A NEW SUBSPACE ALGORITHM FOR BLIND CHANNEL ESTIMATION IN BROADBAND SPACE-TIME BLOCK CODED COMMUNICATION SYSTEMS

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ABSTRACT

In this paper a new subspace method for blind estimation of frequency-selective channels is proposed. Specifically, we consider the case of broadband space-time block coded (STBC) transmissions such as STBC-OFDM, space-frequency block coding (SFBC), or time-reversal STBC. The proposed technique is able to exactly recover the channel within a limited number of data blocks, and its computational complexity is linear in the number of subcarriers or data-block size. Furthermore, it is independent of the specific signal constellation, and therefore it can also be applied when the sources have been precoded to exploit the multipath diversity. Finally, the performance of the proposed method is evaluated by means of some numerical examples.

1. INTRODUCTION

In the last ten years, several families of space-time block codes (STBCs) have been proposed to exploit the spatial diversity in multiple-input multiple-output (MIMO) systems. A common assumption for most of the STBCs is that perfect channel state information is available at the receiver, which has motivated an increasing interest in blind channel estimation algorithms [1–3]. The main advantage of blind techniques resides in their ability to avoid the penalty in bandwidth efficiency or signal to noise ratio (SNR) associated, respectively, to training based approaches or differential techniques [4].

Although the literature on blind and semiblind channel estimation under STBC transmissions is abundant [1–3], only a few works have considered the general problem of frequency selective channels. In particular, the techniques in [5, 6] are based on the application of standard blind channel estimation or equalization approaches, whereas in [7, 8] the authors apply the subspace method to the whole data block. Unfortunately, all these schemes require a relatively large number of data blocks at the receiver side, which translates into strong conditions in the temporal coherence of the MIMO channel. Recently, the problem of blind channel estimation within a reduced number of available blocks at the receiver has been addressed in [9, 10]. On one hand, the technique proposed in [9] considers the particular case of BPSK or QPSK constellations, which precludes its application for other constellations or linearly precoded sources. On the other hand, the computational cost of the method in [10], which is based on convex optimization, remains relatively high.

In this paper we propose a technique for the blind estimation of frequency-selective channels under STBC transmissions, which is able to recover the channel, within a reduced number of data blocks, in systems such as STBC-OFDM, space-frequency block coding (SFBC), and time-reversal STBC. In particular, the proposed technique reduces to the extraction of the principal eigenvector of a generalized eigenvalue problem (GEV) and, unlike previous approaches [7, 8], its computational cost is linear in the data block size (or number of subcarriers). Furthermore, the algorithm is solely based on the second-order statistics (SOS) of the observations and therefore it is independent of the specific signal constellation, which permits its direct application when the sources have been precoded in order to exploit the multipath diversity.

2. DATA MODEL

Throughout this paper we will use bold-faced upper case letters to denote matrices, bold-faced lower case letters for column vectors, and light-faced lower case letters for scalar quantities. Superscript $\hat{(\cdot)}$ will denote estimated matrices, vectors or scalars, the identity matrix of dimension p will be denoted as $\mathbf{I}_p$, and $\mathbf{0}$ will denote the zero matrix of the required dimensions. The superscripts $(\cdot)^T$, $(\cdot)^H$ and $(\cdot)^*$ denote transpose, Hermitian and complex conjugate, respectively. The real and imaginary parts of a matrix $\mathbf{A}$ are denoted as $\Re(\mathbf{A})$ and $\Im(\mathbf{A})$. The trace and Frobenius norm will be denoted as $\mathrm{Tr}(\mathbf{A})$ and $\|\mathbf{A}\|$, respectively. Finally,

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vec$(\mathbf{A})$ will denote the column-wise vectorized version of $\mathbf{A}$, and $\otimes$ will denote the Kronecker product.

2.1. Review of Broadband STBC Systems

Let us consider a multicarrier system with N_c subcarriers. Assuming n_T transmit and n_R receive antennas, the frequency selective channel is divided into N_c orthogonal flat-fading MIMO channels $\mathbf{H}_i \in \mathbb{C}^{n_T \times n_R}$ ($i = 1, \ldots, N_c$), which we assume are encoded with a linear space-time block code (STBC) transmitting M symbols during L uses of each flat-fading MIMO channel $\mathbf{H}_i$. The transmission rate is defined as $R = M/L$, and the number of real symbols transmitted in each STBC block is $M' = 2M$ (or $M' = M$ in the particular case of real STBCs).

For a STBC, the n-th block of data can be expressed as

$$\mathbf{S}(\mathbf{s}_i[n]) = \sum_{k=1}^{M'} \mathbf{C}_k s_{i,k}[n], \qquad i = 1, \ldots, N_c,$$

where $\mathbf{s}_i[n] = [s_{i,1}[n], \ldots, s_{i,M'}[n]]^T$ contains the M' real information symbols transmitted through the i-th channel in the n-th data block, and $\mathbf{C}_k \in \mathbb{C}^{L \times n_T}$, $k = 1, \ldots, M'$, are the code matrices.

The complex signal at the n_R receive antennas is

$$\mathbf{Y}_i[n] = \sum_{k=1}^{M'} \mathbf{W}_k(\mathbf{H}_i) s_{i,k}[n] + \mathbf{N}_i[n], \qquad i = 1, \ldots, N_c, \tag{1}$$

where $\mathbf{N}_i[n] \in \mathbb{C}^{L \times n_R}$ represents the white complex noise with variance σ^2, and $\mathbf{W}_k(\mathbf{H}_i) = \mathbf{C}_k \mathbf{H}_i$, for $k = 1, \ldots, M'$.

Defining now $\mathbf{y}_i[n] = \text{vec}\,(\mathbf{Y}_i[n])$, eq. (1) can be rewritten as

$$\mathbf{y}_i[n] = \mathbf{W}(\mathbf{h}_i)\mathbf{s}_i[n] + \mathbf{n}_i[n], \qquad i = 1, \ldots, N_c,$$

where $\mathbf{h}_i = \text{vec}\,(\mathbf{H}_i)$, $\mathbf{n}_i[n] = \text{vec}\,(\mathbf{N}_i[n])$, and $\mathbf{W}(\mathbf{h}_i)$ can be seen as the i-th complex equivalent channel, whose k-th column is given by $\text{vec}\,(\mathbf{W}_k(\mathbf{H}_i)) = \mathbf{D}_k \mathbf{h}_i$, with $\mathbf{D}_k = \mathbf{I}_{n_R} \otimes \mathbf{C}_k$, $k = 1, \ldots, M'$. Finally, in order to exploit the impropery of the sources we will use the real data model

$$\tilde{\mathbf{y}}_i[n] = \tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\mathbf{s}_i[n] + \tilde{\mathbf{n}}_i[n], \qquad i = 1, \ldots, N_c,$$

where $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i) = \left[\Re\left(\mathbf{W}^T(\mathbf{h}_i)\right) \Im\left(\mathbf{W}^T(\mathbf{h}_i)\right)\right]^T$ is the i-th real equivalent channel

$$\underbrace{\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)}_{2Ln_R \times M'} = \left[\tilde{\mathbf{D}}_1 \tilde{\mathbf{h}}_i \quad \tilde{\mathbf{D}}_2 \tilde{\mathbf{h}}_i \quad \cdots \quad \tilde{\mathbf{D}}_{M'} \tilde{\mathbf{h}}_i\right], \tag{2}$$

and we have defined the extended code matrices

$$\tilde{\mathbf{D}}_k = \underbrace{\begin{bmatrix} \Re(\mathbf{D}_k) & -\Im(\mathbf{D}_k) \\ \Im(\mathbf{D}_k) & \Re(\mathbf{D}_k) \end{bmatrix}}_{2Ln_R \times 2n_T n_R},$$

and the real vectors $\tilde{\mathbf{y}}_i[n] = \left[\Re(\mathbf{y}_i^T[n]), \Im(\mathbf{y}_i^T[n])\right]^T$, $\tilde{\mathbf{n}}_i[n] = \left[\Re(\mathbf{n}_i^T[n]), \Im(\mathbf{n}_i^T[n])\right]^T$ and $\tilde{\mathbf{h}}_i = \left[\Re(\mathbf{h}_i^T), \Im(\mathbf{h}_i^T)\right]^T$.

2.2. Channel Model

As previously pointed out, the N_c orthogonal flat-fading channels are given by the frequency response of the overall MIMO system. Thus, assuming a frequency selective channel of finite length L_c we can write

$$\underbrace{\begin{bmatrix} \mathbf{H}_1 \\ \vdots \\ \mathbf{H}_{N_c} \end{bmatrix}}_{\mathbf{H}} = \underbrace{(\mathbf{F} \otimes \mathbf{I}_{n_T})}_{\mathcal{F}} \underbrace{\begin{bmatrix} \mathbf{\Theta}_1 \\ \vdots \\ \mathbf{\Theta}_{L_c} \end{bmatrix}}_{\mathbf{\Theta}}, \tag{3}$$

where $\mathbf{\Theta} \in \mathbb{C}^{L_c n_T \times n_R}$ represents the impulse response of the MIMO channel and $\mathbf{F} \in \mathbb{C}^{N_c \times L_c}$ is an orthogonal basis given by the first L_c columns of the following matrix

$$\mathbf{G} = \left[\mathbf{f}(\delta) \quad \mathbf{f}(\tfrac{1}{N_c} + \delta) \quad \cdots \quad \mathbf{f}(\tfrac{N_c - 1}{N_c} + \delta)\right]^H,$$

where $\mathbf{f}(\omega)$ is the Fourier vector of length N_c at normalized frequency ω and $0 \leq \delta < 1/N_c$ is a frequency offset in the Fourier grid.

Based on the above definitions, it is easy to prove that the data model presented in this section can be particularized to the well-known cases of STBC-OFDM transmissions, where each STBC-OFDM block is composed of L OFDM symbols; space-frequency block coding (SFBC) , based on only one OFDM symbol; and even to the case of time-reversal STBC systems, where $\delta = 1/(2N_c)$ and the sources $\mathbf{s}_i[n]$ can be seen as a linearly precoded (with matrix $\mathbf{G}$) version of the information symbols.

3. PROPOSED BLIND CHANNEL ESTIMATION TECHNIQUE

Recently, several efforts have been made in order to blindly recover the channel, or the source, under STBC transmissions. The proposed techniques include algorithms specifically designed for flat-fading MIMO channels [1–3], adaptation of standard blind approaches [5,6], and application of the subspace method [7,8]. However, these techniques usually require a relatively large number of data blocks ($N \geq N_c M'$), which introduces a strong constraint in the temporal coherence of the channel. On the other hand, the methods proposed in [9, 10] are able to blindly recover the channel in OSTBC-OFDM systems within a few received blocks, but their computational complexity remains relatively high. Furthermore, the technique in [10] is very sensitive to errors in the previous estimate of the channel energies $\mathbf{H}_i$, and the algorithm in [9] is based on the finite alphabet property of the sources $\mathbf{s}_i[n]$, which precludes its application when the information symbols are precoded to exploit the multipath diversity.

In this section we propose a new blind channel estimation technique, which is independent of the specific signal constellation and is able to blindly recover the channel within a few data blocks. The proposed method reduces to the extraction of the main eigenvector of a generalized eigenvalue problem (GEV) and, unlike previous approaches, its computational complexity is linear in the number of subcarriers N_c.

3.1. Preliminaries

Let us consider a Gaussian distribution for the noise and a set of N data blocks at the receiver side. The unconstrained maximum likelihood (UML) estimator of the channel and information symbols can be formulated as

$$\left\{\hat{\mathbf{\Theta}}^{\text{UML}}, \hat{\mathbf{s}}_i^{\text{UML}}[n]\right\} = = \underset{\mathbf{\Theta},\mathbf{s}_i[n]}{\operatorname{argmin}} \sum_{i=1}^{N_c}\sum_{n=0}^{N-1}\left\|\tilde{\mathbf{y}}_i[n]-\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\mathbf{s}_i[n]\right\|^2.$$

Thus, under the mild assumption[1] of full-column rank equivalent channels $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)$, and after solving for $\mathbf{s}_i[n]$, the above criterion can be rewritten as

$$\hat{\mathbf{\Theta}}^{\text{UML}} = \underset{\mathbf{\Theta}}{\operatorname{argmax}} \sum_{i=1}^{N_c} \operatorname{Tr}\left(\tilde{\mathbf{U}}^T(\tilde{\mathbf{h}}_i)\mathbf{R}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{U}}(\tilde{\mathbf{h}}_i)\right), \quad (4)$$

where

$$\tilde{\mathbf{U}}(\tilde{\mathbf{h}}_i) = \tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\left(\tilde{\mathbf{W}}^T(\tilde{\mathbf{h}}_i)\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\right)^{-1/2} \in \mathbb{R}^{2Ln_R\times M'},$$

is an orthogonal basis for the subspace spanned by $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)$, and $\mathbf{R}_{\tilde{\mathbf{y}}_i} = \frac{1}{N}\sum_{n=0}^{N-1}\tilde{\mathbf{y}}_i[n]\tilde{\mathbf{y}}_i^T[n]$, is the finite sample estimate of the correlation matrix for the observations of the i-th channel.

Let us analyze the asymptotic cases of absence of noise, or perfect estimates of the correlation matrices $\mathbf{R}_{\tilde{\mathbf{y}}_i}$ ($N \to \infty$). In both cases, the right hand side term in (4) is bounded by

$$\sum_{i=1}^{N_c} \operatorname{Tr}\left(\tilde{\mathbf{U}}^T(\tilde{\mathbf{h}}_i)\mathbf{R}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{U}}(\tilde{\mathbf{h}}_i)\right) \le \sum_{i=1}^{N_c} E_i,$$

where $E_i = \sum_{k=1}^{r}\lambda_{i,k}$ is the energy in the signal subspace of $\mathbf{R}_{\tilde{\mathbf{y}}_i}$, which is of rank $r = \min(N, M')$, and $\lambda_{i,k}$ represents the r largest eigenvalues of $\mathbf{R}_{\tilde{\mathbf{y}}_i}$. Therefore, the UML decoder can be viewed as a subspace method which amounts to maximize the energy of the projection of the observations (or *empirical signal subspace*) onto the *parameter-dependent signal subspace* defined by the equivalent channels $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)$. Finally, it is clear that the above criterion is maximized by the true MIMO channels $\mathbf{H}_i$.

[1] This condition is satisfied by the most common STBCs.

3.2. Proposed Criterion (OSTBC Case)

Unfortunately, in a general situation the dependency of $\tilde{\mathbf{U}}(\tilde{\mathbf{h}}_i)$ with $\tilde{\mathbf{h}}_i$ is not trivial, and the solutions of the unconstrained blind ML decoder can not be obtained in closed-form. However, in the case of orthogonal STBCs (OSTBCs) [11] the equivalent channels satisfy $\tilde{\mathbf{W}}^T(\tilde{\mathbf{h}}_i)\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i) = \|\mathbf{H}_i\|^2\mathbf{I}_{M'}$ ($\forall\mathbf{H}_i$), which allows us to propose the following alternative channel estimation criterion

$$\hat{\mathbf{\Theta}} = \underset{\mathbf{\Theta}}{\operatorname{argmax}} \sum_{i=1}^{N_c} \operatorname{Tr}\left(\tilde{\mathbf{W}}^T(\tilde{\mathbf{h}}_i)\mathbf{R}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\right), \quad \text{s.t.} \quad \sum_{i=1}^{N_c}\|\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\|^2 E_i = M'. \quad (5)$$

Here, it is interesting to point out that the energy constraint not only avoids the trivial solution, but also ensures that in the absence of noise, or under perfect estimates of $\mathbf{R}_{\tilde{\mathbf{y}}_i}$

$$\sum_{i=1}^{N_c} \operatorname{Tr}\left(\tilde{\mathbf{W}}^T(\tilde{\mathbf{h}}_i)\mathbf{R}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\right) \le \sum_{i=1}^{N_c}\frac{\|\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)\|^2}{M'}E_i = 1,$$

where the equality is attained, $\forall N \ge 1$, by the actual MIMO channels, i.e., the theoretical solutions of the proposed criterion are those of the UML decoder.

3.3. Extension to Non-Orthogonal STBCs

In the case of general STBCs, the orthogonality property of the equivalent channels is not satisfied and the criterion presented in the previous subsection can not be directly applied. However, we propose to maximize the projection of the *parameter-dependent signal subspaces* onto the *empirical signal subspaces*, which from a practical point of view reduces to a prewhitening of the correlation matrices. Specifically, $\mathbf{R}_{\tilde{\mathbf{y}}_i}$ is replaced in (5) by

$$\mathbf{\Phi}_{\tilde{\mathbf{y}}_i} = \frac{E_i}{r}\tilde{\mathbf{U}}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{U}}_{\tilde{\mathbf{y}}_i}^T, \qquad i = 1,\dots,N_c,$$

where $\tilde{\mathbf{U}}_{\tilde{\mathbf{y}}_i}\tilde{\mathbf{U}}_{\tilde{\mathbf{y}}_i}^T$ are the projection matrices onto the *empirical signal subspaces*, and $\tilde{\mathbf{U}}_{\tilde{\mathbf{y}}_i} \in \mathbb{R}^{2Ln_R\times r}$ is obtained from the r principal eigenvectors of $\mathbf{R}_{\tilde{\mathbf{y}}_i}$.

Although it can be easily proven that, for $N \ge M'$, the actual MIMO channels are solutions of the proposed criterion, when the number of available blocks at the receiver side is $N < M'$, the proposed criterion is not necessarily maximized by the true MIMO channels. This can be seen as a consequence of two different factors: On the one hand, the signal subspace is not completely defined by the observations, i.e., the sources are not *persistently exciting*. On the other, the non-orthogonality of the equivalent channels provokes spurious channel estimates trying to concentrate most of the energy of $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)$ into the directions defined by the rank-reduced *empirical signal subspaces*. However, when $N_c >> L_c$ there are not enough degrees of freedom to select the spurious MIMO channels and, as it will be shown in the simulations section, this translates into very accurate channel estimates even for $N < M'$.

3.4. Practical Implementation

Unlike $\tilde{\mathbf{U}}(\tilde{\mathbf{h}}_i)$, the dependency of $\tilde{\mathbf{W}}(\tilde{\mathbf{h}}_i)$ with respect to $\tilde{\mathbf{h}}_i$ is explicitly given by (2), which allows us to rewrite (5) as

$$\hat{\mathbf{\Theta}} = \underset{\mathbf{\Theta}}{\text{argmax}} \sum_{i=1}^{N_c} \tilde{\mathbf{h}}_i^T \mathbf{\Xi}_i \tilde{\mathbf{h}}_i, \qquad \text{s.t.} \qquad \sum_{i=1}^{N_c} \tilde{\mathbf{h}}_i^T \mathbf{\Psi}_i \tilde{\mathbf{h}}_i = M',$$

where $\mathbf{\Xi}_i = \sum_{k=1}^{M'} \tilde{\mathbf{D}}_k^T \mathbf{\Phi}_{\tilde{\mathbf{y}}_i} \tilde{\mathbf{D}}_k$ and $\mathbf{\Psi}_i = E_i \sum_{k=1}^{M'} \tilde{\mathbf{D}}_k^T \tilde{\mathbf{D}}_k$. Now, taking into account the channel model in (3), and defining the vectors $\boldsymbol{\theta}_k = \text{vec}(\mathbf{\Theta}_k)$, $\tilde{\boldsymbol{\theta}}_k = \left[\Re(\boldsymbol{\theta}_k^T), \Im(\boldsymbol{\theta}_k^T)\right]^T$ and $\tilde{\boldsymbol{\theta}} = \left[\tilde{\boldsymbol{\theta}}_1^T, \ldots, \tilde{\boldsymbol{\theta}}_{L_c}^T\right]$, the final channel estimation criterion is

$$\hat{\tilde{\boldsymbol{\theta}}} = \underset{\tilde{\boldsymbol{\theta}}}{\text{argmax}}\, \tilde{\boldsymbol{\theta}}^T \mathbf{\Xi} \tilde{\boldsymbol{\theta}}, \qquad \text{s.t.} \qquad \tilde{\boldsymbol{\theta}}^T \mathbf{\Psi} \tilde{\boldsymbol{\theta}} = M',$$

where $\mathbf{\Xi} = \sum_{i=1}^{N_c} \mathbf{\Omega}_i^T \mathbf{\Xi}_i \mathbf{\Omega}_i$, $\mathbf{\Psi} = \sum_{i=1}^{N_c} \mathbf{\Omega}_i^T \mathbf{\Psi}_i \mathbf{\Omega}_i$, and the matrices $\mathbf{\Omega}_i$ are obtained from the i-th row ($\mathbf{f}_i^T$) of $\mathbf{F}$ as

$$\mathbf{\Omega}_i = \Re(\mathbf{f}_i^T) \otimes \mathbf{I}_{2n_T n_R} + \Im(\mathbf{f}_i^T) \otimes \begin{bmatrix} \mathbf{0} & -\mathbf{I}_{n_T n_R} \\ \mathbf{I}_{n_T n_R} & \mathbf{0} \end{bmatrix}.$$

Thus, the channel estimate $\hat{\tilde{\boldsymbol{\theta}}}$ is obtained, up to a real scalar, as the eigenvector associated to the largest eigenvalue β of the following generalized eigenvalue problem (GEV)

$$\mathbf{\Xi} \hat{\tilde{\theta}} = \beta \mathbf{\Psi} \hat{\tilde{\theta}}. \tag{6}$$

Finally, we must point out that the computational cost of the proposed technique is dominated by the obtention of the N_c matrices $\mathbf{\Phi}_{\tilde{\mathbf{y}}_i}$, which comes at a cost of order $\mathcal{O}\left(N_c(Ln_R)^3\right)$, and the solution of the GEV in (6), whose computational complexity is $\mathcal{O}\left((L_c n_T n_R)^3\right)$. Therefore, the overall computational cost is linear in the number of subcarriers (or data block size), and in general is lower than that associated to the subspace method in [7, 8], which incurs in a cost of order $\mathcal{O}\left((N_c L n_R)^3\right)$.

4. SIMULATION RESULTS

In this section the performance of the proposed technique is compared with that of the receiver with perfect channel knowledge by means of some numerical examples. We consider QPSK signaling, MMSE decoding and two different STBC-OFDM systems, although similar results have been obtained for SFBC and time-reversal STBC systems. The results have been obtained by averaging 1000 independent experiments, in which the length of the Rayleigh MIMO channels has been selected as $L_c = 4$.

4.1. Results for Orthogonal Codes (OSTBCs)

In the first set of experiments we consider a system with $n_T = 4$ and $n_R = 2$ using the OSTBC presented in Eq. (7.4.10) of [11], whose parameters are $M = 3$ and $L = 4$ ($R = 3/4$). Fig. 1.a shows the bit error rate (BER) after decoding for a system with $N_c = 64$ subcarriers. As can be seen, the proposed technique is able to extract the channel within only $N = 1$ available data-block at the receiver (although the accuracy of the estimates increases with N). On the other hand, Fig. 1.b shows the simulation results for $N = 1$, where we can see that, for a fixed channel length L_c, the accuracy of the proposed technique also increases with N_c.

4.2. Results for Non-Orthogonal Codes

The previous experiments have been repeated for a system with $n_T = n_R = 4$ transmitting with the quasi-orthogonal STBC (QSTBC) proposed in [12] ($M = L = 4$). Here, in order to avoid the identifiability problems pointed out in [1, 13], we have applied the non-redundant precoding technique proposed in [14], i.e., the STBC transmission matrices associated to the subcarriers have been rotated by means of different orthogonal matrices. The results are shown in Fig. 2, where we can see that, for $N < M'$, there exists a noise floor due to the errors in the channel estimate. However, as can be seen in Fig. 2.b, this floor does not only decreases with the number of data blocks, but also with the number of subcarriers (or data block size) N_c. In conclusion, for a sufficiently large number of available blocks or subcarriers, the performance of the proposed technique is close to that of the receiver with perfect channel knowledge.

5. CONCLUSIONS

In this paper we have proposed a new blind channel estimation technique for frequency selective channels under space-time block coded transmissions. The proposed method is solely based on the second-order statistics (SOS) of the received signals, and therefore it is independent of the specific signal constellation and can be directly applied when the information symbols are precoded to exploit the multipath diversity. Unlike previous approaches, the proposed method is able to recover the channel within a few received data blocks, and its computational complexity is linear in the number of subcarriers or data block size. Finally, the performance of the proposed technique has been illustrated by means of some numerical examples.

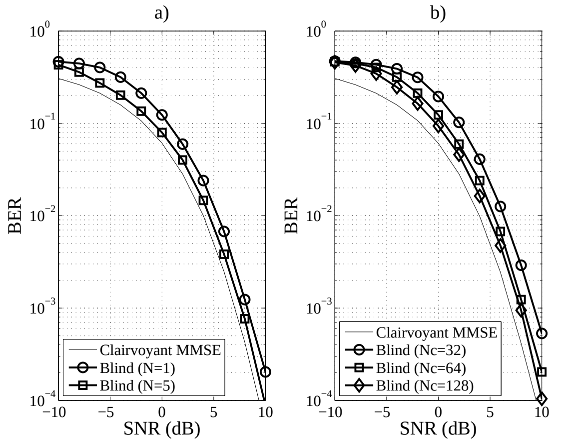

Fig. 1. Performance of the proposed technique in the OSTBC case. $n_T = 4$, $n_R = 2$, $M = 3$, $L = 4$, $L_c = 4$. a) BER vs. SNR for $N_c = 64$. b) BER vs. SNR for $N = 1$.

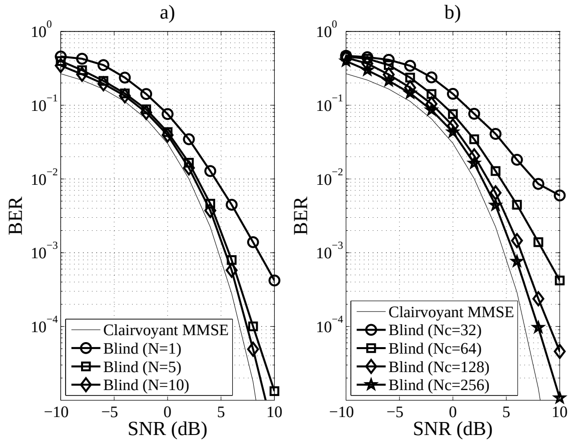

Fig. 2. Performance of the proposed technique for a QSTBC code with $n_T = n_R = M = L = L_c = 4$. a) BER vs. SNR for $N_c = 64$. b) BER vs. SNR for $N = 1$.

6. REFERENCES

[1] S. Shahbazpanahi, A. B. Gershman, and J. H. Manton, "Closed-form blind MIMO channel estimation for orthogonal space-time block codes," *IEEE Trans. on Sig. Proc.*, vol. 53, no. 12, pp. 4506–4517, Dec. 2005.

[2] W. K. Ma, B. N. Vo, T. N. Davidson, and P. C. Ching, "Blind ML detection of orthogonal space-time block codes: Efficient high-performance implementations," *IEEE Transactions on Signal Processing*, vol. 54, no. 2, pp. 738–751, Feb. 2006.

[3] N. Ammar and Z. Ding, "Blind channel identifiability for generic linear space-time block codes," *IEEE Transactions on Signal Processing*, vol. 55, no. 1, pp. 202–217, Jan. 2007.

[4] V. Tarokh and H. Jafarkhani, "A differential detection scheme for transmit diversity," *IEEE Journal on Selected Areas in Communications*, vol. 18, no. 7, pp. 1169–1174, July 2000.

[5] Jinho Choi, "Equalization and semi-blind channel estimation for space-time block coded signals over a frequency-selective fading channel," *IEEE Trans. on Sig. Proc.*, vol. 52, no. 3, pp. 774–785, Mar. 2004.

[6] Zhiguo Ding and D. B. Ward, "Subspace approach to blind and semi-blind channel estimation for space-time block codes," *IEEE Trans. on Wireless Comm.*, vol. 4, no. 2, pp. 357–362, Mar. 2005.

[7] Z. Liu, G.B. Giannakis, B. Muquet, and S. Zhou, "Space-time coding for broadband wireless communications," *Proc. of the IEEE*, vol. 1, no. 1, pp. 35–53, 2001.

[8] Z. Liu, G. Giannakis, S. Barbarossa, and A. Scaglione, "Transmit-antenna space-time block coding for generalized OFDM in the presence of unknown multipath," *IEEE Journal on Selected Areas in Communications*, vol. 19, no. 7, pp. 1352–1364, July 2001.

[9] T.-H. Chang, W.-K. Ma, and C.-Y. Chi, "Maximum-likelihood detection of orthogonal space-time block coded OFDM in unknown block fading channels," *IEEE Transactions on Signal Processing*, vol. 56, no. 4, pp. 1637–1649, Apr. 2008.

[10] N. Sarmadi, A. B. Gershman, and S. Shahbazpanahi, "Blind channel estimation in MIMO-OFDM systems using semi-definite relaxation," in *IEEE Int. Conf. on Acoustic, Speech and Signal Proc. (ICASSP)*, Las Vegas, USA, Apr. 2008.

[11] E. G. Larsson and P. Stoica, *Space-Time Block Coding for Wireless Communications*, Cambridge University Press, New York, USA, 2003.

[12] H. Jafarkhani, *Space-Time Coding: Theory and Practice*, Cambridge University Press, 2005.

[13] J. Vía and I. Santamaría, "On the blind identifiability of orthogonal space-time block codes from second order statistics," *IEEE Transactions on Information Theory*, vol. 54, no. 2, pp. 709–722, Feb. 2008.

[14] J. Vía, I. Santamaría, A. Sezgin, and A. J. Paulraj, "SOS-based blind channel estimation under space-time block coded transmissions," in *IEEE Workshop on Sig. Proc. Advances in Wireless Comm. (SPAWC 2007)*, Helsinki, Finland, June 2007.