SPACE-FREQUENCY BLOCK-CODED COMMUNICATION SYSTEMS FOR TIME-VARYING WIRELESS FADING CHANNELS

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Submitted by

Manu Prakash

Roll No. 801061015

Under guidance of

Dr. Amit Kumar Kohli

Assistant Professor, ECED

T.U. Patiala

ELECTRONICS AND COMMUNICATION ENGINEERING DEPARTMENT

THAPAR UNIVERSITY

(Established under the section 3 of UGC Act, 1956)

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DECLARATION

I, **Manu Prakash**, hereby certify that the work which is being presented in this thesis entitled **"Space-Frequency Block-Coded Communication Systems for Time-Varying Wireless Fading Channels"** by me in partial fulfillment of the requirements for the award of degree of Master of Engineering in Electronics and Communication Engineering from Thapar University (Deemed University), Patiala, is an authentic record of my own work carried out under the supervision of **Dr. Amit Kumar Kohli**.

The matter presented in this thesis has not been submitted in any other University / Institute for the award of any other degree.

Manu Prakash.

Date: 19/6/12

Manu Prakash
Roll No.801061015

It is certified that the above statement made by the student is correct to the best of my knowledge and belief.

Date: 19/6/12

19/6/2012

Dr. Amit Kumar Kohli
Assistant Professor,
ECED, TU, Patiala

Countersigned by:

(Dr. Rajesh Khanna)
Professor and Head ECED
Thapar University, Patiala
Date:

(Dr. S.K. Mohapatra)
Dean of Academic Affairs
Thapar University, Patiala
Date:

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ABSTRACT

Fading is one of the most challenging phenomena in wireless to establish a reliable communication path between transmitter and receiver. Fading is very severe when there is no line-of-sight component, which is the most occurring case in urban and suburban cities. Generally, the multipath fading amplitude distribution is modeled with Rayleigh PDF. But when the fading is very severe, the Rayleigh model fails to characterize the exact channel characteristics. So, a more accurate model, named Nakagami-*m* model may use to represent the channel.

When channel bandwidth is smaller than the signal bandwidth, the channel creates frequency selective fading. This frequency selective fading causes inter symbol interference (ISI). There are several ways to combat ISI. The equalization is one way that tries to invert the effects of the channel. In OFDM, each subcarrier undergoes flat fading since its bandwidth is less than the coherence bandwidth.

OFDM converts a frequency selective channel into a collection of flat fading channels by the use of cyclic prefix, but sub-band of OFDM are exposed to deep fading and may cause complete loss of sub-band information. To cope with this problem space time block coding is applied in OFDM, however it also introduces redundancy. To avoid the problem of fast channel variations in time, the symbols of an orthogonal design can be transmitted on neighboring subcarriers of the same OFDM symbol rather than on the same sub-carrier at subsequent time interval this is basic idea behind "Space Frequency Block Coded OFDM (SFBC-OFDM)"

Therefore, we have made an effort to study STBC-OFDM and evaluation of cyclic prefix (CP) single carrier transmission with SFBC-OFDM working under various fading wireless environment. By appending CP at beginning of OFDM symbol block, the linear convolution associated with the channel impulse becomes a circular convolution, therefore inter-block interference is prevented. The SFBC system codes across two antennas and over two adjacent subcarriers instead of two consecutive symbol intervals and therefore becomes robust against fast fading distortion in frequency nonselective fading environments.

Simulation results are presented to demonstrate that in mobile environments where fading is relative slow such as Rayleigh (Nakagami-m, m=1), SFBC-OFDM does not give any significant performance over STBC-OFDM; more over STBC-OFDM mitigate slow fading (m > 1) more efficiently than SFBC-OFDM. It is also shown that under fast fading (m < 1) conditions SFBC-OFDM outperform STBC-OFDM. For poor SNR (SNR < 11dB), fast fading (m=0.75) is mitigated more efficiently than slow fading (m=1.25) by SFBC-OFDM.

Keywords: Frequency-selective fading, Nakagami-m, OFDM, SFBC, STBC, time-selective fading.

CONTENTS

LIST OF ABBREVIATIONS

3G	third generation
3GPP-LTE	third generation partnership project long term evolution
A/D	analog to digital
ACE	active constellation extension
ADSL	asymmetric digital subscriber line
AWGN	additive white Gaussian noise
BER	bit error rate
BPSK	binary phase shift keying
BTS	base transceiver station
CDF	cumulative distribution function
CDMA	code division multiple access
CFR	channel frequency response
CP	cyclic prefix
CSI	channel state information
D/A	digital to analog
DAB	digital audio broadcasting
DFT	discrete Fourier transform
DVB	digital video broadcasting
ECC	error correction codes
FDE	frequency domain equalization
FDMA	frequency division multiplexing
FFT	fast Fourier transform
FIR	finite impulse response
GI	guard interval
IBI	inter block interference
ICI	inter carrier interference
IDFT	inverse discrete Fourier transform
IEEE	institute of electrical and electronics engineers
IFFT	inverse fast Fourier transform

ISI	inter symbol interference
LAN	local area network
LCF	linear complex field
LOS	line of sight
LMS	least mean square
LS	least square
LTE	long term evaluation
MIMO	multi input multi output
ML	maximum likelihood
MRRC	maximum ratio receive combining
OC	orthogonal code
OFDM	orthogonal frequency division multiplexing
PAM	pulse amplitude modulation
PAPR	peak to average power ratio
PCC	polynomial cancellation code
PDF	probability density function
PSK	phase shift keying
QAM	quadrature amplitude modulation
QPSK	quadrature phase shift keying
RF	radio frequency
RMS	root mean square
SC	single carrier
SER	symbol error rate
SF	space frequency
SFBC	space frequency block code
SI	side information
SLM	selected mapping
SNR	signal to noise ratio
ST	space time
STBC	space time block code
STC	space time code

STTC	space time trellis code
Wi-MAX	worldwide interoperability for microwave access

LIST OF FIGURES

CHAPTER 1

INTRODUCTION

1.1 Introduction to Fading and Diversity

Severe multipath propagation raised due to multiple scattering by buildings and other rigid structures in the vicinity of a mobile unit, makes mobile communication very challenging. This scattering results in random amplitude and phase variations in the received signal as the vehicle moves in the multipath field. In addition, the vehicle motion introduces a Doppler shift, which causes a broadening of the signal spectrum. Measurements confirm that the short term statistics of the resultant signal envelope approximate a *Rayleigh* distribution.

Multipath fading may also be frequency selective, that is, the complex fading envelope of the received signal at one frequency may be only partially correlated with the received envelope at a different frequency. This de-correlation is due to the difference in propagation time delays associated with the various scattered waves making up the total signal. The spread in arrival times, known as delay spread, causes transmitted data pulses to overlap, resulting in inter symbol interference (ISI). In a typical urban environment, a spread of several microseconds and greater can be occasionally expected. There is an additional impairment in a cellular mobile system. The available radio channels are reused at different locations within the overall cellular service area in order to use the assigned spectrum more efficiently. Thus, mobiles simultaneously using the same channel in different locations interfere with each other. This is termed *co-channel interference* and is often the dominant impairment. In addition, there is a long-term variation of the local mean of the received signal, called *shadow fading*. Shadow fading in a mobile radio environment is caused by large obstacles blocking the transmission path.

Therefore wireless communication systems encounter multipath fading and multipath fading can often be relatively deep, i.e. the signals fade completely away, whereas at other times the fading may not cause the signal to fall below a useable strength. The fading channel could be modeled with Nakagami-m distribution if the fading is severe compared to the Rayleigh distribution model. The Rayleigh distribution is a special case of Nakagami-m when m=1.

Multipath, which causes multiple echoes of the transmitted signal to be received with delay spreads of up to tens of microseconds [1]. For bit rates in the range of tens of megabits per second, this translates to inter symbol interference that can span up to 100 or more data symbols. For example, at a 5 MHz symbol rate, a 20 ms multipath delay profile spans 100 data symbols.

In multi path channels, small scale fading is the main issue to concentrate for communication engineer. Due to this small scale fading, the signal strength gets rapid fluctuations over a small travel distance. Wireless broadband systems offer different sources of diversity to combat fading, which can be properly exploited by a proper coding and transmission scheme. The common forms of diversity are:

- Temporal diversity: Replicas of the information bearing signal are transmitted in different time slots, where the separation between the time slots is greater than the coherence of the channel.
- Frequency diversity: In this case, replicas of the information bearing signal are transmitted in different frequency bands, where the separation between the frequencies is greater than the coherence bandwidth of the channel.
- Antenna (spatial) diversity: It has been observed that antennas with a spacing of more than half a wavelength leads to spatially uncorrelated channels. The transmission of the replicas of the information bearing signal over these uncorrelated spatial channels leads to spatial diversity.

Multiple antennas and space time codes can be used to obtain spatial diversity [2] [3] [4]. Space-time Codes (STC) were first introduced by G. J. Foschini from Bell Labs [5] in 1996 as a novel means of providing transmit diversity for the multiple antenna fading channel. There are two main types of STC:

- Space-time Trellis Codes (STTC)
- Space-time Block Codes (STBC)

The original STTC were introduced by Tarokh et al. in [2], these original STTC design were hand crafted and therefore, are not optimum design. In subsequent years, a large number of research proposals were published which proposed new code construction or perform systematic searches for different convolution STTC or some variant of the original design criteria proposed by Tarokh et al. [2] The maximum diversity can be realized using the Space-Time Block Codes (STBC) proposed by Alamouti [6] and Tarokh et al. [7] which provide a simple transmit diversity scheme in a flat fading Multiple-Input Multiple Output (MIMO) channel. The available MIMO techniques are very useful in flat frequency channels. Orthogonal Frequency Division Multiplexing (OFDM) converts a wide band frequency to multiple narrow bands which almost have flat frequency. So we can use MIMO with OFDM to transmit data in wide band frequencies achieving high efficiency and low bit error rate. Space time or space frequency along with OFDM can utilize the orthogonal transmission by Alamouti Scheme [6].

$$\begin{bmatrix} X_1 \\ X_2 \end{bmatrix} \rightarrow \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_1^* \end{bmatrix} \begin{matrix} \rightarrow \textbf{Space} \\ \downarrow \textbf{Time} \end{matrix}$$

Fig. 1.1: Alamouti STBC [8]

Orthogonal frequency division multiplexing (OFDM) significantly reduces receiver complexity in wireless broadband systems, and it currently been use in wireless broadband multi-antenna systems. It is also a promising modulation technique to be used in multiple-input multiple-output (MlMO) systems. The orthogonal and independent sub-channels in OFDM are achieved by appending a cyclic prefix to each transmitted block of data, which is then discarded at the receiver. The cyclic extension results in overhead transmission time and reduces the rate & efficiency of the communications link. This overhead in time is always present, since the length of the cyclic prefix is usually chosen to cover the longest channel spread.

Cyclic guard interval (GI) is inserted at each antenna in OFDM provide a flat fading channel for each sub carrier. Therefore, space-time block codes can be used in conjunction with OFDM. A cyclic prefix (CP) instead of GI appended in the head of each transmitted block to eliminate inter block interference (IBI), makes all channel matrices circulant. However, in the frequency-selective broadband channels, as the broadband channel is subdivided into orthogonal narrowband channels, the symbol duration is increased significantly as compared to that of single carrier of the same total bandwidth.

STBC was first designed assuming a narrowband wireless system, i.e. a flat fading channel. However, when used over frequency selective channels a channel equalizer has to be used at the receiver along with the space-time decoder. For the STBC, the channel equalization problem was addressed by modifying the original Alamouti scheme in such a way that the use over frequency selective channel, and hence the equalization is a much easier task. Therefore STBC was used in conjunction with OFDM. OFDM is used to convert the frequency selective channel into a set of independent, parallel and flat-frequency sub-channels. The Alamouti scheme is then applied to two consecutive subcarriers (or two consecutive OFDM block).

1.2 Introduction to SFBC-OFDM with Frequency Domain Equalization

Depending on how rapidly the transmitted baseband signal changes as compared to the rate of change of the channel, a channel is classified either as a fast fading or slow fading channel. In a fast fading channel, the channel impulse response changes rapidly within the symbol duration. That is, the coherence time of the channel is smaller than the symbol period of the transmitted signal. This causes frequency dispersion (also called time selective fading) due to Doppler spreading, which leads to signal distortion. When viewed in the frequency domain, signal distortion due to fast fading increases with increasing Doppler spread relative to the bandwidth of the transmitted signal. In a fast fading channel use of space frequency block code is beneficial because code is transmitted here on neighboring sub carriers [8] [9].

Space frequency block code has the better performance compared to space- time block coding in highly varying environments, i.e., where the channel varies too quickly [10]. In slow varying channels the performance of space frequency as well as space time block codes are the same. The space time block coding or space frequency block coding can be used in conjunction with OFDM to achieve higher signal to noise ratio (SNR).

To mitigate fast fading distortion caused by high-speed mobility a SFBC SC-FDE system [11] can be used for reliable communication. The SFBC system codes across two antennas and over two adjacent subcarriers instead of two consecutive symbol intervals and thus becomes robust against fast fading distortion in frequency nonselective fading environments

[6], [9]. In a conventional SFBC-OFDM system, pairs of information symbols (X1; X2) are fed to the SFBC encoder according to the Alamouti scheme.

$$\begin{bmatrix} X_1 \\ X_2 \end{bmatrix} \rightarrow \begin{bmatrix} X_1 & X_2 \\ -X_2^* & X_1^* \end{bmatrix} \begin{matrix} \rightarrow & \text{space} \\ \downarrow & \text{subcarrier} \end{matrix}$$

Fig. 1.2.1: Alamouti SFBC [8]

Therefore each Alamouti code-word spans two adjacent subcarriers. Assuming the Channel Frequency Response (CFR) is fixed over each pair of adjacent subcarriers, each orthogonal Alamouti code-word can be detected optimally using a space-frequency matched filter in the absence of ICI. However, in the presence of high Doppler spread, severe ICI from adjacent subcarriers destroys the Alamouti structure resulting in significant performance degradation. The block diagram of the SFBC-OFDM system with 2 T_x transmit antennas is shown in figure 1.2 The implementation of SFBC-OFDM is slightly more complex than for the STBC-OFDM system. In this, the encoding is performed across adjacent sub-carriers within a specific OFDM symbol [8], [12].

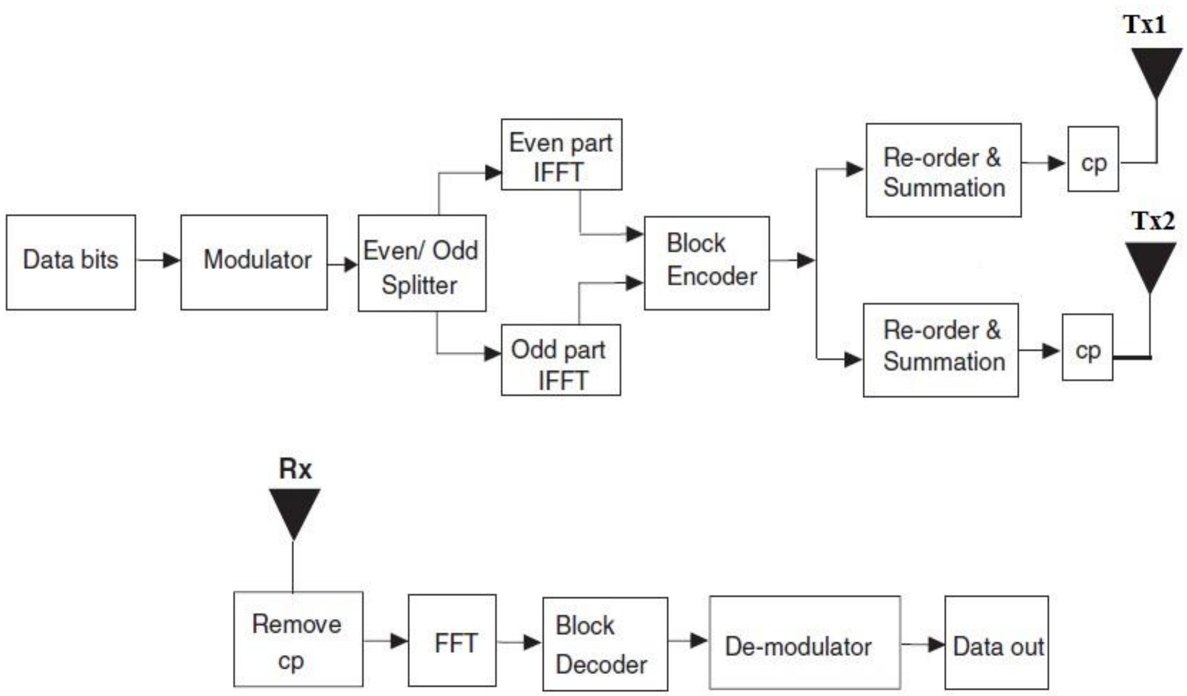

Fig. 1.2.2: Time domain implementation of SFBC-OFDM system

Therefore, when considering this scheme in the time domain and decoding in the frequency domain, each OFDM symbol has to be decomposed into even and odd parts. The data stream is first modulated by a PSK/QAM modulator. The serial to parallel converter output vectors. Implementation of STBC and SFBC OFDM are discussed in details in chapter 4. Decoding of STBC or SFBC requires the knowledge of the channel at the receiver. This knowledge of channel is commonly called channel state information (CSI) can be obtained at the receiver by sending the pilot symbols from each of the transmitter antennas to the receiver antenna and thus CSI can be estimated, but is not emphasized in this thesis. However better estimation of CSI improves the performance of the STBC or SFBC in the time-varying environment.

A conventional anti-multipath approach, which was pioneered in voice-band telephone modems and has been applied in many other digital communications systems, is to transmit a single carrier, modulated by data using, for example, quadrature amplitude modulation (QAM), and to use an adaptive equalizer at the receiver to compensate for ISI. Its main components are one or more adaptive filters for which the number of adaptive tap coefficients is f_n the order of the number of data symbols spanned by the multipath. For 20ms delay spread for example, this would mean a transversal filter with at least 100 taps, and at least several hundred multiplication operations per data symbol. For tens of mega-symbols per second and more than about 30–50-symbol ISI, the complexity and required digital processing speed become exorbitant, and this time domain equalization approach becomes unattractive. Therefore better approach for channel equalization is frequency domain equalization (FDE).

1.3 Problem Statement

This thesis report presents the following work

1. Firstly, the STBC-OFDM and SFBC-OFDM wireless communication systems are investigated for fast fading wireless channel.
2. Next, we have evaluated the cyclic prefix single carrier transmission with SFBC-OFDM over mobile wireless Rayleigh channel.
3. Subsequently, we have evaluated the cyclic prefix single carrier transmission with SFBC-OFDM over mobile wireless Nakagami-m channels.
4. Further, the performance of SFBC-PFDM system is compared with STBC-OFDM system under similar conditions.

1.4 Organization of Report

This report is organized in six chapters

- Chapter I summarize the basic problem statement of the research work and give the overview of the Fading, OFDM and SFBC-OFDM.
- In Chapter II, we review the related work done in the area of SFBC-OFDM.
- Chapter III introduces different fading model and diversity techniques to combat fading along with description of OFDM.
- Chapter IV briefly discusses how STBC is implemented in OFDM and deep mathematical description of SFBC implementation in OFDM.
- In Chapter V we showed simulation results of STBC-OFDM and SFBC-OFDM for various fading channels.
- Finally in Chapter VI we gave concluding remark and future scope.

CHAPTER 2

LITERATURE SURVEY

Diversity

Commonly, wireless channel suffers fading due to constructive or destructive addition of multipath in the propagation media and to interference from other users. The channel statistic usually referred as Rayleigh fading which causes difficulty for the receiver to reliably determine the transmitted signal unless some less attenuated replica of the signal is available at the receiver. This is so called technique diversity, which can be provided using temporal, frequency, polarization, and spatial resources [13][14]. In many situations, however, the wireless channel is neither significantly time-variant nor highly frequency selective. This forced the system engineers to consider the possibility of deploying multiple antennas at both the transmitter and receiver to achieve spatial diversity.

In 1993 transmit diversity was studied extensively by Wittneben [15] as a method of combating detrimental effects in wireless fading channels because of its relative simplicity of implementation and feasibility of having multiple antennas at the base station; and therefore first bandwidth efficient transmit diversity scheme was proposed. Later Foschini introduced a multi-layered space–time architecture [5].

Space-Time Codes

Space–Time Trellis Coding (STTC) was proposed by V. Tarokh, N. Seshadri, and A. R. Calderbank [16], in year 1998 which combines signal processing at the receiver with coding techniques appropriate to multiple transmit antennas and provides significant gain over [17] and [15]. Specific space–time trellis codes designed for two–four transmit antennas perform extremely well in slow fading environments (typical of indoor transmission) and come within 2–3 dB of the outage capacity computed by Telatar [18] and independently by Foschini and Gans [19]. The bandwidth efficiency is about three–four times that of current systems. The space–time codes presented in [16] provide the best possible tradeoff between constellation size, data rate, diversity advantage, and trellis complexity. When the number of transmit antennas is fixed, the decoding complexity of space–time trellis coding (measured by the number of trellis states in the decoder) increases exponentially as a function of both the diversity level and the transmission rate.

Space–time block coding, introduced in [20] and [21], generalizes the transmission scheme discovered by Alamouti to an arbitrary number of transmit antennas and is able to achieve the full diversity promised by transmit and receive antennas. These codes retain the property of having a very simple maximum likelihood decoding algorithm -based only on linear processing at the receiver [20], [21]. For real signal constellations (such as *pulse amplitude modulation*), they provide the maximum possible transmission rate allowed by the theory of space–time coding [16]. For complex constellations, space–time block codes can be constructed for any number of transmit antennas, and again these codes have remarkably simple decoding algorithms based only on linear processing at the receiver. They provide full spatial diversity and half of the maximum possible transmission rate allowed by the theory of space–time coding. For complex constellations and for the specific cases of three and four transmit antennas, these diversity schemes were improved to provide 3/4 of the maximum possible transmission rate [20], [21]. The purpose of this paper is to evaluate the performance of the space–time block codes constructed in [20] and [21] and to provide the details of the encoding and decoding procedures.

Tarokh, Jafarkhani and Calderbank [3] provided an example of space–time block codes transmission using multiple transmit antennas. This paper describes both encoding and decoding algorithms of STBC. The encoding and decoding of these codes have very little complexity. Simulation results were provided to demonstrate that significant gains could be achieved by increasing the number of transmit antenna with very little decoding complexity. It explains analytically that, data is encoded using a space–time block code, and the encoded data is split into n streams which were simultaneously transmitted using n transmit antennas. The received signal at each receive antenna is a linear superposition of the n transmitted signals perturbed by noise. Maximum Likelihood (ML) decoding is achieved in a simple way through decoupling of the signals transmitted from different antennas rather than joint detection. This uses the orthogonal structure of the space–time block code and gives a maximum likelihood decoding algorithm which is based only on linear processing at the receiver. It also reviews the encoding and decoding algorithms for various codes and provide simulation results demonstrating their performance. It is shown that using multiple transmit antennas and space–time block coding provides remarkable performance.

In 1998 a new transmit diversity scheme was been presented by Alamouti [6]. This paper showed that, using two transmit antennas and one receive antenna, the new scheme provides the same diversity order as MRRC with one transmit and two receive antennas. It is further shown that the scheme may easily be generalized to two transmit antennas and receive antennas to provide a diversity order of 2M. An obvious application of the scheme is to provide diversity improvement at all the remote units in a wireless system, using two transmit antennas at the base stations instead of two receive antennas at all the remote terminals. The scheme does not require any feedback from the receiver to the transmitter and its computation complexity is similar to MRRC. When Alamouti scheme was compared with MRRC, if the total radiated power is to remain the same, the transmit diversity scheme found to have 3-dB disadvantage because of the simultaneous transmission of two distinct symbols from two antennas. Otherwise, when the total radiated power was doubled, its performance is identical to MRRC. Moreover, assuming equal radiated power, the scheme requires two half-power amplifiers compared to one full power amplifier for MRRC, which could be advantageous for system implementation. This new scheme also required twice the number of pilot symbols for channel estimation when pilot insertion and extraction was used.

OFDM

Inter-carrier interference (ICI) occurs in an orthogonal frequency division multiplexing (OFDM) system when a channel is rapidly time-varying. K. Kim et al. [22] employed a scheme called LCF-PCC-SFBC OFDM a polynomial cancellation code (PCC) to suppress this ICI, and uses a linear complex field (LCF) code as a transmit-diversity technique to make the transmission rate compatible up to nT / (nT + 1), where nT is the number of transmit antennas used. The conventional orthogonal code (OC), such as the Alamouti code, with accompanying PCC could achieve this rate. Both analysis and simulation results verified that the scheme can achieve higher channel capacity as well as lower bit error rate

(BER) than the other schemes in rapidly time-varying channels, even with a linear receiver and inaccurate channel estimation.

OFDM significantly reduces receiver complexity by providing orthogonal sub-channels. A drawback of OFDM systems is the performance rate reduction due to the cyclic prefix overhead. In year 2003 Alireza Tarighat and Ali H. Sayed [23] proposed receiver structures that exploit the cyclic prefix to increase the performance of the link. The proposed structures use the standard OFDM transmitter, and the modifications are only made at the receiver. Optimum receivers in both the least-mean-squares (LMS) and least-squares (LS) senses are presented, and they do not result in any extra processing complexity compared to the standard OFDM receiver. This architecture is further extended to a MIMO OFDM structure. Simulation results validated the improved performance of the proposed receiver.

SFBC-OFDM

The paper published by Gerhard Bauch [9] is focused on 4G OFDM system, were the elements of the orthogonal design were distributed as space-time block code over successive OFDM symbols which might cause problems in fast fading or as space-frequency block code over neighboring subcarriers which causes problems in severe frequency-selective channels. The paper analyses the suitability and performance of both schemes under 4G conditions and a proposed space-time-frequency mapping which could be a good compromise in terms of performance and delay. It is analyzed the suitability of orthogonal space-time block codes and space-frequency block codes in a 4G OFDM system. While even for high vehicular speed channel variations in time do not significantly degrade the performance of space-time block codes, severe frequency-selectivity is shown to limit the performance of space-frequency block codes unless a very high number of subcarriers were used. It also proposed a space-time-frequency mapping which is more suitable than space-frequency block codes and has lower detection delay than space-time block codes. In contrast to other space-time-frequency codes the scheme retains the simplicity of orthogonal designs. Adaptation to the channel conditions is achieved by a simple modification of the mapping of elements of the orthogonal design to subcarriers and OFDM symbols.

Transmitter diversity technique for wireless communications over frequency selective fading channels was presented K. F. Lee and D.B Williams [10]. This technique utilizes orthogonal frequency division multiplexing (OFDM) to transform a frequency selective fading channel into multiple flat fading sub-channels on which space-frequency processing is applied. Simulation results verified that in slow fading environments the space-frequency OFDM (SF-OFDM) transmitter diversity technique has the same performance as a previously reported space-time OFDM (ST-OFDM) transmitter diversity system but shows better performance in the more difficult fast fading environments. Other implementation advantages of SF-OFDM over the ST-OFDM transmitter diversity technique were also discussed.

This paper mainly focused on two-branch diversity because of its simplicity and its unity coding rate. Higher order SF-OFDM transmitter diversity can be implemented in similar fashion along the frequency dimension. Higher order complex orthogonal block codes all have less than unity coding rate, which results in a reduction in data throughput or an expansion in bandwidth in order to maintain the same data rate. Even if the coding rate loss is acceptable, it is not clear whether using higher order transmitter diversity directly or applying other *error correction codes* (ECC) on top of the second order transmitter diversity system achieved better overall performance. Issue of channel estimation in the SF-OFDM transmitter diversity setting is not addressed in this paper.

Letter by Bertr and Muquet [24] shows a space-frequency block-coded (SFBC) single-carrier frequency-domain equalization (SC-FDE) system. The transmit sequence of the system was designed to have spatial and frequency diversities, and the corresponding combining receiver is derived under a minimum mean square error criterion. It was shown that the system significantly outperforms the space-time block-coded SC-FDE system over fast-fading channels while providing lower computational complexity than orthogonal frequency division multiplexing combined with the SFBC. Therefore to combat fast fading for the SC transmission scheme, system realizes the benefits of both SFBC and SC MMSE-FDE. It has been shown that the system achieves better performance than an STBC SC-FDE system in fast fading environments with increased computational complexity.

In year 2011 a paper was published in IEEE Trans. by M. F. Naeiny and F. Marvasti [25], and introduced *Selected mapping* (SLM) technique for *peak-to-average-power ratio* (PAPR) reduction of orthogonal frequency-division multiplexing systems. In this technique, different representations of OFDM symbols were generated by rotation of the original OFDM frame by different phase sequences, and the signal with minimum PAPR was selected and transmitted. To compensate for the effect of the phase rotation at the receiver, it was necessary to transmit the index of the selected phase sequence as side information (SI). Additionally, a suboptimum detection method that does not need SI was introduced at the receiver side. Simulation results showed that the SLM method effectively reduces the PAPR, and the detection method has performance very close to the case where the correct index of the phase sequence is available at the receiver side.

One important fact is that SLM does not change the orthogonality of space frequency codes. In this method, the same phase sequence is concurrently applied to the frequency-domain signals for both antennas, and the signal with minimum PAPR has been found and transmitted. Optimum ML detection for the transmitted symbols and the phase sequence index has been introduced, and then, a low-complexity suboptimum detector has been introduced to detect the phase sequence index without SI. Simulation results showed that the simplified SLM method effectively reduces the PAPR of SFBC-OFDM system, and the error rate of blind detection of the phase sequence decreases when the number of subcarriers is increased. The detection errors of the method for SNR values higher than 10 and 14 dB were negligible for the OFDM frame lengths of 512 and 128, respectively.

In same year M. F. Naeiny and F. Marvasti published their paper in International Journal of Electronics and Communications [26], which introduced another technique called Active Constellation Extension (ACE) for Peak to Average Power Ratio (PAPR) reduction of OFDM systems. In this technique, the constellation points were extended such that the PAPR is minimized but the minimum distance of the constellation points does not decrease. The method is such that the PAPR is reduced simultaneously at all antennas, while the spatial encoding relationship still holds. In this method, at each iteration, the antenna with maximum PAPR is selected and the ACE algorithm is applied; then the signal of the other antennas was constructed from the resultant signal using SFBC relationship. Simulation results shows that the algorithm converges and its performance is very close to the performance of the ACE in single antenna systems and it outperforms the alternative methods for PAPR reduction of SFBC–OFDM systems when the number of subcarriers increases.

SFBC-OFDM with SC-Frequency Domain Equalization

[27] Discussed principle and history of OFDM and how OFDM signaling is becoming the basis of a world standard for asymmetric digital subscriber line (ASDL) services. In continuation channel estimation and frequency domain equalization is discussed. This paper also showed that, provided it employs a frequency-domain equalizer, SC transmission can handle the same type of channels as does OFDM signaling. In the absence of channel coding, it even substantially outperforms OFDM signaling, the latter technique requires powerful channel coding and frequency-domain interleaving to recover its performance loss with respect to SC transmission. With coding, interleaving, and weighted decoding, OFDM signaling eventually surpasses the performance of SC transmission with frequency-domain equalization, but it suffers from strong sensitivity to nonlinear distortion and carrier synchronization difficulties. Both techniques may be regarded as strong potential techniques for digital terrestrial TV broadcasting, and they can efficiently compensate for multipath channels with long echo delays.

Falconer et al. [28] in April 2002 discussed that small subscriber in non-LOS environments may be subject to severe time dispersion, spanning many bit intervals. For such environments, modulation and equalization strategies based on frequency domain processing should be considered when attempting to achieve adequate anti-multipath performance with reasonable complexity. Frequency domain processing is the basis for OFDM, and it also applies equally well to single carrier modulation. However, OFDM is very sensitive to power amplifier nonlinearities and frequency offsets. SC modulation, coupled with linear frequency domain equalization at the receiver (SC-FDE), has less sensitivity to transmitter nonlinearities and phase noise than OFDM, and its complexity and performance are similar to those of OFDM. Furthermore, the performance of SC-FDE is enhanced when it is combined with simple sparse time-domain decision feedback equalization. Also, as we have seen, SC and OFDM systems can potentially coexist for mutual benefit and cost reduction, because of the obvious similarities in their basic frequency domain signal processing functions. Lastly, single carrier techniques can easily be combined with multiple-input, multiple-output (MIMO) techniques, in which both transmitting and receiving ends use arrays of antenna elements. MIMO techniques can potentially achieve enormous spectral efficiencies (b/s/Hz), limited only by the number of diversity antenna elements that can be implemented practically. This, in turn, relieves the delay-spread issues, since the desired bit rate is achieved without increasing the symbol rate.

Fading Models and SFBC-OFDM

For frequency-selective Nakagami- fading channels, Kang et al. [29] have shown the magnitudes of the channel frequency responses to be also Nakagami-m distributed RVs whose fading and mean power parameters are explicit functions of those of the channel impulse responses. Based on the new results, one can design and evaluate OFDM systems over Nakagami-m fading channels in the frequency domain. In this letter, the BER performance of a multiple receive antenna OFDM system with BPSK and QFSK signals over correlated frequency-selective Nakagami-m fading channels has been evaluated.

High mobility causes performance-limiting ICI in MIMO OFDM receivers. S. Lu et al. [30] designed a novel SFBC-OFDM scheme and showed how to integrate it with FIR ICI-mitigating equalization. To compute the equalizer coefficients, they showed how to estimate the fast time-varying MIMO channel matrix efficiently by exploiting its sparse and banded structure in the time and frequency-domains, respectively. Their simulation results for the Digital Video Broadcasting-Handheld (DVB-H) system demonstrate the effectiveness of the proposed scheme.

Kun-Wah Yip and Tung-Sang Ng [31] proposed a simulation model for m fading channels, $m < 1$. The model reveals that an m fading channel, $m < 1$, having a given Doppler power spectrum can be simulated by generating a complex Gaussian process and a square-root-beta random process. Application of the proposed model to simulate correlated m fading channels has also been demonstrated. In numerical examples, it has been shown that statistical properties of the samples generated from the use of the proposed model are close to the required ones.

[32] This paper presents a general fading distribution – the α-μ Distribution – that includes the Nakagami-m and the Weibull as special cases. Therefore, One-Sided Gaussian, Rayleigh, and Negative Exponential distributions are also special cases of the α-μ Distribution. It is noteworthy that, although the α-μ Distribution has one parameter more than Nakagami-m or Weibull distributions, no additional difficulty is posed by the increase in the number of parameters. The flexibility the α-μ Distribution conveys is outstanding and renders it suited to better adjust to field data. The fading model described here allows one to develop higher order statistics (envelope derivative, etc.) of the α-μ Distribution, a research currently under course.

The Nakagami fading by using sum of sinusoidal using Rayleigh and Ricean fading as given in [33]. The received signal for Rayleigh can be expressed as

$$R_{nakagami} = R_{ray}e^{(1-m)} + R_{rice}(1 - e^{(1-m)}) \qquad (2.1)$$

Where R_{ray} and R_{rice} are envelops of Rayleigh and Ricean channels respectively. Their method of generation is given in [33]. By adopting sum of sinusoidal approach for received signal generation and expressing it in in-phase and quadrature form for both Rayleigh and Ricean fading.

$$R(t) = \sqrt{I(t)^2 + Q(t)^2} \qquad (2.2)$$

Where I(t) and Q(t) are the in-phase and quadrature components.

CHAPTER 3

DIVERSITY AND ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING

Wireless transmission uses air or space for its transmission medium. The radio propagation is not as smooth as in wire transmission since the received signal is not only coming directly from the transmitter, but the combination of reflected, diffracted, and scattered copies of the transmitted signal.

- Diffraction occurs when the signal is obstructed by a sharp object which derives secondary waves.
- Scattering occurs when the signal impinges upon rough surfaces, or small objects.

Therefore receiver can receive more than one copies of the signal in multiple paths with different phases and powers. There are key modeling parameters that help to characterize a wireless channel: path loss, delay spread, coherence bandwidth, Doppler spread, and coherence time. Diversity is the key topic in mobile communications. It is principally used to combat fading. The basic idea is that if different copies of the same signal are available then there is a high probability that at least one of them is of a good quality. Of course choosing the best copy and rejecting all others is not the optimal solution. This brings us to the problem of choosing the best way to combine all of them [34].

3.1 Fading

3.1.1 Path Loss

The average received power diminishes with the distance. In free space when there is a direct path between transmitter and receiver and when there are no secondary waves from the medium objects, the received power is inversely proportional to square of the carrier frequency and square of the distance. If P_R and P_T are the received and transmitted powers respectively, the following relation describes free space propagation [35]

$$P_R \ \alpha \ \frac{G.P_T}{f^2 d^\alpha} \tag{3.1}$$

Where f is the carrier frequency, d is the distance between transmitter and receiver, *G* is the power gain from the transmit and receive antennas, α=2 is the path loss component and path loss is defined as P_T/P_R.

3.1.2 Shadowing

Signal power attenuates randomly with distance when the medium includes obstructions. If the two locations have different surroundings then the variations in the signal should be different. This behavior is called *shadowing*. Measurements indicate that the path loss is random and distributed log-normally. Power received for indoor path loss is called log normal shadowing and following is the formula:

$$PL(dB) = PL(d_0) + 10\ \alpha.\log_{10}(d/d_0) + X_\Omega, \qquad (3.2)$$

Where d0 is the reference distance and X_Ω is a zero mean Gaussian fandom variable with a standard deviation Ω.

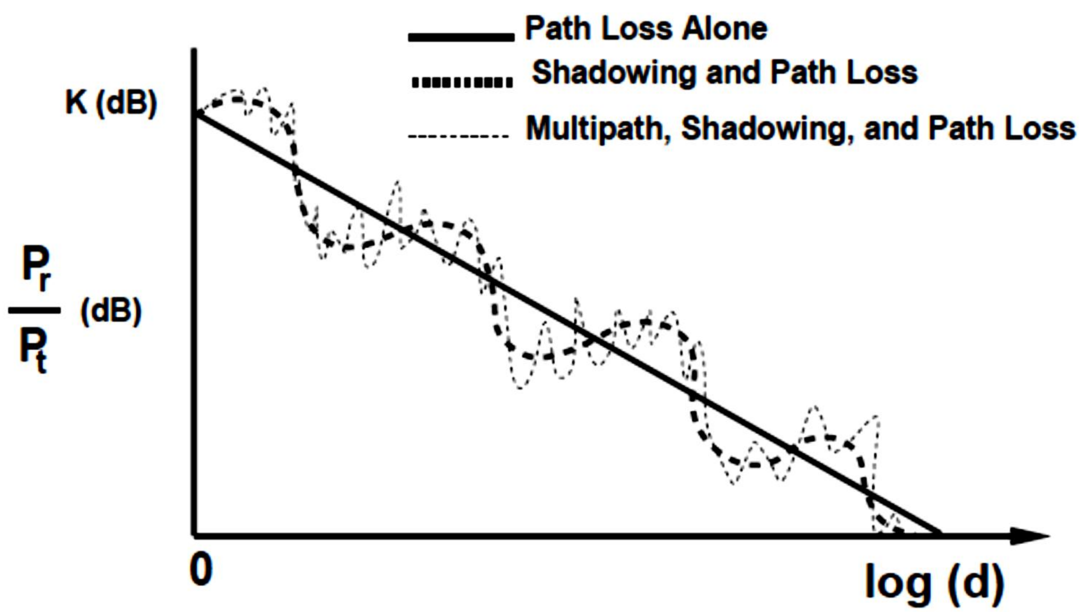

Fig. 3.1.2: Received signal to transmitted power ratio versus distance [35]

A typical situation is shown in figure 3.1.2 is useful to understand the difference between path loss, shadowing and flat fading, which is described in the following section.

3.1.3 Fading Parameters

Fading is present when there are multipath components. Multipath components arrive at the receiver at slightly different times. If there is movement in the system then there is also phase difference between the received components which leads to shift in the frequency. These multiple received copies apply constructive or destructive interference to the signal and create a standing wave pattern that shows rapid signal strength changes, frequency shifts or echoes.

Multipath delay nature of the channel is quantified by *delay spread* and *coherence bandwidth.* The time-varying nature of the channel caused by movement is quantified by *Doppler spread* and *coherence time.*

- *Delay Spread:*

The reflected parts arrive later than the original signal to the receiver. RMS delay spread (σ_τ) is the metric to characterize this delay in terms of second order moment of the channel power profile. Typical values are on the order of microseconds in outdoor and on the order of nanoseconds in indoor radio channels [36]. Depending on the symbol duration (T_s), σ_τ plays an important role to find out how the signal is treated by the channel.

- *Coherence Bandwidth:*

The channel power profile is coupled with channel frequency response through Fourier transform. Coherence bandwidth (B_c) is used to quantify the channel frequency response and it is inversely proportional to RMS delay spread. Coherence bandwidth is used to measure how flat the channel bandwidth is. Flatness is described as the close correlation between the two frequency components. This is important since a signal having a larger bandwidth (B_S) than (B_c) is severely distorted. For a 0.9 correlation $Bc \approx 1/50\sigma_\tau$ [36].

- *Doppler Shift:*

Movement causes shift in the signal frequency. When the stations are moving, the received signal frequency is different than the original signal frequency. Doppler shift is defined as the change in the frequency.

Doppler shift is given by

$$f_d = \frac{1}{2\pi}\frac{\Delta\Phi}{\Delta t} = \frac{v}{\lambda}.cos\theta \quad (3.3)$$

We can see that if the movement is toward the signal generator, the Doppler shift is positive otherwise it is negative [36].

- *Doppler Spread:*

A channel shows a time varying nature when there is a movement in either the source or destination or even objects in the middle. Doppler spread (B_D) is the measure of maximum broadening of the spectrum due to Doppler shift. Thereby B_D is $f_m=v/\lambda$ where f_m is the maximum Doppler shift.

- *Coherence Time:*

The time domain dual of Doppler spread is coherence time (T_c) where $T_c=1/f_m$ Coherence time identifies the time period wherein two received signal have high amplitude correlation.

- *Coherence Distance*

Coherence distance gives a measure of a minimum distance between points in space for which the signal are uncorrelated. This distance is important for multiple antenna systems. It is 0.5 wavelengths for wide beam-width receive antennas and about 10 and 20 wavelengths for low-medium and high BTS antenna heights, respectively [37].

3.1.4 Flat Fading

Delay spread and coherence bandwidth is independent from Doppler spread and coherence time. The signal goes to *flat fading* if the channel bandwidth is greater than the signal bandwidth $(B_c>B_s)$ Spectral shape of the signal remains but the gain changes. Narrow band signals fall into this category since their bandwidth is small as compared to the channel bandwidth. Flat fading channels bring challenges such as variation in the gain and in the frequency spectrum. Distortion in the gain may cause deep fades thereby requiring significant increase in the power in some frequencies. Destructive interferences may cause deep nulls in the signal power spectrum which is a particular problem for narrow band signals since any null in a frequency may cause loss of the signal. There are various ways to reduce the fading distortion [38]. Diversity is one way which leverages multiple independent channels. Since the channels are independent they have lower probability to experience fades at the same time. Space diversity is one of the most efficient diversity technique compared to time or frequency. Multiple slightly separated antennas in space are used to create independent paths to achieve uncorrelated channels. Coding is another way which spreads the signal power among the bandwidth and lessens the power allocated in the null frequencies. As a result, loss occurs in the smaller portion of the signal power. Another method is chopping the bandwidth into small bandwidths as in *orthogonal frequency division multiplexing* (OFDM). In each subcarrier small portion of the code is sent. Thus, if there is a loss in a subcarrier then it is recoverable since it corresponds to the small part of the code.

3.1.5 Frequency Selective Fading

If channel bandwidth is smaller than the signal bandwidth then the channel creates frequency selective fading. Spectral response of a radio signal will show dips due to the multipath. Channel bandwidth limits the signal bandwidth which leads to overlapping of the successive symbols in the time domain due to the convolution $(\sigma_\tau > T_s)$ Overall symbol duration becomes more than the actual symbol duration. This phenomenon is called *inter symbol interference* (ISI).

There are several ways to combat ISI [38]. Equalization is one way that tries to invert the effects of the channel. In OFDM, each subcarrier undergoes flat fading since its bandwidth is less than the coherence bandwidth. Therefore ISI is eliminated within an OFDM symbol but OFDM symbols may overlap in time. A portion of the OFDM symbol, larger than the coherence bandwidth, is appended to the symbol to overcome the ISI between symbols. Equalization is simple in OFDM since each subcarrier only needs a one tap equalizer

3.1.5.1 Fast Fading

Channel bandwidth may also change due to movement. Fast fading occurs if the coherence time is smaller than the symbol duration of the signal $(T_s > T_c)$. The channel becomes time varying and within a symbol duration rapid changes occur in impulse response of the channel. Low data rates is vulnerable to fast fading since the symbol duration is wide in low data rates

3.1.5.2 Slow Fading

If the signal bandwidth is much higher than the Doppler spread $(B_c > B_D)$ then the shift in the frequency due to Doppler shift is insignificant. This is called slow fading and the channel impulse response changes slowly.

3.2 Different Fading Models

3.2.1 Rayleigh Fading

Constructive and destructive nature of multipath components in flat fading channels can be approximated by Rayleigh distribution if there is no line of sight which means when there is no direct path between transmitter and receiver. The Rayleigh distribution is basically the magnitude of the sum of two equal independent orthogonal Gaussian random variables and the *probability density function* (PDF) is given in by

$$p(r) = \frac{r}{\sigma^2} e^{-\frac{r^2}{2\sigma^2}} \quad 0 \leq r \leq \infty \tag{3.4}$$

where σ^2 is the time-average power of the received signal [36].

3.2.2 Ricean Fading

When there is line of sight, direct path is normally the strongest component goes into deeper fade compared to the multipath components. This kind of signal is approximated by Ricean distribution. As the dominating component run into more fade the signal characteristic goes from Ricean to Rayleigh distribution. The Ricean distribution is given by

$$p(r) = \frac{r}{\sigma^2} e^{-\frac{(r^2+A^2)}{2\sigma^2} I_0 \frac{Ar}{\sigma^2}} \text{ for } (A \geq 0, r \geq 0) \tag{3.5}$$

Where A denotes the peak amplitude of the dominant signal, I(.) is the modified Bessel function of the first kind and zero-order [36].

3.2.3 Gamma Distribution

The probability density function of the gamma distribution can be expressed in terms of the gamma function parameterized in terms of a shape parameter *k* and scale parameter θ. Both k and θ will be positive values. The equation defining the PDF and CDF of a gamma-distributed random variable X are given as

$$f_x(x) = \frac{(x/\theta)^{k-1}\exp(-x/\theta))}{\theta\Gamma(k)}u(x) \quad \text{for } k, \theta > 0 \tag{3.6}$$

$$F_x(x) = \frac{P(k, x/\theta)}{\Gamma(k)}u(x) \tag{3.7}$$

In the above equations, the gamma function is a generalization of the factorial function defined by

$$\Gamma(\alpha) = \int_0^\infty e^{-t}t^{\alpha-1}dt \tag{3.8}$$

And the incomplete gamma function is given by

$$P\ (\alpha, \beta) = P(\alpha,\beta) = \int_0^\beta e^{-t}t^{\alpha-1}dt \tag{3.9}$$

3.2.4 Nakagami-m Distribution

The Nakagami distribution or the Nakagami-*m* distribution is a probability distribution related to the gamma distribution. It has two parameters: a shape parameter μ and a second parameter controlling spread, ω. The Nakagami-*m* distribution having the PDF of the form

$$f(x, \mu, \omega) = \frac{2\mu^\mu}{\Gamma(\mu)\omega^\mu}x^{2\mu-1}e^{\frac{\mu}{\omega}x^2} \qquad \text{for } x \geq 0 \tag{3.10}$$

Its CDF is given by

$$F(x, \mu, \omega) = P(\mu, \frac{\mu}{\omega}\ .x^2) \tag{3.11}$$

Where *P* is the incomplete gamma function which is defined in (N.4)

In addition to the m-distribution, the following compact form of distribution is presented by Nakagami (1940) and is named *n-distribution,*

$$P(R) = \frac{2R}{\sigma} e^{\left(\frac{R^2+R_0^2}{\sigma}\right)} I_0\left(\frac{2RR_0}{\sigma}\right) \tag{3.12}$$

and

$$P(R) = \frac{2R}{\alpha\beta} e^{\left(\frac{R^2}{2}\right)\left(\frac{1}{\alpha}+\frac{1}{\beta}\right)} . I_0\left(\frac{R^2}{2}\left(\frac{1}{\beta}-\frac{1}{\alpha}\right)\right) \tag{3.13}$$

is called *q-distribution.*

3.3 Diversity Techniques in Fading Environment

Diversity techniques are important to cope with fading environment as well as coding. It tries to provide the receiver with multiple fade replicas of the same information bearing signal. If there are L independent replicas and if the denote the outage probability then overall probability is p^L which is less than p Diversity gain is at the highest when the diversity branches are uncorrelated. Some of the techniques are listed below.

- **Frequency Diversity:** Multiple channels are separated by more than the coherence bandwidth. Using RAKE receiver, OFDM, equalization are possible techniques for frequency diversity. It cannot be used over frequency-flat channel.
- **Time Diversity:** Channel coding, interleaving are possible techniques but it is not effective over slow fading channels.
- **Angle/Direction Diversity:** Antennas with different polarizations/directions can be used for reception and transmission.
- **Space Diversity:** Using multiple antennas for space diversity is efficient, coherent distance plays an important role for the selection of places of the antennas.

Out of above given diversity techniques space diversity is explained in subsequent chapter

Note that not all kinds of diversity are always feasible. For example a slowly fading channel (with a long coherence time) cannot support temporal diversity with practical interleaving depths. Similarly, frequency diversity is not feasible when the coherence bandwidth of the channel is comparable to the bandwidth of the signal employed. However, irrespective of the channel characteristics, antenna diversity can always be exploited as long as there is sufficient spacing between the antennas.

Temporal and frequency types of diversity normally introduce redundancy in time and/or frequency domain, and therefore induce loss in bandwidth efficiency. Typical examples of spatial diversity are multiple transmit and/or receiver, multiple antenna communication rely on space diversity to mitigate fading without necessarily sacrificing precious bandwidth resources; thus, they become attractive solution for broadband wireless application. Compare to single antenna transmission, multiple antenna transmissions increase the channel capacity.

A simple space diversity scheme is to use multiple antennas at the receiver. This scheme does not involve loss of band width. The optimal way to combine the outputs of different antennas is the Maximal Ratio combining. At the same time remote unit are supposed to be small light weighted pocket communicators. Inevitably, the pocket communicators must remain simple. Also, antenna diversity at a mobile handset is more difficult to implement because of electromagnetic interaction of antenna elements on small platforms and the expense of multiple down conversion RF paths. Furthermore, the channels corresponding to different antennas are correlated, with the correlation factor determined by the distance as well as the coupling between the antennas. Typically, the second antenna is inside the mobiles handset, resulting in signal attenuation at the second antenna. This can cause some loss in diversity benefit. All these factors motivate the use of multiple antennas at the base station for transmission [24]. This method is called transmit diversity.

Two major obstacles to implement transmit antenna diversity are as follows:

i. Unlike the receiver, the transmitter does not have instantaneous information about the fading channels.
ii. The transmitted signals are mixed spatially before they arrive at the receiver.

A number of transmit antenna diversity scheme have been proposed, and can be divided into two categories: open loop and closed loop. The difference between open and closed loop schemes is that the former does not require channel knowledge at the transmitter. On the other hand, the latter relies on some channel information at the transmitter that is acquired through feedback channels. Although feedback channels are present in most wireless systems (for power control purposes,) mobility may cause fast channel variation. As a result, the transmitter may not be capable of acquiring and tracking the channel variations. Thus, usage of open loop transmit antennas diversity schemes is well motivated for future broadband wireless system, which are characterized by high mobility.

Transmit diversity can be classified into the following categories based on the availability of the channel information at the transmitter and receiver:

- Schemes with feedback and feed forward information (giving complete channel state information to both the transmitter as well as receiver).
- Schemes with feed forward information but no feedback (receiver has the channel state information but the transmitter does not have any information.
- Schemes without any channel state information at the transmitter or the receiver.

First category uses, implicit or explicit feedback of information from the receiver to the transmitter to configure the transmitter. In time division duplex systems the same antenna weights are used for reception and transmission, so feedback is implicit in the appeal to channel symmetry. These weights are chosen during reception to maximize the SNR and during transmission to weight the amplitudes of the transmitted signals. Explicit feedback includes switched diversity with feedback based on the feedback of the channel response. However, in particular, vehicle movements or interference causes a mismatch between the state of channel perceived by transmitter and that perceived by receiver. Second category is with feed forward without feedback technique. This technique uses linear processing at the transmitter to spread the information across the antennas. At the receiver, information is obtained by either linear processing or maximum likelihood decoding techniques. Feed forward information, is required to estimate the channel from the transmitter to receiver. These estimates are used to compensate for the channel response at the receiver. This scheme was first proposed by Wittneben [39].

The third category does not require feedback or feed forward information. Instead, it uses multiple transmit antenna combined with channel coding to provide diversity. Example of this approach is to combine phase sweeping transmitter diversity of [40] with channel coding.

Another scheme is to encode information by a channel code and transmit the code symbols using different antennas in orthogonal manner. This can be done either by frequency multiplexing [41], time multiplexing or by using orthogonal spreading sequences for different antenna. The disadvantage of these schemes over previous two schemes is the loss in bandwidth efficiency due to the use of channel code. Using appropriate coding, it is possible to relax the orthogonality requirement need in these schemes and obtain the diversity as well as coding advantage offer without sacrificing bandwidth. This is possible when the whole system is viewed as a MIMO system and suitable codes are used.

In transmit diversity multiple antennas are used to achieve space diversity and to achieve time diversity the transmission is repeated *n* times. Scope of diversity is expanded by using multiple antennas at both the receiver and transmitter. A major conclusion of these works is that the capacity of multi-antenna systems far exceeds that of a single antenna system. In particular, the capacity grows at least linearly with the number of transmit antennas as long as number of received antenna is greater than or equal to number of transmit antennas [24].

3.4 Space-Time Codes

Space-time Codes (STC) were first introduced by Tarokh et al. from AT&T research labs [2] in 1998 as a novel means of providing transmit diversity for the multiple antenna fading channel. There are two main types of STC:

a) Space-time Trellis Codes (STTC)
b) Space-time Block Codes (STBC)

The original STTC were introduced by Tarokh et al. in [2], there has been extensive research aiming the improving the performance of the original STTC design. These original STTC design were hand crafted and therefore, are not optimum design. In the recent years, a large number of research proposals have been published which propose new code construction or perform systematic searches for different convolution STTC or some variant of the original design criteria proposed by Tarokh et al.

When the number of antennas is fixed, the decoding complexity of space-time trellis coding (measured by the number of trellis states at the decoder) increases exponentially as a function of diversity level and transmission rate [24]. In addressing the issue of decoding complexity, Alamouti [6] discovered a remarkable space-time block coding for transmission with two antennas. This scheme supports maximum likelihood (ML) detection based only on linear processing at the receiver. The very simple structure and linear processing of the Alamouti construction makes it a very attractive scheme that is currently part of both the W-CDMA and CDMA-2000 standards [42]. Key advantage of STTC over STBC is the provision of coding gain. Their disadvantage [43] is that they are extremely hard to design and generally require high complexity encoders and decoders.

Space-time block codes operate on a block of input symbols, producing a matrix output whose columns represent time and rows represents antennas. It contrast to single antenna block codes for the AWGN channel, space-time block codes do not generally provide coding gain, unless concatenated with an outer code. Their main feature [44] is the provision of full diversity with a very simple decoding scheme. On the other hand, space-time trellis codes operate on one input symbol at a time, producing a sequence of vector symbols whose length represents antennas.

Both STTC and STBC were first designed assuming a narrowband wireless system, i.e. a flat fading channel. However, when used over frequency selective channels a channel equalizer has to be used at the receiver along with the space-time decoder. Initial attempts to address the problem for STTC made use of whatever structure was available in the space-time coded signal [6] [45], where the structure of the code was used to convert the problem into one that can be solved using known equalization schemes. For the STBC, the channel equalization problem was addressed by modifying the original Alamouti scheme in such a way that the use over frequency selective channel, and hence the equalization, is a much easier task. For example, in [46], STBC was used in conjunction with OFDM. OFDM is used to convert the frequency selective channel into a set of independent parallel frequency-flat sub-channels. The Alamouti scheme is then applied to two consecutive subcarriers (or two consecutive OFDM block) [47].

Recently, Orthogonal Frequency Division Multiplexing (OFDM) has emerged as a technology for high data rates. In particular, many wireless standards (Wi-Max, IEEE802.11a, LTE, DVB) have adopted the OFDM technology as a mean to increase future wireless communications. OFDM is a form of Multi-carrier transmission and is suited for frequency selective channels and high data rates. This technique transforms a frequency-selective wide-band channel into a group of non-selective narrowband channels, which makes it robust against large delay spreads by preserving orthogonality in the frequency domain.

3.5 Frequency Selective Channel

In this section, we first give a brief overview of frequency selective channels.
Let $r(t)$ be the low-pass received signal:

$$r(t) = \int_{-\infty}^{\infty} c(\tau)x(t-\tau)d\tau + n(t) \tag{3.14}$$

Frequency selectivity occurs whenever the transmitted signal $x(t)$ occupies an interval bandwidth [-W/2 , +W/2] greater than the coherence bandwidth B_{coh} of the channel $c(t)$ (defined as the inverse of the delay spread T_d). figure 3.5 presents a typical time impulse response $c(t)$ of a channel. For usual high data rates schemes, the symbol rate T is small

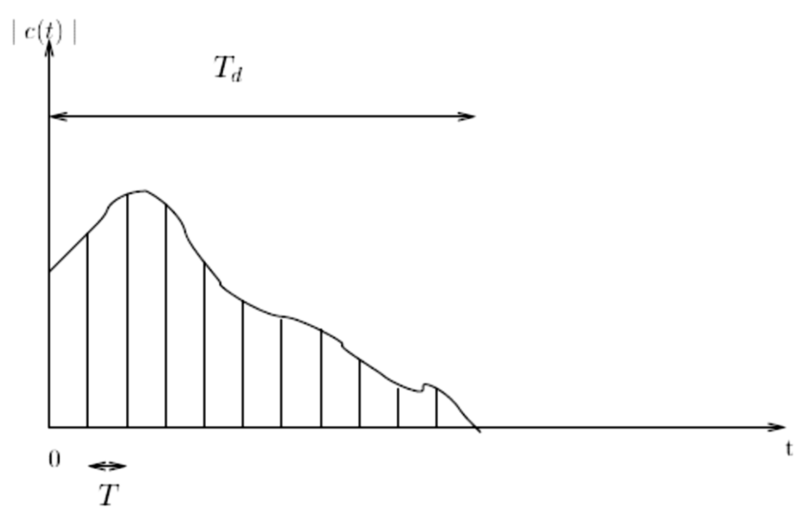

Fig. 3.5: Comparison between delay spread and symbol duration [35]

compared to T_d (they are also called broadband signals) and the signals are therefore subject to frequency selectivity.

3.6 Orthogonal Frequency Division Multiplexing

OFDM dates back some 40 years; a patent was applied for in the mid-1960s [48]. A few years later, an important improvement the *Cyclic Prefix* (CP) was introduced; it helps to eliminate residual delay dispersion. Cimini [49] was the first to suggest OFDM for wireless communications. But it was only in the early 1990s that advances in hardware for digital signal processing made OFDM a realistic option for wireless systems. Furthermore, the high-data rate applications for which OFDM is especially suitable emerged only in recent years. Currently, OFDM is used for *Digital Audio Broadcasting* (DAB), *Digital Video Broadcasting* (DVB), and *wireless Local Area Networks* (LANs) (IEEE 802.11a, IEEE 802.11g). It will also be used in fourth generation cellular systems, including *Third Generation Partnership Project-Long-Term Evolution* (*3GPP-LTE*) and WiMAX.

3.6.1 Principle of Orthogonal Frequency Division Multiplexing

OFDM splits a high-rate data stream into *N* parallel streams, which are then transmitted by modulating *N* distinct carriers (henceforth called *subcarriers* or *tones*). Symbol duration on each subcarrier thus becomes larger by a factor of *N*. In order for the receiver to be able to separate signals carried by different subcarriers, they have to be orthogonal. Conventional Frequency Division Multiple Access (FDMA) is depicted in figure 3.6.1, can achieve this by having large (frequency) spacing between carriers. This, however, wastes precious spectrum A much narrower spacing of subcarriers can be achieved. Specifically, let subcarriers be at the frequencies $f_n = nW/N$, where *n* is an integer, and *W* the total available bandwidth; in the most simple case, $W = N/Ts$. We furthermore assume for the moment that modulation on each of the subcarriers is Pulse Amplitude Modulation (PAM) with rectangular basis pulses. We can then easily see that subcarriers are mutually orthogonal, since the relationship holds good.

$$\int_{iT_s}^{(i+1)T_s} \exp(j2\pi f_k t)\exp(-j2\pi f_n t)dt = \delta_{\mathrm{nk}} \qquad (3.15)$$

Where

$$\delta_{\mathrm{nk}} = \begin{cases} T_s \ \ if\ n = k \\ 0 \ \ otherwise \end{cases}$$

Figure 3.6.1 shows this principle in the frequency domain. Due to the rectangular shape of pulses in the time domain, the spectrum of each modulated carrier has a sin*(x)/x* shape. The spectra of different modulated carriers overlap, but each carrier is in the spectral nulls of all other carriers. Therefore, as long as the receiver does the appropriate demodulation (multiplying by $\exp(-j2\pi f_n t)$ and integrating over symbol duration), the data streams of any two subcarriers will not interfere.

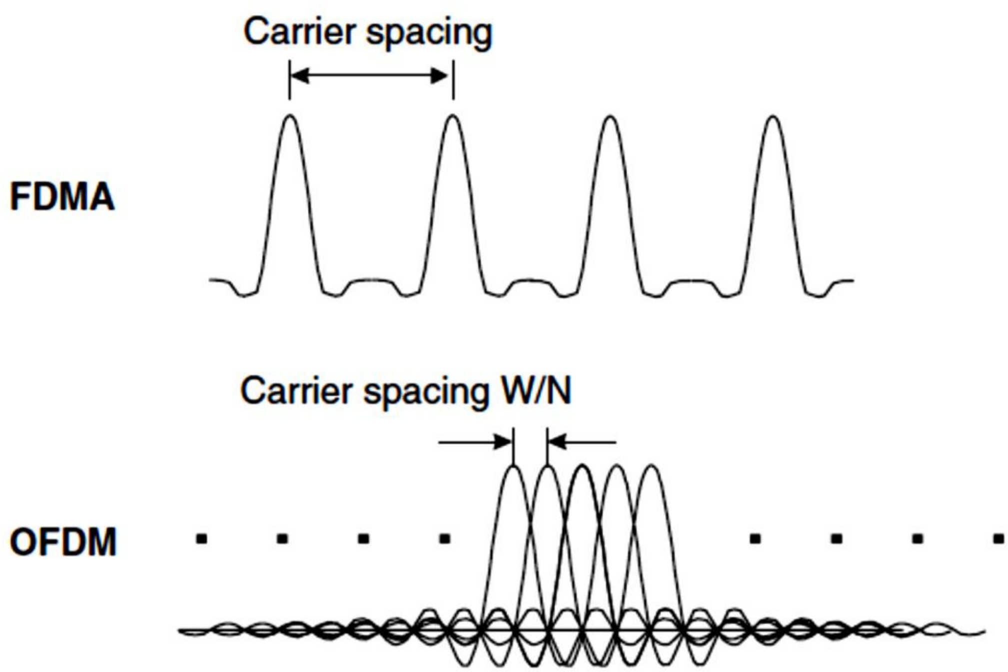

Fig. 3.6.1: FDMA OFDM difference [50]

3.6.2 Effect of Frequency-Selective Channels on OFDM

Intuitively, we would anticipate that delay dispersion will have only a small impact on the performance of OFDM we convert the system into a parallel system of narrowband channels, so that the symbol duration on each carrier is made much larger than the delay spread. But, as we know delay dispersion can lead to appreciable errors even when $S\tau/T\text{s} < 1$. Furthermore, as we will elaborate below, delay dispersion also leads to a loss of orthogonality between the subcarriers, and thus to *Inter Carrier Interference* (ICI). Fortunately, both these negative effects can be eliminated by a special type of guard interval, called the *cyclic prefix (CP)*. In this section, we show how to construct this cyclic prefix, how it works, and what performance can be achieved in frequency-selective channels.

When an input data stream $x[n]$ is sent through a linear time-invariant discrete-time channel $h[n]$, the output $y[n]$ is the discrete-time convolution of the input and the channel impulse response:

$$y[n] = h[n] * x[n] = \sum_k h[k]x[n-k]. \quad (3.16)$$

The N-point circular convolution of $x[n]$ and $h[n]$ is defined as

$$y[n] = x[n] \otimes h[n] = h[n] \otimes x[n] = \sum_k h[k]x[n-k]_N \quad (3.17)$$

Where $[n-k]_N$ denotes $[n-k]$ modulo N. In other words, $x[n-k]_N$ is a periodic version of $x[n-k]$ with period N. It is easily verified that $y[n]$ given by (3.17) is also periodic with period N. From the definition of the DFT, circular convolution in time leads to multiplication in frequency ie.,

$$\mathrm{DFT}\{x[n] \otimes h[n]\} = X[i]H[i],\ 0 \le i \le N-1. \quad (3.18)$$

By (3.18), if the channel and input are circularly convoluted then if $h[n]$ is known at the receiver, the original data sequence $x[n]$ can be recovered by taking the IDFT of $Y[i]/H[i]$, $0 \le i \le N-1$. Unfortunately, the channel output is not a circular convolution but a linear convolution. However, the linear convolution between the channel input and impulse response can be turned into a circular convolution by adding a special prefix to the input called a cyclic prefix (CP), described in the next section.

3.6.3 The Cyclic Prefix

Consider a channel input sequence $x[n] = x[0], \ldots, x[N-1]$ of length N and a discrete-time channel with finite impulse response (FIR) $h[n] = h[0], \ldots, h[\mu]$ of length $\mu + 1 = T_m/T_s$, where T_m is the channel delay spread and T_s the sampling time associated with the discrete time sequence. The cyclic prefix for x[n] is defined as $\{x[N-\mu], \ldots, x[N-1]\}$: it consists of the last μ values of the x[n] sequence. For each input sequence of length N, these last μ samples are appended to the beginning of the sequence. This yields a new sequence $\tilde{x}[n], -\mu \leq n \leq N-1$, of length N+μ, where $\tilde{x}[-\mu], \ldots, \tilde{x}[N-1] = x[N-\mu], \ldots, x[N-1], x[0], \ldots, x[N-1]$. Note that with this definition, $\tilde{x}[n] = x[n]_N$ for $-\mu \leq n \leq N-1$, which implies that $\tilde{x}[n-k] = x[n-k]_N$ for $-\mu \leq n-k \leq N-1$.

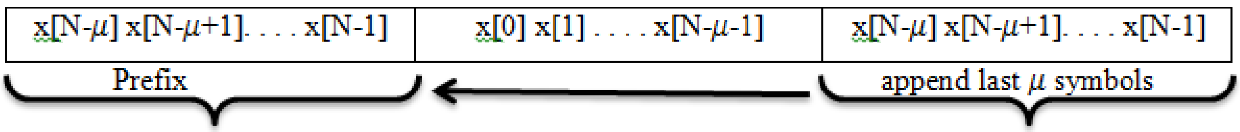

Fig. 3.6.3: Insertion of CP of length μ symbols.

Suppose $\tilde{x}[n]$ is input to a discrete-time channel with impulse response h[n]. The channel output y[n], $0 \leq n \leq N-1$ is then

$$y[n] = \tilde{x}[n] * h[n] \quad (3.19)$$

$$= \sum_{k=0}^{\mu} h[k]\tilde{x}[n-1] \quad (3.20)$$

$$= \sum_{k=0}^{\mu} h[k]x[n-1]_N \quad (3.21)$$

$$= x[n] \otimes h[n] \quad (3.22)$$

Thus, by appending a cyclic prefix to the channel input, the linear convolution associated with the channel impulse response y[n] for $0 \leq n \leq N-1$ becomes a circular convolution.

3.6.4 OFDM Implementation

The OFDM implementation of multicarrier modulation is shown in figure 3.5. The input data stream is modulated by a QPSK modulator, resulting in a complex symbol stream X[0],X[1], . . .,X[N − 1]. This symbol stream is passed through a serial-to-parallel converter, whose output is a set of N parallel QPSK symbols X[0], . . .,X[N−1] corresponding to the symbols transmitted over each of the subcarriers. Thus, the N symbols output from the serial-to-parallel converter are the discrete frequency components of the OFDM modulator output s(t).

In order to generate s(t), these frequency components are first converted into time samples by performing an inverse DFT on these N symbols, which is efficiently implemented using the IFFT algorithm. The IFFT yields the OFDM symbol consisting of the sequence x[n] = x[0], . . . , x[N − 1] of length N, where

$$x[n] = 1/\sqrt{N}\sum_{k=0}^{N-1} X[k]e^{-j2\pi nk/N};\ 0 \le n \le N-1. \quad (3.23)$$

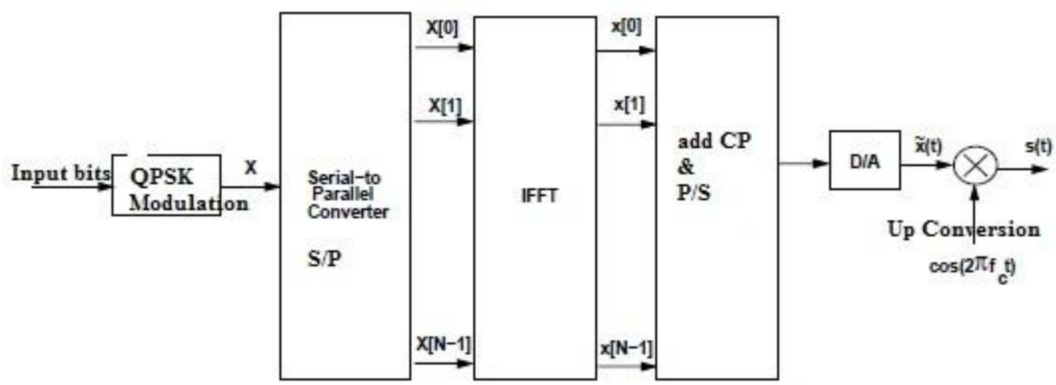

Fig. 3.6.4.1: Transmitter block diagram of OFDM with cyclic prefix

This sequence corresponds to samples of the multicarrier signal: i.e. the multicarrier signal consists of linearly modulated subchannels. The cyclic prefix is then added to the OFDM symbol, and the resulting time samples x̃[n] = x̃[−μ], . . . , x̃[N − 1] = x[N − μ], . . . , x[0], . . . , $x[N-1]$ are ordered by the parallel-to-serial converter and passed through a D/A converter, resulting in the baseband OFDM signal x̃(t), which is then up-converted to frequency f_0.

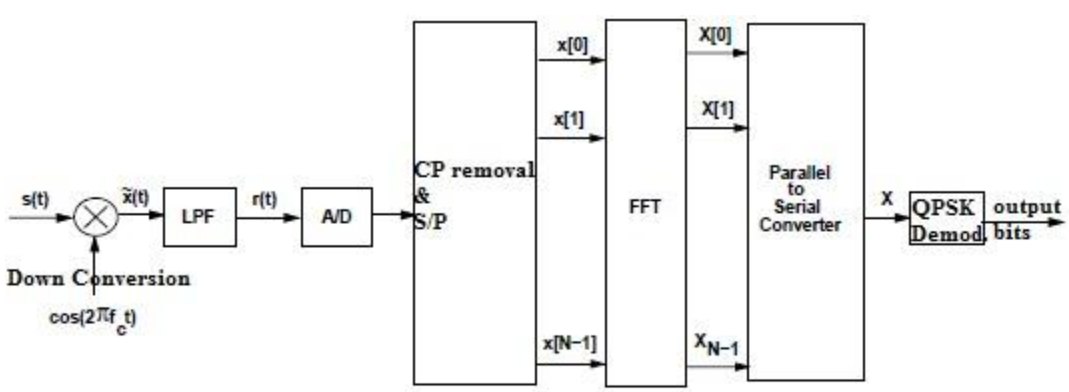

Fig. 3.6.4.2: Receiver block diagram of OFDM with cyclic prefix

Figure 3.6.4.2 shows the receiver of OFDM the transmitted signal is filtered by the channel impulse response $h(t)$ and corrupted by additive noise, so that the received signal is $y(t) = \tilde{x}(t) * h(t) + n(t)$. This signal is down-converted to baseband and filtered to remove the high frequency components. The A/D converter samples the resulting signal to obtain $y[n] = \tilde{x}[n] * h[n] + v[n]$, $-\mu \leq n \leq N - 1$. The prefix of $y[n]$ consisting of the first μ samples is then removed. This results in N time samples whose DFT in the absence of noise is $Y[i] = H[i]X[i]$. These time samples are serial-to-parallel converted and passed through an FFT. This results in scaled versions of the original symbols $H[i]X[i]$, where $H[i] = H(fi)$ is the flat-fading channel gain associated with the i^{th} sub-channel. The FFT output is parallel-to-serial converted and passed through a QPSK demodulator to recover the original data. The OFDM system effectively decomposes the wideband channel into a set of narrowband orthogonal sub-channels with a different QPSK symbol sent over each sub-channel. Knowledge of the channel gains $H[i]$, $i = 0, \ldots, N-1$ is not needed for this decomposition, in the same way that a continuous time channel with frequency response $H(f)$ can be divided into orthogonal sub-channels without knowledge of $H(f)$ by splitting the total signal bandwidth into non overlapping sub-bands. The demodulator can use the channel gains to recover the original QPSK symbols by dividing out these gains: $X[i] = Y[i]/H[i]$. This process is called frequency equalization. However, for continuous-time OFDM, frequency equalization leads to noise enhancement, since the noise in the i^{th} sub-channel is also scaled by $1/H[i]$. Hence, while the effect of flat fading on $X[i]$ is removed by this equalization, its received SNR is unchanged. An alternative to using the cyclic prefix is to use a prefix consisting of all zero symbols. In this case the OFDM symbol consisting of $x[n]$, $0 \leq n \leq N-1$ is preceeded by μ null samples. At the receiver the "tail" of the ISI associated with the end of a given OFDM symbol is added back in to the beginning of the symbol, which recreates the effect of a cyclic prefix, so the rest of the OFDM system functions as usual. This zero prefix reduces the transmit power relative to a cyclic prefix by $N/(\mu+N)$, since the prefix does not require any transmit power. However, the noise from the received tail is added back into the beginning of the symbol, which increases the noise power by $(N+\mu)/N$. Thus, the difference in SNR is not significant for the two prefixes.

3.6.5 Matrix Representation of OFDM

Matric representation is essential so that mathematics of OFDM system may be verified by simulation. Let us Consider a discrete-time channel with FIR $h[n]$, $0 \leq n \leq \mu$, input $\tilde{x}[n]$, noise $v[n]$, and output $y[n] = \tilde{x}[n] * h[n] + v[n]$. Denote the nth element of these sequences as $h_n = h[n]$, $\tilde{x}_n = \tilde{x}[n]$, $v_n = v[n]$, and $y_n = y[n]$. With this notation the channel output sequence can be written in matrix form as

$$\begin{bmatrix} y_{N-1} \\ y_{N-2} \\ \cdot \\ \cdot \\ y_0 \end{bmatrix} = \begin{bmatrix} h_0 & h_1 & . & . & h_\mu & 0 & . & . & 0 \\ 0 & h_0 & . & . & h_{\mu-1} & h_\mu & . & . & 0 \\ . & . & . & . & . & . & . & . & . \\ . & . & . & . & . & . & . & . & . \\ 0 & . & . & . & h_0 & . & . & h_{\mu-1} & h_\mu \end{bmatrix} \begin{bmatrix} x_{N-1} \\ x_{N-2} \\ \cdot \\ \cdot \\ x_0 \end{bmatrix} + \begin{bmatrix} v_{N-1} \\ v_{N-2} \\ \cdot \\ \cdot \\ v_0 \end{bmatrix} \tag{3.23}$$

can also be written in algebraic equation (uppercase bold letter represents matrix and lowercase letters represents vector) form

$$\mathbf{y} = \mathbf{Hx} + \boldsymbol{v} \tag{3.24}$$

The received symbols $y_{-1} \dots y_{-\mu}$ are discarded since they are affected by ISI in the prior data block, and they are not needed to recover the input. The last μ symbols of $x[n]$ correspond to the cyclic prefix: $x_{-1} = x_{N-1}$, $x_{-2} = x_{N-2}$, $\dots$ $x_{-\mu} = x_{N-\mu}$. From this it can be shown that the matrix representation of (3.23) is equivalent to the following representation:

$$\begin{bmatrix} y_{N-1} \\ y_{N-2} \\ \vdots \\ \cdot \\ \vdots \\ y_0 \end{bmatrix} = \begin{bmatrix} h_0 & h_1 & . & . & h_\mu & 0 & \cdot & \cdot & \cdot & . & 0 \\ 0 & h_0 & . & . & h_{\mu-1} & h_\mu & \cdot & \cdot & . & & 0 \\ . & . & . & . & . & . & . & & . & & . \\ 0 & . & . & . & h_0 & . & . & & h_{\mu-1} & & h_\mu \\ . & . & & & . & & & & & & \\ h_2 & h_3 & : & : & h_{\mu-2} & : & : & & \vdots & & \vdots \\ h_1 & h_2 & \cdot & \cdot & h_{\mu-1} & \cdot & \cdot & & h_{\mu-1} & & h_\mu \end{bmatrix} \begin{bmatrix} x_{N-1} \\ x_{N-2} \\ \vdots \\ \cdot \\ \vdots \\ x_0 \end{bmatrix} + \begin{bmatrix} v_{N-1} \\ v_{N-2} \\ \vdots \\ \cdot \\ \vdots \\ v_0 \end{bmatrix} \tag{3.25}$$

above equation can be written as algebraic equation:

$$\mathbf{y} = \hat{\mathbf{H}}\mathbf{x} + \boldsymbol{v}. \tag{3.26}$$

This equivalent model shows that the inserted cyclic prefix allows the channel to be modelled as a circulant convolution matrix $\hat{\mathbf{H}}$ over the N samples of interest. The matrix $\hat{\mathbf{H}}$ is $N \times N$, so it has an eigenvalue decomposition $\hat{\mathbf{H}}$

$$= \mathbf{W\Lambda W}^H \tag{3.27}$$

where $\mathbf{\Lambda}$ is a diagonal matrix of eigenvalues of $\hat{\mathbf{H}}$ and Q^H is a unitary matrix whose rows comprise the eigenvectors of $\hat{\mathbf{H}}$. It is straightforward to show that the DFT operation on $x[n]$ can be represented by the matrix multiplication $\mathbf{X} = \mathbf{W}\mathbf{x}$,

where $\mathbf{X} = (X[0], . . ., X[N-1])^T$, $\mathbf{x} = (x[0], . . . , x[N-1])^T$, and $\mathbf{W}$ is an $N \times N$ twiddle matrix given by

$$\boldsymbol{W} = \frac{1}{\sqrt{N}}\begin{bmatrix} 1 & 1 & . & . & 1 & 1 & . & . & 1 \\ 1 & W_N^1 & W_N^2 & . & . & . & . & . & W_N^{N-1} \\ . & . & . & . & . & . & . & . & . \\ . & . & . & . & . & . & . & . & . \\ 1 & W_N^{N-1} & W_N^{2(N-1)} & . & . & . & . & . & W_N^{(N-1)^2} \end{bmatrix} \tag{3.28}$$

where $W_N = e^{-j\,2\pi/N}$

Now since $\mathbf{W}^{-1} = \mathbf{W}^{H}$, the IDFT can be similarly represented as

$$\mathbf{x} = \mathbf{W}^{-1}\mathbf{X} = \boldsymbol{W}^{H}\mathbf{X}. \tag{3.29}$$

Let $\mathbf{v}$ be an eigenvector of $\hat{\mathbf{H}}$ with eigenvalue λ. Then

$$\lambda\mathbf{v} = \hat{\mathbf{H}}\mathbf{v}, \tag{3.30}$$

The unitary matrix $\mathbf{M}$ has rows that are the eigenvectors of $\hat{\mathbf{H}}$, i.e. $\lambda_i\mathbf{m}_i = \mathbf{H}\mathbf{m}_i$ for $i = 0, 1, . . ., N\text{-}1$. Where $\mathbf{m}_i$ denotes the ith row of $\mathbf{M}H$. It can also be shown by induction that the rows of the DFT matrix $\mathbf{W}$ are eigenvectors of $\hat{\mathbf{H}}$, which implies that $\mathbf{W} = \mathbf{M}^H$ and $\mathbf{W}^H = \mathbf{M}$. Thus we have that

$$\mathbf{Y} = \mathbf{W}\mathbf{y} \tag{3.31}$$

$$= \mathbf{W}[\,\hat{\mathbf{H}}\mathbf{x} + \boldsymbol{v}] \tag{3.32}$$

$$= \mathbf{W}[\,\hat{\mathbf{H}}\mathbf{W}^H\mathbf{X} + \boldsymbol{v}] \tag{3.33}$$

$$= \mathbf{W}[\mathbf{M}\Lambda\mathbf{M}^H\mathbf{W}^H\mathbf{X} + \boldsymbol{v}] \tag{3.34}$$

$$= \mathbf{W}\mathbf{M}\Lambda\mathbf{M}^H\mathbf{W}^H\mathbf{X} + \mathbf{W}\boldsymbol{v} \tag{3.35}$$

$$= \mathbf{M}^H\mathbf{M}\Lambda\mathbf{M}^H\mathbf{M}\mathbf{X} + \mathbf{W}\boldsymbol{v} \tag{3.36}$$

$$= \Lambda\mathbf{X} + \boldsymbol{v}\mathbf{W} \tag{3.37}$$

where since **W** is unitary, $v\mathbf{Q} = \mathbf{Q}v$ has the same noise autocorrelation matrix as $\boldsymbol{v}$, and hence is still generally white and Gaussian, with unchanged noise power. Thus, this matrix analysis also shows that by adding a CP and using the IDFT/DFT, OFDM decomposes an ISI channel into N orthogonal sub-channels and knowledge of the channel matrix **H** is not needed for this decomposition.

3.6.6 Pros and Cons of OFDM

- OFDM converts a frequency selective channel into a collection of flat fading channels by the use of cyclic prefix.
- Spectral efficiency is increased by allowing frequency overlapping of the different carriers (compared to FDMA system). This phenomena is shown in figure 3.6.1
- Due to frequency flat fading, the transmitted information on one OFDM sub-channel can be irremediably lost if a deep fade occurs.
- Rayleigh behavior of such fading can have a dramatic impact on the performance of uncoded OFDM schemes. Methods based on coding (convolutional codes, block codes, multidimensional constellations, turbo-codes) are usually employed with the use of interleaving to combat fading.
- Time interleaving, the coded bits are sent at different times with interval distances greater than the coherence time of the channel. This method is particularly useful in fast fading environment. Otherwise, it incurs a non-tolerable delay in the transmission.
- Frequency interleaving is particularly suited for rich scattering environments. The coded bits are sent on different frequency bands separated by the coherence bandwidth.
- Space interleaving, the coded information is sent on different antennas. A particular simple space-time coding schemes which benefits from the space diversity is the well-known Alamouti scheme [6].

CHAPTER 4

SPACE-FREQUENCY BLOCK-CODED ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING

In OFDM with a sufficient guard interval each subcarrier provides a flat fading MIMO channel, in other words sub-channel appears to be constant for two consecutive symbols [6] for large number of carriers. Therefore a space-time block code can be applied for each subcarrier of OFDM.

4.1 Space-Time Block Codes in OFDM

The mapping of space-time block code matrices on subcarriers is depicted in figure 4.1.1 for a simple example with n_T=2 (number of transmitter antenna) and N_s=2 (number of subcarriers) using the space-time block code

$$B_2 = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}$$

After removing the guard interval and evaluating the FFT at all receive antennas, a block parallel to serial converter performs the inverse operation of the block serial to parallel converter at the transmitter and a space-time block code combiner is applied. The detector assumes that the channel does not change during transmission of a space-time block code matrix, i.e. during P OFDM symbol durations $T_{s,OFDM}$. This is a more critical restriction in OFDM than in single carrier (SC) systems since the OFDM symbol duration is $N_sT_{s,SC}$, where $T_{s,SC}$ would be the symbol duration in a single carrier system. Therefore, the performance of space-time block codes will degrade in fast fading environments.

4.2 Space-Frequency Block Codes in OFDM

To avoid the problem of fast channel variations in time, the symbols of an orthogonal design can be transmitted on neighboring subcarriers of the same OFDM symbol rather than on the same subcarrier of subsequent OFDM symbols [3]. This also reduces the transmission delay. However, the channel needs to be about constant over P neighboring subcarriers.

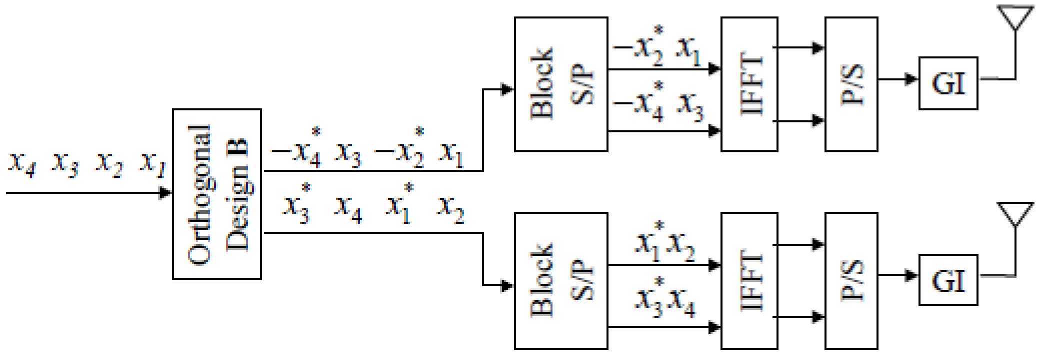

Fig. 4.1.1: Space-time block code in OFDM, with two transmitters [9]

This is true in channels with low frequency-selectivity or can be accomplished by using a large number of subcarriers in order to make the subcarrier spacing very narrow. An example for n_T=2 transmit antennas and N_s=2 subcarriers is given in figure 4.1.2. After the mapping according to the orthogonal design on n_T streams associated with the transmit antennas, a simple serial to parallel converter can be used for each transmit antenna. Consequently, at the receiver a parallel to serial converter is used after the FFT. Space-frequency block codes avoid the problem of fast time variations. However, the performance will degrade in heavily frequency-selective channels where the assumption of constant channel coefficients ($H_i(2m)= H_i(2m+1)$) over a space-frequency block code matrix is not justified.

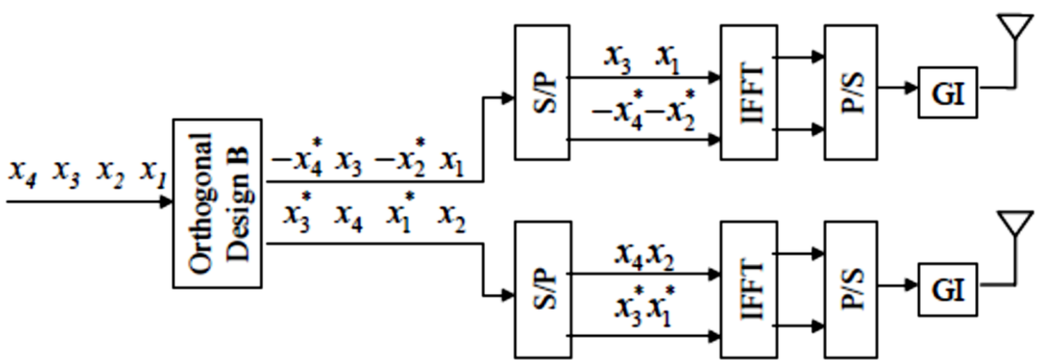

Fig. 4.1.2: Space-frequency block code in OFDM, with two transmitters (Y_T=2) [9]

Above example demonstrates idea behind implementation, to examine SFBC-OFDM more precisely let us assume that $x_i(n)$ denotes n[th] signal transmitted from i^{th} antenna then $x_1(n)$ is given as follows:

$$x_1(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_1(k)\, W_N^{-nk} \qquad ; n=0, 1. \ldots N\text{-}1 \tag{4.1}$$

Where $X_1(n)$ is original QPSK signal to be transmitted. equation (4.1) can be rewritten as[51]

$$x_1(n) = \frac{1}{\sqrt{N}} \sum_{m=0}^{\left(\frac{N}{2}\right)-1} [X_1(2m) + W_N^{-n} X_1(2m+1)]\, W_{\left(\frac{N}{2}\right)}^{-nm} \tag{4.2}$$

$$= \frac{1}{\sqrt{2}} [x^e(n) + W_N^{-n}\, x^o(n)] \tag{4.3}$$

Where $x^e(n)$ is N/2 points IFFT of even samples of $X_1(n)$ and $x^o(n)$ corresponds to odd samples of $X_1(n)$.

$$x^e(n) = \sqrt{\frac{2}{N}} \sum_{m=0}^{\left(\frac{N}{2}\right)-1} X_1(2m)\, W_{\left(\frac{N}{2}\right)}^{-nm} \tag{4.4}$$

$$x^o(n) = \sqrt{\frac{2}{N}} \sum_{m=0}^{\left(\frac{N}{2}\right)-1} X_1(2m+1)\, W_{\left(\frac{N}{2}\right)}^{-nm} \tag{4.5}$$

According to DFT properties [51] $x^e(n)$ and $x^o(n)$ are periodic with period $N/2$ we can write them as $x^e((n)_{N/2})$ and $x^o((n)_{N/2})$. However in figure 4.2.3 these are shown as $x^e(0)$, $x^o(0)$, $x^e(1)$, $x^o(1)$, . . . $x^e(N/2-1)$, $x^o(N/2-1)$

Similarly for second antenna

$$x_2(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_2(k)\, W_N^{-nk} \tag{4.6}$$

$$x_2(n) = \frac{1}{\sqrt{N}} \sum_{m=0}^{\left(\frac{N}{2}\right)-1} [X_2(2m) + W_N^{-n} X_2(2m+1)]\, W_{\left(\frac{N}{2}\right)}^{-nm} \tag{4.7}$$

Now, applying Alamouti scheme to adjacent carriers we have

$$X_1(n) = X_1(0),\ X_1(1),\ X_1(2) \ldots \ldots \ldots .\, X_1(N-2), X_1(N-1) \tag{4.8}$$

$$X_2(n) = -X_1^*(1), X_1^*(0), -X_1^*(3), \ldots \ldots -X_1^*(N-1), X_1^*(N-2) \tag{4.9}$$

$$x_2(n) = \frac{1}{\sqrt{N}} \sum_{m=0}^{\left(\frac{N}{2}\right)-1} [-X_1^*(2m+1) + W_N^{-n} X_1^*(2m)]\, W_{\left(\frac{N}{2}\right)}^{-nm} \tag{4.10}$$

$$= \frac{1}{\sqrt{2}}\left[-x^{0^*}(-n) + W_N^{-n}\, x^{e^*}(-n)\right] \tag{4.11}$$

As discussed in chapter-3, CP is appended at the head of each transmitted block to eliminate inter block interference (IBI), thereby making all channel matrices circulant (***H$_i$***). Received block after removal of cyclic prefix is given below

$$\boldsymbol{r} = \boldsymbol{H_1 x_1} + \boldsymbol{H_2 x_2} + \boldsymbol{n} \tag{4.12}$$

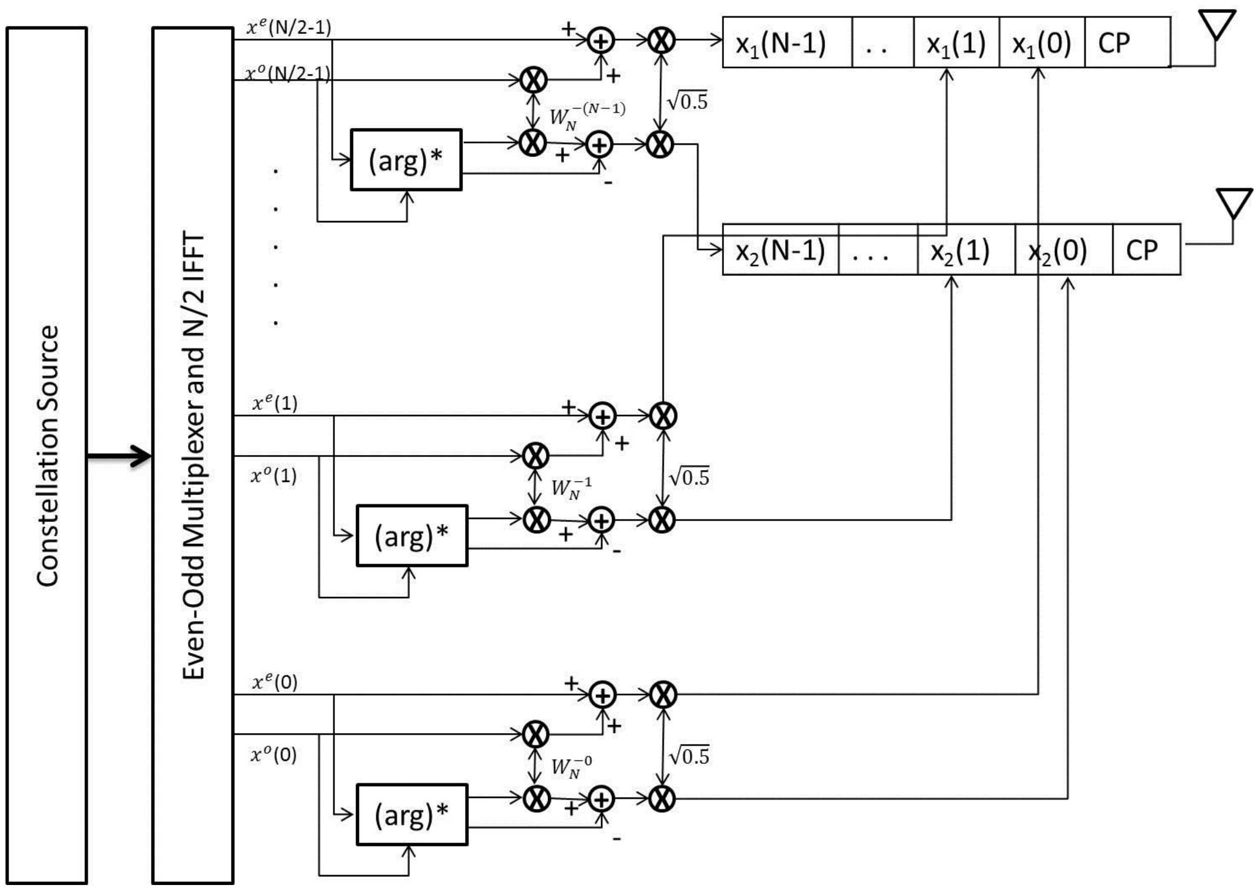

Fig. 4.2.3: Schematic diagram of SFBC-OFDM implementation

Multiplying twiddle matrix ***W*** to equation (4.12)

$$\boldsymbol{R} = \boldsymbol{WH_1x_1} + \boldsymbol{WH2x2} + \boldsymbol{Wn} \tag{4.13}$$

$$= \boldsymbol{WH_1W^HX_1} + \boldsymbol{WH_2W^HX_2} + \boldsymbol{Wn} \tag{4.14}$$

$\boldsymbol{H_i}$ are circulant matrices; $\mathbf{W}$ and $\mathbf{W^H}$ are orthogonal to each other therefore $\boldsymbol{\Gamma_i} = \boldsymbol{WH_iW^H}$; $i=1,2$ are diagonal matrices [52], let Φ_i be diagonal elements of $\boldsymbol{\Gamma_i}$

$$\boldsymbol{R} = \boldsymbol{\Gamma_1 X_1} + \boldsymbol{\Gamma_2 X_2} + \boldsymbol{N} \tag{4.15}$$

Where $\boldsymbol{N}=\boldsymbol{Wn}$ (4.16)

Equation (4.15) can be written in even and odd parts as follows

$$R(2m) = \Phi_1(2m)X_1(2m) - \Phi_2(2m)X_1^*(2m+1) + N(2m) \tag{4.17}$$

$$R(2m) = \Phi_1^*(2m+1)X_1^*(2m+1) - X_1(2m)\Phi_2^*(2m+1) + N^*(2m+1) \tag{4.18}$$

where $m=0, 1, . . . N/2 -1$

Now since Φ_i is channel frequency response (CFR) of m^th subcarrier we make assumption that for large N, CFR is constant over adjacent subcarriers ie., $\Phi_1(2m) \approx \Phi_1(2m+1)$ and $\Phi_2(2m) \approx \Phi_2(2m+1)$. Equation (4.17) and (4.18) are written in matrix form as

$$\boldsymbol{R'_m} = \begin{bmatrix} R(2m) \\ R^*(2m+1) \end{bmatrix} = \begin{bmatrix} \Phi_1(2m) & -\Phi_2(2m) \\ \Phi_2^*(2m) & \Phi_1^*(2m) \end{bmatrix} \begin{bmatrix} X_1(2m) \\ X_1^*(2m+1) \end{bmatrix} + \begin{bmatrix} N(2m) \\ N^*(2m+1) \end{bmatrix} \tag{4.19}$$

$$= \boldsymbol{\Phi'_m X'_m} + \boldsymbol{N'_m} \text{ (say)} \tag{4.20}$$

Multiplying both sides of equation (4.20) by $\boldsymbol{\Phi_m'^H}$ we get equation (4.21)

$$\boldsymbol{Y}_m = \boldsymbol{\Phi_m'^H R'_m} = \boldsymbol{\Phi_m'^H \Phi'_m X'_m} + \boldsymbol{\Phi_m'^H N'_m} \tag{4.21}$$

$$= \begin{bmatrix} |\Phi_1(2m) + \Phi_2(2m)|^2 & 0 \\ 0 & |\Phi_1(2m) + \Phi_2(2m)|^2 \end{bmatrix} \boldsymbol{X'_m} + \boldsymbol{\Phi_m'^H N'_m} \tag{4.22}$$

$$= \begin{bmatrix} X_1(2m)|\Phi_1(2m) + \Phi_2(2m)|^2 \\ X_1^*(2m+1)|\Phi_1(2m) + \Phi_2(2m)|^2 \end{bmatrix} + \boldsymbol{\Phi_m'^H N'_m} = \begin{bmatrix} \alpha^e \\ \alpha^{o^*} \end{bmatrix} \text{ (say)} \tag{4.23}$$

Symbols of equation (4.23) are fed to ML detector, therefore estimated symbols are given in matrix form are output of ML detector

$$\begin{bmatrix} X_{est}^{e} \\ X_{est}^{o^*} \end{bmatrix} = \begin{bmatrix} a^{e} \\ a^{o^*} \end{bmatrix} \tag{4.24}$$

Finally estimated symbols are given as follows

$$X_{est} = [X_{est}^{e}(0), X_{est}^{o}(0), X_{est}^{e}(1), X_{est}^{o}(1) \ldots . X_{est}^{e}(\frac{N}{2} - 1), X_{est}^{o}(\frac{N}{2} - 1)] \tag{4.25}$$

CHAPTER 5

SIMULATION RESULTS

5.1 Simulation Parameters

Simulation results shown in this chapter are verified using Matrix Laboratory v-7.5.0. The symbol-error rate (SER) performance of 2-transmit and 1-receive antenna systems (the STBC SC-FDE and the SFBC SC-FDE) was investigated through computer simulation. We assume that the CSI is perfectly known at the receiver and the effects of channel estimation on our performance comparisons are not addressed. The CP length was set to the channel order.

For purpose of simulation the IEEE 802.11a [53] wireless LAN standard has been used, it occupies 20 MHz of bandwidth in the 5 GHz unlicensed band and is based on OFDM. The IEEE 802.11g standard is virtually identical, but operates in the smaller and more crowded 2.4 GHz unlicensed ISM band [54]. In 802.11a, $N = 64$ subcarriers are generated, out of 64 only 48 are actually used for data transmission, with the outer 12 zeroed in order to reduce adjacent channel interference, and 4 used as pilot symbols for channel estimation (we have not considered these 4 pilot, instead used as actual data ie., 52 subcarrier for actual data). The cyclic prefix consists of $\mu = 16$ samples, therefore total number of samples associated with each OFDM symbol, including both data samples and the cyclic prefix is 80. The same code and modulation has been used for all the subcarriers at any given time. However the modulation types that can be used on the sub-channels are BPSK, QPSK, 16QAM, or 64QAM, we have concentrated only on QPSK. Since the bandwidth B (and sampling rate $1/Ts$) is 20 MHz, and there are 64 subcarriers evenly spaced over that bandwidth, the subcarrier bandwidth is:

$$B_N = 20 \text{ MHz}/64 = 312.5 \text{ KHz}.$$

Since $\mu = 16$ and $1/T_s = 20\text{MHz}$, the maximum delay spread for which ISI is removed is

$$Tm < \mu T_s = 16/20\text{MHz} = 0.8\ \mu\text{sec},$$

Including both the OFDM symbol and cyclic prefix, there are 80=64+16 samples per OFDM symbol time, so the symbol time per sub-channel is

$$T_N = 80T_s = 80/(20 \times 10^6) = 4\mu\text{s}$$

The data rate per sub-channel is $\log_2(M/T_N)$. Thus, the minimum data rate for this system, corresponding to QPSK (2 bit/symbol), an r = 1/2 code, and taking into account that only 48 subcarriers actually carry usable data, therefore bit rate R comes out to be 6.5Mbps. Obviously, a wide range of data rates are possible but for this thesis we have considered data rate of 6.5Mbps. Following table gives overview of set simulation standard:

Parameter	**Value**
Modulation Technique	QPSK
Bits per symbol	2
SNR	0 to 20dB
FFT size	64
Number of used subcarriers	52
FFT Sampling frequency	20MHz
Subcarrier spacing	312.5kHz
Used subcarrier index	{-26 to -1, +1 to +26}
Cylcic prefix duration, T_{cp}	0.8us
Data symbol duration, T_d	3.2us
Total Symbol duration, T_s	4us

For signal to noise ratio we considered the relation between symbol energy and the bit energy is as follows:

$$\frac{E_S}{N_0} = \frac{E_b}{N_0}\left(\frac{nDSC}{nFFT}\right)\left(\frac{T_d}{T_d+T_{cp}}\right) \tag{5.1}$$

Where E_s is symbol energy, E_b is bit energy and N_0 is white noise power spectral density Equation 5.1 can be rewritten in decibels as

$$\frac{E_S}{N_0}dB = \frac{E_b}{N_0}dB + 10log_{10}\left(\frac{nDSC}{nFFT}\right) + 10log_{10}\left(\frac{T_d}{T_d+T_{cp}}\right) \tag{5.2}$$

5.2 Comparison of Results

5.2.1 Comparison of STBC-OFDM for m=0.75, 1, 1.25

Raleigh fading channel is special case of Nakagami-m fading channel when m=1. Fast fading occurs if the coherence time is smaller than the symbol duration of the signal ($T_s > T_c$) this is the case when m<1, such channels become time varying and within a symbol duration rapid changes occur in impulse response of the channel. In STBC systems an assumption is made that channel frequency response (CFR) remain same for two consecutive symbol [6] (ie., for $2T_s$) practically it is not the case for m<1. Therefore performance of STBC-OFDM degrades for m<1 as compared to m≥1.

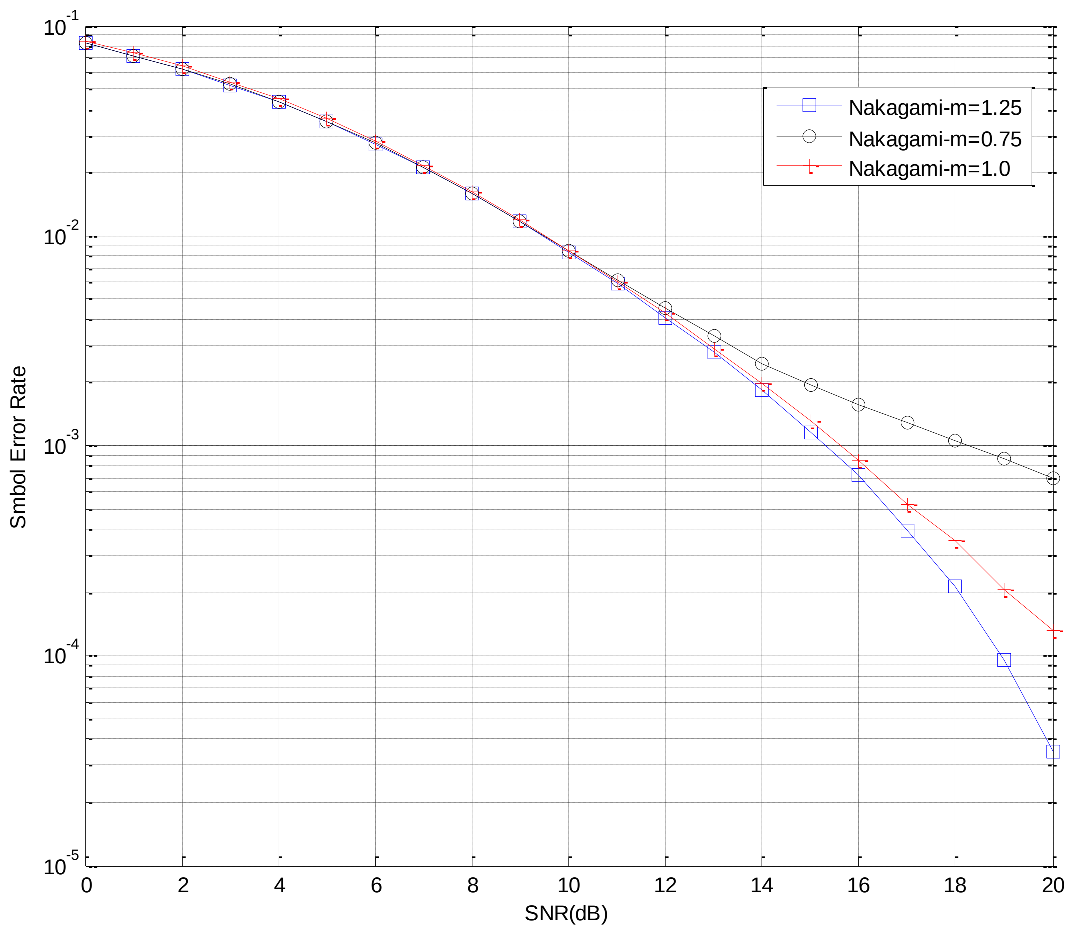

Fig. 5.2.1: STBC-OFDM performance for different fading channels

It is further observed in figure 5.2.1 that there is no significant performance difference from 0dB to 11dB SNR. Starting with 12dB SNR, STBC-OFDM in slow fading (m=1.25) gives consistently better performance than in fast fading (m=0.75). At 20dB SNR, observed SER are 3.5×10^{-5}, 1.325×10^{-4}, 6.925×10^{-4} for m=1.25, 1, 0.75 respectively.

5.2.2 Comparison of STBC-OFDM and SFBC-OFDM for m=0.75

For SFBC, CFR remain same for two adjacent subcarriers ($\Gamma_1(2m) \cong \Gamma_1(2m+1)$) instead of two consecutive symbols, therefore improvement in SER is observed in SFBC-OFDM as compared to STBC-OFDM for m<1 (fast fading.)

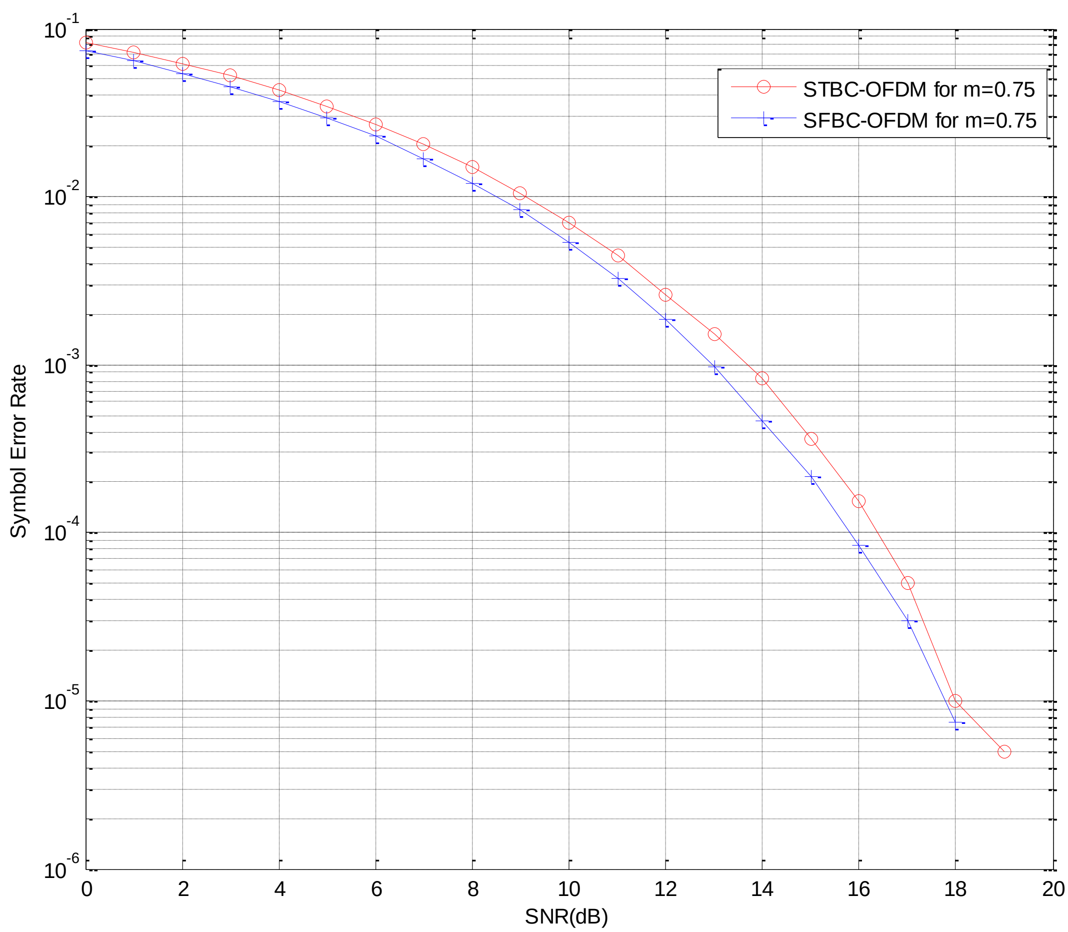

Fig. 5.2.2: STBC-OFDM and SFBC-OFDM comparison for fast fading channel (m=0.75)

This improvement in SER is observable in figure 5.2.2 Starting from 0dB SER of SFBC-OFDM remains lesser than that of STBC-OFDM all the way to 19dB.

5.2.3 Comparison of SFBC-OFDM for m=1.25, 1, 0.75

In figure 5.2.3 performance of SFBC-OFDM system is compared for fast (m<1), slow (m≥1) fading and Raleigh fading channels. It is observed that for slow fading (m=1.25) SFBC combined with OFDM gives very good performance and it is satisfactory for fast fading (m=0.75) as compared to STBC-OFDM system.

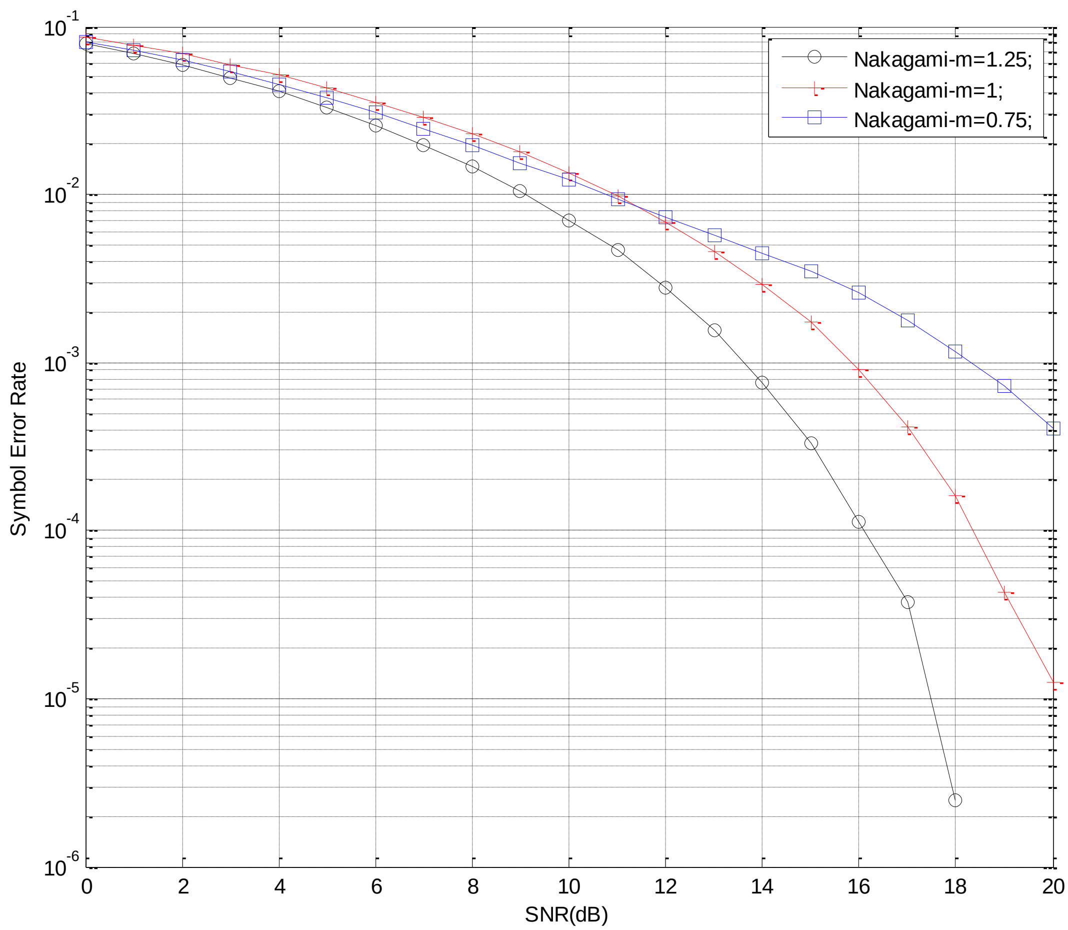

Fig. 5.2.3: SFBC-OFDM performance for different fading channels (m=1.25, 1, 0.75)

At 20dB SNR, SER in fast fading (m=0.75) case is 4.075x10-4 and for slow fading (m=1.25) it is almost 0. At 12dB SNR, SER in fast fading case is 7.3275×10^{-3} and for slow fading it is 2.805×10^{-3} below 11dB SER for fast fading is less than SER of slow fading.

5.2.4 Comparison of STBC-OFDM and SFBC-OFDM for m=1.25

In this section channel is modeled as slow fading channel (m=1.25) and STBC-OFDM system is compared with SFBC-OFDM system. SER of both STBC-OFDM system and SFBC-OFDM system are almost same for SNR < 10dB and after 11dB STBC-OFDM outperform the SFBC-OFDM system.

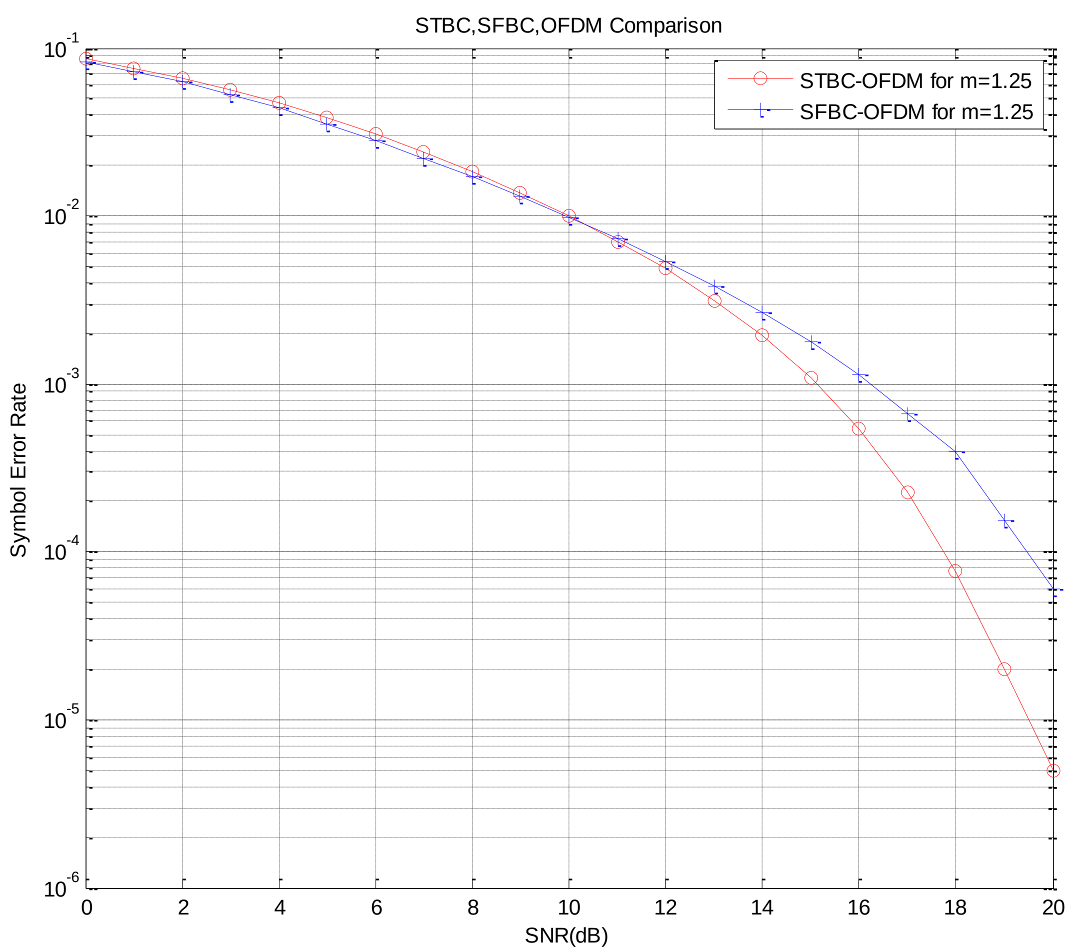

Fig. 5.2.4: STBC-OFDM and SFBC-OFDM comparison for slow fading channel (m=1.25)

There is difference of about 10 times (figure 5.2.4) in SNR at 20dB (SER is $5x10^{-6}$ and $6x10^{-5}$ for STBC-OFDM and SFBC-OFDM respectively at 20dB SNR)

5.2.5 Comparison STBC-OFDM and SFBC-OFDM for m=1.0

In this section channel is modeled as Raleigh fading channel (m=1) and STBC-OFDM system is compared with SFBC-OFDM system. SER of both STBC-OFDM system and SFBC-OFDM system are almost same for 0≤ SNR < 20dB. Figure 5.2.5 suggests that both systems perform nearly same for Rayleigh fading channel (m=1.)

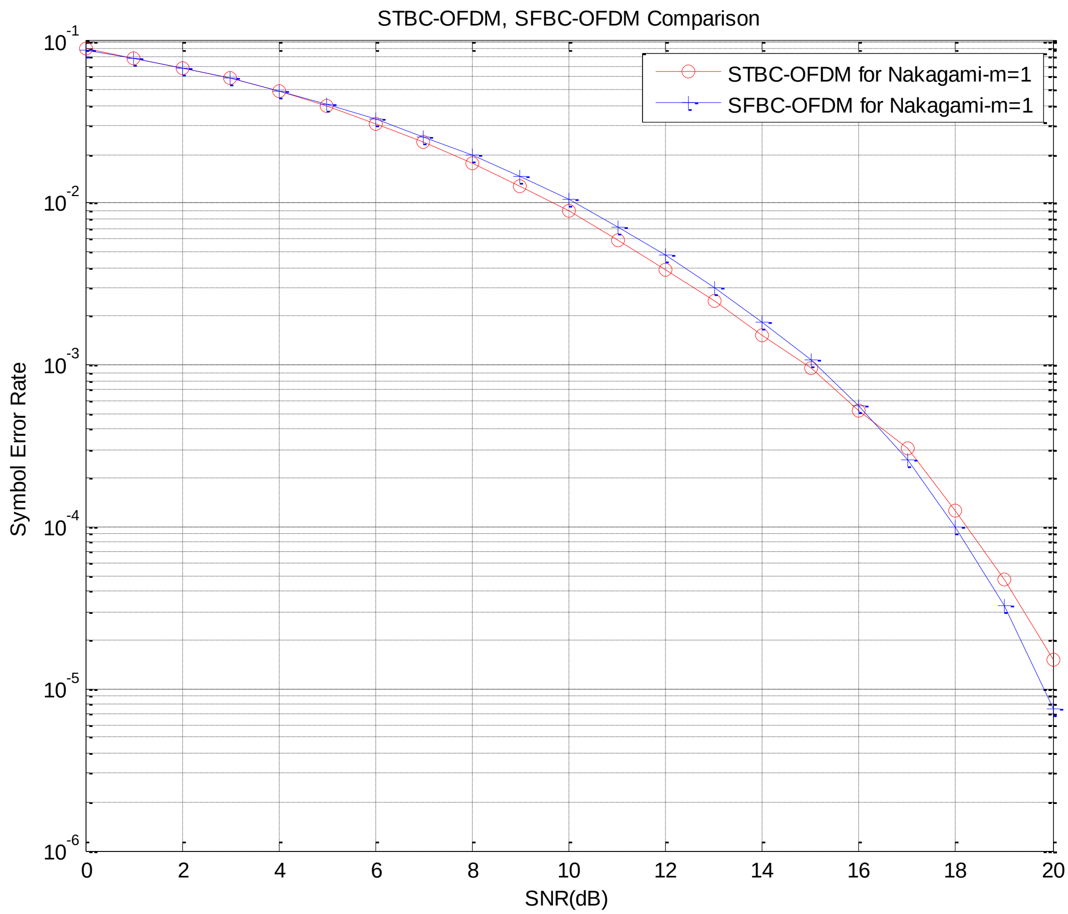

Fig. 5.5: STBC-OFDM and SFBC-OFDM comparison for Rayleigh channel (m=1)

CHAPTER 6

CONCLUDING REMARKS AND FUTURE SCOPE

6.1 Concluding Remarks

OFDM has long been studied and implemented to combat transmission channel impairments. Its applications have been extended from high frequency radio communications to telephone networks, digital audio broadcasting and terrestrial broadcasting of digital television. The advantages of OFDM, especially in the multipath propagation, interference and fading environment, make it promising technology in mobile communication.

Space-time coding for fading channels is a communication technique that realizes the diversity benefits of multiple transmit antennas. Previous work in this area has focused on the narrowband flat fading case where only spatial diversity is available. In this thesis we investigate the use of space-frequency coding in OFDM based broadband system where both spatial and frequency diversity are available. We consider a strategy which consists of coding across OFDM tones and therefore called space-frequency coding.

In wireless communication high-speed mobile environment results in fast fading. The SFBC system codes across two antennas and over two adjacent subcarriers instead of two consecutive symbol intervals (STBC) and thus becomes robust against fast fading distortion in frequency nonselective fading environments. Therefore to mitigate fast fading, use of SFBC-OFDM SC-FDE system is advisable.

Simulation results show that in mobile environments where fading is relative slow such as Rayleigh (special case of Nakagami-m where m=1,) SFBC-OFDM does not give significant performance over STBC-OFDM, more over it is observed that STBC-OFDM mitigate slow fading where $m > 1$ more efficiently than SFBC-OFDM. Further, in case when channel is modeled as fast fading (Nakagami-m, $m < 1$) simulation results show that SFBC-OFDM outperforms STBC-OFDM. When compared for different fading channels SFBC-OFDM gives lowest SER for slow fading ($m > 1$) over entire SNR range (0-20dB). For poor SNR, more precisely SNR ranging from 0 to 11 dB SFBC-OFDM performs better under fast fading channel ($m < 1$) then under Raleigh channel ($m = 1$.)

6.2 Future Scope

In this thesis work only QPSK modulation technique is dealt with, the effect of usage of M-QAM on our performance comparison is not addressed here and is a subject for future research.

We have considered time selective channel research can also be done for doubly selective fading channel. SFBC-OFDM system performance evaluation in the presence of impulse noise is also a subject of future work.

The elements of the orthogonal design are distributed in time over adjacent OFDM symbols in STBC-OFDM which do not give good performance in fast fading as channel is no longer constant for two consecutive symbols [6]. In similar way OC in SFBC-OFDM are distributed over neighboring subcarriers which causes problems in severe frequency-selective channels. Therefore a space-time-frequency mapping, which might be a good compromise, needs to be investigated [9].

The SFBC-OFDM finds its application where communicating device is highly mobile for example DVB-H standard, which supports high mobility [30].

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