ANKARA YILDIRIM BEYAZIT UNIVERSITY

GRADUATE SCHOOL OF NATURAL AND APPLIED SCIENCES

ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING WITH SIGNAL SPACE DIVERSITY

M.Sc. Thesis by

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ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING WITH SIGNAL SPACE DIVERSITY

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We have read the thesis entitled "**ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING WITH SIGNAL SPACE DIVERSITY**" completed by **Murat Can KARAKOÇ** under supervision of **Assist. Prof. Serdar ÖZYURT** and we certify that in our opinion it is fully adequate, in scope and in quality, as a thesis for the degree of Master of Science.

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ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING WITH SIGNAL SPACE DIVERSITY

ABSTRACT

People need more bandwidth to transmit voice, image and data. Orthogonal frequency division multiplexing (OFDM) is an alternative method of increasing data rate by providing parallel data transmission, no guard band and overlapping channels that provide spectral efficiency and has a good performance to combat frequency selective fading channels. Alamouti space time block coding (STBC) is the best alternative to provide diversity order of two for the two transmit antennas scenario without no channel state information at transmitter. Alamouti STBC does not have penalty in data rate, at two time instants two symbols sent by applying Alamouti encoding scheme. Signal space diversity (SSD) is an efficient technique that does not need extra bandwidth and energy and it can show performance improvement in the system. Using this technique for the physical layer security is being worked on nowadays. By using this technique at physical layer privacy of the communication system can be provided. In this study we incorporate three concepts which are OFDM, Alamouti STBC and SSD. Assumed that channel is frequency selective, by applying OFDM frequency selective fading channel is converted into flat fading channel. Theoretical and simulation results are provided for this incorporation.

Keywords: Orthogonal frequency division multiplexing, space time block code, signal space diversity

DİKGEN FREKANS BÖLMELİ ÇOKLAYICI İLE İŞARET UZAYI ÇEŞİTLEMESİ

ÖZ

İnsanlar ses, görüntü ve veri transferi için daha fazla bant genişliğine ihtiyaç duymaktadır. Dikgen Frekans bölmeli çoklama paralel veri transferini sağlayarak veri gönderim hızını artırmakta, gard bant kullanmamak ve üst üste çakışan kanallar aracılığıyla spektral verim sağlamakta ve frekans seçici kanallara karşı gösterdiği başarım sayesinde artan ihtiyaçları giderme konusunda iyi bir alternatif olmaktadır. İki verici antenli Alamouti uzay zaman blok kodlama (space time block code, STBC) yapısı verici sistemin kanal durum bilgisine (channel state information, CSI) sahip olmadan iki sıralı çeşitliliğe sahip olmasını sağlamaktadır. Alamouti STBC yapsı iki zaman anında iki veri gönderek veri hızının aynı kalmasını sağlamaktadır. İşaret uzayı çeşitlemesi (signal space diversity, SSD) fazladan bir enerji ve bant genişliği kullanmadan başarım iyileşmesine neden olmaktadır. Günümüzde SSD fiziksel katmanda güvenlik amacıylada kullanılması konusunda çalışılmaktadır. Bu çalışmada OFDM, Alamouti STBC ve SSD birleştirilmiştir. Frekans seçici sönümlemeli kanal OFDM kullanarak düz sönümlemeli kanal haline getirilmiştir. Bu üç konseptin birleştirilmesi sonucu teorik ve benzetim sonuçları doğrulanmış ve verilmiştir.

Anahtar sözcükler: Dikgen frekans bölmeli çoklayıcı, uzay zaman blok kodlaması, işaret uzayı çeşitlemesi.

CONTENTS

NOMENCLATURE

Acronyms

STBC Space Time Block Code

OFDM Orthogonal Frequency Division Multiplexing

BPSK Binary Phase Shift Keying

SSD Signal Space Diversity

CP Cyclic Prefix

SISO Single Input Single Output

SIMO Single Input Multiple Output

MISO Multiple Input Single Output

MIMO Multiple Input Multiple Output

CSI Channel State Information

BER Bit Error Rate

PDP Power Delay Profile

LIST OF FIGURES

CHAPTER 1

INTRODUCTION

Chang (1966) wrote a paper which shows deriving a combination of band-limited orthogonal signals for multichannel data transmission. After the development of the telegraph, Nyquist worked on the maximum achievable data rate at W Hz band-limited signal. The achievable data rate is found to be equal to 2W pulses/s by him [1]. Nyquist data rate per each channel can be reached by using this technique. Interesting and efficient result of this work that is neither intersymbol interference nor intercarrier interference can occur regardless of the phase noise in the transmitter and receiver filters. This paper also did not investigate phase noise caused by Doppler Effect at the channel. These are the weaknesses of the paper. Profound analysis of the subchannel correlation was done at this work. Derivation of the orthogonal signal for the band-limited channel is a good basement for using in the future.

At [2] achievement of parallel data transmission is mentioned. Parallel data are transferred through overlapping channels. Each channel has duration f Hz and spaced f/2 Hz apart in frequency domain. In general f/2 Hz is defined as subcarriers spacing for OFDM system that has duration of f Hz. The aim of the work is to eliminate cross-talks between different channels. Parallel transmission of the data over overlapping channels might be solution for the problem. Hovewer there are some penalties for parallel transmission, such as efficient roll-off factor for transmitter and receiver filter design is examined. Amplitude and delay distortions are also investigated.

Frequency Division Multiplexing is a multiplexing technique. By using discrete Fourier transform data can be transferred. Weinstein and Ebert (1971) mentioned this concept [3]. Transmission of the data can be divided into two parts which are parallel and serial. If the transmission of the symbols is in serial mode, at each time instant one symbol is sent. However in the parallel transmission of data at each time instant, a piece of all data can be sent. Symbols carried by orthogonal subcarriers are arranged by applying discrete Fourier transform specifically fast Fourier transform.

Mathematical analysis of the received signals at distortionless channel is made. The effect of channels on the amplitude and the phase characteristics of the transmitted signal is also investigated for coherent reception. The main aim of a communication system is to transfer data bits as reliable as possible. Mean square error derivation is done in this paper.

After fundamental works [4], [2], [3] about orthogonal frequency division multiplexing, at [5] a survey about orthogonal frequency division multiplexing has been studied.

Alamouti invented space time block coding for different antenna schemes such as 2x1, 1x2 and 2x2, where the first one is the number of transmitter antennas, second is the number of receiver antennas [6]. This scheme increases diversity, hence improvement in bit-error-rate performance are obtained.

At [7] Alamouti scheme and orthogonal frequency division multiplexing are used. Alamouti space time block coding is used in flat fading channels. Frequency selective fading channel is converted to narrowband flat fading channels with the help of orthogonal frequency division multiplexing. Multiple narrowband channels can be named as subcarrier. In this work a new subcarrier allocation strategy is developed. The developed strategy is applied to Alamouti scheme.

Space time block coding scheme is preferred for robustness with narrowband channels and capacity improvement in low signal-to-noise-ratio. In [8] orthogonal frequency division multiplexing and space time block coding are combined to improve data rate at 1 MHz bandwidth. However it is necessary to use a new antenna for each subcarrier, and this causes increase of the cost of the communication system. In this paper symbol timing and carrier frequencies are perfectly synchronized. Channel estimation is performed by receiver. This estimation is sent from receiver to transmitter which is named channel state information. In the absence of channel state information, Viterbi algorithm is used by sending pilots between data symbols.

Alamouti space time block coding and spreading matrix are used and merged at [9]. An orthogonal frequency division multiplexing symbol is divided into two parts. First part is transmitted from first antenna and second part is transmitted from the second antenna. After that following orthogonal frequency division multiplexing symbol is divided into two parts and these are sent from the third and fourth antennas respectively. Spreading matrix is also used to increase diversity.

At [10] interleaving pattern is introduced. A complex data symbol includes I and Q components. One subcarrier of the orthogonal frequency division multiplexing symbol carries one symbol. In this work any subcarrier's Q component is exchanged with the following symbol's Q component. In orthogonal frequency division multiplexing a symbol is multiplied with spreading matrix to get full diversity gain.

Frequency diversity is obtained if interleaving operation is made before IFFT. Spreading symbols by using spreading namely Hadamard matrix gives both frequency and time diversity, spreading matrix which is used or not is investigated for SISO systems [11]. Low density parity check coded multi-carrier code division multiple access with subcarrier interleaving is investigated at [12].

Interleaving can also be used for security purposes in networking. There are several layers in network systems. Each layer of networking systems have special security structure. The last part of the networking is physical layer. Physical layer security can be supported between transmitter and legitimate user by using interleaving. In wireless communication systems channel state information should be estimated by channel estimation block before equalization in coherent reception scheme. Channel state information can be used for arrangement of subcarrier interleaving scheme.

Interleaving can be used for security purposes at [13], [14]. Assuming that transmitter has full channel state information, this information comprises the important part of the physical layer security concept because of dynamic subcarrier interleaving allocation strategy is developed at the legitimate receiver. If eavesdropper can solve interleaving strategy, one cannot recover, because interleaving strategy is developed by subcarrier channel gain. For instance any Q component channel gain is weaker when compared to other Q components' channel gains. This subcarrier Q component is changed with another I or Q component of the same orthogonal frequency division multiplexing symbol. The channel gains of the subcarriers are different for legitimate and illegitimate user. This strategy cannot work for illegitimate user.

At works [15], [16] coordinate interleaving with space time block code orthogonal frequency division multiplexing-index modulation is investigated. Orthogonal frequency division multiplexing- index modulation is a new concept where subcarriers are divided into subblocks and each block is composed of two subblocks, one for index and another for data. Data and index subblocks are interleaved inside and performance analysis are indicated.

Error performance, optimum rotation angle and interleaving pattern for the phase shift keying for single input single output systems are investigated at [17].

CHAPTER 2

ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING OVERVIEW

Orthogonal frequency division multiplexing (OFDM) is a popular multiplexing technique. Multicarrier modulation or spread spectrum are outstanding techniques used for high-rate wireless communication to compensate intersymbol interference (ISI) [18]. OFDM technique is currently used in Digital Video Broadcasting (DVB), Digital Audio Broadcasting (DAB), Asymmetric Digital Subcarrier Line (ADSL), Wireless Local Area Networks (WLAN) and downlink transmission of LTE systems. Moreover this is a good choice for new radio systems such as 5G and beyond.

Power efficient signaling for large numbers of users on the same channel are provided by OFDM using the orthogonality characteristic of the M-FSK [19]. By using OFDM, all the orthogonal carriers are transmitted simultaneously. In this perspective OFDM could be a hybrid model of the multi-carrier modulation and FSK [20].

2.1 OFDM System Model

The low-pass equivalent OFDM signal is represented as [21],

$$X(t) = \sum_{n} X_n(t - nT) \qquad (\mathbf{2.1})$$

$$X_n(t) = \sum_{k=0}^{N-1} X_{n,k} \emptyset_k(t), \quad 0 \le t < T \qquad (\mathbf{2.2})$$

$$\emptyset_k(t) = \begin{cases} e^{j2\pi f_k t} & {\scriptstyle 0 \le t < T} \\ 0 & elsewhere \end{cases} \qquad (\mathbf{2.3})$$

where T is the symbol period, N is the number of subcarriers, $\emptyset_k$ is the k^{th} subcarrier with frequency f_k and $X_n(t)$ is the n^{th} OFDM symbol.

The orthogonality principle can be written in Equation 2.4 as:

$$\int_0^T \emptyset_k(t)\emptyset_k{}^*(t)dt = a\delta(k-l) \qquad (\mathbf{2.4})$$

The frequency allocation strategy should be done using the following equation to maintain orthogonality amongst all subcarriers of an OFDM symbol.

$$f_k = f_0 + \frac{k}{T} \qquad (\mathbf{2.5})$$

where f_0 is the frequency of the first subcarrier and T is the symbol period.

The structure of the OFDM modulator can be designed as in the Figure 2.1,

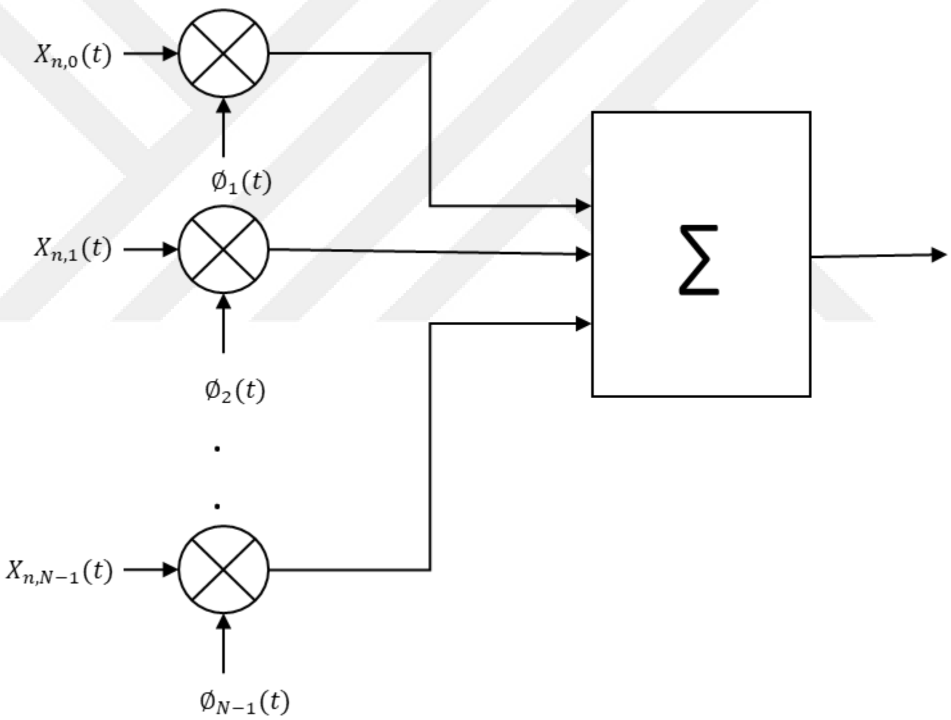

Figure 2.1 OFDM transmitter structure [21]

The OFDM demodulator structure using a match filter for each subcarrier can be plotted as in Figure 2.2,

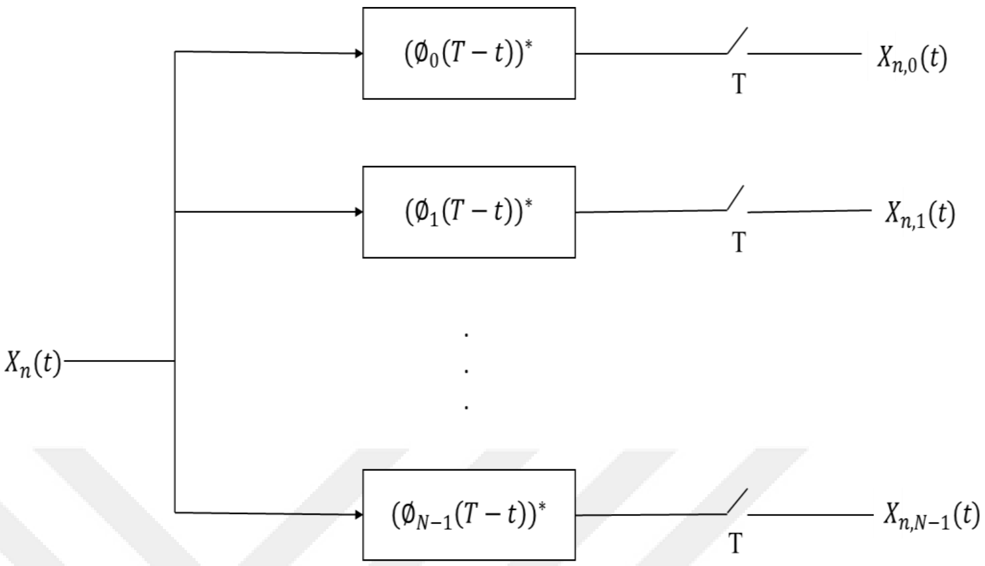

Figure 2.2 OFDM receiver structure [21]

Modulator used at transmitter and demodulator used at receiver need many mixers and filters to multiply data with the carrier and receive the sent data exactly. Transmitter uses inverse discrete Fourier transform (IDFT) or inverse fast Fourier transform (IFFT) that is the algorithm to increase speed of the IDFT operation by not using many mixers. Likewise receiver uses discrete Fourier transform (DFT) or fast Fourier transform instead of many match filters.

The DFT and IDFT operations are given at following equations

$$X[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x[k] e^{-j\frac{2\pi nk}{N}}, \quad 0 \leq k \leq N \qquad (\mathbf{2.6})$$

$$x[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X[k] e^{j\frac{2\pi nk}{N}}, \quad 0 \leq n \leq N \qquad (\mathbf{2.7})$$

The positioning of the subcarriers in an OFDM symbol are displayed in the Figure 2.3. Subcarriers are in an overlapping structure.

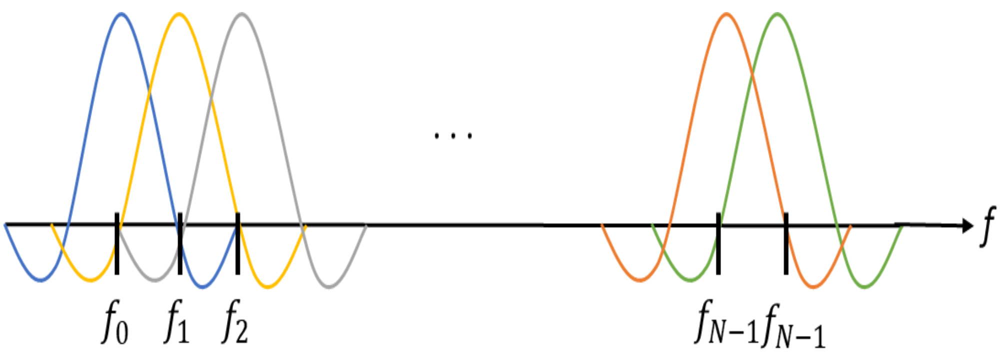

Figure 2.3 OFDM subcarrier structure

2.2 Cyclic Prefix (CP)

Cyclic prefix is used in OFDM systems. The last part of the OFDM symbol is added to the beginning of the OFDM symbol. The length of the last part which is added to the prefix of the OFDM symbol is related to the multipath delay spread of the OFDM symbol. Size of the CP must be greater than the maximum delay spread of an OFDM symbol, in order to prevent ISI between consecutive OFDM symbols.

Demonstration of the transmitted OFDM symbol structure with CP is given at Figure 2.4

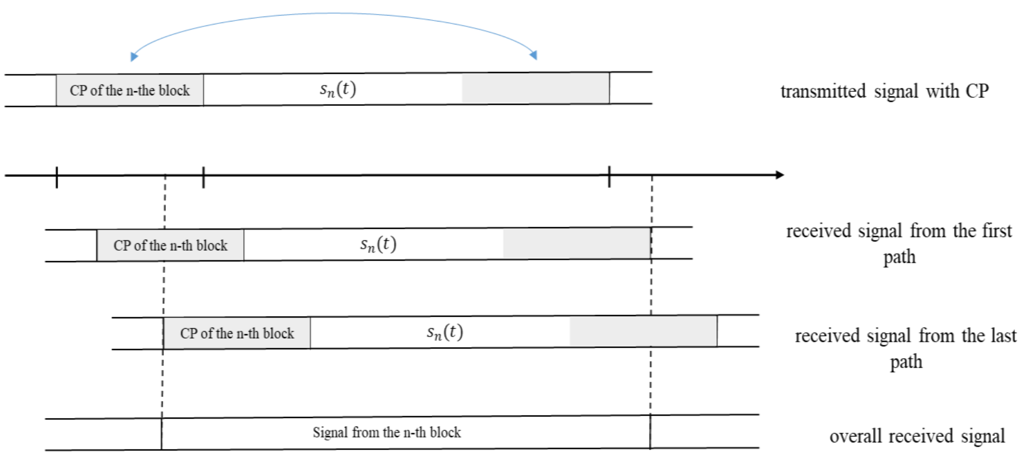

Figure 2.4 OFDM CP structure [20]

CHAPTER 3

COMMUNICATION CHANNEL ANALYSIS[1]

The general equation for the communication systems can be written as in Equation 3.1.

$$\boldsymbol{y} = \boldsymbol{hx} + \boldsymbol{n} \tag{3.1}$$

where $\boldsymbol{n}$ is the additive white Gaussian noise (AWGN), $\boldsymbol{x}$ is the transmitted signal, $\boldsymbol{h}$ is the fading coefficient and $\boldsymbol{y}$ is the received signal. Bold face indicate a vector or matrix representation.

The parameter $\boldsymbol{n}$ emerges from interference experienced in transmission or from electronic components and amplifiers at the receiver of communication system physically [1]. Unless the noise is aroused from transmission, it can be denoted as thermal noise caused by the electronic components and amplifiers [1]. Statistically this noise is defined as a Gaussian noise process. Hence, mathematically this channel can be named as additive Gaussian noise channel [1].

A communication system can be classified into two parts:

- Wireline communication
- Wireless communication

3.1 Wireline Communication

Generally wireline communication systems have only one path between transmitter and receiver. The $\boldsymbol{h}$ component in Equation-3.1 is 1 for wireline communication.

3.2 Wireless Communication

Wireless communication systems have more than one path because of the reflection, scattering and diffraction between transmitter and receiver. Wireless communication system can be represented as in Figure-3.1.

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[1] The concept of this part is written with the help of video lecture notes of Aditya K. Jagannatham at Indian Institute of Technology Kanpur.

The details of the multipath components can be calculated using such models that ray tracing, Okumura, Hata, etc. In ray tracing model Maxwell's electromagnetic wave equations are used. This approach is impractical because this model includes complex calculations. Another model that is commonly used is the statistical model. The aim of this operation is making estimation of the $\boldsymbol{h}$ given in Equation 3.1

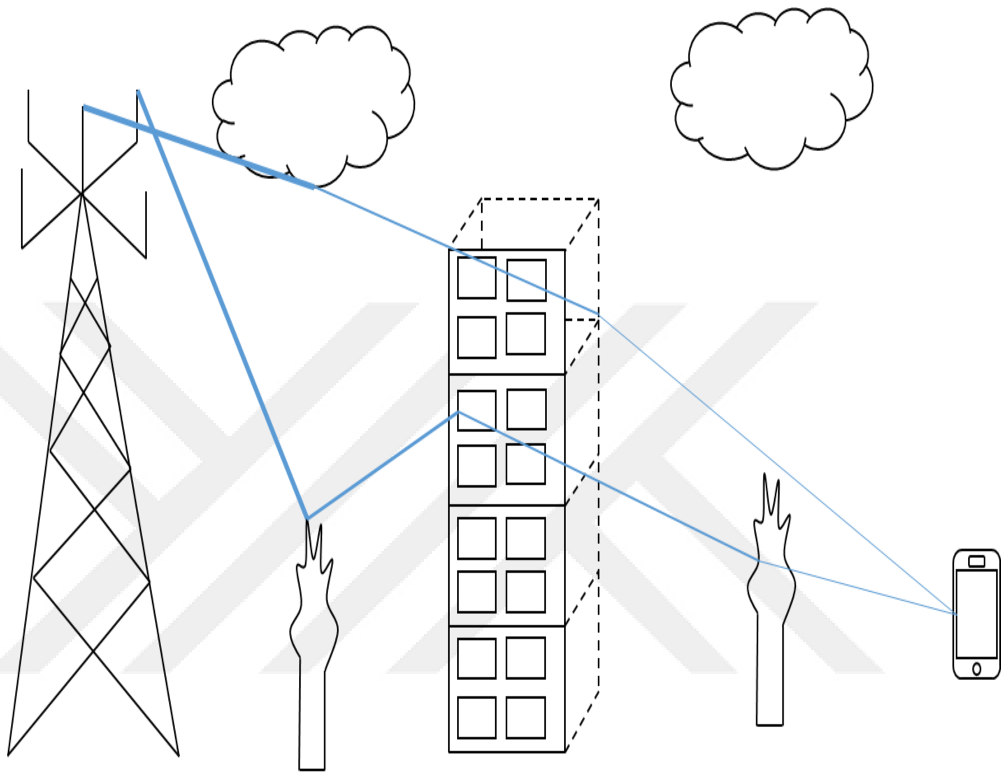

Figure 3.1 Wireless Communication Environment

The receiver of the wireless communication system takes different signals that are multipath components of the transmitted signal. Equation 3.2 can be written for representing multipath components.

$$y_1(t) = Re\left\{a_1 s_b(t-\tau_1)e^{j2\pi f_c(t-\tau_1)}\right\}$$

$$\vdots \qquad \textbf{(3.2)}$$

$$y_{L-1}(t) = Re\left\{a_{L-1} s_b(t-\tau_{L-1})e^{j2\pi f_c(t-\tau_{L-1})}\right\}$$

where a_i is the constructive or destructive attenuation component, s_b is the transmitted baseband signal, f_c is the carrier frequency, τ is the delay.

τ_1 is the first received signal, which may be taken at the 0^{th} time instant, commonly named as line-of-sight signal and τ_{L-1} is the last received signal at any time. Briefly $\tau_1, \dots.., \tau_{L-1}$ is the all components taken by the receiver.

Summation of the all multipath components can be written as in Equation 3.3,

$$y(t) = Re\{\sum_{i=0}^{L-1} a_i s_b(t - \tau_i) e^{j2\pi f_c (t-\tau_i)}\} \qquad \mathbf{(3.3)}$$

Regardless of the radio frequency (RF) namely carrier frequency component written in Equation 3.4

$$y(t) = Re\{\{\sum_{i=0}^{L-1} a_i s_b(t - \tau_i) e^{-j2\pi f_c \tau_i}\} e^{j2\pi f_c t}\} \qquad \mathbf{(3.4)}$$

The power of the signals taken in the receiver power can be illustrated as in Figure 3.2 which named as power delay profile (PDP). PDP is usually shown with an exponentially decaying curve.

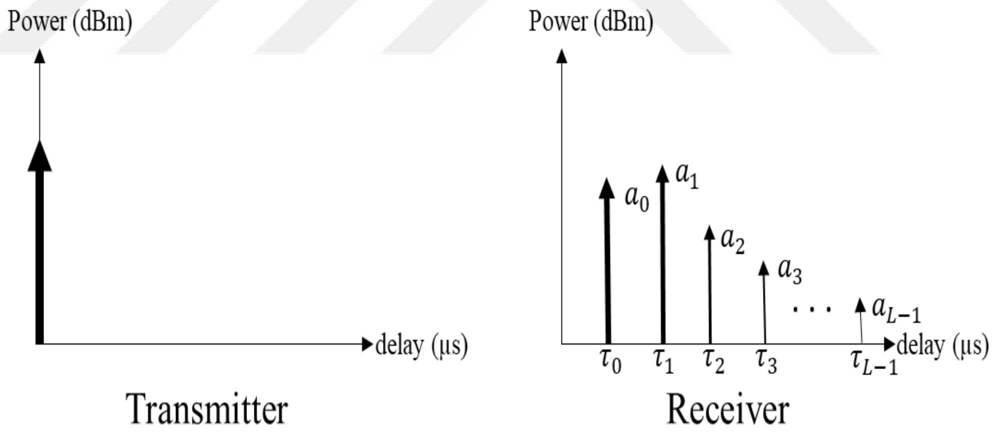

Figure 3.2 Power Delay Profile

Figure 3.2 a_i^{th} index represents the power of the received signal. Generally τ_0 is the first received signal named as line-of-sight (LOS). Receiver plot of the Figure 3.2 can be written in equation form as in Equation 3.5.

$$|h(\tau)|^2 = \sum_{i=0}^{L-1} |a_i|^2 \, \delta(\tau - \tau_i) \qquad \mathbf{(3.5)}$$

With regard to time-dispersive properties, wideband multipath signals/channels are frequently determined by their mean excess delay ($\bar{\tau}$) and root mean square (rms) delay spread (σ_τ) [19].

Expected value of the PDP is the mean excess delay and it is calculated by using Equation 3.6.

$$\bar{\tau} = \frac{\sum_k a_k^2 \tau_k}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k}{\sum_k P(\tau_k)} \qquad \textbf{(3.6)}$$

Variance of the PDP is the rms delay spread which is calculated by using Equation 3.7.

$$\sigma_\tau = \sqrt{\overline{\tau^2} - (\bar{\tau})^2} \qquad \textbf{(3.7)}$$

where $\overline{\tau^2}$ is

$$\overline{\tau^2} = \frac{\sum_k a_k^2 \tau_k^2}{\sum_k a_k^2} = \frac{\sum_k P(\tau_k)\tau_k^2}{\sum_k P(\tau_k)} \qquad \textbf{(3.8)}$$

The first detectable signal which arrives at the receiver at τ_0 is a reference observation for measuring other delayed signals [19].

When Fourier transform is applied to the transmitted and received signals, the system is converted from time domain to frequency domain. Frequency domain representation of the system is displayed in Figure 3.3

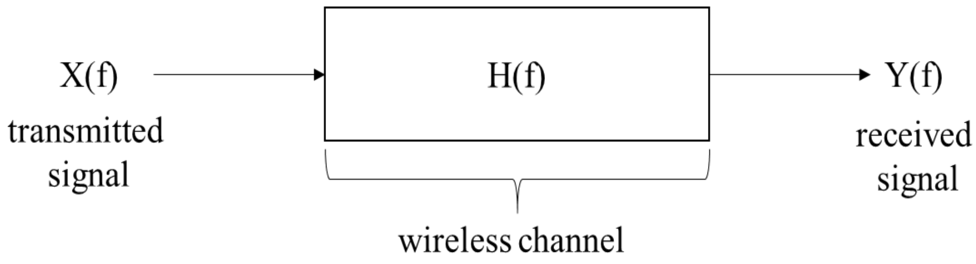

Figure 3.3 Communication system model in frequency domain

where H(f) is the channel frequency response.

3.2.1 Statistical Properties of the Wireless Channel

Statistical properties of the channel can be examined as flat during coherence bandwidth where a channel effects all spectral components approximately with equal gain and linear phase [19].

The autocorrelation property of the time-varying multipath channel can be written in Equation 3.9 in frequency domain [18].

$$A_C(f_1, f_2;\ \Delta t) = \quad \mathbf{(3.9)}$$

$$= E\left[\int_{-\infty}^{\infty} c^*(\tau_1; t) e^{j2\pi f \tau_1} d\tau_1 \int_{-\infty}^{\infty} c(\tau_2; t+\Delta t) e^{-j2\pi f \tau_2} d\tau_2\right] \quad \mathbf{(3.9a)}$$

$$= \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} E[c^*(\tau_1; t) c(\tau_2; t+\Delta t)\, e^{j2\pi f_1 \tau_1} e^{-j2\pi f_2 \tau_2} d\tau_1 d\tau_2$$

$$= \int_{-\infty}^{\infty} A_c\,(\tau, \Delta t) e^{-j2\pi (f_2 - f_1) d\tau}$$

where $A_c(\Delta f; \Delta t)$ $\Delta f = f_2 - f_1$.By transmitting two sinusoids between transmitter and receiver autocorrelation function $A_c(\Delta f; \Delta t)$ can be measured. Δf is the frequency difference between two sinusoidal signals.

If we define $A_C(\Delta f) \triangleq A_c(\Delta f; 0)$

$$A_C(\Delta f) = \int_{-\infty}^{\infty} A_C(\tau) \quad \mathbf{(3.10)}$$

The autocorrelation of $A_C(\Delta f)$ is flat across the bandwidth of the narrowband signal, so this signal will experience flat fading or negligible ISI.

Fading will be independent amongst the different parts of the signal bandwidth, when the autocorrelation $A_C(\Delta f)$ goes to zero inside the bandwidth of the wideband [18]. Thus the channel is named frequency selective. This selectivity causes to ISI.

Generally coherence bandwidth can considered as 20 % of the rms delay spread. Flat fading and frequency selective fading can be represented as in Figure 3.4a and Figure 3.4b respectively.

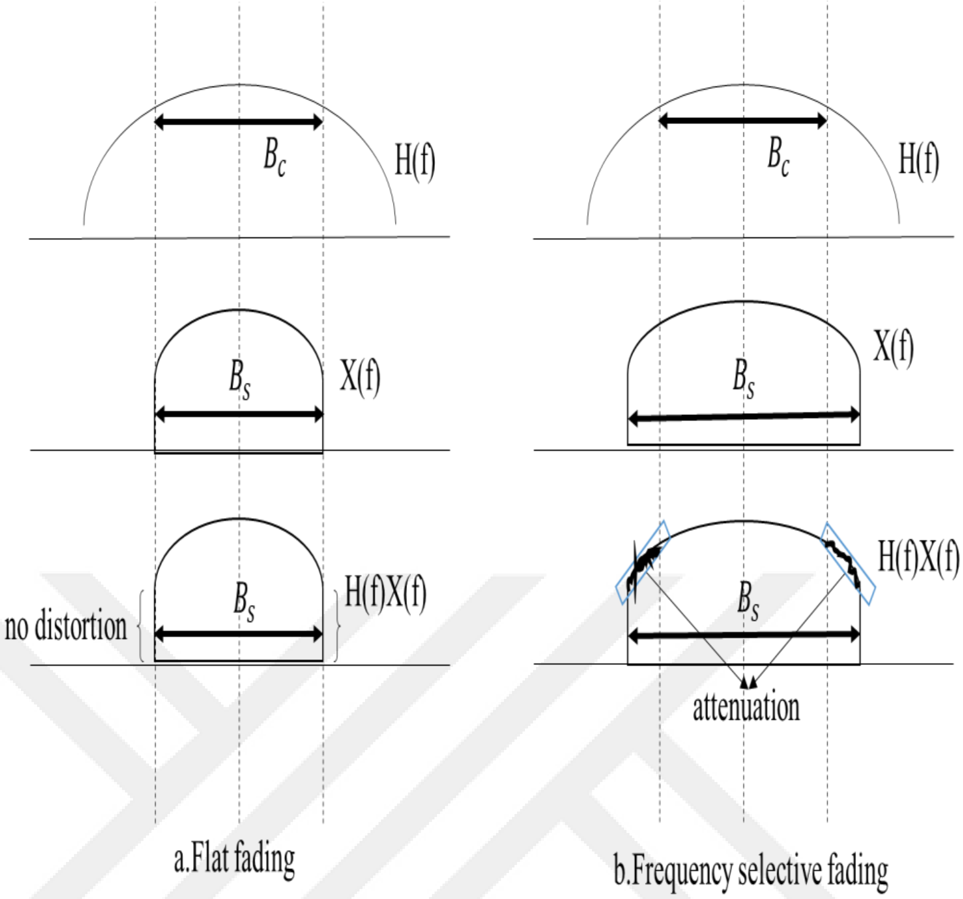

Figure 3.4 Flat and frequency selective fading channels

There is phenomenon which implies that frequency domain parameters can be written in time domain and also inverse is true. In this perspective fast fading in time domain is the inversely proportional equivalent of frequency selective fading in frequency domain, namely the communication channel is changing rapidly. Slow fading is inverse of the fast fading, namely the communication channel is changing slowly.

CHAPTER 4

MULTIPLE INPUT MULTIPLE OUTPUT

Space is a new dimension when using multiple antennas at the transmitter and/or receiver wireless communication link [22]. Performance can be increased considerably, if this scheme is applied correctly. As far as navigation of antenna arrays are made, mitigation of co-channel interference can be obtained. [22]. The use of antenna arrays goes back to the years of World War 2 in especially radar applications [22]. Fundamental research were done at Bell Labs in mid-1990's. The development of antenna array technology is displayed in Figure 4.1.

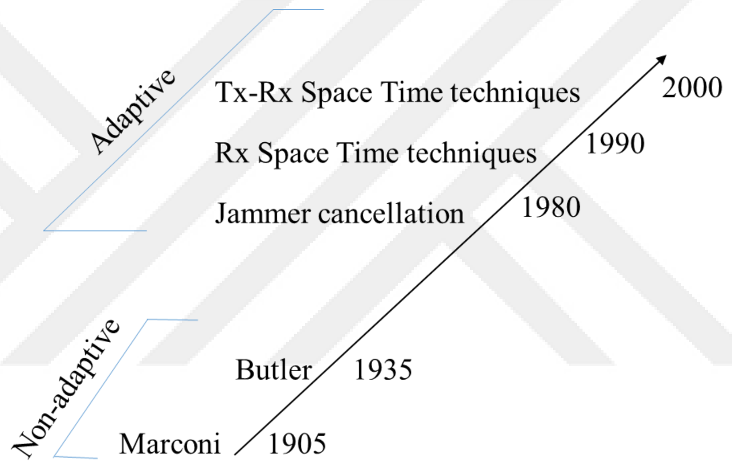

Figure 0 Antenna development

The average SNR increase in the system performance which emerges from effect of multiple antennas at the receiver, transmitter or both can be defined as array gain [22]. The channel is said to be in a fade, when the signal power drops significantly, so that the signal power in a wireless channel fluctuates (or fades). To combat fading, diversity is used in wireless channels.

Schemes with different number of transmitter and receiver antennas are shown in Figure 4.2. Independently faded versions of the transmitted signal can be received by the receiver antenna array. Hence, the fading amount of the signal is decreased.

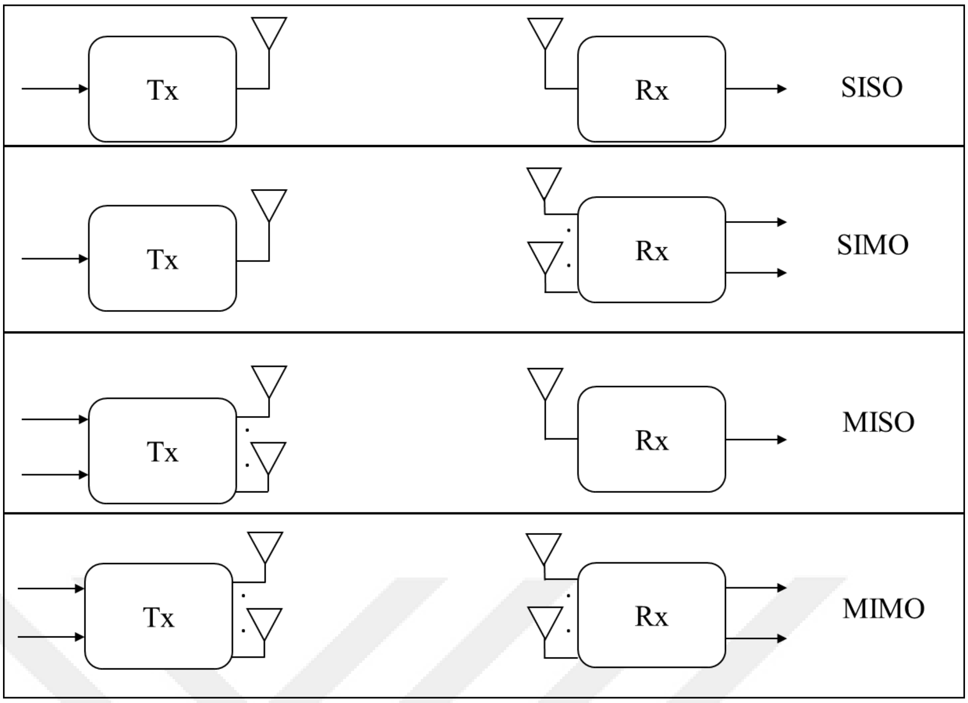

Figure 4.2 Different antenna schemes

The received signal for single input single output (SISO) system can be written as in Equation 4.1.

$$y(t) = \int_{0}^{\tau_{total}} h(\tau, t) s(t - \tau) d\tau = h(\tau, t) * s(t) \qquad \mathbf{(4.1)}$$

where s(t) is the transmitted signal, $h(\tau, t)$ is channel gain that affects the τ delayed sample of the **tth** signal.

There is scheme that uses a single transmit multiple receive antennas, namely SIMO which can be decomposed into M_R SISO channels. Channel gains and received signals can be written in Equation 4.2a and 4.2b respectively as:

$$\boldsymbol{h}(\tau, t) = \left[h_1(\tau, t)\ h_2(\tau, t) \ \ldots.\ \ h_{M_R}(\tau, t)\right]^T \qquad \mathbf{(4.2}\boldsymbol{a}\mathbf{)}$$

$$y_i = h_i(\tau, t) * s(t) \qquad i = 1,2, \ldots\ M_R \qquad \mathbf{(4.2}\boldsymbol{b}\mathbf{)}$$

where the signal at the i^{th} receive antenna is $y_i(t)$ and $h_i(\tau, t)$ is the channel impulse response between the transmit antenna and i^{th} receive antenna. Vectorial representation can be written in Equation 4.3 [22].

$$\boldsymbol{y}(t) = \boldsymbol{h}(\tau, t) * \boldsymbol{s}(t) \tag{4.3}$$

In a multiple input single output system, MISO, there is more than one transmit antennas and single receive antenna. This system is behaved as having M_T, number of transmitter antenna, SISO channels [22]. Channel impulse response can be written as in Equation 4.4.

$$\boldsymbol{h}(\tau, t) = \left[h_1(\tau, t)\ h_2(\tau, t) \ldots.\ \ h_{M_R}(\tau, t)\right]^T \tag{4.4}$$

The input output relationship for the MISO channel can be written as in Equation 4.5.

$$y(t) = \sum_{j=1}^{M_T} h_j(\tau, t) * s_j(t) \tag{4.5}$$

which may be expressed in vector notation in Equation 4.6 as:

$$\boldsymbol{y}(t) = \boldsymbol{h}(\tau, t) * \boldsymbol{s}(t) \tag{4.6}$$

where $\boldsymbol{s}(t) = [s_1(t)\ s_2(t) \ldots\ s_{M_T}(t)]^T$ is a $M_T x 1$ vector.

If there are more than one transmit and receive antennas, this scheme can be defined as MIMO.

Randomly time-varying channel is characterized by the $N_R x N_T$ matrix **H**(τ, t) defined as in following matrix :

$$\begin{bmatrix} h_{11}(\tau;t) & h_{12}(\tau;t) & \cdots & h_{1N_T}(\tau;t) \\ h_{21}(\tau;t) & h_{22}(\tau;t) & & h_{2N_T}(\tau;t) \\ \vdots & & \ddots & \vdots \\ h_{N_R1}(\tau;t) & h_{N_R2}(\tau;t) & \cdots & h_{N_RN_T}(\tau;t) \end{bmatrix}$$

It is supposed that the signal transmitted from the j^{th} transmit antenna is $s_j(t)$, j = 1,2,…. N_T. In the absence of noise the received signal at the i^{th} antenna can be written as in Equation 4.8

$$r_i(t) = \sum_{j=1}^{N_T} \int_{-\infty}^{\infty} h_{ij}(\tau;\ t)\, s_j\,(t-\tau)\mathrm{d}\tau$$

$$= \sum_{j=1}^{N_T} h_{ij}(\tau;\ t) * s_j(t-\tau) \quad i = 1,2, \dots N_R \qquad \mathbf{(4.7)}$$

4.1 Alamouti Space-Time Block Code

In this scheme, two different symbols s_1 and s_2 are transmitted simultaneously from antennas T_{x1} and T_{x2} respectively. During the first time instant s_1 is transmitted from antenna 1, s_2 is transmitted from antenna 2. During the second time instant $-s_2{}^*$ is transmitted from antenna 1, $s_1{}^*$ is transmitted from antenna 2. This scheme can be drawn as in Figure 4.3.

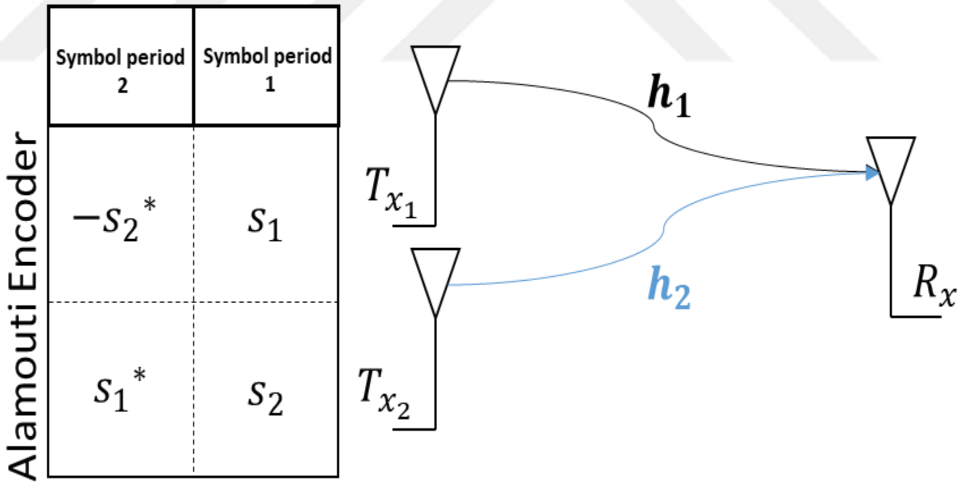

Figure 4.3 Alamouti scheme

y_1 and y_2 are the received signals during the first and second time periods, respectively, and are given by Equation 4.8a, 4.8b.

$$y_1 = \sqrt{\frac{\mathcal{E}_s}{2}} h_1 s_1 + \sqrt{\frac{\mathcal{E}_s}{2}} h_2 s_2 + n_1 \qquad \mathbf{(4.8}\boldsymbol{a}\mathbf{)}$$

$$y_2 = -\sqrt{\frac{\mathcal{E}_s}{2}} h_1 {s_2}^* + \sqrt{\frac{\mathcal{E}_s}{2}} h_2 {s_1}^* + n_2 \qquad \mathbf{(4.8b)}$$

where n_1 and n_2 are zero mean circulary symmetric complex Gaussian noise with $E\{|n_1|^2\} = E\{|n_2|^2\} = N_0$ and $^{E_s}/_2$ is the average transmitted energy per symbol period per antenna.

Received signal for the first and second time instants at a 2x1 system can be written as in Equation 4.10.

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \sqrt{\frac{E_s}{2}} \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \begin{bmatrix} s_1 \\ s_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ {n_2}^* \end{bmatrix} \qquad \mathbf{(4.9)}$$

Without depending on the existence of channel knowledge at the transmitter, Alamouti scheme provides a diversity order of 2 (maximum of transmitt antenna diversity) [22].

CHAPTER 5

SYSTEM DESCRIPTION

Diversity techniques designed for single carrier (SC) modulation over flat fading channels are easily extended to OFDM modulation with time index for the single carrier modulation being replaced by the tone index in OFDM [22].

Schematic of the Alamouti OFDM can be shown as in Figure 5.1.

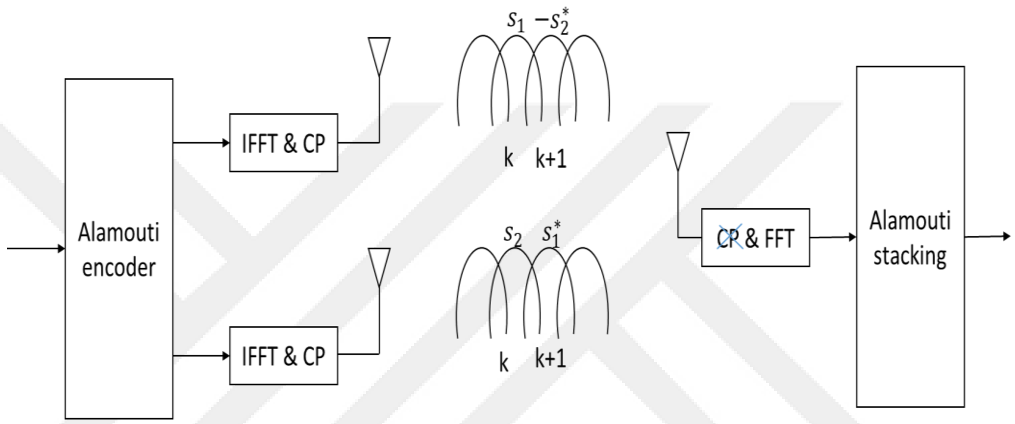

Figure 5.1 Alamouti with OFDM [22]

As shown in Figure 5.1, total number of symbols is divided into odd and even parts. The first odd symbol, namely first subcarrier, of the OFDM symbol is transmitted from first transmitter antenna and the first even symbol of the OFDM symbol is transmitted from second transmitter antenna at first time instant. Minus complex conjugate of the first even symbol and complex conjugate of the first odd symbol is transmitted from the first and the second antennas simultaneously at second time instant. During the transmission of two consecutive symbol periods channel is assumed to be constant namely quasi-static.

The same procedure is followed as in SISO- OFDM system for both first and second antennas at the transmitter. To get independently parallel sub-channels IFFT is applied. Size of the IFFT is the half of total symbol size of an OFDM symbol. CP is added after IFFT.

Different number of resolvable multipath schemes between two transmitters and one receiver antenna are worked. For instance, there is four paths each of the transmitter antennas to the receiver antenna.

During first time instant $h_{11}, h_{12}, h_{13}, h_{14}$ from first transmit antenna and $h_{21}, h_{22}, h_{23}, h_{24}$ from second transmit antenna are the channel gain coefficient taken into consideration. Total received signal for the first time instant from the first transmitter antenna to receiver antenna can be written as in Equation 5.1.

$$h_{11}, h_{12}, h_{13}, h_{14} = \lambda_{11} \qquad (\mathbf{5.1})$$

At the second time instant for the second transmitter antenna to receiver antenna when the same procedure is applied, it can be shown that:

$$h_{21}, h_{22}, h_{23}, h_{24} = \lambda_{21} \qquad (\mathbf{5.2})$$

When assumed that number of subcarrier is sixty four, we can write the following matrices for the first and second time instances respectively after the transmission of all subcarriers of the OFDM symbol.

$$[\lambda_{1,1}s_1 + \lambda_{2,1}s_2, \quad \lambda_{1,2}s_3 + \lambda_{2,2}s_4, \quad \ldots \quad \lambda_{1,32}s_{63} + \lambda_{2,32}s_{64}\ ]$$

$$[-\lambda_{1,1}{s_2}^* + \lambda_{2,1}{s_1}^*, \quad -\lambda_{1,2}{s_4}^* + \lambda_{2,2}{s_3}^*, \quad \ldots \quad -\lambda_{1,32}{s_{64}}^* + \lambda_{2,32}{s_{63}}^*\ ]$$

To get received symbols at the receive antenna independently, received signal is multiplied by pseudoinverse $(\boldsymbol{H}^H\boldsymbol{H})^{-1}\boldsymbol{H}^H$.

When the received signal is multiplied with the pesudoinverse, Equation 5.3 is written for the received symbol. $\boldsymbol{n}$ is zero mean, $N_0/2$ variance normally Gaussian distributed noise, but variance of the $\boldsymbol{n}$ change after multiplication by pseudoinverse.. To eliminate changing of the variance, received signal is multiplied by the Frobenius norm. Frobenius norm can be written as in Equation 5.5 for this design.

$$[(\boldsymbol{H}^H\boldsymbol{H})^{-1}\boldsymbol{H}^H]\boldsymbol{y} = [(\boldsymbol{H}^H\boldsymbol{H})^{-1}\boldsymbol{H}^H]\boldsymbol{H}\boldsymbol{x} + [(\boldsymbol{H}^H\boldsymbol{H})^{-1}\boldsymbol{H}^H]\boldsymbol{n} \quad (\mathbf{5.3})$$

$$\breve{\boldsymbol{n}} = (\boldsymbol{H}^H\boldsymbol{H})^{-1}\boldsymbol{n} \quad (\mathbf{5.4})$$

$$\frac{1}{\beta} = \|h_1\|^2 - \frac{\left|{h_1}^H h_2\right|^2}{\|h_2\|^2} = \left\|P_{h_2}^{\perp} h_1\right\|^2 \quad (\mathbf{5.5})$$

$$= {h_1}^H \left(I - \frac{{h_1}^H h_2}{\|h_2\|^2} \right)$$

where $h_1$is the channel gain coefficient for the first symbol after passing through two time instants and h_2 is the channel gain coefficient for the second symbol during two time instants. $P_{h_2}^{\perp}$ is the null space projection of the h_2 . Following matrix is observed in matrix notation after multiply with Equation 5.5.

$$\begin{bmatrix} \left\|P_{h_2}^{\perp} h_1\right\| x_1 + n_1 \\ \left\|P_{h_1}^{\perp} h_2\right\| x_1 + n_2 \end{bmatrix}$$

where n_1 and n_2 is the zero mean circularly symmetric AWGN is CN(0, $N_0/2$).

SSD is an interesting work that does not need extra BW and energy. Nevertheless this scheme increases diversity. In SSD constellation is rotated. The angle of the optimum rotation is found by using mathematical calculations. In-phase and quadrature components of one symbol is transmitted with independent symbols by using coordinate interleaving. For instance in-phase component of one symbol is transmitted with another symbol instead of its original in-phase or quadrature component.

5.1 Optimum Rotation Angle

When pairwise error probability is calculated, the optimum rotation angle for the BPSK modulation is found to be equal to 45^0 [17]. Normal BPSK modulation does not have quadrature component but quadrature component occurs with the rotation of the constellation. The constellations of the BPSK and rotated BPSK modulation is shown in Figure 5.1.

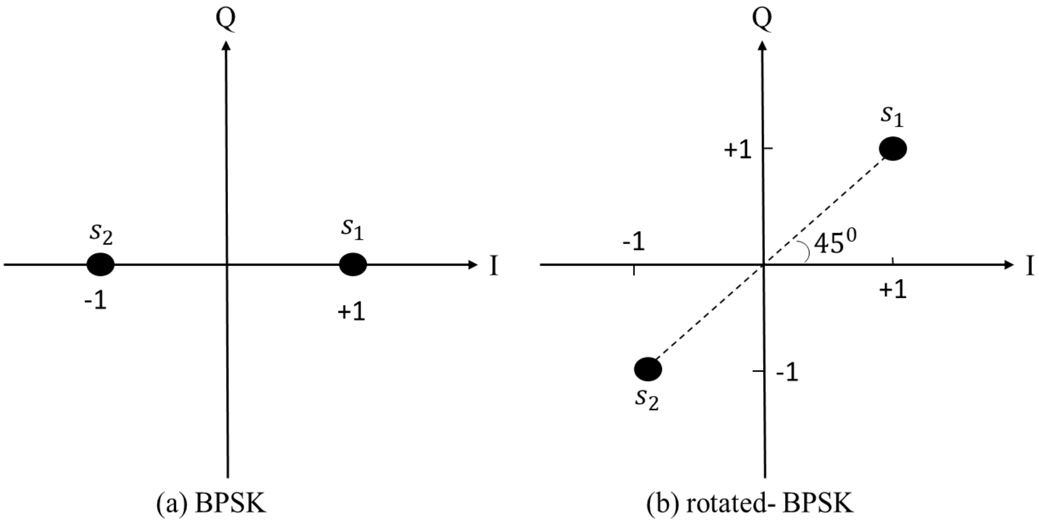

Figure 5.2 Different BPSK constellation scheme [17]

5.2 Combining SSD with OFDM

Normally SSD is applied to the flat fading channels. The channel that is simulated in this work that the channel is frequency selective. But the problem caused by frequency selective fading channel is overcome by using OFDM. OFDM converts frequency selective fading channels into flat fading channels, following applied SSD to the desired system.

Firstly squared absolute correlation coefficient (SACC) between subcarriers of each OFDM symbol can be calculated using Equation 5.1 [23].

$$\rho = |E\{H_k H_m{}^*\}|^2 = \frac{1}{L^2}\frac{\sin^2(\frac{L}{N}(k-m)\pi)}{\sin^2(\frac{1}{N}(k-m)\pi)} \quad \mathbf{(5.6)}$$

where N is the total number of subcarriers for one OFDM symbol, L is the multipath delay spread of the channel and (k-m) is the length between subcarriers.

As shown in Equation 5.6 when $\frac{L}{N}(k-m)$ is integer, SACC is should be zero. Hence subcarriers are uncorrelated.

CHAPTER 6

RESULTS and DISCUSSION

Combination of orthogonal frequency division multiplexing, signal space diversity and Alamouti space time block code is statistically investigated, simulated and verified. In the first part statistical calculation is given and simulation is different cases are verified using MATLAB.

6.1 Statistical Calculation

Decision boundary which is shown with blue line in Figure 6.1 is used to give in favor of which symbol is transmitted at the receiver. Maximum Likelihood (ML) detection is used. In ML detection distance of the received signal and all constellation points are measured after that which distance is minimum, detector decides in favor of this symbol. When the receiver makes wrong decision for instance symbol s_1 is transmitted, and the receiver decides in a favor of a symbol s_2, namely decides lower sides of the blue line in Figure 6.1. Firstly pairwise error probability (PEP) can be calculated using ML detection as Equation 6.1.

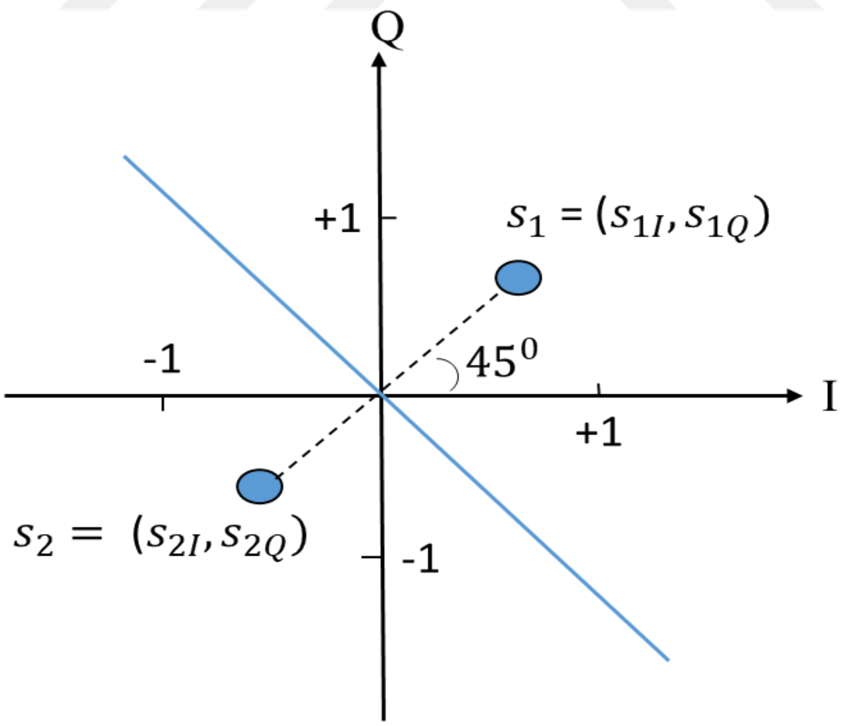

Figure 6.1 Decision boundary

$$P(e) = P(s_1 \to s_2) = P(|d_1 - s_1|^2 > |d_2 - s_1|^2) \qquad \mathbf{(6.1)}$$
$$= P\left(2\sigma_1\Delta_I n_I + 2\sigma_2\Delta_Q n_Q < -{\sigma_1}^2{\Delta_I}^2 - {\sigma_2}^2{\Delta_Q}^2\right)$$

where $\Delta_I = s_{1I} - s_{2I}$, $\Delta_Q = s_{1Q} - s_{2Q}$ are the in-phase and quadrature distances between s_1 and s_2 components of the BPSK constellation, respectively. Error function is calculated for the given channel gain coefficients.

$$P(e|\sigma) = Q\left(\sqrt{\frac{{\sigma_1}^2{\Delta_I}^2 + {\sigma_2}^2{\Delta_Q}^2}{2N_0}}\right) \quad \mathbf{(6.2)}$$

by Applying Craig's formula

$$P(e|\sigma) = \frac{1}{\pi}\int_0^{\pi/2} \exp\left(-\frac{{\sigma_1}^2{\Delta_I}^2 + {\sigma_2}^2{\Delta_Q}^2}{4N_0 sin^2\emptyset}\right) d\emptyset \quad \mathbf{(6.3)}$$

σ_1 and σ_2 are exponentially distributed,

$$P(e) = E[P(e|\sigma_1, \sigma_2)]$$
$$= \frac{1}{\pi}\int_0^{\pi/2}\int_0^{\infty}\int_0^{\infty} \exp\left(-\frac{\varepsilon_b}{2N_0}\frac{{\sigma_1}^2 + {\sigma_2}^2}{sin^2\emptyset}\right)\sigma_1\sigma_2 \exp(-(\sigma_1$$
$$+ \sigma_2))d\sigma_1 d\sigma_2 d\emptyset$$
$$= \frac{1}{\pi}\int_0^{\pi/2} \frac{16}{\left(2 + SNR\,\frac{1}{sin^2(\emptyset)}\right)^4} \quad \mathbf{(6.4)}$$

where $\frac{\varepsilon_b}{N_0} = SNR$.

To solve Equation 6.4 we use numerical integration in Wolfram Mathematica computer program, thus, we obtain numerical results.

6.2 Simulation

Two different cases is simulated for the number of subcarrier, *N* and number of resolvable multipath, *L* is relatively prime or not with respect to Equation 5.6.

Different CI strategies with respect to relationship between *L* and *N* relatively prime or not are provided. In the first case N is relatively prime. Figure 6.2 shows a relatively prime case. There is not any (*k*-*m*) subcarrier interleaving strategy that makes Equation 5.6 is zero, thus we found an optimum subcarrier interleaving strategy that minimizes

SACC. Two interleaving strategies and also no interleaving strategy are obtained in Figure 6.2.

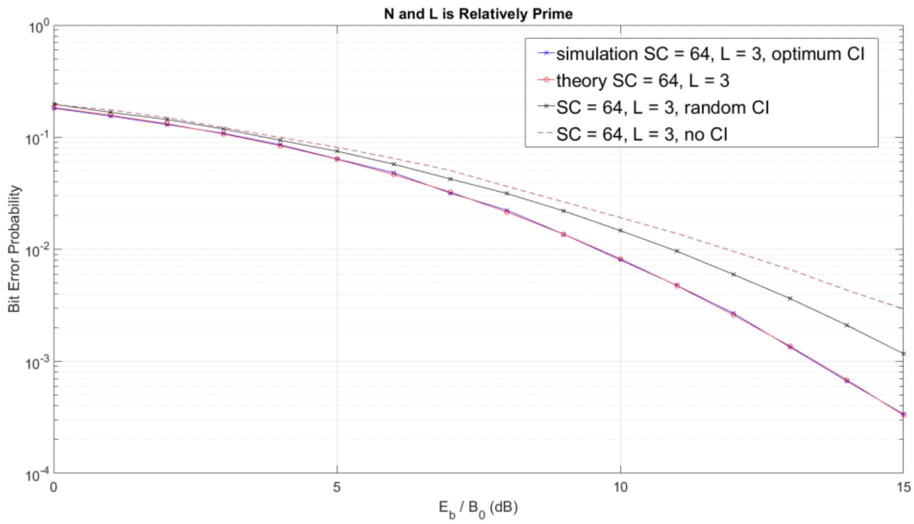

Figure 6.2 N and L is relatively prime

Number of subcarrier, N and number of resolvable multipath, L is 64 and 3 respectively. No interleaving strategy shows the worst performance when compared random CI and optimum CI strategy. The performance analysis is made at the 10^{-2} bit error probability. There is 1 dB better performance is observed at random CI strategy compared with no CI. Optimum CI strategy shows approximately 3 dB better performance than no CI strategy. Random CI strategy as expected is shown worse performance than optimum CI strategy which is approximately 2 dB.

When number of subcarrier N and number of resolvable multipath, L is not relatively prime, there is different number subcarrier interleaving (k-m) values to make SACC is zero. Number of subcarrier, N is 64 and 256 and number of resolvable multipath is 4 and 8 are examined with half of the N interleaving strategy. Theoritical and simulation results match for different N and L values. For the indicated case N = 256, L = 8, there is three possible (k-m) = {16, 32, 64, 128} that makes zero SACC value and also there is 3 dB compared to the no CI strategy.

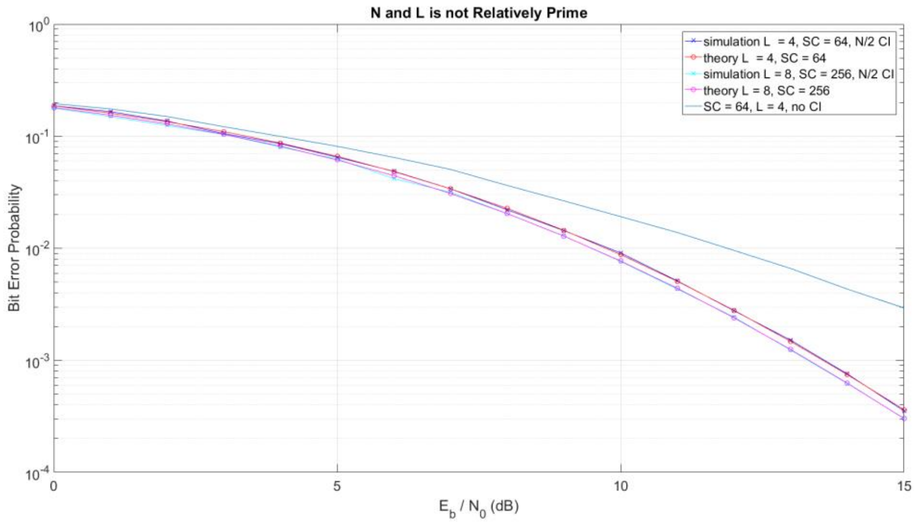

Figure 6.3 N and L not relatively prime

6.3 Physical Layer Security

This interleaving strategy can also be used for physical layer security. Illustration of this scheme is given in Figure 5.5. Consider that Alice and Bob wants speak safely but Eve listens their conservation. In the worst case scenario Eve knows they are using interleaving and number of taps (L = 4) for this conservation. There are 3 alternatives to recover conservation. The interleaving strategy imposes that the minimized correlation between the subcarriers can be 16, 32, 48 for L = 4 and Equation 5.2 should be zero. If Eve makes a wrong decision about interleaving strategy, she will receive 1/3 of the all bits transmitted incorrect. $\log_2 3$ bits can be used for feedback. Inside the logarithm, 3 is the alternative strategy number which gives feedback from the receiver to the transmitter at different time instants to create dynamic security for the conservation.

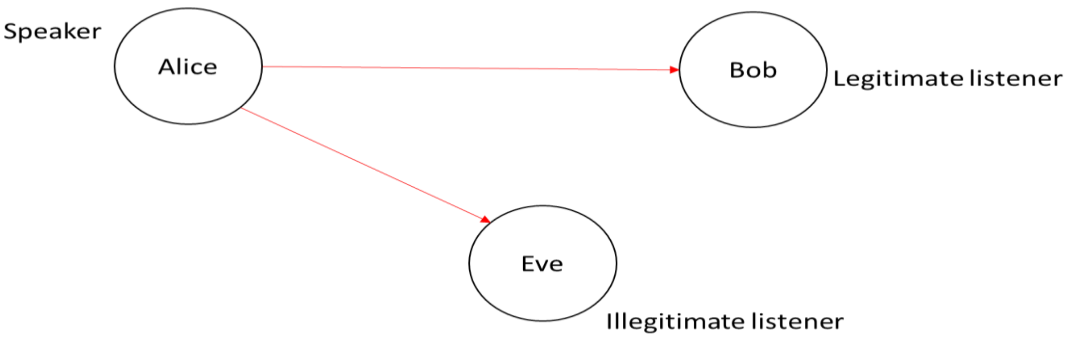

Figure 6.4 Conversation scenario

CHAPTER 7

CONCLUSION

In this work signal space diversity (SSD) is studied theoretically and in simulation by applying Alamouti space time block coding (STBC) with orthogonal frequency division multiplexing (OFDM). OFDM is a popular multiplexing technique. This technique allows use of bandwidth effectively by applying overlapping orthogonal sub-channels. The increase in the bandwidth of transmitted signals and the commonly encountered communication environment of frequency selective fading channels make people need to use narrowband multicarrier channels. OFDM is one of the best choice for this application. Alamouti is a transmit and/or receive diversity technique. Alamouti STBC scheme can also be used in frequency selective fading channels with the help of OFDM. SSD is used in this work and its advantages in physical layer security is shown.

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