# Performance Evaluation of Space-Frequency Block -Coded Detection Technique Based on MMSE Criterion Under Mobile Environment

Thesis submitted towards the partial fulfilment of requirement for the award of degree of

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**IN**

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## DECLARATION

I, **Nandkishor Vansdadiya**, hereby declare that the work, which is being presented in the thesis entitled "**Performance Evaluation of Space-Frequency Block-Coded Detection Technique Based on MMSE Criterion Under Mobile Environment**" by me in partial fulfilment of the requirements for the award of degree of Master of Engineering in Wireless Communication from Thapar University (Deemed University), Patiala, is an authentic record of my own work carried out under the supervision of **Dr. Amit Kumar Kohli,** Associate Professor, Electronics and Communication Engineering Department.

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

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# ABSTRACT

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 symbol period of the transmitted signal is greater than the coherence time of the channel. This causes frequency dispersion which is also called as a time selective fading due to Doppler spreading, which leads to signal distortion, while 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, the use of space-frequency block code (SFBC) is beneficial because symbol is transmitted on neighbouring sub carriers.

SFBC has the better performance as compared to space-time block coding when channel is fast varying in nature, i.e., where the channel varies too quickly. While in slow varying channels, SFBC as well as space-time block-code (STBC) exhibit same performance.

To reduce the fast fading distortion caused by high-speed mobility a single carrier SFBC frequency domain equalization (SC-SFBC FDE) system can be used for reliable communication. The SFBC system codes transmitted, using two antennas and over two adjacent subcarriers instead of two consecutive symbol intervals, is more robust against fast fading distortion in frequency nonselective fading environments.

Therefore we have made an effort to study SC-SFBC FDE system working under frequency selective fading environment.

We have presented first, an Alamouti-like scheme for combining SFBC with single-carrier frequency-domain equalization (SC-FDE) in wireless fading channels where channel response is not same for adjacent subcarriers. The matched filter at the receiver produces error floor in the bit error rate performance, which is compensated by the proposed design. This thesis focuses on the design of low complexity zero forcing (ZF) equalizer for the combining receiver for a communication system with two transmit antennas and one receive antenna. It is shown, through computer simulations that the proposed design of the equalizer outperforms the matched filter receiver for fast fading mobile environments, and also provides an advantage of lower computational complexity at the receiver than classical ZF equalizer.

Further, the work deals with the detection of single carrier space frequency block coded (SC-SFBC) in frequency selective wireless MIMO channel. Due to frequency-selectivity nature of channel the inter-symbol interference (ISI) occurs. In single carrier SFBC system, ISI is caused due to the loss of the ‘quasi-static’ fading assumption caused due to frequency-selectivity of wireless MIMO channel. In this article, we evaluate and propose the performance analysis of SC-SFBC FDE for interference cancellation receiver which mitigates the effect of ISI in frequency-selective MIMO channels.

**Keywords:** Space frequency block coded (SFBC), frequency domain equalization (FDE), inter symbol interference (ISI), frequency selectivity, time selective fading.

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# LIST OF ABBREVIATIONS

| | |
|---|---|
| BER | Bit Error Rate |
| BTS | Base Transceiver Station |
| CDMA | Code Division Multiple Access |
| CFR | Channel Frequency Response |
| CP | Cyclic Prefix |
| CSI | Channel State Information |
| DFT | Discrete Fourier Transform |
| 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 |
| LOS | Line Of Sight |
| LMS | least Mean Square |
| LS | Least Square |
| MIMO | Multi Input Multi Output |
| ML | Maximum Likelihood |
| MMSE | Minimum Mean Squared Error |
| 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 |
| Rx | Receiver |
| SC | Single Carrier |
| SER | Symbol Error Rate |
| SF | Space Frequency |
| SFBC | Space Frequency Block Code |
| SNR | Signal to Noise Ratio |
| ST | Space Time |
| STBC | Space-Time Block-Code |
| Tx | Transmitter |
| Wi-MAX | Worldwide Interoperability for Microwave Access |

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# CHAPTER 1

# INTRODUCTION

***Abstract****: This chapter provides a brief introduction of the thesis. This includes background of wireless communication system, STBC-OFDM, SFBC-OFDM, and objectives, original contributions and organization of the thesis.*

## 1.1 Brief Introduction of Wireless Communication System

The performance of wireless communication systems is ruled by the wireless channel environment. Since the wireless channel is dynamic and unpredictable in nature rather than being static and predictable, it becomes difficult to analyze the characteristics of channel thus making the wireless communication system difficult to analyze exactly. Moreover, optimizing the wireless communication system has become more difficult as the services such as mobile communication and broadband mobile internet access are growing at rapid pace.

Radio propagation is one of the prominent phenomena in the wireless communication, which is subjected to the behavior of the radio waves when they travel from the transmitter to the receiver. Basically, there are mainly three factors which are responsible for affecting the propagation of radio wave i.e. reflection, diffraction, and scattering [1, 2]. When a propagating electromagnetic wave strikes an object having very large dimensions than the wavelength of the wave, this phenomenon is known as Reflection. Here the object could be surface of the earth or a building. Thus the transmit signal power is enforced to reflected back to its origin and does not allowed to pass all the way along the path to the receiver. The other phenomenon is Diffraction which occurs when there are sharp irregularities or small openings in the surface which comes in between the path of the transmitter and the receiver. Moreover, different paths will be formed between the transmitter and the receiver due to the secondary waves generated by diffraction, even when there is no line of sight (LOS) path. The last but the most important phenomenon is Scattering, which forces the radiation of an electromagnetic wave to diverge from a straight path by one or more barriers, which have smaller dimensions compared to the wavelength of the wave. Such scatters may be foliage, street signs, and lamp posts, etc. In the end, it could be deduced that the propagation of the wave is complex and often an unpredictable process trailed by the above mentioned pheromones.

One another phenomenon called fading is a unique characteristic of a wireless channel, which is defined as the variation in the amplitude of the signal over time and frequency. So, fading causes degradation of the signal which is generally non-additive in nature. Fading may occur due to many reasons. One of the prominent reasons is multipath fading. Another reason may be shadowing from various obstacles which otherwise affect the radio wave propagation.  This is called shadow fading. The classification of fading phenomena is shown in the figure below.

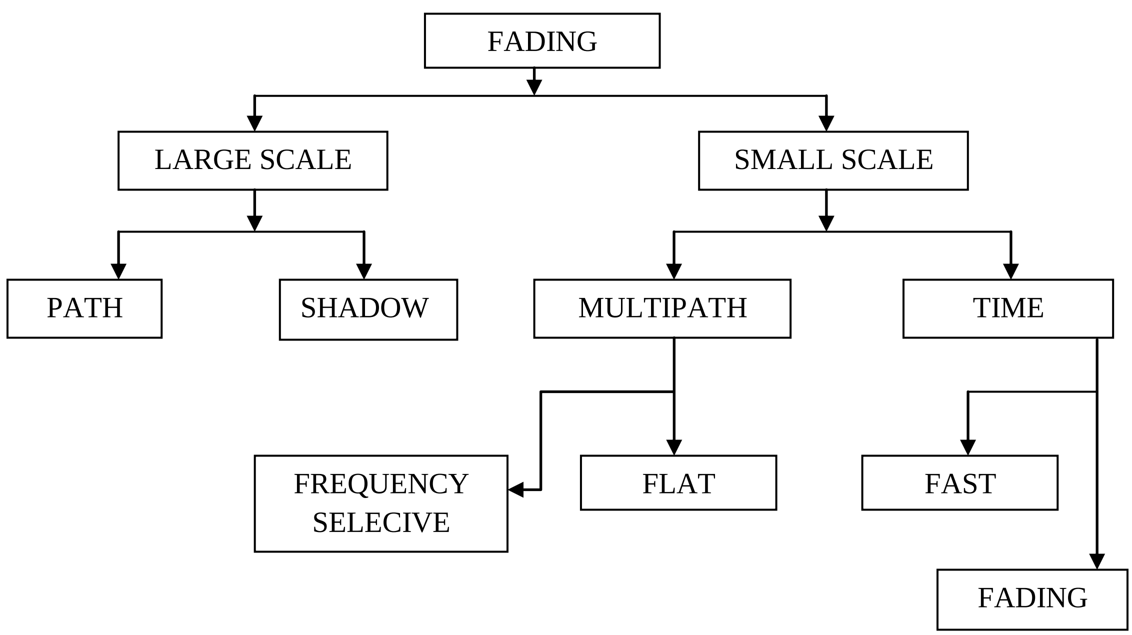


Fig 1.1. Classification of fading channel

Moreover, the large scale fading and small scale fading are two important classifications of the fading phenomena. As mobile travel through a greater distance, large scale fading occur [1]. On the other side, small scale fading occurs when the mobile station travels shorter distances, thus there happens rapid fluctuations in signal levels due to interference of multi paths. The detailed description of the other classifications is given in chapter 3.

Traditionally, annulling of the fading effects is done by simply allowing for deep fades by increasing the transmit power but this approach caused high power consumption as power had to transmit a multiple number of times for better communication. Moreover, interference occurs to considerate amounts. Therefore, orthogonal frequency division multiplexing (OFDM) is implemented which is a promising technique to mitigate inter-carrier interference (ICI) and also it provides robustness to deep fades [3] but simultaneously the receiver complexity increases. Moreover, channel diversity can be exploited as a good option to overcome the effects of the signal fading. In this case, many copies of the same signal which are independent to each other are transmitted over independently fading channels are received by the receiver, thus avoiding the probability of the losing the entire signal components. So, different diversity schemes can be employed to obtain replicas of the transmitted signal.

Secondly, there are numerous challenges that future wireless broadband communication has to face such as high spectral efficiency and high transmission speed which arises due to the use of applications related to audio, video and other popular services like internet [4-7]. By organizing Multiple Input Multiple Output (MIMO) systems in a multipath wireless channel environment, the capacity of the channel increases manifolds. This modification leads to the high data rate transmission without increasing the total transmission power or bandwidth. In present times, most of the concern is earned by MIMO systems since the Space time block codes (STBC) in [8] introduced by Alamouti can be easily implemented in this system which further increases the reliability of the transmission, as redundant copies of the original data are sent over independent fading channels.  . In these codes, the data is coded through space and time. In past years, a lot of research has been done on STBC [9-11]. MIMO and especially STBC have also been adopted in IEEE 802.11n standard [12]. So that higher data rate can be achieved and data is received in more reliable way than traditional single antenna communications [13, 14].

Moreover, it is observed that the wireless communications channels are time varying or frequency selective especially if broadband and mobile applications are considered. Therefore, MIMO has been exploited to address these challenges with Orthogonal Frequency Division Multiplexing (OFDM). OFDM is a very popular technique which has been adopted for present and future broadband communication standards such as LTE or WiMax [15-17].This technique is considered useful since it reduces the effect of frequency selective channel and converts it number of flat fading channel. This happens because the data stream that is to be transmitted is divided into multiple parallel streams and the wideband channel is divided into a number of parallel narrowband sub-channels and thus each sub-channel has a lower rate data stream. OFDM is more popular because of its simplicity of implementation in the digital domain by simply using DFT. Moreover, in OFDM, parallel subcarriers are orthogonal to each other and thus there is overlapping but without causing any interference thus making the system bandwidth efficient. Additionally if cyclic prefix is used with the OFDM, it proves to be as a robust modulation technique under multipath frequency selective fading environment [6, 18].

One popular combination of MIMO and OFDM is STBC-OFDM which was first proposed in [19, 20]. In addition to spatial and temporal diversity, the combination of MIMO-OFDM offers a third dimension of coding which achieves frequency diversity. These coding schemes known as Space-Frequency Block Coding (SFBC) and Space-Time-Frequency Block Coding (STFBC), which are respectively capable of achieving two dimensional coding over space and frequency and three dimensional coding over space, time and frequency have recently been proposed in the literature [21-25]. In addition, coding through spatial and frequency dimension offers implementation advantages [4]. MIMO-OFDM has already been adopted by several standards such as IEEE 802.11n, IEEE802.16a and 3GPP [15, 26, 27]. However, in both STBC-OFDM and SFBC-OFDM, to recover the transmitted symbols channel parameters need to be known at the receiver. Therefore, channel estimation with acceptable level of accuracy and hardware complexity has become an important research topic for MIMO-OFDM systems.

## 1.2 Problem Statement

This thesis report represents the following research work

1. First, STBC and SFBC wireless communication system are investigated for fast wireless fading environment

2. Next, we have evaluated single carrier STBC system with two transmitter and one receiver and compared with one transmitter and one receiver system in same situation.
3. Subsequently, we have evaluated the single carrier SFBC system and compare with single carrier STBC system for different normalized Doppler frequency.
4. Further, we have evaluated the performance of single carrier SFBC detection using combining receiver derived under low complexity zero forcing criterion for 2Tx and 1Rx system.
5. We proposed Interference canceller receiver for SC-SFBC FDE system to mitigate the effect of inter symbol interference.

## 1.3 Organization of the Thesis

This thesis include

- Chapter-1, which introduces to real scenario of today's communication system and how problem can realize using different techniques like STBC and SFBC combine with OFDM.
- Chapter-2, which includes literature survey in the field of fading, OFDM, STBC, Single carrier frequency domain equalization and SFBC-OFDM.
- Chapter-3, which gives briefly discussion of diversity techniques, equalization, and coding to improve the quality of received signal by reducing the fading effects and basic phenomena behind Fading effect and Types of Fading.
- Chapter-4, which discusses the basic principal and implementation of OFDM, STBC-OFDM and SFBC-OFDM are better for communication.
- Chapter-5, which gives briefly introduction of single carrier MMSE frequency domain equalization and SC-STBC FDE system.
- In Chapter-6, we proposed interference canceller receiver and low complexity ZF Combining Receiver and simulation results in MATLAB simulator for SC-SFBC FDE system in frequency selective fading environment.
- Finally in Chapter-7, we describe the conclusions and future work to be carried out to fulfill the research work.

**CHAPTER 2**

# LITERATURE REVIEW

***Abstract:*** *This chapter provides a literature survey in the field of fading, OFDM, STBC, Single carrier frequency domain equalization and SFBC-OFDM.*

## Diversity

Usually, fading occurs in wireless channel due to addition of multipath propagation and interference from other users. The most common channel fading is Rayleigh fading, which causes trouble for the receiver to verify the transmitted signal except less attenuated replica of transmitted signal is available at the receiver. Replica of transmitted signal at receiver is known as diversity technique, which can be provided by using temporal, frequency, polarization, and spatial resources [28, 29]. In many cases, wireless channel is not time varying even not frequency selective. This encouraged the system engineers to develop multiple antenna at both transmitter and receiver to attain spatial diversity.

In 1993 Wittneben studied transmit diversity [30] and observed the effects in wireless fading channels. He suggested that there is relative simplicity of implementation and feasibility of having multiple antennas at base station. In this way, first bandwidth efficient transmit diversity scheme was proposed. Later Foschini introduced multi-layered space time architecture [31].

## STBC

In [32], quasi-orthogonal STBC (QO-STBC) is used instead of the conventional O-STBC with OFDM MIMO systems, thus resulting in a system called quasi-orthogonal space-frequency block coding (QOSFBC). This approach therefore, helps in increasing the transmission rate of STBC-MIMO-OFDM. This paper discusses the employment of a Zero-forcing detector, which is a low complexity receiver and removes the error floor completely.

As it is a known fact, that QO-STBC can offer full transmission rate when it is compared to the partial rate provided by OSTBC. On using computer simulation, the QO-SF-OFDM system is proved to perform better than the O-STBC based SF-OFDM system and on the same time, this improves the code transmission rate.

In [33], detailed study of diversity coding for MIMO systems is presented. Wireless networks have emerged as part of our everyday life. A large number of equipments such as wireless LANs, cell phones etc are using wireless networking. However, range and data rate is limited for these wireless devices but reaches are continually made to overcome these shortcomings. So, one of the prominent methods now-a-days is to use Multiple-Input Multiple-Output (MIMO). Precoding (multi-layer beam forming), diversity coding (space-time coding), and spatial multiplexing are performed with MIMO system which facility is further exploited by the use of multiple antennas. So, higher data rate can be obtained using these MIMO techniques that is explained in details here in this paper. Furthermore, Different space-time block coding (STBC) schemes including Alamouti's STBC are evaluated for 2 transmit antennas as well as 3 and 4 transmit antennas using computer simulation and performance in terms of BER is evaluated for different modulation schemes such as BPSK, QPSK, 16-QAM, and 64-QAM

**OFDM**

When the channel is time varying inter-carrier interference (ICI) occurs in orthogonal frequency division multiplexing system. K. Kim et al. [51] engaged a scheme known as LCF-PCC-SFBC OFDM and used a polynomial cancellation code (PCC) to remove this ICI and also used a linear complex field (LCF) code as a transmit diversity technique to make transmission rate compatible up to $nT/(nT+1)$, where $nT$ is the number of transmitted antenna used. According to the analysis and simulation results this scheme can achieve higher channel capacity as well as lower bit rate (BER) when compared with the conventional orthogonal code (OC) such as Alamouti code in time varying channels, even with a linear receiver and inaccurate channel estimation.

In OFDM there are orthogonal sub-channels which reduce receiver complexity. Cyclic prefix in OFDM reduces the performance rate which is the drawback of the OFDM system. In 2003 Alireza Tarighat and Ali Sayed [52] formed a receiver structures in such a way that exploit cyclic prefix to increase the performance rate. In these structures modifications are only done at receiver but not at OFDM transmitter. Optimum receivers do not result in extra processing complexity in both least-mean –squares (LMS) and least-squares (LS) when compared with standard OFDM receiver. This structure is further modified and formed MIMO OFDM structure which gives the improved performance of the proposed receiver.

**SFBC- OFDM**

In [34], transmitter diversity scheme for wireless communications is presented for the frequency selective fading channels. This scheme makes use of orthogonal frequency division multiplexing for converting the frequency selective fading channel to flat fading sub channels. Thus, space frequency codes are implemented. Simulation results shows that proposed space-frequency OFDM (SF-OFDM) transmitter diversity technique gives the same performance space-time OFDM (ST-OFDM) transmitter diversity system in case of slow fading environment but SF-OFDM performs better in the fast fading environments. Secondly, advantages of SFOFDM over the ST-OFDM transmitter diversity technique are also described.

In [35], space frequency block coded (SFBC) Orthogonal frequency division multiplexing (OFDM) is compared with space time block coded (STBC) OFDM and the Space frequency codes give better performance than the space time codes in the fast fading channel environment. However, if complex orthogonal code having code rate one is considered then it does not exist if more than two antennas are there. A quasi-orthogonal SFBC-OFDM system is proposed to amplify the throughput. This case is observed for four transmit antennas. Moreover, the advantages of the SFBC in mobile communication environment is described and it is shown that the improved throughput achieves code rate one. The inference is finally drawn that the SFBC performs better than the STBC in case of fast fading channel.

In [36], it has been shown that the space-frequency block coded (SFBC) OFDM signals has more benefits if used in high-mobility broadband wireless access, since channel in this case is highly time- as well as frequency-selective. So, receiver has to pay for this as it experiences inter-symbol interference (ISI) and inter-carrier interference (ICI) as well. ISI takes place as the condition of the ‘quasi-static’ fading is terminated due to frequency/ time-selectivity of the channel. As it is already known that due to time selectivity of the channel

ICI occurs that further results into loss of orthogoanlity among the subcarriers. So, an interference cancelling receiver for SFBC-OFDM is presented which combats the ISI and ICI caused due to highly frequency/time selective channels using PIC algorithm. By using this algorithm, first the ISI is estimated and then annulled and secondly, the same procedure is followed for the ICI. This gives reduced error floor as multiple stage algorithm is applied. So, the proposed detector effectively terminates the effects of ISI and ICI and this PIC detector can be easily employed in space-time-frequency (STFC) coded OFDM.

In [37], a novel SFBC-OFDM scheme has been designed which integrates SFBC-OFDM with FIR- ICI mitigation equalization. Since it is quite difficult to estimate the channel accurately using scattered pilots as channels parameters have fast varying nature, but exact estimate of channel is required for ICI mitigation for MIMO-OFDM. A novel space-frequency block coding OFDM (SFBC-OFDM) scheme for doubly selective channels has been presented which reduces the complexity of channel estimation. Here, banded and sparse structure of the channel has been proposed which is employed in both the frequency and time domains. Further, as described earlier a finite-impulse-response minimum-mean-square error (FIR-MMSE) ICI cancellation algorithm for mobile SFBC-OFDM has been designed; its effectiveness is demonstrated for digital video broadcasting-handheld (DVB-H) systems.

In [38], co-channel interference (CCI) cancellation decoders for the STBC-OFDM in a fast fading channel including the Alamouti method, SIC method, ML method, DMLD method and the DZFD method are proposed. A simple DZFD method is given to reduce the computational complexity of the DMLD method. The outdoor channel measurements are conducted to verify the CCI cancellation ability of different methods. Moreover, best results are obtained when ML decoding is applied for both the simulated and measured scenarios. The DMLD and the DZFD performs almost equally but better than the he SIC and the Alamouti detection method. Yet, the DZFD is easy to implement than DMLD.

In [39], space frequency block coded (SFBC) Orthogonal frequency division multiplexing (OFDM) is compared with space time block coded (STBC) OFDM and the Space frequency codes give better performance than the space time codes in the fast fading channel environment.

However, if complex orthogonal code having code rate one is considered then it does not exist if more than two antennas are there. A quasi-orthogonal SFBC-OFDM system is proposed to amplify the throughput. This case is observed for four transmit antennas. Moreover, the advantages of the SFBC in mobile communication environment is described and it is shown that the improved throughput achieves code rate one. The inference is finally drawn that the SFBC performs better than the STBC in case of fast fading channel.

In [40], a low complexity Zero Forcing receiver is designed for two or three transmit antenna and one receive antenna. Generally, a assumption has been observed in case of STBC-OFDM that the channel coefficients will be same for the adjacent subcarriers but in this paper, SFBC- OFDM has been analyzed in the broadband wireless channel but with the condition that the channel coefficients are not same for adjacent subcarriers and on using Matched Filter receiver, it is found that it causes error floor in bit error rate performance. So, the above receiver has been designed to overcome the shortcomings of the MF.

**SC- Frequency Domain Equalization**

In [41], a space-frequency block-coded (SFBC) single-carrier frequency-domain equalization (SC-FDE) system has been proposed. The sequence which has to be transmitted in case of the proposed system is designed in such a way that spatial and frequency diversities could be exploited. Moreover, the combining receiver that is placed at the receiver end makes use minimum mean square error criterion. Thus it is deduced that the performance given by the space-time block-coded SC-FDE system over fast fading channels is better. Additionally, this system provides lower computational complexity when compared to the space frequency block coded (SFBC) orthogonal frequency division multiplexing (OFDM). Hence, a novel means are provided to combat fast fading for the single carrier transmission scheme and benefits of both SFBC and SC MMSE-FDE has been realized by this proposed scheme. It has been observed that proposed system outperforms the STBC SC-FDE system in fast fading environments just with little increased computational complexity.

In [42], Single-Carrier Frequency Division Multiple Access (SCFDMA) has been adopted as a possible air interface for future wireless networks. This provides the combination of the various advantages of the Orthogonal Frequency Division Multiple Access and the low Peak to Average Power Ratio (PAPR) of single carrier (SC) transmission. It has been observed that the already existing transmit antenna diversity techniques i.e. Space-Time Block Coding and Space-Frequency Block Coding are proved to be incompatible not only with the system constraints but also with the SC nature of SC-FDMA. So, novel space-frequency flexible coding scheme has been proposed which is compatible with the SC-FDMA. Since, it is difficult to assure to have even number of SC-FDMA blocks in each frame in case of space-time block coding but single carrier property has been exploited using space frequency block coding. So, a new SFBC coding scheme has been proposed in this paper which is somehow compatible with SC-FDMA transmission and any frame size. Moreover, it gives better performance if PAPR and Bit Error Rate (BER) on frequency selective Multiple Input Multiple Output channels are considered. So, low PAPR of SC-FDMA can be obtained by the use of this proposed scheme.

In [43], the performance of two relay-assisted SC transmission is analyzed and generalized over fast fading channels. So for the same purpose, Space-time block coded (STBC) single carrier (SC) transmission was employed for practical purposes in distributed manner. However, it is assumed that the channel is constant over the space time codeword but if distributed STBC-SC system is considered, it suffers from time selectivity of wireless fading channels. So, distributed space-frequency block coded (D-SFBC) SC transmission has been proposed to provide better performance results if the transmission is done over fast fading channels. Thus, performance of the two distributed SC transmission is compared taking fast fading channel into consideration.  Secondly, a channel model has been presented that suitably captures the time variations in the wireless channel  when evaluation of mean square error (MSE)  and BER of two systems over fast fading channels has to be done. Thus, MSE evaluation can be done in a simple way and moreover, characteristics of inter-carrier interference can be observed. So, if the D-STBC and D-SFBC SC transmission is compared for severe Doppler spread, then D-SFBC SC transmission gives better results. Spatial diversity gain could be achieved in both the above cases with minimum complexity of multiple antennas in case of slow fade. Simultaneously, the results are observed for different block length for both the cases. As the block length increases, the performance of DSTBC SC system deteriorates as compared to D-SFBC SC system. This generally happens due to the additional terms in the MSE of D-STBC SC. So, in the end it could be inferred that for high speed transmission, D-SFBC could be good solution especially with large block length.

In [44], the concept of SC-SFBC is applied in multiuser multiple-input multiple-output (MUMIMO) systems. Since, Single-carrier space frequency block coding (SC-SFBC) is an effective and innovative mapping scheme that is quite suitable for implementing transmit diversity in single-carrier frequency division multiple access (SC-FDMA) systems. By using SC-SFBC, low envelop variations of SC-FDMA can be preserved. Thus, a novel algorithm has been designed such that parameters of SFBC-OFDM are optimized which then lowers the receiver complexity and maximize the overall spectral occupancy in MU-MIMO SC-FDMA systems.  Thus good results are obtained using proposed MU scheme.

In [45], Alamouti-like scheme for combining space–time block-coding with single-carrier frequency-domain equalization is proposed. The scheme is shown to achieve significant diversity gains at low complexity over frequency-selective fading channels with two transmit antennas. Significant performance gains over single-antenna transmission were demonstrated for the EDGE TU channel.

## CHAPTER 3

# EQUALIZATION, DIVERSITY, FADING AND CHANNEL CODING

***Abstract:*** *This chapter provides a brief introduction of the equalization, diversity, and channel coding. Equalization, diversity, and channel coding are three techniques which can be used independently or in combine to improve received signal quality. This chapter also includes basic phenomena behind fading effect and types of fading, and channel performance over small-scale times and distances.*

## 3.1 Equalization

Equalization is one of the most useful techniques that try to invert the response of the channel. Equalization compensates for inter symbol interference (ISI) created by multipath within time dispersive (frequency selective) channels. If the modulation bandwidth exceeds the coherence bandwidth of the radio channel, ISI occurs and modulation pulses are spread in time into adjacent symbols which result in overlapping of adjacent symbols.

An equalizer within a receiver compensates for the average range of expected channel amplitude and delay characteristics. Equalizers must be adaptive since the channel is generally unknown and time varying in nature.

## 3.2 Diversity

Diversity is another technique used to compensate for fading channel impairments, and is usually implemented by using two or more receiving antennas.

The evolving 3G common air interfaces also use transmit diversity where base station may transmit multiple replicas of the signal on spatially separated antennas or frequencies.

We can use combination of diversity and equalizer to improve the performance of channel without altering the common air interfaces and without increasing the transmitted power or bandwidth.

However, while equalization is used to reduce the effects of time dispersion (ISI) and diversity is usually employed to reduce the depth and duration of the fades experienced by a receiver in a local area which are due to motion.

### 3.2.1 Diversity Techniques

Advantage to use Diversity unlike to equalization is that diversity requires no training overhead since a training sequence is not required by the transmitter. Furthermore, there are a wide range of diversity implementations which are very practical and improve thr performance of channel with little added cost.

Diversity measure the random nature of radio propagation by finding independent or uncorrelated signal paths for wireless communication. Generally in all applications, diversity decisions are taken by the receiver, and transmitter is unknown to decisions.

Simple concept of diversity technique is that, if one radio path undergoes a deep fade, another independent path may not be affected by fading and having a strong signal. By having more than one path to select from, both the instantaneous and average SNRs at the receiver may be improved, often by as much as 20 dB to 30 dB [46].

Diversity can be classified according to the transmission scheme used as follows,

#### 3.2.1.1 Space or Antenna Diversity

Space diversity is one of the most popular forms of diversity used in wireless communication. Today's wireless systems consist of an elevated base station antenna and a mobile antenna close to the ground, where there is no line of sight exist between transmitter and receiver so signals will get scattered. From this model Jakes deduced that the signals received from spatially separated antennas on the mobile would have essentially uncorrelated envelopes for antenna separation of on half wavelength or more. The concept of antenna space diversity is also used in base station design. At each cell site, multiple base station receiving antennas are used to provide diversity reception. However, since the important scatters are generally on the ground in the vicinity of the mobile, the base station antennas must be spaced considerably far apart to achieve uncorrelation. Separations on the order of several tens of wavelengths are required at the base station. Multiple antennas can be used either at a mobile station or base station or both sides.

#### 3.2.1.2 Polarization Diversity

At the base station, space diversity is considerably less practical than at the mobile because the narrow angle of incident fields requires large antenna spacing and the cost of space diversity is high.

To overcome the limitations of space diversity, polarization diversity comes into picture. In this only provides two diversity branches, it does allow the antenna elements to be co-located. In the early days of cellular radio, all subscriber units were mounted in vehicles and used vertical whip antennas.

#### 3.2.1.3 Frequency Diversity

In Frequency diversity, replicas of information signal are transmitted on more than one carrier frequency. So all these sub carriers follow the different path to propagate and will thus not experience the same fades.

Theoretically, if the channels are uncorrelated, the probability of simultaneous fading will be the product of the individual fading probabilities

Frequency diversity is often employed in microwave line-of-sight links which carry several channels in a frequency division multiplex mode (FDM).

#### 3.2.1.4 Time Diversity

Time diversity repeatedly transmits same information at different time spacing that exceed the coherence time of the channel, so that multiple repetitions of the signal will be received with independent fading conditions.

One modern implementation of time diversity involves the use of the RAKE receiver for spread spectrum CDMA, where the multipath channel provides redundancy in the transmitted message.

### 3.3 Introduction to Fading

The presence of reflectors or obstacles in the environment surrounding a transmitter and receiver create multiple paths. As a result, the receiver receives multiple copies of the transmitted signal, those propagated at a different path. Each signal copy will experience differences in phase shift, delay and attenuation while travelling through the channel. This can result in either constructive or destructive interference, amplifying or attenuating the signal power seen at the receiver. Strong destructive interference is frequently referred to as a deep fade and may result in temporary failure of communication due to the reduction in the channel signal-to-noise ratio.

Fading is used to describe the rapid fluctuations of the amplitudes, phases, or multipath delays of a radio signal over a short travel distance or period of time, so that large-scale path loss effects may be negligible to ignore it.

Fading is caused by interference between two or more versions of the transmitted signal or multiple copies of transmitted signal which arrive at the receiver at slightly different times. These waves, called multi path waves, which are combine at the receiver antenna to give a resultant signal which can vary widely in amplitude and phase, depending on the distribution of the intensity and relative propagation time of the waves and the bandwidth of the transmitted signal [46].

Multipath waves in the wireless mobile channel create small-scale fading effects. The three most important effects are:

- Rapid changes in signal strength over a small travel distance or time interval
- Random frequency modulation due to varying Doppler shifts on different multipath signals
- Time dispersion or echoes created by multipath propagation delays.

#### 3.3.1 Path Loss

The average received power decreased with the increasing in travelling distance. In free space when there is a direct line of sight 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 given as follows,

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

Where $P_R$ and $P_T$ are the received and transmitted powers respectively and $f$ is the carrier frequency, $d$ is the distance between transmitter and receiver, $G$ is the power gain from the transmitter and receiver antennas, $a = 2$ is the path loss component and path loss is defined as $P_T / P_R$ [11].

### 3.3.2 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.

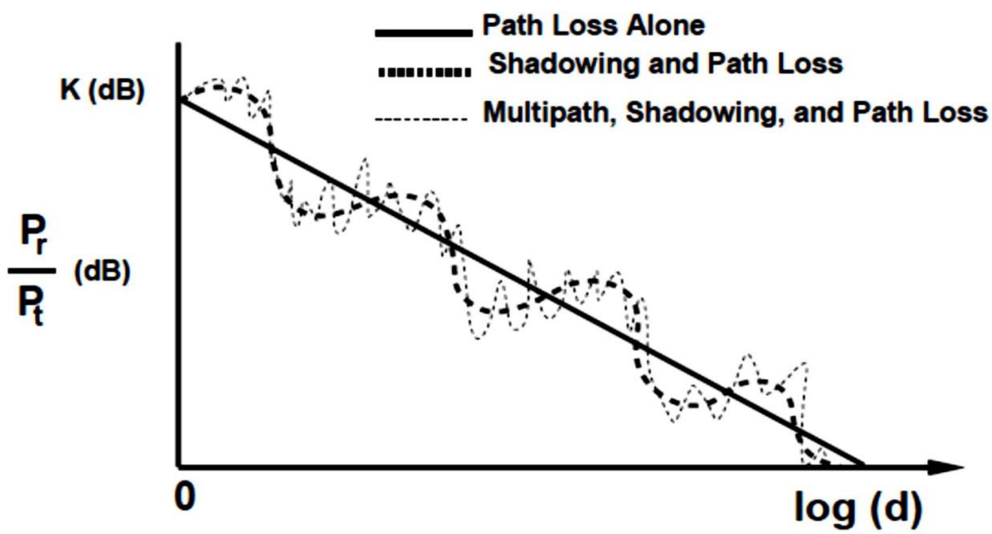


Fig. 3.1 Received signal to transmitted power ratio versus distance [2].

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

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

Where $d_0$ is the reference distance and $X_{\Omega}$ is a zero mean Gaussian random variable with a standard deviation Ω.

A typical situation is shown in Fig. 3.1 is useful to understand the difference between path loss, shadowing and flat fading, which is described in the following section.

### 3.3.3 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 measured by delay spread and coherence bandwidth, where the time-varying nature of the channel occurs due to mobility is quantified by Doppler spread and coherence time [46].

- **Coherence Bandwidth**

The coherence bandwidth is a measure of the maximum frequency difference for which signals still strongly correlated in amplitude.

The channel power profile is coupled with channel frequency response through Fourier transform. Coherence bandwidth ($B_C$) is used to measure the channel frequency response and it is inversely proportional to RMS delay spread. Coherence bandwidth gives the nature of channel flatness. 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 Coherence Bandwidth ($B_C$) is severely distorted for a 0.9 correlation $B_C \approx 1/50\sigma_{\tau}$ [46].

- **Delay Spread**

The reflected parts arrive later than the original signal to the receiver. RMS delay spread ($\sigma_\tau$) is the metric to characterize this delay in terms of second order moment of the channel power profile. Its typical values are in microseconds for outdoor and in nanoseconds for indoor radio channels [2]. Depending on the symbol duration ($Ts$), $\sigma_\tau$ plays an important role to find out how the signal is treated by the channel.

- **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 rapid change in the frequency. If the movement is toward the signal generator, the Doppler shift is positive otherwise it is negative [46].

- **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 or splitting of the spectrum due to Doppler shift [46].

- **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 [46].

- **Coherence Distance**

Coherence distance defines as a minimum distance between points in space for which the signal are uncorrelated [46]. It is important whenever multiple antenna systems. Its value is 0.5 wavelengths for wide beam-width receive antennas and about 10 and 20 wavelengths for low-medium and high base transceiver station (BTS) antenna heights, respectively.

### 3.3.4 Classification of Fading

According to the relation between the signal parameters, such as bandwidth, symbol period, etc. and the channel parameters such as rms delay spread and Doppler spread, different transmitted signals will effected by different types of fading. The time dispersion and frequency dispersion mechanisms in a mobile radio channel introduce four possible distinct effects, which are classified depending on the nature of the transmitted signal, nature of the channel, and the nature of mobility of system. While multipath delay spread leads to time dispersion and frequency selective fading, while Doppler spread resultant into frequency dispersion and time selective fading.

#### 3.3.4.1 Fading Effects Due to Multipath Time Delay Spread

Time dispersion can be classified into two ways as shown below:

- Flat Fading
- Frequency Selective fading.

**1) Flat Fading**

If the mobile radio channel has a constant gain and linear phase response over a bandwidth which is greater than the bandwidth of the transmitted signal, then the received signal will introduced by flat fading [46].

In flat fading, the multipath structure of the channel is such a way that the spectral characteristics of the transmitted signal are preserved at the receiver. However the strength of the received signal changes with time, due to fluctuations in the gain of the channel caused by multipath. The Characteristics of a flat fading channel is shown in Fig. 3.2.

Generally, narrow band signals effect by flat fading because their bandwidth is small 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.

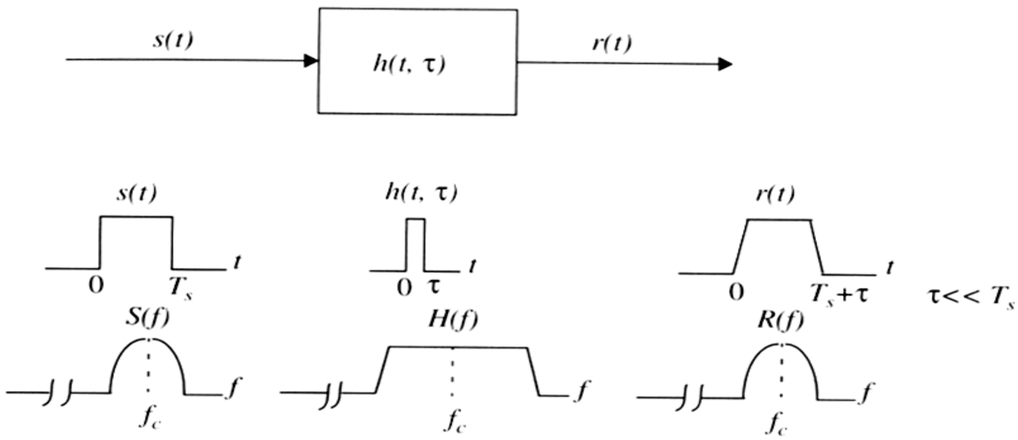


Fig. 3.2 Flat fading channel [2].

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 [46]. Diversity is one method which averages multiple independent channels. Since the channels are independent they have lower probability to experience fades at the same time.

**2) Frequency Selective Fading**

If the mobile radio channel has a constant-gain and linear phase response over a bandwidth which is smaller than the bandwidth of transmitted signal, then the channel will introduced frequency selective fading effect on the received signal.

Under such conditions, the channel impulse response has a multipath delay spread which is greater than the reciprocal bandwidth of the transmitted message waveform. When this situation occurs, the received signal includes multiple versions of the transmitted waveform which are attenuated (faded) and delayed in time, and hence the received signal is distorted. Frequency selective fading is occurs due to time dispersion of the transmitted symbols within the channel. Thus the channel introduces inter symbol interference (ISI). Viewed in the frequency domain, Gains of certain frequency components in the received signal spectrum are greater than others. Frequency selective fading channels are complex in nature to model than flat fading channels since each multipath signal must be modeled and the channel must be considered to be a linear filter, as shown in Fig. 3.3.

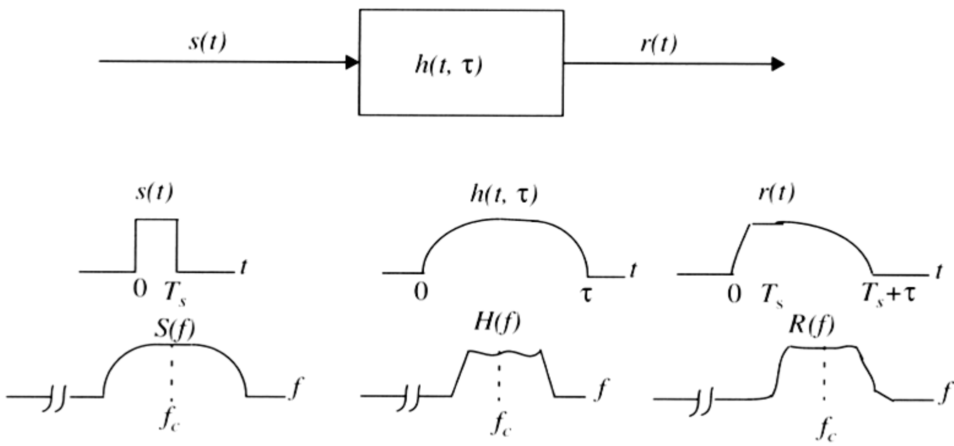


Fig. 3.3 Frequency selective fading channel [2].

There are several ways to combat ISI [46]. Equalization is one method that tries to cancel or invert the response of the channel.

#### 3.3.4.2 Fading Effects Due to Doppler Shift

According to the rapid change in transmitted signal compared to the rate of change of the channel, a channel may be classified either as a fast fading or slow fading channel.

**1) Fast Fading**

In a fast fading channel, the channel impulse response changes rapidly within the symbol duration. That is, the coherence of the channel is smaller than the symbol period of the transmitted signal. This leads to frequency dispersion (or time selective fading) due to Doppler spreading, which causes the distortion in signal.

Viewed in the frequency domain, signal distortion due to fast fading increases with increasing Doppler spread relative to the bandwidth of the transmitted signal

### 2) Slow Fading

In a slow fading channel, the channel impulse response changes at a rate much slower than the transmitted baseband signal. In this case, the channel may be assumed to be static or constant over one or several reciprocal bandwidth intervals.

In the frequency domain, this implies that the Doppler spread of the channel is much less than the bandwidth of the baseband signals. Therefore the channel impulse response changes slowly.

## 3.4 Channel Coding

Channel coding improves the small-scale link performance by adding redundant data bits in the transmitted message so that if an instantaneous fade occurs in the channel, the data will be recovered at the receiver.

At the baseband portion of the transmitter, a channel coder maps the user's digital message sequence into another specific code sequence containing a greater number of bits than originally contained in the message. The coded message is then modulated for transmission in the wireless channel.

Channel coding is used by the receiver to detect or correct some or all of the errors introduced by the channel in a particular sequence of message bits. Because decoding is performed after the demodulation portion of the receiver, coding can be considered to be a post detection technique. The added coding bits lower the data transmission rate through the channel (that is, coding expands the occupied bandwidth for a particular message data rate).

There are three general types of channel codes: block codes, convolution codes, and turbo codes.

Channel coding is generally treated independently from the type of modulation used, although this has changed recently with the use of trellis coded modulation schemes, OFDM, and new space-time processing that combines coding, antenna diversity, and modulation to achieve large coding gains without any bandwidth expansion.

### 3.4.1 STBC and SFBC Schemes

In a given Alamouti code block, two symbols of $S_1$ and $S_2$ are encoded using the following orthogonal matrix

$$\mathbf{A} = \begin{bmatrix} S_1 & S_2 \\ -S_2^* & S_1^* \end{bmatrix} \qquad (3.3)$$

The encoding matrix defines the transmission format with the row index indicating the antenna number and the column index indicating the OFDM symbol index (sub-carrier index) for STBC (for SFBC) .

#### 3.4.1.1 Subcarrier Mapping for STBC

For STBC, a pair of symbols, $S_1$ and $S_2$, are encoded into four variants, $S_1$, $S_2$, $-S_2^*$ and $S_1^*$. As illustrated in Fig. 3.4 $S_1$ is transmitted over a certain sub-carrier from antenna one, and $-S_2^*$ is over the same subcarrier from antenna two [15]. During the next OFDM symbol, $S_2$ and $S_1^*$ are mapped onto the same sub-carrier from the two antennas. That is, each symbol (or its positive/negative conjugate) is transmitted from two antennas and over two OFDM symbols.

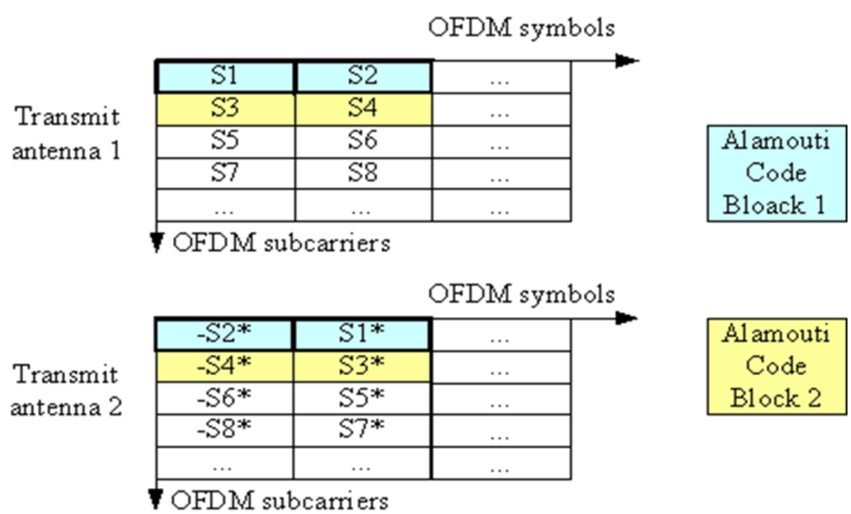


Fig. 3.4. Subcarrier mapping in STBC

#### 3.4.1.2 Subcarrier Mapping for SFBC

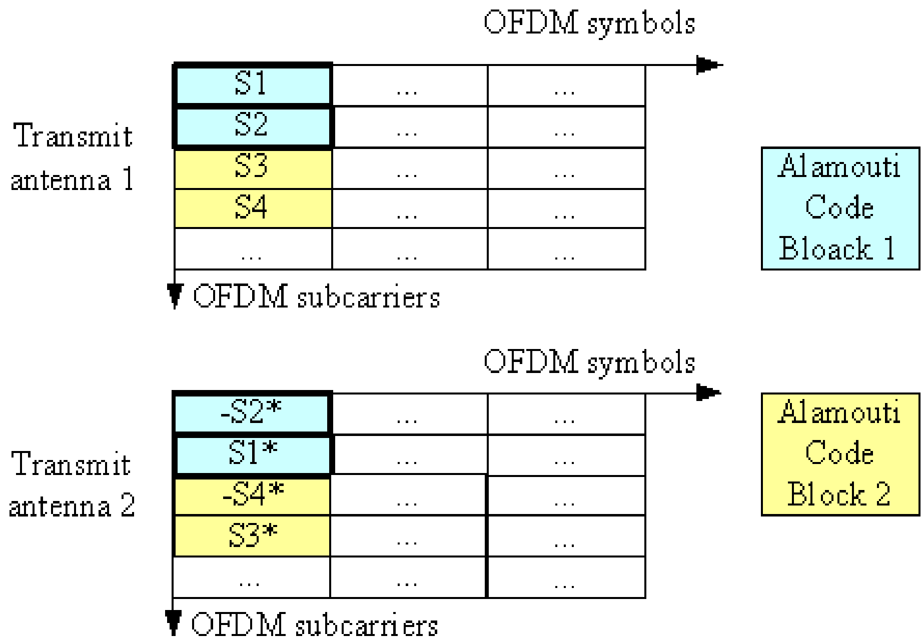


Fig.3.5. Subcarrier mapping in SFBC

As depicted in Fig. 3.5, SFBC also encodes a pair of symbols, $S_1$ and $S_2$ into four variants $S_1, S_2, -S_2^*$ and $S_1^*$. transmits $S_1$ and $-S_2^*$ over a certain sub-carrier from the two antennas. However, the other two variants, $S_2$ and $S_1^*$, are transmitted from the subsequent contiguous or discontinuous sub-carriers.

That is, each symbol (or its positive/negative conjugate) is transmitted from two antennas and over two sub-carriers (rather than over two OFDM symbols in STBC).

## CHAPTER 4

# MIMO-OFDM

***Abstract****: This chapter provides a brief introduction, implementation and mathematical model of the MIMO-OFDM, STBC-OFDM and SFBC-OFDM. OFDM is a multiplexing technique that is considered as a promising candidate for high mobility wireless system due to its high data rate, high spectral efficiency and robustness to the frequency selective channels.*

## 4.1 Principle of Orthogonal Frequency Division Multiplexing

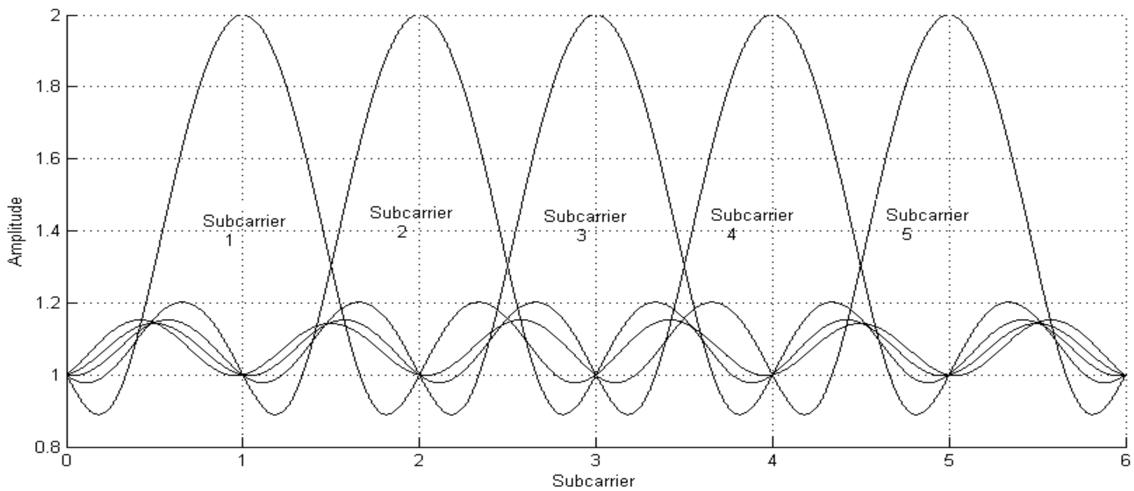


Fig. 4.1. Frequency response of 5 sub-carriers of OFDM signal

As per the name suggested, OFDM divides a high-rate data stream into $N$ parallel streams, which are then transmitted by modulating $N$ distinct carriers which is called as a subcarriers or tones. Symbol duration on each subcarrier thus becomes larger by a factor of $N$. To separate information carried by different subcarriers at receiver, they must be orthogonal.

Conventional Frequency Division Multiple Access (FDMA) can achieve this by having large frequency spacing between adjacent subcarriers as shown in Fig. 5.1. This, however, utilize more spectrum A much narrower spacing of subcarriers can be achieved by OFDM.

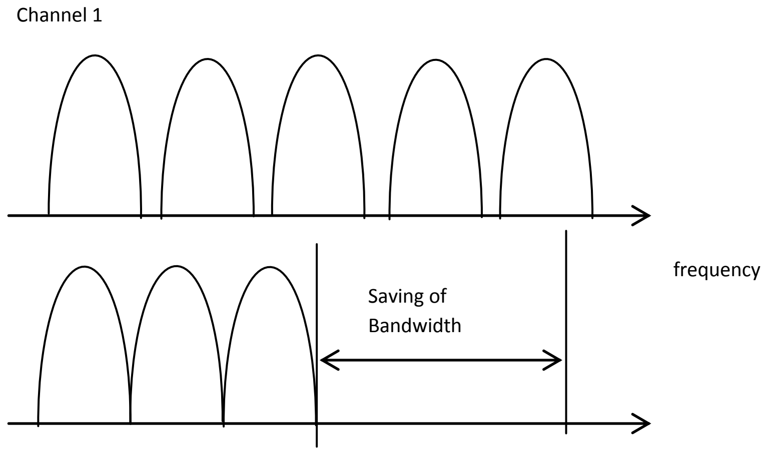


Fig. 4.2. FDMA and OFDM difference

Conventional Frequency Division Multiple Access (FDMA) can achieve this by having large frequency spacing between adjacent subcarriers as shown in Figure 4.2. This, however, utilize more spectrum much narrower spacing of subcarriers can be achieved by OFDM.

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

$$\frac{1}{T_s}\int_0^{Ts} e^{j2\pi f_k t} \times e^{-j2\pi f_l t} = \begin{cases} 0 & \forall \quad k \neq l \\ 1 & \forall \quad k = l \end{cases} \tag{4.1}$$

Fig. 4.1 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 $\frac{\sin x}{x}$ 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 $e^{-j2\pi f_n t}$ and integrating over symbol duration), the data streams of any two subcarriers will not interfere.

## 4.2 Effect of Frequency-Selective Channels on OFDM (Need of Cyclic Prefix)

Initially, 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 $(S\tau / Ts) < 1$. Furthermore, delay dispersion may also be resultant into a loss of orthogonality between the adjacent 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 will see the construction, working principle and performance of Cyclic prefix over OFDM system.

If $x[n]$ is an input data stream which is transmitted through a linear time-invariant discrete channel $h[n]$ then the output $y[n]$ is the discrete-time convolution of the input and the channel impulse response which is given by:

$$\begin{aligned} y[n] &= h[n] * x[n] \\ &= \sum_k h[k] x[n-k] \end{aligned} \tag{4.2}$$

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

$$\begin{aligned} y[n] &= x[n] \otimes h[n] \\ &= h[n] \otimes x[x] \\ &= \sum_k h[k] x[n-k]_N \end{aligned} \tag{4.3}$$

Where $[n-k]_N$ denotes $\left[n-k\right]$ modulo $N$. In other words, $x[n-k]_N$ is a periodic version having $x[n-k]$ with period $N$. It is easily verified that $y[n]$ given by equation (4.3) is also periodic with period $N$.

From the definition of the DFT, circular convolution in time domain is equal to the multiplication in frequency domain.

$$DFT\{x[n]\otimes h[n]\} = X[i]H[i], 0 \le i \le N-1 \tag{4.4}$$

By equation (4.4), if the channel response and input stream are circularly convoluted and $h[n]$ is known at the receiver, the original data sequence $x[n]$ can be recovered by taking the inverse DFT of $(Y[i]/H[i]), 0 \le i \le N-1$.

But the channel output is not a circular convolution but a linear convolution, so we need to convert the linear convolution between the channel input and impulse response into a circular convolution by adding a special prefix to the input called a cyclic prefix (CP), which is described in the next section.

## 4.3 Construction of Cyclic Prefix

Consider a channel input data stream $x[n] = x[0], x[1], ......x[N-1]$ of length $N$ and a discrete-time channel with finite impulse response (FIR) $h[n] = h[0],...,h[\mu]$ having a length of $\mu+1 = Tm/Ts$, where $Tm$ is the channel delay spread and $Ts$ is the sampling time related with the discrete time data stream. The cyclic prefix for $x[n]$ is defined as $\{x[N-m],...,x[N-1]\}$ it consists of the last $\mu$ values of the $x[n]$ sequence.

For each input sequence of length $N$ these last $\mu$ samples are appended to the starting of the sequence. This yields a new sequence $\tilde{x}[n], -\mu \le n \le N-1$, of length $N+\mu$ where $\tilde{x}[-\mu],...,\tilde{x}[N-1] = x[N-m],...,x[N-1], x[0],..., x[N-1]$ .Note that with this definition, $\tilde{x}[n] = x[n]_N$ for $-\mu \le n \le N-1$, which give that $\tilde{x}[n-k] = x[n-k]_N$ for $-\mu \le n-k \le N-1$.

$$\underbrace{\boxed{x[N-\mu]x[N-\mu+1]......x[N-1]}}_{\text{Prefix}}\ \boxed{x[0]x[1]......x[N-\mu-1]}\ \underbrace{\boxed{x[N-\mu]x[N-\mu+1]......x[N-1]}}_{\text{Append last } \mu \text{ symbol}}$$

Fig. 4.3. Insertion of CP of length μ symbols

Now $\tilde{x}[n]$ is output of OFDM transmitter which is act as input of a discrete-time channel with impulse response $h[n]$. The channel output $y[n]$, $0 \le n \le N-1$ is then

$$
\begin{aligned}
y[n] &= \tilde{x}[n]h[n] \\
&= \sum_{k=0}^{\mu} h[k]\tilde{x}[n-k] \\
&= \sum_{k=0}^{\mu} h[k]x[n-k]_N \\
&= x[n] \otimes h[n]
\end{aligned}
$$

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

## `4.4 OFDM Implementation

Implementation Block diagram of OFDM multicarrier modulation is shown in Fig. 5.3. From Figure we can see that the input data sequence is first modulated by a QPSK modulator, which provide a complex symbol stream $X[0], X[1], ..., X[N-1]$ which are in frequency domain. After this symbol stream is passed through a serial-to-parallel converter block, whose output is a set of $N$ parallel QPSK symbols $X[0], X[1], ..., X[N-1]$ associated to the symbols transmitted over each individual subcarriers. In order to generate $x(t)$, these frequency components are first converted into time samples by performing an inverse DFT (IDFT) on these $N$ symbols, which is efficiently implemented using the IFFT algorithm, as the number of complex multiplication and addition is less in IFFT compare to direct IDFT. The IFFT yields the OFDM symbol consisting of the sequence $x[n] = x[0], ..., x[N-1]$ of length $N$.

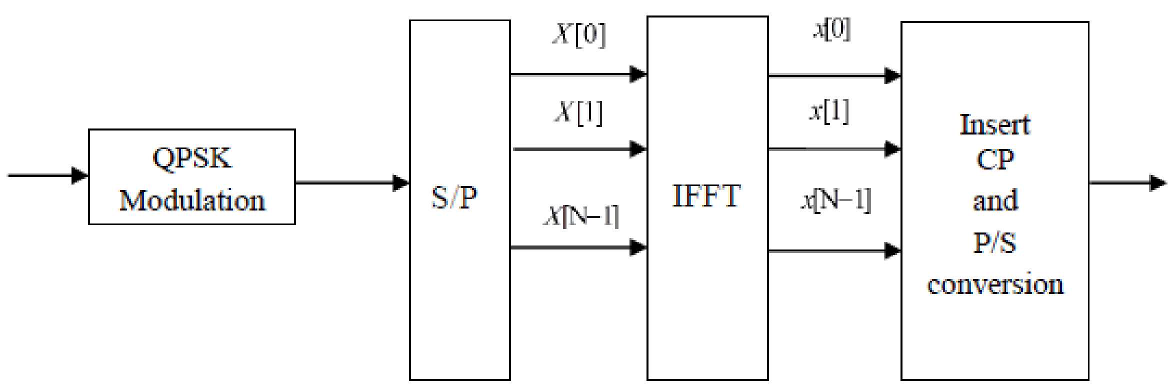


Fig. 4.4. Transmitter block diagram of OFDM with cyclic prefix

Where

$$x[\mathrm{n}] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X[\mathrm{k}] e^{-j2\pi nk/\mathrm{N}}, 0 \le n \le N-1 \tag{4.5}$$

This sequence corresponds to samples of the multicarrier signal: i.e. the multicarrier signal having linearly modulated sub channels. The cyclic prefix is then added to the OFDM symbol, and the resulting time samples $\tilde{x}[n] = \tilde{x}[-\mu], ..., \tilde{x}[N-1] = x[N-\mu], ..., x[0], ..., x[N-1]$ are ordered by the parallel-to-serial converter resulting in the baseband OFDM signal, which is then transmitted through the wireless channel.

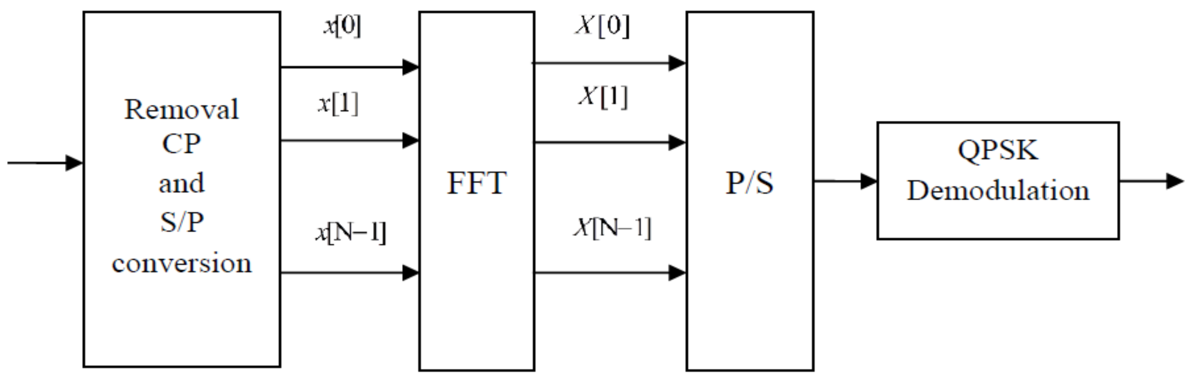


Fig. 4.5. Receiver block diagram of OFDM with cyclic prefix

This transmitted signal is then received at the OFDM receiver as shown in Figure 5.4. The transmitted signal is filtered by the channel impulse response $h[n]$ and corrupted by additive noise, so that the received signal contain the information signal being transmitted and noise introduced by channel which is given by

$$y[n] = x[n] * h[n] + n'[n] \tag{4.6}$$

The cyclic prefix of $y[n]$ consisting of the first $\mu$ samples is then removed. This results in $N$ time samples whose DFT without noise is $Y[i] = H[i]X[i]$. These time samples are serial-to-parallel converted and passed through an FFT. This result in scaled versions of the original symbols $H[i]X[i]$, where $H[i] = H(f)$ 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 information being transmitted. 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 $H[i], i = 0,..., N-1$ 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 dividing or 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 as 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, it's 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 $x[n], 0 \le n \le N-1$ is preceded by $\mu$ null samples.

At the receiver the "tail" of the ISI associated with the end of a given OFDM symbol is added back into the beginning of the symbol, which recreates the effect of a cyclic prefix, so the rest of the OFDM system will functioning 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.

## 4.5 STBC-OFDM

Here, as simplified transmitter diversity technique is adapted for the OFDM system that was proposed in [8] to clarify the use of space-time OFDM in gaining diversity gain over frequency selective fading environment. A block diagram of the space-time OFDM transmitter diversity scheme is shown in Fig. 4.6. At the output of the serial to parallel converter dispose the serial data symbol vectors one pair at a given instant of time.

The first vector denotes in the pair is odd vector, $X_O$ and the second one the even vector $X_e$,For example, if $X_O$, is the P-th block data symbol vector and $X$ , is the (P + 1) -th block vector, they are defined as

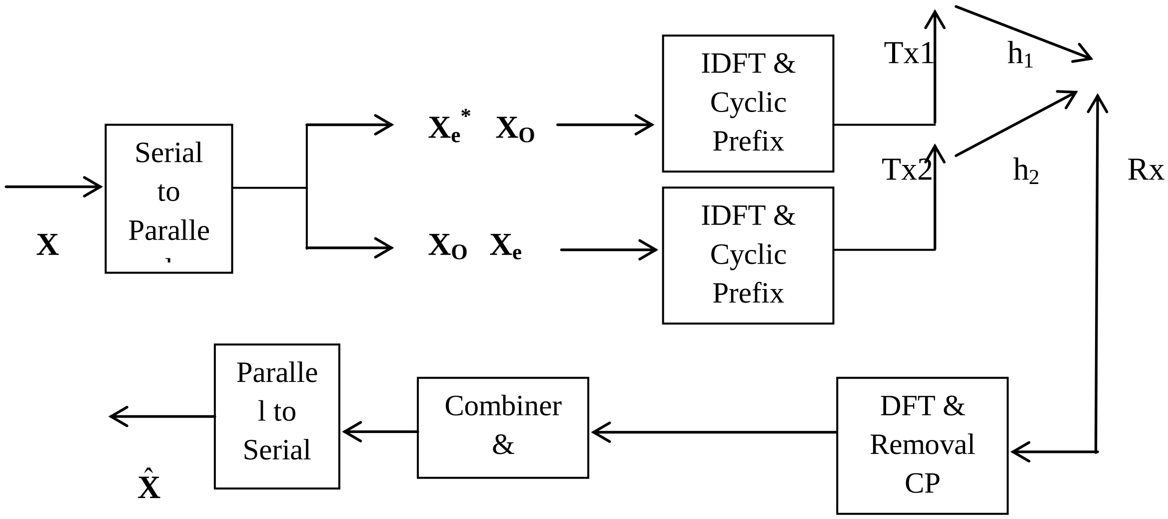


Fig. 4.6. Block diagram STBC-OFDM System

$$\mathbf{X_o} = [X(PN).....X(PN+N-1)]^T$$

$$\mathbf{X_e} = [X(PN+N).....X(PN+2N-1)]^T \tag{4.7}$$

During the first time interval $\mathbf{X_o}$ is transmitted from the first transmitter and $\mathbf{X_e}$ is transmitted from the second transmitter whereas in the second time interval $\mathbf{-X_e^*}$ transmitted from first transmitter and $\mathbf{X_o^*}$ transmitted from second transmitter. Thus, the transmission matrix of space time block code is given by

$$\mathbf{G} = \begin{pmatrix} \mathbf{X_o} & \mathbf{X_e} \\ \mathbf{-X_e^*} & \mathbf{X_o^*} \end{pmatrix}$$

i.e. $\mathbf{X_o}, \mathbf{X_e}$ are OFDM symbol vectors and their conjugates. Suppose $\mathbf{\Lambda_1}$ and $\mathbf{\Lambda_2}$ be two diagonal matrices whose diagonal elements are the DFTs of the respective channel impulse responses, $\mathbf{h_1}$ and $\mathbf{h_2}$. In STBC, an assumption is taken that the channel responses will be static during the two time slots thus transmitted vector is demodulated in their corresponding time slots as

$$\mathbf{Y_1} = \mathbf{\Lambda_1 X_o} + \mathbf{\Lambda_2 X_e} + \mathbf{Z_1}$$

$$\mathbf{Y_2} = -\mathbf{\Lambda_1 X_e^*} + \mathbf{\Lambda_2 X_o^*} + \mathbf{Z_2} \tag{4.8}$$

We can also write equation (4.8) in matrix for as given below

$$\begin{aligned}\mathbf{Y} &= \begin{bmatrix} \mathbf{Y_1} \\ \mathbf{Y_2} \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{\Lambda_o} & \mathbf{\Lambda_e} \\ \mathbf{\Lambda_2^*} & -\mathbf{\Lambda_1^*} \end{bmatrix} \begin{bmatrix} \mathbf{X_o} \\ \mathbf{X_e} \end{bmatrix} + \begin{bmatrix} \mathbf{Z_1} \\ \mathbf{Z_2} \end{bmatrix} \\ &= \mathbf{\Lambda X} + \mathbf{Z} \end{aligned} \tag{4.9}$$

Received signals are estimated with the help of linear combiner like MMSE. So, estimated symbols are given by

$$\hat{\mathbf{X}} = (\mathbf{\Lambda}^H \mathbf{\Lambda} + \frac{1}{SNR}\mathbf{I})\mathbf{\Lambda}^H \quad . \tag{4.10}$$

## 4.6 SFBC-OFDM

In [8] a new two branch SF-OFDM transmitter diversity system has been proposed which is an extension of the simple orthogonal transmitter diversity technique. Space frequency block coded OFDM is preferentially employed over frequency selective channels. The system has been shown in figure. 2. The space-frequency encoder block encodes the data symbol vector $X(n)$ as:

$$\mathbf{X_1} = [X_o(n) \quad -X_1^*(n) \quad .......X_{N-2}(n) \quad -X_{N-1}^*(n)]^T ,$$

$$\mathbf{X_2} = [X_1(n) \quad X_o^*(n) \quad .......X_{N-1}(n) \quad -X_{N-2}^*(n)]^T \tag{4.11}$$

At the transmission side, symbols from $\mathbf{X_1}$ and $\mathbf{X_2}$ are transmitted from transmitter 1 and transmitter 2 simultaneously. The space frequency encoder and decoder describe the whole process by dividing the transmitting vectors in even and odd component vectors.

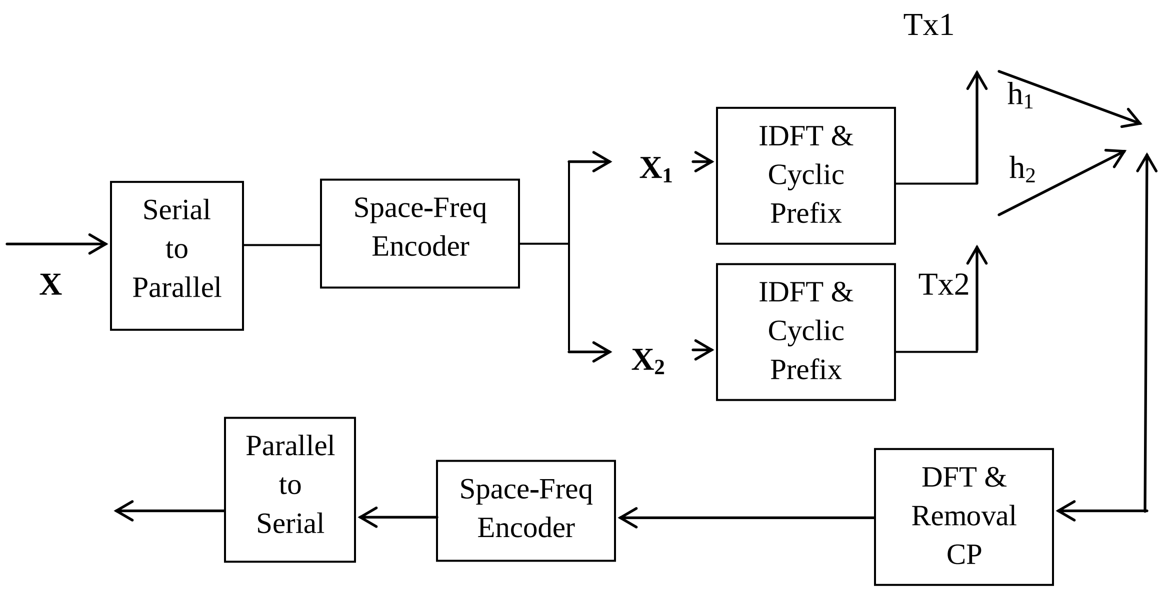


Fig. 4.7. Block diagram SFBC-OFDM System

The length of these vectors will be $N$/2 each. Thus,

$$\mathbf{X_e} = [X_o(n) \quad X_2(n) \quad .......X_{N-4}(n) \quad X_{N-2}(n)]^T ,$$

$$\mathbf{X_o} = [X_1(n) \quad X_3(n) \quad .......X_{N-3}(n) \quad X_{N-1}(n)]^T \tag{4.12}$$

In the same way, $\mathbf{X_{1,e}}$, $\mathbf{X_{1,o}}$, $\mathbf{X_{2,e}}$ and $\mathbf{X_{2,o}}$ are the even and odd component vectors of $\mathbf{X_1}$ and $\mathbf{X_2}$ respectively. Equation (1) can then be given in terms of the even and odd component vectors as

$$\mathbf{X_{1,e}} = \mathbf{X_e},\ \mathbf{X_{1,o}} = -\mathbf{X_o^*} \tag{4.13}$$

$$\mathbf{X_{2,e}} = \mathbf{X_e},\ \mathbf{X_{2,o}} = \mathbf{X_e^*} \tag{4.14}$$

The equivalent space-frequency block code transmission matrix [3] is given by

$$\mathbf{G_2} = \begin{pmatrix} \mathbf{X_e} & \mathbf{X_o} \\ -\mathbf{X_o^*} & \mathbf{X_e^*} \end{pmatrix}$$

Suppose $\mathbf{\Lambda_1}$ and $\mathbf{\Lambda_2}$ be two diagonal matrices such that its elements are the DFTs of the channel impulse responses $\mathbf{h_1}$ and $\mathbf{h_2}$ respectively. The received signal after demodulation is given by

$$\mathbf{Y} = \mathbf{\Lambda_1 X_1} + \mathbf{\Lambda_2 X_2} + \mathbf{Z} \tag{4.15}$$

Or, it can be defined in terms of even and odd components as below

$$\mathbf{Y_e} = \mathbf{\Lambda_{1,e} X_e} + \mathbf{\Lambda_{2,e} X_o} + \mathbf{Z_e}$$

$$\mathbf{Y_o} = -\mathbf{\Lambda_{1,o} X_o^*} + \mathbf{\Lambda_{2,o} X_e^*} + \mathbf{Z_o} \tag{4.16}$$

We can also write equation (4.14) in matrix for as given below

$$\begin{aligned} \mathbf{Y} &= \begin{bmatrix} \mathbf{Y_e} \\ \mathbf{Y_o} \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{\Lambda_{1,e}} & \mathbf{\Lambda 2} \\ \mathbf{\Lambda_{2,o}^*} & -\mathbf{\Lambda_1^*} \end{bmatrix} \begin{bmatrix} \mathbf{X_e} \\ \mathbf{X_o} \end{bmatrix} + \begin{bmatrix} \mathbf{Z_1} \\ \mathbf{Z_2} \end{bmatrix} \\ &= \mathbf{\Lambda X} + \mathbf{Z} \end{aligned} \tag{4.17}$$

Since it is assumed that the response of the channel can be known or estimated at the receiver perfectly, the estimate vector $\hat{\mathbf{X}}$ with the help of linear combiner like MMSE are given by

$$\hat{\mathbf{X}} = (\mathbf{\Lambda}^H \mathbf{\Lambda} + \frac{1}{SNR}\mathbf{I})\mathbf{\Lambda}^H \ . \tag{4.18}$$

# CHAPTER 5

# SINGLE CARRIER FREQUENCY DOMAIN EQUALIZATION

***Abstract:*** *In this chapter we discuss an Alamouti-like scheme [8] for combining space time block-coding (STBC) with single-carrier minimum mean squared error frequency-domain equalization (SC-MMSE FDE).*

## 5.1 Single Carrier Frequency Domain Equalization

In case of single carrier block transmission, cyclic prefix has been employed where length of each block is $N$, to eliminate the inter block interference. At the receiver side, $N+v$ symbols are received, thus $N$ symbols are taken into consideration by eliminating the cyclic prefix respectively. The output will be given as:

$$\mathbf{y} = \mathbf{Hx} + \mathbf{n} \tag{5.1}$$

$\mathbf{y}$, $\mathbf{x}$, $\mathbf{n}$ are received, input and noise symbols having block length $N$. Noise is assumed to be complex, zero mean and uncorrelated. The $N \times N$ channel matrix $\mathbf{H}$ is taken as circulant matrix, therefore the Eigen distribution is given as:

$$\mathbf{H} = \mathbf{Q}^* \mathbf{\Lambda Q} \tag{5.2}$$

Where Q is twiddle matrix and $\mathbf{\Lambda}$ is diagonal matrix of length $N \times N$.

- **SC-MMSE-FDE**

The time domain received $\mathbf{y}$ is thus transformed into the frequency domain such that

$$\begin{aligned} \mathbf{Y} &= \mathbf{Qy} \\ &= \mathbf{\Lambda Qx} + \mathbf{Qn} \\ &= \mathbf{\Lambda X} + \mathbf{N} \end{aligned} \tag{5.3}$$

Therefore, the SC MMSE-FDE is represented by $\mathbf{W}$ which is a $N \times N$ diagonal matrix whose $(i,i)$ elements are given below as: $\mathbf{W}(i,i) = \dfrac{\mathbf{\Lambda}^*(i,i)}{|\mathbf{\Lambda}^*(i,i)|^2 + \dfrac{1}{SNR}}$

Where $SNR = \sigma_x^2 / \sigma_n^2$ and thus transforming back, the output of SC MMSE-FDE is given by $\mathbf{Z = WY}$ is transformed in to time domain, that is represented by

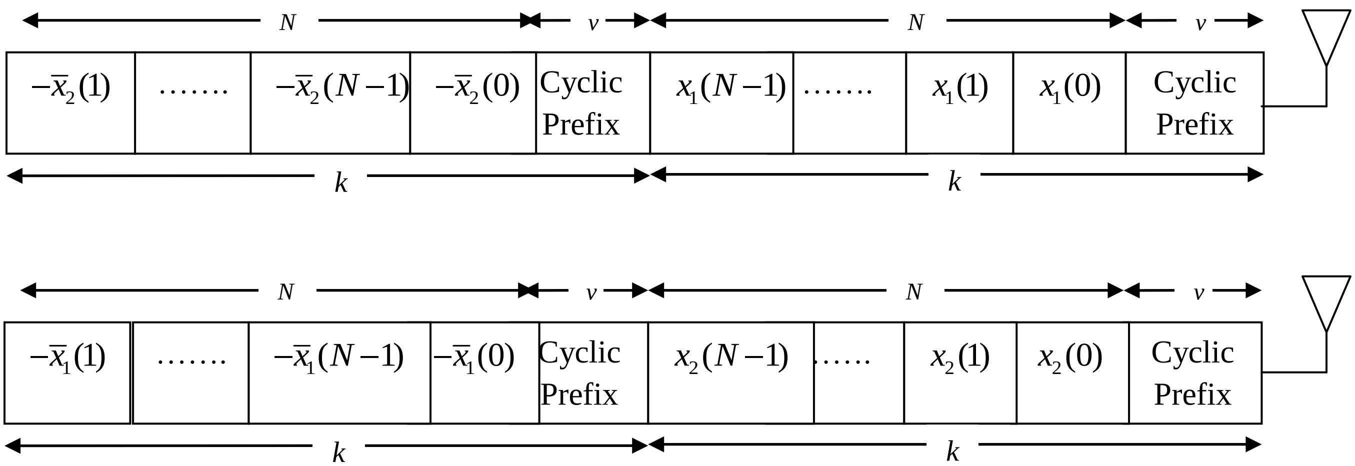


Fig. 5.1. Block diagram of SC-STBC FDE transmission scheme

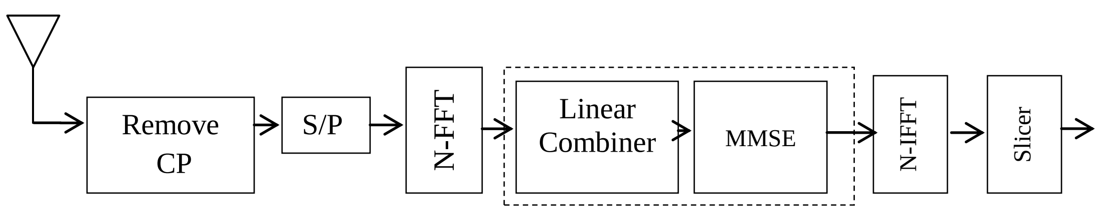


Fig. 5.2. Receiver block diagram of SC-MMSE FDE

$$\mathbf{z = Q^*Z = Q^*\Lambda^*(\Lambda\Lambda^* + \frac{1}{SNR} I_N)^{-1}\Lambda Q x + \tilde{n}} \tag{5.4}$$

Where $\mathbf{\tilde{n} = Q^*WQn}$, thus the final output is obtained by applying hard decision on $\mathbf{x}$. Therefore, the basic difference between the OFDM and FDE is that the decision is made in frequency domain and time domain respectively.

## 5.2 SC-STBC FDE

Denote the $n$th symbol of the th $k$ transmitted block from $i$ antenna by $\mathbf{x}_i^k(n)$. At times $k = 0, 2, 4, ......,$ pairs of length $N$ blocks $\mathbf{x}_1^k(n)$ and $\mathbf{x}_2^k(n)$ ( for $o \le n \le N-1$ ) are generated by an information source. Following SC-STBC FDE transmit diversity scheme

is given by

$$\mathbf{x}_1^{k+1}(n) = -\bar{\mathbf{x}}_2^k((-n)_N) \text{ and } \mathbf{x}_2^{k+1}(n) = -\bar{\mathbf{x}}_1^k((-n)_N)\,. \tag{5.5}$$

for $n = 0,1,..., N-1$ and $k = 0, 2, 4, ......,$ .

Where $\overline{(\bullet)}$ and $(\bullet)_N$ denote complex conjugation and modulo-$N$ operations, respectively. In addition, a cyclic prefix of length $v$ is added to each transmitted block to eliminate IBI and make all channel matrices circulant. Finally, the transmitted power from each antenna is half its value in the single-transmit case so that total transmitted power is fixed.

Now, the received blocks at $k$, $k+1$ instant are, when we have two transmit antenna and one receive antenna:

$$\mathbf{y}^j = \mathbf{H}_1^j \mathbf{x}_1^j + \mathbf{H}_2^j \mathbf{x}_2^j \mathbf{1} + \mathbf{n}^j \tag{5.6}$$

Where $\mathbf{H}_1^j ; \mathbf{H}_2^j$ are circulant channel Matrix from the transmit antenna 1 and 2 respectively, over block j, to receive antenna. Thus, by applying DFT, we get

$$\mathbf{Y}^j = \mathbf{Q}\mathbf{y}^j = \mathbf{\Lambda}_1^j \mathbf{X}_1^j + \mathbf{\Lambda}_2^j \mathbf{X}_2^j + \mathbf{N}^j$$

Where $j = k, k+1$, $\mathbf{X}_i^j = \mathbf{Q}\mathbf{x}_i^j ; \mathbf{N}_i^j = \mathbf{Q}\mathbf{n}_i^j$ and $\mathbf{X}_1^{k+1}(m) = -\bar{\mathbf{X}}_2^k(m); \mathbf{X}_2^{k+1}(m) = \bar{\mathbf{X}}_1^k(m)$ for $m = 0,1,\ldots.., N-1$ and $k = 0, 2, 4, \ldots.$

The other assumption taken is the channel coefficients are assumed to be constant over subsequent blocks, i.e. $\mathbf{H}_i^{k+1} = \mathbf{H}_i^k = \mathbf{H}_i$ and $\mathbf{\Lambda}_i^{k+1} = \mathbf{\Lambda}_i^k = \mathbf{\Lambda}_i$ .

Therefore,

$$\mathbf{Y} = \begin{bmatrix} \mathbf{Y}^k \\ \bar{\mathbf{Y}}^{k+1} \end{bmatrix} = \begin{bmatrix} \mathbf{\Lambda}_1 & \mathbf{\Lambda}_2 \\ \mathbf{\Lambda}_2^* & -\mathbf{\Lambda}_1^* \end{bmatrix} \begin{bmatrix} \mathbf{X}_1^k \\ \mathbf{X}_2^k \end{bmatrix} + \begin{bmatrix} \mathbf{N}^k \\ \bar{\mathbf{N}}^{k+1} \end{bmatrix} = \mathbf{\Lambda X} + \mathbf{N} \qquad (5.7)$$

$\mathbf{\Lambda}$ is orthogonal matrix,

Hence,

$$\tilde{\mathbf{Y}} = \begin{bmatrix} \tilde{\mathbf{Y}}^k \\ \tilde{\mathbf{Y}}^{k+1} \end{bmatrix} = \mathbf{\Lambda}^* \mathbf{Y} = \begin{bmatrix} \tilde{\mathbf{\Lambda}} & 0 \\ 0 & \tilde{\mathbf{\Lambda}} \end{bmatrix} \begin{bmatrix} \mathbf{X}_1^k \\ \mathbf{X}_2^k \end{bmatrix} + \tilde{\mathbf{N}} \qquad (5.8)$$

An estimation symbols $\hat{\mathbf{X}}$ using MMSE equalizer can given by

$$\hat{\mathbf{X}}_k = (\mathbf{\Lambda}^* \mathbf{\Lambda} + \frac{1}{SNR} \mathbf{I}_2)^{-1} \tilde{\mathbf{Y}} = \begin{bmatrix} \hat{\mathbf{X}}_1^k \\ \hat{\mathbf{X}}_2^k \end{bmatrix} \qquad (5.9)$$

In the SC-FDE detection is made in the time domain whereas channel equalization is performed in the frequency domain. Therefore, the estimated transmitted symbols can be obtained as by taking N/2 point IDFT of $\hat{\mathbf{X}}_1^k$ and $\hat{\mathbf{X}}_2^k$.

## 5.3 Simulation Results

The bit-error rate (BER) performance of multiple antenna systems (the SC-FDE and the STBC SC-FDE and) was investigated through MATLAB simulator. We assume that the channel state information (CSI) is known at the receiver. The CP length was set to the channel order. The typical urban (TU) channel with symbol duration of $t_S = 3.69\mu s$ was used for the simulation, as in the proposed third-generation TDMA cellular standard EDGE [50].

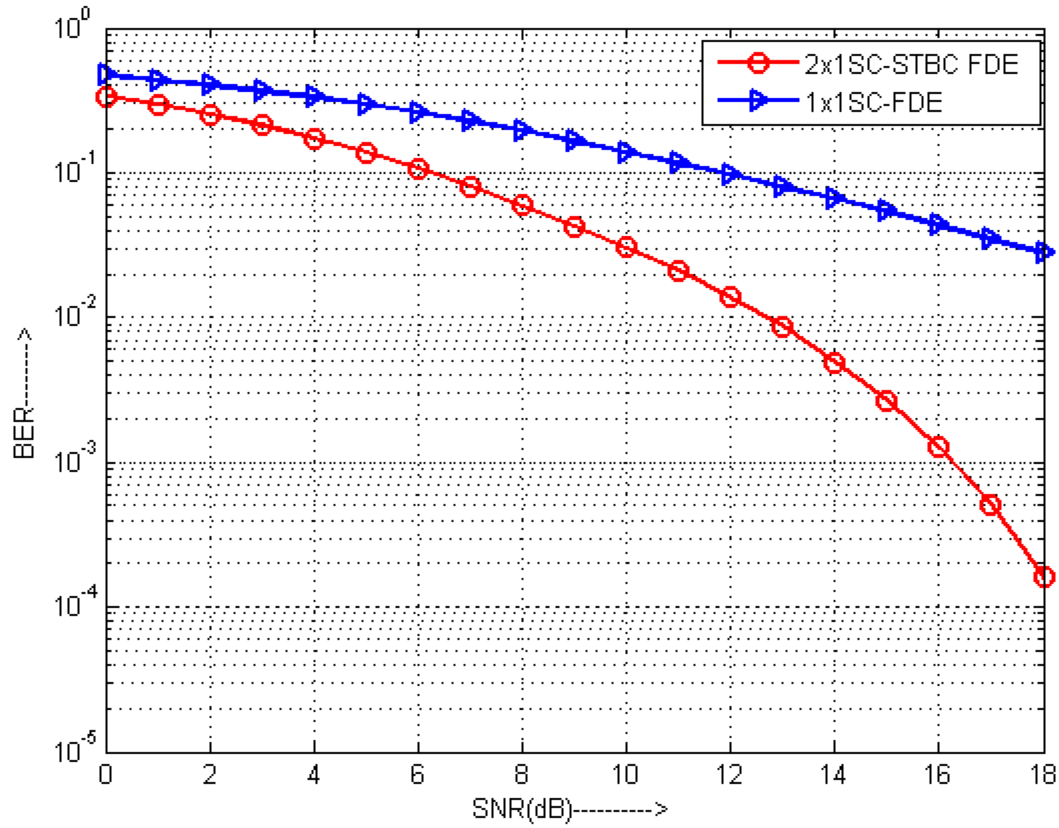


Fig. 5.3. BER of SC MMSE-FDE and STBC for EDGE TU channel with QPSK modulation and $N = 64$

Fig. 5.3 shows the significant improvement achieved in SC MMSE-FDE performance when combining it with the STBC scheme, especially at high SNR where effects of diversity are more pronounced.

## CHAPTER 6

# SC- SFBC FREQUENCY DOMAIN EQUALIZATION

***Abstract:*** *In this chapter we present Space–frequency block-coding (SFBC) with single-carrier frequency-domain equalization (SC-FDE) in wireless fading channels. This chapter focuses on the design of low complexity zero forcing (ZF) equalizer for the combining receiver and proposed interference canceller receiver for SC-SFBC FDE system with two transmits antennas and one receive antenna to mitigates the effect of ISI in frequency selective fading environment.*

## 6.1 SC-SFBC FDE

The Alamouti STBC scheme for fading environments has been captured with time-domain equalization [47]. The benefits of STBC and SC-FDE were combined [45] for slowly varying channels, where the channel response is assumed to be constant for two consecutive symbols [34, 35].

However, in practical scenarios, a high speed mobile causes fast fading which is characterized by different channel response for consecutive symbols. In order to suppress the effects of fast fading in wireless mobile environment, an SFBC SC-FDE scheme is proposed. The effectiveness of mitigating the fast fading distortion in frequency non-selective fading environment lies under SFBC as it utilizes two adjacent subcarriers over two transmit antennas [47], [48], unlike STBC which uses two adjacent time intervals. OFDM system can also be coded with SFBC to achieve high symbol error rate performance [45]. As the transmit sequence of an SC system is processed in time domain, so it is not relevant to directly apply SFBC to SC system [41]

### 6.1.1 Transmit Diversity

We adopt SC system with two transmit antennas and one receive antenna. In SC system transmit sequence is in time domain so we cannot directly applied SFBC scheme to the SC system. A block diagram is shown in Fig. 6.1, consist of fast Fourier transform (FFT) and inverse FFT (IFFT) blocks and give orthogonal structure of SFBC. We denotes $x_j(n)$ as n[th] signal of the transmitted block from j[th] antenna. Using the IFFT, the transmitted signal of the first antenna can be given as,

$$
\begin{aligned}
x_1(n) &= \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} X_1(k) W_N^{-mn} \\
&= \frac{1}{\sqrt{N}} \sum_{v=0}^{(N/2)-1} (X1(2v) + W_N^{-n} X_1(2v+1)) W_{(N/2)}^{-nv} \\
&= \frac{1}{\sqrt{2}} (x^e(n) + W_N^{-n} x^o(n)), n = 0,1,....N-1
\end{aligned}
\tag{6.1}
$$

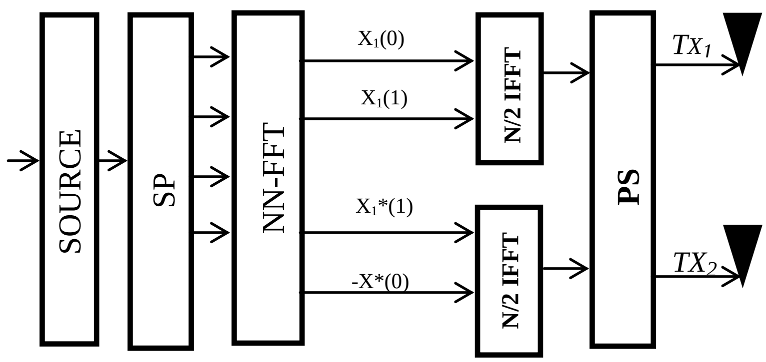


Fig. 6.1. SC-SFBC Transmit scheme

Where $x^o(n)$ and $x^e(n)$, $n = 0,1......, N-1$ are transmitted symbols given by

$$
x^e(n) = \sqrt{\frac{2}{N}} \sum_{v=0}^{(N/2)-1} X_1(2v) W_{(N/2)}^{-nv}
$$

$$
x^o(n) = \sqrt{\frac{2}{N}} \sum_{v=0}^{(N/2)-1} X_1(2v+1) W_{(N/2)}^{-nv}. \tag{6.2}
$$

Since $x^o(n)$ and $x^e(n)$ can be represented as periodic in $n$ with period $N/2$, we can rewrite them as $x^o((n))_{(N/2)}$ and $x^e((n))_{(N/2)}$. In equation (1) and (2) the scaling factor of IFFT was chosen in such a way so that the transmitted normalize power from each transmitted antenna is equal to unity. Similarly, transmitted signal $x_2(n)$ from second antenna can be derived with the help of SFBC and the discrete Fourier transform (DFT) property [49] $x^{*}(-n)_{N} \Leftrightarrow X^{*}(k)$, $k$, $n = 0,1......, N-1$.

The constellation symbols are splinted in to odd and even by demultiplexer, as follows:

$$x^{o}(z) = x(2z+1)\,,\; x^{e}(z) = x(2z)\,,\; z = 0,1,.....,\frac{N}{2}-1\,. \tag{6.3}$$

$$\begin{aligned} x_2(n) &= \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} X_2(k) W_N^{-mn} \\ &= \frac{1}{\sqrt{N}} \sum_{v=0}^{(N/2)-1} (-X_1^{*}(2v+1) + W_N^{-n} X_1^{*}(2v)) W_{(N/2)}^{-nv} \\ &= \frac{1}{\sqrt{2}} (-x^{o^{*}}(n) + W_N^{-n} x^{e^{*}}(n)) \end{aligned} \tag{6.4}$$

Each block of length $N$ is appended in head of with a length-$N_p$ cyclic prefix to reduce effect of inter block interference (IBI) and channel matrices become circulant.

### 6.1.2 Received Signal Model

We consider perfect channel estimation is known at the receiver. The transmitted signal $\mathbf{x}_j : \left\{ x_j(n) \right\}_{n=0}^{N-1}$, where $j$ =1,2, the received block $\mathbf{y} : \left\{ y(n) \right\}_{n=0}^{N-1}$ **(**after removing the first received $N_p$ symbols corresponding to the cyclic prefix) is given by

$$\mathbf{y} = \mathbf{h}_1\mathbf{x}_1 + \mathbf{h}_2\mathbf{x}_2 + \mathbf{n} \tag{6.5}$$

where $\mathbf{hi}$, i=1, 2, are the $N \times N$ right circulant channel matrices from the first and second transmit antennas, respectively, with first column is the channel impulse response (CIR) and $\mathbf{n}$ is the zero-mean complex Gaussian noise (AWGN) symbols with length $N \times 1$. Taking the DFT of equation (5) by multiplying twiddle matrix $\mathbf{W}$ of length $N \times N$ .

$$\mathbf{Y} = \mathbf{W}\mathbf{h}_1\mathbf{W}^{H}\mathbf{X}_1 + \mathbf{W}\mathbf{h}_2\mathbf{W}^{H}\mathbf{X}_2 + \mathbf{N}',\; \mathbf{N}' = \mathbf{W}\mathbf{n} \tag{6.6}$$

where $(\bullet)^H$ denotes complex conjugate transpose. Let $\boldsymbol{\rho}_i = \mathbf{W}\mathbf{h}_i\mathbf{W}^{\mathbf{H}}$ ,$i$=1,2, are $N \times N$ diagonal matrices whose diagonal element represent DFT of channel impulse response. We can split equation (6.6) in to even and odd frequency components as

$$\begin{aligned} Y(2k) &= \rho_1(2k)X_1(2k) - \rho_2(2k)X_1^*(2k+1) + N'(2k) \\ Y^*(2k+1) &= \rho_1^*(2k+1)X_1^*(2k+1) + \rho_2^*(2k+1)X_1^*(2k) + N'(2k+1) \end{aligned} \tag{6.7}$$

Where $k = 0,1,....,\frac{N}{2}-1$ and $\rho_j(k)$ represents channel frequency response (CFR) of $k^{\text{th}}$ subcarrier between the $j^{\text{th}}$ transmitter and the receiver.

## 6.2 Detection of SFBC SC-FDE for Frequency Flat Fading channel

For the case of frequency flat (i.e. no loss of quasi-static assumption) there is no ISI. In the STBC system, by assuming that channel frequency response is identical between two adjacent symbol intervals [8]. Similarly, in SFBC system, channel frequency response is identical between two adjacent subcarrier i.e. $\rho_1(2k) \approx \rho_1(2k+1)$ and $\rho_2(2k) \approx \rho_2(2k+1)$ for large $N$ [34]. Then equation (6.7) can be written in matrix form as

$$\begin{aligned} \mathbf{Y}_k &= \begin{pmatrix} Y(2k) \\ Y(2k+1) \end{pmatrix} \\ &= \begin{pmatrix} \rho_1(2k) & -\rho(2k) \\ \rho_2^*(2k) & \rho_1^*(2k) \end{pmatrix} \begin{pmatrix} X_1(2k) \\ X_1^*(2k+1) \end{pmatrix} + \begin{pmatrix} N'(2k) \\ N'(2k+1) \end{pmatrix} \\ &= \boldsymbol{\rho}_k \mathbf{X}_k + \mathbf{N}_k^{'} \end{aligned} \tag{6.8}$$

An estimation of $\mathbf{Y}_k$ using MMSE equalizer can be obtained as

$$\begin{aligned} \hat{\mathbf{Y}}_k &= \boldsymbol{\rho}_k^H \mathbf{Y}_k \\ &= \begin{pmatrix} \hat{\rho}_k & 0 \\ 0 & \hat{\rho}_k \end{pmatrix} \mathbf{X}_k + \hat{\mathbf{N}}_k^{'} \end{aligned}$$

Where $\hat{\rho}_k = |\rho_1(2k)|^2 + |\rho_2(2k)|^2$, $\hat{\mathbf{N}}_k^{'} = \boldsymbol{\rho}_k^H \mathbf{N}_k^{'}$. Here we have seen that $\boldsymbol{\rho}_k^H \boldsymbol{\rho}_k$ is a diagonal matrix, and hence due to no loss of 'quasi-static' assumption there will not be any inter-symbol interference (ISI). An estimated symbol obtains as:

$$\hat{\mathbf{X}}_k = (\boldsymbol{\rho}_k^H \boldsymbol{\rho}_k + \frac{1}{SNR}\mathbf{I}_2)^{-1}\hat{\mathbf{Y}}_k = \begin{pmatrix} \hat{X}^e(k) \\ \hat{X}^o(k) \end{pmatrix} = \begin{pmatrix} \hat{X}_1(2k) \\ \hat{X}_1^*(2k+1) \end{pmatrix} \tag{6.9}$$

## 6.3 Proposed ZF Combining Receiver

In this section, we propose low complexity zero forcing (LZF) equalizer for the combining receiver for a communication system with two transmit antennas and one receive antenna for SFBC SC-FDE. The proposed design of the equalizer outperforms the matched filter receiver for fast fading (i.e. loss of 'quasi-static' assumption) mobile environments, and also provides an advantage of lower computational complexity at the receiver than classical ZF equalizer. ISI caused due to loss of 'quasi-static' assumption i.e. $\rho_1(2k) \neq \rho_1(2k+1)$ and $\rho_2(2k) \neq \rho_2(2k+1)$. Then equation (6.7) can be given in matrix form as

$$\begin{aligned} \mathbf{Y}_k &= \begin{pmatrix} Y(2k) \\ Y(2k+1) \end{pmatrix} \\ &= \begin{pmatrix} \rho_1(2k) & -\rho(2k) \\ \rho_2^*(2k+1) & \rho_1^*(2k+1) \end{pmatrix} \begin{pmatrix} X_1(2k) \\ X_1^*(2k+1) \end{pmatrix} + \begin{pmatrix} N'(2k) \\ N'(2k+1) \end{pmatrix} \\ &= \boldsymbol{\rho}_k \mathbf{X}_k + \mathbf{N}_k^{'} \end{aligned} \tag{6.10}$$

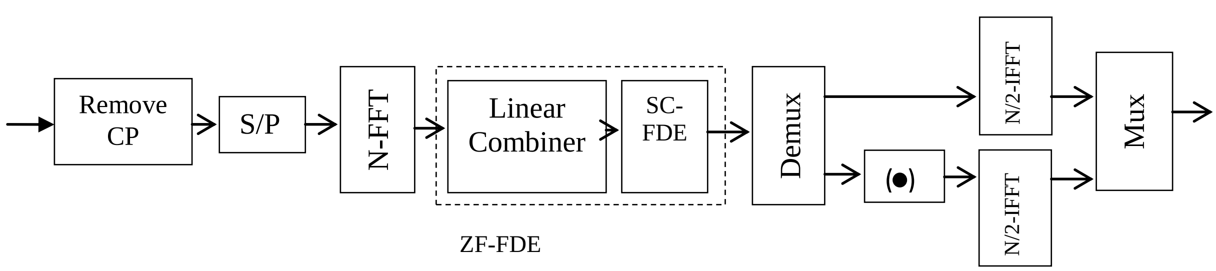


Fig. 6.2. ZF-combining receiver

Here, the channel is considered to be fast fading mobile environment, the channel coefficients of adjacent subcarriers are considered not equal. Applying matched filter (MF) equalization to Eq. (6.10) to obtain the estimated symbols as

$$\begin{aligned}\hat{\mathbf{Y}}_k &= \boldsymbol{\rho}_k^H \mathbf{Y}_k \\ &= \begin{bmatrix} D_{2k} & \varepsilon \\ \varepsilon^* & D_{2k+1} \end{bmatrix} \begin{bmatrix} X_1(2k) \\ X_1^*(2k+1) \end{bmatrix} + \begin{bmatrix} \rho_1^*(2k) & \rho(2k+1) \\ -\rho_2^*(2k) & \rho_1(2k+1) \end{bmatrix} \begin{bmatrix} N'(2k) \\ N'(2k+1) \end{bmatrix}\end{aligned} \tag{6.11}$$

Where $D_{2k} = |\rho_1(2k)|^2 + |\rho_2(2k)|^2$, $D_{2k+1} = |\rho_1(2k+1)|^2 + |\rho_2(2k+1)|^2$ are the diversity gains, and $\varepsilon = \rho_2(2k+1)\rho_1^*(2k+1) - \rho_1^*(2k)\rho_2(2k)$, is the interference term.

In order to perfectly estimate the received symbols, eliminate the interference term produced by the MF receiver and to provide the full diversity gains as in classical ZF receiver, we propose a low complexity ZF equalizer matrix given as

$$\boldsymbol{\rho}_k^{LZF} = \begin{bmatrix} \rho_1^*(2k) & \dfrac{\rho_2(2k+1)}{G_k} \\ -\rho_2^*(2k) & \dfrac{\rho_1(2k+1)}{G_k} \end{bmatrix} \tag{6.12}$$

Where $G_k = \rho_2(2k+1)\rho_1^*(2k+1) \big/ \rho_1^*(2k)\rho_2(2k)$. Using the matrix defined in Eq. (6.12), the estimated symbol matrix is given as

$$\begin{aligned}\hat{\mathbf{Y}}_k &= \boldsymbol{\rho}_k^{LZF} \mathbf{Y}_k \\ &= \begin{bmatrix} D'_{2k} & 0 \\ 0 & D'_{2k+1} \end{bmatrix} \begin{bmatrix} X_1(2k) \\ X_1^*(2k+1) \end{bmatrix} + \boldsymbol{\rho}_k^{LZF} \begin{bmatrix} N'(2k) \\ N'(2k+1) \end{bmatrix}\end{aligned} \tag{6.13}$$

Where $D'_{2k} = |\rho_1(2k)|^2 + \dfrac{|\rho_2(2k)|^2}{G_k}$, $D'_{2k+1} = |\rho_1(2k+1)|^2 + \dfrac{|\rho_2(2k+1)|^2}{G_k^*}$ are the diversity gains of the low complexity ZF equalizer. An estimated symbol obtains as:

$$\hat{\mathbf{X}}_k = (\boldsymbol{\rho}_k^{LZF}\boldsymbol{\rho}_k)^{-1}\hat{\mathbf{Y}}_k = \begin{pmatrix} \hat{X}^e(k) \\ \hat{X}^o(k) \end{pmatrix} = \begin{pmatrix} \hat{X}_1(2k) \\ \hat{X}_1^*(2k+1) \end{pmatrix} \tag{6.14}$$

In the SC-FDE detection is made in the time domain whereas channel equalization is performed in the frequency domain. Therefore, the estimated transmitted symbols can be obtained as by taking N/2 point IDFT of $\hat{\mathbf{X}}_1(2k)$ and $\hat{\mathbf{X}}_2(2k+1)$. The proposed low complexity ZF receiver results in zero interference term from the adjacent subcarriers in the estimated symbols. This method also avoids the inversion of channel matrix, as in classical ZF, which makes it lesser computationally complex than classical ZF for SC-FDE.

## 6.4 Proposed Interference Canceller Receiver

In this section, we propose interference canceller detector for SFBC SC-FDE. The proposed receiver estimates and mitigates the ISI caused due to loss of 'quasi-static' assumption i.e. $\rho_1(2k) \neq \rho_1(2k+1)$ and $\rho_2(2k) \neq \rho_2(2k+1)$. Then equation (6.7) can be given in matrix form as

$$\begin{aligned} \mathbf{Y}_k &= \begin{pmatrix} Y(2k) \\ Y(2k+1) \end{pmatrix} \\ &= \begin{pmatrix} \rho_1(2k) & -\rho(2k) \\ \rho_2^*(2k+1) & \rho_1^*(2k+1) \end{pmatrix} \begin{pmatrix} X_1(2k) \\ X_1^*(2k+1) \end{pmatrix} + \begin{pmatrix} N'(2k) \\ N'(2k+1) \end{pmatrix} \\ &= \boldsymbol{\rho}_{k,LQS}\mathbf{X}_{k,LQS} + \mathbf{N}'_{k,LQS} \end{aligned} \tag{6.15}$$

Where $\mathbf{Y}_{k,LQS}$ represent received signal under condition of loss of 'quasi-static' assumption. First we model ISI caused due to loss of 'quasi-static' assumption. We split $\boldsymbol{\rho}_{k,LQS}$ in to two parts $\boldsymbol{\rho}_{nisi}$ and $\boldsymbol{\rho}_{isi}$, so that

$$\boldsymbol{\rho}_{k,LQS} = \boldsymbol{\rho}_{nisi} + \boldsymbol{\rho}_{isi} \tag{6.16}$$

Where

$$\boldsymbol{\rho}_{nisi} = \begin{pmatrix} \rho_1(2k) & -\rho(2k) \\ \rho_2^*(2k) & \rho_1^*(2k) \end{pmatrix}, \tag{6.17}$$

And

$$\boldsymbol{\rho}_{isi} = \begin{pmatrix} 0 & 0 \\ \Delta\rho_2(2k) & \Delta\rho_1(2k) \end{pmatrix}. \tag{6.18}$$

Where $\Delta\rho_1(2k) = \rho_1^*(2k+1) - \rho_1^*(2k)$ and $\Delta\rho_2(2k) = \rho_2^*(2k+1) - \rho_2^*(2k)$ Based on equation (6.16), (6.17) and (6.18), we can write (6.15) as

$$\mathbf{Y}_{k,LQS} = \boldsymbol{\rho}_{nisi}\mathbf{X}_k + \underbrace{\boldsymbol{\rho}_{isi}\mathbf{X}_k}_{\text{Loss of ‘quasi-static’}} + \mathbf{N}_k^{'}$$

An estimation of $\mathbf{Y}_{k,LQS}$ can be obtained as

$$\begin{aligned} \hat{\mathbf{Y}}_{k,LQS} &= \boldsymbol{\rho}_{nisi}^H \mathbf{Y}_{k,LQS} \\ &= \underbrace{\begin{pmatrix} \hat{\rho}_k & 0 \\ 0 & \hat{\rho}_k \end{pmatrix}\mathbf{X}_k}_{\text{Required signal}} + \underbrace{\boldsymbol{\rho}_{nisi}^H \boldsymbol{\rho}_{isi}\mathbf{X}_k}_{\text{ISI}} + \underbrace{\boldsymbol{\rho}_{nisi}^H \hat{\mathbf{N}}_k^{'}}_{\text{Noise}} \end{aligned} \tag{6.19}$$

Where $\hat{\rho}_k = \left|\rho_1(2k)\right|^2 + \left|\rho_2(2k)\right|^2$

As can be seen, in equation (6.19) give details of required signal, ISI and noise present in estimate $\hat{\mathbf{Y}}_{k,LQS}$. Based on model of received signal in equation (6.19) and knowledge of the matrices $\boldsymbol{\rho}_{nisi}$ and $\boldsymbol{\rho}_{isi}$ formation of the propose interference estimation and cancellation receiver process given as below.

**(1)** For each space frequency block coded $k$, estimate the $\hat{\mathbf{X}}_k$ from equation (6.19) ignoring inter-symbol interference.

**(2)** For each space frequency block coded $k$, estimate the inter-symbol interference from (6.19) (i.e. second term in equation (6.19)).

**(3)** Cancel the ISI, which is estimated in step-2 from $\hat{\mathbf{Y}}_{k,LQS}$.

Base on the above discussion, we summarized interference cancellation algorithm for L stage as bellow.

*At starting stage: set L=1.*

*Evaluate*

$$\hat{\mathbf{Y}}_k^{(L)} = \boldsymbol{\rho}_{nisi}^H \mathbf{Y}_{k,LQS}$$

*Repeat:*

*Estimate*

$$\hat{\mathbf{X}}_k^{(L)} = (\boldsymbol{\rho}_{nisi}^H \boldsymbol{\rho}_{nisi} + \frac{1}{SNR}\mathbf{I}_2)^{-1} \hat{\mathbf{Y}}_k^{(L)}$$

*Cancel ISI*

$$\hat{\mathbf{Y}}_k^{(L+1)} = \hat{\mathbf{Y}}_k^{(1)} - \boldsymbol{\rho}_{nisi}^H \boldsymbol{\rho}_{isi} \hat{\mathbf{X}}_k^{(L)}$$

$L = L+1$

go to *Repeat.*

It is noted that $\boldsymbol{\rho}_{nisi}^H \boldsymbol{\rho}_{nisi}$ is a diagonal matrix, due to this reason inversion becomes simple. In this algorithm, perfect channel estimation is required.

In the SC-FDE detection is made in the time domain whereas channel equalization is performed in the frequency domain. Therefore, the estimated transmitted symbols can be obtained as by taking N point IDFT of $\hat{\mathbf{X}}_k^{(L)}$.

## 6.5 Simulation Results

### 6.5.1 BER Performance of SC-STBC FDE and SC-SFBC FDE in Different Fading Channel

The bit-error rate (BER) performance of two transmits and one received antenna systems (the SC-STBC FDE and the SC-SFBC FDE) was investigated through MATLAB simulator. We assume that the channel state information (CSI) is known at the receiver. The CP length was set to the channel order. The typical urban (TU) channel with symbol duration of $t_S = 3.69\mu s$ was used for the simulation, as in the proposed third-generation TDMA cellular standard EDGE [50].

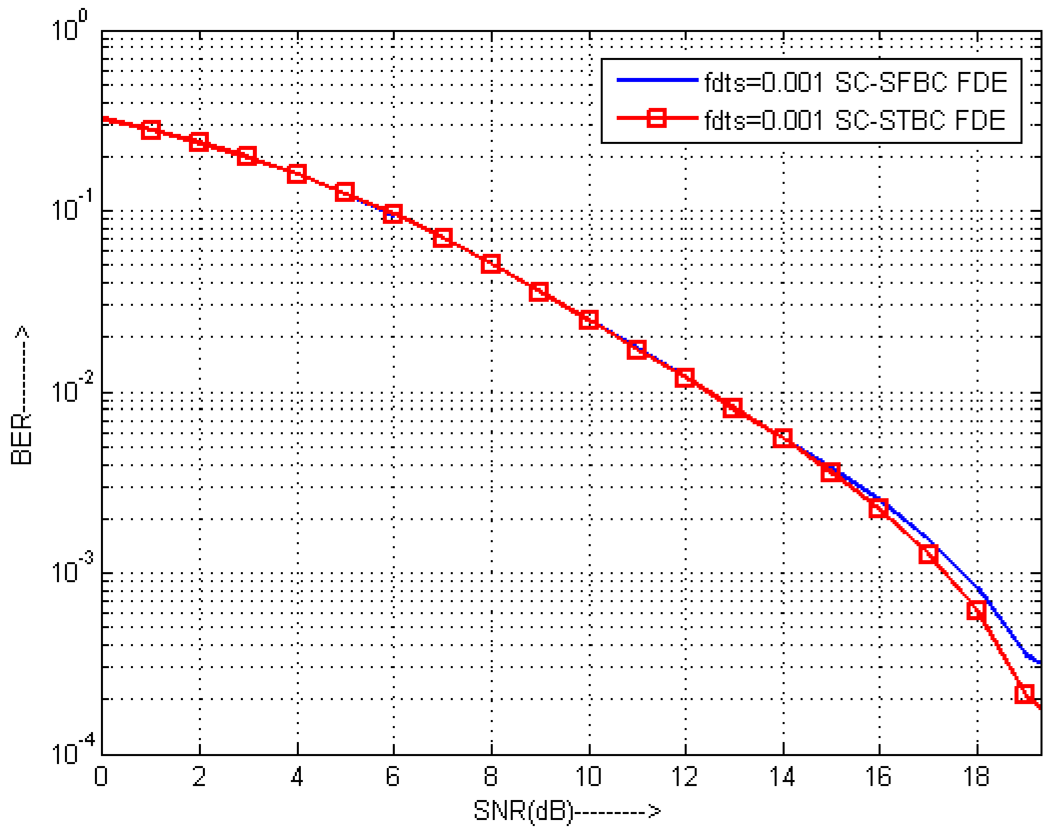


Fig. 6.3. BER performance of SC transmission schemes for EDGE TU slow fading channel with QPSK modulation and $N = 64$.

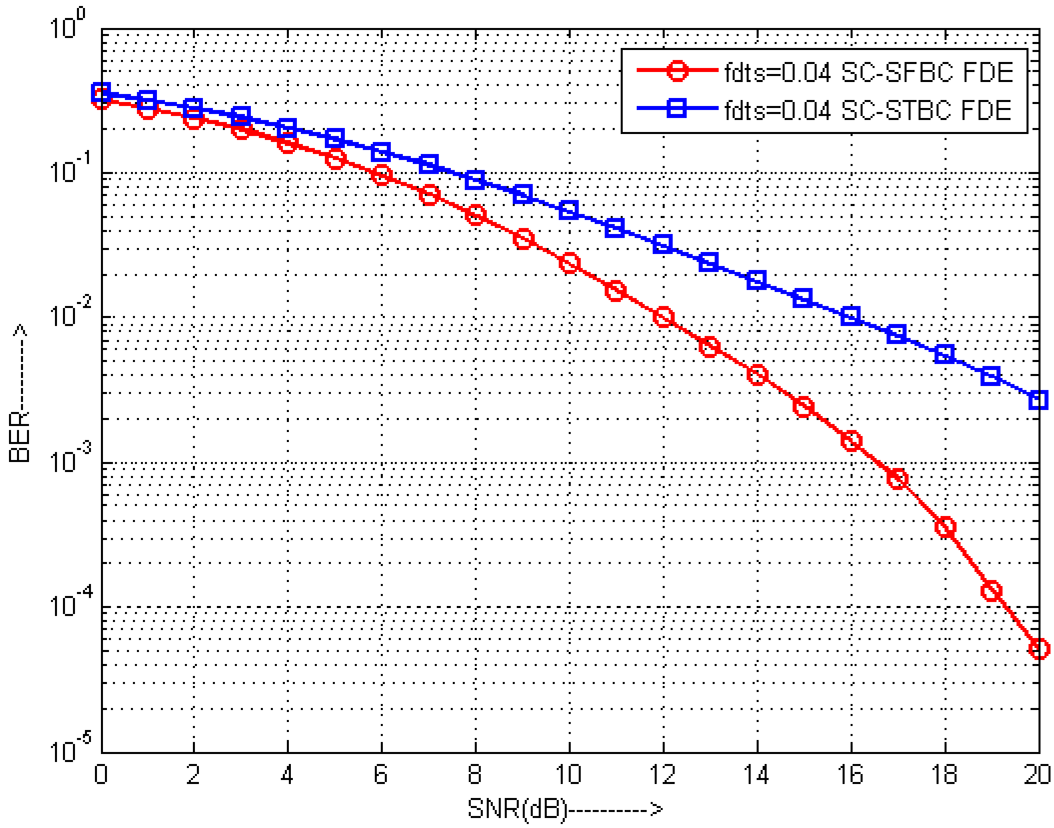


Fig. 6.4. BER performance of SC transmission schemes for EDGE TU fast fading channel with QPSK modulation and $N = 64$

Fig. 6.3 compares the performance of the SC-STBC FDE and SC-SFBC FDE in slow fading channel (normalize Doppler frequency $f_d t_s = 0.001$). SC-STBC FDE and SC-SFBC FDE give almost similar performance at lower SNR value but at high SNR, SC-STBC FDE give better performance then SC-SFBC FDE.

Fig. 6.4 compares the performance of the SC-STBC FDE and SC-SFBC FDE fast fading channel (normalize Doppler frequency $f_d t_s = 0.04$) and transmission employing the SFBC scheme significantly outperforms the one combined with the STBC scheme, when there exists severe Doppler spread. In the simulations, we assumed that the channel gain is perfectly known at the receiver. The effects of channel estimation on our performance comparisons are not addressed here and are a subject for future research.

### 6.5.2 BER Performance of Proposed Low Complexity ZF combining Receiver

The Bit error rate (BER) performance for two transmit antenna and one received antenna (SC-SFBC FDE) system was investigated through MATLAB simulator. Considering a fast fading Rayleigh channel modeled using Jakes' model with normalized Doppler frequency $f_d t_s = 0.01$, with QPSK modulation for the symbols to be transmitted, and block duration of symbols is taken to be $t_s = 3.69\mu s$ seconds as proposed in the third generation TDMA cellular standard EDGE [50].

The bit error rate (BER) performance, averaged over 100 iterations, was computed for proposed ZF SC-SFBC FDE system, and the results were compared with the classical ZF receiver. It is demonstrated in Fig. 6.7, that the proposed low complexity ZF receiver is equivalent to classical ZF for SC-FDE at each signal to noise ratio.

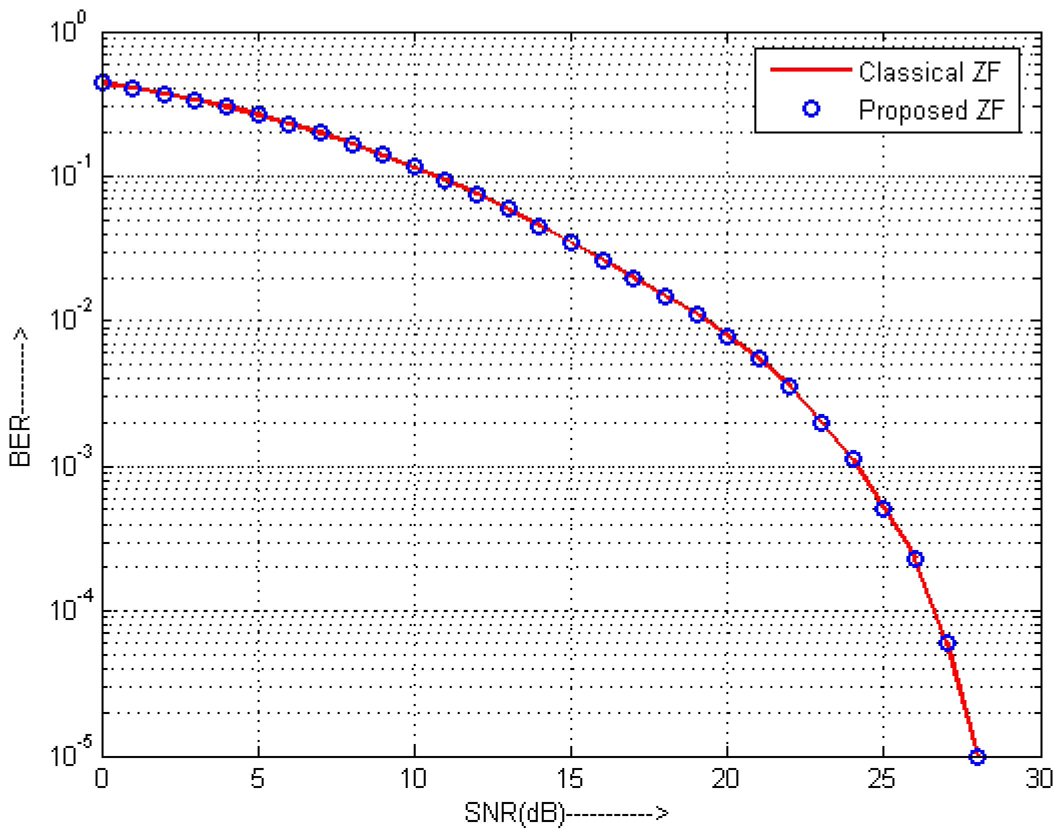


Fig. 6.5. BER of SC SFBC ZF-FDE for time varying Rayleigh fading channel with QPSK modulation and $N = 64$

### 6.5.3 BER Performance of Proposed Interference Canceller Receiver

The bit-error rate (BER) performance of SC-SFBC FDE for two transmits antenna and one received antenna system was investigated through MATLAB simulator. We assume that the channel state information (CSI) is known at the receiver. System parameters used are $N = 64$ symbols with QPSK modulation, $f_c = 5$ GHz and maximum Doppler shift=200 Hz.

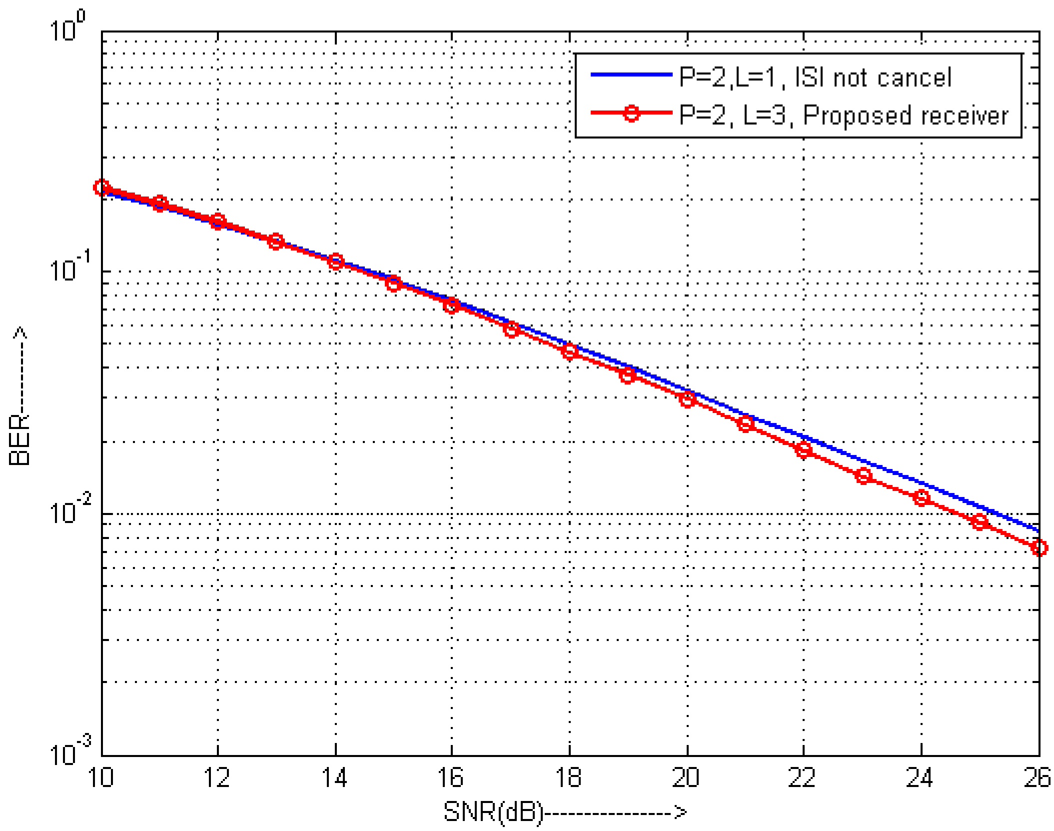


Fig. 6.6. BER performance of SC-SFBC FDE for frequency selective Rayleigh fading, $N = 64$, QPSK modulation and P = 2 paths

In Fig. 6.5 and Fig. 6.6, we plot SER performance of proposed interference canceller receiver in frequency-selective Rayleigh fading channel with P=2 (i.e. small channel delay spread) and P=4 (i.e. large channel delay spread) equal power paths respectively for SC-SFBC FDE.

It can be observed that for *P*=2 the induced ISI is small due to small channel delay spread and hence there is no major improvement in performance. However for P=4 the induced ISI is high due to large channel delay spread, and proposed receiver give significant performance gain (e.g. SNR between 10 dB to 16 dB).

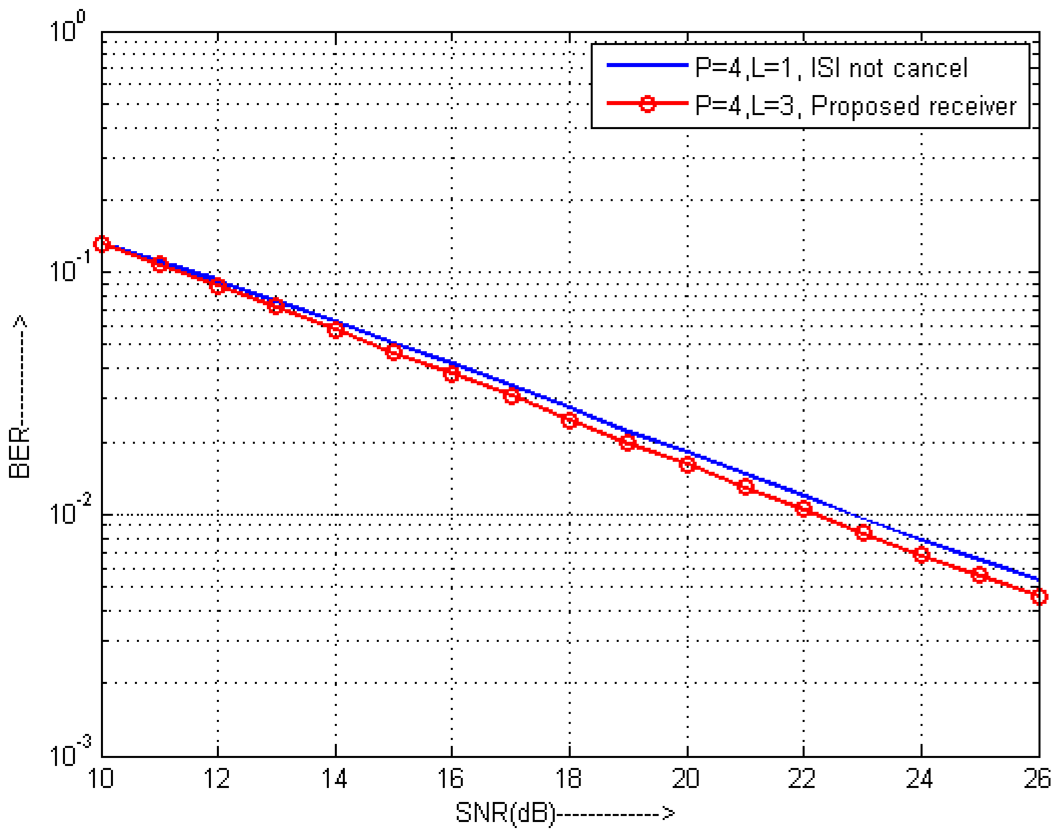


Fig. 6.7. BER performance of SC-SFBC FDE for frequency selective Rayleigh fading, $N = 64$, QPSK modulation and P = 4 paths

# CHAPTER 7

# CONCLUDING REMARKS AND FUTURE SCOPE

## 7.1 Concluding Remarks

Single-carrier transmit-diversity scheme for frequency-selective channels, the scheme combines the advantages of an Alamouti-like space–time block-coding scheme and FFT based single-carrier frequency- domain equalization.

Simulation results in Fig. 5.3 show that significant improvement achieved in SC MMSE-FDE performance when combining it with the STBC scheme. In Fig. 6.3 and Fig. 6.4, it has been shown that an SC-SFBC FDE system achieves better performance than an SC-STBC FDE system in fast fading environments and similar performance in slow fading environment.

We proposed ZF combining receiver in chapter-6 (section-6.3) is better than the classical ZF receiver and MF receiver for fast fading mobile environments for SC-FDE systems. The proposed system takes into account the advantages offered by SFBC and SC-FDE, and combines them into much lesser computationally complex ZF receiver. Simulation results in Fig. 6.5, it has been shown that the proposed ZF receiver outperforms the MF receiver and matches with the classical ZF receiver, with the added advantage of lesser computations at the receiver for symbol estimation and detection.

We proposed an interference canceller algorithm in chapter-6 (section-6.4) for frequency-selectivity induced ISI in SFBC SC-FDE system. In this algorithm we estimate the ISI and then cancelled it. We can repeat this procedure in multiple stages to reduce ISI induced error-floors. Our simulation results in Fig. 6.6 and Fig. 6.7 for frequency-selective channel shown that proposed detector mitigates the effect of ISI.

## 7.2 Future Scope

- The proposed interference canceller receiver algorithm and proposed ZF combining receiver can be extended to space time frequency coded (STFC) [53] as well.
- In Single carrier space frequency block coding detection techniques based on Minimum Mean Squared Error criterion under mobile environment can be extended for symbol detection techniques such as M-ary QAM with power efficient signal constellation.

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