SEMI-BLIND CHANNEL ESTIMATION FOR BLOCK PRECODED SPACE-TIME OFDM TRANSMISSIONS

Shengli Zhou[1*], *Bertrand Muquet*[2] and *Georgios B. Giannakis*[1*]

[1]Dept. of ECE, University of Minnesota, 200 Union Str. SE, Minneapolis, MN 55455
[2]Motorola Labs Paris, Espace Technologique, Saint-Aubin, 91193 Gif-sur-Yvette, France

ABSTRACT

We develop in this paper a (semi-) blind channel estimation algorithm for space time (ST) block precoded OFDM transmissions over frequency-selective channels. We establish that multi-channel identifiability is guaranteed up to one or two scalar ambiguities, when distinct or identical precoders are employed for even and odd indexed symbol blocks. With known pilots inserted before precoding, we resolve the residual scalar ambiguities and show that distinct precoders require less pilots than identical precoders to achieve the same channel estimation accuracy. Simulation results confirm our theoretical analysis and illustrate that the proposed semi-blind algorithm is capable of tracking slow channel variations and improving the overall system performance relative to competing differential ST alternatives.

1. INTRODUCTION

New applications such as high speed Internet access and wireless digital television call for high data rate transmissions. Usage of multiple transmit- and receive-antennas has the potential to increase the channel capacity, and thus the maximum achievable rate. Equipped with Space-Time Coding (STC) at the transmitter and intelligent signal processing at the receiver, multi-antenna transceivers offer also diversity and coding advantages over single antenna systems (see [4,6] for tutorial treatments). But all these enhancements in capacity, diversity and coding gains can be realized if the underlying channels can be acquired at the receiver.

Conventionally, training symbols are transmitted periodically to assist the receiver in acquiring channel state information (CSI), see e.g., [2] for ST-OFDM systems. However, training sequences consume bandwidth and thereby incur spectral efficiency loss especially in rapidly varying environments. For this reason, blind channel estimators receive growing attention. Relying on non-redundant and non-constant modulus precoding, [1] proposed blind channel estimation and equalization for OFDM-based multi-antenna systems using cyclostationary statistics. For ST-OFDM, a deterministic blind channel estimator was derived in [3] when the channel transfer functions are coprime (no common zeros) and the transmitted signals have constant-modulus (CM).

In this paper, we deal with a linearly precoded ST-OFDM system with two transmit antennas and derive (semi-) blind channel identification algorithms for frequency-selective FIR channels. With properly designed redundant precoders, the proposed subspace-based blind channel estimator possesses the following three attractive features: i) it can be applied to arbitrary signal constellations; ii) it guarantees channel identifiability regardless of the underlying channel zero locations; iii) it can estimate multiple channels simultaneously up to one or two scalar ambiguities.

To enable channel equalization, we also show how to resolve the residual scalar ambiguities using a minimal number of pilots that we insert before precoding.

Notation: Bold upper (lower) letters denote matrices (column vectors); $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^{\mathcal{H}}$ denote conjugate, transpose and Hermitian transpose; $\mathcal{R}(\cdot)$ stands for range space; $\mathbf{I}_K$ denotes the identity matrix of size K and $\mathbf{0}$ denotes an all-zero matrix or vector; $\text{diag}(\mathbf{x})$ will stand for a diagonal matrix with $\mathbf{x}$ on its diagonal; $[\cdot]_p$ denotes the pth entry of a vector, and $[\cdot]_{p,q}$ denotes the (p,q)th entry of a matrix.

2. SYSTEM DESCRIPTION

Figure 1 depicts the wireless system considered in this paper, where the ST transceiver is equipped with two transmit antennas and one receive antenna as in [4]. Prior to transmission, the information bearing symbols are first grouped into blocks $\mathbf{s}(n)$ of size $K \times 1$. Two different linear block precoders denoted by the tall $J \times K$ matrices $\mathbf{\Theta}_1$ and $\mathbf{\Theta}_2$, one for the even block indices $2n$ and one for the odd indices $2n+1$, are used to introduce redundancy ($J > K$). The corresponding $J \times 1$ precoded blocks

$$\tilde{\mathbf{s}}(2n) := \mathbf{\Theta}_1\mathbf{s}(2n) \quad \text{and} \quad \tilde{\mathbf{s}}(2n+1) := \mathbf{\Theta}_2\mathbf{s}(2n+1), \quad (1)$$

are fed to the ST encoder $\mathcal{M}(\cdot)$. The ST encoder takes as input two consecutive precoded blocks, $\tilde{\mathbf{s}}(2n)$ and $\tilde{\mathbf{s}}(2n+1)$, to output the following $2J \times 2$ code matrix:

$$\begin{bmatrix} \bar{\mathbf{s}}_1(2n) & \bar{\mathbf{s}}_1(2n+1) \\ \bar{\mathbf{s}}_2(2n) & \bar{\mathbf{s}}_2(2n+1) \end{bmatrix} := \begin{bmatrix} \tilde{\mathbf{s}}(2n) & -\tilde{\mathbf{s}}^*(2n+1) \\ \tilde{\mathbf{s}}(2n+1) & \tilde{\mathbf{s}}^*(2n) \end{bmatrix}.$$

Each block column of this matrix is transmitted over successive time intervals with the blocks $\bar{\mathbf{s}}_1(n)$ and $\bar{\mathbf{s}}_2(n)$ sent through transmit-antennas 1 and 2, respectively.

The frequency-selective channels between the two transmit antennas and the receive antenna can be modeled as FIR linear time-invariant filters with impulse responses $\mathbf{h}_i := [h_i(0), \ldots, h_i(L)]$, $i = 1, 2$, where L is an upper bound on the channel orders of $\mathbf{h}_1$ and $\mathbf{h}_2$. Moreover, we assume that OFDM modulation has been deployed to convert the FIR channels into a set of parallel flat faded subchannels (see e.g., [8] for detailed derivations). Let $\boldsymbol{\mathcal{D}}_1$ and $\boldsymbol{\mathcal{D}}_2$

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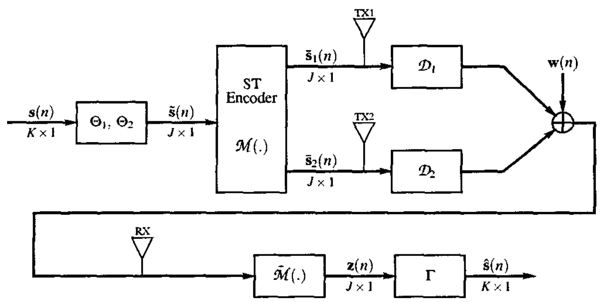

Fig. 1. Block precoded ST-OFDM transceiver model

be the diagonal matrices corresponding to these subchannels: $\mathcal{D}_i := \text{diag}[H_i(0) \dots H_i(J-1)]$, where $H_i(k) := \sum_{l=0}^{L} h_i(l) e^{-j\frac{2\pi}{J}lk}$. Considering two successive received blocks: $\bar{\mathbf{y}}(2n)$ and $\bar{\mathbf{y}}(2n+1)$, let us define the super blocks $\check{\mathbf{y}}(n)$ and $\check{\mathbf{s}}(n)$ as: $\check{\mathbf{y}}(n) := [\bar{\mathbf{y}}^T(2n), \bar{\mathbf{y}}^{\mathcal{H}}(2n+1)]^T$ and $\check{\mathbf{s}}(n) := [\mathbf{s}^T(2n), \mathbf{s}^T(2n+1)]^T$. Letting $\check{\mathbf{w}}(n)$ be the additive noise, the received block $\check{\mathbf{y}}(n)$ can then be expressed as (see also [4] for further details):

$$\check{\mathbf{y}}(n) = \mathcal{D}\Phi_{12}\check{\mathbf{s}}(n) + \check{\mathbf{w}}(n) := \mathcal{H}\check{\mathbf{s}}(n) + \check{\mathbf{w}}(n), \quad (2)$$

where $\mathcal{D}, \Phi_{12}, \mathcal{H}$ are defined respectively as:

$$\mathcal{D} := \begin{bmatrix} \mathcal{D}_1 & \mathcal{D}_2 \\ \mathcal{D}_2^* & -\mathcal{D}_1^* \end{bmatrix}, \ \Phi_{12} := \begin{bmatrix} \Theta_1 & \mathbf{0} \\ \mathbf{0} & \Theta_2 \end{bmatrix}, \ \mathcal{H} := \mathcal{D}\Phi_{12}.$$

When the channel matrices $\mathcal{D}_1$ and $\mathcal{D}_2$ become available at the receiver, it is possible to demodulate $\check{\mathbf{y}}(n)$ with diversity gains by a simple matrix multiplication:

$$\check{\mathbf{z}}(n) = \mathcal{D}^{\mathcal{H}}\check{\mathbf{y}}(n) = \begin{bmatrix} \bar{\mathcal{D}}_{12}\Theta_1 & \mathbf{0} \\ \mathbf{0} & \bar{\mathcal{D}}_{12}\Theta_2 \end{bmatrix} \check{\mathbf{s}}(n) + \mathcal{D}^{\mathcal{H}}\check{\mathbf{w}}(n), \quad (3)$$

where the diagonal matrix $\bar{\mathcal{D}}_{12} := \mathcal{D}_1^*\mathcal{D}_1 + \mathcal{D}_2^*\mathcal{D}_2$ equals $\text{diag}[\sum_{i=1}^{2} |H_i(e^{j0})|^2, \dots, \sum_{i=1}^{2} |H_i(e^{j\frac{2\pi}{J}(J-1)})|^2]$. Eq. (3) reveals that zero-forcing recovery of $\check{\mathbf{s}}(n)$ from $\check{\mathbf{z}}(n)$ requires the matrices $\bar{\mathcal{D}}_{12}\Theta_i$, $i \in [1,2]$, to have full column rank. Because the channels have maximum order L, $\bar{\mathcal{D}}_{12}$ has at most L zero diagonal entries. Hence, the full rank of $\bar{\mathcal{D}}_{12}\Theta_i$ can be always assured if we adopt the following design conditions on the block lengths and the linear precoders:

a1) $J > K + L$;

a2) $\Theta_i, i \in \{1,2\}$, *is designed so that any K rows of Θ_i are linearly independent.*

Based on a1) and a2), our objective in this paper is to develop a subspace-based (semi-) blind multichannel estimation algorithm.

3. (SEMI-) BLIND MULTICHANNEL ESTIMATION

At the receiver, we collect N received blocks $\check{\mathbf{y}}(n)$, with:

a3) *N large enough ($\geq 2K$) so that $\mathbf{S}_N\mathbf{S}_N^{\mathcal{H}}$ has full rank $2K$, where $\mathbf{S}_N := [\check{\mathbf{s}}(0), \dots, \check{\mathbf{s}}(N-1)]$.*

Under a1), a2) and a3), a consistent blind channel estimator has been developed in [4,9]. We summarize the resulting algorithm in the following steps:

S1. Collect the received data blocks $\check{\mathbf{y}}(n)$ and compute $\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(N)} = (1/N)\sum_{n=0}^{N-1} \check{\mathbf{y}}(n)\check{\mathbf{y}}^{\mathcal{H}}(n)$;

S2. Determine the eigenvectors $\tilde{\mathbf{u}}_k$, $k = 1, \dots, 2J-2K$ corresponding to the smallest $2J - 2K$ eigenvalues of matrix $\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(N)}$; split each vector $\tilde{\mathbf{u}}_k$ into its upper and lower parts as: $\tilde{\mathbf{u}}_k = [\hat{\mathbf{u}}_k^T, \check{\mathbf{u}}_k^T]^T$ and form the matrix

$$\mathcal{D}(\tilde{\mathbf{u}}_k) := \begin{bmatrix} \text{diag}(\hat{\mathbf{u}}_k^*) & -\text{diag}(\check{\mathbf{u}}_k) \\ \text{diag}(\check{\mathbf{u}}_k^*) & \text{diag}(\hat{\mathbf{u}}_k) \end{bmatrix}.$$

S3. From these eigenvectors, estimate $[\mathbf{h}_1^T, \mathbf{h}_2^{\mathcal{H}}]^T$ as the left eigenvector corresponding to the smallest eigenvalue of $\mathbf{Q}$, where $\mathbf{Q}$ is defined as:

$$\mathbf{Q} := \mathcal{F}[\mathcal{D}(\tilde{\mathbf{u}}_1)\Psi, \dots, \mathcal{D}(\tilde{\mathbf{u}}_{2J-2K})\Psi], \quad (4)$$

with $\mathcal{F} := \begin{bmatrix} \mathbf{V}^T & \mathbf{0} \\ \mathbf{0} & \mathbf{V}^{\mathcal{H}} \end{bmatrix}$, $\Psi := \begin{bmatrix} \Theta_1 & \mathbf{0} \\ \mathbf{0} & \Theta_2^* \end{bmatrix}$, and $\mathbf{V}$ a tall Vandermonde matrix with $[\mathbf{V}]_{p+1,q+1} = e^{-j\frac{2\pi}{J}pq}$.

An inherent problem to all subspace based estimators is their relatively slow convergence with respect to the number of data required. To facilitate data efficiency and also enable tracking of slow channel variations, a semi-blind implementation of the subspace based method can be devised by capitalizing on training sequences, which are anyways present for synchronization and quick channel acquisition in practical systems. Proceeding as in [5], the *semi-blind implementation* of our algorithm is outlined next:

1. Obtain initial channel estimates $\hat{\hat{\mathbf{h}}}_1$ and $\hat{\hat{\mathbf{h}}}_2$ (and thus $\hat{\mathcal{H}}$) through training (using e.g., [2]); and estimate the autocorrelation matrix $\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}$ as (σ_s^2 denotes symbol energy): $\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(0)} = \sigma_s^2\hat{\mathcal{H}}\hat{\mathcal{H}}^{\mathcal{H}}$.
2. Refine iteratively the autocorrelation matrix each time a new symbol block $\check{\mathbf{y}}(N)$ becomes available using a rectangular sliding window of length W:

$$\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(N)} = \mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(N-1)} + \frac{1}{W}[\check{\mathbf{y}}(N)\check{\mathbf{y}}^{\mathcal{H}}(N) - \check{\mathbf{y}}(N-W)\check{\mathbf{y}}^{\mathcal{H}}(N-W)]. \quad (5)$$

3. Perform the subspace algorithm based on $\mathbf{R}_{\check{\mathbf{y}}\check{\mathbf{y}}}^{(N)}$.

4. CHANNEL IDENTIFIABILITY

The key question here is whether the solution of S3 is unique. With the proof provided in [9], we present channel identifiability results for two precoder choices: identical precoders and distinct precoders.

Theorem 1 (identical precoders): *Suppose a1), a2) and a3) hold true; if $\Theta_1 = \Theta_2 = \Theta$, the matrix $\mathbf{Q}$ in (4) loses row rank by two and the resulting estimate $[\mathbf{h}_3^T, \mathbf{h}_4^{\mathcal{H}}]^T$ belongs to a two-dimensional vector space that is spanned by $\mathbf{h}_{12} = [\mathbf{h}_1^T, \mathbf{h}_2^{\mathcal{H}}]^T$ and $\mathbf{h}_{21} = [\mathbf{h}_2^T, -\mathbf{h}_1^{\mathcal{H}}]^T$. The underlying channels are identified up to two scalar ambiguities as:*

$$\begin{bmatrix} \mathbf{h}_1 \\ \mathbf{h}_2 \end{bmatrix} = \frac{1}{|\alpha_1|^2 + |\alpha_2|^2} \begin{bmatrix} \alpha_1^*\mathbf{I} & -\alpha_2\mathbf{I} \\ \alpha_2^*\mathbf{I} & \alpha_1\mathbf{I} \end{bmatrix} \begin{bmatrix} \mathbf{h}_3 \\ \mathbf{h}_4 \end{bmatrix}. \quad (6)$$

Theorem 2 (distinct precoders): *Suppose a1), a2) and a3) hold true; let* $\bar{\mathbf{D}}$ *denote any diagonal matrix with unit amplitude diagonal entries, and* $\bar{\Theta}_1$, $\bar{\Theta}_2$ *be formed by any* $J-L$ *rows of* Θ_1, Θ_2, *respectively. If* $\bar{\Theta}_1$ *and* $\bar{\Theta}_2$ *satisfy:* $\bar{\mathbf{D}}\bar{\Theta}_1 \notin \mathcal{R}(\bar{\Theta}_2)$, *the resulting estimate* $[\mathbf{h}_3^T, \mathbf{h}_4^{\mathcal{H}}]^T$ *is unique up to a constant and thus channel identifiability within one scalar is guaranteed:*

$$\begin{bmatrix}\mathbf{h}_1\\ \mathbf{h}_2\end{bmatrix} = \alpha \begin{bmatrix}\mathbf{h}_3\\ \mathbf{h}_4\end{bmatrix}. \tag{7}$$

Therefore, from the received data only, multiple channels can be estimated simultaneously up to one or two scalar ambiguities with linearly block precoded ST-OFDM transmissions. To enable channel equalization, we show next how to resolve these scalar ambiguities by inserting known symbols in the transmitted sequence.

5. RESOLVING SCALAR AMBIGUITIES

To resolve the scalar ambiguities inherent to all blind channel estimators, known symbols are needed in the transmitted sequence. We here focus on pre-precoding pilots where the known symbols are inserted before precoding by Θ; alternatively, known symbols can be inserted after precoding and we will call them post-precoding pilots [9].

We pursue the scalar determination with identical precoders first. Notice that the estimated channels of (6) satisfy:

$$\mathcal{D}_{34} = \begin{bmatrix}\mathcal{D}_3 & \mathcal{D}_4\\ \mathcal{D}_4^* & -\mathcal{D}_3^*\end{bmatrix} = \begin{bmatrix}\mathcal{D}_1 & \mathcal{D}_2\\ \mathcal{D}_2^* & -\mathcal{D}_1^*\end{bmatrix}\begin{bmatrix}\alpha_1\mathbf{I}_J & -\alpha_2^*\mathbf{I}_J\\ \alpha_2\mathbf{I}_J & \alpha_1^*\mathbf{I}_J\end{bmatrix}, \tag{8}$$

and thus $\bar{\mathcal{D}}_{34} := \mathcal{D}_3^*\mathcal{D}_3 + \mathcal{D}_4^*\mathcal{D}_4$ equals:

$$\bar{\mathcal{D}}_{34} = (|\alpha_1|^2+|\alpha_2|^2)(\mathcal{D}_1^*\mathcal{D}_1+\mathcal{D}_2^*\mathcal{D}_2) = (|\alpha_1|^2+|\alpha_2|^2)\bar{\mathcal{D}}_{12}.$$

Multiplying $\check{\mathbf{y}}(n)$ by $\mathcal{D}_{34}^{\mathcal{H}}$ yields [c.f. (3)]:

$$\begin{aligned}\check{\mathbf{z}}(n) &= \begin{bmatrix}\mathbf{z}(2n)\\ \mathbf{z}(2n+1)\end{bmatrix} := \mathcal{D}_{34}^{\mathcal{H}}\check{\mathbf{y}}(n) \\ &= \frac{1}{|\alpha_1|^2+|\alpha_2|^2}\begin{bmatrix}\alpha_1^*\mathbf{I}_J & \alpha_2^*\mathbf{I}_J\\ -\alpha_2\mathbf{I}_J & \alpha_1\mathbf{I}_J\end{bmatrix}\begin{bmatrix}\bar{\mathcal{D}}_{34}\Theta\mathbf{s}(2n)\\ \bar{\mathcal{D}}_{34}\Theta\mathbf{s}(2n+1)\end{bmatrix},\end{aligned} \tag{9}$$

where, for brevity, we omitted the noise.

Because the known symbols are inserted in the data stream before precoding, we need to equalize the channel and compensate for the precoding first, before resolving the residual scalar ambiguities. With identical precoders, a zero-forcing (ZF) equalizer can be applied to $\mathbf{z}(2n)$ and $\mathbf{z}(2n+1)$ in (9) by pre-multiplying with $(\bar{\mathcal{D}}_{34}\Theta)^\dagger$, where † stands for matrix pseudo-inverse. Based on (9), the equalizer outputs $\grave{\mathbf{s}}(2n) := (\bar{\mathcal{D}}_{34}\Theta)^\dagger\mathbf{z}(2n)$ and $\grave{\mathbf{s}}(2n+1) := (\bar{\mathcal{D}}_{34}\Theta)^\dagger\mathbf{z}(2n+1)$ can be written as:

$$\begin{bmatrix}\grave{\mathbf{s}}(2n)\\ \grave{\mathbf{s}}(2n+1)\end{bmatrix} = \frac{1}{|\alpha_1|^2+|\alpha_2|^2}\begin{bmatrix}\alpha_1^*\mathbf{I}_J & \alpha_2^*\mathbf{I}_J\\ -\alpha_2\mathbf{I}_J & \alpha_1\mathbf{I}_J\end{bmatrix}\begin{bmatrix}\mathbf{s}(2n)\\ \mathbf{s}(2n+1)\end{bmatrix}. \tag{10}$$

Suppose that two known symbols p_1 and p_2 are placed inside two consecutive blocks $\mathbf{s}(2m)$ and $\mathbf{s}(2m+1)$ at position k. Letting $\grave{s}_1 := [\grave{\mathbf{s}}(2m)]_k$ and $\grave{s}_2 := [\grave{\mathbf{s}}(2m+1)]_k$; we obtain:

$$\begin{bmatrix}\grave{s}_1\\ \grave{s}_2\end{bmatrix} = \frac{1}{|\alpha_1|^2+|\alpha_2|^2}\begin{bmatrix}\alpha_1^* & \alpha_2^*\\ -\alpha_2 & \alpha_1\end{bmatrix}\begin{bmatrix}p_1\\ p_2\end{bmatrix}, \tag{11}$$

from which α_1 and α_2 can be solved as:

$$\begin{bmatrix}\alpha_1\\ \alpha_2^*\end{bmatrix} = \frac{1}{|\grave{s}_1|^2+|\grave{s}_2|^2}\begin{bmatrix}\grave{s}_1^* & \grave{s}_2\\ -\grave{s}_2^* & \grave{s}_1\end{bmatrix}\begin{bmatrix}p_1\\ p_2^*\end{bmatrix}. \tag{12}$$

With α_1 and α_2 resolved, the true channels can then be found from (6). However, this step is not necessary since the symbol estimates can be obtained directly from (10) by:

$$\begin{bmatrix}\hat{\mathbf{s}}(2n)\\ \hat{\mathbf{s}}(2n+1)\end{bmatrix} = \begin{bmatrix}\alpha_1\mathbf{I}_J & -\alpha_2^*\mathbf{I}_J\\ \alpha_2\mathbf{I}_J & \alpha_1^*\mathbf{I}_J\end{bmatrix}\begin{bmatrix}\grave{\mathbf{s}}(2n)\\ \grave{\mathbf{s}}(2n+1)\end{bmatrix}. \tag{13}$$

With distinct precoders, we should equalize $\mathbf{z}(2n)$ by $(\bar{\mathcal{D}}_{34}\Theta_1)^\dagger$ and $\mathbf{z}(2n+1)$ by $(\bar{\mathcal{D}}_{34}\Theta_2)^\dagger$. Substituting $\alpha = \alpha_1$ and $\alpha_2 = 0$ in (12), the scalar α can be figured out as:

$$\alpha = (\grave{s}_1^* p_1 + \grave{s}_2 p_2^*)/(|\grave{s}_1|^2 + |\grave{s}_2|^2). \tag{14}$$

Similarly, if $|p_1| = |p_2|$ and thus $|\grave{s}_1| = |\grave{s}_2|$, eq. (14) can be further simplified to $\alpha = (1/2)\,(p_1/\grave{s}_1 + p_2^*/\grave{s}_2^*)$.

Remark (advantage of distinct over identical precoders): As indicated by Theorems 1 and 2, with distinct precoders Θ_1 and Θ_2 the channels can be identified up to one scalar α instead of two scalars (α_1, α_2) that must be determined with identical precoders $\Theta_1 = \Theta_2 = \Theta$. With one pair of known symbols (p_1, p_2), the residual scalar ambiguities can be resolved by (12) and (14) for pre-precoding pilots. Therefore, the advantage of distinct precoders over identical precoders is not clearly justified since two scalars are also not difficult to resolve for identical precoders as in (12). However, the noise analysis we detail in [9] for the scalar ambiguity determination of (12) and (14) reveals that distinct precoders lead to a 3dB gain over identical precoders for suppressing the channel estimation error caused by the imperfectly resolved scalar ambiguities. To achieve the same channel estimation accuracy, identical precoders need to employ twice the number of pilots relative to distinct precoders.

In a nutshell, designing distinct precoders instead of identical precoders pays off either in terms of increasing the system efficiency by using half the number of pilots, or, in terms of improving the system performance with the same number of pilots, a feature that we also verified by simulations.

6. SIMULATIONS

To test the proposed channel estimation algorithm we use as figures of merit the averaged Normalized Mean Square Error (NMSE) of the channels defined as: $(1/2)\sum_{i=1}^{2}\|\hat{\tilde{\mathbf{h}}}_i - \tilde{\mathbf{h}}_i\|^2/\|\tilde{\mathbf{h}}_i\|^2$, and the Bit Error Rate (BER). We set the system parameters as: $L = 8$, $K = 3L$, $J = K + L = 32$; and generate the channels according to the Channel Model

A specified by ETSI. We assume here that each data burst has $N = 400$ symbol blocks, in which the first one $\breve{s}(0)$ is not precoded and serves as a training block. The semi-blind channel estimator is implemented using (5) with $W = 100$ and is initialized using the training based method of [2]. The channel estimates are updated every 50 blocks in order to render the complexity reasonable. To resolve the residual scalar ambiguities, $N_p = 1, 2, 4$ pairs of pre-precoding pilots are distributed inside each set of 50 symbol blocks.

To illustrate the advantage of distinct over identical precoders, we depict in Fig. 2 the NMSE averaged over the entire data burst with different number of pilots employed. From Fig. 2, we infer that indeed at high SNR identical precoders need to double the number of pilots to be able to catch up with the performance of distinct precoders, which is consistent with our noise analysis in [9]. To check the overall performance of channel estimation, equalization and ST decoding, we plot in Fig. 3 the BERs averaged over the entire data burst with ZF equalizers constructed from different channel estimates. Compared to the benchmark BER performance obtained with perfect channel knowledge at the receiver, our semi-blind channel estimator only incurs less than 2 dB SNR loss, while a high error floor is observed for the training based approach since the channels are time varying and no tracking mechanism has been invoked.

To illustrate the advantage of channel acquisition and coherent detection at the receiver, we also plot in Fig. 3 the BER performance of a competing differential ST-OFDM alternative, where the differential encoding of [7] is applied on each subcarrier to dispense with channel estimation. To make up for the same information rate, convolutional coding with rate 3/4 ($= K/J$) is also tested for differential ST-OFDM. Since the differential decoder output takes binary values [7], the Viterbi decoding algorithm with hard decision is applied here. Without assuming any side channel information, the path metric for Viterbi decoding is set to be the Hamming distance between the received bit stream at the output of differential decoder (denoted by $\hat{c}_1, \ldots, \hat{c}_n$, where $\hat{c}_i \in [0, 1]$) and the possible codeword candidates (denoted by $c_1, \ldots, c_n$). If side information on the channel fading coefficients $\{f_i^2\}_{i=1}^n$, where $f_i^2 := |H_1(\rho_i)|^2 + |H_2(\rho_i)|^2$, can be acquired at the receiver, the path metric could be modified using the weighted Hamming distance: $\sum_{i=1}^n f_i^2(\hat{c}_i - c_i)^2$. Fig. 3 demonstrates that precoded ST-OFDM equipped with our semi-blind channel estimator outperforms the differential ST-OFDM considerably, for both uncoded and coded transmissions at the considered SNR range.

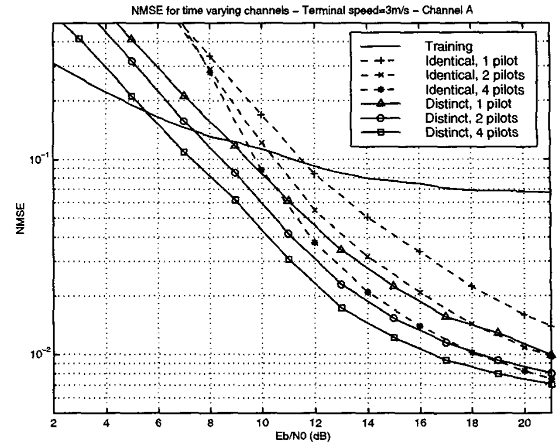

Fig. 2. Channel NMSE versus SNR

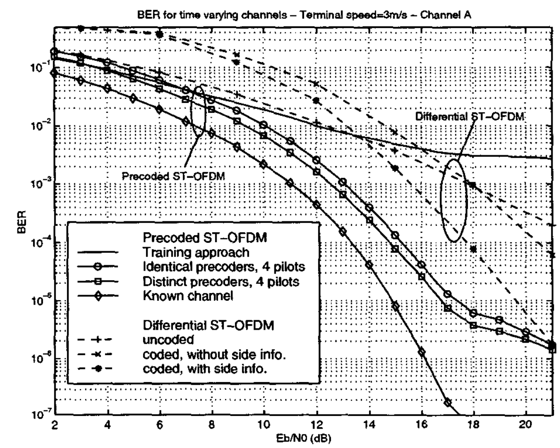

Fig. 3. BER versus SNR comparisons

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