Transmit Precoding for Receive Spatial Modulation Using Imperfect Channel Knowledge

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***Abstract*—In this paper, motivated by the relatively new concept of Spatial Modulation (SM), we are addressing the problem of Receive-Spatial Modulation (R-SM) under two cases of imperfect channel knowledge at the transmitter side. In the first case, we adopt a statistical model for the channel uncertainties, whereas in the second one a worst-case approach is followed and the channel uncertainties are expressed as a bounded set. Based on Zero-Forcing (ZF) precoding and using standard tools from optimization theory, we derive closed form solutions that turn out to be robust. Simulation results show that the proposed schemes have a performance close to the perfect channel knowledge scenario in low and mid-low SNRs. Furthermore, these designs can be applied to wide range of channels with different correlation states combined with a transmit power gain.**

I. Introduction

In the last years, we have been facing an increasing demand for wireless system throughput. This demand creates the need for spectrally efficient and reliable radio communication systems which will meet all the Quality of Services (QOS) standards combined with increased energy efficiency. The most promising way to fulfil this need is Multiple-Input Multiple-Output (MIMO) technology where multiple antennas are employed at the receiver and at the transmitter [1]. For example, the well known Vertical Bell Laboratories Layered Space-Time (V-BLAST) architecture achieves high data rates by using spatial multiplexing [2]. The main drawbacks that V-BLAST faces are the need for Inter-Antenna Synchronization (IAS) and Inter-Channel Interference (ICI) cancellation. As a result, the computational complexity of the receiver is high.

Recently, a new MIMO scheme has been proposed named Spatial Modulation (SM) [3], [4], [5]. SM avoids ICI by using only one antenna during transmission period to convey information, all the other antennas are not activated. With this method we not only avoid ICI but also gain a second mechanism of conveying information in addition to the classical M-ary Quadrature Amplitude Modulation (QAM) constellation. This second mechanism is the index of the transmitting antenna. Furthermore, SM does not require IAS. In addition, a special case of SM called Space Shift Keying (SSK) is proposed in [7]. Other SM-like schemes include [8] and [9], where both space and time dimension is used in order to form an extended concept of SSK. It is notable that SM has the capability to outperform other MIMO systems like V-BLAST [4]-[9].

In [10] a variation of SM is proposed called pre-processing aided spatial modulation. Using Channel State Information at the Transmitter side (CSIT) and precoding it is formed a Receive-Spatial Modulation (R-SM) scheme. Instead of activating only one antenna at the transmitter as SM, R-SM receives the transmitted signal at only one antenna at the receiver. Explicitly, the transmitter activates all of its antennas and using precoding targets at only one receive antenna. Thus, the index of the receiving antenna can be used by the transmitter as the additional mechanism of conveying information. In this work, we will use the name of R-SM for the pre-processing aided spatial modulation for reasons of simplicity.

Generally, precoding using CSIT on MIMO systems has been used either for receiver simplification and power consuming benefits for the case of downlink, or capacity gain. The key factor for precoding is the CSIT, which can be obtained either using low rate feedback from the receiver or using the reciprocity principle. In most cases, Channel Side Information at the Receiver (CSIR) is obtained using a training sequence and can be assumed to be accurate. In contrast, supplying the transmitter with accurate CSIT is a difficult task. Thus, the precoding designs should be robust under imperfect or partial CSIT. Generally, the design of robust precoders is divided in two categories. The first category models the channel uncertainties as Random Variables (RV) and solves the resulting optimization problem using a statistical approach [11]. In contrast, the second category makes the assumption that the channel uncertainty is a bounded set and a worst-case approach is followed [12].

In this work, we use Stochastic Robust Approximation (SRA) and Worst-Case Robust Approximation (WCRA) [15] in order to form a R-SM scheme similar to [10] that remains robust under imperfect CSIT. The motivation to base our design on SRA and WCRA is [13], where these optimization methods are used in MIMO precoding under frequency selec-

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tive channel and imperfect CSIT. Similar to [13], the resulting solutions have a closed form and appears to have competitive performance under different correlation scenarios combined with a transmit power gain. Finally, R-SM has a complexity gain at the receiver compared to SM, due to the simplification of the detector.

The rest of the paper is organized as follows: section II presents the system model and the problem statement. The proposed precoding schemes are presented in section III and their performances are shown in section IV. Finally, we conclude the paper in section V.

Notation: In the following, lowercase bold letters denote vectors and uppercase bolt letters denote matrices. $(\cdot)^T$, $(\cdot)^H$, tr$(\cdot)$ and $\mathbf{A}^{1/2}$ denotes transpose, Hermitian transpose, matrix trace and the square root of $\mathbf{A}$, respectively. E$[\cdot]$ is the mean value of a RV. Finally, a complex Gaussian distribution with mean m and variance σ_C^2 is represented as $CN(m, \sigma_C^2)$, where its real and imaginary part are independent and identically distributed (i.i.d.) Gaussian RV with distribution $N(m, \frac{\sigma_C^2}{2})$.

II. System Model and Problem Statement

A. System Model

A MIMO frequency flat system is considered with N_t transmit antennas and N_r receive antennas, with $N_t \geq N_r$. When a precoder is applied at the transmitter, the system equation at every symbol period is given as

$$\mathbf{y} = \mathbf{HPx} + \mathbf{w}, \tag{1}$$

where $\mathbf{H}$ is a $N_r \times N_t$ matrix representing the wireless MIMO channel, $\mathbf{P}$ is the precoding matrix of size $N_t \times N_r$, $\mathbf{x}$ is the $N_r \times 1$ transmitted signal vector and $\mathbf{w} \in \mathcal{C}^{N_r}$ is the zero mean vector of additive complex Gaussian noise with i.i.d. elements and variance of σ_w^2. A strategy for designing the precoding matrix $\mathbf{P}$, that aims at the elimination of Inter-Channel Interference (ICI), is the ZF method. In this case the precoding matrix is just the pseudo-inverse of the channel matrix $\mathbf{H}$. Using Singular Value Decomposition (SVD) the precoding matrix is equal to $\mathbf{P} = \mathbf{V\Sigma}^{-1}\mathbf{U}^H$ (where $\mathbf{H} = \mathbf{U\Sigma V}^H$).

In this paper, we use ZF in order to form a R-SM system as in [10]. This is done by using the precoding matrix $\mathbf{P}$ to eliminate the effect of MIMO channel and the appropriate formation of the transmitted symbol $\mathbf{x}$.

The system model we use is the same as in [10]. Let us assume that the transmitted vector $\mathbf{x}$ has the form of $\mathbf{x} = \mathbf{e}_i s_k$, where the vector $\mathbf{e}_i = [0, ..., 0, 1, 0, ...0]^T$, $i = 1, ..., N_r$, has length of N_r and all of its elements are zero except from the one at the i-th position which has the value of 1. $s_k \in \{s_1, ..., s_M\}$ represents the transmitted symbol selected from a constellation like M-QAM. Furthermore, if we assume perfect channel knowledge at the transmitter and the ZF precoding matrix as described before, the system equation has the following form:

$$\mathbf{y} = \underbrace{\mathbf{HP}}_{\mathbf{I}_{N_r,N_r}} \mathbf{e}_i s_k + \mathbf{w}. \tag{2}$$

In this scenario, the observation at each of the N_r receive antennas is given as

$$\begin{array}{ll} y_j = s_k + w_j, & j = i, \\ y_j = w_j, & j \neq i, \end{array} \tag{3}$$

$j = 1, ..., N_r$, which means that only one antenna receives the transmitted symbol s_k degraded from Gaussian noise and all the other receive antennas are facing only Gaussian noise.

If we configure our system so that the number of receive antennas is a power of 2, it is possible to map $k_1 = \log_2 N_r$ bits on the index of the i-th receive antenna while $k_2 = \log_2 M$ bits are mapped on the transmitted symbol s_k. In this way, we have formed the same R-SM system as in [10] which conveys $k = k_1 + k_2$ bits every symbol period.

Because of this, it is straightforward to use a ML detector (similar to the one used in [10])

$$(\hat{i}, \hat{s}_k) = \arg\min_{i, s_k} \|\mathbf{y} - \mathbf{e}_i s_k\|_2^2, \tag{4}$$

which minimizes the Euclidean norm over all possibles combinations of $i = 1, ..., N_r$ and $s_k \in \{s_1, ..., s_{k_2}\}$. As we can see from (4), the detector does not requires any channel knowledge in order to detect the received $(\hat{i}, \hat{s}_k)$.

B. Complexity of ML Detection

Using the zero structure of $\mathbf{e}_i$, the ML detector can be rewritten as

$$(\hat{i}, \hat{s}_k) = \arg\min_{i, s_k} |s_k|^2 - 2\text{Re}\{y_i^* s_k\}. \tag{5}$$

If we consider that $|s_k|^2$ is pre-computed and stored at the memory, the evaluation of (5) requires $N_r M$ complex multiplications, $N_r M$ real multiplications and $N_r M$ real additions.

Especially, for the case of Receive-Space Shift Keying (R-SSK), where $s_k = 1$ and only the index of the receive antenna is used as the mechanism of conveying information, the ML detector can be further simplified as

$$\hat{i} = \arg\max_{i} \text{Re}\{y_i\}, \tag{6}$$

which has the complexity of only N_r real comparisons.

Hence, if we compare the complexity of R-SM detection with the one of SM's optimal detector [6], which has the complexity of $2N_tN_rM - 2N_tM$ complex additions and $2N_tN_rM + N_tM$ complex multiplications plus the complexity of N_tM real additions and N_tM real multiplications, we can see a notable complexity simplification.

C. Objective Function

In this subsection, we present the objective function that we use in the next section to design the precoding matrix $\mathbf{P}$, using partial channel knowledge.

In order to design the ZF precoding matrix of (1) we can compute the pseudo-inverse matrix of $\mathbf{H}$ using SVD. An alternative way is to solve the following N_r minimization problems

$$\min_{\mathbf{p}_l} \|\mathbf{Hp}_l - \mathbf{e}_l\|_2^2, \ \forall l = 1, ..., N_r, \tag{7}$$

where $\mathbf{p}_l$ is the l-th column of the precoding matrix $\mathbf{P} = [\mathbf{p}_1, ..., \mathbf{p}_{N_r}]$ and $\mathbf{e}_l$ is the l-th column of the identity matrix $\mathbf{I}_{N_r,N_r} = [\mathbf{e}_1, ..., \mathbf{e}_{N_r}]$. When perfect channel knowledge is available at the transmitter, the solution of the previous minimization problems is trivial. Our target in the next section is to solve the minimization problems of (7) using imperfect channel knowledge.

III. Precoding Using Imperfect Channel Knowledge

A. Imperfect Channel Knowledge at the Transmitter

In realistic scenarios providing the precoder with perfect channel knowledge is almost impossible. A common assumption is to model MIMO channel as $\mathbf{H} = \overline{\mathbf{H}} + \widetilde{\mathbf{H}}$ [14]. In this model, $\overline{\mathbf{H}}$ represents the long term channel evolution which can be accurately acquired by the transmitter. Whereas, $\widetilde{\mathbf{H}}$ denotes the channel for which only some kind of statistical or worst-case knowledge is considered possible. Usually $\widetilde{\mathbf{H}}$ is modeled as Zero Mean Circular Symmetric Complex Gaussian Matrix (ZMCSCG) with i.i.d. elements of variance $\sigma^2_{\widetilde{\mathbf{H}}}$.

In the following two subsections, the minimization problems of (7) are solved using two methods from optimization theory, SRA and WCRA [15]. SRA uses a statistical approach and solves the minimization problems over their expectation. The partial knowledge that SRA employs is the matrix $\overline{\mathbf{H}}$ and the matrix $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$. On the other hand, WCRA is a worst-case approach and solves the same minimization problems assuming full knowledge of $\overline{\mathbf{H}}$ and α which is defined as $\|\widetilde{\mathbf{H}}\|_2 \leq \alpha$.

Both designs are using the long term channel evolution $\overline{\mathbf{H}}$ which is a slow varying characteristic. Furthermore, SRA uses $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$ which is also a slow varying statistic [16] and in some cases, as it is described in the next subsection, has a structured form. On the other hand, WCRA uses the additional information of α which is just a scalar. Thus, our schemes enjoy the merit of reduced feedback from the receiver.

B. ZF-like Precoding Based on SRA

SRA solves the minimization problems of (7) over its expectation [15],

$$\min_{\mathbf{p}_l} \mathrm{E}\left[\|\mathbf{H}\mathbf{p}_l - \mathbf{e}_l\|_2^2\right], \ \forall l = 1, ..., N_r. \tag{8}$$

If we use the fact that the MIMO channel can be written as $\mathbf{H} = \overline{\mathbf{H}} + \widetilde{\mathbf{H}}$, where $\mathrm{E}[\widetilde{\mathbf{H}}] = \mathbf{0}$, (8) can be reformulated as

$$\min_{\mathbf{p}_l} \left\{ \|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l\|_2^2 + \mathbf{p}_l^H \mathrm{E}\left[\widetilde{\mathbf{H}}^H\widetilde{\mathbf{H}}\right]\mathbf{p}_l \right\}, \tag{9}$$

which is obviously a convex optimization problem (sum of quadratic functions) and can be solved by setting its gradient equal to zero. The analytical solution is formed as

$$\mathbf{p}_l = \left[\overline{\mathbf{H}}^H\overline{\mathbf{H}} + \mathrm{E}\left[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}\right]\right]^{-1}\overline{\mathbf{H}}^H\mathbf{e}_l. \tag{10}$$

Inspecting (10), we can see that the computation of $\mathbf{p}_l$ requires the knowledge of $\overline{\mathbf{H}}$ and the matrix $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$ at the transmitter. Finally, it is worth mentioning that in many scenarios $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$ has a structured form, which can simplify the hardware design and reduce the feedback from the receiver. For example, in the case of the uncorrelated channel $\mathbf{H}$, with $\tilde{h}_{i,j} \in CN(0, \sigma_{\widetilde{\mathbf{H}}})$, $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$ is the identity matrix multiplied by $N_r\sigma^2_{\widetilde{\mathbf{H}}}$.

C. ZF-like Precoding Based on WCRA

In this subsection we use WCRA [15] in order to design the precoding matrix $\mathbf{P}$. We define the non-empty and bounded set $\Phi \subseteq \mathbf{C}^{N_r,N_t}$ which represents all the possible values of the channel matrix $\mathbf{H}$. Given a feasible precoding vector $\mathbf{p}_l$ the worst case error can be formulated as $e_{\mathrm{wc}}(\mathbf{p}_l) = \sup\left[\|\mathbf{H}\mathbf{p}_l - \mathbf{e}_l\|_2 | \mathbf{H} \in \Phi\right]$. Our target here is to design a precoder which minimizes the $e_{\mathrm{wc}}(\mathbf{p}_l)$. In this case, the minimization problem can be formulated as

$$\min_{\mathbf{p}_l} \sup\left[\|\mathbf{H}\mathbf{p}_l - \mathbf{e}_l\|_2 | \mathbf{H} \in \Phi\right] \ \forall l = 1, ..., N_r. \tag{11}$$

One possible solution to the minimization problem of (11) can be reached using the Norm Bound Error (NBE) [15]. In this method, the uncertainty of $\widetilde{\mathbf{H}}$ is considered within a norm ball of radius α and the set Φ is written as $\Phi = \left\{\mathbf{H} = \overline{\mathbf{H}} + \widetilde{\mathbf{H}} | \|\widetilde{\mathbf{H}}\|_2 \leq \alpha\right\}$, where $\alpha > 0$. Let $e_{\mathrm{wc}}^{\mathrm{NBE}}(\mathbf{p}_l) = \sup\{\|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l + \widetilde{\mathbf{H}}\mathbf{p}_l\|_2| \ \|\widetilde{\mathbf{H}}\|_2 \leq \alpha\}$ be the worst-case error given the precoding vector $\mathbf{p}_l$. It is easily to show [15] that $e_{\mathrm{wc}}^{\mathrm{NBE}}(\mathbf{p}_l)$ is equal to $e_{\mathrm{wc}}^{\mathrm{NBE}}(\mathbf{p}_l) = \|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l\|_2 + \alpha\|\mathbf{p}_l\|_2$ and it is attained for $\widetilde{\mathbf{H}} = \alpha\mathbf{u}\mathbf{v}^H$ where $\mathbf{u} = \frac{\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l}{\|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l\|_2}$ and $\mathbf{v} = \frac{\mathbf{p}_l}{\|\mathbf{p}_l\|_2}$, given that $\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l \neq 0$ and $\mathbf{p}_l \neq \mathbf{0}$ [15]. Thus the minimization problem of (11) can be reformulated as $\min_{\mathbf{p}_l} \|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l\|_2 + \alpha\|\mathbf{p}_l\|_2$, which can be re-written in a Tikhonov Regularization form for some value of β [15]

$$\min_{\mathbf{p}_l} \|\overline{\mathbf{H}}\mathbf{p}_l - \mathbf{e}_l\|_2^2 + \beta\|\mathbf{p}_l\|_2^2. \tag{12}$$

Again, the minimization problem of (12) is convex because $e_{\mathrm{wc}}^{\mathrm{NBE}}(\mathbf{p}_l)$ is the sum of quadratic functions. Thus the solution can be reached using the gradient condition

$$\mathbf{p}_l = \left[\overline{\mathbf{H}}^H\overline{\mathbf{H}} + \beta\mathbf{I}\right]^{-1}\overline{\mathbf{H}}^H\mathbf{e}_l \ \forall l = 1, ..., N_r. \tag{13}$$

D. Precoding in the Presence of Transmit and Receive Space Correlations

It is clear that both problems of (9) and (12) are different forms of Tikhonov Regularization. A valuable property of Tikhonov Regularization theory is that it does not pose any rank restriction on the involved matrices $\overline{\mathbf{H}}$ and $\widetilde{\mathbf{H}}$ as long as the matrices $\overline{\mathbf{H}}^H\overline{\mathbf{H}} + \mathrm{E}\left[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}\right]$ and $\overline{\mathbf{H}}^H\overline{\mathbf{H}} + \beta\mathbf{I}$ are positive definite [15]. Thus, our analytical solutions of (10) and (13) enjoy the additional merit of being applicable to spatially correlated channels.

In this paper we employ the Kronecker correlation model [17]. Under this correlation model the MIMO channel can be rewritten as

$$\mathbf{H} = \mathbf{R}_R^{1/2}\mathbf{H}_w(\mathbf{R}_T^{1/2})^T, \tag{14}$$

where $\mathbf{H}_w$ is a ZMCSCG matrix with i.i.d elements and variance of $\sigma^2_{\mathbf{H}_w} = 1$. The matrices $\mathbf{R}_T$ and $\mathbf{R}_R$ represent the transmit spatial correlation matrix and the receive spatial correlation matrix, respectively. Usually, the entries of the spatial correlation matrices $\mathbf{R}_R$ and $\mathbf{R}_T$ are generated using an exponential model with $R_T(i,j) = \rho_t^{|i-j|}$ and

$R_R(i,j) = \rho_r^{|i-j|}$, where $0 \leq \rho_t, \rho_r \leq 1$. Values of ρ_t and ρ_r close to 0 mean low correlation, whereas values closely to 1 mean high correlations.

If we combine the Kronecker correlation model of (14) with the model of partial channel knowledge described at subsection III-A the MIMO channel can be written as

$$\mathbf{H} = \mathbf{R}_R^{1/2}\left(\overline{\mathbf{H}} + \widetilde{\mathbf{H}}\right)(\mathbf{R}_T^{1/2})^T,$$

where the matrix $\mathbf{R}_R^{1/2}\overline{\mathbf{H}}(\mathbf{R}_T^{1/2})^T$ represents the full known part of the channel and the matrix $\mathbf{R}_R^{1/2}\widetilde{\mathbf{H}}(\mathbf{R}_T^{1/2})^T$ represents the partial known part of the channel.

Inspecting (10) and (13) we can see that only the analytical form of the design based on SRA is affected by the correlated channel. This is because SRA requires the computation of $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}]$ which can be shown that it takes the form of $\mathrm{E}[\widetilde{\mathbf{H}}^{\mathrm{H}}\widetilde{\mathbf{H}}] = \sigma^2_{\widetilde{\mathbf{H}}}\mathrm{tr}(\mathbf{R}_{\mathrm{R}})\mathbf{R}_{\mathrm{T}}^{1/2}(\mathbf{R}_{\mathrm{T}}^{1/2})^{\mathrm{T}}$.

E. Transmit Power Analysis

ZF precoding has the disadvantage of increasing the transmit power significantly in order to overcome deep fades. In real systems, this is something that should be avoided. A common solution is to impose a transmit power constraint $\|\mathbf{P}\|_F^2 = \sum_{l=1}^{N_r}\|\mathbf{p}_l\|_2^2 \leq p_t$, where $\|\cdot\|_F^2$ is the Frobenius norm and p_t is the transmit power constraint.

SRA and WCRA are forms of Tikhonov Regularization problem. Tikhonov Regularization theory is the most common form of regularization that penalize big values of $\|\mathbf{p}_l\|_2^2$ [15]. As a consequence, our precoding schemes achieve solutions with small values of $\|\mathbf{p}_l\|_2^2$ depending on the value of channel uncertainty $\sigma^2_{\widetilde{\mathbf{H}}}$. In SRA and WCRA, increased values of $\sigma^2_{\widetilde{\mathbf{H}}}$ produce precoding vectors $\mathbf{p}_l$ with small Euclidean norm [15]. Hence, our systems indirectly satisfy the previous transmit power constraint . In the next section, we study the behavior of the transmitted power of our schemes using numerical simulations.

IV. Simulation Results

In this section, we present simulation results that demonstrate the performance of the proposed precoding methods. The system configuration assumes $N_t = 4$ transmit antennas and $N_r = 2$ receive antennas. Furthermore, every transmitted symbol $s_k \in \{s_1, ..., s_M\}$ is selected from a 4-QAM constellation with average transmitted power equal to 1. As a consequence, every symbol period a total of $k_1 + k_2 = 3$ bits are transmitted. The MIMO channel $\mathbf{H}$ is generated as the sum of two ZMCSCG matrices ($\overline{\mathbf{H}}$ and $\widetilde{\mathbf{H}}$) whose elements follow i.i.d. $\bar{h}_{i,j} \sim CN(0, 0.99)$ and $\tilde{h}_{i,j} \sim CN(0, 0.01)$ distribution such that $h_{i,j} \sim CN(0,1)$. The entries of $\overline{\mathbf{H}}$ are refreshed every 100 symbol periods whereas the entries of $\widetilde{\mathbf{H}}$ are refreshed every symbol period. In addition, the entries of the spatial correlation matrices $\mathbf{R}_R$ and $\mathbf{R}_T$ are generated using an exponential model with $\rho_t = \rho_r = \rho$, where ρ takes values from the set $\{0.1, 0.3, 0.5, 0.9\}$. Finally, Signal-to-Noise Ratio (SNR) per bit is defined as $\mathrm{SNR}_{\text{bit,dB}} = 10\log\frac{\mathrm{tr}(\mathbf{P}^H\mathbf{P})\sigma_s^2/[N_r(k_1+k_2)]}{\sigma_w^2}$, where the variance of the transmitted symbol s_k is equal to $\sigma_s^2 = 1$[1].

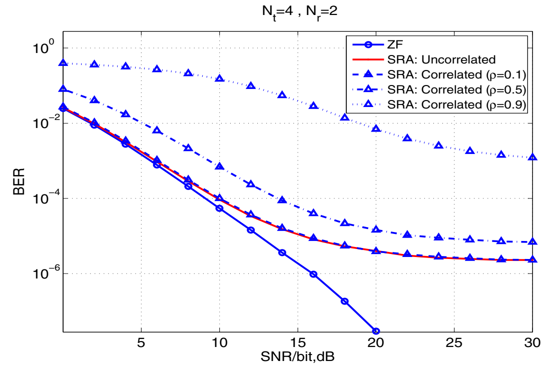

Fig. 1. BER versus SNR for SRA precoder for $\sigma^2_{\widetilde{\mathbf{H}}} = 0.01$

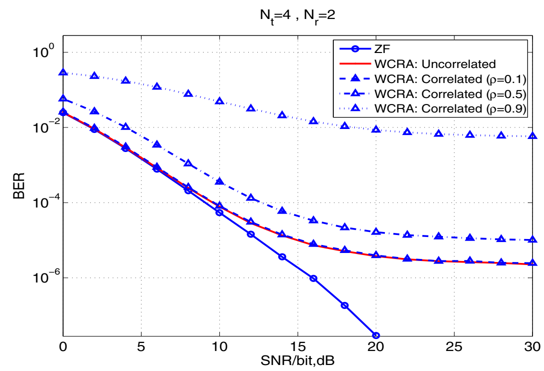

Fig. 2. BER versus SNR for WCRA precoder for $\sigma^2_{\widetilde{\mathbf{H}}} = 0.01$

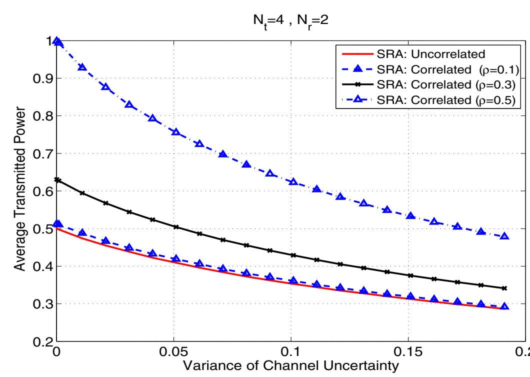

Fig. 3. Average transmitted power versus $\sigma^2_{\widetilde{\mathbf{H}}}$ for SRA precoder

The BER performance of the two proposed schemes is presented in Fig. 1 and Fig. 2 for the cases of uncorrelated and correlated channel. Especially, for the correlated case we assume spatial correlated channels with low correlation ($\rho = 0.1$), with intermediate correlation ($\rho = 0.5$) and with

[1]The transmitted signal power is divided by N_t because every symbol period only one column of the precoder $\mathbf{P}$ is activated.

high correlation ($\rho = 0.9$). In addition, in the same figures we present the BER performance of the ZF precoder under full channel knowledge. In order to have a fair comparison between the proposed designs and the ZF precoder, the channel that the ZF precoder faces has the same statistical characteristics as our schemes.

Inspecting Fig. 1 and Fig. 2, we can see that under the uncorrelated channel the proposed designs have similar performance to the ZF precoder at the low and mid-low SNRs. Clearly, at the mid-high and high SNRs ZF precoder outperforms our scheme. In these SNRs the main degradation factor for our schemes is the residual ICI, ZF precoder has better performance due to accurate channel knowledge. Furthermore, it seems that our systems, due to the residual ICI, face a bottleneck at 20 dB, where higher values of SNR do not affect their performance rapidly. The way that correlation affects the performance of our schemes can be divided in two cases. In the low correlation case ($\rho = 0.1$), we can see that there is no performance difference between the uncorrelated and correlated channel. In the second case, as correlation increases the performance of our systems deteriorates. For the intermediate level of correlation ($\rho = 0.5$) there is a need of 10 dB in order to achieve a BER performance of practical interest. For the high level of correlation ($\rho = 0.9$) this need is increased to 22 dB.

Fig. 3 and Fig. 4 present the average transmitted power versus the variance of channel uncertainty for different correlation scenarios. As it can be seen from these figures, as the channel uncertainty $\sigma^2_{\tilde{\mathbf{H}}}$ increases the transmitted power decreases. This means that when the channel knowledge at the transmitter becomes more inaccurate, the transmitter reduces the transmitted power in order to avoid further degradation. Although increasing channel correlation requires higher average transmit power, the proposed new scheme requires for low ($\rho = 0.1$) correlation almost the same power as the uncorrelated case. Finally, the results offer another indirect comparison between our precoding designs and the ZF precoder using perfect channel knowledge. When $\sigma^2_{\tilde{\mathbf{H}}} \to 0$, SRA and WCRA reduce to ZF precoder with full channel knowledge. Clearly, there is a significant power gain for our schemes compared to ZF precoder especially for higher values of $\sigma^2_{\tilde{\mathbf{H}}}$.

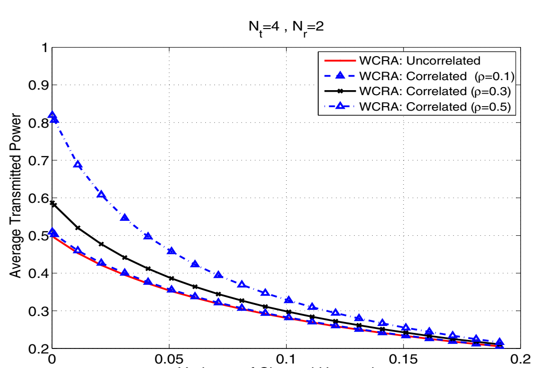

Fig. 4. Average transmitted power versus $\sigma^2_{\tilde{\mathbf{H}}}$ for WCRA precoder

V. Conclusions

This paper treats the problem of R-SM under imperfect CSIT. Using two different forms of imperfect channel knowledge we formulate two precoding schemes for R-SM. By expressing the resultant optimization problems into Tikhonov regularization forms we achieve a transmit power gain and the capability to treat correlated channels. In this way, we propose two R-SM schemes that make use of practical forms of CSIT and have low detection complexity at the receiver. Our systems turns to have a well-behaved BER performance under different channel correlation states combined with a transmit power gain. Finally, the simulation results confirm the advantages of the new schemes.

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