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Space-Time Adaptation and MIMO Standardization Status

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1. Introduction

The use of multiple antennas at the transmitter and/or receiver sides has been used for years in order to increase the signal to noise ratio at the receiver or for beam steering (a.k.a. beamforming) to reduce the amounts of interference at the receiver. However, one of the main benefits of using a Multiple Input Multiple Output (MIMO) channel is the increase of the channel capacity. Nowadays, by using different space-time-frequency coding techniques, orthogonal (or quasi orthogonal) virtual paths between transmitter and receiver can be obtained. These virtual paths can be used in order to increase the spectral efficiency or in order to increase the signal diversity (i.e. space diversity). In fact, a fundamental trade-off between diversity and multiplexing capabilities exists and must be considered when designing a multiple antenna system. In this chapter, the following issues are addressed:

- We start by describing and reviewing the (ergodic) capacity of the MIMO channel in case of perfect channel state information at the transmitter (CSIT). Afterwards, the outage capacity is studied leading to the diversity-multiplexing trade-off.
- Then, we analyze different spatial adaptation and precoding mechanisms that can be applied to increase the performance of the system (either in terms of throughput or robustness). A new spatial adaptation algorithm proposed by the authors and called *Transmit Antenna and space-time Coding Selection* (TACS) is described showing some performance results that illustrate the improve on performance and/or throughput.
- Finally, the well-known Space-Time coding techniques are reviewed, and a summary of the MIMO techniques adopted in WiMAX2 (IEEE 802.16e/m) is provided.

The present analysis is done following the general Linear Dispersion Codes framework, which is of special interest since it allows describing in an elegant way most of the space-time block codes existing in the literature.

2. Characteristics of the MIMO channel

Before going into the details of how MIMO transmission can be carried out, it is important to have a look to the capacity of the MIMO channel. Usually, for capacity evaluation of

MIMO channels it is assumed that the fading coefficients between antenna pairs are i.i.d. Rayleigh distributed. In addition, and without loss of generality, it is also assumed that the channel is constant during the transmission of one MIMO codeword. Under this assumption the channel is referred as a block fading channel.

As shown in [1], the capacity of the MIMO channel can be obtained as follows

$$C(\rho, \mathbf{H}) = \arg \max_{\mathbf{Q} > 0: \text{Tr}\{\mathbf{Q}\} = M} \log_2 \det \left[\mathbf{I}_N + \frac{\rho}{M} \mathbf{H} \mathbf{Q} \mathbf{H}^H \right] \quad (1)$$

where \mathbf{H} is the MIMO channel matrix, M the number of transmit antennas, I_N an identity matrix of size N equal to the number of receive antennas, ρ is the Signal to Noise Ratio (SNR) and \mathbf{Q} is the input covariance matrix whose trace is normalized to be equal to the number of transmit antennas. To gain a further insight on the channel characteristics, a good method is to apply the *Singular Value Decomposition* (SVD) to the MIMO channel matrix, so we express the channel matrix as

$$\mathbf{H} = \mathbf{U} \mathbf{\Sigma} \mathbf{V}^H, \quad (2)$$

where $\mathbf{\Sigma}$ is a diagonal matrix, whose entries are the eigenvalues of \mathbf{H} , and \mathbf{U} and \mathbf{V} are the lower and upper diagonal matrices respectively. The first important characteristic of the MIMO channel is given by the number of eigenvalues which tells us about the number of the independent virtual channels between transmitter and receiver. In addition, using the SVD of channel matrix, we can rewrite the channel capacity as

$$C(\rho, \mathbf{H}) = \arg \max_{\{p_k\}_{k=1}^n} \sum_{k=1}^n \log_2 (1 + \rho p_k \sigma_k^2) \quad (3)$$

where σ_k and p_k are the eigenvalues and the power transmitted through each of the said virtual channels respectively.

In case the channel is known at the transmitter, one can use this information to maximize the channel capacity by applying what it is known as multiple eigenmode transmission. This is much less complex than it sounds, and is carried out by multiplying the input vector \mathbf{x} by \mathbf{V}^H . At the receiver side, similar operation is needed, therefore the result of multiplying the received signal by \mathbf{U}^H is shown in the following equation

$$\mathbf{y} = \mathbf{U} \mathbf{\Sigma} \mathbf{V}^H \mathbf{V} \mathbf{x} + \mathbf{n} = \mathbf{U} \mathbf{\Sigma} \mathbf{x} + \mathbf{n} \Rightarrow \hat{\mathbf{x}} = \mathbf{U}^H \mathbf{U} \mathbf{\Sigma} \mathbf{x} + \mathbf{U}^H \mathbf{n} = \mathbf{\Sigma} \mathbf{x} + \tilde{\mathbf{n}}, \quad (4)$$

where it is observed that the number of eigenvalues determines the number of independent complex symbols that can be transmitted per each MIMO codeword. It can be also concluded that at low SNR values, the optimum allocation strategy will be to allocate all the available power to the strongest (or dominant) eigenmode, whereas at high SNRs the maximum capacity is obtained by allocating the same power to all the non-zero eigenmodes [1]. Actually, it is also proved that uniform power allocation (UPA) is the optimum strategy for fast fading channels where the transmitter is not able to capture the instantaneous channel state. The (ergodic) channel capacity in case of UPA allocation is given by

$$C_{UPA}(\rho, \mathbf{H}) = E \left\{ \sum_{k=1}^n \log_2 \left(1 + \frac{\rho}{M} \sigma_k^2 \right) \right\} = E \left\{ \log_2 \left(1 + \frac{\rho}{M} \|\mathbf{H}\|_F^2 \right) \right\}. \quad (5)$$

where $\|\mathbf{H}\|_F^2$ stands for the Froebenius norm. Similar research studies have been undertaken regarding the effects of antenna correlation on the channel capacity. It is shown in [2] that for low antenna correlation values the optimum strategy is to allocate the same power to all the eigenmodes, whereas for high correlation values the optimum strategy is allocating all the power to the strongest eigenmode.

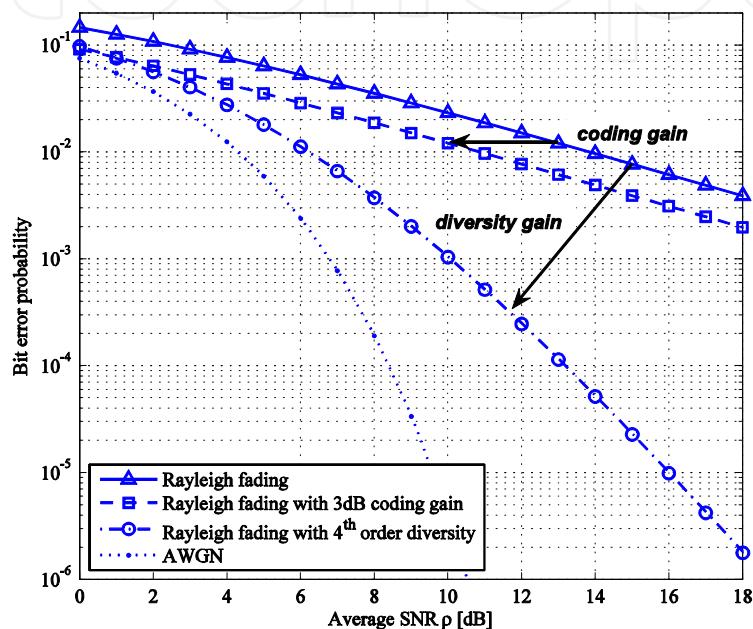


Fig. 1. Array and diversity gain in Rayleigh fading channels.

2.1 The diversity-multiplexing trade-off

When dealing with frequency/time-variant channels, one intrinsic characteristic of the channel is the diversity that can be achieved. In case of single input single output transmission, the general approach is to use coding and interleaving in the frequency and time domains so that one codeword is spread over the highest number possible of channel states. However, frequency and time diversity incur in a loss in bandwidth and/or transmission time delay. Alternatively, in case of multiple input multiple output channels the spatial dimension can be also exploited in order to increase the diversity without neither losing bandwidth nor increasing the transmission delay. Some metrics are defined to characterize the diversity. First, the *diversity gain* (or diversity order) is linked with the number of independent fading branches. Formally, it is defined as the negative asymptotic slope (i.e. for $\rho \rightarrow \infty$) of the *log-log* plot of the average error probability P versus the average SNR ρ

$$g_d = - \frac{\log(P)}{\log(\rho)}. \quad (6)$$

Finally, the *coding gain* is defined as the SNR gain (observed as a left shift of the error curve). Then the coding gain g_c is analytically expressed as

$$P_e = \left(\frac{c}{g_c \rho} \right)^{\alpha g_a}, \quad (7)$$

where P_e is error probability, and α and c are scaling constants depending on the modulation level, the coding scheme, and the channel characteristics. The *array gain* g_a represents the decrease of average SNR due to coherent combining (*beamforming*) in case of multiple antennas at both transmitter or receiver sides, and it is formally expressed as

$$g_a = \frac{\rho_{ma}}{\rho_{sa}} \quad (8)$$

where ρ_{sa} is the average SNR for the SISO link, and ρ_{ma} is the average SNR for the MIMO link. The three different concepts are illustrated in Fig. 1 which shows the bit error rate of a QPSK transmission having an AWGN channel and an uncorrelated Rayleigh channel.

In case of multiple antennas at both sides of the link, multiple independent channels exist according to the rank of the channel matrix [1]. The multiplexing gain g_s is defined as the ratio of the transmission rate $R(\rho)$ to the capacity of an AWGN with array gain g_a

$$g_s = \frac{R(\rho)}{\log_2(1 + g_a \rho)} \quad \Rightarrow \quad R(\rho) = g_s \log_2(g_a \rho) \quad (9)$$

where $R(\rho)$ is the transmission rate.

For a slow fading channel (i.e. block fading channel), the maximum achievable rate for each codeword is a time variant quantity that depends on the instantaneous channel realizations. In this case, the outage probability P_{out} metric is preferred and is defined as the probability that a given channel realization cannot support a given rate R

$$P_{out}(R) = \inf_{\mathbf{Q} > 0: \text{Tr}\{\mathbf{Q}\} \leq M} P \left(\log_2 \det \left(\mathbf{I}_N + \frac{\rho}{M} \mathbf{H} \mathbf{Q} \mathbf{H}^H \right) < R \right). \quad (10)$$

where *inf* stands for the \mathbf{Q} that achieves the lower bound in terms of outage probability. Then, we can gain insights into the channel behaviour by analysing the outage probability as a function of the SNR and ρ for a given transmission rate. Actually, we can establish a relationship between the diversity gain and the multiplexing gain via the outage probability P_{out} as

$$g_d(g_s, \rho) = - \frac{\partial \log(P_{out}(R))}{\partial \log(\rho)} \quad \Rightarrow \quad P_{out}(\rho) = \rho^{-g_d}. \quad (11)$$

Both the multiplexing gain and the diversity gain are upper bounded by $g_s \leq \min(N, M)$ and $g_d \leq N \times M$. Intuitively, the multiplexing gain indicates the increase of the transmission rate as

a function of the SNR, whereas the diversity gain gives us an idea on how fast the outage probability decreases with the SNR.

The analysis of Eq. (11) at high SNR and uncorrelated Rayleigh channel leads to the diversity-multiplexing trade-off of the channel [3]. It has been shown that $g_d(g_s, \infty)$ is a piecewise linear function joining the $(g_s, g_d(g_s, \infty))$ points with $g_s = \{0, \dots, \min(N, M)\}$ and $g_d = (N - g_s) \times (M - g_s)$. This trade-off is illustrated in the following Fig. 2. It is observed that maximum diversity is achieved when there is no spatial multiplexing gain (i.e. the transmission rate is fixed), whereas the maximum spatial multiplexing gain is achieved when the diversity gain is zero (the outage probability is kept fixed).

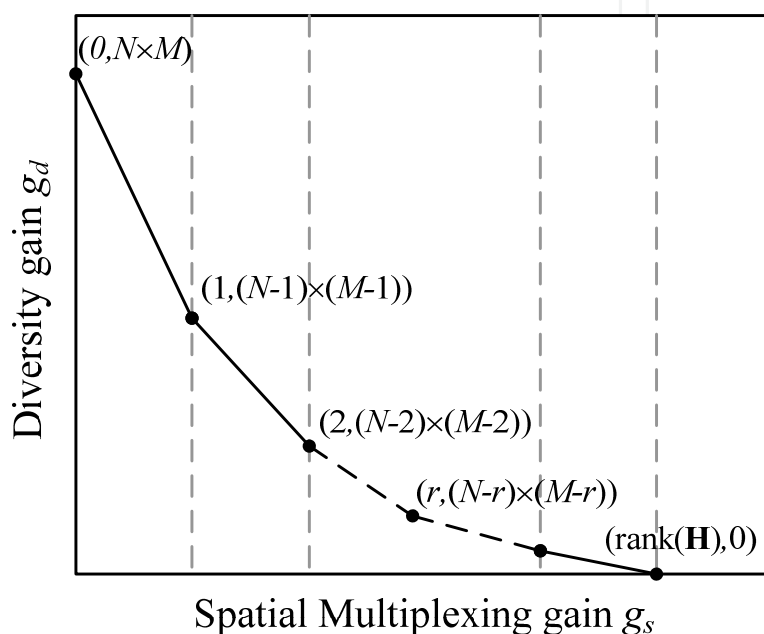


Fig. 2. Asymptotic diversity-multiplexing trade-off in uncorrelated Rayleigh channels.

2.2 Space-time coding over MIMO channels

Based on the above introduction on MIMO channel characteristics, and the very important principle of spatial diversity versus spatial multiplexing tradeoff, we could now start studying the Space-Time MIMO encoding techniques. Analogously to channel coding in SISO links, two types of channel coding have been used for MIMO channels: block coding (referred as Space Time/Frequency Block Coding, STBC/SFBC) and convolutional coding (referred as Space Time Trellis Coding-STTC) [4][5]. For the STBC case, the codeword is only a function of the input bits, whereas the encoder output for the STTC is a function of the input bits and the encoder state. The inherent memory of the STTC provides an additional coding gain compared to the STBC at the expense of higher computational complexity [7][8]. However, since STBC transforms the MIMO channel into an equivalent scalar AWGN channel [6], the concatenation of traditional channel coding with STBC shows good performance and even outperforms STTC for low number of receive antennas ($M, N \leq 2$) [7] and the same number of encoder states. Furthermore, STBC/SFBC are of significantly less complexity than STTC and for this reason they are usually preferred.

2.2.1 Space-time block coding system model

In this section an example of a MIMO system with M , N transmitter and receiver antennas respectively communicating over a frequency flat-fading channel is assumed. A codeword \mathbf{X} is transmitted over T channel accesses (symbols) over the M transmitting antennas, hence $\mathbf{X}=[\mathbf{x}_0 \dots \mathbf{x}_{T-1}]$ with $\mathbf{x}_i \in \mathbb{C}^{M \times 1}$. All the codewords are contained inside a codebook \mathcal{X} , and each codeword contains the information from Q complex symbols. The ratio of symbols transmitted per codeword is defined as the spatial multiplexing rate $r_s=Q/T$, where in case of $r_s=M$ the code is referred as *full-rate*. The transmission rate is given by $R=Q \times n_b/T$ [bits/s/Hz] where n_b is the number of bits transmitted by each $\mathbf{x}_i(j)$ complex symbol. Moreover, the spreading of the symbols in the time and spatial domains leads to the increase of diversity, whereas by modifying Q we can modify the spatial multiplexing gain. In consequence, $M \times N \times T$ determines the maximum diversity order, while Q defines the spatial multiplexing rate [10]. During any i^{th} time instant (equivalent to a channel access), the transmitted and received signals are related as

$$\mathbf{y}_i = \sqrt{\frac{\rho}{M}} \mathbf{H}_i \mathbf{x}_i + \mathbf{n}_i, \quad i = \{0, \dots, T-1\}. \quad (12)$$

In Eq. (12) we have assumed the channel is constant within each channel Access. However, we can go one a step further and assume that the channel is constant during the transmission of one whole codeword transmission ($T_{\text{coh}} \gg T$) time, and in this case the channel dependency on the i^{th} time instant subindex can be dropped such that $\mathbf{H}_i = \mathbf{H}$ for any $i = \{0, \dots, T-1\}$. Under these conditions the quasi-static block fading channel model can be assumed, and we can rewrite Eq. (12) as follows

$$\mathbf{Y}^T = \sqrt{\frac{\rho}{M}} \mathbf{H} \mathbf{X}^T + \boldsymbol{\psi}^T \Rightarrow \mathbf{Y} = \sqrt{\frac{\rho}{M}} \mathbf{X} \mathbf{H}^T + \boldsymbol{\psi} \quad (13)$$

where $\mathbf{X} \in \mathbb{C}^{T \times M}$ means the space-time transmitted codeword, $\mathbf{Y} \in \mathbb{C}^{T \times N}$ means the received space-time samples and $\mathbf{V} \in \mathbb{C}^{T \times N}$ represents the noise over each receive antenna during each channel access.

2.2.2 Linear Dispersion Codes

The Linear Dispersion Codes (LDC) class belongs to a subclass of the STBC codes where the codeword is given by a linear function of the input data symbols [11][20][21]. When the codeword is a linear function of the data symbols, the transmitted codeword can be expressed as

$$\mathbf{X} = \sum_{q=1}^Q (\alpha_q \mathbf{A}_q + j \beta_q \mathbf{B}_q), \quad (14)$$

where $\mathbf{A}_q \in \mathbb{C}^{T \times M}$ determines how the real part of the symbol s_q , a_q , is spread over the space-time domain, and the same for the imaginary part β_q which is spread according to $\mathbf{B}_q \in \mathbb{C}^{T \times M}$. For power normalization purposes, it is considered that the transmitted complex symbols s_q have zero mean and unitary energy, this is $E\{s_q^* s_q\} = 1$. The matrices \mathbf{A}_q and \mathbf{B}_q are referred as the basis matrices and usually are normalized such that

$$\sum_{q=0}^{Q-1} \left(\text{Tr} \{ \mathbf{A}_q \mathbf{A}_q^H \} + \text{Tr} \{ \mathbf{B}_q \mathbf{B}_q^H \} \right) = MT. \quad (15)$$

If we impose some conditions on the set of basis matrices $\mathbf{A}_q, \mathbf{B}_q$ with $q=\{0, \dots, Q-1\}$ the mapping between the input symbols and the transmitted codeword \mathbf{X} is unique and the symbols can be perfectly recovered. Substituting Eq. (15) into Eq. (13) and applying the *vec* operator on both sides of the expression, an equivalent real valued system equation can be written as

$$\mathbf{y} = \begin{bmatrix} \Re(\mathbf{y}_0) \\ \Im(\mathbf{y}_0) \\ \vdots \\ \Re(\mathbf{y}_{Q-1}) \\ \Im(\mathbf{y}_{Q-1}) \end{bmatrix} = \sqrt{\frac{\rho}{M}} \mathbf{H} \underbrace{\begin{bmatrix} \alpha_0 \\ \beta_0 \\ \vdots \\ \alpha_{Q-1} \\ \beta_{Q-1} \end{bmatrix}}_{\hat{=}\mathbf{s}} + \underbrace{\begin{bmatrix} \mathbf{n}_0 \\ \mathbf{n}_0 \\ \vdots \\ \mathbf{n}_{Q-1} \\ \mathbf{n}_{Q-1} \end{bmatrix}}_{\hat{=}\mathbf{n}} = \sqrt{\frac{\rho}{M}} \mathcal{H} \mathbf{s} + \mathbf{n}, \quad (16)$$

where $\mathbf{n} \in \mathbb{N}(0, 1/2)$ is the real vector noise with i.i.d. components. The equivalent real valued channel matrix \mathbf{H} is obtained as

$$\mathcal{H} = \begin{bmatrix} \mathcal{A}_0 \hat{h}_0 & \mathcal{B}_0 \hat{h}_0 & \cdots & \mathcal{A}_{Q-1} \hat{h}_0 & \mathcal{B}_{Q-1} \hat{h}_0 \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ \mathcal{A}_0 \hat{h}_{N-1} & \mathcal{B}_0 \hat{h}_{N-1} & \cdots & \mathcal{A}_{Q-1} \hat{h}_{N-1} & \mathcal{B}_{Q-1} \hat{h}_{N-1} \end{bmatrix}_{2NT \times 2Q}. \quad (17)$$

with:

$$\mathcal{A}_q = \begin{bmatrix} \Re\{\mathbf{A}_q\} & -\Im\{\mathbf{A}_q\} \\ \Im\{\mathbf{A}_q\} & \Re\{\mathbf{A}_q\} \end{bmatrix}, \mathcal{B}_q = \begin{bmatrix} -\Im\{\mathbf{B}_q\} & -\Re\{\mathbf{B}_q\} \\ \Re\{\mathbf{B}_q\} & -\Im\{\mathbf{B}_q\} \end{bmatrix} \in \mathbb{R}^{2T \times 2M}; \hat{h}_n = \begin{bmatrix} \Re\{\mathbf{h}_n\} \\ \Im\{\mathbf{h}_n\} \end{bmatrix} \in \mathbb{R}^{2M \times 1}, \quad (18)$$

where \mathbf{h}_n is the n -th row of the MIMO channel matrix \mathbf{H} . Theoretically the maximum number of possible independent streams or channel modes of the effective MIMO channel matrix \mathbf{H} is $2NT \times 2Q$ (maximum number of singular values different than zero). However, following the LDC design, the number of scalar modes that are excited is equal to the rank of $\mathbf{H}^T \mathbf{H}$, which means equal to $2Q$ in the best case [18]. In addition, it is observed that it is possible to use a linear receiver only if $Q \leq NT$, otherwise the system would be undetermined. When $Q < NT$ the system is over-determined, and in consequence more reliability is given to each estimated symbol (i.e. spatial diversity is increased at the expense of spatial multiplexing).

3. Detection techniques: Linear vs non-linear schemes

Recovering the transmitted symbols within each codeword might become a challenging task depending on the set of basis matrices used. Moreover, we have already observed that the

ergodic channel is also a function of the basis matrices set. Therefore, a compromise between complexity and performance (in terms of spectral efficiency or decoding errors) exists which motivates the implementation and use of different LDC codes and decoding schemes.

3.1 Maximum Likelihood (ML) decoding

Optimum signal detection requires the maximization of the likelihood function over the discrete set of the code alphabet [13]. Mathematically, this can be expressed as

$$\hat{s} = \arg \min_{\hat{s} \in \Sigma} \left\| y - \sqrt{\frac{\rho}{M}} \mathbf{H} \hat{s} \right\|_F^2 \quad (19)$$

where Σ is the space of all the transmitted symbol vectors with all the input data combinations having the same likelihood. Regarding the computational complexity of the ML detector, since each vector has a set of $2Q$ symbols each one mapped over $\log_2(Z)$ bits, the computational complexity is exponential with $Q \times \log_2(Z)$.

The theoretical framework to understand the behavior of a MIMO system using an ML receiver has been extensively studied and analyzed in the scientific literature (e.g. [9]-[13]) leading to important conclusions. The first and more obvious conclusion is that the diversity order is $g_{d \leq N} \min(M, T)$ [13]. So, it becomes clear that T should be equal to M to achieve full diversity. However, increasing T requires increasing also the value of Q to maintain the same rate R , which leads to an increase in memory requirements, computational complexity, and delay. It is then apparent that a trade-off exists between achievable diversity and the decoding complexity.

Often the performance of a system is measured in terms of the post-processing SNR or the Effective SNR (ESINR). This ESINR value estimates the SNR required in an AWGN channel to obtain the same performance as in the given system (i.e. our MIMO system). In [14], the author proposed a simple parametrizable expression to estimate the performance of the system under different MIMO transmission schemes and antenna configurations. This model has been used in later sections when the performance evaluation of adaptive MIMO systems with ML receivers is developed and analysed.

3.2 Linear detectors: Zero forcing and minimum mean square error

The high computational cost of the ML receiver ($O(2^{Q \log_2(Z)})$) makes the use of less computational demanding receiving techniques more appealing, sometimes even despite a degradation on the system performances. Following the expression in Eq. (12), a linear relationship between input and output symbols exists and the system can be solved applying simple algebra as long as $Q \leq NT$, i.e.

$$\hat{\mathbf{s}} = \mathbf{G} \mathbf{y} = \sqrt{\frac{\rho}{M}} \mathbf{G} \mathbf{H} \mathbf{s} + \mathbf{G} \mathbf{n}, \quad (20)$$

where $\mathbf{G} \in \mathbb{R}^{2Q \times 2N}$ is the equalizer matrix which compensates the MIMO channel effects. Similar to frequency equalization, the equalizer matrix might be designed to suppress the

inter-symbol interference (despite the noise vector might be increased due to the equalization) or to minimize the mean square error (i.e. the MMSE). The first design criterion is known as the Zero-Forcing equalization where

$$\mathbf{G}_{ZF} = \sqrt{\frac{M}{\rho}} \mathbf{H}^\dagger. \quad (21)$$

On the other hand, according to the MMSE criterion the following equalizer is obtained

$$\mathbf{G}_{MMSE} = \sqrt{\frac{M}{\rho}} \left(\mathbf{H}^H \mathbf{H} + \left(\frac{\rho}{M} \right)^{-1} \mathbf{I}_Q \right)^{-1} \mathbf{H}^H. \quad (22)$$

where \mathbf{I}_Q is an identity matrix of size Q . Besides the lower computational cost of the linear receivers, another advantage of using them is that the channel effects can be perfectly estimated on a symbol basis, hence a closed expression for the *ESINR* per each transmitted symbol can be obtained as

$$ESINR_q = \frac{\text{diag}[\mathbf{D}\mathbf{D}^H]_q}{\text{diag}\left[\frac{1}{\rho} \mathbf{G}^H \mathbf{G} + \mathbf{I}_{self} \mathbf{I}_{self}^H\right]_q}. \quad (23)$$

where $\mathbf{D} = \text{diag}[\mathbf{G}\mathbf{H}]$, and $\mathbf{I}_{self} = \mathbf{G}\mathbf{H} - \mathbf{D}$ is the self-interference term. The full expression of ML and ZF receivers can be found in [14].

4. Exploiting the transmit channel knowledge

It has been already stated in previous sections that in case the transmitter has perfect channel information knowledge, the SNR at the receiver is maximized if the transmitter applies all the power over the dominant eigenmode of the channel. Moreover, in order to increase the throughput it may be preferable to transmit over all the non-zero eigenmodes of the channel allocating to each mode a power obtained following the water-filling algorithm [10]. However, both (dominant and multiple eigenmode transmission) beamforming techniques require that the transmitter knows perfectly the channel state information (CSI), and also that the channel doesn't change during a sufficiently large period to allow the CSI estimation and application of the beamforming. In consequence, the beamforming might be only applied for low mobility scenarios and where the channel can be accurately estimated.

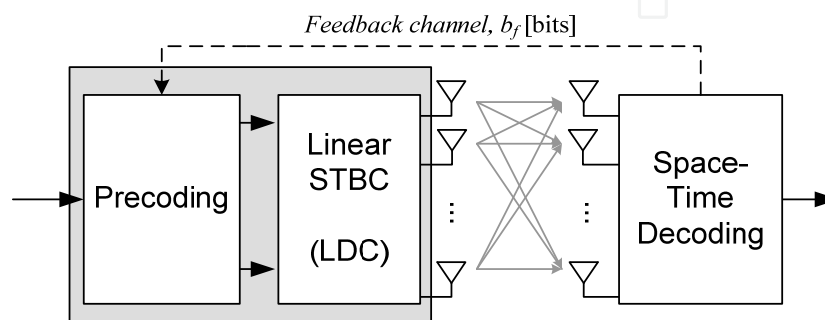


Fig. 3. Linear space-time precoding

In order to obtain the channel information at the transmitter, the most common approach is that the receiver sends some signalling to allow the transmitter to know the status of the downlink channel (in case of FDD system this has to be done explicitly by transmitting the matrix \mathbf{H} , and for TDD systems the channel reciprocity allows sending some pilots in the reverse link so that the channel is estimated for the forward channel). However, any of these two alternatives will consume bandwidth either in the form of feedback signalling or channel estimation signalling. This triggered lot of work on how to reduce the feedback leading to techniques and metrics such as quantizing the channel information, using the channel condition number, the Demmel condition number, the channel rank, etc [26]. In general, the schemes where the input symbols are adjusted according to the channel status are known as precoders. Actually, precoding and space-time block coding can be considered into the same block where given a set of codes (defined by the codebook) one of them is selected each *Time Transmission Interval* (defined by T or multiples of) according to the b_f feedback bits.

4.1 Transmit and receive antenna selection

Besides the increase of the capacity or the reliability by any of the before mentioned precoding techniques (beamforming, codeword selection based on finite codebooks, etc.), a very simple precoding technique is to select which antenna (or subset of antennas) should be used according to an optimization criterion (e.g. capacity, reliability, etc.). Antenna selection also aids to reduce the hardware cost as well as the signal processing requirements, therefore, it may be good for handheld receivers where, space, power consumption, and cost must be seriously taken into account. Obviously, the reduction of the number of antennas reduces the array gain, however when the channel in any of these antennas is experiencing a deep fade, the capacity loss by not using such antenna is negligible [19]. In consequence, antenna selection at both transmitter and receiver helps in reducing the implementation costs while retaining most of the benefits of MIMO technology.

A MIMO system model considering antenna selection is depicted in Fig. 4, where M and N are the number of transmitter and receiver RF chains respectively, whereas the available antennas are referred by M_a and N_a for transmitter and receiver respectively ($M \leq M_a, N \leq N_a$).

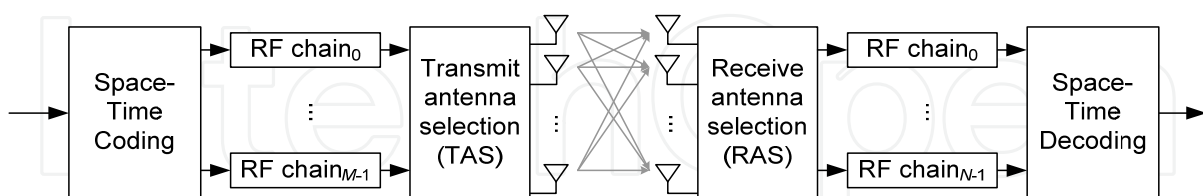


Fig. 4. Antenna selection in MIMO systems with M_a available transmit and N_a receive antennas.

In the SIMO case, it is shown in [10] and [19] that the array gain using a Maximum Ratio Combiner (MRC) without Receive Antenna Selection (RAS) is equal to $g_a = N_a$, whereas when RAS is applied (e.g. the antenna with better channel is selected) the array gain is given by

$$g_a = N \left(1 + \sum_{j=N+1}^{N_a} \frac{1}{j} \right). \quad (24)$$

As a result, we can note that RAS implies a loss in the SNR which becomes larger as the difference between N and N_a is increased. However, the diversity order for both schemes is exactly the same and there is only a coding gain difference [19]. The analysis for Transmit Antenna Selection (TAS) is reciprocal, therefore, the same effects are observed in case of MISO with TAS.

For the MIMO case and if multiple streams are simultaneously transmitted, transmit and/or receive antenna selection has further implications than just a reduction of the array gain. Actually, the inherent spatial multiplexing-diversity trade-off leads to different optimization criteria: diversity optimization (i.e. select the set of antennas that gives a higher Frobenius norm of the channel), improve the link reliability (discard antennas that produces large fadings in any eigenmode), maximize the Shannon capacity, etc. Furthermore, in [19] it is stated that the diversity gain obtained by transmit antenna and receive antenna selection is the same as without selection procedure, hence $g_d=(N-g_s)\times(M-g_s)$ with $g_s=\{0,\dots, \min(N,M)\}$.

4.2 Transmit antenna selection in MIMO systems

Transmit antenna selection techniques were first proposed during the very late 90s in the context of MIMO links in order to improve the array gain. During that period, antenna selection was derived according to the class of ST coding scheme that was involved. Heath et al. in [22] focused on the antenna selection in case of Spatial Multiplexing for linear receivers. The optimization criterion in [22] was to maximize the post-processing SNR in order to minimize the bit error probability. Later, it has been shown that the difference between optimizing the $ESINR$ is only 0.5dB better than optimizing the lowest eigenvalue (the $ESINR$ in case of Zero-Forcing is lower bounded by the minimum eigenvalue of H). Similar works have been carried out for Orthogonal STBC (OSTBC) which in this case concluded that maximizing the Frobenius norm of the active channel was the optimum strategy [23]. More recently, Deng et al. extended these transmit antenna selection schemes under the LDC framework concluding that the best selection criteria for minimize the bit error probability is based on maximizing the post-processing (or Effective) SNR [24]. Finally, an interesting application of transmit antenna selection has been proposed by Freitas et al. in [25] where different spatial layers are assumed combining spatial diversity and spatial multiplexing. In [25] the different branches are disposed in parallel hence both spatial diversity and multiplexing gains can be simultaneously achieved. The antennas subsets are then assigned to the spatial layers in order to minimize the bit error probability, where the (more susceptible) SM based layers are assigned the best subset of antennas and the remaining are assigned to the OSTBC layers.

4.3 MIMO precoding based on LDC codes

In the previous section, TAS precoding scheme has been introduced for some of the existing STBC and LDC. Therefore, given a specific code, the number of bits f_b that must be fed-back from the receiver to indicate the optimum transmit antennas set is

$$f_b = \binom{M_a}{M} = \frac{M_a!}{M!(M_a - M)!} \quad (25)$$

However, if we could afford sending few more bits over the feedback channel, the transmitter/receiver may be able to select which code is more suitable according to the current channel state, or to choose how many spatial streams can be transmitted according to the channel rank [26]. Recent researches have extended the space-time coding selection (i.e. codebook based precoding) into the LDC framework [27]-[32]. An important result was obtained in [29], where it is shown that $f_b = \log_2(M)$ feedback bits are enough to achieve full diversity. In [32], the authors showed that the average SNR can be improved up to 2dB compared to the open loop scheme with only 3 feedback bits (i.e. 8 sets of LDC codes).

4.4 Transmit antenna and space-time Code Selection (TACS)

As it has been explained in the previous section, when partial CSIs information is available at the transmitter two common selection techniques could be applied, which are: the space-time code selection, and the transmit antenna selection. One of the first works joining both concepts is that presented by Heath et al. in [33] where the number of the spatial streams (in the SM case) are adapted by selecting the best set of transmitter antennas (i.e. $f_b = M$). Furthermore, it was stated that if the optimum number of streams are transmitted from the optimum selected antenna set, the diversity gain is also maximized ($g_d \leq MN$). Then, given an antenna subset and a fixed rate, the required constellation could be determined as well as the number of spatial streams.

A simplification of this optimization problem is given in [34] where each stream is switched on/off when the post-processing of the SNR value of the stream is above/below a fixed threshold which is related with the rate. Further extensions of space-time code selection with TAS are given by Machado et al. in [35] where the available codes in the codebook are; the Alamouti code, the SM with $M=2$, the Quasi-OSTBC with $M=3$ and single antenna transmission.

In addition, the space-time code selection with transmit antenna selection has been generalized by the authors in [36]-[39] under the LDC framework considering both the linear and the ML receivers and developed within the IEEE 802.16m framework [38]. This generalization allows us to use any type of linear STBC (independently of the optimization criteria) codes and determine which codes are used most of the time and under which channel conditions. Two optimizations criteria have been developed in [14], one following the classical bit error rate optimization (minimizing of the scaled minimum Euclidian distance), and a second one is based on the throughput maximization given a fixed link quality (i.e. fixed packet error rate or bit error rate). This second optimization criterion can be used for resource allocation and scheduling purposes. Nevertheless, it is also shown that for low multiplexing rates the classic STBC codes (i.e. Alamouti, SM and Golden code) with transmit antenna selection are sufficient to explore the Grassmanian subspace [14].

4.4.1 The TACS selection criteria

Given the $ESINR$ per stream and the average pairwise error probabilities, two different code and antenna subset optimization scenarios namely *Minimizing the bit error rate* and *Maximizing the throughput* respectively have been evaluated in [14][36]-[39]. In the first scenario, we consider that the same modulation is applied to all the symbols with a fixed rate R . In that case, and since transmission power is also fixed, we are interested in selecting

the transmit antenna subset and the LDC code that minimizes the error rate probability (i.e. the bit error rate - BER) while the modulation that is required by each LDC is adapted in order to achieve the required rate R . In that case, since the Q -function is monotonically decreasing as a function of the input, the optimization problem is defined as follows

$$\max_{LDC_i, p_i} \min_q \left\{ ESINR_q(H, LDC_i, p_i) d_{\min}^2(Z_i) \right\}, \quad (26)$$

where i means the LDC index, p_i denotes the transmitting antenna subset, q refers to the spatial stream (i.e. the symbol) index, and d_{\min} is the minimum Euclidean distance according to the QAM constellation size used (note that the QAM constellation is a function of the LDC).

In the second scenario, the optimization is performed in order to maximize the system throughput considering a certain quality of service requirement (i.e. a maximum Block Error Rate - BLER). In that case, the problem is formulated as follows

$$\max_{LDC_i, p_i, MCS_j} \min_q R \left(1 - BLER(ESINR_q) \right) \quad \text{s.t.: } BLER \leq \mu \quad (27)$$

where j means the Modulation and Coding Scheme (MCS) index that maximizes the spectral efficiency for the specific channel state subject to a maximum Block Error Rate (BLER). The following Fig. 5 illustrates the scheme of the MIMO system applying the TACS selection algorithm.

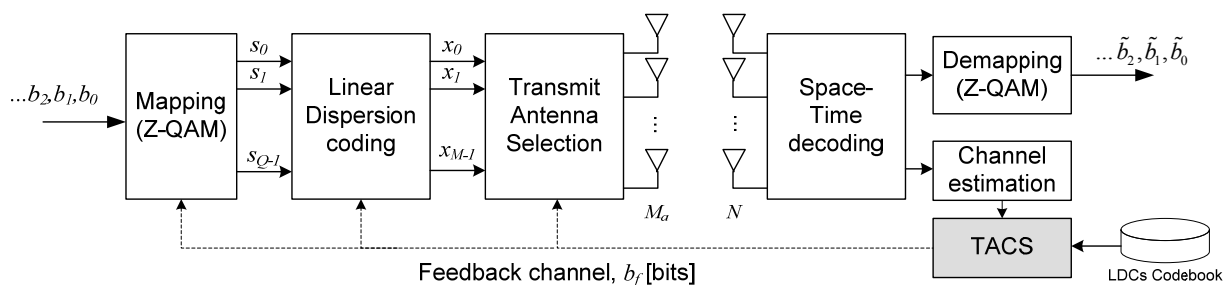


Fig. 5. Proposed TACS spatial adaptation scheme and integration into the transmission scheme.

4.4.2 TACS performance evaluation

In this section we are setting the main parameters to evaluate the performance of the TACS scheme, the IEEE 802.16 standard is here used to carry out the experiment. Some parameters are depicted in Table 1, where perfect synchronization is assumed and inter-cell interference is not considered. The used modulation is a Z-QAM ($Z=\{2,4,16,64\}$) with Gray mapping. According to the CSI measured, the BS determines: *i*) the antenna subset, *ii*) the LDC subset and in case of throughput maximization, *iii*) the MCS that maximizes the rate for a maximum Block Error Rate - BLER (second optimization criterion). The codebook is composed mainly by the Single Input Multiple Output (Maximum Ratio Combining is used at the receiver) receiver, the Alamouti 's Spatial Diversity (SD) coding scheme (referred as G2 in hereafter plotted the figures) [15], the pure spatial multiplexing (SM) and the Golden code [17]. The performance of the system is evaluated over 100.000 channel realizations,

OFDMA Air Interface and System Level configuration	
Subcarrier Permutation	Distributed (PUSC) and Contiguous (Band AMC)
FFT length, CP	2048, 12.5%
# of used subcarriers	1728
Modulation	{4,16,64}-QAM
Channel coding ¹	Turbo coding with rates: 1/3, 1/2, 2/3, 3/4
Channel model	Rayleigh and ITU Pedestrian A
Channel estimation (CQI)	Ideal without any delay
Frame duration, T_{frame}	5ms
DL/UL rate	2:1
OFDM symbols in the DL	30
Number of transmit antennas, M	{1,2,4}
Number of receive antennas, N	{1,2,4}
MIMO detector	MMSE
Rate (spectral efficiency)	{2,4,8} bits per channel use (bpcu)

Table 1. TACS evaluation framework system parameters

where for each realization a tile or a subchannel (specified in each analysis) is transmitted. In case of Partial Usage Subcarrier permutation (PUSC), the tile is formed by 4 subcarriers and 3 symbols, where 4 tones are dedicated to pilots as defined in IEEE 802.16e [17]. For the Band Adaptive Modulation and Coding (AMC) permutation scheme, each bin (equivalent to the tile concept) is comprised by 9 subcarriers where 1 tone is used as pilot. Perfect channel estimation is assumed at the receiver. Every $\log_2(Z)$ bits are mapped to one symbol. The channel models used are uncorrelated Rayleigh ($\mathbf{H} \sim \text{CN}(0,1)$) and the ITU Pedestrian A [38]. In both cases the channel is considered constant within a tile (*block fading channel model*). In case of uncorrelated Rayleigh the channel between tiles is uncorrelated, whereas in the ITU PedA case the channel is correlated both in frequency and time.

4.4.3 MIMO reference and simulation results

In Fig. 6, the reference performance for a fixed rate is depicted for $N=2$ when no transmit antenna selection neither code selection are used. For uncorrelated Rayleigh channel, we can observe that for low data rate, i.e. $R=\{2,4\}$, the Alamouti code outperforms the rest of the schemes. This is strictly related to the diversity order that G2 achieves equal to $g_d=N \times M=4$, whereas the SM and the Golden code with a linear receiver get a diversity order of

¹ Forward error correction is considered only for the throughput maximization case, where the LUT used to predict the BLER as a function of the ESINR, are obtained using the Duo-Binary Turbo code defined for IEEE 802.16e.

$g_d=(N - M + 1)=1$. At higher data rates ($R>8$), all the codes perform similarly in the analysed SNR range despite of the different diversity order between them.

4.4.4 TACS performance under bit error rate minimization criterion

In Fig. 7 and Fig. 8, the bit error rate performance using TACS is shown having a fixed rate $R=4$. Fig. 7 shows the improvement due to the increase in M_a and also the performance achieved when combined with code selection. It can be observed how the TAS increases the diversity order, leading to a large performance increase for the SM and Golden subsets. It is very important to notice that despite the diversity increase for all the LDC subsets, SD and SIMO schemes still perform better when each code is evaluated independently. However, in Fig. 8, we can observe that when the code selection is switched on, SIMO and Golden subsets are selected most times, while the usage of SIMO increases with the SNR and the usage of SM and the Golden code increases with M_a . Furthermore, the achieved improvement by the TACS is clearly appreciated in Fig. 7, where an SNR improvement of approximately 1dB is obtained for $M_a=\{3,4\}$. It is also surprising that the SM code is rarely selected knowing that the Golden code should always outperform SM since it obtains a higher diversity. However, as it is observed in Fig. 8, for less than 5% of the channel realizations the SM may outperform slightly the Golden code. Whether the singular value decomposition of the effective channel H is analysed when SM is selected, it has been observed that when all singular values are very close, both the SM and the Golden code lead to very similar performances, therefore no matter which one is selected.

In Fig. 9 and Fig. 10, the performance using the TACS is again analysed for $R=8$. In Fig. 9 the different diversity orders of SD, SM, and the Golden Code are illustrated. We can appreciate here that the SM and the Golden code show the best performance when $M_a=\{3,4\}$, and also for $M_a=2$ when $\text{SNR}\leq 18\text{dB}$. Furthermore the increase in the diversity order due to TACS can be observed in both Fig. 7 and Fig. 9. The maximum diversity order ($g_d = M_a N$) is achieved since at least one LDC (SIMO and G2) from those in the codebook are able to achieve the maximum diversity order.

Moreover, the BER using the TACS is equivalent to that obtained from the SISO scheme (referred as SISO_{eq} in the plots) over a Rayleigh fading channel with the same rate R , a diversity order $g_d=M_a N$ and a coding gain equal to Δ . The performance of this *equivalent* SISO scheme, in terms of the bit error rate probability P_b , can be obtained directly by close expressions that are found in [41][42] and applying the Craig's formula in [43],

$$P_b = \frac{1}{\log_2 \sqrt{Z}} \sum_{i=1}^{\log_2 \sqrt{Z}} P_b(i) \quad (28)$$

$$P_b(i) = \frac{2}{\sqrt{Z}} \sum_{k=0}^{(1-2^{-i})\sqrt{Z}-1} \omega(k, i, Z) \frac{1}{\pi} \int_0^{\pi/2} \left(1 + (2k+1)^2 \frac{3\rho_b}{2(Z-1)\sin^2 \theta} \right)^{-g_d} d\theta \quad (29)$$

$$\omega(k, i, Z) = (-1)^{\lfloor \frac{k \cdot 2^{i-1}}{\sqrt{Z}} \rfloor} \left(2^{i-1} - \left\lfloor \frac{k \cdot 2^{i-1}}{\sqrt{Z}} + \frac{1}{2} \right\rfloor \right) \quad (30)$$

where $\rho_b = \Delta \cdot \rho / \log_2(Z)$, $\lfloor x \rfloor$ means the smallest integer of x , and Z is the modulation order of the Z-QAM modulation.

The values of Δ for different combinations of $M_a=\{2,3,4\}$, $N=\{2,3,4\}$ and $R=\{4,8\}$ are depicted in Table 2. These values have been obtained adjusting the BER approximation in Eq. (28) to the empirical BER. As shown in Fig. 7 and Fig. 9 the performance of the TACS schemes is perfectly parameterized under the equivalent SISO model. Notice also that the power gain is constant across the whole SNR range.

Δ		N=2	N=3	N=4
R = 4	$M_a=2$	2.66	3.9	6.31
	$M_a=3$	3.20	5.2	8.41
	$M_a=4$	3.75	6.2	9.44
R = 8	$M_a=2$	4.20	9	14
	$M_a=3$	6.75	14.5	23
	$M_a=4$	9.00	19	28.5

Table 2. Coding gain Δ for the TACS proposal with $M_a=\{2,3,4\}$, $N=\{2,3,4\}$, and $R=\{4,8\}$.

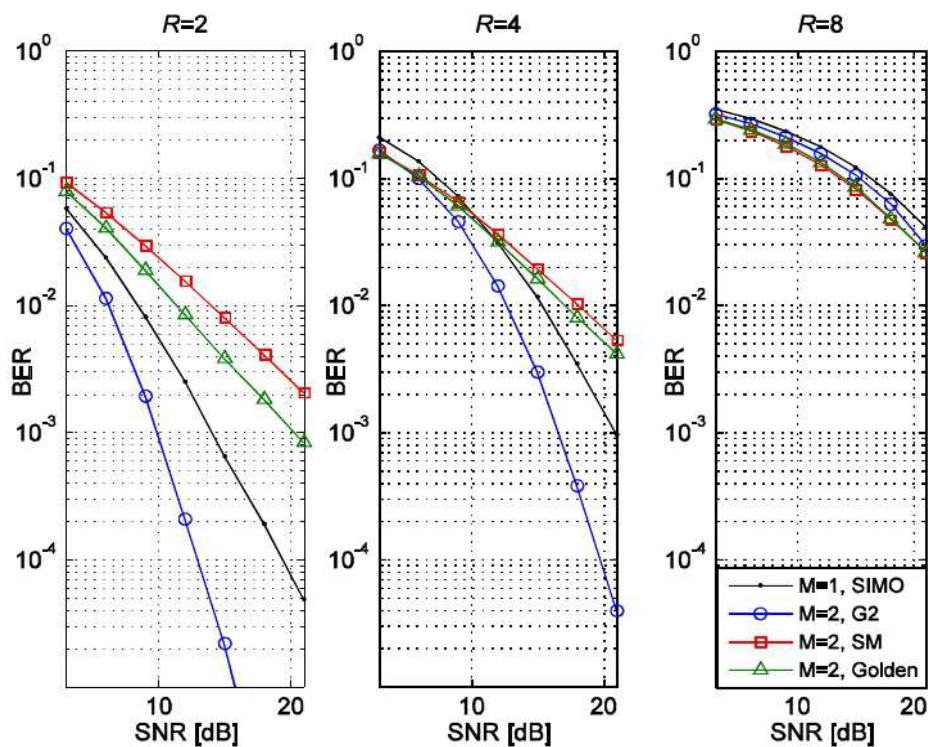


Fig. 6. Uncoded BER for uncorrelated Rayleigh channel with MMSE detector and $N=2$.

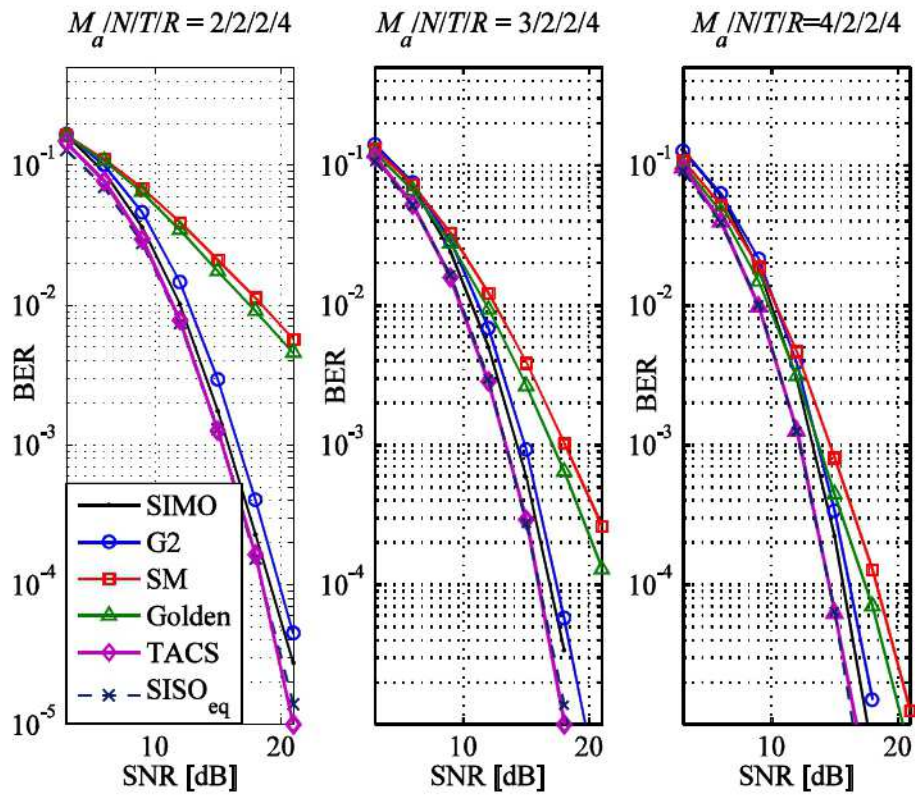


Fig. 7. Uncoded BER performance when $N=2$, $R=4$, $M_a=\{2,3,4\}$ for uncorrelated MIMO Rayleigh channel and MMSE linear receiver.

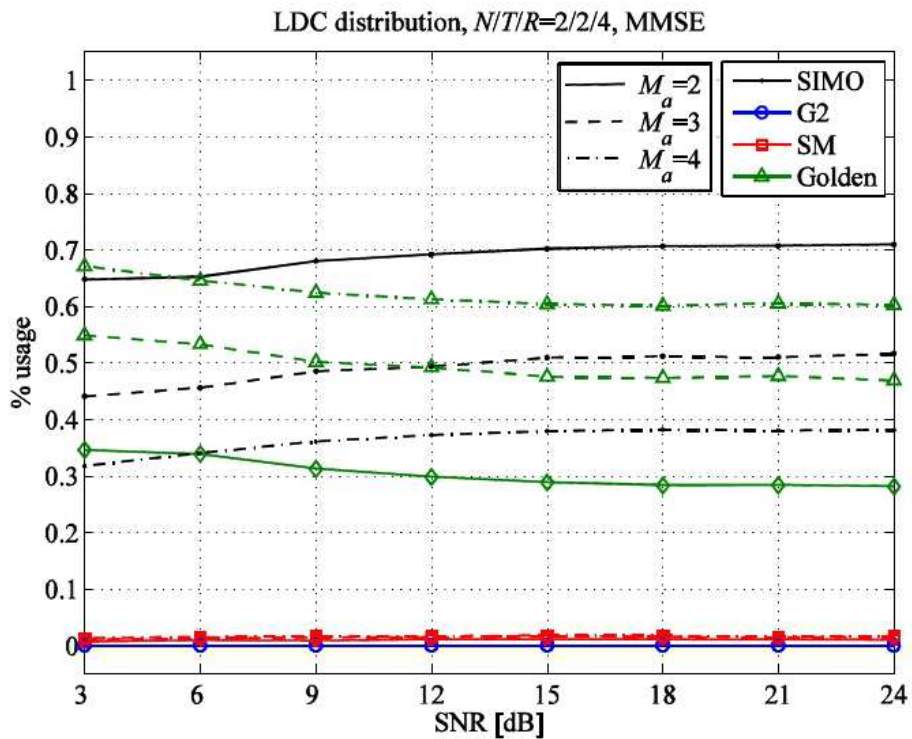


Fig. 8. LDC selection statistics with $N=2$, $R=4$, $M_a=\{2,3,4\}$ for uncorrelated MIMO Rayleigh channel and MMSE linear receiver.

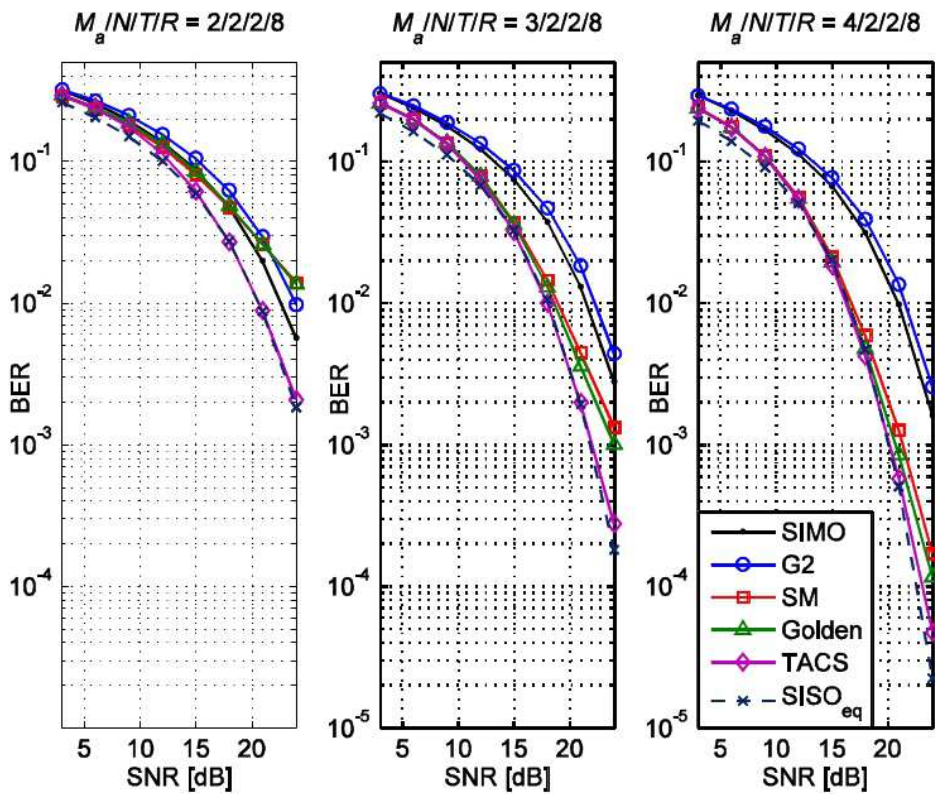


Fig. 9. Uncoded BER performance when $N=2$, $R=8$, $M_a=\{2,3,4\}$ for uncorrelated MIMO Rayleigh channel and MMSE linear receiver.

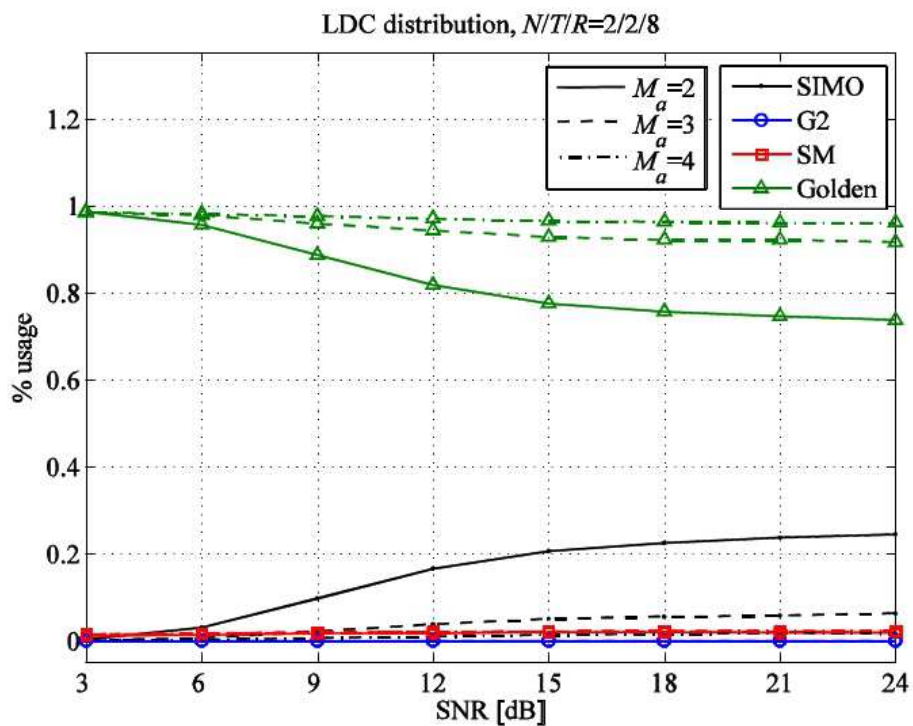


Fig. 10. LDC selection statistics when $N=2$, $R=8$, $M_a=\{2,3,4\}$ for uncorrelated MIMO Rayleigh channel and MMSE linear receiver.

4.4.5 TACS performance under throughput maximization criterion

In this section the performance of the TACS adaptation scheme in case the throughput is maximized (see Eq.(27)) is analysed. Then, for such adaptation scheme, the antenna set and the LDC code that maximizes the throughput is selected. In addition, the highest MCS (in the sense of spectral efficiency) that achieves a BLER<0.01 (1%) is also selected. The look-up-table used for mapping the ESINR to the BLER is shown and described in [14]. In the scenarios considered, the minimum allocable block length according the IEEE 802.16e standard was selected [17] (i.e. the number of sub-channels N_{sch} occupied per block varies between 1 and 4). The number of available antennas is $M_a=2$ whereas $N=2$.

In Fig. 11 and Fig. 12, the spectral efficiency achieved by TACS with adaptive Modulation and Coding (AMC) as well as the LDC statistics are shown. For Spatial Multiplexing (SM), two encoding options named *Vertical Encoding* (VE) and *Horizontal Encoding* (HE) are considered. For the first scheme, VE, the symbols within the codeword apply the same MCS format, whereas for the second, HE, each symbol may apply a different MCS. Clearly the first is more restrictive since is limited by the worst stream ($\min(ESNR_q)$) whereas the second is able to exploit inter-stream diversity at the expense of higher signalling requirements (at least twice as that required with VE in case of $M=2$).

Depicted performances shown that at low SNRs (SNR<13dB), the SIMO and Alamouti achieve the highest spectral efficiencies (something that has been already obtained in several previous works [10]). However, as the SNR is increased, the codes with higher multiplexing capacity (e.g. the SM and the Golden code) are preferred. It could be also observed that the SM with VE implies a loss of around 2dB compared to the Golden code, but when HE is used, the Golden code is around 0.5dB worse than the SM-HE.

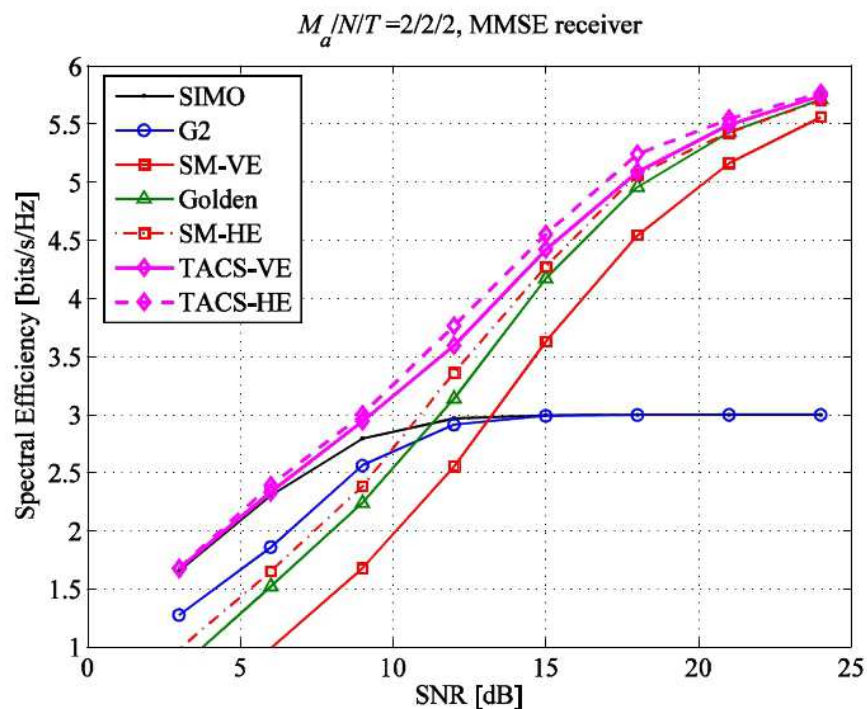


Fig. 11. Spectral efficiency under TACS with throughput maximization criterion with $M_a=2$, $N=2$, adaptive MCS and MMSE receiver for an uncorrelated MIMO Rayleigh channel.

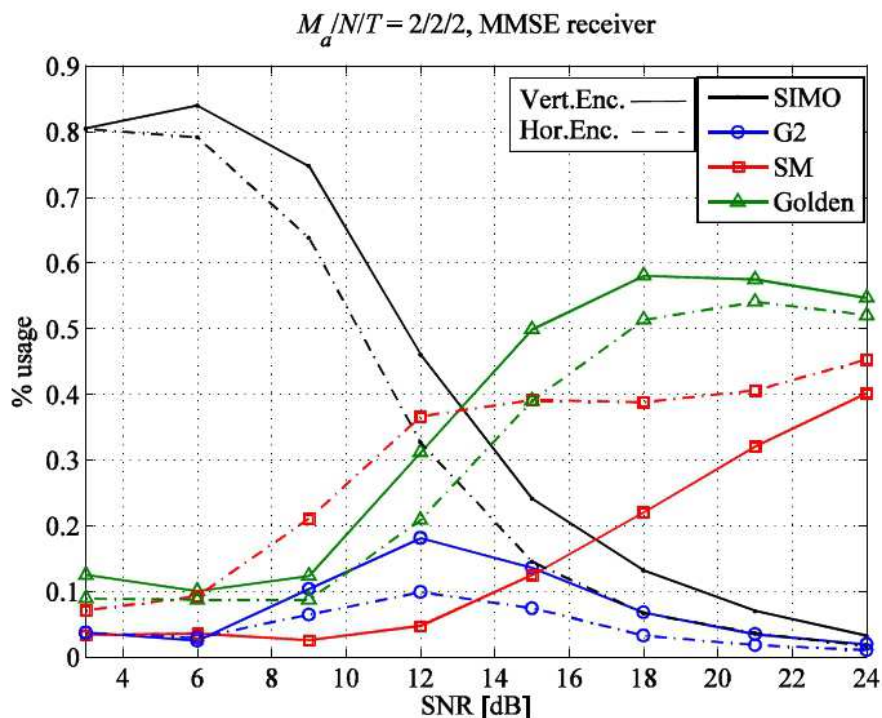


Fig. 12. LDC selection statistics under TACS with throughput maximization criterion with $M_a=2$, $N=2$, adaptive MCS and MMSE receiver for an uncorrelated MIMO Rayleigh channel.

To gain further insights of the TACS behaviour, the statistics of LDC selection as a function of the average SNR are plotted in Fig. 12. We can clearly appreciate that at low SNR the preferred scheme is SIMO where all the power is concentrated in the best antenna, while as the SNR is increased full rate codes ($Q=M$) are more selected since they permit to use lower size constellations. Moreover, comparing SM-VE with SM-HE, we can observe that SM-HE is able to exploit the stream's diversity and hence achieves a higher spectral efficiency than if the Golden code is used. Actually, at average SNR=12, the SM with HE is the scheme selected for most frames, even more than SIMO. These results show that in case of linear receivers (e.g. MMSE) the TACS scheme with AMC gives a noticeable SNR gain (up to 3dB) in a large SNR margin (SNR from 6 to 18dB) and also is a good technique to achieve a smooth transition between diversity and multiplexing.

5. MIMO in IEEE 802.16e/m

The use of MIMO may improve the performance of the system both in terms of link reliability and throughput. As it was discussed in previous sections, both concepts pull in

different directions, and in most cases a trade-off between both is met by each specific space-time code. From a system point of view, and due to the inherent time/freq variability of the wireless channel, no code is optimal for all channel conditions, and at most, the codes can be optimized according to the ergodic properties of the channel. In fact, this is the reason why the TACS scheme is able to bring significant gain compared to a scheme where the same space-time code is always used. This situation is well-known and it is the reason why in most of the Broadband Wireless Access (BWA) systems, the number of space-time codes is increasing.

In IEEE 802.16e/m, two types of MIMO are defined, Single User MIMO and Multiuser MIMO, the first corresponding to the case where one resource unit (the minimum block of frequency-time allocable subcarriers) is assigned to a single user, and the second when this one is shared among multiple users.

In case of two transmit antennas, IEEE 802.16e/m defines two possible encoding schemes referred as Matrix A and Matrix B. Matrix A corresponds to the Alamouti scheme, while Matrix B corresponds to the Spatial Multiplexing (SM) case. In case of using SM, WiMAX allows both Vertical Encoding (VE) and Horizontal Encoding (HE). In the first case, VE, all the symbols are encoded together and belong to the same *layer*. In addition to Matrix A and Matrix B, IEEE 802.16 also defines a Matrix C which corresponds to the Golden Code. This code is characterized for providing the highest spatial diversity for the spatial rate $R=2$. In case of 3 and 4 transmit antennas, WiMAX also defines the encoding schemes of Matrix A, Matrix B, and Matrix C, all of them providing different trade-offs between diversity and spatial multiplexing.

The list of combinations is even longer since WiMAX allows antenna selection and antenna grouping, therefore, the list of encoding matrices also includes the possibility that not all antennas are used, and only a subset are selected (the list of matrices in Table 3 do not show this possibility). In case not all the antennas are used, the power is normalized so that the same power is transmitted regardless of the number of active antennas.

Besides the possibility to select among any of the previous coding matrices, IEEE 802.16e/m also allows the use of precoding. In this case, the space-time coding output is weighted by a matrix before mapping onto transmitter antennas

$$z = Wx \quad (31)$$

where x is $M_t \times 1$ vector obtained after ST encoding, where M_t is the number of streams at the output of the space time coding scheme. The matrix W is a $M \times M_t$ weighting matrix where M is the number of transmit antennas. The weighting matrix accepts two types of adaptation depending on the rate of update, named short term closed-loop precoding and long term closed-loop precoding.

In the later IEEE 802.16m, the degrees of flexibility has been broadened, allowing several kinds of adaptation [44]. On top of this, IEEE 802.16m includes also ST codes for up to 8 transmitter antennas, enabling the transmission at spectral efficiencies as high as 30bits/sec/Hz which become necessary to achieve the very high throughputs demanded for IMT-Advanced systems [45].

M	N_{\min}	T	Q	R	MIMO Encoding Matrix	Name
2	1	2	2	1	$\begin{bmatrix} s_0 & -s_1^* \\ s_1 & s_0^* \end{bmatrix}$	Alamouti (a.k.a. Matrix A)
2	2	1	2	2	$[s_0 \ s_1]^T$	Spatial Multiplexing (a.k.a. Matrix B)
2	2	2	4	2	$\frac{1}{\sqrt{1+r^2}} \begin{bmatrix} s_0 + jrs_3 & rs_1 + s_2 \\ s_1 - rs_2 & s_3 + jrs_0 \end{bmatrix}, r = \frac{-1 + \sqrt{5}}{2}$	Golden Code (a.k.a. Matrix C)
3	2	4	4	1	$\begin{bmatrix} s_0 & -s_1^* & 0 & 0 \\ s_1 & s_0^* & s_2 & -s_3^* \\ 0 & 0 & s_3 & s_2^* \end{bmatrix}$	Matrix A ²
3	2	4	4	1	$\begin{bmatrix} \sqrt{\frac{3}{4}} & 0 & 0 \\ 0 & \sqrt{\frac{3}{4}} & 0 \\ 0 & 0 & \sqrt{\frac{3}{2}} \end{bmatrix} \begin{bmatrix} s_0 & -s_1^* & s_4 & -s_5^* \\ s_1 & s_0^* & s_5 & s_4^* \\ s_6 & -s_7^* & s_2 & -s_3^* \end{bmatrix}$	Matrix B
3	2	4	4	1	$[s_0 \ s_1 \ s_2]^T$	Matrix C
4	1	4	4	1	$\begin{bmatrix} s_0 & -s_1^* & 0 & 0 \\ s_1 & s_0^* & 0 & 0 \\ 0 & 0 & s_2 & -s_3^* \\ 0 & 0 & s_3 & s_2^* \end{bmatrix}$	Matrix A
4	2	4	8	2	$\begin{bmatrix} s_0 & -s_1^* & s_4 & -s_5^* \\ s_1 & s_0^* & s_5 & s_4^* \\ s_2 & -s_3^* & s_6 & -s_7^* \\ s_3 & s_2^* & s_7 & s_6^* \end{bmatrix}$	Matrix B
4	4	1	4	4	$[s_0 \ s_1 \ s_2 \ s_3]^T$	Matrix C

Table 3. WiMAX IEEE 802.16e MIMO encoding matrices.

² In case of 3 and 4 transmit antennas, Matrix A, B and C accept different antenna grouping and selection schemes. This antenna grouping does similar effects as TACS, indicating which antennas and Space-time codes are preferred.

6. Summary

The use of multiple antenna techniques at transmitter and receiver sides is still considered a hot research topic where the channel capacity can be increased if multiple streams are multiplexed in the spatial domain. The study on the trade-off between diversity and multiplexing has motivated the emergence of many different space-time coding architectures where most of the proposed schemes lie in the form of Linear Dispersion Codes. Furthermore, as it was shown by the authors in previous sections, when the transmitter disposes of partial channel state information, robustness and throughput can be very significantly improved. One of the simplest adaptation techniques is the use of antenna selection, which increases the diversity of the system up to the maximum available ($g_d = M_a \times N_a$). On the other hand, when transmit antenna selection is combined with code selection a coding gain is achieved. In this chapter, a joint Transmit Antenna and space-time Coding Selection (TACS) scheme previously proposed by the authors has been described. The TACS algorithm allows two kind of optimization: *i*) bit error rate minimization, and *ii*) throughput maximization. One important result obtained from these studies is that the number of required space-time coding schemes is quite low. In fact, previous studies by the author have shown that in case of spectral efficiencies of 8bits/second/Hertz or lower, using SIMO, Alamouti, SM, and the Golden code is enough to maximize the performance (for higher rates, codes with higher spatial rate would be required). Furthermore, the worse performance achieved by linear receivers (e.g. ZF, MMSE) is compensated by the TACS scheme, which allows to achieve performances close to those obtained with the non-linear receivers (e.g. the Maximum Likelihood) with much lower computational requirements. As a final conclusion, it can be considered that transmit antenna selection with linear dispersion code selection can be an efficient spatial adaptation technique whose low feedback requirements make it feasible for most of the Broadband Wireless Access systems, especially in case of low mobility.

7. Acronyms

3GPP	3rd Generation Partnership Project
AWGN	Additive White Gaussian Noise
BLER	Block Error Rate
BS	Base Station
CSI	Channel State Information
FDD	Frequency Division Duplexing
LDC	Linear Dispersion Codes
LTE	Long Term Evolution
MCS	Modulation and Coding Scheme
MIMO	Multiple Input Multiple Output
MMSE	Minimum Mean Square Error
OSTBC	Orthogonal Space-Time Block Code
QAM	Quadrature Amplitude Modulation
SIMO	Single Input Multiple Output
SISO	Single Input Single Output
SM	Spatial Multiplexing
SNR	Signal To Noise Ratio

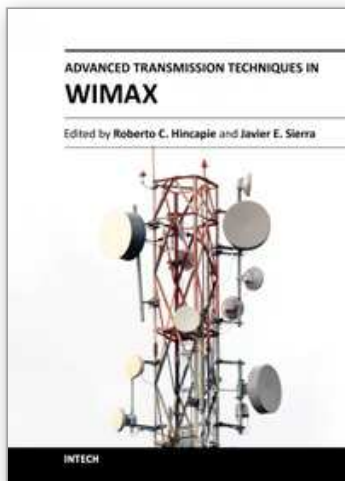
STBC	Space-Time Block Code
TACS	Transmit Antenna and (space-time) Code Selection
TDD	Time Division Duplexing
UPA	Uniform Power Allocation
ZF	Zero Forcing

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This book has been prepared to present the state of the art on WiMAX Technology. The focus of the book is the physical layer, and it collects the contributions of many important researchers around the world. So many different works on WiMAX show the great worldwide importance of WiMAX as a wireless broadband access technology. This book is intended for readers interested in the transmission process under WiMAX. All chapters include both theoretical and technical information, which provides an in-depth review of the most recent advances in the field, for engineers and researchers, and other readers interested in WiMAX.

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