FLEXIBLE MIMO ARCHITECTURES: GUIDELINES IN THE DESIGN OF MIMO PARAMETERS

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ABSTRACT

One of the multiple advantages of communicating through MIMO systems is their inherent ability to provide flexible configurations. Following this line of thought, in this paper we present a generic framework to study the degrees of freedom in the design of MIMO communication systems (e.g.: code length, number of multiplexed streams, or receiver structure). Precisely, we focus our efforts to bridge the gap between the design of MIMO systems with full and no channel state information at the transmitter side and also with different complexity degrees at the receiver side. For instance, we can establish a trade-off, not only between the achievable rates and the diversity or beamforming gains, but also between the rate and the robustness to uncertainties in the channel state information.

1. INTRODUCTION

In the last years, much attention has been paid to the design of transmitters for multi-input-multi-output (MIMO) channels. The design depends on the quality of the channel state information (CSI) available at the transmitter and also on the detector employed at the receiver. Many authors have considered the use of suboptimum linear receivers in order to reduce the complexity, although the optimum performance is achieved by a maximum likelihood (ML) detector, which may require an exhaustive search over all the possible transmitted symbol sequences.

Depending on the quality of the CSI, different transmitter architectures have been proposed in the literature. For example, in the case of having no CSI at the transmitter, a classical approach consists in employing a space-time code, which may be convolutional [1] or block [2]. Within the family of space-time codes, the orthogonal space-time block codes (OSTBC) [2] have received special attention since they attain full diversity and can be decoded with a low complexity requiring only linear computations. Other solutions, such as quasi-orthogonal space-time codes (QOSTBC) [3] try to increase the symbol rate, although the optimum ML detection cannot be implemented with linear operations.

In this paper, we present a generic signal model that describes the transmission process as a concatenation of a temporal processing, a power allocation, and a spatial processing stages. This generic signal model encompasses several transmitter architectures under very diverse situations, including different degrees of quality of the CSI and the two already mentioned receivers: a linear and a ML detector. For some of these transmitter architectures, we present the cost function to be optimized in order to minimize the bit error rate (BER) or the pairwise error probability (PEP).

This paper is organized as follows. Section 2 presents the generic signal model, which is particularized to some known cases in Section 3. Section 4 is devoted to the extension of the design to the case of ML detectors, whereas in Section 5, a robust design is presented. Finally, some conclusions are obtained in Section 6.

2. GENERIC SIGNAL MODEL

In the following, a discrete time narrowband multiplexing system with n_T transmit and n_R receive antennas corrupted with additive Gaussian noise is considered. Let us define $\mathbf{x}_n \in \mathbb{C}^{n_T}$ as the transmitted signal at time $n \in \mathbb{N}$, where $[\mathbf{x}_n]_k$ represents the transmitted signal through k-th antenna. We also define $\mathbf{H} \in \mathbb{C}^{n_R \times n_T}$ as the channel matrix, which is assumed to remain constant for the duration of N discrete time intervals, and where $[\mathbf{H}]_{jk}$ represents the path gain from the k-th transmitter to the j-th receiver. Finally, $\mathbf{w}_n \in \mathbb{C}^{n_R}$ is defined as the noise vector, where $[\mathbf{w}_n]_j$ represents the noise component received at the j-th antenna. The noise vector is modeled as a temporally and spatially white circularly symmetric Gaussian distributed random vector, with $\mathbb{E}|[\mathbf{w}_n]_j|^2 = N_0/2$, $\forall j, n$. The received signals vector at time n, $\mathbf{y}_n \in \mathbb{C}^{n_R}$, for this model can be expressed as $\mathbf{y}_n = \mathbf{H}\mathbf{x}_n + \mathbf{w}_n$. The transmitted signal $\mathbf{x}_n$ is considered to be a linear function of the binary data symbols vector, $\mathbf{s} \in \{-1, +1\}^{n_S}$, which contains the n_S information bits to be sent. Precisely, $\mathbf{x}_n$ is obtained from $\mathbf{s}$ as a three-step linear process (see Fig. 1),

$$\mathbf{x}_n = \sqrt{E_s}\mathbf{U}\mathbf{P}_n^{\frac{1}{2}}\boldsymbol{\mathcal{V}}_n^H\mathbf{s}, \quad n = 1, \ldots, N, \tag{1}$$

where $\mathbf{U} \in \mathbb{C}^{n_T \times n_M}$ is a spatial processing matrix containing n_M orthonormal columns of length n_T coupling the transmission through spatial modes, and $\mathbf{P}_n^{\frac{1}{2}}$ is a positive definite diagonal matrix taking into account the power allocation among the spatial modes. The purpose of $\boldsymbol{\mathcal{V}}_n^H \in \mathbb{C}^{n_M \times n_S}$ is three fold. First of all, it builds the constellation from the binary data. Secondly, it spreads the constellation symbols among the spatial modes. Finally, as $\boldsymbol{\mathcal{V}}_n^H$ is allowed to be a function of discrete time n, it also takes care of the temporal processing. Notice that the bitrate of the system as it is described in (1) is n_S/N as $\mathbf{s}$ remains constant from $n = 1$ to $n = N$.

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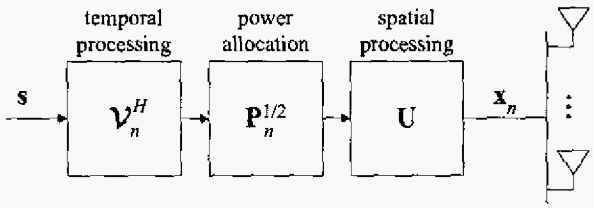

Fig. 1. Generic Signal Model Scheme

It is important to mention that, in general, $\mathbf{U}$ matrix will depend on the channel realization or on the CSI that is made available at the transmitter side. This fact allows us to consider an equivalent channel, $\mathbf{H}_c$, which incorporates the effects of the spatial processing matrix as $\mathbf{H}_c = \mathbf{H}\mathbf{U}$. As it will prove useful later, we define $\mathbf{H}_c^H = [\mathbf{h}_1^c \cdots \mathbf{h}_{n_R}^c]$, and then the channel correlation matrix is given by

$$\mathbf{R_H} = \mathbf{H}_c^H \mathbf{H}_c = \sum_{r=1}^{n_R} \mathbf{h}_r^c \mathbf{h}_r^{cH}. \quad (2)$$

In addition, for the sake of notation, we will group the power allocation matrix, $\mathbf{P}_n^{\frac{1}{2}}$, and the temporal processing matrix, $\mathcal{V}_n^H$, to form a new matrix, $\mathbf{V}_n^H \in \mathbb{C}^{n_M \times n_S}$, as

$$\mathbf{V}_n^H = \mathbf{P}_n^{\frac{1}{2}} \mathcal{V}_n^H. \quad (3)$$

Using the above definitions, the received signal can now be expressed as $\mathbf{y}_n = \sqrt{E_s}\mathbf{H}_c\mathbf{V}_n^H\mathbf{s} + \mathbf{w}_n$.

If we impose that $\mathbb{E}\{\mathbf{s}\mathbf{s}^H\} = \mathbf{I}_{n_S}$, the transmitted energy at time instant n, E_T^n, will be

$$E_T^n = \mathbb{E}\,\mathrm{Tr}\left(\mathbf{x}_n\mathbf{x}_n^H\right) = E_s\,\mathrm{Tr}\left(\mathbf{V}_n^H\mathbf{V}_n\right). \quad (4)$$

For the particular case where $\mathcal{V}_n^H\mathcal{V}_n = \mathbf{I}_{n_M}$, last expression simplifies to $E_T^n = E_s\,\mathrm{Tr}\,\mathbf{P}_n$.

3. PARTICULAR CASES

The generic signal model of the previous section includes some designs that can be found in the literature as particular cases. In the following, some examples are given.

3.1. Delayed decision architectures

Delayed decision architectures are obtained if the decision on the transmitted data symbols vector, $\hat{\mathbf{s}}$, is performed based upon a collection of N received signal vectors, *i.e.*, $\hat{\mathbf{s}} = f(\mathbf{y}_1, \mathbf{y}_2, \ldots, \mathbf{y}_N)$. Notice that these architectures can be used to enhance the diversity gain since they allow the transmission of multiple copies of the same message, in that case, the system will suffer from a penalty in the multiplexing gain. One of the most prolific examples of delayed decision architectures are space-time codes [1], among which OSTBC are very popular.

3.1.1. Orthogonal space-time block codes

OSTBC [2] attain full diversity while the optimum receiver only requires linear operations. However, a full symbol rate equal to 1 can only be achieved for two transmit antennas. OSTBC transmitters can also be formulated as a particular case of the notation presented in Section 2. In this situation, the matrices $\mathbf{U}$ and $\mathbf{P}^{\frac{1}{2}}$ are both square with dimensions $n_T \times n_T$, *i.e.*, $n_M = n_T$, and equal to the identity matrix, up to a scalar factor which depends on the transmission power. The matrices $\{\mathcal{V}_n^H\}$, depend on the pattern of the OSTBC and also on the modulation used.

For example, if we wish to particularize our system model for the case of an Alamouti transmission scheme [4] with QPSK modulation ($n_T = N = 2, n_S = 4$) it would yield

$$\mathcal{V}_1^H = \frac{1}{\sqrt{2}}\begin{pmatrix} 1 & j & 0 & 0 \\ 0 & 0 & 1 & j \end{pmatrix}, \; \mathcal{V}_2^H = \frac{1}{\sqrt{2}}\begin{pmatrix} 0 & 0 & -1 & j \\ 1 & -j & 0 & 0 \end{pmatrix}, \quad (5)$$

and the transmitted signal would be equivalent to the well-known expression of Alamouti's code

$$[\mathbf{x}_1 \; \mathbf{x}_2] = \frac{1}{\sqrt{2}}\begin{pmatrix} s_1 + js_2 & -s_3 + js_4 \\ s_3 + js_4 & s_1 - js_2 \end{pmatrix} \sim \begin{pmatrix} q_1 & -q_2^* \\ q_2 & q_1^* \end{pmatrix},$$

where q_i represents a QPSK symbol.

3.2. Instantaneous decision architectures

Instantaneous decision schemes are obtained whenever the decision on the transmitted data symbols vector is performed as soon as the signal is received. In this case $\hat{\mathbf{s}} = f(\mathbf{y}_n)$, *i.e.*, $N = 1$, which implies $\{\mathcal{V}_n^H\} = \mathcal{V}^H$. These architectures allow a natural form of exploiting the multiplexing capabilities of MIMO channels (and simultaneously discarding the diversity gain), for which a different data symbols vector $\mathbf{s}$ can be transmitted each time the channel is accessed. In the following subsections, two instantaneous decision architectures, designed for the two extreme cases of no-CSI and full-CSI, are described.

3.2.1. Channel unknown to the transmitter

A paradigmatic example which does not require CSI is BLAST [5], that attains a symbol rate equal to the number of transmit antennas. Within the family of BLAST techniques, in V-BLAST the transmission is performed by sending independent symbol streams through the transmit antennas, each one with a duration equal to one time interval ($N = 1$). In such a situation, the signal model is particularized as follows. Both the matrices $\mathbf{U}$ and $\mathbf{P}^{\frac{1}{2}}$ are exactly equal to the previous case corresponding to OSTBC. The main difference is concerned with the matrix $\mathcal{V}^H$, which is also equal to the identity matrix.

3.2.2. Optimum linear transceivers with perfect CSI

The performance of any communication system can be importantly improved when perfect CSI is available and exploited to design both sides of the system. If only linear operations are permitted at both sides, several designs arise depending on the quality measure. In [6], it is shown that the optimum transmitter for a variety of cost functions is given also by the generic signal model presented in Section 2. The matrix $\mathbf{U}$ contains the n_M maximum eigenvectors of $\mathbf{H}^H\mathbf{H}$, where $n_M = \min\{n_S, \mathrm{rank}(\mathbf{H})\}$, and, in case that the objective is the minimization of the BER, $\mathbf{P}^{\frac{1}{2}}$ and $\mathcal{V}^H$ are designed so that the mean square error associated to the detection of each of the n_S transmitted symbols is equal.

4. MAXIMUM LIKELIHOOD RECEIVER

In this section it will be assumed that the receiver will perform ML detection of the transmitted symbols, s_i, with $i = 1, \ldots, n_S$. As was stated in Section 2, each element of the data symbols vector, as in (1), is randomly chosen from the set $\{-1, +1\}$ and then is scaled by the factor $\sqrt{E_s}$. For this particular case, the set of possible codewords, $\mathcal{C}$ has size $|\mathcal{C}| = 2^{n_S}$. For notation purposes, the p-th element of $\mathcal{C}$ will be denoted by $\mathbf{s}_p$, with $p = 1, \ldots, 2^{n_S}$.

The estimate of the transmitted vector is given by the vector $\mathbf{s}$ maximizing the likelihood, which can be written as

$$\mathcal{L}(\mathbf{s}) = \sum_{n=1}^{N} \|\mathbf{y}_n - \sqrt{E_s}\mathbf{H}_c\mathbf{V}_n^H\mathbf{s}\|^2. \tag{6}$$

The vector $\mathbf{s}_p$ is chosen if $\mathcal{L}(\mathbf{s}_p) > \mathcal{L}(\mathbf{s}_q),\ \forall q \neq p$, which, after some manipulations from (6), implies

$$\sum_{n=1}^{N} \tilde{\mathbf{s}}_{p,q}^{n\ H}\mathbf{R_H}\tilde{\mathbf{s}}_{p,q}^{n} > \frac{-2}{\sqrt{E_s}}\sum_{n=1}^{N}\Re\left\{\mathbf{w}_n^H\mathbf{H}_c\tilde{\mathbf{s}}_{p,q}^{n}\right\}, \quad \forall q,$$

where $\tilde{\mathbf{s}}_{p,q}^{n} = \mathbf{V}_n^H(\mathbf{s}_p - \mathbf{s}_q)$. Taking the term on the right of the inequality in last equation as a Gaussian random variable, it can be shown that the PEP of deciding $\mathbf{s}_q$ instead of $\mathbf{s}_p$, $\Pr(\mathbf{s}_p \to \mathbf{s}_q) = P_e^{p,q}$, can be upper bounded by

$$P_e^{p,q} \leq K_0 \exp\left(-\frac{E_s}{2N_0}\sum_{n=1}^{N}\tilde{\mathbf{s}}_{p,q}^{n\ H}\mathbf{R_H}\tilde{\mathbf{s}}_{p,q}^{n}\right). \tag{7}$$

Assuming only one error is present and that it is located at s-th data symbol, *i.e.*, $[\mathbf{s}_p - \mathbf{s}_q]_j = \pm 2\delta_{j-s}$, the expression for the PEP can be further simplified to

$$P_e^{p,q} \leq K_0 \exp\left(-\frac{2E_s}{N_0}\sum_{n=1}^{N}\mathbf{v}_{n,s}^H\mathbf{R_H}\mathbf{v}_{n,s}\right), \tag{8}$$

where $\mathbf{v}_{n,s}$ represents the s-th column of matrix $\mathbf{V}_n^H$.

4.1. Instantaneous knowledge of the channel matrix

In order to minimize the worst PEP in (8), an equal error probability has to be imposed for each stream, which implies:

$$\sum_{n=1}^{N}\mathbf{v}_{n,s}^H\mathbf{R_H}\mathbf{v}_{n,s} = \beta_0 N, \quad \forall s. \tag{9}$$

Taking advantage of the instantaneous knowledge of the channel matrix, it is sufficient to assign, for example,

$$\mathbf{v}_{n,s}^H\mathbf{R_H}\mathbf{v}_{n,s} = \beta_0, \quad \forall s, \tag{10}$$

in order to ensure an equal error probability for each stream. Last condition can be obtained, in a simple way, by

$$\mathbf{v}_{n,s} = \beta_0^{\frac{1}{2}}\frac{\mathbf{c}_{n,s}}{\sqrt{\mathbf{c}_{n,s}^H\mathbf{R_H}\mathbf{c}_{n,s}}}, \tag{11}$$

where, in principle, $\mathbf{c}_{n,s}$ are arbitrary vectors, which should be chosen as orthogonal as possible. With this assignment the probability of error can be finally expressed as

$$P_e^{p,q} \leq K_0 \exp\left(-\frac{2E_T^n}{N_0}\frac{N}{\displaystyle\sum_{s=1}^{n_S}\frac{\|\mathbf{c}_{n,s}\|^2}{\mathbf{c}_{n,s}^H\mathbf{R_H}\mathbf{c}_{n,s}}}\right), \tag{12}$$

where $E_T^n = E_s\beta_0\sum_{s=1}^{n_S}\|\mathbf{c}_{n,s}\|^2(\mathbf{c}_{n,s}^H\mathbf{R_H}\mathbf{c}_{n,s})^{-1}$.

4.2. Lack of knowledge of the channel matrix

An alternative formulation for expression (8) is

$$P_e^{p,q} \leq K_0 \exp\left(-\frac{2E_s}{N_0}\operatorname{Tr}\left(\mathbf{R_H}\sum_{n=1}^{N}\mathbf{v}_{n,s}\mathbf{v}_{n,s}^H\right)\right). \tag{13}$$

Proceeding analogously as in the last section, in order to minimize the worst PEP, an equal error probability has to be imposed for each stream, but in this case, no information about the channel matrix is available at the transmitter side. However, it is still possible to find a design for $\mathbf{v}_{n,s}$ such that each stream has the same error probability by imposing

$$\sum_{n=1}^{N}\mathbf{v}_{n,s}\mathbf{v}_{n,s}^H = \beta_0\mathbf{I}_{n_M}, \quad \forall s. \tag{14}$$

If last condition is met, after some manipulations, the error probability reads as

$$P_e^{p,q} \leq K_0 \exp\left(-\frac{2E_T^n}{N_0}\frac{N}{n_S}\frac{\operatorname{Tr}\mathbf{R_H}}{n_M}\right), \tag{15}$$

where $E_T^n = \beta_0 n_s n_M E_s/N, \forall n$.

4.3. Receiver covariance matrix knowledge

The instantaneous PEP obtained previously in (8) can be averaged over the distribution of the fading coefficients assuming that there is no correlation among the channel seen by each receive antenna. First of all, we introduce

$$\mathbf{A}_s = \sum_{n=1}^{N}\mathbf{v}_{n,s}\mathbf{v}_{n,s}^H. \tag{16}$$

Using (16), the expected PEP can be formulated as

$$\mathbb{E}_{\mathbf{H}}P_e^{p,q} = K_1\prod_{r=1}^{n_R}\int\exp\left(-\frac{2E_s}{N_0}\mathbf{h}_r^{c\,H}\mathbf{A}_s\mathbf{h}_r^c\right)dF_{\mathbf{h}_r}, \tag{17}$$

where $dF_{\mathbf{h}_r^c}$ represents the differential of the distribution of $\mathbf{h}_r^c$. For the case of Rayleigh fading without correlation between the receiving antennas, $\mathbf{h}_r \sim \mathcal{CN}(\mathbf{0}, \boldsymbol{\Sigma}_r)$ and $\mathbb{E}\mathbf{h}_r\mathbf{h}_t^H = \mathbf{0}, \forall r, t = 1\ldots n_R$, the expectation can be computed analytically as

$$\mathbb{E}_{\mathbf{H}}P_e^{p,q} = K_1\prod_{r=1}^{n_R}\det\left(\mathbf{I} + \frac{2E_s}{N_0}\boldsymbol{\Sigma}_r\mathbf{A}_s\right)^{-1}. \tag{18}$$

For the common case where the correlation matrix $\boldsymbol{\Sigma}_r$ is the same for each receive antenna last expression can be further simplified to

$$\mathbb{E}_{\mathbf{H}}P_e^{p,q} = K_1\det\left(\mathbf{I} + \frac{2E_s}{N_0}\boldsymbol{\Sigma}\mathbf{A}_s\right)^{-n_R}. \tag{19}$$

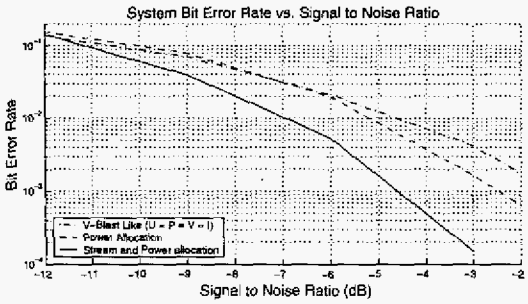

Fig. 2. System BER versus SNR for three different transmission architectures and ML receiver

4.3.1. *Simulations*

We have considered an instantaneous decision ($N = 1$) MIMO system, with $n_T = 4$, $n_R = 8$, and $n_S = 4$, where the transmitter is informed with $\mathbf{\Sigma}$. Notice that, in this case, expression (19) can be written as $\mathbb{E}_{\mathbf{H}} P_e^{p,q} = K_1 \left(1 + 2E_s/N_0 \mathbf{v}_s^H \mathbf{\Sigma} \mathbf{v}_s\right)^{-n_R}$.

For comparison purposes, we have simulated three different transmit architectures. The first one corresponds to a BPSK V-BLAST transmit scheme ($\mathbf{U} = \mathbf{I}_{n_T}$, $\mathbf{P} = \mathbf{I}_{n_T}$, and $\mathcal{V}^H = \mathbf{I}_{n_T}$, see 3.2.1). The second one also has $\mathbf{U} = \mathcal{V}^H = \mathbf{I}_{n_T}$, but it admits a power allocation matrix which maximizes the worst-case averaged PEP as in (19), *i.e.*, the product $[\mathbf{P}]_{ii}[\mathbf{\Sigma}]_{ii}$ is constant, $\forall i$. Finally, the third one transmits through the $n_M = 2$ best eigenmodes of $\mathbf{\Sigma}$, with associated eigenvalues λ_1 and λ_2, with a power allocation distribution that maximizes the worst-case PEP (*i.e.*, the product $[\mathbf{P}]_{ii}\lambda_i$ is also a constant), and a $\mathcal{V}^H$ matrix that distributes the $n_S = 4$ symbols inside these two eigenmodes, building a QPSK modulation scheme (similarly to $\mathcal{V}_1^H$ in (5)). The results of the simulation are plotted in Fig. 2. In this case it can be seen that the last system performs much better than the two others, due to the fact that it exploits better the information about the channel correlation.

5. ROBUST DESIGNS WITH IMPERFECT CSI

An instantaneous channel estimate can be available at the transmitter during the design in time division duplexing (TDD) systems with a low enough time variability, or when using a feedback channel. Obviously, due to the estimation noise and the limited feedback capacity, the channel estimate is expected to be imperfect, *i.e.*, to have some error [7]. According to this, the actual channel can be represented as

$$\mathbf{H} = \widehat{\mathbf{H}} + \mathbf{\Delta}. \tag{20}$$

The optimum design in this situation corresponds to a robust approach, in which the error $\mathbf{\Delta}$ is taken into account explicitly [8].

There are different ways of obtaining robust designs depending on the approach taken to model the error [8, 9]. The Bayesian designs use a statistical model for the error, whereas the maximin approaches [10, 11] consider that the error, which is unknown, belongs to a predefined uncertainty region. In the following, a maximin design of a MIMO system is presented, where the objective is the optimization of the worst SNR for any possible error.

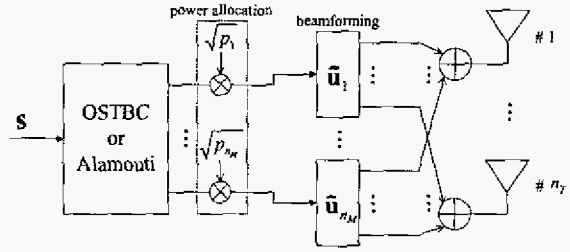

Fig. 3. Robust transmitter based on the combination of OSTBC and beamforming. With full OSTBC: $n_M = n_T$, and with Alamouti: $n_M = 2$.

5.1. Combining OSTBC and Beamforming

The generic signal model presented in Section 2 can also encompass a transmitter which is robust under the maximin philosophy. The proposed transmitter architecture is shown in Fig. 3, where $\widehat{\mathbf{u}}_i$ is the i-th eigenvector of $\widehat{\mathbf{H}}^H\widehat{\mathbf{H}}$ and $\{p_i\}$ represents a distribution of the transmit power among the outputs of the OSTBC, *i.e.*, among the estimated eigenmodes. Two possible choices can be given concerning the OSTBC: a full OSTBC, where the number of its outputs is equal to the number of transmit antennas ($n_M = n_T$), or an Alamouti code [4], where ($n_M = 2$), *i.e.*, only the two strongest eigenmodes are used. Note that, when the number of antennas is higher than 2, the symbol rate achieved by the full OSTBC is lower than one. Obviously, although the symbol rate is higher in the second scheme, the robustness capabilities are expected to be lower since the number of eigenmodes that can be used is also lower.

This transmitter architecture is a particular case of the generic signal model of Section 2. In this situation, the matrices $\{\mathcal{V}_n^H\}$ are given by the Hurwitz-Radon matrices defined for OSTBC, the diagonal matrix $\mathbf{P}^{\frac{1}{2}}$ is given by the power allocation ($\mathbf{P} = \mathrm{diag}(\{p_i\})$), which is time-independent, and the matrix $\mathbf{U}$ contain the n_M strongest eigenvectors of the channel estimate (depending on the number of outputs of the OSTBC), *i.e.*, $\mathbf{U} = \widehat{\mathbf{U}} = [\widehat{\mathbf{u}}_1 \cdots \widehat{\mathbf{u}}_{n_M}]$.

Since the considered block space-time codes are orthogonal, the ML detector simplifies to a linear receiver [2], where the BER is directly related to SNR, which is given by

$$\mathrm{SNR} = \frac{2E_s}{N_0}\,\mathrm{Tr}\left(\widehat{\mathbf{U}}^H(\widehat{\mathbf{H}}+\mathbf{\Delta})^H(\widehat{\mathbf{H}}+\mathbf{\Delta})\widehat{\mathbf{U}}\mathbf{P}\right).$$

The power allocation that maximizes the worst SNR for any possible error in the channel estimate can be found easily when the uncertainty region $\mathcal{R}$ for the error $\mathbf{\Delta} \in \mathcal{R}$ is convex. The original maximin problem, formulated in terms of the following two optimization stages:

$$\max_{\mathbf{P}} \min_{\mathbf{\Delta}\in\mathcal{R}} \quad \mathrm{SNR}(\mathbf{P},\mathbf{\Delta}) \tag{21}$$

$$\text{s.t.} \quad \textstyle\sum_i p_i = 1, \tag{22}$$

can be simplified to one simple convex optimization problem (see [12] for a complete proof of this statement):

$$\begin{aligned} \min_{t,\mathbf{\Delta}} \quad & t \\ \text{s.t.} \quad & t \geq \widehat{\mathbf{u}}_i^H(\widehat{\mathbf{H}}+\mathbf{\Delta})^H(\widehat{\mathbf{H}}+\mathbf{\Delta})\widehat{\mathbf{u}}_i, \quad i = 1,\ldots,n_M \\ & \mathbf{\Delta} \in \mathcal{R}. \end{aligned}$$

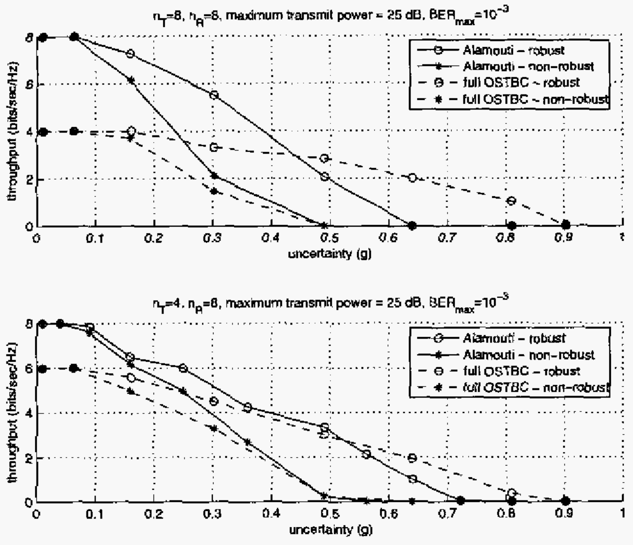

Fig. 4. Maximum transmission rate combining the robust design with adaptive modulation.

In the previous simplified convex problem, the n_M inequality constraints are applied to the estimated eigenmodes used at the transmitter, *i.e.*, the number of constraints is equal to n_T or 2, depending on the OSTBC. The robust power allocation, $\{p_i\}$, is shown to be equal to the optimum Lagrange multipliers associated to the inequality constraints, that can be calculated efficiently with existing convex software packages.

This robust design can be combined with adaptive modulation, so that the transmission rate is maximized subject to maximum power and to QoS constraints. These QoS constraints can be formulated as a maximum guaranteed BER for any possible error in the uncertainty region. In Fig. 4, a comparison among different schemes is given: the robust transmitter with full and Alamouti OSTBC, and the non-robust solution, in which only the maximum eigenmode is used, *i.e.*, $p_1 = 1$, $p_i = 0$, $\forall i > 1$. The maximum achieved transmission rate is represented as a function of the degree of uncertainty g assuming spherical uncertainty regions ($\mathcal{R} = \{\mathbf{\Delta} : \|\mathbf{\Delta}\|_F^2 \leq g\|\widehat{\mathbf{H}}\|_F^2\}$). As expected, when the uncertainty is low, robustness is not essential and, therefore, the scheme based on the Alamouti code provides a higher transmission rate due to using a full rate code. However, as the uncertainty increases, the solution based on a full code can attain a higher transmission rate, since it can achieve a higher degree of robustness based on the use of all the eigenmodes of the estimated channel.

5.1.1. Extension to QOSTBC with ML Detection

The previous scheme can be further extended to other architectures by considering QOSTBC instead of the orthogonal codes, providing a full symbol rate transmission. Note that when using a QOSTBC, the optimum ML detector is not longer a linear receiver. In such a situation, although the SNR obtained for the OSTBC case is not useful, the PEP associated to the optimum ML detector, assuming only one error in the decoding, is also dominated by an exponential term whose exponent is proportional to the SNR expression given previously. In other words, if a robust maximin design is to be applied to a transmitter using a QOSTBC and a receiver based on the ML criterion, the same power distribution as before should be used.

6. CONCLUSIONS

In this paper, a generic and flexible signal model has been presented, in which the tradeoff between the diversity and multiplexing gains can be implemented from a practical point of view. This transmitter scheme, composed by a time processing, a power allocation, and a space processing stages, encompasses very diverse situations concerning different degrees of knowledge of the CSI and also different detection schemes at the receiver. The PEP cost functions to be optimized have been deduced for ML detection combined with the assumption of having a perfect CSI, a correlation knowledge, and a noisy channel estimate. Finally, some of these combinations have been evaluated by means of simulations, including the comparison of a system designed using the knowledge of the long-term channel correlation matrix with V-BLAST, and also the comparison of a maximin robust transmitter with a non-robust beamforming solution.

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