CHANNEL ADAPTIVE USER COOPERATION STRATEGIES FOR FADING WIRELESS CHANNELS

A DISSERTATION
SUBMITTED TO THE DEPARTMENT OF ELECTRONICS ENGINEERING
AND THE COMMITTEE ON GRADUATE STUDIES
OF IŞIK UNIVERSITY
IN PARTIAL FULFILLMENT OF THE REQUIREMENTS
FOR THE DEGREE
OF
MASTER OF SCIENCE

by
ÇAĞATAY EDEMEN

JUNE 2006

CHANNEL ADAPTIVE USER COOPERATIVE STRATEGIES
FOR WIRELESS FADING CHANNELS

Approved for University Committee on Graduate Studies:

I certify that I have read this dissertation and that, in my opinion, it is fully adequate in scope and quality, as a dissertation for the degree of Master of Science.

Prof. Dr. Hüsnü A. ERBAY
(Director)
School of Electronics Engineering, *Işık University*

I certify that I have read this dissertation and that, in my opinion, it is fully adequate in scope and quality, as a dissertation for the degree of Master of Science.

Prof. Dr. Ahmet AKSEN
(Head of Electronics Engineering Department)
School of Electronics Engineering, *Işık University*

I certify that I have read this dissertation and that, in my opinion, it is fully adequate in scope and quality, as a dissertation for the degree of Master of Science.

Asst. Prof. Dr. Onur Kaya
(Supervisor)
School of Electronics Engineering, *Işık University*

Examining Committee Members

…This thesis is dedicated to my mother

Thank you for everything …

ABSTRACT

CHANNEL ADAPTIVE USER COOPERATION STRATEGIES FOR FADING WIRELESS CHANNELS

EDEMEN, ÇAĞATAY

In this dissertation, a channel adaptive method is proposed for a two user half-duplex cooperative communication system under the assumption of perfect channel state information at the transmitters and the receiver. An achievable rate region for the proposed method is characterized, and evaluated by computer simulation.

ÖZET

KABLOSUZ KAYIPLI KANALLAR İÇİN KANAL UYARLAMALI KULLANICI YARDIMLAŞMA STRATEJİLERİ

EDEMEN, ÇAĞATAY

Bu çalışmada, yollayıcı ve alıcı tarafının mükemmel kanal bilgilerine sahip olduğu varsayımı altında, iki kullanıcılı tek yönlü yardımlaşmalı haberleşme sistemi için kanal uyarlamalı bir sonuç önerilmiştir. Bu önerilen model için, ulaşılabilecek kapasite alanları karakterize edilmiş ve bilgisayar simülasyonu ile değerlendirilmiştir.

ACKNOWLEDGMENTS

There are a lot of people I wish to thank for while studying as a graduate student at Işık University. First of all, I would like to thank my mother for her financial support from the beginning. Her patience, generosity and self-sacrifice were unbelievable over the past three years. I am also grateful to my supervisor, Asst. Prof. Onur KAYA, for continual encouragement at every step of my dissertation over the past year. He was an immense help during my dissertation by sharing his experience about my subject continuously. Last but not least, I would like to thank my schoolmates, and the Işık University Institute of Science and Engineering for their continual support.

CONTENTS

LIST OF TABLES

LIST OF FIGURES

LIST OF SYMBOLS/ABBREVIATIONS

B_d	Doppler Frequency
$f_{doppler}$	Doppler Effect
$f_{\max}$	Angle of Arrival Wave
$\alpha_{doppler}$	Coherence Bandwidth
$(\Delta f)_c$	Bandwidth
W	Multipath Spread
T_m	Mutual Information
I_{mutual}	Expected value
$\mathrm{E}(\bullet)$	Minimum Value
$\min(\bullet)$	Maximum Value
$\max(\bullet)$	Vehicle Speed
v	Speed of Light
c_0	

AWGN	Additive White Gaussian Noise
BSC	Binary Symmetric Channel
CDMA	Code Division Multiple Access
CSI	Channel State Information
DCS	Digital Cellular System
DMC	Discrete Memoryless Channel
FDMA	Frequency Division Multiple Access
FPLMTS	Future Public Land Mobile Telecommunications System
GSM	Global Special Mobile
ISI	Intersymbol Interference
IMT 2000	International Mobile Telecommunications 2000
MIMO	Multiple-input Multiple-output
MAC	Multiple Access Channel
PCS	Personal Communication System
TDMA	Time Division Multiple Access

UMTS	Universal Mobile Telecommunications System
VAS	Value Added Services

Chapter 1

Introduction

For several years the mobile communication systems have been improving with fastest growth. In addition, wireless communications can be defined as one of the most active areas of technology development of our time. The globalization in the telecommunications sector and increased demand for high frequency ranges cause development in coding and diversity techniques.

The mobile wireless channels suffer from fading, the effects of which can be overcome by adjusting the transmit strategies. A common way of combating fading is the use of diversity techniques. The most popular diversity techniques are time, frequency and space diversity. The space diversity generally requires more than one antennas. For this reason, the implementation of the space diversity in wireless communication is limited by size of mobile devices. Additionally, it is advantageous on a cellular base station, it may not be practical for other scenarios. However, the cooperative diversity enables single antenna mobile devices in a multi user environment to share their antennas and generate a virtual multiple antenna transmitter.

The main challenges in future mobile communication system are to increase spectral and power efficiency. The cooperative networks will play greater role in providing spectral efficiency in future mobile communication systems. For this reason, a significant amount of research has recently been done in this area have [2], [7]. In this dissertation we propose an adaptive transmit strategy for half duplex cooperative communication. The objective of this dissertation is to calculate the achievable rates for the half duplex scheme based on our proposed model. This dissertation provides suggestions for improvement of achievable rate in half duplex scheme.

1.1 Outline of the Dissertation

This dissertation is organized as follows. In Chapter 2, we introduce the related background materials about wireless communication technology. In Chapter 3, we present the capacity results for the Discrete Memoryless Channel, Multiple Access Channel and Relay Channel. Besides, the cooperative communication is explained in this chapter. In Chapter 4, our proposed model for half-duplex cooperation scheme is introduced. In the last section of this chapter, we show our numerical results for our proposed two user half duplex cooperative communication and compare them with those for full duplex cooperation. In Chapter 5, we present conclusions and future work.

Chapter 2

Related Background Knowledge

This chapter consists of the related knowledge that is needed to explain cooperation communication. We also present some important properties of wireless communication system we introduce wireless channels in section 2.1, give a mathematical characterization of the wireless channels in section 2.2, present some common diversity techniques in section 2.3, and review some basic channel models in section 2.4.

2.1. Wireless Channels

The communication channel provides the connection between the transmitter and the receiver side. The physical channel may be a pair of wires that carry the electrical signal, or an optical fiber that carries the information on a modulated light beam, or an underwater ocean channel in which the information is transmitted acoustically, or free space over which the information-bearing signal is radiated by use of an antenna. One common problem in signal transmission though any channel is additive noise. In general, additive noise is generated internally by a component such as a resistor or solid-state devices used to implement the communication system. This is sometimes called the *thermal noise*. Other sources of noise and interference may arise externally, such as interference from other users of channel. When such noise and interference occupy the same frequency band as the desired signal, their effect can be minimized by the proper design of the transmitted signal and its demodulator at the receiver. Other types of the signal degradation that may be encountered in transmission over the channel are signal attenuation, amplitude and phase distortion, and multipath distortion [17]. These effects are illustrated in Figure 2.1

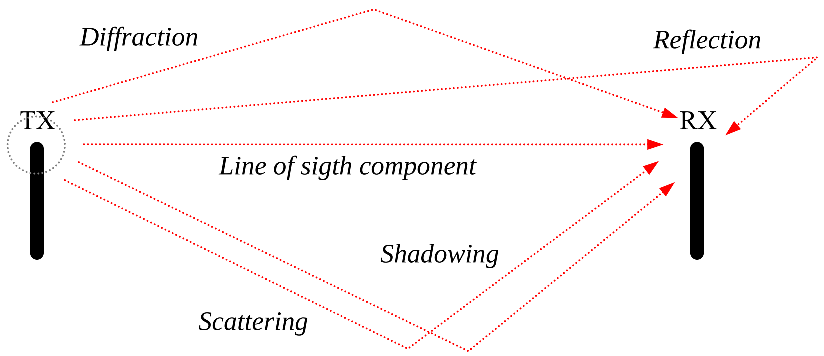

Figure 2.1: Illustration of multipath effects in wireless communication.

Propagation losses are also an issue in wireless channels. These are of two basic types: *diffusive losses* and *shadow fading*. Diffusive losses arise because of the open nature of wireless channels. For example, the energy radiated by a simple point source in free space will spread over an ever-expanding spherical surface as the energy propagates away from the source. This means that an antenna with a given aperture size will collect an amount of energy that decreases with the square of the distance between the antenna and the source. In most terrestrial wireless channels, the diffusion losses are actually greater that this, due to the effects of ground-wave propagation, foilage, etc. For example, in cellular telephony, the diffusion loss is inverse-square with distance within line-of-sight of the cell tower, and it falls off with a higher power (typically 3 or 4) at greater distances. As its name implies, shadow fading results from the presence of objects (buildings, walls, etc.) between transmitter and receiver. Shadow fading is typically modeled by attenuation (i.e., a multiplicative factor) in signal amplitude that follows a log-normal distribution. The variation in this fading is specified by the standard deviation of the logarithm of this attenuation.

Multipath refers to the phenomenon in which multiple copies of a transmitted signal are received at the receiver due to the presence of multiple radio paths between the transmitter and receiver. These multiple paths arise due to reflections from objects in the radio channel. Multipath is manifested in several ways in communications receivers, depending on the degree of path difference relative to the wavelength of propagation, the degree of path difference relative to the signaling rate, and the relative motion between the transmitter and receiver. Multipath from scatterers that are spaced very close together will cause a random change in the amplitude of the received signal.

Due to central-limit type effects, the resulting received amplitude is often modeled as being a complex Gaussian random variable. This results in random amplitude whose envelope has a Rayleigh distribution, and this phenomenon is thus termed *Rayleigh fading*. When the scatterers are spaced so that the differences in their corresponding path lengths are significant relative to a wavelength of the carrier, then the signals arriving at the receiver along different paths can add constructively or destructively. This gives rise to fading that depends on the wavelength (or, equivalently, the frequency) of radiation, which is thus called *frequency-selective fading*. When there is relative motion between the transmitter and receiver, this type of fading also depends on time, since the path length is a function of the radio geometry. This results in *time-selective fading*. (Such motion also causes signal distortion due to Doppler effects.) A related phenomenon arises when the difference in path lengths is such that the time delay of arrival along different paths is significant relative to a symbol interval. This results in dispersion of the transmitted signal, and causes *intersymbol interference* (ISI); i.e., contributions from multiple symbols arrive at the receiver at the same time [23].

The effects of noise may be minimized by increasing the power in transmitted signal. However, equipment and other practical constraints limit the power level in the transmitted signal. Another basic limitation is the available channel bandwidth. A bandwidth constraint is usually due to the physical limitation of the medium and the electronic components used to implement the transmitter and receiver [17].

Besides the multipath propagation effect, the Doppler effect has negative influence on the transmitted signal. The Doppler effect occurs due to movement of the mobile unit and causes frequency shift of the communication waves. Corresponding to this, the Doppler effect causes a frequency expansion during transmission. The Doppler effect can be expressed as (2.1)

$$f_{doppler} = f_{\max} \cos \alpha_{doppler} \tag{2.1}$$

where $f_{\max}$ is maximum Doppler frequency and related to speed of the mobile. $\alpha_{doppler}$ is an angle of arrival wave towards the mobile unit.

$$f_{\max} = {}^{v}\!/_{c_0}\, f_0 \tag{2.2}$$

In (2.2) v denotes speed of mobile unit, c_0 denotes speed of light and f_0 denotes carrier frequency. The Doppler effect reaches the maximum value in case of $\alpha_{doppler} = 0$, and minimum value in case of $\alpha_{doppler} = \pi$.

2.2. Characterization of Wireless Channels

In addition to these effects which are explained in previously in Section 2.1, the statistical characterization of multipath communication channel should be developed to determine effects of propagation illustrated in Figure 2.2.

Let us develop a number of useful correlation functions and power spectral density functions that define the characteristics of a fading multipath channel. The starting point of the development of the characteristics, is determining a low-pass impulse response $c(t;\tau)$, which is characterized as a complex-valued random process in the t variable. If $c(t;\tau)$ is assumed to be wide-sense-stationary. then the autocorrelation function of $c(t;\tau)$ can be defined as

$$\phi_c(\tau_1,\tau_2;\Delta t) = \frac{1}{2}\mathrm{E}\left[c^*(\tau_1;t)\,c(\tau_2;t+\Delta t)\right] \tag{2.3}$$

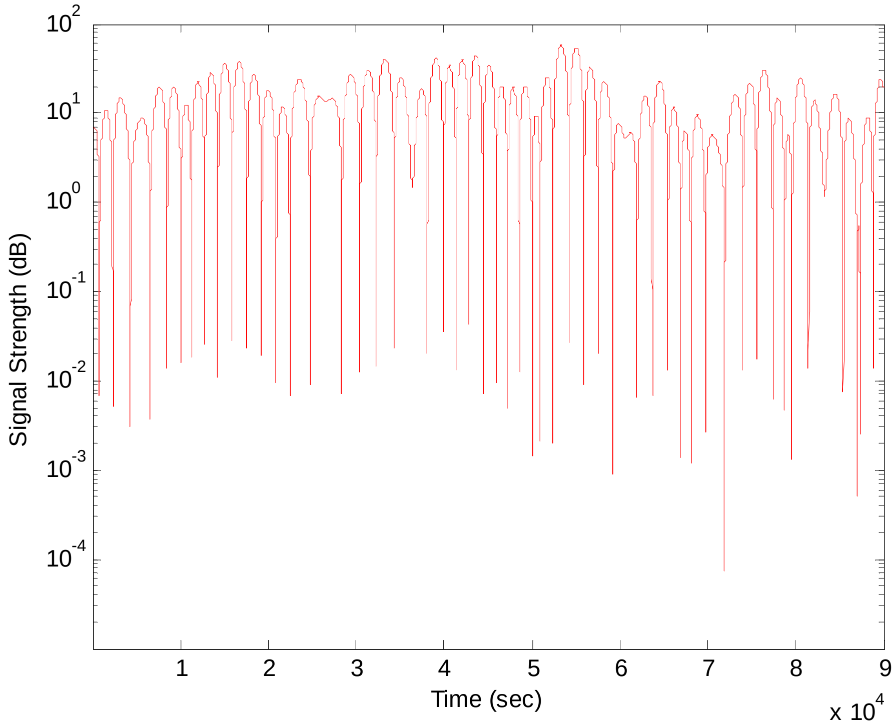

Figure 2.2: Illustration of received signal in multipath communication channel.

Generally, the attenuation and phase shift of the channel associated with path delay τ_1 is uncorrelated with the attenuation and phase shift associated with path delay τ_2 in the communication. This case is called as *uncorrelated-scattering*. From this assumption which are scattering at two different delay is uncorrelated and incorporate it into (2.3) the following equation can be obtained [17].

$$\frac{1}{2}\mathrm{E}\left[c^*(\tau_1;t)c(\tau_2;t+\Delta t)\right]=\phi_c(\tau_1;\Delta t)\delta(\tau_1-\tau_2) \tag{2.4}$$

If $\Delta t = 0$, the autocorrelation can be defined as $\phi_c(\tau;0)\equiv\phi_c(\tau)$ and a function of time delay τ. Beside this, $\phi_c(\tau;\Delta_t)$ is called as the *multipath intensity profile* or the *delay power spectrum* of the channel.

To characterize the channel, there is need to pass to the frequency domain by taking Fourier transform of $c(\tau;t)$. In case of taking Fourier transform of

$c(\tau;t)$, the time variant transfer function $C(f;t)$, is obtained by using following equation.

$$C(f;t) = \int_{-\infty}^{\infty} c(\tau;t) e^{-j2\pi f\tau} d\tau \tag{2.5}$$

Since $c(\tau;t)$ is complex-valued zero-mean Gaussian random variable, the transfer function $C(f;t)$ has the same statistical property. Under this consideration, the autocorrelation function can be expressed as

$$\phi_c(f_1, f_1;\Delta t) = \frac{1}{2}\mathrm{E}\left[C^*(f_1;t)C(f_2;t+\Delta t)\right] \tag{2.6}$$

Since $C(f;t)$ is Fourier transform of $c(\tau;t)$, it is easy to establish a relationship between $\phi_c(f_1, f_1;\Delta t)$ and $\phi_c(\tau_1, \tau_1;\Delta t)$ by using (2.5). Thus,

$$\begin{aligned}
\phi_c(f_1, f_1;\Delta t) &= \frac{1}{2}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \mathrm{E}\left[c^*(\tau_1;t)c(\tau_2;t+\Delta t)\right] e^{j2\pi(f_1\tau_1 - f_2\tau_2)} d\tau_1 d\tau_2 \\
&= \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \phi_c(\tau_1;\Delta t)\delta(\tau_1 - \tau_2) e^{j2\pi(f_1\tau_1 - f_2\tau_2)} d\tau_1 d\tau_2 \\
&= \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} \phi_c(\tau_1;\Delta t) e^{j2\pi(f_1\tau_1 - f_2\tau_2)} d\tau_1 d\tau_2 \\
&= \int_{-\infty}^{\infty} \phi_c(\tau_1;\Delta t) e^{j2\pi(f_1 - f_2)\tau_1} d\tau_1 \\
&= \int_{-\infty}^{\infty} \phi_c(\tau_1;\Delta t) e^{j2\pi(\Delta f)\tau_1} d\tau_1 \\
&\equiv \phi_C(\Delta f;\Delta t)
\end{aligned} \tag{2.7}$$

where $\Delta f = f_1 - f_2$. The function $\phi_C(\Delta f;\Delta t)$ is closely related to frequency difference Δf. For this reason, it is called as the *spread-frequency, space time correlation function of the channel*.

In case of choosing $\Delta t = 0$ in (2.7), $\phi_C(\Delta f;\Delta t)$ and $\phi_c(\tau;\Delta_t)$ can be determined as $\phi_C(\Delta f;0) \equiv \phi_C(\Delta f)$ and $\phi_c(\tau;0) \equiv \phi_c(\tau)$ with the following relationship

$$\phi_C(\Delta f) = \int_{\infty}^{\infty} \phi_c(\tau) e^{-j2\pi\Delta f \tau} d\tau \tag{2.8}$$

where $\phi_C(\Delta f)$ is a metric of frequency coherence of the channel. From (2.8), the relationship between Δf and τ can be approximately denoted as

$$(\Delta f)_{coherence} \approx 1/T_m \tag{2.9}$$

where $(\Delta f)_{coherence}$ is the coherence bandwidth. Depending on the value of $(\Delta f)_{coherence}$, the channel can be divided into two categories. When $(\Delta f)_{coherence} < bandwidth$, the channel is called as *frequency-selective*. In this case, the signal is distorted by the channel. In the second case when $(\Delta f)_{coherence} > bandwidth$, the channel is denoted as *frequency-nonselective*.

In addition to Δf in $\phi_C(\Delta f;\Delta t)$, Δt is the time variations of the channel which is related to Doppler effect. To determine this relation ship between Δt and Doppler effects, it is needed to take the Fourier transform of $\phi_C(\Delta f;\Delta t)$ with respect to Δt

$$S_C(\Delta f;\lambda) = \int_{-\infty}^{\infty} \phi_C(\Delta f;\Delta t) e^{-j2\pi\lambda\Delta t} d\Delta t \tag{2.10}$$

If we chose $\Delta f = 0$, the equation (2.10) is expressed as

$$S_C(\lambda) = \int_{-\infty}^{\infty} \phi_C(0;\Delta t) e^{-j2\pi\lambda\Delta t} d\Delta t \tag{2.11}$$

where $S_C(\lambda)$ is a power spectrum function which determines the signal intensity as a function of Doppler frequency λ. Therefore, $S_C(\lambda)$ is called as *Doppler spectrum* of the channel.

The value range of Doppler frequency λ is called as *Doppler spread* B_d of the channel. And the relationship between B_d and τ can be approximately denoted as

$$(\Delta t)_{coherence} \approx 1/B_d \tag{2.12}$$

where $(\Delta t)_{coherence}$ is the coherence time. A channel which is slowly changing has large coherence time and for this reason has small Doppler spread.

If the symbol duration is smaller than $(\Delta t)_{coherence}$, then the channel is classified as slow fading. Slow fading channels are very often modeled as time-invariant channels over a number of symbol intervals. On the other hand, if it is smaller than the symbol duration, the channel is considered to be fast fading. In general, it is difficult to estimate the channel parameters in a fast fading channel.

Due to explanations in above, a fading channel can be classified into four different types which is illustrated in Table 2.1. Let's develop a relationship between $\phi_C(\Delta f;\Delta t)$ and $\phi_c(t;\Delta t)$ and a relationship between $\phi_C(\Delta f;\Delta t)$ and $S_C(\Delta f;\lambda)$. These relationships can be obtained by taking Fourier transform with respect to Δf and Δt. Thus,

$$S(\tau;\lambda)=\int_{-\infty}^{\infty}\phi_c(\tau;\Delta t)e^{-j2\pi\lambda\Delta t}d\Delta t \tag{2.13}$$

where $S(\tau;\lambda)$ is a function of Fourier transform of $\phi_c(t;\Delta t)$ with respect to Δt, and

$$S(\tau;\lambda)=\int_{-\infty}^{\infty}S_C(\Delta f;\lambda)e^{j2\pi\tau\Delta f}d\Delta f \tag{2.14}$$

Beside this, $S(\tau;\lambda)$ and $\phi_C(\Delta f;\Delta t)$ are related by double Fourier transform as

$$S(\tau;\lambda)=\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\phi_c(\Delta f;\Delta t)e^{-j2\pi\lambda\Delta t}e^{j2\pi\tau\Delta f}d\Delta t d\Delta f \tag{2.15}$$

This new function is called as the *scattering function of the channel*. It is a metric of the average power output of the channel.

	$\left(\Delta f\right)_{coherence} < bandwidth$	$\left(\Delta f\right)_{coherence} > bandwidth$
$\left(\Delta t\right)_{coherence} < symbol\ duration$	Frequency selective fast fading	Frequency non-selective fast fading
$\left(\Delta t\right)_{coherence} > symbol\ duration$	Frequency selective slow fading	Frequency non-selective slow fading

Table 2.1: Classifications of fading channel.

2.3. Diversity Techniques for Wireless Channels

Diversity techniques are based on the notion that errors occur in reception when the channel attenuation is large, i.e., when the channel is in a deep fade. If we can supply to the receiver several replicas of the same information signal transmitted over independently fading channels, the probability that all the signal components will fade simultaneously is reduced considerably. That is, if p is the probability that any one signal will fade below some critical value, then p^L is the probability that all L independently fading replicas of the same signal will fade below the critical value. There are several ways in which we can provide the receiver with L independently fading replicas of the same information-bearing signal.

One method is to employ *frequency diversity*. That is, the same information bearing signal is transmitted on L carriers, where the separation between successive carriers equals or exceeds the coherence bandwidth $\left(\Delta f\right)_c$ of the channel

A second method for achieving L independently fading versions of the same information-bearing signal is to transmit the signal in L different time slots, where the separation between successive time slots equals or exceeds the coherence time $\left(\Delta t\right)_c$ of the channel. This method is called *time diversity*.

Note that the fading channel fits the model of a bursty error channel. Furthermore, we may view the transmission of the same information either at different frequencies or in difference time slots (or both) as a simple form of repetition coding. The separation of the diversity transmissions in time by $(\Delta t)_c$ or in frequency by $(\Delta f)_c$ is basically a form of block-interleaving the bits in the repetition code in an attempt to break up the error bursts and, thus, to obtain independent errors.

Another commonly used method for achieving diversity employs multiple antennas. For example, we may employ a single transmitting antenna and multiple receiving antennas. The latter must be spaced sufficiently far apart that the multipath components in the signal have significantly different propagation delays at the antennas. Usually a separation of a few wavelengths is required between two antennas in order to obtain signals that fade independently.

A more sophisticated method for obtaining diversity is based on the use of a signal having a bandwidth much greater than the coherence bandwidth $(\Delta f)_c$ of the channel. Such a signal with bandwidth W will resolve the multipath components and, thus, provide the receiver with several independently fading signal paths. The time resolution is $1/W$. Consequently, with a multipath spread T_m seconds, there are $T_m W$ resolvable signal components. Since $T_m \approx 1/(\Delta f)_c$ the number of resolvable signal components may also be expressed as $W/(\Delta f)_c$. Thus, the use of a wideband signal may be viewed as just another method for obtaining frequency diversity of order $L \approx W/(\Delta f)_c$ [17].

The following sections consist of details of common diversity techniques.

2.3.1 Frequency Diversity

Frequency diversity transmits information on more than one carrier frequency. Each carrier should be separated from the others by at least the coherence bandwidth. Theoretically, if the channels are uncorrelated, the probability of simultaneous fading will be the product of the individual fading probabilities

Frequency diversity is often employed in microwave line-of-sight links which carry several channels in a frequency division multiplex mode (FDM). Due to tropospheric propagation and resulting refraction, deep fading sometimes occurs. In practice, $1/N$ *protection switching* is provided by a radio licensee, where in one frequency is nominally idle but is available on a stand-by basis to provide frequency diversity switching for any one of the N other carrier (frequencies) being used on the same link, each carrying independent traffic. When diversity is needed, the appropriate traffic is simply switched to the backup frequency. This technique has the disadvantage that it not only requires spare bandwidth but also requires that there be as many receivers as there are channels used for the frequency diversity. However, for critical traffic, the expense may be justified [13]. At the receiver side, the L independently faded signals are "optimally" combined to give a statistic for decision.

2.3.2 Time Diversity

Time diversity is used in cases where the transmissions channel suffers from the Doppler effect. The signals are repeated in different time slot. Therefore the signals are received with independent fading conditions. In time diversity, the same data is transmitted multiple times. The main advantage of time diversity is flexible form thus it can be modified for different channel. However, the disadvantage is long delays.

2.3.3 Space Diversity

Space diversity, also known as *antenna diversity*, is one of the most popular forms of diversity used in wireless systems. Conventional cellular radio systems consist of an elevated base station antenna and a mobile antenna close to the ground. The existence of a direct path between the transmitter and the receiver is not guaranteed and the possibility of a number of scatterers in the vicinity of the mobile suggests a Rayleigh fading signal.

The concept of antenna space diversity is also used in base station design. At each cell site, multiple base station receiving antennas are used to provide diversity reception. However, since the important scatterers are generally on the ground in the vicinity of the mobile, the base station antennas must be spaced considerably far apart to achieve decorrelation. Separations on the order of several tens of wavelengths are required at the base station [13]. Space separation of *half of the wavelength* is sufficient to obtain two uncorrelated signals. Space diversity can be used at either the mobile or base station, or both.

2.4. Multi-access Techniques

In wireless communication, the multiple user schemes are used to allow many mobile user to share their information simultaneously in a limited available bandwidth. There are lots of types for sharing information. However the common types are multiple access channel and broadcast network. One of them is multiple access channel in which many user share a common communication channel to send their information towards a destination as illustrated in Figure 2.3. The other common type is broadcast network in which a transmitter sends information towards many receivers. Separately from multiple access and broadcast channel, the relay channel where there is one source and receiver but one or more intermediate sender-receiver as acted relay, is other way to share information. In these types of multiuser communication system, the users can use different methods to send their information through the communication channel. These methods are detailed in the next sections.

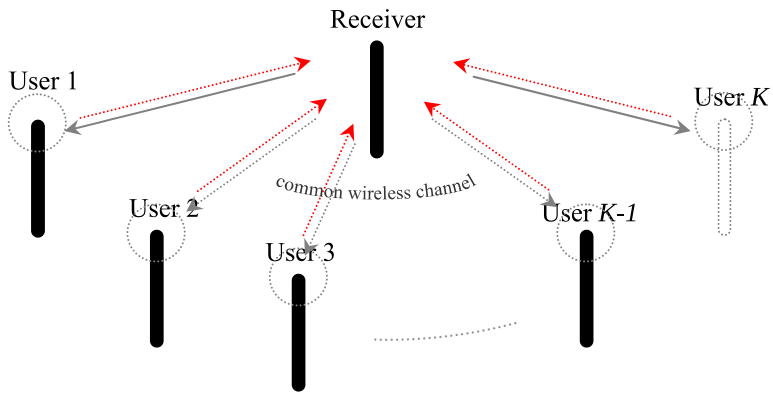

Figure 2.3: Illustration of multiple access and broadcast communication system the red lines denote multiple access system the gray lines denote broadcast system.

2.4.1 FDMA-Frequency Division Multiple Access

FDMA-Frequency Division Multiple Access is a method for dividing available bandwidth into user number without overlapping. During the communication no other user can use same frequency range. The best benefit of this method is low intersymbol interference. Thus, it causes little or no required to use equalization in FDMA. The number of subchannels in FDMA can be given by following simple equation

$$N = B - 2B_{guard} / B_c \tag{2.16}$$

where B is total spectrum allocation, B_{guard} is guard band to protect from overlapping and B_c is bandwidth of subchannel. This methods is commonly used in wireline communication for data and voice transmission.

2.4.2 TDMA-Time Division Multiple Access

TDMA-Time Division Multiple Access systems divide the radio spectrum into time slots without overlapping and each slot is available for only one user to transmit or receiver. In this method, the data is transmitted by *buffer-and-burst method*. Therefore, the transmission is non continuous for any user and this results in low battery consumption since turning off when not in use. In comparison to the FDMA method, TDMA shares a single carrier frequency and is used in digital data and digital modulation methods. The number of subchannels in TDMA can be given by following simple equation

$$N = m\left(B - 2B_{guard}\right)\Big/B_c \tag{2.17}$$

where m is maximum number of TDMA users which are supported on each radio channel and B is total spectrum allocation, B_{guard} is guard band to protect from overlapping and B_c is bandwidth of subchannel. This methods is commonly used in wireline communication for data and voice transmission

2.4.3 CDMA-Code Division Multiple Access

As an alternative method, Code Division Multiple Access (CDMA) allows more users to use a common channel or subchannel by using spread spectrum signals. All users use the same carrier frequency range and can transmit simultaneously. Each user in CDMA, has a codeword approximately orthogonal to all other codewords. Therefore a small cross-correlation relationship is incurred in the receiver side. The signals of users are overlapping in both time and frequency domain. In comparison to TDMA and FDMA method, CDMA has a soft capacity limit. This means, there is no absolute limit on the number of users. Increasing of the number of users causes increasing noise floor in a linear manner.

2.5. Channel Models

In this section, some simple channel models are described to explain the channel capacity in the next chapter.

2.5.1 Binary channels

Let us consider an additive noise channel and let the modulator and the demodulator/detector be included as parts of the channel. If the modulator employs binary waveforms and the detector makes hard decisions, then the composite channel has a discrete-time binary input sequence and a discrete-time binary output sequence. Such a composite channel is characterized by the set $X = \{0,1\}$ of possible inputs, the set of $Y = \{0,1\}$ of possible outputs, and a set of conditional probabilities that relate the possible outputs to the possible inputs. If the channel noise and other disturbances cause statistically independent errors in the transmitted binary sequence with average probability p_{error}, then

$$\begin{aligned} P\left(Y = 0 \middle| X = 1\right) &= P\left(Y = 1 \middle| X = 0\right) = p_{error} \\ P\left(Y = 1 \middle| X = 1\right) &= P\left(Y = 1 \middle| X = 1\right) = 1 - p_{error} = p_{correct} \end{aligned} \tag{2.18}$$

Thus, we have reduced the cascade of the binary modulator, the waveform channel, and the binary demodulator and detector into an equivalent discrete time channel. This binary-input, binary-output, symmetric channel is simply called *binary symmetric channel* (BSC). Since each output bit from the channel depends only on the corresponding input bit, we say that the channel is memoryless

2.5.2 Discrete memoryless channels

The BSC is a special case of a more general discrete-input, discrete-output channel. Suppose that the output symbols from the channel encoder are Q-ary symbols, i.e., $X = \{x_0, x_1, ..., x_{q-1}\}$ and the output of the detector consists of Q-ary symbols, where $Q \geq M = 2^q$. If the channel and the modulation are memoryless, then the input-output characteristics of the composite channel are described by a set of qQ conditional probabilities.

$$P(Y = y_i | X = x_i) \equiv P(y_i | x_i) \tag{2.19}$$

where $i = 1, 2, ..., Q-1$ and $j = 1, 2, ..., q-1$. Such a channel is called a *discrete memoryless chanel* (DMC). Hence, if the input to a DMC is a sequence of n symbols $u_1, u_2, ..., u_n$ selected from the alphabet X and the corresponding output is the sequence $v_1, v_2, ..., v_n$ of symbols from the alphabet Y, the joint conditional probability is

$$\begin{aligned} &P(Y_1 = v_1, Y_2 = v_2, ..., Y_n = v_n | X = u_1, ..., X = u_n) \\ &= \prod_{k=1}^{n} P(Y_1 = v_k | X = u_k) \end{aligned} \tag{2.20}$$

This expression is simply a mathematical statement of the memoryless condition [17].

2.5.3 Discrete-input, continuous-output channel

Now, suppose that the input to the modulator comprises symbols selected from a finite and discrete input alphabet $X = \{x_0, x_1, ..., x_{q-1}\}$ and the output of the detector is unquantized $(Q = \infty)$. Then, the input to the channel decoder can assume any value on the real line, i.e., $Y = \{-\infty, \infty\}$. This leads us to define a composite discrete-time memoryless channel that is characterized by the discrete input X, the continuous output Y, and the set of conditional probability density functions,

$$p\left(y \middle| X = x_k\right), \quad k = 0,1,....,q-1 \tag{2.21}$$

The most important channel of this type is the additive white Gaussian noise (A WGN) channel, for which

$$Y = X + G \tag{2.22}$$

where G is a zero-mean Gaussian random variable with variance σ^2 and $X = x_k$, $k = 0,1,....,q-1$. For a given X, it follows that Y is Gaussian with me an x_k and variance σ^2. That is,

$$p\left(y \middle| X = x_k\right) = \frac{1}{\sqrt{2\pi\sigma}} e^{-y-x_k/2\sigma^2} \tag{2.23}$$

For a given input sequence $X_i, i = 1,2,....,n$, there is a corresponding output sequence

$$Y_i = X_i + G_i \quad i = 1,2,....,n \tag{2.24}$$

The condition that the channel is memoryless may be expressed as

$$p\left(y_1, y_1, ..., y_n \middle| X_1 = u_1, X_2 = u_2, ..., X_n = u_n\right) = \prod_{i=1}^{n} p\left(y_i \middle| X_i = u_i\right) \tag{2.25}$$

Chapter 3

Channel Capacity and Cooperative Diversity

This chapter consists of some knowledge about capacity for several channel models such as DMC, AWGN, band-limited AWGN, Multiple Access Channel, Broadcast channel and Relay Channel. In addition to these capacity expressions, the cooperative diversity technique and its implementation methods are explained in this following chapter.

3.1 Channel Capacity in DMC and AWGN

Consider in DMC (Discrete Memoryless Channel) which is introduced previously in Chapter 2. In this channel the input alphabet and the output alphabet are denoted as $X = \{x_0, x_1, ..., x_n\}$ and $Y = \{y_0, y_1, ..., y_m\}$. The channel is modeled by the conditional probability $P(Y = y_i | X = x_i) \equiv P(y_i | x_i)$ as given in Chapter 2.5.2. In case of transmitting the symbol x_j towards to destination, the mutual information is obtained by receiving the symbol y_i as;

$$I_{mutual} = \mathrm{E}\left[\log\left(P(y_i | x_i) / P(y_i)\right)\right] \tag{3.1}$$

where

$$P(y_i) \equiv P(Y = y_i) = \sum_{k=0}^{n-1} P(x_k) P(y_i | x_k) \tag{3.2}$$

so the average mutual information is defined as

$$I(X;Y)=\sum_{j=0}^{n-1}\sum_{i=0}^{m-1}P(x_j)P(y_i|x_j)\log\frac{P(y_i|x_j)}{P(y_i)} \tag{3.3}$$

The conditional probability $P(y_i|x_i)$ is determined by the channel characteristics but the probabilities of input symbols $P(x_i)$ are determined by encoder structure. The value of $I(X;Y)$ is maximized over the set of probability of input alphabet $P(x_i)$ and is a quantity that depends on the characteristics of the DMC through the conditional probabilities $P(y_i|x_i)$. This quantity is called the *capacity* of the channel and denoted by C. So the capacity of the DMC is defined as below

$$C=\max_{P(x_j)} I(X;Y)$$

$$=\max_{P(x_j)}\sum_{j=0}^{n-1}\sum_{i=0}^{m-1}P(x_j)P(y_i|x_j)\log\frac{P(y_i|x_j)}{P(y_i)} \tag{3.4}$$

under constraints $P(x_j)\geq 0$ and $\sum_{j=0}^{n-1}P(x_j)=1$

if we consider a discrete memoryless channel with the probability density function in (2.21), the capacity can be expressed by using maximum mutual information between discrete input $X=\{x_0,x_1,....,x_n\}$ and output $Y=\{-\infty,\infty\}$

$$C=\max_{P(x_i)}\sum_{i=0}^{n-1}\int_{-\infty}^{\infty}P(y|x_j)P(x_i)\log_2\frac{P(y|x_j)}{P(y)}dy \tag{3.5}$$

where

$$p(y)=\sum_{k=0}^{n-1}P(x_k)P(y|x_k) \tag{3.6}$$

3.2 Channel Capacity in band-limited AWGN

Consider a band-limited waveform channel with additive white Gaussian noise. Formally, the capacity of the channel per unit time has been defined by Shannon (1948) as

$$C = \lim_{T\to\infty} \max_{P(x)} \frac{1}{T} I\left(X;Y\right) \tag{3.7}$$

where the average mutual information $I\left(X;Y\right)$ is given as

$$I\left(X;Y\right) = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} P\left(x\right)P\left(y\middle|x\right)\log\frac{P\left(y\middle|x_j\right)P\left(x\right)}{P\left(y\right)P\left(x\right)}dxdy \tag{3.8}$$

In (3.8) X and Y are random variables with joint PDF $p\left(x,y\right)$ and marginal PDFs $p\left(x\right)$ and $p\left(y\right)$.

Beside this, we may use coefficients $\left\{x_i\right\}$, $\left\{y_i\right\}$ and $\left\{n_i\right\}$ in the series expansion of x_t, y_t and n_t respectively, to determine the average mutual information between $\underline{x}^N = \left[x_1, x_2, ..., x_N\right]$ and $\underline{y}^N = \left[y_1, y_2, ..., y_N\right]$ where $N = 2WT$, $y_i = x_i + n_i$ and $p\left(y_i\left|x_i\right.\right) = \frac{1}{\sqrt{2\pi\sigma}_x} e^{-\left(y_i - x_i\right)^2 / 2\sigma_x^2}$, is given by

$$I\left(X^N;Y^N\right) = \int_{x_1}\cdots\int_{x_N}\int_{y_1}\cdots\int_{y_N} P\left(y_N\left|x_N\right.\right)P\left(x_N\right)\log\frac{P\left(y_N\left|x_N\right.\right)}{P\left(y\right)}d\underline{x}^N d\underline{y}^N$$

$$= \sum_{i=0}^{m-1}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} P\left(y_i\middle|x_j\right)P\left(x_j\right)\log\frac{P\left(y_i\middle|x_j\right)}{P\left(y_i\right)}dy_i dx_i \tag{3.9}$$

The maximum value of $I\left(X;Y\right)$ over marginal PDFs $p\left(x_i\right)$ can be obtained when the input symbols x_i are statistically independent zero-mean Gaussian random variables characterized by the pdf

$$p(x_i) = \frac{1}{\sqrt{2\pi\sigma}_x} e^{-(y_i - x_i)2 \big/ 2\sigma^2 x} \tag{3.10}$$

where σ_x^2 is variance of each x_i. Then, by using (3.9) the mutual information can be expressed as

$$\begin{aligned}
\max_{p(x)} I(X_N; Y_N) &= \sum_{i=1}^{N} \frac{1}{2} \log\left(1 + \frac{2\sigma_x^2}{N_0}\right) \\
&= \frac{1}{2} N \log\left(1 + \frac{2\sigma_x^2}{N_0}\right) \\
&= WT \log\left(1 + \frac{2\sigma_x^2}{N_0}\right)
\end{aligned} \tag{3.11}$$

where channel noise variance is $N_0/2$. In case of putting power constraint on the average power of x_t

$$\begin{aligned}
P_{average} &= \frac{1}{T} \int_0^T \mathrm{E}\left[x^2(t)\right] dt \\
&= \frac{1}{T} \sum_{i=1}^{N} \mathrm{E}\left(x_i^2\right) \\
&= \frac{N\sigma_x^2}{T}
\end{aligned} \tag{3.12}$$

after some mathematical simplification, the variance of x_i

$$\begin{aligned}
\sigma_x^2 &= \frac{NP_{average}}{T} \\
&= \frac{P_{average}}{2W}
\end{aligned} \tag{3.13}$$

by substitution of (3.13) into (3.11) for variance yields

$$\max_{p(x)} I(X_N; Y_N) = WT \log\left(1 + \frac{P_{average}}{WN_0}\right) \tag{3.14}$$

the channel capacity of per unit can be obtained by dividing (3.13) by T. Thus

$$C = W \log\left(1 + \frac{P_{average}}{WN_0}\right) \tag{3.15}$$

This expansion in (3.15) is called as capacity of band-limited AWGN channel with power constraint.

3.3 Channel Capacity in Multiuser Communication Channels

In additional to single user channel capacity explained in section 3.1 and 3.2, the multi user channel capacity is related to multiple user techniques such as multiple access channel, broadcast channel, relay channel and Gaussian two-way channel which is used to share information in multiple channel.

3.3.1 Gaussian Multiple Access Channel

In general, the Gaussian multiple access channels with m transmitter and one receiver can be modeled as below

$$Y = \sum_{i=0}^{m-1} X_i + N \tag{3.16}$$

where Y, X_i and N denotes that received signal, transmitted signal of i th user which is independent respectively others, and the noise component with variance N_0 at the receiver side. The capacity region for Gaussian channel can be expressed in the following inequalities.

$$\sum_{i \in S} R_i \prec \frac{1}{2}\log\left(1 + \frac{\sum_{i \in S} P_i}{N_0}\right) \tag{3.17}$$

where $S \subset \{1,2,...,m\}$, R_i denotes that rate of i th user and the total achievable rate can be found by using last inequality in (3.16). It is interesting point that the sum of rate goes to infinity since the user number m grows to infinity.

3.3.2 Two User Gaussian Broadcast Channel

In broadcast channel, the system has a sender with power P and two receivers with Gaussian noise variances N_1 and N_2 of noise Z_1 and Z_2. If $N_1 \prec N_2$, it causes that Y_1 is less noisy than the receiver Y_2. The received signals Y_i is determined as

$$Y_i = X + Z_i; \;\; i = \{1,2\} \tag{3.18}$$

For encoding the messages the encoder needs two codebooks with indices $i \in \{1,2,....,2^{nR_1}\}$ and $j \in \{1,2,....,2^{nR_2}\}$. The encoder sends the sum of these codewords. Then the receivers decode these codewords. The receiver Y_2 merely searches closest codewords to the received signal in the second codebook. Thus the codewords with power αP from the first codebook are treated as a noise component for receiver Y_2. However the receiver Y_1 decodes Y_2's codewords. The receiver Y_1 subtracts this decoded codeword $\tilde{X}_2$ from Y_1. Then it searches new codewords which are the closest to the resulting difference $Y_1 - \tilde{X}_2$ in the first codebook. This method can achieve an error probability value as low as desired. The capacity region for a two user broadcast channel is given as

$$R_1 \prec \frac{1}{2}\log\left(1+\frac{\alpha P}{N_1}\right) \tag{3.19}$$

$$R_2 \prec \frac{1}{2}\log\left(1+\frac{(1-\alpha)P}{\alpha P + N_2}\right) \tag{3.20}$$

where R_1 and R_2 are rates to receiver Y_1 and Y_2. Beside this α is coefficient between $0 \le \alpha \le 1$ to trade of R_1 and R_2.

3.3.3 Gaussian Relay Channel

In relay channel, the system has a sender, a receiver and a relay which helps the receiver as illustrated in Figure 3.1. The relay channel can be characterized as

$$Y_1 = X + Z_1 \tag{3.21}$$

$$Y = X + X_1 + Z_2 \tag{3.22}$$

where Z_1 and Z_2 are Gaussian noise components with the variance N_1 and N_2 at the relay and the receiver side. And the encoding structure in relay is a function of received signal from transmitter as given by

$$X_{1i} = f_i\left(Y_{1i}\right) \tag{3.23}$$

The transmitter X has power P and relay X_1 has power P_1. The capacity of the relay channel is not known. However, the following rate region can be achieved as

$$C = \sup_{p(x,x_1)} \min\left\{I\left(X, X_1; Y\right), I\left(X; Y_1 \middle| X_1\right)\right\} \tag{3.24}$$

$$C = \max_{0\le\alpha\le1} \min\left\{\frac{1}{2}\log\left(1+\frac{P+P_1+\sqrt{\alpha' P P_1}}{N_1}\right), \frac{1}{2}\log\left(1+\frac{\alpha P}{N_1}\right)\right\} \tag{3.25}$$

where $\alpha' = 1-\alpha$.

Let $R_1 \prec \frac{1}{2}\log\left(1+\alpha P/N_1\right)$. Two codebooks are needed. The first codebook has 2^{nR_1} words of power αP. The second has 2^{nR_0} codewords of power $\alpha' P$. We shall use codewords from these codebooks successively in order to create the opportunity for cooperation by the relay. We start by sending a codeword from

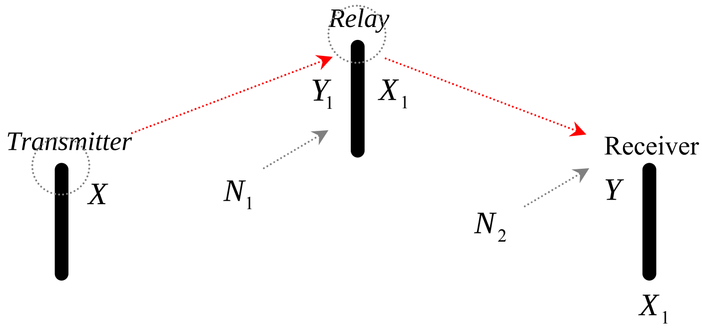

Figure 3.1: Illustration of relay channel with one relay

---

the first codebook. The relay now knows the index of this codeword since $R_1 \prec \frac{1}{2}\log\left(1+\alpha P/N_1\right)$, but the intended receiver has a list of possible codewords of size $2^{n\left(R_1-\frac{1}{2}\log\left(1+\alpha P/N_2\right)\right)}$. This list calculation involves a result on list codes.

In the next block, the transmitter and the relay wish to cooperate to resolve the receiver's uncertainty about the previously sent codeword on the receiver's list. Unfortunately, they cannot be sure what this list is because they do not know the received signal Y. Thus they randomly partition the first codebook into 2^{nR_0} cells with an equal number of codewords in each cell. The relay, the receiver, and the transmitter agree on this partition. The relay and the transmitter find the cell of the partition in which the codeword from the first codebook lies and cooperatively send the codeword from the second codebook with that index. That is, both X and X_1 send the same designated codeword. The relay, of course, must scale this codeword so that it meets his power constraint P_1. They now simultaneously transmit their codewords. An important point to note here is that the cooperative information sent by the relay and the transmitter is sent coherently. So the power of the sum as seen by the receiver Y is $\left(\sqrt{\alpha' P}+\sqrt{P_1}\right)^2$.

However, this does not exhaust what the transmitter does in the second block. The transmitter also chooses a fresh codeword from the first codebook, adds it "on paper" to the cooperative codeword from the second codebook, and sends the sum over the channel.

The reception by the ultimate receiver Y in the second block involves first finding the cooperative index from the second codebook by looking for the closest codeword in the second codebook. The receiver subtracts the codeword from the received sequence, and then calculates a list of indices of size 2^{nR_0} corresponding to all codewords of the first codebook that might have been sent in the second block.

Now it is time for the intended receiver to complete computing the codeword from the first codebook sent in the first block. He takes his list of possible codewords that might have been sent in the first block and intersects it with the cell of the partition that he has learned from the cooperative relay transmission in the second block. The rates and powers have been chosen so that it is highly probable that there is only one codeword in the intersection. This is Y's guess about the information sent in the first block.

We are now in steady state. In each new block, the transmitter and the relay cooperate to resolve the list uncertainty from the previous block. In addition, the transmitter superimposes some fresh information from his first codebook to this transmission from the second codebook and transmits the sum.

The receiver is always one block behind, but for sufficiently many blocks, this does not affect his overall rate of reception [8].

3.4 Cooperative Diversity

Transmit diversity generally requires more than one antenna at the transmitter. However, many wireless devices are limited by size or hardware complexity to one antenna. Recently, a new class of methods called cooperative communication has been proposed that enables single antenna mobiles in a multi-user environment to share their antennas and generate a virtual multiple-antenna transmitter that allows them to achieve transmit diversity. The advantages of multiple-input multiple-output(MIMO) systems have been widely acknowledged, to the extent that certain transmit diversity methods (i.e., Alamouti signaling) have been incorporated into wireless standards. Although transmit diversity is clearly advantageous on a cellular base station, it may not be practical for other scenarios. Specifically, due to size, cost, or hardware limitations, a wireless agent may not be able to support multiple transmit antennas. Examples include most handsets (size) or the nodes in a wireless sensor network (size, power) [20].

The basic idea in cooperative communication illustrated in Figure 3.2, is that single-antenna mobiles in a multi-user scenario can share their antennas in a manner that creates a virtual MIMO system. Several important milestones in this area have been achieved, leading to a flurry of further research activity. Transmitting independent copies of the signal generates diversity and can effectively combat the deleterious effects of fading. In particular, spatial diversity is generated by transmitting signals from different locations, thus allowing independently faded versions of the signal at the receiver. Cooperative communication generates this diversity in a new and interesting way. Each mobile has one antenna and can not individually generate spatial diversity. However, it may be possible for one mobile to receive the other, in which case it can forward some version of “overheard” information along with its own data. Because the fading paths from two mobiles are statistically independent, this generates space diversity.

Cooperative communication is based on the relay channel explained in Section 3.3 The relay channel was introduced by Van der Meulen [18] and investigated extensively by Cover and El Gamal [1]. Cover and El Gamal provided a number of relaying strategies, found achievable regions and provided upper bounds to the capacity of a general relay channel. They also provided an expression for the capacity of the degraded relay channel, in which

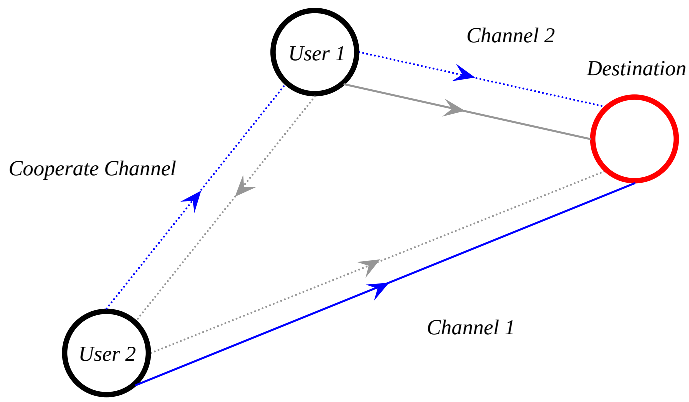

Figure 3.2: Illustration of basic two user's cooperative communication with each user node is both a transmitter and a relay

the communication channel between the source and the relay is physically better than the source-destination link.

Since the source is generally further away from the destination than the relay, the received signal at the destination due to the source would be much weaker than the relay signal. However, when fading is also taken into account, this scheme would incur considerable loss, especially in diversity, compared to one in which the destination processes both signals. Hence one can use multi-hop not just to overcome path loss, but also to provide diversity. Motivated by the above observation, cooperative communication involves two main ideas: (i) Use relays (or multi-hop) to provide spatial diversity in a fading environment, (ii) Envision a collaborative scheme where the relay also has its own information to send so both terminals help one another to communicate by acting as relays for each other (called "partners").

One can think of a cooperative system as a virtual antenna array, where each antenna in the array corresponds to one of the partners. The partners can overhear each other's transmissions though the wireless medium, process this information and re-transmit to collaborate. This provides extra observations of the source signals at the destinations, the observations which are dispersed in space and usually discarded by current implementations of cellular, wireless LAN or ad-hoc systems [15].

The main cooperative communication signaling models can be classified into two categories as illustrated in Figure 3.3a and 3.3b. The details of the cooperative communication are explained in Chapter 4.

3.4.1 Amplify and Forward Method

The A*mplify-and-Forward method* is perhaps the most conceptually simple of the cooperative signaling methods. In this method, each user receives a noisy version of the signal transmitted by its partner. As the name implies, the user then amplifies and retransmits this noisy signal as illustrated Figure 3.3a. The receiver will combine the information sent by the user and partner and will make a final decision on the transmitted bit. Although the noise of the partner is amplified in this scheme, the receiver still receives two independently faded versions of the signal and is thus able to make better decisions for the transmitted bits. A potential challenge in this scheme is that sampling, amplifying, and retransmitting analog values may be technologically non-trivial. Nevertheless, Amplify-and-Forward is a simple method that lends itself to analysis, and therefore has been very useful in furthering the understanding of cooperative communication systems [20].

3.4.2 Decode and Forward Method

The Decode-and-Forward method illustrated in Figure 3.3b is related to traditional relay channel idea. With this method, a user detects the partner's information and then retransmits the detected signal towards destination. Erroneous frames are not transmitted to destination. For this reason, the performance of the decoder can be decreased. However the successful decoded frames are re-encoded and retransmit to the destination. The main advantage of Amplify-and-Forward over Decode-and-Forward is that no hard decisions are

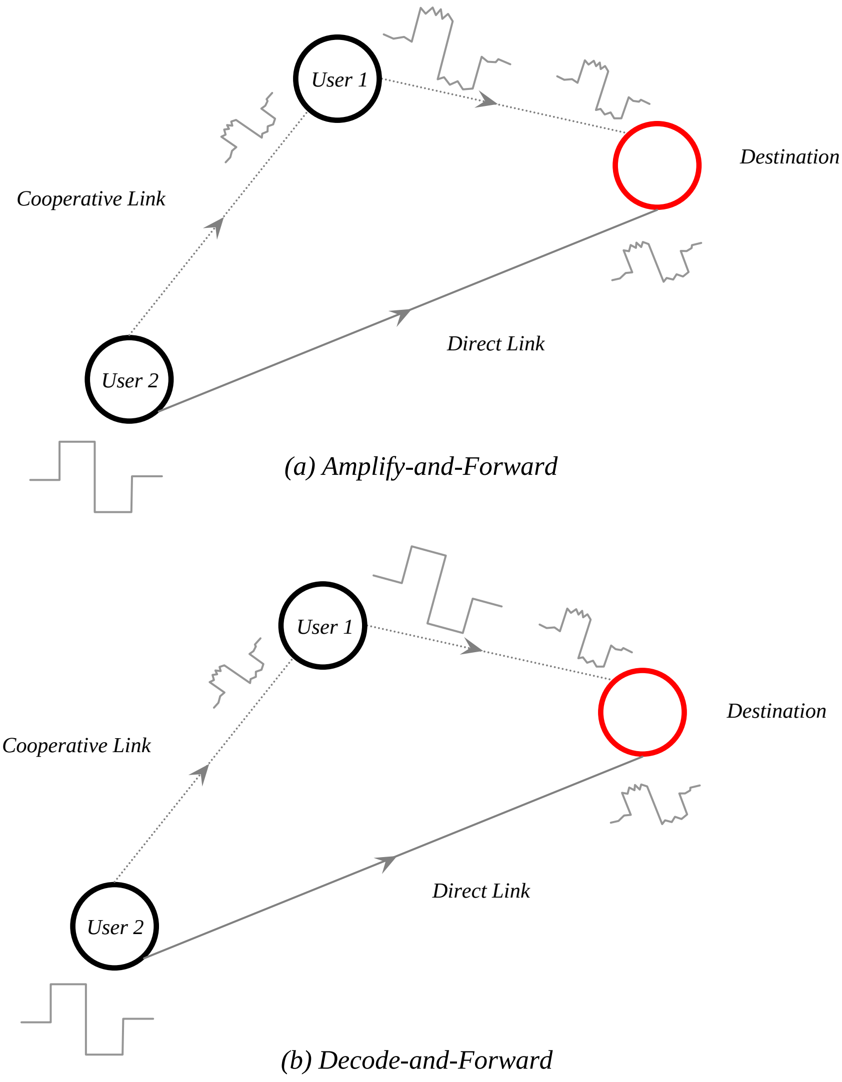

Figure 3.3: a) Illustration of two user's cooperative communication with Amplify and Forward method. b) Illustration of two user's cooperative communication with Decode and Forward method. [20]

---

made, but on the contrary Amplify-and-Forward does not regenerate the signal. In other words, Decode-and-Forward regenerates the signal while AF does not lose soft information. This is main disadvantage of Decode-and-Forward method over Amplify-and-Forward.

Chapter 4

Transmit Strategies for Half Duplex User Cooperation Systems

4.1 Introduction

The wireless communication technologies improve day by day. Therefore, there is need to satisfy the demands of the capacity improvements in wireless communication. Based on constant increasing demand, several diversity techniques have been proposed such as frequency diversity, time diversity, spatial diversity and recently cooperation diversity.

As noted in Chapter 2, frequency diversity technique can be used to combat frequency selective fading effects. Meanwhile, time diversity can be an effective way to combat time selective fading if error control and interleaving method are employed together. The space diversity technique can combat both frequency selective fading and time selective fading. In other words, the main benefit of spatial diversity is that it provides a way to prevent effects of multi path fading. However, as explained in Chapter 3 there is a physical constraint which limits the spatial diversity: the small size of portable devices. It may be not possible to develop more than one antenna on the same mobile device. Therefore, alternative means of achieving diversity is needed. Cooperation diversity can be employed to develop diversity gain in the absence of more than one antenna at the same mobile device.

The simplest example of a cooperative network is the relay channel, which was first introduced in by Van der Meulen [18]. Cover and El Gamal [22] then provided a number of relaying strategies, found achievable rate regions. Cover and Leung [1] considered a two user which have access to the channel output for MAC and obtained an achievable rate region. The same rate region was demonstrated by Willems and van der Meulen, if there is feedback between only one of the transmitters and channel output [3]. Willems [4] also provided the capacity region of the MAC with partially cooperating encoders. Willems, van der Meulen, Schalkwijk proposed an achievable rate region with generalized feedback for a MAC [6]. Sendonaris, Erkip, and Aazhang [2], [7] first presented a general information theoretic model for cooperation between a pair of users. They then proposed a CDMA-based implementation of detect-and-forward cooperative signaling, and achievable rate regions and outage capacity for this scheme. The outage probability was characterized by Laneman, Tse and Wornell [22] for half duplex scheme where there is no channel state information at the transmitters.

In this Chapter, we consider a two user system where the users communicate with the receiver by using a cooperative scheme to share their information over a fading channel. Our main purpose in this Chapter is to obtain an effective way of reaching as high a capacity region as possible by using a half duplex communication scheme.

4.2 System Model

We firstly consider a two user cooperative scheme in fading Gaussian MAC with known channel state information at the receiver and transmitter side as illustrated in Figure 4.1. Each mobile user wants to communicate with the same intended receiver. The discrete time cooperative communication model for two users can be modeled as below [7]

$$Y_0 = \alpha_{10} X_1 + \alpha_{20} X_2 + N_0 \tag{4.1}$$

$$Y_1 = \alpha_{21} X_2 + N_1 \tag{4.2}$$

$$Y_2 = \alpha_{12} X_1 + N_2 \tag{4.3}$$

where Y_i is the received signal at user i; X_i is the transmitted signal of user i and N_i is channel noise at wireless node i which is modeled by an independent zero-mean Gaussian random variable with variance σ_i^2; α_{ij} are modeled as independent complex circularly symmetric Gaussian random variables which are assumed to be constant over a symbol period.

As illustrated in Figure 4.1, each user receives an attenuated and noisy version of transmitted signal. Nevertheless, each user has decoding capability for its partner's transmitted signal X_i. The transmitted and received signals at each mobile nodes are encoded by *superposition block Markov encoding* and decoded by *backward decoding* [6] algorithm where the receiver doesn't work on decoding until all B block codewords completely received.

After some preliminary information about the system in consideration we can now mathematically describe the encoding strategy.

Let us start by explaining the two user full-duplex cooperative communication model [7]. This will allow us to obtain our proposed half-duplex model.

As understood from (4.1), (4.2) and (4.3) X_i is the signal transmitted by user i which is encoded by super position Markov encoding scheme. This coding scheme, for a two user MAC with generalized feedback is illustrated in Figure 4.2 [7]. The transmitted signal X_i for the two user scheme can be expressed as below;

$$X_1 = P_{10}(\alpha)X_{10} + P_{12}(\alpha)X_{12} + P_{u1}(\alpha)U \tag{4.4}$$

$$X_2 = P_{20}(\alpha)X_{20} + P_{21}(\alpha)X_{21} + P_{u2}(\alpha)U \tag{4.5}$$

where X_{10} and X_{20} are transmitted signals towards destination directly which are treated as a noise component for the other cooperative user; X_{12} and X_{21} are transmitted signals towards partners; U is the cooperative signal which contains information from the previous blocks. $P_{10}(\alpha)$, $P_{20}(\alpha)$, $P_{12}(\alpha)$

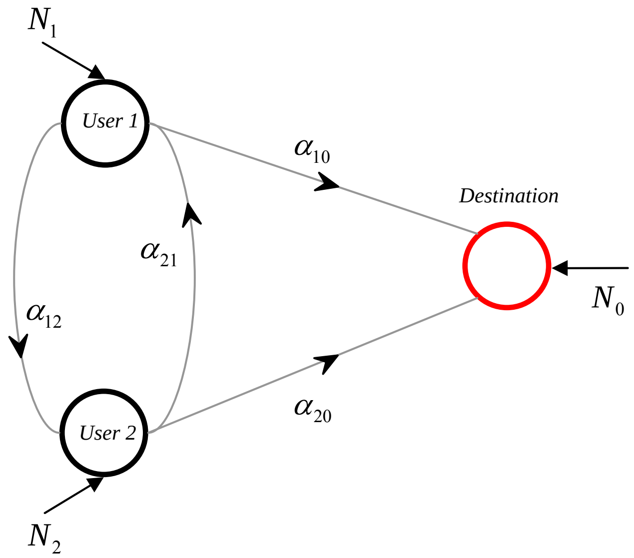

Figure 4.1: Illustration of two user's cooperation communication with explained channel attenuations

---

$P_{21}(\alpha)$, $P_{u1}(\alpha)$ and $P_{u2}(\alpha)$ are the allocated power values for sub signals in X_1 and X_2, which are functions of the channel state vector $\underline{\alpha} = [\alpha_{10}, \alpha_{12}, \alpha_{20}, \alpha_{21}]$ Beside this, the power allocations are defined with the following constraints

$$P_1(\underline{\alpha}) = P_{10}(\alpha) + P_{12}(\alpha) + P_{u1}(\alpha) \tag{4.6}$$

$$P_2(\underline{\alpha}) = P_{20}(\alpha) + P_{21}(\alpha) + P_{u2}(\alpha) \tag{4.7}$$

$$\mathrm{E}\left[P_i(\underline{\alpha})\right] \leq \bar{P}_i \quad i = \in \{1, 2\} \tag{4.8}$$

where $\bar{P}_i$ is average power of user i .

The benefits of this signal scheme in the capacity expansion are detailed in Section 4.3. The sub signals in (4.4), (4.5) are expressed as a function of some not only the current messages, but also the messages from the previous block. The sub signals X_{10}, X_{12} and U can be expressed as below

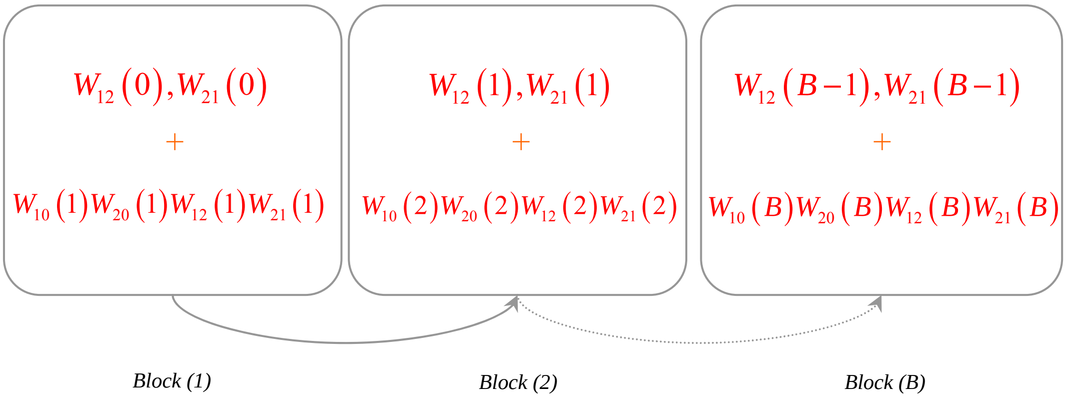

Figure 4.2: Illustration of super position Markov encoding algorithm for "B" blocks

---

$$\begin{Bmatrix} X_{10} = \sqrt{P_{10}}\, X_{10}\left(W_{10}(i), W_{12}(i-1), W_{21}(i-1)\right) \\ X_{12} = \sqrt{P_{12}}\, X_{12}\left(W_{12}(i), W_{12}(i-1), W_{21}(i-1)\right) \\ U = \sqrt{P_{u1}}\left(W_{12}(i-1), W_{21}(i-1)\right) \end{Bmatrix} \quad ; i \in \{1, 2 \ldots B\} \qquad (4.9)$$

where i and $i-1$ indicate the current block and previous block. W_{10}, W_{20} are parts of sending information intended towards destination and W_{12}, W_{21} are parts of information intended both towards its partner and towards the destination. User 2's signals X_{20}, X_{21} and U are characterized similary. The following section, which summarizes the results in [1], [7], [24] illustrates the computation of the achievable rates for two user cooperative communication. The signal X_1 does not only depend on new information $W_{10}(i)$ and $W_{12}(i)$, but it is also depends on the previous block $W_{12}(i-1)$ and $W_{21}(i-1)$ which are helpful to generate cooperative signal.

4.3 Achievable Rate

In this section we compute the achievable region for case of using cooperative communication structure in Figure 4.1. The cooperative model was explained in (4.1)-(4.3) at the beginning of this chapter. As described previously the cooperation structure is based on superposition Markov encoding and backward decoding strategy. Beside this, in cooperative communication structure, each user transmits its information, partner's information and cooperative information which is constituted from received partner's signal.

Therefore, the transmitted signal of user 1 can be modeled as below without power allocation.

$$X_1 = X_{10} + X_{12} + U \tag{4.10}$$

where X_{10} is employed for transmitting of W_{10} directly to destination at rate R_{10}, X_{12} is employed for transmitting of W_{12} directly to user 2 at rate R_{12} and U is employed for transmitting cooperative signals $W_{12}(i-1)$ and $W_{21}(i-1)$ directly to destination. The User 2 works to send its signals similarly. This signal scheme with power allocation was determined in (4.4). Separately, the power allocation must satisfy these equations in (4.6)-(4.8);

Based on Markov encoding and backward decoding scheme, we can generate the achievable rate region akin to [2]. R_1, R_2 denote the rates of User 1 and User 2. They have two sub components R_{i0} and R_{i2} which describe the rates between user-destination and user-cooperative partner.

To understand the achievable rate region in cooperation diversity structure, firstly we should explain the backward decoding scheme. As described previously, the destination must wait until all blocks are received. In the last block of code sequence $W_{10}(B)$, $W_{20}(B)$, $W_{12}(B)$ and $W_{21}(B)$ contain no new information and they can be set $W_{10}(B)$, $W_{12}(B)$, $W_{20}(B)$, $W_{21}(B) = (0,0,0,0)$. Then the decoding algorithm starts decoding from the $B'th$ block to first block. In case of transmitting no new information at the last block, the total information rate reduces by a coefficient of $(B-1)/B$. However, the reduction of information ratio can be undervalued at large B. In contrary to the encoding scheme, in the backward decoding the destination wants to decode message by help of $W\ (i+1)$. This decoding strategy can be clarified in Figure 4.3.

By using properties of the mutual information, the capacity of User1 and User 2 in destination side can be described as;

$$R_1 \prec \mathrm{I}\left(X_1;Y\middle|X_2\right) \quad \textit{and} \quad R_2 \prec \mathrm{I}\left(X_2;Y\middle|X_1\right) \tag{4.11}$$

$$R_1 + R_2 \prec \mathrm{I}\left(X_1, X_2; Y\right) \tag{4.12}$$

where

$$R_1 = R_{10} + R_{12} \tag{4.13}$$

$$R_2 = R_{20} + R_{21} \tag{4.14}$$

As explained in the previous page, R_{10} is a rate of user 1 towards destination side and can be expressed as;

$$R_{10} \prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{10}^2 P_{10}(\alpha)}{N_0}\right]\right\} \tag{4.15}$$

Nevertheless, in order to find R_{12}, which is the rate of User 1 towards cooperative side, we first look at block 1 illustrated in Figure 4.2. Here, we set $\left(W_{12}(0), W_{21}(0)\right) = (0,0)$. These signals are known by the User 1 and User 2. The first signal $X_1(1)$ contains $W_{12}(0)$ and $W_{21}(0)$ which are known, $W_{12}(1)$ which is of interest to User 2 $W_{10}(1)$ which is not attempted to be decoded by User 2. User 2 will treat X_{10} as noise. The rate R_{12} between User 1 and User 2 can then be found by the following equation

$$R_{12} \prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{12}^2 P_{12}(\alpha)}{\alpha_{12}^2 P_{10}(\alpha)+N_2}\right]\right\} \tag{4.16}$$

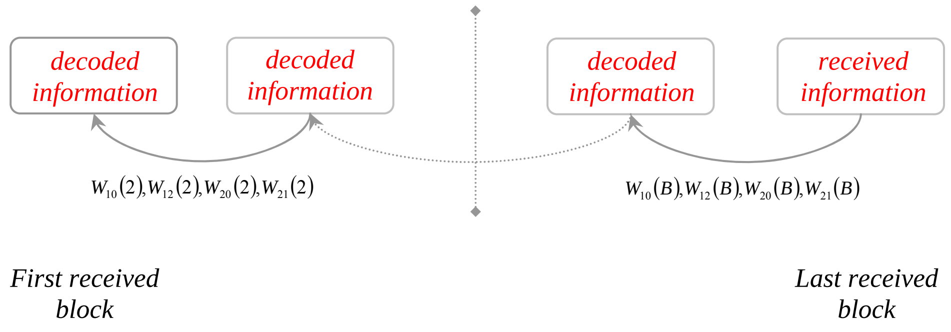

Figure 4.3: A simple illustration of backward decoding algorithm for "B" blocks

---

where P_{10} is treated as a noise component at the cooperative partner side so it must be in denominator of logarithm. R_{20}, R_{21} is in similar fashion and can be expressed as;

$$R_{20} \prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{20}^2 P_{20}(\alpha)}{N_0}\right]\right\} \tag{4.17}$$

$$R_{21} \prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{21}^2 P_{21}(\alpha)}{\alpha_{21}^2 P_{20}(\alpha)+N_1}\right]\right\} \tag{4.18}$$

in (4.18) P_{20} is treated a noise component like (4.17). And the sum of rates at the destination side, R_{10}, R_{20} are bounded by

$$R_{10}+R_{20} \prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{10}^2 P_{10}(\alpha)+\alpha_{20}^2 P_{20}(\alpha)}{N_0}\right]\right\} \tag{4.19}$$

The decoding starts from last block where there is no new information. In the last block, the decoder wants to decode $W_{12}(B-1)$ and $W_{21}(B-1)$ since $W_{10}(B), W_{20}(B), W_{12}(B), W_{21}(B)$ are set to zero. The following rate constraint needs to be satisfied at the destination side

$$R_{12}+R_{21}\prec \mathrm{E}\left\{\log\left[1+\frac{\alpha_{10}^2P_1(\alpha)+\alpha_{20}^2P_2(\alpha)+2\sqrt{\alpha_{10}^2\alpha_{20}^2P_{u1}(\alpha)P_{u2}(\alpha)}}{N_0}\right]\right\} \tag{4.20}$$

where P_1 and P_2 are total power which are described in (4.6), (4.7) and which must also satisfy the average power constraint in (4.8)

The term $2\sqrt{\alpha_{10}^2\alpha_{20}^2P_{u1}(\alpha)P_{u2}(\alpha)}$ in (4.20) is due to the coherent addition of cooperative signals U at the destination side. In the next step, the decoder wants to decode $W_{10}(B-1)$, $W_{20}(B-1)$, $W_{12}(B-1)$, $W_{21}(B-1)$.

However, the sum rate in side of destination is dominated by (4.21) and can be expressed as proved in [1]

$$\begin{aligned}R_{10}+R_{20}+R_{12}+R_{21}\prec \mathrm{E}\Bigg\{\log\Bigg[1&+\frac{\alpha_{10}^2P_1(\alpha)+\alpha_{20}^2P_2(\alpha)}{N_0}\\&+\frac{2\sqrt{\alpha_{10}^2\alpha_{20}^2P_{u1}(\alpha)P_{u2}(\alpha)}}{N_0}\Bigg]\Bigg\}\end{aligned} \tag{4.21}$$

Beside this, in consideration of given power allocation structures and constraints in (4.6)-(4.8) this rate in (4.21) can be dominated by sum of individual rates. Thus, sum of R_1 and R_2, which are denoted by $R_{10}+R_{12}$ and $R_{20}+R_{21}$ respectively, is bounded by

$$\begin{aligned}R_1+R_2\prec \min\Bigg(&\mathrm{E}\left\{\log\left[1+\frac{\alpha_{10}^2P_1(\alpha)+\alpha_{20}^2P_2(\alpha)+2\sqrt{\alpha_{10}^2\alpha_{20}^2P_{u1}(\alpha)P_{u2}(\alpha)}}{N_0}\right]\right\},\\&\mathrm{E}\Bigg\{\log\left[1+\frac{\alpha_{10}^2P_{10}(\alpha)+\alpha_{20}^2P_{20}(\alpha)}{N_0}\right]+\log\left[1+\frac{\alpha_{12}^2P_{12}(\alpha)}{\alpha_{12}^2P_{10}(\alpha)+N_2}\right]\\&+\log\left[1+\frac{\alpha_{21}^2P_{21}(\alpha)}{\alpha_{21}^2P_{20}(\alpha)+N_1}\right]\Bigg\}\Bigg)\end{aligned} \tag{4.22}$$

4.4 Adaptive Half Duplex Scheme

In the previous sections, the full duplex cooperative communication model and its achievable region are reviewed. In this section, our proposed half duplex structure, the numerical solution and the algorithm are described.

The half duplex scheme in cooperative communication can be advantageous according when compared to the full-duplex scheme due to high complexity of the latter.

Let us start by assuming two users who want to communicate by using half duplex cooperative communication structure as illustrated in Figure 4.1. In fact, our proposed model has different achievable rates than in (4.22), due to the usage of the half duplex scheme. As explained in the definition of the half duplex communication system, each user is allowed to conduct only one directional communication, which means, each user must decide whether to be in the transmitting mode or the receiving mode. In Figure 4.4, there are some possible illustrated scenarios which are based on their transmission decision.

To obtain the achievable rates for the half duplex scheme, there is need to define some power constraints. By choosing the half duplex scheme, we introduce the constraints $P_{10}.P_{21} = 0$, $P_{20}.P_{12} = 0$, $P_{12}.P_{21} = 0$, $P_{12}.P_{U2} = 0$, $P_{21}.P_{U2} = 0$. These standard power constraints are listed in Table 4.1. Besides, in the cooperative communication, the users are usually located far away from the destination. For this reason the direct channel attenuation between users and the destination α_{10}, α_{20} are often worse than cooperative channel attenuations α_{12}, α_{21} with high probability. Based on this fact, our proposed model can be developed under the assumption: $\alpha_{12} \succ \alpha_{10}$ and $\alpha_{21} \succ \alpha_{20}$. This assumption simplifies the expressions for the achievable rates. Under this assumption, for every channel attenuation, the direct power components P_{10} and P_{20} should be equal to zero to improve the achievable rate [24]. In case of

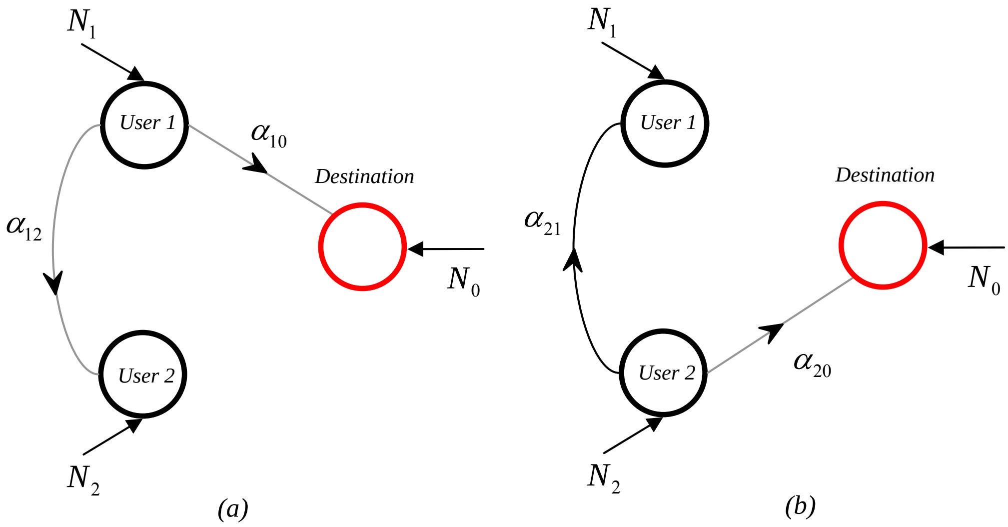

Figure 4.4: Illustration of 2 possible half duplex communication schemes. a) User 1 has decided to be in transmission mode and User 2 in listening mode. b) User 1 has decided to be in listening mode and User 2 in transmission mode.

---

discarding the direct power components P_{10} and P_{20}, we can allocate the power to the other components such as P_{12}, P_{21}, P_{U1} and P_{U2} [24].

As explained previously, there are several possible transmission schemes in half duplex structure. Among these transmission schemes, we throw away the ones where only one of the two power components P_{U1} and P_{U2} is equal to zero. In such a case, there is no advantage rate gain due to coherent combining, and these new constraints are also listed in Table 4.1.

After these power constraints in Table 4.1, there are only 3 transmission schemes to communicate toward destination.

- Case 1: In this case, the cooperative power component P_{12} is greater than zero based on power constraints. The other power components P_{21}, P_{U1} and P_{U2} are selected as zero.
- Case 2: In this case, the cooperative power component P_{21} is greater than zero based on power constraints. The other power components P_{12}, P_{U1} and P_{U2} are selected as zero.

Power Constraints	$P_{10}.P_{21}=0$, $P_{20}.P_{12}=0$, $P_{12}.P_{21}=0$, $P_{12}.P_{U2}=0$, $P_{21}.P_{U2}=0$
Extra Power Constraints	always $P_{10}=0$, $P_{20}=0$ If $P_{U1}=0$ then $P_{U2}=0$ or If $P_{U2}=0$ then $P_{U1}=0$
Case 1	$P_{12}>0$, $P_{21}=0$, $P_{U1}=0$, $P_{U2}=0$, $P_{10}=0$, $P_{20}=0$
Case 2	$P_{21}>0$, $P_{12}=0$, $P_{U1}=0$, $P_{U2}=0$, $P_{10}=0$, $P_{20}=0$
Case 3	$P_{12}=0$, $P_{21}=0$, $P_{U1}>0$, $P_{U2}>0$, $P_{10}=0$, $P_{20}=0$

Table 4.1: Power constraints and possible transmission sceneries

- Case 3: In this case, the cooperative power components P_{U1} and P_{U2} are greater than based on power constraints. The other power components P_{12} , P_{21} are selected as zero

Using the possible power allocations given in those there cases, the expression for the sum rate of half duplex scheme can be obtained from (4.22) as

$$
\begin{aligned}
R_{proposed} = \min\Bigg(& \mathrm{E}_{case1}\left\{\log\left(1+\left(\frac{\alpha_{10}^2 P_{12}}{N_0}\right)\right)\right\}+\mathrm{E}_{case2}\left\{\log\left(1+\left(\frac{\alpha_{20}^2 P_{21}}{N_0}\right)\right)\right\} \\
& +\mathrm{E}_{case3}\left\{\log\left(1+\left(\frac{\alpha_{10}^2 P_{U1}+\alpha_{20}^2 P_{U2}+2\sqrt{\alpha_{10}^2\alpha_{20}^2 P_{U1}P_{U2}}}{N_0}\right)\right)\right\}, \\
& \mathrm{E}_{case1}\left\{\log\left(1+\left(\frac{\alpha_{12}^2 P_{12}}{N_2}\right)\right)\right\}+\mathrm{E}_{case2}\left\{\log\left(1+\left(\frac{\alpha_{21}^2 P_{21}}{N_1}\right)\right)\right\}\Bigg)
\end{aligned}
\tag{4.23}
$$

Now we should decide that which cases should be selected under which channel conditions. To this end, we firstly searched a mathematical solution to create a decision algorithm. However we were not able to obtain an optimum closed form solution. there is no mathematical solution. For this reason we propose a numerical solution to obtain as high as achievable rates as possible using our proposed model. The decision algorithm is summarized in Table 4.2. The main idea of this decision algorithm is to make use of the relative strength of channel states while choosing the transmit strategies. For example, if the cooperative links are much stronger than the direct links, one would want to use either P_{12} or P_{21}, since these will contribute more to the objective function. However, in order to convey the cooperative information to the receiver, one should also choose to transmit using P_{U1} and P_{U2} together, which, heuristically should be used when the direct links are not very weak when compared to the cooperative links, we define a threshold function

$$\Gamma\left(\underline{\alpha}\right)=\frac{\alpha_{12}\alpha_{21}}{\alpha_{10}\alpha_{20}} \tag{4.24}$$

This threshold function $\Gamma\left(\underline{\alpha}\right)$ is compared with a threshold value γ^*. We very this threshold and search an optimum threshold value γ^*, for which the sum rate is maximized. Therefore, we use a computer search algorithm to find the best threshold value using a step size 0.1 for the threshold. The sample search result is illustrated in Figure 4.5. In this sample simulation, we obtain the best threshold value of 9, which results in a sum of rate equal to 0.4498, when $\mathrm{E}\left[\alpha_{10}\right]=0.3$, $\mathrm{E}\left[\alpha_{20}\right]=0.3$, $\mathrm{E}\left[\alpha_{12}\right]=0.9$, $\mathrm{E}\left[\alpha_{21}\right]=0.9$. As understood from Figure 4.5, the pick value denotes that the best threshold value. As explained previously, it gives the best achievable region. The searching algorithm is explained in Section 4.4.1.

4.4.1 The Best Threshold Searching Algorithm

The following steps explain the searching algorithm

- Generated channel attenuation with their probabilities.
- Regenerated these channel attenuations as vector forms.

- Decided on the search interval for searching gamma. It depends on the channel mean. For this reason, we should select small values for fast algorithm since channel means are closer to each other.
- Computed all achievable rate based on the decision algorithm in Table 4.2.
- Computed the best threshold value which gives the highest achievable rates.
- Computed the best power values by using the best threshold value.
- Compared with MAC and Full duplex model with no power control.

The Matlab code is clarified in Appendix A.

Comparison		*Decisions*	
$\Gamma(\underline{\alpha}) < \gamma^*$	$\alpha_{12} > \alpha_{21}$	$P_{12} > 0; P_{21} = 0$	$P_{10} = 0; P_{20} = 0$
	$\alpha_{21} > \alpha_{12}$	$P_{21} > 0; P_{12} = 0$	
$\Gamma(\underline{\alpha}) > \gamma^*$		$P_{U1} > 0; P_{U2} > 0$	

Table 4.2: Comparison and Decision algorithm in proposed model

---

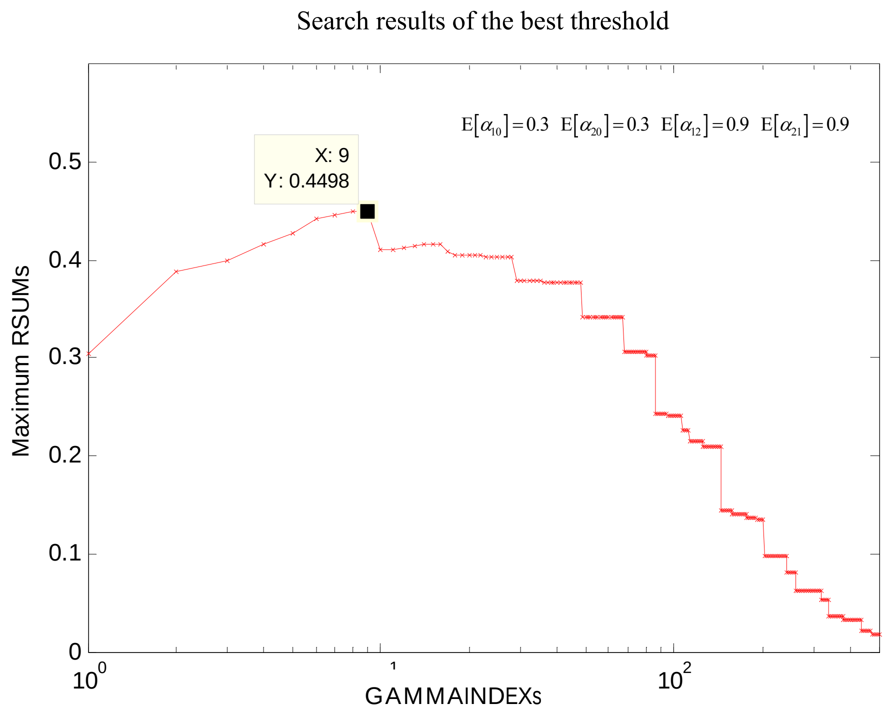

Figure 4.5: Illustration of search results of the best threshold. The best threshold value “9” is obtained when “Rsum=0.4498”

4.5 Results

In this section, the comparative results are illustrated in Figure 4.6 – 4.7. The first simulation results were obtained when $\mathrm{E}\left[\alpha_{10}\right]=0.3$, $\mathrm{E}\left[\alpha_{20}\right]=0.3$, $\mathrm{E}\left[\alpha_{12}\right]=0.6$, $\mathrm{E}\left[\alpha_{21}\right]=0.6$. In Figure 4.6, we compare our proposed half duplex structure with the full duplex structure and traditional MAC. In this figure we see that our adaptive power control algorithm shows close performance to the non-power controlled full duplex structure and the best achievable rate was obtained at the best threshold value 2.600. Similarly, we see another comparative simulation result in Figure 4.7. This result is obtained at the best threshold value 8.500 when $\mathrm{E}\left[\alpha_{10}\right]=0.3$, $\mathrm{E}\left[\alpha_{20}\right]=0.3$, $\mathrm{E}\left[\alpha_{12}\right]=0.9$, $\mathrm{E}\left[\alpha_{21}\right]=0.9$. Our proposed half duplex model in this simulation shows better performance than the full duplex structure. The main reason of getting better performance is our adaptive power control algorithm. As explained in previous sections, the threshold gamma function is based on channel attenuations. For this reason we get better performance in our adaptive power control algorithm due to the high ratio between cooperative and direct channel attenuations, on the average.

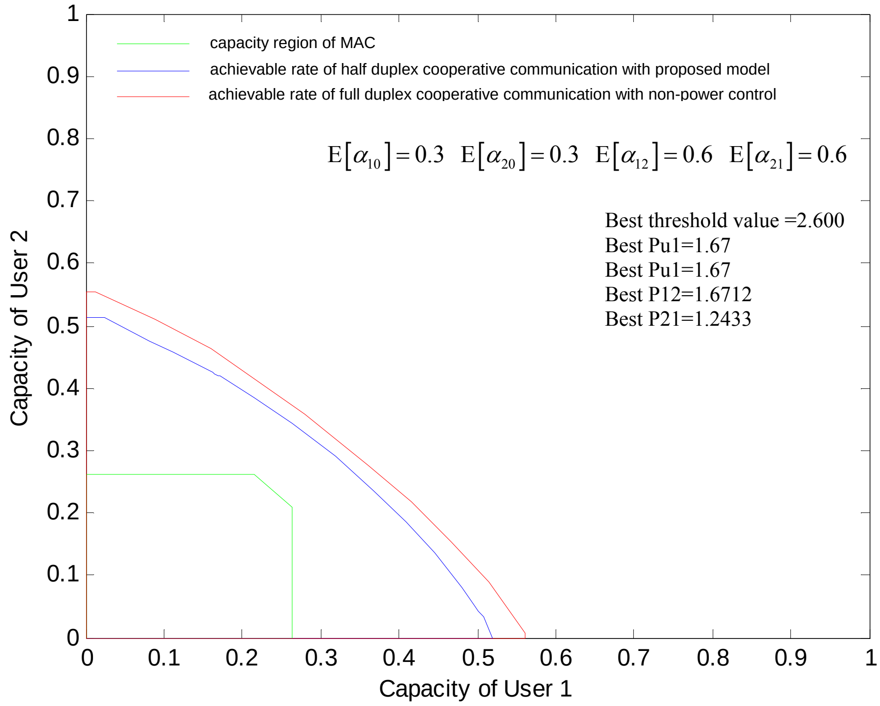

Figure 4.6: Illustration of proposed model with different cooperative channel means

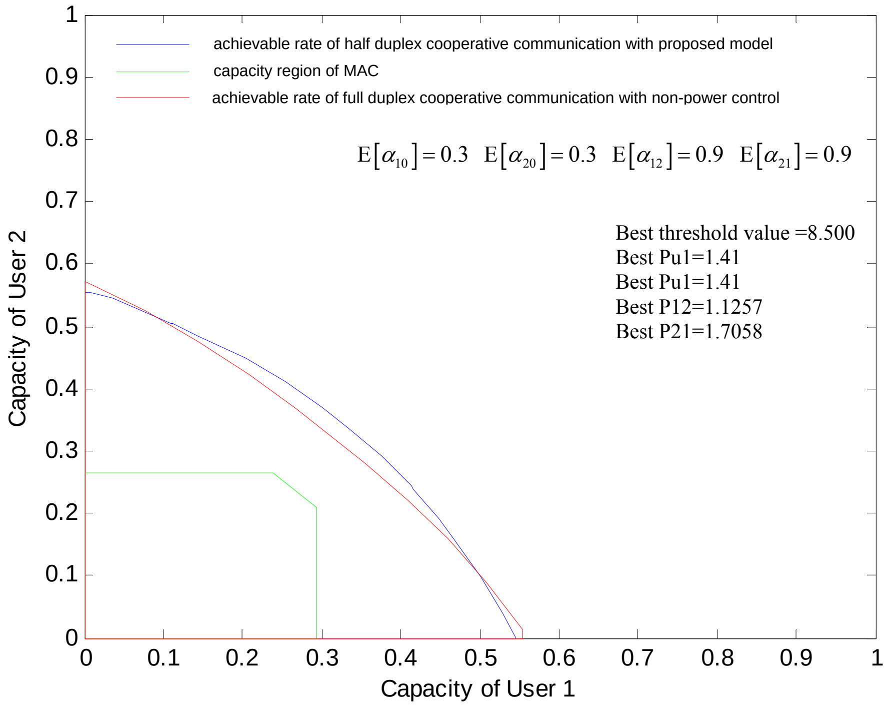

Figure 4.7: Illustration of proposed model with different cooperative channel means

Chapter 5

Conclusion and Future Work

The cooperative communication provides the benefit of sharing information for two or more users in absence of good communication channel situation. Thus, in this dissertation we aimed to create and illustrate a half duplex cooperative scheme with adaptive power control to reduce the complexity according to full-duplex scheme. First of all, we introduced the background materials which are related to cooperative communication such as main multi user diversity techniques. Then, the capacity expressions in different channel models and cooperative communication are explained. The main contribution of this dissertation was the development of a channel adaptive half duplex cooperative communication scheme by imposing some half duplex constraints on the powers, which were then used to simplify the achievable rate expression. When compared with other schemes such as full-duplex cooperation and traditional MAC, our proposed half-duplex model provides close performance to (sometimes even better than) the full-duplex non channel adaptive scheme. The main reason of this good performance is that our proposed adaptive power control algorithm makes use of the information about the relative qualities of the channel states.

Generally, the cooperative communication scheme generates many areas of research. Especially, there is need to propose investigate cooperative communication with more than two users by using optimal power control scheme. Also, the problem of obtaining an optimal solution forth half-duplex scheme with power control remains open.

APPENDIX A

MATLAB CODE

```
%_____________________________________________________________
%            CHANNEL ADAPTIVE USER COOPERATIVE STRATEGIES
%                          FOR FADING WIRELESS CHANNELS
%                                        ISIK UNIVERSITY
%               GRADUATE SCHOOL SCIENCE AND ENGINEERING
%                    ELECTRONICS ENGINEERING DEPARTMENT
%                                      by ÇAĞATAY EDEMEN
%                                                   2006
%_____________________________________________________________

samplenumber=10000;
quantization=10;
mean10=.3;
mean20=.3;
mean12=.6;
mean21=.6;
noisevariance0=1;
noisevariance1=1;
noisevariance2=1;
s1=1;
s2=1;
s0=1;
continuousrayleigh10=random('exp',mean10,1,samplenumber);
continuousrayleigh20=random('exp',mean20,1,samplenumber);
continuousrayleigh12=random('exp',mean12,1,samplenumber);
continuousrayleigh21=random('exp',mean21,1,samplenumber);

[prbbltyofchnn10,c_att10]=hist(continuousrayleigh10,quantization);
[prbbltyofchnn20,c_att20]=hist(continuousrayleigh20,quantization);
[prbbltyofchnn12,c_att12]=hist(continuousrayleigh12,quantization);
[prbbltyofchnn21,c_att21]=hist(continuousrayleigh21,quantization);

[q10,h10]=hist(continuousrayleigh10,quantization);
[q20,h20]=hist(continuousrayleigh20,quantization);
[q12,h12]=hist(continuousrayleigh12,quantization);
[q21,h21]=hist(continuousrayleigh21,quantization);

prbbltyofchnn10=prbbltyofchnn10./samplenumber;
prbbltyofchnn20=prbbltyofchnn20./samplenumber;
```

```
prbbltyofchnn12=prbbltyofchnn12./samplenumber;
prbbltyofchnn21=prbbltyofchnn21./samplenumber;

q10=prbbltyofchnn10;
q20=prbbltyofchnn20;
q12=prbbltyofchnn12;
q21=prbbltyofchnn21;

channelattenuation10=kron(c_att10,ones(1,quantization));
channelattenuation12=kron(c_att12,ones(1,quantization));
channelattenuation20=kron(ones(1,quantization),c_att10);
channelattenuation21=kron(ones(1,quantization),c_att21);

probabilityofchannel10=kron(prbbltyofchnn10,ones(1,quantization));
probabilityofchannel12=kron(prbbltyofchnn12,ones(1,quantization));
probabilityofchannel20=kron(ones(1,quantization),prbbltyofchnn20);
probabilityofchannel21=kron(ones(1,quantization),prbbltyofchnn21);

channelattenuation12longvector=kron(channelattenuation12,ones(1,…
quantization*quantization));
channelattenuation21longvector=kron(channelattenuation21,ones(1,…
quantization*quantization));
channelattenuation10longvector=kron(ones(1,quantization*quantization),…
channelattenuation10);
channelattenuation20longvector=kron(ones(1,quantization*quantization),…
channelattenuation20);

probabilityofchannel12longvector=kron(probabilityofchannel12,ones(1,…
quantization*quantization));
probabilityofchannel21longvector=kron(probabilityofchannel21,ones(1,…
quantization*quantization));
probabilityofchannel10longvector=kron(ones(1,quantization*quantization),…
probabilityofchannel10);
probabilityofchannel20longvector=kron(ones(1,quantization*quantization),…
probabilityofchannel21);

h10long=kron(h10,ones(1,10));
h20long=kron(ones(1,10),h20);

q10long=kron(q10,ones(1,10));
q20long=kron(ones(1,10),q20);
q21long=kron(ones(1,10),q21);
q12long=kron(q12,ones(1,10));

Ptotal1=1;
Ptotal2=1;
```

```
%________________________________GAMMA SEARCH ALGORITHM

kk=0;
gammaindexx=1;
step=.1;
maxi=8;

index=(channelattenuation12longvector.*channelattenuation21longvector)./(…
channelattenuation10longvector.*channelattenuation20longvector);
pjoint=probabilityofchannel21longvector.*probabilityofchannel12longvector.*
probabilityofchannel10longvector.*probabilityofchannel20longvector;
Pu1matrix=zeros(11,11,((maxi-gammaindexx)/step)+1);
Pu2matrix=zeros(11,11,((maxi-gammaindexx)/step)+1);
Rsumhalf=zeros(11,11,((maxi-gammaindexx)/step)+1);
bestgammamatrix=zeros(11,11,((maxi-gammaindexx)/step)+1);
h = waitbar(0,'Please wait...');

j=1;

for gammaindex=gammaindexx:step:maxi

   kk=kk+1;
   subindex1=find((channelattenuation12longvector>…
   channelattenuation21longvector)&(index>gammaindex));
   subindex2=find((channelattenuation21longvector>…
   channelattenuation12longvector)&(index>gammaindex));

   subindex3=find(index<=gammaindex);

   probabilityofcase1=sum(pjoint(subindex1));
   probabilityofcase2=sum(pjoint(subindex2));
   probabilityofcase3=sum(pjoint(subindex3));

   Pu1max=Ptotal1/probabilityofcase3;
   Pu2max=Ptotal2/probabilityofcase3;

   ii=0;

   for P_u1=0:Pu1max/10:Pu1max
      ii=ii+1;
      jj=0;
      for P_u2=0:Pu2max/10:Pu2max
         jj=jj+1;

         P_12=(Ptotal1-P_u1*probabilityofcase3)/probabilityofcase1;
         P_21=(Ptotal2-P_u2*probabilityofcase3)/probabilityofcase2;
```

```
%___________________________________________ case1 P12>0
%________________________________________P21=Pu1=Pu2=0
%_______________________________________ P10=0 and P20=0

R12case1=log(1+(channelattenuation12longvector(subindex1)*…
P_12./noisevariance2))*pjoint(subindex1)';
R21case1=0;
R10case1=0;
R20case1=0;
R10R20case1=0;
R1R2case1=log(1+(channelattenuation10longvector(subindex1)*…
P_12./noisevariance0))*pjoint(subindex1)';

%___________________________________________ case2 P21>0
%________________________________________P12=Pu1=Pu2=0
%_______________________________________ P10=0 and P20=0

R12case2=0;
R21case2=log(1+(channelattenuation21longvector(subindex2)*…
P_21./noisevariance1))*pjoint(subindex2)';
R10case2=0;
R20case2=0;
R10R20case2=0;
R1R2case2=log(1+(channelattenuation20longvector(subindex2)*…
P_21./noisevariance0))*pjoint(subindex2)';

%__________________________________ case3 Pu1>0 and Pu2>0
%____________________________________________ P12=P21=0
%_______________________________________ P10=0 and P20=0

R12case3=0;
R21case3=0;
R10case3=0;
R20case3=0;
R10R20case3=0;
R1R2case3=log(1+(channelattenuation10longvector(subindex3)*…
P_u1+channelattenuation20longvector(subindex3)*P_u2+2*…
sqrt(channelattenuation10longvector(subindex3).*…
channelattenuation20longvector(subindex3)*P_u1*P_u2))./…
noisevariance0)*pjoint(subindex3)';
Rsumhalf(jj,ii,kk)=min(R1R2case1+R1R2case2+R1R2case3,…
R12case1+R21case2);
Pu1matrix(jj,ii,kk)=P_u1;
Pu2matrix(jj,ii,kk)=P_u2;
bestgammamatrix(jj,ii,kk)=gammaindex;
```

```
        R1half=min(R12case1+R12case2+R12case3+R10case1+R10case2+…
        R10case3,Rsumhalf(jj,ii,kk));
        R2half=min(R21case1+R21case2+R21case3+R20case1+R20case2+…
        R20case3,Rsumhalf(jj,ii,kk));
        C10half=[R1half,0];
        C1half=[R1half,Rsumhalf(jj,ii,kk)-R1half];
        C2half=[Rsumhalf(jj,ii,kk)-R2half,R2half];
        C20half=[0,R2half];
        Chalf(4*j-2:4*j+1,:)=[C1half;C2half;C10half;C20half];
        j=j+1;

      end
   end

waitbar((kk*(100/((maxi-gammaindexx)/step)))/100)
end

% ________________________________________end of GAMMA SEARCH

disp('***best gamma search is completed...')

%__________________________________________________________MAC
Ptotal1=1;
Ptotal2=1;
j=1;
for P1=0:0.1:Ptotal1
   for P2=0:0.1:Ptotal2

      Rmac=(log(1+channelattenuation10longvector.*P1+...
      channelattenuation20longvector.*P2)./noisevariance0)*pjoint';

      R1mac=min(((log(1+channelattenuation10longvector.*P1)./noisevariance
      0)*pjoint'),Rmac);
      R2mac=min(((log(1+channelattenuation20longvector.*P2)./noisevariance
      0)*pjoint'),Rmac);

      C10mac=[R1mac,0];
      C1mac=[R1mac,Rmac-R1mac];
      C2mac=[Rmac-R2mac,R2mac];
      C20mac=[0,R2mac];
      Cmac(4*j-2:4*j+1,:)=[C1mac;C2mac;C10mac;C20mac];

   end
end
```

```
%_______________________________________________________end of MAC
disp('***MAC capacity computation is completed...')

%_____________________________FULL DUPLEX NO POWER CONTROL

p1=1;
p2=1;
C=zeros(256000,2);
j=1;
for p10=0:0.1:p1;
   for p12=0:0.1:p1-p10
      for p20=0:0.1:p2
         for p21=0:0.1:p2-p20

            pu1=p1-p10-p12;
            pu2=p2-p20-p21;
            R12=log(1+h12*p12./(h12*p10+s2))*q12';
            R21=log(1+h21*p21./(h21*p20+s1))*q21';
            R10=log(1+h10*p10./s0)*q10';
            R20=log(1+h20*p20./s0)*q20';

            R10R20=log(1+(h10long*p10+h20long*p20)./s0)*(q10long.*…
            q20long)';
            R1R2=log(1+(h10long*p1+h20long*p2+2*sqrt(h10long.*…
            h20long*pu1*pu2))./s0)*(q10long.*q20long)';

            Rsum=min(R1R2,R12+R21+R10R20);
            R1= min(R12+R10,Rsum);
            R2= min(R21+R20,Rsum);
            C10=[R1,0];
            C1=[R1,Rsum-R1];
            C2=[Rsum-R2,R2];
            C20=[0,R2];
            C(4*j-2:4*j+1,:)=[C1;C2;C10;C20];
            j=j+1;

         end;
      end;
   end;
end;

%________________________end of FULL DUPLEX NO POWER CONTROL


Chalf=real(Chalf);
Cmac=real(Cmac);
```

```
yhalf=Chalf(:,1);
xhalf=Chalf(:,2);
ymac=Cmac(:,1);
xmac=Cmac(:,2);
Khalf=convhull(xhalf,yhalf);
Kmac=convhull(xmac,ymac);
C=real(C);
y=C(:,1);
x=C(:,2);
K=convhull(x,y);
plot(xhalf(Khalf),yhalf(Khalf),'b-')
axis([0 1 0 1])
hold on
plot(xmac(Kmac),ymac(Kmac),'g-')
plot(x(K),y(K),'r-')
close(h)

disp(['-----gamma starts from ' int2str(gammaindexx) ' to ' int2str(maxi) ' and…
the best gamma is:'])
bestgammaindex=find(Rsumhalf==max(max(max(Rsumhalf))))
bestgamma=bestgammamatrix(bestgammaindex)

subindex1=find((channelattenuation12longvector>channelattenuation21…
longvector)&(index>bestgamma));
subindex2=find((channelattenuation21longvector>channelattenuation12…
longvector)&(index>bestgamma));
subindex3=find(index<=bestgamma);

probabilityofcase1=sum(pjoint(subindex1));
probabilityofcase2=sum(pjoint(subindex2));
probabilityofcase3=sum(pjoint(subindex3));
bestPu1=Pu1matrix(bestgammaindex)
bestPu2=Pu2matrix(bestgammaindex)

P_12=(Ptotal1-bestPu1*probabilityofcase3)/probabilityofcase1
P_21=(Ptotal2-bestPu2*probabilityofcase3)/probabilityofcase2

disp('-----check power constraint')
disp('-----P12.*probabilityofcase1+bestPu1.*probabilityofcase3')
expactation1=P_12.*probabilityofcase1+bestPu1.*probabilityofcase3
disp('-----P21.*probabilityofcase2+bestPu2.*probabilityofcase3')
expactation2=P_21.*probabilityofcase2+bestPu2.*probabilityofcase3

%_____________________________________________end of SIMULATION
```

Bibliography

[1] T. M. Cover and C. S. K. Leung. *An Achievable Rate Region for the Multiple Access Channel with Feedback*. IEEE Trans. on Information Theory, 27(3):292-298 • May 1981.

[2] A. Sendonaris, E. Erkip and B. Aazhang. *User Cooperation Diversity Part II: System Description. IEEE Trans. On Communications*, 51(11): 1939-1948 • November 2003.

[3] F. M. J. Willems and E. C. van der Meulen. *Partial Feedback for the Discrete Memoryless Multiple Access Channel*. IEEE Trans. on Information Theory, 29(2):287-290 • March 1983.

[4] F. M. J. Willems. *The Discrete Memoryless Multiple Access Channel with Partially Cooperating Encoders*. IEEE Trans. on Information Theory, 29(3):441-445 • May 1983.

[5] F. M. J. Willems and E. C. van der Meulen. *The Discrete Memoryless Multiple Access Channel with Cribbing Encoders*. IEEE Trans. on Information Theory, 31(3):313-327 • May 1985.

[6] F. M. J. Willems, E. C. van der Meulen and J. P. M. Schalkwijk. *An Achievable Rate Region for the Multiple Access Channel with Generalized Feedback*. In Proc. Allerton Conference, Monticello, IL • October 1983.

[7] A. Sendonaris, E. Erkip and B. Aazhang. *User Cooperation Diversity – Part I: System Description*. IEEE Trans. on Communications, 51(11): 1927–1938 • November 2003.

[8] Thomas M. Cover, Joy A. Thomas. *Elements of Information Theory*. Wiley Series in Communication • 1991.

[9] A. Host-Madsen and J. Zhang. *Capacity Bounds and Power Allocation for Wireless Relay Channels*. IEEE Trans. on Information Theory, 51(6):(2020–2040) • June 2005.

[10] E. Erkip. *Capacity and Power Control for Spatial Diversity*. In Proc. Conference on Information Sciences and Systems, pp. WA4-28/WA4-31, Princeton, NJ • March 2000.

[11] A. J. Goldsmith and P. P. Varaiya. *Capacity of Fading Channels with Channel Side Information*. IEEE Trans. on Information Theory, 43(6):1986–1992 • November 1997.

[12] R. Knopp and P. A. Humblet. *Information Capacity and Power Control in Single-cell Multiuser Communications*. In Proc. IEEE ICC • June 1995.

[13] Theodore S. Rappaport. *Wireless Communications Principle & Practice*. Prentice Hall PTR • July 1999

[14] E. Biglieri, J. Proakis and S. Shamai (Shitz). *Fading Channels: Information-theoretic and Communications Aspects*. IEEE Trans. on Information Theory, 44(6):2619–2692 • October 1998.

[15] Elza Erkip, Andrew Sendonaris, Andrej Stefanov, Behnaam Aazhang. *Cooperative Communication in Wireless Systems*. DIMACS Series in Discrete Mathematics and Theoretical Computer Science

[16] D. P. Bertsekas. *Nonlinear Programming*. Athena Scientific • 1995.

[17] John G. Proakis. *Digital Communication*. McGraw-Hill International Edition, 4th Edition • 2001.

[18] E.C. Van Der Meulen. *Three-terminal communication channels*. Advances in Applied Probability, 3:120–154 • 1971.

[19] S. Hanly and D.N.C. Tse. *Multiaccess Fading Channels – Part I: Polymatroid Structure, Optimal Resource Allocation and Throughput Capacities*. IEEE Trans. on Information Theory, 44(7):2796–2815 • November 1998.

[20] Todd E. Hunter, Ahmadreza Hedayat. *Cooperative Communication in Wireless Networks*, IEEE Communications Magazine:74-80 • October 2004

[21] T.M. Cover and A. El Gamal. *Capacity theorems for the relay channel*. IEEE Transactionson Information Theory, vol. 25, pp. 572–584 • September 1979.

[22] J. N. Laneman, D. N. C. Tse and G. W. Wornell. *Cooperative Diversity in Wireless Networks: Efficient Protocols and Outage Behavior*. IEEE Trans. on Information Theory, 50(12): 3062–3080 • December 2004.

[23] Xiaodong Wang, H. Vincent Poor. *Wireless Communication Systems Advanced Techniques for Signal Reception*. Prentice Hall Communications Engineering and Emerging Technologies Series Theodore S. Rappaport, Series Editor • 2002

[24] Onur Kaya, Sennur Ulukus. Fading Cooperative Multiple Access Channels, preprint • July 2006