Synchronization and User Detection of Multi point to Multi point Asynchronous OFDM Links

A project Report
submitted by

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Master of Technology

Under the guidance of

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Thesis Certificate

This is to certify that the thesis titled **Synchronization and User Detection of Mulit point to Multi piont Asynchronous OFDM Links**, submitted by **Sureshbabu Gutti**, to the Indian Institute of Technology, Madras, for the award of the degree of **Master of Technology**, is a bona fide record of the research work done by him under our supervision. The contents of this thesis, in full or in parts, have not been submitted to any other Institute or University for the award of any degree or diploma.

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Finally, and most importantly, I would like to thank my parents for their support, encouragement, quiet patience and unwavering love were undeniably the bedrock upon which my life have been built.

Abstract

Orthogonal Frequency Division Multiplexing (OFDM) is one of the famous technologies that supports very high data rate. It is one of the best techniques to overcome adverse effects of multipath spread. OFDMA (Orthogonal Frequency Division Multiple Access) is an extension of OFDM. It permits many users to transmit simultaneously on different subcarriers per OFDM symbol.

In multipoint to multipoint asynchronous OFDM links, not only figure out the frame boundaries and Carrier Frequency Offsets but also receiver needs to know strongest transmitting user. In this project we have developed a multipoint to multipoint asynchronous OFDMA frame work for multi user aeronautical system which is constrained by severe Doppler offset, fairly high data rate requirements, asynchronous users leading to interference and low latency. The major aspect in this project is to identify the transmitted user at the receiver with low complexity but efficient algorithms.

Interference can occur at the level of frequency hopping or in the symbol boundaries. To mitigate this interference at the level of frequency hopping, data is repeated in frequency domain and the symbol boundary problem is solved by implementing linear FIR filters on received data. Generalized Chirp Like (GCL) sequences of prime length are chosen as preamble. User identification problem has been resolved at the receiver by the use of these GCL sequences correlation properties. Estimation of user identification number has been performed at different SIR profiles and at different percentages of time domain preamble overlap.

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Abbreviations

CFO	Carrier Frequency Offset
CP	Cyclic Prefix
dB	Decibel
FFT	Fast Fourier Transform
GCL	Generalized Chirp Like
IBI	Inter Block Interference
ICI	Inter Carrier Interference
IFFT	Inverse Fast Fourier Transform
ISI	Inter Symbol Interference
LB	Lower Band
OFDM	Orthogonal Frequency Division Multiplexing
QPSK	Quadrature Phase Shift Keying
SINR	Signal to Interference Ratio
SNR	Signal to Noise Ratio
UB	Upper Band

Notations

f_d	Doppler Shift
f_{offset}	Frequency offset
Δf	Subcarrier spacing
ε	Normalized frequency offset
ε_{est}	Estimated normalized frequency offset
$n_{\max}$	Index of maximum value
N	FFT size
N_G	GCL sequence length
T_s	Sampling time

Chapter 1

Introduction

Wireless communication framework is dependent on the type of resources available and the requirements of the users. Cellular communication is between users spread all over the world. There are smaller scales of communication systems like WiFi and Bluetooth which are spread over tens of meters and have different channel characteristics and environmental issues compared to other types of communication systems. Each type of system needs its own special framework and design to tackle the issues specific to that system.

In this project, designed system is a set of transmitters and receivers in an aeronautical scenario and we need to design a communication framework for the same. This system has the availability of a 102.4 MHz bandwidth. Unlike conventional cellular communication systems, the number of users is limited to a maximum of 30 users. Also, there are no major obstacles in the vicinity, and there is almost always a direct line of sight path from the transmitter to the receiver. All the users are not synchronized, so different users start transmission at different instances of time.

1.1 Motivation

In asynchronous data transmission, transmitter has to allocate resources in such a way that there is a minimum probability of collision among users. This is the major criterion of transmission side. In the receiver side, receiver needs to estimate the user before decoding process. To ensure the user estimation works perfectly, transmitter has to use special preamble pattern in the OFDM frame. Also user detection problem completely depends on preamble structure in the frame, if preamble is corrupted by the data of other user then the total decoding process goes wrong. So preamble locations have to be separated from data locations in the frequency domain for effective utilization of filters to filtering out data of its own or other user for initial frame acquisition. For this purpose resources in frequency domain allocated in such a way that there are fixed locations for preamble and fixed locations for data.

Another major problem in multi user asynchronous transmission is preamble preamble overlap. Frame structure in time domain should be ensuring that minimal probability of collision in the preambles. Long length frame is one simple solution for this problem.

In any communication system, where there is a burst transmission, it is of utmost importance to maintain timing synchronization, since it serves as the timing reference to the data symbols that follow. In OFDM system this means figuring out the end of the preamble, which in turn gives an estimate of where the data starts. The frequency shift problem is entangled with this timing synchronization, without which the estimation of its shift is not possible. So the frequency estimation and timing synchronization algorithms go hand-in-hand.

1.2 Design Objective

The main objective of this work is to implement low complex algorithm for the synchronization of high speed OFDM transmission links using GCL based preambles and to estimate user index in the received frame at the receiver.

1.3 Thesis Organization

The rest of the thesis is organized as described below.

Chapter 2 explains the system requirements for which the proposed synchronization and user detection schemes are designed and corresponding system design proposed to full fill the requirements.

Chapter 3 describes the GCL preamble design for proposed system and filter design for preamble locations.

Chapter 4 gives the brief explanation on timing and frequency synchronization algorithm using correlation property.

Chapter 5 describes the user index detection algorithm using correlation properties of GCL sequences and discusses some results of this algorithm.

Chapter 6 Gives the conclusions from the work explained in all these chapters. Also explains the future scope of the proposed algorithms.

Chapter 2

System Description

Multi point to multi point OFDM links that have been designed had unique requirements. Having applications in defense aeronautical purposes, these links subjected to high Doppler shifts and malignant channel conditions. Bandwidth is used by number of users causing adverse effects on the data of a particular user. One effect is the interference which can lead to bad timing offset causing Inter Block Interference (IBI). Another effect is Doppler shift which can cause Inter Carrier interference (ICI). In the interference environment, major task of the receiver is to know strongest user and then start further decoding process.

The OFDM framework and receiver architecture developed, can deal with all the above impairments up to a certain degree.The following sections briefly introduce each design element of the overall system.

2.1 OFDM Frame Work

The present work is carried out to the OFDM system with the following block diagram.

2.1.1 Transmitter

User data is QPSK modulated and converted to parallel and then mapped to subcarriers according to Frequency Hopping Sequence. 1024 point IFFT is performed after subcarrier mapping and converted to serial data. Cyclic Prefix (CP) of length 256 samples is added before OFDM symbol.

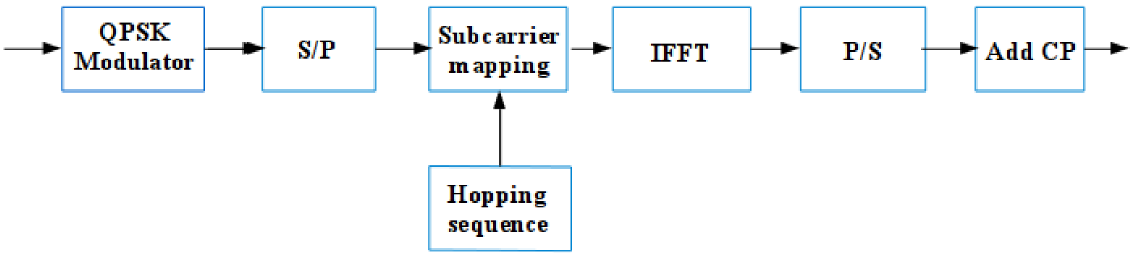


Figure 2.1: Transmitter block diagram

2.1.2 Frame Structure

In asynchronous data transmission among users, frame length should be large enough to minimize the probability of preamble overlap with other users. So, in the current design 200 OFDM symbols long frame length is used excluding preamble. Preamble structure is little different from the OFDM data symbol. Instead of using CP, preamble is copied again before preamble which acts as long CP. The advantages of this long CP is explained in detail in Chapter 3.

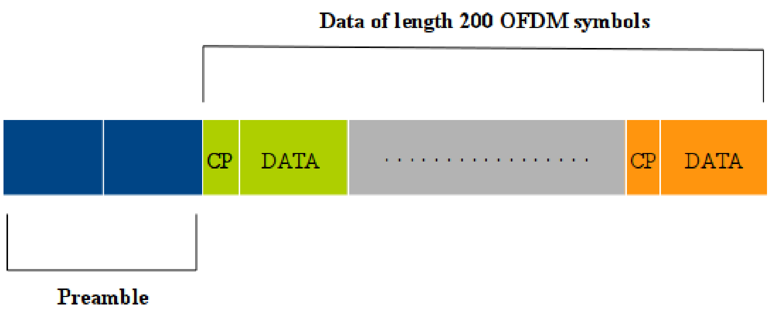


Figure 2.2: Frame Structure

Each OFDM data symbol length is 12.5 μs and preamble length is 20 μs resulting the frame length of 2.52 ms which is fairly enough to avoid interference of preamble most of the time.

2.1.3 Subcarrier Mapping

The system is built over 102.4 MHz band. The bandwidth is then divided into 1024 subcarriers, enumerated from -511 to 512. The subcarriers are further classified into several sub groups and mapped to certain type of symbols. All these further classifications are described below.

Upper and Lower bands

The 1024 subcarriers are further subdivided into two sub bands, upper and lower. The upper band (UB) consists of subcarriers 1 to 512. The lower band (LB) consists of subcarriers -511 to 1. Each user is given a certain number of subcarriers in UB and LB. In order to overcome jamming, the user sends the same data over both the bands.

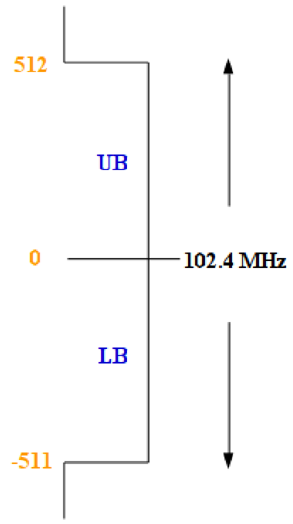


Figure 2.3: Sub bands

Subcarriers block allocation

Total Band Width is divided in to blocks of 27 subcarriers and total of 16 data blocks excluding preamble. First and last subcarrier in each block is null subcarrier. User can use this entire block for data transmission in UB and LB. Therefore, in each OFDM symbol user transmits same data in two blocks to overcome jamming and interference mitigation. 20 subcarriers above in the upper band and 20 subcarriers below in the lower band are used as guard intervals.

Pilot subcarriers

OFDM System is designed for defence aeronautical purpose, channel changes rapidly due to high velocity of vehicles. Because of these high velocities channel becomes fast fading. In the fast fading environment, it is necessary to estimate the channel in every OFDM symbol. For this, comb type pilot design is used in every OFDM symbol. Every fifth subcarrier is used as a pilot subcarrier in the current design.

Let S_f be the period of pilot tones in frequency. In order to keep track of the frequency-selective channel characteristics, the pilot symbols must be placed as frequently as coherent bandwidth is. As the coherence bandwidth is determined by an inverse of the maximum delay spread σ_{max} the pilot symbol period must satisfy the following inequality:

$$S_f \leq \frac{1}{\sigma_{max}} \tag{2.1}$$

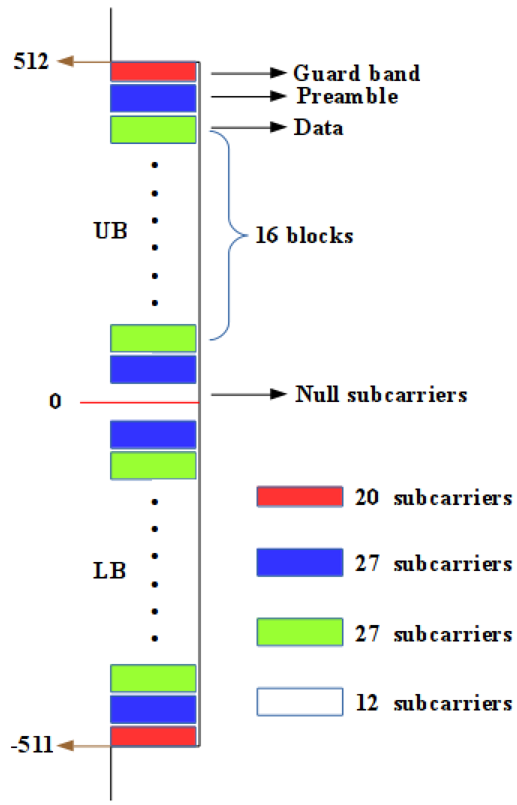


Figure 2.4: Subcarrier Blocks Division

2.2 Frequency Hopping

A special Hopping sequence is used for mapping the data to the subcarriers. In the Frequency domain data is duplicated to cater the jamming effects. As the system is designed for higher bandwidths such that jammer can occupy only an integer fraction of it's subcarrier bandwidth. Data symbols are mapped to some of the subcarriers in Upper band and Lower band simultaneously. Every frame will use different subcarrier blocks. Hopping sequence is designed such that there will be no interference for data and preamble symbol allocations.

To generate hopping sequence a long PN sequence circle is constructed and is divided in to arcs. Each arc is assigned to each user. Since, total band width is divided in to UB, LB and 16 data blocks per UB or LB, So the arc is divided in to 32 segments to allow upper band and lower band hopping. The length of each segment is of 4 bits. In each hop user gets the data block according to this 4 bit segment.

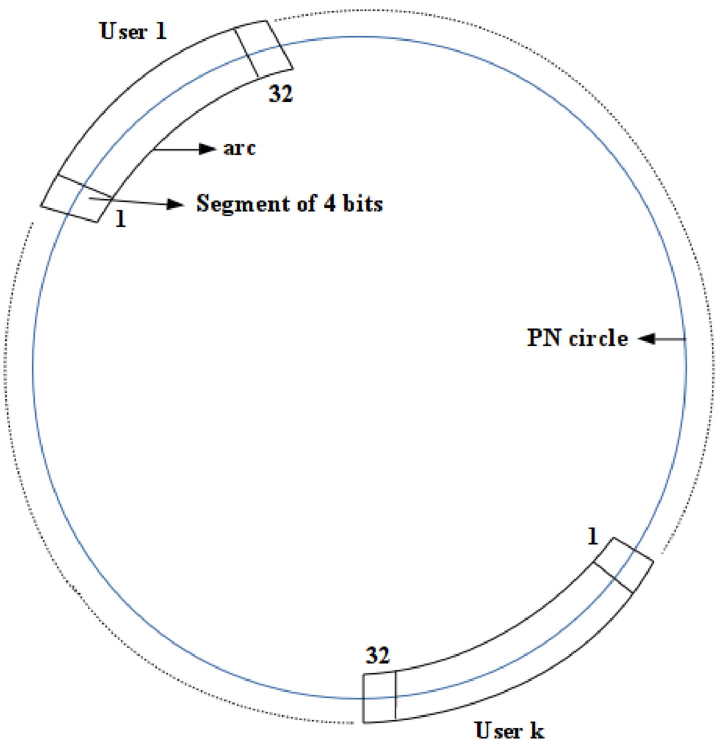


Figure 2.5: Frequency Hopping circle

2.3 Receiver Structure

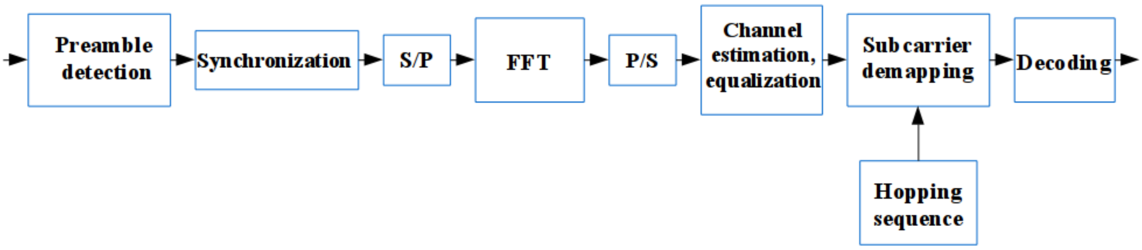


Figure 2.6: Receiver

The first task of the receiver is to find the user index of received data using preamble. If estimated user index is the intended user index then receiver starts further processing otherwise receiver drops that frame and searches for another preamble. Complete description of receiver blocks is explained in subsequent chapters.

2.4 Data Rate Calculations

Each frame has 200 OFDM data symbols and every data symbol uses 25 subcarriers. Total band width is 102.4 MHz and subcarrier spacing is 100 KHz. Total frame length is 2.52 ms. Modulation is QPSK.

$$\begin{aligned} \text{Uncoded bit rate} &= total\ number\ of\ data\ bits\ per\ frame/frame\ duration \\ &= (200 \times 25 \times 2)/(2.52ms) \\ &= 3.968Mbps \end{aligned} \tag{2.2}$$

Chapter 3

Preamble Design

3.1 Concept of Preamble in OFDM

In burst transmission where the data are transmitted in packets, the receiver needs to know the arrival time of the packet and the start of the the data symbol in order to demodulate them. The offset of determining the start of the symbol causes the received symbols experience Inter Symbol Interference (ISI). Techniques are being researched in order to estimate the time offset in the OFDM system.There are training symbol or preamble based and also cyclic prefix based. Burst transmission exploits training symbol or preamble since the preambles are available in the OFDM frame for synchronization purposes. Continuous transmission such as in the broadcasting, cyclic prefix is exploited for the estimation.

Preamble is an OFDM symbol which is specifically designated for synchronization.It also referred as Training symbol. It has some specific structure which can be used for timing and frequency offset estimations. The estimation techniques are basically exploit correlation in the training symbol or cyclic prefix in time domain since the received signal is in time domain as well. Since training symbol or preamble is actually consists of known sequence, it is good to have a sequence that has good property in the correlation itself. A sequence with good correlation property therefore will help the particular estimation. In the current work Generalized Chirp Like (GCL) sequences are used to design preamble symbol.

3.2 Introduction to GCL sequences

A specific complex sequence known as Generalized Chirp Like (GCL) sequence is a kind of poly phase sequence which has lower PAPR and better correlation property than PN sequence. GCL sequence is derived from Zadoff-Chu sequence and is given by:

$$a(k) = s(k)b(k \bmod m), \quad k = 0 \cdots N_G - 1, \quad N_G = cm^2 \tag{3.1}$$

where $b(i), i = 0 \cdots m-1$ is any sequence of m complex numbers with unit amplitude and $s(k)$ is a Zadoff-Chu sequence.

$$s_u(k) = \begin{cases} exp\left\{ -j2\pi u \dfrac{k(k+1)+qk}{2N_G} \right\} & \text{for odd } N_G \\ \\ exp\left\{ -j2\pi u \dfrac{k^2+qk}{2N_G} \right\} & \text{for even } N_G \end{cases} \tag{3.2}$$

q is any integer and $u = 1 \cdots N_G - 1$.

In order to come up a maximal number of sequences with low PAPR and good cross correlation, we choose a prime number N_G because any class index u will be relatively prime to N_G , which results in optimal cross correlation between any two sequences with different class indices. (we also choose $b = 1, q = 0$ for simplicity. Note that a prime number N_G means that in the equation 3.1 $c = N_G$ and $m = 1$.

3.2.1 Properties of GCL Sequences

$property\ 1$: The GCL sequence has constant amplitude, and its N_G-point DFT has also constant amplitude.
$property\ 2$: The GCL sequences of any length have an ideal cyclic autocorrelation (i.e., the correlation with the circularly shifted version of itself is a delta function).
$property\ 3$: The absolute value of the cyclic cross-correlation function between any two GCL sequences is constant and equal to $1/\sqrt{N_G}$, when $\mid u_1 - u_2 \mid$, u_1 and u_2 are all relatively prime to N_G (a condition that can be easily guaranteed if N_G is a prime number).
$property\ 4$: GCL sequences are circular, i.e., $s(k) = s(k \bmod N_G)$.

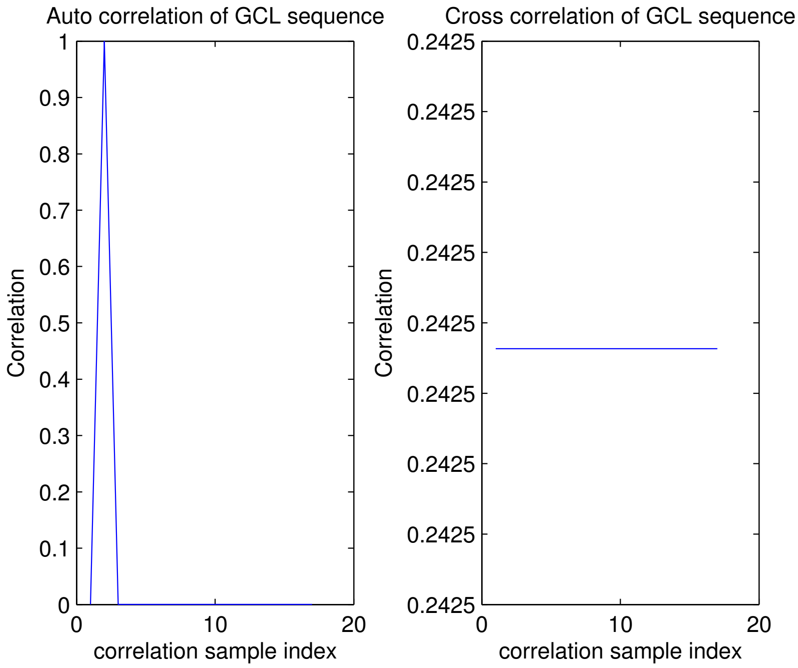


Figure 3.1: Auto and Cross Correlation of GCL Sequences of length 17

3.3 GCL Sequence for preamble in current work

GCL sequences of length 31 is used in the current design to accommodate 30 users in the system. We have 27 subcarriers block for preamble, so GCL sequences are truncated to 17 length. This 17 length truncated GCL sequence is mapped to 17 subcarriers in preamble block. These sequences are generated according to

$$s_u(k) = exp\Bigg\{ -j2\pi u\frac{k(k+1)}{2N_G}\Bigg\} \quad \text{Here } N_G = 31,\ u = 1\cdots 30 \tag{3.3}$$

3.4 Preamble Locations

Two locations in upper band and two locations in lower band are used for preamble. Out of these 4 bands user gets one upper and one in lower band. In each frame these two locations are random. Receiver need not have knowledge on these locations. By passing the received data through filters on 4 locations, receiver decides the two location where preamble actually presents. Preamble locations are indicated in Figure 3.2.

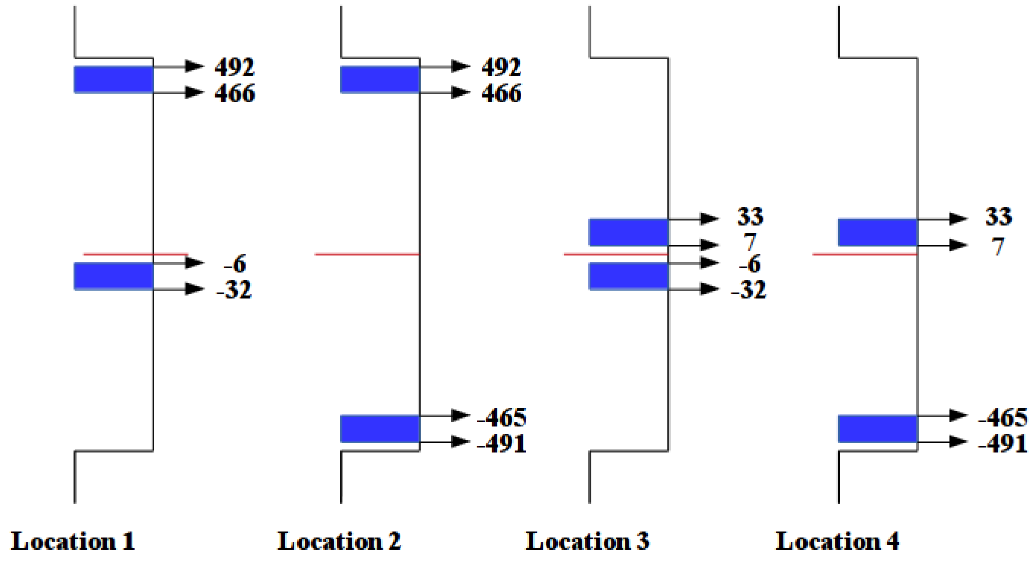


Figure 3.2: Preamble Locations

3.5 Preamble in time domain

Preamble itself does not have any repetition in time domain. To perform synchronization we need repetition in time domain. Preamble OFDM symbol is repeated again to satisfy this criterion.

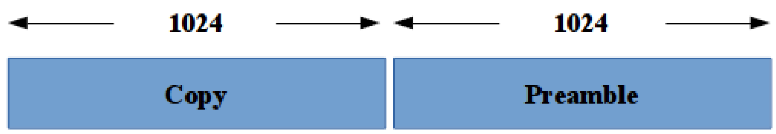


Figure 3.3: Time domain Preamble

3.6 Filter Design

In OFDMA systems total Band Width is divided and multiple small bands and assigned to users. This allows users to transmit data on different bands without interference. Receiver has to operate on particular bands to get the data and to remove out band noise. So, filters has to be implemented according to user's allocated band. In this chapter design of linear phase FIR (Finite Impulse Response) filters is proposed.

3.7 Linear Phase FIR Filter

"Linear Phase" refers to the condition where the phase response of the filter is a linear (straight-line) function of frequency (excluding phase wraps at +/- 180 degrees).This results in, the delay through the filter being the same at all frequencies. Therefore, the filter does not cause any phase distortion. The lack of phase distortion is critical advantage over IIR (Infinite Impulse Response) filters.

A FIR filter is a linear phase if (and only if) its coefficients are symmetrical around the center coefficient.
Consider a causal FIR transfer function $H(z)$ of length $N+1$, i.e., of order N:

$$H(z) = \sum_{n=0}^{N} h[n]z^{-n} \tag{3.4}$$

The above transfer function has a linear phase, if its impulse response $h[n]$ is either symmetric, i.e.,

$$h[n] = h[N-n], \quad 0 \leq n \leq N \tag{3.5}$$

or is antisymmetric, i.e.,

$$h[n] = -h[N-n], \quad 0 \leq n \leq N \tag{3.6}$$

3.7.1 Type-1 FIR Filter

For Type-1 filter $h[n] = h[N-n]$ and N is even.

In the general case for Type-1 FIR filters,the frequency response is of the form

$$H(e^{jw}) = e^{-jwN/2}\tilde{H}(jw) \tag{3.7}$$

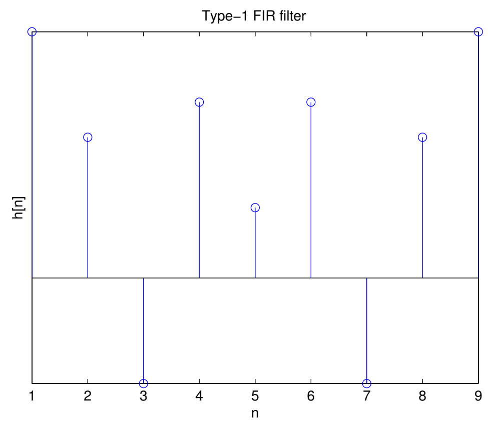


Figure 3.4: Type-1 FIR filter with N=8

where the amplitude response $\tilde{H}(jw)$, also called the zero-phase response is of the form

$$\tilde{H}(jw) = h\left[\frac{N}{2}\right] + 2\sum_{n=1}^{N/2} h\left[\frac{N}{2} - n\right]\cos(wn) \tag{3.8}$$

3.7.2 Type-2 FIR Filter

For Type-2 filter $h[n] = h[N-n]$ and N is odd.

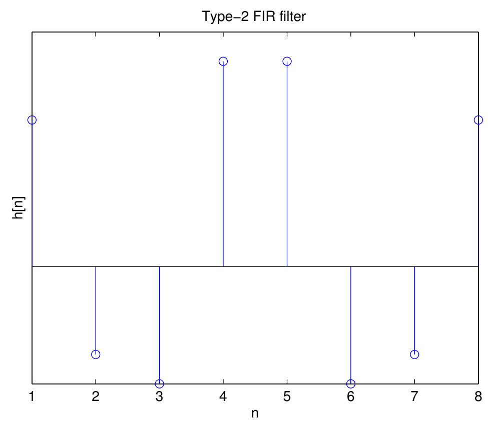


Figure 3.5: Type-2 FIR filter with N=7

In the general case for Type-2 FIR filters,the frequency response is of the form

$$H(e^{jw}) = e^{-jwN/2}\tilde{H}(jw) \tag{3.9}$$

where the amplitude response $\tilde{H}(jw)$, also called the zero-phase response is of the form

$$\tilde{H}(jw) = 2 \sum_{n=1}^{(N+1)/2} h\Big[\frac{N+1}{2} - n\Big] \cos(w(n-1/2)) \tag{3.10}$$

3.7.3 Type-3 FIR Filter

For Type-3 filter $h[n] = -h[N-n]$ and N is even.

In the general case for Type-3 FIR filters,the frequency response is of the form

$$H(e^{jw}) = e^{-jwN/2}\tilde{H}(jw) \tag{3.11}$$

where the amplitude response $\tilde{H}(jw)$, also called the zero-phase response is of the form

$$\tilde{H}(jw) = 2 \sum_{n=1}^{N/2} h\Big[\frac{N}{2} - n\Big] \sin(wn) \tag{3.12}$$

3.7.4 Type-4 FIR Filter

For Type-4 filter $h[n] = -h[N-n]$ and N is odd.

In the general case for Type-4 FIR filters,the frequency response is of the form

$$H(e^{jw}) = je^{-jwN/2}\tilde{H}(jw) \tag{3.13}$$

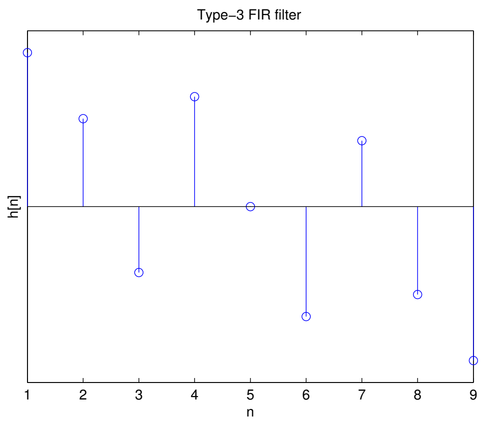


Figure 3.6: Type-3 FIR filter with N=8

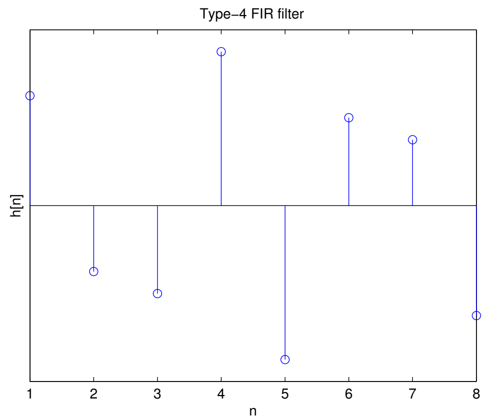


Figure 3.7: Type-4 FIR filter with N=7

where the amplitude response $\tilde{H}(jw)$, also called the zero-phase response is of the form

$$\tilde{H}(jw) = 2 \sum_{n=1}^{(N+1)/2} h\Big[\frac{N+1}{2} - n\Big] \sin(w(n - 1/2)) \tag{3.14}$$

3.8 Complex Coefficient Linear Band pass Filter

Magnitude response of real coefficient FIR filter has symmetry about π in 0 to 2π interval. In current work, we need Band Pass Filter (BPF) which does not have symmetry in 0 to 2π interval. So the idea is, first design Low Pass Filter (LPF) with desired cut-off frequency and shift the filter to the required band.

Let $h[n]$ be the real coefficient Low Pass Filter, derived then Band Pass Filter with $h[n]$ to the desired shift w_0 is

$$h_s[n] = e^{jw_0 n} h[n] \tag{3.15}$$

3.8.1 Preamble Filters

The current system is designed to have preamble on 4 locations and the base LPF Band Width is $52\pi/1024$, so filter shifts for preamble are $66\pi/1024$, $986\pi/1024$, $1066\pi/1024$, $1986\pi/1024$ respectively.

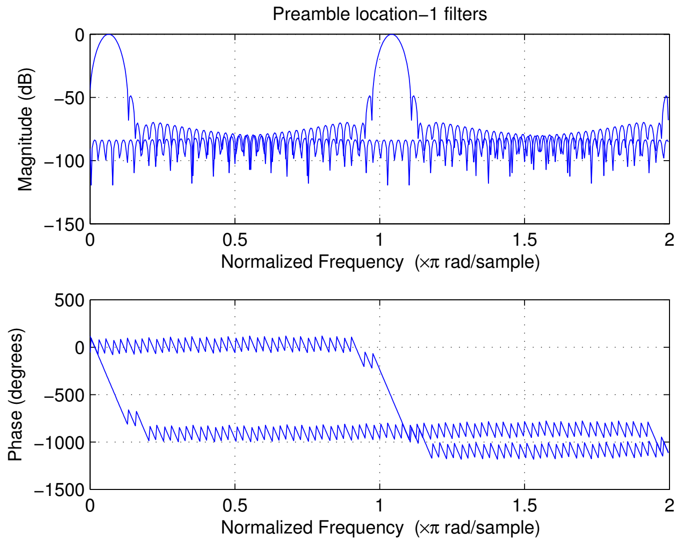


Figure 3.8: Frequency response of preamble location-1 filter

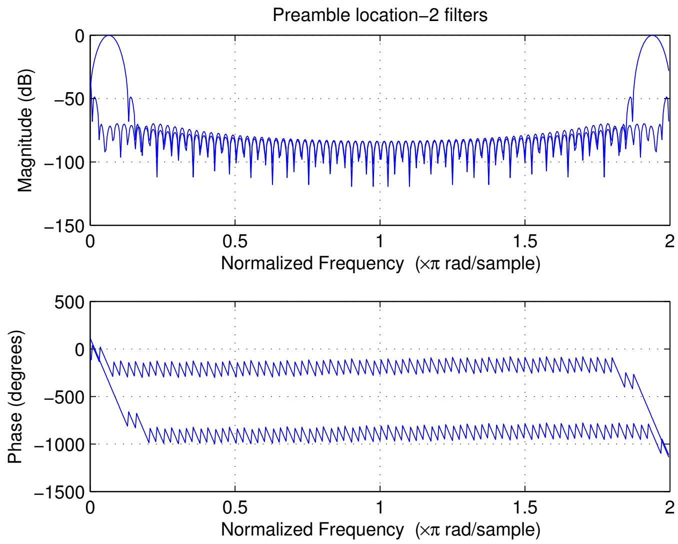


Figure 3.9: Frequency response of preamble location-2 filter

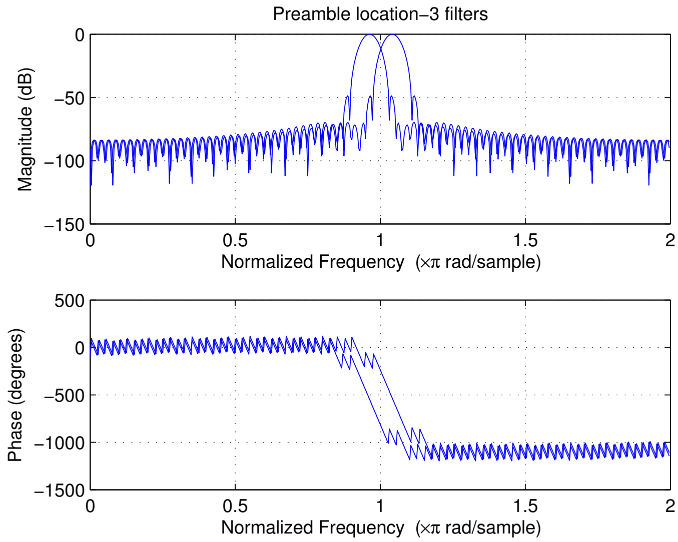


Figure 3.10: Frequency response of preamble location-3 filter

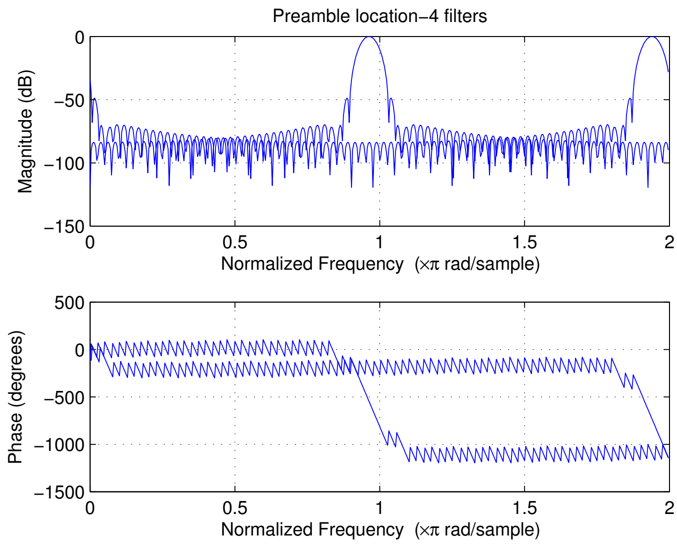


Figure 3.11: Frequency response of preamble location-4 filter

Chapter 4

Timing and Frequency Synchronization

OFDM system carries the message data on orthogonal subcarriers for parallel transmission, combating the distortion caused by the frequency selective channel or equivalently, the inter symbol interference in the multi-path fading channel. However, the advantage of the OFDM can be useful only when the orthogonality is maintained. In case the orthogonality is not sufficiently warranted by any means, its performance may be degraded due to Inter Symbol Interference (ISI) and Inter Carrier Interference (ICI).

In multipath fading channels, however, the guard interval is corrupted by ISI and the periodic property is destroyed. Consequently, correct symbol synchronization cannot be guaranteed in the case of ISI. If the symbol timing error is outside of the ISI free range in the guard interval, the inaccurate symbol timing can cause ISI that destroys the orthogonality of the subcarriers and degrades the performance of OFDM systems. In addition, the performance of the channel estimation via interpolation can be degraded by the symbol timing errors. Hence, more accurate symbol timing synchronization methods are needed to satisfy the synchronization requirement in OFDM systems.

4.1 Timing Synchronization

The symbol timing recovery relies on searching for a training symbol with two identical halves in the time domain, which will remain identical after passing through the channel, except that there will be a phase difference between them caused by the carrier frequency offset. The two halves of the training symbol are made identical (in time order) by coping GCL preamble as long CP before GCL preamble again.

Let $r[n]$ is the received signal and auto correlation is:

$$corr[m] = \sum_{n=0}^{N-1} r[n+m]r^*[n+m+N-1] \tag{4.1}$$

N is preamble OFDM symbol length.

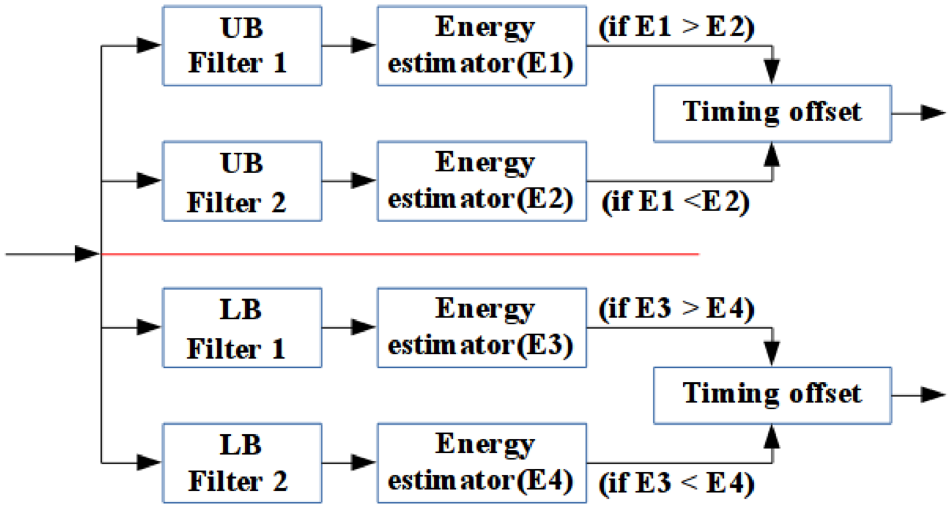
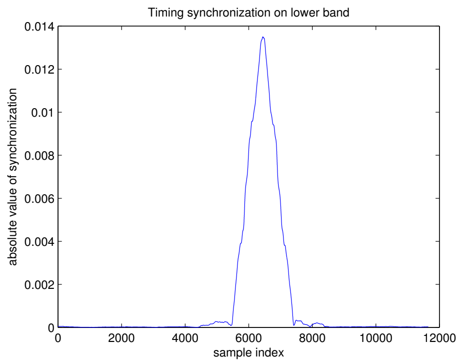
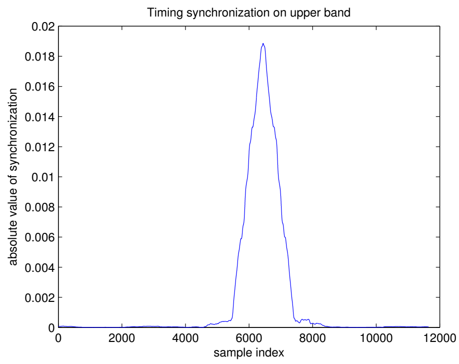


Figure 4.1: Timing offset

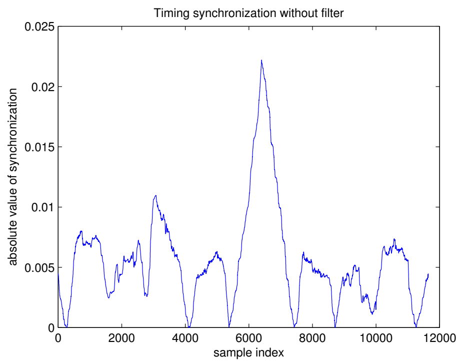
In the Figure 4.1 it is clearly indicating that energy is estimated on the filtered signal and then which ever signal energy is maximum in both bands that signal is sent for timing offset prediction. This procedure does not effect repetition in preamble symbol, since all filters are linear filters. Timing offset is improved with the use of these filters as they mitigate the effect of interferer and adjacent data symbols. Timing offset algorithm with and without filters are shown in Figure 4.2



(a) Timing offset in lower band



(b) Timing offset in upper band



(c) timing offset without filters

Figure 4.2: Timing offset with and without filters

4.2 Frequency Synchronization

The baseband transmit signal is converted up to the passband by a carrier modulation and then,converted down to the baseband by using a local carrier signal of the same carrier frequency at the receiver. In general, there are two types of distortion associated with the carrier signal. One is the phase noise due to the instability of carrier signal generators used at the transmitter and receiver, the other is the carrier frequency offset (CFO) caused by Doppler frequency shift f_d. Even if we intend to generate exactly the same carrier frequencies in the transmitter and receiver, there may be an unavoidable difference between them due to the physically inherent nature of the oscillators. Let f_c and f_c' denote the carrier frequencies in the transmitter and receiver, respectively. Let f_{offset} denote their difference (i.e.,$f_{\text{offset}} = f_c - f_c'$. Doppler frequency f_d is determined by the carrier frequency f_c and the velocity v of the receiver as

$$f_d = \frac{v f_c}{c} \tag{4.2}$$

where c is the speed of light. Let us define the normalized CFO, ε, as a ratio of the CFO to subcarrier spacing Δf, shown as

$$\varepsilon = \frac{f_{\text{offset}}}{\Delta f} \tag{4.3}$$

4.2.1 Carrier Frequency Offset Estimation

Let $r[n]$ be the signal without CFO and $y[n]$ be the signal with CFO, in the ideal case $y[n]$ related to $r[n]$ as follows:

$$y[n] = r[n] \;\; e^{\dfrac{j2\pi f_{\text{offset}} n}{N\Delta f}} \tag{4.4}$$

In the preamble:

$$\begin{aligned} y[n]y^*[n+N] &= r[n]r^*[n+N] \;\; e^{-\dfrac{j2\pi f_{\text{offset}}}{\Delta f}} \\ &= |r[n]|^2 \;\; e^{-\dfrac{j2\pi f_{\text{offset}}}{\Delta f}} \\ &= |r[n]|^2 \;\; e^{-j2\pi\varepsilon} \end{aligned} \tag{4.5}$$

Equation 4.5 clearly shows that, by finding angle of $y[n]y^*[n+N]$ we can predict CFO. i.e., CFO is the angle of max$\Big(corr[m] = \sum_{n=0}^{N-1} r[n+m]r^*[n+m+N-1]\Big)$.

Maximum Carrier Frequency Offset that can be estimated in the current design from the above algorithm is:

$$\begin{aligned} -\pi \leq 2\pi\varepsilon_{\text{est}} &\leq \pi \\ \frac{1}{2} \leq \varepsilon_{\text{est}} &\leq \frac{1}{2} \\ |f_{\text{offset}}| &\leq 50 \text{ KHz} \end{aligned} \tag{4.6}$$

4.2.2 Carrier Frequency Offset Correction

Frequency shift of ε in the frequency domain signal $Y[k] = R[k - \varepsilon]$ subjects to the CFO of ε and leads to an inter carrier interference (ICI), which means a subcarrier frequency component is affected by other subcarrier frequency components. Figure 4.3 shows that

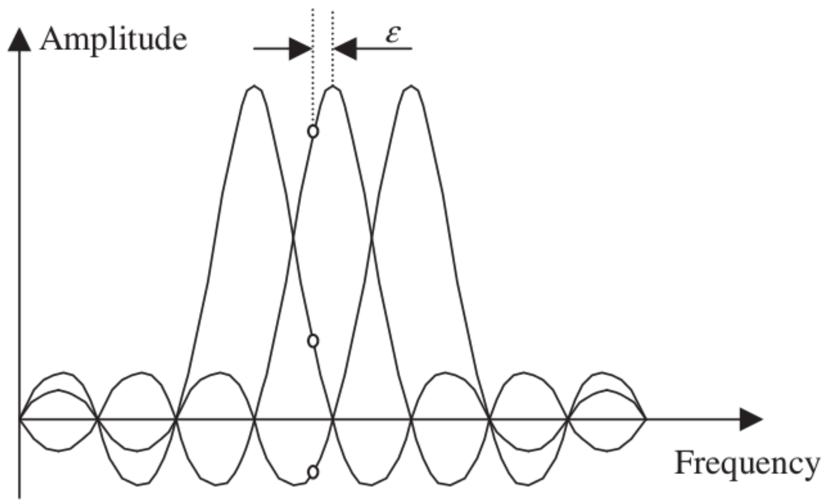


Figure 4.3: Intercarrier interference (ICI) subject to CFO.

ICI with other carriers. So to overcome this effect and user detection algorithm to work properly, time domain signal has to be multiplied with $e^{\pm j\pi\varepsilon_{\text{est}}n}$.

Chapter 5

User Detection

Multi point to multi point asynchronous transmission scenario, every user can transmit data without any knowledge about channel or other users. After getting FFT window and CFO correction receiver has to estimate user index which the data is actually intended for. So before further processing the data user detection algorithm should be implemented to get desired data. In this chapter two algorithms are proposed to find user index.

5.1 Auto Correlation Based User Detection

The user index search is to determine directly the class indices "u" from the preamble signal. The process is described as in the following.

- After a rough timing estimate has been found a block of N received time-domain data is transformed to the frequency domain using the usual FFT process.
- Denote the frequency domain data on a particular segment as $Y(m)$ for m = 1 to N_p (ignoring the unused subcarriers) and $S_G(m)$ is the GCL sequences used at those reference subcarriers, a vector of "differential based" values is then computed based on the pairs of reference subcarriers. These values are conveniently collected into vector format (a differential based vector) for efficient FFT based processing. The differential-based vector is computed as

$$Z(m) = Y(m)Y^{*}(m-1), \quad m = 1, \cdots, N_p - 1 \tag{5.1}$$

- Assuming the channel between two adjacent reference subcarriers does not change drastically, which is normally satisfied as long as the spacing of reference subcarriers is not too large, $Y(m)Y^*(m-1)$ is approximately equal to

$$\begin{aligned} Y(m)Y^*(m-1) &\approx S_G(m)S_G^*(m-1) \\ &= e^{\dfrac{-j\pi um(m+1)}{N_G}} e^{\dfrac{j\pi u(m-1)(m)}{N_G}} \\ &= e^{\dfrac{-j2\pi um}{N_G}} \\ Z(m) &= e^{\dfrac{-j2\pi um}{N_G}} \end{aligned} \tag{5.2}$$

- Thus, the user index (or sequence index) information "u" is carried in the differential-based vectors. By processing the differential based values and identifying a prominent frequency component of the vector, we can detect "u". To obtain the frequency domain components, a commonly used tool is to take an IFFT on Z(m) to get

$$\begin{aligned} IFFT(Z(m)) &= \frac{1}{N}\sum_{m=0}^{N-1} e^{\dfrac{-j2\pi um}{N_G}} e^{\dfrac{j2\pi nm}{N}} \\ &= \frac{1}{N}\sum_{m=0}^{N-1} e^{j2\pi m\left[\dfrac{n}{N} - \dfrac{u}{N_G}\right]} \end{aligned} \tag{5.3}$$

Let $p = 2n - \frac{uN}{N_G}$ then

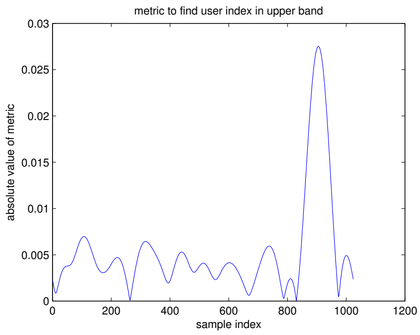
$$\begin{aligned} IFFT(Z(m)) &= \frac{1}{N}\sum_{m=0}^{N-1} e^{j2\pi m\left[\dfrac{p-n}{N}\right]} \\ z(n) &= \left[e^{j2\pi(p-n)\left(\dfrac{N-1}{N}\right)}\right]\frac{\sin\pi(n-p)}{\sin\pi(n-p)/N} \\ &= \begin{cases} 1 & \text{for n=p} \\ 0 & \text{for n} \neq \text{p} \end{cases} \end{aligned} \tag{5.4}$$

- From the equation 5.4 the peak position (say n_{max}) of $z(n)$ gives information about u, i.e., the mapping between the identified prominent frequency component at n_{max} to a corresponding transmitted sequence index is determined as:

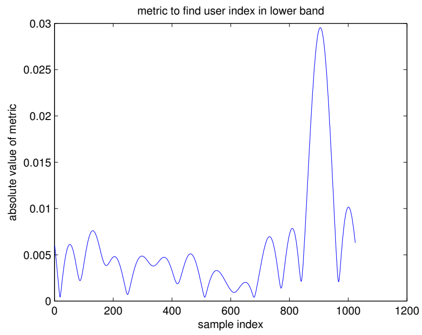
$$\begin{aligned} p &= n_{max} \\ N\left(\frac{2\text{n}_{max}}{N} - \frac{u}{N_G}\right) &= \text{n}_{max} \\ u &= \frac{\text{n}_{max}N_G}{N} \end{aligned} \tag{5.5}$$

5.1.1 Results of Auto Correlation Based User Detection Algorithm

All results are at 10 dB SNR, at different SINR and with interferer preamble overlap
Preamble with 100% overlap

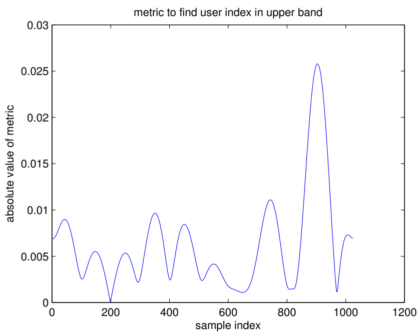


(a) Metric in upper band at 9dB SINR

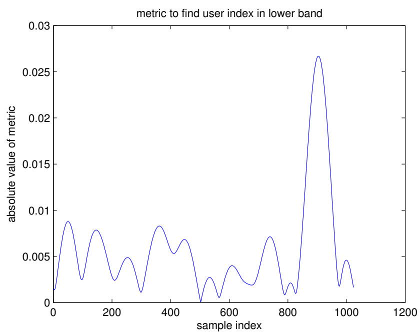


(b) Metric in lower band at 9dB SINR

Figure 5.1: Metric at 9dB SINR with full overlap of preamble

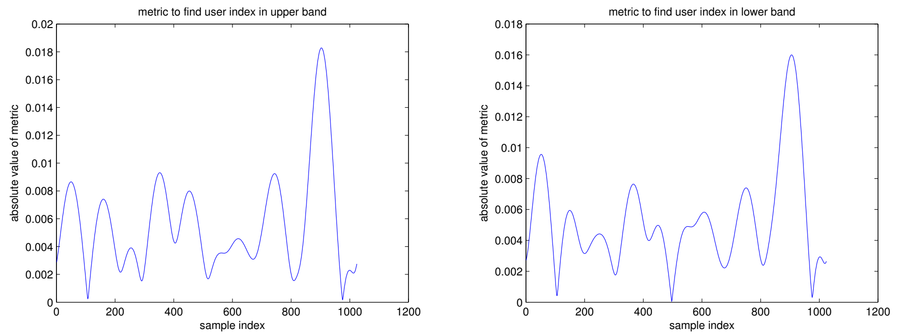


(a) Metric in upper band at 6dB SINR



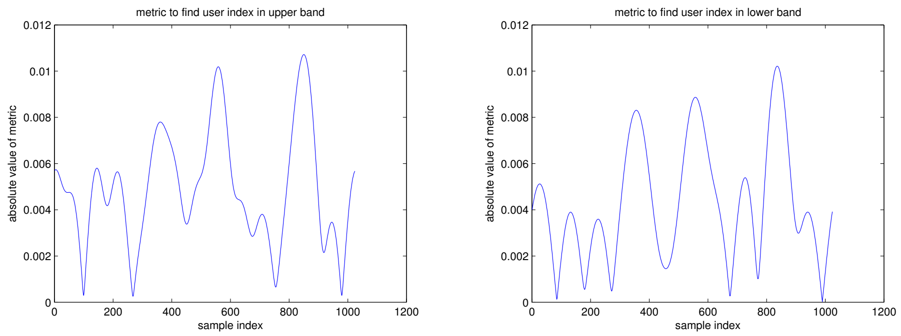
(b) Metric in lower band at 6dB SINR

Figure 5.2: Metric at 6dB SINR with full overlap of preamble



(a) Metric in upper band at 3dB SINR (b) Metric in lower band at 3dB SINR

Figure 5.3: Metric at 3dB SINR with full overlap of preamble



(a) Metric in upper band at 0dB SINR (b) Metric in lower band at 0dB SINR

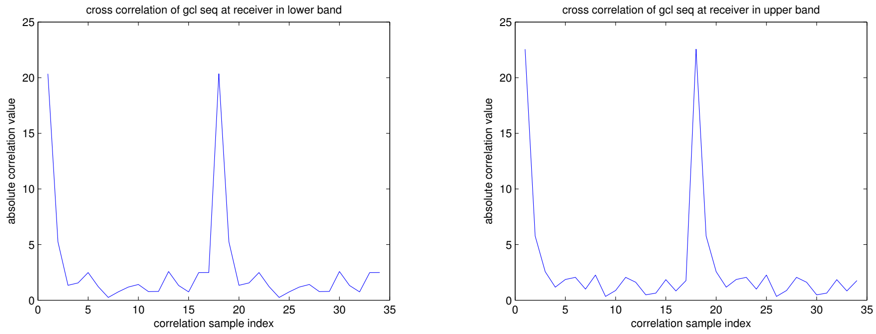
Figure 5.4: Metric at 0dB SINR with full overlap of preamble

5.2 Cross Correlation Based User Detection

In this method of user detection each receiver has its own copy of preamble which is used in its intended transmitter. Receiver correlates its preamble with transmitted preamble. This algorithm works purely based on cross correlation properties of preamble as described in chapter 3.

5.2.1 Results of cross Correlation Based User Detection Algorithm

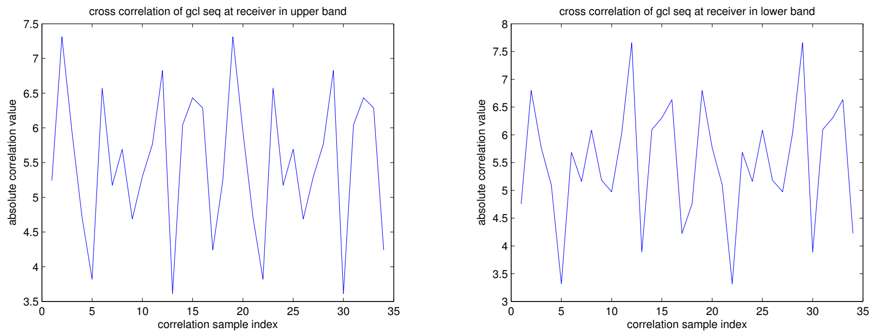
Results are at 20 dB SNR with no interferer (without normalization by N_G)



(a) Cross correlation in lower band

(b) Cross correlation in upper band

Figure 5.5: Cross correlation with desired user preamble



(a) Cross correlation in lower band

(b) Cross correlation in upper band

Figure 5.6: Cross correlation with interferer preamble

5.2.2 Disadvantage with Cross Correlation Algorithm

In the current design, time domain GCL sequence has repetition. When we run this algorithm, in the case of desired user we will be getting two peaks as indicated in Figure 5.5 and the difference between these two peak value indices is N_G. In the case of interferer also the difference between consecutive peak value indices is N_G. So there will be an ambiguity whether the received signal is intended user's signal or interferer signal.

Chapter 6

Conclusion and Future Work

6.1 Conclusion

In the thesis, I have described the OFDM system design, that was designed specifically for the unique requirements for the project at hand. The design took care about frame detection, Synchronization and user index extraction from received data. The design also ensured that hopping pattern allows minimum collision among users. A special preamble symbol in the design gave the best timing offset and carrier frequency offset estimated with high accuracy all the time. The pilots are used to estimate the channel and perform equalization.

From the simulations carried out for the proposed methods we can conclude the work as below. For the multipath channels with SINR above 3dB Preambles gives exact user index and if SINR decreases below 3dB the probability of finding user index decreases. At 0dB SINR system performance is degraded. Percentage of overlap of preamble also effects the system performance. In interference environment filtering is the better option before doing synchronization to get the sharp peak.

6.2 Future Work

Now, the proposed scheme can be extended to improve the SIR and SINR capability by using more sharp cutoff frequency filters and repetitions in preamble . And the proposed work can also be extended to Transmitter diversity and receive diversity cases. This work proposes user index estimation part, where as it can be applied to more number of users in the system and can check the user detection capability synchronization efficiency.

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