Enhanced Channel Estimator for OFDM Systems in Specular Multipath Channels

A Project Report

submitted by

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in partial fulfilment of the requirements
for the award of the degree of

MASTER OF TECHNOLOGY

DEPARTMENT OF ELECTRICAL ENGINEERING
INDIAN INSTITUTE OF TECHNOLOGY MADRAS.

May 2013

THESIS CERTIFICATE

This is to certify that the thesis titled **Enhanced Channel Estimator for OFDM Systems in Specular Multipath Channels** , submitted by **D Kiran**, 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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Date: 7th May 2013

ACKNOWLEDGEMENTS

I would like to take this opportunity to thank my parents for their support and encouragement without which learning in and becoming a part of such a prestigious institution would not have been possible. I would like to dedicate this work to them.

I am ever thankful to Dr. Srikrishna Bhashyam for his continual support and invaluable advice throughout the duration of my project. I feel honoured and encouraged to have worked under his guidance. I would also like to take this opportunity to thank my professors at IIT Madras, Prof. Arun Pachai Kannu, Prof. David Koilpillai, Prof. R. Aravind, Prof. K. Giridhar, Prof. Venkatesh Ramaiyan who have imparted knowledge and have motivated me to learn the intricacies of the subjects.

Last but not the least, I would like to thank my friends and labmates Hussam, Prathyusha, Vinay, Srikanth, Kiran kumar and some others about whom I have not mentioned here for their help and encouragement. I had fun filled sessions in the lab with my friends and enjoyed through and through my stay here.

ABSTRACT

KEYWORDS: Wireless communications ; OFDM; channel estimation ; pilot pattern ; specular multipath channel; Interpolation

This Thesis emphasized on improved channel estimation techniques for orthogonal frequency division multiplexing (OFDM) based wireless communication systems. The objective of this thesis is to estimate and exploit multipath delay information for OFDM channel estimation. We are exploiting the fact that channel multipath delays are slow varying parameters unlike multipath gains for estimating the channel. Estimating multipath delays i.e sparsity information of the channel helps in reducing the signal space dimension of the channel estimator and improves the performance of the channel estimator.

First we introduced OFDM technology,We explained its operation in signal domain and vector domain. We explained how this OFDM technique divides the frequency selective channel to parallel flat fading channel. We explained about the significance of this OFDM model. We studied channel estimation algorithms for OFDM model for sample-spaced channel (multipath delays are at the sampling instants). By using GAIC criterion we estimated multipath delays for sampled channel and then by exploiting these multipath delays we estimated the channel. The performance of these techniques are demonstrated by computer simulations

However, when the channel is non-sample-spaced these algorithms results in irreducible error floor due to interpolation errors. So, by using ESPRIT technique we estimated multipath delays for non sampled channel and then by exploiting these multipath delays we estimated the channel. In this technique we will derive shift invariance subspaces by sending equispaced pilots. From these two invariance subspaces, we will find rotational matrix. From rotational matrix we will find multipath delays.

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ABBREVIATIONS

BER	Bit Error Rate
CCI	Co-Channel Interference
CIR	Channel Impulse Response
CP	Cyclic Prefix
CSI	Channel State Information
DOA	Direction Of Arrival
DFT	Discrete Fourier Transform
ESPRIT	Estimation of Signal Parameters via Rotational Invariance Techniques
FFT	Fast Fourier Transform
ICI	Inter Carrier Interference
GAIC	Generalized Akaike Information Criterion
IDFT	Inverse Discrete Fourier Transform
IFFT	Inverse Fast Fourier Transform
LS	Least Squares
MIMO	Multiple Input Multiple Output
MSE	Mean Squared Error
MMSE	Minimum Mean Squared Error
MLS	Modified Least Squares
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
PDP	Power Delay Profile
BPSK	Binary Phase Shift Keying
SNR	Signal to Noise Ratio
SIR	Signal to Interference Ratio

NOTATION

Bold face small letters denote vector or matrix

$\mathbf{x}^T$	Transpose of column vector $\mathbf{x}$
X^+	Pseudo inverse of matrix X
x^*	Complex conjugate of element x
$arg(x)$	Phase angle of x in the range $[0, 2\pi)$
I_k	$K \times K$ identity matrix
$0_{p\times q}$	$p \times q$ matrix with zero entries
$diag(X)$	Diagonal matrix with elements of the vector X on its main diagonal
x̂	Estimated value of the variable x
Cov(x,y)	Covariance between $\mathbf{x}$ and $\mathbf{y}$
$\mathbf{F}$	Fourier matrix
$\bar{\mathbf{F}}$	The matrix derived from $\mathbf{F}$ with rows corresponding to pilot positions
$\bar{\mathbf{X}}$	Transmitted signal on pilot sub-carriers
$\bar{\mathbf{Y}}$	Received signal on pilot sub-carriers
$\hat{\bar{\mathbf{H}}}$	Channel estimate on pilot sub-carriers
$\hat{\mathbf{H}}$	Channel estimate on all sub-carriers
$\hat{\mathbf{h}}$	Time domain channel estimates
σ^2	Noise variance
$\mathbf{R_h}$	Power distribution matrix or channel covariance matrix

CHAPTER 1

Introduction

1.1 Channel characteristics

The characteristic of the mobile wireless channel is the variations of the channel strength over time and over frequency. The variations can be roughly divided into two types Large-scale fading, due to path loss of signal as a function of distance and shadowing by large objects such as buildings and hills. Small-scale fading, due to the constructive and destructive interference of the multiple signal paths between the transmitter and receiver. These multipath effects can be modeled as a linear time-varying system with impulse response having L taps as proposed in (Tse and Viswanath, 2005). Each tap value (h_τ) is constituted by path gains of all the paths in which the signal suffers a delay τ.

$$h(\tau, t) = \sum_i a_i(t)\delta(t - \tau) \tag{1.1}$$

Time response of the wireless channel is characterized by the coherence time. The coherence time Tc of a wireless channel is defined as the interval over which $h_l[m]$ changes significantly as a function of m. When the different paths contributing to the l^{th} tap have different doppler shifts, the magnitude of $h_l[m]$ changes significantly. This is happening at the time-scale inversely proportional to the largest difference between the doppler shifts, the doppler spread Ds. Coherence time is inversely proportional to doppler spread and is given by

$$T_c = \frac{1}{4D_s} \tag{1.2}$$

Frequency response of the wireless channel is characterized by the coherence bandwidth. The delay spread(T_d) of the channel dictates its frequency coherence. Wireless channels change both in time and frequency. The time coherence shows us how quickly the channel changes in time, and similarly, the frequency coherence shows how quickly it changes in frequency. So coherence bandwidth is given by

$$W_c = \frac{1}{2T_d} \tag{1.3}$$

Channel is classified into two, based on its time response and application. We will call a channel fast fading, if the coherence time T_c is much shorter than the delay requirement of the application, and slow fading if T_c is longer. Here delay requirement of the application means the maximum duration over which the symbols can be coded. For example for voice application symbols should be coded with in a small duration otherwise it will be ugly to hear i.e delay requirement is less for voice application. Therefore this classification is depends on both channel and application.

Channel is classified into two based on its frequency response. We call a channel a frequency selective fading, if coherence bandwidth(W_c) is less than the input signal bandwidth otherwise channel is called flat-fading channel. We expect channel to be flat fading because it is free from ISI. The frequency selective channel can be converted to K flat fading channel through OFDM technique.

1.2 Need for channel estimation

The maximum rate of transmission over wireless channel is limited by its ergodic capacity. The demand for the wireless services increased, so the radio frequency bands are precious. So the need for spectrally efficient signalling schemes and reuse-1 cellular deployment to serve more users are imminent. The performance of wireless communication systems employing coherent receiver is superior to non-coherent receiver and differential techniques (Proakis and Manolakis, 1995). If we know channel state information (CSI) at transmitter, it helps in improving the spectral efficiency of the system by adapting adaptive modulation techniques like, power allocation(by water filling method) and changing the signal constellation are varied on each sub-carrier depending on the CSI.

To achieve the full capability of MIMO technology the CSI at transmitter and receiver is important (R.Nabar, 2003). To avoid the system performance degradation, due to channel estimation errors the mean squared error (MSE) of channel estimation should be negligible compared to the operating SNR. Therefore we need a efficient channel estimation algorithm is required at receiver.

1.3 Channel estimation methods

Channel estimation methods are mainly classified into two types namely pilot based and blind.

1.3.1 Pilot symbol assisted channel estimation (PACE)

In PACE, known symbols called pilots are transmitted over time-frequency grid to estimate the channel variation . Pilots can be sent in any arrangement by following two rules. The pilot separation in time (P_t) in-terms of symbols should be less than coherence time and the pilot separation in frequency (P_f) should be less than coherence bandwidth.

$$P_t \leq \frac{1}{2f_dT_s} \tag{1.4}$$

and

$$P_f \leq \frac{KT}{\tau_{max}} \tag{1.5}$$

where

τ_{max} is delay spread and f_d is doppler spread.

In block type pilots are sent on all sub-carriers $\forall N_t$ duration. In pilots are sent on few carriers, channel estimates on other sub-carriers are estimated by interpolating the pilot channel estimates.

CHAPTER 2

OFDM Model

2.1 Introduction

To meet the increasing demand of wireless services, the radio frequency bands are precious, so we have to use spectrally efficient signaling schemes and reuse-1 cellular deployment to serve more users. We are able to transmit at high data rates even in frequency selective channel by using multi carrier modulation techniques (such as OFDM). So they received considerable remark in the last few years. The multi-carrier transmission operates on the fact that free running complex exponentials forms the eigenfunctions for linear time invariant (LTI) systems. However, for finite duration complex exponentials one way to maintain the same relation is by adding cyclic prefix (CP). By introducing a guard interval (CP) between adjacent OFDM symbols, we are able to convert the linear convolution of the signal and the channel into circular convolution. So that we can implement the DFT using FFT algorithms. Note that the CP length should be greater than the length of the channel impulse response.

The main advantages of OFDM are low complexity equalization and the use of adaptive modulation methods. Since OFDM converts the frequency selective channel to K number of flat fading sub channels, so single tap equalizers are sufficient at receiver. If we know Channel state information (CSI), We can use adaptive modulation techniques such as power allocation (by water filling method),multiuser preceding to enhance the spectral efficiency of the link. Since OFDM has above advantages, it is used in the following systems.

OFDM has been adopted as a modulation scheme for next generation wire-line (ADSL-asymmetric digital subscriber line, VDSL-very high speed DSL) and wireless communication systems (802.11a, 802.16d/e, LTE).

2.2 Signal analysis of OFDM system

Consider a OFDM system operating at bandwidth (B)=$\frac{1}{T}Hz$ where T is the sampling time and having K sub-carriers of which K_u are useful sub-carriers (excluding guard bands and DC sub-carrier). The operation of OFDM in signal domain can visualize from the diagram shown in Figure 2.1 and is explained below

2.2.1 Operation

Transmitter side

1. First we collect K symbols from input each of duration(T) to form a block, so each block duration is $T_{sym} = KT$. At this stage the signal occupies a bandwidth of $B = \frac{1}{T} = \frac{K}{T_{sym}}$.

2. Now expand the each symbol of duration(T) to a duration of $T_{sym} = KT$ and then modulated to a frequency, which is orthogonal to other modulated symbol frequency. (i_{th} symbols are modulated to frequency$f_i = \frac{i}{T_{sym}}$).
   Frequency domain explanation: Since the symbol time duration is expanded to duration of $T_{sym} = KT$, So in frequency domain it will be compressed by K times and shifted to the modulated frequency as shown in Figure **??**.
   This entire step can be implemented by calculating the inverse Fourier transform (IFFT) of the each block. Therefore IFFT is working as a modulator for OFDM system.

3. The IFFT generates a block of K symbols for, a block of K input symbols.

4. We form a OFDM symbol by adding cyclic prefix(L_{cp} symbols)) to the each block. So the resulting block length is $T_s = (K + L_{cp})T$ and transmitted over the channel.
   To overcome the Inter block interference(IBI) the cyclic prefix length should be greater than the channel delay spread (the maximum delay offered by the channel)
   By adding cyclic prefix we are converting the linear convolution of the signal and

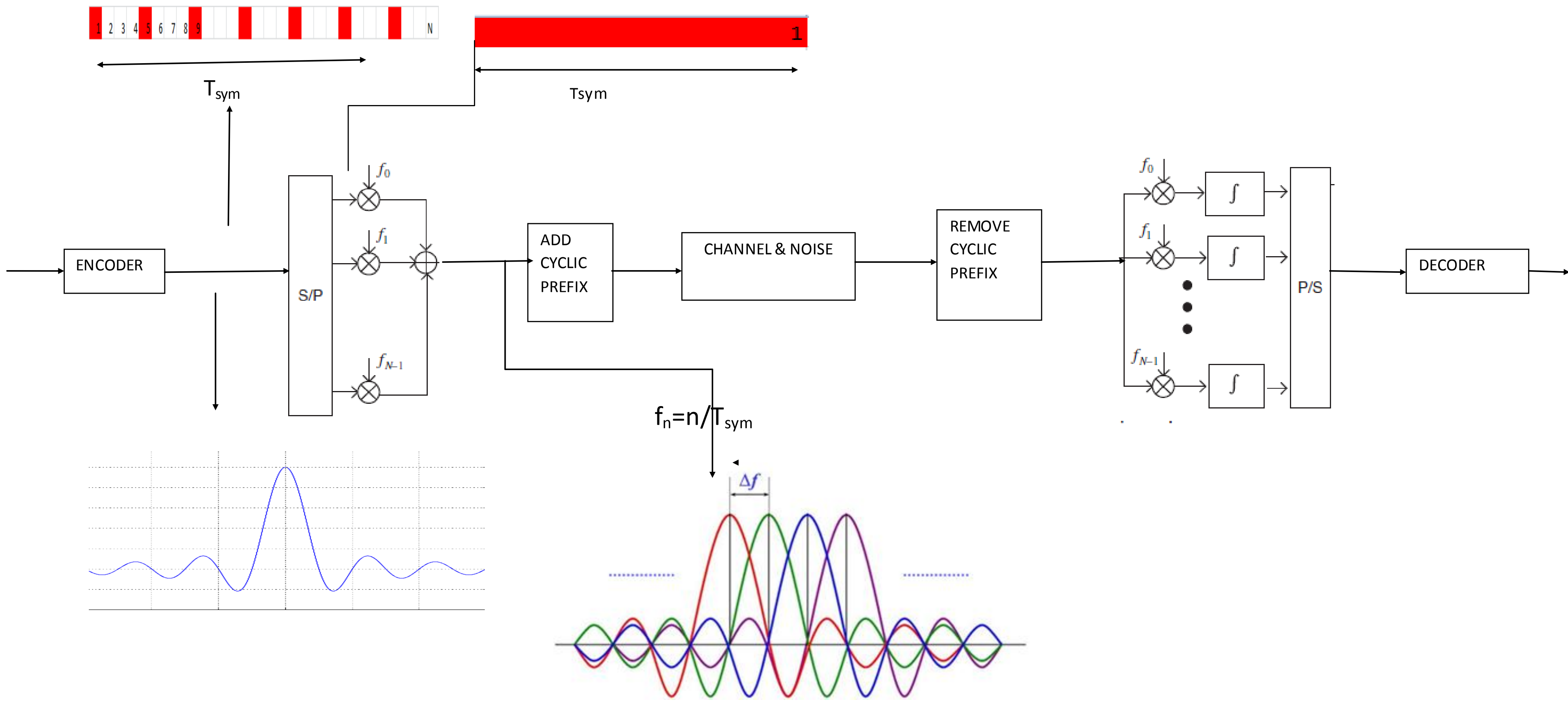

Figure 2.1: OFDM operation

channel into circular convolution. Which enables the modulation can be implemented by IFFT operation and demodulation can be implemented by FFT operation.

5. we will send these OFDM blocks (each block of length$(K + L_{cp})T$) into the channel one after other continuously.

Receiver side

At Receiver side we have to do exact reverse operations as that of transmitter side.

1. At the output of channel we will receive a sequence of blocks (each block of length $(K + L_{cp})T$).
2. Assume the channel has delay spread LT.
3. Removing cyclic prefix:
   In each block, the last L samples of the cyclic prefix part of each block are contaminated by previous block due to multipath channel. So take the first K samples out of $(K + L_{cp})$ samples as shown in Figure 2.2.
4. Demodulation: Now we have K symbols after removing cyclic prefix.now do demodulation of each symbol.
   Frequency domain explanation: Each symbol spectrum is shifted back from the modulated frequencies.
   This entire step can be implemented by FFT operation of K symbols. Therefore demodulation in OFDM systems is implemented by FFT operation as shown block diagram.
5. Now we are able to get K symbols which were sent at transmitter side.

2.2.2 Significance of cyclic prefix and its length determination

Cyclic prefix (CP) overcomes inter block interference (IBI) as clearly shown in Figure 2.2 (the red symbols are contaminated by previous block due to multipath channel). From the Figure 2.2 we can visualize that to overcome IBI the cyclic prefix length should be greater than channel memory or channel delay spread.

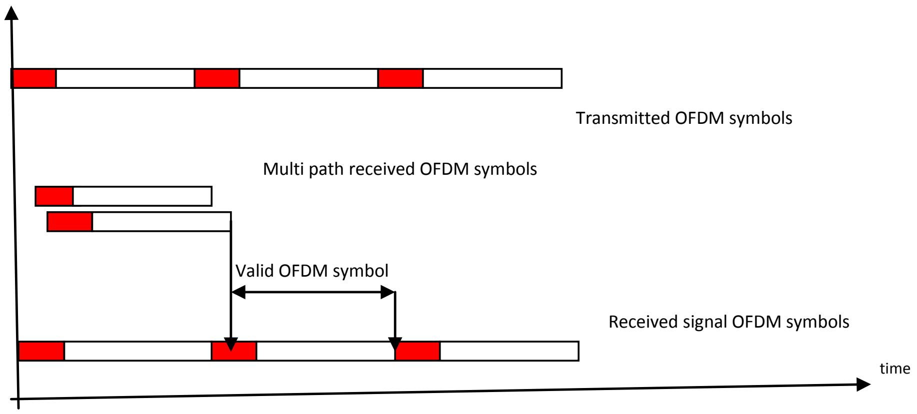

Figure 2.2: Cyclic prefix length determination

By adding cyclic prefix we are converting the linear convolution of the signal and channel into circular convolution. Which enables the modulation can be implemented by IFFT operation and demodulation can be implemented by FFT operation. So computation complexity reduces at transmitter and receiver.

2.2.3 How OFDM converts the frequency selective channel into K flat fading channels?

OFDM converts the frequency selective channel into K flat fading channels. So single tap equalizers are sufficient at receiver and mathematically explained below.

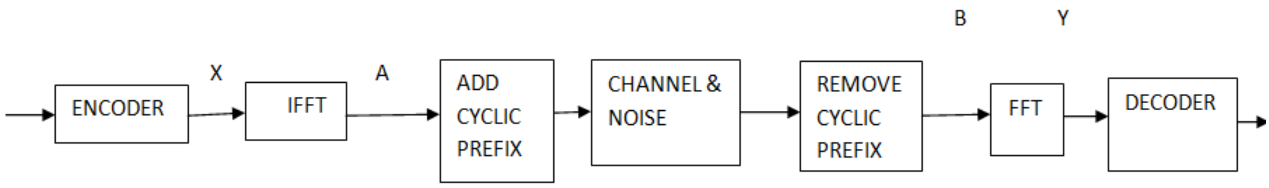

Figure 2.3: OFDM block diagram

$$B = H_c A \quad (2.1)$$
$$= (F^H D F) A \quad (2.2)$$
$$= (F^H D F)(F^H X) \quad (2.3)$$
$$= F^H D X \quad (2.4)$$
$$Y = F B \quad (2.5)$$
$$= F(F^H D X) \quad (2.6)$$
$$= D X \quad (2.7)$$

where

- The diagonal elements of the diagonal matrix (D) are N-point DFT of the channel impulse response. Therefore the OFDM system can be modeled as

$$Y = XH + V \quad (2.8)$$
$$= XFh + V \quad (2.9)$$

2.2.4 OFDM special feature

In OFDM system, modulation and demodulation sections of the communication system are implemented by IFFT and FFT operations respectively,(Weinstein and Ebert, 1971) . So it reduces the equipment complexity at transmitter and receiver.

2.2.5 Why OFDM spectrally efficient scheme?

In normal FDM systems, we should leave guard band between every user because practical digital filters will not work ideally (they don't have sharp cutoff), so to avoid interference from other user we should allow guard band.
But in OFDM systems, even by allowing the overlap between sub-carriers we are able to extract the symbols at receiver, by maintaining the orthogonality between sub-carriers.

2.2.6 Orthogonality in OFDM

Consider the time limited complex exponential signals $\{e^{-j2\pi f_k t}\}_{k=1}^{k=(K-1)}$, which represent the different sub-carriers at $f_k = \frac{k}{Tsym}$ in the OFDM signal, where $0 \leq t \leq T_{sym}$. These signals are orthogonal signals because the integral products for their fundamental period is zero,i.e

$$\begin{aligned}
\frac{1}{T_{sym}} \int_0^{T_{sym}} e^{j2\pi f_k t} e^{-j2\pi f_i t} \mathrm{d}t &= \frac{1}{T_{sym}} \int_0^{T_{sym}} e^{j2\pi \frac{k}{T_{sym}} t} e^{-j2\pi \frac{i}{T_{sym}} t} \mathrm{d}t \\
&= \frac{1}{T_{sym}} \int_0^{T_{sym}} e^{j2\pi \frac{(k-i)}{T_{sym}} t} \mathrm{d}t \\
&= \begin{cases} 1 & \forall \text{integer } k = i, \\ 0 & \text{otherwise} , \end{cases}
\end{aligned}$$

It shows that these complex signals are orthogonal to each other and having constant energy.

At transmitter :

First, we form a block of K input symbols and each symbol is spread over T_{sym}.

These K input symbols are modulated by the above K different complex signals, i.e K symbols are modulated to different orthogonal frequencies given by $f_k = \frac{k}{Tsym}$,k=1,2....(K-1) as shown if Figure **??**. Now these K orthogonal modulated signals are added in time domain over T_{sym}, to form a OFDM signal.

Since practical channel response is not a impulse, so it offers ISI and it effects the orthogonality. So, to maintain orthogonality,we will add cyclic prefix and then we will send into channel.

At receiver :

Remove the cyclic prefix from the received signal. and then demodulate. Then take the discrete samples with the sampling instances at $t = nT_s = \frac{nTsym}{K}, n = 0, 1, 2.....(K - 1)$..

If we sample the received signal at time $t = nT_s = \frac{nTsym}{K}, n = 0, 1, 2.....(K - 1)$., then we will get ISI free symbols and mathematically explained below

$$
\begin{aligned}
\frac{1}{K} e^{j2\pi f_k (nT_s)} e^{-j2\pi f_i (nT_s)} \mathrm{d}t &= \frac{1}{K} \sum_{n=0}^{(K-1)} e^{j2\pi \frac{k}{T_{sym}}(nT_s)} e^{-j2\pi \frac{i}{T_{sym}}(nT_s)} \\
&= \frac{1}{K} \sum_{n=0}^{(K-1)} e^{j2\pi \frac{(k-i)}{KT_s} nT_s} \\
&= \begin{cases} 1 & \forall \text{integer } k = i, \\ 0 & \text{otherwise }, \end{cases}
\end{aligned}
$$

Therefore the above orthogonality is an essential condition for the OFDM signal to be ISI free.

2.3 Vector analysis of OFDM system

For large K (as number of carriers increases) analysis of the OFDM system in vector domain makes simple and easy. In this thesis we made analysis in vector domain for channel estimation problem.

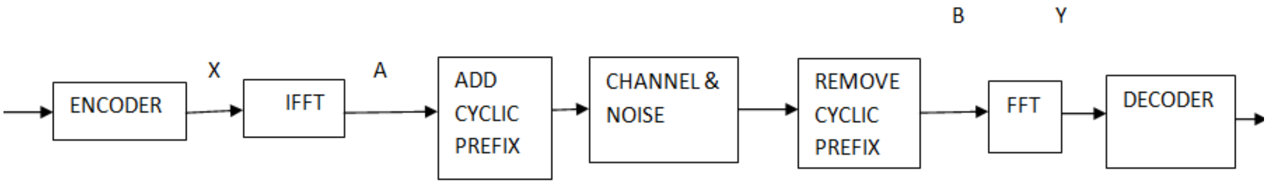

Figure 2.4: OFDM block diagram

2.3.1 Operation

Transmitter operation

1. The input block of size (K) is modulated by using IFFT operation as shown in Figure 2.4. Therefore the output of modulator (IFFT) in matrix notation is given as

   $\mathbf{A} = \mathbf{F^H X}$

$$\begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_K \end{bmatrix} = \begin{bmatrix} W_{11}^* & W_{12}^* & \dots & W_{1K}^* \\ W_{21}^* & W_{22}^* & \dots & W_{2K}^* \\ \vdots & \vdots & \ddots & \vdots \\ W_{K1}^* & W_{K2}^* & \dots & W_{KK}^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ \vdots \\ x_K \end{bmatrix}$$

   where

   $W_{mn} = e^{\frac{-j2\pi mn}{K}}$

2. Significance of matrix (F^H):Each column is orthogonal to the remaining $(K-1)$ other columns. Therefore each column acts as a orthogonal basis. Each input symbol is transmitted on each basis. i.e first symbol is transmitted on first column, second symbol is transmitted on second column, like that K^{th} symbol is

transmitted in the K^{th} column. In this way we form a K dimensional vector$\mathbf{A}$ which is present in the column space of matrix $\mathbf{F_H}$. This entire step can be implemented by IFFT operation.

3. After adding cyclic prefix to this vector $\mathbf{A}$, we will send the resultant vector into the channel. Cyclic prefix is added because to avoid Inter block interference and to make the channel into K flat fading sub channels.

Receiver operation

1. After removing cyclic prefix we will get K dimensional vector ($\mathbf{B}$) as shown in figure 2.6.

2. Symbols demodulation: By projecting this vector ($\mathbf{B}$) on to the orthogonal basis (columns of matrix (F^H) or rows of matrix (F)), we will get the required symbols. It means if you want i^{th} symbol then project the vector ($\mathbf{B}$) on to the i^{th} orthogonal basis i.e (i^{th} row of matrix (F). This entire step can be implemented by using FFT operation.

Analysis of the OFDM system in vector domain is very simple and easy, for any value of K. We can manipulate the vectors in simple and understandable way.
For example in channel estimation problem, the output corresponding to the pilots, are obtained by projecting the vector ($\mathbf{B}$) on to corresponding orthogonal basis, instead of projecting on to all other basis.

2.4 Significance of OFDM

Advantages of OFDM

- Makes efficient use of the spectrum by allowing overlap.
- It is resistant to frequency selective channel because it makes frequency selective channel into K flat fading sub channels. So single tap equalizers are sufficient.
- Therefore channel equalization becomes simple.
- Eliminates inter symbol interference(ISI) and inter block interference(IBI) through use of cyclic prefix.
- OFDM is computationally efficient by using IFFT and FFT operations for modulation and demodulations functions.

Disadvantage of OFDM

- The OFDM signal has a noise like amplitude with a very large dynamic range, therefore it requires RF power amplifiers with a high peak to average ratio. However, a number of data preprocessing methods can be employed to reduce the effect of PAPR as given in (Litsyn and Shpunt, 2007), (Slimane, 2007), (Mobasher and Khandani, 2006).

In spite of this disadvantage, OFDM forms as one of the promising modulation techniques for next generation wireless systems for its simplicity and flexibility.

CHAPTER 3

Algorithms to estimate the channel for OFDM model for sampled spaced channel

We studied pilot based channel estimation algorithms for sampled spaced channels as given in (Ragavendra, 2007). sampled channel means the multi path delays τ's are the multiple integral of input sampling time.

We will send K_p number of equispaced pilot sub-carriers and data on other sub-carriers .

Problem statement:At receiver, by using the symbols on pilot sub-carriers (red colored as shown in Figure 3.1.), we have to estimate the channel on all sub-carriers.

We studied the following three algorithms to solve the above problem.

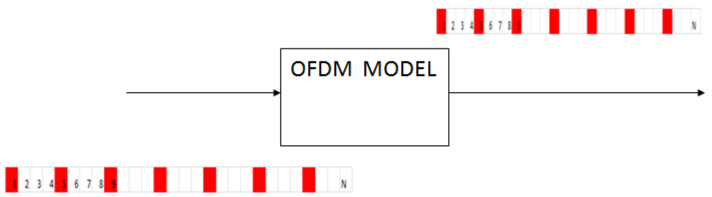

Figure 3.1: OFDM block diagram

The output of the OFDM system can be written as

$$\bar{Y} = \bar{X}\bar{H} + \bar{V} \tag{3.1}$$

$$= \bar{X}\bar{F}h + \bar{V} \tag{3.2}$$

where

$\bar{X}$, be the symbol vector send on pilot sub-carriers

$\bar{Y}$ be the symbol vector received on pilot sub-carriers.

$\hat{\bar{H}}$ be the channel estimate on pilot sub-carriers.

3.1 Modified Least Squares(MLS) Estimation

ML estimation:Maximum likelihood estimator for h is $\hat{h}_{ML}$ which maximizes the probability density function of received vector given pilot symbols.

$$\hat{h}_{ML} = \arg\max_{h} f(\bar{\mathbf{Y}}/\bar{\mathbf{X}}) \tag{3.3}$$

If noise is AWGN then it reduces to below optimization problem,

$$\hat{h}_{ML} = \arg\min_{h}((\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{F}}\mathbf{h})^H(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{F}}\mathbf{h})) \tag{3.4}$$

Estimation of channel frequency response

1. If Noise is considered as AWGN then ML estimation of $\mathbf{H}$ becomes:

$$\hat{\bar{H}}_{ML} = \arg\min_{\bar{H}}((\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{H}})^H(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{H}})) \tag{3.5}$$

   we have to find $\hat{\mathbf{H}}$ such that which minimizes the error $(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{H}})^H(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{H}})$ in worst scenario. This is also called least squares(LS) estimation

2. The solution to the above step is $\bar{\mathbf{H}} = \bar{\mathbf{X}}^{-1}\bar{\mathbf{Y}}$. Up to now, we estimated channel on pilot sub-carriers only.

3. We can estimate channel on other data sub-carriers also by interpolation as given below

$$\hat{h}_{mls} = (\bar{F}_{cp}^H\bar{F}_{cp})^{-1}\bar{F}_{cp}^H\hat{\bar{H}} \tag{3.6}$$

$$\hat{H}_{mls} = F\hat{h}_{mls,} \tag{3.7}$$

   This technique of estimating the channel on data sub-carriers by interpolating the pilot channel estimates is called modified least squares(mls).

where

- $\hat{h}_{mls}$ is the $L_{cp} \times 1$ vector,it is an estimate of impulse response of channel.
  L_{cp} is the cyclic prefix length.

- $\bar{F}_{cp}$ is derived from K-point DFT matrix F, with first Lcp columns and rows corresponding to pilot positions.

- $\hat{\hat{H}}$ is the pilot channel estimates
- $\hat{H}$ is the channel estimates on all sub-carriers.

3.2 MMSE estimation

MMSE estimation have to find $\hat{\hat{h}}$ such that which minimizes the average error over all conditions . i.e which minimizes the expected error $E[(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{F}}\hat{\mathbf{h}})^H(\bar{\mathbf{Y}} - \bar{\mathbf{X}}\bar{\mathbf{F}}\hat{\mathbf{h}}).]$
As noise is random over time. So error will also be random over time.
So minimizing the mean of the squared error over time is equivalent to minimizing the expectation of squared error as given in (Ragavendra, 2007).. The above expectation will be simplified to

$$\hat{h} = \arg\min_{h} E[\| h - \hat{h} \|^2] \tag{3.8}$$

the solution for $\hat{h}$ is given as (Morelli and Mengali, 2001)

$$\hat{h}_{mmse} = (\bar{F}_{cp}^H \bar{F}_{cp} + (\sigma)^2 R_h^{-1})^{-1} \bar{F}_{cp}^H \hat{\hat{H}} \tag{3.9}$$

The channel frequency response can be calculated by

$$\hat{\hat{H}}_{mmse} = F\hat{h}_{mmse} \tag{3.10}$$

Therefore if we know noise variance σ^2 and channel covariance (Power profile distribution matrix R) then we can improve the performance of the estimator by MMSE technique.

3.3 GAIC criterion to estimate the sparsity information

The performance of the channel estimator can be improved, if we know more information about channel at receiver. If channel length is known then we can reduce the signal space dimension of the estimator as given in (Ragavendra, 2007)..

How sparsity information reduces the signal space dimension of channel estimator?

We know frequency response of channel is present in column space of fourier matrix(F) as given below

$$\hat{H} = F\hat{h} \tag{3.11}$$

From above equation, we can visualize that the signal space dimension of estimator is determined by the number non zero valued taps of h. So if we estimate the actual length (Cui and Tellambura, 2006) and actual tap positions of channel impulse response then we can reduce the signal space dimension of estimator.

For a sparse channel channel,channel length alone is not sufficient, we have to estimate the sparsity information (multipath delays or effective tap positions of channel impulse response) to improve the performance of the estimator. Channel sparsity estimation techniques for OFDM can be found in (Minn and Bhargava, 2000). Channel having longer delay spread with few multipath components is called sparse channel (Cotter and Rao, 2002). Channel length estimation methods for OFDM are provided in We studied a technique to estimate sparsity information of sampled spaced channel at receiver using generalized Akaike information criterion(GAIC) (Larsson *et al.*, 2001) as given below

The GAIC cost function has the form

$$GAIC(L) = V_L + \gamma ln(ln(p))(L+1) \tag{3.12}$$

$$= \text{MSE for channel length}(L) + f(L) \tag{3.13}$$

where

$$V_L = \frac{p}{2} ln(\sigma_L^2) + \frac{1}{\sigma_L^2} ||\bar{Y} - \bar{X}\bar{F}h_L||^2 + K_0 \tag{3.14}$$

- p is the number of pilots.
- γ is user-specified parameter.
- σ_L^2 is the estimate of noise variance for a channel length(L).

Technique: We will find the value of (L), for which noise variance σ_L^2 is minimum, we treat this as channel length. After finding the channel length, we will remove the effect of this tap then we will find next significant tap position by using same technique.

GAIC algorithm steps are explained below

1. Initially set the limit $\mathcal{L} = L_{cp}$.
2. Calculate the cost function GAIC(L) for $L = 1, 2, ..., \mathcal{L}$.
3. The GAIC estimate of L is then obtained as

$$\hat{L} = \arg\min_{L} GAIC(L) \tag{3.15}$$

4. Note this $\hat{L}$ as significant tap position.
5. Now remove the effect of estimated tap from the received vector by setting $h_{ls}(\hat{L}) = 0$, $\bar{H}_{ls} = \bar{F}_{cp} h_{ls}$ and $\bar{Y} = \bar{X}\bar{H}_{ls}$
6. If $\hat{L} \neq 1$ then repeat steps 2 to 4 with $\mathcal{L} = \hat{L} - 1$ to estimate next significant tap position.
7. Therefore by successively canceling the estimated tap positions in time domain, we estimated the channel length ($\hat{L}$) and significant tap positions of channel. i.e we estimated the sparsity information.

With these estimated tap positions, the improved frequency domain channel estimates are obtained as

$$\hat{h}_{SCE-mls} = (\bar{F}_s^H \bar{F}_s)^{-1} \bar{F}_s^H \hat{\bar{H}} \tag{3.16}$$

$$\hat{H}_{SCE-mls} = F_s \hat{h}_{SCE-mls} \tag{3.17}$$

Where

- F_s is the modified K point DFT matrix whose columns corresponding to estimated channel tap positions and
- $\bar{F}_s$ is derived from F_s with rows corresponding to pilot positions.

3.4 Simulated results

We evaluated the performance of above three estimators, using an OFDM system with specifications given below
Channel model:

- Sparse channel with multipath delays=[1,4,6,7] . (i.e tap positions)
- Rayleigh fading with 4 taps(i.e each tap came from distribution CN(0,1)).

OFDM specifications:

- The cyclic prefix length(Lcp)=25 is done,which is greater than the channel delay spread(7).
- Number of ofdm symbols over which averaging done=10^4.
- Number of sub-carriers in each ofdm symbol(N)=1024.
- Number of sub-carriers used for pilots(p)=32

We estimated the frequency response of channel by each algorithm and we evaluated the performance of each algorithm by calculating the mean square error(MSE) and bit error rate(BER) as given below.

$$MSE = \frac{1}{ns}\sum_{i=1}^{ns} E[||(H_i - \hat{H}_i)||^2] \tag{3.18}$$

Where

- ns is the number of OFDM symbols over which averaging is done.
- H_i is the known frequency channel response for the i^{th} OFDM Symbol.
- $\hat{H}_i$ is the estimated frequency channel response for the i^{th} OFDM Symbol.

Bit error rate calculation(BER): Demodulate the received signal by using H and $\hat{H}$. Then find BER from the two demodulated signals. The simulated results are shown below

3.4.1 Comparison of performance of the different estimators for the specifications : N=1025, Number of pilots(P)=32, cyclic prefix length(Lcp)=25,channel taps(L)=4

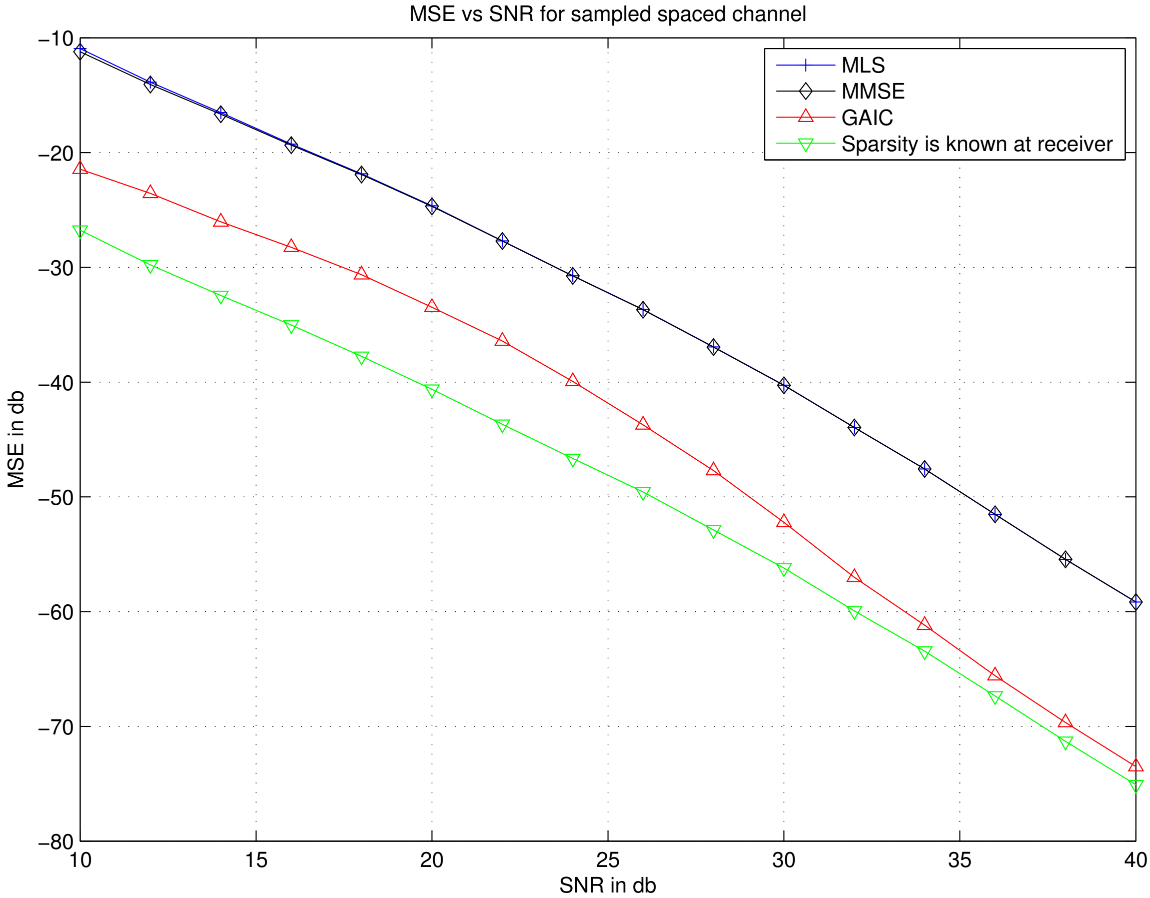

Figure 3.2: MSE performance in db

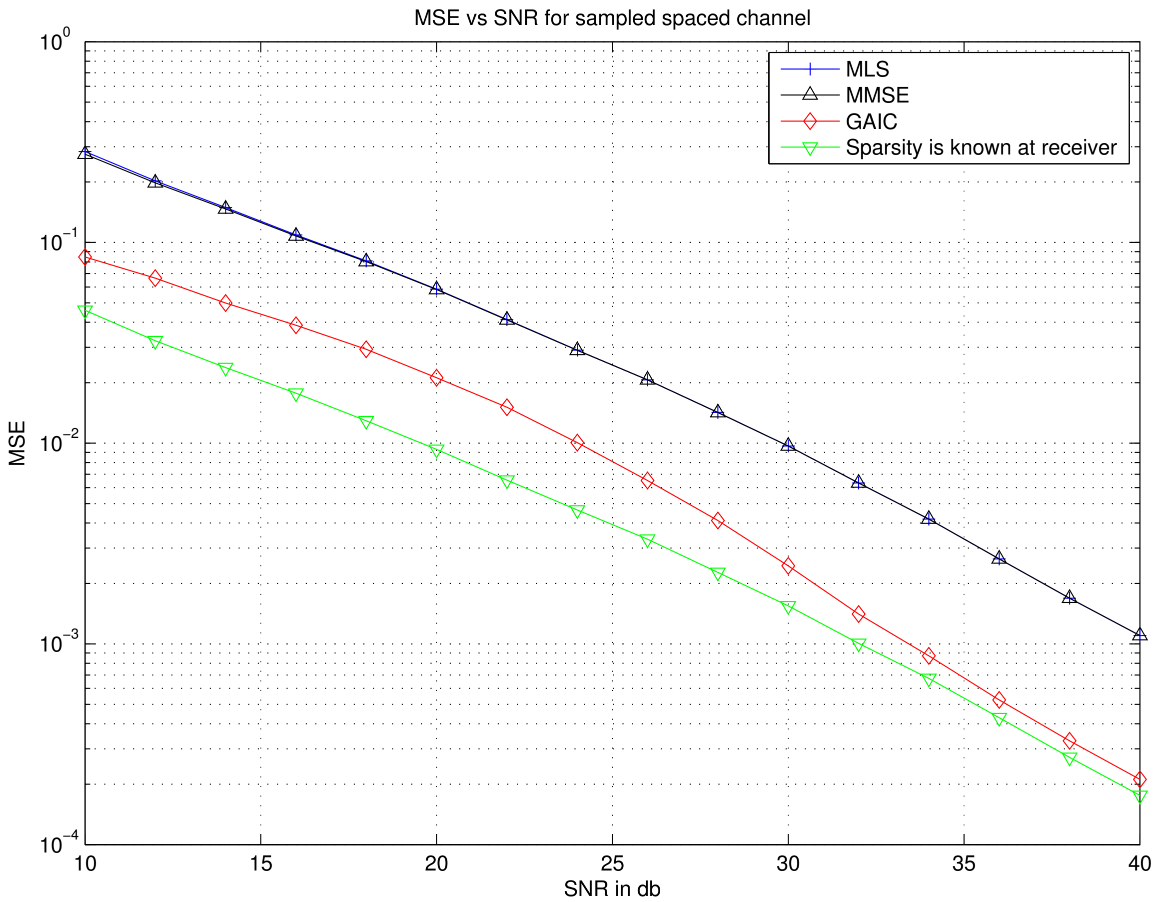

Figure 3.3: MSE performance

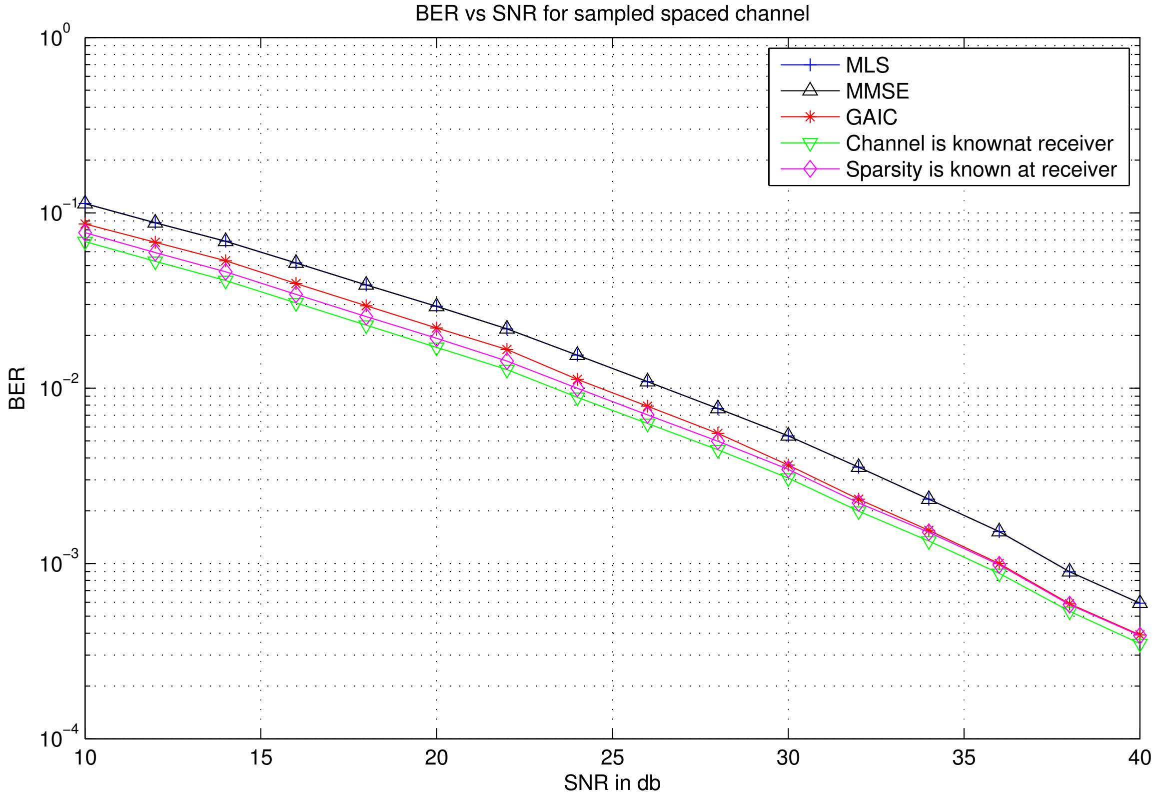

Figure 3.4: BER performance

Figure 3.2. shows the MSE performance of different channel estimators. It is clear that the MLS based channel estimator with sparsity information(SCE-mls) outperforms conventional mls algorithm. We can observe that the performance of the channel sparsity estimator improves with SNR. From Figure 3.4 we can say, the performance of sparsity based MLS estimator improved by 2db.

3.4.2 Effect of increasing the number of pilots(P) for specifications: total sub carriers(N)=1024, cyclic prefix length(Lcp)=25,number of taps of channel(L)=4,delay spread=7,tap positions=1,4,6,7

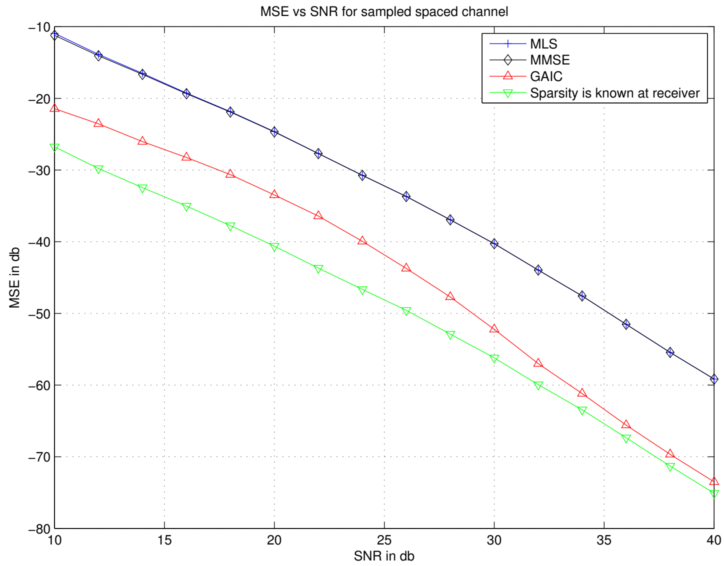

Figure 3.5: MSE performance in db for P=32

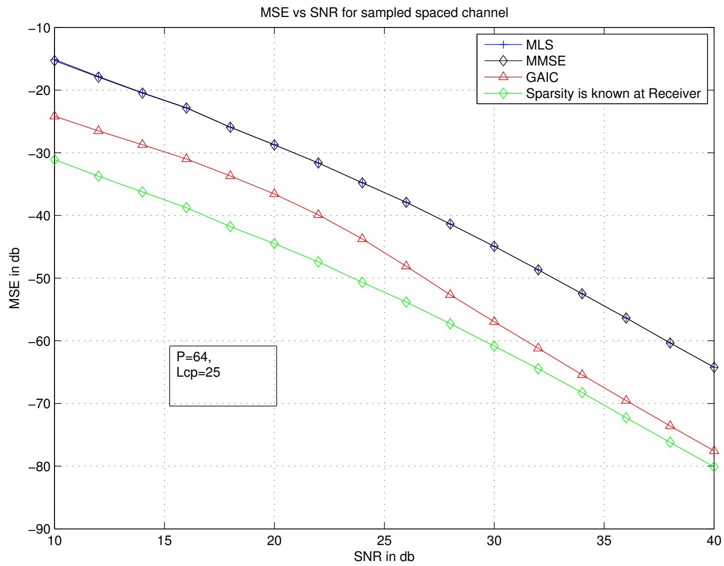

Figure 3.6: MSE performance in db for P=64

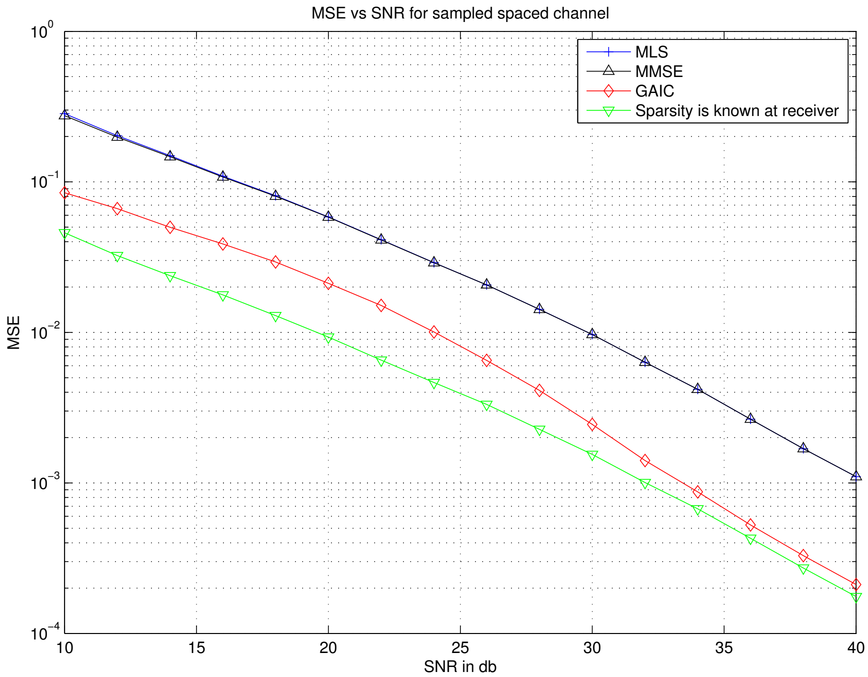

Figure 3.7: MSE performance for P=32

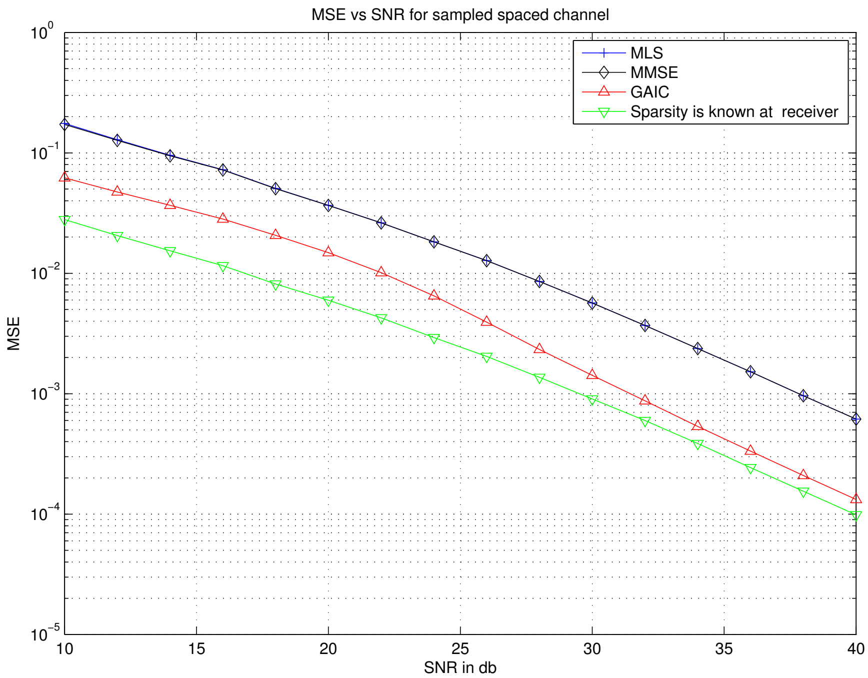

Figure 3.8: MSE performance for P=64

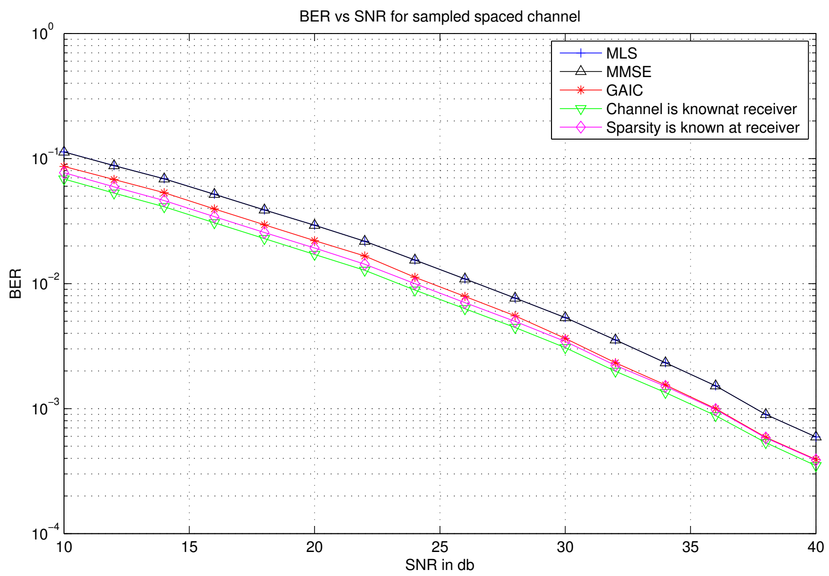

Figure 3.9: BER performance for P=32

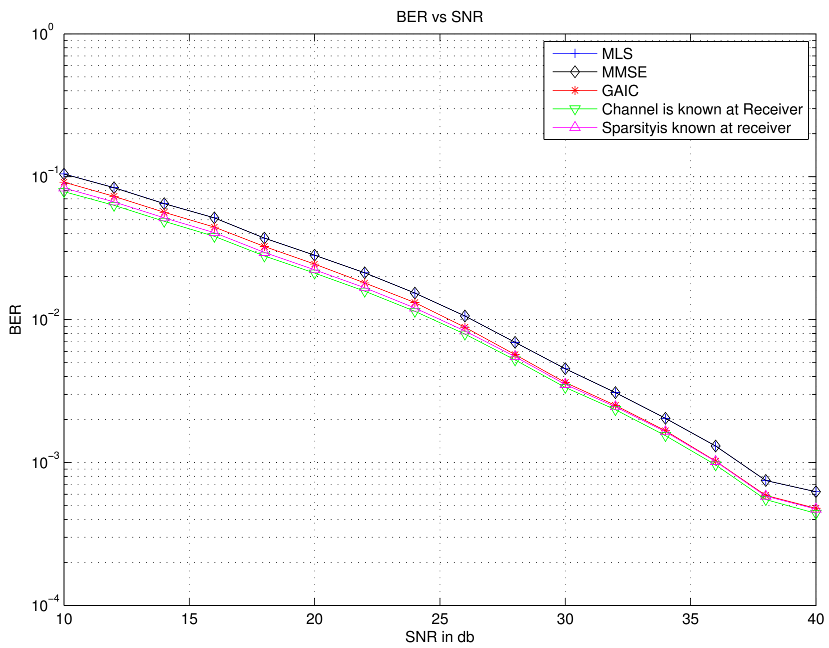

Figure 3.10: BER performance for P=64

From Figure 3.5 and Figure 3.6 we can observe that, by increasing the number of pilots symbols we can improve the performance of estimator. By increasing the number of pilots by 32, the Mean square Error(MSE) of estimator decreases by 3 db as shown in Figures .

3.4.3 Effect of increasing the Cyclic prefix length(Lcp) for specifications: N=1024,P=64,L=4,delay spread=7,tap positions=1,4,6,7

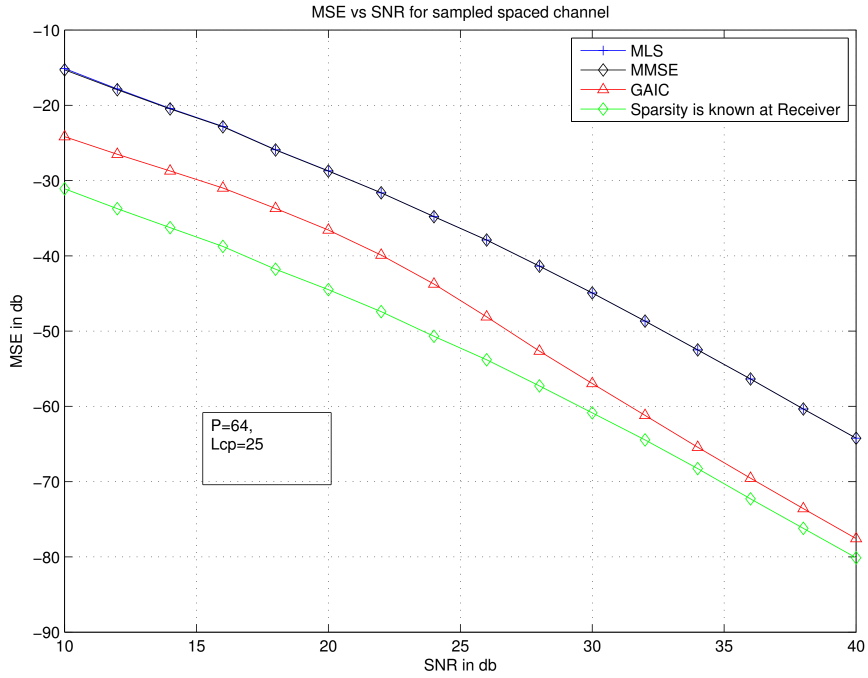

Figure 3.11: MSE performance in db for Lcp=25

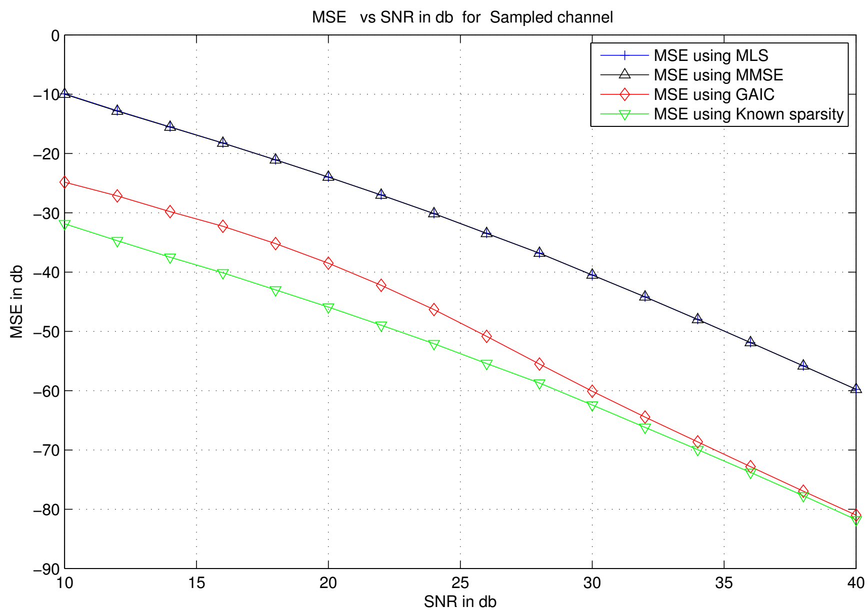

Figure 3.12: MSE performance in db Lcp=50

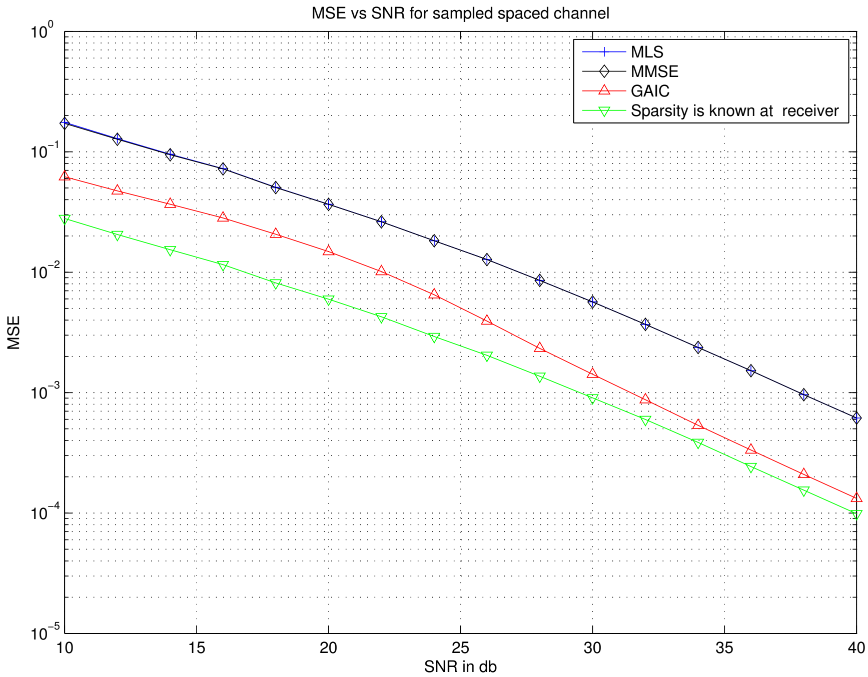

Figure 3.13: MSE performance for Lcp=25

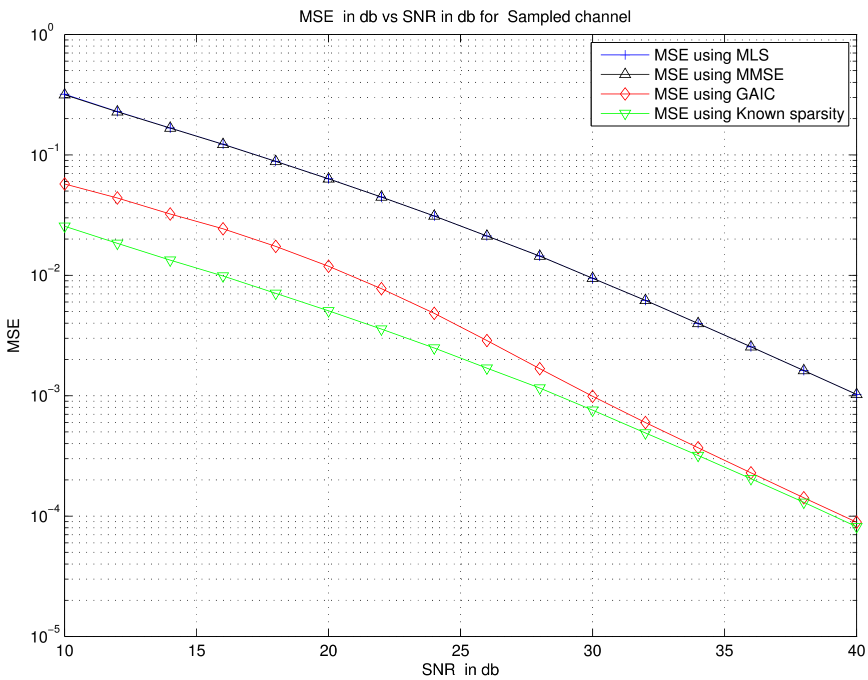

Figure 3.14: MSE performance for Lcp=50

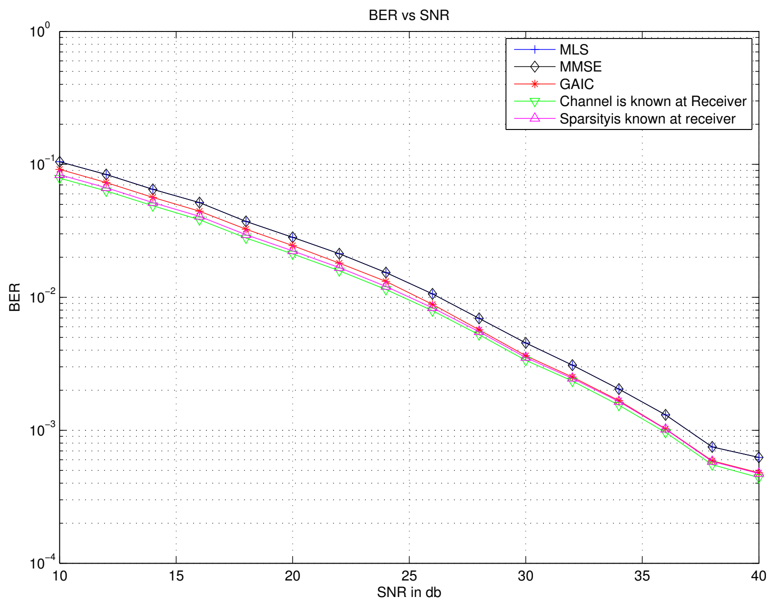

Figure 3.15: BER performance for Lcp=25

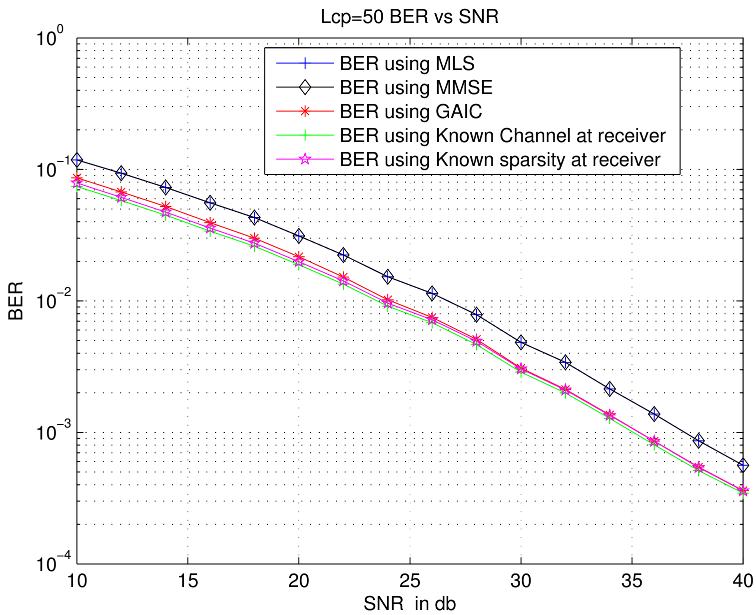

Figure 3.16: BER performance for Lcp=50

From Figure 3.15 and Figure 3.16 we can observe that, by increasing the cyclic prefix for a given number of pilots i.e by increasing the number of variables ,we can observe that, performance gain of sparsity based algorithm over the conventional mls algorithm increases. The number of pilots should be greater than cyclic prefix length.

3.5 Summary

We can increase the performance of channel estimator, if we know more information about channel. Signal space dimension of the estimator can be reduced, if we know the channel length and sparsity information at receiver. Simulated results shows that significant gain in MSE performance can be achieved if we know the channel length and sparsity information at receiver.

CHAPTER 4

Algorithms to estimate the channel for OFDM model for non sampled spaced channel

- The received signal on pilot sub-carriers in frequency domain is given by

$$\bar{Y} = \bar{X}\bar{H} + \bar{V} \tag{4.1}$$

$$= \bar{X}\bar{F}h + \bar{V} \tag{4.2}$$

- Channel estimates on pilot sub-carriers is given by (least squares estimation)

$$\hat{\bar{H}} = \bar{X}^{-1}\bar{Y} \tag{4.3}$$

$$= \bar{H} + \bar{X}^{-1}\bar{V} \tag{4.4}$$

$$= \bar{F}h + \bar{W} \tag{4.5}$$

  where

  - $\bar{X}$ is the diagonal matrix with pilot symbols on its diagonal.
  - $\bar{H}$ is the sampled frequency response of the channel.
  - $\bar{V}$is noise vector.
  - The matrix $\bar{F}$ is derived from F with rows corresponding to pilot positions.

- The channel frequency response H in matrix notation can be written as

$$\begin{bmatrix} H_1 \\ H_2 \\ \vdots \\ \vdots \\ \vdots \\ H_K \end{bmatrix} = \begin{bmatrix} W_{11} & W_{12} & \dots\dots\dots & W_{1K} \\ W_{21} & W_{22} & \dots\dots\dots & W_{2K} \\ \vdots & \vdots & \ddots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ W_{K1} & W_{K2} & \dots\dots\dots & W_{KK} \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \\ \vdots \\ h_L \\ 0 \\ \vdots \\ 0 \end{bmatrix} \tag{4.6}$$

where
$W_{mn} = e^{\frac{-j2\pi mP\tau_n}{KT}}$ $n \in 1, 2, 3......K$

- From above equation we can visualize that H presents in the column space of the matrix F. Therefore by using this matrix F, we can find the channel estimates on data sub-carriers by interpolating the pilot channel estimates.

- For sampled spaced channel where the multipath delays are multiple integrals of input sampling period($\tau_n = nT$ for $n \in 1, 2, 3......K$), the normal Fourier matrix will spans the channel frequency response H.

- For non sampled spaced channel where the multipath delays are not a multiple integrals of input sampling period, if we use the normal Fourier matrix for finding the channel estimates on data sub-carriers by interpolating the pilot channel estimates, then results in irreducible channel interpolation error (van de Beek J.J *et al.*, 1995), (Pirak *et al.*, 2006).
  Therefore we have to find the orthogonal basis that spans H .
  The required orthogonal basis are similar to columns of Normal fourier matrix $[W_{mn}]$
  ($W_{mn} = e^{\frac{-j2\pi m\tau_n}{KT}}$ $n \in 1, 2, 3......K$). So we have to find multipath delays τ_n $n \in 1, 2, 3......K$.

- Multi-path delays can be estimated by ESPRIT technique as given below.

4.1 Estimation of Signal Parameters via Rotational Invariance Techniques (ESPRIT)

High resolution signal parameter estimation is a significant problem in many signal processing applications like direction of arrival(DOA) estimation, system identification and time series analysis. ESPRIT technique can be applied to wide rang of problems. It exploits the rotational invariance among signal subspaces induced by an array of sensors with a translational invariance structure. Here we explained how direction of arrival(DOA) can estimated by ESPRIT technique as given in (Roy R.and Kailath, 1989).

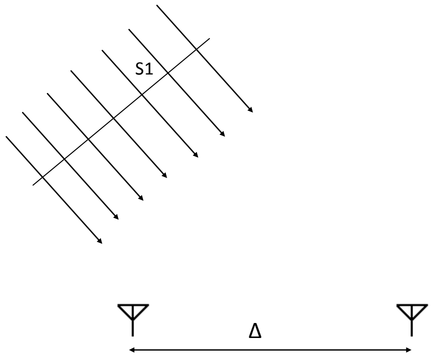

Figure 4.1: Two sensors receiving narrow band signals

- **LOGIC**

  Consider the scenario: The parallel beam of signals are receiving by two sensors separated by a distance(Δ) as shown in Figure 4.2. Then received signal at second sensor will have a phase difference that of the signal received at first sensor. The phase difference is given by $\Phi = w_0 \frac{\Delta \sin(\theta)}{c}$

  where

  w_0 is the operated frequency of source (which is generating the signal). So we can find the direction of arrival (θ) from Φ. Same logic can extended to array of sensors, as given in (Roy R.and Kailath, 1989),is explained below.

- Consider d number of narrow band sources operating at frequency (w_0) generating parallel beams as shown in figure 4.1 and m number of sensors are present in each array.

- The received signal at i^{th} sensor in the first array is given by

$$x_i(t) = \sum_{k=1}^{d} s_k(t) a_i(\theta_k) + n_{x_i}(t) \tag{4.7}$$

where

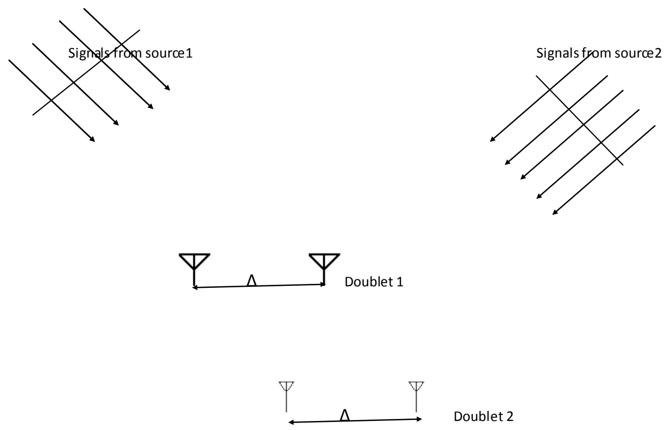

Figure 4.2: Array of sensors

- s_k is signal generated by k^{th} source.
- $a_i(\theta_k)$ is fading coefficient of channel between k^{th} source and i^{th} sensor in the direction of θ_k.
- $n_{x_i}(t)$ is noise signal.

So received signal at m sensors can be written in vector domain as

$$\begin{bmatrix} x_1(t) \\ x_2(t) \\ \vdots \\ x_m(t) \end{bmatrix} = \begin{bmatrix} a_1(\theta_1) & a_1(\theta_2) & \dots & a_1(\theta_d) \\ a_2(\theta_1) & a_2(\theta_2) & \dots & a_2(\theta_d) \\ \vdots & \vdots & \ddots & \vdots \\ a_m(\theta_1) & a_m(\theta_2) & \dots & a_m(\theta_d) \end{bmatrix} \begin{bmatrix} s_1(t) \\ s_2(t) \\ \vdots \\ s_d(t) \end{bmatrix} + \begin{bmatrix} n_1(t) \\ n_2(t) \\ \vdots \\ n_m(t) \end{bmatrix} \tag{4.8}$$

In matrix notion it can be written as **X =AS+N**

- The received signal at i^{th} sensor in the second array is given by

$$y_i(t) = \sum_{k=1}^{d} s_k(t) a_i(\theta_k) e^{-jw_0\tau_k(\theta_k)} + n_{y_i}(t) \tag{4.9}$$

where

- s_k is signal generated by k^{th} source.
- $a_i(\theta_k$ is fading coefficient of channel between k^{th} source and i^{th} sensor in the direction of θ_k.
- $n_{y_i}(t)$ is noise signal, and$\tau_k(\theta_k)$ is propagation delay.

So received signal at m sensors can be written in vector domain as

$$\begin{bmatrix} y_1(t) \\ y_2(t) \\ \vdots \\ y_m(t) \end{bmatrix} = \begin{bmatrix} a_1(\theta_1) & a_1(\theta_2) & \dots & a_1(\theta_d) \\ a_2(\theta_1) & a_2(\theta_2) & \dots & a_2(\theta_d) \\ \vdots & \vdots & \ddots & \vdots \\ a_m(\theta_1) & a_m(\theta_2) & \dots & a_m(\theta_d) \end{bmatrix} \begin{bmatrix} e^{-jw_0\tau_1(\theta_1)} & 0 & \dots & 0 \\ 0 & e^{-jw_0\tau_2(\theta_2)} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & e^{-jw_0\tau_d(\theta_d)} \end{bmatrix}$$

$$\times \begin{bmatrix} s_1(t) \\ s_2(t) \\ \vdots \\ s_d(t) \end{bmatrix} + \begin{bmatrix} n_1(t) \\ n_2(t) \\ \vdots \\ n_m(t) \end{bmatrix}$$

In matrix notion it can be written as $\mathbf{Y} = \mathbf{AS\Phi} + \mathbf{N}$

- **Analysis**

  We can visualize the following points, from the equations for **X** and **Y** in matrix notations.

  If we rotate the vector **X** by Φ then we get the vector **Y**.

  It indicates that subspace of **X** and **Y** are same.

  The orthogonal basis that spans **Y** can be obtained by, rotating the orthogonal basis that spans **X** by Φ.

- **Problem**

  If we know the signal vectors(**X** and **Y**) received by two arrays of sensors then we have to find the rotational matrix Φ such that **Y** should be obtained from **X**.

- **Solution**

  Φ can be estimated by subspace model approach as given in (Roy R.and Kailath, 1989).

- We can find direction of arrival($\theta_k^{'}$s) from matrix Φ.

4.2 Estimating multi path delays by using ESPRIT technique and estimating the channel

We will derive shift invariance subspaces by sending equispaced pilots as given in (Ragavendra, 2007). From these two invariance subspaces, we will find rotational matrix. From rotational matrix we will find multipath delays.

Algorithm

.

1. We can calculate the pilot channel estimates as below

$$\hat{\bar{H}} = \bar{X}^{-1}\bar{Y} \tag{4.10}$$

2. We have to find the subspace which spans this frequency response of channel $\hat{\bar{H}}$. we used auto correlation method is given below

   - Find autocorrelation matrix (R) of $\hat{\bar{H}}$ by using equation R=$\hat{\bar{H}}.\hat{\bar{H}}^{H}$.
   - Find average autocorrelation(R) by average over 'ns' number of OFDM symbols by using equation $R = \frac{1}{ns}\sum_{i=1}^{ns} R_i$.
   - Form **U** matrix with columns as, $\hat{L}$ number of eigenvectors corresponding to $\hat{L}$ dominant eigenvalues.

     If absolute value of eigenvalue is greater than noise power then eigenvalue is considered as dominant eigenvalue.
   - Now the columns of matrix(U) spans the frequency response of channel$\hat{\bar{H}}$.

3. Now split the matrix(U) into two matrices U_1, U_2, both matrices U_1, U_2 spans same subspace.

4. Find the rotational matrix Υ such that $U_2 = U_1\Upsilon$.

5. Find eigenvalues(λ) of Υ.

6. Find the argument of (λ^*) and round off to $[0, 2\pi]$ and from the argument of (λ^*), we can find the multipath delays.

7. Now matrix which spans the $\hat{\bar{H}}$, can be written as $[\bar{F}_{i,l}] = e^{-j.(argument_l)(i)}$. Where

   - $l = 1, 2...\hat{L}$.

- i=$0, 1, 2....(p-1)$ where 'p' is the number of pilot sub-carriers.

8. Now we will find impulse response of channel and the data channel estimates can be calculated by interpolating the pilot channel estimates, as given below

$$\hat{h} = (\bar{F}^H \bar{F})^{-1} \bar{F}^H \hat{\bar{H}} \tag{4.11}$$

$$\hat{H} = F\hat{h} \tag{4.12}$$

where

- $\hat{h}$ is the $\hat{L} \times 1$ vector of time domain channel estimates
- The matrix (F) is like $\bar{F}$ but $i = 0, 1, 2....(K-1)$

Therefore by this algorithm, we can estimate multipath delays by using ESPRIT technique, so that we can improve the performance of channel estimator in non sampled spaced channel.

4.2.1 Comparision of performance of the different estimators for the specifications : N=1025, Number of pilots(P)=32, cyclic prefix length(Lcp)=25,channel taps(L)=4,delayspread =7

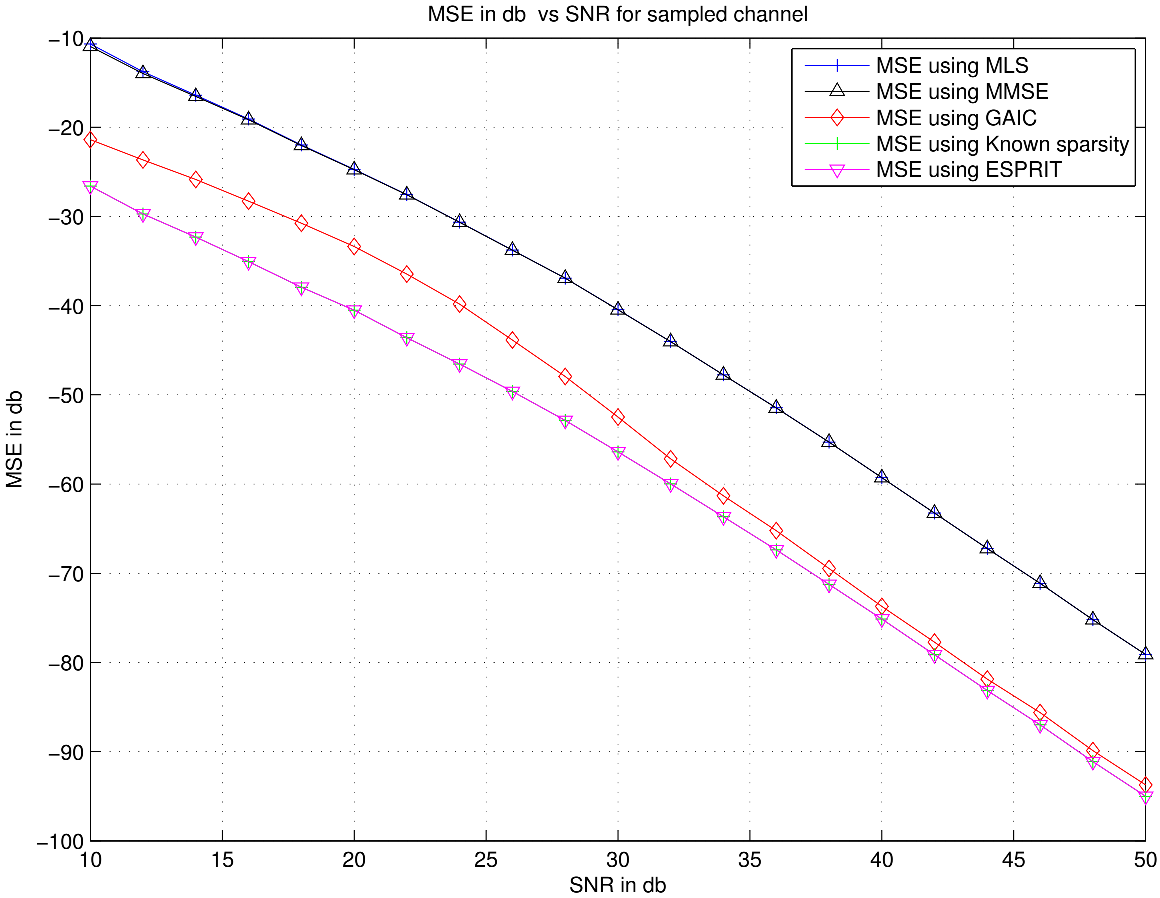

Figure 4.3: MSE performance in db

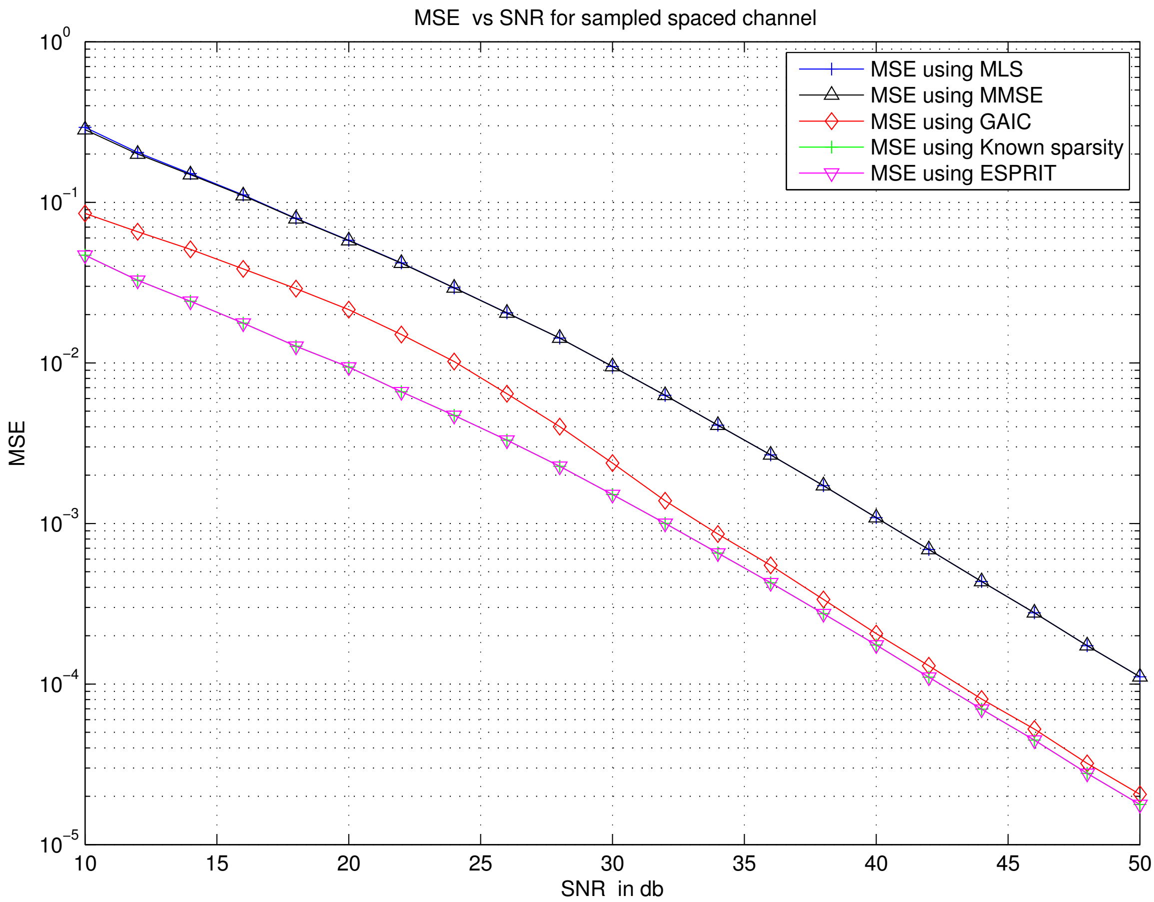

Figure 4.4: MSE performance

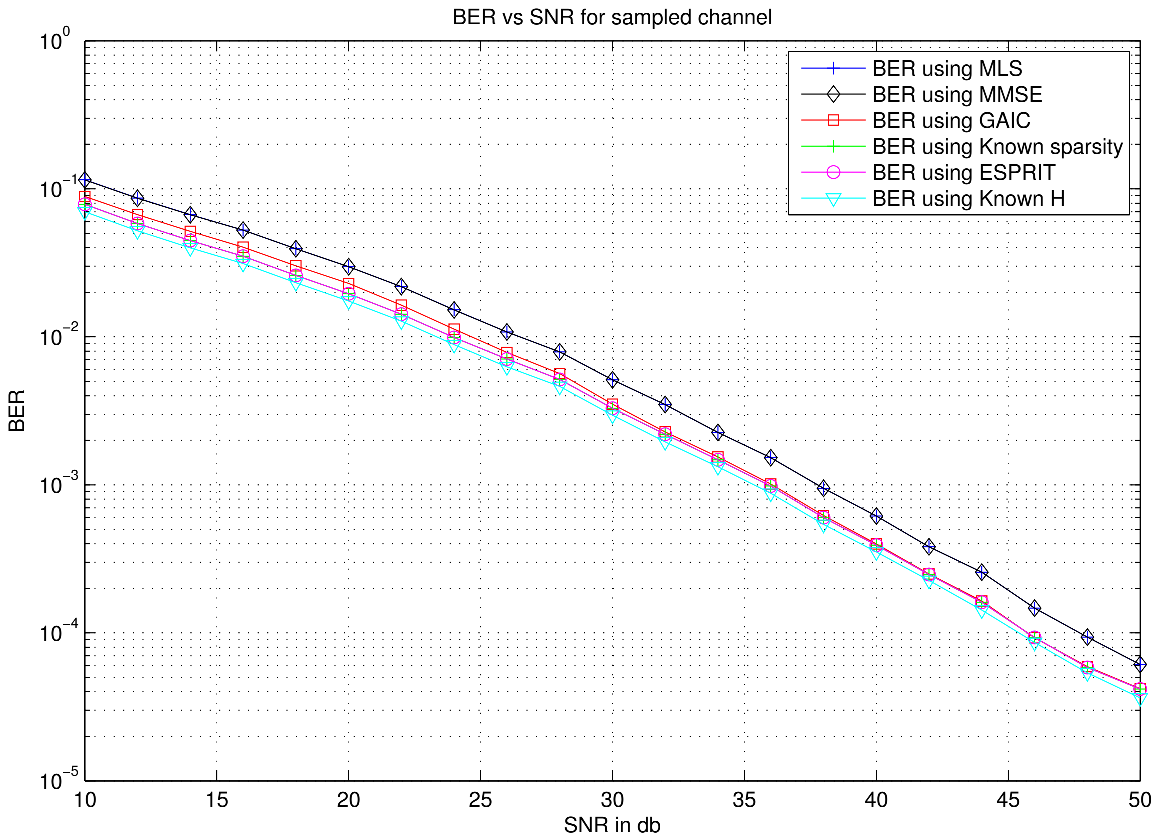

Figure 4.5: BER performance

From Figure 4.3 and Figure 4.5 we can observe that,the ESPRIT based estimator outperforms the conventional mls and GAIC based estimator. We can observe that, performance of ESPRIT based estimator is close to ideal sparsity known estimator.

CHAPTER 5

Conclusions

The demand for the wireless services increased so the radio frequency bands are precious, so the need for spectrally efficient signaling schemes and reuse-1 cellular deployment to serve more users are imminent. The performance of wireless communication systems employing coherent receiver is superior to non-coherent receiver and differential techniques.

We introduced multi-carrier schemes like OFDM and its operation in both signal and vector domain. We discussed the advantages of OFDM technology and significance of OFDM in future generation.

If we know channel state information (CSI) at transmitter, it helps in improving the spectral efficiency of the system by adaptive modulation techniques like, power allocation (by water filling method) and changing the signal constellation are varied on each sub-carrier depending on the CSI.

To achieve the full capability of MIMO technology the CSI at transmitter and receiver is important. To avoid the system performance degradation due to channel estimation errors,the mean squared error (MSE) of channel estimation should be negligible compared to the operating SNR. Therefore we need a efficient channel estimation algorithm is required at receiver.

We studied the channel estimation algorithms for sampled spaced channel. If we know channel information like noise variance and channel covariance then we will use MMSE technique which improves the performance of estimator. We introduced an algorithm to estimate the sparsity information so that we can reduce the signal space dimension of estimator and we can improve the performance of the estimator.

5.1 Future work

We can develop new techniques to estimate the multipath delays of the channel, if we analysis and simulate the channel estimation techniques, given in (Ragavendra, 2007),some of the possible work is given below

5.1.1 Optimal pilot pattern for channel estimation

Pilot pattern plays an important role in estimating the multipath delays because equispaced pilot pattern reduces the MSE of the channel estimates and cluster based pilot pattern increases the range of delay spread estimation. So it is interesting to find optimal pilot pattern that will get both get both benefits.

5.1.2 Estimation of broadcast channels

The broadcast channels will have longer delay spreads because the broadcast done by multiple synchronous transmitters are received at different time delays. The delay spread of these channels can be greater than cyclic prefix length. So inter-block interference will take place.

It is interesting to study the effect of inter-block interference in estimating multipath parameters for such long delay spread channels.

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