PERFORMANCE OF SPACE-TIME-FREQUENCY CODES WITH PRE-DFT PROCESSING OF MIMO-OFDM

A Project report submitted in partial fulfillment of the requirements for the award of the degree of

BACHELOR OF TECHNOLOGY

IN

ELECTRONICS AND COMMUNICATION ENGINEERING

Submitted by

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DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

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(UGC AUTONOMOUS)
(*Permanently Affiliated to AU, Approved by AICTE and Accredited by NBA & NAAC with 'A' Grade*)
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PRAGNANAM BRAHMA

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CERTIFICATE

This is to certify that the project report entitled **"PERFORMANCE OF SPACE-TIME-FREQUENCY CODES WITH PRE-DFT PROCESSING OF MIMO-OFDM"** submitted by **J.V. Lalitha devi (317126512136), Mohammad Ansar (317126512150), N.Santhoshi Sunandha (317126512151), K. Asha Surya Sirisha (318126512L27) , P.Moses Divya Rohit (316126512160)** in partial fulfillment of the requirements for the award of the degree of **Bachelor of Technology** in **Electronics & Communication Engineering** of Andhra University, Visakhapatnam is a record of bonafide work carried out under my guidance and supervision.

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ABSTRACT

Subcarrier based space processing was conventionally employed in Orthogonal Frequency Division Multiplexing (OFDM) systems under Multiple-Input and Multiple-Output (MIMO) channels to achieve optimal performance. At the receiver of such systems, multiple Discrete Fourier Transform (DFT) blocks, each corresponding to one receive antenna, are required to be used. This induces considerable complexity. In this project, we propose a pre-DFT processing scheme for the receiver of MIMO-OFDM systems with space-time-frequency coding. With the proposed scheme, the number of DFT blocks at the receiver can be any numbers from one to the number of receive antennas, thus enabling effective complexity and performance tradeoff.

Using the pre-DFT processing scheme, the number of input signals to the space-time-frequency decoder can be reduced compared with the subcarrier-based space processing. Therefore, a high dimensional MIMO system can be shrunk into an equivalently low dimension one. Due to the 'dimension reduction, both the complexity of the decoder and the complexity of channel estimation can be reduced. In general, the weighting coefficients calculation for the pre-DFT processing scheme should be relevant to the specific space-time-frequency code employed. In this project, we propose a simple universal weighting coefficients calculation algorithm that can be used to achieve excellent performance for most practical space-time-frequency coding schemes. This makes the design of the pre-DFT processing scheme independent of the optimization of the space-time-frequency coding, which is desirable for multiplatform systems.

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List of Abbreviations

Abbreviations	Descriptions
AWGN	Adaptive White Gaussian Noise
BER	Bit Error Rate
DMT	Discrete Multi-tone Modulation
SFNs	Signal Frequency Networks
PDC	Personal Digital Cellular
SDMA	Space Division Multiple Access
BTS	Base Trans-receiver Station
EM	Electra magnetic
GSM	Global System for Mobile communication
VSB	Vestigial Side Band
ICI	Inter-Carrier- Interference
ISI	Inter Symbol Interference
MFNs	Multi-Frequency broadcast Networks
SFNs	Single Frequency broadcast Networks
CSI	Channel State Information
CTF	Space-Time-Frequency coding
WLAN	Wireless Local Area Networks
CIRs	Channel Impulse Response

CHAPTER 1

INTRODUCTION

1.1 Overview

In recent years, orthogonal frequency division multiplexing (OFDM) has emerged as a promising air-interface technique for various wireless communication systems. The concepts of multiple-input multiple-output (MIMO) have been under development for many years for both wired and wireless systems. For high data rate wideband wireless communications, Orthogonal Frequency Division Multiplexing (OFDM) can be used with Multiple-Input and Multiple-Output (MIMO) technology to achieve superior performance. In conventional MIMO-OFDM systems, subcarrier-based space processing was employed to achieve optimal performance. However, it requires multiple discrete Fourier transform/inverse DFT (DFT/IDFT) blocks, each corresponding to one receive/transmit antenna. Even though DFT/IDFT can be efficiently implemented using fast Fourier transform/inverse FFT (FFT/IFFT), its complexity is still a major concern for OFDM implementation. In addition, the use of multiple antennas requires the baseband signal processing components to handle multiple input signals, thus inducing considerable complexity for the decoder and the channel estimator at the receiver.

MULTIPLE receive antennas can be employed with orthogonal frequency-division multiplexing (OFDM) to improve system performance, where space diversity is achieved using subcarrier-based space combining. However, in subcarrier-based space combining, it is required that multiple discrete Fourier transform (DFT) processing, each per receive antenna, be used. As a result, such systems are quite complicated, because the complexity of DFT is a major concern for system implementation. Recently, some schemes have been proposed to reduce the number of DFT blocks required. In the principle of orthogonal designs, the number of DFT blocks is reduced to a half with 3 dB performance degradation. In the received time-domain OFDM symbols from each antenna are first weighted and then combined before the

DFT processing. By doing so, the number of DFT blocks required is reduced to one. Its OFDM with multiple transmit and multiple receive antennas (MIMO) is investigated, and a reduced- complexity algorithm is proposed to reduce the number of DM' blocks required to one, One important issue in the proposed pre-DFT processing scheme for M1MO-OFDM systems with space-time-frequency coding is the calculation or the weighting coefficients before the DFT processing. In general, the weighting coefficients calculation are specific to the space-time-frequency coding scheme.

Motivated by the work in and based upon Eigen analysis, we propose a receive space-diversity scheme to effectively trade off system performance and complexity. The scheme in can be regarded as a special case of our proposed scheme. However, we will show that the good performance is only applicable when the number of distinct paths in the channel is very limited. When the number of distinct paths is large, it will be shown that more DFT blocks are needed. In our proposed scheme, the received signals are weighted and combined both before and after the DFT processing, and the margin of the performance improvement decreases along with the increase of the number of DFT blocks. As a result, system complexity and performance can be effectively traded off. When the weighting coefficients are obtained assuming perfect channel information, we will show that the maximum number of DFT blocks required is the minimum of the number of receive antennas and the number of distinct paths in the channel. Such an achievement is obtained without performance loss, compared with subcarrier-based space combining. When the number of distinct paths is larger than the number of receive antennas, the number of DFT blocks required is not necessarily equal to the number of receive antennas. For example, extensive simulation results will show that good performance can be achieved by using only two DFT blocks for four receive antennas and eight-ray channels. An OFDM system with differential modulation, where the weighting coefficients before the DFT processing are obtained using the signal covariance matrix. In this case, when the number of distinct paths is less than the number of receives antennas, the proposed scheme can achieve better performance than the subcarrier-based space combining scheme, but with lower complexity.

In general, the weighting coefficients before the DFT processing can be calculated assuming that the CSIs are explicitly available. In this paper, we will show that the weighting coefficients can also be obtained using the signal space method without the explicit knowledge of the CSIs. This helps to reduce the complexity of channel estimation required by the space-time-frequency decoding since the number of equivalent channel branches required to be estimated in the proposed scheme can be reduced from the number of receive antennas to the number of DFT blocks.

1.2 Problem Definition

In several applications, especially for Subcarrier based space processing was conventionally employed in Orthogonal Frequency Division Multiplexing (OFDM) systems under Multiple-Input and Multiple-Output (MIMO) channels to achieve optimal performance. At the receiver of such systems, multiple Discrete Fourier Transform (OFT) blocks, each corresponding to one receive antenna, are required to be used. The pre-DFT processing scheme, the number of DFT blocks at the receiver can be any numbers from one to the number of receive antennas, thus enabling effective complexity and performance tradeoff.

1.3 Technical approach

Using the pre-DFT processing scheme, the number of input signals to the space-time-frequency decoder can be reduced compared with the subcarrier-based space processing. Therefore, a high dimensional MIMO system can be shrunk into an equivalently low dimension one. Due to the dimension reduction, both the complexity of the decoder and the complexity of channel estimation can be reduced. In general, the weighting coefficients calculation for the pre-OFT processing scheme should be relevant to the specific space-time-frequency code employed.

1.4 Organization of the Thesis

The thesis of the mini project is organized in seven chapters including introduction and conclusion and two data sheets.

Chapter-2 explains the digital modulation scheme and radio propagation.

Chapter-3 describes the mathematical expressions of the orthogonal frequency-division multiplexing (OFDM) and explanation of the OFDM system.

Chapter-4 deals with multiple-input multiple-output (MIMO)- orthogonal frequency-division multiplexing (OFDM), and explanation of the MIMO-OFDM system.

Chapter-5 explains the space- time-frequency coding.

Chapter-6 deals with the total description of the PRE-DFT processing.

Chapter-7 provides simulation results conclusions and future scope of the project.

CHAPTER 2

DIGITAL MODULATION SCHEME AND RADIO PROPOGATION

2.1 Modulation

In communication, modulation is the process of varying a periodic waveform, in order to use that signal to convey a message over a medium. Normally a high frequency waveform is used as a carrier signal. The three key parameters of a sine wave are frequency, amplitude, and phase, all of which can be modified in accordance with a low frequency information signal to obtain a modulated signal. There are 2 types of modulations Analog modulation and digital modulation. In analog modulation, an information-bearing analog waveform is impressed on the carrier signal for transmission whereas in digital modulation, an information-bearing discrete-time symbol sequence (digital signal) is converted or impressed onto a continuous-time carrier waveform for transmission.

2.2 DIGITAL MODULATION

Nowadays, digital modulation is much popular compared to analog modulation. The move to digital modulation provides more information capacity, compatibility with digital data services, higher data security, better quality communications, and quicker system availability. The aim of digital modulation is to transfer a digital bit stream over an analog band pass channel or a radio frequency band. The changes in the carrier signal are chosen from a finite number of alternative symbols. Digital modulation schemes have greater capacity to convey large amounts of information than analog modulation schemes. There are three major classes of digital modulation techniques used for transmission of digitally represented data.

- Amplitude-shift Keying (ASK)
- Frequency-shift keying (FSK)
- Phase-shift keying (PSK)

All convey data by changing some aspect of a base signal, the carrier wave, (usually a sinusoid) in response to a data signal. In the case of PSK, the phase is changed to represent the data signal. There are two fundamental ways of utilizing the phase of a signal in this way

- By viewing the phase itself as conveying the information, in which case the demodulator must have a reference signal to compare the received signal's phase against; or

- By viewing the change in the phase as conveying information — differential.

schemes, some of which do not need a reference carrier (to a certain extent).

A convenient way to represent PSK schemes is on a constellation diagram. This shows the points in the Argand plane where, in this context, the real and imaginary axes are termed the in-phase and quadrature axes respectively due to their 90° separation. Such a representation on perpendicular axes lends itself to straightforward implementation. The amplitude of each point along the in-phase axis is used to modulate a cosine (or sine) wave and the amplitude along the quadrature axis to modulate a sine (or cosine) wave.

2.3 PHASE SHIFT KEYING (PSK)

PSK is a modulation scheme that conveys data by changing, or modulating, the phase of a reference signal (i.e. the phase of the carrier wave is changed to represent the data signal). A finite number of phases are used to represent digital data. Each of these phases is assigned a unique pattern of binary bits; usually each phase encodes an equal number of bits. Each pattern of bits forms the symbol that is represented by the particular phase. A convenient way to represent PSK schemes is on a constellation diagram (as shown in figure). This shows the points in the Argand plane where, in this context, the real and imaginary axes are termed the in-phase and quadrature axes respectively due to their 90° separation. Such a representation on perpendicular axes lends itself to straightforward implementation. The amplitude of each point along the in-phase axis is used to modulate a cosine (or sine) wave and the amplitude along the quadrature axis to modulate a sine (or cosine) wave.

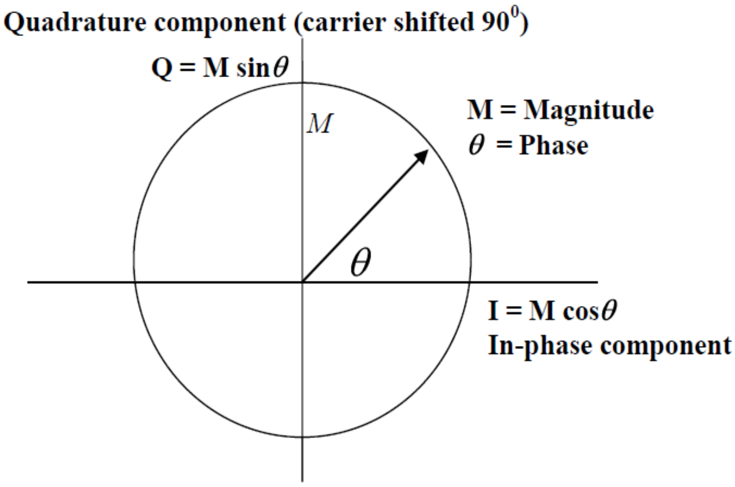


Fig 2.2 Constellation diagram for PSK

In PSK, the constellation points chosen are usually positioned with uniform angular spacing around a circle. This gives maximum phase-separation between adjacent points and thus the best immunity to corruption. They are positioned on a circle so that they can all be transmitted with the same energy. In this way, the module of the complex numbers they represent will be the same and thus so will the amplitudes needed for the cosine and sine waves. Two common examples are binary phase-shift keying (BPSK) which uses two phases, and quadrature phase-shift keying (QPSK) which uses four phases, although any number of phases may be used. Since the data to be conveyed are usually binary, the PSK scheme is usually designed with the number of constellation points being a power of 2. Notably absent from these various schemes is 8-PSK. This is because its error-rate performance is close to that of I6-QAM — it is only about 0.5 dB better— but its data rate is only three-quarters that of 16-QAM. Thus 8-PSK is often omitted from standards and, as seen above, schemes tend to 'jump' from QPSK to 16-QAM (8-QAM is possible but difficult to implement).

Any digital modulation scheme uses a finite number of distinct signals to represent digital data. PSK uses a finite number of phases; each assigned a unique pattern of binary bits. Usually, each phase encodes an equal number of bits. Each pattern of bits forms the symbol that is represented by the particular phase. The demodulator, which is designed specifically for the symbol-set used by the modulator, determines the phase of the received signal and maps it back to the symbol it represents, thus recovering the original data. This requires the receiver to be able to compare the phase of the received signal to a reference signal — such a system is termed coherent (and referred to as CPSK).

Alternatively, instead of using the bit patterns to set the phase of the wave, it can instead be used to change it by a specified amount. The demodulator then determines the changes in the phase of the received signal rather than the phase itself. Since this scheme depends on the difference between successive phases, it is termed differential phase-shift keying (DPSK). DPSK can be significantly simpler to implement than ordinary PSK since there is no need for the demodulator to have a copy of the reference signal to determine the exact phase of the received signal (it is a non-coherent scheme). In exchange, it produces more erroneous demodulations. The exact requirements of the particular scenario under consideration determine which scheme is used.

Applications of PSK

Owing to PSK's simplicity, particularly when compared with its competitor quadrature amplitude modulation, it is widely used in existing technologies.

The wireless LAN standard, uses a variety of different PSKs depending on the data-rate required. At the basic-rate of 1 Mbit/s, it uses DBPSK (differential BPSK). To provide the extended-rate of 2 Mbit/s, DQPSK is used. In reaching 5.5 Mbit/s and the full-rate of 11 Mbit/s, QPSK is employed, but has to be coupled with complementary code keying. The higher-speed wireless LAN standard has eight data rates 6, 9, 12, 18, 24, 36, 48 and 54 Mbit/s. The 6 and 9 Mbit/s modes use OFDM modulation where each sub-carrier is BPSK modulated. The 12 and 18 Mbit/s modes use OFDM with QPSK. The fastest four modes use OFDM with forms of quadrature amplitude modulation.

Because of its simplicity BPSK is appropriate for low-cost passive transmitters, and is used in RFID standards such as ISO/IEC 14443 which has been adopted for biometric passports, credit cards such as American Express's Express-pay, and many other applications. Frequency bands 868-915 MHz using BPSK and at 2.4 GHz using OQPSK.

For determining error-rates mathematically, some definitions will be needed

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{\frac{-t^2}{2}} dt = \frac{1}{2} erfc\left(\frac{x}{\sqrt{2}}\right), x \geq 0 \qquad \ldots.2.1$$

Eb = Energy-per-bit

Es = Energy-per-symbol = kE_b with k bits per symbol

Tb = Bit duration

Ts= Symbol duration

No / 2 = Noise power spectral density (W/Hz)

Pb = Probability of bit-error : Ps= Probability of symbol-error

Q(x) will give the probability that a single sample taken from a random process with zero-mean and unit-variance Gaussian probability density function will be greater or equal to x. It is a scaled form of the complementary Gaussian error function.

The error-rates quoted here are those in additive white Gaussian noise (AWGN). These error rates are lower than those computed in fading channels, hence, are a good theoretical benchmark to compare with in this project, BPSK, QPSK are selected to be the digital modulation schemes for OFDM.

2.3.2 Binary Phase-Shift Keying (BPSK)

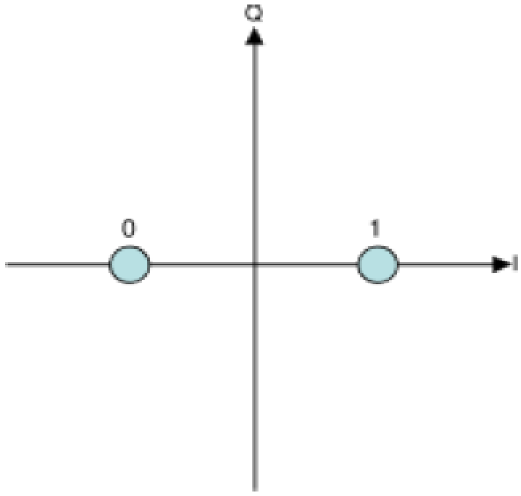


Fig 2.2 Constellation diagram for BPSK.

BPSK (also sometimes called PRK, Phase Reversal Keying) is the simplest form of PSK. It uses two phases which are separated by 180° and so can also be termed 2-PSK. It does not particularly matter exactly where the constellation points are positioned, and in this figure they are shown on the real axis, at 0° and 180°. This modulation is the most robust of all the PSK.s since it takes serious distortion to make the demodulator reach an incorrect decision. It is, however, only able to modulate at 1 bit/symbol (as seen in the figure) and so is unsuitable for high data-rate applications when bandwidth is limited. In the presence of an arbitrary phase-shift introduced by the communications channel, the demodulator is unable to tell which constellation point is which. As a result, the data is often differentially encoded prior to modulation.

Binary data is often conveyed with the following signals

$S_0 = \sqrt{\frac{2E_b}{T_b}}\,[\cos(2\pi f_c t + \pi)] = -\sqrt{\frac{2E_b}{T_b}}\,[\cos(2\pi f_c t)]$ for binary ‘0’ ..2.2

$S_0= \sqrt{\frac{2E_b}{T_b}}\,[\cos(2\pi f_c t)]$ for binary "1" …2.3

$\Phi(t) = \sqrt{\frac{2}{T_b}}\,[\cos(2\pi f_c t)]$ … 2.4

Where 1 is represented by $\sqrt{E_b}\ \Phi(t)$ and 0 is represented by $-\sqrt{E_b}\ \Phi(t)$.

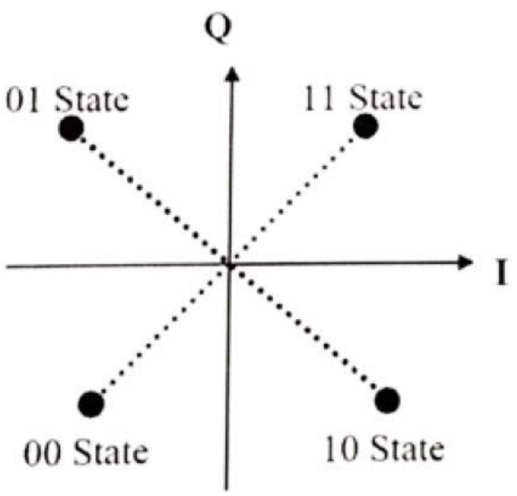
Bit error rate (BER) of BPSK in AWGN can be calculated as

$P_b = Q\left(\sqrt{\frac{2E_b}{N_0}}\right)$ Or $P_b = \frac{1}{2}\,erfc\left(\sqrt{\frac{E_b}{N_0}}\right)$ …2.5

Since there is only one bit per symbol , this is also the symbol error rate.

2.3.3 Quadrature Phase Shift Keying (QPSK)

QPSK is a multilevel modulation techniques, it uses 2 bits per symbol to represent each phase. Compared to BPSK, it is more spectrally efficient but requires more complex receiver. Figure 2.3 shows the constellation diagram for QPSK with Gray coding. Each adjacent symbol only differs by one bit. Sometimes known as quaternary or quadra phase PSK or 4-PSK, QPSK uses four points on the constellation diagram, equi-spaced around a circle.



Phase of carrier

n/4, 3n/4, 5n/4, 7n/4

Fig 2.3 Constellation Diagram for QPSK

With four phases, QPSK can encode two bits per symbol, shown in the diagram with Gray coding to minimize the BER - twice the rate of BPSK but half the bandwidth needed. Although QPSK can be viewed as a quaternary modulation, it is easier to see it as two independently modulated quadrature carriers. With this interpretation, the even (or odd) bits are used to modulate the in-phase component of the carrier, while the odd (or even) bits are used to modulate the quadrature-phase component of the carrier. BPSK is used on both carriers and they can be independently demodulated.

The implementation of QPSK is more general than that of BPSK and also indicates the implementation of higher-order PSK. Writing the symbols in the constellation diagram in terms of the sine and cosine waves used to transmit them

$$S_i\,(t) = \sqrt{\frac{2E_s}{T}}\,[\,\cos(2\pi f_c t + (2i-1)\pi/4)] \qquad .....2.6$$

This yields the four phase's $\pi/4$, $3\,\pi/4$, $5\,\pi/4$ and $7\,\pi\,/4$ as needed. This results in a two-dimensional signal space with unit basis functions.

$$\Phi_1(t) = \sqrt{\frac{2}{T_s}}\,[\,\cos(2\pi f_c t)] \qquad ......\ .\ 2.7$$

$$\Phi_2(t) = \sqrt{\frac{2}{T_s}}\,[\,\sin(2\pi f_c t)] \qquad ......\ .\ 2.8$$

The first basis function is used as the in-phase component of the signal and the second as the quadrature component of the signal. Hence, the signal constellation consists of the signal-space 4 points

$$\pm\left(\sqrt{\frac{E_s}{2}}\right), \pm\left(\sqrt{\frac{E_s}{2}}\right)$$

The factors of 1/2 indicate that the total power is split equally between the two carriers. Comparing these basis functions with that for BPSK show clearly how QPSK can be viewed as two independent BPSK signals. Note that the signal-space points for BPSK do not need to split the symbol (bit) energy over the two carriers in the scheme shown in the BPSK constellation diagram.

Bit error rate

Although QPSK can be viewed as a quaternary modulation, it is easier to see it as two independently modulated quadrature carriers. With this interpretation, the even (or odd) bits are used to modulate the in-phase component of the carrier, while the odd (or even) bits are used to modulate the quadrature-phase component of the carrier. BPSK is used on both carriers and they can be independently demodulated. As a result, the probability of bit-error for QPSK is the same as for BPSK

$$P_b = 2Q(\sqrt{\frac{2E_b}{N_o}}) \qquad \text{.....2.9}$$

However, in order to achieve the same bit-error probability as BPSK, QPSK uses twice the power (since two bits are transmitted simultaneously).

The symbol error rate is given by

$$P_s = 1 - (1 - P_b)^2 = 2Q(\sqrt{\frac{E_s}{N_o}}) - Q^2(\sqrt{\frac{E_s}{N_o}}) \qquad \text{...2.10}$$

If the signal-to-noise ratio is high (as is necessary for practical QPSK systems) the probability of symbol error may be approximated

$$P_s = 2Q(\sqrt{\frac{E_s}{N_o}}) \qquad \text{.....2.11}$$

QPSK signal in the time domain

The modulated signal is shown below for a short segment of a random binary data-stream. The two carrier waves are a cosine wave and a sine wave, as indicated by the signal-space analysis above. Here, the odd-numbered bits have been assigned to the in-phase component and the even-numbered bits to the quadrature component (taking the first bit as number 1).

The total signal — the sum of the two components — is shown at the bottom. Jumps in phase can be seen as the PSK changes the phase on each component at the start of each bit-period. The topmost waveform alone matches the description given for BPSK above.

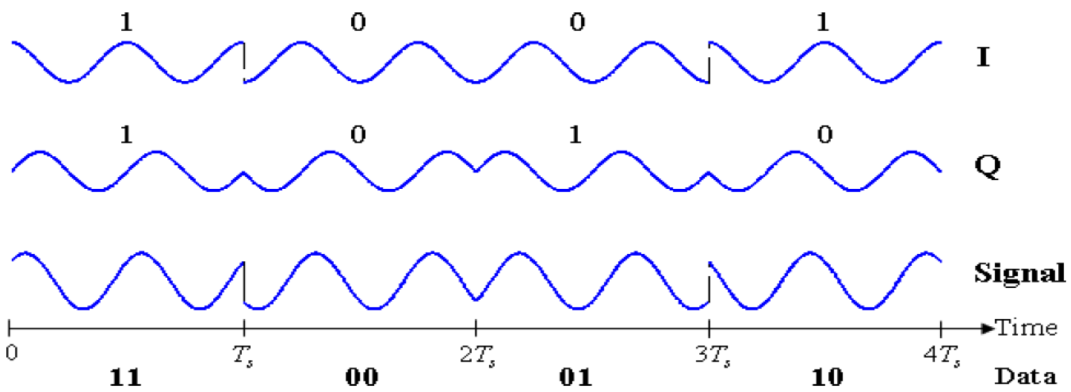


Fig 2.4 QPSK signal in the time domain

The binary data stream is shown beneath the time axis. The two signal components with their bit assignments are shown the top and the total, combined signal at the bottom. Note the abrupt changes in phase at some of the bit-period boundaries.

The binary data that is conveyed by this waveform is 1 1 0 0 0 1 1 0.

- The odd bits, highlighted here, contribute to the in-phase component **11000 110**
- The even bits, highlighted here, contribute to the quadrature-phase component **11000110.**

2.4 RADIO PROPAGATION

In an ideal radio channel, the received signal would consist of only a single direct path signal, which would be a perfect reconstruction of the transmitted signal. However in a real channel, the signal is modified during transmission in the channel. The received signal consists of a combination of attenuated, reflected, refracted, and diffracted replicas of the transmitted signal. On top of all this, the channel adds noise to the signal and can cause a shift in the carrier frequency if the transmitter or receiver is moving (Doppler Effect). Understanding of these effects on the signal is important because the performance of a radio system is dependent on the radio channel characteristics.

2.4.1 Attenuation

Attenuation is the drop in the signal power when transmitting from one point to another. It can be caused by the transmission path length, obstructions in the signal path, and multipath effects. Any objects that obstruct the line of sight signal from the transmitter to the receiver can cause attenuation. Shadowing of the signal can occur whenever there is an obstruction between the transmitter and receiver. It is generally caused by buildings and hills, and is the most important environmental attenuation factor. Shadowing is most severe in heavily built up areas, due to the shadowing from buildings. However, hills can cause a large problem due to the large shadow they produce. Radio signals diffract off the boundaries of obstructions, thus preventing total shadowing of the signals behind hills and buildings. However, the amount of diffraction is dependent on the radio frequency used, with low frequencies diffracting more than high frequency signals. Thus high frequency signals, especially, Ultra High Frequencies (UHF), and microwave signals require line of sight for adequate signal strength. To overcome the problem of shadowing, transmitters are usually elevated as high as possible to minimize the number of obstructions.

2.4.2 Fading Effects

Fading is about the phenomenon of loss of signal in telecommunications. Fading or fading channels refers to mathematical models for the distortion that a carrier modulated telecommunication signal experiences over certain propagation media. Small scale fading also known as multipath induced fading is due to multipath propagation. Fading results from the superposition of transmitted signals that have experienced differences in attenuation, delay and phase shift while traveling from the source to the receiver.

2.4.3 Multipath Fading

In wireless communications, multipath is the propagation phenomenon that results on radio signals reaching the receiving antenna by two or more paths. Causes of multipath include atmospheric ducting, ionosphere reflection and refraction and reflection from terrestrial obstacles such as mountains. buildings or vehicles. Below Figure show some of the possible way, to which multipath signals can occur.

The effects of multipath include constructive and destructive interference and phase shining of the signal which results in Fading of the signal called multipath fading. Small scale fading is usually divided into fading based on multipath time delay spread and that based on Doppler spread.

2.4.3.1 Delay Spread

The received radio signal from a transmitter consists of typically a direct signal, plus reflections off objects such as buildings, mountains, and other structures. The reflected signals arrive at a later time than the direct signal because of the extra path length, giving rise to a slightly different arrival times, spreading the received energy in time. Delay spread is the time spread between the arrival of the first and last significant multipath signal seen by the receiver. In a digital system, the delay spread can lead to inter-symbol interference. This is due to the delayed multipath signal overlapping with the following symbols. This can cause significant errors in high bit rate systems, especially when using time division multiplexing (TDMA). As the transmitted bit rate is increased the amount of inter-symbol interference also increases. The effect starts to become very significant when the delay spread is greater than —50% of the bit time. Inter-symbol interference can be minimized in several ways. One method is to reduce the symbol rate by reducing the data rate for each channel (i.e. split the bandwidth into more channels using frequency division multiplexing, or OFDM). Another is to use a coding scheme that is tolerant to inter-symbol interference such as CDMA. Adding guard interval is also a solution. There are two type of fading based on multipath time delay spread

Flat fading

The bandwidth of the signal is less than the coherence bandwidth of the channel or the delay spread is less than the symbol period.

Frequency selective fading

The bandwidth of the signal is greater than the coherence bandwidth of the channel or the delay spread is greater than the symbol period. Frequency selective fading occurs at selected frequencies and those frequencies are determined by the environment. But at these frequencies the signal is almost entirely wiped out. This is because of the destructive interference of the multipath signals.

2.4.3.2 Doppler Shift

When a wave source and a receiver are moving relative to one another the frequency of the received signal will not be the same as the source. When they are moving toward each other the frequency of the received signal is higher than the source, and when they are moving away each other the frequency decreases. This is called the Doppler Effect. An example of this is the change of pitch in a car's horn as it approaches then passes by. This effect becomes important when developing mobile radio systems. The amount the frequency changes due to the Doppler Effect depends on the relative motion between the source and receiver and on the speed of propagation of the wave. The Doppler shift in frequency can be written as

$$\Delta f = \pm(fv/c)\cos\theta \qquad \dots 2.12$$

Where Δf is the change in frequency of the source seen at the receiver, f is the frequency of the source, v is the speed difference between the source and receiver, c is the speed of light and θ is the angle between the source and receiver.

However, Doppler shift can cause significant problems if the transmission technique is sensitive to carrier frequency offsets (for example OFDM) or the relative speed is very high as is the case for low earth orbiting satellites.

There are two types of fading based on Doppler spread

Fast fading

Fast fading has a high Doppler spread. The coherence time is less than the symbol time and the channel variations are faster than baseband signal variation.

Slow fading

Slow fading has a low Doppler spread. The coherence time is greater than the symbol period and the channel variations are slower than the baseband signal variation.

2.4.33 Rayleigh fading

Rayleigh fading is the special case of multipath fading where there is no direct line of sight path available from transmitter to receiver end. That is all the signals received at the receiver are multipath reflected components. The relative phase of multiple reflected signals can cause constructive or destructive interference at the receiver. This is experienced over very short distances (typically at half wavelength distances).

CHAPTER 3

ORTHOGONAL FREQUENCY-DIVISION MULTIPLEXING (OFDM)

3.1 Introduction

Orthogonal frequency-division multiplexing (OFDM), essentially identical to coded OFDM (COFDM) and discrete multi-tone modulation (DMT), is a frequency-division multiplexing (FDM) scheme utilized as a digital multi-carrier modulation method. A large number of closely-spaced orthogonal sub-carriers are used to carry data. The data is divided into several parallel data streams or channels, one for each sub-carrier. Each sub-carrier is modulated with a conventional modulation scheme (such as quadrature amplitude modulation or phase-shift keying) at a low symbol rate, maintaining total data rates similar to conventional single-carrier modulation schemes in the same bandwidth

OFDM has developed into a popular scheme for wideband digital communication, whether wireless or over copper wires, used in applications such as digital television and audio broadcasting, wireless networking and broadband internet access. The primary advantage of OFDM over single-carrier schemes is its ability to cope with severe channel conditions (for example, attenuation of high frequencies in a long copper wire, narrowband interference and frequency-selective fading due to multipath) without complex equalization filters. Channel equalization is simplified because OFDM may be viewed as using many slowly-modulated narrowband signals rather than one rapidly-modulated wideband signal. The low symbol rate makes the use of a guard interval between symbols affordable, making it possible to handle time-spreading and eliminate inter symbol interference (ISI). This mechanism also facilitates the design of single frequency networks (SFNs), where several adjacent transmitters send the same signal simultaneously at the same frequency, as the signals from multiple distant transmitters may be combined constructively, rather than interfering as would typically occur in a traditional single-carrier system.

3.2 Time Division Multiple Access

Time division multiple access (TDMA) is a channel access method for shared medium networks, It allows several users to share the same frequency channel by dividing the signal into different time slots. The users transmit in rapid succession, one after the other, each using his own time slot. This allows multiple stations to share the same transmission medium (e.g. radio frequency channel) while using only a part of its channel capacity. TDMA is used in the digital 2G cellular systems such as Global System for Mobile Communications (GSM), 1S-136, Personal Digital Cellular (PDC) and in the Digital Enhanced Cordless Telecommunications (DECT) standard for portable phones. It is also used extensively in satellite systems, and combat-net radio systems. For usage of Dynamic TDMA packet mode communication, see below.

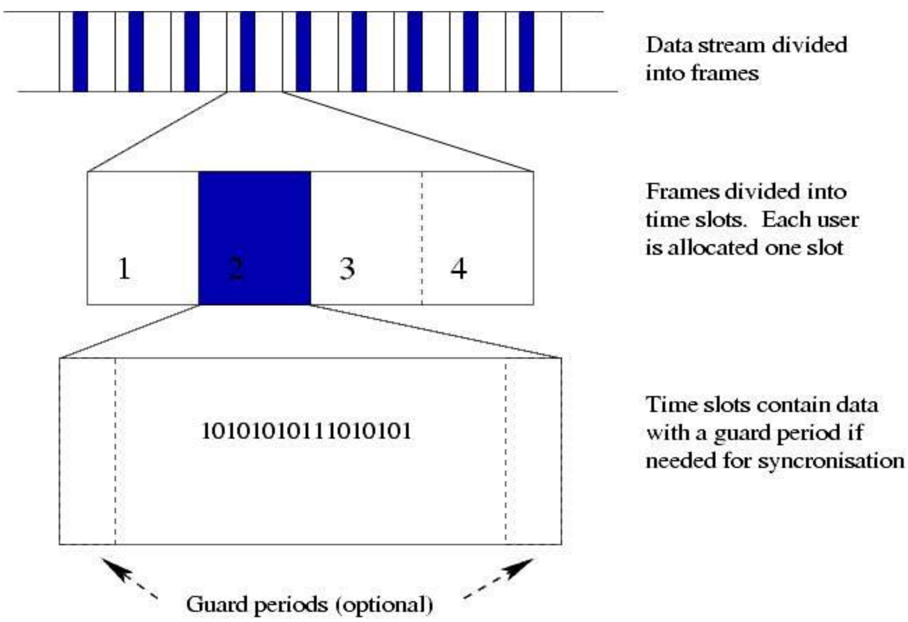


Fig 3.1 Dynamic TDMA pocket mode communication

TDMA frame structure showing a data stream divided into frames and those frames divided into time slots. TDMA is a type of Time-division multiplexing, with the special point that instead of having one transmitter connected to one receiver, there are multiple transmitters. In the case of the uplink from a mobile phone to a base station this becomes particularly difficult because the mobile phone can move around and vary the timing advance required to make its transmission match the gap in transmission from its peers.

Frequency Division Multiple Access or FDMA is a channel access method used in multiple-access protocols as a channelization protocol. FDMA gives users an individual allocation of one or several frequency hands, or channels, Multiple Access systems coordinate access between multiple users. The users may also share access via different methods such as TDM A, ('DMA, or SDMA. These protocols are utilized differently, at different levels or the theoretical OSI model. Disadvantage Crosstalk which causes interference on the other frequency and may disrupt the transmission.

Space-Division Multiple Access (SDMA) is a channel access method based on creating parallel spatial pipes next to higher capacity pipes through spatial multiplexing and/or diversity, by which it is able to offer superior performance in radio multiple access communication systems. In traditional mobile cellular network systems, the base station has no information on the position of the mobile units within the cell and radiates the signal in all directions within the cell in order to provide radio coverage. This results in wasting power on transmissions when there are no mobile units to reach, in addition to causing interference for adjacent cells using the same frequency, so called co-channel cells. Likewise, in reception, the antenna receives signals coming from all directions including noise and interference signals. By using smart antenna technology and by leveraging the spatial location of mobile units within the cell, space-division multiple access techniques offer attractive performance enhancements. The radiation pattern of the base station, both in transmission and reception is adapted to each user to obtain highest gain in the direction of that user. This is often done using phased array techniques.

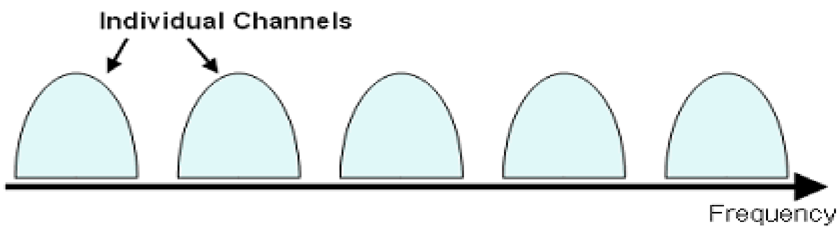
In GSM cellular networks, the base station is aware of the mobile phone's position by use of a technique called Timing Advance (TA). The Base Transceiver Station (BTS) can determine how distant the Mobile Station (MS) is by interpreting the reported TA. This information, along with other parameters, can then be used to power down the BTS or MS, if a power control feature is implemented in the network. The power control in either BTS or MS is implemented in most modern networks, especially on the MS, as this ensures a better battery life for the MS and thus a better user experience (in that the need to charge the battery becomes less frequent). This is why it may actually be safer to have a BTS close to you as your MS will be powered down as much as possible. For example, there is more power being transmitted from the MS than what you would receive from the BTS even if you are 6 m away from a mast. However, this estimation might not consider all the MS's that a particular BTS is supporting with EM radiation at any given time.

Code division multiple access (CDMA) is a channel access method utilized by various radio communication technologies. It should not be confused with the mobile phone standards called CDMA-One and CDMA2000 (which are often referred to as simply "CDMA"), which use CDMA as an underlying channel access method. One of the basic concepts in data communication is the idea of allowing several transmitters to send information simultaneously over a single communication channel. This allows several users to share a bandwidth of different frequencies. This concept is called multiplexing. CDMA employs spread-spectrum technology and a special coding scheme (where each transmitter is assigned a code) to allow multiple users to be multiplexed over the same physical channel. By contrast, time division multiple access (TDMA) divides access by time, while frequency-division multiple access (FDMA) divides it by frequency. CDMA is a form of "spread-spectrum" signaling, since the modulated coded signal has a much higher data bandwidth than the data being communicated.

An analogy to the problem of multiple access is a room (channel) in which people wish to communicate with each other. To avoid confusion, people could take turns speaking (time division), speak at different pitches (frequency division), or speak in different languages (code division). CDMA is analogous to the last example where people speaking the same language can understand each other, but not other people. Similarly, in radio CDMA, each group of users is given a shared code. Many codes occupy the same channel, but only users associated with a particular code can understand each other.

Frequency division multiplexing (FDM) involves the allocation of each channel to a unique frequency range. This frequency range prescribes both the center frequency and channel width (bandwidth). Because these channels are non-overlapping, multiple users can operate concurrently simply by using different channels of the frequency domain. Below, we illustrate the frequency domain of an FDM system.

Note from the diagram that each channel operates a different carrier frequency and that these channels are band limited to operate within a defined bandwidth.



Individual channel

Fig 3.2 FDM system

FDM is commonly used in a variety of communications protocols including Bluetooth and cellular protocols such as GSM, TDMA, and CDMA. Bluetooth, a digital communications protocol that is utilized by cell phones, laptops, and PDA's, is one example. It operates in the 2.4GHz unlicensed band and implements FDM by defining 79 channels from 2.402 GHz to 2.480 GHz which are spaced at 1 MHz apart. Each channel is band limited through the implementation of Gaussian filter. As second common implementation of FDM is in the Global System for Mobile Communications protocol (GSM) which is a 3G cellular communication standard. With GSM, the frequency range is divided into downlink channels from 890 - 915 MHz and the uplink channels at 935 - 960 MHz. Moreover, these frequency bands are further divided so that there are 124 channels which are spaced at 200 kHz intervals. Again, the bandwidth of each channel can be limited through the implantation of a root raised cosine filter.

3.3 OFDM DEFINITION

Modulation

A mapping of the information on changes in the carrier phase, frequency or amplitude or combination.

Multiplexing

Method of sharing a bandwidth with other independent data channels.

OFDM is a combination of modulation and multiplexing. Multiplexing generally refers to independent signals, those produced by different sources. So it is a question of how to share the spectrum with these users. In OFDM the question of multiplexing is applied to independent signals but these independent signals are a sub-set of the one main signal. In OFDM the signal itself is first split into independent channels, modulated by data and then re-multiplexed to create the OFDM carrier. OFDM is a special case of Frequency Division Multiplex (FDM). As an analogy, a FDM channel is like water flow out of a faucet, in contrast the OFDM signal is like a shower. In a faucet all water comes in one big stream and cannot be sub-divided.

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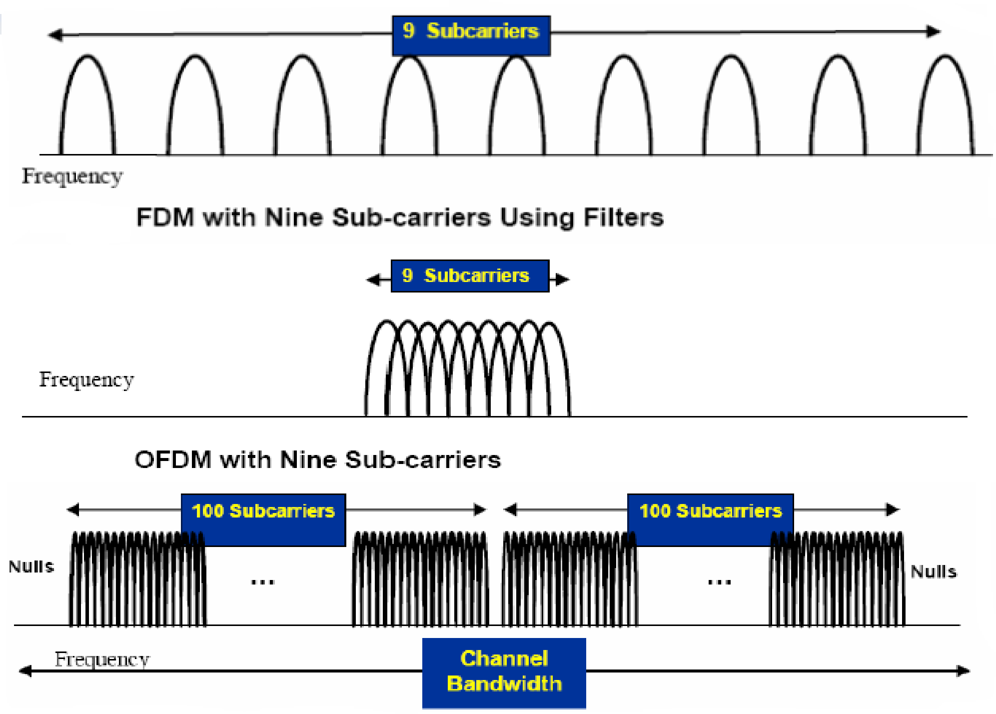

Fig 3.3 FDM system vs OFDM system

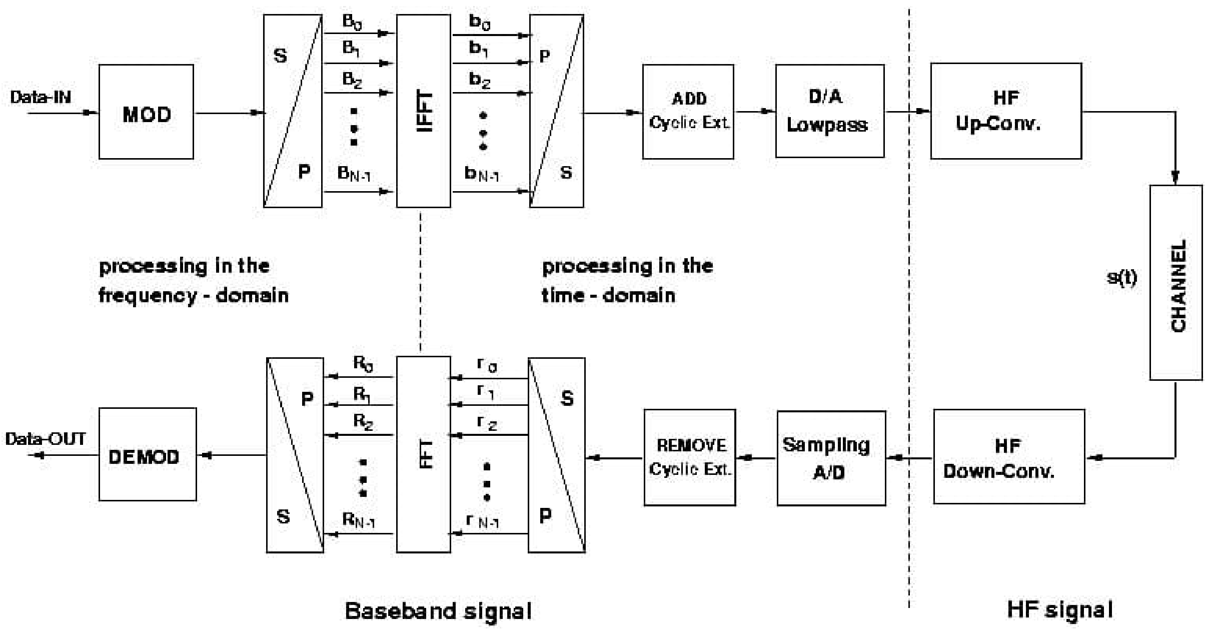

Fig 3.4 Basic OFDM system

3.4 Orthogonality Principle

In OFDM, the sub-carrier frequencies are chosen so that the sub-carriers are orthogonal to each other, meaning that cross-talk between the sub-channels is eliminated and inter-carrier guard bands are not required. This greatly simplifies the design of both the transmitter and the receiver; unlike conventional FDM, a separate filter for each sub-channel is not required.

The orthogonality requires that the sub-carrier spacing is $\Delta f = k/T_u$ Hertz, where Tu seconds is the useful symbol duration (the receiver side window size), and k is a positive integer, typically equal to 1. Therefore, with N sub-carriers, the total passband bandwidth will be B=N. Δf (Hz).

The orthogonality also allows high spectral efficiency, with a total symbol rate near the Nyquist rate for the equivalent baseband signal (i.e. near half the Nyquist rate for the double-side band physical passband signal). Almost the whole available frequency band can be utilized. OFDM generally has a nearly 'white' spectrum, giving it benign electromagnetic interference properties with respect to other co-channel users.

A simple example A useful symbol duration Tu = 1 ms would require a sub-carrier spacing of Δf = 1/1 ms =1kHz (or an integer multiple of that) for orthogonality. N = 1,000 sub-carriers would result in a total passband bandwidth of $N\Delta f$ = 1 MHz. For this symbol time, the required bandwidth in theory according to Nyquist is NTu= 0.5 MHz (i.e., half of the achieved bandwidth required by our scheme). If a guard interval is applied (see below), Nyquist bandwidth requirement would be even lower. The FFT would result in N = 1,000 samples per symbol. If no guard interval was applied, this would result in a base band complex valued signal with a sample rate of 1 MHz, which would require a baseband bandwidth of 0.5 MHz according to Nyquist. However, the passband RF signal is produced by multiplying the baseband signal with a carrier waveform (i.e., double-sideband quadrature amplitude-modulation) resulting in a passband bandwidth of 1 MHz. A single-side band (SSB) or vestigial sideband

(VSB) modulation scheme would achieve almost half that bandwidth for the same symbol rate (i.e., twice as high spectral efficiency for the same symbol alphabet length). It is however more sensitive to multipath interference.

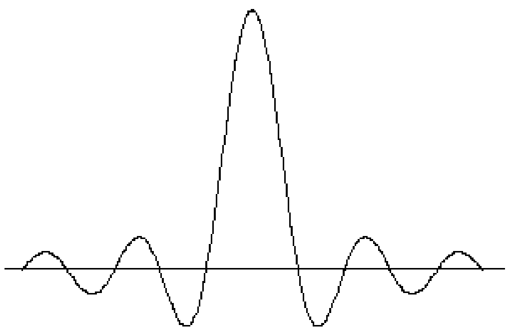

Fig 3.5 Single carrier of OFDM Signal

OFDM requires very accurate frequency synchronization between the receiver and the transmitter; with frequency deviation the sub-carriers will no longer be orthogonal, causing inter-carrier interference (ICI) (i.e., cross-talk between the sub-carriers). Frequency offsets are typically caused by mismatched transmitter and receiver oscillators, or by Doppler shift due to movement. While Doppler shift alone may be compensated for by the receiver, the situation is worsened when combined with multipath, as reflections will appear at various frequency offsets, which is much harder to correct. This effect typically worsens as speed increases, and is an important factor limiting the use of OFDM in high-speed vehicles. Several techniques for ICI suppression are suggested, but they may increase the receiver complexity.

3.5 Implementation Using The FFT Algorithm

The orthogonality allows for efficient modulator and demodulator implementation using the FFT algorithm on the receiver side, and inverse FFT on the sender side. Although the principles and some of the benefits have been known since the 1960s, OFDM is popular for wideband communications today by way of low-cost digital signal processing components that can efficiently calculate the FFT.

One key principle of OFDM is that since low symbol rate modulation schemes (i.e., where the symbols are relatively long compared to the channel time characteristics) suffer less from inter-symbol interference caused by multipath propagation, it is advantageous to transmit a number of low-rate streams in parallel instead of a single high-rate stream. Since the duration of each symbol is long, it is feasible to insert a guard interval between the OFDM symbols, thus eliminating the inter-symbol interference. The guard interval also eliminates the need for a pulse-shaping filter, and it reduces the sensitivity to time synchronization problems.

A simple example If one sends a million symbols per second using conventional single-carrier modulation over a wireless channel, then the duration of each symbol would be one microsecond or less. This imposes severe constraints on synchronization and necessitates the removal of multipath interference. If the same million symbols per second are spread among one thousand sub-channels, the duration of each symbol can be longer by a factor of a thousand (i.e., one millisecond) for orthogonality with approximately the same bandwidth. Assume that a guard interval of 1/8 of the symbol length is inserted between each symbol. Inter-symbol interference can be avoided if the multipath time-spreading (the time between the reception of the first and the last echo) is shorter than the guard interval (i.e., 125 microseconds). This corresponds to a maximum difference of 37.5 kilometers between the lengths of the paths.

The cyclic prefix, which is transmitted during the guard interval, consists of the end of the OFDM symbol copied into the guard interval, and the guard interval is transmitted followed by the OFDM symbol. The reason that the guard interval consists of a copy of the end of the OFDM symbol is so that the receiver will integrate over an integer number of sinusoid cycles for each of the multipath when it performs OFDM demodulation with the FFT.

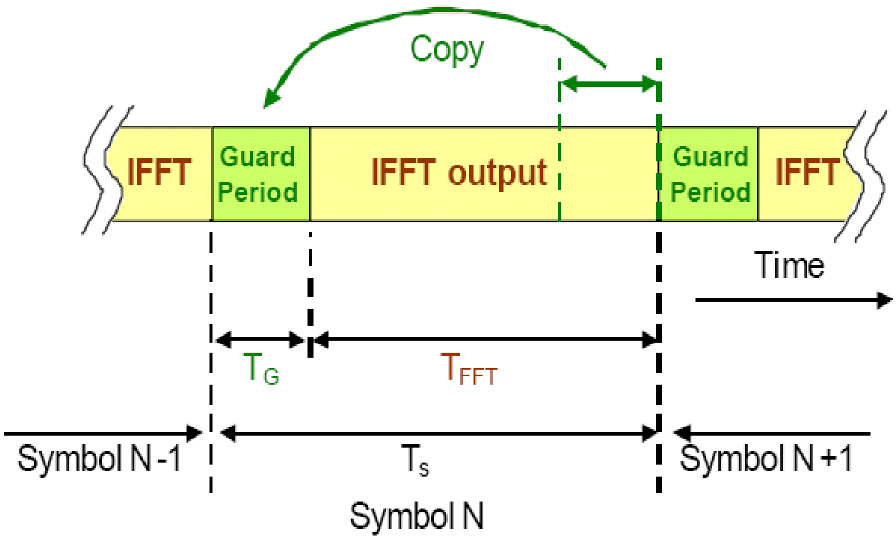

Fig 3.6 Guard period insertion in OFDM

3.6 Simplified Equalization

The effects of frequency-selective channel conditions, for example fading caused by multipath propagation, can be considered as constant (flat) over an OFDM sub-channel if the sub-channel is sufficiently narrow-banded (i.e., if the number of sub-channels is sufficiently large). This makes equalization far simpler at the receiver in OFDM in comparison to conventional single-carrier modulation. The equalizer only has to multiply each detected sub-carrier (each Fourier coefficient) by a constant complex number, or a rarely changed value.

Our example The OFDM equalization in the above numerical example would require one complex valued multiplication per subcarrier and symbol (i.e., N = complex multiplications per OFDM symbol; i.e., one million multiplications per second, at the receiver). The FFT algorithm requires N log2 N =10,000complex-valued multiplications per OFDM symbol (i.e., 10 million multiplications per second), at both the receiver and transmitter side. This should be compared with the corresponding one million symbols/second single-carrier modulation case mentioned in the example, where the equalization of 125 microseconds time-spreading using a FIR filter would require, in a naive implementation, 125 multiplications per symbol (i.e., 125 million multiplications per second). FM- techniques can be used to reduce the number of multiplications for an FIR equalizer to a number comparable with OFDM, at the cost of delay between reception and decoding which also becomes comparable with OFDM.

In a sense, improvements in FIR equalization using FFTs or partial FFTs leads mathematically closer to OFDM, but the OFDM technique is easier to understand and implement, and the sub-channels can be independently adapting in other ways than varying equalization coefficients, such as switching between different QAM constellation patterns and error-correction schemes to match individual sub-channel noise and interference characteristics. Some of the sub-carriers m some of the OFDM symbols may carry pilot signals for measurement of the channel conditions (i.e., the equalizer gain and phase shift for each sub-carrier). Pilot signals and training symbols (Preamble (communication)) may also be used for time synchronization (to avoid intersymbol interference, ISO, and frequency synchronization (to avoid inter-carrier interference, ICI, caused by Doppler shift).

If differential modulation such as DPSK or DQPSK is applied to each sub-carrier, equalization can be completely omitted, since these non-coherent schemes are insensitive to slowly changing amplitude and phase distortion. OFDM was initially used for wire, and stationary wireless communications. However with increasing number of applications operating in highly mobile environment, the possibility of using OFDM for such purpose is also investigated. Over the last decade, several research have been done on how to equalize OFDM transmission over doubly selective channels

3.7 Channel Coding And Interleaving

OFDM is invariably used in conjunction with channel coding (forward error correction), and almost always uses frequency and/or time interleaving. Frequency (subcarrier) interleaving increases resistance to frequency-selective channel conditions such as fading. For example, when a part of the channel bandwidth fades, frequency interleaving ensures that the bit errors that would result from those subcarriers in the faded part of the bandwidth are spread out in the bit-stream rather than being concentrated. Similarly, time interleaving ensures that bits that are originally close together in the bit-stream are transmitted far apart in time, thus mitigating against severe fading as would happen when travelling at high speed.

However, time interleaving is of little benefit in slowly fading channels, such as for stationary reception, and frequency interleaving offers little to no benefit for narrowband channels that suffer from flat-fading (where the whole channel bandwidth fades at the same time). The reason why interleaving is used on OFDM is to attempt to spread the errors out in the bit-stream that is presented to the error correction decoder, because when such decoders are presented with a high concentration of errors the decoder is unable to correct all the bit errors, and a burst of uncorrected errors occurs. A similar design of audio data encoding makes compact disc (CD) playback robust.

A classical type of error correction coding used with OFDM-based systems is convolutional coding, often concatenated with Reed-Solomon coding. Usually, additional interleaving (on top of the time and frequency interleaving mentioned above) in between the two layers of coding is implemented. The choice for Reed-Solomon coding as the outer error correction code is based on the observation that the Viterbi decoder used for inner convolutional decoding produces short errors bursts when there is a high concentration of errors, and Reed-Solomon codes are inherently well-suited to correcting bursts of errors.

Newer systems, however, usually now adopt near-optimal types of error correction codes that use the turbo decoding principle, where the decoder iterates towards the desired solution. Examples of such error correction coding types include turbo codes and LDPC codes, which perform close to the Shannon limit for the Additive White Gaussian Noise (AWGN) channel. Some systems that have implemented these codes have concatenated them with either Reed-Solomon (for example on the MediaFLO system) or BCH codes (on the DVB-S2 system) to improve upon an error floor inherent to these codes at high signal-to-noise ratios.

The resilience to severe channel conditions can be further enhanced if information about the channel is sent over a return-channel. Based on this feedback information, adaptive modulation, channel coding and power allocation may be applied across all sub-carriers, or individually to each sub-carrier. In the latter case, if a particular range of frequencies suffers from interference or attenuation, the carriers within that range can be disabled or made to run slower by applying more robust modulation or error coding to those sub-carriers.

The term discrete multitone modulation (DMT) denotes OFDM based communication systems that adapt the transmission to the channel conditions individually for each sub-carrier, by means of so called bit-loading. Examples are ADSL and VDSL. The upstream and downstream speeds can be varied by allocating either more or fewer carriers for each purpose. Some forms of rate-adaptive DSL use this feature in real time, so that the bit rate is adapted to the co-channel interference and bandwidth is allocated to whichever subscriber needs it most.

3.8 Space Diversity

In OFDM based wide area broadcasting, receivers can benefit from receiving signals from several spatially-dispersed transmitters simultaneously, since transmitters will only destructively interfere with each other on a limited number of sub-carriers, whereas in general they will actually reinforce coverage over a wide area. This is very beneficial in many countries, as it permits the operation of national SFNs, where many transmitters send the same signal simultaneously over the same channel frequency. SFNs utilize the available spectrum more effectively than conventional multi-frequency broadcast networks (MFN), where program content is replicated on different carrier frequencies. SFNs also result in a diversity gain in receivers situated midway between the transmitters. The coverage area is increased and the outage probability decreased in comparison to an MFN, due to increased received signal strength averaged over all sub-carriers.

Although the guard interval only contains redundant data, which means that it reduces the capacity, some OFDM-based systems, such as some of the broadcasting systems, deliberately use a long guard interval in order to allow the transmitters to be spaced farther apart in an SFN, and longer guard intervals allow larger SFN cell-sizes.

A rule of thumb for the maximum distance between transmitters in an SFN is equal to the distance a signal travels during the guard interval — for instance, a guard interval of 200 microseconds would allow transmitters to be spaced 60 km apart.

An OFDM signal exhibits a high peak-to-average power ratio (PAPR) because the independent phases of the sub-carriers mean that they will often combine constructively. Handling this high PAPR requires

- A high-resolution digital-to-analogue converter (DAC) in the transmitter
- A high-resolution analogue-to-digital converter (ADC) in the receiver
- A linear signal chain.

Any non-linearity in the signal chain will cause inter modulation distortion that

- raises the noise floor
- may cause inter-carrier interference
- Generates out-of-band spurious radiation.

The linearity requirement is demanding, especially for transmitter RF output circuitry where amplifiers are often designed to be non-linear in order to minimize power consumption. In practical OFDM systems a small amount of peak clipping is allowed to limit the PAPR in a judicious trade-off against the above consequences. However, the transmitter output filter which is required to reduce out-of-band spurs to legal levels has the effect of restoring peak levels that were clipped, so clipping is not an effective way to reduce PAPR.

Although the spectral efficiency of OFDM is attractive for terrestrial and space communications, the high PAPR requirements have so far limited OFDM applications to terrestrial systems.

3.9 IDEALIZED SYSTEM MODEL

This section describes a simple idealized OFDM system model suitable for a time-invariant AWGN channel.

3.9.1 Transmitter

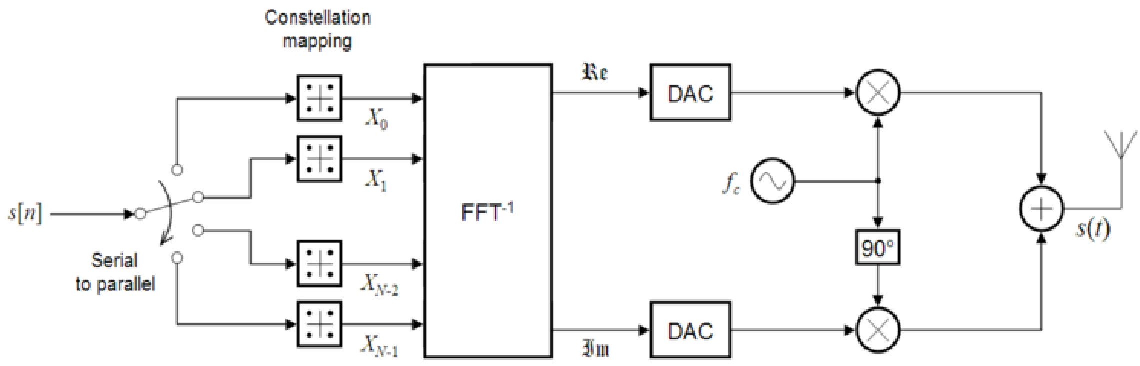

Fig 3.7 Idealized MIMO-OFDM Transmitter

An OFDM carrier signal is the sum of a number of orthogonal sub-carriers, with baseband data on each sub-carrier being independently modulated commonly using some type of quadrature amplitude modulation (QAM) or phase-shift keying (PSK). This composite baseband signal is typically used to modulate a main RF carrier.

S[n] is a serial stream of binary digits. By inverse multiplexing, these are first demultiplexed into Nparallel streams, and each one mapped to a (possibly complex) symbol stream using some modulation constellation (QAM, PSK,, etc.). Note that the constellations may be different, so some streams may carry a higher bit-rate than others.

An inverse FFT is computed on each set of symbols, giving a set of complex time-domain samples. These samples are then quadrature-mixed to passband in the standard way. The real and imaginary components are first converted to the analogue domain using digital-to-analogue converters (DACs); the analogue signals are then used to modulate cosine and sine waves at the carrier frequency, fe, respectively. These signals are then summed to give the transmission signal, s(t).

3.9.2 Receiver

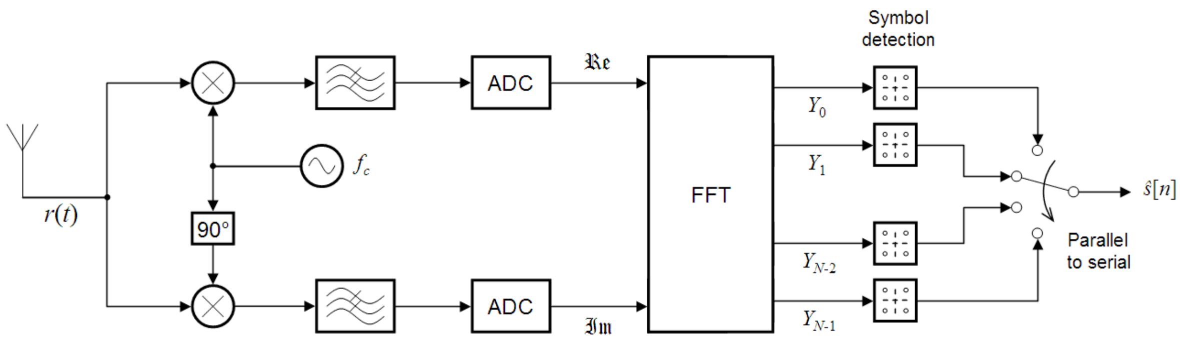

Fig 3.8 Idealized MIMO-OFDM Receiver

The receiver picks up the signal r(t), which is then quadrature-mixed down to baseband using cosine and sine waves at the carrier frequency. This also creates signals centered on $2f_c$, so low-pass filters are used to reject these. The baseband signals are then sampled and digitized using analogue-to-digital converters (ADCs), and a forward FFT is used to convert back to the frequency domain. This returns **N** parallel streams, each of which is converted to a binary stream using an appropriate symbol detector. These streams are then re-combined into a serial stream, s^[n], which is an estimate of the original binary stream at the transmitter.

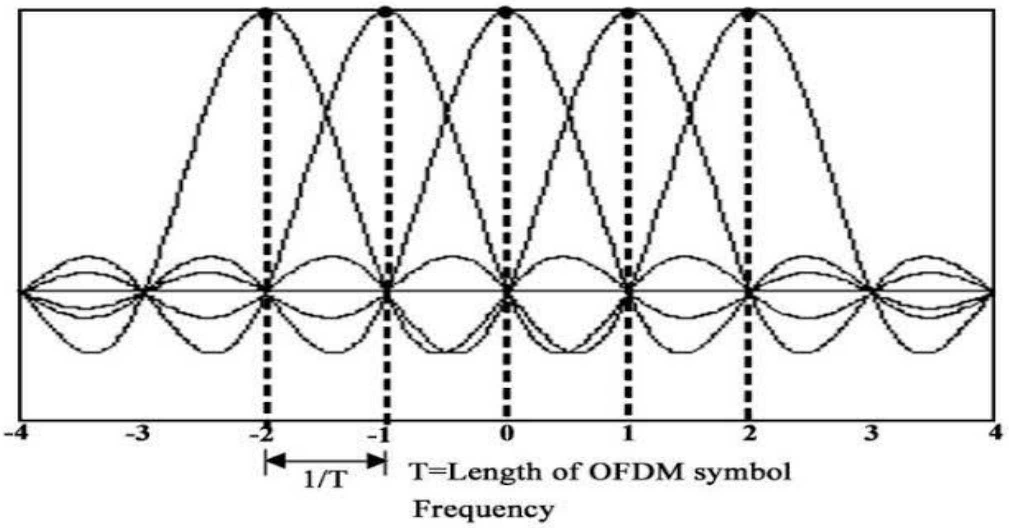

Fig 3.9 Frequency spectrum of OFDM transmission

OFDM transmits a large number of narrowband sub channels. The frequency range between carriers is carefully chosen in order to make them orthogonal each another. In fact, the carriers are separated by an interval of 1/T, where T represents the duration of an OFDM symbol, The frequency spectrum of an OFDM transmission is illustrated in above figure, The figure indicates the spectrum of carriers significantly over laps over the other carrier. This is contrary to the traditional FDM technique in which a guard band is provided between each carrier. Each sine of the frequency spectrum in the above Figure corresponds to a sinusoidal carrier modulated by a rectangular waveform representing the information symbol. One could easily notice that the frequency spectrum of one carrier exhibits zero-crossing at central frequencies corresponding to all other carriers. At these frequencies, the inter carrier interference is eliminated, although the individual spectra of subcarriers overlap. It is well known that orthogonal signals can be separated at the receiver by correlation techniques. The receiver acts as a bank of demodulators, translating each carrier down to baseband, the resulting signal then being integrated over a symbol period to recover the data. If the other carriers beat down to frequencies which, in the time domain means an integer number of cycles per symbol period (T), then the integration process results in a zero contribution from all these carriers. The waveforms of some of the carriers in an OFDM transmission are illustrated in below figure.

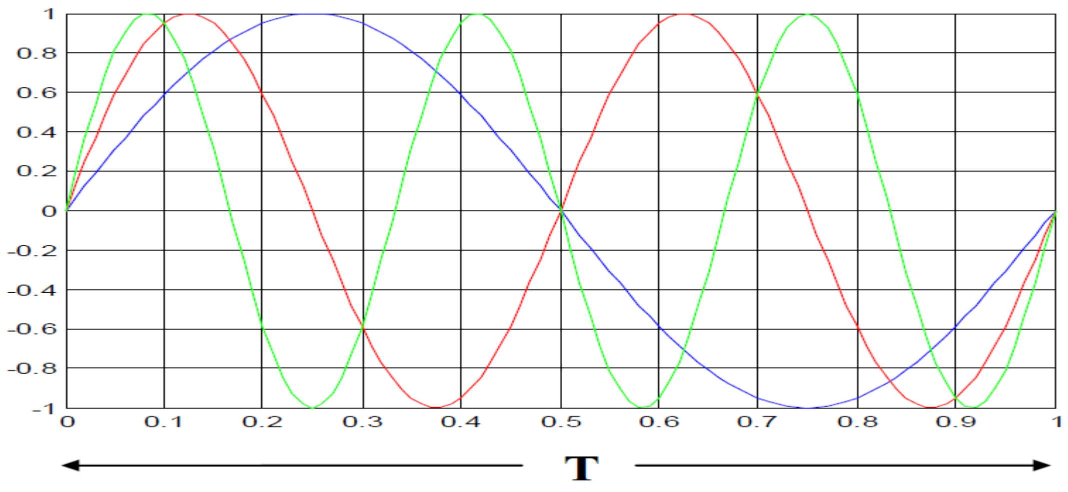

Fig 3.10 Carrier signals in an OFDM transmission

3.10 ADVANTAGES OF OFDM

OFDM (Orthogonal Frequency Division Multiplexing) Several parallel channels with low bitrates whose carriers are overlapping but orthogonal. This is an efficient way of having several sub channels in a fixed bandwidth. The subcarriers are not separated by a distance corresponding to their bandwidth but are overlapping. The spacing between the subcarriers are arranged such that they become orthogonal, hence the name OFDM. Apart from the bandwidth efficiency, you get a fast implementation with FFT, which is used as a digital modulator/demodulator of each sub channel.

QAM (Quadrature Amplitude Modulation) Amplitude modulation in which two signals modulate on in-phase (0 degrees) and quadrature (90 degrees) carriers, thus doubling the information carried in an amplitude-modulated (AM) system. QAM is used by the Cable TV industry instead of VSB for coax or HFC DTV. QPSK (Quadrature Phase Shift Keying) 4 phase digital modulation. Two data channels modulate the carrier. Transitions in the data cause the carrier to shift by either 90 or 180 degrees. This is the binary version of QAM. This allows customers to transmit two discrete data streams; identified as I channel (In phase) and the Q channel (Quadrature) data. Digital Dolby AC-3 has an encoded bit stream using (QPSK) modulation.

- Makes efficient use of the spectrum by allowing overlap.
- By dividing the channel into narrowband flat fading sub channels, OFDM is more resistant to frequency selective fading than single carrier systems are. i.e. robustness to frequency selective fading channels
- Eliminates ISI through use of a cyclic prefix.
- Using adequate channel coding and interleaving one can recover symbols lost due to the frequency selectivity of the channel.
- Channel equalization becomes simpler than by using adaptive equalization techniques with single carrier systems.
- It is possible to use maximum likelihood decoding with reasonable complexity.

- OFDM is computationally efficient by using FFT techniques to implement the modulation and demodulation functions.

- Is less sensitive to sample timing offsets than single carrier systems are.
- Provides good protection against co channel interference and impulsive parasitic noise

3.11 DISADVANTAGES OF OFDM

- It is more sensitive to ICI (inter carrier interference) which is due to frequency offset.
- Peak to average power ratio(PAPR) is high
- Bandwidth and power loss can be significant due to guard interval
- High power transmitter amplifiers need linearization
- Low noise receiver amplifiers need large dynamic range

CHAPTER 4

MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO)-ORTHOGONAL FREQUENCY-DIVISION MULTIPLEXING (OFDM)

4.1 Introduction

SIMO is an acronym for single-input and single-output system. In control engineering usually refers to a simple single variable control system with one input and one output. In radio it is the use of only one antenna both in the transmitter and receiver. SISO systems arc typically less complex than Multiple-Input Multiple-Output (MIMO) systems. Usually, it is also easier to make order of magnitude or trending predictions "on the fly" or "back of the envelope". MIMO systems have too many interactions for most of us to trace through them quickly, thoroughly, and effectively in our heads. Frequency domain techniques for analysis and controller design dominate SISO control system theory. Bode, Nyquist, Nichols, and root locus arc the usual tools for SISO system analysis. Controllers can be designed through the polynomial design, root locus design methods to name just 2 of the more popular. Often SISO controllers will be PI, MD, or Lead-Lag.

Multiple-Input and Multiple-Output, or MIMO (commonly pronounced my-moh or me-moh), is the use of multiple antennas at both the transmitter and receiver to improve communication performance. It is one of several forms of smart antenna technology. MIMO technology has attracted attention in wireless communications, because it offers significant increases in data throughput and link range without additional bandwidth or transmit power. It achieves this by higher spectral efficiency (more bits per second per hertz of bandwidth) and link reliability or diversity (reduced fading). Because of these properties, MIMO is a current theme of international wireless research. (Refer to Research trends in MIMO literature).

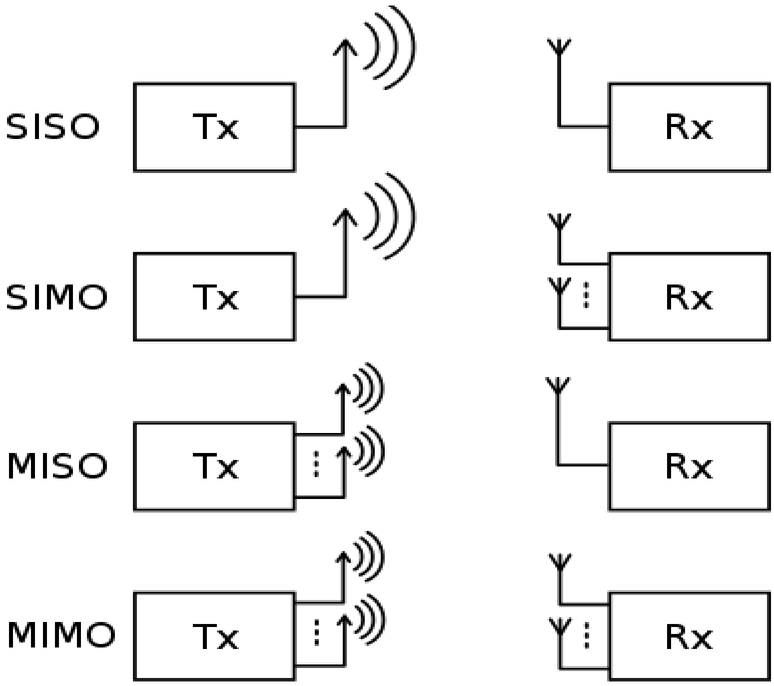

Figure 4.1 MIMO-OFDM System

Multiple-input multiple-output (MIMO) wireless technology uses multiple antennas at the transmitter and receiver to produce significant capacity gains over single-input single-output (SISO) systems using the same bandwidth and transmit power. It has been shown that the capacity of a MIMO system increases linearly with the number of antennas in the presence of a scattering-rich environment. This will ensure that the signals at the antennas in the array are sufficiently uncorrelated with each other. This is where antenna design comes in for MIMO systems.

The main of MIMO is to reduce correlation between received signals by exploiting various forms of diversity that arise due to the presence of multiple antennas, like

- Space diversity (spacing antennas far apart),
- Pattern diversity (using antennas with different or orthogonal radiation patterns),
- Polarization diversity (using antennas with different polarizations) etc

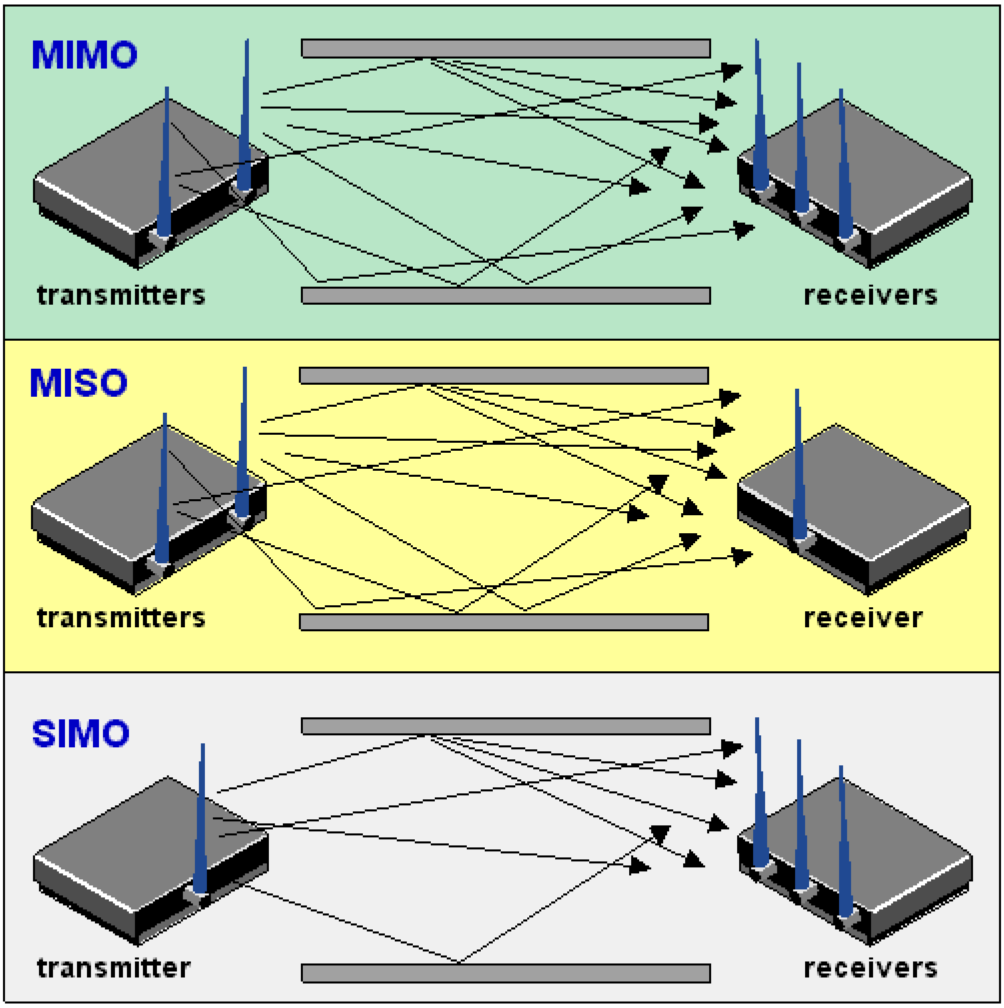

Fig 4.2 MIMO-OFDM System

In MIMO systems, a transmitter sends multiple streams by multiple transmit antennas. The transmit streams go through a matrix channel which consists of all N, N, paths between the N, transmit antennas at the transmitter and N, receive antennas at the receiver. Then, the receiver gets the received signal vectors by the multiple receive antennas and decodes the received signal vectors into the original information. A narrowband flat fading MIMO system is modelled as

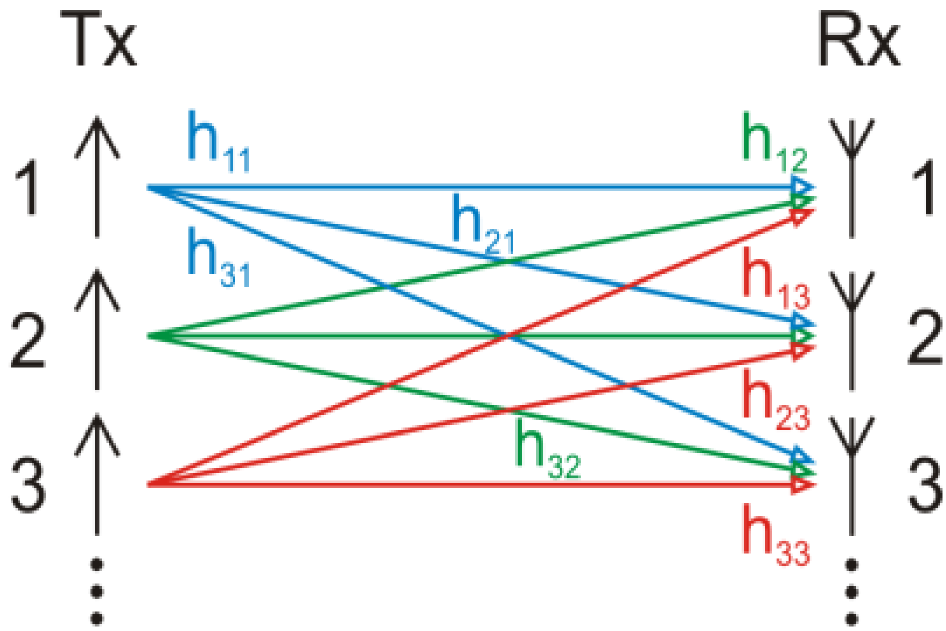

Fig 4.3 MIMO Transmitting and receiver Antenna System

Where Y and X are the receive and transmit vectors, respectively, and H and n are the channel matrix and the noise vector, respectively.

4.2 Concepts of MIMO System

MIMO can be sub-divided into three main categories Precoding, spatial multiplexing or SM, and diversity coding.

4.2.1 Precoding

The multi-layer beam forming in a narrow sense or all spatial processing at the transmitter in a wide-sense. In (single-layer) beam forming, the same signal is emitted from each of the transmit antennas with appropriate phase (and sometimes gain) weighting such that the signal power is maximized at the receiver input. The benefits of beam forming are to increase the signal gain from constructive combining and to reduce the multipath fading effect. In the absence of scattering, beamforming results in a well defined directional pattern, but in typical cellular conventional beams are not a good analogy. When the receiver has multiple antennas, the transmit beamforming cannot simultaneously maximize the signal level at all of the receive antennas, and precoding is used. Note that precoding requires knowledge of the channel state information (CSI) at the transmitter.

4.2.2 Spatial Multiplexing

Requires MIMO antenna configuration. In spatial multiplexing, a high rate signal is split into multiple lower rate streams and each stream is transmitted from a different transmit antenna in the same frequency channel. If these signals arrive at the receiver antenna array with sufficiently different spatial signatures, the receiver can separate these streams, creating parallel channels free. Spatial multiplexing is a very powerful technique for increasing channel capacity at higher Signal to Noise Ratio (SNR). The maximum number of spatial streams is limited by the lesser in the number of antennas at the transmitter or receiver. Spatial multiplexing can be used with or without transmit channel knowledge.

4.2.3 Diversity Coding

Techniques are used when there is no channel knowledge at the transmitter. In diversity methods a single stream (unlike multiple streams in spatial multiplexing) is transmitted, but the signal is coded using techniques called space-time coding. The signal is emitted from each of the transmit antennas using certain principles of full or near orthogonal coding. Diversity exploits the independent fading in the multiple antenna links to enhance signal diversity. Because there is no channel knowledge, there is no beamforming or array gain from diversity coding.

Spatial multiplexing can also be combined with precoding when the channel is known at the transmitter or combined with diversity coding when decoding reliability is in trade-off.

4.3 Communication With Mimo-Ofdm

High data-rate wireless access is demanded by many applications. Traditionally, more bandwidth is required for higher data-rate transmission. However, due to spectral limitations, it is often impractical or sometimes very expensive to increase bandwidth. In this case, using multiple transmit and receive antennas for spectrally efficient transmission is an alternative solution. Multiple transmit antennas can be used either to obtain transmit diversity, or to form multiple-input multiple-output (MIMO) channels. Many researchers have studied using multiple transmit antennas for diversity in wireless systems. Transmit diversity may be based on linear transforms or space—time coding. In particular, space—time coding is characterized by high code efficiency and good performance; hence, it is a promising technique to improve the efficiency and performance of orthogonal frequency division multiplexing (OFDM) systems. On the other hand, the system capacity can be significantly improved if multiple transmit and receive antennas are used to form MIMO channels. It is proven in that, compared with a single-input single-output (SISO) system with flat Rayleigh fading or narrowband channels, a MIMO system can improve the capacity by a factor of the minimum number of transmit and receive antennas. For wideband transmission, space—time processing must be used to mitigate intersymbol interference (ISI). However, the complexity of the space—time processing increases with the bandwidth, and the performance substantially degrades when estimated channel parameters are used.

In OFDM, the entire channel is divided into many narrow parallel sub channels, thereby increasing the symbol duration and reducing or eliminating the ISI caused by the multipath. Therefore, OFDM has been used in digital audio and video broadcasting in Europe, and is a promising choice for future high-data-rate wireless systems. Multiple transmit and receive antennas can be used with OFDM to further improve system performance. We have studied OFDM systems with adaptive antenna arrays for co-channel interference suppression and transmit diversity based on space— time coding, delayed transmission, and permutation. In particular, a channel parameter estimator for OFDM systems with multiple transmit antennas was proposed in and simplified in . Optimum training sequences for OFDM with multiple transmit antennas were also proposed in. Multiple transmit and receive antennas for OFDM to form MIMO channels (MIMO-OFDM). Our focus here is enhanced channel estimation and signal detection. The rest of this paper is organized as follows; we introduce MIMO-

OFDM systems based on space—time coding and briefly discuss wireless channel characteristics. Before introducing the signal detection and enhanced channel estimation technique, we briefly describe a MIMO-OFDM system and the statistics of mobile wireless channels.

4.4 Space -Time-Frequency Coding

Recent research results have shown that the adverse effects of the wireless propagation environment can be significantly reduced by employing multiple transmit and receive antennas, resulting in multiple input-multiple-output (MEMO) communication systems. Combining MIMO systems with OFDM modulation, MIMO-OFDM systems have been proposed, and two coding approaches have been suggested for such systems space-frequency (SF) coding, to exploit the spatial and frequency diversities, and space-time-frequency (STF) coding, to exploit the spatial, temporal, and frequency diversities available in frequency selective MIMO channels.

The first SF coding scheme was proposed, in which previously existing space-time (ST) codes were used by replacing the time domain with frequency domain (OFDM tones). Later works also described similar schemes, i.e. using ST codes directly as SF codes. The resulting SF codes achieved only the spatial diversity, and were not guaranteed to achieve the full spatial and frequency diversities. Later in systematic SF code design methods were proposed that can guarantee to achieve full diversity. The STF coding strategy, by coding across multiple OFDM blocks, was first proposed in for two transmit antennas and further developed and for multiple transmit antennas. Both assumed that the MIMO channel stays constant over multiple OFDM blocks. In an intuitive explanation on the equivalence between antennas and OFDM tones was presented from a capacity point of view. In the performance criteria for STF codes were derived, and an upper bound on the maximum achievable diversity order was established. However, there was no discussion in whether the upper bound can be achieved or not. A general framework for the performance analysis of MIMO-OFDM systems, taking into account coding over the spatial, temporal and frequency domains. Our model incorporates the ST and SF coding approaches as special cases. We also derive an alternative form of the performance criteria for STF-coded MIMO-OFDM systems, based on the results. We determine the maximum achievable diversity order for STF codes, and demonstrate that a simple repetition coding approach can be used to achieve it. Finally, we investigate the effect of temporal and frequency-domain correlation on the performance of MIMO-OFDM systems and show that if the channel changes independently from OFDM block to OFDM block, significant performance improvement can be achieved by STF coding compared to the SF coding approach.

4.5 System Model

We consider a STF-coded MIMO-OFDM system with M_t transmit antennas, M_r receive antennas and N sub-carriers. Suppose that the frequency selective fading channels between each pair of transmit and receive antennas have L independent delay paths and the same power delay profile. The MIMO channel is assumed to be constant over each OFDM block period, but it may vary from one OFDM block to another. At the k-th OFDM block, the channel impulse response from transmit antenna i to receive antenna j at time τ can be modelled as

$$H_{i,j}^{(K)}(f) = \left[\sum_{l=0}^{L-1} \alpha_{i,j}^{k}(l)\delta\left(\tau - \tau_l\right)\right] \qquad \text{.....4.1}$$

Where $\alpha_{i,j}^{k}(l)$ is the complex amplitude of the i-th path between transmit antenna I and receive antenna j at the k-th OFDM block.

The $\alpha_{i,j}^{k}(l)$'s are modelled as zero-mean, complex Gaussian random variables with variances $E|\alpha_{i,j}^{k}(l)|^2=\delta_l^2$ where E stands for the expectation. The powers of the L paths are normalized such that $\sum_{l=0}^{L-1}\delta_l^2=1$ From (1) the frequency response of the channel is given by

$$H_{i,j}^{(K)}(f) = \left[\sum_{l=0}^{L-1} \alpha_{i,j}^{k} e^{-j2\pi f_q T_l}\right] \qquad \text{......4.2}$$

Where j=$\sqrt{-1}$. We assume that the MIMO channel is spatially uncorrelated, i.e. the channel taps $\alpha_{i,j}^{k}(l)$ are independent for different indices (i, j). We consider STF coding across Mt transmit antennas, N OFDM sub-carriers and K OFDM blocks. Each STF codeword can be expressed as a KN *Mt matrix

$$c=[c_1^T\ c_2^T\ c_3^T\ c_4^T\ \ldots\ldots\ldots\ldots.c_k^T]^T \qquad \ldots\ldots 4.3$$

$$C_k = \begin{bmatrix} c_1^k(0) & c_2^k(0) & \cdots & . & c_{M_t}^k(0) \\ c_1^k(1) & c_2^k(1) & \cdots & . & c_{M_t}^k(1) \\ \cdot & \cdot & \cdots & \cdots & \vdots \\ . & . & \cdots & \cdots & \vdots \\ c_1^k(N-1) & c_2^k(N-1) & \cdots & \cdots & c_{M_t}^k(N-1) \end{bmatrix} \qquad \ldots . 4.4$$

is the channel symbol matrix transmitted in the k-th OFDM block and $c_i^k(n)$ is the channel symbol transmitted over the n-th sub-carrier by transmit antenna i at the k-th OFDM block. The STF code is assumed to satisfy the energy constraint $E\|C\|^2_F=KNM_t$, where $\|C\|_F$ is the Frobenius norm of C. At the k-th OFDM block, the OFDM transmitter applies an N-point IFFT to each column of the matrix Ck. After

appending a cyclic prefix, the OFDM symbol corresponding to the i-th (i=1;2…….Mt) column of Ck is transmitted by transmit antenna i.

At the receiver, after matched filtering, removing the cyclic prefix, and applying FFT, the received signal at the n-th sub-carrier at receive antenna j in the k-th OFDM block is given by

$$y_j^k(n) = \sqrt{\frac{\rho}{M_t}} \sum_{i=1}^{M_t} c_i^k(n) H_{i,j}^k(n) + z_j^k(n) \qquad \ldots\ldots 4.5$$

$$H_{i,j}^{(K)}(n) = \left[\sum_{l=0}^{L-1} \alpha_{i,j}^k(l) e^{-j2\pi n \Delta f T_l} \right] \qquad \ldots\ldots 4.6$$

is the channel frequency response at the n-th sub-carrier between transmit antenna i and receive antenna j, Δf =1/T is the sub-carrier separation in the frequency domain, and T is the OFDM symbol period. We assume that the channel state information $H_{i,j}^{K}(n)$ is known at the receiver, but not at the transmitter. In (5), $z_j^{K}(n)$ denotes the additive complex Gaussian noise with zero mean and unit variance at the n-th sub-carrier at receive antenna j in the k-th OFDM block. The factor $\sqrt{\left[\frac{\rho}{M_t}\right]}$ (5) ensures that ρ is the average signal to noise ratio (SNR) at each receive antenna, and it is independent of the number of transmit antennas.

4.6 Performance Criteria

In this section. we derive the performance criteria for STF-coded MIMO-OFDM systems according to previous proposed schemes

Using the notation

$$c_i\ ((k - 1)N+n) \cong c_i^{k}(n) \quad \text{…. 4.7}$$

$$H_{i,j}\big((k-1)N+n\big) \cong H_{i,j}^{k}(n) \quad \text{…. 4.8}$$

$$y_i\big((k-1)N+n\big) \cong y_j^{k}(n) \quad \text{….4.9}$$

$$z_j\big((k-1)N+n\big) \cong z_j^{k}(n) \quad \text{… 4.10}$$

For 1$\leq k \leq K, 0 \leq n \leq N-1, 1 \leq i \leq Mt\ and\ 1 \leq j \leq Mr$

The received signal in (5) can be expressed as

$$y_j(m) = \sqrt{\frac{\rho}{M_t}} \sum_{i=1}^{M_t} c_i(m) H_{i,j}(m) + z_j(m) \quad \text{…..4.11}$$

From m=0,1………KN-1.

We further rewrite the received signal in vector form as

$$Y = \sqrt{\frac{\rho}{M_t}} DH + Z \quad \text{……4.12}$$

Where D is a KNMr*KNMtMr matrix constructed from the STF codeword C in (3) as follows

$$D = I_{M_r} \otimes [D_1 D_2 \ldots . D_{M_t}]$$ ……4.13

Where $\otimes$ denotes the tensor product, I_{Mr} identity matrix of size Mr*Mr.

D=diag{Ci(0),Ci(1),……..,Ci(kN-1)} …4.14

For any i=1,2,…….Mt

Each D_i in the above equation is related to i-th coloumn of the STF codeword C.

The terms Y,H,Z in the received signal can be given as follows

The channel vector H of size KNMtMr*1 is formatted as

$$H = [H_1(0) \ldots \ldots . H_1(KN-1) H_2(0) \ldots \ldots . H_{M_r}(0) \quad \ldots H_{M_r}(KN-1)]^T$$ ..4.15

Where

$$H_{l,j} = [H_{i,j}(0) H_{i,j}(1) \ldots \ldots \ldots \ldots . H_{i,j}(KN-1)]^T$$

The received signal vector Y of size KNMr*1 is given by

$$Y = [y_1(0) \ldots \ldots . y_1(KN-1) y_2(0) \ldots \ldots . y_{M_r}(0) \quad \ldots y_{M_r}(KN-1)]^T$$ ..4.16

and the noise vector Z has the same form as Y, i.e.,

$$z = | \ z_t(0) \ldots \ldots . z_1(KN-1) z_2(0) \ldots \ldots 2_{M,}(0) \quad \ldots z_{M_1}(KN-1)]^\top$$ 4.17

Suppose that D and D-cap are two different matrices related to two different STF codewords C and C-cap respectively. Then the pairwise error probability between D and D-cap can be bounded as

$$P(D-\bar{D}) \leq \left(2^{r-1} r^2\right) \left(\Pi_{i=1}^{r} \gamma_i\right)^{-1} \left(\frac{\rho}{M_t}\right)^{-r}$$ …4.18

Where r is the rank of $(D-\bar{D}) R (D-\bar{D})^H$

Υ_1, Υ_2, Υ_3,……. Υ_r

Are the non-zero eigen values of $(D-\bar{D}) R (D-\bar{D})^H$

$R = E\left(HH^H\right)$ is the correlation matrix of H.

The superscript H stands for the complex conjugate and transpose of a matrix. Based on the upper bound on the pairwise error probability in the above given equation , two general STF code performance criteria can be proposed as follows.

Diversity (rank) criterion

The minimum rank of $(D - \bar{D})R(D - \bar{D})^H$ over all pairs of different codewords C and L should be as large as possible.

Product criterion

The mininuun value of the productrThi yi over all pairs of different codewords C and C should be maximized. ST coding, SF coding, and STF coding approaches for MIMO-OFDM systems performance is compared depending on different calculations. Typically, ST-coded OFDM has a simple implementation in providing a minimal decoding complexity, but it cannot achieve multipath diversity nor high rate. SF-coded OFDM, by mapping the symbols on other subchannels, can exploit the multipath diversity, but results in a reduction of the data rate. When combined with the signal space diversity technique via a constellation rotation , SF coding can achieve the maximum diversity and full rate over multipath fading channels, but the decoding complexity is increased and a joint ML decoding is needed. STF-coded OFDM tailored for block fading channels can achieve full diversity in space, time, and frequency and full rate. It can be seen that the two STF coding approaches achieve a larger diversity gain than the other coding approaches.Pre-DFT processing scheme for MIMO-OFDM systems with space-time-frequency coding is used to reduce the complexity present at receiver side in MIMO-OFDM systems.

CHAPTER 5

Pre-DFT Processing

5.1 Introduction

Orthogonal Frequency Division Multiplexing (OFDM) is very effective in mitigating adverse multipath effects of a broadband wireless channel. OFDM has been successfully used in Wireless Local Area Networks (WLANs), such as, European Hiper LAN/2 or Japanese MMAC standards as high-data rate physical layer transmission scheme for local area coverage. The WLAN standard specifies channel coding and frequency interleaving to exploit the frequency diversity of the wideband channel. Efficiency can only be achieved if the channel is sufficiently frequency-selective, corresponding to long channel delay spreads. In a flat fading situation (or in relatively lesser frequency-selective fading situation which we often encounter in indoor wireless scenario), all or most subcarriers are attenuated simultaneously leading to long error bursts. In this case, frequency interleaving does not provide enough diversity to significantly improve the decoding performance.

For OFDM system with multiple transmit antennas, the schemes mentioned earlier, explicitly or implicitly, assume that the channel state information (CSI) is known at the transmitter. In mobile communications, where the channel can vary rapidly, it is difficult to maintain the CSI at the transmitter up-to-date without substantial system overhead. Space-time-frequency codes were proposed for OFDM systems to fully take advantage of the frequency diversity and spatial diversity presented in frequency selective fading channels without the requirement of the availability of CSI at the transmitter. For such a system, traditional subcarrier based space processing induces considerable complexity due to the reasons mentioned before. In this paper, we propose to use pre-DFT processing to reduce the receiver complexity of MIMO-OFDM systems with space-time-frequency coding. In our proposed scheme, the received signals at the receiver are first weighted and then combined before the DFT processing. Owing to the pre-DFT processing, the number of DFT blocks required at the receiver can be reduced, and a high dimensional MIMO system can be shrunk into an equivalently low dimension one. Both enable effective complexity reduction.

5.2 weighting coefficients

One important issue in the proposed pre-I)FT processing scheme for MIMO-OFDM systems with space-time-trequency coding is the calculation of the weighting coefficients before the MT processing. In general, the weighting coefficients calculation are specific to the space-time-frequency coding scheme. In this project, we propose a universal weighting coefficients calculation algorithm that can be applied in most practical space-time-frequency codes such as those proposed in earlier. This makes the design of the pre-DFT processing scheme independent of the optimization of the space-time-frequency coding, which is desirable for multiplatform systems. In general, the weighting coefficients before the DFT processing can be calculated assuming that the CSIs are explicitly available. In this paper, we will show that the weighting coefficients can also be obtained using the signal space method without the explicit knowledge of the CSIs. This helps to reduce the complexity of channel estimation required by the space-time-frequency decoding since the number of equivalent channel branches required to be estimated in the proposed scheme can be reduced from the number of receive antennas to the number of DFT blocks.

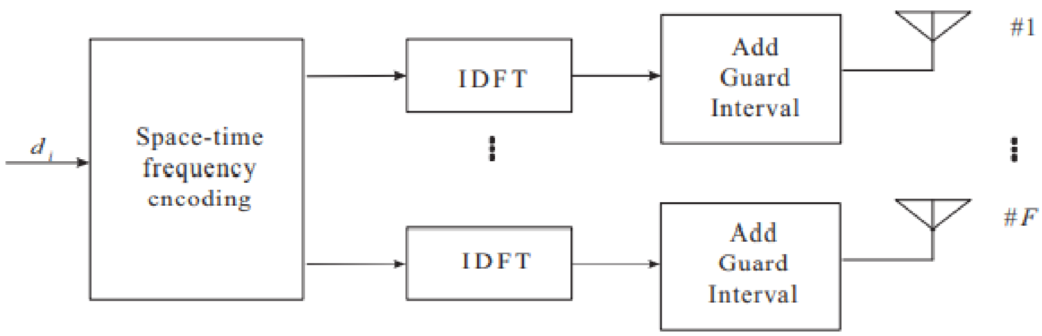

Fig 5.1 Transmitter of Proposed Model

We investigate a MIMO-OFDM system with N subcarriers as shown in Fig. In the system, there are F transmit antennas and M receive antennas. At the t th OFDM symbol period, the output of the space-time-frequency encoder is assumed to be as follows

$$C^{(t)} = \left[c_{0,1}^{(t)}, \ldots c_{N-1,1}^{(t)}, \ldots \ldots c_{0,F}^{(t)} \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots c_{N-1,F}^{(t)}\right] \qquad \ldots 5.1$$

T=0,1,…….,T-1

Where $c_{n,f}^{t}$ is the coded information symbol at the nth subcarrier of the t th OFDM symbol period transmitted from the f th transmit antenna, and T is the number of OFDM symbols in a space-time-frequency codeword. When T = 1, the space time-frequency code reduces to a space-frequency code. After the IDFT processing, at the t th OFDM symbol period, the 1 th sample at the f th transmit antenna is given by

$$S_{t,l}^{(f)} = \frac{1}{N}\left[\sum_{l=0}^{L-1} c_{n,f}^{(t)} e^{-j\frac{2\pi \ln}{N}}\right] \qquad ..5.2$$

$$-N_q \le l < N, f = 1, \ldots\ldots\ldots\ldots\ldots F, \quad t = 0, \ldots\ldots\ldots\ldots \mathrm{T} - 1$$

Where Ng is the length of the cyclic prefix, and we assume that (Ng + 1) < N to keep high transmission efficiency.

In the following, we assume that the channel does not vary over the period of one space-time-frequency codeword (i.e., the period of T OFDM symbols). Furthermore, we assume that the channel impulse responses (CIRs) decay to zero during the cyclic extension, or L 5 (Ng + 1) < N where L is the maximum length of the CIRs. At the m th receive antenna, the 1 th sample at the t th OFDM symbol period is then given by

$$r_{t,1}^{(m)} = \sum_{f=1}^{F} h_i^{(m,f)} * s_{t,l}^{(f)} + z_{t,l}^{(m)} \qquad \ldots\ldots 5.3$$

$$-N_q \le l < N, f = 1, \ldots\ldots\ldots\ldots..F, \quad t = 0, \ldots\ldots\ldots\ldots T - 1$$

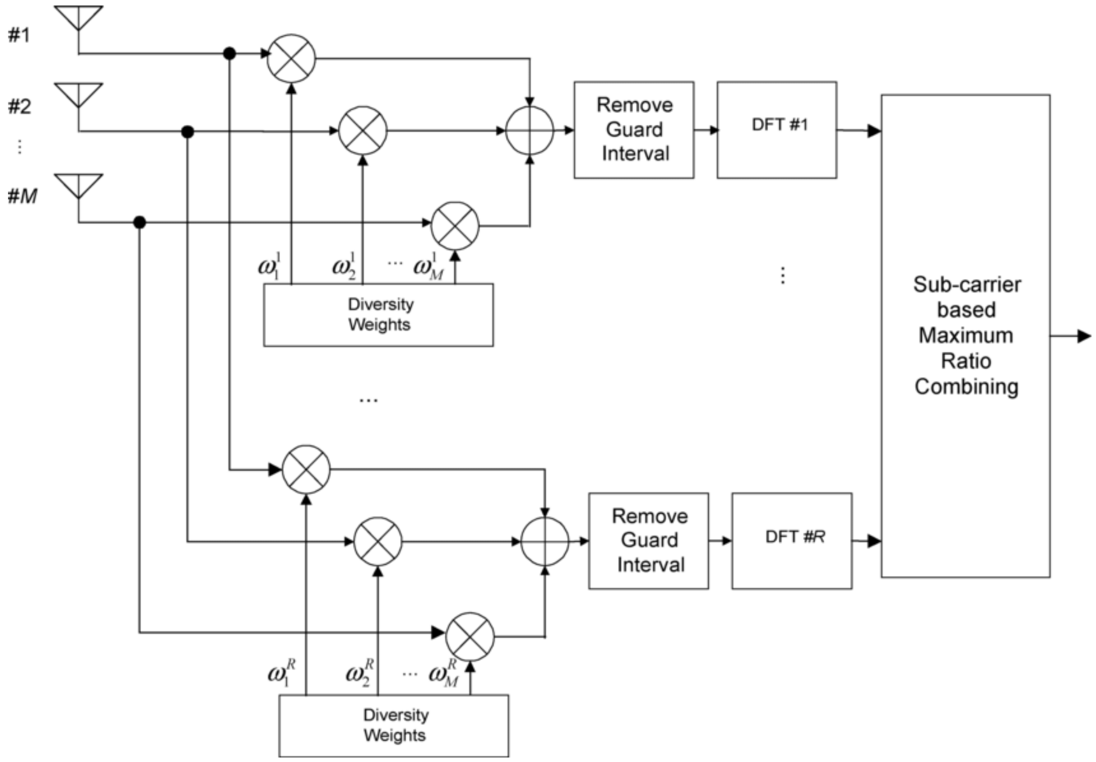

Fig 5.2 Receiver of propsed model

Where * denotes the convolution product, denotes the CIR between the f th transmit antenna and the m th receive antenna, and denotes the additive white Gaussian noise (AWGN) component at the m th receive antenna. At the receiver, before the DFT processing, the M data streams from the output of the M receive antennas are weighted and then combined to form Q branches. After the guard interval removal, the weighted and combined signals are then applied to the DFT processors. Note that there are Q branches, and hence the number of DFT blocks required at the receiver is Q. As a result, compared to the conventional receiver structure, where M DFT blocks are used, the number of DFT blocks employed at the receiver can be reduced when pre-DFT processing is used. For the q th branch, the output of the DFT processor at the t th OFDM symbol period is given by

$$v_{n,q}^{(t)} = \sum_{f=1}^{F} \sum_{m=1}^{M} w_{m,q} H_n^{(m,f)} c_{n,f}^{(t)} + \sum_{m=1}^{M} w_{m,q} \hat{z}_{t,n}^{(m)} \quad .5.4$$

$$H_n^{(m,f)} = \sum_{l=0}^{L-1} H_l^{(m,f)} e^{\frac{-j2\pi ln}{N}}$$ ..5.5

$$\hat{z}t, n^m = \sum l = 0^{N-1} z_{t,l}^m e^{\frac{-j2\pi \, \ln}{N}}$$ ..5.6

and wm,q is the weighting coefficient for the m th receive antenna at the q th branch. In order to keep the noise white and its variance at different branch the same, we assume that the weighting coefficients are normalized (i.e., $\Omega.\Omega^H$ = IQ, where Ω is an M x Q matrix with the (m, q) th entry given by com,q, and IQ is a Q x Q identify matrix).

5.3 Maximum-Likelihood Decoding

Upper and lower bounds on the error probability of linear codes under maximum-likelihood (ML) decoding are shortly surveyed and applied to ensembles of codes on graphs. For upper bounds, we focus on the Gallager bounding techniques and their relation to a variety of other known bounds. Within the class of lower bounds, we address de Caen's based bounds and their improvements, and sphere-packing bounds with their recent developments targeting codes of moderate block lengths.

Consider the classical coded communication model of transmitting one of equally likely signals over a communication channel. Since the error performance of coded communication systems rarely admits exact expressions, tight analytical upper and lower bounds serve as a useful theoretical and engineering tool for assessing performance and for gaining insight into the effect of the main system parameters. As specific good codes are hard to identify, the performance of ensembles of codes is usually considered. The Fano and Gallager bounds were introduced as efficient tools to determine the error exponents of the ensemble of random codes, providing informative results up to the ultimate capacity limit. Since the advent of information theory, the search for efficient coding systems has motivated the introduction of efficient bounding techniques tailored to specific codes or some carefully chosen ensembles of codes. A classical example is the adaptation of the Fano upper bounding technique to specilfic codes, as reported in the seminal dissertation by Gallager (to be referred to as the 1961 Gallager-Fano bound). The incentive for introducing and applying such bounds has strengthened with the introduction of various families of codes defined on graphs which closely approach the channel capacity limit with feasible complexity (e.g., turbo codes, repeat-accumulate codes, and low-density parity-check (LDPC) codes). Clearly, the desired bounds must not be subject to the union bound limitation, since for codes of large enough block lengths, these ensembles of turbo-Iike codes perform reliably at rates which are considerably above the cuto. rate ([R.sub.0]) of the channel (recalling that union bounds for long codes are not informative at the portion of the rate region above [R.sub.0], where the performance of these capacity-approaching codes is most appealing). Although maximum-likelihood (ML) decoding is in general prohibitively complex for long codes, the derivation of upper and lower bounds on the ML decoding error probability is of interest, providing an ultimate indication of the system performance.

A way to calculate the weighting coefficients for the proposed pre-DFT processing scheme. When the ML decoder is employed, the pair-wise error probability (PEP) can be used to denote system performance, which is further determined by the pair-wise codeword distance. The pair-wise codeword distance d2(C, E|H) between a favored coded sequence.

CHAPTER 6

Simulation Results

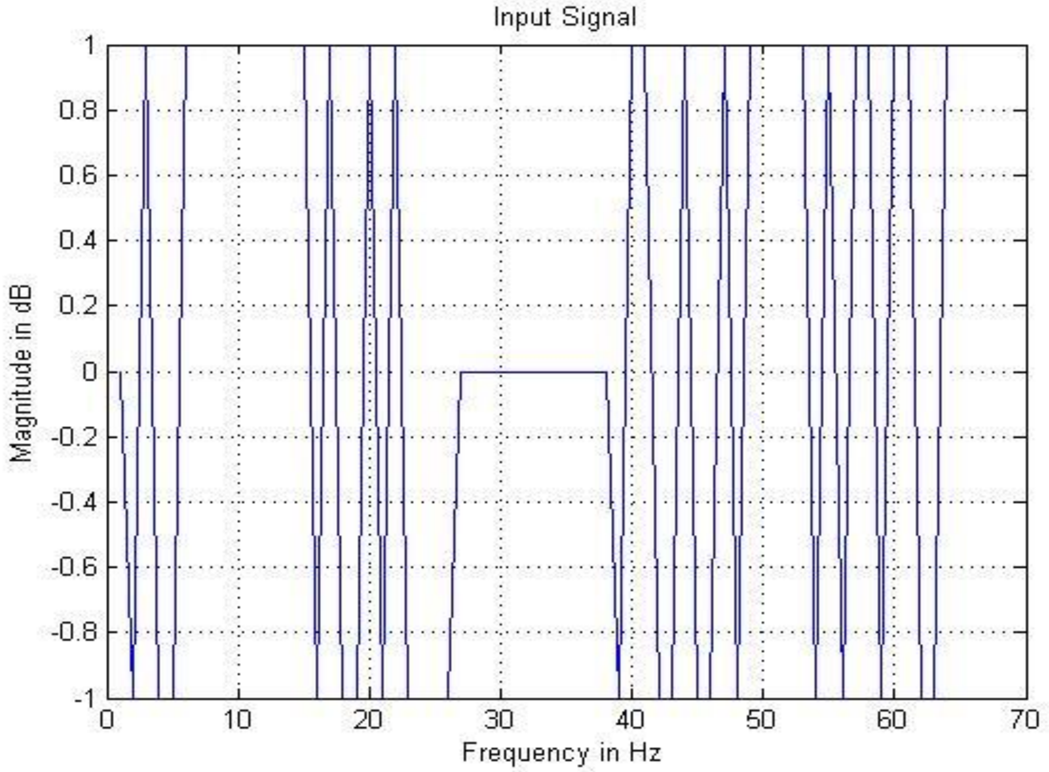

Fig 7.1 Input Signal

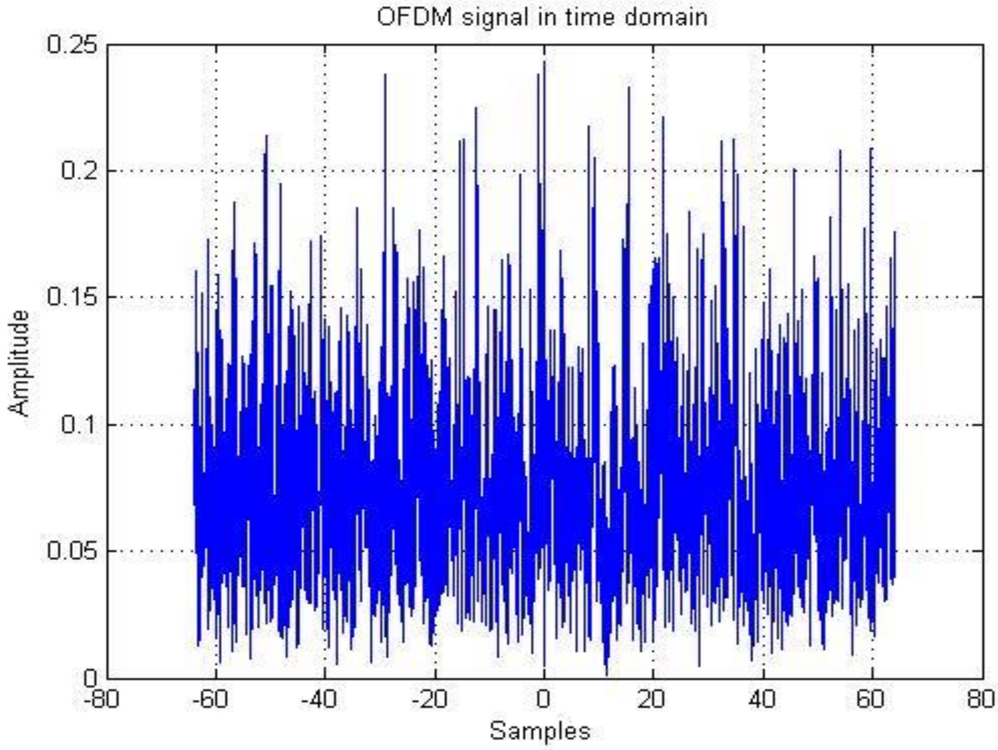

Fig 7.2 OFDM Signal in Time Domain

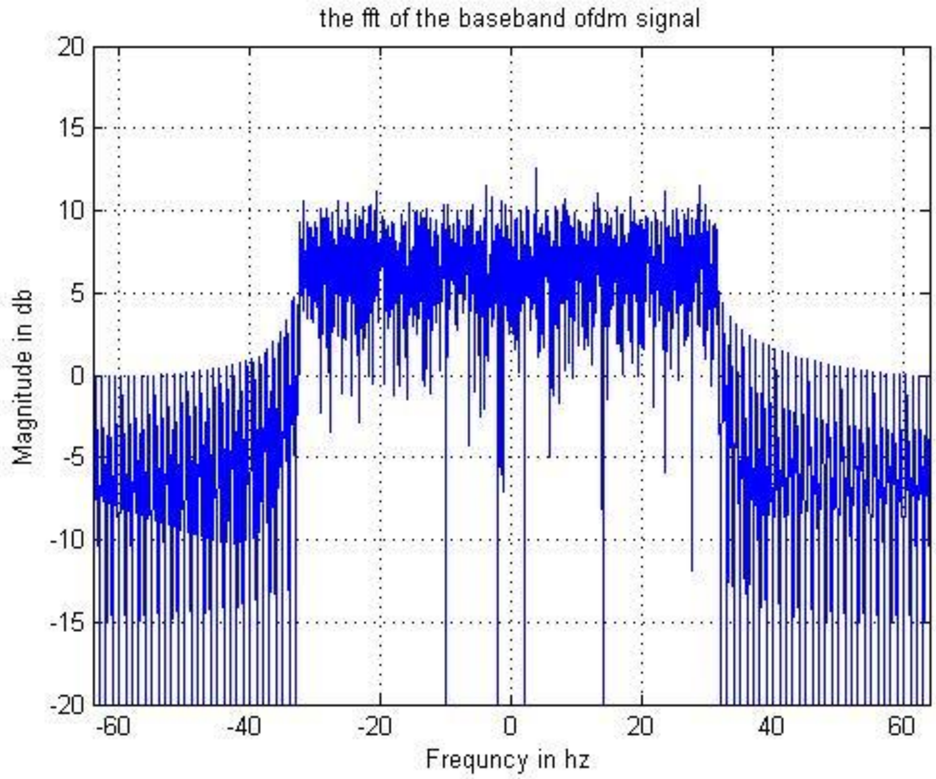

Fig 7.3 FFT of the Baseband OFDM signal

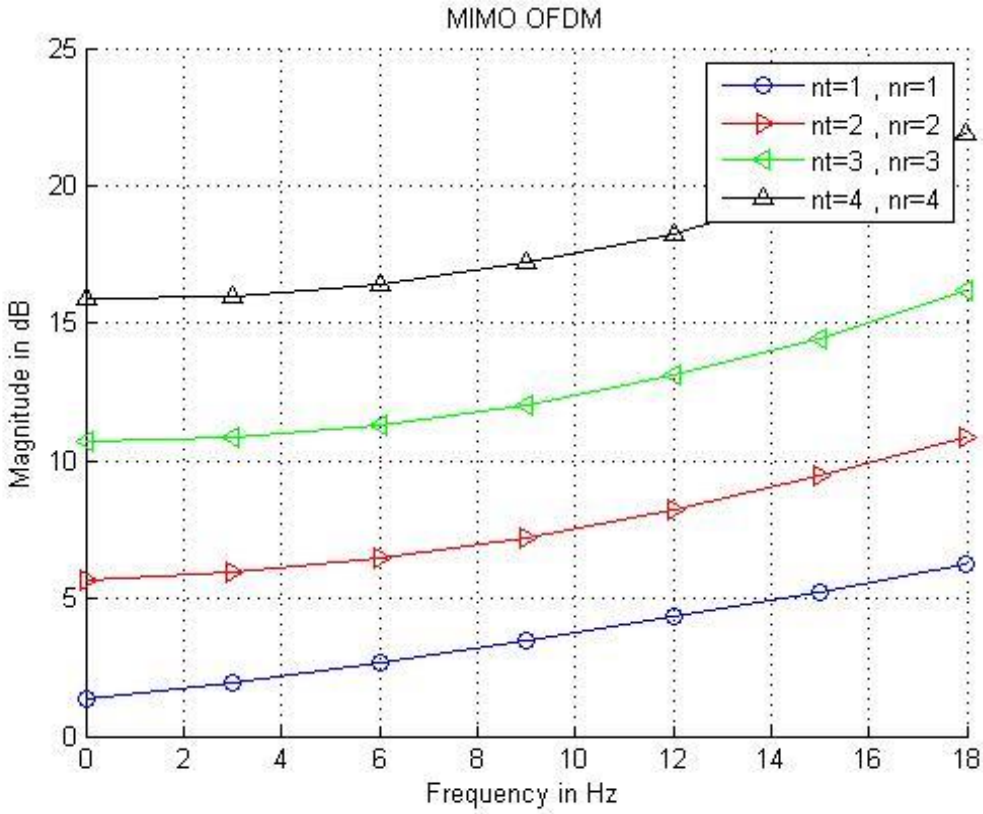

Fig 7.4 MIMO-OFDM

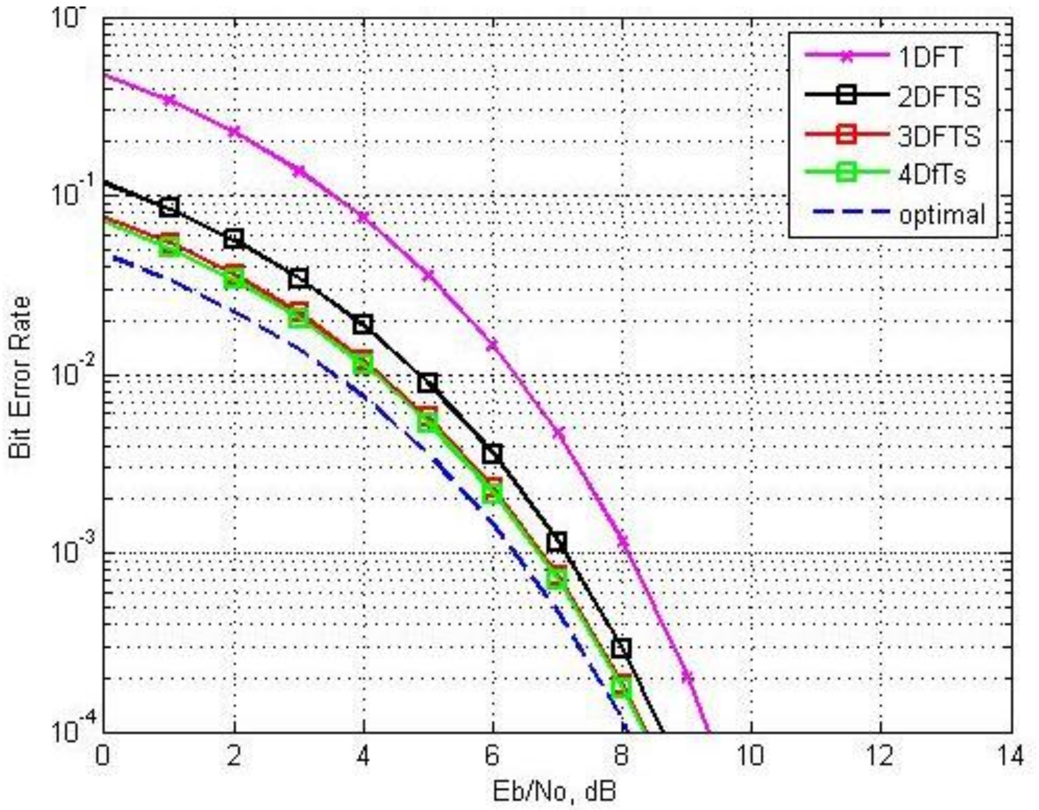

Fig 7.5 BER performance of the proposed scheme with space-time-frequency code over a two-ray equal gain Rayleigh fading channel

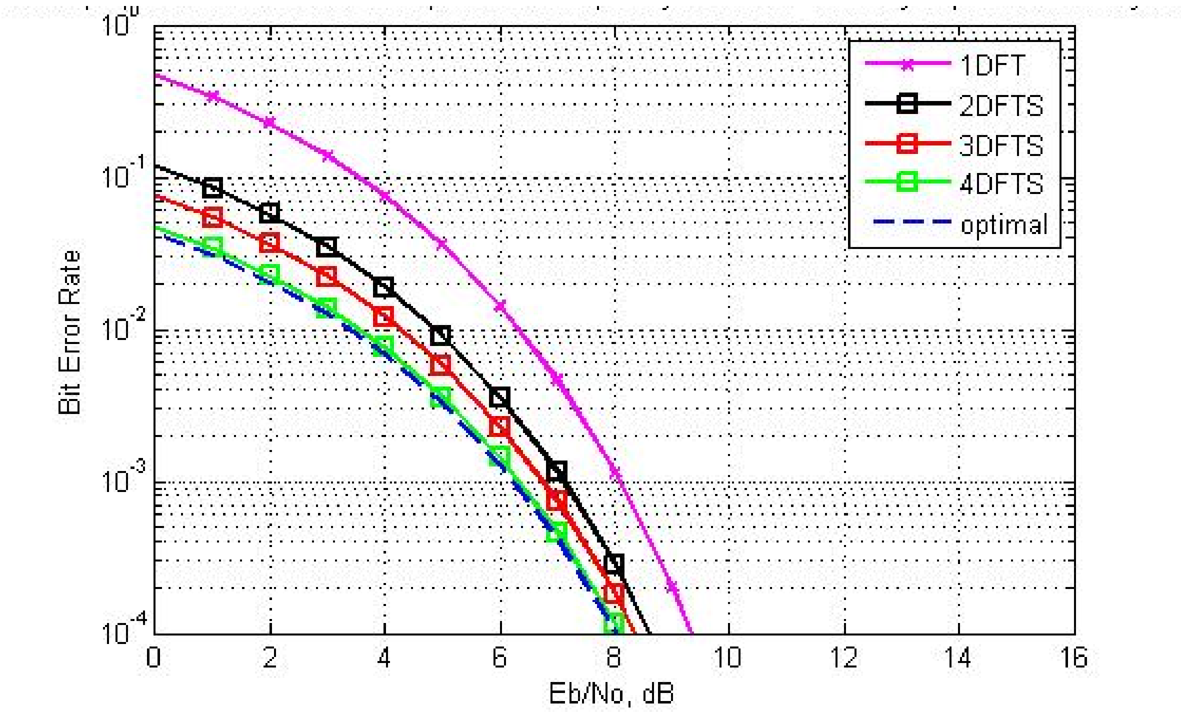

Fig 7.6 BER performance of the proposed scheme with space-time-frequency code over a six-ray exponential decay Rayleigh fading channel

Chapter 7

Conclusions

A pre-DFT processing scheme was proposed for a MIMO-OFDM system with space-time-frequency coding. With the proposed scheme, system complexity and performance can be effectively traded off. A simple weighting coefficients calculation algorithm was also derived. Theoretical analysis and simulation results have shown that the algorithm can be applied for most existing practical space time-frequency codes. Using the proposed scheme, we have also shown that it is possible to use a very limited number of DFT blocks to achieve near optimal system performance.

References

[1] Y. Gong and K. B. Letaief, "An efficient space-frequency coded OFDM system for broadband wireless communications," IEEE Trans. Commun., vol. 51, no. 11, pp. 2019-2029, Nov. 2003.

[2] D. R. V. Jagannadha Rao, V. Shashidhar, Z. A. Khan, and B. S. Rajan, "Lowcomplexity, full-diversity space-time-frequency block codes for MIMO-OFDM," in Proc. IEEE Global Telecommun. Conf., vol. 1, Nov. 29- Dec. 3, 2004, pp. 204-208.

[3] Z. Liu, Y. Xin, and G. B.Giannakis, "Space-time-frequency coded OFDM over frequency-selective fading channels," IEEE Trans. Signal Processing, vol. 50, no. 10, pp. 2465-2476, Oct. 2002.

[4] A. Stamoulis and N. Al-Dhahir, "Impact of space-time block codes on 802.11 network throughput," IEEE Trans.Wireless Commun.,vol. 2.

[5] B. Lu, X. Wang, and K. R. Narayanan, "LDPC-based space-time coded OFDM systems over correlated fading channels performance analysis and receiver design," IEEE Trans. Commun., vol. 50, no. 1, pp. 74-88, Jan. 2002.

[6] D. Agrawal, V. Tarokh, A. Naguib, and N. Seshadri, "Space-time coded OFDM for high data-rate wireless communication over wideband channels," in Proc. IEEE Veh. Technol. Conf., vol. 3, May 18-21, 1998, pp. 2232-2236.

[7] H. Bolcskei, M. Borgmann, and A. J. Paulraj, "Space-frequency coded MIMO OFDM with variable multiplexing-diversity tradeoff," in Proc. IEEE Int. Conf. Commun.,vol 4 May. 11-15, 2003, pp. 2837-2841.

[8] W. Su, Z. Safar, and K. J. R. Liu, "Full-rate full-diversity space-frequency codes with optimum coding advantage," IEEE Trans. Inf. Theory, vol. 51, no. 1, pp. 229 249, Jan. 2005.

[9] L Shao and S. Roy, "Rate-one space-frequency block codes with maximum diversity gain for MIMO-OFDM," IEEE Trans. Wireless Commun., vol. 4, no. 4, pp. 1674-1687, July 2005.

[10] M. Okada and S. Komaki, Pre-DFT combining space diversity assisted COFDM." IEEE Trans. Veh. Technol., vol. 50, no. 2, pp. 487-496, Mar. 2001.

[11] D. Huang and K. B. Letaief, "Pre-DFT processing using eigen-analysis for coded OFDM with multiple receive antennas," IEEE Trans. Commun., vol. 52, no. 11, pp. 2019-2027, Nov. 2004

[12] S. Hara, M. Budsabathon, and Y. Hara, "A pre-FFT OFDM adaptive antenna array with eigenvector combining," in Proc. IEEE Int. Conf. Commun., vol. 4. pp. 2412-2416.

[13] D. Huang and K. B. Letaief, "Symbol-based space diversity for coded OFDM systems," IEEE Trans. Wireless Commun., vol. 3, no. 1, pp. 117-127, Jan. 2004.