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LINEAR SPACE-TIME MODULATION IN MULTIPLE-ANTENNA CHANNELS

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Abstract

This thesis develops linear space–time modulation techniques for (multi-antenna) multi-input multi-output (MIMO) and multiple-input single-output (MISO) wireless channels. Transmission methods tailored for such channels have recently emerged in a number of current and upcoming standards, in particular in 3G and "beyond 3G" wireless systems. Here, these transmission concepts are approached primarily from a signal processing perspective.

The introduction part of the thesis describes the transmit diversity concepts included in the WCDMA and cdma2000 standards or standard discussions, as well as promising new transmission methods for MIMO and MISO channels, crucial for future high data-rate systems. A number of techniques developed herein have been adopted in the 3G standards, or are currently being proposed for such standards, with the target of improving data rates, signal quality, capacity or system flexibility.

The thesis adopts a model involving matrix-valued modulation alphabets, with different dimensions usually defined over *space* and *time*. The symbol matrix is formed as a linear combination of symbols, and the space-dimension is realized by using multiple transmit and receive antennas. Many of the transceiver concepts and modulation methods developed herein provide both spatial multiplexing gain and diversity gain. For example, full-diversity full-rate schemes are proposed where the symbol rate equals the number of transmit antennas. The modulation methods are developed for open-loop transmission. Moreover, the thesis proposes related closed-loop transmission methods, where space–time modulation is combined either with automatic retransmission or multiuser scheduling.

Keywords: Space–time coding, modulation, multiple-input multiple-output (MIMO) channel, open-loop transmission, closed-loop transmission.

Preface

Most of results documented here have been developed at Nokia Research Center in Helsinki, Finland in recent years. I would like to express my appreciation to a number of colleagues for fruitful collaboration in the research areas considered in this thesis. In particular, many of the open-loop transmission methods considered here were developed with Dr. Olav Tirkkonen. Similarly, the closed-loop concepts described in Chapter 2 were developed in collaboration with Dr. Risto Wichman. In addition, fruitful discussions with Dr. Jussi Vesma and Dr. Niko Nefedov are acknowledged. Dr. Kari Kalliojärvi provided a number of constructive comments throughout the course of the work and supported the writing of the thesis introduction by enabling the author to step outside the daily project responsibilities when needed. In addition, project support from Dr Jorma Lilleberg at Nokia Technology Platforms is greatly appreciated. Prof. Olli Simula is acknowledged for supporting the work and smoothing many of the necessary steps in the final stages of the thesis work. In addition, I would also like to thank a number of people at John Wiley & Sons, for setting deadlines, and for unlimited patience when writing our book “Multi-antenna transceiver techniques for 3G and beyond” (see [1]). Parts of this book form the skeleton for the introduction part of the thesis. Finally, financial support from Nokia Foundation is acknowledged.

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Introduction

A number of fundamentally new modulation and coding methods have been invented in the last decade. Turbo codes [2, 3], proposed in 1993, are able to approach channel capacity limit as derived by Shannon in the 1940s [4]. In the late 1990s another major leap in modulation and coding theory was provided by Tarokh et. al. [5–7] and Alamouti [8] with the invention of space–time codes. Roughly at the same time Foschini [9] and Telatar [10] proved a way to increase channel capacity by efficient use of spatial dimension. Some of these concepts fall under the general term “transmit diversity” and some under “MIMO modulation”.

Transmit diversity is not an entirely new concept. Concepts proposed by Wittneben [11] and Hiroike et. al [12] approach diversity via effective signal processing solutions. These early papers lack the coding aspects of the signal design problem, but are often simple to implement and enjoy the support of engineers, if not the coding theorists. Transmit diversity solutions, or multiple-input single-output (MISO) concepts, provide a diversity or performance gain, but not necessarily the spectral efficiency gain. The spectral efficiency gain requires rigorous exploitation of multiple-input multiple-output (MIMO) channels and involves the use of multi-antenna transmission techniques.

From an engineering perspective MIMO channels have been known in wireless communications for some twenty years [13]. However, only when the MIMO capacity expressions were explicitly derived in [9, 10] the research area started to gain momentum. It was essentially shown that under certain conditions the capacity increases linearly with $\min(N_\mathrm{t}, N_\mathrm{r})$, where N_t is the number of deployed transmit antennas and N_r is the number of receive antennas.

MISO concepts that require only one receive antenna have gained popularity in standardization arenas along with efficient channel coding schemes. Eventually both Turbo codes and transmit diversity concepts found a way to both 3G systems, WCDMA and cdma2000. The keen adoption of new technology explains in part the technological merits of 3G systems, when compared to 2G systems. It is in part due to these advances that third generation systems, such as WCDMA [14, 15], provide enhanced system capacity, better services and significantly higher data rates when compared to 2G systems such as GSM or IS-95 [16]. Indeed, the first release of the 3G wideband CDMA standard developed within the 3GPP [17] applies an orthogonal space–time block code [8] and two transmit diversity schemes using feedback control [8, 18, 19]. Extensions of these concepts have been proposed for more than two transmit antennas, see e.g. [20].

Wireless standards are under continuous development and it is anticipated that some future physical layer standard release will contain further enhancements in terms of multi-antenna solutions. A solution adopted to a practical multi-access system should address the diversity-multiplexing tradeoffs [21], multiuser interference [22] and scheduling aspects [23, 24] appropriately.

1.1 SCOPE AND STRUCTURE OF THE THESIS

The thesis contributes to signal transmission techniques in MIMO and MISO channels. New efficient space–time modulation methods are developed in particular for open-loop transmission. The thesis also develops new closed-loop transmission, retransmission and multiuser diversity solutions for use with space–time modulation alphabets. Some closed-loop transmission techniques developed during the course of this work have been adopted in 3G WCDMA system (WCDMA closed-loop Mode 1) and are currently already on the market. These solutions are briefly discussed in the introduction part of the thesis, in Chapter 3.

The thesis is structured as follows. The introductory part, that you are currently reading, includes a view of the MISO and MIMO landscape and captures some relevant results by other researchers and the author, e.g. the closed-loop Mode 1 stated above [25]. A small part of the material presented here is based on the author's contributions to our book A. Hottinen, O. Tirkkonen, R. Wichman, *Multi-antenna transceiver techniques for 3G and beyond*. 2003. Copyright John Wiley and Sons Ltd (Reproduced with permission).

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Chapter 2 discusses the diversity resources that are currently accessible for 3G systems in a general level. Chapter 3 addresses the MIMO capacity notions, ergodic capacity and outage capacity. It is shown how ergodic capacity behaves under correlated fading both with and without channel state information at transmitter. Chapter 4 summarizes a representative set of MIMO and MISO modulation methods, with references to the original publications [P1]-[P8], listed below.

The main contributions of the thesis are included in publications [P1]-[P8]. Their content is summarized in the following section. In addition to the main publications, a non-exhaustive list of related publications is given. Only those publications that are directly relevant to the topics addressed in the thesis are listed. Moreover, the author holds approximately 40 patents and only the most relevant of them, in view of this thesis, are listed in references (see e.g. www.uspto.gov for an up-to-date list of US patents). For example, the author and the co-author of [P3] were granted a Finnish patent for code constructions developed therein.

1.2 LIST OF PUBLICATIONS AND AUTHOR'S CONTRIBUTIONS

[P1] A. Hottinen and O. Tirkkonen, "A randomization technique for non-orthogonal space-time block codes," In *Proc. IEEE Vehicular Technology Conference*, Rhodes, Greece, pp. 1479–1482, May 2001

[P2] O. Tirkkonen and A. Hottinen, "Improved MIMO performance with non-orthogonal space-time block codes," In *Proc. IEEE Global Telecommunications Conference*, San Antonio, Texas, USA, pp. 1122–1126, November 2001

[P3] O. Tirkkonen and A. Hottinen, "Square matrix embeddable space-time block codes for complex signal constellations," *IEEE Transactions on Information Theory*, Vol. 48, No. 2, pp. 384–395, February 2002

[P4] A. Hottinen and O. Tirkkonen, "Non-orthogonal space-time block code with symbol rate two," In *Proc. Conf. Inf. Sci. Syst.*, Princeton, NJ, USA, March 2002

[P5] A. Hottinen and O. Tirkkonen,"Matrix modulation and adaptive retransmission," in *Proc. Seventh International Symposium on Signal Processing and its Applications*, Paris, France, pp. 221–224, July 2003

[P6] A. Hottinen,"Multiuser scheduling with matrix modulation," in *Proc. IEEE International Symposium on Signal Processing and Information Technology*, Darmstadt, Germany, pp. 5–8, December 2003

[P7] A. Hottinen,“Matrix-modulated closed-loop MIMO with multiuser scheduling,” in *Proc. Conf. Inf. Sci. Syst.*, Princeton, NJ, USA, March 2004

[P8] A. Hottinen and O. Tirkkonen, “Precoder designs for high rate space–time block codes,” in *Proc. Conf. Inf. Sci. Syst.*, Princeton, NJ, USA, March 2004

Publication [P1] describes a randomization technique for non-orthogonal space–time coding concepts, for use with channel coded multiple-antenna systems. The focus is on linear modulation methods that achieve symbol rate one. The paper was written by the author, with constructive comments from the second author. This publication builds directly on publication [R11] (see the list next section), where the first symbol rate one non-orthogonal space–time block code for use with four transmit antennas is proposed.

Publication [P2] develops a symbol rate two quasi-orthogonal MIMO transmission method that provides a high coding gain within a class of concepts for which all symbols get the same received power. The concept was invented jointly by the paper authors [26] and the paper was written by O. Tirkkonen with comments provided by the present author. The author of the thesis contributed in particular to the underlying idea of transmitting two different space–time block codes simultaneously over the same MIMO channel.

Publication [P3] develops a theory for orthogonal space–time block codes for complex signal constellations. The concepts and the motivation for the paper were invented by the paper authors who were granted a Finnish patent on related space–time coding methods with international applications pending [27]. The paper reiterates and extends the work in [R7] and [R9] (see list below), and was written by O. Tirkkonen with constructive comments provided by the present author.

Publication [P4] proposes a symbol rate two non-orthogonal space–time block code using quasi-orthogonal layers, designed in particular for cases where the number of transmit antennas is larger than the number of receive antennas. The concept proposed in the paper was invented jointly by the paper authors, paper was written by the current author and constructive comments were provided by O. Tirkkonen.

Publication [P5] proposes a novel retransmission concept for use with matrix modulated systems. The concept was invented by the present author [28]. The author wrote the main part of the paper, while the second author clarified the text and provided comments.

Publication [P6] proposes a novel scheduling criteria for use with matrix modulation. The underlying idea in the paper was invented by the present author, and it also otherwise completely author’s own work.

Publication [P7] continues on [P6] in applying the scheduling criteria in [P6] to closed-loop systems. The underlying idea in the paper was invented by the present author, and it is also otherwise completely author's own work.

Publication [P8] describes power efficient MIMO modulation methods with high coding gains. In particular, the paper proposes new complex precoders for use with two transmit and receive antennas and improves on previous work [1, 29], and demonstrates the effectiveness of related designs in frequency-selective channels. The concepts proposed in the paper were invented jointly by the paper authors and most of the paper was written by the current author.

1.3 RELATED PUBLICATIONS

[R1] A. Hottinen and R. Wichman, " Transmit diversity by antenna selection in CDMA downlink," in *Proc. IEEE ISSSTA*, Sun City, South Africa, September 1998

[R2] A. Correia, A. Hottinen, and R. Wichman, "Optimized constellations for transmit diversity," in *Proc. Vehicular Technology Conference*, Amsterdam, September, 1999

[R3] A. Hottinen and R. Wichman, "Soft-weighted transmit diversity for WCDMA," in *Proc. Allerton Conference on Communications and Computing*, Illinois, USA, September 1999.

[R4] R. Wichman and A. Hottinen, "Transmit diversity in the WCDMA system," *Int. Journal of Wireless Information Networks*, Volume 6, Number 3, July 1999

[R5] A. Hottinen and R. Wichman, "Transmit diversity using filtered feedback weights in the FDD/WCDMA System," in *Proc. Int. Zurich Seminar on Communications*, Zurich, Switzerland, Feb. 2000

[R6] M. Raitola, A. Hottinen and R. Wichman, "Transmission diversity in wideband CDMA," in *Proc. 49th IEEE Vehicular Technology Conference*, May 16 - 19, 1999, Houston, Texas, USA 1999, 1545 - 1549

[R7] O. Tirkkonen, A. Hottinen, "The algebraic structure of space-time block codes," in *Proc. Finnish Wireless Communications Workshop 2000 (FWCW' 00)*, Oulu, Finland, pp. 80-84, May 2000

[R8] A. Hottinen, O. Tirkkonen and R. Wichman, "Closed-loop transmit diversity techniques for multi-element transceivers," in *Proc. Vehicular Technology Conference*, Boston, Mass. USA September 2000

[R9] O. Tirkkonen, A. Hottinen, “Complex modulation space-time block codes for four Tx antennas,” in *Proc. Globecom 2000*, San Francisco, CA, USA, November 2000.

[R10] B. Raghothaman, A. Boariu, O. Tirkkonen, A. Hottinen, ”Performance of simple space-time block codes for more than two transmit antenna,” in *Proc. Allerton Conf.*, September 2000

[R11] O. Tirkkonen, A. Boariu, A. Hottinen, “Minimal non-orthogonality rate 1 space-time block code for 3+ Tx Antennas,” in *Proc. IEEE sixth international symposium on spread spectrum techniques and applications (ISSSTA 2000)*, Parsippany, NJ, USA, pp. 429-432, September 2000.

[R12] A. Hottinen, R. Wichman, “Enhanced filtering for feedback mode transmit diversity,” in *Proc. Conf. Inf. Sci. Syst.*, Princeton, NJ, USA, March 2000

[R13] A. Hottinen, R. Wichman, “A closed-loop transmit diversity concept for WCDMA systems,” in *Proc. Conf. Inf. Sci. Syst.*, Baltimore, MD, USA, March 2001

[R14] A. Hottinen, O. Tirkkonen, and R. Wichman, “Multi-antenna transmission with feedback for WCDMA systems,” in *Proc. 3G Wireless*, San Francisco, USA, May 2001.

[R15] A. Hottinen, K. Kuchi and O. Tirkkonen, “A space–time coding concept for a multi-element transmitter,” in *Proc. Canadian Workshop on Information Theory* Vancouver, Ca., June 2001.

[R16] O. Tirkkonen and A. Hottinen, “Tradeoffs between rate, puncturing and orthogonality in space-time block codes,” in *Proc. ICC ’01*, Helsinki, Finland

[R17] A. Hottinen and R. Wichman, “Asymmetric quantization of feedback beams in WCDMA,” in *Proc. Conf. Inf. Sci. Syst.*, Princeton, NJ, USA, March 2002

[R18] A. Hottinen, J. Vesma, O. Tirkkonen, N. Nefedov, ”High Bit Rates for 3G and Beyond Using MIMO Channels,” in *Proc. PIMRC 2002*, Portugal, 2002

[R19] A. Hottinen, J. Vesma, O. Tirkkonen, ”High Bit Rates for HSDPA Using MIMO Channels,” *WSEAS Tr. Comm.*, July 2002

[R20] A. Hottinen, O. Tirkkonen and R. Wichman, *Multi-antenna transceiver techniques for 3G and beyond*, John Wiley Sons, Chichester, England, January 2003

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Some publications in this list are included also in the References with reference numbers. They are given here labels [R1]–[R20] for reader's convenience. A few words related to these publications is due.

[R1] is the outcome of research that started feedback mode studies for 3G systems. [R2] discusses the use of complex precoders in conjunction with transmit diversity. [R3] proposes a weighted space-time block code for use with feedback to increase robustness to feedback errors. A related concept has been discovered independently in [30]. [R4] summarizes the status of WCDMA transmit diversity, as of publication date. [R5] is the first publication on WCDMA closed-loop transmit diversity mode 1, invented (and patented) by the authors of the paper [25]. This publication and the related part of the standard specification [14] constitute the main engineering contribution of the author in the sense that the support for the developed concept is currently implemented in all WCDMA terminals. [R6] presents simulation results on transmit diversity in the WCDMA system. [R7] is first instance where Clifford algebra-based space–time block codes are developed, together with [R9]. [R8] proposes novel multi-antenna transceivers with feedback, together with [R12], [R13], and [R14]. [R10] introduces the ABBA transmission method [31], used also in [P1]. ABBA has been independently discovered in a slightly different form [32,33]. [R15] describes a four-antenna open-loop transmission concept, invented by the authors, that is still today being proposed in 3GPP, and is potentially included in some future standard release. [R16] attempts to increase the bit rate for space–time block codes by multimodulation. [R17] proposes novel techniques for feedback mode transmit diversity for structured channels. In the proposed concept dominant eigenbeams are quantized with higher resolution that less dominant. [R18] and [R19] are summary papers, wherein closed-loop concepts and open-loop concepts are compared. [R20] is the summary of author's and co-authors' work over the past years, and also constitutes a skeleton for the introduction part of this thesis.

2

Diversity and Capacity Enhancement in Wireless Systems

This chapter summarizes a number of capacity enhancement and diversity techniques available to wireless systems. The primary focus is placed on concepts adopted to 3G and, in part, 2G wireless systems. Some of the diversity concepts involving multi-antenna transceivers, including methods developed in this thesis, were originally designed for the Universal Terrestrial Radio Access (UTRA) WCDMA system. While the description below is elaborated for the WCDMA system, related concepts in cdma2000 and GSM evolutions are also summarized.

2.1 WCDMA

The WCDMA standard incorporates a number of diversity concepts aimed at mitigating the effects of fading in a radio propagation environment. In particular, WCDMA Release '99 and Release 4 support

- multipath diversity (frequency selectivity)
- time diversity using Automatic Repeat ReQuest (time selectivity)
- Rx diversity, using multiple receive antennas (antenna diversity)
- Tx diversity, with one open and two closed loop solutions (transmit diversity)
- soft handover (macro diversity)

2.1.1 Multipath Diversity

Due to a wideband channel (with chip rate 3.84Mcps in WCDMA) the receiver is able to resolve a large number of multipath components. Each multipath component typically faces an independent (or different) channel realization and the combined energy, weighted and integrated appropriately over each component, is subject to reduced signal fading when compared to any individual component. The embedded diversity may be captured by a linear or a non-linear receiver, e.g. by the RAKE receiver [34, 35] or channel equalizer [36]. Clearly, different environments have different multipath spreads and the number of resolvable components is sometimes small. For example, in indoor channels, the delayed components arrive predominantly within chip duration (inverse of chip rate), and only one channel coefficient (or tap) is resolvable. In such environments alternative forms of diversity are needed.

2.1.2 Macro Diversity

Macro diversity creates antenna diversity by utilizing the network in a efficient manner. A signal transmitted by a mobile station in uplink propagates to multiple base stations, and since the channel coefficients to each base station are independent, the signal combined over all base stations enjoys diversity. On the other hand, due to limited bandwidth in the fixed network between the base stations, optimal signal combining (in the spirit of diversity antennas) is not feasible. Nevertheless, at least selection-type combining is possible, in the sense that it is sufficient to receive the transmitted signal correctly in at least one base station. In downlink, multiple copies of the same signal are transmitted from spatially separate source locations (the base stations), again to result in independent fading at the mobile station.

The specification includes also a feedback-based macro diversity option, called Site Selection Diversity Transmission (SSDT). SSDT attempts to mitigate interference to other users in the system by more optimal power allocation across cells. Thus, it is essentially an antenna selection concept combined with trivial power allocation and improves both diversity and power efficiency provided that the feedback signalling is up-to-date. In SSDT cells (Node Bs, base stations) are assigned a temporary identification (ID). The UE periodically informs the ID of a primary cell to the base stations using an uplink (feedback) signalling field. The dedicated channel in other cells (called non-primary cells) are turned off. The ID of the primary cell is signalled 1-5 times in 10 ms frame, depending on the selected signalling formats.

2.1.3 Time Diversity

UTRA Release '99 supports Type I Automatic Repeat Request (ARQ) protocol. In Type I ARQ erroneous frames are discarded in the receiver. When a negative acknowledgement (NACK) is sent to the transmitter the frame is repeated later on. Time diversity or time selectivity of the channel can be exploited, provided that the retransmitted frame arrives after a sufficiently long time interval (after channel coherence time). In addition to ARQ, a more conventional form of time diversity is exploited via the combined use of interleaving and forward error correction (FEC) codes.

2.1.4 Receive Antenna Diversity

The number of receive antennas one wishes to deploy is typically an implementation issue. When multiple receive antennas are used we say that the receiver uses receive (Rx) antenna diversity. Rx diversity may be used in the base station to improve uplink capacity or coverage. Due to cost and space considerations multi-antenna reception is not popular in terminals. However, Rx diversity is one of the most efficient diversity techniques and likely to be used when performance or coverage improvements are desired.

2.1.5 Transmit Diversity

A significant effort has been devoted in 3GPP to develop efficient transmit diversity solutions to enhance downlink capacity. Transmit diversity methods provide space diversity for terminals with only one receive antenna, and improve the link performance while retaining the complexity at the base station. Typically, the transmitting antenna elements are relatively close to each other. In this case the delay profile is essentially the same for each transmitting element. The closed loop transmit (Tx) diversity solutions developed for the FDD mode support two transmit antennas. Both open-loop and closed-loop Tx diversity solutions are specified for UTRA FDD and TDD modes.

Open-loop Mode: The first open-loop concepts proposed in 3G standardization were based on Code Division Transmit Diversity (Orthogonal Transmit Diversity [37]) and Time Switched Transmit Diversity [38]. Time Switched Transmit Diversity (TSTD) is applied in the WCDMA standard for certain common channels. In TSTD the transmitted signal hops across two transmit antennas, according to [12]. Eventually, also a more efficient Space–Time Transmit Diversity (STTD) solution, based on a variant of the space–time block code developed by Alamouti [8], was adopted for Release '99 [39].

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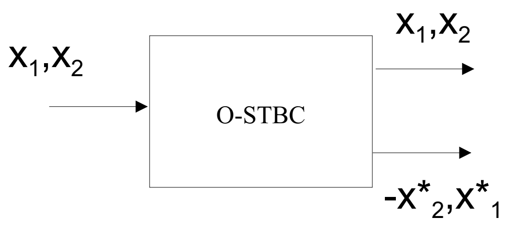

Fig. 2.1: STTD modulator using a 2×2 Orthogonal Space–Time Block Code (O-STBC).

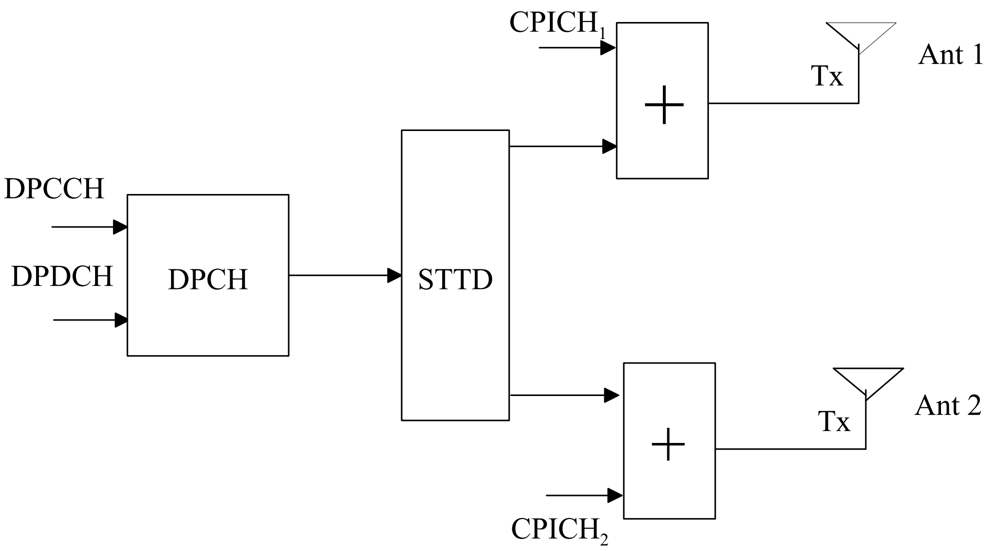

Fig. 2.2: WCDMA open-loop transmit diversity.

The Alamouti code variant used in STTD is

$$\mathbf{X}(x_1, x_2) = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \end{bmatrix}, \tag{2.1}$$

where column 1 is transmitted from antenna 1 and column 2 from antenna 2. The symbols are QPSK modulated in Rel. 99 and Rel. 4. A diagram of 2×2 Orthogonal Space–Time Block Code (O-STBC) is depicted in Figure 2.1. and the transmitter structure (omitting spreading and scrambling) is shown in Figure 2.2. The TDD mode of WCDMA uses a similar transmission matrix, with the exception that (permuted) vectors are transmitted in place of individual symbols, and the variant is called Block STTD (B-STTD). B-STTD mitigates receiver complexity, by simplifying the application of multiuser or multichannel detection.

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The simplification is called for, since the processing gain in the TDD mode is only 16 and therefore advanced receivers are required.

Closed-loop Modes: The first feedback mode proposed to 3G systems was based on selective transmit diversity (STD), where only one additional feedback bit is used per feedback slot to select the desired transmit antenna [40, 41]. These contributions sparked the research on feedback modes, and a number of improvements were eventually suggested in 3G standardization.

Currently, the WCDMA Release '99 and Release 4 specifications include two closed-loop transmit diversity concepts. In both concepts co-phasing information, signalled using a fast feedback channel (of rate 1500 bps), is applied in selecting one of 4 or 16 possible beam weights, respectively. Both concepts approximate coherent transmission (channel-matched beamforming) using different channel quantization and feedback signalling strategies. The transmitter architecture is depicted in Figure 2.3.

In both feedback modes, the transmit weight is selected using the procedure given below:

- terminal measures common pilot channels CPICH1 and CPICH2, transmitted via antennas 1 and 2.
- terminal obtains channel estimates for an l-path channels $\mathbf{h}_1 \in \mathbb{C}^l$ and $\mathbf{h}_2 \in \mathbb{C}^l$ for antenna 1 and antenna 2, respectively.
- the desired transmit weight vector $\mathbf{w} = (w_1, w_2)$ (beam coefficients) is determined from
$$\mathbf{w} = \arg\max_{\mathbf{w}} (w_1\mathbf{h}_1 + w_2\mathbf{h}_2)^\dagger (w_1\mathbf{h}_1 + w_2\mathbf{h}_2)$$
with power constraint $\mathbf{w}^\dagger\mathbf{w} = 1$.

The target is to maximize the combined signal power at terminal. This quantity is invariant to phase shift, and therefore we may constrain w_1 to be real. Then, the problem may be posed as

$$\begin{aligned} w_2 &= ze^{j\phi} \\ (z,\phi) &= \arg\max_{z\in\mathbb{A},\phi\in\mathbb{B}} ||(\sqrt{1-z^2}\mathbf{h}_1 + ze^{j\phi}\mathbf{h}_2)||^2 \end{aligned} \tag{2.2}$$

where $\mathbb{A} = [0, 1]$ and $\mathbb{B} = [0, 2\pi)$.

In the two feedback modes w_2 is quantized and signalled differently to the base station using the Feedback Signalling Message (FSM) field of the uplink signalling frame. FSM is a part of the Feedback Indicator (FBI) field of the uplink dedicated physical control channel

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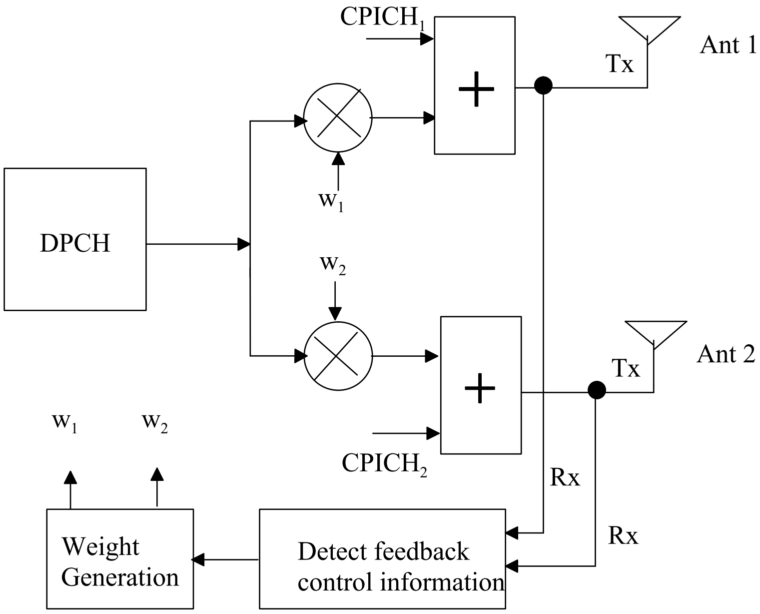

Fig. 2.3: WCDMA closed-loop transmit diversity.

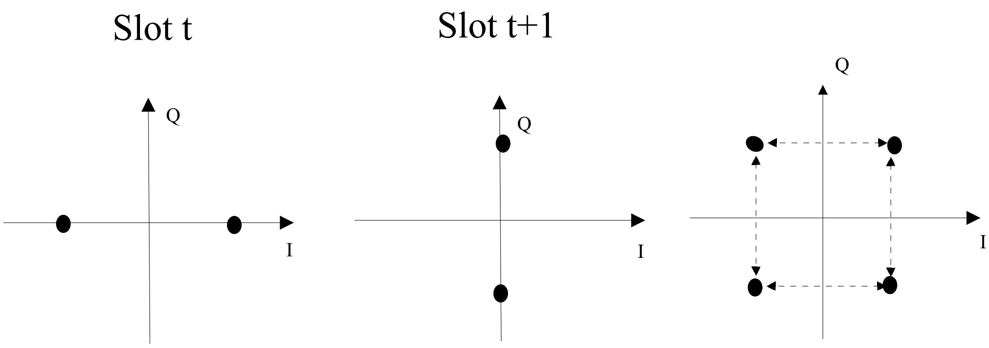

Fig. 2.4: The feedback bits (in FBI field) correspond to feedback pertaining to I and Q branches in successive slots. When these are combined over two slots, the possible states for w_2 and transitions are given in the figure on right.

(DPCCH). The message word is of length $N_{\mathrm{ph}} + N_{\mathrm{po}}$ bits and one bit is transmitted in each uplink slot resulting in a 1500 Hz signalling overhead.

In Mode 1 and Mode 2 closed-loop solutions, the weight w_2 is quantized to 16 state APSK or QPSK constellations, as shown in Figures 2.6 and 2.4, respectively. Each Gray labelled constellation state corresponds to a feedback word. The labels are transmitted to the base station using the $FSMph$ field of the uplink signalling frame, shown in Figure 2.5.

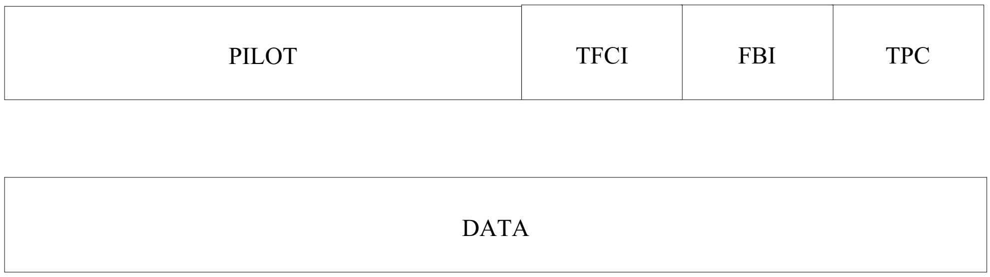

Fig. 2.5: Uplink slot structure for the I and Q branch in UTRA WCDMA. FBI field of the uplink frame supports feedback mode transmit diversity. TPC field contains a power control bit, and TFCI field contains transport format information.

In Mode 1, each feedback bit designates either the real or the imaginary part of the (currently available) feedback weight. The corresponding bits are sent in even and odd numbered slots, respectively, as described in Figure 2.4. The BS combines two consecutive received feedback bits and constructs a transmit weight for the diversity antenna [18, 42] as

$$w_2[t] = 1/\sqrt{2}\mathrm{e}^{\mathrm{j}\,\phi[t]}, \tag{2.3}$$

where

$$\phi[t] = \arg(\mathrm{j}^{\,t \bmod 2}\mathrm{sgn}(y[t]) + \mathrm{j}^{\,(t-1) \bmod 2}\mathrm{sgn}(y[t-1])), \tag{2.4}$$

where $y[t]$ denotes the (noisy) feedback command received at the base station for slot t, and $w_2[t]$ is the complex weight applied in the diversity antenna for the duration of slot $t+1$. In the current specification, the sign function $\mathrm{sgn}(.)$ is used to quantize each received feedback bit, and therefore the resulting weight constellation has four states. Transitions, if any, are allowed to neighboring weight states, as shown in Figure 2.4. The gain information z related to the optimal beamforming vector is not signalled to the transmitter in Mode 1, as neglecting this reduces control delay, and consequently allows beneficial use of the concept also in more rapidly fading channels. Also, by *a priori* constraining the amplitudes in both transmit antennas to be identical, amplifier design problem at the base station is simplified due to smaller peak-to-average ratio (PAR) per antenna element.

Mode 2 provides more accurate weight signalling to the BS transmitter with 16 possible weights. The weight states are shown in Figure 2.6, along with possible state transitions when sequential updating is used. The transmit weight has eight phase states and one power state, which improves beam resolution at the expense of increased feedback delay when compared to Mode 1. Three feedback bits are used for phase adjustment and one for controlling the relative power between antenna 1 and 2. The relative transmit powers of antennas 1 and 2 are

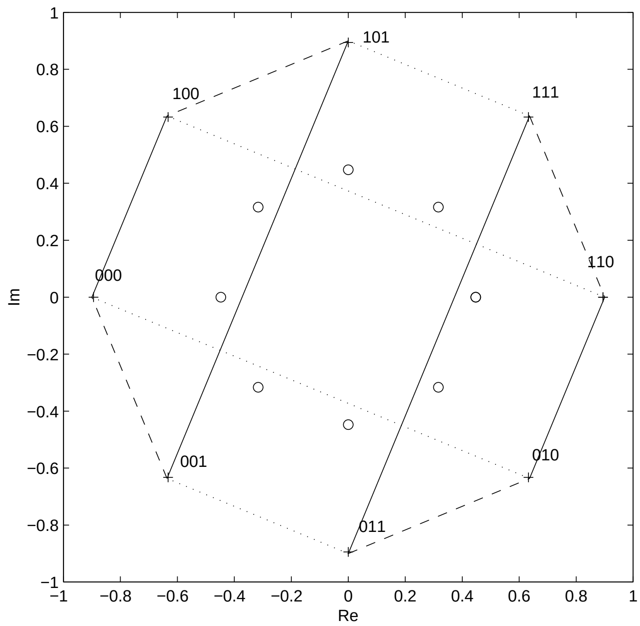

Fig. 2.6: Sixteen possible transmit weights w_2 (with associated state transitions) for feedback Mode 2. Three successive feedback bits determine state transitions in phase and one bit (not shown) designates the relative powers between antenna one and two. The power of each point in the inner constellation ('o') is 0.2, and the power of the each point in the outer constellation is 0.8.

either $\{0.8, 0.2\}$ or $\{0.2, 0.8\}$ depending on the value of the $FSMpo$ field. In analogy with Mode 1, when sequential updating is used the feedback, Mode 2 also applies time-varying quantization constellation for w_2, which is apparent from Figure 2.6. Here, however, the constellation (set partitioning) used at a given time depends on the previously transmitted feedback bits. Tables 2.1 and 2.2 summarize the Mode 2 parameters [14, 19]. In channels

Table 2.1: Feedback power bits and corresponding relative transmit powers for Mode 2

FSM_{po}	Power A1	Power A2
0	0.2	0.8
1	0.8	0.2

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Table 2.2: Feedback phase bits and corresponding phase differences for Mode 2

Phase	180	-135	-90	-45	0	45	90	135
FSM_{ph}	000	001	011	010	110	111	101	100

with small Doppler spread Mode 2 is expected to be superior to Mode 1 due to its better transmit weight resolution. On the other hand, when control delay dominates performance, Mode 1 outperforms Mode 2 [43, 44]. Thus, in practice the network occasionally switches dynamically between the two modes. Switching can be based on e.g. Doppler spread estimates at the base station. The parameters of both feedback modes are given in Table 2.3.

Table 2.3: Feedback mode parameters

Mode	1	2
Phase bits per word (N_{ph})	1	3
Gain bits per word (N_{po})	0	1
Feedback bit rate	1500 bps	1500 bps
Update rate	1500 Hz	1500 Hz
Filtering at BS	yes (2 slots)	no

2.1.6 Beamforming

Conventional beamforming [45] via the use of an antenna array is supported in WCDMA for both fixed and adaptive array concepts. Fixed beams are supported by enabling the use of Secondary Common Pilot Channels (S-CPICH). A predetermined S-CPICH can be used for channel estimation by all users within the coverage area of the beam. The users are assumed to receive data only under one fixed-beam. At most fifteen S-CPICH codes may be associated with a given Primary-CPICH.

When equipped with an adaptive array, the base station may deploy user specific beamforming. In this case the channel seen by each user is generally different and a common channels may not be used for channel estimation. Instead, dedicated pilot symbols, embed-

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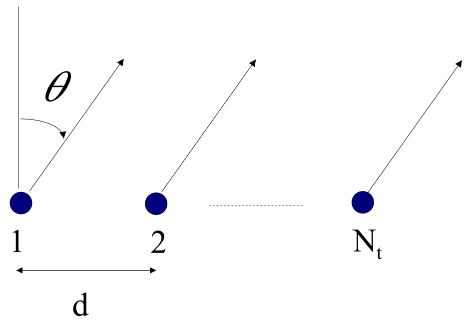

Fig. 2.7: Uniform linear array with N_t elements separated by distance d.

ded in the downlink dedicated channels, are used to obtain channel estimates for coherent reception. Details on the two beamforming options may be found from [46].

Beamforming options given here, and the transmit diversity concepts described in previous section, differ in a number of important details. For example, conventional beamforming in FDD systems typically attempts to direct the beam towards a spatial direction where the user resides thus reducing the average interference to other users in the cell. In this approach the transmit directions (or radiation patterns) are matched with the dominant receive directions and the directional beams are formed with calibrated antenna arrays, e.g. with uniform linear (see Fig. 2.7) or circular arrays. In determining the dominant transmit direction, one typically averages over the (fast) fading distribution with a sufficiently long integration window. This is needed, since due to lack of channel reciprocity a receive direction determined from uplink rarely provides an optimal transmit direction for the downlink channel. In contrast, in closed-loop transmit diversity the beamforming coefficient is determined by the terminal from downlink measurements and therefore it is matched appropriately to the downlink channel. Therefore, with closed-loop solutions both uncalibrated and calibrated arrays can be used, since the feedback weight implicitly takes into account any phase errors in the transmitting elements.

Further insight into possible extensions is obtained, if we parameterize the transmit beam with Direction of Transmission (DoT) or Direction of Arrival (DoA) parameter θ,

$$\mathbf{w}(\theta) = [1\ , e^{j2\pi d \sin(\theta)/\lambda}, ..., \ e^{j2\pi(N_T-1)d\sin(\theta)}]^T, \tag{2.5}$$

where d denotes the inter-element distance in a Uniform Linear Array, and λ is the carrier wavelength. In a way, the feedback signal determines the transmit direction. Indeed, in WCDMA $N_\mathrm{t} = 2$, d is arbitrary and θ is analogous to the feedback phase in Mode 1 or Mode 2. If this parameterization were used in closed-loop modes, only one coefficient would need

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to be signalled to the network [1,47], regardless of the number transmitting antenna elements, and with a uniform linear array the DoT interpretation is valid when $d = \lambda/2$.

2.1.7 High Speed Packet Access (HSDPA)

WCDMA Release 5 incorporates a data-centric option for downlink, called High Speed Downlink Packet Access (HSDPA) [48]. HSDPA includes advanced air interface concepts that enhance the downlink throughput, such as

- link adaptation (via adaptive modulation and coding)
- improved ARQ solution (Hybrid ARQ),
- reduced length (2 ms) Transport Time Interval (TTI),
- higher peak rates via high order modulation (16QAM), and
- improved macro diversity via Fast Cell Selection (FCS).

A new type of transport channel is defined, the High Speed Downlink Shared Channel (HS-DSCH), with a fixed spreading factor of length 16. In order to reduce service delays, the HS-DSCH Transport Time Interval equals 2 ms, a fifth of the TTI length defined for Release 99 and Release 4. The control of HS-DSCH is terminated in the base station, as opposed to Base Station Controller. For peak rates a terminal may employ high order modulation and multicode transmission. The number of supported multicodes depends on terminal capability. If one terminal does not use all 15 available multicodes (one is reserved for common channels) other users may be code-multiplexed in the same TTI. Link adaptation is used to select the optimal coding and modulation options, one of many possible transport format configurations, so that maximal throughput and desired QoS is maintained. In good channel conditions 16 QAM and a high coding rate may be selected.

Perhaps the most relevant concept in HSDPA is that user scheduling and associated data rates are assigned based on channel state information signalled from the receiver to the transmitter. The channel information is embedded into Channel Quality Indicator (CQI), which the base station may use in allocating transport formats, channelization codes and time slots to the users to maximize system capacity or throughput. CQI feedback enables therefore multiuser diversity in the spirit of [49], when applied together with downlink scheduling at the base station.

As stated before, the Release 4 specification uses Type I ARQ. For improved system efficiency, the Release 5 adopts also additional Hybrid ARQ (HARQ) concepts. These include combining schemes that are based on Incremental Redundancy (IR). The new concepts are called Type II and Type III (with Chase Combining) using an N-channel stop and wait (SAW) principle. HARQ can be interpreted to provide implicit rate matching, while AMC attempts to determine the optimal rate before (first) transmission of a given packet.

2.2 CDMA2000

Many of the diversity solutions available to cdma2000 [50, 51] systems are similar to those described above for WCDMA. The cdma2000 standard defined in 3GPP2 supports fixed beam transmission via the use of auxiliary spreading codes. In contrast to WCDMA, dedicated pilots are not used in cdma2000 and this essentially disables the application of adaptive arrays. As far as transmit diversity is concerned, cdma2000 has adopted a concept called space–time spreading (STS) [52] which separates successive Alamouti-coded symbols using two orthogonal codes, whereas two (orthogonal) time slots are used in WCDMA. Where WCDMA applies Time-Switched Transmit Diversity, cdma2000 applies Orthogonal Transmit Diversity (OTD). In contrast to UTRA/WCDMA, cdma2000 specification includes both STS and OTD as optional modes, for both terminals and the network. In UTRA, the support for STTD is mandatory for the network, and other transmit diversity modes are optional. However, all UTRA terminals need to support all specified transmit diversity modes.

In analogy with HSDPA, defined for UTRA, cdma2000 supports two similar data-centric transmission standards. A concept called cdma2000 1xEV-DO (single carrier cdma2000 EVolution-Data Only) provides a peak data rate of 2.457 Mbps in downlink using 1.25 MHz spectrum [53] and a separate carrier is needed for the service. It is thus orthogonal in frequency domain to speech services. Link adaptation is used to match the transmission format to the channel conditions as well as possible using a base station that always operates at full power. The other concept in cdma2000 allows to mix speech and high speed data in the same carrier and is called 1xEV-DV [51] (cdma2000 EVolution- Data & Voice). The technical physical layer solutions in 1xEV-DV are similar to those in the UTRA/HSDPA concept, but there are several differences in how the general principles (like multiuser scheduling, rate adaptation, multiplexing, etc.) are brought into practice.

2.3 GSM/EDGE

The GSM/EDGE standard was developed before many of the most prominent multi-antenna concepts were invented, or at least before they were popularized. Therefore, it is natural that the standard does not explicitly support the use of particular multi-antenna transmission schemes. This, however, does not mean that transmit diversity cannot be applied. Indeed, there are various implicit transmit diversity solutions, such as frequency sweep diversity [12], frequency hopping, antenna hopping and delay diversity [11] that can be used to some extent.

In delay diversity a delayed copy of the signal is transmitted from a diversity antenna in the base station. The associated receiver (mobile) is transparent to delay diversity, as it sees only a slightly longer impulse response, and the channel equalizer can combine the signals transmitted from multiple transmit antennas. In CDMA systems, delay diversity is not as popular (in downlink) since delayed copies of the used channelization codes are non-orthogonal. Therefore, if delay diversity were used in CDMA systems, interference among different users would exist even in flat fading channels.

Antenna hopping may be applied so that different bursts are transmitted via different antennas. This is also completely transparent to the receiver, as channel estimation may done separately for each burst. In the same vein, frequency hopping can be used, where the co-phasing coefficient changes for different bursts. Thus, any such techniques may be applied provided that the signal processing algorithms used in the receiver need not be changed. Recall that STTD cannot be received (optimally) if the use of the method is not known to the receiver.

The GSM/EDGE standard also evolves, just like WCDMA and cdma2000. Therefore, many of the space–time block coding solutions considered herein have been studied also in the context of TDMA systems, and in particular assuming severe Intersymbol Interference (ISI) prevalent in GSM/EDGE signalling. In such channels, Time-reversed space–time block codes [54–58] may be more readily applicable.

3

Multi-antenna Channels

As described in the previous Chapter, the 3G systems include explicit support for two transmit antennas using space–time block coding. However, these solutions are aimed at improving performance, not spectral efficiency. Future wireless systems are likely to promote also considerably higher spectral efficiencies using multiple transmit and receive antenna in both ends of the communication link. Such an increase in the number of antennas improves both power and spectral efficiency, especially when the modulation/coding design is optimized for the arising multiple-input multiple-output (MIMO) channel. This Chapter discusses the capacity promise of MIMO channels and summarizes a portion of the theoretical background behind space–time coding and matrix modulation.

3.1 MOTIVATION

MIMO and MISO transmission methods are currently being developed in numbers around the globe, prompted by the capacity promise due to [10]. Efficient MIMO and MISO transmission methods are being considered for the evolving 3G wireless standards [1,59], OFDM-based systems [60], GSM/EDGE [54–58, 61] and 4G. In particular, envisioned 4G data rates of 100 Mbps/1 Gbps using 100 MHz bandwidth in wide-area high-mobility/local-area low-mobility environments are difficult to achieve unless MIMO channel properties are fully exploited.

One common characteristic in advanced multi-antenna techniques is that they attempt to explicitly utilize the random characteristic of the wireless medium. Indeed, fading is not by default assumed to have a detrimental effect on system or link capacity. Rather, in a properly designed wireless system fading (or a random channel) is used as an additional multiplexing resource. Generally, multi-antenna transmission and reception techniques provide

- improved fading resistance, or deliberate exploitation of fading,
- interference mitigation (e.g. using beamforming and null steering at both transmitter and receiver),
- reduced transmitter power levels per transmit antenna path, which simplifies power amplifier design problems,
- a new dimension for rate and power allocation problems,
- theoretically higher system capacity.

On the other hand, the practical problems in multi-antenna channels are manyfold. The design of spectrally efficient transmission schemes that are able to reach capacity is not straightforward, at least when the receivers are constrained to have limited complexity. A large number of MIMO-friendly coding and modulation solutions have been proposed recently, each with particular disadvantages and advantages. Solutions advocating combined channel coding and MIMO modulation, considered e.g. in [62–66], have good performance but may be difficult to embed into an existing transmission scheme, for example, in a way STTD was embedded into WCDMA. Separating space–time modulation and channel coding is potentially a more practical approach, in that it applies a modular design principle. In a modular design, a change an individual part of the system has a minimal effect on other parts of the system, and thus avoids cumbersome re-engineering. Linear matrix modulation is particularly well-suited for such a modular design, as it can be designed to reach capacity while attaining full diversity with satisfactory performance, and since it is relatively easy to decode. STTD provides an example of matrix modulation, although it is known to achieve capacity only with one receive antenna. In a non-degenerate MIMO channel, where both ends have more than one antenna, STTD does not suffice, and new codes need to be invented.

3.2 SIGNAL MODEL, CHANNEL AND CAPACITY

Abstracting from the coding, interleaving and multi-user multiplexing units, the baseband signal model considered in this work is formulated concisely as follows for a one-path channel

$$\underset{T\times N_{\mathrm{r}}}{\mathbf{Y}} = \underset{T\times N_{\mathrm{b}}}{\mathbf{X}} \underset{N_{\mathrm{b}}\times N_{\mathrm{t}}}{\mathbf{W}} \underset{N_{\mathrm{t}}\times N_{\mathrm{r}}}{\mathbf{H}} + \underset{T\times N_{\mathrm{r}}}{\mathrm{noise}} \tag{3.1}$$

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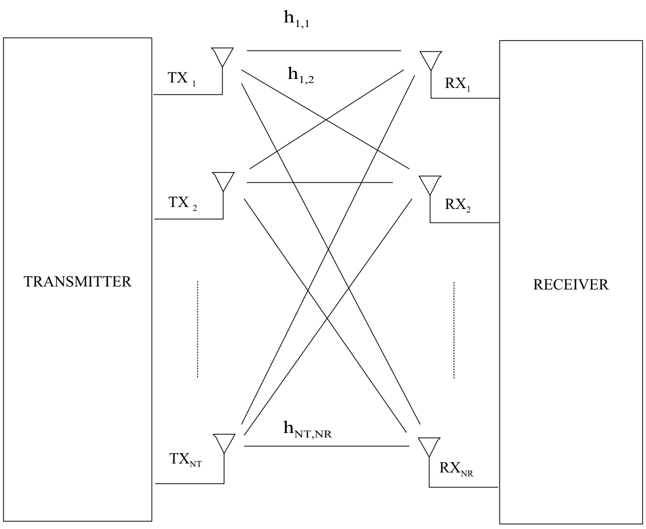

Fig. 3.1: MIMO model with N_t transmit and N_r receive antennas [1].

Above,

- $\mathbf{X}$ is the space–time modulation matrix
- T is the block length of the matrix modulator (or space–time code)
- N_b is the number of transmission beams,
- N_t is the number of transmit antennas,
- N_r is the number of receive antennas,
- $\mathbf{Y}$ is the $T \times N_\mathrm{r}$ matrix of received signals,
- $\mathbf{W}$ is the $N_\mathrm{b} \times N_\mathrm{t}$ beamforming matrix,
- $\mathbf{H}$ is a matrix where each column is a channel vector from the multiple transmit antennas to one receive antenna,

and the noise is assumed to be complex Gaussian.

Modulation matrix $\mathbf{X}$ transmits $R_\mathrm{s}T$ complex modulation symbols over N_b beams during a block of T symbol epochs. The number of parallel streams R_s is defined as the (average) number of complex symbols transmitted per symbol epoch, i.e. the symbol rate. In a space–time modulator with block of length T, altogether $R_\mathrm{s}T$ complex symbols are thus

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transmitted. The beamforming unit prepares the N_b beams for transmission from N_t antennas using matrix $\mathbf{W}$.

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} & \dots & h_{1N_\mathrm{r}} \\ h_{21} & h_{22} & \dots & h_{2N_\mathrm{r}} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N_\mathrm{t}1} & h_{N_\mathrm{t}2} & \dots & h_{N_\mathrm{t}N_\mathrm{r}} \end{bmatrix}, \tag{3.2}$$

as depicted in Figure 3.1. In a MISO system, $\mathbf{H}$ is a column vector $\mathbf{h}$. When a multipath channel model is considered, the channel matrix $\mathbf{H}$ is extended to cover the multipath components, and $\mathbf{X}$ is extended to cover multiple transmission blocks.

3.2.1 Capacity

Channel capacity determines the ultimate spectral efficiency limit, which can generally be only approached by practical modulation and coding methods. Below, we take a look at the information theoretic reasoning behind MIMO channels, and provide some numerical results on achievable spectral efficiencies in some relevant channels. MIMO modulation, along the lines of [13, 67], has recently been justified using information theoretic measures by [9, 10, 68, 69]. In the following, we consider capacity expressions in cases where channel state information exists at the transmitter and when it does not exist. These may also be called, respectively, “closed-loop capacity” and “open-loop capacity”. Closed-loop capacity has been derived in number of publications and in classical information theory literature, see e.g. [70], analogous results are associated with capacity in parallel (correlated) Gaussian channels.

3.2.1.1 No Channel State Information at Transmitter Assume that the transmitter has no information on the channel coefficients of transmission medium, i.e. there is no channel state information (CSI) in the transmitter, but perfect channel information at the receiver. In deriving capacity results, we assume that the MIMO transmitter applies vector modulation

$$\mathbf{x} = [x_1 \; x_2 \; \dots \; , x_{N_\mathrm{t}}], \tag{3.3}$$

with covariance $\mathbf{Q} = \mathrm{E}\left\langle \mathbf{x}^\dagger \mathbf{x} \right\rangle$, to be defined. In this case, the received signal reduces to

$$\mathbf{y} = \mathbf{x}\,\mathbf{H} + \mathrm{noise}. \tag{3.4}$$

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The vector of N_t transmitted symbols may be solved (in the absence of noise), provided that $\mathbf{H}$ has rank N_t. If $\mathbf{H}$ is singular (rank less than N_t), the model is ill-conditioned and the receiver cannot unambiguously resolve elements of $\mathbf{x}$.

Let $\rho = P/\sigma^2$ denote average (transmitted) SNR, expressed in terms of the total transmit power P applied on symbol vector $\mathbf{x}$, and the average noise power in the receiver σ^2. The mutual information between zero-mean $\mathbf{x}$ and $\mathbf{y}$, given $\mathbf{H}$, is maximized, when the input distribution of $\mathbf{x}$ is Gaussian. This is written as

$$\begin{aligned} \mathcal{I}(\mathbf{x},\mathbf{y}|\mathbf{H}) &= \log\det\left(\mathbf{I}_{N_\mathrm{r}} + \frac{\rho}{P}\mathbf{H}^\dagger\mathbf{Q}\,\mathbf{H}\right) && (3.5)\\ &= \log\det\left(\mathbf{I}_{N_\mathrm{t}} + \frac{\rho}{P}\mathbf{Q}\,\mathbf{H}\mathbf{H}^\dagger\right) && (3.6) \end{aligned}$$

where $\mathbf{Q} = \mathrm{E}\left\langle \mathbf{x}^\dagger\mathbf{x}\right\rangle$ and the logarithm is taken in base 2 to provide capacity in terms of bits per channel use. The equality (3.6) follows from matrix equality $\det(\mathbf{I}_N + \mathbf{A}\mathbf{B}) = \det(\mathbf{I}_M + \mathbf{B}\mathbf{A})$ where $\mathbf{B}$ and $\mathbf{A}$ are $M \times M$ and $N \times N$ matrices, respectively.

Ergodic capacity of a random channel is defined as

$$\mathcal{C} = \mathrm{E}\left\langle \mathcal{I}(\mathbf{x},\mathbf{y}|\mathbf{H})\right\rangle_{\mathbf{H}} \tag{3.7}$$

where the subscript designates that the expectation is taken over $\mathbf{H}$ and $\mathrm{Tr}\,\mathbf{Q} = P$. In an i.i.d complex Gaussian channel with no channel state information at transmitter ergodic capacity is maximized when $\mathbf{Q} = P/N_\mathrm{t}\mathbf{I}_{N_\mathrm{t}}$, and we obtain

$$\mathcal{C} = \mathrm{E}\left\langle \log\det\left(\mathbf{I}_{N_\mathrm{r}} + \frac{\rho}{N_\mathrm{t}}\,\mathbf{H}^\dagger\mathbf{H}\right)\right\rangle_{\mathbf{H}} . \tag{3.8}$$

Thus, with no CSI at transmitter, power is distributed with equal power over all antennas (or beams). For a scalar (rank one SISO) channel with $N_\mathrm{t} = N_\mathrm{r} = 1$, capacity is

$$\mathcal{C} = \mathrm{E}\left\langle \log(1 + \rho|h|^2)\right\rangle_h . \tag{3.9}$$

The analysis of multi-antenna capacity is tractable when the capacity equation is rewritten using the singular value decomposition (SVD). The SVD-based capacity characterization was used originally in [10] in deriving exact expressions for capacity. This was carried out by using results on eigenvalue distributions of random matrices. Such results are readily available for the flat fading Gaussian MIMO channel in question, see e.g. [71]. In this approach, let the singular value decomposition of matrix $\mathbf{H}$ be given as

$$\mathbf{H} = \mathbf{W}^\dagger\,\Sigma\,\mathbf{V} , \tag{3.10}$$

where $\mathbf{W}$ is a $N_\mathrm{t} \times N_\mathrm{t}$ unitary matrix, Σ is a $N_\mathrm{t} \times N_\mathrm{r}$ matrix with $\min(N_\mathrm{t}, N_\mathrm{r})$ singular values on the main diagonal, and $\mathbf{V}$ is a $N_\mathrm{r} \times N_\mathrm{r}$ unitary matrix. Corresponding to SVD, we may write eigenvalue decomposition of the channel correlation matrix

$$\mathbf{H}^{\dagger}\mathbf{H} = \mathbf{V}^{\dagger}\,\Lambda\,\mathbf{V}\,, \tag{3.11}$$

with $\Lambda = \Sigma^2$ a diagonal matrix with the $N_{\rm t}$ eigenvalues λ_i of the channel correlation matrix on the diagonal. Using $\det(\mathbf{I}_N + \mathbf{AB}) = \det(\mathbf{I}_M + \mathbf{BA})$ we note that matrix $\mathbf{V}$ is reducible, and the capacity is written with a sum over $\min(N_{\rm t}, N_{\rm r})$ parallel channels. The number of terms in the sum corresponds to the $\min(N_{\rm t}, N_{\rm r})$ non-zero eigenvalues of $\mathbf{H}^{\dagger}\mathbf{H}$:

$$\mathcal{C} = \sum_{i=1}^{\min(N_{\rm t},N_{\rm r})} \mathrm{E}\left\langle \log\left(1 \;+\; \rho/N_{\rm t}\;\lambda_i\right)\right\rangle_{\mathbf{H}}\;. \tag{3.12}$$

The results above, summarizing those in [10], suggest that in a multivariate Gaussian channel

- MIMO capacity grows linearly in $\min(N_{\rm r}, N_{\rm t})$, and the matrix channel decomposes to $\min(N_{\rm r}, N_{\rm t})$ independent parallel channels,
- the linear capacity increase is due the increased rank of $\mathbf{H}$.

Clearly, the actual realizations of the singular values of $\mathbf{H}$ also affect the actual realized capacity. For example, in near-singular channels some singular values are very small and their effect in the sum is insignificant.

The capacity results obviously only provide a guideline for communication engineers. The implicit assumptions on infinite block size, the presence of a hypothetical capacity-reaching code, and optimum multiuser decoding are rather challenging in practice. A related performance measure, the *outage capacity* is perhaps a more practical object as it yields a bound on a packet error rate in quasi-static channels. The mutual information outage probability is defined as

$$P_{\rm out}(R, \mathbf{H}, \rho) = Pr(\mathcal{I} < R), \tag{3.13}$$

giving the probability that the channel supports rate R with probability $P_{\rm out}(R, \mathbf{H}, \rho)$, where $\mathcal{I}$ is considered as a random variable, since the channel $\mathbf{H}$ is random. The outage capacity is the maximum rate R that is supported with some prescribed probability ε,

$$R_{\varepsilon} = \arg\max_{R}[P_{\rm out}(R, \mathbf{H}, \rho) = \varepsilon]. \tag{3.14}$$

3.2.1.2 Transmission using Channel State Information When $\mathbf{H}$ is fixed, equation (3.10) can be used to diagonalize the transmission into a number of parallel channels. In the MIMO interpretation the digonalization is realized with a transmit beamforming matrix

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$\mathbf{W}$ and a receive beamforming matrix $\mathbf{V}^\dagger$. The parallel channels have different gains, just as the corresponding eigenvalues are different. With side information at the transmitter, the transmitter may choose the covariance of the transmitted symbols to be

$$\mathbf{Q} = \tilde{\mathbf{W}}^\dagger \, \mathbf{P} \, \tilde{\mathbf{W}} \,, \tag{3.15}$$

where $\tilde{\mathbf{W}}$ is a generalized beamforming matrix, which constructs altogether N_t orthogonal beams. The power allocation matrix $\mathbf{P}$ is a diagonal matrix that may be chosen to exploit the differences of the eigenvalues of the eigenbeams, and the optimal power allocation depends on $\mathbf{H}$ and noise power. The transmitted signal is $\tilde{\mathbf{x}} = \mathbf{x} \, \mathbf{A} \, \mathbf{W}$. MIMO channel capacity in the presence of correlated noise has been addressed in [72, 73].

The set of non-interfering parallel channels is written as

$$\tilde{\mathbf{y}} = \mathbf{y} \, \mathbf{V}^\dagger = \mathbf{x} \, \mathbf{A} \, \Sigma + \mathbf{n} \,, \tag{3.16}$$

where $\mathbf{A}$ is a diagonal amplitude matrix, which satisfies $\mathbf{A}^2 = \mathbf{P}$. The capacity becomes

$$\mathcal{C} = \max_{\sum_i P_i = P} \sum_{i=1}^{\min(N_\mathrm{t}, N_\mathrm{r})} \mathrm{E} \left\langle \log \left(1 \;+\; P_i/\sigma^2 \;\; \lambda_i\right) \right\rangle_H \;, \tag{3.17}$$

where the i'th diagonal elements of the power allocation matrix is P_i. The optimal power allocation strategies are found in closed form [10, 70]. Therein, it is shown that the capacity of a multi-antenna channel (in bits per dimension) is reached with a water-filling power-allocation policy. The power allocated to the i'th row of the $\mathbf{W}$ matrix in the SVD, corresponding to the eigenvalue λ_i, is

$$P_i = \sigma^2 \; \left[\mu - \lambda_i^{-1}\right]_+ \;, \tag{3.18}$$

where the variable μ is defined by the total power constraint: $\sum_i P_i \leq P$. The function $[\;]_+$ sets negative numbers to zero. The resulting capacity expression is

$$\mathcal{C} = \sum_{i=1}^{\min(N_\mathrm{t}, N_\mathrm{r})} \mathrm{E} \left\langle \left[\log \mu \lambda_i\right]_+ \right\rangle \;. \tag{3.19}$$

3.2.1.3 Comparisons For additional insight in the practical differences in open- and closed-loop capacity expressions (3.12) and (3.19) it is instructive to consider the case involving multiple transmit antennas but only one receive antenna. When $N_\mathrm{r} = 1$, $\mathbf{H}$ collapses into a row vector, the channel rank is one, only one non-zero eigenvalue prevails. In the open-loop case, i.e. without knowledge of $\mathbf{H}$, the transmitter can be thought of forming an $N_\mathrm{t} \times N_\mathrm{t}$ beamforming matrix $\mathbf{W}$, wherein the $N_\mathrm{t} - 1$ rows correspond to noise subspace. Equal power is applied for all eigenbeams and the open-loop capacity (3.12) is

$$\mathcal{C} = \mathrm{E} \left\langle \log \left(1 \;+\; \rho/N_\mathrm{t} \;\lambda\right) \right\rangle_{\mathbf{h}} \;. \tag{3.20}$$

,

Interestingly, [74] showed that STTD reaches channel capacity in a block-fading i.i.d Rayleigh fading channels, but only when $N_{\mathrm{t}} = 2, N_{\mathrm{r}} = 1$, When $N_{\mathrm{r}} > 1$, capacity-optimal modulators cannot be found from a class of orthogonal space–time block codes.

In contrast, when $\mathbf{H}$ is known, the optimal power allocation is trivial - all power is allocated to the single eigenbeam. The closed-loop capacity (3.19) becomes

$$\mathcal{C} = \mathrm{E}\left\langle \log\left(1 \ + \ \rho\,\lambda\right)\right\rangle_{\mathbf{h}} \ . \tag{3.21}$$

In the closed-loop (beamforming) case no power is wasted on the noise subspace, and the receiver sees an N_{t}-fold beamforming (or SNR) gain.

When $N_{\mathrm{r}} = N_{\mathrm{t}}$, with open-loop transmission, we again construct N_{t} eigenbeams and transmit with equal power using these beams. Then, nothing is gained in terms of capacity in a Rayleigh fading channel, beamforming only changes one unitary basis to another. On the other hand, in the closed-loop case with power allocation the capacity is again at least as good as with open loop transmission. In this case, however, even optimal power allocation is not able to provide dramatic gains, as shown in [72]. The gains may be higher in cases where the channel has more structure, i.e. when the eigenvalue spread is larger than in the symmetric i.i.d. Gaussian channel.

3.2.2 MIMO channel models

Most of the theoretical results in communication through a MIMO channel assume i.i.d fading, since the analysis is then simplified, and since this case brings forth the capacity promise in MIMO channels, see e.g. [9, 10]. Correlated MIMO channels have been developed in an attempt to modify the stochastic channel model closer to reality [75, 76]. The channel correlation has been found to depend on both the environment and the spacing of the antenna elements. A receiver, surrounded by a large number of nearby scatterers, is likely to experience (almost) uncorrelated fading even when the antennas are separated by half the wavelength. In this case, virtually uncorrelated fading may arise even if the elements are separated by 7.5 cm when the carrier frequency is 2 GHz. If polarization diversity [77, 78] is used as well significant diversity benefits can be reaped even with very narrow antenna separation. The base station antennas are typically significantly higher above ground than the scatterers, and sufficiently low correlation is likely to require much larger separation between antenna elements, perhaps around 10 wavelengths.

Publications [60, 69] examine the characteristics of resolvable multipaths in broadband MIMO systems, and [79, 80] present the first a multipath-inspired random matrix model for

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MIMO channels [79]. It has been shown that an estimate of the number of non-resolvable dominant scatterers can be used to accurately predict the eigenvalue spectra.

Perhaps a most tractable model is a statistical one, capturing the joint distribution of the $N_\mathrm{t}N_\mathrm{r}$ channel coefficients. If the distribution is assumed to belong to a class of complex Gaussian distribution, it becomes necessary to characterize typical correlation (values) matrices for relevant channels. A $N_\mathrm{r}N_\mathrm{t} \times N_\mathrm{r}N_\mathrm{t}$ spatial MIMO correlation matrix can be written as

$$\mathbf{R}_\mathrm{MIMO} = \mathrm{E}\left\langle \mathrm{vec}(\mathbf{H})\mathrm{vec}(\mathbf{H})^\dagger \right\rangle$$

where $\mathrm{vec}(\mathbf{H})$ stacks the vectors $\mathbf{h}_m = [h_{1m}, \cdots, h_{N_\mathrm{t}m}]^\mathrm{T}$ on top of each other, where h_{ij} models the channel coefficient between the i^th transmit and the j^th receive antenna element. If we assume that the spatial correlation is the same regardless of the antenna element index, in that all elements illuminate the same scatters, a simplified Kronecker-product type approximation arises. In this model [75, 81, 82] the $N_\mathrm{t}N_\mathrm{r} \times N_\mathrm{t}N_\mathrm{r}$ channel correlation matrix is approximated by a Kronecker product of the transmit and receive correlation matrices,

$$\mathbf{R}_\mathrm{MIMO} \cong \mathbf{R}_\mathrm{Tx} \otimes \mathbf{R}_\mathrm{Rx}\,, \tag{3.22}$$

as shown in [83–85]. The model is simple and this inherent simplicity also leads to inaccuracies in that the approximation is not always justified [86]. However, the models are developed mostly to aid link level simulations and even in those cases they cover only certain special cases with various degrees of channel correlation [81]. One possible set of characteristic channels is summarized in Table 3.2, where Case 1 is a simple uncorrelated flat Rayleigh fading case, while other cases provide channel characteristics in different correlated environments. Cases 2 and 3 model a typical urban macro cellular environments with different degrees of time dispersion and azimuth spread (AS), power azimuth spread (PAS) and angle of arrival (AOA). Case 4 models micro-cellular and urban environments, assuming different angle of arrivals for different delays. The delays of 3GPP channels are tabulated in Table 3.1.

Table 3.1: ITU delay profiles

Model	delay profile [ns]	power profile [dB]
Pedestrian A	0, 110, 190, 410	0 -9.7 -19.2 -22.8
Pedestrian B	0, 200, 800, 1200, 2300, 3700	0 -0.9 -4.9 -8.0 -7.8 -23.9
Vehicular A	0, 310, 710, 1090, 1730, 2510	0 -1 -9 -10 -15 -20

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Table 3.2: MIMO channel parameters from [81]

case	1	2	3	4
#paths	1	4	6	6
delay profile	N/A	Pedestrian A	Vehicular A	Pedestrian B
MS topology	N/A	$\frac{1}{2}\lambda$ element spacing		
MS PAS	N/A	uniform over $360°$		
MS AOA [deg]	N/A	0	0	0
BS topology	N/A	uniform linear array with $\frac{1}{2}\lambda$ or 4λ element spacing		
BS PAS	N/A	Laplacian, AS $5°$	Laplacian, AS $10°$	Laplacian, AS $15°$
BS AOA [deg]	N/A	20 or 50	20 or 50	2, -20, 10, -8, -33, 31

In cases 2 and 3 in Table 3.2 the same AOA is used for all paths at BS and two separate cases are defined in this respect. Also, for these cases the Ricean K-factor is either 0 dB or 3 dB for the first path. This is used to model different line-of-sight type channels.

For example, in macro cell (case 2) with $\frac{1}{2}\lambda$ element spacing, $20°$ AOA and $5°$ AS, the spatial correlations for a 4-element linear antenna array are given by

$$\mathbf{R}_{\mathrm{Tx},5^\circ} = \begin{bmatrix} 1 & 0.97\,\mathrm{e}^{0.34\pi\mathrm{j}} & 0.89\,\mathrm{e}^{0.68\pi\mathrm{j}} & 0.77\,\mathrm{e}^{0.99\pi\mathrm{j}} \\ 0.97\,\mathrm{e}^{-0.34\pi\mathrm{j}} & 1 & 0.97\,\mathrm{e}^{0.34\pi\mathrm{j}} & 0.89\,\mathrm{e}^{0.68\pi\mathrm{j}} \\ 0.89\,\mathrm{e}^{-0.68\pi\mathrm{j}} & 0.97\,\mathrm{e}^{-0.34\pi\mathrm{j}} & 1 & 0.97\,\mathrm{e}^{0.34\pi\mathrm{j}} \\ 0.77\,\mathrm{e}^{-0.99\pi\mathrm{j}} & 0.89\,\mathrm{e}^{-0.68\pi\mathrm{j}} & 0.97\,\mathrm{e}^{-0.34\pi\mathrm{j}} & 1 \end{bmatrix} \tag{3.23}$$

In a micro cell model (case 4) with $10°$ AOA, $15°$ AS and $\lambda/2$ antenna element spacing, the spatial correlation matrix in the base station becomes

$$\mathbf{R}_{\mathrm{Tx},15^\circ} = \begin{bmatrix} 1 & 0.76\,\mathrm{e}^{0.17\pi\mathrm{j}} & 0.43\,\mathrm{e}^{0.35\pi\mathrm{j}} & 0.25\,\mathrm{e}^{0.53\pi\mathrm{j}} \\ 0.25\,\mathrm{e}^{-0.53\pi\mathrm{j}} & 1 & 0.76\,\mathrm{e}^{0.17\pi\mathrm{j}} & 0.43\,\mathrm{e}^{0.35\pi\mathrm{j}} \\ 0.43\,\mathrm{e}^{-0.35\pi\mathrm{j}} & 0.25\,\mathrm{e}^{-0.53\pi\mathrm{j}} & 1 & 0.76\,\mathrm{e}^{0.17\pi\mathrm{j}} \\ 0.76\,\mathrm{e}^{-0.17\pi\mathrm{j}} & 0.43\,\mathrm{e}^{-0.35\pi\mathrm{j}} & 0.25\,\mathrm{e}^{-0.53\pi\mathrm{j}} & 1 \end{bmatrix} \tag{3.24}$$

A typical assumption in a mobile station is that there is no dominant direction of the impinging signals. Assumption of uniformly distributed angles of arrival in $[-\pi, \pi)$ and $\frac{1}{2}$

wavelength antenna spacing produces spatial correlations $J_0(\pi k)$, $k = 0, 1, 2, 3$

$$\mathbf{R}_{\text{Rx}} = \mathbf{R}_{\text{Tx},360^\circ} = \begin{bmatrix} 1 & -0.3043 & 0.2203 & -0.1812 \\ -0.3043 & 1 & -0.3043 & 0.2203 \\ 0.2203 & -0.3043 & 1 & -0.3043 \\ -0.1812 & 0.2203 & -0.3043 & 1 \end{bmatrix}, \tag{3.25}$$

which can also be used to model the angles of departure in a pico-cell base station. In this modelling approach, using (3.22) [83], the channel matrix is colored to produce

$$\mathbf{H} = \mathbf{R}_{\text{Tx}}^{1/2}\mathbf{N}(\mathbf{R}_{\text{Rx}}^{1/2})^\dagger,$$

where $\mathbf{N}$ is a random $N_{\text{t}} \times N_{\text{r}}$ matrix with i.i.d. complex Gaussian elements and $(\cdot)^{1/2}$ denotes a matrix square root with $\mathbf{R}^{1/2}(\mathbf{R}^{1/2})^\dagger = \mathbf{R}$. In uplink transmission $\mathbf{R}_{\text{MIMO}} = \mathbf{R}_{\text{Rx}} \otimes \mathbf{R}_{\text{Tx}}$, and a similar story follows.

3.2.3 Examples

Having discussed the capacity expressions and channel models for both open-loop and closed-loop cases, it is useful to consider some numerical examples using Monte-Carlo simulations of different channel models. The representative channel models are taken from Table 3.2. In the closed-loop case, optimal power allocation is used for each channel realization, and the the ergodic capacity is computed by averaging over channel distribution. The ergodic capacity results are shown both for a symmetric case, where $N_{\text{t}} = N_{\text{r}}$, and for an asymmetric case where $N_{\text{t}} = 2N_{\text{r}}$. The latter antenna configuration is simulated to highlight the fact that closed-loop capacity remains high in the symmetric antenna configuration, and in structured (correlated) channels. In all case $SNR = 0$ dB, and the figures depict capacity increase as the number of transmit elements is increased. Note that the i.i.d. Rayleigh channel has the highest capacity in all cases. Closed-loop MIMO provide gains in particular when the number transmit antenna is larger than the number of receive antennas, and in cases where the channel gains are correlated

3.2.4 Capacity with imperfect CSI

Hybrid open-loop and closed-loop MISO systems, with only partial channel state information in the transmitter and with perfect CSI in the receiver, have been considered from capacity and quantization viewpoints in [87–91]. Similar studies, motivated by pairwise error probability criteria, were carried out in [92], and from a bit error probability point of view in [93, 94].

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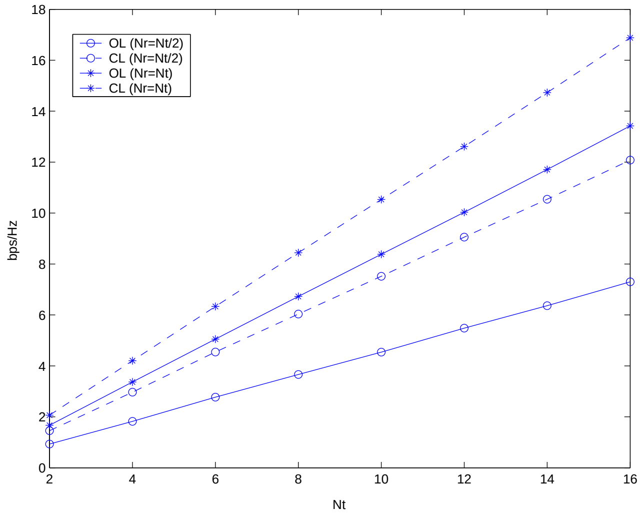

Fig. 3.2: MIMO capacity in iid Rayleigh channel at $SNR = 0$ dB with different number of transmit and receive antennas.

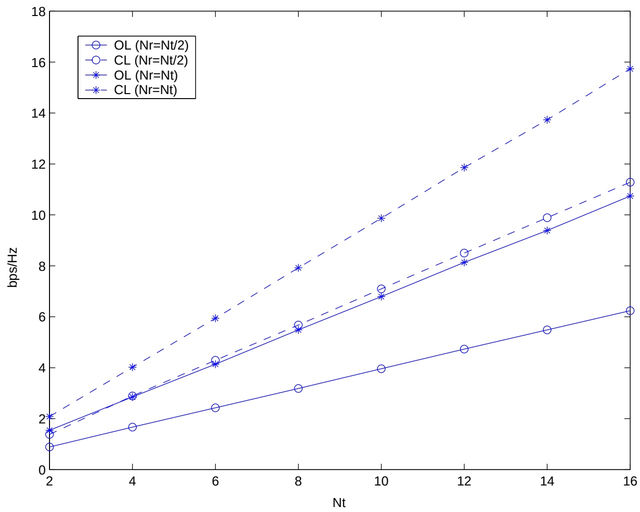

Fig. 3.3: MIMO capacity in Micro channel (PedB) $SNR = 0$ dB with different number of transmit and receive antennas

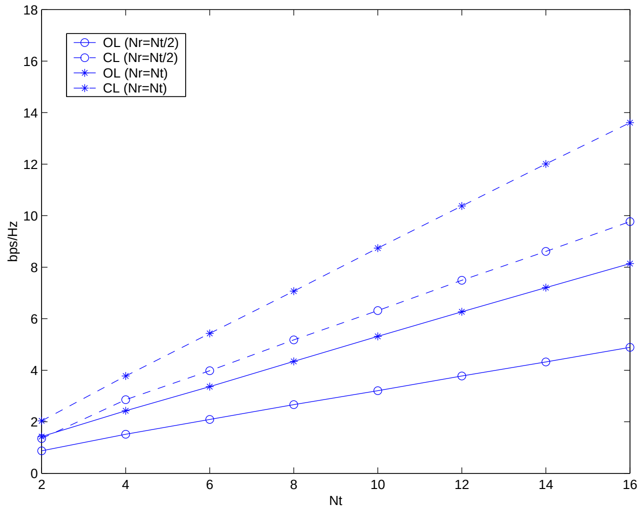

Fig. 3.4: MIMO capacity in Macro channel $SNR = 0$ dB with different number of transmit and receive antennas

Partial channel state information can be modelled with a stochastic channel characterization, or stochastic channel state information. In [89] two partial feedback strategies were considered. Therein, with "mean feedback" the transmitter assumes that the channel coefficients are multivariate complex Gaussian $N(\mathbf{w}, \delta^2\mathbf{I})$, where $\mathbf{w}$ and δ^2 model the channel mean, as specified by feedback, and the corresponding variance, respectively. In "covariance feedback", the channel (as assumed at the transmitter) is distributed as $N(\mathbf{0}, \mathbf{R}_{\text{Tx}})$, with zero mean and a given transmit covariance matrix. The covariance model is appropriate when the channel is changing rapidly and the mean feedback is unable to track or model the instantaneous channel dynamics. On the other hand, it is feasible to assume that the geometrical properties of the channel are more stable. In these cases, the receiver may estimate the covariance and signal it to the transmitter. Alternatively, under certain assumptions (e.g. calibrated arrays) $\mathbf{R}_{\text{Tx}}$ may be estimated at the transmitter from uplink measurements.

According to the results, hybrid closed-loop and open-loop transceiver concepts achieve a high diversity order, even in the presence of imperfect channel state information at the transmitter. See also related power allocation solutions for MISO systems with erroneous feedback in [95], and related independent results in [30]. Related power allocation studies for MIMO systems were given in [96, 97]. Optimal transmission strategies depending on the feedback quality and the channel covariance matrix were solved numerically in [89]. Necessary and sufficient conditions for achieving capacity with mean and covariance feedback

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were developed in [90]. Under these conditions beamforming and simple scalar coding are optimal in terms of achieving capacity and more complex matrix modulation schemes (or space–time coding) are not required.

Long-term beamforming, considered in [87, 89, 90, 98, 99] and suggested for MIMO long-term beams in [20, 100] is one way of exploiting partial channel information. The idea is to exploit the possible structure in the channel correlation matrix $\mathbf{H}\mathbf{H}^{\dagger}$. The correlation matrix is calculated by filtering over a number of instantaneous channel realizations. Power allocation between long-term beams, optimally water-filling [89] is applied to increase capacity.

4

Transmission Methods for MIMO channels

This Chapter discusses a number of transceiver concepts that are developed for Multiple-Input Multiple-Output channels. For open-loop systems, the modulation (or code) design criteria are presented, along with particular designs. In addition, multi-antenna extensions of selected closed-loop concepts described in Chapter 2 are discussed.

4.1 TERMINOLOGY

The signal model

$$\underset{T\times N_{\mathrm{r}}}{\mathbf{Y}} = \underset{T\times N_{\mathrm{b}}}{\mathbf{X}} \underset{N_{\mathrm{b}}\times N_{\mathrm{t}}}{\mathbf{W}} \underset{N_{\mathrm{t}}\times N_{\mathrm{r}}}{\mathbf{H}} + \underset{T\times N_{\mathrm{r}}}{\mathrm{noise}} \tag{4.1}$$

is adopted in this thesis. In developing transmission methods, the individual blocks $\mathbf{W}$ and $\mathbf{X}$ need to be defined. It is useful to discuss the terminology adopted here, before specifying the actual transmission concepts.

Single-stream modulation: In the context of this thesis, single-stream modulation refers to a modulation method in which the transmitted symbols are orthogonal to each other. Such a modulator may be realized in a Single-Input Single-Output (SISO) channel, by transmitting one symbol per channel use, or in multi-antenna channel by transmitting an orthogonal symbol matrix for which the symbol rate may in principle be arbitrary, but less than the number of transmit antennas.

Multi-stream modulation: Multi-stream or spatial multiplexing refers, in its purest form, to high symbol rate modulation concepts developed originally in [10, 13, 68, 101]. In these papers, the information stream is split into multiple parallel streams and the streams are transmitted in parallel using multiple transmit antennas. The streams are independently coded and modulated. The number of these streams is R_s, the symbol rate. With different spatial multiplexing units, diagonal (DBLAST) [68] or vertical (VBLAST) [13, 101], or some other high rate symbol modulator with limited diversity arise. As an example, in vector modulation $\mathbf{X}$ reduces to a vector

$$\mathbf{x} = [x_1,\ x_2,\ \ldots\ , x_{N_\mathrm{t}}],$$

where independent symbol streams are transmitted via different antennas to target a rate increase with factor N_t compared to single-stream modulation. However, the received symbols correlate, since the rows of $\mathbf{H}$ are generally non-orthogonal. Hence, components of the transmitted vector can be unambiguously detected only in special circumstances. Increasing the number of receive antennas (with uncorrelated channels) simplifies the detection task at the expense of increased hardware complexity.

Orthogonal space–time block codes: In space–time block codes, as defined in [8, 102] the space–time modulator maps the modulation symbols x to a space–time code (symbol) matrix $\mathbf{X}$ in a linear fashion and in such a way that the symbols are orthogonal at the receiver regardless of $\mathbf{H}$. The linear mapping is designed to fulfill various performance and diversity criteria. As an example, the WCDMA [14, 17, 48] system developed in 3GPP supports an open-loop scheme with a 2×2 space–time block code, a variant of the code proposed by [8]. The variant used in 3GPP, known as Space-Time Transmit Diversity (STTD), applies a space–time block code given by

$$\mathbf{X}(x_1, x_2) = \begin{bmatrix} x_1 & -x_2^* \\ x_2 & x_1^* \end{bmatrix}, \tag{4.2}$$

where column 1 is transmitted from antenna 1 and column 2 from antenna 2. The matrix is orthogonal for arbitrary orthogonal input alphabets and a linear receiver maintains optimality, as shown in [102]. Linear block codes may be written in general using symbols and their conjugates

$$\mathbf{X} = \sum_{k=1}^{Q} \mathbf{X}_k(x_k, x_k^*), \tag{4.3}$$

where

$$\mathbf{X}_k(x_k, x_k^*) = x_k \mathbf{B}_k^- + x_k^* \mathbf{B}_k^+ \tag{4.4}$$

where the set of $2Q$ matrices $\{\mathbf{B}_k^-\}$ and $\{\mathbf{B}_k^+\}$ satisfy the Radon-Hurwitz conditions [1, 102]. The number of matrices that satisfy these conditions determines the maximum symbol rate $R_\mathrm{s} = Q/T$.

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Non-orthogonal space–time block codes: In non-orthogonal space–time block codes(see e.g. [29, 31, 103, 104] and [P1,P2,P8]) the space–time modulator maps the modulation symbols x to a space–time code (symbol) matrix $\mathbf{X}$ in a linear fashion as with orthogonal space–time block codes. The difference is that at least two symbols interfere with each. These codes or modulators break the rate constraints posed by the orthogonality criteria (e.g. by using Q larger than what orthogonality constraint allows), and are able to provide higher symbol rates than their orthogonal counterparts. Also here, the linear mapping is designed to fulfill various performance and diversity criteria. One of the first examples such modulators is given by the 2 by 2 modulation matrix proposed in [P2]

$$\mathbf{X} = \mathbf{X}_A(x_1, x_2) \; + \; \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix} \mathbf{X}_B(y_3, y_4), \tag{4.5}$$

where $\mathbf{X}_A$ and $\mathbf{X}_B$ are two-dimensional orthogonal space–time block code matrices and $[y_1, y_2]^T = \mathbf{U}[x_3, x_4]^T$, where a particular unitary precoding matrix $\mathbf{U}$ is used to modify the input symbols to guarantee full diversity.

Matrix modulation: Matrix modulation is a general term that refers to the formation of a symbol matrix $\mathbf{X}$ in such a way that the rows of $\mathbf{X}$ are dependent of each other. In linear matrix modulation, the code matrix is formed as a linear combination of the symbols. Examples include orthogonal space–time block codes, non-orthogonal space–time block codes, and threaded algebraic space-time codes [29], to name a few examples, but not vector modulation or spatial multiplexing.

Space–time trellis codes: With space–time trellis codes, developed originally in [105], $T = 1$, the space–time modulator is a vector modulator, and $\mathbf{X}$ outputs sequences of $R_\mathrm{s} = N_\mathrm{t}$ symbols from the N_t transmit antennas in each symbol period. The mapping of bits or symbols to matrix $\mathbf{X}$ is typically nonlinear.

Open-loop transmission: Open-loop transmission, strictly defined, refers to transmitting $\mathbf{XHW}$ without any knowledge of transmission medium, channel realizations or its properties. The strict definition is not used here, however, and in the context of this thesis open-loop transmission refers simply to the case where $\mathbf{W} = \mathbf{I}$.

Closed-loop transmission: Closed-loop transmission in general assumes some knowledge of channel properties (or some derived quantities), and their use in controlling the transmission resources. The control may refer to selection or optimization of $\mathbf{W}$, the beamforming matrix, modulation matrix $\mathbf{X}$, transmit power, symbol rate etc. Knowledge of channel properties affects channel capacity, selection different modulation matrices $\mathbf{X}$, and also changes (typically mitigates) receiver complexity. It is also possible to devise hybrid open-loop/closed-loop concepts that utilize limited channel state state information, as shown e.g. in [P7] and [47, 92, 93, 95, 106].

Feedback concepts: Feedback concepts are closely related to closed-loop transmission methods described above. The term *feedback* highlights the fact that the receiver explicitly signals control information to the transmitter. In this thesis feedback concepts refer to different means of signalling $\mathbf{W}$ or $\mathbf{H}$ to the transmitter, assuming that only the receiver has full access to their true values. This is often the case in wireless systems using Frequency Division Duplex (FDD), due to lack of channel reciprocity. A particular feedback concept has been invented by the author and the co-inventor and it is currently supported by all WCDMA terminals. In this concept [15, 18] a beamforming vector $\mathbf{W}$ is signalled in a particular manner from the terminal to the base station.

Randomized transmission: Randomization techniques may be applied to many of the concepts discussed above. Here, a different (pseudo) randomly chosen beamforming matrix $\mathbf{W}$ is applied in the transmitter for different symbol or symbol sequences. Often the use of randomization techniques allows to simplify the modulator or demodulator for the subsequences and thus have engineering motivation. For example, some randomized concepts have a trivial space–time modulator $\mathbf{X}$, and $\mathbf{W}$ constitutes phase or antenna hopping [12, 107], which can be described by choosing different $1 \times N_{\mathrm{t}}$ beamforming vectors $\mathbf{W}$ for symbols transmitted in the different times. Some have a non-trivial space–time modulator $\mathbf{X}$, and the beamforming matrix performs multibeamforming by applying antenna permutations or multi-antenna hopping. Randomization can be used to improve the performance of a coded space–time transmission scheme, as shown e.g. in [12] and [P1].

4.2 OPEN-LOOP TRANSMISSION

4.2.1 Design Criteria

The original space–time coding contributions [102, 105, 108] suggest criteria for optimizing the code (or symbol) matrix such that full diversity is achieved. Perhaps the most commonly used design criteria are based on optimizing an upper bound of symbol error rate for one space–time coded symbol matrix. In a Ricean fading channel, the probability for making an

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error between two code words (matrices) was given in [105]. The bound therein is based on the codeword difference matrix

$$\mathbf{D}^{(\mathrm{ce})} = \mathbf{X}^{(\mathrm{c})} - \mathbf{X}^{(\mathrm{e})} \ . \tag{4.6}$$

The Ricean bound was shown to reduce in i.i.d. Rayleigh fading to

$$P_{c \mapsto e} \leq \det \left(\mathbf{I}_{N_\mathrm{t}} + \rho \mathbf{D}^{(\mathrm{ce})\ \dagger} \mathbf{D}^{(\mathrm{ce})} \right)^{-N_\mathrm{r}} \ . \tag{4.7}$$

The bound (4.7) should be evaluated for all possible pairs c, e and minimized for optimal performance. At high SNR ρ, this leads to the following criteria

- *Rank criterion* [108]:

$$\min_{e \neq c} \mathrm{Rank}[\mathbf{D}^{(\mathrm{ce})}] \ \leq \ \min[T, N_\mathrm{t}] \ . \tag{4.8}$$

- *Determinant criterion (MAX-MIN-DET)* [105]: In a (Rayleigh) fading environment, the code matrix should maximize

$$\min_{e \neq c} \det{}' \left[\mathbf{D}^{(\mathrm{ce})\dagger} \mathbf{D}^{(\mathrm{ce})} \right] \ . \tag{4.9}$$

In the determinant criterion the prime indicates the zero eigenvalues are discarded. The determinant criterion is a generalization of the product distance used in designing fading resistant constellations. For maximal diversity, $\mathbf{D}^{(\mathrm{ce})}$ should have full rank for all non-vanishing codeword pairs. For maximal coding gain, the minimum determinant computed over all codeword differences should be maximized.

There are various other criteria that have been used in designing fading resistant modulation schemes, see e.g. [109–111]. This determinant criteria is related to the product distance criteria, that has been used with signal space diversity [112], and deserves special attention. The minimum product distance of constellation $\mathbb{A}$ is given by

$$d_{min} = \min_{\boldsymbol{\Delta} = \mathbf{x}^{(\mathrm{c})} - \mathbf{x}^{(\mathrm{e})},\ \mathbf{x}^{(\mathrm{c})} \neq \mathbf{x}^{(\mathrm{e})} \in \mathbb{A}} \prod_{i=1}^{D} |\Delta_i| \tag{4.10}$$

Here $\mathbf{x}^{(\mathrm{c})}$ is the transmitted symbol vector, and $\mathbf{x}^{(\mathrm{e})}$ is an erroneously detected symbol vector. The symbol difference vector is $\boldsymbol{\Delta}$.

The product distance may be used define constellations that are robust to fading. Consider below a linear procoding (transformation) approach applied to a D-dimensional symbol constellation $\mathbb{A}$. The linear precoder $\mathbf{U}$ defines a set

$$\{\mathbb{A}_{tr} | \mathbf{U}\mathbf{x}, \mathbf{x} \in \mathbb{A}\}$$

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for the transformed signal constellation. The transformation matrix $\mathbf{U}$ is typically a unitary (rotation) matrix. The constellation rotation $\mathbf{U}$ can be optimized by maximizing the minimum product distance over all codeword pairs:

$$\mathbf{U} = \arg\max_{\mathbf{U}} \min_{\Delta=\mathbf{x}^{(c)}-\mathbf{x}^{(e)},\ \mathbf{x}^{(c)}\neq\mathbf{x}^{(e)}\in\mathbb{A}} \prod_{i=1}^{D} |(\mathbf{U}\boldsymbol{\Delta})_i| \,. \tag{4.11}$$

In addition, it is desirable to minimize the number of constellation vectors that meet the minimum product distance.

4.2.2 Orthogonal space–time block codes

Orthogonal space–time block codes, such as STTD or the Alamouti code, have always full rank. This is due to the linearity and orthogonality of the code matrix, which results in a full-rank diagonal (orthogonal) codeword difference matrix. In linear codes embedding Q symbols, the codeword difference matrix simplifies to

$$\mathbf{D}^{(\mathrm{ce})} = \mathbf{X}(x_1^{\mathrm{c}} - x_1^{\mathrm{e}}, x_2^{\mathrm{c}} - x_2^{\mathrm{e}}, ..., x_Q^{\mathrm{c}} - x_Q^{\mathrm{e}}) \,. \tag{4.12}$$

Although having full rank is a desirable property it comes with a price on supportable symbol rates. Full rate designs exist only for real modulation alphabets up to 8 antennas [102]. However, for complex modulation alphabets, no full rate linear decoding space–time block codes exist for more than two antennas [102,113]. Orthogonality is a demanding requirement and can only be fulfilled if the symbol rate is sufficiently low. For example, it can be shown that the maximal rate of a square unitary space-time design is [30,113] [P3]

$$\frac{\lceil \log_2 N_{\mathrm{t}} \rceil + 1}{2^{\lceil \log_2 N_{\mathrm{t}} \rceil}} \tag{4.13}$$

where N_{t} is the number of transmit antennas, and $\lceil \circ \rceil$ denotes the integer greater or equal to $\circ$.

The rate loss is not so pronounced when the number of transmit antennas is four. The first space-time block code that meets the maximal rate of $3/4$ with four antennas was proposed in [102]. This is given by

$$\mathbf{X}(x_1, x_2, x_3) = \begin{bmatrix} x_1 & x_2 & \frac{1}{\sqrt{2}}x_3 & \frac{1}{\sqrt{2}}x_3 \\ -x_2^* & x_1^* & \frac{1}{\sqrt{2}}x_3 & -\frac{1}{\sqrt{2}}x_3 \\ \frac{1}{\sqrt{2}}x_3^* & \frac{1}{\sqrt{2}}x_3^* & \frac{1}{2}(x_2 - x_2^* - x_1 - x_1^*) & \frac{1}{2}(x_1 - x_1^* - x_2 - x_2^*) \\ \frac{1}{\sqrt{2}}x_3^* & -\frac{1}{\sqrt{2}}x_3^* & \frac{1}{2}(x_1 - x_1^* + x_2 + x_2^*) & -\frac{1}{2}(x_1 + x_1^* + x_2 - x_2^*) \end{bmatrix} \tag{4.14}$$

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A three antenna version of this code is produced by turning off one of the antennas (deleting one column). The symbol rate is apparent from the fact that matrix has three input symbols but requires four time epochs for transmission.

Publications [6,113,114] and [P3] provide examples alternative space-time coding matrices that meet the same rate. Matrices with desirable properties may be found by by multiplying the code matrix from left or right by a unitary matrix. As an example, a simple form of rate 3/4 matrix for three or four antennas is

$$\mathbf{X}(x_1, x_2, x_3) = \begin{bmatrix} x_1 & x_2 & x_3 & 0 \\ -x_2^* & x_1^* & 0 & -x_3 \\ -x_3^* & 0 & x_1^* & x_2 \\ 0 & x_3^* & -x_2^* & x_1 \end{bmatrix}, \tag{4.15}$$

proposed originally in [113]. This may be obtained via unitary transformations on (4.14), as shown in [113] and [P3]. Similarly, considering transmission via 5 to 8 antennas [P3], one obtains full diversity rate 1/2 scheme,

$$\mathbf{X}(x_1, x_2, x_3, x_4) = \begin{bmatrix} x_1 & x_2 & x_3 & 0 & x_4 & 0 & 0 & 0 \\ -x_2^* & x_1^* & 0 & -x_3 & 0 & x_4 & 0 & 0 \\ -x_3^* & 0 & x_1^* & x_2 & 0 & 0 & x_4 & 0 \\ 0 & x_3^* & -x_2^* & x_1 & 0 & 0 & 0 & x_4 \\ -x_4^* & 0 & 0 & 0 & x_1^* & x_2 & x_3 & 0 \\ 0 & x_4^* & 0 & 0 & -x_2^* & x_1 & 0 & -x_3 \\ 0 & 0 & x_4^* & 0 & -x_3^* & 0 & x_1 & x_2 \\ 0 & 0 & 0 & x_4^* & 0 & x_3^* & -x_2^* & x_1^* \end{bmatrix}, \tag{4.16}$$

which is different from those obtained in [6].

Although the symbol rate loss cannot be avoided it is possible to increase the bit rate. For STBCs this should be done so that peak-to-average ration in transmitter is minimized. One feasible solution is relax the implicit assumption that all symbols are taken from the same modulation alphabet. This leads to *multimodulation* space-time codes [115]. In this scheme, the symbols that acquire a lower power in a single-modulation scheme like (4.14), can be selected from another higher order symbol alphabet modulating more bits. Naturally, more power can be allocated for these symbols. For example, if symbols x_1, x_2 are taken from the

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QPSK alphabet, and x_3 is in 8-PSK, and transmitted with double power, the code is given by

$$\mathbf{X}(x_1, x_2, x_3) = \begin{bmatrix} x_1 & x_2 & x_3 & x_3 \\ -x_2^* & x_1^* & x_3 & -x_3 \\ x_3^* & x_3^* & \frac{1}{2}(x_2 - x_2^* - x_1 - x_1^*) & \frac{1}{2}(x_1 - x_1^* - x_2 - x_2^*) \\ x_3^* & -x_3^* & \frac{1}{2}(x_1 - x_1^* + x_2 + x_2^*) & -\frac{1}{2}(x_1 + x_1^* + x_2 - x_2^*) \end{bmatrix}. \tag{4.17}$$

The bit-rate of this code is then $7/8 = 1/4 \sum_i \log_2(M_i)$, where $M_1 = M_2 = 2$ and $M_3 = 3$.

It is likely that in future wireless systems symbol rate cannot be compromised. If increased diversity benefits are desired, one has to design a non-orthogonal space-time block code that attains simultaneously a high symbol rate and a high diversity order.

4.2.3 Non-orthogonal space–time block codes and linear precoding

Fading resistant modulation methods using diversity transforms, or symbol precoders that output rotated constellations [109, 112, 116–118] are effective in increasing system performance without compromising decoding delay or bandwidth efficiency. Fading resistant multidimensional constellations can be constructed via linear transforms at the transmitter, without channel side information. The transform matrix often needs to be manually designed for the given channel, or for a given input constellation. In fading resistant modulation the coordinates of a rotated multidimensional signal are distributed over multiple signal dimensions, defined e.g. in space, time or frequency. Ideally, the rotation is such that the original information stream can be retrieved unless all signal coordinates fade simultaneously. Orthogonal or unitary precoders maintain the Euclidean distances between the signal states at precoder input and output. Some related techniques, such as Trellis coded modulation [119], also increase the robustness to fading, but they tend to impose a performance loss in a Gaussian channel.

It is apparent that in a conventional D-dimensional QPSK constellation the minimum product distance is zero, since all coordinate constellations are identical. The purpose of constellation rotation is to provide modulation diversity, or signal space diversity [112], such that the coordinate constellations are different.

4.2.3.1 *Examples assuming $N_\mathrm{r} = 1$* In order to appreciate the benefits of symbol precoding for spatial diversity consider the following illustrative example in the spirit of [111, 112, 120]. The presentation below differs from ones in the literature in that the method is described explicitly in terms of the received signal model and the ensuing correlations.

We assume that 2 QPSK symbols are transmitted orthogonally over a fading channel corrupted by Gaussian noise, during two symbol epochs. The rotation matrix is of dimension

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2×2. The QPSK vector with two complex dimensions is precoded, and the coordinates of the rotated symbol vector at the output of the precoder are transmitted via two orthogonal signalling channels. Orthogonal signalling is implemented e.g. with CDTD, such that the first complex coordinate is spread with code $\mathbf{s}_1$ transmitted from antenna 1, whereas the second coordinate is spread with code $\mathbf{s}_2$ and transmitted from antenna 2. Hence, we use a code matrix $\mathbf{S} = [\mathbf{s}_1\ \mathbf{s}_2]$.

In a flat fading channel the most transparent received signal model is given by

$$\mathbf{y} = \mathbf{SHUx} + \mathbf{n}\ ,$$

where the diagonal elements of $\mathbf{H} = \mathrm{diag}(h_1, h_2)$, designate the complex channel amplitudes between transmit antennas 1 and 2 and the receive antenna, $\mathbf{x}$ is a 2-dimensional QPSK symbol vector, and $\mathbf{n}$ is Gaussian noise. For comparison with other linear space–time modulation schemes, the signal model may be written according to Equation (4.1) as

$$\mathbf{y} = \mathbf{X}\,\mathbf{h} + \mathbf{n}\ , \tag{4.18}$$

where $\mathbf{h} = [h_1\ \ h_2]^{\mathrm{T}}$ is the channel vector. Denoting the rotated symbols by $\tilde{\mathbf{x}} = \mathbf{Ux}$, the matrix modulation is

$$\mathbf{X} = \mathbf{S}\begin{bmatrix} \tilde{x}_1 & 0 \\ 0 & \tilde{x}_2 \end{bmatrix}\ . \tag{4.19}$$

The symbol precoding matrix $\mathbf{U}$ assumes the following generic form

$$\mathbf{U}(\mu, \nu) = \begin{bmatrix} \mu & \nu \\ -\nu^* & \mu^* \end{bmatrix}, \tag{4.20}$$

with power constraint $|\mu|^2 + |\nu|^2 = 1$. It is a unitary 2×2 matrix with unit determinant. The received signal after matched filtering is

$$\mathbf{z} = \mathbf{U}^\dagger\mathbf{H}^\dagger\mathbf{S}^\dagger\mathbf{SHUx} + \mathbf{n}, \tag{4.21}$$

where the noise is generally colored. With arbitrary orthonormal signalling channels, $\mathbf{S}^\dagger\mathbf{S} = \mathbf{I}_2$. Then, the effect of the diversity transform is manifested in the correlation matrix

$$\mathbf{U}^\dagger\mathbf{H}^\dagger\mathbf{HU} = \begin{bmatrix} a_1 & b \\ b^* & a_2 \end{bmatrix} \tag{4.22}$$

where

$$a_1 = |h_1|^2|\mu|^2 + |h_2|^2|\nu|^2 \tag{4.23}$$

$$a_2 = |h_2|^2|\mu|^2 + |h_1|^2|\nu|^2 \tag{4.24}$$

$$b = (|h_1|^2 - |h_2|^2)\mu^*\nu. \tag{4.25}$$

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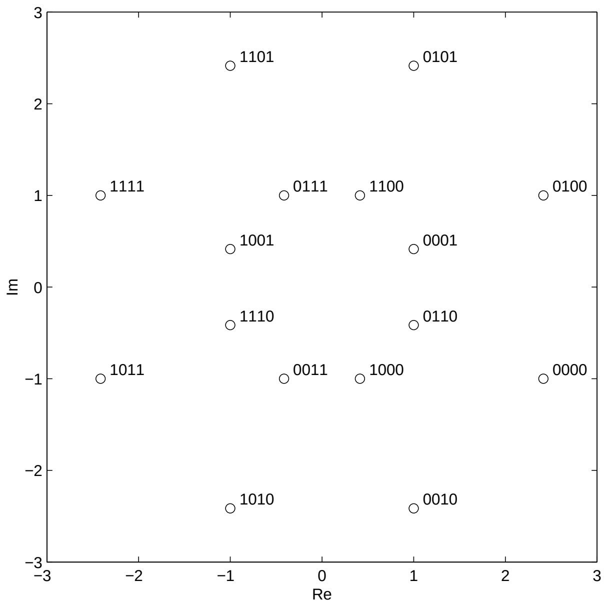

Fig. 4.1: Symbol alphabet and symbol labels for the first dimension of a two-dimensional rotated constellation.

The code is orthogonal if $\nu = 0$ or $\mu = 0$, or if $|h_1| = |h_2|$. With $|\mu| = |\nu|$ the energy of both symbols is distributed evenly over the two antennas.

Figures 4.1 and 4.2 show all possible signal states and related signal labels for the first and second elements of $\tilde{\mathbf{x}} = \mathbf{U}\mathbf{x}$, when $\mu = 1/\sqrt{2}$ and $\nu = \exp(\mathrm{j}\,\pi/4)/\sqrt{2}$. Note that to maintain orthogonal signalling, two symbol epochs are needed to transmit 4 bits. Therefore, the spectral efficiency remains at 2 bps/Hz, which is equivalent to the original QPSK constellation. It is apparent from Figures 4.1 and 4.2 that a properly selected rotation extends each symbol over multiple transmission intervals or more generally over multiple dimensions defined in time, space, code or frequency domain. As a result, each bit can be recovered unless channel states in both dimensions vanish simultaneously due to fading. On the other hand, if the precoding matrix $\mathbf{U}$ vanishes, both symbols are sampled from QPSK constellation, and both channel realizations need to be of sufficient quality to recover all four bits. Here, symbol precoding improves the diversity order at symbol level, and performance gain is not solely dependent on some well-designed outer code.

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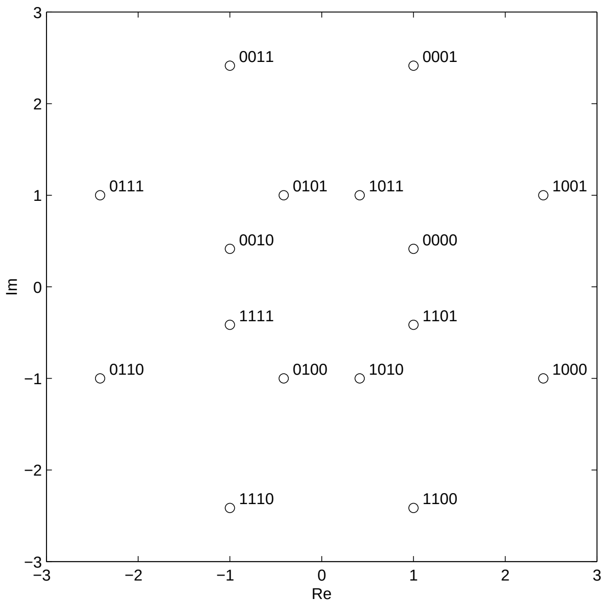

Fig. 4.2: Symbol alphabet and symbol labels for the second dimension of a two-dimensional rotated constellation.

In the previous example the code matrix was two-dimensional. In the presence of more than two transmit antennas, full modulation diversity cannot be achieved, unless the dimension of the precoding matrix is also increased. Often the matrix is found by a numerical search over some parameterized family of rotation matrices [109, 112]. Figure 4.3 depicts the determinant distance in a grid of values for ϕ and $|\mu|^2$ for case where the input alphabet is QPSK and $N_{\mathrm{t}} = 2$. The maximum value is 1, and this occurs with 45 degree phase rotation and equal power transmission. The numerical search for the optimal precoding matrix is obviously exacerbated when the matrix size is increased beyond two.

The concept above holds as a special case Orthogonal Transmit Diversity (OTD) that is adopted for the cdma2000 standard. Indeed, OTD arises when $\mathbf{S}$ comprises orthogonal CDMA spreading codes and when $\mathbf{U}$ vanishes, i.e. when the symbol precoding and related diversity enhancement is omitted. Antenna hopping (or Time-Switched Transmit Diversity, as it is called in 3GPP) is also a special case, in that is assumes a diagonal “code matrix” $\mathbf{S}$ and in analogy with OTD, no symbol precoding. Symbol precoding induces intentional correlations between symbols. When applied in multi-antenna systems, it effectively generates a non-orthogonal block code with improved diversity order. Below, we make the link between precoding and non-orthogonal space–time block codes more transparent.

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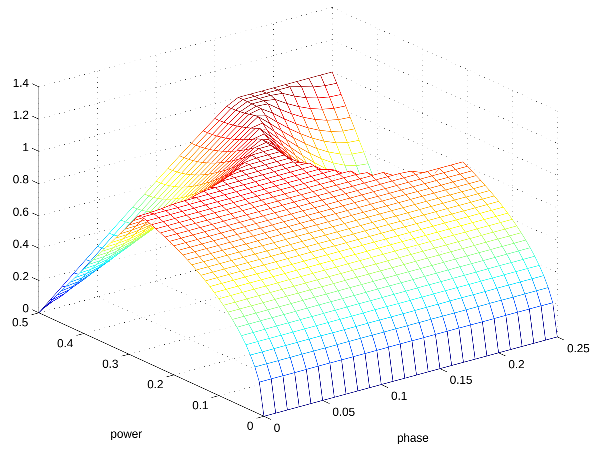

Fig. 4.3: Determinant distance for symbol rate one modulation as a function of phase (ϕ) and power difference for the two symbols.

With four transmit antennas and one receive antenna the maximal rate for O-STBC is 3/4. However, several rate one non-orthogonal STBCs do exist in this case, and they can be designed to reach full diversity. The ABBA code, proposed originally in [31], can be used with scalar-rotated symbols [121] to attain full diversity, when it only achieves diversity two without such rotations. Similarly, we may convert a diversity order two four antenna orthogonal space–time block code into a full diversity code with matrix-valued rotations. The diversity order two code, STTD-OTD , is summarized below and the effect of using matrix-valued rotations is demonstrated.

The transmission matrix for the combined STTD and Orthogonal Transmit Diversity (OTD) concept is

$$\mathbf{X} = \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{X}_A & \mathbf{X}_B \\ \mathbf{X}_A & -\mathbf{X}_B \end{bmatrix}. \tag{4.26}$$

Since the constituent codes $\mathbf{X}_A$ and $\mathbf{X}_B$ are rate one 2×2 STBC matrices, the overall transmission matrix supports four transmit antennas and has rate one. It is apparent from (4.26) that two STBCs are transmitted in parallel using orthogonal code-division multiplexing. When time-orthogonal signalling is used, an analogous transmission (omitting normalization) is

$$\mathbf{X}(x_1, x_2, x_2, x_4) = \begin{bmatrix} \mathbf{X}_A & \mathbf{0}_2 \\ \mathbf{0}_2 & \mathbf{X}_B \end{bmatrix}. \tag{4.27}$$

In terms of the determinant criteria, both designs above attain only diversity order two, the same as the constituent STBCs alone.

The different transmission methods may be mapped to each other with particular pre- or postcoding [1]. Also, the connection between ABBA [31]

$$\begin{bmatrix} \mathbf{X}_A & \mathbf{X}_B \\ \mathbf{X}_B & \mathbf{X}_A \end{bmatrix}$$

and (4.27) is apparent from

$$\frac{1}{2}\begin{bmatrix} \mathbf{I}_2 & \mathbf{I}_2 \\ -\mathbf{I}_2 & \mathbf{I}_2 \end{bmatrix}\begin{bmatrix} \mathbf{X}_A & \mathbf{X}_B \\ \mathbf{X}_B & \mathbf{X}_A \end{bmatrix}\begin{bmatrix} \mathbf{I}_2 & -\mathbf{I}_2 \\ \mathbf{I}_2 & \mathbf{I}_2 \end{bmatrix} = \begin{bmatrix} \mathbf{X}_A+\mathbf{X}_B & \mathbf{0}_2 \\ \mathbf{0}_2 & \mathbf{X}_A-\mathbf{X}_B \end{bmatrix}$$

Therefore, ABBA may be written in a diagonal form as in (4.27), only with linearly precoded symbol matrices. Such transformations do not change the (average) performance in i.i.d. Rayleigh channels [1].

The equivalent channel matrix $\mathcal{H}$ arises when converting the matrix model $\mathbf{Y} = \mathbf{XH} + \mathrm{noise}$ into a statistically equivalent vector model $\mathbf{y} = \mathcal{H}\mathbf{x} + \mathrm{noise}$ [1]. This may be used the get further insight into the properties of different modulators. Considering the Alamouti code as an example, one may conjugate the received signal during the second symbol period and the received signal may be written in terms of an equivalent signal model

$$\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \mathcal{H}\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \mathrm{noise}\ . \tag{4.28}$$

The “equivalent channel matrix” is

$$\mathcal{H} = \tfrac{1}{\sqrt{2}}\begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix}\ , \tag{4.29}$$

and the self-interference properties of Alamouti code (and STTD) are seen after applying the Hermitian conjugate of the equivalent channel matrix on the received signal (4.28):

$$\mathcal{H}^\dagger\begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \tfrac{1}{2}\left(|h_1|^2+|h_2|^2\right)\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \ \mathrm{noise}\ . \tag{4.30}$$

The diagonal part of the equivalent channel correlation matrix $\mathcal{R} = \mathcal{H}^\dagger\mathcal{H}$ is called in the following as the diversity part and is denoted as $\mathcal{D}_{\mathbf{h}}$. The off-diagonal part of $\mathcal{R}$ is the self-interference part, denoted as $\mathcal{S}_{\mathbf{h}}$, as specified in more detail in [1].

[1] For certain linear transmission methods this is not possible in complex domain (e.g. if a row of $\mathbf{X}$ contains symbols and its conjugates), and a conversion into an equivalent real channel with a higher dimensionality is needed [1].

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For the STTD-OTD transmission the same procedure is also applicable, and it easy to see that for a given antenna index configuration the diversity matrix and self-interference matrices are

$$\mathcal{D}_{\mathbf{h}} = \begin{bmatrix} \left(|h_1|^2 + |h_2|^2\right)\mathbf{I}_2 & \mathbf{0}_2 \\ \mathbf{0}_2 & \left(|h_3|^2 + |h_4|^2\right)\mathbf{I}_2 \end{bmatrix} \quad (4.31)$$

$$\mathcal{S}_{\mathbf{h}} = 0 \quad (4.32)$$

To attain full diversity, one can precode the symbols in using appropriately defined matrix-valued rotation $\mathbf{U}$ across antennas [1].

An attractive alternative to STTD-OTD is to concatenate STTD and phase hopping. This approach, called in short STTD-PHOP, first proposed in [107], is orthogonal and assumes even simpler decoding than STTD-OTD as the optimal detector with STTD-PHOP operates over only two symbols. This construction was later proposed in 3GPP [122] and analyzed in [1]. Here two identical copies of STTD are transmitted in parallel using four antennas, with pseudo-random phase hopping applied for consecutive STTD blocks. Consider a modulation matrix

$$\frac{1}{\sqrt{2}}\begin{bmatrix} \mathbf{X}_A & \mathbf{X}_A \end{bmatrix}, \quad (4.33)$$

and the right action matrix

$$\mathbf{W} = \begin{bmatrix} \mathbf{I}_2 & \mathbf{0}_2 \\ \mathbf{0}_2 & \Theta \end{bmatrix}, \quad (4.34)$$

where the phasor matrix is

$$\Theta = \begin{bmatrix} \mathrm{e}^{\mathrm{j}\,\theta_{t,3}} & 0 \\ 0 & \mathrm{e}^{\mathrm{j}\,\theta_{t,4}} \end{bmatrix}. \quad (4.35)$$

Applying this right action on the transmission matrix (4.33) rotates the symbols in the third and fourth columns of the modulation matrix. The received signal is

$$\mathbf{y} = \mathbf{X}\mathbf{W}\mathbf{h} + \text{noise}. \quad (4.36)$$

After maximum ratio combining over antenna pairs 1 and 3, and 2 and 4, respectively, the equivalent channel correlation matrix becomes

$$\mathcal{D}_{\mathbf{h}} = \begin{bmatrix} |h_1 + h_3\mathrm{e}^{\mathrm{j}\,\theta_{t,3}}|^2\mathbf{I}_2 & \mathbf{0}_2 \\ \mathbf{0}_2 & |h_2 + h_4\mathrm{e}^{\mathrm{j}\,\theta_{t,4}}|^2\mathbf{I}_2 \end{bmatrix} \quad (4.37)$$

$$\mathcal{S}_{\mathbf{h}} = 0. \quad (4.38)$$

Clearly, the scheme is orthogonal, $\mathcal{S}_{\mathbf{h}} = 0$. In STTD-PHOP, the phasor matrix is fixed during two symbol periods, but changes pseudo-randomly after that. Phase hopping sequences are defined such that $\{\theta_{t,3}\}$ and $\{\theta_{t,4}\}$ sample the interval $(0, 2\pi]$ evenly over a time period that corresponds to one channel-coded block. This phase hopping solution is similar to the randomization technique proposed in [P1]. When [P1] randomizes the off-diagonal elements of the equivalent channel correlation matrix, the STTD-PHOP modulator randomizes the corresponding diagonal. Neither STTD-PHOP nor randomized ABBA in [P1] give gain in the absence of an outer channel code.

4.2.3.2 Examples assuming $N_\mathrm{r} > 1$ The modulation matrices described in the previous section can obviously be used with any number of receive antennas. However, they do not reach capacity in a Rayleigh fading channel when $N_\mathrm{r} > 1$. This was shown in [123], together with modulation matrices that are specifically designed with capacity criteria. Some of the first matrix modulation concepts that are explicitly designed to provide high capacity were proposed in [104, 124–129]

In threaded space–time coding obtaining full diversity and full rate each thread is active during all symbol intervals and symbols in each thread are distributed over all antennas. The transmission matrices may be written by

$$\mathbf{X} = \sum_{i=1}^{R_\mathrm{s}} \Pi^i \mathrm{diag}[\gamma_i \mathbf{U} \mathbf{x}_{i,\cdot}], \tag{4.39}$$

where Π is a permutation (shift) matrix and R_s the symbol rate, $\mathbf{x}_{i,\cdot}$ is a symbol vector transmitted on ith thread, $\mathbf{U}$ is a precoding matrix and γ_i is a (unit power) complex number. As an example, a 4×4 threaded code with $R_\mathrm{s} = 4$ may be written as

$$\mathbf{X} = \begin{bmatrix} \tilde{x}_{1,1} & \tilde{x}_{2,1} & \tilde{x}_{3,1} & \tilde{x}_{4,1} \\ \tilde{x}_{4,2} & \tilde{x}_{1,2} & \tilde{x}_{2,2} & \tilde{x}_{3,2} \\ \tilde{x}_{3,3} & \tilde{x}_{4,3} & \tilde{x}_{1,3} & \tilde{x}_{2,3} \\ \tilde{x}_{2,4} & \tilde{x}_{3,4} & \tilde{x}_{4,4} & \tilde{x}_{1,4} \end{bmatrix}. \tag{4.40}$$

It is clear that the precoded symbols adhere with vector modulation, only that they are concatenated over $T = 4$ symbol periods. The precoder $\mathbf{U}$ that outputs modified constellations $\{\tilde{x}_{i,j}\}$ links the symbols together and increases diversity per thread and $\{\gamma_i\}$ must be optimized to maintain full diversity across threads. In terms of terminology, symbols $\tilde{x}_{1,1}, \ldots, \tilde{x}_{1,4}$ belong to the first thread, symbols $\tilde{x}_{2,1}, \ldots, \tilde{x}_{2,4}$ to the second, $\tilde{x}_{3,1}, \ldots, \tilde{x}_{3,4}$ to the third and $\tilde{x}_{4,1}, \ldots, \tilde{x}_{4,4}$ to the fourth.

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One option in rotating the symbol-vectors in each thread to produce $\tilde{\mathbf{x}}_{i,\cdot} = \gamma_i \mathbf{U}\mathbf{x}_{i,\cdot}$ is to apply Hadamard transform

$$\mathbf{U} = \frac{1}{2}\begin{bmatrix} 1 & 1 & 1 & 1 \\ 1 & -1 & 1 & -1 \\ 1 & 1 & -1 & -1 \\ 1 & -1 & -1 & 1 \end{bmatrix}. \tag{4.41}$$

Another solution, adopted in [104], is to use symbol rotation matrices $\mathbf{U}$ that are based on Vandermonde matrices. To make the $R_\mathrm{s} > 1$ rate transmission full rank, algebraic ("Diophantine") rotations γ_i have been suggested for the threads in [29, 104, 129], and the resulting scheme is called here Threaded Algebraic Space-Time Coding (TAST), regardless of the selected precoding matrix.

The threaded construction in (4.40) defines a symbol rate four transmission scheme and requires four receive antennas. A symbol rate two TAST code proposed in [29] punctures two layers and the corresponding transmission matrix reads

$$\mathbf{X} = \begin{bmatrix} \tilde{x}_{1,1} & 0 & 0 & \gamma\tilde{x}_{4,1} \\ \gamma\tilde{x}_{4,2} & \tilde{x}_{1,2} & 0 & 0 \\ 0 & \gamma\tilde{x}_{4,3} & \tilde{x}_{1,3} & 0 \\ 0 & 0 & \gamma\tilde{x}_{4,4} & \tilde{x}_{1,4} \end{bmatrix}. \tag{4.42}$$

where the Diophantine number $\gamma = \exp(-\mathrm{j}\,\pi/6)$ and precoding matrix

$$\mathbf{U} = \begin{bmatrix} a & b & -c & -d \\ -b & a & d & -c \\ c & d & a & b \\ -d & c & -b & a \end{bmatrix} \tag{4.43}$$

with $a = 0.2012, b = 0.3255, c = -0.4857, d = -0.7859$, the optimal real precoding matrix defined for symbol rate one transmission in [112]. This code requires only two receive antennas.

For related contributions, the first high rate construction adopted Weyl basis [123], and some contributions use quasi-orthogonal Clifford bases [1, 130, 131]. The symbol rate four and symbol rate two TAST codes were compared in [P5] to Clifford-based DABBA and QABBA codes, as proposed originally in [P4] and [1, 130], respectively. Therein, QABBA and DABBA were shown to outperform the corresponding TAST codes. The DABBA variant, considered in [1] and [P8], that transmits $\mathbf{X}_A, \mathbf{X}_B, \mathbf{X}_C$ and $\mathbf{X}_D$ (four STTD blocks) encod-

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ing the symbol pairs (x_1, x_2), (x_3, x_4), (x_5, x_6) and (x_7, x_8), modulator takes the form

$$\mathbf{X}_{iDABBA} = \frac{1}{\sqrt{2}} \begin{bmatrix} \mathbf{X}_A + \mathbf{X}_C & \mathbf{X}_B + \mathbf{X}_D \\ \mathrm{j}\,(\mathbf{X}_B - \mathbf{X}_D) & \mathbf{X}_A - \mathbf{X}_C \end{bmatrix} \tag{4.44}$$

It is made full rank with a simple diagonal precoder that rotates the information symbols in STTD blocks $\mathbf{X}_C$ and $\mathbf{X}_D$. Placing j in the lower left corner still improves performance over the DABBA proposed in [P4]. It should also be noted that the DABBA code proposed in [P4] requires symbol rotations to attain full diversity. Interestingly, as shown in [P8], DABBA holds a two antenna TAST code as a special case. This occurs when either odd or even columns are punctured from the DABBA code matrix. The QABBA code may be written as [1]

$$\begin{aligned} \mathbf{X}_{QABBA} \quad = \quad & \begin{bmatrix} \mathbf{X}_A & \mathbf{X}_B \\ \mathbf{X}_B & \mathbf{X}_A \end{bmatrix} + \begin{bmatrix} \mathbf{X}_C & \mathbf{X}_D \\ -\mathbf{X}_D & -\mathbf{X}_C \end{bmatrix} \\ & + \mathrm{j} \begin{bmatrix} \mathbf{X}_E & \mathbf{X}_F \\ \mathbf{X}_F & \mathbf{X}_E \end{bmatrix} + \mathrm{j} \begin{bmatrix} \mathbf{X}_G & \mathbf{X}_H \\ -\mathbf{X}_H & -\mathbf{X}_G \end{bmatrix}, \end{aligned} \tag{4.45}$$

where $\mathbf{X}_A(x_1, x_2), \mathbf{X}_B(x_3, x_4), \ldots, \mathbf{X}_H(x_{15}, x_{16})$ are orthogonal space–time block codes e.g. of the Alamouti form. Transmissions with lower symbol rates are realized by placing zeroes in place of symbols, thus essentially puncturing layers. In summary, with appropriate precoding, QABBA [1], DABBA [P4], ABBA [31] all capture full rate and full diversity using orthogonal STBCs as constituent matrices.

Returning to the case with $N_\mathrm{t} = 2, R_\mathrm{s} = 2$, a quasi-orthogonal transmission scheme was proposed in [P2]. The code, called here Twisted STTD (TwSTTD), is of the form

$$\mathbf{X} = \mathbf{X}_A(x_1, x_2) \; + \; \begin{bmatrix} 1 & 0 \\ 0 & -1 \end{bmatrix} \mathbf{X}_B(y_3, y_4) \tag{4.46}$$

where $\mathbf{X}_A$ and $\mathbf{X}_B$ are two-dimensional orthogonal space–time block code matrices (such as STTD in WCDMA specification) and $[y_1, y_2]^T = \mathbf{U}[x_3, x_4]^T$. The Precoding matrix

$$\mathbf{U} = \frac{1}{\sqrt{7}} \begin{bmatrix} 1 + \mathrm{j} & 1 + 2\mathrm{j} \\ -1 + 2\mathrm{j} & 1 - \mathrm{j} \end{bmatrix} \tag{4.47}$$

was proposed in [130] to maximize coding gain for QPSK input. With proper precoding, TwSTTD above uses full-diversity STBCs as constituent codes, and overlays two different codes. This code has the highest coding gain among all codes tabulated in [132]. It indeed has the highest coding gain if all symbols are treated equally. However, the code is not a global optimum, according to this criterium, since the rate two full diversity 2×2 codes proposed in [29] and [P8] have a slightly better coding gain than this. Other high rate space–time modulators have been proposed and analyzed in [123, 124, 133].

4.2.4 Detection

Although this thesis is not focused on detection algorithms, the general detection problem is useful to formulate. The thesis discusses mainly coherent space–time transmission methods and require proper channel estimation in the receiver, to obtain an estimate of $\mathbf{H}$. Having $\mathbf{H}$, the equivalent channel matrix can be constructed and the detection problem can be posed. If needed, iterative channel estimation, by using decision feedback channel estimation e.g. using soft information provided by a channel decoder can be used [58, 134]. The methods for channel estimation and detection are in fact very much the same as in multiuser detection. A pilot-based channel estimate can be combined with a blind channel estimate [135] to further improve the quality of the channel estimate.

Given the channel estimate and an equivalent signal model a parametric Maximum Likelihood (ML) detector solves

$$\hat{\mathbf{x}}_{\mathrm{ml}} = \arg \max_{\mathbf{x} \in \mathbb{A}} \Omega(\mathbf{x}) \tag{4.48}$$

where $\mathbb{A}$ is the modulation alphabet and

$$\Omega(\mathbf{x}) = 2 \,\mathrm{Re}\left(\mathbf{x}^\dagger \mathbf{z}\right) - \mathbf{x}^\dagger \mathcal{H}^\dagger \mathcal{H} \mathbf{x}. \tag{4.49}$$

This problem may be rewritten as

$$\hat{\mathbf{x}}_{\mathrm{ml}} = \arg \min_{\mathbf{x}} \|\mathbf{z} - \mathcal{H}^\dagger \mathcal{H} \mathbf{x}\|_{\mathbf{Q}}^2, \tag{4.50}$$

where

$$\|\mathbf{v}\|_{\mathbf{Q}}^2 = \mathbf{v}^\dagger \mathbf{Q} \mathbf{v}$$

and $\mathbf{Q} = \mathcal{R}^{-1}$ is the weighting matrix where the equivalent channel correlation matrix is $\mathcal{R} = \mathcal{H}^\dagger \mathcal{H}$.

The number of signal states explodes as the size of the equivalent channel matrix increases. Thus, it is important to use efficient multi-channel detectors in the spirit of multiuser detection [101, 136–143]. Joint decoding and detection in the spirit of [144] or [145–153] may also be resorted to when decoding complexity requirement can be somewhat relaxed.

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4.3 CLOSED-LOOP CONCEPTS

Closed-loop transmit diversity solutions were originally proposed in [87,98,154,155]. However, their practical relevance remained undercover until particular solutions were proposed in 3G WCDMA standardization forums [40]. Therein, it was shown that very crude feedback signalling, with similar feedback rate as in closed-loop power control, can be extremely useful in improving the downlink performance. The 3G solutions proposed by the author and co-author also introduced the concept of antenna verification, which mitigates the effect of imperfect feedback [41].

WCDMA closed-loop transmit diversity has been analyzed e.g. in [156, 157] assuming unquantized feedback. Rate distortion theory [158] has been applied to the quantization problem in [159] assuming that channel and transmit weight vectors are jointly Gaussian-distributed. In addition to theoretical performance assessment, field experiments related to different closed-loop modes have been conducted with different antenna and feedback mode configurations [160]. Ultimate field experiments are eventually carried out in practice, since the support for the techniques currently is implemented in all WCDMA terminals.

Below, some these techniques are refined for FDD systems, following in part a description in [1]. In general, feedback signalling in FDD systems remains an important topic both practical applications and in theory. Practical relevance of enhanced concepts involving statistical channel information [88–91] is yet to be determined.

4.3.1 Beamforming feedback

The most straightforward solution for providing channel feedback is to quantize the individual elements of matrix $\mathbf{H}$ and signal them with appropriate signalling channels to the transmitting unit. A related concept is to feedback the desired beam or beam parameters to the transmitting unit. Below, some signal processing solutions related to these concepts are summarized, and extensions for multi-antenna transmission are given.

4.3.1.1 Potential WCDMA enhancements As stated in Chapter 2, closed-loop transmit diversity uses a simple two-slot filtering concept on the feedback signal. Improred filtering concepts may be applied at the base station in order to mitigate the effect of unreliable feedback signalling, while simultaneously improving the resolution of derived transmit beam. Adaptive filter lengthwas proposed in [161]. Reliability-based filtering for a binary AWGN signalling channel was proposed in [42]. In these solutions the diversity weight is given e.g.

by

$$w_2[t] = 1/N \sum_{k=t}^{t-N+1} \mathrm{j}^{\,(k \bmod 2)/2} \tanh(|h[k]|z[k]/\sigma^2), \tag{4.51}$$

where h is here the amplitude of the feedback channel, N is the filter length, and σ^2 is the noise variance at the base station. Reliability aspects are addressed with the $\tanh$ function. This soft-limiting function powers off the diversity antenna when the signalling is unreliable.

Another possibility for providing robustness to feedback errors is to use adaptive weighting for the matrix modulation method in question, depending on feedback reliability. The concept described below, Soft-Weighted STBC [95] (proposed independently also in [30]), may be applied to the Alamouti code as follows

$$\begin{bmatrix} y_1 \\ y_2 \end{bmatrix} = \begin{bmatrix} a_1 x_1 & a_2 x_2 \\ -a_1 x_2^* & a_2 x_2^* \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2 \end{bmatrix}, \tag{4.52}$$

where h_1 and h_2 refer to the complex channel coefficients between transmit antennas 1 and 2 and the receiver, respectively. Relative weighting factors a_1 and $a_2 = \sqrt{1 - a_1^2}$ are imposed on signals transmitted from antennas 1 and 2, but otherwise the signal respects the Alamouti code structure. The inherent simplicity of the method is one of its strong merits, while the discussion above related to antenna weighting or selection, and it can be combined with beamforming equally well. See also [162] for related work.

Other approaches, which also enhance beamforming resolution via channel probing and subspace tracking were proposed in [163, 164]. The probing concept uses perturbed pilot channels and low-rate feedback, and the transmitter reconstructs the transmit weight using the *a priori* known probing dynamics and sign-feedback. For related work, see also [165]. This solution is advantageous in that it mitigates the need for the receiver to know multiple pilot signals.

4.3.1.2 Multi-antenna extensions A framework for multi-antenna extensions is given in Figure 4.4. When developing multi-antenna solutions the optimal beamforming solution with perfect CSI (for single-stream transmission), maximizing the power of received signal, corresponds to the dominant eigenvector of $\mathbf{H}\mathbf{H}^\dagger$. This solution is analogous to the WCDMA closed-loop modes, if one neglects quantization issues pertinent to FDD systems. In place of using a fixed codebook, the receiver may quantize different coordinates of the transmit weight vector or weight matrices independently of each other. A number of tradeoffs involved with this procedure, and the codebook-based concept, are analyzed in [166, 167] and summarized in [1]. The quantization resolution, feedback rate, quality of feedback channel all affect the eventual performance of the system. Despite the tradeoffs, the potential gains have

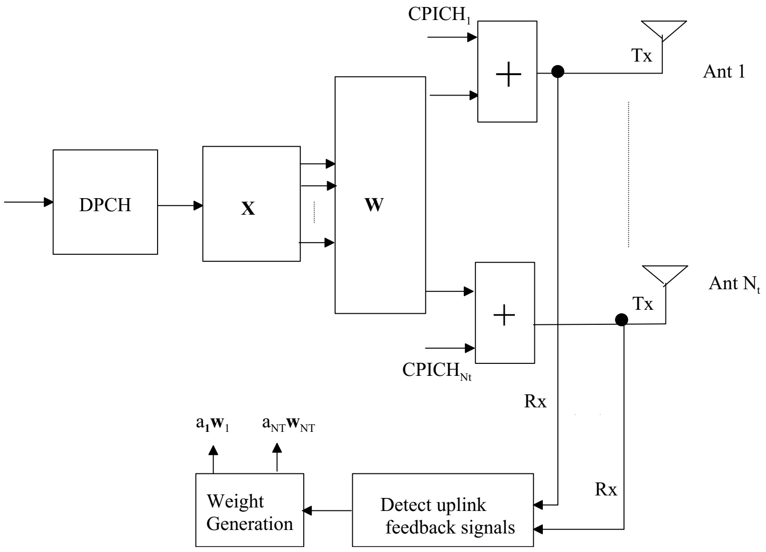

Fig. 4.4: Multi-antenna transmitter with feedback

been found to be large, warranting their further study when specifying the physical layers in wireless systems.

Beamforming feedback (or implicit feedback) refers here to the case where the feedback signalling defines a beamforming matrix, but the transmitter remains unaware of the channel. This is the solution adopted in WCDMA, wherein only one beamforming weight is signalled to the base station. Consider the following example in an attempt to extend this solution to larger arrays. Assume that the receiver knows a "codebook " of feasible beamforming matrices, and this set is given by a Discrete Fourier Transform (DFT) matrix **F**,

$$\mathbf{F} = \begin{bmatrix} 1 & 1 & 1 & \cdots & 1 \\ 1 & w & w^2 & \cdots & w^{(M-1)} \\ 1 & w^2 & w^4 & \cdots & w^{2(M-1)} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 1 & w^{(M-1)} & w^{2(M-1)} & \cdots & w^{(M-1)(M-1)} \end{bmatrix}, \tag{4.53}$$

where $w = e^{-\mathrm{j}\,2\pi/M}$. If $N_{\mathrm{t}} = M$, one of the M orthogonal columns is designated as the transmit weight vector, after normalization by dividing the vector with $\sqrt{M}$.

The receiver feedbacks an index corresponding to a desired beamforming solution as follows. With given N_{t} and M the set of feasible beams is

$$\left\{\mathbf{w}_l = [1\ , w^l, ..., \ w^{(N_{\mathrm{t}}-1)l}]^T/\sqrt{N_{\mathrm{t}}}\right\}_{l=0}^{M-1}. \tag{4.54}$$

,

The terminal picks the SNR maximizing solution by constructing a channel matrix $\mathbf{H} = [\mathbf{h}_1, \mathbf{h}_2, ..., \mathbf{h}_{N_\mathrm{t}}]^T$, where $\mathbf{h}_m$ is the impulse response between the mth array element and the terminal. The optimal DFT basis vector can be found as a solution to problem [1]

$$\hat{l} = \arg \max_{l \in \{0,...,(M-1)\}} \mathbf{w}_l^\dagger \mathbf{H}\mathbf{H}^\dagger \mathbf{w}_l. \tag{4.55}$$

Thus, only the index $\hat{l}$ needs to be signalled to the transmitter. If a calibrated Uniform Linear Array (ULA) is used in the transmitter, with element separation $d = \lambda/2$, the weight vectors are associated with different transmit directions in analogy with fixed beam transmission.

The solutions above are based on instantaneous channel realizations. Using the fact that geometric channel properties, such as dominant reflectors or dominant transmit or receive directions, are invariant to frequency separation, some simplifications related to feedback are possible. Indeed, if the beamforming is intended to match only to the long-term properties of the channel, the beamforming coefficients may be estimated either from uplink or downlink measurements. Downlink based long-term beamforming solutions are advocated in [20, 98, 99, 155, 168]. Therein, as an exanmple, the terminal maintains an estimate of the averaged correlation matrix

$$\mathbf{R}_{\mathrm{Tx}}[t_0] = C \sum_{t=t_0-P}^{t_0} \mathbf{H}[t]\mathbf{H}^\dagger[t], \tag{4.56}$$

where $\mathbf{H}[t]$ constitutes the instantaneous channel matrix at measurement slot t and C is a normalization coefficient. The integration window P typically extends over the channel coherence time. In structured or correlated channels this average correlation matrix has only a few dominant eigenvalues and the dominant long-term beams are defined (at slot t_0) as the dominant eigenvectors of $\mathbf{R}_{\mathrm{Tx}}[t_0]$ using the eigenvalue decomposition

$$\mathbf{R}_{\mathrm{Tx}}[t_0]\mathbf{E} = \Lambda \mathbf{E}.$$

Having calculated these vectors, they (or their parameters) are signalled to the base station. The beamforming matrix at the base station is set to $\mathbf{W} = \mathbf{E}$. Then, any open-loop method can be used to distribute the signals to the columns of $\mathbf{W}$, each column representing one eigenbeam. In effect, this approach reduces to a change of basis combined with long-term power allocation. Eventually, the receiver experiences the effective channels

$$\mathbf{h}_{\mathbf{e}_j}[t] = \mathbf{H}^\dagger[t]\mathbf{e}_j, \; j = 1, ..., L. \tag{4.57}$$

where L designates the number of long-term beams.

4.3.1.3 MIMO extensions In a MIMO channel, where a terminal has N_r receive antennas, the feedback weight can be found as a dominant eigenvector corresponding to

$$\mathbf{R}_{\mathrm{Tx}} = \sum_{n=1}^{N_\mathrm{r}} \mathbf{H}_{[n]} \mathbf{H}_{[n]}^{\dagger}, \tag{4.58}$$

where $\mathbf{H}_{[n]}$ is a matrix that contains the channel coefficients from N_t transmit antennas to receive antenna n. Otherwise the solutions given above hold, and the associated transmitter structure for closed-loop transmit diversity is depicted in Figure 4.4.

The capacity results derived for MIMO and MISO channels with imperfect feedback (mean or covariance feedback) suggest that eigenvalues and eigenvectors of the channel correlation matrix $\mathbf{R}_{\mathrm{Tx}}$ with eigenvalue decomposition

$$\mathbf{R}_{\mathrm{Tx}} = \mathbf{E}^{\dagger} \Lambda \mathbf{E} \tag{4.59}$$

are relevant in attempting to reach the capacity. The dominant beams are defined as the eigenvectors corresponding to the (ordered) largest eigenvalues $\Lambda = \mathrm{diag}(\lambda_1, ..., \lambda_{N_\mathrm{t}})$. When the corresponding unit power eigenvectors are used in the transmitter, the average power amplification of eigenbeam k is λ_k.

In MIMO solution we thus choose to use eigenbeams as an inherent part of the precoding matrix $\mathbf{W}$ within the signal model

$$\mathbf{Y} = \mathbf{XWH} + \mathrm{noise}, \tag{4.60}$$

where the matrix $\mathbf{X}$ is the space–time modulation matrix, $\mathbf{W}$ is a precoding matrix at the transmitter, and matrix $\mathbf{H}$ contains the MIMO channel coefficients, as described in Equation (4.1). The eigenbeams enter into the factorization for the beamforming matrix,

$$\mathbf{W} = \mathbf{\Pi A E}, \tag{4.61}$$

where $\mathbf{A}$ is a diagonal matrix that specifies the transmit amplitudes for each eigenbeam, and $\mathbf{\Pi}$ is a permutation matrix that can be used to shuffle the columns of $\mathbf{X}$ to appropriate eigenbeams. Therefore, with power allocation the average received signal power for the kth eigenbeam is $\lambda_k a_k^2$, and for a given channel realization $\mathbf{H}$, the receiver experiences channel

$$\tilde{\mathbf{H}} = \mathbf{WH}. \tag{4.62}$$

From a performance point of view, $\mathbf{X}$ and $\mathbf{W}$ should be jointly optimized [92]. In practice, a sequential approach, where $\mathbf{X}$ is fixed and $\mathbf{W}$ is selected optimally conditioned on $\mathbf{X}$, also works sufficiently well. Conditional bit-error-rate (BER) was suggested as a performance measure in [P7]. The conditional BER can be computed in closed form for a decorrelating detector and it may be approximated for the LMMSE detector. The conditional performance measure, or Channel Quality Indicator, may be computed for a set of feasible beams and the beamforming matrix that reaches minimal BER is selected. The power allocation may also be done using MMSE or outage criteria in place of mutual information [73].

4.3.2 Duplex hopping

The research on feedback concepts is currently focused on the case of imperfect feedback in FDD systems [89, 92–94]. Accurate high resolution signalling that renders $\mathbf{H}$ known to the transmitter is not considered feasible. However, we may depart from the signalling based solution, and modify the FDD principle to improve the quality of CSI at the transmitter. With the modified FDD system, suggested below, the explicit feedback signal may be dispensed.

Assume that the two duplex directions are assigned different carrier frequencies f_1 and f_2. In the modified FDD system, coined herein as *Duplex Hopping FDD*,

- f_1 is used for transmission in the first duplex direction during slot t_1 and
- f_1 is used for transmission in the second duplex direction during slot t_2.

Simultaneously,

- f_2 is used for transmission in the second duplex direction during slot t_1 and
- f_2 is used for transmission in the second duplex direction during slot t_2.

In the context of wireless networks, first and second duplex direction may refer to uplink and downlink directions, respectively. With this approach, the channel used in reception during slot t_1 (t_2) is used for transmission during slot t_2 (t_1), as shown in Fig. 4.5, and channel reciprocity akin to that obtained via Time Division Duplex (TDD) is available. A guard interval may be placed between the slots.

With accurate CSI the capacity-optimal transmision solutions for closed-loop MIMO, described in Chapter 2, become accessible. Such solutions are advantageous also from receiver viewpoint, since with perfect CSI at transmitter and receiver the optimal parallel channels are orthogonal and complex multichannel reception can be avoided. These advantages suggest that Duplex hopping-based solutions for paired frequency bands may constitute a fruitful research area in an attempt to improve the spectral efficiency of future wireless systems. However, the detailed analysis of this method is beyond the scope of this thesis.

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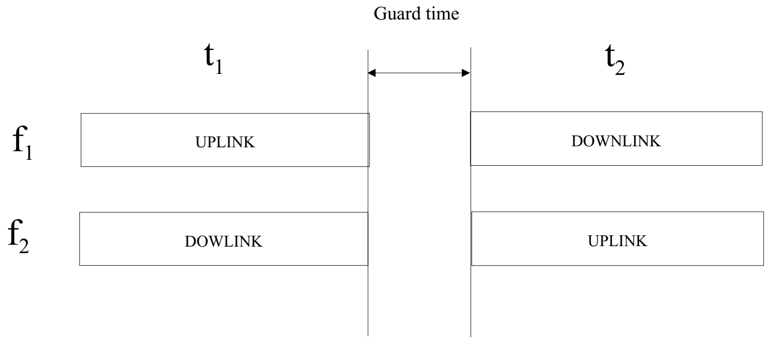

Fig. 4.5: Duplex hopping with two carriers f_1 and f_2 and two time slots t_1 and t_2. Guard interval may be placed between hops.

4.4 RELATED FEEDBACK CONCEPTS

4.4.1 MIMO and Multiuser Diversity

The potential benefits of channel-aware multiuser diversity were originally analyzed in [49, 169]. These information-theoretic studies were converted to engineering formalism and adopted to practical wireless systems in [48, 53, 170]. Scheduling based solutions provide throughput gain by exploiting the relative differences in the channel gains of different users. Multiple antennas may be used to increase diversity, in which case the scheduling gains may be negative when compared to single antenna transmission [23]. On the other hand, multiple antennas may be used instead to increase channel variability as shown in [24] or to increase the size of the modulation alphabet and thus peak rates.

MIMO modulation and scheduling were combined [171], where it was shown that joint spatial multiplexing and scheduling are advantageous, provided that the scheduling metric is defined properly. Furthermore, it was shown that different users may need to be assigned to different transmit antennas. Channel quality indicators relevant for scheduling in a matrix-modulated MIMO system were proposed by the author in [P6]. There, it was shown that the inter-stream interference, embedded in matrix-modulated transmission, has to be taken into account. The combined use of MIMO modulation and multiuser diversity remains an important research area. As an example, sufficiently simple-to-compute channel quality indicators need to be defined, and their effect on performance need to evaluated in a realistic system. Furthermore, means of reducing the required feedback capacity e.g. by using threshold-based feedback mechanisms [172] are of practical interest .

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4.4.2 MIMO and ARQ

Automatic Repeat Request (ARQ) protocols are included in many wireless systems, such as 3G, as noted in Chapter 2. Different ARQ concepts differ e.g. in the way incremental redundancy is exploited in the transmitter and the receiver. In the presence of a MIMO channel, ARQ may be defined so that by combining multiple packets the diversity order of the transmission increases [P5]. This section provides further insight into the Space–Time Adaptive Retransmission (STAR) concepts suggested in [1, 28, 173] and [P5]. Interestingly, a related concept is currently being proposed to IEEE 802.16e (WiMax) specification [174].

A simple example of STAR is given by a solution that converts a symbol rate two transmission with diversity order one (spatial multiplexing or BLAST) into a symbol rate one transmission with diversity order four. Assuming two transmit antenna and two receive antennas the transmission assumes the form

$$\begin{bmatrix} x_1 & x_3 & \cdots & x_{2N-1} & \text{control delay} & -x_{2N}^* & \cdots & -x_4^* & -x_2^* \\ x_2 & x_4 & \cdots & x_{2N} & \text{control delay} & x_{2N-1}^* & \cdots & x_3^* & x_1^* \end{bmatrix}^T, \tag{4.63}$$

where “control delay” and all that follows are due to retransmission, in the event that initial transmission has failed. The part before the “control delay” is identical to BLAST, and when combining the transmission over the whole block the matrix collapses to the time-reversal space–time block code of [175].

In another example, we may convert symbol rate two transmission with diversity order two into symbol rate one transmission with diversity order four (for arbitrary input alphabets). In this case, the initial full transmission matrix is (neglecting “control delay”)

$$\mathbf{X} = \begin{bmatrix} \mathbf{X}_A & \mathbf{X}_B \\ \mathbf{X}_B & \mathbf{X}_A \\ \mathbf{X}_A & -\mathbf{X}_B \\ \mathbf{X}_B & -\mathbf{X}_A \end{bmatrix}, \tag{4.64}$$

with $\mathbf{X}_A$ and $\mathbf{X}_B$ denoting two orthogonal space–time block codes. The first transmission sends out the first two rows, first possible retransmission (if needed) the next two rows, and finally (if needed) the final four rows. Thus, we observe that DSTTD is converted to ABBA, then to a fully orthogonal full-diversity delay sub-optimal rate 1/4. transmission. Additional examples are provided in [1], and a number of variations of these may be devised depending on the application. For example, the precoding parameters may be changed for the retransmissions, while keeping all else intact, to get identical solutions in certain cases.

5

Conclusion

This thesis contributes to the modulation aspect in exploiting MIMO and MISO channels. The modular and linear MIMO and MISO modulation methods that are proposed in the publications [P1]-[P8] enable high data rates even for those who do not have abundant bandwidth. On the other hand, if data rate gain (or spectral efficiency) is not in the designer's agenda, complex modulation orthogonal space–time codes may be used to increase diversity and power efficiency, see [P3]. In general, the power efficiency and high coding gain of the developed schemes motivates their potential e.g. use in mean or peak power limited systems. Indeed, full-diversity symbol rate two transmission methods developed in [P8] are currently potential candidates for adoption in 3G and 4G wireless systems.

MIMO modulation methods may not be easily attached to large-scale networks due to increased implementation complexity. Also, it may be difficult to guarantee sufficiently independent fading across antenna elements with reasonably sized transceiver units. In correlated channels the effective channel rank is reduced and the benefits of multi-stream transmission are diluted. In part for this reason, to avoid the use of very large arrays, the thesis has focused in MIMO transmission using, say, 2-8 transmit antennas, and 1-4 receive antennas. In the future, if the methods are developed for systems operating at much higher carrier frequencies than 2 GHz, this restriction on the number of antenna elements may altered. Then, true MIMO modulators with spectral efficiencies even beyond 10 bps/Hz may be practical. The peak spectral efficiency tested in this thesis is 8 bps/Hz. Even this would be a large leap for 3G cellular systems.

Despite the theoretical promise enjoyed by linear space–time modulation methods, several practical issues need to be addressed before they reach full maturity. The high symbol rate modulation methods need to be optimized for frequency-selective multiuser channels. Although some steps in that direction were taken in [P8], a complete theory for frequency-selective channels is still called for. On the other hand, even if the modulation methods and concepts developed in [P1]-[P8] are targeted for flat fading channels, they could be used with little or no change in frequency-selective channels, if combined with an appropriate block modulation concept, such as OFDM, see [P8].

It is apparent that the complexity of the terminals and network elements will increase for all MIMO solutions considered herein and elsewhere. The number of signal states in a MIMO receiver grows exponentially in the number of interfering symbols and efficient detection algorithms need to be devised. Nevertheless, Moore's Law suggests that at some point in the future multi-antenna techniques may become standard technology. Should this be the case, the considered modulation methods are likely to be introduced to wireless systems that have alternative capacity-enhancing solutions, such as ARQ, multiuser scheduling, and beamforming already in place. Thus, the combined use of multiuser diversity via scheduling and MIMO modulation remains an important topic, as well their combination with ARQ protocols. Publications [P6] and [P7] address this research area by suggesting scheduling criteria (channel quality indicators) for matrix-modulated systems, and publication [P5] proposes a new ARQ solution for use with linear space–time modulation. Thus, even though a modular approach is called for, other aspects of a wireless system may need to be partially re-engineered upon the introduction of efficient modulation methods.

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