal values lingering around 0. As it can be seen, the amplitude of the own signal increases on average by a factor of 8 relative to that of the user of the other interfering system, i.e. the

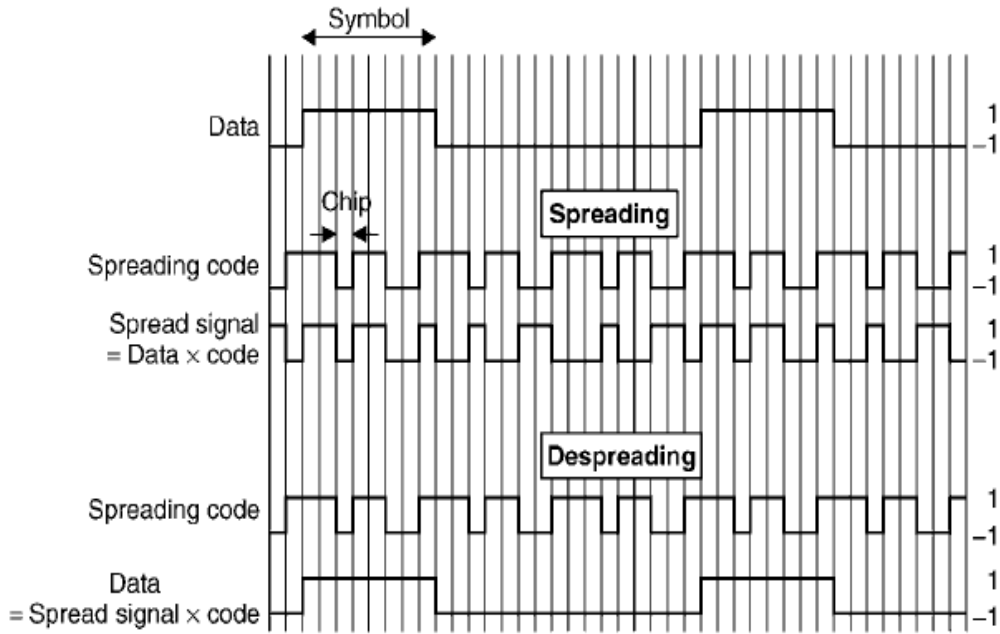


Figure 2.1 Spreading and despreading in DS-CDMA [12].

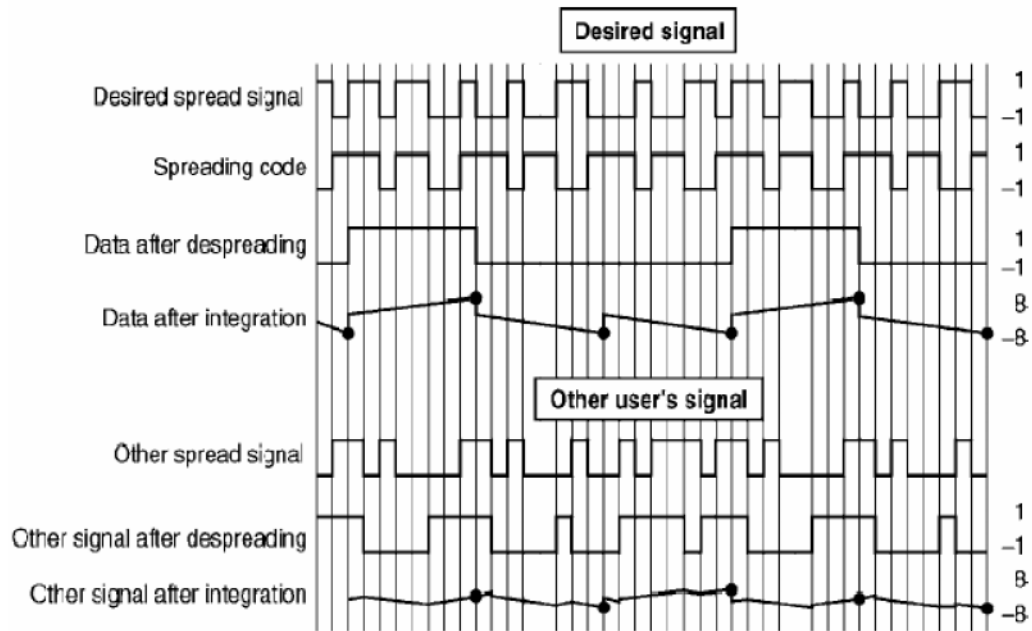


Figure 2.2 Principle of the CDMA correlation receiver [12].

correlation detection has raised the desired user signal by the spreading factor, here 8, from the interference present in the CDMA system. This effect is termed ‘processing gain’ and is a fundamental aspect of all CDMA systems, and in general of all spread spectrum systems. Processing gain is what gives CDMA systems the robustness against self-interference that is

necessary in order to reuse the available 5 MHz carrier frequencies over geographically close distances.

Table 2.1 Main WCDMA parameters

Multiple access method	DS-CDMA
Duplexing method	Frequency division duplex/time division duplex
Base station synchronization	Asynchronous operation
Chip rate	3.84 Mcps
Frame length	10 ms
Data rate	2 Mbps
Uplink frequency	1920–1980 MHz
Downlink frequency	2110–2170 MHz

2.5 WCDMA AIR INTERFACE

The WCDMA air interface employs Frequency Division Duplexing (FDD), which means that the downlink employs a different carrier frequency than the uplink. The frequency range allocation is 1920–1980 MHz for the uplink and 2110–2170 MHz for the downlink. The uplink and downlink frequency bands are split into channels of 5 MHz width, and the frequency spacing between the uplink and the downlink channel is 190 MHz. Because the uplink and the downlink transmissions employ different frequency bands, both the uplink and downlink transmissions are organized in the time domain in frames, which have a duration of 10 ms. Given the chip rate of 3.84 Mchips per second (Mcps), a frame contains 38 400 chips. The frame is the fundamental time unit for the processing of channels within the WCDMA physical layer: channels are started and stopped at frame boundaries. The scrambling codes employed in both uplink and downlink repeat on a frame by frame basis.

2.6 ROLE OF CHIP RATE PROCESSING

Within the physical layer, the role of Chip Rate Processing (CRP) is to receive the baseband signal from the output of the (Radio Frequency, Intermediate Frequency) RF/IF subsystem and to deliver independent received data streams (Physical Channels, PhCHs) to Coded Composite Transport Channels (CCTrCH) processing. The CRP subsystem is also responsible for generating uplink chips to transmit given streams of encoded PhCH bits from Bit Rate Processing (BRP), and for managing the physical layer closed loops (power control, Site Selection Diversity Transmit (SSDT) and transmit diversity).

2.7 SPREADING AND SCRAMBLING

WCDMA uses a two level code system: orthogonal spreading codes and pseudo random scrambling codes. In order to support variable data rates, the WCDMA air interface allows for per-channel selectable spreading factors, and this family of spreading codes is called Orthogonal Variable Spreading Factor (OVSF) codes. The use of OVSF and scrambling codes is different in the downlink and uplink. In the downlink, the OVSF codes, also called channelization codes, are used to multiplex different channels transmitted in the same cell. In the uplink, the OVSF codes are used to separate data and control channels from a specific user. Scrambling, using pseudorandom sequences, is used in addition to spreading. In the downlink, different scrambling codes separate different cells, and in the uplink they separate different users. Figure 2.4 shows the process of spreading and combining in the User Equipment (UE), where $C_{d,1}$ denotes the OVSF code for the first data channel, α_d is a scaling factor, and $S_{dpch,n}$ denotes the scrambling code. Spreading and scrambling are applied differently in the uplink and downlink as illustrated in Figure 2.3 and Figure 2.4 [15].

2.7.1 SPREADING

2.7.1.1 OVSF CODES

The ratio of the data symbol rate to the chip rate is the spreading factor SF. By keeping the chip rate constant, at 3.84 MHz for WCDMA, and by varying the spreading factor for a specific code in the range from 4 to 512, a wide range of data rates can be supported. The OVSF codes used for spreading are named $C_{sp,i,j}$ where i is the spreading factor, and j is the

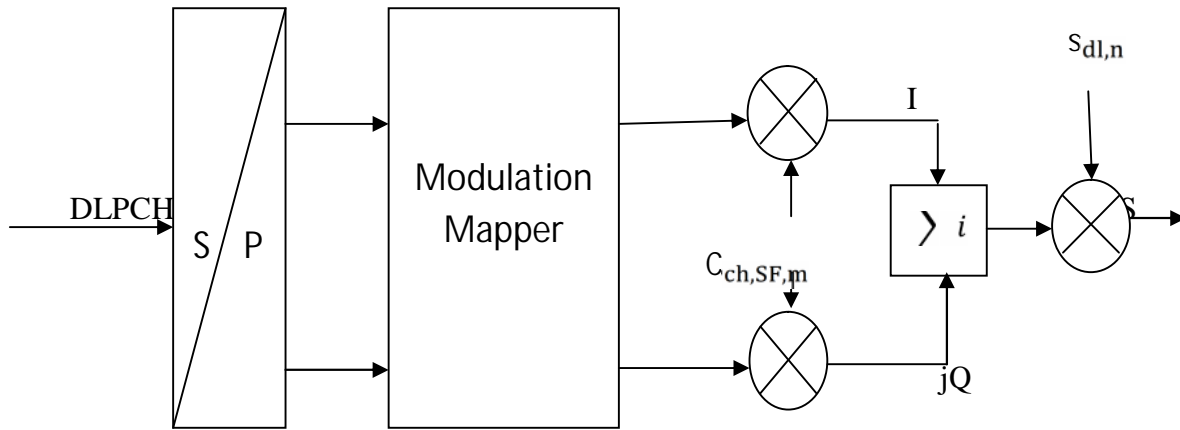


Figure 2.3. Spreading for all downlink physical channels [15].

of the spreading factor $j \in \{0,1,\dots,(i-1)\}$. The $C_{sp,i,j}$ are perfect orthogonal sequences such that:

$$\sum_{n=0}^{i-1} C_{sp,i,j}(n) C_{sp,i,j}(n) = 0 \quad 2.1$$

for $u,v \in \{0,1,\dots,(i-1)\}$ and $u \neq v$

$$C_{ch,1,0} = 1 \quad 2.2$$

$$\begin{bmatrix} C_{ch,2,0} \\ C_{ch,2,1} \end{bmatrix} = \begin{bmatrix} C_{ch,1,0} & C_{ch,1,0} \\ C_{ch,1,0} & -C_{ch,1,0} \end{bmatrix} \quad 2.3$$

$$\begin{bmatrix} C_{ch,2^{n+1},0} \\ C_{ch,2^{n+1},1} \\ C_{ch,2^{n+1},2} \\ C_{ch,2^{n+1},3} \\ \vdots \\ C_{ch,2^{n+1},2^{n+1}-2} \\ C_{ch,2^{n+1},2^{n+1}-1} \end{bmatrix} = \begin{bmatrix} C_{ch,2^n,0} & C_{ch,2^n,0} \\ C_{ch,2^n,0} & -C_{ch,2^n,0} \\ C_{ch,2^n,1} & C_{ch,2^n,1} \\ C_{ch,2^n,1} & -C_{ch,2^n,1} \\ \vdots & \vdots \\ C_{ch,2^n,2^{n-1}} & C_{ch,2^n,2^{n-1}} \\ C_{ch,2^n,2^{n-1}} & -C_{ch,2^n,2^{n-1}} \end{bmatrix} \quad 2.4$$

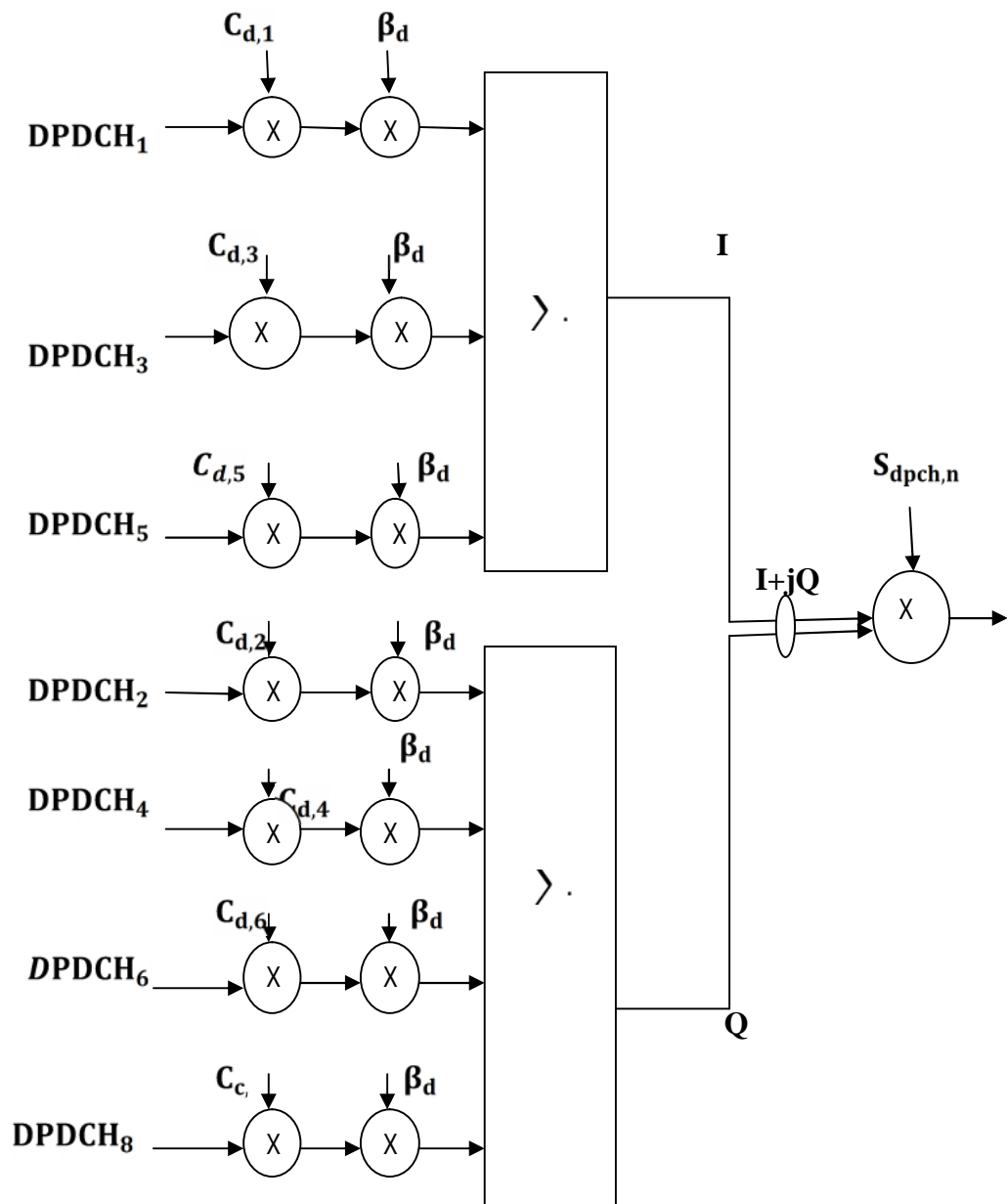


Figure 2.4 Spreading for uplink dedicated channels [15].

This results in the code tree shown in Figure 2.5. This code tree schematically depicts the relationship between the different OVSF codes that can be generated using the rules above; each node corresponds to an OVSF code $C_{ch,i,j}$ and the edges show the relationship between the codes. Two codes that lie on different sub branches on the code tree are

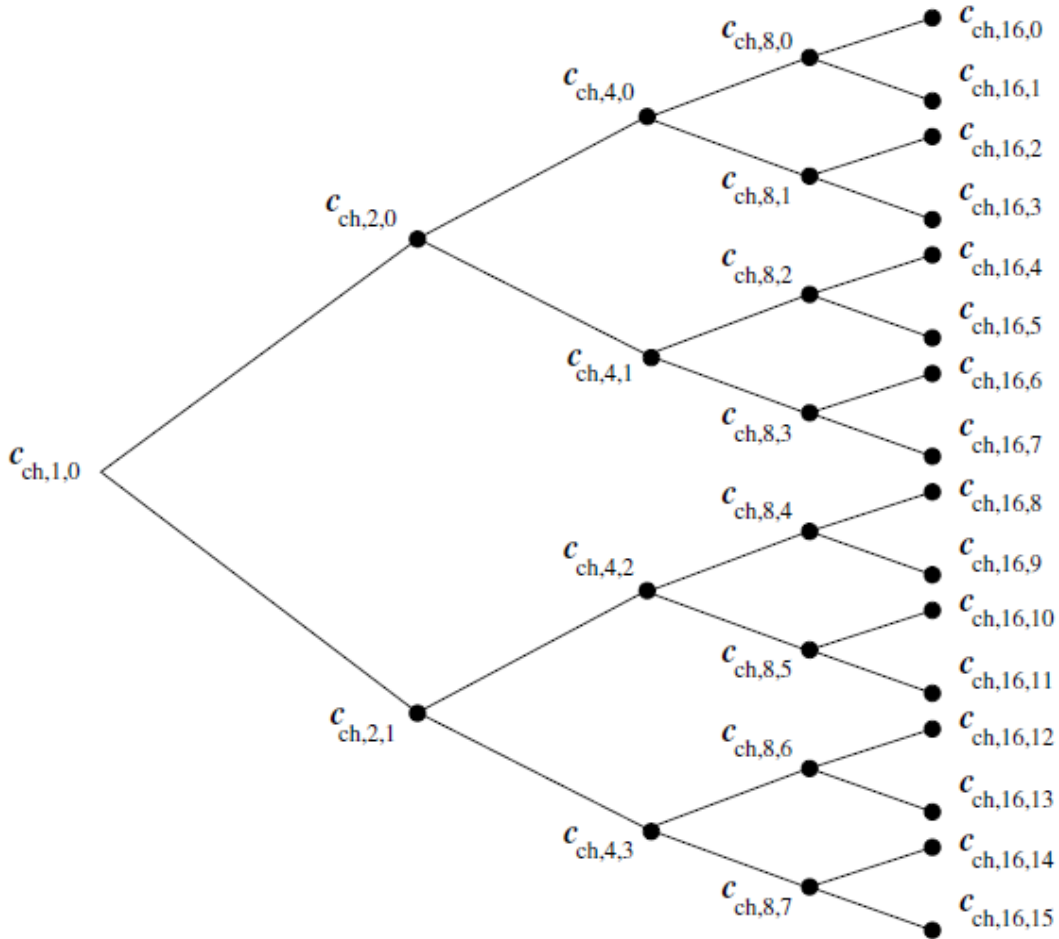


Figure 2.5 The OVSF (Walsh) code tree [15].

orthogonal, and an OVSF code with a spreading factor of less than 256 interferes with all codes in the sub tree to the right of itself. The WCDMA air interface's OVSF codes allow implementation of flexible data rates, instead of receiving two codes of spreading factor 256 (multi code), a user is allocated a single code of factor 128. This single code covers the same space in the code tree that two neighboring codes of length 256 would, and offers the same data rate as the two longer codes in parallel, but only a single code needs to be decoded at the receiver. For higher data rates, bundling multiple neighboring codes into a single higher rate

code gives an even more significant receiver simplification.

2.7.1.2 AUTO- AND CROSS CORRELATION PROPERTIES OF OVFSF CODES

The OVFSF codes are orthogonal Walsh codes. This class of code loses orthogonality if they are not time aligned. This poor cross correlation causes problems in multipath propagation channel conditions, where the echoes of the downlink signal would interfere significantly with each other. The necessary improvement in the signal's auto and cross correlation properties is achieved by the scrambling code.

2.7.2 SCRAMBLING

Scrambling is applied on top of spreading and does not change the chip rate of the spread signal. In the uplink, scrambling is used for the separation of users and in the downlink different scrambling is used for the separation of cells. Scrambling codes are 10 ms long codes formed using shift registers such as that shown in Figure 2.6 for generation of downlink scrambling codes. Unlike spreading codes, scrambling codes are not orthogonal sequences, but they have good auto- and cross correlation properties.

2.7.2.1 SCRAMBLING IN THE DOWNLINK

In the downlink, the shift register arrangement allows $2^{18} - 1$ different scrambling codes to be generated. However, for ease of code acquisition and cell search, not all of these are used. The scrambling codes are arranged into 512 sets of primary scrambling codes, each with an associated 15 secondary codes. The 512 primary codes are further divided into 64 code groups of 16 codes. Each code has associated with it an alternative left and right associated scrambling code that can be used in compressed mode. Each cell is allocated exactly one primary scrambling code; the code number of this primary scrambling code is the cell's cell ID, and the primary scrambling code is used to transmit the, Primary Common Control Physical Channel P-CCPCH, Secondary Common Control Physical Channel S-CCPCH, Primary Common Pilot Channel P-CPICH, Page Indicator Channel PICH, Acquisition Indicator Channel AICH, Access Preamble Acquisition Indicator Channel AP-AICH, Collision Detection/Channel Assignment Indicator Channel CD/CA-ICH and the CPCH Status Indicator Channel CSICH. The other physical channels can either be transmitted on the primary scrambling code or on one of the 15 secondary

codes from the set associated with the primary code. The primary scrambling code is searched for by the mobile terminal

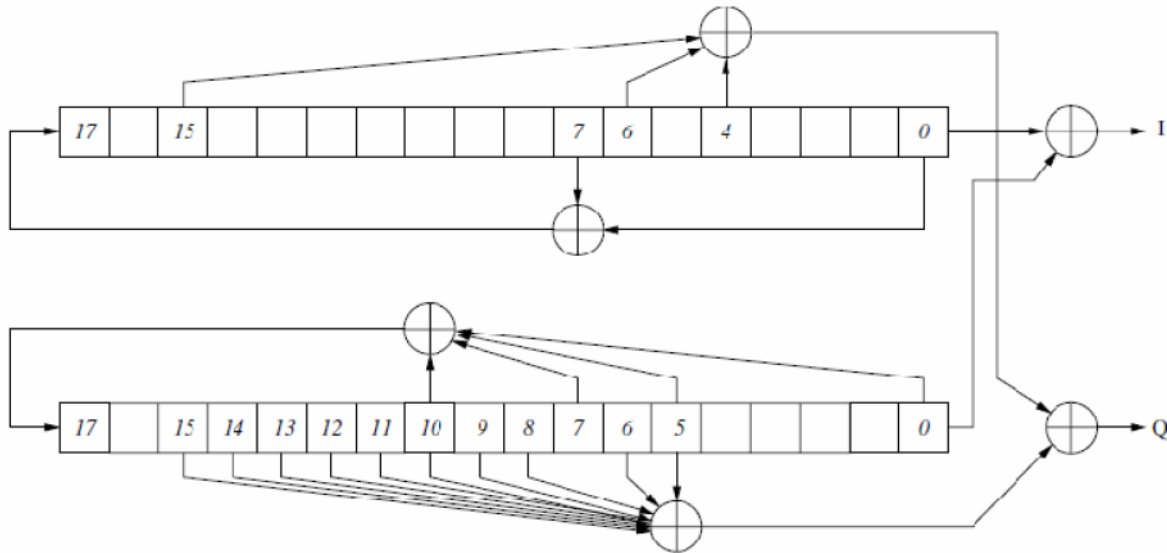


Figure 2.6 Downlink scrambling code generator[15].

during cell search. The mobile terminal finds the primary scrambling code during the cell search procedure. Detection of the S-SCH gives the scrambling code group to which the primary code belongs and correlation of the CPICH with all of the codes in that group identifies the code.

2.7.2.2 SCRAMBLING IN THE UPLINK

In the uplink, each mobile terminal is allocated its own scrambling code by the network; this code must be unique in the immediate network neighborhood as it is received by all the cells with which the mobile terminal is in soft handover.

2.8 TRANSPORT CHANNELS AND THEIR MAPPING TO THE PHYSICAL CHANNELS

The data generated at higher layers is carried over the air with transport channels, which are mapped in the physical layer to different physical channels. The physical layer is required to support variable bit rate transport channels to offer Bandwidth-On-Demand (BoD) services, and to be able to multiplex several services to one connection. Each transport channel is accompanied by the Transport Format Indicator (TFI) at each time event at which data is expected to arrive for the specific transport channel from the higher layers. The physical layer combines the TFI

information from different transport channels to the Transport Format Combination Indicator (TFCI). The TFCI is transmitted in the physical control channel to inform the receiver which transport channels are active for the current frame. The TFCI is decoded appropriately in the receiver and the resulting TFI is given to higher layers for each of the transport channels that can be active for the connection. In Figure 2.7, two transport channels are mapped to a single physical channel, and also error indication is provided for each transport block. The transport channels may have a different number of blocks and at any moment not all the transport channels are necessarily active.

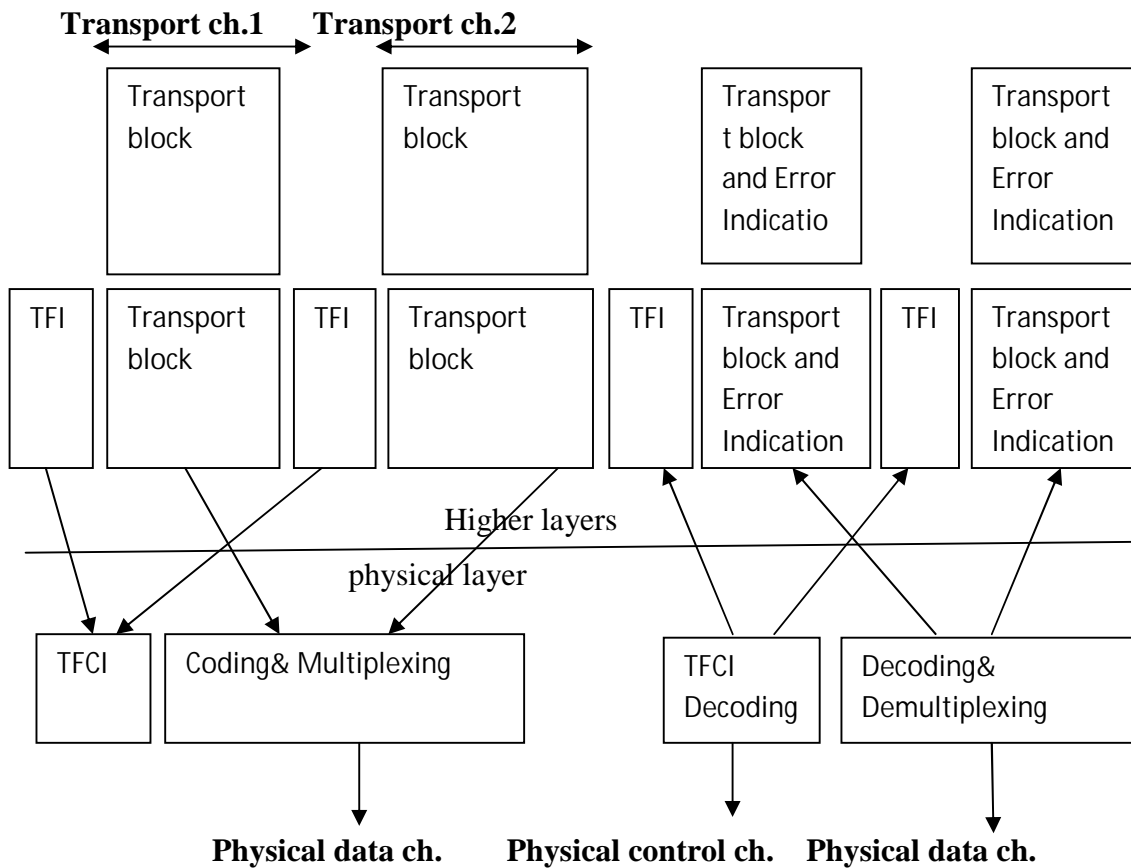


Figure 2.7 The interface between higher layers and the physical layer[12].

2.9 TYPES OF CHANNELS

Two types of transport channel exist: dedicated channels and common channels. The main difference between them is that a common channel is a resource divided between all or a group

of users in a cell, whereas a dedicated channel resource, identified by a certain code on a certain frequency, is reserved for a single user only.

2.9.1 DEDICATED TRANSPORT CHANNEL

The only dedicated transport channel is the dedicated channel, for which the term DCH is used in the 25-series of the UMTS Terrestrial Radio Access (UTRA) specification. The dedicated transport channel carries all the information intended for the given user coming from layers above the physical layer, including data for the actual service as well as higher layer control information. The content of the information carried on the DCH is not visible to the physical layer, thus higher layer control information and user data are treated in the same way. Naturally the physical layer parameters set by UMTS Terrestrial Radio Access Network (UTRAN) may vary between control and data. The familiar GSM channels, the Traffic Channel (TRCH) or Associated Control Channel (ACCH), do not exist in the UTRA physical layer. The dedicated transport channel carries the service data, such as speech Frames, and higher layer control information, such as handover commands or measurement reports from the terminal. In WCDMA a separate transport channel is not needed because of the support of variable bit rate and service multiplexing. The dedicated transport channel is characterized by features such as fast power control, fast data rate change on a frame-by-frame basis, and the possibility of transmission to a certain part of the cell or sector with varying antenna weights with adaptive antenna systems. The (DCH) supports soft handover.

2.9.2 COMMON TRANSPORT CHANNELS

There are six different common transport channel types defined for UTRA in Release '99, which are introduced in the following sections. There are a few differences from second generation systems, for example transmission of packet data on the common channels, and a downlink shared channel for transmitting packet data. Common channels do not have soft handover but some of them can have fast power control.

2.9.2.1 BROADCAST CHANNEL

The Broadcast Channel (BCH) is a transport channel that is used to transmit information specific to the UTRA network or for a given cell. The most typical data needed in every network is the

available random access codes and access slots in the cell or the types of transmit diversity method used with other channels for that cell. As the terminal cannot register to the cell without the possibility of decoding the broadcast channel, this channel is needed for transmission with relatively high power in order to reach all the users within the

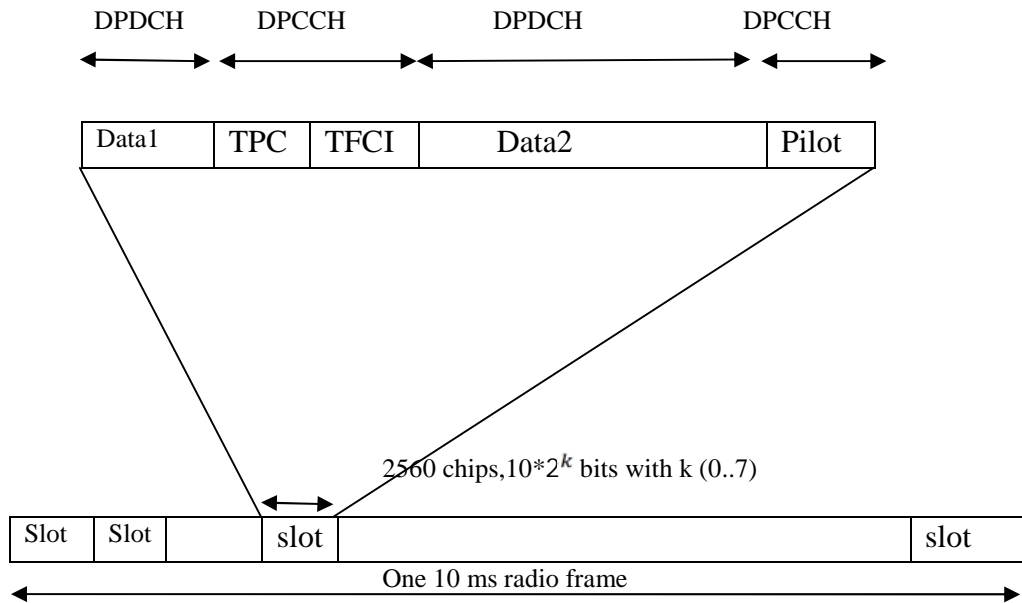


Figure 2.8 Frame structure for the downlink DPCH [15].

intended coverage area. From a practical viewpoint, the information rate on the broadcast channel is limited by the ability of low-end terminals to decode the data rate of the broadcast channel, resulting in a low and fixed data rate for the UTRA broadcast channel.

2.9.2.2 FORWARD ACCESS CHANNEL

The Forward Access Channel (FACH) is a downlink transport channel that carries control information to terminals known to be located in the given cell. This is used, for example, after a random access message has been received by the base station. It is also possible to transmit packet data on the FACH. There can be more than one FACH in a cell. One of the forward access channels must have such a low bit rate that it can be received by all the terminals in the cell area. With more than one FACH, the additional channels can have a higher data rate. The

FACH does not use fast power control, and the messages transmitted need to include in band identification information to ensure their correct receipt.

2.9.2.3 PAGING CHANNEL

The Paging Channel (PCH) is a downlink transport channel that carries data relevant to the paging procedure, that is, when the network wants to initiate communication with the terminal. The simplest example is a speech call to the terminal: the network transmits the paging message to the terminal on the paging channel of those cells belonging to the location area that the terminal is expected to be in. The identical paging message can be transmitted in a single cell or in up to a few hundred cells, depending on the system configuration. The terminals must be able to receive the paging information in the whole cell area. The design of the paging channel also affects the terminal's power consumption in the standby mode. The less often the terminal has to tune the receiver in to listen for a possible paging message, the longer the terminal's battery will last in standby mode.

2.9.2.4 RANDOM ACCESS CHANNEL

The Random Access Channel (RACH) is an uplink transport channel intended to be used to carry control information from the terminal, such as requests to set up a connection. It can also be used to send small amounts of packet data from the terminal to the network. For proper system operation the random access channel must be heard from the whole desired cell coverage area, which also means that practical data rates have to be rather low, at least for the initial system access and other control procedures.

2.9.2.5 UPLINK COMMON PACKET CHANNEL

The Uplink Common Packet Channel (CPCH) is an extension to the RACH channel that is intended to carry packet-based user data in the uplink direction. The reciprocal channel providing the data in the downlink direction is the FACH. In the physical layer, the main differences to the RACH are the use of fast power control, a physical layer-based collision detection mechanism and a CPCH status monitoring procedure. The uplink CPCH transmission may last several frames in contrast with one or two frames for the RACH message.

2.9.2.6 DOWNLINK SHARED CHANNEL

The Downlink Shared Channel (DSCH) is a transport channel intended to carry dedicated user data and/or control information; it can be shared by several users. In many respects it is similar to the forward access channel, although the shared channel supports the use of fast power control as well as variable bit rate on a frame-by-frame basis. The DSCH does not need to be heard in the whole cell area and can employ the different modes of transmit antenna diversity methods that are used with the associated downlink DCH. The downlink shared channel is always associated with a downlink DCH.

2.10 CONCLUSION

This chapter explores the basics of Wideband Code Division Multiple Access (WCDMA) which is used as a multiple access technique for 3G communication system. This is based on the principle of spread spectrum technique which has two common methods DS-Direct Sequence (used in WCDMA) and FH-Frequency Hopping. It has carrier bandwidth of 5 MHz so several advantages come through the wide bandwidth of WCDMA. It supports two basic modes of operation: Frequency Division Duplex (FDD) and Time Division Duplex (TDD). The frequency range allocation is 1920–1980 MHz for the uplink and 2110–2170 MHz for the downlink. WCDMA uses a two level code system: orthogonal spreading codes in order to support variable data rates and pseudo random scrambling codes to achieve improvement in the signal's auto and cross correlation properties. Finally it is observed that WCDMA has some transport channels through which data generated at higher layers is carried over the air which are mapped in the physical layer to different physical channels.

CHAPTER-3

PRACTICAL DESIGN REQUIREMENT FOR WCDMA

3.1 INTRODUCTION

This chapter describes the radio aspects of 3GPP's Frequency Division Duplex (FDD) version in the European frequency bands. The chapter examines the requirements imposed by the 3GPP specifications, describes a typical UMTS transceiver and looks at key performance requirements. The Radio Frequency (RF) section of a handset is crucial to the User Equipment (UE) performance and hence to the success of a product. Poor quality reception can encourage customers to change their handset for a better one. Network operators, on the other hand, have an interest in maximizing cell radius in order to reduce the number of base stations to provide coverage. Both sides, customers and operators, also want attractive UE costs, while battery lifetime is generally of interest to the users. However, both objectives are difficult to achieve if additional signal processing or circuitry is required to recover the losses incurred by the RF front end. The RF front end comprises, amongst other components, amplifiers, filters, oscillators and mixers (modulators) for both transmitter and receiver. The term Baseband (BB), refers to the (digital) signal processing blocks which are closely linked to the RF transceiver circuits. These include the Analog to Digital Conversion (ADC) and Digital to Analog Conversion (DAC) stages, the pulse shape filtering stage and the Automatic Gain Control (AGC) loop.

3.2 RADIO ARCHITECTURE OVERVIEW

The UE's radio access technology, which processes the signal from and to the air interface, is broken down into three chief components:

- **Radio transceiver** comprises the RF section and the associated baseband signal processing stages, responsible for demodulation and modulation.
- **Chip rate processing entity** comprises functions which are relevant to the proper functioning of the UE and the preprocessing of received data and processing of data suitable for modulation. A number of processes work without being controlled or managed by a layer above the physical layer, e.g. synchronization, while other operations require information and control from layers above, e.g. executing a

compressed mode pattern. Its primary task is synchronization, spreading and dispreading the signal.

- **Bit rate processing entity** comprises the functionality which processes the actual information from and to higher layers, suitable for transmission over the air interface via the chip rate processing block. Its primary task is one of signal protection and correction of errors. The radio transceiver hardware provides an interface between the antenna and the follow on digital signal processing entity. The block diagram of a typical WCDMA FDD (only) transceiver is shown in Figure 3.1 where the radio hardware has been split into four areas:
 - **Receive RF** filters, amplifies and down converts the RF signal received at the antenna.
 - **Receive baseband** filters and converts the analog signal into the digital signal which can be processed by the chip rate processing entity.
 - **Transmit baseband** consists of blocks which process and convert the digitized signal to an Analog signal suitable for modulation.
 - **Transmit RF** modulates, up converts and amplifies the signal onto a high power RF carrier

Frequency control is achieved within the Local Oscillator (LO) block. The wide bandwidth of the UMTS signal makes the phase noise performance requirements slightly easier than for GSM. A single LO synthesizer may be used for fixed duplex operation; however, for direct conversion RF architecture, it is more convenient to have two independent LO synthesizers, for the receiver and the transmitter. The implementation of compressed mode requires the receiver to tune to another channel to make a signal measurement. Fast synthesizer tuning is required to ensure that the measurement can be carried out within the defined time span. The frequency accuracy of the terminal is defined by the frequency reference and this is controlled using an AFC loop to maintain frequency synchronization with the base station. When operating using non compressed mode, the GSM and UMTS circuits will be used simultaneously and the reference frequency must be chosen to meet the requirements of both systems. The DAC and the ADC provide the interface to and from the chip rate block to the RF stages respectively. The transmit DAC must have sufficient dynamic range to meet the adjacent channel leakage requirements. The receive

ADC must have sufficient dynamic range to cater for the signal peak to average ratio and residual blocking signals, and will also

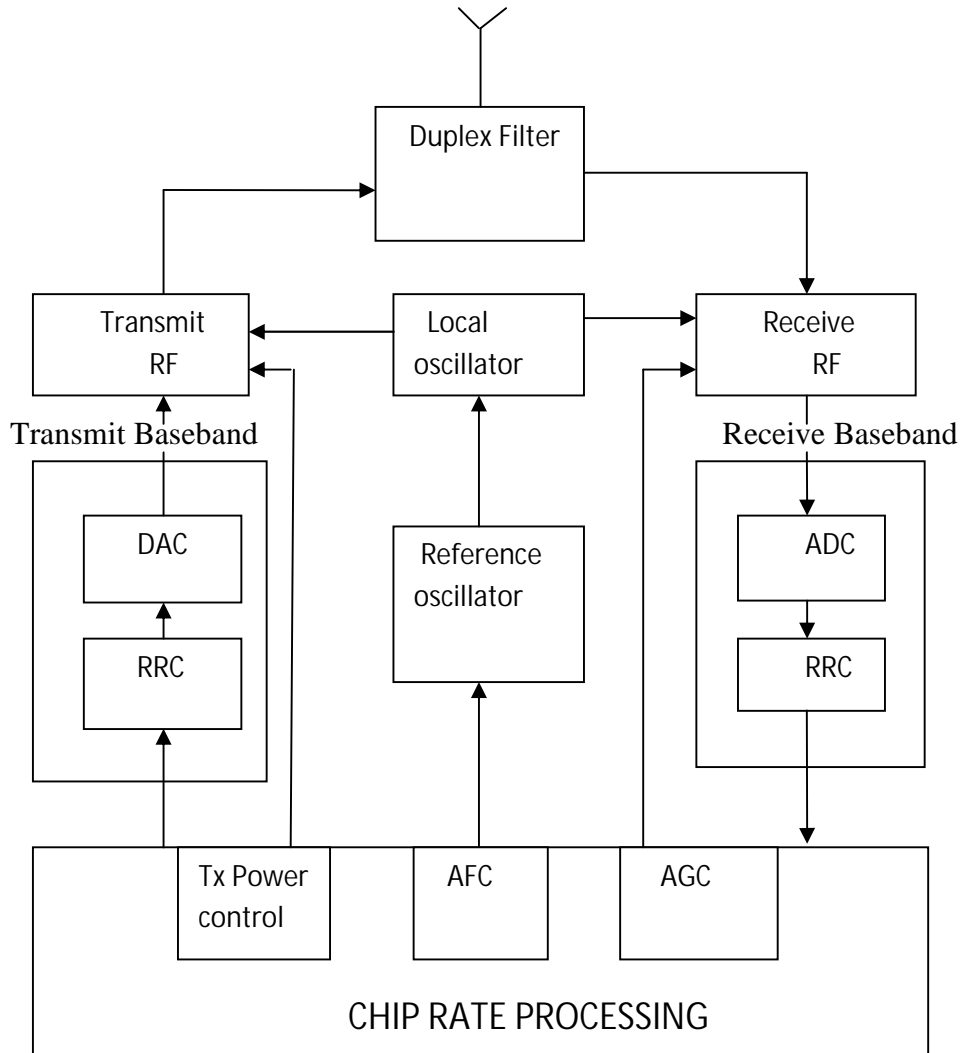


Figure 3.1 High level block diagram of an FDD transceiver[15].

depend on the AGC algorithm employed in the receiver. Both converters operate at a multiple of the chip rate, typically between four to eight times, to achieve the system's filtering requirements and provide sufficient timing resolution for the RAKE receiver. The UMTS channel filters, the pulse shape filters, are implemented using digital Root Raised Cosine (RRC) filters in the baseband. The transmit RF stage provides up conversion from a baseband signal to RF and provides power level control for the transmitter. The modulation employed for WCDMA is

different from GSM in that it does not have constant amplitude. Amplitude distortion in the signal processing components will result in unacceptable degradation of the modulation; consequently linear RF circuits must be used. Another problem with WCDMA systems occurs between mobiles that are at different distances from the base station. As many terminals share the same frequency band, a near mobile may block a far mobile if it transmits at a high power level. This problem is overcome using a power control loop in which the transmit power is adjusted 1500 times per second. Implementation of fast and accurate transmit power control circuits is vital to the performance of the UMTS system.

3.3 THE DUPLEXER: CONNECTING TRANSMITTER AND RECEIVER

The receiver and transmitter must be isolated as the transmit signal will appear as a blocking signal at the receiver. This is typically achieved using a filter with two band pass sections [16], called the duplex filter in Figure 3.1 The duplex filter should have a low insertion loss in the transmitter frequency band (Tx band), high transmitter to receiver isolation in the receiver frequency band, and a low insertion loss in the Rx band, and a high Tx to Rx isolation in the Tx band. At the transmitter, the insertion loss can be reduced if the duplex filter attenuates noise stemming from the Power Amplifier (PA), while noise generated before the PA should be filtered in the transmitter's signal path. Another design aspect which requires attention is the transmit RF leakage which can have an impact in the Rx band because of receiver nonlinearities. The receive RF stage must be able to process multiple signals received on the same frequency. To allow separation of the wanted signal from the other co channel signals, the receiver must maintain the level of orthogonality between the different code channels. The high peak to average power of the composite received signal requires linear RF signal processing, and attention must be paid to differential group delay within the receiver.

3.4 RECEIVER RF DESIGN

The heterodyne receiver is well established and used for a wide range of communication systems. They achieve good performance with respect to channel selectivity and sensitivity, at the expense of complexity and cost. However, recent developments in receiver technology and the requirement for multi standard (GSM/WCDMA) receivers operating in different frequency bands have favored Direct Conversion Receiver (DCR) architecture. A direct conversion, or

homodyne/zero Intermediate Frequency (IF), receiver converts an RF signal to a signal centered at zero frequency. The conversion employs a quadrature down conversion, using an in-phase and quadrature-phase local oscillator to allow the signal to be recovered, although spectral overlap occurs in the conversion. Using a single down conversion stage results in a simpler receiver and this translates directly into a cost advantage.

3.4.1 DIRECT CONVERSION RECEIVER

Direct conversion receivers are becoming more popular for handset applications due to their lower component count, cheaper cost and simpler frequency plan. The elimination of an IF removes the need for a Surface Acoustic Wave (SAW) filter and an IF synthesizer (LO oscillator) and mixer. Direct conversion receivers are well suited to UMTS in that the wideband signal reduces the impact of flicker noise and reduces the gain that must be provided in the baseband amplifiers. Figure 3.2 shows the block diagram of a simple direct conversion receiver for UMTS. In this example the antenna is routed to the GSM or UMTS circuits using a switch. This will allow compressed mode to be implemented. However, an alternative arrangement may be required for non compressed mode. A duplex filter is used to sum the UMTS transmit signal and to provide isolation. Both these elements introduce loss in the receiver front end and consequently degrade the sensitivity. A Low Noise Amplifier (LNA) is positioned immediately after the duplex filter to recover the signal level. The gain of the LNA component is chosen to be high enough that noise introduced in subsequent circuits has a minimal impact on the overall sensitivity. The amplified RF signal from the antenna is converted directly to baseband using a mixer circuit. If a simple down conversion was used then all spectral components of the wanted signal would be mixed to a band of 0–1.92 MHz and the signal could not be correctly demodulated. The receiver uses a pair of mixer circuits provided with in-phase (I) and quadrature-phase (Q) local oscillators so that positive and negative frequency components can be resolved by the follow on baseband processing. The output of the mixer comprises the baseband difference signal and the sum signal with a frequency of $2 f_{LO}$, where f_{LO} is the local oscillator frequency. The output also includes a number of signals resulting from distortion. The sum signal is not required and so a filter is used after each mixer so that it is attenuated to a low level. In practice, the frequency separation of the sum is so large that a simple low pass filter, e.g. a

simple RC filter, will provide adequate rejection. As there is no narrowband filtering within the RF portion of the receiver, the selectivity is defined entirely by the steepness of the baseband of

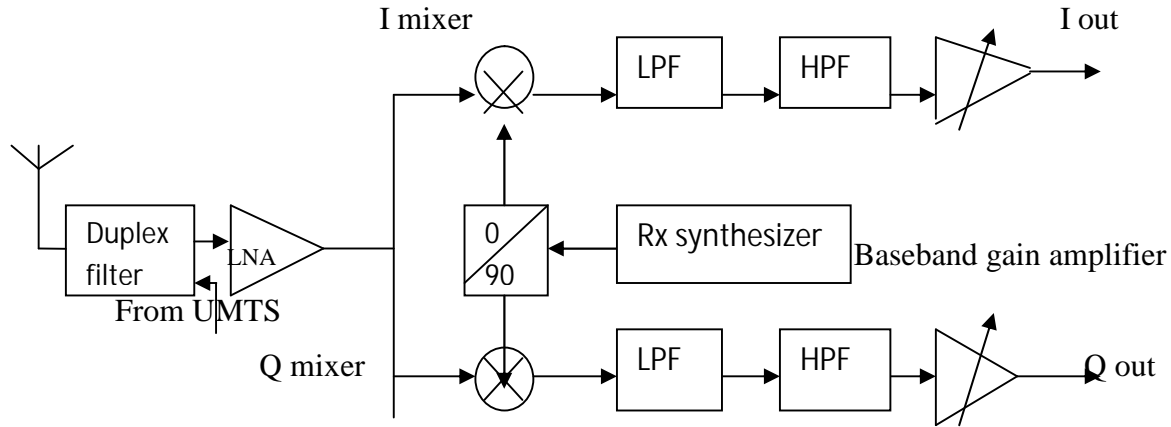


Figure 3.2 Typical direct conversion receiver[15].

filters. The filter may be implemented partially in the analog processing and partially in subsequent digital signal processing stages; however, some analog filtering is required to attenuate the level of blocking signals before further baseband amplification. As there is no narrowband filtering within the RF portion of the receiver, the selectivity is defined entirely by the baseband filters. The signal at the output of the filter stage requires amplification to increase the level so that it matches the operating level of the ADC. As the input signal level will vary depending on the propagation conditions, the amplifier gain must be variable and is typically controlled using an AGC loop. The presence of DC offsets at the output of the mixer must be considered when designing the receiver. A high pass filter is included before the amplifiers to remove DC offsets and low frequency signals to prevent the amplifier output from becoming saturated when set to high gain. The presence of DC offsets at the ADC appears as a blocking signal and can seriously degrade the sensitivity of the receiver. To overcome this problem, the output of the baseband amplifier will typically be AC coupled to the ADC so that any offset voltage introduced in the amplifiers does not impact on performance.

3.5 RECEIVER BASEBAND DESIGN

For the purpose of this section, a receive Digital Reference Point (DRP) has been defined at the output of the combined RF and baseband stage, see Figure 3.3. It is the task of the entire receiver

chain, comprising RF, analog and digital baseband, to maintain a digital signal at this digital reference point with enough fidelity (dynamic range) to allow the remaining digital signal processing stages to extract the wanted signal without undermining the performance. It must do so over component and manufacturing tolerances, UE characteristics and under various propagation conditions. A typical baseband receive signal processing chain is shown in Figure 3.3 the receive baseband is critical to the receiver's performance, because information is lost or corrupted within this block, and the receive RF can neither be recovered nor easily corrected. Also, the receiver BB functions yield some key parameters which are required to design the receiver's RF stage.

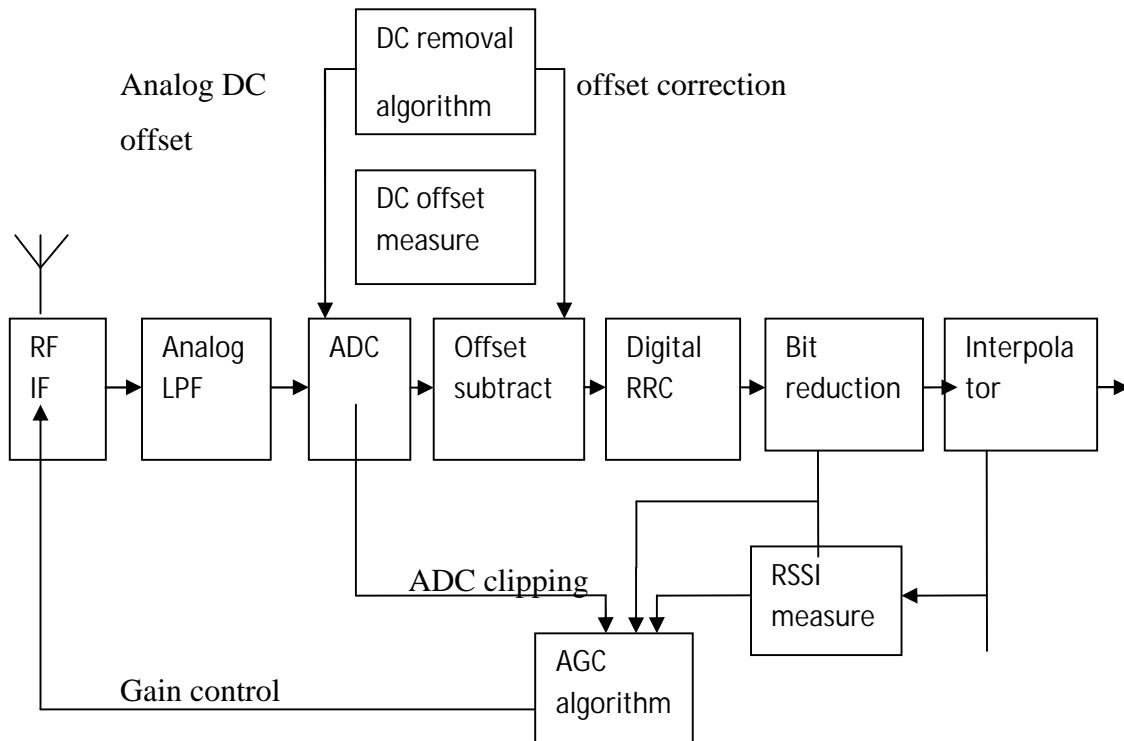


Figure 3.3 A typical baseband receiver signal processing chain [15].

3.6 TRANSMITTER BASEBAND DESIGN

The baseband signal processing [17] in a UMTS mobile terminal comprises signal conversion, conditioning and control functions which, in the receive direction, take the analog signals from

the RF receive chain and provide an adequate digital representation of the band of interest. In the transmit direction, it must provide the transmit RF chain with an analog representation of the digital transmit chain, without compromising the transmit spectrum mask. The operational and performance requirements of both the transmit and receive baseband chains are driven by a number of conflicting demands from the 3GPP specifications, the requirements of the transmit and receive RF architectures, the requirements of the digital signal processing, and the UE requirements. The baseband processing stages must provide an acceptable compromise between all these conflicting requirements.

3.7 BASEBAND MODULATION

After the uplink data has been encoded and spread, the signal must be modulated onto an RF carrier. This is achieved by forming a complex filtered signal from the control and data channels and modulating this onto the carrier using a vector modulator. The UMTS system uses a linear modulation scheme in which both the amplitude and phase of the signal vary. This achieves greater spectral efficiency but requires the use of linear RF components. In particular, the output power amplifier must be linear and the requirement to operate with output powers below the compression point reduces the efficiency and increases the power consumption. The 3GPP uplink modulation scheme employs Hybrid Phase Shift Keying (HPSK) [18], in order to relax the requirements posed on the power amplifier. HPSK exploits two signal properties, namely:

- by using a complex scrambling code, the power between the in-phase and quadrature component is equally distributed.
- using even-numbered Walsh codes, which consist of pairs of bits, the number of zero crossings are reduced by 50% and hence a reduction in peak to average power ratio of about 1 to 1.5 dB is achieved.

Therefore, uplink modulation is different from the downlink modulation. The 3GPP Uplink modulation scheme allows up to six Dedicated Physical Data Channels (DPDCHs) and one Dedicated Physical Control Channel (DPCCH) to be modulated onto the uplink carrier [19]. Figure 3.4 shows a full uplink configuration. For uplink data rates below 450 kbps, a single

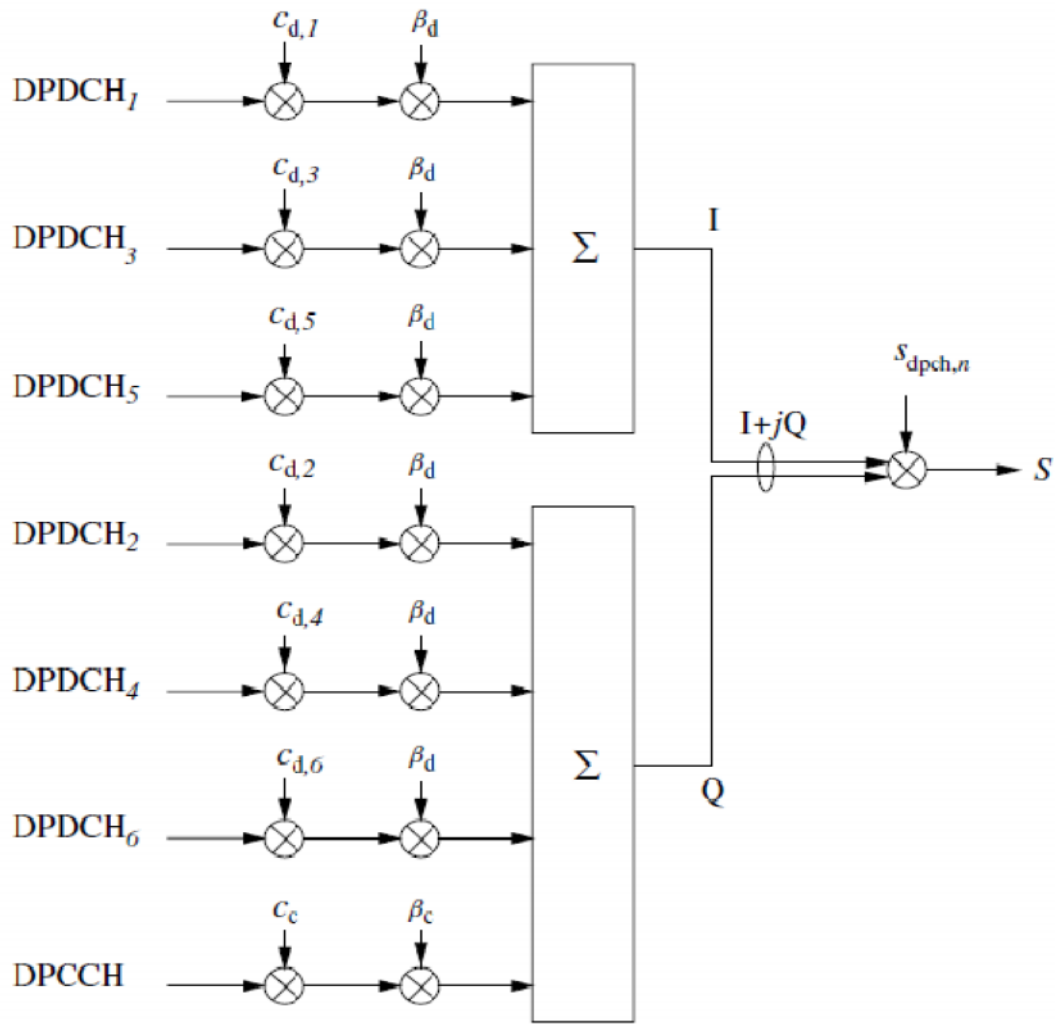


Figure 3.4 DPCH uplink channel configuration [19].

DPDCH is always used, in which case the in-phase component of the modulation is obtained from the DPDCH and the quadrature-phase component is obtained from the DPCCH. The data rate of the DPDCH can vary, whilst the data rate of the DPCCH is fixed and uses spreading code $C_{(256,0)}$ where 256 denotes the spreading factor and zero the code number.

3.8 TRANSMITTER RF DESIGN

3.8.1 RF UP CONVERSION

This section looks at the conversion from baseband I and Q signals to a modulated RF signal. Two different schemes for achieving this are shown in Figure 3.5 and Figure 3.6. These are different in the context that the scheme in Figure 3.5 modulates directly to the final RF carrier, whilst Figure 3.6 uses an intermediate frequency stage.

3.8.2 TRANSMITTER DIRECT UP CONVERSION

The main advantage of direct modulation is its low cost and small size. These factors are extremely important as handset manufacturers are constantly looking to reduce the size and cost of handsets. A typical direct conversion transmitter is shown in Figure 3.5. The local oscillator VCO (Tx synthesizer) is run at twice the transmit frequency. This has the benefit of reducing any coupling effects of the local oscillator directly to the transmit output and allows an accurate quadrature signal to be created using division by two. The baseband variable gain control amplifiers have differential I and Q inputs to reduce common mode noise and interference from other baseband circuits. The baseband gain control range is typically in the range of 20–30 dB and has to be set with a minimum resolution of 1 dB. One of the limiting factors in baseband gain control is the amount of local oscillator leakage that is present at the output of the modulator. As the signal is reduced through the baseband amplifier, the local oscillator leakage becomes more significant and the transmit signal Error Vector Magnitude (EVM) is degraded [20]. A low pass filter is employed at the output of the baseband amplifier. The filter must be designed to ensure that alias products from the DAC and noise from the DAC and baseband are adequately attenuated. It is important that the first and second adjacent channel noise and spurious emissions are more than 40 dB and 50 dB below the wanted signal respectively, as there is no other adjacent channel filtering in the transmitter after this baseband filter. The baseband signal is converted directly to the RF frequency by the modulator. Low local oscillator leakage and linearity performance are important parameters for the modulator. The RF gain control amplifier has a gain control in the range of 60 dB to 70 dB. Maximum output power level will be typically 0 dBm to 5 dBm. The amplifier can be implemented as a switched gain amplifier or as a linear controlled gain amplifier. The minimum step size requirement is 1 dB. The advantage of

a switched gain amplifier is that the transmitter is easier to calibrate to meet the various step and absolute gain requirements. The disadvantage of the switched gain amplifier is that the switching process can produce switching transients which will produce spectral components in the adjacent frequency band.

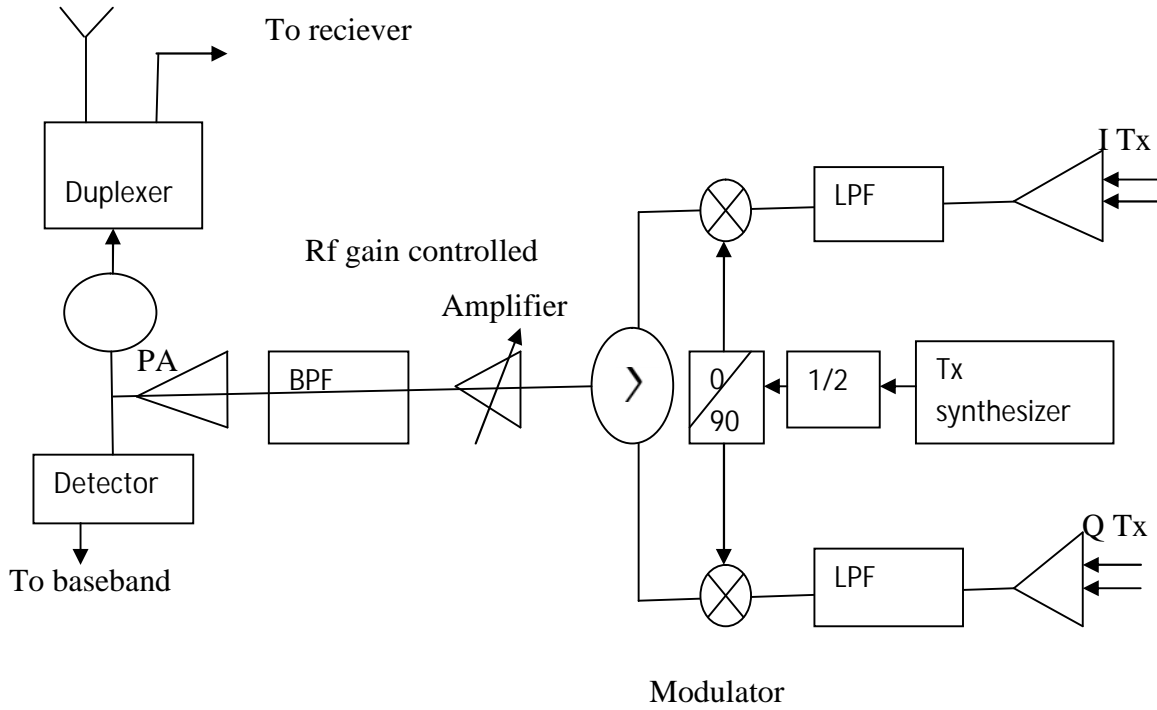


Figure 3.5 Direct conversion transmitter [15].

The power amplifier will typically have a fixed gain of between 25 dB and 30 dB depending on the number of gain stages. The amplifier has to maintain a good degree of linearity because of the nature of vector modulation. Generally, it will be a class AB amplifier with an efficiency of around 40% at maximum output power. The power amplifier is followed by an isolator. This protects the power amplifier from power reflected back from the antenna due to mismatches that regularly occur at a handset antenna. Reflected power will increase distortion in the power amplifier output stage, and lead to a failure of the ACLR specification. The duplex filter does little for the transmitter other than add loss and reduce efficiency. However, the filter does reduce the harmonics from the power amplifier and noise and local oscillator leakage. The detector allows the transmitter power to be measured by the baseband so that the power output

can be prevented from exceeding the maximum allowed output, thus it is fed from the PA output rather than the antenna port.

3.8.3 TRANSMITTER IF BASED UP CONVERSION

The main advantage of an IF based up convertor is that most of the gain control can be implemented at the IF frequency. The main disadvantage is that the transmitter requires an additional up conversion stage. A typical IF based transmitter is shown in Figure 3.6.

To Receiver

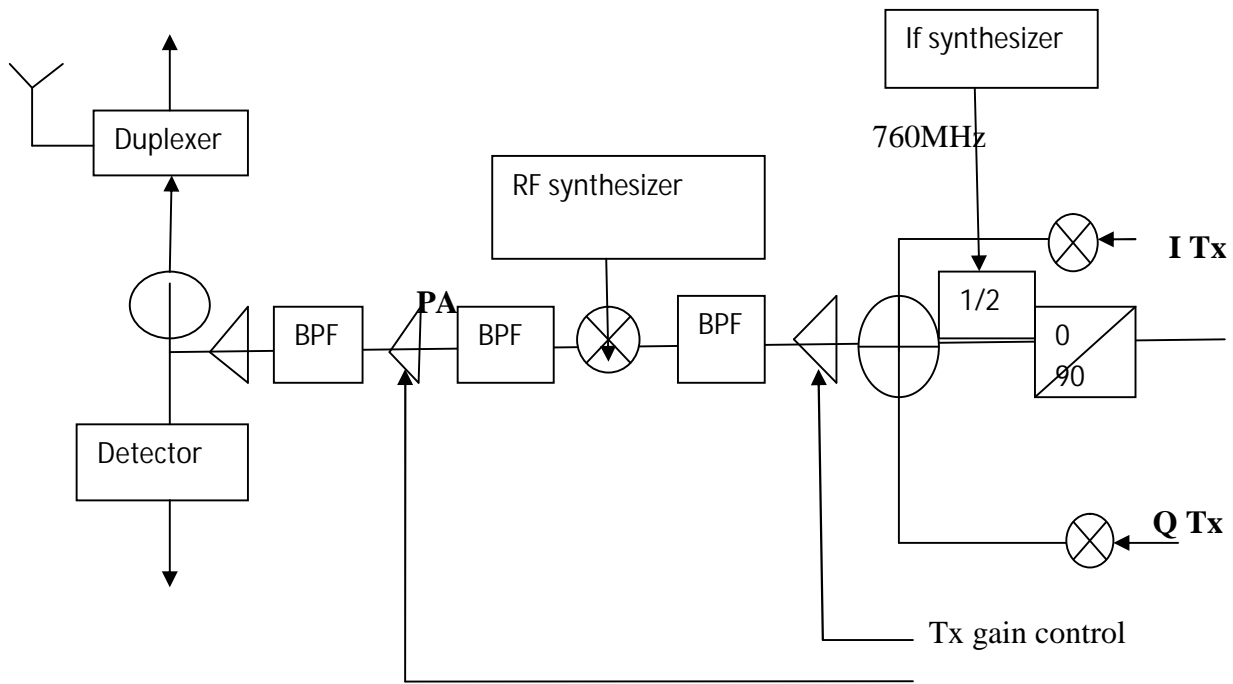


Figure 3.6 IF based transmitter [19].

Here the IF is set at 380 MHz and the RF synthesizer (LO) could be shared with the receiver if a superhetrodyne receiver design is used. The transmit I and Q signals are up converted to 380 MHz. As there is no baseband gain control, carrier leakage in the modulator is not as important as in the direct conversion case. A transmit carrier leakage of better than -20 dB is readily achievable. The 380 MHz IF filter can be implemented as an LC filter as the main requirement is to filter noise that will appear in the receive band after up conversion to the output frequency. If a narrowband SAW filter is used to filter the IF, then this would provide 15–20 dB of rejection of the adjacent channel noise and would largely remove the need for baseband filtering before

the modulator. For a typical transmitter, the control will be split so that the IF gain control amplifier will have a range of up to 60 dB while the RF gain control amplifier will have up to 35 dB of range. The power control range required for a class 3 handset is 74 dB. The transmitter is designed to have a larger range, up to 95 dB, to take into account temperature, frequency and process variations in the transmitter gain.

3.9 CONCLUSION

The UE's radio access technology, which processes the signal from and to the air interface, is broken down into three chief components Radio transceiver, Chip rate processing entity and Bit rate processing entity. The wide bandwidth of the UMTS signal makes the phase noise performance requirements slightly easier than for GSM. In the transceiver the transmit DAC must have sufficient dynamic range to meet the adjacent channel leakage requirements and the receive ADC must have sufficient dynamic range to cater the signal peak to average ratio and residual blocking signals. The receiver and transmitter must be isolated as the transmit signal will appear as a blocking signal at the receiver. This is accomplished by duplex filter which should have a low insertion loss in the transmitter frequency band (Tx band), high transmitter to receiver isolation in the receiver frequency band, and a low insertion loss in the Rx band, and a high Tx to Rx isolation in the Tx band. Recent developments in receiver technology and the requirement for multi standard (GSM/WCDMA) receivers operating in different frequency bands have favored Direct Conversion Receiver (DCR) architecture for receiver RF design. After the uplink data has been encoded and spread, the signal must be modulated onto an RF carrier. Modulation is done by Hybrid Phase Shift Keying (HPSK), in order to relax the requirements posed on the power amplifier. Then the modulated data is up converted by transmitter direct up conversion due its low cost or by IF based up conversion which has the advantage that the gain control can be implemented at the IF frequency.

CHAPTER-4

FIR FILTERS

4.1 INTRODUCTION

The focus of this chapter will be on Finite Impulse Response (FIR) filters that have only a finite number of terms in their impulse response. This chapter first gives the advantages of FIR filter over IIR filter. Then filter design techniques are discussed. There are two filter design techniques frequency sampling technique and the window method. FIR filtering is very must in WCDMA transmitter which is used as multiple access technique in 3G, because WCDMA spectrum has adjacent channel side lobes which cause interference. That is why filtering in the transmitter should be used to overcome the effect of the adjacent channel side lobes. In this chapter Nyquist filter is discussed which is a, pulse shaping filter and helps in rejecting the interference. It is also presented the criteria to attain zero ISI.

4.2 ADVANTAGES OF FIR FILTER

The main advantages of FIR filter over IIR filter are as follows [21]:

- An FIR filter is always stable.
- They are realizable.
- They provide a linear phase response under specific conditions.
- The design methods are generally linear.
- They can be realized efficiently in hardware.
- The filters start up time transients have finite duration.

4.3 FILTER DESIGN TECHNIQUES

Here it is dealt with some of the techniques used to design FIR filters: FIR filters are filters having a transfer function of a polynomial in z^{-1} and is an all-zero filter in the sense that the zeroes in the z -plane determine the frequency response magnitude characteristic. The Z transform of an N -point FIR filter is given by Eq. 4.1.

$$H(Z) = \sum_{n=0}^{N-1} h(n)z^{-n} \quad 4.1$$

FIR filters are particularly useful for applications where exact linear phase response is required. The FIR filter is generally implemented in a non-recursive way which guarantees a stable filter. FIR filter design essentially consists of two parts.

- Approximation problem
- Realization problem

The approximation stage takes the specification and gives a transfer function through four steps.

- A desired or ideal response is chosen, usually in the frequency domain.
- An allowed class of filters is chosen (e.g. the length N for a FIR filters).
- A measure of the quality of approximation is chosen.
- A method or algorithm is selected to find the best filter transfer function.

The realization part deals with choosing the structure to implement the transfer function which may be in the form of circuit diagram or in the form of a program. There are two well-known methods for FIR filter design namely [22]:

- (1) The frequency sampling technique.
- (2) The window method.

4.3.1 THE FREQUENCY SAMPLING TECHNIQUE

In this method the desired frequency response is provided. The given frequency response is sampled at a set of equally spaced frequencies to obtain N samples. Thus, sampling the continuous frequency response $H_d(w)$ at N points essentially gives us the N-point DFT of $H_d(2\pi nk/N)$. Thus by using the IDFT formula, the filter coefficients can be calculated as

$$h(n) = \frac{1}{N} \sum_{k=0}^{N-1} H(K) e^{j(2\pi n/N)k} \quad 4.2$$

Now using the above N-point filter response, the continuous frequency response is calculated as an interpolation of the sampled frequency response. The approximation error would then be exactly zero at the sampling frequencies and would be finite in frequencies between them. The smoother the frequency response being approximated, the smaller will be the error of interpolation between the sample points. One way to reduce the error is to increase the number of frequency samples. The other way to improve the quality of approximation is to make a number

of frequency samples specified as unconstrained variables. The values of these unconstrained variables are generally optimized by computer to minimize some simple function of the approximation error e.g. one might choose as unconstrained variables the frequency samples that lie in a transition band between two frequency bands in which the frequency response is specified e.g. in the band between the pass band and the stop band of a low pass filter. There are two different set of frequencies that can be used for taking the samples. One set of frequency samples are at $f_k = k/N$ where $k = 0, 1, \dots, N-1$. The other set of uniformly spaced frequency samples can be taken at $f_k = (k + 1/2)/N$ for $k = 0, 1, \dots, N-1$. The second set gives us the additional flexibility to specify the desired frequency response at a second possible set of frequencies. Thus a given band edge frequency may be closer to type-II frequency sampling point than to type-I in which case a type-II design would be used in optimization procedure.

Rabiner and Gold [23], has mentioned a technique based on the idea of frequency sampling to design FIR filters. The steps involved in this method suggested by Rabiner are as follows:

- The desired magnitude response is provided along with the number of samples. For given N , the designer determines how fine an interpolation will be used ?
- It was found by Rabiner that for designs, investigated, where N varied from 15 to 256, $16N$ samples of $H(w)$ lead to reliable computations, so 16 to 1 interpolation was used.
- Given N values of H_k , the unit sample response of filter to be designed, $h(n)$ is calculated using the inverse FFT algorithm.
- In order to obtain values of the interpolated frequency response two procedures were suggested by Rabiner. They are

(a) $h(n)$ is rotated by $N/2$ samples (N even) or $(N-1)/2$ samples for N odd to remove the sharp edges of impulse response, and then $15N$ zero-valued samples are symmetrically placed around the impulse response.

(b) $h(n)$ is split around the $N/2$ nd sample, and $15N$ zero-valued samples are placed between the two pieces of the impulse response. The zero augmented sequences are transformed using the FFT algorithm to give the interpolated frequency responses.

4.3.1.1 MERITS OF FREQUENCY SAMPLING TECHNIQUE

- Unlike the window method, this technique can be used for any given magnitude response.
- This method is useful for the design of non-prototype filters where the desired magnitude response can take any irregular shape.

There are some disadvantages with this method i.e. the frequency response obtained by interpolation is equal to the desired frequency response only at the sampled points. At the other points, there will be a finite error present.

4.3.2 THE WINDOW METHOD

In this method, from the desired frequency response specification $H_d(w)$, corresponding unit sample response $h_d(n)$ is determined using the following relation

$$h_d(n) = \frac{1}{2\pi} \int_{-\pi}^{\pi} H_d(w) e^{jwn} dw \quad 4.3$$

$$\text{Where } H_d(w) = \sum_{n=-\infty}^{\infty} h_d(n) e^{-jwn}$$

In general, unit sample response $h_d(n)$ obtained from the above relation is infinite in duration, so it must be truncated at some point say $n = M-1$ to yield an FIR filter of length M (i.e. 0 to $M-1$). This truncation of $h_d(n)$ to length $M-1$ is same as multiplying $h_d(n)$ by the rectangular window defined as

$$W(n) = \begin{cases} 1 & 0 \leq n \leq M-1 \\ 0 & \text{otherwise} \end{cases} \quad 4.4$$

thus the unit sample response of the FIR filter becomes

$$h(n) = \begin{cases} h_d(n) w(n) & \\ = h_d(n) & 0 \leq n \leq M-1 \\ = 0 & \text{otherwise} \end{cases} \quad 4.5$$

Now, the multiplication of the window function $w(n)$ with $h_d(n)$ is equivalent to convolution of $H_d(w)$ with $W(w)$, where $W(w)$ is the frequency domain representation of the window function

$$W(w) = \sum_{n=0}^{M-1} w(n) e^{-jwn} \quad 4.6$$

Thus the convolution of $H_d(w)$ with $W(w)$ yields the frequency response of the truncated FIR filter

$$H(\omega) = \frac{1}{2\pi} \int_{-\pi}^{\pi} H_d(\nu) W(\omega - \nu) d\nu \quad 4.7$$

The frequency response can also be obtained using the following relation

$$H(\omega) = \sum_{n=0}^{M-1} h(n) e^{-j\omega n} \quad 4.8$$

But direct truncation of $h_d(n)$ to M terms to obtain $h(n)$ leads to the Gibbs phenomenon effect which manifests itself as a fixed percentage overshoot and ripple before and after an approximated discontinuity in the frequency response due to the non-uniform convergence of the Fourier series at a discontinuity. Thus the frequency response obtained by using Eq. 4.8 contains ripples in the frequency domain. In order to reduce the ripples, instead of multiplying $h_d(n)$ with a rectangular window $w(n)$, $h_d(n)$ is multiplied with a window function that contains a taper and decays toward zero gradually, instead of abruptly as it occurs in a rectangular window. As multiplication of sequences $h_d(n)$ and $w(n)$ in time domain is equivalent to convolution of $H_d(\omega)$ and $W(\omega)$ in the frequency domain, it has the effect of smoothing $H_d(\omega)$.

The several effects of windowing the Fourier coefficients of the filter on the result of the frequency response of the filter are as follows:

- A major effect is that discontinuities in $H(\omega)$ become transition bands between values on either side of the discontinuity.
- The width of the transition bands depends on the width of the main lobe of the frequency response of the window function, $w(n)$ i.e. $W(\omega)$.
- Since the filter frequency response is obtained via a convolution relation, it is clear that the resulting filters are never optimal in any sense.
- As M (the length of the window function) increases, the main lobe width of $W(\omega)$ is reduced which reduces the width of the transition band, but this also introduces more ripple in the frequency response.
- The window function eliminates the ringing effects at the band edge and does result in lower side lobes at the expense of an increase in the width of the transition band of the filter.

Some of the windows commonly used are:

1. Bartlett window
2. Generalized cosine windows
3. Kaiser window with parameter β

The major advantages of using window method are their relative simplicity as compared to their methods and ease of use. The fact that well defined equations are often available for calculating the window coefficients have made this method successful.

There are following problems in filter design using window method:

- This method is applicable only if $H_d(w)$ is absolutely integrable i.e. only if Eq. 4.3 can be evaluated. When $H_d(w)$ is complicated or cannot easily be put into a closed form mathematical expression, evaluation of $h_d(n)$ becomes difficult.
- The use of windows offers very little design flexibility e.g. in low pass filter design, the pass band edge frequency generally cannot be specified exactly since the window smears the discontinuity in frequency. Thus the ideal LPF with cut-off frequency f_c , is smeared by the window to give a frequency response with pass band response with pass band cutoff frequency f_1 and stop band cut-off frequency f_2 .
- Window method is basically useful for design of prototype filters like low pass, high pass, band pass etc. This makes its use in speech and image processing applications very limited [21].

4.4 NYQUIST FILTERS

Nyquist defined a type of filters which have good time domain properties and good frequency domain properties. Nyquist filters results in zero ISI at the optimum sampling points for the filtered data. A practical example of Nyquist filter is raised cosine filter.

4.5 RAISED COSINE FILTER

Raised cosine filters are commonly used in digital data communication systems to limit inter symbol interference . The impulse response of a raised cosine filter is zero at each adjacent symbol period. Specifying a raised cosine filter is straight-forward, and requires only the "roll off factor" (often called "beta" or "alpha"), the sample rate, the symbol rate, and the number of FIR taps. The ideal raised cosine filter frequency response consists of unity gain at low frequencies, a raised cosine function in the middle, and total attenuation at high frequencies. The width of the middle frequencies are defined by the roll off factor constant Alpha, ($0 < \text{Alpha} \leq 1$). In filter solutions, the pass band frequency is defined as the 50% signal attenuation point. The group delay must remain constant at least out to 15 to 20 dB of attenuation. When the pass band

frequency of a raised cosine filter is set to half the data rate, then the impulse response of Nyquist first criteria is satisfied in that the impulse response is zero for $T = NT$, where N is an integer, and T is the data period[24].

Mathematically, the frequency response may be written as

$$\begin{aligned}
 F(\omega) &= 1 & \omega < \omega_c(1-\alpha) \\
 F(\omega) &= 1 + \cos(\pi(\omega - \omega_c(1 - \alpha))) / 2 & \omega_c(1-\alpha) < \omega < \omega_c(1+\alpha) \\
 F(\omega) &= 0 & \omega > \omega_c(1+\alpha)
 \end{aligned}
 \tag{4.9}$$

The ideal raised cosine filter frequency response is shown below:

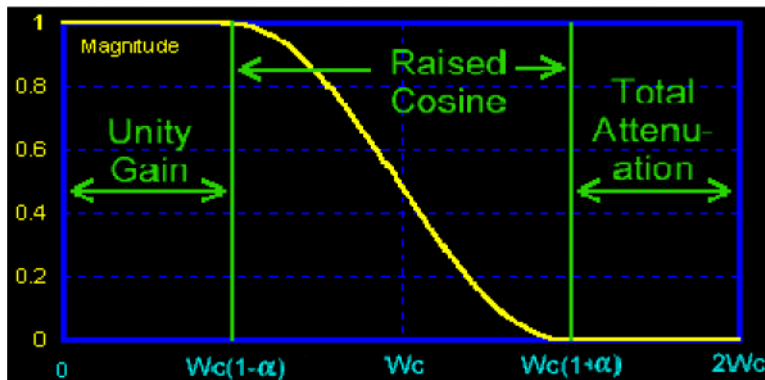


Figure 4.1 Ideal raised cosine frequency response[21].

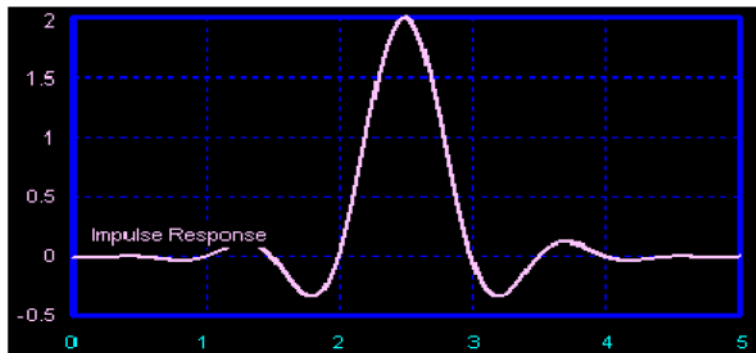


Figure 4.2 Ideal raised cosine impulse response (Alpha = 0.35)[21].

4.6 FEATURE REQUIRED TO ACHIEVE ZERO ISI

The impulse response plot below displays the feature required to achieve zero intersymbol interference. It is shown in Figure 4.3 that starting from the center (peak) tap, it can be seen that the impulse response passes through zero at each interval of eight taps. (The design specification gave eight samples per symbol, so each set of eight taps corresponds to one sample period.) However, if it does not "synchronize" the symbol sampling to the right instant (which is the center tap of the filter) it will begin to have intersymbol interference. For example, if it starts just one tap off from the center, then look at the response of the filter eight taps away from that, it will be seen that the impulse response has a non-zero value. This illustrates that even with the correct filter; intersymbol interference can occur if the symbol sampling synchronization is not correct.

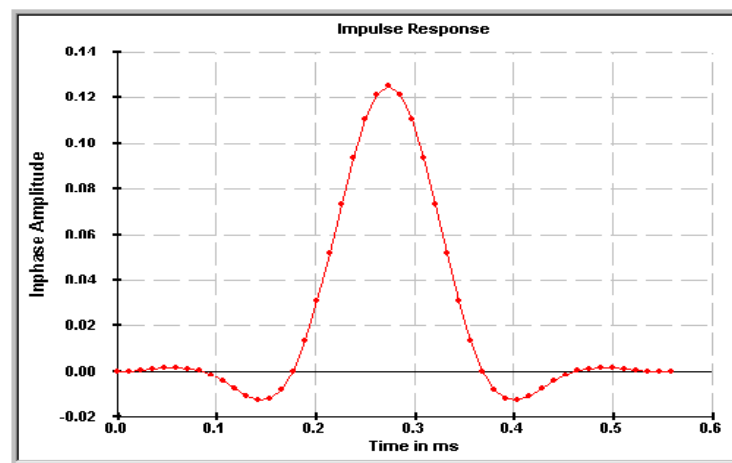


Figure 4.3 Ideal raised cosine impulse response (Alpha = 0.35)[21].

4.7 FILTER TRANSFER FUNCTION SPLIT BETWEEN TRANSMITTER AND RECIEVER

In addition to filtering in the transmitter, it also needs filtering in the receiver. The filtering in the receiver is required to reject the noise and interference present at the receiver. The effects of transmit filter and receive filter are combined together. It is the combined effect of transmit filter and receive filter that must be considered and which must have the properties of Nyquist filter shown in Figure 4.4.

$$H_T(f) = \text{Transmitter Transfer Function}$$

$$H_R(f) = \text{Receiver Transfer Function}$$

$$H_{\text{sys}} = H_T(f) H_R(f)$$

Where H_{sys} is the system transfer function. Now the transfer function for the transmitter is given by the square root of H_{sys} or by the transfer function of root raised cosine filter.

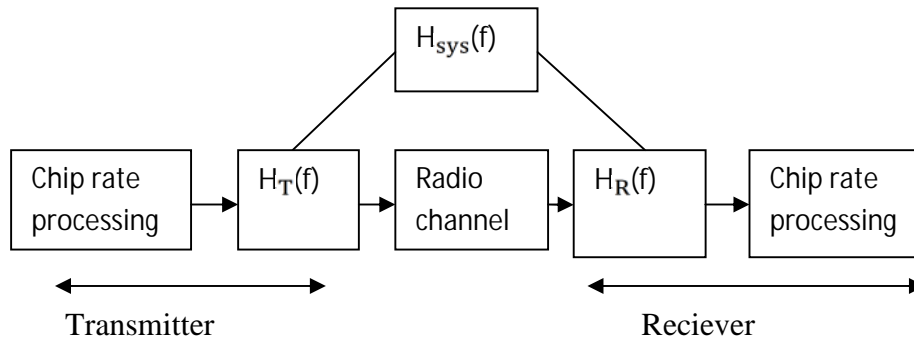


Figure 4.4 Filter distribution between transmit and receive functions to meet Nyquist criteria [7].

4.8 ROOT RAISED COSINE FILTER

FIR filters in the transmission and reception chains are used to implement root raised cosine (of Nyquist filter). The time domain representation for the RRC filter is given in Eq. 4.10.

$$h(t) = \frac{\sin[(1-a)\pi t/T] + 4a(t/T)\cos[(1+a)\pi t/T]}{((\pi t/T)[1 - (\frac{4at}{T})^2])^{1/2}} \quad 4.10$$

Where $T = 1/\text{chip rate} = 1/3.84 \text{ MHz} = .260\mu\text{s}$ for WCDMA.

The simplest method of implementing a filter such as RRC filter is to use FIR filter. The FIR consists of a delay line that delays the samples of the signal to be filtered. The output from each delay element is multiplied by the quantity $h(i)$. The output of the multipliers is then added to produce the output of the filter, which is the filtered version of the input signal. The quantity $h(i)$ is a sample of the impulse response of the filter being implemented. The input to the filter is the signal to be filtered, sampled at an appropriate rate. In designing the filter, the sampling rate also needs to be considered. The higher the sampling rate the easier it is after a digital to analog converter to remove high frequency components generated in the sampling process. The higher the sampling rate, the faster the converter must operate, which in general will lead to greater cost

and greater power consumption. To keep the approach realistic, it has been chosen an oversampling factor of 4 (thus a sampling rate $f_s = 4 \times \text{chip rate} = 15.36\text{MHz}$) as a compromise between high performance on a hand and cost and complexity on the other hand.

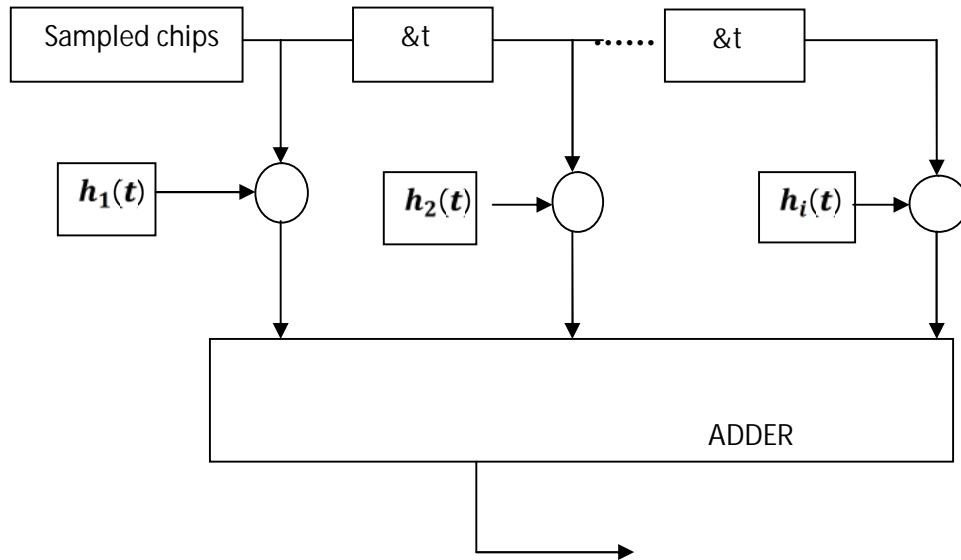


Figure 4.5 FIR Implementation using sampled impulse response [7].

4.9 CONCLUSION

Finite Impulse Response (FIR) filters have only a finite number of terms in their impulse response. FIR filters are advantageous over IIR filters in way that they are always stable, can be realizable, provide a linear phase response, design methods are generally linear, can be realized efficiently in hardware, and start up time transients have finite duration. There are two main filter designing techniques sampling method and window method. Merits of sampling method are unlike the window method, this technique can be used for any given magnitude response, this method is useful for the design of non-prototype filters where the desired magnitude response can take any irregular shape. Nyquist filters have good time domain properties and good frequency domain properties and results in zero ISI at the optimum sampling points for the filtered data. Raised Cosine filter is the type of Nyquist filter.

CHAPTER-5

FIR FILTER IN FREQUENCY DOMAIN FOR WIRELESS COMMUNICATION SYSTEM.

5.1 INTRODUCTION

This chapter gives design methodology for implementation of FIR filter in the frequency domain and time domain. The filter coefficients for these two implementation are taken by sampling the time representation and frequency representation of RRC filter. The RRC filter transfer function is used to obtain Raised Cosine filter which is a pulse shaping filter. Then the performance evaluation is made by comparing the simulation results of the two filter implementation on the basis of computational complexity, inter symbol interference rejection depending on average gain in immunity against ISI, Error Vector Magnitude (EVM), Peak distortion and AWGN rejection. Then it is shown that by choosing frequency domain filter coefficients as the samples of the ideal frequency transfer function can boost the performances of such a filters mainly in terms of immunity against inter symbol interference.

5.2 DESIGN METHODOLOGY

In order to evaluate the performance of proposed FIR implementation method in the frequency domain, especially in terms of immunity against inter symbol interference and noise rejection, two FIR filter implementations are considered:

- A time domain FIR filter of which the coefficients are obtained by sampling the RRC impulse response with an oversampling factor of 4 and which processes the I and Q paths successively;
- The frequency domain FIR filter with coefficients extracted as will be discussed in section 5.4 and which combines the I and Q paths into one complex signal and processes the two paths simultaneously.

In designing the filter, the sampling rate also needs to be considered. The higher the sampling rate the easier it is after a digital to analog converter to remove high frequency components generated in the sampling process. The higher the sampling rate, the faster the converter must

operate, which in general will lead to greater cost and greater power consumption. To keep the approach realistic, it is chosen an oversampling factor of 4 (thus a sampling rate $f_s = 4 \times \text{chip rate} = 15.36 \text{ MHz}$) as a compromise between high performance on one hand and cost and complexity on the other hand. Then it is created two real random input signals of length 100 points with values between -1 and 1 to simulate inputs on the I and Q paths, and included 3 null samples between each adjacent values to simulate an oversampled signal by a factor of 4. These signals were passed through a classical time domain filter to get the transmitted signal. The two filter implementations (the time domain and proposed frequency domain) are then applied to the transmitted signal.

5.3 TIME DOMAIN IMPLEMENTATION OF FIR FILTER

Third Generation (3G) communication systems use WCDMA technique to transmit and receive data. The spectrum of a signal of the type used in WCDMA has adjacent channel side lobes. These side lobes introduce interference. Number of considerations discussed in [7] proposed to the use of a FIR filter in the WCDMA transmitter, together with a matched filter in the receiver in a way to get a Nyquist filter as the combined filter response. In the time domain, the FIR operation is achieved by convolution.

$$\text{FIR}[x(t)] = x(t) * h(t) \quad 5.1$$

Where $h(t)$, is filter coefficients, $*$ is the convolution product, $x(t)$ is discrete input, $\text{FIR}[x(t)]$ implies the FIR filtering in frequency domain. FIR filters in the transmission and reception chains are used to implement root raised cosine (RRC) filters of which the combined transfer function results in a raised cosine filter (a type of Nyquist filter). The time domain representation for the RRC filter is given in Eq. 5.2 [25].

$$h(t) = \frac{\sin[(1-a)\pi t/T] + 4a(t/T)\cos[(1+a)\pi t/T]}{((\pi t/T)[1 - (\frac{4at}{T})^2])} \quad 5.2$$

Where $T = 1/\text{chip rate} = 1/3.84\text{Mhz} = .260\mu\text{s}$ for WCDMA.

So the filter coefficients for time domain implementation of FIR filter are extracted by sampling the impulse response given by Eq. 5.2 shown in Figure 5.1. These samples are taken as the coefficients of the time domain implementation filter.

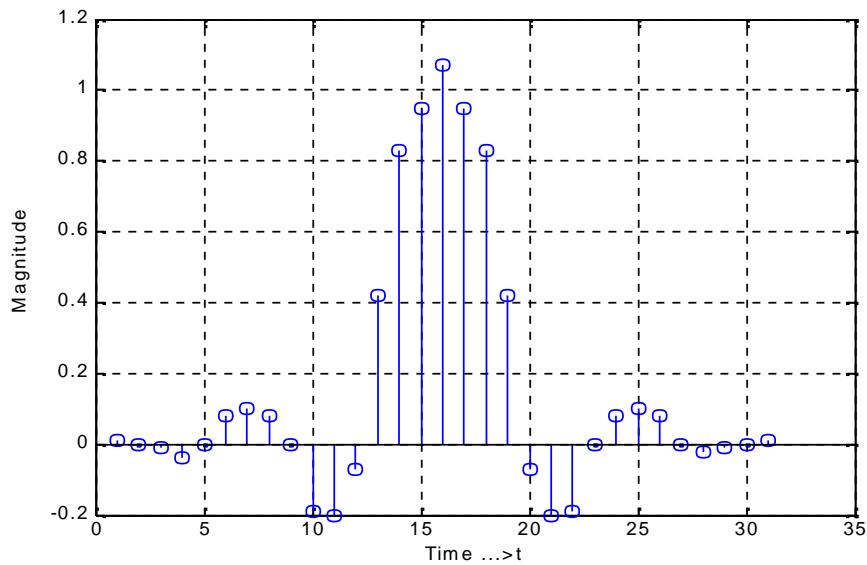


Figure 5.1 Samples of time domain impulse response of RRC filter

5.4 FREQUENCY DOMAIN IMPLEMENTATION OF FIR FILTER

Implementing FIR filters by FFT is not an innovative idea. Many researchers have proposed FIR filters in the frequency domain in the past two decades [2, 3, 4]. But no previous work is proposed for a frequency-domain FIR filter for the UMTS standard, mainly because of computational complexity problems. Instead of performing the FFT of the filter's coefficients extracted by sampling the RRC time representation like shown above in Eq. 5.2, take the ideal square root of the raised cosine filter transfer function shown in Eq. 5.3 and sample it to get the coefficients that should be used in the frequency domain FIR filter. The resultant coefficients are then used in the frequency domain implementation of the FIR filter.

$$\begin{aligned}
 H(f) &= T & 0 \leq \text{mod}(f) \leq m \\
 H(f) &= T/2(1 + \cos[\pi t/a(\text{mod}(f) - m)]) & m \leq \text{mod}(f) \leq M \\
 H(f) &= 0 & \text{mod}(f) > M
 \end{aligned} \tag{5.3}$$

$$\text{Where } m = 1 - \frac{a}{2T} \quad M = 1 + \frac{a}{2T}$$

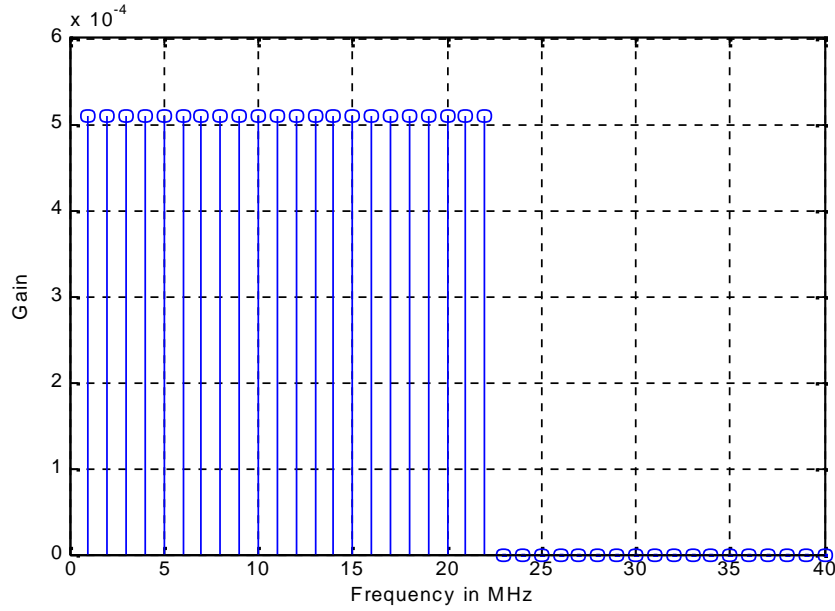


Figure 5.2 Samples of root raised cosine frequency response.

The FIR filtering in frequency domain is implemented as follows

$$\text{FIR}_f = F^{-1} [H.F(x)] \quad 5.4$$

Where FIR_f denotes the FIR filtering in frequency domain H is the Fourier transform of filter coefficients $h(t)$, (\cdot) is the convolution product, (\cdot) is the dot product and $F(x)$ is the Fast Fourier transform of discrete input $x(t)$. The frequency domain filter proposed here directly processes the I and Q paths. This is done by combining the paths before entering the filter into complex samples that will be treated by the direct and inverse FFT then breaking the output again into two paths. In fact, from the linearity property of the Fourier transform, it can be stated that

$$\begin{aligned} \text{FIR}_f(X_I + jX_Q) &= F^{-1}[F(X_I + jX_Q)H] \\ &= F^{-1}[F(X_I)H + jF(X_Q)H] \\ &= \text{FIR}_f(X_I) + j\text{FIR}_f(X_Q) \end{aligned} \quad 5.5$$

where FIR_f denotes the FIR filtering in frequency domain and X_I and X_Q are the inputs on the I and Q paths respectively.

5.5 SIMULATION RESULTS AND PERFORMANCE EVALUATION

5.5.1 COMPUTATIONAL COMPLEXITY IN TIME DOMAIN FIR FILTER

To process one sample in the time domain, a symmetric FIR filter needs $\frac{N}{2}$ multiplications and $N - 1$ additions. This means that a FIR filter executes $0.5(3N-2)$ operations per sample. This processing should be executed for both I and Q paths, which means that the actual number of operations per sample (a couple of an I-sample and a Q-sample) doubles and becomes $3N - 2$. Consequently, the computational complexity (the number of operations per second without distinguishing between addition and multiplication) of a FIR filter operating at frequency F is given a:

$$CC_{\text{FIRt}} = (3N - 2) F \quad 5.6$$

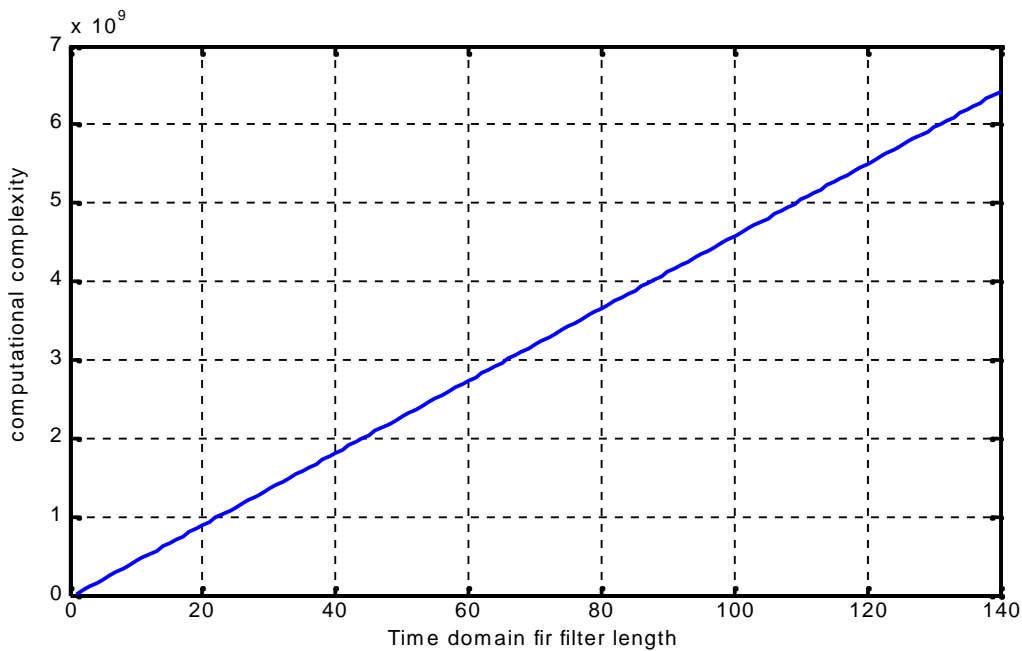


Figure 5.3 Computational complexity in time domain FIR filter.

5.5.2 COMPUTATIONAL COMPLEXITY IN FREQUENCY DOMAIN FIR FILTER

The frequency domain FIR implementation, uses the overlap-add method [26] which consists of decomposing the signal into simple components, processing each of the components, and recombining the processed components into the final signal. If the input signal is segmented into

sections of length L and a FIR filter of length N is to be implemented, a FFT of length L+N -1 or more should be performed to avoid time aliasing. In this study, the input signal is segmented into sections of length N and thus FFT of length P where P is the minimal power of 2 greater than or equal to 2N can be used. Each FFT of length P requires $5P \log_2(P)$ operations and processes N samples at a frequency F to which it has to be added 6 operations per sample to perform the complex multiplication by the filter coefficients H. Hence, the computational complexity of the frequency domain FIR filters of length N as:

$$CC_{\text{FIR}} = 2(5P/N \log_2 P + 3) F \quad 5.7$$

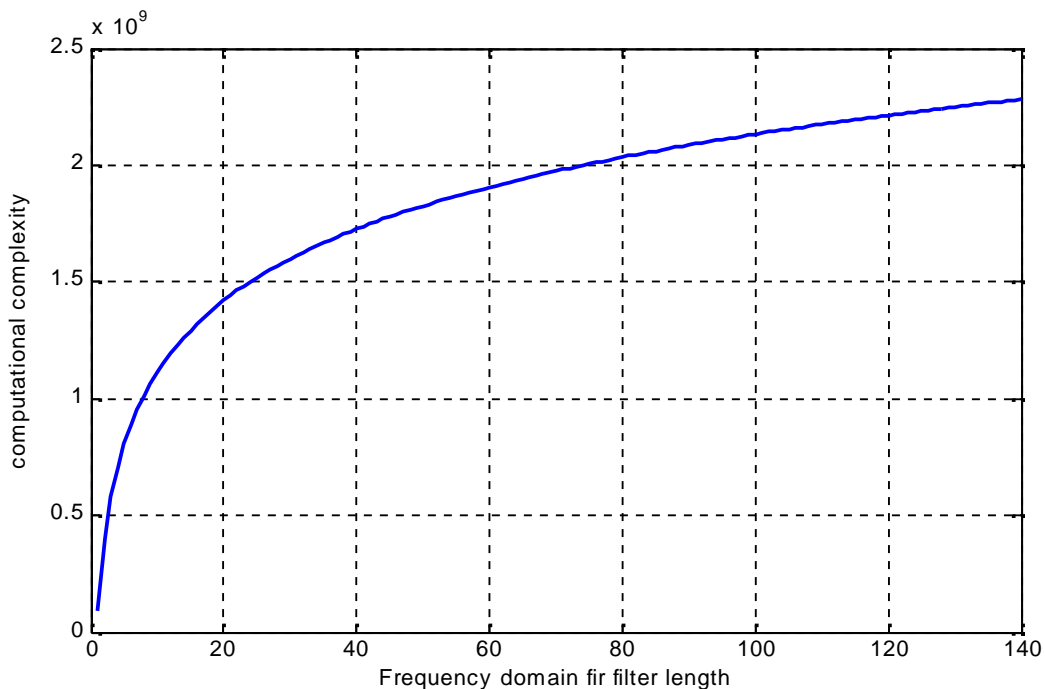


Figure 5.4 Computational complexities in frequency domain FIR filter.

Comparing the computational complexities of the 2 different implementations of the FIR filter for different filter lengths observing Figure 5.5. From this Figure, it can be found that it is more advantageous to implement FIR filters in the frequency domain for higher lengths which is generally the case of the UMTS standard because Table 5.1 shows that the computational complexity of time domain filter increases linearly whereas in frequency domain FIR filter first it increases sharply for lower filter length and afterwards it attains almost a constant value. So frequency domain implemented FIR filters have less computational complexity than time domain

FIR filters for higher filter length but classic frequency domain implementation is normally more computationally complex than a time domain implementation for small filter lengths.

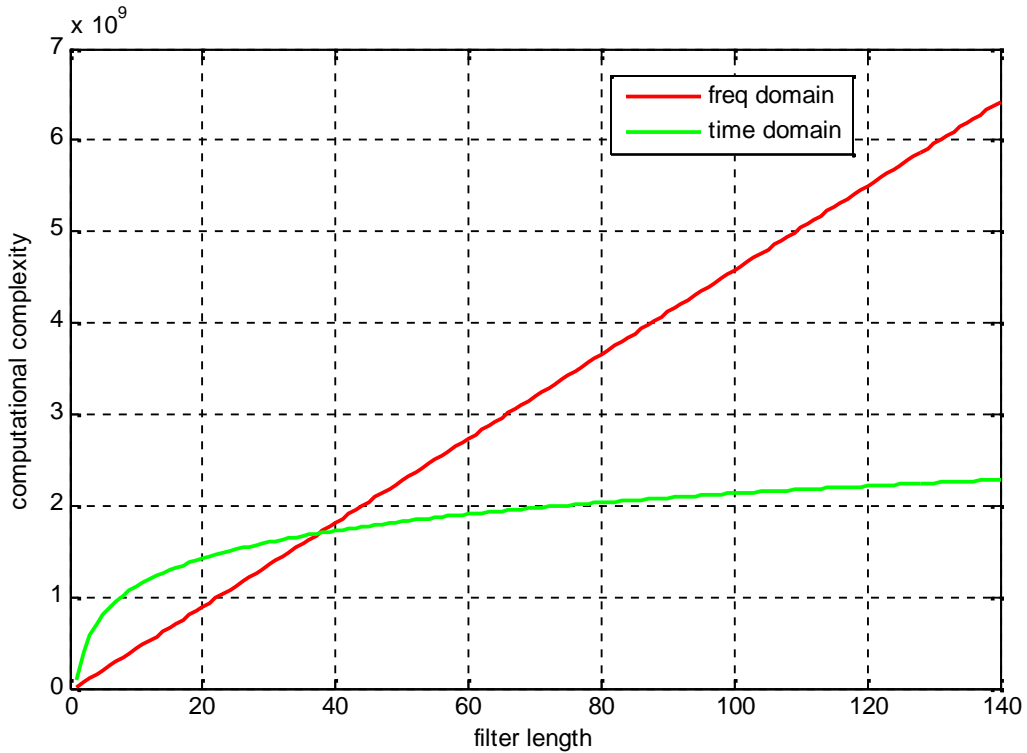


Figure 5.5 Comparison of computational complexity of two filters in time domain and frequency domain.

Table 5.1 Comparison of computational complexity of two filters in Mega Operations per Second (MOPS)

Sr. No	Filter length	Time domain filter	Frequency domain filter
1	20	800 MOPS	1400 MOPS
2	40	1800 MOPS	1750 MOPS
3	60	2700 MOPS	1900 MOPS
4	80	3600 MOPS	2100 MOPS
5	100	4600 MOPS	2300 MOPS
6	120	5500 MOPS	2400 MOPS
7	140	6400 MOPS	2500 MOPS

5.5.3 GAIN IN IMMUNITY AGAINST INTERSYMBOL INTERFERENCE IN TIME DOMAIN FIR FILTER

The main property of Nyquist filters is that they result in zero inter symbol interference (ISI) at the optimum sampling point for filtered data [7]. The ISI is measured as the variance of the error between the samples of the input signal and those of the received signals at the optimum sampling points. The gain in immunity against ISI is 51% in the case the case of time domain FIR filter as obtained in Figure 5.6.

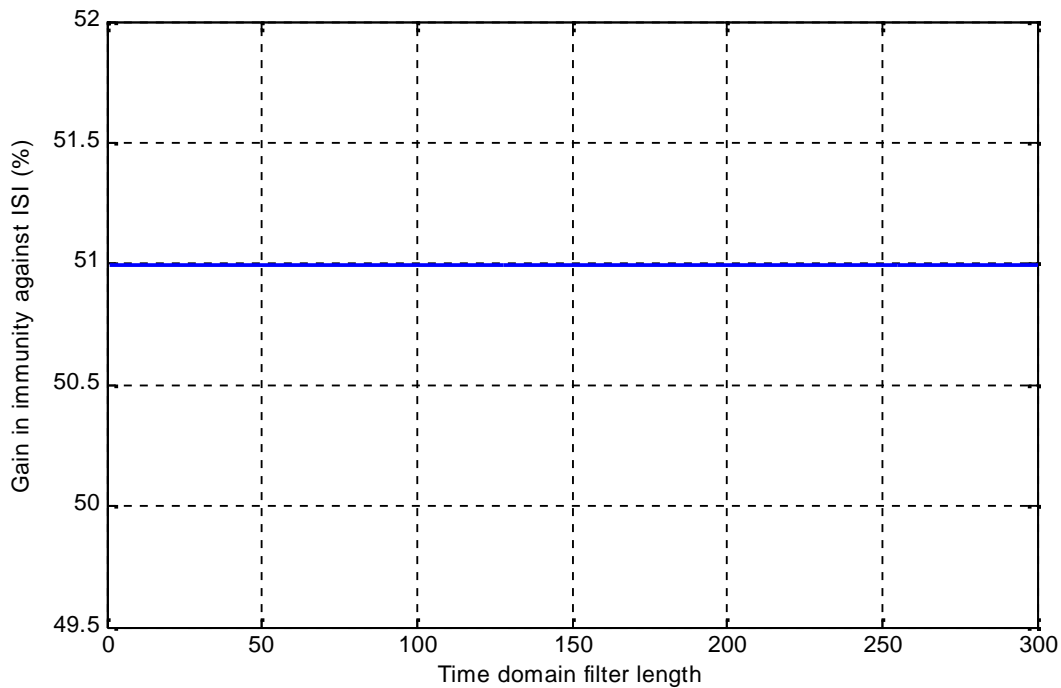


Figure 5.6 Gain in immunity against ISI in time domain FIR filter.

5.5.4 GAIN IN IMMUNITY AGAINST INTERSYMBOL INTERFERENCE IN FREQUENCY DOMAIN FIR FILTER

The ISI is measured as the variance of the error between the samples of the input signal and those of the received signals at the optimum sampling points. From Figure 5.7 the gain in immunity against ISI is 74.5% in the case of frequency domain FIR filter.

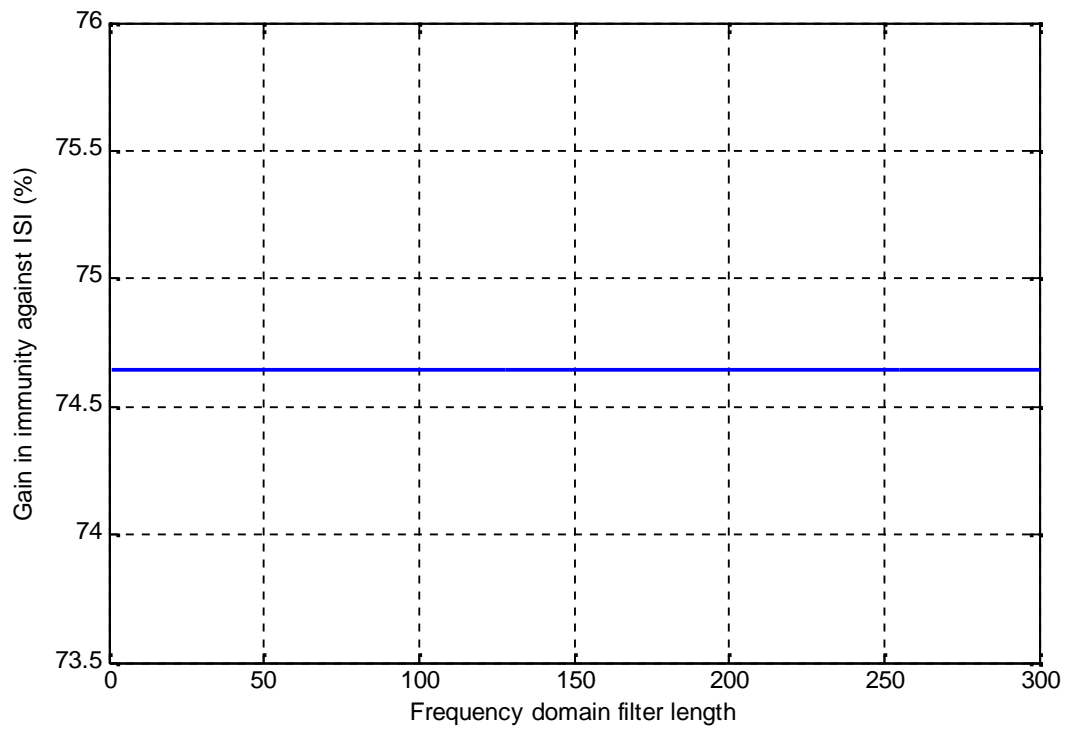


Figure 5.7 Gain in immunity against ISI in frequency domain FIR filter.

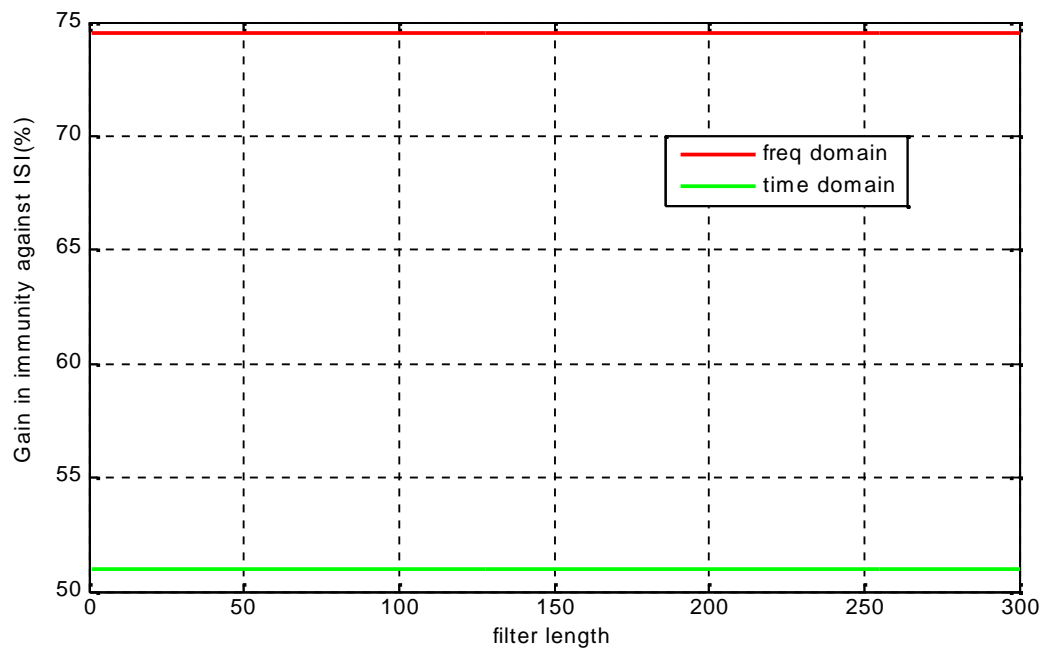


Figure 5.8 Comparison of gain in immunity against ISI of two filters in time domain and frequency domain.

Table 5.2 Comparison of gain in immunity against ISI of two filters.

Sr.No	Filter length	Time domain filter	Frequency domain filter
1	50	50.9%	74.5%
2	100	50.9%	74.5%
3	150	50.9%	74.5%
4	200	50.9%	74.5%
5	250	50.9%	74.5%
6	300	50.9%	74.5%

As it is clear from the Figure 5.8 and Table 5.2 that gain in immunity against inter symbol interference is greater in the case of frequency domain filtering .So they are more immune to ISI than time domain FIR filters.

5.5.5 ERROR VECTOR MAGNITUDE IN TIME DOMAIN FIR FILTER

It is another parameter to indicate ISI. In literature discussing UMTS radio receivers, a parameter called EVM is commonly used to indicate the amount of ISI [27]. The UMTS standard provide the following definition of EVM: “The Error Vector Magnitude is a measure of the difference between the reference waveform and the measured waveform. This difference is called the error vector. The EVM result is defined as the square root of the ratio of the mean error vector power to the mean reference power expressed as a percentage”. These technical specifications define also a minimum requirement for the EVM: the EVM shall not exceed 17.5% in the user equipment radio transmission and reception in the FDD mode [25]. The formula used to calculate the EVM is presented in (Eq. 5.10) where P_{error} is the root mean square power of the error

vector is and $P_{\text{Reference}}$ is the root mean square power of the ideal transmitted signal. Figure 5.9 gives the Error vector magnitude in time domain FIR filter.

$$\text{EVM}(\%) = \sqrt{\frac{P_{\text{error}}}{P_{\text{Reference}}}} * 100 \quad 5.8$$

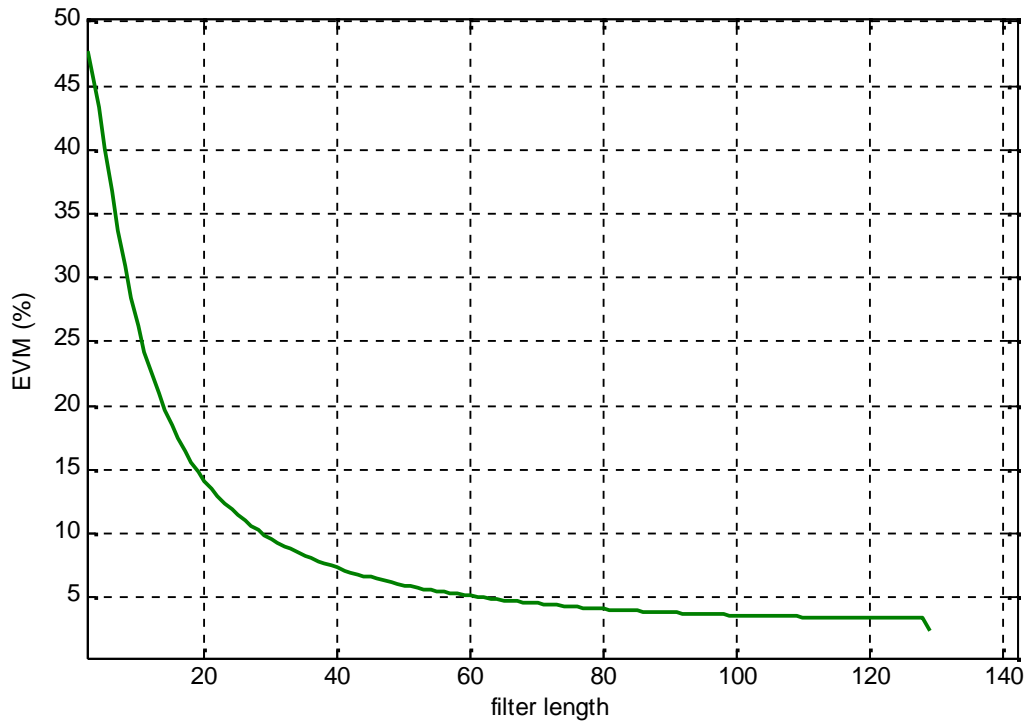


Figure 5.9 Error vector magnitude in time domain FIR filter.

5.5.6 ERROR VECTOR MAGNITUDE IN FREQUENCY DOMAIN FIR FILTER

The Error Vector Magnitude is a measure of the difference between the reference waveform and the measured waveform. This difference is called the error vector. The EVM result is defined as the square root of the ratio of the mean error vector power to the mean reference power expressed as a percentage”. The formula used to calculate the EVM is presented in Eq.5.10 where P_{error} the root mean square power of the error vector is is and $P_{\text{reference}}$ is the root mean square power of the ideal transmitted signal. Figure 5.10 gives the Error Vector Magnitude in frequency domain FIR filter.

$$\text{EVM (\%)} = \sqrt{\frac{P_{\text{error}}}{P_{\text{Reference}}}} * 100$$

5.9

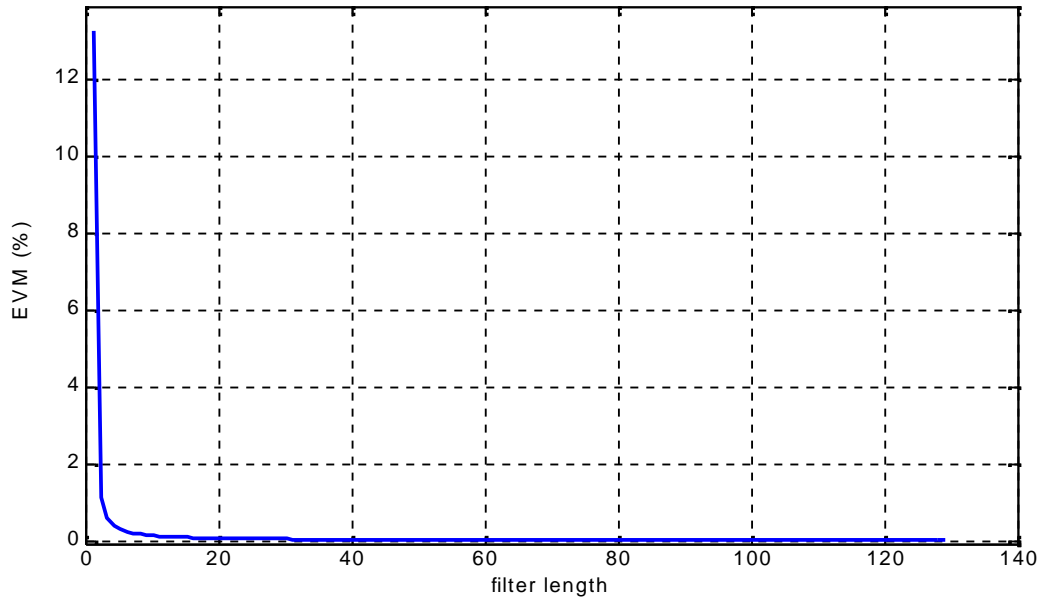


Figure 5.10 Error vector magnitude in frequency domain FIR filter.

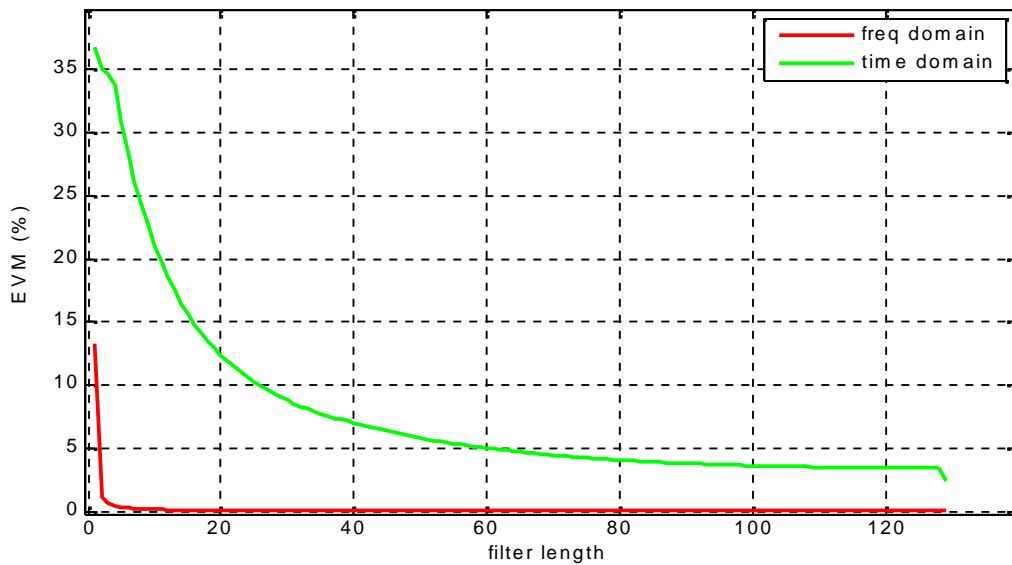


Figure 5.11 Comparison of EVM of two filters in time domain and frequency domain.

Table 5.3 Comparison of EVM of two filters.

Sr.No	Filter length	Time domain filter	Frequency domain filter
1	10	25%	14%
2	20	14%	.1%
3	40	7.5%	.1%
4	60	5%	.1%
5	80	4.5%	.1%
6	100	4.5%	.1%
7	120	3.5%	.1%

Observing the Figure 5.11 it is found that EVM obtained in frequency domain FIR filters 14% which satisfies the condition that the EVM shall not exceed 17.5% in the user equipment radio transmission and reception in the FDD mode. Simulations with different filter lengths and for the two different FIR implementations discussed in this study shows that the frequency domain implementation using the sampled ideal filter transfer function is always better than the time domain implementation. From Figure 5.11, it is seen that the frequency domain implementation ensures that the minimum requirement of EVM is met even with a FIR filter of length 8. Table 5.3 gives the comparison of two filter implementation

5.5.7 PEAK DISTORTION IN TIME DOMAIN

Another parameter commonly used to measure the ISI of transmit-receive filter combination is the peak distortion given by Equation 5.11.

$$D_p(\text{dB})=20\log_{10}[(2 \sum_{k=1}^{\infty} h(\frac{N}{2} + kM))/(h(\frac{N}{2}))] \quad 5.10$$

h is the impulse response of the transmit-receive filter combination, N is the length of this response and M is the oversample factor. Furthermore it has been assumed that h is symmetric. In order to extract h , a pulse signal is passed into two sets of transmit-receive filters. Here a transmit and a receive filters implemented in the time domain. The oversample

factor is always $M = 4$. Figure 5.14 gives the peak distortion obtained in frequency domain implementation of FIR filter.

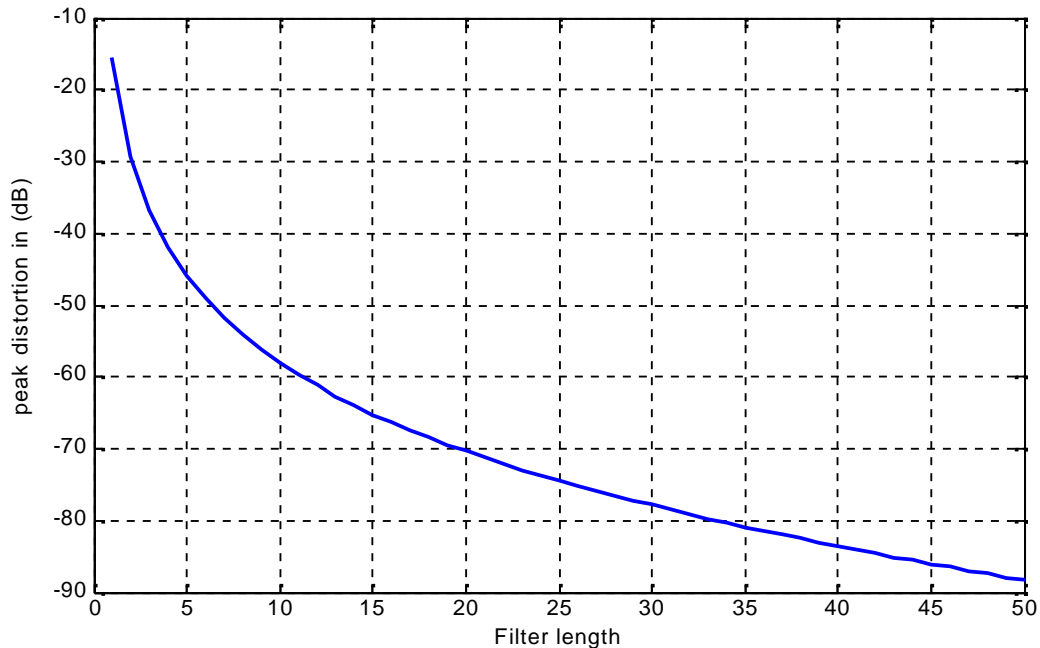


Figure 5.12 Peak distortion in time domain FIR filter.

5.5.8 PEAK DISTORTION IN FREQUENCY DOMAIN

Figure 5.13 gives the peak distortion obtained in frequency domain implementation of FIR filter. This is obtained from the same Eq. 5.10.

$$D_p(\text{dB}) = 20 \log_{10} \left[\frac{2 \sum_{k=1}^{\infty} h\left(\frac{N}{2} + kM\right)}{h\left(\frac{N}{2}\right)} \right]$$

Here a transmit and a receive filters are implemented in the frequency domain. h is the impulse response of the transmit-receive filter combination, N is the length of this response and M is the oversample factor.

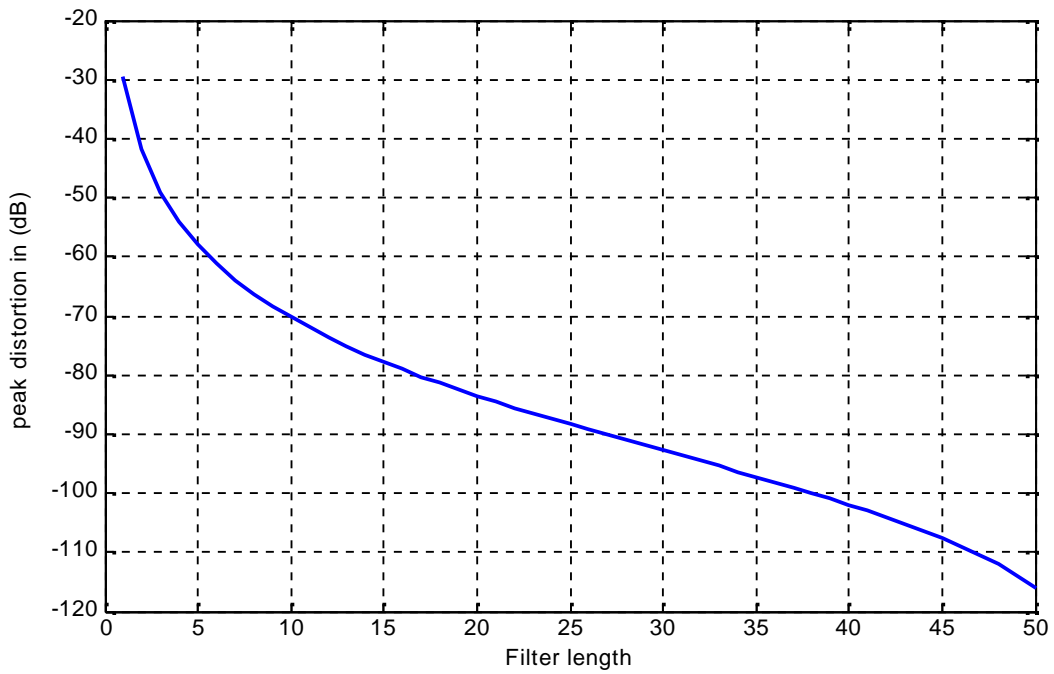


Figure 5.13 Peak distortion in frequency domain FIR filter.

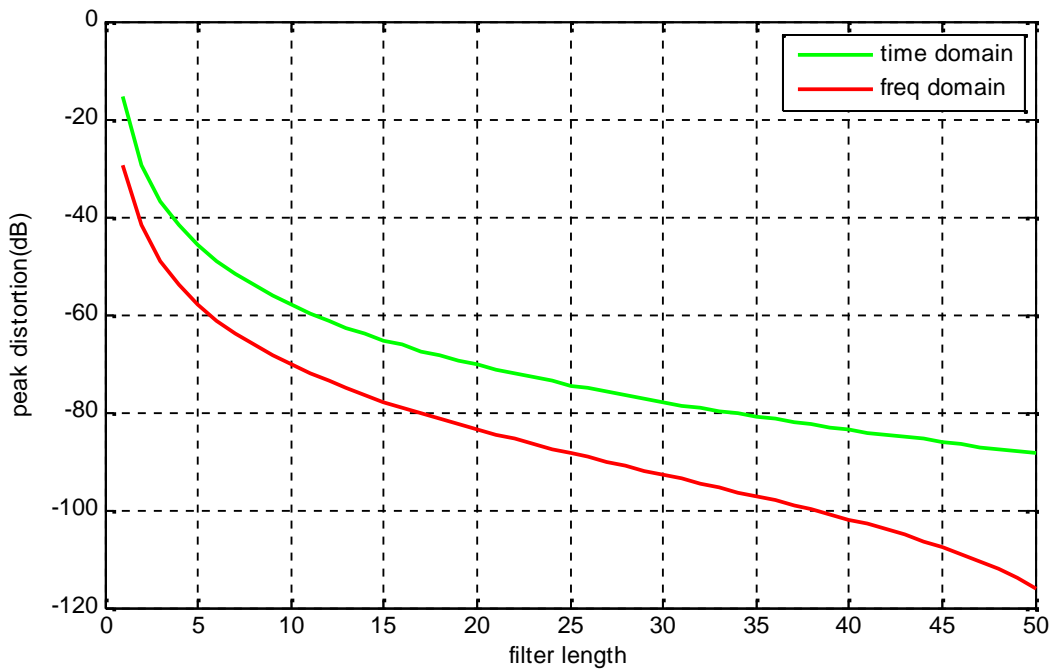


Figure 5.14 Comparison of peak distortion of two filters in time domain and frequency domain.

Table 5.4 Comparison of peak distortion of two filters.

Sr. no	Filter length	Time domain filter	Frequency domain filter
1	1	-15 dB	-30 dB
2	5	-45 dB	-58 dB
3	10	-58 dB	-70 dB
4	15	-65 dB	-78 dB
5	20	-70 dB	-84 dB
6	25	-74 dB	-88 dB
7	30	-78 dB	-94 dB
8	35	-81 dB	-98 dB
9	40	-82 dB	-104 dB

Observing the Figure 5.14 it is found that peak distortion obtained in frequency domain FIR Filter is less than the time domain filter. Table 5.4 clearly shows that peak distortion decreases in both the cases with increase in filter length but it is lesser in case of frequency domain FIR filter. So it confirms that frequency domain FIR filter presents less peak distortion than the time domain filter.

5.5.9 ADDITIVE WHITE GAUSSIAN NOISE REJECTION IN TIME DOMAIN.

The essential reason for implementing the raised cosine filter as two RRC filters, one in the transmitter and the other in the receiver, is that filtering in the receiver is essential to reject the noise and interference present in the receiver [7]. To eliminate the Additive White Gaussian Noise (AWGN) caused by the channel, the transmit and receive filters transfer functions should

be identical. Here in Figure 5.15 the Additive White Gaussian Noise (AWGN) is added to the transmitted signal which has signal to noise ratio SNR of 16dB. This noisy signal is then passed through FIR filter implemented in time domain this will reject the added (AWGN) noise as it can be seen in Figure 5.16.

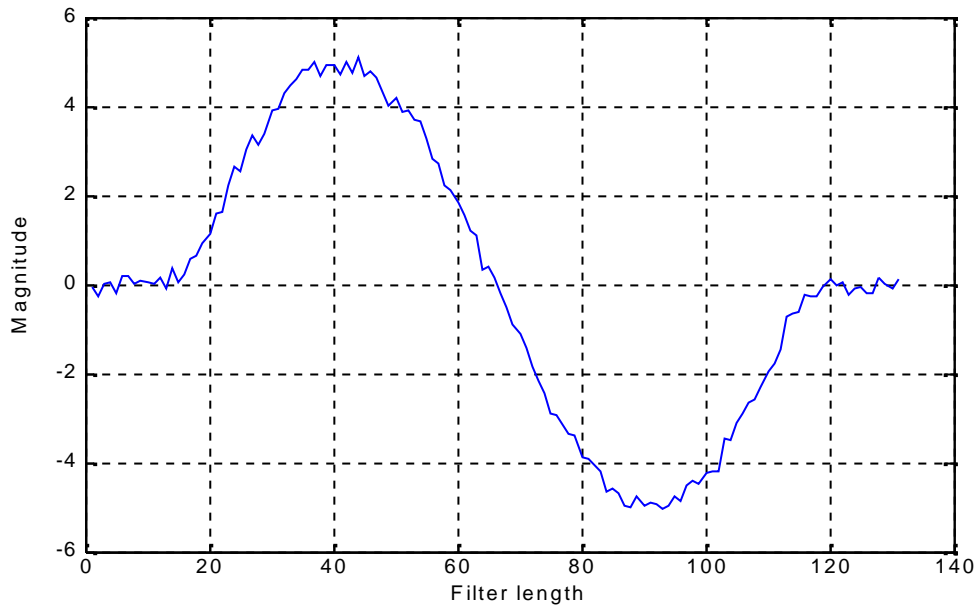
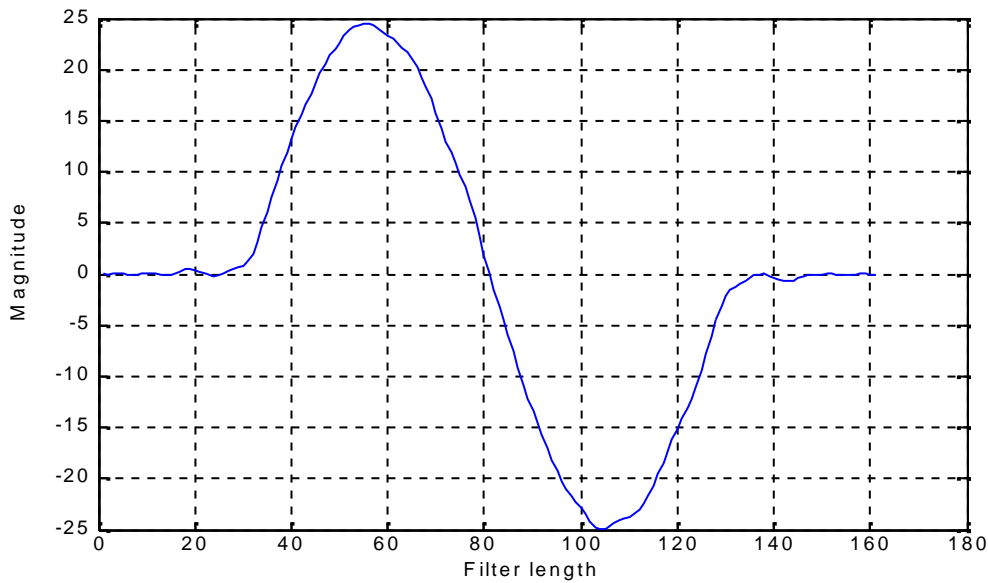


Figure 5.15 AWGN added to transmitted signal with SNR 16 dB in time domain.



5.16 AWGN rejected by the FIR filter implemented in time domain.

Figure

5.5.10 ADDITIVE WHITE GAUSSIAN NOISE REJECTION IN FREQUENCY DOMAIN.

Here again the same procedure is followed to reject the AWGN. Two frequency domain FIR filters are required one at the transmitter and other at the receiver. These two filters are designed by the RRC filter transfer function, the combined effect results the raised cosine filter, which acts as a pulse shaping filter. This raised cosine filter helps in rejecting the AWGN that is caused by the channel. Figure 5.17 shows the transmitted signal in the frequency domain and to which AWGN is added. The SNR of the signal is 16dB. This noisy signal is then passed through FIR filter implemented in frequency domain which will reject the AWGN in the signal. AWGN rejected signal is shown in figure 5.18.

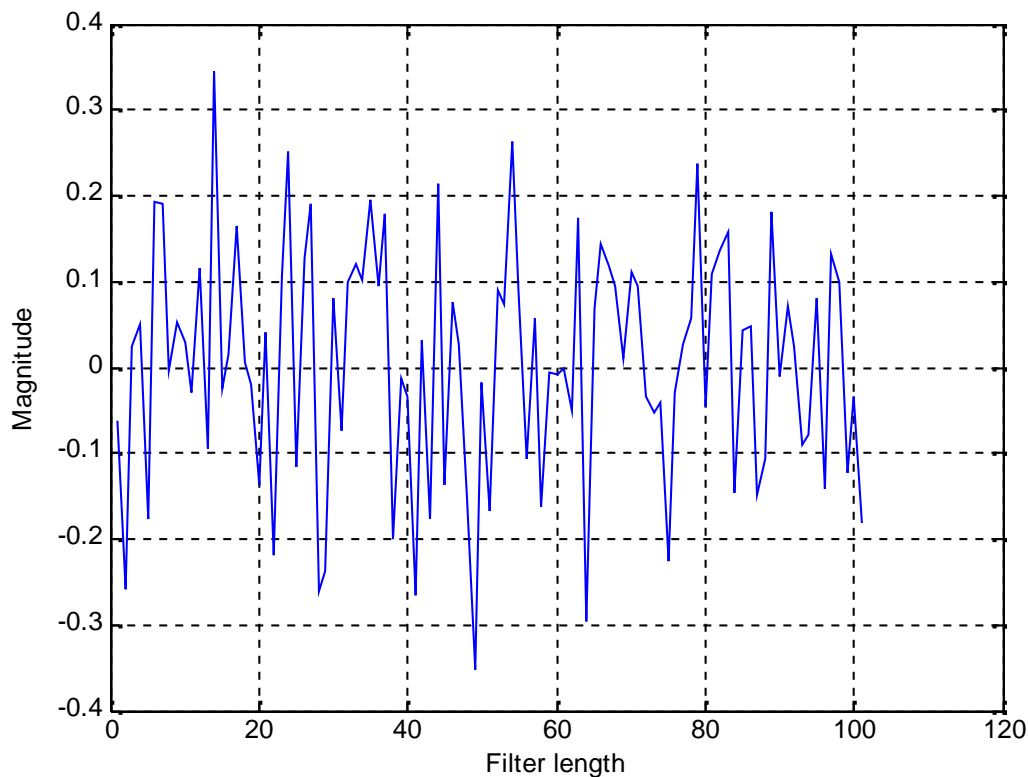
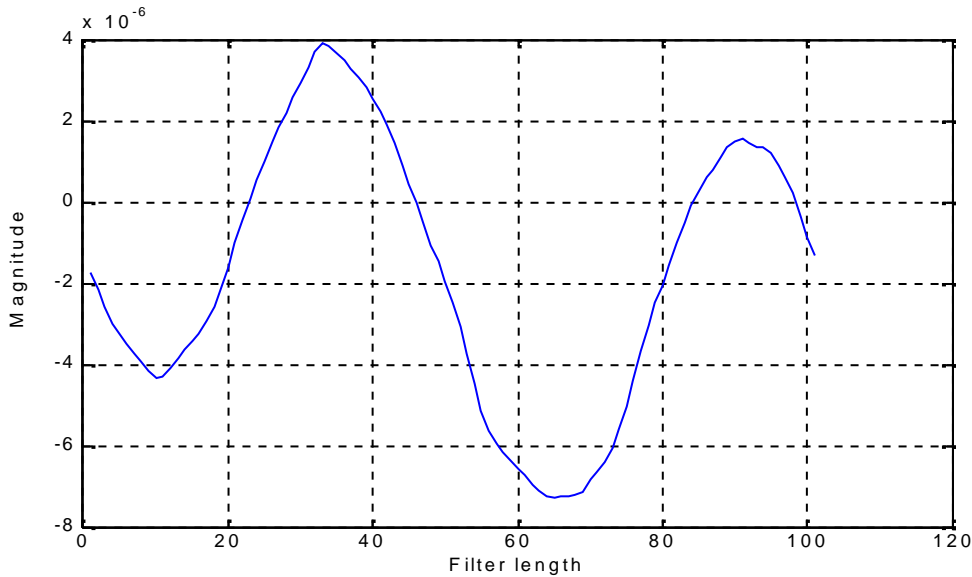


Figure 5.17 AWGN added to transmitted signal with SNR 16 dB in frequency domain.



Figure

5.18 AWGN rejected by the FIR filter implemented in frequency domain.

5.6 CONCLUSION

So observing Figure 5.5 and Table 5.1 it is estimated that the computational complexity of time domain filter increases linearly whereas in frequency domain FIR filter first it increases sharply and afterwards it attains almost a constant value. So frequency domain implemented FIR filters have less computational complexity than time domain FIR filters for higher filter length but classic frequency domain implementation is normally more computationally complex than a time domain FIR filters for small filter lengths. From Figure 5.11 and Table 5.3 it is found that EVM obtained in frequency domain FIR filters is 14% which satisfies the condition the EVM shall not exceed 17.5% in the user equipment radio transmission and reception in the FDD mode. Figure 5.8 and Table 5.2 depicts that gain in immunity against ISI is 50.9% for time domain FIR filter whereas it is 74.5% in case of frequency domain FIR filter. Figure 5.14 and Table 5.4 gives the comparison between two filters in terms of peak distortion. It is found that peak distortion decreases in both the cases with increase in filter length but it is lesser in case of frequency domain FIR filter. So it confirms that frequency domain FIR filter presents less peak distortion than the time domain filter. Then Figure 5.15 shows that AWGN noise added to the signal is rejected by time domain FIR filter and Figure 5.17 shows that AWGN noise added to the signal is equally well rejected by frequency domain FIR filter. So it does not degrade the performance of frequency domain FIR filter.

CHAPTER-6

CONCLUSION & FUTURE SCOPE

An attempt was made in the present work to study comparison between a classical time domain FIR implementation, and frequency domain FIR implementation, in terms of computational complexity, intersymbol interference rejection on the basis of average gain in immunity against ISI, Error Vector Magnitude (EVM) and peak distortion. After taking all the results and comparing them on the basis of the parameters that were selected for the performance of evaluation, it is concluded that frequency domain implementation of FIR filter is better than the time domain implementation for UMTS standard. It is evident from the comparison on the basis of computational complexity that it is more advantageous to implement FIR filters in the frequency domain for lengths 44 and more which is generally the case of the UMTS standard. The gain in immunity against ISI is 74.5% in the case of frequency domain FIR filter whereas in case of time domain FIR filter it is 51%. So FIR filter implemented in frequency domain are more immune to ISI. The other factor which determines the ISI is error vector magnitude EVM. For 3G EVM should be less than 17%. So EVM obtained in the case of frequency domain is 14% which satisfies the criteria for 3G, whereas in case of time domain filter EVM is 35% for low filter length. In case of AWGN rejection Additive white Gaussian noise is added to the transmitted signal with a Signal-to-Noise Ratio (SNR) of 16dB. The noisy signal is passed through the two types of receive filters and the result is observed again. In both the cases it is found that the noise is rejected equally by both types of filters. This means that employing the frequency domain implemented filter will not degrade the system's performance concerning the rejection of the AWGN. So all these results show that frequency domain implementation of FIR filter is better than the time domain implementation for UMTS standard. The peak distortion obtained in frequency domain FIR filter is less than the time domain filter. So it confirms that the frequency domain FIR filter presents less peak distortion than the time domain filter. So it is better to implement the FIR filter in frequency domain which boosts the performance and gives a better ISI rejection. Finally it is concluded that implementation of FIR filter in frequency domain for wireless communication system is advantageous and recommended for 3G communication.

FUTURE SCOPE

The future scope of the method of implementation of FIR filter in frequency domain is that nowadays different multiple access techniques are used, this method extracts similarities between different multiple access without having to implement in hardware all the functions needed by these different technologies. This method provides the improvement of wireless communications and their quality of service to enable them to support a multi-standard function with both high performance and reduced cost. Due to the nature of the wide research topic, there are still several areas of improvement for future work in this frequency domain implementation of FIR filter. The implementation can be carried out and tested for other wireless communication standards.

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