

New OSTBC for Blind Channel Estimation and Tracking in MIMO-OFDM Systems

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Abstract—Applying orthogonal space time block coding (OSTBC) to multiple-input multiple-output (MIMO) systems helps reduce receiver complexity. However, this approach has been applied only to flat fading channels, as when the channel is a frequency selective fading MIMO channel, OSTBC cannot be used directly since its orthogonal propriety may be lost. Furthermore, the MIMO channel is not always known. To deal with this problem, many techniques were proposed to estimate the channel using a training sequence. Unfortunately, these techniques reduce the useful spectral bandwidth. This paper proposes OSTBC for blind channel estimation and data detection in the case of a MIMO frequency selective channel. The aim of this new OSTBC is twofold: to solve the ambiguity of channel estimation and to reduce the complexity of the detector. By exploiting the well-known technique of orthogonal frequency division multiplexing (OFDM), the frequency selective fading MIMO channel is split into a set of flat fading sub-channels. Moreover, to accommodate the fact that a MIMO channel can be time varying, the steady state Kalman channel estimator (SS-KCE) is extended to track the channel's fast variations. The performance of the proposed blind algorithm is related by the adequate choice of the number of subcarriers and it is compared with other existing approaches by means of Monte Carlo simulations.

Keywords—MIMO, OFDM, blind channel estimation, OSTBC, SS-KCE.

1. Introduction

The growing interest in wireless mobile communication requires high performance and quality communication. This pushes researchers to develop many techniques with the aim of increasing data transmission rates. Multi-input multi-output (MIMO) is one of the techniques proposed to boost useful channel capacity without increasing total transmission power [1]. In addition, space-time block coding (STBC) has appeared as a powerful approach to exploit spatial diversity and to combat fading in MIMO wireless channels [2], [3].

The higher the data rate, the more the wireless channel becomes frequency selective and the more Inter Symbol

Interference (ISI) is encountered. To eliminate the ISI, orthogonal frequency division multiplexing (OFDM) is used in many standards, such as those pertaining to broadband wireless metropolitan area networks (IEEE 802.16) and mobile broadband wireless access networks (IEEE 802.20) [4]. The OFDM technique has some drawbacks too, such as its peak-to-average power ratio (PAPR) [5], but it still remains a candidate for 5G after introducing some modifications, such as filter bank multiple carrier (FBMC) [6] or generalized frequency division multiplexing (GFDM) [6]. The basic idea behind the OFDM technique consists in converting a frequency selective channel into a bank of flat fading sub-channels [7]. In addition, a combination between orthogonal STBC and OFDM is sought, because it allows to reach the maximum transmit diversity and simplifies the receiver structure [8].

Besides being time variable, a wireless channel may also be unknown to both the receiver and the transmitter. Generally, a periodic training sequence is transmitted to identify this wireless channel, but, unfortunately, this sequence occupies nearly 20% of the useful data rate. Here, in order to identify a wireless channel, we focus our interest on the blind channels which require no training pilot. Many research works dealing with channel estimation in MIMO-OFDM systems can be cited. For instance, paper [9] proposed a subspace blind channel estimation method for MIMO-OFDM systems, based on frequency correlation between adjacent subcarriers. Also, Sarmadi *et al.* in [10] proposed a semi-definite relaxation technique to achieve blind channel estimation for orthogonal STBC-OFDM systems. Another method was also proposed in [11]. It is based on the weighted covariance approach which explores all redundancies contained in the orthogonal code to resolve channel estimation non-scalar ambiguities. In [12], Zhang *et al.* developed a new subspace channel estimation method for OSTBC with MIMO-OFDM, using the block cyclic property of the channel matrix and the cyclic repetition method (CRM). However, this method needs a few pilots to alleviate channel estimation ambiguity. A new approach was developed by Jiang *et al.* in [8], where the

Table 1
List of abbreviations

A	Diagonal matrix controls channel coefficients
$\hat{B}^{(\ell,q)}$	Block contains codeword symbols corresponding to channel estimating matrix $\hat{\mathcal{H}}^{(\ell)} [u]$
$\{C\}$	Set of modulation constellation points
F	Transition diagonal matrix
f_0	Carrier frequency
$F_D T_s$	Normalized Doppler rate
G	Generator matrix
g	Discrete modulation waveform
H	Channel matrix of a frequency Rayleigh selective fading
\mathcal{H}	Orthogonal matrix groups channel attenuation effect and coding effect
$\hat{\mathcal{H}}$	Estimated matrix channel
\mathbf{h}_{rt}	Sub-channel between the t -th transmitting antenna and the r -th receiving antenna
\hat{h}_{rj}	FFT of the sub-channel \mathbf{h}_{rj}
\hat{h}	Vector of the matrix \mathcal{H}
$I_{N_r N_t}$	$N_r \times N_t$ unitary matrix
J_0	Zero-order Bessel function of the first kind
k	Block OFDM number
L	Frequency sub-channel length
ℓ, ℓ_1, ℓ_2 and n	Integer numbers
m	Total number of constellation points
N	OFDM subcarrier number
N_g	Length of cyclic prefix
N_t	Transmitting antennas
N_r	Receiving antennas
n_s	Energy of each uncoded data block
P_∞	Steady state estimation error covariance matrix
$\text{Re}(\cdot)$	Real part
s_i	Transmitted symbol with $i = 1, 2, \dots, 4N$
\hat{s}	Estimated data symbol
\tilde{s}	Final detected data symbol
T	Duration of codeword block with $T = 4N \times T_s$
T_s	Symbol duration
y_r	Channel output signal of received antenna (r)
w_r	Noise signal of receiver antenna (r)
\hat{w}_r	FFT of the noise w_r
w	Measurement noise vector
R	Data rate
R_w	Covariance matrix of measurement noise
η	Process noise vector
$z_r^{(v)}$	Output of OFDM demodulator (r) at instant (v)
σ_n^2	Variance of noise w_r
β	Transition coefficient
$\theta_{j,j+1}$	Phase factor $j = 1, 3, \dots, 2N - 1$
φ_ℓ	Phase belonging to a pre-determined set of m^2 different phases
$(\cdot)^H$	Conjugate transpose
$(\cdot)^T$	Transpose
$(\cdot)^*$	Complex conjugate
$E\{\cdot\}$	Expectation
\star	Convolution operation
$\ \cdot\ _F$	Frobenius norm
\otimes	Kronecker product

MIMO-OFDM channel is estimated by a noise subspace method, exploiting the null space induced by the OSTBC. The work in [13] presented an EVD decomposition of the signal covariance matrices from a group of carefully selected subcarriers to identify the channel with a low order signal constellation.

All these algorithms need a long sequence of data for the receiver to converge, which leads to high complexity. In this paper, a blind method is proposed, based on a new OSTBC which has a very low complexity and which does not need a long sequence to obtain an optimized estimation of the channel. This orthogonal code allows to estimate blind MIMO channels and reduce the complexity of the detector.

The remainder of the paper is organized as follows: in Section 2, a model of the MIMO-OFDM system with the proposed OSTBC is described. Section 3 presents the proposed blind estimation method tracking the time variation of a MIMO-OFDM channel. Section 4 is devoted to a detailed description of the data detector. Simulation results are presented and discussed in Section 5. Finally, some conclusions are drawn in Section 6.

2. MIMO-OFDM System Model

The proposed OSTBC is based on a MIMO-OFDM system model with $N_t = 2$ transmitting antennas, $N_r = 2$ receiving antennas and N subcarriers, as shown in Fig. 1. It is assumed that the channel order does not exceed the cyclic prefix length and remains constant over 4 successive OFDM symbols (codeword length equal to $4N$). In the blind environment, where no prior knowledge about channel information state is available at the receiver, an initial blind channel estimator is used to initiate the steady state Kalman channel estimator (SS-KCE) that refines estimation of the time changing coefficients of the matrix channel, and tracks them. Combining OSTBC and a MIMO-OFDM system can significantly simplify the structure of the receiver and yields a maximum transmit diversity.

The output of the MIMO-OFDM channel from Fig. 1 is expressed as:

$$\begin{aligned} y_1[n] &= \mathbf{h}_{11} \star x_1[n] + \mathbf{h}_{12} \star x_2[n] + w_1[n] \\ y_2[n] &= \mathbf{h}_{21} \star x_1[n] + \mathbf{h}_{22} \star x_2[n] + w_2[n] \end{aligned} \quad (1)$$

where $y_r[n]$ ($r = 1, 2$) is the channel output at time instant $[n]$, $w_r[n]$ ($r = 1, 2$) is the noise at time instant $[n]$, assumed to be a zero mean additive white Gaussian Noise (AWGN) with variance σ_n^2 , $\mathbf{h}_{rt} = [h_{rt}^{(0)}, \dots, h_{rt}^{(L-1)}]$ ($r = 1, 2$ and $t = 1, 2$) is the L length frequency selective sub-channel between the t -th transmitting antenna and the r -th receiving antenna.

It is assumed that the channel matrix \mathbf{H} , as defined in Eq. (2), is that of a frequency Rayleigh selective fading MIMO channel that is constant or presents no significant change during the transmitted frame of $4N$ instants.

$$\mathbf{H} = \begin{pmatrix} \mathbf{h}_{11} & \mathbf{h}_{12} \\ \mathbf{h}_{21} & \mathbf{h}_{22} \end{pmatrix}. \quad (2)$$

The transmitted signal $x_t[n]$, from the t -th ($t = 1, 2$) transmitting antenna at time instant $[n]$ is:

$$x_t[n] = \text{Re} \left(e^{2j\pi f_0 n T_s} \sum_{u=1}^{N+N_g} I_t^+[u] \cdot g[n-u] T_s \right), \quad t = 1, 2, \quad (3)$$

where $I_t^+[u]$ is the u -th element of the vector I_t^+ , formed by the output vector from the N length IFFT block, to which is appended a cyclic prefix (CP) of length N_g . Subscript t refers to the t -th antenna.

For more details about the OFDM technique, see [14]–[20].

2.1. Proposed OSTBC

The existing OSTBC solutions were applied to MIMO systems only in the case of a flat fading channel, to ensure that its orthogonal property is preserved. This property is important for reconstructing the transmitted data and reducing complexity of the detector.

In this paper, we propose an OSTBC approach that works even for a frequency selective channel. This code is developed for a $(N_t, N_r) = (2, 2)$ MIMO system and an OFDM system of N subcarriers, as shown in Fig. 2.

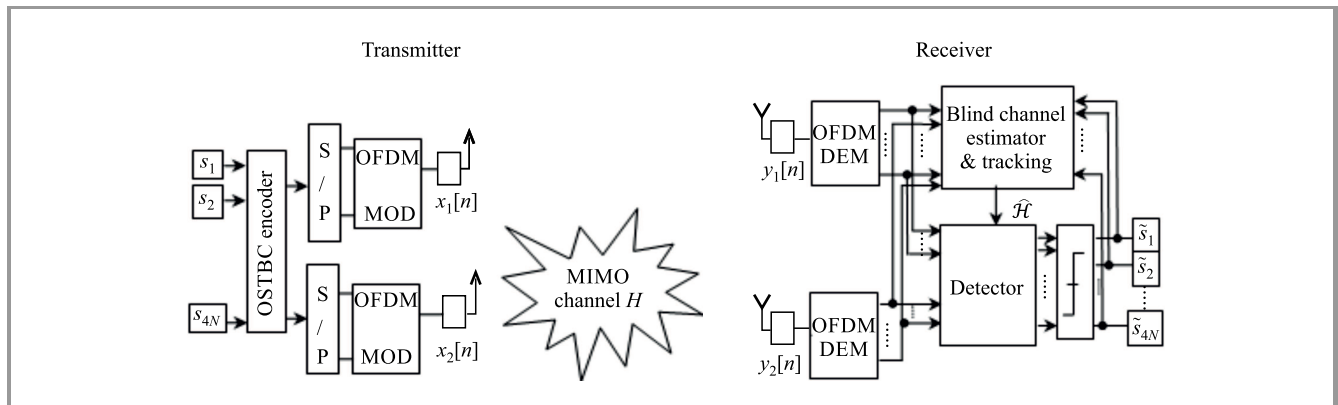


Fig. 1. Block diagram for blind estimation and detection of a MIMO-OFDM system based on the proposed OSTBC.

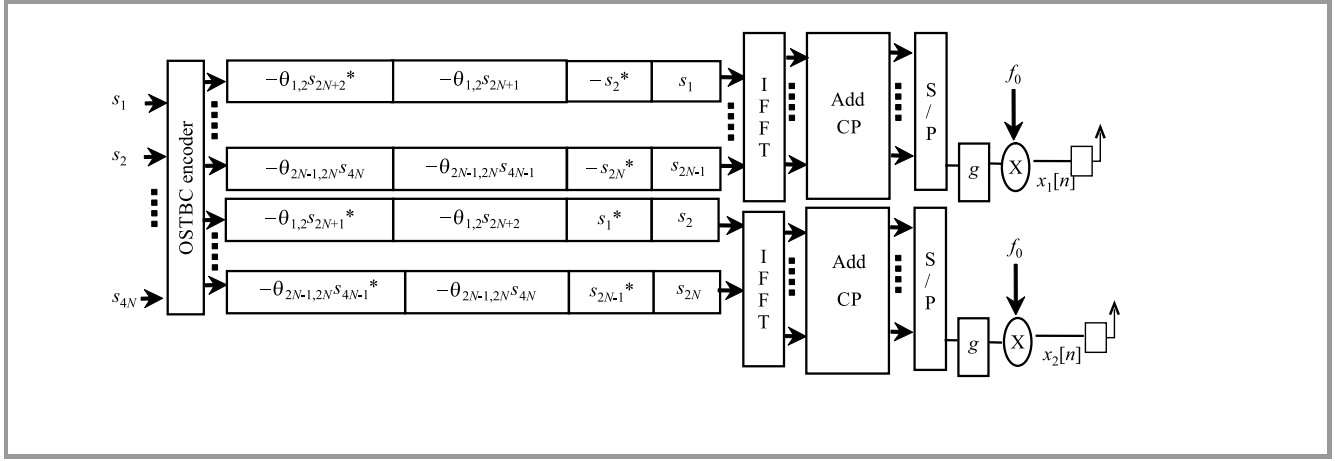
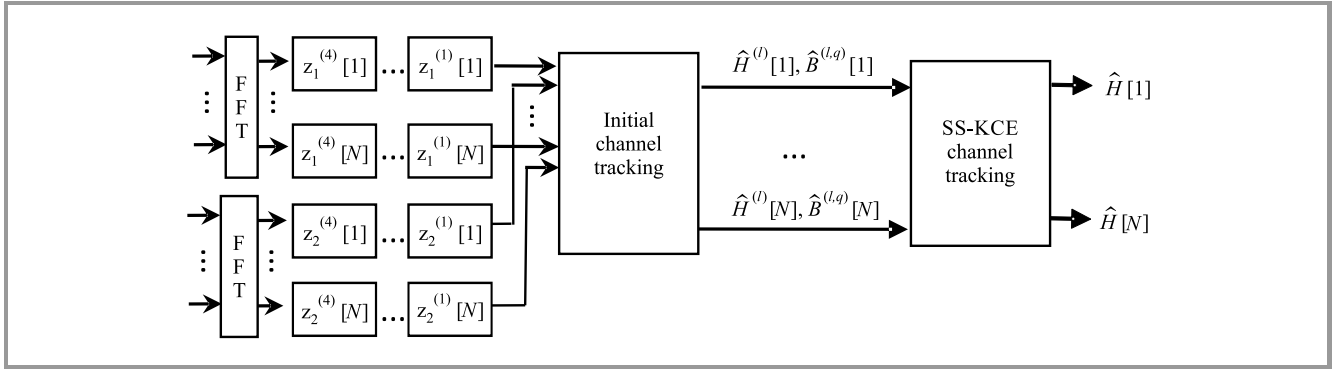

 Fig. 2. OSTBC encoder with OFDM system with N sub-carriers.


Fig. 3. Estimation and tracking of MIMO-OFDM channel.

The generator matrix G of the proposed OSTBC is given by:

$$G = \begin{bmatrix} s_1 & s_2 \\ \vdots & \vdots \\ s_{2N-1} & s_{2N} \\ -s_2^* & s_1^* \\ \vdots & \vdots \\ -s_{2N}^* & s_{2N-1}^* \\ \vdots & \vdots \\ \theta_{2N-1,2N} s_{4N-1} & \theta_{2N-1,2N} s_{4N} \\ -\theta_{1,2} s_{2N+2}^* & \theta_{1,2} s_{2N+1}^* \\ \vdots & \vdots \\ -\theta_{2N-1,2N} s_{4N}^* & \theta_{2N-1,2N} s_{4N-1}^* \end{bmatrix}_{4N \times 2}, \quad (4)$$

where s_i ($i = 1, 2, \dots, 4N$) is the transmitted symbol that belongs to a constellation of m points, $\theta_{j,j+1} = e^{j\varphi_l}$, $j = 1, 3, \dots, 2N-1$ and $l \in \{1, 2, \dots, m^2\}$ is a phase factor, with φ_l being a phase belonging to a pre-determined set of m^2 different phases.

Each possible combination of $\begin{bmatrix} s_j \\ s_{j+1} \end{bmatrix}$ has a unique phase factor $\theta_{j,j+1}$. The choice of the pre-determined set of m^2 different phases should respect the following condition:

$$\varphi_{\ell_1} \neq \varphi_{\ell_2} + n\pi \quad \forall \ell_1 \neq \ell_2, \quad \{\ell_1, \ell_2\} \in \{1, 2, \dots, m^2\}. \quad (5)$$

An adequate choice of the phase factor can solve the ambiguity problem during the channel estimation step.

The generator matrix G is composed of four parts. The first two parts (column-wise) are used by the receiver to find all possible combinations for channel matrix estimation. The remaining parts (with the phase factors) are used to eliminate ambiguity in finding the matrix channel when the minimum square error (MSE) criterion is applied. This code is orthogonal, since we have:

$$G^H \times G = \sum_{i=1}^{4N} \|s_i\|^2 I_2. \quad (6)$$

3. Blind Channel Estimation and Tracking

The proposed blind estimator of the MIMO-OFDM channel comprises two parts: initial blind channel estimation algorithm and decoding. It works only for the first received codeword block to initialize the SS-KCE algorithm that tracks the fast channel's variations for the remaining codeword stream.

After the OFDM demodulator, which includes S/P, CP removal, and FFT (see Fig. 3), the signal $z_r^{(v)}$ from the r -th receiving antenna at instant (v) may be expressed as [21]:

$$z_r^{(v)}[u] = s_1^{(v)}[u]h_{r1}[u] + s_2^{(v)}[u]h_{r2}[u] + w_r^{(v)}[u];$$

$$r = 1, 2; u = 1, 2, \dots, N; v = 1, 2, \dots, 4, \quad (7)$$

where $h_{r1}[u]$, $h_{r2}[u]$ are the u -th coefficients of the FFT of the sub-channels \mathbf{h}_{r1} and \mathbf{h}_{r2} , respectively, $w_r[u]$ is the u -th coefficient of the FFT of the noise $w_r[n]$, and $s_1^{(v)}[u]$ and $s_2^{(v)}[u]$ correspond to the upper and lower coded data from the OSTBC encoder.

3.1. Initial Blind Estimation of the MIMO-OFDM Channel

Using the first data block received, the blind algorithm based on MSE criteria provides an initial estimation of the channel's FFT coefficient matrices, together with the most probable transmitted codeword block. The two outputs of this algorithm are used to initialize SS-KCE that tracks variations of the channel's coefficients.

$$\begin{aligned} & s_1^{(1)}[u]^* \times z_1^{(1)}[u] - s_2^{(1)}[u] \times z_1^{(2)}[u] \\ &= h_{11}[u] \left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right) \\ &+ s_1^{(1)}[u]^* \times w_1^{(1)}[u] - s_2^{(1)}[u] \times w_1^{(2)}[u]; \\ & s_2^{(1)}[u]^* \times z_2^{(1)}[u] + s_1^{(1)}[u] \times z_1^{(2)}[u] \\ &= h_{12}[u] \left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right) \\ &+ s_2^{(1)}[u]^* \times w_1^{(1)}[u] + s_1^{(1)}[u] \times w_1^{(2)}[u]; \\ & s_1^{(1)}[u]^* \times z_2^{(1)}[u] - s_2^{(1)}[u] \times z_2^{(2)}[u] \\ &= h_{21}[u] \left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right) \\ &+ s_1^{(1)}[u]^* \times w_2^{(1)}[u] - s_2^{(1)}[u] \times w_2^{(2)}[u]; \\ & s_2^{(1)}[u]^* \times z_2^{(1)}[u] + s_1^{(1)}[u] \times z_2^{(2)}[u] \\ &= h_{22}[u] \left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right) \\ &+ s_2^{(1)}[u]^* \times w_2^{(1)}[u] + s_1^{(1)}[u] \times w_2^{(2)}[u]; \end{aligned} \quad (8)$$

By neglecting the noise effect, an estimate of the matrix channel $\hat{\mathcal{H}}[u]$ may be obtained by:

$$\begin{aligned} \hat{h}_{11}[u] &= \frac{s_1^{(1)}[u]^* \times z_1^{(1)}[u] - s_2^{(1)}[u] \times z_1^{(2)}[u]}{\left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right)}; \\ \hat{h}_{12}[u] &= \frac{s_2^{(1)}[u]^* \times z_1^{(1)}[u] + s_1^{(1)}[u] \times z_1^{(2)}[u]}{\left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right)}; \\ \hat{h}_{21}[u] &= \frac{s_1^{(1)}[u]^* \times z_2^{(1)}[u] - s_2^{(1)}[u] \times z_2^{(2)}[u]}{\left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right)}; \\ \hat{h}_{22}[u] &= \frac{s_2^{(1)}[u]^* \times z_2^{(1)}[u] + s_1^{(1)}[u] \times z_2^{(2)}[u]}{\left(\|s_1^{(1)}[u]\|^2 + \|s_2^{(1)}[u]\|^2 \right)}; \end{aligned} \quad (9)$$

Since $s_1^{(1)}[u]$ and $s_2^{(1)}[u]$ are unknown, there are m^2 possibilities for $\hat{\mathcal{H}}[u]$:

$$\hat{\mathcal{H}}^{(\ell)}[u] = \begin{bmatrix} \hat{h}_{11}^{(\ell)}[u] & \hat{h}_{12}^{(\ell)}[u] \\ \hat{h}_{21}^{(\ell)}[u] & \hat{h}_{22}^{(\ell)}[u] \end{bmatrix}, \quad (10)$$

$$\begin{aligned} \hat{h}_{11}^{(\ell)}[u] &= \frac{s_{\ell_1}^* z_1^{(1)}[u] - s_{\ell_2} z_1^{(2)}[u]}{\left(\|s_{\ell_1}\|^2 + \|s_{\ell_2}\|^2 \right)}; \\ \hat{h}_{12}^{(\ell)}[u] &= \frac{s_{\ell_2}^* z_1^{(1)}[u] + s_{\ell_1} z_1^{(2)}[u]}{\left(\|s_{\ell_1}\|^2 + \|s_{\ell_2}\|^2 \right)}; \\ \hat{h}_{21}^{(\ell)}[u] &= \frac{s_{\ell_1}^* z_2^{(1)}[u] - s_{\ell_2} z_2^{(2)}[u]}{\left(\|s_{\ell_1}\|^2 + \|s_{\ell_2}\|^2 \right)}; \\ \hat{h}_{22}^{(\ell)}[u] &= \frac{s_{\ell_2}^* z_2^{(1)}[u] + s_{\ell_1} z_2^{(2)}[u]}{\left(\|s_{\ell_1}\|^2 + \|s_{\ell_2}\|^2 \right)}; \end{aligned} \quad (11)$$

with $\ell = 1, \dots, m^2$, $\{s_{\ell_1}, s_{\ell_2}\} \in \{C\}$, and $u = 1, 2, \dots, N$.

To obtain the right estimate of the channel matrix and, at the same time, detect the four symbols $\{s_{\ell_1}, s_{\ell_2}, s_{q_1}, s_{q_2}\}$, of the transmitted codeword, we apply the following MSE criterion:

$$\begin{aligned} & \langle \hat{\mathcal{H}}^{(\ell)}[u], \hat{\mathbf{B}}^{(\ell,q)} \rangle \geq \operatorname{argmin}_{(\hat{\mathcal{H}}^{(\ell)}[u], \hat{\mathbf{B}}^{(\ell,q)})} \|Z[u] \\ & - \hat{\mathcal{H}}^{(\ell)}[u] \hat{\mathbf{B}}^{(\ell)}[u] \hat{\mathbf{B}}^{(\ell,q)}\|_F^2; \\ Z[u] &= \begin{bmatrix} Z_1^{(1)}[u] & Z_1^{(2)}[u] & Z_1^{(3)}[u] & Z_1^{(4)}[u] \\ Z_2^{(1)}[u] & Z_2^{(1)}[u] & Z_2^{(1)}[u] & Z_2^{(1)}[u] \end{bmatrix}; \\ \hat{\mathbf{B}}^{(\ell,q)} &= \begin{bmatrix} s_{\ell_1} & -s_{\ell_1}^* & \theta_{\ell} s_{q_1} & -\theta_{\ell} s_{q_2}^* \\ s_{\ell_2} & s_{\ell_2}^* & \theta_{\ell} s_{q_2} & \theta_{\ell} s_{q_1}^* \end{bmatrix}; \end{aligned} \quad (12)$$

$\{s_{\ell_1}, s_{\ell_2}, s_{q_1}, s_{q_2}\} \in \{C\}$; $u = 1, 2, \dots, N$; $\ell, q = 1, \dots, m^2$.

The value of θ_{ℓ} is related to the pair of symbols $\begin{bmatrix} s_{\ell_1} \\ s_{\ell_2} \end{bmatrix}$ transmitted by the two antennas.

3.2. SS-KCE for MIMO-OFDM Channel Tracking

The SS-KCE was proposed for tracking the variations of a flat fading channel in an OSTBC-MIMO system [22]. It has the advantage of performing nearly as good as the optimal time-varying Kalman estimator with reduced complexity [23], [24].

In this paper, the property of an OFDM system, such as frequency selective MIMO channel which may be modeled by a set of N parallel flat fading channels is exploited to apply the SS-KCE in the proposed OSTBC-MIMO-OFDM approach.

Two formulas are required for channel tracking: the process equation and the measurement equation [23], [25], [26]. The former describes the dynamic behavior of the state variables to be estimated, while the latter captures the relationship between the output system and the state variables.

In the case concerned, Eq. (13) may be considered as the process equation:

$$\hat{h}_{(k)}[u] = F\hat{h}_{(k-1)}[u] + A\eta[u]; \quad u = 1, 2, \dots, N, \quad (13)$$

where:

- $\hat{h}_{(k)}[u] = \text{vec}(\mathcal{H}_{(k)}[u])$, vec being the operator that places the columns of a matrix on top of each other. $\hat{h}_{(k)}[u]$ is assumed to be nearly constant during a frame with the length $4N$ symbols;
- $\eta[u]$ is a vector of length $N_r \times N_t$ containing independent samples of a circularly symmetric zero-mean FFT of Gaussian noise, with the covariance matrix $R_\eta = I_{N_r N_t}$.
- $F = \beta I_{N_r N_t}$ is a transition diagonal matrix with $\beta = J_0(2\pi f_D T_s)$. J_0 is the first kind zero-order Bessel function $f_D T_s$ is the normalized Doppler rate (assumed to be the same for all transmitting-receiving antenna pairs), and T_s is the baud duration [28].

According to the widely used wide-sense stationary uncorrelated scattering (WSSUS) model [27], the FFT channel coefficients are modeled as independent, zero-mean, complex Gaussian random variables with the auto-correlation matrix given by:

$$E[\hat{h}_{(k)}[u] \hat{h}_{(k+1)}[u]^{\mathcal{H}}] \approx F^{\mathcal{H}}; \quad u = 1, 2, \dots, N. \quad (14)$$

From Eq. (13), the rate of the channel variations is fixed by transition matrix F and the amplitude of all $\hat{h}_{(k)}[u]$ is controlled by matrix A . To be sure that the correlation matrix of $\hat{h}_{(k)}[u]$ is unitary, matrix A should be taken as $A = \sqrt{1 - \beta^2} I_{N_r N_t}$ [24].

For instance, if we have a mobile terminal that moves with the velocity of $v_0 = 90$ km/h, carrier frequency of $f_0 = 2.3$ GHz, and data rate of $R = 4N \times 0.2$ Mb/s, the maximum Doppler frequency is $f_D = f_0 v_0 / c$, where c is the speed of light. The normalized Doppler rate is $f_D T \approx 0.001$, $T = 4N \times T_s$ where $R = 1/T_s$ and N is the number of subcarriers. In this case, according to the Kalman filtering terminology, $\hat{h}_{(k)}[u]$ represents the state vector and the FFT of the channel output, $Z[u]$, defined in Eq. (12), represents the measurement. By stacking the columns of matrices $Z[u]$, $\mathcal{H}_{(k)}[u]$ and $W[u]$, the following measurement equation can be obtained from Eq. (7):

$$z[u] = S[u]\hat{h}_{(k)}[u] + w[u], \quad u = 1, 2, \dots, N, \quad (15)$$

where: $S[u] = S[u]^T \otimes I_{N_r}$,

$$S[u] = \begin{bmatrix} s_j & -s_{j+1}^* & \theta_{j,j+1} \times s_{2N+j} & -\theta_{j,j+1} \times s_{2N+j+2}^* \\ s_{j+1} & s_j^* & \theta_{j,j+1} \times s_{2N_j+1} & \theta_{j,j+1} \times s_{2N_j+2}^* \end{bmatrix},$$

$j = 1, 3, \dots, 2N - 1$, $u = 1, 2, \dots, N$,

$w[u]$ is the measurement noise with a covariance matrix $R_w = \sigma_w^2 I$.

Then, Eqs. (13) and (15) represent the state-space formulation of the problem of estimating a flat varying MIMO channel. Both are linear functions of state vector $\hat{h}_{(k)}[u]$ and noises $w[u]$ and $\eta[u]$ are independent, white and Gaussian, the Kalman filter provides the optimal recursive estimates in the MMSE sense of the channel coefficients [23], [24]. SS-KF proposed by [22] is extended and used in the present work to track channel variations. In this filter, the channel coefficients are updated by:

$$\hat{h}_{k/k}[u] = \beta B_\infty \hat{h}_{k-1/k-1}[u] + \frac{1}{n_s} A_\infty S[u]^{\mathcal{H}} z[u], \quad u = 1, 2, \dots, N. \quad (16)$$

where $A_\infty = P_\infty \left(\frac{\sigma_w^2}{n_s} I_{N_r N_t} + P_\infty \right)^{-1}$, $B_\infty = I_{N_r N_t} - A_\infty$.

The steady-state value of the estimation error covariance matrix P_∞ can be obtained by solving the following discrete algebraic Riccati equation (DARE) [23], [24]:

$$P_\infty = \beta^2 P_\infty - \beta^2 P_\infty \left(P_\infty + \frac{\sigma_w^2}{n_s} I_{N_r N_t} \right)^{-1} P_\infty + \sigma_w^2 I_{N_r N_t}, \quad (17)$$

where n_s represents the energy, supposed to be constant for each uncoded data block.

4. Data Detection

Exploiting the orthogonal property of the proposed OSTBC simplifies the detection process [29]. In this case, estimates \hat{s}_j and \hat{s}_{j+1} are decoupled and can be obtained as:

The u -th channel output can be written as:

$$\begin{bmatrix} z_1^{(1)}[u] & z_1^{(2)}[u] \\ z_2^{(1)}[u] & z_2^{(2)}[u] \end{bmatrix} = \begin{bmatrix} h_{11}[u] & h_{12}[u] \\ h_{21}[u] & h_{22}[u] \end{bmatrix} \begin{bmatrix} s_j & -s_{j+1}^* \\ s_{j+1} & s_j^* \end{bmatrix} + \begin{bmatrix} w_1^{(1)}[u] & w_1^{(2)}[u] \\ w_2^{(1)}[u] & w_2^{(2)}[u] \end{bmatrix}, \quad u = 1, 2, \dots, N. \quad (18)$$

Equation (18) may be rewritten as:

$$\begin{bmatrix} z_1^{(1)}[u] \\ z_2^{(1)}[u] \\ z_1^{(2)*}[u] \\ z_2^{(2)*}[u] \end{bmatrix} = \mathcal{H}[u] \times \begin{bmatrix} s_j \\ s_{j+1} \end{bmatrix} + \begin{bmatrix} w_1^{(1)}[u] \\ w_2^{(1)}[u] \\ w_1^{(2)*}[u] \\ w_2^{(2)*}[u] \end{bmatrix}, \quad (19)$$

where $\mathcal{H}[u]$ is an orthogonal matrix, which groups together two effects: channel attenuation and coding. Its transpose conjugate is given by:

$$\mathcal{H}^{\mathcal{H}}[u] = \begin{bmatrix} \hat{h}_{11}^* & \hat{h}_{21}^*[u] & \hat{h}_{12}[u] & \hat{h}_{22}[u] \\ \hat{h}_{12}^* & \hat{h}_{22}^*[u] & -\hat{h}_{11}[u] & -\hat{h}_{21}[u] \end{bmatrix}. \quad (20)$$

The estimate of $\begin{bmatrix} s_j \\ s_{j+1} \end{bmatrix}$ can be obtained from:

$$\begin{bmatrix} \hat{s}_j \\ \hat{s}_{j+1} \end{bmatrix} = \mathcal{H}^{\mathcal{H}c}[u] \times \begin{bmatrix} z_1^{(1)}[u] \\ z_2^{(1)}[u] \\ z_1^{(2)*}[u] \\ z_2^{(2)*}[u] \end{bmatrix}. \quad (21)$$

From this estimate, an estimate of the factor $\theta_{j,j+1}[u]$ is deduced and used to get $\begin{bmatrix} \hat{s}_{j+2N} \\ \hat{s}_{j+2N+1} \end{bmatrix}$:

$$\begin{bmatrix} \hat{s}_{j+2N} \\ \hat{s}_{j+2N+1} \end{bmatrix} = [\hat{\theta}_{j,j+1}[u]]^{-1} \mathcal{H}^{\mathcal{H}c}[u] \begin{bmatrix} z_1^{(3)}[u] \\ z_2^{(3)}[u] \\ z_1^{(4)*}[u] \\ z_2^{(4)*}[u] \end{bmatrix}. \quad (22)$$

A threshold detector is applied to get an approximation of the transmitted data symbols:

$$\begin{bmatrix} \hat{s}_j \\ \hat{s}_{j+1} \end{bmatrix} = \mathcal{Q}\left(\begin{bmatrix} \hat{s}_j \\ \hat{s}_{j+1} \end{bmatrix}\right); \quad \begin{bmatrix} \hat{s}_{j+2N} \\ \hat{s}_{j+2N+1} \end{bmatrix} = \mathcal{Q}\left(\begin{bmatrix} \hat{s}_{j+2N} \\ \hat{s}_{j+2N+1} \end{bmatrix}\right). \quad (23)$$

5. Simulation Results

Simulation results are presented below to illustrate the performance of the proposed blind channel estimation and data recovery method. These results were obtained using Monte Carlo runs. In each run, we generate a stream composed of 1,000 blocks of $4 \times N$ BPSK information symbols, with $N = 8, 16, 32, 64, 128, 512$ and 1024 subcarriers, using a cyclic prefix of $CP = 25\% N$. This stream was transmitted through a $(2, 2)$ Rayleigh frequency selective fading MIMO channel, with two different sub-channels length values: $L = 3$ and $L = 4$.

A set of phases $\{\phi\} = \{0, \pi/3, \pi/2, 2\pi/3\}$ that verify Eq. (5) is chosen arbitrarily. The associate set of pairs of symbols is $\left\{ \begin{bmatrix} 1 \\ 1 \end{bmatrix}, \begin{bmatrix} -1 \\ 1 \end{bmatrix}, \begin{bmatrix} 1 \\ -1 \end{bmatrix}, \begin{bmatrix} -1 \\ -1 \end{bmatrix} \right\}$.

Figure 4 shows the accuracy of time channel variation tracking performed by SS-KCE, in the case of a channel speed variation factor of $f_D T = 0.001$, with $T = 4NT_s$, and different values of subcarriers N . It plots the normalized mean square error (NMSE) of the channel estimation, defined in Eq. (22), versus the SNR.

$$\text{NMSE} = \frac{1}{N_{MC}} \sum_{ir=1}^{N_{MC}} \frac{\|\hat{H}_{ir} - H\|_F^2}{\|H\|_F^2}, \quad (24)$$

where N_{MC} represents the number of Monte Carlo runs.

For $N = 16$, Fig. 4 shows that the blind channel estimation is better (small values of NMSE) compared to other values of $N > 16$, so raising the number of subcarriers N may increase the error channel estimation NMSE. This is because of N cumulative errors issued from N banks of the SS-KCE channel estimator. However, performance is not good when N is low, because in the OFDM technique, in order to convert a frequency selective channel to a set of parallel

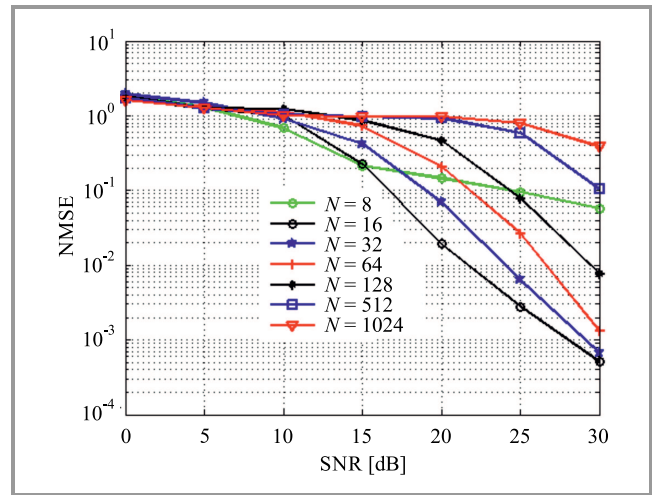


Fig. 4. Channel estimation NMSE of the proposed blind MIMO-OFDM algorithm with $(N_t, N_r) = (2, 2)$, BPSK, $L = 3$, $f_D T = 0.001$ and different numbers of N subcarriers.

flat fading sub-channels, a high value of subdivisions N is required that is not sufficient for $N = 8$. So, in this case, the channel is still not totally flat fading after the conversion and, as a consequence, the error channel estimation NMSE is raised.

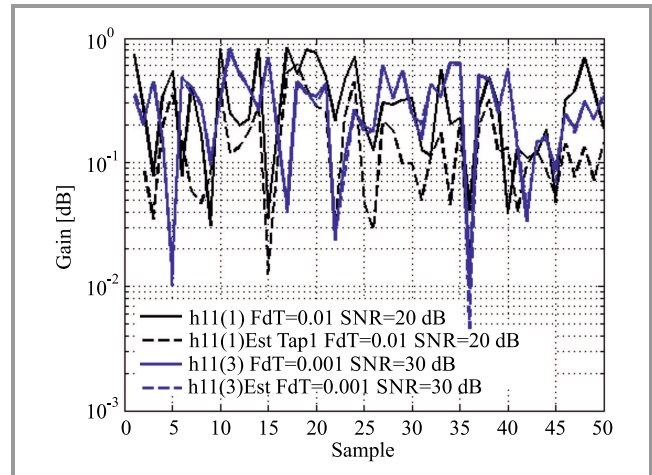


Fig. 5. SS-KCE tracking of channel coefficients $h_{11}(1)$ and $h_{11}(3)$ for $f_D T = 0.01$ at SNR = 20 dB and $f_D T = 0.001$ at SNR = 30 dB, respectively.

To illustrate the time tracking capability of the SS-KCE, coefficients $h_{11}(1)$ and $h_{11}(3)$ of the first MIMO selective sub-channel for $f_D T = 0.01$ at SNR = 20 dB, and $f_D T = 0.001$ at SNR = 30 dB, respectively, are compared with their estimates in Fig. 5. It is clear that the higher the SNR is and the smaller $f_D T$ is, the more accurate the channel estimation performed by the SS-KCE is.

In Fig. 6, the bit error rate (BER) of the proposed algorithm is plotted versus SNR for $f_D T = 0.001$ and different values of the number of subcarriers N . One may observe that when the number of subcarriers N is very high, the performance of the SS-KCE degrades. This is due to the estimation errors of the set of N sub-channel matrices, which increase

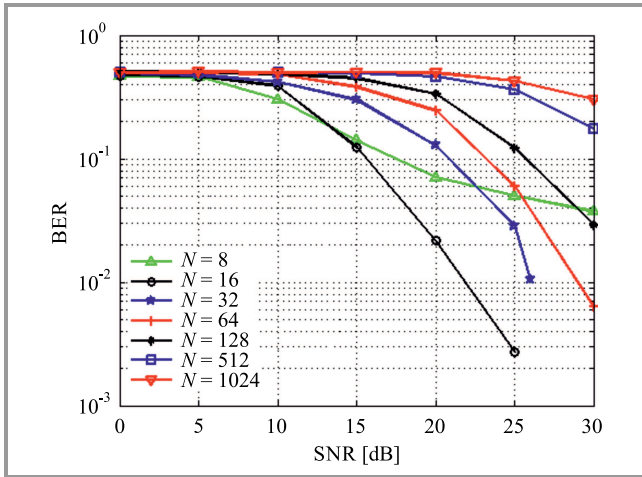


Fig. 6. BER of the proposed blind MIMO-OFDM algorithm with $(N_t, N_r) = (2, 2)$, BPSK, $L = 3$, $f_D T = 0.001$ and different values for the number of subcarriers N .

as N increases. These errors exert a negative impact on the performance of the detector, which in turn degrades the performance of the channels' estimation by SS-KCE. On the one hand, N should be large enough to guarantee that the MIMO selective channel is equivalent to a set of parallel flat fading channels. For the case of $N = 8$, BER is not good, because the detector is based directly on the output of the blind channel estimator, which is not very accurate (NMSE of $N = 8$ – Fig. 4). On the other hand, too large a value of N degrades the performance of the receiver. Therefore, the choice of N must result from a trade-off between these two contradictory requirements.

Finally, in Fig. 7 the performance of the proposed blind receiver with SS-KCE tracking of the channel variations is compared with the performances of two other blind MIMO-OFDM algorithms proposed in [30] and [31], respectively, for: $f_D T = 0$, $N = 64$, $L = 4$, $(N_t, N_r) = (2, 2)$, and BPSK. As one may observe, the proposed algorithm outperforms these two algorithms when SNR is higher than 20 dB.

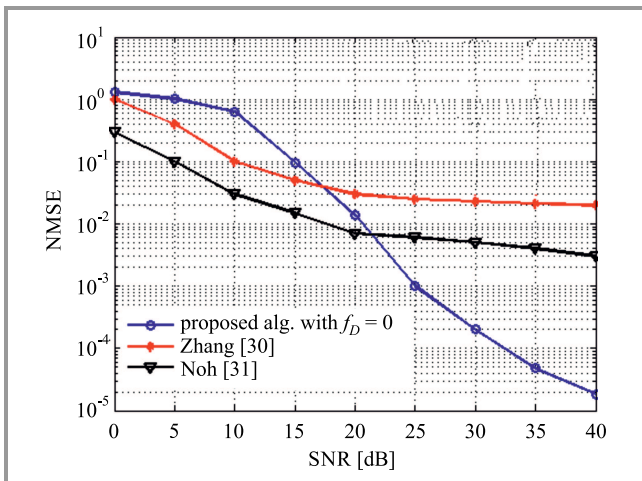


Fig. 7. Comparison between the proposed algorithm and two other blind MIMO-OFDM algorithms for $(N_t, N_r) = (2, 2)$, $N = 64$, BPSK, and $L = 4$.

6. Conclusion

In this paper, a blind MIMO system that operates over a frequency selective fading channel is proposed. It extends the already proposed OSTBC to be used with the well-known OFDM technique to solve the problem of channel frequency selectivity. This OSTBC offers a double benefit: on the one hand, it allows to alleviate the ambiguity that occurs in the estimation of the channel matrix with judicious choice of the number of subcarriers N , and on the other hand, it reduces complexity of the detector thanks to its orthogonal property. The performance of the proposed structure has been evaluated by means of Monte Carlo simulations and compared with two other similar systems proposed in literature. The results obtained show that the proposed system has the best performance for a reasonable range of SNR values.

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