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A NOVEL CHANNEL ESTIMATION TECHNIQUE FOR COMPLEXITY REDUCTION OF LEAST MINIMUM MEAN SQUARE ERROR

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ABSTRACT

Performance comparison of several channel estimation techniques in the downlink of LTE systems was performed based upon the algorithm's complexity which varies according to the density of the channel matrix. In this study has been presented a proposed channel estimation technique whose complexity is invariant with the number of multipath components in fast fading channels. The proposed channel estimator is derived by a review of the mathematical model of the received OFDM signal in the down link of LTE systems its performance is evaluated using the mean squared error as the performance metric of interest. Also a study of the effect of the proposed channel estimation technique on the bit error rate performance of the downlink of LTE systems is also investigated numerically through simulation.

Keywords: LTE System, Channel Estimation, OFDM System, MIMO System

1. INTRODUCTION

Wireless multimedia services require a high speed data access in order to satisfy their exponential growing dem. Therefore, various techniques have been proposed in recent years to achieve high system capacities. Among them, we interest to the Multiple-Input Multiple Output (MIMO). The MIMO concept has attracted lot of attention in wireless communications due to its potential to increase the system capacity without extra bwidth (Rana, 2010). Multipath propagation usually causes selective frequency channels, to combat the effect of frequency selective fading, MIMO is associated with Orthogonal Frequency-Division Multiplexing (OFDM) technique. OFDM is a modulation technique which transforms frequency selective channel into a set of parallel flat fading channels. A cyclic prefix CP is added at the beginning of each OFDM symbol to eliminate ICIISI. However, in most of these research works, the inserted cyclic prefix CP length is assumed to be equal or longer than the maximum propagation delay of the channel. But in some casesbecause of some unforeseen channel behaviour, the cyclic prefix can be shorter than propagation delay of the channel (Jing and Fang, 2011). In this case, both ICIISI will be introducedthis makes the task of channel estimation more difficult. Equalization techniques that could flexibly detect the signals in both cases in MIMO-OFDM systems are discussed in (Schwarz *et al*., 2010; Ketonen *et al*., 2009). The 3GPP Long Term Evolution (LTE) defining the next generation radio access network. LTE Downlink systems adopt Orthogonal Frequency Division Multiple Access (OFDMA) MIMO to provide up to 100 Mbps (assuming a 2×2 MIMO system with 20MHz bwidth). The performance of a MIMO-OFDM communication system significantly depends upon the channel estimation. Channel estimation techniques for MIMO-OFDM systems were carried out in many articles (Jiang *et al*., 2010; Deng and Wang, 2011).

The performance evaluation of the two channel estimation techniques: Least Square (LS) Least Minimum

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Mean Square Error (LMMSE) are discussed in (Hou and Liu, 2010; Chen *et al*., 2011) for LTE but many work still required for the study of the performance evaluation of those two estimators in LTE downlink under the effect of channel propagation delay to enhance the signal quality of LS estimatorin the same time reduce the complexity of the LMMSE estimator (Song *et al*., 2011).

In this study we introducedanproposed algorithm for channel estimation in downlink LTE. The proposed algorithm has less complexity than the LS algorithm due to using matrix transpose rather than matrix inverse.The proposed algorithm has less BER than LMMSE as the matrix A doesn't containsall data of the user but it contains neighbour's users only. The time correlated channel was generated by an implementation of the Rosa Zheng model. Also as the LTE is a synchronized system, the proposedalgorithm improves this synchronizationthe SNR raisethe MSE reduced because raising of the N_{fft} make two things reduce the error that produced from non-orthogonalityreduce the AWGNall interfaces signals. Sources of Errors that includes in the proposed algorithm are possible imperfect orthogonality of the used carries, Additive White Gaussian Noise (AWGN) Quantization of the channel delay.

The study is organized as following: 2 describe the systemchannel model, 3 present the proposed channel estimation algorithm 4 provides the simulation results, 5 present the conclusion.

2. SYSTEM MODELCHANNEL MODEL

2.1. System Model

AOFDM system packed by layers of encoding, scrambling…, it is preferable to review the model of this core system in its baseb form. Adiscrete-time baseb OFDM signal in time domain for one symbol interval can be expressed as Equation 1:

$$x_n = \frac{1}{N_{fft}} \sum_{k=1}^{N_{fft}-1} X_k e^{j\frac{2\pi kn}{N_{fft}}} \quad (1)$$

where, x_n is the transmitted symbol, n denotes the discrete-time index $0 \le n \le N_{fft}-1$, N_{fft} is the duration of one OFDM symbol interval in samples, the number of subcarriers is also N_{fft}, K represents the discrete-frequency domai nindex of the K^{th} sub-carrier. In most OFDM systems, it is commonly to insert Cyclic Prefix (CP) of N_g samples length before the data symbol as a guard interval to prevent Inter-Symbol Interference (ISI) Equation 2:

$$x_n = \frac{1}{N_{fft}} \sum_{k=1}^{N_{fft}-1} X_k e^{j\frac{2\pi kn}{N_{fft}}}, ... - N_g < n < N_{fft} - 1 \quad (2)$$

Or in matrix form as Equation 3:

$$x_n = \frac{1}{N_{fft}} F x_k \quad (3)$$

where, $x, x_n \in C^{N_{fft} \times 1}$ are the inputoutput vectors respectivelythe (m, n) entry of $F \in C^{N_{fft} \times N_{fft}}$ is viewed as an Inverse Fast Fourier Transform (IFFT) matrix operator defined as Equation 4:

$$F = \sum_{k=1}^{N_{fft}-1} e^{j\frac{2\pi kn}{N_{fft}}}, 0 \le k, n \le N_{fft} - 1 \quad (4)$$

2.2. Channel Model

For the convenience of analysis, the channel impulse response of any multipath channel can be re-written, without loss of generality, by introducing a one-to-one mapping from attenuation-delay pairs to equivalent Multi Path Components (MPCs) representation as follows Equation 5:

$$h(t) = \sum_{l=0}^{N_p-1} \alpha_l \delta(t - \tau_l) \quad (5)$$

where, h(t) is the channel response, N_p is the total number of MPCs, α_L τ_L area the path attenuationthe corresponding delay of any received MPC, respectively, with*l*representing the index of this path. The channel weights α_L are assumed to be complex in general. This allows the CIR of the RF channel to be expressed in a general form that is compatible with any multipath channel profile regardless of the nature of the amplitudedelay statistics. Also, for convenience of analysis, the whole time duration of the CIR is divided into time bins i.e., we let $\tau_1 = 1\Delta\tau$ and I = 0, 1,…, N_p-1, 1,..., N_p-1 where $\Delta\tau$ is thetiming resolution. This time resolution will be set to the symbol duration in the process of channel estimation as will be seen in the proposed channel estimation algorithm. According to the timing resolution limitation of either the receiver sampling or the bwidth of the pulse, $\Delta\tau$ is selected in simulations so as no more than one MPC arrives in this minimum timing resolution. This last form of the CIR suggests that the channel can be represented as an FIR filter as illustrated in **Fig. 1**.

It can be intuitively thought that the FIR model of the physical RF channel is a well sampled theoretical version of the physical channel model for the same channel. In this study, it is assumed that the maximum delay of multipath channel is smaller than guard interval, which means no ISI occurs.

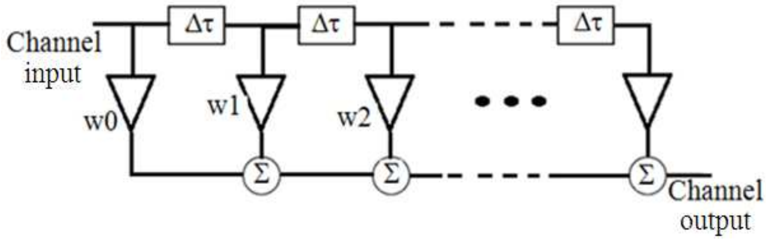

Fig. 1. The FIR filter model of a multipath channel

The ROSA model is new sum-of-sinusoids statistical simulation models which are proposed for Rayleigh fading channels. These new models employ rom path gain, rom initial phase, conditional rom Doppler frequency for all individual sinusoids. It is shown that the autocorrelationscross correlations of the quadrature components, the autocorrelation of the complex envelope of the new simulators match the desired ones exactly, even if the number of sinusoids is as small as a single-digit integer. Moreover, the probability density functions of the envelopephase, the level crossing rate, the average fade duration, the autocorrelation of the squared fading envelope which contains fourth order statistics of the new simulators, asymptotically approach the correct ones as the number of sinusoids approaches infinity, while good convergence is achieved even when the number of sinusoids is as small aseight. The new simulators can be directly used to generate multiple uncorrelated fading waveforms for frequency selective fading channels, multiple-input multiple-output channels, diversity combining scenarios. Statistical properties of one of the new simulators are evaluated by numerical results, finding good agreements.

Consider a frequency-nonselective fading channel comprised of N_p propagation paths; the low-pass fading process is given by Equation 6:

$$g(t) = E_0 \sum_{l=1}^{N_p} C_n e^{\left[j\left(w_d t \cos\alpha_n + \phi_n\right)\right]} \tag{6}$$

where, E_0 is a scaling constant C_n, α_n and ϕ_n are, respectively, the rom path gain, angle of incoming wave, initial phase associated with the l^{th} propagation path, w_d is the maximum radian Doppler frequency occurring when $\alpha_n = 0$.

2.3. Received Signal Model

At the receiver side, the received base b signal over multipath time-varying channel after removing the guard interval at one receive antenna port can be expressed as Equation 7:

$$y_n(t) = \sum_{l=0}^{N_p-1} (h_n^{(l)} x_{n-\tau(l)})(t) + \eta(t) \tag{7}$$

where, $y_n(t)$ is the received symbol, $\eta(t)$ Gaussian noise, A Low Pass Filter (LPF) is needed in the demodulation process, implemented as an integrator whose integrationinterval spans an OFDM symbol duration:

$$\int_{t_o}^{t_o+T_s} y_n(t)\ d\,t = \int_{t_o}^{t_o+T_s} \left(\sum_{l=0}^{N_p-1} \begin{matrix} (h_n^{(l)} x_{n-\tau(l)}) \\ (t) + \eta(t) \end{matrix} \right) d\,t \tag{8}$$

The integral in Equation (8) is then sampled prior to applying aFast Fourier Transform (FFT) operationan observation vector y_n is formed on a symbol by symbol basis by collecting together N_{fft} successive output samples corresponding to a time interval of OFDM symbol duration (Aziz, 2011). This observation vector can be written in matrix form as Equation 9:

$$y_n = A_n + w_n \tag{9}$$

where, $A_n \in C^{N_{fft}\times 1}$ and w_n is a realization vector of the sampled AWGN rom process n (t) Extracting the channel parameters from the vector A_n is a difficult task as the vector A_n is in general complex for a baseb processing system as it includes the known symbol matrix, thepilot sequence of a pilot symbol help inchannel estimation process,the unknown channel parameters (Aziz *et al*., 2011). To facilitate the process of estimation of the unknown channel parameters it is necessary to decompose A_n into the product of two independent matrices as Equation 10:

$$y_n = H_n x_n + w_n \tag{10}$$

where, the matrix H_n is the channel matrix $x_n \in C^{N_{fft}\times 1}$ is the information bearing vector. In Equation (8), we assume that the system is generally asynchronous, that is, the OFDM receiver has an arbitrary timing reference which will not necessarily be aligned with the OFDM symbol boundaries, thus the integration period over one symbol may in general overlap two adjacent received symbols, the actual receiving situation of non integer delayed symbol sequence. Each sample can be viewed as composed of a linear combination of the values of these two symbols, that is, the pastthe current symbols X_n, X_{n-1} Equation 11 and 12:

$$\int_{t_0+nT_s}^{t_0+(n+1)T_s} y_n(t)\, dt = \int_{t_0+nT_s}^{t_0+(n+1)T_s} \left(\sum_{l=0}^{N_p-1} (h_n^{(l)} x_{n-\tau(l)}) (t) + \eta(t) \right) dt \quad (11)$$

$$\int_{t_0+nT_s}^{t_0+(n+1)T_s} y_n(t)\, dt = I_n^L + I_n^R + w_n \quad (12)$$

where, I_n^L is the part of the integral that yields the value of the integration over ςT_s and I_n^R is the part of the integral that yields the value of the integration over $(1-\varsigma)T_s$. Consequently the observation vector y_n is a superposition of two exclusive vectors a_n^R, a_n^L Equation 13:

$$y_n = [a_n^R + a_n^L] + W_n \quad (13)$$

As the sampler is a discrete time device, the time delays of the channel will be quantized by normalization to the sampling interval which is the same as the OFDM symbol durationT_s hence Equation 14-16:

$$\frac{\tau_p}{T_s} = q + \zeta, q \in \{0,1,\ldots,N_{FFT}-1\}, \zeta \in [0,1) \quad (14)$$

$$I_{n,k}^L = \sum_{l=0}^{N_p-1} \left(\alpha_l (1-\zeta) x_n \int_{t_0+nT_s}^{t_0+(n+1)T_s} y_n(t)\, c_k(t-qT_s)\, dt + \alpha_l \zeta x_{n-1} \int_{t_0+nT_s}^{t_0+(n+1)T_s} y_n(t)\, c_k(t-qT_s)\, dt \right) \quad (15)$$

$$I_{n,k}^L = \sum_{l=0}^{N_p-1} \left(\alpha_l (1-\zeta) x_n \int_{o+nT}^{t_0+(n+1)T_s} y(t)\, c_k(t+qT_s)\, dt + \alpha_l \zeta x_{n-1} \int_{t_0+nT_s}^{o+(n+1)T} y_n(t)\, c_k(t+qT_s)\, dt \right) \quad (16)$$

where, q is the integer number of the part of the symbol, ζ is the fraction number of the part of the symbol. The contribution of the K^{th} subcarrier C_K (t) in the vector y_n appears in two parts, the right part of the OFDM symbol shown as R appearing in the left symbol x_{i-1} the left part shown as L in the right symbol, χ_i The weights for the linear combination of the two vectors that contribute to the right part of symbols χ_i are ζ 1-ζ, where ζ is the fractional part of the delay. Let's define the matrices $U^R \in C^{N_{FFT}\times N_{FFT}}$ and $U^L \in C^{N_{FFT}\times N_{FFT}}$ that are formed by arranging columns of $C(t-qT_s)$ $C(t+qT_s)$ representing the orthogonal carriers defined on $0 \le t \le T_s$ delayed by all possible integer delays q. Hence we can writeU^R, U^L as Equation 17 and 18:

$$U^L[q] = [c^L[0] \ \ c^L[1] \ \ c^L[N_{fft}-1]] \quad (17)$$

$$U^R[q] = [c^R[0] \ \ c^R[1] \ \ c^R[N_{fft}-1]] \quad (18)$$

where, $c_k^{L,R}[q]$ represents the K^{th} subcarrier C_K (t), truncatedzero padded, representing either the left or right cyclically shifted by q samples.Let Z $K \in C^{N_{fft}\times 1}$ be the composite channel impulse response vector sampled N_{fft} times over T_s secondsconsisting of all the weight vectors in the linear combination of the vectors $c_k^{L,R}[q]$, while the vector Z_n is expressed as follows Equation 19-21:

$$z_n = [0 \ldots M \ldots 0] \quad (19)$$

$$M = [\ \alpha_1 (1-\zeta)\ \ \alpha_1 \zeta\ \ \alpha_{N_p} (1-\zeta)\ \ \alpha_{N_p} \zeta\ ] \quad (20)$$

$$y_n = [U^R \ \ U^L](z_n \odot x_n) + w_n = [U^R z_n U^L z_n] x_n + w_n \quad (21)$$

where, (⊕) is defined as the element wise multiplication operator. The columns of A can be expressed in terms of U^R, U^L and Z_n as Equation 22 and 23:

$$a_n^R = U^R z_n \quad (22)$$

$$a^R{}_N = U^R Z_N \quad (23)$$

The received observation vector $y_n$in Equation (9) is thus Equation 24:

$$y_n = [U^R z_n \; U^L z_n] x_n + w_n; \quad x_n = [\, x_n \quad x_{n-1}]^T \quad (24)$$

where, $[.]^T$ is the matrix transpose operator.

3. THE PROPOSED CHANNEL ESTIMATION ALGORITHM

The mathematical model for the received signal of the intended user should be used to extract the channel parameters, that is, the attenuationsthe corresponding delays.The proposed model has the advantage of being consistent with a general multipath channel. The vector entries χ_n, χ_{n-1} represent either the information or training symbols. We shall set χ_n =1 during the estimation course such that yn includes only the channel coefficients (Hong-Jin and Li-Fa, 2011). This assumption has the advantage that no specific form for the pilot sequence is required; rather, the process of channel estimation will be done by only periodic transmission of a linear combination of all carriers with equal amplitudes for a number of times equal to the length of the pilot sequence in symbols. For only one symbol duration Equation (24) can now be expressed as follows Equation 25:

$$y_n = [U^R + U^L] z_n + w_n \quad (25)$$

Or equivalently Equation 26:

$$y_n = [U^R + U^L] z_n + w_n \quad (26)$$

Multiplying both sides of Equation (26) by U^T Equation 27:

$$U^T y_n = U^T U z_n + U^T w_n \quad (27)$$

The product U^T U is a $N_{fft} \times N_{fft}$ matrix whose elements are merely the values of the autocorrelation function of the subcarrier matrix evaluated at different discrete time shifts ranging from 0 to $N_{fft}{}^{-1}$ thus Equation 28:

$$U^T U = R \quad (28)$$

Now R is a Toeplitz autocorrelation matrix with the entry (i, j) of the matrix R is defined as follows Equation 29:

$$R_{ij} = c^J[i] c[j] \quad (29)$$

With $\exp^T$ [j] is a vector of the k^{th} sampled sub-carriershifted by i samples and c[J] is at the same carrier shifted version of the same vector shifted by J samples, thus the value of R_{ij} is equal to N_{fft} times δ_{ij}, the Kronicker delta function, ifonly if the carriers are perfect orthogonal, however, for purpose of generalization, no need to assume that perfect orthogonality, we shall instead assume that the used carriers violates the orthogonality property, may be due to a Doppler shift,the autocorrelation matrix R can be considered as consisting of the sum of two matrices, the identity matrix I scaled by N_{fft} that will result if the carriers are a perfect orthogonal setan error matrix E that will appear only with the non orthogonal part as follows Equation 30:

$$R = U^T U = N_{fft}(I + E) \quad (30)$$

where, the entry (i, j) of the matrix E is expressed as follows Equation 31:

$$E_{ij} = \begin{cases} \dfrac{c^T[i]c[j]}{N_{fft}}, & i = j \\ 0 & i = j \end{cases} \quad (31)$$

The diagonal elements of the matrix E are always equal to zeroas the number of these orthogonal sub-carriers N_{fft} is large enough it will be valid to assume that the product $U^T U$ can be well approximated by an identity matrix multiplied by N_{fft}. strictly speaking Equation 32:

$$\lim_{N_{fft} \to \infty} E = O \quad (32)$$

With O being the zero matrix, thus Equation 33:

$$\lim_{N_{fft}\to\infty} \frac{U^T y_n}{N_{fft}} = z \tag{33}$$

The limit in Equation (33) can be replacedpractically by using a very large number of the orthogonal sub-carriers however this necessitates raising the limit from Equation (33):

$$\frac{1}{N_{fft}}\left(U^T y_n\right) = z_n + z_n E + \frac{1}{N_{fft}}\left(U^T w_n\right) \tag{34}$$

The LHS of Equation (34) is the estimate of the CIR vector z appearing on the RHS. Let's define the estimated CIR vector $\hat{z}$ as follows Equation 35:

$$\hat{z}_n = \frac{1}{N_{fft}}\left(U^T U x_i\right) + e_i \tag{35}$$

With, e_n representing an error vector defined as Equation 36:

$$e_n = E + \frac{1}{N_{fft}}\left(U^T w_n\right) \tag{36}$$

This error is induced by one or more of three sources, that is, the possible imperfect orthogonality of the receivedsub-carriers, possibly due to Doppler shift resulting from mobility, the additive white Gaussian noisethe quantization of the channel delays.

4. SIMULATION RESULTS

We present simulation results and discuss the performance of the different channel estimation techniques while the MSE performance of the proposed channel estimation algorithm is compared with the optimal LMMSE estimation technique and the low rank LS estimation techniques (Zhao *et al*., 2010). The three techniques are performed in the frequency domain as it is well known that time domain channel estimation leads to the problem of propagating errors due to the back and forth FFT transformation that may lead to residual errors. Simulation parameters that were used to evaluate the system performance are shown below in **Table 1**. The mean squared error is the performance metric of interest that is used to evaluate the performance of the proposed estimation algorithm and also to compare its performance with other known channel estimation techniques.

Table 1. Simulation parameters

System	LTE
Modulation	16-QAM
Symbol time	224 μs
Bandwidth	5.71 MHz
Speed	60 km/hr
FFT size	64,128,256
Carrier frequency	2.4 GHz
Number of pilots	None
Channel model	Rosa Zheng, dispersive, multipath
Channel estimators	LMMSE, LS Vs Proposed

Since the theoretical MSE expressions are not evaluated in this study, Monte Carlo simulation technique was used by running the simulation code for 1000 times, calculating the MSE every iteration and the result is averaged over all iterations to ensure that the obtained simulation results approaches the un-evaluated theoretical expressions. The time correlated channel was generated by an implementation of the Rosa. We generate a time correlated channel impulse response for every sample of the baseband transmit signal. As can be seen from the figures shown below that in all cases, the MSE of a channel estimation technique should decrease with increasing SNR. Moreover as deduced in section V of the proposed channel estimation algorithm, the proposed channel estimation technique also decreases with increasing the FFT size. **Figure 2** Illustrates the MSE performance of the proposed channel estimator at different FFT size values. **Figure 3** illustrates the MSE of the proposed channel estimation algorithm Vs. SNR at different values of the FFT size. It can be noticed that the FFT size has a dramatic effect on the value of the MSE at a given SNR. Current OFDM based systems use an FFT size that is larger than 1024 which guarantees a MSE below 1e-4 even at low SNR values. Mean squared error performance of the proposed channel estimator vs. FFT size is depicted in **Fig. 3** at three different SNR values. This figure also emphasizes the same deduced inverse proportionality relation between the MSE and the FFT size. **Figure 4** depicts the MSE performance comparison between the proposed channel estimator and LS, LMMSE estimators for a user velocity of 60 km h^{-1} at FFT size N_{fft} = 64. **Figure 5** depicts the MSE of the presented channel estimators for a user velocity of 60 km h^{-1} at FFT size N_{fft} = 128 at a given SNR. **Figure 6** the difference in MSE between the proposed estimator and the LMMSE estimator is decreasing as the FFT size increases from 64 to 256.

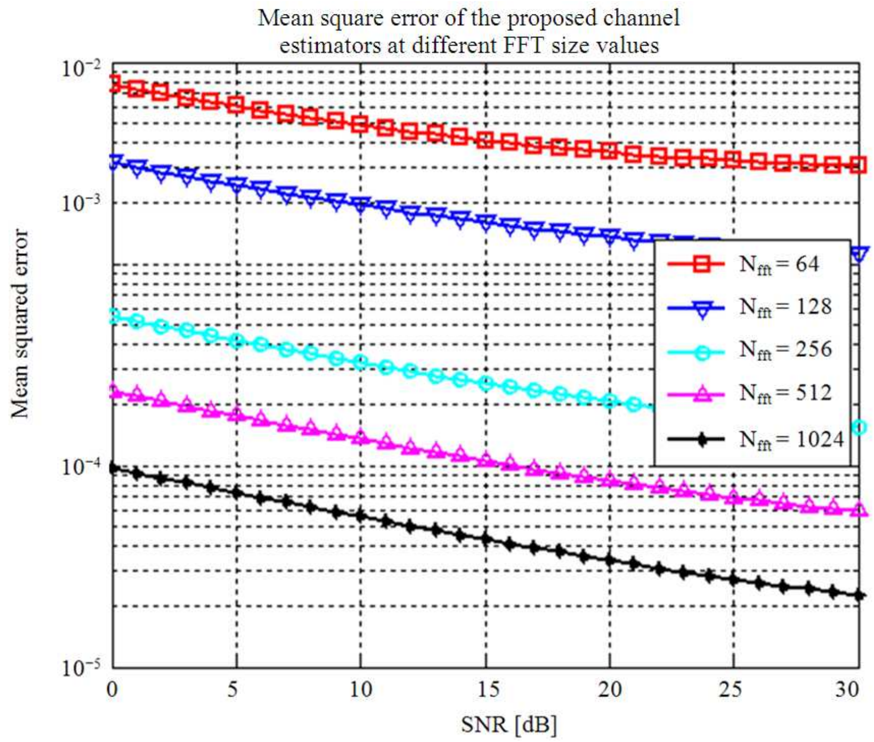

Fig. 2. Mean squared error performance of theproposed channel estimator at different FFT size values

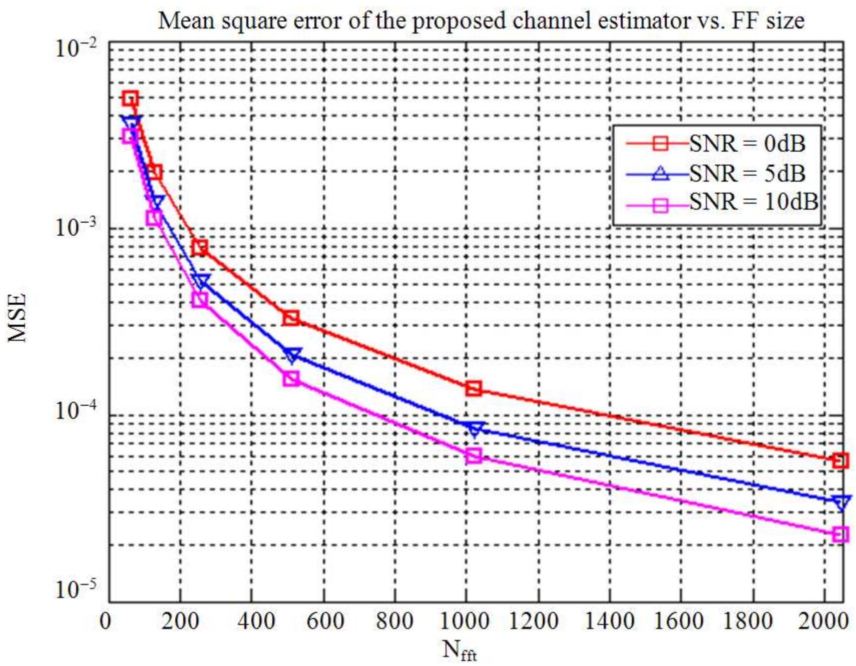

Fig. 3. Mean squared error performance of the proposedchannel estimator Vs FFT size

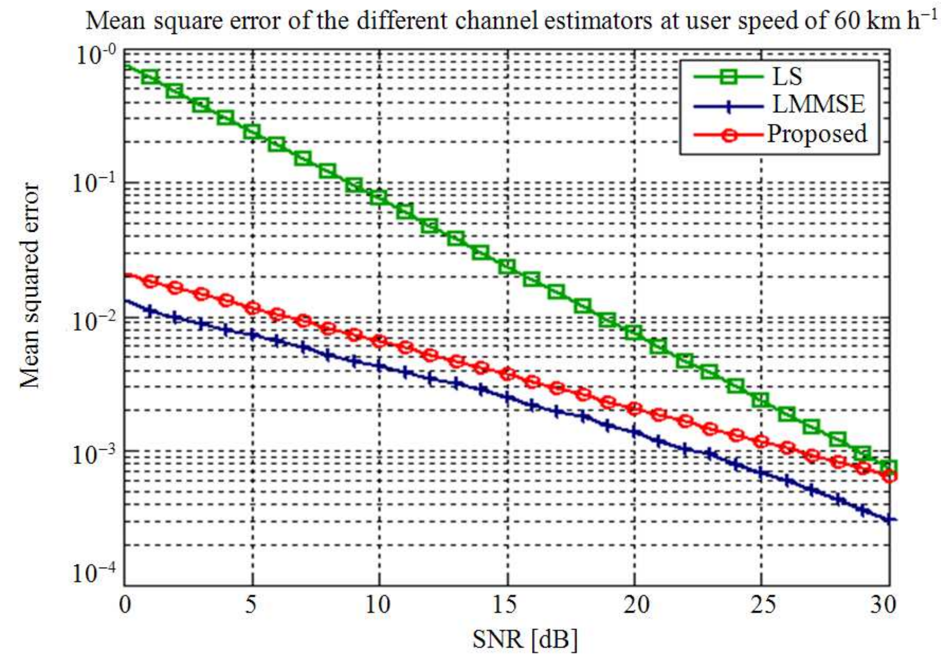

Fig. 4. MSE performance comparison between the proposed channel estimator, LS and LMMSE estimators $N_{fft} = 64$

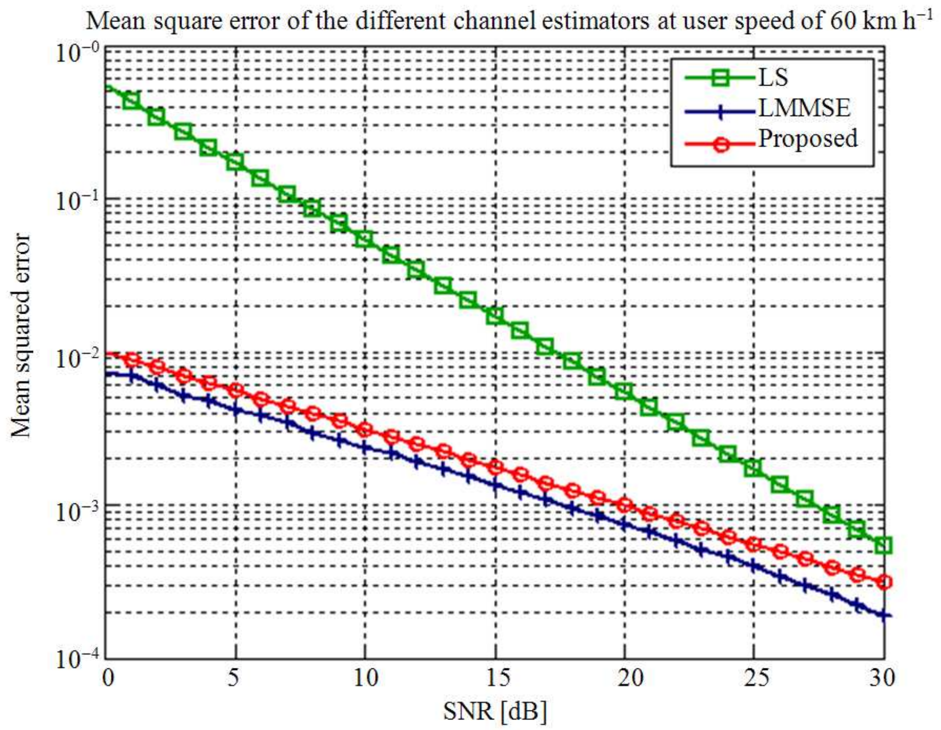

Fig. 5. MSE performance comparison between the proposed channel estimator, LS and LMMSE estimators $N_{fft} = 128$

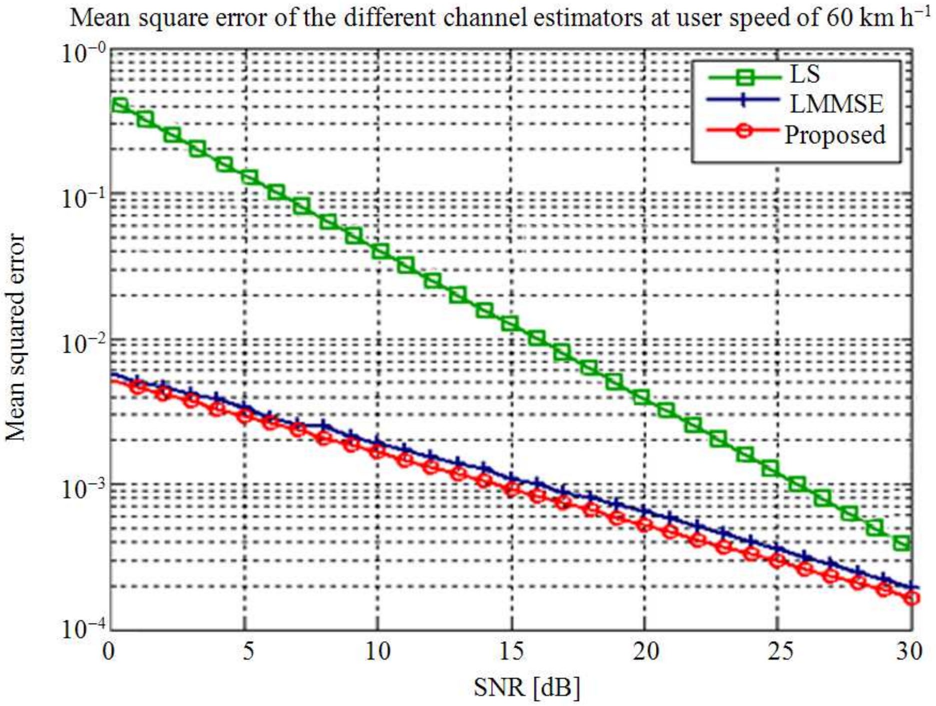

Fig. 6. MSE performance comparison between the proposed channel estimator, LS and LMMSE estimators $N_{fft} = 256$

5. CONCLUSION

A novel channel estimation algorithm has been proposedpresented. This algorithmcan be adopted to LTE systems. The technique has the advantage of complexity invariant with the channel matrix representing a dense multipath channelthe complexity is reduced to merely a matrix multiplication process. Simulation results show that the proposed algorithm is competitive the optimum LMMSE from a complexity point of viewthe LS algorithm if the accuracy is considered when the SNRthe MSE are the performance metrics of interest.

6. REFERENCES

Aziz, A.M., 2011. A soft-decision fusion approach for multiple-sensor distributed binary detection systems. IEEE Trans. Aeros. Electr. Syst., 47: 2208-2216. DOI: 10.1109/TAES.2011.5937293

Aziz, A.M., M. Ahmed, A. Mohamed and G.M.A. Hamid, 2011. A new soft fusion approach for multiple receivers wireless communication systems. Electr. Telecomm. Res. Instit. J., 33: 310-319.

Chen, Y., L. Cui and C. Qi, 2011. Performance analysis of pilot aided channel estimation methods for MBSFN in LTE system. Proceedings of the International Conference on Computational Problem-Solving, Oct. 21-23, IEEE Xplore Press, Chengdu, pp: 682-687. DOI: 10.1109/ICCPS.2011.6089767

Deng, B. and W. Wang, 2011. Low complexity channel estimation method of MIMO-OFDM for LTE downlink. Proceedings of the 7th International Conference on Wireless Communications, Networking and Mobile Computing, Sept. 23-25, IEEE Xplore Press, Wuhan, pp: 1-4. DOI: 10.1109/wicom.2011.6036656

Hong-Jin, W. and L. Li-Fa, 2011. A modified channel estimation based on svd in LTE-advanced systems. Proceedings of the 7th International Conference on Wireless Communications, Networking and Mobile Computing, Sept. 23-25, IEEE Xplore Press, Wuhan, pp: 1-5. DOI: 10.1109/wicom.2011.6040119

Hou, J. and J. Liu, 2010. A novel channel estimation algorithm for 3GPP LTE downlink system using joint time-frequency two-dimensional iterative Wiener filter. Proceedings of the 12th IEEE International Conference on Communication Technology, Nov. 11-14, IEEE Xplore Press, Nanjing, pp: 289-292. DOI: 10.1109/ICCT.2010.5689239

Jiang, M., N. Prasad and X. Wang, 2010. Design of high performance MIMO receivers for LTE/LTE-a uplink. Proceedings of the 44th Asilomar Conference on Signals, Systems and Computers, Nov. 7-10, IEEE Xplore Press, Pacific Grove, CA., pp: 1493-1497. DOI: 10.1109/ACSSC.2010.5757785

Jing, S.X.Y. and Y. Fang, 2011. Channel estimation for TD-LTE system. Proccedings of the Cross Strait Quad-Regional Radio Science Wireless Technology Conference, Jul. 26-30, IEEE Xplore Press, Harbin, pp: 923-925. DOI: 10.1109/CSQRWC.2011.6037107

Ketonen, J., M. Juntt and J.R. Cavallaro, 2009. Receiver implementation for MIMO-OFDM with AMC and precoding. Proceedings of the Conference Record of the 43th Asilomar Conference on Signals, Systems and Computers, Nov. 1-4, IEEE Xplore Press, Pacific Grove, CA., pp: 1268-1272. DOI: 10.1109/ACSSC.2009.5469950

Rana, M.M., 2010. Channel estimation techniques and LTE terminal implementation challenges. Proceedings of the 13th International Conference on Computer and Information Technology, Dec. 23-25, IEEE Xplore Press, Dhaka, pp: 545-549. DOI: 10.1109/ICCITECHN.2010.5723916

Schwarz, S., C. Mehlf and M. Rupp, 2010. Calculation of the spatial preprocessing and link adaption feedback for 3GPP UMTS/LTE. Proceedings of the 6th Conference on Wireless Advanced, Jun. 27-29, IEEE Xplore Press, London, pp: 1-6. DOI: 10.1109/WIAD.2010.5544947

Song, H., LY. Xiaowen and Yu. Ziming, 2011. A complexity reduction approach for LTE downlink channel estimation. Peoceeding of the International Conference on Internet Technology and Applications, Aug. 16-18, IEEE Xplore Press, Wuhan, pp: 1-4. DOI: 10.1109/ITAP.2011.6006127

Zhao, Q., Z. Zhang and Z. Shi, 2010. Researchlink simulation of key technologies for LTE downlink. Proceedings of the 6th International Conference (WiCOM) on (IEEE) Wireless Communications Networking Mobile Computing, Sept. 23-25, IEEE Xplore Press, Chengdu, pp: 1-4. DOI: 10.1109/WICOM.2010.5600207