Fast Time-Varying Dispersive Channel Estimation and Equalization for 8-PSK Cellular System

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Abstract: **The channel estimation and equalization for EDGE system with time-varying and frequency-selective fading channels are discussed. It is shown that the fast fading channel during a selected slot in the EDGE system can be modeled as a linear function of time, and a linear least-squares algorithm is proposed to estimate the fading channel. For typical channel profiles of the EDGE system, the channel impulse response is not in its minimum phase form, thus cannot be directly used in computationally efficient equalizers, such as delayed decision feedback sequence estimation or reduced state sequence estimation. To overcome this problem, a Cholesky decomposition-based method is introduced to transform the estimated channel impulse response into its minimum phase form. The simulation results show that the proposed algorithms can effectively combat the time-varying and frequency-selective channel fading with Doppler frequency being in a wide range up to 300 Hz.**

I. INTRODUCTION

In recent mobile communication systems, Enhanced Data Rates for Global Evolution(EDGE) was introduced to achieve higher data rates and spectral efficiency. In order to keep backward compatibility with the second generation cellular systems GSM and IS-136, EDGE has similar slot structure and system parameters as GSM. Time-division multiple-access (TDMA) is used in EDGE with the symbol period $T_s = 3.69\mu s$ and a slot length of $576.92\mu s$. Consequently, for static or slow moving communication devices, it is reasonable to assume that the fading channel is time-invariant during the period of one time slot. This assumption is adopted by [8], [9], where various channel estimation and equalization algorithms are developed or applied for EDGE system with slow fading channels. The simulation results obtained in these references are in good agreement with those obtained under ideal cases, *i.e.*, perfect channel estimation with maximum likelihood sequence estimation (MLSE) equalizer. However, the algorithms developed for slow fading channels cannot be directly applied to systems with high mobile speed subscribers, where the time-invariant channel assumption cannot hold.

The channel estimation and equalization for EDGE system with time-varying and frequency-selective fading channels are discussed in this paper. We first analyze the characteristics of the time-varying fading channel, and show that the time-varying property of fast fading channels of EDGE system can be modeled as a linear function of time. Based on this property, we will develop a novel linear LS-based method for EDGE system with fast fading dispersive channels.

The MLSE equalizer with Viterbi algorithm (VA) is currently used in GSM systems, where binary Gaussian minimum-shift keying (GMSK) is used as the modulation scheme. In order to improve the system throughput, 8-PSK modulation is employed in EDGE system. The computational complexity of MLSE equalizer makes it prohibitive to be used in EDGE system. It is shown in [8] that DDFSE and RSSE equalization techniques provide good trade-offs between the system performance and computational complexity. It is important to note that the performance of these equalizers will degrade dramatically if the CIR is not in its minimum-phase form, as the leading taps of the CIR needs to be dominant. For time-invariant channel, prefilter can be used in the system to produce a minimum-phase channel. However, the prefilter approach is not applicable to systems with time-varying CIR without resorting to complex channel tracking. To overcome this problem, a Cholesky decomposition method is introduced in this paper to convert the estimated CIR to its minimum phase form in the presence of a time-varying CIR. Simulation results show that the newly proposed receiver algorithms can obtain very good system performance in terms of both bit error rate and mean square error.

The paper is organized as follows. In section II, the characteristics of the time-varying frequency-selective fading channel in the EDGE system are discussed, and a novel linear LS-based channel estimation algorithm is proposed. Section III introduces the Cholesky decomposition-based method to transform the estimated CIR to its minimum-phase counterpart, and the obtained equivalent CIR is then used in DDFSE or RSSE equalizers to detect the transmitted data. Simulations are carried out in Section IV to demonstrate the performance of the proposed algorithms, and Section V concludes this paper.

II. CHANNEL ESTIMATION

In this section, the properties of the time-varying frequency-selective fading channels of the EDGE system are analyzed. Based on these properties, a linear LS-based channel estimation algorithm is proposed.

A. Channel Characteristics

At the transmitter of the EDGE system, the modulated 8-PSK symbols $a_k \in \{\exp(j\frac{2\pi}{8}i); i = 0, 1, 2, \cdots, 7\}$ are placed in short slots, and a linearized Gaussian filter is used as the transmit filter. The original slot structure of EDGE has 26 pilot symbols in the middle of each slot as shown in Fig. 1a. This slot structure is good enough to be used for estimating time-invariant channel state information of the entire slot. However, when the mobile subscriber is moving fast, the Doppler frequency is high, the fading within one slot is no longer constant, then the original slot structure of EDGE system can not be used for estimating the time-varying channel fading with reasonable good accuracy. Therefore, in this paper, the EDGE slot structure is slightly modified to facilitate the estimation of time-varying fading channels. We split the 26 pilot symbols into two groups, we shift the first group of 13 pilot symbols to the front of the data block and shift the second group of 13 symbols to the end of the data block, and keep the total number of symbols (and data) the same as those of the original slot structure. The modified slot structure is shown in Fig. 1b. The modified slot structure is used throughout this paper.

3	58	26	58	3	8.25
Tail	Data	Training Sequence	Data	Tail	Guard

(a)

3	13	58	58	13	3	8.25
Tail	Training Sequence	Data	Data	Training Sequence	Tail	Guard

(b)

Fig. 1. (a)The original EDGE slot structure, (b)The slightly modified slot structure

The baseband representation of the EDGE system can be written as

$$y_k = \sum_{l=0}^{L-1} h_k(l) x_{k-l} + n_k, \tag{1}$$

where x_k is the transmitted 8-PSK symbol, y_k is the symbol rate sampled output of the receive filter, n_k is the additive white Gaussian noise (AWGN), and $h_k(l)$, $0 \leq l \leq L-1$, is the time-varying CIR of the fading channel. $h_k(l)$ is the symbol rate sampled version of the composite channel impulse response that is the convolution of the transmit filter $p_{_T}(t)$, the receive filter $p_{_R}(t)$, and the physical channel impulse response $g(t,\tau)$, which can be viewed as the response of the channel at time t to an impulse input at time $t-\tau$. In the EDGE system, the transmit filter is a linearized Gaussian filter, the receive filter is a root raised cosine filter, and the physical channel impulse response has the form

$$g(t,\tau) = \sum_i \alpha_i(t)\delta(\tau - \tau_i), \tag{2}$$

where $\alpha_i(t)$ for a certain value of i can be viewed as a time-varying flat fading process with average power $E[|\alpha_i(t)|^2]$ determined by the delay power profile of the channel. When the maximum Doppler frequency, f_d, of the fading channel is in the range of [0, 20] Hz, the multipath fading channels can be considered as time-invariant for one slot duration [8], thus the time variable k can be omitted in the representation of the CIR. Hence, the time-invariant CIR $h(l)$, $0 \leq l \leq L-1$, can be reliably estimated with the conventional LS based algorithms. For typical fast fading channels, the maximum Doppler frequency can be as high as 250 Hz, and the CIR can no longer be treated as time-invariant. It will be shown next that the fading channel can be approximated as a linear function of the time variable.

According to [3], the ith-path of the fading channel can be written as

$$\alpha_i(t) \quad = \quad E_i \sum_{n=1}^{N} C_{ni} \exp[j(\omega_d t \cos\alpha_{ni} + \phi_{ni})], \tag{3}$$

where E_i is a scaling constant, $\omega_d = 2\pi f_d$, α_{ni} and ϕ_{ni} are statistically independent random variables and they are uniformly distributed on $[-\pi, \pi)$. After some algebraic manipulations, we can get

$$\alpha_i(t) = \alpha_{ci}(t) + j\alpha_{si}(t) \tag{4}$$

$$\alpha_{ci}(t) = E_i \sum_{n=1}^{N} C_{ni} \{\cos(\omega_d t \cos\alpha_{ni}) \cos\phi_{ni} - \sin(\omega_d t \cos\alpha_{ni}) \sin\phi_{ni}\} \tag{5}$$

$$\alpha_{si}(t) = E_i \sum_{n=1}^{N} C_{ni} \{\sin(\omega_d t \cos\alpha_{ni}) \cos\phi_{ni} + \cos(\omega_d t \cos\alpha_{ni}) \sin\phi_{ni}\}\,. \tag{6}$$

Since the time duration of one EDGE slot is $576.92\mu s$, when $f_d \leq 100$Hz, we have $|\omega_d t \cos(\alpha_{ni})| \leq 0.3625$ radians for $0 \leq t \leq 576.92\mu$. Consequently we can make the following small angle approximations

$$\cos(\omega_d t \cos\alpha_{ni}) \quad \approx \quad 1 \tag{7}$$

$$\sin(\omega_d t \cos\alpha_{ni}) \quad \approx \quad \omega_d t \cos\alpha_n. \tag{8}$$

Substituting (7) and (8) into (3) with some algebraic manipulations, we will have

$$\alpha_i(t) \quad = \quad E_i \left[\tilde{\alpha}_{ci}(t) + j\tilde{\alpha}_{si}(t)\right] \tag{9}$$

$$\tilde{\alpha}_{ci}(t) \quad = \quad \sum_{n=1}^{N} C_{ni}\{\cos\phi_{ni} - t\omega_d \cos\alpha_{ni} \sin\phi_{ni}\} \tag{10}$$

$$\tilde{\alpha}_{si}(t) \quad = \quad \sum_{n=1}^{N} C_{ni}\{\sin\phi_{ni} + t\omega_d \cos\alpha_{ni} \cos\phi_{ni}\}. \tag{11}$$

It is apparent that $\alpha_i(t)$ can be approximated as a linear function of the time variable t, so as to $h_k(l)$, which is a linear function of $\alpha_i(t)$.

These analysis are supported by Fig. 2, which shows the real part and imaginary part of one tap of a typical channel impulse response within one slot duration with maximum Doppler frequency $f_d = 200$ Hz.

B. Channel Estimation

Having analyzed the characteristics of the Rayleigh fading channel, we can proceed to the estimation of the time-varying frequency-selective channel impulse response $h_k(l)$.

Based on the analysis in Section II-A, we approximate the CIR $h_k(l)$ as a linear function of the time variable k,

$$h_k(l) = u_0(l) + k u_1(l), \tag{12}$$

where $u_0(l)$ and $u_1(l)$ are parameters to be estimated. For a time-varying frequency-selective fading channel with channel length L,

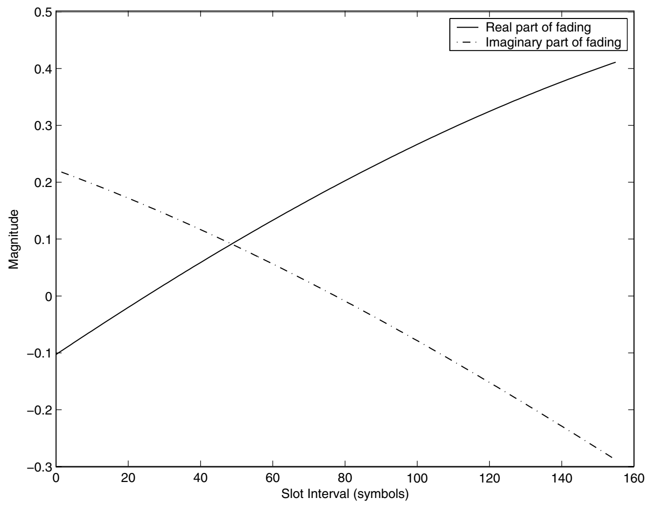

Fig. 2. Real and imaginary part of the Rayleigh fading in one slot interval at f_d=200Hz

there are $2L$ parameters to be estimated for each slot, and the channel impulse response of one slot can be linearly approximated by these parameters.

From (1) and (12), the kth received sample y_k can be represented as

$$y_k = \mathbf{x}_k(\mathbf{u}_0 + k \cdot \mathbf{u}_1) + n_k, \tag{13}$$

where $\mathbf{x}_k = [x_k, x_{k-1}, \cdots, x_{k-L+1}] \in \mathbb{C}^{1\times L}$ are the transmitted symbols, and $\mathbf{u}_i = [u_i(0), u_i(1), \cdots, u_i(L-1)]^T \in \mathbb{C}^{L\times 1}$, for $i = 0, 1$, with $(\cdot)^T$ representing the operation of transpose. With the known training symbols transmitted in the beginning and end of each slot, the received samples contributed exclusively by training symbols can be written into matrix format

$$\mathbf{y} = \mathbf{A}\mathbf{u_0} + \mathbf{T}\cdot\mathbf{A}\cdot\mathbf{u_1} + \mathbf{n}, \tag{14}$$

$$= \begin{bmatrix} \mathbf{A} & \mathbf{T}\cdot\mathbf{A} \end{bmatrix} \begin{bmatrix} \mathbf{u_0} \\ \mathbf{u_1} \end{bmatrix} + \mathbf{n}, \tag{15}$$

where

$$\mathbf{y} = \begin{bmatrix} y_{L-1} \cdots y_{15}\ y_{132+L-1} \cdots y_{147} \end{bmatrix}^T \in \mathbb{C}^{(34-2L)\times 1} \tag{16}$$

$$\mathbf{n} = \begin{bmatrix} n_{L-1} \cdots n_{15}\ n_{132+L-1} \cdots n_{147} \end{bmatrix}^T \in \mathbb{C}^{(34-2L)\times 1} \tag{17}$$

$$\mathbf{A} = \begin{bmatrix} x_{L-1} & \cdots & x_0 \\ \vdots & \vdots & \vdots \\ x_{15} & \cdots & x_{15-L+1} \\ x_{132+L-1} & \cdots & x_{132} \\ \vdots & \vdots & \vdots \\ x_{147} & \cdots & x_{147-L+1} \end{bmatrix} \in \mathbb{C}^{(34-2L)\times L} \tag{18}$$

and $\mathbf{T}$ is a diagonal matrix defined as

$$\mathbf{T} = \text{diag}\{L-1, \cdots, 14, 15, 132+L-1, \cdots, 146, 147\}. \tag{19}$$

With (15), the cost function for LS criterion can be defined as follows

$$J_{LS} = (\mathbf{y} - \Phi\mathbf{u})^H(\mathbf{y} - \Phi\mathbf{u}), \tag{20}$$

where $\Phi = \begin{bmatrix} \mathbf{A} & \mathbf{T}\cdot\mathbf{A} \end{bmatrix}$, and $\mathbf{u} = \begin{bmatrix} \mathbf{u}_0^T & \mathbf{u}_1^T \end{bmatrix}^T$. The $\hat{\mathbf{u}}$ that minimizes J_{LS} can be obtained from the equation $\frac{\partial J_{LS}}{\partial \mathbf{u}^H} = 0$, and the solution is

$$\hat{\mathbf{u}} = \begin{bmatrix} \mathbf{A}^H\mathbf{A} & \mathbf{A}^H\mathbf{T}\mathbf{A} \\ \mathbf{A}^H\mathbf{T}\mathbf{A} & \mathbf{A}^H\mathbf{T}^2\mathbf{A} \end{bmatrix}^{-1} \begin{bmatrix} \mathbf{A}^H\mathbf{y} \\ \mathbf{A}^H\mathbf{T}\mathbf{y} \end{bmatrix}. \tag{21}$$

With the estimation of the parameters $\mathbf{u}_0$ and $\mathbf{u}_1$, the CIR of the entire slot can be easily obtained from (12). The CIR information are then used in the equalizer to recover the original transmitted data.

III. Channel Equalization

For EDGE system with 8PSK constellation, it is improper to use MLSE with VA as the equalizer due to its prohibitively computational complexity. It is shown in [8] that DDFSE and RSSE algorithms are promising equalization techniques for EDGE system. However, these algorithms can only be applied to systems with minimum phase CIR, otherwise, a dramatic performance degradation will occur. For system with time-invariant CIR, prefilter can be used to obtain an equivalent CIR with minimum phase [8]-[10], but the prefilter approach is not applicable to systems with time-varying CIR without resorting to complex channel tracking. In this section, a Cholesky decomposition-based method is introduced to obtain the equivalent minimum phase CIR for time-varying fading channels. For the purpose of simplicity, here we combine the first and second data block of one slot together, *i.e*, $\mathbf{y}_d = [y_{16}, y_{17}, \cdots, y_{131}] \in \mathbb{C}^{N_d\times 1}$, where $N_d = 116$ is the length of the entire data block.

Based on the estimated time-varying CIR $\hat{h}_k(l)$ and (1), the input output relationship of the data block can be written into matrix format (22) or compact form as

$$\mathbf{y}_d = \mathbf{H}_d \cdot \mathbf{x}_d + \mathbf{n}_d. \tag{23}$$

where $\mathbf{x}_d = [x_{16-L+1}, x_{16-L+2}, \cdots, x_{131}]^T \in \mathbb{C}^{(N_d+L-1)\times 1}$ and the additive white Gaussian noise column vector $\mathbf{n}_d \in \mathbb{C}^{N_d\times 1}$.

For typical channel profiles, such as the Typical Urban (TU) profile and Hilly Terrain (HT) profile [2], the impulse responses of the frequency-selective fading channels are usually not in their minimum phase state, *i.e.*, the power of $h_k(1)$ and $h_k(2)$ is larger than that of $h_k(0)$. Hence the CIR matrix $\mathbf{H}_d$ formed with this CIR is not diagonally dominant, which may cause serious numerical instability problems when the DDFSE or RSSE equalizers are used.

Our objective is to find an equivalent system with minimum phase CIR, whose input-output relationship of the equivalent system can be represented as

$$\mathbf{W}\mathbf{y}_d = \mathbf{B}\mathbf{x}_d + \mathbf{e}_d, \tag{24}$$

where $\mathbf{B}$ and $\mathbf{e}_d$ are the CIR matrix and the noise vector of the equivalent system, respectively [11]. In order to effectively equalize the time-varying frequency-selective fading channel using reduced state equalizers, the CIR matrix $\mathbf{B}$ should satisfy the following two conditions: (a) $\mathbf{B}$ should minimize the variance of the noise component of the system; (b) $\mathbf{B}$ should be an upper triangular matrix with most of the time-varying CIR energy concentrated

$$
\begin{bmatrix} y_{16} \\ y_{17} \\ \vdots \\ y_{130} \\ y_{131} \end{bmatrix} = \begin{bmatrix} \hat{h}_{16}(L-1) & \cdots & \hat{h}_{16}(1) & \hat{h}_{16}(0) & 0 & \cdots & \cdots & 0 \\ 0 & \hat{h}_{17}(L-1) & \cdots & \hat{h}_{17}(1) & \hat{h}_{17}(0) & 0 & \cdots & \cdots \\ \vdots & 0 & \ddots & \ddots & \ddots & \ddots & 0 & \cdots \\ 0 & \cdots & 0 & \hat{h}_{130}(L-1) & \cdots & \hat{h}_{130}(1) & \hat{h}_{130}(0) & 0 \\ 0 & \cdots & \cdots & 0 & \hat{h}_{131}(L-1) & \cdots & \hat{h}_{131}(1) & \hat{h}_{131}(0) \end{bmatrix} \begin{bmatrix} x_{16-L+1} \\ x_{16-L+2} \\ \vdots \\ x_{130} \\ x_{131} \end{bmatrix} + \begin{bmatrix} n_{16} \\ n_{17} \\ \vdots \\ n_{130} \\ n_{131} \end{bmatrix}, \quad (22)
$$

in the first few taps. The first condition will improve the performance of the equalizer, and the second condition can guarantee a system with minimum phase CIR.

The variance of the noise component of the equivalent system is

$$\sigma_e^2 = \frac{1}{N_d} \text{trace}\left[E(\mathbf{e}_d \mathbf{e}_d^h)\right]. \quad (25)$$

where $trace(\cdot)$ will return the sum of the diagonal elements of a matrix. According the orthogonality principle [12, pp.256-258], the matrix $\mathbf{B}$ minimizing σ_e^2 must satisfy

$$\mathbf{B} \cdot \mathbf{R}_{xy} = \mathbf{W} \cdot \mathbf{R}_{yy}, \quad (26)$$

where $\mathbf{R}_{xy} = E(\mathbf{x}_d \cdot \mathbf{y}_d^h)$, and $\mathbf{R}_{yy} = E(\mathbf{y}_d \cdot \mathbf{y}_d^h)$. Substitute (26) into (25), the minimum value of σ_e^2 can be obtained as

$$\min(\sigma_e^2) = \frac{1}{N_d} \text{trace}\left[\mathbf{B}(\mathbf{R}_{xx} - \mathbf{R}_{xy}\mathbf{R}_{yy}^{-1}\mathbf{R}_{yx})\mathbf{B}^h\right], \quad (27)$$

where $\mathbf{R}_{xx} = E(\mathbf{x}_d \cdot \mathbf{x}_d^h)$. Combining (27) with (23), we can obtain

$$\min(\sigma_e^2) = \frac{1}{N_d} \text{trace}\left[\mathbf{B}(\mathbf{R}_{xx}^{-1} + \frac{1}{\sigma_n^2}\mathbf{H}_d^h \mathbf{H}_d)^{-1}\mathbf{B}^h\right], \quad (28)$$

$$= \sigma_n^2 \cdot \text{trace}\left[\mathbf{B}(\frac{1}{\text{SNR}}\mathbf{I}_{N_d+L-1} + \mathbf{H}_d^h \mathbf{H}_d)^{-1}\mathbf{B}^h\right], \quad (29)$$

where (29) is based on the assumption that the input data symbols x_k are independent, *i.e.*, $\mathbf{R}_{xx} = E_s \mathbf{I}_{N_d+L-1}$ with E_s being the symbol energy, σ_n^2 is the variance of the AWGN n_k, and $\text{SNR} = E_s/\sigma_n^2$.

In order to guarantee that the CIR matrix $\mathbf{B}$ is in minimum phase form, we perform the Cholesky decomposition on the matrix $\mathbf{R} = \frac{1}{\text{SNR}}\mathbf{I}_{N_d+L-1} + \mathbf{H}_d^h \mathbf{H}_d$

$$\begin{aligned} \mathbf{R} &= \frac{1}{\text{SNR}}\mathbf{I}_{N_d+L-1} + \mathbf{H}_d^h \mathbf{H}_d, \\ &= \mathbf{U}^h \mathbf{U}, \end{aligned} \quad (30)$$

where $\mathbf{U} \in \mathbb{C}^{(N_d+L-1)\times(N_d+L-1)}$ is an upper triangular matrix. Combining (30) with (29), we have

$$\min(\sigma_e^2) = \sigma_n^2 \cdot \text{trace}\left[\mathbf{B}\mathbf{U}^{-1}(\mathbf{U}^\mathbf{h})^{-1}\mathbf{B}^h\right]. \quad (31)$$

Therefore, we find the equivalent CIR matrix $\mathbf{B}$ as

$$\mathbf{B} = \mathbf{U}, \quad (32)$$

and $\mathbf{W}$ can be easily derived from (24) as

$$\mathbf{W} = (\mathbf{U}^h)^{-1}\mathbf{H}_d^h. \quad (33)$$

From above equations, the input-output relationship of the equivalent system can be written as

$$(\mathbf{U}^h)^{-1}\mathbf{H}_d^h \mathbf{y}_d = \mathbf{U}\mathbf{x}_d + \mathbf{e}_d, \quad (34)$$

where the matrix $\mathbf{U}$ is in its minimum phase state, that is, the matrix $\mathbf{U}$ has the time-varying CIR energy concentrated in the first few taps. The obtained equivalent minimum phase CIR matrix can then be used in the DDFSE or RSSE equalizer to recover the original transmitted symbols.

IV. Simulation Results

In this section, simulations are carried out to evaluate the performance of the proposed channel estimation and equalization algorithms for EDGE systems with time-varying and frequency-selective fading channels, in terms of both estimation mean square error (MSE) and bit error rate (BER). The performance is evaluated under the Typical Urban and Hilly Terrain channel profiles as defined in [2], and the simulation system is oversampled 37 times to obtain a time resolution of $T_{sample} = T_{sym}/37 \approx 0.1\mu s$, which is the minimum differential delay of the multipath branches of the channel.

The MSE of the channel estimation algorithms under different maximum Doppler frequencies is shown in Figure 3 for TU and HT profiles. It can be seen from this figure that the maximum Doppler frequency has very little influence on the MSE of the proposed estimation algorithm, while the MSE of the LS algorithm [8], [9] degrades dramatically with the increase of f_d. From the figure, we can conclude that the proposed algorithm can obtain a rather accurate estimation of the time-varying fading channel for a wide range of Doppler frequencies.

In Fig. 4, the BER performances of various channel estimation algorithms at different Doppler frequencies by employing MLSE equalizer are presented. The performance obtained from perfect channel estimation is shown as a lower bound reference. When the Doppler frequency is low, i.e. $f_d = 10$Hz, the LS and proposed algorithms have nearly the same BER performance which is 0.0038 at $E_b/N_o = 20$ dB. With the increase of Doppler frequency to 200Hz, the BER performance based on LS algorithm degrades dramatically to 0.039 at the same E_b/N_o. However, there is only a minor loss of performance for our proposed algorithm. In Fig. 5, the simulation is carried out using a DDFSE equalizer (8 state) under the minimum phase CIR. With the channel CIR in its minimum phase form, there is a 1 dB degradation compared to the MLSE equalizer, which is the trade off between the implementation complexity and system performance[6],[8]. Hence, the BER performance in Fig. 5 has shown that the Cholesky based algorithm is a promising method to obtain the time-variant minimum phase CIR needed for reduced state detection where feedback of detected symbols are deployed.

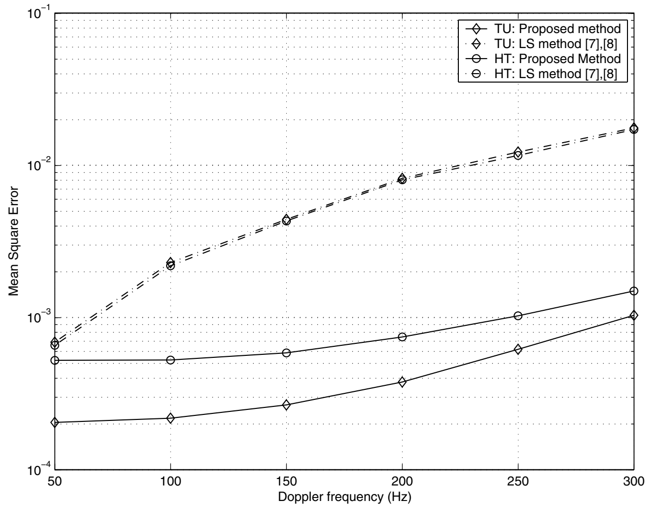

Fig. 3. Mean-Square-Error at frequency range of 50-300Hz in TU and HT profiles

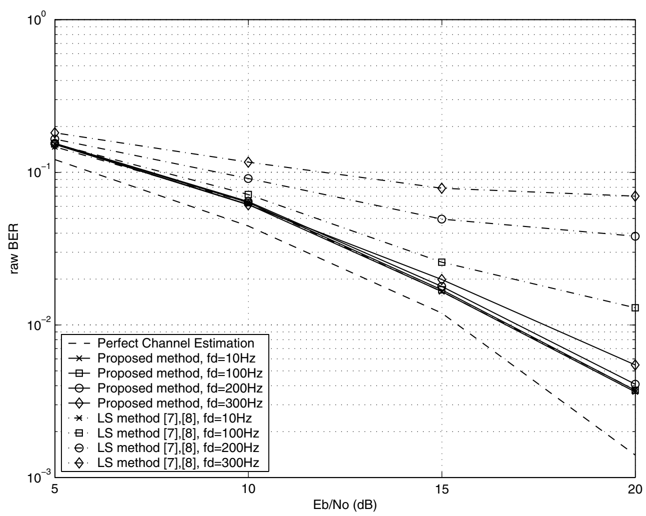

Fig. 4. BER of LS and linear LS channel estimation employing MLSE equalizer at f_d=10, 100, 200 and 300 Hz in TU profile

V. CONCLUSION

In this paper, a linear least-squares algorithm was presented to estimate time-varying and frequency-selective fading channels of EDGE system. The proposed algorithm can accurately estimate various fading channels which have wide range of Doppler frequency (up to 300 Hz). In terms of mean square error and bit error rate, it was shown via simulations that the proposed algorithm has much better performance than the least-squares algorithm, especially for Doppler frequency higher than 50 Hz. A reliable equalizer which employs the estimated time-varying channel impulse response is also discussed briefly.

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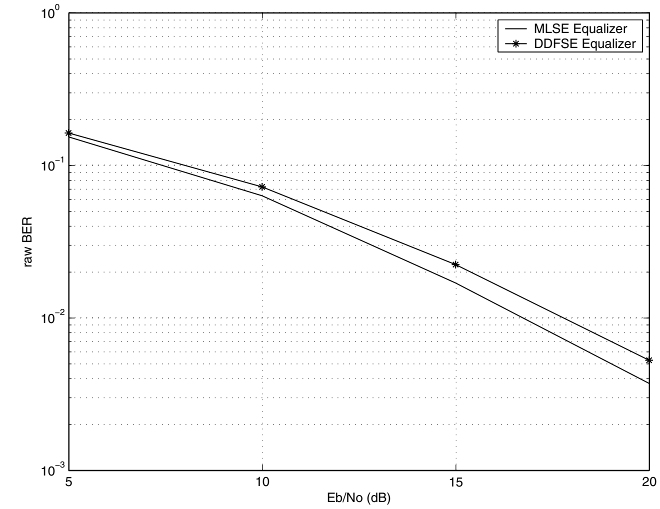

Fig. 5. BER of proposed channel estimation under MLSE and DDFSE equalizer at Doppler frequency 100Hz in TU profile

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