KLT-based Estimation of Rapidly Time-Varying Channels in MIMO-OFDM Systems

Pedro Suárez-Casal, José A. García-Naya, Luis Castedo
Dept. of Electronics and Systems
University of A Coruña, SPAIN
{pedro.scasal, jagarcia, luis}@udc.es

Markus Rupp
Institute of Telecommunications,
Vienna University of Technology, Austria
mrupp@nt.tuwien.ac.at

***Abstract*—MIMO-OFDM is a transmission method that is very sensitive to the channel time-variations arising in high mobility scenarios due to the emergence of Inter-Carrier Interference (ICI). In this paper we propose a channel estimation method using a basis expansion model that incorporates the knowledge of the channel Doppler spectrum, taking into account the statistics of the averaged coefficients of the channel. The advantage of the proposed approach is that it does not require a specific pilot subcarrier structure to estimate the ICI, and outperforms existing methods in the literature.**

***Keywords*—OFDM, ICI, Time-varying channels, MIMO, Basis Expansion Model.**

I. Introduction

MIMO-OFDM is the transmission method adopted by the last generations of broadband wireless communication systems. MIMO-OFDM combines Multiple Input Multiple Output (MIMO) transmission over multiple transmit and receive antennas together with Orthogonal Frequency Division Multiplexing (OFDM) modulation. The primary advantage of MIMO-OFDM is its ability to achieve large spectral efficiencies while equalizing frequency-selective channels with low-complexity receivers. Nevertheless, the performance of MIMO-OFDM degrades severely when transmitting over rapidly time-varying channels such as those encountered in high mobility applications. In MIMO-OFDM, fast channel time variations cause symbols received in an OFDM subcarrier to be affected by symbols in adjacent subcarriers. This harmful effect is known as Inter-Carrier Interference (ICI) and has to be estimated and compensated for MIMO-OFDM to work adequately over rapidly time-varying wireless channels.

In the literature, several works address ICI estimation in OFDM systems. Most of the proposed solutions make use of a Basis Expansion Model (BEM) in which the channel impulse response is decomposed as a linear combination of basis vectors in order to effectively reduce the number of coefficients to be estimated. Different BEMs have been proposed: Complex-Exponentials (CE-BEM) [1], Generalized Complex-Exponentials (GCE-BEM) [2], Polynomials (P-BEM) [3], Discrete Prolate Spheroidal (DPS) sequences (DPS-BEM) [4] and Karhunen-Loève (KL) coefficients (KL-BEM) [5].

ICI estimaton using BEMs has been addressed in Single Input Single Output (SISO)-OFDM for single and multiple OFDM blocks [6], [7], as well as MIMO-OFDM systems [8]. These works, however, assume a particular OFDM symbol structure in which clusters of pilot subcarriers are available in the frequency domain. This is an unrealistic assumption that is not accomplished in most currently standardized wireless communication systems, such as LTE, where pilots are typically placed in non-adjacent subcarriers.

Other works in the literature have specifically addressed ICI estimation in OFDM systems with arbitrary pilot structures using a two-step approach. In the first step a time-averaged channel impulse response estimation is computed for a group of OFDM symbols, and then these individual estimations are used in a second step to estimate the ICI. Least Squares (LS) polynomial fitting [9] and low-pass interpolation [10] have been proposed. Other approximations have considered the use of a LS interpolation based on a smoothness constraint [11]. This scheme has been generalized to consider ICI estimation in MIMO-OFDM systems using a DPS-BEM [12].

The DPS-BEM has been shown to be optimum for time-varying channels with a flat Doppler spectrum. Nevertheless, only channels with a 3D model of the arrival paths exhibit a flat Doppler spectrum. As an example, channel models for urban environments tipically follow a symmetrical or asymmetrical Jakes' spectrum depending on the angle of arrival of the multipath components [13]. Non-flat Doppler spectrum shapes also appear in aeronautical [14] and satellite [15] channel models. Another example is the COST-207 channel model for GSM communications in mobile environments that utilizes a bi-Gaussian Doppler spectrum [16].

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. The proposed approach utilizes the individual frequency response estimations obtained from a frame of several consecutive MIMO-OFDM symbols without requiring any specific pilot subcarrier structure in the frequency domain. Information about the channel Doppler spectrum is incorporated into the LMMSE channel estimation while the Karhunen-Loève Transform (KLT) is used to obtain the low-rank approximation. The results of computer simulations show that the proposed BEM exhibits a superior performance when estimating the ICI originated by time-varying MIMO-OFDM channels with non-flat Doppler spectrum.

II. Signal Model

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

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

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 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 $T = N_t T_s$ 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 between the j-th transmit antenna and the i-th receive antenna can be expressed in the 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 following time-domain matrix with size $M_R N \times M_T N$ [8]

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

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

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

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 $M \times M$, $\mathbf{w}_k$ is a $M_T N \times 1$ 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 containing the received subcarriers, and $\otimes$ denotes the Kronecker product. The signal model in Eq. (3) simplifies to

$$\mathbf{y}_k = \tilde{\mathbf{G}}_k \mathbf{x}_k + \mathbf{w}_k, \tag{4}$$

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

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

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 channels are time-varying, non-zero off-diagonal entries will appear causing ICI between transmitted subcarriers. If we denote the time-averaged impulse response of the channel during the k-th OFDM block as

$$\bar{h}(\tau) = \sum_{l=1}^{L} \bar{h}_l \delta(\tau - \tau_l), \tag{6}$$

with $\bar{h}_l = \frac{1}{N}\sum_{t=(k-1)N_t+N_g}^{kN_t-1} h_l(tTs)$, the channel frequency response can take the form

$$\mathbf{G}_{k,i,j} = \mathbf{F}(\bar{\mathbf{H}}_{k,i,j} + \Delta\mathbf{H}_{k,i,j})\mathbf{F}^H = \bar{\mathbf{G}}_{k,i,j} + \mathbf{G}_{\mathrm{ICI}}, \tag{7}$$

where $\bar{\mathbf{H}}_{k,i,j}(r,s) = \bar{h}_{i,j}(\mathrm{mod}(r-s,N)T_s)$, $\bar{h}_{i,j}(\tau)$ is the time-averaged impulse response between antennas i and j, 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, $\bar{\mathbf{G}}_{k,i,j}$ is a diagonal matrix with the time-averaged frequency response, and $\mathbf{G}_{\mathrm{ICI}}$ contains 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 $\mathbf{G}_{\mathrm{ICI}}$ as noise.

III. ICI ESTIMATION

Let us define $\mathbf{g}_n$ as the vector containing the complex gain of the n-th subcarrier during K consecutive OFDM symbols with N samples after removing the cyclic prefix in a transmission between antennas i and j. We will drop these indices in this section to simplify the notation:

$$\mathbf{g}_n = [\mathbf{h}_1, \mathbf{h}_2, \cdots, \mathbf{h}_L]\mathbf{f}_n, \tag{8}$$

where the $\mathbf{h}_l = [h_l(N_g T_s), \cdots, h_l((N_t-1)T_s), h_l((N_t+N_g)T_s), \cdots, h_l((KN_t-1)T_s)]^T$ 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 $\mathbf{g}_n$ 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.

Using a Q-dimensional BEM, we can approximate $\mathbf{g}_n$ as follows

$$\mathbf{g}_n = \mathbf{B}\mathbf{c}_n, \tag{9}$$

where $\mathbf{B} = [\mathbf{b}_0^T, \mathbf{b}_1^T, ..., \mathbf{b}_Q^T]$ is an $KN \times (Q+1)$ matrix with the basis vectors that define the BEM, and $\mathbf{c}_n$ a $(Q+1) \times 1$ vector of BEM coefficients for the n-th subcarrier. To obtain the channel evolution during the k-th OFDM symbol, matrix $\mathbf{B}$ can be split in K different matrices $\mathbf{B}_k$ with dimensions $N \times (Q+1)$, hence defining a BEM matrix for each symbol inside the group of K symbols as $\mathbf{B}_k = [\mathbf{b}_{(k-1)N+1}, \mathbf{b}_{(k-1)N+2}, \cdots, \mathbf{b}_{kN}]^T$, where the $\mathbf{b}_i$ vectors represent the i-th column of $\mathbf{B}^T$.

In a SISO-OFDM system, the ICI can be estimated from the channel estimates obtained in the frequency domain using the following expression of the channel matrix $\mathbf{G}_k$ in terms of a BEM for the k-th OFDM symbol [12]

$$\mathbf{G}_k = \sum_{q=1}^{Q+1} \mathbf{D}_{k,q} \mathrm{diag}\{\mathbf{d}_q\}, \tag{10}$$

where

$$\mathbf{D}_{k,q} = \mathbf{F}\mathrm{diag}\{\mathbf{b}_{k,q}\}\mathbf{F}^H, \tag{11}$$

$\mathbf{b}_{k,q}$ represents the q-th column of the matrix $\mathbf{B}_k$, and $\mathbf{d}_q = [c_1(q), c_2(q), ..., c_N(q)]$.

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

A. BEM Coefficients Estimation

Let us consider the estimated channel coefficients corresponding to the n-th subcarrier during a block of K OFDM symbols, obtained with a frequency response estimation algorithm. Any frequency response estimation algorithm can be considered in this step, as long as the estimation error of the method is known:

$$\hat{\mathbf{g}}_n = \bar{\mathbf{g}}_n + \mathbf{e}_n, \tag{12}$$

where $\bar{\mathbf{g}}_n = [\bar{\mathbf{G}}_1(n,n), \bar{\mathbf{G}}_2(n,n), \cdots, \bar{\mathbf{G}}_K(n,n)]^T$ is a $K \times 1$ vector with the time-averaged channel coefficients in the n-th subcarrier during K OFDM symbols, $\hat{\mathbf{g}}_n$ is a $K \times 1$ vector which contains the estimated channel coefficients in the n-th subcarrier during K OFDM symbols, and $\mathbf{e}_n$ is a $K \times 1$ vector of error components from the estimation method used in the frequency response estimation step, with a variance σ_e^2. It will be assumed that the estimation error and the channel coefficients are uncorrelated, and also the estimation error between different symbols for a given subcarrier.

Let us also define a $KN \times K$ matrix $\mathbf{W}$ to obtain the LMMSE estimation of $\mathbf{g}_n$ from the estimated channel frequency response estimations with the form

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

Matrix $\mathbf{W}$ will be defined as follows

$$\mathbf{W} = \mathbf{C}_{g\bar{g}}(\mathbf{C}_{\bar{g}\bar{g}} + \sigma_e^2\mathbf{I})^{-1} \tag{14}$$

where $\mathbf{C}_{g\bar{g}}$ is the $KN \times K$ 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 $K \times K$ autocorrelation matrix of the average of the channel coefficients. These matrices can be readily computed from the elements of $\mathbf{C}_{gg}$ as

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

and

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

The following SVD decompositions can be applied

$$\mathbf{C}_{g\bar{g}} = \mathbf{U}_1\mathbf{\Sigma}_1\mathbf{V}_1^H \tag{17}$$

$$\mathbf{C}_{\bar{g}\bar{g}} = \mathbf{U}_2\mathbf{\Sigma}_2\mathbf{V}_2^H \tag{18}$$

where $\mathbf{U}_1$, $\mathbf{V}_1$, $\mathbf{U}_2$ and $\mathbf{V}_2$ are unitary matrices, and $\mathbf{\Sigma}_1$ and $\mathbf{\Sigma}_2$ are matrices with the singular values arranged in decreasing order on the main diagonal. Replacing in Eq. (13) we obtain

$$\hat{\mathbf{g}}_{\text{LMMSE},n} = \mathbf{U}_1\mathbf{\Sigma}_1\mathbf{V}_1^H\mathbf{V}_2\mathbf{\Lambda}\mathbf{U}_2^H\hat{\mathbf{g}}_n = \mathbf{Z}\hat{\mathbf{g}}_n, \tag{19}$$

where $\mathbf{\Lambda} = \text{diag}\{[\sigma_1 + \sigma_e^2, \sigma_2 + \sigma_e^2, \cdots, \sigma_K + \sigma_e^2,]\}^{-1}$ and the σ_k elements are the singular values of $\mathbf{C}_{\bar{g}\bar{g}}$. A rank-reduced version of this estimator can be obtained by considering only the first $Q+1$ coefficients in the main diagonal of $\mathbf{\Sigma}_1$, and assigning to zero the remaining elements.

This expression can be interpreted as a BEM (see Eq. (9)) with $\mathbf{B} = \mathbf{U}_1$ and $\mathbf{c}_n = \mathbf{\Sigma}_1\mathbf{V}_1^H\mathbf{V}_2\mathbf{\Lambda}\mathbf{U}_2^H\hat{\mathbf{g}}_n$.

The MSE of this estimator for the n-th subcarrier is

$$\begin{aligned}\text{MSE}_n &= \frac{1}{KN}\text{E}\{(\mathbf{g}_n - \hat{\mathbf{g}}_{\text{LMMSE},n})(\mathbf{g}_n - \hat{\mathbf{g}}_{\text{LMMSE},n})^H\} \\ &= \frac{1}{KN}\text{tr}(\mathbf{C}_{gg} - 2\Re(\mathbf{Z}\mathbf{C}_{\bar{g}g}) + \mathbf{Z}\mathbf{C}_{\bar{g}\bar{g}}\mathbf{Z}^H). \end{aligned}\tag{20}$$

B. ICI Estimation

Let us assume we have estimations of the main diagonal of $\mathbf{G}_k$ for a set of K OFDM symbols and define the $N \times 1$ column vectors $\hat{\mathbf{g}}_k = \text{diag}\{\mathbf{G}_k\}^T$, which can be computed by any method. In this sense, if an LS or LMMSE estimator is used, ICI corresponding to pilot and data subcarriers will interfere in the estimation.

The LS estimator of the coefficients can be written as [12]:

$$[\hat{\mathbf{c}}_1, \hat{\mathbf{c}}_2, ..., \hat{\mathbf{c}}_N] = (\mathbf{M}^H\mathbf{M})^{-1}\mathbf{M}^H[\hat{\mathbf{g}}_1, \hat{\mathbf{g}}_2, ..., \hat{\mathbf{g}}_K]^T, \tag{21}$$

where $\mathbf{M} = [\mathbf{m}_0^T, \mathbf{m}_1^T, ..., \mathbf{m}_Q^T]$ is a matrix composed by the basis vectors sampled at the OFDM symbol period, with $\mathbf{m}_q = [b_q(N_g + N/2), b_q(N/2 + N_t), ..., b_q(N/2 + (K-1)N_t)]$. This method is equivalent to a LS fitting of the coefficients according to the vectors of the BEM.

With the result obtained in the previous section, and under the assumption that the coefficients in all subcarriers have the same Doppler Power Spectral Density (PSD), we can define a low rank approximation of the LMMSE estimator for the coefficients and the coefficients to compute the ICI component are obtained as follows

$$[\hat{\mathbf{c}}_1, \hat{\mathbf{c}}_2, ..., \hat{\mathbf{c}}_N] = \mathbf{\Sigma}_1\mathbf{V}_1^H\mathbf{V}_2\mathbf{\Lambda}\mathbf{U}_2^H[\hat{\mathbf{g}}_1, \hat{\mathbf{g}}_2, ..., \hat{\mathbf{g}}_K]^T. \tag{22}$$

Finally, once the estimates of $\mathbf{c}_q$ are calculated, they can be replaced in Eq. (10) and $\mathbf{G}_k$ can be computed. To build the original $\tilde{\mathbf{G}}_k$ matrix, the algorithm described in this section is repeated for each pair of transmit and receive antennas.

It is important to note that in this estimation method it is not required to sample the basis vectors in specific time indices, which are arbitrarily chosen in the LS estimation method shown in Eq. (21). Furthermore, if the KL-BEM is directly applied with this method, it will not take into account the correlations between the average of the coefficients, which represent more accurately the behaviour of the channel.

C. Computational complexity

The computational complexity of this estimation method, once the projection matrix $\mathbf{\Sigma}_1\mathbf{V}_1^H\mathbf{V}_2\mathbf{\Lambda}\mathbf{U}_2^H$ is known and considering all antennas, is given by $\mathcal{O}(KNM_TM_R(Q+1))$. The computation of the $\tilde{\mathbf{G}}_k$ matrix has complexity $\mathcal{O}(N^2M_TM_R(Q+1))$, which can be further reduced if only N_{off} off-diagonal elements are considered.

In order to obtain the projection matrix, it is necesary to compute the SVD of the time-averaged channel correlation matrices. In this line, the complexity to obtain the BEM vectors is bounded by the decomposition of the $N \times K$ matrix $\mathbf{C}_{g\bar{g}}$, which is similar to the complexity of computing a KL-BEM. Furthermore, to apply this method is also necessary to know the estimation error of the frequency response algorithm used, which would increase the overall complexity of this method.

IV. Results

Computer simulations were carried out to illustrate the performance of the time-varying MIMO-OFDM channel estimation method 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 2×2 and 4×4 transmissions in the open-loop mode [17]. The carrier frequency is $f_c = 5.2$ GHz and the maximum Doppler spread is $f_d = vf_c/c$, 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

$$r(\tau) = J_0(2\pi f_d \tau), \tag{23}$$

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

$$r(\tau) = \int_{\sin^{-1}(f_{\min})}^{\sin^{-1}(f_{\max})} \cos(2\pi f_d \tau \sin\phi)d\phi \tag{24}$$

$$-j\int_{\cos^{-1}(f_{\min})}^{\cos^{-1}(f_{\max})} \sin(2\pi f_d \tau \cos\phi)d\phi, \tag{25}$$

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 will be considered.

These autocorrelation functions depend on the normalized Doppler spread f_dT, 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 [4] and KL-BEM [5] using the method based on the LS estimation [12] with order $Q = 3$. The ICI estimation algorithm is applied to groups of $K = 6$ MIMO-OFDM symbols, starting from the LMMSE estimation of the frequency response of the channel for each symbol. In order to compute Bit Error Ratio (BER) estimations, an approximation of the LS ICI cancellation algorithm is applied [18] and properly adapted to the MIMO-OFDM case. Up to $N_{\text{off}} = 4$ diagonals at each side of the main diagonal of the channel frequency response matrix were considered.

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_TM_R channels between each transmit-receive antenna pair. We considered spatially uncorrelated MIMO channels in which the impulse responses of the different channel pairs between all transmit and receive antennas are uncorrelated. The Doppler spectrum follows an asymmetrical Jakes model with $f_{\min} = 0.4$ and $f_{\max} = 1.0$. Since this algorithm takes as input the channel frequency response of individual OFDM symbols, and we have considered a pilot-aided LMMSE estimation, these starting estimations will be affected by ICI generated by the unknown data subcarriers from all the transmit antennas. In our implementation, the power of the ICI is assumed to be part of the noise vector for this initial frequency response estimation, deriving it from the time correlation 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

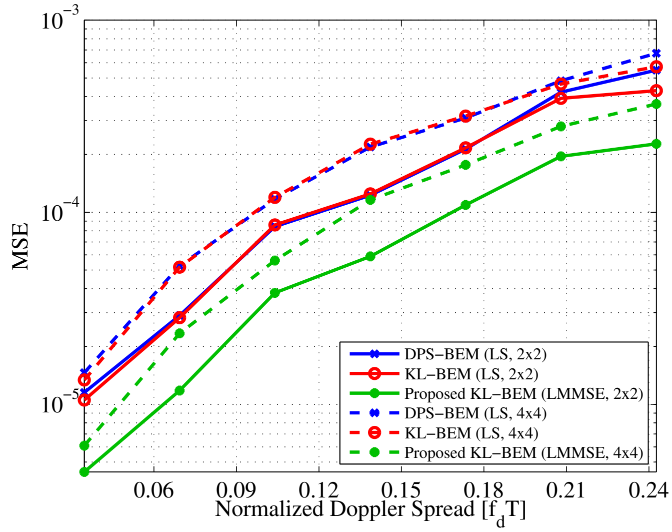

Fig. 1. MSE of the channel matrix estimation over Doppler spread at SNR = 30 dB.

Fig. 1 plots the resulting channel estimation MSE for a given SNR = 30 dB and different Doppler spreads for a 2×2 and a 4×4 MIMO-OFDM systems. The MSE is measured at the receiver, comparing the estimated coefficients with the channel generated during the simulations. All diagonals of the matrices are considered to compute the MSE. It can be seen that the proposed ICI estimation method with a modified KL-BEM outperforms those based on previous estimation algorithms, hence obtaining a constant gain independent of the Doppler spread.

Fig. 2 plots the resulting channel estimation MSE for a given Doppler spread $f_dT = 0.20$ and different SNR levels. Notice how the proposed KL-BEM method outperforms DPS-BEM and classical KL-BEM.

Fig. 3 plots the uncoded BER at the MIMO-OFDM receiver for a given Doppler spread $f_dT = 0.20$ and different SNR levels. It can also be observed that larger estimation errors shown in previous plots translate into higher bit error ratios in scenarios with more transmit antennas.

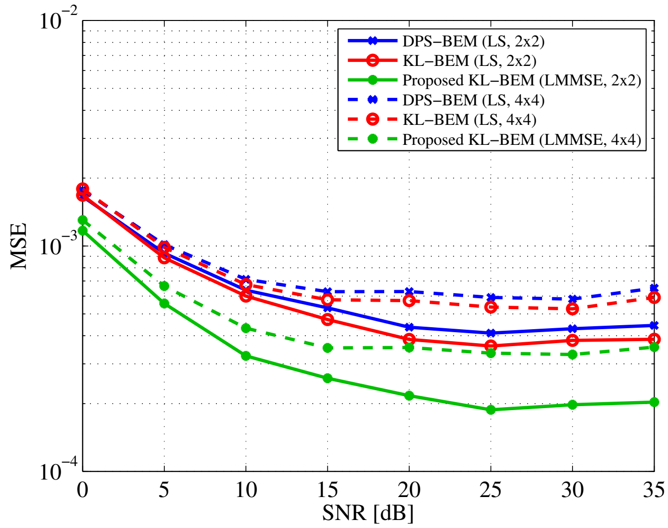

Fig. 2. MSE of the channel matrix estimation over SNR at $f_d T = 0.20$.

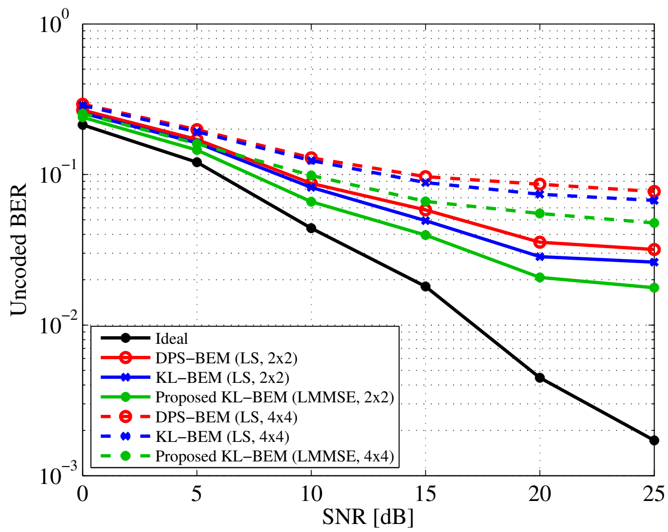

Fig. 3. BER over SNR for 2×2 and 4×4 MIMO transmissions at $f_d T = 0.20$

V. Conclusions

In this work we have presented a technique for the estimation of rapidly time-varying channels in MIMO-OFDM systems. The method utilizes a Basis Expansion Model (BEM) obtained from a low-rank approximation to the LMMSE estimation of the ICI. The proposed approach incorporates the knowledge of the channel Doppler spectrum and does not require a specific pilot subcarrier structure. The results of computer simulations show that the proposed BEM exhibits a superior performance when estimating and cancelling the ICI of rapidly time-varying MIMO-OFDM channels with non-flat Doppler spectra.

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