อัลกอริทึมที่เหมาะที่สุดสำหรับการทำให้ค่ากำลังส่งมีค่าน้อยที่สุดบนพื้นฐานของข่าวสารสถานะช่องสัญญาณบางส่วนในระบบหลายทางเข้าหลายทางออก

นายฟาม ดึน เติง

วิทยานิพนธ์นี้เป็นส่วนหนึ่งของการศึกษาตามหลักสูตรปริญญาวิศวกรรมศาสตรมหาบัณฑิต
สาขาวิชาวิศวกรรมไฟฟ้า ภาควิชาวิศวกรรมไฟฟ้า
คณะวิศวกรรมศาสตร์ จุฬาลงกรณ์มหาวิทยาลัย
ปีการศึกษา 2548

ISBN 974-17-5618-6

OPTIMIZATION ALGORITHMS FOR TRANSMIT POWER MINIMIZATION BASED ON PARTIAL CHANNEL STATE INFORMATION IN MIMO SYSTEMS

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A Thesis Submitted in Partial Fulfillment of the Requirements
for the Degree of Master of Engineering in Electrical Engineering
Department of Electrical Engineering
Faculty of Engineering
Chulalongkorn University
Academic year 2005
ISBN 974-17-5618-6

Thesis Title	Optimization Algorithms for Transmit Power Minimization Based on Partial Channel State Information in MIMO Systems.
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ฟาม ดึน เติง : อัลกอริทึมที่เหมาะที่สุดสำหรับการทำให้ค่ากำลังส่งมีค่าน้อยที่สุดบนพื้นฐานของข่าวสารสถานะช่องสัญญาณบางส่วนในระบบหลายทางเข้าหลายทางออก (OPTIMIZATION ALGORITHMS FOR TRANSMIT POWER MINIMIZATION BASED ON PARTIAL CHANNEL STATE INFORMATION IN MIMO SYSTEMS) อ. ที่ปรึกษา : รศ. ดร. สมชาย จิตะพันธ์กุล,59 หน้า. ISBN 974-17-5618-6.

ถึงแม้ว่า สเปกตรัมความถี่จะมีไม่จำกัด แต่มีเพียงช่วงบางสเปกตรัมความถี่เท่านั้นที่สามารถใช้ในการสื่อสารไร้สายด้วยเทคโนโลยีในปัจจุบันได้ การสื่อสารไร้สายกำลังวิวัฒนาการเข้าสู่ยุคที่ 4 การใช้งานความถี่กลายเป็นเรื่องท้าทายอย่างมาก เนื่องจากมีความต้องการอัตราข้อมูลสูง สเปกตรัมความถี่สำหรับระบบไร้สายใหม่เหลือน้อยและราคาสูง ดังนั้นจึงมีความต้องการที่จะทำให้อัตราการส่งสูงที่สุดภายใต้สเปกตรัมความถี่จำกัด ซึ่งความต้องการนี้จะไม่ประสบความสำเร็จหากปราศจากการเปลี่ยนแปลงประสิทธิภาพในการใช้สเปกตรัมความถี่

การส่งข้อมูลผ่านช่องสัญญาณหลายสัญญาณเข้าหลายสัญญาณออก (MIMO) เป็นแนวทางที่มีแนวโน้มที่ดี สายอากาศจำนวนมากถูกใช้ทั้งที่เครื่องส่งและเครื่องรับ โดยใช้สเปกตรัมความถี่เดียวกัน

วิทยานิพนธ์ฉบับนี้พิจารณา ระบบ Orthogonal Frequency Division Multiplexing ร่วมกับการส่งแบบ MIMOโดยตั้งสมมุติฐานว่าเครื่องส่งทราบข้อมูลสถานะช่องสัญญาณบางส่วน การออกแบบเครื่องส่งด้วย ตัวเข้ารหัส-ตัวกำหนดทิศทางลำคลื่น ถูกประยุกต์ใช้เพื่อเพิ่มความทนทานของระบบต่อความไม่แน่นอนของค่าทางสถิติของข้อมูลสถานะช่องสัญญาณ เนื่องจากการจัดวางบิทแบบเหมาะสมที่สุดมีความซับซ้อนสูง วิทยานิพนธ์ฉบับนี้จึงเสนอระเบียบวิธีใหม่สำหรับลดกำลังงานส่งให้มากที่สุด สำหรับระบบ MIMO-OFDM วิธีการที่นำเสนอใหม่มีสมรรถนะที่ดีกว่าและมีความซับซ้อนต่ำกว่าวิธีการที่มีอยู่ในปัจจุบัน เมื่อจำนวนบิตเฉลี่ยต่อช่องสัญญาณย่อยน้อยเมื่อเทียบกับขนาดการมอดูเลตบิต

ภาควิชา วิศวกรรมไฟฟ้า ลายมือชื่อนิสิต....................

สาขาวิชา วิศวกรรมไฟฟ้า ลายมือชื่ออาจารย์ที่ปรึกษา....................

ปีการศึกษา 2548

4670655221 : MAJOR ELECTRICAL ENGINEERING
KEY WORD: MULTIPLE-INPUT MULTIPLE-OUTPUT (MIMO) / WIRELESS LOCAL AREA NETWORKS (WLAN) / ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING (OFDM) / ADAPTIVE MODULATION

PHAM DINH TAN : OPTIMIZATION ALGORITHMS FOR TRANSMIT POWER MINIMIZATION BASED ON PARTIAL CHANNEL STATE INFORMATION IN MIMO SYSTEMS. THESIS ADVISOR: ASSOC. PROF. SOMCHAI JITAPUNKUL, Dr.Ing., 59 pp. ISBN 974-17-5618-6.

Although frequency spectrum is infinite, only a very limited range of spectrum can be utilized for wireless communications with current technology. As wireless communications evolve to the Fourth Generation (4G), frequency utilization becomes a challenge since much higher data rate is expected. Frequency spectrum for new wireless systems is scarce and expensive. It is desired to maximize the transmission rate within limited bandwidth. This demand can not be met without a significant change in spectral efficiency.

Transmission over Multiple-Input Multiple-Output (MIMO) channel is the most promising solution. Multiple antennas are employed at both transmitter and receiver, using the same frequency spectrum.

In this thesis, Orthogonal Frequency Division Multiplexing (OFDM) systems in combination with MIMO transmission are considered. Only partial channel state information (CSI) at transmitter is assumed. The transmitter design with coder-beamformer is adopted to increase system robustness against uncertainties of statistical CSI. Since optimal bit allocation is of high complexity, this thesis proposes a novel algorithm for minimizing transmit power in MIMO-OFDM systems. The proposed scheme has better performance and lower complexity than existing algorithm when average number of bits per sub-channel is small with respect to constellation size.

Department Electrical Engineering Student's signature..................................

Field of study Electrical Engineering Advisor's signature..................................

Academic year ....2005...................

Acknowledgment

First, I would like to thank my advisor, Dr. Somchai Jitapunkul, for his guidance and encouragement during the time I studied in Thailand. I would like to thank Dr. Watit Benjapolakul, Dr. Chedsada Chinrungrueng and Dr. Nisachon Tangsangiumvisai for accepting the invitation to join my thesis committee. I would like to thank Dr. Prasit Prapinmongkolkarn and Dr. Chedsada Chinrungrueng for giving lectures in classes with only a few students.

I am grateful to Dr. Suwich Kunaruttanapruk for his help during the past two years. Thank also comes to Wilaiporn Lee, Jatupon Pattanavichate, Manil Cotonetr, Kitty Tongpraduppet and many others.

Finally, I would like to thank AUN/SEED-Net for financial support. I do not know how to express my acknowledgment to Chulalongkorn University, but simply, I am proud to be a student of Chulalongkorn University.

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List of Symbols

n_T	number of transmit antennas
n_R	number of receive antennas
J	minimum number of transmit antennas and receive antennas
H	channel matrix
h_{ij}	gain between the j^{th} transmit antenna to the i^{th} receive antenna
r	rank of channel matrix
N	number of OFDM tones
$\sqrt{\lambda_m^n}$	gain of the m^{th} spatial subchannel at the n^{th} OFDM tone
$(\cdot)^T$	matrix transpose
$(\cdot)^H$	Hermitian matrix transpose

Chapter I

Introduction

1.1 Background and Signification of the Research Problem

1.1.1 Wireless Local Area Networks

The wireless local area network (LAN) industry has been expanding continuously into business, finance, education, manufacturing, and education. WLAN allows people to access Internet, multimedia services with low mobility such as in houses, office buildings, airports, factories, etc. With the standardization of new series of IEEE 802.11 and HIPERLAN, wireless LAN (WLAN) is evolving into high-speed data transmission supporting both packet- and connection-oriented voice, Quality of Service. These new trends of WLAN technology will certainly have rapid development of markets.

All functionalities of a conventional fixed LAN are available in a WLAN including file sharing, peripheral sharing, Internet access, etc. The WLAN can be such implemented to replace or to extend the capability of the LAN by providing mobility. Compared to the fixed LAN, the main advantages and benefits of wireless networks are the mobility and cost-saving installation. Most of the application scenarios of WLAN are related to these features. Wireless networking provides significant cost savings in the areas where cables cannot be easily installed. Wireless networking is applicable to all situations where mobile computer usage is needed and/or the cable installation is not feasible. For example, the office need not to be rewired each time there is a change.

WLAN standard 802.11 was standardized by IEEE in 1997, for operating in the unlicensed 2.4 GHz band with data rate of 1-2 Mbps. Frequency band around 2.4 GHz is referred to as the Industrial, Scientific and Medical (ISM) band. This band is regulated in different ways in Europe, Japan, and the United States. The availability of unlicensed radio spectrums at 2.4 GHz ISM band has prompted the wireless industry to develop broadband wireless data communication systems. More and more technologies and applications are developed for expanding the market of Wireless Local Area Networks (WLAN). By exploring spread spectrum technology at 2.4 GHz band, transmission rate is pushed up to 11 Mbps in 1999 with 802.11b. The standard 802.11b is currently occupies a dominant market share. Transmission rate of 54 Mbps is supported by 802.11g (backward compatible with 802.11b) with OFDM at 2.4 GHz band [1].

However, microwave ovens, which appear in most buildings, also operate at 2.4 GHz band. This causes interference to WLAN systems. The standard 802.11a, based on Orthogonal Frequency Division Multiplexing (OFDM), is a natural move from 2.4 GHz to the cleaner 5 GHz band with transmission rate of up to 54 Mbps. However, operational range is reduced at this frequency. It means that more access points are required and overall cost increases. However, 802.11a standard is only popular in North America since HYPERLAN, a standard which is similar but not compatible to 802.11a, is adopted in Europe. Although the development of WLAN market has been held back due to issues such as high price and relatively low throughput (especially when compared to the wired LANs). Most recently with the standardization of HIPERLAN/2 and IEEE 802.11a and 802.11b, a new fast development of market is being expected. The standard 802.11a is based on OFDM (Orthogonal Frequency Division Multiplexing) modulation scheme that was selected for its robustness against frequency selective fading and narrowband interference. The specifications of the Physical Layer encompass data rates from 6 Mbps up to 54 Mbps with 20 MHz spacing between adjacent channels.

The IEEE 802.11 standard for WLAN Medium Access (MAC) and Physical Layer (PHY) specifications defines over-the-air protocols necessary to support local area networking. The 802.11 standard provides MAC and PHY functionality for wireless connectivity of fixed, portable, and moving stations at pedestrian and vehicular speeds with a local area. The IEEE 802.11 standard provides two physical layer specifications for radio frequency (RF), operating in 2.4 GHz ISM band, and one for infrared. For both Frequency Hopping and Direct Sequence Spread Spectrum physical layers, two different data rates are specified, 1 Mbps and optional 2 Mbps.

The basic access method of 802.11 MAC is a scheme called Carrier Sense Multiple Access with Collision Avoidance (CSMA/CA). Before transmitting, a station senses the channel. When the channel is idle, the packet is transmitted right away. If the channel is busy, the stations keep sensing the channel until it is idle, then waits a uniformly distributed random backoff period before sensing the channel again. If the channel is still idle it transmits its packet, otherwise it backs off again. The backoff mechanism results in the avoidance of the collision of packets from multiple transmitters who all sense a clear channel at about the same time. All directed traffic receives a positive acknowledgement and packets are retransmitted if an acknowledgement is not received. WLAN standards consist of specifications on physical (PHY) layer and medium access control (MAC) layer. PHY layer can be selected from

- *Frequency hopping (FH) spread spectrum;*
- *Direct sequence (DS) spread spectrum;*
- *Infrared (IR).*

The protocol architecture is shown in Figure 1.1.

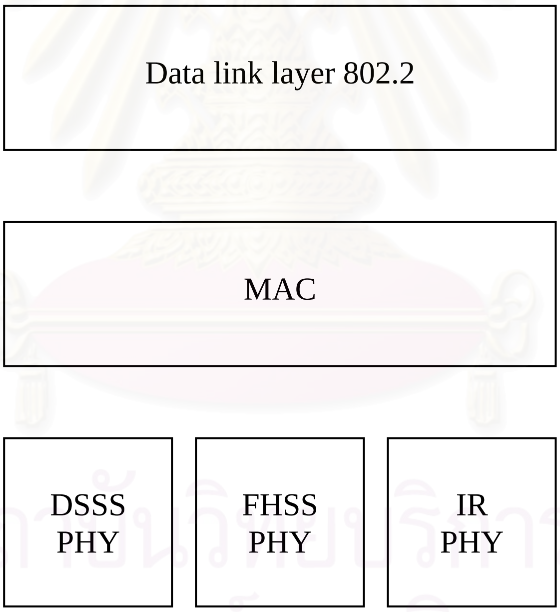

Figure 1.1: WLAN IEEE802.11 Protocol.

Components of WLAN systems can be classified into

- *Mobile Terminal (MT): components with low-mobility;*
- *Access Point (AP);*
- *Portal (PO): interconnnect between wireless LAN and wired LAN.*

Each AP serves a coverage area called a basic service set (BSS). Multiple BSSs form an extended service set (ESS). The inter access-point protocol (IAPP) is used for communicating between different APs in an ESS for handoff related purposes.

The simplest BSS is constituted by two MTs that can communicate directly. The stations can communicate directly with each other in ad hoc networks. This mode of operation is often refered to as an ad hoc network because this type of IEEE 802.11 WLAN is typically created and maintained as needed without prior arrangement for specific purposes (such as transferring a file from one personal computer to another). This basic type of IEEE 802.11 WLAN is called independent BSS (IBSS). There is usually no connection to the wired network.

The second type of BSS is an infrastructure BSS. This configuration is also known as access point (AP) based networks. Within an infrastructure BSS, an AP operates as the coordinator of the BSS. The mobile terminals (MT) communicate directly with an AP that is connected to the wired network. Instead of existing independently, two or more BSS can be connected together through some kind of backbone network that is called the distribution system (DS). The whole interconnected WLAN (some BSSs and a DS) is identified by the IEEE 802.11, as a single wireless network called extended service set (ESS).

It is important to remark that IEEE 802.11 does not impose any constraint on the DS (for example, it does not specify if the DS should be data link layer–based or network-layer based). Instead, IEEE 802.11 specifies a set of services that are associated with different parts of the architecture. Such services are divided into those assigned to MT, called station service (SS) and to the DS, called distribution system service (DSS). Both categories of services are used by the IEEE 802.11 MAC sublayer. The services assigned to the MT are:

- *Authentication/deauthentication;*
- *Privacy;*
- *MAC service data unit (MSDU) delivery to upper layer.*

The services supported by the DS are:

- *Association/disassociation;*

- *Distribution;*
- *Integration;*
- *Reassociation.*

The SSs are provided by all stations, including AP, conforming with IEEE 802.11, while the DSSs are provided by the DS. Both MAC and PHY layers include two management entities: MAC sublayer management entity (MLME) and PHY layer management entity (PLME).

The PHY is divided into two sublayers. The first layer is the physical medium dependent sublayer (PMD), which carries out the modulation and the encoding. The second layer is the Physical Layer Convergence Protocol (PLCP), which carries out PHY-specific functions, supporting common PHY SAP and providing a clear channel assessment signal.

The following services are supported by the MAC layers:

- *Asynchronous data service, which provides peer IEEE 802.2 entities with the ability to exchange MSDUs;*
- *Security services, which in IEEE 802.11 are provided by the authentication service and the wired-equivalent privacy (WEP) mechanism;*
- *MSDU ordering, whose sole effect, for the set of MSDUs received at the MAC service interface of any single station, is a change in the delivery order of broadcast and multicast MSDUs, relative to directed MSDUs, originating from a single source station address*

MAC frame format consists of the following components:

- *A MAC header, which comprises frame control, duration, address,*
- *and sequence control information;*
- *A variable length frame body, which contains information specific*
- *to the frame type;*
- *A frame check sequence (FCS), which contains an IEEE 32-bit CRC.*

Most equipment available in the market permits the choice of one channel or another through a configuration menu provided by the manufacturers for this purpose. The maximum values of power depend on each regulating organism. In the United States, the Federal Communications Commission (FCC) has fixed the limit as 1W; in Europe, the ETSI has established it as 100 mW, equivalent isotropic radiated power (EIRP); in Japan, it is 10 mW/MHz. Due to the lack of a coordinating body governing the use made by various users in the neighboring area within the band, it is essential that they respect the maximum power values established with the aim of not perturbing the communications of other equipment working nearby. In fact, any RF energy sensed by the radio equipment of the WLAN that is not recognized as potentially generated by some IEEE 802.11b device working in DS spread spectrum mode is considered interference. Particularly, this includes the 802.11b signals generated by equipment working in other channels of the band. The existence of interference leads to the reception of packets with errors, which at the MAC level brings about retransmissions, which in turn brings about a fall in throughput. Given that the equipment also has a fallback mechanism at a lower speed, the presence of interference can cause the negotiation of lower bit rates (i.e., passing from 11 Mbps to 5.5, 2, or even 1 Mbps with the consequent reduction in throughput). In general, to avoid this situation, it is necessary to try to maintain a signal level of about 10–12 dB more than the noise level. Using the menu offers the user the ability to monitor at any time the conditions of the link in order to find the optimum site of the equipment.

1.1.2 Modern Communications Systems

Although frequency spectrum is infinite, only a very limited range of spectrum can be utilized for mobile/wireless communications with current transmission technology. As wireless communications evolving to the Fourth Generation (4G), frequency utilization becomes a challenge since much higher data rate, up to 1 Gb/sec for indoor WLAN applications, is required. Frequency spectrum for new wireless systems is scarce and expensive. It is desired to maximize the rate of communications within a given bandwidth. This demand can not be met without a significant change in spectral efficiency.

Transmission over Multiple-Input Multiple-Output (MIMO) channel is the most promising solution. Multiple antennas are employed at both transmitter and receiver as shown in Figure 1.2. The same frequency spectrum is used over all transmit antennas. Rich scattering environment is assumed since under that condition, signals from different transmit antennas can be separated by receive antenna array. Signal detection is performed at receiver using some sophisticated algorithms.

The success of MIMO systems comes from the ability to exploit the space dimension of wireless channel. MIMO can be viewed as an extension of the so-called smart antennas, a widely-used technique using antenna arrays at base stations to improve quality via beam-forming or antenna diversity. Research on MIMO began with theoretical work of Foschini [2] on channel capacity. Under ideal propagation, the capacity of MIMO increases linearly with number of antennas. It has been demonstrated that the Bell Laboratories Layered Space-Time (BLAST) coding technique can achieve the spectral efficiency up to 42 bits/sec/Hz. This represents a leap in spectral efficiency compared to currently achievable spectral efficiencies of 2-3 bits/sec/Hz in cellular mobile and WLAN systems [3]. As subscriber units are supporting more and more services other than just pocket telephones, the stringent size and complexity constraints are becoming more relaxed. Implementation of multiple antennas at portable units becomes feasible [4].

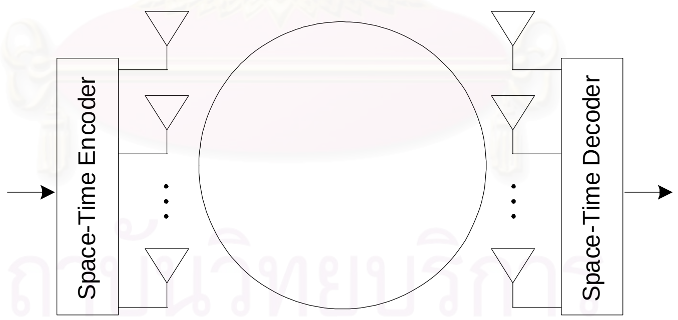

Figure 1.2: Block diagram of multiple-input multiple-output (MIMO) system.

An important characteristic of wireless channel is multipath fading. Intersymbol interference (ISI) occurs as signal propagates through various paths with different gains and delays. Due to the high data rate requirement of new broadband services, ISI becomes a serious problem. Channel equalization techniques were proposed to deal with ISI for single antenna systems. But for MIMO systems, especially with multiple antennas, equalizers are not attractive due to high cost and complexity. Multi-carrier modulation is an alternative.

Orthogonal Frequency Division Multiplexing (OFDM) is a multi-carrier modulation technique against frequency selective fading. Total bandwidth is divided into overlapping subchannels with orthogonal subcarriers. Data stream is split into multiple substreams with lower rate. Each substream is transmitted over one subchannel. OFDM is implemented efficiently with the Discrete Fourier Transform (DFT) and the Inverse Discrete Fourier Transform (IDFT). Since signals are transmitted in parallel so OFDM symbol rate is very low. Intersymbol interference (ISI) can be eliminated by adding guard time with duration larger than the maximum delay spread of fading channel. If the guard period is left blank, ICI will occur as orthogonality between subcarriers is lost. The solution is to add cyclic prefix in guard period. At receiver, the DFT of convolved signal and channel is simply the multiplication of their DFT's. OFDM was adopted in wireless standards such as European digital audio broadcasting (DAB) and digital video broadcasting (DVB), IEEE broadband wireless local area networks (WLAN) IEEE802.11g and European HIPERLAN [3].

Recently, the combination of MIMO and OFDM for WLAN has been of great interest since it offers very high spectral efficiency, with acceptable complexity. For indoor WLAN channel, delay spread and Doppler spread are small. Typical value of delay spread is in the range of 10-50 ns, while Doppler spread lies in the range of 0.3-6.1Hz [5]. This leads to large values of coherence bandwidth and coherence time. So it is feasible to exploit spectral efficiency of both MIMO and OFDM in a combined system to maximize data rate for indoor WLAN. At the time of writing this thesis, OFDM systems with MIMO transmission are being standardized for WLAN. The standard IEEE802.11n is expected to appear in the year 2007.

1.2 Literature Review

In literature on MIMO, optimization algorithms address the problem of either maximizing throughput or minimizing transmit power when assuming that channel is known at transmitter via a feedback channel. Since fixed data rate is required in some applications, this work only considers the problem of minimizing transmit power. Initially, algorithms were proposed for MIMO-OFDM systems assuming that perfect channel state information (CSI) is available at transmitter. When perfect CSI is available at transmitter, optimal transmission scheme to exploit perfect CSI is via Singular Value Decomposition (SVD).

However, perfect CSI is very difficult to obtain due to estimation error, feedback delay, quantization error, etc. Some papers proposed optimization algorithms based on partial CSI (or statistical CSI) [13],[14] with channel mean at transmitter. Adaptive modulation is performed at transmitter with a coder-beamformer to maximize transmission rate.

Channel capacity of OFDM-based Spatial Multiplexing was derived in [16]. Recently, adaptive modulation was proposed for OFDM-based spatial multiplexing systems. Performance of an OFDM-based spatial multiplexing with 4 transmit and 4 receive antennas was reported [17]. Adaptive modulation over 256 tones was considered, which achieved at a rate gain of 2 at SNR=33dB for maximum bit rate and a gain of 19dB at $BER = 2 \times 10^{-4}$ with constant bit rate. Other factors such as estimation error, limited feed-back ability and outdated estimate also make SVD technique unattractive. In [11], V-BLAST-based adaptive modulation was proposed as an alternative. Backward ordering [12] achieves a 2dB gain compared with forward ordering.

1.3. Objectives

In this work, adaptive modulation algorithm for downlink of MIMO-OFDM systems with partial CSI is proposed.

1.4. Scope

1. Downlink of MIMO-OFDM in indoor WLAN.
2. Only partial CSI is available at transmitter.
3. Optimization Algorithms with Partial Channel State Information

1.5 Organization of the thesis

In chapter 2, multiple antenna systems with spatial diversity and spatial multiplexing systems are described. Flat fading channel model is assumed. Results from this section is generalized for multiple antenna systems in frequency selective fading. Both transmit diversity and receive diversity schemes are applied to increase robustness of communication systems against uncertainties of partial CSI. Frequency selective fading MIMO channel is transformed into flat fading channels via multi-carrier modulation technique OFDM.

Materials on adaptive modulation are included in chapter 3. Adaptive modulation with different system configurations are described. When perfect CSI is assumed, optimal solution is the water-filling solution. The problem of discrete number of bits per subchannel is introduced. Since optimal bit allocation is of high complexity, different sub-optimal schemes have been proposed. We consider the Piazzo's algorithm which was initially proposed for OFDM systems. Motivated from the logarithmic search of Piazzo's algorithm, we propose the upward bit allocation algorithm (UBAA) for MIMO-OFDM systems with perfect CSI.

In chapter 4, assumption on partial CSI, or channel mean feedback, is made. Configuration MIMO-OFDM systems is described by adopting the transmitter design with coder-beamformer. Upward bit allocation algorithm is proposed to replace the Hughes-Hartogs algorithm as in [13]. Performance of the proposed scheme is evaluated by computer simulation and compared to those of optimal and existing sub-optimal scheme.

Chapter II

Multiple Antenna Systems

2.1 Introduction

In this chapter, multiple antenna systems are described with simulation results. In section 2.2, spatial diversity for improving reliability is discussed. In section 2.3, spatial multiplexing MIMO for increasing spectral efficiency is described. MIMO systems in flat fading are considered. Results are extended to the case of frequency selection fading channel with MIMO-OFDM systems.

2.2 Spatial Diversity

Diversity is essential for improving reliability, as human being with 2 ears, 2 eyes, etc for better sensing the surrounding world. This is the result of evolution process which took place in millions of years. The same principle has been applied to communication systems.

Due to the varying nature of wireless communication channels, transmitted signals are corrupted and can not be recovered at receiver. This phenomenon is generally referred to as fading. Performance of a single antenna system under Gaussian channel and flat fading channel is shown in Figure 2.1.

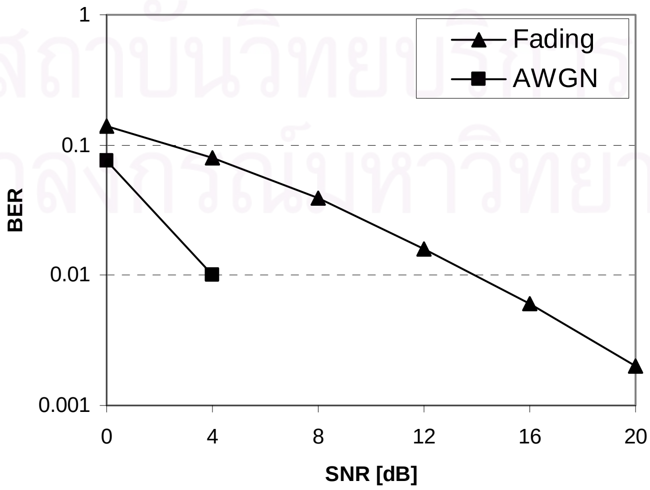

Figure 2.1: Performance loss due to fading.

With spatial diversity, performance of communication systems can be significantly improved. Antenna arrays are implemented at receivers with classical combining techniques such as maximum ratio combining (MRC), equal gain combining (EGC), and selection combining (SC). Maximum ratio combining is the optimal technique. Its performance is shown in Figure 2.2 with different antenna configuration : 2 receive antennas, 4 receive antennas.

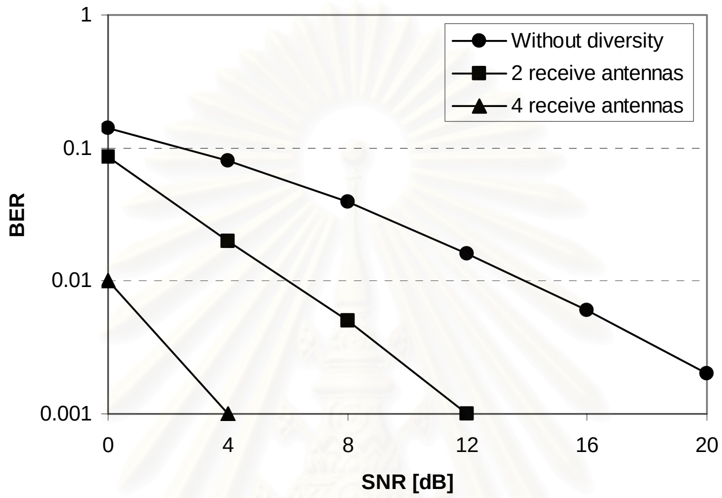

Figure 2.2: Receiver diversity.

Receive diversity achieves good performance and has low complexity. However, it is only attracted in upward link designs, where antenna arrays can be implemented at base station. For the downlink, multiple antennas can not be implemented at mobile stations to get diversity due to cost and size.

Transmit diversity appears to meet this demand under the name space-time coding. Space-time trellis codes were proposed first in [15], the classic paper on space-time coding. Framework for designing space-time code is considered thororuthly in this work. However, complexity of space-time trellis codes increases exponentially with number of transmit antennas.

Alamouti proposed an elegant coding scheme for systems with 2 transmit antennas, widely known as Alamouti code. Although performance of transmit diversity is poorer than receive diversity, transmit diversity is still of high interest due to its robustness for the downlink.

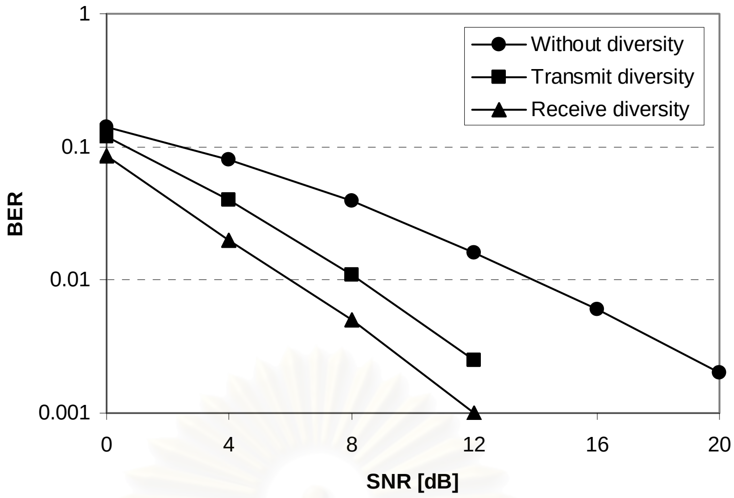

Figure 2.3: Comparison between transmit and receive diversity.

Generalization of Alamouti code to systems with any number of transmit antennas is known as space-time block codes. Complexity of space-time block code is lower than that of space-time trellis code. However, space-time block code achieves only diversity gain while space-time trellis code provides both diversity gain and coding gain. Performance of systems with different number of transmit antennas is shown in Figure 2.4.

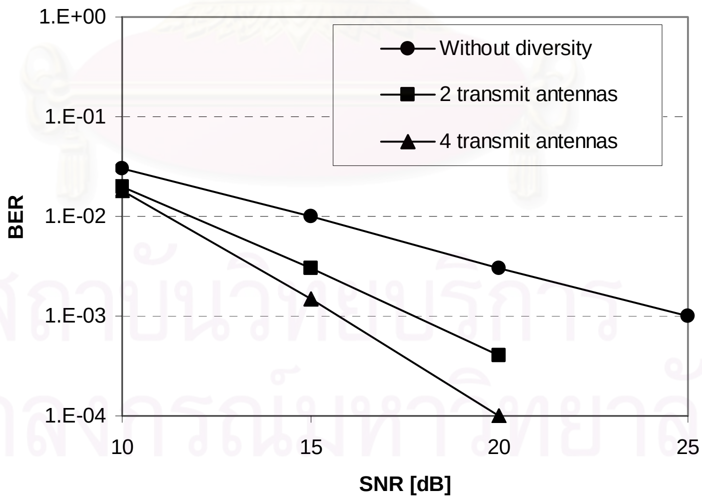

Figure 2.4: Diversity with transmit antennas.

2.3 Spatial Multiplexing

2.3.1 MIMO Capacity

Spatial Multiplexing approach aims at maximizing data rate for a given bandwidth, or in other words, maximizing spectral efficiency. MIMO channel over flat fading channel can be written as

$$v = Hu + n\,, \tag{2.1}$$

Singular Value Decomposition (SVD) theorem says that any $n_R \times n_T$ matrix **H** of rank *r* can be written as

$$H = UDV^H\,, \tag{2.2}$$

where **D** is an $n_R \times n_T$ non-negative and diagonal matrix with elements

$$d_{1,1} \geq d_{2,2} \geq \ldots \geq d_{r,r} > d_{r+1,r+1} = \ldots = d_{J,J} = 0\,, \tag{2.3}$$

U and **V** are $n_R \times n_R$ and $n_T \times n_T$ unitary matrices, respectively. That is,

$$\begin{cases} UU^H = I_{n_R} \\ VV^H = I_{n_T} \end{cases}, \tag{2.4}$$

where I_{n_R} and I_{n_T} are $n_R \times n_R$ and $n_T \times n_T$ identity matrices, respectively. The eigenvalues of $\mathbf{HH}^H$, denoted by λ, are defined as

$$HH^H y = \lambda y, \quad y \neq 0\,, \tag{2.5}$$

where **y** is an $n_R \times 1$ eigenvector associated with λ. The diagonal elements of **D** (or square roots of the eigenvalues of matrix $\mathbf{HH}^H$) are referred to as the singular values $\sqrt{\lambda_i}$ of channel matrix **H**. The columns of **U** are the eigenvectors of $\mathbf{HH}^H$ and the columns of **V** are the eigenvectors of $\mathbf{H}^H\mathbf{H}$ [3]. By substituting (2) into (1) we can write for the received vector **v** as

$$v = UDV^H u + n\,, \tag{2.6}$$

By introducing the following transformations

$$
\begin{aligned}
v' &= U^H v \\
u' &= V^H u \ , \\
n' &= U^H n
\end{aligned}
\tag{2.7}
$$

(1) can be rewritten as

$$v' = Du' + n' \tag{2.8}$$

The number of nonzero eigenvalues of matrix $\mathbf{HH}^H$ is equal to the rank r of matrix **H**. For the channel matrix H of size $n_R \times n_T$, the rank r is at most $J = \min(n_T, n_R)$, which means that at most J of its singular values are nonzero. Let us denote the singular values of **H** by $\sqrt{\lambda_i}, \quad i = 1, 2, ..., r$. By substituting the entries $\sqrt{\lambda_i}$ in (8), we obtain the received signal components

$$v_i' = \begin{cases} \sqrt{\lambda_i} u_i' + n_i', & i = 1, 2, ..., r \\ n_i', & i = r+1, r+2, ..., J \end{cases}, \tag{2.9}$$

If we only consider r spatial subchannels with non-zero eigenvalues, (9) can be expanded into

$$
\begin{aligned}
v_1' &= \sqrt{\lambda_1} u_1' + n_1' \\
v_2' &= \sqrt{\lambda_2} u_2' + n_2' \\
&\vdots \\
v_r' &= \sqrt{\lambda_r} u_r' + n_r'
\end{aligned}
, \tag{2.10}
$$

As (9) indicates, received signals $v_i', \quad i = 1, 2, ..., r$ depend on the transmitted component u_i'. The equivalent MIMO channel from (8) can be considered as consisting of r uncoupled parallel subchannels. In the equivalent MIMO channel model described by (10), the subchannels are uncoupled and thus their capacities are added up. Each subchannel is assigned to a singular value of matrix **H**, which corresponds to the channel gain. The channel power gain is thus equal to the eigenvalue of $\mathbf{HH}^H$. Block diagram of MIMO channel with premultiplying and postmultiplying is given in Figure 2.5.

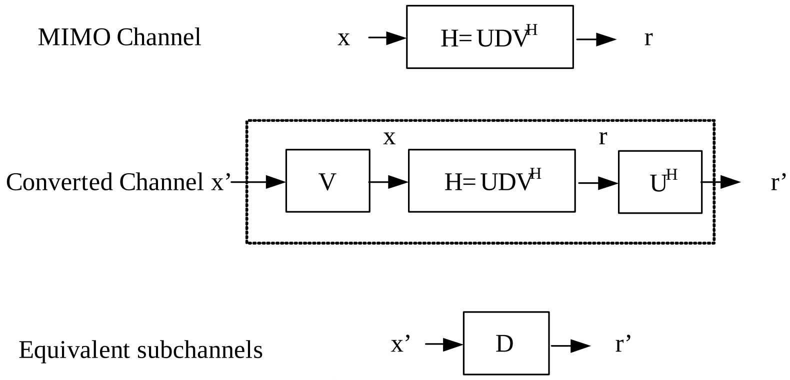

Figure 2.5: MIMO channel is converted into parallel subchannels.

By using Singular Value Decomposition (SVD), each MIMO channel can be considered as consisting of *r* parallel subchannels. Assume that CSI is not available at transmitter and power is allocated equally over all transmit antennas P/n_T. The overall channel capacity, denoted by C, can be estimated by using the Shannon capacity formula

$$C = W\sum_{i=1}^{r}\log_2\left(1+\frac{P_{ri}}{\sigma^2}\right), \tag{2.11}$$

where W is the bandwidth of each subchannel and P_{ri} is the received signal power in the i^{th} subchannel. Capacity formula can be written as

$$C = W\log_2\det(I_m + \frac{P}{n_T\sigma^2}Q) \tag{2.12}$$

where Q is the Wishart matrix, defined as

$$Q = \begin{cases} HH^H, & n_R < n_T \\ H^H H, & n_r \geq n_T \end{cases}, \tag{2.13}$$

As an example, we consider a single antenna system with $n_T = n_R = 1$ and $H = h = 1$. Shannon channel capacity is given by

$$C = W\log_2\left(1+\frac{P}{\sigma^2}\right) \tag{2.14}$$

Next, consider an MIMO channel with $n_T = n_R = n$. Assume that all subchannels are orthogonal with channel matrix $H = \sqrt{n}I$. From (14), we have

$$
\begin{aligned}
C &= W\log_2 \det\left(I_n + \frac{P}{n_T\sigma^2} nI_n \right) \\
&= W\log_2\left[diag\left(1+\frac{P}{\sigma^2}\right)\right] \quad = \quad nW\log_2\left(1+\frac{P}{\sigma^2}\right)
\end{aligned}
\tag{2.15}
$$

Channel capacity, given in (15), is also referred to as ergodic capacity. This theoretical capacity is achieved by unlimited complexity codes with infinite length. For flat fading channel, ergodic capacity increases linearly with $\min(n_T, n_R)$ at high SNR [2].

There is always a non-zero probability that the transmission rate is not supported by the channel, no matter how small the transmission rate is. This non-zero probability is referred to as outage probability. To evaluate channel reliability, *outage capacity* is introduced. This is the data rate supported by the channel over specified percentage of transmission duration. In other words, outage occurs less than some probability when transmitting at that rate. Outage capacity of $C_{0.1}$ means that this rate is supported over 90% of transmission time [7].

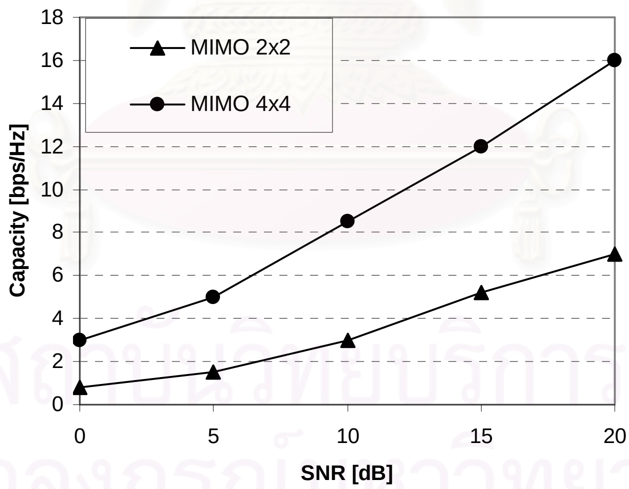

Figure 2.6: Outage Capacity.

2.3.2 Horizontal Encoding

V-BLAST is one of the most well-known horizontal encoding schemes. Bit stream are demultiplexed into n_T substreams, corresponding to n_T transmit antennas. Each substream is encoded and interleaved separately as shown in Figure 2.7. The maximum diversity gain is n_R which is lower than that of vertical encoding schemes [7].

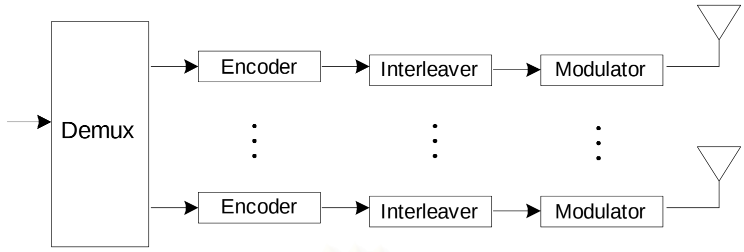

Figure 2.7: Horizontal Encoding.

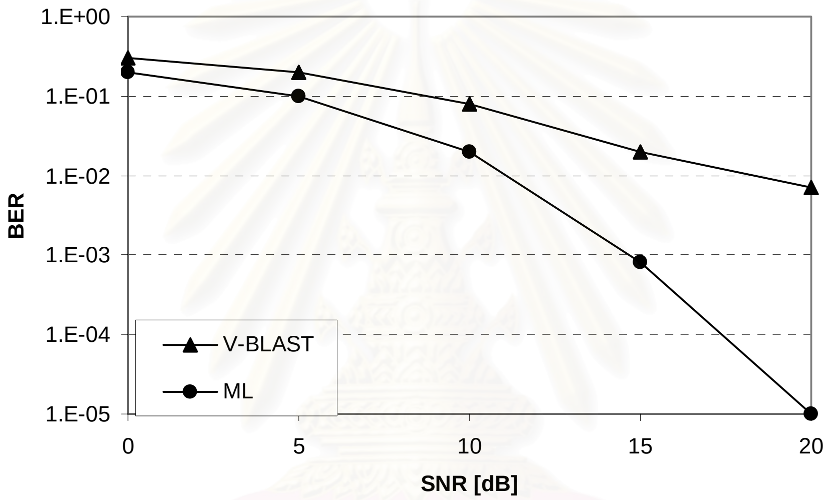

Figure 2.8: Performance of V-BLAST.

2.3.3 Vertical Encoding

In this scheme, data bit stream is first encoded and interleaved. The ouput bit stream is demultiplexed into n_T substreams to modulate and transmit over n_T antennas. Since each information bit is spread over all transmit antennas, this scheme achieves higher diversity order than horizontal encoding, that is $n_T \times n_R$. A drawback is that joint detection is required at receiver, which is more complicated than horizontal decoding.

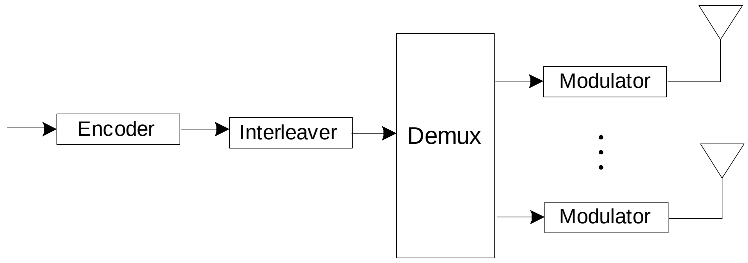

Figure 2.9: Horizontal Encoding.

2.4 MIMO-OFDM Systems with Perfect CSI

Let us consider a single point-to-point MIMO system with n_T transmit antennas and n_R receive antennas. Signal transmitted from the m^{th} antenna in the k^{th} symbol period are represented by a $n_T \times 1$ column matrix x_k^m. Denote $x_k = \left[x_k^1 \; x_k^2 \cdots x_k^{n_T} \right]$ as the transmitted vector at the k^{th} symbol period. Consider the frequency selective fading channel model with L paths. Capacity of frequency selective fading channel is higher than of flat fading channel.

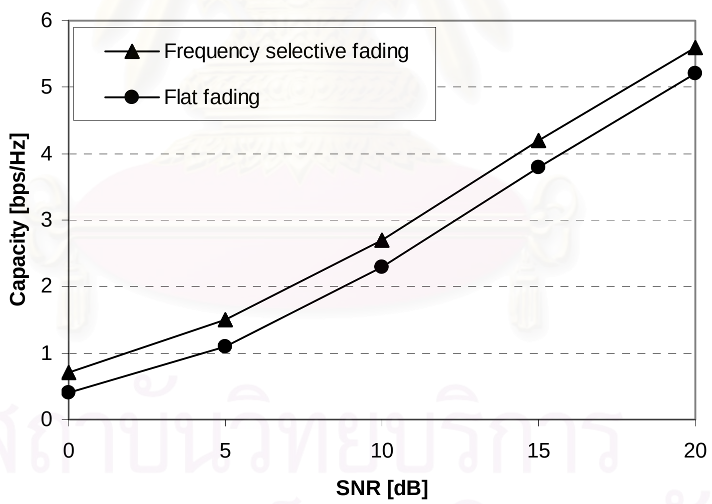

Figure 2.10: Ergodic capacity.

Received signal at receiver array can be written as

$$y_k = \sum_{l=0}^{L} G_l x_{k-l} + n_k \,, \tag{2.16}$$

where G_l is channel matrix at the l^{th} path, described by an $n_R \times n_T$ complex matrix. When fading channels are uncorrelated, G_l can be rewritten as

$$G_l = \sigma_l W_l \,, \tag{2.17}$$

where $\{\sigma_l^2\}$ is delay power profile. Elements of W_l are considered to be zero mean independent identically distributed (i.i.d.) Gaussian variables.

Since frequency selective fading causes inter-symbol interference (ISI), performance of MIMO systems can be significantly degraded. Equalization has been proposed for MIMO systems but this solution is not attractive due to complexity. Another choice is combining MIMO with multicarrier modulation, such as OFDM. OFDM transforms MIMO channel with memory into memoryless channels corresponding to different OFDM tones.

In each symbol period, N symbol vectors $\{u_1 \, u_2 \ldots u_N\}$ are transmitted with $u_n = [u_n^1 \, u_n^2 \ldots u_n^{n_T}]^T$. These symbol vectors are permutated to form symbol groups $\{u_1^m \, u_2^m \ldots u_N^m\}$ for $m = 1, 2, \ldots, n_T$. Symbol group $\{u_1^m \, u_2^m \cdots u_N^m\}$ corresponding to the m^{th} transmit antenna is transformed into output symbol vector $\{x_1^m \, x_2^m \cdots x_N^m\}$ for $m = 1, 2, \ldots, n_T$ by the Inverse Discrete Fourier Transform (IDFT). MIMO channel is diagonalized by adding cyclic prefix of length L to form transmit symbol vector

$$\{x_{N-L+1}^m \cdots x_N^m \; x_1^m \; x_2^m \cdots x_N^m\} \tag{2.18}$$

This symbol vector is transmitted over frequency selective fading channel. At the m^{th} receive antenna, cyclic prefix is discarded to produce $\{y_1^m \, y_2^m \cdots y_N^m\}$ for $m = 1, 2, \ldots, n_R$. This symbol vector is transformed into the symbol vector $\{v_1^m \, v_2^m \cdots v_N^m\}$ by the Discrete Fourier Transform (DFT). Those symbols are regrouped into $v_n^1 \, v_n^2 \ldots v_n^{n_R}$ for $n = 1, 2, \ldots, N$. Since channel is diagonalized via IDFT and DFT, received signals can be written as

$$\begin{aligned} v_1 &= H_1 u_1 + \tilde{n}_1 \\ &\vdots \quad \vdots \quad \vdots \\ v_N &= H_N u_N + \tilde{n}_N \end{aligned} \tag{2.19}$$

Chapter III

Adaptive Modulation

3.1 Introduction

In this section, a brief introduction on adaptive modulation is given. Margin maximization algorithm for Discrete Multi-tone (DMT) systems is considered and generalized for MIMO-OFDM systems.

Since fading coefficients fluctuate dramatically, communication system design must deal with the worst case by leaving some margins to maintain overall performance. Fixed modulation schemes do not take full advantage of instantaneous channel capacity.

In a narrow-band channel, as a result of its rapid fading, the short term SNR can be severely degraded, especially if the channel exhibits a deep fade. The general philosophy of AQAM is to employ a higher-order modulation mode, when the channel quality is favourable in order to increase the transmission throughput and conversely, a more robust lower-order modulation mode is invoked, when the channel quality is low. This is achieved at a constant symbol-rate, regardless of the modulation mode selected and hence at a constant bandwidth requirement.

Many transmission schemes have been proposed to utilize CSI available at both transmitter and receiver, referred to as adaptive modulation (AM). If multi-level coding is also included, it is called adaptive coded modulation (ACM). The parameters to be adapted can be coding rate, transmit power, constellation size. Simplified block diagram of a system with adaptive modulation is shown in Figure 4.

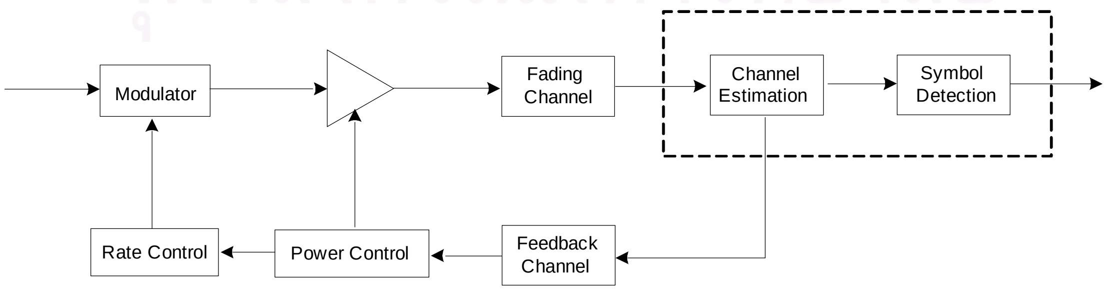

Figure 3.1: Rate and power adaptation.

3.2 Optimal Adaptive Modulation

Some new parameters are defined in this section. First, we consider only one subchannel given in (10)

$$v = \sqrt{\lambda} u + n \,, \tag{3.1}$$

Denote P as transmit power then SNR at receiver is obtained by

$$SNR = \frac{\lambda P}{\sigma_n^2} \tag{3.2}$$

where σ_n^2 is noise power spectral density given by

$$\sigma_n^2 = \frac{N_0}{2} \tag{3.3}$$

We obtain the spectral efficiency as

$$c = \log_2\left(1 + SNR\right) = \log_2\left(1 + \frac{\lambda P}{\sigma_n^2}\right), \tag{3.4}$$

To analyze the achievable data rate, SNR gap Γ is introduced as

$$\Gamma \stackrel{\Delta}{=} \frac{2^c - 1}{2^{b_{\max}} - 1} = \frac{SNR}{2^{b_{\max}} - 1} \tag{3.5}$$

Error probability can be approximated as

$$P_e = Q\left(\frac{d_{\min}}{2\sigma_n}\right) = Q\left(\sqrt{\frac{3P}{M-1} SNR}\right) = Q\left(\sqrt{3\Gamma}\right) \tag{3.6}$$

where the function Q(x) is defined as

$$Q(x) = \frac{1}{2} erfc\left(\frac{x}{\sqrt{2}}\right) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-z^2/2} dz \tag{3.7}$$

Each gap value corresponds to an error probability, at which the maximum transmission rate is given by

$$b_{\max} = \log_2\left(1 + \frac{SNR}{\Gamma}\right) = \log_2\left(1 + \frac{\lambda P}{\sigma_n^2 \Gamma}\right), \tag{3.8}$$

Another parameter to be included is margin γ, which is defined as the amount of SNR that can be reduced while maintaining the specified target error probability.

$$\gamma \overset{\Delta}{=} \frac{2^{b_{\max}}-1}{2^{b}-1} = \frac{SNR/\Gamma}{2^{b}-1}\,, \tag{3.9}$$

Bit rate corresponding to this margin is obtained by

$$b = \log_2\left(1+\frac{SNR}{\Gamma\gamma}\right) = \log_2\left(1+\frac{\lambda P}{\sigma_n^2\Gamma\gamma}\right), \tag{3.10}$$

Signals are transmitted over N OFDM subcarriers. Each OFDM subchannel corresponds to an MIMO channel. Assume that all MIMO channel matrices have the same rank of r. Average rate of rN spatial subchannels is given by

$$\bar{b} = \frac{1}{rN}\log_2\prod_{i=1}^{rN}\left(1+\frac{\lambda_i P_i}{\sigma_n^2\Gamma}\right) \overset{\Delta}{=} \log_2\left(1+\frac{SNR_{mul}}{\Gamma}\right) \tag{3.11}$$

where SNR_{mul} is defined as multi-channel SNR

$$SNR_{mul} \overset{\Delta}{=} \left[\left(\prod_{i=1}^{rN}\left(1+\frac{\lambda_i P_i}{\sigma_n^2\Gamma}\right)\right)^{1/rN} - 1\right]\Gamma\,, \tag{3.12}$$

Here we consider the optimization problem to minimize transmit power subject to fixed throughput at specified target BER.

Minimize

$$\sum_{i=1}^{rN} P_i = P_1 + P_2 + \ldots + P_{rN} \tag{3.13}$$

Subject to fixed throughput

$$b = \sum_{i=1}^{rN} b_i = b_1 + b_2 + \ldots + b_{rN} \tag{3.14}$$

(36) can also be expressed as follows

$$b = \sum_{i=1}^{rN}\log_2\left(1+\frac{\lambda_i P_i}{\sigma_n^2\Gamma}\right) = \log_2\left(1+\frac{\lambda_1 P_1}{\sigma_n^2\Gamma}\right) + \log_2\left(1+\frac{\lambda_2 P_2}{\sigma_n^2\Gamma}\right) + \ldots + \log_2\left(1+\frac{\lambda_{rN} P_{rN}}{\sigma_n^2\Gamma}\right), \tag{3.15}$$

or

$$b = \log_2 \prod_{i=1}^{rN} \left(1 + \frac{\lambda_i P_i}{\sigma_n^2 \Gamma} \right) = \log_2 \left[(1 + \frac{\lambda_1 P_1}{\sigma_n^2 \Gamma})(1 + \frac{\lambda_2 P_2}{\sigma_n^2 \Gamma}) \ldots (1 + \frac{\lambda_{rN} P_{rN}}{\sigma_n^2 \Gamma}) \right], \tag{3.16}$$

Using the Generalized Cauchy Inequality, we have

$$\sum_{i=1}^{rN} \frac{\sigma_n^2 \Gamma}{\lambda_i} + \sum_{i=1}^{rN} P_i = \sum_{i=1}^{rN} \left(1 + \frac{\lambda_i P_i}{\sigma_n^2 \Gamma} \right) \frac{\sigma_n^2 \Gamma}{\lambda_i} \geq \left(\sum_{i=1}^{rN} \frac{\sigma_n^2 \Gamma}{\lambda_i} \right) \left(\prod_{i=1}^{rN} \left(1 + \frac{\lambda_i P_i}{\sigma_n^2 \Gamma} \right)^{\frac{\sigma_n^2 \Gamma}{\lambda_i}} \right)^{\frac{1}{\sum_{i=1}^{rN} \frac{\sigma_n^2 \Gamma}{\lambda_i}}} \tag{3.17}$$

Equality happens when

$$\frac{\sigma_n^2 \Gamma}{\lambda_1} + P_1 = \frac{\sigma_n^2 \Gamma}{\lambda_2} + P_2 = \ldots = \frac{\sigma_n^2 \Gamma}{\lambda_{rN}} + P_{rN} = \mu \tag{3.18}$$

Since no power is allocated to channels with non-positive values of P_i, (39) is rewritten as

$$P_i = \mu - \frac{\sigma_n^2 \Gamma}{\lambda_i} \quad for \quad i = 1, 2, \ldots, rN \tag{3.19}$$

$$P_i = \max \left\{ 0, \mu - \frac{\sigma_n^2 \Gamma}{\lambda_i} \right\} \text{ for } i = 1, 2, \ldots, rN \tag{3.20}$$

This result can also be found with Lagrange multiplier method. The solution given above is widely known in literature as the “water-filling” solution. From the constraint on fixed throughput, we have

$$b = \sum_{i=1}^{rN} \log_2 \left(1 + \frac{\lambda_i P_i}{\sigma_n^2 \Gamma} \right) = \sum_{i=1}^{rN} \log_2 \left(\frac{\lambda_i \mu}{\sigma_n^2 \Gamma} \right) = \log_2 \prod_{i=1}^{rN} \left(\frac{\lambda_i \mu}{\sigma_n^2 \Gamma} \right), \tag{3.21}$$

$$2^b = \left(\frac{\mu}{\Gamma} \right)^{rN} \prod_{i=1}^{rN} \left(\frac{\lambda_i}{\sigma_n^2} \right) \tag{3.22}$$

The constant μ is determined as

$$\mu = \Gamma \frac{2^{b/rN}}{\left(\prod_{i=1}^{rN} \left(\lambda_i / \sigma_n^2 \right) \right)^{1/rN}}, \tag{3.23}$$

Rate and power allocation are given by

$$P_i = \mu - \frac{\sigma_n^2 \Gamma}{\lambda_i}, \tag{3.24}$$

$$b_i = \log_2\left(\frac{\lambda_i \mu}{\sigma_n^2 \Gamma}\right) \tag{3.25}$$

The analysis above assumes that b_i can receive any non-negative real value, which is impossible to implement. In practice, constellation sizes are taken from a finite set of integers.

Define the information unit β be the smallest incremental unit of spectral efficiency b_i. The rate of the i^{th} subchannel is given by

$$r_{k,i}[n] = r_{k,i}^{D}[n] + r_{k,i}^{I}[n], \tag{3.26}$$

$$b_i = B_i \beta \text{ where } i = 1,2,\ldots,MN \tag{3.27}$$

where B_i is a non-negative integer with the maximum value of $B_{\max} = \max_i B_i$.

Incremental power for the i^{th} subchannel is defined as the power difference when allocating B_i and $B_i - 1$ information units β.

$$\Delta P_i(b_i) \overset{\Delta}{=} P_i(b_i) - P_i(b_i - \beta), \tag{3.28}$$

Bit allocation is efficient if any movement of a bit from one subchannel to another subchannel will increase the overall symbol power, that is

$$\max_i[\Delta P_i(b_i)] \le \min_j[\Delta P_j(b_j + \beta)], \tag{3.29}$$

Levin-Campello Algorithm for margin maximization is introduced [8].

$$r_{k,i}[n] = r_{k,i}^{D}[n] + r_{k,i}^{I}[n], \tag{3.30}$$

1. $j \leftarrow \arg\{\min_{1\le p\le rN}[P_p(b_p + \beta)]\}$

2. $i \leftarrow \arg\{\max_{1\le q\le rN}[P_q(b_q)]\}$

3. While $\Delta P_j(b_j + \beta) < \Delta P_i(b_i)$ do

   $b_j \leftarrow b_j + \beta$

$$b_i \leftarrow b_i - \beta$$

$$j \leftarrow \arg\left\{\min_{1\le p\le rN}[P_p(b_p+\beta)]\right\}$$

$$i \leftarrow \arg\left\{\max_{1\le q\le rN}[P_q(b_q)]\right\}$$

4. Set $\tilde{b} = \sum_{i=1}^{rN} b_i$

5. While $\tilde{b} \neq b$

If $\tilde{b} > b$

$$i \leftarrow \arg\left\{\max_{1\le q\le rN}[P_q(b_q)]\right\}$$

$$\tilde{b} \leftarrow \tilde{b} - \beta$$

$$b_i \leftarrow b_i - \beta$$

Else

$$j \leftarrow \arg\left\{\min_{1\le p\le rN}[P_p(b_p+\beta)]\right\}$$

$$\tilde{b} \leftarrow \tilde{b} + \beta$$

$$b_i \leftarrow b_i + \beta$$

Adaptive modulation is performed on an MIMO-OFDM system with n_T transmit and n_R receive antennas. Margin Maximization algorithm for DMT is applied to MIMO-OFDM system by noting that each OFDM subchannel has its own MIMO channel. In such way, (10) can be generalized into

$$r_{k,i}[n] = r_{k,i}^{D}[n] + r_{k,i}^{I}[n], \tag{3.31}$$

$$\left.\begin{aligned} r_1^1 &= \sqrt{\lambda_1^1}\, x_1^1 + n_1^1 \\ r_2^1 &= \sqrt{\lambda_2^1}\, x_2^1 + n_2^1 \\ &\ \vdots \quad \vdots \quad \vdots \\ r_K^1 &= \sqrt{\lambda_K^1}\, x_K^1 + n_K^1 \end{aligned}\right\} \quad \text{Spatial subchannels of the } 1^{\text{st}} \text{ OFDM tone}$$

$$\vdots \quad \vdots \quad \vdots \quad \vdots$$

$$\vdots \quad \vdots \quad \vdots \quad \vdots$$

$$\left.\begin{aligned} r_1^N &= \sqrt{\lambda_1^N}\, x_1^N + n_1^N \\ r_2^N &= \sqrt{\lambda_2^N}\, x_2^N + n_2^N \\ &\ \vdots \quad \vdots \quad \vdots \\ r_K^N &= \sqrt{\lambda_K^N}\, x_K^N + n_K^N \end{aligned}\right\} \quad \text{Spatial subchannels of the } N^{\text{th}} \text{ OFDM tone}$$

Both OFDM and MIMO are vulnerable to error so powerful channel coding is required. However, channel coding is not considered in this work.

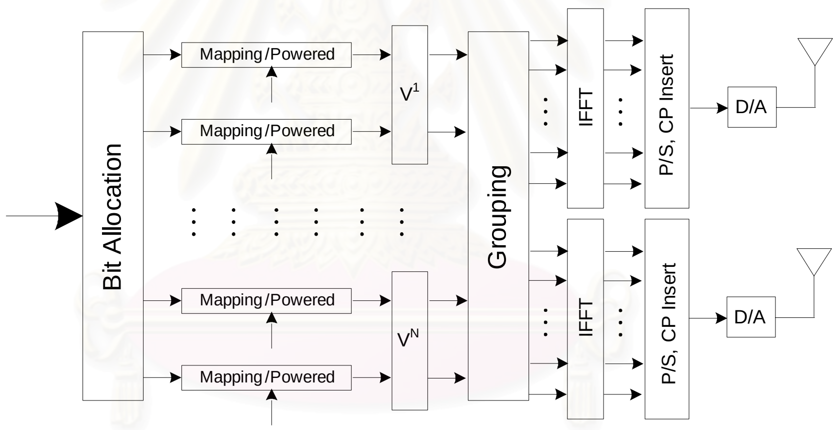

Figure 3.2: Transmitter of MIMO-OFDM with adaptive modulation .

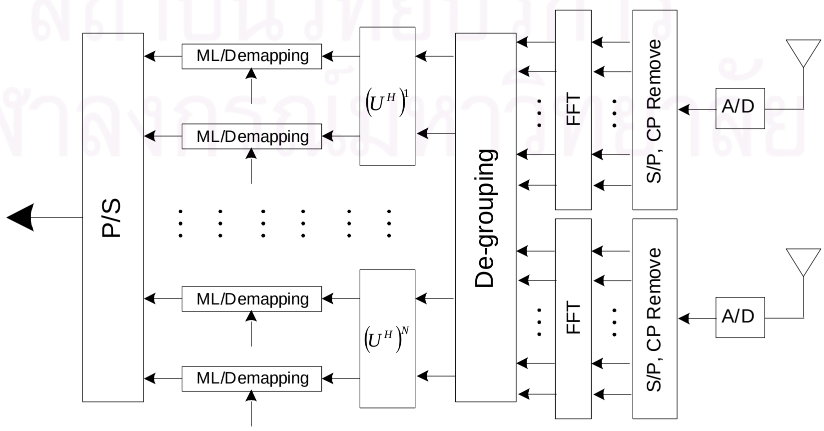

Figure 3.3: Receiver of MIMO-OFDM with adaptive modulation.

In simulation, six-path channel model obtained from the modified Jakes simulator is chosen [9]. Assume that fading is quasi-static, that is fading is constant in one OFDM symbol period and changes from symbol to symbol. Adaptive modulation is performed with square M-QAM constellations. Number of bits per symbol is taken from the set of $\{2,4,6\}$. Gains of first-order spatial subchannels dominate gains of second order spatial subchannels. We can reduce complexity by performing margin maximization algorithm over first order spatial subchannels. This helps reduce the burden on feedback channel, at the cost of degraded performance. Block diagrams of transmitter and receiver for this scheme are given in Figure 3.4 and Figure 3.5. Transmit signal is premultiplied with right singular vector of **H** as given in (11). Receive signal is postmultiplied with left singular vector of **H**, as given in (12).

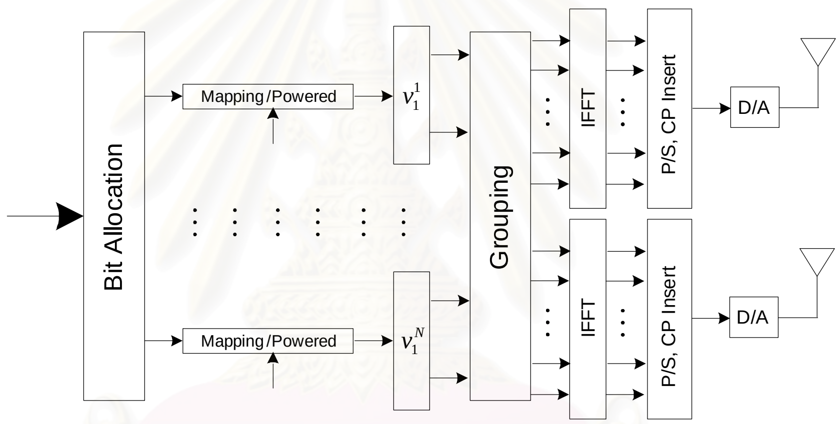

Figure 3.4: Transmitter of MIMO-OFDM with one channel per tone.

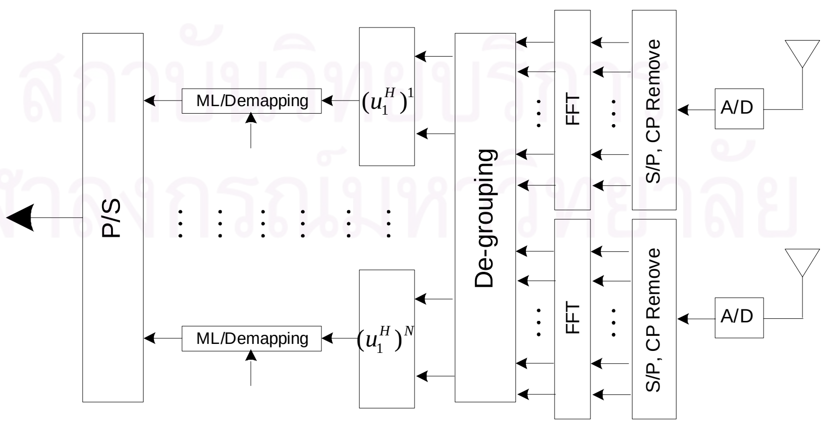

Figure 3.5: Receiver of MIMO-OFDM with one channel per tone.

Chapter IV

Proposed Schemes

4.1 Introduction

As described in chapter 3, optimal bit allocation algorithm (OBAA) is of high complexity since only one incremental unit of bits is loaded at a time to the subchannel which requires least amount of extra power. Different bit allocation schemes have been proposed as sub-optimal approaches. Optimal bit allocation algorithm (OBAA) for discrete multi-tone (DMT) systems is proposed in [18]. One bit incremental unit is allocated at a time to the subcarrier which requires least extra power.

In this chapter, a novel adaptive modulation algorithm for minimizing transmit power in MIMO-OFDM systems is proposed. Our proposed algorithm is refered to as upward bit allocation algorithm (UBAA). Some materials on UBAA can be found in [16]. UBAA is applied to MIMO-OFDM systems with different assumptions on channel state information. A novel algorithm for bit allocation is proposed, called upward bit allocation algorithm (UBAA), to minimize transmit power while maintaining bit error rate (BER) in MIMO-OFDM systems with fixed throughput.

First, MIMO-OFDM systems with perfect channel state information are considered. Optimal solution is the combination of Singular Value Decomposition (SVD) and water-filling techniques. V-BLAST based technique for adaptive modulation is proposed using UBAA to reduce complexity.

Secondly, assumption on partial channel state infomation (CSI) is made. Partial CSI is also known as statistical CSI in literature. Transmitter design with coder-beamformer from [14] is adopted. Suboptimal bit allocation algorithm UBAA is proposed in stead of the Hughes-Hartogs algorithm which is used in [14] to reduce complexity.

4.2 UBAA for MIMO-OFDM Systems with Perfect CSI

When channel state information (CSI) is available at transmitter via a feedback link, system performance can be significantly improved by adapting transmission to varying channel. Adaptive modulation has been studied extensively. Optimal solution for MIMO-OFDM systems is OBAA in combination with singular value decomposition (SVD). MIMO channel at each OFDM tone is decomposed into parallel spatial subchannels. Performance of MIMO-OFDM is shown in Figure 4.1 with $n_T = n_R = 4$, $N = 64$ while varying velocity of mobile station .

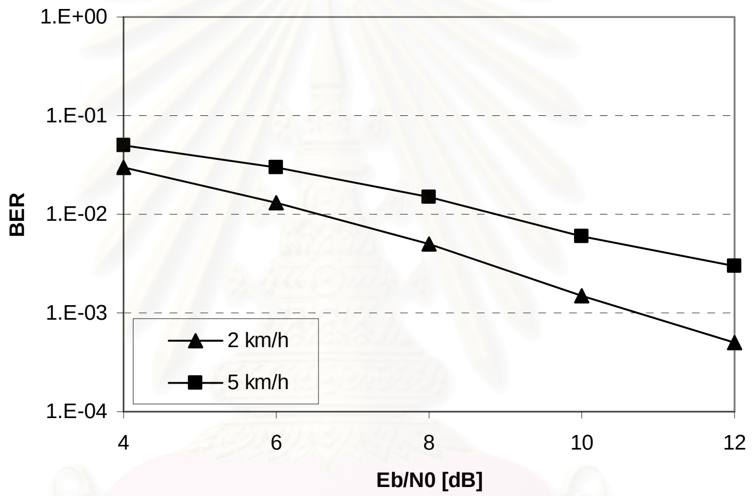

Figure 4.1: Performance of MIMO-OFDM systems with SVD.

Since SVD-based adaptive modulation for MIMO-OFDM is too complicated, V-BLAST-based adaptive modulation is considered for minimizing transmit power to reduce complexity.

Bell Labs Space Time Architecture (BLAST) detection was initially proposed for a testbed on MIMO systems. The idea was developed from multiuser detection with successive interference cancellation in each iteration. The original scheme is Diagonal BLAST (D-BLAST) which achieves efficient spectral efficiency with high complexity. In the family of BLAST algorithms, Vertical BLAST (V-BLAST) is the most popular due to low complexity and acceptable performance. It was reported that

MIMO systems with Vertical Bell Laboratories Layered Space-Time (V-BLAST) detection achieved spectral efficiency of up to 20-40 bits/sec/Hz.

Data streams for different transmit antennas are encoded separately. Detection process is performed with nulling and cancellation. Nulling is performed based on either the zero-forcing (ZF) criteria or the minimum mean square error (MMSE) criteria. Nulling vector is chosen to detect one symbol at a time. Detected symbol is cancelled out from the received symbols. The limitation of V-BLAST detection is error propagation. Performance of MIMO systems with V-BLAST detection, zero-forcing (ZF) detection and maximum likelihood (ML) is shown in Figure 4.2.

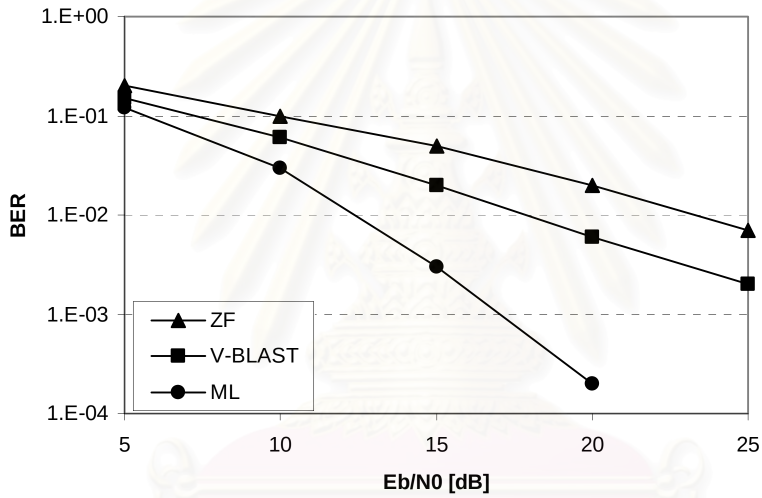

Figure 4.2: Comparison of different detection algorithms.

As a sub-optimal approach to avoid complexity of OBAA, an efficient bit allocation scheme for OFDM systems was proposed in [19]. Bit allocation for the overall system is found by solving optimization problem iteratively.

Each MIMO channel has n_T spatial subchannels so the total number of spatial subchannels available is $N \times n_T$. Denote denote $P_{k,s}$ and $b_{k,s}$ as power and number of bits, respectively, allocated to the s^{th} spatial subchannel at the k^{th} OFDM tone with $s \in \{1, 2, \ldots, n_T\}$ and $k \in \{1, 2, \ldots, N\}$.

At transmitter, $N \times n_T$ bit tuples are mapped to $N \times n_T$ complex QAM symbols. The Inverse Discrete Fourier Transform (IDFT) translates those QAM symbols to OFDM tones to form an OFDM symbol.

At receiver, OFDM demodulation is performed with the Discrete Fourier Transform (DFT). V-BLAST detection is performed for each OFDM tone separately. Only zero-forcing (ZF) nulling is considered for simplicity. Post-detection signal-to-noise (SNR) of each symbol can be expressed as

$$\rho_{k,s} = \frac{P_{k,s}}{\sigma_n^2 \left\| w_{k,s} \right\|^2} \tag{4.1}$$

where σ_n^2 is noise variance, $\left\| w_{k,s} \right\|$ is norm of the nulling vector $w_{k,s}$. Spatial subchannel gain can be written as

$$g_{k,s} = \frac{1}{\left\| w_{k,s} \right\|^2}. \tag{4.2}$$

For simplicity, only square QAM signal constellations QPSK, 16-QAM, 64-QAM are considered. Bit sequences of length b_{sum}, corresponding to one OFDM symbol, are allocated to $N \times n_T$ spatial subchannels. Constraint on fixed throughput is given by

$$\sum_{k=1}^{N} \sum_{s=1}^{n_T} b_{k,s} = b_{sum} \quad with \quad b_{k,s} \in \{0, 2, 4, 6\} \tag{4.3}$$

where $b_{k,s}$ is the number of bits corresponding to the QAM constellation of size $M_{k,s} = 2^{b_{k,s}} \in \{4, 16, 64\}$. A spatial subchannel with $b_{k,s} = 0$ is not used for information bearing and no power is allocated to that subchannel. All subchannels have the same desired bit-error-rate (BER)

$$\xi_{k,s} \le \xi \quad with \quad \forall k, s. \tag{4.4}$$

Total transmit power P_{sum} is calculated from

$$P_{sum} = \sum_{k=1}^{N} \sum_{s=1}^{n_T} P_{k,s} \quad with \quad P_{k,s} \geq 0. \tag{4.5}$$

Optimization problem can be stated as follows: *minimize total transmit power* P_{sum} *given in (4.5)subject to constraints (4.3) and (4.4).*

To facilitate V-BLAST detection, same number of bits b_k is allocated to all subchannels at the k^{th} OFDM tone

$$b_{k,1} = b_{k,2} = \ldots = b_{k,n_T} = b_k \quad with \quad \forall k. \tag{4.6}$$

Bit allocation over $N \times n_T$ spatial subchannels now reduces to bit allocation over N OFDM tones

$$\sum_{k=1}^{N} b_k = \frac{b_{sum}}{n_T} \quad with \quad b_k \in \{0, 2, 4, 6\}. \tag{4.7}$$

Each OFDM tone is characterized by average tone gain defined as

$$\eta_k = \frac{1}{n_T} \sum_{s=1}^{n_T} g_{k,s} = \frac{1}{n_T} \sum_{s=1}^{n_T} \frac{1}{\|w_{k,s}\|^2}. \tag{4.8}$$

Let $R_i(\xi)$ be the SNR required to transmit i bits per OFDM tone at BER ξ. Power required at the k^{th} OFDM tone can be written as

$$P_k = \frac{\sigma_n^2 R_{b_k}(\xi)}{\eta_k}. \tag{4.9}$$

Total power to be minimized is

$$P_{sum} = n_T \times \sigma_n^2 \sum_{k=1}^{N} \frac{R_{b_k}(\xi)}{\eta_k}. \tag{4.10}$$

Optimal solution OBAA is presented in [18] by introducing the parameter Γ

$$\Gamma = \frac{R_{b_k}(\xi)}{2^{b_k} - 1} \tag{4.11}$$

which characterizes the difference between achievable performance and Shannon capacity. Each Γ corresponds to a specified value of BER. Equation (4.11) can be rewritten as

$$b_k = \log_2\left(1 + \frac{R_{b_k}(\xi)}{\Gamma}\right) = \log_2\left(1 + \frac{P_k \eta_k}{\sigma_n^2 \Gamma}\right). \tag{4.12}$$

From (4.12), power required to load b_k bit to the k^{th} OFDM tone is

$$P_k = (2^{b_k} - 1)\frac{\sigma_n^2 \Gamma}{\eta_k}. \tag{4.13}$$

We compare power required to load different number of bits while maintaining BER. From (4.13), it is clear that

$$\frac{P_{k|b_k=4}}{P_{k|b_k=2}} = \frac{2^4 - 1}{2^2 - 1} = 5 \tag{4.14}$$

and

$$\frac{P_{k|b_k=6}}{P_{k|b_k=2}} = \frac{2^6 - 1}{2^2 - 1} = 21. \tag{4.15}$$

In OBAA, 2 bits are loaded at a time to the OFDM tone which requires smallest additional power. Although OBAA is optimal, it is too complicated, especially for systems utilizing large number of OFDM tones. Next, The algorithm proposed by Piazzo in [19] is briefly described. For clarity, Piazzo's algorithm for adaptive modulation with square QAM constellations will be discussed. A J-bit group is defined as the set of all OFDM tones which are allocated J bits per tone. Final solution is found by considering systems with 2 QAM constellations at a time, referred to as a two-mode system (TMS).

First, all bits in budget are allocated to OFDM tones in J-bit group corresponding to the constellation with largest size of the TMS being considered. Then bits are reallocated to $(J-2)$-bit group. In each reallocation, $2(J-2)$ bits from $(J-2)$ OFDM tones with smallest tone gains in J-bit group are moved to form 2 OFDM tones in $(J-2)$-bit group. New OFDM tones in $(J-2)$-bit group are taken from OFDM tones with largest gains in 0-bit groups.

Once optimal solution for the first TMS is determined, number of OFDM tones in J-bit group is fixed. Number of bits allocated to J-bit group is deduced from total bit budget. The same process is performed to solve optimization problem for new TMS with $(J-2)$-bit group and $(J-4)$-bit group. Although solution for each TMS is optimal, overall result is not guaranteed to be optimal. Our proposed upward bit allocation algorithm (UBAA) is performed by sorting

$$\eta_{k_1} \ge \eta_{k_2} \ge \ldots \ge \eta_{k_N} \tag{4.16}$$

where $k_i \in \{1, 2, \ldots, K\}$ is OFDM tone index. Let α be index of the OFDM tone with largest gain, β be index of the OFDM tone with smallest gain in J-bit group. Total number of bits allocated to OFDM tones in J-bit group is denoted by b_{budget}. Two bits from the tone with index β are moved to the tone with index α. Now, OFDM tone with index α belongs to $(J+2)$-bit group, OFDM tone with index β belongs to $(J-2)$-bit group. Bit reallocation for OFDM tones in 4-bit group is shown in Figure 4.3 and Figure 4.4. Reallocation results in reducing transmit power if

$$\frac{R_{J+2}(\xi)}{\eta_\alpha} + \frac{R_{J-2}(\xi)}{\eta_\beta} < \frac{R_J(\xi)}{\eta_\alpha} + \frac{R_J(\xi)}{\eta_\beta}. \tag{4.17}$$

Reallocation is performed iteratively until transmit power can not be further reduced. Inequality (4.17) can be rewritten as

$$\frac{R_{J+2}(\xi)-R_J(\xi)}{\eta_\alpha}-\frac{R_J(\xi)-R_{J-2}(\xi)}{\eta_\beta}<0. \tag{4.18}$$

From (4.11), it comes to

$$R_{J+2}(\xi)>R_J(\xi)>R_{J-2}(\xi). \tag{4.19}$$

Denote values of α and β after the m^{th} reallocation by $\alpha(m)$ and $\beta(m)$. Initially, these values are set at $\alpha(0)=1$ and $\beta(0)=b_{budget}/J$.

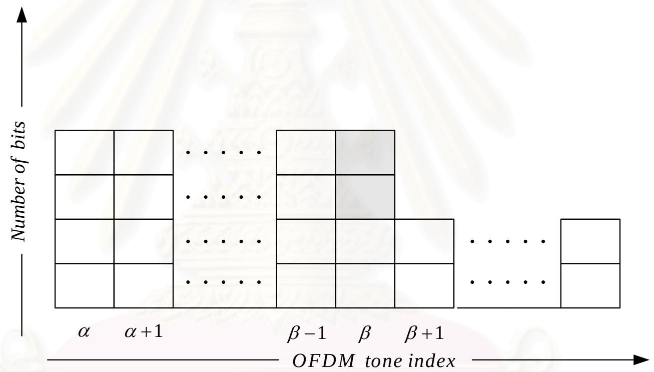

Figure 4.3: Before Reallocation.

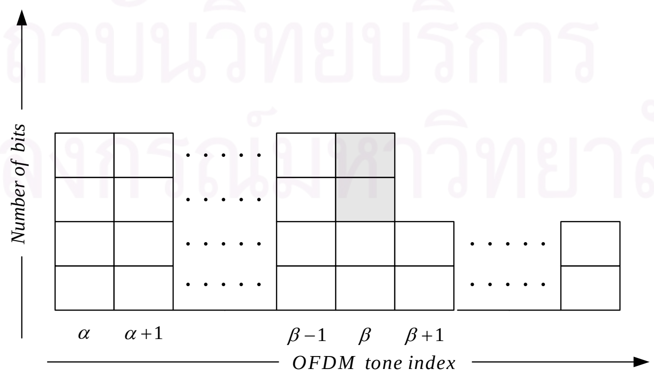

Figure 4.4: After Reallocation.

After the m^{th} reallocation, subchannel indices are changed to $\alpha(m) = 1 + m$ and $\beta(m) = (b_{budget} / J) - m$. Substituting these values to (4.18), the left-hand side becomes a function of m

$$f(m) = \frac{R_{J+2}(\xi) - R_J(\xi)}{\eta_{(1+m)}} - \frac{R_J(\xi) - R_{J-2}(\xi)}{\eta_{((b_{budget}/J)-m)}}. \quad (4.20)$$

From (4.16) and (4.19), $f(m)$ is an increasing function with respect to m. The optimal solution is found by searching for the value of m that satisfies $f(m) \leq 0$ and $f(m+1) > 0$. The search is linear as m increases by one in each step. Linear search is replaced by efficient logarithmic search as in Piazzo's algorithm [5] since the function $f(m)$ is an increasing function. Complexity of adaptive modulation for systems with large number of OFDM tones is significantly reduced by logarithmic search. Logarithmic search halves the range at each step by testing the value of $f(m)$ at center of the range and updates the range accordingly.

Denote the minimum value and the maximum value of m by m_0 and m_1, respectively. Initial value of m_0 is zero and initial value of m_1 is determined from $m_1 = b_{budget} / J / 2.$ UBAA can be mathematically described as follows

i, Initialize $m_0 = 0$, $m_1 = b_{budget} / J / 2$

ii, Compute $m_x = (m_0 + m_1)/2$

iii, If $f(m_x) < 0$ *let* $m_0 = m_x$*; else* $m_1 = m_x$

iv, If $m_1 = m_0 + 1$ *go to v,; otherwise go to ii,*

v, Stop, $b_{budget} = (J+2)m_0$

Total power assigned to the k^{th} OFDM tone is $n_T \times P_k$, determined from bit allocation process. This amount of power is further divided to spatial subchannels.

Since the same QAM constellation of size M_k is used over all n_T spatial subchannels of each OFDM tone, BER can be written as

$$\xi_{k,s} = \frac{4}{b_k}\left(\frac{\sqrt{M_k}-1}{\sqrt{M_k}}\right)Q\left(\sqrt{\frac{3P_{k,s}}{(M_k-1)\sigma_n^2\left\|w_{k,s}\right\|^2}}\right). \quad (4.21)$$

From (4.4), all spatial subchannels are kept at the same BER

$$\xi_{k,s} = \xi \text{ with } \forall k, s. \quad (4.22)$$

Substituting (4.21) into (4.22), power allocation to each spatial subchannel is found in closed-form

$$P_{k,s} = \frac{\left\|w_{k,s}\right\|^2}{\sum_{s=1}^{n_T}\left\|w_{k,s}\right\|^2} \times n_T \times P_k. \quad (4.23)$$

Rayleigh fading channel model with exponential power delay profile is simulated. Simulation results for uncoded MIMO-OFDM systems with $N = 256$ and $n_T = n_R = 4$ are shown in Figure 4.5. The four curves show transmit power required by different schemes: without adaptation, V-BLAST-based Piazzo's algorithm, proposed V-BLAST UBAA and V-BLAST-based OBAA.

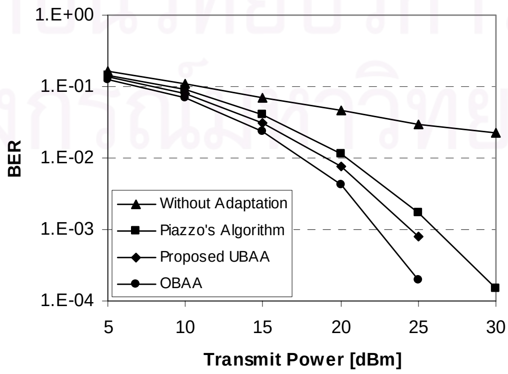

Figure 4.5: V-BLAST based adaptive modulation with perfect CSI.

Complexity of adaptive modulation schemes in this thesis consist of operations such as comparision and look-up table accesses. Hughes-Hartogs requires $(N-1)b_{sum}/2$ comparision operations and $N-1+(b_{sum}/2)$ table accesses. Piazzo's algorithm requires $N\log_2 N$ comparision operations and $N-1+(b_{sum}/2)$ table accesses. Our proposed schemes requires $N\log_2 N$ comparision operations and $2N-1$ table accesses. In simulation, average number of bits per subchannel is set at $b_{avg}=2$. With this setting, complexity of UBAA is lower than that of Piazzo's algorithm. However, this is not the case when average number of bits per subchannel is high, $b_{avg}>3$, when compared to constellation sizes in use.

4.3 UBAA for MIMO-OFDM Systems with Partial CSI

When perfect CSI is available at transmitter, performance of MIMO-OFDM systems can significantly be improved by adapting transmission to varying channel. However, assumption on perfect CSI is somewhat impractical due to estimation error, feedback delay, outdated CSI, quantization error, etc. All these impairments are neglected when perfect CSI is assumed. That comes to designs with partial CSI (also known in literature as statistical CSI). In this section, configurations of MIMO-OFDM systems from [1] are described.

The concept of partial CSI is loosely defined in literature with different assumptions. In this work, concept of partial CSI as in [13]-[14] with mean feedback model is adopted. On the n^{th} tone, an unbiased channel estimate $\bar{H}(k,n)$ with pertubation component $\Xi(k,n)$ is assumed.

$$\breve{H}[k,n]=\bar{H}[k,n]+\Xi[k,n] \tag{4.24}$$

$\Xi(k,n)$ is a Gaussian random matrix with variance $\delta_{\varepsilon}^{2}(k,n)$. As in [13], a two-dimensional coder-beamformer is implemented for each OFDM tone. Two OFDM symbols are taken to form one space-time coded block. Actually, this structure is a special case of combination between beam-forming and space-time block coding. Different OFDM tones suffer from different attenuation due to frequency selective fading. Power allocation is performed to balance BER over these tones. Denote $A[k,n]$, $P[k,n]$ as number of bits and power, allocated to the n^{th} tone in the k^{th} space-time coded block, respectively. Number of bits $A[k,n]$ corresponds to the constellation size of $M[k,n]=2^{b[k,n]}$. In this work, only square QAM constellations are considered. Symbol error rate is bounded by

$$P_s \le 1-\left[1-2Q\left(\sqrt{\frac{3E_s}{(M-1)N_0}}\right)\right]^2 \tag{4.25}$$

where E_s is symbol power , N_0 is noise power and $M=2^k$ is constellation size. Bit error rate can be expressed as

$$P_b = \frac{2^{k-1}}{2^k-1}P_s \tag{4.26}$$

Beam-forming matrix

$$U^*[k,n]=\left[u_1^*[k,n] \quad u_1^*[k,n]\right] \tag{4.27}$$

Alamouti space-time code matrix can be written as

$$S[k,n]=\begin{bmatrix} s_1[k,n] & -s_2^*[k,n] \\ s_2[k,n] & s_1^*[k,n] \end{bmatrix} \tag{4.28}$$

Power allocation matrix over basis beams

$$D[k,n]=\begin{bmatrix} \sqrt{\delta_1[k,n]} & 0 \\ 0 & \sqrt{\delta_2[k,n]} \end{bmatrix} \tag{4.29}$$

with $0\le\delta_1[k,n],\delta_1[k,n]\le 1$ and $\delta_1[k,n]+\delta_1[k,n]=1$.

Space-time coded block can be written as

$$X\left[k,n\right]=U^{*}\left[k,n\right]D\left[k,n\right]S\left[k,n\right] \tag{4.30}$$

Received block can be expressed by

$$\begin{aligned} Y\left[k,n\right] &= H^{T}\left[k,n\right]X\left[k,n\right]+W\left[k,n\right] \\ &= H^{T}\left[k,n\right]U^{*}\left[k,n\right]D\left[k,n\right]S\left[k,n\right]+W\left[k,n\right] \end{aligned} \tag{4.31}$$

From partial CSI $\left(\bar{H}[n],\sigma_{\varepsilon}^{2}[n]\right)$, optimal basis beams are determined from two eigenvectors of channel correlation matrix, corresponding to the largest eigenvalues. . Once system configuration is determined, bit allocation is performed over all subchannels using Hughes-Hartogs algorithm (HHA). At each time, one bit is loaded to the subchannel which requires smallest amount of additional power.

Adaptive modulation is performed at transmitter based on subchannel gains to determine bit and power allocation in data transmission period. Only square QAM signal constellations QPSK, 16-QAM, 64-QAM are considered. Denote the minimum Euclidean distance by $d_{\min}[n]$. Scaled distance metric is defined as

$$d^{2}[n]=\frac{d_{\min}^{2}[n]}{4}=\frac{g(b[n])}{4}P[n]=\frac{6}{4(2^{b[n]}-1)}P[n] \tag{4.32}$$

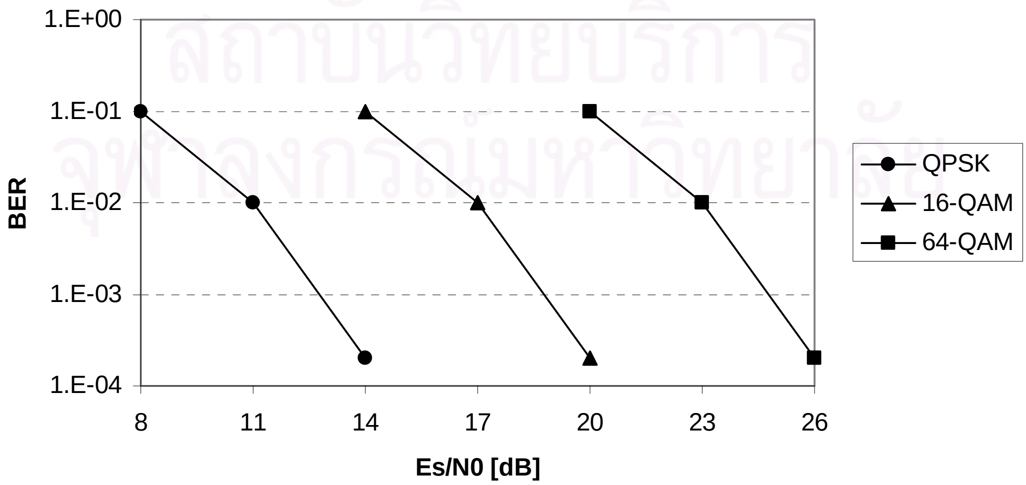

Figure 4.6: Performance of square QAM constellations.

Bit sequences of length b_{sum}, corresponding to one OFDM symbol, are allocated to N spatial subchannels. Constraint on fixed throughput is given by

$$\sum_{n=1}^{N} b_n = b_{sum} \quad with \quad b_n \in \{0,2,4,6\} \tag{4.33}$$

where b_n is the number of bits allocated to a spatial subchannel, corresponding to a QAM constellation size $M_n = 2^{b_n} \in \{4,16,64\}$. A spatial subchannel with $b_n = 0$ is not used for information bearing and no power is allocated to that subchannel.

Total transmit power P_{sum} is calculated by

$$P_{sum} = \sum_{n=1}^{N} P_n \quad with \quad P_n \geq 0 \tag{4.34}$$

In [15], it is proved that there exists a threshold value of $d_0^2[n]$ such that

$$BER[n] \leq BER_0[n] \text{ if and only if } d^2[n] \geq d_0^2[n] \tag{4.35}$$

This threshold value can also be seen as inverse of subcarrier gain $\eta[n] = 1/d_0^2[n]$. Minimum power required to load $b[n]$ bits to the subcarrier with threshold value of $d_0^2[n]$ is $d_0^2[n]/g(b[k])$. Incremental power required to load the last bit can be expressed by

$$\Delta P(n,b[n]) = \frac{d_0^2[n]}{g(b[n])} - \frac{d_0^2[n]}{g(b[n]-1)} \tag{4.36}$$

Let $R_{b[k]}(\xi)$ be SNR required to transmit i bits per OFDM tone at BER ξ. Power required at the n^{th} OFDM tone can be written as

$$P[n] = \frac{\sigma^2 R_{b[n]}(\xi)}{\eta[n]} \tag{4.37}$$

Total power to be minimized is

$$P_{sum} = \sigma^2 \sum_{n=1}^{N} \frac{R_{b[n]}(\xi)}{\eta[n]} \tag{4.38}$$

Optimal solution OBAA is presented in [3] by introducing the parameter Γ

$$\Gamma \triangleq \frac{R_{b[n]}(\xi)}{2^{b[n]}-1} \tag{4.39}$$

which characterizes the difference between achievable performance and Shannon capacity. Each Γ corresponds to a value of BER. Equation (4.11) can be rewritten as

$$b[n] = \log_2\left(1+\frac{R_{b[n]}(\xi)}{\Gamma}\right) = \log_2\left(1+\frac{P_{b[n]}\eta_{[n]}}{\sigma^2\Gamma}\right) \tag{4.40}$$

Power required to load $b[n]$ bit to the n^{th} OFDM tone is

$$P[n] = (2^{b[n]}-1)\frac{\sigma^2\Gamma}{\eta_n} \tag{4.41}$$

The same process is performed to solve optimization problem for new TMS with $(J-2)$-bit group and $(J-4)$-bit group. Although solution for each TMS is optimal, overall result is not guaranteed to be optimal. Average tone gains η_n are sorted in descending order

$$\eta_{k_1} \geq \eta_{k_2} \geq \ldots \geq \eta_{k_N} \tag{4.42}$$

where $k_i \in \{1,2,\ldots,N\}$ is OFDM tone index.

Let α be index of the OFDM tone with largest gain, β be index of the OFDM tone with smallest gain in J-bit group. Total number of bits allocated to OFDM tones in J-bit group is denoted by b_{budget}. Two bits from the tone with index β are moved to the tone with index α. Now, OFDM tone with index α belongs to $(J+2)$-bit group, OFDM tone with index β belongs to $(J-2)$-bit group. Bit reallocation for OFDM tones in 4-bit group is shown in Figure 4.2. Reallocation results in reducing transmit power if

$$\frac{R_{J+2}(\xi)}{\eta_\alpha}+\frac{R_{J-2}(\xi)}{\eta_\beta} < \frac{R_J(\xi)}{\eta_\alpha}+\frac{R_J(\xi)}{\eta_\beta} \tag{4.43}$$

Reallocation is performed iteratively until transmit power can not be further reduced

$$\frac{R_{J+2}(\xi)-R_J(\xi)}{\eta_\alpha}-\frac{R_J(\xi)-R_{J-2}(\xi)}{\eta_\beta} < 0 \tag{4.44}$$

From (4.42), we have the inequalities

$$R_{J+2}(\xi) > R_J(\xi) > R_{J-2}(\xi) \tag{4.45}$$

Denote values of α and β after the m^{th} reallocation by $\alpha(m)$ and $\beta(m)$. It can be seen easily that initial values are $\alpha(0)=1$ and $\beta(0)=b_{budget}/J$. After the m^{th} reallocation, subchannel indices are changed to $\alpha(m)=1+m$ and $\beta(m)=(b_{budget}/J)-m$. Substituting these values to (18), the left-hand side becomes a function of m

$$f(m) = \frac{R_{J+2}(\xi) - R_J(\xi)}{\eta_{(1+m)}} - \frac{R_J(\xi) - R_{J-2}(\xi)}{\eta_{((b_{budget}/J)-m)}} \tag{4.46}$$

From (76), $f(m)$ is an increasing function with respect to m. The optimal solution is found by searching for the value of m that satisfies $f(m) \le 0$ and $f(m+1) > 0$.

Channel model for simulation is taken from [14]. Uncoded MIMO-OFDM is simulated with $N = 256$ and $n_T = n_R = 4$. The four curves correspond to without adaptation, Piazzo's algorithm, UBAA and OBAA. It can be seen that performance of UBAA is better than that of Piazzo's algorithm when transmit power is higher than 32 dBm. Although the reduction of transmit power is not high, it is still of significantly importance. The reason is that in multi-user scenerio with frequency reuse, power reduction for this cell help reduce interference to other cells operating at the same spectrum.

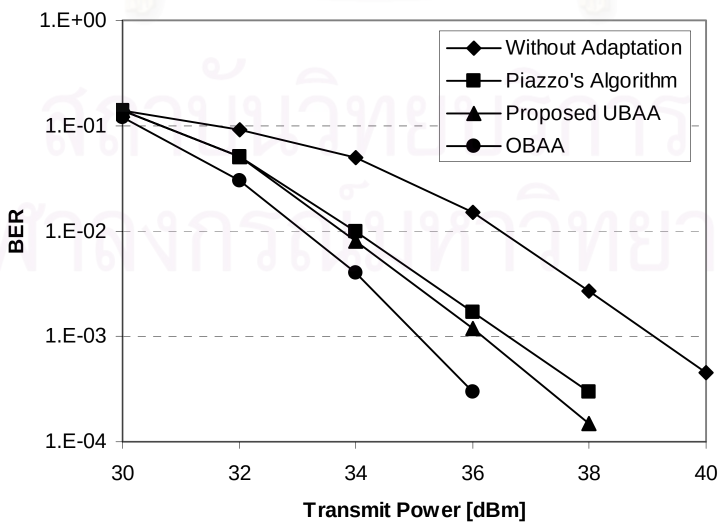

Figure 4.7: Adaptive modulation for MIMO-OFDM with partial CSI.

Complexity of adaptive modulation schemes in this thesis consist of operations such as comparision and look-up table accesses. Hughes-Hartogs requires $(N-1)b_{sum}/2$ comparision operations and $N-1+(b_{sum}/2)$ table accesses. Piazzo's algorithm requires $N\log_2 N$ comparision operations and $N-1+(b_{sum}/2)$ table accesses. Our proposed schemes requires $N\log_2 N$ comparision operations and $2N-1$ table accesses. In simulation, average number of bits per subchannel is set at $b_{avg}=2$. With this setting, complexity of UBAA is lower than that of Piazzo's algorithm. However, this is not the case when average number of bits per subchannel is high, $b_{avg}\geq 3$, when compared to constellation sizes in use.

Chapter V

Conclusions and Future Work

5.1 Conclusions

Multiple-input multiple-output (MIMO) systems promise very high spectral efficiency, with antenna arrays implemented at both transmitters and receivers. However, MIMO systems have no mechanism to combat frequency selective fading. Combined MIMO-OFDM systems are considered in this work for the downlink of wireless LAN (WLAN).

When channel state information (CSI) is available at transmitter via a feedback link, performance of wireless communication systems can significantly be improved. Since perfect CSI is difficult to achieve, assumption on partial CSI is rather feasible. We considered multiple antenna systems for both spatial diversity and spatial multiplexing. Alamouti code is initially proposed for transmit diversity. Alamouti code is adopted to transmitter design to increase robustness of MIMO-OFDM systems against uncertainties of CSI. In this work, we adopt the coder-beamformer design in [15] to increase robustness of MIMO-OFDM systems when only partial CSI is available at transmitter.

Optimal bit allocation is complicated, with only one bit loaded at a time to the subchannel which requires least extra power. Different sub-optimal schemes have been proposed in literature. In this work, we propose a sub-optimal scheme for bit allocation in MIMO-OFDM systems. Performance of our proposed scheme is evaluated by computer simulation. It has been shown that when average number of bits per subchannel is small when compared to constellation sizes in use, our proposed scheme has better performance when compared to existing sub-optimal algorithm.

5.2 Future Work

Results in this thesis can be extended in different ways. The first approach is to perform adaptation over the combined modulation and coding such as Trellis Coded Modulation schemes. To avoid complexity, only square QAM constellations (QPSK, 16-QAM, 64-QAM) are considered in this thesis. However, adaptive modulation over these constellations is of low flexibility and granularity.

Adaptive modulation for MIMO-OFDM for the downlink of WLAN is considered in this thesis. However, it is limited to single user scenerio. A straight way is to consider multi-user WLAN with cooperative signal processing. Performance of communication systems can be improved via spatial diversity when data for each user is transmitted from several access points.

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List of Publications

Tan Phamdinh, Wilaiporn Lee, Manil Cotonetr, and Somchai Jitapunkul, "Upward Bit Allocation Algorithm for Minimizing Transmit Power in MIMO-OFDM Systems," to appear, *Proceeding of IEEE ISCIT 2005*, Beijing, China, October 2005.

Tan Phamdinh, Wilaiporn Lee, Manil Cotonetr, and Somchai Jitapunkul, "Novel V-BLAST-based Adaptive Modulation for MIMO-OFDM Systems over Frequency Selective Fading Channels," to appear, *Proceeding of IEEE ICICS 2005*, Bangkok, Thailand, December 2005.

Upward Bit Allocation Algorithm for Minimizing Transmit Power in MIMO-OFDM Systems

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***Abstract*—Orthogonal frequency division multiplexing (OFDM) systems in combination with multiple-input multiple-output (MIMO) transmission are potential candidates for next generation Wireless Local Area Networks (WLAN). In this paper, we propose a novel algorithm for bit allocation, called upward bit allocation algorithm (UBAA), to minimize transmit power while maintaining bit error rate (BER) in MIMO-OFDM systems with fixed throughput. Since the optimal discrete bit allocation is of high complexity, UBAA with efficient logarithmic search is proposed as a sub-optimal approach. Performance of UBAA is evaluated by computer simulation and compared to that of optimal algorithm and existing sub-optimal schemes.**

***Keywords*—Multiple-input multiple-output (MIMO), adaptive modulation, OFDM, wireless LAN (WLAN).**

I. Introduction

Transmission over multiple-input multiple-output (MIMO) channel is a technique to maximize spectral efficiency. Antenna arrays are implemented at both transmitter and receiver. All transmit antennas use the same frequency spectrum. It was reported that MIMO systems with Vertical Bell Laboratories Layered Space-Time (V-BLAST) detection achieved spectral efficiency of up to 20-40 bits/sec/Hz [1].

Orthogonal frequency division multiplexing (OFDM) is a multi-carrier modulation technique for combating frequency selective fading. A data stream with high rate is split into multiple substreams with lower rate for transmission in parallel. OFDM technique is adopted in standards for Wireless Local Area Networks (WLAN) such as IEEE802.11a, HYPERLAN2 with transmission rate of up to 54 Mb/sec [2]. Specifications for OFDM systems in combination with MIMO transmission are being standardized for WLAN since spectral efficiency of MIMO can be effectively exploited in rich scattering indoor environment.

When channel state information (CSI) is available at transmitter via a feedback link, system performance can be significantly improved by adapting transmission to varying channel. Adaptive modulation has been studied extensively in literature [3]-[7]. Most research on adaptive modulation for wireless communications studied algorithms for either maximizing data rate or minimizing transmit power, developed from algorithms which were initially proposed for Discrete Multi-Tone (DMT) systems. In this work, we only consider adaptive modulation algorithms for minimizing transmit power while maintaining system performance.

Optimal bit allocation algorithm (OBAA) for DMT systems is proposed in [3]. One bit incremental unit is allocated at a time to the subcarrier which requires least extra power. Minimizing transmit power for MIMO-OFDM systems is addressed in [3]-[6]. Optimal solution for MIMO-OFDM systems is OBAA in combination with singular value decomposition (SVD). MIMO channel at each OFDM tone is decomposed into parallel spatial subchannels. OBAA is performed over those spatial subchannels. Since SVD-based OBAA for MIMO-OFDM is too complicated, V-BLAST-based OBAA for minimizing transmit power was proposed in [4] with much lower complexity.

As a sub-optimal approach to avoid complexity of OBAA, an efficient bit allocation scheme for OFDM systems was proposed in [5]. Bit allocation for the overall system is found by solving optimization problem iteratively. Details on this algorithm will be described in section 3.

The paper is organized as follows. In section 2, MIMO-OFDM system model with V-BLAST-based adaptive modulation and detection is briefly described. Section 3 introduces the power minimization problem with existing solutions. Proposed algorithm is described in section 4. Simulation results are given in section 5. Section 6 contains some concluding remarks.

II. System Description

Consider a MIMO-OFDM system with n_T transmit antennas and n_R receive antennas ($n_T \le n_R$). Each transmit antenna utilizes an OFDM modulator with K tones. It can be viewed as that there is one MIMO channel at each OFDM tone.

Each MIMO channel has n_T spatial subchannels so the total number of spatial subchannels available is $K \times n_T$. We denote $P_{k,s}$ and $b_{k,s}$ as power and number of bits, respectively, allocated to the s^{th} spatial subchannel at the k^{th} OFDM tone with $s \in \{1,2,\ldots,n_T\}$ and $k \in \{1,2,\ldots,K\}$.

At transmitter, $K \times n_T$ bit tuples are mapped to $K \times n_T$ complex QAM symbols. The Inverse Discrete Fourier Transform (IDFT) translates those QAM symbols to OFDM tones to form an OFDM symbol.

At receiver, OFDM demodulation is performed with the Discrete Fourier Transform (DFT). V-BLAST detection is performed for each OFDM tone separately. Only zero-forcing (ZF) nulling is considered for simplicity. Post-detection signal-to-noise (SNR) of each symbol can be expressed as

$$\rho_{k,s} = \frac{P_{k,s}}{\sigma_n^2 \left\| w_{k,s} \right\|^2} \qquad (1)$$

where σ_n^2 is noise variance, $\left\| w_{k,s} \right\|$ is norm of the nulling vector $w_{k,s}$. Spatial subchannel gain can be written as

$$g_{k,s} = 1/\left\| w_{k,s} \right\|^2. \qquad (2)$$

III. Power Minimization Problem

Here, for simplicity, only square QAM signal constellations QPSK, 16-QAM, 64-QAM are considered. Bit sequences of length b_{sum}, corresponding to one OFDM symbol, are allocated to $K \times n_T$ spatial subchannels. Constraint on fixed throughput is given by

$$\sum_{k=1}^{K}\sum_{s=1}^{n_T} b_{k,s} = b_{sum} \quad with \quad b_{k,s} \in \{0,2,4,6\} \qquad (3)$$

where $b_{k,s}$ is the number of bits corresponding to the QAM constellation of size $M_{k,s} = 2^{b_{k,s}} \in \{4,16,64\}$. A spatial subchannel with $b_{k,s} = 0$ is not used for information bearing and no power is allocated to that subchannel. All subchannels have the same desired bit-error-rate (BER)

$$\xi_{k,s} \le \xi \quad with \quad \forall k,s. \qquad (4)$$

Total transmit power P_{sum} is calculated by

$$P_{sum} = \sum_{k=1}^{K}\sum_{s=1}^{n_T} P_{k,s} \quad with \quad P_{k,s} \ge 0. \qquad (5)$$

Optimization problem can be stated as follows:

minimize total transmit power P_{sum} given in (5) subject to constraints (3) and (4).

To facilitate V-BLAST detection, same number of bits b_k is allocated to all subchannels at the k^{th} OFDM tone

$$b_{k,1} = b_{k,2} = \ldots = b_{k,n_T} = b_k \quad with \quad \forall k. \qquad (6)$$

Bit allocation over $K \times n_T$ spatial subchannels now reduces to bit allocation over K OFDM tones.

$$\sum_{k=1}^{K} b_k = \frac{b_{sum}}{n_T} \quad with \quad b_k \in \{0,2,4,6\}. \qquad (7)$$

Each OFDM tone is characterized by average tone gain defined as

$$\eta_k = \frac{1}{n_T}\sum_{s=1}^{n_T} g_{k,s} = \frac{1}{n_T}\sum_{s=1}^{n_T}\frac{1}{\left\| w_{k,s} \right\|^2}. \qquad (8)$$

Let $R_i(\xi)$ be the SNR required to transmit i bits per OFDM tone at BER ξ. Power required at the k^{th} OFDM tone can be written as

$$P_k = \frac{\sigma_n^2 R_{b_k}(\xi)}{\eta_k}. \qquad (9)$$

Total power to be minimized is

$$P_{sum} = n_T \times \sigma_n^2 \sum_{k=1}^{K}\frac{R_{b_k}(\xi)}{\eta_k}. \qquad (10)$$

Optimal solution OBAA is presented in [3] by introducing the parameter Γ

$$\Gamma = \frac{R_{b_k}(\xi)}{2^{b_k} - 1} \qquad (11)$$

which characterizes the difference between achievable performance and Shannon capacity. Each Γ corresponds to a specified value of BER. Equation (11) can be rewritten as

$$b_k = \log_2\left(1 + \frac{R_{b_k}(\xi)}{\Gamma}\right) = \log_2\left(1 + \frac{P_k \eta_k}{\sigma_n^2 \Gamma}\right). \qquad (12)$$

From (12), power required to load b_k bit to the k^{th} OFDM tone is

$$P_k = (2^{b_k} - 1)\frac{\sigma_n^2 \Gamma}{\eta_k}. \qquad (13)$$

We compare power required to load different number of bits while maintaining BER. From (13), we have

$$\frac{P_{k|b_k=4}}{P_{k|b_k=2}} = \frac{2^4 - 1}{2^2 - 1} = 5 \qquad (14)$$

and

$$\frac{P_{k|b_k=6}}{P_{k|b_k=2}} = \frac{2^6 - 1}{2^2 - 1} = 21. \qquad (15)$$

In OBAA, 2 bits are loaded at a time to the OFDM tone which requires smallest additional power. Although OBAA is optimal, it is too complicated, especially for systems utilizing large number of OFDM tones. Next, we briefly describe the algorithm proposed by Piazzo in [5]. For clarity, we will describe Piazzo's algorithm for adaptive modulation with square QAM constellations. A J-bit group is defined as the set of all OFDM tones which are allocated J bits per tone. Final solution is found by considering systems with 2 QAM constellations at a time, referred to as a two-mode system (TMS).

First, all bits in budget are allocated to OFDM tones in J-bit group corresponding to the constellation with largest size of the TMS being considered. Then bits are reallocated to $(J-2)$-bit group. In each reallocation, $2(J-2)$ bits from $(J-2)$ OFDM tones with smallest tone gains in J-bit group are moved to form 2 OFDM tones in $(J-2)$-bit group. New OFDM tones in $(J-2)$-bit group are taken from OFDM tones with largest gains in 0-bit groups.

Once optimal solution for the first TMS is determined, number of OFDM tones in J-bit group is fixed. Number of bits allocated to J-bit group is deduced from total bit budget. The same process is performed to solve optimization problem for new TMS with $(J-2)$-bit group and $(J-4)$-bit group. Although solution for each TMS is optimal, overall result is not guaranteed to be optimal. However, complexity of Piazzo's algorithm is much lower than that of OBAA.

IV. Proposed Algorithm

In this section, our proposed algorithm, referred to as upward bit allocation algorithm (UBAA), is described.

A. Bit Allocation

First, average tone gains are sorted in descending order

$$\eta_{k_1} \geq \eta_{k_2} \geq \ldots \geq \eta_{k_N} \tag{16}$$

where $k_i \in \{1, 2, \ldots, K\}$ is OFDM tone index.

Let α be index of the OFDM tone with largest gain, β be index of the OFDM tone with smallest gain in J-bit group. Total number of bits allocated to OFDM tones in J-bit group is denoted by b_{budget}. Two bits from the tone with index β are moved to the tone with index α. Now, OFDM tone with index α belongs to $(J+2)$-bit group, OFDM tone with index β belongs to $(J-2)$-bit group. Bit reallocation for OFDM tones in 4-bit group is shown in Fig. 1. Reallocation results in reducing transmit power if

$$\frac{R_{J+2}(\xi)}{\eta_\alpha} + \frac{R_{J-2}(\xi)}{\eta_\beta} < \frac{R_J(\xi)}{\eta_\alpha} + \frac{R_J(\xi)}{\eta_\beta}. \tag{17}$$

Reallocation is performed iteratively until transmit power can not be further reduced. Inequality (17) can be rewritten as

$$\frac{R_{J+2}(\xi) - R_J(\xi)}{\eta_\alpha} - \frac{R_J(\xi) - R_{J-2}(\xi)}{\eta_\beta} < 0. \tag{18}$$

From (11), we have

$$R_{J+2}(\xi) > R_J(\xi) > R_{J-2}(\xi). \tag{19}$$

Denote values of α and β after the m^{th} reallocation by $\alpha(m)$ and $\beta(m)$. Initially, we have $\alpha(0) = 1$ and $\beta(0) = b_{budget} / J$.

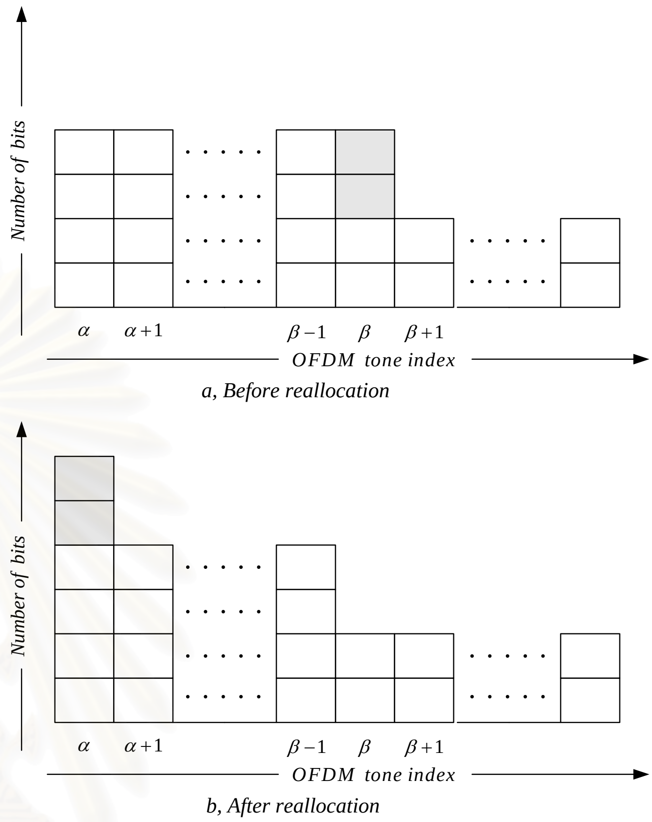

Fig. 1. Upward bit reallocation for OFDM tones in 4-bit group

After the m^{th} reallocation, we have $\alpha(m) = 1 + m$ and $\beta(m) = (b_{budget} / J) - m$. Substituting these values to (18), the left-hand side becomes a function of m

$$f(m) = \frac{R_{J+2}(\xi) - R_J(\xi)}{\eta_{(1+m)}} - \frac{R_J(\xi) - R_{J-2}(\xi)}{\eta_{((b_{budget}/J)-m)}}. \tag{20}$$

From (16) and (19), $f(m)$ is an increasing function with respect to m. The optimal solution is found by searching for the value of m that satisfies $f(m) \leq 0$ and $f(m+1) > 0$. The search is linear as m increases by one in each step. Linear search is replaced by efficient logarithmic search as in Piazzo's algorithm [5] since the function $f(m)$ is an increasing function. Complexity of adaptive modulation for systems with large number of OFDM tones is significantly reduced by logarithmic search. Logarithmic search halves the range at each step by testing the value of $f(m)$ at center of the range and updates the range accordingly.

Denote the minimum value and the maximum value of m by m_0 and m_1, respectively. Initial value of m_0 is zero and initial value of m_1 is determined from $m_1 = b_{budget} / J / 2$. UBAA can be mathematically described as follows

i, Initialize $m_0 = 0$, $m_1 = b_{budget} / J / 2$

ii, Compute $m_x = (m_0 + m_1)/2$

iii, If $f(m_x) < 0$ *let* $m_0 = m_x$ *; else* $m_1 = m_x$

iv, If $m_1 = m_0 + 1$ *go to v,; otherwise go to ii,*

v, Stop, $b_{budget} = (J+2)m_0$

B. Power Allocation

Total power assigned to the k^{th} OFDM tone is $n_T \times P_k$, determined from bit allocation process. This amount of power is further divided to spatial subchannels. Since the same QAM constellation of size M_k is used over all n_T spatial subchannels of each OFDM tone, BER can be written as [7]

$$\xi_{k,s} = \frac{4}{b_k}\left(\frac{\sqrt{M_k}-1}{\sqrt{M_k}}\right) Q\left(\sqrt{\frac{3P_{k,s}}{(M_k-1)\sigma_n^2 \left\| w_{k,s} \right\|^2}}\right). \quad (21)$$

From (4), all spatial subchannels are kept at the same BER

$$\xi_{k,s} = \xi \text{ with } \forall k, s. \quad (22)$$

Substituting (21) into (22), power allocation to each spatial subchannel is found in closed-form

$$P_{k,s} = \frac{\left\| w_{k,s} \right\|^2}{\sum_{s=1}^{n_T} \left\| w_{k,s} \right\|^2} \times n_T \times P_k. \quad (23)$$

V. Simulation Results

We use the 3-path Rayleigh fading channel model with exponential power delay profile as in [6]. Average number of bits per spatial subchannel is set at $b_{avg} = 2$. Simulation results for uncoded MIMO-OFDM systems with $N = 256$ and $n_T = n_R = 4$ are shown in Fig. 2. The four curves show transmit power required by different adaptation schemes: without adaptation, V-BLAST-based Piazzo's algorithm, proposed V-BLAST UBAA and V-BLAST-based OBAA. It can be seen that transmit power required in our proposed UBAA is 1.5 dB less than in Piazzo's algorithm at $BER = 10^{-3}$.

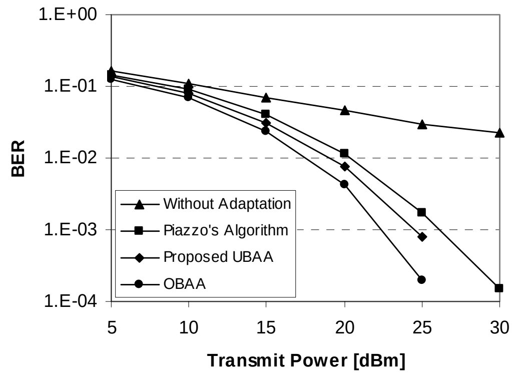

Fig. 2. Uncoded MIMO-OFDM systems with adaptive modulation.

VI. Conclusions

In this work, we proposed a novel V-BLAST-based adaptive modulation algorithm UBAA with logarithmic search. Simulation results show that UBAA requires 1.5 dB less power than Piazzo's algorithm at $BER = 10^{-3}$.

Acknowledgment

The authors would like to acknowledge AUN/SEED-Net for financial support via collaborative research.

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Novel V-BLAST-based Adaptive Modulation for MIMO-OFDM Systems over Frequency Selective Fading Channels

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***Abstract*—While most of existing adaptive modulation techniques for MIMO-OFDM systems focus on either maximizing data rate or minimizing transmit power, our work aims at minimizing average bit error rate (BER) under transmit power constraint with fixed throughput. In this paper, a novel adaptive modulation algorithm for minimizing BER is proposed by introducing additional constraints. Performance of proposed algorithm in wireless local area network (WLAN) environment is evaluated by computer simulation. Frequency selective fading channel model is simulated. Numerical results show that our proposed algorithm with low complexity yields better performance than existing suboptimal scheme under certain conditions.**

***Keywords*—Multiple-input multiple-output (MIMO), adaptive modulation, orthogonal frequency division multiplexing (OFDM).**

I. INTRODUCTION

Transmission over multiple-input multiple-output (MIMO) channel promises very high spectral efficiency. Antenna arrays are implemented at both transmitter and receiver using the same frequency spectrum. It was reported that MIMO systems with Vertical Bell Laboratories Layered Space-Time (V-BLAST) detection technique achieved spectral efficiency of up to 20-40 bits/sec/Hz [1].

Orthogonal frequency division multiplexing (OFDM) is a multi-carrier modulation technique to deal with frequency selective fading. A high-rate data stream is split into multiple sub-streams with lower rate for transmission in parallel. OFDM technique is adopted in wireless local area network (WLAN) standards such as IEEE802.11a of North America with transmission rate up to 54 Mb/sec [2]. OFDM systems in combination with MIMO transmission are being standardized for WLAN environment since spectral efficiency of MIMO can be fully exploited in rich scattering environment.

When channel state information (CSI) is available at transmitter via a feedback link, system performance can be significantly improved by adapting transmission to the varying channel. Adaptive modulation algorithms for bit and power allocation (BPA) have been studied extensively in literature [3], [4] and references therein. Most research on adaptive modulation for wireless communications studied algorithms for either maximizing data rate or minimizing transmit power since these algorithms are directly developed from algorithms which were initially proposed for Discrete Multi-Tone (DMT) systems. When perfect CSI is available at transmitter, optimal solution is the water-filling technique. Since optimal solution requires signal constellations of infinite-granularity, a suboptimal algorithm for bit and power allocation in DMT systems is proposed in [4]. To reduce complexity of adaptive modulation for multiple antenna systems, V-BLAST-based adaptive modulation for minimizing transmit power was proposed in [3] using quadrature amplitude modulation (QAM) constellations.

The problem of adaptive modulation for minimizing average bit error rate (BER) is addressed in [5] for OFDM systems with bit-interleaved coded modulation. Since exhaustive search for optimal bit allocation is not practical, a suboptimal scheme is proposed by sorting OFDM subcarriers into groups. Once bit allocation is determined, optimal power allocation is found by utilizing Lagrange multipliers. Transmission over a subset of OFDM subcarriers is addressed in [2]. The main idea is that performance loss due to the use of high-order constellation size can be partially compensated by the allocation of total power to fewer subcarriers.

In this work, we propose a novel adaptive modulation algorithm for MIMO-OFDM systems. Assume that perfect CSI is available at transmitter via a feedback link. Subchannel gains are determined from training period when power and bits are allocated equally. A subset of OFDM tones with largest gains is selected for data transmission. The same number of bits is loaded to all spatial subchannels associated with OFDM tones in that subset. Power is allocated to keep the same BER over all spatial sub-channels.

This paper is organized as follows. In section 2, MIMO-OFDM system model with V-BLAST detection is briefly described. Section 3 introduces the BER minimization problem with existing solutions. Proposed algorithm is described in section 4. Simulation results for MIMO-OFDM systems with different levels of adaptation are given in section 5. Section 6 contains some concluding remarks.

II. System Description

Consider an MIMO-OFDM system with n_T transmit antennas and n_R receive antennas ($n_T \le n_R$). Transmitter is equipped with n_T OFDM modulators for n_T transmit antennas. Denote $N_{\max}$ as the total number of OFDM tones which can be used for information bearing. These OFDM tones are indexed by elements of the set $\Omega_{\max} = \{1, 2, \ldots, N_{\max}\}$.

It can be considered that each OFDM tone is associated with one MIMO channel. Each MIMO channel has n_T spatial subchannels so the total number of spatial subchannels available is $N_{\max} \times n_T$.

Channel between the q^{th} transmit antenna and the p^{th} receive antenna is modelled as a finite impulse response (FIR) filter with L taps, denoted as $h_{pq}(l)$ with $l = \{0, 1, \ldots, L-1\}$. Let $h_{pq}(l)$ be the element at the p^{th} row and the q^{th} column of channel matrix $H(l)$. Discrete-time signal model can be written as

$$r(\tau) = \sum_{l=0}^{L-1} H(l)x(\tau - l) + n(\tau), \tag{1}$$

with $x(\tau)$ is transmit signal vector of size $n_T \times 1$, $r(\tau)$ is receive signal vector of size $n_R \times 1$ and $n(\tau)$ is noise vector of size $n_R \times 1$.

With bit and power allocation determined from adaptive modulation algorithm, power and bits are allocated to a subset of N OFDM tones with $N \le N_{\max}$. OFDM tones in this subset are indexed by elements of the set $\Omega = \{ f_i \mid f_i \in \Omega_{\max}, i = 1, 2, \ldots, N \}$. Denote $P_{f_i,s}$ and $b_{f_i,s}$ as power and number of bits, respectively, allocated to the s^{th} spatial subchannel of the f_i^{th} OFDM tone. Let $P_{f_i,s}$ and $b_{f_i,s}$ be elements of power allocation vector P and bit allocation vector b, respectively.

At transmitter, bit sequences of length b_{sum} are split into $N \times n_T$ bit tuples, mapped to complex QAM symbols for transmission over the subset of N OFDM tones. For simplicity, only square QAM constellations (QPSK, 16-QAM, and 64-QAM) are considered. The Inverse Discrete Fourier Transform (IDFT) translates those QAM symbols to OFDM tones. Cyclic prefix is inserted to eliminate intersymbol interference (ISI).

At receiver, OFDM demodulation is performed with cyclic prefix removal and the Discrete Fourier Transform (DFT). Bit and power allocation achieved from adaptive modulation algorithm is utilized for V-BLAST detection. V-BLAST detection is performed for each OFDM tone separately. Detection order is kept the same as in training period. Only zero-forcing (ZF) nulling is considered for simplicity. Post-detection signal-to-noise (SNR) of each symbol can be expressed as

$$\rho_{f_i,s} = \frac{P_{f_i,s}}{\sigma_n^2 \left\| w_{f_i,s} \right\|^2}, \tag{2}$$

where σ_n^2 is noise variance, $\left\| w_{f_i,s} \right\|$ is norm of the nulling vector $w_{f_i,s}$.

III. BER Minimization Problem

Bit sequences of length b_{sum}, corresponding to one OFDM symbol, are allocated to $N \times n_T$ spatial subchannels. Fixed rate constraint is given by

$$\sum_{i=1}^{N} \sum_{s=1}^{n_T} b_{f_i,s} = b_{sum} \quad with \quad b_{f_i,s} \in \{2, 4, 6\}, \tag{3}$$

where $b_{f_i,s}$ is the number of bits corresponding to the QAM constellation of size $M_{f_i,s} = 2^{b_{f_i,s}} \in \{4, 16, 64\}$.

Constraint on total transmit power P_{sum} is expressed by

$$\sum_{i=1}^{N} \sum_{s=1}^{n_T} P_{f_i,s} = P_{sum} \quad with \quad P_{f_i,s} \ge 0. \tag{4}$$

Average BER can be approximated as [5]

$$\begin{aligned} \bar{\xi} &\approx \frac{4}{b_{sum}} \sum_{i=1}^{N} \sum_{s=1}^{n_T} \frac{\sqrt{M_{f_i,s}} - 1}{\sqrt{M_{f_i,s}}} Q\left(\sqrt{\frac{3\rho_{f_i,s}}{M_{f_i,s} - 1}} \right) = \\ &= \frac{4}{b_{sum}} \sum_{i=1}^{N} \sum_{s=1}^{n_T} \frac{\sqrt{M_{f_i,s}} - 1}{\sqrt{M_{f_i,s}}} Q\left(\sqrt{\frac{3P_{f_i,s}}{\left(M_{f_i,s} - 1\right)\sigma_n^2 \left\| w_{f_i,s} \right\|^2}} \right), \end{aligned} \tag{5}$$

where the function $Q(\cdot)$ is defined as

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} e^{-z^2/2} dz. \tag{6}$$

Optimization problem can be stated as follows:

minimize average BER $\bar{\xi}$ given in (5)

subject to constraints (3) and (4).

Optimal bit allocation can be found in various ways. The first method comes from the observation that number of bits allocated to each spatial subchannel belongs to a small set of {2, 4, 6}. Exhaustive search can be performed to find the optimal bit allocation. This method is not practical due to high complexity for systems with large number of OFDM tones and antennas. Another method is to use Lagrange multipliers to solve the optimization problem. Constraints on total power and total number of bits are taken into account by defining

$$X(b) = \sum_{i=1}^{N} \sum_{s=1}^{n_T} b_{f_i,s} - b_{sum}, \tag{7}$$

and

$$Y(P) = \sum_{i=1}^{N} \sum_{s=1}^{n_T} P_{f_i,s} - P_{sum}. \tag{8}$$

Constraint on discrete number of bits {2, 4, 6} allocated to spatial subchannels is taken into account by

$$Z(b_{f_i,s}) = (b_{f_i,s} - 2)(b_{f_i,s} - 4)(b_{f_i,s} - 6). \quad (9)$$

Defining the vector of Lagrange multipliers

$$L = \{\alpha, \beta, \gamma_{f_1,1}, \ldots, \gamma_{f_N,n_T}\}. \quad (10)$$

Cost function can be expressed by

$$F(b,P,L) = \overline{\xi} - \alpha X(b) - \beta Y(P) - \sum_{i=1}^{N}\sum_{s=1}^{n_T} \gamma_{f_i,s} Z(b_{f_i,s}), \quad (11)$$

with parameters defined in [7]-[10].

Unfortunately, this optimization problem has not been solved up to now. A suboptimal approach for bit allocation in OFDM systems is described in [5]. Subcarriers are sorted into 3 groups based on channel gains, subcarriers in each group adapt the same signal constellation, for example: 64-QAM for subcarriers with largest SNRs, QPSK for subcarriers with smallest SNRs and 16-QAM for subcarriers in the remaining group.

Once bit allocation is determined, optimal power allocation can be found by reducing (11) to

$$F(P,\beta) = \overline{\xi} - \beta Y(P), \quad (12)$$

where β is a Lagrange multiplier. By taking partial derivatives and setting these values to zeros, we have

$$\frac{\sqrt{M_{f_i,s}} - 1}{\sqrt{2\pi}\sqrt{M_{f_i,s}}} \exp\left(-\frac{1}{2} P_{f_i,s} U_{f_i,s}\right) \frac{\sqrt{U_{f_i,s}}}{2\sqrt{P_{f_i,s}}} + \beta = 0, \quad (13)$$

where

$$U_{f_i,s} = \frac{3}{\left(M_{f_i,s} - 1\right)\sigma_n^2 \left\| w_{f_i,s} \right\|^2}. \quad (14)$$

Equation (13) can be rewritten as

$$V_{f_i,s} P_{f_i,s} \exp\left(P_{f_i,s} U_{f_i,s}\right) = \Lambda, \quad (15)$$

where Λ is a positive constant and

$$V_{f_i,s} = \frac{M_{f_i,s}\left(M_{f_i,s} - 1\right)\left\| w_{f_i,s} \right\|^2}{\left(\sqrt{M_{f_i,s}} - 1\right)^2}. \quad (16)$$

Once bit allocation is specified, optimal power allocation is determined from

$$P_{f_i,s} = \frac{1}{U_{f_i,s}} W\left(\frac{U_{f_i,s}}{V_{f_i,s}} \Lambda\right), \quad (17)$$

where $W(\cdot)$ is the Lambert W-function, defined as the inverse of the function $f(\omega) = \omega e^{\omega}$ [6]. We will refer to the combination of suboptimal bit allocation and optimal power allocation presented in this section as group-based BPA. Performance of group-based BPA for MIMO-OFDM systems will be evaluated by computer simulation and compared to our proposed scheme which will be described in the next section.

IV. Proposed Algorithm

In this section, a simplified adaptive modulation algorithm for MIMO-OFDM systems is proposed. V-BLAST-based adaptive modulation algorithm is performed at both transmitter and receiver with perfect CSI. Spatial subchannel gains are achieved in training period by allocating bits and power equally over all $N_{\max} \times n_T$ spatial subchannels. From (2), spatial subchannel gain can be written as

$$g_{f,s} = \frac{1}{\left\| w_{f,s} \right\|^2}, \quad (18)$$

for $f \in \Omega_{\max}$ and $s \in \{1, 2, \ldots, n_T\}$. Each OFDM tone is characterized by average tone gain defined as

$$\eta_f = \frac{1}{n_T}\sum_{s=1}^{n_T} g_{f,s} = \frac{1}{n_T}\sum_{s=1}^{n_T} \frac{1}{\left\| w_{f,s} \right\|^2}. \quad (19)$$

A. *Bit allocation*

Proposed approach is suboptimal, achieved by imposing additional constraints. The first constraint is that the same number of bits is allocated to all spatial subchannels associated with N OFDM tones. Bit allocation over MIMO-OFDM spatial subchannels now reduces to OFDM tone subset selection.

A subset of N OFDM tones with largest average tone gains is selected from $N_{\max}$ OFDM tones. The same number of bits,

$$b = \frac{b_{sum}}{N \times n_T}, \quad (20)$$

is loaded to spatial subchannels associated with OFDM tones in this subset. This can be described mathematically as follows

Initialization

$\Omega = \varnothing$

$\Omega_{\max} = \{1, 2, \ldots, N_{\max}\}$

$f \in \Omega_{\max}$

$s \in \{1, 2, \ldots, n_T\}$

$b_{f,s} \leftarrow 0$

$i \leftarrow 1$

Iteration

$f_i \leftarrow \arg\max_{\Omega_{\max} \backslash \Omega} \eta_f$

$\Omega = \Omega \cup \{f_i\}$

$b_{f_i,s} \leftarrow b$

$i \leftarrow i + 1$

Until $i = N + 1.$

B. Power allocation

BER of each spatial subchannel is given as in [5]

$$\xi_{f_i,s} = \frac{4}{b_{f_i,s}}\left(\frac{\sqrt{M_{f_i,s}}-1}{\sqrt{M_{f_i,s}}}\right)Q\left(\sqrt{\frac{3P_{f_i,s}}{(M_{f_i,s}-1)\sigma_n^2\left\|w_{f_i,s}\right\|^2}}\right). \qquad (21)$$

As the same QAM constellation of size $M = 2^b$ is used over all $N \times n_T$ spatial channels, (21) can be rewritten as

$$\xi_{f_i,s} = \frac{4}{b_{f_i,s}}\left(\frac{\sqrt{M}-1}{\sqrt{M}}\right)Q\left(\sqrt{\frac{3P_{f_i,s}}{(M-1)\sigma_n^2\left\|w_{f_i,s}\right\|^2}}\right). \qquad (22)$$

The second constraint is that all spatial subchannels are kept at the same BER

$$\xi_{f_1,1} = \ldots = \xi_{f_i,s} = \ldots = \xi_{f_N,n_T}. \qquad (23)$$

Substituting (22) into (23), we get

$$\frac{P_{f_1,1}}{\left\|w_{f_1,1}\right\|^2} = \cdots = \frac{P_{f_i,s}}{\left\|w_{f_i,s}\right\|^2} = \cdots = \frac{P_{f_N,n_T}}{\left\|w_{f_N,n_T}\right\|^2}. \qquad (24)$$

Solving equations in (24), power allocation to each spatial subchannel is given in closed-form

$$P_{f_i,s} = \frac{\left\|w_{f_i,s}\right\|^2}{\sum_{i=1}^{N}\sum_{s=1}^{n_T}\left\|w_{f_i,s}\right\|^2}P_{sum}. \qquad (25)$$

It means that more power should be allocated to compensate for subchannels with low gains. Note that V-BLAST detection is performed for each OFDM tone separately and detection order is kept the same as in training period.

V. SIMULATION RESULTS

We use six-path Rayleigh fading channel model with exponential power delay profile, r. m. s. delay spread is set at 100ns. OFDM parameter settings are taken from IEEE802.11a standard with 64-point IDFT and DFT. Both transmitter and receiver are equipped with antenna arrays of 4 elements. A subset of $N = 24$ OFDM tones is selected from $N_{\max} = 48$. Performance of uncoded MIMO-OFDM systems is shown in Fig. 1. The four curves show average BER of MIMO-OFDM systems with different levels of adaptation: without adaptation, group-based BPA (described in section 3), our proposed BPA and optimal BPA. Optimal BPA is achieved with exhaustive search. Optimal power allocation is determined from bit allocation as in section 3. Our proposed algorithm has better performance than the group-based BPA at E_b / N_0 lower than 20.5 dB. However, performance improvement diminishes with E_b / N_0 larger than 20.5 dB. This can be explained such that performance degradation due to the use of high-order constellations in group-based BPA can be compensated when total transmit power is large enough.

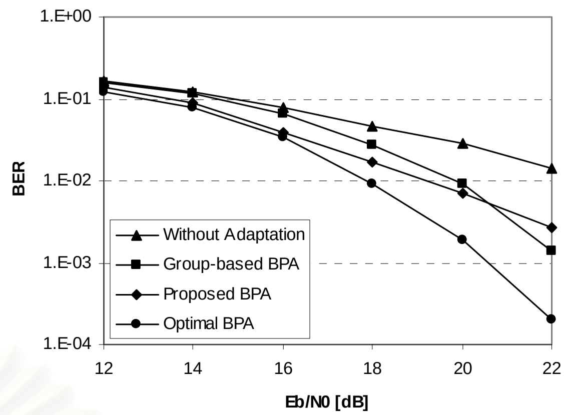

Figure 1. Performance comparison of MIMO-OFDM systems with V-BLAST detection at different levels of adaptation

VI. CONCLUSIONS

A simple adaptive modulation algorithm for MIMO-OFDM systems has been proposed. Bit and power allocation is found by performing V-BLAST-based adaptive modulation with perfect CSI at both transmitter and receiver. By introducing additional constraints, our proposed algorithm is simple to implement with good performance at low E_b / N_0.

ACKNOWLEDGMENT

The authors gratefully acknowledge AUN/SEED-Net for financial support via collaborative research.

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VITAE

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