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Power Delay Profile Estimation for MIMO-OFDM Systems under LMMSE Channel Estimation Approach

Sriramula Poornima, Lendale Praveen Kumar

Abstract- *A multiple-input multiple-output (MIMO) communication system combined with the orthogonal frequency division multiplexing (OFDM) modulation technique can achieve reliable high data rate transmission over broadband wireless channels. But the performance gain depends heavily on accurate channel estimation. In linear-minimum-mean-Square-error(LMMSE) channel estimation for multicarrier system, one needs to know the channel statistics. We propose a power delay profile (PDP) estimation technique to obtain the frequency domain channel statistics at the receiver. The pilot symbols of all transmit antenna ports are used in estimating the PDP. The estimated PDP is used to generate the LMMSE filter Coefficients for data-subcarrier channel estimation. The distortions caused by null subcarriers and an insufficient number of samples for PDP estimation are also considered. The proposed technique effectively reduces the distortions for accurate PDP estimation. Simulation results show that the performance of LMMSE channel estimation using the proposed PDP estimate approaches that of Wiener filtering due to the mitigation of distortion effects.*

Index Terms- *Channel estimation, PDP, MIMO, OFDM, LMMSE, 3GPP-LTE.*

I.INTRODUCTION

Telecommunications in the current information age is increasingly relying on the wireless link. This is because wireless communication has made possible a variety of services ranging from voice to data and now to multimedia. Consequently, demand for new wireless capacity is growing rapidly at a very alarming rate. In a bid to cope with challenges of increasing demand for higher data rate, better quality of service, and higher network capacity, there is a migration from Single Input Single Output (SISO) antenna technology to a more promising Multiple Input Multiple Output (MIMO) antenna technology. On the other hand, Orthogonal Frequency Division Multiplexing (OFDM) technique has emerged as a very popular multi-carrier modulation technique to combat the problems associated with physical properties of the wireless channels such as multipath fading, dispersion, and interference. The combination of MIMO technology with OFDM techniques, known as MIMO-OFDM Systems, is considered as a promising solution to enhance the data rate of future broadband wireless communication Systems. However, this enhancement requires accurate and computationally efficient channel state information to achieve the maximum diversity or multiplexing gain [5]-[7].

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***** Correspondence Author (s)
Sriramula Poornima, M.Tech Student, Department of Electronics & Communication Engineering, Anurag Group of Institutions, Hyderabad, AP, India.
Lendale Praveen Kumar, Assistant Professor, Department of Electronics & Communication Engineering, Anurag Group of Institutions, Hyderabad, AP, India.

The pilot-aided channel estimation, based on the linear minimum mean square error (LMMSE) technique, is optimum in the sense of minimizing mean square error (MSE) when the receiver knows the channel statistics. To obtain the frequency domain channel statistics at the receiver, power delay profile (PDP) estimation schemes have been proposed. These schemes are based on the maximum likelihood (ML) estimation by taking advantage of the cyclic prefix (CP) segment of OFDM symbols. However, the ML PDP estimators require very high computational complexity for obtaining an accurate PDP.

Another approach is by employing an approximated PDP (i.e., uniform or exponential model) with the estimation of second-order channel statistics. But the performance of LMMSE channel estimator with the approximated PDP degrades due to correlation mismatch and the estimation error of delay parameters. To reduce the mismatch in the frequency domain, PDP is estimated using only the pilot symbols of all transmit antenna ports with low computational complexity. . This technique effectively mitigates the distortion effects, incurred by null subcarriers and an insufficient number of estimated channel impulse response (CIR) samples. The performance of LMMSE channel estimation with the proposed PDP estimate approaches that of Wiener filtering. The performance of the estimators is presented in terms of the mean square error (MSE) and pilot SNR.

II. SYSTEM MODEL

Before introducing the PDP estimation and LMMSE channel estimation technique, we briefly describe a MIMO-OFDM system.

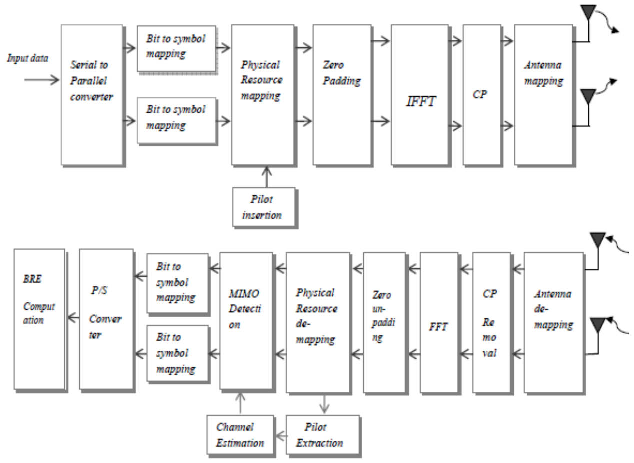

Fig. 1.MIMO-OFDM system.

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TheMIMO-OFDM system with P transmit and Q receive antennas, and K total subcarriers is considered. Suppose that the MIMO-OFDM system transmits subcarriers at the central spectrum assigned for data and pilots with $K-K_d$ virtual subcarriers, in order to control interferences with other systems. The CIRs corresponding to different transmit and receive antennas in MIMO systems usually have the same PDP [12].

Let us consider the pilot subcarriers used to be a QPSK modulated signal from known sequences between the transmitter and receiver which is represented as $C_P\ [k_p,\ n_p]$ for p^{th} transmit antenna at n_p^{th} OFDM symbol. Fig.2 shows the pilot symbol arrangement in a physical resource block (PRB), that are distributed over a time and frequency grid, to retain its orthogonality among different transmit and receive antennas.

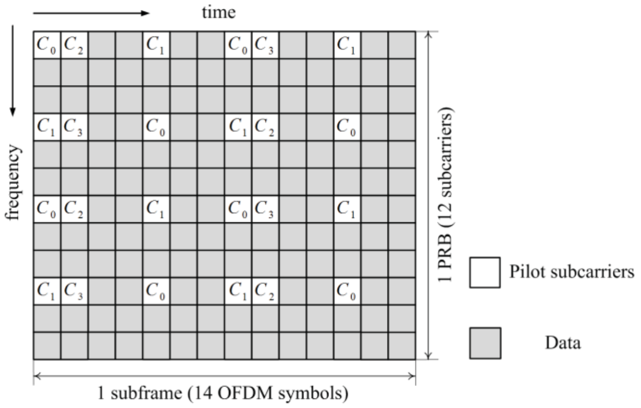

Fig. 2. Pilot symbol arrangement in a physical resource block (PRB) of the LTE OFDM system.

$K_p \in \mathcal{F}_p$ and $n_p \in T_p$ represent the index sets for the pilot subcarriers of the p^{th} antenna port in the frequency and time domains, respectively. At the n_p^{th} OFDM symbol, the number of pilot subcarriers is defined as $K_p = |\mathcal{F}_p|$. The pilot inserted OFDM symbol is transmitted over the wireless channel after performing an inverse fast Fourier transform (IFFT) and adding a CP. It is assumed that the length of CP,L_g, is longer than the channel maximum delay, L_{ch}, making the channel matrix circulant ($L_{ch} \leq L_g$). The received pilot symbol for the q^{th} receive antenna after perfect synchronization, the removal of CP, and FFT operation can be represented as,

$$y_q[n_p] = \text{diag}(x_p)\mathbf{F}_p\mathbf{h}_{p,q} + \mathbf{n}_q\ , \qquad (1)$$

where x_p is a pilot vector at the n_p^{th} OFDM symbol for $i_k \in F_p$ and k=1,2,.....K_p and is given as $x_p=[C_p(i_1,n_p),C_p(i_2,n_p)........C_p(i_{K_p},\ n_p)]^T$. *diag*(xp) is the K_pXK_p diagonal matrix whose entries are K_p elements of the vector x_p, $\mathbf{F}_p$ is a K_pXL_g matrix with the (i_k, l)th entry and is given as $[\mathbf{F}_p]_{i_k,l} = 1/K\ \exp\{-j2\pi i_k l/K\}$ where $i_k \in F_p$ and l= 0,1,...., L_g-1, $h_{p,q}$ is an L_gX1 CIR matrix at the p^{th} transmit antenna and q^{th} receive antenna and is given as $h_{p,q} = [\ h_{p,q}[n_p,0],\ h_{p,q}[n_p,1]......h_{p,q}[n_p,L_{ch}],\ ...0]^T$. $(.)^T$ represents the transpose operation. $\mathbf{n}_q$ is a complex Additive White Gaussian Noise (AWGN) vector at the q^{th} receive antenna with each entry having a zero mean and variance of σ_n.The average power of $h_{p,q}$ depend on the delay profiles of the wireless channels.

III.PDP ESTIMATION

In this section, we will present derivation of PDP in MIMO-OFDM systems, including PDP estimation in practical MIMO-OFDM Systems. Then, the estimated PDP is used to obtain the frequency-domain channel correlation in the LMMSE channel estimator.

Using the regularized least squares(RLS) channel estimation the channel impulse response can be estimated approximately from (1),at $(p,q)^{th}$ antenna port with a fixed length of L_g as

$$\hat{h}_{R,p,q} = \left(\boldsymbol{F}_p^H F_p + \epsilon\, \boldsymbol{I}_{L_g}\right)^{-1} \boldsymbol{F}_p^H diag(\boldsymbol{x}_p)^H \boldsymbol{y}_q[n_p]$$
$$\triangleq \boldsymbol{W}_{RLS,p}\boldsymbol{y}_q[n_p], \qquad (2)$$

where ϵ=0.001 is a small regularization parameter, and $\boldsymbol{I}_{L_g}$ is the $Lg \times Lg$ identity matrix. $\boldsymbol{F}_p^H F_p$ in (2) is ill-conditioned due to the sparsity of pilot tones in the frequency domain and the presence of virtual subcarriers [8]. To derive the PDP from the

estimated CIR in (2), the ensemble average of $\hat{\boldsymbol{h}}_{R,p,q}\hat{\boldsymbol{h}}_{R.p.q}{}^H$ is

$$E\{\hat{\boldsymbol{h}}_{R,p,q}\hat{\boldsymbol{h}}_{R,p,q}^H\} = \boldsymbol{W}\boldsymbol{R}_{hh}\boldsymbol{W}^H + \sigma_n^2 \boldsymbol{W}_{RLS,p}\boldsymbol{W}_{RLS,p}^H \qquad (3)$$

$$\boldsymbol{R}_{hh} = E\{\boldsymbol{h}_{p.q}\boldsymbol{h}_{p,q}^H\} \qquad (4)$$

$$\boldsymbol{W} = \left(\boldsymbol{F}_p^H \boldsymbol{F}_p + \epsilon\, \boldsymbol{I}_{L_g}\right)^{-1} \boldsymbol{F}_p^H\, \boldsymbol{F}_p \qquad (5)$$

The PDP of multipath channel within the length of L_g ,is given by the diagonal elements of the channel covariance matrix, R_{hh}, and all off-diagonal elements are zeros. Hence, the covariance matrix can be expressed as

$$\boldsymbol{R}_{hh} = diag(\boldsymbol{P}_h) \qquad (6)$$

where $\boldsymbol{P}_h = [p_0, p_1, \cdots, p_{L_{ch}}, 0 \cdots, 0]^T$ and $p_l = E\left\{\left|h_{p,q}[n_p, l]\right|^2\right\}$. Due to the presence matrix $\boldsymbol{F}_p^H\boldsymbol{F}_p$, **W** is ill conditioned. So $\boldsymbol{R}_{hh}$ is distorted by W, thus we investigate the method for eliminating the spectral leakage of W.The covariance matrix of the estimated CIR is defined as

$$\boldsymbol{R}_{\hat{h}\hat{h}} = \boldsymbol{W}\boldsymbol{R}_{hh}\boldsymbol{W}^H \qquad (7)$$

$$\therefore \boldsymbol{R}_{\hat{h}\hat{h}} = \sum_{l=0}^{l_g-1} \boldsymbol{W} diag(p_l \boldsymbol{u}_l)\boldsymbol{W}^H \qquad (8)$$

where $\mathbf{u}_l$ is a unit vector with the lth entry being one and otherwise zeros. Let $\boldsymbol{P}_{\hat{h}} = Dg(\boldsymbol{R}_{\hat{h}\hat{h}})$ and $\boldsymbol{t}_l = Dg(\boldsymbol{W}diag((\boldsymbol{u}_l)\boldsymbol{W}^H)$,where Dg(**A**) is the column vector containing all the diagonal elements of **A**. Then, the relation in (8) is simplified as

$$P_{\hat{h}} = p_0\boldsymbol{t}_0 + p_1\boldsymbol{t}_1 + \cdots + p_{L_{g-1}}\boldsymbol{t}_{L_{g-1}} \triangleq \boldsymbol{T}_{\boldsymbol{P}_h} \qquad (9)$$

where T=[t_0, t_1, ..., t_{L_g-1}] is defined as a distortion matrix by W. It is noted that the distortion matrix is a strictly diagonally dominant matrixsince the non-diagonal elements of **T** are composed of the leakage powers of u_i for all *i*. From the Gershgorin circle theorem, a strictly diagonally dominant matrix is non-singular [13]. In addition, the distortion matrix is a well-conditioned matrix. Hence, the distortion of W can be eliminated as

$$\boldsymbol{P}_h = \boldsymbol{T}^{-1}P_{\hat{h}} = E\{g_{p,q}[n_p]\} - \sigma_n^2\widehat{W}, \qquad (10)$$

Where $g_{p,q}[n_p] = T^{-1}Dg\left(\hat{h}_{R,p,q}\hat{h}_{R.p.q}^H\right)$ is defined as the received sample vector for estimating PDP at the (p ,q)th antenna port on the n_pth OFDM symbol, and $\overline{\boldsymbol{W}} = \boldsymbol{T}^{-1}Dg\left(\boldsymbol{W}_{RLS,p}\boldsymbol{W}_{RLS,p}^H\right)$.

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A.PDP Estimation in Practical MIMO-OFDM Systems

The received sample vector in (10) can be expressed as

$$g_{p,q}[n_p] = Dg(\boldsymbol{h}_{p,q}\boldsymbol{h}_{p,q}^H) + \boldsymbol{\overline{n}}_{p,q} + \boldsymbol{e}_{p,q} \quad (11)$$

where $\boldsymbol{\overline{n}}_{p,q} = \boldsymbol{T}^{-1}Dg(\boldsymbol{W}_{RLS,P}\boldsymbol{n}_q\boldsymbol{n}_q^H\boldsymbol{W}_{RLS,p}^H)$ and $\boldsymbol{e}_{p,q} = 2Re\{\boldsymbol{T}^{-1}Dg(\boldsymbol{W}\boldsymbol{h}_{p,q}\boldsymbol{n}_q^H\boldsymbol{W}_{RLS,p}^H)\}$. We assume that $\boldsymbol{\overline{n}}_{p,q}$ is an effective noise by AWGN. Then, the sample average of $g_{p,q}[n_p]$ is given by

$$\langle g_{p,q}[n_p]\rangle_N \triangleq \frac{1}{N}\sum_{n_p=1}^{|T_p|}\sum_{p=1}^{P}\sum_{q=1}^{Q} g_{p,q}[n_q] \quad (12)$$

$$= \langle Dg(\boldsymbol{h}_{p,q}\boldsymbol{h}_{p,q}^H)\rangle_N + \langle\boldsymbol{\overline{n}}_{p,q}\rangle + \langle\boldsymbol{e}_{p,q}\rangle_N$$

where $N \triangleq |T_p|PQ$ represents the total number of samples for PDP estimation. $|T_p|$ is the number of pilot symbols at the K_{th} subcarrier in a time slot. When N is sufficiently large, the PDP can be perfectly estimated, since $\langle D_g(\boldsymbol{h}_{p,q}\boldsymbol{h}_{p,q}^H)\rangle_N \to \boldsymbol{P}_h, \langle\boldsymbol{\overline{n}}_{p,}\rangle_N \to \sigma_n^2\boldsymbol{\overline{W}}$, and $\langle e_{p,q}\rangle_N \to 0$. However, it is difficult for a receiver of practical MIMO-OFDM systems to obtain such a large number of samples. With an insufficient number of samples, the PDP can be approximated as $\mathbf{P}_h \approx \langle D_g(h_{p,q}h_{p,q}^H)\rangle_N$.

To improve the accuracy of PDP estimation with insufficient samples, we mitigate the effective noise as follows

$$\langle g_{p,q}[n_p]\rangle_N - \sigma_n^2\boldsymbol{\overline{W}} = \langle D_g(\boldsymbol{h}_{p,q}\boldsymbol{h}_{p,q}^H)\rangle_N + \boldsymbol{Z}_N \quad (13)$$

where $\boldsymbol{Z}_N \triangleq \langle\boldsymbol{e}_{p,q}\rangle_N + \langle\boldsymbol{\overline{n}}_{p,q}\rangle_N - \sigma_n^2\boldsymbol{\overline{W}}$ is defined as a residual
noise vector, in which each entry has a zero-mean. Then, the error of PDP estimation with N samples can be calculated as

$$\tilde{e}_N = (\langle D_g(\boldsymbol{h}_{p,q}\boldsymbol{h}_{p,q}^H)\rangle_N - \mathbf{P}_h) + \boldsymbol{Z}_N \quad (14)$$

Since $[\boldsymbol{P}_h]_i \geq 0$ for all i, the PDP can initially be estimated as

$$\boldsymbol{\hat{p}}_{init} = \frac{1}{N}\sum_{n_p=1}^{|T_p|}\sum_{p=1}^{P}\sum_{q=1}^{Q} s_{p,q}[n_q], \quad (15)$$

where $s_{p,q}[n_q]$ is the sample vector of proposed PDP estimator with the l th entry

$$s_{p,q}^l[n_q] = \begin{cases} g_{p,q}^l[n_q] - \sigma_n^2\breve{w}^l & if\ g_{p,q}^l[n_q] > \sigma_n^2\breve{w}^l \\ 0 & otherwise \end{cases}, \quad (16)$$

where $g_{p,q}^l[n_q] = [g_{p,q}[n_p]]_l$ and $\breve{w}^l = [\boldsymbol{\breve{w}}]_l$. To mitigate the
detrimental effect of residual noise $\mathbf{Z}_N$, the proposed scheme estimates the average of residual noise at the zero-taps of $\boldsymbol{P}_h$. At the l th entry of $\boldsymbol{\hat{p}}_{init}$, the zero-tap can be detected as

$$t_z^l = \begin{cases} 1 & if\ \hat{p}_{init}^l < \beta_{th} \\ 0 & otherwise \end{cases}, \quad (17)$$

where $\beta_{th} = \frac{1}{L_g}\sum_{l=0}^{L_g-1}\hat{p}_{init}^l$ is defined as a threshold value for the zero-tap detection. Then, the average of residual noise at the zero-taps can be estimated as

$$\hat{n}_{R,avg} = \frac{1}{N_z}\sum_{l=0}^{L_g-1}\hat{p}_{init}^l\, t_z^l, \quad (18)$$

where $N_z = \sum_{l=0}^{L_g-1} t^l$ represents the total number of detected zero-taps. With the mitigation of residual noise, the l[th] tap of the PDP estimate, , can be expressed as

IV.PERFORMANCE AND COMPLEXITY ANALYSIS

The LMMSE channel estimator with the imperfect PDP in (18) is given by

$$\boldsymbol{W}_{f,p} = \boldsymbol{F}_L D_g(\boldsymbol{\hat{p}}_h)\boldsymbol{F}_p^H\left(\boldsymbol{F}_p D_g(\boldsymbol{\hat{p}}_h)\boldsymbol{F}_p^H + \sigma_n^2\boldsymbol{I}_{K_p}\right)^{-1} \quad (19)$$

where F_L is the $K_d \times L_g$ matrix obtained by taking the first L_g columns of the DFT matrix. $\hat{p}_h = p_h + \bar{e}_{pdp}$ is expressed as the estimated PDP, where the l[th] element of $\bar{e}_{pdp}$ is defined as

$$e_{pdp}^{-l} = \begin{cases} [\tilde{e}_N]_l - \hat{n}_{R,avg} & if\ [\tilde{e}_N]_l > \hat{n}_{R,avg} \\ 0 & otherwise \end{cases} \quad (20)$$

From the matrix inversion lemma, $\left(F_p D_g(\hat{p}_h)F_p^H + \sigma_n^2 I_{K_p}\right)^{-1}$ in (22) is converted as

$$\left(F_p D_g(\hat{p}_h)F_p^H + \sigma_n^2 I_{K_p}\right)^{-1} = A^{-1} - A^{-1}F_p B F_p^H A^{-1} \quad (21)$$

where $A \triangleq \left(F_p D_g(\hat{p}_h)F_p^H + \sigma_n^2 I_{K_p}\right)$ and $B \triangleq D_g(\bar{e}_{pdp})\left(I_{L_g} + F_p^H A^{-1}F_p D_g(\bar{e}_{pdp})\right)^{-1}$. Then, the coefficient matrix for LMMSE channel estimation with $\hat{p}_h$ can be rewritten as

$$\boldsymbol{W}_{f,p} = \boldsymbol{W}_{opt,p} + \boldsymbol{W}_{err,p} \quad (22)$$

where $\boldsymbol{W}_{opt,p} \triangleq F_L D_g(p_h)F_p^H\left(F_p D_g(p_h)F_p^H + \sigma_n^2 I_{K_p}\right)$ is the coefficient matrix for Wiener filtering, and $\boldsymbol{W}_{err,p}$ is given by

$$\boldsymbol{W}_{err,p} = -F_L D_g(p_h)F_p^H A^{-1}F_p B F_p^H A^{-1} + F_L D_g(\bar{e}_{pdp})F_p^H\left(F_p D_g(\hat{p}_h)F_p^H + \sigma_n^2\right)^{-1} \quad (23)$$

The error covariance matrix of LMMSE channel estimation with the imperfect PDP can be obtained as

$$\boldsymbol{E_p} = E\left\{(F_L h_{p,q} - W_{f,p}\hat{h}_{LS,p,q})(F_L h_{p,q} - W_{f,p}\hat{h}_{LS,p,q})^{\boldsymbol{H}}\right\}$$
$$= (F_L - W_{f,p}F_p)D_g(p_h)(F_L - W_{f,p}F_p)^H + \sigma_n^2 W_{f,p}F_p F_p^H W_{f,p}^H \quad (24)$$

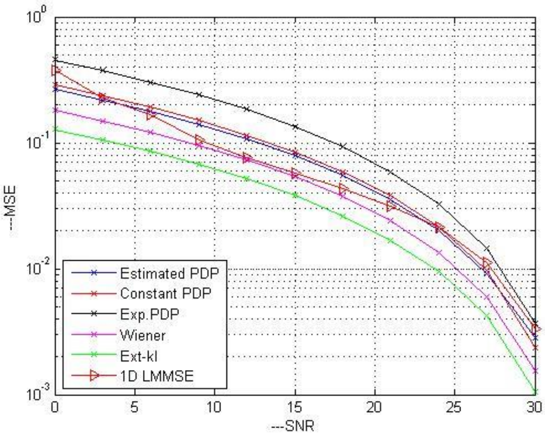

Fig. 3. Performance of LMMSE technique using the estimated PDP over ETU channel.

where $\hat{h}_{LS,p,q} \triangleq diag(X_p)^H y_p[n_p]$. Using the error covariance matrix, the frequency-domain MSE of the proposed scheme is given by

$$MSE_{f,p} = \frac{T_r(\boldsymbol{E}_p)}{T_r(\boldsymbol{F}_L D_g(\boldsymbol{P}_h)\boldsymbol{F}_L^H)} \quad (25)$$

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where $Tr(\boldsymbol{E}_p)$ denotes the trace operation of $\boldsymbol{E}_p$. With a sufficiently large number of samples, $\bar{e}_{pdp} \rightarrow 0$. Thus, the MSE of the proposed scheme achieves that of Wiener filtering
because $\boldsymbol{W}_{f,p} \rightarrow \boldsymbol{W}_{opt,p}$.

The additional complexity by the proposed PDP estimation technique is $O(L_g^3 + K_p L_g^2 + |T_p|PQL_g)$, which mainly comes from computing (2) and (6). When the pilot spacing is fixed in the frequency domain, all entries of $\boldsymbol{F}_p$ and $\mathbf{T}$ are constant. Thus, $\left(\boldsymbol{F}_p^H \boldsymbol{F}_p + \epsilon\, \boldsymbol{I}_{L_g}\right)^{-1} \boldsymbol{F}_p^H$ and $\boldsymbol{T}^{-1}$ can be computed only once, and their values can be stored. The additional complexity is then reduced to $O(L_g^2 + |T_p|PQL_g)$.

V. SIMULATION RESULTS

We consider a MIMO-OFDMsystem with the physical layer parameters for the downlink of 3GPP LTE [14]. The system bandwidth is 5 MHz with 301 subcarriers for transmitting data information and pilots including a DC subcarrier at 2 GHz carrier frequency. The width of each subcarrier is 15 kHz with an FFT size of 512. The MIMO-OFDM system utilizes four transmit and two receive antennas (P=4,Q=2). We assume that the pilots of the four transmit antenna ports are distributed as the time and frequency grid of the LTE system in Fig. 2. The length of CP is 40 (Lg=40). For all simulations, the channel estimator is based on a cascaded 2×1D LMMSE technique during 14 OFDM symbols ($|\mathcal{T}1| = |\mathcal{T}2| = 2$, $|\mathcal{T}3| = |\mathcal{T}4| = 1$), as shown in Fig. 2, where the filtering in frequency domain is followed by the filtering in time domain over slowly fading channels with the Doppler frequency of 5 Hz.

Figure 3 shows the MSE performance of the 2×1D LMMSE technique using the estimated PDP. All underlying links are modeled as extended typical urban (ETU) channels [14]. The performance of the 2×1D Wiener filter with exact PDP is included as a lower bound. For performance comparisons, we plot the performance of frequency domain regularized LS channel estimation in which the PDP information is not required. The performance of the 2×1D LMMSE technique using the approximated PDP, which is uniform or exponential model with the channel delay parameter estimation in [11], is also plotted. Note that the LMMSE technique using the estimated PDP outperforms the conventional methods, since the correlation mismatch is reduced by the proposed PDP estimation. We also observe from Fig. 4 that the proposed method has a performance loss within only a 2.4-dB gap, compared with 2×1D Wiener filtering.

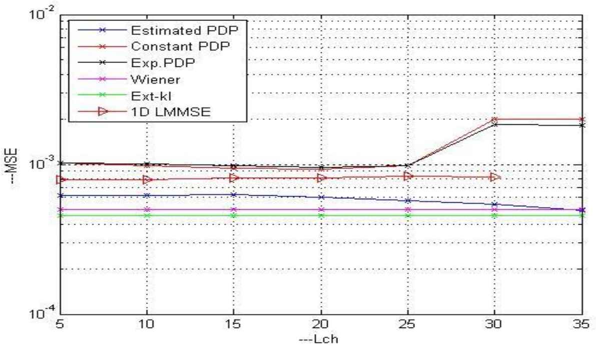

Fig. 4. Performance of LMMSE technique using the estimated PDP over 6- ray exponential channel with variable channel maximum delays (Pilot SNR= 30 dB).

In Fig. 4, we investigate the MSE performance of the proposed scheme over the exponentially power decaying six-path Rayleigh fading channel model, where the channel maximum delay, Lch, is variable. The PDP of the channel model is defined as $E\left\{\left|\alpha_{p,q}^{l}[n]\right|^{2}\right\} = \frac{1}{S_h} e^{l/\tau_{rms}}$ for l=0,Δτ,...,5Δτ and $\Delta = \frac{L_{ch}}{5}$.Here, $\tau_{rms} = \frac{L_{ch}}{log(2L_{ch})}$and S_h is the normalization factor $\left(S_h = \sum_l e^{-l/\tau_{rms}}\right)$. The performance of the proposed scheme is better than that of the conventional methods, and approaches that of Wiener filtering in various channel environments.

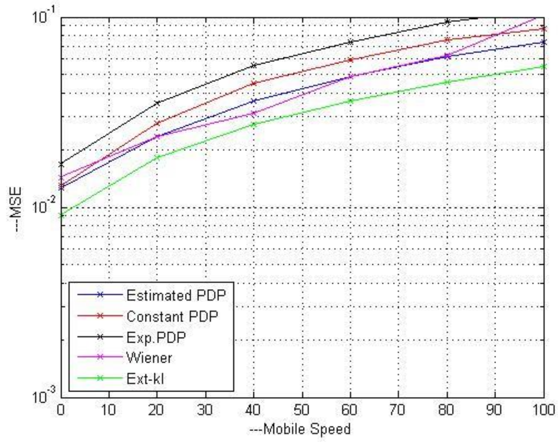

Fig. 5. Performance of LMMSE technique using the estimated PDP over ETU channel with different mobile equipment speeds (Pilot SNR= 30 dB and Doppler frequency = 9.26 − 203.7 Hz).

Figure 5 shows the MSE performance of the $2 \times 1D$ LMMSE technique using the estimated PDP for different mobile equipment speeds at 30-dB SNR. All underlying links are modelled as ETU channels. In Fig. 5, it can be seen that the MSE of LMMSE technique using the estimated PDP achieves that of Wiener filtering even at high Doppler frequencies.

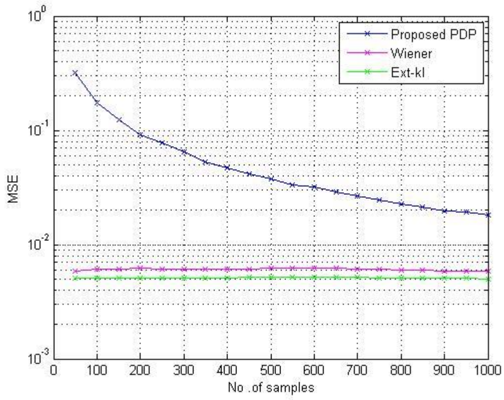

Fig. 5. Simulation and analysis results of LMMSE channel estimation over ETU channel with various number of samples for the PDP estimation (Pilot SNR= 20 dB).

Figure 6 shows simulation and analysis results of the frequency-domain LMMSE channel estimation with various samples for obtaining the PDP at 20-dB SNR ($N = |\mathcal{T}p|PQ$). We assume that 2×2 MIMO-OFDM system over ETU channels with 70-Hz Doppler frequency.

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The simulation results correspond to the channel estimation performance at the first OFDM symbol of antenna port 1 shown in Fig. 2. We obtain the analytic results in (24) by using the coefficient matrix for LMMSE channel estimation with the perfect or imperfect PDP at the antenna port. In Fig. 6, it is observed that the MSE of the proposed scheme improves the MSE performance with an increase in the number of samples for PDP estimation.

VI. CONCLUSION

We proposed a PDP estimation technique for the LMMSE channel estimator in MIMO-OFDM systems. The CIR estimates at each path of the MIMO channels were used to obtain the PDP. For accurate PDP estimation, we considered the spectral leakage effect from virtual subcarriers, and the resid-ual noise caused by the insufficient number of estimated CIR samples. The proposed technique effectively mitigates both the spectrum leakage and residual noise. Simulation results show that the performance of LMMSE channel estimation using the proposed PDP estimate approaches that of Wiener filtering and extended kalman filtering due to the mitigation of distortion effects.

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