A PDP Estimation Technique For The LMMSE Channel Estimator Of MIMO- OFDM Systems

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Abstract

The combination of multiple-input multiple-output (MIMO) system with OFDM techniques is regarded as one of the promising solution for wireless communication systems, including the 3rd Generation Partnership Project Long Term Evolution (3GPP-LTE) and IEEE 802.16 (Wi-MAX). In this paper, we propose a power delay profile (PDP) estimation technique for linear minimum mean square error (LMMSE) channel estimator of MIMO-OFDM systems. For practical applications, the proposed technique uses only the pilot symbols of all the transmit antenna ports to estimate the PDP with low computational complexity. In addition, this technique also reduces the correlation mismatch in the frequency domain, the distortions caused by null subcarriers and an insufficient no. of samples for PDP estimation. Simulation results show that the performance of LMMSE channel estimation using the Proposed PDP estimate is much better than the ML PDP estimator and an approximated PDP estimator(uniform or exponential).

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

1. Introduction

Orthogonal frequency division multiplexing (OFDM) is one of the most promising layer technologies for high data rate wireless communications due to its robustness to frequency selective fading, high spectral efficiency and low computational complexity. OFDM can be used in conjunction with a MIMO transreceiver to increase the diversity gain and/or the system capacity by exploiting spatial domain. Because the OFDM system effectively provides numerous parallel narrow band channels, MIMO-OFDM is considered as a key technology in emerging high data rate system such as 3G, IEEE 802.16 and IEEE 802.11.

The performance gain of MIMO-OFDM systems depends highly on accurate channel estimation. The block diagram of MIMO-OFDM system using pilot symbols is as shown in fig.1

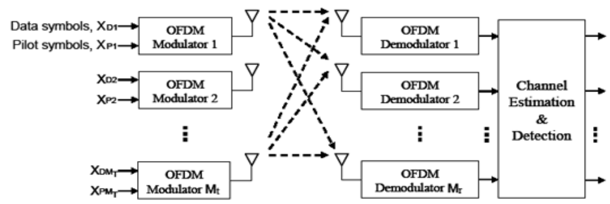

Fig 1: MIMO-OFDM system using Pilot symbols

Whenever the receiver knows the channel statistics, the pilot- aided channel estimation, based on the linear minimum mean square error (LMMSE) technique is the optimum estimation technique to minimize the mean square error (MSE). Hence power delay profile estimation schemes based on ML PDP estimators, approximated PDP estimators etc. have been proposed to obtain the frequency domain channel statistics at the receiver. The first method uses the advantage of cyclic prefix (CP) segment of OFDM symbols. But, the major drawback in this method is, it requires very high computational complexity for obtaining an accurate PDP estimation.

The second method, based on PDP estimation is an approximated PDP (i.e., either uniform or exponential model) which uses second order channel statistics like mean delay and root mean square (RMS) delay spread. The estimation of delay parameters uses pilot symbols providing the low computational complexity. Hence, the LMMSE channel estimator with the approximated PDP is suitable for real time applications like Wi-MAX system. But still this method degrades the performance of the system due to correlation mismatch and the estimation error of the channel delay parameters.

Hence, we propose the method called a PDP estimation technique for the LMMSE channel estimation (i.e., wiener filtering) of MIMO-OFDM systems, which minimizes the correlation mismatch in frequency domain. This improved technique uses only the pilot symbols of all transmit antenna ports to estimate the PDP with low computational complexity.

In addition to this, the proposed technique effectively reduces the distortions caused by null subcarriers and an insufficient no. of CIR (Channel impulse response) samples. Thus the simulation results shows that the performance of LMMSE channel estimation using the proposed PDP estimator is much better than the other PDP estimation schemes like MLPDP estimator and an approximated PDP (Uniform or Exponential) estimator.

2. System Model

MIMO-OFDM system with P transmit and Q receive antennas and K, total sub carriers are considering here. Now let us assume that, in order to control interferences with other systems, MIMO-OFDM system transmits K_d subcarriers for data and pilots with K-K_d virtual subcarriers. In MIMO systems, the channel impulse response (CIR) corresponding to different transmit and receive antennas may have the same PDP.

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.

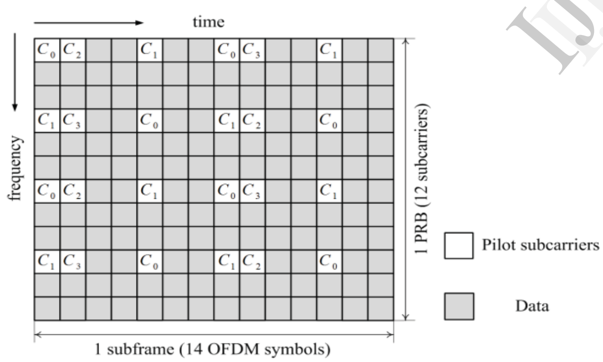

Fig 2: Pilot Symbol Arrangement in a physical resource block (PRB) of the LTE OFDM system.

Let $k_p \epsilon F_p$ represents the index set in frequency domain and $n_p \epsilon T_p$ represents the index set in time domain for the pilot subcarriers of the p^{th} transmit antenna port. At the n_p^{th} OFDM symbol, the number of pilot subcarriers may be defined as $K_p = |F_p|$. Prior to the transmission of pilot inserted OFDM symbol over the wireless channel, an inverse fast fourier transform (IFFT) and addition of a cyclic prefix (CP) has to be performed. The use of CP in MIMO-OFDM system prevents the inter symbol interference (ISI) and also inter carrier interference (ICI).

Let us assume that the length of C_p, L_g is larger than the delay spread of the channel, L_{ch}, making the channel vector circulant (i.e., $L_g \geq L_{ch}$). After perfect synchronization at the receiver, the removal of CP and FFT operation has been performed. The received pilot symbol for the q^{th} receive antenna can be expressed as

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

Where x_p is a pilot vector at the n_p^{th} OFDM symbol for $i_k \epsilon F_p$ and k= 1,2,.....K_p and is given as $x_p = [C_p(i_1,n_p),\ C_p(i_2,n_p)........C_p(i_{Kp},n_p)]^T$. $diag(x_p)$ is the $K_p \times K_p$ diagonal matrix whose entries are K_p elements of the vector x_p, $\mathbf{F}_p$ is a $K_p \times L_g$ matrix with the $(i_k, l)^{th}$ entry and is given as $[\mathbf{F}_p]_{ik,l} = 1/\sqrt{K} \exp\{-j2\pi\ i_k l/K\}$ where $i_k \epsilon F_p$ and l= 0,1,...., L_g-1, $h_{p,q}$ is an $L_g \times 1$ 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. 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^2.

3. Proposed Method for PDP estimation

A. Derivation of the PDP in MIMO-OFDM systems

The CIR estimates at the $(p,q)^{th}$ antenna port can be estimated approximately from [1] using the regularized least squares (RLS) channel estimation with a fized length of L_g as

$$\hat{h}_{R,p,q} = (F^H_p F_p + \epsilon I_{Lg})^{-1} F^H_p diag(X_p)^H y_q[n_p] \triangleq W_{RLS,p} y_q[n_p] \quad [2]$$

Where ϵ = 0.001, which is a small regularization parameter, $(.)^H$ represents the transpose and conjugate operation of a vector or matrix and I_{Lg} is L_g x L_g identity matrix. $F_p^H F_p$ in equation 2 is ill- conditioned due to the sparsity of pilot tones in the frequency domain and the presence of virtual subcarriers. From the estimated CIR in (2), the PDP can be derived and the ensemble average of $\hat{h}_{R,p,q}\ \hat{h}^H_{R,p,q}$ is given by

$$E\left\{\hat{\mathrm{h}}_{R,p,q}\hat{\mathrm{h}}_{R,p,q}^{H}\right\} = \mathbf{W}\mathbf{R}_{hh}\mathbf{W}^{H} + \sigma_n^2\mathbf{W}_{RLS,p}\mathbf{W}_{RLS,p}^{H}, \quad [3]$$

Where $\mathbf{R}_{hh} = E\{\mathbf{h}_{p,q}\mathbf{h}^H_{p,q}\}$ and $\mathbf{W} = (\mathbf{F}_p^H\, \mathbf{F}_p + \epsilon\mathbf{I}_{Lg})^{-1}\, \mathbf{F}_p^H\, \mathbf{F}_p$. The PDP of multi path channel within the length of L_g, is represented by $\mathbf{R}_{hh}$, which is the diagonal elements of the channel covariance matrix and is expressed as $\mathbf{R}_{hh} = diag(\mathbf{p}_h)$, where $\mathbf{p}_h = [\mathbf{p}_0\ ,\ \mathbf{p}_1\ .....,\ \mathbf{p}_{Lch},\ 0,\ ....,\ 0]^T$ and

$\mathbf{P}_l$ = E{| $\mathbf{h}_{p,q}[\mathbf{n}_p,l|^2$}. And all off–diagonal elements are zeros. Unfortunately, the covariance matrix ($\mathbf{R}_{hh}$) is distorted by an ill-conditioned matrix (**W**) due to the presence of $\mathbf{F}_p^H\ \mathbf{F}_p$. Thus, instead of calculating we investigate the method for eliminating the spectral leakage of **W**.

The covariance matrix of the estimated CIR is defined as

$$\mathbf{R}_{\hat{h}\hat{h}}=\mathbf{W}\mathbf{R}_{hh}\mathbf{W}^H$$

which can be expressed as

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

[4]

Where $\mathbf{u}_l$ is a unit vector with the l^{th} entry being one and otherwise zeros. Let $p_{\hat{h}}$ and t_l be the $L_g \times 1$ vectors defined as $\mathbf{p}_{\hat{h}}$ = $\mathbf{Dg}(\mathbf{R}_{\hat{h}\hat{h}})$ and $\mathbf{t}_l$ = $\mathbf{Dg}(\mathbf{W}$ $diag(\mathrm{u}_l)$ $\mathbf{W}^H)$, respectively, where $\mathbf{Dg}(\mathbf{A})$ is the column vector containing all the diagonal elements of **A**. Then, the relation in (4) is simplified as

$$\mathbf{p}_{\hat{h}} = \mathbf{p}_0\mathbf{t}_0 + \mathbf{p}_1\mathbf{t}_1 + ...+ \mathbf{p}_{Lg-1}\mathbf{t}_{Lg-1} \triangleq \mathbf{T}\mathrm{p}_h$$

[5]

where $\mathbf{T}$= [$\mathbf{t}_0$, $\mathbf{t}_1$,....$\mathbf{t}_{Lg-1}$] is defined as a distortion matrix by **W**. It is noted that the distortion matrix is a strictly diagonally dominant matrix, satisfying

$$|[\mathbf{T}]_{ii}| > \sum_{j\neq i} |[\mathbf{T}]_{ii}| \text{for all i ,j,}$$

since 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 in addition, the distortion matrix is a well conditioned matrix. Hence, the distortion of W can be eliminated as

$$p_h = T^{-1}p_{\hat{h}} = E\{g_{p,q}[n_p]\} - \sigma_n^2\tilde{w}$$

[6]

where $g_{p,q}[n_p] = T^{-1}Dg(\hat{h}_{R,p,q}\ \hat{h}^H_{R,p,q})$ is defined as the received sample vector for estimating PDP at the (p,q)th antenna port on the npth OFDM symbol and $\tilde{w}$ equal to T^{-1} $Dg(W_{RLS,p}W^H_{RLS,p})$.

B. PDP estimation in practical MIMO-OFDM systems.

The received sample vector in [6] can be expressed as

$$g_{p,q}[n_p] = Dg(h_{p,q}h^H_{p,q}) + \tilde{n}_{p,q} + e_{p,q}$$

[7]

From the above equation, we assume that $\tilde{n}_{p,q}$ is an effective noise by AWGN and is given as

$\tilde{n}_{p,q} = T^{-1}Dg(W_{RLS,p}n_qn^H_qW^H_{RLS,p})$ and

$e_{p,q} = 2Re\{\ T^{-1}Dg(W_{RLS,p}n_qn^H_qW^H_{RLS,p})\}$.

Here, Re{a} denotes the real part of a. Then the sample average of $g_{p,q}[n_p]$ is given by

$$\langle \mathbf{g}_{p,q}[n_p]\rangle_N \triangleq \frac{1}{N}\sum_{n_p=1}^{|\mathcal{T}_p|}\sum_{p=1}^{P}\sum_{q=1}^{Q}\mathbf{g}_{p,q}[n_p]$$

[8]

$$= \langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\rangle_N + \langle\tilde{\mathbf{n}}_{p,q}\rangle_N + \langle\mathbf{e}_{p,q}\rangle_N$$

Where $N \triangleq |T_p|P,Q$ represents the total no. of samples for PDP estimation. $|T_p|$ is the no. of pilot symbols at the k_p^{th} subcarrier in a time slot.

Since,

$\langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\rangle_N \rightarrow \mathrm{p}_h \langle\tilde{\mathbf{n}}_{p,q}\rangle_N \rightarrow \sigma_n^2\tilde{\mathbf{w}}$, and $\langle\mathbf{e}_{p,q}\rangle_N \rightarrow \dot{\mathbf{0}}$, the PDP can be perfectly estimated, where N is sufficiently large. However, in order to obtain such a large no. of samples, it is difficult for a receiver of practical MIMO-OFDM systems. With an insufficient no. of samples, the PDP can be approximated as $\mathrm{p}_h \approx \langle Dg\left(\mathbf{h}_{p,q}\mathbf{h}_{p,q}^H\right)\rangle_N$.

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

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

[9]

Where $\mathbf{z}_N \triangleq \langle\mathbf{e}_{p,q}\rangle_N + \langle\tilde{\mathbf{n}}_{p,q}\rangle_N - \sigma_n^2\tilde{\mathbf{w}}$ 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

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

[10]

Since $[P_h]i \geq 0$ for all i, the PDP can initially be estimated as

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

[11]

Where $S_{p,q}[n_p]$ is the sample vector of proposed PDP estimator with the l^{th} entry

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

[12]

Where $g^l_{p,q}[n_p] = [\mathbf{g}_{p,q}[n_p]]_l$ and $\tilde{w}^l = [\tilde{\mathbf{w}}]_l$.

To reduce the detrimental effect of residual noise z_N, the proposed scheme estimates the average of residual noise at the zero-taps of Ph. At the lth entry of $\hat{P}_{init}$, the zero-tap can be detected as

$$t_z^l = \begin{cases} 1 & \text{if } \hat{p}^l_{init} < \beta_{th} \\ 0 & \text{otherwise}\end{cases}$$

[13]

where $\beta_{th} = \frac{1}{L_g}\sum_{l=0}^{L_g-1}\hat{p}^l_{init}$ 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

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

[14]

Where $N_z = \sum_{l=0}^{L_g-1} t^l$ represents the total no. of detected zero-taps. With the estimation of residual noise, the lth tap of the PDP estimate, $P\hat{}_h$, can be expressed as

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

[15]

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

4. Performance and Complexity Analysis

From [15], the LMMSE channel estimator with the imperfect PDP, is given by

$$W_{f,p}=F_L Dg(p\hat{}_h)F^H_p(F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n IK_p)^{-1}$$

[16]

Where F_L is the $K_d \times L_g$ matrix and is obtained by considering the first L_g columns of the DFT matrix. And $p\hat{}_h = p_h + e^l_{pdp}$ is expressed as the estimated PDP, where the l^{th} element of e^{-l}_{pdp} is defined as

$$\tilde{e}^l_{pdp} = \begin{cases} [\tilde{e}_N]_l - \hat{n}_{R,avg} & \text{if } [\tilde{e}_N]_l > \hat{n}_{R,avg} \\ 0 & \text{otherwise} \end{cases}$$

[17]

From the matrix inversion, $(F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n I_{Kp})^{-1}$ in [16] is converted as

$$(F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n I_{Kp})^{-1} = A^{-1}-A^{-1}F_p BF^H_p A^{-1}$$

[18]

Where $A \triangleq (F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n I_{Kp})$ and

$B \triangleq Dg(e^{-1}_{pdp})\,(I_{Lg}+F^H_p A^{-1}F_p Dg(e^{-1}_{pdp}))^{-1}$.

Then, the coefficient matrix for LMMSE channel estimation with $p\hat{}h$ can be rewritten as

$$W_{f,p} = W_{opt,p} + W_{err,p}$$

[19]

Where

$W_{opt,p} \triangleq F_L Dg(p\hat{}_h)F^H_p(F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n I_{Kp})^{-1}$ is the coefficient matrix for wiener filtering and $W_{err,p}$ is given by

$$W_{err,p} = -F_L Dg(p_h)F^H_p A^{-1}F_p BF^H_p A^{-1} + F_L Dg(e^-_{pdp})F^H_p(F_p Dg(p\hat{}_h)F^H_p+\sigma^2_n I)^{-1}$$

[20]

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

$$E_p = E\{(F_L h_{p,q} - W_{f,p}h\hat{}_{LS,p,q})(F_L h_{p,q} - W_{f,p}h\hat{}_{LS,p,q})^H\}$$
$$= (F_L - W_{f,p}F_p)Dg(p_h)(F_L - W_{f,p}F_p)^H + \sigma^2_n W_{f,p}F_p F_p{}^H W^H{}_{f,p}$$

[21]

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

$$MSE_{f,p} = T_r(E_p) / T_r(F_L Dg(p_h)F^H{}_L)$$

[22]

Where $T_r(E_p)$ denotes the trace operation of E_p. With a sufficiently large no. of samples, $e^-{}_{pdp} \rightarrow 0$. Thus, the MSE of the proposed scheme achieves that of wiener filtering because $W_{f,p} \rightarrow W_{opt,p}$.

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

5. Simulation Results

A MIMO-OFDM system with the physical layer parameters for downlink of 3GPP- LTE has been considered. The MIMO-OFDM system utilizes four transmit (P= 4) and two receive (Q=2) antennas. We assume that the pilots of the four transmit antenna ports are distributed as the time and frequency grid of the LTE OFDM system as shown in fig.2. As per the data sheets of 3GPP LTE system, the parameters for downlink LTE are mentioned in table1.

S. No	Parameter	Specification
1.	Modulation Technique	OFDM
2.	Subcarrier spacing	5KHz
3.	Transmission Band width	5MHz
4.	FFT size	512
5.	Access scheme (DL)	OFDMA
6.	No. of Sub carriers used	300
7.	Carrier frequency	2GHz
8.	Length of CP (L_g)	40
9.	Constellation Mapping Schemes	QPSK,16QAM, 64QAM
10.	Sampling Frequency	7.68 MHz

Table 1: LTE Downlink Parameters

The Mean Square error (MSE) performance of LMMSE technique using the estimated PDP is as shown in fig 3, where all underlying links are modeled as extended typical urban (ETU) channel. The performance of LMMSE technique using the approximated PDP, which is uniform or exponential model with the channel delay parameter estimation is plotted. It is required to note that, the LMMSE techniques using the estimated PDP shows better performance than the conventional methods as this technique reduces the correlation mismatch.

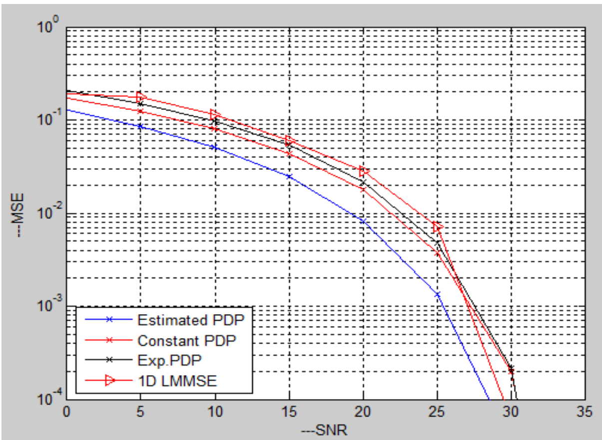

Fig 3: MSE Performance of LMMSE technique using the estimated PDP over ETU channel

Fig 4 shows the MSE performance of the LMMSE technique over the exponentially power decaying six path Rayleigh fading channel model, where the channel maximum delay (L_{ch}), is variable. The PDP of the channel model is defined as

$$E\{|\alpha^{l}_{p,q}[n]|^2\}=\frac{1}{S_h}e^{l/\tau_{rms}}$$

For l = 0, $\Delta\tau$, ... $5\Delta\tau$ and $\Delta\tau = L_{ch} / 5$

$\tau_{rms} = L_{ch} / \log(2 L_{ch})$ & S_h is the normalization factor, is given as $S_h = \Sigma_l e^{-l/\tau_{rms}}$. The performance of the proposed scheme is better than that of the existing methods.

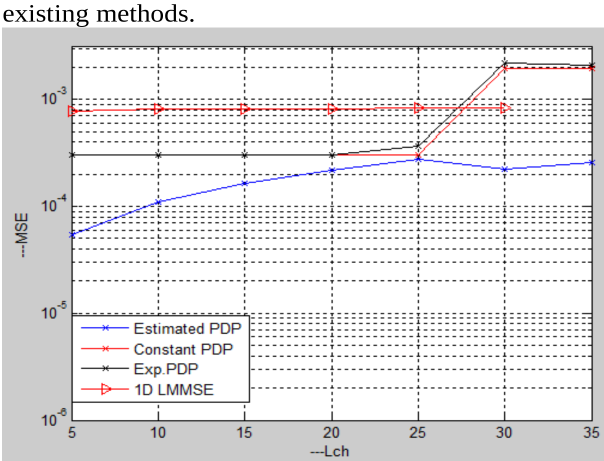

Fig 4: Performance of LMMSE technique using the estimated PDP over 6 – Ray exponential channel with variable channel maximum delays

In fig 5, the MSE performance of LMMSE technique using the estimated PDP for different mobile equipment speeds ranging from 0 to 100 kmph at 30db SNR and doppler frequency = 9.26 – 203. 7 Hz is shown. All underlying links are modeled as ETU channels. Fig 5 shows the MSE of LMMMSE technique using the estimated PDP achieves better results than the conventional methods even at high Doppler frequencies.

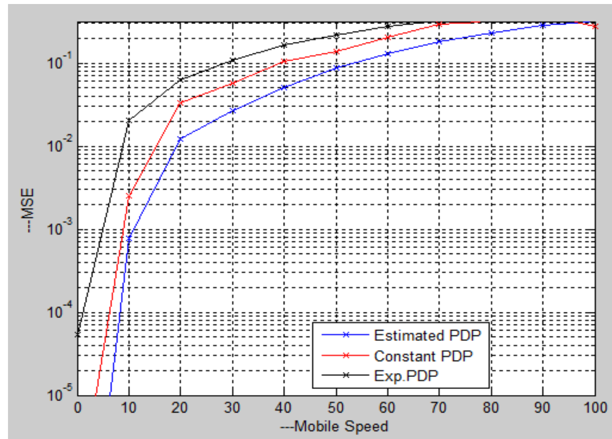

Fig 5: Performance of LMMSE technique using the estimated PDP over ETU channels with different mobile equipment speeds

Fig 6 shows simulation and analysis results of the frequency domain LMMSE channel estimation with various no. of samples for obtaining the PDP at 20db SNR (N = $|\tau_p|$ PQ). The simulation results correspond to the channel estimation performance at the first OFDM symbol of antenna port 1 is as shown in fig 2. We obtain the analytic results 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 proposed technique improves the MSE performance with an increase in the no. of samples for PDP estimation.

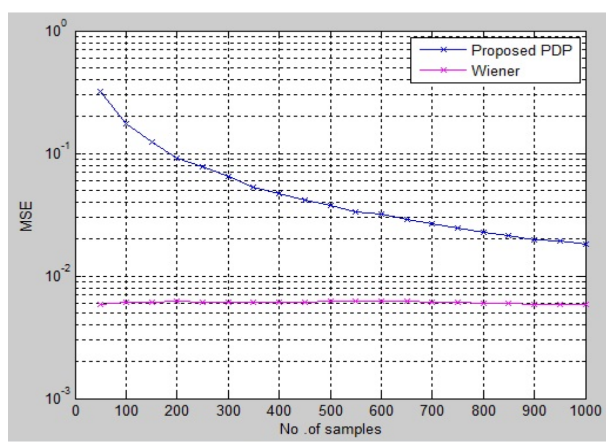

Fig 6: Simulation and analysis results of LMMSE channel estimation over ETU channel with various samples for PDP estimation

6. Conclusion

In this paper, a power delay profile (PDP) estimation technique for the LMMSE channel estimator of MIMO-OFDM systems has been proposed. By using only the pilot symbols of all transmit antenna ports, the proposed technique estimates the PDP with low computational complexity. Also the channel impulse response (CIR) estimates at every path of the MIMO channels where used to obtain the PDP. In addition to this, for accurate PDP estimation, this technique effectively reduces the distortions caused by null sub carriers and inefficient no. of CIR samples. The simulation results shows that the performance of LMMSE channel estimation using the proposed PDP estimates is much better than the existing methods like uniform or exponential PDP estimates.

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8. Biographies

Mrs. Y. Lavanya completed B.Tech in Kuppam Engineering College, affiliated to JNTU, Hyderabad in 2007. She is having 4 years of teaching experience. At present she is pursuing M.Tech with the specialization "Digital Electronics & Communication Systems" from St.Ann's College of Engineering & Technology, Chirala, affiliated to JNTU, Kakinada.

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