Three Dimensional Pilot Aided Channel Estimation for Filter Bank Multicarrier MIMO Systems with Spatial Channel Correlation

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Abstract:- In this paper we propose three dimensional pilot aided channel estimation (PACE) for MIMO/OQAM systems with spatially correlated channels. The scheme allocates pilots in time, frequency, and space. The scheme takes advantage of channel correlation to reduce the pilot overhead without lowering the channel estimation accuracy. Simulation results confirm the success of this scheme in reducing the number of pilots by half.

I. INTRODUCTION

Multicarrier modulation (MCM) has received much interest in recent wireless communications research; since MCM signals are robust to multipath fading [1]. Digital filter banks can be used to generate MCM signals. A special case of filter bank multicarrier (FBMC) modulation is orthogonal frequency division multiplexing (OFDM). The major difference between FBMC and OFDM is the improved spectral efficiency of FBMC [1]. This is due to the absence of cyclic prefix that is used in OFDM. Another advantage of FBMC modulation is that it uses frequency well-localized pulse shaping instead of rectangular pulse shaping [2]. Subcarriers are modulated using offset QAM (OQAM) [1].

Multiple-input multiple-output (MIMO) systems offer significant capacity gains over independent narrow-band channels having the same total bandwidth [3]. Spatial diversity can be used to achieve performance gains in MIMO systems. Maximum diversity gain can be obtained when fading is uncorrelated across antenna pairs [4]. Over a multipath channel, it makes sense to use MCM with MIMO [5]. Channel estimation (CE) is very important in wireless systems [6]. Spatial correlation of antennas has been exploited to enhance the CE accuracy [7], and/or to raise the spectral efficiency through reducing the pilot density [8] by inserting pilot symbols at a subset of the transmit antennas. CE has been proposed for OFDM systems by exploiting the channel correlation in the time and frequency domains [9]. We refer to these schemes as two-dimensional (2D). For uncorrelated transmit antennas, pilot symbols decrease the capacity gain, especially when the number of transmit antennas is large [10]. In MIMO-OFDM systems, with spatially correlated transmit antennas, channel correlations can be exploited to add a third (spatial) dimension to the CE process [7][8].

Channel estimation in FBMC systems is more difficult than in OFDM because of intrinsic imaginary interference suffered by adjacent subcarriers. This interference results from limiting the orthogonality of subcarriers to the real field. Channel estimation in FBMC systems has been recently studied for preamble-based [11] and scattered pilots-based [12] schemes.

In this paper, we study scattered-pilots channel estimation for FMBC-MIMO systems with correlated transmit antennas. We propose three dimensional pilot aided channel estimation (3D-PACE) for FMBC-MIMO systems operating on spatially correlated fading channels. Pilots are distributed in time, frequency, and space. The proposed scheme copes with the intrinsic interference of FMBC-MIMO, and takes advantage of spatial correlations to reduce the total pilot overhead; without degrading the CE performance. Simulation results confirm success of the proposed scheme in reducing the number of pilots by a half.

This paper is organized as follows. In section II, we introduce FBMC modulation and the channel model. 3D-PACE FMBC-MIMO CE and proposed PACE scheme is described in section III. Section III also discusses pilot reduction with spatially correlated transmit antennas. Results are given in section IV. The paper is concluded in section V.

II. FBMC MODULATION AND CHANNEL MODELING

A. FBMC Modulation

Let's consider a multicarrier wireless communication system employing an even number of subcarriers $M = 2N$. The signal at the output of the SFB can be expressed as follows [1]:

$$s(t) = \sum_{m=0}^{M-1} \sum_{n=-\infty}^{\infty} a_{m,n} g(t - n\tau_0) j^{m+n} e^{j2\pi m F_0 t} \quad (1)$$

where $a_{m,n}$ are the OQAM symbols (the real and imaginary parts of the QAM symbols) and $F_0 = 1/T$ is the subcarrier spacing. The multicarrier symbol duration is $T = MT_s$. The time offset between the real and imaginary parts of the symbols is $\tau_0 = T/2$ [1].

The prototype filter impulse response $g(t)$ will be assumed to be real and symmetric. Let's define

$$g_{m,n}(t) = g(t - n\tau_0)\, j^{m+n} e^{j2\pi m F_0 t} \quad (2)$$

This allows (1) to be rewritten in the form

$$s(t) = \sum_{m=0}^{M-1} \sum_{n=-\infty}^{\infty} a_{m,n} g_{m,n}(t) \quad (3)$$

Assuming a noise-free time varying channel, the demodulated symbol over the m-th subcarrier during the n-th symbol period is given by:

$$r_{m,n} = a_{m,n} h_{m,n} + I'_{m,n} \quad (4)$$

where $h_{m,n}$ is the channel response, and $I'_{m,n}$ is the intrinsic interference term, given by:

$$I'_{m,n} = \sum_{(p,q)\neq(m.n)} \sum h_{p,q} a_{p,q} \int g_{p,q}(t) g^*_{m,n}(t) dt \quad (5)$$

Assuming a well localized prototype filter, the summations in (5) have significant values only for $(p,q) \in \Omega_{m,n}$, where $\Omega_{m,n}$ is the subset of subcarrier and time indices near (m,n). Equation (4) can be approximated by

$$r_{m,n} \approx h_{m,n} \left(a_{m,n} + I_{m,n} \right) \quad (6)$$

where

$$I_{m,n} \approx \sum_{(p,q)\in\Omega_{m,n}} \sum a_{p,q} \int g_{p,q}(t) g^*_{m,n}(t)\, dt \quad (7)$$

In this work, we use a nearly perfect reconstruction (NPR) prototype filter of length $L = 4M$ [2], the used prototype filter is designed based on the frequency sampling method and has a very low NPR noise variance which is ignored in our analysis. Table 1 illustrates the most effective values of the transmultiplexer response ($g_{p,q}$) of the reference filter bank system [2]. Ignoring the common factor $h_{m,n}$, $r_{m,n}$ can be expressed as the sum of three terms as follows:

$$r_{m,n} = j^{m+n} a_{m,n} g_{m,n} + \sum_{(p,q)\in\Omega^*_{m,n}} \sum j^{p+q} a_{p,q} g_{p,q} \quad (8)$$

where $\Omega^*_{m,n} = \Omega_{m,n} - \{(m,n)\}$, simple mathematical manipulation of (8) leads to

$$r_{m,n} \approx a_{m,n} + jI_{m,n} \quad (9)$$

where

$$I_{m,n} = \sum_{(p,q)\in\Omega^*_{m,n}} \sum j^{p+q} a_{p,q} g_{p,q} \quad (10)$$

B. Channel Modeling

The channel is represented using a P-tap delay line filter model. User mobility will be assumed to induce weight variations in the tap coefficients. Channel tap weights will be denoted by $\left\{ c_{p,n} \right\}_{p=0}^{P-1}$. Tap weights will be assumed to be complex-valued. Following [8], the channel impulse response can be represented in the form

$$h^{(\mu)}_{m,n} = \sum_{p=0}^{P-1} c_{p,n} \alpha^{(\mu)}(\varphi_p) e^{-j2\pi \frac{m\tau_p}{T}} \quad (11)$$

where $\alpha^{(\mu)}(\varphi)$ is the response of antenna μ at departure angle φ, φ_p is the departure angle of path p, τ_p is the of path p and T is the multicarrier symbol duration. Assuming the transmit antennas are placed in a uniform linear array (ULA) with spacing d, the response of antenna μ takes the form [13]

$$\alpha^{(\mu)}(\varphi_p) = e^{j2\pi \frac{d}{\lambda} \mu \sin\varphi_p} \quad (12)$$

where λ is the carrier wavelength. Let's define the 3-D channel correlation function

$$R(\mu', m', n') = \mathrm{E}\left[h^{(\mu)}_{m,n} \left(h^{(\mu+\mu')}_{m+m',n+n'} \right)^* \right] \quad (13)$$

Note that this function involves a frequency index m, a time index n and a spatial index μ. In practice, it can be assumed that the time correlation is the same of all P paths. Spatial correlations, however, are usually different for different paths. The correlation function in (13) can be expressed as the product of a time correlation function and a frequency-space correlation function $R_{f,s}(\mu', m')$ [8]

$$R(\mu', m', n') = R_t(n') R_{f,s}(\mu', m') \quad (14)$$

The time correlation function is given by [14]:

$$R_t(n') = J_0(2\pi n' f_{d,\max} T) \quad (15)$$

where $f_{d,\max}$ is the maximum Doppler frequency and $J_0(.)$ is the first kind Bessel function of zero order. The frequency-space correlation function is given by[8]:

$$R_{s,f}(\mu', m') = \sum_{p=0}^{P-1} R_{p,f}(m') R_{p,s}(\mu') \quad (16)$$

The frequency correlation function $R_{p,f}(m')$ is given by [8]:

$$R_{p,f}(m') = \sigma_p^2 e^{-j2\pi\tau_p \frac{m'}{T}} \quad (17)$$

Where $\sigma_p^2 = \mathrm{E}\left[\left| c_{p,n} \right|^2 \right]$. The spatial correlation function $R_{p,s}(\mu')$ between antennas μ and $\mu + \mu'$, for the considered channel model, has the form [8]

$$R_{p,s}(\mu') = e^{j2\pi d \mu' \sin\varphi_p} \quad (18)$$

Table 1: Transmultiplexer Response

	n-2	n-1	n	n+1	n+2
m-1	-.1250	-j.2058	0.2393	j.2058	-.1250
m	0	.5644	1	.5644	0
m+1	-.1250	j.2058	0.2393	-j.2058	-.1250

III. 3D PACE MIMO-OFDM/OQAM CE

A. Background

In the proposed scheme N_t transmit antennas and one receive antenna are assumed. Real symbols $a_{m,n}^{(\mu)}$ are assumed to be transmitted over antenna μ, where $0 \le \mu < N_t$. After demodulation we have

$$r_{m,n} = \sum_{\mu=0}^{N_t-1} r_{m,n}^{(\mu)} = \sum_{\mu=0}^{N_t-1} h_{m,n}^{(\mu)}(a_{m,n}^{(\mu)} + I_{m,n}^{(\mu)}) + z_{m,n} \quad (19)$$

where $h_{m,n}^{(\mu)}$ denotes the channel transfer function between transmit antenna μ and the receive antenna, $I_{m,n}^{(\mu)}$ is the intrinsic interference due to transmitted $a_{m,n}^{(\mu)}$ over antenna μ, and $z_{m,n}$ is zero mean additive white Gaussian noise (AWGN) with variance N_0.

In 2D PACE, samples of the transmission channel in the frequency and time directions are estimated through the use of pilots that are inserted at given time-frequency locations. After estimating the channel coefficients at pilot locations, the whole channel can be recovered by interpolating these samples. In systems with multiple correlated transmit antennas, this idea can be extended to the spatial domain, such that the pilots are distributed in three dimensions (space, frequency, and time).

For simplicity, let us start with a single transmit antenna. If a pilot $p_{m,n}$ is transmitted over subcarrier m, at time n, and from antenna μ, (19) becomes

$$r_{m,n} = h_{m,n}\left(p_{m,n} + I_{m,n}\right) + z_{m,n} \quad (20)$$

Define $\tilde{h}_{m,n}$ as the estimated channel coefficient at pilot $p_{m,n}$ location as:

$$\tilde{h}_{m,n} = \frac{r_{m,n}}{\left(p_{m,n} + I_{m,n}\right)} = h_{m,n} + \frac{z_{m,n}}{\left(p_{m,n} + I_{m,n}\right)} \quad (21)$$

It is clear that $\tilde{h}_{m,n}$ cannot be recovered immediately even with the knowledge of $p_{m,n}$; because of the presence of the intrinsic interference term $I_{m,n}$. This problem can be solved using the so-called auxiliary pilot approach, as presented in [12]. In this approach an auxiliary pilot $x_{m,n}$ is required to mitigate the interference that is generated by neighboring symbols. Interference can be compensated by setting the value of one of the neighboring symbols, say $x_{m,n}$, to the total intrinsic interference, making the summation in (10) almost equal zero. To minimize the magnitude of an auxiliary pilot in the used NPR filter, we locate the auxiliary pilot immediately before or after the main pilot $p_{m,n}$, i.e. at $(m, n \pm 1)$ [15]. Without loss of generality, in this paper we locate the auxiliary pilot immediately before the main pilot. By using an auxiliary pilot two OQAM symbols (or one QAM symbol) are wasted. Thus, the pilot overhead using this scheme is equal to the pilot overhead in OFDM.

B. Proposed Pilot Aided Channel Estimation Scheme

Let's consider pilot signal $p_{m_p,n_p}^{(\mu_p)}$ transmitted over antenna μ_p, where $0 < \mu_p < N_t - 1$, and at frequency-time index (m_p, n_p). This represents the pilot used to recover the CTF coefficient $h_{m_p,n_p}^{(\mu_p)}$. Now, from (19),

$$r_{m,n} = h_{m,n}^{(\mu_p)}\left(p_{m,n}^{(\mu_p)} + I_{m_p,n_p}^{(\mu_p)}\right) + \sum_{\substack{\mu=0 \\ \mu \ne \mu_p}}^{N_t-1} h_{m,n}^{(\mu)}\left(a_{m,n}^{(\mu)} + I_{m,n}^{(\mu)}\right) + z_{m,n} \quad (22)$$

This can be rewritten in the form

$$r_{m,n} = h_{m,n}^{(\mu_p)} p_{m,n}^{(\mu_p)} + I_a + I_I + z_{m,n} \quad (23)$$

where

$$I_a = \sum_{\substack{\mu=0 \\ \mu \ne \mu_p}}^{N_t-1} h_{m,n}^{(\mu)} a_{m,n}^{(\mu)}, I_I = \sum_{\mu=0}^{N_t-1} h_{m,n}^{(\mu)} I_{m,n}^{(\mu)} \quad (24)$$

To recover $h_{m,n}^{(\mu_p)}$, both I_a and I_I terms in (23) should be zero. A simple way to zero these terms is to set $a_{m,n}^{(\mu)} = 0$ for $\mu \ne \mu_p$ (to cancel I_a), and to use auxiliary pilots over all antennas (to cancel I_I), as shown in Figure 1. Then, $r_{m,n}$ becomes

$$r_{m,n} = h_{m,n}^{(\mu_p)} p_{m,n}^{(\mu_p)} + z_{m,n} \quad (25)$$

Based on the above, $h_{m_p,n_p}^{(\mu_p)}$ can be recovered by:

$$\tilde{h}_{m,n}^{(\mu_p)} = \frac{r_{m,n}}{p_{m,n}^{(\mu_p)}} + \frac{z_{m,n}}{p_{m,n}^{(\mu_p)}} = h_{m,n}^{(\mu_p)} + \frac{z_{m,n}}{p_{m,n}^{(\mu_p)}} \quad (26)$$

The transmission symbols can be written in the form

$$a_{m,n}^{(\mu)} = \begin{cases} 0, & \text{pilot at } (m,n) \\ d_{m,n}^{(\mu)}, & \text{OQAM data symbol} \\ p_{m,n}^{(\mu)}, & \text{pilot symbol} \\ x_{m,n}^{(\mu)}, & \text{aux. pilot at } (m, n+1) \end{cases} \quad (27)$$

C. *Pilot Reduction with Spatially Correlated Transmit Antennas*

For the proposed three dimensional (3D) PACE, spatial correlations among antennas are exploited to reduce the pilot overhead. By allowing for a D_s spatial pilot spacing, pilots are only inserted on a subset of transmit antennas. The channel response of all transmit antennas is obtained by interpolation in space.

Let N_s, N_f and N_t be the numbers of pilots which are distributed in space, frequency and time, respectively. The total number of pilot is $N_P = N_s N_f N_t$. A 3D pilot grid can be uniquely described by [16]

$$\underline{p} = D\,\underline{\tilde{p}} + \underline{p}_0 \tag{28}$$

where

$$D = \begin{bmatrix} D_s & \delta_{sf} & \delta_{st} \\ d_{sf} & D_f & \delta_{ft} \\ d_{st} & d_{ft} & D_t \end{bmatrix}, \quad \begin{matrix} \underline{p} = \begin{bmatrix} \mu_p & m_p & n_p \end{bmatrix}^T \\ \underline{\tilde{p}} = \begin{bmatrix} \tilde{\mu} & \tilde{m} & \tilde{n} \end{bmatrix}^T \\ \underline{p}_0 = \begin{bmatrix} \mu_0 & m_0 & n_0 \end{bmatrix}^T \end{matrix} \tag{29}$$

The symbols D_f and D_t denote the pilot spacings in frequency and time, respectively, $\underline{p}$ denotes the location of the pilot signal, $\underline{\tilde{p}}$ is the pilot index vector, and $\underline{p}_0$ accounts for the shift of the first pilot with respect to $[0 \;\; 0 \;\; 0]^T$. The sampling matrix D completely determines pilot pattern.

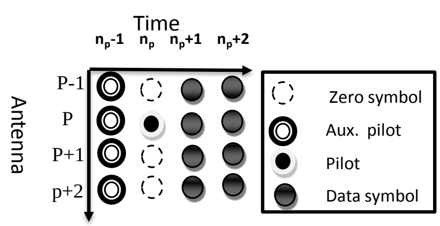

Figure 1: Pilots in the frequency-time plane

In general, non-zero off-diagonal elements of D assemble a non-rectangular pilot grid; setting $d_{ft} \neq 0$ and/or $\delta_{ft} \neq 0$ in (29) results in a shifted time-frequency grid over indices m and n. The parameter pairs d_{sf}, δ_{sf} and d_{st}, δ_{st} shift pilots along the space-frequency and space-time dimensions, respectively. Not all N_P pilots may be utilized for channel estimation. Rather, the estimation of the CTF at a certain subcarrier may be restricted to a subset of pilots (closed set to estimated position) in time, $M_t \leq N_t$, frequency, $M_f \leq N_f$, and space, $M_s \leq N_s$. The received pilot sequence of dimension $M \times 1$, with $M = M_f M_t M_s$, is a subsampled version of the 3D CTF corrupted by noise.

D. *Derivation of the Wiener Interpolation Filter*

Based on the minimum mean squared error (MMSE) criterion, the estimate of the 3D CTF, denoted by $\hat{h}_{m,n}^{(\mu)}$, is determined using a 3D FIR Wiener interpolation filter (WIF), with M taps, and coefficients $\mathrm{w}^H[m,n,\mu]$.

$$\hat{h}_{m,n}^{(\mu)} = \mathrm{w}^H[m_d, n_d, \mu]\underline{\tilde{\mathrm{h}}} \tag{30}$$

where $\underline{\tilde{\mathrm{h}}}$ denotes the received pilot sequence of length M. The WIF $\mathrm{w}^H[m_d, n_d, \mu] = \mathrm{R}^{-1}\mathrm{p}[m_d, n_d, \mu]$ minimizes the means squared error (MSE) between the desired response $h_{m,n}^{(\mu)}$ and the filtered output $\hat{h}_{m,n}^{(\mu)}$ given $\underline{\tilde{\mathrm{h}}}$. The matrix

$$\mathrm{R} = E[\tilde{\mathrm{Y}}\tilde{\mathrm{Y}}^H] = E[\tilde{\mathrm{H}}\tilde{\mathrm{H}}^H] + \frac{1}{\gamma}\mathrm{I} \tag{31}$$

is the $M \times M$ 3D correlation matrix of the received pilots $\underline{\tilde{\mathrm{h}}}$, where $E[\tilde{\mathrm{H}}\tilde{\mathrm{H}}^H]$ is the $M \times M$ correlation matrix of the CTF at pilot positions excluding the AWGN term, γ is the SNR and I denotes the $M \times M$ identity matrix. The 3D cross-correlation function $\mathrm{p}[m_d, n_d, \mu]$ with dimension $M \times 1$ represents the cross-correlation between $\underline{\tilde{\mathrm{h}}}$ and the desired response $h_{m,n}^{(\mu)}$.

IV. SIMULATION RESULTS

Simulations have been carried out with the channel in [8]. We consider FBMC-MIMO systems with $N_t = \{2,4\}$ transmit antennas, with $\lambda/d = 0.5$ element spacing, based on the IST-WINNER project [17] and one receive antenna. Simulation parameters are borrowed from WiMAX. Signal bandwidth is $B = 10$ MHz, $M = 1024$ subcarriers, subcarrier spacing is $F_0 = 10.94$ kHz, and the sampling frequency of QAM symbols is 11.2 MHz. At 2.5 GHz, we consider an urban mobility scenario with velocities up to 50 km/h. Pilot spacings are chosen to be $D_f = 16$, $D_t = 2$, and $D_s = \{1,2\}$ [8]. WLOG, we set $M_f = 16$, $M_t = 2$, and $M_s = N_T / D_s$. The matrix D can be in the form

$$D = \begin{bmatrix} D_s & 1 & 0 \\ 0 & D_f & 0 \\ 0 & 0 & D_t \end{bmatrix} \tag{32}$$

This gives the same pilot grid for all antennas. To handle the pilot and the auxiliary pilot locations, pilot patterns are shifted in time and/or frequency keeping in mind that no auxiliary pilot windows are overlapped. This shift can be achieved by setting $\underline{p}_0$ in (28). For example, and without loss of generality, the values for $\underline{p}_0$ can be put in the form

$$\underline{p}_0 = \begin{bmatrix} [0 \quad 0 \quad 0], & \text{for antenna 1} \\ [0 \quad 3 \quad 0], & \text{for antenna 2} \\ [5 \quad 3 \quad 0], & \text{for antenna 3} \\ [5 \quad 0 \quad 0], & \text{for antenna 4} \end{bmatrix}^T \quad (33)$$

where the setting in (33) is not unique. To show the performance of our channel estimation scheme, we have compared three cases of simulation by plotting the MSE as function of the SNR. The first case is 3D-PACE with $D_S = 1$, where pilots are inserted at all transmit antennas. The second case is 3D-PACE with $D_S = 2$, where pilots are inserted only at a subset of the transmit antennas. The third case is conventional 2D-PACE, which does not attempt to exploit spatial correlation. The cases 3D-PACE with $D_S = 1$ and 2D-PACE employ the same pilot grid and exhibit the same pilot overhead.

To compare the channel estimation performance of high spatial channel correlation to the channel estimation performance of channel with low spatial correlation, the two channel models are applied to the proposed 3D-channel estimation, WINNER typical urban channel model (model C2 in [18]) and no line of sight indoor office environment channel model. In the C2 channel model, the angles of departure of the channel paths are distributed with a 35^o angular spread. This value of small angular spread encourages the transmit antennas to be highly correlated in space.

An example of the system has been modeled by C2 channel is an outdoor macro-cell with a MIMO downlink where the transmitter base station is located above rooftop. The departing ray angles are included within a narrow angular spread rising the spatially correlation between transmit antennas. In the A1 channel model, the angles of departure of the channel paths are distributed with angular spread of 200^o which reduces the spatial correlation at the transmit antennas. A1 channel model describes an indoor office environment where the base station is located under the rooftop.

Figure 2 shows the performance of the proposed FBMC-MIMO ($N_t = 4$) system applied to both channel models(C2 and A1). For the high spatial correlated channel (C2), If 3D-PACE with $D_S = 1$ and 2D-PACE are compared, we can see that the MSE performance of the former is lower than that of the latter, because 3D-PACE, with $D_S = 1$ exploits the spatial correlation for an improved channel estimation. The same holds true when we compare 3D-PACE, with $D_S = 1$ to 3D-PACE, with $D_S = 2$, because the 3D-PACE, with $D_S = 1$ uses twice the number of pilots in its estimation. Now, for 2D-PACE and 3D-PACE, with $D_S = 1$, it can be noticed that both 2D-PACE and 3D-PACE, $D_S = 1$ have approximately the same MSE performance.

For low spatial correlation channel (A1), the performance of 3D-PACE, with $D_S = 1$ is almost the same as the performance of 2D-PACE. This is an expected result because the channel has very low spatial correlation that cannot be exploited to improve the channel estimation performance.

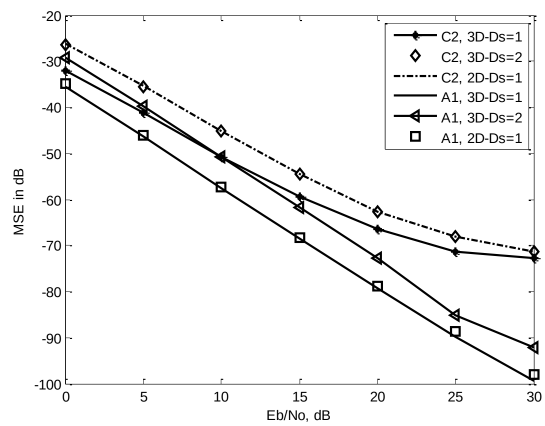

Figure 2: MSE performance for 4x1 FBMC-MIMO

Figure 3 shows that MSE performance for FBMC-MIMO $N_t = 2$ applied to C2 and A1 channels. For C2 model, the performance of 3D-PACE with $D_S = 1$ is slightly improved compared to the performance of 2D-PACE. And the performance of 3D-PACE with $D_S = 2$ is worse than the performance of 2D-PACE with $N_t = 2$ transmit antenna.

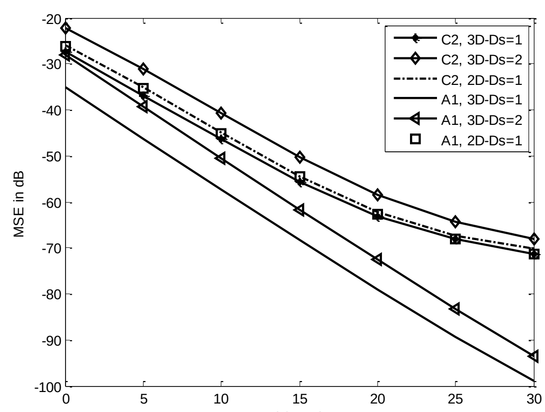

Figure 3: MSE performance for 2x1 FBMC-MIMO

Now, for 2D-PACE and 3D-PACE with $D_s = 2$, it can be noticed that the performance of 2D-PACE is better than the performance of 3D-PACE, with $D_s = 2$. For A1 channel model, the MSE performance of 3D-PACE with $D_s = 1$ is almost the same as the performance of 2D-PACE, and the performance of 3D-PACE with $D_s = 2$ is worse than the performance of 2D-PACE with $N_t = 2$ transmit antennas.

It can be concluded that the proposed 3D-PACE is efficient over high spatially correlated channel (like C2 channel), and that the pilot overhead can be reduced (up to a factor of two for 4x1 FBMC-MIMO) without lowering the CE performance compared with the performance of 2D-PACE. On the other hand, the 3D-PACE is inefficient over channels with low spatial correlation (like A1 channel).

V. CONCLUSION

Three dimensional pilot aided channel estimation (PACE) has been proposed for MIMO/OQAM systems with spatially correlated channels. The proposed scheme is based on allocating pilots in the three dimensions of time, frequency, and space. The proposed scheme copes with the intrinsic interference of MIMO/OQAM, and takes advantage of spatial channel correlation to reduce the total pilot overhead without lowering the channel estimation accuracy. Simulation results confirm the success of the proposed scheme in reducing the total pilots by half.

It can be concluded that the proposed 3D-PACE is efficient over highly spatially correlated channels (like C2 channel) and the pilot overhead can be reduced (up to a factor of two for 4x1 FBMC-MIMO) without lowering the CE performance compared with the performance of 2D-PACE. On the other hand, the 3D-PACE is inefficient over the channel with low spatial correlation (like A1 channel).

VI. ACKNOWLEDGMENTS

The authors acknowledge the financial support by the European Union FP7-ICT project EMPhAtiC under grant agreement no. 318362, and the COST Action IC 902.

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