BLIND DETECTION OF HIGH RATE ORTHOGONAL SPACE-TIME BLOCK CODES

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ABSTRACT

We present a new approach to blind equalization for generalized orthogonal space-time block codes. Our method takes the form of linear programming (LP) and is globally convergent. We exploit the implicit structure of orthogonal space-time block codes to cast the problem as linear programming that can be solved efficiently. Unlike several known methods, the proposed technique is applicable to many full-rate orthogonal space time codes such as the popular Almouti code. Our algorithm allows receiver detection of full diversity codes without channel knowledge with detection performance comparable to the optimum maximum-likelihood (ML) detection.

Index Terms— MIMO, space time block codes, linear programming, blind equalization, channel estimation.

1. INTRODUCTION

Over the past decade, space-time coding has established itself as an effective technique to achieve full transmit antenna diversity in multi-input multi-output (MIMO) wireless communication systems [1][2]. Without channel state information (CSI) at the transmitter side, space-time block codes (STBC) can effectively combat channel fading in MIMO wireless systems. In particular, orthogonal STBC represent an attractive class of space-time block codes because they support full diversity and, at the same time, admit maximum-likelihood (ML) receivers with low complexity. However, it should be noted that STBC detection requires channel state information at the receiver in order to achieve the full MIMO diversity and simple ML detection. This channel information is typically estimated at the receiver by exploiting the pilot data that are sent by the transmitters.

In order to conserve transmission power and channel bandwidth, blind or semi-blind channel estimation and equalization techniques provide viable alternatives that requires little or no pilot data assistance. In the literature, several blind channel estimation techniques for flat-fading and frequency selective channels [3] [4] [5] have been proposed for generic STBC transmissions. Once a channel estimate is obtained, coherent ML detector and other low complexity detectors can then be applied to recover the information data sequence from received signals. These aforementioned channel estimation techniques essentially rely on the separation of signal and noise subspaces [6] by exploiting redundancy in the code structure and MIMO system dimensions ($n_r \times n_t$). In particular, the blind channel estimates for frequency selective fading channels proposed in [4] [5] are derived from the noise subspace of the received signal matrix. One major drawback of the existing blind channel estimation algorithms is that they fail under full rate Alamouti code [1] when the MIMO flat fading channel matrix does not have full column rank. In fact, strict channel identification conditions are required for full-rate code like Alamouti and some other lower rate codes. An investigation into the channel identifiability condition for subspace methods under STBCs have shown that more receive antennas than the transmit antennas are required for the existence of noise space[3].

A different approach to blind detection of STBC transmission is to directly equalize the unknown MIMO channel blindly. One such algorithm, proposed in [7], exploits the structure of STBCs to achieve zero-forcing blind equalization. Nevertheless, this method is not applicable to full-rate code like the Alamouti code and several other near full rate codes. Another ML detector proposed in [8] exhibits high bit error rate (BER) as compared to pilot assisted and semi-blind schemes. In fact, the authors of [9] studied the blind ML detection of rotationally-invariant STBCs and concluded that STBCs like the popular Alamouti code cannot be uniquely identified without exploiting additional signaling information.

In this paper, we present a new blind detection algorithm that can be applied to the popular Alamouti code and other high rate orthogonal STBCs. We achieve the signal recoverability by exploiting the knowledge on QAM signal constellation. Under linear QAM transmission in STBC, we develop a linear programming (LP) based channel equalization for Orthogotnal STBCs that achieves detection performance comparable to ML detection with perfect CSI. Moreover, this new algorithm does not impose any additional MIMO chan-

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nel condition. Our scheme is different from that of [7] in that we equalize a virtual channel using LP. Moreover, we exploit the structure of the orthogonal virtual channel to recover all signal streams under orthogonal STBC transmissions. The proposed algorithm retains full diversity and admits blind ML detection.

This manuscript is presented as follows. First, the STBC and channel models for blind equalization are presented in section 2, along with the important assumptions necessary for algorithm development. The basic linear programming approach to blind equalization under orthogonal STBC is is derived in section 3. Its global optimality and unique equalizer solution is discussed. Numerical test results are presented in section 4.

2. SYSTEM MODEL

2.1. Notation

We use uppercase boldface letters and lower case boldface letters for matrices and vectors, respectively.

$\{\cdot\}^H$	matrix conjugate transpose
$\{\cdot\}^T$	matrix transpose
$\{\cdot\}^*$	Matrix conjugate
†	Moore-Penrose pseudo-inverse
n_t	number of transmit (Tx) antennas
n_r	number of receive (Rx) antennas
$h_{i,k}$	channel gain between the k^{th} Tx antenna and the i^{th} Rx antenna
s_k	k^{th} information symbol
n_s	number of information symbols in each codeword
$\mathcal{W}$	additive white Gaussian noise (AWGN) matrix
$\mathbf{H}$	$n_r \times n_t$ flat fading MIMO channel matrix
$\boldsymbol{\theta}_j$	j^{th} ZF blind equalizer

2.2. Channel Model

Consider an MIMO channel $\mathbf{H}$ in a STBC system. We assume the flat-fading channel matrix $\mathbf{H} \in \mathcal{C}^{n_r \times n_t}$ to be time-invariant during one frame of L information symbols $\mathbf{s} = \begin{bmatrix} s_1 & \ldots & s_{L-1} & s_L \end{bmatrix}^T$. Each element $h_{i,j}$ in $\mathbf{H}$ is an independent circularly symmetric Gaussian random variable, identically distributed as

$$h_{i,j} \sim N_C(0, \sigma_h^2). \tag{1}$$

2.3. Signal Model

We assume that the user data s_k in frame $\mathbf{s}$ are generated from a symmetric (QAM) constellation set $\mathcal{Q}$ such that its real and imaginary parts are independent and identically distributed (i.i.d.). Additionally, the constellation satisfies

$$-M \leq \mathrm{Re}(s_k) \leq M, \text{ and } -M \leq \mathrm{Im}(s_k) \leq M.$$

The signal s_k is persistently exciting.

The STBC maps a set of complex symbols $\{s_1, \ldots, s_{n_s}\}$ onto the code word matrix $\mathbf{X} \in \mathcal{C}^{n_t \times N}$ to be transmitted in N epochs. Thus, the code rate is $\frac{n_s}{N}$. Let $K = \frac{L}{n_s}$ be the number of codewords corresponding to L data symbols. Let $\mathbf{X}_i$ be the i^{th} codeword. The symbol frame $\mathbf{s}$ is mapped into the data matrix

$$\mathcal{X} = [\mathbf{X}_1 \cdots \mathbf{X}_K] \in \mathcal{C}^{n_t \times KN}$$

For the K STBC codewords, the received signal matrix $\mathcal{Y} \in \mathcal{C}^{n_r \times KN}$ corrupted by additive white gaussian noise $\mathcal{W} \in \mathcal{C}^{n_r \times KN}$ equals

$$\mathcal{Y} = \mathbf{H}\mathcal{X} + \mathcal{W} \tag{2}$$

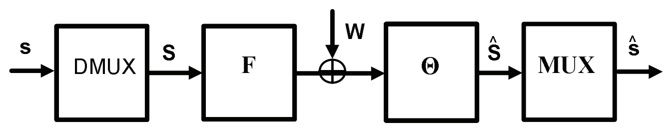

Fig. 1. System Model for OSTBC with Virtual Channel

As shown in Figure 1, the equivalent V-BLAST model for STBC can be written as a virtual channel matrix $\mathbf{F} = f(\mathbf{H}) \in \mathcal{C}^{Nn_r \times n_t}$. The virtual channel matrix $\mathbf{F}$ consists of orthogonal columns for orthogonal STBC (OSTBC). The matrix form of the Figure 1 is

$$\mathbf{Y} = \mathbf{F}.\mathbf{S} + \mathbf{W}, \tag{3}$$

where $\mathbf{Y} = \begin{bmatrix} \mathbf{Y}_1 \\ \vdots \\ \mathbf{Y}_{n_r} \end{bmatrix}$, $\mathbf{F} = \begin{bmatrix} \mathbf{F}_1 \\ \vdots \\ \mathbf{F}_{n_r} \end{bmatrix}$, and $\mathbf{W} = \begin{bmatrix} \mathbf{W}_1 \\ \vdots \\ \mathbf{W}_{n_r} \end{bmatrix}$.

$$\mathbf{F}^H\mathbf{F} = \alpha \mathbf{I}_{n_t}, \quad \alpha = \| \mathbf{H} \|^2 . \tag{4}$$

$\mathbf{S}$ is an $n_s \times K$ matrix of demultiplexed data symbol from $\mathbf{s}$. The block matrices $\mathbf{Y}_n \in \mathcal{C}^{N \times \mathbf{K}}$ and $\mathbf{W}_n \in \mathcal{C}^{N \times \mathbf{K}}$ are obtained from $\mathcal{Y}$ and $\mathcal{W}$ as follows

$$\mathbf{Y}_n = \begin{bmatrix} \mathcal{Y}_n^{(Nj-N+1)} \\ \mathcal{Y}_n^{(Nj-N+2)*} \\ \vdots \\ \mathcal{Y}_n^{(Nj)*} \end{bmatrix}, \quad \mathbf{W}_n = \begin{bmatrix} \mathcal{W}_n^{(Nj-N+1)} \\ \mathcal{W}_n^{(Nj-N+2)*} \\ \vdots \\ \mathcal{W}_n^{(Nj)*} \end{bmatrix},$$

where $n = 1, \cdots, n_r$ and $j = 1, \cdots, \frac{\mathbf{L}}{n_s}$. Here, $\mathcal{Y}_n^{(j)}$ (or $\mathcal{W}_n^{(j)}$) represents the $(n,\ j) - th$ element of $\mathcal{Y}$ (or $\mathcal{W}$). The matched filter receiver is also the ML detector that generates

$$\hat{\mathbf{S}}_{ML} = \frac{\mathbf{F}^H\mathbf{Y}}{\| \mathbf{H} \|^2}. \tag{5}$$

For the case of the rate 1 Alamouti code [1] ($n_s = 2, N = 2$)

$$\mathbf{X}_i = \begin{bmatrix} s_{2i-1} & s_{2i}^* \\ s_{2i} & -s_{2i-1}^* \end{bmatrix}, \tag{6}$$

The relation between the two signal models is:

$$\mathbf{S} = \begin{bmatrix} s_1 & s_3 & \dots & s_{L-1} \\ s_2 & s_4 & \dots & s_L \end{bmatrix}_{2\times\frac{L}{2}}, \quad \mathbf{Y}_n = \begin{bmatrix} \mathcal{Y}_n^{(2j-1)} \\ \mathcal{Y}_n^{(2j)*} \end{bmatrix}$$

$$\mathbf{F}_n = \begin{bmatrix} h_{n1} & h_{n2} \\ -h_{n2}^* & h_{n1}^* \end{bmatrix} \quad \text{and} \quad \mathbf{W}_n = \begin{bmatrix} \mathcal{W}_n^{(2j-1)} \\ \mathcal{W}_n^{(2j)*} \end{bmatrix}.$$

3. BLIND EQUALIZATION FOR OSTBC

We generalize the concept of linear programming (LP) [10] for OSTBCs. Let $\mathbf{\Theta} = \begin{bmatrix} \boldsymbol{\theta}_1 & \dots & \boldsymbol{\theta}_{n_t} \end{bmatrix} \in \mathcal{C}^{Nn_r \times n_t}$ be the blind ZF equalizer to be computed using LP, then

$$\mathbf{Z} = \mathbf{\Theta}^H \mathbf{Y} \in \mathcal{C}^{n_t \times (L/n_t)} \tag{7}$$

is the matrix consisting of all the equalizer outputs corresponding to the L data symbols in $\mathbf{s}$.

The convex cost function that achieves global minimization of the interference for the j^{th} blind equalizer $\boldsymbol{\theta}_j = \left[\theta_j^{(1)}, \cdots, \theta_j^{(Nn_r)}\right]^T$ is

$$\mathbf{J}(\boldsymbol{\theta}_j) \triangleq \frac{1}{M} \max |\mathrm{Re}(\mathbf{Z})| \tag{8}$$

$$= \frac{1}{M} \max |\mathrm{Im}(\mathbf{Z})| \tag{9}$$

The cost function $\mathbf{J}(\boldsymbol{\theta}_j)$ is piecewise convex in $\boldsymbol{\theta}_j$. To avoid trivial solution, we exploit the full-column rank matrix $\mathbf{F}$ by applying singular value decomposition $\mathbf{Y} = \mathbf{U}\Lambda_y\mathbf{V}^H$, where

$$\Lambda_y = \mathrm{diag}(\lambda_1, \lambda_1, \dots, \lambda_{n_t}, \underbrace{\sigma_n^2, \dots, \dots, \dots, \sigma_n^2}_{Nn_s - n_t})$$

The first n_t eigenvectors of $\mathbf{U}$ correspond to the large eigenvalues of Λ_y and span the signal space $\mathbf{U}_s$, while the rest spans the noise subspace $\mathbf{U}_n$. Hence,

$$\mathbf{U} = [\mathbf{U}_s \;\; \mathbf{U}_n] \tag{10}$$

To recover the STBC signal, we require

$$\mathbf{U}_n^H \boldsymbol{\theta}_j = \mathbf{0}. \tag{11}$$

Also the non-trivial solution in the minimization of $\mathbf{J}(\boldsymbol{\theta}_j)$ is avoided by imposing a linear constraint

$$\mathrm{Re}\left(\theta_j^{(1)}\right) \quad + \quad \mathrm{Im}\left(\theta_j^{(1)}\right) = 1. \tag{12}$$

Since the blind equalizer vectors are related, only one equalizer is sufficient to construct the blind equalizer $\mathbf{\Theta}$. For Alamouti OSTBC, after solving the ambiguity, if $\boldsymbol{\theta}_1$ recovers the odd data symbols, then

$$\mathbf{\Theta} = \begin{bmatrix} \theta_1^{(1)} & -\theta_1^{(2)*} \\ \theta_1^{(2)} & \theta_1^{(1)*} \\ \vdots & \vdots \\ \theta_1^{(2n_r-1)} & -\theta_1^{(2n_r)*} \\ \theta_1^{(2n_r)} & \theta_1^{(2n_r-1)*} \end{bmatrix} \tag{13}$$

To summarize, the LP algorithm for blind equalization is as follows:

$$\begin{aligned} \text{minmize} \quad & \tau_1 + \tau_2 \\ \text{subject to} \quad & -\tau_1 \le \mathrm{Re}(\mathbf{Y}^H \boldsymbol{\theta}_j) \le \tau_1 \\ & -\tau_2 \le \mathrm{Im}(\mathbf{Y}^H \boldsymbol{\theta}_j) \le \tau_2 \\ & \mathbf{U}_n^H \boldsymbol{\theta}_j = 0 \\ & \mathrm{Re}\left(\theta_j^{(1)}\right) + \mathrm{Im}\left(\theta_j^{(1)}\right) = 1. \end{aligned} \tag{14}$$

Note that this LP algorithm is globally convergent because it minimizes a convex cost function. The only potential problem in global optimaility lies in the possibility of a (non-strictly) convex cost function. We can show that when the MIMO channel and STBC code jointly satisfies a rather weak condition, global optimality of this algorithm is guaranteed. Moreover, this detector is also asymptotically a maximum likelihood detector that preserves the diversity order of the original OSTBC. The detailed proof and discussions are omitted here due to page limit.

4. SIMULATIONS

Now we provide simulation results to test the performance of the proposed LP algorithm. In the simulations, we use $\{s_k\}$ from rectangular QAM-16. The MIMO channel $\mathbf{H}$ consisting of i.i.d. gaussian elements $\mathcal{N}(0,\ 1)$ are invariant during one data frame. To compute BER, the inherent scalar ambiguity of equalization is resolved using power normalization and one known symbol at the receiver. Our results are compared against the ML detector using PCSI and, whenever possible, against direct equalizer of [7] of similar complexity. Figure 2 compares the proposed LP equalization with the ML detection with PCSI under Alamouti OSTBC. Both 1×2 and 2×2 MIMO channels are tested using a frame length of 300 symbols. Note that no other known blind channel equalization

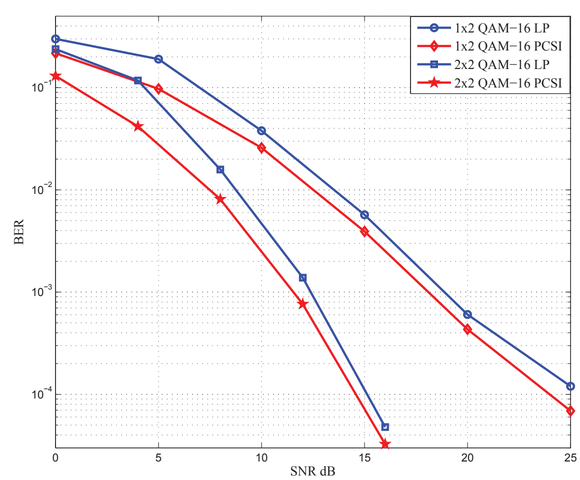

Fig. 2. BER performance for Alamouti OSTBC

and estimation algorithm is applicable to this scenario. As our results show, the performance of the LP equalizer is very close to that of the ML detection with PCSI. The SNR loss is less than 1 dB in both scenarios.

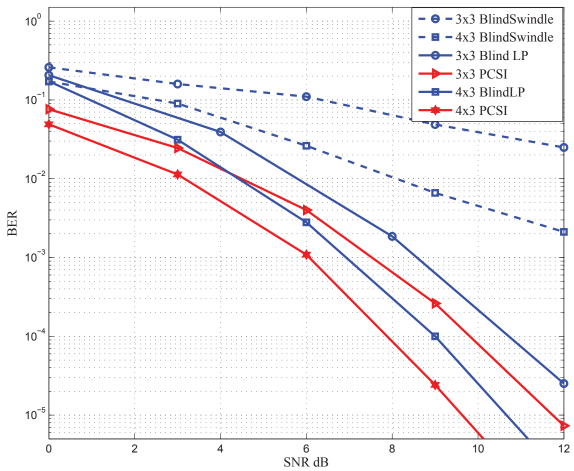

Fig. 3. BER performance for the rate R=3/4 OSTBC

For rate 3/4 OSTBC, we can compare the equalizer of [7] with our LP equalizer. Here we consider $n_r = 3$ and $n_r = 4$ and use frame length of 420 symbols. As the results in Figure 3 show, the performance of the LP algorithm is consistent with that under Alamouti code. The performance gap from the ML detector with PCSI is still below 1 dB.

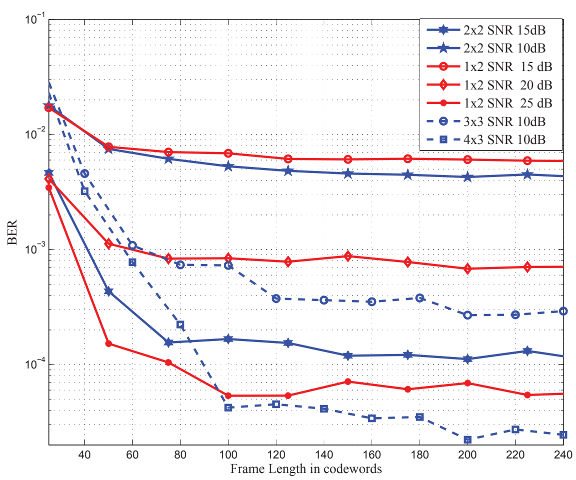

Fig. 4. Effect of frame length on BER performance

Finally, we test the sensitivity of the proposed LP equalizer to difference frame length. As shown in Figure 4, the convergence under the both Alamouti and rate 3/4 OSTBC require about 100 codewords in each frame. More receive antennas imply more unknown equalizer parameters, which leads to the need of more data symbols for convergence. This is consistent with the heuristics.

5. CONCLUSION

We present an efficient technique for blind equalization of OSTBC systems. Our algorithm based on linear programming is fast and globally convergent. The basic principle can also be extended to non-orthogonal STBCs. Our results show that this method suffers little in performance loss when compared against maximum likelihood detectors with perfect channel state information. More importantly, this proposed method is the only known blind equalization that is applicable to the most popular full-rate Alamouti code.

6. REFERENCES

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