EXPLOITING THE STRUCTURE OF OSTBC'S TO IMPROVE THE ROBUSTNESS OF WORST-CASE OPTIMIZATION BASED LINEAR MULTI-USER MIMO RECEIVERS

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ABSTRACT

In this paper, we improve the performance of robust linear receivers for multiuser multiple-input multiple-output (MIMO) wireless systems by exploiting the inherent structure of orthogonal space-time block codes (OSTBCs). This particular structure results in a worst-case optimization problem with structured uncertainty set. Exploiting this structure, an improved robust linear receiver with a combination of fixed diagonal loading and adaptive non-diagonal loading of the data covariance matrix is obtained.

1. INTRODUCTION

Space-time coding is a powerful approach to exploit spatial diversity and combat fading in MIMO wireless communication systems. OSTBCs [1]-[2] represent an attractive class of space-time coding techniques because they enjoy full diversity and very low complexity of the maximum likelihood (ML) decoding in the point-to-point MIMO case.

In the multiuser MIMO case, the ML decoder becomes prohibitively expensive. Motivated by this fact, several authors proposed suboptimal linear receiver techniques for space-time coded multiuser MIMO systems [3]-[6]. Unfortunately, all these techniques assume that the exact channel state information (CSI) is available at the receiver. In practice, this condition can be hardly met because of a limited/outdated training, and the effects of multi-access interference (MAI) and noise.

Recently, a linear receiver based on the worst-case design has been proposed in [7] to improve the robustness of symbol detection against erroneous CSI. It has been shown that this robust receiver provides substantial performance improvements with respect to the linear receivers of [6]. However, the inherent structure of OSTBCs is not exploited in [7] and, as a result, the design of [7] appears to be overly conservative.

In this paper, we propose a new robust linear receiver which, similarly to the receiver of [7], resorts to the worst-case design, but additionally exploits the specific structure of OSTBCs. This results in a worst-case optimization problem with structured uncertainty set. Using this approach, an improved robust linear receiver is obtained. The proposed receiver amounts to a combination of *fixed diagonal loading* and *adaptive non-diagonal loading* of the data covariance matrix.

Simulation results validate a substantially improved performance of the proposed linear receiver as compared to the receiver of [7].

2. BACKGROUND

Let us consider an uplink multiuser MIMO communication system. The transmitters are assumed to have the same number of transmitting antennas and to encode information-bearing symbols using the same OSTBC[1]. The received signal is given by

$$\mathbf{Y} = \sum_{p=1}^{P} \mathbf{X}_p \mathbf{H}_p + \mathbf{V} \tag{1}$$

where $\mathbf{Y} \triangleq [\mathbf{y}^T(1) \ \cdots \ \mathbf{y}^T(T)]^T$, $\mathbf{X}_p \triangleq [\mathbf{x}_p^T(1) \ \cdots \ \mathbf{x}_p^T(T)]^T$, and $\mathbf{V} \triangleq [\mathbf{v}^T(1) \ \cdots \ \mathbf{v}^T(T)]^T$ are the matrices of the received signals, transmitted signals of the pth transmitter, and noise, respectively, $\mathbf{H}_p$ is the $N \times M$ complex channel matrix between the pth transmitter and the receiver, P is the number of transmitters, T is the block length, and $(\cdot)^T$ denotes the transpose. Here, $\mathbf{y}(t) \triangleq [y_1(t) \ \cdots \ y_M(t)]$, $\mathbf{x}_p(t) \triangleq [x_{p,1}(t) \ \cdots \ x_{p,N}(t)]$, and $\mathbf{v}(t) \triangleq [v_1(t) \ \cdots \ v_M(t)]$ are the complex row vectors of the received signal, transmitted signal of the pth user, and noise, respectively. Each of the information-bearing symbols is assumed to be drawn from a constant modulus constellation.

We denote complex information-bearing symbols of the pth user prior to space-time encoding as $s_{p,1}, s_{p,2}, \ldots, s_{p,K}$. It can be readily verified that the matrix $\mathbf{X}(\mathbf{s}_p)$ can be written as [6]-[7]

$$\mathbf{X}(\mathbf{s}_p) = \sum_{k=1}^{K} \left(\mathbf{C}_k \mathrm{Re}\{s_{p,k}\} + \mathbf{D}_k \mathrm{Im}\{s_{p,k}\}\right) \tag{2}$$

where $\mathbf{C}_k \triangleq \mathbf{X}(\mathbf{e}_k)$, $\mathbf{D}_k \triangleq \mathbf{X}(j\mathbf{e}_k)$, $j = \sqrt{-1}$ and $\mathbf{e}_k$ is the $K \times 1$ vector having one in the kth position and zeros elsewhere. Using (2), one can rewrite (1) as [6]

$$\underline{\mathbf{Y}} = \sum_{p=1}^{P} \mathbf{A}(\mathbf{H}_p)\, \underline{\mathbf{s}_p} + \underline{\mathbf{V}} \tag{3}$$

where the "underline" operator for any matrix $\mathbf{P}$ is defined as

$$\underline{\mathbf{P}} \triangleq \begin{bmatrix} \mathrm{vec}\{\mathrm{Re}(\mathbf{P})\} \\ \mathrm{vec}\{\mathrm{Im}(\mathbf{P})\} \end{bmatrix} \tag{4}$$

and $\mathrm{vec}\{\cdot\}$ is the vectorization operator stacking all columns of a matrix on top of each other. Here, the $2MT \times 2K$ real matrix $\mathbf{A}(\mathbf{H}_p)$ is defined as [6]

$$\begin{aligned} \mathbf{A}(\mathbf{H}_p) &= [\underline{\mathbf{C}_1\mathbf{H}_p} \cdots \underline{\mathbf{C}_K\mathbf{H}_p}\ \underline{\mathbf{D}_1\mathbf{H}_p} \cdots \underline{\mathbf{D}_K\mathbf{H}_p}] \\ &\triangleq [\mathbf{a}_1(\mathbf{H}_p) \cdots \mathbf{a}_{2K}(\mathbf{H}_p)] \end{aligned}$$

[1]These assumptions are only needed for notational simplicity and can be relaxed, see [6].

Using the array-processing-type model (3) and assuming that the first transmitter is the transmitter-of-interest, we can express the output vector of a linear receiver as [6]-[7]

$$\underline{\hat{\mathbf{s}}_1} = \mathbf{W}^T \underline{\mathbf{Y}} \tag{5}$$

where $\mathbf{W} = [\mathbf{w}_1\, \mathbf{w}_2 \cdots \mathbf{w}_{2K}]$ is the $2MT \times 2K$ real matrix of the receiver coefficients and $\underline{\hat{\mathbf{s}}_1}$ is the estimate of the vector $\underline{\mathbf{s}_1}$ at the receiver output. The vector $\mathbf{w}_k$ can be interpreted as the weight vector for the kth entry of $\underline{\mathbf{s}_1}$.

Given the matrix $\mathbf{W}$, the estimate of the vector of information-bearing symbols of the transmitter-of-interest can be computed as $\hat{\mathbf{s}}_1 = [\mathbf{I}_K \; j\mathbf{I}_K]\, \underline{\hat{\mathbf{s}}_1}$.

The similarity of the vectorized multiuser MIMO model (3) and models used in array processing gives an opportunity to design the matrix $\mathbf{W}$ using minimum variance (MV) principle. In particular, in [6] it has been proposed to estimate each entry of $\underline{\mathbf{s}_1}$ by minimizing the receiver output power while preserving a unity gain for this particular entry of $\underline{\mathbf{s}_1}$, that is,

$$\min_{\mathbf{w}_k} \mathbf{w}_k^T \hat{\mathbf{R}} \mathbf{w}_k \;\; \text{s.t.} \;\; \mathbf{a}_k^T(\mathbf{H}_1)\mathbf{w}_k = 1 \;\; \text{for all} \;\; k = 1, \ldots, 2K \tag{6}$$

where $\hat{\mathbf{R}} = \frac{1}{J}\sum_{i=1}^{J} \underline{\mathbf{Y}_i}\, \underline{\mathbf{Y}_i}^T$ is the sample estimate of the $2MT \times 2MT$ covariance matrix $\mathbf{R} \triangleq \mathrm{E}\{\underline{\mathbf{Y}}\, \underline{\mathbf{Y}}^T\}$ of the vectorized data (3) , $\mathbf{Y}_i$ is the ith received data block, J is the number of data blocks available, and $\mathrm{E}\{\cdot\}$ denotes the statistical expectation.

The solution to (6) is given by [6]

$$\mathbf{w}_{\mathrm{MV},k} = \frac{\hat{\mathbf{R}}^{-1}\mathbf{a}_k(\mathbf{H}_1)}{\mathbf{a}_k^T(\mathbf{H}_1)\hat{\mathbf{R}}^{-1}\mathbf{a}_k(\mathbf{H}_1)}, \quad k = 1, \ldots, 2K \tag{7}$$

To improve the performance of (7) in the case of small sample size, a fixed diagonal loading (DL) has been used in [6]. The DL-based modification of the MV receiver (7) can be written as

$$\mathbf{w}_{\mathrm{DLMV},k} = \frac{(\hat{\mathbf{R}} + \gamma\mathbf{I}_{2MT})^{-1}\mathbf{a}_k(\mathbf{H}_1)}{\mathbf{a}_k^T(\mathbf{H}_1)(\hat{\mathbf{R}} + \gamma\mathbf{I}_{2MT})^{-1}\mathbf{a}_k(\mathbf{H}_1)}, k = 1, \ldots, 2K \tag{8}$$

It can be seen that in the MV receiver design (8), the CSI knowledge is required. However, in practice it is unrealistic to have the *exact* CSI at the receiver. Therefore, the performance of the MV receiver (8) may degrade significantly due to CSI errors [7]. To provide robustness against CSI errors, a modification of the MV receiver (7) based on the worst-case performance optimization has been proposed in [7]. The weight vector $\mathbf{w}_k$ of the latter robust receiver has been obtained in [7] as a solution to the following constrained optimization problem:

$$\min_{\mathbf{w}_k} \mathbf{w}_k^T \hat{\mathbf{R}} \mathbf{w}_k \quad \text{s.t.} \quad \min_{\|\mathbf{e}_k\| \le \epsilon} \mathbf{w}_k^T(\mathbf{a}_k(\hat{\mathbf{H}}_1) + \mathbf{e}_k) \ge 1 \tag{9}$$

where $\mathbf{e}_k = \mathbf{a}_k(\mathbf{H}_1) - \mathbf{a}_k(\hat{\mathbf{H}}_1)$ is the vector of mismatch between the true $\mathbf{a}_k(\mathbf{H}_1)$ and its presumed value $\mathbf{a}_k(\hat{\mathbf{H}}_1)$, and ϵ is a known constant which upper-bounds the Frobenius norm of the CSI error as $\|\mathbf{H}_1 - \hat{\mathbf{H}}_1\| \le \epsilon$. Using the Lagrange multiplier method, the solution to (9) can be obtained from the equation [7]

$$2\hat{\mathbf{R}}\mathbf{w}_k + \mu\epsilon\mathbf{w}_k/\|\mathbf{w}_k\| = \mu\mathbf{a}_k(\hat{\mathbf{H}}_1) \tag{10}$$

where μ is the unknown Lagrange multiplier. To get around the problem of computing μ, one can use the fact that each of the information-bearing symbols is drawn from a constant modulus constellation. In this case, each $\mathbf{w}_k$ can be arbitrarily rescaled without affecting the performance of a linear receiver. Using this fact and applying the rescaling $\mathbf{w}_k := 2\mathbf{w}_k/\mu$, equation (10) can be rewritten as [7]

$$\mathbf{w}_k = \left(\hat{\mathbf{R}} + \frac{\epsilon}{\|\mathbf{w}_k\|}\mathbf{I}_{2MT}\right)^{-1}\mathbf{a}_k(\hat{\mathbf{H}}_1) \tag{11}$$

where the term $(\epsilon/\|\mathbf{w}_k\|)\mathbf{I}_{2MT}$ can be viewed as an *adaptive diagonal loading* of the matrix $\hat{\mathbf{R}}$. To solve (11), a simple Newton-type technique of [9] can be used, see also [7].

A serious shortcoming of the design of (11) is that the inherent structure of the mismatch vector $\mathbf{e}_k$ is ignored in (9). Therefore, the worst-case performance optimization in (9) may lead to an overly conservative design, which can cause unnecessary degradation of the receiver performance.

3. EXPLOITING THE STRUCTURE OF UNCERTAINTY IN ROBUST RECEIVER DESIGN

In this section, we develop a new modification of the approach (6) which is robust against imperfect CSI at the receiver and which, in contrast to (11), exploits the inherent OSTBC-related structure of the mismatch vector $\mathbf{e}_k$.

First, let us obtain a relationship between $\mathbf{a}_k(\mathbf{H}_1)$ and $\mathbf{H}_1$ explicitly by using the structure of OSTBCs. Let

$$\mathbf{G}_k \triangleq \begin{cases} \mathbf{C}_k, & k = 1, \ldots, K \\ \mathbf{D}_{k-K}, & k = K+1, \ldots, 2K \end{cases}$$

Using the definition of the underline operator (4), we have

$$\begin{aligned} \mathbf{a}_k(\mathbf{H}_1) &= \begin{bmatrix} \mathrm{vec}\{\mathrm{Re}(\mathbf{G}_k\mathbf{H}_1)\} \\ \mathrm{vec}\{\mathrm{Im}(\mathbf{G}_k\mathbf{H}_1)\} \end{bmatrix} \\ &= \begin{bmatrix} \mathrm{Re}(\mathbf{I}_M \otimes \mathbf{G}_k) & -\mathrm{Im}(\mathbf{I}_M \otimes \mathbf{G}_k) \\ \mathrm{Im}(\mathbf{I}_M \otimes \mathbf{G}_k) & \mathrm{Re}(\mathbf{I}_M \otimes \mathbf{G}_k) \end{bmatrix} \begin{bmatrix} \mathrm{vec}\{\mathrm{Re}(\mathbf{H}_1)\} \\ \mathrm{vec}\{\mathrm{Im}(\mathbf{H}_1)\} \end{bmatrix} \\ &= \mathbf{\Psi}_k \mathbf{h}_1 \end{aligned} \tag{12}$$

where

$$\mathbf{\Psi}_k \triangleq \begin{bmatrix} \mathrm{Re}(\mathbf{I}_M \otimes \mathbf{G}_k) & -\mathrm{Im}(\mathbf{I}_M \otimes \mathbf{G}_k) \\ \mathrm{Im}(\mathbf{I}_M \otimes \mathbf{G}_k) & \mathrm{Re}(\mathbf{I}_M \otimes \mathbf{G}_k) \end{bmatrix} \tag{13}$$

is a $2MT \times 2MN$ real-valued matrix, and $\mathbf{h}_1 \triangleq \underline{\mathbf{H}_1}$.

Since for any OSTBC $\mathbf{G}_k^H\mathbf{G}_k = \mathbf{I}_N$ [2], one can easily prove from (13) that for any OSTBC

$$\mathbf{\Psi}_k^T\mathbf{\Psi}_k = \mathbf{I}_{2MN} \tag{14}$$

Note that for the Alamouti code [1], $N = T = 2$ and $\mathbf{\Psi}_k$ is a square matrix. Using the property (14), we have that in this case $\mathbf{\Psi}_k$ is a unitary matrix.

Next, let us consider the error matrix $\mathbf{\Delta}_1 \triangleq \mathbf{H}_1 - \hat{\mathbf{H}}_1$ between the true channel matrix $\mathbf{H}_1$ and its presumed (e.g., estimated) value $\hat{\mathbf{H}}_1$ and let the Frobenius norm of this error matrix be upper bounded by a known constant ϵ, that is,

$$\|\mathbf{\Delta}_1\| \le \epsilon \tag{15}$$

Using the linearity of the underline operator (4), we have

$$\mathbf{a}_k(\mathbf{H}_1) = \mathbf{a}_k(\hat{\mathbf{H}}_1) + \mathbf{a}_k(\mathbf{\Delta}_1) = \mathbf{a}_k(\hat{\mathbf{H}}_1) + \mathbf{\Psi}_k\mathbf{d}_1 \tag{16}$$

where $\mathbf{d}_1 \triangleq \underline{\mathbf{\Delta}_1}$.

The sought robust linear receiver should minimize the output power subject to the constraint that the distortionless response is maintained for any estimate $\hat{\mathbf{H}}_1$ of the channel matrix $\mathbf{H}_1$ that satisfies (15). To combat the finite-sample effects, we additionally replace the minimization of the output power with minimization of the worst-case output power [10]. Then, the weight vector $\mathbf{w}_k$ ($k = 1, \ldots, 2K$) of our robust linear receiver can be obtained as the solution to the following constrained optimization problem:

$$\min_{\mathbf{w}_k} \max_{\|\mathbf{\Delta}\| \leq \gamma} \mathbf{w}_k^T (\hat{\mathbf{R}} + \mathbf{\Delta}) \mathbf{w}_k$$
$$\text{s.t.} \quad \min_{\|\mathbf{d}_1\| \leq \epsilon} \mathbf{w}_k^T (\mathbf{a}_k(\hat{\mathbf{H}}_1) + \mathbf{\Psi}_k \mathbf{d}_1) \geq 1 \qquad (17)$$

where $\mathbf{\Delta}$ is the covariance matrix mismatch due to finite-sample effects. The main difference between (17) and (9) is that the unstructured uncertainty $\mathbf{e}_k$ in (9) is replaced by the structured uncertainty $\tilde{\mathbf{e}}_k \triangleq \mathbf{\Psi}_k \mathbf{d}_1$. Note that the structure of the OSTBC is contained in $\mathbf{\Psi}_k$. Taking this structure into account in (17) reduces the set of all possible values of $\mathbf{e}_k$ to a smaller subset of the values which $\tilde{\mathbf{e}}_k$ can take.

Now, let us solve the problem (17). The left hand side of the inequality constraint in (17) can be written as

$$\min_{\|\mathbf{d}_1\| \leq \epsilon} \mathbf{w}_k^T (\mathbf{a}_k(\hat{\mathbf{H}}_1) + \mathbf{\Psi}_k \mathbf{d}_1) = \mathbf{w}_k^T \mathbf{a}_k(\hat{\mathbf{H}}_1) - \epsilon \|\mathbf{\Psi}_k^T \mathbf{w}_k\| \quad (18)$$

The equality holds if $\mathbf{d}_1 = -\epsilon \mathbf{\Psi}_k^T \mathbf{w}_k / \|\mathbf{\Psi}_k^T \mathbf{w}_k\|$. Using this fact and take into account that [10]

$$\max_{\|\mathbf{\Delta}\| \leq \gamma} \mathbf{w}_k^T (\hat{\mathbf{R}} + \mathbf{\Delta}) \mathbf{w}_k = \mathbf{w}_k^T (\hat{\mathbf{R}} + \gamma \mathbf{I}_{2MT}) \mathbf{w}_k \qquad (19)$$

the optimization problem (17) can be equivalently written as

$$\min_{\mathbf{w}_k} \mathbf{w}_k^T (\hat{\mathbf{R}} + \gamma \mathbf{I}_{2MT}) \mathbf{w}_k$$
$$\text{s.t.} \quad \mathbf{w}_k^T \mathbf{a}_k(\hat{\mathbf{H}}_1) - \epsilon \|\mathbf{\Psi}_k^T \mathbf{w}_k\| \geq 1 \qquad (20)$$

Using the technique developed in [8], the problem (20) can be converted into the following second-order cone programming (SOCP) problem:

$$\min_{\mathbf{w}_k, \tau} \tau \quad \text{s.t.} \quad \|\mathbf{U}\mathbf{w}_k\| \leq \tau \qquad (21)$$
$$\epsilon \|\mathbf{\Psi}_k^T \mathbf{w}_k\| \leq \mathbf{w}_k^T \mathbf{a}_k(\hat{\mathbf{H}}_1) - 1$$

where $\mathbf{U}$ is an upper triangular matrix from the Cholesky decomposition $\mathbf{U}^T\mathbf{U}$ of $\hat{\mathbf{R}} + \gamma \mathbf{I}_{2MT}$. The complexity of solving the problem (21) using interior point method is $\mathcal{O}(M^3T^3)$.

An intuitive interpretation of the problem (20) can be obtained using the Lagrange multiplier method. The Lagrangian function $\mathcal{L}$ can be written as

$$\mathcal{L} = \mathbf{w}_k^T (\hat{\mathbf{R}} + \gamma \mathbf{I}_{2MT}) \mathbf{w}_k - \mu (\mathbf{w}_k^T \mathbf{a}_k(\hat{\mathbf{H}}_1) - \epsilon \|\mathbf{\Psi}_k^T \mathbf{w}_k\| - 1) \qquad (22)$$

where $\mu > 0$ is the unknown Lagrange multiplier. Using (22), we obtain that the solution to (20) can be found from the equation

$$\left(2\hat{\mathbf{R}} + 2\gamma \mathbf{I}_{2MT} + \mu\epsilon \frac{\mathbf{\Psi}_k \mathbf{\Psi}_k^T}{\|\mathbf{\Psi}_k^T \mathbf{w}_k\|} \right) \mathbf{w}_k = \mu \mathbf{a}_k(\hat{\mathbf{H}}_1) \qquad (23)$$

To get around the problem of computing the Lagrange multiplier μ, let us use the same trick as in (11). That is, let us exploit again the fact that each of the information-bearing symbols is drawn from a constant modulus constellation. As mentioned above, in this case each $\mathbf{w}_k$ can be arbitrarily rescaled without affecting the performance of a linear receiver. Applying the rescaling $\mathbf{w}_k := 2\mathbf{w}_k/\mu$, equation (23) can be rewritten as

$$\mathbf{w}_k = \left(\hat{\mathbf{R}} + \gamma \mathbf{I}_{2MT} + \epsilon \frac{\mathbf{\Psi}_k \mathbf{\Psi}_k^T}{\|\mathbf{\Psi}_k^T \mathbf{w}_k\|} \right)^{-1} \mathbf{a}_k(\hat{\mathbf{H}}_1) \qquad (24)$$

The term $\gamma \mathbf{I}_{2MT} + \epsilon \mathbf{\Psi}_k \mathbf{\Psi}_k^T / \|\mathbf{\Psi}_k^T \mathbf{w}_k\|$ in (24) can be interpreted as a combination of *fixed diagonal loading* and *adaptive non-diagonal loading* of the data covariance matrix $\hat{\mathbf{R}}$. Interestingly, the non-diagonal loading term depends on the matrix $\mathbf{\Psi}_k$ which, in turn, is entirely determined by the structure of the OSTBC. In the particular case of the Alamouti code, $\mathbf{\Psi}_k$ is unitary and the loading becomes *fully diagonal*. Moreover, for the Alamouti code we have $\|\mathbf{\Psi}_k^T \mathbf{w}_k\| = \|\mathbf{w}_k\|$. Therefore, in this particular case the proposed receiver (24) reduces to the receiver (11) with additional fixed diagonal loading term $\gamma \mathbf{I}_{2MT}$.

Because of the (generally) non-diagonal term in (24), we cannot easily solve this equation. Therefore, we resort to the SOCP-based solution to (24) which can be obtained using (21) and interior-point method.

4. SIMULATION RESULTS

In simulations, we use the 3/4-rate ($K = 3; T = 4$) OSTBC from [2] and assume $P = 2$ transmitters with $N = 3$ antennas per transmitter. The receiver has $M = 4$ antennas. It is assumed that the interfering transmitter uses the same OSTBC as the transmitter-of-interest. Throughout the simulations, the interference-to-noise ratio (INR) is equal to 10 dB and the QPSK modulation scheme is used. All plots are averaged over 1000 independent simulation runs. In each simulation run, the elements of the true channel matrices $\mathbf{H}_1$ were independently drawn from a complex Gaussian random generator with zero mean and unit variance, whereas each element of the presumed channel matrix $\hat{\mathbf{H}}_1$ was generated by drawing a complex Gaussian random variable with zero mean and the variance of 0.1 and adding this variable to a corresponding element of the matrix $\mathbf{H}_1$. The imperfect CSI case is assumed, i.e., all the receivers tested use the presumed channel matrix rather than the true one.

The proposed robust receiver (21) is compared to the DLMV receiver (8) and the robust receiver (11). To make the latter receiver robust against the finite-sample effects and consistent to the other receivers tested, the fixed diagonal loading with the factor γ has been added to it. For all receivers, the parameter $\gamma = 5\sigma^2$ is chosen where σ^2 is the noise variance. For the receivers (11) and (21), the parameter $\epsilon = \|\hat{\mathbf{H}}_1\|/2$ has been chosen.

In Fig. 1, the symbol error rates (SERs) of these linear receivers are displayed versus the SNR for $J = 200$. In the same figure, we also show the performance of the receivers (11) and (21) when the true data covariance matrix $\mathbf{R}$ is used instead of the sample matrix $\hat{\mathbf{R}}$. These two receivers do not correspond to any practical situation and are displayed for the sake of comparisons only. From Fig. 1 it follows that, as expected, the proposed robust receiver (21) substantially outperforms the receiver (11) both in the cases when the true and sample covariance matrix is used. We also observe from this figure that the performance of the DLMV receiver severely degrades at high SNRs. This is the effect of fixed

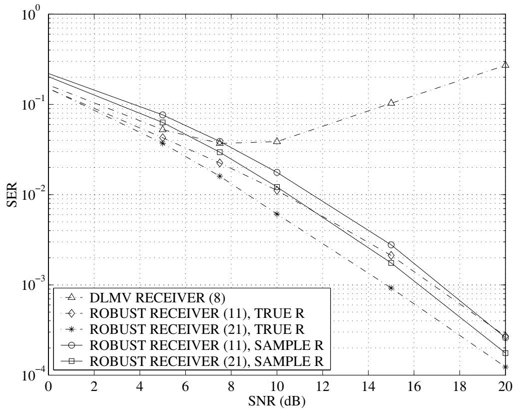

Fig. 1. SER versus SNR.

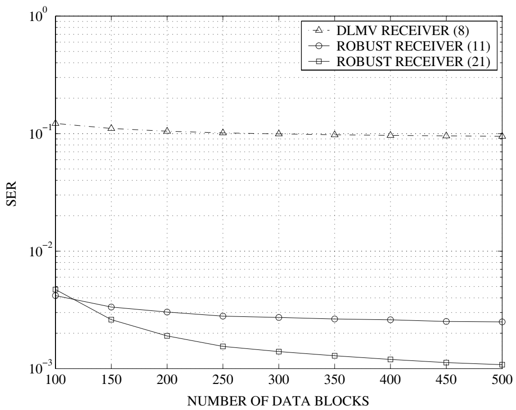

Fig. 2. SER versus number of data blocks.

diagonal loading which cannot be optimal for all values of SNR [10].

Fig. 2 shows the SERs of the same receivers versus the number of data blocks J for $\mathrm{SNR} = 15$ dB. We can see that the proposed receiver (21) outperforms the robust receiver (11) if $J > 100$.

Fig. 3 shows the receiver SERs versus the parameter $\epsilon/\|\hat{\mathbf{H}}_1\|$. It can be seen that the performance of both the receivers (11) and (21) is not very sensitive to the choice of this parameter. Note also that the performance of the DLMV receiver in Figs. 2 and 3 is much worse than that of the receivers (11) and (21).

5. CONCLUSIONS

In this paper, a new robust linear receiver for joint space-time decoding and interference rejection in multiuser MIMO wireless communication systems has been designed. The proposed receiver is based on the worst-case performance optimization and exploits the inherent structure of the OSTBCs. Simulation results show an improved robustness of the proposed receiver against mismatched CSI.

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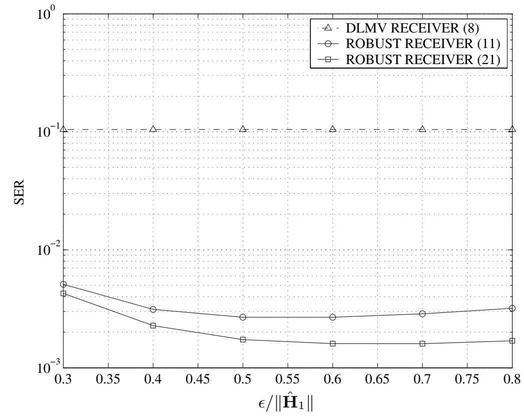

Fig. 3. SER versus $\epsilon/\|\hat{\mathbf{H}}_1\|$.