Channel Estimation in Spatially Correlated High Mobility MIMO-OFDM Systems

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***Abstract*—MIMO-OFDM is a transmission method very sensitive to the channel time-variations arising in high mobility scenarios due to the appearance of inter-carrier interference (ICI). In this paper we propose an algorithm for channel estimation in MIMO-OFDM systems that uses a Basis Expansion Model that exploits temporal, frequency and spatial correlations within the wireless channel in high mobility scenarios. The advantage of the proposed approach is that it does not require a specific pilot subcarrier structure to estimate the ICI and is suitable for its implementation in currently standardized MIMO-OFDM radio interfaces.**

I. Introduction

MIMO-OFDM is one of the most widely adopted transmission methods in the last generations of broadband wireless communication systems. Although it has significant advantages such as high spectral efficiency and low-complexity equalization, it is very sensitive to inter-carrier interference (ICI) arising in high mobility environments. Under these conditions, each received subcarrier is distorted by the signal transmitted in adjacent subcarriers.

The estimation of doubly-selective fading channels has been previously addressed in the literature by approximating the channel impulse response with the use of a Basis Expansion Model (BEM) and it has also been studied for the particular case of OFDM. Different BEM strategies have been tested such as Complex-Exponentials (CE-BEM) [1], Generalized Complex-Exponentials (GCE-BEM) [2, 3], Polynomials (P-BEM) [4, 5], Discrete Prolate Spheroidal (DPS) sequences (DPS-BEM) [6, 7] and Karhunen-Loève (KL) coefficients (KL-BEM) [8]. The estimation of the channel impulse response from a block of OFDM symbols [9], as well as the effects of ICI on MIMO-OFDM signals and its equalization has also been studied [10]. These works assume in general a particular structure of clustered pilot subcarriers in the frequency domain to achieve optimal performance, but this assumption cannot be directly translated to standardized wireless communication systems, such as LTE, where pilots are typically placed in non-adjacent subcarriers.

Other works have tackled the problem by means of a two-steps estimation scheme, where in a first stage the individual frequency response estimations of a set of OFDM symbols are computed, and then these estimations are processed to extract the ICI component. In this line, Least Squares (LS) polynomial fitting [11] and low-pass interpolation [12] have been tried, and eventually generalized to a MIMO-OFDM channel estimator which allows to use arbitrary BEMs, and tested with the DPS-BEM [13]. In the context of this estimator, the spatial correlation of the MIMO channel is not considered [14], and even if a classical KL-BEM is used, its performance can be improved if information about the time-averaged channel is taken into consideration [15]. Moreover, the DPS-BEM has been shown to be optimum for time-varying channels with a flat Doppler spectrum, but this PSD only appears on channels with a 3D model of the arrival paths.

In this work we propose a BEM obtained from a low-rank approximation to the Linear Minimum Mean Squared Error (LMMSE) estimation of the ICI in a MIMO-OFDM system, which takes into account the freqency, temporal, and spatial correlations between the different TX-RX antenna pairs.

II. System model

Let us start considering the baseband equivalent impulse response of a Single Input Single Output (SISO) time-varying frequency-selective multipath channel given by

$$h(t,\tau) = \sum_{l=0}^{L-1} h_l(t)\delta(\tau-\tau_l), \qquad (1)$$

where $h_l(t)$ and τ_l are the complex gain and delay of the l-th channel multipath component, respectively.

Next, let us assume data symbols are transmitted over this channel using an Orthogonal Frequency Division Multiplexing (OFDM) modulation with N subcarriers cyclically extended with N_g samples. The total number of samples transmitted in an OFDM symbol is $N_t = N+N_g$. These samples are transmitted at a sample period T_s. Hence the total duration of an OFDM symbol in the time domain is $T = N_tT_s$ while the bandwidth is $F_s = 1/T_s$. Assuming $N_g > \tau_l \ \forall l$, the channel response during the transmission of the k-th OFDM symbol after removing the cyclic prefix can be expressed in form of a time-domain convolution matrix $\mathbf{H}_{k,i,j} \in \mathbf{C}^{N\times N}$ whose entries are $\mathbf{H}_{k,i,j}(r,s) = h_{i,j}(((k-1)N_t + N_g + r-1)T_s, \mathrm{mod}(r-s,N)T_s)$, with $h_{i,j}(t,\tau)$ the channel impulse response for the i-j antenna pair.

The previous channel model can be readily extended to consider a MIMO-OFDM system with M_T transmit and M_R receive antennas. In this case, the time-varying frequency-selective channel is represented by the follow-

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ing time-domain matrix with size $M_R N \times M_T N$

$$\tilde{\mathbf{H}}_k = \begin{pmatrix} \mathbf{H}_{k,1,1} & \mathbf{H}_{k,1,2} & \cdots & \mathbf{H}_{k,1,M_T} \\ \mathbf{H}_{k,2,1} & \mathbf{H}_{k,2,2} & \cdots & \mathbf{H}_{k,2,M_T} \\ \cdots & \cdots & \cdots & \cdots \\ \mathbf{H}_{k,M_R,1} & \mathbf{H}_{k,M_R,2} & \cdots & \mathbf{H}_{k,M_R,M_T} \end{pmatrix}, \quad (2)$$

where $\mathbf{H}_{k,i,j}$ represents the time-domain channel matrix between the j-th transmit antenna and the i-th receive antenna during the k-th OFDM symbol.

We assume an uncoded MIMO-OFDM system where the input modulated symbols are simply distributed across the N subcarriers and the M_T transmit antennas. The signal at the output of the MIMO-OFDM channel can be expressed as

$$\mathbf{y}_k = (\mathbf{I}_{M_R} \otimes \mathbf{F})\tilde{\mathbf{H}}_k(\mathbf{I}_{M_T} \otimes \mathbf{F}^H)\mathbf{x}_k + \mathbf{w} \quad (3)$$

where $\mathbf{x}_k$ is the vector of transmitted subcarriers in the k-th OFDM symbol with size $M_T N \times 1$, $\mathbf{F}$ is the $N \times N$ standard DFT matrix, $\mathbf{I}_M$ is the identity matrix with size $M \times M$, $\mathbf{w}$ is a $M_R N \times 1$ vector that represents the additive white Gaussian channel noise with variance σ_w^2, $\mathbf{y}_k$ is a $M_R N \times 1$ vector with the received subcarriers, and $\otimes$ denotes the Kronecker product. The signal model in Eq. (3) simplifies to

$$\mathbf{y}_k = \tilde{\mathbf{G}}_k \mathbf{x}_k + \mathbf{w}, \quad (4)$$

where $\tilde{\mathbf{G}}_k = (\mathbf{I}_{M_R} \otimes \mathbf{F})\tilde{\mathbf{H}}_k(\mathbf{I}_{M_T} \otimes \mathbf{F}^H)$ is the $M_R N \times M_T N$ frequency-domain MIMO-OFDM channel matrix during the transmission of the k-th OFDM symbol.

Matrix $\tilde{\mathbf{G}}_k$ is a block matrix built from the frequency response of the channels between each TX-RX antenna pair

$$\tilde{\mathbf{G}}_k = \begin{pmatrix} \mathbf{G}_{k,1,1} & \mathbf{G}_{k,1,2} & \cdots & \mathbf{G}_{k,1,M_T} \\ \mathbf{G}_{k,2,1} & \mathbf{G}_{k,2,2} & \cdots & \mathbf{G}_{k,2,M_T} \\ \cdots & \cdots & \cdots & \cdots \\ \mathbf{G}_{k,M_R,1} & \mathbf{G}_{k,M_R,2} & \cdots & \mathbf{G}_{k,M_R,M_T} \end{pmatrix}. \quad (5)$$

If the channels are static, the individual frequency-domain channels $\mathbf{G}_{k,i,j}$ will have non-zero entries only on their main diagonal. However, if the chanels are time-varying, non-zero off-diagonal entries will appear causing ICI between transmitted subcarriers, while the elements of the main diagonal will contain the average frequency response during the OFDM symbol

$$\mathbf{G}_{k,i,j} = \mathbf{F}(\bar{\mathbf{H}}_{k,i,j} + \Delta\mathbf{H}_{k,i,j})\mathbf{F}^H = \bar{\mathbf{G}}_{k,i,j} + \mathbf{G}_{\text{ICI}}, \quad (6)$$

where $\bar{\mathbf{H}}_{k,i,j}$ is the average impulse response of the channel during the OFDM block, and $\Delta\mathbf{H}_{k,i,j}$ contain the variations of the channel relative to the average of the coefficients during the block. Since $\bar{\mathbf{H}}_{k,i,j}$ is a circulant matrix, $\bar{\mathbf{G}}_{k,i,j}$ will be a diagonal matrix, and $\mathbf{G}_{\text{ICI}}$ will contain the ICI component of the frequency response. A receiver that ignores ICI will only estimate the elements contained in the main diagonal of $\bar{\mathbf{G}}_{k,i,j}$, treating the remaining components generated by $\mathbf{G}_{\text{ICI}}$ as white noise. It can also be observed that signals transmitted from different antennas interfere with each other, hence increasing the ICI with the number of transmit antennas.

It is important to note that if the whole $\tilde{\mathbf{G}}_k$ matrix had to be estimated, an underdetermined system would be obtained even in the case in which all transmitted symbols are known, since there are $M_T M_R N^2$ unknowns and $M_T N$ equations. In order to effectively estimate the MIMO-OFDM channel, the coefficients of $\tilde{\mathbf{G}}_k$ can be decomposed in the form of a BEM to reduce the number of parameters to be estimated. In practical MIMO-OFDM systems, channel estimation of individual channels can be done transmitting pilot subcarriers in disjoint frequency sets assigned to each antenna. Taking this into account, estimation of a MIMO OFDM channel can be considered equivalent to estimating $M_T M_R$ SISO channels between each transmit and receive antenna.

III. ICI ESTIMATION

Let us define $\mathbf{g}_n = [\mathbf{g}_{n,1,1}^T, \cdots, \mathbf{g}_{n,1,M_T}^T, \cdots, \mathbf{g}_{n,M_R,M_T}^T]^T$ as the vector which stacks the complex gains of the n-th subcarrier for all TX-RX antenna pairs $\mathbf{g}_{n,i,j} = [\mathbf{h}_{1,i,j}, \mathbf{h}_{2,i,j}, \cdots, \mathbf{h}_{L,i,j}]\mathbf{f}_n$, where the $\mathbf{h}_{l,i,j} = [h_l(N_g T_s), \cdots, h_l((N_t - 1)T_s), h_l((N_t + N_g)T_s), \cdots, h_l((KN_t - 1)T_s)]^T$ are $KN \times 1$ vectors which stack the channel gain for the l-th path of the channel. $\mathbf{f}_n$ is the $L \times 1$ vector with the first L rows of the n-th column of the matrix $\mathbf{F}$. The correlation matrix of $\mathbf{g}_{n,i,j}$ can be written as $\mathbf{C}_{gg} = \sum_{l=1}^{L} |f_n(l)|^2 \mathbf{C}_{h_l h_l}$ if the paths of the channel are uncorrelated, with $\mathbf{C}_{h_l h_l}$ denoting the time autocorrelation matrix of the l-th path of the channel. Denoting $\mathbf{R} = \mathbf{R}_r \otimes \mathbf{R}_t$, with $\mathbf{R}_t$ and $\mathbf{R}_r$ the $M_T \times M_T$ transmit and $M_R \times M_R$ receive correlation matrices, respectively, the correlation matrix $\text{E}(\mathbf{g}_n \mathbf{g}_n^H) = \mathbf{R} \otimes \mathbf{C}_{gg}$.

Using a Q-dimensional BEM, we can express $\mathbf{g}_n$ as follows [16]

$$\mathbf{g}_n = \mathbf{B}\mathbf{c}_n, \quad (7)$$

where $\mathbf{B} = \left[\mathbf{B}_{1,1}^T, \cdots, \mathbf{B}_{1,M_T}^T, \cdots, \mathbf{B}_{M_R,M_T}^T\right]^T$ is a $KNM_T M_R \times (Q+1)$ matrix with the basis vectors that define the BEM, where $\mathbf{B}_{i,j}$ are $KN \times (Q+1)$ matrices with the BEM for each TX-RX antenna pair, and $\mathbf{c}_n$ a $(Q+1) \times 1$ vector of BEM coefficients for the n-th subcarrier. Furthermore, each $\mathbf{B}_{i,j}$ matrix can be split into K different matrices $\mathbf{B}_{k,i,j}$ with dimensions $N \times (Q+1)$, defining a BEM matrix for each symbol inside the group of K symbols, and for each TX-RX antenna pair, as $\mathbf{B}_{k,i,j} = [\mathbf{b}_{(k-1)N+1}, \mathbf{b}_{(k-1)N+2}, \cdots, \mathbf{b}_{kN}]^T$, where the $\mathbf{b}_r$ vectors represent the r-th column of $\mathbf{B}_{i,j}^T$.

The matrix $\mathbf{G}_{k,i,j}$ can be expressed in terms of a BEM [13, 16] as

$$\mathbf{G}_{k,i,j} = \sum_{q=1}^{Q+1} \mathbf{D}_{k,q,i,j} \text{diag}\left\{\mathbf{d}_q\right\}, \quad (8)$$

where

$$\mathbf{D}_{k,q,i,j} = \mathbf{F}\text{diag}\left\{\mathbf{b}_{k,q}\right\}\mathbf{F}^H, \quad (9)$$

and $\mathbf{b}_{k,q}$ represents the q-th column of the matrix $\mathbf{B}_{k,i,j}$, and with $\mathbf{d}_q = [\mathbf{c}_1(q), \mathbf{c}_2(q), ..., \mathbf{c}_N(q)]$.

In the ensuing subsection, a low-rank approximation of the LMMSE estimation of $\mathbf{g}_n$ will be used to obtain the BEM coefficents $\mathbf{c}_n$.

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A. Frequency response estimation

Let us start by applying a conventional frequency response estimation algorithm to each symbol inside the group of K OFDM symbols. The LMMSE estimator, when a set of pilot subcarriers are allocated, can be expressed as [17]

$$\hat{\mathbf{g}}_{\mathrm{LS},k}^{(p)} = \mathbf{P}^{-1} y_k^{(p)} \quad (10)$$

$$\hat{\mathbf{g}}_{\mathrm{LMMSE},k} = \mathbf{R}_{gg_p}(\mathbf{R}_{g_pg_p} + (\sigma_w/E_p)^2)^{-1}\mathbf{g}_{\mathrm{LS},k}^{(p)}, \quad (11)$$

where $y_k^{(p)}$ represents the $P \times 1$ vector of received pilot symbols in the k-th OFDM symbol, $\mathbf{P}$ is a $P \times P$ matrix with the transmitted pilot symbols on its main diagonal, $\mathbf{R}_{gg_p}$ is the $N \times P$ cross-correlation matrix of the channel coefficients with the pilot positions, $\mathbf{R}_{g_pg_p}$ is the $P \times P$ autocorrelation matrix of the frequency response in the pilot positions, E_p is the energy of the pilot subcarriers, and σ_w^2 is the variance of the noise. In a time-selective fading channel, the noise caused by ICI can be considered as part of the channel noise variance σ_w^2. The coefficients of $\hat{\mathbf{g}}_{\mathrm{LMMSE},k}$ will correspond to the elements of the main diagonal of $\mathbf{G}_{k,i,j}$, so an estimation of the time-averaged frequency response of the channel during the symbol for each subcarrier will be obtained.

The MSE of this estimator is

$$\mathrm{MSE}_k = \frac{1}{N}\mathrm{tr}(\mathbf{R}_{gg} - \mathbf{R}_{gg_p}(\mathbf{R}_{g_pg_p} + (\sigma_w/E_p)^2)^{-1}\mathbf{R}_{g_pg}). \quad (12)$$

B. BEM parameter estimation

Once an estimation of the frequency response of the channel for a group of symbols is available, the correlation properties in the time domain can be exploited to reconstruct the channel gain at each frequency. Let us define an estimator $\mathbf{W}$ for the complex gain for the n-th subcarrier during K OFDM symbols, from the average coefficients obtained from a channel estimation step, as defined in the previous section

$$\hat{\mathbf{g}}_{\mathrm{LMMSE},n} = \mathbf{W}\mathbf{g}_n = \mathbf{W}(\bar{\mathbf{g}}_n + \mathbf{e}), \quad (13)$$

where $\mathbf{g}_n = \left[\mathbf{g}_{n,1,1}^T, \cdots, \mathbf{g}_{n,1,M_T}^T, \cdots, \mathbf{g}_{n,M_R,M_T}^T\right]^T$ is a $KM_TM_R \times 1$ vector with the complex gain for the n-th subcarrier during K OFDM symbols for all paths, with $\mathbf{g}_{n,i,j} = [\bar{\mathbf{G}}_{1,i,j}(n,n), \bar{\mathbf{G}}_{2,i,j}(n,n), \cdots, \bar{\mathbf{G}}_{K,i,j}(n,n)]$, $\mathbf{W}$ is a $KNM_TM_R \times KM_TM_R$ matrix, and $\mathbf{e}$ is a $KM_TM_R \times 1$ vector of white Gaussian noise variables, with variance σ_{MSE}^2 equal to the MSE of the frequency response estimation method used in the previous step. If $\mathbf{W}$ corresponds to an LMMSE estimator, this matrix takes the form

$$\mathbf{W} = \mathbf{C}_{g\bar{g}}(\mathbf{C}_{\bar{g}\bar{g}} + \sigma_{\mathrm{MSE}}^2\mathbf{I}_{KM_TM_R})^{-1}, \quad (14)$$

where $\mathbf{C}_{g\bar{g}}$ is the $KNM_TM_R \times KM_TM_R$ cross-correlation matrix of the channel coefficients with the average of the channel coefficients during an OFDM symbol, and $\mathbf{C}_{\bar{g}\bar{g}}$ is the $KM_TM_R \times KM_TM_R$ autocorrelation matrix of the average of the channel coefficients. These matrices can be readily computed from the elements of $\mathbf{C}_{gg}$ as

$$\mathbf{C}_{g\bar{g}}(n,m) = \mathbf{R} \otimes \left(\frac{1}{N}\sum_{i=1}^{N}\mathbf{C}_{gg}(n,(m-1)N_t + i)\right), \quad (15)$$

and

$$\mathbf{C}_{\bar{g}\bar{g}}(n,m) = \mathbf{R} \otimes \left(\frac{1}{N^2}\sum_{i=1}^{N}\sum_{j=1}^{N}\mathbf{C}_{gg}((n-1)N+i,(m-1)N+j)\right). \quad (16)$$

Next, let us apply an SVD decomposition and write the correlation matrices in the form

$$\mathbf{C}_{g\bar{g}} = \mathbf{U}_{g\bar{g}}\mathbf{S}_{g\bar{g}}\mathbf{V}_{g\bar{g}}^H,$$
$$\mathbf{C}_{\bar{g}\bar{g}} = \mathbf{U}_{\bar{g}\bar{g}}\mathbf{S}_{\bar{g}\bar{g}}\mathbf{V}_{\bar{g}\bar{g}}^H. \quad (17)$$

Alternatively, this decomposition can be written as

$$\mathbf{C}_{g\bar{g}} = (\mathbf{U}_s \otimes \mathbf{U}_1)(\mathbf{S}_s \otimes \mathbf{S}_1)(\mathbf{V}_s^H \otimes \mathbf{V}_1^H), \quad (18)$$
$$\mathbf{C}_{\bar{g}\bar{g}} = (\mathbf{U}_s \otimes \mathbf{U}_2)(\mathbf{S}_s \otimes \mathbf{S}_2)(\mathbf{V}_s^H \otimes \mathbf{V}_2^H), \quad (19)$$

where $\mathbf{U}_1\mathbf{S}_1\mathbf{V}_1^H$ and $\mathbf{U}_2\mathbf{S}_2\mathbf{V}_2^H$ are the SVD decompositions of the expression in parentheses in Eq. (15) and Eq. (16), respectively; $\mathbf{U}_1$, $\mathbf{V}_1$, $\mathbf{U}_2$ and $\mathbf{V}_2$ are unitary matrices; and $\mathbf{S}_1$ and $\mathbf{S}_2$ are matrices with their singular values arranged in decreasing order on the main diagonal. If the singular values of $(\mathbf{S}_s \otimes \mathbf{S}_1)$ and $(\mathbf{S}_s \otimes \mathbf{S}_2)$ are reordered on the main diagonals, the columns and rows of the matrices containing the singular vectors will have to be reordered accordingly.

Replacing these expressions in Eq. (14), we obtain

$$\hat{\mathbf{g}}_{\mathrm{LMMSE},n} = \mathbf{U}_{g\bar{g}}\mathbf{S}_{g\bar{g}}\mathbf{V}_{g\bar{g}}^H\mathbf{U}_{\bar{g}\bar{g}}(\mathbf{S}_{\bar{g}\bar{g}} + \sigma_{\mathrm{MSE}}^2\mathbf{I}_{KM_TM_R})^{-1}\mathbf{V}_{\bar{g}\bar{g}}^H\mathbf{g}_n. \quad (20)$$

A rank-reduced version of this estimator can be obtained by considering only the $(Q+1)$ higher elements on the main diagonal of $\mathbf{S}_{g\bar{g}}$, and setting to zero de remaining values. This expression can be interpreted as a BEM (see Eq. (7)) where $\mathbf{B} = \mathbf{U}_{g\bar{g}}$ and $\mathbf{c}_n = \mathbf{S}_{g\bar{g}}\mathbf{V}_{g\bar{g}}^H\mathbf{U}_{\bar{g}\bar{g}}(\mathbf{S}_{\bar{g}\bar{g}} + \sigma_{\mathrm{MSE}}^2\mathbf{I}_{KM_TM_R})^{-1}\mathbf{V}_{\bar{g}\bar{g}}^H\mathbf{g}_n$.

It is important to note that in this estimation method it is not necessary to make an assumption on the samples chosen from the original matrix $\mathbf{B}$ to make the projection, hence avoiding the definition of an arbitrary time index [11, 13]. Also, the information about the average of the coefficients is taken into account when the BEM is defined which reflects more accurately the behaviour of the channel. The order of this BEM is related with the joint spatial and temporal correlations, so the values of Q will be larger compared with systems were only the temporal correlation is considered.

C. Computational complexity

The computation of the $\mathbf{c}_n$ vectors, assuming that the estimation matrix is known, has computational complexity $\mathcal{O}(NKM_TM_R(Q+1))$. Computing the $\tilde{\mathbf{G}}_k$ matrix has a computational complexity of $\mathcal{O}(N^2M_TM_R(Q+1))$. This complexity can be lowered if only some off-diagonal elements are assumed to be different from zero. If the number of considered diagonals different from the main diagonal is denoted as N_{off}, the complexity has the form $\mathcal{O}(NN_{\mathrm{off}}M_TM_R(Q+1))$. Furthermore, if we examine the structure of the matrix $\mathbf{U}_{g\bar{g}}$ as a result of a Kronecker product (see Eq. (18)), it can be observed that some of the coefficients of $\mathbf{c}_n$ could operate with scaled versions

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of the same column of $\mathbf{U}_1$ depending on the antenna pair considered, turning out to be possible to further reduce the computation of $\tilde{\mathbf{G}}_k$.

It must be highlighted that the estimation of the correlation matrices, and the power of noise and ICI is not considered in this analysis, as well as the SVD decomposition to build the BEM. These steps would increase the overall computational complexity of the estimator.

IV. RESULTS

Computer simulations were carried out to illustrate the performance of the time-varying MIMO-OFDM channel estimation and equalization described in the previous sections. We considered a MIMO-OFDM system with N = 128 subcarriers and baseband bandwidth F_s = 1 MHz. The cyclic prefix length was set to 1/8 of the number of subcarriers, i.e $N_g = 16$. The pilot structure is specified by the WiMAX standard for MIMO 4×4 transmissions in the open-loop mode [18]. The carrier frequency is $f_c = 5.2$ GHz and the maximum Doppler spread is $f_d = v f_c / c$, being v the relative velocity between transmitter and receiver.

Regarding Doppler spectrum, the Jakes model assumes that the angles of arrival of paths are uniformly distributed in the interval $[-\pi, \pi]$ in the azimuth plane of the receiving antennas. The autocorrelation function of this model is given by

$$r_j(\tau) = J_0(2\pi f_d \tau), \tag{21}$$

where $J_0(\cdot)$ is the zeroth-order Bessel function. When the angles of arrival are constrained to arrive from a narrower interval, an asymmetrical Jakes spectrum model arises, and its autocorrelation function is

$$r_{aj}(\tau) = \int_{\sin^{-1}(f_{\min})}^{\sin^{-1}(f_{\max})} \cos(2\pi f_d \tau \sin\phi) d\phi \tag{22}$$

$$-j \int_{\cos^{-1}(f_{\min})}^{\cos^{-1}(f_{\max})} \sin(2\pi f_d \tau \cos\phi) d\phi, \tag{23}$$

where $f_{\min} = f_{d,\min}/f_d$ and $f_{\max} = f_{d,\max}/f_d$ are the normalized minimum and maximum Doppler frequencies, respectively. In our simulations an asymmetrical Jakes spectrum model is considered.

These autocorrelation functions depend on the normalized Doppler spread $f_d T$, which is related to the relative speed between the transmitter and the receiver. In our simulations, the results obtained with the proposed estimation method are compared to those obtained with DPS-BEM [6, 7] using the method based on the LS estimation [13] with order $Q = 15$ and $Q = 31$. Since Q is related with the joint spatial and temporal representation of the channel, in the case of the estimators which ignore spatial correlation, it is equivalent to use a temporal BEM with order $Q_t = 0$ and $Q_t = 1$, respectively, for each TX-RX antenna pair in this 4×4 setup. The ICI estimation algorithm is applied to groups of $K = 6$ MIMO-OFDM symbols, starting from the MMSE estimation of the frequency response of the channel for each symbol. In this initial estimate, the power of noise and the power of ICI are assumed to be known, computing the latter from the time correlation function of the channel [19].

TABLE I. CHANNEL MODEL

Tap	Power (dB)	Delay (ns)
1	0	0
2	-1	100
3	-9	300
4	-10	500
5	-15	800
6	-20	1200

Table I shows the Power Delay Profile (PDP) of the channel impulse response considered in the simulations. The same PDP was used for the $M_T M_R$ channels between each transmit-receive antenna pair. The spatial correlation matrix considered was

$$\mathbf{R}_t = \mathbf{R}_r = \begin{pmatrix} 1 & 0.9 & 0.7 & 0.5 \\ 0.9 & 1 & 0.9 & 0.7 \\ 0.7 & 0.9 & 1 & 0.9 \\ 0.5 & 0.7 & 0.9 & 1 \end{pmatrix}. \tag{24}$$

The Doppler spectrum follows an asymmetrical Jakes model with $f_{\min} = 0.4$ and $f_{\max} = 1.0$. It is also interesting to note how this estimation algorithm is affected by the number of transmit antennas, regardless of the considered BEM. Since this algorithm takes as input the channel frequency response of individual OFDM symbols, and we have considered a pilot-aided MMSE estimation, these starting estimations will be affected by ICI generated by the unknown data subcarriers from all the transmit antennas.

Figure 1 shows the channel estimation MSE over the SNR in a 4×4 setup, for $f_d T = 0.18$. It can be seen that the KL-BEM defined in this paper outperforms the DPS-BEM in every SNR level. On the other hand, only a small gain is observed by increasing the order when spatial correlation is considered, while the other methods achieve significant gains.

Figure 2 plots the resulting channel estimaton MSE over the Doppler spread with a fixed SNR = 30 dB. Again, the proposed estimator outperforms the LS estimator with the DPS-BEM obtaining a constant gain for every considered Doppler spread.

Figure 3 shows the channel estimation MSE over the symbol index. In this case the results are again better when the proposed KL-BEM is used, but the error in the central symbols of the block is lower than in the edges. This would allow to define a windowed estimation algorithm, where only those symbols are considered to make estimations in order to lower the estimation errors.

These results shows that with a low order BEM in this highly spatially correlated environment, the same performance can be achieved as with higher order BEMs which ignore this correlation.

V. CONCLUSIONS

In this work we have shown the results obtained with an LMMSE estimator of the ICI which takes as input the channel frequency response of a block of OFDM symbols considering the temporal, frequency and spatial correlation of a MIMO channel. This estimator shows better performance than previous LS estimation methods defined in literature and significant gains are obtained when the channel exhibits a high spatial correlation.

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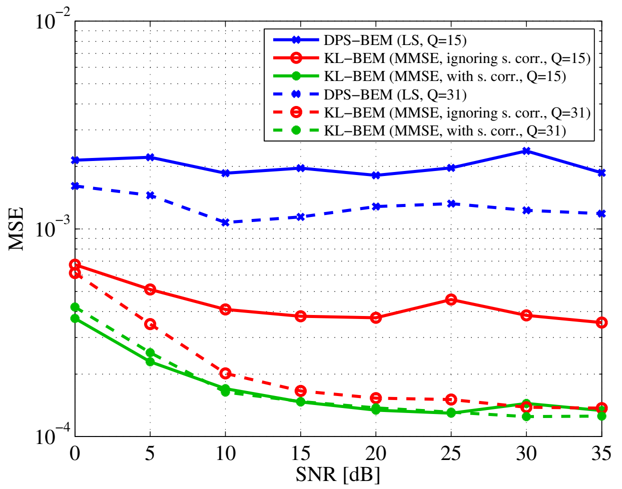

Fig. 1. MSE over SNR of the channel matrix estimation at $f_d T = 0.18$.

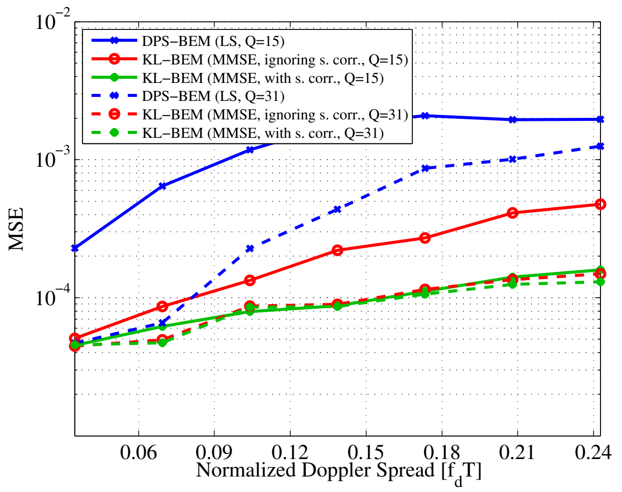

Fig. 2. MSE over Doppler spread of the channel matrix estimation at SNR = 30 dB.

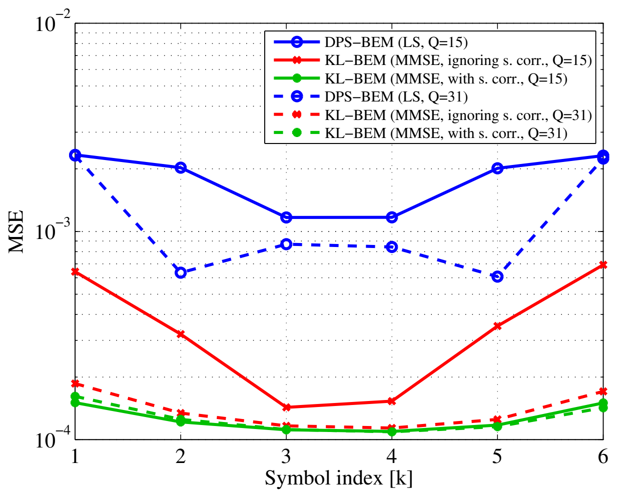

Fig. 3. MSE over symbol index of the channel matrix estimation at SNR = 30 dB and $f_d T = 0.18$.

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