ESTIMATED POWER DELAY PROFILE WITH WIENER FILTER ALGORITHM FOR MIMO-OFDM SYSTEMS

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ABSTRACT

This letter proposes a power delay profile (PDP) estimation technique for linear minimum mean square error (LMMSE) channel estimator of multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems. For practical applications, only the pilot symbols of all transmit antenna ports are used in estimating the PDP. 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.

I.INTRODUCTION

MULTIPLE-INPUT multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) is one of the most promising techniques for wireless communication systems, including the 3rd Generation Partnership Project Long Term Evolution (3GPP LTE) [1], [2] and IEEE 802.16 (WiMAX). MIMO-OFDM provides a considerable performance gain over broadband single-antenna systems by obtaining the spatial diversity or multiplexing gain [3], [4]. Most receiver techniques of MIMO-OFDM systems are designed with the assumption that channel state information (CSI) is available, in order to achieve the maximum diversity or multiplexing gain. The performance gain depends heavily on accurate channel estimation, which is crucial for the MIMO-OFDM systems. 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 [8]. To obtain the frequency domain channel statistics at the receiver, power delay profile (PDP) estimation schemes have been proposed [9], [10]. 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 for improving the performance of LMMSE channel estimation employs an approximated PDP (i.e., uniform or exponential model) with the estimation of second-order channel statistics, which are mean delay and root-mean-square (RMS) delay spread [11]. The channel delay parameters are estimated using pilots with low computational Manuscript received September 29, 2011. The associate editor coordinating the review of this letter and approving it for publication was M. Tao. This work was supported by the National Research Foundation of Korea (NRF) grant funded by the Korea government (MEST) (No. 2011-0013807). The authors

are with the Department of Electrical Engineering, Pohang University of Science and Technology (POSTECH), Pohang 790-784, therefore, the LMMSE channel estimator with the approximated PDP is appropriate for practical applications such as a WiMAX system. However, the performance degradation is caused by both the correlation mismatch and the estimation error of delay parameters. To reduce the mismatch in the frequency domain, we propose a PDP estimation technique for the LMMSE channel estimator of MIMO-OFDM systems. For practical applications, the proposed technique uses only the pilot symbols of all transmit antenna ports to estimate the PDP with low computational complexity. In addition, the proposed technique effectively mitigates the distortion effects, incurred by null subcarriers and an insufficient number of estimated channel impulse response (CIR) samples. Simulation results show that the performance of LMMSE channel estimation with the proposed PDP estimate approaches that of Wiener filtering.

II. SYSTEM MODEL

The system under consideration is a MIMO-OFDM system with P transmits and Q receive antennas, and K total subcarriers. Suppose that the MIMO-OFDM system transmits Kd subcarriers at the central spectrum assigned for data and pilots with $K-Kd$ 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 $C[kp, np]$ be the pilot subcarrier for the $p$pth transmit antenna at the np th OFDM symbol, which is a QPSK modulated signal from known sequences between the transmitter and receiver. We assume that the pilot subcarriers are distributed over a time and frequency grid as in Fig. 1, to preserve the Orthogonality of pilots among different transmits antennas. kp And np $\mathcal{T}p$ represent the index sets for the pilot subcarriers of the $p$pth antenna port in the frequency and time domains, respectively. At the npth OFDM symbol, the number of pilot subcarriers is defined as Kp=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, Lg, is longer than the channel maximum delay, Lch , making the channel matrix circulate ($Lch \leq Lg$). At the receiver, after perfect synchronization, the removal of CP, and FFT operation, the received pilot symbol for the qth receive antenna can be represented as

$$\mathrm{y}_q[n_p] = diag(\mathrm{x}_p)\mathrm{F}_p\mathrm{h}_{p,q} + \mathrm{n}_q, \qquad (1)$$

Where $\mathbf{h}_{p,q}= [h_{p,q}[n_p,0], h_{p,q}[n_p,1], \ldots, h_{p,q}[n_p,L_{ch}], 0, \ldots, 0]^T$ is an Lg×1 CIR vector at the $p$pth transmit antenna and qth receive antenna. $(\cdot\ )T$ and $(\cdot\ )H$ represent the transpose operation, and the transpose and conjugate open[ration of a vector or matrix, respectively. $\mathbf{x}p$ = $Cp[i1, np]$, $[C_p[i_1,n_p], C_p[i_2,n_p], \ldots, C_p[i_{K_p},n_p]]^T$ denotes a pilot vector at the npth OFDM symbol for $i_k \in \mathcal{F}_p$ and k=1, 2, . . .,Kp. $dia(\mathbf{x}p)$ is the Kp×Kp diagonal matrix whose entries are the Kp Elements of the vector $\mathbf{x}p$. $\mathbf{F}p$ is a Kp×Lg matrix with the (ik, l)pth entry $[\mathbf{F}_p]_{i_k,l} = 1/\sqrt{K} exp\{-j2\pi i_k l/K\}$ $i_k \in \mathcal{F}_p$ and $l=0,1,\ldots,L_g-1$. $\mathbf{n}_q$

. **n**q is a complex additive white Gaussian noise (AWGN) vector at the $q$pth receiver antenna with each entry having a zero-mean and variance of $\sigma 2\ n$.

III.PROPOSED METHOD FOR THE PDP ESTIMATION:-

From (1), the CIR at the (p, q)pth antenna port can be estimated approximately using the regularized least squares (RLS) channel estimation with a fixed length of Lg as

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

Where ϵ=0.001 is a small regularization parameter, and **I**Lg is the Lg×Lg identity matrix. **F**Hp
Fp 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 ˆ**h**R,,qˆ**h**H R,p,q is given by

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

Where $\mathbf{R}_{hh}=E\{\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\}$ and $\mathbf{W}=(\mathbf{F}_p^H\mathbf{F}_p+\epsilon\mathbf{I}_{L_g})^{-1}\mathbf{F}_p^H\mathbf{F}_p$. Note that the diagonal elements of the channel covariance matrix, **R**hh, represent the PDP of multipath channel within the length of Lg, and all off-diagonal elements are zeros. Hence, the covariance matrix can be expressed as **R**hh=$diag$ (**ph**), where **ph**=[p0, p1, . . . , pLch, 0, . . . , 0]T and pl=E{hp,q|[np, l]|2}. Unfortunately, **R**hh is distorted by **W**, which is an ill-conditioned matrix due to the presence of **F**H **F**p. Thus, instead of calculating **W**−1, we investigate the method for eliminating the spectral leakage of **W**. The covariance matrix of the estimated CIR is defined as **R**hhˆ =WRhhWh which can be expressed as

$$\mathbf{R}_{\hat{h}\hat{h}} = \sum_{l=0}^{L_g-1}\mathbf{W}diag(p_l\mathbf{u}_l)\mathbf{W}^H, \qquad (4)$$

Where u l is a unit vector with the $l$pth entry being one and otherwise zeros. Let **p**h and **t**l be the Lg×1 vectors defined as **p**h=D(**Rh**ˆhˆ) and **t**l=Dg (**W**$diag$(**u**l)**W**H) , respectively, where Dg(**A**) is the column vector containing all the diagonal elements of **A**. Then, the relation in (4) is simplified as

$$\mathbf{p}_{\hat{h}} = p_0\mathbf{t}_0 + p_1\mathbf{t}_1 + ... + p_{L_g-1}\mathbf{t}_{L_g-1} \triangleq \mathbf{T}\mathbf{p}_h, \qquad (5)$$

Where $\mathbf{T}$=[$\mathbf{t}0$, $\mathbf{t}1$, ..., $\mathbf{t}Lg-1$] is defined as a distortion matrix by $\mathbf{W}$. It is noted that the distortion matrix is a strictly diagonally dominant matrix, satisfying [$\mathbf{T}$]i| > Σ $j\neq i$ [$\mathbf{T}$]ij | for all i, j, since the non-diagonal elements of $\mathbf{T}$ are composed of the leakage powers of $\mathbf{u}i$ for all i. From the Gershgorin circle theorem, a strictly diagonally dominant matrix is non-singular. In addition, the distortion matrix is a well-conditioned matrix. Hence, the distortion of $\mathbf{W}$ can be eliminated as

$$\mathbf{p}_h = \mathbf{T}^{-1}\mathbf{p}_{\hat{h}} = E\left\{\mathbf{g}_{p,q}[n_p]\right\} - \sigma_n^2 \tilde{\mathbf{w}}, \qquad (6)$$

where $\mathbf{g}p,q[np]$=$\mathbf{T}-1Dg$ (ˆ$\mathbf{h}$ R,p,q ˆ$\mathbf{h}H$ R,p,q) is defined as the received sample vector for estimating PDP at the (p, q)pth antenna port on the nppth OFDM symbol, and ˜$\mathbf{w}$ = $\mathbf{T}-1Dg$($\mathbf{W}RLS,p\mathbf{W}H$ RLS,p).

$$\mathbf{g}_{p,q}[n_p] = Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right) + \tilde{\mathbf{n}}_{p,q} + \mathbf{e}_{p,q}, \qquad (7)$$

where ˜$\mathbf{n}p,q$ = $\mathbf{T}-1Dg$($\mathbf{W}RLS,p\mathbf{n}q\mathbf{n}Hq$ $\mathbf{W}H$ RLS,p) and $\mathbf{e}p,q$ = $2Re${$\mathbf{T}-1Dg$($\mathbf{W}\mathbf{h}p,q\mathbf{n}Hq$ $\mathbf{W}H$ RLS,p)}. Here, R{$\mathbf{a}$} denotes the real part of $\mathbf{a}$. We assume that ˜$\mathbf{n}p$, is an effective noise by AWGN. Then, the sample average of $\mathbf{g}p$,[np] is given by

$$\begin{aligned}\left\langle \mathbf{g}_{p,q}[n_p]\right\rangle_N &\triangleq \frac{1}{N}\sum_{n_p=1}^{|\mathcal{T}_p|}\sum_{p=1}^{P}\sum_{q=1}^{Q}\mathbf{g}_{p,q}[n_p] \qquad (8)\\ &= \left\langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\right\rangle_N + \left\langle\tilde{\mathbf{n}}_{p,q}\right\rangle_N + \left\langle\mathbf{e}_{p,q}\right\rangle_N,\end{aligned}$$

Where N = |Tp| PQ represents the total number of samples for PDP estimation. $\mathcal{T}$ is the number of pilot symbols at the kppth subcarrier in a time slot. When N is sufficiently large, the PDP can be perfectly estimated, since⟨ D($\mathbf{h}p,q\mathbf{h}Hp,q$)⟩ N→ $\mathbf{p}$h,(˜$\mathbf{n}p,q$)N→ $\sigma 2n$˜$\mathbf{w}$, and ($\mathbf{e}p,q$)N→ $\mathbf{0}$. 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 (≈Dg($\mathbf{h}p,q\mathbf{h}H$ p,q)) N.

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

$$\left\langle \mathbf{g}_{p,q}[n_p]\right\rangle_N - \sigma_n^2\tilde{\mathbf{w}} = \left\langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\right\rangle_N + \mathbf{z}_N, \qquad (9)$$

Where $\mathbf{z}N$ = ($\mathbf{e}p$,)N +(˜$\mathbf{n}p,q$)$N-\sigma 2n$ ˜$\mathbf{w}$ 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

$$\tilde{\mathbf{e}}_N = \left(\left\langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\right\rangle_N - \mathbf{p}_h\right) + \mathbf{z}_N. \qquad (10)$$

Since [**ph**] ≥ 0 for all i, the PDP can initially be estimated as

$$\hat{\mathbf{p}}_{init} = \frac{1}{N}\sum_{n_p=1}^{|\mathcal{T}_p|}\sum_{p=1}^{P}\sum_{q=1}^{Q}\mathbf{s}_{p,q}[n_p], \qquad (11)$$

where **s**p,q[np] is the sample vector of proposed PDP estimator with the $l$pth entry

$$s^l_{p,q}[n_p] = \begin{cases} g^l_{p,q}[n_p] - \sigma^2_n \tilde{w}^l & \text{if } g^l_{p,q}[n_p] > \sigma^2_n \tilde{w}^l \\ 0 & \text{otherwise} \end{cases}, \qquad (12)$$

Where gl p,[np]= [**g**p,q[np]]l and ˜ wl=[˜**w**]l. To mitigate the detrimental effect of residual noise **z**N, the proposed scheme estimates the average of residual noise at the zero-taps of **p**h. At the $l$pth entry of ˆ**p**$init$, the zero-tap can be detected as

$$t^l_z = \begin{cases} 1 & \text{if } \hat{p}^l_{init} < \beta_{th} \\ 0 & \text{otherwise} \end{cases}, \qquad (13)$$

Where βth= 1 Lg Σ Lg−1 l=0 ˆpl $init$ 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

$$\hat{n}_{R,avg} = \frac{1}{N_z}\sum_{l=0}^{L_g-1}\hat{p}^l_{init}t^l_z, \qquad (14)$$

The above equation represents the total number of detected zero-taps. With the mitigation of residual noise, the lth tap of the PDP estimate, ˆ**p**h, can be expressed as

$$\hat{p}^l_h = \begin{cases} \hat{p}^l_{init} - \hat{n}_{R,avg} & \text{if } \hat{p}^l_{init} > \hat{n}_{R,avg} \\ 0 & \text{otherwise} \end{cases}. \qquad (15)$$

Then, the estimated PDP in (15) can be used to obtain the frequency-domain channel correlation in the LMMSE channel estimator.

$$\mathbf{W}_{f,p} = \mathbf{F}_L Dg(\hat{\mathbf{p}}_h)\mathbf{F}^H_p\left(\mathbf{F}_p Dg(\hat{\mathbf{p}}_h)\mathbf{F}^H_p + \sigma^2_n \mathbf{I}_{K_p}\right)^{-1}, \quad (16)$$

Where **F**L is the Kd × Lg matrix obtained by taking the first Lg columns of the DFT matrix. ˆ**p**h = **p**h + ¯**ePDP** is expressed as the estimated PDP, where the Lpth element of ¯**e**PDP is defined as

$$\bar{e}^{l}_{pdp} = \begin{cases} [\tilde{\mathrm{e}}_N]_l - \hat{n}_{R,avg} & \text{if } [\tilde{\mathrm{e}}_N]_l > \hat{n}_{R,avg} \\ 0 & \text{otherwise} \end{cases} . \quad (17)$$

From the matrix inversion lemma, (**FpDg**(ˆ**p**h)**F**pH + $\sigma 2n$**I**Kp)−1 in is converted as

$$\left(\mathbf{F}_p Dg(\hat{\mathrm{p}}_h)\mathbf{F}_p^H + \sigma_n^2 \mathbf{I}_{K_p}\right)^{-1} = \mathbf{A}^{-1} - \mathbf{A}^{-1}\mathbf{F}_p \mathbf{B} \mathbf{F}_p^H \mathbf{A}^{-1}, \ (18)$$

where **A** ≜(**F**pDg(**p**h)**F**Hp+ $\sigma 2n$**I**Kp)and **B** ≜Dg(¯**e**pdp)(**I**Lg+**F**Hp**A**−1**F**pDg(¯**e**pdp))−1 Then, the coefficient matrix for LMMSE channel estimation with ˆ**p**h can be rewritten as

$$\mathbf{W}_{f,p} = \mathbf{W}_{opt,p} + \mathbf{W}_{err,p}, \qquad (19)$$

Where **W**opt, ≜ **F**LDg(**p**h)**F**Hp(**F**pDg(**p**h)**F**Hp + $\sigma 2n$**I**Kp)−1is the coefficient matrix for Wiener filtering and **W**err,p is given by

$$\begin{aligned} \mathbf{W}_{err,p} &= -\mathbf{F}_L Dg(\mathrm{p}_h)\mathbf{F}_p^H \mathbf{A}^{-1}\mathbf{F}_p \mathbf{B}\mathbf{F}_p^H \mathbf{A}^{-1} \qquad (20) \\ &+ \mathbf{F}_L Dg(\bar{\mathrm{e}}_{pdp})\mathbf{F}_p^H \left(\mathbf{F}_p Dg(\hat{\mathrm{p}}_h)\mathbf{F}_p^H + \sigma_n^2 \mathbf{I}\right)^{-1}. \end{aligned}$$

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

$$\begin{aligned} \mathbf{E}_p &= E\left\{\left(\mathbf{F}_L \mathbf{h}_{p,q} - \mathbf{W}_{f,p}\hat{\mathbf{h}}_{LS,p,q}\right)\left(\mathbf{F}_L \mathbf{h}_{p,q} - \mathbf{W}_{f,p}\hat{\mathbf{h}}_{LS,p,q}\right)^H\right\} \\ &= \left(\mathbf{F}_L - \mathbf{W}_{f,p}\mathbf{F}_p\right) Dg(\mathbf{p}_h)\left(\mathbf{F}_L - \mathbf{W}_{f,p}\mathbf{F}_p\right)^H \\ &+ \sigma_n^2 \mathbf{W}_{f,p}\mathbf{F}_p\mathbf{F}_p^H \mathbf{W}_{f,p}^H, \qquad (21) \end{aligned}$$

where ˆ**h**LS,p,q ≜ $diag$(**x**p)H**y**p[np]. Using the error covariance matrix, the frequency-domain MSE of the proposed scheme is given by

$$MSE_{f,p} = \frac{Tr(\mathbf{E}_p)}{Tr(\mathbf{F}_L Dg(\mathbf{p}_h)\mathbf{F}_L^H)}, \qquad (22)$$

where Tr(**E**p) denotes the trace operation of **E**p. With a sufficiently large number of samples, ¯**e**pdp → **0**. Thus, the MSE of the proposed scheme achieves that of Wiener filtering because **W**f, →**W**opt,p.

The additional complexity by the proposed PDP estimation technique is $O\ (L3g + KpL2g + |\mathcal{T}p|PQL\ )$, which mainly comes from computing (2) and (6). When the pilot spacing is fixed in the frequency domain, all entries of $\mathbf{F}p$ and $\mathbf{T}$ are constant. Thus, $(\mathbf{F}p\mathbf{F}Hp + \epsilon\mathbf{I}Lg\ )-1\mathbf{F}Hp$ and $\mathbf{T}-1$ can be computed only once, and their values can be stored. The additional complexity is then reduced to $O\ (L2g + |\mathcal{T}p|PQL\ )$

IV.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,=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. 1. 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,as shown in Fig. 1, 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 2 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

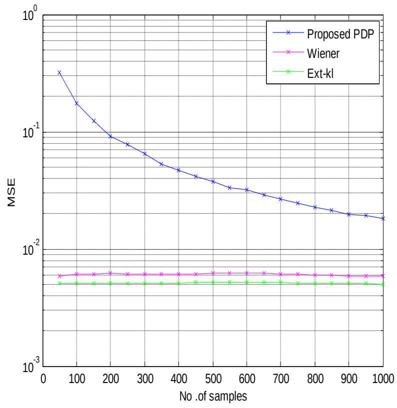

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Figure-1 Simulation and analysis results of LMMSE channel estimation over ETU channel with various number of samples for the PDP estimation

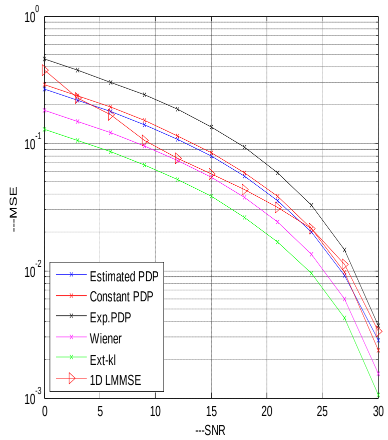

Figure 2 hows the different previous techniques estimation with proposed algorithm using pilot signal to noise ratio (db) to mean square error. It shows estimated result approach approaches to wiener filtering.

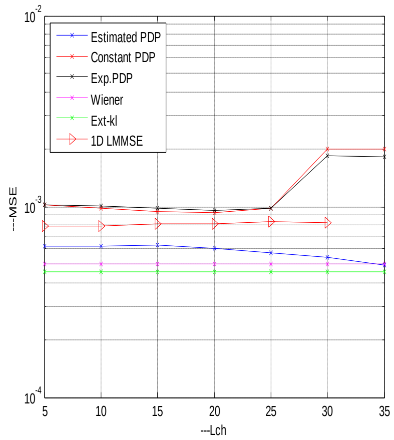

Figure -3 Performance of LMMSE technique using the estimated PDP over 6-ray exponential channel with variable channel maximum delays (Pilot SNR= 30 dB).Here channel length delay Lch is estimated using mean square error.

V.CONCLUSIONS

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 residual 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.

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